CHAPTER 1

Introduction to EMC/EMI.

1.0. Introduction.
Electromagnetic compatibility (EMC) is the branch of engineering, which studies the operation of equipment in their electromagnetic environment. As early as 1920, technical papers on radio interference began to appear in many technical journals and today interference has become a major problem, which can lead to catastrophic consequences. All electrical and electronic equipment and nearly all engineering systems contain electronics-based instrumentation and control circuits. Over the years, the logic thresholds on which the operation of these circuits is based have decreased dramatically and this has brought about miniaturisation and enormous increases in power per unit volume [1].

The consequences of this development is that modern circuits are potentially more susceptible to electromagnetic interference (EMI) upsets and are themselves, potential sources of interference. The increase in processing speed is a major contributor in this latter feature. Clock rates used in digital
circuits are of the order of tens of megahertz with pulse risetimes of a few nanoseconds. These frequencies and their harmonics generate interference over a wide spectrum. Frequency converters, switch mode power supplies and drive systems use thyristors and other switching elements. The process involves switching of substantial currents at frequencies of tens of kilohertz and hence generates high frequency interference in close proximity to sensitive digital circuits.

In the Working Nation Statement of May 1994, the Australian Government announced that it will implement an Electromagnetic Compatibility (EMC) framework from January 1996, coinciding with the introduction of a similar regime in the European Union [2]. EMC has become increasingly important because of its wide ranging industrial implications and the wide spread use of electrical and electronic equipment.

The following section introduces the basic concept of Electromagnetic Compatibility / Interference and attempts to highlight its importance in industry and everyday life. In addition, a short statement will be provided to explain the aim of this project. Finally, this chapter will contain an in-depth investigation of the EMI in power electronics and controlled drive systems.

- **Basic Definitions.** In order to discuss this problem, it is necessary to define some of the terms that will be used. In particular the words *Interference* or *Disturbance* need to be understood as terms that both describe the phenomena (noise, unwanted signal, etc.) causing the degradation of the system in question. The IEEE has defined the following terms:
• EMC (Electromagnetic Compatibility) is the ability of a device, equipment or a system to function properly in its Electromagnetic (EM) environment without introducing intolerable disturbances to anything in that environment.

• EMI (Electromagnetic Interference) describes the impairment of desired operation by an EM disturbance.

• RFI (Radio Frequency Interference) is acknowledged as an older term for EMI, coined when most EMI problems were radio frequency related.

• EMS (Electromagnetic Susceptibility) is the inability of a device, equipment, or system to perform without degradation in the presence of an Electromagnetic disturbance [3].

Examples (Problems caused by EMI)

In order to give an idea about what sort of EMC problems can be encountered in real life; it is worth providing some examples from the past.

On July 29, 1967, the US aircraft carrier Forrestal was deployed off the coast of North Vietnam. The carrier deck contained numerous attack aircraft that were fuelled and loaded with 1000-pound bombs, as well as air-to-air and air-to-ground missiles. One of the aircraft missiles was inadvertently deployed striking another aircraft and causing an explosion in its fuel tanks and the subsequent death of 134 servicemen. The problem was thought to be caused by the generation of radio frequency (RF) voltages across the contacts of a shielded connector by the ship's high power search radar [2].

Some years later a recently modified radio controlled crane in a steel works unexpectedly tipped a ladle of molten steel over workers, killing one and injuring others. It was found that in some circumstances a new
radio link had interfered with the original push-button controller, thus causing the accident. These are just examples of accidents caused by EMI in our dense electronic world and the probability of occurrence of such incidents is mainly due to the spread of electric and electronic equipment. The life threatening results clearly demand remedies [3].

**Power line electromagnetic interference.**

The proliferation of sophisticated electronic equipment in today’s society has spawned numerous electromagnetic interference (EMI) and electromagnetic compatibility (EMC) problems. Many of these problems are associated with the commercial electrical power distribution system. Since most electronic equipments are connected to commercial power systems, the concern for both power quality and power susceptibility has increased dramatically in recent years by both the providers and users of electric power.

The problem has been aggravated as modern electronic system incorporated embedded computers, microprocessors and other complex solid-state components. By their nature, these devices operate at low energy level and high speeds, making them particularly fragile and susceptible to power electrical noise. At the same time, these same devices often contribute to the power line levels in a system as well.

Low voltage (up to 1000 V) electric power supply lines in Australia are three-wire lines. The three wires are the line (phase), neutral, and safety ground conductors. The neutral and ground conductors are bonded together at each service entrance. The distance between the equipment/apparatus connected to the power supply line, and the actual location of electrical
earth is thus limited, or small. In this situation, common-mode surges, and interference, would be smaller than the differential-mode interference's. In power distribution systems with two wire lines, the bond between neutral and earth is located remotely from the service entrance. In this case, the common-mode interferences would predominate over the differential-mode interferences.

Electrical disturbances are induced in the power supply lines as a result of natural electromagnetic phenomena, and from the operation of a variety of equipment. The most common natural phenomena of lightning can induce transients on overhead power supply lines either by a direct strike, or by way of induction from a strike on a nearby structure. Machine operations such as local load switching, switching-off or switching-on of heavy electrical equipment, motor control activation arc welders, and industrial cranes can induce substantial electrical disturbances in the power supply lines.

**Types of power line disturbances.** There are different types of power line disturbances, which can be classified under the following categories [5]:

- **Voltage variation** - is the deviation of voltage from the operating range. When the input level is below the range it is termed a "sag", and "surge" when the input level is above the range.

- **Frequency variation** - is a change in the sine wave frequency, caused by poor speed regulation of the generating equipment.
- **Harmonic distortion** - refers to the distortion of the voltage and current waveforms of a pure sine wave. Harmonic distortion is a key concern of power utilities, but not for electronic equipment.

- **Continuous electrical noise** - means any continuous variation from the desired signal, which would include the sags, surges and distortions. Continuous noise is quite common and is due to repetitive events such as switching mode power supplies, motor brush arcing or external coupling from radio transmitters.

- **Transients** - are short term disturbances (much less than a cycle). They may instantaneously either increase (spike) or decrease (notch) the sine wave amplitude. Figure 1.1 shows various types of disturbances.

Voltage/current surges propagating along the power supply lines and their behavior have the following characteristics:

- For typical voltage or current surges produced by lightning or switching of loads, their propagation in power distribution lines may be considered as a case of classical transmission line only if the lines are long enough to contain the surge front. In that case, the characteristic impedance of the transmission line $Z_0$ is given by the equation:

$$Z_0 = \sqrt{\frac{L}{C}}$$

where L and C are the inductance and the capacitance of the transmission line per unit length.

- In most practical cases, $Z_0$ is not the significant parameter. The frequency spectrum of the impulse, and the line impedance at significant frequencies of that spectrum, will need to be taken into consideration for an accurate analysis. These considerations call for a Characterisation of the complex impedance of a network.
Figure 1.1. Types of disturbances
Sources of disturbances.
A source of EMI can be any electrical or electronic device that has changing voltages and currents. As might be expected, the faster the rate of change of voltage or current in any faulty equipment, the wider is the spectrum of RF interference produced. The higher the magnitude of the noise voltages or currents the greater will be the conducted and radiated emissions. Sources can be grouped in terms of usage, and they may be sources of continuous or transient interference. Emission from continuous sources is best measured and analysed in terms of its spectral content using a spectrum analyser. While for transient emission, the signals are often of short duration, and hence have a wide signal bandwidth, it is easier to measure and analyse such a signal in the time domain [6].

1.0.1. EMC problem.
The problem of EMC can be stated as follows. All electrical and electronic equipment governed fundamentally by physical laws which, in the macroscopic case, are described by Maxwell's equations

\[ \nabla \times \mathbf{E} = \frac{\partial \mathbf{B}}{\partial t} \]  
(1.1)

\[ \nabla \times \mathbf{H} = \mathbf{J} + \frac{\partial \mathbf{D}}{\partial t} \]  
(1.2)

\[ \nabla \cdot \mathbf{D} = \rho \]  
(1.3)

\[ \nabla \cdot \mathbf{B} = 0 \]  
(1.4)

These equations completely describe the electromagnetic field throughout region, including boundary conditions, change around a particular point. Maxwell's equations form the cornerstones of electromagnetic phenomena,
and are therefore essential to our understanding of how to design electronic systems such that they will comply with regulatory requirements and will also not cause interference or be interfered with. Designers usually use simplified versions of Maxwell's equations, since they are complex and very difficult to calculate. Any simplifications fail to take into account all the possible EM interactions [4]. No equipment can completely avoid electromagnetic disruption if the disrupting fields are of sufficient intensity. Therefore, in order to achieve a certain level of performance, a realistic assessment of the threat levels in normal operation should be made [7].

The widespread use of motor drives increase the harmonic pollution and degrade the quality of power in the main networks. Special attention is given, in this study, to pulse-width-modulated (PWM) inverters since they can have a high switching frequency of up to 150KHz. The conducted emission produced by these types of drives create numerous unwanted negative effects, such as resonant conditions in network or the overheating in electrical machines, the aging of insulation, and of course interference with communication systems. In addition, the measurement methodology needs to be improved to allow quick, inexpensive, and data acquired assessment. This project aims to fulfil these needs.

1.0.2. Requirements and standards for electrical and electronic equipment.

In order to control interference produced by products manufactured or sold in a country, government agencies impose legal requirements. These legal requirements can take a form of standards, which ensure that, electromagnetic compatibility was managed properly in the design. Each
country sets its own EMC standards, although in many cases they are based on the work of international bodies such as CISPR (International Special Committee on Radio Interference). However, differences between standards is creating problems in commercial trading of electrical and electronic equipment across national boundaries, as shown in table 1.1.

Table 1.1. EMI Standards and regulations in different countries

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<th>EUROPEAN EN</th>
<th>INTERNATIONAL</th>
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In Australia the first significant standard relating to radio interference was developed in the early 1960s. The standards setting organisation in Australia is the Standards Association of Australia (SA), and within SA there are various committees dealing with different aspects. In the case of EMC, the relevant committee is Committee TE/3. The Committee’s policy is to work on maintaining the Australian Standard in a form, which resembles as closely as possible to the latest version of the CISPR documents [8].

1.1. Introduction to EMC & EMI in power electronics.
A small amount of EM energy coupled to a modern circuit using low level thresholds can disturb data and in general can lead to a change in circuit function or cause damage. Therefore, the tendency to use higher frequencies and fast pulses in all electrical and electronic applications affects both emission and susceptibility. Change in electrical conditions, usually generated by fast switching is the most common cause of emission. Furthermore, this change is not instantaneous which means that emission during the transient period is severe. Controlling the rate of change and minimising coupling paths is necessary to control interference.
The use of power electronics is associated with many advantages such as high efficiency, zero acoustic noise and speed of response but the switching mode of operation of this system can lead to unwanted harmonics. In order to reduce the effects of EMI, many measures can be implemented against either the generation or the propagation of RFI, mainly through the proper design of equipment and the suppression of propagation [9].

1.1.1. Propagation of EM energy.
EM energy could be transported away from its sources via conductive and radiative mechanisms. When EM energy is travelling along wires away from the source the interference is called conductive, while when energy propagates long distances in free space away from source, the interference is called radiative. It is very important that the nature of the coupling paths is understood, for the purpose of providing appropriate remedies. Remedies vary from shielding to filtering to grounding. Shielding is considered as a way to prevent leakage of EM radiation by introducing a conducting barrier. Filtering is the introduction of limiting devices, while some filters absorb and some divert interference energy [10]. Grounding has a main purpose that is to ensure the safety of personnel and the appropriate operation of protective devices and it has a profound effect on EMC.

1.1.2. Power electronics and EMC.

Power electronic equipment has particular problems, and may be classified according to the power delivered and the switching frequency

- High frequency and small power systems (up to 1MHz, several hundred W). dc-dc converters are the main applications of this class. The high switching frequency can generate harmonics at radio frequencies. In order to minimise EMI, good design, layout, shielding and filtering are needed.
- Low frequency and large power systems (up to 1KHz, 10’s kW to 100’s kW). The applications include traction, power distribution, conversion and variable speed motor drives [11].

Power electronics and EMI is a very large topic, which needs more than one chapter to fully cover and understand the area. However, the rest of this
chapter will aim to investigate the EMI produced by motors and associated drives.

1.2. EMI in motors and associated controls.
All motorised appliances may cause interference with broadcasts and receptions of radio and television signals. Most RFI are related to commutation effects, the use of brush-gears, and magnetic conditions. Sources of RFI problems can be:

- rapid current commutation
- rapid change of magnetic field
- rapid switching of controllers
- making and breaking of contacts
- bad physical contacts
- capacitor discharge
- sparking or arcing

Reducing and controlling RFI in motors and associated controls is possible by the following two steps: design consideration and suppression of RFI [12].

1.2.1. Design consideration.

1. Magnetic/electric loading ratio. Achieving a proper balance between these two quantities is necessary since this affects not only the input/output performance but also brush wear (in dc motors). A loading ratio of 0.7 for industrial type dc motors and 1.2 for small motors would be an optimum.
2. **Effect of commutator shift.** A proper brush shift for improved commutation has to be maintained so that the armature current can still be commutated in the neutral zone under load. However, commutator shift also depends on the loading ratio and therefore cannot be optimised in isolation.

3. **Effect of brushes.** The choice of guides and brush guides is also important. Low current densities and high resistance in the brushes will reduce sparking at the commutator and a consequent reduction in RFI. Brush bedding affects contact area, current density, sparking and RFI. Lubricated brushes help RFI, but wear out faster mechanically [13].

### 1.2.2. Suppression of RFI.

Switching of currents in motors and discontinuous waves and pulses in controls create RFI that is conducted or radiated to other circuits.

1. **Filtering of power lines.** Interference can propagate in two forms, radiation and conduction. For radiated interference, matched impedance can be used to reduce the effect of radiation. At higher frequencies, i.e. above 30 MHz, direct radiation from the drive itself tends to be more significant. If a radio receiver is used in close proximity to a drive or its associated cables, there may be interference to AM broadcasts in particular, e.g. a case where an unfiltered drive is directly fed from the same low voltage supply as domestic premise [14]. To create a deliberate mismatch of its impedance to the load at frequencies in the band, a common-mode choke filter is used as shown in figure 1.2.
Figure 1.2. Common-mode choke

Figure 1.3. Suppression networks
2. **Other techniques.** Eliminating and reducing the undesirable RF noise is possible by using other techniques such as inductive and capacitive suppression. Capacitance across the motor terminals effectively short circuits the HF noise, while chokes in series dampen the noise in the VHF range. Different types of RFI suppression networks are shown in figure 1.3. In addition, the use of ferrite beads can reduce noise in the high-frequency range, but this technique is effective for low-power circuits only since high power circuits may be polarised and the technique become ineffective [15].

### 1.2.3. EMI in AC drives.

AC drives are a source of high frequency electrical interference, and most industrial drives are installed without radio interference suppression filters. To better understand the work of an AC drive we will provide a power diagram of a typical AC drive, shown in figure 1.4, noting that only voltage source inverters are considered.

The rise and fall time of output transistors is typically in the order of 100 ns, which gives rise to EM disturbances due to stray capacitances and inductances within the power circuit. Therefore, since the output is unfiltered, the fast voltage changes are applied to the motor cable and to the motor, which, at the power frequency can be modeled as being mainly inductive. The motor and the cable contain stray capacitance and hence can be represented by a distributed circuit of inductors and capacitors, as shown in figure 1.5. The fast voltage change, due to these stray capacitances will cause HF currents to circulate from line to line and from line to ground [16]. The conducted HF disturbances in the supply network originate from the common mode current, and can reach 120 dB/mv but tend to decrease as the
frequency increases. In addition, the input rectifier diodes produce a non-
sinusoidal input current waveform and disturbances due to the recovery
current [15]. The drives contain another source of EMI, i.e. the switch-mode
power supply (SMPS) which has high switching speed (dv/dt). This device
is analysed later in this work. The disturbances from a drive (rectifier and
inverter) propagate along the motor and the supply cable and may then be
coupled into other equipment either by conduction or radiation from these
cables. At higher frequencies (above 30 MHz), the radiation from the drive
itself tends to be more significant. The conducted disturbances are more
likely to cause further problems [17]. These problems may be avoided by
using:

1. **Good installation practices such as:**
   - Segregating control and signal cables from the noisy mains and motor
cables.
   - Shielding of control and signal cables.
   - Grounding of shielded cables at the end where grounded reference exists.
   - Using shielded or armored cables for the connection between drive and
   motor.

2. **Suppression of noise.**
   - Using an interference suppression filter at the input of the drive.
   - Installing a low impedance bond between the case of the filter, the frame
   of the drive and the shield of the motor cable.

It is important to remember that shielding without filtering can produce an
increase in disturbance, while filtering without shielding will provide poor
attenuation at high frequencies. The filters can be installed at the input of the
drive as shown in figure 1.6. Two common mode chokes were used with an

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*MASTER OF ENGINEERING (HONOURS)*
Figure 1.4. Circuit diagram of AC drive.

Figure 1.6. Input filter

Figure 1.5. Modeled circuit of the motor and cable
Y class capacitor placed on the load side of each choke. For suppression of differential mode emissions, X-class capacitors were fitted. The leakage inductance of the common mode chokes does provide some differential mode inductance.

**PWM AC Variable Speed Motor Drive.** AC variable speed drives enable a standard AC motor to operate over a wide speed range by providing a sinusoidal output with variable frequency, typically in the range 0 to 100Hz. This is achieved by rectifying the incoming power to dc, followed by a PWM inverter to give a three-phase output of controllable voltage and frequency. A typical illustration of such a drive is provided in figure 1.7. There is no EMC product standard for variable speed drives because they are usually incorporated into complete systems [15]. The drive output comprises a pulse width modulated rectangular wave with magnitude in the range 300-800V and rise/fall times in the range of 50-500ns. However, there is a need to filter the drive power input because of the high currents and high power density of the drive. Conducted emissions are decomposed into a common-mode component that flows out through the phase and neutral conductors and returns on the earth wire, and a differential-mode component that flows out through the phase conductor and returns on the neutral conductor.

**Common mode emission.** It is well known that the common-mode emission is dominant and it is the most difficult to filter because of the constraint of earth leakage. A common-mode model, shown in figure 1.8, allows the emission level to be estimated.
Figure 1.7. Typical arrangement of an AC inverter drive

Figure 1.8. Equivalent circuit for common-mode emission.

Figure 1.9. Equivalent circuit for series-mode emission.
Series-mode emission (differential mode). It is also necessary to check series-mode attenuation, especially as the larger value capacitors connected between the lines are physically large. Figure 1.9 illustrates the series-mode. Hargis [19] calculated the estimated emission levels for the above two models and revealed that the common-mode emission is predominant.

Filtering. Power line filters are employed at the power mains input of electrical or electronic devices to attenuate EMI. These passive devices are composed of inductors and capacitors arranged to form a filter. A modeling exercise allows emission to be estimated in the case of no filtering [18]. Figure 1.10 shows the basic circuit and figure 1.11 shows the common-mode model with allowance made for the parasitic element of the components. Figure 1.12 shows the series mode-model with an attenuation that depends on the leakage inductance of the current compensated inductors.

From the equivalent circuits it is clear that the main emission characteristics are determined by the following parameters [19]:

- Drive switching frequency
- Motor and cable capacitance to earth
- DC link capacitor (and circuit stray series inductance and resistance).

1.2.4. PWM Inverter with Low Electromagnetic Noise.

The progress of switching devices (MOSFETs, IGBTs) now allows designers to develop a voltage source PWM inverter with a higher switching frequency. This gives a great number of benefits in limiting harmonic voltages and currents. In addition, if the switching frequency is over 20 kHz,
Figure 1.10. Filter circuit

Figure 1.11. Common-mode equivalent circuit of drive with filter.

Figure 1.12. Series-mode equivalent circuit of drive with filter.
the acoustic noises can be totally eliminated. On the other hand a high frequency PWM may lead to more switching losses and more electromagnetic noises. An alternative approach uses power converters with a soft switching technique. A Zero-Current-Switching Based PWM Inverter (ZCS) having resonant circuits on the AC-side will reduce switching losses and noises (see figure 1.13) [20].

The switching device is turned on and off by keeping the size of the resonant current higher than the load current. In addition, the amplitude of the resonant current can be controlled independently because the neutral point of the resonant circuits is connected to that of the DC link.

1.2.5. Switched mode power supplies (SMPS) and EMI.

Switched mode power conversion has many advantages, which make it very desirable, but its major drawback is the generation of high frequency electrical noise associated with the fast switching waveforms in the power regulator [21]. The block diagram of a typical SMPS is shown in figure 1.14.

SMPS continually switch the current flow, thus are potentially a major source of EMI. The square wave output of an SMPS is designed to have very steep rise and fall times to minimise the transistor switching loss, but the steeper the edges the greater the harmonic content, and the greater the likelihood of EMI being a problem.

In a simple SMPS as shown in figure 1.15, it is possible to identify the areas that give rise to EMI by considering the changes in voltage and current in the circuit. The area around the switching transistor has high values of \( \frac{dv}{dt} \) associated with it, while the conductor from the reservoir capacitor...
Have high di/dt values. Rectifier diodes represent another possible source of interference since it gives rise to discontinuous change in current and hence high di/dt values. Figure 1.16 shows the potential EMI problem areas in a switching supply.

1.3. Reduction of EMI Effects.

For EMI to be a problem in any electronic system there are three key elements.

1. There must be a source of interference.
2. There must be a transmission medium.
3. There must be a susceptible receiver.

To eliminate a noise problem there are steps that should be taken. Firstly the noise should be reduced at the source by reducing the values of dv/dt and di/dt in the circuit if possible [22]. Attention should also be paid to reducing stray electrical and magnetic fields by careful layout. In addition, conduction noise paths should be broken using frequency filters and finally, radiation should be contained by the use of screens.

1.3.1. Reduction of EMI at source.

The addition of small series inductors to limit di/dt and small shunt capacitors to limit dv/dt will be a solution for the problem of interference and can reduce it at the source, but it will also result in a slight increase in the power dissipated in the switching device. The physical layout of the
Figure 1.13. Configuration of Zero Current Switching Based PWM inverter

Figure 1.14. Block diagram of SMPS.

Figure 1.15. Simple SMPS circuit.

Figure 1.16. EMI problem areas in an SMPS.
supply can play a greater role in minimising radiation, and all leads carrying rapidly switching currents and voltages should be kept as short as possible as shown in figure 1.17. In addition, high current and high voltage should be kept away from the sensitive parts of the circuit [14].

1.3.2. Grounding.

Two points should be borne in mind when considering grounding:
1. Separate ground points on a chassis are seldom at the same potential.
2. All conductors and wires have finite impedance.

The first point means that a number of common impedance paths are introduced to earth returns for units 1,2 and 3. (the case of a common single point earth), as shown in figure 1.18. This can result in excessively noisy grounds especially in high power stages. At low frequencies where wiring inductance is not a problem, the solution is to use a single point grounding. For frequencies above 1MHz, a low impedance ground plane is a good practice especially on printed circuit boards. Here multi point grounding is the preferred method, provided the ground connections are kept short.

1.3.3. Filtering of EMI.

A useful component for filtering power supply lines between equipment is the bifilar wound choke. It consists of bifilar winding placed on a core to form a broadband transformer, which allows equal and opposite currents to flow through its windings while suppressing unequal opposite currents such as ground noise. Bifilar chokes are used to break ground loops as shown in figure 1.19. In addition, ferrite beads are a simple and convenient way to
Figure 1.17. Example of good and bad wiring.

BAD GROUNDING PRACTICE

(a)

GOOD GROUNDING PRACTICE

(b)

Figure 1.18. a) Series connections-single point  b) Parallel connection-single point
increase the high frequency loss in a conductor [23]. They are most effective above 1MHz and when properly used can give effective high frequency decoupling and shielding.

**Input filters.** The filtering at the input must be capable of blocking the switching frequency ripples and its entire harmonics well into the MHz region of the spectrum. This ensures that no interference is fed back into the primary power supply. A typical input filter, which meets most input noise specifications, is shown in figure 1.20. In this input filter, two bifilar or common mode, chokes are used along with four bypass capacitors. The values of the chokes depend upon the switching frequency and the amount of attenuation required.

**Output filter.** At the output, the filtering problem is normally easier as there is already a low-pass filter to smooth the output current. At harmonics in the MHz range, where EMI is usually a problem, the stray inductance of the capacitor will make it ineffective as a bypassing component. The solution to this problem is to use two filter sections, one to perform low frequency filtering at the switching frequency and a second section to remove high frequency components as shown in figure 1.21. The switching frequency filter is placed across the line and neutral, while the EMI filter is placed from each line to ground. In this way the high frequency components are bypassed to ground, while the low frequencies are bypassed from line to line [15].
Figure 1.19. Equivalent circuit of bifilar choke

Figure 1.20. A typical input filter.

Figure 1.21. LC output filter with high frequency bypass component
1.3.4. Screening.

By paying careful attention to screening and grounding, it is possible to achieve the prevention of radiated interference from the supply entering into surrounding systems and wiring. When planning the screening for a supply, it is necessary to know the type of field that is being shielded against. The characteristics of the field are determined by the nature of the source. A source with a high current and low voltage will have a near-field that is predominantly magnetic while one that has a high voltage and low current has a predominantly electric field. The electric and magnetic fields must be considered separately [24]. When planning enclosures and shields, the following points should be taken into account:

1. Magnetic fields are harder to shield in comparison to electric fields.
2. Use good conductors when dealing with electric fields, plane waves and high-frequency magnetic fields.
3. Actual screening effectiveness obtained in practice is usually determined by the leakage at the seams and joints, not by the material itself.
1.4. References


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CHAPTER 2

AC DRIVES AND PWM INVERTERS.

2.0. AC drives.
This project aims to study the conducted emission produced by PWM inverters. It is essential first to understand the operation and control of AC drives. AC motors require control of frequency, voltage and current for variable speed operation, but these power controllers require advanced feedback control techniques. They are relatively complex and more expensive. There are two types of AC drives:
• Induction motor drives
• Synchronous motor drives
In this chapter, various AC drives are discussed without any complex mathematical derivations.

2.0.1. Induction motor drives.
An induction motor has three phase stator and rotor windings. The stator winding produces induced voltage in the rotor windings, due to transformer action, when supplied with balanced three-phase AC voltage as shown in figure 2.1. The equivalent circuit is shown in figure 2.2. An appropriate
Figure 2.1. Cross sectional view of two pole squirrel-cage induction motor.

Figure 2.2. The equivalent circuit of induction motor.
distribution of stator winding creates the effect of multiple poles, which produce several cycles of magnetomotive force (mmf) around the air gap. This field will establish distributed flux density in the air gap [1-2]. Generally speaking, the control of speed and torque of induction motor is possible by varying or controlling the following parameters:

- Stator voltage
- Rotor voltage
- frequency
- Stator voltage and frequency
- Stator current
- Voltage, current, and frequency.

For more information refer to Appendix A.

2.0.2. Synchronous motors.

The stator of a synchronous motor has a polyphase winding as well as a field winding, which carries a dc current on the rotor. The field current and the armature current create two mmfs and the torque is produced by the resultant mmf. Induction and synchronous motors have similar stators but their rotors are different. Synchronous motors rotate with zero slip, the number of poles and frequency dictates the synchronous speed [1]. One of the advantages of synchronous motors is that it can be used as a motor or generator, and a PWM inverter or a cycloconverter can control it. There are three types of synchronous motors:
• DC excited synchronous motors.
• Reluctance synchronous motors.
• Permanent-magnet synchronous motors.

Reluctance synchronous motors are similar to the salient-pole synchronous motors, but they do not have field windings on the rotor. Induced field in the rotor, by the armature circuit, tends to align with the armature field. This type of motors is used where a number of motors are required to rotate in synchronism. While modern reluctance motors can be operated as DC brushless motors, the permanent-magnet motors are also similar to that of salient rotor motors, except the DC excitation is provided by permanent magnets. A cross sectional view of a variable reluctance motor, and permanent magnet motor are shown in figures 2.3 and 2.4 respectively.

2.1. Frequency-Controlled Motor Drives.
The AC drives have many applications in traction, pumps, fans, compressors, conveyers, machine tools, and so on. This popularity is associated with the ability of variable frequency induction motor to function in a maintenance free operation and in applications involving explosive and contaminated environment. Variable speed drives are able to compete with dc drives in medium and lower range high power drives due to the progress in power switching devices[4]. Variable frequency drives employ different power semiconductor converters, mainly:
• Voltage source inverters
• Current source inverters
• Cycloconverter
Figure 2.3. Cross sectional view of doubly-salient variable reluctance motor.

Figure 2.4. Cross sectional view of DC permanent magnet motor.
An inverter is called voltage source inverter, if seen from the load side, its DC terminal function as a voltage source. If its terminals function as a current source this inverter will be called current source inverter. A low impedance voltage source inverter will operate as a single motor drive or a multi motor drive. This is due to its ability to sustain a constant voltage source with variation of a load, while any change in load conditions leads to a variation of voltage source in current source inverter because of its high internal impedance. Hence, current source inverters are not used for multi-source drives. Using a fixed frequency voltage source cycloconverter can supply variable frequency with voltage or current source.

2.2. PWM Inverters.
PWM inverters are the most important class of inverters for motor drivers and most commonly employed [5]. PWM inverters are meant to control the output voltage, however, they can be supplied either by AC voltage or fixed dc voltage as shown in figure 2.5. In the case of AC supply, the inverter is supplied from the diode bridge. The output voltage has a low harmonic content, which makes the drive response faster. The low harmonic content leads to a smooth low-speed operation, lower torque pulsation and cogging, and higher efficiency [5]. On the other hand, PWM inverters have a few disadvantages such as complex control and high switching loss because of high frequency operation of the switches. The power circuit is shown in figure 2.6. There are a number of configurations of PWM. Most common of them are PWM with uniform sampling, sinusoidal PWM, PWM with selective
Figure 2.5. PWM inverter drives.

Figure 2.6. Three phase voltage source inverter.

Figure 2.7. Generation of control signal.
harmonic elimination, and so on. The sinusoidal PWM is the scheme used in the present study and therefore will be explained in some detail.

2.2.1. Sinusoidal PWM inverter.

The generation of a control signal for sinusoidal PWM is illustrated as shown in figure 2.7. The voltages, $V_a$, $V_b$, and $V_c$ are reference voltages with variable amplitude $A$. Three comparators compare these voltages with a common isosceles triangular carrier wave, $V_r$, which has a fixed amplitude $A_m$. These comparators supply the inverter by control signal at switch pairs (S1,S4), (S3,S6), and (S5,S2). For example the switch pair (S1,S4) controls the voltage of the machine phase A with respect to the imaginary middle point of the DC source, 0. The other pairs of switches operate similarly. The production of the line voltage waveform, $V_{ab}$, when each cycle of the reference wave has 5 cycles of triangular wave is shown in figure 2.8.

The reference waves ($V_a$, $V_b$) and the carrier wave $V_r$ are drawn on a common time axis for a positive full cycle of $V_a$. When $V_a>V_r$, switch S1 receives the signal, and when $V_a<V_r$, switch S4 receives the signal. The resultant waveform, $V_{ac}$ is also shown. Similarly the other resultant voltages, $V_{bo}$, and $V_{ca}$ are drawn. The line voltages, $V_{ab}$, $V_{bc}$, and $V_{ca}$ are obtained from the resultant voltages ($V_{ab}$-$V_{bc}$), ($V_{bc}$-$V_{ca}$), and ($V_{ca}$-$V_{ab}$). The modulation is called sinusoidal PWM, because the pulse width is determined by its angular position in the cycle and the relation between width and angular position is sinusoidal [6].

The motor terminal voltages have the same frequency as the reference voltages which means that by changing the frequency of reference
voltage, the frequency of the motor voltages can be controlled (Appendix B).

Instead of maintaining the width of all pulses the same as in the case of multiple pulse modulation, the width of each pulse is varied in proportion to the amplitude of a sine wave evaluated at the center of the same pulse. The distortion factor and lower-order harmonics are reduced significantly. This type of modulation is used industrial applications.

The rms output voltage can be varied by varying the modulation index $M$. It can be observed that the area of each pulse corresponds approximately to the area under the sine wave between the adjacent midpoints of off periods on the gating signal. If $\delta_m$ is the width of $m$th pulse, the rms voltage can be found from the following equation:

$$V_0 = V_r \left( \sum_{m=1}^{p} \frac{\delta_m}{\pi} \right)^{1/2}$$  \hspace{1cm} (2.1)

where $p$ is the number of pulses per half cycle and is found from:

$$p = \frac{f_c}{2f_0}$$  \hspace{1cm} (2.2)

The general form of a Fourier series for the instantaneous output voltage is:

$$v_o(t) = \sum_{n=1,3,5,\ldots}^{\infty} B_n \sin n\omega t$$  \hspace{1cm} (2.3)

The Fourier coefficient of output voltage can be found from:

$$B_n = \frac{2V_r}{n\pi} \sin \frac{n\delta_m}{2} \left[ \sin n\left(\alpha_m + \frac{\delta_m}{2}\right) - \sin n\left(\pi + \alpha_m + \frac{\delta_m}{2}\right) \right]$$

for $n = 1, 3, 5, \ldots$  \hspace{1cm} (2.4)
This type of modulation eliminates all harmonics less than or equal to $2p-1$. For $p=5$, the lowest order harmonics is ninth. The distortion factor is significantly reduced compared to that of multiple-pulse modulation.

The output voltage of an inverter contains harmonics. The PWM pushes the harmonics into a high-frequency range around the switching frequency $f_c$ and its multiple, that is, around harmonics $m_f, 2m_f, 3m_f$, and so on. The frequency at which the voltage harmonics occur can be related by:

$$f_n = (jm_f \pm k)f_c$$  \hspace{1cm} (2.5)

where the $n$th harmonic equals to the $k$th sideband of $j$th times the frequency-modulation ratio $m_f$.

$$n = jm_f \pm k$$

$$= 2jp \pm k$$ \hspace{1cm} for $j=1, 2, 3$, and $k=1, 3$, \hspace{1cm} (2.6)

The peak fundamental output for SPWM can be found approximately from:

$$V_{m_1} = dV_s$$ \hspace{1cm} for $0 \leq d \leq 1$ \hspace{1cm} (2.7)

For $d=1$, Eq.(2.7) gives the maximum peak amplitude of the fundamental output voltage. But $V_{m_1 (max)}$ could be as high as $1.278V_s$. In order to increase the fundamental output voltage, $d$ must be increased beyond 1. The operation beyond $d=1$ is called overmodulation. Overmodulation basically leads to a square-wave operation and adds more harmonics as compared to operation in the linear range.

2.3. Reduction of EMI in AC inverters.

One of the major problems facing the development of drives is the requirement to reduce the amount of EMI caused by these drives as discussed in chapter 1. However, a number of methods have been developed recently in order to achieve this reduction. Some of these methods are described below [7].
Figure 2.8. Principle of PWM
2.3.1. Soft Switching Phase-Shifted PWM High-Frequency Inverter.
Yonemori, and Chibani [8] proposed a newly improved sinewave CVCF power conversion conditioning system (PCCS) based upon a high frequency (HF)-isolated AC- link cycloconversion circuit using soft-switching, phase shifted PWM (PSM) and instantaneous voltage regulation schemes. The circuit topology of a voltage-clamped quasi-resonant (QR) switch concept which comprises of two cascaded power conversion stages operating under the principle of zero-voltage switching (ZVS), is introduced in order to minimise switching losses, device stresses, remove a complicated HF-AC snubber and minimise EMI noise.

The schematic circuit configuration of a 1st phase primary side sinewave PSM regulated HFAC inverter link cycloconverter cascaded system is shown in figure 2.9. The bi-directional power conversion processing is accomplished through two stages. Firstly the input DC is converted into a sine wave PSM-HFAC in the inversion power stage using a soft switched 20 kHz inverter with voltage clamped QR circuits and lossless capacitive snubbers. Secondly the conversion of sine waves PSM-HFAC voltage into sine wave voltage is accomplished using a soft switching commutated cycloconverter.

2.3.2. Switching Frequency Modulation.
According to Feng Lin, and Dan Chen [7], it is possible to modify noise emission spectrum by modulating the PWM frequency of the power supply. In the case of PWM, the emission is periodical with respect to the switching frequency. This emission centers at the switching frequency and the harmonic frequencies. This particular concentration of emission power at these discrete frequencies will lead to noticeable EMI. In order to
reduce EMI level, modulating the switching frequency creates side bands and the emission spectrum is thus broadened. The power is distributed into smaller pieces scattered around many side-band frequencies as shown in Figure. 2.10. The extent of distribution and the amplitude of the spectrum depend on the modulation index. The larger the modulation index, the more evenly is the spectrum distributed.

Figure. 2.11 shows an example of a drive in which a switching frequency modulation has been used [7]. Frequency modulation is particularly effective in reducing the fundamental harmonic, if the switching frequency is less than 150 kHz as shown in figure 2.12. The common mode emission and the differential mode emission can be reduced by frequency modulation, but the emission generated by ringing cannot be reduced by this approach. Snubber circuits and the use of soft diodes solve this problem. The EMI filter must be properly damped to avoid the possible amplification of EMI at the filter pole frequencies.

2.3.3. A Zero Current Switching Based Three Phase PWM Inverter.
Fujita and Akagi [9] investigated this approach and concluded that a zero current switching based three-phase PWM inverter is characterised by the following:

- A soft switching based PWM inverter can greatly reduce the switching losses and electromagnetic noise, compared with a conventional hard switching inverter.

- A soft switching based PWM inverter can drive AC motors without any restrictions such as a conventional three phase voltage source PWM.
Figure 2.9. Schematic circuit diagram of HF-link converter.

Figure 2.10. Spectra of unmodulated and modulated waveforms.
Figure 2.11. (a) Drive circuit. (b) Variable switching frequency.
Figure 2.12. The effect of frequency modulation on spectrum [7].
2.4. References.

CHAPTER 3

Test Bench Development.

3.0. Introduction.

This chapter aims to explain the experimental procedure, which has been adopted in the present work for measurement. In addition an explanation of certain aspects of the procedure are discussed when necessary. Basically the project deals with 3 areas as follows:

- Improving the methodology of measurement by using a simulation technique (graphic language (LabVIEW)).
- Measuring the conducted emission (voltage, current) and determining the best indication of interference.
- Measuring the conducted emission of three systems, PWM inverter + AC motors (Permanent magnet, Induction, Reluctance).


The undesirable coupling of electromagnetic energy out of an emitter or into a receptor, via any of its connecting wires or cables, is termed conducted electromagnetic interference.
The generated emission is conducted out of the system along an AC power cord. Since the commercial power distribution system is a large array of wires connecting the various power outlets it represents a large antenna. The conducted emission can radiate quite efficiently from this antenna, and probably cause interference in all electronic systems connected to this installation [1]. The fact that the conducted emission has only one path to propagate along, makes it easier to reduce its effect.

EMI specifications state limits on conducted EMI in various ways. Some specifications express it in terms of the magnitude of EMI current flowing out of equipment under test through conductors, while other specifications express it in terms of the two terminal voltages where a conductor emerging from the equipment under test interfaces with the outside installation.

Experimental testing plays an important role since it allows us to determine the EMI emission produced by a product and compare the results against specified limits. In this study the reference limits used are discussed and the choice is based on the information provided by Standards Australia, the International Special Committee on Radio Interference (CISPR), and the Federal Communications Commission (FCC), USA. Another point to note is that compliance with regulations is becoming obligatory in most industrialised countries thus requiring a serious consideration even at design level [2].

3.1.1. Standards and Limits.
It is clear from the latest publication of Standards Australia that the trend
is toward adopting the regulations provided by CISPR where they are applicable [3].

The test standard AS3548, which is identical to corresponding International Standard CISPR 22, has been used as the reference for the present study. In addition, the FCC conducted emission limits and the CISPR 22 (average) are also used. According to FCC [4], digital devices separate into two classes:

1. Class A: includes digital devices produced for use in a commercial, industrial or business environment.
2. Class B: includes digital devices produced for use in residential environment.

The CISPR also provides a similar separation, but the differs in its frequency range of applicability. The FCC limits begin from 450 kHz while the CISPR 22 conducted emission limits extends down to 150 kHz [5]. The reason behind this extension is the need to cover the emissions of power supplies, which employ switching technology. The break frequency of the CISPR 22 class A limit is at 500 kHz. Both extend to an upper limit of 30 MHz. The FCC and CISPR 22 limits on conducted emission are compared in figure 3.1. These limits are provided when the receiver uses a quasi-peak (QP) detector and when the receiver uses an average (AV) detector.

The document MIL-STD-461 dealing with the EMC requirement for military use produced in the USA will be used. The advantage of the MIL-STD-461 document is that it covers a considerably larger class of products than FCC and CISPR 22 [6]. The document MIL-STD-461 (UM05) deals with the conducted and radiated emissions produced by commercial electrical and electromechanical equipment. Figure 3.2. illustrates the
Figure 3.1. The CISPR conducted emission limit compared to the FCC limit.

Figure 3-2. MIL-STD-461 UM05 conducted emission for commercial equipment
UM05 limits for broadband conducted emission for commercial electrical and electromechanical equipment. It is worth noting that the limits for residential equipment are more stringent than in the commercial case since interference in the industrial environment can be more readily corrected.

3.1.2. Signal Spectra.
This study deals with the spectral content of the signal (voltage, current) and it is important to provide insight into the time domain factors that affect their spectral content. A quick review of the Fourier transform technique is needed to understand the ideas behind the experimental results. For a pulsed waveform, the key parameters that contribute to the high frequency spectral content of the waveform are the rise and fall times of the pulse. Pulses having short rise/fall times will have large high frequency spectral content compared with the pulses having long rise/fall times. Therefore, to reduce the high frequency spectrum and hence to reduce the emission of a product, an increase in the rise/fall times of data pulses is necessary.

Any slight variations in duty cycle from an exact 50% can cause the even harmonic level to vary significantly. Reducing the duty cycle (pulse width) reduces the low frequency spectral content of the waveform, but does not affect the high frequency content [1]. Parasitic inductance or capacitance can cause a phenomenon called ringing. Usually, there is a resonant region of enhanced emission in a narrow frequency band, due to the undershoot/overshoot present on the digital waveforms. Thus the elimination of ringing is crucial for EMC purposes.
Suppose \( \{h(n)\} \) is a discrete time sequence. The following equation (3.1) defines the Fourier transform of the discrete time sequence.

\[
H(e^{j\omega}) = \sum_{n=-\infty}^{\infty} h(n)e^{-j\omega n}
\]  

(3.1)

Discrete Fourier transform (DFT) is the finite duration discrete frequency sequence that is obtained by sampling one period of the Fourier transform. This sampling is conventionally performed at \( N \) equally spaced points over the period extending over \( 0 \leq \omega \leq 2\pi \).

If \( \{h(n)\} \) is the discrete time sequence with the Fourier transform then the DFT \( \{H(k)\} \) is defined as

\[
H(k) = H(e^{j\omega})_{\omega = \omega_k = 2\pi k/N}
\]

(3.2)

DFT starts at \( k=0 \) corresponding to \( \omega = 0 \) but does not include \( k=N \) corresponding to \( \omega = 2\pi \) [7].

### 3.2. Test Apparatus and Setup.

The object of the current study is to investigate emissions that are generated and conducted out of the product along an AC power cord. The measurement setup is identical to the recommended setup provided in document MIL-STD-461. It is important to note that there are different ways to measure conducted emissions. The most common are:

1. conducted EMI current measurement.
2. conducted EMI voltage measurement

Measuring conducted EMI current is possible using a clamp-on current probe. This probe works on the principle of treating the conductor under test as the primary winding of a transformer with the current probe acting
s toroidal wound secondary winding. The clamp-on current probe is
talent to measuring the voltage drop across a small resistance inserted
eries with the current being measured [8]. If an EMI analyser is
rated to read the voltage, such voltage levels require knowledge of the
fer impedance of the current probe, which is the ratio of output voltage
turn. It can be stated as:

\[ Z_r = \frac{V_{out}}{I_{in}} \]  \hspace{1cm} (3.3)

unknown EMI current in dBμA is:

\[ V_{out} - Z_r \]

\[ V_{out} \text{ in } dB\mu V \text{ and } Z_r \text{ in } dB\Omega \].

Ire 3.3 illustrates the principle of current probe.

ther way of measuring conducted emission of a product is measuring
ternal voltages appearing at that point where the conductor leading
of the equipment under test (EUT) interfaces with the outside world.
nts measured for verification of compliance with the regulatory
its are to be measured with a Line Impedance Stabilisation Network
SN) [9], sometimes called an “Artificial Mains Network”, inserted
ween the commercial power outlet and product’s AC power cord. An
uration of the use of an LISN is given in figure 3.4.
lock diagram of the test setup is given in figure 3.5. A LISN is inserted
ween the power outlet and product’s AC power cord, for phase and
ral conductor. The diagram shows two possible ways of measurement
using a current probe clamped around one conductor (phase, neutral)
the other measuring the voltage drop across conductors. In the case of a
tent probe, the signal is amplified and then driven to a digital
Figure 3.3. Current probe field around conductor [8].

Figure 3.4. The use of LISN in the measurement of conducted emission [8].
Figure 3.5. Block diagram of experimental layout.
oscilloscope which is connected to a PC through a General Purpose Interface Bus (GPIB). A program has been developed using the graphical programming language, “G”, which allows the transformation of the signal from the time domain to the frequency domain. Similarly in the case of voltage measurement, the signal is driven directly to the digital oscilloscope and then, using GPIB, the data is transferred to a PC to be transformed into the frequency domain.

3.3. Line Impedance Stabilisation Network (LISN).

As discussed earlier, the LISN is a buffer inserted between Equipment Under Test (EUT) and the power mains. LISN has multiple functions such as:

- Stabilisation of the impedance of the main power as presented to the EUT.
- Coupling to EMI analyser any EMI introduced into the power mains by the EUT.
- Preventing any EMI coming from the power mains to be measured by the EMI analyser [10].

The idea of LISN was developed in early 1950s when engineers needed to develop a means of providing a standard power supply impedance network for aircraft in an attempt to correlate equipment EMC bench tests with those carried out using actual aircraft supplies. Figure 3.6 illustrates the use of a LISN in the measurement of conducted emission. It may appear that the conducted emission can be simply measured with a current probe, but the requirement that the measured data be correlated between measurement sites makes this test unrealistic. Looking into the AC power system, the
impedance varies considerably over the measurement frequency range and also varies from outlet to outlet. Therefore it is likely to affect the measurement of conducted emission. Thus, LISN guarantees that the impedance seen by the product looking into the AC power system is stabilised from one site to another site.

Another issue related to LISN is that the amount of noise in the power system also varies depending on the site. There is a need to exclude this noise otherwise it will be added to the measured conducted emission [11]. Again the use of LISNs prevents external conducted noise on the power system from contaminating the measurement. LISN can vary widely depending on the requirement for testing for a specific standard. There are broadband and narrowband LISNs, and each document recommends a specific LISN type. In this study the choice is based on the need to start measurement from the lowest possible frequency (10 kHz) up to the highest (30 MHz), but it is impossible to do so with a simple LISN. Therefore the use of two LISNs is necessary. Figure 3.7 shows a schematic for the LISN used for testing from 10 to 150 kHz (Appendix D), while figure 3.8 shows the complementary LISN used from 0.15 to 30 MHz (Appendix E) [8]. The first LISN design as provided by [8] is required to meet tests CE02, CE04, CS02 in notice 3 of MIL-STD-462. The second LISN design is required for use in MIL-STD-462 N3, which is suitable for measurements over a higher frequency range up to 50 MHz. An analyse of these two circuits is provided in Appendix D, and Appendix E. Analysis were performed using Electronic workbench software.

To better understand LISN, consider for example the one, which used for measuring from 0.15 to 30 MHz. The purpose of the 1µF capacitor between phase and earth wires and between neutral and earth wire on the
Figure 3.6. The measurement of conducted EMI using LISN [8].

Figure 3.7. LISN for testing from 10 to 150 kHz [source: Reference 8].

Figure 3.8 LISN for testing from 0.15 to 30 MHz [source: Reference 8].
commercial power side is to divert external noise and prevent that noise from flowing through the measurement device. The purpose of the 50\(\mu\)H inductor is also to block the external noise. The other 0.1\(\mu\)F capacitor is to prevent any dc current from overloading the input of the test receiver. The 1K\(\Omega\) resistor acts as a static charge path to discharge the 0.1\(\mu\)F capacitors. One 50\(\Omega\) resistor represents the input impedance of the test receiver, while the other is a 50\(\Omega\) dummy load to ensure that the impedance between the phase and the earth wire, and between neutral and the earth wire is approximately 50\(\Omega\) at all times. Both the phase and the neutral voltages must be measured over the frequency range of the conducted emission limits. The LISN is mounted on a metallic ground plane alongside the EUT and the EMI meter is connected via a high performance low-loss coaxial cable to the measurement port. Any LISN measurement ports, which are not being used as part of the measurement are terminated with 50 ohms [1]. Malack [12], has shown that the mean value of actual noise impressed on the power distribution system by the EUT and the EUT noise measured with LISN inserted are essentially equal. In another study, Malack [13] has shown that the differences in power line interference measurement levels are caused by the use of different LISNs. The measurement difference factors can be determined in a similar manner for the class of equipment and other LISNs. Nicholson and Malack [14], recommend that a 50\(\mu\)H LISN, when used in its specified range (0.15 - 30 MHz), is a good representation of the mean commercial power line impedance. Consequently it is a good choice for standardisation among the various laboratories which perform conducted interference measurements.

Common-mode currents are not necessary for the functional performance of the system and they depend on non-ideal factors such as proximity to nearby ground planes and other metallic objects. However, they can be measured using current probes.

Ampere’s law states that a magnetic field can be induced around a contour by either conduction current or displacement current that penetrates the open surface $S$.

\[ \oint_c \mathbf{H} \cdot d\mathbf{l} = \oint_c \mathbf{J} \cdot d\mathbf{s} + \frac{d}{dt} \int_c \mathbf{E} \cdot d\mathbf{s} \]  \hspace{1cm} (3.5)

where $c$ is the contour bounding the open surface

- $\mathbf{J}$ is the current flux density
- $\mathbf{E}$ is the electric flux intensity
- $\mathbf{H}$ is the magnetic field intensity

Figure 4.9 illustrates Ampere’s law and the use of a current probe.

A time changing electric field produces a displacement current in the medium in which it is operating. If no time varying electric field penetrates this surface, the induced magnetic field is numerically equal to the conduction current passing through the loop. Physically, a current probe is formed from a core of ferrite material separated into two halves, which are joined by a conducting hinge and closed with a conducting clip. The core is placed around the wire whose current is to be measured. The current produces a magnetic field, which circulates around the core. A few wires are wound around the core, so that any circulating time changing magnetic field induces an emf proportional to this magnetic field. Therefore the induced voltage is proportional to the current passing through the probe.
The ratio of the voltage $V$ to the current $I$ is related through a calibration curve

$$\hat{Z}_r = \frac{V}{I}$$

$Z$ is called the transfer impedance of the current probe and is in units of ohms. The calibration chart could be given in dB as

$$|\hat{Z}_r|_{dB} = |V|_{dB} - |I|_{dB}$$

In this study, the current probe used is a Tektronix type A6302. The probe manufacturer provided a calibration chart, which shows the magnitude of the transfer impedance versus frequency. Figure 3.10 shows the chart of transfer impedance. The calibration curve is valid only when the probe is terminated in the same impedance as used in the course of its calibration [8].

3.5. General Purpose Interface Bus (GPIB).

The GPIB is a link, or interface system, through which interconnected electronic devices can communicate. The GPIB carries two types of messages:

- Data messages which contain device specific information e.g. programming instructions, measurement results, machine status, and data files.
- Command messages, which perform tasks such as initialising the bus, addressing devices, and setting device modes for remote programming.

GPIB devices can be talkers, listeners, or controllers. The controller addresses a talker and a listener before the talker can send its message to
Fig 4.9. (a) Illustration of Ampere's law (b) the use of current probe.

Fig 3.10. Current probe transfer impedance [15]
the listener. A controller is necessary when there is a need to change the active or addressed talker or listener (Appendix C).

3.6. Program design.
The actual programming task is to build a program, which serves as a measurement tool using the Laboratory Virtual Instrument Engineering Workbench (LabVIEW). The measurement tool needed for this experiment is the one that shows the power spectrum of measured the EMI signal, i.e. the Spectrum Analyser. The use of a spectrum analyser (hardware) can be easier and more straightforward but the proposed tool of measurement has a lot of advantages such as:

- The cost of this measurement tool is far less than a dedicated commercial spectrum analyser.
- The possibility of using this tool by more than one measuring system (using networking) makes a difference especially in industry and in quality control.
- Another advantage of using this measurement tool is its flexibility since in some measurements a series of considerations should be taken such as data transfer, further signal processing, the need to test different devices simultaneously, and so on.

This program implements the same procedure used in the most advanced Spectrum Analysers which eliminates the possibility of obtaining misleading results.

A brief explanation of LabVIEW is needed before going through the particular use of this programming system in this experiment. LabVIEW is a program development application. LabVIEW programs are also called virtual instruments (VIs) since their appearance and operation imitate
actual instruments. It uses a graphical programming language, called (G), to create programs in block diagram form instead of using text based languages to create lines of code. This programming system relies on graphic symbols rather than textual language to describe programming actions. LabVIEW includes conventional program tools, it is possible to animate the execution to see how data passes through the program. VIs has both an interactive user interface and a source code equivalent, and accept parameters from higher level VIs. VI has three main parts: the front panel, the block diagram, and the icon/connector.

The interactive user interface of a VI is called the front panel, since it simulates the panel of a physical instrument. The front panel can contain knobs, graphs, and other controls and indicators. Data can be inserted using a keyboard or mouse and the results can be viewed on the computer screen.

A block diagram, constructed in language G, instructs the VI. The block diagram is considered the source code for the VI. VI can be used as a top-level program, or a subprogram within other programs or subprograms, because VIs is modular and hierarchical.

An application can be divided into a series of tasks, which can further be divided into a series of simple subtasks again to make a complex application. Each subtask can be built as VI and then the combination of those VIs on a different block diagram accomplishes the larger task. Many low level sub VIs often perform tasks common to several applications, which make it possible to develop a specialised set of sub VIs suited to applications needed to construct. The icon and connector of a VI work like a graphical parameter list so that other VIs can pass data to it as sub VI [14].
3.6.1. Building of EMI spectrum analyser VI.
The reason for using G language is to construct a spectrum analyser, which will be able to measure the power spectrum of EMI waveform and also to complete the measurement of current and voltage waveforms. Hence, two VIs are constructed, one for measuring the power spectrum of the current waveform and the other the power spectrum of the voltage waveform. These two VIs are slightly different but operate on the same principle. During the measurement of the current waveform (using a current probe and inverse amplifier) the vertical gain of the oscilloscope, on which the signal is visualised, should be set to 10 mV/div. This means that the amplitude of current waveform is controlled by the variable current division of the inverse amplifier [15]. In fact, the use of Case function in G programming language makes it possible to overcome the difference between the measurement of voltage and current waveform and enables the use of one VI for different measurements. The complete program is based on a VI developed by National Instruments and is called a “Simple Spectrum Analyser”. National Instruments provided this as an example, which shows how to construct a simple spectrum analyser using the measurement VIs. This example simulates waveform and computes the autopower spectrum and the power of the peak frequency component. The block diagram of a simple spectrum analyser VI is shown in figure 3.11.
In developing the VI, it is necessary to set a link between the measurement VIs and an oscilloscope. Fortunately, National Instruments supply LabVIEW with different instrument drivers e.g. a driver for TDS 320 oscilloscope which is needed for this experiment. In addition, a modification has been done to the block diagram by introducing the case
function to be able to measure the voltage and the current. The front panel has been changed as well so that the graph shows results in desired units and the block diagram can be controlled totally from the front panel. The current probe division is also controlled via the front panel. This VI is called “EMI spectrum analyser voltage/current”. The connector pane, front panel, list of sub diagrams, block diagram, and the hierarchy diagram of the “EMI spectrum analyser (voltage/current) VI” are provided in figures 3.12.(a), (b), (c), (d), and (e) respectively. The complete VI uses 5 sub VIs: from the LabVIEW library:

- Tek TDS 3xx Example VI
- Scaled Time Domain Window VI
- Auto Power Spectrum VI
- Spectrum Unit Conversion VI
- Power & Frequency Estimate VI

The “Tek TDS 3xx example VI” file produces an analogue output for the Tek TDS oscilloscope. The use of this VI is possible by connecting a probe from the channel 1 input to the probe compensation connector, then by pressing the GO arrow. After initialising and configuring the instrument (first run only), the VI reads and graphs the channel 1 waveform and makes automated measurements on it. On subsequent runs, the waveforms are measured without reconfiguring the instrument. The front panel, connector pane, and the block diagram of this VI are shown as figures 3.13.(a), (b), and (c) respectively.

The “scaled time domain window VI” file applies a scaled window to the time-domain signal. The windowed time domain signal is scaled so that the power or amplitude spectrum of the windowed waveform is computed
with all windows providing the same level within the accuracy of the window. This VI also returns important window constants for the selected window. These constants are useful when VIs are used to perform computations on the power spectrum. The waveform is the real time domain signal. The block diagram is shown in figure 3.14.

The "Auto Power Spectrum. VI" file computes the single sided, scaled auto power spectrum of a time domain signal. The power spectrum is computed as:

\[ \text{FFT}(X) \times \frac{\text{FFT}^*(X)}{N^2} \]

Where \( N \) is the number of points in the signal array \( X \) and \( * \) denotes complex conjugate. Signal \( (V) \) is the array containing the time domain signal. \( dt \) is the sample period of the time domain signal and \( dt=1/F_s \) where \( F_s \) is the sampling frequency. The power spectrum \( (V^{2\text{rms}}) \) is the single sided power spectrum in volts RMS squared if the input is in volts or in amperes RMS squared if the input signal is in amperes. The block diagram of this VI is shown in figure 3.15.

The "power & frequency estimate VI" file computes the estimated power and frequency around a peak in the power spectrum of a time domain signal. This VI allows good frequency estimation for measured frequencies that lie between frequency lines on the spectrum to be achieved. The block diagram of this VI is shown in Fig 3.16.
Figure 3.11. Block diagram of simple spectrum analyser VI.
Figure 3.12. EMI spectrum analyser (voltage/current).VI. (a) connector pane, (b) front panel, (c) list of sub diagrams.
Figure 3.12. (d) Block diagram of EMI spectrum analyser
(voltage/current) VI.
Figure 3.12. (e) The Hierarchy diagram of EMI spectrum analyser (voltage current) VI.
Figure 3.13. Tek TDS 3xx example.VI (a) front panel, (b) connector pane.
Figure 3.13. (c) Block diagram of Tek TDS 3xx example. VI.
Figure 3.14. Block diagram of scaled time domain window. VI.
Figure 3.15. Block diagram of auto power spectrum. VI.
Figure 3.16. Block diagram of power and frequency estimate. VI.
3.8. Technical summary.

Equipment and software used in this experiment are:

- Current probe system (AM 503S). It includes current probe A602 and current probe amplifier AM 503 A.
- Oscilloscope Tektronix TDS 320 (100 MHz)
- Single phase drive (ZENER MSC-S), (TOSHIBA VF-S3).
- Motors (Induction, Permanent Magnet, Reluctance)
- General-purpose interface bus (AT-GPIB/TNT).
- Personal computer 486 DX-33 8 Mb RAM.
- Software package for Laboratory Virtual Instrument Engineering workbench (LabVIEW).
3.9. References


CHAPTER 4

EXPERIMENTAL RESULTS.

4.0. Introduction.
The noise current conducted out through the AC power cord of the equipment under test (EUT) is the conducted emission that has to be measured. According to standards requirements, the conducted emission tests are to be carried out by inserting line impedance stabilisation networks (LISN) in series between the commercial power system and the EUT, so that an AC 220 V supply flows through the LISN to the EUT. This procedure is described in detail in [1]. The measurement set-up is provided in chapter 3.
The experiment highlights three areas of interest:
1. Testing the improved methodology
2. Comparing the spectrum magnitude of voltage signals, and current signals
3. Comparing the spectrum magnitude of different systems
The level of conducted emission is measured by a computer simulated spectrum analyser that displays the magnitude spectrum of the signal (voltage output from LISN, and current through current probe).
4.1. Test procedure.

The following measurements are taken:

1. The spectrum magnitude of the voltage waveforms across the 50 Ω resistor in the LISN.

2. The spectrum magnitude of the current waveforms picked by a current probe.

The first measurement should be performed for both the low frequency band (10-150 kHz) and the high frequency band (0.15-30 MHz). The reason behind this separation is caused by the use of two types of LISN (high frequency, and low frequency) as explained in chapter 3.

The fact that current probes have lower impedance at lower frequencies leads to the use of an inverse amplifier [2]. This device amplifies the output of the current probe to produce a flat 1 Ω transfer impedance characteristic such that the input voltage to the EMI analyser, dB µV, is numerically equal to the current flowing in the test lead, in dB µ A.

The procedure is repeated for each power lead while the safety ground is exempted from the testing. The EUT consists of a PWM drive and motor. Three types of motors are tested:

1. Induction
2. Permanent magnet
3. Reluctance

Two PWM drives have been tested and the procedures and tests were repeated for each drive. A diagram of the testing procedure is provided in figure 4.1.
Figure 4.1. Diagram of test procedure
4.2. Conditions of measurement.

The emission current may be split into a differential mode that flows out through the phase conductor and returns through the neutral conductor, and a common mode component that flows out through the phase and neutral conductors and returns on the earth wire as shown in figure 4.2 [1]. Therefore the measured voltages are:

\[ V_p = 50(I_c + I_D) \]  \hspace{1cm} (4.1)  
\[ V_N = 50(I_c - I_D) \]  \hspace{1cm} (4.2)

The pulsating transistor current causes differential mode (DM) noise. Part of the pulsating current is shunted by the bulk capacitor and the rest flows through the 50 \( \Omega \) resistor (figure 3.6). The level of DM noise coupling is related to the impedance of the capacitor C. If the impedance is smaller than 100 \( \Omega \) (50+50), then the DM noise level is small. The current coupled through parasitic capacitors causes common mode noise. Voltages across these capacitors are high-voltage, high frequency waveforms. The displacement current generated by these voltage waveforms flows through the 50\( \Omega \) resistor, and results in measured EMI emission [4]. However, if the common mode and differential mode currents are of the same magnitude, the phase and neutral voltages will not be the same. Generally, one component dominates over the other, so that the magnitudes of the phase and neutral voltages are identical:

\[ V_p = 50I_c \hspace{1cm} I_c \gg I_D \]  \hspace{1cm} (4.3)
Experimental results

\[ V_N = 50I_c \quad I_c \gg I_D \]  \hspace{1cm} (4.4)

\[ V_P = 50I_D \quad I_D \gg I_c \]  \hspace{1cm} (4.5)

\[ V_N = -50I_D \quad I_D \gg I_c \]  \hspace{1cm} (4.6)

This experiment ignores the split of the total current into common mode and differential mode components. These modes of noise currents contribute to the total EMI noise current as shown in figure 4.3 [3].

This experiment has been carried out in a laboratory where no shielding has been provided. The building contain 2 lifts, a powerful central air-conditioning system, over 40 PCs, and a workshop containing cranes and mechanical tools. Therefore the ambient conditions may in a small way approach a small industrial environment. A circuit breaker was inserted between the main power leads and the laboratory bench. The maximum value of the current flowing through this circuit is 35 mA. A current exceeding this limit has been noticed occasionally during the transient periods at turn off and on, which resulted in long delays in conducting the experiment. Unfortunately the existence of an active telecommunication station (tower) right behind the laboratory introduced another factor that could degrade the value of the results. This experiment has been repeated in a shielded room where no radiated emission could contaminate the results.

Table 4.1 is provided to describe the conditions under which these tests are carried out.
Figure 4.2. Illustration of common-mode and differential-mode current in measurement.

Figure 4.3. (a) Common mode noise. (b) Differential mode noise (c) Total noise [3].
Table 4.1. Condition of measurement

<table>
<thead>
<tr>
<th>TYPE OF MEASUREMENT</th>
<th>Magnitude Spectrum of Signal</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Voltage</td>
</tr>
<tr>
<td>SPECTRUM BAND</td>
<td>Low Frequency</td>
</tr>
<tr>
<td></td>
<td>up to 300 KHZ</td>
</tr>
<tr>
<td>TIME DIVISION (CRO)</td>
<td>100 µs</td>
</tr>
<tr>
<td>POWER LEADS</td>
<td>Phase</td>
</tr>
<tr>
<td>AMPLITUDE DIVISION (CRO)</td>
<td>100 V</td>
</tr>
<tr>
<td>CURRENT PROBE DIVISION</td>
<td>1.00 A</td>
</tr>
<tr>
<td>3 Drives</td>
<td>PWM</td>
</tr>
<tr>
<td>RATED FREQUENCY</td>
<td></td>
</tr>
<tr>
<td>SWITCHING FREQUENCY</td>
<td>1-1.1 KHz, .340-2.190 KHz</td>
</tr>
<tr>
<td>1st motor</td>
<td>P. Magnet</td>
</tr>
<tr>
<td>2nd motor</td>
<td>Reluctance</td>
</tr>
<tr>
<td>3rd motor</td>
<td>Induction</td>
</tr>
</tbody>
</table>
4.3. Conducted emission tests.

The test was carried out repeatedly at different times (morning-evening, Monday-Friday), and under the same conditions as given in table 4.1. The results did not have large variations (differences did not exceed a few dB) in voltage and current waveforms. A PC was used to control the test except for changing the time and amplitude scales of the oscilloscope, and the current division for the inverse amplifier. Each area of interest has six results; three for the first drive (ZENER-MSC) and a further three results for the second drive (TOSHIBA VF-S3). Three results are recorded for the induction motor, permanent magnet motor, and a reluctance motor. Each group of results (six graphs) is provided together for comparison purposes. The motor supply voltage and current waveforms are also provided in the time domain in figures 4.4 - 4.7. The magnitude spectrum of the measured signals are provided in figures 4.8 - 4.15.

The magnitude spectrum of the measured signals, in a shielded room, is provided in figure 4.16. A photograph of the test bench is provided in figure 4.17.

For estimating the EMI of power electronic equipments, the amplitude density function is sufficient to find the worst case maximum amplitudes of the noise components. The EMI generated by power electronic equipment is that of a pulse train, since rectification involves repetitive switching from conduction to cutoff. The repetition rate is some multiple of the operating frequency, depending on circuit configuration. The electromagnetic emissions of semiconductor devices are broadband and coherent. Thus, for studying EMI from semiconductor circuits, the amplitude density function is a useful tool.
The amplitude density function of these emissions can be easily calculated with suitable approximations. The EMI evaluation generally can be performed with knowledge of the broken-like envelope of the amplitude density function (the spectrum). An approximate evaluation of the spectrum is made easier by further simplifying the above assumption. The EMI of power electronic equipment is qualified, in reference to the measuring bandwidth, as broadband. Therefore, instead of performing a discrete harmonic analysis of the pulse train, it is sufficient to define the spectrum of a single impulse in the pulse train. This is a significant simplification since, in most cases, the calculations become much easier. The predicted spectrum magnitude of the conducted emission produced by PWM inverter has been calculated according to this approximation [6]. Figure 4.18 illustrate the spectrum magnitude of the conducted emission produced by PWM inverter. A comparison between the calculated results and the measured one reveal the fact that, to certain extent the calculated results give a very close indication of the real level of emission.
Experimental results

Figure 4.6. Voltage waveform (Phase)

Figure 4.7. Voltage waveform. (Neutral)
**Experimental results**

**Figure 4.4. Current waveform (Phase).**

**Figure 4.5. Current waveform (Neutral)**
Figure 4.8. Spectrum magnitude of Voltage waveforms (Neutral, HF). a) ZENER drive. b) TOSHIBA drive.
Figure 4.9. Spectrum magnitude of Voltage waveforms (Neutral, LF). a) ZENER drive. b) TOSHIBA drive.
Figure 4.10. Spectrum magnitude of Voltage waveforms (Phase, HF). a) ZENER drive. b) TOSHIBA drive.
Figure 4.11. Spectrum magnitude of Voltage waveforms (Phase, LF). a) ZENER drive. b) TOSHIBA drive.
Figure 4.12. Spectrum magnitude of Current waveforms (Phase, HF). a) ZENER drive. b) TOSHIBA drive.
Figure 4.13. Spectrum magnitude of Current waveforms (Phase, LF). a) ZENER drive. b) TOSHIBA drive.
Figure 4.14. Spectrum magnitude of Current waveforms (Neutral, HF). a) ZENER drive. b) TOSHIBA drive.
Figure 4.15. Spectrum magnitude of Current waveforms (Neutral, LF). a) ZENER drive. b) TOSHIBA drive.
Figure 4.16. Spectrum magnitude of current waveforms measured in a shielded room.
Figure 4.17. A photograph of the test bench used in this study.
Figure 4.18. the estimated spectrum of the EMI emission.
4.4. References


6. L. Tihanyi, "Electromagnetic compatibility in power electronics". The Institute of Electrical and Electronics Engineers, Inc. 1994, USA.
CHAPTER 5

CONCLUSIONS

5.0. Discussion and conclusions.
Every switching operation of the devices in the motor drive imposes an impulse voltage with the derivative $dv/dt$ with a period of a hundred of nanoseconds, not only on the line-to-line voltages but also on the line-to-ground voltage of the output signal. Furthermore, physically each portion of the motor winding, the output feeder cables to the motor or portion of semiconductor in the device itself, acts as a plate of a capacitor with respect to the grounded metal frame. During the switching transient an impulse charging/discharging current flows through these distributed capacitances to ground and returns to the power mains along many paths. These currents are referred to as the common mode current. Moreover, since all the conductors have stray inductance and since losses exist in the current paths, these impulse charging/discharging currents serves to explain the high frequency ringing and the complex wide bandwidth spectrum of conducted EMI. The switching devices dominate the value of parasitic capacitance and the common mode emission. The capacitance formed by the p-n junction in the semiconductor devices or by two portions of conductors in different phases also produce another
type of RF current, flowing around the power leads during each switching transient. This type of RF current is referred to as differential mode emission. They are less than the common mode emission. The diodes used both in the inverter switching modules and the rectifier bridge are also one of the differential mode emission sources due to their reverse recovery current switching transient.

The EMI emissions of a PWM inverter are periodical with respect to the switching frequency. The spectrum of the emission, therefore, centers at the switching frequencies and multiples. Any switching converter will generate two kinds of electromagnetic interference:

1. EMI produced by the natural switching action, which occurs at multiples of the switching frequency.

2. EMI produced by excitation of parasitic elements when the semiconductor devices are turned on and off. This is a consequence of the non-ideal realisation of the switch [1].

This concentration of emission power at the switching frequency and its multiples makes it harder, and sometimes impossible, to satisfy EMI regulations. The reduction of EMI emission is possible by reducing amplitude and distributing the harmonic content. It is a good practice to deal with this problem at the design stage. Despite the dramatic improvement in the performance of power electronic devices their applications in switched mode power supplies leads to an increase in the switching frequencies and ultimately a reduction in power loss. The resulting side effects are an unintentional increase of harmonic amplitude. Gate controlled devices (MOSFET and IGBT) require suitable driving techniques in order to achieve a balance between power losses and
conducted emissions during the transient stage of the switching. A properly designed driving circuit, which controls the slope of the current during the turn-on and turn-off, can reach this goal [2].

The modulating of the switching frequency leads to a creation of side bands resulting in a broader emission spectrum, and the distribution of the power into smaller magnitudes scattered along a wide band [3]. Figure 5.1 shows the effect of frequency modulation on the frequency spectra, where:

- $f_c$ is the carrier frequency
- $f_m$ is the modulating frequency
- $\Delta f$ is the amplitude of frequency change

The magnitude of the modulated signal of figures 5.1 (b) is less than the unmodulated signal (figure 5.1 (a)) and side band harmonics are generated as well. The same applies for the spectra of unmodulated square waveforms (figure 5.1 (c)) and frequency modulated square waveform (figure 5.1 (d)).

The frequency modulation index $\beta = \Delta f / f_m$ dictates the extent of distribution of the resultant spectrum. The larger the index, the wider the spectrum and the less the magnitude of the harmonic components. According to Garson’s Rule [3], the total power of a signal is unaffected by the frequency modulation, and 98% of the total power of a frequency-modulated signal will be contained inside the bandwidth ($B_r$), where

- $B_r = 2(\beta + 1)f_m$ [4].

Another way to reduce the effect of conducted EMI was proposed by Simonetti [5], who proposed that when a converter operates with a
constant switching frequency, the harmonic content is concentrated around the switching frequency.

The measured magnitude spectrum of all waveforms were within the boundary of requirement. (CISPR 22 average class A). These results were expected since the switching frequency is low (1.0-1.1 kHz for ZENER MSC, 0.40-2.19 kHz for TOSHIBA VF-S3), and these drives use the modulated switching frequency method, and in addition, the switching frequency is variable. All tests were performed under no load conditions. The tests proved that an LISN inserted between the main power network and the test bench prevented any outside interference from contaminating results.

From an EMI standpoint, there is no significant difference between the tested drives since the level of spectrum magnitude of all results is almost the same. The effect of switching frequency dominates in all cases, however, a slight difference between the spectrum magnitude of different motor drives has been recorded.

The Induction motor drive achieved the lowest level while the reluctance motor scored the highest (a difference of only a few dB). The operating current in the case of the induction motor was the lowest (see table 5.1.)

Table 5.1. Input current of different drive systems.

<table>
<thead>
<tr>
<th>Mode</th>
<th>ZENER MSC</th>
<th>TOSHIBA VF-S3</th>
</tr>
</thead>
<tbody>
<tr>
<td>INDUCTION</td>
<td>I_{oper} = 0.16 A</td>
<td>I_{oper} = 0.22 A</td>
</tr>
<tr>
<td>P. MAGNET</td>
<td>I_{oper} = 0.37 A</td>
<td>I_{oper} = 0.36 A</td>
</tr>
<tr>
<td>RELUCTANCE</td>
<td>I_{oper} = 0.49 A</td>
<td>I_{oper} = 0.91 A</td>
</tr>
</tbody>
</table>

MASTER OF ENGINEERING (HONOURS)
Figure 5.1. Magnitude spectra
The magnitude spectrum of both the voltage and current waveforms at different points has indicated almost the same level of conducted EMI. However, the need to take into account the transfer impedance of the current probe and current division leads to a complication in the programming process. In addition, the current waveform is measured by a current probe, which is susceptible to outside interference, which can introduce an additional noise to the measured signal. While the voltage waveform is measured directly across a 50 Ω resistor. Both methods are quite effective. However, measuring the voltage signal is considered to be more practical and probably more accurate.

The results of the experiment conducted in a shielded room, for reluctance motor drive, showed that the magnitude spectrum of the voltage waveforms do not change. However, the magnitude spectrum of the current waveforms indicated a decrease between 7 to 10 dB.

The EMC laboratory of Siemens Matsushita Components tested the conducted emission produced by a 5.5 KVA frequency converter with a switching frequency of 20 kHz [6]. Their test was conducted under normal conditions of operation, e.g. the motor lead is about 20 metres long with the motor running at part load (not determined). The test was conducted at an operating current of 13 A. Figure 5.2 provides a comparison of their results with the results of the method proposed in the current project.

The improved method of measurement appears to be worthy of implementation, since all measurement results have met expectations. All the equipments used in this experiment have been properly calibrated. All known parameters, which may influence the results have been able to be taken into account such as: current probe division, voltage probe division,
and transfer impedance. In addition, this improved method appears to have many advantages: it is flexible; it can be easily modified; and it can offer the option of building a multi purpose virtual instrument. It enables the transfer of data between the local computer and other devices at burst rates approaching 29 Mbytes/s [7]. It is probably more practical to use a spectrum analyser and avoid this complicated process, even for a simple single measurement. However the need to store results as data and to use them for comparison and further analysis, and sometimes to transmit them to another system through communication networks makes the proposed method flexible and useful. It has the added advantage of being a low cost measurement.

5.1. Suggestions for further work.
Electromagnetic disturbance problems associated with the use of static power converters, especially for power drive systems, have not been well addressed nor have mandatory test requirement appeared. However, this does not imply that EMC problems can be neglected in power drive applications because they are still subject to the general condition that no harmful interference be caused. Hence, it is of concern to all, that EMI problems be minimised even when standards do not presently exist. The effect of non linear load on conducted EMI could be an interesting area to investigate, since the present investigation is restricted to tests when motors are running under no load conditions. This load can be of fixed or variable value, and in the case of variable load, the behaviour of conducted EMI is still unknown.
Another interesting area is the prediction of conducted EMI at design level by establishing a direct relationship between conducted EMI and the parameters of the desired variable speed motor drive.
Figure 5.2. Interference voltage diagrams (a) Frequency converter
(b) PWM drive.
5.2. References


2. Feng Lin, and Dan Y. Chen, "Reduction of power supply EMI emission by switching frequency modulation". IEEE transactions on power electronics, Vol. 9, No 1, January 1994.


Appendix A

Induction motor drive

Synchronous speed is the speed of rotation of the field and can be written:

\[ \omega_s = \frac{2\omega}{P} \]  \hspace{1cm} (1)

where \( P \) is the number of poles and \( \omega \) is the supply frequency.

Given the stator voltage, \( v_s = \sqrt{2} \sin \omega t \), then the induced voltage in the rotor winding is:

\[ e_r = -s\sqrt{2}E_r \sin(s \omega t - \delta) \]  \hspace{1cm} (2)

where \( E_r \) is the rms value of the rotor induced voltage,

and \( \delta \) is the relative position of the rotor,

\[ s = \frac{\omega_s - \omega_m}{\omega_m} \quad \text{s is the slip} \]

where \( \omega_m \) is the angular speed of the rotor.

If the rotor current is \( I_r \), and the stator current is \( I_s \), then the performance parameters of a motor are as follow:

- Stator copper loss \( P_s = 3I_s^2R_s \) \hspace{1cm} (3)
- Rotor copper loss \( P_r = 3I_r^2R_r \) \hspace{1cm} (4)
- Core loss \( P_c = \frac{3V^2}{R_m} \approx \frac{3V^2}{R_m} \) \hspace{1cm} (5)
- Air gap power \( P_g = 3I_s \frac{R_r}{\omega_s} \) \hspace{1cm} (6)
Developed torque

\[ T_d = \frac{3R_sV_i^2}{s\omega_s[(R_s + \frac{R_r}{s})^2 + (X_s + X_r)^2]} \] (7)

where \( R_m \) is the resistance to represent excitation loss,

and \( X_m \) is the magnetising reactance

\( R_r, X_r \) are refereed to the stator winding

The motor speed can be computed as a function of torque

\[ \omega_m = \omega_s (1 - s) \] (8)

In any magnetic circuit, the induced voltage is proportional to flux and frequency, and the rms air-gap flux can be expressed as

\[ V_a = bV_s = K_m\omega_s\phi \]

\[ \phi = \frac{V_a}{K_m\omega_s} = \frac{bV_s}{K_m\omega_s} \] (9)

Where \( b < 1 \)

If the ratio of voltage to frequency is kept constant, the flux \( \phi \) remains constant and hence the maximum torque remains constant. In order to maintain the torque level, at low frequency, the voltage has to be increased since the air gap flux is reduced due to the drop in the stator impedance.

This type of control is known as volts/hertz control. The synchronous speed corresponding to the rated frequency is called the base speed, \( \omega_b \).

If \( \omega_s = \beta\omega_s \), and the voltage to frequency ratio is constant so that
\[
\frac{V_a}{\omega_z} = d
\] (10)

then the maximum torque can be written as

\[
T_d = \frac{3R_d d^2 \omega_s^2 (\beta \omega_b - \omega_m)}{\omega_b^2 (R_s + R_d)^2 + [(\beta \omega_b - \omega_m) (X_s + X_d)]^2}
\] (11)

The torque-speed characteristics are shown in Figure 1. If the frequency is reduced, \(\beta\) decreases and the slip for maximum torque increases. By varying the frequency and voltage, the torque and speed can be controlled. Normally the torque is kept constant while the speed is varied. One of the possible circuit arrangements for obtaining variable voltage and frequency is shown in Figure 2.

A three phase inverter can supply variable voltage at a variable frequency PWM inverter can vary the voltage and frequency within the inverter while the dc voltage remains constant. The generation of harmonics into the AC supply by PWM inverters, due to the rectifier diode, is one disadvantage of the spectrum.
Figure 1. Torque-speed characteristics with volts/hertz control.

Figure 2. Induction Motor drive
Appendix B

Sinusoidal PWM

The modulation index, \( m \), is the ratio of the magnitude of the reference wave to that of the carrier wave

\[
m = \frac{A}{A_m}
\]  \hspace{1cm} (1)

The rms value of the resultant waveform is:

\[
V_i = \frac{mV_d}{2\sqrt{2}}
\]  \hspace{1cm} (2)

It can be seen that the relation between the fundamental voltage and \( m \) is linear. If \( m > 1 \), the modulation ceases to be sinusoidal since the number of pulses is reduced. The resultant waveform contains harmonics and these harmonics are odd multiples of the carrier frequency. In addition this waveform contains sidebands centred around multiples of \( f_c \). By moving away from the band center, the magnitudes of the band frequency harmonics decrease rapidly. The increase in the modulation index, \( m \), increases the width of a band.

The ratio of carrier frequency to that of reference frequency is \( p \):

\[
p = \frac{f_c}{f}
\]  \hspace{1cm} (3)
If \( p \) is large, the harmonics are very large in comparison to the fundamental frequency. The effect of harmonics on the operation of the motor can be minimised by increasing \( p \) to the maximum, i.e. by operating the switches at the highest possible frequency. An increase in the switching frequency would decrease the machine losses but the inverter switching losses would increase, and above certain frequencies, the drive efficiency would fall.
Appendix C

GPIB

The interface system consists of 16 signal lines and 8 shield drain lines. Fig. 1 shows GPIB connector. The 16 lines are divided into three groups:

- Eight data lines
- Three handshake lines
- Five interface management lines

Data and command messages are carried by the eight data lines, DIO1 through DIO8. Most commands use the 7-bit ASCII or International standards organisation code set (ISO).

The transfer of message among devices is controlled asynchronously by three lines. This will guarantee that message bytes on the data line are sent and received without error.

Messages could be:

- NRFD (not ready for data)
- NDAC (not data accepted)
- DAV (data valid).

Five lines manage the flow of information across the interface:

- ATN (attention)
- IFC (interface clear)
- REN (remote enable)
- SQR (service request)
- EOI (end of identity).

Fig. 3-11 GPIB connector.
Appendix D

Analysis of LISN 1

AC Analysis LISN 1

Phase (deg)

Transient Analysis LISN 1

Voltage (V)
Total harmonic distortion (%)  0.46769

Parameter                  Value
Initial TStep             0.000625
Initial TMax             0.00125
Nominal temperature       27
Operating temperature     27
Total iterations          65
Transient iterations      62
Circuit Equations         18
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Appendix E

Analysis of LISN 2

AC Analysis LISN 2

Transient Analysis LISN 2
Fourier Analysis LISN 2

Total harmonic distortion (%) 0.45968

Noise Analysis LISN 2

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<td>-------------</td>
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<td>Circuit Equations</td>
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EMI STUDIES IN MOTOR DRIVES

BY

ALI HELLANY
B.Eng. (Telecommunications)

Thesis submitted for the degree of
Master of Engineering (Hon)
In the Faculty of Engineering
University of Western Sydney-Nepean
November 1996
PLEASE NOTE

The greatest amount of care has been taken while scanning this thesis,

and the best possible result has been obtained.
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**ACKNOWLEDGMENT**

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**STATEMENT OF ORIGINALITY**

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This thesis reviews in general the topic of electromagnetic compatibility, and electromagnetic interference and their origin and effects in modern electronically controlled motor drives. The measurement techniques for EMI noise are reviewed. The sources of noise of a switching power circuit are described.

This thesis investigates the establishment of a procedure for measuring conducted emission produced by motor drives, using a virtual instrument. This procedure is based on the traditional methodology of EMI measurement and the use of simulation techniques. A test bench is designed.

This thesis covers the detailed design of a virtual instrument for measuring conducted current produced by motor drives. A line impedance stabilisation network LISN is designed and built.

A series of measurements were carried out using the developed instrument. The results show very little difference between the conducted emission produced by induction, permanent magnet and reluctance motor drives. Comparing one of the experimental results with published results from a major test laboratory assesses the validity of the designed instrument. The experimental results refer to drive systems under no load conditions. Useful conclusions are drawn and future research studies are also recommended.
PREFACE

Chapter 1 provides an introduction to the work reported and an extensive literature survey to give an insight into the area of electromagnetic compatibility/interference (EMC/EMI). The survey is limited to power electronics and drive systems and their associated EMI problems.

Chapter 2 discusses AC drive systems and pulse width modulation (PWM) inverters and various techniques for frequency and voltage controls. In this chapter, a few methods of reducing EMI in AC drives are also discussed.

Chapter 3 investigates the design of a new measurement methodology for testing conducted emission produced by motor drives. The design of a line impedance stabilisation network (LISN) is described. The design of a virtual instrument working as EMI spectrum analyser is also described.

Chapter 4 shows the experimental results of conducted emission produced by three types of motor drives (induction, permanent magnet and reluctance).

Chapter 5 provides a discussion of the results, and general conclusions, which are derived from the investigation carried out in this research, and recommends directions for future work on the conducted emission of motor drives under various load conditions.
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Finally, the author wishes to express his gratitude to all his friends and colleagues for their continuing support.
STATEMENT OF ORIGINALITY

I certify that the substance of this thesis has not already been submitted for any degree and is not currently being submitted for any other degree.

I certify that any assistance received in preparing this thesis, and all sources used, have been acknowledged in this thesis.

Ali Hellany