

ALL SIC TRACTION CONVERTER FOR LIGHT RAIL TRANSPORTATION
SYSTEMS: DESIGN METHODOLOGY AND DEVELOPMENT OF A 500 - 900
VDC, 165 KVA PROTOTYPE

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ABSTRACT

ALL SiC TRACTION CONVERTER FOR LIGHT RAIL TRANSPORTATION SYSTEMS: DESIGN METHODOLOGY AND DEVELOPMENT OF A 500 - 900 VDC, 165 KVA PROTOTYPE

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Design methodology and development of a 165 kVA, all-silicon carbide (SiC) traction converter for new generation light rail transportation systems (LRTS) have been carried out in this thesis. Electrical, mechanical, and thermal design principles of all-SiC power MOSFET-based traction converter have been described in detail. In order to verify the produced converter, design and implementation of a full-scale physical simulator of an all-SiC traction motor drive for LRTS have been performed. A complete mathematical model of the physical system has been derived to carry out real-time simulations of LRTS, which takes into consideration the actual two-quadrant speed versus torque characteristics of a typical traction motor drive. Power losses and thermal performance of SiC power MOSFET modules have been assessed in comparison with those of the alternative Hybrid-IGBT and Si-IGBT modules for various switching frequencies, and their advantages have been considered in view of converter design objectives. This analysis is also repeated for three-level converter topologies in order to visualize the effect of SiC power MOSFETs on converter topology selection. Implementation techniques of the resulting converter power

stage, involving the DC and AC laminated busbars, the general layout, and associated cooling system, have been paid special attention. The application method of the space vector PWM technique for operation under variable DC catenary voltages over a wide range has been uniquely described. The associated experimental results were taken on the developed all-SiC power MOSFET-based traction inverter and full-scale physical simulator for LRTS. Very promising results have been obtained on the track of converter design objectives in terms of performance criteria such as very high efficiency, low motor line current total demand distortion, low cooling requirement, relatively high switching frequency, and hence ease of controller implementation.

Keywords: Railway Traction Inverters, SiC Power MOSFET, Induction Machine Drive, Field Oriented Control, Space Vector Pulse Width Modulation

ÖZ

HAFİF RAYLI TAŞIMA SİSTEMLERİ İÇİN SiC ÇEKİŞ KONVERTÖRÜ: TASARIM METODOLOJİSİ VE 500 - 900 VDC, 165 KVA PROTOTİPİNİN GELİŞTİRİLMESİ

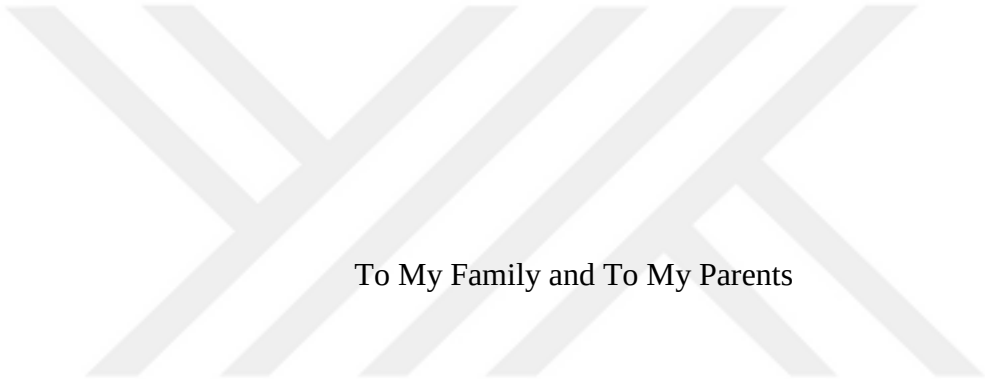
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Bu tezde, yeni nesil hafif raylı ulaşım sistemleri için 165 kVA, tamamı SiC güç MOSFET'lerinden oluşan bir cer dönüştürücünün tasarım metodolojisi ve geliştirilmesi gerçekleştirilmiştir. SiC güç MOSFET tabanlı cer konvertörlerinin elektriksel, mekanik ve termal tasarım prensipleri detaylı olarak anlatılmıştır. Hafif raylı taşıma sistemleri için geliştirilen tamamı silikon karbür (SiC) tabanlı cer motoru sürücüsünün testi için tam ölçekli, fiziksel simülatörünün tasarımı ve uygulaması yapılmıştır. Hafif raylı taşıma sistemlerinin gerçek zamanlı simülasyonlarını gerçekleştirmek için, fiziksel sistemin eksiksiz bir matematiksel modeli türetilmiştir. SiC güç MOSFET modüllerinin güç kayıpları ve termal performansı, çeşitli anahtarlama frekansları için alternatif hibrit ve Si-IGBT modüllerininkilerle karşılaştırılarak değerlendirilmiş ve dönüştürücü tasarım amaçları açısından avantajları göz önünde bulundurulmuştur. Bu analiz ayrıca SiC güç MOSFET'lerinin dönüştürücü topoloji seçimi üzerindeki etkisini görünür kılmak için üç seviyeli dönüştürücü topolojileri için de tekrarlanmıştır. Konvertör tasarımında, DC ve AC lamine baralara, genel yerleşime ve soğutma sistemini içeren güç modülü üretimi için uygulama tekniklerine özel dikkat gösterilmiştir. Geniş bir aralıkta değişken DC

katener voltajları altında çalışma için uzay vektörü PWM tekniğinin uygulama yöntemi, geliştirilmiş tam ölçekli, SiC güç MOSFET tabanlı çekiş dönüştürücüsünde ve inşa edilen test düzeneğinde alınan ilgili deneysel sonuçlarla benzersiz bir şekilde tanımlanmıştır. Çok yüksek verim, düşük motor akımı bozulması, düşük soğutma gereksinimi, nispeten yüksek anahtarlama frekansı ve dolayısıyla kontrol yöntemi uygulama kolaylığı gibi performans kriterleri açısından yeni teknoloji evirici tasarım hedeflerinin izinde çok umut verici sonuçlar elde edilmiştir.

Anahtar Kelimeler: Demiryolu Araçları Çekiş Evirgeçleri, SiC Güç MOSFET, Asenkron Makina, Sürücüsü, Alan Yönlendirmeli Kontrol, Uzay Vektörü Darbe Genişlik Modülasyonu



To My Family and To My Parents

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LIST OF ABBREVIATIONS

LRTS	Light Rail Transportation System
SiC	Silicon Carbide
ESS	Energy Storage System
MOSFET	Metal Oxide Semiconductor Field Effect Transistor
IGBT	Insulated Gate Bipolar Transistor
LVDC	Low Voltage DC
GTO	Gate Turn-Off Thyristor
GaN	Gallium Nitride
LRV	Light Rail Vehicle
NPC	Neutral Point Clamp
MVDC	Medium Voltage DC
HVDC	High Voltage DC
SST	Solid State Transformer
IGCT	Integrated Gate Commutated Thyristor
SHE	Selective Harmonic Elimination
SHM	Selective Harmonic Mitigation
PWM	Pulse Width Modulation
OPWM	Optimum Pulse Width Modulation
SVPWM	Space Vector Pulse Width Modulation
HSCB	High Speed Circuit Breaker
PMSM	Permanent Magnet Synchronous Motors
VVVF	Variable Voltage Variable Frequency
SMPMSM	Surface Mounted Permanent Magnet Synchronous Motors
IPMSM	Interior Permanent Magnet Synchronous Motors
THD	Total Harmonic Distortion
TDD	Total Demand Distortion
MTBF	Mean Time Between Failure

GPU	Graphical Processing Unit
SPWM	Sinusoidal Pulse Width Modulation
THIPWM	Third Harmonic Injection Pulse Width Modulation
DPWM	Discontinuous Pulse Width Modulation
FOC	Field Oriented Control
DTC	Direct Torque Control
EV	Electric Vehicle
SMPS	Switched Mode Power Supply
CBCT	Communications-Based Train Control
ERTMS	European Railway Traffic Management System

CHAPTER 1

INTRODUCTION

1.1 History of Railway Vehicles

Transportation has provided important contributions to the development of civilizations from ancient times to the present by meeting the travel needs of the people and the transport needs of the goods. It is an integral part of any society and is recognized as being of key importance for both employment and economic growth. Thus, sustainable, efficient, and smart transportation solutions have great importance for humanity. However, transportation solutions also have some environmental impacts such as air pollution, noise pollution, and energy consumption. According to 2018 figures of the European Environmental Agency Report, a quarter of European's greenhouse gas emissions derived from transportation with civil aviation, maritime transport, road transportation, rail, and other transportation, respectively 13.9%, 13.4%, 71.7%, 0.5%, and 0.5% of the total. Considering the energy consumption, CO₂, and exhaust atmospheric emission, rail transportation is stated as the most friendly sector than other types of transportation in Europe.

Railway development began to gain importance with the discovery of steam engines at the beginning of the 19th century. The first steam locomotive was made by Richard Trevithick in 1803. However, a commercially successful steam locomotive was first introduced in 1812 by John Blenkinsop [1]. After about 25 years, a battery-powered full-electric locomotive was built by Robert Davidson in 1838. The full-electric locomotive did not last long due to insufficient battery technology. The main reason was the lack of generating sufficient electrical power from batteries. Between 1870 and 1885, basic models of full-electric locomotives were developed by Siemens (1879), Edison (1880), and Daft (1885). In the mid of 19th century, light-duty street-car tramways had started to use in rapid-transit service. The street cars used low

voltage DC (LVDC) systems that were supplied by batteries located on the tramways or supplying DC current from a station through overhead wire or a ground-level rail. The LVDC electric traction systems of tramway vehicles had a great influence on the use of electric traction systems on heavy-duty railway vehicles in the following years. The LVDC electric traction spread to displace steam traction. After a while, diesel-electric locomotives were introduced as soon as the diesel engine was presented by Rudolf Diesel [2].

In the last 50 years, there has been a rapid acceleration in railway traction drive system development in parallel with innovations in the power electronics and microprocessor fields. Solid-state rectifiers had begun to be used in locomotives between 1970 and 1973. After that, the tendency to use solid-state products in motor control systems had increased, and the use of thyristors in DC traction motor control began in the 1980s. The better tractive effort, better wheel-slip control, higher efficiency, and weight reduction had been achieved with the use of thyristors. In addition to this, a prototype of a diesel-electric locomotive that used AC traction motors was presented in 1971 [2]. Manufacturing of AC traction motors is easier, and AC traction motors require less maintenance during their operating life. They also give a higher power to weight ratio and power to volume ratio than the same power level DC traction motors [3]. Insulated gate bipolar transistors (IGBT) technology was developed in the 1980s. However, it became commercially widespread after the 1990s. By using IGBT technology, power quality, power to weight ratio, and power to volume ratio of traction systems were further increased. As a result of these developments, old technology traction systems have been replaced by modern and new technology electric traction systems. The main advantages of modern electric traction systems are higher energy efficiency, higher reliability, less noise, lower maintenance, lower operating cost, pollution-free operation, higher power to weight ratio, and higher power to volume ratio than old equivalent traction systems. Less weight and volume also provide more free space for passengers. Thus, the need for innovative and efficient modern traction systems will increase day by day.

The improvement of overall energy efficiency and reliability needs are two of the most important factors that triggered the development of new semiconductor technologies. Nowadays, gallium nitride (GaN) and silicon carbide (SiC) based wide band-gap semiconductor technologies have begun to gain importance in low and medium power applications [4]–[8]. Higher blocking voltage, higher operating temperature, and higher switching speed features make SiC-based power MOSFETs attractive for low and medium power applications by decreasing the volume and weight of converters and increasing conversion efficiency. In addition to this, wide-bandgap power products offer a chance to use simpler modulation techniques and control algorithms. Because of these superior advantages, railway traction-drive manufacturers such as Mitsubishi, Toshiba, and Hitachi, have a strong tendency to use SiC semiconductors.

1.1.1 Types of Railway Traction Systems

So far, four main types of traction systems have been built up. These are respectively steam-powered, electric-powered, diesel-powered, and hybrid traction systems. Among them, the steam-powered traction systems firstly appeared. These traction systems were working by burning combustible materials such as coal, wood, and oil. Those systems had a big disadvantage in terms of air pollution comparing the other traction systems. From the early 1900s, steam-powered traction systems were gradually superseded by electric and diesel-powered traction systems. Although full electric-powered traction systems had been built up earlier and have more advantages than diesel-powered traction systems, they have been widespread after diesel-powered traction systems because of the lack of electrification infrastructures. Typical diesel-powered traction systems are, namely, diesel-mechanical, diesel-electric, diesel-hydraulic, diesel-steam, and diesel-pneumatic. Diesel-powered traction systems have many advantages over steam-powered ones, such as easy start/stop, safely operable by one person, dirt free, low maintenance requirements, elimination of dependency on an electrification system, and very less capital investment needs and operational costs. Besides this, compared to electric-powered

traction systems, diesel-powered traction systems also have a bad environmental impact on air pollution, are less powerful and less efficient, creating more audible noise. Full electric-powered traction systems could only be widespread after the 1970s [9].

The main objectives of technological improvements in railway traction systems are to increase energy efficiency and reliability and to offer solutions to reduce environmental damage. In parallel with the development of modern traction converters, since the 2000s, supercapacitor energy storage systems (ESS) have been used in railway applications to reduce the energy consumption of the railway vehicles, to regulate the catenary voltage, and to provide temporarily catenary-free operation [10], [11]. These systems can be installed on-board or as stationary at power substations [12]–[15]. The use of energy storage equipment in electric-powered traction systems is growing up due to the great advantage of decreasing waste of energy. With the developments of rechargeable energy storage equipment such as high voltage lithium batteries, supercapacitors, and lithium-ion capacitors, energy storage systems have been used on power stations, metro, and tram vehicles. In addition, hybrid diesel-electric powered traction systems equipped with energy storage systems are becoming widespread on shunting locomotives and diesel-electric EMUs.

1.1.2 Types of Railway Electrification Systems

Railway electrification systems provide electrical energy to traction and auxiliary power supply systems located on the railway vehicles. Therefore, they have a very important role in making electric traction systems usable. Early electrification systems provide low voltage 600VDC for tramways and light rail vehicles. These rail vehicles had primitive traction systems consisted of resistors and contactors to perform speed and torque control of DC electric motors.

Today, several electrification systems exist in the world for different rail vehicles and can be classified according to voltage and contact system. The most common

voltages are 750 VDC, 1500 VDC, 3000 VDC, 15 kVAC ($16^{2/3}$ Hz), 25 kVAC (50Hz) respectively [16].

The permissible range of voltages and allowed limits with their durations for the standardized voltages as stated in standard BS EN50163 [17] are illustrated in Table 1.1 and Figure 1.1.

Table 1.1 Nominal Voltages and Permissible Limits [17]

Electrification System	Lowest non-permanent voltage U_{min2} V	Lowest permanent voltage U_{min1} V	Nominal voltage U_{n1} V	Highest permanent voltage U_{max1} V	Highest non-permanent voltage U_{max2} V	Highest long term overvoltage U_{max3} V
D.C.	500	500	750	900	1000	1270
	1000	1000	1500	1800	1950	2540
	2000	2000	3000	3600	3950	5075
A.C.	11000	12000	15000	17250	18000	24300
	17500	19000	25000	27500	29000	38750

Lower voltages 750 VDC and 1500 VDC are often used for tramways, LRVs, and metros with third rail, fourth rail, or overhead line systems, whereas the other higher voltage levels are often used for regional trains, very high/high-speed trains, and locomotives due to long travel distance and high power requirements. Increased voltage levels necessitate the use of overhead lines because for safety reasons. Higher voltage levels have the advantage of low resistive loss and have a chance to use long-distance between two feeder stations.

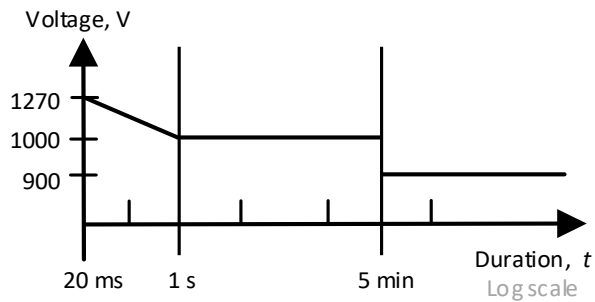


Figure 1.1. Permissible voltage variations in 750 V DC catenary line are specified in EN 50163 [17]

As tabulated in Table 1, the voltage levels are not fixed, and they may vary according to power demand, the direction of the power flow, the distance between the railway vehicle to the power station, and electrification system characteristics. The allowable voltage ranges and time limits are illustrated in Figure 1.1. The permissible operating ranges of vehicle current against catenary line voltage are as specified in EN 50388 [16] and reconstructed as shown in Figure 1.2. In the case of a non-standard customer's requirement, this limitation is taken into account.

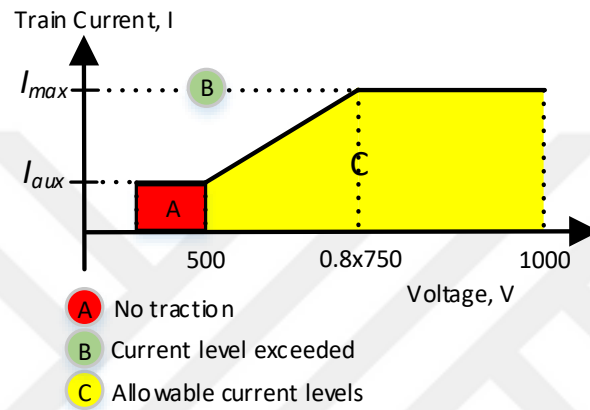


Figure 1.2. Permissible operating ranges of vehicle current against catenary line voltage are as specified in EN 50388 [16]

1.2 Electric Powered Traction Drives for Railway

Electric-powered traction drive includes electric motors, power converters, and a control platform. The main purpose of the traction drive is to make the needed movement of vehicles. The role of power converters in this chain is synthesizing the required voltage and current waveforms to drive the electric motors. Electric-powered traction drives have been used for more than one hundred years for railway applications. For many years, most of the drives used series or separately-excited DC motors with 600-750 VDC input voltage. Mainly, the motor current is proportional to the reciprocal of speed. Thus, high motor currents can create high tractive effort at low speed and high power at high speeds. Although this ability fulfills the traction effort requirements, the motor current at low speed has to be limited to protect the

electric motor from very high currents due to very low back-emf. Until the devolvement of power converters, the speed and torque control of DC motors are performed with resistance control. A rotary drum switch and a bank of resistors were used to control the current. A sample current vs. speed characteristic is given in Figure 1.3. This method had caused to waste of considerable amount of energy in the past [3].

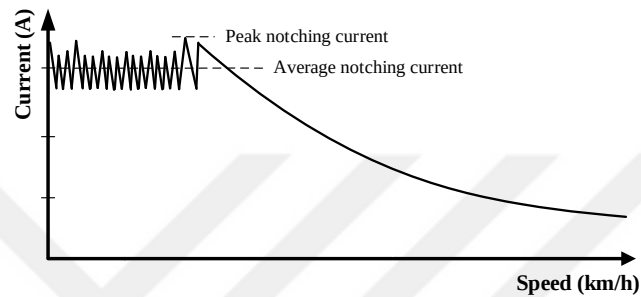


Figure 1.3. Current – Speed graph with resistance notching

In addition to this, when low-frequency AC electrification lines are available, universal/DC motors have been utilized with tapped transformer and diode rectifier. A basic circuit diagram is shown in Figure 1.4. The motor voltage and current are adjusted with the help of a tap changer of the transformer [3].

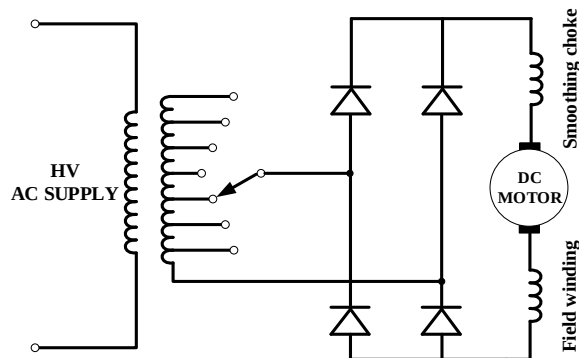


Figure 1.4. Circuit diagram of tap changer controlled DC motor

Although this solution provides higher efficiency than the resistance control, the energy conversion efficiency is still very low due to the low power quality. After the

generator supply solution had spread on the vehicle side, an AC alternator and diode rectifier arrangement was used [3]. The basic power circuit is seen in Figure 1.5.

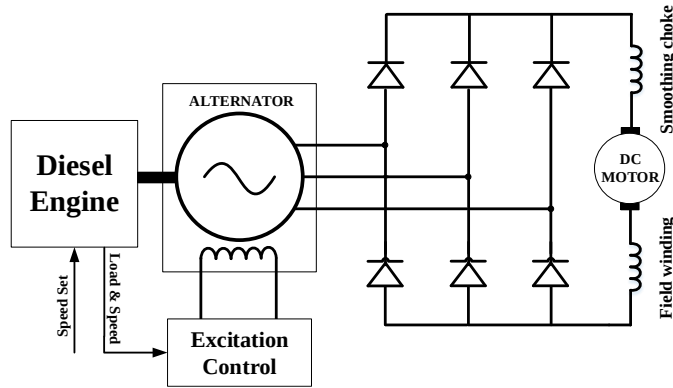


Figure 1.5. DC motor control with generator & diode bridge circuit

After a while, the thyristor bridge circuit, as shown in Figure 1.6, was used to increase system efficiency by controlling the diesel motor and alternator at or near to their high-efficiency points and by improving power quality [3].

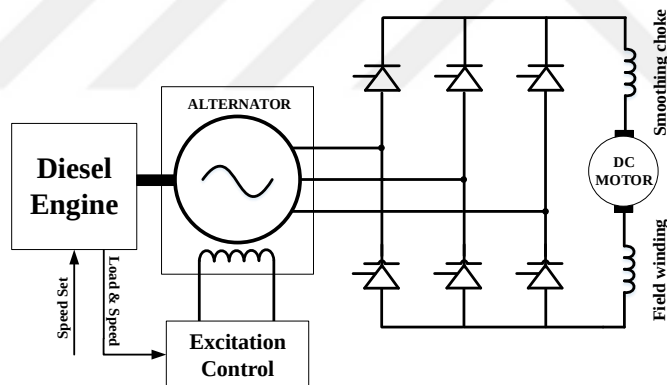


Figure 1.6. DC motor control with generator & thyristor bridge circuit

In the 1970s, thyristors-based DC-DC chopper converters were started to use for railway DC traction motor control applications. DC-DC chopper converters provide variable voltage to the DC motors from a fixed voltage source. A considerable amount of energy saving at the acceleration instant of railway vehicles, smooth torque control, fast response, and regenerative braking features made DC-DC chopper converters advantageous than older traction systems. A basic circuit schematic of the DC-DC chopper-based traction drive system is given in Figure 1.7.

The main thyristor is T1. T2 and T3 are used as auxiliary circuits to divert the current from T1 to L2 with the help of a resonant circuit in order to switch off T1 [18].

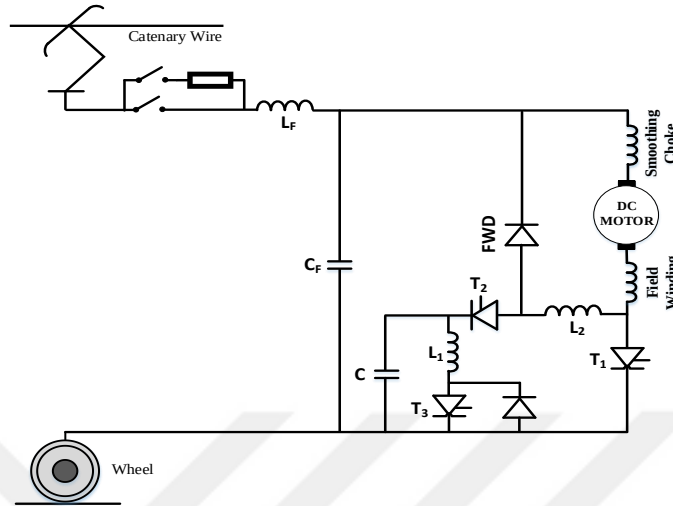


Figure 1.7. DC motor control with thyristor-based DC-DC chopper circuit [18]

Besides this, AC traction motors and gate turn-off (GTO) thyristor-based power converters were started to use in traction applications at the end of the 1970s with the developments in semiconductor and microprocessor technology. Since then, the usage of AC traction motors has spread quickly in railway applications. Figure 1.8 shows the schematic diagram of the inverter circuit [19].

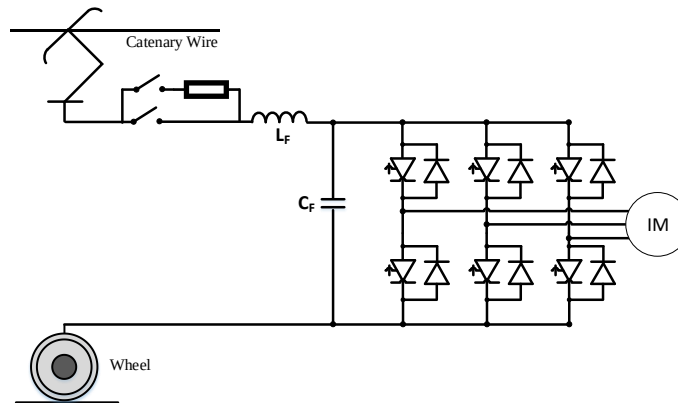


Figure 1.8. GTO thyristor-based three-phase traction inverter [19]

These developments had made AC traction motors and GTO thyristor-based power converters more attractive in the past. These traction systems are also called AC

traction drives. The advantages of AC traction drives can be expressed as follows by comparing them with DC traction drives:

- Higher power to volume and power to weight ratio,
- Higher efficiency,
- Low maintenance requirements,
- Higher terminal voltage and shaft speeds due to brushless design,
- Higher immunity for environmental factors such as snow, rain, humidity and etc.
- More robust mechanical structure.

However, GTO thyristor-based power converters have limited switching frequencies up to 500 Hz and require complex gate driver circuits. As a result of this, they create high harmonics and hardware implementation difficulties. After a while, the development of insulated gate bipolar transistors (IGBTs) gave traction converter manufacturers a chance to increase switching frequencies up to 1 - 3 kHz with the simple gate drive circuits. IGBTs helped to reduce harmonics while maintaining inverter efficiencies of more than 98%. Furthermore, IGBTs gave a chance to improve the control techniques of power converters. Currently, the IGBT-based traction converter, which is given in Figure 1.9, is used as a standard solution [20].

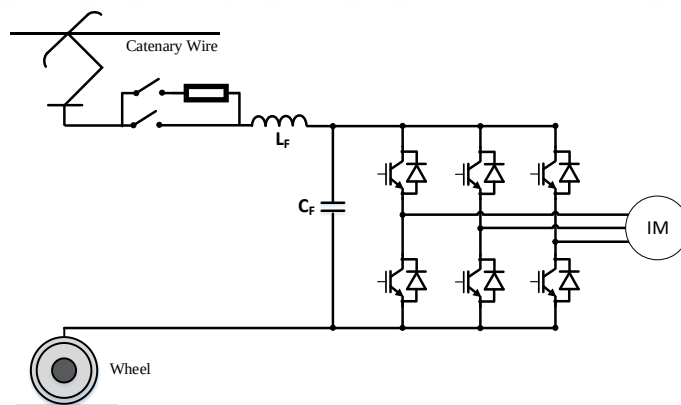


Figure 1.9. IGBT based three-phase traction inverter

Recent trends are on the specific emphasis of SiC power semiconductors due to the better switching characteristics, higher breakdown voltage capability, lower switching losses, capability to operate higher junction temperatures, and higher thermal conductivity. The use of SiC power semiconductors has a great influence on

traction converter volume and weight. By increasing power quality and efficiency, and decreasing cooling requirements, it lets to use simpler converter topologies and small size heatsinks [20]. Mitsubishi Electric declares that a 40 % reduction was achieved in losses, volume, and mass of the inverter [21]. The circuit diagram of the SiC MOSFET-based traction inverter is given in Figure 1.10.

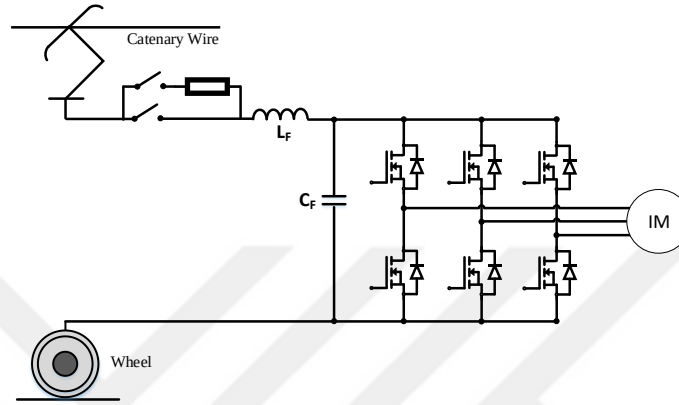


Figure 1.10. SiC MOSFET-based three-phase traction inverter

Two major types of AC traction drives are known, namely voltage source traction drives and current source traction drives. Basic block diagrams are given in Figure 1.11. Figure 1.11 (a) and (b) show the basic structure of the voltage source traction drives while (c) and (d) illustrate the current source traction drives.

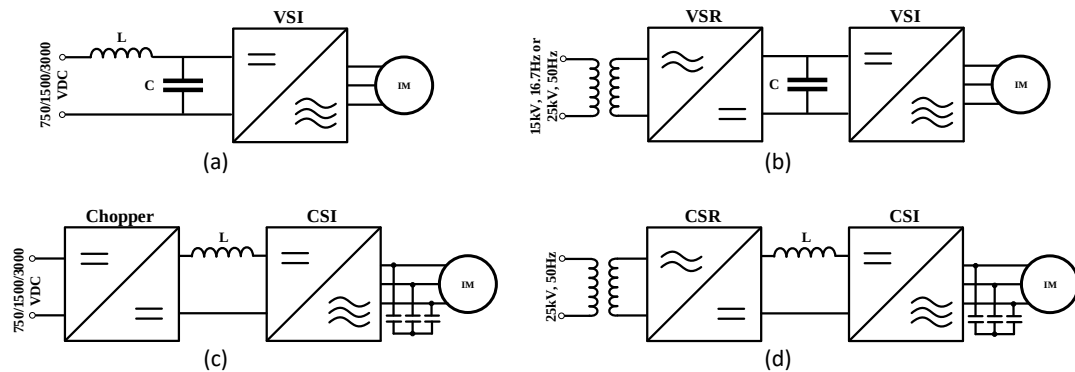


Figure 1.11. Voltage source & current source traction drives

In the literature, lots of studies were conducted to investigate control strategies and modulation techniques for both voltage and current source traction converters. However, the current source power converters are less common in railway traction

applications [22], [23]. For these applications, voltage source power converters are preferred due to the following superior features [24]:

- Ease of control,
- Low cost,
- Excellent dynamic performance,
- Small size,
- Lower total losses due to the lack of bulky inductive components.

In addition to this, AC traction drives are classified in the literature according to the voltage levels of converters, namely two-level and multilevel converters. Classification of different types of traction converter topologies is illustrated in Figure 1.12.

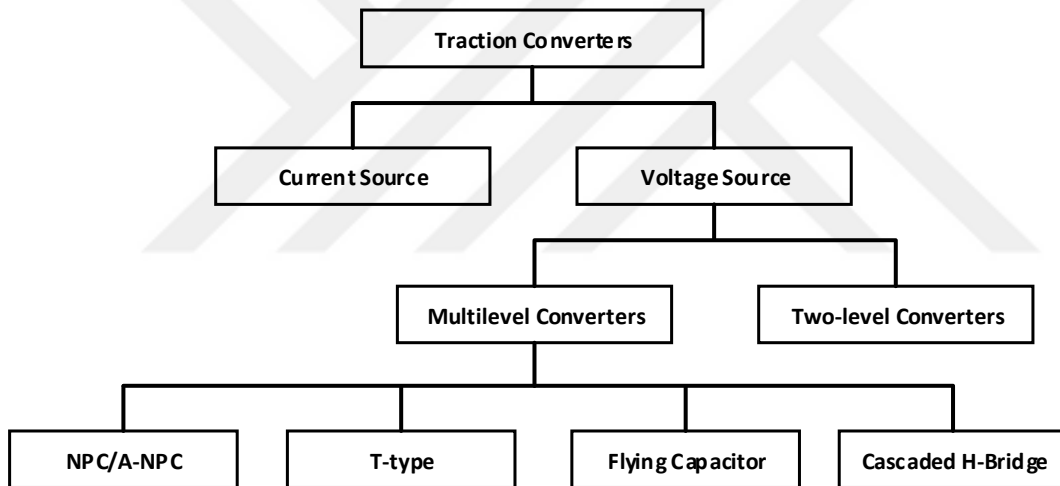


Figure 1.12. Traction converter topologies [24]

From the past, multilevel converters have been commonly used for medium voltage and medium power drive applications due to the lack of sufficient blocking voltages of semiconductors. In 1981s, a three-level neutral point clamped converter was introduced from A. Nabae, I. Takahashi, and H. Akagi. Years later, the multilevel converter technology, which is composed of several full-bridge converters connected serially, was presented. These converters, which appeared in 2002, are called modular multilevel converters [25]–[28]. Flying capacitor multilevel converter was also developed in the same years.

Multilevel converters gave a chance to use series-connected semiconductors with lower blocking voltages. Because of the applied lower voltage steps, the output voltage waveform quality is increased, and stresses on electric motor and traction transformers are decreased by a certain amount. However, on the other hand, the number of semiconductors, converter complexity, and isolated gate driver need is increased. Due to the increased number of components, the reliability of the converter decreases.

The frequently used multilevel traction converter topologies are described in the next section. With the increased blocking voltages of semiconductors, the world's leading traction drive designer companies have switched to two-level voltage source converter solutions. Multilevel traction converters are preferred in applications with a power level of 1.5 MVA / 2000 V and above or in applications requiring high current control bandwidth. Today, the classical two-level voltage source AC traction drives are commonly used in railway traction drive applications [24].

The traction drive supplies balanced voltage and current waveforms to the traction motor by taking its energy from the railway electrification line. The input/output voltage and current waveforms of typical AC or DC fed two-level voltage source traction drives are given in Figure 1.13.

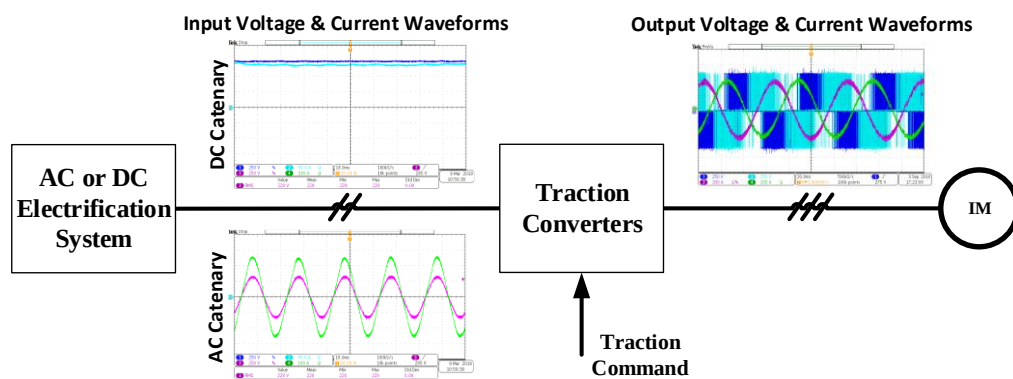


Figure 1.13. Typical input/output voltage and current waveforms

While the traction motor acts like the muscles in the human body, the traction converter represents the heart of the system. Because of these important tasks, continuous operation of traction drives is mandatory. Therefore, availability,

reliability, performance, and lifetime constitute important criteria for traction drive chain equipment.

1.3 Today's Power Converters for Railway Traction Drive

Demands for low volume, light weight, high reliability, high power, low cost, and high efficiency in railway traction are continuously increasing [29]–[31]. Due to these demands, the developments of semiconductor technologies and power converter topologies in the field of power electronics have been triggered. This section shows an overview of recently used power converter topologies for railway traction.

In order to decrease cost and complexity and increase reliability, the traction converter manufacturers primarily offer solutions by using simple power converter topologies together with mature and cheap semiconductor technologies. This approach provides to takes advantage of the simplest two-level converter topologies and their simple control methods. Simple hardware structures and control architecture of two-level converter topologies provide a chance for easy implementation. In addition to this, fewer component counts help to increase reliability and decrease the cost. Single-phase two-level voltage source rectifier and three-phase two-level voltage source inverter topologies are presented in Figure 1.14 (a) and (b).

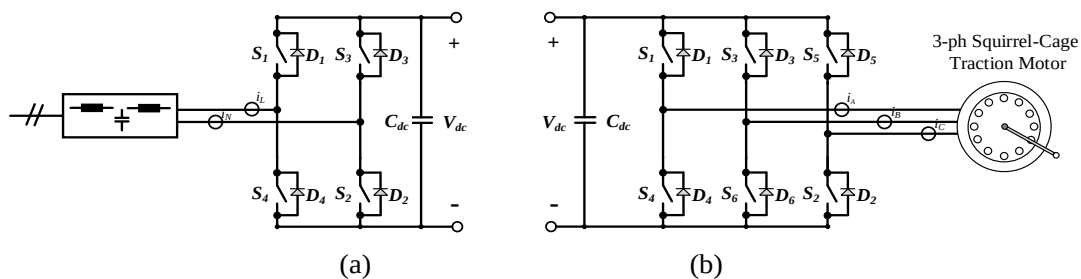


Figure 1.14. Two-level voltage source converter topologies a) Single-phase rectifier b) Three-phase inverter

In fact, it is not always possible to use the simplest topologies due to power loss and breakdown voltage limitations of semiconductors. In this case, the industry has to use mature semiconductors and multilevel converter topologies. Single-phase three-level NPC voltage source rectifier and three-phase three-level NPC voltage source inverter topologies are presented in Figure 1.15 (a) and (b).

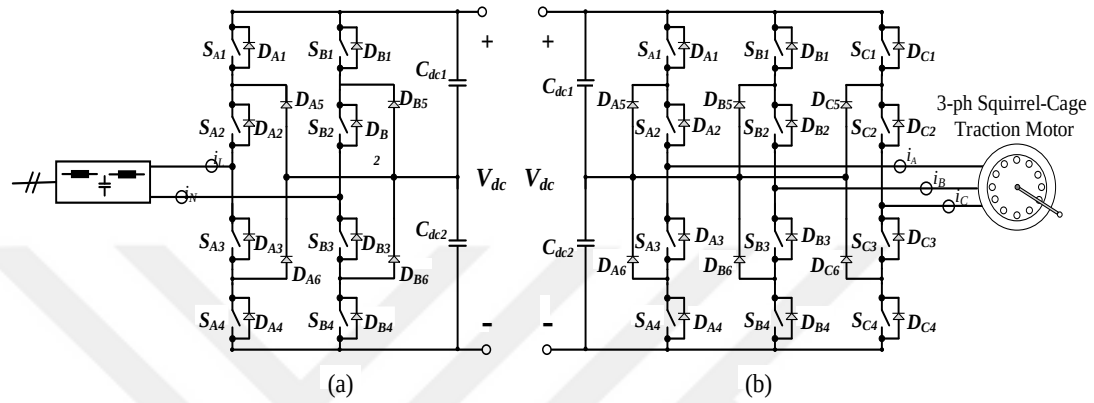


Figure 1.15. Three-level voltage source converter topologies a) Single-phase rectifier b) Three-phase inverter

Until today, the main applications where multilevel power converter topologies are used include high-speed train and mainline locomotive traction systems. Besides this, another application of multilevel power converter appears in magnetic levitation in railway traction. Basic needs arise from requirements of high-frequency control bandwidth (up to 400 Hz), high power, high reliability, and efficiency. In order to synthesize the required voltage and current waveforms with two-level power converters, relatively high switching frequencies are needed. However, this requires bulky and/or very expensive cooling systems in the case of IGBT-based two-level power converters used.

Multilevel converters not only meet the wide frequency control band, high power, and efficiency needs but also increase the power quality for the electrification systems and decrease the need for power filters, common and differential mode noises, stresses on electric motors, and passive components. Other multilevel converter topologies used in motor drive applications, as shown in Figure 1.16 (a),

(b), and (c), are three-level ANPC, three-level flying capacitor, and three-level T-type voltage source inverters, respectively.

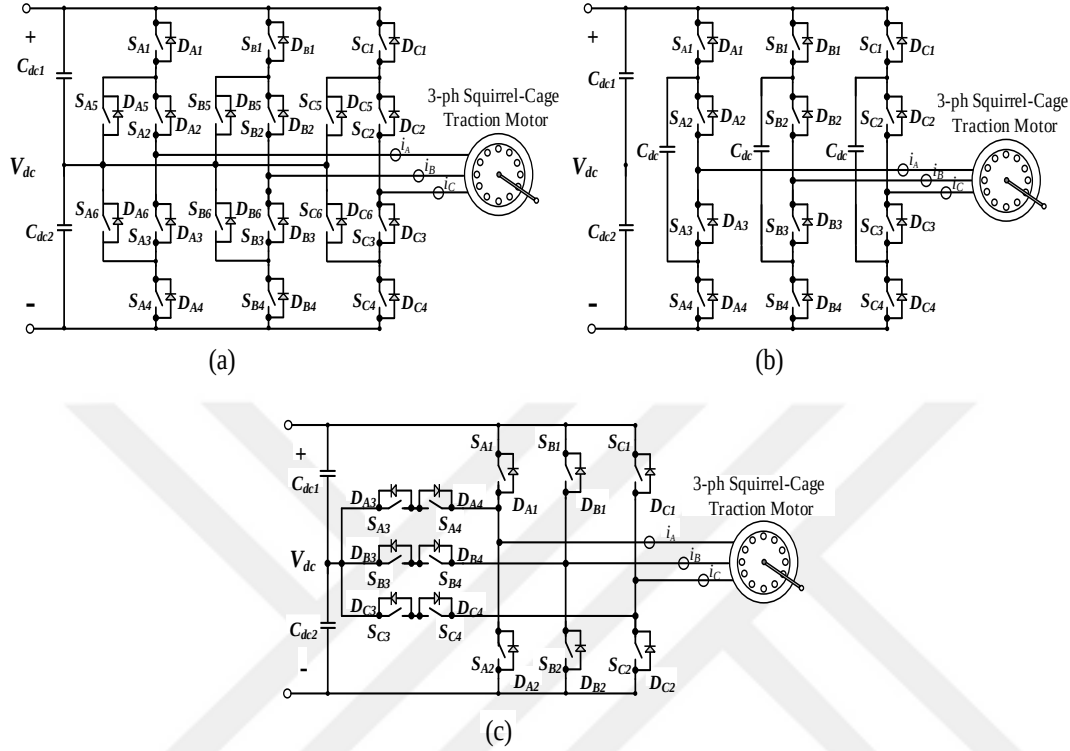


Figure 1.16. There-level three-phase voltage source converter topologies a) ANPC inverter b) Flying capacitor inverter c) T-type inverter

The use of multilevel converters also brings difficulties in the implementation of hardware and control systems because of complex circuit structures. Additionally, there is an imbalance problem in capacitor voltages and semiconductor temperatures [32], [33]. Moreover, reduced reliability and increased cost due to the increased component count create additional disadvantages.

Sustainability is the most critical issue in the industry. In order to be prepared for the future, it is obvious that the continuous development of power electronic technology is needed. Actually, new technologies provide huge opportunities that can motivate the industry for further improvements and developments. Among them, innovative semiconductor technologies are the main ones. Especially, the wide bandgap devices are grateful examples of this innovative process. The wide bandgap semiconductors

give a chance to operate at high efficiencies with higher current and voltage ratings. By using this technology, the level of multilevel converters can be decreased. Thus, wide bandgap semiconductors will make it possible to move away from complex converter topologies and return to more simple converter topologies for railway traction converters. This increases the appetite in the sector and ensures that measures are taken to accelerate the transition process in order to develop new technology semiconductors and make them widespread [32]. However, there are also drawbacks. The new technology semiconductors are more expensive in the early days of their release. In addition to this, they come with some challenges. They have very fast current and voltage switching capability. Due to this fast switching transition, electrical stress on electric motors and passive components increases. This requires the use of improved insulation materials. Moreover, the conducted common mode emissions and radiated emissions also increase. In order to overcome this shortcoming, parasitic capacitances must be minimized in both semiconductor packaging and inverter design. Moreover, working at high temperatures creates additional challenges for semiconductor packaging. Because such issues need to be solved, especially in the first years of newly developed semiconductor technologies, the widespread use of new technology semiconductors requires a transition period.

The solid-state transformer is another challenge for railway applications, which also requires the use of multilevel converters. In Figure 1.17, the schematic of this topology is illustrated. The solid-state transformer is more complex than the conventional low-frequency traction transformers, but it can be smaller and efficient. In Figure 1.17, a symbolic circuit diagram of SST is given. According to 2021 figures, this SST topology shows low reliability compared to a low-frequency transformer equipped with an active rectifier [34]. However, the technological improvements on semiconductors, capacitors, medium frequency transformers, gate-drivers, and so on will give a chance to reach longer life times for all those components. These will help to widespread similar topologies. Moreover, the widespread use of this topology also depends on other applications. Medium-voltage direct current (MVDC) transmission systems are expected as the future energy

transmission system. Therefore, the solid-state transformers will be a pioneer in preparing railway vehicles for the power distribution lines of the future.

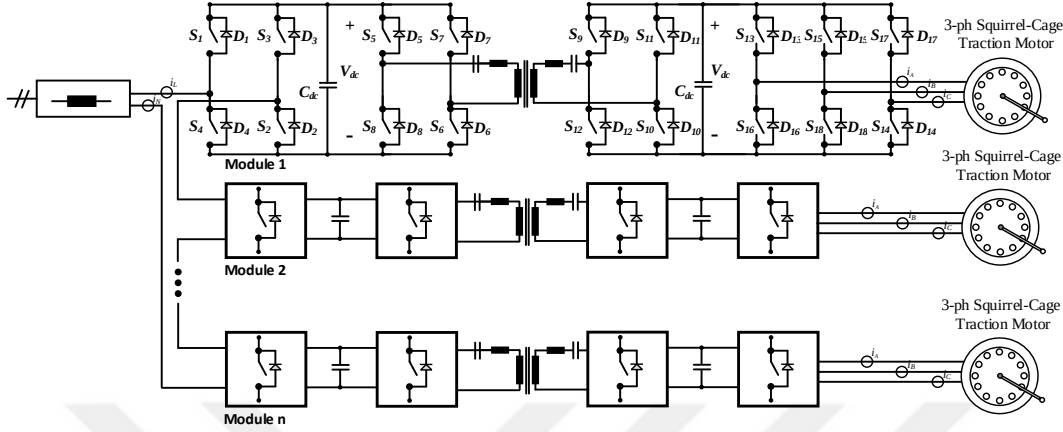


Figure 1.17. Solid-state transformer example for railway applications

Every time, the development of innovative industrial solutions poses new challenges but also creates new chances for the future of human beings.

1.4 Power Semiconductors and Development Trends for Railway Traction

Modern traction drives require an overwhelming level of technology, from semiconductors to microcontrollers. In this section, the semiconductor technologies are only evaluated.

Thyristors were used for the control of the DC traction motor in the past. Then, the GTOs were developed and used for a three-phase induction motor drive. Subsequently, IGBT technology was introduced, and IGBTs began to replace GTOs in railway traction applications. After those years, IGCT technology was announced in 1997 and started to use for medium voltage and medium power variable frequency inverters, energy management, and motor drive applications. There were a few examples of railway traction applications with IGCTs. On-state voltage drops are lower than IGBTs. However, they require a di/dt limiting inductor to keep control reverse recovery losses of antiparallel diodes which stay in series path, and a complex, bulky gate drive and voltage-clamp circuits. Moreover, possible switching

frequencies of IGCTs remain lower than IGBTs for railway traction applications which require relatively high switching frequencies. IGCTs are better suited for high power but relatively low switching frequency applications [35]. As the widely used technology in railway traction, IGBT-based traction systems provide a chance to operate relatively high switching frequencies and high system efficiencies. This provides less volume and less weight due to the simple and low power gate drive circuit and basic cooling system needs. In addition to this, relatively high switching frequencies give a chance to obtain better controllability, superior dynamic performance, wide current control bandwidths, low torque pulsation, high reliability, low acoustic noise, low cost, and modular design as compared to the GTO and IGCT based traction systems. With these advantages mentioned above, IGBT technology seems to have made a tremendous contribution to the development of modern traction drives. Nevertheless, this technology is reaching its limits, especially in terms of breakdown voltage, switching losses, thermal properties, and current-carrying capacity. Besides these, new semiconductors based on gallium nitride (GaN) and silicon carbide (SiC) are offering abroad opportunities. In Figure 1.18, the key characteristics of Si, GaN, and SiC materials are given.

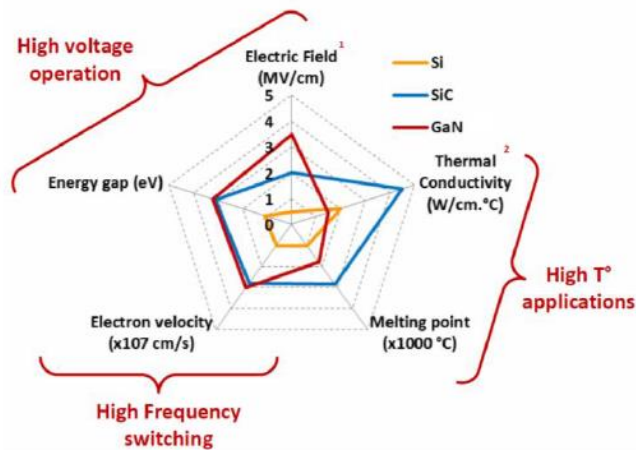


Figure 1.18. Characteristics of Si, GaN, and SiC materials [31]

Nowadays, the state-of-the-art power switch based on SiC semiconductor technology is beginning to be widely applied to traction and power conversion

applications. With the use of SiC semiconductors, further improvements have been made to all features described in IGBT converters. Moreover, the lower switching losses and increased breakdown voltages of SiC semiconductors allow the use of simpler converter topologies. Voltage and current ratings of commercially available power semiconductors for railway traction applications are tabulated in Table 1.2 [36]–[42].

Table 1.2 Power Semiconductors for Railway Traction

Manufacturer	Semiconductor Type	Voltage (V)	Current Range (A)	Status
Fuji	Si-IGBT Module	1700	600-3600	Active
Fuji	Si-IGBT Module	3300	800-1500	Active
Hitachi ABB	Si-IGBT Module	1700	800-3600	Active
Hitachi ABB	Si-IGBT Module	3300	250-1800	Active
Hitachi ABB	Si-IGBT Module	4500	150-1500	Active
Hitachi ABB	Si-IGBT Module	6500	400-1000	Active
Hitachi ABB	SiC MOSFET Module	1700	900-1800	Sample
Hitachi ABB	SiC MOSFET Module	3300	500-1000	Sample
Hitachi ABB	Si-IGCT Module	2800	1430-2660	Active
Hitachi ABB	Si-IGCT Module	3300	1290	Active
Hitachi ABB	Si-IGCT Module	4000	1290	Active
Infineon	Si-IGBT Module	1700	600-3600	Active
Infineon	Si-IGBT Module	3300	200-2400	Active
Infineon	Si-IGBT Module	4500	400-1200	Active
Infineon	Si-IGBT Module	6500	250-750	Active
Mitsubishi	Si-IGBT Module	1700	600-2400	Active
Mitsubishi	Si-IGBT Module	2500	400-1200	Active
Mitsubishi	Si-IGBT Module	3300	400-1800	Active
Mitsubishi	Si-IGBT Module	4500	450-1500	Active
Mitsubishi	Si-IGBT Module	6500	200-1000	Active
Mitsubishi	SiC MOSFET Module	3300	375-750	Active
Wolfspeed	SiC MOSFET Module	1700	225	Active

In Table 1.3, these switches are qualitatively compared according to their native characteristics [43]–[49].

Table 1.3 Semiconductor Comparison

	Si-IGBT Module	Si-IGCT	SiC MOSFET Module	Impact
Snubber need	No	Yes	Yes	Power density
On-state voltage	Moderate	Low	Low	Power density
Current density	Moderate	High	High	Power density
Gate driver difficulty	Low	Moderate	Low	Power density
Switching frequency	<3kHz	<1kHz	<20kHz	Power density
Failure mode	Open	Short	Open	Reliability
Short-circuit current	Gate driver limited	Snubber limited	Gate driver limited	Reliability
Series connection	Easier	Moderate	Easier	Reliability
Parallel connection	Moderate	Difficult	Easier	Reliability
Thermal cycling capacity	Moderate	High	High	Reliability

The switching frequencies could not be increased due to the high switching losses of previously existing power semiconductors and switching frequencies that remain significantly lower than today's state-of-the-art semiconductors. This results in low power quality (i.e., high amount of harmonics), poor system efficiency, torque pulsation, narrow current control bandwidths, and slow response.

High switching frequencies can be considered more suitable for traction applications requiring relatively wide current (i.e., torque) control bandwidth and high power quality (e.g., low-frequency harmonics are significantly reduced). Wide control bandwidth helps to increase system stability. In addition to this, lower motor current harmonics lead to increase system efficiencies. With the improvements of semiconductors and microprocessors, intelligent modulation methods, voltage, and current control algorithms have been developed. In the future, higher control

bandwidths that can be achieved with innovative control algorithms and semiconductor technology will also help traction motors to operate at very high speeds. This may help to reduce traction motor volume and weights. With the continuously increasing switching frequency capability of the power semiconductors, the power electronic converters will provide more benefits at the system levels.

In addition, short circuit protection, overload, thermal cycling, and power cycling capabilities, easy series and parallel operation, power density, and safe state of power semiconductors under fault conditions create other important requirements for railway traction applications. Parallel to the above features and being commercially available, IGBTs appear to be widely used in traction applications. However, on the other hand, when SiC MOSFETs become widespread, it is seen that IGBTs will be replaced by SiC MOSFETs.

1.5 Problem Statement and Motivation

In the railway industry, operators typically compare several features such as driving performance, load capacity, passenger comfort, initial and operating cost, reliability, availability, and maintainability. All operators know the importance of energy consumption to decrease operating cost.

A reduction in energy consumption should be achieved without resulting in setbacks with respect to the features such as driving performance, load capacity, comfort, operating cost, reliability, availability, and maintainability. Up to now, many electric-powered railway traction systems have been built around the world. As of today, most of the electric-powered traction systems widely use silicon-based insulated-gate bipolar transistors (IGBT). In medium power drives, like electric-powered traction systems, the switching frequency of IGBTs is limited to between 500-3000 Hz due to the considerable amount of switching losses. This restriction creates many problems. Low-order harmonics may appear due to low switching frequencies. This situation decreases power quality and overall energy conversion

efficiency and creates torque ripple at integer multiples of the fundamental frequency. In order to eliminate this low order harmonics, special pulse width modulation techniques such as Selective Harmonic Elimination (SHE), Selective Harmonic Mitigation (SHM), Optimum Pulse Width Modulation (OPWM), and etc. have to be used. However, this necessity makes the control algorithms and modulation techniques complex. In addition, the elimination of low order harmonics may still not be possible due to the limit of effective switching frequency if the fundamental frequency of the output waveforms is quite high as usual in motor drive applications. In this case, the pulse number has to be decreased because of the effective switching frequency reached in such modulation methods. Anymore, torque ripples and motor losses cannot be reduced. This ends up with a decrease in overall system efficiency and an increase in operating cost. Moreover, lower switching frequencies also yield narrow current and torque control bandwidth. This makes poorer the system stability. All these issues show that resulting in some setbacks are inevitable with IGBT-based two-level power converters. Such problems can be solved with IGBT-based multilevel converters, as described in the previous section. However, this time the complexity of the hardware and control algorithms increases. In addition, the reliability decreases as the number of components increases.

Nowadays, SiC semiconductor technology has begun to gain importance in the application of power electronics. SiC semiconductor-based traction systems provide promising opportunities for energy saving and better power quality. The SiC-based power switches give a chance to reduce switching losses and increase operating temperature. This situation leads to increased efficiency, power to weight, and power to volume ratio by decreasing cooling system requirements. Besides this, SiC-based power switches allow operating high switching frequencies at higher voltage and current levels. This makes it possible to use simpler converter topologies instead of multilevel topologies, eliminate complex control algorithms and low order harmonics and increase current control bandwidths. Especially, this new technology which attracts the attention of the academy and the industry has many issues that

need to be researched in terms of application areas such as traction and propulsion drives, medium voltage industrial drives, medium voltage DC and high voltage DC energy transmission systems.

1.6 Scope of the Thesis

In this thesis, it is aimed to significantly reduce control system complexity, increase efficiency and power density by providing high-quality ac currents at higher output fundamental frequencies. In order to achieve this target, all SiC power MOSFET-based 500-900 VDC, 165 kVA two-level traction inverter has been developed and implemented. The capabilities of the developed system have been investigated by simulations and laboratory test studies. Field-oriented control method and space vector pulse width modulation technique has been implemented to control traction motor. The prototype system is composed of a power electronic block, inverter control unit, voltage and current transducers, LC input filter, cooling system, main line contactor, soft start contactor, and resistor. The cut-off frequency of the low pass filter is set below the mains frequency. An air-core line reactor is chosen as the input filter element in order to prevent magnetic saturation under short-circuit failure cases. This feature is very important to limit the current in the event of a short-circuit. New technology semiconductors are used to build the power electronics block in the simplest form. 10 kHz switching frequency is used to minimize the current harmonics even at the higher output fundamental frequencies. This study is the first example of a SiC power MOSFET-based traction drive developed for rail transportation vehicles in Turkey.

In addition to this, the effects of SiC power semiconductors on the selection of converter topologies, passive components, cooling systems, motor windings, and bearings are discussed. Moreover, the impact of SiC-based power converters on the copper and iron losses in the motor is explored systematically.

In order to validate these effects on motor drive applications, a full-scale physical simulator for all-SiC traction motor drive was designed and implemented. This test

setup is equipped with software programs for a catenary model, rail track model, and vehicle model and permits the performance verification of various control, modulation, and energy-saving strategies. Finally, laboratory tests are conducted on this setup.

This research work makes following original contributions to voltage source inverters in railway applications:

- one of the first design and development works of all silicon carbide (SiC) traction converter for new generation light rail transportation systems,
- presentation of qualitative design criteria based on analysis, simulation and experimental work for SiC power MOSFET based traction converter,
- first design and implementation of a full scale physical simulator of an all SiC traction motor drive for LRTS with on-board super-capacitor energy storage system,
- verification of designed new generation traction converter with full scale physical simulator.

The outline of the thesis is given below:

In Chapter 2, a description of light rail transportation systems, system architectures, traction motors, operating principles of two and three-level voltage source traction inverters, control methods, and modulation techniques are presented.

In Chapter 3, the design methodology of the traction inverter is described. The design requirements of the prototype system are stated. The selection of the critical components is explained. Semiconductor power loss calculations are investigated for two-level and three-level traction inverter topologies with candidate semiconductors. The overall system efficiencies are investigated. Traction inverter power block design is illustrated. The motor control algorithm, including active resonant damping, is presented.

In Chapter 4, the modeling studies of the MOSFET module, traction drive system, and test setup are presented. Modeling studies of the rail vehicle and track are briefly

explained and the method used to derive the system requirements is defined. The simulation results under steady-state and dynamic regimes are investigated.

In Chapter 5, the design and the implementation of the test setup are presented. The contributions of this test bench are explained.

In Chapter 6, the laboratory test results are investigated.

In Chapter 7, the future prospects of SiC power MOSFETs for power electronics applications are presented.

In Chapter 8, conclusions are stated, and proposals for further works are given.

In Appendix A, experimental results of the SHEPWM technique are presented.

CHAPTER 2

LIGHT RAIL TRANSPORTATION SYSTEMS

A brief introduction to railway traction systems is presented in Chapter 1. It has been basically summarized as follows:

- The electric traction systems can be fed from DC power stations through the third rail or the catenary systems, and AC power stations through the catenary systems,
- The DC electrification systems are commonly composed of six and twelve pulse diode rectifiers,
- Low-frequency harmonics are present on DC electrification systems, and line regulation is weak due to diode rectifier based power stations,
- Future DC fed traction systems for railway vehicles will conform with system nominal voltage of 750 VDC or above voltage levels [17],
- DC traction motors have been used in early traction systems, while new systems are containing AC traction motors,
- DC fed traction systems require braking resistor and chopper due to the nature of diode rectifier base electrification system,
- The common input voltage levels are 750 VDC, 1500 VDC, 3000 VDC, 15 kVAC ($16^{2/3}$ Hz, 50 Hz) and 25 kVAC (50Hz) [16],
- AC-fed traction systems mostly do not require a braking chopper and resistor. Instead of this, low power voltage limiting circuit and resistor are used,
- State-of-the-art power converters are always required,
- Current and voltage source converter topologies are both applicable. However, almost only the voltage source converters have been used due to better features.

In this chapter, system architecture, system components, their functions, and requirements of the railway vehicles are firstly investigated. In addition, the physics of vehicle motion and various parameters which affect the size of the traction system and overall system efficiency is introduced. A basis for traction motors is presented. Then, voltage source two-level and multilevel traction converter topologies are discussed, considering reliability, cost, and performance. This is followed by explaining the operational principles of mostly used traction converter topologies. Finally, different modulation techniques and control methods that are commonly used for traction converters are summarized.

2.1 System Architecture

Electric powered railway vehicles can receive electrical energy from electrification lines through catenary or third rail systems and energy storage systems (such as batteries, supercapacitors, or a mix of them) and then convert electrical energy into kinetic energy through the traction drive systems. A basic block diagram of the multi-system traction converter (e.g., the traction converter can work with different types of electrification systems.) which shows almost all components in a traction chain, is presented in Figure 2.1.

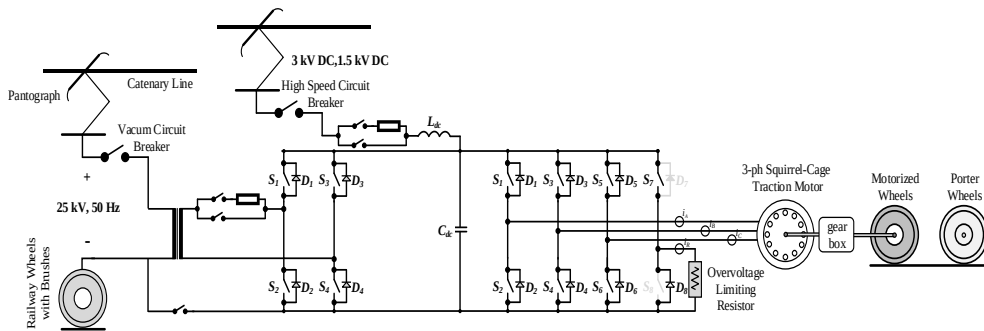


Figure 2.1. Multi-system traction inverter

The chain of the electric traction system consists of electrical transmission devices, protection devices, traction transformers, traction converters, traction motors, gearboxes, and wheels.

Pantograph or third rail brush draws energy from the power line. Depending on the types of electrification system, the energy first feeds the traction transformer (for AC fed converters) or directly to input filters of the traction inverter (for DC line fed converters). In fact, the traction converter with its superior control algorithms and functions is the core of the traction system. It acts as the heart of the traction system. It takes electrical energy and synthesizes required current and voltage waveforms for the traction motors to move trains. The traction motor acts as a muscle in the human body. It converts electrical energy to kinetic energy. Mechanical energy is transferred to the wheels via the gearbox.

Traction converters have direct or indirect interfaces with many system components, including different engineering disciplines such as vehicle control computers, traction transformers, traction motors, encoders, protection devices on railway vehicles, and electrification lines. It is worthy to note that the types of energy sources and the voltage level have a strong effect on the design of traction converters.

The types of railway electrification systems are defined in Chapter 1. According to applications, AC/DC electrification lines with different voltage levels are used in different countries. The choice of energy sources is strictly dependent on vehicle requirements. Low voltage DC electrification systems (750 VDC and 1500 VDC) and batteries/supercapacitors are commonly used for metro and light rail vehicles, which require higher acceleration/deceleration rates and frequent starting-stopping. In other words, low voltage electrifications are more common for shorter distances and low power applications. When the rail vehicles are supplied with DC, there is no need to perform AC/DC conversion at the input stage of the traction converter; DC input of the traction converter can be directly fed from the DC electrification line. DC fed systems mainly consist of traction inverters, traction motors, gearboxes, protection equipment such as high-speed circuit breakers (HSCB), surge arrestors, fuses, braking choppers [24]. These systems do not contain big and heavy power transformers. Thus, the traction system chains are cheaper and lighter. On the other hand, 3000 VDC and 15/25 kVAC electrification systems are mostly used for long distances and high power trains such as regional trains, very high-speed/high-speed

trains, and mainline locomotives. Unlike the DC-fed traction systems, AC-fed traction systems need an AC/DC conversion stage at the input stage of the traction inverter. Therefore, input (line) converters and traction transformers are additionally used in AC-fed traction systems.

In summary, the main requirements of a traction system can be defined according to the following constraints:

- Electrification system specification
- Vehicle specifications (e.g., empty weight, payload capacity, type, etc.)
- Drive cycle and special requirements (e.g., acceleration and deceleration rates, maximum speed, etc.)
- Track profile (e.g., distances, grades, curves, etc.)
- Network management (e.g., number of vehicles in operation)

2.1.1 Functions of Traction System Components

The main components of traction systems are summarized under the following topics:

1) Pantograph or Third Rail Brush:

There are two different types of contact to transmit the electrical energy to railway vehicles, namely third rail contact or catenary contact. The third rail contact systems are located on the ground. Hence, this type of contact has to be protected to avoid injury, dead or electric shocks of people. Because of that reason, this type of contact is mostly used for underground vehicles at low voltage levels such as 600 VDC or 750 VDC. The most important advantage of the third rail contact is that it allows tunnel diameters to be opened small and consequently reduces tunneling cost. On the other hand, due to safety reasons, above 1000 V, catenary type contacts are used to

feed traction systems of trains. Third rail brushes are used in third rail contact systems, whereas pantographs are used in the catenary contact systems.

A pantograph or a third rail brush is an electro-mechanical device. The electrical energy is supplied through the pantographs or the third rail brushes to the rail vehicle from the electrification system (catenary or third rail) during the working time of the vehicle. Thus, these electro-mechanical devices have to maintain good electric contact under all running speed conditions. Because of the increasing ground disturbances with speed, it is difficult to provide good electrical contact. Springs or air pressure is used to ensure good electrical contact between the electrification line and the pantograph or third rail brush [24].

2) Surge Arrester:

A surge arrester is a device to protect electrical equipment from overvoltage transients that can be caused by external (lightning) or internal (switching) events. The surge arrester should be connected between the transmission line and the ground in front of the other protective devices such as circuit breakers (both sides of the circuit breaker should be considered) to protect all electrical equipment against overvoltage transients. A surge arrestor should have a low impulse ratio to provide a fast response time [24].

3) DC or AC Circuit Breaker:

DC or AC circuit breaker is a protection device to be used for safely interrupting high currents that are caused by an electrical failure of the power system components during the operation of a train. The response time of the circuit breaker is very important. Faster response time reduces the effect of short circuit stress on traction transformers, traction converters, traction motors, insulators, and cables/conductors [24], [50].

4) Traction Transformer:

Traction transformers are used to step down the AC catenary voltage to safe voltage levels to transmit on the vehicles and appropriate voltage levels for traction converters. There are two different solutions. The first and oldest one is a single-phase low frequency ($16^{2/3}$ Hz or 50 Hz or 60 Hz) traction transformer. It usually steps down the 15 kV or 25 kV catenary voltage to traction converter operating voltage in the range of 800 V – 2000 V. Traction converters that fed on AC catenary line, mainly composed of a four-quadrant single-phase line rectifier and a four-quadrant three-phase inverter. The single-phase line rectifier takes AC input voltage and converts it to DC voltage while it is operating almost unity power factor.

There are also some limitations on catenary current harmonics. If the transformer leakage inductances are sufficient to keep harmonics below the required limits with the help of converter modulation methods, there is no need to use an additional input filter at the traction converter input. Otherwise, it will be necessary to use an additional input filter at the transformer and/or (typically a line reactor) at the converter input. Moreover, for the high power levels (e.g., 1 MVA or above), 100 Hz voltage ripple is so strong on DC link voltage. In order to filter it, second-order tuned filters are used in high-power applications. This filter is composed of a main capacitance block and a reactor. This reactor is mostly embedded in the traction transformer due to the high cooling requirement.

It is worthy to note that the second and innovative solution is to use medium frequency (several kHz) traction transformers. The catenary voltage directly feeds to the cascaded H-bridge multilevel rectifier. The output of each H-bridge converter is connected to a resonant converter and medium frequency traction transformer. Thus, the transformer volume and weight can be reduced considerably due to the medium frequency switching operation. Although there are some examples of medium-frequency traction transformers for 15 kV/ $16^{2/3}$ Hz electrification, this solution has not become widespread. Due to the breakdown voltages of existing IGBTs, many H-bridge converters have to be connected in series. This situation decreases the benefits

that can be achieved with medium frequency transformers such as lightweight, low volume, high efficiency, and reliability. It is considered that this solution can become widespread with the high breakdown voltages that can be reached by state-of-the-art SiC semiconductors [24], [51].

5) Traction Converter:

Traction converters are responsible for vehicle motion. They take electrical energy from electrification systems (AC or DC systems) and generate appropriate voltage and current waveforms to drive traction motors. The traction converter design varies according to the electrification system. For AC fed traction converters, two main types of power converters are located in the traction converter box, such as a four-quadrant single-phase line rectifier and a three-phase inverter. For DC fed traction converters, there is only an inverter power block as the main part. Since the traction converters are examined in detail within the scope of the thesis, this part was kept simple [51].

6) Brake System:

Brake systems can be specified as the most important subsystem of a railway vehicle in terms of safety. In railway applications, the brake systems are categorized under two main groups, namely as adhesion brake systems and non-adhesion brake systems. Mechanical brakes and electric brakes are located under the adhesion type brake systems, while air resistance brakes and electro-magnetic rail brakes are in non-adhesion type brake systems. Adhesion-type brake systems have utilized the friction between the rail and wheel. During rain and snow, the friction coefficient between the rail and wheel decreases, and dirty/iced/wet rails make it impossible to apply strong braking force. If the brake force is too high for those rail conditions, the wheels start to slide. This situation makes the drive systems unsafe by prolonging the stopping distance and negatively affecting vehicle dynamics. It can also damage the wheels by disrupting the surface of the wheel. In order to prevent this situation, wheel slide protection systems are used for both mechanical brakes and electric

brakes by utilizing sanding. They detect the wheel slide and reduce the applied braking force. Besides this, the non-adhesion type brake systems do not use friction between the rail and the wheel. Hence, in some railway applications, especially in tramways and high-speed trains, the electro-magnetic rail brake systems are also preferred in addition to the adhesion type brake system to increase the drive system safety and decrease the stopping distance under unexpected conditions [52].

7) Energy Storage System:

The energy consumption of trains depends on drive cycles, vehicle speed, vehicle volume, vehicle weight, the efficiency of power converters, energy recovery ability, electrification systems, and track profiles. Especially, vehicle weight depends on the material (e.g., aluminum or steel) of car bodies, the weight of traction system components, auxiliary power converters, air conditioning power unit, and passenger capacity [24]. In order to decrease energy consumption, important attempts have been made to reduce vehicle weights. Aluminum-made bodies are used to decrease vehicle mass. In addition to this, with the help of newly developed technological power semiconductors, the size and weight of the traction converters and auxiliary power converters are further reduced [21]. The energy efficiency of power converters is also improved together with the weight and size reduction. These improvements contribute to the reduction of energy consumption. Besides this, network management and drive cycle strategies, together with modern electrification systems, also make a great contribution to reducing energy consumption. Because, during the braking process, most of the energy is still lost as heat on the braking resistors to prevent an overvoltage on the catenary for DC fed traction systems. If this braking energy can be delivered to the other vehicles operated at the same network or delivered back to the mains, a great amount of braking energy will be recovered.

Another way to recover braking energy is the use of energy storage systems. If energy storage systems are used on the vehicle or at power substations, a great portion of the regenerative energy can be recovered. This stored energy can further

help to reduce energy consumption and improve catenary line regulation when it is used at the instant of acceleration process of the vehicle.

Figure 2.2 shows a classification of existing energy storage technologies considering the power and energy densities [53].

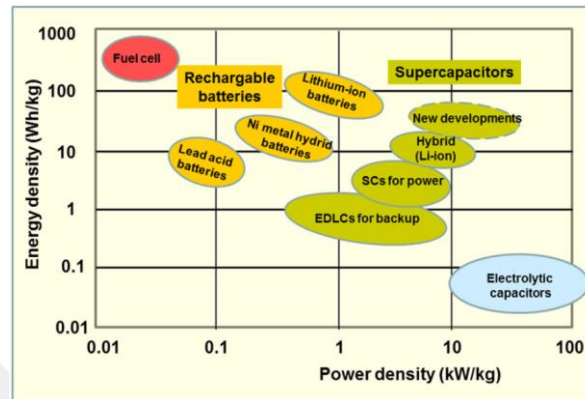


Figure 2.2. Ragone chart of energy density vs. power density of different energy storage devices [53]

In fact, the braking times are usually less than 30 seconds for metro and light rail vehicles. However, peak power levels are very high in this time interval. Therefore, the energy storage technologies to be used in the recovery of the braking energy must accordingly have a high power density. Besides this, there may also be a need for different energy storage technologies that require high energy density in applications where catenary free operation is required. Moreover, both specifications are needed together in some applications due to the high power and high-energy requirements. Ragone's chart shows that supercapacitors fit quite well for these requirements.

8) Braking Chopper & Braking Resistor:

There are two types of electric (or dynamic) braking, namely, regenerative braking and rheostatic braking. During dynamic braking, the traction motors are used as a generator. If the generated electrical energy is returned to the electrification system, it is called regenerative braking, and if the generated electrical energy is dissipated as heat in a resistor, it is called rheostatic braking. For both cases, incoming electrical

energy charges the DC-link capacitors and causes a voltage rise at the DC-link voltage. If the DC-link voltage reaches a very high value (e.g., comparable levels with breakdown voltages) during the dynamic braking, the power semiconductors and DC-link capacitors will be damaged [24].

For the DC fed traction systems, the line voltage increases at the braking instant due to the nature of diode bridge rectifier-based electrification systems. Some or all of the generated energy can be used by the other railway vehicles if the working conditions are fully appropriate. However, this state will not be possible most of the time. In order to control DC-link voltage, electrical energy can be sent to the electrification system, be stored in energy storage systems, and be burnt in onboard braking resistors with the help of braking choppers [54]. Generally, for AC-fed traction systems, braking resistors and choppers are not used due to the bidirectional power flow nature of AC electrification systems.

9) Traction Motor:

The main components which are used to convert electrical energy to kinetic energy are traction motors. Traction motors provide the required mechanical propulsion for the train. Traction motors are usually located in the bogie in order to minimize mechanical transmission needs and transmission losses. However, this also brings some difficulties together. Actually, there is not a big space for mounting the traction motors on the bogie. Because of this reason, power to volume ratios of traction motors are so strict. In the past, DC traction motors were used in railway systems. Nowadays, squirrel cage induction motors and permanent magnet synchronous motors are used in railway systems. These traction motors have superior advantages in terms of power to volume and power to weight ratio and low maintenance requirements. Nowadays, between these two traction motors, the squirrel cage induction motors are commonly used. The permanent magnet synchronous motors have not been spread yet in railway systems, although they are the most advantageous in terms of the properties mentioned above. Besides this, there are some special applications in which the permanent magnet synchronous motors are

used without gearboxes. They are directly coupled to the wheel shaft. This solution brings with it many advantages, such as electrical and mechanical efficiency improvements. On the other hand, this will bring along safety precautions to be taken on the traction motor and converter sides.

Cooling of the traction motors can be self-cooling or externally forced air or water cooling. In practice, self-cooled or forced air-cooled type of traction motors are mostly preferred in order to reduce extra need for infrastructure like as water cooling units on the vehicle [24], [51].

10) Gearbox:

In railway applications, a gearbox is generally required at the traction motor shaft, although there are a few examples of directly-coupled traction motors and wheels. The gearbox is a mechanical transmission component. It is located between the traction motor and the wheel together with a flexible coupling or direct coupling. On the traction motor side, the rotational speed is relatively very high, and the output torque of the traction motor is relatively very low comparing with the wheel side speed and torque requirements. The gearbox reduces the higher rotational traction motor speed to slower wheel speed and increases the torque of the wheel shaft with a fix ratio.

2.1.2 Types of Railway Vehicles

Not only the developments in power electronics but also the developments in vehicular technology are very important. For specific customer needs, innovative railway vehicles equipped with optimized traction drive systems are working around the world. Types of railway vehicles are mainly associated with travel distance and purpose. The main groups are given below:

- Mainline locomotives
- Very high/high-speed trains

- Regional trains
- Metros
- Light rail vehicles
- Tramways

Typical power and voltage specifications of vehicles are given in Table 2.1.

Table 2.1 Classification of Railway Vehicles [24]

Railway Vehicle	Supply Voltage	Traction Motor Power (kW)	Traction Converter Power (kW)	Top Speed (kph)
LRV & Tram	600/750/900 VDC	75-150	150-300	50-70
METRO	750/1500 VDC	150-250	350-1000	80-100
Regional Train	1500/3000 VDC, 15/25 kVAC 16.7/50 Hz	200-400	300-800	150-200
High Speed Train	15/25 kVAC 16.7/50 Hz	300-500	350-800	200-300
Very High Speed Train	15/25 kVAC 16.7/50 Hz	400-600	500-1400	350-500
Locomotives	1500/3000 VDC, 15/25 kVAC 16.7/50 Hz	340-1400	500-2500	100-200

Locomotives are fed by AC (15 kV or 25 kV) or DC electrification line (1.5 kV or 3 kV) by considering characteristics such as high load, high power demand, long-distance transport, and drive cycles. Although locomotives are mainly used for the transportation of goods, they are still used for passenger transport today. Locomotives have concentrated traction systems (e.g., the entire traction systems are located within locomotive).

Very high/high-speed trains are used for cross-countries and long-distance domestic passenger transport. Especially because of their speeds, these trains shorten travel times and create a strong competitor to other alternative means of transport. Besides this, regional trains/EMUs are used for intercity transport in the middle distance. Regional trains/EMUs provide an efficient solution for transport like high-speed

trains, which creates a good alternative for passenger transport when considering the intensity, risks, and comfort of land transportation.

On the other hand, Metro, light rail vehicles, and tramways are used for passenger transportation in urban residential areas.

2.1.3 Physics of Railway Vehicle Motion

The motion of a railway vehicle can be provided with various types of traction systems as described in the Chapter 1. However, modern railway vehicles are equipped with full electric drives, and the motion is provided by electric motors. The knowledge of railway vehicle dynamics and track data is so crucial for designing optimum electric drives.

In general, the aim is to shorten the journey time as much as possible while ensuring a comfortable journey. In order to achieve this target, railway vehicles should achieve their maximum speeds in a time as short as possible. After accelerating, the vehicle speed should be kept constant at high speed for a certain period, and then the vehicle should reach standstill in the shortest possible time.

While satisfying the requirements stated above, energy consumption should also be kept minimum as much as possible to decrease operating cost. In order to decrease energy consumption, coasting mode is usually utilized. In coasting mode, traction power is cut off, and the vehicle is kept without traction to avoid additional losses. Besides this, optimization of all traction system components is also important. By optimizing the traction transformer, traction converter, gearbox and traction motor, the vehicle weight can be reduced further, and the component efficiencies can be increased to decrease operating cost.

In Figure 2.3 (a) and (b), acceleration/deceleration profile and typical tractive effort at zero grade are given. Railway vehicles require a large traction force at low speed to provide the required acceleration rates, whereas considerably a small traction force is needed to keep its speed constant as much as possible at high speed.

There are four operating regions of vehicle movement, namely acceleration, constant speed operation, coasting, and braking. These areas can be visualized in Figure 2.3.

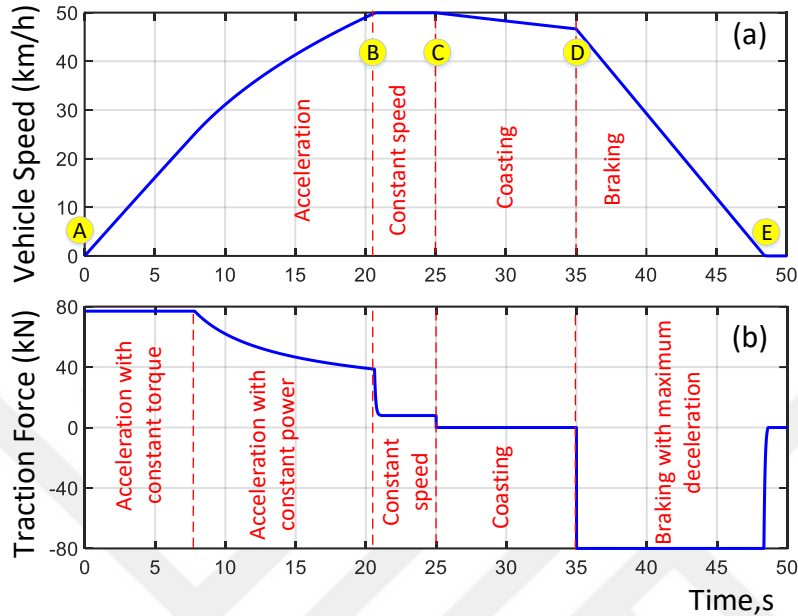


Figure 2.3. Typical acceleration and deceleration profile of a light rail vehicle

Practically, the shape of the speed-time curve of vehicles may be different than shown in Figure 2.3 because of the gradients, curvatures, and speed limits. However, this figure is good enough to explain operating regions. Four different operation regions are seen in the figure: acceleration period from A to B, constant speed period from B to C, coasting (free-running) period from C to D, and the braking period from D to E. The understanding of railway vehicle motion is very important to define the optimum traction system requirements. Because fully defined system models make it possible to accurately derive the traction system requirements. Therefore, the wheel-rail contact relation, the vehicle model and train resistance, track profile, and electrification system properties should also be well known.

1) Wheel-Rail Contact Relation

Traction motors convert electrical energy to mechanical energy and deliver it to the wheels via the gearbox. A considerable amount of torque is generated at the wheel shaft. However, the applied torque that can be transmitted from the wheels to the rail

is limited by adhesion force which is the transmitted tangential force in the longitudinal direction between the rail and wheels [55]. Low adhesion can limit the transmittable torque to a considerable level. The difference between the applied force and counter forces (e.g., train resistance and gravitational force) are used to accelerate or decelerate the train.

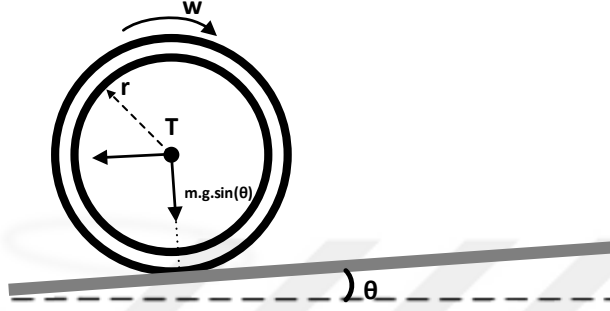


Figure 2.4. Forces acting on the rail

The F_N is the normal force acting on the rail. F_T is the tangential force which is also known as adhesion force. The ratio between the adhesion force and normal force is known as the adhesion coefficient, μ_0 [55].

$$F_N = mgsin(\theta)\mu_0 \quad (2-1)$$

$$\mu_0 = F_T/F_N \quad (2-2)$$

If this ratio is so small, the transmittable torque is strictly limited, and undesirable slip and slide conditions may occur.

The maximum transmittable torque, T_T can be calculated as follows.

$$F_T \leq mgsin(\theta)\mu_0 \quad (2-3)$$

$$T_T \leq mgsin(\theta)\mu_0 r \quad (2-4)$$

During the operation of a railway vehicle, the maximum torque should be limited according to these constraints; otherwise, the wheel will tend to slip or slide depending on the operation phase (e.g., acceleration or deceleration) [55].

2) Vehicle Model, Track Model and Train Resistance

In order to completely evaluate the power requirement of a railway vehicle, it is necessary to establish an adequate model which consists of the main components located on the railway vehicle. Figure 2.5 shows the power flow diagram for an AC-fed railway vehicle with the related components. In DC fed systems, a traction transformer does not exist, unlike the flow diagram stated below.

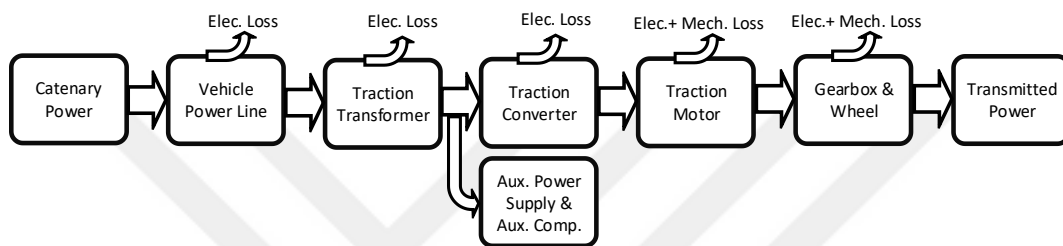


Figure 2.5. Power flow diagram of AC fed traction drive system

The power consumption of a rail vehicle can be split into seven main categories, which are associated with transmitted power to wheel and system losses:

- ❖ **Power Distribution Line Losses:** The current flows through a conductor, and this conductor has a finite resistance and a dielectric strength. For this reason, small amount of losses may be caused by conductor and stray current losses. In general, a very small part of the electrical losses is due to the onboard electrification cables. This is usually unimportant.
- ❖ **Traction Transformer Losses:** They include copper loss, core loss, and stray loss. The copper loss depends upon the resistance properties of the transformer windings, load current, and harmonics. Besides this, core loss is related to the power quality of the applied waveforms and the magnetic properties of the material used for the construction of the transformer core. There is also small amount of stray losses caused by stray currents. In general, the efficiency of the low-frequency traction transformers is between 93.5% to 96.5%.

- ❖ **Traction Converter Losses:** Converter losses include copper loss (bus bars, filters, cables and etc.), power semiconductor loss (e.g., switching and conduction), and stray current loss. A considerable part of the inverter losses may arise from power semiconductors and line reactor losses. In general, the efficiencies of IGBT-based traction converters are between 95% to 98%. Nowadays, the converter efficiencies can be improved with SiC power semiconductors over 99%.
- ❖ **Auxiliary Power Supply and Auxiliary Component Losses:** They include copper loss (bus bars, filters and etc.), power semiconductor loss (e.g., switching and conduction), reactive component loss, stray current loss, and auxiliary component losses for passenger comfort such as fans, cooling systems, heating and lighting systems, toilets, and etc.
- ❖ **Traction Motor Losses:** It is well known that the losses in a traction machine include stator and rotor copper losses, core loss, friction, and windage loss, and stray current loss. Nowadays, the squirrel cage induction motors are widely used in railway applications, and their efficiencies are usually between 92% to 94%. The traction motor efficiency can be increased up to 97% by using permanent magnet synchronous motors [56].
- ❖ **Mechanical Losses:** Traction motor output torque has to overcome the total load caused by gearbox friction, train resistance, curvature resistance of rails, gravitational forces in order to accelerate railway vehicles. The gravitational force is proportional to the instantaneous vehicle mass. The energy invested for the acceleration of a railway vehicle or uphill driving does not directly cause a considerable amount of losses. A large portion of the energy can be restored in the energy storage systems or transmitted back to the electrification line except for the copper, electro-magnetic and mechanical losses. If this energy is not stored or transmitted, it may be dissipated as heat on the braking resistor or mechanical brake. By the way, gearbox loss, train, and curvature resistances directly create unrecoverable power losses.

- ❖ **Transmitted Power:** The transmitted power from the wheel to rail is used to control the speed of the railway vehicle.

The losses, resistances, and counter forces mentioned above always oppose the vehicle movement except for gravitational force, F_{grv} . In areas where the slope is negative, the gravitational force acts in the direction of supporting the movement of the vehicle. Where it is positive, the opposite is valid. The effect of these forces on the train is represented in equation (2-5).

$$F_{applied} - F_{ext} = m_e a \quad (2-5)$$

where m_e is the equivalent mass of the train, which contains the static mass of the train m and the reflected mass from the inertias of wheels, gearboxes, motors, couplings, axles and etc.

The external acting forces can be written as in equation (2-6).

$$F_{ext} = F_{tr} + F_{is} + F_{grv} + F_{curve} \quad (2-6)$$

where F_{tr} is the sum of static and dynamic resistances and drag force acting on the vehicle, F_{is} is the initial static force of the vehicle system, which is dominant at starting instant and at very low speeds, F_{grv} is the gravitational force caused by the land slope, F_{curve} is the force countered by the curve resistance as stated in equation (2-7).

$$F_{curve} = DMg/R(x) \quad (2-7)$$

where, D curve resistance coefficient, and R is the radius of any curve on the rail track.

The formula for train resistance is given equation (2-8) which is known as Davis Equation. Theoretical calculation of these coefficients is very difficult. These coefficients are usually determined by experimental tests [57].

$$F_{tr} = Am + Bvm + Cv^2 \quad (2-8)$$

In (2-8), A is the friction force coefficient per unit weight of the vehicle which is independent of speed, B is a constant per unit weight of the vehicle to give dynamic friction force proportional to speed, C is the air resistance coefficient when multiplied by the square of the speed gives the aerodynamic drag force acting on the vehicle. Figure 2.6 shows some illustrative examples of train resistances at the half and full load conditions [24].

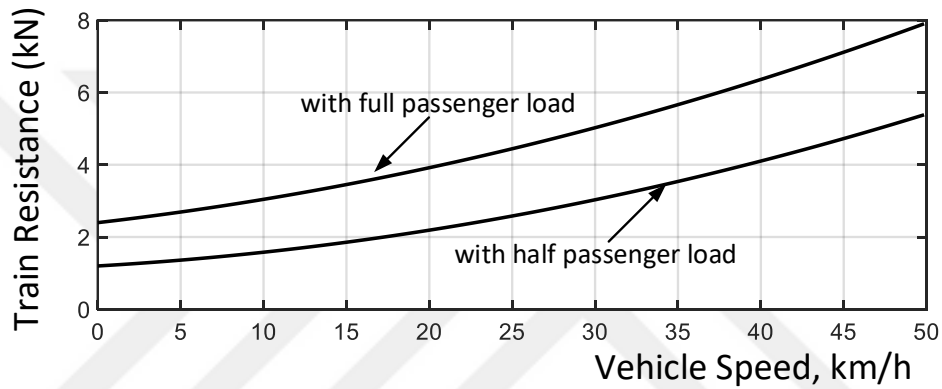


Figure 2.6. Typical train resistance of a light rail vehicle

On the other hand, F_{grv} gravitational force is as given in equation (2-9).

$$F_{grv} = mgsin(\theta) \quad (2-9)$$

A complete mathematical model of a public transportation vehicle running on a specific rail track is given in this section. The force–balance equation on the translational motion side is given in (2-10).

$$mF_{shaft} = M_{eqv} \frac{dv}{dt} + [F_r + F_{curve} + Mgsin[\alpha(x)]] \quad (2-10)$$

where, m is the number of traction motors, F_{shaft} the tractive force per motor reflected to the translational motion side by assuming rigid shafts, M_{eqv} the mass of static and rotating parts of transportation vehicle excluding motor inertias, v the

linear speed of the vehicle, F_r the sum of static and dynamic resistances and drag force acting on the vehicle, F_{curve} the curve resistance force, M the static mass of the vehicle, g the gravitational acceleration, and α the slope angle of the rail track as a function of displacement, x from a reference point. M_{eqv} in (2-10) has three components as given in (2-11).

$$M_{eqv} = M + J_{gb} \frac{1}{r^2} + J_w \frac{1}{r^2} \quad (2-11)$$

where, J_{gb} and J_w are respectively moment of inertia of all gear-boxes and wheels on the low rotational speed side. Equation (2-10) can be referred to the motor shaft by substituting $v = \omega r/n$ and dividing both sides by gear down ratio, n , and m . These yields (2-12)

$$\left[J_m + \left(\frac{M_{eqv} r^2}{m n^2} \right) \right] \frac{d\omega}{dt} = T_m - \frac{1}{m} \left[T_r + T_{curve} + \left(\frac{M g r}{n} \right) \sin[\alpha(x)] \right] \quad (2-12)$$

where, J_m is the motor inertia in kg-m², ω is the motor speed in mech-rad/s, T_m is the torque output of each traction motor in Nm, r the radius of the wheels in m, and T_r and T_{curve} are the load torque components owing to the train resistance and the curve resistance of the rail track, respectively.

T_r and T_{curve} are expressed as given in (2-13) and (2-14).

$$T_r = [(A + Bv) M + Cv^2] r/n \quad (2-13)$$

$$T_{curve} = (DMg/R(x)) r/n \quad (2-14)$$

3) Train Duty-Cycle and Electrification System

In railway applications, as mentioned before, there is not much free space on the bogie to integrate the traction motors. In addition, the component masses and sizes should be minimized in order to decrease mechanical losses and improve the system's overall efficiency. The development of an optimized traction system for a given application involves a complex, iterative process to decide the number of traction motors, motor size, inverter rated power, cooling system, etc. [58]. For this

reason, well-defined train duty-cycle profiles (e.g., how the train will be operated) are used to define traction system requirements.

Acting loads on the traction motors can be calculated by using duty-cycles, driving requirements (e.g., maximum acceleration, deceleration, and jerk), track and vehicle specifications (e.g., speed limits, track slopes and curves, train resistance, vehicle mass, and etc.). From there, the required number of motors and the gear ratio, torque, and speed requirements can be deduced. Figure 2.7 (a), Figure 2.7 (b), and Figure 2.8 show the required torque and speed characteristics in the time domain. The thermal behaviors of traction motors and traction drives are analyzed using these data.

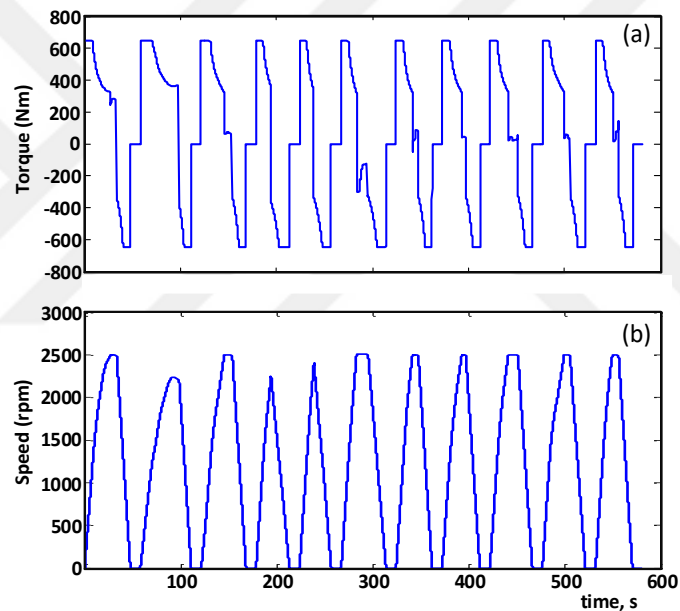


Figure 2.7. Torque/speed characteristics of traction motor in the time domain

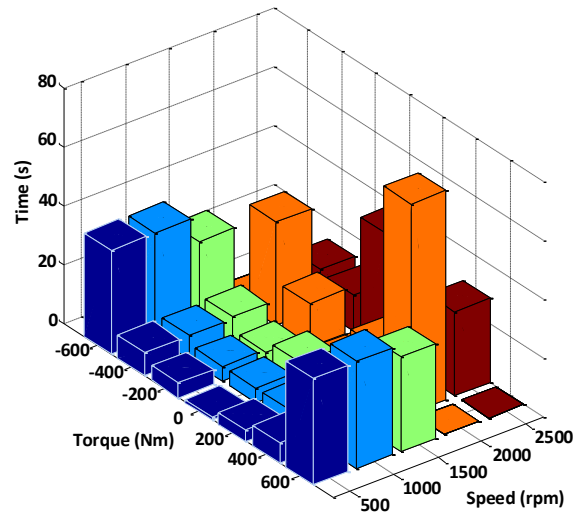


Figure 2.8. Load cycle in the time domain

Besides this, specifications of the electrification system represent other important criteria. Although most of the requirements of the electrification system are defined in the standards EN 50163 & EN 50388, railway operators sometimes specify special requirements such as voltage and current limits that differ from the standards. Therefore, all these requirements should also be taken as design constraints and used as input in drive duty-cycle analysis.

2.1.4 Power Quality Issues

Power quality has been a major issue of railway systems since the beginning of railway electrification. It has to be improved to satisfy the requirements for the safe and reliable operation of railway systems such as train control systems, onboard and off-board signaling systems, auxiliary supplies, and so on.

Modern railway traction systems use AC traction motors and Si-IGBT-based power electronics converters. Loads vary dynamically, and some inevitable physical events are taking place, such as arcs and switching harmonics due to the loose contact between pantographs and overhead line or shoes and third rail, and switching actions of semiconductor power devices.

Urban traction systems use DC electrification systems which mostly consist of 12-pulse diode rectifiers. These systems generate high amount of harmonics on the grid side. Especially, the orders of the highest current harmonics are 11th and 13th. In addition to this, the AC and DC fed traction systems also produce relatively high-frequency harmonics due to the switching operation of the power semiconductors.

In the literature, many studies have been conducted to improve the power quality of railway electric networks. Most of these studies are related to compensation strategies that are used to make power quality better at the electrification network side. Only a few of them deal with the traction load side improvements [59]. However, the presence of harmonic components created from traction systems can cause faulty train management by disturbing signaling and communication systems.

Recently used methods for load side harmonic reduction are improving the modulation methods of traction converters, use of passive filtering components, and/or multilevel power converter topologies. Especially as the switching frequency increases, the control capabilities of the power electronics converters also increase. However, because of the huge semiconductor losses, the switching frequency of the power electronics converters could not be increased. This deficiency was eliminated with the discovery of multilevel converters. Thus, differential and common noises are reduced. There are also some other benefits featured with the use of multilevel converters that decrease the stresses on transformer and motor winding insulations, semiconductor devices, and bearing currents.

On the other hand, SiC MOSFET technology, which is becoming widespread today, offers many benefits to increase capabilities of power electronics converters and make control of power converters easy and better by eliminating the need for some special modulation methods such as SHEPWM, SHMPWM, and OPWM. Switching frequencies of medium voltage and medium power converters can be increased 5 to 15 times as compared to the latest Si-IGBT-based power electronics converters. As a result of this, the control bandwidths and harmonics have been remarkably improved. As opposed to this, common-mode noises are increased due to the fast

switching transients. However, common-mode noises can also be improved by minimizing the parasitic capacitances in semiconductor packaging and power module design, using smaller filters as compared with the Si-IGBT-based power electronics converters.

2.2 Traction Motors used for Railway Traction

Electric motors have provided important contributions to the development of industrial systems up to the present. Nowadays, they are used in many applications such as transportation of goods and human beings (i.e., railway vehicle traction, electric vehicle traction, ship propulsion, cranes, elevators, etc.), washing machines, air conditioning systems, wind generators, pumps, fans, machining and so on. Torque/speed requirements differ according to the need of the application. In this study, only railway vehicle traction applications are presented. Other applications are beyond the scope of this research work.

A few critical requirements of traction motors are high power density, high efficiency, high reliability, robustness, maintenance-free operation, high torque at low speeds, high power at high speed, fast torque response, and operation in a wide speed range. Electric motors are categorized into two main groups, respectively, direct current (DC) and alternating current (AC). Figure 2.9 shows the classification of different types of electric motors available for traction applications [60].

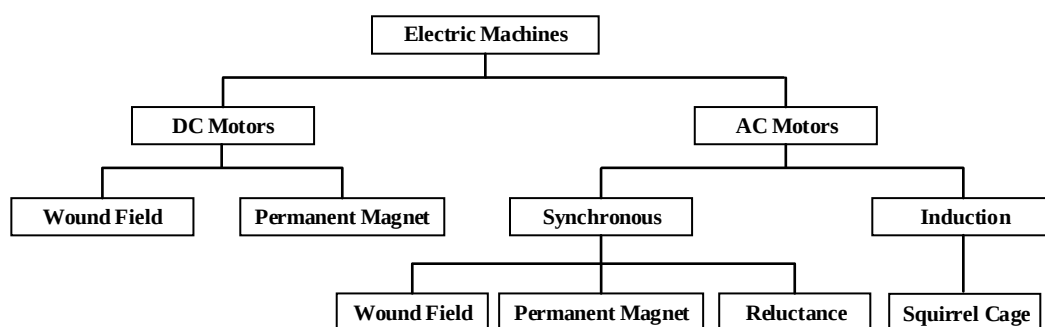


Figure 2.9. Electric motors used for traction applications

In the past, DC motors had been mostly used in railway electric traction systems due to their suitable torque-speed characteristics and their simple control [18], [58], [61]. However, DC motors have commutators and brushes. These structures require frequent maintenance. Besides this, the voltage rating and the speed of DC traction motors are limited because of the brushes. Moreover, power to volume and power to weight ratios are also low compared to AC motors. Unlike DC motors, AC motors have simple mechanical construction and require low maintenance tasks. However, the speed control of AC motors is not as simple as DC motors. AC motor control requires more complicated power electronic converters [60].

The first diesel-electric locomotive with inverter-fed squirrel cage traction motors was started to test by Brown Boveri in 1970 [62]. Since then, AC drives have been introduced into railway traction systems in most European countries, Japan, and North America. In addition to this, with the advances in power semiconductors, microprocessors, and control strategies, dc motors have left their place to ac motors due to their advantages.

Nowadays, induction motors are widely used for railway traction applications among all-electric motors due to their simple structure. Although permanent magnet synchronous machines (PMSM) are seen as the preferred type due to their superior advantages such as lightweight, reduced size, higher efficiency, and faster dynamic response, they are not widely utilized due to the safety reasons, magnet needs, and initial cost. During the phase to phase or phase to neutral short circuits on the motor side, PMSM produces uncontrolled high braking torque, and it requires the development of different protection concepts.

2.2.1 DC Traction Motor

The early traction systems used DC traction motors that were supplied from low voltage (e.g., 600 VDC) DC electrification systems. The advantage at that time was that DC motors provided easy control of the speed with simple control circuits made of high power resistors and rotary switches. However, the efficiency of such a

traction drive system was very low. Later, the efficiency of the DC traction motor drives was increased using modern DC-DC chopper converters. DC traction motors continued to be used extensively until serious improvements in power semiconductors and microprocessors were seen because of their simple speed control advantages.

In the past, series or separately excited DC motors were used for traction applications. The characteristics of both types are suitable for traction. However, the separately-excited DC motor has advantages comparing the series one, such as good characteristics in constant torque, constant power, and high-speed (weak field) operation regions. In order to control DC motor, armature winding current and field winding current should be controlled. The basic physical illustration of the DC motor windings and circuit schema of separately excited wound field DC motor are given in Figure 2.10 (a) and (b) [62].

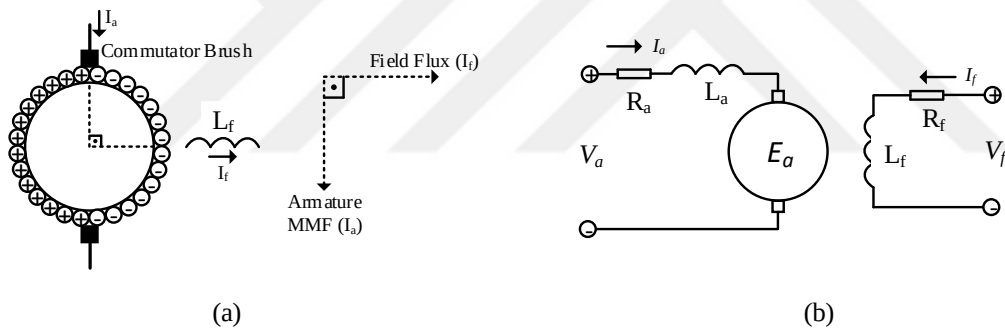


Figure 2.10. Illustration of a) Orientation of armature MMF and field flux in a DC machine b) Separately excited wound field DC motor circuit diagram

In DC motors, armature winding is located on the rotor, and field winding lies on the stator. The armature winding is fed with the help of the commutator and brushes. Because of the physical structure of DC motors, the armature current distribution is fixed in space regardless of rotor speed. Thus, the armature MMF and the field flux are physically guaranteed mutually perpendicular to each other. Moreover, in the separately excited DC motor, the armature current and field current can be controlled independently due to the inherent structure of the DC motor.

The electromagnetic behavior of the DC motor is described by (2-15) for back emf [62], [63].

$$E_a = \frac{p}{2} \lambda_f \omega_r \quad (2-15)$$

where, ω_r is mechanical rotor speed in radians per second, λ_f is magnetic flux linkage of armature winding, and p is the number of magnetic poles. By multiplying the induced voltage E_a with armature current I_a , electromechanical power, P_{em} is obtained. The equation of P_{em} is given in (2-16), and it includes the shaft power, core losses, friction, and windage losses.

$$P_{em} = E_a I_a \quad (2-16)$$

Electromagnetic torque (2-17) can be calculated by dividing the power by ω_r .

$$T_{em} = \frac{p}{2} \lambda_f I_a \quad (2-17)$$

The torque-speed curve of a separately excited DC motor is demonstrated in Figure 2.11.

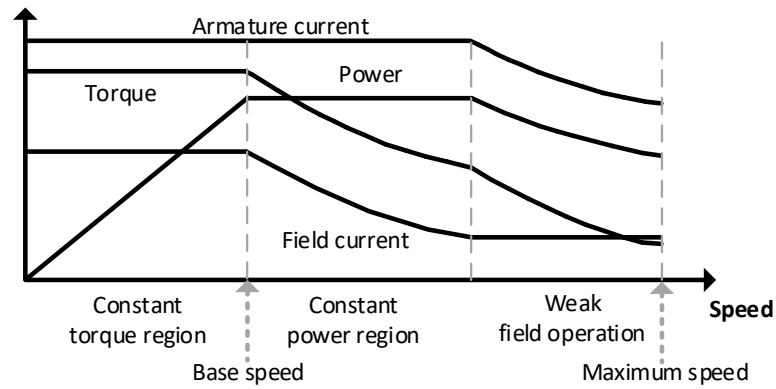


Figure 2.11. Torque-speed curve and operating regions of separately-excited DC motor

During the constant torque operation, armature current and field current should be kept constant. In the constant power region, the field current is decreased (e.g., the

first region of field weakening operation) proportionally with the increased shaft speed. In the last operating region (e.g., second field weakening region), the armature current is reduced while the field flux is kept almost constant at a weak level. In this region, the DC traction motor is operated at high speeds, and relatively high power is supplied to train to keep operating speed almost constant [62].

2.2.2 Induction Motor

Two different types of three-phase induction motors are used for industrial applications. These are called as squirrel cage and wound rotor induction motors. The wound-rotor induction motors have also slip rings. For this reason, instead of wound rotor induction motors, squirrel cage induction motors are preferred in traction applications. In this study, squirrel cage induction motors will be referred to as induction motors.

The induction motors have become more attractive with the development of microprocessor-controlled power converters since 1980. They provide the following advantages compared to DC motors [62].

- Quite high working speed & voltage because of no commutators and brushes,
- Higher robustness and reliability with a low maintenance requirement,
- High power density,
- Inherent regenerative braking capability,
- Steep torque-speed characteristics,
- Better electromagnetic compatibility due to the lack of brush-commutator structure.

As mentioned in previous sections, three-phase induction motors are the most mature technology among other traction motor options. The induction motor consists of phase windings on the stator and conductor bars short-circuited with end rings on the rotor. Induction motor stator windings are supplied with sinusoidal

balanced phase voltages at required magnitude and frequencies. The applied voltages create currents in the stator windings, and as a result of the balanced phase currents, a rotating magnetic field is obtained at the air gap. The rotating magnetic field induces voltages and so currents on the rotor bars. The created rotor currents interact with the rotating magnetic field, and as a result of this interaction, electromagnetic torque is created. Essentially, the induction motor resembles the transformers, which have rotating secondary windings.

The basic physical representation of the three-phase induction motor is sketched in Figure 2.12 [63].

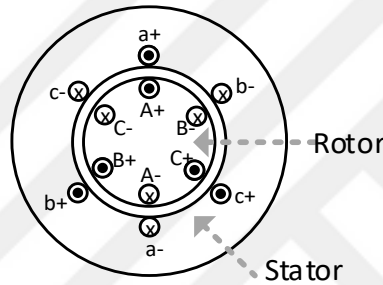


Figure 2.12. A physical representation of a three-phase induction machine

The speed of an induction motor is mainly determined by its supply voltage and frequency and shaft load. Figure 2.13 shows the per-phase induction motor equivalent circuit in which rotor parameters are transferred to the stator side.

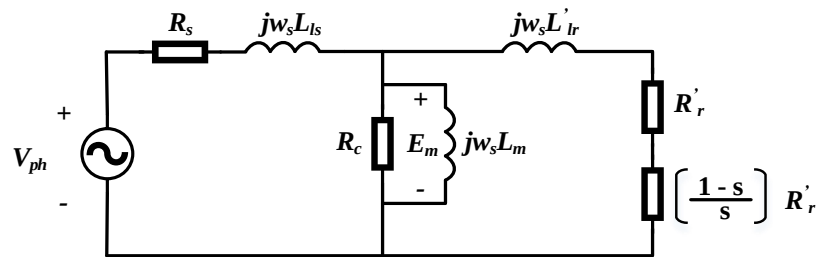


Figure 2.13. Per-phase equivalent circuit of induction machine

The input power P_{ph} three components as given in (2-18)

$$P_{ph} = P_{RS} + P_{ag} + P_c = 3V_{ph}I_{ph} \cos \emptyset \quad (2-18)$$

where, P_{Rs} is the stator copper loss, P_{ag} is the airgap power and P_c is the core loss component. P_{Rs} , P_{ag} and P_c are expressed as given in (2-19) – (2-21).

$$P_{Rs} = 3R_s I_{ph}^2 \quad (2-19)$$

$$P_{ag} = P_{Rr} + P_{em} = 3R'_r I_r'^2 + 3R'_r \left(\frac{1-s}{s}\right) I_r'^2 \quad (2-20)$$

$$P_c = \frac{E_m^2}{R_c} \quad (2-21)$$

where the first term of air gap power represents the rotor copper loss and $3R'_r \left(\frac{1-s}{s}\right) I_r'^2$ gives the electromechanical power, which includes friction and windage losses. Then, the electromechanical torque component of the motor can be written as (2-22) [63].

$$T_{em} = 3 \frac{p}{2} \lambda_r I_r' \quad (2-22)$$

Figure 2.14 shows the typical torque-speed characteristics of induction motor driven with variable voltage variable frequency (VVVF) drive.

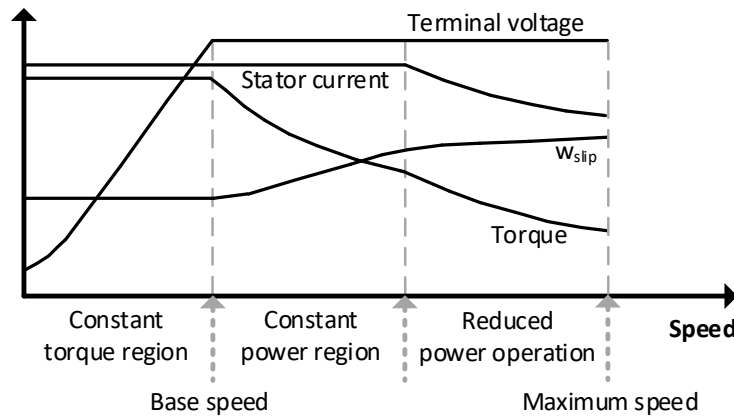


Figure 2.14. Typical torque-speed characteristics of an induction machine with VVVF drive control

The constant torque region, power region, and high-speed region (i.e., reduced power region) can be implemented by varying the stator voltage and frequency. Control methods of induction machines are introduced later in this chapter. In the constant torque region, flux and torque components of current are kept as constant. In the constant power region (field weakening region I), flux is decreased linearly with the increased speed to keep $T \cdot \text{Speed}$ product constant. In field weakening region II at high-speed operation, flux and torque components of current are further decreased with the increasing speed by considering the electromagnetic design and thermal limitations.

2.2.3 Permanent Magnet Synchronous Motor

The most distinct advantage of permanent magnet synchronous motor with respect to the induction motor and DC motor is the absence of rotor windings. This leads to improve efficiency and power density. The permanent magnets are placed on the rotor instead of windings. NeFeB or SmCo alloys are commonly used to construct rotor magnets. The inertia of the rotor is also reduced because of compact and small size rotor structures. By considering the installation of the permanent magnet materials, PMSMs are categorized into two groups, namely Surface Mounted Permanent Magnet Synchronous Motors (SMPMSM) and Interior Permanent Magnet Synchronous Motors (IPMSM). The rotor magnets are located on the rotor surface for SMPMSMs, while magnets are buried in the rotor for IPMSMs. Basic physical illustrations of PMSMs are given in Figure 2.15 (a) and (b) [63].

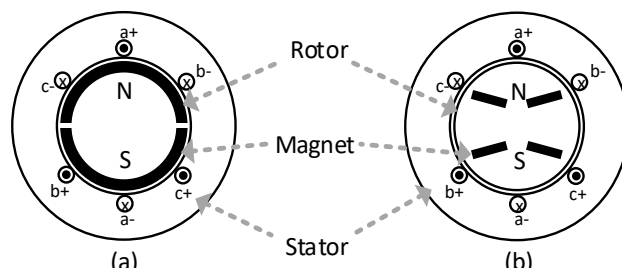


Figure 2.15. Basic physical representations of surface mounted and interior mounted permanent magnet motors

As seen from Figure 2.15, the three-phase windings are only placed on the stator for both types of permanent magnet motors. Because of this structure, less cooling requirement is needed to dissipate heat generated in the rotor. This gives a chance to the PMSM that can be fully sealed easily. This can further decrease the need for maintenance. Figure 2.16 shows the phase equivalent circuit of SMPMSM.

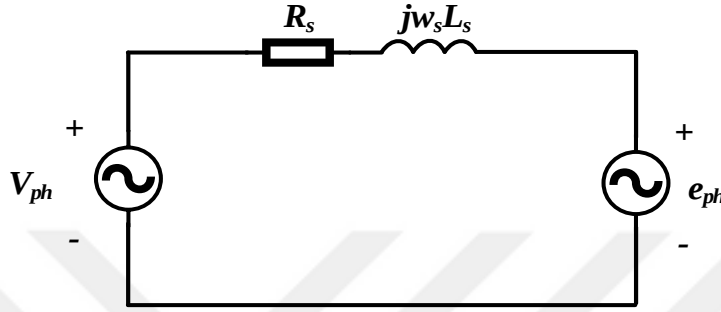


Figure 2.16. Equivalent circuit of a PM motor seen from stator phase

where L_s is the self-inductance, R_s is the stator resistance, e_{ph} is the back-emf. In PMSMs, field flux is created permanently by rotor magnets, and the power/torque is produced with the interaction of magnet flux linkage and the phase currents. The relationship between the back-emf, flux linkage, and rotor angular speed is described in (2-23) for SMPMSMs [63].

$$E_a = \omega_e \lambda_f = \frac{p}{2} \omega_r \lambda_f \quad (2-23)$$

where λ_f is the flux linkage, p is the number of magnetic poles, ω_r is the angular speed in radian per second, and ω_e is the electrical angular speed in radian per second. Referring the Figure 2.16, the power equation can be written as in (2-24).

$$P_{mech} = 3I_s E_a \cos \emptyset \quad (2-24)$$

By substituting (2-23) in (2-24) and then, by dividing the power equation by w_r , the output torque of SMPMSMs can be computed as in (2-25).

$$T_{mech} = 3 \frac{p}{2} I_s \lambda_f \cos \emptyset \quad (2-25)$$

Unlike SMPMSMs, additional torque can be obtained for IPMSMs because of the reluctance torque. In addition to this, IPMSMs have good overload capability and good efficiency values at high speeds compared to SMPMSMs. Moreover, IPMSMs have a more robust structure, and demagnetization of their magnets is difficult since the magnets are physically buried in the rotor. These make IPMSMs more suitable for high-speed applications. However, the price of IPMSMs is particularly higher than SMPMSMs [64].

2.2.4 Comparison of Traction Motors

Technological advances in power electronics have made different motor types more advantageous over time. From past to now on, DC motors, induction motors, and PMSMs have been used in traction applications. As mentioned in several sections, performance, cost, reliability, efficiency, maintenance, weight, size, and electromagnetic compatibility form the main considerations for traction applications. Because of easy control features, adequate performance and torque-speed characteristics, low price, and easy constructions, DC motors were used to some extent in the past. Later, rugged structure, high efficiency, low cost, and less maintenance features make induction motors more attractive with the advance of semiconductors and microprocessors. Nowadays, PMSMs are seen as one of the convenient motor types for traction applications. However, issues such as high price, production complexity, magnet needs, and safety-critical issues limit the spread of PMSMs. A comparative evaluation of major traction motor types is summarized in Table 2.2.

Table 2.2 The Comparison of Different Types of Traction Motors

Attribute	Motor Type			
	SMPMSM	IPMSM	IM	DC Motor
Efficiency	+++	+++	++	+
Size	+++	+++	++	+
Cost	++	++	+++	+++
Rotor Cooling	+++	+++	+	+
High-Speed Operation	++	+++	+++	+
Service and Maintenance	+++	+++	+++	+
Fault Tolerance	+	++	+++	+
Overload Capacity	++	+++	+++	++
Parallel Operation	+	+	+++	+++
Prevalence of New Railway Applications	+	+	+++	+

DC motors are not preferred anymore due to speed and voltage limitations, low efficiency, and high maintenance requirements. Among the other motor types, induction motors have superior fault tolerance capability, prevalence, and cost advantage. The efficiency is slightly lower, but it is still very close to PMSMs. The excitation can be easily removed for induction machines at any speed. This is preferred for rail vehicles manufacturers for safety reasons. However, in PMSMs, excitation cannot be removed due to the permanent magnets; induced voltage will create current. This current will continue to circulate in the winding as long as the motor is in motion. Therefore, it is not simply sufficient to turn-off the semiconductors because of the antiparallel diodes. It should be considered in the motor and drive system design. However, this makes the motor heavier, less efficient, and more expensive. On the other hand, a few precautions may be taken into account in the traction converter design, such as the use of circuit breakers or contactors between motor and inverter. Despite all these precautions, this problem may not be solved and may require additional actions if a permanent short circuit occurs at the motor side. In order to solve this, the gearbox may be decoupled at the failure instant. These also increase the cost of the drive system. This especially requires high price coupling between traction motor and gearbox. Moreover, the permanent magnet flux also creates an additional disadvantage in the coasting mode;

even the semiconductors and main input contactors are switched off, induced emf is generated. It will end up with an increase in DC-link voltage and may damage semiconductors and capacitors or decrease the lifetime of semiconductors and capacitors at high speed. In order to get rid of this problem, voltage ratings of components can be selected according to this high voltage level. However, this will end up with increased size, increased cost, and decreased efficiency. Moreover, in order to exit coasting mode and re-accelerate the vehicle, special control algorithms should be taken into account. DC-link voltage can be equalized by applying field weakening with a very small amount of negative torque. This is necessary to keep the DC-link voltage as stable while producing de-magnetizing current. As soon as the DC-link voltage is equal to the DC input voltage, the main line contactors can be closed. These solutions increase the control system complexity and the cost of the system [63]. On the other hand, PMSMs have higher power densities and efficiencies compared to DC motors and induction motors. Because of these features, SMPMSMs and IPMSMs are adopted for a few traction applications. Due to their low weight and size, PMSMs may provide certain advantages for low-floor tramway vehicles which have very limited space on their bogies.

Induction motors and IPMSMs have good overload capacities, while SMPMSMs have limitations for overload. Additionally, high-speed operation is more suitable for induction motors and IPMSMs due to the mechanical structures. Surface-mounted magnets create disadvantages for the high-speed operation of SMPMSM. In addition to this, induction machines are suitable for parallel driving, which is often preferred in rail transportation. All evaluations show that asynchronous motors are the most advantageous in existing system solutions [65].

2.3 Topological Overview of Two-Level Voltage Source Traction Inverters

The function of a voltage source inverter is to convert a DC voltage to an AC voltage waveform with variable magnitude and frequency. In Figure 2.17, the circuit diagram of a two-level voltage source inverter is given. Fully controllable six semiconductor

switches with antiparallel free-wheeling diodes composed the two-level voltage source inverter. Therefore, six gate-driver units and a minimum number of four isolated gate-driver power supplies are needed for this topology.

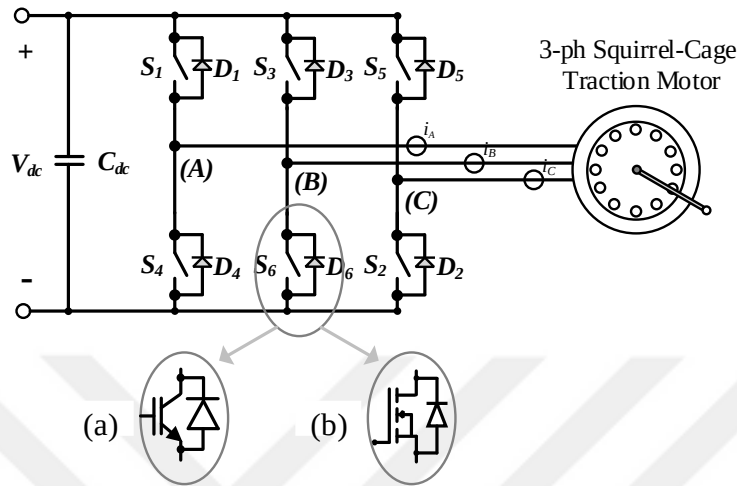


Figure 2.17. Two-level three-phase voltage source inverter

Two-level voltage source inverters have been used for three-phase motor drive applications from the appearance of gate-turn-off thyristors (GTOs). Since about 2000s, GTOs were superseded by IGBTs, and Si-IGBT-based simple and robust two-level inverter solutions have appeared as the classical solutions for motor drive applications. Today, for catenaries of 750 VDC, 1500 VDC, and 3000 VDC, IGBTs which have blocking voltages respectively, 1700 V, 3300 V or 4500 V and 6500 V are available [24].

From past to present, technological developments have had a serious role in the development of power converters. Today, silicon IGBTs are widely used for railway traction applications. Si-IGBT-based two-level voltage source inverters seem the simplest and common choice in railway applications such that the power and voltage requirements are around or less than 1.5 MVA and 2000 VDC [66]. Above these power and voltage ratings, Si-IGBT-based three-level inverters are becoming commonly used topology considering advantages such as low harmonic distortion, reduced voltage and thermal stresses on the individual power semiconductor, traction transformer, and traction motor, and increased efficiency.

Recently, SiC technology appears to have significant effects on topology selection by extending the limits of switching devices such as higher blocking voltage, higher operating temperature, and higher switching frequencies [47]. Moreover, many studies have been conducted to investigate the performances of SiC MOSFETs for future use in railway applications [4], [7], [8], [20], [47]. The theoretical works and experimental results show a significant reduction of semiconductor losses and confirm the possibility of increased switching frequency; the combination of these two advantages will have a great impact at the system level. Loss reduction in converters will lead to reductions in the cooling system size or even change its technology [21]. In addition to this, high-frequency operation at higher voltages and currents helps to use simple converter topologies instead of multilevel converters. At present, SiC MOSFETs with blocking voltages of 1700 V DC and 3300 V DC are available for railway traction applications. SiC technology can enable the use of two-level inverters instead of multilevel converters in a certain power and voltage range. This interpretation will be analyzed numerically in the following sections.

2.3.1 Operating Principles

In voltage source inverters, capacitors have been placed in dc-link as the energy storage element. Due to the sufficiently large dc-link capacitance, dc-link voltage V_{dc} stays nearly constant over a switching period. There are two major issues with motor control. One of the important issues is the current control method. The other issue is the modulation technique.

In fact, there are different control methods and modulation techniques. The aim of those is all similar. However, they differ from each other according to their performances and provides different benefits. The proper control of torque is mandatory to obtain needed acceleration and deceleration rates and to regulate the speed of the vehicle. The two most common control methods in railway traction applications are V/Hz (improved version) and field-oriented control (FOC). In operation, at least two of the three-phase currents of the traction motor and the DC-

link voltage are measured via current and voltage transducers to be used for motor control and overcurrent/overvoltage protection purposes. In addition to this, motor speed feedback is obtained with the help of encoders. The current controller takes the necessary reference and feedback signals such as speed and torque commands, line currents, voltages, speed of the motor shaft, and temperature and then generates the proper voltage reference signals for the modulator. Later, the modulator decides to correct switching states. Depending on the modulation technique, which will be discussed in detail later in the next section, fully controllable semiconductor switches are triggered according to the corresponding switching signals. There are eight possible switching states for two-level voltage source inverters. The eight possible switching states are listed in Table 2.3.

Table 2.3 All Allowable Switching Configurations for Two-level Three-phase Inverter

Switching Configuration	Switch States	Output line-to-line Voltages			Output Phase Voltage Vector		DC-link Current
	$\begin{bmatrix} S_1 & S_3 & S_5 \\ S_4 & S_6 & S_2 \end{bmatrix}$	V_{ab}	V_{bc}	V_{ca}	$ \bar{V}_o $	$\angle \bar{V}_o$	
1	$\begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 1 \end{bmatrix}$	V_{DC}	0	$-V_{DC}$	$\frac{2}{3}V_{DC}$	0	i_a
2	$\begin{bmatrix} 1 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix}$	0	V_{DC}	$-V_{DC}$	$\frac{2}{3}V_{DC}$	$\frac{\pi}{3}$	$-i_c$
3	$\begin{bmatrix} 0 & 1 & 0 \\ 1 & 0 & 1 \end{bmatrix}$	$-V_{DC}$	V_{DC}	0	$\frac{2}{3}V_{DC}$	$\frac{2\pi}{3}$	i_b
4	$\begin{bmatrix} 0 & 1 & 1 \\ 1 & 0 & 0 \end{bmatrix}$	$-V_{DC}$	0	V_{DC}	$\frac{2}{3}V_{DC}$	π	$-i_a$
5	$\begin{bmatrix} 0 & 0 & 1 \\ 1 & 1 & 0 \end{bmatrix}$	0	$-V_{DC}$	V_{DC}	$\frac{2}{3}V_{DC}$	$-\frac{2\pi}{3}$	i_c
6	$\begin{bmatrix} 1 & 0 & 1 \\ 0 & 1 & 0 \end{bmatrix}$	V_{DC}	$-V_{DC}$	0	$\frac{2}{3}V_{DC}$	$-\frac{\pi}{3}$	$-i_b$
7	$\begin{bmatrix} 1 & 1 & 1 \\ 0 & 0 & 0 \end{bmatrix}$	0	0	0	0	-	0
8	$\begin{bmatrix} 0 & 0 & 0 \\ 1 & 1 & 1 \end{bmatrix}$	0	0	0	0	-	0

2.3.2 Modulation of Two-level Three-phase Voltage Source Inverter

There are several applicable modulation methods to generate output voltage and current waveforms. In this section, the most common modulation technique, which is space vector pulse width modulation (SVPWM), is only illustrated. In the SVPWM technique, the modulation signal references $V_{An(ref)}$, $V_{Bn(ref)}$ and $V_{Cn(ref)}$ shown in Figure 2.18 are obtained by adding the optimal third harmonic zero-sequence signal to the sinusoidal reference signals. The construction of reference modulation signals and other PWM techniques will be given in the next section.

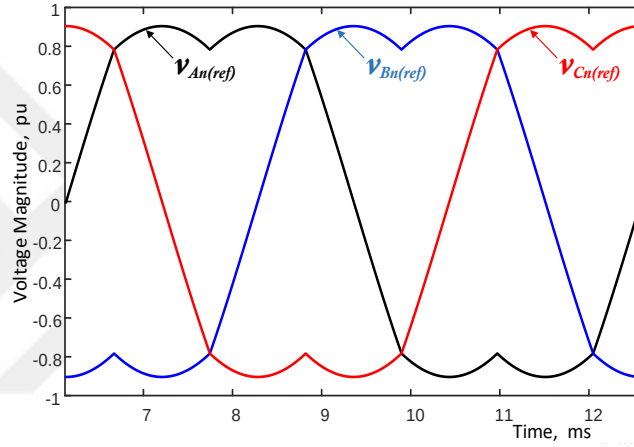


Figure 2.18. Modulation signal references for SVPWM

Vector and scalar implementations are the two different approaches for applying PWM techniques. In vector implementation, modulation references are transferred into the complex-valued space vector coordinate system in (2-26) and (2-27). This represents a vector with constant magnitude and rotating with a constant angular velocity in the plane.

$$\overline{V_o} = \frac{2}{3} (V_{An(ref)} + V_{Bn(ref)} e^{j\frac{2\pi}{3}} + V_{Cn(ref)} e^{j\frac{4\pi}{3}}) \quad (2-26)$$

$$\overline{V_o} = \frac{2}{3} (V_{An(ref)} + V_{Bn(ref)} a + V_{Cn(ref)} a^2) = v_o(t) e^{j\alpha_o(t)} \quad (2-27)$$

In Figure 2.19, six possible active vectors and related sectors are given. As illustrated in Figure 2.20, reference $\bar{V}_o$ can be realized by applying the two adjacent active vectors and a zero-vector for the calculated time durations.

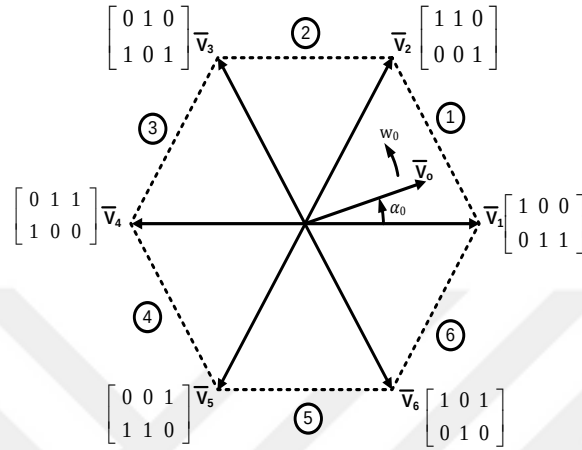


Figure 2.19. Active output voltage space vectors, related sectors, and the reference output phase voltage vector, $\bar{V}_o$ in complex plane

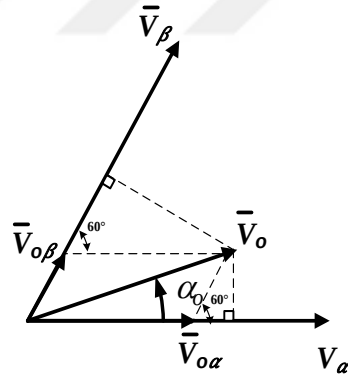


Figure 2.20. Reference vector realization

To synthesize rotating reference voltage vectors at different magnitudes and phases, it is necessary to compute duty ratios for each switching period. The reference phase voltage vector can be written in basic form using the adjacent active vectors;

$$\bar{V}_o = \bar{V}_{o\alpha} + \bar{V}_{o\beta} \quad (2-28)$$

with

$$\begin{aligned}\overline{V_{o\alpha}} &= \overline{V_\alpha} \delta_\alpha \\ \overline{V_{o\beta}} &= \overline{V_\beta} \delta_\beta\end{aligned}\tag{2-29}$$

Then, the duty ratios are;

$$\begin{aligned}\delta_\alpha &= \frac{|\overline{V_{o\alpha}}|}{|\overline{V_\alpha}|} \\ \delta_\beta &= \frac{|\overline{V_{o\beta}}|}{|\overline{V_\beta}|}\end{aligned}\tag{2-30}$$

$$\text{Also, } |\overline{V_{o\alpha}}| = \frac{|\overline{V_o}| \sin(\frac{\pi}{3} - \alpha_0)}{\sin(\frac{\pi}{3})}, |\overline{V_{o\beta}}| = \frac{|\overline{V_o}| \sin(\alpha_0)}{\sin(\frac{\pi}{3})}.$$

Hence, the duty ratios can be written as;

$$\delta_\alpha = \frac{|\overline{V_o}|}{\sqrt{\frac{3}{2}}|\overline{V_\alpha}|} \sin(\frac{\pi}{3} - \alpha_0), \delta_\beta = \frac{|\overline{V_o}|}{\sqrt{\frac{3}{2}}|\overline{V_\beta}|} \sin(\alpha_0)\tag{2-31}$$

$$|\overline{V_\alpha}| = |\overline{V_\beta}| = \frac{2}{3} |V_{DC}|, \frac{\sqrt{3}|\overline{V_o}|}{|V_{DC}|} = m_i \text{ and } 0 \leq m_i \leq 1$$

where m_i is the voltage modulation index, which represents the input to output voltage transfer ratio, and V_{DC} is the average value of the DC-link voltage. Thus, the duty ratios of the active vectors can be written as;

$$\begin{aligned}\delta_\alpha &= m_i \sin\left(\frac{\pi}{3} - \alpha_0\right) = \frac{T_\alpha}{T_s} \\ \delta_\beta &= m_i \sin(\alpha_0) = \frac{T_\beta}{T_s}\end{aligned}\tag{2-32}$$

After calculating the duty ratios d_α and d_β , the duty ratio of the zero-phase voltage vector can be written as follow;

$$\delta_0 = 1 - \delta_\alpha - \delta_\beta = \frac{T_{v0}}{T_s} \quad (2-33)$$

The vectors and their duty cycles are determined according to the magnitude and angle of the reference $\overline{V_0}$. The vector sequences for all plane is given in Table 2.4. The vector approach provides flexibility to change the sequence of switching states. In some cases, it will be an important feature for switching loss optimization and even harmonic cancellation [67].

Table 2.4 Switching Table for SVPWM of Two-level Inverter

	$\frac{1}{4}\delta_0$	$\frac{1}{2}\delta_\alpha$	$\frac{1}{2}\delta_\beta$	$\frac{1}{2}\delta_0$	$\frac{1}{2}\delta_\beta$	$\frac{1}{2}\delta_\alpha$	$\frac{1}{4}\delta_0$
Sector I	V ₀	V ₁	V ₂	V ₇	V ₂	V ₁	V ₀
Sector II	V ₇	V ₂	V ₃	V ₀	V ₃	V ₂	V ₇
Sector III	V ₀	V ₃	V ₄	V ₇	V ₄	V ₃	V ₀
Sector IV	V ₇	V ₄	V ₅	V ₀	V ₅	V ₄	V ₇
Sector V	V ₀	V ₅	V ₆	V ₇	V ₆	V ₅	V ₀
Sector VI	V ₇	V ₆	V ₁	V ₀	V ₁	V ₆	V ₇

In the scalar implementation, modulation reference signals are compared with a carrier signal. Then, the switching signals of semiconductors can be obtained by comparing this carrier signal and modulation references. When the magnitude of modulation signals is higher than the carrier wave, the related upper switches are turned-on. The lower switches operate in a complementary manner, and it is not allowed to turned-on at the same instant to avoid short circuit faults. While making the commutations or during switch state transitions, a dead-time should be added between the transition of upper and lower switches considering the switch-on and switch-off times of active switches. During this time interval, both active switches are switched-off, and the load currents continue to flow through the antiparallel free-wheeling diodes.

The modulation reference signal of each leg, corresponding phase voltages and currents are identical in shape; however, 120° phase shift was inserted between the

adjacent ones. Typical normalized forms of phase voltages (b) - (d) and line-to-line voltage (e) are shown in Figure 2.21.

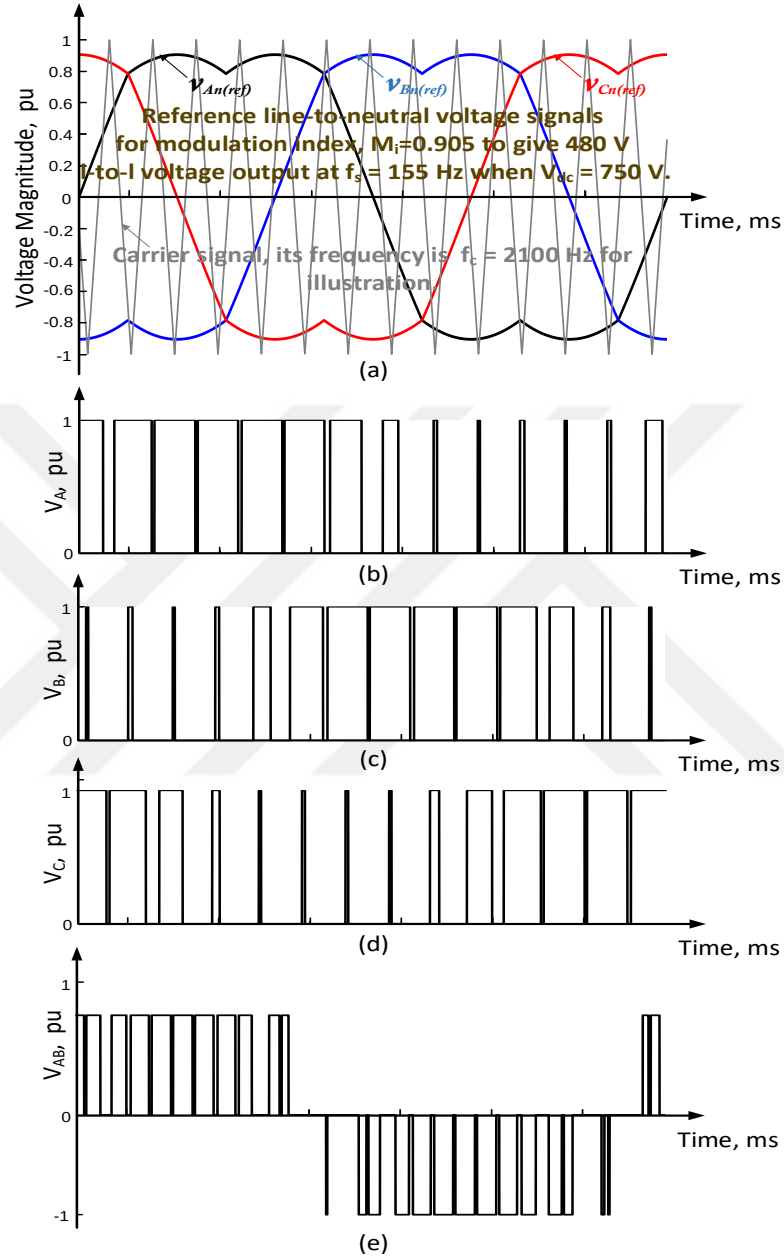


Figure 2.21. Simulated waveforms for SVPWM two-level three-phase inverter (a) Modulation references and carrier signal (b)-(d) Phase voltages (e) Line-to-line voltage

Since the peak values of voltage pulses are theoretically DC-link voltage, magnitudes in the Figure 2.21 and harmonic spectra of line-to-line voltage in Figure 2.22 (a) are given as normalized form with respect to V_{dc} . In Figure 2.22 (b) and (c), the three-phase line currents and THD of a line current are illustrated.

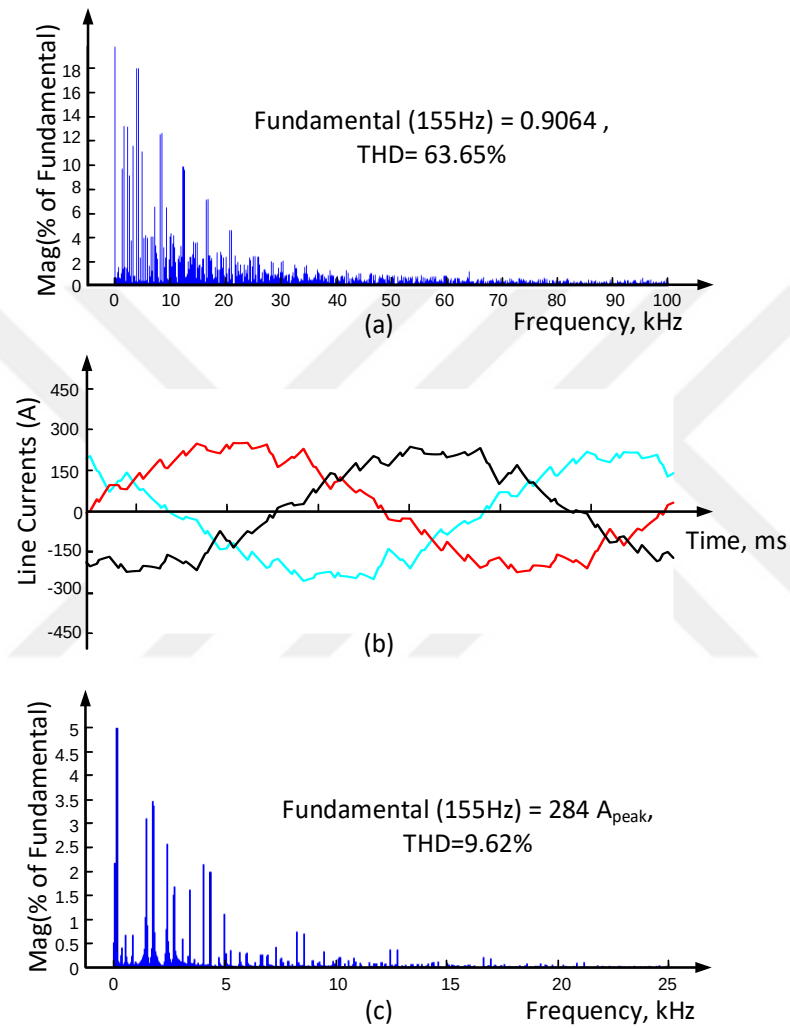


Figure 2.22. Simulated waveforms for SVPWM two-level three-phase inverter (a) Normalized harmonic spectra and THD of line-to-line voltage waveform (b) Three-phase line currents (c) THD of line current

Figure 2.23 (a) presents a typical vector sequence in Sector I. In addition to this, the current conduction paths of the two-level three-phase inverter during the regenerative braking are illustrated in Figure 2.23 (b).

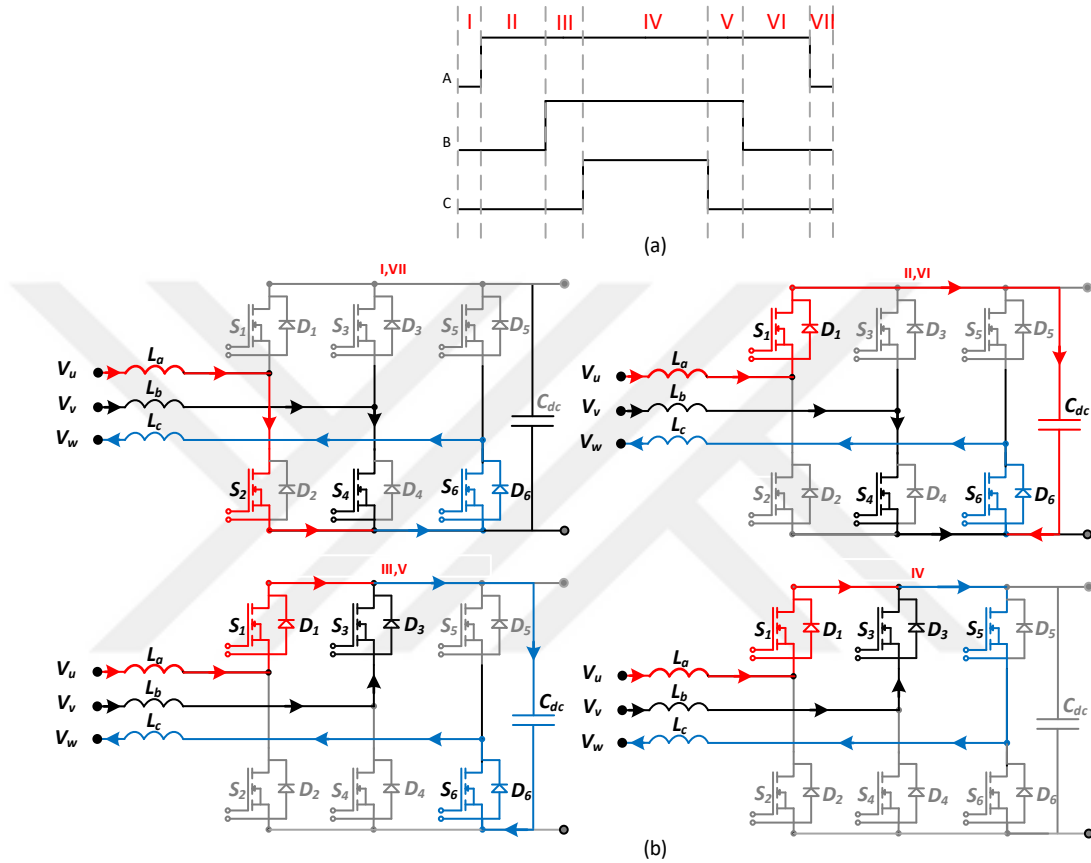


Figure 2.23. Typical current conduction paths during regenerative braking instant

2.4 Topological Overviews of Three-Level Neutral Point Clamp (NPC) and T-Type Voltage Source Traction Inverters

As mentioned earlier, Si-IGBT has been the most widely used semiconductor technology in railway traction applications for 20 years. However, the switching frequencies have been kept relatively low due to the high switching losses of IGBTs for medium power and medium voltage applications. This situation will result in high amount of harmonics at relatively low frequencies hence decreases the overall

system efficiency and increases stability problems. In addition to this, pulse width modulated voltage source inverters produce phase voltages whose instantaneous sum is not zero, and this will end up with common-mode voltages at the motor winding neutral point. Moreover, more stray currents occur due to the high dv/dt rates, and these yield undesired high-frequency common-mode currents on the motor bearings.

These two problems can be solved with the use of passive filters. However, relatively low switching frequencies made these passive filters very bulky. Because of this, passive filter solution was not well accepted in practical motor drive applications, which require high power densities and bandwidths.

Considering the defined problems above, the following benefits were handled with the multilevel converters [26], [27], [68], [69]:

- i- THD and the dv/dt rate are lowered because of the increased voltage steps.
- ii- Common mode voltages are reduced due to the increased voltage steps.
- iii- Less electromagnetic interference due to lower differential and common mode noises.

Another reason to apply multilevel inverters was the insufficient blocking voltages of switching devices. By increasing the voltage steps, voltage stress across the switching devices and switching losses are decreased. Besides this, better voltage and current waveforms are handled at lower switching frequencies with a multilevel voltage source inverter. This may lead to higher overall efficiencies in medium power and medium voltage applications even if conduction losses are increased due to the increased number of serially connected switching devices. As opposed to these, the increased number of switching devices, their gate drivers, and dc-link capacitors increase the system complexity and the cost of the systems. Moreover, the reliability of the system decreases due to the increased number of components. Voltage imbalance problem arises, and this topology requires the control of mid-point voltage [26]. Although the low voltage semiconductor usage is possible, each semiconductor requires a separate isolated gate-drive circuit; hence isolated power

supply needs are increased. This also makes multilevel converters more expensive and complex. Additionally, difficulty in designing simple and fast modulation techniques creates another difficulty [27]. In railway applications, diode-clamp voltage source inverter topology is generally used where the voltage and power ratings are above 2000 V and 1.5 MVA consecutively.

The diode-clamp multilevel inverter consists of clamping diodes, series-connected capacitors, and semiconductors to produce multiple levels of AC voltage waveforms. This inverter topology can also be configured as more than three-level such as four, five, and so on. However, only the three-level inverter configuration has found wide application in medium power and medium voltage drives [67]. In Figure 2.24, a circuit diagram of a three-level NPC inverter is given. The main characteristics of multilevel inverters are lower dv/dt rates and harmonic contents of output voltage waveforms compared to two-level inverters. Another advantageous aspect for motor drive applications is the better control bandwidths achieved due to the higher effective switching frequencies.

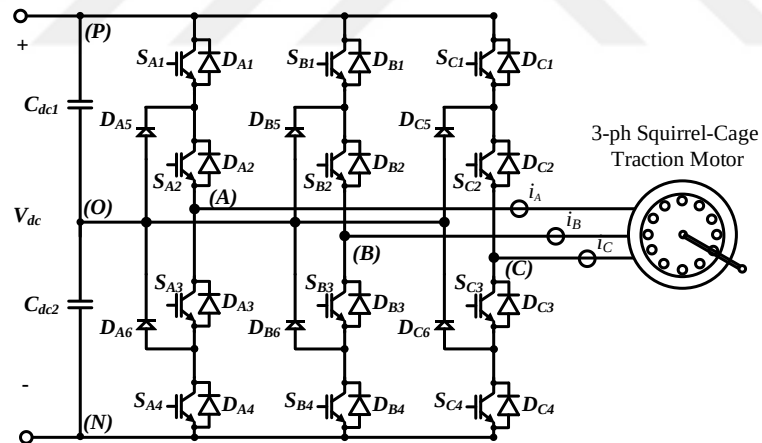


Figure 2.24. Three-level three-phase NPC voltage source inverter

The inverter is composed of four groups of power semiconductors with antiparallel diodes and two clamping diodes for each leg. There are twelve fully controllable power semiconductors hence twelve gate-driver units. Besides, a minimum number of ten isolated gate-driver power supplies are needed, and six clamping diodes are used in this topology.

In addition to this, T-type three-level voltage source inverters can be an alternative for railway traction applications. As we discussed, most of the medium voltage medium power motor drive applications require high efficiency, high reliability, small size, and less weight. Above a power limit, multilevel inverters are preferred compared to two-level inverters because of the low harmonic distortion, high efficiency, and providing a solution for high power as well as high voltage levels. However, by using the multilevel inverters, the number of components also increases. The common drawback of the different multilevel inverters is the increased complexity due to the higher number of switching elements. In order to decrease component counts for reduced complexity and cost, different types of multilevel inverter topologies were developed. The T-type multilevel inverter is one of the advanced neutral point clamp inverter topologies that have the advantages of a lower number of components [70].

The three-level T-type inverter is shown in Figure 2.25. It is composed of a two-level inverter connected to three anti-series branches of semiconductor switches. This inverter topology can also be configured as more than three-level such as five, seven, and so on.

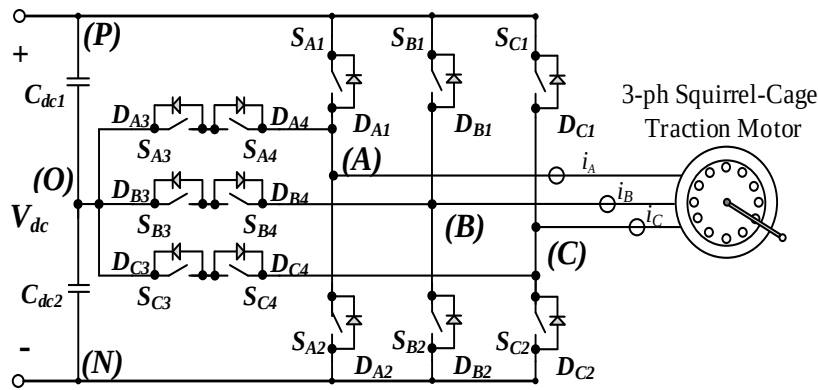


Figure 2.25. Three-level T-type three-phase voltage source inverter

For the three-level T-type three-phase inverters, there are twelve fully controllable power semiconductors hence twelve gate-driver units. Besides, a minimum number of seven isolated gate-driver power supplies are needed.

2.4.1 Operating Principles of Three-level NPC and T-Type Voltage Source Traction Inverters

The DC-link capacitor is split into two series-connected capacitors for a three-level NPC inverter. The voltage of each capacitor is equal to $V_{dc}/2$. When the S_{A1} and S_{A2} are turned-on, the output terminal (A) is connected to the most positive rail of the DC bus, which is denoted as (P). When S_{A2} and S_{A3} are turned-on, the terminal of (A) is connected to the mid-point of DC-link (O), which has a voltage value of $V_{dc}/2$. If S_{A3} and S_{A4} are turned-on, the terminal of (A) is connected to the negative bus, which is denoted as (N). Similarly, the same switching states create the same results for other legs. By considering the voltage levels of all three phases, the inverter has a total of twenty-seven possible switching states, as listed in Table 2.5. The voltage waveforms of V_{AO} , V_{BO} , and V_{CO} have three levels, $+V_{dc}/2$, 0, and $-V_{dc}/2$.

Similarly, the DC-link capacitor is split into two series-connected capacitors for a three-level T-type inverter. The voltage of each capacitor is equal to $V_{dc}/2$. In order to achieve the three different voltage levels with respect to the midpoint, high side semiconductor switch, low side semiconductor switch, or semiconductor switches of anti-series branches are controlled. The voltage levels $+V_{dc}/2$, $-V_{dc}/2$, and 0 are obtained, respectively. In fact, the turned-on and turned-off switches of NPC and T-type inverters are different; however, the voltage vectors are the same. Considering the operation of all phases, a common list for switching states of NPC and T-type three-level three-phase can be given in Table 2.5. P-type vectors represent the small voltage vectors that are obtained using the upper half of the DC-link capacitors, while the N-type small vectors use the lower half of the DC-link.

Table 2.5 Voltage Vectors and The Switching States

Vectors		Switching State		Vector Classification	Vector Magnitude
V ₀		[PPP] [OOO] [NNN]		Zero vectors	0
		P-type	N-type	Small vector	$\frac{1}{3}V_{DC}$
V ₁	V _{1P}	[POO]			
	V _{1N}		[ONN]		
V ₂	V _{2P}	[PPO]			
	V _{2N}		[OON]		
V ₃	V _{3P}	[OPO]			
	V _{3N}		[NON]		
V ₄	V _{4P}	[OPP]			
	V _{4N}		[NOO]		
V ₅	V _{5P}	[OOP]			
	V _{5N}		[NNO]		
V ₆	V _{6P}	[POP]			
	V _{6N}		[ONO]		
V ₇		[PON]		Medium vector	$\frac{\sqrt{3}}{3}V_{DC}$
V ₈		[OPN]			
V ₉		[NPO]			
V ₁₀		[NOP]			
V ₁₁		[ONP]			
V ₁₂		[PNO]			
V ₁₃		[PNN]		Large vector	$\frac{2}{3}V_{DC}$
V ₁₄		[PPN]			
V ₁₅		[NPN]			
V ₁₆		[NPP]			
V ₁₇		[NNP]			
V ₁₈		[PNP]			

The switching states are represented by the letters in square brackets. Based on the magnitudes of the applied vectors, switching states can be divided into four groups as zero vectors, small vectors, medium vectors, and large vectors.

2.4.2 Modulation of Three-level NPC and T-type Three-phase Voltage Source Inverters

Various modulation techniques can be applicable for three-level NPC and T-type inverters as similarly described for the two-level inverters. In this section, conventional SVPWM techniques are only presented. The T-type inverter is another form of NPC topology. It provides the advantages of less component counts. The applicable voltage vectors are the same for both topologies. Because of this, the modulation process is explained together. T-type converters have control flexibility due to the difference in the circuit structure. Similar to the two-level inverter, scalar and vector implementations are possible. However, modulation strategies for multilevel inverters are more complex than the two-level inverters. In order to solve the DC-link voltage balancing problem of NPC and T-type inverters, sequences of switching states have to be reorganized. Because of this reason, scalar implementation of the modulation techniques in multilevel NPC and T-type inverters is not preferred. In the vector approach, similar to the two-level inverters, the reference vectors for the three-level NPC inverter are constructed based on the “volt-second balancing” principle. The divisions of sectors, regions, and applicable space vectors are illustrated in Figure 2.26.

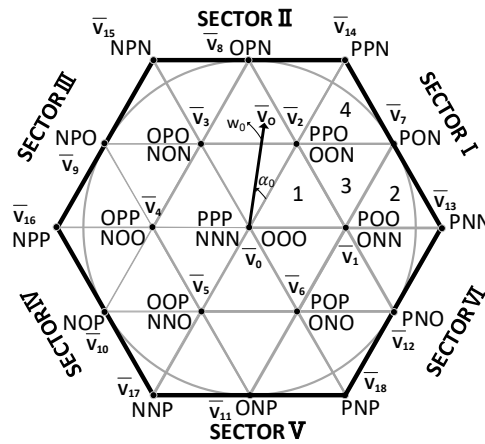


Figure 2.26. Sectors and possible vectors

In Figure 2.27, a reference vector together with the nearest three active vectors are presented. Reference voltage vector, $\bar{V}_0$ can be synthesized by multiplying three

nearest space vectors $\bar{V}_2$, $\bar{V}_3$ and $\bar{V}_8$ with corresponding duty cycles and then summing these results as shown in (2-34) and (2-35). The required duty cycles for these vectors can be calculated using equation (2-36). To establish generalized expressions for the duty-cycles for different space vectors which are located in different sector areas, mode of the angle with respect to $\pi/3$ is taken. In this way, all duty cycles can be calculated similarly for other sectors and regions [70]. All duty cycles and corresponding space vectors in Sector II are given in Table 2.6.

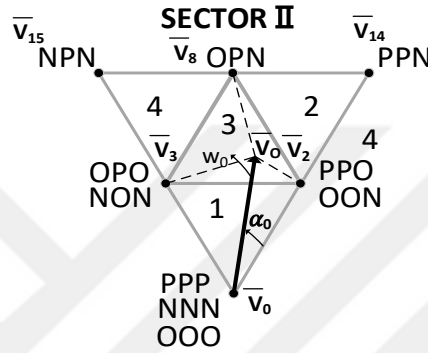


Figure 2.27. Reference vector and corresponding space vectors

$$\bar{V}_2\delta_\alpha + \bar{V}_8\delta_\beta + \bar{V}_3\delta_\theta = \bar{V}_o \quad (2-34)$$

$$\delta_\alpha + \delta_\beta + \delta_\theta = 1 \quad (2-35)$$

$$\begin{bmatrix} |\bar{V}_o| \cos \alpha_0 \\ |\bar{V}_o| \sin \alpha_0 \\ 1 \end{bmatrix} = \begin{bmatrix} Re(\bar{V}_2) & Re(\bar{V}_8) & Re(\bar{V}_3) \\ Im(\bar{V}_2) & Im(\bar{V}_8) & Im(\bar{V}_3) \\ 1 & 1 & 1 \end{bmatrix} \begin{bmatrix} \delta_\alpha \\ \delta_\beta \\ \delta_\theta \end{bmatrix} \quad (2-36)$$

Table 2.6 Duty Cycle Calculation in Sector II

Region	V	δ_α	V	δ_β	V	δ_θ
1	V_2	$2m_a \sin(\frac{\pi}{3} - \alpha_0)$	V_0	$1 - 2m_a \sin(\frac{\pi}{3} + \alpha_0)$	V_3	$2m_a \sin(\alpha_0)$
2	V_2	$2 - 2m_a \sin(\frac{\pi}{3} + \alpha_0)$	V_8	$2m_a \sin(\alpha_0)$	V_{14}	$2m_a \sin(\frac{\pi}{3} - \alpha_0) - 1$
3	V_2	$1 - 2m_a \sin(\alpha_0)$	V_8	$2m_a \sin(\frac{\pi}{3} + \alpha_0) - 1$	V_3	$1 - 2m_a \sin(\frac{\pi}{3} - \alpha_0)$
4	V_{15}	$2m_a \sin(\alpha_0) - 1$	V_8	$2m_a \sin(\frac{\pi}{3} - \alpha_0)$	V_3	$2 - 2m_a \sin(\frac{\pi}{3} + \alpha_0)$

Typical SVPWM reference waveforms (a), normalized form of phase voltages (b) - (d), and line-to-line voltage (e) are shown in Figure 2.28.

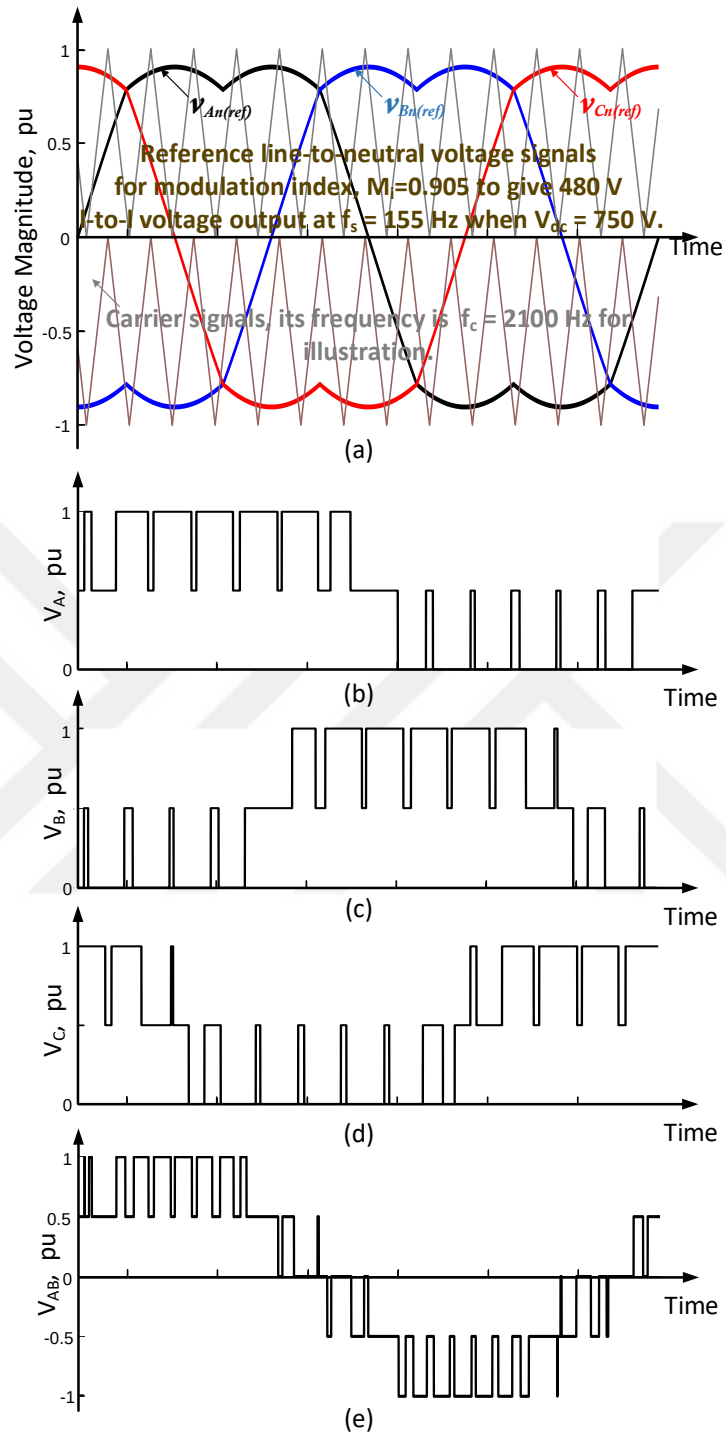


Figure 2.28. SVPWM modulated three-level voltage source inverter at $m_i = 0.905$ and $f_{sw} = 2100$ Hz (a) Modulation reference waveforms (b) - (d) Normalized form of phase voltages (e) Line-to-line voltage

In Figure 2.29 (a), harmonic spectra of line-to-line voltage waveform is given in the normalized form with respect to V_{dc} . In Figure 2.29 (b) and (c), the three-phase line currents and THD of the line current are illustrated.

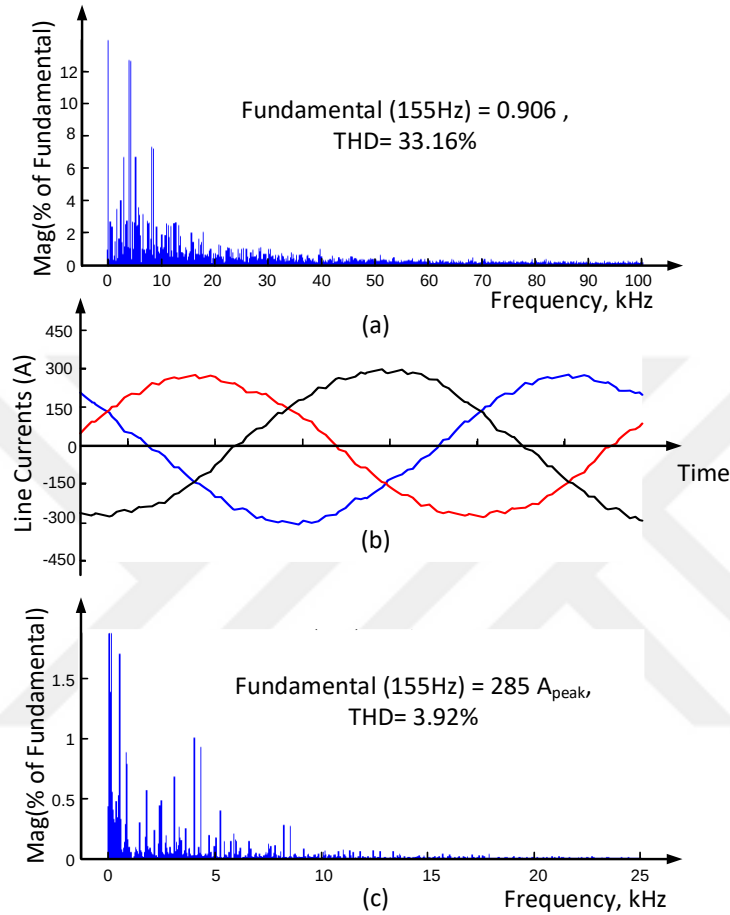


Figure 2.29. SVPWM modulated three-level voltage source inverter at $m_i = 0.905$ and $f_{sw} = 2100$ Hz (a) Harmonic spectra of line-to-line voltage waveform (b) – (c) The three-phase line currents and THD of the line current

Figure 2.30 presents a typical vector sequence in Sector II [71], [72].

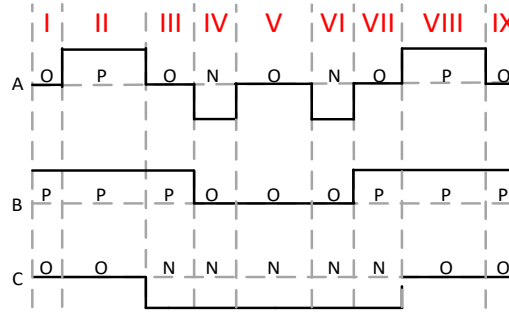


Figure 2.30. A typical switching pattern for sector II and sub-region 3

The current conduction paths of the three-level three-phase NPC inverter are illustrated in Figure 2.31 during regenerative braking for a typical switching pattern in sector II and sub-region 3.

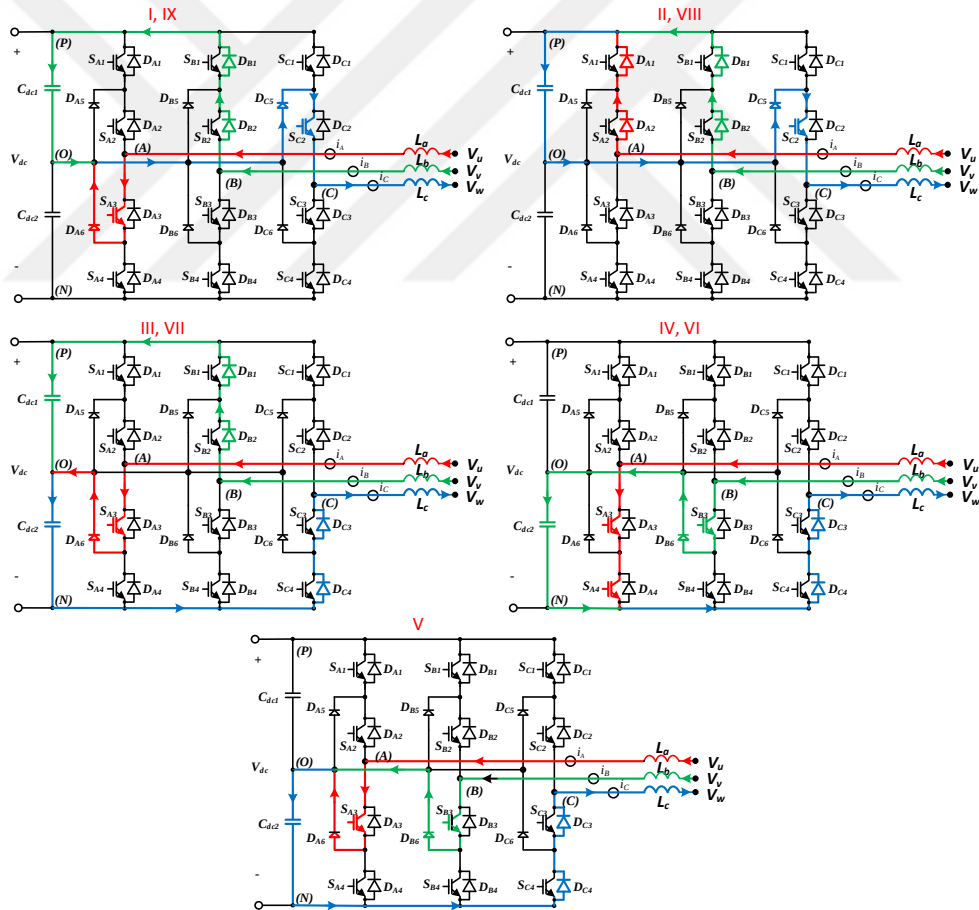


Figure 2.31. Current conduction paths of three-level three-phase NPC converter for a typical pattern

The current conduction paths of the three-level three-phase T-type inverter are illustrated in Figure 2.32 during regenerative braking for a typical switching pattern in Sector II and sub-region 3.

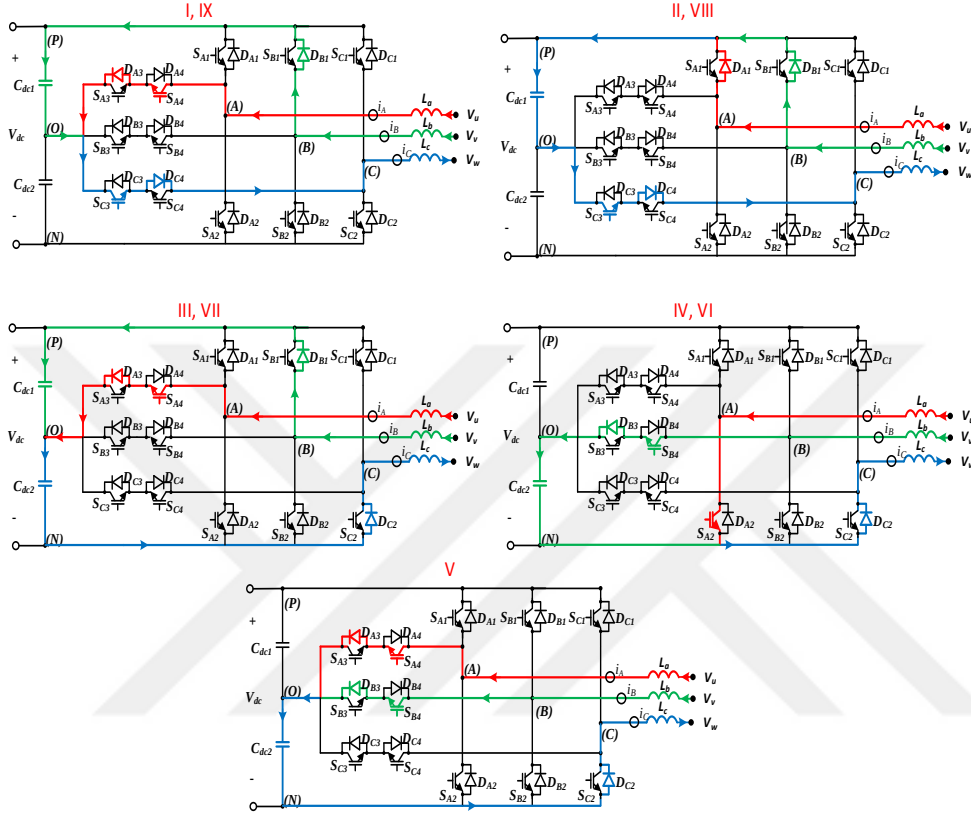


Figure 2.32. Current conduction paths of three-level three-phase T-type converter for a typical pattern

2.5 Comparative Evaluation of Si-IGBT based Three-Level, and SiC MOSFET based Two-Level Voltage Source Traction Converters

In Table 2.7, specific attributes of the two-level, three-level NPC, and three-level T-type voltage source inverters are tabulated. Today, the breakdown voltages of available power semiconductors allow using of two-level voltage source inverters alone in many applications. Semiconductors with a wide voltage range that offer solutions between 1700 V and 6500 V in rail transportation traction drives make this

situation similar. As can be seen from the table below, if we do not take into account the stresses due to power loss, the two-level solution seems to offer a significant advantage in terms of MTBF. However, this is not the case in reality. The switching losses have greatly limited the switching frequency of the two-level topology and severely reduced the power quality. In fact, in order to solve this restriction and continue with the choice of two-level inverter topology, special modulation techniques such as Selective Harmonic Elimination Pulse Width Modulation (SHEPWM), Selective Harmonic Mitigation Pulse Width Modulation (SHMPWM), and Optimal Pulse Width Modulation (OPWM) are used to decrease switching losses. They will be explained in more detail in the next section. This solution may require lots of offline computations to find the optimum angles of the pulses for different modulation index. If anyone wants to do all those by online calculations, more powerful graphical processing units are needed. In addition, this solution also requires modification in the motor control methods (e.g., switching from vector control to scalar control methods). Because of this, multilevel converters have to be used in medium voltage and medium power railway traction applications.

From now on, the reduced switching losses together with higher blocking voltage capability make it possible to use simpler SiC-based two-level voltage source inverters instead of multilevel inverters. In other words, by using SiC MOSFETs, the switching frequency can be increased by a considerable amount because of the inherently reduced switching losses. Thus, harmonic distortions on the voltage and current waveforms can be seriously reduced without the need for multilevel converters for a certain power and voltage level. In addition, the conduction losses are also decreased a considerable amount because of the number of serially used semiconductor devices in multilevel converters. Besides these, the reliability of the converter is also increased as a result of the decrease in the number of serially connected semiconductor devices and the higher operating temperature. In fact, these improvements will not enable two-level inverters to be used in all power and voltage ranges. However, it will greatly expand the power and voltage levels, causing the topology changes between two-level to three-level inverters. The exact border of this

limit could not be calculated in this study, as there are not yet SiC-based power MOSFETs that are becoming very common.

Table 2.7 Three-phase Voltage Source Inverter Topologies

Three-phase Inverter Topologies				
Attribute	Si-IGBT based			SiC-based
	Two-level	Three-level NPC	Three-level T-type	Two-level
No. of switches	6	12	12	6
No. of diodes	0	0	6	0
Min. num. of isolated DC power supply	4	10	7	4
No. of gate-driver units	6	12	12	6
Max. switch voltage	V _{dc}	V _{dc} /2	V _{dc} & V _{dc} /2	V _{dc}
Max. diode voltage	0	V _{dc} /2	0	0
Modularity	Good	Poor	Moderate	Good
Modulation	Moderate	Complex	Complex	Simple
Conduction loss	Lower	Higher	Moderate	Lower
Switching losses @ same effective switching frequency	Higher	Lower	Moderate	Lower
Switch temperature distribution	Homogeneous	Not homogeneous	Not homogeneous	Homogenous
Power quality @ same switching frequency	Lower	Higher	Higher	Higher
dv/dt	Moderate	Lower	Lower	Higher

However, with a new SiC MOSFET module suitable for high power and voltage applications put on the market by Mitsubishi, it has been revealed that the field of two-level inverter application will expand in terms of voltage and power.

In general, following main advantages can be specified for SiC-based two-level voltage source converters comparing conventional Si-IGBT based three-level voltage source converters:

- Low conduction losses (due to the decrease in the number of semiconductor devices)

- Low switching losses (e.g., an inherent feature of SiC devices, higher switching frequencies, and smaller reactive components)
- Higher operating temperatures (e.g., less cooling requirements)
- Higher blocking voltages
- Higher power densities
- Low complexity
- High reliability

However, as opposed to these, the stresses on insulation materials, especially the motor winding insulations, are increased due to the high dv/dt rates. This situation may increase the stray currents flowing through the winding insulation materials to the chassis and may shorten the life of insulation materials. Moreover, this also increases the bearing currents and shaft voltage. In general, bearing currents and shaft voltage can cause accelerated degradation of the motor. Therefore, stronger insulation material (e.g., Kapton, isolated bearings, ground brushes for rotor shaft) should be preferred for such applications.

2.6 Voltage and Current Waveform Synthesizing

The developments in fully controllable power switches and high-speed digital processors have made various modulation methods applicable for power converters. In principle, the aims of the modulation methods are to synthesize the desired voltage and current waveforms from the input source [67]. Nowadays, IGBT-based induction motor drives are the most common solution for railway applications. However, due to high switching losses as in high power and medium voltage drives, IGBT switching frequencies are limited between 500 Hz-3 kHz. This situation reduces voltage and current waveform quality, especially at medium power and relatively high speeds drives. The distortions on waveforms may be brought to sufficient levels by using combinations of special modulation methods. However, it increases the workload of the algorithm and requires pre-computational effort, and

also increases the complexity. In addition to this, strong computational ability such as GPU is needed if all operations are required to be done online.

2.6.1 Pulse Width Modulation Methods

In the past, six-step modulation was commonly used for motor drive applications at constant power and high-speed regions due to the allowable low switching frequency limits. However, this method creates low order harmonics (such as 5th, 7th, 11th, so on) with considerable magnitudes. Development in the semiconductor area reduces the switching losses significantly, and this loss reduction gives a chance to operate higher switching frequencies. As a result of this, utilization of various PWM methods has been possible for various applications. Thus, elimination or reduction of significant low order harmonics became possible depending on the power requirements.

Today, many pulse width modulation methods are widely used for synthesizing output voltage and current waveforms of inverters. The frequency and the magnitude of the inverter output voltage and current waveforms can be controlled with low harmonic content. The harmonic content of the output waveforms is strictly related to the applied number of pulses. In fact, the better harmonic content is achieved with the increased switching frequency. However, the increased switching losses still create a strong limit to increase switching frequency. Therefore, there is a trade-off between the reduction of the harmonics and the switching losses. In addition to this, not only the switching frequency but also applied PWM methods also have a strong influence on inverter/motor losses, harmonics as well as additional noise and vibration [73]. The most common PWM techniques used for motor drive applications are Sinusoidal Pulse Width Modulation (SPWM), SVPWM, Third Harmonic Injection Pulse Width Modulation (THIPWM), Discontinuous Pulse Width Modulation (DPWM), and programmed Pulse Width Modulation [74]. Moreover, SHEPWM, SHMPWM, and OPWM techniques are also used in railway traction applications [58].

2.6.1.1 Sinusoidal Pulse Width Modulation Method

The simplest PWM method used for motor drive applications is SPWM. In the scalar implementation of SPWM, sinusoidal modulation reference signals are compared with a carrier signal to generate pulse patterns to be applied to the semiconductor switches. The intersection of modulation references and carrier signal specify the switching instants. A basic illustration for the generation of pulse patterns for a three-phase system is given in Figure 2.33 [73].

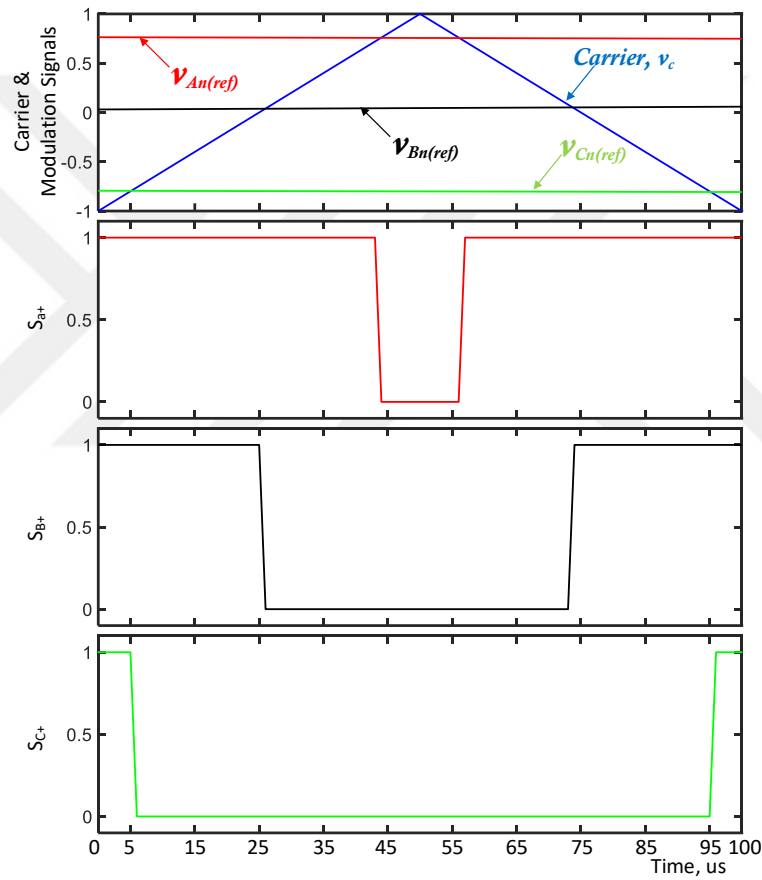


Figure 2.33. Scalar implementation of a PWM and the upper side switch signals

where v_c is the triangular carrier signal between -1 and 1 with a magnitude V_c and a frequency f_c . On the other hand, $v_{An(ref)}$, $v_{Bn(ref)}$, $v_{Cn(ref)}$ are the modulation reference signals with a magnitude V_m and a frequency f_m . In practice, for the low output frequencies, f_m is very small compared to f_c . Therefore, the variation in

modulation reference signal is very small and can be assumed as constant during the PWM period.

Definition of two fundamental parameters, amplitude modulation index m_i and frequency modulation index m_f are useful for the investigation of pulse width modulation methods.

$$m_i = \frac{V_m}{V_c} \quad (2-37)$$

$$m_f = \frac{f_c}{f_m} \quad (2-38)$$

The fundamental component of the output voltage waveform magnitude is controlled by the amplitude modulation index, whereas the output fundamental frequency is controlled by the frequency modulation index. DC-link voltage utilization and the output voltage/current waveform harmonics are also constituting two other fundamental parameters to explain the performance of output waveform synthesizing. In Figure 2.34, DC-link voltage utilization is given by considering the variation in m_i while the m_f and f_m are kept as 21 and 50 Hz. These results are obtained by a computer simulation with a RL load.

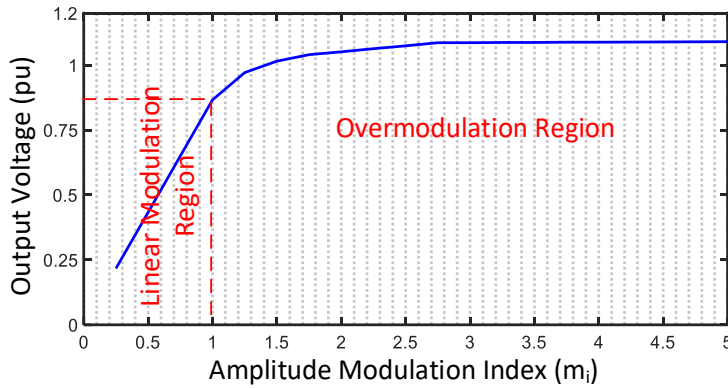


Figure 2.34. SPWM modulation regions and line-to-line output voltage linearity of the three-phase inverter

If the value of m_i is between 0 and 1, as seen from Figure 2.34, the inverter is operated under the linear modulation region. In this region, the fundamental component of the output voltage can be linearly changed with the increased m_i and the low order harmonics are quite low in this region compared to over-modulation and six-step operation regions. If the m_i is set to higher than 1, the output voltage cannot be changed linearly anymore with the increased m_i and considerable amount of low frequency harmonics begin to appear in the output voltage waveforms. After a limit, further increase in m_i does not create any impact on output waveforms. In this region, the inverter is operated under six-step modulation, and it is not possible to control the magnitude of the output voltage waveforms in three-phase voltage source inverters.

In addition to this, another important effect can be seen in the output voltage waveforms with the variation of m_f . In order to visualize the effect of m_f , a computer simulation has been conducted by keeping m_i as 0.8 with the RL load while reducing m_f from 21 to 5. If m_f is quite less than 21, the low order harmonics increased significantly. The carrier is synchronized with the fundamental frequency during this simulation study. The results are given in Figure 2.35.

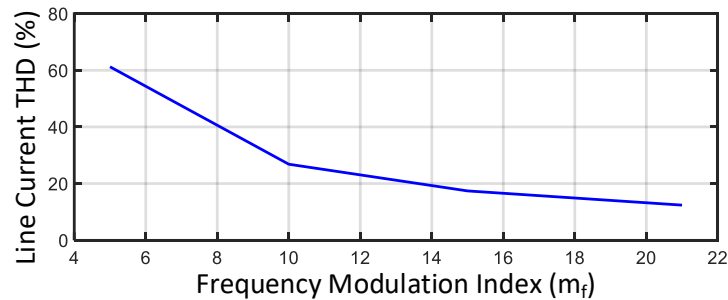


Figure 2.35. THD of line current versus frequency modulation index

In the light of the above, it can be summarized that the SPWM does not provide good performance for smaller values of m_f . Moreover, the DC-link utilization is also limited with 0.86 in the linear region.

2.6.1.2 Third Harmonic Injection Pulse Width Modulation Methods

In most of the old industrial drives, DC-link voltage is obtained by using six-pulse diode rectifiers. Thus, the average rectified DC-link voltage is limited with the input source voltage and source impedance. As a result, the efficient utilization of available DC-link voltage has important implications for using standard industrial motors without overheating. Especially in order to avoid low-frequency distortions, control of inverters without moving into over modulation region is very important. Buja and Indri in 1975 recognized that the maximum modulation index could be increased by adding a third harmonic zero-sequence voltage to modulation reference signals. Since the third harmonic components are injected to phase voltages, they are canceled between the phases, and line current waveforms are not affected due to the neutral point is not connected to the mid-point of the inverter [75]. Figure 2.36 (a) and (b) illustrate the modulation signals for one-sixth magnitude third harmonic injection and one-quarter magnitude third-harmonic injection.

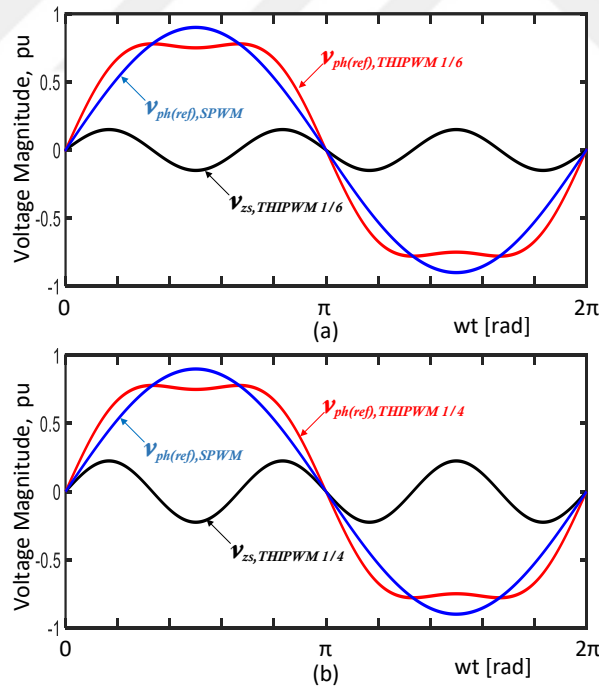


Figure 2.36. (a) Typical modulation and zero sequence reference signals (a) 1/6 THIPWM (b) 1/4 THIPWM

From analytically, anyone can easily compute that the modulation index is increased by %15 by adding one-sixth magnitude third harmonic and %12 by adding one-fourth magnitude third harmonic. Although the maximum increase is achieved with one-sixth magnitude third harmonic injection, some researchers have proposed that optimum results are achieved with one-fourth magnitude third harmonic injection because of the reduced harmonic distortions [75]. DC-link voltage utilization of one-fourth magnitude third harmonic injection PWM is shown in Figure 2.37.

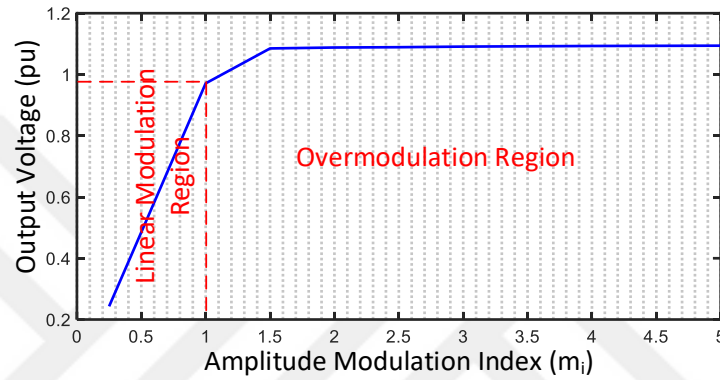


Figure 2.37. 1/4 THIPWM regions and line-to-line output voltage linearity of the three-phase inverter

2.6.1.3 Space Vector Pulse Width Modulation Method

Low DC-link utilization as in SPWM at linear region has led to the investigation of new modulation methods to provide a wider voltage linearity range. In motor drive applications, the neutral point of the electric motor is isolated. Thus, low-frequency zero-sequence currents cannot find a path to flow, although the neutral point voltage is different from zero. This situation gives a chance to inject a wide range of zero-sequence voltage to the inverter phase voltages. By injecting zero-sequence voltage, it is mainly aimed to improve waveform quality, switching losses, and wider voltage linearity range [76]. In the mid of 1980s, a form of PWM called space vector pulse width modulation was proposed (SVPWM). Nowadays, SVPWM is the most common PWM method, which is obtained by adding an optimized third harmonic

zero-sequence voltage to the phase voltages. In practice, analog and digital implementation of SVPWM have been applied. One of the simplest implementations is that the reference modulation signals are generated by selecting the purely sinusoidal modulation signal with the smallest magnitude and adding half of it to the original sinusoidal reference signal [76]. Figure 2.38 shows the modulation signal and third harmonic zero sequence component.

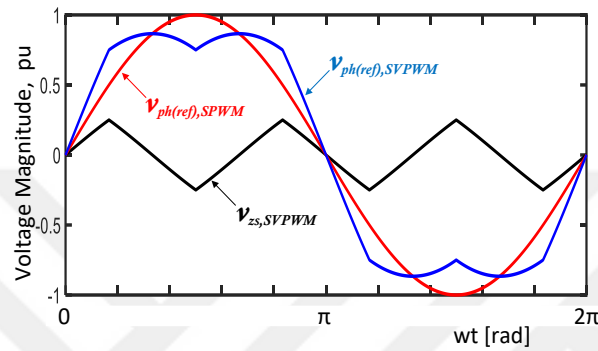


Figure 2.38. SPWM, SVPWM, and zero sequence signals

DC-link voltage utilization of SVPWM is illustrated in Figure 2.39. The SVPWM method has a wider voltage linearity range comparing the other CPWM techniques [76].

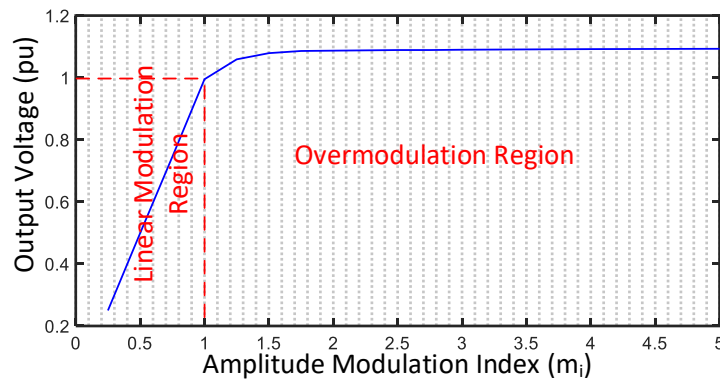


Figure 2.39. SVPWM regions and line-to-line output voltage linearity of the three-phase inverter

2.6.1.4 Discontinuous Pulse Width Modulation Methods

In the previous PWM methods, the reference modulation signals are always within the carrier signal boundaries in the linear regions, and therefore, in every carrier cycle, an on/off switching transition occurs. These methods are categorized as Continuous PWM (CPWM) methods due to this continuous switching operation. In the Discontinuous PWM (DPWM) methods, one semiconductor switch of each leg is always turned-on during its 120°, 60°, or 30° unmodulated region, while the other semiconductor switch of this leg is turned-off. These DPWM methods have the advantage of eliminating switching transitions during some carrier periods. There are six different DPWM methods, namely, DPWM0, DPWM1, DPWM2, DPWM3, DPWMMIN, and DPWMMAX. All these DPWM methods have different characteristics and provide different advantages. Therefore, choosing the correct type is very important for the application. DPWM1 has minimum switching losses at unity power factor operating conditions, while DPWM2 is suitable for lagging power factor operating conditions. DPWM2 has minimum switching losses at 30° lagging power factor. On the other hand, DPWM1 is the correct choice for leading power factor operating conditions, and it has minimum switching losses for at 30° leading power factor. In addition to this, DPWM3 provides the benefit of low harmonic distortion. DPWMMAX and DPWMMIN give the benefits of low harmonic distortion. However, the thermal stresses on semiconductors are non-uniform. DPWM2 is sometimes useful for high power induction machine drives because of the high semiconductor losses created with CPWM techniques. Because of this, in accordance with the subject of the study, only waveforms related to DPWM2 will be included. The modulation reference, phase voltages, line-to-line output voltage, and line current waveforms are given in Figure 2.40 and Figure 2.41 at $m_i=0.905$, $f_c=1050$ Hz, and $f_m=50$ Hz.

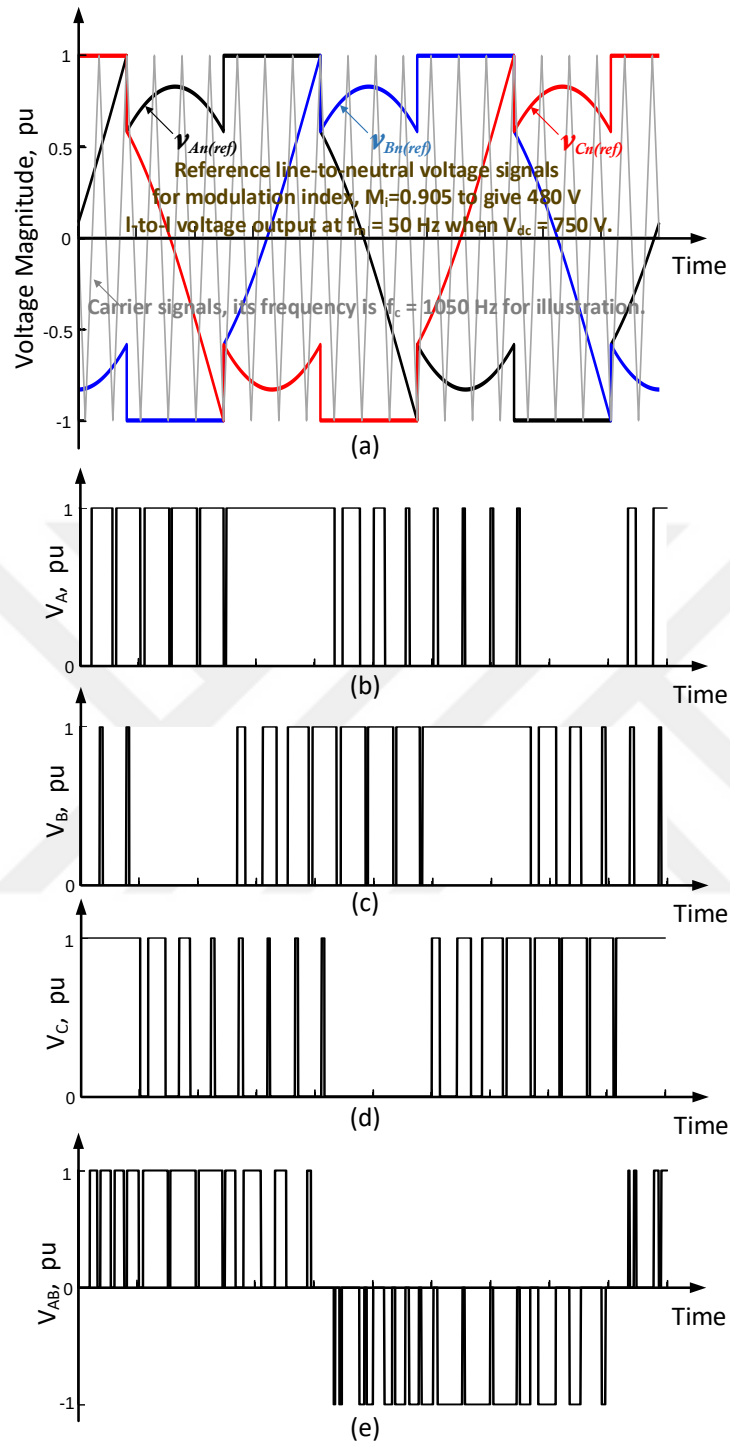


Figure 2.40. Typical voltage waveforms for DPWM2 modulated two-level voltage source inverter (a) Modulation and carrier signals (b) - (d) Phase voltages (e) Line-to-line output voltage

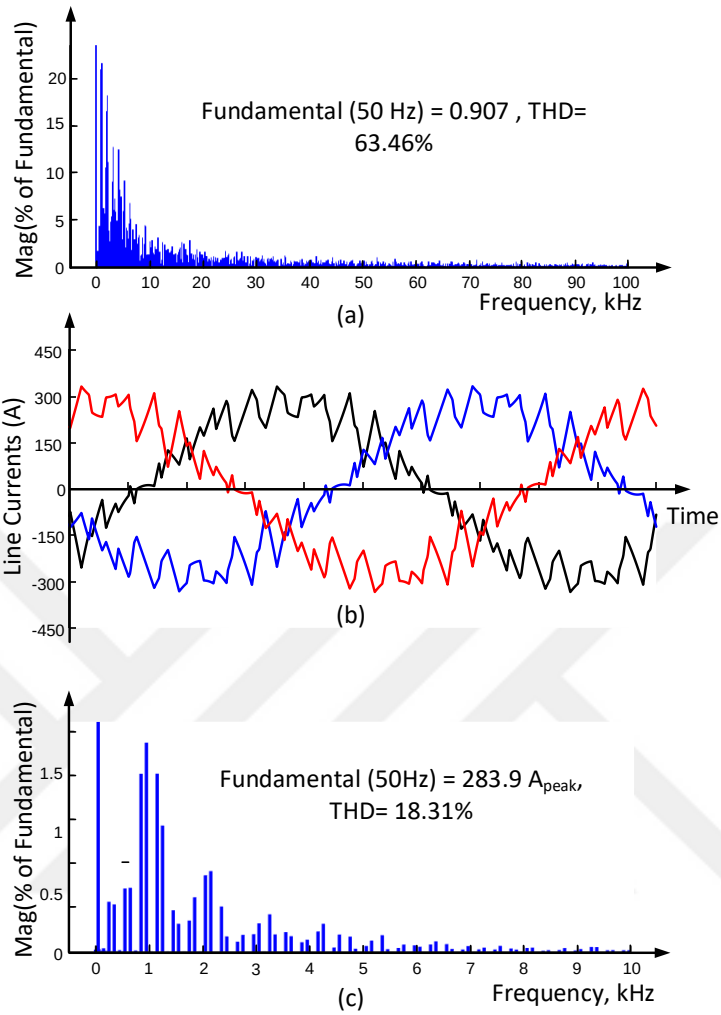


Figure 2.41. Typical harmonic spectrum and output current waveforms for DPWM2 modulated two-level voltage source inverter (a) THD of line-to-line voltage waveform (b) Three-phase output currents (c) THD the of output phase A current

The waveform quality of DPWM2 is superior to SVPWM at high modulation index, although it is not good as CPWM methods at low modulation. On the other hand, operation in the over modulation region is problematic, and in many studies, this subject was investigated [74], [76]. Moreover, the SVPWM technique is proved to have higher linearity than DPWM in general under the same switching frequency [77].

2.6.1.5 Programmed Pulse Width Modulation Methods

In high power traction applications, overall system efficiency is very important. Therefore, the switching and harmonic losses have to be considered to reduce operational cost and the thermal stresses (e.g., longer life time and higher operational system efficiency). Due to these reasons, the switching frequencies of IGBTs are limited in the range of 500-3000 Hz. For lower switching frequencies, m_f is getting very low, especially in high-speed operation regions. This reduces the resolution in PWM patterns, may lead to loss of full-wave, half-wave, and quarter-wave waveform symmetry. As a result of that, low order (5th, 7th, 11th, 13th, 17th, etc.) frequency harmonics are generated in the output waveforms. In practice, low-frequency harmonic components cannot be filtered by using small passive filters. In this case, the filter increases the product size, weight, and system cost. Because of this reason, bulky low pass filters are not preferred in traction applications. In addition to this, many studies show that multilevel converters have been providing good performance to decrease the harmonics in high power applications. However, this also increases the cost and decreases the reliability of the converter because of the increased number of semiconductors, gate-drives, capacitors, isolated power supplies, and so on. On the other hand, different PWM techniques may also help to manage low order harmonics up to a certain level. Instead of carrier-based PWM methods, three major programmed PWM methods, namely Selective Harmonic Elimination PWM, Selective Harmonic Mitigation PWM, and Synchronous Optimal PWM, are used to eliminate the low order frequency harmonics in practice [75], [78]–[80]. The use of these efficient modulation methods will be very convenient to obtain output waveforms with acceptable harmonic content in high-power applications. The approach of these modulation methods is based on the cancellation of low order current harmonics or the minimization of the RMS value of the current harmonics in the motor by a proper choice of the commutation angles by taking into account suitable objective functions [78].

By using the SHEPWM method, a limited number of low-order harmonics can be canceled. However, the non-canceled and relatively high order harmonics are not controlled. On the other hand, the SHMPWM method gives a chance to reduce the magnitudes of low order and high order harmonics altogether. The main goal of this modulation method is to reduce the filtering requirements in order to decrease the size, weight, and cost of the filtering elements by fulfilling the harmonic requirements [80]. In addition to this, Buja and Indri [81] proposed the SOPWM method, which considers a whole spectrum to minimize the harmonic distortion of the current in a motor.

In all these methods, the pulse angles have to be calculated by solving idealized analytical formulations. In practice, the complexity of the problem increases as the number of switching angles increases; finding the optimum solutions are getting harder, and the switching losses increase, although using more switching angles eliminates more harmonics. In order to solve a set of analytical formulations, there have been two groups of techniques presented in the literature, namely iterative solution and search algorithm techniques. Almost all of these techniques require heavy computational effort. Because of this difficulty, the angles have been solved with offline calculations and stored in lookup tables. However, in practice, this approach is not easily applicable for traction applications where the electrification line voltage is not fix. In order to optimize the harmonics, line voltage fluctuations have to be considered. All of the angles should be solved by considering many conditions (different modulation indexes, pulse numbers, etc.) and stored in the lookup tables. This creates big problems. In order to overcome these problems, online calculation of pulse angles can be done. Online calculations can be made by using a graphical processing unit that provides high-level computing technology. However, the system and labor cost increase a lot in this case.

2.7 Principles of Electric Traction Drives

The high-performance electric traction drive systems should offer the following features:

- four-quadrant operation,
- high power density,
- high conversion efficiency,
- fast-flux and torque response,
- fast dynamics for slip/slide protection,
- maximum output torque per ampere operation in wide speed range,
- high and constant switching frequency,
- high current control bandwidth,
- robustness to parameter variations,
- high reliability and low maintenance need,
- low cost,
- low flux and torque ripple,
- low current harmonic distortion for motor currents.

Hardware capabilities and applied control strategy should be considered together to provide the above-listed features. DC-link voltage, phase currents, modulation index m_i , slip, and rotor frequency are the important parameters for induction machine control. These may differ according to the applied motor control algorithms.

2.7.1 Control Methods for Voltage Source Inverter Fed Induction Machines in Railway Applications

In this section, control methods for voltage source inverter fed induction machine drive will be presented. This section has been performed with two main goals: presenting the theoretical background of each control method and comparative analysis and choosing the most appropriate control methods considering the operating regions.

Traction converters commonly receive a torque command from the master controller or CBCT/ERTMS systems. The applicable torque that can be produced at a given speed will essentially depend on the machine and inverter designs, as explained in this section. The operating regions as constant torque, constant power, and high-speed regions are defined. The traction inverters are responsible for driving induction machines properly in all these regions. The gain and phase margins are very important for stable motor control. The stability of such drives depends on available voltage, applicable switching frequency, applied control method, and modulation technique. In fact, all of these also dynamically depend on each other. Therefore, control method, modulation technique, switching frequency, and available voltage should be considered altogether for an efficient and stable traction drive system. Even for most of the medium-power drives, control methods, modulation techniques, and switching frequency are changed dynamically depending on the operating region of the traction motors.

Control methods for induction machines can be classified as scalar and vector types. Scalar control methods are derived from the steady-state equivalent machine models and are suitable for applications that do not require very fast changes in operating conditions. Vector control methods are contrarily used for high-performance drives, which require fast control loops and dynamic responses. Instead of steady-state equations, dynamic equations of the machine are used in this case. The upper-level block diagram of induction machine control methods is illustrated in Figure 2.42 [58].

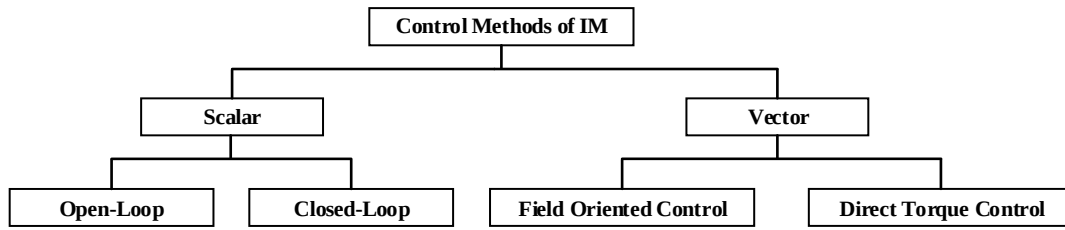


Figure 2.42. Control methods for induction machine

In fact, the dynamic response requirements derived from the operational profiles of railway vehicles show that relatively slow dynamic response is needed due to the jerk limitations. The open-loop scalar control method was used in early railway traction drives. Slip value was increased or decreased according to the given torque command. In the open-loop scalar control method, there is not any feedback mechanism. Later, closed-loop speed regulation was provided by controlling the slip according to the speed set value and the speed feedback signal. The block diagram of the closed-loop speed control-based scalar control method is shown in Figure 2.43 [58].

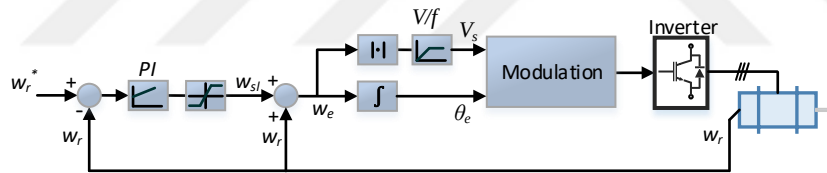


Figure 2.43. V/Hz control scheme with speed control [58]

However, the closed-loop speed-based control method does not give a chance the control torque accurately. Instead of this control method, flux and torque feedbacks are used in improved scalar control methods. A block diagram of the improved scalar control method is given in Figure 2.44. Using the line current measurement, applied stator voltage, and frequency, the torque and flux are estimated.

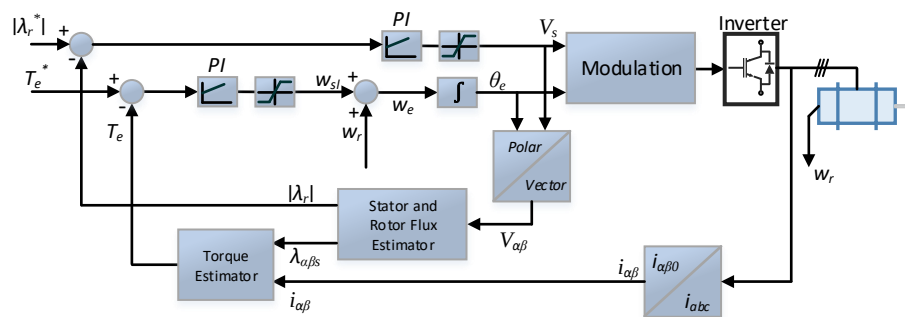


Figure 2.44. Scalar control scheme with torque-flux control [58]

In this improved closed-loop scalar control technique, the torque and flux values are controlled accurately in steady-state. However, it still needs improvement for good dynamic response characteristics. In order to improve the dynamic behavior of this control method, feed-forward terms are added by using the torque command and stator frequency [58]. While improving the dynamic response, the coupling between the flux and torque has to be taken into account to prevent any torque pulsation and instability.

Today, the most widely used control methods are vector-based control methods. In vector-based control methods, flux and torque components are aimed to control directly. They are very well described in the literature [78]–[87], where the superiority of vector control methods is clearly shown. Although the Field Oriented Control (FOC) and Direct Torque Control (DTC) methods are popular for medium power motor drive applications, they are often used at relatively low motor speeds. Block diagrams of the FOC and DTC methods are given in Figure 2.45 and Figure 2.46, respectively.

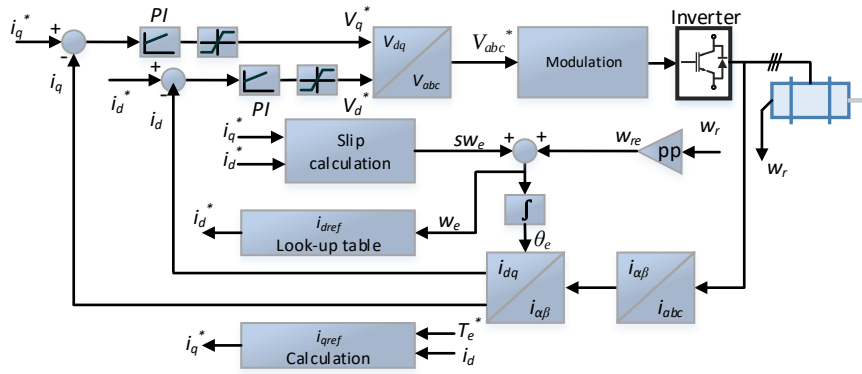


Figure 2.45. FOC block scheme [58]

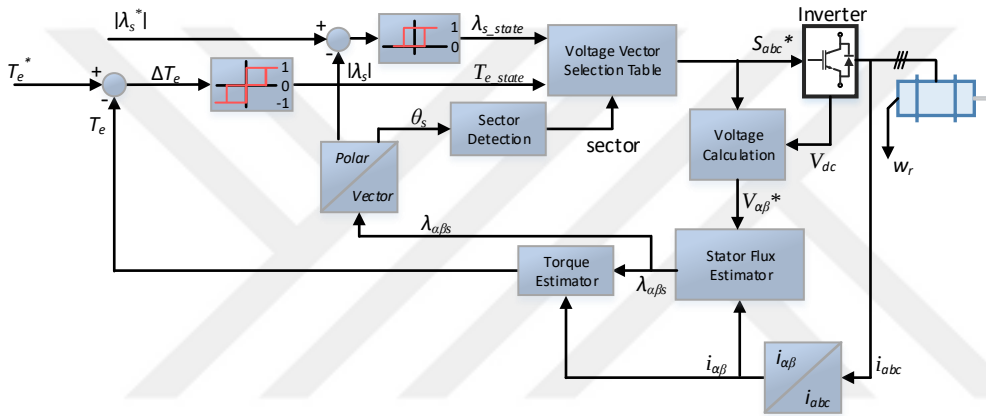


Figure 2.46. DTC block scheme [58]

In the FOC method, stator currents are measured and transferred to the rotating frame, and then flux and torque components are calculated. After that, the set values of flux and torque components and the calculated feedbacks are fed to high-bandwidth current regulators to produce modulation signals. As a next step, inverse transformations are performed, and the modulating reference signals are obtained. Finally, the modulator takes the reference signals and generates the necessary switching states of power semiconductors. On the other hand, in the DTC method, torque and flux components are directly calculated from the measured line currents and applied line voltages and then calculated values subtracted from the set values of torque and flux components and feed to the hysteresis controllers. The output of the hysteresis controller is sent to a pre-defined look-up table to decide the switching

states of power semiconductors. This control method is not suitable for operation at the boundary of the overmodulation region. In addition to this, it is difficult to keep the switching frequency constant. Hence, it is not preferred in high-power railway traction drives [58].

Moreover, due to the increased output fundamental frequency, the switching frequency to fundamental voltage-frequency ratio is decreased with increased speed. This causes to generate lower-order voltage harmonics hence current harmonics, and leads to decrease system efficiency and stability. In order to solve this problem, different modulation techniques, such as SHEPWM, SHMPWM, and OPWM with scalar motor control methods, have to be applied.

Consequently, the use of vector-based control methods in railway applications at high speeds can be problematic due to the insufficient voltage margin, limited switching frequency, and unsuitability for use with SHEPWM, SHMPWM, and OPWM techniques. In over modulation regions, the current and voltage distortions became prominent. This situation causes instability problems. On the other hand, V/Hz control methods are fitted well with these modulation techniques for high-speed operating regions, although the dynamic performance is poor. All these explain that why the control methods and modulation techniques are dependent on operating regions. In order to overcome the disadvantages of control methods and modulation techniques in all operating region, the use of a mix of control methods and modulation techniques are preferred. At relatively low speeds, the FOC method and SVPWM techniques can be used. While this is the case for low speed (e.g., generally is limited with constant torque region), the V/Hz control method can be used together with one of the above-mentioned preprogrammed modulation techniques for constant power and high-speed regions. Table 2.8 summarizes the properties of control methods [58]. It is also worthy to note that torque command step changes are normally limited in railway applications due to the jerk limitation, which is required for passenger comfort. Instead of this, the traction drive has to meet the acceptable torque/current ripple requirements.

Table 2.8 Summary of Presented Control Methods for Railway Traction

Properties	Control Methods				
	Open-Loop		Closed-Loop		
	V/Hz	V/Hz with only speed feedback	V/Hz with speed and current feedbacks	Rotor flux oriented control	Direct torque control
Reference frame	rotating	rotating	rotating	rotating	stationary
Fix switching frequency	yes	yes	yes	yes	no
Low speed performance	moderate	moderate	moderate	good	poor
High speed (overmodulation) performance	good	good	good	moderate	poor
Six-step operation	good	good	good	needs additional improvement	
Dynamic response	poor	poor	moderate	good	good

As explained above, the main source of the problem is the applicable low switching frequencies at high speeds. This picture began to change with the discovery of SiC power semiconductors. Hereafter, the traction drive systems using FOC control and SVPWM modulation techniques, which are almost the only solution in low power applications, will be possible instead of mixed control and modulation methods that switch between each other for medium power and medium voltage applications.

CHAPTER 3

DESIGN AND IMPLEMENTATION OF ALL SiC TRACTION INVERTER

The architecture of electric traction systems, brief information about traction motors, traction inverter topologies, and operating principles have been presented in Chapter 2. It was also emphasized that the traction inverters must perform high-performance tasks in harsh environmental conditions. In this chapter, the design methodology and implementation of a high-performance 500 – 900 VDC, 165 kVA all SiC traction inverter prototype is presented.

3.1 Design Methodology of Prototype System

Within the scope of this thesis, the prototype of a full SiC-based traction inverter has been designed, developed, and verified. To be able to define traction system requirements, traction analysis is firstly performed. Type of electrification system, the diameter of the vehicle wheel, the empty and fully-loaded weight of the vehicle, track data and acceleration/deceleration specifications are taken into account and then, voltage and current ratings of traction inverter and traction motor, and turns ratio of the gearbox are determined.

Light rail transportation systems are usually operating in the form of a vehicle set consisting of multiple cars. They run under very severe operating conditions of tight curves and frequent acceleration and braking. Generally, they are powered by 750 VDC electrification systems. The vehicles have maximum operating speeds of about 50-80km/h.

The flow chart of the proposed systematic design approach is shown in Figure 3.1. It starts by collecting the system requirements and doing traction analysis. Then, a proper inverter topology is chosen according to derived requirements. Switching

frequency, modulation technique, and current control method are determined, and then candidate power semiconductor, gate-driver, and dc-link capacitor selections and laminated busbar structure design are performed. In the next step, the total inverter loss and inverter efficiency are calculated, and then power quality requirements are checked. After that, the thermal analysis is done, and as a final step, system-level requirements such as efficiency, size, and weight are controlled. This flow defines an iterative process. Some or all of the critical elements such as switching frequency, modulation techniques, control methods, and candidate semiconductor selections can be changed according to the iterative calculations, and the process continues until optimum values are found.

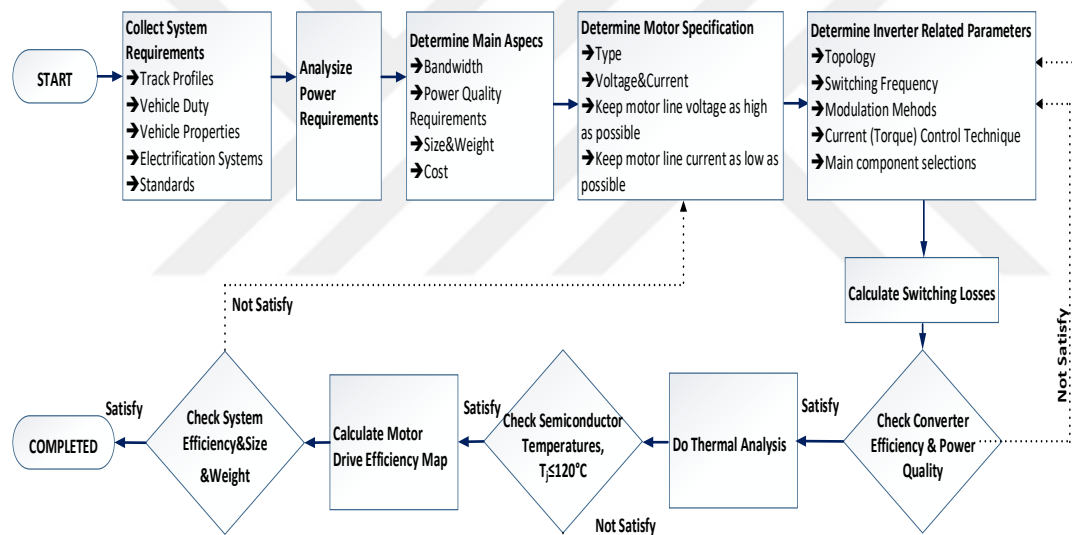


Figure 3.1. Flow chart of design methodology approach

3.2 Requirements of Prototype System

The design requirements are derived by considering the specifications of a sample light rail vehicle and a sample track. The vehicle requirements are summarized in Table 3.1 [93].

Table 3.1 Technical Specifications of The Light Rail Vehicle

Parameter	Values
Vehicle body structure	5 Cars, M+T+T+T+M
Number of traction motors, m	4
New wheel radius, r	300 mm
Gear ratio, n	7.5
Maximum slope	8.5 %
Maximum speed	50 km/h
Acceleration up to 35 km/h	1.2 m/s ² max.
Deceleration from 50 km/h	1.3 m/s ² max.
Jerk limit	1.0 m/s ³
Vehicle mass, M	39 tons unloaded (AW0) 56 tons loaded (AW3) 58 tons fully loaded (AW4)
Jgb+Jw (Gearbox and wheel inertia)	182 kg-m ²
A	0.0405m/s ²
B	0.0023s ⁻¹
C	3.81kg/m
D	0.053m
Curve radius, R	30m (Minimum)

By considering Table 3.1, traction analysis had been performed in MATLAB/Simulink environment and then the operating characteristics of traction motor drive have been revealed and given in Figure 3.2 and Table 3.2. The traction analysis is performed in the simulation chapter. In this section, only obtained data set is illustrated.

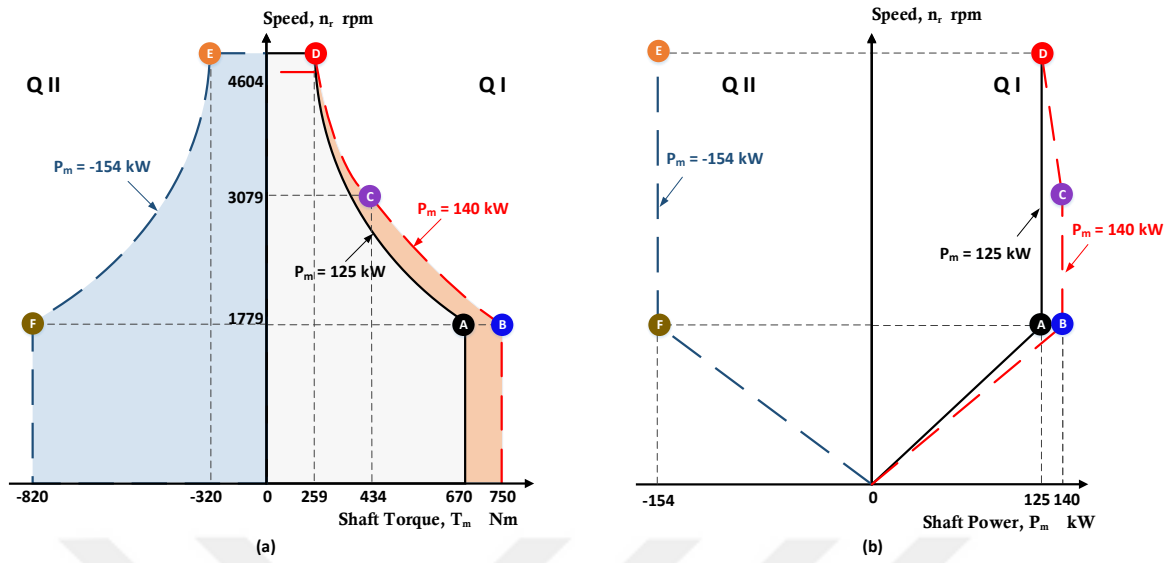


Figure 3.2. Operating characteristics of the traction motor drive for forward motion.
(a) Speed/torque characteristics, (b) Speed/shaft power characteristics

Table 3.2 Quantification of All Operating Points on Figure 3.2

Operating Points on Fig.1	Stator Frequency (Hz)	Shaft Speed (rpm)	Converter Output Power (kVA)	Stator Voltage V l-to-l	Motor Output Power (kW)
A	60	1779	165 at 0.81 pf lag	480	125
B	60	1774	185 at 0.81 pf lag		140
C	104	3079	170 at 0.89 pf lag		140
D	155	4604	154 at 0.88 pf lag		125
E	155	4604	176 at 0.81 pf lag		154
F	58	1779	172 at 0.83 pf lag		154

In the light of traction analysis results, the power requirements of the traction motor and SiC MOSFET-based traction inverter prototype have been revealed. Considering the results, products of the traction motor suppliers were investigated, and the best-suited traction motor was selected. The traction motor need had been tried to be met from the ready product on the shelf. The specifications for the vehicle requirements and the selected traction motor are listed in Table 3.3.

Table 3.3 Technical Specifications of Traction Motor

Vehicle Demand per Motor for the Chosen Track	Average Torque	500 Nm
	Average Speed	1500 rpm
	Average Power	80 kW
	Max. Torque	943 Nm
	Max. Speed	2250 rpm
	Max. Power	140 kW
4 – Pole, 60 Hz, 475 V, 125 kW, 0.83 pf (S1), $\eta = 93.5\%$		
Traction Motor (VEM DKCBZ 0212-4)	Rated Torque	671 Nm
	Rated Speed	1779 rpm
	Max. Torque	940 Nm
	Max. Speed	5100 rpm
	J_m	0.67 kg-m ²

Mass, volume, performance, stability, reliability, and availability are the main indicators for railway traction inverters. Higher component mass increases the vehicle mass, and similarly, higher component volume increases vehicle volume. This leads to higher energy demands and low vehicle performance. The design constraints such as system efficiency, control bandwidth, power to weight and power to volume ratio, output current THD, DC-link voltage ripples are considered as fundamental design requirements.

The technical specification of the traction inverter is listed in Table 3.4.

Table 3.4 Traction Inverter Target Specifications

Input Voltage Range	500 - 900 VDC (750 VDC rated)
Input Current	0 - 200 A
Output Voltage Range	0 - 500 V (480Vrms rated)
Output Current	0 – 225 Arms (200Arms rated)
Output Power	0 – 185 kVA (165 kVA rated)
Nominal Frequency	60 Hz
Inverter Efficiency	> 98.5 %
Traction Drive Efficiency	> 90 %
Current Bandwidth	> 170 Hz
Peak-to-Peak Torque Ripple	< 10 %

The selection of the inverter topology is essential to optimize the cost, power density, efficiency and to increase reliability. The popular inverter topologies are two-level and three-level NPC voltage source inverters for railway traction systems. Considering the voltage and power levels, two-level and three-level voltage source inverters are both applicable. In fact, the three-level NPC voltage source inverters have superior advantages for efficiency and harmonics at high voltage and high power levels. The benefits they provide for such a relatively low voltage and power level are very limited. Because of the increased switch numbers, three-level inverters are getting costly and less reliable. If the power requirements are strictly suitable to use two-level inverter topology, the industry decides to go for the simplest, mature and inexpensive solution. For the medium voltage and medium power traction derives, three-level NPC inverter topology is going to be mostly preferred topology. Better output voltage and current waveform quality can be obtained using a three-level NPC voltage source inverter by using the same switching frequency with two-level voltage source inverters. In addition to this, higher operational efficiencies, lower voltage, and thermal stresses can be achieved. On the other hand, rapidly spreading SiC semiconductor technologies will redefine the power levels that determine the transition between two-level and multilevel converter topologies. In

fact, by using SiC power switches, nearly ten times lower switching losses will be obtained comparing with the Si-IGBT technology. By the way, the SiC power switch technology eliminates the weaknesses of two-level voltage source inverters in terms of harmonics due to the high switching frequency is possible. Higher operational efficiencies and lower thermal stresses can also be achieved.

In the light of mentioned above, the two-level voltage source traction inverter topology has been chosen due to high reliability, good waveform quality, and cost-effective solution at this voltage and power level. As the next steps, the switching frequency, power semiconductors, modulation technique, and control method are determined. Next, line reactor, DC-link capacitors, laminated busbar structure, and heat-sink have been selected. After that, the design requirements have been implemented and verified by simulation studies conducted in Chapter 4. The simulations of the traction drive system have been carried out by using the MATLAB/Simulink simulation tool. The modulation techniques, control methods for traction inverters have been examined with the simulation model. Steady-state and transient state performance of the traction drive system have been investigated. Finally, all works will be confirmed on the laboratory prototype.

3.3 Selection of Power Switches and Switching Frequency

For 750 VDC catenaries, all traction inverters should withstand 500 - 900 VDC continuously, 1000 VDC for 5 minutes, and 1270 VDC for 20 ms, as defined in EN 50163:2004-A1:2007. Also, a safety factor should be added to this voltage rating due to the switching transients. This requires the use of power semiconductors with a standard blocking voltage capability of 1700 V. In addition to this, the current rating of candidate power semiconductors should also be determined initially without doing some thermal analysis. In this application, the maximum apparent power that will be delivered by the inverter is 185 kVA. The associated maximum input current per inverter, I_{max} , will be supplied from the catenary line when the catenary line voltage is 750 V DC is 200 A DC. Therefore, the input power is limited by 150 kW per

inverter. Below this voltage level, the input current kept constant, and power linearly decreased to 120kW up to 600 V DC and then linearly decreased towards zero for 500 V DC catenary line voltage as defined in Chapter 1. Current through each switch (controlled semiconductor and its antiparallel diode) is a chopped alternating current waveform and is a combination of current components drawn from the catenary and the DC link capacitors. In this application, the maximum value of the peak current through the controlled semiconductor is 320 A. Therefore, the continuous DC current rating of the chosen power semiconductor is selected as around 300 A. When making this selection, overload, short-circuit fault conditions, and the fault protection concept should also be considered.

In Figure 3.3, the torque-speed curves and main operating points considered in this study are illustrated.

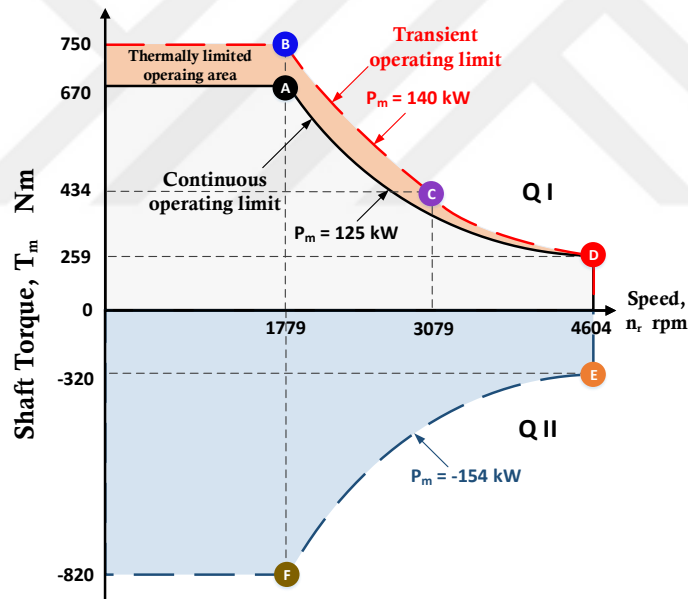


Figure 3.3. The torque-speed curve of the traction motor

From there, the minimum current control bandwidth and the minimum switching frequency requirements are initially deduced as 155 Hz and 1550 Hz, respectively. By adding a safety margin to include 10 % over speed, the current control bandwidth is determined as 170 Hz. On the other hand, in order to obtain the minimum output current total harmonic distortion content by using the asynchronous PWM technique,

the ratio of the possible switching frequency to the output fundamental frequency will be tried to be provided as a minimum of 21. This requires a minimum switching frequency of 3255 Hz. In addition to this, peak-to-peak torque ripple has to be limited to 10 % for all operating points on the torque-speed curve. In order to check torque ripple requirement, MATLAB/Simulink simulations were performed simultaneously. The torque ripples were analyzed against different switching frequencies for the 60 Hz and 155 Hz output fundamental voltage frequencies while the traction motor is delivering 125 kW to its shaft. Obtained torque ripples at rated speed and maximum speed operating points against switching frequency are shown in Figure 3.4 (a) and (b). In this simulation study, both the two-level and three-level inverter topologies are considered to make the benefits of SiC MOSFETs more visible. It is seen that torque ripple decreases with the increased switching frequency for both two and three-level topologies. As mentioned before, due to the high switching losses of Si-based IGBTs, it was not possible to reach high switching frequencies, and topology change (e.g., transition from two-level topology to three-level topology) is required. With SiC semiconductors, high switching frequencies can be achieved without being exposed to negative effects such as low efficiency and high thermal stresses. This improvement makes the two-level topology more advantageous than the three-level topology, given the increase in material counts, complexity, and the reduction in MTBF.

4000 Hz and above switching frequencies for three-level inverters are enough to satisfy the peak-to-peak torque ripple requirement, while two-level inverters require 8000 Hz and above switching frequencies. Comparisons for power losses and semiconductor temperatures will be given in the next sections.

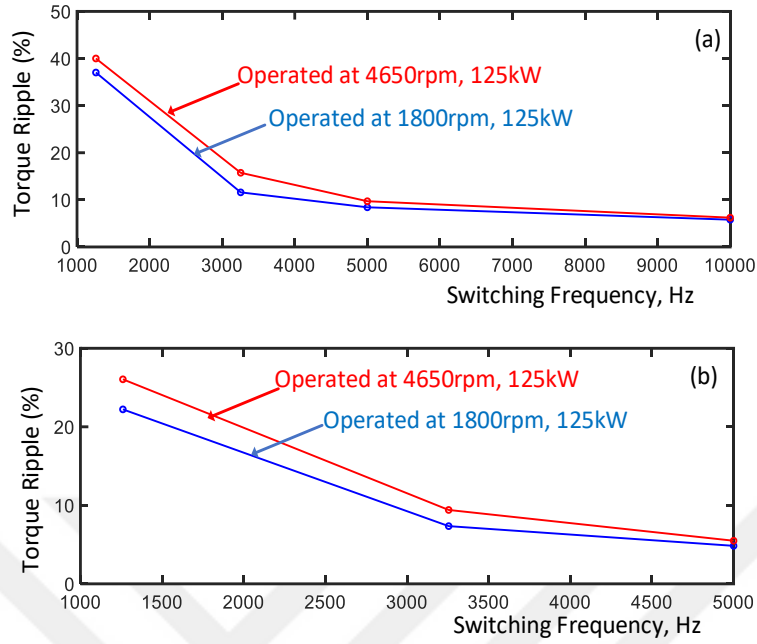


Figure 3.4. Peak-to-peak torque ripple comparison with different switching frequencies for (a) SVPWM two-level inverter (b) SVPWM three-level NPC inverter

Minimum inverter efficiency and total system efficiency (e.g., inverter & motor) for all operation points defined on the torque-speed curves should be 98.5 % and 90 %, respectively. In fact, the size and weight requirements of traction drives are also very strict in order to optimize vehicle volume and weight. However, a closed compact package of traction inverter was not made to work comfortably on the prototype in the laboratory. Because of this, volume and weight requirements are not defined in this study.

3.3.1 Loss Evaluation for 2L Inverter Topology

Simple converter topology, lower complexity in the control algorithm and modulation methods for the whole operating range, high switching frequency to minimize audible noise, and smaller size are qualitative objectives in the converter design. On the other hand, high efficiency, high reliability, less harmonic distortion in the applied motor currents, especially in low order harmonic range, and wide

current and hence torque bandwidth are being the quantitative design objectives. In the design phase, the traction converter, which optimizes not only the initial cost/kW-converter but also the operating cost (lower energy consumption, less maintenance), should be preferred.

Two-level classical three-phase bridge topology without parallel or series-connected power semiconductors in all legs in conjunction with the most common space vector PWM as the modulation technique and field-oriented control to adjust rotor flux during operation meets some of the design objectives given above. To complete the optimum design work, the most suitable types of power semiconductors among Si-IGBT, hybrid IGBT, asymmetrical IGCT, and SiC Power MOSFET and the optimum switching frequency for the chosen power semiconductor should be determined by carrying out some computer simulations. In traction converters supplied from 750 V DC catenary lines, asymmetrical IGCTs are not used because of their very high current and voltage ratings suitable only for operating at 1500 and 3000 V DC-link voltages.

Commercially available candidate power semiconductors evaluated in this study are given in Table 3.5. Note that the traction application compliant semiconductors were not used in this study to find the comparable package, current, and voltage specifications that are similar to the SiC MOSFET module.

Table 3.5 Technical Specifications of Candidate Semiconductors

	Semiconductor Modules			
Ratings	Si-IGBT Infineon FF300R17KE4	Si-IGBT Mitsubishi CM300DY-34T	Hybrid-IGBT Fuji Electric 2MSI400VAE-170-53	SiC MOSFET Wolfspeed/Cree CAS300M17BM2
Max. Blocking Voltage (V)	1700	1700	1700	1700
Continues Current (A) @ 25°C	440	300	520	325
$R_{ds(on)}/V_{ce(sat)}$	1.95V	1.95V	2.15	8 mOhm
Total Gate Charge (nC)	3050	2350	3000	1076
Recovery Charge (μC) @125°C	105	105	almost 0	almost 0
Max. Junction Operating Temp. (°C)	150	150	150	150
Base Area (mm ²)	61.4x106.4	62x108	80x110	61.4x106.4
Thermal Resistance per MOSFET/IGBT Junction-to-case (K/kW)	40	44.0	33	67
Thermal Resistance per Diode Junction-to-case (K/kW)	65	67.7	70	60

Selection of the power switches and the switching frequency is a fundamental issue for a power converter. Whole system performance is directly affected by this choice. When selecting the power switches and the switching frequency for the traction inverter, several practical issues should be considered. Traction motor current harmonics, desired rotational speed of traction motor or current control bandwidth of the inverter, system efficiency, modulation techniques, DC-link capacitor currents, input harmonics, power switch thermal resistances, and the cooling system requirements have to be considered for selection of power switches and switching frequency. In addition to this, the characteristic of the power switch should be identified very well since the power switch junction temperature depends on thermal resistances and capacitances, conduction, and switching losses. The switching frequency of the active switches is very important for output waveform quality. Higher switching frequency is quite well in order to obtain better waveform quality; however, it is limited with the semiconductor losses. In medium voltage and medium power applications, the switching frequency cannot be kept higher than a few kHz

(500 Hz - 3 kHz) for conventional Si-IGBT-based traction drives. Synchronization between the carrier and modulating signals is quite important if the ratio of the carrier signal and modulating reference frequency is low. Otherwise, it may create subharmonics, and these subharmonics create additional loss and torque ripple [93]. In addition to this, increased torque ripple will reduce the system stability for low inertia systems.

3.3.1.1 IGBT Conduction Loss Calculation

Before starting the loss calculations, the current waveforms passing through the semiconductors in the Si-IGBT-based traction drivers are shared in Figure 3.5 in order to make the difference between IGBT and MOSFET visible. Since the IGBT is a unidirectional switch, the current can only flow in the forward direction. In the first half period, only the IGBT S1 is conducting, as seen in Figure 3.5 (a). We can also see from Figure 3.5 (b) that D1 is only conducting in the second half period. Hence, during the first half period, only the S1 produces losses, and similarly, only D1 produces losses in the second half period.

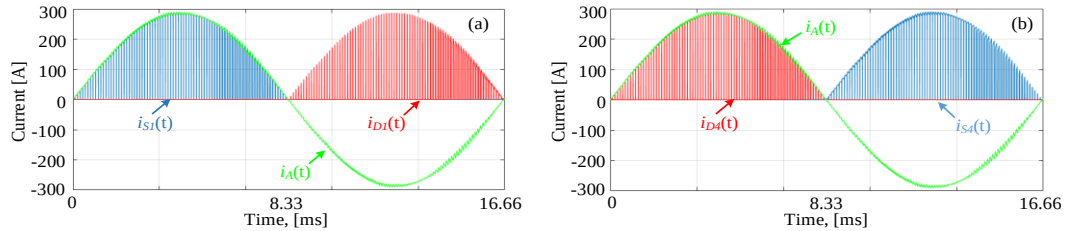


Figure 3.5 Current waveforms of S1, D1, S4, and D4, obtained by MATLAB/Simulink when the traction inverter is supplied from 750 V DC and applies 480 V l-to-l at 60 Hz to the stator and is supplying 173 kVA at 0.81 pf

The calculation of conduction losses is not an easy task due to the difficulty in the calculation of conduction durations. It directly depends on modulation techniques. In the literature, calculations are commonly made for the SPWM technique as it is relatively easy to formulate. IGBT and diode loss equations are given below in (3-1) – (3-4) for sinusoidal pulse width modulated two-level voltage source inverters [93].

IGBT conduction loss equation:

$$P_{TC} = u_{ce0} * I_{Tav} + r_c * I_{Trms}^2 \quad (3-1)$$

$$= u_{ce0} * I_o * \left(\frac{1}{2\pi} + \frac{m_i * \cos(\phi)}{8} \right) + r_c * I_o^2 * \left(\frac{1}{8} + \frac{m_i * \cos(\phi)}{3\pi} \right) \quad (3-2)$$

Diode conduction loss equation:

$$P_{DC} = u_{D0} * I_{Dav} + r_D * I_{Drms}^2 \quad (3-3)$$

$$= u_{D0} * I_o * \left(\frac{1}{2\pi} - \frac{m_i * \cos(\phi)}{8} \right) + r_D * I_o^2 * \left(\frac{1}{8} - \frac{m_i * \cos(\phi)}{3\pi} \right) \quad (3-4)$$

where, u_{ce0} is on-state zero current collector-emitter voltage, r_c collector-emitter on-state resistance, I_{Tav} average IGBT current, I_{Trms} RMS IGBT current, m_i amplitude modulation index, ϕ displacement factor, I_o the peak value of the output current, u_{D0} on-state zero current forward voltage, r_D on-state resistance, I_{Dav} average diode current, I_{Drms} RMS diode current.

3.3.1.2 SiC MOSFET Conduction Loss Calculation

The waveforms of the currents passing through the semiconductors in the SiC power MOSFET-based traction drives for low and high power levels are given in Figure 3.6. Since the MOSFET is a bidirectional switch, the current can flow in forward and reverse directions. Unlike the IGBT-based traction drives, S1 is conducting in both half-cycles, as seen in Figure 3.6. On the other hand, we can also see that D1 is only conducting in the second half period as same as in IGBT-based drives. Hence, during the full period, power switch S1 produces losses, and only D1 produces losses in the second half period. As can be seen from Figure 3.6, there are additional issues to take care of. Since the die thicknesses of SiC power MOSFETs are considerably thinned, the on-resistance of the semiconductor is getting very low. Accordingly,

while in reverse conduction mode, MOSFET takes a considerable amount of the current on itself, and the current passing through the diode is greatly reduced. As seen in Figure 3.6 (c) and (d), SiC-based antiparallel freewheeling Schottky diode does not conduct at all (e.g., $i_{D1}=0$), and the full current is passing through the MOSFET only when the output current is low.

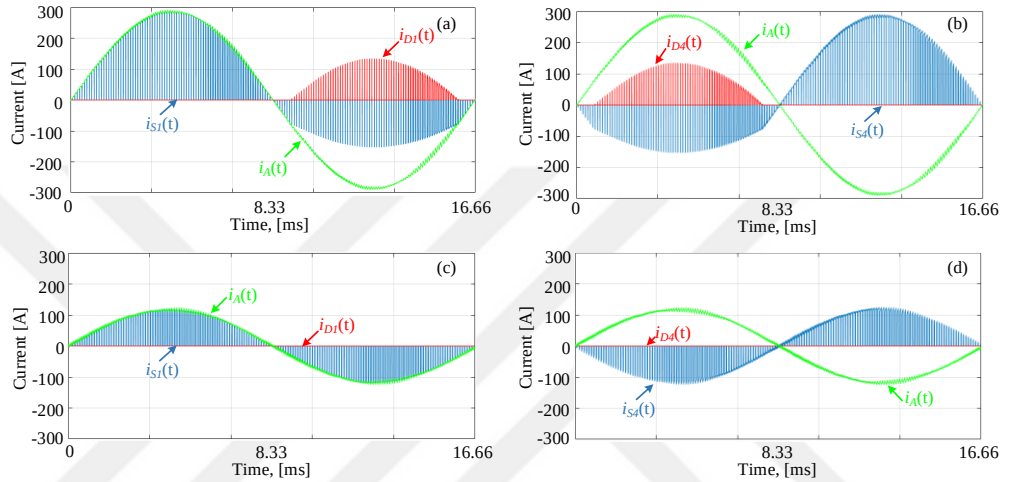


Figure 3.6. Current waveforms of S1, D1, S4, and D4, obtained by MATLAB/Simulink when the traction inverter is supplied from 750 V DC and applies 480 V l-to-l at 60 Hz to the stator. SiC power MOSFET based converter (a) and (b) supplies 173 kVA at 0.81 pf, and (c) and (d) supplies 75 kVA at 0.29 pf

This is an advantage of SiC power MOSFET over Si and hybrid IGBTs. By this reverse conduction ability, the current is shared between MOSFET and diode, thus reducing the conduction loss at relatively high operating powers. The electrical connection diagram of the first leg of the traction converter and the associated all-SiC half-bridge module are as shown in Figure 3.7 (a) and (b), respectively [93].

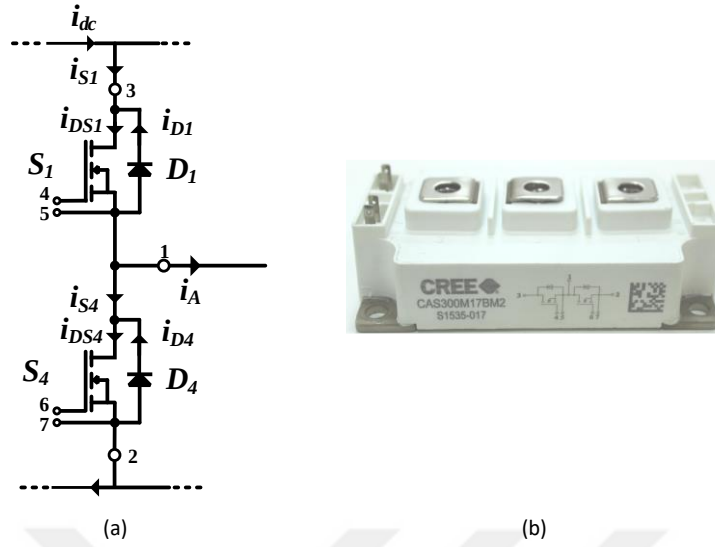


Figure 3.7. CAS300M17BM2 MOSFE module (a) Circuit diagram (b) Module package

Low device current in the reverse direction is carried only by the SiC power MOSFET chip. However, high device current in the reverse direction will be shared between SiC power MOSFET chip and SiC Schottky diode chip. The transition point can be expressed as follows [93]:

$$-i_{DS} \times R_{ds(on)} \geq V_{F(th)} \quad (3-5)$$

where, i_{DS} is the instantaneous value of the drain-to-source current, $R_{ds(on)}$ the equivalent resistance of the SiC power MOSFET chip during the conduction, and $V_{F(th)}$ is the threshold forward voltage drop of the SiC Schottky diode.

The current sharing between MOSFET and Schottky diode chips when both of them are conducting can be formulated as given in (3-6) and (3-7) [93]. Forward conduction voltage drop, V_F of SiC Schottky diode depends on forward current, i_D and its virtual junction temperature, T_{vj} while $V_{F(th)}$ depends only on T_{vj} .

$$i_{DS1} = \frac{-V_F + i_{S1}r_{diode}}{R_{ds(on)} + r_{diode}} \quad (3-6)$$

$$i_{D1} = \frac{-V_F - i_{S1}R_{ds(on)}}{R_{ds(on)} + r_{diode}} \quad (3-7)$$

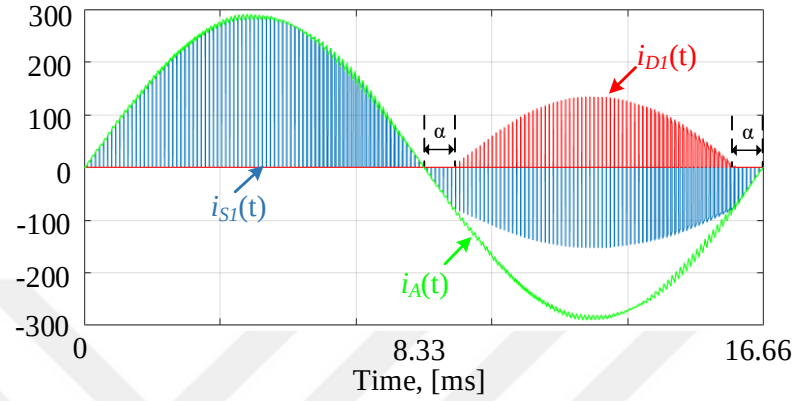


Figure 3.8. The instantaneous one phase, upper MOSFET, and diode current waveforms

In Figure 3.8, α defines the regions where the MOSFET alone is in reverse conduction. Please note that the displacement angle ϕ is also very important due to the inductive nature of the induction machine. While calculating the duty cycles, the modulation signal reference is used. Therefore, in order to use output current values together with corresponding duty cycles, the displacement angle should also be considered. A closed-form of the conduction loss calculation of MOSFETs is given in (3-8) – (3-10).

$$P_{MC} = \frac{1}{2\pi} \int_0^{2\pi} r_{dson} * (I_o * \cos(\theta))^2 * d_1(\theta + \phi) d\theta \quad (3-8)$$

$$P_{MC} = \frac{1}{2\pi} \int_0^{2\pi} r_{dson} * (I_o * \cos(\theta))^2 * d_1(\theta + \phi) d\theta \quad (3-9)$$

$$\begin{aligned} P_{MC} &= \frac{1}{2\pi} \int_0^{\pi+\alpha} r_{dson} * (I_o * \cos(\theta))^2 * d_1(\theta + \phi) d\theta \\ &+ \frac{1}{2\pi} \int_{\pi+\alpha}^{2\pi-\alpha} r_{dson} * (I_o * \cos(\theta) - i_{D1}(\theta))^2 * d_1(\theta + \phi) d\theta \\ &+ \frac{1}{2\pi} \int_{2\pi-\alpha}^{2\pi} r_{dson} * (I_o * \cos(\theta))^2 * d_1(\theta + \phi) d\theta \end{aligned} \quad (3-10)$$

where d_1 is the duty cycle of power MOSFET switch S1. It differs according to moving angle, and the expression of the duty cycle can be defined as in (3-11).

$$d_1(\phi) = \frac{1}{2} (1 + m_i * \sin(\phi)) \quad (3-11)$$

Similarly, the conduction loss equation of diode can be written as in (3-12).

$$P_{MC} = \frac{1}{2\pi} \int_{\pi+\alpha}^{2\pi-\alpha} (u_{D0} * i_{D1}(\theta) + r_D * i_{D1}^2(\theta)) * d_1(\theta + \phi) d\theta \quad (3-12)$$

The detailed forms of the MOSFET and diode conduction loss equations are given in [93].

3.3.1.3 SiC MOSFET and IGBT Switching Loss Calculation

In order to make a simple calculation for switching losses, one half cycle DC equivalent output current of the inverter is used. The equivalent DC output current value is obtained with (3-13).

$$I_{DC} = \frac{1}{\pi} * I_o \quad (3-13)$$

By using I_{DC} , E_{on} and E_{off} energies of MOSFET, IGBT, and diode can be found from the product datasheets.

IGBT and MOSFET switching loss equation:

$$P_{Tsw} = (E_{Ton} + E_{Toff}) * f_{sw} * \frac{I_o}{\pi I_N} * \frac{V_{DC-link}}{V_N} \quad (3-14)$$

$$P_{Msw} = (E_{Mon} + E_{Moff}) * f_{sw} * \frac{I_o}{\pi I_N} * \frac{V_{DC-link}}{V_N} \quad (3-15)$$

Diode switching loss equation:

$$P_{Dsw} = (E_{Don} + E_{Doff}) * f_{sw} * \frac{I_o}{\pi I_N} * \frac{V_{DC-link}}{V_N} \quad (3-16)$$

Total losses of IGBT, MOSFET, and diode can be finally obtained with the following equations:

$$P_{Ttotal} = P_{TC} + P_{Tsw} \quad (3-17)$$

$$P_{Dtotal} = P_{DC} + P_{Dsw} \quad (3-18)$$

Simulation software of IPOSIM, Melcosim, Fuji IGBT Simulator, and SpeedFit are used in performance calculations respectively of FF300R17KE4P, CM300DY-34T, 2MSI400VAE-170-53, and CAS300M17BM2 for different switching frequencies. Note that the traction application compliant semiconductors were not used in this study to find the comparable package, current, and voltage specifications that are similar to the SiC MOSFET module. The power loss analysis results of the 2L inverter are summarized in Table 3.6. The allowable junction temperature limit is taken as 120°C for calculations in order to increase the lifetime of the power semiconductor. In the previous section, 8 kHz is determined as the minimum switching frequency to satisfy the peak-to-peak torque ripple requirement.

Table 3.6 Power Loss Analysis of 2L Inverter for Different Frequencies

f_{sw} (Hz)	f_s (Hz)	Loss & Junc. Temp	S=165 kVA pf=0.81 lag, I_{out} =200 Arms, V_{l-l} =480Vrms M_i =0.905, R_{th-a} = 18K/kW															
			Si IGBT Module Infineon FF300R17KE4				Si IGBT Module Mitsubishi CM300DY-34T				Hybrid IGBT Module Fuji Electric 2MSI400VAE-170-53				SiC MOSFET Module Wolfspeed/Cree CAS300M17BM2			
			IGBT		Diode		IGBT		Diode		IGBT		Diode		MOSFET		Diode	
			Sw.	Cond.	Sw.	Cond.	Sw.	Cond.	Sw.	Cond.	Sw.	Cond.	Sw.	Cond.	Sw.	Cond.	Sw.	Cond.
1260	60	Loss (W)	51	163	17	33	50	170	12	50	48	142	6	30	-	-	-	-
		T_{vj} (°C)	94.3		75.5		84.7		76.5		77.9		73.1		-		-	
3255	60	Loss (W)	143	168	47	33	135	177	37	49	127	145	14	31	-	-	-	-
		T_{vj} (°C)	117.1		92.4		108.2		95.8		91.2		83.8		-		-	
3255	155	Loss (W)	142	169	46	31.7	135	177	36	49	126	145	14	31	-	-	-	-
		T_{vj} (°C)	115.2		91.1		108.2		95.8		90.2		83.4		-		-	
5000	60	Loss (W)	240	174	82	34	208	177	56	49	199	148	22	31	20	263	-	14
		T_{vj} (°C)	141.9		120.23		134.2		119.6		103.4		93.5		109.2		90.25	
5000	155	Loss (W)	237	174	79	32	208	177	56	49	199	148	22	31	19	262	-	14
		T_{vj} (°C)	138.2		118		134.2		119.6		102.1		93.1		108		89	
10000	60	Loss (W)	591	190	260	35	416	177	113	49	450	157	42	32	44	271	-	14
		T_{vj} (°C)	>200		191.6		186		163.4		145		126.2		115.7		94.9	
10000	155	Loss (W)	591	191	258	34	416	177	113	49	446	156	42	32	42	270	-	13
		T_{vj} (°C)	>200		185.2		186		163.4		142.4		125.2		114.8		93.8	

The results show that only the SiC power MOSFET can operate above 8 kHz switching frequency.

3.3.2 Loss Evaluation for 3L NPC & 3 L T Type Inverter Topologies

Semiconductor losses have been calculated with the help of analytical methods presented in this section and semiconductor manufacturer's simulation programs. IGBT and diode loss equations are given below for sinusoidal pulse width modulated three-level NPC and T-type voltage source inverters [94].

IGBT conduction loss equation for 3L NPC inverter:

IGBT 1&4:

$$P_{TC1\&4} = \frac{m_i * I_o}{12\pi} * (3 * u_{ce0} * [(\pi - \phi) * \cos(\phi) + \sin(\phi)] + 2 * r_c * I_o * [1 + \cos(\phi)]^2) \quad (3-19)$$

IGBT 2&3:

$$P_{TC2\&3} = \frac{I_o}{12\pi} * (u_{ce0} * [12 + 3 * m_i * (\phi \cos(\phi) - \sin(\phi))] + r_c * I_o * [3\pi - 2 * m_i(1 - \cos(\phi)^2)]) \quad (3-20)$$

Diode conduction loss equation for 3L NPC inverter:

Diode 1&4:

$$P_{DC1\&4} = \frac{m_i * I_o}{12\pi} * (3 * u_{D0} * [-\phi \cos(\phi) + \sin(\phi)] + 2 * r_D * I_o * [1 - \cos(\phi)]^2) \quad (3-21)$$

Diode 2&3:

$$P_{DC2\&3} = \frac{m_i * I_o}{12\pi} * (3 * u_{D0} * [-\phi \cos(\phi) + \sin(\phi)] + 2 * r_D * I_o * [1 - \cos(\phi)]^2) \quad (3-22)$$

Diode 5&6:

$$P_{DC5\&6} = \frac{I_o}{12\pi} * (u_{D0} * [12 + 3 * m_i * (2 * \phi - \pi) * \cos(\phi) - 2 * \sin(\phi)] + r_D * I_o * [3\pi - 4 * m_i * (1 + \cos^2(\phi))]) \quad (3-23)$$

IGBT switching loss equation for 3L NPC inverter:

IGBT 1&4:

$$P_{T1\&4SW} = (E_{Ton(N)} + E_{Toff(N)}) * f_{sw} * \frac{I_o}{I_N} * \frac{(0.5 * V_{DC-link})}{V_N} * \left(\frac{1}{2\pi} * [1 + \cos(\phi)] \right) \quad (3-24)$$

IGBT 2&3:

$$P_{T2\&3SW} = (E_{Ton(N)} + E_{Toff(N)}) * f_{sw} * \frac{I_o}{I_N} * \frac{(0.5 * V_{DC-link})}{V_N} * \left(\frac{1}{2\pi} * [1 - \cos(\phi)] \right) \quad (3-25)$$

Diode 1&2:

$$P_{D1\&2SW} = E_{REC(N)} * f_{sw} * \frac{I_o}{I_N} * \frac{(0.5 * V_{DC-link})}{V_N} * \left(\frac{1}{2\pi} * [1 - \cos(\phi)] \right) \quad (3-26)$$

Diode 2&3:

$$P_{D2\&3SW} = 0 \quad (3-27)$$

Diode 5&6:

$$P_{D5\&6SW} = E_{REC(N)} * f_{sw} * \frac{I_o}{I_N} * \frac{(0.5 * V_{DC-link})}{V_N} * \left(\frac{1}{2\pi} * [1 + \cos(\phi)] \right) \quad (3-28)$$

IGBT conduction loss equation for 3L T-Type inverter:

IGBT 1&4:

$$P_{TC1\&4} = \frac{m_i * I_o}{12\pi} * (3 * u_{ce0} * [(\pi - \emptyset) * \cos(\emptyset) + \sin(\emptyset)] + 2 * r_c * I_o * [1 + \cos(\emptyset)]^2) \quad (3-29)$$

IGBT 2&3:

$$P_{TC2\&3} = \frac{I_o}{12\pi} * (u_{ce0} * [12 + 6 * m_i * (\emptyset \cos(\emptyset) - \sin(\emptyset)) - 3 * m_i \pi * \cos(\emptyset)] + r_c * I_o * [3\pi - 4 * m_i (1 - \cos^2(\emptyset))]) \quad (3-30)$$

Diode 1&4:

$$P_{DC1\&4} = \frac{m_i * I_o}{12\pi} * (3 * u_{D0} * [-\emptyset \cos(\emptyset) + \sin(\emptyset)] + 2 * r_D * I_o * [1 - \cos(\emptyset)]^2) \quad (3-31)$$

Diode 2&3:

$$P_{DC2\&3} = \frac{I_o}{12\pi} * (u_{D0} * [12 + 3 * m_i * [2 * \emptyset \cos(\emptyset) - 2 * \sin(\emptyset)] - 3 * m_i \pi * \cos(\emptyset)] + r_D * I_o * [3\pi - 4 * m_i * (1 + \cos^2(\emptyset))]) \quad (3-32)$$

IGBT switching loss equation for 3L T-type inverter:

IGBT 1&4:

$$P_{T1\&4SW} = (E_{Ton(N)} + E_{Toff(N)}) * f_{sw} * \frac{I_o}{I_N} * \frac{V_{DC-link}}{V_{N(T1\&4)}} * \left(\frac{1}{2\pi} * [1 + \cos(\phi)] \right) \quad (3-33)$$

IGBT 2&3:

$$P_{T2\&3SW} = (E_{Ton(N)} + E_{Toff(N)}) * f_{sw} * \frac{I_o}{I_N} * \frac{(0.5 * V_{DC-link})}{V_{N(T2\&3)}} * \left(\frac{1}{2\pi} * [1 - \cos(\phi)] \right) \quad (3-34)$$

Diode 1&4:

$$P_{D1\&2sw} = E_{REC(N)} * f_{sw} * \frac{I_o}{I_N} * \frac{V_{DC-link}}{V_{N(D1\&4)}} * \left(\frac{1}{2\pi} * [1 - \cos(\phi)] \right) \quad (3-35)$$

Diode 2&3:

$$P_{D2\&3sw} = E_{REC(N)} * f_{sw} * \frac{I_o}{I_N} * \frac{(0.5 * V_{DC-link})}{V_{N(D2\&3)}} * \left(\frac{1}{2\pi} * [1 + \cos(\phi)] \right) \quad (3-36)$$

Commercially available candidate power semiconductors evaluated for 3L NPC & 3L T-type inverters are given in Table 3.7 and Table 3.8.

Table 3.7 Technical Specifications of Candidate Semiconductors for NPC and T-type

Ratings	Semiconductor Modules for NPC and T-type		
	Si-IGBT Infineon FF300R12KE4	Si-IGBT Mitsubishi CM225DX-24T	Si IGBT Fuji Electric 2MBI225VN-120-50
Max. Blocking Voltage (V)	1200	1200	1200
Continues Current (A) @ 125°C	300	225	225
$V_{ce(sat)}$ (V)	1.75	1.55	2.05
Total Gate Charge (nC)	2500	1700	1500
Recovery Charge (μ C) @125°C	56	59	60
Max. Junction Operating Temp. (°C)	150	150	150
Base Area (mm ²)	61.4x106.4	62x152	62x150
Thermal Resistance per IGBT Junction-to-case (K/kW)	93	102	140
Thermal Resistance per Diode Junction-to-case (K/kW)	150	150	190

Simulation software of IPOSIM, Melcosim, and Fuji IGBT Simulator are used in performance calculations respectively of FF300R12KE4, CM225DX-24T, and 2MBI225VN-120-50 for 5 kHz switching frequency and 750 V DC-link. The power loss analysis of the 3L NPC inverter is summarized in Table 3.9. Three candidate power semiconductors are used in calculations. The allowable junction temperature limit is taken as 120°C for calculations in order to increase the lifetime of the power semiconductor. The results show that CM225DX-24T is rated as the best in terms of temperature and loss. There-level NPC voltage source inverter switching frequency can be increased up to 5000 Hz.

Table 3.8 Technical Specifications of Candidate Semiconductors for T-type

	Semiconductor Modules for T-type only		
Ratings	Si-IGBT Infineon FF300R17KE4	Si-IGBT Mitsubishi CM225DX-34T	Si-IGBT Fuji Electric 2MBI225VX-170-50
Max. Blocking Voltage (V)	1700	1700	1200
Continues Current (A) @ 125°C	300	225	225
$V_{ce(sat)}$ (V)	1.95	2.00	2.35
Total Gate Charge (nC)	3050	1760	2000
Recovery Charge (μC) @125°C	105	102	100
Max. Junction Operating Temp. (°C)	150	150	150
Base Area (mm ²)	61.4x106.4	62x152	62x150
Thermal Resistance per IGBT Junction-to-case (K/kW)	83	104	120
Thermal Resistance per Diode Junction-to-case (K/kW)	130	176	200

In order to calculate the power loss components for 3L T-type inverter, simulation software of IPOSIM, Melcosim, and Fuji IGBT Simulator, and the mathematical calculations are used in performance calculations respectively of FF300R17KE4, FF300R12KE4, CM225DX-34T, CM225DX-24T, 2MBI225VX-170-50 and 2MBI225VN-120-50 for 5 kHz switching frequency and 750 V DC-link. The power loss analysis of the 3L T-type inverter is summarized in Table 3.10. Three pairs of candidate power semiconductor are used in calculations. The allowable junction temperature limit is again taken as 120°C for calculations in order to increase the lifetime of the power semiconductor. The results show that CM225DX-34T and CM225DX-24T pair are rated as the best in terms of total semiconductor loss. However, there-level T-type voltage source inverter switching frequency cannot be increased up to 5000 Hz due to the thermal design requirements.

Table 3.9 Power Loss Analysis for 3L NPC Inverter at 750 VDC, 165 kVA

		Semiconductor Modules for 3L NPC Inverter					
Loss Components @ 5 kHz		Si IGBT Infineon FF300R12KE4		Si IGBT Mitsubishi CM225DX-24T		Si IGBT Fuji Electric 2MBI225VN-120-50	
IGBT 1,4 Conduction Loss (W)	Tvj (peak) (°C)	97.95	119.43	97.08	120.71	123.7	131.4
IGBT 1,4 Switching Loss (W)		54.84		53.14		42.7	
Diode 1,4 Conduction Loss (W)	Tvj (peak) (°C)	2.32	100.56	1.43	101.26	1.7	102.3
Diode 1,4 Switching Loss (W)		4.49		2.58		3.1	
IGBT 2,3 Conduction Loss (W)	Tvj (peak) (°C)	147.87	120.85	153.77	121.56	196.6	137.5
IGBT 2,3 Switching Loss (W)		8.59		5.46		4.1	
Diode 2,3 Conduction Loss (W)	Tvj (peak) (°C)	2.32	100.85	1.43	101.46	1.7	102.6
Diode 2,3 Switching Loss (W)		0		0.0		0.0	
Diode 5,6 Conduction Loss (W)	Tvj (peak) (°C)	44.29	109.91	49.0	108.43	60.0	117.3
Diode 5,6 Switching Loss (W)		25.6		19.01		20.3	
Total 3L NPC Inv. Loss (W)		2325,3		2297.5		2723.9	

Table 3.10 Power Loss Analysis for 3L T-type Inverter at 750 VDC, 165 kVA

		Semiconductor Modules for 3L T-type Inverter					
Loss Components @ 5 kHz		Si IGBT Infineon FF300R17KE4, FF300R12KE4		Si IGBT Mitsubishi CM225DX-34T, CM225DX-24T		Si IGBT Fuji Electric 2MBI225VX-170-50, 2MBI225VN-120-50	
IGBT 1,4 Conduction Loss (W)	Tvj (peak) (°C)	100.05	124.99	130.46	126.28	137.3	141.8
IGBT 1,4 Switching Loss (W)		114.84		66.60		108.1	
Diode 1,4 Conduction Loss (W)	Tvj (peak) (°C)	2.30	106.36	2.34	100.32	2.8	105.8
Diode 1,4 Switching Loss (W)		6.34		2.65		5.2	
IGBT 2,3 Conduction Loss (W)	Tvj (peak) (°C)	49.95	102.20	59.75	103.70	68.4	98.9
IGBT 2,3 Switching Loss (W)		7.45		5.55		3.8	
Diode 2,3 Conduction Loss (W)	Tvj (peak) (°C)	44.44	109.89	63.29	108.66	59.5	105.1
Diode 2,3 Switching Loss (W)		25.39		18		19.2	
Total 3L T-type Inv. Loss (W)		2104.56		2091.84		2425,8	

3.4 Overall System Efficiency

In this thesis, the efficiency of the traction motor drive is maximized by switching the traction inverter at the optimum frequency. Also, the operating range of the virtual junction temperature of the SiC power MOSFET is affected by the switching frequency. Lower junction temperatures lead to significantly longer lifetimes for power semiconductors. The overall efficiency of the traction motor drive is the multiplication of traction inverter efficiency and motor efficiency. The main power loss components are conduction and switching losses for the inverter and copper and core loss for the traction motor. These power loss components are determined by tests as well as some detailed simulations. To determine the power loss components of the traction motor, it is driven by the loading machine at synchronous speed ($n_r=1800$ rpm) corresponding to the rated frequency ($f_s=60$ Hz), the rated voltage is applied ($V_s=480$ V l-to-l), and the stator line current and stator power input are measured as a function of the inverter switching frequency. Core loss is separated from stator copper loss using the stator resistance provided by the motor manufacturer as a function of the drive switching frequency and stator temperature, and line currents. Core loss includes hysteresis loss, eddy current loss, and other mechanical loss components. The hysteresis losses increase with an increase in AC flux density, B_{AC} and operating and switching frequency. The general form of the hysteresis and eddy current loss equations are given in (3-37) and (3-38).

$$P_{hys} = kf^\alpha(B_{AC})^\beta \quad (3-37)$$

where k , α , and β are constants that depend on the core material. B_{AC} contains two components: i) Low-frequency component @ motor stator fundamental frequency ii) High-frequency component @ switching frequency. The high-frequency components are decreased a considerable amount with the increased switching frequency.

$$P_{eddy} = \frac{d^2w^2B^2}{24\rho_{core}} \quad (3-38)$$

where d is the lamination thickness, ω angular frequency,

In addition, the friction and windage loss component is measured after all rotating parts have been separated from the motor shaft. In Figure 3.9, the changes in core loss and stator copper loss components at synchronous speed are illustrated as a function of switching frequency.

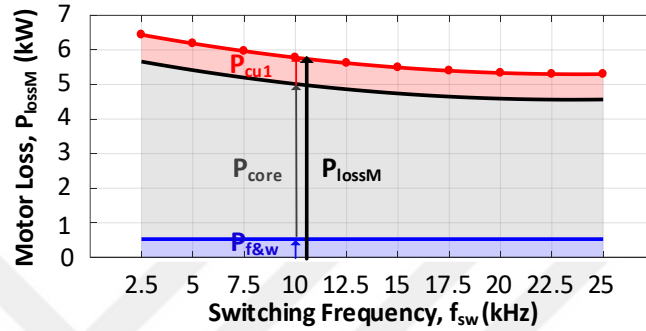


Figure 3.9. Variations in total power loss components of the traction motor at no-load against switching frequency of the inverter, when $V_{dc}=750$ V, $V_s=480$ V l-to-l, $f_s=60$ Hz, and $n_r=1800$ rpm

The conduction and switching loss components of the all SiC power MOSFET-based traction inverter are calculated with Wolfspeed's SpeedFit Design Simulator as a function of the switching frequency given in Figure 3.10. According to simulation results presented in previous sections, the minimum switching frequency was calculated as 8 kHz in order to satisfy the peak-to-peak torque ripple requirement. Therefore, 8 kHz or higher switching frequency should be selected to satisfy the design requirements.

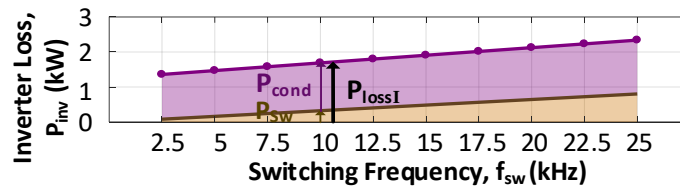


Figure 3.10. Variations in total power loss components of the traction inverter against switching frequency when $V_{dc}=750$ V, $V_s=480$ V l-to-l, $f_s=60$ Hz, $n_r=1779$ rpm in delivering 165 kVA to the motor at 0.81 pf lag

In order to determine the optimum switching frequency, the overall efficiency of the traction motor drive is calculated as a function of switching frequency. The total power loss of the traction motor drive is obtained by considering the motor losses and the inverter losses against different switching frequencies. The stator and rotor copper loss components are calculated considering the current harmonics at rated operating conditions and then added to the core loss, friction, and windage loss components of the motor and the total losses of the inverter for each switching frequency. Variations in total motor drive loss as a function of switching frequency are then plotted as in Figure 3.11. A detailed examination of this curve shows that the total power loss is minimized in the switching frequency range from 10 to 17.5 kHz. Although 15 kHz switching frequency causes minimum total power loss of the traction motor drive, it causes a higher peak junction temperature during operation, and the contribution of the total efficiency between 10 kHz to 15 kHz is negligible. By considering the junction temperature of MOSFETs for longer lifetimes, 10 kHz was chosen as the optimum switching frequency. The power loss components at the selected switching frequency of 10 kHz will then be verified by tests.

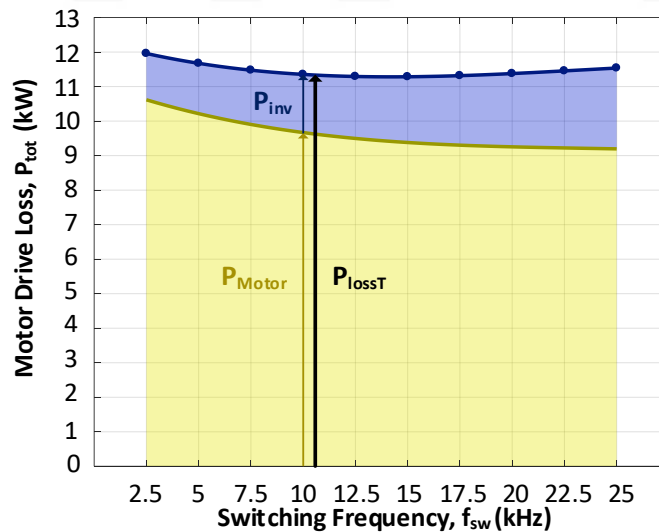


Figure 3.11. Variations in total power loss components of the traction motor drive against switching frequency when $V_{dc}=750$ V, $V_s=480$ V l-to-l, $f_s=60$ Hz, $n_r=1779$ rpm in delivering 125 kW to its load

3.4.1 Loss Comparison of Two-Level & Three-Level Converter Topologies

Improvement of output waveform quality and output power increase are the advantages well-known advantages of multilevel converters. Improvements in output waveform quality were already explained in Chapter 2, and later it will be examined in detail in Chapter 4. In this section, the loss, cost, volume, and reliability comparisons are briefly examined.

The power loss analysis shows that Si-IGBT-based two-level inverter topology can survive at switching frequencies around 3.5 kHz at most. Above this switching frequency, the junction temperatures rise above 120 °C with full Si-IGBT-based solutions. Thus, full Si-based IGBT modules do not satisfy the design requirements. Together with Si-IGBT modules, a hybrid IGBT module and full SiC MOSFET module are also taken into account in power loss calculations. The analysis results show that even if the hybrid IGBT module lets to increase switching frequency considerable amount, it does not still satisfy the design requirements. Only the full SiC MOSFET-based two-level solution has shown great success in satisfying the requirements with two-level inverter topology. On the other hand, Si-IGBT-based three-level T-type and NPC inverter topologies also provide great advantages to meet design requirements. The summary of power loss analysis, the unit cost of power electronic blocks, and the total semiconductor areas are given in Table 3.11. In order to make a fair comparison, the effective switching frequencies are also set to 10 kHz (i.e., 5 kHz switching frequency) for three-level converters. In Table 3.11, only products showing the best performance results are listed.

Table 3.11 Comparison of Different Inverter Topologies In Terms of Losses, Unit Costs and Semiconductor Base Areas

		Topologies					
Loss Components		SiC MOSFET based Two-level Inverter @ fsw = 10 kHz CAS300M17BM2		Si IGBT based Three-level T-Type Inverter @ fsw = 5 kHz CM225DX-34T&CM225DX-24T		Si IGBT based Three-level NPC Inverter @ fsw = 5 kHz CM225DX-24T	
Total Switch Loss (W)	Tv _j (peak) (°C)	1767.00	120.2	1574.16	126.28	1856.7	121.56
Total Diode Loss (W)		324.00		517.68		440.8	
Total Power Loss (W)		2091.00		2091.84		2297.5	
Cost for Power Electronic Block (Unit)		1.024		1.000		1.048	
Semiconductor Area (Unit)		1		2.89		4.33	

Although the same total power losses are obtained with the SiC MOSFET-based two-level inverter solution and Si-IGBT-based three-level T-type topology, the maximum junction temperature target is not satisfied with T-type topology. On the other hand, among the converter topologies, three-level NPC topology has the highest total power loss. If we compare the topologies in terms of power block costs, the T-type inverter offers the cheapest solution. The power block costs contain laminated busbars, heatsink, gate-driver, semiconductor, DC-link capacitor, phase current sensor, and cable prices. However, the difference between power block costs is too small to be considered in the total converter cost. It is seen that the biggest difference between the converter topologies is the semiconductor area. Hence, it can be said that the smallest and the most efficient topology that meets the requirements is the SiC MOSFET-based two-level inverter. Moreover, another issue that needs to be compared is reliability. The two-level solution is also very advantageous in terms of reliability due to the low number of components and the ability to withstand higher thermal stresses. If we make an additional comparison between the three-level converter topologies, it is seen that the T-type topology, where the number of serial switches is less, is more advantageous at low voltage and power levels and relatively low-frequency applications. The three-level NPC inverter is superior for relatively high switching frequencies.

3.5 Selection of Input Filter Components

Most of the railway DC electrification systems consist of six or twelve pulse bridge rectifiers, as mentioned in Chapter 1. These rectifiers inject low-frequency voltage ripple on the third rail or catenary line. Besides this, high-frequency harmonics are imposed on the third rail or catenary line due to the switching behavior of the traction inverter. In addition to this, the third rail or catenary voltage can be interrupted or discontinuous in normal operation due to some physical difficulties. This can cause voltage spikes due to the sudden change of highly demanded current and parasitic inductances of the power lines. This can end up with semiconductor failure. Moreover, voltage dips can also be seen at the electrification line due to some electrical and mechanical failures. This situation can also cause huge current flow from traction inverter DC-link capacitors to electrification lines. In order to eliminate all these problems, an LC filter with a low cut-off frequency is needed at the input stage of the traction inverter. In these applications, the cut-off frequency is selected below the grid frequency [24].

Following criteria should be taken into account for the selection of input filter components;

- The DC-link capacitance should be enough to reach the target voltage ripple,
- The DC-link capacitor should have enough RMS current carrying capacity,
- The laminated busbar structure with DC-link capacitor should provide sufficiently low parasitic inductance and resistance to filter out high-frequency harmonics and prevent switching spikes,
- The cut-off frequency of the input filter should be appropriate for harmonics produced from six and twelve pulse rectifiers,
- The line reactor should have sufficiently high inductance to limit current under failure instant (i.e., short circuit failures),

- The line reactor should have sufficiently high inductance to limit current ripple,
- The line reactor inductance value should be sufficiently low, and DC-link capacitance should be sufficiently high to let needed transient response,
- The size, weight, and cost optimization of both input filter components should be considered.

Resonance frequency f_0 is set to 32.8 Hz, which is sufficiently lower than 50 or 60 Hz grid frequency and sufficiently higher than 10 Hz as recommended in the literature [24], [95], [96]. The transfer characteristic of the resulting input filter is plotted in Figure 3.12. Unity gain is obtained at $f_{tr} = 47.5$ Hz for the input filter.

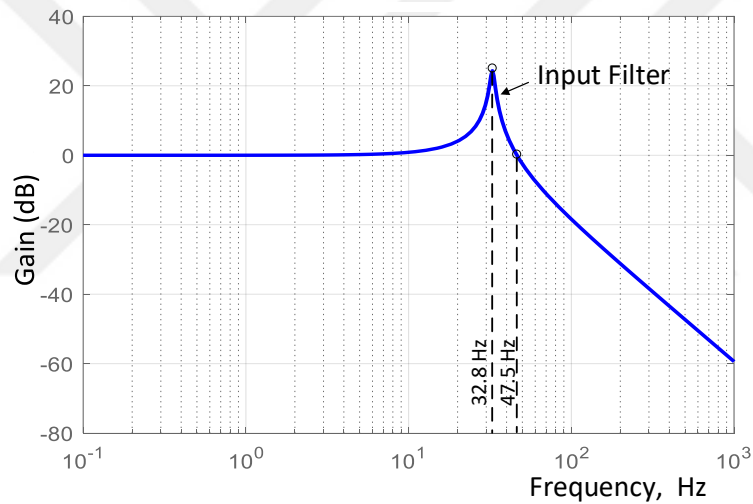


Figure 3.12. Transfer characteristics of input power filter

3.5.1 DC-link Reactor

The line reactor can consist of air-core or iron core materials. Although the iron core line reactors usually offer size advantages, the air core reactors are mostly used in railway power converter applications due to fully linear magnetic characteristics at high currents. This feature is very important at the short circuit failure instants.

The functional specifications of the DC-link reactor of traction inverter are listed below:

- It provides filtering functionality between the catenary line and the traction inverter,
- It limits the short circuit currents,
- It provides protection for semiconductors against the voltage spikes build-up at the catenary line,
- It reduces the capacitor RMS current due to the voltage ripple at the catenary line.

In order to determine the inductance value of the line reactor, the current ripple requirements have to be considered firstly. The peak-to-peak current ripple target in this study is to keep it below 10 %. In order to calculate the inductance value, the voltage variation of the line reactor is determined with the simulation study. In order to obtain a suitable design for both six and twelve pulse diode bridge rectifiers, the design requirements were determined by using the specifications of six pulse diode rectifiers. While performing this simulation, the electrification line is loaded with the full train power to obtain similar variations as in reality, and the DC-link voltage ripple was kept below 1 %. In Figure 3.13, the voltage variation on the line reactor is given.

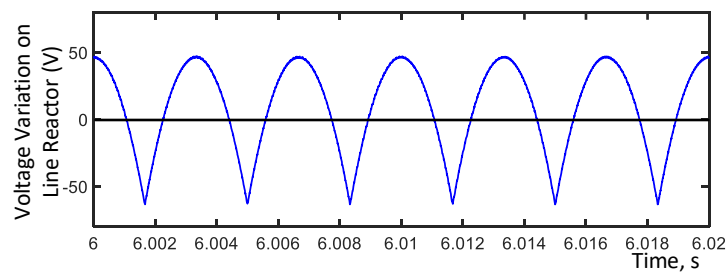


Figure 3.13. Voltage variation on DC-link reactor

In order to roughly calculate the minimum inductance value, (3-39) can be used.

$$L_{dc} = \frac{\Delta V * dt}{\Delta i} \quad (3-39)$$

In rated operation, nearly 185A is supplied through the line reactor. 10 % of it equals 18.5A. From there, the minimum inductance value can be computed as 2.84 mH. As a second target, the short circuit current has to be checked. In order to decrease the failure effect at the instant of the short circuit, the opening time of the circuit breaker should be kept as short as possible. In this study, opening time is taken as less than 10ms. In order to open circuit breaker in 10ms, it needs a di/dt range 10^5 to $2 * 10^5$ [97]. Hence, the inductance value is determined as 4 mH. Using this value, the line reactor peak-to-peak current ripple is obtained as in Figure 3.14. It satisfies the input current ripple requirement. In the next section, the DC-link capacitance value will be determined. Considering the input filter requirements, an iterative process will be required to arrive at appropriate L_{dc} and C_{dc} values. In this case, this step has to be repeated.

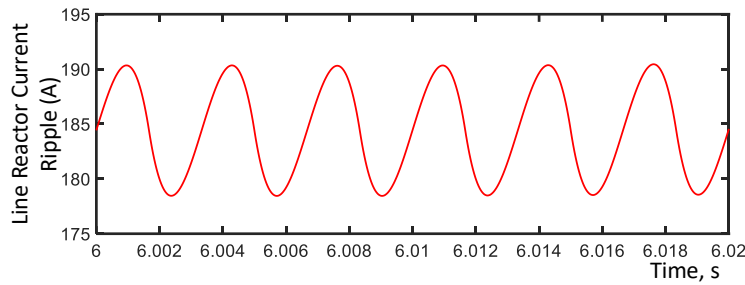


Figure 3.14. DC-link reactor current

Front and side views of the DC-link reactor are given in Figure 3.15 (a) and (b).

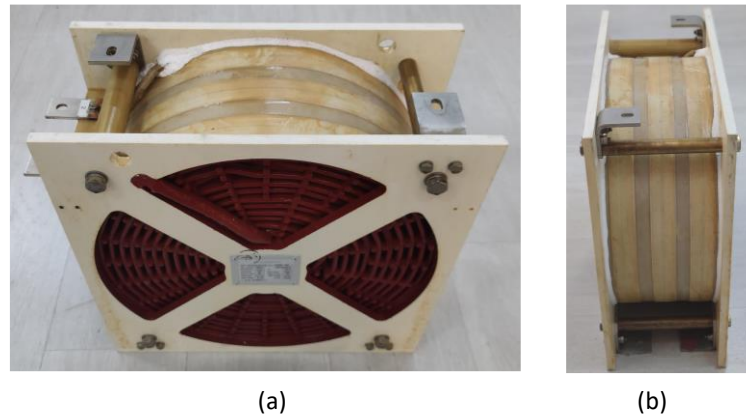


Figure 3.15. DC-link reactor (a) Top-front view (b) Side view

3.5.2 DC-Link Capacitor

DC-link capacitors are also one of the most critical components in a traction inverter. Their dimensions are usually large in size, and they directly affect inverter power density and reliability. To increase power density, the capacitance value should be optimized by considering the input filter requirements. The capacitor selection is mainly determined by the RMS current carrying capacity of capacitors, the capacity of the stored energy, and the tolerable voltage ripple on the DC-link. The RMS currents supplied from the DC-link capacitors cause to increase in the internal temperature of capacitors [97]. In order to increase lifetime and reliability, the internal temperature of capacitors should be kept as low as possible.

As previously mentioned, railway electrification lines consist of diode bridge rectifiers. This means that unidirectional power flow can be possible between the rail vehicle and the electrification line. Besides this, while one rail car is braking and the other rail car is accelerating at the same time, power flow between these two cars can be possible. However, this situation will not always be possible. During the design phase, the worst cases should be considered as design requirements. Therefore, during the braking phase, due to the nature of regenerative braking, line voltage and the DC-link voltage will increase. EN 50163 standard defines the permissible voltage variations of standard railway electrification lines. Actually, a crowbar circuit is used

to limit the voltage increase. However, there should be a voltage margin to enable this circuit. This voltage limit should be out of normal working voltage ranges. This means the activation voltage of the crowbar circuit should be higher than the upper level of nominal working voltage. In this study, 910 VDC was set for crowbar circuit activation. This fact should also be considered during the design phase. This also has to be taken into account for the selection of DC-link capacitors.

Following criteria should be considered for the selection of DC-link capacitors;

- Nominal and peak voltages considering with the exposure times,
- Harmonic current ratings in steady-state and transient operations,
- Lifetime,
- Thermal stresses,
- DC-link voltage ripple due to railway electrification line voltage ripple and high frequency ripple current caused from switching nature of the traction inverter considering the reactive power need,
- Capacitor energy has to sustain output power demand during short periods at the instant of energy interruptions,
- The capacitance value should be enough for expected dynamic load variations (e.g., At the emergency stop condition, the torque is immediately set to zero, and all power switches are made off within a very short time period. The stored energy on the line reactor or at the braking instant the stored energy on the stator windings flows through the DC-link capacitor).

In previous sections, L_{dc} and resonance frequency f_0 were determined as 4 mH and 32.8 Hz, respectively. Using (3-40), C_{dc} is computed as 5.9 mF.

$$C_{dc} = 1/[(2\pi f_0)^2 L_{dc}] \quad (3-40)$$

By using these L_{dc} and C_{dc} values, simulations are performed again, and DC-link voltage ripple and DC-link capacitor current ripple values are investigated. Simulation results are plotted in Figure 3.16 and Figure 3.17. Because of the LC

input filter requirements, C_{dc} was calculated quite high, and DC-link voltage ripple is obtained near to 0.5 %. In motor drive applications, voltage ripple is generally taken as 1 %. In Figure 3.18, the DC-link capacitor current harmonic spectrum is given. From this simulation study, the worst case for the RMS current ripple is obtained as 120 Arms considering the all acceleration and deceleration period. This graph gives information about the capacitor currents and their frequencies which are used for the DC-link capacitor selection. In capacitor selection, all current components appearing in the spectrum analysis have to be taken into account for the capacitor's lifetime and natural resonance frequencies.

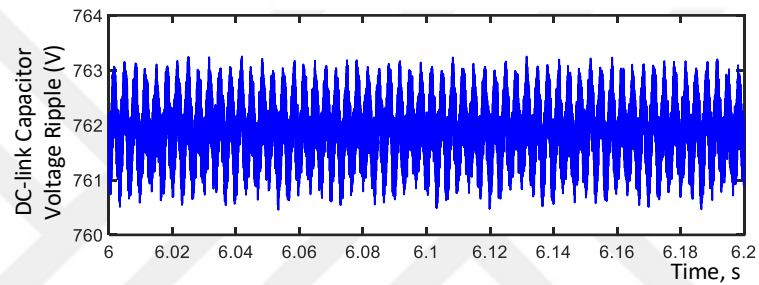


Figure 3.16. DC-link voltage variation

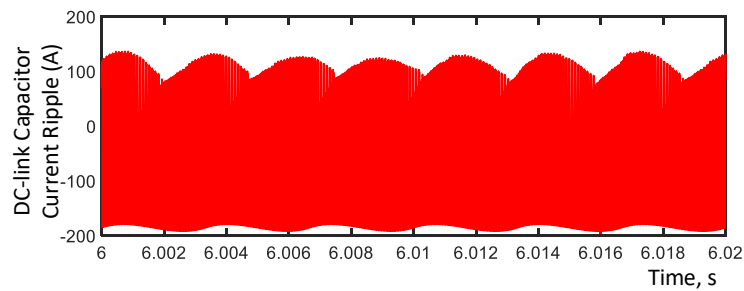


Figure 3.17. DC-link capacitor current ripple

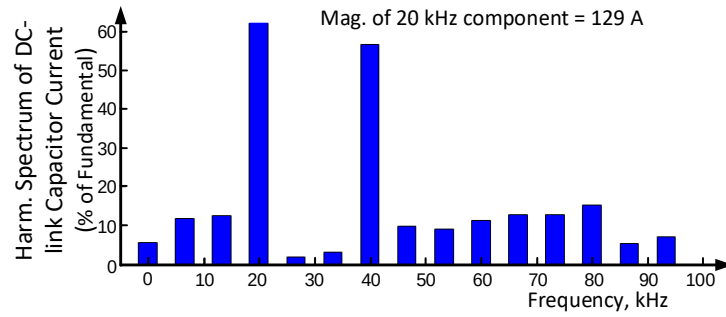


Figure 3.18. DC-link capacitor current harmonics with SVPWM when the induction machine is operated at 165kVA

Al-electrolytic and metalized film power capacitors are usually used in power converters. Comparing the electrolytic capacitors, film power capacitors are giving a chance to have a wide application range, safe operation due to the nature of self-healing, high RMS current carrying capacity, long lifetime, low ESR and ESL, and strong overvoltage capability [98]. Due to these benefits, metalized film capacitors are preferred in DC-link.

Photographs of the selected DC-link capacitors are given in Figure 3.19 (a) and (b). On the left side, a bulky metalized film capacitor used to satisfy the DC-link capacitance requirement is seen. On the right side, very low ESL and ESR capacitors used on the laminated busbar to minimize the switching spikes are illustrated.

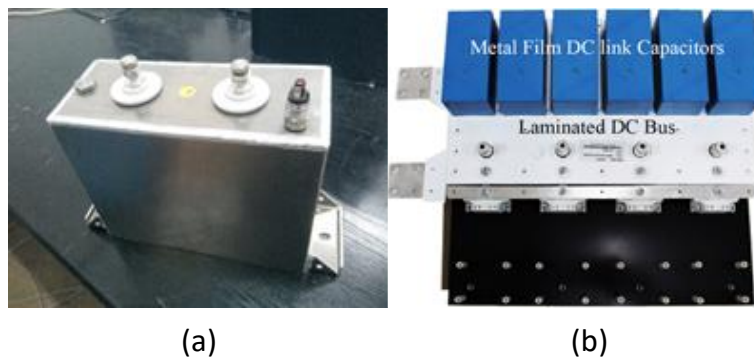


Figure 3.19. DC-link capacitors (a) 4.7mF metalized film capacitor (b) 6x200uF low ESL, and ESR metalized polypropylene film capacitors

3.5.3 Critical Design Approaches for Laminated Busbar

With the development of WBG devices, new stringent requirements emerged for laminated busbar technology. In fact, the busbar design was always very critical for inverter operation. However, now it is more critical than before. In order to decrease converter size, SiC power MOSFETs are used for the high voltage and high power products. Up to now, several specific module packages have been developed to decrease the stray inductances, increase the current-carrying capacities, and improve the heat transfer. However, similar requirements must also be taken into account in busbar design. Lower stray inductances let to use of faster switches, higher current carrying capacity together with improved heat strength and improved heat transfer let to produce smaller and powerful inverters. Nowadays, laminated busbar producers have started to offer new solutions for SiC-based power converters. The aims are to reduce the stray inductances by external laminated busbar with low inductive connections and to improve heat strength by changing the insulation and the resin/glue materials. In Figure 3.20, the approach of MERSEN company in material selection can be seen [99].

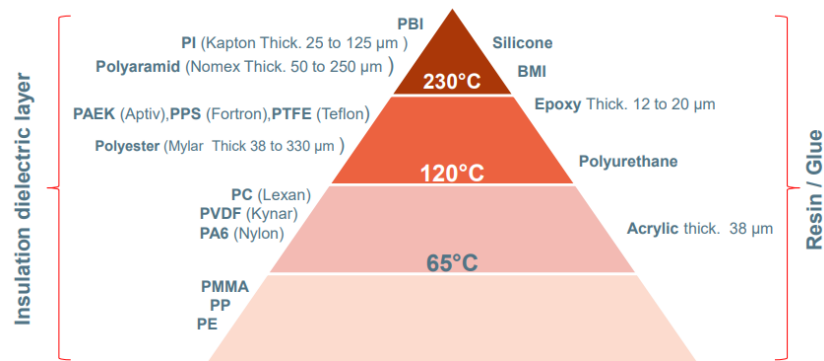


Figure 3.20. Mersen material solutions for SiC power electronics [99]

In this study, a low inductive laminated busbar structure was designed specifically for the CAS300M17BM2 SiC MOSFET module. By maximizing the overlap of metallic conductors, making laminated symmetrical structures, and providing low inductive connections for DC-link capacitor and SiC power MOSFETs, the stray inductances decreased as much as possible. In addition to this, 1.5 mm conductor thickness is used to decrease the transmission losses. A photograph of the designed low inductive

laminated busbar is given in Figure 3.21. The stray inductance of this laminated busbar between the drain and source terminals of the power MOSFETs with distributed and soldered capacitors (i.e., as seen in Figure 3.19) is measured as 20nH.

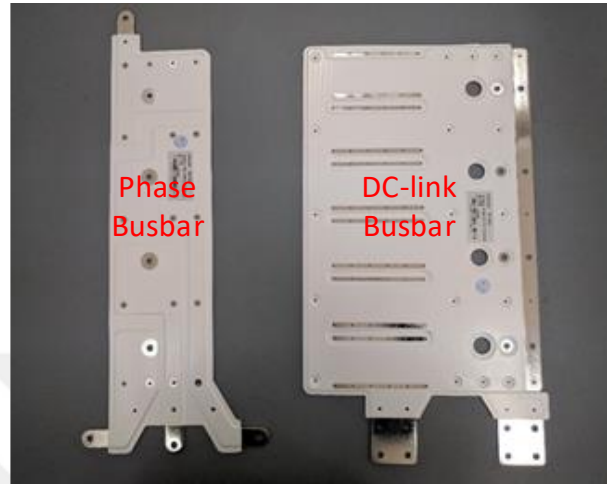


Figure 3.21. Low inductive laminated busbars

3.5.4 Heatsink

As in the laminated busbar design, the need for innovative solutions in heatsink/cold plate designs is increasing. Although SiC chips cannot operate at very high temperatures now, they promise to work at very high temperatures in the future. The point where technology has come these days is that high-power converters can be developed with small power modules. This brings with the challenge of transferring power losses from small spaces. As a solution to this challenge, embedded heat pipe solutions for air-cooled heatsinks began to be offered. In addition to this, optimized cold plate designs are constructed for liquid-cooled systems.

In this study, power losses were first calculated analytically for rated power operation. The total power losses per SiC MOSFET module are calculated as 697 W. Hence, the total losses for heatsink are equal to 3x697 W. For heatsink selection, maximum ambient temperature is taken as 45°C, and it is aimed to keep the SiC MOSFET junction temperature below 120 °C. Then, the selected heatsink is modeled

in a simulation environment so that transient effects such as overload can be examined. The steady-state thermal model of the SiC power MOSFET module and the power block assembly are given in the next section. In Figure 3.22, the thermal behavior of power block assembly is given for steady-state and short-time overload conditions. For short-time transient analysis, the vehicle duty cycle is taken as 90 %. In other words, 150 kW of full power (185 kVA) is applied during the 90 seconds, and power is off for 10 seconds. This cycle was continued until the temperature reached equilibrium.

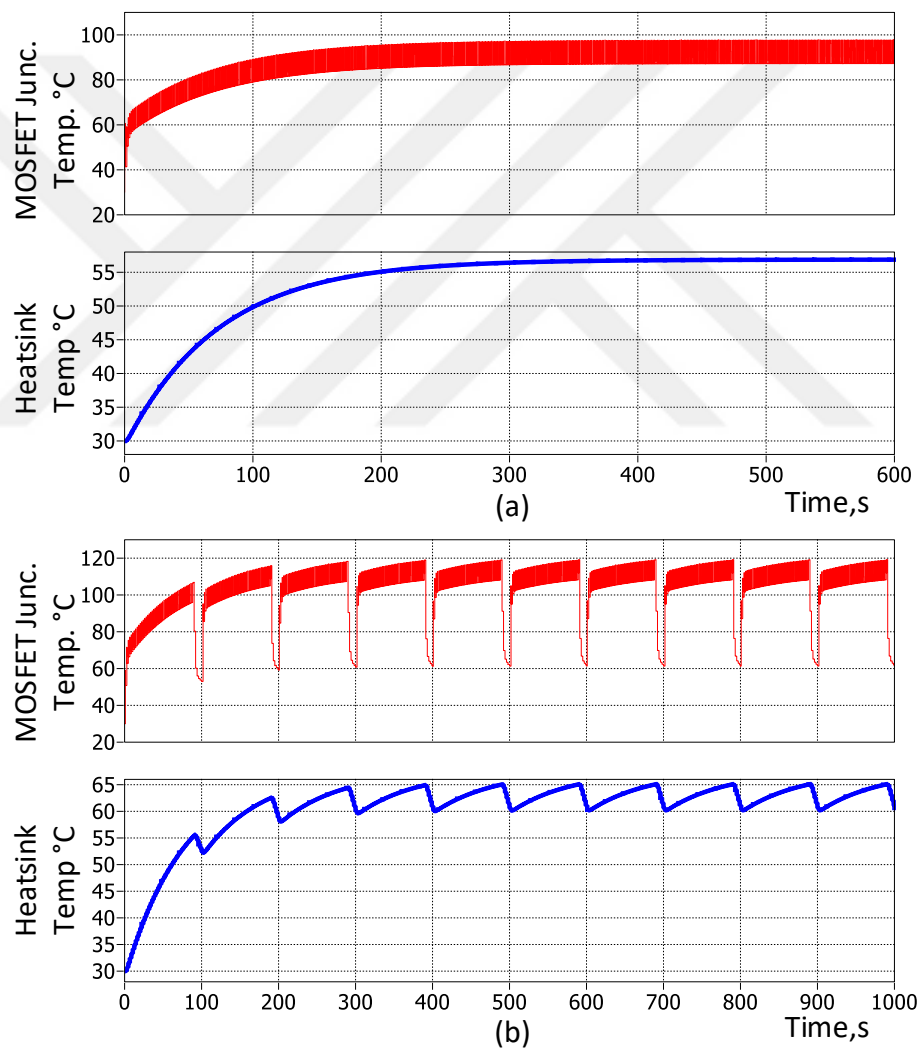


Figure 3.22. Thermal simulations of power electronic block (a) Steady-state performance (b) Transient performance

3.6 Traction Inverter Power Block Design

The layout of the SiC power MOSFET modules on the heatsink is shown in Figure 3.23. Each SiC power MOSFET module consists of serially connected to two MOSFETs and their antiparallel freewheeling diodes. The three SiC power MOSFET modules placed on the right side of the heatsink are used to build the two-level three-leg bridge traction inverter. The single module on the left is used as a brake chopper.

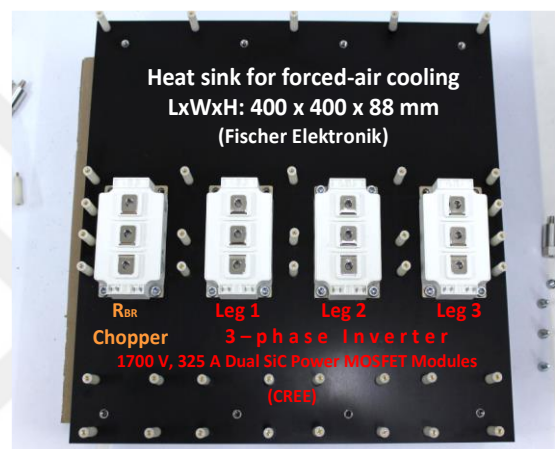


Figure 3.23. The bottom layer of the power electronic block

The thermal characteristics of the high-performance heatsink (Fischer SK 461 400 SA) block according to different lengths are given in Figure 3.24 [100]. It is seen that the most effective cooling will be provided at $v = 5$ m/s airflow and 350 mm heatsink length or more. The thermal resistance increases a considerable amount below this heatsink length. On the other hand, beyond this length, it is seen that an increase in the length does not have much effect on reducing the thermal resistance.

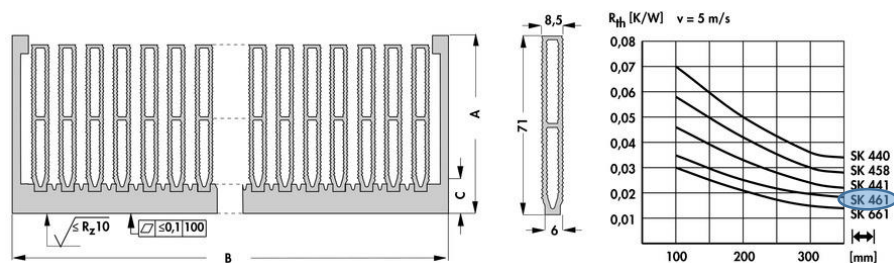


Figure 3.24. High-performance heatsink (Fischer SK 461 400 SA)

The thermal characteristics of the SiC power MOSFET module and the heatsink are given in Table 3.12.

Table 3.12 Thermal Specifications of the MOSFET Module and Heatsink

Maxium Ambient Temperature = 45 °C		
CAS300M17BM2 MOSFET die	T_{vjmax}	150 °C
	Safety limit in the design	30 K
	R_{j-c}	67 K/kW
CAS300M17BM2 diode die	T_{vjmax}	150 °C
	Safety limit in the design	30 K
	R_{j-c}	60 K/kW
CAS300M17BM2 Module	R_{c-h}	20 K/kW
Heatsink (Fisher – SK 461 400 SA)	R_{h-a} 40 cm x 40 cm with 5m/s airflow	18 K/kW

The heatsink thermal resistance is given as 18 K/kW for a 5m/s airflow. However, the fan used for forced cooling gives a higher airflow rate than 5 m/s. The real thermal resistance is calculated as 14 K/kW via the conducted tests in the laboratory. Steady-state thermal models of the MOSFET module and the power electronic block assembly are as given in Figure 3.25 and Figure 3.26. A maximum junction temperature of 120°C has been defined as a design requirement, aiming for the longevity of SiC power MOSFETs and diodes. The maximum junction temperatures of both MOSFET and diode chips are found less than 120°C (110°C for MOSFETs and 85°C for diodes) at rated operating conditions that satisfy the target design requirements.

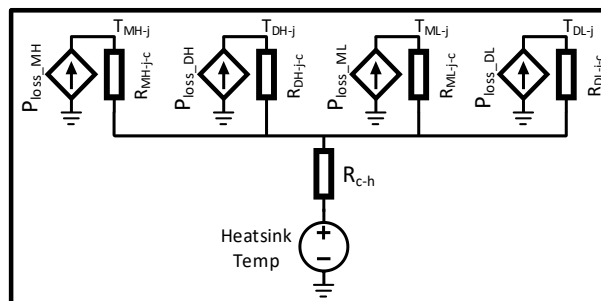


Figure 3.25. Steady-state thermal model of the MOSFET module

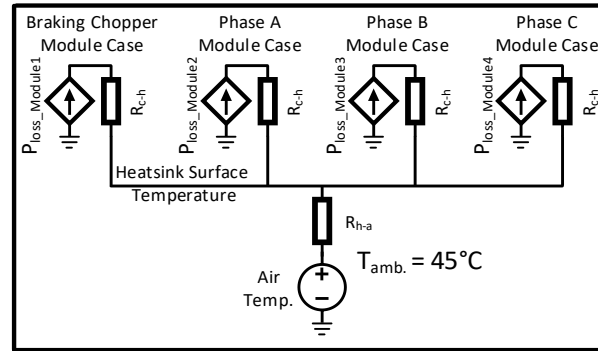


Figure 3.26. Steady-state thermal model of the power electronic block

These thermal models are good enough to calculate the steady-state thermal behavior of the developed power blocks. However, the transient cases should be investigated in the time domain, considering the thermal capacitances and the vehicle duty cycles. Because of this, the temperature variations in transient overload conditions are investigated in the simulation chapter considering the vehicle duty-cycles and rail track used in the design.

Since WBG devices have very fast turn-on and turn-off characteristics resulting in the generation of high di/dt and dv/dt at many points of the power circuitry, the parasitic capacitances and inductances are very important for the circuit design. Special attention should therefore be paid to the design of the power stage to minimize stray inductances and capacitances. The effects of parasitic inductance and capacitance can include voltage and current overshoots and ringing phenomena. The half-bridge structure of the SiC power MOSFET module provides flexibility for the assembly process, taking a leg-level optimized product from Cree/Wolfspeed company. In this way, the semiconductor connections are already optimized in a leg. In addition to this, a low inductive laminated busbar design is used while connecting each leg with other legs and the voltage source. Moreover, high-frequency metalized film capacitors are uniformly distributed on the laminated busbar to supply the switching currents with minimum inductance in the current paths. The total capacitance value of these high-frequency capacitors is calculated as 1.2 mF considering the high-frequency DC-link voltage ripples and capacitor currents. The

stray capacitances between the MOSFET module to heatsink and cables to chassis are tried to minimize to decrease the high-frequency common-mode currents. The view of this stage of assembly is given in Figure 3.27.

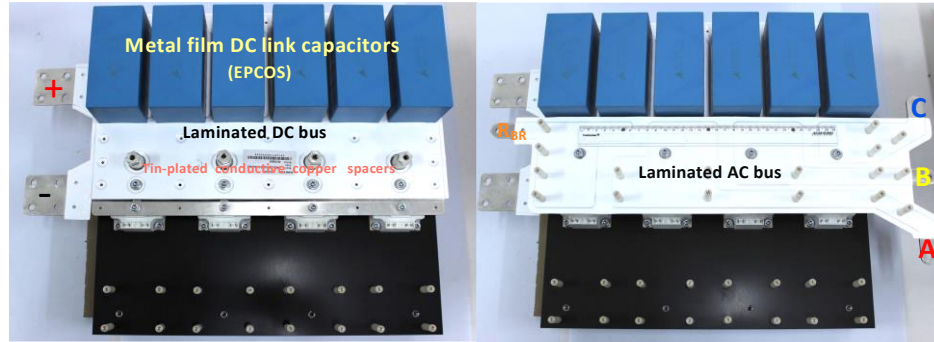


Figure 3.27. MOSFET and low inductive laminated busbar assembly

In the implementation, down-to-top assembly is preferred to minimize the volume and wiring. After this stage, isolated gate drivers are first mounted. Then the control platform is placed at the top, and cable connections are made. The control platform consists of a power board, a mainboard, and a controller board. A fiber optic interface is preferred to provide the connection between the control platform and the gate drivers. Thus, although the gate drivers have their own isolation barrier, an extra isolation barrier has been added, and hence an extra precaution has been taken against very high-frequency conductive emissions that may occur at the turned-on and turned-off instants. The final assembly of the power stage is given in Figure 3.28.



Figure 3.28. Final assembly of the power electronic block

3.7 Design of Control System

A piece of brief information about the railway traction drive systems has been presented in Chapter 2. The difficulties and the need for hybrid control methods and modulation techniques are also presented in the same chapter. Moreover, the effect of SiC-based semiconductors on the control methods and modulation techniques has been revealed. In this chapter, the implementation details of the selected control method and the protection concept will be explained. The control system of the traction inverter consists of the active damping controller for input filter resonance, motor controller, and modulation signal generator.

SiC power MOSFETs can be switched at higher switching frequencies than the Si-based IGBTs due to their low switching losses. Since continuous modulation methods can be applicable with high switching frequencies in all operating regions, a hybrid modulation technique is no longer needed. Instead of the hybrid modulation technique used in Si-IGBT-based drives, a more simple and common modulation method, SVPWM, is chosen for modulation signal generation. Similarly, as in the modulation method, hybrid motor control methods are no longer needed due to the high-frequency modulation ratio. For this reason, the FOC method, which is widely preferred in motor driver applications, has been preferred. PI controllers have been chosen to regulate the currents. The block diagram of the control system is given in Figure 3.29. In this diagram, there is a one-fast control loop to perform the motor current/torque control and active damping control functions. The d and q axes decoupling equations are omitted to avoid compression of the block diagram. The speed control function has been performed from the driver or communication-based train control system.

Hence, the torque command is given from an upper-level controller. In this control model, using the torque command, motor speed, motor line currents, and DC-link voltage, required modulation reference signals are constructed. Firstly, three-phase AC motor currents are transformed to the rotating frame to simplify the controller tasks by using the rotor speed and the slip. Then the converted current components

of torque and flux are subtracted from the set values and then fed to the PI controllers. The outputs of the PI controllers are the reference voltages for d and q axes. Again using the angle of rotation, d and q reference voltages are retransformed to rotating two orthogonal voltage references and then feed to the SVPWM generator.

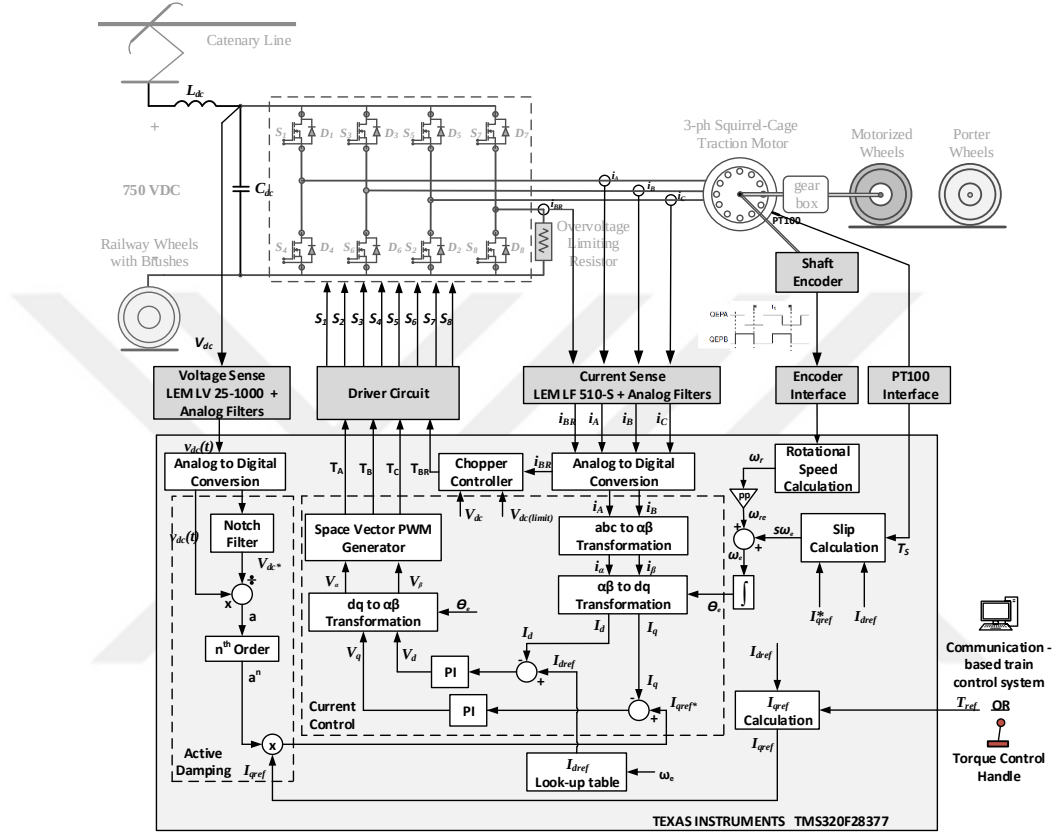


Figure 3.29. Block diagram of the digital control system with SVPWM, active damping, and field-oriented control

The control functions have been implemented on a TMS320F28377S microcontroller. This microcontroller has two separate central process units (CPUs). CPU 1 is used to perform the motor control functions and all protections against overcurrent, overvoltage, over-temperature, and over speed, while CPU 2 is used for CAN communication and system-level low-level tasks.

The optimum switching frequency is determined as 10 kHz, as explained in previous sections. In order to increase the bandwidth of the current controller, the modulation duty cycles are updated twice in this period. First, current, voltage, and angle

measurements are started at 20 μs before the bottom edge of the carrier signal, while second measurements are started at 20 μs before the upper edge. Hence, the sampling period of the controller is 50 μs . The loop-time of the control and protection tasks are less than 10 μs . The measurements and mathematical computations are completed in 20 μs , and at the duty cycle update instant, the most up-to-date modulation references are used.

In addition, an active damping control algorithm is also realized in CPU 1. As explained in the filter design section, the input filter consists of series L_{dc} and shunt C_{dc} and has a resonance frequency of 32.8 Hz. Nonlinearities, non-ideal operating conditions, and non-identical system parameters on both the catenary line and the traction inverter sides may produce inter-harmonics at or near the resonance frequency of 32 Hz and superimposed on V_{dc} . AC current through L_{dc} and AC voltage on C_{dc} will then be amplified significantly, resulting in destructive effects on system components. Therefore, resonance or amplification risks of the input filters should be avoided by adding a significant damping effect to the input filter. A conventional solution is the addition of a sufficiently large series resistance or a small shunt resistance to the circuit of the input filter. Unfortunately, these passive damping techniques are dissipative and, therefore, rarely used in modern systems. In this work, active damping, which is equivalent to connecting a virtual dissipative element, is implemented to suppress the generated uncharacteristic current harmonics at or around the resonance frequency via the control system and thereby the power circuit of the traction converter. The control system calculates the ratio of unfiltered, $v_{dc}(t)$ to filtered, $V_{dc}^*(t)$ dc bus voltage values ($a = v_{dc}(t)/V_{dc}^*(t)$). Filtered DC-link voltage signal, $V_{dc}^*(t)$ is obtained by using a digital notch filter, the transfer characteristic of which is as given in Figure 3.30. It provides an effective filtering performance in the frequency range from 16 Hz to 68.6 Hz. Active damping circuitry modifies the set value of I_q by defining new set value as in (3-41). “a” is a real number around unity, and its optimum power $n=11$ is found by simulations and verified experimentally.

$$I_{qref}^* = a^n I_{qref} \quad (3-41)$$

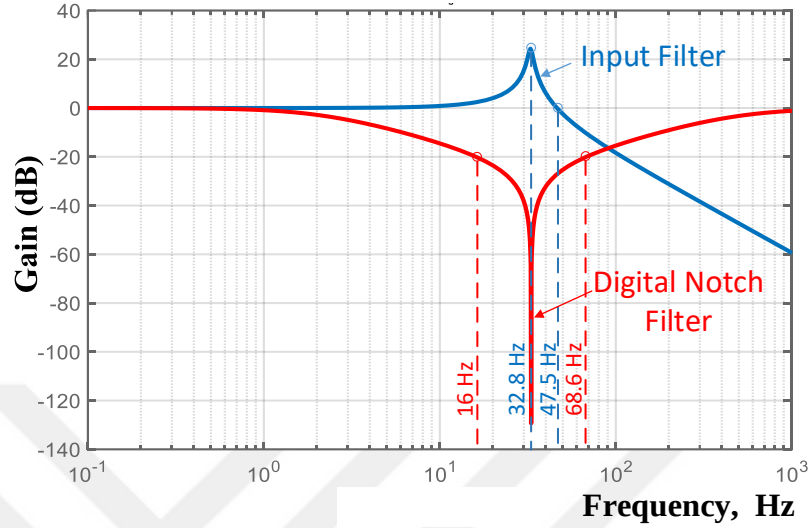


Figure 3.30. Transfer characteristics of input power filter and digital notch filter plotted against the same frequency range

Harmonic frequencies lower than 68.6 Hz are successfully suppressed by active filtering action via digital notch filter. Low order and high-frequency harmonics outside the frequency band defined above (e.g., 0 - 68.6 Hz), including switching frequency, its sidebands, and multiples, appearing on both catenary and traction sides, are successfully suppressed by the input filter.

CHAPTER 4

MODELING OF ELECTRIC TRACTION DRIVE SYSTEM

In this chapter, first, the LTspice model was created to investigate the switching performance of the SiC MOSFET Module CAS300M17BM2. Then, MATLAB/Simulink models for simulation of light rail transportation vehicle and its traction drive system are developed, and simulation results are presented. These two models are performed in separate simulation tasks because of the simulation sample time and duration needs. In the vehicle simulation model, track data, vehicle weight, wheel diameters, adhesion coefficient, train resistance, acceleration/deceleration rates, operating duty cycles are taken into account, and traction system requirements are analyzed. Besides this, in the traction drive system model, power electronic-based simulations are performed to analyze the system response together with the hardware and control system. The whole system model consists of a 6 and 12 pulse diode rectifier electrification system, a three-phase two-level voltage source rectifier, a three-phase two-level voltage source traction inverter, a three-phase asynchronous traction motor, and a control system. The electrical power system model is implemented by using Simulink/Simpower System Toolbox. Besides, the standard block sets of Simulink are used to construct the control system. These two different discipline can be easily incorporated. The separation between the control system and the converter power stage allows the user to simulate and analyze the complex systems easily. During the simulations, field-oriented control and V/Hz scalar control methods are utilized. The SVPWM modulation technique is implemented. The purpose of the simulations is to see the characteristics of the whole traction drive system and optimize it before the hardware design stage is completed. In addition to this, SDEM is also implemented to make an additional comparison. Finally, a PLECS model is developed to investigate the thermal variations of semiconductors

during overload conditions. In this section, developed models are explained in a simple way. The results will then be presented at the end of this section.

4.1 SiC Power MOSFET Model

The power semiconductor chosen for implementation is modeled in the LTspice environment in order to investigate switching transients and losses. Spice chip models (i.e., MOSFET and diode dies) and the module circuit schematics shared by Wolfspeed were used in the modeling study. Due to confidentiality, the LTspice CAS300M17BM2 MOSFET module model could not be presented in this thesis. By using this simulation model, turn-on and turn-off transitions, switching losses, and the effect of parasitic inductances were obtained. In Figure 4.1 (a) and (b), the turn-on current and voltage waveforms and turn-on power loss are given. The turn-on time is obtained nearly 90ns.

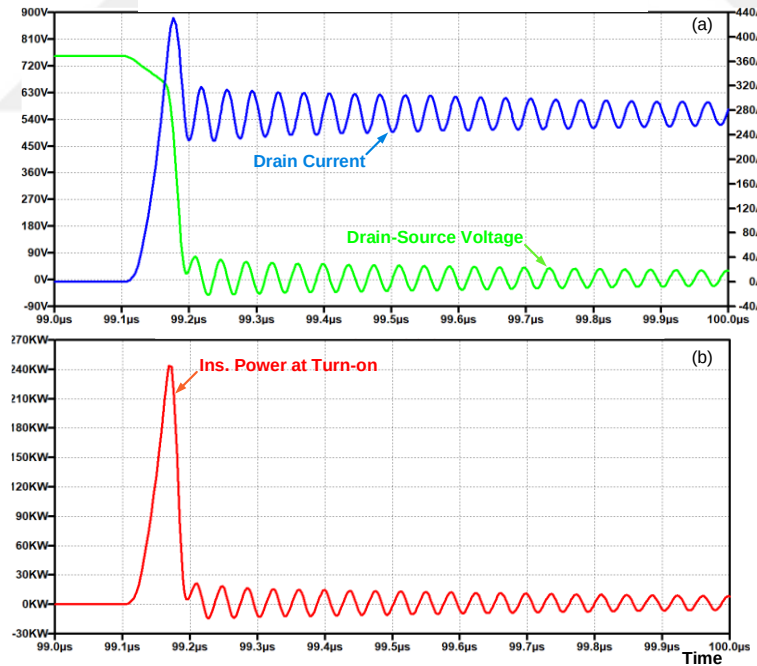


Figure 4.1. SiC MOSFET turn-on (a) Current and voltage waveforms (b) Power loss graph

Similarly, the turn-off current and voltage waveforms and turn-off power loss are obtained with a simulation model and presented in Figure 4.2 (a) and (b). The turn-off time is obtained nearly 70ns.

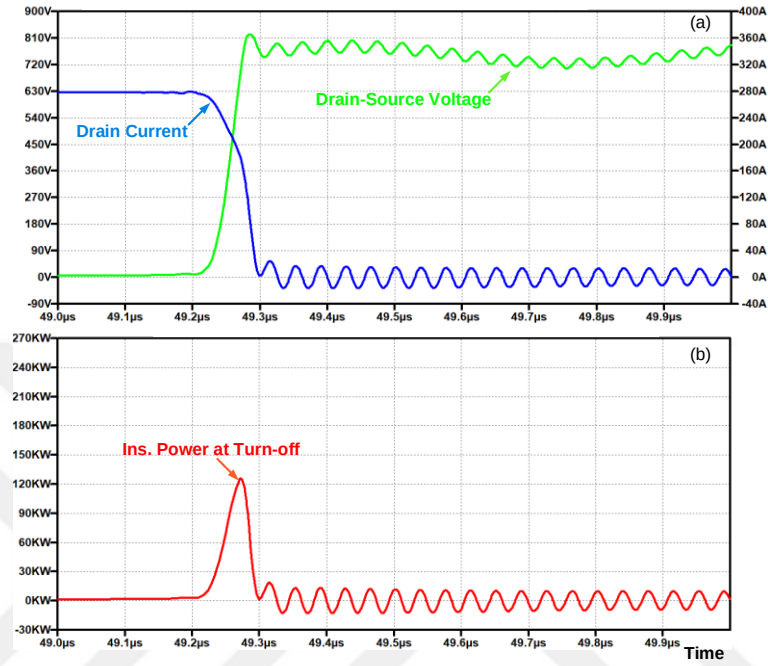


Figure 4.2. SiC MOSFET turn-off (a) Current and voltage waveforms (b) Power loss graph

4.2 Model of Rail Vehicle and Track

Modeling and simulation play an important role in the design, analysis, and evaluation of big and complex systems. System engineers can derive the requirements and optimize them according to the customer's needs. Since the main subject of this study is not traction analysis, the rail vehicle and track model will not be explained in detail. However, the main parts and their outputs will be briefly explained. This model has three main parts, namely, vehicle model, track model, and start/stop the algorithm. Firstly, related blocks of the vehicle model are given in Figure 4.3. Some of the mechanical properties of vehicles, such as mass, friction-related parameters, wheel diameter, and gearbox parameters, are taken into account

in this model. The outputs of this part are traction force, vehicle speed, acceleration/deceleration, train resistance, and travel distance.

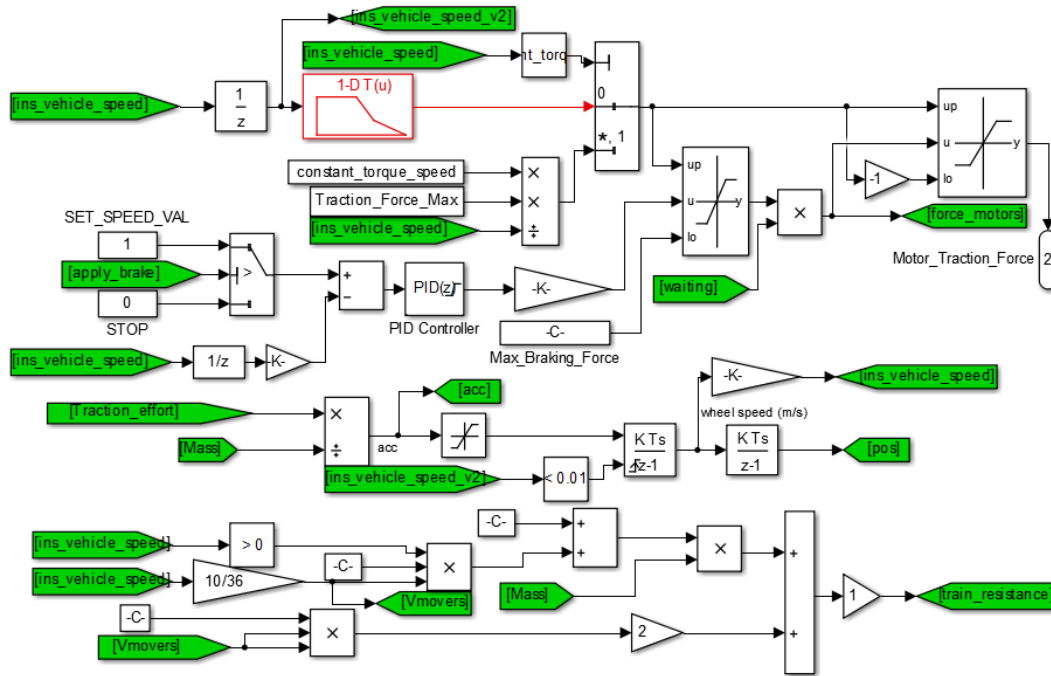


Figure 4.3. Vehicle model

Next, the track data model is given in Figure 4.4. In this model, route information, passenger stations, gradients, and forces caused by gravity are taking into account. In the last part, the start/stop algorithm is implemented. The block schema of the start/stop algorithm is given in Figure 4.5. This algorithm is the most complex part of this model. In order to stop heavy vehicles at the correct position in the passenger stations, the braking distance must be accurately estimated. Given that slope conditions change along the way, this calculation is quite complex and requires a recursive algorithm. This algorithm determined the correct braking instant to stop at the correct position in passenger stations.

advantages are also very important for traction applications. However, in most of the existing DC railway power substations, six or twelve pulse diode rectifiers are used, and this situation creates big disadvantages from the point of traction inverter. The typical power circuit of the twelve pulse diode rectifier is given in Figure 4.6. The effects of the power supply and the transmission line impedances are also taken into account.

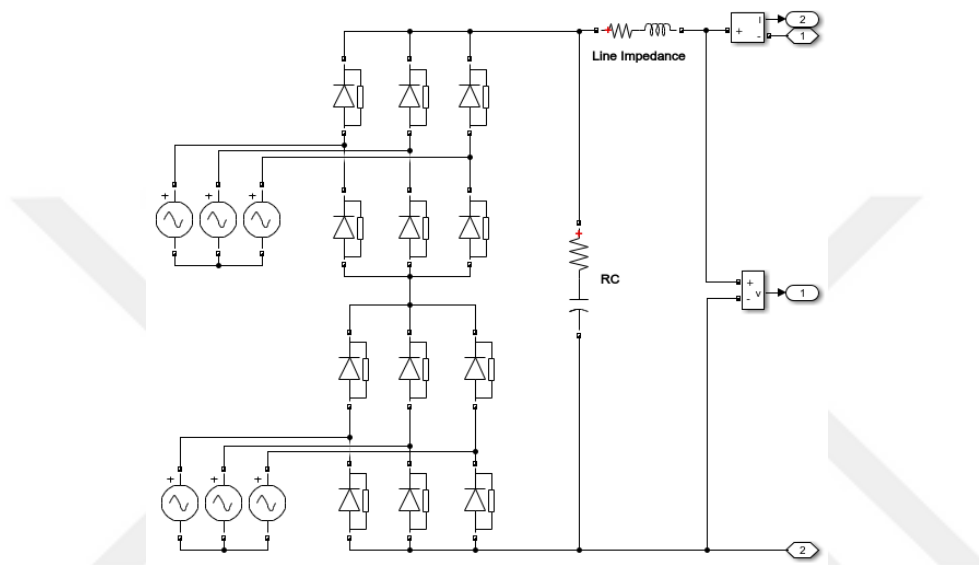


Figure 4.6. Twelve pulse rectifier power circuit

In the field, the catenary voltage fluctuates due to the load changes (i.e., acceleration and deceleration instants) and the nature of the diode rectifier. As defined in Chapter 1, the limits of variations are defined in standard EN 50163. In a simulation environment, it is very easy to implement such a power supply. However, in a small laboratory, the implementation of a diode rectifier and creating voltage fluctuations are very difficult. Moreover, the power grid has very limited power flow capability in the university laboratory. Due to these reasons, this power supply should be regenerative, and during the braking, the energy has to be fed back to the grid. In this way, a 150 kW traction system test bench was able to be integrated into a low-capacity energy supply infrastructure. In order to create similar variations as in the real catenary lines, instead of a diode rectifier, active pulse width modulated rectifier is developed and implemented in the laboratory. During the simulations, six and

twelve pulse diode rectifiers and active PWM rectifier are implemented to investigate the effects of different power supplies.

4.4 Model of Induction Machine

The induction machine model was directly used from Simulink/Power system library.

4.5 Model of Traction Inverter

In this thesis, a full SiC MOSFET-based two-level voltage source inverter was designed and implemented. However, in order to explain the position of the designed full SiC-based inverter between the Si-based two-level and three-level inverter topologies, two-level inverter topology, and three-level NPC/T-type inverter topologies were also modeled together in the simulation environment and investigated in detail. The aim of this simulation is to reveal the advantages of different inverter topologies considering the switching frequencies. Firstly, the simulation model of a two-level voltage source inverter is given in Figure 4.7. Next, the three-level NPC voltage source inverter circuit model is illustrated in Figure 4.8. Finally, the T-type three-level voltage source inverter model is shared in Figure 4.9.

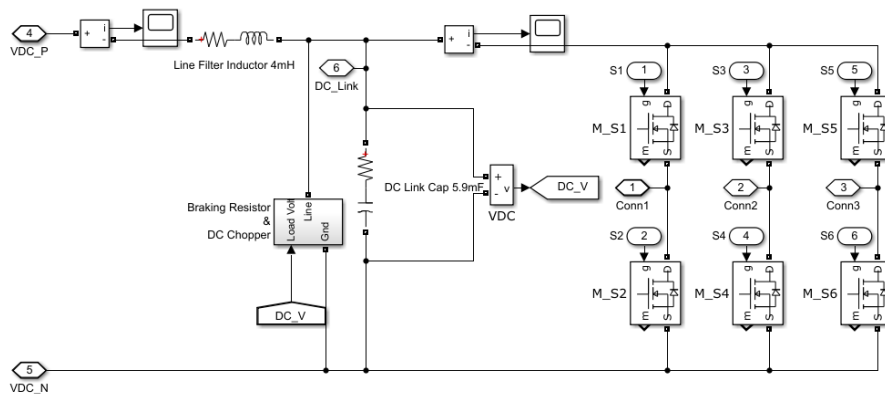


Figure 4.7. Two-level voltage source inverter MATLAB/Simulink model

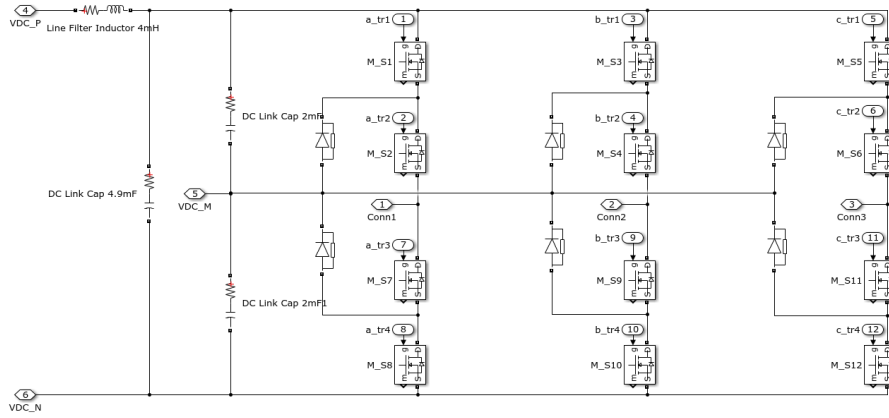


Figure 4.8. Three-level NPC voltage source inverter MATLAB/Simulink model

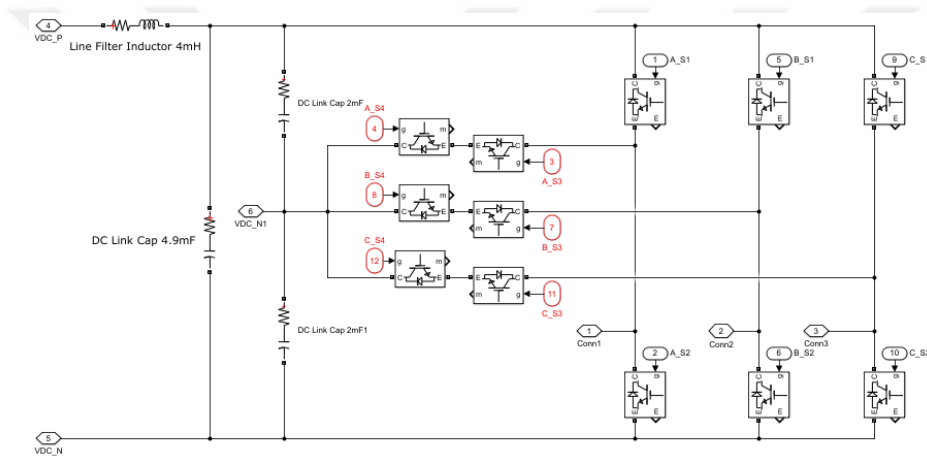


Figure 4.9. Three-level T-type voltage source inverter MATLAB/Simulink model

4.6 Model of Traction Drive

In this section, the overall simulation model of the traction drive system is presented. In the implementation, SVPWM and SHEM techniques are used to modulate the inverters, scalar and vector control methods are used for motor control. Both the switching frequency (f_{sw}) and sampling frequency were firstly set to 1260 Hz for a two-level inverter-based traction drive system. Then, this value was changed to 3255 Hz, 5000 Hz, and 10000 Hz, respectively, in order to examine the effects of different switching frequencies. After that, both the switching frequency and sampling frequency were set as 1260 Hz and 5000 Hz for three-level inverter topologies. The

traction drive model is given in Figure 4.10, and the simulation parameters are given in Table 4.1.

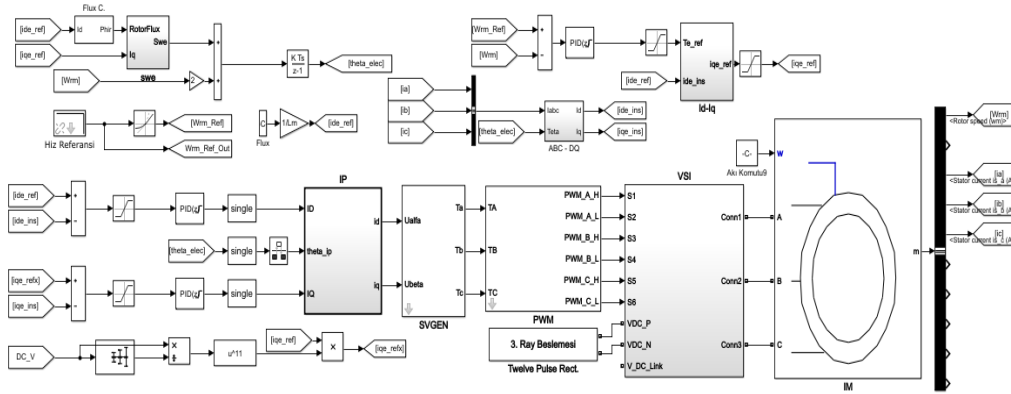


Figure 4.10. MATLAB/Simulink model of traction drive

Table 4.1 Simulation Parameters

Nominal Input Voltage	750 VDC
Feedback Sample Frequency	20 kHz
Switching Frequency	10 kHz
Simulation Sample Time	0.1 us
Inverter Line-to-line Voltage	0-480 Vrms
Inverter Output Frequency	0-200 Hz
Motor Torque	0-750 Nm
Line Inductance	4 mH
DC-link Capacitance	5.9 mF
Traction Motor Stator Resistance	24.77 mΩ
Traction Motor Stator Inductance	284 uH
Traction Motor Trans. Rotor Resistance	18.44 mΩ
Traction Motor Trans. Rotor Inductance	238 uH
Traction Motor Magnetizing Inductance	6.86 mH
Traction Motor Pole-Pairs	2
Traction Motor Inertia	0.69 kg.m ²
Traction Motor Rated Voltage	480 V
Traction Motor Rated Power	125 kW
Traction Motor Rated Torque	671 Nm
Traction Motor Rated Frequency	60 Hz

4.7 Simulation Results

4.7.1 Simulation Results of Traction Drive System

The traction drive system models developed in this study were operated under rated operating conditions at different switching frequencies, and then, the operating performance of them was observed. Figure 4.11 shows the two-level voltage source inverter line-to-line voltage waveform and its harmonic content for $m_i = 0.905$ and $f_{sw} = 1260\text{Hz}$.

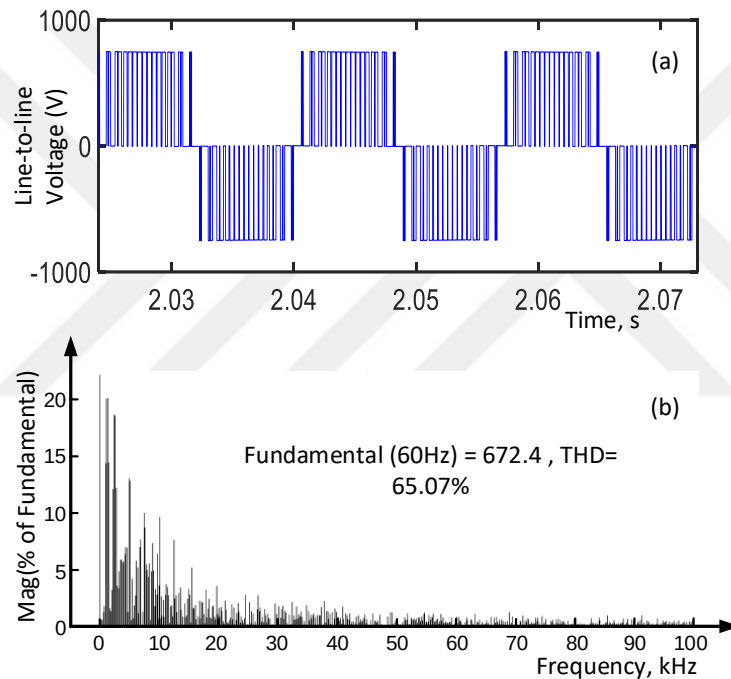


Figure 4.11. SVPWM applied two-level inverter line-to-line voltage waveform and total harmonic distortion for $m_i=0.905$ and $f_{sw} = 1260$ Hz

1260 Hz switching frequency provides a relatively high-frequency modulation ratio to obtain symmetric line-to-line voltage waveforms for a 60 Hz fundamental frequency. This also helps to decrease the low-frequency harmonics. In Figure 4.12, the two-level voltage source inverter line current waveform and its harmonic content for SVPWM when m_i and f_s equal to 0.905 and 1260 Hz, respectively. For the developed system, total harmonic distortion of the line current waveform is obtained

as 14.03 % for these conditions. In fact, for the motor drive applications, there is not a strict limit for motor current harmonics. However, the current harmonics cause the current ripples and hence the torque ripple. Torque ripple may increase losses, cause faster aging for mechanical parts, decreases passenger comfort and the stability of the system. In addition to this, the current harmonics increase the copper and hysteresis losses of the traction motor. The design requirements stated in this study describes strict limits for peak-to-peak torque ripple, which is already defined in the design chapter.

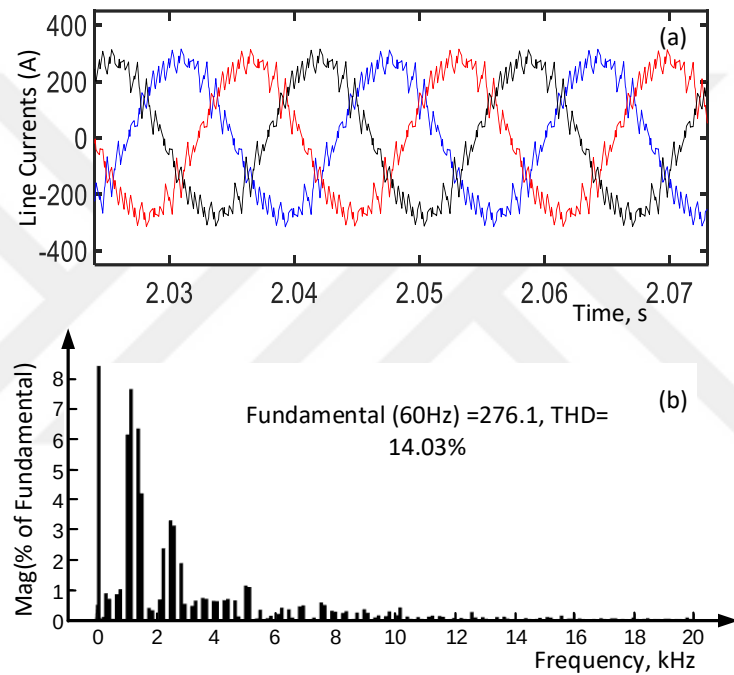


Figure 4.12. SVPWM applied two-level inverter line current waveforms and total harmonic distortion for $m_i = 0.905$ and $f_{sw} = 1260$ Hz

The second observation was made at optimum switching frequency (i.e., 10 kHz) in order to see the effect of switching frequency on output waveforms. Figure 4.13 shows the two-level voltage source inverter line-to-line voltage waveform and its harmonic content for SVPWM, $m_i = 0.905$ and $f_{sw} = 10000$ Hz.

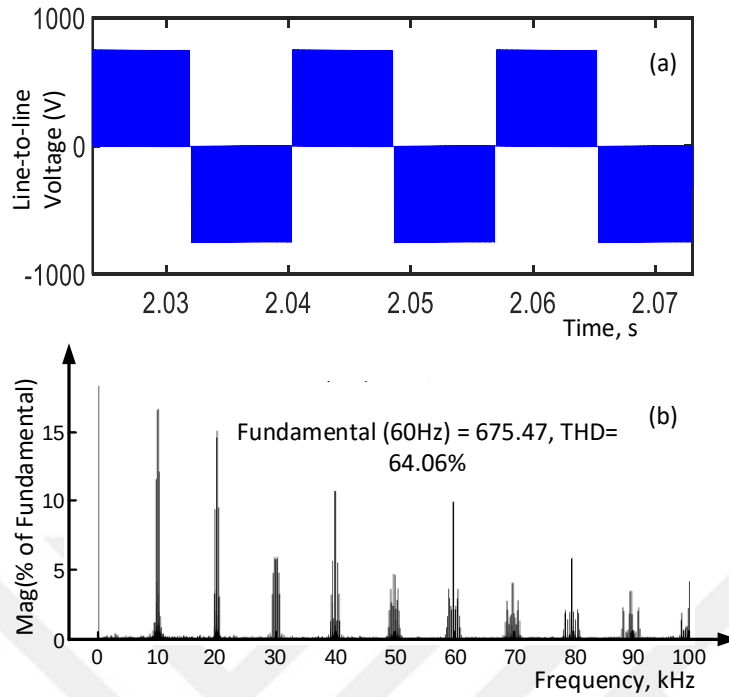


Figure 4.13. SVPWM applied two-level inverter line-to-line voltage waveform and total harmonic distortion for SVPWM, $m_i = 0.905$ and $f_{sw} = 10000$ Hz

It is seen that the THD value of the line-to-line voltage is not affected by a significant switching frequency increase. However, significant changes were obtained for the frequency components of the line-to-line voltage harmonics. As can be followed from Figure 4.11, all significant harmonics components were moved to higher frequencies (i.e., components at switching frequency and integer multiples of it). This will especially help to greatly reduce the motor current harmonics. In Figure 4.14, two-level voltage source inverter line current waveform and its harmonic content for SVPWM at $m_i = 0.905$ and $f_{sw} = 10000$ Hz are given.

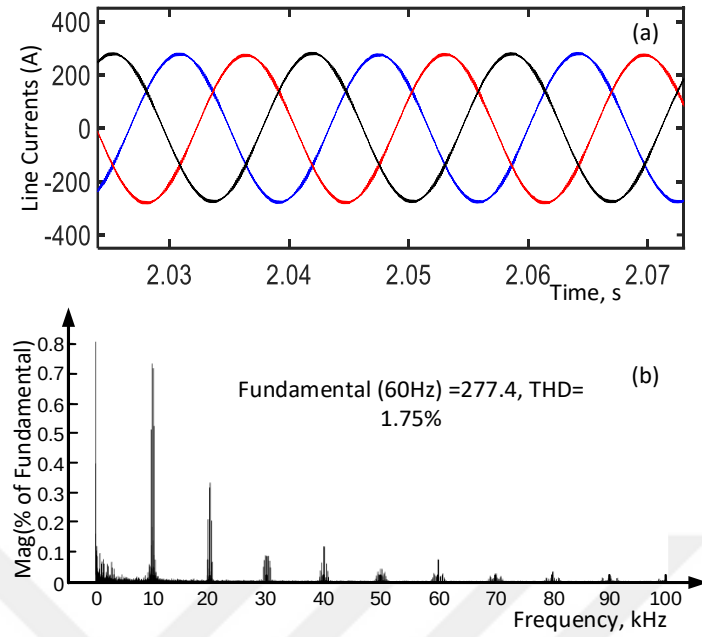


Figure 4.14. SVPWM applied two-level inverter line current waveforms and total harmonic distortion for $m_i = 0.905$ and $f_{sw} = 10000$ Hz

Although we did not observe a change in the THD value of the line-to-line voltage waveforms, the change in harmonic component frequencies of line-to-line voltage causes a drastic decrease in the current harmonics (i.e., 1.75 %).

Figure 4.15 shows the two-level voltage source inverter line-to-line voltage waveform and its harmonic content for SHEPWM, $m_i = 0.905$ and $f_{sw} = 900$ Hz. As explained before, most of the Si-IGBT-based medium voltage and medium power motor drives use the preprogrammed modulation techniques in constant power and high speed operating regions due to the low-frequency modulation ratio. In this study, simulation of the two-level three-phase voltage source inverter with SHEPWM is also performed to see the contributions of this modulation technique. Experimental results of this modulation technique are illustrated in Appendix A.

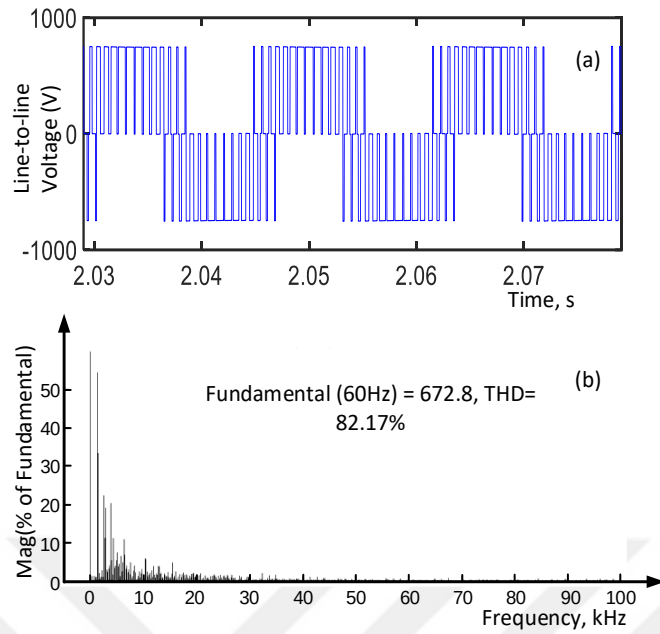


Figure 4.15. SHEPWM applied a two-level inverter line-to-line voltage waveform, and total harmonic distortion for $m_i = 0.905$ and effective switching frequency is 900 Hz

Figure 4.16 shows the two-level voltage source inverter line current waveform and its harmonic content for SHEPWM, $m_i = 0.905$ and $f_{sw} = 900$ Hz.

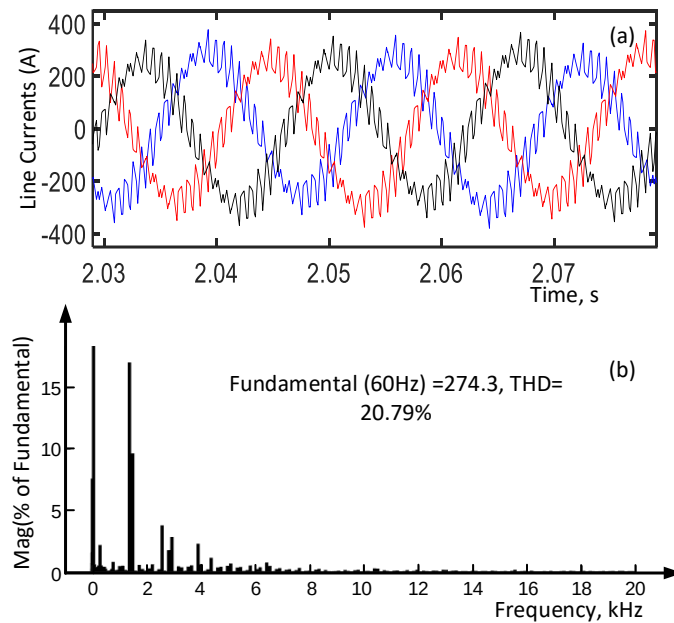


Figure 4.16. SHEPWM applied two-level inverter line current waveforms and total harmonic distortion for $m_i = 0.905$ and $f_{sw} = 900$ Hz

In addition to this, as explained in previous chapters, the medium voltage and medium power two-level voltage source traction inverters suffer high losses that increase with power and voltage levels. Because of this, the topology change is needed between two-level and three-level inverter topologies. Three-level topologies are superior to two-level topologies if the switching frequencies are relatively low. This issue was already examined in the design chapter. However, this situation can be changed if the switching frequencies can be increased. In this study, simulation of the three-level three-phase voltage source inverter with SVPWM is also performed in order to make a comparison between two-level and three-level three-phase voltage source inverter topologies. In Figure 4.17, the three-level inverter line-to-line voltage waveform and its THD are given at $m_i = 0.905$ and $f_{sw} = 1260$ Hz.

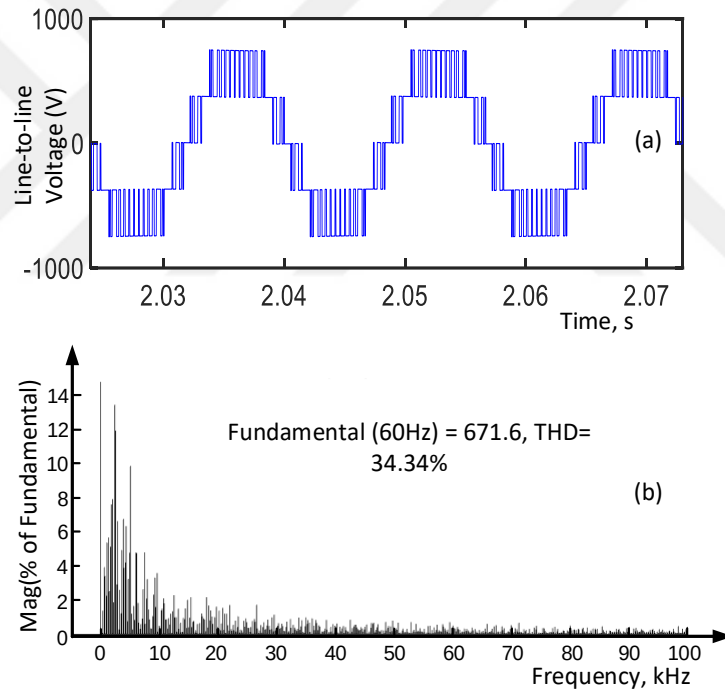


Figure 4.17. SVPWM applied three-level inverter line-to-line voltage waveform and total harmonic distortion for $m_i = 0.905$ and $f_{sw} = 1260$ Hz

A significant improvement in the line-to-line voltage waveform has been achieved compared to the result obtained with the two-level inverter. Here, the THD of the line-to-line voltage is decreased a considerable amount, and the frequency content is

significant harmonics shift towards to high-frequency region. Undoubtedly, this improvement will have a huge impact on current waveforms and hence the losses. Figure 4.18 shows the three-level voltage source inverter line current waveform and its harmonic content for SVPWM at $m_i = 0.905$ and $f_{sw} = 1260\text{Hz}$.

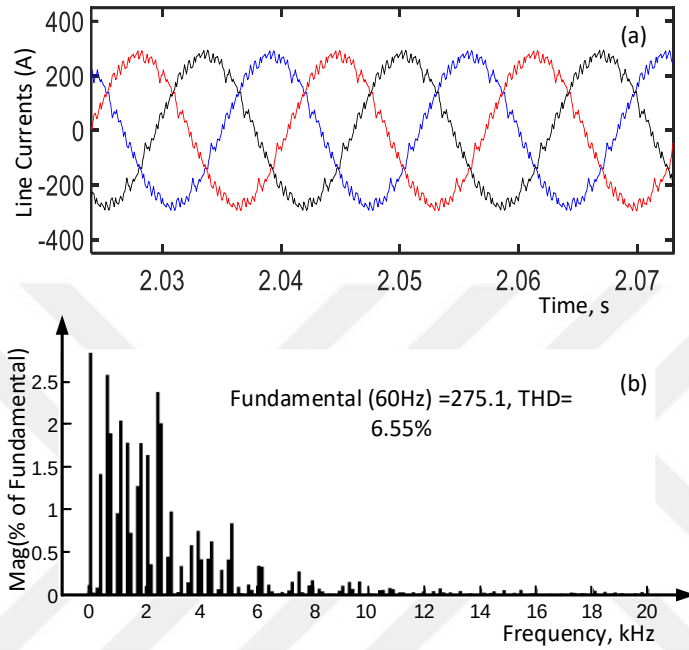


Figure 4.18. SVPWM applied three-level inverter line current waveforms and total harmonic distortion for $m_i = 0.905$ and $f_{sw} = 1260\text{ Hz}$

Same simulations are performed for 5000 Hz switching frequency. Figure 4.19 shows the three-level voltage source inverter line-to-line voltage waveform and its harmonic content for SVPWM, $m_i = 0.905$ and $f_{sw} = 5000\text{ Hz}$.

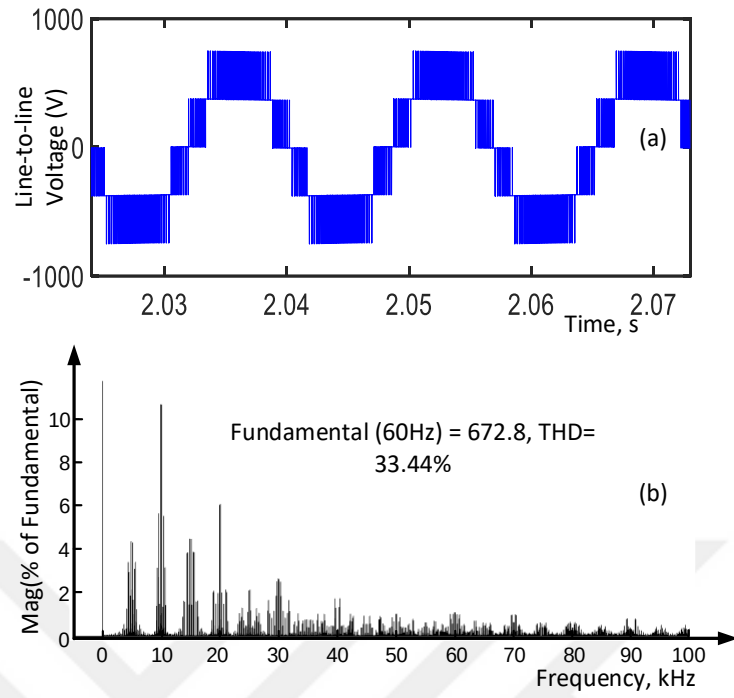


Figure 4.19. SVPWM applied three-level inverter line-to-line voltage waveform and total harmonic distortion for $m_i = 0.905$ and $f_{sw} = 5000$ Hz

As seen in the two-level inverter results, there is no remarkable change in THD of line-to-line output voltage waveform with switching frequency increase. In addition to this, similarly, the frequencies of the harmonics were moved to the high-frequency region. Unlike two-level inverters, the voltage steps are smaller. This means lower electrical stress on semiconductors and passive materials. Figure 4.20 shows the three-level voltage source inverter line current waveform and its harmonic content for SVPWM, $m_i = 0.905$ and $f_{sw} = 5000$ Hz.

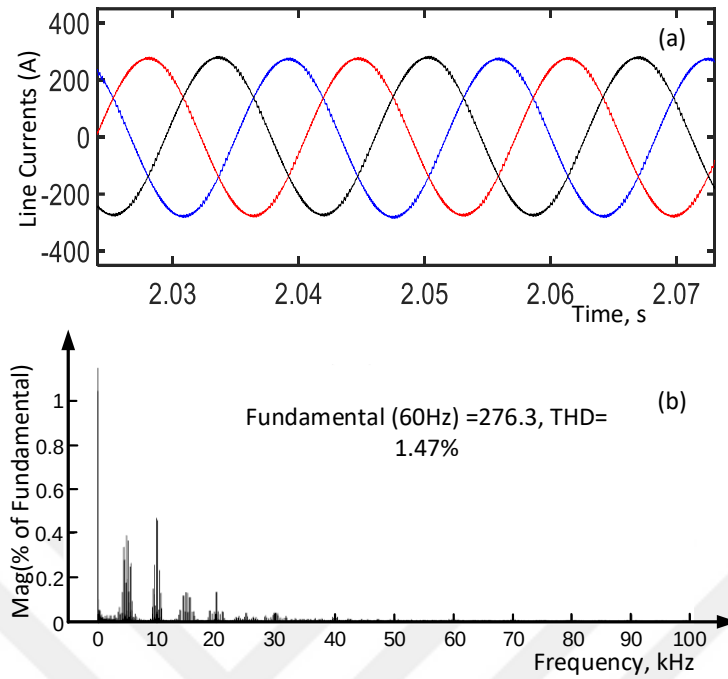


Figure 4.20. SVPWM applied two-level inverter line current waveforms and total harmonic distortion for $m_i = 0.905$ and $f_{sw} = 5000$ Hz

The line current THD level obtained with a switching frequency of 10000 Hz in a two-level inverter was obtained at half switching frequency with a three-level inverter. This means that lower switching losses can be easily reached with the three-level topologies. The blocking voltage and the switching frequency are half of the case of the two-level inverter. In this chapter, the two-level and three-level inverter topologies are examined in terms of the output power quality. Loss, cost, implementation difficulty, and reliability comparisons were made in Chapter 3.

In the following part, the active damping algorithm is examined. In order to verify the developed active damping algorithm, a MATLAB/Simulink model was built. The performance of traction drive without active damping stabilization control is shown in

Figure 4.21 when railway vehicle is operated in a wide speed range (i.e., 0-50 km/h). The waveforms in Figure 4.21 from (a) to (d) are the DC-link voltage, the catenary current, the traction motor shaft torque, and traction motor rotor speed, respectively.

The voltage and current waveforms show that there is a strong resonance due to the input LC filter. This resonance will cause major problems such as instability, failure of semiconductors, passenger comfort issue and undesirable overcurrent/overvoltage trips.

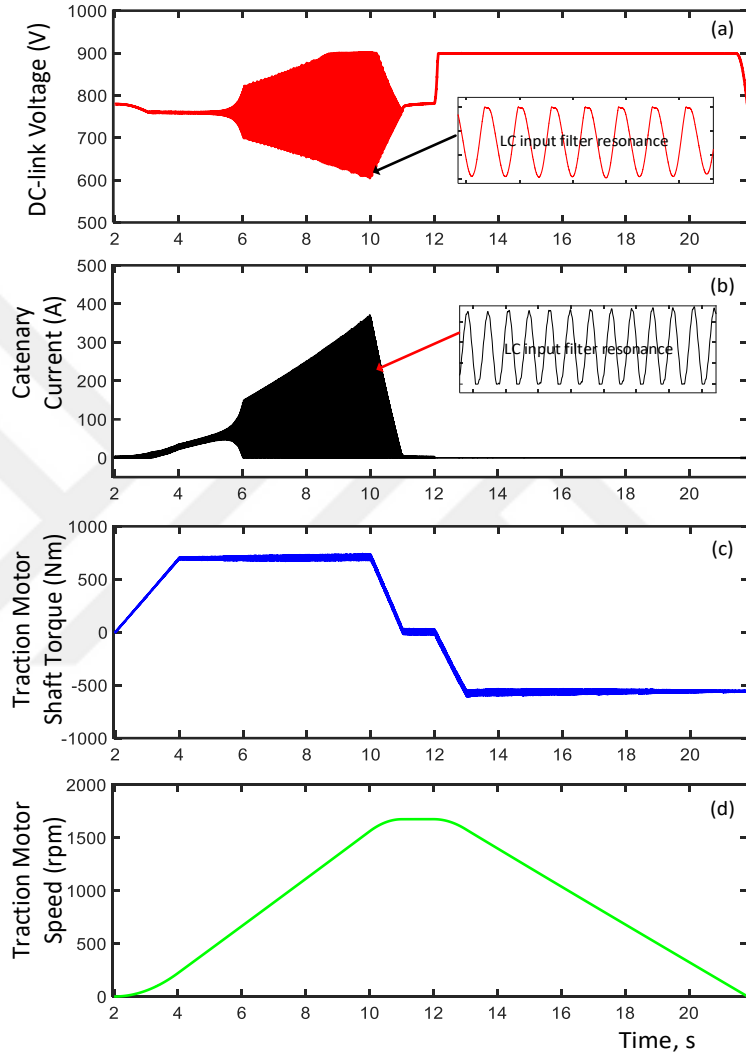


Figure 4.21. Performance of traction drive without active damping stabilization control. (a) DC-link voltage V_{dc} (b) Catenary current I_c (c) Motor shaft torque T_m (d) Motor speed n_r

In Figure 4.22, Fourier analysis results of DC-link voltage and catenary current are plotted. It can be observed from the Fourier analysis results; significant oscillations lead to significant distortion on both current and voltage waveforms. The highest amplitudes of the catenary currents and DC-link voltages are 350 A and 300 V,

respectively. In addition to this, significant harmonics are located around resonance frequencies.

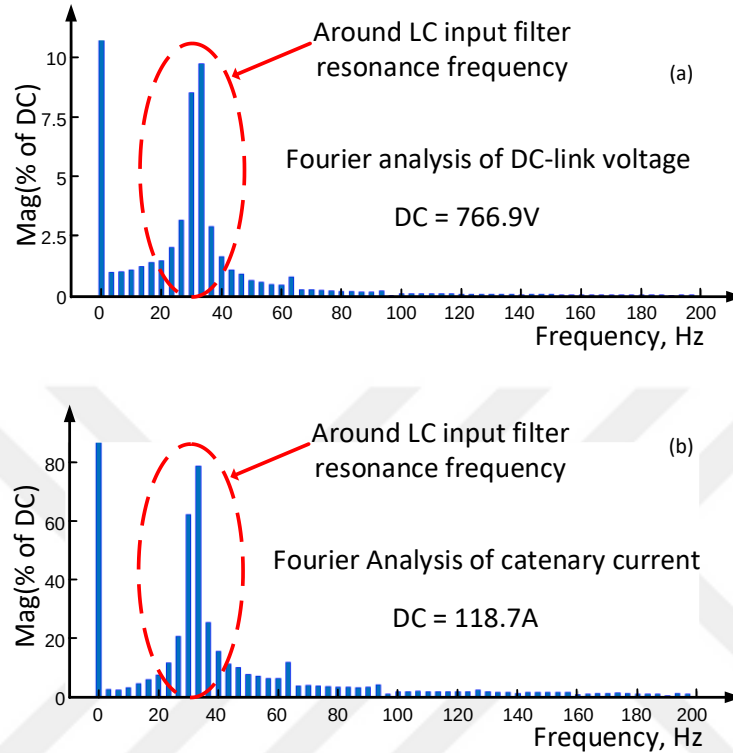


Figure 4.22. Fourier analysis results without active damping stabilization control. (a) Fourier analysis of DC-link voltage (b) Fourier analysis of catenary current

In Figure 4.21 (c) and (d), significant distortions are not seen on the traction motor shaft torque and shaft speed. A closer examination of traction motor shaft torque shows that there is also a distortion. However, due to the high control bandwidth of the current/motor controller, the effect was very low. This effect is also greatly reduced further on the traction-motor speed due to the inertia of the vehicle.

In order to see the success of the active damping control algorithm, the designed controller is activated, and the simulation is performed again. LC filter resonance is damped with an active damping control algorithm. The DC-link voltage, catenary current, motor torque, and motor speed graphics are given in Figure 4.23. The success of the algorithm can be easily understood from the new graphs.

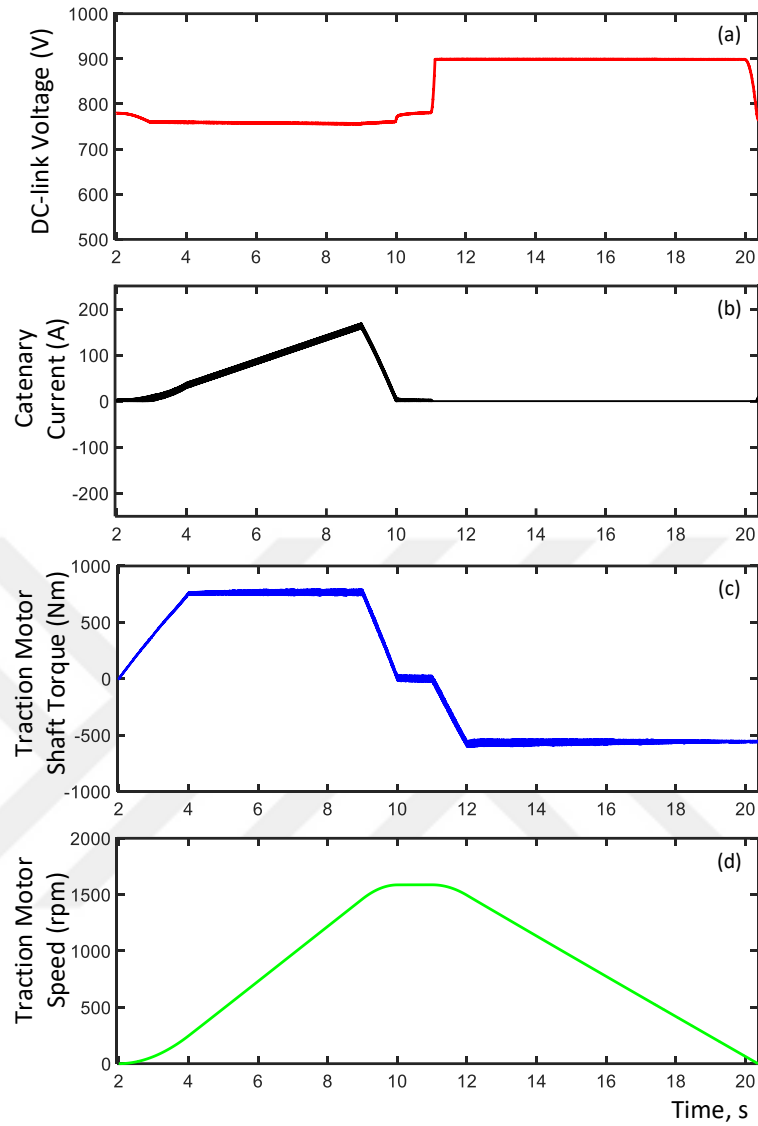


Figure 4.23. Performance of traction drive without active damping stabilization control. (a) DC-link voltage V_{dc} (b) Catenary current i_c (c) Motor shaft torque T_m (d) Motor speed n_r

In Figure 4.24, the acceleration and deceleration performance of the traction drive is illustrated. The reflected inertia of the motor driver has been reduced by 20 times to achieve attractive graphics. From (a) to (d), the traction motor shaft speed, traction motor line current, traction motor line-to-line voltage waveform, and DC-link current of the inverter are given.

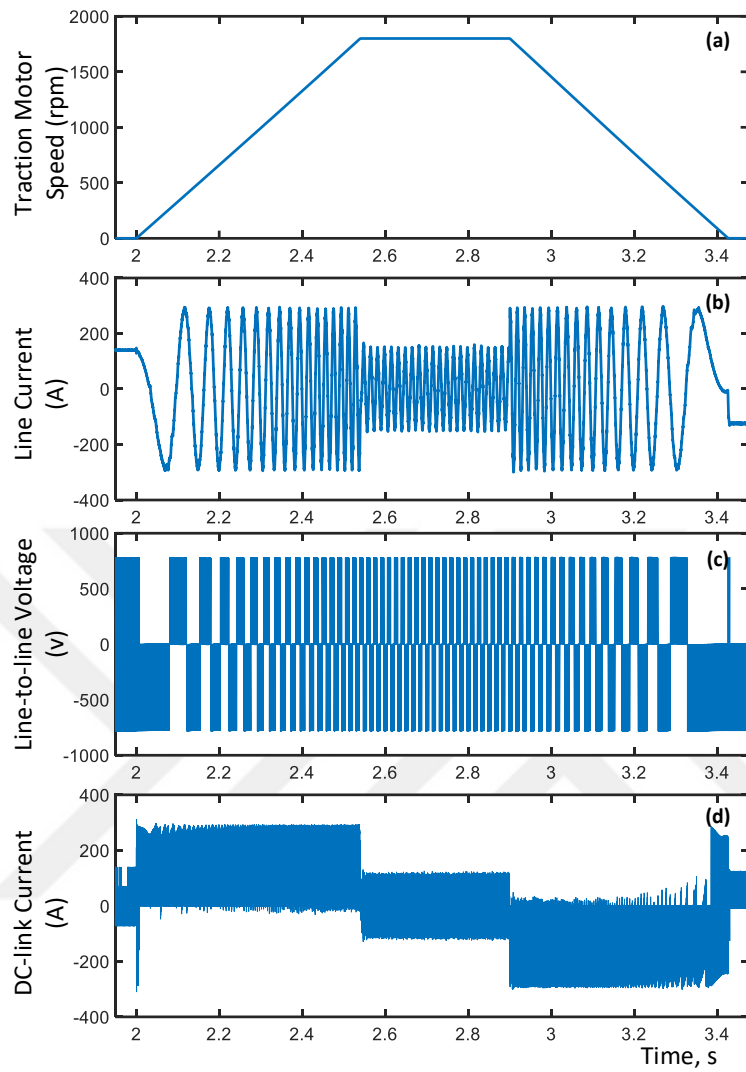


Figure 4.24. Traction drive acceleration and deceleration performance

In addition to Figure 4.24, shaft power and shaft torque curves are given in Figure 4.25.

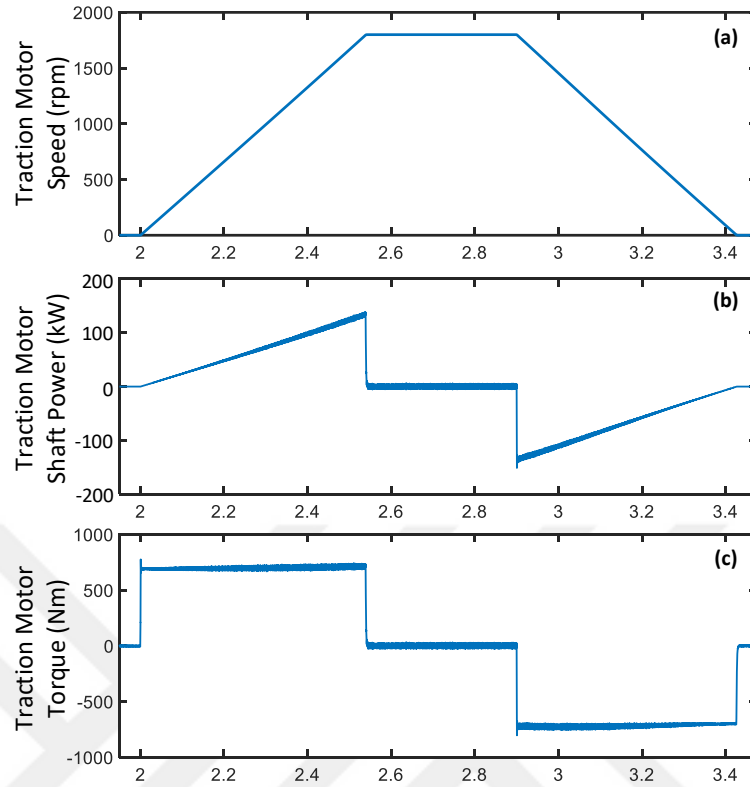


Figure 4.25. Motor shaft power and torque during acceleration and deceleration

4.7.2 Transient Performances

Detailed simulations studies are carried out to show the performance of developed SiC-based traction inverters. In fact, in a real system, the torque command has a jerk limit in order to satisfy the comfort requirements of the rail vehicle. Thus, the torque commands have major slew rates. This jerk limitation prevents sharp changes at the torque reference side. In order to see the transient behavior under the step torque or power demand changes, jerk slew rates are removed from the control algorithms. And then, the dynamic performance of the inverter in the transient regime was observed by changing the torque reference value between 50 % and 100 %. In the 3.5 seconds of the measurement in Figure 4.26, the transient performance of the inverter was investigated by changing the torque reference momentarily from 50 % to 100 %.

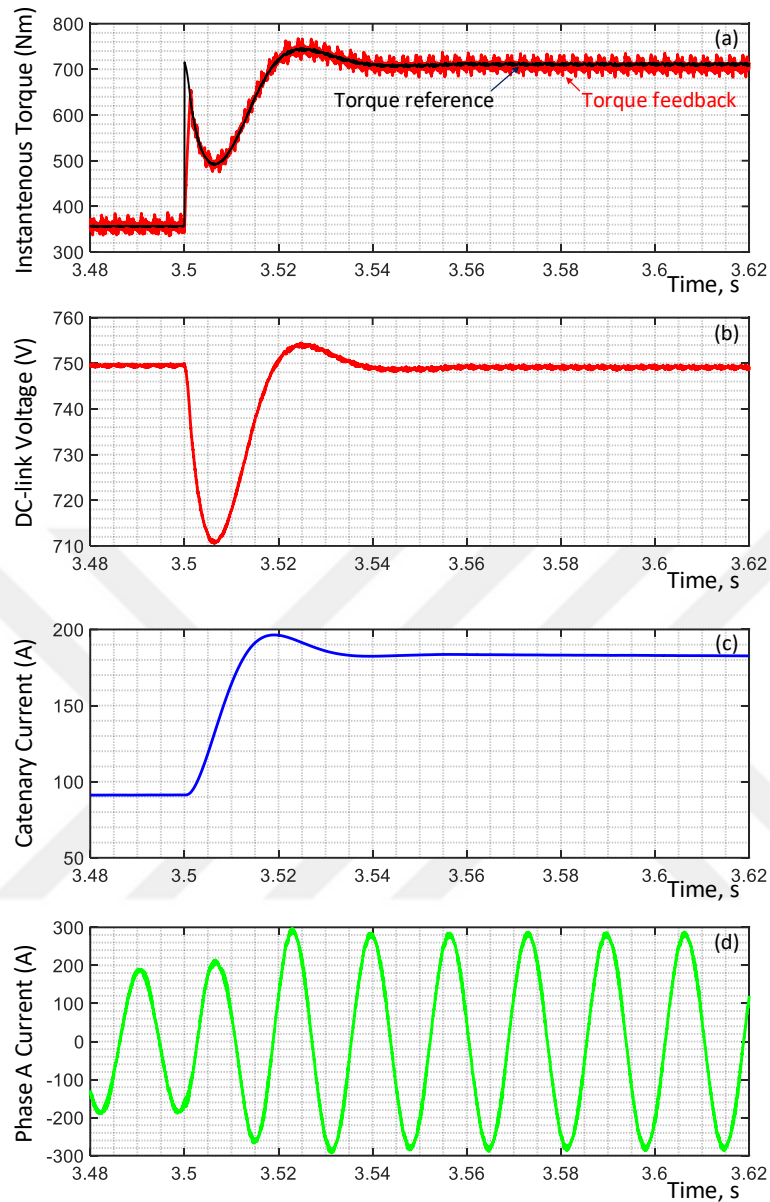


Figure 4.26. Traction inverter 50 % - 100 % load change (a) Instantaneous torque reference and feedback (b) DC-link voltage variation (c) Catenary current variation (d) Change in phase A current

In the 3 seconds of the measurement in Figure 4.27, the transient performance of the inverter was investigated by changing the torque reference momentarily from 100 % to 50 %.

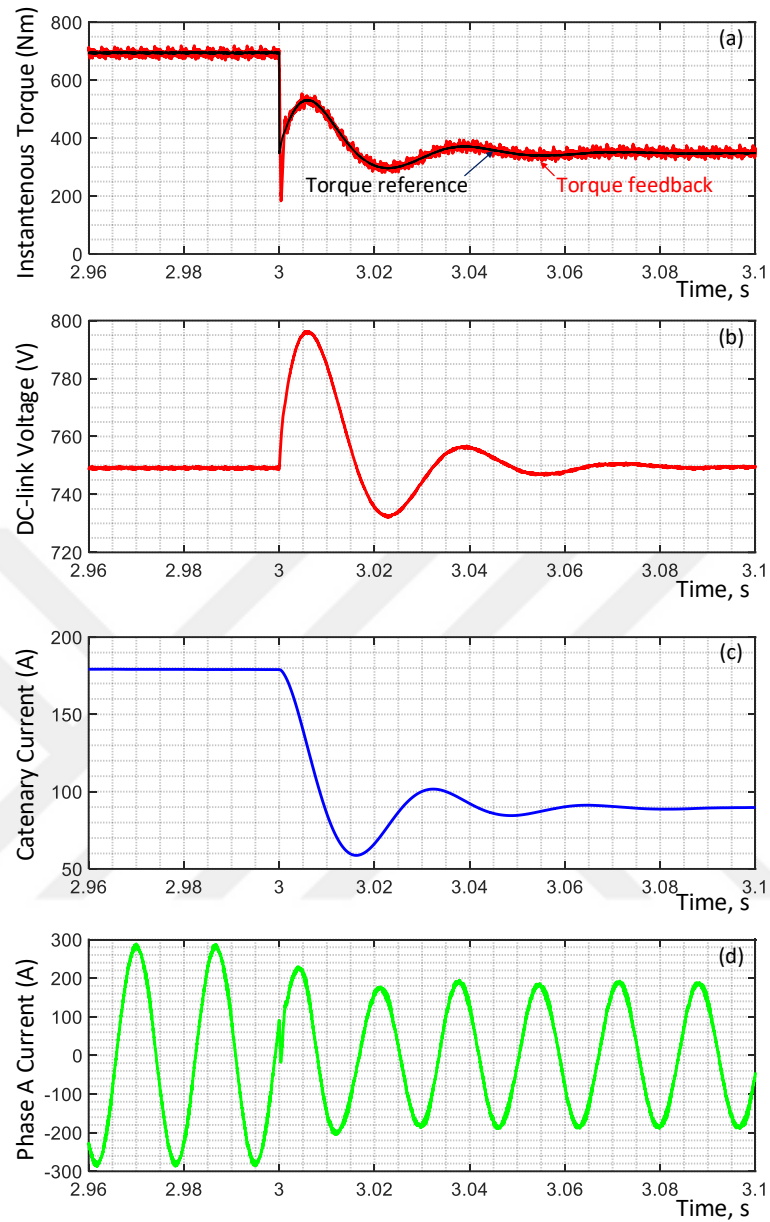


Figure 4.27. Traction inverter 100 % - 50 % load change (a) Instantaneous torque reference and feedback (b) DC-link voltage variation (c) Catenary current variation (d) Change in phase A current

At the instant of torque reference change in Figure 4.26 and Figure 4.27, the torque reference is directly set from 50 % to 100 % and from 100% to 50%. However, as seen in the plots, the reference set value is not in the form of a step waveform. Because the active resonance damping algorithm is changed the form of applied

torque reference. After the step change occurs in the torque reference, the DC-link voltage decreases considerably with the effect of the line reactor. At this time, the active damping algorithm uses the ratio of filtered and unfiltered DC-link voltage as explained in the control design section, and if there is a sudden decrease in the DC-link voltage, the torque set value is also decreased automatically from this algorithm. If there is a sudden increase in DC-link voltage, this time, the set value is increased. Therefore, the shapes of the torque references and feedbacks differ from the square wave at the instantaneous reference change.



CHAPTER 5

FULL-SCALE PHYSICAL SIMULATOR OF ALL SiC TRACTION MOTOR DRIVE FOR LIGHT RAIL TRANSPORTATION SYSTEMS

It is essential to experimentally validate the simulation results using a real laboratory setup. A special experimental setup has been built to carry out the experimental studies. In this chapter, first, the architecture of the experimental test setup is introduced, and the design aspects of each subsystem component are explained in detail in this chapter. The experimental results are presented along with figures and tables in the next chapter.

5.1 Laboratory Experimental Setup

Although the main subject of this study is the design and implementation of a full SiC-based traction inverter, a full-scale physical simulator per traction motor for all SiC traction motor drive with an onboard supercapacitor ESS has been developed as a valuable tool in the design and test of all SiC traction converters [93]. Hence, the design and implementation of a full SiC-based PWM rectifier and a full SiC Buck-Boost Converter with a supercapacitor energy storage bank are other important parts conducted in this work. In addition to this, a motor test bench and a motor drive system have also been installed in the scope of this study. The complete laboratory setup in the block diagram form is given in Figure 5.1.

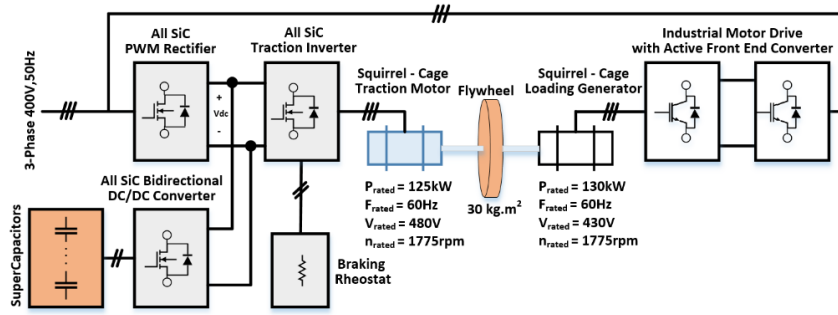


Figure 5.1. Block diagram of the developed laboratory test setup

The 750 V DC catenary line is created by developing a bidirectional, all SiC PWM rectifier and is controlled to change the catenary voltage between 565 - 900 V DC using special software for the catenary line model. The designed and constructed SiC PWM rectifier during this thesis are as shown in Figure 5.2.

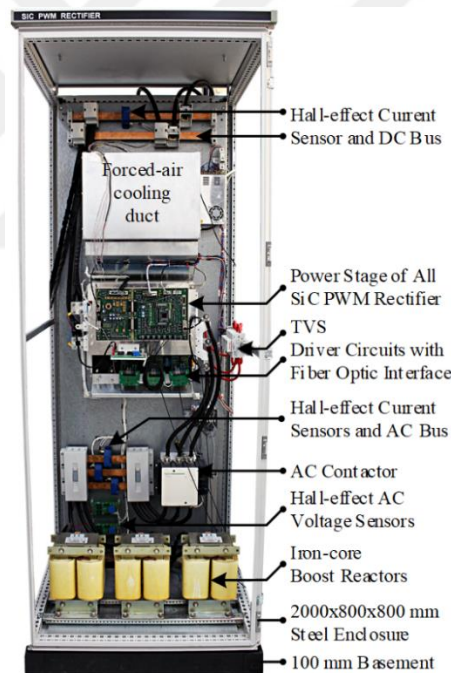


Figure 5.2. The developed PWM rectifier prototype

The PWM rectifier is supplied from a three-phase 400 V, 50 Hz grid and can deliver 150 kW DC power to the traction inverter at rated DC link voltage. The traction inverter can deliver 185 kVA to the traction motor. A bidirectional DC/DC converter for supercapacitor ESS is also developed and integrated into the DC link. It is

operated in buck mode during charging and in boost mode during discharging periods of the supercapacitor bank.

Figure 5.3 shows the all-SiC power MOSFET-based two-level voltage source three-phase traction inverter developed and implemented in this thesis. Traction inverter composed of a bottom-to-top mounted power electronic block, current and voltage sensors, a bulky filter capacitor, forced air cooling duct and fan, air-core line reactor, soft-start contactor and resistance, and the main line contactor.

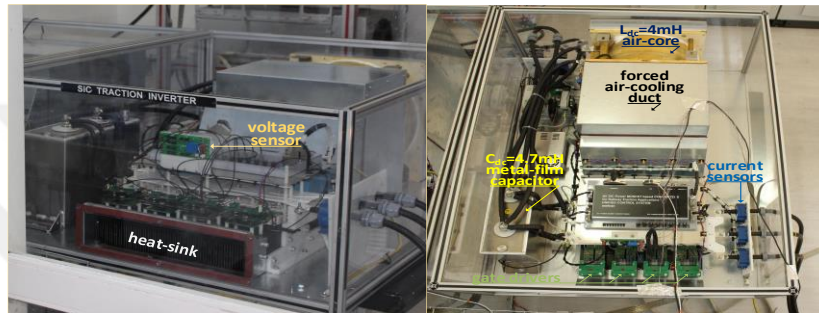


Figure 5.3. Full SiC MOSFET based traction inverter prototype

A detailed block diagram of the developed full SiC MOSFET-based traction inverter is given in Figure 5.4.

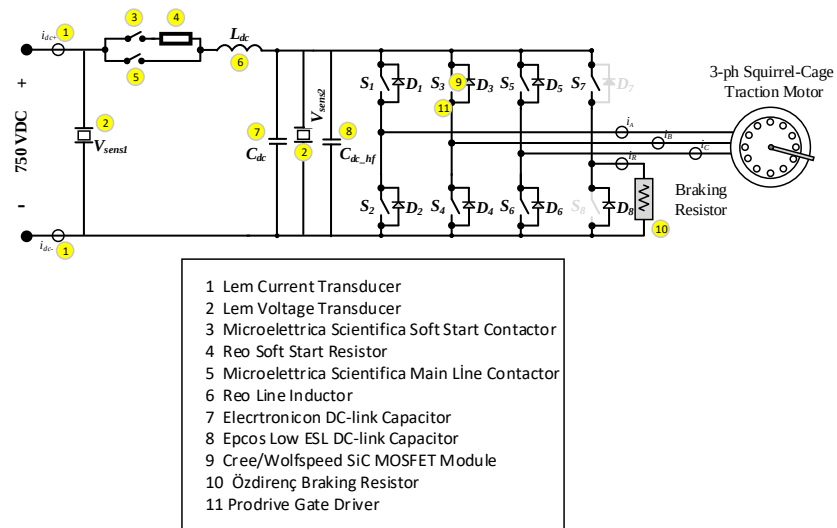


Figure 5.4. A detailed block diagram of the developed full SiC MOSFET traction inverter

As explained in Figure 5.1, the traction motor drive consists of a three-phase two-level PWM traction inverter to drive a three-phase squirrel-cage traction motor, a flywheel coupled to the motor shaft to represent the inertia of the transportation vehicle, a loading generator connected to the grid via an industrial motor drive with an active frontend to model the dynamic behavior of the transportation vehicle caused from the rail track. A photograph of the motor test bench is given in Figure 5.5.



Figure 5.5. Installed motor test bench

The loading generator of the physical simulator is connected to the common utility grid through an industrial motor drive with an active front-end converter to represent the mechanical load of the vehicle acting on the shaft of the traction motor in the dynamic state. On the other hand, the flywheel coupled to the shaft of the traction motor represents the inertia ($J=30 \text{ kg.m}^2$) of the vehicle per traction motor during acceleration and deceleration. The kinetic energy storage capability of the flywheel is 0.54 MJ at 1800 rpm. All rotating parts of the physical simulator are mounted on a common steel base frame. The base frame is isolated from the third floor by using low-frequency anti-vibration pads. Misalignment, unbalance, and vibration of rotating parts are periodically measured by proper instruments.

The output power of the loading generator is controlled by Emotron VFXR/FDUL46-430 regenerative and low harmonic industrial AC drive with active front-end converter in order to return the generated power to the common 400 V, 50 Hz grid. A basic circuit diagram and a photograph of the installed industrial motor drive are given in Figure 5.6 and Figure 5.7.

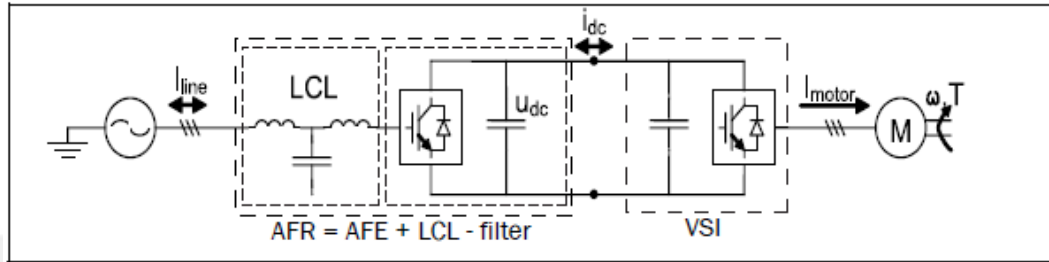


Figure 5.6. The basic circuit diagram of Emotron VFXR/FDUL46-430 regenerative and low harmonic industrial AC drive

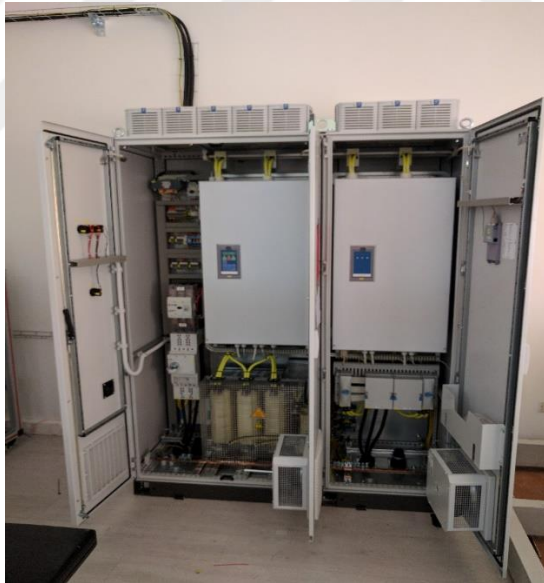


Figure 5.7. The photograph of Emotron VFXR/FDUL46-430 regenerative and low harmonic industrial AC drive

Since the AC supply side of the physical simulator is a common bus, electrical power consumed from the utility grid only accounts for the power losses of converters, electrical machines, and cabling during normal operation.

The developed physical simulator has all the protection, control, and communication facilities of a real rail transportation system. The Signal interface diagram of the physical simulator system is as shown in Figure 5.8. These signals are primarily used for monitoring and protection and secondarily for control purposes. The communication network between various blocks of the physical simulator and the MATLAB/Simulink models running on a Workstation is as given in Figure 5.9. The associated control and communication scheme is as shown in Figure 5.10, to illustrate the online coordinated operation of MATLAB Simulink and all subsystems of the physical simulator.

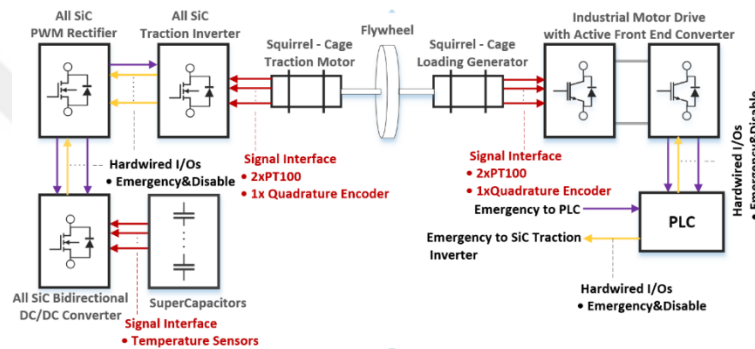


Figure 5.8. Signal interface diagram of the developed physical simulator

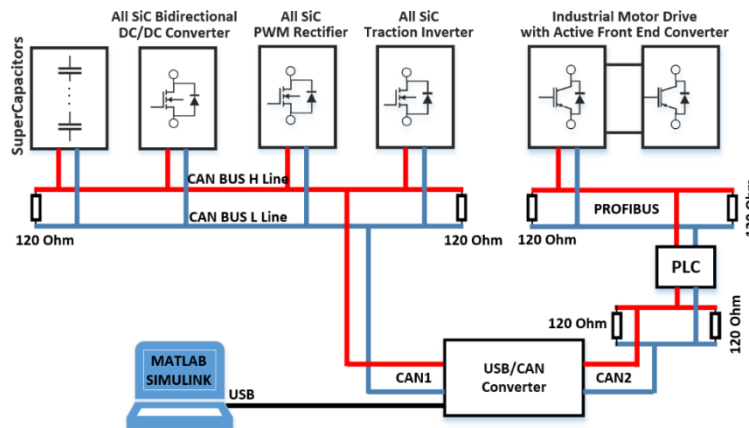


Figure 5.9. The communication network of developed physical simulator

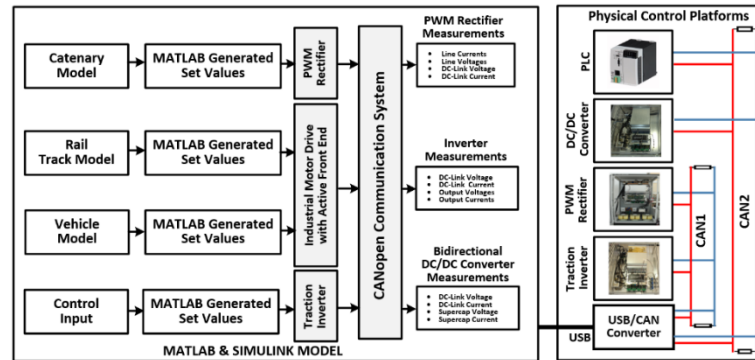


Figure 5.10. Control and communication scheme of the developed physical simulator

The footprint of the laboratory test setup is as given in Figure 5.11.

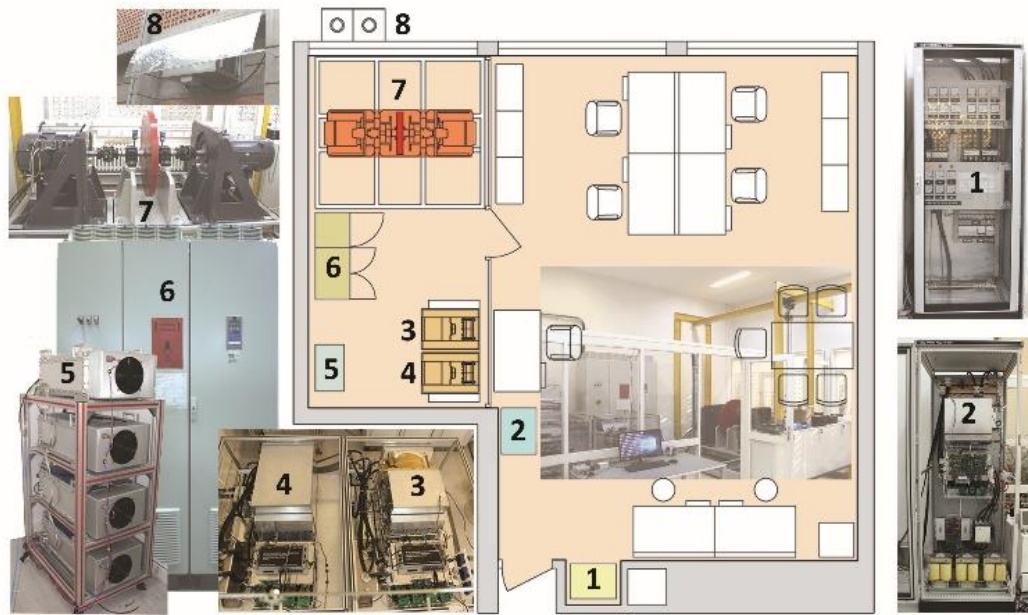


Figure 5.11. The footprint of the test setup 1: 400V, 50 Hz grid panel, 2: PWM rectifier, 3: Traction inverter 4: Bidirectional buck-boost converter, 5: Supercapacitor bank, 6: Industrial motor drive, 7: Traction motor&loading generator, 8: Braking resistor

5.2 Contributions of The Test System

The main contributions of the developed test system can be summarized as follows:

- i. The developed physical system can simulate the operational performance of Light Rail Transportation Systems (LRTS) in real-time, both in the steady-state and dynamic state. It is equipped with software programs for the catenary model, rail-track model, and vehicle model, thereby allowing performance verification of various control, modulation, and energy-saving strategies. Operational strategies are summarized in Figure 5.12.

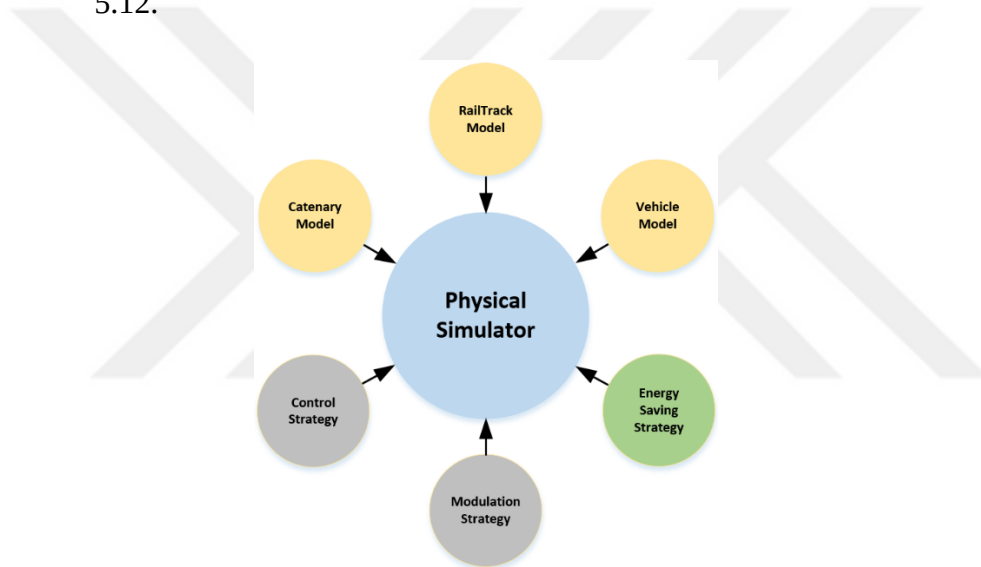


Figure 5.12. Operational strategies of the physical simulator

The vehicle model is a valuable design tool in the formation of LRTS, together with the rail track model and catenary model. The performance of the resulting LRTS can then be tested for different rail tracks and for various catenary lines by operating the physical simulator system. Pre-programmed catenary, rail track, vehicle, and energy-saving models run on the MATLAB/Simulink environment to operate the physical simulator. This unified software can be adapted to different vehicles, rail tracks, and catenary lines simply by reconfiguring some parameters.

- ii. The future trend in traction applications is the use of SiC power MOSFET-based power converters. The physical properties of SiC power MOSFETs make possible the development of higher efficiency, smaller size, and weight, and lower current harmonic distortion traction converters in comparison with the ones based on Si-IGBT. Therefore, the physical simulator described in this thesis permits the design, development, and performance testing of all SiC traction converters such as the traction inverter, PWM rectifier, and ESS.
- iii. Since LRTS and its loading system are operating connected to a common 400 V, 50 Hz bus, the physical simulator consumes very low power, i.e., nearly 20% of its installed capacity from the utility grid. This permits the installation of the physical simulator in laboratories that have weak utility grid connections.

CHAPTER 6

VERIFICATION OF 500 - 900 VDC, 165 KVA ALL SIC MOSFET BASED TRACTION INVERTER

In this chapter, the performance of the developed traction system has been presented using the track data, which is obtained from the real field. The traction inverter is fed with an active PWM rectifier. The input voltage of the traction inverter is regulated according to catenary model outputs by utilizing the active PWM rectifier. In order to load the traction motor, Emotron VFXR/FDUL46-430 regenerative and low harmonic industrial AC drive is used with another traction motor. A user interface was also developed and implemented in this study to control Emotron industrial AC drive. This is carried out using EATON XC-CPU202-EC4M-XV modular PLC. The vehicle and route model is embedded on the Workstation. In addition to this, the inverter performance has been tested under various steady-state and dynamic regimes.

The experimental records have been obtained using the measuring devices listed in Table 6.1.

Table 6.1 List of The Measuring Devices

- Tektronix MDO3034 Digital Oscilloscope 350 MHz 2.5 Gs/s - Tektronix P5210 High Voltage Differential Probe 50 MHz - Tektronix P5205A High Voltage Differential Probe 100 MHz - Tektronix TCP404XL Current Probe 2 MHz & TCPA400 Current Probe Amplifier - CWT Ultra-Miniature High-Frequency Rogowski Current Transducers 30 MHz - Hioki PW3198 Power Quality Analyzer - Sper Scientific 800007 2 Channel Thermocouple Measurement Device
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6.1 Experimental Results

In this section, experimental results taken during this study are illustrated. Before sharing the measurement results, the accuracy rates of the devices will be presented in Table 6.2.

Table 6.2 Device Accuracies

Device	Oscilloscope	Current Probe	Voltage Probe
Accuracy	$\pm 1.5 \%$ (AC+DC)	$\pm 1 \%$ (DC), $\pm 3 \%$ (AC)	$\pm 2 \%$ (AC+DC)

As seen from Table 6.2, the device accuracies are not good enough to take very sensitive measurements. Additional measuring devices (Yokogawa WT3000 power analyzer, LEM LV25-1000 voltage transducers and LEM LF-510-S current transducers with high precision measuring resistance) were used to determine the calibration coefficients of test setup (oscilloscope, current probe and voltage probe). Accuracies of 0.06 % and 0.1 % were achieved for DC and AC measurements, respectively. The accuracy of efficiency is calculated as nearly $\pm 0.32 \%$.

6.1.1 SiC Power MOSFET Switching Waveforms

Turn-on and turn-off voltage and current waveforms of CAS300M17BM2 SiC power MOSFETs employed in traction inverter are recorded and given in Figure 6.1 and Figure 6.2. Fall time and rise time of the voltage waveforms are measured as 58 ns and 73 ns, respectively. It is seen that the time durations are very short compared to the fall and rise times of the Si-IGBTs. The switching time descriptions are defined using the datasheet of CAS300M17BM2 SiC power MOSFET module.

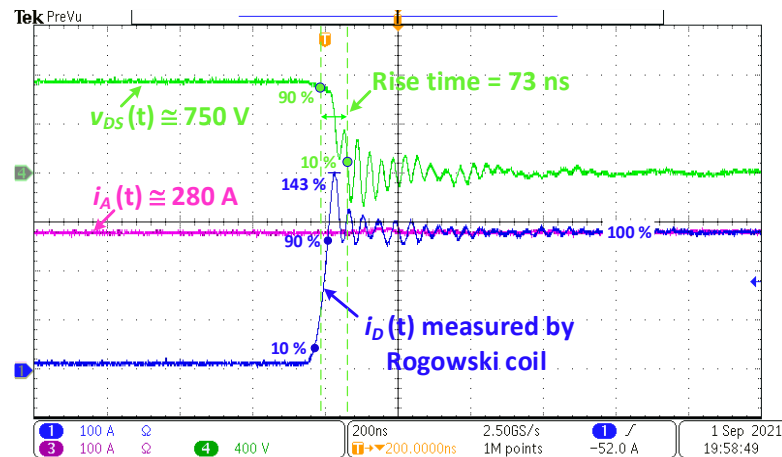


Figure 6.1. Voltage and current waveforms at turn-on

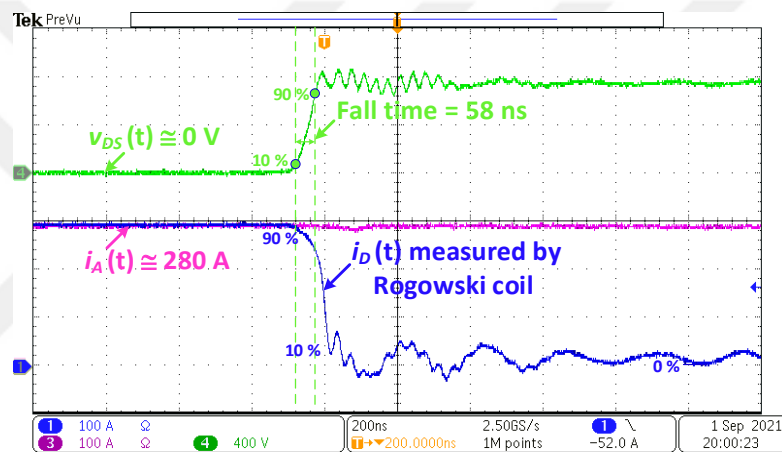


Figure 6.2. Voltage and current waveforms at turn-off

One of the major benefits of the SiC power MOSFETs is the very low switching losses obtained from the high switching speed. However, this can create extra stress for some applications. On the other hand, fast switching transitions may create an electro-magnetic interference problem. This fast transition creates big difficulty to avoid common mode voltages and currents. In order to improve this situation parasitic capacitances should be minimized. Additionally, due to the vesity fast transitions, the effects of parasitic inductances are also getting more important. In order to operate the semiconductor at high voltages, reduce the emission and reduce the electrical stresses on passive materials, parasitic inductances should also be minimized. Depending on the application, it may be possible to obtain faster or slower turn-on and turn-off times by adjusting the gate resistance and gate-source

capacitance in the gate driver circuits. However, major changes may cause to loss of the advantage of SiC MOSFETs. Hence, the changes should be made with careful consideration. An alternative but more complex and costly improvement technique might be the design and use of an active gate driver to adjust only the gate resistance in the Miller Plateau by measuring the gate-source voltage.

6.1.2 dv/dt 's at Inverter and Motor Terminals

125 kW, 480 V, 60 Hz traction motor is supplied from the developed all-SiC traction converter via railway qualified single core flexible rubber cables as illustrated in Figure 6.3. Effects of SiC power MOSFET operation on traction motor are experimentally obtained on this setup.

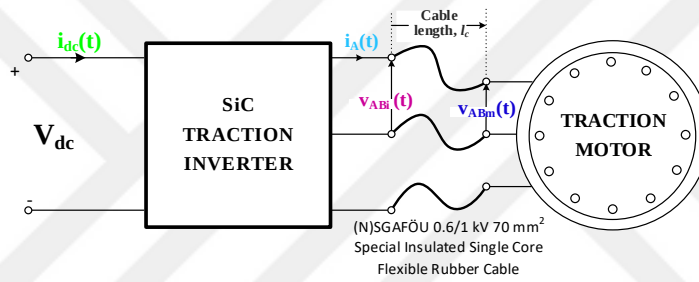


Figure 6.3. Experimental setup to test the effects of cable length on the system performance

Nowadays, in most of the railway traction converters, Si-IGBTs are employed. With the advent of SiC technology, Si-IGBT-based traction converters will be replaced by SiC power MOSFET based ones in the near future. SiC power MOSFETs can be turned-on and -off 2-5 times more rapidly and at 3-10 times higher switching frequencies in comparison with Si-IGBTs. The traction converter applies PWM voltage waveforms with very short rise times to the stator of the traction motor via a three-phase or 3x single phase cable system. This traction specific cable system forms an RLC network and hence high frequency oscillations take place and applied voltage is enhanced at the motor terminals. These voltage overshoots are created by reflected waves at the interface between cable and electrical machine terminals due to impedance mismatch, and depend on the converter output, the cable length and electrical machine terminal impedance [101], [102]. These are verified

experimentally on the developed system shown in Figure 6.3 by recording $V_{ABi}(t)$, $V_{ABm}(t)$, $i_A(t)$ and $i_{dc}(t)$ for two different cable lengths, of $\ell_c = 4$ and 12 meters. These records are given in Figure 6.4 when the peak value of the inverter output voltage is set to $V_{ABi} = V_{dc} = 765$ V.

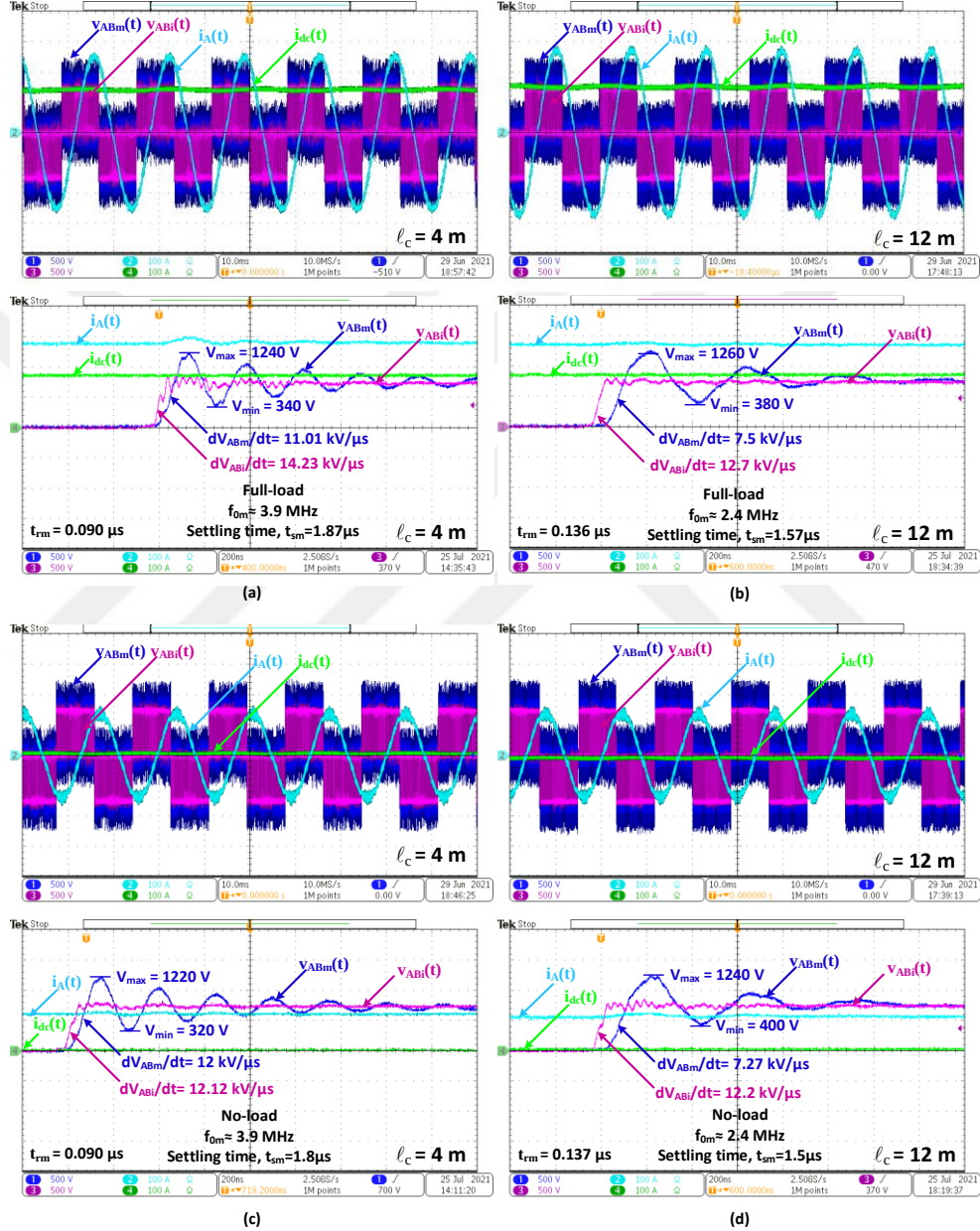


Figure 6.4. Effects of motor cable length on line-to-line motor voltages (a) 12 m cable length at full-load (165 kVA), (b) 4 m cable length at full-load (165 kVA), (c) 12 m cable length at no-load (83 kVA), (d) 4 m cable length at no-load (83 kVA)

The following conclusions can be drawn from the results of experimental work carried out in this section:

- Motor line currents are nearly sinusoidal e.g. $i_A(t)$.
- Although $V_{ABi} = V_{ABm} \cong 480$ V l-to-l rms, first peak of $V_{ABi}(t) \cong 780$ -810 V is enhanced to $V_{ABm} = 1240$ V for $l_c = 4$ and 1260 V for $l_c = 12$ meters.
- Frequency of oscillation in line-to-line voltage waveform is lower for longer cables. Settling time of these oscillations are shorter for longer cables.
- dv/dt of each pulse in the line-to-line PWM voltage waveform at the output of the inverter is usually higher than 10 kV/ μ s (in excess of 10 kV/ μ s). Cables connecting inverter to the motor lower (lessen) these high dv/dts usually below 10 kV/ μ s.
- Records for no-load operation of traction motor drive are given in Figure 6.4 (c) and (d) which correspond to records in Figure 6.4 in (a) and (b) for full-load operation. Similar conclusions can be drawn also for no-load operation.
- For the application described in this paper, a cable length around 12 m is more beneficial than shorter cables such as 4 m in length.

Since the catenary voltage varies in a wide range, records corresponding to Figure 6.4 (b) for $l_c = 12$ m are obtained from the developed system and given in Figure 6.5 (a) and (b) respectively for $V_{dc} = 600$ V and 900 V.

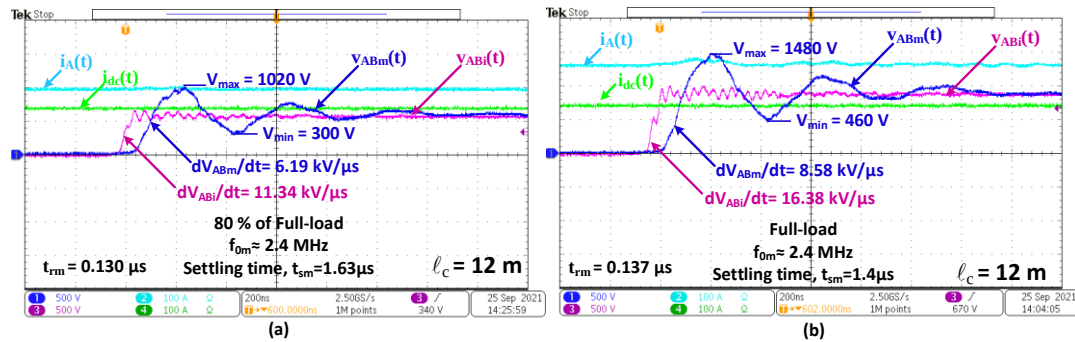


Figure 6.5. Effects of catenary voltage V_{dc} on line-to-line motor voltages (a) $V_{dc} = 600$ V and at 80 % of full-load and (b) $V_{dc} = 900$ V and full-load

As can be understood from Figure 6.4 and Figure 6.5, first voltage peaks in the pulses of $V_{ABi}(t)$ and $V_{ABm}(t)$ and corresponding $dv_{ABi}(t)/dt$ and $dv_{ABm}(t)/dt$ values become higher and higher as the catenary voltage, V_{dc} increases. Limiting curve of admissible impulse voltage for the industrial induction motors is as recommended in Fig. 40 of [101] and is partly redrawn by extrapolating the data from $t_{rm}=0.1$ to $0.05 \mu s$ as given in Figure 6.6.

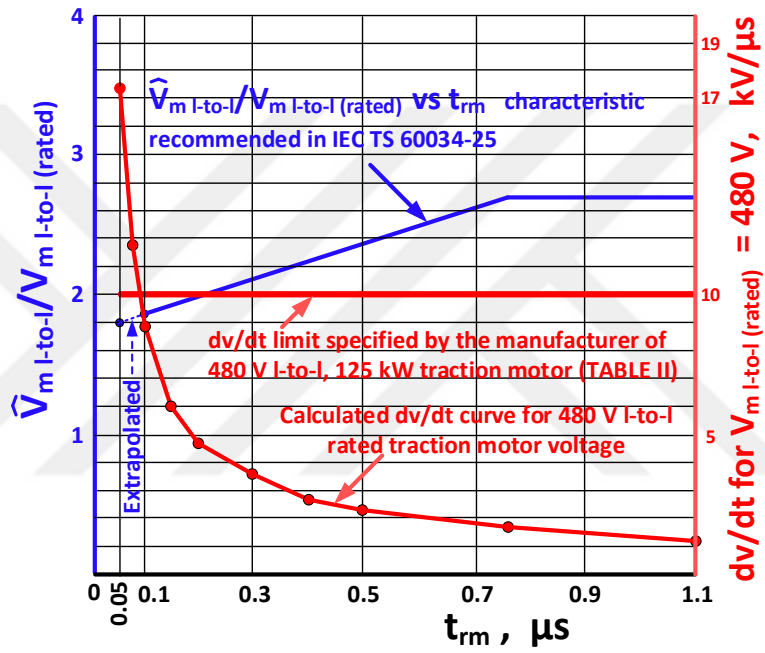


Figure 6.6. Limiting curve in Fig. 40 of [101] and corresponding dv/dt curve for 480 V l-to-l rated traction motor voltage

This standard defines the area under the blue colored curve as the safe operating area for the induction motor. In this research work, rise time t_{rm} of the line-to-line voltage pulses at the stator terminals is measured to be 90 ns in the worst case ($l_c = 4 \text{ m}$). Since $\hat{V}_{m \text{ l-to-l}} / V_{m \text{ l-to-l (rated)}} \cong 2$ for $t_{rm} = 90 \text{ ns}$ and $V_{m \text{ l-to-l (rated)}} = 480 \text{ V rms}$ then maximum recommended peak value of the impulse voltage is calculated to be $2 \times 480 = 960 \text{ V}$. However, $\hat{V}_{m \text{ l-to-l}}$ is measured to be 1260 V (Figure 6.4 (a)) which is much higher than maximum recommend value of 960 V. On the other hand, maximum permissible value of dv_{ABm}/dt is recommended to be

10 kV/ μ s by traction motor manufacturer. Permissible dv_{ABm}/dt curve can be calculated from $\hat{V}_m l - t_o - l/V_m l - t_o - l$ (rated) curve for $V_m l - t_o - l$ (rated) = 480 V and different t_{rm} values. The resulting characteristics is marked on Figure 6.6. Although its value is calculated to be nearly 10 kV/ μ s, its actual value is measured to be 11 kV/ μ s (Figure 6.4 (a)) while $dv_{ABm}/dt = 8.58$ kV/ μ s (Figure 6.5 (b)) for $l_c = 12$ m which is safe.

However, in the practical applications of SiC power MOSFET-based traction converters t_{rm} may be lower e.g., 50 ns and the cable may be shorter $l_c = 3$ -4 m which result in dv_{ABm}/dt values considerable higher than 10 kV/ μ s.

Presently, [102] recommends one of the following techniques to reduce dv_{ABm}/dt in industrial motors fed from PWM converters switching frequencies of which is about 2.5 kHz: i) output reactors, ii) dv/dt filter, iii) sinusoidal filter, and iv) electrical machine termination unit. These techniques reduce dv_{ABm}/dt values in Si-IGBT-based traction converters below 1 kV/ μ s at the expense of 0.5 - 1 % drop in the efficiency of traction motor drive.

In view of the above findings, insulation systems should be enhanced in the future traction motors for using SiC power MOSFET-based traction converters with high efficiency, at the expense of higher cost, and/or specially designed active gate-driver circuits may be used at the expense of marginally higher switching losses.

6.1.3 Steady-State Performance of the Traction Converter

Line-to-line voltage, line current, DC-link voltage, and DC-link current waveforms of the traction inverter are as given in Figure 6.7 and Figure 6.8, respectively for 765 and 900 V DC catenary voltages while the traction inverter is delivering rated power to the traction motor at rated voltage and frequency, and the optimum switching frequency of 10 kHz. $V_{Al}(t)$ is obtained from $V_{AB}(t)$ by MATLAB/Simulink and plotted on the same figure. The net mechanical output power of the traction motor at this operating point is nearly 125 kW.

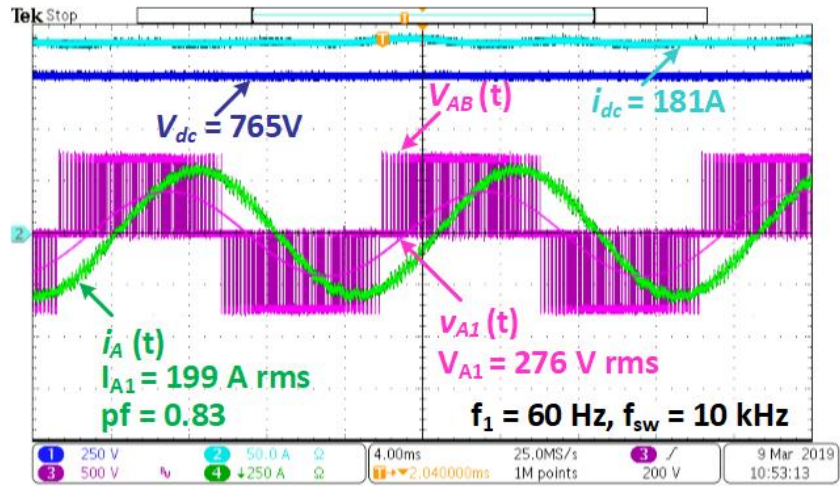


Figure 6.7. Input DC current, i_{dc} and DC-link voltage, V_{dc} and output line-to-line voltage $V_{AB}(t)$, and line current $i_A(t)$ when traction inverter is delivering 165 kVA to the traction motor at 765 VDC

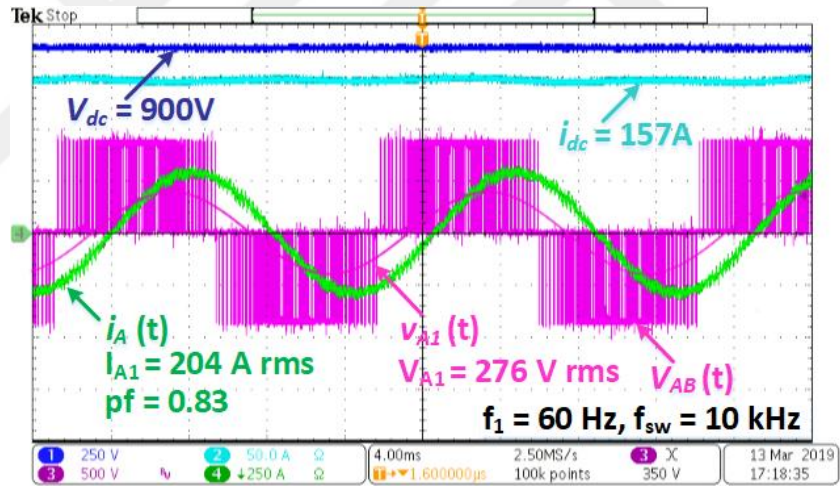


Figure 6.8. Input DC current, i_{dc} and DC-link voltage, V_{dc} and output line-to-line voltage $V_{AB}(t)$, and line current $i_A(t)$ when traction inverter is delivering 169 kVA to the traction motor at 900 VDC

Harmonic components up to 50th component in line current waveforms given in Figure 6.9 are obtained for the operating condition in Figure 6.7 by using Hioki PW3198 Power Quality Analyzer with 3-phase, 3-wire connection. Since the operating condition is rated at $n_r=1779\text{ rpm}$, I-TDD of line currents up to 50th

harmonic is 0.43%. In order to assess the effects of switching frequency harmonics, its integer multiples, and their sidebands on the magnitude of I-TDD, $i_A(t)$ in Figure 6.7 is expanded into FFT by MATLAB/Simulink and given in Figure 6.10. I-TDD value calculated from the harmonic spectrum in Figure 6.14 is 3.11 %. It can then be concluded that harmonic distortion in line current waveforms is largely affected by switching frequency and its integer multiples.

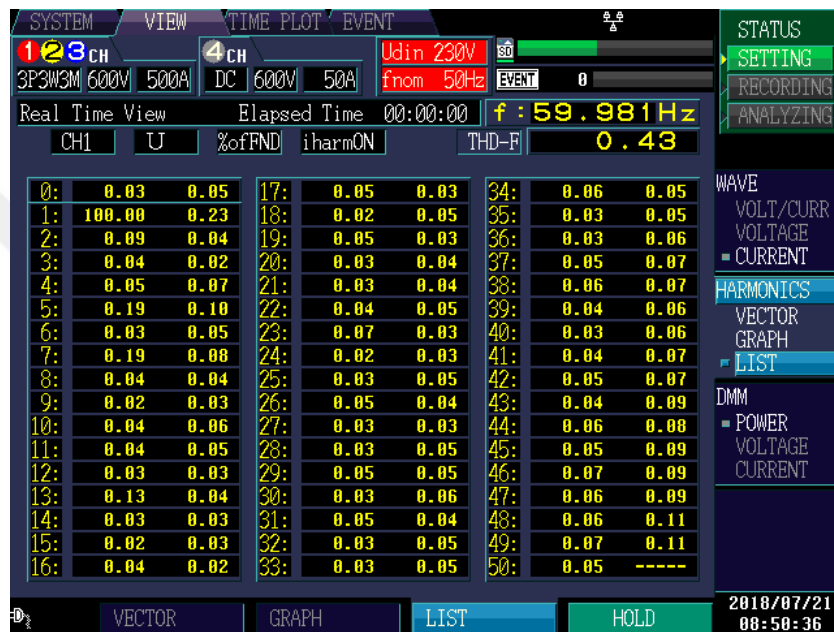


Figure 6.9. Harmonic components of the line currents up to 50th are obtained by a 3-phase, 3-wire connection for the operating condition in Figure 6.7

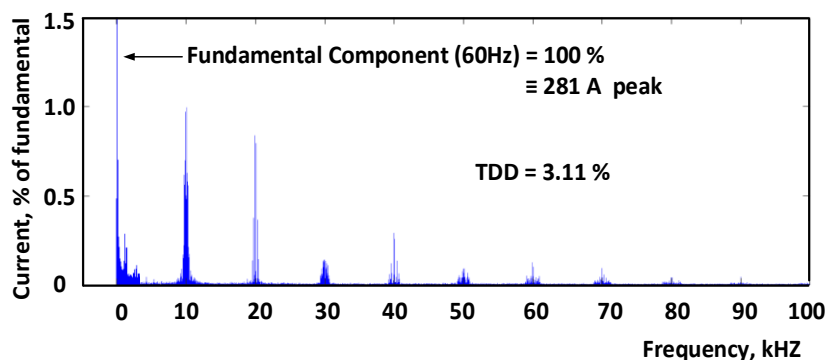


Figure 6.10. Fourier series expansion of $i_A(t)$ in Fig. Figure 6.7 obtained from MATLAB/Simulink FFT toolbox

Percent efficiency of the traction inverter against kVA at stator terminals of the traction machine has been obtained by controlling the loading machine and by setting the shaft speed to 1779 rpm, which requires an applied stator frequency at 60 Hz from partial load to full-load. Theoretical values in Figure 6.11 are calculated by using Wolfspeed's SpeedFit design simulator and also by hand calculations using the datasheet of the SiC power modules. It is observed that both methods give nearly the same results. In constructing the theoretical efficiency graphs in Figure 6.11, their average values are used.

Wolfspeed's SpeedFit design simulator also gives the switching and conduction losses of the traction converter separately, as in Table 6.3 for the same operating points in Figure 6.11. On the other hand, the experimental efficiency characteristics in Figure 6.11 are obtained from the fundamental components of voltage, current, and power factor at the stator terminals of the traction machine on the AC side of the traction inverter, and the mean values of the DC link voltage and current. These quantities are obtained from the raw data collected by Tektronix MDO3034 oscilloscope. Percent efficiency is expressed as the ratio of output power to input power in kW; however, in Figure 6.11, these efficiency variations are given against output kVA since the input power factor of the traction machine varies with the operating condition. Efficiency values are very high as expected, within the measurement accuracy range of transducers. It is observed that experimental efficiency values are nearly 0.2 % lower than theoretical values. This is because parameters of wiring, cabling, and laminated buses are neglected in the theoretical calculations. Furthermore, the efficiency of the traction converter decreases as the line current and hence the kVA on the stator side increases.

Table 6.3 Loss Distribution of Traction Inverter at Motoring Mode

STATOR APPARENT POWER (kVA)	POWER LOSS COMPONENTS (W)	
	Inversion Mode During Motoring	
	Switching	Conduction
31.3	66.5	34.8
62.5	133.6	146.9
93.8	199.8	251.2
125	264.4	534.6
150	329.8	1003.2
176	395.9	1714.1

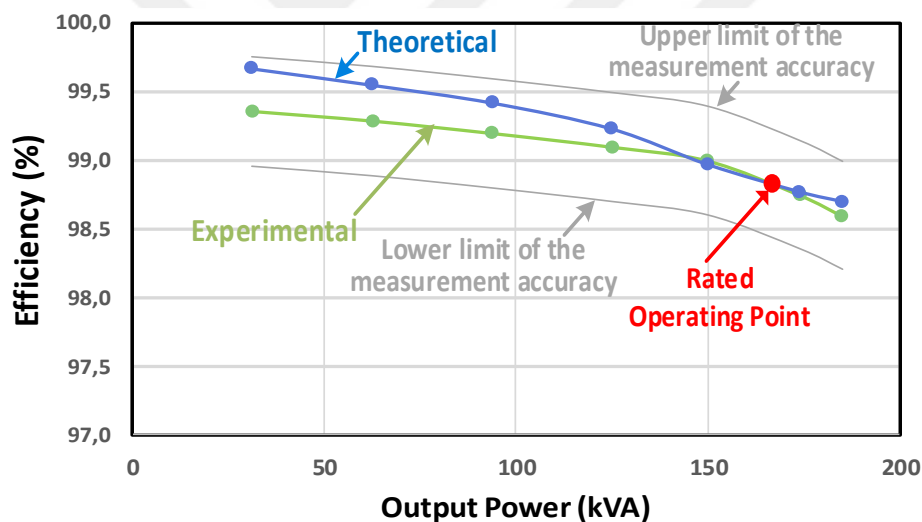


Figure 6.11. Percent efficiency of traction converter against traction machine kVA at rated speed during motoring mode

Thermal behaviors of traction inverter are as given in Figure 6.12. During this test, the ambient temperature remains nearly constant at 30 °C, the traction inverter delivers 165 kVA to the stator of the motor at a line-to-line voltage of 480 V, at nearly 0.83 pf lagging. The heat sink temperature variations are recorded for a time period of one hour by a Type J thermocouple. However, the temperature variations

in case (T_c) and virtual junction (T_{vj}) of SiC power MOSFETs are calculated by using heatsink-to-case and case-to-junction thermal resistances. In the steady-state, all thermal capacitances are fully charged, and hence the temperatures remain constant within a ± 1 deg. band. In Figure 6.12, only the initial 15-min. portion of temperature profiles is given. As can be understood from the steady-state value of T_{vj} of inverter semiconductors, T_{vj} does not exceed 120 °C even for extreme ambient temperature conditions of 45 °C. Since the maximum permissible value of T_{vj} is specified as 150 °C, there is still a considerable amount safety margin.

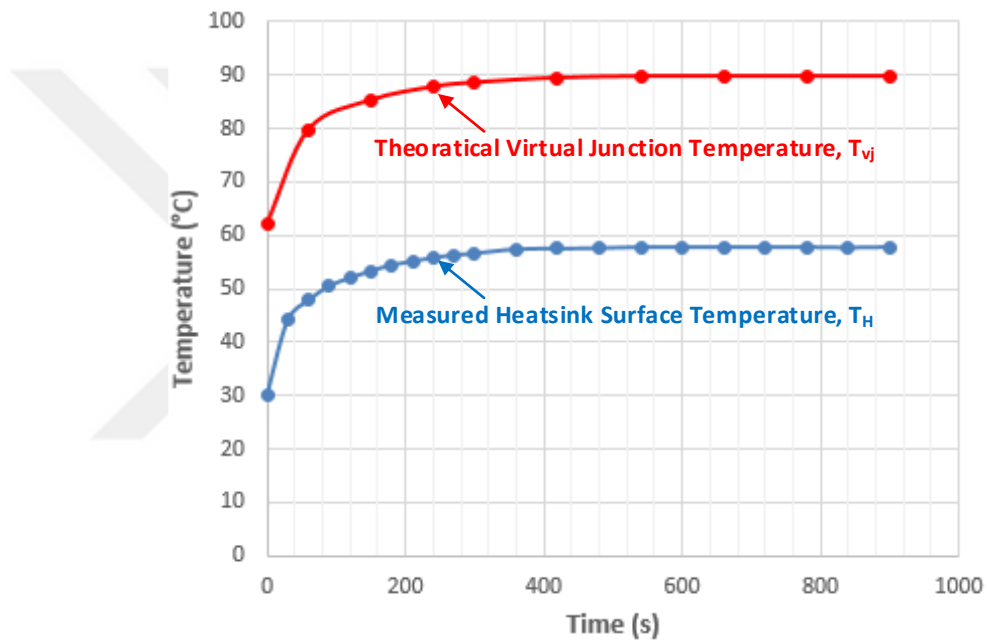


Figure 6.12. Temperature variations of SiC MOSFET virtual junction and heatsink surface against time

6.1.4 Regenerative Braking Mode of Operation

During the operation of a LRTS, the regenerative braking technique is frequently applied in order i. to reduce the running speed, ii. to maintain constant speed operation while moving on an inclined track having a negative slope, and iii. to stop the vehicle at the next station in combination with the mechanical brake. It is a temporary operation mode of the traction motor as a generator and takes place for

not more than a few tens of seconds. The same data set recorded for motoring operation is also recorded for regenerative braking mode as in (ii) and given in Figure 6.13 and Figure 6.14, while the traction inverter is transferring nearly rated power to the catenary during the regenerative braking period.

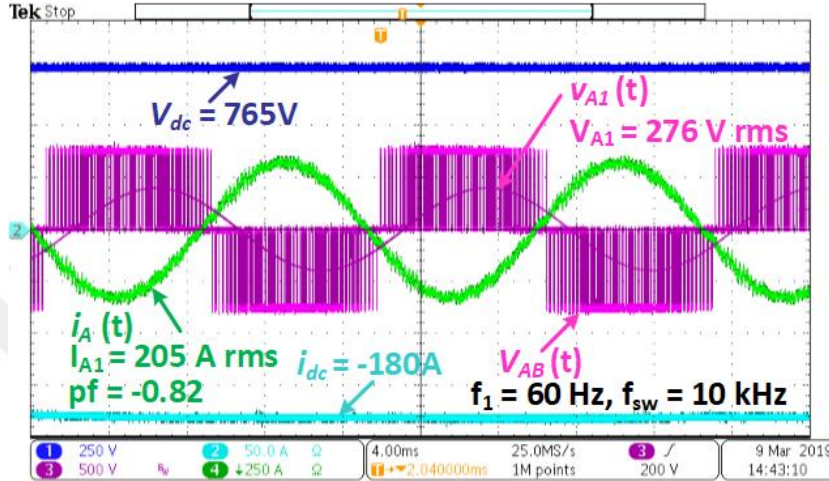


Figure 6.13. Input DC current, i_{dc} and DC-link voltage, V_{dc} and output line-to-line voltage $V_{AB}(t)$, and line current $i_A(t)$ when traction inverter is delivering 138 kW to the catenary at 765 VDC

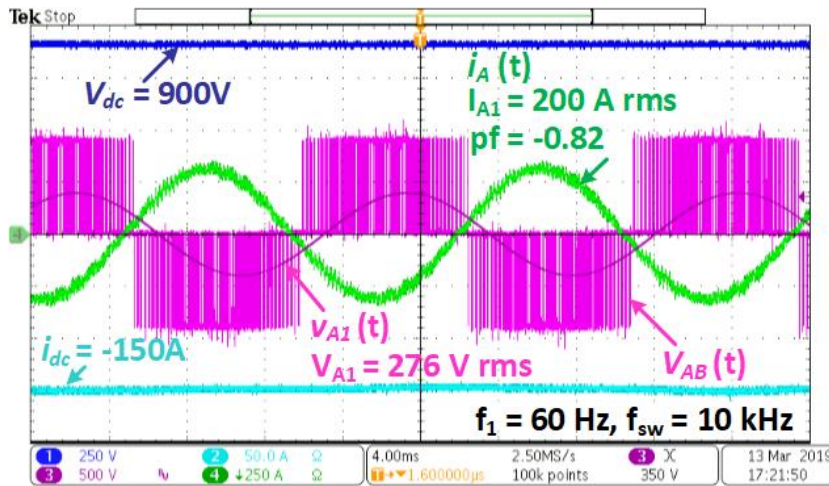


Figure 6.14. Input DC current, i_{dc} and DC-link voltage, V_{dc} and output line-to-line voltage $V_{AB}(t)$, and line current $i_A(t)$ when traction inverter is delivering 139 kW to the catenary at 900 VDC

Similarly, the efficiency graph for regenerative braking mode is constructed, and the loss components of the traction inverter are obtained, as explained in the previous section. The switching loss of the SiC power MOSFETs is directly proportional to instantaneous stator current. Therefore, switching losses are found to be nearly the same for both inversion and rectification modes, as shown in Table 6.3 and Table 6.4. However, this is not the case for conduction losses, which are directly proportional to the square of SiC power MOSFET RMS current value. At high kVAs, the conduction loss component in inversion mode is much higher than the conduction loss in rectification mode. This is because at high operating kVAs, in inversion mode, during the conduction period, mostly the SiC power MOSFETs carry the line current in forward conduction mode, whereas in rectification mode, both SiC power MOSFET in reverse conduction mode and its antiparallel Schottky diode share the current. This causes a marginal reduction in the efficiency of the traction converter in inversion mode in comparison with rectification mode, as given in Figure 6.15.

Table 6.4 Loss Distribution of Traction Inverter at Regenerative Braking Mode

STATOR APPARENT POWER (kVA)	POWER LOSS COMPONENTS (W)	
	Rectification Mode During Regeneration	
	Switching	Conduction
31.3	66.5	31.2
62.5	133.6	112.4
93.8	199.9	145.1
125	264.9	258.1
150	330.9	385.1
176	398.4	536.6

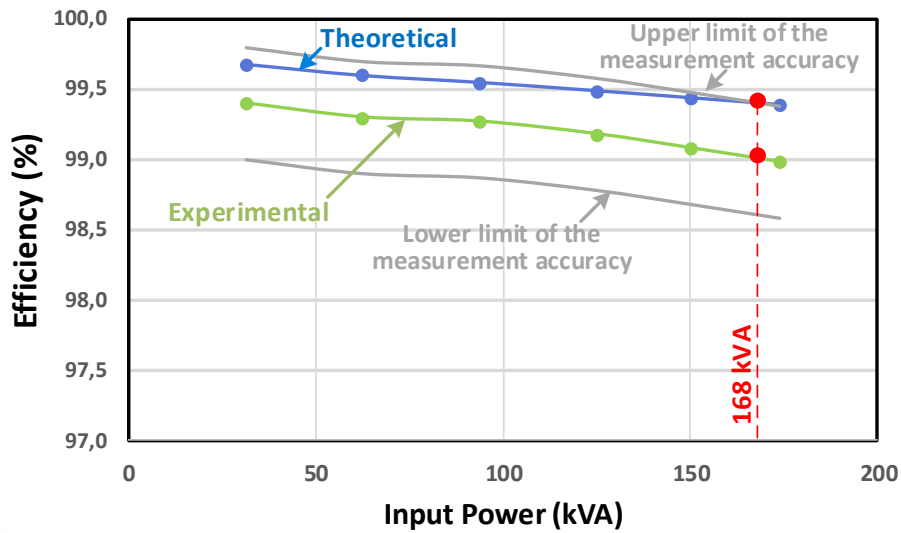


Figure 6.15. Percent efficiency of traction converter against traction machine kVA at rated speed during generating mode

6.1.5 A Closer View of DC-link Voltage and Current

DC-link voltage and line reactor current waveform measurements were already given in previous measurements. The instantaneous DC-link voltage and line reactor current have the following components: the DC component, low-frequency component due to the input filter resonance, load unbalance and catenary voltage fluctuations, and a high-frequency component due to the switched phase currents. In order to see AC components, the oscilloscope settings are switched to AC coupling mode for DC-link voltage and line reactor current measurements. Thus, the DC components of waveforms are canceled. This gives a chance to visually investigate the AC components in closer forms. In Figure 6.16, the AC components of the DC-link voltage and line reactor current waveforms, the line to line output voltage, and line A current waveforms are illustrated.

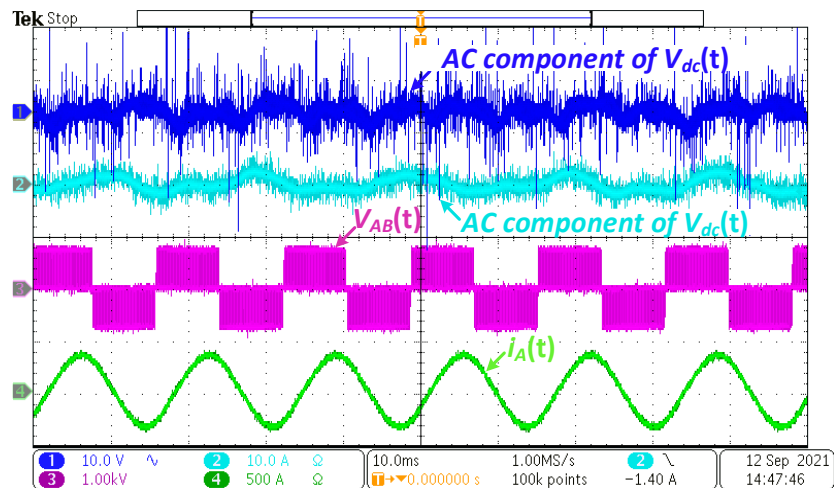


Figure 6.16. DC-link voltage and line reactor ripple in two-level three-phase space vector modulated voltage source traction inverter at 150kW, 480 V l-to-l, 60 Hz, 0.83pf

In this section, closer forms of the DC-link voltage and line reactor current waveforms are given in Figure 6.17 and Figure 6.18. The shape of the high-frequency switching ripples of DC-link voltage depends on the magnitudes of the line reactor, motor phase currents, and power factor.

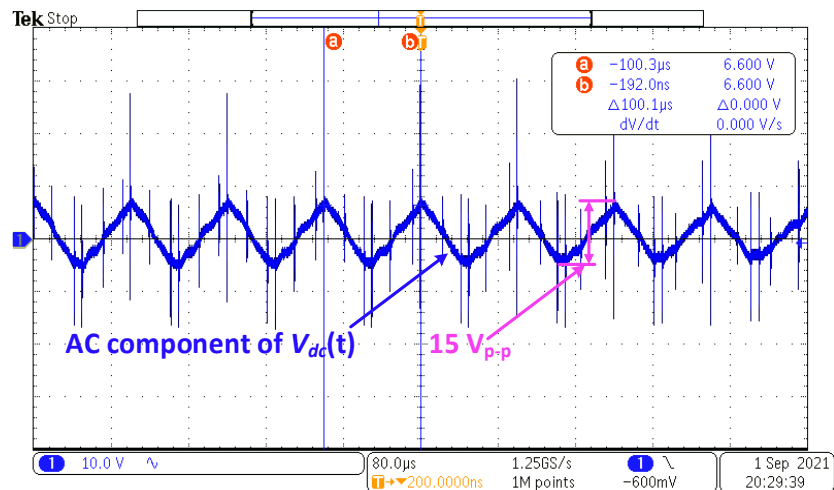


Figure 6.17. The closer view of DC-link voltage, $V_{dc}(t)$

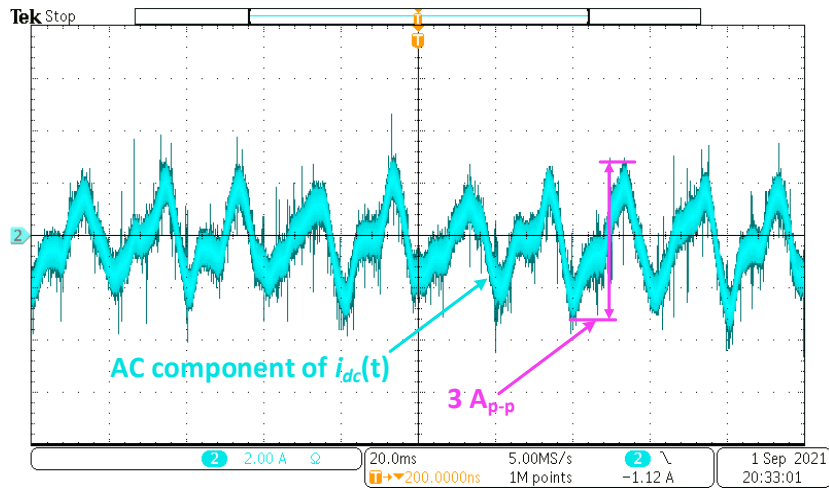


Figure 6.18. The closer view of line reactor current, $i_{dc}(t)$

It is seen that there is no significant high frequency switching current harmonics due to the quite big inductance value of the line reactor. There are only very small peak-to-peak low-frequency current components. In addition to this, low-frequency current components that may occur at the input filter resonance frequency are prevented by the active damping algorithm.

6.1.6 Transient Performance

Detailed experimental studies are carried out to show the performance of developed SiC-based traction inverter. In fact, in a real system, the torque command has a jerk limit in order to satisfy the safety and comfort requirements of the rail vehicle. Thus, the torque commands have major slew rates. This jerk limitation prevents sharp changes at the torque reference side. In order to see the transient behavior under the step torque or power demand changes, jerk slew rates are removed from the control algorithms. And then, the dynamic performance of the inverter in the transient regime was observed by changing the torque reference value between 50 % and 100 %.

In the 375 ms of the measurement in Figure 6.19, the torque reference was changed momentarily from 50 % to 100 %, and the transient performance of the inverter was examined.

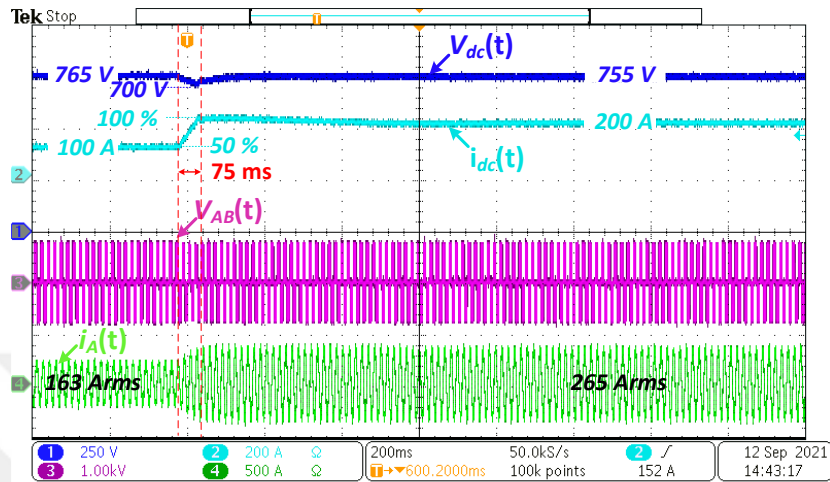


Figure 6.19. Transient response when traction motor torque was changed from 50 % (375 Nm) to 100 % at 480 V l-to-l, 60 Hz operation

In the 370 ms of the measurement record in Figure 6.20, the torque reference was changed momentarily from 100 % to 50 %, and the transient performance of the inverter was presented.

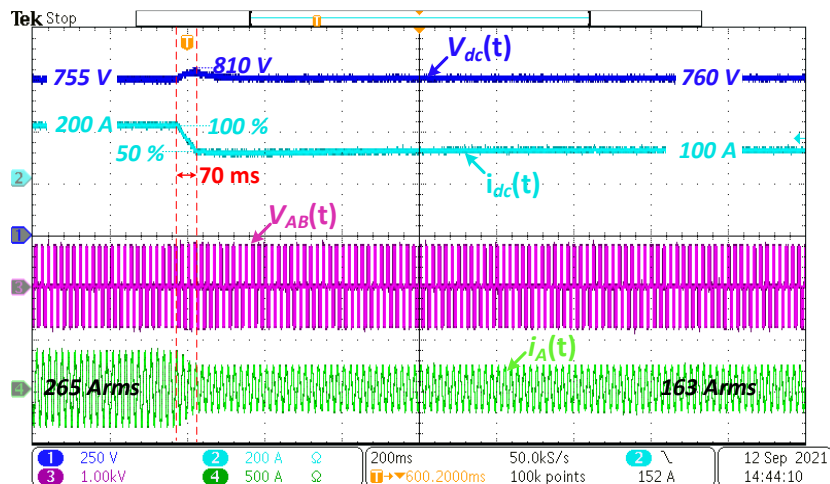


Figure 6.20. Transient response when traction motor torque was changed from 100 % (750 Nm) to 50 % at 480 V l-to-l, 60 Hz operation

6.1.7 Dynamic Operation on Real Rail-Track Conditions

The physical simulator system should have the capability of close representation of a real LRTS, in both the steady-state operation at a constant speed and dynamic operation state during acceleration and regenerative braking. In order to verify this, the data of a real elliptic track was uploaded to the rail track model. The 5850 m long track has 12 stations. The physical simulator in Table 6.5 is then run on this track, and the variations in speed, torque, and power per traction motor are plotted against time as given in Figure 6.21. Station-to-station distances and the variations in slope on this track are also given in the same figure.

Table 6.5 Technical Specifications of Physical Simulator

4 – Pole, 60 Hz, 475 V, 125 kW, 0.81 pf (S1), η = 93.5%		
Traction Motor (VEM – DKCBZ 0212-4)	Rated Torque	671 Nm
	Rated Speed	1779 rpm
	Max. Torque	940 Nm
	Max. Speed	5100 rpm
4 – Pole, 60 Hz, 430 V, 130 kW, 0.82 pf (S1), η = 93.7%		
Active Loading Machine (VEM – DKOBZ 0610-4)	Rated Torque	700 Nm
	Rated Speed	1775 rpm
	Max. Torque	1200 Nm
	Max. Speed	5500 rpm
Moment of Inertias	J_m	0.67 kg- m ²
	J_g	0.84 kg- m ²
	J_{fw}	30 kg-m ²
	Double Flex Couplings +Bearings + Rigid Shafts	0.15 kg- m ²
Vehicle Demand per Motor for the Chosen Track	Average Torque	500 Nm
	Average Speed	1500 rpm
	Average Power	80 kW
	Max. Torque	943 Nm
	Max. Speed	2250 rpm
	Max. Power	128 kW

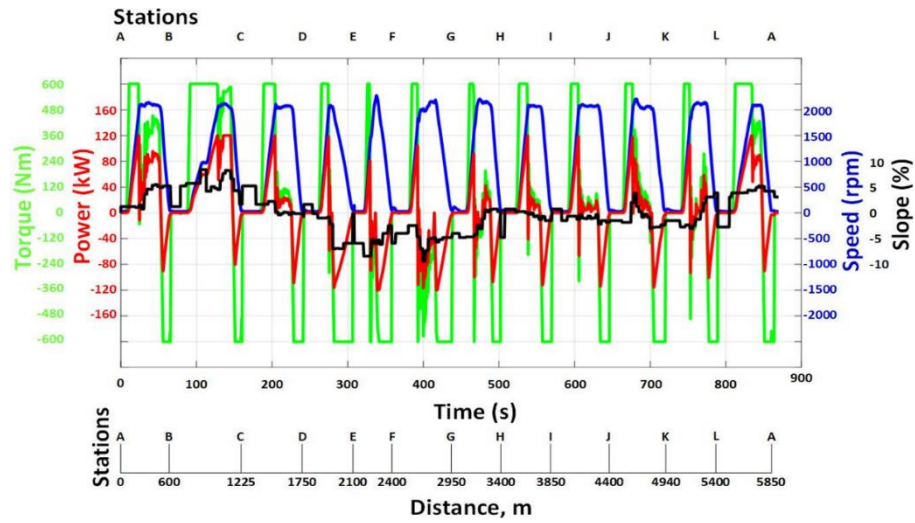


Figure 6.21. Records of the physical simulator in Table 6.5 while the light rail vehicle is running on a real elliptic track

In order to test the performance of the traction system developed in this thesis and hence the vehicle body structure formed as in Table 3.1, on real rail-track conditions, the physical simulator per traction motor is operated for several running conditions. Among these, records corresponding to acceleration and deceleration on a level track, acceleration on a maximum positive slope and, regenerative braking on a maximum negative slope are plotted.

The first test consists of the acceleration, constant speed operation, and deceleration instants on a level track, and the results are illustrated in Figure 6.22. In Figure 6.22 (a), traction motor shaft speed, traction motor shaft torque, and load motor torque variations against time are plotted, while one phase current of traction motor is exhibited in Figure 6.22 (b).

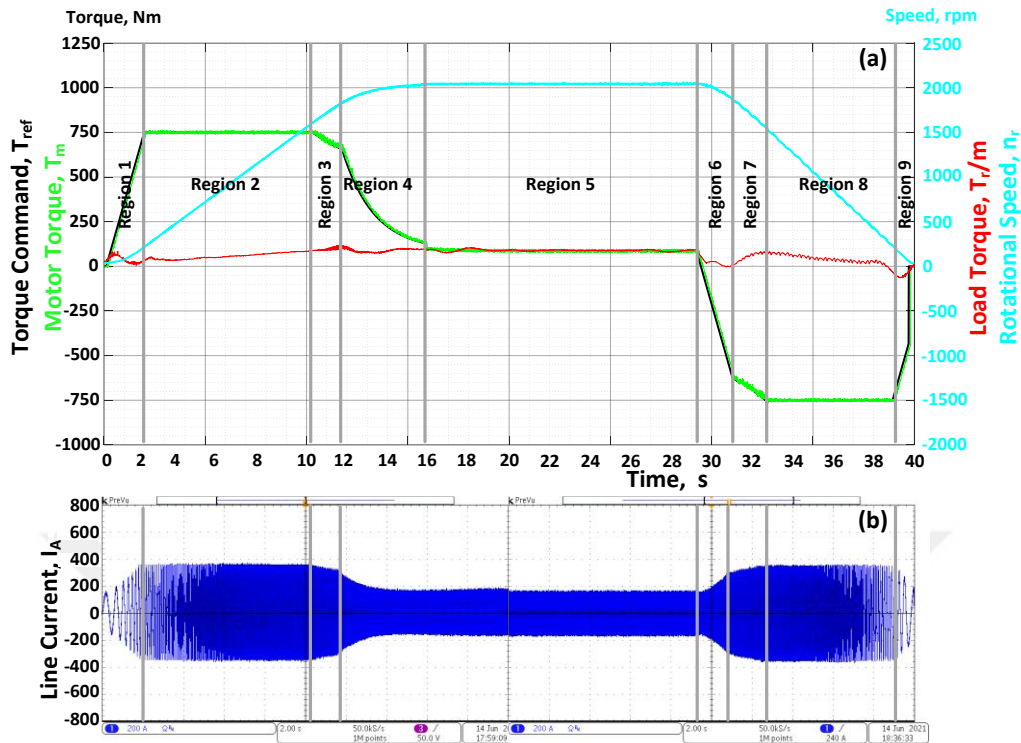


Figure 6.22. Acceleration, constant speed operation, and then deceleration on a level track (Region 1: Starting period, Region 2: Acceleration at maximum torque, Region 3: Acceleration at constant power, Region 4: Controller action for a smooth transition from acceleration to constant speed operation, Region 5: Constant speed operation, Region 6: Starting of regenerative braking operation, Region 7: Deceleration at constant power, Region 8: Deceleration under maximum braking torque, Region 9: Bringing the vehicle to a stop by the mechanical brake.)

At time zero, the traction force (e.g., traction motor torque command) is applied within the jerk limits. In two seconds, the torque reference (i.e., black colored signal) of the traction motor reached its maximum value. The green one is the applied torque feedback which is calculated in the traction inverter. It is seen that the torque reference and feedback seem to overlap during the test period. The load torque on a level track represents the train resistance. The inertia of the vehicle is represented by the flywheel. During fast torque changes, there is a deviation in the applied torque values of the load motor due to a fault in the industrial motor driver. However, it does not affect the driving performance or the traction inverter performance. It only

has a negligible effect on acceleration and deceleration times. Moreover, the tests were continued for two different conditions. This time the tests were carried out for the toughest conditions for the traction drive system. These tests covered the scenarios of starting and accelerating under the positive maximum slope and braking under the maximum negative slope. The results are illustrated in Figure 6.23 and Figure 6.24, respectively. Similar to Figure 6.22 (b), one-phase motor currents are also given in Figure 6.23 (b) and Figure 6.24 (b).

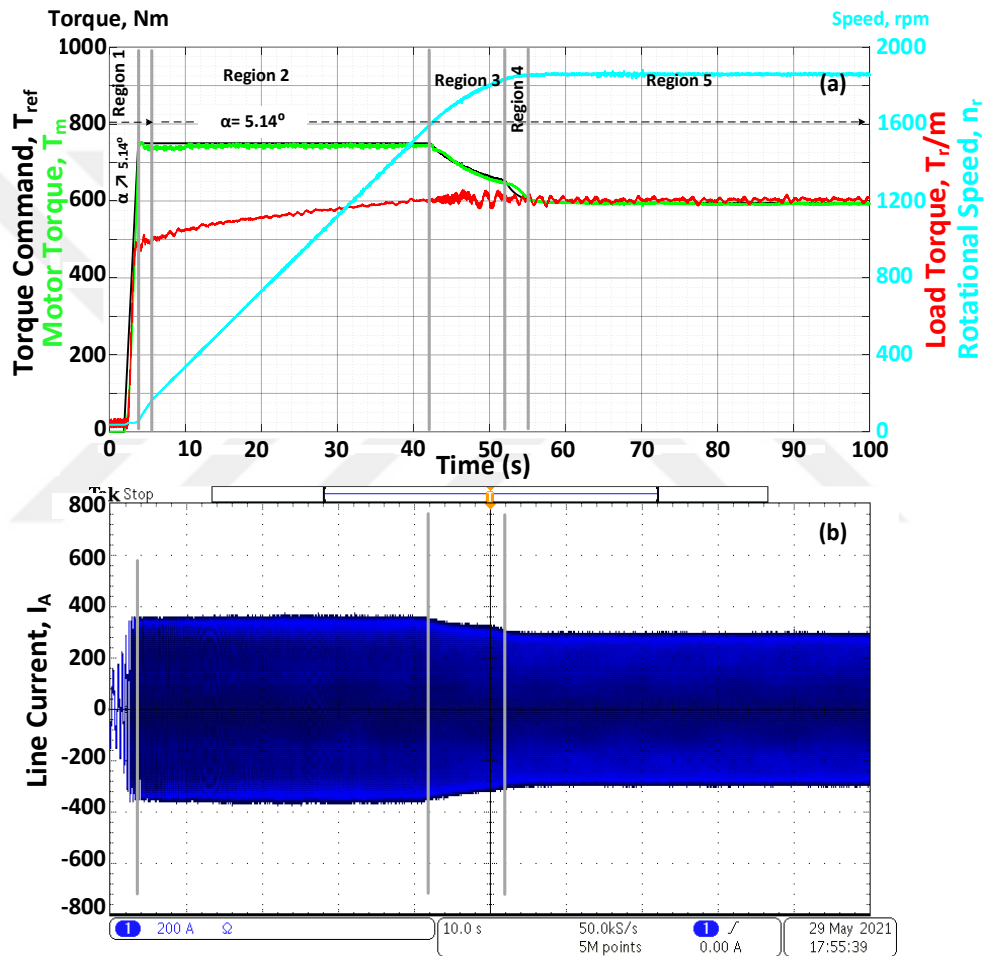


Figure 6.23. Acceleration and then constant speed operation on a track having maximum positive slope (Region 1: Starting period, Region 2: Acceleration at maximum torque, Region 3: Acceleration at constant power, Region 4: Controller action for a smooth transition from acceleration to constant speed operation, Region 5: Constant speed operation)

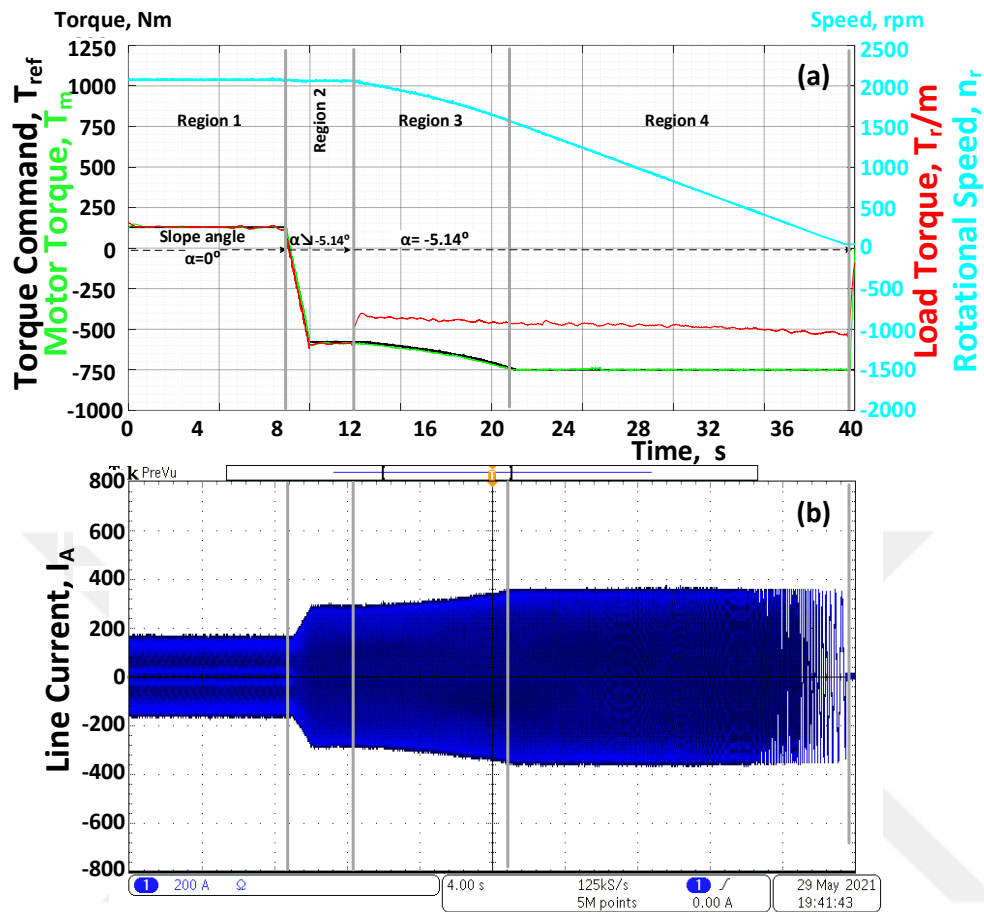


Figure 6.24. Constant speed operation and then deceleration by regenerative braking while going downhill (Region 1: Constant speed operation, Region 2: Starting of regenerative braking operation, Region 3: Deceleration at constant power, Region 4: Deceleration under maximum braking torque, Region 5: Bringing the vehicle to a stop by the mechanical brake.)

In the last two tests, maximum slope conditions lead to produce maximum force in the opposite direction. In Figure 6.23, the load torque equals to the sum of two opposite forces opposing the acceleration of the vehicle, namely the train resistance and gravitational force. On the other hand, in Figure 6.24, the load torque represents the difference between gravitational force and train resistance. In both cases, the traction drive system encounters a very strong opposing force. For this reason, the acceleration and stopping times of the vehicle are considerably longer.

The above operating conditions are programmed on MATLAB/Simulink by using the kinematic equation in (2-12) of the vehicle to form the associate the real-time models to control the physical simulator system in Figure 5.1. Kinematic equations for the running conditions given in Figure 6.22, Figure 6.23, and Figure 6.24 are obtained from (2-12) as given respectively in (6-1), (6-2), and (6-3). A closer examination of motor torque, load torque, torque command, shaft speed, and line A current waveforms in Figure 6.22 - Figure 6.24 have shown that all SiC traction motor drive and its control system developed in this thesis matches perfectly the requirements of the pre-specified rail track conditions.

$$258.5 \frac{d\omega}{dt} = T_m - \frac{1}{4} T_r \quad (6-1)$$

$$258.5 \frac{d\omega}{dt} = T_m - \frac{1}{4} [T_r + 1968] \quad (6-2)$$

$$258.5 \frac{d\omega}{dt} = T_m - \frac{1}{4} [T_r - 1968] \quad (6-3)$$

CHAPTER 7

FUTURE PROSPECTS OF SIC POWER MOSFETS IN POWER CONVERTERS

Increasing the efficiency of power converters has been one of the most important issues that industry and academy have been working on for years. Today, the increasing need for electrical energy use significantly creates more pressure for the new pursuits to achieve increased power conversion efficiency. In academic studies and thousands of exemplary applications in the industry, it is seen that Si-based power semiconductors have reached their design limits in terms of breakdown voltages, conduction, and switching losses. In order to overcome these problems, the solutions based on parallel and serial use of semiconductors have been derived for years. However, these solutions increase the system cost, bring application difficulties and reduce product reliability due to the increase in the number of components.

On the other hand, years of work have led to the widespread use of gallium nitride (GaN) and silicon carbide (SiC) based wide bandgap power devices for high voltage and power applications, which significantly improve the breakdown voltage, conduction, and switching losses. The future perspective of GaN and SiC power semiconductors are given in Figure 7.1.

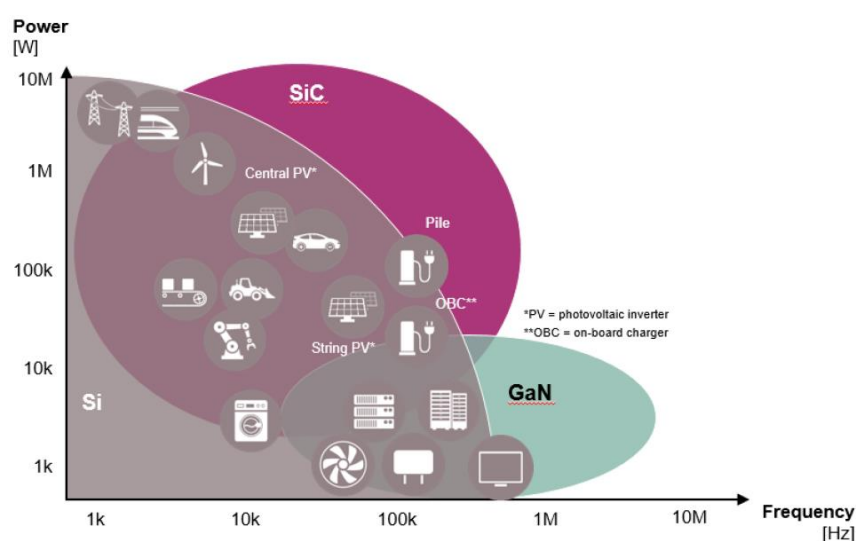


Figure 7.1. The future perspective of GaN and SiC power semiconductors[103]

While SiC semiconductors appear in applications with very high power and high switching frequency requirements, on the other hand, GaN semiconductors are seen in relatively low power and very high switching frequency applications. Today, due to their undeniable advantages, SiC power MOSFETs have started to show themselves in many high power applications such as energy storage systems, industrial drives, electric vehicle (EV) fast-charging stations, traction drives, industrial switched-mode power supply (SMPS), and automotive powertrains. As the subject of the thesis, these current developments are examined under the application field of traction motor drives.

With the 500 - 900 VDC, 165 kVA traction motor drive designed within the scope of this thesis, 98.7 % efficiency was obtained on the basis of the converter at 10 kHz switching frequency and 25 °C ambient temperature with a considerably small semiconductor base area. In addition to this, 91.7 % efficiency was obtained for traction motor drive. The total losses of traction motor and traction inverter for 2.5 kHz and 10 kHz switching frequencies and the percent efficiency of traction motor drive against switching frequency at rated operating conditions are given in Figure 7.2 (a) and (b).

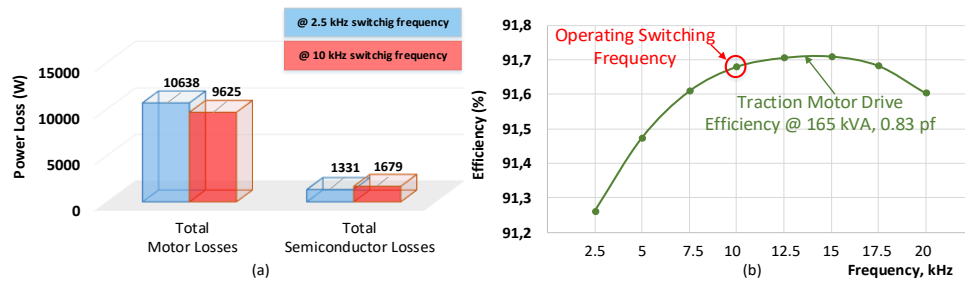


Figure 7.2. (a) Total traction motor and traction inverter losses for 2.5 kHz and 10 kHz switching frequencies (b) Percent efficiency of traction motor drive against switching frequency

The total loss of traction motor is decreased by almost 1 kW by changing switching frequency from 2.5 kHz to 10 kHz, as shown in Figure 7.2 (a). It facilitates traction motor cooling and helps to extend the lifetime of the winding insulation materials. On the other hand, it causes an increase of nearly 350 W for the total power loss of the traction inverter. Considering this power loss increase per switch, it corresponds to an increase of about 58W and does not create unmanageable thermal stress on the switches. As can be seen from Figure 7.2 (b), the total motor drive efficiency can be improved further by almost 0.5 % with the switching frequency optimization. At the 10 kHz operating point, the calculated junction temperature of MOSFETs approaches to 110 °C while the traction drive is operating at an ambient temperature of 45 °C in nominal working conditions (i.e., 480 V_{line-to-line}, 165 kVA, and 60 Hz). The junction temperatures at 12.5 kHz and 15 kHz switching frequencies are 113 °C and 117 °C, respectively. In order not to shorten the lifetime of the switches, the junction temperatures of the chips are kept minimum as much as possible. Each 10 °C increase in junction temperatures causes half the lifetime of the semiconductor. Moreover, further increase of switching frequency beyond the 10 kHz does not satisfy considerable advantages to improve system efficiency. Thus, 10 kHz switching frequency was determined as optimum switching frequency, as explained in Chapter 3.

In addition to the increase in total system efficiency, with the use of SiC power MOSFET technology, many benefits have been obtained within the scope of this thesis. The main ones are:

- Convenience in control method and modulation methods,
- An increase in system stability due to the increased control bandwidths,
- Lower unit cost due to the decrease in the cooling system requirements and component numbers,
- Increase in the reliability due to the decrease in the number of components and high-temperature strength features of SiC power MOSFETs,
- Decrease in operating cost depending on energy efficiency and maintenance/repair cost.

Let's try to summarize how the benefits are obtained with the following evaluations. When Table 3.6 is reviewed again, it is seen that Si and hybrid-IGBT solutions do not meet the target requirements for a high-performance traction inverter, even though they have the same or superior current carrying capacity and package sizes. The main reason for this is that the selected Si- and hybrid-IGBT modules cannot be switched at a switching frequency of 10 kHz due to the die junction temperatures reaching outside of the operating band. It is seen that they are suitable for operation at lower switching frequencies. However, low switching frequencies present many challenges. Due to low switching frequency, increasing current harmonic distortions with vehicle speed both worsen the system efficiency and worsen the system stability due to increased torque ripple. In order to optimize the current ripples, asynchronous PWM is switched to synchronous PWM with the increased vehicle speed and then to programmed modulation methods at a higher speed, as shown in Figure 7.3.

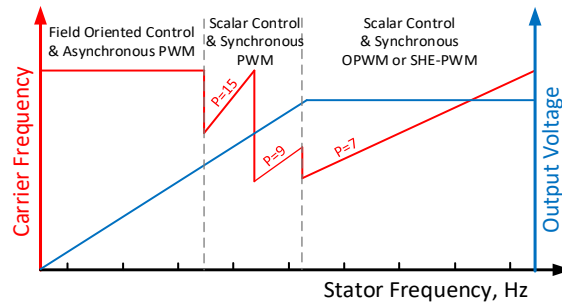


Figure 7.3. Conventional motor control and PWM techniques in railway traction

In parallel with the changing modulation method, it is necessary to use more than one control method in motor control for system stability and compatibility of the control method and modulation technique. In order to avoid all these challenging, complex operations, a traction inverter design with a high switching frequency is targeted.

In order to meet the target requirements of a high-performance traction drive with Si-IGBT-based inverters, it is necessary to use parallel semiconductors or multilevel topologies. In both solutions, there is an increase in volume, weight, and cost, and a decrease in reliability depending on the increased number of components. The four different configurations in which the desired switching frequency is obtained are shown in Figure 7.4 in terms of losses and semiconductor base areas. By paralleling the three units of the CM300DY-34T module, the targeted switching frequency was achieved with Si-IGBT-based two-level inverter topology.

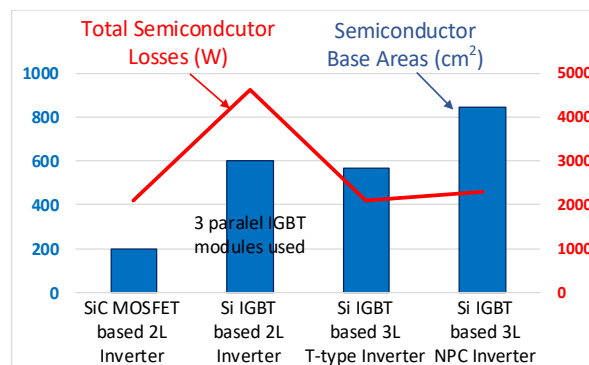


Figure 7.4. Losses and semiconductor base areas for applicable four different inverter configurations

In Figure 7.4, it can be seen that the SiC MOSFET-based two-level inverter is the best topology in terms of efficiency and semiconductor base area. 10 kHz switching frequency gives a chance to use common SVPWM as a modulation technique and FOC method for motor control in all operating regions.

Many analyses and evaluations have been made within the scope of this thesis to show that the benefits to be obtained with multilevel converters can also be achieved with SiC MOSFET-based two-level inverters. Moreover, additional advantages such as volume, weight, and reliability obtained with SiC MOSFET-based inverter are also reported.

In addition to these, there are also studies that are still the subject of research and whose distinction cannot be clearly demonstrated. One of these problems is electromagnetic interference, and the other is related to the operating temperature. In fact, there will be a positive reflection on electromagnetic interference due to the almost zero reverse recovery charge (e.g., lower peak switching current). However, parasitic capacitances also need to be optimized to avoid common-mode currents that can occur with increasing switching speed. Moreover, due to increased switching dynamics, parasitic inductances should be further reduced compared to existing power modules in order to fulfill the electromagnetic compatibility requirements and to guarantee operation in the safe operating area of semiconductors. High switching speeds also have a great strain on motor windings insulations and bearings. Therefore, it may be necessary to improve the motor winding insulation materials. These are the points that need extra attention from an electrical point of view. As for the evaluation of operating temperature, if the targeted very high operating temperatures can be achieved with SiC MOSFETs, it is considered that highly innovative solutions will be needed in cooling system structures and laminated busbar designs. For example, if operating temperatures of 400-500 °C are reached, there may be serious increases in radiative heat transfer. It is thought that this change will require design changes in many hardware, especially laminated busbar and cooling system structures.

CHAPTER 8

CONCLUSIONS AND FUTURE WORKS

The design and implementation of 165 kVA all SiC MOSFET-based traction inverter has been carried out in this research work. The prototype has been designed and developed according to the sample light rail transportation vehicle properties, its track, and operating profile. In order to validate the produced prototype, a 150kW full-scale physical simulator which consists of a motor test bench, has been successfully implemented in a low power laboratory, which consumes active power from the AC grid much smaller than the installed capacity of the simulator to account only for the losses. The developed prototype was validated by using this simulator.

The developed physical simulator can be used in the design, technology development, and performance testing of various types of traction inverters. The catenary model, rail-track model, and vehicle model have been satisfactorily integrated into the developed system to allow the performance verification of a vehicle formation on a real track in real-time. The use of SiC power MOSFETs reduces the volume and improves efficiency and representation of transient changes in catenary voltage in comparison with Si-IGBT devices in traction applications. Since SiC MOSFETs can be switched at higher frequencies with much shorter turn-on and turn-off times in comparison with Si-IGBT based traction converters, EMI considerations such as shorter cabling, ceramic motor bearings, optimized laminated bus, silver-plated flat braid grounding cables, and etc., are considered as an important issue in the design of the simulator.

At the present time, converter size, weight reduction issues, and energy efficiency improvements by employing new wide bandgap power semiconductors are the most attractive topics. Additionally, the electromagnetic compatibility and reliability issues of AC drives with wide bandgap power semiconductors such as GAN and SiC are popular and intensive research areas. Many significant academic studies have

been conducted on these subjects over the ten years. However, performance issues involving the effects on other system components in AC traction drives (e.g., motors, cables, transformers, etc.) are important and yet not fully investigated and addressed. Operating an inverter with high dv/dt rates is problematic for the insulation of motor windings. The current harmonics are significantly reduced with the increased frequency. On the other hand, the frequency of exposure to a high dv/dt rate is also increasing. This situation will create a considerable effect on the lifetime of electric motors. New generation traction motors with insulating materials having dv/dt ratings higher than $10 \text{ kV}/\mu\text{s}$ should be commercially available in the future in order to take full advantage of SiC power converters. Moreover, the effects of wide bandgap power semiconductors on optimum converter topology selection (e.g., two-level, three-level, etc.), contributions of wide bandgap power semiconductors to the FOC, direct torque control algorithms, and programmed PWM techniques that are needed in medium voltage high power applications are significant and wait for clarification.

Following conclusions can be summarized from the results of theoretical and experimental works carried out within the scope of this thesis:

- The conversion efficiency of traction converter is obtained higher than 98.7 % with 10 kHz switching frequency in both motoring and regenerative braking mode of operations.
- Very high quality current waveforms were obtained. Motor current waveform THD was observed as 3.11 %.
- The motor losses have been reduced by about 1 kW (e.g. 0.7 % improvement).
- The overall system efficiency (e.g. traction conv. & motor) is increased almost 2 %.
- Operating cost is decreased due to the increased system efficiency.
- The system stability and dynamic performance of the system are increased.

- Only one modulation and control technique were used in all operating regions. Therefore, controller design has been simplified.
- The audible noise level is reduced due to the high switching frequency.
- Mechanical stresses on motor shaft, couplings, gearboxes are improved due to the decreased torque ripple.
- It has been observed that the power and voltage levels at which the transition to multilevel converters will be increased in the case of using SiC MOSFET-based semiconductors.
- This work showed that SiC semiconductors allowed the use of simple converter topologies in higher power and voltage levels.
- The voltage waveforms at the motor terminals were examined and it was observed that the dv/dt rates exceeded the design limits of current traction motors. It has been revealed that with the further development of the SiC semiconductors, even more challenging conditions will arise. It was emphasized that the motor winding insulation material should be improved.
- Fully isolated motor bearings should be used due to the increased frequency of common mode voltages and currents.

In this thesis, all the above-mentioned issues are investigated to make contributions to the high power and medium voltage traction drive applications. Although this review gives an idea of the future benefits of SiC MOSFETs, the boundary on which they will offer advantages over power and voltage levels is not clearly defined due to the limited variety of SiC power semiconductor products. In addition to this, as future studies, after the product variety of SiC MOSFETs is sufficiently increased, the distortion effects caused by non-linear effects of power semiconductors on the output current and voltage waveforms of SiC MOSFETs can be examined and compared to IGBTs. Moreover, the reduced die area and thickness with the SiC materials yields to decrease the thermal capacitances of semiconductors. Because of this, the aging due to the thermal cycles is very important and waiting for an

investigation. Additionally, the short circuit current capacity of the SiC power modules should be improved for railway traction applications, or advanced gate-driver which satisfy very fast short circuit protection should be designed. The increased switching speeds increase the voltage spikes and leakage currents, and hence increasing electromagnetic interference. Finally, passive filters and active gate-drivers can be considered to reduce dv/dt rates at motor terminals then these solutions can be compared for system efficiency, cost, EMI and reliability.



REFERENCES

- [1] “Stream Locomotives” https://en.wikipedia.org/wiki/Steam_locomotive.
- [2] M. C. Duffy, 1880 – 1990 Electric Railways, 1990.
- [3] A. Pankhurst, Electric Traction Systems 2014, Published by The Institution of Engineering and Technology, London, United Kingdom, 2014. 2014.
- [4] T. E. Salem and R. A. Wood, “1000-H Evaluation of a 1200-V, 880-A All-SiC Dual Module,” IEEE Trans. Power Electron., vol. 29, no. 5, pp. 2192–2198, May 2014.
- [5] A. Merkert, T. Krone, and A. Mertens, “Characterization and Scalable Modeling of Power Semiconductors for Optimized Design of Traction Inverters with Si- and SiC-Devices,” IEEE Trans. Power Electron., vol. 29, no. 5, pp. 2238–2245, May 2014.
- [6] M. Imaizumi et al., “Remarkable Advances in SiC Power Device Technology for Ultra High Power Systems,” in 2013 IEEE International Electron Devices Meeting, pp. 6.5.1-6.5.4, Dec. 2013.
- [7] S. Hazra, S. Madhusoodhanan, G. K. Moghaddam, K. Hatua, and S. Bhattacharya, “Design Considerations and Performance Evaluation of 1200-V 100-A SiC MOSFET-Based Two-Level Voltage Source Converter,” IEEE Trans. Ind. Appl., vol. 52, no. 5, pp. 4257–4268, Sep. 2016.
- [8] E. Wiesner, K. Masuda, and M. Joko, “New 1200V Full SiC Module with 800A Rated Current,” in 2015 17th European Conference on Power Electronics and Applications (EPE’15 ECCE-Europe), pp. 1–9, Sep. 2015.
- [9] Brenna, M., Foiadelli, F. and Zanineli, D., Electrical Railway Transportation Systems, Published by John Wiley & Sons, Inc., Hoboken, New Jersey, 2018.
- [10] M. Steiner, M. Klohr, and S. Pagiela, “Energy Storage System With Ultracaps

On Board of Railway Vehicles,” in 2007 European Conference on Power Electronics and Applications, pp. 1–10, 2007.

- [11] L. Mir, I. Etxeberria-Otadui, I. P. de Arenaza, I. Sarasola, and T. Nieva, “A Supercapacitor Based Light Rail Vehicle: System Design and Operations Modes,” in 2009 IEEE Energy Conversion Congress and Exposition, pp. 1632–1639, Sep. 2009.
- [12] A. L. Allegre, A. Bouscayrol, J.-N. Verhille, P. Delarue, E. Chattot, and S. El-Fassi, “Reduced-Scale-Power Hardware-in-the-Loop Simulation of an Innovative Subway,” *IEEE Trans. Ind. Electron.*, vol. 57, no. 4, pp. 1175–1185, Apr. 2010.
- [13] A. L. Allegre, A. Bouscayrol, P. Delarue, P. Barrade, E. Chattot, and S. El-Fassi, “Energy Storage System With Supercapacitor for an Innovative Subway,” *IEEE Trans. Ind. Electron.*, vol. 57, no. 12, pp. 4001–4012, Dec. 2010.
- [14] D. Iannuzzi and P. Tricoli, “Speed-Based State-of-Charge Tracking Control for Metro Trains With Onboard Supercapacitors,” *IEEE Trans. Power Electron.*, vol. 27, no. 4, pp. 2129–2140, Apr. 2012.
- [15] W. Wang et al., “A Novel Energy Management Strategy of Onboard Supercapacitor for Subway Applications With Permanent-Magnet Traction System,” *IEEE Trans. Veh. Technol.*, vol. 63, no. 6, pp. 2578–2588, Jul. 2014.
- [16] “EN 50388:2012 European Standard ‘Railway Applications - Power Supply And Rolling Stock - Technical Criteria For The Coordination Between Power Supply (Substation) And Rolling Stock To Achieve Interoperability’, 2012”.
- [17] “EN 50163:2004 European Standard. ‘Railway Applications - Supply Voltages Of Traction Systems’, November 2004”.
- [18] R. J. Hill, “Electric railway traction. Part 1: Electric Traction and DC Traction Motor Drives,” *Power Eng. J.*, vol. 8, no. 1, pp. 47–56, Feb. 1994.

- [19] R. J. Kemp, "Developments in Electric Traction," *Power Eng. J.*, vol. 3, no. 2, p. 71, 1989.
- [20] D. Ronanki, S. A. Singh, and S. S. Williamson, "Comprehensive Topological Overview of Rolling Stock Architectures and Recent Trends in Electric Railway Traction Systems," *IEEE Trans. Transp. Electr.*, vol. 3, no. 3, pp. 724–738, Sep. 2017.
- [21] "<https://www.mitsubishielectric-railway.eu/products-solutions/traction/vvfvf-inverter>".
- [22] S. A. Azmi, K. H. Ahmed, S. J. Finney, and B. W. Williams, "Comparative Analysis Between Voltage and Current Source Inverters in Grid-Connected Application," in *IET Conference on Renewable Power Generation (RPG 2011)*, pp. 101–101, 2011.
- [23] Y. Suh, J. K. Steinke, and P. K. Steimer, "Efficiency Comparison of Voltage-Source and Current-Source Drive Systems for Medium-Voltage Applications," *IEEE Trans. Ind. Electron.*, vol. 54, no. 5, pp. 2521–2531, Oct. 2007.
- [24] Abad, G., 'Power Electronics and Electric Drives for Traction Applications', Published by John Wiley & Sons, Inc., UK, 2017.
- [25] D. Busse, J. Erdman, R. J. Kerkman, D. Schlegel, and G. Skibinski, "The Effects of PWM Voltage Source Inverters on The Mechanical Performance of Rolling Bearings," in *Proceedings of Applied Power Electronics Conference. APEC '96*, vol. 2, pp. 561–569, 1996.
- [26] S. Ogasawara and H. Akagi, "Modeling and Damping of High-Frequency Leakage Currents in PWM Inverter-Fed AC Motor Drive Systems," *IEEE Trans. Ind. Appl.*, vol. 32, no. 5, pp. 1105–1114, 1996.
- [27] D. Busse, J. Erdman, R. J. Kerkman, D. Schlegel, and G. Skibinski, "System Electrical Parameters and Their Effects on Bearing Currents," *IEEE Trans.*

Ind. Appl., vol. 33, no. 2, pp. 577–584, 1997.

- [28] L. Franquelo, J. Rodriguez, J. Leon, S. Kouro, R. Portillo, and M. Prats, “The Age of Multilevel Converters Arrives,” *IEEE Ind. Electron. Mag.*, vol. 2, no. 2, pp. 28–39, Jun. 2008.
- [29] “States-of-the-art-review,” 2019, https://projects.shift2rail.org/s2r_matrixtd.aspx.
- [30] ECPE, “Position Paper Energy Efficiency – the Role of Power Electronics,” 2007.
- [31] ECPE, “Next Generation Power Electronics based on Wide Bandgap Devices - Challenges and Opportunities for Europe, ECPE Position Paper,” 2016.
- [32] J. Rodriguez et al., “Multilevel Converters: An Enabling Technology for High-Power Applications,” *Proc. IEEE*, vol. 97, no. 11, pp. 1786–1817, Nov. 2009.
- [33] J. Zhang, J. Liu, S. Zhong, J. Yang, N. Zhao, and T. Q. Zheng, “A Power Electronic Traction Transformer Configuration With Low-Voltage IGBTs for Onboard Traction Application,” *IEEE Trans. Power Electron.*, vol. 34, no. 9, pp. 8453–8467, Sep. 2019.
- [34] P. Simiyu, I. E. Davidson, “MVDC Railway Traction Power Systems; State-of-the Art, Opportunities, and Challenges,” *Energies* 2021, vol. 14, p. 4156, 2021.
- [35] S. Singh, “IGCT Transient Analysis and Clamp Circuit Design for VSC valves,” Master Thesis, Royal Inst. Technol. (KTH), Stock., 2012.
- [36] Y. Shakweh, “New Breed of Medium Voltage Converters,” *Power Eng. J.*, vol. 14, no. 1, pp. 12–20, Feb. 2000.
- [37] S. Bernet, “Recent Developments of High Power Converters for Industry and Traction Applications,” *IEEE Trans. Power Electron.*, vol. 15, no. 6, pp.

1102–1117, Nov. 2000.

- [38] “https://www.fujielectric-europe.com/en/power_semiconductor.”
- [39] “<https://www.hitachiabb-powergrids.com/offering/product-and-system/semiconductors>.”
- [40] “<https://www.infineon.com/cms/en/product/power/igbt/igbt-modules/#!products>,” 2021.
- [41] “<https://www.mitsubishielectric.com/semiconductors/products/powermod/index.html>.”
- [42] “<https://www.wolfspeed.com/products/power/sic-power-modules>.”
- [43] B. Wu, *High-Power Converters and AC Drives*, Piscataway, NJ, IEEE Press, 2006.
- [44] Integrated gate-commutated thyristors, ABB Switzerland Ltd, IGCT flyer, Doc. No. 5SYA2023, April 2008. <http://www.abb.com/semiconductors>.
- [45] R. D. Klug, N. Klaassen, “High Power Medium Voltage Drives - Innovations, Portfolio, Trends,” 2005 European Conference on Power Electronics and Applications, pp. 10 pp, 2005.
- [46] R. Alvarez, F. Filsecker, and S. Bernet, “Characterization of a new 4.5 kV press pack SPT+ IGBT for medium voltage converters,” in 2009 IEEE Energy Conversion Congress and Exposition, pp. 3954–3962, Sep. 2009.
- [47] J. Fabre, P. Ladoux, and M. Piton, “Characterization and Implementation of Dual-SiC MOSFET Modules for Future Use in Traction Converters,” *IEEE Trans. Power Electron.*, vol. 30, no. 8, pp. 4079–4090, Aug. 2015.
- [48] J. Fabre and P. Ladoux, “Parallel Connection of 1200-V/100-A SiC-MOSFET Half-Bridge Modules,” *IEEE Trans. Ind. Appl.*, vol. 52, no. 2, pp. 1669–1676, 2016.
- [49] P. Hofstetter and M. M. Bakran, “The Two-Dimensional Short-Circuit

Detection Protection for SiC MOSFETs in Urban Rail Transit Application,”
IEEE Trans. Power Electron., vol. 35, no. 6, pp. 5692–5701, Jun. 2020.

- [50] T. W. McNairy, “High-Speed Circuit Breakers,” 1928.
- [51] D. L. S. Marson, “High Speed Rail Electric-Power Supply Network,” Thesis, Czech Tech. Univ. Prague, 2016.
- [52] “http://www.ejrcf.or.jp/jrtr/jrtr20/pdf/F52_Tech.pdf.”
- [53] https://www.i-sis.org.uk/Supercapacitors_for_Flexible_Energy_Storage.php
- [54] A. Jongeling, “Optimising the Input Filter of Traction Installations in DC Railway Power Systems,” Master Thesis, Delft Univ. Technol., pp. 1–146, 2017.
- [55] Y. Zhu, “Adhesion in the Wheel – Rail Contact”, Thesis, Royal Institute of Technology, 2013.
- [56] M. Kondo, M. Miyabe, R. Ebizuka, and K. Hanaoka, “Design and Efficiency Evaluation of a High-Efficiency Induction Motor for Railway Traction,” Electr. Eng. Japan (English Transl. Denki Gakkai Ronbunshi), vol. 194, no. 2, pp. 15–23, 2016.
- [57] Z. Tian, “System Energy Optimisation Strategies for DC Railway Traction Power Networks,” University of Birmingham.
- [58] A. F. Abouzeid et al., “Control Strategies for Induction Motors in Railway Traction Applications,” Energies, vol. 13, no. 3. p. 700, Feb. 06, 2020.
- [59] S. Mohammad, M. Gazafrudi, A. T. Langerudy, E. F. Fuchs, and K. Al-haddad, “Power Quality Issues in Railway Electrification : A Comprehensive Perspective,” IEEE Trans. Ind. Electron., vol. 62, no. 5, pp. 3081–3090, 2015.
- [60] A. Choudhury, “Three-Level Neutral Point Clamped (NPC) Traction Inverter Drive for Electric Vehicles, Electrical and Computer Engineering Presented in Partial Fulfillment of the Requirements for the Degree of Doctor

of Philosophy Electrical and Comp,” no. April, 2015.

- [61] J. M. Knezevic, “Low-Cost Low-Resolution Sensorless Positioning of DC Motor Drives for Vehicle Auxiliary Applications,” *IEEE Trans. Veh. Technol.*, vol. 62, no. 9, pp. 4328–4335, Nov. 2013.
- [62] J. Brenneisen, E. Futterlieb, E. Muller, and M. Schulz, “A New Converter Drive System for a Diesel Electric Locomotive with Asynchronous Traction Motors,” *IEEE Trans. Ind. Appl.*, vol. IA-9, no. 4, pp. 482–491, Jul. 1973.
- [63] I. Press and R. Perez, *Electrical Railway Transportation Systems*, Hoboken, NJ, USA: John Wiley & Sons, Inc., 2018.
- [64] R. Krishnan, “Permanent Magnet Synchronous and Brushless DC Motor Drives”, London New York, CRC Press, 2010.
- [65] G. Pellegrino, A. Vagati, B. Boazzo, and P. Guglielmi, “Comparison of induction and PM synchronous motor drives for EV application including design examples,” *IEEE Trans. Ind. Appl.*, vol. 48, no. 6, pp. 2322–2332, 2012.
- [66] D. A. Rendusara, “Analysis of Some Application Aspect in Modern (PWM) Adjustable Speed Drive Systems and New Solutions,” Texas A&M University, 2000.
- [67] P. M. Page, *HIGH-POWER CONVERTERS and AC Drives*. 2007.
- [68] A. L. Julian, G. Oriti, and T. A. Lipo, “Elimination of common-mode voltage in three-phase sinusoidal power converters,” *IEEE Trans. Power Electron.*, vol. 14, no. 5, pp. 982–989, Sep. 1999.
- [69] M. M. Swamy and S. L. Rossiter, “Harmonic Interaction Between 1500 kVA Supply Transformer and VFD Load at an Industrial Plant,” *IEEE Trans. Ind. Appl.*, vol. 34, no. 5, pp. 897–903, 1998.
- [70] A. Salem and M. A. Abido, “T-Type Multilevel Converter Topologies: A

Comprehensive Review,” Arab. J. Sci. Eng., vol. 44, no. 3, pp. 1713–1735, 2019.

- [71] K. Yamanaka, A. M. Hava, H. Kirino, Y. Tanaka, N. Koga, and T. Kume, “A Novel Neutral Point Potential Stabilization Technique Using the Information of Output Current Polarities and Voltage Vector,” IEEE Trans. Ind. Appl., vol. 38, no. 6, pp. 1572–1580, Nov. 2002.
- [72] M. Koyama, T. Fujii, R. Uchida, and T. Kawabata, “Space Voltage Vector-Based New PWM Method for Large Capacity Three-Level GTO Inverter,” in Proceedings of the 1992 International Conference on Industrial Electronics, Control, Instrumentation, and Automation, pp. 271–276, 1992.
- [73] E. Ün, “Common Mode Voltage and Current Reduction in Voltage Source Inverter Driven Three Phase AC Motors,” Middle East Technical University, 2007.
- [74] A. M. Hava, R. J. Kerkman, and T. A. Lipo, “Carrier-based PWM-VSI overmodulation strategies: analysis, comparison, and design,” IEEE Trans. Power Electron., vol. 13, no. 4, pp. 674–689, Jul. 1998.
- [75] Lipo Thomas A.; D. Grahame Holmes, Pulse Width Modulation For Power Converters. 2003.
- [76] “A.M. Hava, ‘Carrier Based PWM-VSI Drives in the Overmodulation Region,’ Ph.D. Thesis, University of Wisconsin-Madison, 1998.”
- [77] Y. He, “A Comparison of Modulation Techniques and Motor Performance Evaluation,” Master Thesis, Chalmers Univ. Technol., 2018.
- [78] G. S. Buja and G. B. Indri, “Optimal Pulsewidth Modulation for Feeding AC Motors,” IEEE Trans. Ind. Appl., vol. IA-13, no. 1, pp. 38–44, Jan. 1977.
- [79] G. Narayanan and V. T. Ranganathan, “Two Novel Synchronized Bus-Clamping PWM Strategies Based on Space Vector Approach for High Power Drives,” IEEE Trans. Power Electron., vol. 17, no. 1, pp. 84–93, Jan. 2002.

- [80] J. Napoles, J. I. Leon, R. Portillo, L. G. Franquelo, and M. A. Aguirre, "Selective Harmonic Mitigation Technique for High-Power Converters," *IEEE Trans. Ind. Electron.*, vol. 57, no. 7, pp. 2315–2323, Jul. 2010.
- [81] J. Lago and M. L. Heldwein, "Generalized Synchronous Optimal Pulse Width Modulation for Multilevel Inverters," *IEEE Trans. Power Electron.*, vol. 32, no. 8, pp. 6297–6307, Aug. 2017.
- [82] F. Niu, B. Wang, A. S. Babel, K. Li, and E. G. Strangas, "Comparative Evaluation of Direct Torque Control Strategies for Permanent Magnet Synchronous Machines," *IEEE Trans. Power Electron.*, vol. 31, no. 2, pp. 1408–1424, Feb. 2016.
- [83] M. Rashed, P. F. A. MacConnell, A. F. Stronach, and P. Acarnley, "Sensorless Indirect-Rotor-Field-Orientation Speed Control of a Permanent-Magnet Synchronous Motor With Stator-Resistance Estimation," *IEEE Trans. Ind. Electron.*, vol. 54, no. 3, pp. 1664–1675, Jun. 2007.
- [84] I. Takahashi and T. Noguchi, "A New Quick-Response and High-Efficiency Control Strategy of an Induction Motor," *IEEE Trans. Ind. Appl.*, vol. IA-22, no. 5, pp. 820–827, Sep. 1986.
- [85] N. R. N. Idris and A. H. M. Yatim, "Direct Torque Control of Induction Machines With Constant Switching Frequency and Reduced Torque Ripple," *IEEE Trans. Ind. Electron.*, vol. 51, no. 4, pp. 758–767, Aug. 2004.
- [86] G. K. Singh, K. Nam, and S. K. Lim, "A Simple Indirect Field-Oriented Control Scheme for Multiphase Induction Machine," *IEEE Trans. Ind. Electron.*, vol. 52, no. 4, pp. 1177–1184, Aug. 2005.
- [87] R. Bojoi, M. Lazzari, F. Profumo, and A. Tenconi, "Digital Field-Oriented Control for Dual Three-Phase Induction Motor Drives," *IEEE Trans. Ind. Appl.*, vol. 39, no. 3, pp. 752–760, May 2003.
- [88] J. Rodriguez, R. M. Kennel, J. R. Espinoza, M. Trincado, C. A. Silva, and C.

- A. Rojas, “High-Performance Control Strategies for Electrical Drives: An Experimental Assessment,” *IEEE Trans. Ind. Electron.*, vol. 59, no. 2, pp. 812–820, Feb. 2012.
- [89] H. Kubota and K. Matsuse, “Speed Sensorless Field-Oriented Control of Induction Motor with Rotor Resistance Adaptation,” *IEEE Trans. Ind. Appl.*, vol. 30, no. 5, pp. 1219–1224, 1994.
- [90] I. Takahashi and Y. Ohmori, “High-Performance Direct Torque Control of an Induction Motor,” *IEEE Trans. Ind. Appl.*, vol. 25, no. 2, pp. 257–264, 1989.
- [91] H. Kubota, K. Matsuse, and T. Nakano, “DSP-Based Speed Adaptive Flux Observer of Induction Motor,” *IEEE Trans. Ind. Appl.*, vol. 29, no. 2, pp. 344–348, 1993.
- [92] D. Casadei, F. Profumo, G. Serra, and A. Tani, “FOC and DTC: Two Viable Schemes for Induction Motors Torque Control,” *IEEE Trans. Power Electron.*, vol. 17, no. 5, pp. 779–787, Sep. 2002.
- [93] D. Yildirim et al., “Full-Scale Physical Simulator of All SiC Traction Motor Drive With Onboard Supercapacitor ESS for Light-Rail Public Transportation,” *IEEE Trans. Ind. Electron.*, vol. 67, no. 8, pp. 6290–6301, Aug. 2020.
- [94] “AN6156 Calculating Power Losses in an IGBT Module Application Note.” https://www.dynexsemi.com/Portals/0/assets/downloads/DNX_AN6156.pdf.
- [95] L. Buhrkall, “Traction System Case Study,” in *IET Professional Development Course on Electric Traction Systems*, pp. 45–63, 2008.
- [96] M. Youssef, J. Abu-Qahouq, and M. Orabi, “The Electromagnetic Compatibility Design Considerations of the Input Filter of a 3-Phase Inverter in a Railway Traction System,” in *2010 IEEE Energy Conversion Congress and Exposition*, pp. 4210–4216, Sep. 2010.
- [97] “<https://www.secheron.com/products-solutions/ess/dc-circuit-breakers-2/for->

rail-vehicles/.”

- [98] Jin Wang, “Practical Design Considerations of Power Electronics in Hybrid and Fuel Cell Vehicles,” in 2008 IEEE Vehicle Power and Propulsion Conference, pp. 1–6, Sep. 2008.
- [99] “https://ep-us.mersen.com/sites/mercen_us/files/2020-03/Mersen-Solutions-for-SiC-Inverter.pdf,” 2021.
- [100] “[https://www.fischerelektronik.de/web_fischer/en_GB/heatsinks/D01/High-performance%20heatsinks/\\$catalogue/fischerData/PR/SK461_/search.xhtml](https://www.fischerelektronik.de/web_fischer/en_GB/heatsinks/D01/High-performance%20heatsinks/$catalogue/fischerData/PR/SK461_/search.xhtml), ” 2021.
- [101] IEC, “Rotating Electrical Machines - Part 25: AC Electrical Machines Used in Power Drive Systems - Application Guide,” no. 60034–25, p. 64, 2014.
- [102] B. S. I. S. Publication, “Rotating Electrical Machines Part 18–41: Partial Discharge Free Electrical Insulation Systems (Type I) Used in Rotating Electrical Machines Fed from Voltage Converters — Qualification and Quality Control Tests,” no. Type I, 2011.
- [103] “<https://www.infineon.com/cms/en/product/wide-band-gap-semiconductors-sic-gan/>,” 2021.

APPENDICES

A. Experimental Results with SHEPWM

The simulation results of SHEPWM are verified by experiments. Experimental work is carried out for two different dc link voltages. Output line-to-line voltage and line current waveforms, and associated harmonic spectra are given in Figure A. 1-3.

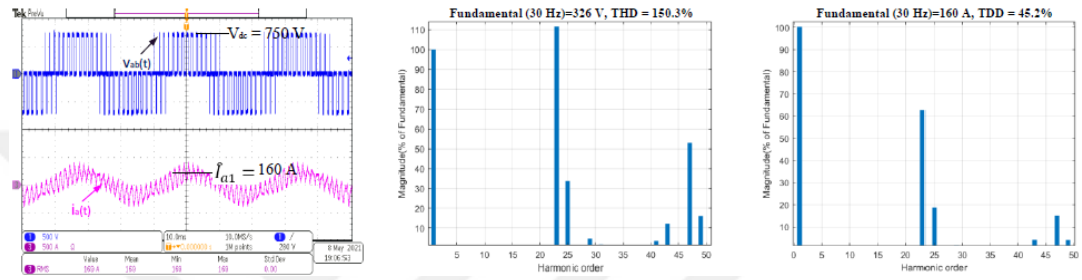


Figure A. 1. Experimental results for $V_{dc} = 750$ V, $f_s = 30$ Hz and $V_{l-to-l} = 230$ V while delivering 45 kVA to the stator

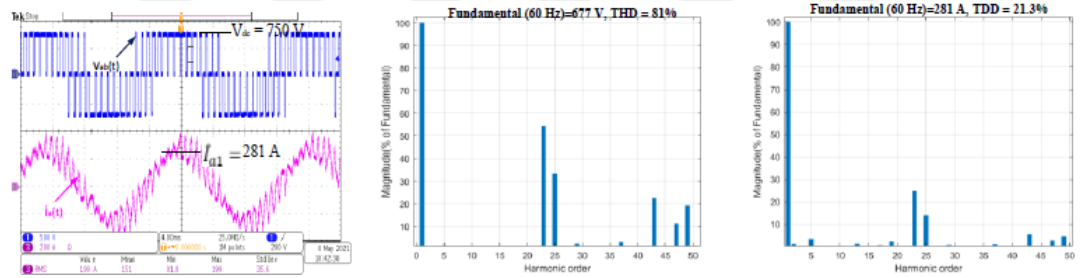


Figure A. 2. Experimental results for $V_{dc} = 750$ V, $f_s = 60$ Hz and $V_{l-to-l} = 478$ V while delivering 161 kVA to the stator

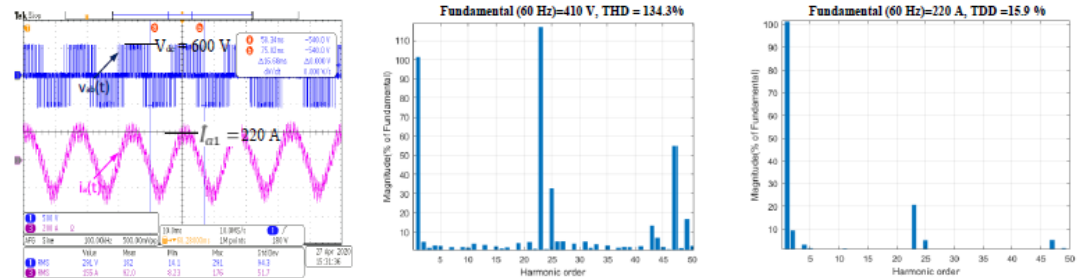


Figure A. 3. Experimental results for $V_{dc} = 600$ V, $f_s = 60$ Hz and $V_{l-to-l} = 293$ V while delivering 80 kVA to the stator