EMC Aspects of PWM Controlled Loads in Vehicles

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Licentiate Thesis

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Abstract

The number of electrically driven loads in a modern vehicle is constantly increasing. Many loads that former were mechanically driven will in the future be driven by electricity. This implies that a number of electronic systems have to be packed together in the limited space in a vehicle. When different electronic systems are placed close to each other, there is always a risk for electromagnetic interference between the different systems causing malfunction or even failure. It is important to ensure that this does not happen, and this concept is called electromagnetic compatibility, EMC. EMC implies that different electrical systems should be able to work in close proximity without affecting each other. From the EMC point of view, integration of electric traction drives in present vehicles represents a considerable challenge.

In order to save energy, many electrical loads can be controlled on demand. A common and energy efficient way to do this is to use a method called pulse width modulation, PWM, where the load voltage is pulsed in order to create the desired average output voltage. When this method is employed, the voltage pulses are present on the conductors between the power electronic converter and the load. Since the space in a vehicle is limited, it is often not possible to place the power electronic converter close to the load. Consequently, long conductors are often required between the power electronic converter and the load. The steep edges of the voltage pulses and the fundamental of the square wave, called the switching frequency, together with the long conductors cause electromagnetic interference problems. These disturbances could interfere with, for example, the radio in the vehicle. In this thesis, different electromagnetic compatibility aspects of a pulse width modulated system are investigated.

Some solutions are proposed in order to mitigate the disturbances. The solutions involve increasing the rise and fall times of the voltage pulses and employing a randomly varying switching frequency. Also the effects from different conductor layouts, such as using the vehicle body sheet metal as a current return path or having the lead-in and return conductor close to each other, are investigated. In order to evaluate the results from the different set-ups, the voltage across the load and the radiated magnetic field are measured.

The experimental results in this thesis show that a conductor should be used for current return and that this conductor should be placed as close to the lead-in conductor as possible in order to suppress electromagnetic noise. It is also shown that a randomly varying switching frequency will give a more broadband noise in the switching frequency range. Increasing the resistance of the gate resistor mitigates the disturbance in the higher frequency areas at the expense of increased switching losses.

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Although I spend a large share of the time that I am awake in Lund, I would not have been able to carry through this project without the support from my family and friends. I am very grateful that you are always there for me and always are encouraging me to believe in myself. There is also one very special person who I have asked a lot of during the last months, someone who has been taking care of and supported me during this project. Tomas, I don't know how to thank you enough for all the love, understanding and patience that you have showed me.

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Contents

Chapter 1

Introduction

> *"It takes good 'juice' and lots of it to run a modern auto; not the kind that Uncle Sam has put a ban upon, [but] the electric 'juice'."*

This quote is from the chapter "The Elusive 'juice'" in the 1918 edition of Putnam's Automobile Handbook [10]. Although this was written a long time ago, it is, besides for the nostalgic reference, still true today.

1.1 Background

Over the last two decades, the incorporation of electronics into cars and other vehicles has increased with an exceptional rate and continues to do so. The electrical system in a relatively modern car constitutes at least 20% of the vehicle's total cost. This means that the electrical system of the car costs more than the motor and the transmission together. A modern car today has due to this trend more on board computing power than the Apollo spacecraft of the 1960's [33].

There are two key reasons for the rapid increase in vehicle's electronic content. *Firstly*, the price of crude oil rose dramatically in the 1970s, which also caused the cost of petrol to rise. This triggered governments and drivers to demand more fuel efficient cars. At the same time, limits were imposed on the level of exhaust gas pollutants in order to reduce the problems of city smog. To be able to meet these requirements it was necessary to use complex engine management systems to accurately control the spark timing and the air/fuel ratio. Electronic systems were of vital importance to perform this task, which was beyond the capabilities of a purely mechanical system. The availability of the microprocessor enabled the execution of complex calculations at high speed and at a much lower cost than was previously possible. *Secondly*, customers today are demanding features that only the electronics can provide. Some of these features are for safety purposes, such as traction control and antilock braking systems; some are for instrumentation such as engine speed indicators and trip computers; and some are for convenience such as intermittent wiper and cruise control. There is also a new category of cars, the hybrid cars, where the internal combustion engine is accompanied by one or more electrical motors. It is, of course, the semiconductors that are enabling the technology that makes it all possible (as well as making this thesis necessary).

1.2 Todays electrical system

Due to the rapid increase in electrical loads in cars the present 12 V system has become insufficient. The 12 V system has been standard since the mid 1950's in almost all cars. At that time, the automotive industry switched from 6 V to 12 V in order to cope with the demand of higher spark energy to achieve reliable ignition in the V8-engines with a high compression rate (8:1 instead of the earlier 6.5:1). In those days, there were not many components in the car that were affected by this change in voltage level. Since then, the electrical power requirement has been steadily increasing, which motivates the introduction of a higher system voltage.

There are some disadvantages with the present 12 V systems, such as

- There is a need for point-to-point wiring, which makes the wiring harness heavy and complex.
- The system voltage varies between 9 V and 16 V and destructive high voltage transients can occur.
- Loads are designed to operate at the lowest foreseeable voltage. Therefore, at higher voltages non-controlled resistive loads draw more current than needed, which leads to overrated load components.
- The system is designed to withstand a load-dump transient; i.e. a voltage spike caused due to sudden load loss of a fully loaded alternator.
- There are high losses in the brushes of 12 V motors.
- The failure rate of connectors is high.

One of the constraints, enumerated above, imposed by the conventional system design is that all the loads in the electrical system of a car must survive a transient event known as a load dump. A load dump occurs when the battery is disconnected during charging. This causes a sudden drop in the alternator current and a rise in the alternator voltage. The increase in voltage occurs because the decreased alternator current is accompanied by a corresponding decrease in the voltage across the series armature reactance. The time constant of the regulator is typically a fraction of a second, which leads to a delay in the order of 100 ms before the alternator voltage returns to normal. The maximum of this voltage is, for system design purposes, generally assumed to be between 40 V and 80 V, which mean that electronic systems in the car must be design to survive this transient. For practical purposes, semiconductors that are directly connected to the electrical system must be rated for voltages between 50 V and 100 V – even though a load dump may never occur during the cars lifetime. This overrating is very costly, especially for power MOSFETs that often are used instead of relays.

The conventional Lundell alternator that supplies the 12 V systems has also some drawbacks. This component has actually the poorest characteristics of all components in the whole electrical system [28]. The efficiency of the Lundell alternator is in the vicinity of 50% and it has a large armature reactance. This low efficiency does not only mean that a considerably amount of energy is wasted, but also that the alternator has to be designed to withstand the thermal stress of dissipating something in the order of 1.5 kW. A poor efficiency in the alternator will of course affect the performance of the whole system. This is demonstrated when the cost for the generated electricity is calculated. The heat energy content of gasoline is 43500 Ws/g, and the density is 0.73 kg/l. This gives a specific energy of $3.3 \cdot 10^4$ kWs/l or 9.3 kWh/l. If the efficiency of the engine is assumed to be 20% and the efficiency of the alternator is assumed to be 45%, then:

$$9.3 \text{ kWh/l (fuel)} = 9.3 \cdot 0.2 \cdot 0.45 = 0.84 \text{ kWh/l (electric)}$$

Assuming that the price of the gasoline is 10 SKr/l the fuel cost for generating electricity in a car is approximately 11.20 SKr/kWh. This is a high cost and any improvement of the electrical system's performance is worthwhile. It has been shown, when a standard controlled profile has been used to measure the fuel economy, that an improvement of the electrical system efficiency by 100 W corresponds to removing 50 kg from the car, with respect to the fuel economy [28].

1.3 Higher voltage systems

Increasing the voltage of the system is necessary in order to cope with the increased number of loads. This will lead to several advantages as for example reduced weight and volume of the wiring harness. The optimum choice of voltage is a complex trade-off between cost, safety, economy and performance. For example, some electronic equipment such as lamps operate better at low voltage.

Mercedes Benz, now Daimler Chrysler, started the ball rolling in the mid 1980's with the growing realisation that 12 V would in the future be inadequate to meet the increasing electrical demands of its high-end cars. In 1994, Mercedes Benz approached MIT in USA in order to cooperate in the investigation of the advantages and disadvantages of a change to a higher voltage. They eventually decided that the most suitable voltage level is 42 V since this was the highest possible multiple of 14 V meeting the internationally agreed definition of a safety extra-low voltage – 60 V DC maximum (including transient overvoltages) [36].

Some of the benefits that should be gained when changing to a higher system voltage were e.g.

- There would be less load on the engine since many mechanical loads driven by chains and belts connected to the motor would be replaced by loads driven and controlled "on demand" by an electrical motor.
- The combustion engine could be started in a new way. By using a DC/AC converter as well as an AC/DC rectifier the generator could also function as a starting motor.
- The wiring harness could be made lighter. A three-fold increase in the voltage means that the wiring resistance can be increased by a factor of nine while retaining the same power loss over a given length of wire. Since the resistance increases as the wire diameter is reduced, the weight savings could come to as much as 40%. Also a high degree of networking in the electrical system could decrease the weight significantly.
- The cost from the semiconductor switches could be lowered. A higher voltage implies lower current, leading to a reduction in the required

silicon area. The silicon area could in principle be reduced by a factor nine when the voltage is increased three-folded.

- New solutions for reversed polarity and load dump could be implemented. Since new standards are drafted for 42 V, a central load dump protection at the generator is included, together with a central reverse polarity protection. Eliminating the need for protection at the component level should thus lead to cheaper components.

1.4 Challenges

Although the change to a higher voltage level would gain a lot of benefits, volume production is still some way off. All of the manufacturers are unwillingly to take the first step, since this would mean to shoulder a substantial share of the initial development costs. There are some issues with a higher voltage system that still has to be solved before the volume production can start.

- Co-existence;
    - The biggest barrier to introducing the 42 V system is the shortage of 42 V components. In the case with small motors and actuators for example, a switch from 12 to 42 V will necessitate a three-fold increase in the number of copper winding turns.
    - Light bulbs will also be a source of difficulties. Compared to an ordinary 12 V light bulb, the filament in a light bulb for 42 V has to be either three times longer or the square root of three times thinner. Neither of these solutions is welcomed since shocks and vibrations will limit the life expectancy of these light bulbs. For comparison: 24 V bulbs have a 30% shorter life than 12 V bulbs.
    - At least under a transitional period there will be 12 V components in a 42 V car. Different conversion strategies have been proposed to deal with the power supply for these components. A central conversion with one or two DC/DC converters/generators is one solution. The problem is that this method would require a dual-voltage supply network throughout the whole vehicle. Local conversion is another solution where the conversion function is implemented in electronic control units.

This solution is suitable for vehicles with a highly modular structure. Another approach is to do the conversion in the final driving stages. Semiconductor switches could be used to supply a low-frequency pulse width modulated input to the loads. Finally, the conversion could be made in the component, which would lead to 12 V "islands" within the car.

- Fusing;
  - It is difficult to define the appropriate characteristics of the fuse wire in a multiple-voltage car since it is hard to predict the impedance of a short-circuit path since this path also could be between different voltage systems. Safety in a multiple-voltage car can only be ensured if a failure mode and effects analysis are conducted into all possible short circuits, and acceptable measures are taken to insure the safety of the electrical system even in the event of a crash.
- Arcing;
  - The mechanisms that sustain an arc are much stronger at 42 V than at 12 V. This will have implications on relays and contacts that they can be separated under load.

The first implementations of 42 V are likely to be in high-end cars, with all their gewgaws, and in smaller fuel-efficient cars, where efficient power distribution coupled with soft-hybrid functions demand a higher voltage rail.

However, when looking further ahead, toward a not too distant future, it is already suggested that 42 V is inadequate for the demands of hybrid cars, where power in the order of 20-50 kW should be transferred through the electrical network. In for example a fuel cell car, the voltage for the traction motor is often about 300-400 V.

1.5 Problem definition

The number of electrically driven loads in a vehicle is increasing, which means that many different systems should be squeezed into an already crowded space. It is usually not possible to keep the power electronic converter, which often is a central unit for different systems, close to the load. The distance between the power electronic converter and the load requires long conductors. In order to keep the cost, weight and assembly work at a minimum; the conductors that are used are often unshielded. It is also common to use the body sheet metal of the vehicle as a current return path, and thereby saving the cost and weight of a current return conductor.

Many of the electrical loads benefits from being controlled on demand, thereby saving energy and in the end also fuel. A common and energy efficient way to control an electrical load is to use a method called pulse width modulation where the battery voltage is pulsed in order to create the desired average output voltage. Pulse width modulation enables control of the output voltage irrespective if it is AC or DC, the output current or the torque.

When the method with pulse width modulation is employed, the voltage pulses of the output voltage are distributed on the long cables between the power electronic converter and the load. The edges of these voltage pulses are steep with high derivatives and these pulses could be coupled into other systems in the vehicle, thereby disturbing them. Also the fundamental of the output voltage, which is equal to the switching frequency, could disturb other systems in the vehicle. One example of a system that is often affected by pulse width modulation is the AM-band of the radio. The disturbance is then noticed as a noise in the radio.

There are different ways to deal with the disturbances arising from a pulse width modulated system. One solution is to integrate the power electronics with the load, thereby avoiding the long conductors in between. Another way is to use filters or snubbers. Many of these solutions are not feasible for the automotive industry since they require extra components or more space. The aim of this thesis has been to find out how the disturbances from a pulse width modulated system could be reduced without incorporating any extra components.

1.6 Outline of the thesis

The first chapter contains a short historic survey and a vision of the future. In order to solve a problem, it is essential to have some knowledge about the background. Therefore, a brief overview of the area of electromagnetic compatibility is given in Chapter 2. Different coupling methods, noise suppression techniques, standards and measurement methods are described together with CE marking of electrical products.

In Chapter 3, an introduction to electromagnetic compatibility in vehicles is given for those who have no prior knowledge in this area. Here are the different coupling paths in a vehicle, and also some parts of the electrical system described. Vehicles and automotive components have their own type of CE marking, called e marking. The differences between these two types of marking are discussed and it is also described how a vehicle or a product is tested for e marking.

Chapter 4 explains the theory behind pulse width modulation and the semiconductors required for creating the voltage pulses.

Since the background has been covered in the first four chapters, Chapter 5 focuses on the electromagnetic compatibility problems in a vehicle and some solutions to these problems are proposed. In order to test these solutions a test set-up is built, which is described in Chapter 6.

This test set-up is used in order to measure the radiated magnetic field and the spectrum of the output voltage from the converter. The results from these measurements are depicted and described in Chapter 7. Finally, in Chapter 8, the work of this thesis is concluded together with some ideas of future work.

Chapter 2

Electromagnetic Compatibility

To have a number of electrical systems working close to each other without interfering is not an easy task. This chapter starts with a description of what electromagnetic compatibility means. After that different coupling methods for the noise and the characteristics for common mode and differential mode disturbances are described. In section 2.3 grounding of a system is discussed, and this is followed by section 2.4, which gives a short overview of common noise reduction techniques. The next section treats the international harmonization and CE marking. The last part of this chapter describes EMC testing and the equipment used.

2.1 Electromagnetic compatibility

The use of electronic circuits for communication, automation and other purposes is increasing, which necessitates for different circuits to be able to operate in close proximity. This implies that electromagnetic interference (EMI) may become a major problem in circuit design. The problem with EMI gets worse as the circuitry becomes smaller, meaning that more circuits are being crowded into less space, which increases the probability of interference.

An electronic equipment designer must assure that a system functions correctly when it has other electronic equipment nearby and not only under the ideal conditions in the laboratory. It is important that external noise sources do not affect the system and that it not itself is a source of noise to the environment. This is called electromagnetic compatibility (EMC) and is becoming a very important aspect to be considered in the early part of the product development.

IEEE has defined the concept EMC in the following way [14]:

"Electromagnetic Compatibility, EMC, is the ability of a device, equipment or system to function satisfactorily in its electromagnetic environment without introducing intolerable electromagnetic disturbances to anything in that environment."

In order to define an electromagnetic noise problem, three items are necessary, i.e. a noise source, a coupling channel and a susceptible receptor. A component acting as a noise source at one time could very well be a receptor the next time. This implies that electromagnetic compatibility has two aspects, i.e. emission and susceptibility.

There are a number of different ways for the emissions to be coupled into other systems and there are many techniques by which noise can be reduced in an electronic/electromagnetic system. This implies that there does not exist a unique solution to most noise reduction problems.

2.2 Coupling channels for the disturbances

In this section some of the different coupling channels for the disturbances are described. This is followed by a discussion about common mode and differential mode disturbances and their characteristics.

Capacitive coupling

Capacitive coupling, also known as electric coupling, is a result from the interaction of electric fields between circuits. If two conductors are placed next to each other, there will be a stray capacitance between them. Figure 2.1 shows an example of this phenomenon and an equivalent circuit.

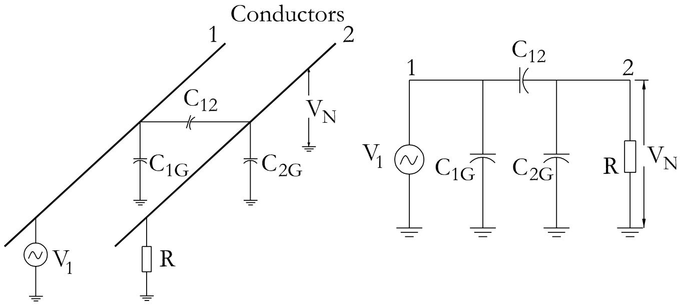

Figure 2.1. Capacitive coupling between two conductors [41].

Capacitance C_{12} in Figure 2.1 is the stray capacitance between the two conductors. Capacitor C_{1G} and C_{2G} are the total capacitance between each conductor and ground. The resistance, R, is the resistance of conductor 2 and ground, and is the result of the circuitry connected to this conductor. In the figure is the voltage V_1 the source of interference, and circuit 2 the receptor. The noise voltage, V_N, picked up by conductor 2 could be expressed as

$$V_N = \frac{j\omega\left[\dfrac{C_{12}}{C_{12}+C_{2G}}\right]}{j\omega + \dfrac{1}{R(C_{12}+C_{2G})}} \cdot V_1 \tag{2.1}$$

In most practical cases the resistance R represents a much lower impedance than the impedance for the stray capacitance C_{12} plus C_{2G}. Equation (2.1) could therefore be reduced to the following

$$V_N = j\omega RC_{12}V_1 = j2\pi fRC_{12}V_1 \tag{2.2}$$

These equations show clearly how the noise voltage on conductor 2 depends on the various parameters. It can be seen in equation (2.2) that the noise voltage is directly proportional to the frequency of the noise source, the resistance R of the affected circuit to ground, the stray capacitance between the circuits and the magnitude of the source voltage. In many cases it is not possible to change the voltage or the frequency of the noise source, which leaves only two parameters to reduce the capacitive coupling. In order to decrease the resistance, R, it is necessary for the receiving circuit to work at a lower resistance level. It is in many cases easier to decrease the value of the stray capacitance since this is affected by the orientation of the conductors and shielding. If the conductors are moved farther apart, the stray capacitance will decrease. However, only little additional attenuation will be gained by spacing the conductors a distance greater than 40 times their diameter, according to

$$C_{12} = \frac{\pi\varepsilon}{\cosh^{-1}\left(\dfrac{D}{d}\right)} \tag{2.3}$$

where D is the distance between the cables and d the diameter of the cable.

The other way to affect the stray capacitance is with electric field shielding of the conductor. When the conductor is shielded, the length of the conductor that extends beyond the shield determines the stray capacitance, and it is therefore important to keep this length short. Also a good ground connection of the shield is necessary in order to provide a good electric field shield.

Inductive coupling

Inductive coupling, also known as magnetic coupling, is a result from the interaction between the magnetic fields of two circuits. A magnetic flux is created as a current flows in a closed circuit. The flux is given by the multiplication of the current in the circuit times the inductance of the circuit. The value of the inductance is depending on the geometry of the circuit and the magnetic properties of the medium that contains the field. A current flow in one circuit may produce a magnetic field in another circuit, and the connection between these two circuits is called the mutual inductance, M.

Due to a magnetic field with the flux density $\overline{B}$, the noise voltage, V_N, induced in a loop of area $\overline{A}$ can be derived from Faraday's law as in equation (2.4) [41]

$$V_N = -\frac{d}{dt}\int_A \overline{B} \cdot d\overline{A} \tag{2.4}$$

where $\overline{B}$ and $\overline{A}$ are vectors. For a stationary closed loop with a sinusoidally varying flux density that is constant over the loop area the expression for the noise voltage reduces to

$$V_N = j\omega BA\cos\theta \tag{2.5}$$

where θ is the angle between the area and the magnetic flux density. Since $BA\cos\theta$ represents the total flux coupled to the receptor circuit, which means that equation (2.5) can be rewritten as

$$V_N = j\omega MI_1 = M\frac{di_1}{dt} \tag{2.6}$$

where I_1 is the current in the source circuit and M the mutual inductance between the receptor and source circuit. Corresponding physical representation and equivalent circuit for the magnetic coupling between two circuits are shown in Figure 2.2.

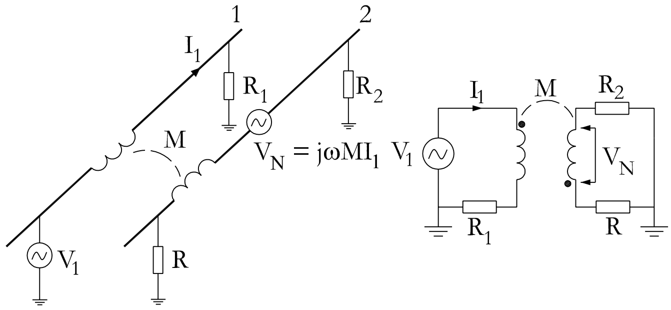

Figure 2.2. Inductive coupling between two conductors [41].

As can be seen in equation (2.5) and equation (2.6) the coupling between the two circuits is directly proportional to the frequency via ω ($2\pi f$). The three parameters that are left to modify in order to reduce the noise voltage are then the magnetic flux density, B, the loop area, A, and the angle between A and B. The magnetic flux density can be reduced by physically separating the circuits or by twisting the source wires. Twisting the wires causes a cancellation of the magnetic fields from the wires, but can also be used to decrease the area of the receiver circuit. If the return current of the receiving circuit instead is led through a ground plane, placing the conductor close to the ground plane can decrease the area. A proper orientation of the source and receiver circuits could reduce the $\cos\theta$ term.

A nonmagnetic shield that is placed around a conductor and grounded at one end has no effect on the magnetically induced voltage in that conductor since the shield does not have any effect on the geometry or magnetic properties between the two circuits. A conductor with a nonmagnetic shield that is grounded in both ends will have a current on the shield. This shield current will actually induce a second noise voltage into the conductor.

Electromagnetic coupling

Electromagnetic coupling or radiation is a combination of electric and magnetic fields, and is a term that is normally used in the far field. For a definition of the near and far field, see Appendix A.

One effective way to prevent radiation is to shield the source of interference. As can be seen in Figure 2.3, a shield that is grounded in one end will only affect the electric field around the conductor. In order to be able to affect both the electric and magnetic field around the conductor, a shield that carries a current equal and in the opposite direction of the conductor current is needed.

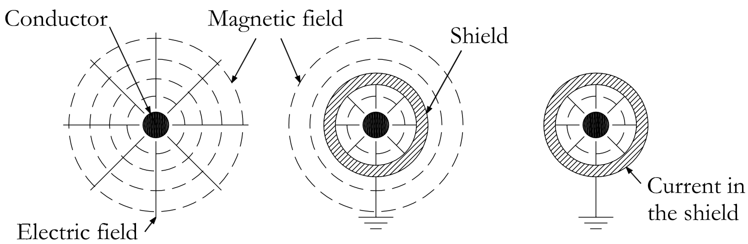

Figure 2.3. Field around a current carrying conductor. The leftmost figure describes a conductor with no shield, the figure in the middle describes a shielded conductor where the shield is grounded at only one point and the rightmost figure describes a shielded conductor where the shield is grounded and carries a current equal to the conductor current but in the opposite direction [41].

The magnetic field around two conductors that are placed in close proximity with equal current of opposite signs cancel each other according to Figure 2.4.

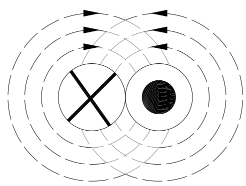

Figure 2.4. The magnetic fields around two conductors carrying the equal current in the opposite direction cancels each other when the conductors are placed close to each other.

Common mode and differential mode

Electromagnetic disturbances appear in the form of common mode (CM) or differential mode (DM) disturbances. The differential mode voltage component of a circuit is a voltage that can be measured between phase conductors. The differential mode current component flows in the supply wires, including the neutral wire. The common mode current component on the other hand flows from the phase and neutral conductor toward earth. The circuit for the common mode current is closed by stray capacitances between the earthed parts and the circuit. Definitions for the DM and CM components can be seen in Figure 2.5 and equation (2.7)[47]

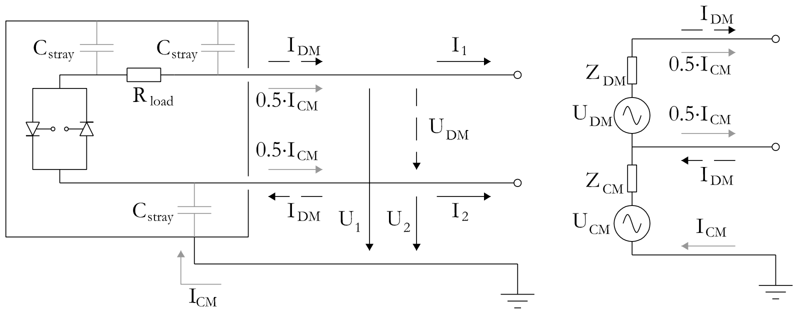

Figure 2.5. Differential mode (DM) and common mode (CM) electromagnetic interference voltage and current components. A typical electromagnetic interference source to the left, and the high frequency substitution circuit of the electromagnetic interference source to the right [47].

$$\begin{cases} U_{DM} = U_1 - U_2 \\ I_{DM} = \dfrac{I_1 - I_2}{2} \\ U_{CM} = \dfrac{U_1 + U_2}{2} \\ I_{CM} = I_1 + I_2 \end{cases} \tag{2.7}$$

The radiation from an electronic circuit could occur as either common mode or differential mode radiation.

Differential mode radiation

A current that flows in a loop formed by the conductors of the circuit causes differential mode radiation. These small current loops will then act as small antennas, radiating a magnetic field. The most effective antennas are loop antennas where the circumference is smaller than one quarter of a wavelength. The current in such a current loop is in phase everywhere, which adds to rather than subtracts from the overall emission of the circuit, which may happen when the loop area is larger. The radiation from a loop antenna is proportional to the current in the loop, the loop area and the square of the frequency. If the magnitude and the frequency or the harmonic content of the current is reduced, also the radiation will decrease. Minimizing the area enclosed by the current flow will also decrease the radiation, and this could easily be done by placing signal leads and their associated return leads close together.

Common mode radiation

The common mode radiation has a large influence on the emission performance of a product, and is harder to control than the differential mode radiation. When the differential mode radiation is controlled by a proper circuit layout, the control of common mode radiation demands that the common mode currents in all cables are minimized.

Common mode radiation originates from the cables in the system. The frequencies that are radiated differ from the differential mode signal frequencies in the cable and are instead determined by the common mode potential (usually the ground voltage). The common mode radiation is proportional to the frequency, the length of the antenna (the cable) and the magnitude of the common mode current on the antenna. For lower frequencies of a symmetrical square wave, the common mode emission spectrum is flat. The spectrum is flat up to a frequency, proportional to the rise time of the signal ($1/\pi \cdot t_{rise}$), where the spectrum starts to decrease. This implies that the common mode emissions are mostly a problem in the lower frequency areas, which is not all true since the capacitive coupling is increased for the higher frequency areas. If the magnitude of the common mode current that is needed to produce a certain radiated field is compared to the magnitude of the differential mode current needed to produce the same radiated field (in a case where the loop area, cable length and frequency is constant), there ratio between these two currents is large. This means that a common mode current of a few microamperes can cause the same amount of radiated emissions as a few milliamperes of differential mode current.

When the common mode radiation should be controlled, it is desirable to limit both the rise time and the frequency of the circuit just as in the case with differential mode radiation. Since the common mode radiation emanates from the cables of the system, it is important to keep these short. This is often not possible, but on the other hand, the emission cease to increase for cables longer than one quarter of a wavelength due to the presence of out of phase currents.

The common mode current is often the only parameter controllable and is therefore an important control parameter for the radiated emissions. The common mode current is not required for system operation, but it is important to ensure that although the components used for common mode suppression not affects the functional differential mode currents. One of the first things to be done in order to control the common mode currents is to minimize the source voltage driving the antenna. This is usually the ground potential, and one solution to this problem could then be to use for example a ground plane in order to reduce the voltage drop in the ground system. One positive side effect of introducing a ground plane is that it also decreases the differential mode radiation, due to the induced mirror current. Another way of controlling the common mode current is to put a large common mode impedance (a common mode choke) in series with the cable. A common mode choke is the only technique that does not require a ground to function and also the only technique that does not affect the differential mode current. Decoupling the cables (shunting the current to ground) or shielding are methods that also affects the common mode current. The problem with these methods is that they require a "quiet" or "clean" ground in order to work properly.

2.3 Grounding

A well-designed ground system can provide protection against unwanted interference and emission without any additional per unit cost to a product. A ground is normally defined as an equipotential, a point where the voltage does not change regardless of the current applied to it or drawn from it, a point or plane that serves as a reference potential for a circuit or system. This definition is often not representative of practical ground systems since they are not equipotentials. Another problem with this definition is that it does not emphasize the importance of the actual path taken by the current when it is returned to the source. It is central to know the actual current path in order to determine the radiated emission or the susceptibility of a circuit [41].

It is also important to keep in mind the frequency of the circuit when designing the ground system. At high frequencies and in digital circuitry, a multipoint ground system could be used in order to minimize the ground impedance. This type of ground system should be avoided when the frequency is low. In a low-frequency system it is important to have a single-point ground system, and there should also be a minimum of three separate ground returns, i.e. signal ground, noisy ground and hardware ground.

Something that sometimes could be a source of noise is a ground loop. A ground loop could for example be created when multiple ground points are separated by a large distance and are connected to the AC power ground, see Figure 2.6. In order to improve this situation, one of the grounds could be removed, thereby converting the system to a single-point ground system, or the effect of the multiple grounds can be eliminated or at least minimized by isolating the two circuits. Isolation could be achieved by i.e. transformers, common mode chokes, optical couplers or balanced circuitry.

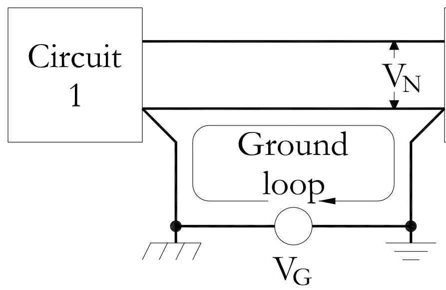

Figure 2.6. A ground loop between circuit 1 and circuit 2 [41].

When talking about different grounds it is important to separate between the two categories that they are distinguished into; safety grounds that are usually at earth potential and signal grounds that may or may not be at earth potential.

Ground plane

A ground plane is a large conducting area, which serves as a circuit common or power return. The ground plane is often placed on a printed circuit board where it stiffens the board and also provides thermal heat spreading. A ground plane works as an electrostatic Faraday shield and reduces magnetic field and net parasitic capacitances through image currents, see Figure 2.7.

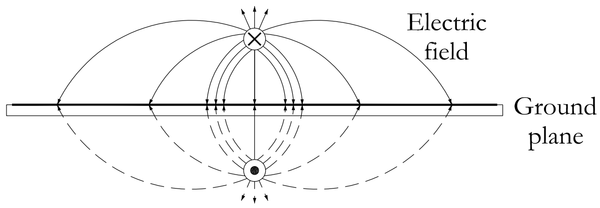

Figure 2.7. An image current below the ground plane created by the conductor above the ground plane.

The ground plane affects the electric and magnetic field so that they distribute themselves as if a mirror image conductor exists on the opposite side of the ground plane. This image conductor carries then an equal current in the opposite direction.

Ground plane slots under conductors significantly reduce the effectiveness of the ground plane. It is always better to have a thin solid ground plane than a thick plane with slots.

2.4 Common noise reduction techniques

There are some common noise reduction techniques that often could be used in order to deal with EMI problems. Some of the techniques are essentially free of added cost and should therefore be used whenever applicable, and the other techniques that will be mentioned in this section are techniques that should be used whenever additional noise reduction is required. This section gives only some examples of noise reduction techniques and has no intention to explain the underlying theory behind the techniques. Most of the techniques could, however, be explained by the information in Chapter 2.2.

In order to suppress the noise, noisy leads should always be twisted together since this will decrease the area of the loop, and thereby decrease the magnetic coupling. An even better solution would be if noisy leads besides being twisted also could be shielded. Limiting the pulse rise times would decrease the high frequency contents of the signal, thereby lowering both the capacitive and inductive coupling. It is also desired to have all noise sources enclosed in a shielded enclosure and that all leads leaving this enclosure should be filtered.

If the chassis is used as a ground return, the cables should be placed near the chassis in order to minimize the inductive coupling. All ground leads should be kept as short as possible and questionable or accidental ground should be avoided. The length of sensitive leads should be kept as short as possible since a long cable is a more efficient antenna. Leads extending beyond cable shields should be kept short to minimize the capacitive coupling. Since there is always a coupling between two leads that are placed in close proximity, noisy and quiet leads should be separated. It is important to separate signal, noisy and hardware grounds in order to minimize the noise.

EMC = black magic?

Many, maybe even most electrical designers feel insecure when the subject EMC is brought up. The solutions to EMI problems are in many cases considered to be some kind of black magic. This is not very strange since parasitic effects that play a fundamental role cause most of the EMI problems. A rule of thumb is that only one third of the components affecting the EMI are on the schematics of the circuit. Another third of the components consists of parasitic elements within components, and the final third is created by the trace routing on the printed circuit board and component mounting, placement and orientation. Another thing that contributes to the "black magic" theory is that the results from tests are strongly affected from small details such as differences in cable length, cable arrangement, apparatus arrangement and so on.

Although there are regulations that should be followed if a company wants to have their electrical product on the market, many companies are not working with control of EMC. As justification they mean that the conformity to the EMC regulations are not easily visible and that the damages that their product could cause or be exposed to is limited. They are also not very worried about getting caught since they mean that the probability of this is very low.

2.5 International harmonization

In 1934 the Comité International Spécial des Pertubations Radioélectrique (CISPR, International Special Committee on Radio Interference) was founded in Paris by representatives from countries that had become concerned with the problem of radio frequency interference. They early agreed that their primary job would be to document standard EMI measurement methods and to determine internationally acceptable noise level limits. CISPR later became the first organization that was authorized to promulgate international recommendations on EMC.

Their founding conference proposed to establish a common commission in the IEC (International Electrotechnical Commission) and UIR (Union International de Radiodiffusion) to facilitate the preparation of recommendations. This was before World War II and after World War II, UIR was not recovered again and CISPR became a special committee of IEC. CISPR differs from other study groups in the sense that several other international organizations participate in CISPR's work with observer status. The preliminary effort of CISPR was to publish a set of documents that would describe widely applicable requirements for EMI measurement equipment and techniques. This effort was completed in 1961.

In 1973, CISPR was reorganized and six subcommittees with different areas of interest were created. The subcommittees were the following [15]:

- Subcommittee A: Interference measuring devices, measurement methods (Publ. 16, 17)
- Subcommittee B: EMI from industrial, scientific and medical apparatus (Publ. 11, 19, 23, 28)
- Subcommittee D: Ignition interference from motor vehicles, combustion engines and related subjects (Publ. 12, 21, 25)
- Subcommittee F: EMI in domestic appliances, fluorescent tubes and similar devices (Publ. 14, 15, 30)
- Subcommittee H: Limits for the protection of radio services (Publ. 31)
- Subcommittee I: EMC of information technology equipment (ITE), multimedia equipment and receivers (Publ. 13, 20, 22, 24)

Since CISPR is mainly working with radio frequencies above 9 kHz, IEC has another committee that covers the other frequency areas. This committee is called TC 77 and has as a main task to prepare basic and generic EMC publications specifying electromagnetic environments, emissions, immunity, test procedures and measurement techniques. TC 77 is also divided into different subcommittees:

- Subcommittee 77A: Low frequency phenomena up to and including 9 kHz
- Subcommittee 77B: High-frequency continuous and transient phenomena, including electrostatic discharges
- Subcommittee 77C: High-power transient such as the electromagnetic fields produced by high-altitude nuclear detonations.

The standards and directives that are issued by i.e. CISPR and TC77 form the basis for the tests that are performed when a product should be tested for CE marking.

CE marking

In order to be able to sell an electrical product within the European Union, it is mandatory to use CE marking and a Declaration of Conformity [49]. The CE marking is the manufacturer's declaration, which shows that the product complies with all the applicable directives. Most products can be self-assessed by the manufacturer to meet the essential requirements. CE marking is therefore not an approval, certification or quality mark, but only a declaration of the supplier's own responsibility.

The main goals of the CE marking are to

- Indicate the products conformity with the "essential" requirements of the directives
- Allow products to be placed on the market
- Ensure the free movement of goods and permits withdrawal of non-conforming products

The CE marking is only a topic when it comes to products and not components since they are excluded from these rules.

When a product should be CE marked, the producer starts to identify the directives that are applicable to the product that should be tested [11]. The directive and the type of product then settle the conformity assessment procedure that must be taken. This can be self-declaration, which involves testing and inspection or a quality system assessment from a notified body or perhaps a combination of these. The next step is to identify if there are any harmonized European standards that are applicable to the product. These standards are not always mandatory for the manufacturer to follow, but there is a presumption that conformity of these standards will give conformity with a relevant part of the directive. In order to ensure that the product complies with all the essential requirements of the directive, appropriate measurements have to be made to meet the terms of or identify existing data and test reports. In order to fulfil the directives for the product also the technical documentation should support the compliance with the requirements of the directive. This technical documentation along with the declaration of conformity should be available to competent authorities (EU members) upon request when the product has passed the test and received the CE mark.

2.6 EMC testing

It is important to remember that there are many different needs for EMC testing during a product's lifecycle, and that every test has its own technical, cost and time requirements. But if EMC testing is done at all stages in the development of a product, this can save a great deal of time and money. An electronics company will probably need to deal with all four types of EMC tests; for diagnostics, pre-compliance, full compliance and production [50].

Diagnostic EMC test are carried out while a product is in its development phase. In these tests the EMC effects of different designs are inspected. An advantage of these tests is that it will give the designer a database of knowledge of the effects of alternative design choices. This knowledge may prove to be very useful later when the possible EMC effects of post-design changes need to be assessed.

The pre-compliance tests are carried out when there is a prototype available that represents the model, which will be subject to full compliance testing later. These tests are important in order to avoid unpleasant surprises, to define the worst-case configuration and operating mode, and to define the criteria for immunity test failure.

Tests for full-compliance can either be done in-house or by an external test house. It is important that the test set-up for full compliance is capable to perform the tests according to the required standards. The factors that are often crucial for whether the tests are done in-house or not are the available budget, skills and resources, the nature of the product that should be tested and the requirements of the applicable standards.

Production EMC testing means that the manufacturer should make sure that the manufacturing process ensures compliance of each product with the EMC directive. This will probably involve some form of sampling system or "golden" product comparison. A golden product is then a product sample that has been tested and approved according to the directives. These tests are rather simple and a test rig with instrumentation that will detect expected variations due to production tolerances may be all that is required.

There are mainly five different phenomena that are examined for each product that should be CE marked. The five things that are looked into are radiated emissions, conducted emissions, radiated immunity, conducted immunity and electrostatic discharge. All of these examinations are essential since the product should not only be able to work in close proximity to other products but also not disturb them. As mentioned before, some test may be performed in-house and Table 2.1 suggests which of the phenomena that are suitable for in-house testing. Tests marked with figure 3 are essential to make in-house, tests marked with 2 are likely to be done in-house, and finally are the tests marked with the figure 1 that may be possible to make in-house.

Table 2.1. Different tests for EMC

Test	Diagnostics	Precompliance	Full compliance	Production
Conducted emissions	2	3	1	2
Radiated emissions	1	2	1	1
ESD immunity	2	3	3	1
Transient immunity	2	3	3	2
Surge immunity	1	1	2	1
Conducted RF immunity	2	2	1	2
Radiated RF immunity	1	1	1	1

2.7 Equipment for EMC testing

Antennas

In development, diagnostics and quality assurance tests, it can be very useful to have a near field or close field probe (for definitions of near and far field, see Appendix A). With the help of a near field probe it is possible to determine where the main sources of radiation are. This is very helpful for diagnostics, but irrelevant when it comes to pre- or full compliance tests since a near field probe cannot be used to predict or measure the far field performance. There are near field probes for the magnetic field and the electric field.

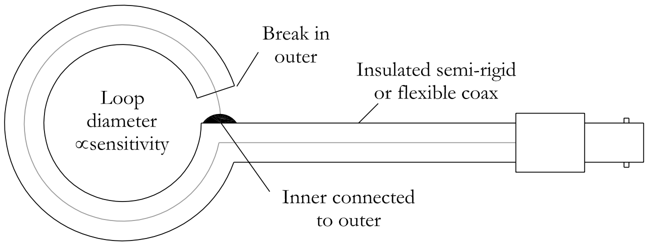

Figure 2.8. A loop antenna for magnetic field measurements.

In Figure 2.8 a loop-antenna for the radiated magnetic field is shown.

As described in equation(2.5), it is possible to describe the voltage produced by the loop in a magnetic field as

$$V_m = 2\pi fBA\cos\theta \tag{2.8}$$

The angle, θ, is measured between a right-angled line from the centre of the loop antenna and the magnetic field. If the loop antenna has no shield, it would not only pick up the magnetic field, but also the incident electric field. The shield eliminates this pickup, but it is important that the shield is broken somewhere in the loop in order to prevent the flow of shield current. Otherwise, the shield current would cancel the magnetic field as well as the electric field. The magnetic field probe is the most useful near field probe, and is often used to detect regions of high di/dt flow on printed circuit boards or to detect discontinuities in shielded enclosures.

Most near field probes for magnetic or electric field can also be used for immunity testing. Instead of connecting the probe to an EMI receiver, the probe could be connected to the output of a radio frequency signal generator. The probes will generate localised magnetic or electric fields.

In the EMC standards, only limits for the emissions in the far field are given. This implies that only far field antennas are used during full compliance testing [3]. Getting enough far away could be a problem in the case where frequencies lower than 30 MHz should be measured, since the distance between the equipment under test and the antenna has to be very large. This is sometimes solved by using an open area test site or by measuring the magnetic field and electric field components in the near field separately. In order to measure the magnetic field, a large loop probe is used, and a whip antenna that is typically 1 m long is used for the electric field.

There are many different antennas that are used when measuring in the far field. Dipoles could for example be used for the frequency areas from 30 MHz to 1 GHz, but many of these have limited frequency ranges, which make testing very time consuming since a number of antennas have to be used. Also biconical antennas are sometimes used, but they do also have limits in their frequency ranges.

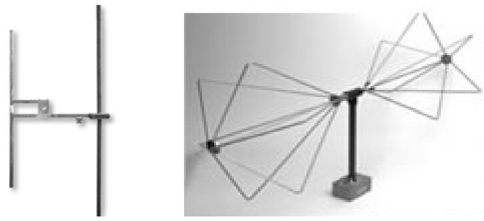

Figure 2.9. Left: A dipole antenna (www.sofratec.com) Right: A biconical antenna (www.com-power.com)

The standard antenna for the industry is the bilog antenna. A bilog antenna covers the frequency range from 30 MHz to 1 GHz, but there are also antennas available that measure from 20 MHz up to 2 GHz.

For measurements of frequencies higher than 1 GHz, dipoles, horns or double ridged waveforms are used.

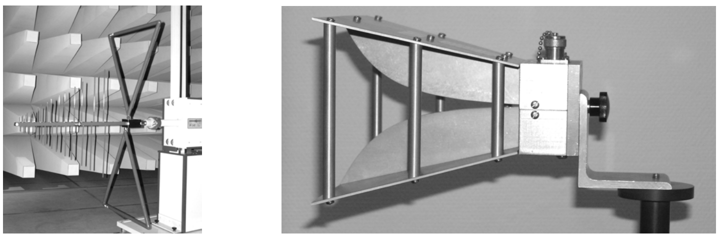

Figure 2.10. Left: A bilog antenna Right: A horn antenna.

Current probes

Since CM currents (see section Common mode and differential mode) cause most of the radiated emissions, a current probe is interesting to use in order to examine the presence of these currents. Using a current probe for measuring CM currents is also a good instrument for pre compliance testing since the results from the measurements give a very good conception about the radiated emissions.

A current probe often consists of a ferrite cylinder, through which the cable with the current that should be measured is passed. Since a CM current is present with equal signs in both send and return paths, the two cables should pass through the ferrite in the same direction, see Figure 2.11. In this way, the CM current is measured without any DM component since the DM components of the current are not measured when using the current probe as in Figure 2.11 since they cancel each other.

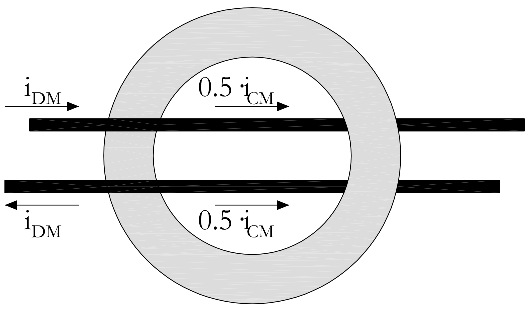

Figure 2.11. Measuring common mode currents using a current probe

Line Impedance Stabilization Network

A Line Impedance Stabilization Network (LISN), also called an Artificial Mains Network (AMN), is a transducer used to perform conducted emissions testing. It is required to provide defined impedance at radio frequencies to the terminals of the system that is being tested. It also isolates the test circuit from unwanted radio frequency signals on the supply cables and couples the disturbance voltage to the measuring receiver.

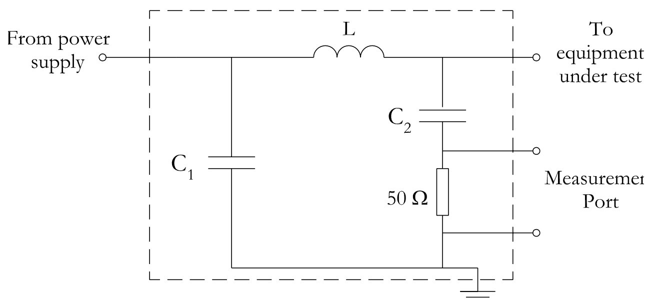

Figure 2.12. A simplified model of a LISN

In the standard CISPR 16, different LISNs for measurements of conducted emissions are defined, one example can be seen in Figure 2.13 and a simplified model is shown in Figure 2.12.

Although the LISN in Figure 2.13 is useful for most applications, it is not suitable for e.g. the aerospace or automotive industry. In these cases, other values of the network impedance are required. The impedance of a LISN is the magnitude of the impedance with respect to reference earth measured at an equipment terminal when the corresponding disturbance output terminal is terminated with 50 Ω.

By using a LISN, the tests that are performed are convenient to make and repeatable.

The noise voltage in a system measured across the 50 Ω resistor of a LISN is by definition the conducted EMI emission of the switching power circuit [31]. The values of the reactive components inside a LISN are such that for the fundamental frequency, the inductance is essentially shorted and the capacitors are essentially open. For noise frequency, however, the inductance is more or less an open circuit, and the capacitor C_2 is essentially shorted. This means that the fundamental frequency power goes through the LISN unaffected, and noise generated by the switching circuit sees a 50 Ω impedance from line to ground. This 50 Ω impedance is generally the input impedance of a spectrum analyser.

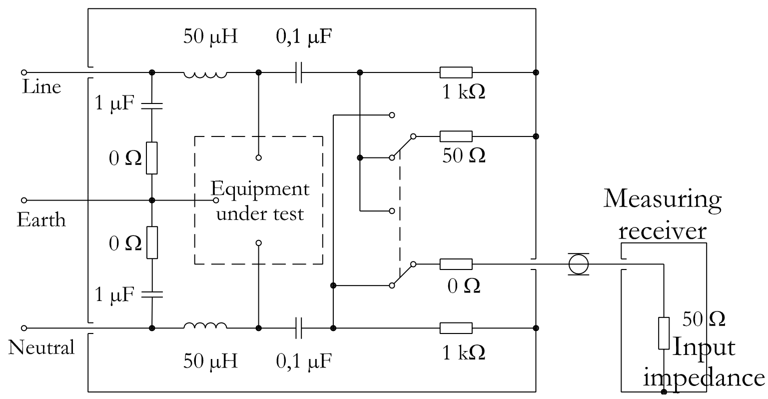

Figure 2.13. An example of a LISN

Since the voltage measured across the 50 Ω resistance is the measured conducted electromagnetic interference, this implies that any current coupled through the 50 Ω resistor causes electromagnetic interference conducted emission.

Absorbing clamp

If a device only has a main lead as the only external lead, the disturbance capability of this device may be defined as the power it could supply to its main lead acting as a radiating antenna [22]. This power is nearly equal to that supplied by the device to suitable absorbing equipment placed around the lead where the absorbed power is at its maximum. The absorbing equipment that is used is called an absorbing clamp or a ferrite clamp. When a device has more external leads than a mains lead, these leads could also radiate disturbing energy, whether they are shielded or not. The absorbing clamp could be used to measure the disturbances from these leads as well.

An absorbing clamp is in essence a current probe that is clipped on to the cable from the equipment that is being tested. It often consists of an assembly of ferrite rings, which has the purpose to absorb the power produced by the device under test and to prevent unwanted currents to flow on the coaxial cable that connects the absorbing clamp to the EMI instrument.

Ferrites have less effective impedance stabilization at lower frequencies, and since the function of the absorbing clamp is highly dependent on the ferrites, it should not be used at frequencies lower than 30 MHz.

When an absorbing clamp is used as a measuring device, it is clipped on to the lead of interest. By sliding the clamp along the lead, it is possible to se a number of high readings along the cable, due to the standing wave along the cable. It is the highest of these that is adopted as characterizing the EMI source.

It is important to keep in mind when using an absorbing clamp that a substantial amount of radiation could emanate directly from the device, and this cannot be measured with the absorbing clamp.

Spectrum analyser

One measurement receiver that is commonly used in EMC tests is the spectrum analyser. It is also often used for pre-compliance tests, since the digital storage function on a spectrum analyser facilitates the investigation since the effects of any hardware modification can be traced visually and easily. The instrument also allows examinations of a large frequency spectrum, which implies that occasional resonances in a system could be recognized immediately.

A spectrum analyser views the measured voltage as a function of the frequency as an ordinary oscilloscope views the voltage as a function of time. The instrument could therefore be described as a narrowband peak indicating voltmeter that sweeps over a range of frequencies. A simplified block scheme for the spectrum analyser can be seen in Figure 2.14.

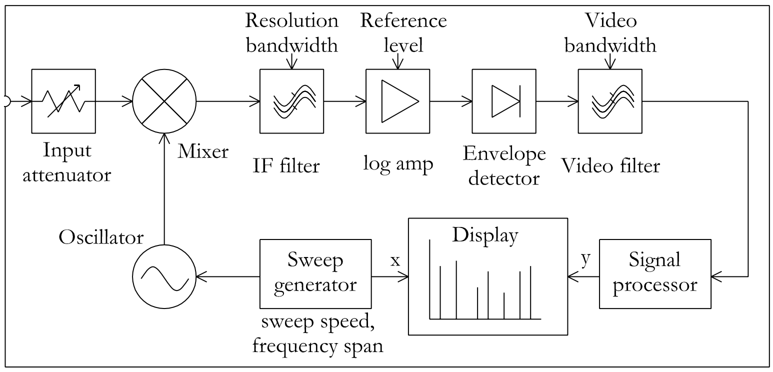

Figure 2.14. Block diagram of a spectrum analyser

In a spectrum analyser the input signal is mixed with a local oscillator in a mixer [12]. The difference frequency is then filtered through a narrowband band pass filter (the IF filter). When the difference frequency agrees with the frequency of the band pass filter, the mixed signal is let through and detected by the envelope detector. The mixed signal is the shown on the display of the spectrum analyser.

Spectrum analysers have true peak or average detectors and are calibrated on rms. This means that for narrowband signals, the results obtained by a CISPR receiver match those of a spectrum analyser.

Peak, quasi-peak and average measurements

There are three kinds of detectors in common use in radio frequency emissions measurements: peak, quasi peak and average. The characteristics are different for each frequency band (see Appendix B) and are defined in the standard CISPR16 [22].

Interference emissions are seldom continuous at a fixed level. A carrier signal, for example, may be amplitude modulated and both a carrier and a broadband emission may be pulsed. The three types of detectors gives different measured levels, depending on the signal that is measured (see Figure 2.15) [50].

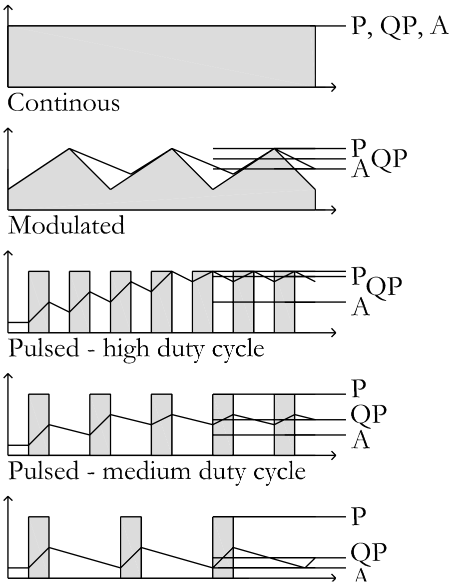

Figure 2.15. Peak (P), quasi-peak (QP) and average (A) measurements for different signals

Peak

A peak detector responds almost instantly to the peak value of the signal and discharges fairly rapidly. If the receiver resides on a single frequency, the output from the peak detector will follow the envelope of the signal. Because of this behaviour it is sometimes called an envelope detector.

Since the peak detector has a very fast response, it is very suitable for diagnostics and quick tests. The CISPR emission standards do not require the use of peak detectors.

Quasi-peak

The quasi-peak detector is a peak detector with weighted charge and discharge times. This detector was developed and recommended for common use by CISPR since it was observed that by weighting the time constants of the peak detector, a better correlation could be made between the EMI receiver readings and the broadcast disturbances heard by the human ear. This means that the quasi-peak detectors were established with reference to the human ear. It has been noticed that interference at low pulse repetition frequencies is subjectively less annoying on radio reception than interference at high pulse repetition frequencies. Therefore, a pulse-type emission will be treated more moderately by a quasi-peak measurement than by a peak measurement.

One drawback with the quasi-peak detector is that it is time consuming to get an accurate result since the measurement must dwell on each frequency for a longer time than the quasi-peak charge and discharge times.

Since the intention of CISPR-based tests historically have been to protect the voice and broadcast users of the radio spectrum, CISPR lay considerably emphasis on the use of the quasi-peak detector.

Average

The average detector measures, as its name implies, the average value of the signal. In the case where the signal is continuous, the average detector will measure a value equal to the peak detector, but in the cases where the signal is pulsed or modulated, the average level will be lower than the peak detector.

Test sites

When EMC tests are carried out it is important to use an environment that ensures valid and repeatable measurement results. For radiated measurements testing were former often conducted in an open field, but since the ambient is getting noisier, also shielded sites are used nowadays.

Open area test site

An open area test site (OATS) is historically described by the CISPR standard as an open site in which there are no obstructions within a defined distance of the measurement range. The size and set-up of an OATS is shown in Figure 2.16.

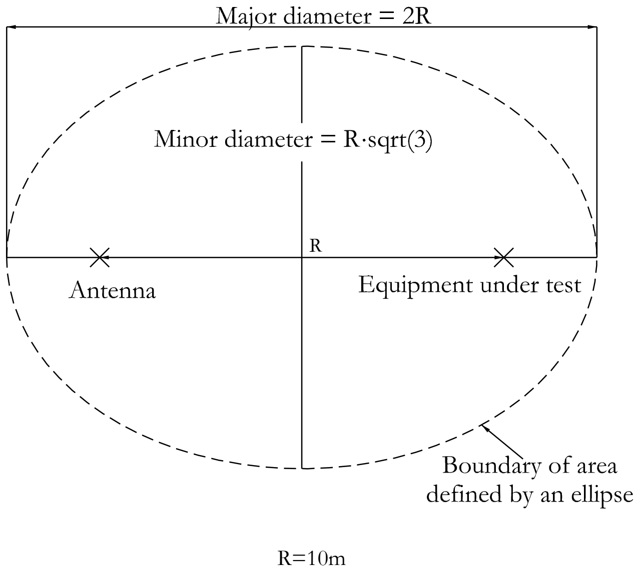

Figure 2.16. Measurements for an OATS

By having these strains on the test sites, it is possible to control any extraneous reflections. These demands often result in that an OATS is interpreted as a site that is actually out in the open. This is not necessarily true since a large building, such as an aircraft hangar or a bus garage, also could be suitable for measurements. An OATS that is indoors is also protected against bad weather, which otherwise could cause problems. Today, however, the weather is not the greatest difficulty when performing OATS measurements but the presence of ambient radio signals. Unfortunately there is no reliable method of removing them from an emissions measurement, so the best alternative when OATS measurements should be done is to get as far away as possible from civilization, out in an open field. The test site should be flat, free of overhead wires and nearby reflecting structures, sufficiently large to permit the placement of an antenna at the specified distance and provide adequate separation between antenna, the equipment under test and reflecting structure. A reflecting structure is defined as something in which the constructing material is primarily conductive.

Closed site testing

An increasingly popular alternative to the open test site with its ambient problems is the closed site, which is a site that is shielded from the normal ambient and has a controlled electromagnetic environment. Plain metal-shielded rooms can be used to cut down the amount of ambient interference, but it is important to make sure that there is a continuous metal skin with no gaps of any size. This complicates openings for doors and ventilation, and all mains power and signals entering or leaving the room must be adequately filtered.

As an alternative to a shielded room, shielded tents are available. A shielded tent is made of conductive fabric and are easily moved to another location or folded away when not in use.

The problem with a plain metal room or a shielded tent is that they suffer from very strong internal reflections, which gives rise to radio frequency resonances. A good way to reduce these resonances is to use radio frequency absorbers. The absorbers are available in two different types, ferrite tiles or large wedges made of carbon-loaded foam. Both alternatives are expensive, and the problem with the ferrite tiles is that they are very heavy and the problems with the wedges are that they take up a lot of space.

A shielded room with good normalised site attenuation require a large amount of absorbers. Usually the absorbers cover all the walls, the ceiling and the doors. If the floor is covered with absorbers, the shielded room is called a fully anechoic chamber, compared to if the floor consists of a large metal ground plane, then the chamber is said to be semi-anechoic

Chapter 3

Automotive electromagnetic compatibility

This chapter starts with a historical description of electromagnetic compatibility and vehicles. After that are the different EMC aspects of a vehicle and the automotive ground system discussed. Since the automotive industry has its own type of CE marking their products, called e marking, this is described in one section. Finally, electromagnetic compatibility problems in vehicles with an electrical motor for propulsion, so called hybrid electric vehicles, are described.

3.1 Background

The first consumer product to be subjected to electromagnetic compatibility legislation in the United Kingdom was the car. This was due to broadband impulsive noise from the spark ignition system, which originally was of sufficient magnitude to cause interference to domestic television reception. In 1952, the Parliament enacted a Statutory Instrument to limit this interference. The distance between the vehicle and a domestic receiver antenna was assumed to be 10 m, for the purpose of the standard that was created, which implied that reducing the radiated emissions were relatively straightforward.

In the 1950s and 1960s, the number of cars with a car radio increased significantly and this required a greater degree of suppression in order to reduce the audio interference to an acceptable level. The extra suppression had to compensate for both the closer vicinity of the receiver antenna to the ignition system and the relatively low level of the desired signal provided by a car antenna. There was an additional EMC phenomenon that was introduced, conducted interference, which also could cause audible interference noticeable on the radio. The conducted interference arose from transients generated by the ignition system or other vehicle electrical devices that were conducted into the radio by the 12 V power supply leads.

The solid-state electronic devices were introduced into the vehicle electrical system in the early 1970s. Initially, discrete semiconductors were used in alternators, radio and ignition system. When the semiconductors were used in the vehicle it was important to ensure that that they could withstand the high-voltage transients present on the vehicle harness.

In the mid-1970s large-scale integrated circuits and microprocessors began to be used for the control of vehicle functions, which previously had been implemented by purely mechanical means. An example was the electronically controlled fuel injection system that replaced the mechanical carburettor in the more expensive cars. Such systems are of course a potential victim to radiated emissions.

3.2 Automotive EMC aspects

As with any kind of electronic system, it is possible to break up the automotive EMC aspects into the four categories: conducted emissions, conducted susceptibility, radiated emission, radiated susceptibility and electrostatic discharge (ESD). Four of these categories will be discussed more in detail in the following section.

Conducted emissions

Since there is no physical external connection between a car on the road and the surroundings, all conducted emissions that are present in the car are only caused by, and affect only, components in the car. These conducted emissions are either transients and arise from the commutation of electric motors or the switching of solenoids, relays and other inductive components or continuous and created by for example a pulse width modulated system.

The emissions are conducted along the wiring harness and into the power supply terminals of electronic systems.

Conducted susceptibility

Transients that are coupled into electric systems via the cables could cause malfunction of the exposed component. One solution to this problem is to limit the magnitude of the transients that the inductive components cause. Another solution is to control the conducted susceptibility of the electronic systems so that no transient is big enough to cause a disturbance. This can be done by e.g. input filters, (supply) voltage limiters etc.

Radiated emissions

The radiated emissions mainly arise from three sources on the vehicle. One source is the wiring harness that might act as an aerial, thereby radiating emissions originating from the conducted transients. Another source are the electronic systems that contains high-speed logic such as e.g. microprocessor circuitry. The pulses from the clock in high-speed logic systems have a harmonics, which extend to over 100 MHz. These harmonics are either radiated directly from the system, using e.g. the printed circuit board tracks or the wiring harness as an antenna. The last source for radiated emissions on the vehicle is electronic chopper circuits. These are often used for lamp dimming or motor speed control and have a switching frequency in the range of tens of kilohertz. This implies that the emissions from the chopper circuits will have significant harmonics in the range of a few megahertz.

The only problem caused by radiated emissions is audible interference on the radio in the car. There is only one way to prevent radio interference, and that is by controlling the level of the radiated emissions. In the case where the radiated emissions originate from a microprocessor circuit, a careful design in order to reduce the radiated emissions is relatively straightforward. The radiated emissions from an electric motor are harder to suppress and does often require the addition of some components in order not to interfere with any other electronic equipment.

Radiated susceptibility

It is hard to predict the electromagnetic environment that a car will be subjected to since it is inherently mobile. Some examples of areas where the car could be subjected to a severe electromagnetic environment is near a broadcast transmitter, military transmitters, cellular phones and radar.

In most cases it is the on-board transmitters that creates the highest values of the field strength that the car experiences. Although the field strength from these apparatus is very high, many of the components such as Emergency Service transceivers and cellular telephones are well controlled. The manufacturers of these products have developed the products in close proximity with those responsible for installing the transmitters. This is in order to ensure that the transmitting aerials are located in a suitable position on the car and that the feeder cable is kept well away from sensitive wiring. However, in some cases the electronic equipment has to have an even higher immunity against radiated emissions. This could be in the cases when radio amateurs install their equipment without considering the rest of the electronic equipment in the car.

Electromagnetic fields and vehicle electronics

There are two ways for the electromagnetic fields to be coupled into the vehicle electronics. The first way is coupling into the electronic system via printed circuit board tracks or internal wiring. Both act as antennas and convert the field into a conducted voltage or current. The second way of coupling is when the vehicle wiring harness acts as an antenna and conducts the interference into the electronic system.

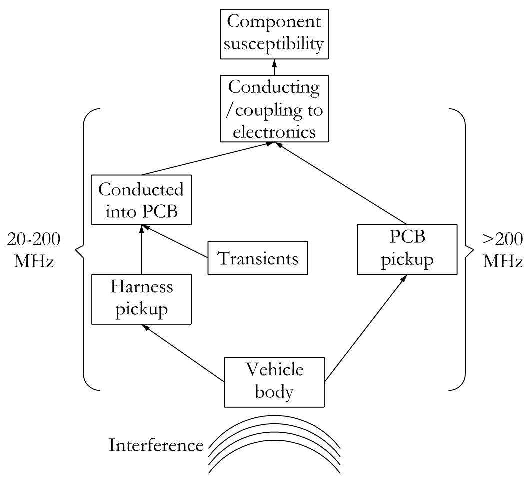

Figure 3.1. Coupling of interference into vehicle electronics. PCB = printed circuit board

Since the coupling process is strongly dependent on the frequency of the interference will neither the wiring harness nor the electronic system work as efficient antennas at frequencies at or below 20 MHz. This is due to their length, which is relatively short compared to the wavelength of the interference (a good antenna has a length equal to a quarter of a wavelength). The wiring harness works as a reasonably efficient antenna between 20 and 200 MHz. Currents in the magnitude of 1 mA could then be caused to flow for each volt per meter of interference field.

The vehicle harness attenuates high frequencies above 200 MHz and is therefore no longer a good antenna. But the lengths of the printed circuit board tracks are becoming comparable to the quarter wavelength value and the interference could then be coupled directly into the electronic system. The electronic circuit themselves are fortunately lossy at such high frequencies, so no problems generally occur.

3.3 Automotive ground system and battery

Conductors referred to as ground often perform (intentionally or unintentionally) more than one task in an electrical system [30]. In a vehicle, the term ground is even more diverse and is often used as a description of conductors which function as:

- a current return path,
- an equi-potential conducting surface that provides a voltage reference level for analogue sensors and digital logic,
- shielded conductors which are used as a bypass for radio frequency currents,
- a drain for electro static charges.

When the engine of the vehicle is running the alternator supplies current for most DC-loads but the internal inductance in the alternator is too large to supply rapidly switched currents. The battery therefore supplies these currents. This configuration makes the current return path somewhat ambiguous. One additional complication is the starter motor since it, in order to ensure the function when the weather is cold, should have a current return path with a very low DC-resistance. This implies that the battery negative should be connected to the ground of the starter motor (the engine block), and the placing of the starter motor and battery may cause unfavourable loops for this conductor.

The battery of a vehicle is not only an electrical energy storage, but it also performs a number of noise suppression tasks. It has a highly non-linear electrical characteristic, which depends on the frequency, state of charge and load current. At low frequencies the battery has a capacitive reactance of 1-2 F per 100 Ah. This reactance reduces the low-frequency noise components created by the ignition system when the battery is placed in parallel with the alternator.

For frequencies in the range of 25 kHz to 250 kHz the capacitance of the battery is roughly constant with the frequency and in the order of some tens of nanofarads. At higher frequencies the complexity of the battery increases and it is then best described by an LC-circuit, including the lead-inductance. This implies that series resonances are common for frequencies above 1 MHz, and that the battery should not be relied upon to provide a radio frequency suppression in the AM-band and above, see Appendix C.

Although there is a trend in the automotive industry towards composite or plastic body panels in vehicles, the majority of vehicle bodies still have a significant metal content. Due to its size, shape and convenient localization (i.e. adjacent to most electrical components), the body sheet metal offers a unique opportunity for current return paths, potential referencing, shielding, and noise reduction. The large surface area and the folded shape of the body sheet metal offer a significant self-capacitance. It has been found that the body sheet metal is an effective sink for low frequency noise currents with small amplitudes. Using the body sheet metal for these purposes increases the risk for common impedance coupling. This is especially a problem when the current has to pass through body panel joints. An indiscriminate use of the body sheet metal as a universal current return path could furthermore eliminate its beneficial features and create a significant source of radiated emissions.

The body sheet metal could also be used for low frequency capacitive shielding. For frequencies above a couple of kHz, the body sheet metal creates a modest electromagnetic shielding of the engine compartment as a result of its skin depth. It is also the primary conductor through which the vehicle is shielded from lighting, human generated electrostatic discharges and external sources of radiation.

In an ideal sheet metal body should the sheet metal panels be bonded electrically, and movable panels, such as door and hood, should be connected with ground straps. The body sheet metal should also be connected to the battery negative through a robust low-impedance bond. These kinds of connections are seldom present in a body sheet metal. The metal panels are often joined by spot welds or bolts with intervening layers of nonconducting corrosion inhibitor, or with nonconductive paints.

3.4 e marking

As mentioned in Chapter 2 the European EMC framework directive requires that all electrical and electronic products offered for sale in the European Union should conform to its requirements and that they should be CE marked. This is valid for all electrical products, unless a separate product specific directive covers them, and one example of such a product specific directive is the automotive industry. The main reason for the automotive industry to have its own directives is that none of the existing European standards were sufficient to take care of the types of environment to which a motor vehicle could be exposed. Since a motor vehicle is mobile, it is hard to predict the electromagnetic environments that the car will be exposed to. Some examples of severe electromagnetic environments that a motor vehicle could be exposed to are broadcast transmitter sites, airports (with radar) and installation of internal mobile communications equipment. Since there are many safety critical applications like anti lock brakes in even the most basic motorcars today, a high level of immunity in the electrical system of a vehicle is vital.

This has led to that the automotive industry has had its own marking, called e marking, since 1970 which then is a required approval for all motor vehicles sold into the European Union. Since October 1, 2002, also manufacturers of aftermarket products intended for use in vehicles are required to obtain a formal type approval for their products before placing them on the market [24].

One thing that may come as a surprise for people used to CE marking is that the testing and approvals rules required for e marking is very different from the rules that are normal for CE marking. The major difference between the two types of marking is the route to approval. The norm for CE marking is manufacturer self-declaration, but the only route to comply with the automotive directive is formal type approval. This formal type approval for a product could only be issued by a type approval authority in one of the EU member states, which in Sweden is Vägverket. There are two ways in which a vehicle can be approved – the whole vehicle together with all its available components can be tested or components can be tested individually on the bench, and approved in their own right. Testing each individual component is an important option since there may be a number of alternative components available. One example is the audio system of a vehicle. It would be very expensive and time consuming to repeat the whole vehicle test for each alternative radio that is available.

There are three areas on which the e marking is focused on: broadband emissions predominantly generated by the ignition system of petrol engines, narrowband emissions generated by harmonics of microprocessor clock oscillators and immunity to externally radiated electromagnetic fields.

Table 3.1. Differences between 95/54/EC and CISPR22

95/54/EC	CISPR 22
Equipment under test (EUT) placed 50 mm above fixed table with conducting surface, 1 m above reference (ground) plane	EUT placed on nonconducting, rotating table 800 mm above reference (ground) plane
Measurement antenna site 1 m from EUT	Measurement antenna site 10 m from EUT (or 3 m with the results extrapolated to 10 m)
EUT fixed throughout test	EUT rotated 360° during test
Measurement antenna height fixed throughout test	Measurement antenna height varied from 1 to 4 m during test
Measurement antenna pointed at midpoint of wiring loom connected to EUT	Measurement antenna pointed at EUT
Single product class	Two product classes (A and B) with different limit lines
Two emission types, broadband and narrowband, with different limit lines	Same limit lines apply to all emissions
Average and quasi-peak detectors used for narrowband and broadband emissions respectively	Only quasi-peak detector is used

Also the test methods for e marking differs from the methods used when CE marking. One example is when electronic subassemblies are tested. Normally, these products are tested according to the test methods given in the standard CISPR 22, but for the automotive e marking, testing according to the directive 95/54/EC are required. Some major differences between those test methods can be seen in Table 3.1 [16] [24].

As with the generic EMC directive, there are a number of tests required to demonstrate an acceptable performance of a motor vehicle.

Conducted emissions testing

Since a vehicle normally is not physically connected to anything via cables, the legislation for vehicles contains no requirements to test the conducted emissions. The legislation for testing electronic subassemblies is derived from the legislation for vehicles and so, does the Automotive Directive 95/54/EC not contain any requirements on conducted emissions. It is assumed that the wiring in a vehicle is an electrically noisy item, and it is therefore ignored whether an electronic subassembly pollutes the wiring even more. Vehicles manufacturer employs traditionally the opposite logic and says that an electronic subassembly should have a high level of immunity to conducted interference.

Radiated emissions testing

The idea for radiated emission testing in a vehicle is to ensure that the vehicle does not cause any unnecessary interference with other road users or any other fixed radio receivers. The problem with interfering the vehicles own radio is not covered in the directives, but is instead a quality matter for the vehicle manufacturer to address.

When it comes to radiated emissions the vehicle directive gives two limits, one for broadband and one for narrowband emissions. Broadband emissions are those characteristic of spark ignition systems, and narrowband emissions are those characteristic of microprocessor based systems. Each equipment under test is tested against both the broadband and the narrowband limits. This implies that a cautious manufacturer ensures that the emissions fall well below the narrowband, no matter which classification the product has.

Conducted susceptibility testing

An electronic subassembly of a vehicle does not have to meet any specific conducted susceptibility or electrostatic discharge criteria. The legislation places the responsibility on the vehicle manufacturers to ensure that electronic subassemblies are immune to conducted interference.

Radiated susceptibility testing

The radiated immunity requirement for an automotive electronic subassembly is very troublesome. One example is the field strength for free-field immunity testing is 30 V/m (and some vehicle manufacturers have internal requirements that recommends 100 V/m), which could be compared with 3 V/m for EN55024B, which is a standard for testing immunity of information technology equipment. The requirement for immunity is to show that the basic safety of the vehicle is not compromised by radio transmissions for example, and so it is only the components that may affect the direct control of the vehicle that needs to pass the immunity requirements. Examples of such systems are cruise control, anti lock brakes, steering and lighting.

When a manufacturer has a product they believe are compliant with 95/54/EC the formal type-approval process must begin before the product can be sold into the European Union market. The formal tests of the product are performed in a laboratory approved by the authorities. This laboratory will ensure that the test sample and its arrangement are set up for the worst-case condition. The manufacturer must also submit product documentation in support of an application. When all tests have been completed successfully, a certificate is issued which becomes a part of the vehicle type approval.

Whole vehicle radiated susceptibility testing

When the whole vehicle is tested for radiated susceptibility, all systems are tested at once. The vehicle is usually put into an anechoic or semi-anechoic chamber with dynamometers under the wheels (rolling road) in order to be able to load the engine of the vehicle. Since it takes a lot of time to perform these whole-vehicle tests, the chamber offers forced air-cooling for the engine and an extraction system for the exhaust gases. The frequencies with which the vehicle is being irradiated range from 100 kHz to 10 GHz and the field strength can be as high as 100 V/m in order to simulate a worst-case scenario. The signals that are used are either continuous wave (CW) or amplitude modulation (AM) with a modulation frequency of 1 kHz and a modulation index of 0.8.

During the time when the tests are performed on the vehicle, no persons are allowed to be inside the chambers due to the high field strengths. The vehicle is therefore monitored by video cameras that are placed inside the chamber. The signal is sent to the control room where it is shown on television screens and supervised by the personnel.

A commonly used method for testing radiated susceptibility is the substitution method, since it agrees very well with reality. When the substitution method is employed, a spot somewhere near the midpoint of the dashboard is selected as a reference point for the field strength. The field strength in this point is first determined when there is no vehicle present in the chamber. The output power that is needed in order to produce the wanted field strength in the reference point is then measured. This value of the output power is then used during the tests when there is a vehicle present in the chamber. Normally, three antennas are used during tests with the substitution method, one at the front of the vehicle and one on each side of the vehicle.

Sometimes car manufacturers also place mobile transmitters on the vehicle to test how the vehicle withstands the high-localised field strengths generated by these transmitters.

3.5 EMC in vehicles with electric drives

From the EMC point of view, the integration of electric traction drive systems into today's vehicles represents a considerable challenge. From an EMC perspective the traction drives operate in similar ways as small pulse width modulation controlled drives, e.g. the power window drive or the electric power steering drive, but at one order of magnitude higher voltage and several orders of magnitude higher power. Such traction drives are used in pure battery vehicles, combustion electric hybrid vehicles and fuel cell vehicles. If the traction drive would be treated as a conventional automotive component it would lead to substantial incompatibility problems [19].

The electric traction drive is a system consisting of a high-voltage power source, a frequency converter, an electric motor and high-power cables. These high voltage and power ratings implies that conventional test procedures are not appropriate for such components due to size, weight, power and emitted noise. The electric drive components are connected to a high-voltage DC bus, designed to be a galvanically insulated power supply network and not galvanically connected to the conventional electrical system.

Since the power electronic systems are known to be the main source of electromagnetic interferences within electric drive systems, there is a great deal of research in these areas. Different converter topologies such as resonant converters or multilevel inverters have been explored to be advantageous for electric drives, but the first commercial applications are equipped with a conventional hard switching power converter. The biggest noise sources in the power electronics are the switched semiconductors. The short rise and fall times of the voltage creates a significant amount of electromagnetic noise.

One very important aspect that has to be taken into consideration during the design process of an electrically driven vehicle is the cables in the high-voltage bus. These cables have to be taken into account since the magnitude of the currents, the length of the cables and the frequency of the noise implies that the power electronic applications emit the noise mainly through the cables. Usually the cables between the converter and motor are kept very short to gain as good results as possible in terms of volume and electromagnetic compatibility. Since the space in a vehicle is limited, there have to be longer cables for the connection to the supply voltage. As mentioned before this means that there are high-voltage cables carrying as much as 900 V within the system. An important question during the design process is to find out if shielded cables are necessary or not. Generally, shielding reduces the emissions in order to ensure electro magnetic compatibility within the system. The drawbacks with shielding are higher costs, increased weight and problems in the assembling. Shielded cables have a reduced flexibility and it is not possible to assure the grounding of the shield after some time due to vibrations from the vehicle in motion. Regarding the EMC criteria, the best solution would be a common shielding of the high-voltage bus cables. Unfortunately, such a solution would aggravate the problem since the cooling conditions will become impaired. This would lead to an increase in the size of the cables and an even higher cost and lower flexibility. The tradeoff between shielded cables and noise suppression has therefore to be found for each specific application.

An electric drive system consists of several components, which are only connected to the high-voltage bus. When an electric drive system is integrated into a vehicle, the noise sink to be protected against interference is the conventional electric system and its low-voltage devices. This implies that the connections between the conventional system and the new system have to be analyzed.

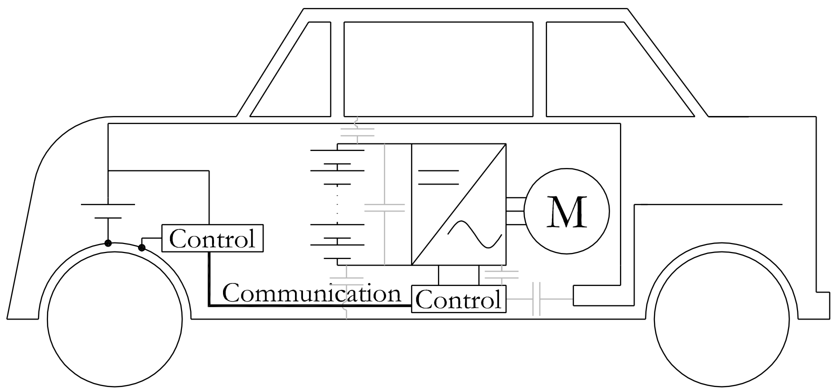

Figure 3.2. An example of a vehicle with an electric traction drive. The parasitic capacitances are drawn in grey.

However, as mentioned before, the high-voltage system will be insulated and it will therefore not use the car body as ground return. One important coupling path between the two systems is capacitively coupled crosstalk between cables. Since the space for the wiring harness is very limited in a modern car, this is not a negligible coupling path.

Chapter 4

Pulse width modulation

This thesis treats the electromagnetic compatibility problems that arise when a load is pulse width modulated. In this chapter, the theoretical background to pulse width modulation is discussed.

4.1 The converter

A power electronic energy converter in an electrical control system operates as a power amplifier. The energy conversion of a process can be influenced by controlling the voltage fed to the load. In a low power system it is possible to control the voltage continuously, but in many cases the output power has to be controlled by modulation in order to reduce the power losses in the converter itself. Voltage modulation means that the instantaneous value of the output voltage alternates between well-defined levels, and is thus not controllable to an arbitrary value in every time instant. The goal when using carrier wave modulation is to obtain linearity between the reference of the output voltage and the average of the output voltage measured over a finite time period.

There are several different ways to modulate the output voltage. This thesis will focus on a type of modulation where a carrier wave is used, but there are also modulation methods that involve a direct control of secondary quantity such as current or torque.

A power electronic converter involves the use of power electronic components, arranged to operate as switches, i.e. as elements that either conduct with ideally with no voltage drop across them or block with ideally no current –

like a mechanical switch. When studied closely it is found that the circuit on each side of the switch element has different electrical properties. One side is always capacitive and one side is inductive. This is the same as to say that a power electronic switch element cannot/should not be used to connect two capacitors of different charge since the initial current through the switch would be infinite, or the switch element cannot/should not be used to break up an inductive circuit since the induced voltage would force the current to continue through the semiconductor chip, in both cases the switch element would be destroyed. The capacitive side of the circuit has a continuous voltage and the inductive side has a continuous current through the switch transition. Correspondingly, the capacitive side has a discontinuous current and the inductive side a discontinuous voltage.

Carrier wave modulation

This principle for modulation can be derived from Figure 4.1. A two-position switch that controls the potential v_c is shown in the circuit in Figure 4.1. When this switch switches periodically, the output voltage alternates between the potentials v_a and v_b with the same frequency.

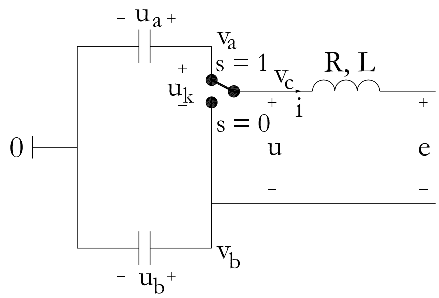

Figure 4.1. A power electronic switch, controlling the potential v_c. When the switch is in its upper position, v_c equals v_a and when it is in its lower position, v_c equals v_b.

In this circuit the capacitive side is represented by the capacitive potentials v_a and v_b and the inductive side by an inductive current i flowing through a load consisting of a resistance (R), an inductance (L) and a counter electromagnetic force (e) in series. The output voltage will then be a function of the switch position according to

$$u = s \cdot (v_a - v_b) = s \cdot u_k \tag{4.1}$$

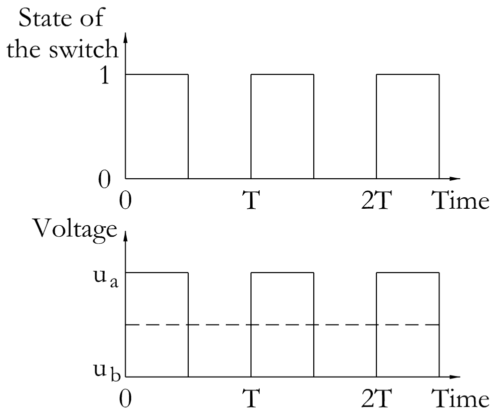

Figure 4.2. Upper diagram, state of the switch as a function of time. Lower diagram, output voltage as a function of time, the dashed line is the average output voltage.

For the following discussion, a circuit according to Figure 4.3 is assumed. This is a 2-quadrant DC converter that is suitable to explain carrier wave modulation method used in this thesis. The same method, with small additional assumptions, can be used for more complex converters, see [2].

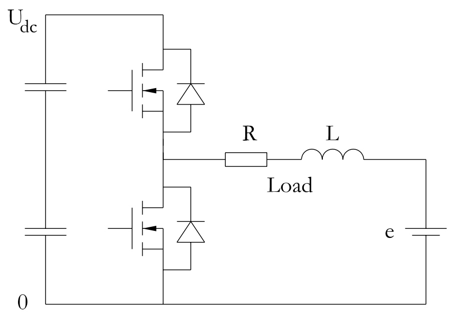

Figure 4.3. A 2-quadrant DC converter with an RLE load

Modulation of power electronic switches is about the selection of time instants to “flip the switch” in order to obtain a desired average output voltage within a certain time interval. In this thesis, the output voltage is created by using carrier wave modulation. The carrier wave itself is derived from a study of the average voltage in one half switch period.

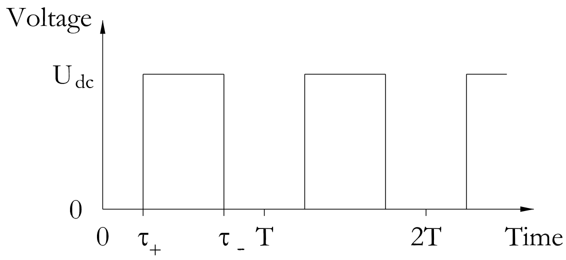

Figure 4.4. Output voltage pulses symmetrically positioned in one switch period.

Assume that the output voltage pulses is positioned in the middle of the switch interval, see Figure 4.4. The average voltage $\bar{u}$ during the first and second half of the switch interval can then be described according to:

$$\begin{cases} \bar{u}(0 < \tau_+ < T/2) = U_{dc} \cdot \left(1 - \dfrac{\tau_+}{T/2}\right) \\ \bar{u}(T/2 < \tau_- < T) = U_{dc} \cdot \dfrac{\tau_-}{T/2} \end{cases} \tag{4.2}$$

where T is the switch period at the assumed switching frequency $f_{sw}=1/T$. Modulation is now about selecting the time instants τ_+ and τ_- to obtain an average value of the voltage that is equal to the voltage reference within the particular halfpulse. This is done by using a modulator.

The modulating wave and the modulator

The modulating wave, u_m, used in carrier wave modulation is selected to be equal to the predicted average voltage for each half period, according to:

$$u_m = \bar{u} \tag{4.3}$$

The modulator is represented by a comparison between the desired output voltage and the carrier wave, where the switch changes state when the comparison changes sign. This can either be done in an analogue manner with a triangular wave generator and comparators or digital by using a timer circuit or an up/down counter with a digital comparator.

With a carrier wave according to equation (4.2), the output voltage is linearised with respect to variations in the supply voltage. Figure 4.5 illustrates a few modulation periods with varying voltage reference and supply voltage.

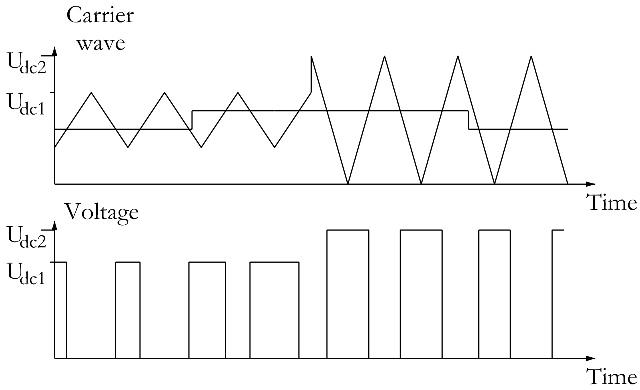

Figure 4.5. Upper: Carrier wave and reference, lower: output voltage, note that the pulse width decreases when the supply voltage increases.

If the supply voltage is lowered, the pulse width is increased in order to maintain the desired average output voltage.

Current sampling

When a generic RLE circuit is driven by the output voltage of the converter according to Figure 4.6, the current will cross its average value at the time instants when the carrier wave turns.

The current will have one out of two derivatives according to:

$$\frac{di}{dt} = \begin{cases} \dfrac{U_{dc} - e - R \cdot i}{L} \\ \dfrac{-e - R \cdot i}{L} \end{cases} \tag{4.4}$$

From Figure 4.6 it is indicated that the current ripple will pass through its average value twice every switching period. Assuming that the electric time constant L/R>>T, this occurs at time instants that coincide with the turning of the carrier wave. Thus, in applications where current feedback is needed, there are excellent opportunities to sample the current at these time instants.

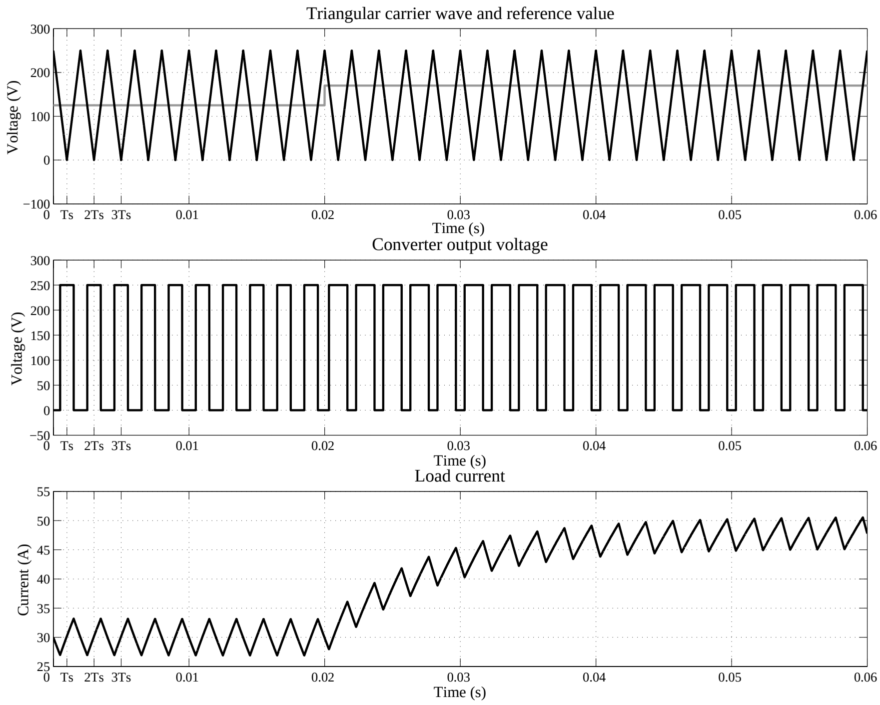

Figure 4.6. 2-quadrant converter supplying a resistive-inductive load with a back emf.

Modulation frequency

In some applications, it might be of interest to have a carrier wave frequency much faster than the sampling frequency. When this is implemented, it is important to make sure that the sampling instants still coincides with the turning of the carrier wave. This implies that the carrier wave has to consist of an integer number of half periods between two sampling instants.

The frequency of the carrier wave is equal to the switching frequency of the converter. This frequency is depending on the application. A high switching frequency implies a faster and more accurate control of the controlled quantities, and in many applications, a switching frequency higher than 20 kHz is used since the human ear does not perceive frequencies higher than 20 kHz. The problem with a high switching frequency is that the switching losses are proportional to the switching frequency, i.e. selecting switching frequency for an application is a trade off between high frequency in order to spare the human ear and to get an accurate control of the system and switching losses.

This method to create variable output voltage by controlling the pulses voltage pulses is called Pulse Width Modulation (PWM).

4.2 Types of converters

There are different types of converters depending on which direction the energy flow should have. The three types that are available are 1-quadrant, 2-quadrant and 4-quadrant converters. The quadrants are defined in Figure 4.7 where u and i are the converter output voltage and current. A 1-quadrant converter, for example, works in an area where the voltage and the current are positive. A 2-quadrant converter could operate with strictly positive voltage but a bi-directional output current, or the opposite combination. This means that the energy flow can have different directions in a 2-quadrant converter. In Figure 4.7 it is possible to see the difference between the different types.

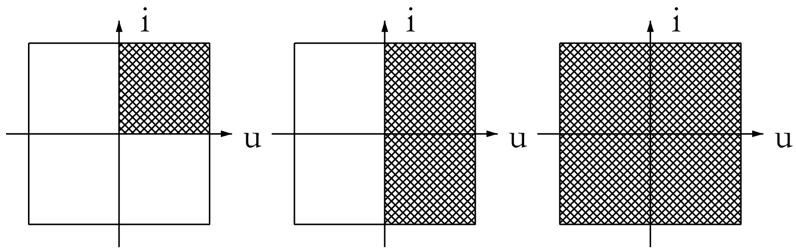

Figure 4.7. From left: the working area of a 1-quadrant converter, a 2-quadrant converter and a 4-quadrant converter.

A 4-quadrant converter, or full-bridge converter, is built with two phase legs, each consisting of two semiconductor switches and their anti-parallel diodes Figure 4.8. The semiconductors in a converter are often IGBTs (Insulated Gate Bipolar Transistors) or MOSFETs (Metal-Oxide-Semiconductor Field Effect Transistor). The main focus of this thesis will be on the MOSFETs, since this is the component that has been used in the laboratory set-up. The purpose of the anti parallel diodes is to provide an alternative path for the current in some switch states.

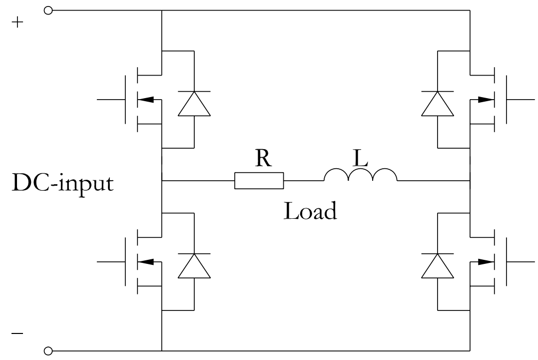

Figure 4.8. An H-bridge

The two switches in one of the legs in a full-bridge converter are switched in such a way that when one of them is conducting the other one is always turned off to avoid short-circuiting the DC-input. In most cases both switches are also never in their off state simultaneously, except for a short time interval called blanking time, which also has the purpose to avoid short-circuiting.

For a full bridge converter it is possible to have any combination of polarities for the voltage and the current. This implies that the converter could be used in only one of the four quadrants if wanted. In this thesis, a full bridge converter has been used as a 1-quadrant step down converter. A step down converter converts energy from a higher voltage to a lower.

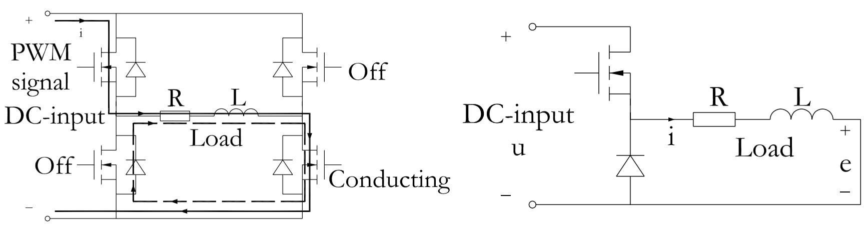

Figure 4.9. A 4-quadrant converter (left) used as a 1-quadrant step down converter (right).

Figure 4.9 shows a picture of a how the 4-quadrant converter is used as a step down converter. When the semiconductor switch is conducting, the current flows from the dc-input voltage through the MOSFET and to the load. Since the load is inductive the current will increase. When the switch is turned off the inductance will continue to lead the current through the freewheeling diode. Due to the inductive nature of the load, the current will slowly rise during the time period when the MOSFET is conducting and slowly fall when it is freewheeling through the diode.

Chapter 5

EMI mitigation techniques in vehicles

Many of the EMI mitigation techniques used in other applications and areas are not applicable in a vehicle. In this chapter is first the electromagnetic environment in a vehicle described before some solutions are proposed. The effects from the solutions proposed in this chapter will be described in Chapter 7.

5.1 The electromagnetic environment in a vehicle

The number of electrical loads in a vehicle is increasing. Due to lack of space and demands on the fuel consumption, some of the loads that are traditionally mechanically driven today will in the future be driven by electricity. An electrically driven load is easier to place, as for example an electric air conditioning compressor, which then not requires to be connected to the crankshaft by a belt, and is also easier to control on demand. When controlling a load on demand, it helps reducing the fuel consumption. A new type of vehicle with an increasing market is the hybrid. In a hybrid, an ordinary combustion engine is combined with an electrical motor. The electric motor will work together with the combustion engine in order to allow the combustion engine to work with the best possible efficiency.

Many of these systems are controlled by pulse width modulation see Chapter 4, which is a method that enables a way of controlling the electrical loads by e.g. demand.

As mentioned before the space inside a vehicle is packed, and it is not unusual for the power electronics and the load to be placed at a distance of about 1-2 m apart. The cables between them are mostly unshielded, and so, the disturbances from these cables will be heard on the radio.

One of the problems with a pulse width modulated system is the shape of the voltage pulses that propagates along the cable between the power electronics and the load. These pulses create a square wave signal from the battery voltage.

When a Fourier transform is performed on a square wave, the result shows that an infinite number of sines build up a square wave and that the magnitude of these sines fall of with frequency. When this square wave shaped voltage propagates along the cables, it is not only a sine with the fundamental frequency that is present, but also a number of higher harmonics. Since the cables between the power electronics and the load in a vehicle always are placed near to the body sheet metal, there will always be a parasitic capacitance between the cables and the load. This parasitic capacitance creates a closed current loop for a common mode current, see Chapter 2, which flows from the power electronics, along the cables and down to ground by the parasitic capacitance.

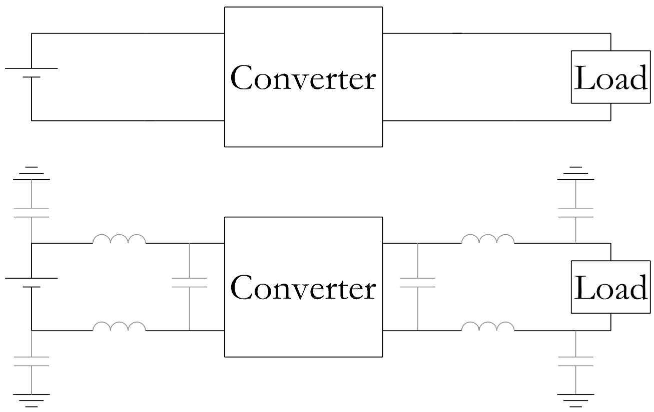

Figure 5.1. Upper: An ideal picture of a pulse width modulated system in a vehicle. Lower: A more realistic picture of the system with the parasitic components marked in grey.

This is where some of the problems with radiated emissions arise. The common mode current together with the cables and the body sheet metal creates a loop antenna, which thereby radiates magnetic emissions. The common mode current through the parasitic capacitance is determined by the size of the parasitic capacitance and the voltage derivatives as in

$$i = C_{parasitic} \cdot \frac{dV}{dt} \tag{5.1}$$

This implies that when affecting the parasitic capacitance by changing the distance between the cable and the body sheet metal or by affecting the voltage derivatives, the common mode current can be changed.

In the automotive industry, there is no room, neither financially nor literally, for any extra components, e.g. filters. In addition to this, there is no DC-link capacitor on the DC side of the converter, since the battery is supposed to keep the voltage stiff. This implies that the electrical environment in a vehicle is noisy, and it is up to each accessory manufacturer to make sure that their components can handle this environment.

The automotive industry demands that every component except for the alternator should be able to withstand a reverse connection of the battery. This demand requires either semiconductor switches at the input of every system in order for it not to be exposed to a voltage of reverse polarity or that components such as RCD snubbers (see Appendix D) are not used, since they could cause a short circuit if the battery is reversed.

In this thesis, the focus has been on investigating the effects from different cable layouts and cables on the radiated emissions, and also to propose and test two different solutions that not require any extra components.

5.2 Proposed solutions

In order to decrease the emissions in the vehicle without introducing any new components some solutions are proposed here. Also a solution with using a shielded cable is included in order to see the effects of a cable shield on the system.

Using a higher gate resistance

As mentioned in Chapter 2 the radiated emissions originating from the common mode current are affected by a change in the voltage derivative. The pulse edges in the voltage square wave are steep and create thereby an almost infinite time derivative, which in turn means that the current created by this high derivative is also high. If the pulse edges were a little less steep, the time derivative of the voltage would decrease.

Given that the voltage pulses are created by the MOSFET in the converter, it is possible to change the shape of the pulses by affecting turn-on and turn-off of the MOSFET.

There are a number of ways to affect the turn-on and turn-off of a MOSFET. One example is to use either a device for controlling the gate voltage or by using a so-called "smart" MOSFET where this device is integrated on the same chip as the MOSFET. By using these methods the gate voltage, and thereby also the turn-on and turn-off, could be formed in almost any shape. The drawbacks with these methods are that they require an extra component or a more expensive component than an ordinary MOSFET. A simple solution to this problem could be to increase the size of the gate resistance of the MOSFET. Since the gate resistance is an available component, a change in the size will not imply any increase in cost or required space.

Increasing the gate resistance implies that the time constant with which the gate-source voltage increases also will increase see Appendix D.

$$\frac{dv_{DG}}{dt} = \frac{dv_{DS}}{dt} = \frac{i_G}{C_{gd}} = \frac{V_G - V_{GS,I_0}}{R_G \cdot C_{gd}} \tag{5.2}$$

An increase in the rise time for the gate-source voltage implies that it will take longer time for the MOSFET to turn on, and hence longer time for the drain-source voltage to fall. The same holds for the turn-off of the MOSFET when the drain-source voltage should rise. This limits the time derivatives of the voltage pulses at turn-on and turn-off, which will suppress some of the radiated noise.

Although an increase in the gate resistance will suppress some of the radiated noise, there are several drawbacks that have to be taken into consideration before implementing this method. Firstly, a fixed gate resistance does not allow for controlled voltage transition when switching, but a "scaling" of the time duration of the voltage transition. The second drawback is the increase in switching losses.

Switching losses are the losses that occur during turn-on or turn-off of a MOSFET. Since a MOSFET is a component with very short turn-on and turn-off times these losses are often considered to be very small. But when the component is switched with a certain switching frequency and the resistance to the gate is increased, then the switching losses starts to constitute a larger part of the total switching losses, which also takes into account the conduction losses. Larger losses do not only decrease the total efficiency of the system, but also increases the temperature on the chip [43]. This is not always eligible since the reliability, service life and performance of a semiconductor is a direct function of its junction temperature.

In Figure 5.2 approximated switching characteristics shown for a MOSFET when a low gate resistance has been used [25]. The upper diagram depicts the drain-source voltage and the drain current during a MOSFET turn-on and turn-off, and the lower diagram depicts the instantaneous values of the power losses, which are obtained by multiplying the instantaneous value for the drain-source voltage with the instantaneous value of the drain current. The integral of the lower diagram will give the total energy loss during one switching period. In order to get the total switching losses, the total energy losses should be multiplied by the switching frequency.

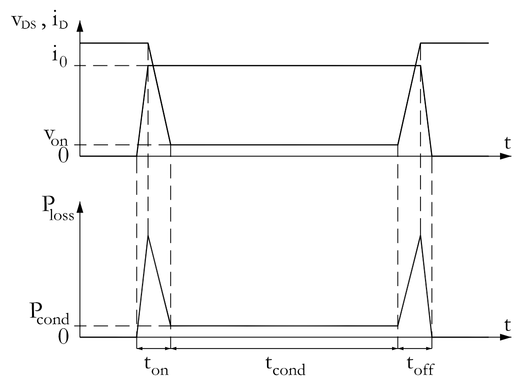

Figure 5.2. Upper diagram: Approximated drain-source voltage and drain current during MOSFET turn-on and turn-off. Lower diagram: The approximated switching losses at every time instant.

As mentioned above, an increase in the gate resistance implies an increase in the turn-on and turn-off times of the MOSFET since the time derivatives of the drain-source voltage and the drain current increases. This, and also the effect that it has on the power losses, can be seen in Figure 5.3 where approximations of the case with a low gate resistance is compared to approximations of the case with a high gate resistance.

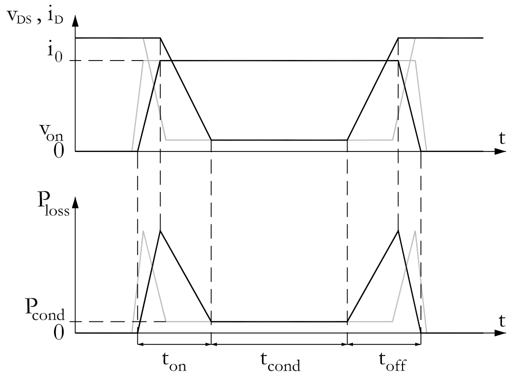

Figure 5.3. Upper diagram: Approximated drain-source voltage and drain current during MOSFET turn-on and turn-off when different gate resistances are employed, grey: low gate resistance, black: high gate resistance. Lower diagram: The approximated switching losses at every time instant, grey: low gate resistance, black: high gate resistance.

As can be seen in Figure 5.3 there is an increase in the power losses when a higher gate resistance is employed. This implies that when choosing the method with a higher gate resistance, there must be a careful consideration where the switching losses and increase in temperature are considered against the profits gained in the emission spectrum.

Random pulse width modulation (RPWM)

One idea that has gained foothold in power electronics of the 1990s is the random pulse width modulation (RPWM) technique, especially in inverter fed AC drives. Several independent studies have shown that RPWM could be a simple and effective tool for electromagnetic noise suppression in electric drives [29]. It has also been shown that the best way of removing or transferring the harmonics originating from the switching frequency is to randomly change the switching frequency. Commercial products that are utilizing RPWM have now appeared on the market.

The fundamental difference between the classic PWM and RPWM methods is that the power carried by the pulse width modulated signal is no longer restricted to a few dominating frequencies, normally determined by the switching frequency and the modulated signal. The total power is conserved, but the spread of power results in two major advantages due to the RPWM generation: the acoustic noise often emitted by a PWM controlled motor drive may from a subjective point of view be converted into less annoying broad band noise, and compliance with standards for electromagnetic interference may be met more easily.

A well-designed RPWM technique has the same fundamental frequency characteristics as a technique with a fixed switching frequency (FPWM). The difference is best observed in the higher frequency ranges where FPWM has harmonics that agrees with multiples of the switching frequency while RPWM transforms these harmonics to a continuous density spectrum. However, this approach to reduce electromagnetic interference peaks may appear questionable since the same power is only spread over a wider frequency range [8].

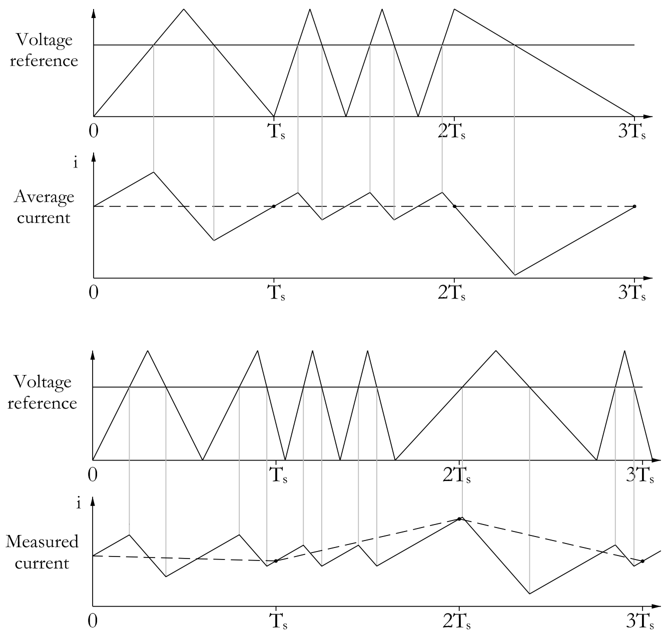

Figure 5.4. Upper two diagrams: RPWM where the switching frequency is altered at every sampling instant. The current that is sampled corresponds to the average value of the current. Lower two diagrams: RPWM where the change in switching frequency does not correspond to the sampling instants. Here the sampled current also includes some of the current ripple.

Sampling of the current of the controlled load is often an important feedback quantity. In these cases, where the sampling frequency is fixed, it is important to assure that the average of the current is sampled within the ripple in order to get an accurate value of the current. With a fixed switching frequency, and a load with an electric time constant much longer than the switching period, the current ripple passes its average value when the derivative of the carrier wave changes sign. This implies that only a limited number of switching frequencies could be used in the selected frequency range. As Figure 5.4 shows it is important that the number of switching periods between two sampling instants equals an integer number of half switching periods. In this way, it is the average current that is sampled and not the ripple.

The improvement when employing RPWM is particularly evident at the fundamental switching frequency, and this means that in the cases where an electromagnetic interference filter is present, it is possible to downsize the filter significantly when RPWM is used compared to when FPWM is used. This is a consequence of that it is the fundamental frequency electromagnetic interference that dictates the size of the filter. Since the side-band harmonics generated by RPWM are spread over a wide frequency range, the electromagnetic interference filter must be properly damped to avoid possible amplification of emission at the filter pole frequencies.

It has been shown [48] that the impacts of different PWM techniques on the efficiency of the drive are negligible.

One of the problems with the RPWM method is that a wide band excitation is also very likely to excite system resonances.

A common conception about implementing RPWM is that it requires a lot of computational power. This does not have to be the reality, since there are many different ways to implement RPWM. One example is the number of different frequencies that should be employed. Already when randomly choosing between only two different switching frequencies, a number of advantages are gained see Figure 5.5

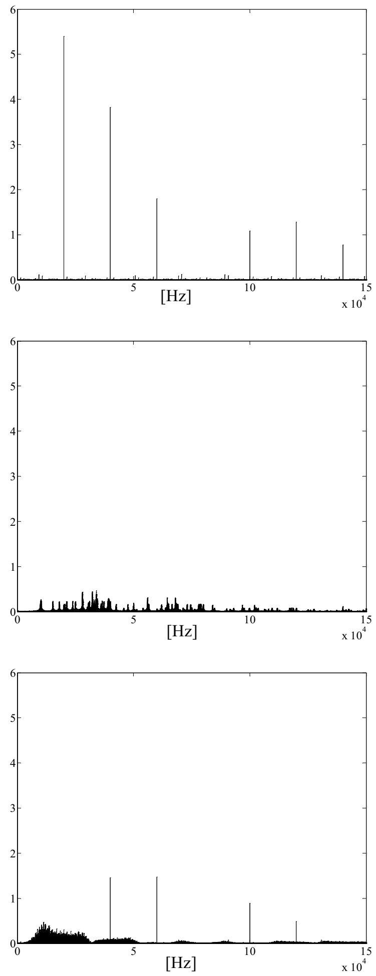

Figure 5.5. Upper: Fourier transform of a pulse width modulated signal with a fixed switching frequency. Middle: Fourier transform of a pulse width modulated signal with a randomly varying switching frequency between 10 kHz and 40 kHz. Lower: Fourier transform of a pulse width modulated signal with a switching frequency that randomly varies between two different switching frequencies (10 kHz and 20 kHz, coincident harmonics could be seen at for example 40 kHz and 60 kHz).

Another thing that speaks for RPWM is that the price for computational power gets more reasonable all the time, and even a rather complex RPWM does not require more than a relatively simple micro controller today.

Shielding

Shielding cables that emanates radiated emissions is often used as a method to suppress noise. When a shielded cable is placed inside a static electric field, induced charges are set up on the surface in order to reduce the internal field in the cable to zero [14]. If the electric field oscillates with a low frequency, the charges have to continually re-adjust themselves for each new field value, thereby causing currents to flow on the cable shield. At low frequencies, these currents will be distributed throughout the thickness of the conducting shield. The current on the internal surface will generate both magnetic and electric fields within the shielded cable.

When the frequency of the electric field increases, skin effects causes the current to become more and more localised on the outside surface of the shielding, thereby reducing the internal currents and the internal fields. This implies that the shielding effectiveness starts to increase as the skin effect starts to dominate. The skin effect is among others depending on the thickness of the shield.

Although shielding is an effective way of protecting against unwanted noise, the best way to protect against magnetic fields are to decrease the loop area, and shielding could for example do this.

The most important thing to remember when using shields for suppressing noise is the grounding of the shield. It has been shown in Chapter 2 that if a shield around a conductor is grounded at only one point, this will only protect against electric fields. In order to suppress the magnetic fields, the shield has to be grounded in both ends in order to cause a shield current to flow in the opposite direction of the current in the cable, thereby creating a magnetic field with an opposite sign.

A shield that is grounded in both ends could also reduce the size of the loop area as shown in Figure 5.6 [41]

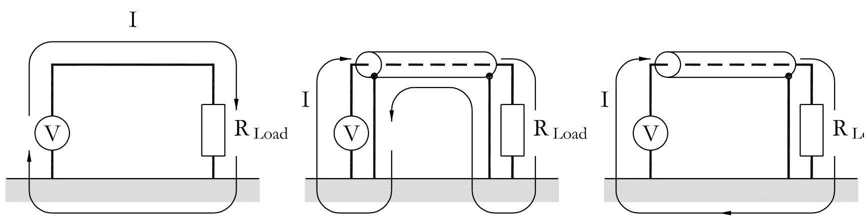

Figure 5.6. Different current loops when the cable is shielded or unshielded. The smallest current loop obtainable is when the shielded cable is grounded in both ends.

Since the grounding of a shield is such a vital issue, it is important that the ground points of the shield really remain grounded. This causes problems in the automotive industry where vibrations from the vehicle could cause the shield to break. Other drawbacks with shielded cables are that they are hard to mount (because of the grounding points), less flexible, heavier and more expensive.

Sometimes a braided shield is chosen since its flexibility, durability, strength and long flex life. These shields, however, provides a slightly reduced electric field shielding but a greatly reduced magnetic field shielding. The reason for this is that the braid distorts the uniformity of the shield current. Another drawback with braided shields becomes apparent at higher frequencies when the braid holes become large compared to a wavelength, and the effectiveness decreases further.

Current return through cables and cable layout

Since a large part of a vehicle is made out of metal, it is easy and convenient to use the body sheet metal as a ground plane and a current return path. This is not a proper way to ground since the body sheet metal has a high impedance due to the fact that the metal panels often are joined by spot welds or bolts with intervening layers of nonconducting corrosion inhibitor, or with nonconductive paints.

Another drawback when using the ground plane as a current return path is that the magnetic field around the lead-in conductor will increase. If the current return was through a cable and the cables were placed in close proximity, the magnetic fields from the cables would cancel each other see Chapter 2. If there is no cable for the current return, the magnetic flux density will become larger, see Chapter 2, equation (2.4). When the current return is through the ground plane, the area of the ground loop becomes a very efficient antenna. This is especially observable at lower frequencies when the skin effect causes the current to spread out to the corners of the metal.

To use a cable for ground return is a good way to protect against unwanted magnetic fields. When a cable is used for return, it is important to place the lead-in and return cable as close to each other as possible. The magnetic field in a point at a distance r from a cable in free space carrying the current I could be calculated as:

$$B = \mu \frac{I}{2\pi r} \tag{5.3}$$

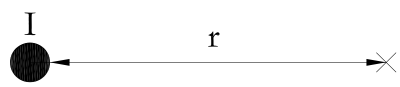

Figure 5.7. The flux density B at a distance r from a conductor carrying the current I.

Where B is the magnetic flux density, measured in tesla (T), r is the distance from the cable to the point where the flux is measured and μ is the relative permeability times the permeability for vacuum ($4\pi \cdot 10^{-7}$). When the two cables are placed close to each other with currents of opposite signs, the resulting flux density is then the difference between the two fields.

$$B = \mu \frac{i}{2\pi} \left(\frac{1}{r-a} - \frac{1}{r+a} \right) \tag{5.4}$$

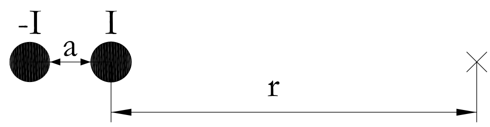

Figure 5.8. The flux density B at a distance r from two conductors carrying the same current, I, in opposite directions.

Where a is the distance between the two cables and r is the distance from the midpoint between the cables to the point where the flux is measured.

By comparing equation (5.3) and equation (5.4) it is possible to see that the magnetic field decreases when a cable is used for current return and when this cable is placed next to the lead-in cable.

Chapter 6

Experimental set-up

The experimental rig and corresponding measurement equipment is described in this chapter.

6.1 The experimental set-up

In order to test the proposed solutions from chapter 5, a test set-up is built. It consists of a car battery, two line impedance stabilization networks (LISNs), a drive circuit for MOSFETs, a converter, and a load, see Figure 6.1.

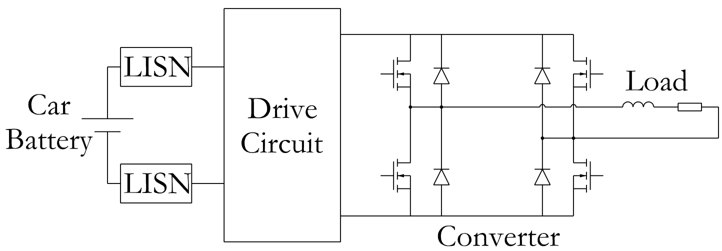

Figure 6.1. The laboratory test set-up

A ground plane is located under the set-up.

Ground plane

In a car, the body sheet metal is often used as a ground plane. This ground plane is far from being a perfect ground since it is often spot-welded or bolted together see Chapter 3. In order to recreate this grounding situation, two different ground planes are used. One ground plane is made of two 0.7 mm thick steel plates that are 0.5 m wide and 1 m long each. These two plates are spot-welded together in their short ends with the spots at distances of about 1.5 cm apart. This creates a ground plane that is 0.5 m wide and 1.98 m long. The other ground plane built is made up of 1.5 mm thick aluminium and has the same measures as the steel ground plane. The difference between the two ground planes is that the aluminium ground plane is not spot-welded but instead clenched together.

Both ground planes have connectors bolted to the ground plane at each end in order to facilitate current return through the ground plane.

Car battery

The car battery is an Exide Maxxima 900 with orbital wounded cells. It has a nominal voltage of 12 V, a capacity of 50 Ah, and a maximal starting current of 900 A.

Line Impedance Stabilization Network (LISN)

According to CISPR 25 [13], two LISNs are required in set-ups where the equipment under test is remotely grounded, i.e. has a power return line longer than 200 mm. Since this is the case in the test set-up, one LISN is used for the power supply line and one for the power return line. The LISNs used in the set-up are built according to the CISPR 25 standard. This standard prescribes that the inductance in the LISN should be 5 μH instead of 50 μH, which otherwise is the normal size of the inductance, in order to simulate a 5 m long cable between the battery and equipment under test. The circuit diagram of the LISN is shown in Figure 6.2.

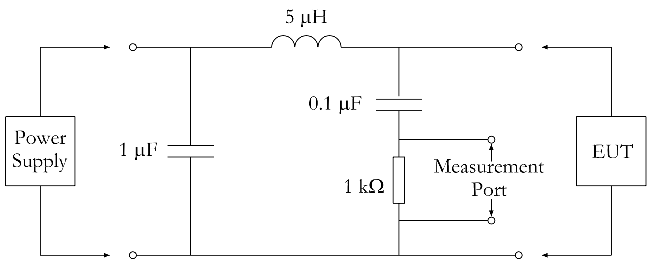

Figure 6.2. The circuit diagram for the LISNs that have been used in the tests, which are built according to the CISPR25 standard.

Both LISNs are mounted directly on the ground plane and their cases are bonded to the ground plane. The power supply return of the LISN is connected to the ground plane between the power supply and the LISNs.

The purpose of the LISNs is to have a defined impedance between the equipment under test and the car battery. The LISNs also provide an easily accessible port for measurements of conducted emissions.

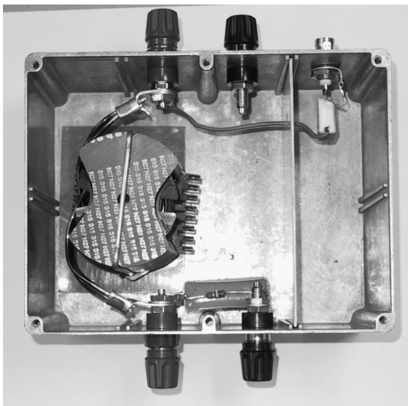

Figure 6.3. One of the LISNs used during the tests

The drive circuit

The high frequency full bridge FET driver from Harris Semiconductor, HIP4081A, is used to drive the MOSFETs. This driver provides the charging needed to turn-on or turn-off the MOSFETs. A separate modulator card creates the pulse pattern for the HIP4081A. On this modulator card, a combination of a micro controller and a programmable logic device (PLD) is used for pulse generation. The micro controller used is a PIC16F876 from Microchip and the PLD is a M4A5-128/64 from Lattice Semiconductor. The modulator card is driven from an external 24 V voltage source. A 40 MHz crystal provides the clock for the PLD, which also divides this signal down to a 20 MHz signal for the processor. Since the current is not sampled in this set-up, a fictitious sampling frequency of 500 Hz is assumed.

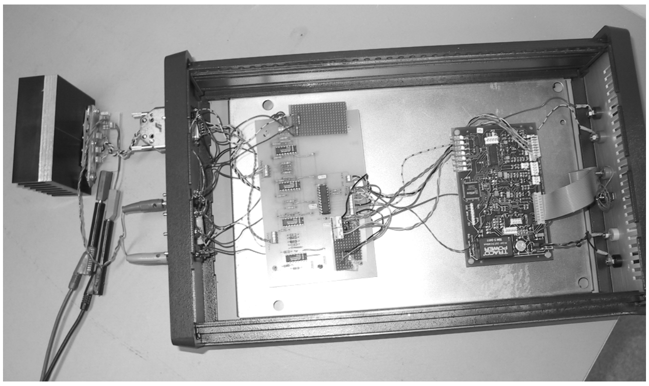

Figure 6.4. The modulator card to the right, the drive circuit in the middle and the power electronic converter outside the box, mounted on a big heat sink. The cables in the left corner are the cables from the power electronics to the load.

All the available switching frequencies are multiples of 250 Hz and are in the range from 10 kHz to 40 kHz. This results in 121 different switching frequencies that may be chosen. From these 121 different switching frequencies a list containing the number of clock pulses during one half period for a certain switching frequency is built up using the software Microsoft® Excel. The list is converted into a file format that makes it readable for a c-program, which generates a big if-statement. This if-statement is later used for determining the time for one half period, given a certain switching frequency. The solution with the if-statement is used since there is not very much data storage capacity in the micro controller, and to facilitate easy changes in the frequency list. Each time the micro controller should choose a new switching frequency it generates a random number in the range of 0-120, where number 0 corresponds to 10 kHz and 120 corresponds to 40 kHz. If the sampling time is assumed to be 500 Hz, this means that the chosen frequency should be employed for 2 ms. A switching frequency of 10 kHz has a period time of 0.1 ms. This implies that when 10 kHz is chosen as the switching frequency, 20 whole carrier wave periods or 40 half periods should pass before it is time to chose a new switching frequency. For a switching frequency of 40 kHz, the number of half periods between two sampling instants are 160. This means that when a switching frequency is selected by the help of the random number, a bias of 40 should be added to the random number in order to determine the number of half periods that should pass before the next frequency is selected. When it is time for the micro controller to chose a new switching frequency the generated random number together with the if-statement decides the time period that corresponds to a half period, and how many half periods there should pass before a new frequency is selected. The micro controller also checks the reference value for the output voltage in order to calculate reference values for the comparator. When these values are available, the micro controller sends them by a serial bus to a shift register. This shift register is read by the PLD, which has a counter that starts to increase, and thereby creates a triangular wave. In the PLD the comparison with the reference value is done, and the pulse pattern is created. As soon as the PLD has started with a new frequency, the micro controller starts to determine the switching frequency for the next sampling period. This operation takes less than 200 μs. Consequently the load on the processor is less than 200 μs / 2 ms = 10%. When the PLD has employed this switching frequency for one fictitious sampling period, it reads in the next switching frequency from the shift register.

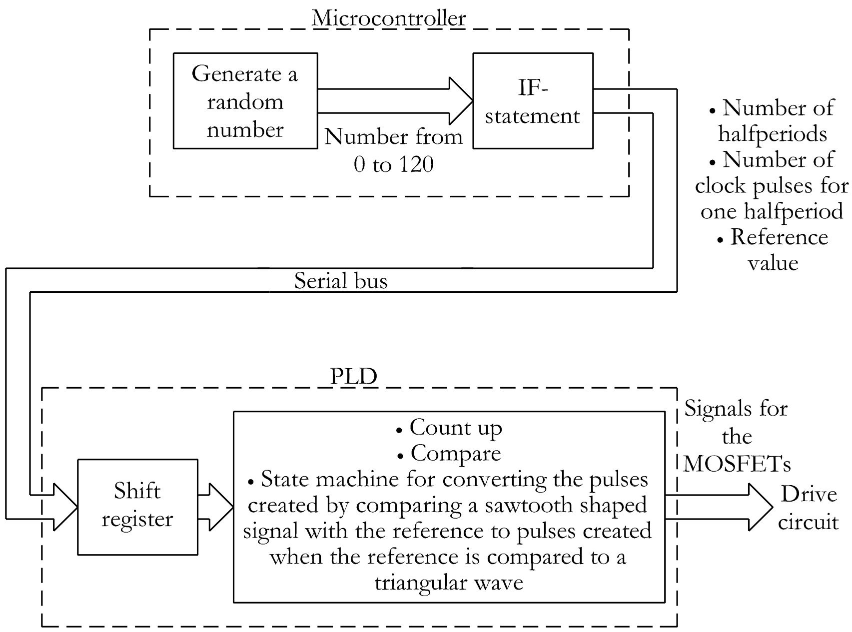

Figure 6.5. A flowchart for the modulator card.

In order to be able to change the gate resistance without having to change anything in the hardware set-up, one analogue switch is used for each MOSFET. This analogue switch of type MAX4662 connects the drive circuit with the gate of the MOSFET by a number of resistances, see Figure 6.6.

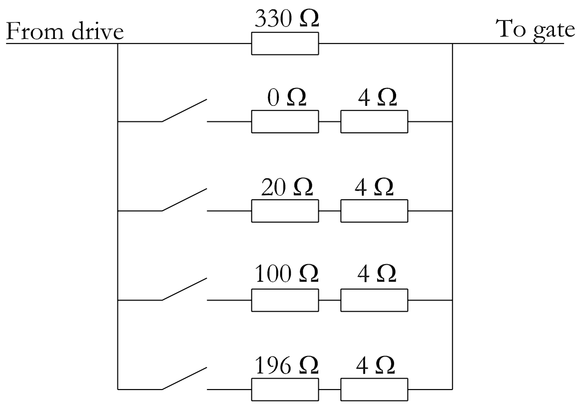

Figure 6.6. The analogue switch with the different gate resistances

It is possible to obtain values of the gate resistance from 3.2 Ω to 330 Ω by manually switching the different contacts.

Converter

The converter that is used in the test set-up consists of four MOSFETs of type IPU20N03L from infineon Technologies [23], and their ratings are V_{DS} = 30 V, I_D = 30 A and $R_{DS(on)}$ = 20 mΩ. The rise and fall times for these MOSFETs are typically 11 ns and 18 ns respectively. As freewheeling diodes, the internal diodes of the MOSFETs are used.

Load

The motor that is used as a model in the experiments is a small permanent magnet DC motor. It is designed for intermittent operation as a motor for an electrically operated window in a car. Since the purpose of this thesis has been to investigate the emissions emanating from the cables between the power electronics and the motor, a model of the motor has been used in order to avoid all the disturbances that arise from the brushes on the motor. The armature inductance of the motor is about 0.3 mH and the current when the motor is operating at nominal load with 9 V is around 2 A. To get a realistic model of this motor when it is operating at nominal speed and load, a 0.3 mH inductance and a resistance of 3 Ω is used instead of the motor.

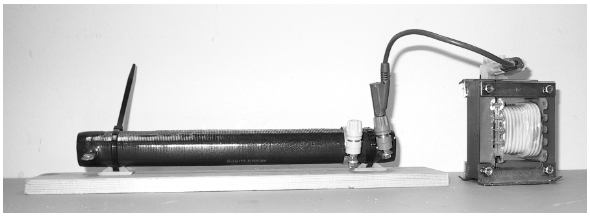

Figure 6.7. The load; to the left the 3 Ω resistance, to the right a 0,3 mH inductance

Cables

During the tests, different cables are used in order to determine the effect on the radiated emissions. In the reference set-up, 1.8 m long unshielded 2.5 mm^2 cables are used. In order to measure the effects of a shielded cable, these are later replaced by a 1.8 m long shielded two wire cable with a braided shield. The reason for choosing a braided shield is that it provides a more flexible and stronger shield than a solid shield. This makes it an often chosen shield in the rare cases when shielded cables are used in a vehicle. The two wires are 1.5 mm^2 and the shield is grounded in each end with a 2 cm long 1.5 mm^2.

6.2 Measurements of the magnetic field

Both radiated and some of the conducted emission tests are carried out in a semi-anechoic chamber at Delta Development Technology AB in Västerås, Sweden.

The chamber is a semi-anechoic chamber, meaning that the walls and the ceiling are covered with 2-2.5 m long absorbers in the form of wedges that are made of carbon-loaded foam.

Measurement set-up

The test set-up with the ground plane, the power electronics and the load are placed on a 20 cm thick cellular plastic layer on a wooden roller table. This roller table is placed in the semi-anechoic chamber on the floor ground plane. Since the roller table has a height of 80 cm and the cellular plastic layer is 20 cm thick, the distance between the set-up and the floor ground plane is 1 m. Two copper braids are mounted to the ground plane and also attached to the floor ground plane of the chamber.

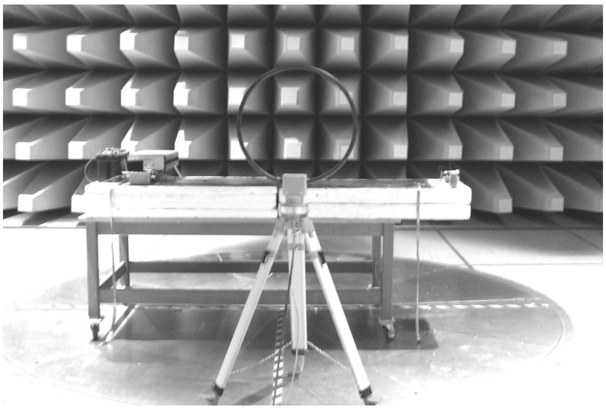

Figure 6.8. Picture of the laboratory test set-up. To the left, the battery, the LISNs and the power electronics, and to the right the load. In the foreground is the antenna, parallel to the cables between the power electronics and the load.

The load is placed on the cellular (foam) plastic, next to the ground plane. On the other end of the ground plane is the two LISNs and the power electronics placed. The ground of the two LISNs are mounted on to the groundplane, and the power electronics are placed on the LISNs in order to minimize the cable length between the LISNs and the power electronics. Next to the LISNs, on the cellular plastic is the battery placed. These two set-ups can be seen in Figure 6.9.

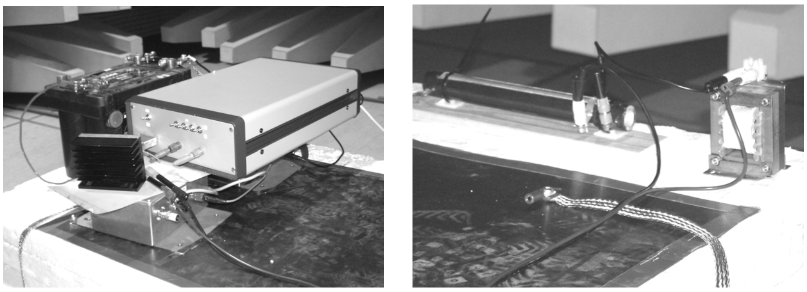

Figure 6.9. Left: In the background is the battery, placed next to the ground plane. In the foreground are the two LISNs, and placed on top of them is the power electronic converter. Right: Next to the ground plane is the load placed. The copper braid from the ground plane to the floor ground plane is visible in the foreground.

The antenna that is used for the radiated emission measurement is a magnetic field loop antenna of type HFH2-Z2 from Rohde & Schwarz. This loop antenna has a diameter of 50 cm and is placed at a distance of 1 m from the border of the ground plane, parallel to the cables between the power electronics and the load. These cables are placed 10 cm from the border of the ground plane.

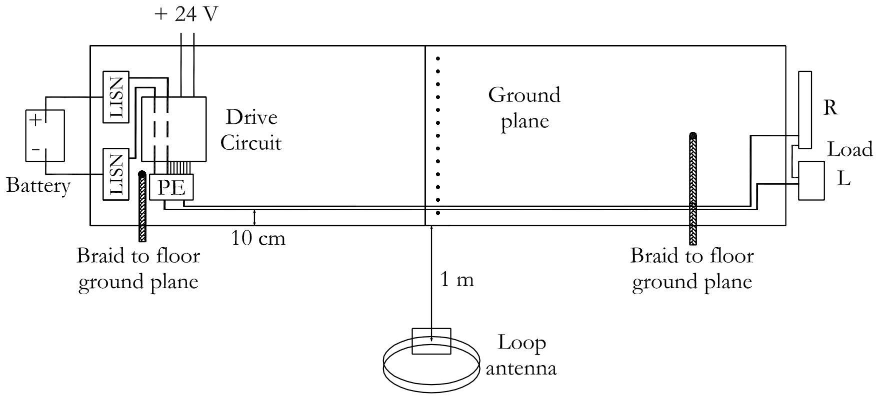

Figure 6.10. A schematic picture of the set-up

The 24 V power supply for the modulator card is placed outside the semi-anechoic chamber, and conducted into the chamber through filters. It is then recovered from an outlet in the floor under the floor ground plane.

Instrument

During the tests of the radiated and conducted emissions, the test receiver is placed in a control room outside the semi-anechoic chamber. The receiver is of type EMI Test Receiver 9 kHz – 30 MHz ESHS10 from Rohde & Schwarz. During the peak measurements with the receiver different bandwidths of the measurements are employed. The values of these can be seen in Table 6.1.

Table 6.1. The measurement bandwidths for different frequencies

Frequency range	Measurement bandwidth
9-150 kHz	200 Hz
150 kHz – 30 MHz	9 kHz

Quasi-peak measurements are also carried out at some specific frequencies in order to save time, and these are performed with a bandwidth of 200 Hz and a scan rate of 5 seconds.

6.3 Measurements of the load voltage spectrum

There are also some measurements of the load voltage spectrum, which are carried out in an ordinary unshielded laboratory. The measurements are conducted by measuring the output voltage from the converter. This voltage is then analysed with an oscilloscope and a spectrum analyser. The oscilloscope is a TDS 640A from Tektronix, and is used to analyse the output voltage in the A-band (see Appendix B). In order to investigate the frequency content of the output voltage, the oscilloscope shows a fast Fourier transform (FFT, see Appendix E) using a Hamming window, and a 12 point average. The B-band of CISPR 16 is investigated using a spectrum analyser called 5012-2 from Hameg. For this frequency range, the spectrum analyser has a bandwidth of 120 kHz.

Chapter 7

Experiments

This chapter describes the different test set-ups that are used in the experiments and the measured results from the measurements. After the introduction a reference set-up is described. The results from the experiments made on this set-up are used for comparison with the results from the other set-ups.

7.1 Introduction

During the experiments, the laboratory test set-up described in chapter 6 is used. The different cases that are tested are

- randomly varying switching frequency
- increased gate resistance
- different cable layouts, such as spacing or shielding the cables and to use the ground plane as a current return.
- characteristics from the two different ground planes

In the tests, a duty-cycle of 75% is employed. This results in a load current of 2 A. In order to evaluate the differences in the various set-ups, the radiated magnetic field, the spectrum of the load voltage and in some cases also the input voltage for the power electronic converter are measured. The radiated magnetic field is measured using a loop antenna and a peak measuring EMI receiver (Chapter 6), with a bandwidth of 200 Hz (for 9-150 kHz) and 9 kHz (for 0.15-30 MHz), respectively. The results from the measurements of the radiated emissions are presented in a peak dBμA/m scale. To be able to measure the spectrum of the voltage over the load, both an oscilloscope and a spectrum analyser are employed. The oscilloscope covers the A-band of CISPR 16 (9-150 kHz) and makes a fast Fourier transform of the signal using a Hamming window with an average of 12. These results are displayed in a dBV RMS scale. For the B-band of CISPR 16 (0.15-30 MHz) a spectrum analyser is used. This spectrum analyser is measuring the voltage spectrum using an average function and a maximum hold/peak function, both displayed in a dBμV scale. The bandwidth of the spectrum analyser is set to 120 kHz. The input voltage for the power electronic converter is measured using the LISN connected to the battery negative. This signal, which corresponds to the conducted emissions, is analysed with the same EMI receiver as in the case with the radiated magnetic field, and the results are displayed in dBμV.

The purpose of these measurements is to get a conception of how different design parameters affect the emissions from a pulse width modulated system. Since none of the measurements are carried out in a real vehicle, the focus is on the relative differences between two measurements rather than the exact values from the results.

As a reference set-up in the tests, a set-up with low gate resistance, fix switching frequency of 20 kHz, and two 1.8 m long unshielded cables is employed.

7.2 Reference set-up

The reference set-up is intended to represent a situation in a vehicle where there are long unshielded lead-in and return cables between the power electronic converter and the load. A fixed switching frequency of 20 kHz is employed, and a gate resistor of 3.2 Ω is used. The 1.8 m long cables are placed on the steel ground plane, next to each other. An explanatory sketch of the set-up is shown in Figure 7.2.

Figure 7.1. The reference set-up.

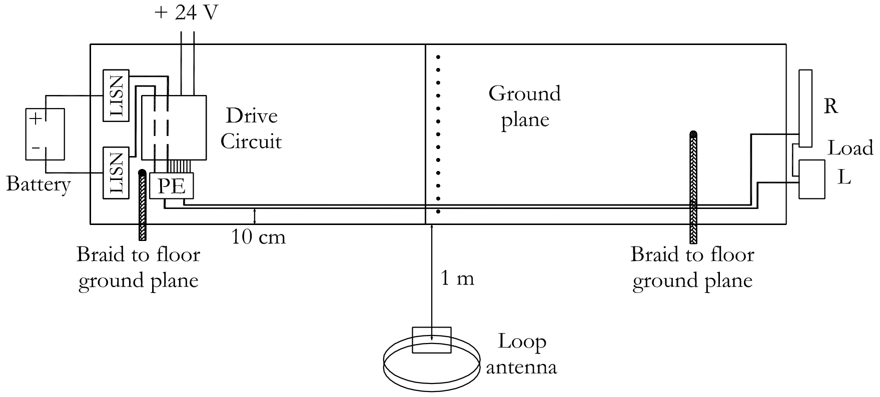

Figure 7.2. A sketch of the reference set-up. The battery, the LISNs, the drive circuit for the MOSFETs and the power electronic (PE) converter are located to the left. The load is placed to the right. Between the load and the power electronics there are two 1.8 m long unshielded cables located next to each other on the ground plane.

The results from the measurements of the radiated magnetic field and the input voltage for the converter are shown in Figure 7.3.

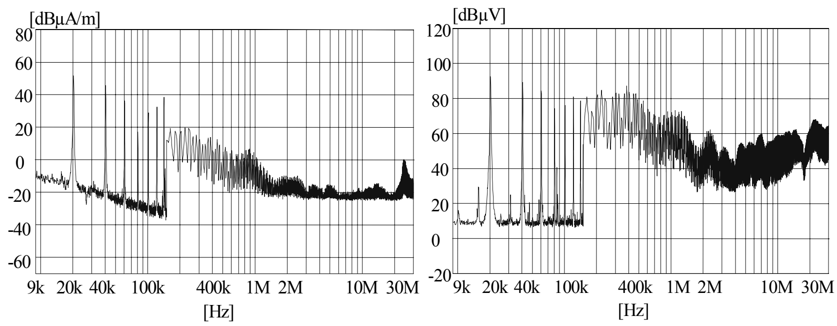

Figure 7.3. Left: The results from the measurement of the radiated magnetic field of the reference set-up. Right: Results from the measurements of the input voltage for the converter through one of the LISNs.

The switching frequency and its harmonics are easily visible in the measurements. There is also a peak at about 24 MHz. This peak could be due to oscillations between the input capacitance of the MOSFET and the inductance in the wire to the gate. The input capacitance for the IPU20N03L MOSFET is typically 530 pF, which gives a value of the inductance as in

$$\omega_{osc} = 2 \cdot \pi \cdot f_{osc} = \frac{1}{\sqrt{L \cdot C_{input}}} \quad \Rightarrow$$
$$L = \frac{1}{C_{input} \cdot (2 \cdot \pi \cdot f_{osc})^2} = \frac{1}{530 \cdot 10^{-12} \cdot (2 \cdot \pi \cdot 24 \cdot 10^6)^2} = 83 \text{ nH} \tag{7.1}$$

If the rule of thumb that a wire length of one meter corresponds to 1 μH, it implies that the length of the wire should be about 8 cm according to equation (7.1). This matches the length of the actual wire needed, since the power electronic converter is not built into the box used for the drive circuit and the modulator card.

A spectrum analysis of the voltage over the load is measured at the output of the converter. The results are shown in Figure 7.4.

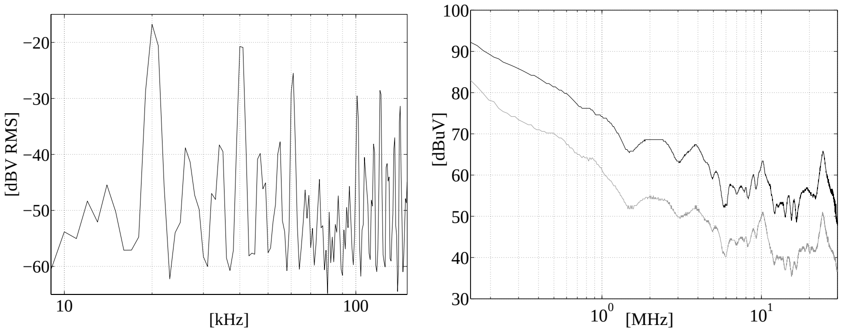

Figure 7.4. Left: A 12 point average FFT analysis of the voltage for frequencies from 9 kHz to 150 kHz. Right: A peak (black) and average (grey) frequency analysis of the signal up to 30 MHz.

As in the measurements of the radiated magnetic field the contributions from the switching frequency are perceptible in the voltage over the load. The peak at 24 MHz is also visible in the load voltage.

7.3 Increasing the turn-on and turn-off times

In the reference case of the set-up, a gate resistor of 3.2 Ω is employed. In this set-up the gate resistance is increased to 330 Ω by using the analogue switches. This is the only change in the set-up, which means that the rest of the hardware is not modified. Increasing the gate resistance of the switching MOSFET implies that the switching times increase as well as the switching losses. In order to examine the switching losses and the rise and fall times for the voltage, the drain-source voltage and the drain current was measured using an oscilloscope and a current probe. For this set-up the rise and fall times of the drain-source voltage of the MOSFET changed according to Table 7.1.

Table 7.1. Rise and fall times of the drain-source voltage.

Gate resistance	Rise time	Fall time
3.2 Ω	545 ns	38 ns
330 Ω	1.42 μs	94 ns

The voltage pulses are then similar to Figure 7.5.

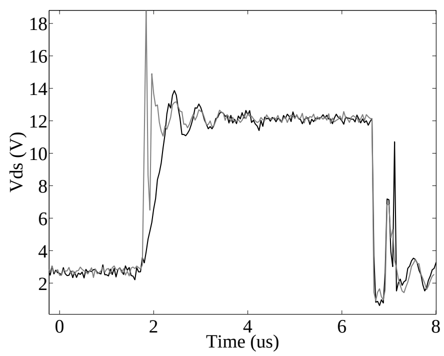

Figure 7.5. Drain-source voltage at MOSFET turn-off and turn-on with 3.2 Ω gate resistance (grey) and 330 Ω gate resistance (black).

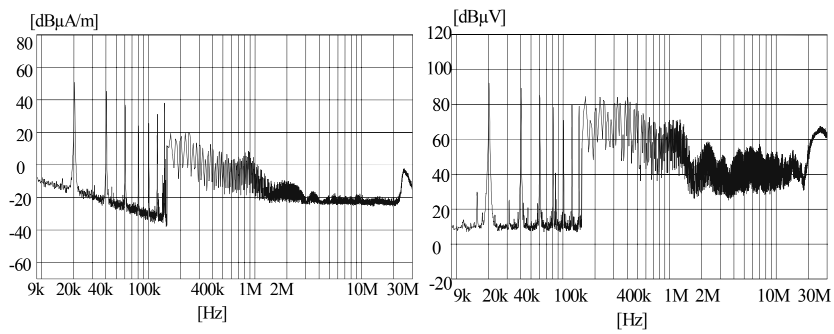

Figure 7.6. Left: Results from the measurement of the radiated magnetic field when a gate resistance of 330 Ω is employed. Right: Results from the measurements of the input voltage for the converter through one of the LISNs.

The measurements of the radiated field when a high gate resistor is used shows that the effect from the increase in rise and fall times are most noticeable in the frequency range above 3 MHz when compared with the reference set-up.

The total losses for one switching period are 4.0 W with a 3.2 Ω gate resistance and 4.3 W with 330 Ω gate resistance.

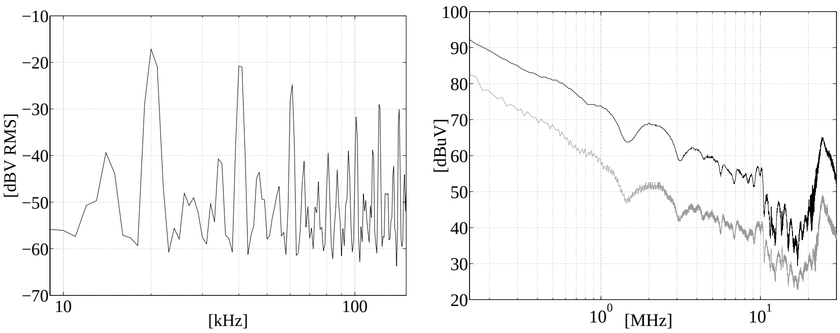

Figure 7.7. The voltage over the load when a gate resistance of 330 Ω is employed. Left: A 12 point average FFT analysis of the voltage for frequencies from 9 kHz to 150 kHz. Right: A peak (black) and average (grey) frequency analysis of the signal up to 30 MHz.

When comparing the frequency spectrum of the voltage pulses displayed in Figure 7.5 the same trends with the largest attenuation above 3 MHz as in the radiated field measurements occur. The envelope of the spectrum for a Fourier series of a trapezoidal wave (i.e. a square wave with finite time derivatives of the pulse edges) decreases at a rate of 20 dB/decade up to a point proportional to the rise time of the wave, where the envelope starts to decrease with a rate of 40 dB/decade. In cases where the rise and fall times of the pulses differ, the shortest time should be used in order to find this point. For the set-up with a high gate resistance, the shortest time is the fall time, 94 ns, which gives a value of the frequency where the spectrum should start to decrease of:

$$f = \frac{1}{\pi \cdot t_r} = \frac{1}{\pi \cdot 94 \cdot 10^{-9}} = 3.4 \text{ MHz} \tag{7.2}$$

The calculated frequency of 3.4 MHz agrees with the observed attenuation in the measurements of the radiated magnetic field and the voltage spectrum.

7.4 Randomly varying switching frequency

In the set-up where the effects from a randomly varying switching frequency (RPWM) are investigated, only a change in the software for the modulator card is done. This implies that the set-up of the hardware is similar to the set-up in the reference case, meaning that a low gate resistance (3.2 Ω) and two long unshielded cables between the power electronic converter and the load are used, according to Figure 7.2.

In the case of randomly varying switching frequency, 121 different switching frequencies between 10 kHz - 40 kHz are employed and altered with a frequency of 500 Hz. This implies that the frequency spectrum should be lower and more broadband compared to when fixed switching frequency is employed. The radiated electromagnetic field is shown in Figure 7.8.

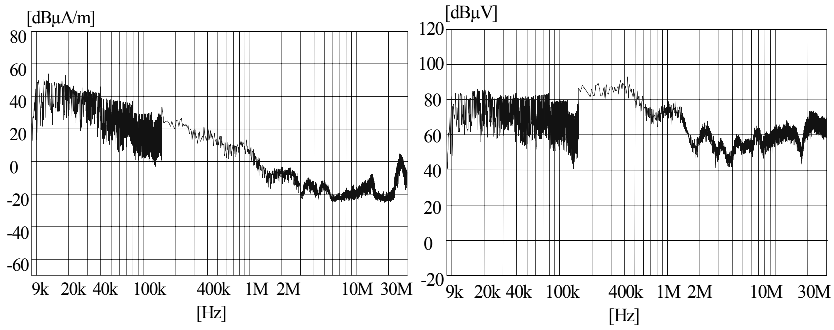

Figure 7.8. Left: Results from the measurement of the radiated magnetic field when a randomly varying switching frequency between 10 kHz and 40 kHz is used instead of a fix switching frequency. Right: Results from the measurements of the input voltage for the power electronic converter through one of the LISNs.

From the measurement results of the radiated magnetic field in Figure 7.8 it is evident that this method has only a minor effect on the emissions in the frequency range of the fundamentals (10 kHz - 40 kHz). For the higher frequency ranges, it appears as if the RPWM method causes even more disturbances than the method with a fixed switching frequency. These results are mainly a consequence of the measurement technique that is employed. In an EMI measurement receiver as the ESHS10, the incoming measurement signal is filtered through a band pass filter and then mixed with a local oscillator, which is stepped with a frequency of 6 kHz. After a number of mixing stages, the signal is filtered again through a band pass filter with a resolution bandwidth of 9 kHz (this resolution bandwidth is defined as the area of the filter down to –6 dB). A frequency close to the boundary of the band pass filter is then measured with a higher attenuation than a frequency present in the centre of the filter. This implies that when a fixed switching frequency is employed, the switching frequency and its harmonics could result in lower measurement values than the actual levels due to the effect of the band pass filter. In the case with the randomly varying switching frequency, the different switching frequencies are placed close to each other, which implies that there is always at least one switching frequency or switching frequency harmonic present at the centre of the filter. This could be an explanation of the high measurement values for the RPWM method. In Figure 7.8 a peak measurement of the magnetic field is shown, implying that for each frequency the peak value is displayed. This measurement method is not representing the energy emitted at each frequency. A quasi-peak measurement of the radiated magnetic field would give a more correct representation of the effects from RPWM see Chapter 2. Due to the considerable amount of time required for each quasi-peak measurement, it is only performed at some relevant frequencies, see Table 7.2.

Table 7.2. Peak and quasi-peak values of the radiated magnetic field at some relevant frequencies.

	Fix PWM (dBμA/m)		RPWM (dBμA/m)	
Frequency (kHz)	Peak	Quasi-Peak	Peak	Quasi-Peak
20	52.1	51.5	43.8	37.5
40	47.2	46.9	39.0	31.9
60	39.7	38.5	39.5	24.1
120	31.7	32.2	39.4	24.0

It can be seen in Table 7.2 that in the case of randomly varying switching frequency there are noticeable differences between the peak and the quasi-peak measurements. This is derived from the fact that each switching frequency is only employed for a short time interval before it is altered and replaced with a new switching frequency. Since the quasi-peak method dwells on each frequency for a longer time interval than the peak method, the quasi-peak method will take into account that for RPWM, there are less energy for the different frequencies.

For the higher frequency ranges in Figure 7.8 the peak measurements lead to the conclusion that the RPWM method results in higher emissions than the method with fixed switching frequency. This is due to both the fact that the measurements are performed using a peak analyser and superposition of the harmonics of the different switching frequencies.

In the frequency spectrum of the voltage over the load, it is possible to see a considerable decrease in the low frequency range (A-band), see Figure 7.9. Here, the oscilloscope presents an average of the Fourier transform, and it shows that the noise in the voltage is more broadband when RPWM is employed. In the B-band of the frequency, no effects are noticeable since the RPWM method is most effective in the frequency ranges around the fundamentals.

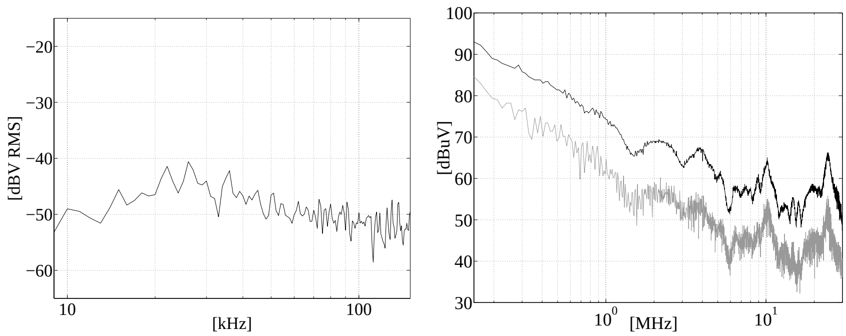

Figure 7.9. The voltage over the load when a randomly varying switching frequency is employed. Left: A 12 point average FFT analysis of the voltage for frequencies from 9 kHz to 150 kHz. Right: A peak (black) and average (grey) frequency analysis of the signal up to 30 MHz.

7.5 Shielding

When the effects from using a shielded cable should be examined, the set-up is almost equal to the reference set-up except for the cables between the power electronic converter and the load. In the reference set-up, two unshielded 1.8 m long cables with a cross section area of 2.5 mm^2 are used. The shielded cables used in this set-up are two 1.8 m long, 1.5 mm^2 cables in a braided shield. The shield is mounted to the ground plane via 2 cm long 1.5 mm^2 cables in both ends. The shielded cables are placed on the ground plane (Figure 7.10).

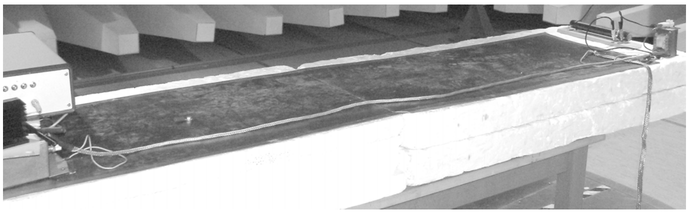

Figure 7.10. A picture of the set-up when shielded cables are used between the converter and the load.

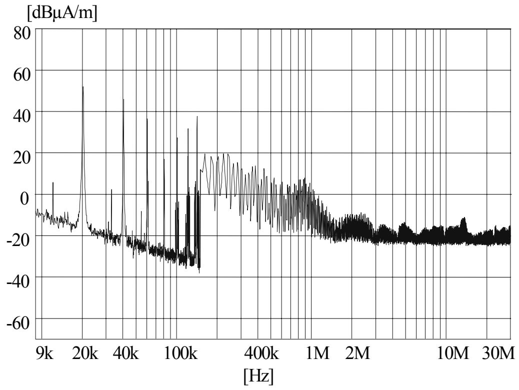

Figure 7.11. The results from the measurement of the radiated magnetic field when shielded cables between the power electronic converter and the load are used.

The results from the radiated magnetic field measurements show some improvements on the radiated emissions in the higher frequency ranges (above 15 MHz). For the frequency 30 MHz, the wavelength is 10 m. This implies that a quarter of a wavelength (2.5 m) approaches the length of the cables. This means that in the higher frequency areas the cables work as a good antenna. One explanation for the poor results is that the shield on the cable is a braided shield and that this is not the best type of shield for magnetic radiation, see Chapter 5, since the braid distorts the uniformity of the shield current. Also the different geometries of the shielded cables influence the results, which is clear from looking at the frequency spectrum of the load voltage. In this case, two set-ups are compared; one where both ends are mounted to ground, and one where none of the ends are mounted to ground. Both set-ups show the same characteristics that are different from the cases with the shielded cables.

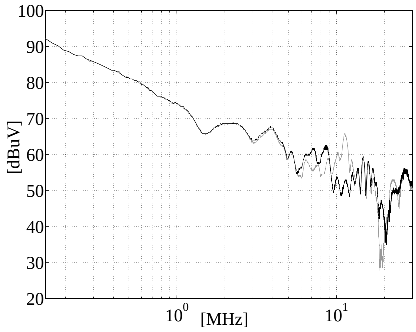

Figure 7.12. Frequency spectrum for the load voltage where ungrounded shielded cables (grey) are compared with shielded cables grounded at both ends (black).

The shielded cables do not affect the frequency spectrum for the A-band while the spectrum for the higher frequency ranges is reduced for frequencies from 15 MHz to 29 MHz, see Figure 7.13. This decrease is not expected and is most likely caused by the difference in cable dimensions. At the end of the B-band (around 30 MHz) the voltage frequency spectrum is higher in the case of shielded cables than in the case of unshielded cables (the reference case), which implies that the decrease in the measured radiated magnetic field in this frequency area is caused by the shield.

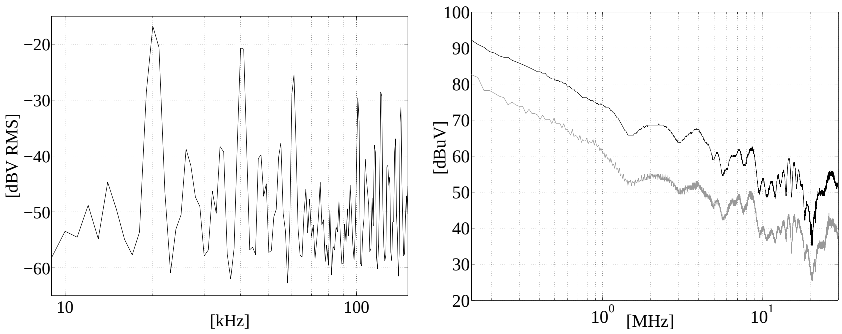

Figure 7.13. The voltage over the load when shielded cables between the power electronic converter and the load are employed. Left: A 12 point average FFT analysis of the voltage for frequencies from 9 kHz to 150 kHz. Right: A peak (black) and average (grey) frequency analysis of the signal up to 30 MHz.

7.6 Cable layout

In the tests different cable layouts are evaluated in order to determine their impact. In the first test, the cables in the reference set-up are moved horizontally apart from each other by 10 cm see Figure 7.14.

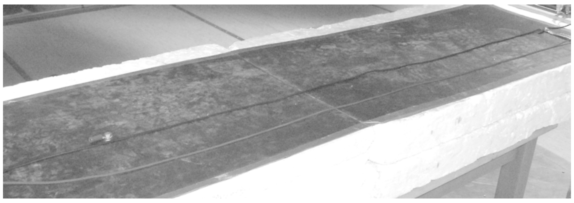

Figure 7.14. The set-up when the cables are separated and placed on the ground plane.

When the cables are separated on the ground plane, the results from the measurements, both radiated and voltage frequency spectrum, show no effects from this separation, see Figure 7.15, Figure 7.16.

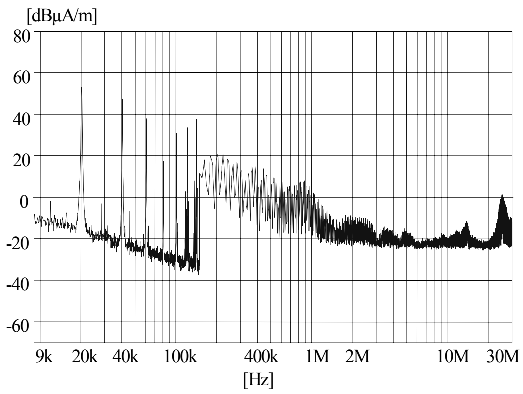

Figure 7.15. The results from the measurement of the radiated magnetic emissions when the cables between the converter and the load are placed on the ground plane and separated by 10 cm.

The separation of the cables is expected to cause more radiated magnetic emissions since the area of the current loop is increased, see equation (5.4). This is however not noticeable, due to the orientation of the cables and the antenna. If the antenna is to be placed above the set-up, with the normal from the antenna pointing in the same direction as the normal from the created current loop, the results would have been different.

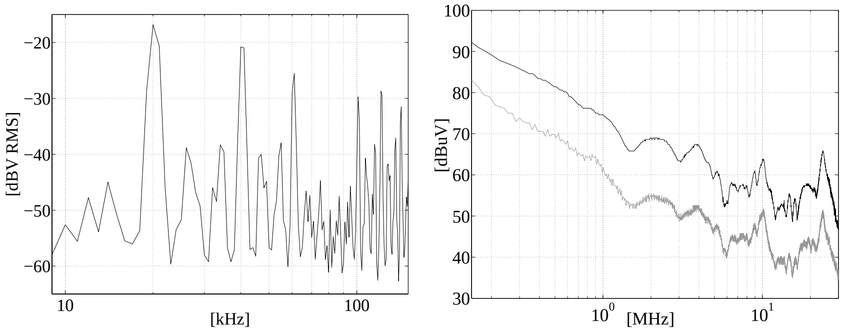

Figure 7.16. The voltage over the load when the cables between the power electronic converter and the load are separated on the ground plane. Left: A 12 point average FFT analysis of the voltage for frequencies from 9 kHz to 150 kHz. Right: A peak (black) and average (grey) frequency analysis of the signal up to 30 MHz.

In the next set-up, both cables were raised together to a position 11 cm above the ground plane by a long piece of cellular plastic, see Figure 7.17.

Figure 7.17. Both cables between the converter and the load are raised 11 cm above the ground plane. The cables are placed next to each other.

This set-up should provide a larger loop area for the common-mode current, which is returning through the ground plane since the cables are raised. This increase in distance between the cables and the ground plane decreases the size of the parasitic capacitance between the ground plane and the cables, thereby decreasing the common-mode current. The results from the measurements of the radiated magnetic field are shown in Figure 7.18.

As seen in Figure 7.18 the radiated magnetic field increases in the higher frequency range when the cables are raised. This is due to the increase of the loop area for the antenna formed by the common mode current in the cables and the parasitic capacitance to ground.

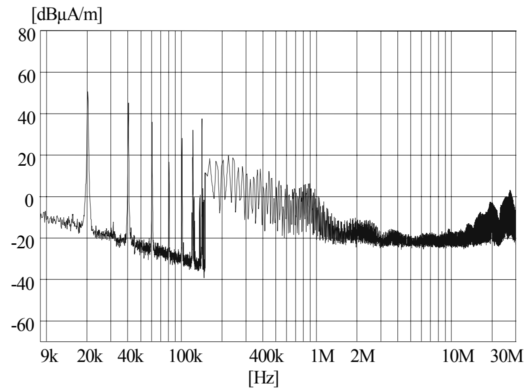

Figure 7.18. The results from the measurement of the radiated magnetic emissions when the cables between the power electronic converter and the load are placed 11 cm above the ground plane.

The load voltage frequency spectrum appears to be almost the same as in the reference case except for some of the frequency peaks that are shifted, since the resonance of the system is changed due to the reduction of parasitic elements.

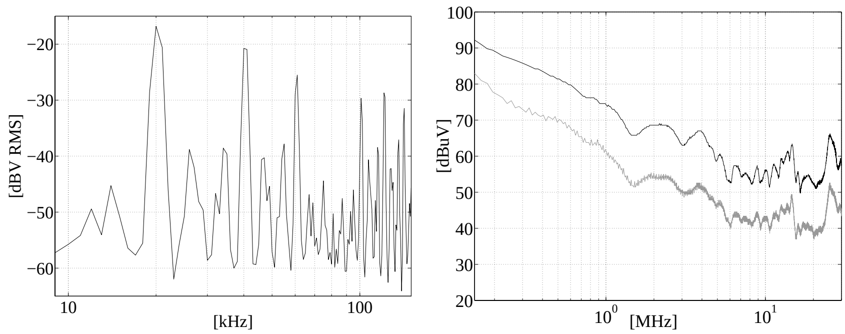

Figure 7.19. The voltage over the load when the cables are raised 11 cm above the ground plane. Left: A 12 point average FFT analysis of the voltage for frequencies from 9 kHz to 150 kHz. Right: A peak (black) and average (grey) frequency analysis of the signal up to 30 MHz.

In the reference set-up, there is one peak at 11 MHz. This peak is shifted to about 14.5 MHz in this set-up, which means that there is a ratio of about 1.3 between the two frequencies. Since this peak arise from a resonance in the system, it is interesting to investigate what has happened with the capacitance of the circuit when the cables are raised above the ground plane. The capacitance between two parallel conductors and ground could be determined by using equation (7.3):

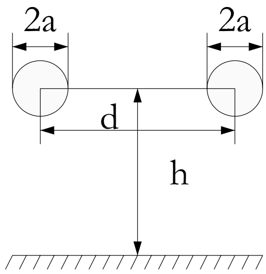

Figure 7.20. Two wires above a ground plane.

$$C = \frac{\pi \cdot \varepsilon_0 \cdot l}{\ln \dfrac{2 \cdot h \cdot d}{a\sqrt{4 \cdot h^2 + d^2}}} \tag{7.3}$$

where ε_0 is the permittivity of free space, equal to $8.85 \cdot 10^{-12}$ F/m. In both cases the radius of the conductors (a) are 0.9 mm, the distance between the conductors 2 mm and the length of the conductors 1.8 m. The heights are 1 mm and 110 mm above ground respectively. These values give the following capacitances:

Table 7.3. Parasitic capacitances for conductors on and above the ground plane.

Height between conductors and ground plane [mm]	Capacitance [pF]
1	111
110	62.7

The resonance frequencies are then compared as

$$\frac{f_{\text{above ground}}}{f_{\text{on ground}}} = \frac{1}{2 \cdot \pi \sqrt{L \cdot C_{\text{above ground}}}} \cdot \frac{2 \cdot \pi \sqrt{L \cdot C_{\text{on ground}}}}{1} = \sqrt{\frac{111}{62.7}} = 1.3 \tag{7.4}$$

which shows that the calculated ratio between the frequencies are 1.3, which agrees with the measured results.

In order to test the contribution from the differential mode radiation, the cables on the cellular plastic above the ground planes are separated 10 cm in the next set-up, see Figure 7.21.

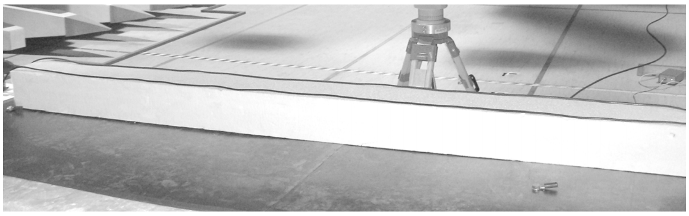

Figure 7.21. A picture of the set-up where the cables are placed 11 cm above the ground plane and are separated by 10 cm.

In the measurements of radiated emissions for this set-up, the effect from the differential mode radiation is apparent already in the low frequency range where the magnitude of the fundamental and the first harmonics have increased. The ground plane together with the two cables now works as a distributed antenna, for both common mode and differential mode radiation.

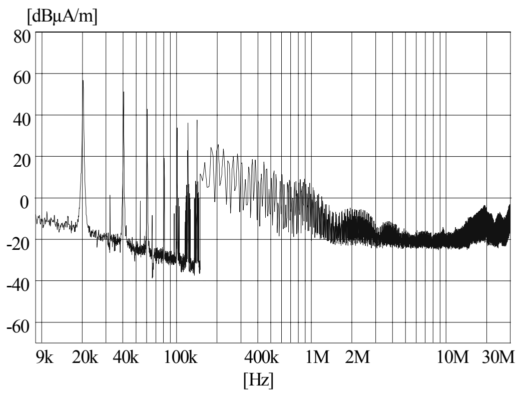

Figure 7.22. The results from the measurement of the radiated magnetic emissions when the cables between the power electronic converter and the load are placed 11 cm above the ground plane and separated by 10 cm.

The load voltage frequency spectrum for the set-up with the cables 10 cm apart looks similar to the spectrum in Figure 7.19 except for the attenuation of the frequency peak at 26 MHz.

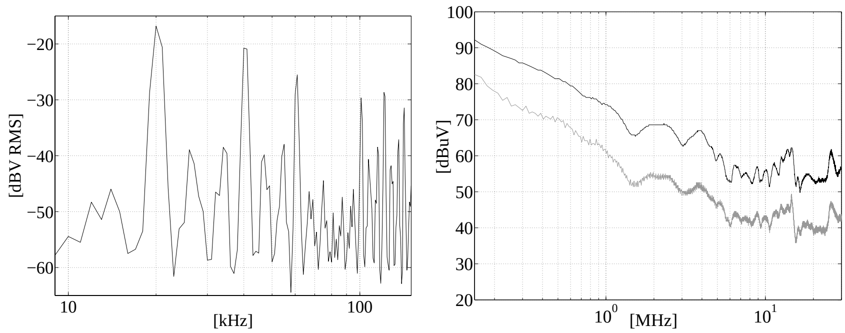

Figure 7.23. The voltage over the load when the cables are raised 11 cm above the ground plane and separated by 10 cm. Left: A 12 point average FFT analysis of the voltage for frequencies from 9 kHz to 150 kHz. Right: A peak (black) and average (grey) frequency analysis of the signal up to 30 MHz.

There is a difference in the attenuation of this resonance frequency of about 3 dB with the lower value when the conductors are separated. In this case, it is the damping that differs between the two set-ups. The damping at a resonance frequency is derived by

$$\zeta = \frac{R}{2} \cdot \sqrt{\frac{C}{L}} \tag{7.5}$$

where ζ is the damping, R the resistance, C the capacitance between the two conuctors and L the inductance of the circuit. Since the same conductors are used in both set-ups, the resistance is constant. The capacitances for the different set-ups are calculated as equation (7.3) and will thus be 62.7 pF in the case where the conductors are close to each other and 10.8 pF in the case where they are separated. The mutual inductance between the two conductors is also affected when they are separated, and that is calculated by the following equation

$$M = 9.84 \cdot 10^{-8} \cdot \ln\left(1 + \frac{(2h)^2}{d^2}\right) \tag{7.6}$$

where h is the height above the ground plane and d the distance between the conductors. For these two set-ups, the values of the total inductance (mutual and internal, internal equal to 1 μH/m) are 2.73 μH (for the set-up with the conductors next to each other) and 1.97 μH for the set-up with the conductors separated. Comparing the damping for these two set-ups gives a difference like

$$\frac{\zeta_{close}}{\zeta_{apart}} = \sqrt{\frac{C_{close} \cdot L_{apart}}{C_{apart} \cdot L_{close}}} = 2.05 \Rightarrow 10 \cdot \log 2.05 = 3.1\,\mathrm{dB} \qquad (7.7)$$

This results shows conformity with the measured results.

Since a vehicle manufacturer has the whole body sheet metal available as a ground plane, it is often used as a return path for the current. In order to investigate the effects on the emissions from this solution, the steel ground plane is used as a current return path in this set-up. The lead-in cable is one of the long unshielded cables from the reference case that is placed on the ground plane.

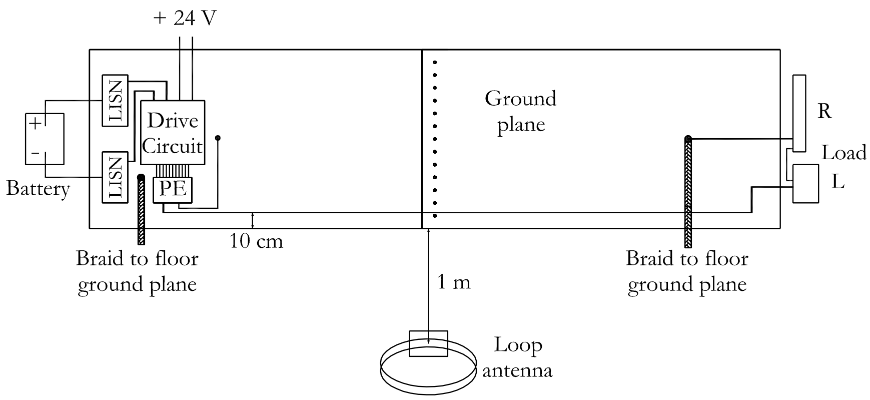

Figure 7.24. The set-up with the current return through the ground plane.

Since the ground plane is used as a current return path, it is possible for the return current to spread out over the ground plane due to the skin effect. This makes the set-up an efficient antenna for differential mode radiation, which can be seen in Figure 7.25.

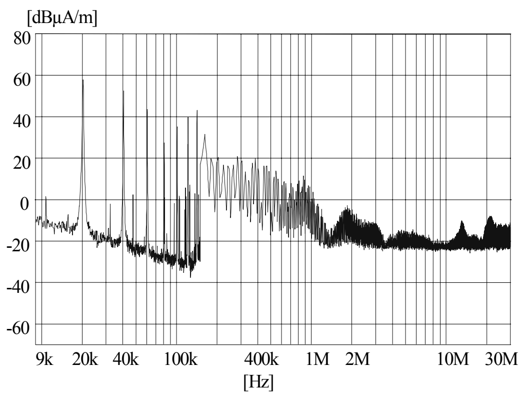

Figure 7.25. The results from the measurement of the radiated magnetic emissions when the current return is through the ground plane and the lead-in conductor is placed on the ground plane.

In this set-up, the frequency spectrum for the output voltage above 2 MHz is affected, which can be seen in Figure 7.25 and Figure 7.26. The frequency peak at 24 MHz is attenuated, and this could be due to interference phenomenon since the peak is still present in the load voltage frequency spectrum, see Figure 7.26.

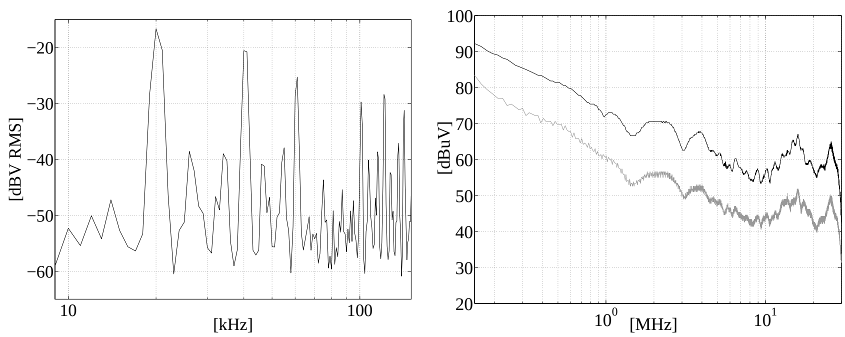

Figure 7.26. The voltage over the load when the current return is through the ground plane and the lead in cable is placed on the ground plane. Left: A 12 point average FFT analysis of the voltage for frequencies from 9 kHz to 150 kHz. Right: A peak (black) and average (grey) frequency analysis of the signal up to 30 MHz.

Raising the lead-in cable above the ground plane when the ground plane is used as a current return path creates a loop antenna with a large area. This antenna radiates mainly differential mode disturbances, see Figure 7.27.

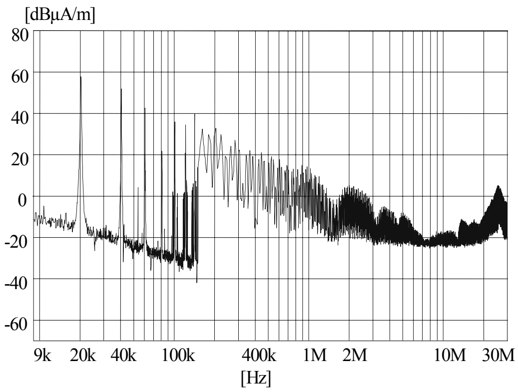

Figure 7.27. The results from the measurement of the radiated magnetic emissions when the current return is through the ground plane and the lead-in conductor is placed 11 cm above the ground plane.

The load voltage frequency spectrum for this set-up shows that the harmonic content of the signal has not increased, but some of the frequency peaks is shifted once again due to the change in the parasitic components.

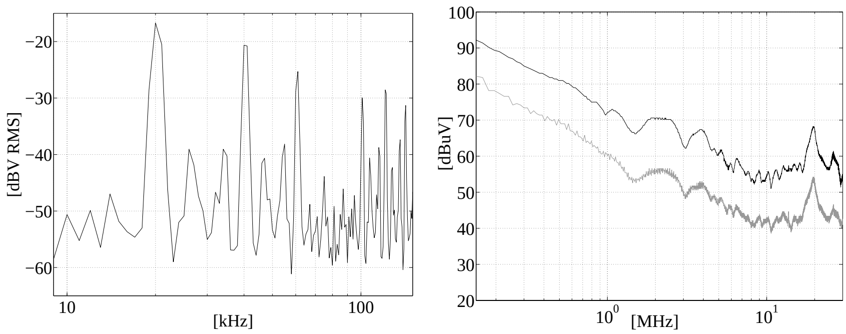

Figure 7.28. The voltage over the load when the current return is through the ground plane and the lead in cable is placed 11 cm above the ground plane. Left: A 12 point average FFT analysis of the voltage for frequencies from 9 kHz to 150 kHz. Right: A peak (black) and average (grey) frequency analysis of the signal up to 30 MHz.

7.7 Aluminium ground plane

Since some modern vehicles today have a body sheet metal that is made of aluminium, a set-up with an aluminium ground plane is used during some of the tests, see Figure 7.29.

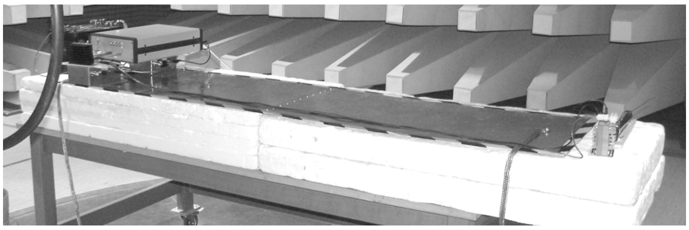

Figure 7.29. A picture of the set-up when the aluminium ground plane is used.

In all the different set-ups where the aluminium ground plane is employed, the set-up is equal to the corresponding set-up with the steel ground plane.

The first set-up to be tested with the aluminium ground plane is the reference set-up with a fixed switching frequency of 20 kHz, a low gate resistance (3.2 Ω) and two long (1.8 m) unshielded cables placed on the ground plane between the converter and the load.

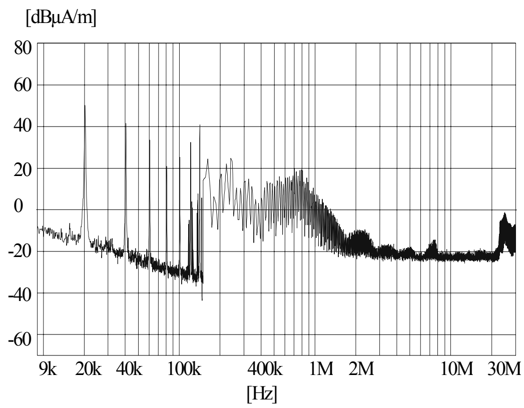

Figure 7.30. The results from the measurement of the radiated magnetic emissions when the aluminium ground plane is used in a set-up similar to the reference set-up.

As seen in Figure 7.30 the radiated magnetic emissions are somewhat higher in the set-up with the aluminium ground plane. This is due to the lower resistivity in the material, which offers less damping in the resonance circuit.

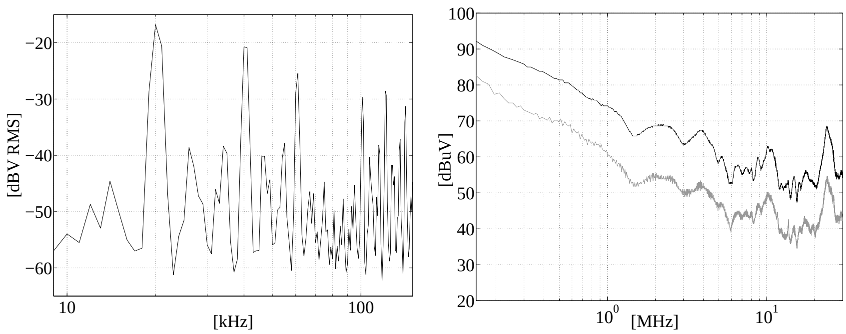

Figure 7.31. The voltage over the load when the aluminium plate is used as ground plane. Left: A 12 point average FFT analysis of the voltage for frequencies from 9 kHz to 150 kHz. Right: A peak (black) and average (grey) frequency analysis of the signal up to 30 MHz.

In the frequency spectrum for the voltage over the load there is a difference in the attenuation of the resonance at 24 MHz of about 5 dB when comparing the aluminium ground plane with the steel ground plane. There is a difference in the resistivity between the two materials as in Table 7.4

Table 7.4. Resistivity for the different ground planes.

Material	Resistivity [Ω mm^2/m]
Aluminium	$2.7 \cdot 10^{-2}$
Steel	$10.5 \cdot 10^{-2}$

Damping of a signal at resonance frequencies is calculated by using equation (7.5). If the damping of the two different set-ups are compared to each other, the damping in the case with the aluminium ground plane should be:

$$\frac{\zeta_{\mathrm{Al}}}{\zeta_{\mathrm{Steel}}} = \frac{\rho_{\mathrm{Al}}}{\rho_{\mathrm{Steel}}} \Rightarrow 10 \cdot \log\left(\frac{\zeta_{\mathrm{Al}}}{\zeta_{\mathrm{Steel}}}\right) = 10 \cdot \log\left(\frac{2.7}{10.5}\right) = -5.8 \text{ dB} \qquad (7.8)$$

which corresponds well to the measured results.

Also a set-up with the ground plane as a current return path is tested. In this set-up the lead-in cable is separated from the ground plane with an 11 cm thick piece of cellular plastic. The measurement of the radiated magnetic field is shown in Figure 7.32. Here it is also possible to see the similarities with the results from the tests with the steel ground plane.

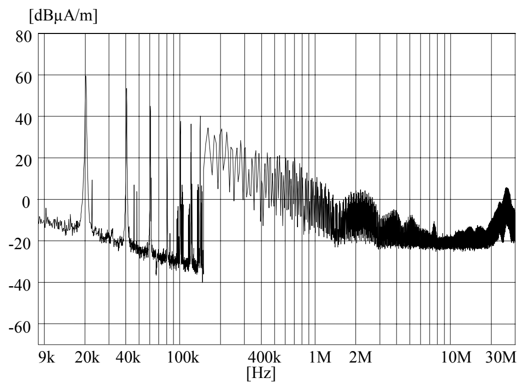

Figure 7.32. The results from the measurement of the radiated magnetic emissions when the aluminium ground plane is used as current return and the lead-in cable is placed 11 cm above the ground plane.

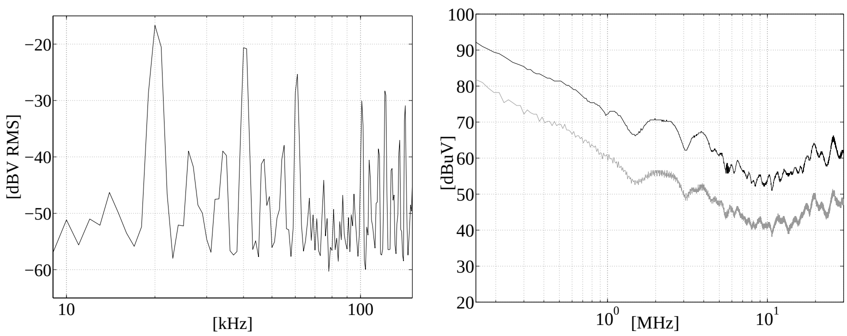

Figure 7.33. The voltage over the load when the aluminium plate is used as current return and the lead-in cable is raised 11 cm above the ground plane. Left: A 12 point average FFT analysis of the voltage for frequencies from 9 kHz to 150 kHz. Right: A peak (black) and average (grey) frequency analysis of the signal up to 30 MHz.

When comparing the load voltage spectrum in Figure 7.33 with the corresponding spectrum for the case with the steel ground plane they are similar, implying that the thickness of the material does not affect the emissions when the ground plane is used as current return.

Chapter 8

Conclusions

In this thesis, different methods for EMI mitigation in vehicles are proposed and investigated. The first part of this chapter presents the conclusions drawn from the measurements in Chapter 7, and the latter proposes ideas for future work.

8.1 Summary of results

The results from the tests of the different set-ups described in Chapter 7 are reviewed in this section. The different set-ups are compared with a reference set-up, which employs a low gate resistance, a fixed switching frequency and two long unshielded cables placed on the steel ground plane between the power electronic converter and the load. Different properties of this reference set-up are alternated in the following set-ups in order to observe the impact on the electromagnetic emissions.

Increasing the gate resistance

In the reference set-up, a gate resistor of 3.2 Ω was used. To increase the rise and fall times of the switching MOSFET, the value of the gate resistor is increased to 330 Ω. The effects from this increase are best observed in the higher frequency ranges above 3 MHz. The increase in gate resistance also results in an increase of the total losses for one period from 4 W to 4.3 W. This implies that using a higher value of the gate resistance gives better attenuation of the high frequency content of the signal, at the cost of increased losses. The increase in losses results in a temperature increase in the semiconductors, which could imply problems in a vehicle where the surrounding environment often has a high temperature. The automotive industry is today asking for more heatproof semiconductors in order to cope with the high temperatures. When using a higher gate resistance it is in other words important to balance between the increase in losses and the attenuation in the higher frequency areas.

Using a randomly varying switching frequency

When the pulse width modulation technique is employed, the switching frequency and its harmonics characterise the appearance of the frequency spectrum. As could be seen in Chapter 7 there are high peaks in the spectrum at these frequencies. In order to lower these peaks a randomly varying switching frequency is employed. The switching frequency is varied between 10 kHz to 40 kHz, and this gives a more broadband noise, especially in the low frequency range where the contribution from the switching frequency to the spectrum is most noticeable. When analysing this method with a quasi-peak measurement it gave very good results for the investigated frequencies. The sound that may originate when employing pulse width modulation with a randomly varying switching frequency is found to be more bearable by the human ear than the sound emitted when a fixed switching frequency is employed. Results from subjective tests in the laboratory with a radio tuned in on the AM-band show that the noise emitted in the radio when the RPWM method is employed are less annoying than the noise emitted when the converter works with a fixed switching frequency. Using a randomly varying switching frequency requires more computational power than a fixed switching frequency. This should, however, not be a problem since the price for computational power continues to decrease, and it is shown in this thesis that the method is realizable also with a cheap micro controller instead of a DSP (digital signal processor). It is important to keep in mind that also the sampling process is affected by the change in switching frequency, and it is therefore important to make sure that only the average of the signal is sampled and not the ripple.

Shielding the conductors

The conductors in the set-up are shielded during one of the test, and this does not give as good results as expected. The poor results could be due to the fact that a braided shield is used instead of a solid shield, since a braided shield is most effective for mitigating the electric field around conductors. It is also shown in Chapter 7 that the spectrum of the output voltage is affected when the shielded conductors are used. This is due to the introduction of new parasitic capacitance located between the shield and the conductors. Even though a braided shield is not very effective against magnetic fields, it is the type of shield that should be chosen in the cases where a shield in a vehicle is employed since a braided shield is stronger, more flexible, and has a longer durability.

Layout of the conductors

In Chapter 7, the effects on the emitted emissions from the conductor layout are investigated. It is shown that using the ground plane/body sheet metal as current return path is not to be recommended since this increase the radiated magnetic emissions radically. It is also shown that the conductors should be placed close to each other and to the ground plane in order to minimize the current loop for the differential and common mode current respectively, thereby decreasing the radiated emissions.

Aluminium ground plane

In order to investigate the effects when some parts of the body sheet metal of a vehicle are made of a nonmagnetic material, a set-up with a ground plane made of aluminium is tested. The results in Chapter 7 show that due to the lower resistivity of the aluminium the damping of the resonance frequencies in the system is to some extent less effective. Tests with the aluminium ground plane as a current return path with the lead-in conductor placed at a distance of 11 cm above the ground plane shows no relevant differences compared to the set-up with the steel ground plane.

8.2 Future work

The increasing number of electrical loads in vehicles makes achieving EMC in a vehicle an even more cumbersome task in the future. The increase in system power in a hybrid vehicle could be a problem due to the high currents that are switched in the converter for the traction motor. However, this traction system causes seldom problems since the vehicle manufacturer shields these cables between the power electronic traction converter and the motor and places the power electronics in close proximity to the motor. Since these high currents only are present at this load, there is often room both financially and practically for filters and snubbers, for which the cost then is a part of the whole power train.

The semiconductor manufacturers have started to address the problems with the steep pulse edges, and a new technology for so called “smart switches” has been developed. These “smart switches” has a built-in feature that allows the gate voltage to be controlled carefully and thereby e.g. provide a softer turn on and off of a MOSFET. This technique is interesting and should be investigated further in the future.

Conductors where the insulating plastic coating is mixed with some kind of iron powder with a high permeability in order to provide some kind of integrated shield on the cable is another interesting topic. This kind of shield would provide great flexibility and durability.

Finally, converter topologies that provide soft switching should be considered an alternative in these applications. A soft switching converter minimizes the voltage time derivative and/or the current time derivative, which also reduces the stress on the switching transistors. Topologies of interest to be investigated are zero-voltage switching, zero-current switching, and resonant mode converters.

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Appendix A

Near and far field

The characteristics of a field are determined by the source, the media surrounding the source and the distance between the source and the point of observation. At a point close to the source are the field properties primarily determined by the source characteristics. Far from the source, the properties of the field depend mainly on the medium through which the field is propagating. This implies that the space surrounding the source of radiation can be divided into two regions, the near and the far field. One common approximation is to consider the large and small distances from the source in comparison with the wavelengths of the radiation. Thus the crucial point is to be able to identify the transition region where the near field “turns into” the far field. This transition region can be found by looking at the wave impedance, which is the ratio between the electric field, E and the magnetic field, H. In the far field is this ratio E/H equal to the characteristic impedance of the medium selected (e.g., E/H=377 Ω for air or free space). The transition region is then found to be at a distance

$$r \approx \frac{\lambda}{2 \cdot \pi} \tag{A.1}$$

from the source of radiation. Since this region is a transition region, the near and far field regions should be some distance away from this. For a working definition it can be said that the near and far field regions should be an order of magnitude smaller or bigger than this transition distance, i.e.

$$\begin{cases} r_{near} \leq 0.1 \cdot \left(\dfrac{\lambda}{2 \cdot \pi}\right) \\ r_{far} \geq 10 \cdot \left(\dfrac{\lambda}{2 \cdot \pi}\right) \end{cases} \tag{A.2}$$

Many handbook however often use a crude diving line based on the value $\dfrac{\lambda}{2 \cdot \pi}$, namely

$$\begin{cases} r \leq \dfrac{\lambda}{2 \cdot \pi} & \rightarrow \quad \text{near field} \\ r \geq \dfrac{\lambda}{2 \cdot \pi} & \rightarrow \quad \text{far field} \end{cases} \tag{A.3}$$

Since this definition involves the wavelength of the radiation it is frequency dependent, which is shown in Table A.1.

Table A.1. Transition distances from near to far field for different frequencies

Frequency	$\lambda/2\pi$
30 kHz	1590 m
3 MHz	15.9 m
30 MHz	1.59 m
300 MHz	0.159 m
3 GHz	0.0159 m

Differences in the near and far field

In the near field must the electric and magnetic fields be considered separately since the ratio of the two is not constant. The ratio is determined by the characteristics of the source and the distance from the source to where the field is observed. If the source has a high current and a low voltage, the near field is predominantly magnetic, i.e. $E/H<377\ \Omega$. On the other hand, if the source has a high voltage and a low current, the near field is mainly electric, i.e. $E/H>377\ \Omega$.

For a straight wire antenna, the source impedance is high. This type of antenna generates mainly an electric field, which implies that the wave impedance close to the antenna is high. As the distance from the antenna is increased, the electric field loses some of its intensity as it generates a complementary magnetic field. The generation of a magnetic field implies that the wave impedance from a straight wire antenna decreases with distance and asymptotically approaches the impedance of free space in the far field. This is an effect from the fact that electric field from an electric dipole attenuates at a rate of $(1/r)^3$ (r=distance from the source of radiation to the point where the field is observed), whereas the magnetic field attenuates at a rate of $(1/r)^2$.

When a magnetic dipole, e.g. a loop antenna, is used, the field produced is mainly magnetic and the wave impedance near the antenna is low. As the distance from the source increases, the magnetic field attenuates at a rate of $(1/r)^3$ and the electric field at a rate of $(1/r)^2$. This implies that the wave impedance increases with distance and approaches that of free space.

In the far field, the wave impedance is constant. That implies that both the electric and magnetic fields attenuate at the same rate, which is 1/r. The electric and magnetic fields are orthogonal in the far field, thus forming a plane wave. The radiated power in the far field is strongly dependent on the input impedance of the antenna.

Appendix B

CIPSR 16

CISPR16 is an international standard that deals with specification for radio disturbance and immunity measuring apparatus and methods that is widely used. It is divided into three parts, covering the following areas:

- Part 1 – covers radio disturbance and immunity testing apparatus
- Part 2 – covers the methods of measurements
- Part 3 – contains further information on radio disturbance

Part 1 of CISPR16 is designated a basic standard and specifies the characteristics and performance of equipment for the measurement of radio disturbance voltages, currents and fields in the frequency range of 9 kHz to 18 GHz. In this standard, the frequency range between 9 kHz and 1 GHz is divided into four bands:

Table B.1. The different frequency bands of CISPR 16

Band A	Band B	Band C	Band D
9 kHz–150 kHz	0.15 MHz–30 MHz	30 MHz–300 MHz	300 MHz–1 GHz

Appendix C
Frequency list

Table C.1. Different radio bands

Broadcast Radio band	Region	Frequency (MHz)
Long Wave (LW) radio	Europe	0.15…0.28
Medium Wave (MW) radio (AM)	Global	0.53…1.75
4 meter	Global	65.2...88.1
FM 1	Japan	75.2...90.9
FM 2	Global	86.6...109.1
2 meter	Global	140.6…176.3
TV, Digital Audio Broadcast (DAB) 1	Europe	172.4…242.2
TV	Global	470…890
Global Positioning System (GPS)	Global	1567…1574 1574…1576 1576…1583
Satellite Digital Audio Radio Service (SDARS)	North America	2320…2345
Bluetooth	Global	2400…2500

Source: [17][18]

Appendix D

MOSFET

This appendix contains a brief description of the MOSFET, mainly originating from the book Power Electronics: Converters, Applications and Design by Ned Mohan, Tore Undeland and William Robbins [35].

The Metal-Oxide-Semiconductor Field Effect Transistor, MOSFET, is based on the original field-effect transistor that was introduced in the 70's. It was partly the limitations of the bipolar power junction transistors (BJTs), which until then was the device of choice in power electronics applications that was the driving force when the power MOSFET was invented [45][7][35]. A bipolar power transistor is a current controlled device, which requires a large base drive current to keep the device in the ON state. It also requires a high reverse base drive current to obtain a fast turn-off. This makes the base drive circuit design very complicated and expensive, which makes the power MOSFET a better choice although the BJT is cheaper than the MOSFET. Another BJT limitation is that both electrons and holes contribute to conduction. The high carrier lifetime of the holes causes the switching speed to be much slower than the power MOSFET, which has no minority carrier injection. A BJT has also problems with thermal runaway since their forward voltage drop decreases with increasing temperature, thereby causing diversion of current to a single device when several BJTs are paralleled. It is easier paralleling power MOSFETs, since the forward voltage drop in a power MOSFET increases with increasing temperature, thereby ensuring an even distribution of current among all components. The MOSFET was introduced in power electronics in the early 1980's and is now often choosed because of its fast switching qualities. It is widely used in applications with low power level and low blocking voltage.

A power MOSFET consists mainly of a vertically oriented structure with four layers. The layers are of alternating p-type and n-type doping as shown in Figure D.1.

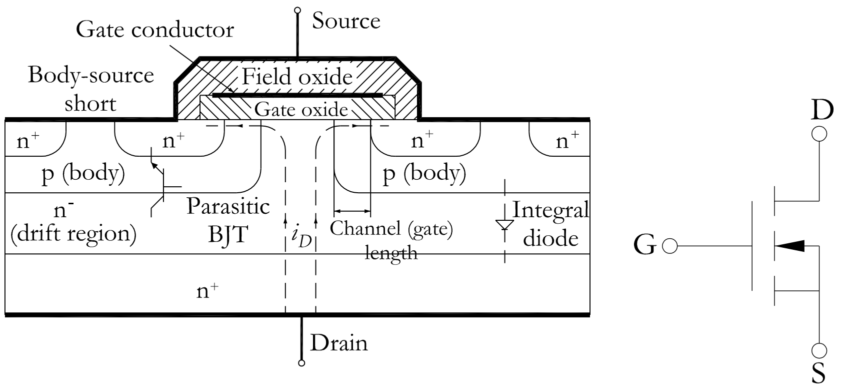

Figure D.1. Left: Vertical cross-section view of an n-channel MOSFET. The opposite doping profile may be employed, which is then termed a p-channel MOSFET. Right: Circuit symbol for an n-channel MOSFET.

The two n^+ end layers, labelled source and drain, are the layers with the largest doping, typically $10^{19}\,cm^{-3}$ [35]. The p-type body layer in the middle is typically doped at $10^{16}\,cm^{-3}$ and this is the layer where the channel is established between source and drain when the MOSFET is turned on. The drift region, below the body layer, determines the breakdown voltage of the MOSFET and this has often a doping between 10^{14} to 10^{15} cm^{-3}.

An internal parasitic BJT exists between the source and the drain of a MOSFET [7]. It is the p-type body region that forms the base of the bipolar junction transistor, BJT. The source region becomes the emitter of the BJT and the n-type drain region becomes the collector. When the parasitic BJT is turned on, the breakdown voltage of the MOSFET decreases by 50-60%, which means that if the drain current is not externally limited, the component will be destroyed if the breakdown voltage is applied. In order to prevent the turn-on of the parasitic BJT, the n+ source region and p-type body region is shorted by the source metallization. Also a large increase rate of the drain-source voltage of the MOSFET could turn on the parasitic BJT (latch-up). This could be prevented by decreasing the switching speed or by increasing the doping density of the body region. Since the source region is short, also a parasitic diode is formed. This parasitic diode can be useful as a freewheeling diode in half-bridge and full-bridge converters where the reverse recovery time of the diode is not a critical issue.

Characteristics of the capacitance

There are a number of different parasitic capacitances in a MOSFET structure. An equivalent circuit with parasitic capacitances is shown in Figure D.2.

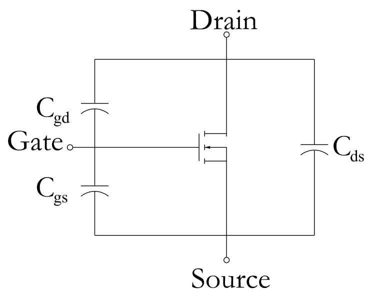

Figure D.2. An equivalent circuit of a MOSFET with parasitic capacitances.

The capacitance between gate and source is denoted C_{gs}. This capacitance is the sum of the capacitance between the gate and source metal, the capacitance between the gate and n^+ source diffusion region and the capacitance between the gate and p-body. The size of C_{gs} is affected by a number of different parameters, both geometrical and electrical, such as the drain-source voltage. C_{gd} is the capacitance between the gate and the drain, and is influenced by the voltage of the gate and the drain. The capacitance between the drain and the source, C_{ds}, varies due to the variation of the depletion region widths of the p-body and the n^- -drift region that depends on the drain-source voltage.

The turn-on of the MOSFET

The depletion region between the SiO_2 and the silicon in Figure D.3 is formed by applying a small positive gate-source voltage to the MOS capacitor structure.

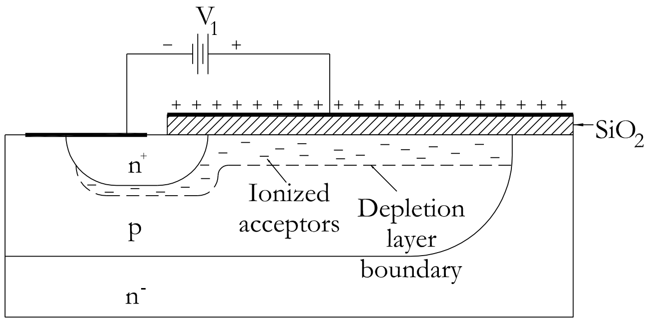

Figure D.3. Formation of the depletion layer in a MOSFET.

The applied gate-source voltage induces a positive charge on the upper metallization. This positive charge requires an equal negative charge on the other side of the plate, which is created by repelling the majority carrier holes and thereby exposing negative charged acceptors. The result is a depletion region.

When the gate-source voltage continues to increase, the depletion layer to grows in thickness in order to provide additional negative charge Figure D.4.

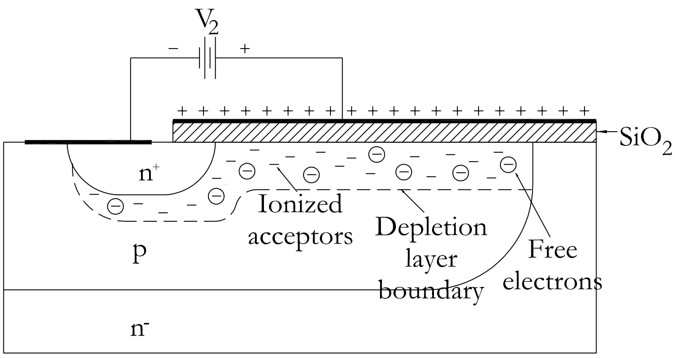

Figure D.4. Start of the formation of the inversion layer.

As the gate-source voltage is increased, also the electric field at the oxide-silicon interface increases. This does not only repel the free holes, but does also attract free electrons. The source of these free electrons is mainly the electron-hole generation via thermal ionization with the free holes being pushed into the bulk ahead of the depletion region. Electrons that are attracted from the n+ drain by the positive charge of the holes neutralize these extra holes.

When the gate-source voltage is increased further, the density of free electrons at the interface will become equal to the free hole density in the bulk of the body region, away from the depletion layer. This layer of free electrons is highly conducting and will thus have the property of an n-type semiconductor, see Figure D.5.

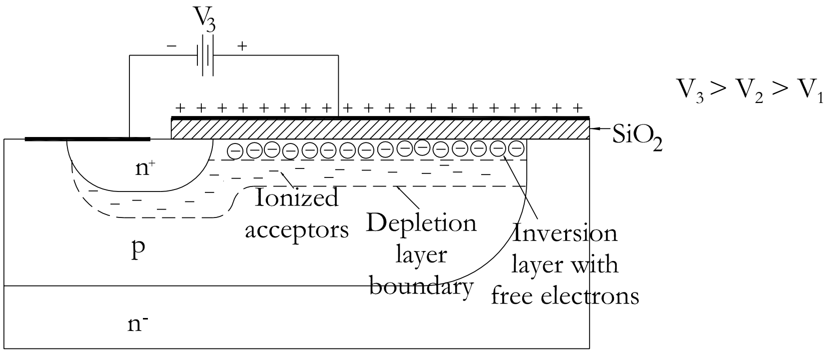

Figure D.5. Formation of the inversion layer at the SiO_2 interface.

The layer of free electrons is termed an inversion layer, and will be a conductive path or channel between the n+ drain and source regions. This will permit the flow of a current between the source and the drain. The property that makes it possible to change the conductivity type beneath the gate insulator this way is called field effect, which is a part of the component name MOSFET (Metal-Oxide-Semiconductor Field Effect Transistor).

Switching waveforms

A turn on or a turn off of a semiconductor is, due to the structure with the carriers, never an instant event. This is also true for a MOSFET, and the currents and voltages at turn on are depicted in Figure D.6.

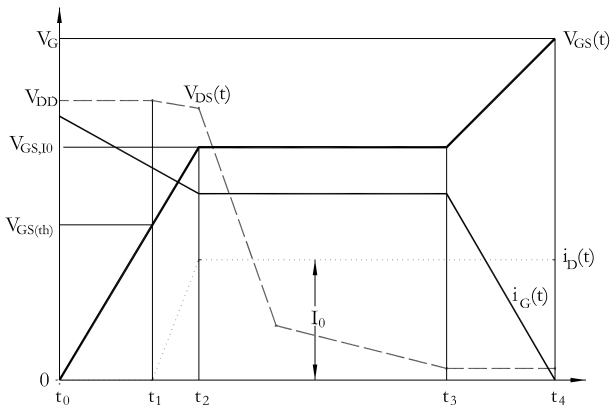

Figure D.6. The switching waveforms for a MOSFET at turn on.

From time t_0 to t_1 are C_{gs} and C_{gd} charged by the gate current. The equivalent circuit of the MOSFET during this time period is shown in Figure D.7. During this time is the gate-source voltage increasing from 0 V to the threshold level $V_{GS(th)}$.

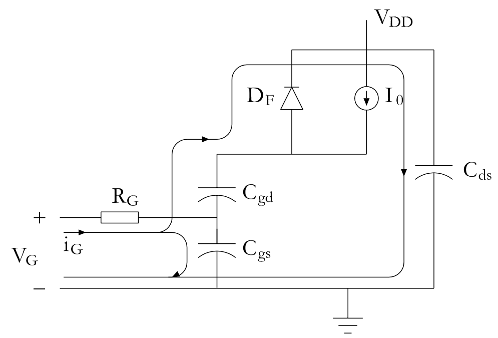

Figure D.7. Equivalent circuit of the MOSFET from time t_0 to time t_1.

The RC-circuit formed by the gate resistance, gate-source capacitance and the gate-drain capacitance determines the slew rate of the gate-source voltage. The voltage will be increasing exponentially with a time constant $\tau_1 = R_G(C_{gs} + C_{gd})$. In this time period the MOSFET is still in the state where it has not been turned on, which means that the drain current is zero and that the drain-source voltage still is equal to V_{DD}.

During the time period from t_1 to t_2, the gate-source voltage continues to increase exponentially thereby passing the threshold voltage. When V_{GS} passes the threshold voltage, the drain current starts to increase and reaches eventually the full load current, I_0.

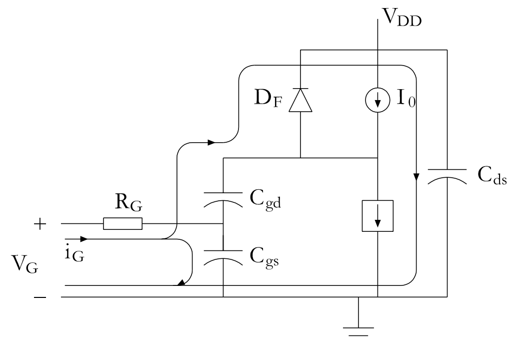

Figure D.8. The equivalent circuit of the MOSFET between t_1 and t_2.

In the time period between t_2 and t_3 the MOSFET carries the full load current and the gate-source voltage becomes temporarily clamped at $V_{GS,I0}$, which is the voltage needed to maintain the drain current equal to the full load current in the active region. The entire gate current will flow through the gate-drain capacitance as is shown in Figure D.9.

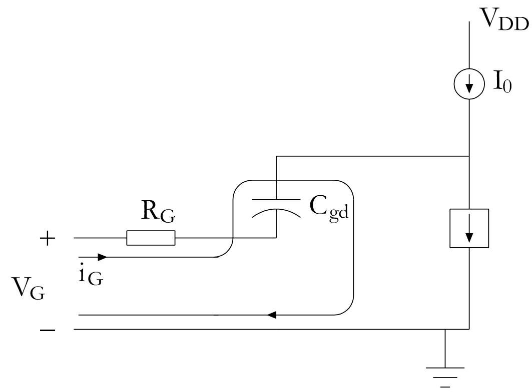

Figure D.9. Equivalent circuit of the MOSFET from time t_2 to time t_3.

The value of the gate current is given by the expression

$$i_G = \frac{V_G - V_{GS,I_0}}{R_G} \tag{D.1}$$

The drop rate for the drain-source voltage will then be

$$\frac{dv_{DG}}{dt} = \frac{dv_{DS}}{dt} = \frac{i_G}{C_{gd}} = \frac{V_G - V_{GS,I_0}}{R_G \cdot C_{gd}} \tag{D.2}$$

As can be seen in this expression it is possible to influence the fall time of the drain-source voltage by changing the size of the gate resistance. This is a possibility that is used in this thesis, see Chapter 5. In this time period, the MOSFET is still said to be working in the active region, and as the drain-source voltage decreases it gets closer to the ohmic region.

At time t_3, the drain-source voltage reaches $V_{DS(on)} = I_0 \cdot R_{DS(on)}$ and the equivalent circuit for this time is shown in Figure D.10.

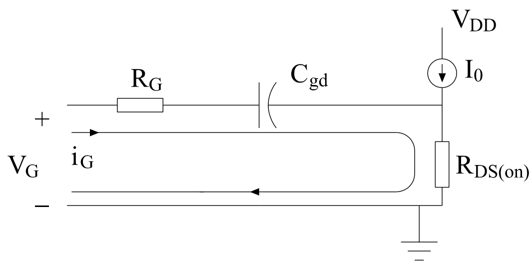

Figure D.10. The equivalent circuit of the MOSFET at time t_3.

In the last time period of Figure D.6. the MOSFET works in its ohmic region.

Snubbers

Although the turn on and turn off of a semiconductor as a MOSFET is not an instant event, the voltage and current derivatives could be high enough to destroy the semiconductor. In order to protect the semiconductor, a snubber is used, which is a protective circuit for a power electronic component, such as e.g. a MOSFET. There are a number of different snubbers, protecting the power electronic components against phenomena such as overvoltage, or high time derivatives of the current or the voltage. Depending on their function, the snubbers are often divided into groups, as for example overvoltage, turn-on and turn-off snubbers. The overvoltage snubber limits the voltage over the

MOSFET at turn off caused by stray inductances. Since there always are stray inductances in a system, the energy stored in them causes a high voltage during turn off (the current through the stray inductance decreases at a high rate) of the MOSFET. An overvoltage snubber like the one in Figure D.11. limits this high voltage by letting the energy stored by the stray inductance transfer to the snubber capacitance, C_s.

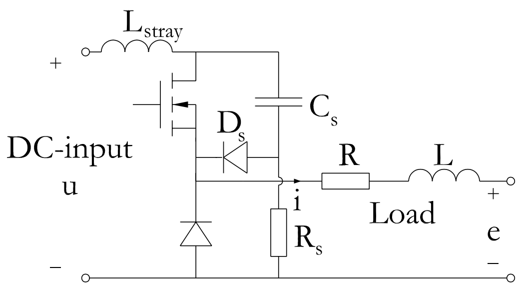

Figure D.11. A step down converter with an overvoltage snubber. The snubber components are marked with the index “s”.

A turn-on snubber is used in order to reduce the turn-on switching losses at high switching frequencies. The function of the snubber is to reduce the voltage across the switch as the current builds up. In Figure D.12. it is the inductance L_s that reduces the voltage at turn-on.

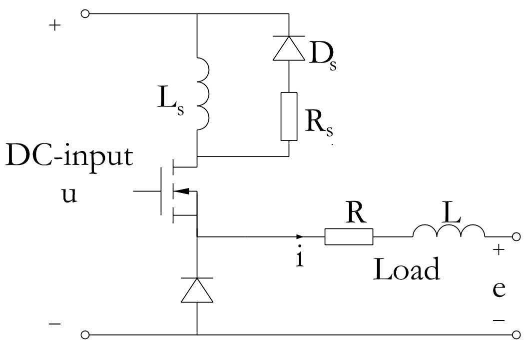

Figure D.12 A step down converter with a turn-on snubber. The snubber components are marked with the index “s”.

The turn-off snubber is used to avoid problems during turn-off by providing a zero voltage across the transistor while the current turns off. This snubber does not protect against the high voltages created by the stray inductances like the overvoltage snubber. During turn-off of the switch, the capacitor C_s in Figure D.13. provides an alternative path for the load current when the transistor current is decreasing.

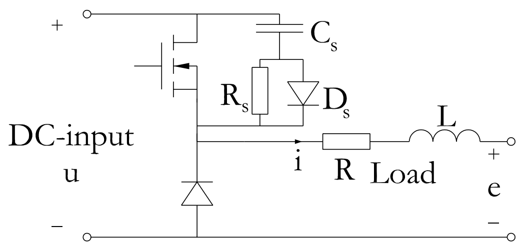

Figure D.13. A step down converter with a turn-off snubber. The snubber components are marked with the index "s".

Appendix E

Fourier analysis

A pure sine with or without a superposed DC-level consists of only one frequency. In general could any continuous periodic with another appearance, as e.g. a triangular wave or a square wave, be expressed as a sum of different sines. Except for the fundamental frequency, the signal also consists of a number of harmonics, i.e. multiples of the fundamental frequency.

When performing a Fourier analysis on a continuous periodic signal it is decomposed into its frequency components. An arbitrary continuous periodic signal, u(t), with the period time, T, has a repetition frequency of $f=1/T$ and an angular frequency of $\omega=2\cdot\pi\cdot f$. This signal could be expressed as a sum of an infinite number of discrete sine (and cosine) signals where the lowest frequency is equal to f and the other frequencies are multiples of this fundamental frequency, i.e. $k\cdot f$, where k is an integer. The signal could then be expressed as

$$u(t) = U_0 + (a_1 \cos\omega t + b_1 \sin\omega t) + (a_2 \cos 2\omega t + b_2 \sin 2\omega t) + \ldots \quad \text{(E.1)}$$

or

$$u(t) = U_0 + \sum_{k=1}^{\infty} a_k \cos(k\omega t) + \sum_{k=1}^{\infty} b_k \sin(k\omega t) \quad \text{(E.2)}$$

The term U_0 is the DC-level for the periodic signal, and the Fourier coefficients a_k and b_k are calculated by

$$\begin{cases} a_k = \dfrac{2}{T}\displaystyle\int_0^T u(t)\cdot\cos\left(\dfrac{2k\pi t}{T}\right)dt \\ b_k = \dfrac{2}{T}\displaystyle\int_0^T u(t)\cdot\sin\left(\dfrac{2k\pi t}{T}\right)dt \end{cases} \tag{E.3}$$

For a square wave with 75% duty-cycle (Figure E.1.), which oscillates between 0 and 12 V, the Fourier coefficients will be as follows.

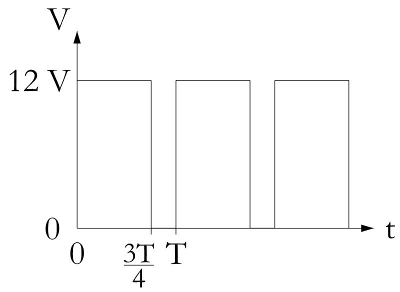

Figure E.1. A square wave signal with 75% duty-cycle.

$$\begin{cases} a_k = \dfrac{2}{T}\displaystyle\int_0^{3T/4} 12\cos\left(\dfrac{2\pi kt}{T}\right)dt = \dfrac{24}{T}\left[\dfrac{T}{2\pi k}\sin\left(\dfrac{2\pi kt}{T}\right)\right]_0^{3T/4} = \dfrac{12}{\pi k}\sin\left(\dfrac{3\pi k}{2}\right) \Rightarrow \\ \Rightarrow \begin{cases} -\dfrac{12}{\pi k} \text{ for } k = 1,5,9\ldots \\ 0 \text{ for } k = 2,4,6\ldots \\ \dfrac{12}{\pi k} \text{ for } k = 3,7,11\ldots \end{cases} \\ b_k = \dfrac{2}{T}\displaystyle\int_0^{3T/4} 12\sin\left(\dfrac{2\pi kt}{T}\right)dt = \dfrac{24}{T}\left[-\dfrac{T}{2\pi k}\cos\left(\dfrac{2\pi kt}{T}\right)\right]_0^{3T/4} = \dfrac{12}{\pi k}\left(1-\cos\left(\dfrac{3\pi k}{2}\right)\right) \Rightarrow \\ \Rightarrow \begin{cases} \dfrac{12}{\pi k} \text{ for } k = 1,3,5\ldots \\ 0 \text{ for } k = 4,8,12\ldots \\ 2\cdot\dfrac{12}{\pi k} \text{ for } k = 2,6,10\ldots \end{cases} \end{cases}$$

This gives a frequency spectrum as in Figure E.2.

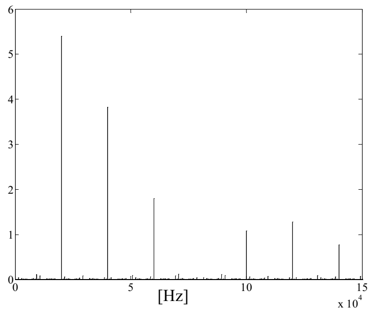

Figure E.2. The frequency spectrum of the signal in Figure E.1.

This kind of series representation enables the description a periodic function in terms of the frequency-domain attributes of amplitude and phase. With a Fourier transform, this can be extended to describe functions that are not periodic.

When a signal is sampled into an oscilliscope for example, the oscilliscope could perform a discrete Fourier transform on the signal. This is a very time consuming calculation, and in order to be able to see the Fourier transform in real time, a fast Fourier transform (FFT) is performed on the signal. A FFT is a type of discrete Fourier transform that builds on a 2-points discrete Fourier transform, and will not be investigated further in this Appendix.

Window functions

When a signal is digitally sampled, it is often multiplied with a window function before displayed. A window function is a pure digital weighting function. The digital samples are being weighted by multiplying each value with a factor of the window function. This gives weighted input data for the Fourier transform process. Each sample gets a weight that is equal to 0 at the beginning of the window and maximal by the middle of the window. A digital window function could be compared to an analogue band pass filter. There are a number of different window functions which all have their own properties. Some window functions, such as Hamming or Hanning windows facilitates investigations of the frequency content of a signal, and others, such as Flattop window, are more accurate for displaying amplitudes.

Appendix F
Decibel

Decibel is a commonly used term, and it is a logarithmic unit expressing the ratio between two powers. The term decibel is abbreviated dB, and is defined as

$$\text{Number of dB} = 10\log\frac{P_2}{P_1} \tag{F.1}$$

Sometimes, the decibel is used for other ratios than power ratios. It could for example be used to describe the ratio between two current or voltages, and the expressions look then like:

$$\begin{aligned}\text{dB current gain} &= 10\log\frac{I_2^2}{I_1^2} = 20\log\frac{I_2}{I_1}\\ \text{dB voltage gain} &= 10\log\frac{V_2^2}{V_1^2} = 20\log\frac{V_2}{V_1}\end{aligned} \tag{F.2}$$

The decibel may also be used to represent an absolute power level by placing a reference power, P_0, in the denominator, as

$$\text{Number of dB (absolute)} = 10\log\frac{P}{P_0} \tag{F.3}$$

In this case, the decibel represents the absolute power level above or below the reference power, which implies that it is important to know the reference power. In order to mark the reference power used, one or more additional letters are added to the abbreviation dB. Some commonly used dB units, their reference levels and abbreviations are shown in Table F.1:

Table F.1. Commonly used dB units and their reference levels

Unit	Type unit	Reference level
dBm	Power	1 mW
dBV	Voltage	1 V
dBmV	Voltage	1 mV
dBμV	Voltage	1 μV

Appendix G
Nomenclature

Abbreviations

A	Average
AC	Alternating current
AM	Amplitude modulation
AMN	Artificial mains network
BJT	Bipolar junction transistor
CE	Communautées Européennes
CISPR	International Special Committee on Radio Interference
CM	Common mode
CW	Continuous wave
DC	Direct current
DM	Differential mode
DS	Drain-source
DSP	Digital signal processor
E	Electric field

EMC	Electromagnetic compatibility
EMI	Electromagnetic interference
ESD	Electrostatic discharge
EU	European Union
EUT	Equipment under test
FET	Field effect transistor
FFT	Fast Fourier transform
FPWM	Pulse width modulation, fix switching frequency
GD	Gate-drain
GS	Gate-source
H	Magnetic field
IEC	International Electrotechnical Comission
IEEE	Institute of Electrical and Electronics Engineers
IGBT	Insulated gate bipolar transistor
LISN	Line impedance stabilization network
MOSFET	Metal-oxide-semiconductor field effect transistor
OATS	Open area test site
P	Peak
PCB	Printed circuit board
PE	Power electronics
PLD	Programmable logic device
PWM	Pulse width modulation
QP	Quasi-peak

RCD	Term for a snubber consisting of a resistor, a capacitance and a diode
RMS	Root mean square
RPWM	Pulse width modulation, randomly varying switching frequency
TC77	Technical Committee 77
UIR	Union international de radiodiffusion

Symbols

A	Area of current loop
a_k	Fourier coefficient
B	Magnetic flux density
b_k	Fourier coefficient
C_{DS}	Drain-source capacitance
C_{GD}	Gate-drain capacitance
C_{GS}	Gate-source capacitance
D	Distance between conductors
D	Diameter of conductor
D_F	Freewheeling diode
e	Electromotive force, emf
f	Frequency
f_{sw}	Switching frequency
h	Height above ground plane
I_0	Load current

I_{CM}	Common mode current
I_D	Drain current
I_{DM}	Differential mode current
I_G	Gate current
M	Mutual inductance
$r_{DS(on)}$	MOSFET on resistance
R_G	Gate resistance
T	Period time
U_0	Fourier constant
U_{CM}	Common mode voltage
U_{dc}	DC-link voltage
U_{DM}	Differential mode voltage
V_{DD}	Drain voltage
V_{DS}	Drain-source voltage
V_G	Gate voltage
V_{GS}	Gate-source voltage
$V_{GS(th)}$	Gate-source voltage, threshold
$V_{GS,I0}$	Gate source voltage depending on the load current
V_N	Noise voltage

Greek symbols

ε	Dielectric constant
ζ	Damping

θ	Angle between magnetic field and loop antenna
λ	Wavelength
ρ	Resistivity
τ_-	Time instant for the falling edge of a pulse
τ_+	Time instant for the rising edge of a pulse
ω	Angular frequency