Design of an Optimally Self-Protected Medium Frequency Power Supply for Induction Heating Purposes

Abdul Kareem M. Obais

Department of Elect. Eng. , College of Engineering, University of Babylon

Abstract

In this paper, the basic elements of the induction heating process are reviewed. The expected risks accompanying the power supplies when they are operated with induction-inductors are stated and a remedial solution is presented. The solution is a complete design of a power supply that operates without any restriction among a variety of widely extending ratings of induction-inductors. This supply is identified with an accurate load current limitation technique and is self-protected against real short circuits and extremely excessive loads. Its frequency can be smoothly varied from 500Hz to 10 KHz. It is easily possible to control load power continuously while induction heating process is proceeding.

الخلاصة

في هذا البحث تمت مراجعة العناصر الأساسية الداخلة في عملية التسخين الحثي. تم تحديد الأخطار المتوقعة من عملية تحميل مجهزات القدرة بملفات التسخين الحثي وكذلك تم تقديم الحل العلاجي لها. الحل يكمن بتصميم كامل لمجهز قدرة يعمل بدون أي قيد وسط تشكيلة من الملفات الحثية ذات المديات التقديرية الواسعة. يمتاز هذا المجهز باحتواءه على تقنية دقيقة لتحديد تيار الحمل وتقنية للحماية الذاتية ضد الدوائر القصيرة الحقيقية والأحمال المفرطة جدا. هنالك إمكانية في هذا المجهز لتغيير التردد بصورة سلسة من 500Hz إلى 10 KHz. وكذلك يمكن السيطرة على قدرة الحمل بشكل مستمر بينما عملية التسخين الحثي جارية.

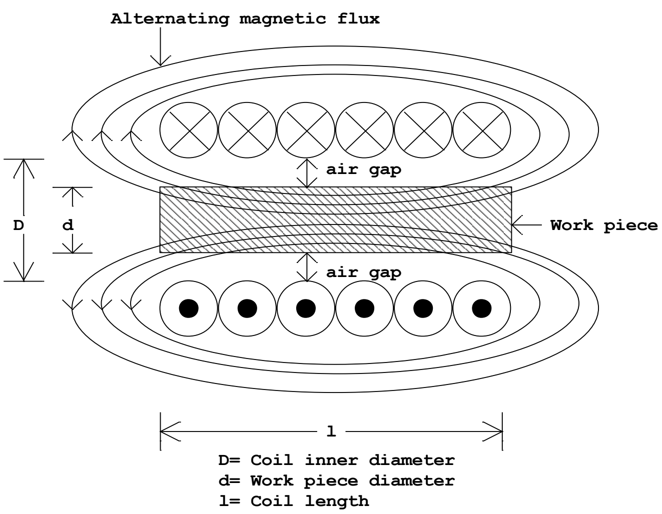
1. Introduction

Induction heating systems are based mainly on electromagnetic induction. Placing a conducting material in a varying magnetic field will cause eddy currents to flow inside the material. The generated eddy currents in turn will heat the material. The induction-inductor [Davies and Simpson 1979, Rev 2000] is one of basic means used to invest the heat produced by eddy currents phenomenon. The induction-inductor is simply a solenoid within which the material to be heated is placed. The AC power supplies are the main feeders of the induction-inductors. Single-phase inverters are generally representing the most of these supplies [Skvarenina 2001, Rashid 2001]. They are either voltage-source type or current-source type. The solid-state switching devices are the basic building blocks of inverters [Hinchliffe and Hobson 1986]. Thyristors, power MOSFETs, power transistors, and insulated-gate transistors (IGBT's) are the widely used solid-state switching devices in designing power supplies for induction heating purposes. The full-bridge voltage-source inverter [Skvarenina 2001 and Rashid 2001] is widely used as an induction heating power supply. Also single switch AC-AC converters [Shenkman *et al* 2004], the series resonant converters [Koertzen *et al* 2001], and LC resonant oscillators are used for induction-heating purposes. Power supplies are designed to operate at certain frequencies and certain volt-ampere ratings. Assuming a constant operating DC voltage V_{DC}, each supply has a specified frequency band and a maximum current rating. Obviously, the nature of load impedance determines the magnitude and shape of the load current. Currents exceeding maximum limits are certainly representing risks to power supplies, and definitely, the solid state-switching devices are the most probable to be damaged first [Jain and Diwan 1988, Koller and Tevan 1999]. Hence, an accurate selection of the load impedance plays an important role in the protection process beside the external protection circuitry [John *et al* 1999].

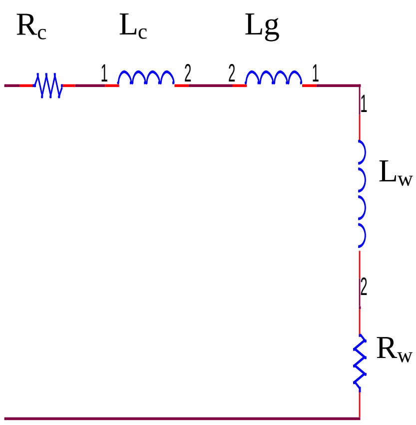
The optimally self-protected power supply is different from conventional power supplies since it automatically detects any excessive load and soon prepares the required control signals such that the maximum current rating is not exceeded. It is provided with an accurate load current limiting circuit and a reliable protection circuit against excessive loads and real short circuits. It is preferred to present a summarized background about the induction-inductor and the conventional voltage-source full –bride single-phase inverter in order to state the insisting need to the optimally self-protected power supply.

2. The induction-inductor concept

The induction inductor is an important element in induction heating process. It may be a single turn of a copper wire or a multi-turn coil wound in such a way that it can serve a certain function [David 2001]. Generally, in many industrial applications, the induction inductor is a solenoid surrounding the conducting material to be heated. It is considered as loosely coupled transformer [Koller and Tevan 1999], where the coil itself represents the primary side of the transformer, and the work piece (the material to be heated) represents the secondary side. Fig.1.a shows a longitudinal section of a certain AC current carrying inductor, and inside it, a work piece is inserted. It is clearly understood that the coil is not tightly surrounding the work piece or in other words, there is an air gap separating between the coil and the work piece.



(a)



(b)

Fig.1. (a) A longitudinal-section of an induction-inductor, (b) its equivalent circuit.

The circuit parameters are given by

$$L_c = \frac{\delta_c \mu_o K_c N^2 \pi D}{2l} \quad \text{.................}(2.1)$$

$$R_c = \frac{\delta_c \mu_o K_c N^2 \omega \pi D}{2l} \quad \text{...............}(2.2)$$

$$L_g = \frac{\mu_o N^2 \omega\pi (D^2 \_ d^2)}{4l} \ldots\ldots\ldots..(2.3)$$

$$L_w = \frac{\delta_w \mu_o \mu_r N^2 \pi d}{2l} \ldots\ldots\ldots\ldots\ldots..(2.4)$$

$$R_w = \frac{\delta_w \mu_o \mu_r N^2 \omega\pi d}{2l} \ldots\ldots\ldots\ldots..(2.5)$$

Where, Rc : coil resistance (primary side resistance of the assumed transformer).
L_c : coil inductance (primary side inductance of the assumed transformer)
L_g : air gap inductance
K_c: coil constant which is normally of the range 1.1 to 2 depending upon the shape and spacing of the coil conductors.
N : coil number of turns.
δ : skin depth and is given by $\delta = \sqrt{2\rho / \mu\omega}$
ρ : resistivity of the material.
δ_c: copper skin depth.
δ_w: work piece skin depth.
μ_o: absolute permeability.
μ_r: the work piece relative permeability.
ω : angular frequency of the power supply.
R_w: work piece resistance referred to primary side.
L_w: work piece inductance referred to primary side.
The total impedance of the inductor is given by

$$Z_{tot} = \sqrt{(\omega L_{tot})^2 + (R_{tot})^2} \ldots\ldots\ldots...(2.6)$$

, and $L_{tot} = L_c + L_g + L_w$ Where, $R_{tot} = R_c + R_w$

It is clear that Z_{tot} is a function of two temperature dependent parameters ρ and μ_r. Where ρ increases with temperature and this fact is generally valid for metals like mild steel, which resisitivity varies from 0.2μΩm to 1.1Ωm for a temperature change of 20°C to 800°C respectively. All Ferro-materials lose their magnetic properties above Curie point (~720°C). This means that μ_r for these materials tend to one when temperature exceeds Curie point. The inductor itself is made of copper and is carefully cooled; hence, ρ_c & δ_c remain almost constant. Consequently, the important temperature dependent parameters in Z_{tot} are R_w and L_w. When Ferro-materials are heated, these two parameters decreases largely as temperature rises above Curie point. For example, consider a mild steel work piece heated from 20°C to 800°C. For this temperature change, μ_r varies from 50 to one, and the corresponding steel skin depth varies from 4.5mm to 74.7mm. Consequently L_s and R_s values at 800°C decreases to ⅓ their corresponding values at 20°C. In fact an associative reduction in Z_{tot} will occur. Note that the steel is saturated at a magnetic flux density of 2 Tesla and a starting value of 50 given to μ_r is satisfactory in induction heating. It is reasonably to start the induction heating process with a certain Z_{tot} for a cold work piece and end with ½ Z_{tot} when the material is heated above Curie point.

3. The full-bridge single-phase inverter

Fig.2.a shows a full-bridge single-phase inverter. It consists of four solid-state switches S_1, S_2, S_3, and S_4. If the load of this inverter is an induction-inductor, then the load current waveform is determined by Z_{tot} parameters, V_{DC}, and the operating frequency f. The peak value of the load current I_m is given by

$$I_m = \frac{V_{DC}}{R_{tot}}\left[1 - \exp\left(-\frac{0.5T - t_D}{\tau}\right)\right] \text{......(3.1)}$$

Where, T: load current repetition time in seconds = 1/f

τ: load impedance time constant = L_{tot}/R_{tot}

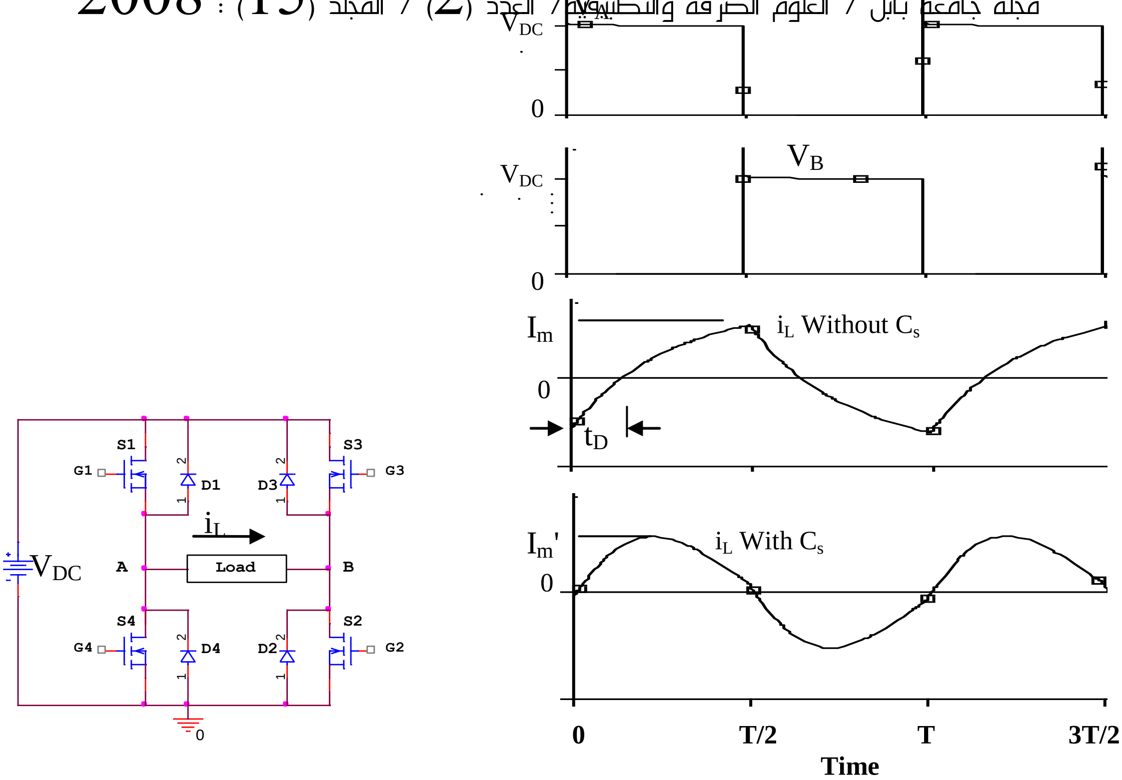
t_D: the conduction time of feedback diodes D_1,D_2 or D_3,D_4 and is given by

$$t_D = -\tau \ln\left(\frac{1+\exp(-0.5T/\tau)}{2}\right) \text{..................(3.2)}$$

D_1, D_2 or D_3, D_4 are called feedback diodes because they feedback energy to the DC source when none of the switching devices is conducting. Note that, the switches S_1, S_2 or S_3,S_4 conduct for the time interval (T/2 $-t_D$). The positive portion of load current starts at the instant when S_1 and S_2 conduct simultaneously. This portion grows increasingly until S_1 and S_2 are turned off. Meanwhile, the diodes D_3 and D_4 start conduction feeding back energy to the DC source and the load current in this instance continues to flow decreasingly through D_3, V_{DC}, and D_4. As the current decays to zero, D_3 and D_4 are no longer conducting and S_3, and S_4 conduct starting the negative portion of load current. A similar process will repeat for this portion except that the feedback diodes in this case are D_1 and D_2. When a series capacitor C_S is connected with Z_{tot}, then the peak current I_m' is given by [Rashid 2001]

$$I_m' = \sqrt{\sum_{n=1,3,5,\dots}^{\infty} I_n^{\,2}} = \sqrt{\sum_{n=1,3,5,\dots}^{\infty} \frac{V_n^{\,2}}{Z_n^{\,2}}} = \sqrt{\sum_{n=1,3,\dots}^{\infty} \frac{\left(4V_{DC}/n\pi\right)^2}{R_{tot}^{\,2} + \left(n\omega L_{tot} - 1/n\omega C_s\right)^2}} \text{..............(3.3)}$$

Where I_n represents the nth harmonic of the load current, V_n is nth harmonic of the load voltage V_{AB} and Z_n is the impedance to V_n. Fig.2.b shows the voltage and current waveforms of this supply. If $\omega L_{tot}=1/\omega C_S$, then the fundamental component of the load current I_1 is $4V_{DC}/\pi R_{tot}$. This current will be very risky when R_{tot} decays to smaller values due to temperature growth in side the work piece as the induction heating process is progressing. It can be easily deduced that using such kinds of inverters for induction heating purposes mean that the load currents of these inverters may be started with lower values and finally approach their maximum ratings at the end of induction heating process.



(a) (b)

Fig.2. (a) The full-bridge single-phase inverter, (b) its voltage and current waveforms.

4. The optimally self-protected power supply schematic design

The above brief background of the induction-inductor and the full-bridge inverter indicates the insisting need to the optimally self-protected power supply. Fig.3 shows the block diagram of this supply. Obviously, it is a full-bridge single-phase inverter when the power circuit is considered only, but its electronic circuitry means a new approach in controlling and protection techniques. The solid-state switching devices used here are the insulated-gate bipolar transistors (IGBT's). They are snubberless devices and their turning on and turning off are normally achieved through appropriate triggering signals applied directly to their gates.

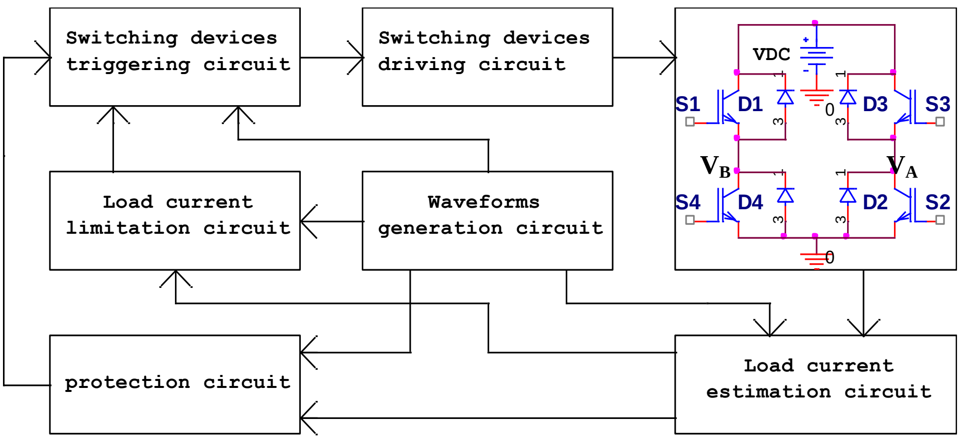


Fig .3. The optimally self-protected power supply block diagram.

The switching devices triggering signals are shown in Fig.4. These signals are totally governed by the load current limiting circuit and the protection circuit. The time t_1 is determined according to turn-on and turn-off times of the switching devices used in designing process in order to avoid the probability of short circuits through S_1,S_4 or S_3,S_2. In addition, t_1 is recommended to be sufficient for charging a bootstrapped capacitor from zero to +15V. Note that two bootstrapped capacitors are used in the driving circuit to prepare the required bias voltages for driving S_1 and S_3. The time t_2 is variable and is only determined by the limiting circuit. Its value is zero during normal load operation. Normal loads mean those loads, which are operated safely by a similar rating conventional full-bridge single-phase inverter. As it is shown above, t_2 has a maximum value of ($0.5T\_2t_1$). This range of t_2 permits an extending range of excessive loads to be treated here safely. The load current estimation circuit is designed such that the saturation collector-to-emitter voltage (V_{CEsat}) of S_2 or S_4 is directly detected by sampling a small fraction of V_A when S_2 is triggered on or V_B when S_4 is triggered on. Of coarse V_{CEsat2} and V_{CEsat4} are directly proportional to the currents flowing through S_2 and S_4 respectively. V_{CEsat2} corresponds to the positive half-cycle of the load current whereas V_{CEsat4} corresponds to the negative half-cycle. V_{CEsat2} and V_{CEsat4} are coupled and amplified to form what is known as V_{est}. The analogue voltage V_{est} is fed to the load current limiting circuit and compared with an adjustable DC voltage there. If V_{est} is less than it then, no action is taken by the load current limiting circuit and the triggering signals waveforms will be as shown in Fig.4 except that t_2=0. When V_{est} grows slightly above that DC voltage then, the limiting circuit sends two signals controlling the pulse-width of the positive portions of V_{S1} and V_{S2}. Again the triggering signals will be as shown in Fig.4, but in this case, $t_2 \neq 0$ and is determined by the amount by which the load

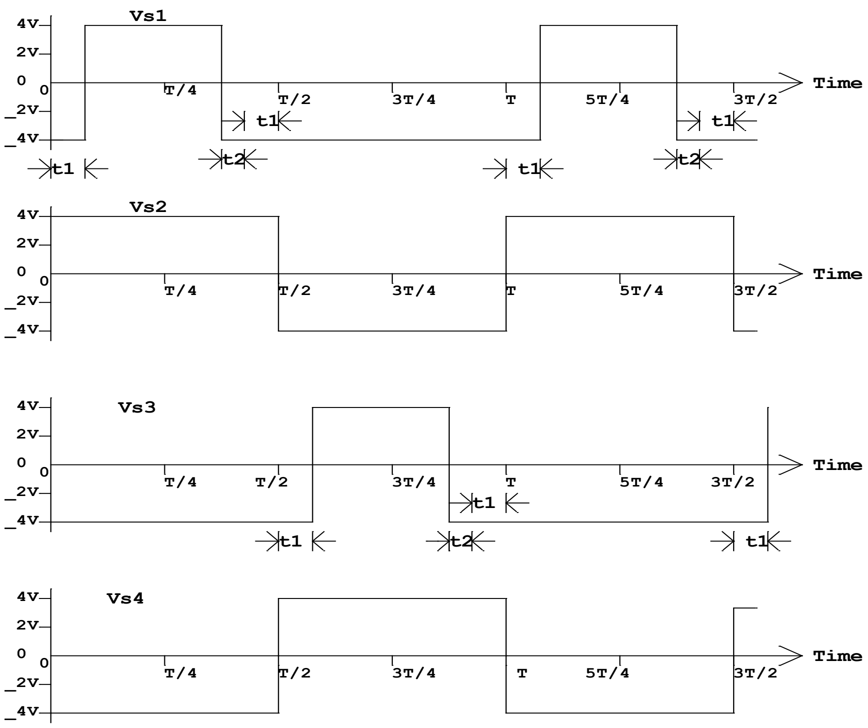


Fig.4. The triggering signals of the optimally self-protected power supply.

current is excessive.

The protection circuit is designed such that when a short circuit exists or an extremely excessive load is applied, then it immediately responds and generates four controlling signals to disable the triggering signals of the switching devices. The waveforms generation circuit produces a triangular waveform of amplitude of about 3V. Its frequency can be smoothly varied from 500Hz to 10 KHz. This frequency range lies within the medium frequency band of the induction heating frequency spectrum [Davies and Simpson 1979]. It is the most useful frequency range to heat or melt all types of steel [Rev 2000]. Another necessary waveform produced in this circuit is the rectangular waveform, which is obtained from the triangular waveform shown in Fig.5. The time t_3 is recommended to be within the range of ($t_1 \leq t_3 \geq 2t_1$) in order to guarantee optimal operation of the load current limitation circuit.

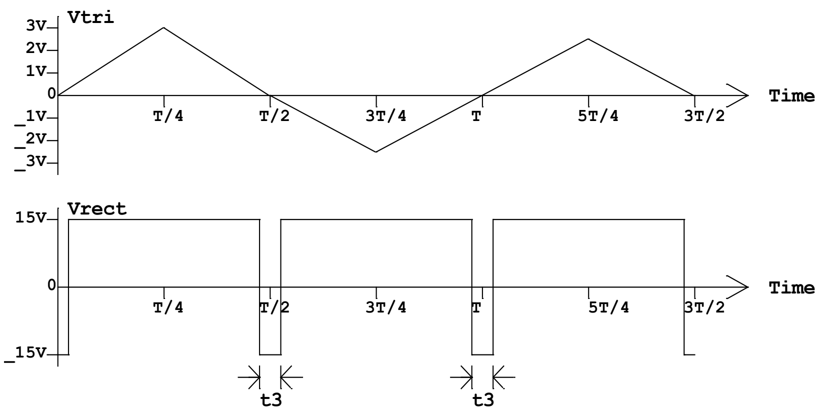


Fig.5. The triangular and rectangular waveforms.

The triangular and rectangular waveforms constitute the timing system of this supply.

The switching devices triggering circuit produces the triggering signals shown in Fig.4. It involves four comparators governed by four controlling signals coming from limiting and protection circuits. The driving circuit of the switching devices is designed such that at the maximum current rating specified for the power supply, the total switching energy losses of the switching devices will be within or less than typical values. The power circuit includes four IGBT's, four fast recovery free-wheel diodes, and a constant DC source, which is obtained from a three-phase full-bridge rectifier.

5. The power supply circuits design

The datasheets of components are carefully considered in the designing process. Fig.6 shows the complete circuit diagram of the optimally self-protected power supply. The system circuit diagram is exactly representing the schematic design discussed above except that; the general descriptions are transformed to actual circuits. These circuits are designed using components, which have fast speeds; hence, the supply response within its specified frequency band is reliable. The designed power supply has been built using PSpice. The results obtained from its operation among a variety of widely extending ratings of induction-inductors are interesting. For better understanding to system performance, it is preferred to study the circuits forming Fig.6 individually.

.

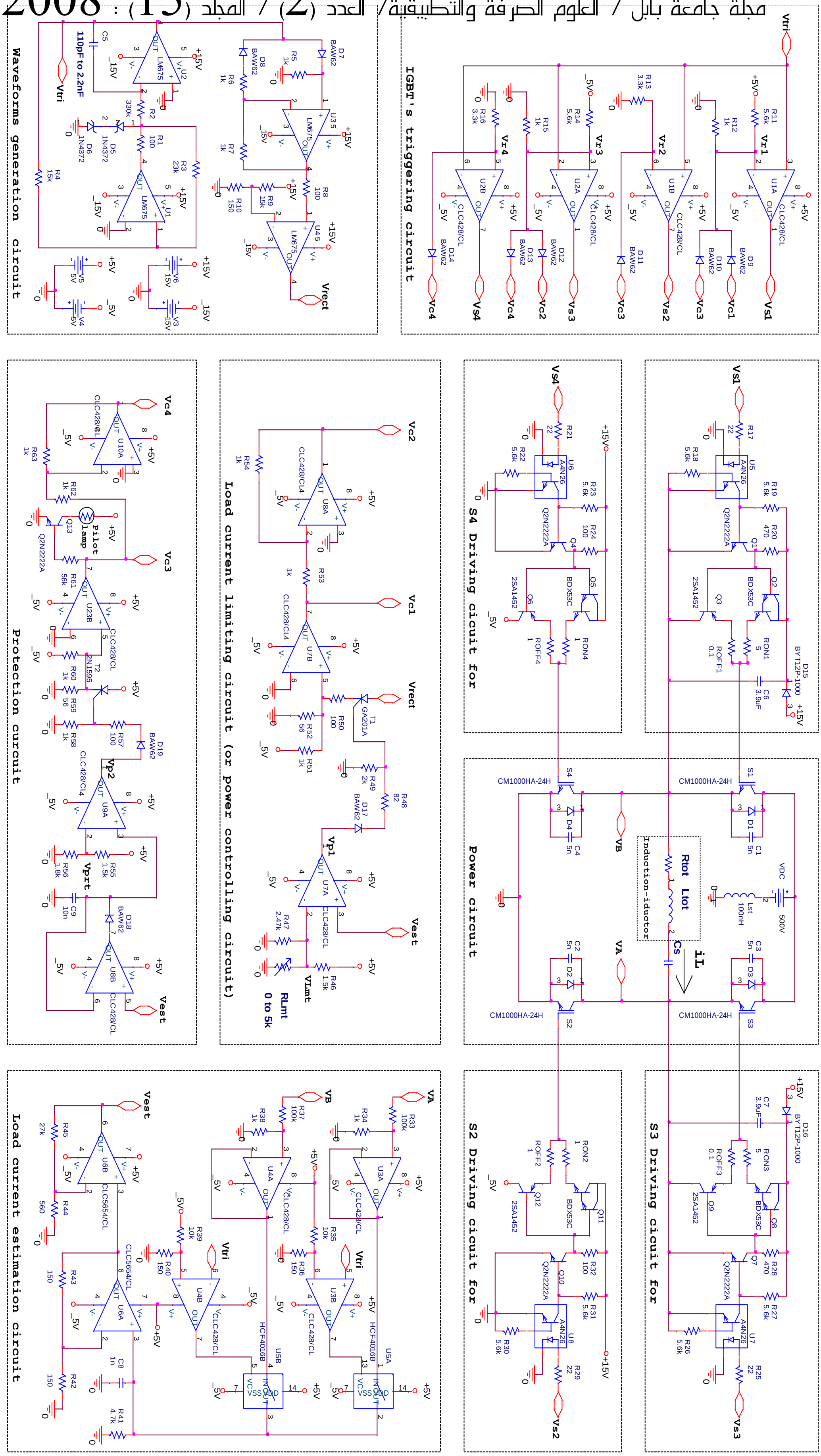
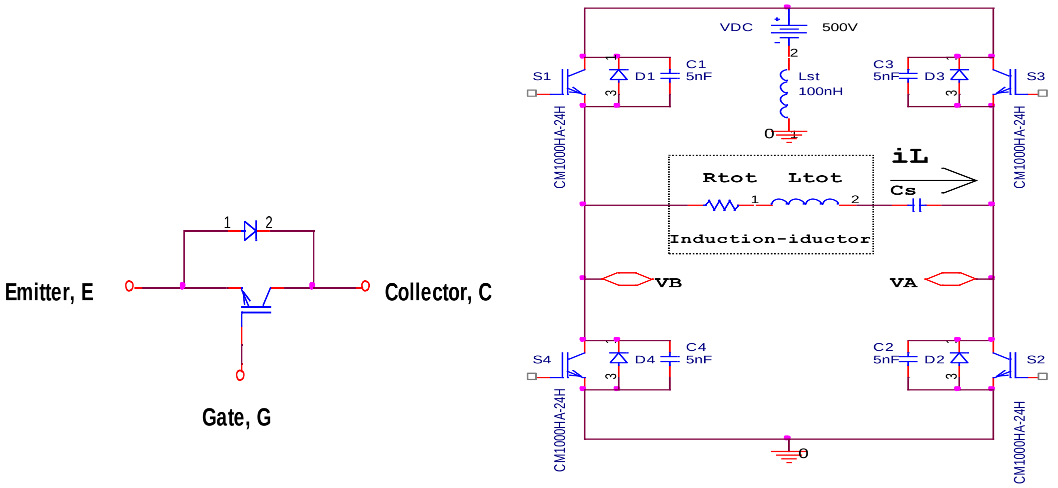


Fig.6.The optimally self-protected power supply circuit diagram.

5.1. The power circuit

The power circuit includes four IGBT modules. They are of the type CM1000HA-24H. Each module consists of one IGBT in a single configuration with a reverse-connected super-fast recovery free-wheel diode as shown in Fig.7a. Each configuration has the following maximum rating at 25C^o: V_{CE}=1200V, I_C(continuous)=1000A, I_C(pulsating)=2000A, V_{GE}=±20V, Collector dissipation, PC=5800W, I_E(continuous)=1000A, I_E(pulsating)=2000A. Its junction temperature range is _40C^o to 150C^o. A 500V DC voltage source is used and the stray inductance is considered here by adding a reasonable stray inductance of a value of 100nH in series with V_{DC}. Of course, the load of the supply is an induction-inductor. Since, the power factor of Z_{tot} is often poor, a series capacitor C_s may be preferred to be used here. The shunting capacitors C_1, C_2, C_3, and C_4 are used for protecting the power switching devices against excessive dv/dt.



(a) (b)

Fig.7 (a) The IGBT module, (b) the power circuit.

5.2. The IGBT driving circuit

S_1 and S_2 driving circuits are identical and including the two bootstrapped capacitors C_6 and C_7 shown in Fig.8. When S_4 is triggered on, C_6 charges to +15V through D_{15} and S_4. The time t_1 must be sufficient to charge C_6 from zero to +15V.

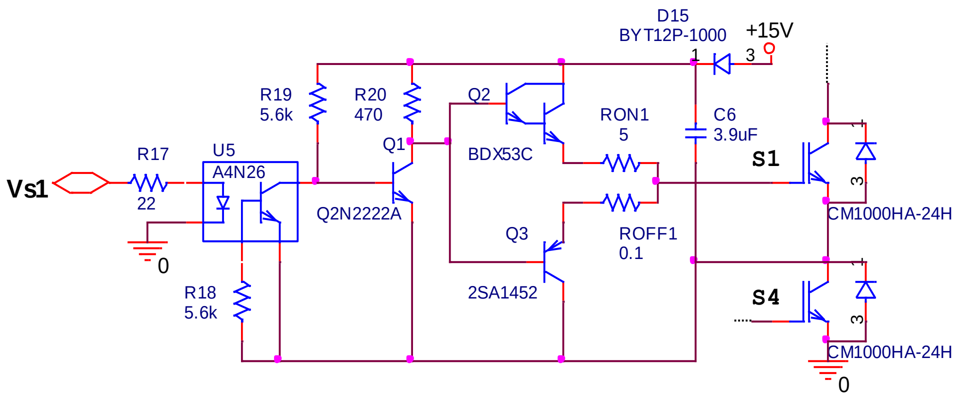


Fig.8.The IGBT driving circuit (namely S_1 driving circuit).

Turn-on and turn-off times of the IGBT are counted for by choosing the appropriate circuit parameters. S_2 and S_4 driving circuits are a bit different from that of S_1 and S_3. The values of R_{OFF} and R_{ON} are set such that the switching energy losses, dv/dt, and di/dt are as low as possible.

5.3. The waveforms generation circuit

This circuit produces a triangular waveform V_{tri} of a frequency, which can be varied directly by using the variable capacitor C_5 shown in Fig.9. The frequency of this waveform is given by [Millman and Halkias 1983]

$$f = \frac{R_3}{4R_2R_4C_5}$$

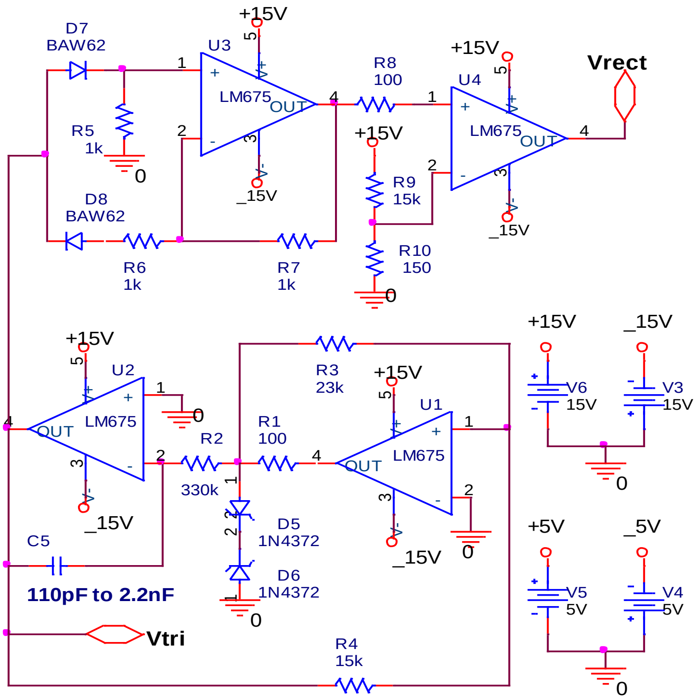
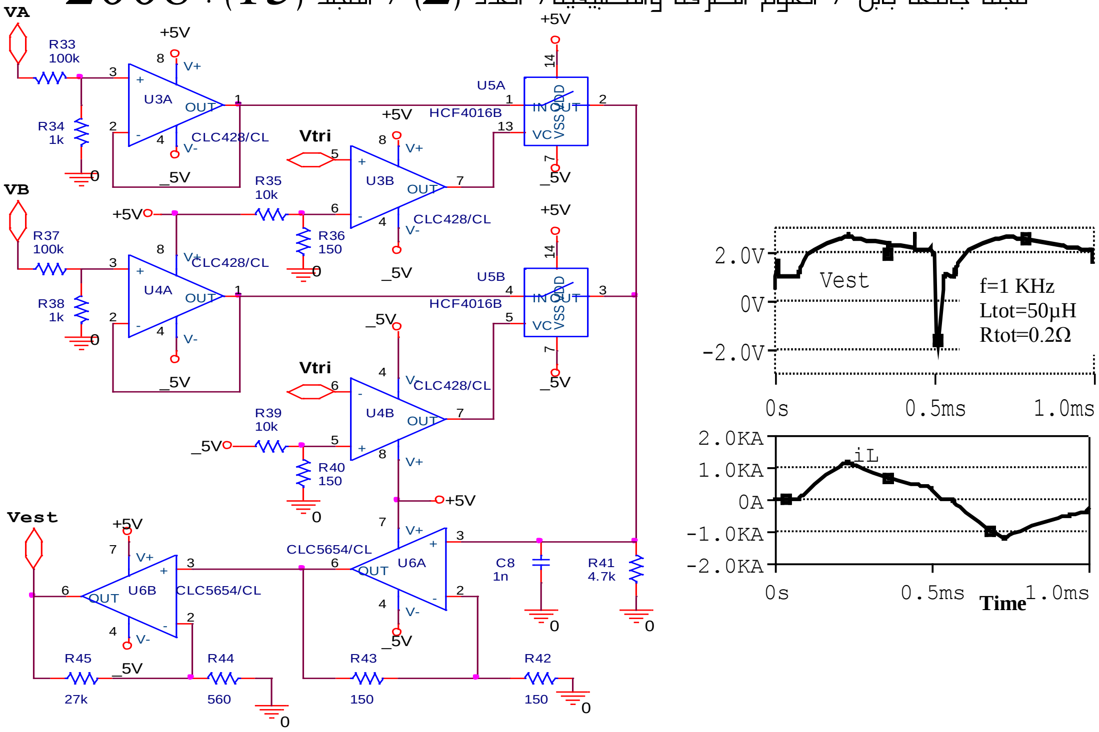


Fig.9. The waveforms generation circuit.

The above waveform is fed to the full-wave precision rectifier U_3 and then compared with the DC voltage appears across the resistor R_{10}. Consequently, the rectangular waveform V_{rect} will be obtained. These waveforms are used later in the controlling process of the power supply.

5.4. The load current estimation circuit

The instantaneous voltages V_A and V_B across the load terminals are stepped down and then sampled at the intervals where S_2 and S_4 are triggered on respectively. The outputs of the samplers U_{5A} and U_{5B} are summed through the amplifier U_{6A} and then amplified by the amplifier U_{6B}. Really, the output of U_{6B} represents the exact saturated voltages V_{CEsat2} and V_{CEsat4} instantaneously. This voltage is called V_{est}. Fig.10.a shows this circuit and Fig.10.b shows V_{est} corresponding to a certain excessive load current. The outputs of the comparators U3B and U4B approximately coincide with the triggering signals V_{S2} and V_{S4} respectively. The only differences are those very small times left on both sides of the positive portions of V_{S2} and V_{S4} in order to offer the sufficient time for their corresponding devices to be turned on.



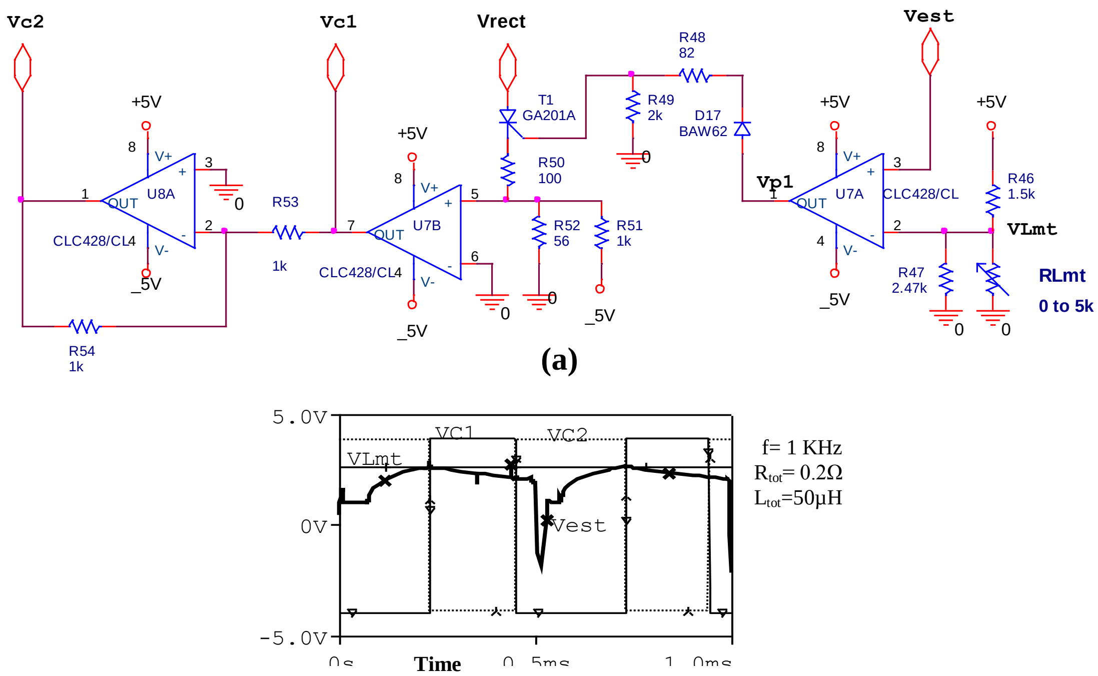
(a) (b)

Fig.10. (a) The load current estimation circuit, (b) its output for a certain load test.

.

5.5. The load current limitation circuit:

Since the peaks of V_{est} correspond to the absolute peaks of the load current, it is possible to limit both positive and negative portions of the load current symmetrically. Fig.11.a shows the load current limitation circuit. The tests made for



(a)

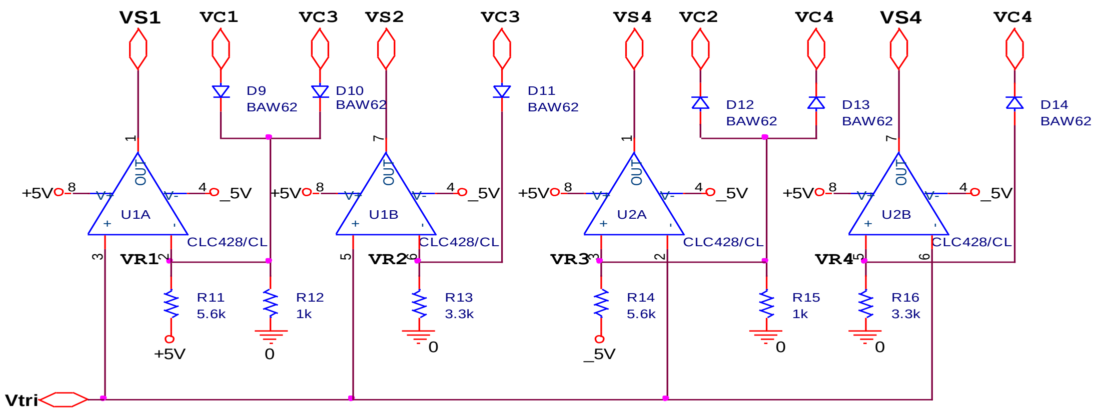
(b)

Fig.11. (a) The load current estimation circuit, (b) its input and output voltages.

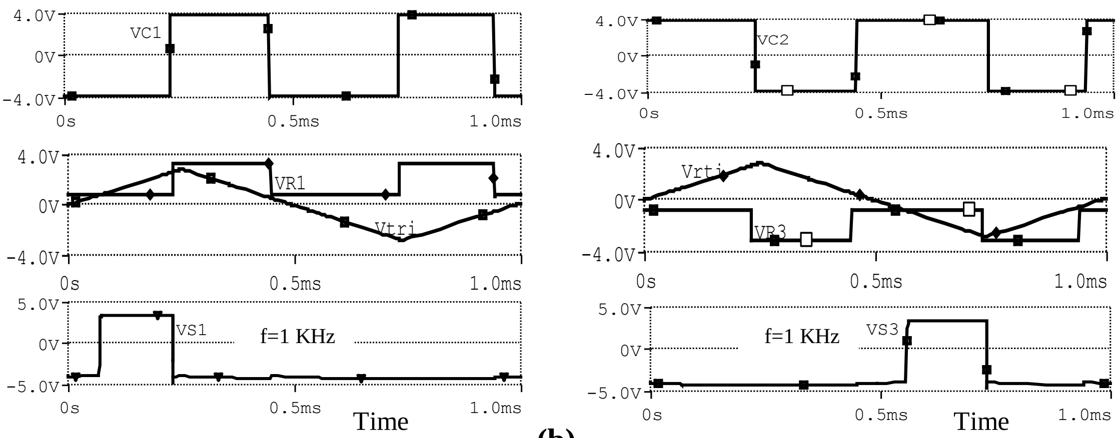
The IGBT used in this power supply at the specified frequency band, allow a safe maximum peak of 1500A to the load current at 25C$^{o.}$ For more safety, the parameters are chosen such that circuit the load current is limited at 1200A. Note that the 5kΩ position of R_{Lmt} corresponds to this value. The positive half-cycle of load current starts when S_1 and S_2 conduct simultaneously. In this interval, the load current grows increasingly. If V_{est} grows slightly greater than V_{Lmt} then, the output of the comparator U7A toggles to +4V.Consequently, the fast thyristor T1 is turned on. Because of this action, V_{C1} toggles to +4V and successively S_1 is turned off. At this instant, the load current continues to flow through S_2 and D_4 decreasingly until S_2 is turned-off. Here S_2 behaves as a free-wheel diode. Although at this instant S_4 is triggered on, the load current continues to flow through D_3, V_{DC}, and D_4 until it decays to zero. As the diodes, D3 and D4 stop conduction, S_3 and S_4 conduct and the negative half-cycle of load current starts increasingly. When the negatively increasing portion of the load current negative half-cycle touches the specified peak, V_{est} again grows above V_{Lmt} and the corresponding action is that S_3 is turned off. The load current continues to flow through S_4 and D_2 and then through D_1, V_{DC}, and D_2. S_1 and S_3 are turned off while they are carrying peak currents. Hence, their turn-off losses are higher than those of S_2 and S_4, which are turned off at somewhat lower currents.

5.6. The IGBT's triggering circuit:

This circuit includes four comparators as shown in Fig.12.a. The reference-



(a)



(b)

Fig.12. (a) The IGBT's triggering circuit, (b) its response to V_{C1} and V_{C2}.

voltages V_{R1}, V_{R2}, V_{R3}, and V_{R4} are governed by the controlling signals V_{C1}, V_{C2}, V_{C3}, and V_{C4}. If D_1, D_2, D_3, and D_4 are all off, then V_{R1}=_V_{R3}= 0.757 V and V_{R2}= V_{R4}= 0. In this case, the comparators outputs will be as shown in Fig.4, except that; t_2= 0. The controlling signal status is either +4V or _4V. V_{C1} and V_{C3} are active when they are +4V, while _4V means that V_{C2} and V_{C4} are active. The activity of a controlling signal here means its ability to lead its corresponding diode to on condition. The effects of V_{C1} and V_{C2} on V_{S1} and V_{S3} are indicated in Fig.12.b. V_{S2} and V_{S4} are not affected by V_{C1} and V_{C2}. Note that V_{C1} is effective only during the positive half-cycle of the triangular waveform V_{tri}, while V_{C2} is effective during the negative half-cycle.

5.7. The protection circuit:

This is designed to reject short circuits and extremely excessive loads. It is set at a current level slightly greater than the maximum value specified for the load current limitation circuit. This circuit is shown in Fig.13. When the voltage V_{est} grows slightly above V_{prt}, V_{p2} toggles to +4V and the thyristor T_2 is switched on. V_{C3} and V_{C4} will be +4V and _4V respectively. Consequently, the diodes D_{10}, D_{11}, D_{13}, and D_{14} of Fig.12.a conduct and the outputs of their corresponding comparators become _4V. Note that T_2 cannot be switched off unless the +5V DC voltage is switched off. In this case, the pilot lamp will light indicating that the inconvenience recovery implies supply shut down.

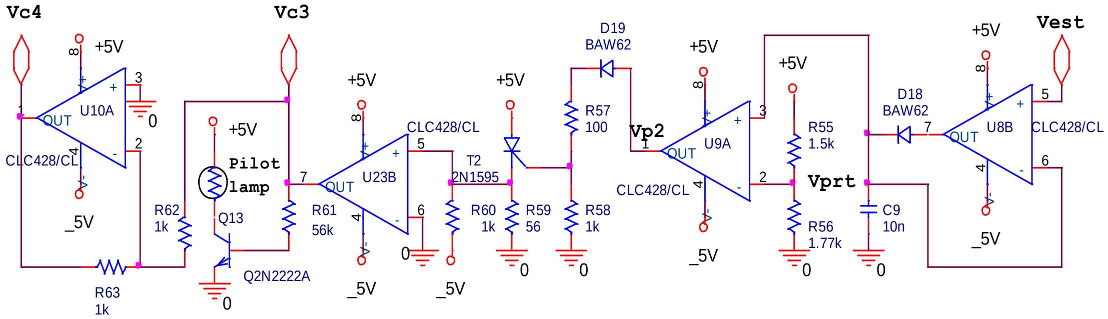


Fig.13. The power supply protection circuit.

6. Results and conclusions:

The supply was tested at 27^{o}C on PSpice simulator within its specified frequency band and among a variety of loads. The extremely excessive load tests are shown in Fig.14. The first test shows an extremely excessive load operated at 500Hz,

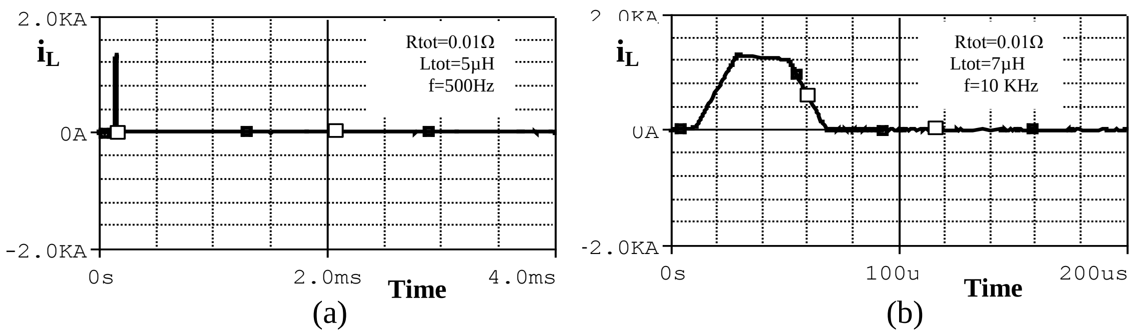


Fig.14. The protection circuit rejection to extremely excessive loads.

while the second test represents an extremely excessive load operated at 10 KHz. The power supply protection circuit rejects both loads. Note that the two tests lay on the frequency band extremes. These tests ascertain the successful performance of the power supply when it is operated with extremely excessive load within its frequency band. It is clear that the protection circuit response is fast enough that the fast varying load current is blocked at a current level slightly above 1200A which is the value of the peak current specified for the supply. In case of short-circuit the blocked current peak may exceed 1200A, and this is not dangerous at all since, the current in this case is a unique pulse as shown in Fig.15.

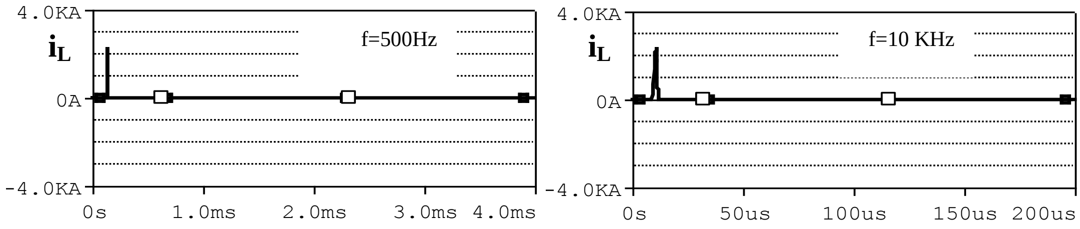


Fig.15. The protection current response to real short circuits.

Fig.16 exhibits six load tests covering the power supply frequency band. (a), (c), and (e) correspond to three normal loads, while (b), (d), and (f) represent three excessive loads which their corresponding currents are limited at peaks of 1200A.

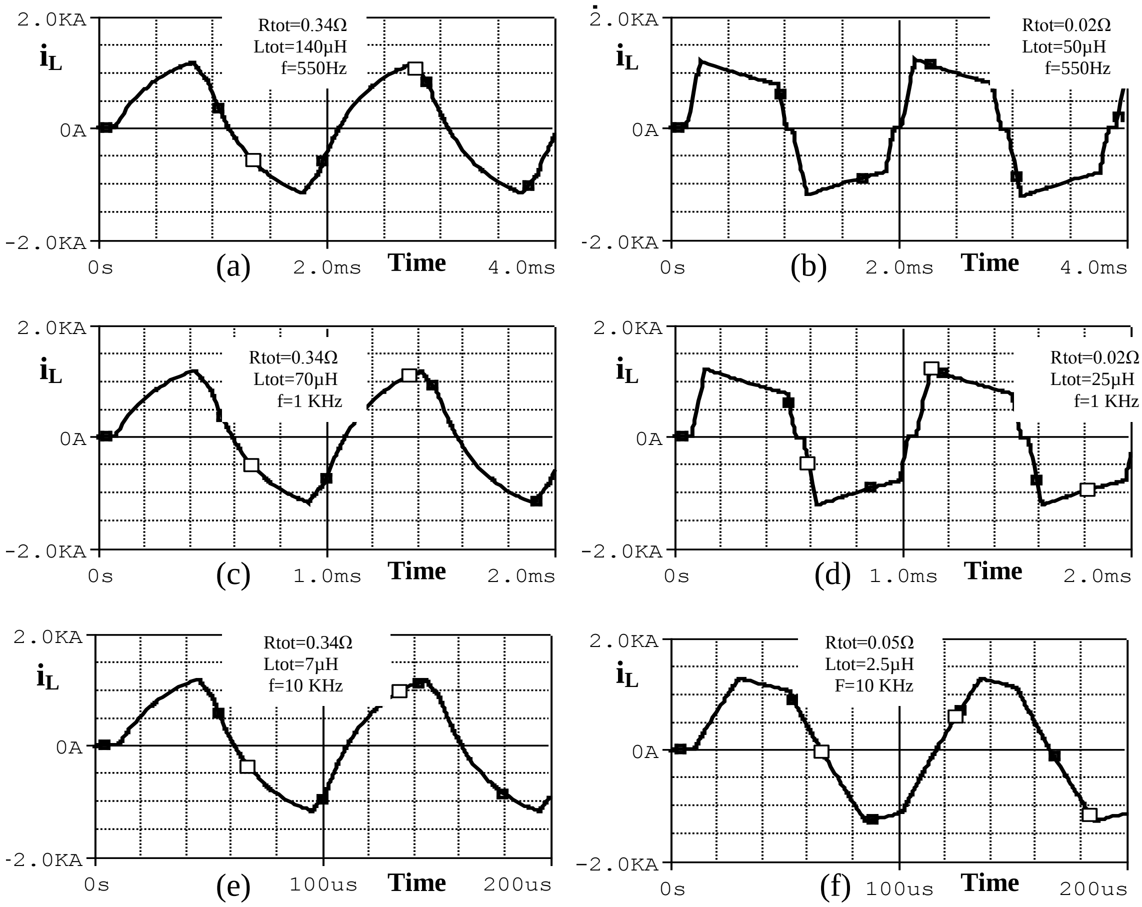


Fig.16. Load currents corresponding to six different load tests within the supply band.

Series capacitors are often used when poor power factor loads are applied. Many tests were made concerning this matter. Fig.17 exhibits two of those tests.

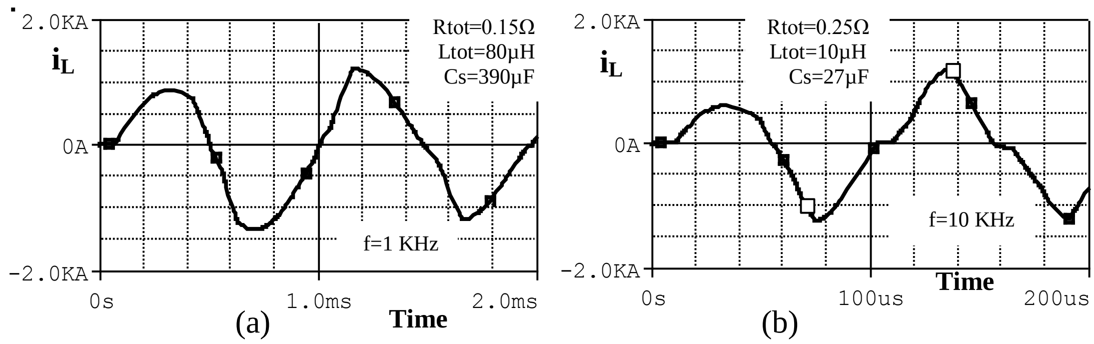


Fig.17. Power supply operation with RLC loads.

It is clear that the supply operates successfully when series capacitors are used. Note that the first cycles of load currents are somewhat distorted because the bootstrapped capacitors are not sufficiently charged. However, after this cycle the power supply operates properly as shown in Fig.17. The mechanism followed by the power supply offers the possibility of operating the supply with extremely excessive loads without using series capacitors. This is because the switches S_2 and S_4 are operating as free-wheel-diodes during the time t_2 discussed in the power supply schematic design. During this time, the energy stored in the inductance L_{tot} is transformed partly or totally to real power in the load resistance R_{tot}. Fig.18.a shows a test applied on a certain excessive load. The test shows that no energy is fedback to the DC source and all the energy stored in the inductance is transformed to heat through the load resistance. In this test, S_2 and S_4 are turned off while they are carrying zero current and hence their turn-off energy losses are negligible. The source current seems to be as a strip of narrow pulses. However, the load current is continuous. Turn-on losses are almost negligible for all switching devices and this is because these devices are turned on while they are carrying zero starting inductive currents. During excessive load operation mode, S_1 and S_3 are turned off at 1200A. The turn-off period for S_1 and S_3 is shown in Fig.18.b.

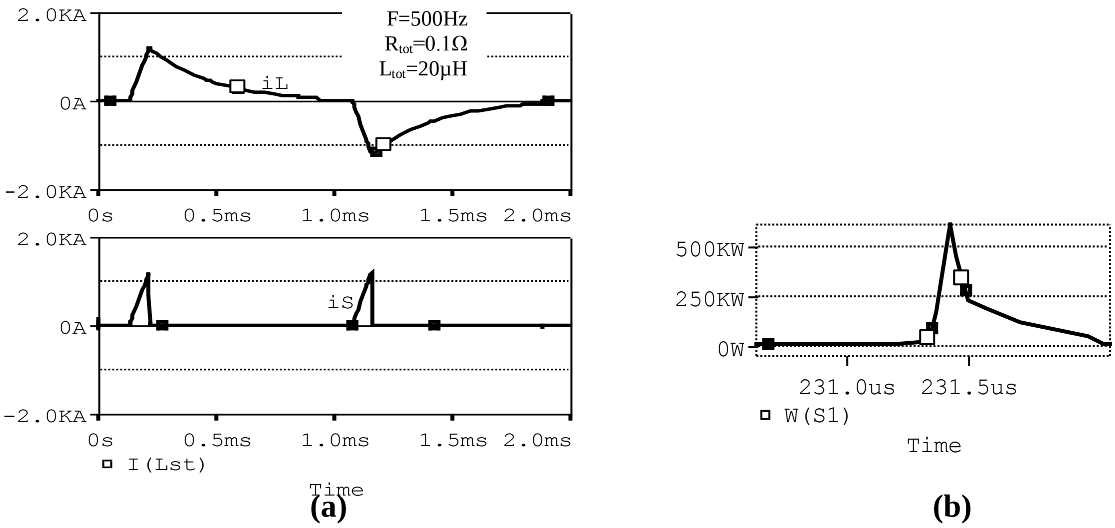


Fig.18. (a) The no energy feedback load test, (b) the turn-off period of S_1 and S_3.

References

Davies J. and Simpson P. (1979) Induction heating handbook. McGraw-Hill.

Hinchliffe S. and Hobson L. (1986) Review of solid-state devices and circuits for HF electric process heating applications. Int'l Journal of Electronics, 61: 261-279.

Jain P. and Diwan S. (1988) Starting problems associated with a transformer-coupled load in a series inverter. IEEE Transaction on Magnetics, 24: 2895-2897.

John V., Suh B., and LipoT. (1999) Fast-clamped short-circuit protection of IGBT's. IEEE Transactions on industry applications, 35: 477-486.

Koetzen J., Van D., and Ferreira J. (1995) Design of the half-bridge series resonant converter for induction heating. IEEE PESC Record, 2: 729-735.

Koller L. and Tevan G. (1999) Electrical dimensioning of inverter-inductor-load system induction heating of Ferromagnetic plates as load. Periodica Polytechnica Ser.El.Eng, 43: 91-100.

Lazor D. (2001) Induction related considerations in investment casting. Modern Investing Technical Seminar March 27-29, 2001-OPUS Technologies, pp 14.

Millman J. and Halkias C. (1983) Integrated electronics: analogue and digital circuits and systems. McGraw-Hill, pp 911.

Rashid M. (2001) Power electronics handbook. Academic Press, pp 892.

Rev D.(2000) Induction heating system topology review. Fairchild Semiconductor AN9012, pp 27.

Shenkman A, Berkovich Y., and Axelrod B. (2004) Dynamic analysis of a new single-switch AC-AC converter for induction heating applications. IEE Proc.-Electr. Power Appl., 151: 398-403.

Skvarenina T. (2001) The power electronics handbook. CRC Press, pp 626.