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DESIGN AND IMPLEMENTATION OF A
20KW 20KHZ UNITY POWER FACTOR
BOOST CHOPPER

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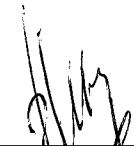
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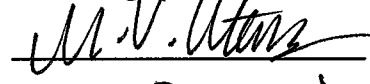
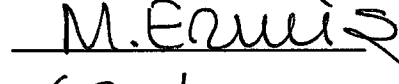
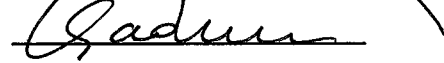
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ABSTRACT

DESIGN AND IMPLEMENTATION OF A 20KW 20KHZ UNITY POWER FACTOR BOOST CHOPPER

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A three-phase, medium-power standard ac motor can be operated from a single-phase, highly-distorted and unregulated ac source by the use of a diode bridge rectifier / boost chopper / inverter cascade. The integration of a boost chopper unit employing an inductor made of amorphous metal into the system results in unity pf at the ac supply side, low harmonic content line currents, high efficiencies and power densities. In this thesis, a 20kW 20kHz unity power factor boost chopper has been designed and implemented. Laboratory and field test results have shown that this system constitutes a reliable, cost-competitive and almost maintenance-free alternative to other drive options, especially for auxiliary systems in electric locomotives where squirrel-cage induction motors are chosen as auxiliary driving machines.

Keywords: Boost chopper, unity pf, ac motor drive

ÖZ

20KW 20KHZ GÜÇ FAKTÖRÜ BİR OLAN BİR YÜKSELTİCİ KİYİCİNİN TASARLANMASI VE GERÇEKLEŞTİRİLMESİ

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Bir 3 faz orta güçlü standart AA motoru bir diyot köprü doğrultucu / yükseltici kıyıcı / evirgeç sistemi kullanılarak regülesiz tek faz AA kaynaktan çalıştırılabilir. Amorf metalden yapılmış bir bobini olan yükseltici kıyıcının sisteme katılması giriş güç faktörünün bir olmasını, giriş akımının az harmoniğe sahip olmasını ve verimim yüksek olmasını sağlamıştır. Bu tezde, regülesiz tek faz AA kaynaktan beslenen 20kW 20kHz güç faktörü bir olan bir yükseltici kıyıcı tasarlanmış ve gerçekleştirilmiştir. Laboratuar ve alan testleri bu sistemin güvenilir olduğunu, bakım gerektirmediğini ve pahalı olmadığını göstermiştir. Bu bakımdan, bu sistem, özellikle yardımcı devre sürücü sistemlerinde sincap kafesli endüksiyon motorları bulunan elektrik lokomotiflerinde kullanımı açısından diğer alternatiflere göre daha üstündür.

Anahtar Sözcükler: Yükseltici kıyıcı, Güç faktörü düzenleyici, AA motor sürücü

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CHAPTER 1

INTRODUCTION

The unity power factor boost chopper has been used for power factor correction at the supply side and preregulation of voltage at the input of switch-mode power converters [1-14]. Steady-state performance of a dc motor supplied from a unity pf boost converter has also been discussed [15]. In the literature many unity power factor boost chopper topologies have been examined and their performance have been reported since 1980s [1-7]. The basic boost scheme shows a lot of variations depending on the number of switches, operating mode (continuous and discontinuous), the number of storage capacitors, the number and the location of the inductors and the presence of isolation [1, 8-14]. The most important feature of high frequency power factor correctors is their ability to shape line current so as to follow line voltage waveform. Several current control methods are available for active control of line current such as average current, peak current, hysteric and borderline control methods [1, 10-13]. Although unity power factor integrated circuits are available commercially to adopt any one of these control methods, the most popular one is average current control due to its smallest achievable line current total harmonic distortion.

Nowadays, active power factor correction has become an increasingly more demanded feature in off-line industrial equipments. Some of these topologies are commercially available at relatively low power applications up to a few kW especially in computer and telecommunication power supplies and, domestic appliances such as

fluorescent lighting, air conditioning etc. In such systems. Usually power MOSFETs are employed as switching elements with switching frequencies up to a few 100 kHz. The rapid progress in IGBT technology however extends the application area of boost power factor correctors thus allowing the implementation of such converters at higher output voltage and power ratings. New generation IGBTs (3rd generation IGBTs in late 1995 and 4th generation IGBTs in mid-1997) will increase the efficiency of boost converters because of the improvements in their switching performance and forward voltage drop. In this area, a few tens of kW power rating at an upper frequency limit of 20 kHz for hard switching is now achievable by the use of a single IGBT power module. The soft switching converter topologies can further increase the switching frequency at the expense of extra cost and complexity.

This text describes an IGBT based unity pf boost converter for three-phase ac motor drive applications from a single phase ac supply. Auxiliary ac motor drives in electric locomotives and agricultural applications in rural area, where a strongly fluctuating ac supply is available are considered as typical examples to the application of this system at medium power levels. In such applications, the boost converter topology offers a viable alternative to the solution of preregulation problem of strongly fluctuating supply voltage and allows the use of standard types of frequency controlled three-phase ac motor drives with scalar and/or vector control. The chopper can be also used as a preregulator of the voltage at the input of switch-mode power converters in computer and telecommunication area.

1.1 System Description and Problem Definition

E14000 type electrical units manufactured by GEC Alsthom have been kept in service since early 1980 by Turkish State Railways authorities (see photograph in Fig. 1). In the original electrical units, diode bridge rectifier-fed dc motors have been used

for more than 10 years to drive the fan load which cools down the traction motors and, the compressor load which provides pressurized air for pneumatic brakes and doors. Operation, maintenance and repair statistics in the long term have shown the feasibility of replacing the dc motor drives by modern squirrel-cage induction motor drives with preregulation of catenary voltage. A further increase in the usual maintenance and repair costs of DC motor drives has been observed owing to the strong fluctuations in catenary voltage on suburban lines (Catenary voltage waveforms are given in Figures 19-22.).



Fig. 1. Photograph of revised locomotive

In order to meet the needs of the Turkish State Railways to AC motor drives in E14000 Electrical Units ASELSAN signed a contract with TÜBİTAK- BİLTEN in 1995 to develop AC motor drives for fan and compressor loads. Since then, the author of this thesis has worked as the principal investigator in the development of 20kW, 20kHz unity pf boost chopper unit. The boost converter provides a regulated 530V DC voltage input to a commercial and standard 3 phase inverter unit from the highly distorted and unregulated single phase AC supply. The squirrel-cage induction motors are fed from two independent inverters mentioned above to drive fan and compressor loads.

1.2 Scope of the thesis

This thesis is devoted to the design and implementation of a 20kW, 20kHz unity power factor boost chopper. The designed system has been in operation successfully since September 1996.

In Chapter 2, power factor is defined and related standards are given. The disadvantages of low power factor is presented. Active and passive solutions are also discussed. The energy storage methods in chopper circuits are reviewed. Active power factor correction techniques are compared and a detailed explanation of the chosen method is also given.

Chapter 3 describes the design of the chopper. A sample application is given and the corresponding design work is presented. The PCBs developed within the scope of this research work are also described. Finally, the choice of the boost inductance and the storage capacitor is presented.

Chapter 4 is devoted to the implementation of the chopper circuit. Design and manufacturing of the inductor are also explained. The choice of the power semiconductor switch and its cooling requirement and cooling method are presented. Also, the chosen motors and inverters are briefly described.

Chapter 5 gives the experimental results obtained in the laboratory and discusses the results of the tests conducted on the auxiliary drive systems of E14000 type electrical units in the field.

CHAPTER 2

ACTIVE POWER FACTOR CORRECTION

Power factor correction and low cost line harmonics reduction are becoming more popular as the amount of line-fed electronic equipment increases and the related standards become more important or even vital to obey [1].

The aim of the active power factor correction is to make the input power factor unity. Input current is programmed in response to the input voltage by an active power factor corrector. If the ratio between the voltage and current is a constant the input will behave as a resistor and power factor will be unity. When the ratio deviates from a constant the input will contain phase displacement, harmonic distortion or both and this will make the power factor different than unity.

In this chapter, first of all, the definition of power factor is given and disadvantages of low power factor are presented. Then, passive and active solutions for power factor correction are discussed. The energy storage methods for the chopper circuits are explained. Active power factor correction techniques are compared and a detailed explanation of chosen technique is also given.

2.1 The Definition of Power Factor

The traditional definition of power factor is

$$PF = \frac{P}{V_{rms}(\sin) * I_{rms}(\sin)} \quad (1)$$

Where P is real power input, V_{rms} and I_{rms} are the root mean square of voltage and current. However, this equation is applicable only to pure sinusoidal waveforms. For the most general case, power factor is divided into two and the one is mentioned above known as displacement factor. Phase displacement is a measure of the reactance of the input impedance.

$$PF = PF_{disp} * PF_{dist} \quad (2)$$

Where PF_{disp} and PF_{dist} are displacement factor and distortion factor respectively.

The distortion factor is a measure of non-linearity of the input impedance. It is given by Eqn. 3.

$$PF_{dist} = \frac{\frac{I_{1,1}}{\sqrt{2}}}{\sqrt{\sum_{n=1}^{\infty} \left[\frac{I_{1,n}}{\sqrt{2}} \right]^2}} \quad (3)$$

Power factor is not a sensitive measure of the distortion or displacement of the current waveform. For example, 30% total harmonic distortion will lead to a drop in power factor from 1.0 to 0.95. However, 30 degrees phase displacement between current and voltage corresponds to much lower power factor (0.86).

2.2 Related Standards

World standard organizations specifies maximum limits for the amount of current allowed at each of the harmonics of the line frequency. In the near future, these harmonic limiting standards will be compulsory at low power in Europe. Therefore active power factor correction will become very popular. Harmonic limiting standards are as follows.

-IEC 555: Disturbances in Supply Systems Caused by Household Appliances and Similar Equipment; Part 2: Harmonics (for less than 16A/phase) and Part 4: Harmonics (more than 16A/phase, in development)

-EN 60555-2: European Norms derived from IEC 555

-VDE 0838: Also derived from IEC 555

-IEEE Std 519-1992: IEEE Recommended Practices and Requirements for Harmonic Control in Electrical Power Systems

-MIL-STD-1339

-TS 9882/Şubat 1992: TSE derived from IEC 555

2.3 Disadvantages of Low Power Factor

The low power factor results in several unwanted effects. The disadvantages of low power factor is given as below.

- The maximum obtainable rectified power from a wall outlet is reduced.
- The harmonic currents cause excess wiring and power transformer losses in the utility network.
- The voltage waveform is flattened because of high peak currents.
- The harmonics may cause resonance in power system.
- The equipment fails to comply with agency regulations.

2.4 Passive Solutions to AC/DC Conversion

Line commutated rectifiers are used in ac-to-dc conversion applications because of their ruggedness and simplicity [6]. They require few power circuit elements and control signals are easily generated by simple circuitry (when uncontrolled rectifier is used no control signal is required). However, it has some disadvantages.

- The output voltage is not pure dc but contains a substantial ripple component.
- In the controlled rectifier case, this ripple component increases with delay angle.

The dc filter should be used to reduce the effects of this ripple component on the load (Fig. 2). However, the combination of rectifier and filter causes quite non-sinusoidal input current (Fig. 3). Harmonic content of the current may vary as the phase angle is increased. Furthermore as the delay angle is increased the fundamental

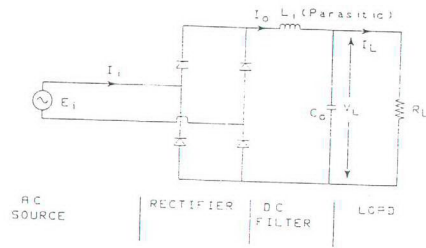


Fig. 2. Conventional single phase diode rectifier

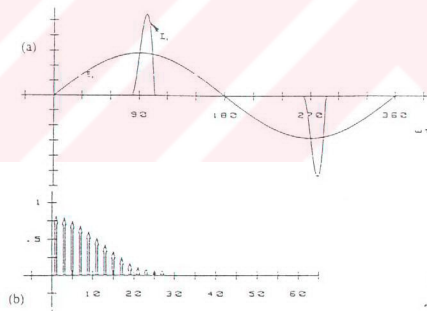


Fig. 3. Input current and its spectrum for the conventional rectifier [2].

component of the input current becomes increasingly phase shifted with respect to sinusoidal input voltage.

The non-ideal characteristics of the input current drawn by standard rectifier causes undesirable results.

- Phase displacement of the current and voltage fundamentals requires that source and distribution equipment handle reactive power increasing their VA rating.

- Harmonic current causes an increase in heating and sometimes peak current levels.

- The impedance of the distribution network will cause these harmonics to distort the supply voltage, causing increased losses and interference in other loads elsewhere on the power lines.

Using LC filter (Fig. 4) may be a solution. It improves power factor and decreases harmonic distortion (Fig. 5).

Some other passive solutions can be found but, in general, the advantages and disadvantages of passive solutions are listed in the next two sections.

2.4.1 Advantages

- easy to understand, implement and service

- cost-effective

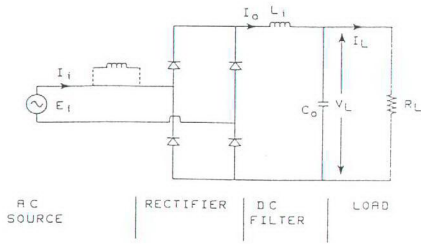


Fig. 4. LC Filter

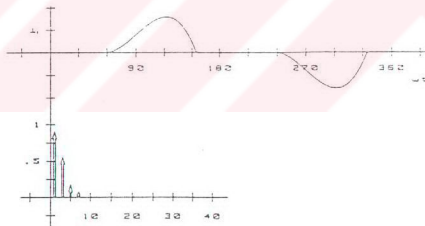


Fig. 5. Input current and its spectrum for LC filter [2].

- reliable and rugged
- no high frequency EMI problem
- insensitive to line surges and spikes

2.4.2 Disadvantages

- heavy and bulky
 - poor static and dynamic regulation
 - excess phase shift
 - significant line current distortion
 - sensitive to line frequency (some circuits)
- a narrow range of operating points for which the power factor can be optimized

2.5 Active Solutions to AC/DC Conversion

To overcome the problems of passive solutions, a switch (may be more than one in some topologies) is used. This switch is operated at frequencies above the audio range to prevent acoustic noise. Active power factor correction has many advantages over passive solutions.

- light weight
- tight regulation
- low-line current distortion
- wide range of operation
- straightforward design
- lower current ratings for power semiconductor device
- higher efficiency because of low rms value of input current

However, the penalty paid for all these improvements is the cost of control, EMI filter and fast power semiconductors.

2.6 Inductive and Capacitive Energy Storage

Line current, voltage and power waveforms are given in Fig. 6 for a resistive load.

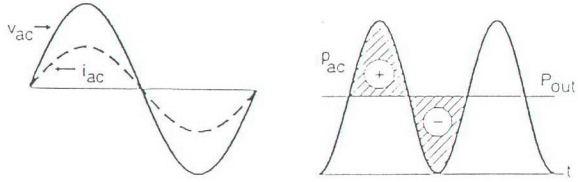


Fig. 6. Voltage, current and power waveforms for a resistive load

The waveforms are similar to waveforms of an active power factor correction with averaging over a few switching cycle. The instantaneous power drawn from line is above the mean power during half of the line period and below the mean power during the other half of the line period. If the load is assumed to draw constant power from the supply the excess energy should be stored and retrieved. The stored energy should be small for an economical solution. The minimum stored energy can be found as follows.

$$\begin{aligned}
 W_{dc} &= P_{out} \frac{T}{4} & W_{ac} &= \int_0^{\frac{T}{4}} P_{out} (1 - \sin 2\omega t) dt = P_{out} \left(\frac{T}{4} - \frac{T}{2p} \right) \\
 W_{min} &= W_{dc} - W_{ac} = P_{out} \frac{T}{2p} = \frac{P_{out}}{\omega}
 \end{aligned} \tag{4}$$

If this energy is stored in an inductor, the minimum inductance will be

$$L_{\min} = \frac{2P_{\text{out}}}{\omega I_{\text{pk}}^2} \quad I_{\text{pk}}: \text{peak inductor current} \quad (5)$$

If a capacitor is used, the minimum capacitance is

$$C_{\min} = \frac{2P_{\text{out}}}{\omega V_{\text{pk}}^2} \quad V_{\text{pk}}: \text{peak capacitor voltage} \quad (6)$$

However, the use of the values causes the inductor current or capacitor voltage to vary from zero to peak value which is not desired by the load. Therefore a larger storage device is needed.

2.6.1 Inductive Energy Storage

Buck and Buck-Boost topologies can be used with inductive energy storage by using a large inductor to carry constant current. This scheme is used very rarely because a large inductor is required, load transient response is slow and a floating switch is needed.

2.6.2 Capacitive Energy Storage

This type of energy storage is very common in active power factor correction. It can be used with any type of converters but, here only three topologies will be examined.

2.6.2.1 Boost Topology

Boost topology has many advantages over other topologies therefore it is used very frequently [3]. The boost circuit is given in Fig. 7.

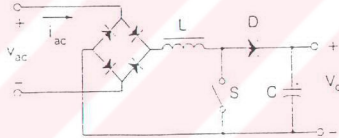


Fig. 7. Boost power factor corrector

It provides tightest control over line current waveform, offers best utilization of power switch, applies smallest HF current amplitude to the input capacitor, uses a ground-referenced switch and has the smallest output capacitor volume. However,

the output voltage must be higher than peak input current so the voltage rating of the switch should be V_{out} . There is no series switch able to prevent from inrush current, overload and short-circuit. The isolation between input and output is difficult to implement.

2.6.2.2 Buck Topology

In Fig. 8 and Fig. 9, the buck power factor corrector circuit and its switch current are shown respectively.

Since there is a series switch overload, short-circuit and inrush current protection is possible. In addition there is no voltage boosting and the voltage rating of the switch is V_{in} . However, it has some drawbacks.

- When the instantaneous line voltage is below the output voltage line current is zero causing significant line current distortion.

- EMI is high.

- Switch rms current is high.

- Large filter is required.

2.6.2.3 Buck-Boost Topology

Buck-Boost power factor corrector and its typical switch current is shown in Fig. 10 and Fig. 11 respectively.

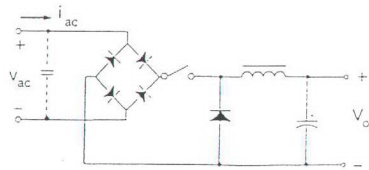


Fig. 8. Buck power factor corrector

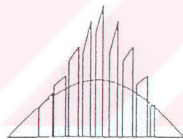


Fig. 9. Switch current in Buck Circuit [1]

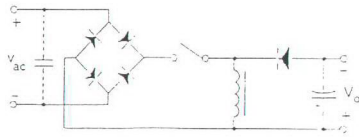


Fig. 10. Buck-Boost power factor corrector

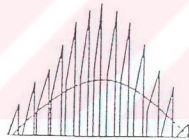


Fig. 11. Switch current in Buck-Boost Circuit [1]

It can provide overload, short-circuit and inrush current protection due to series switch. Output isolation and adjustable output voltage is obtained while achieving low line current distortion. The disadvantages of this topology are listed below.

- floating switch
- high rms current
- high EMI
- inverted output voltage
- increased voltage stress of the switch and freewheeling diode

2.7 Control Of Boost Topology

The most popular method used in active power factor correction is boost topology operating in continuous conduction mode for high power applications or discontinuous conduction mode for low power applications. Other topologies are derivative of this. In Fig. 12 four types of current control method are shown.

Average current, peak current and hysteretic control are examples of continuous conduction mode operation. Borderline control is an example of discontinuous conduction mode. In continuous conduction mode a smaller inductance is required than that in discontinuous conduction mode. Also the eddy current and

skin effect is reduced by continuous conduction mode. However, efficiency may be increased in some cases by discontinuous conduction mode.

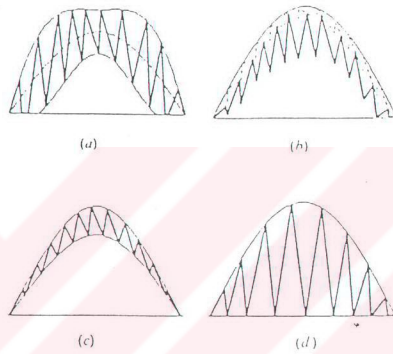


Fig. 12. Four types of current control method [3]

a) Average current control
c) Hysteretic control

b) Peak current control
d) Borderline control

A brief comparison between current control methods is given in Table 1[1-14].

Table 1. Comparison of current control methods

	average current	peak current	hysteretic	borderline
Frequency	constant	constant	variable	variable
THD	low	high	low	low
Variety of Control IC	wide	few	only one	few
Complexity of Control	high	low	high	low
EMI	low	low	low	high
Input Voltage Range	wide	narrow	wide	wide

Typical values and assumptions for Table-1:

THD Low: < 5%

Input Voltage Range Wide: > $V_{nom} \pm 20\%$

Complexity of Control: The complexity is said to be high if a current error amplifier is added to the feedback loop.

2.7.1 Average Current Control

Average current control is based on a simple concept (Fig. 13). An amplifier is used in the feedback loop so that input current tracks the programming signal with very little error. Input current and output voltage should be controlled. The current loop is programmed by the rectified line voltage so that input to the converter will appear to be resistive. The output voltage is controlled by changing the average amplitude of the current programming signal. An analog multiplier creates the current programming signal by multiplying the rectified line voltage with the output of the voltage error amplifier so that the current programming signal has the shape of the input voltage and an average amplitude which controls the output voltage.

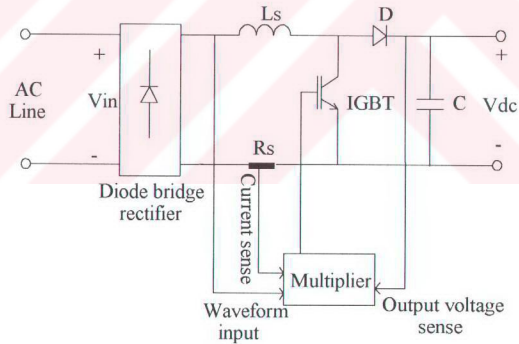


Fig. 13. Average current control system

Lowest line current distortion is achieved using this control method but inductor current must be sensed. Complexity of the circuit is increased due to the current amplifier in the feedback loop. There are many control ICs available in the market such as UC3854 (Unitrode), L4981 (SGS-Thomson), TDA4815 (Siemens) and ML4821 (Micro Linear).

2.7.2 Peak Current Control

This method is also suitable for boost topology. The peak inductor current is determined by the current reference signal. The dashed line in Fig. 12b is the reference signal with the stabilizing ramp necessary for constant frequency operation. Switch current is sensed so conductor loss is decreased. Complexity is reduced because a current amplifier is not used. Adding the necessary slope compensation for stability over wide range of duty cycles can be difficult to implement accurately so significant line current distortion is present. Examples of control ICs are ML4812 (Microlinear) and TK84812 (Toko).

2.7.3 Hysteretic Control

The inductor current is switched to follow a programmed peak and valley current level which track the shape of the instantaneous line voltage. Variable frequency operation of the switch is required to perform regulation of the output voltage, which may be undesirable in certain applications requiring synchronization to a system database. The switching frequency varies widely over the line period as

shown in Fig. 14. Only one control IC is available for this purpose (Cherry Semiconductor CS3810).

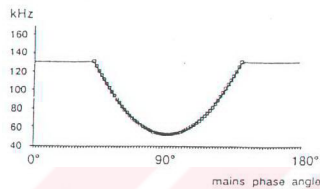


Fig. 14. Frequency variation for hysteretic control

2.7.4 Borderline Control

It is a version of hysteretic control with variable hysteresis. The switch is turned off when the inductor current reaches the reference level and is turned on when the inductor current reaches zero. Therefore converter operates at continuous and discontinuous conduction mode. Switching period varies widely over a line period and depends on line voltage and load current. TDA4814 (Siemens), UC 3852 (Unitrode) and L6560 (SGS-Thomson) are typical examples to available control ICs

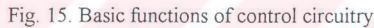
2.8 Average Current Control of Boost Chopper

There are many control ICs used to perform average current mode control of a boost chopper but, here, average current control technique suitable for UC3854(Unitrode) is presented.

The power circuit is given in Fig. 13. There is a single phase uncontrolled bridge to rectify AC input voltage. This rectified AC input voltage is used to program input current waveform. The large input capacitor normally placed at the output of the rectifier is moved to the output of the chopper.

Basic functions of control circuitry is shown in Fig. 15. There are two control parameters, input current and output voltage. The shape of current programming signal, I_{mo} , is just like the rectified line voltage to program input current. The average amplitude of the programming signal keeps the output voltage constant. This signal is created by an analog multiplier by multiplying rectified line voltage with the output of voltage error amplifier. However, multiplier input from voltage error amplifier is divided by the square of average of input voltage first. This feedforward signal results in constant gain of voltage loop. If this is not met, voltage loop gain changes with the square of average input voltage [7].

If the input voltage range is wide transient response must be fast therefore voltage loop bandwidth must be as large as possible. However, making voltage loop bandwidth very large results in modulation of input current to keep output voltage constant. In this case input current distortion would be high. Keeping voltage loop constant allows bandwidth be as close as input line frequency.



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2.8.1 UC3854 Block Diagram

The complete block diagram of the UC3854 is shown in Fig. 16. The voltage error amplifier (VEA), the squarer/divider circuit and multiplier has been explained before. I_{mo} , the output of the multiplier and actual current are the inputs of the current error amplifier (CEA). The output of the current amplifier connects the pulse width modulation comparator where it is compared to the oscillator ramp. The capacitor C_T and the resistor R_{SET} are used to set the oscillator frequency. The oscillator and the comparator drive the RS flip-flop and the flip-flop drives gate drive output (GTD). The gate drive output can supply 0.5A continuously and 1.5 A at 50% duty cycle from 15V.

The supply voltage range of the UC3854 is very large from nearly 20V to 35V. Enable input (ENA) allows the control of the starting time of the PFC. The inductor current is limited to a certain value by sensing the actual current at pin 2 of UC3854. The starting of the PFC is soft and the duty-cycle does not change quickly. This is done by a capacitor connected to soft-start (SS) pin of the UC3854. The maximum power dissipation of the integrated circuit is 1W. For further information, UC3854 data sheet is given in Appendix A.

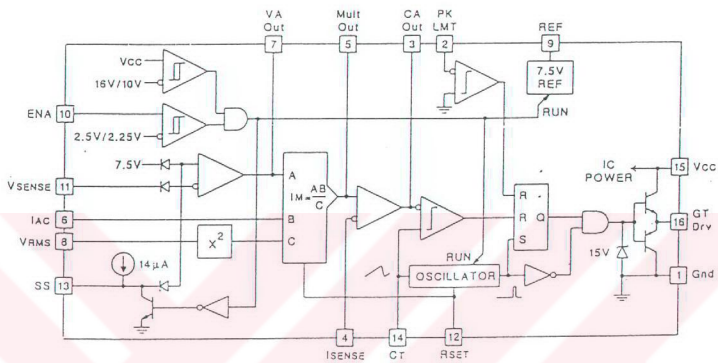


Fig. 16. Complete block diagram of the UC3854

CHAPTER 3

DESIGN

In this chapter, a sample application of the unity pf boost chopper is given and the need for this type of chopper is presented. Design of the system and the chopper is discussed as far as the given application is concerned.

3.1 An Application Area

The unity pf boost chopper can be used in many applications such as off-line equipments in computer and telecommunication areas at a few kW and preregulation of auxiliary ac motor drives in electric locomotives and agricultural applications in rural area at a few tens of kW.

The designed system has been implemented as a preregulator of auxiliary motor drives for fan and compressor loads in E14000 type electric locomotives which are in service in Turkish State Railways (TCDD). The fan is used for cooling of traction motors and the compressor is used for closing the doors and for the operation of brakes.

The old system given in the Fig. 17 is used to drive the fan and the compressor load. In the old system a diode bridge rectifier was used to feed dc motors from a 220V transformer. This system is replaced by a unity pf boost chopper-DC link inverter cascade. The new system has many advantages over the old system.

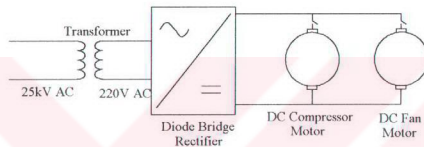


Fig. 17. The old system

The advantages of the new system:

1. Maintenance free operation
2. High performance
3. Low repair costs due to preregulation of catenary voltage
4. Flexible operation due to frequency control of the inverter

3.2 System Description

Fig. 18 shows the dc traction motor drive and auxiliary squirrel-cage induction motor drives supplied from 25 kV catenary line via the step-down train transformer. Fluctuations in the catenary voltage may cause an operating voltage range of $220\text{V} + 30\%$, $- 20\%$ across terminals of the auxiliary winding to which auxiliary drives are connected. In addition to these fluctuations, the supply voltage is highly distorted containing zero voltage periods owing to the overlap phenomenon caused by 1 MW line-commutated traction rectifier. Some sample records of the catenary voltage measured at the secondary of auxiliary winding are shown in Fig. 19, Fig. 20, Fig. 21 and Fig. 22. Traction drivers are idle in Fig. 19 and Fig. 20. Fig. 20 shows the persistence test of the catenary voltage which means about 100V variation at the peak.

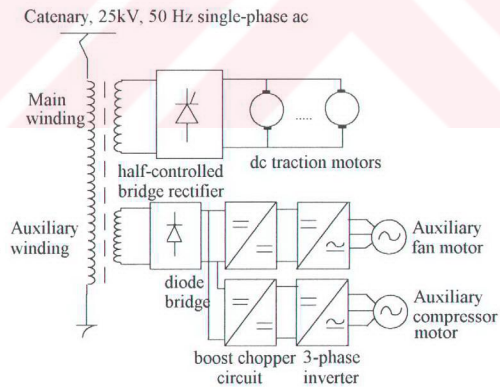


Fig. 18. Block diagram of auxiliary systems after revision

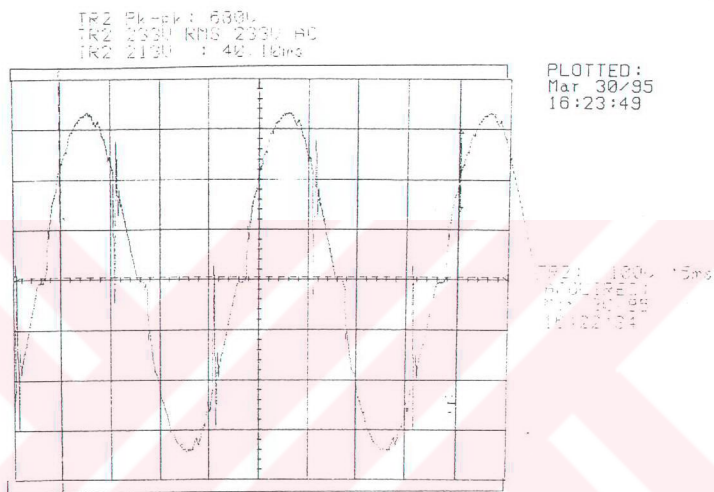


Fig. 19. Catenary voltage waveform when the traction drivers are idle

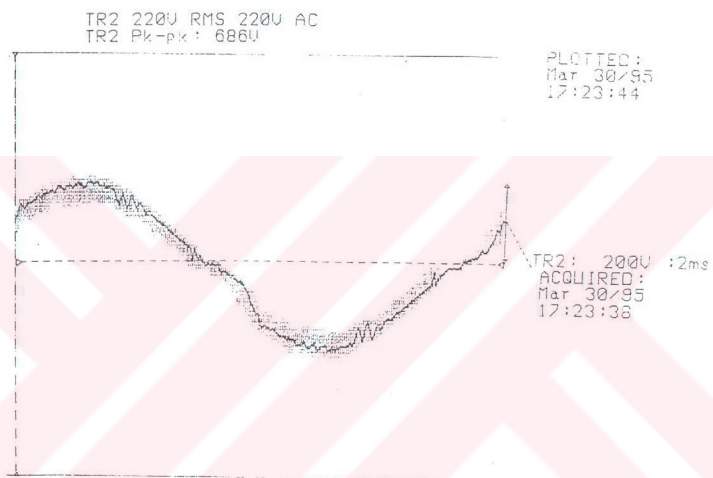
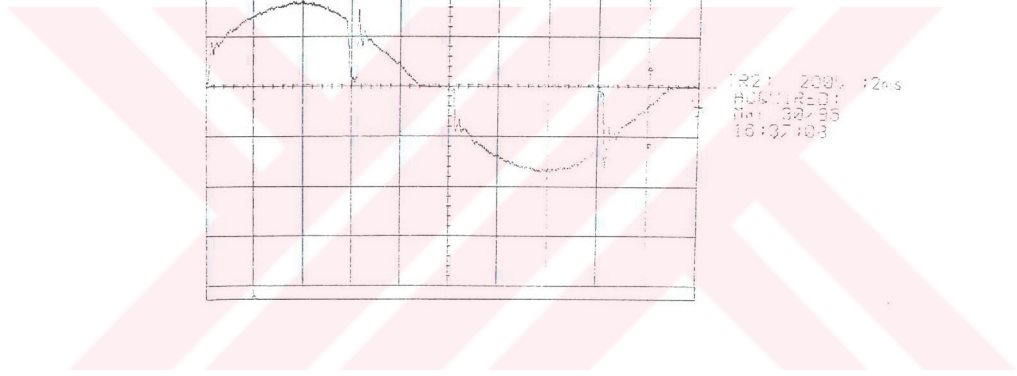


Fig. 20. Catenary voltage waveform in the persistence test



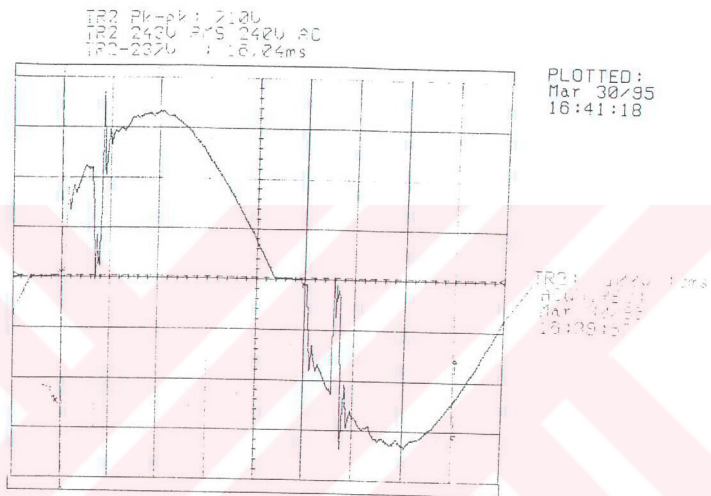


Fig. 22. Catenary voltage waveform during acceleration period (b)

Fig. 21 and Fig. 22 indicates the catenary voltage during the rapid acceleration of traction motors. Overlap angle may even reach 30° and the overlap of the half-controlled rectifier is demonstrated in Fig. 23 for firing angles 0° (Fig. 23b) and β (Fig. 23c).

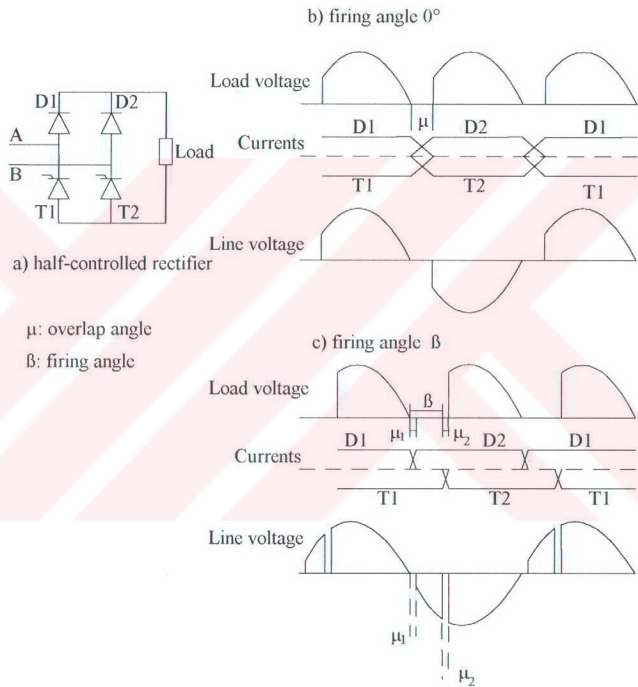


Fig. 23. Effect of overlap on line voltage of half-controlled rectifier

The compressor supplies pressurized air for a 1000 liters tank in the locomotive. The compressor characteristics is given in Appendix F. The compressor runs when the pressure of the tank below 7.5bars and stops above 9.2bars. If the pressure declines below 5bars the locomotive stops automatically. The compressor is driven by an induction motor at 900rpm. Hence, the air flow at 10bars is 2250 l/min. The power consumed by the compressor is above 15kW. However, the measurements have shown that the catalog value is exceeded by 10-15 % in practice.

Another induction motor drives the fan at 2700rpm. The fan cools down 1MW (total power) traction motors. If the fan does not run at a certain speed the locomotive can't start to move. There are two types of fan used at the locomotive. Some operating points for these fans are given in Appendix G. The air flow at this speed is 2-2.5m³/s depending on the type.

The boost type converter stabilizes the dc output voltage at 530V $\pm$ 2% against line voltage variations including both fluctuations and distortions. Standard three-phase inverter units are then connected to this constant voltage dc link to supply power to squirrel cage induction motors at the desired frequency. The dc link voltage of 530V corresponds to the necessary input voltage to the commercially available variable frequency inverters for generation of 380V, 50Hz standard 3-phase motor voltage in many countries. The resulting three-phase ac motor drive is highly flexible because soft starting technique by frequency control at a preset value of current limit can be applied in addition to adjustable speed control in the steady-state.

3.3 Boost Chopper Unit

Fig. 24 shows the boost chopper circuit with active pf correction. The controller consists of a power factor corrector and protection subcircuits and, can

communicate with the master control unit of overall system. Power factor corrector is a dedicated average current mode control integrated circuit allowing boost chopper operation in continuous conduction mode. In this operation mode, the input current follows the input voltage waveform thus maintaining very nearly unity pf conditions.

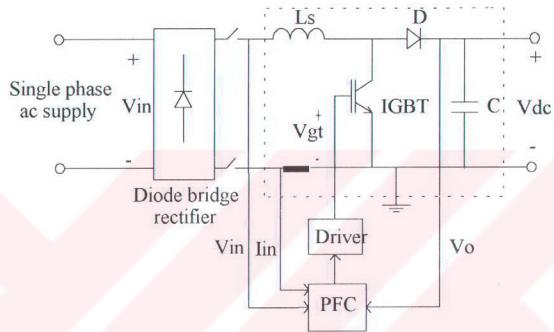


Fig. 24. Boost chopper configuration

The block diagram of the overall system with control logic is shown in Fig. 25. A general view of the implemented boost chopper unit is shown in Fig. 26. In this system, mainly four PCB cards are used and each one is described in the following subsections.

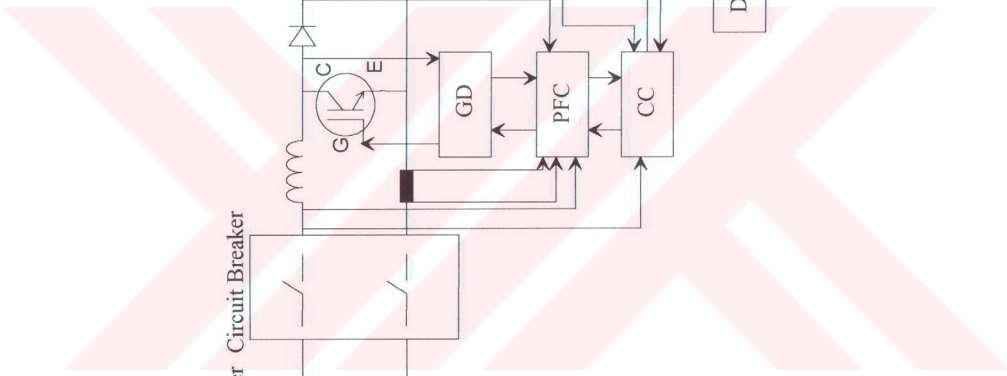


Fig. 25. Block diagram of the overall system with control logic

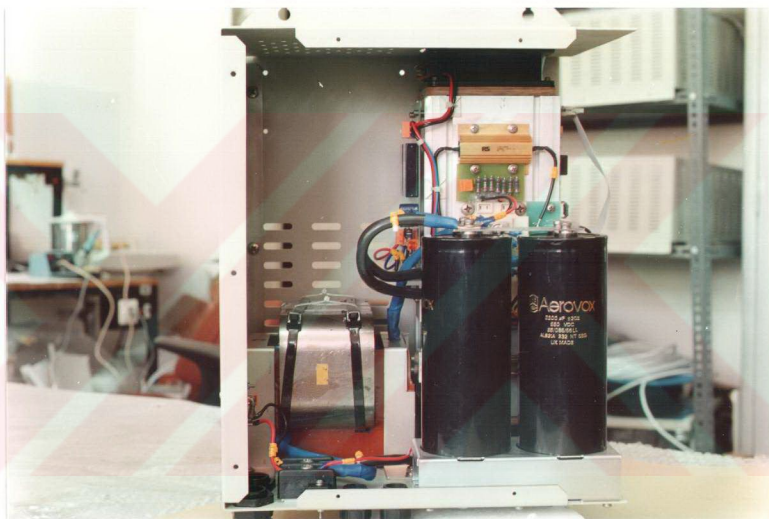


Fig. 26. Implemented Boost chopper unit

3.3.1 Power Factor Correction Card (PFC)

This card is used to keep the output voltage of boost chopper constant and to make the input power factor unity. The heart of the card is UC3854 and the other parts are passive elements (Fig. 15) which help UC3854 for control functions. Rvc and Rvd set DC output voltage. Cvf and Rvf is used in the compensation of the voltage error amplifier. Rff1, Rff2 and Rff3 set input voltage range. These resistors with Cff1 and Cff2 form a second order low pass filter to obtain average value of the input voltage. Rvac supplies required operating current for the multiplier and gives the information about input voltage waveform. All resistors and capacitors must be precise to get high performance.

This card is controlled by Chopper Control Card, which is explained later, to start and stop operation via optocouplers. In addition, this card gives information about short circuit error of the IGBT obtained from IGBT Gate Driver Card.

3.3.2 IGBT Gate Driver Card (GD)

A commercially available IGBT gate driver card SKHI10 is used from Semikron (Appendix E). Semikron SKHI10 IGBT Gate Driver Card can supply peak current up to 8A. It can function at both 5V and 15V logic level. It protects IGBT from short-circuit by VCE monitoring. It turns off IGBT softly in the case of short-circuit. If the supply voltage decreases below a certain value the card disables the gate drive output. Input-output isolation is realized by transformers. The gate resistance can be arranged externally.

3.3.3 Chopper Control Card (CC)

This card controls the operation of the chopper. It starts and stops the operation of the chopper via Power Factor Correction Card if the required signal comes from the Master Control Card and the input voltage of the chopper is in the prespecified range. It detects the misoperation of the chopper and resets the chopper a few times to recognize the misoperation whether it is temporary or not. If the misoperation is permanent the card stops the chopper. It also protects system from the overvoltage at the input and the output of the chopper. It sends signals to master control card carrying information about faults and status of the chopper. This card is electrically isolated from the system using isolated SMPS, relay and optocoupler to prevent the card from an high voltage in the case of a fault and noise attenuation if exists.

3.3.4 Master Control Card (MC)

This card controls the operation of the overall system. Since there are two different systems (fan and compressor) it functions as the master controller for both systems. It starts the choppers if the circuit breakers are closed and input voltages are in the prespecified range. It runs motors via inverters if the run signal comes from the locomotive control system. It shows the status of the inverters, the CBs and the choppers by using LEDs. It also indicates faults which occurred in the inverters and the choppers. This card is also electrically isolated from the system like the Chopper Control Card not only for the reasons indicated above but also for permitting the control of the fan and the compressor systems from a single card.

3.4 Protection

In the system, there are many protection circuits. These circuits are explained step by step.

1. Overvoltage at both sides of the chopper: The chopper control card senses input and output voltage and stops the chopper if the input or output voltage range is exceeded.
2. Overload at the chopper: The actual inductor current is sensed in the power factor correction card and if it increases above the specified value the duty cycle of the IGBT is decreased.
3. Short-circuit at the input of the chopper: There are two circuit breakers which are used at the input of both choppers to clear the short-circuits. In addition, there is a semiconductor fuse at the input of the diode bridge rectifier.
4. Snubber: To limit dv/dt of the IGBT below catalog value an RC snubber is used.
5. Varistors: There are two varistors used in the chopper for transient voltage suppression. The first is for the IGBT and the second is for the output capacitors of the chopper.
6. Overload at the motor side: The rated current value of the motor can be set by the inverter keypad as well as the permissible overload period. If the actual motor current exceeds beyond the preset value this protection facility becomes active to stop the operation the inverter.
7. Short-circuit at the motor: If the output current of the inverter rises very fast which occurs only for external short-circuits, the inverter disables the output.
8. Conformal coating: All cards in the system is coated by a conformal coating spray. This prevents the system from short-circuit possibly caused by dust and humidity thus providing an isolation level of 90kV/mm.

3.5 Chopper Specifications

The input and output specifications of the designed step-up chopper are given as below.

Output dc power, $P_{\text{out}} = 20 \text{ kW}$

Switching frequency, $f_{\text{sw}} = 20 \text{ kHz}$

Output voltage, $V_{\text{dc}} = 530 \text{ V} \pm 2\%$

Input voltage range, $V_{\text{in}} = 220 \text{ V} -20\%, +30\%$

Input frequency, $f_i = 50 \text{ Hz} \pm 1\%$

Maximum rms input current, $I_{\text{rms}} = 130 \text{ A}$

Input current ripple, $\Delta I = 5\%$ peak-to-peak at 20 kHz

Chopper efficiency, $\eta = 95\%$ at full load

Input power factor, $\text{pf} \approx 1$

Total harmonic current distortion at the input $\leq 5\%$ at full load

Hold-up time, $\Delta t = 5 \text{ ms}$

3.6 Design Considerations

The boost converter's power circuit consists mainly of an inductor, a semiconductor controlled switch, a diode and the output capacitors (Fig. 24). In this section, the circuit design and criteria for the selection of main circuit elements will be presented.

3.6.1 Choice of Boost Inductance Value

The designed boost topology operates at a fixed switching frequency in continuous inductor current mode and is regulated by means of average current control method. The inductance L in Fig. 24 determines the amount of high frequency ripple current at the input for continuous conduction mode of operation. For the inductor current i_L to be continuous, the ripple current ΔI should not exceed twice minimum input peak current value. The ripple current can be made arbitrarily small by increasing the value of boost inductance. ΔI value for this particular application is chosen to be about 5% of the maximum peak line current by making a compromise between core losses and core size. Small ripple current magnitude also lowers the injected high frequency noise into the supply lines.

The inductance value L is computed by using the peak of input sine wave current I_{peak} occurring at minimum input voltage and, the corresponding duty cycle and switching frequency as given in Eqn. 7.

$$L = \frac{V_{in(min)} D}{f_s \Delta I}$$

where, $D = \frac{V_o - V_{in(min)}}{V_o}$ (7)

and, $V_{in(min)} = \sqrt{2} V_{in(ac)min}$

The peak inductor current I_{peak} is the sum of high frequency ripple current and the power frequency line current and varies with input voltage as illustrated in Fig. 27. The corresponding inductor should then be designed to handle this current level without saturating. That is,

$$I_{peak} = \sqrt{2} I_{in(rms)max} + \frac{\Delta I}{2}$$

where $I_{in(rms)max} = \frac{P_{inmax}}{V_{in(rms)min}}$ (8)

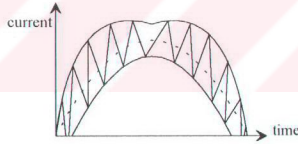


Fig. 27. Inductor current waveform

During on time of switching element, inductor current ramps upwards from its initial value. After the switch turns off, energy stored in L during the on-time of the switch is delivered to the output stage and inductor current ramps down. Fig. 27

illustrates the variations in i_L in continuous conduction mode for the average current control method.

Many inductor design procedures can be developed by changing optimization parameter e.g. core loss or volume. Two design procedures are given in Fig. 28 and Fig. 29. The first is a general design procedure which is especially applicable to high frequency inductors. In this procedure, the core size is tried to be minimized if the temperature consideration doesn't cause any problem. The other procedure is derived for only continuous mode boost preregulators. The parameter which should be optimized is power loss. The design is straightforward because the structure of the procedure is quite appropriate for the design of the boost chopper.

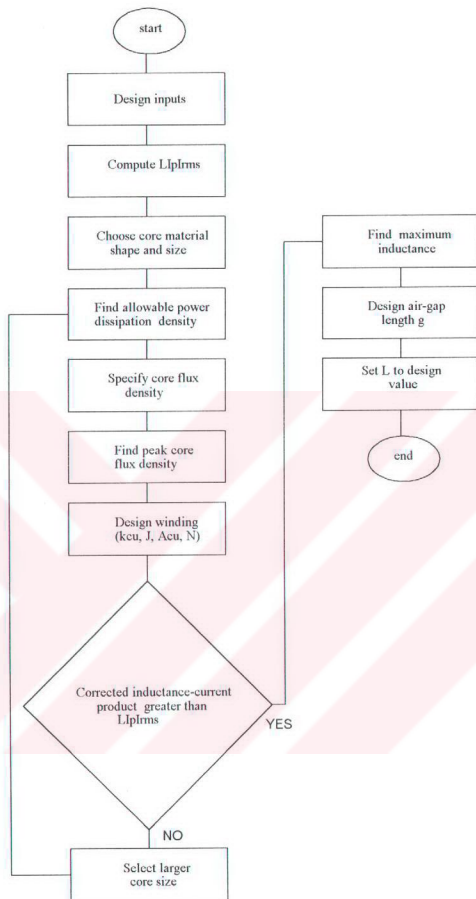


Fig. 28 Design procedure from [19]

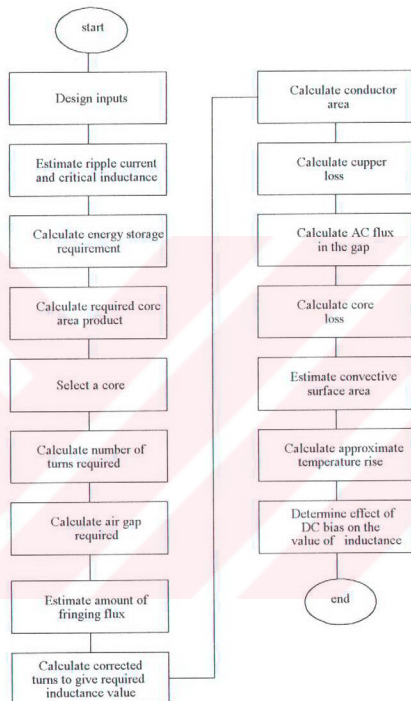


Fig. 29. Design procedure from Allied Signal Inc.

3.6.2 Choice of Output Capacitance

The factors involved in the selection of the output capacitor are the switching frequency ripple current, the second harmonic ripple current, the dc output voltage and, the output ripple voltage. The calculation of capacitance C of the capacitor is based on energy transfer from V_o to V_{min} . The output capacitor is to be chosen among the low ESR and low ESL type, electrolytic capacitors with high ripple current capability. The switching frequency ripple current ($\Delta I=10$ A p-p), second harmonic ripple current ($I_2=48$ A), the dc output voltage and the output ripple voltage are being the factors affecting this choice. However, the capacitance value of output capacitor should be determined from Eqn. 9 which expresses the energy transfer capability of the output capacitance during the hold-up time.

$$C_o = \frac{2P_{out} \Delta t}{V_o^2 - V_{omin}^2} \quad (9)$$

For this particular application, minimum acceptable value of hold-up time is the overlap period in the line voltage in the worst case (≈ 2 ms). A longer hold-up time chosen in the design can also tolerate supply interruptions with short durations at the expense of bulky and costly capacitor banks. A compromise is found between these factors and the hold-up time Δt is chosen to be 5 ms.

CHAPTER 4

IMPLEMENTATION OF SYSTEM

Design and implementation of various system components are explained in this chapter. Comparison of alternative core materials which may be used in the system are also presented in terms of size, power loss and cost. A photograph of the implemented drive consisting of a boost chopper / inverter cascade is as given in Fig. 30.

4.1 Design and Implementation of the Inductor

The design of the inductor is made by using the procedure given in Fig. 29 with small modifications. Various types of core materials can be used in this application. Advantages and disadvantages of these materials are listed in Table 2. The main objective in the design of power magnetics is to minimize the power loss and hence temperature rise of the inductor within the constraints of reasonable physical size and cost as well as material availability.



Fig. 30 Photograph of the implemented system

In order to select the most proper core material and a standard core size for use in the boost inductor, several commercially available cores have been examined on the bases of manufacturer's data and unit prices. These cover only amorphous metal, ferrite and powdered iron cores. The boost inductor ($L=0.8$ mH) is then designed by employing these core materials in turn. The magnetic circuit of the boost inductor could be formed by making use of more than one standard cores for ferrite and iron

powder materials (The dimensional informations of the cores are given in Appendix B). However, a single standard core of largest size is found quite satisfactory to meet the required area product for amorphous metal material. These cores and the corresponding designs are compared in Tables 3 and 4.

Table 2. Properties of various core materials

Material	Advantages	Disadvantages
Amorphous metal	High B_{sat} Low core loss at high B_{ac} and f	Moderate cost
Ferrite	Low core loss at high B_{ac}	Low B_{sat} , decrease of B_{sat} with T_{amb} , high cost, brittle
Powdered iron	Moderate B_{sat} Low cost	High core loss at high f and B_{ac}

It should be noted that this application needs H values much higher than those of amorphous metal and ferrite materials in the linear region of their B - H characteristics. Therefore an airgap must be inserted to their magnetic circuits to meet the desired magnetic operating conditions in the linear region. There is however no need to insert an external airgap to the iron-powder core owing to the distributed airgap properties inherent in them.

In this particular application, the boost inductor will carry a rectified ac current of 150A peak with a superimposed $\pm 2.5\%$ 20kHz component in the worst case. Design values of the peak flux density B_{peak} and peak magnetic field intensity H_{peak} in these core materials are given respectively in 3rd and 4th columns of Table 3.

Table 3. Comparison of standard cores in the market (a)

Core type	Saturation values		Design values			Coercive field	Air-gap
	Bsat (T)	H (A/m)	Bpeak (T)	Hpeak (A/m)	Bac (mT)	Hc (A/m)	required
Amorphous C-core Alloy SA-1	1.56	80	1.4	40	38	3.2	yes
Ferrite N27 UU-core	0.42	1200	0.3	60	8	20	yes
Iron powder E-core Mix. no 33	1.10	39800	0.8	20000	22	406	no

A qualitative loss comparison is made in 1st and 2nd columns of Table 4. Ferrite materials are basically oxide mixtures of iron and other magnetic materials. They have quite large electrical resistivity but rather low saturation flux densities. No significant eddy-current loss occurs because of the high electrical resistivity. The material has only hysteresis loss. Also in iron powder cores, there is no eddy current loss component because of the distributed airgap properties inherent in them. However, the hysteresis loss is significant in comparison with those of ferrite and amorphous metal cores especially for operating frequencies higher than 10kHz. Amorphous metal cores have low frequency-dependent losses. Although eddy-current loss component is present for them, it can be controlled and made small by reducing the ribbon thickness at the expense of higher cost. Core loss versus frequency curves of these materials are given in Fig. 31 on W/kg basis for an ac flux density (B_{ac}) of 0.1T. According to these curves, iron powder material seems to be drastically more dissipative. However, in this application, the total core loss figure for the cores build up of these materials depends on peak flux density of high frequency component as well as on material density, core volume and low frequency flux density component. That is, the relative core loss figures will be different than those of Fig. 31.

Table 4. Comparison of standard cores in the market (b)

Core type	Loss Comparison			Normalized volume	
	Eddy Current Loss	Hysteresis loss at 20kHz and 0.1T	Normalized core loss	core	occupied
Amorphous C-core Alloy SA-1	low	low	1.0	1.0	1.0
Ferrite N27 UU-core	negligible	low	0.1	4.2	6.1
Iron powder E-core Mix. no 33	negligible	high	9.0	2.9	4.7

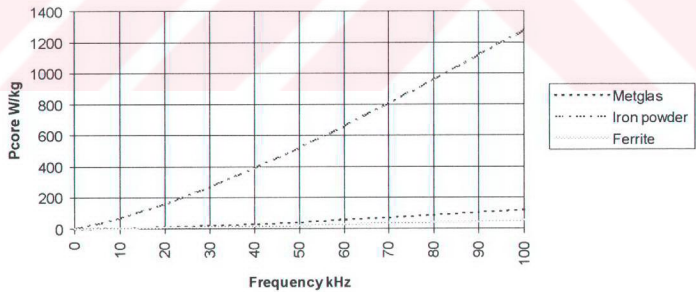


Fig. 31. Core loss comparison at $B_{ac}=0.1T$

For the boost inductor designed in this work, total core losses for the candidate materials are given in 3rd column of Table 4. Core loss component of amorphous metal core is nearly ten times bigger than that of ferrite and four times smaller than that of iron powder core in the final design. It is calculated to be 15 W. The core losses per unit surface area are estimated to be 44, 37 and 104 mW/cm² respectively for amorphous C-core, ferrite UU-core and iron powder E-core based designs. It can then be concluded that natural air-cooling is applicable to all cases.

A dimensional comparison of the three designs is as shown in Fig. 32. A volume comparison is also made and the normalized results are presented in the last two columns of Table 4. Here, core volume is calculated from core cross-sectional area and mean length while the volume occupied from core volume and window area. The choice of core material and core size is dictated primarily by price, volume occupied and power dissipation and, secondarily by circuit compactness and ease in manufacturing. A comparison is made in Table 5 on the basis of primary factors affecting the choice. In the last column, the total loss of each design includes both core loss and copper loss components. For each case, copper loss is estimated by assuming that the magnetic circuit is equipped with a standard wire coil. The use of a copper foil and Litz wire would yield better results at the expense of difficulty in bringing out coil ends to terminal box and higher cost, respectively.

In view of the above results, amorphous metal core (Appendix B) has been chosen for use in the boost chopper.

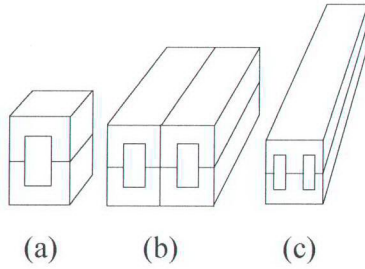


Fig. 32. Size comparison among different core types

- a. Amorphous C-core(1 core) b. Ferrite UU core(14 cores)
c. Iron powder E-core (10 cores)

Table 5. Normalized core cost and volume comparison

Core type	Unit price	Price/ system	Volume pu	Core loss pu	Total loss pu
Amorphous C-core	10.	1.1	1.0	1.0	4.0
Ferrite UU-core	7.0	9.8	4.2	0.1	6.5
Iron powder E-core	1.0	1.0	2.9	9.0	13.2

4.2 Power Semiconductors

The power switch consists of a shunt IGBT power transistor and series boost diode as a single integrated power module. The boost converter module is built up of high power, high frequency third generation IGBT transistors. In order to achieve a high conversion efficiency, the topology should have a low voltage stress across the IGBT in the on state. A 200 Amp, 1200V IGBT is used as the power semiconductor switch. The characteristics of the IGBT is given in Appendix D.

4.3 Thermal Considerations

In order to guarantee safe operation of the IGBT module with a forward diode, the forced-air cooling method is used for the transfer of heat generated in the module. The total power loss under worst case operating conditions considering both switching and conduction losses are estimated as 380W for the IGBT and, 170W for the diode. The junction temperature should not exceed 150⁰C for both the IGBT and the diode. With a safety margin of 25⁰C in the junction temperature and considering a maximum ambient temperature of 45⁰C a heatsink with a thermal resistance $R_{th(lt-a)}$ 0.080⁰C/W is required. This requirement is satisfied by a 300mm heatsink (Appendix C) with an airflow of 4.5m/s.

4.4 Layout

Special attention is paid to the layout of the power circuit because of high frequency and high rms currents carried. To avoid the generation of dangerous overvoltages, stray inductance in the loop formed by power switch, boost diode and output capacitor is minimized by making corresponding connections as short as possible. Furthermore, a specially designed U-shaped high frequency shunt is employed for current sensing. Power factor controller PCB, driver and snubber circuits are mounted on the same heatsink thus minimizing cable lengths. Twisted and twin pair cabling is used throughout the system to minimize cable inductances.

4.5 Induction Motors

Two three-phase squirrel cage induction motors are used for driving the fan and the compressor loads (Appendix F). The power requirement of the compressor is Specified as 15kW at 950rpm and at an air tank pressure of 10bars. However, the compressor operates at a duty cycle of approximately 80%. Therefore, a standard three-phase 15kW induction motor has been chosen to drive the compressor load. The photograph of the induction motor mounted horizontally at the bottom of electrical unit is shown in Fig. 33. The motor is therefore selected as outdoor type and has a protection class of IP54 against dust and water.

The fan (Appendix G) cooling down traction motors operates continuously when the locomotive is in service. The same motor ratings have been chosen with a vertically down mounted type. The protection class is same as the compressor motor.



Fig. 33 Photograph of the compressor motor with the compressor

4.6 Inverters

Since the inverter is connected to the output of the chopper, a standard 15kW DC-link converter which can be directly supplied from the DC link has been chosen to drive the induction motors. The fan and the compressor motors' operate independently from each other and therefore, an individual inverter unit should be used for each of them. Both inverters should be capable of operating at 20% overload continuous and at 50% overload for 60 seconds. They can soft-start and stop the motors with frequency control and can hold the current limit at 48A max. or at a predefined value smaller than 48A set from a keypad. In addition, they can adopt both scalar control and open-loop vector control methods. The inverter output voltage can be varied from 0-400V AC with variable frequency of 0-50Hz. The acceleration and deceleration ramp rates are also programmable from the keypad.

CHAPTER 5

EXPERIMENTAL RESULTS

This chapter presents experimental results of the system obtained both in the laboratory and the field. The performance of the implemented system is tested and the related figures and characteristic waveforms are also given.

5.1. Lab test results

The harmonic content of the open-circuit busbar voltage is measured to be $V_3/V_1=3.2\%$ and $V_7/V_1=0.8\%$ and, higher harmonics are negligible. An autotransformer connected to the above busbar has been used to change the input voltage of the boost chopper. The autotransformer introduces a further distortion to the ac input voltage, especially at the test conditions in which the input voltage to the chopper is lowered. The input ac voltage to the system is varied and, rectified input voltage and output voltage of boost chopper are recorded simultaneously. Sample waveforms corresponding to low input voltage ($V_{in}=160V$ AC) and high input voltage ($V_{in}=250V$ AC) are given in Fig.34 and Fig. 35 respectively. These results show that the implemented boost chopper satisfies the predetermined output voltage specifications against the variations in the input voltage.

Input voltage and current waveforms are also recorded at the primary side of the autotransformer for different operating conditions. A sample waveform set is given in Fig. 36 for an input voltage of 240V rms and at an output power of 19kW. It is observed that power factor is very nearly unity for heavy load conditions and best at rated load ($P_{out}=19kW$, displacement angle $\phi=9^\circ$ and $pf=0.99$ lag). Over the entire operating range, performance figures of boost converter have been measured and given in Table 6. The test results have shown that the boost converter operates at an efficiency higher than 95% from half-load to full-load.

Table 6. Efficiency of the chopper circuit

V_{in} , V	I_{in} , A	ϕ	pf	P_{out} , kW	η , pu
242	77	9°	0.99	18.0	0.97
242	70	11°	0.98	16.2	0.97
180	81	14°	0.97	13.6	0.96
200	72	14°	0.97	13.6	0.97
220	66	14°	0.97	13.6	0.97
220	59	14°	0.97	12.2	0.96
180	42	18°	0.95	6.8	0.95

Harmonic spectrum of the input alternating current is also obtained experimentally and the one corresponding to full load operating conditions at rated input voltage is given in Fig. 37. Table 7 shows harmonics values in the figure.

Table 7. Harmonics values in Fig. 37

Harmonics	Value in dB	Percent of fundamental
1	0	100
3	-27	4.4
5	-25	5.6
7	-30	3.2
9	-28	3.9
11	-37	1.4
13	-37	1.4
15	-47	0.4
17	-45	0.6
19	-45	0.6

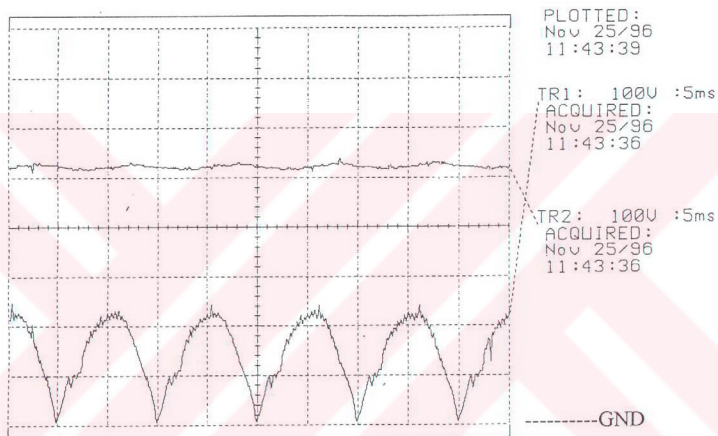
The total line current harmonic distortion and distortion power factor are calculated as,

$$THD = \frac{(I_{rms}^2 - I_{1(rms)}^2)^{1/2}}{I_{1(rms)}} \quad (10)$$

$$PF_{dist} = \frac{I_{1(rms)}}{I_{rms}} \quad (11)$$

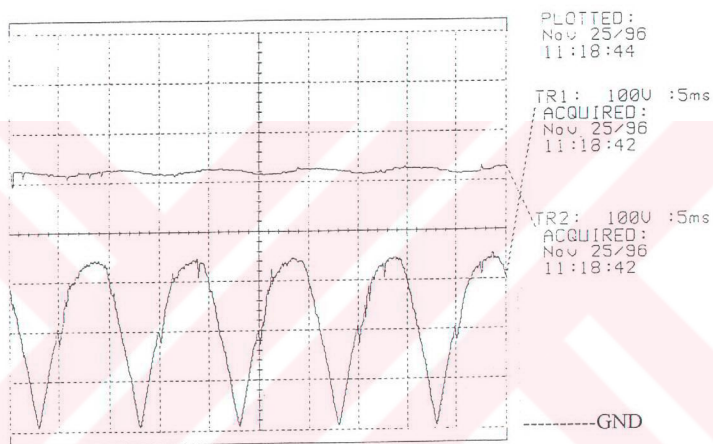
$$I_{rms} = (I_{1(rms)}^2 + I_{2(rms)}^2 + \dots + I_{n(rms)}^2)^{1/2} \quad (12)$$

From Eqn. 12 I_{rms} is found to be $1.004I_{1(rms)}$. Total harmonic distortion is found to be 9% and none of the harmonics exceeds 6% of the magnitude of fundamental component. The corresponding distortion power factor is negligibly small and is equal to 0.996. The reduction of line current harmonics is of great importance as dictated in [17, 18]. Since the boost converter is supplied from a distorted busbar voltage with a 3.3% total harmonic distortion, its own line current harmonic content would be better than the above figure.



TR1: Output voltage of diode bridge rectifier
TR2: Chopper output dc voltage

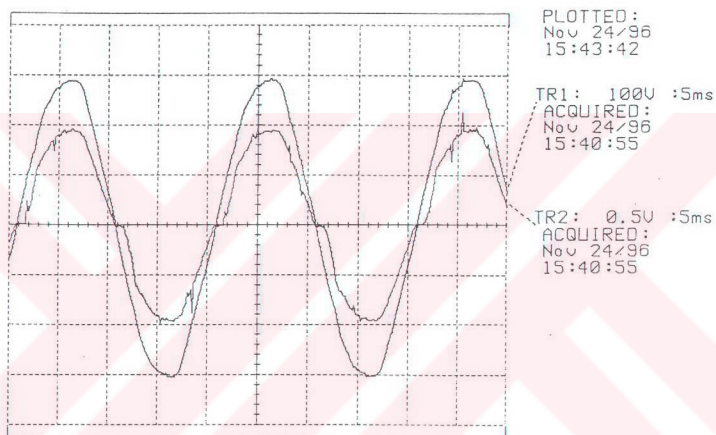
Fig. 34. Output voltage vs. input voltage waveforms at low input voltage



TR1: Output voltage of diode bridge rectifier

TR2: Chopper output dc voltage

Fig. 35. Output voltage vs. input voltage waveforms at high input voltage

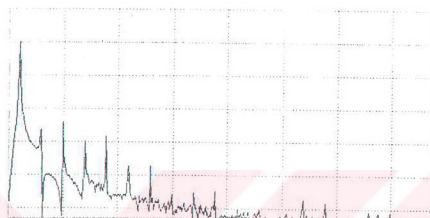


TR1: Input voltage at autotransformer primary

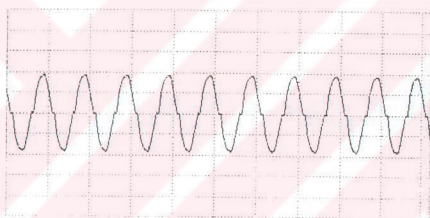
TR2: Line current at autotransformer primary

Fig. 36. Input current vs. input voltage waveforms

TRIG: CH1 CH2 MANU '96-12-24 17:03:05



a. LINEAR SPECTRUM
CH2
y.L06-MAG 10dB/div
(-50dB*10dB)
x.LIN-Hz
250 Hz/div
Peak:
y=-4.9dBV
x=48.82 Hz



b. STORAGE
CH2
y.LIN-REAL
500mV/div
x.TIME
20ms/div

Window
Averaging

RECTAN
OFF

Max. freq. 1.95kHz

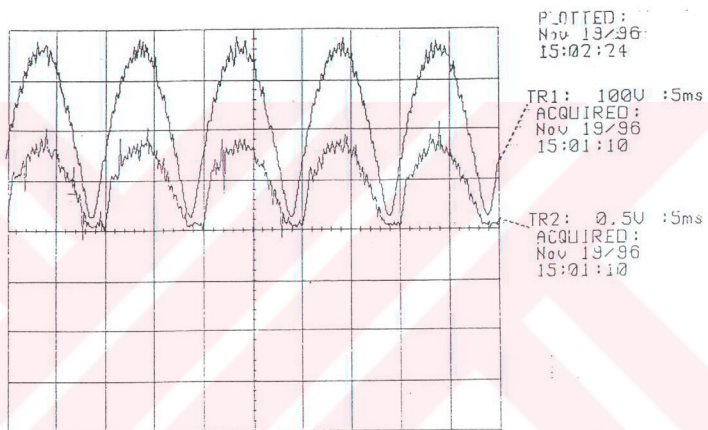
HIOKI 8850

Fig. 37. Line current harmonic spectrum at full-load

5.2 Field test results

Performance of the system for both the compressor load and the fan load have been tested in the field for various operating conditions and the corresponding current and voltage waveforms have been recorded. The boost chopper units operate in an output power range of 14-19kW for the compressor as the air tank pressure varies between 7.5 and 9.2bars while the fan load's inverter demands a constant power of 14kW. The compressor load which has a high starting torque requirement is satisfactorily started in a few seconds time by gradually increasing the frequency of applied stator voltages at a current limit of 45A rms.

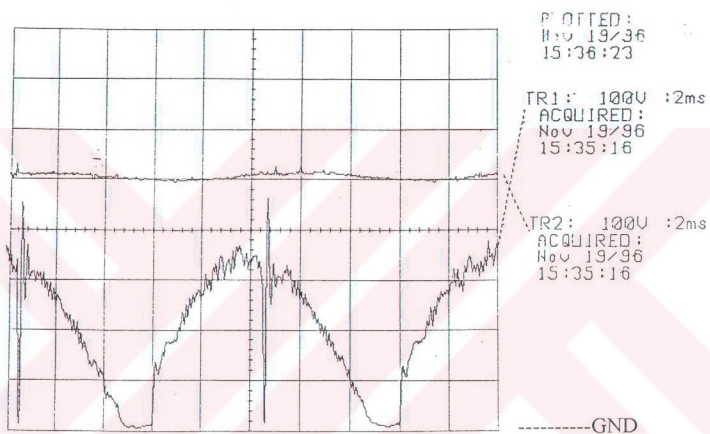
Typical rectified input current and voltage waveforms for the compressor drive at constant shaft speed and for a tank pressure of 8bars are as shown in Fig. 38. Output voltage and the rectified input voltage of the boost converter for the compressor load driven at constant speed have been recorded for the following running conditions of the electrical unit. The first has been recorded during acceleration period of traction motors (Fig. 39) and, the second while running at constant speed in the steady-state (Fig. 40). These results have shown that the asynchronous motor drive with a unity pf, 20kHz boost chopper - inverter cascade operates satisfactorily even under the worst operating conditions on the suburban lines with both compressor and fan loads.



TR1: Input voltage of the chopper circuit

TR2: Input current of the chopper circuit

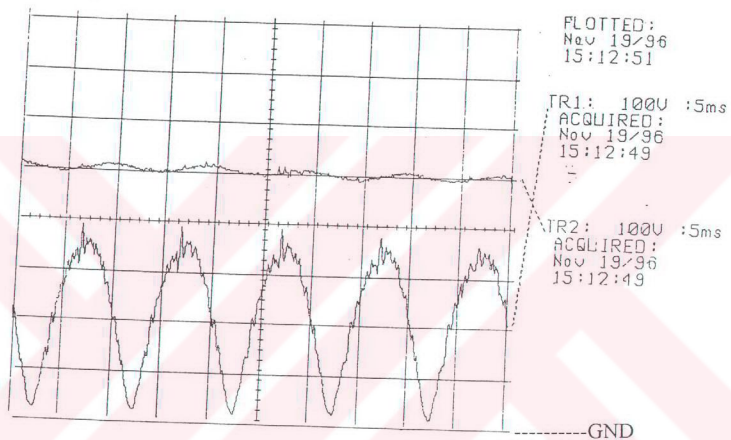
Fig. 38. Input current and voltage waveforms (Field test)



TR1: Input voltage of the chopper circuit

TR2: Output voltage of the chopper circuit

Fig. 39. Output voltage vs. input voltage during acceleration period



TR1: Input voltage of the chopper circuit

TR2: Output voltage of the chopper circuit

Fig. 40. Output voltage vs. input voltage at constant speed

CHAPTER 6

CONCLUSION

The unity power factor boost converter, three-phase inverter cascade allows the operation of a medium-power, standard three-phase ac motor from a single-phase, badly distorted ac supply whose rms voltage varies in a wide range. This results in a robust system with minimum maintenance requirement in comparison with the dc motor drive alternative. Unity pf at the ac supply side, low harmonic content of the line current waveform, high efficiency and high power density are further advantages of this system.

The two prototypes of 20kW 20kHz unity pf boost chopper - inverter cascade driving fan and compressor motors are successfully in operation as the auxiliary drive systems of E14000 type electrical unit since 1996.

The implemented boost converter preregulates the input supply voltage which varies in a wide range between 220V-20% and 220V+30% according to the predetermined specifications.

Lab and field test results have shown that the efficiency of boost converter is greater than 95% over the operating range from half-load to full-load. Total harmonic distortion is 9% and none of the harmonics exceeds 6% of the magnitude of fundamental component. The corresponding distortion power factor is negligibly small and reaches to 0.996 at full-load condition. Since the boost converter is supplied from

a distorted busbar voltage with a 3.3% total harmonic distortion, its own line current harmonic content would be better than this figure. The displacement power factor is greater than 95% over entire operating range and best at full-load condition (%99) due to the adopted active power factor correction strategy.

Developments in IGBT technology, the use of soft switching technique and larger amorphous metal (Metglas) core sizes would further increase the operating frequency and the power rating of a single unit, thus allowing the implementation of larger size systems with higher power densities.

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APPENDIX A

UC3854 DATA SHEET

FEATURES

- Control Boost PWM to 0.99 Power Factor
- Limit Line Current Distortion To <5%
- World-Wide Operation Without Switches
- Feed-Forward Line Regulation
- Average Current-Mode Control
- Low Noise Sensitivity
- Low Start-Up Supply Current
- Fixed-Frequency PWM Drive
- Low-Offset Analog Multiplier/Divider
- 1A Totem-Pole Gate Driver
- Precision Voltage Reference

DESCRIPTION

The UC1854 provides active power factor correction for power systems that otherwise would draw non-sinusoidal current from sinusoidal power lines. This device implements all the control functions necessary to build a power supply capable of optimally using available power-line current while minimizing line-current distortion. To do this, the UC1854 contains a voltage amplifier, an analog multiplier/divider, a current amplifier, and a fixed-frequency PWM. In addition, the UC1854 contains a power MOSFET compatible gate driver, 7.5V reference, line anticipator, load-enable comparator, low-supply detector, and over-current comparator.

The UC1854 uses average current-mode control to accomplish fixed-frequency current control with stability and low distortion. Unlike peak current-mode, average current control accurately maintains sinusoidal line current without slope compensation and with minimal response to noise transients.

The UC1854's high reference voltage and high oscillator amplitude minimize noise sensitivity while fast PWM elements permit chopping frequencies above 200kHz. The UC1854 can be used in single and three phase systems with line voltages that vary from 75 to 275 volts and line frequencies across the 50Hz to 400Hz range. To reduce the burden on the circuitry that supplies power to this device, the UC1854 features low starting supply current.

These devices are available packaged in 16-pin plastic and ceramic dual in-line packages, and a variety of surface-mount packages.

BLOCK DIAGRAM

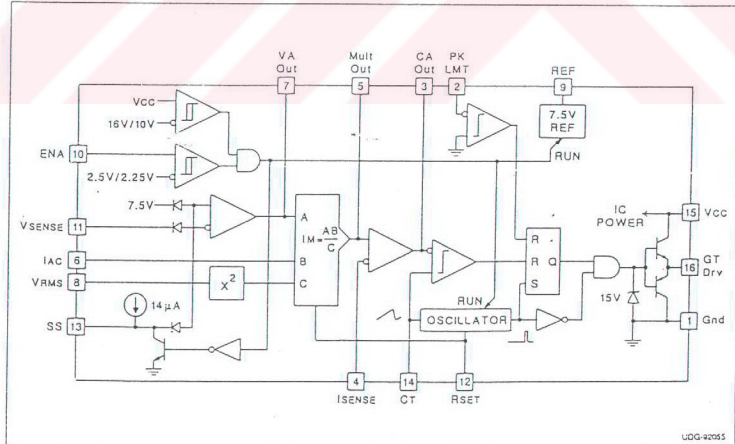


Fig. 41. UC3854 data sheet (a)

UC1854
UC2854
UC3854

ABSOLUTE MAXIMUM RATINGS

Supply Voltage Vcc	35V
GT Drv Current, Continuous	0.5A
GT Drv Current, 50% Duty Cycle	1.5A
Input Voltage, VSENSE, VRMS	11V
Input Voltage, ISENSE, Mult Out	11V
Input Voltage, PKLMT	5V
Power Current, RSET, IAC, PKLMT, ENA	10mA
Power Dissipation	1W
Storage Temperature	-65°C to +150°C
Lead Temperature (Soldering, 10 Seconds)	+300°C

Note 1: All voltages with respect to Gnd (Pin 1).

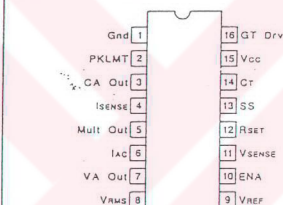
Note 2: All currents are positive into the specified terminal.

Note 3: ENA input is internally clamped to approximately 14V.

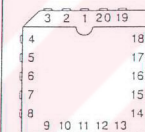
Note 4: Consult Unitorde Integrated Circuits databook for information regarding thermal specifications and limitations of packages.

CONNECTION DIAGRAMS

DIL-16 & SOIC-16
(Top View)
J, N & DW Pack-



PLCC-20 & LCC-20
(Top View)
Q & L Packages



PACKAGE PIN	FUNCTION
1	N/C
2	Gnd
3	PKLMT
4	CA Out
5	ISENSE
6	N/C
7	Mult Out
8	IAC
9	VA Out
10	VRMS
11	N/C
12	VREF
13	ENA
14	VSENSE
15	RSET
16	N/C
17	SS
18	CT
19	VCC
20	GT Drv

ELECTRICAL CHARACTERISTICS

Unless otherwise stated, VCC=18V, RSET=15k to ground, CT=1.5nF to ground, PKLMT=1V, ENA=7.5V, VRMS=1.5V, IAC=100μA, ISENSE=0V, CA Out=3.5V, VA Out=5V, VSENSE=7.5V, no load on SS, CA Out, VA Out, REF, GT Drv, -55°C<T<125°C for the UC1854, -40°C<T<85°C for the UC2854, and 0°C<T<70°C for the UC3854, and TA=TJ.

PARAMETER	TEST CONDITIONS	MIN	TYP	MAX	UNITS
OVERALL					
Supply Current, Off	ENA=0V		1.5	2.0	mA
Supply Current, On			10	16	mA
Vcc Turn-On Threshold		14.5	16	17.5	V
Vcc Turn-Off Threshold		9	10	11	V
ENA Threshold, Rising		2.4	2.55	2.7	V
ENA Threshold Hysteresis		0.2	0.25	0.3	V
ENA Input Current	ENA=0V	-5.0	-0.2	5.0	μA
VRMS Input Current	VRMS=5V	-1.0	-0.1	1.0	μA
VOLTAGE AMPLIFIER					
Voltage Amp Offset Voltage	VA Out=3.5V	-8		8	mV
Vsense Bias Current	VSENSE=0V	-500	-25	500	nA
Voltage Amp Gain		70	100		dB
Voltage Amp Output Swing			0.5 to 5.8		V
Voltage Amp Short Circuit Current	VA Out=0V	-30	-12	-5	mA
SS Current	SS=2.5V	-20	-14	-6	μA

Fig. 42. UC3854 data sheet (b)

ELECTRICAL CHARACTERISTICS

Unless otherwise stated, VCC=18V, RSET=15k to ground, CT=1.5nF to ground, PKLMT=1V, ENA=7.5V, VRMS=1.5V, IAC=100μA, ISENSE=0V, CA Out=3.5V, VA Out=5V, VSENSE=7.5V, no load on SS, CA Out, VA Out, REF, GT Drv, -55°C<TA<125°C for the UC1854, -40°C<TA<85°C for the UC2854, and 0°C<TA<70°C for the UC3854, and TA=TJ.

PARAMETER	TEST CONDITIONS	MIN	TYP	MAX	UNITS
CURRENT AMPLIFIER					
Current Amp Offset Voltage		-4		4	mV
ISENSE Bias Current		-500	-120	500	nA
Input Range, ISENSE, Mult Out		-0.3 to 2.5			V
Current Amp Gain		80	110		dB
Current Amp Output Swing			0.5 to 16		V
Current Amp Short Circuit Current	CA Out=0V	-30	-12	-5	mA
Current Amp Gain-BW Product	TA=25°C (Note 6)	400	800		kHz
REFERENCE					
Reference Output Voltage	IREF=0mA, TA=25°C	7.4	7.5	7.6	V
	IREF=0mA, Over Temp.	7.35	7.5	7.65	V
VREF Load Regulation	-10mA<IREF<0mA	-15	5	15	mV
VREF Line Regulation	15V<VCC<35V	-10	2	10	mV
VREF Short Circuit Current	REF=0V	-50	-28	-12	mA
MULTIPLIER					
Mult Out Current IAC Limited	IAC=100μA, RSET=10k	-220	-200	-180	μA
Mult Out Current Zero	IAC=0μA, RSET=15k	-2.0	-0.2	2.0	μA
Mult Out Current RSET Limited	IAC=450μA, RSET=15k	-280	-255	-220	μA
Mult Out Current	IAC=50μA, VRMS=2V, VA=4V	-50	-42	-33	μA
	IAC=100μA, VRMS=2V, VA=2V	-38	-27	-12	μA
	IAC=200μA, VRMS=2V, VA=4V	-165	-150	-105	μA
	IAC=300μA, VRMS=1V, VA=2V	-250	-225	-150	μA
	IAC=100μA, VRMS=1V, VA=2V	-95	-80	-60	μA
Multiplier Gain Constant	(Note 5)		-1.0		V
OSCILLATOR					
Oscillator Frequency	RSET=15k	46	55	62	kHz
	RSET=8.2k	86	102	118	kHz
Cr Ramp Peak-to-Valley Amplitude		4.8	5.2	5.6	V
Cr Ramp Valley Voltage		0.8	1.1	1.3	V
GATE DRIVER					
Maximum GT Drv Output Voltage	0mA load on GT Drv, 18V<VCC<35V	13	14.5	18	V
GT Drv Output Voltage High	-200mA load on GT Drv, VCC=15V	12	12.8		V
GT Drv Output Voltage Low, Off	VCC=0V, 50mA load on GT Drv		0.9	1.5	V
GT Drv Output Voltage Low	200mA load on GT Drv		1.0	2.2	V
	10mA load on GT Drv		0.1	0.4	V
Peak GT Drv Current	10nF from GT Drv to Gnd		1.0		A
GT Drv Rise/Fall Time	1nF from GT Drv to Gnd		35		ns
GT Drv Maximum Duty Cycle			93		%
CURRENT LIMIT					
PKLMT Offset Voltage		-10		10	mV
PKLMT Input Current	PKLMT=-0.1V	-200	-100		μA
PKLMT to GT Drv Delay	PKLMT falling from 50mV to -50mV		175		ns

Note 5: Multiplier Gain Constant (K) is defined by:
$$Mult Out = \frac{K \times IAC \times (VA Out - 1)}{VRMS^2}$$

Note 6: Guaranteed by design. Not 100% tested in production.

Fig. 43. UC3854 data sheet (c)

APPENDIX B

CORE DATA SHEETS

The dimensions of the cores used in comparison in the Chapter 4.1 are as given below.

1. Amorphous metal C-core, Alloy SA-1, AMCC-1000 type from Allied Signal Inc.

Mean core length, l_m :	42.7 cm
Core area, A_c :	23 cm ²
Mass, M :	7060 g
Volume, V :	981.92 cm ³
Window area, W_a :	42 cm ²
$W_a A_c$:	966 cm ⁴

2. Ferrite UU-core, N27 type material, UU93 type from Siemens

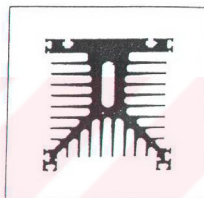
Mean core length, l_m :	35.4 cm
Core area, A_c :	8.4 cm ²
Mass, M :	1500 g
Volume, V :	297 cm ³
Window area, W_a :	35.5 cm ²
$W_a A_c$:	298.37 cm ⁴

3. Iron powder E-core, Material mix. no 33, E450 type from Micrometals Inc.

Mean core length, l_m :	22.9 cm
Core area, A_c :	12.2 cm ²
Mass, M :	1764 g
Volume, V :	280 cm ³
Window area, W_a :	12.7 cm ²
$W_a A_c$:	154.94 cm ⁴

HEATSINK DATA SHEET

Heatsink
P 3



- Intended for isolated power modules: the SEMIPACK and SEMITRANS ranges, and also for the SEMIPONT bridge rectifier range
- Available in various lengths
- Mounting channels are provided for the power modules as well as for additional accessories
- A suitable axial fan is available
- A large selection of mounting hardware is available

1) Non-standard lengths available on request
2) At the given power dissipation per semiconductor component
3) With fan type W2S 107-AA 01-16

Fig. 44. Heatsink data sheet

APPENDIX D

IGBT CHARACTERISTICS

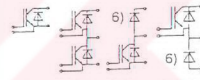
Absolute Maximum Ratings		Values		Units	
Symbol	Conditions ¹⁾	... 123 D	... 123 D1		
V _{CES}	R _{GE} = 20 kΩ	1200		V	
V _{GSR}	T _{case} = 25/80 °C	1200		V	
I _C	T _{case} = 25/80 °C; t _p = 1 ms	200 / 150		A	
I _{CM}		400 / 300		A	
V _{GES}		± 20		V	
P _{tot}	per IGBT, T _{case} = 25 °C	1250		W	
T _j (T _{arg})		- 40 ... +150 (125)		°C	
V _{isol}	AC, 1 min.	2500		V	
humidity	DIN 40 040	Class F			
climate	DIN IEC 68 T.1	55/150/56			
Inverse Diode			FWD ⁵⁾		
I _{FSM} = I _C	T _{case} = 25/80 °C	200 / 130	250 / 180	A	
I _{FSM} = I _{CM}	T _{case} = 25/80 °C; t _p = 1 ms	400 / 300	400 / 300	A	
I _{SM}	t _p = 10 ms; sin., T _j = 150 °C	1450	1800	A	
t _T	t _p = 10 ms; T _j = 150 °C	10 500	24 200	A·s	
Characteristics					
Symbol	Conditions ¹⁾	min.	typ.	max.	Units
V _{GE(th)}	V _{GE} = 0, I _C = 4 mA	≥ V _{CES}	–	–	V
V _{GE(on)}	V _{GE} = V _{CES} , I _C = 6 mA	4,5	5,5	6,5	V
I _{CES}	V _{GE} = 0 } T _j = 25 °C	–	0,2	3	mA
I _{GES}	V _{CE} = V _{CES} } T _j = 125 °C	–	12	–	mA
V _{CE(sat)}	V _{GE} = 20 V, V _{CE} = 0	–	–	1	μA
V _{CE(sat)}	I _C = 150 A V _{GE} = 15 V;	–	2,5(3,1)	3(3,7)	V
g _{fs}	I _C = 200 A T _j = 25 (125) °C	–	2,8(3,6)	–	V
C _{GH}	V _{CE} = 20 V, I _C = 150 A	95	–	–	S
C _{GS}	per IGBT	–	10	700	pF
C _{GS}	V _{GE} = 0	–	10	13	nF
C _{GS}	V _{CE} = 25 V	–	1,5	2	nF
C _{GS}	f = 1 MHz	–	0,8	1,2	nF
C _{CE}		–	–	20	nH
t _{on(1)}	V _{CC} = 600 V	–	220	400	ns
t _{on(2)}	V _{GE} = -15 V / +15 V ³⁾	–	100	200	ns
t _{on(3)}	I _C = 150 A, ind. load	–	600	800	ns
t _{off(1)}	R _{DS(on)} = R _{DS(on)} = 5,6 Ω	–	70	100	ns
E _{on(1)}	T _j = 125 °C	–	24	–	mWs
E _{on(2)}		–	17	–	mWs
Inverse Diode ⁴⁾					
V _F = V _{EC}	I _F = 150 A V _{GE} = 0 V;	–	2,0(1,8)	2,5	V
V _F = V _{EC}	I _F = 200 A T _j = 25 (125) °C	–	2,25(2,05)	–	V
V _{RO}	T _j = 125 °C	–	–	1,2	V
r _T	T _j = 125 °C	–	5	7	mΩ
I _{RRM}	I _F = 150 A; T _j = 25 (125) °C ²⁾	–	55(80)	–	A
Q _{rr}	I _F = 150 A; T _j = 25 (125) °C ²⁾	–	8(20)	–	μC
FWD of types "GAL", "GAR" ¹⁾²⁾³⁾⁴⁾⁵⁾					
V _F = V _{EC}	I _F = 150 A V _{GE} = 0 V;	–	1,85(1,6)	2,2	V
V _F = V _{EC}	I _F = 200 A T _j = 25 (125) °C	–	2,0(1,8)	–	V
V _{RO}	T _j = 125 °C	–	–	1,2	V
r _T	T _j = 125 °C	–	3	5,5	mΩ
I _{RRM}	I _F = 150 A; T _j = 25 (125) °C ²⁾	–	60(90)	–	A
Q _{rr}	I _F = 150 A; T _j = 25 (125) °C ²⁾	–	8(23)	–	μC
Thermal Characteristics					
R _{th(jc)}	per IGBT	–	–	0,1	°C/W
R _{th(jc)}	per diode / FWD "GAL", "GAR"	–	–	0,25/0,18	°C/W
R _{th(ja)}	per module	–	–	0,038	°C/W

SEMITRANS® M IGBT Modules

SKM 200 GA 123 D ¹⁾
 SKM 200 GB 123 D
 SKM 200 GB 123 D1 ⁶⁾
 SKM 200 GAL 123 D ⁶⁾
 SKM 200 GAR 123 D ⁶⁾



SEMITRANS 3



GA GB GAL GAR

Features

- MOS input (voltage controlled)
 - N channel, Homogeneous Si
 - Low inductance case
 - Very low tail current with low temperature dependence
 - High short circuit capability, self limiting to $6 \cdot I_{CEM}$
 - Latch-up free
 - Fast & soft inverse CAL diodes ⁵⁾
 - Isolated copper baseplate using DCB Direct Copper Bonding Technology
 - Large clearance (13 mm) and creepage distances (20 mm).
- Typical Applications: → B 6 - 75
- Switching (not for linear use)

¹⁾ $T_{CASE} = 25^\circ\text{C}$, unless otherwise specified
²⁾ $I_F = -I_C, V_A = 600 \text{ V}, -di/dt = 1500 \text{ A}/\mu\text{s}, V_{GE} = 0 \text{ V}$
³⁾ Use $V_{GE} = -5 \dots -15 \text{ V}$
⁴⁾ See fig. 2 + 3; $R_{DS(on)} = 5,6 \Omega$
⁵⁾ The free-wheeling diodes of the GAL and GAR types have the data of the inverse diodes of SKM 300 GA 123 D
⁶⁾ $V_{DS} = 4000 \text{ V}_{rms}$ on request
 CAL = Controlled Axial Lifetime Technology
 Cases & mech. data → B 6 - 76
⁷⁾ SEMITRANS 4 → B 6 - 88

Fig. 45. IGBT characteristics

APPENDIX E

IGBT GATE DRIVER CHARACTERISTICS

Block diagram SKH110

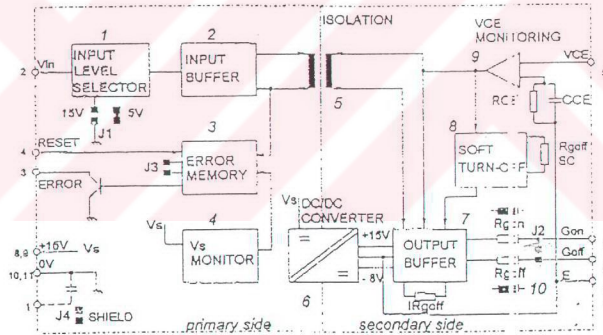


Fig. 46. IGBT gate driver block diagram

Absolute Maximum Ratings (T _a =25°C)			
Symbol	Term	Values	Unit
V _S	Supply voltage primary	18	V
V _{IH}	Input signal voltage (HIGH)		
	for 15V input level	V _S	V
	for 5V input level	V _S	V
R _{in}	Input resistance	10,0	kΩ
I _{out,peak}	Output peak current	± 8	A
I _{out,av}	Output average current	± 100	mA
V _{CE}	Collector-emitter voltage sense	1600	V
dv/dt	Rate of rise and fall of voltage (secondary to primary side)	75	kV/μs
V _{isol IO}	Isolation test voltage IN/OUT (1min)	4000	V~
T _{op}	Operating temperature	-25 ... +85	°C
T _{stg}	Storage temperature	-25 ... +85	°C

SEMIKRON®
High Power IGBT Driver
SH110
Preliminary Data

Electrical Characteristics (T _a =25°C)			Values			
Symbol	Term		min	typ	max	Unit
V _S	Supply voltage primary		14,4	15,0	15,6	V
I _S	Supply current (max.)			0,5 ¹⁾		A
I _{SO}	Supply current primary side (no load)			150		mA
V _{IT+}	Input threshold voltage (HIGH) min.					
	for 15V input level		10,0	11,0	11,8	V
	for 5V input level		1,7	1,8	2,1	V
V _{IT-}	Input threshold voltage (LOW) max.					
	for 15V input level		3,8	4,0	4,4	V
	for 5V input level		0,50	0,65	0,80	V
V _{O(on)}	Turn-on output gate voltage			+15		V
V _{O(off)}	Turn-off output gate voltage			-8		V
f	Maximum operating frequency		see Fig.15			
t _{d(on)to}	Input-output turn-on propagation time			1,0 ²⁾		μs
t _{d(off)to}	Input-output turn-off propagation time			1,0 ²⁾		μs
t _{d(err)}	Error input-output propagation time			1,0 ³⁾		μs
V _{CEtest}	Reference voltage for VCE monitoring ⁴⁾			5,2		V
R _{ON}	Internal gate resistor for ON signal			22 ⁴⁾		Ω
R _{OFF}	Internal gate resistor for OFF signal			22 ⁴⁾		Ω
C _{ps}	Primary to secondary capacitance				8	pF

Features

- Single driver circuit for high power IGBTs
- Drives all SEMIKRON IGBTs with V_{CE} up to 1600V
- CMOS/TTL/HCMOS compatible input buffers
- Short circuit protection by V_{CE} monitoring
- Soft short circuit turn-off
- Isolation due to transformers (no opto couplers)
- Supply under voltage monitoring (<13V)
- Error memory / output signal
- Internal isolated power supply (LOW or HIGH logic)

Typical Applications

- High frequency SMP3
- Braking choppers
- Asymmetrical bridges
- High power UPS
- DC bus voltages up to 1200V
- Single high power switches or paralleled IGBTs (up to 400A/1600V @ 20kHz)

1) This current value is a function of the maximum output load condition

2) see also Fig. 10

3) This value does not consider I_{ON} of IGBT and I_{OFF} adjusted by R_{CE} and C_{CE}; see also Fig. 13

4) matched to be used with IGBTs <100A; for bigger currents, see Table 2

5) with R_{CE}=15kΩ, C_{CE}=330pF; see Fig. 8

Fig. 47. IGBT gate driver characteristics

APPENDIX F

COMPRESSOR CHARACTERISTICS

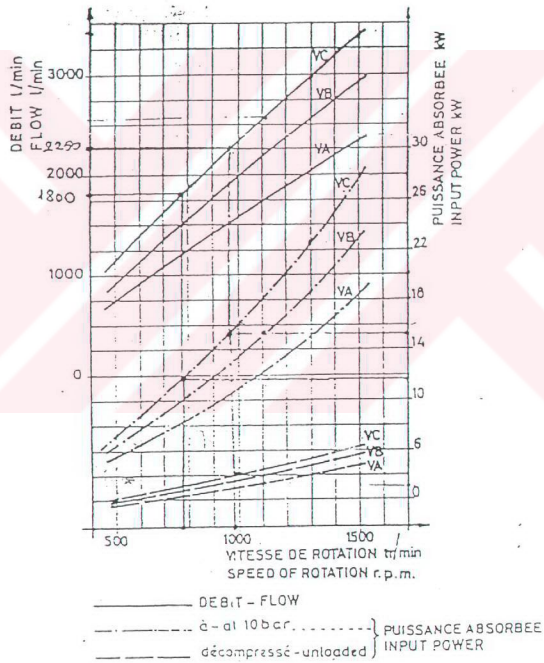


Fig. 48. Compressor characteristics

APPENDIX G

FAN CHARACTERISTICS

The two types of the fans used in the E14000 type electrical units and corresponding fan characteristics are given in Table 8.

Table 8. Fan Characteristics

	Fan 1	Fan 2
Code	Ventmeca MB-14	Ventmeca MB-14
Type	Centrifugal	Centrifugal
Power Absorbed	7.84 kW	10.12kW
Speed	2520 rpm	2770 rpm
Ventilation Direction	vertically down	vertically down