

DESIGN OF A VIDEO LARYNGEAL STROBOSCOPY SYSTEM FOR STUDYING VOCAL FOLD PATHOLOGY

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ABSTRACT

Video Laryngeal Stroboscopy (VLS) is a stroboscopic imaging technique that allows observations of standstill and slow motion images of vibrating vocal folds. It is a valuable tool for studying vocal fold pathology, for evaluating the degrees of infiltration caused by cancerous lesions, in identifying effects of changes in pitch, tension, or intensity. It is a relatively recent technique and has gained clinical acceptance within the last two decades. The recording of vibration provides a good teaching, learning and documentation material. The resulting image data has a clear and magnified image of vocal folds allowing clinician to compare pre treatment and post treatment status of vocal folds. Besides an endoscope and a video camera, video laryngeal stroboscopic examination requires a high-intensity, pulsed light source that consists of a high-voltage, high-power supply unit, a triggering unit (TU) whose frequency can be synchronized with the fundamental frequency of the vocal folds and a Xenon flash tube. Since human vocal fold vibration fundamental frequency may reach 500Hz (for a baby), the pulsed-light source used for VLS is quite different from that used for conventional stroboscopes.

In this thesis work we designed and realized a power unit and a TU for VLS. The power unit is based on a flyback topology and operates either in QR (free-running) or fixed frequency mode. To maximize efficiency, only low voltage switching is used and at light loads, frequency is reduced. The operating mode can be changed from free-running to fixed frequency mode and the regulation characteristics from primary to secondary regulation by jumpers. The output voltage can easily be adjusted by a potentiometer; thus it can also be used in other applications like laser pumps, defibrillators, etc. The power density of the power unit is about $3\text{W}/\text{inch}^3$, reflecting a state-of-the-art technology. It has been tested at a power level of 250W.

The TU is the unit that extracts the fundamental frequency of the vocal folds for firing the xenon flash tube. Vibrations of the vocal folds are picked up by a laryngeal microphone and amplified, passed through a notch filter to reduce 50Hz pick-up, and then through a 1000Hz low pass filter and a 70Hz high pass filter in series. This filtered signal is the input of a Schmitt trigger circuit. The signal at the output of the Schmitt trigger is then inverted and integrated using two reset integrators having different time constants. The integrator outputs are compared to obtain the triggering signal for firing the Xenon flash tube. The triggering signal can easily be adjusted to obtain either standstill images of the vocal folds or slow-motion images of the vocal folds. A peak detector included within this unit can be used to determine the peak amplitude of the speech. Frequency is displayed using an ADC. The unit is optically isolated for patient safety.

Keywords: Video Laryngeal Stroboscopy, switched mode supply, free running, xenon flash lamp

SES TELLERİ PATOLOJİLERİNİN DEĞERLENDİRİLMESİ İÇİN VIDEO LARİNGOSTROBOSKOPİ SİSTEMİ TASARIMI

ÖZET

Video Laringostroboskopi (VLS) ses tellerinin yavaş çekim ve durağan görüntülerini gözlemlemeye yarayan bir görüntüleme tekniğidir. Ses tellerinin patolojilerinin incelenmesinde ve kanserli dokuların ne derece yayıldığını değerlendirmede ve perde, gerilim ve yoğunluktaki değişimleri saptamada değerli bir araçtır. Nispeten yeni bir tekniktir ve son 20 yılda klinik olarak kabul görmüştür. Titreşimlerin kaydı iyi bir ders, bilgi ve belgelendirme materyali olarak kullanılmasına imkan sağlar. Sonuç olarak elde edilen imge ses tellerinin berrak ve büyütülmüş imgesi olduğundan ses tellerinin tedavi öncesi ve sonrası durumunu değerlendirmeye imkan verir. Endoskop ve video kameranin yanında VLS yüksek-gerilimli ve yüksek güçte bir güç kaynağı ile frekansı ses tellerinin temel titreşim frekansı ile zenon parıldama lambasının frekansını senkronize edebilen bir tetikleme ünitesinden oluşan yüksek-yoğunluklu, darbeli bir ışık kaynağına ihtiyaç duyar. İnsan ses tellerinin temel titreşim frekansı bebeklerde 500Hz'e kadar çıktığından darbeli ışık kaynağı geleneksel stroboskoplardan farklıdır.

Bu tez çalışmasında VLS için bir güç birimi ve tetikleme ünitesi tasarımı yapıp gerçekleştirdik. Güç ünitesi geri-uçuşlu (flyback) ve değişken frekanslı veya sabit frekansta çalışabilme özelliğine sahiptir. Anahtarlama akıttıcı gerilimi düşük düzeyde olduğunda yapıldığından ve düşük güçlerde anahtarlama frekansı düşürüldüğünden verim artmaktadır. Çalışma biçimi değişken frekanslıdan sabit frekanslıya ve düzenleme tipide birincil düzenlemeden ikincil düzenlemeye bir atlama vasıtasıyla çevrilebilmektedir. Güç kaynağının güç yoğunluğu 3W/inch³ civarında olup günümüz teknolojisini yansıtmaktadır. Güç kaynağı 250W düzeyinde test edilmiştir.

Zenon parıldama lambasının tetikleme sinyali ses tellerinin temel titreşim frekansından elde edilir. Ses tellerinin titreşimleri bir gırtlak mikrofonu vasıtasıyla alınmakta, kuvvetlendirilmekte, 50Hz çentik, 1000Hz alçak geçiren ve 70Hz yüksek geçiren filtreden geçirilmektedir. Bu işaret Schmitt tetikleme devresinden geçmekte ve terslendirildikten sonra farklı zaman sabitlerine sahip 2 farklı entegral alıcı devre vasıtasıyla entegrali alınmaktadır. Entegral alıcı devre çıkışları karşılaştırılmakta ve zenon parıldama lambası için tetikleme darbesi elde edilmektedir. Tetikleme işareti ses tellerinin durağan veya hareketli görüntülerini elde edebilmek için kolayca ayarlanabilmektedir. Üniteye dahil edilen bir tepe dedektörü işaretin tepe değerini algılamakta kullanılabılır. Frekans analog sayısal dönüştürücü vasıtasıyla görüntülenmektedir. Ünite hasta güvenliği için optik olarak yalıtılmıştır.

Anahtar Sözcükler: Video laringostroboskopi, anahtarlama güç kaynağı, değişken frekanslı, zenon parıldama lambası

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LIST OF ABBREVIATIONS

AC	Alternating current
ADC	Analog to digital converter
CCD	Charge coupled device
CCM	Continuous current (conduction) mode
CM	Common Mode
CMRR	Common Mode Rejection Ratio
CTR	Current Transfer Ratio
DC	Direct current
DCM	Discontinuous current (conduction) mode
DM	Differential Mode
EMI	Electromagnetic interference
ESR	Effective series resistance
HV	High voltage
HVPS	High voltage power supply
MOSFET	Metal oxide field effect transistor
MOV	Metal oxide varistor
NTC	Negative temperature coefficient
OTC	Off time circuit
PFC	Power factor correction
QR	Quasi-resonant
QRC	Quasi-resonant converter
RC	Resonant converter
RHP	Right half plane
RMS	Root mean square
SCR	Silicon controlled rectifier
SMPS	Switch mode power supply
TU	Triggering unit
UV	Ultra violet
VLS	Video Laryngeal Stroboscopy
ZCS	Zero current switching
ZVS	Zero voltage switching

LIST OF SYMBOLS

$P_{o\max}$	Maximum output power
$P_{in\max}$	Maximum input power
I_{inrms}	RMS input current
$V_{p\min}$	Minimum voltage on the primary electrolytic capacitor
$V_{p\min\ peak}$	Minimum peak voltage across the primary electrolytic capacitor
$V_{p\max}$	Maximum voltage across the primary electrolytic capacitor
C_{in}	Primary electrolytic capacitor
T_d	Discharging time at each half line cycle
W_{in}	Required energy at discharging time
$V_{in\max}$	Maximum AC input voltage
$V_{in\min}$	Minimum AC input voltage
V_{ir}	Input ripple voltage
η	Efficiency
V_r	Reflected output voltage
F_{line}	Line frequency
$D_{\max}$	Maximum duty cycle
V_{aux}	Auxiliary voltage
T_a	Maximum ambient temperature
V_{fo}	Main output diode voltage drop
V_{or}	Main output voltage ripple
V_o	Main output voltage
P_o	Main output maximum power
V_{faux}	Auxiliary output diode voltage drop
V_{raux}	Auxiliary output voltage ripple
V_{aux}	Auxiliary output voltage
P_{aux}	Auxiliary output power
F_s	Switching frequency
μ_0	Magnetic constant
T_o	Operating temperature
V_e	Effective magnetic volume of ferrite core
l_e	Effective magnetic path length
A_e	Effective magnetic cross-sectional area
$A_{\min}$	Minimum magnetic cross-sectional area
A_n	Usable cross-sectional area of winding

l_n	Average length of turn
B	Flux density
$B_{\max}$	Maximum allowable flux density in the transformer
$V_{d\max}$	Maximum allowed drain voltage of the power MOSFET
F_{os}	Factor for the overshoots at the primary winding
A_l	Inductance factor
s	Air gap length
$I_{inave\max}$	Maximum average current flowing through the primary winding
$I_{p\max}$	Peak current flowing through the primary inductance
I_{prms}	RMS current flowing through the primary inductance
N_p	Number of primary turns
N_s	Number of secondary turns
N_{aux}	Number of auxiliary turns
V_{o2}	Secondary output 2 voltage
V_{fo2}	Secondary output 2 diode voltage drop
N_{s2}	Number of secondary output 2 turns
L_p	Primary winding inductance
L_s	Secondary winding inductance
L_{s2}	Secondary 2 winding inductance
L_{aux}	Auxiliary winding inductance
A_p	Primary winding cross-sectional area
A_s	Secondary winding cross-sectional area
A_{aux}	Auxiliary winding cross-sectional area
A_{s2}	Secondary winding 2 cross-sectional area
ϕ_p	Primary winding wire diameter
ϕ_s	Secondary winding wire diameter
ϕ_{aux}	Auxiliary winding wire diameter
ϕ_{s2}	Secondary winding 2 wire diameter
Mar	Margin on each side of the winding area
f_{Cu}	Copper space factor
I_{speak}	Peak current flowing through the secondary inductance
I_{srms}	RMS current flowing through the secondary inductance
L_σ	Leakage inductance of the primary winding
$I_{o\max}$	Maximum average load current
T_{off}	Total off time of the power switch
T_{on}	On time of the power switch
J_p	Primary winding current density
J_s	Secondary winding current density

V_{snub}	Snubbing voltage
V_{dsnub}	Maximum inverse voltage stress on the snubber diode
C_{oss}	MOSFET output capacitance
C_{stray}	Stray capacitance of the transformer
C_{res}	Additional resonance capacitance placed across the power switch
C_r	Total resonance capacitance across the power switch
t_{rr}	Reverse recovery time of the rectifier
V_{rdiode}	Maximum reverse voltage across the output rectifier diode
I_{inave}	Average converter input current
k_{fd}	Diode current form factor
k_{fi}	Converter input current form factor
T_a	Ambient temperature
T_j	Maximum operating temperature of the MOSFET
$R_{\theta jc}$	MOSFET's junction to case thermal resistance
$R_{\theta cs}$	MOSFET's case to sink thermal resistance
R_{sa}	Heat sink thermal resistance
R_{pin2}	Resistor at pin 2 of the TDA16846 SMPS Control IC
C_{pin2}	Capacitor at pin 2 of the TDA16846 SMPS Control IC
P_{bridge}	Input diode bridge loss
$P_{copperp}$	Primary winding copper loss
$P_{coppers}$	Secondary winding copper loss
P_{core}	Transformer core loss without an air gap
P'_{core}	Transformer core loss with an air gap
P_{orec}	Output rectifier loss
P_{son}	MOSFET turn-on loss
P_{soff}	MOSFET turn-off loss
P_{scond}	MOSFET conduction loss
P_{NTC}	NTC loss
P_{bleed}	Bleeder loss
P_{Mosfet}	Total MOSFET loss including turn-on, turn-off and conduction losses
$P_{losstot}$	Total loss in the converter circuit
$R_{DS(on)}$	On-state resistance of power MOSFET
V_{air}	Air throughput of fan at a given heat loss
μ_e	Effective permeability with an air gap
μ_i	Initial permeability without air gap (measured on a ring core)
G_{vc}	Control to output transfer function
G_{vg}	Line to output transfer function

T_{ch}	Charging time of the integrating capacitor
T_d	Discharging time of the integrating capacitor
V_{TH}	Threshold voltage of the integrator circuit comparator
BW	Effective bobbin width
BWe	Effective winding cross-sectional area

1. INTRODUCTION

1.1 Background and Motivation

Stroboscope is an optical instrument for making a fast moving object appear to be slowed down or stationary. This effect is called *stroboscopic effect* and the light used to obtain this effect is called the stroboscopic light. *Stroboscope* word comes from Greek. *Strobo* means whirling and *Skopeo* means “I look at”.

There are two types of stroboscopes. *Mechanical shutter stroboscopes*, which are pulse modulators of light and *electrical flash stroboscopes*, which are sources of pulsed light.

Stroboscopes are used in a variety of areas. They are used in printing and paper industry, textile industry, automobile and engine industries, optical industry, chemical industry, electrical industry, film production, biological sciences, ballistics, sporting events, aerial and underwater photography, machine vision and medical applications.

Stroboscopic effect is created by interrupting the observer's view so that the object is seen only at regularly spaced intervals rather than continuously. In its simplest form the stroboscope is a rotating disk; along its edge are evenly spaced holes through which the moving object is observed. If the object's motion is cyclic, the speed of the disk can be synchronized with it so that the object always appears in the same position when viewed through one of the holes. During the time that a solid area is blocking the line of sight, the persistence of vision enables the eye to retain the image previously seen, while the object moves to the same or a similar position by the time the next hole is in front of the eye. The effect is thus one of a stationary object. If the stroboscope is not quite synchronized with the object's motion, the object will appear to move slowly either backward or forward, depending upon whether the stroboscope's rotation is too fast or too slow.

1.2 Pulse Image Perception

Electromagnetic spectrum and its visible portion can be seen in Fig.1.1. Human eye reacts to radiant energy within the visible spectrum, corresponding to wavelengths of approximately $\lambda = 400\text{-}700\text{nm}$. The spectral distribution of sensitivity of eye is seen in Fig.1.2. Here 1, 2 and 3 represents the eye's central field of vision at daytime, night-time and in poor light, respectively. During daytime spectral distribution of the eyes sensitivity attains a maximum at $\lambda = 555\text{nm}$.

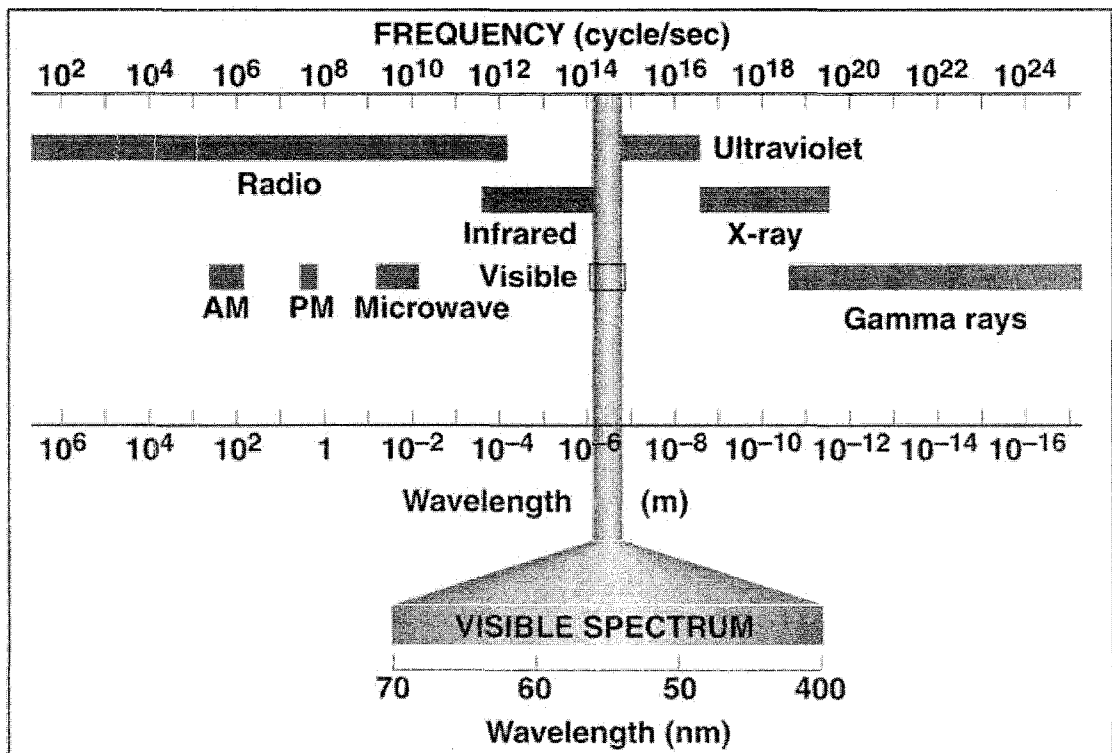


Figure 1.1. Electromagnetic spectrum and visible spectrum.

If the frequency of the vibrating object is higher than approximately 5Hz eye cannot differentiate successive images and the image sequence is perceived as *flickering image*. But if the frequency is lower than 5Hz the human eye can differentiate successive images and the image sequence is perceived as *persistent image*. The reaction of eye to flickering and persistent image is shown in Fig. 1.3.

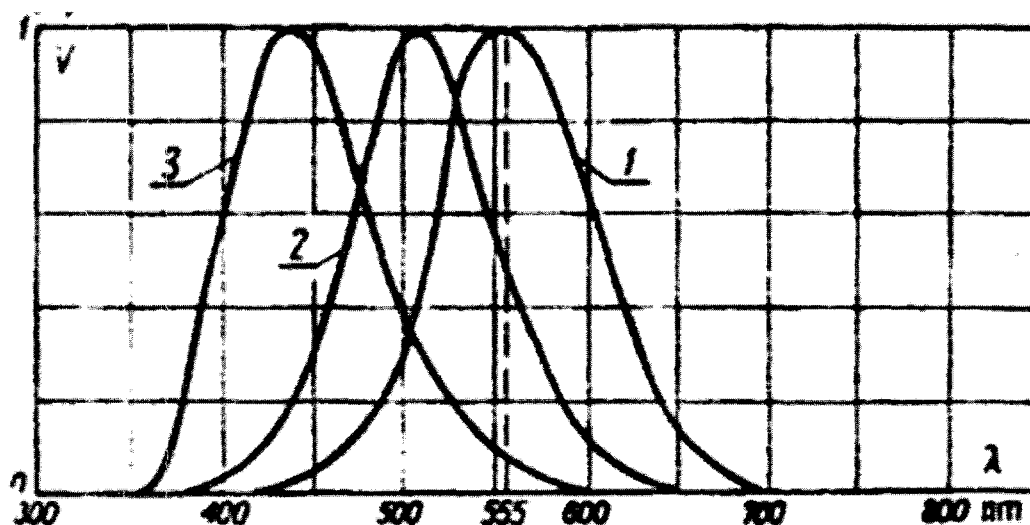


Figure 1.2. Spectral distribution of sensitivity of eye.

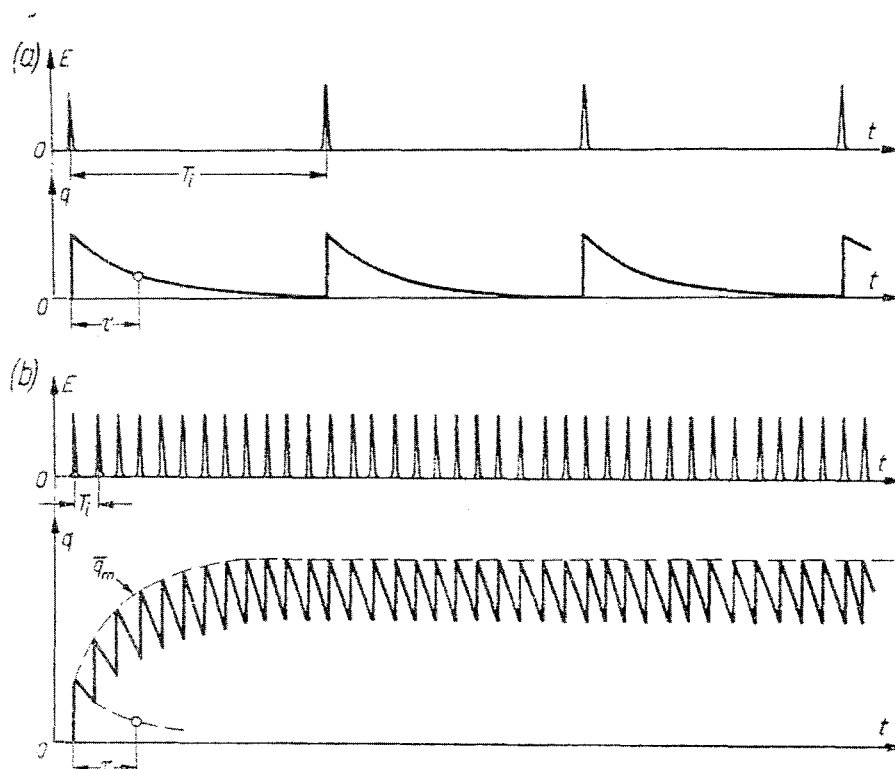


Figure 1.3. Reaction of eye to flickering (top) and persistent image (bottom).

Requirements for a pulsed light source used in a stroboscopy system are as below:

- For saving electrical power the spectral distribution of the light source power should correspond to the spectral distribution of the sensitivity of the eye or the recording device.

- The light source should be capable of providing appropriate light output, light flux or luminous intensity (optical fitting) independent of the medium conditions.
- The pulse duration should be very short to obtain less blurred sharp images.
- Because the flash lamp does not operate in continuous state it is subjected to short but large loads with alternating relatively long periods of rest. Therefore the durability of the light source should be as long as possible.

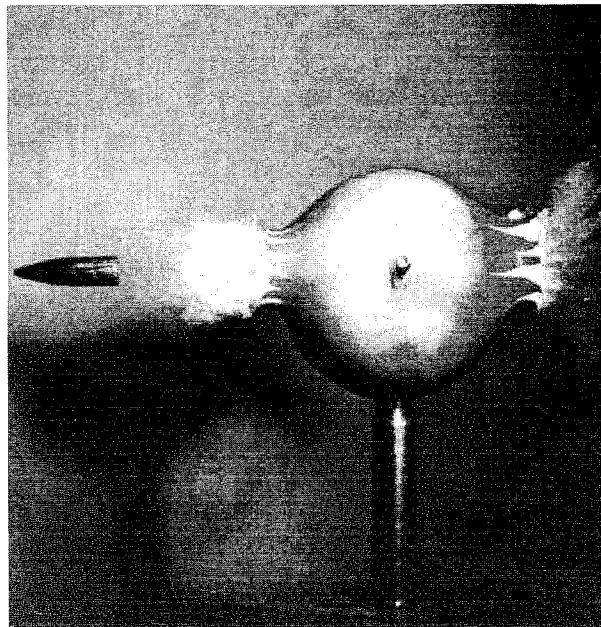


Figure 1.4. A bullet passing through an apple taken using single flash photography.

In Fig.1.5 spectral distribution of a xenon flash lamp is given. Xenon flash lamps are valuable because they produce in excess of 250,000 candelas, a measure of optical intensity. This very high compared to other illumination systems. Increase in the optical intensity allows for superior camera performance, especially in terms of increased depth of field and shorter integration times. Xenon flash lamps have very short pulse durations which make them very useful in freezing motion. This allows elimination of stroboscopic blur. In some applications the pulse durations are in the order of micro seconds. Xenon flash lamps also have a continuous and relatively flat (compared to tungsten-halogen and led based systems) spectrum at visible spectrum as seen from Fig.1.5. But again they produce radiation at different wavelengths and this contains UV range. Special filters can be used to remove these wavelengths, which cause damage to human eye. The color

temperature of pulsed xenon lightning is between 5500K and 12000K depending on operating conditions. Filtering can be used to filter out the unwanted wavelengths. Visible light filters can be used to obtain cold light sources. Pulsed xenon flash lamps also reduce power required, because the camera records only when the lamp flashes.

Taking all these into account stroboscopic xenon systems have short pulse widths and wide spectral distribution quality lighting. These properties improves camera image, increase the camera's depth of view, shorten camera integration times, eliminate the need for a camera shutter, improve color rendition, reduce power consumption and allow for high speed inspection.

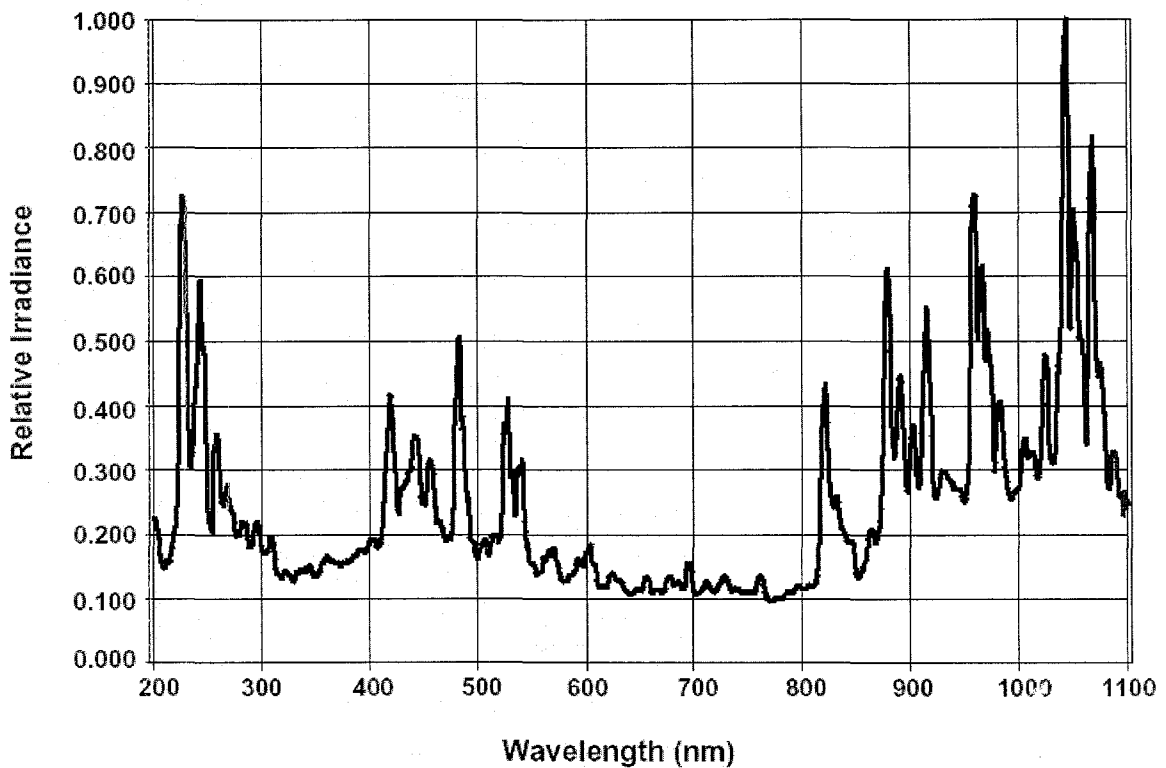


Figure 1.5. Typical xenon spectral output illustrates wide spectral distribution (continuous and relatively flat in the visible spectrum).

This technique has recently been used for observing shape, movement, vibratory pattern, and timing relationships between opening and closure of the glottis and to study vibration characteristics of vocal folds and thus to obtain important diagnostic information on vocal pathology [1].

Stroboscopic studies of the vocal folds have been demonstrated to be very valuable for studying vocal fold pathology and for evaluating the degrees of infiltration caused by

cancerous lesions [1], [2]. If (video) images of the previous vocal fold states are recorded and compared with new stroboscopic images, discrimination of subtle changes not visible to the unaided eye is possible. The resulting technique, called “VLS” is a practical imaging technique and allows observations of standstill and slow motion images of vibrating vocal folds.

VLS is also used in identifying the altering effects of changes in pitch, tension, or intensity. The recording of vibration provides a good teaching, learning and documentation material. The resulting image data has a clear and magnified image of vocal folds allowing clinician to compare pre treatment and post treatment status of vocal folds. The patient can watch the image data to have an idea about his/her illness.

Video laryngeal stroboscopy system requires the following units:

- Pulsed light source (stroboscope)
- Triggering Unit (TU)
- Endoscope
- Video recorder
- CCD Camera
- Microphone

The examination starts with the placement of a pre laryngeal microphone. By using filters and wave shaping mechanisms, the frequency of pulsed light source is synchronized with the vibration frequency of vocal folds. The frequency and phase adjustment option provides either standstill or dynamic images of vocal folds. Standstill images of vocal folds allow clinician to investigate vocal fold structure, and to determine if there exists an aperiodic vibration pattern. Dynamic images allow clinician to investigate the functional properties of vocal folds.

1.3 Objectives

The objective of this thesis work is to design and realization of a VLS system for the examination of vocal fold pathology. Average fundamental frequency of vocal folds for a male, a female and a baby are 140Hz, 230Hz and 500Hz, respectively [3]. Special faster (and therefore more expensive) equipment is required here.

The pulsed light source and the TU are the critical units in a VLS system. The other units such as the endoscope and video camera can be procured from a number of vendors but not these critical units. Therefore the scope of this thesis work was limited to the development of these units. An overview of the system developed is presented in the following subsection.

1.4 Overview of the System Developed

A simplified block diagram of the system is presented in Fig. 1.6. It consists of a high-voltage, high-power DC power supply (HVPS) unit, a xenon flash tube (XFT), an electret microphone, a TU and a HV semiconductor switch. The HVPS is a high-voltage, high-power SMPS. It operates in QR and in free-running to minimize switching losses. The TU is the signal shaping circuit which uses audio signal from a laryngeal microphone to extract fundamental frequency of the vocal folds to trigger the xenon flash tube.

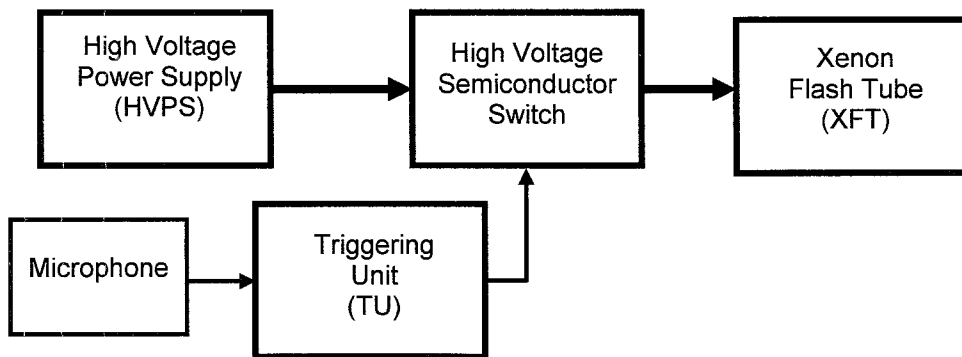


Figure 1.6. Simplified block diagram of the system.

1.5 Outline of the Thesis

In this chapter we discussed principles of stroboscopy, its application areas and perception of pulsed light by the human eye. The remaining chapters are organized as follows. Chapter 2 introduces VLS, its clinical parameters and value and videolaryngostroboscopic evaluation. Power supply design, component selections, efficiency calculations, thermal design and PCB design are described in Chapter 3. Converter operation and dynamics and derivation of transfer functions are discussed in

Chapter 4. TU consisting of microphone preamplifier and isolation circuit, various filters used, instrumentation amplifier, integrators and isolated SCR trigger signal conditioning circuit is presented in Chapter 5. Chapter 6 is a summary of the basic conclusions and recommendations for future work. Appendix A includes PCB documents of the designed units. Appendix B, C and D includes data extracted from semiconductor and device datasheets. Photographs of the units are in Appendix E. Appendix F contains some drain voltage and current waveforms taken during experiments.

2. VIDEO LARYNGEAL STROBOSCOPY FOR STUDYING VOCAL PATHOLOGY

2.1 Introduction

Vibration characteristics of vocal folds provide important diagnostic data when evaluating vocal fold pathologies. Normally vocal folds move higher than 70Hz and because of this human can not observe this motion without the help of a device. The fundamental frequency of each patient depends on his/her physical structure. In the case of a crying baby this frequency is about 500Hz. For a mature man and a woman has a fundamental frequency about 140 and 230Hz respectively [3].

VLS lies on the basis of sensation of pulsed light. Normally the image stays on the retina about 0.2 seconds. Because of this if an image changes at a frequency higher than 5Hz this sequence is sensed as a continuous image.

With VLS either the vocal folds or the surrounding structures can be recorded as standstill and moving images which allow determination of morphological and functional disorders.

2.2 Videolaryngostroboscopic Evaluation

A microphone is placed on the larynx and this signal is shaped and a trigger pulse is produced. This pulse activates the light source and the vocal folds are recorded with a camera. Adjustable parameters are frequency of trigger and phase of trigger. Frequency of trigger provides standstill or moving images. Phase of trigger determines which phase of the signal produces a trigger pulse.

2.2.1 Synchronization

Frequency of pulsed light and input signal is in the same frequency, so the triggering is done at the same phase at each period. As a result a standstill image of vocal folds is

observed. Phase adjust at this position determines which portion of the vibration period will be illuminated.

This allows clinician if there is structural disorders and if there is aperiodicity. Standstill image is used in evaluation of edges and shape of vocal folds and to find out if the laryngeal closure is ok.

2.2.2 Asynchronization

If we change the frequency of the pulsed light the resulting image sequence seems to move in a slow motion. Each light pulse illuminate different phase of vibration period. If a video camera is used to record this period the resulting image consist of different phases taken from different pulses. So we can say that the resulting image does not the real but an illusionary image. But this image sequence is really helpful in determining vocal fold pathologies.

2.3 Importance of VLS

Video laryngeal stroboscope is the necessary equipment in revealing vocal fold pathologies [1]. Observation of standstill and moving images of vocal folds provides evaluation of amplitude, glottic closure, size and spread of nodules and polyps [4]. It is useful in discrimination of structural and functional causes [1], [4]. In the case of vocal fold paralysis it is used to observe the progress in the beginning so determining the required medical treatment method [4]. It is also helpful in early detection of cancer [2].

2.4 Stroboscopic Parameters [1], [2], [3], [5], [6], [7]

Schematic illustration of a normal cycle of the vocal fold vibration is shown in Fig 2.2 and a typical image obtained from a VLS is shown in Fig 2.3. The fundamental stroboscopic parameters are:

Periodicity: It is the regularity of successive apparent cycles of vocal fold vibration. It may be regular, irregular or inconsistent. It is determined by changing from slow motion to standstill during a sustained 'i'. If it is regular then the static image persists, if irregular there will be observable movement of the vocal fold at the

standstill state. The periodicity may also vary and may be inconsistent. Aperiodic vibrations may be caused by marked, asymmetry in the mechanical properties of the vocal folds, interference with the homogeneity of the folds, incapability of maintaining a steady tone in the laryngeal muscles or of blowing out air from the lungs with a consistent force.

Amplitude: It is the maximum excursion from equilibrium in an oscillatory movement or waveform. It may be great, normal, small or zero. Amplitudes of vibration become restricted if the vibratory portion of the vocal folds is shortened or the stiffness or mass increases. The existence of an obstacle results in a decrease of amplitude, and an increase in subglottic pressure increases the amplitude.

Symmetry of vibration (phase): It is the degree to which the vocal folds provide mirror images of each other during vibration. Differences in the mechanical properties of the two vocal folds result in asymmetrical movements which may be due to the position, shape, mass, tension elasticity or viscosity of the vocal fold.

Glottic closure: It can be complete, incomplete or inconsistent. There are seven levels of glottic closure. It is related to the action of the adductor, abductor and tensor muscles of the larynx. Incomplete glottic closure may result from impaired adduction of the vocal folds, a nonlinear edge of the vocal fold, an obstacle intervening between the two vocal folds or a stiff vocal fold edge.

Mucosal wave: It can be great, normal, small or absent. It results from the movement of covering mucosa relative to the underlying muscle. The larger the amplitude, the more intensified is the mucosal wave and vice versa. The stiffer the mucosa, the less marked is the mucosal wave. When the mucosa is only partially stiff, the wave stops traveling at the stiff portion. With a missing muscle tonus such as in recurrent nerve palsy, the vocal folds vibrate as homogenous masses but no mucosal wave is visible.

Stiffness / mobility: It is the immobility along any part of the vocal cord observed during phonation. It is present or absent. It is an indication of the stiffness and immobility of the vocal fold. This may result from scar tissue or infiltration of the vocal fold by neoplasm.

Fundamental Frequency: It is the lowest frequency in a periodic waveform; also called the first harmonic frequency. It is 140Hz and 230Hz for men and women, respectively. It is greater if the vocalis muscle is tenser, stiffer and shorter and increases with a rise in subglottic pressure. On the other hand, if the mass of the vocal fold is greater, the fundamental frequency will decrease.

In VLS aperiodic vibrations cannot be observed well. This is the main limitation of VLS.

In Fig. 2.1 different optical methods are compared [8]. Illustration of a normal cycle of vocal fold vibration is given in Fig. 2.2. In Fig. 2.3 actual cycle of vocal fold vibration images taken using a stroboscope is presented [8].

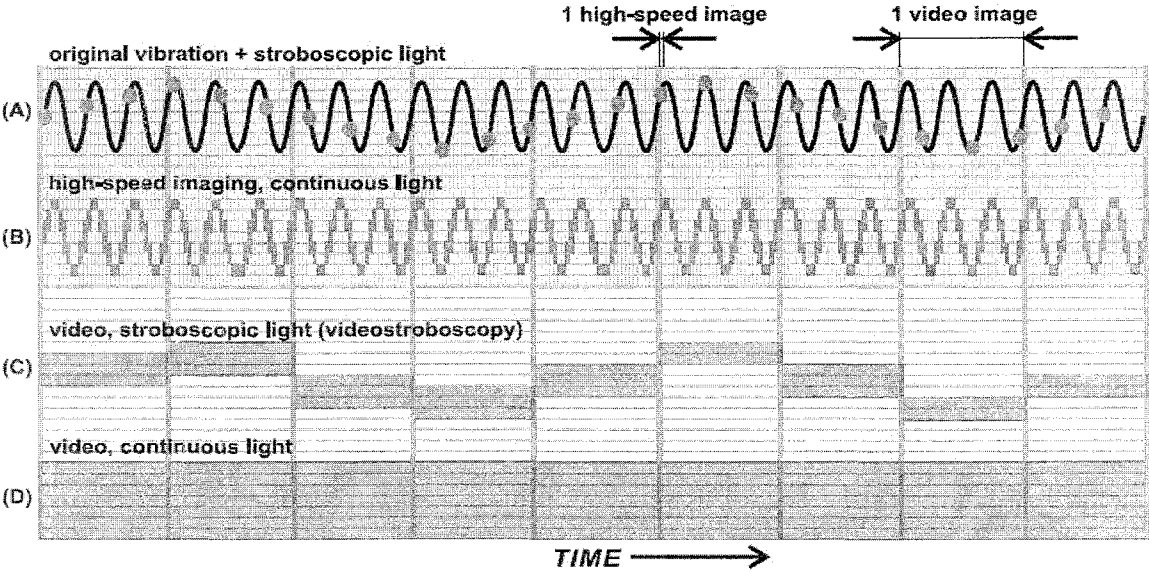


Figure 2.1. Illustration of differences among optical methods [8].

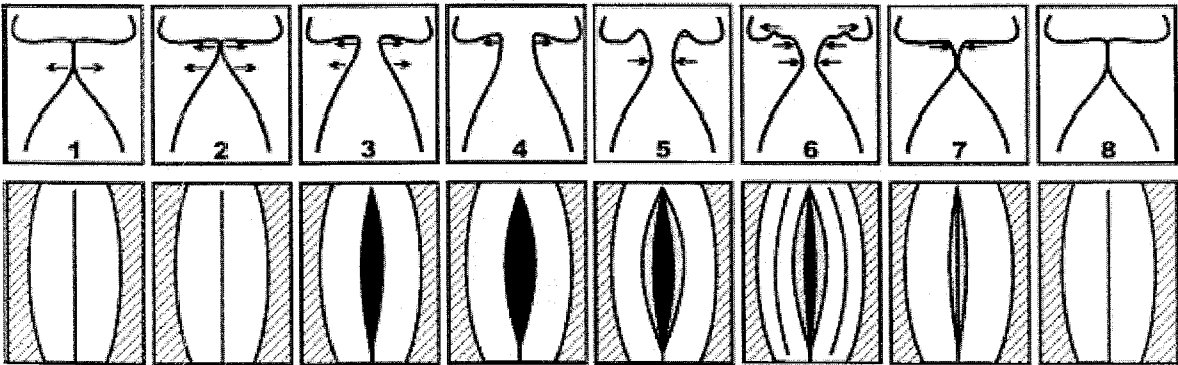


Figure 2.2. Schematic illustration of a normal cycle of the vocal fold vibration [8].

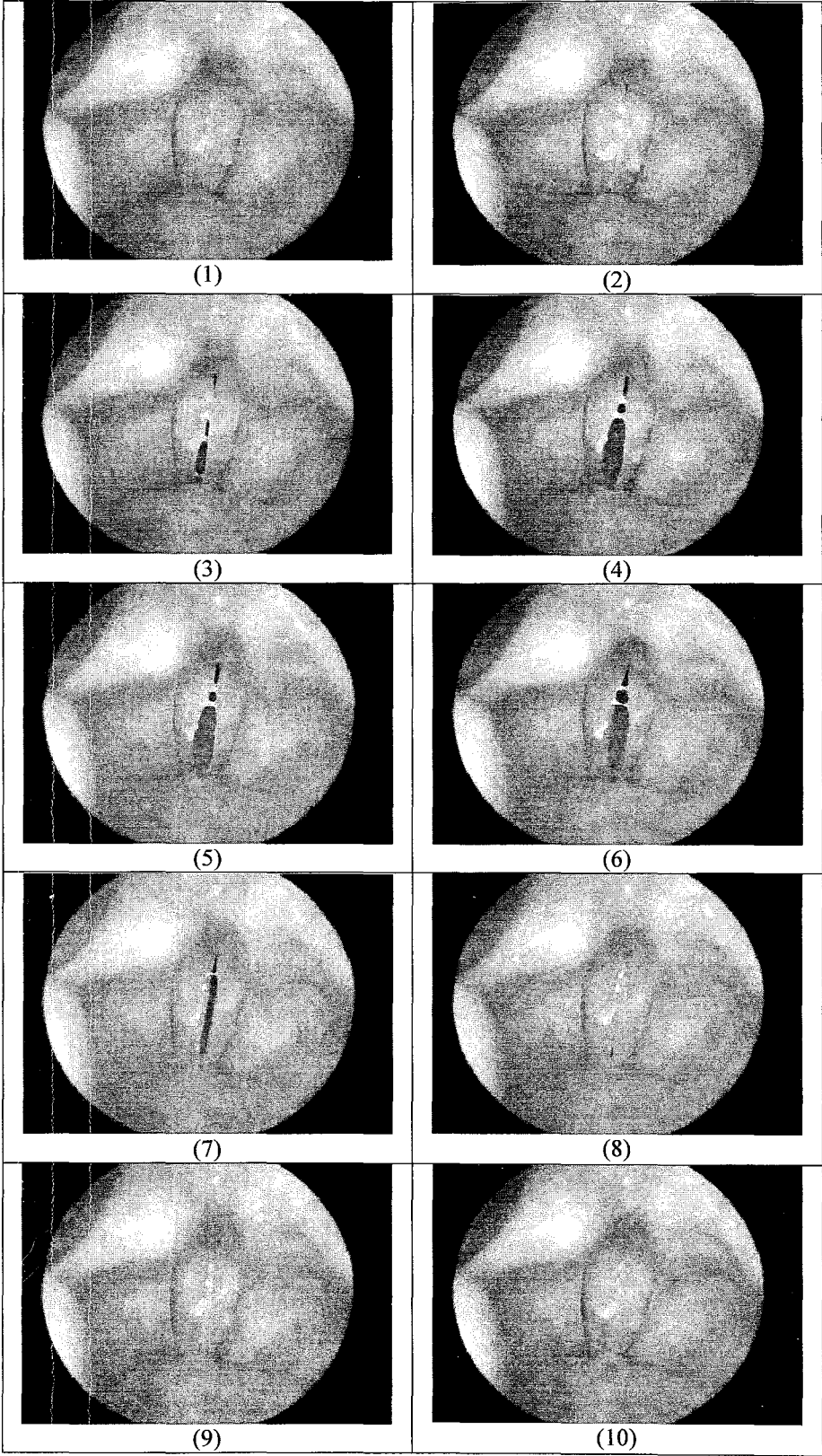


Figure 2.3. One cycle of vocal fold vibration obtained from a video laryngeal stroboscope.

3. HIGH VOLTAGE, HIGH POWER, QUASI-RESONANT, FREE-RUNNING POWER SUPPLY

3.1 Introduction

Flyback converters are widely used for high output voltages at relatively low output powers. If the DC supply voltage is high enough ($\geq 160V$) so that primary current is not very high, it can be used up to 150W. In flyback converters there is no need of an output inductor and no HV free-wheeling diode. It is suitable for multi-outputs, because output voltages track one another with line and load changes far better than forward converters. The better tracing is due to the absence of output inductors. It can be used from DC input voltages as low as 5V and by applying suitable turn ratio high voltages can be obtained [9]. The design of a flyback converter for a high power output, in practice is limited by the leakage (stray) inductance of the transformer and the size of the output capacitor [10]. Therefore to achieve a high converter output power, the transformer must be designed carefully to obtain a low leakage inductance. Also a high frequency low ESR and high value output capacitor must be used. To design the transformer the input parameters must be defined. By using these parameters we can design for the worst-case conditions.

We designed the transformer for a hard switched flyback converter. In QR free running flyback converters device stresses are much lower than the hard switched one so the design equations for the normal flyback converter still holds for the QR free-running operation. The input parameters are summarized in Table 3.1.

Table 3.1.
Power Supply Input Parameters

Parameter	Description	Value	Unit
V_{imax}	Maximum AC input voltage	253	V
V_{imin}	Minimum AC input voltage	187	V
V_{ir}	Input ripple voltage	14	V
η	Efficiency (assumed)	0.85	%
V_r	Reflected output voltage	231	V
T_o	Operating temperature	-	°C
F_{line}	Line frequency	50	Hz
D_{max}	Maximum duty cycle	0.5	-
V_{aux}	Auxiliary voltage	15	V
	SMPS Control IC	<i>TDA16846</i>	
V_{fo}	Main output diode voltage drop	4,5	V
V_{or}	Main output voltage ripple	10	V
V_o	Main output voltage	1000	V
P_o	Main output maximum power	290	W
V_{faux}	Auxiliary output diode voltage drop	0.7	V
V_{raux}	Auxiliary output voltage ripple	0.15	V
V_{aux}	Auxiliary output voltage	15	V
P_{aux}	Auxiliary output power	5	W
V_{o2}	Second output voltage	15	V
V_{fo2}	Second output diode voltage drop	0.7	V
P_{o2}	Second output power	5	W
F_s	Switching frequency	50	kHz
μ_0	Magnetic constant	1.257×10^{-8}	H/cm
T_a	Ambient temperature	50	°C
DCM	Converter operation mode (DCM)	-	-

3.2 Calculations for the Power Supply Components

Maximum output power: The maximum total output power is the sum of output powers consumed by the main and the auxiliary output. Thus,

$$P_{o\max} = P_{aux} + P_o + P_{o2} = 5W + 290W + 5W = 300W \quad (3.1)$$

The maximum input power is given by

$$P_{in\max} = \frac{P_{o\max}}{\eta} (W) \quad (3.2)$$

Here η represents the efficiency. The maximum input power when $\eta = 85\%$ is:

$$P_{in\max} = \frac{300W}{0.85} = 355W$$

The maximum DC voltage on primary electrolytic capacitor is given by

$$V_{p\max} = V_{in\max} \times \sqrt{2} = V_{p\max} = 253V \times \sqrt{2} = 357.8 \text{ V} \quad (3.3)$$

The minimum peak voltage on primary electrolytic capacitor is

$$V_{p\min\text{ peak}} = V_{in\min} \times \sqrt{2} = V_{p\min\text{ peak}} = 187V \times \sqrt{2} = 264.5 \text{ V} \quad (3.4)$$

The minimum voltage on primary electrolytic capacitor is

$$V_{p\min} = V_{p\min\text{ peak}} - V_{ir} = V_{p\min} = 264.5V - 14V = 250.5 \text{ V} \quad (3.5)$$

Capacitor discharge time at each half line cycle is given by

$$T_d = 5 \times 10^{-3} \left(1 + \frac{\arcsin\left(\frac{V_{p\min}}{V_{p\min\text{ peak}}}\right)}{90} \right) = 5 \times 10^{-3} \left(1 + \frac{\arcsin\left(\frac{250.5V}{264.5V}\right)}{90} \right) = 8.96 \times 10^{-3} s \quad (3.6)$$

Energy required during discharge is

$$W_{in} = P_{in\max} \times T_d = 355W \times 8.96 \times 10^{-3} \text{ sec} = 3.179 \text{ Ws} \quad (3.7)$$

Input capacitor value required to achieve these specifications is

$$C_{in} = \frac{2 \times W_{in}}{V_{p\min\text{ peak}}^2 - V_{p\min}^2} = \frac{2 \times 3.179Ws}{264.5V^2 - 250.5V^2} = 882 \times 10^{-6} F \quad (3.8)$$

According to [10] there is an alternative way of choosing an input capacitor based on $\mu\text{F}/\text{W}$. The lookup table provides for the selection of input capacitor using the desired input voltage ripple. Here V_F is the bridge rectifier voltage drop.

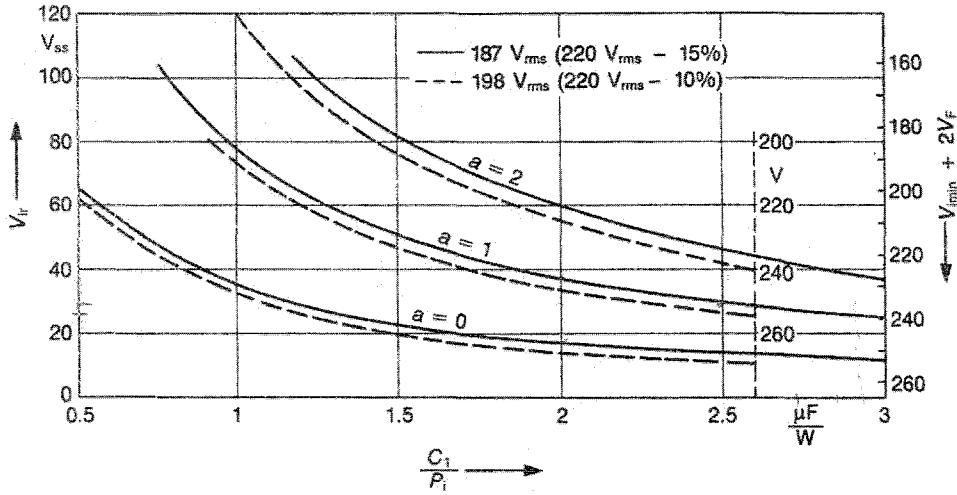


Figure 3.1. Variation in ripple voltage V_{ir} (V_{pp}) and direct input voltage $V_{inmin} + 2V_F$ (V) with reservoir capacitance C_{in} relative to input power P_{in} , with the number of missing mains half cycles (a) as parameter. Solid lines are for -15% mains voltage ($187V_{rms}$) and dashed lines for -10% mains voltage ($198V_{rms}$) [10].

Post calculation:

Two 450V, 470 μF , 85 °C electrolytic capacitors from Nippon-Chemi-Con were used in the present design in parallel. The total of capacitance is 940 μF is obtained which gives a V_{pmin} of 251.3V.

$$V_{pmin} = \sqrt{\left(V_{pmin\ peak}^2 - \frac{2 \times W_{in}}{C_{in}}\right)} = \sqrt{(264.5V)^2 - \frac{2 \times 3.179Ws}{940 \times 10^{-6}F}} = 251.3\text{ V} \quad (3.9)$$

3.3 Transformer Design

Peak current flowing through the primary winding inductance of the transformer:

Since the converter operates in a discontinuous conduction mode the current $I_p(0)$ is zero. Thus the peak current flowing through the primary inductance occurs when the input voltage is at minimum and the duty cycle is at maximum. Its value is:

$$I_{peak} = \frac{2 \times P_{inmax}}{V_{pmin} \times D_{max}} = \frac{2 \times 355W}{251.3V \times 0.5} = 5.65A \quad (3.10)$$

RMS current flowing through the primary winding inductance of the transformer:

The RMS current flowing through the inductor is found using the RMS formula of a signal with the maximum duty cycle and maximum current in the primary inductance. The resulting RMS current is:

$$I_{rms} = I_{peak} \times \sqrt{\frac{D_{max}}{3}} = 5.65A \times \sqrt{\frac{0.5}{3}} = 2.3 A \quad (3.11)$$

The primary winding inductance that can support the peak current at the maximum duty cycle of the transformer is:

$$L_p = \frac{D_{max} \times V_{pmin}}{I_{peak} \times F_s} = \frac{0.5 \times 251.3V}{5.65A \times 50000Hz} = 445 \times 10^{-6}H \quad (3.12)$$

If a larger inductance than this value is used the converter may operate in a continuous conduction mode.

3.3.1 Ferrite Core Selection

The ferrite core is selected according to [10] as follows:

For DCM operation at $F_s = 20kHz$ and $B = 0.3T$ the minimum and the maximum core volume that can handle the required power level is given by

$$V_e = (0.1 - 0.5) \times P_{inmax} = (0.1 - 0.5) \times 355W = 35.5cm^3 - 177.5cm^3 \quad (3.13)$$

$35.5cm^3$ is the minimum core volume required to achieve our power level at $20kHz$ and $B = 0.3T$. From transformer datasheets E55/21 N27 material ferrite core is chosen. At high power levels the converter operates at a switching frequency higher than $20kHz$ so the volume of the core needed becomes smaller.

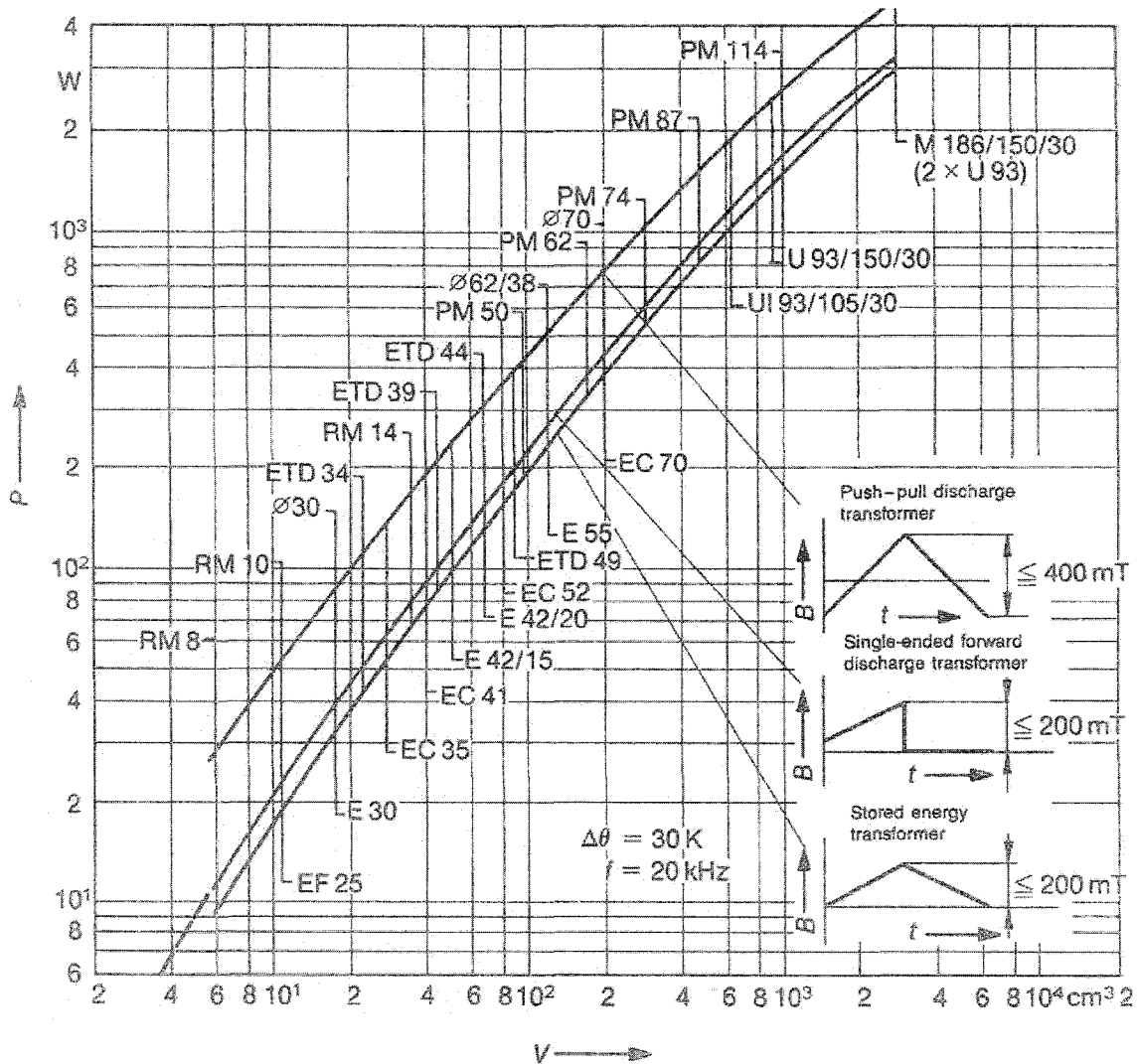


Figure 3.2. Power handling capability P and unit volume V of transformers with ferrite cores of N27/N41 material (guidance values) at $f = 20\text{ kHz}$ for a temperature rise of 30 K in various modes of operation (Siemens) [10].

3.3.2 Hysteresis and Air Gap

Normally the ferrite core can store low energy levels. In forward converters the energy is transferred to secondary side in all cycles; thus in a forward converter there is no need of an air gap. However since in a flyback converter, the idea is to store energy in one cycle and release it in the next, we should improve its energy storage capability. This can be done by introducing an air gap into the ferrite material. Because the permeability of air is much lower than the ferrite material used the air gap can store very high energy compared to the ferrite itself.

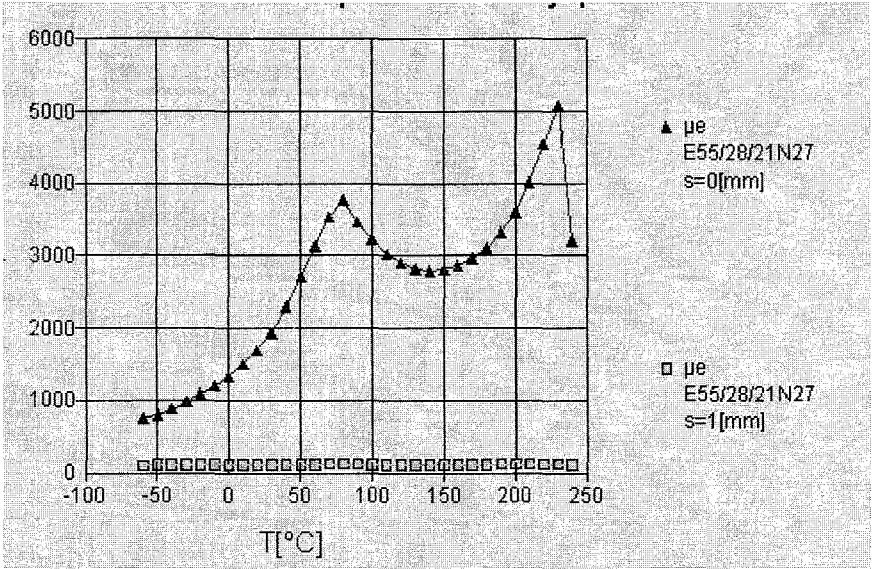


Figure 3.3. Effective permeability without and with a total air gap of 1mm as a function of temperature [11]. Introduction of an air gap made effective permeability almost insensitive to variation in temperature.

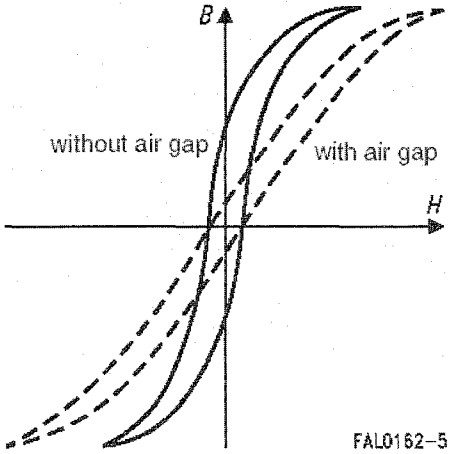


Figure 3.4. Hysteresis loop of a core with and without an air gap [12].

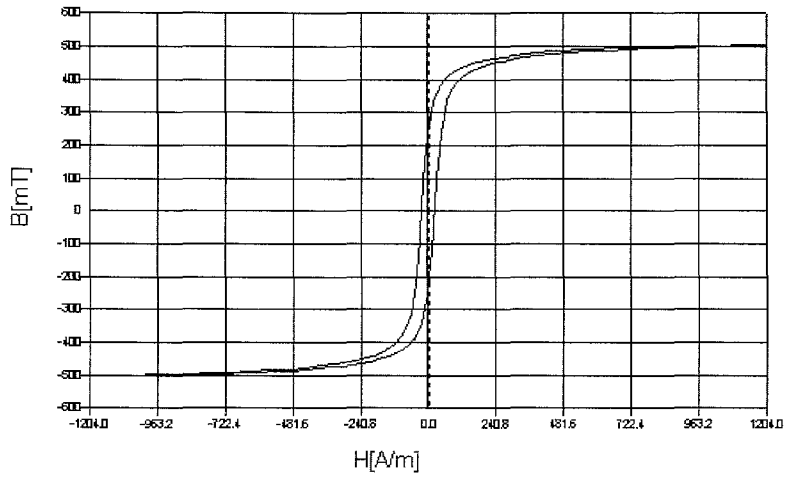


Figure 3.5. Hysteresis curve of N27 material at 25°C and 100°C [11].

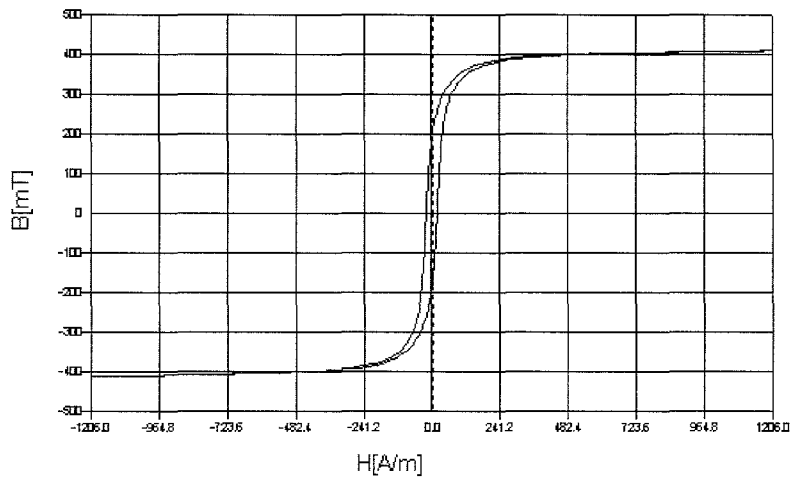


Figure 3.6. Hysteresis curve of N27 material at 100°C [11].

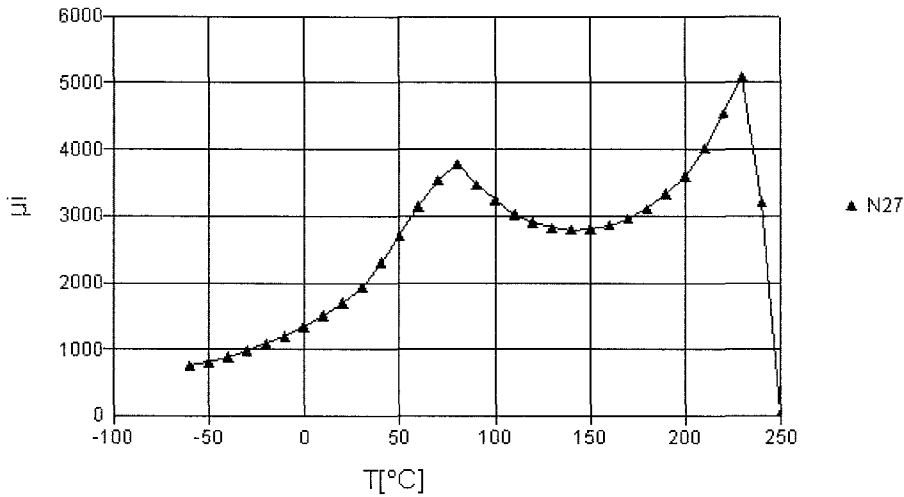


Figure 3.7. Initial permeability of N27 material as a function of temperature [11].

3.3.3 Number of Primary Winding Turns

The required number of turns can be calculated using the formula given in the application note of TDA16846 [13].

$$N_p = \frac{(V_{d\max} - V_{p\max}) \times B_{\max} \times A_e}{(V_{p\min} \times F_{os} + V_{d\max} - V_{p\max}) \times 2 \times I_{inave\max} \times A_l} \text{ (Turns)} \quad (3.14)$$

Here F_{os} is the factor for the overshoots at the primary winding of the transformer. Without PFC $F_{os} = 1.3$, with PFC $F_{os} = 1.8$. A_e is the effective cross-sectional area of the core in m^2 , $V_{d\max}$ is the maximum allowable drain voltage of the MOSFET, A_l is the inductance factor for the ferrite core for a selected air gap, $I_{inave\max}$ is the maximum average current through the primary winding.

$$I_{inave\max} = \frac{P_{in\max}}{V_{p\min}} = \frac{355W}{250.3V} = 1.42A \quad (3.15)$$

For an air gap length of 1mm the inductance factor can be found from ferrite core datasheet $A_l = 496 \times 10^{-9} H$. If the MOSFET's maximum drain voltage is taken as the maximum allowable drain voltage ($V_{d\max} = 800V$), and assuming there is no PFC ($F_{os} = 1.3$), $B_{\max} = 0.2T$ the primary number of turns can be calculated as follows:

$$N_p = \frac{(800V - 357.8V) \times 0.2T \times 354 \times 10^{-6} m^2}{(251.3V \times 1.3 + 800V - 357.8V) \times 2 \times 1.42A \times 496 \times 10^{-9} H} = 28.7 Turns$$

We choose $N_p = 30$. Secondary and auxiliary winding number of turns is based on this value.

The number of secondary turns is given by:

$$N_s = \frac{N_p \times (V_o + V_{fo})}{V_r} \text{ (Turns)} \quad (3.16)$$

In Eq. 3.16 V_r is the reflected output voltage to the drain terminal of the MOSFET, V_{fo} is the on-state voltage drop across the main output diodes (3 diodes are in series so the voltage drop is $V_{fo} = 3 \times 1.5V = 4.5V$). Thus

$$N_s = \frac{30 Turns \times (1000V + 4.5V)}{232V} = 130 Turns$$

The number of auxiliary windings is given by:

$$N_{aux} = \frac{N_p \times (V_{aux} + V_{f_{aux}})}{V_r} = \frac{30 Turns \times (15V + 0.7V)}{232V} = 2 Turns \quad (3.17)$$

The number of secondary winding 2: In the second version of the power supply another secondary winding was added to the transformer in order to help secondary regulation and to provide a supply voltage for the fan that will be used for cooling purposes.

$$N_{s2} = \frac{N_p \times (V_{o2} \times V_{fo2})}{V_r} = \frac{30 Turns \times (15V + 0.7V)}{232V} = 2 Turns \quad (3.18)$$

The transformer data is given in the Table 3.2.

Table 3.2.
Designed Flyback Transformer Data

Transformer Type: EE 55/21 N27 ferrite core

Input voltage range :	187-253 Vrms
Transformer power :	300W
Secondary output 1 voltage and current :	1000V, 0.290A
Secondary output 2 voltage and current :	15V, 0.3A
Auxiliary winding voltage and current :	15V, 0.3A
Primary winding inductance (L_p) :	446 μ H
Secondary winding inductance (L_s) :	8,38mH
Secondary winding 2 inductance :	1.98 μ H
Auxiliary winding inductance :	1.98 μ H
Primary winding turn (N_p) :	30
Secondary winding turn (N_s) :	130
Secondary winding 2 turn (N_{s2}) :	2
Auxiliary winding turn (N_{aux}) :	2
Primary winding leakage inductance :	12,4 μ H
Bobbin data (N_p) :	3 x 0.8mm
Bobbin data (N_s) :	0.45mm
Bobbin data (N_{s2}) :	0.45mm
Bobbin data (N_{aux}) :	0.45mm

3.3.4 Post Calculations of Inductance, Current and Flux Density

After choosing the exact turns for each winding the actual primary inductance values should be determined. With these calculated values the currents and flux density also change and needs a post calculation. Primary, secondary, secondary 2, auxiliary winding inductances (L_p , L_s , L_{s2} , L_{aux}), primary peak current (I_{peak}) and maximum flux density (B_{max}) are:

$$L_p = N_p^2 \times A_l (H) \quad (3.19)$$

$$L_p = 30^2 \times 496 \times 10^{-9} H = 446 \times 10^{-6} H$$

$$L_s = N_s^2 \times A_l (H) \quad (3.20)$$

$$L_s = 130^2 \times 496 \times 10^{-9} H = 8.38 \times 10^{-3} H$$

$$L_{s2} = N_{s2}^2 \times A_l (H) \quad (3.21)$$

$$L_{s2} = 2^2 \times 496 \times 10^{-9} H = 1.984 \times 10^{-6} H$$

$$L_{aux} = N_{aux}^2 \times A_l (H) \quad (3.22)$$

$$L_{aux} = 2^2 \times 496 \times 10^{-9} H = 1.984 \times 10^{-6} H$$

$$I_{ppeak} = \frac{V_{pmin} \times D_{max}}{L_p \times F_s} (A) \quad (3.23)$$

$$I_{ppeak} = \frac{251.3V \times 0.5}{446 \times 10^{-6} \times 50000Hz} = 5.63A$$

And RMS current is:

$$I_{prms} = 5.63A \times \sqrt{\frac{0.5}{3}} = 2.29A$$

$$B_{max} = \frac{L_{pmax} \times I_{ppeak}}{N_p \times A_e} (T) \quad (3.24)$$

$$B_{max} = \frac{446 \times 10^{-6} H \times 5.63A}{30 \times 354 \times 10^{-6}} = 0.237T$$

In energy storage operation it is recommended that the maximum flux density is under 0.2T. In order to get best performance we can increase the number of primary turns slightly to get a lower magnetization in the core. Another solution is to operate the converter in CCM so the energy required at each cycle reduces because the current at the end of the off time is not zero. CCM reduces device stresses but it is harder to control. Because of this in the present application we use DCM. However, rather than using a normal hard switched flyback we use ZVS to reduce the stress on the switching elements. For example if we use 36 turns for primary and 144 turns for secondary we achieve ZVS at minimum input voltage and because of D_{max} being 0.5, ZCS is also achieved ($A_l = 344nH$, $L_p = 446\mu H$, $N_p = 36$ and $N_s = 144$).

Secondary peak and RMS currents:

The secondary RMS current can be estimated using the transformer turns ratio as

$$I_{speak} = I_{ppeak} \times \frac{N_p}{N_s} = 5.63A \times \frac{30}{130} = 1.3A \quad (3.25)$$

The peak current calculated with this formula is not accurate enough because it also contains the losses in the primary current. Since the secondary current does not include these losses an exact analysis should be performed to determine the real secondary peak current.

The secondary peak current is given by

$$I_s = I_{speak} \times \left(1 - \frac{t}{T_{off}}\right) \quad (A) \quad (3.26)$$

$$I_{o\max} = \frac{1}{T_s} \times \int_0^{T_s} I_s \times dt = \frac{1}{T_s} \times \int_0^{T_s} I_{speak} \times \left(1 - \frac{t}{T_{off}}\right) \times dt = \frac{I_{speak} \times T_{off}}{2 \times T_s} \quad (A) \quad (3.27)$$

$$I_{speak} = 2 \times I_{o\max} \times \frac{T_s}{T_{off}} \quad (A) \quad (3.28)$$

If the converter operates at $D = D_{\max} = 0.5$ ($V_p = V_{p\min} = 251.3V$)

$$I_{speak} = 4 \times I_{o\max} \quad (A) \quad (3.29)$$

As a result the secondary peak current is calculated to be $I_{speak} = 4 \times 0.3A = 1.2A$, which is smaller than the calculated value before.

The secondary RMS current can be calculated again with the well known RMS value formula as below;

$$I_{srms} = \sqrt{\frac{1}{T_s} \times \int_0^{T_s} I_{speak}^2 \times \left(1 - \frac{t}{T_{off}}\right)^2 \times dt} \quad (A) \quad (3.30)$$

$$I_{rms} = \sqrt{\frac{1}{T_s} \times I_{speak}^2 \times \int_0^{T_s} \left(1 + \frac{t^2}{T_{off}^2} - \frac{2 \times t}{T_{off}} \right) dt} \quad (A) \quad (3.31)$$

$$I_{rms} = \sqrt{\frac{1}{T_s} \times I_{speak}^2 \times \left(T_{off} + \frac{T_{off}^3}{3 \times T_{off}^2} - \frac{T_{off}^2}{T_{off}} \right)} \quad (A) \quad (3.32)$$

$$I_{rms} = I_{speak} \times \sqrt{\frac{T_{off}}{3 \times T_s}} \quad (A) \quad (3.33)$$

If the converter operates at $D = D_{max} = 0.5$ ($V_p = V_{pmin} = 251.3V$)

$$I_{rms} = I_{speak} \times \sqrt{\frac{1}{6}} \quad (A) \quad (3.34)$$

Calculated secondary RMS current with the peak current calculated in accordance with the turns ratio is;

$$I_{rms} = 1.3A \times \sqrt{\frac{0.5}{3}} = 0.53A$$

If we use the secondary peak current calculated using the exact analysis the RMS current becomes

$$I_{rms} = 1.2A \times \sqrt{\frac{1}{6}} = 0.49A$$

This is also smaller than the value calculated before.

Calculation of the wire sizes for each winding:

The bobbin width of E55/21 is measured to be 33mm. The creepage distance for universal input voltage range is typically 8mm. This sets a minimum margin width as a half of the creepage distance to 4mm. Therefore in order to provide good mains isolation there should be a 4mm margin on each side of the winding area. The isolation between the primary and the secondary windings is done using three layers of isolation tape. More layers of isolation tape provide better isolation but leakage inductance increases. If we

want to eliminate magnetic interference, we should use a screen between the secondary and the primary windings that is at the mains potential.

Thus the effective bobbin winding width and effective winding cross-section are

$$BW_e = BW - 2 \times Mar = 33mm - 2 \times 4mm = 25mm \quad (3.35)$$

$$AN_e = \frac{A_n \times BW_e}{BW} = \frac{280mm^2 \times 25mm}{33mm} = 212mm^2 \quad (3.36)$$

This winding cross-section should be shared between windings (0.675 of available winding cross-section belongs to primary winding, 0.3 to secondary winding and 0.025 to auxiliary and secondary winding 2). If we take copper space factor as $f_{Cu} = 0.42$ available winding areas and possible wire diameters for each winding can be calculated.

$$A_p = \frac{0.675 \times AN_e \times f_{Cu}}{N_p} = \frac{0.675 \times 212mm^2 \times 0.42}{30} = 2.0045mm^2 \quad (3.37)$$

$$\phi_p = 1.6mm$$

$$A_s = \frac{0.3 \times AN_e \times f_{Cu}}{N_s} \quad (mm^2) \quad (3.38)$$

$$A_s = \frac{0.3 \times AN_e \times f_{Cu}}{130} = 0.2045mm^2$$

$$\phi_s = 0.51mm$$

$$A_{aux} = \frac{0.025 \times AN_e \times f_{Cu}}{N_{aux}} \quad (mm^2) \quad (3.39)$$

$$A_{aux} = \frac{0.025 \times 212 \times 0.42}{2} = 1.1136mm^2$$

$$\phi_{aux} = 1.19mm$$

$$A_{s2} = \frac{0.025 \times AN_e \times f_{Cu}}{N_{s2}} = \frac{0.025 \times 212 \text{ mm}^2 \times 0.42}{2} = 1.1136 \text{ mm}^2 \quad (3.40)$$

$$\phi_{s2} = 1.19 \text{ mm}$$

These calculated values are the possible wire diameters. By using these values and not exceeding them we can choose the optimum wire sizes. At high frequencies, after 40 - 50kHz, the skin effect increases and the current tends to flow through the outer parts of the conducting wire. To reduce the resistance of the wire and thus the skin effect occurring at high frequencies several wires can be paralleled or *litz wire* which consists of several wires connected in parallel can be used. We connected *three* 0.8mm wires in parallel to reduce the impedance of the wire.

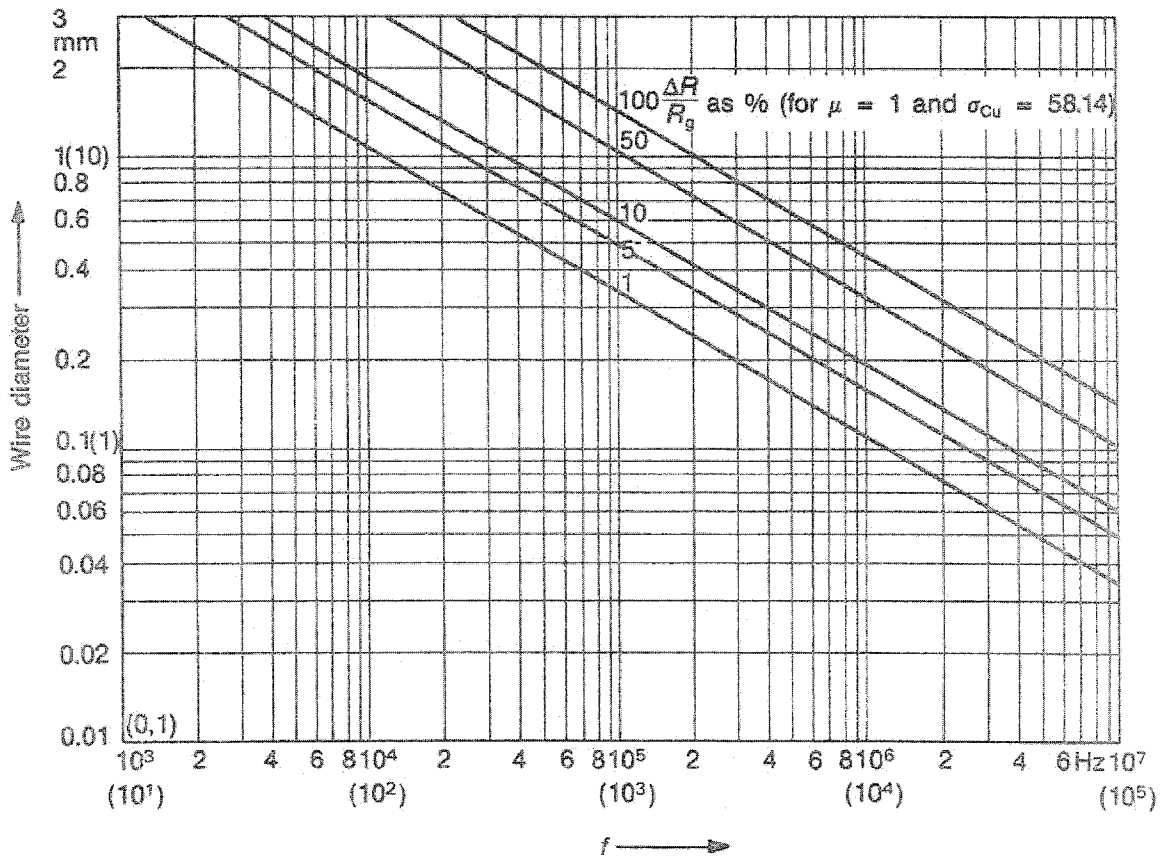


Figure 3.8. Variation with frequency and wire diameter of resistance increase $\Delta R/R_d$ due to skin effect [10].

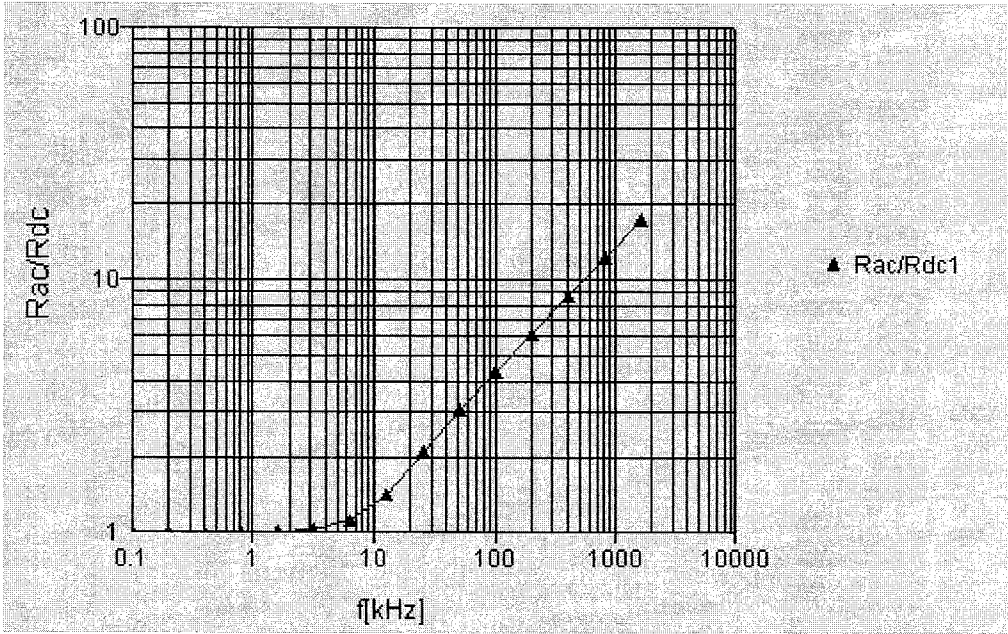


Figure 3.9. R_{ac}/R_{dc} for E55/28/21 ferrite core as a function of frequency, one layer primary with 0.912mm wire size ($R_{dc} = 0.013\Omega$ and $R_{ac} = 0.04\Omega$ at $F_s = 50000Hz$).

Table 3.2
Current Density j , Flux-Density Excursion b and Power-Handling Capability p of Transformers with E-Shape Ferrite Cores at a Switching Frequency of 20kHz [10]

Core type	A_e mm^2	A_{min} mm^2	V_e mm^3	AN mm^2	V_n mm^3	R_{th} K/W	δB mT	J A/mm^2	δB_2 mT	J_2 A/mm^2	P W	Energy storage operation W
E30	60	49	4000	73	3360	22	565	4,9	327	6,5	75	24
E42/15	181	175	17600	157	13700	14	361	3,1	361	3,1	254	110
E42/20	240	229	23300	170	17900	14,3	317	2,6	317	2,6	270	131
E55	354	350	42500	238	26900	10	285	2,6	285	2,5	500	264

Another important parameter is the current density in an inductor. For E55 core the maximum current density is given to be 2,6A/mm² in the datasheet. So when choosing the wire current density should not be higher than this. From the available wires we choose 3*0.8mm wire for primary winding, 1*0.45mm wire for secondary and auxiliary windings.

The cross-sectional area of primary winding when 3*0.8mm wire used is

$$A_p = 3 \times \pi \times \left(\frac{0.8}{2}\right)^2 = 1.51mm^2$$

The cross-section of secondary winding when 1*0.45mm wire used is;

$$A_s = \pi \times \left(\frac{0.45}{2} \right)^2 = 0.159 \text{ mm}^2$$

Now with these values we can calculate the current density of each winding.

$$J_p = \frac{I_{prms}}{A_p} \text{ (A/mm}^2\text{)} \quad (3.41)$$

$$J_p = \frac{2.29 \text{ A}}{1.51 \text{ mm}^2} = 1.52 \text{ A/mm}^2$$

$$J_s = \frac{I_{srms}}{A_s} \text{ (A/mm}^2\text{)} \quad (3.42)$$

$$J_s = \frac{0.49 \text{ A}}{0.159 \text{ mm}^2} = 3.08 \text{ A/mm}^2$$

The primary current density is very good but the secondary current density is higher than the maximum permissible for the E55/21 ferrite core. In order to eliminate this we can use a wire with a greater diameter or we can use forced-air cooling rather than convection to overcome the heating effect of this current density. In our prototype we provide another 15V winding to supply power to a fan that will be included in the power supply.

Leakage inductance:

In a transformer all the flux in the primary winding can not be transferred to the secondary winding because of the existence of leakage inductance. Ideally the secondary winding is coupled to the primary winding 100%. In practical situations ideal coupling can not be achieved. The equivalent circuit of a flyback transformer is shown as in Fig. 3.11. When the MOSFET is on the primary winding current develops through the magnetizing inductance which is the sum of leakage inductance $a \times L_p$ and the inductance $(1-a) \times L_p$. When the MOSFET switches off the energy stored in the inductance $(1-a) \times L_p$ is transferred to the secondary side but the energy stored on the leakage inductance stays

there and causes a voltage overshoot ($L_{\sigma} \times \frac{di}{dt}$) across the power switch. The total capacitance across the MOSFET should absorb this energy and limit the voltage generated by the leakage inductance. Knowledge of the leakage inductance is useful in designing the snubber network.

The leakage inductance can be estimated using the equation [10]

$$L_{\sigma} = \mu_0 \times N_p^2 \times \frac{l_n}{w} \times \left(x + \frac{y}{3} \right) \text{ (H)} \tag{3.43}$$

Here x is the separation between primary and secondary windings, y is the total depth of all windings and l_n is the mean length per turn for E55/21 transformer from the bobbin datasheet in cm . In order to provide good mains isolation between the input and the output we should make x bigger so if we use four layers of isolation tape after primary winding and three layers after auxiliary and assume the tape thickness to be $0.02cm$ then the total separation between primary and secondary becomes $0.14cm$.

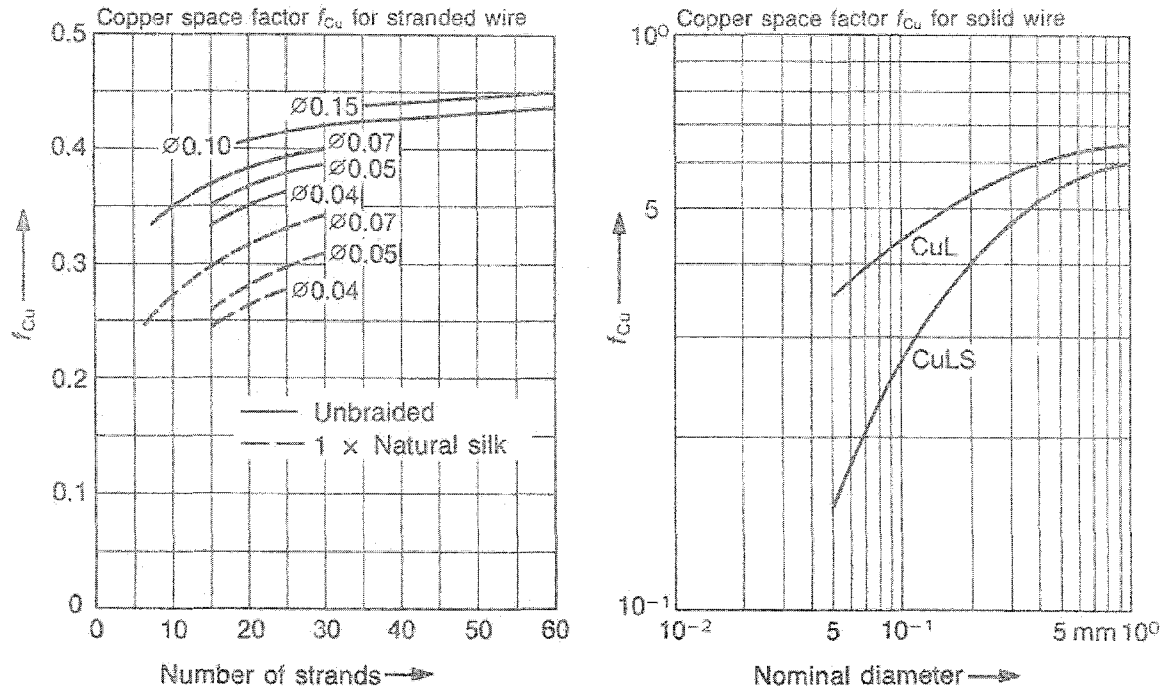


Figure 3.10. Variation with nominal diameter of copper space factor f_{Cu} for stranded and solid wires (Siemens) [10].

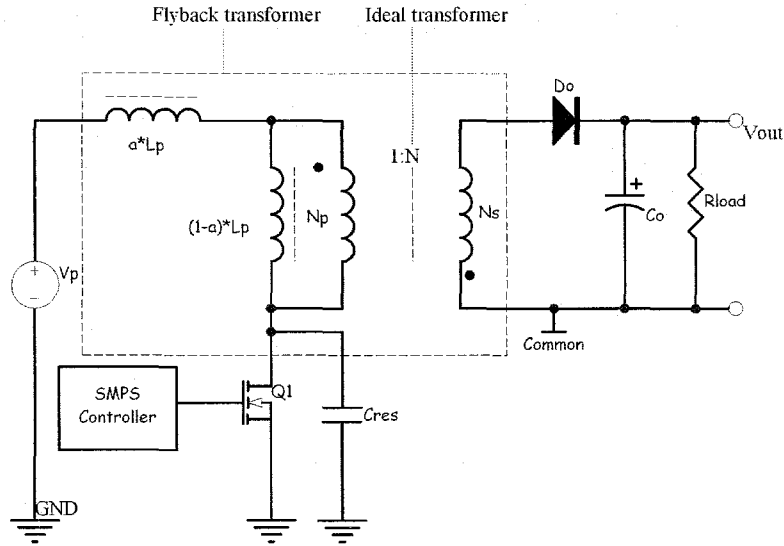


Figure 3.11. Equivalent circuit of transformer (a is the leakage coefficient (k_σ), $a \times L_p$ is the leakage inductance and $(1-a) \times L_p$ is the real inductance which is capable of transferring all the energy to the secondary side).

$$L_\sigma = 1.257 \times 10^{-8} \frac{H}{m} \times 30^2 \times \frac{11.3cm}{3.3cm} \times \left(0.14cm + \frac{0.375cm}{3} \right) = 10.26 \times 10^{-6} H$$

If an LCR meter is available the inductance can be measured easily. Primary inductance (total magnetizing inductance) can be measured while all the secondary windings are open circuit. The leakage inductance can be measured by measuring the primary inductance while all the secondary terminals are short circuited. When we used the normal winding technique (primary winding, auxiliary winding, secondary winding one over another) leakage inductance is measured to be $12.4 \times 10^{-6} H$ while the primary inductance was measured to be $442 \times 10^{-6} H$. The leakage coefficient can be found from

$$k_\sigma = \frac{L_\sigma}{L_p} = \frac{12.4 \times 10^{-6} H}{442 \times 10^{-6} H} = 0.028 \quad (3.44)$$

According to [10] the leakage coefficient should not be more than 0.5% while in practical situations and in application notes its practical limit is about 2-3%. Our measurement agrees with the latter one. Also in [10] it is also told that the leakage inductance is not an important parameter in single output flyback converters in which there is one controlled output. But if we want to design a multi output flyback converter leakage inductance should be lowered [10].

to prevent electrical interference to reach into the power supply and reduce the radiated electrical interference from the power supply to the power line special kinds of EMI filters are used. These filters are inductance-capacitance type and they usually have a damping effect only at frequencies above around 50kHz. They do not protect against voltage variations but they provide attenuation of common and DM interference at the input.

Commercial power supplies cannot be put on the market until they do not meet certain EMC regulations. The total EMI has two components; DM and CM components.

DM EMI: DM currents flow from hot to neutral or from the neutral to hot. These currents are opposite in phase (180° out of phase) on the hot line and the return [15]. DM currents are resultant of the switching current waveform (which is twice the switching frequency) and it is proportional to the input current [15].

CM EMI: CM current flows into parasitic (stray) capacitors between the power converter components and protection earth. These currents are in phase (no phase difference) that is they flow in the same direction in both the hot line and the neutral [15]. They return to the source via the system ground. Because these currents flow through the ground planes and the PCB area of ground planes are much larger than the other power tracks they provide good antennas for the radiated EMI [16]. These currents may even reach at the output by the transformer stray capacitance. CM currents are the resultant of the voltage waveform developed across the drain of the switching transistors [15]. The CM current is developed due to the charging and discharging of the parasitic capacitance (10 to 150pF) between transistor junction and its heat sink which provides the path to system ground [15].

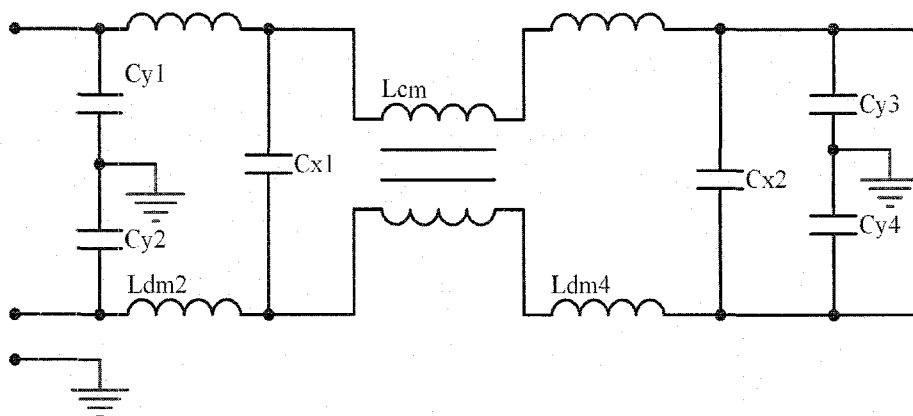


Figure 3.13. Typical EMI filter including CM inductor and X capacitors with it and DM inductors and Y capacitors with them.

DM currents are proportional to input current and CM currents are proportional to the input line voltage according to [15] which current mode is dominant for the converter can be estimated at the design stage using these parameters (i.e. High-Voltage - Low-Current: CM, Low-Voltage-High-Current: DM).

DM currents are troublesome in the low frequency range and the main input capacitance (C_{in}) should provide at least 80% of the DM current filtering [15]. The X capacitor placed between the hot line and neutral in series with an inductor provides attenuation below 1MHz is achieved.

CM currents are troublesome above 1MHz. The CM currents below 1MHz are filtered out easily by the use of a CM inductor in series with line to ground and neutral to ground type capacitors. After 1MHz the parasitic capacitance of the CM inductor becomes effective and its attenuation capability is reduced. Thus in order to be effective at high frequencies, these capacitors should be of good quality and have low lead inductance. Y type capacitors are suitable for these capacitors.

Another important parameter is the leakage current that should be taken into consideration. Because the Y type capacitors are connected between hot line and system ground and the case is also connected to ground, when someone touches the case in the event of a failure the current is limited only by the inductive reactance of the Y type capacitor and body resistance which is about $2K\Omega$. So the values of Y type capacitors are limited to a few nanofarads. Also in medical supplies Y type capacitors are omitted to reduce the leakage current. Therefore the CM mode attenuation is not as good as it was

with a normal connection. If Y type capacitors are used two of them can be connected in series to prevent electrical shock in case of a capacitor short.

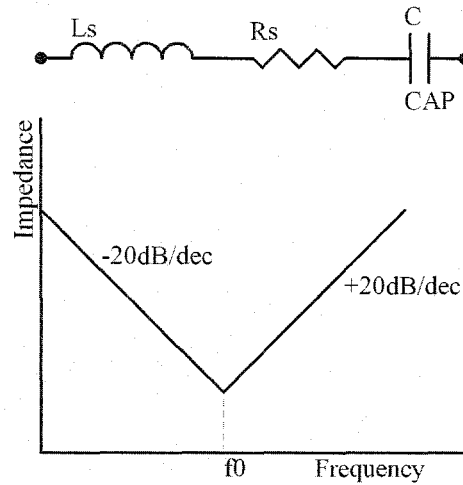


Figure 3.14. Simplified equivalent diagram of a capacitor and its impedance versus frequency plot (R_s : series resistance of the supply leads, electrodes and the losses of the dielectric, L_s : the inductance of the supply leads and electrodes, C : capacitance).

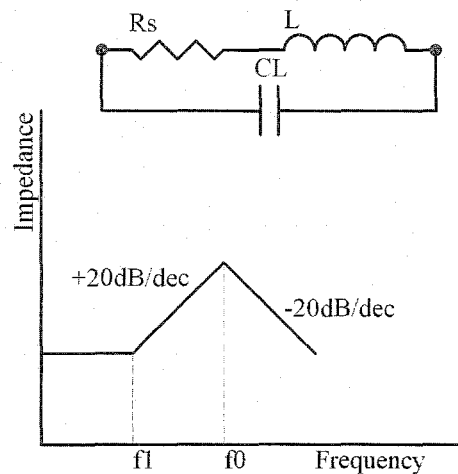


Figure 3.15. Simplified equivalent diagram of an inductor and its impedance versus frequency plot (L : inductance, R_s : series resistance (resistance in the wire and other losses in the wire and core) and CL : self-capacitance in the coil (distributed capacitance, parasitic capacitance or stray capacitance)).

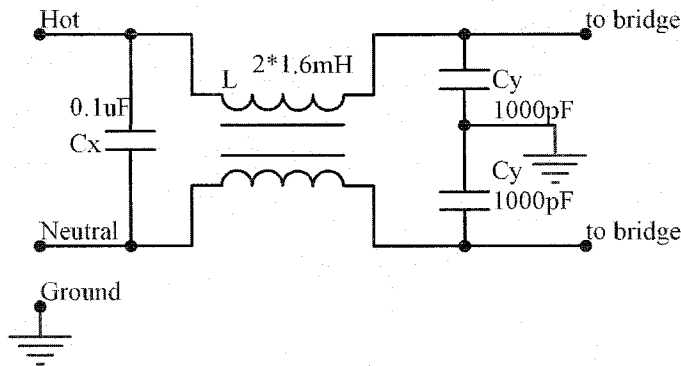


Figure 3.16. EMI filter used.

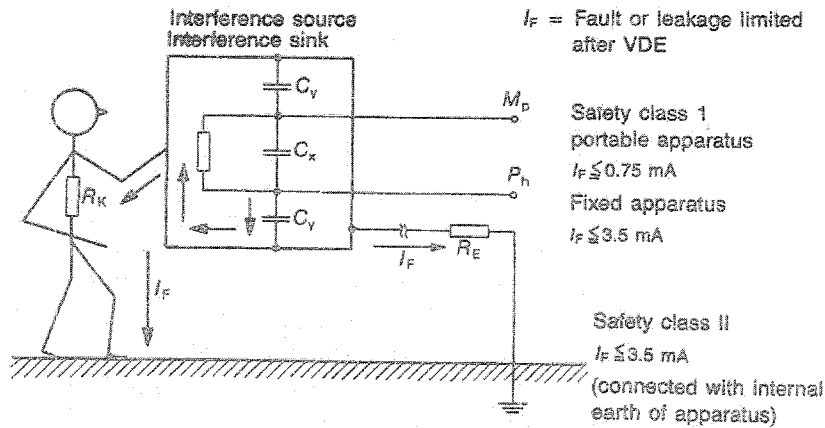


Figure 3.17. X and Y capacitors and leakage current (Siemens) [10].

Another important feature of using an input filter is it provides some PFC action. As the inductance value of the input filter increases the input current shape closes to a sinusoidal waveform. This means that the input current peaks are lower and the RMS current drawn from the mains decreases. On the other hand, an increase in inductance means increase in filter size. An optimization should be done to select the sufficient filter inductance.

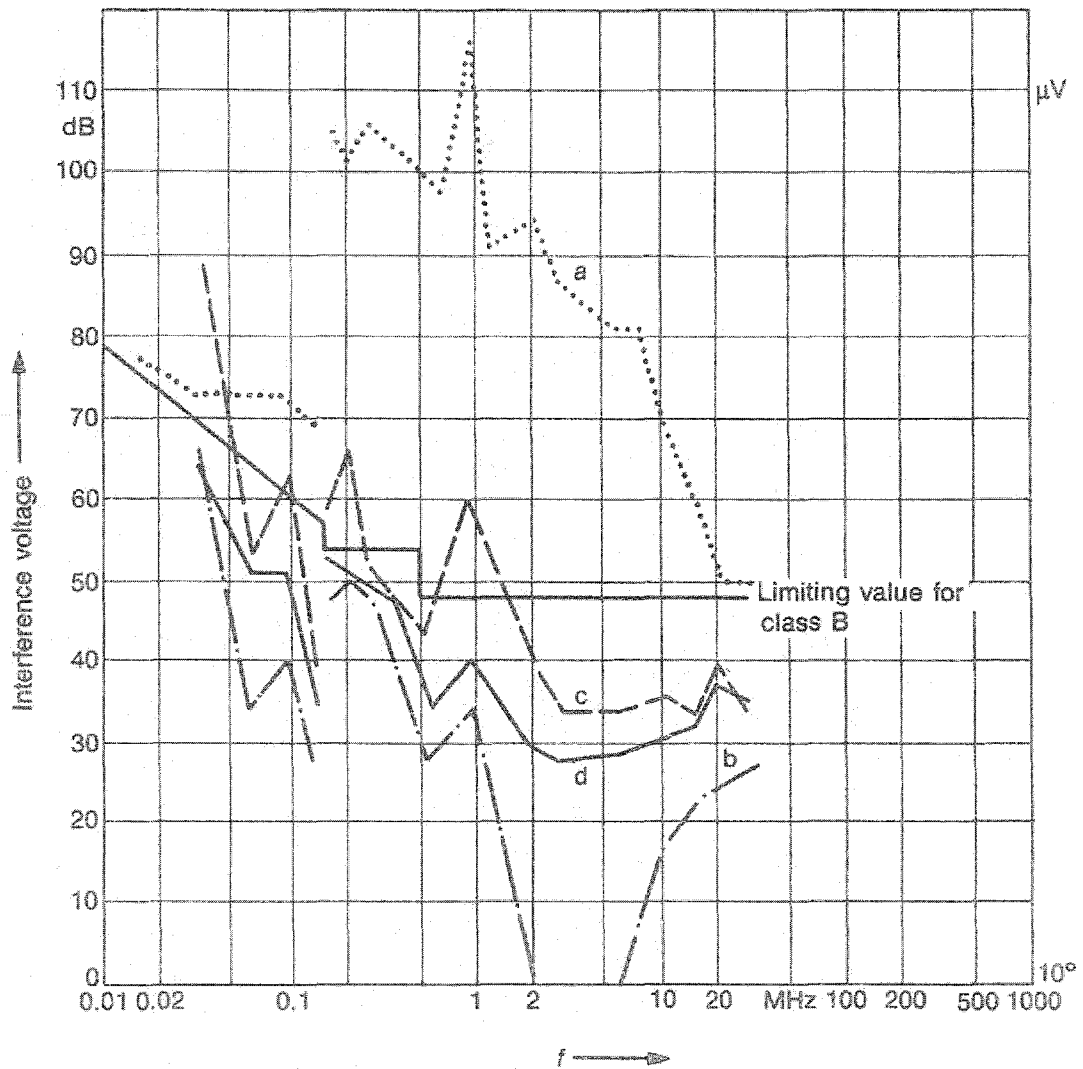


Figure 3.18. Interference levels in a switched-mode power supply with various suppression measures (Siemens) without suppression, -.-.-.- suppression outside the enclosure (inductor unscreened), ----- suppression within the enclosure (inductor unscreened), straight line suppression within the enclosure (inductor screened) [10].

Transient Surge Protection: Effective transient surge protection at the input of a SMPS can be achieved using a MOV. The V-I characteristics of a typical MOV are shown in Fig.3.19. MOVs are bi-directional and exhibit a sharp breakdown which helps limit voltage surges. We used a MOV to protect the power supply from voltage surges.

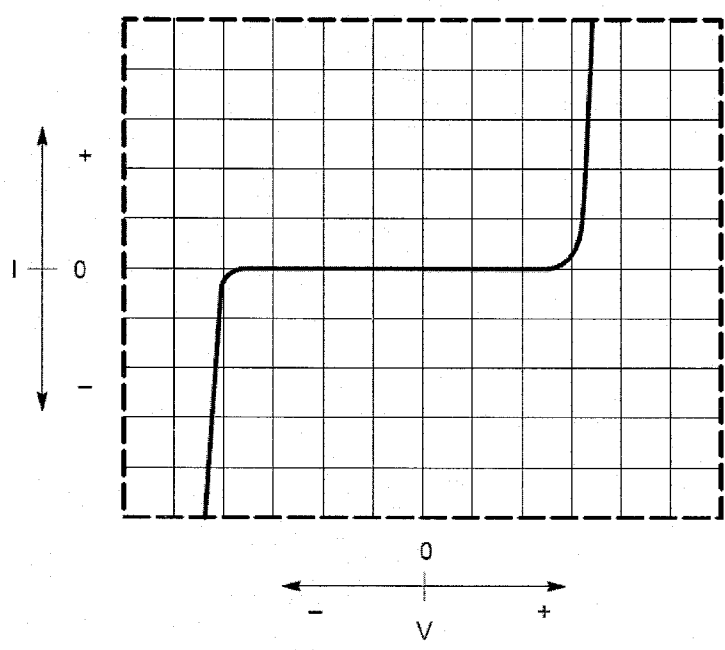


Figure 3.19. Typical V-I characteristic of an MOV [17].

Input smoothing capacitor RMS current rating determination: Input current to a SMPS is not sinusoidal. The current is supplied by current pulses and thus the RMS currents are very high. Normal multimeters cannot measure RMS current because they are calibrated to measure sinusoidal currents. In order to measure complex waveforms true RMS multimeters should be used. These multimeters use ICs to measure the true RMS value of the input signal.

Input smoothing capacitor can be calculated in two different methods [10]. In the first method the input current is observed via an oscilloscope and then using the diode conduction time RMS current is calculated. Details can be found in [10]. In the second method the input diode bridge current form factor is used. We chose this method because we did not have a current probe available. Calculation of the capacitor RMS current relying on this method provides higher current than observed because the input resistor is not taken into consideration; thus the result corresponds to the worst case situation.

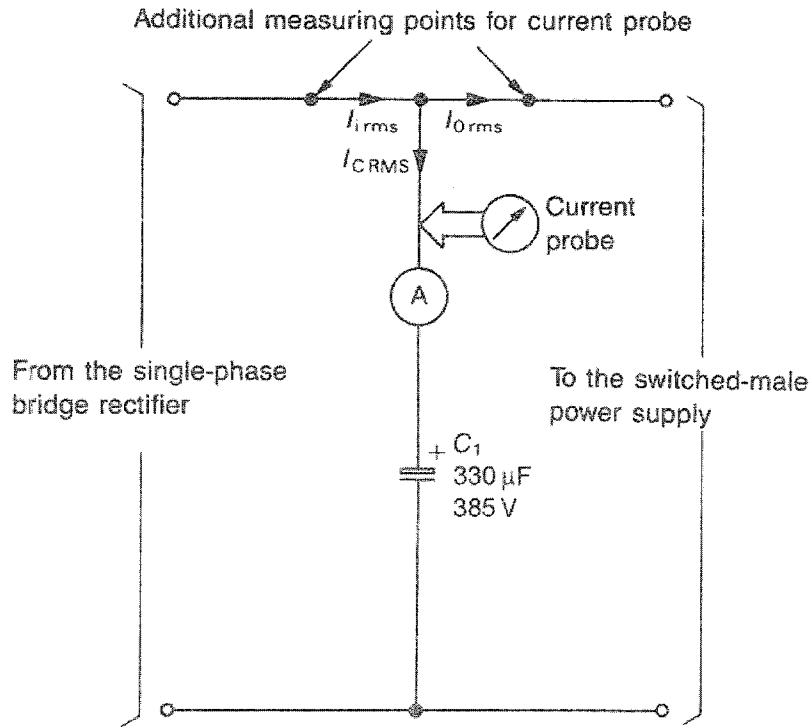


Figure 3.20. Definitions of alternating capacitor currents with indications of measuring points (Valvo) [10].

$$I_{inave} = \frac{P_{inmax}}{V_{pmin peak} - 0.5 \times V_r} = \frac{355W}{187 \times \sqrt{2} - 0.5 \times 14} = 1.379A \quad (3.46)$$

$$x = \frac{V_r}{V_{pmin peak}} = \frac{14}{187 \times \sqrt{2}} = 0.0529 \quad (3.47)$$

The diode current form factor k_{fd} is

$$k_{fd} = \frac{1}{x} \times \sqrt{\pi \times \left(\frac{\pi}{2} - \sin^{-1}(1-x) - (1-x) \times \sqrt{2 \times x - x^2} \right)} \quad (3.48)$$

$$k_{fd} = \frac{1}{0.0529} \times \sqrt{\pi \times \left(\frac{\pi}{2} - \sin^{-1}(1-0.0529) - (1-0.0529) \times \sqrt{2 \times 0.0529 - 0.0529^2} \right)} = 5.05$$

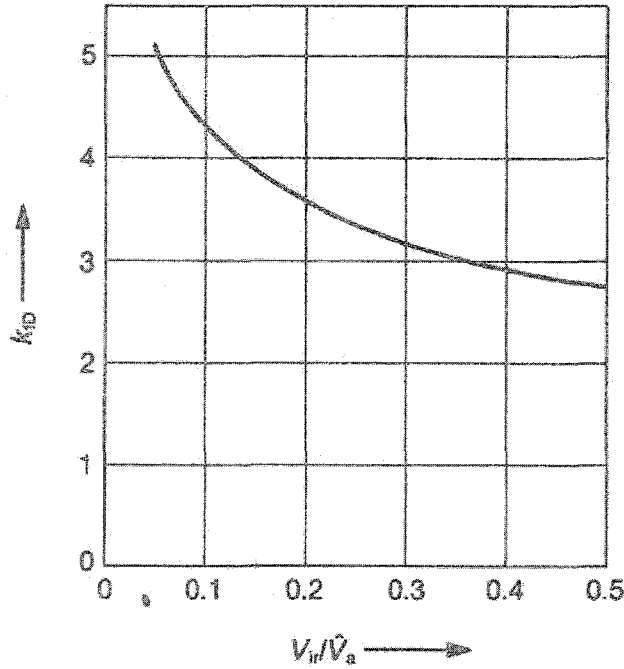


Figure 3.21. Variation of diode current form factor k_{fD} with ripple voltage V_{ir} relative to the peak minimum voltage $V_{pmin\ peak}$ [10].

$$I_{Cinrms} = I_{inave} \times \sqrt{0.5 \times k_{fd}^2 + k_{fi}^2 - 2} \quad (A) \quad (3.49)$$

Here k_{fi} is the converter input current form factor and it lies between the limits;

$$1.3 < k_{fi} < 1.7 \quad (3.50)$$

We take it as 1.5 for the calculation of capacitor RMS current.

$$I_{Cinrms} = 1.379A \times \sqrt{0.5 \times 5.05^2 + k_{fi}^2 - 2} = 4.97A$$

This is the capacitor current rating at maximum temperature. Divider ratios for the selected ambient temperature can be found from the electrolytic capacitor datasheet. We selected from Epcos B43304 capacitor datasheet [18]. The divider ratio at the ambient temperature of $T_a = 50^\circ C$ is about 1.6 for 25000 hours of useful working life.

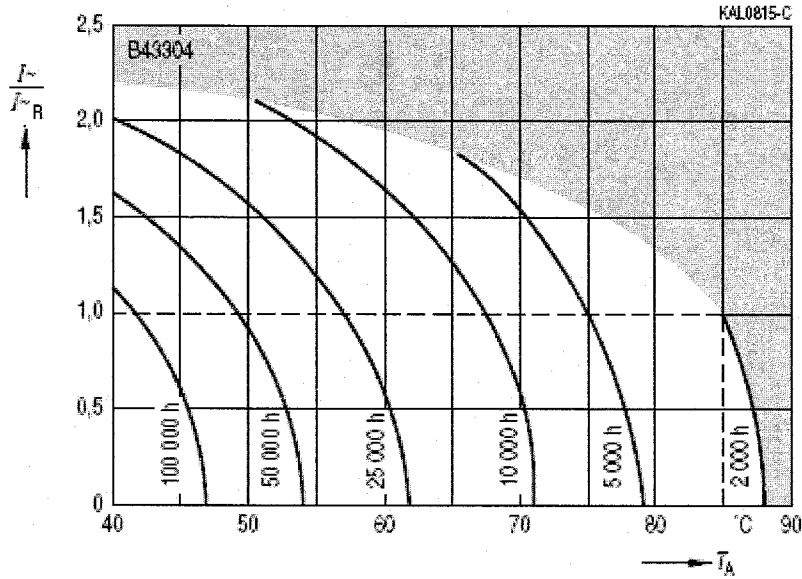


Figure 3.22. Divider ratios as a function of temperature for Epcos B43304 electrolytic capacitors [18].

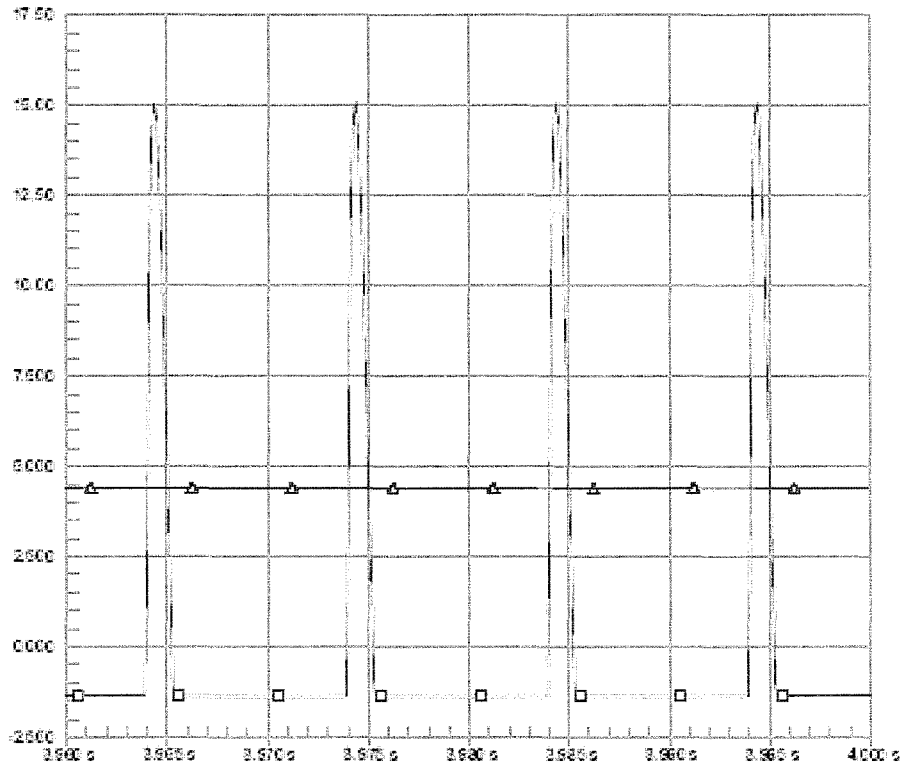
By using this value and I_{Cinrms} we can determine the required capacitor rating at selected operating temperature.

$$Cap_Rating = \frac{I_{Cinrms}}{Divider_Ratio} = \frac{4.97A}{1.6} = 3.106A \quad (3.51)$$

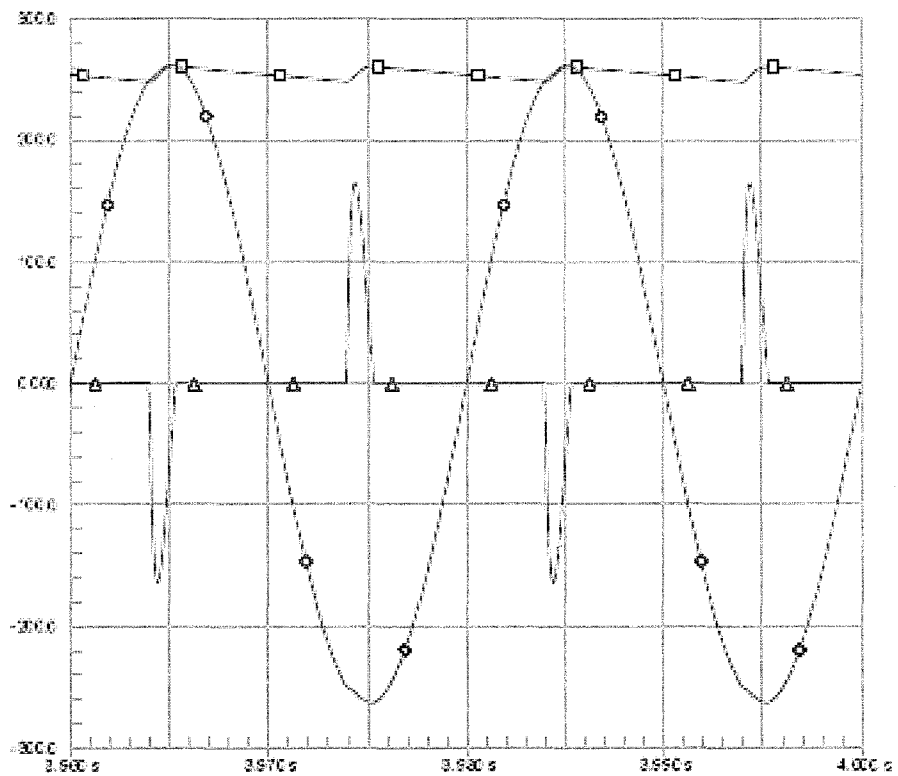
Therefore we can choose a capacitor whose current rating is 3,1A. We selected two 470 μ F capacitors in parallel, so the current rating can be divided by two, giving the sufficient capacitor rating of;

$$Cap_Rating = \frac{3.106}{2} = 1.553A$$

We also simulated the input capacitor current with Protel99 to validate this selection criteria, Fig.3.23 shows the simulated input and capacitor currents.



(a)



(b)

Figure 3.23. Simulated input and capacitor currents for capacitor ripple selection. In (a) the peaks show the input capacitor current and straight line is the RMS current. In (b) the alternating waveform is the input voltage and over it the capacitor voltage can be seen and the peaks are the $10 \times I_{in}$.

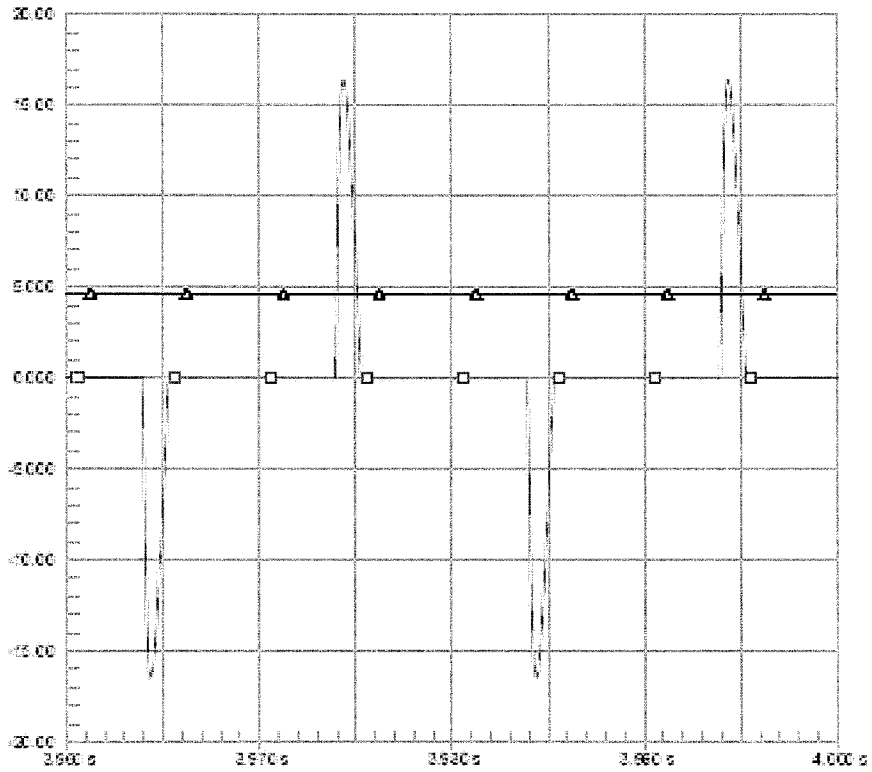


Fig3.23. (continued) Simulated input and capacitor currents. (c) input current and the RMS input current is seen. As seen from the figure the calculated values are very close to the simulated ones. The simulation parameters are: $V_{in} = V_{inmin} = 187V_{rms}$, $C_{in} = 940 \times 10^{-6} F$, $I_{inave} = 1.379 A$.

3.4.1.1 NTC resistor selection

In order to limit the inrush current we can use a NTC resistor. After warming up, the value of this resistor drops to a level so that it does not consume much power. The loading of the NTC resistor is;

$$I_{Rrms} = I_{inave} \times \frac{k_{fd}}{\sqrt{2}} = 1.379 A \times \frac{5.05}{\sqrt{2}} = 4.924 A \quad (3.52)$$

From data book we choose 6A, 5Ω NTC. At maximum current its resistance drops to 0.128Ω.

3.4.1.2 Bridge rectifier selection

The minimum bridge rectifier voltage rating should be $253V_{rms}$ which corresponds to the minimum input voltage. One important parameter of the rectifier is the I^2t rating which is useful in determining the input fuse. If it is not given in datasheet it can be calculated with Eq. 3.53 using the $10 \times 10^{-3} s$ (for 50Hz) surge current I_{FSM} of the rectifier.

$$I^2t = \int_0^{10 \times 10^{-3} s} i^2 \times dt = \frac{1}{2} \times 10 \times 10^{-3} s \times I_{FSM}^2 (A^2 s) \quad (3.53)$$

For KBU8K bridge rectifier I_{FSM} is 300A and I^2t is:

$$I^2t = \frac{1}{2} \times 10 \times 10^{-3} s \times 300^2 A^2 = 450 A^2 s$$

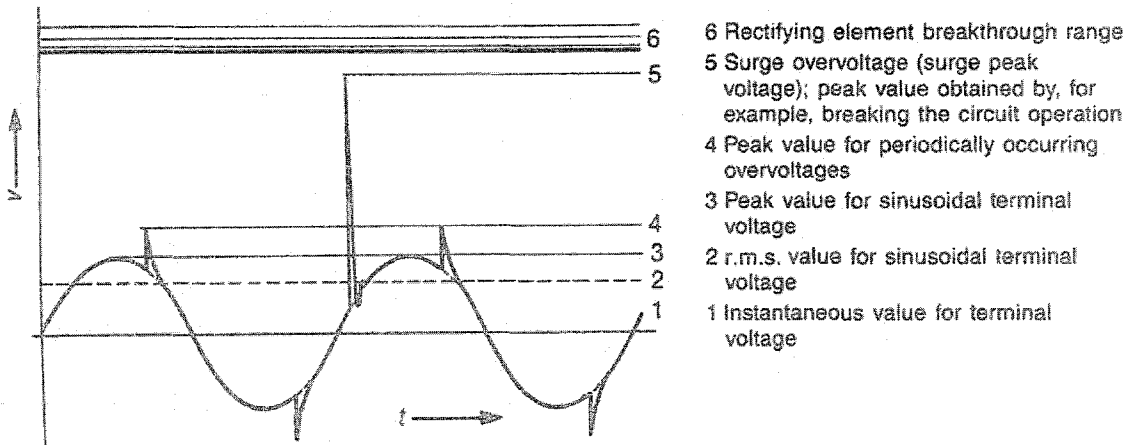


Figure 3.24. Explanation of terms relative to mains supplies [10].

3.4.1.3 Input fuse selection

Main parameters when selecting the fuse are [14]:

- The rated voltage is the greatest extended working voltage and the type of voltage (AC voltage or DC voltage) at which the fuse may be used.
- The current rating is the working current at which the fuse is designed to be used. The current rating is somewhat lower than the current which is able to flow for an extended period without tripping the fuse. The difference between

these two currents varies for different standards (e.g. CSA, IEC, Miti, and UL).

- The breaking characteristics describe the correlation between how quickly the fuse trips and how high the current is. There are different types of fuses in terms of breaking characteristics (FF (very fast), F (fast), M (intermediate), T (delayed action), TT (very delayed action)). In SMPSs the inrush current during startup is very high so fuses with delayed characteristics (T type) are used.

Since the RMS current flowing through the NTC is found to be $4.924A$ the fuse current rating should be slightly higher than this. The minimum voltage rating should be higher than the maximum input voltage which is $253V_{rms}$. We selected $5A$, $250V_{rms}$ fuse. If sufficient data is available in the fuse table, the peak-let-through current (I_{PLT}) and the total clearing energy (I^2t) can be found using the fault current, voltage and current information.

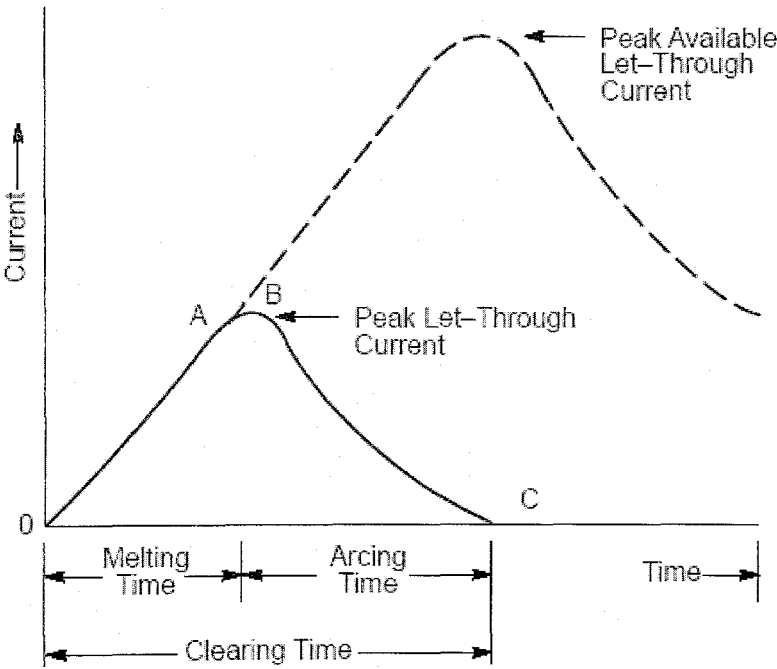


Figure 3.25. Action of a fuse in a circuit subjected to a current surge [17].

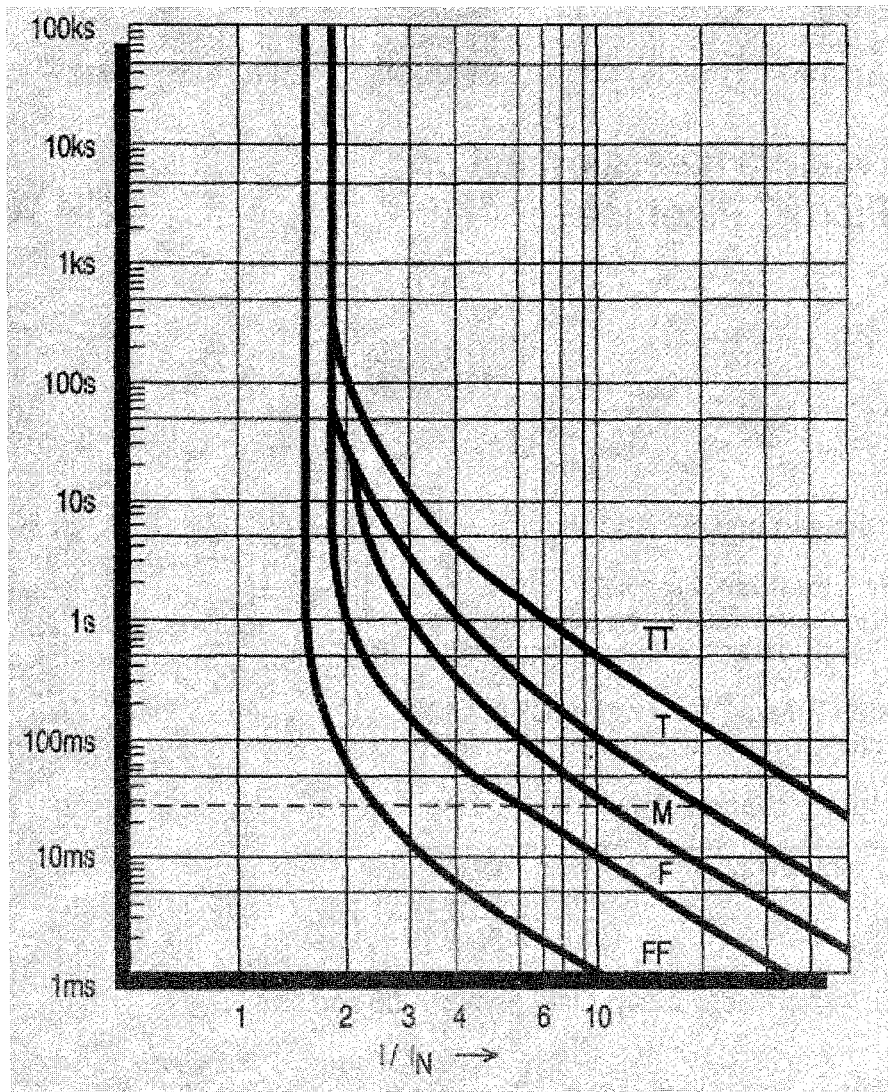


Figure 3.26. Trip time for different types of fuses at various currents, relative to the nominal current rating [14].

3.4.1.4 MOSFET selection

There are some parameters which should be considered during selection of the correct MOSFET for application. The system requirements that have the strongest influence on MOSFET selection are [19], [20]:

- Load current, I_{load} , since this directly affects the MOSFET power dissipation;
- The system input and output voltages, which determine the MOSFET duty cycle;
- Switching frequency, F_s , which affects the power dissipated during the switching transition time; and

- Maximum allowed operating temperature of the MOSFET, which is usually specified to meet a long term reliability goal for the system.

After the operating conditions of the system have been determined (load current, frequency, V_{out} , etc.) the choices for the power MOSFET design are the following:

- The value of $R_{DS(on)}$. The lower the value, the less dissipation and the better the system will work. However, lower resistance MOSFETs will cost more than higher resistance devices.
- Heat sink. If space available, external heat sinks can achieve the same results as a lower $R_{DS(on)}$ at a lower cost.
- Multiple MOSFETs. Sometimes the same operating temperature can be achieved at a lower price with two higher value $R_{DS(on)}$ MOSFETs connected in parallel, if board space permits.

The voltage at the drain terminal at turn-off consists of input capacitor voltage, reflected secondary voltage and the voltage spike caused by the leakage inductance of the transformer. So the total voltage across the drain equals to $357.8V + 231V = 588.8V$. If we use a 600V MOSFET the overshoot should be 11.2V for the MOSFET to operate in its normal conditions. This means that we should snub the voltage overshoot to 11.2V. In order to prevent MOSFET from exceeding operating conditions we can use a higher drain voltage rated MOSFET. We can choose 800V MOSFETs which provides us a voltage overshoot of 211.2V. From Fairchild catalog we choose 13A, 800V, FQA13N80 power MOSFET which has a maximum current capability higher than our peak current. Also its on state resistance is 0.75Ω which is low enough to reduce conduction losses [21].

3.4.1.5 Drain capacitor selection

The total capacitance at the drain of the MOSFET equals to the sum of MOSFET output capacitance C_{oss} , transformer stray capacitance C_{stray} and the resonance capacitor C_{res} .

$$C_r = C_{oss} + C_{stray} + C_{res} \text{ (F)} \quad (3.54)$$

3.4.1.6 Output rectifier selection

Because our output voltage is $1000V$ we should use several diodes in series. The maximum reverse voltage across these diodes determines the number of diodes to be used. Another important parameter is t_{rr} which shows the speed of the diode. Because SMPSs operate in high frequency t_{rr} should be small enough. We can use HV diodes but as the reverse breakdown voltage increases the reverse recovery time increases. So we should use multiple diodes in series configuration.

The maximum reverse voltage across the output rectifier is;

$$V_{rdiode} = V_o + \left(V_{pmax} \times \frac{N_s}{N_p} \right) = 1000V + \left(357.8V \times \frac{130}{30} \right) = 2550V \quad (3.55)$$

Taking this value and maximum secondary current we selected 3*MUR4100E (4A, 1000V, 75μs) power diode which is in series configuration. Rectifiers connected in series tend to unequally share the total reverse voltage because of the variations in reverse characteristics: leakage currents and turn-off switching parameters [22].

In order to ensure that each diode is operating within its voltage rating, it is generally necessary to add a voltage sharing network. There are two types of voltage sharing steady state and transient. In our converter we used steady state voltage sharing and placed resistors in parallel with the output rectifiers. As an improvement capacitors can be added to provide transient voltage sharing. The calculations of sharing resistors are done according to [22] as below:

Normal formula: This formula gives the value of the parallel resistor to be used.

$$R < \frac{V_{RRM} \times (V_{RRM} \times (n+a) - V_M \times (1+a))}{\Delta I_R \times V_{RRM} \times (1+a) \times (n-1) - I_{RM} \times (1-k) \times (1+a) \times (n \times V_{RRM} - V_M)} \quad (\Omega) \quad (3.56)$$

Here V_{RRM} is the peak repetitive reverse voltage of the rectifier used (1000V for MUR4100E), V_M is the maximum blocking voltage (peak output rectifier voltage stress, calculated as 2550V), n is the number of rectifiers used, a is the upper limit of the resistance tolerance, ΔI_R is about $0.85 \times I_{RM}$, I_{RM} is the maximum leakage current at

V_{RRM} and at the operating junction temperature and is taken from the datasheet, k is a constant and can be taken as 0.8 for today's rectifiers.

If operating junction temperature (I_{RM}) is known the maximum leakage current can be calculated as follows: Junction-to-ambient thermal resistance of MUR4100E is taken as $R_{\theta a} = 50^\circ C/W$ from datasheet. The total power lost in three output rectifiers is $2.205W$.

So the power loss per diode is $P_{orecl,2,3} = \frac{2.205W}{3} = 0.735W$. So the temperature rise in the rectifier caused by power dissipation is:

$$T_j - T_a = P_{orecl,2,3} \times R_{\theta a} = 0.735W \times 50^\circ C/W = 36.75^\circ C.$$

The total rectifier temperature is $36.75^\circ C + 50^\circ C = 86.75^\circ C$. At this temperature the maximum leakage current flowing through the rectifier is found using:

$$I_{RM}(T_j) = I_{RM}(100^\circ C) \times \exp[-0.054 \times (100 - T_j)] \quad (3.57)$$

From datasheet I_{RM} is read as $22 \times 10^{-6} A$ @ $100^\circ C$ and V_{RRM} so at $86.75^\circ C$ the maximum leakage current is:

$$I_{RM}(T_j) = 22 \times 10^{-6} A \times \exp[-0.054 \times (100^\circ C - 86.75^\circ C)] = 10.76 \times 10^{-6} A$$

$$\Delta I_R = 0.85 \times I_{RM} = 0.85 \times 10.76 \times 10^{-6} A = 9.146 \times 10^{-6} A$$

If we choose 5% resistor values because all the calculations are based on the worst case that is all the resistors except one resistor is at its lowest limit and the remaining is at its upper limit a can be taken as $a = 0.1$. Thus the maximum resistance of the resistors should not exceed:

$$R < \frac{V_{RRM} \times (V_{RRM} \times (n+a) - V_M \times (1+a))}{\Delta I_R \times V_{RRM} \times (1+a) \times (n-1) - I_{RM} \times (1-k) \times (1+a) \times (n \times V_{RRM} - V_M)} \quad (3.58)$$

$$R < \frac{1000V \times (1000V \times (3+0.1) - 2550V \times (1+0.1))}{9.146 \times 10^{-6} A \times 1000V \times (1+0.1) \times (3-1) - 10.76 \times 10^{-6} A \times (1-0.8) \times (1+0.1) \times (3 \times 1000V - 2550V)}$$

$$R < 15.5 \times 10^6 \Omega$$

If we think of the worst case we can use the following approach:

Pessimistic approach: In this approach the calculated resistor value gives a lower resistance value and then a high power dissipation ($k=1$: supposing that the reverse current I_{RM} is constant, whatever the blocking voltage across the diode, $a=0$: neglecting the effect of the tolerance of resistors, $\Delta I_R = I_R$: the worst case is taken into account (I_R is the $I_{R \max}$ at $T_{j \max}$ specified ($100^\circ C$))).

$$R < \frac{n \times V_{RRM} - V_M}{(n-1) \times I_R} = \frac{3 \times 1000V - 2550V}{(3-1) \times 22 \times 10^{-6} A} = 10.2 \times 10^6 \Omega \quad (3.59)$$

Because we choose a good rectifier that can work at high frequencies and has low leakage current, high value resistances can be used with these rectifiers and thus the total power lost in these resistances becomes very low. We choose three $8.2 \times 10^6 \Omega$ resistors for this purpose. The power loss is:

$$P_{ores} = 3 \times \left(0.5 \times \frac{\left(\frac{V_M}{3} \right)^2}{R} \right) = 3 \times \left(0.5 \times \frac{\left(\frac{2550V}{3} \right)^2}{8.2 \times 10^6 \Omega} \right) = 0.132W$$

3.4.1.7 Snubber Network

In order to reduce the overshoot effect of leakage inductance at turn-off we can use a snubber circuit across the transformer primary winding. There are different kinds of snubber circuits. In the present design we use the well known RCD snubber. The R_{snub} , D_{snub} and C_{snub} can be calculated using the following equations.

Snubbing voltage for a chosen MOSFET is;

$$V_{snub} = V_{BVDSS} - V_{pmax} - V_r = 800V - 357.8 - 231 = 211.2V \quad (3.60)$$

This value is the assumed voltage overshoot caused by the leakage inductance. If the exact value should be found a detailed analysis should be performed with the known value of the

leakage inductance. It can also be found by observing the drain voltage waveform of the MOSFET with an oscilloscope.

$$C_{snub} = \frac{I_{ppeak}^2 \times L_{\sigma}}{(V_r + V_{snub}) \times V_{snub}} = \frac{5.63 A^2 \times 12.4 \times 10^{-6} H}{(231 V + 211.2 V) \times 211.2 V} = 4.2 \times 10^{-9} F \quad (3.61)$$

$$R_{snub} = \frac{(V_{snub} + V_r)^2 - V_r^2}{0.5 \times L_{\sigma} \times I_{ppeak}^2 \times F_s} = \frac{(211.2 V + 231 V)^2 - 231^2 V^2}{0.5 \times 12.4 \times 10^{-6} H \times 5.63^2 A^2 \times 50000 Hz} = 14469 \Omega \quad (3.62)$$

Power rating of R_{snub} can be determined using the loss calculated in the losses section.

The maximum inverse voltage stress on the snubber diode is;

$$V_{dsmub} = V_{pmax} + V_r = 357.8 V + 231 V = 588.8 V \quad (3.63)$$

3.5 Power Losses

In this section we calculate the losses for a hard switched flyback converter. Because the switch voltage during turn-on is close to zero (or zero) in a QRC the MOSFET turn-on losses will be smaller than calculated ones. The total power loss in converter circuit consists of input diode bridge loss (P_{bridge}), primary winding copper loss ($P_{copperp}$), copper loss of secondary winding ($P_{coppers}$), transformer core loss (P_{core}), snubber circuit loss (P_{snub}), output rectifier diode loss (P_{orec}), turn-on (P_{son}), turn-off (P_{soff}) and conduction losses (P_{scond}) of the MOSFET, NTC loss (P_{NTC}), input capacitor loss (P_{Cin}).

3.5.1 Input Diode Bridge Loss

Input diode bridge loss can be calculated using voltage drop of bridge rectifier and average input current.

$$P_{bridge} = 2 \times I_{inave} \times V_F = 2 \times 1.379 A \times 1 V = 2.758 W \quad (3.64)$$

3.5.2 Input Smoothing Capacitor Loss

The input capacitor loss is related to its ESR.

$$P_{Cin} = I_{Cinrms}^2 \times ESR \text{ (W)} \quad (3.65)$$

From the capacitor data book [18] the ESR is 0.35Ω . Since two capacitors are used in parallel, the ESR is $\frac{0.35\Omega}{2} = 0.175\Omega$ therefore:

$$P_{Cin} = (4.97A)^2 \times 0.175\Omega = 4.32W$$

3.5.3 Core Losses

Core losses are highly dependent on temperature, flux density and switching frequency. The relationship can be observed from the Figures 3.27. - 3.29.

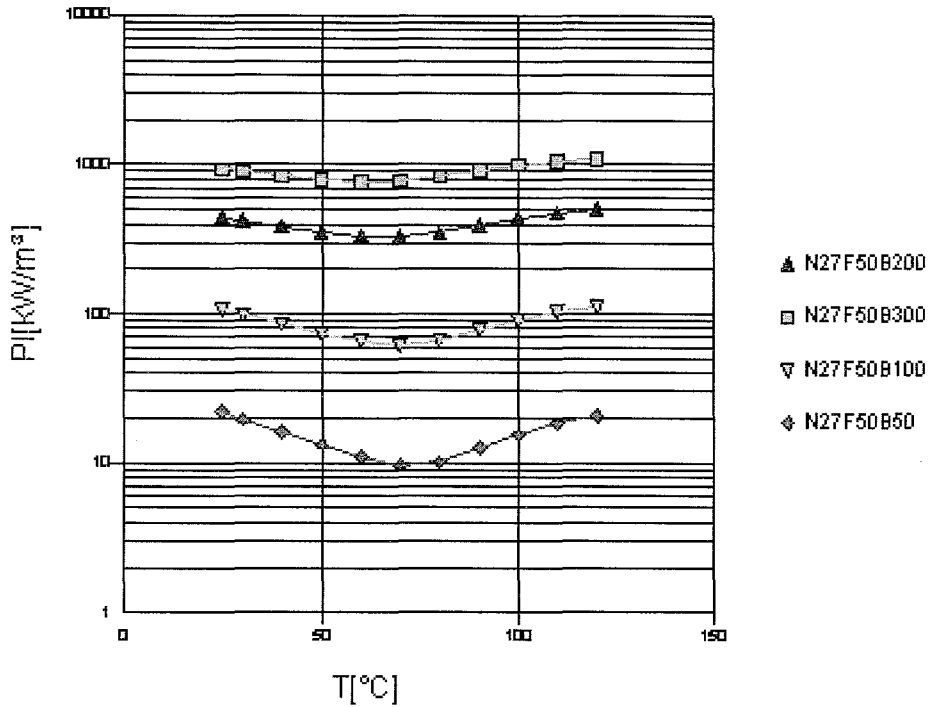


Figure 3.27. Core losses per volume (kW/m^3) for different switching frequency and flux densities for N27 material [11].

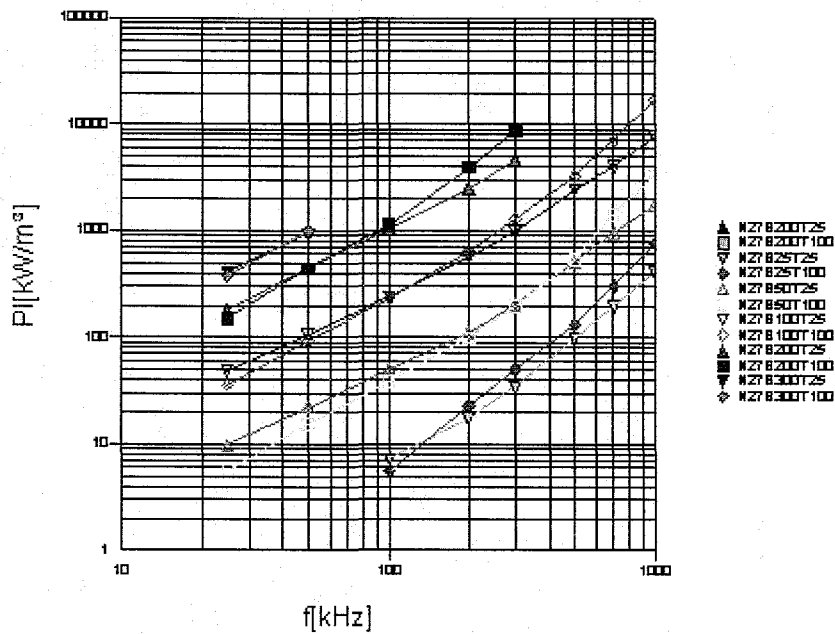


Figure 3.28. Core losses in kW/m^3 for N27 material at different temperatures and flux densities as a function of temperature [11].

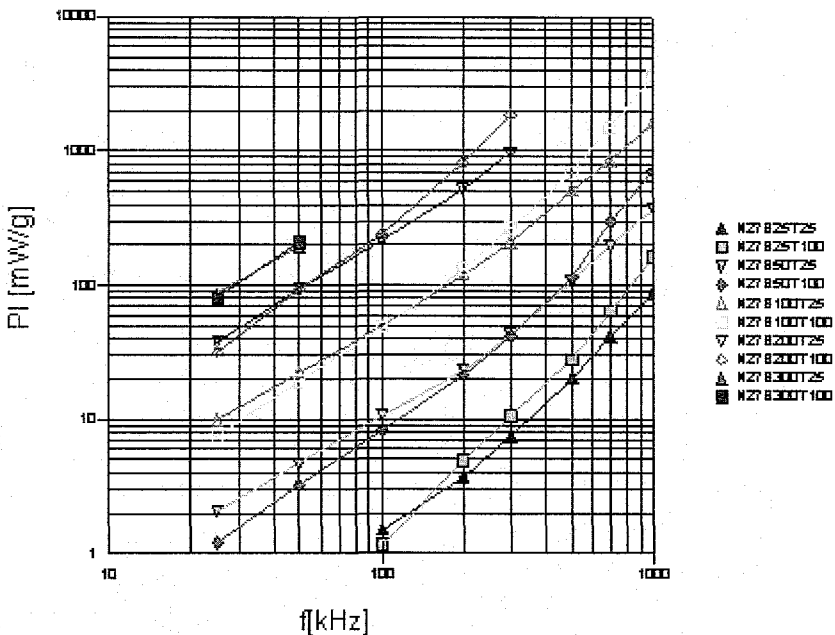


Figure 3.29. Core losses in mW/g at different temperatures and flux densities as a function of frequency [11].

Core losses in a transformer increases with the increasing flux density as shown in Fig. 3.28. If an air gap is used the core losses are reduced. The core loss with an air gap is:

$$P'_{core} = P_{core} \times \frac{\mu_e}{\mu_i} \text{ (W)} \quad (3.66)$$

Here μ_i is the initial permeability without an air gap (measured on a ring core) and μ_e is the effective permeability with an air gap introduced. μ_i is 2000 for N27 material. Effective permeability can be found from [11] or can be calculated using Eq. 3.67.

$$\mu_e = \frac{A_i \times \sum \frac{l_e}{A_e}}{\mu_0} \quad (3.67)$$

l_e is the effective magnetic path length, A_e is the effective magnetic cross-sectional area and for E55/21 core $l_e = 12.54\text{cm}$, $A_e = 3.54\text{cm}^2$. Thus,

$$\mu_e = \frac{496 \times 10^{-9} \text{ H} \times \frac{12.4\text{cm}}{3.54\text{cm}^2}}{12.57 \times 10^{-9} \text{ H/cm}} = 138$$

The core loss reduces to:

$$P'_{core} = P_{core} \times \frac{138}{2000} = 0.069 \times P_{core} \text{ (W)} \quad (3.68)$$

P_{core} can be determined by using the maximum flux density, switching frequency and temperature from Fig. 3.29. For N27 material at 50kHz, at a maximum temperature of 120°C and at maximum flux density of 200mT core loss per unit gram is between 106,53mW/g. E55/21 core is 215g per set. So the power loss per set is 22,9W per set. The final core loss is then:

$$P'_{core} = 0.069 \times (22.9\text{W}) = 1.58\text{W}$$

3.5.4 Primary and Secondary Winding Copper Loss

Copper losses occur due to the resistance of the wire used. The wire resistance can be calculated using unit resistance per length of the wire used taken from the wire table [14]. We used three 0.8mm wires in parallel for primary winding so the resistance drops by 3. The resistance per unit length for 0.8mm wire is $34\Omega/\text{km}$. For the secondary wire of 0.45mm diameter the resistance per unit length is $108\Omega/\text{km}$. These values are for 20°C .

$$P_{copperp} = I_{p\max}^2 \times \frac{D_{\max}}{3} \times R_{Cup} \text{ (W)} \quad (3.69)$$

R_{Cup} is the resistance of primary winding. We used three wires in parallel

$$R_{Cus} = \frac{N_p \times l_n}{3} \times R_{\text{per_unit_length}(0.8\text{mm_wire})} \text{ (}\Omega\text{)} \quad (3.70)$$

$$R_{Cup} = \frac{30 \times 0.113\text{m}}{3} \times 0.034\Omega/\text{m} = 0.03842\Omega$$

$$P_{copperp} = (5.63\text{A})^2 \times \frac{0.5}{3} \times 0.03842\Omega = 0.20297\text{W}$$

For the secondary winding copper loss:

$$R_{Cus} = \frac{N_s \times l_n}{3} \times R_{\text{per_unit_length}(0.45\text{mm_wire})} = \frac{130 \times 0.113\text{m}}{3} \times 0.108\Omega/\text{m} = 0.52884\Omega \quad (3.71)$$

$$P_{coppers} = I_{\text{peak}}^2 \times \frac{D_{\max}}{3} \times R_{Cus} = (1.2\text{A})^2 \times \frac{0.5}{3} \times 0.52884 = 0.1269\text{W} \quad (3.72)$$

Because the turns of the auxiliary and secondary output 2 is very small and so the winding resistances, copper losses are very small and can be neglected.

3.5.5 Snubber Losses

$$P_{\text{snub}} = \frac{1}{2} \times I_{p\max}^2 \times L_\sigma \times F_s \text{ (W)} \quad (3.73)$$

$$P_{\text{snub}} = \frac{1}{2} \times (5.63\text{A})^2 \times 12.4 \times 10^{-6} \text{H} \times 50000 = 9.82\text{W}$$

3.5.6 Output Rectifier Diode Loss

The output rectifier loss is related to the secondary terminal peak current,

$$P_{orec} = I_{speak} \times \sqrt{\frac{1-D_{\max}}{3}} \times V_F \text{ (W)} \quad (3.74)$$

$$P_{orec} = 1.2A \times \sqrt{\frac{1-0.5}{3}} \times 4.5V = 2.205W$$

3.5.7 MOSFET Turn-on Loss

The turn-on loss of a MOSFET is related with the total output capacitance appears across drain source terminals. The output capacitance of the MOSFET is the sum of MOSFET output capacitance C_{oss} and the resonant capacitance C_{res} so the total resonance capacitance becomes:

$$C_r = C_{oss} + C_{res} \text{ (F)} \quad (3.75)$$

$$C_r = 275 \times 10^{-12} \text{F} + 220 \times 10^{-12} \text{F} = 495 \times 10^{-12} \text{F}$$

$$P_{son} = \frac{1}{2} \times C_r \times (V_{p\max} + V_r)^2 \times F_s \text{ (W)} \quad (3.76)$$

$$P_{son} = \frac{1}{2} \times 495 \times 10^{-12} \text{F} \times (357.8V + 231V)^2 \times 50000 \text{Hz} = 4.29W$$

3.5.8 MOSFET Turn-off Loss

The turn-off losses are related to the MOSFETs turn-off fall time T_r .

$$P_{soff} = \frac{1}{6} \times (V_{p\min} + V_r) \times I_{p\max} \times F_s \times t_r \text{ (W)} \quad (3.77)$$

$$P_{soff} = \frac{1}{6} \times (251.3V + 231V) \times 5.63A \times 50000 \text{Hz} \times 1.1 \times 10^{-7} \text{s} = 2.489W$$

3.5.9 MOSFET Conduction Loss

MOSFET conduction losses are related with the on state resistance $R_{DS(on)}$ and the RMS current passing through it.

$$P_{scond} = I_{prms}^2 \times R_{DS(on)} = \frac{1}{3} \times I_{pmax}^2 \times R_{DS(on)} \times D_{max} \text{ (W)} \quad (3.78)$$

$$P_{scond} = \frac{1}{3} \times (5.63A)^2 \times 0.75\Omega \times 0.5 = 3.962W$$

3.5.10 NTC Resistor Loss

At turn-on, while the NTC resistor is not warm enough the power loss will be very high because of the inrush current. After warm-up the resistance of the NTC drops to a low value. For the one we have chosen the normal resistance is 5Ω and at the rated current the resistance drops to 0.128Ω . If we consider the operating conditions the power loss can be calculated as below:

$$P_{NTC} = I_{Rinrms}^2 \times R_{NTC} = (4.924A)^2 \times 0.128\Omega = 3.103W \quad (3.79)$$

3.5.11 Total Power Loss and Efficiency

Total power loss is therefore:

$$P_{losstot} = P_{bridge} + P_{Cin} + P'_{core} + P_{copperp} + P_{coppers} + P_{snub} + P_{orec} + P_{son} + P_{soff} + P_{scond} + P_{NTC} \text{ (W)} \quad (3.80)$$

$$P_{losstot} = 2.758W + 4.32W + 1.58W + 0.203W + 0.1269W + 9.82W + 2.205W + 4.29W + 2.489W + 3.962W + 3.103W = 34.8569W$$

The efficiency is (without considering the power factor):

$$\eta = \frac{P_{omax}}{P_{omax} + P_{Plosstot}} = \frac{300W}{300W + 34.8569} \times 100 = 89.59\% \quad (3.81)$$

$$\eta = \frac{300W}{300W + 34.8569} \times 100 = 89.59\%$$

3.6 Thermal Design

3.6.1 Heat sink thermal design

The analogy between the electrical and thermal circuit is shown in the Fig. 3.30.

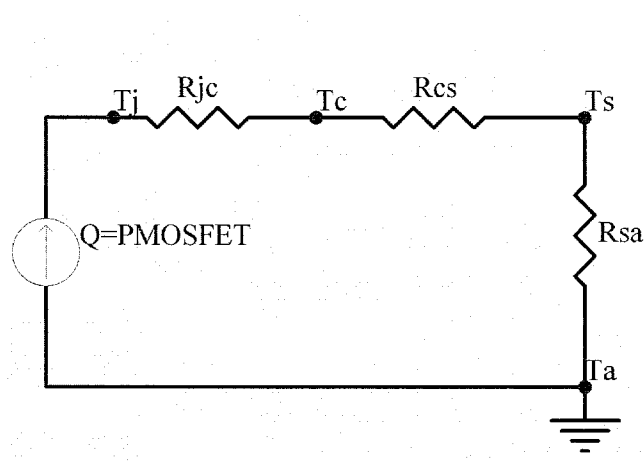


Figure 3.30. Analogy between electrical and thermal circuits.

Table 3.4.
Analogy between electrical and thermal circuits

Electrical Circuit	Thermal Circuit
V (Voltage in Volts)	T _{1,2} (Temperature in °C)
I (Current in Amperes)	Q (Thermal power in Watts)
R (Resistance in Ω)	R _θ (Thermal resistance in °C/W)

The thermal design is based on the analogy of temperature to current and thermal resistances to omic resistances. The maximum allowable junction-to-ambient thermal resistance $R_{\theta ja}$ is;

$$R_{\theta ja} = \frac{(T_j - T_a)}{P_{Mosfet}} \left(\frac{^{\circ}C}{W} \right) \tag{3.82}$$

The junction-to-ambient thermal resistance $R_{\theta ja}$ is also equals to;

$$R_{\theta ja} = R_{\theta jc} + R_{\theta cs} + R_{\theta sa} \left(\frac{^{\circ}C}{W} \right) \tag{3.83}$$

The thermal resistance of the heat sink required for the MOSFET can be combining these two equations with Eq. 3.84.

$$R_{\theta sa} = \frac{(T_j - T_a)}{P_{Mosfet}} - (R_{\theta cs} + R_{\theta jc}) \left(\frac{^{\circ}C}{W} \right) \quad (3.84)$$

The design is based on the worst case condition, so we used the maximum values. Here T_a is the maximum ambient temperature ($50^{\circ}C$ in our design), T_j is maximum operating temperature of MOSFET ($150^{\circ}C$ for Fairchild FQA13N80 MOSFET), $R_{\theta cs}$ is MOSFET's case to sink thermal resistance ($0.24^{\circ}C/W$) and $R_{\theta jc}$ is the MOSFET's junction to case thermal resistance ($0.42^{\circ}C/W$). P_{Mosfet} is the total power loss in the switching MOSFET. So the thermal resistance of the heat sink becomes:

$$R_{\theta sa} = \frac{(150^{\circ}C - 50^{\circ}C)}{(4.29W + 2.489W + 3.962W)} - (0.24^{\circ}C/W + 0.42^{\circ}C/W) = 8.65^{\circ}C/W$$

In the first design we used a small heat sink which has a greater thermal resistance than this and we saw that it was excessively warm during operation and as a result of this a few MOSFETs blew up. So in the second design we choose a heat sink with a lower thermal resistance.

This is the minimum required heat sink thermal impedance. The on state resistance of a MOSFET is dependent on temperature it will rise with rising temperature. Therefore the conduction losses will become excessive if inadequate heat sinks are used; the actual required thermal resistance of the heat sink should be much lower than the calculated value. For FQA13N80 $R_{DS(on)}$ @ $150^{\circ}C$ is about 2.75Ω which means four times greater conduction loss in the MOSFET and the thermal resistance should be as low as $4^{\circ}C/W$. From data books we have chosen a suitable heat sink that has a thermal resistance about $2.3^{\circ}C/W$ allowing some safety margin.

3.6.1 Fan Requirement

If the operating temperature of the power components becomes higher than the specified maximum operating temperature and heat sink can not provide good cooling a

fan can be should be used with these heat sinks to cool the devices. The heat sink we use is sufficient to cool the MOSFET to operating temperatures but if we use power components different from our design (for example power MOSFET with a higher $R_{DS(on)}$) the power dissipation will increase and the heat sink may not provide good cooling. In order to prevent this we provided an additional 12V winding at the output for a fan. This winding does not exist in the first version of the supply.

The most common fan types are as follows [14]:

- Axial fans, the most widely used in electronics applications. They are available in a range of sizes for various requirements in terms of air throughput, air pressure, sound levels, etc.
- Radial fans are used in applications requiring greater air pressure at a given air quantity. They are noisier than comparable axial fans.
- Tangential fans operate very quietly with a given air quantity. They have the disadvantage that they are physically very large and that the air pressure is low.

Selection of fans can be done with Eq. 3.85 [14].

At a given heat loss the air throughput needed V_{air} is:

$$V_{air} = 3.0 \times \frac{P_{losstot}}{(T_{a\max} - T_a)} \text{ (m}^3/\text{h)} \quad (3.85)$$

Here V_{air} is the air throughput in m^3/h , $P_{losstot}$ is the total power loss, T_a is the ambient temperature in $^{\circ}\text{C}$ and $T_{a\max}$ is the maximum permitted temperature in $^{\circ}\text{C}$ inside the package. If we use a maximum permitted temperature inside the package as 60°C the air throughput requirement is;

$$V_{air} = 3.0 \times \frac{35.2055\text{W}}{(60^{\circ}\text{C} - 50^{\circ}\text{C})} = 9.56 \text{ m}^3/\text{h}$$

A suitable fan with this air throughput can be selected using a fan catalog [14]. We selected a $12V\ 0.14A$ axial fan. As mentioned above if the selected devices and an adequate heat sink is used there is no need for a fan. However if a smaller heat sink or another MOSFET with a greater on-state resistance with different thermal characteristics is used calculations should be repeated in order to get the desired temperature rise in the enclosure. If we use a fan with our PCB board the fan and so the SMPS will work continuously. An electronic thermostat like National's LM26/LM27 can be used to control the fan speed. The LM26/LM27 consists of a temperature sensor and a voltage reference. The output drive is hysteretic and the fan is not fully turned on or turned off. Instead the fan speed is adjusted/programmed to predetermined levels so the startup noise when a temperature increase occurs is eliminated.

3.7 Calculation of External Circuit Elements

In this section external circuit element calculations for the TDA16846 are presented.

3.7.1 Pin 2 - Primary Current Simulation PCS

A voltage proportional to the current of the power transistor is generated at pin 2 by the RC combination R_{pin2} , C_{pin2} . Resistor with capacitor connected to this pin determines the peak permissible current flowing through the primary winding inductance and so the maximum power that can be delivered [13]. We choose pin 2 resistor to be $R_{pin2} = 1 \times 10^6 \Omega$. The capacitor at pin 2 can be determined with the known values of peak primary current, primary inductance and R_{pin2} as below:

$$C_{pin2} = \frac{V_{pmin} \times t_{onmax}}{R_2 \times (V_{pin2} - 1.5V)} \quad (F) \quad (3.86)$$

V_{pin2} is the voltage at pin 2. The maximum value of V_{pin2} is the control voltage. If the control voltage is not lowered by feedback the maximum value will be 5V.

$$C_2 = \frac{251.3V \times \frac{0.5}{50000Hz}}{10^6 \Omega \times (5V - 1.5V)} = 718 \times 10^{-12} F$$

We can choose 680pF or 820pF from the standard values. We choose 820pF for our application.

The maximum current that can flow through the primary inductance from [23]

$$I_{p\max} = \frac{3.5V \times R_{pin2} \times C_{pin2}}{L_p} \text{ (A)} \quad (3.87)$$

$$I_{p\max} = \frac{3.5V \times 10^6 \Omega \times 820 \times 10^{-12} F}{446 \times 10^{-6} H} = 6.43 A$$

The calculated maximum primary current is 5,63A and higher than the above value. So the capacitor value can be increased slightly to the next value of 820pF. If the voltage at pin 2 exceeds the control voltage (voltage at pin 4) then the driver switches off.

3.7.2 Pin 4 - Soft Start and Regulation Capacitor SRC

This is the pin for the control voltage. The control voltage is kept on a capacitor connected on this pin (C_{pin4}). The value of this capacitor is between 1.5 and 10nF and determines the duration of the soft start and the speed of the control especially in primary regulation. In the absence of an opto coupler feedback the voltage on this pin is the control voltage. Control voltage also determines when the transistor will be switched on.

The control voltage can be reduced either by internal error amplifier or by opto coupler feedback or by the voltage at pin 11 (Fold back point correction).

If there is much noise at pin 4 higher capacitance value can be used for C_{pin4} .

Table 3.5.
Control voltage versus output power

Control Voltage	Output Power	Off-Time $T_{D_{pin1}}$
1.5 – 2V	Low	Constant ($T_{D_{pin1\max}}$), constant frequency stand by
2 – 3,5V	Medium	Decreasing
3,5V – 5V	High	Free running, switch-on at first minimum

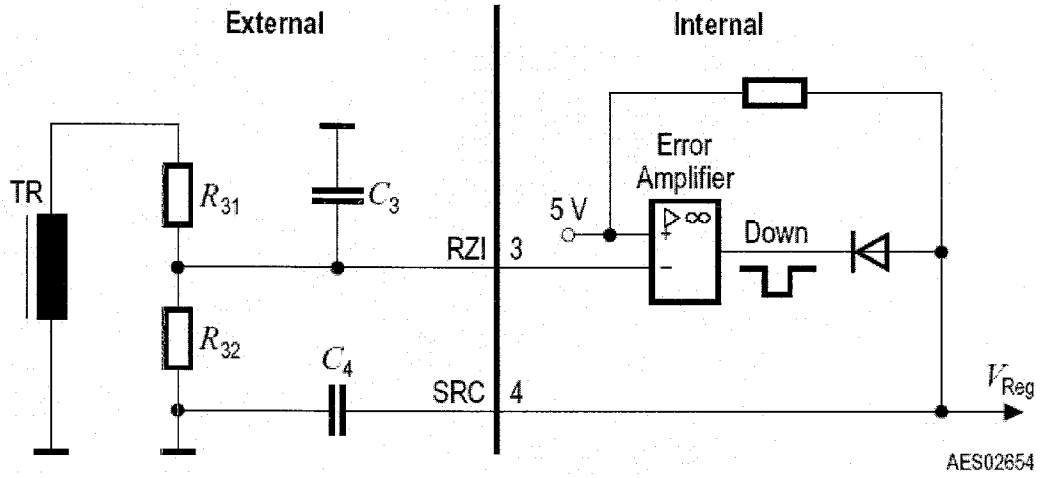


Figure 3.31. Internal error amplifier circuitry [23].

3.7.3 Pin 1 - Off-Time circuit OTC

In free-running mode switching frequency of the converter is determined by the load current/power. So there is an internal OTC which adjusts the off-time of the switching cycle. A parallel RC circuit is connected to this pin. C_{pin1} is charged by an internal current source of $10^{-3} A$ until the capacitor's voltage reaches 3.5V. The charge time is:

$$T_{C_{pin1}} = \frac{C_{pin1} \times 1.5V}{10^{-3} A} \quad (3.88)$$

Here $T_{C_{pin1}}$ is called the ringing suppression time. So for proper operation of the anti jitter circuit $T_{C_{pin1}}$ should have the same value as the resonance time (T_r) of the power switch. In our circuit the resonance time T_r is calculated as follows:

$$T_r = \frac{1}{F_r} = 2 \times \pi \times \sqrt{L_p \times C_r} \quad (s) \quad (3.89)$$

$$T_r = 2 \times \pi \times \sqrt{446 \times 10^{-6} H \times 495 \times 10^{-12} F} = 2.95 \times 10^{-6} s \text{ so,}$$

$$C_{pin1} = \frac{10^{-3} A \times 2.95 \times 10^{-6} s}{1.5V} = 1.97 \times 10^{-9} F$$

We can take C_{pin1} as 1.8nF or 2.2nF from standard values. We selected 2.2nF.

After charging C_{pin1} up to 3.5V the current source is disconnected and C_{pin1} is discharged by R_{pin1} . If the control voltage is between 3.5 and 5V the converter is operating at high power level and the MOSFET will be switched on at the first zero crossing. If it is between 2 and 3.5V the converter is operating at medium power level and as soon as the pin 1 voltage drops below control voltage the controller will wait for a zero crossing signal. All the zero crossings before this will be neglected. As soon as a zero crossing signal reaches the controller via pin 3 circuit the MOSFET will be switched on. If no zero crossing signal is coming to pin 3 the MOSFET is switched on after an additional delay (when pin 1 voltage drops below 1.5V).

If the control voltage is below 2V the off-time is at maximum and T_{D1} is also maximum and equal to:

$$T_{D1} = 0.47 \times R_{pin1} \times C_{pin1} \quad (3.90)$$

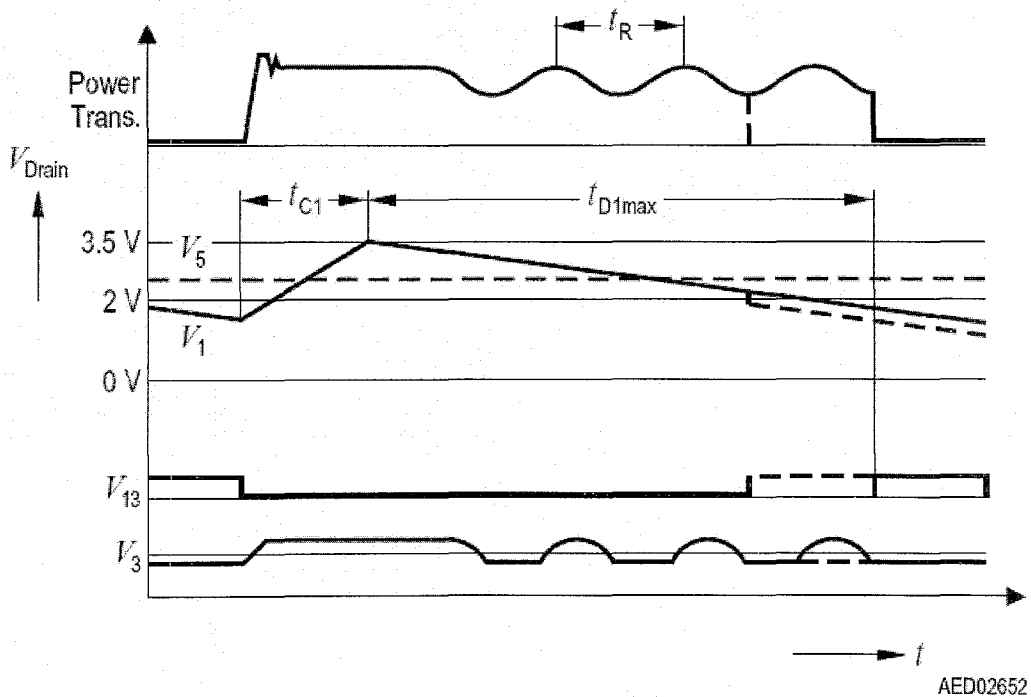


Figure 3.32. Pulse diagram of off-time circuit [23].

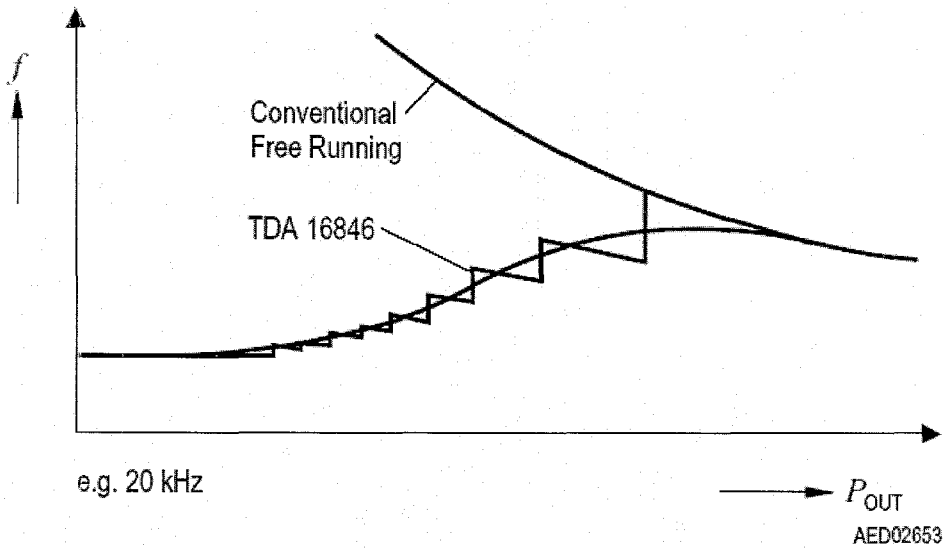


Figure 3.33. Load dependent frequency course [23].

3.7.4 Pin 3 - Regulation and Zero Crossing Input RZI

This is the input of internal error amplifier circuit and zero crossing detector. The voltage from the auxiliary winding of the transformer is divided by a resistive voltage divider and is smoothed by a capacitor to ground. The value of this capacitor is extremely important and requires trial and error. If it is a wrong value the SMPS may not be able to start. If the pulses at the pin 3 exceed a 5V level the control voltage at pin 4 is lowered. Dimensioning of the pin 3 elements is as below [24]:

$$R_{\text{gin3}_1} = \frac{V_{p\min} \times N_{\text{aux}}}{N_p \times 0.5 \times 10^{-3} \text{ A}} = \frac{251.3 \text{ V} \times 2}{30 \times 0.5 \times 10^{-3} \text{ A}} = 33506 \Omega \quad (3.91)$$

We select the closest value of $33 \times 10^3 \Omega$.

$$R_{\text{pin3}_2} = \frac{R_{\text{pin3}_1}}{\left(\frac{V_{\text{regtr}}}{5 \text{ V}} - 1 \right)} (\Omega) \quad (3.92)$$

V_{regtr} is the induced positive voltage at the regulation winding (auxiliary winding), e.g. 15V.

$$R_{pin3_2} = \frac{33 \times 10^3 \Omega}{\left(\frac{15V}{5V} - 1\right)} = 16.5 \times 10^3 \Omega$$

We select $15 \times 10^3 \Omega$. In order to adjust the divider ratio precisely we placed a $10 \times 10^3 \Omega$ potentiometer (R_{pin3_3}) in series with R_{pin3_2} .

The pin 3 capacitor C_{pin3} smoothes the voltage shape at pin 3. After dimensioning C_{pin3} , in primary regulation mode in order to get the best load regulation and the best switch on time for the MOSFET C_{pin3} is adjusted. Dimensioning is as below:

$$C_3 = \frac{1000 \times 10^{-9} s}{R_{pin3_1}} = \frac{1000 \times 10^{-9} s}{33 \times 10^3 \Omega} = 30.3 \times 10^{-12} F \quad (3.93)$$

We selected $33 \times 10^{-12} F$. We also used different values of C_{pin3} and found that the best is $22 \times 10^{-12} F$.

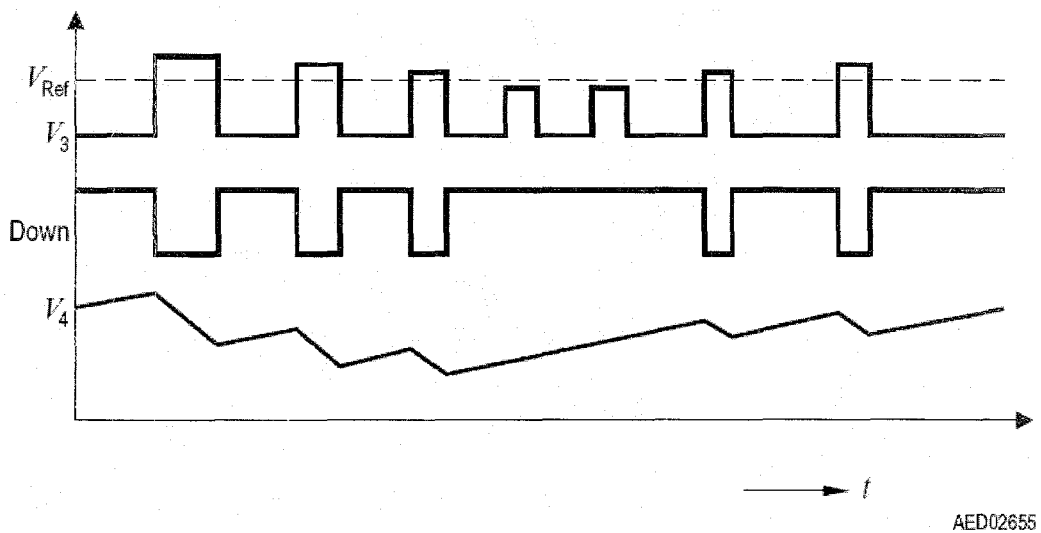


Figure 3.34. Regulation voltage (Pin 4).

3.7.5 Pin 5 - Opto Coupler Input OCI

When the converter is in primary regulation mode this pin is not used and it is at the internal reference voltage level. When opto coupler feedback is used the collector of the

opto coupler transistor is connected to this pin and the pulses at pin 3 should be adjusted to be between 2.5 and 5.0 V. So the pin 3 acts only as a zero crossing input and an additional over voltage protector. Pin 5 should not be pulled down to ground and a minimum of 1.5V should exist. A resistor of $1.8 \times 10^3 \Omega$ between emitter of the opto coupler and ground ensures this voltage. In the second version of TDA16846 control IC it is reported that this resistor is not necessary but because we have the first version of this IC we use this resistor.

We placed a jumper close to pin 5 and the operation mode can be selected as either primary regulation or secondary regulation.

3.7.6 Pin 7 Fixed frequency and Synchronization Pin

Control IC can work in either fixed frequency or free-running mode. If fixed frequency is wanted a parallel RC network is connected to this pin. We designed the supply to work in either mode and the operation mode can be selected using 2 jumpers. The circuit elements for fixed frequency mode are chosen with Eqs. 3.94 – 3.95.

$$F_s = \frac{1.18}{R_{pin7} \times C_{pin7}} \text{ (Hz)} \quad (3.94)$$

$$C_{pin7} = \frac{1.18}{F_s \times R_{pin7}} \text{ (F)} \quad (3.95)$$

If we choose $R_{pin7} = 56 \times 10^3 \Omega$ and $F_s = 50 \times 10^3 \text{ Hz}$ the capacitor C_{pin7} will be:

$$C_{pin7} = \frac{1.18}{50 \times 10^3 \times 56 \times 10^3} = 421 \times 10^{-12} \text{ F}$$

We choose the closest value of $470 \times 10^{-12} \text{ F}$. So the actual switching frequency is 44832 Hz .

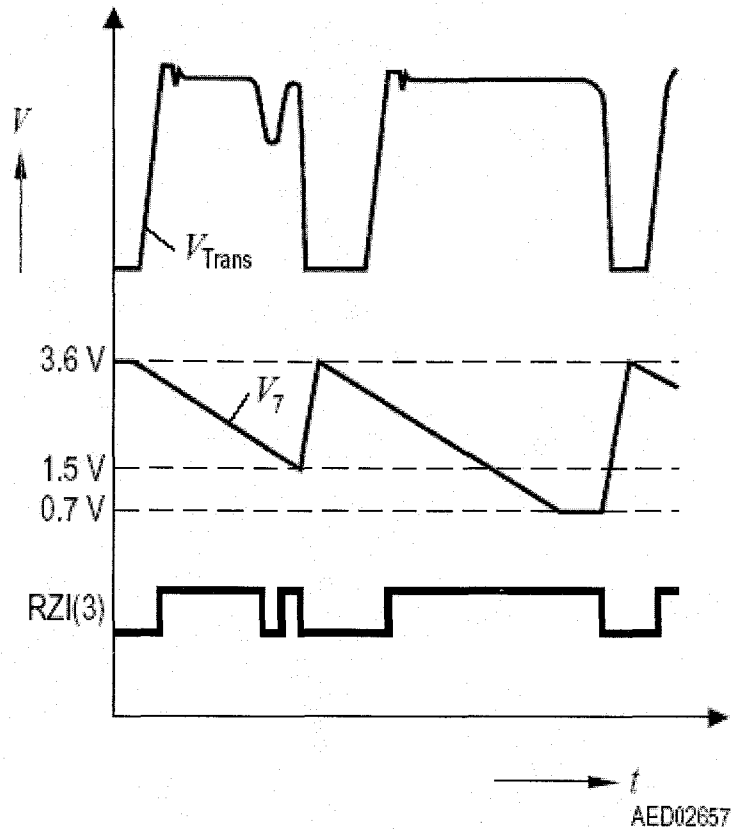


Figure 3.35. Pulse diagram for fixed frequency circuit [23].

3.7.7 Pin 6, 10, 12 - Ground Pins

These pins are connected to ground.

3.7.8 Pin 13 - Gate drive pin

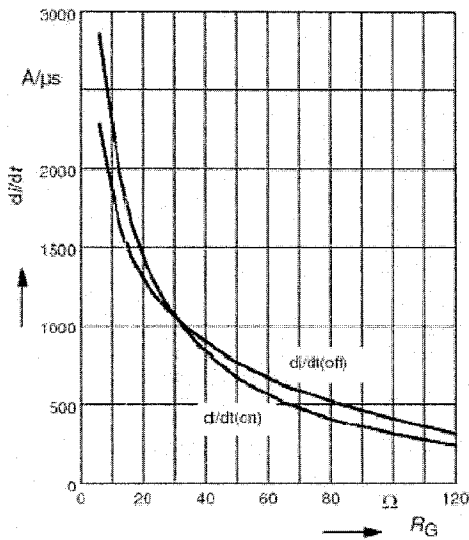
TDA16846 can drive a MOSFET without the need of an external gate drive circuit. If we want transistor to turn-off quickly we can connect a diode, for example 1N4148, in parallel with the gate resistor. The gate resistor should be in the range between 2.2-22Ω. Higher gate resistor means higher power loss in the transistor but it also prevents RF noise that is caused by fast switching time of MOSFET to reach into primary voltage check pin (pin 11), which is very sensitive to noise spikes [25], [26]. According to [25] gate resistance, gate-source and drain-source voltage and drain current (continuous or discontinuous) have influence on EMI spectrum.

Gate resistance can be used to control the switching speed and therefore the EMI behavior is a variation of the gate resistance. As the resistance increases the time constant of this resistor and the MOSFET capacitance is increased which will slow down the

switching transient and eventually reduce the electrical noise (by controlling di/dt and dV/dt).

17 Typ. drain current slope

$dI/dt = f(R_G)$, inductive load, $T_j = 125^\circ\text{C}$
par.: $V_{DS}=380\text{V}$, $V_{GS}=0/+13\text{V}$, $I_D=11\text{A}$



18 Typ. drain source voltage slope

$dV/dt = f(R_G)$, inductive load, $T_j = 125^\circ\text{C}$
par.: $V_{DS}=380\text{V}$, $V_{GS}=0/+13\text{V}$, $I_D=11\text{A}$

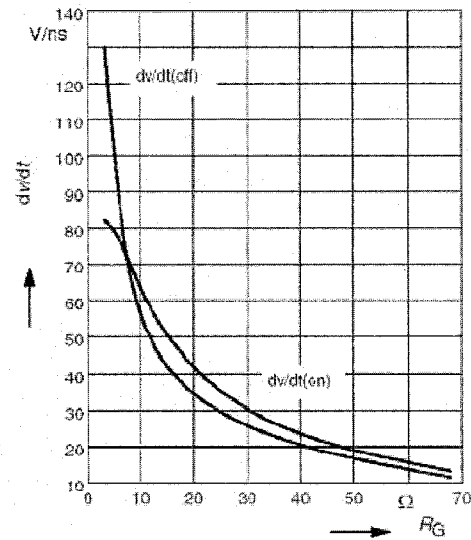


Figure 3.36. Typical switching slopes versus gate resistance of CoolMOS™ SPPN11N60C3 (datasheet information) [25].

In order to prevent TDA16846 IC from being destructed in case of a failure a zener diode is connected between gate and ground terminals. In the absence of the zener we observed that IC blows with the blowing MOSFET, but when we used a zener and when the MOSFET blew the IC was not destructed.

We chose 15Ω gate resistor, an anti parallel 1N4148 diode and a 15V zener diode.

3.7.9 Pin 14 - Supply Voltage Pin

An electrolytic capacitor (C_{pin14}) in the range of $1\mu\text{F}$ to $100\mu\text{F}$ connected to this pin supplies the operating voltage. This capacitor is connected to the auxiliary winding of the transformer with a 1N4148 diode. A RF suppression capacitor (C_{pin14_2}) of 100nF is connected across this pin. The voltage at this pin enters the supply voltage comparator SVC and the other input is 15V. When the power applied to the chip C_{pin14} is charged to

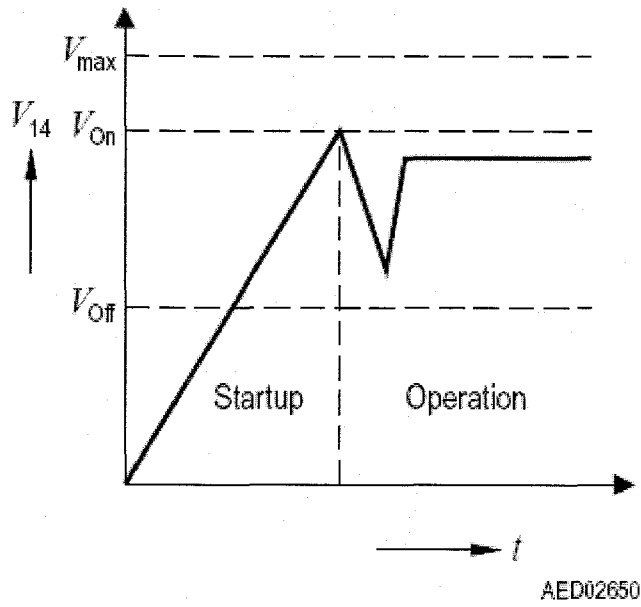


Figure 3.38. Startup voltage diagram [23].

3.7.10 Pin 11 - Primary Voltage Check and Fold Back Point Correction PVC

The voltage at this pin is deviated from the rectified mains and reduces the limit of the possible current maximum in the power transistor if the mains voltage increases. This limit is independent of the mains (only active in free running mode). The possible maximum current in the primary winding determined by this pin is:

$$I_{p\max} = \frac{\left(4V - \frac{V_{pin11}}{3}\right) \times R_{pin2} \times C_{pin2}}{L_p} \quad (\text{A}) \quad (3.96)$$

The voltage at pin 11 is:

$$V_{pin11} = \frac{R_{pin11_1} \times V_p}{R_{pin11_1} + R_{pin11_2}} \quad (\text{V}) \quad (3.97)$$

Taking $R_{pin11_2} = 3.9 \times 10^6 \Omega$ and $R_{pin11_1} = 18 \times 10^3 \Omega$ and we calculate for the minimum and maximum values of V_p , V_{pin11} is:

$$V_{pin11_low} = \frac{18 \times 10^3 \Omega \times 251.3V}{18 \times 10^3 \Omega + 3.9 \times 10^6 \Omega} = 1.15V$$

$$V_{pin11_high} = \frac{18 \times 10^3 \Omega \times 357.8V}{18 \times 10^3 \Omega + 3.9 \times 10^6 \Omega} = 1.64V$$

If there are problems with this pin R_{pin11_1} can be increased to prevent chip from shutdown because of mains under voltage. This pin is very sensitive to RFI noise so a RFI suppression capacitor in the range of 10 to 100nF can be placed across it. We used 100nF.

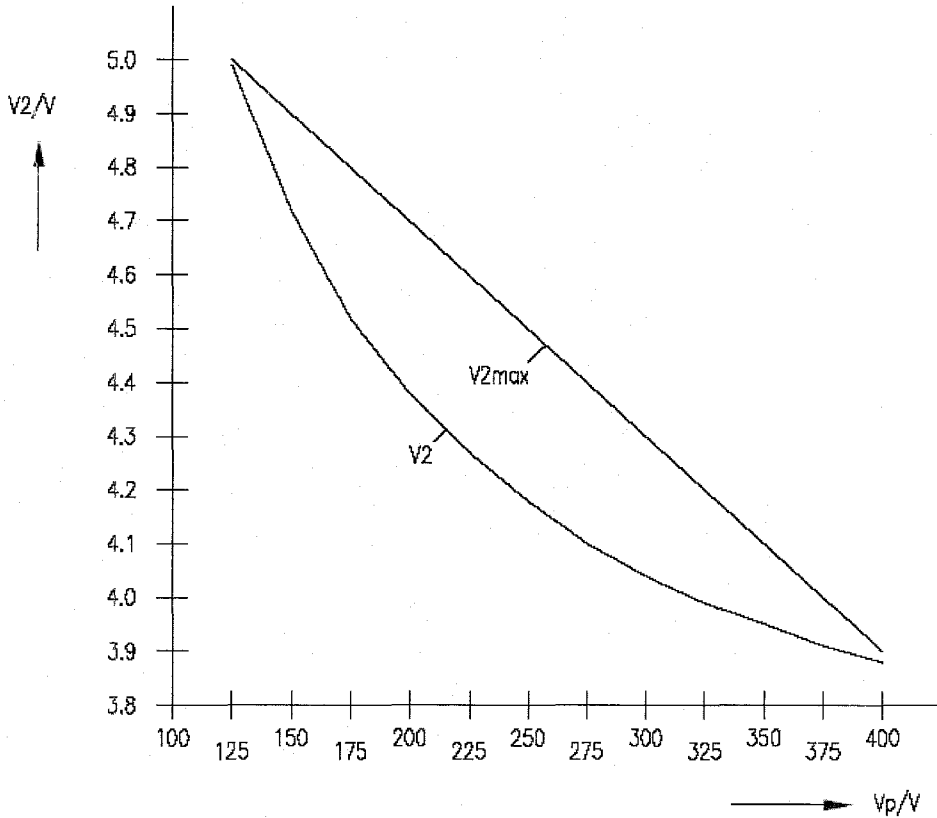


Figure 3.39. Function of the fold back point correction (input voltage versus pin 2 voltage) [13].

3.8 PCB Design

The maximum current that can flow through foil conductors depends on the thickness of the copper layer on PCB material and the allowed conductor temperature. There are three standard copper thicknesses (17.5, 35 and 70μm). In power supply design because of the high current capability the 70μm copper thickness is used. We used 35μm copper thickness because we could not obtain 70μm copper thickness PCB material. As the copper thickness increases the current carrying capacity increases. To obtain the same

current density we made some track widths wider. The minimum conductor width that can be produced in photographic exposure is 0.3mm.

Table 3.6.
Maximum current in a through foil conductors on PCB with 17.5µm foil [14]

Conductor Width (mm)	Allowable temperature in conductor					
	10 °C	20 °C	30 °C	60 °C	75 °C	100 °C
0.5	0.6	1	1.2	1.7	2	2,3
1.0	1.1	1.5	2	3	3,2	3,7
1.5	2	2,6	3,4	4,3	5	6
2	2,3	3,2	4	5	6	7
4	4	5	7	9	10	11
6	5	7	9	12	13	14
8	6	9	11	14	16	18
10	7	10	13	16	19	21

Table 3.7.
Maximum current in a through foil conductors on PCB with 35µm foil [14]

Conductor Width (mm)	Allowable temperature in conductor					
	10 °C	20 °C	30 °C	60 °C	75 °C	100 °C
0.5	1.3	2	2,3	3	3,5	4
1.0	2	2,8	3,1	4	5	6
1.5	2,6	3,7	4,4	6	7	8
2	3,2	5	6	8	9	10
4	5,5	8	10	11	15	16,5
6	8	11	13	18	21	23
8	9,5	13	16	22	24	26
10	11	16	20	27	29	33

Table 3.8.
Maximum current in a through foil conductors on PCB with 70µm foil [14]

Conductor Width (mm)	Allowable temperature in conductor					
	10 °C	20 °C	30 °C	60 °C	75 °C	100 °C
0.5	2,4	3,2	4	5	6	7
1.0	3,3	4,5	6	8	9	10
1.5	4,3	6	8	10	12	13
2	5	8	10	13	14	15
4	9	13	15	21	23	25
6	12	16	22	30	32	35

Table 3.9.
Minimum allowable distance between conductors and insulation distance depending on the maximum voltage on the conductors [14]

Voltage (V) between conductors	50	150	300	500
Min. distance (mm) between conductors	0.3	0.6	1.2	1.8

The minimum allowable distance between conductors and the insulation distance is dependent on production method and maximum voltage. We used minimum 4mm between 1000V tracks that provided us sufficient margin. Also applying a dielectric material on all track sides provides extra insulation.

4. QUASI RESONANT CONVERTER OPERATION: FREQUENCY PREDICTION AND DYNAMIC RESPONSE

4.1 Introduction

The steady state response of a converter is important. But another parameter that is also important is the converter's dynamic response. There will be disturbances in line and load. If we permit these disturbances to reach at the output the output voltage of the converter may become instable and if the regulation loop is not designed well the converter may oscillate. In order to obtain good line and load regulation we should know the converter transfer function and then design the required regulator for the converter. There are two ways to determine the frequency response of a converter. First and the most accurate way is to use a frequency response analyzer. With this device an AC signal is injected into the converter feedback circuit and the variations in the output voltage is recorded. The second method is the use of a mathematical model that describes the system behavior. Because we do not have available measurement setup we used mathematical methods to plot the transfer function of the converter. The converter operation is nonlinear so we used mathematical method to linearize the nonlinear operation of the converter. In control engineering a normal operation of the system may be around an equilibrium point, and the signals may be considered small signals around the equilibrium. If the system operates around an equilibrium point and if the signals involved are small signals, then it is possible to approximate the nonlinear system by a linear system. This linear system is equivalent to the nonlinear system considered within a limited operating range. The linearization procedure is based on the expansion of nonlinear function into a Taylor series about the operating point and the retention of only the linear term. The higher order terms are neglected assuming that these terms are small enough so the variables deviate slightly from the operating condition [27].

4.2 Quasi Resonant Converter Operation and Switching Frequency Prediction

In QR converters the switching frequency is dependent on load conditions. So the switching frequency determination is a problem. In order to predict the switching

frequency we divide the circuit operation into 4 subdivisions and then determine switching times dependent on the primary current. So the final switching frequency becomes dependent on the primary peak current. The solution to this frequency is done with the output power relation.

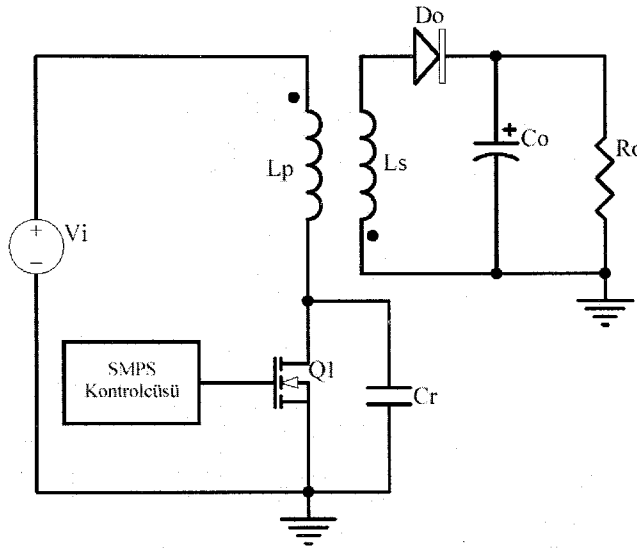


Figure 4.1. Simplified circuit scheme of a flyback converter.

$t_0 < t < t_1$ time period (T_{on} – MOSFET conduction time)

At $t = t_0$ switch is driven and the current increases linearly through the primary inductance. Current through the primary winding inductance and the voltage across the switch is shown as:

$$I_p(t) = I_p(t_0) + \frac{V_p \times t}{L_p} \quad (\text{A}) \quad (4.1)$$

$$V_d(t) = I_p(t) \times R_{ds(on)} \quad (\text{V}) \quad (4.2)$$

At the end of the conduction time T_{on} (at $t = t_1$) current reaches its maximum:

$$I_p(t_1) = I_{p\max} = I_p(t_0) + \frac{V_p \times T_{on}}{L_p} \quad (\text{A}) \quad (4.3)$$

$$V_d(t_1) = I_p(t_1) \times R_{ds(on)} \quad (V) \quad (4.4)$$

And on-time is:

$$T_{on} = \frac{L_p}{V_p} \times (I_p(t_1) - I_p(t_0)) \quad (s) \quad (4.5)$$

In our design the converter operation mode is DCM, so the initial current at $t = t_0$ is zero ($I_p(t_0) = 0$).

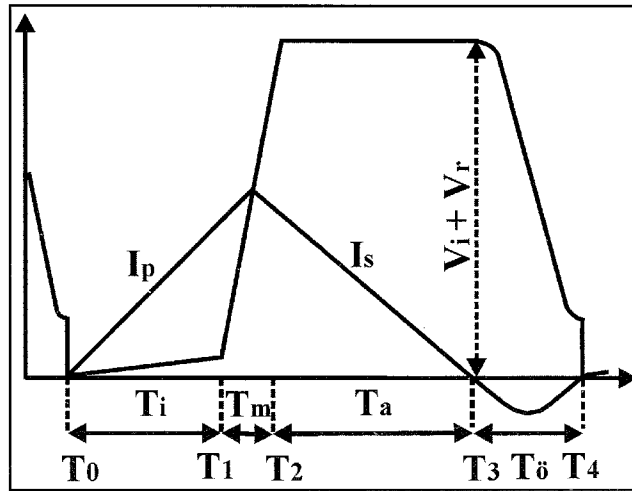


Figure 4.2. Drain voltage (V_d), primary (I_p) and secondary (I_s) winding currents.

$t_1 < t < t_2$ time period (T_{com} - Commutation time)

At $t = t_1$ the switch opens but the current through an inductor can not go zero at zero time, and this current starts to charge the capacitances across the switch. The total capacitance across the switch is the sum of MOSFET output capacitance (C_{oss}), resonance capacitance (C_{res}) and the stray capacitance of the transformer (C_{stray}).

$$C_r = C_{oss} + C_{res} + C_{trans} \quad (F) \quad (4.6)$$

Primary winding current and voltage across the switch can be determined using simple circuit analysis and applying Laplace transformation.

$$V_{L_p} = \frac{L_p \times dI_p}{dt} \quad (\text{V}) \quad (4.7)$$

$$I_p = \frac{C \times dV_d}{dt} \quad (\text{A}) \quad (4.8)$$

By solving $I_p(s)$ from equation (4.7) and plugging it into equation (4.8) gives us the voltage across the switch in the Laplace domain. Similarly $V_d(s)$ is solved from equation (4.8) and plugged into equation (4.7) to determine $I_p(s)$. By taking inverse Laplace transform the drain voltage and the primary winding current is obtained in time domain.

$$V_d(t) = V_p + \frac{I_p(t_1) \times Z_r \times \sin(\alpha + \omega_r \times t)}{\cos(\alpha)} \quad (\text{V}) \quad (4.9)$$

$$I_p(t) = \frac{V_d(t_1) + V_p}{Z_r} \times \sin(\omega_r \times t) + I_p(t_1) \times \cos(\omega_r \times t) \quad (\text{A}) \quad (4.10)$$

Z_r is the characteristic impedance and ω_r is the resonance frequency in rad/sec .

$$\alpha = \tan^{-1} \left(\frac{V_d(t_1) - V_p}{I_p(t_1) \times Z_r} \right) \quad (\text{Radian}) \quad (4.11)$$

$$Z_r = \sqrt{\frac{L_p}{C_r}} \quad (\Omega) \quad (4.12)$$

$$\omega_r = \frac{1}{\sqrt{L_p \times C_r}} \quad (rad/s) \quad (4.13)$$

At the end of this period the drain voltage and primary winding current is:

$$V_d(t_2) = V_p + V_r \quad (\text{V}) \quad (4.14)$$

$$I_p(t_2) = \frac{V_d(t_1) + V_p}{Z_r} \times \sin(\omega_r \times t_2) + I_p(t_1) \times \cos(\omega_r \times t_2) \quad (\text{A}) \quad (4.15)$$

Commutation time can be obtained from equation (9) using the known value of drain voltage at the end of this switching period:

$$T_{com} = \frac{1}{\omega_r} \times \left(\sin^{-1} \left(\frac{V_r \times \cos(\alpha)}{I_p(t_1) \times Z_r} \right) - \alpha \right) \text{ (s)} \quad (4.16)$$

$t_2 < t < t_3$ time period (T_{demag} – Secondary diode conduction time)

In this period the energy stored in the transformer is transferred to the secondary side. The drain voltage is constant during this period. The voltage and current relations in this period are:

$$V(t) = V_p + V_r \text{ (V)} \quad (4.17)$$

$$I_p(t) = I_p(t_2) - \frac{V_r \times t}{L_p} \text{ (A)} \quad (4.18)$$

At the end of this period the drain voltage and primary winding current is:

$$V_d(t_3) = V_p + V_r \text{ (V)} \quad (4.19)$$

$$I_p(t_3) = 0 \quad (4.20)$$

The demagnetization time (T_{demag}) can be obtained by putting the values at $t = t_3$:

$$T_{demag} = \frac{I_p(t_2) \times L_p}{V_r} \text{ (s)} \quad (4.21)$$

$t_3 < t < t_4$ time period (T_{delay} – Dead time or delay time)

In this period the transformer is demagnetized and the total resonance capacitor (C_r) across the MOSFET starts resonating with the primary inductance of the transformer with a frequency f_{res} . This frequency is:

$$f_{res} = \frac{1}{2 \times \pi \times \sqrt{L_p \times C_r}} \text{ (Hz)} \quad (4.22)$$

The drain voltage and primary winding current in this period is:

$$V_d(t) = V_p + V_r \times \cos(\omega_r \times t) \text{ (V)} \quad (4.23)$$

$$I_p(t) = -\frac{V_r}{Z_r} \times \sin(\omega_r(t)) \text{ (A)} \quad (4.24)$$

At the end of this period the drain voltage and primary winding current is:

$$V_d(t_4) = V_p + V_r \times \cos(\omega_r \times t_4) \text{ (V)} \quad (4.25)$$

$$I_p(t_4) = -\frac{V_r}{Z_r} \times \sin(\omega_r \times t_4) \text{ (A)} \quad (4.26)$$

Different values of input voltage (V_p) determines the behavior of the switching action

a) $V_p = V_r \Rightarrow$

$$V_d(t_4) = 0, I_p(t_4) = 0 \text{ and} \quad (4.27)$$

$$T_{delay} = \frac{\pi}{\omega_r} \text{ (s)}$$

b) $V_p > V_r \Rightarrow$

$$V_d(t_4) = V_p - V_r \text{ (V)}, I_p(t_4) = 0 \text{ and} \quad (4.28)$$

$$T_{delay} = \frac{\pi}{\omega_r} \text{ (s)}$$

c) $V_p < V_r \Rightarrow$

$$V_d(t_4) = 0, I_p(t_4) = -\frac{V_r}{Z_r} \times \sqrt{\left(1 - \frac{V_p^2}{V_r^2}\right)} \text{ (A) and} \quad (4.29)$$

$$T_{delay} = \frac{1}{\omega_r} \times \cos^{-1}\left(-\frac{V_p}{V_r}\right) \text{ (s)}$$

The switching period and frequency is given by Eq. 4.30 and Eq. 4.31, respectively.

$$T_s = T_{on} + T_{com} + T_{demag} + T_{delay} \text{ (s)} \quad (4.30)$$

$$F_s = \frac{1}{T_s} \text{ (Hz)} \quad (4.31)$$

At the same time the power output for a flyback converter is:

$$P_o = \frac{I_p^2 \times L_p \times f_s \times \eta}{2} \text{ (W)} \quad (4.32)$$

Equation (4.32) and equation (4.30) should be solved together to determine the switching frequency. This nonlinear equation that is dependent on primary peak current can be solved numerically using MATLAB or EXCEL with the known power level.

$$T_s = \frac{L_p}{V_p} \times (I_p(t_1) - I_p(t_0)) + \frac{1}{\omega_r} \times \left(\sin^{-1} \left(\frac{V_r \times \cos(\alpha)}{I_p(t_1) \times Z_r} \right) - \alpha \right) + \frac{I_p(t_2) \times L_p}{V_r} + T_{delay} \quad (4.33)$$

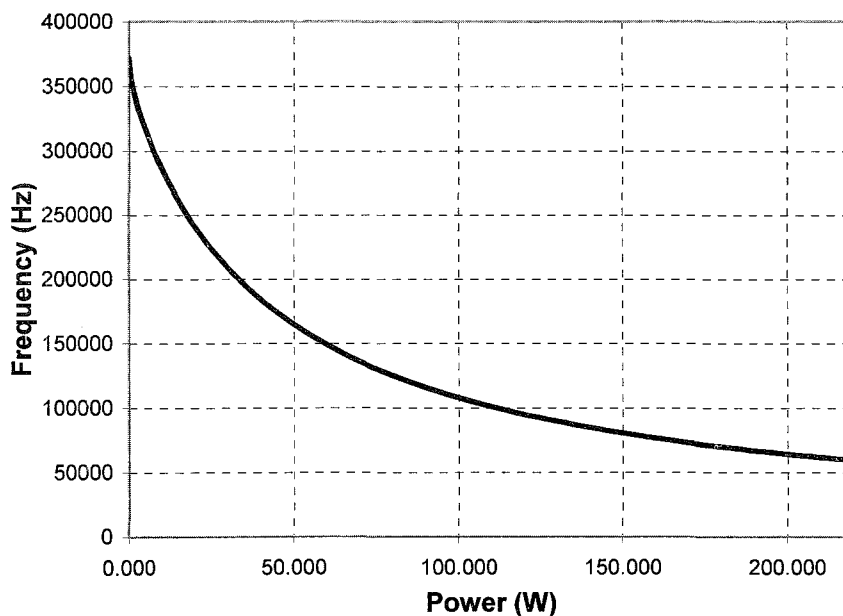


Figure 4.3. Switching frequency-power relationship when $L_p = 446 \mu H$.

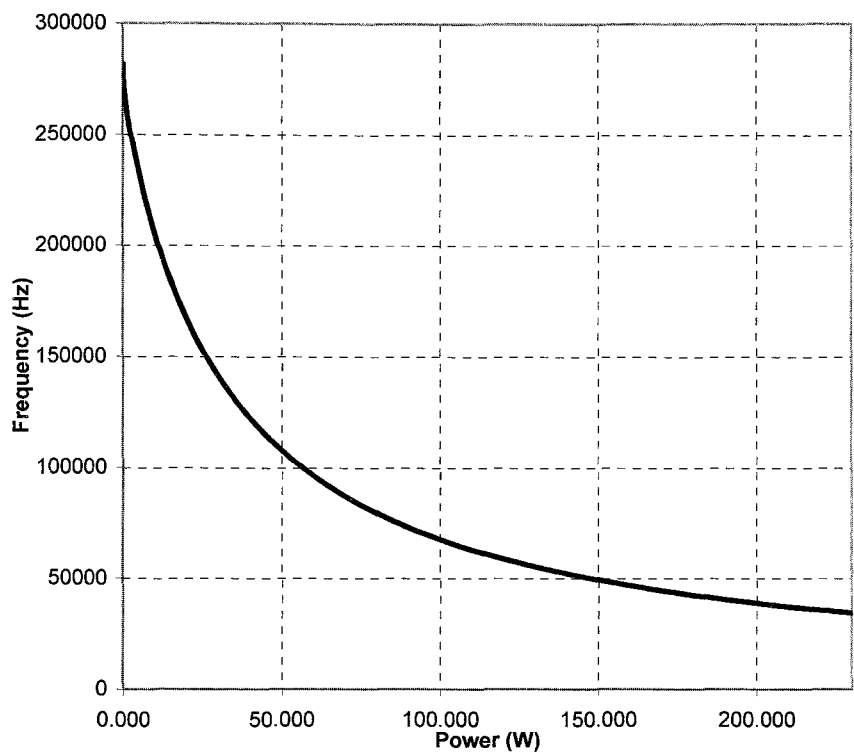


Figure 4.4. Switching frequency-power relationship when $L_p = 758\mu H$.

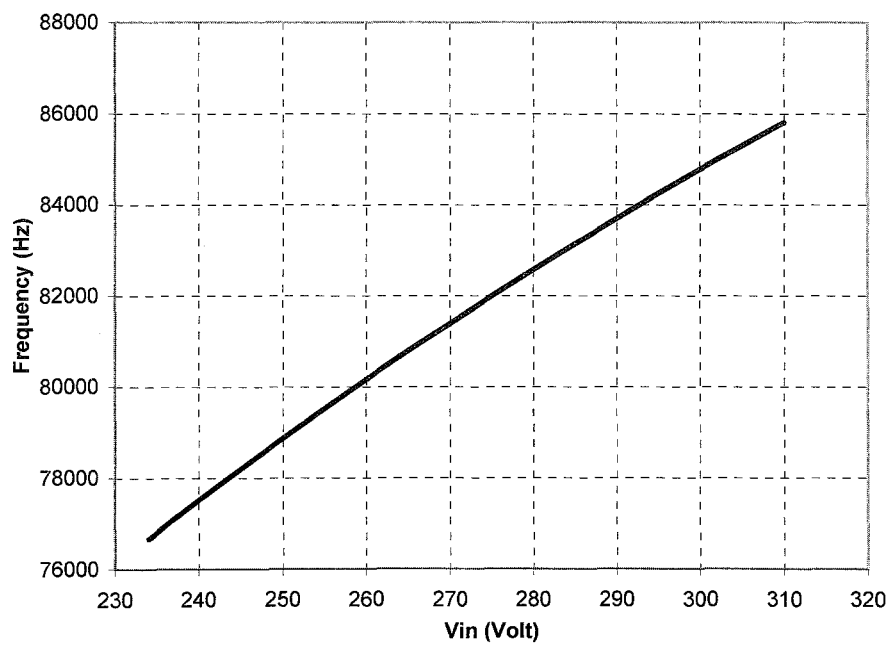


Figure 4.5. Input voltage – switching frequency relationship (@ $I_p = 3A$ and $P_o = 96W$).

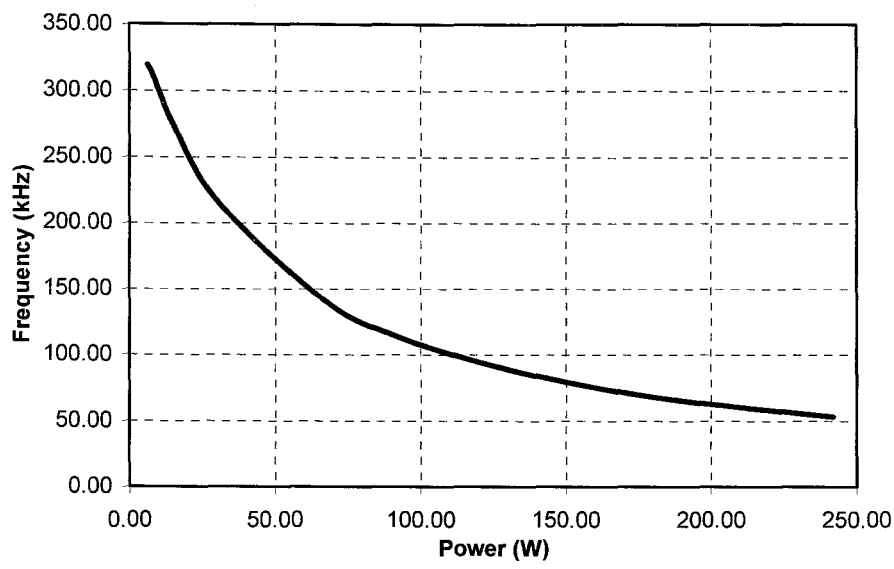


Figure 4.6. Output power – switching frequency relationship calculated using simplified equations.

4.3 Accurate Prediction of Switching Frequency

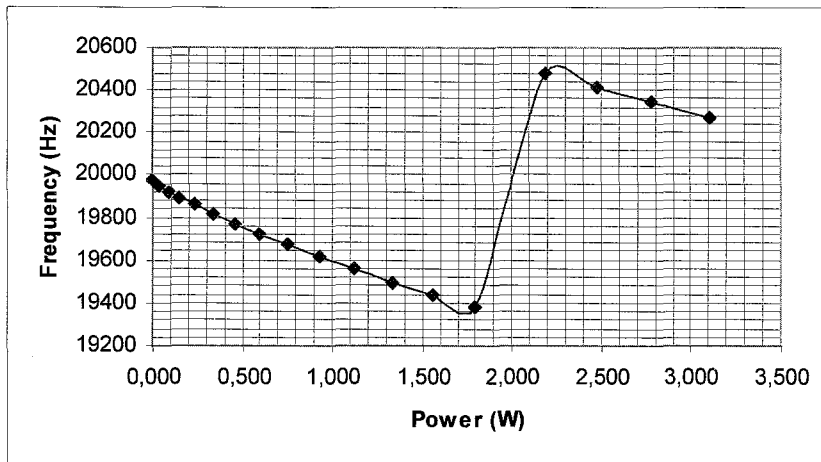
The switching frequency based on the above analysis show that at low output powers switching frequency approaches 300kHz. This is because of the fact that the converter is assumed to be switched at the first zero crossing. This causes very high switching losses in the circuit. In actual circuit this is not the case and there is an off-time circuit in the IC which controls the additional delay after switch-on command. This additional delay is dependent on error voltage or control voltage (V_{err}) which in turn dependent on the peak primary current (output power). The output power’s dependency to control voltage is given in the Table 4.1.

Table 4.1.
Output power versus control voltage

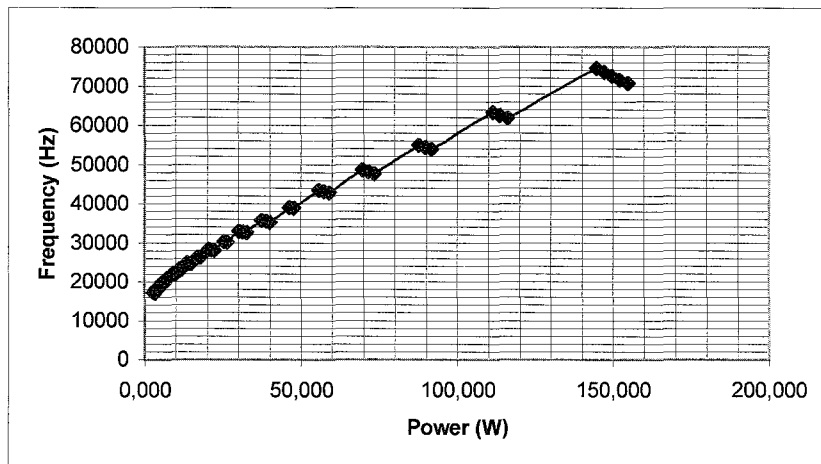
Control Voltage	Output Power	Off-Time $T_{D_{pin1}}$
1,5 – 2V	Low	Constant ($T_{D_{pin1\max}}$), constant frequency stand by
2 – 3,5V	Medium	Decreasing
3,5V – 5V	High	Free running, switch-on at first minimum

The calculated switching frequency with the equations before is true when the control voltage is between 3.5V-5.0V (high power level), when the driver switches at the

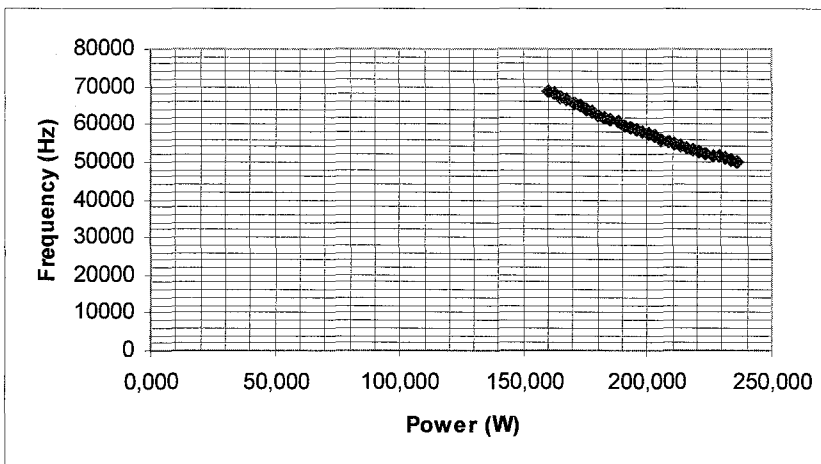
T_{OFF} is the off time calculated with the equation before. With this formula we consider the effect of off-time circuit on switching frequency prediction.



(a)

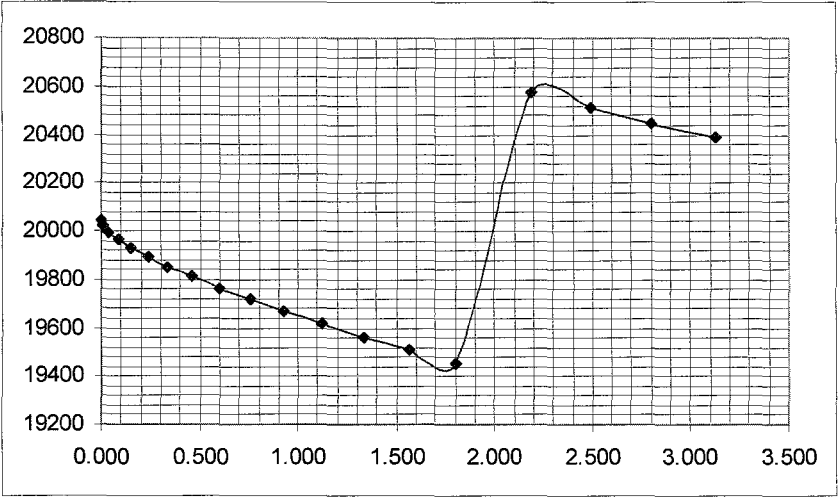


(b)

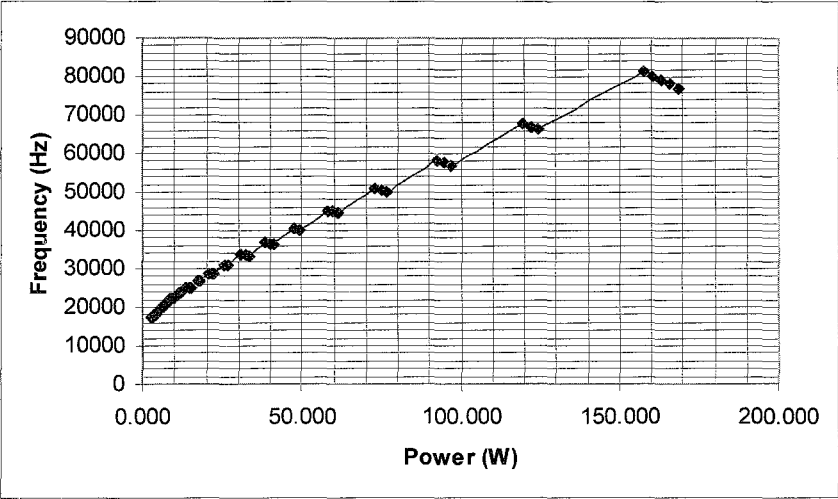


(c)

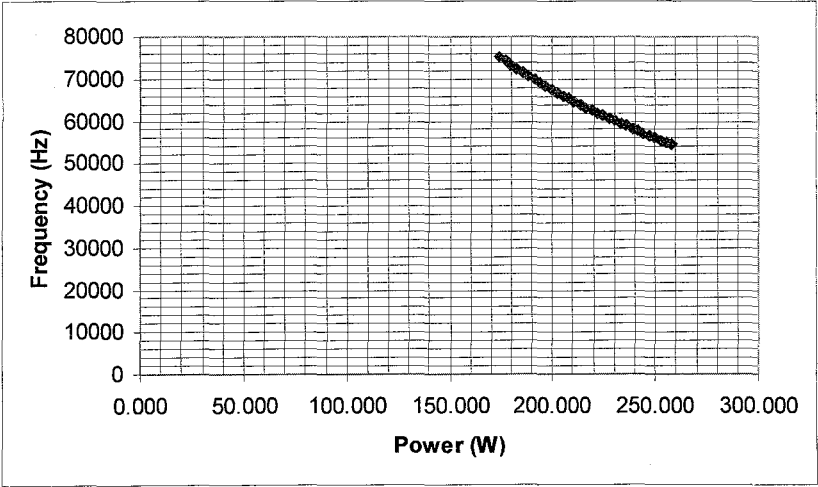
Figure 4.9. Switching frequency vs. output power for $V_{in} = 250V$. (a) $V_{err} < 2V$, (b) $2V < V_{err} < 3.5V$ (c) $3.5V < V_{err} < 5V$.



(a)

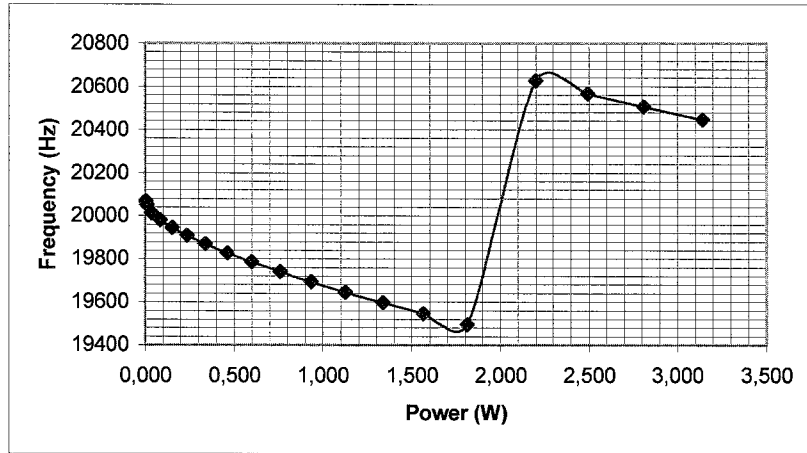


(b)

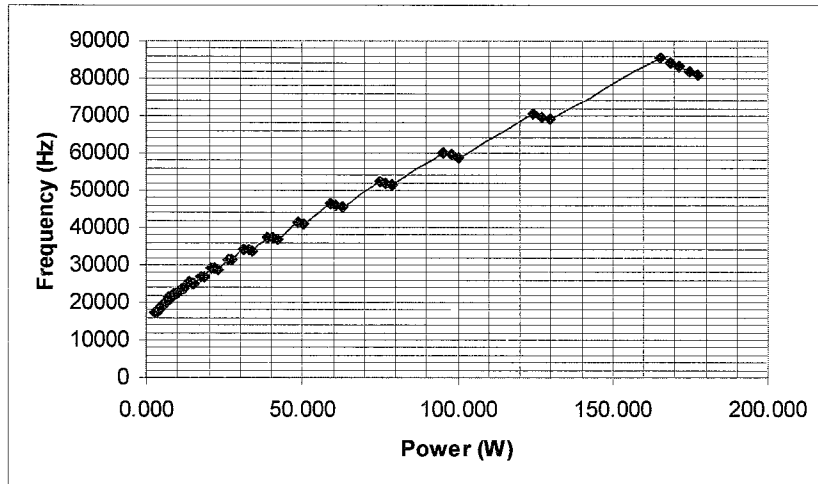


(c)

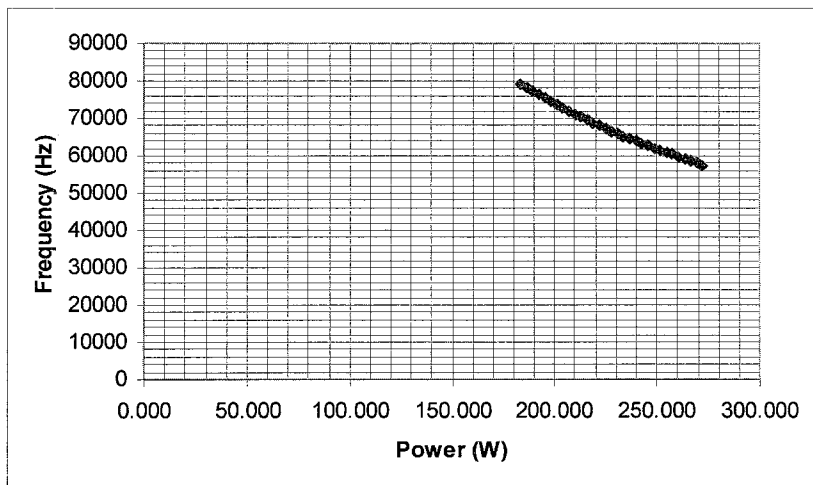
Figure 4.10. Switching frequency vs. output power for $V_{in} = 310V$. (a) $V_{err} < 2V$, (b) $2V < V_{err} < 3.5V$ (c) $3.5V < V_{err} < 5V$.



(a)



(b)



(c)

Figure 4.11. Switching frequency vs. output power for $V_{in} = 357V$. (a) $V_{err} < 2V$, (b) $2V < V_{err} < 3.5V$ (c) $3.5V < V_{err} < 5V$.

4.4 Control-to-Output and Input-to-Output Transfer Function Derivation

4.4.1 Circuit Averaging, Perturbation and Linearization

4.4.1.1 Averaging Waveforms

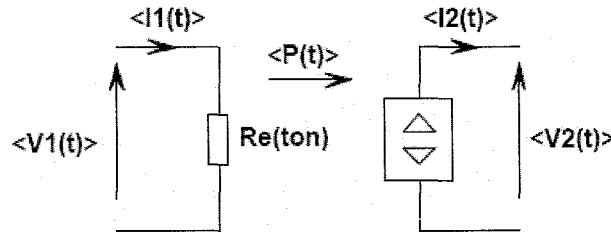


Figure 4.12. Two port model of a flyback converter.

$$T_{on} = \frac{L_p \times I_p}{V_p} \text{ (s)} \quad (4.39)$$

$$T_{demag} = \frac{L_p \times I_p}{V_r} \text{ (s)} \quad (4.40)$$

Reflected secondary voltage is:

$$V_r = (V_o + V_{fo}) \times \frac{N_p}{N_s} \text{ (V)} \quad (4.41)$$

$$R_e(t_{on}) = \frac{2 \times L_p \times (T_{on} + T_{com} + T_{demag} + T_{delay})}{T_{on}^2} \text{ (}\Omega\text{)} \quad (4.42)$$

$$F_s = \frac{1}{T_s} = \frac{1}{T_{on} + T_{com} + T_{demag} + T_{delay}} \text{ (Hz)} \quad (4.43)$$

$$\langle I_1(t) \rangle = \frac{V_p}{R_e(t_{on})} \text{ (A)} \quad (4.44)$$

When T_{com} and T_{delay} is neglected

$$\langle I_2(t) \rangle = \frac{V_p}{V_o + V_{fo}} \times \frac{V_p \times T_{on}^2}{2 \times L_p \times (T_{on} + T_{demag})} \quad (A) \quad (4.45)$$

$$\langle I_2(t) \rangle = \frac{V_p}{V_o + V_{of}} \times \frac{V_p \times T_{on}}{L_p} \times \frac{1}{2} \times \frac{\frac{L_p \times I_p}{V_p}}{\frac{L_p \times I_p}{V_p} + \frac{L_p \times I_p}{V_r}} \quad (A) \quad (4.46)$$

$$\langle I_2(t) \rangle = \frac{V_p}{V_o + V_{of}} \times \frac{1}{2} \times I_p \times \frac{V_r}{V_p + V_r} \quad (A) \quad (4.47)$$

$$\langle I_2(t) \rangle = \frac{V_p}{V_o + V_{of}} \times \frac{1}{2} \times I_p \times \frac{(V_o + V_{of}) \times \frac{N_p}{N_s}}{V_p + (V_o + V_{of}) \times \frac{N_p}{N_s}} \quad (A) \quad (4.48)$$

Average secondary terminal current (without efficiency)

$$\langle I_2(t) \rangle = N \times \frac{1}{2} \times I_p \times \frac{1}{1 + N \times M} \quad (A) \quad (4.49)$$

$N = \frac{N_p}{N_s}$ is primary to secondary turn ratio and

$M = \frac{(V_o + V_{of})}{V_p}$ is the conversion ratio.

$$\langle I_2(t) \rangle = \frac{N_p}{N_s} \times \frac{I_p}{2} \times \frac{V_p}{V_p + (V_o + V_{of}) \times \frac{N_p}{N_s}} \quad (A) \quad (4.50)$$

If we consider efficiency:

$$\langle I_2(t) \rangle = N \times \frac{1}{2} \times I_p \times \frac{1}{1 + N \times M} \times \eta \quad (A) \quad (4.51)$$

The maximum current that can flow through the primary inductance is limited by a RC network at pin 2 of TDA16846 control IC. According to databook [23] the maximum current that can flow through the primary winding is:

$$I_p = \frac{(V_{err} - 1.5V)}{L_p} \times R_2 \times C_2 \text{ (A)} \quad (4.52)$$

R_2 and C_2 are the external resistors connected to pin 2 of the control IC. The average secondary current becomes:

$$\langle I_2(t) \rangle = \frac{N_p}{N_s} \times \frac{V_p}{V_p + (V_o + V_{of})} \times \frac{1}{2} \times \frac{(V_{err} - 1.5V)}{L_p} \times R_2 \times C_2 \text{ (A)} \quad (4.53)$$

4.4.1.2 Linearization of the averaged secondary terminal current

This averaged nonlinear model can be linearized using two different techniques. The first technique is the *perturbation* technique. In this technique all the variables are thought to contain DC and AC components (V_{err} , V_p , V_o). By putting all these values into the averaged equation we derive an equation containing multiplication of DC and AC components plus multiplications of AC components. From the Taylor series expansion, the cross products of these AC components are so small compared to others. Thus we neglect them and the final equation is the linearized small signal model. The second technique is based on partial derivatives. In this technique we take partial derivatives with respect to each variable. These derivatives are the constant gain constants. Then we multiply each variable by its gain constant and obtain the linearized small signal model. After linearization the control to output and line to output transfer functions can be found using known circuit analysis methods. In the present thesis work we used the partial derivatives technique to construct the small signal model of the converter which is described below.

The secondary terminal current I_2 dependent on the error, input and output voltage can be written as follows:

$$\hat{I}_2 = K_e \times \hat{V}_{err} + K_f \times \hat{V}_p + K_r \times \hat{V}_o \text{ (A)} \quad (4.54)$$

The error amplifier gain constant is given by

$$K_e = \frac{\partial \langle I_2(t) \rangle}{\partial V_{err}} \text{ (A/V)} \quad (4.55)$$

$$K_e = \frac{N_p}{N_s} \times \frac{R_2 \times C_2}{2 \times L_p} \times \frac{V_p}{V_p + (V_o + V_{of}) \times \frac{N_p}{N_s}} = N \times \frac{R_2 \times C_2}{2 \times L_p} \times \frac{1}{1 + M \times N} \text{ (A/V)} \quad (4.56)$$

The input voltage gain constant is given as

$$K_f = \frac{\partial \langle I_2(t) \rangle}{\partial V_p} \text{ (A/V)} \quad (4.57)$$

$$K_f = \frac{N_p}{N_s} \times \frac{R_2 \times C_2}{2 \times L_p} \times \left((V_{err} - 1.5V) \times \frac{\left(V_p + (V_o + V_{of}) \times \frac{N_p}{N_s} - V_p \right)}{\left(V_p + (V_o + V_{of}) \times \frac{N_p}{N_s} \right)^2} \right) \text{ (A/V)} \quad (4.58)$$

$$K_f = \frac{N_p^2}{N_s^2} \times \frac{R_2 \times C_2}{2 \times L_p} \times \frac{V_{err} - 1.5V}{V_p^2} \times (V_o + V_{of}) \times \frac{1}{\left(V_p + \frac{V_o + V_{of}}{V_p} \times \frac{N_p}{N_s} \right)^2} \text{ (A/V)} \quad (4.59)$$

$$K_f = N^2 \times \frac{R_2 \times C_2}{2 \times L_p} \times \frac{V_{err} - 1.5}{V_p} \times \frac{M}{(1 + M \times N)^2} \text{ (A/V)} \quad (4.60)$$

For the output voltage gain constant

$$K_r = \frac{\partial \langle I_2(t) \rangle}{\partial V_o} \text{ (A/V)} \quad (4.61)$$

$$K_r = \frac{N_p^2}{N_s^2} \times \frac{R_2 \times C_2}{2 \times L_p} \times \left((V_{err} - 1.5V) \times \frac{\left(-\frac{N_p}{N_s} \times V_p \right)}{\left(V_p + (V_o + V_{of}) \times \frac{N_p}{N_s} \right)^2} \right) \text{ (A/V)} \quad (4.62)$$

$$K_r = -\frac{N_p^2}{N_s^2} \times \frac{R_2 \times C_2}{2 \times L_p} \times \frac{V_{err} - 1.5}{V_p} \times \frac{1}{\left(1 + \frac{V_o + V_{of}}{V_p} \times \frac{N_p}{N_s}\right)^2} \text{ (A/V)} \quad (4.63)$$

$$K_r = -N^2 \times \frac{R_2 \times C_2}{2 \times L_p} \times \frac{V_{err} - 1.5V}{V_p} \times \frac{1}{(1 + M \times N)^2} \text{ (A/V)} \quad (4.64)$$

The linearized equation around its steady state operating point becomes:

$$\begin{aligned} \hat{I}_2 = & \left(N \times \frac{R_2 \times C_2}{2 \times L_p} \times \frac{1}{1 + M \times N} \right) \times \hat{V}_{err} + \left(N^2 \times \frac{R_2 \times C_2}{2 \times L_p} \times \frac{V_{err} - 1.5V}{V_p} \times \frac{M}{(1 + M \times N)^2} \right) \times \hat{V}_p \\ & + \left(-N^2 \times \frac{R_2 \times C_2}{2 \times L_p} \times \frac{V_{err} - 1.5V}{V_p} \times \frac{1}{(1 + M \times N)^2} \right) \times \hat{V}_o \text{ (A)} \end{aligned} \quad (4.65)$$

This equation assumes 100% power transfer. When efficiency is considered it takes the form:

$$\hat{I}_{2eff} = \hat{I}_2 \times \eta \text{ (A)} \quad (4.66)$$

$$\hat{I}_{2eff} = (K_e \times \hat{V}_{err} + K_f \times \hat{V}_p + K_r \times \hat{V}_o) \times \eta \text{ (A)} \quad (4.67)$$

The relation between the error voltage and the output power is obtained as follows:

$$F_s = \frac{1}{T_s} = \frac{1}{T_{on} + T_{off}} \text{ (Hz)} \quad (4.68)$$

$$F_s = \frac{1}{L_p \times I_p \times \left(\frac{1}{V_p} + \frac{1}{V_r} \right)} \text{ (Hz)} \quad (4.69)$$

The output power is:

$$P_o = \frac{1}{2} \times L_p \times I_p^2 \times F_s \times \eta \text{ (W)} \quad (4.70)$$

$$P_o = \frac{1}{2} \times L_p \times I_p^2 \times \frac{1}{L_p \times I_p \times \left(\frac{1}{V_p} + \frac{1}{V_r} \right)} \times \eta = \frac{1}{2} \times I_p \times \eta \times \frac{1}{\frac{1}{V_p} + \frac{1}{V_r}} \quad (\text{W}) \quad (4.71)$$

Primary peak current is:

$$I_p = \frac{2 \times P_o}{\eta} \times \left(\frac{1}{V_p} + \frac{1}{V_r} \right) \quad (\text{A}) \quad (4.72)$$

From datasheet the maximum permissible primary peak current is:

$$I_p = (V_{err} - 1.5V) \times \frac{R_2 \times C_2}{L_p} \quad (\text{A}) \quad (4.73)$$

The error voltage is thus:

$$V_{err} = \frac{I_p \times L_p}{R_2 \times C_2} + 1.5V \quad (\text{V}) \quad (4.74)$$

Substituting I_p in V_{err} the error voltage dependent on output power can be determined:

$$V_{err} = \frac{2 \times P_o}{\eta} \times \left(\frac{1}{V_p} + \frac{1}{V_r} \right) \times \frac{L_p}{R_2 \times C_2} + 1.5V \quad (\text{V}) \quad (4.75)$$

The procedure for determining the transfer functions is described in the next section.

4.4.2 Control-to-Output and Line-to-Output Transfer Functions

Control-to-output and line-to-output transfer functions are found using two port techniques. When finding one transfer function the other variables are assumed to be zero.

The control to output transfer function is found by:

$$G_{vc} = \frac{\hat{V}_o}{\hat{V}_{err}} \Big|_{\hat{V}_p=0} \quad (\text{V/V}) \quad (4.76)$$

The line to output transfer function is found by:

$$G_{vg} = \frac{\hat{V}_o}{\hat{V}_p} \Big|_{\hat{V}_{err}=0} \quad (\text{V/V}) \quad (4.77)$$

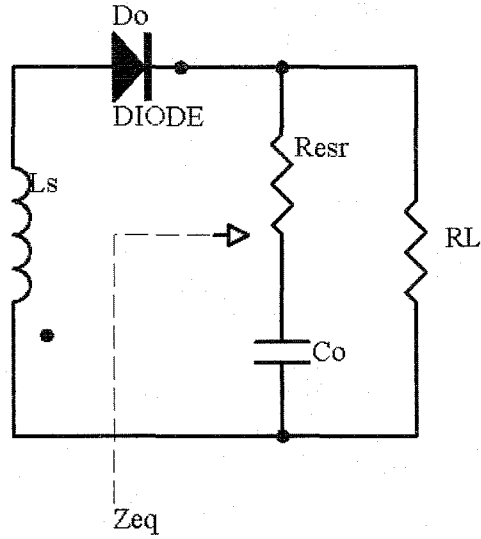


Figure 4.13. Output impedance of flyback converter.

The output impedance of the flyback converter is:

$$Z_{eq} = \frac{\hat{V}_o}{\hat{I}_2} \quad (\Omega) \quad (4.78)$$

$$Z_{eq} = R_L \times \frac{1 + s \times C_o \times R_{esr}}{1 + s \times C_o \times (R_{esr} + R_L)} \quad (\Omega) \quad (4.79)$$

$$\hat{V}_o = Z_{eq} \times \hat{I}_2 = Z_{eq} \times (K_e \times \hat{V}_{err} + K_f \times \hat{V}_p + K_r \times \hat{V}_o) \quad (\text{V}) \quad (4.80)$$

Control to output transfer function (G_{vc})

$$G_{vc} = \frac{\hat{V}_o}{\hat{V}_{err}} \Big|_{\hat{V}_p=0} \quad (\text{V/V}) \quad (4.81)$$

$$G_{vc} = \frac{\hat{V}_o}{\hat{V}_{err}} = Z_{eq} \times \left(K_e + K_r \times \frac{\hat{V}_o}{\hat{V}_{err}} \right) \quad (\text{V/V}) \quad (4.82)$$

$$G_{vc} = \frac{Z_{eq} \times K_e}{1 - Z_{eq} \times K_r} \text{ (V/V)} \quad (4.83)$$

$$G_{vc} = \left(\frac{K_e \times R_L \times (1 + s \times C_o \times R_{esr})}{1 + s \times C_o \times (R_{esr} + R_L)} \right) \bigg/ \left(1 - \frac{K_r \times R_L \times (1 + s \times C_o \times R_{esr})}{1 + s \times C_o \times (R_{esr} + R_L)} \right) \text{ (V/V)} \quad (4.84)$$

$$\begin{aligned} G_{vc} &= \frac{K_e \times R_L \times (1 + s \times C_o \times R_{esr})}{1 + s \times C_o \times (R_{esr} + R_L) - K_r \times R_L \times (1 + s \times C_o \times R_{esr})} \\ &= \frac{K_e \times R_L \times (1 + s \times C_o \times R_{esr})}{(1 - K_r \times R_L) + s \times C_o \times (R_{esr} + R_L - K_r \times R_L \times R_{esr})} \text{ (V/V)} \end{aligned} \quad (4.85)$$

$$G_{vc} = \frac{K_e \times R_L}{1 - K_r \times R_L} \times \frac{1 + s \times C_o \times R_{esr}}{1 + s \times \frac{C_o \times (R_{esr} + R_L - K_r \times R_L \times R_{esr})}{1 - K_r \times R_L}} = K_a \times \frac{1 + \frac{s}{w_z}}{1 + \frac{s}{w_p}} \text{ (V/V)} \quad (4.86)$$

Where

$$K_a = \frac{K_e \times R_L}{1 - K_r \times R_L} \quad (4.87)$$

The zero of the control to output transfer function of power stage is

$$w_z = \frac{1}{C_o \times R_{esr}} \text{ (Radian/sec)} \quad (4.88)$$

The pole of the control to output transfer function of power stage is

$$w_p = \frac{1 - K_r \times R_L}{C_o \times (R_{esr} + R_L - K_r \times R_L \times R_{esr})} \text{ (Radian/sec)} \quad (4.89)$$

The line to output transfer function (G_{vg})

$$G_{vg} = \frac{\hat{V}_o}{\hat{V}_p} \bigg|_{\hat{V}_{err}=0} \text{ (V/V)} \quad (4.90)$$

$$G_{vg} = \frac{\hat{V}_o}{\hat{V}_p} = Z_{eq} \times \left(K_f + K_r \times \frac{\hat{V}_o}{\hat{V}_p} \right) \text{ (V/V)} \quad (4.91)$$

$$G_{vg} = \frac{Z_{eq} \times K_f}{1 - Z_{eq} \times K_r} \text{ (V/V)} \quad (4.92)$$

$$G_{vg} = \left(\frac{K_f \times R_L \times (1 + s \times C_o \times R_{esr})}{1 + s \times C_o \times (R_{esr} + R_L)} \right) / \left(1 - \frac{K_r \times R_L \times (1 + s \times C_o \times R_{esr})}{1 + s \times C_o \times (R_{esr} + R_L)} \right) \text{ (V/V)} \quad (4.93)$$

$$G_{vg} = \frac{K_f \times R_L \times (1 + s \times C_o \times R_{esr})}{1 + s \times C_o \times (R_{esr} + R_L) - K_r \times R_L \times (1 + s \times C_o \times R_{esr})} \text{ (V/V)} \quad (4.94)$$

$$G_{vg} = \frac{K_f \times R_L \times (1 + s \times C_o \times R_{esr})}{(1 - K_r \times R_L) + s \times C_o \times (R_{esr} + R_L - K_r \times R_L \times R_{esr})} \text{ (V/V)} \quad (4.95)$$

$$G_{vg} = \frac{K_f \times R_L}{(1 - K_r \times R_L)} \times \frac{1 + s \times C_o \times R_{esr}}{1 + s \times \frac{C_o \times (R_{esr} + R_L - K_r \times R_L \times R_{esr})}{1 - K_r \times R_L}} = K_b \times \frac{1 + \frac{s}{w_z}}{1 + \frac{s}{w_p}} \text{ (V/V)} \quad (4.96)$$

Where:

$$K_b = \frac{K_f \times R_L}{(1 - K_r \times R_L)} \quad (4.97)$$

The zero frequency of the line to output transfer function of the power stage is:

$$w_z = \frac{1}{C_o \times R_{esr}} \text{ (Radian/sec)} \quad (4.98)$$

Pole frequency of the line to output transfer function of the power stage is:

$$w_p = \frac{1 - K_r \times R_L}{C_o \times (R_{esr} + R_L - K_r \times R_L \times R_{esr})} \text{ (V/V)} \quad (4.99)$$

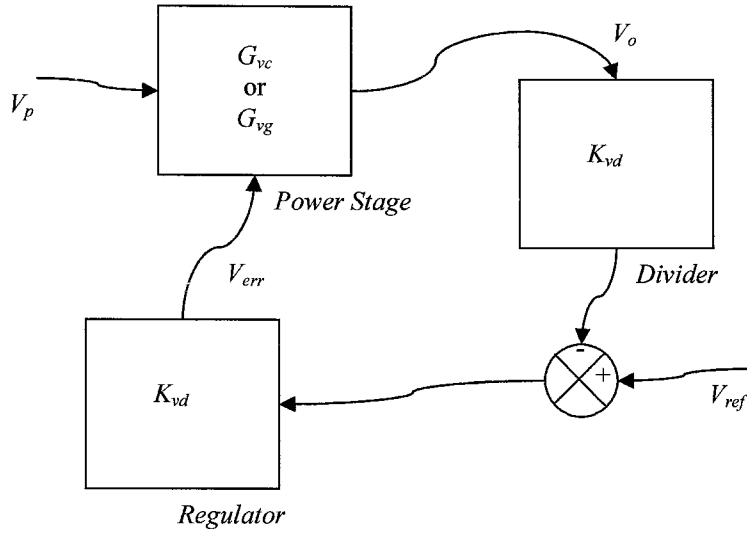


Figure 4.14. Block diagram of the proposed converter.

Open loop transfer function of converter is the product of all the gains:

$$F_{open} = F_{PWR}(s) \times K_{vd} \times F_{reg} \text{ (V/V)} \quad (4.100)$$

Here the $F_{PWR}(s)$ includes the error amplifier output to primary current transfer function. The open loop control to output transfer function is:

$$F_{open cv} = K_a \times \frac{1 + \frac{s}{w_z}}{1 + \frac{s}{w_p}} \times K_{vd} \times F_{reg}(s) \text{ (V/V)} \quad (4.101)$$

The line to output transfer function is:

$$F_{open vg} = K_b \times \frac{1 + \frac{s}{w_z}}{1 + \frac{s}{w_p}} \times K_{vd} \times F_{reg}(s) \text{ (V/V)} \quad (4.102)$$

We see that either control to output or the line to output transfer functions are multiplied by regulator transfer function. Design of regulator is based on the converter power stage transfer function. Control to output and line to output transfer function bode plots at different power and line voltage are given in Figs. 4.15-4.26.

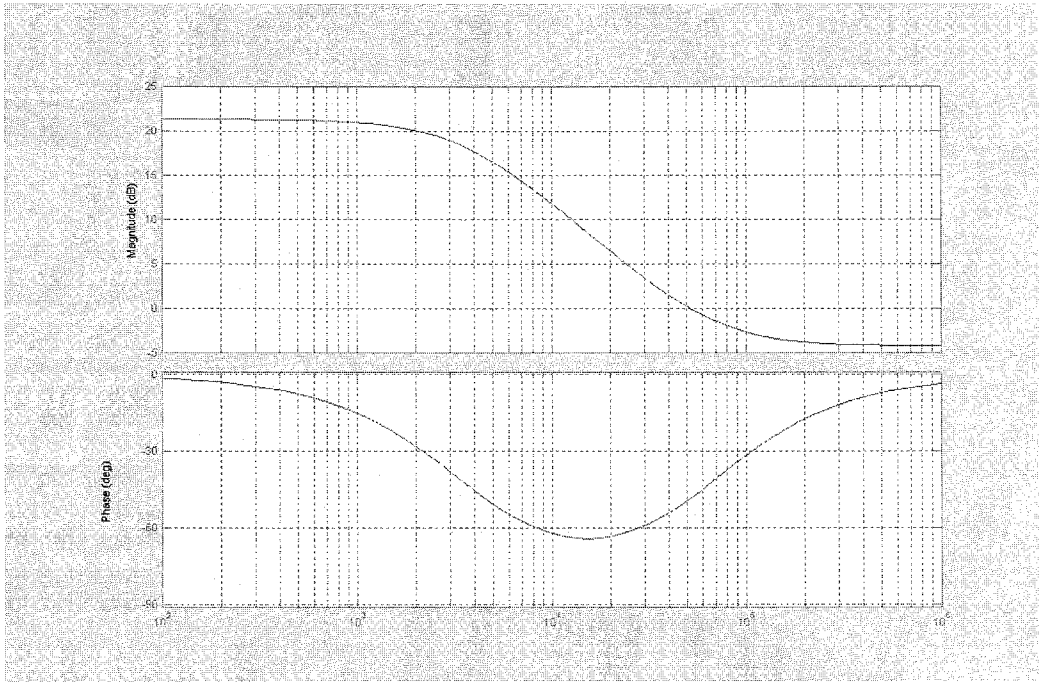


Figure 4.15. Control to output @ maximum input voltage maximum load.

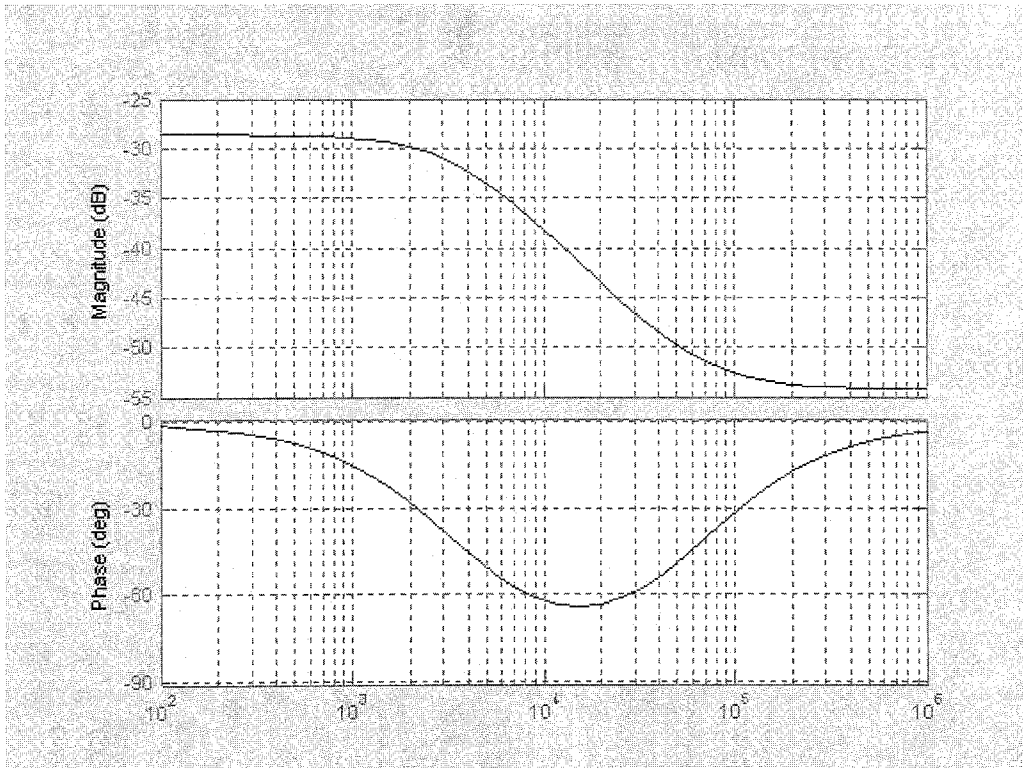


Figure 4.16. Line to output @ maximum input voltage maximum load.

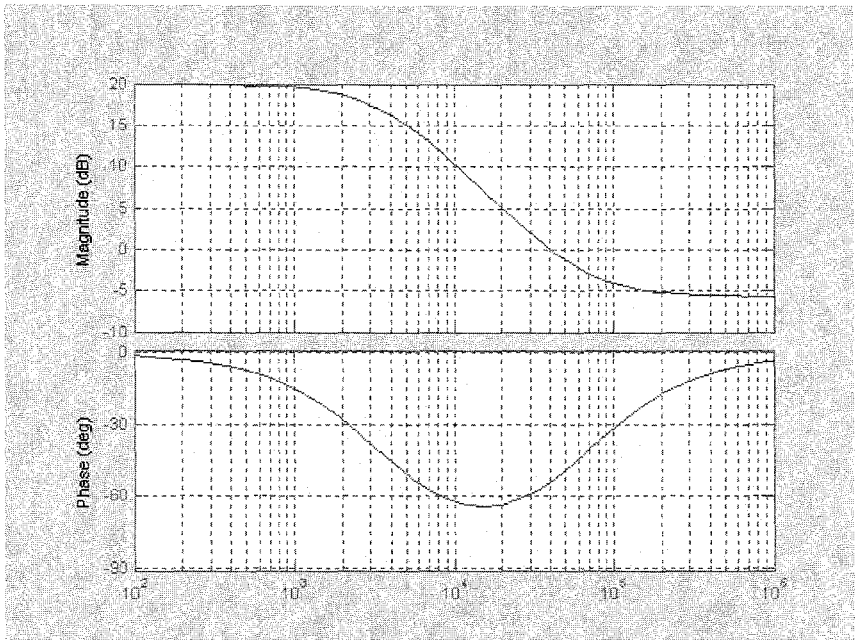


Figure 4.17. Control to output @ maximum load minimum input voltage.

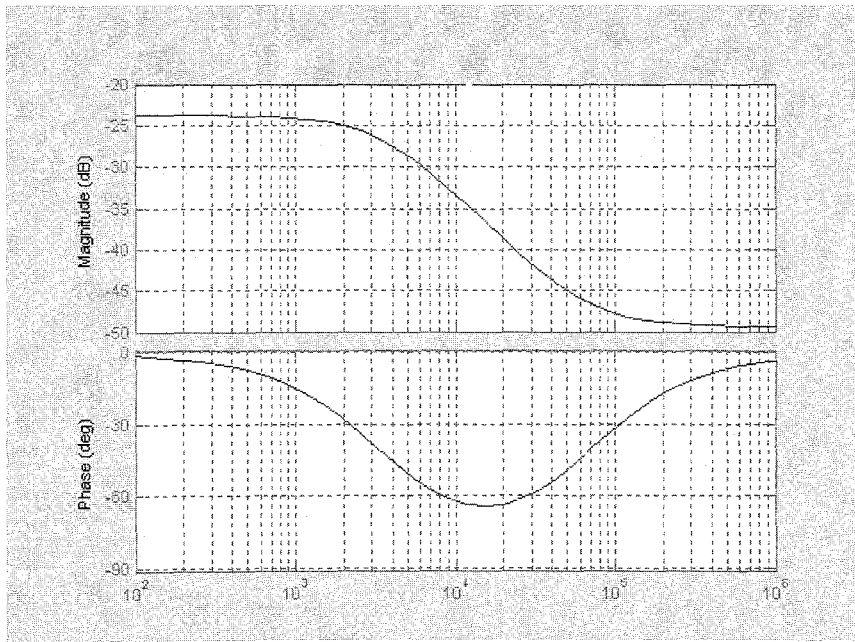


Figure 4.18. Line to output @ maximum load minimum input voltage.

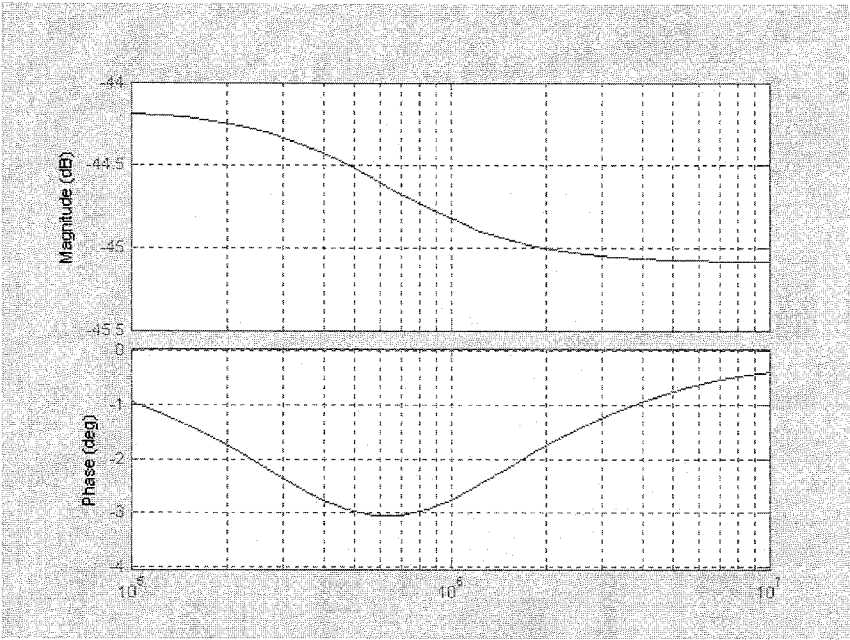


Figure 4.19. Control to output @ minimum load minimum input voltage.

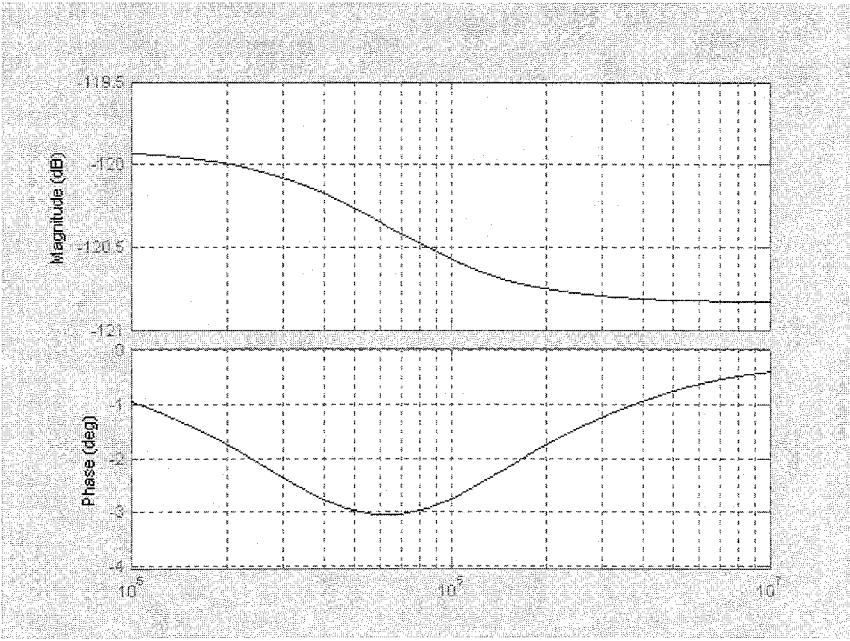


Figure 4.20. Line to output @ minimum load minimum input voltage.

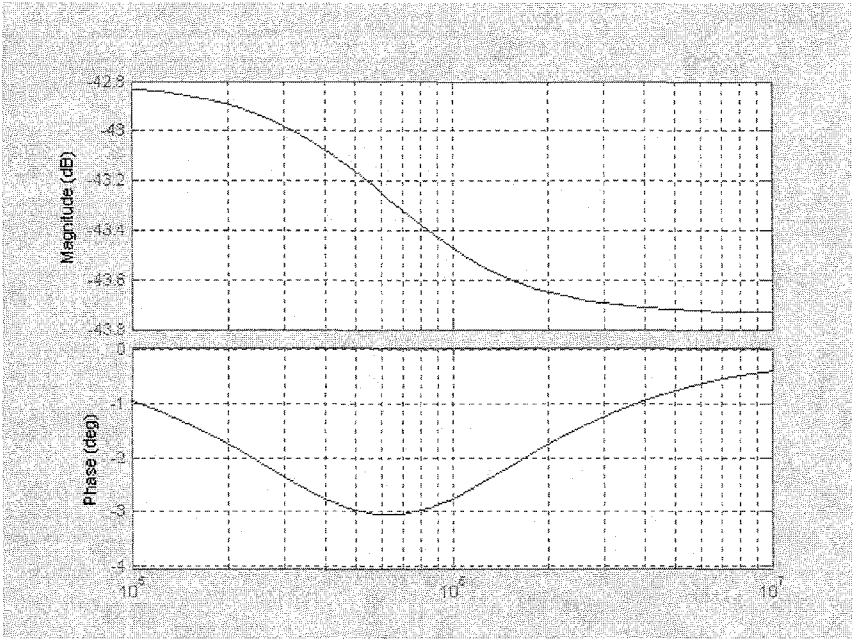


Figure 4.21. Control to output @ minimum load maximum input voltage.

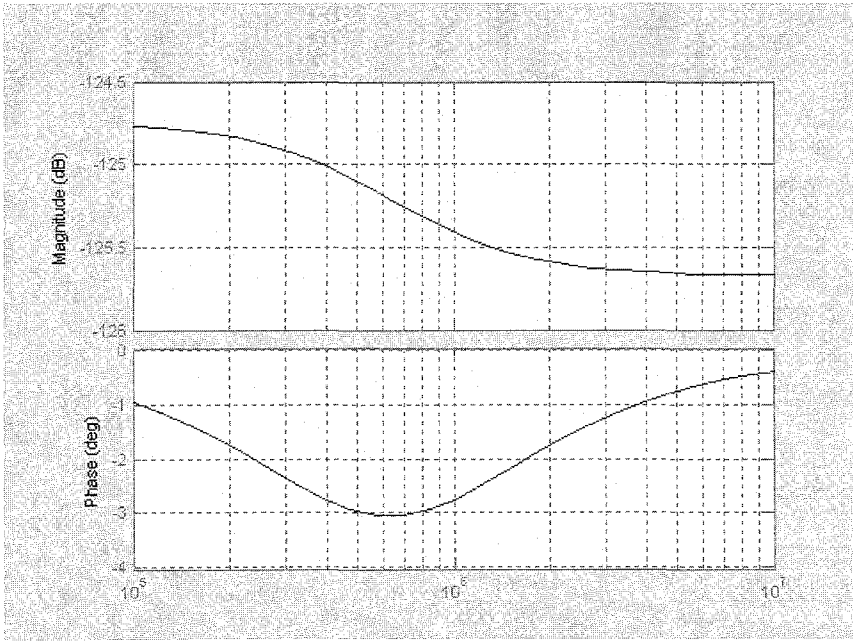


Figure 4.22. Line to output @ minimum load maximum input voltage.

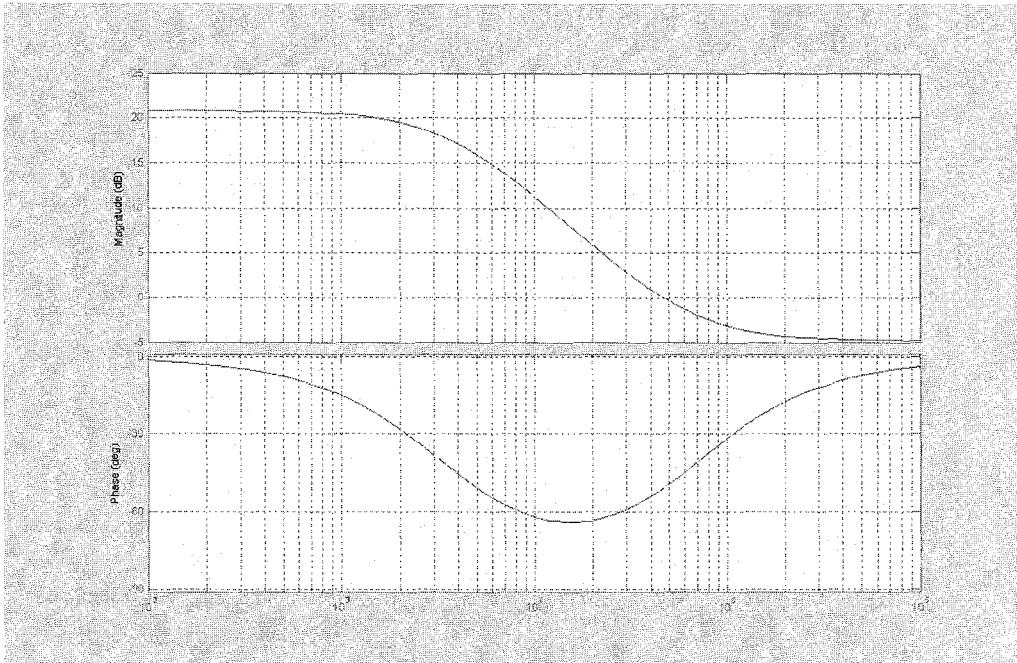


Figure 4.23. Control to output @ maximum load nominal input voltage.

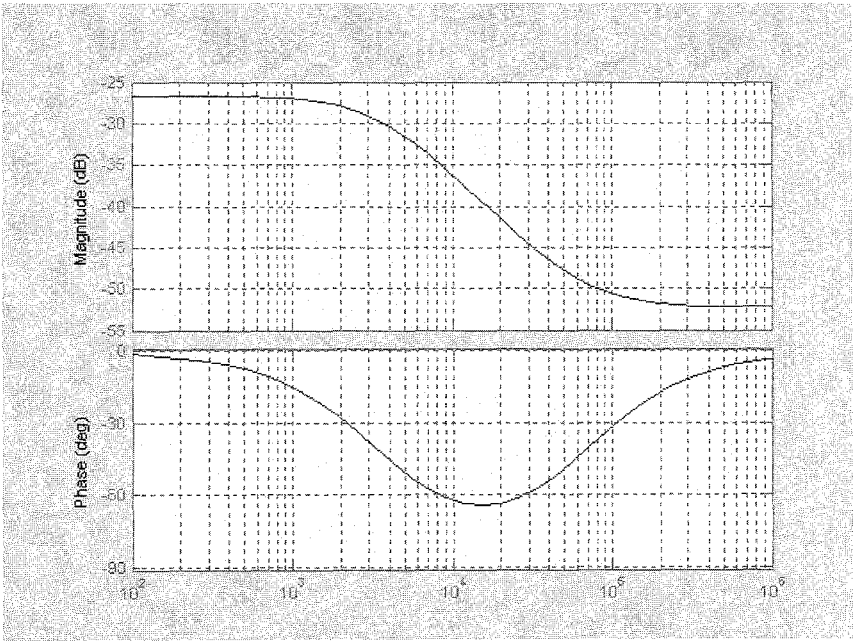


Figure 4.24. Line to output @ maximum load nominal input voltage.

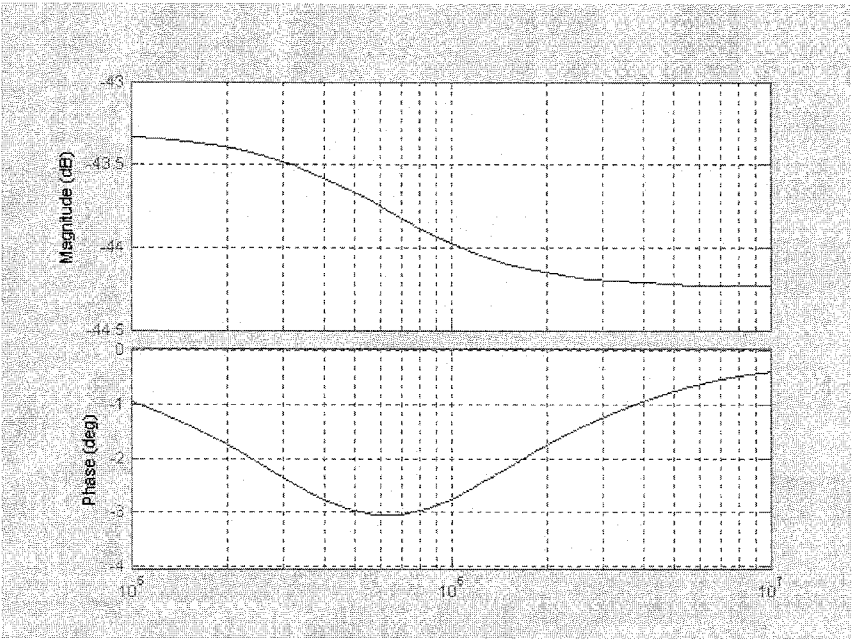


Figure 4.25. Control to output @ minimum load nominal input voltage.

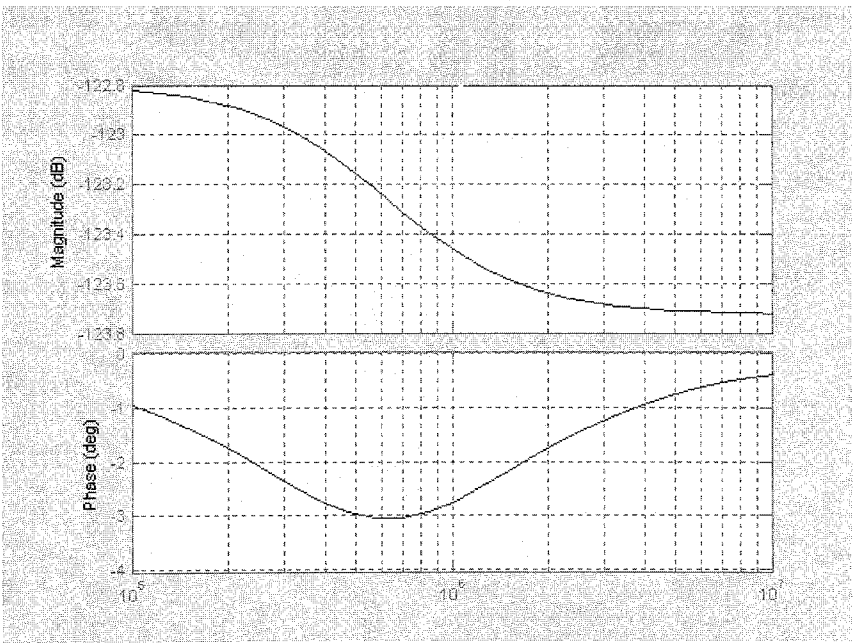


Figure 4.26. Line to output @ minimum load nominal input voltage.

4.5 Error Amplifier Transfer Function Derivation

An optimum feedback amplifier has its maximum gain below the crossover frequency and minimum gain above it.

Open loop gain crossover frequency (Loop bandwidth): Gain crossover frequency is the frequency at which the gain of the open loop transfer function is 0 dB on the Bode diagram. A feedback loop will oscillate if the loop gain is 1 and total phase lag is 360° (in phase). In terms of stability two factors are important. Phase margin is the difference between phase lag and 360° when the open loop gain is unity (in degrees) and gain margin is the amount of gain has fallen below unity when the total phase lag is 360° (in dB) [28]. In classical control applications oscillation is discussed in terms of 180° but in reality this is 360° . Here we also consider the 180° case.

It is suggested that the phase margin of the open loop transfer function should be within $45^\circ - 75^\circ$ [28]. Lower margins give good transient response at the expense of peaking of the closed loop transfer function and output impedance. Higher margins give flat closed loop transfer functions and minimum peaking of output impedance, at the expense of speed and settling time. Optimum choice may be designing between these values, which may be 60° .

There are three types of error amplifiers: Type I, Type II and Type III. In Type I the phase lag is 270° and is constant. Because of this the total phase of the converter cannot be adjusted. Type two amplifiers have a pole-zero pair, which allows us to adjust the overall phase of the converter open loop transfer function. Since there is one zero, the maximum phase boost achievable is 90° . Type III amplifier has one pole and two zeros so that the total phase boost achievable is 180° . This type is used when there is a LC filter at the output which may lead to a phase lag about 180° , indicating more correction action is needed. In Type II error amplifiers a pole at the origin is placed to provide maximum open loop gain below the crossover frequency. It also provides better transient response and disturbance rejection. A pole at higher frequencies minimizes the open loop gain in high frequency range above the crossover frequency. It also improves noise immunity. The theoretically maximum phase boost is 90° .

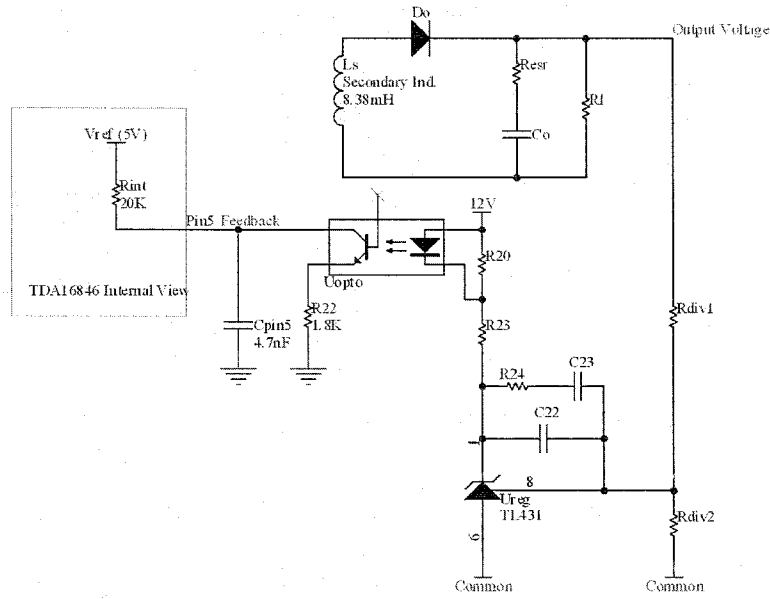


Figure 4.27. Error amplifier used in the circuit that uses a TL431 regulator and 4N35 opto coupler (PI – Proportional plus Integrator action).

In Fig. 4.27 we can see that a resistor is connected with a capacitor (R_{24} and C_{23}) in series and a capacitor (C_{22}) in parallel with them. As seen from Fig. 4.27 and the functional block diagram of TL431 below at DC and below nominal output voltages C_{22} and C_{23} are open circuit and the transistor at the output of the operational amplifier is not conducting. No current flows through the opto coupler LED. However after the nominal output voltage the cathode voltage changes with the open loop gain of the TL431. At AC C_{22} and C_{23} are closed and a feedback path develops. Capacitor C_{22} is placed to form an integrator to boost the DC gain. In some designs C_{22} is placed across R_{24} but because C_{23} is a high value capacitor it has poor performance at high frequencies. To avoid high frequency currents through C_{23} , C_{22} is placed across the C_{23} - R_{24} combination.

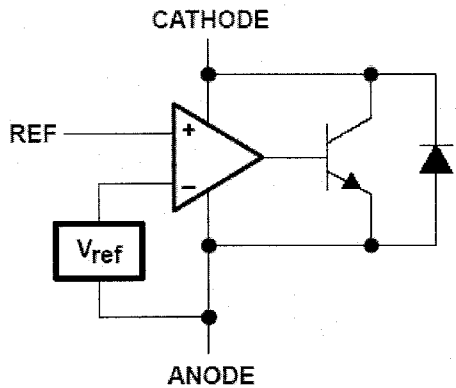


Figure 4.28. Functional diagram of TL431 regulator [29].

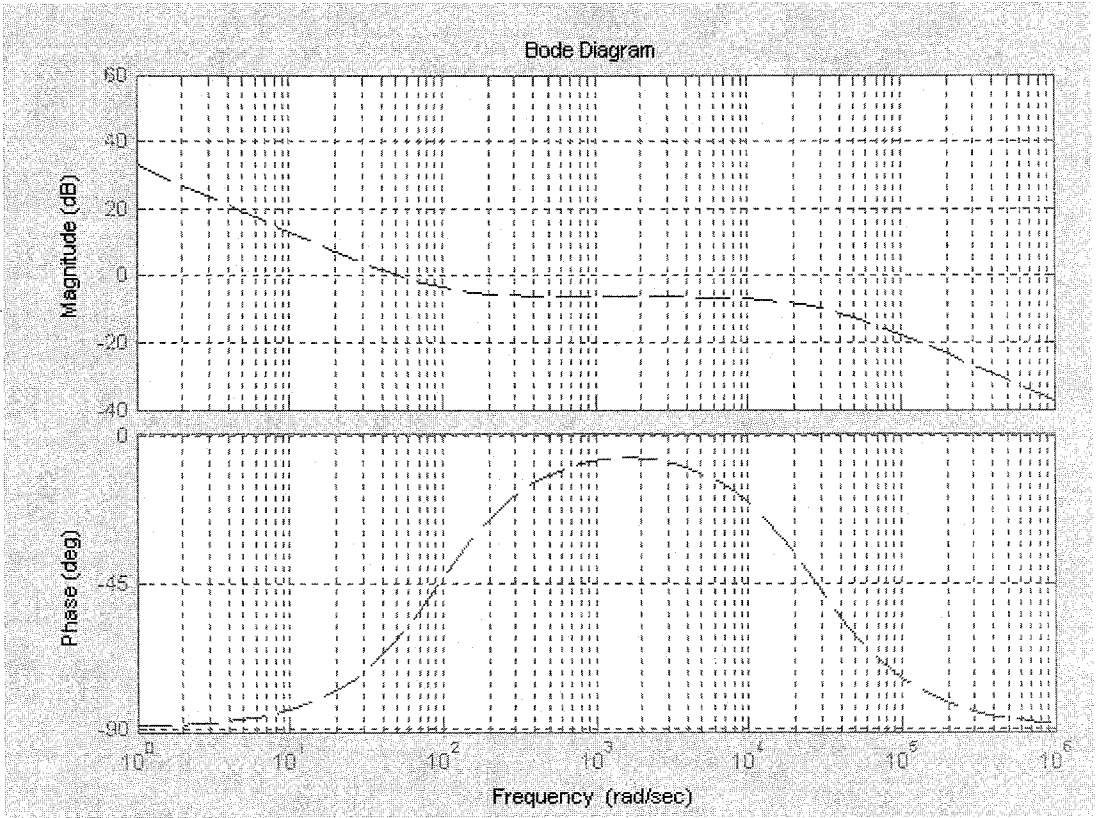


Figure 4.29. Typical Type II error amplifier bode plot.

At DC C22 and C23 are open-circuit and the opto coupler led current is;

$$I_{LED} = \frac{15V - V_z}{R_d // R_{20} + R_{23} + R_z} \times \frac{R_5}{R_d + R_5} \quad (A) \quad (4.103)$$

$$I_{LED} = \frac{R_{20}}{R_d \times R_{20} + R_{23} \times R_d + R_{23} \times R_{20}} \times (15V - V_z) \quad (A) \quad (4.104)$$

The feedback current is the LED current multiplied by the CTR of the opto coupler, which is usually in the range between 1 and 2. Because the capacitors are open circuit at DC, the cathode voltage is determined by the open loop gain of the TL431.

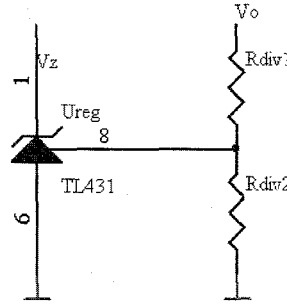


Figure 4.30. At DC capacitors are open circuit and zener voltage is dependent on the open loop gain of TL431.

$$V_z = -V_o \times \frac{R_{div2}}{R_{div1} + R_{div2}} \times A_{v_{TL431}} \quad (V) \quad (4.105)$$

$$I_{FB} = I_{LED} \times CTR \quad (A) \quad (4.106)$$

$$I_{FB} = \frac{R_{20}}{R_d \times R_{20} + R_{23} \times R_d + R_{23} \times R_{20}} \times \left(15V + \frac{V_o \times R_{div2}}{R_{div1} + R_{div2}} \times A_{v_{TL431}} \right) \times CTR \quad (A) \quad (4.107)$$

The feedback voltage is given by:

$$V_{FB} = 5V - R_{int} \times I_{FB} \quad (V) \quad (4.108)$$

$$V_{FB} = 5 - \frac{R_{20} \times R_{int}}{(R_d \times R_{20} + R_{23} \times R_d + R_{23} \times R_{20})} \times \frac{R_{div2}}{R_{div1} + R_{div2}} \times A_{v_{TL431}} \times CTR \quad (V) \quad (4.109)$$

And:

$$\frac{\partial V_{FB}}{\partial V_o} = - \frac{R_{20} \times R_{int}}{R_d \times R_{20} + R_{23} \times R_d \times R_{20} \times R_{23}} \times \frac{R_{div2}}{R_{div1} + R_{div2}} \times A_{v_{TL431}} \times CTR \quad (V/V) \quad (4.110)$$

At AC, C_{22} and C_{23} are closed and led current is:

$$I_{LED} = \frac{R_{20}}{R_d \times R_{20} + R_{23} \times R_d + R_{23} \times R_{20}} \times (15V - V_z) \quad (A) \quad (4.111)$$

R_{div2} is at virtually ground potential (0V). Zener voltage is:

$$V_z = -V_o \times \frac{Z_{es}}{R_{div1}} \quad (V) \quad (4.112)$$

The equivalent impedance is:

$$Z_{eq} = \frac{1 + s \times R_{24} \times C_{23}}{s \times ((C_{22} + C_{23}) + s \times R_{24} \times C_{22} \times C_{23})} (\Omega) \quad (4.113)$$

$$V_z = -\frac{V_o}{R_{div1}} \times \frac{1 + s \times R_{24} \times C_{23}}{s \times (C_{22} + C_{23}) \times \left(1 + s \times \frac{R_{24} \times C_{22} \times C_{23}}{C_{22} + C_{23}}\right)} (V) \quad (4.114)$$

The LED current is:

$$I_{LED} = \frac{R_{20}}{R_d \times R_{20} + R_{23} \times R_d + R_{20} \times R_{23}} \times \left(15 + \frac{V_o}{R_{div1} \times (C_{22} + C_{23})} \times \frac{1 + s \times R_{24} \times C_{23}}{s \times \left(1 + s \times \frac{R_{24} \times C_{22} \times C_{23}}{C_{22} + C_{23}}\right)}\right) (A) \quad (4.115)$$

The feedback current is:

$$I_{FB} = I_{LED} \times CTR (A) \quad (4.116)$$

The feedback voltage is:

$$V_{FB} = 5 - I_{LED} \times CTR \times R_{int} (V) \quad (4.117)$$

$$V_{FB} = 5 - \frac{R_{20}}{R_d \times R_{20} + R_{23} \times R_d + R_{20} \times R_{23}} \times \left(15 + \frac{V_o}{R_{div1} \times (C_{22} + C_{23})} \times \frac{1 + s \times R_{24} \times C_{23}}{s \times \left(1 + s \times \frac{R_{24} \times C_{22} \times C_{23}}{C_{22} + C_{23}}\right)}\right) \times R_{int} \times CTR (V) \quad (4.118)$$

The regulator transfer function can be written as:

$$\frac{\partial V_{FB}}{\partial V_o} = -\frac{R_{20} \times R_{int} \times CTR}{(R_d \times R_{20} + R_{23} \times R_d + R_{20} \times R_{23}) \times R_{div1} \times (C_{22} + C_{23})} \times \left(\frac{1 + s \times R_{24} \times C_{23}}{s \times \left(1 + s \times \frac{R_{24} \times C_{22} \times C_{23}}{C_{22} + C_{23}}\right)}\right) (V/V) \quad (4.119)$$

We can write this as:

$$\frac{\partial V_{FB}}{\partial V_o} = G_{10} \times \left(\frac{1 + s \times R_{24} \times C_{23}}{s \times \left(1 + s \times \frac{R_{24} \times C_{22} \times C_{23}}{C_{22} + C_{23}} \right)} \right) \text{ (V/V)} \quad (4.120)$$

Where:

$$G_{10} = - \frac{R_{20} \times R_{int} \times CTR}{(R_d \times R_{20} + R_{23} \times R_d + R_{20} \times R_{23}) \times R_{div1} \times (C_{22} + C_{23})} \quad (4.121)$$

This is a Type II transfer function, which contains one pole at zero frequency that provides high gain at DC and at low frequencies. It contains a zero at a lower frequency, which cancels the pole of the power stage. Thus bode diagram slope becomes zero. It also has a pole at high frequencies that makes bode plot to fall with a slope of -20dB beyond this range. The zero frequency of the transfer function is:

$$f_z = \frac{1}{2 \times \pi \times R_{24} \times C_{23}} \text{ (Hz)} \quad (4.122)$$

The first pole is:

$$f_{p1} = 0 \text{ Hz}$$

The second high frequency pole is:

$$f_{p2} = \frac{C_{22} + C_{23}}{2 \times \pi \times R_{24} \times C_{22} \times C_{23}} \text{ (Hz)} \quad (4.123)$$

In order to determine where to place the zero and the pole of the error amplifier we should look at the control-to-output transfer function and determine the error amplifier bandwidth. Normally in discontinuous flyback converters the RHP zero is assumed not to exist. In some papers it is claimed that the boundary conduction mode flyback converters also have a RHP zero which changes with the changing switching frequency and duty ratio. If we do not consider the commutation time and delay time, the switching frequency increases as high as 300kHz. However in real situations, because of the off-time circuit at lower power levels and at lower frequencies, the RHP zero will appear at a very high frequency (> 100kHz). Therefore the regulator bandwidth can be increased. At higher power levels the RHP gets closer to 30kHz so we should design the regulator with this

value. It is advised to have the crossover frequency to be 1/10 times of RHP zero. Thus we choose 3kHz for the crossover frequency.

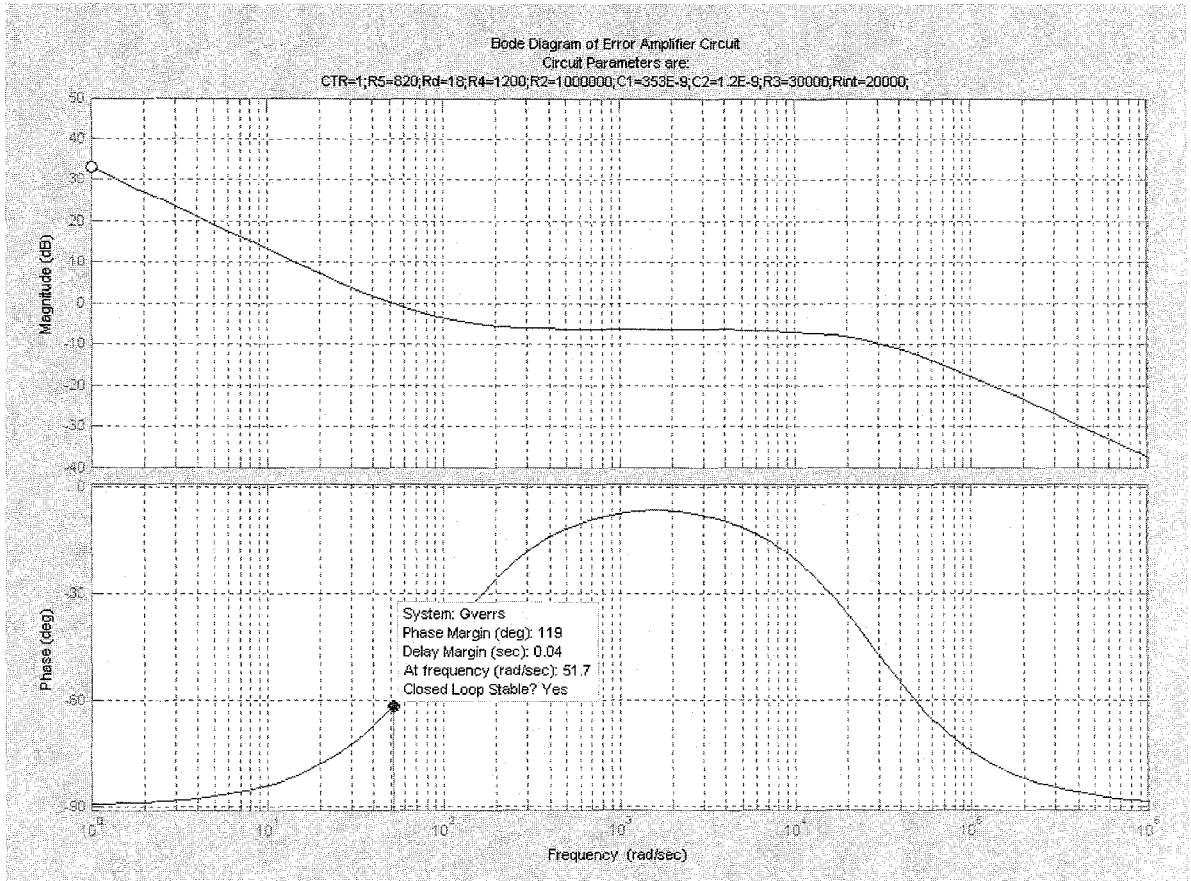


Figure 4.31. Bode diagram of the regulator.

The procedure for selection of R_{24} , C_{22} and C_{23} is as follows:

The poles of the power stage at the maximum and the minimum load are:

$$R_{LH} = \frac{V_o^2}{P_{o \max}} = \frac{(1000V)^2}{300W} = 3333.33\Omega$$

$$R_{LL} = \frac{V_o^2}{P_{o \min}} = \frac{(1000V)^2}{5W} = 200\,000\Omega$$

$$w_p = \frac{1 - K_r \times R_L}{C_o \times (R_{esr} + R_L - K_r \times R_L \times R_{esr})}$$

The gain and phase of control to output transfer function at gain crossover frequency (for $K_e = 0.11$, $K_r = -1.78E-4$, $R_l = 3333\Omega$, $R_{esr} = 1\Omega$) is:

$$G_{vc} = |G_{vc}(2\pi f_c)| = 5.8dB$$

$$\Phi_{vc} = -86^\circ$$

The error amplifier should correct the phase and gain of the power stage to a desired level.

Gain and phase of the error amplifier is:

$$G_{err} = G_{err}(2\pi f_c) = \frac{1}{G_{vc}} = -5.8dB$$

$$\Phi_{err} = -180^\circ + (\Phi_{pm} + \Phi_{m1}) - \Phi_{vc}$$

Here Φ_{pm} is the desired phase margin, Φ_{m1} is the phase shift caused by the RHP zero. Since we selected the gain crossover frequency to be smaller than $0.2f_{smin}$ we can consider it to be zero. Thus

$$\Phi_{err} = -180^\circ + (60^\circ + 0) - (-86^\circ) = -34^\circ$$

With these values we can place the pole and the zero using the zero and pole equations of the error amplifier. The zero frequency is placed to:

$$f_{zerr} = \frac{w_z}{2\pi} = \frac{\alpha}{2\pi} \times \frac{1+D}{R_{load}C_o}$$

α is a constant between 0.1 and 5.

$$f_{zerr} = \frac{0.1 \text{ to } 5}{2\pi} \times \frac{1+0.5}{3333\Omega \times 3 \times 10^{-6}F} = 2.39 \text{ to } 119.5Hz$$

We selected 15Hz for the zero frequency so the value of C_{23} when we select $100k\Omega$ for R_{24} is;

$$C_{23} = \frac{1}{2\pi R_{24} f_z} = \frac{1}{2\pi \times 100000\Omega \times 15Hz} = 106 \times 10^{-9}F \text{ (we selected } 100nF)$$

The pole frequency is placed to:

$$f_p = \frac{w_p}{2\pi} = \frac{f_c}{\left| \tan\left(\frac{\pi}{180}\Phi_{err}\right) \right|} = 4447Hz$$

$$C_{22} = \frac{C_{23}}{(1 + 2\pi f_p R_{24} C_{23})} = \frac{100 \times 10^{-9} F}{1 + 2\pi \times 4447Hz \times 100000\Omega \times 100 \times 10^{-9} F} = 0.359 \times 10^{-9} F$$

We selected 380pF.

5. THE TRIGGERING UNIT (TU)

5.1 Introduction

The TU is the signal shaping circuit, which uses audio signal from a laryngeal microphone to extract the fundamental frequency of the vocal folds for firing the xenon flash tube. A block diagram of this unit is presented in Fig. 5.1.

As shown in the block diagram the audio signals generated by vocal fold vibrations are picked up by a laryngeal (or air coupled) microphone and amplified. This signal is optically isolated from the patient and is passed through a notch filter to reduce the 50Hz mains noise. After the notch filter, the signal is first passed thorough a 1000Hz low pass filter and then through a 70Hz high pass filter. This signal is amplified using an instrumentation amplifier. After the amplification it is passed through an integrator (Integrator 1). The instrumentation amplifier output is inverted through an inverting amplifier¹ and integrated using a second integrator (Integrator 2). Both integrator outputs are compared using a comparator. The comparator output is the desired triggering signal. This signal is first applied to the gate of a trigger SCR that drives the triggering transformer of the Xenon flash tube.

The details of this unit are presented in the next sections.

5.2 Microphone Preamplifier and Isolation

The TU evaluates audio signals it receives from a laryngeal microphone and generates the necessary triggering pulses. In order to provide isolation between the patient side and the device an optical isolator that is powered from an external battery is used. An amplifier is used to amplify the microphone signal, which then is passed through filters. The schematic diagram of this circuit is shown in Fig. 5.2.

¹ This amplifier is located physically within the Integrator 1 unit.

5.3 Filter Circuits

In VLS we are interested in signal frequencies between 70Hz to 1000Hz. The microphone signal after amplification is passed through three filter stages (Fig. 5.1). The first stage is a notch filter which is used to reduce the 50Hz ripple; the second is a low pass filter which has a cut-off frequency of 1000Hz and the last one is a high pass filter with a cut-off frequency of 70Hz. Low and high pass filters are 8th order filters. The low frequencies are not so important but higher frequencies are important because if the low pass filter is not working and a signal frequency higher than 1000Hz is applied the energy per flash and so the power will increase linearly with the applied frequency. Thus, high frequencies should be attenuated to a level that it cannot cause any trigger action. The schematic diagrams of the filter circuits are presented in Figs. 5.3-5.5. and Fig. 5.7.

We constructed the low pass and high pass filters using Butterworth type second order stages. All four stages have different quality factor (Q) and the filter coefficients can be obtained from a filter catalog. For our circuit, Q values are 0.5098, 0.6013, 0.9001 and 2.563. All the stages have the same cut-off frequency. A summary of generalized second order low pass, high pass and notch filter transfer functions are presented in Table 5.1 [30]. Measured response characteristics of these filters are shown in Fig. 5.6 and Fig. 5.8.

Table 5.1.
Generalized Second Order Low Pass, High Pass and Notch Filter Transfer Functions

	Transfer function	Gain
Low Pass Filter	$T(s) = \frac{a_0}{s^2 + s \times \frac{w_0}{Q} + w_0^2}$	$A_{DC} = \frac{a_0}{w_0^2}$
High Pass Filter	$T(s) = \frac{a_2 \times s^2}{s^2 + s \times \frac{w_0}{Q} + w_0^2}$	$A_{HF} = a_2$
Notch Filter	$T(s) = a_2 \times \frac{s^2 + w_0^2}{s^2 + s \times \frac{w_0}{Q} + w_0^2}$	$A_{DC} = a_2$ $A_{HF} = a_2$

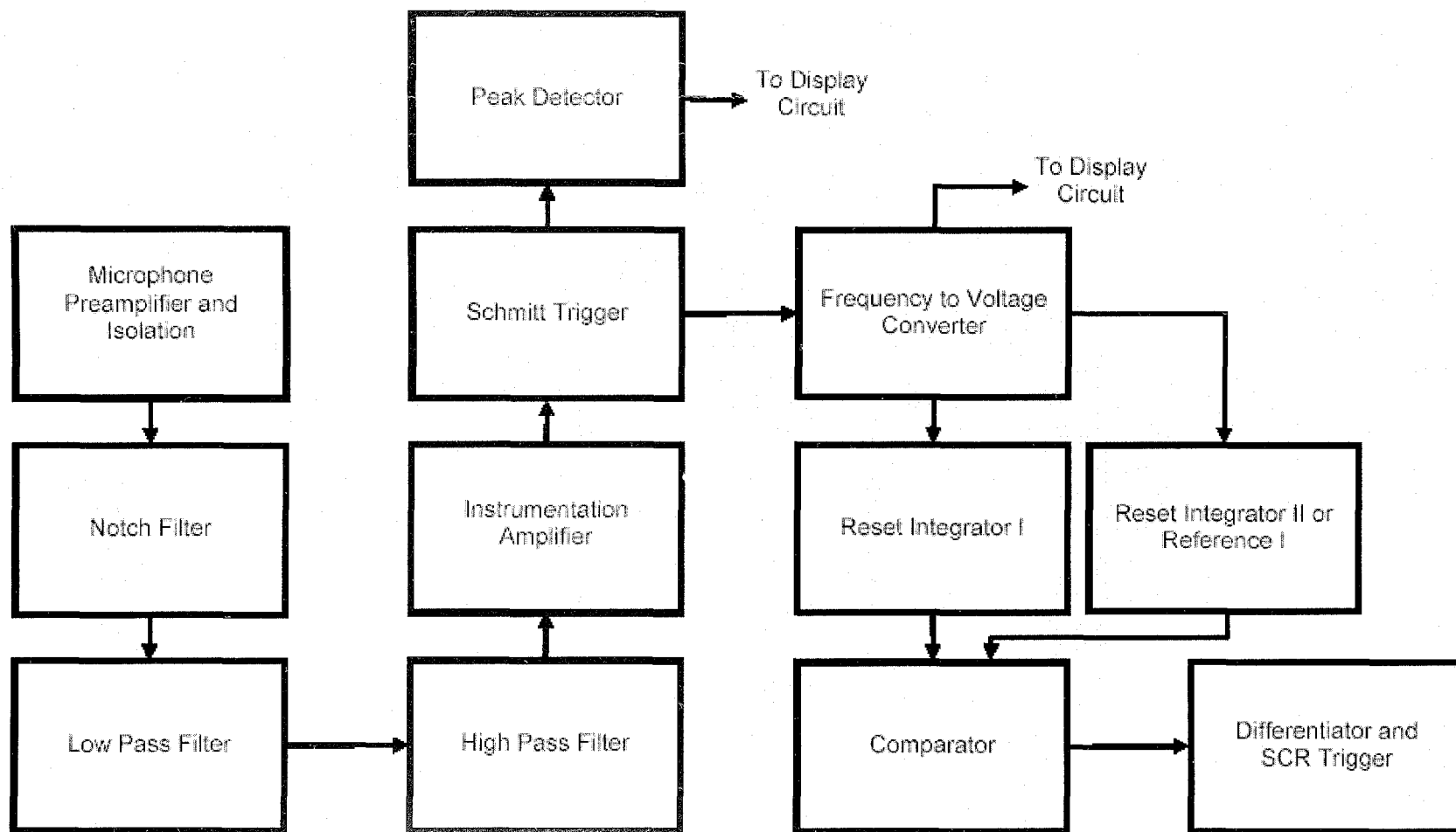


Figure 5.1 Block diagram of the trigger unit.

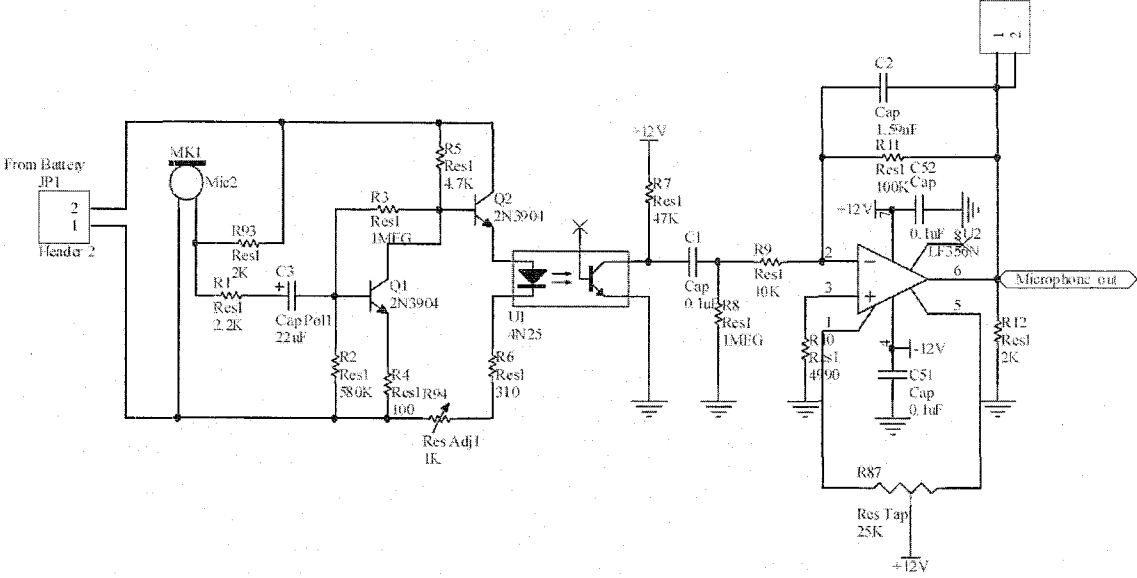


Figure 5.2. Microphone preamplifier and isolation circuit.

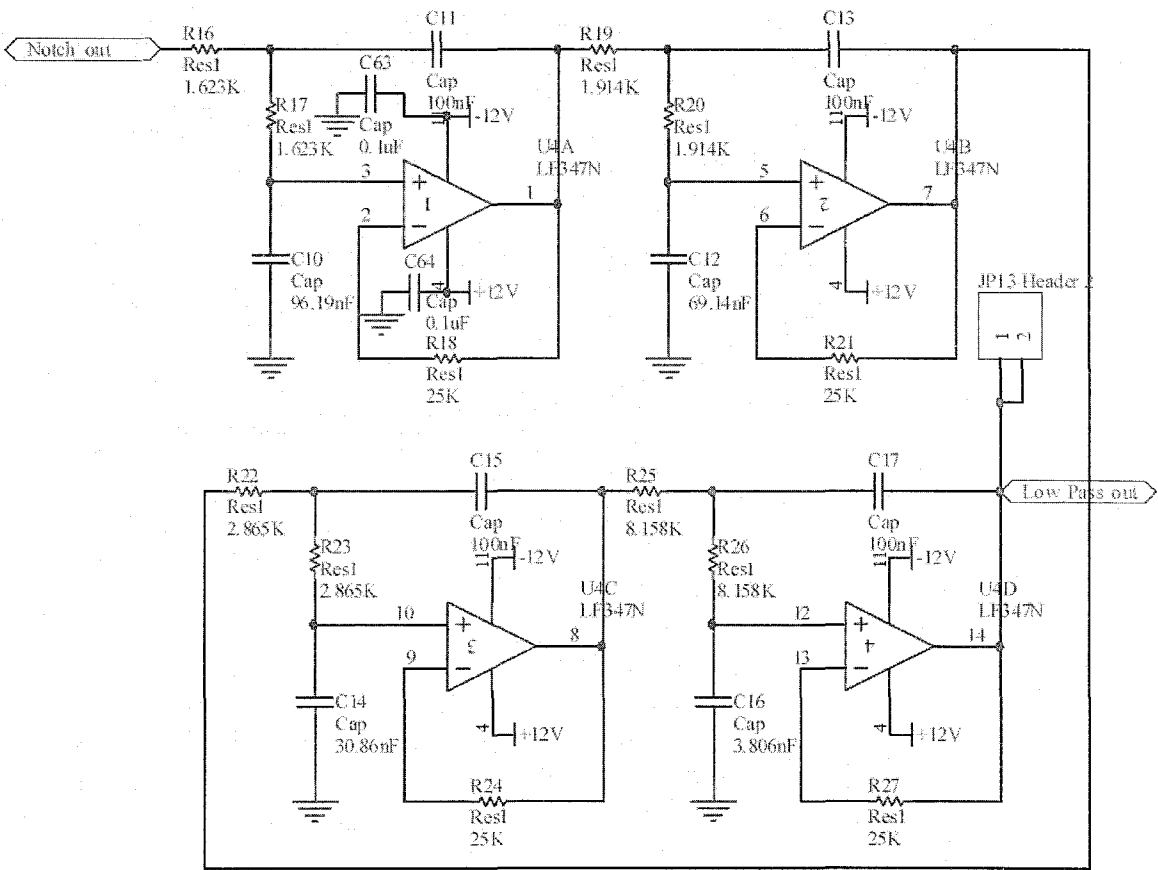


Figure 5.3. Schematic diagram of the low pass filter (C_{63} and C_{64} are ceramic capacitors).

Table 5.1.
Generalized Second Order Low Pass, High Pass and Notch Filter Transfer Functions

	Transfer function	Gain
Low Pass Filter	$T(s) = \frac{a_0}{s^2 + s \times \frac{w_0}{Q} + w_0^2}$	$A_{DC} = \frac{a_0}{w_0^2}$
High Pass Filter	$T(s) = \frac{a_2 \times s^2}{s^2 + s \times \frac{w_0}{Q} + w_0^2}$	$A_{HF} = a_2$
Notch Filter	$T(s) = a_2 \times \frac{s^2 + w_0^2}{s^2 + s \times \frac{w_0}{Q} + w_0^2}$	$\begin{aligned} A_{DC} &= a_2 \\ A_{HF} &= a_2 \end{aligned}$

The transfer functions, quality factor and cut-off frequency of high and low pass filters are dependent on the device values are given in the Table 5.2.

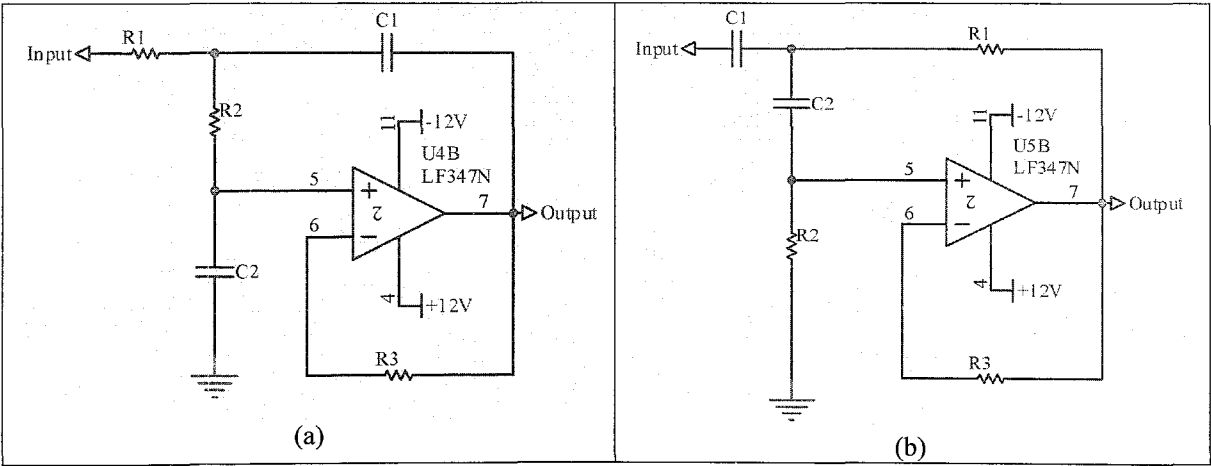


Figure 5.4. One stage of (a) low pass and (b) high pass filter.

Table 5.2.
One Stage of Low Pass and High Pass Filter Transfer Functions

	Low Pass Filter	High Pass Filter
$T(s)$	$\frac{1}{C_1 C_2 R_1 R_2} \times \frac{1}{s^2 + s \frac{R_1 + R_2}{C_1 R_1 R_2} + \frac{1}{C_1 C_2 R_1 R_2}}$	$\frac{s^2}{s^2 + s \frac{C_1 R_1 + R_1 C_2}{C_1 C_2 R_1 R_2} + \frac{1}{C_1 C_2 R_1 R_2}}$
w_0 (radian/sec)	$\frac{1}{\sqrt{C_1 C_2 R_1 R_2}}$	$\frac{1}{\sqrt{C_1 C_2 R_1 R_2}}$
Q	$\sqrt{\frac{C_1}{C_2}} R_1 R_2 \frac{1}{R_1 + R_2}$	$\sqrt{\frac{R_2}{R_1}} C_1 C_2 \frac{1}{C_1 + C_2}$

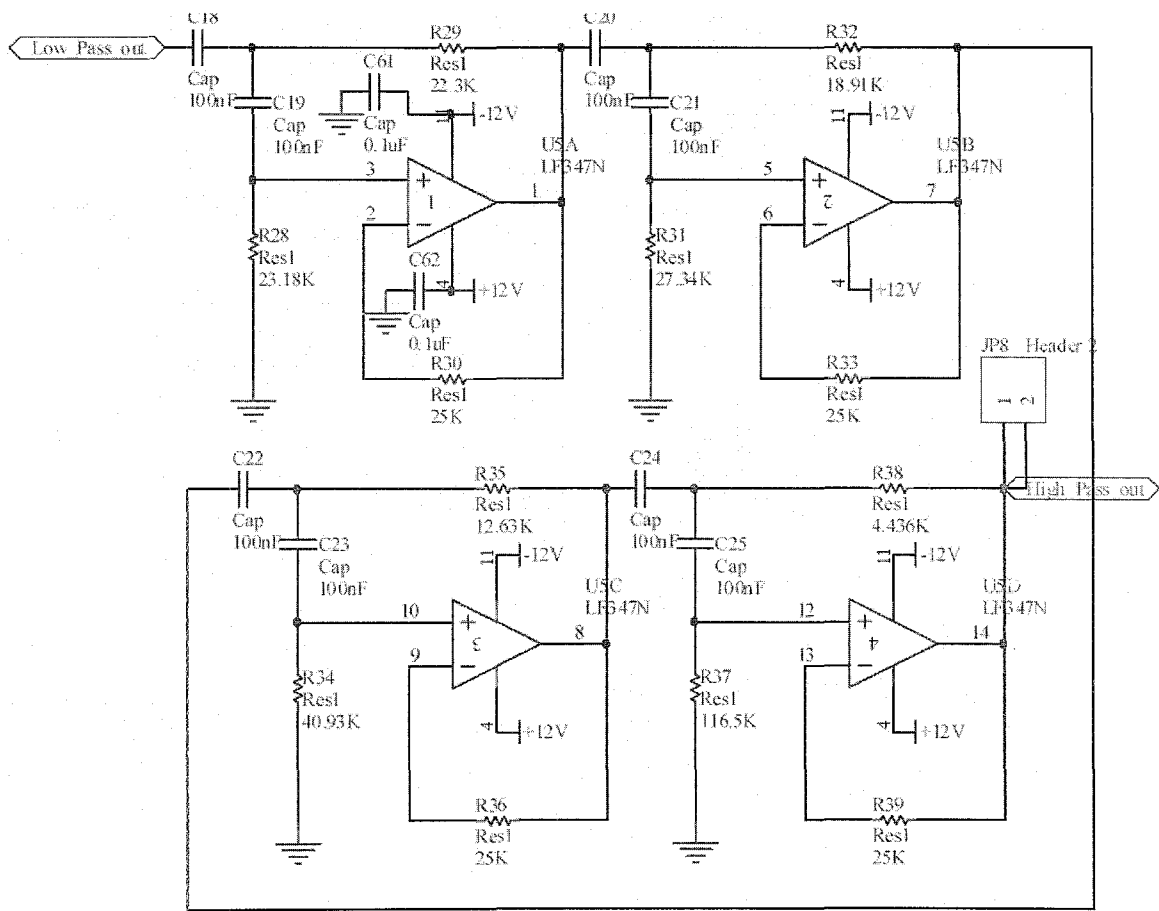


Figure 5.5. Schematic diagram of the high pass filter (C_{61} and C_{62} are ceramic capacitors).

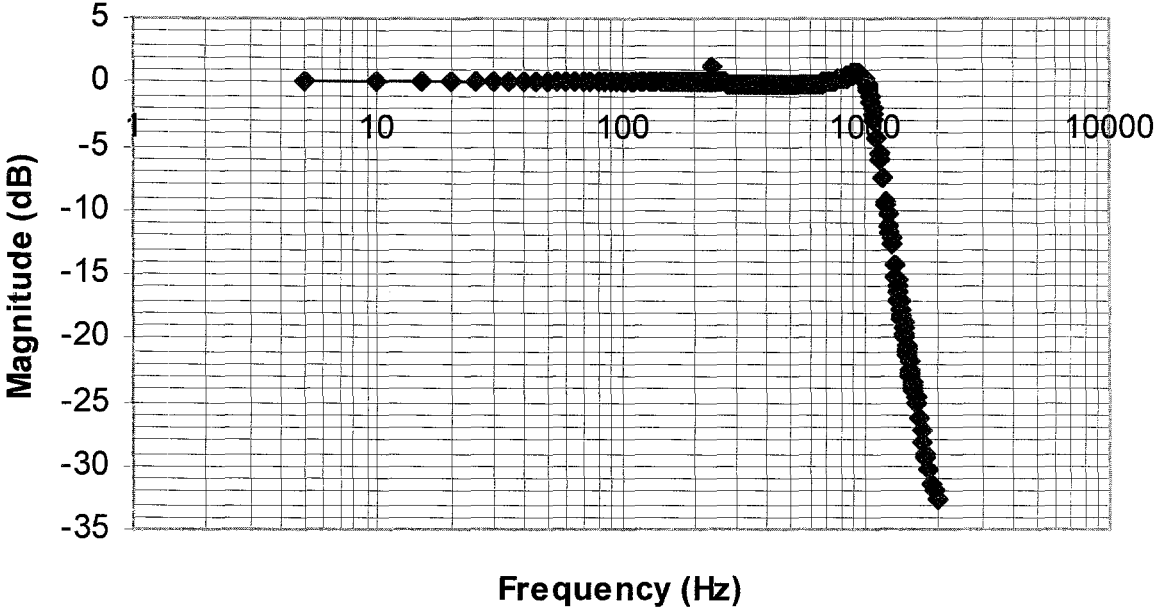


Figure 5.6. Measured low pass filter frequency response.

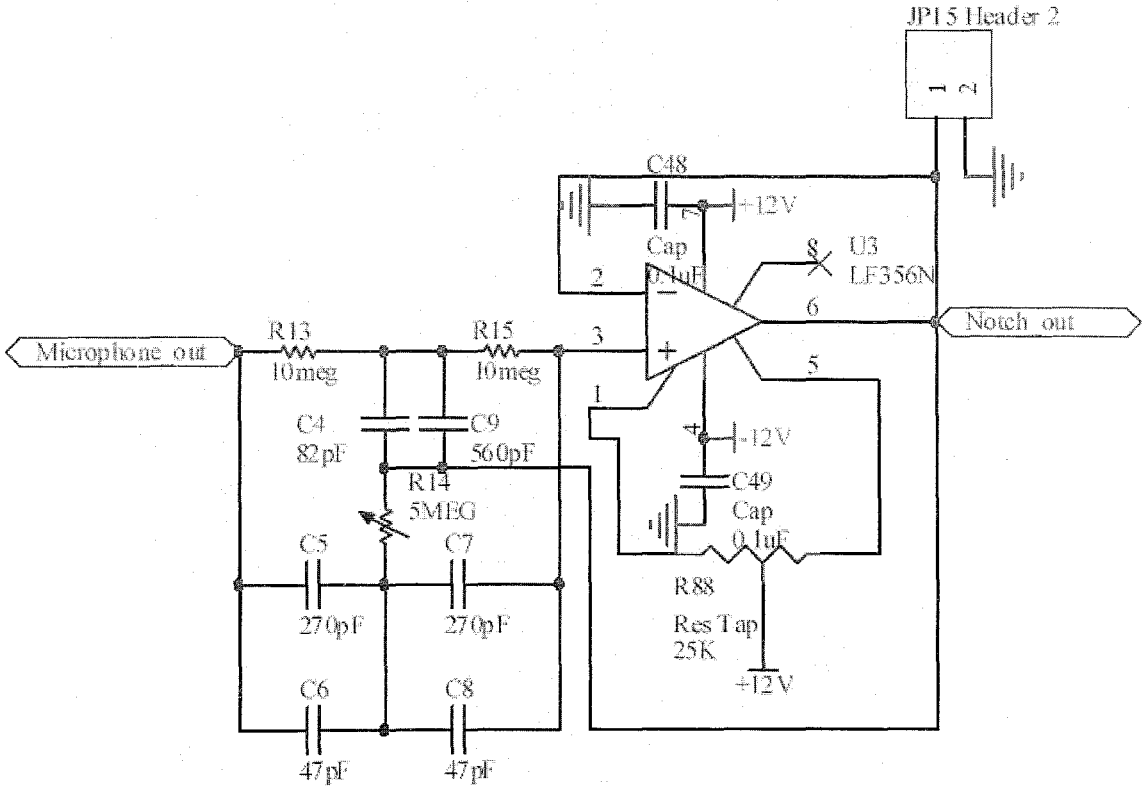


Figure 5.7. Notch filter for 50Hz rejection.

The notch frequency when

$$R_{13} = R_{15} = 2 \times R_{14} \text{ and } C_5 + C_6 = C_7 + C_8 = \frac{C_4 + C_9}{2} \text{ is [31]:}$$

$$f_0 = \frac{1}{2 \times \pi \times R_{13} \times (C_5 + C_6)} \quad (5.1)$$

It is advised in datasheets that capacitors are silvered mica or polycarbonate and 1% tolerance capacitors and resistors are 0.1% tolerance resistors.

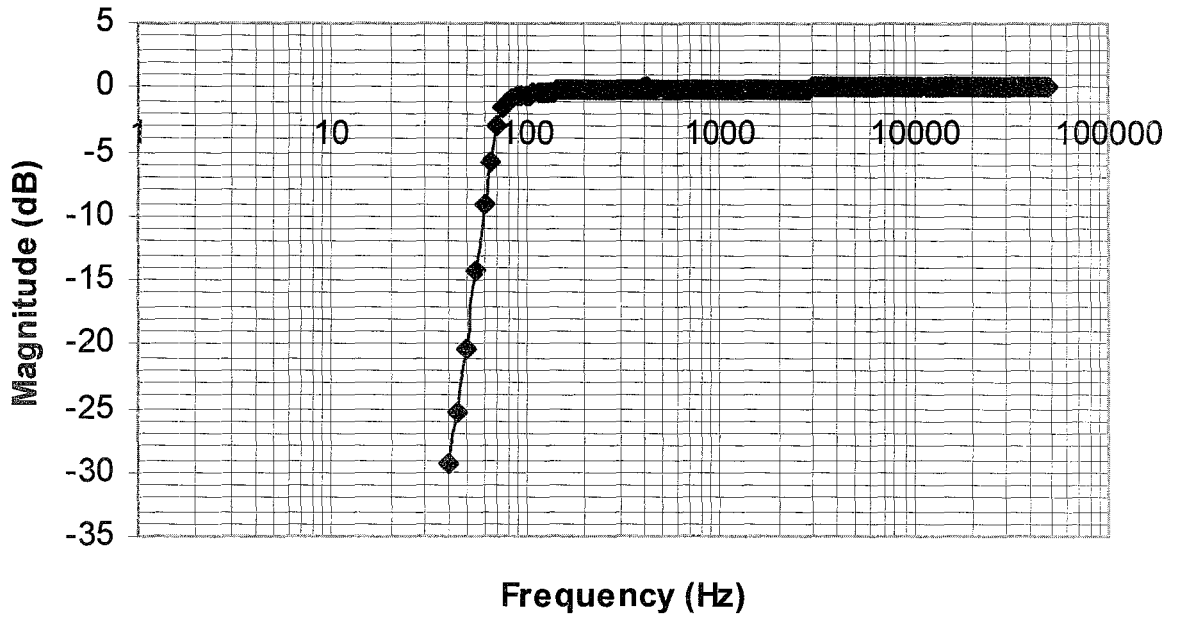


Figure 5.8. Measured high pass filter frequency response.

5.4 Instrumentation Amplifier (AD620)

In order to amplify the signal output from the filter circuits the AD620 instrumentation amplifier was used. According to the datasheet this amplifier has a CMRR of 100dB minimum at a gain of 10.

$$Gain = \frac{49400}{R_{gain}} + 1 \text{ (V/V)} \quad (5.2)$$

$$R_g = \frac{49400}{G-1} (\Omega) \quad (5.3)$$

0.33 μ F and 0.01 μ F capacitors in parallel are used at the power supply pins of the instrumentation amplifier. For AD620 by-pass capacitors should be placed between power supplies and the reference ground (pin 5). We connected reference terminal to ground and therefore placed by-pass capacitors between the power supply pin and ground. We also placed a small value capacitor to provide AC coupling at the input. The parallel resistor placed across the input terminal is especially important since it provides a ground return path for the DC input bias current which is necessary to bias the input transistor of the amplifier [32]. This resistor is typically 1M Ω or less. We also placed a high pass filter at the output to provide AC coupling. Schematic diagram of the amplifier circuit is shown in Fig. 5.9.

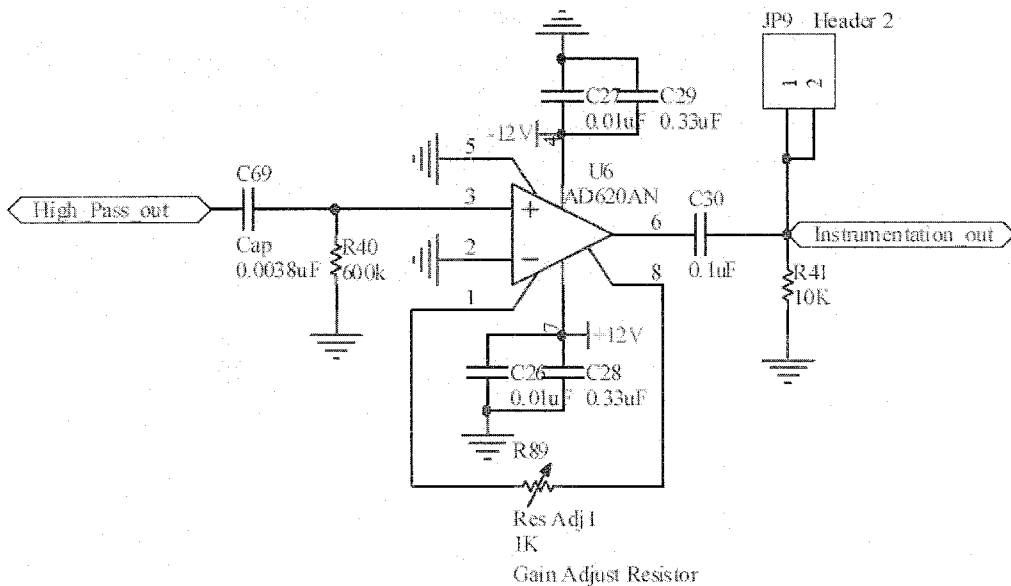


Figure 5.9. Schematic diagram of the instrumentation amplifier (C_{26} , C_{27} , C_{28} , C_{29}).

5.5 Schmitt Trigger

After the amplifier stage we used a Schmitt Trigger circuit to convert the input signal into a square wave at the same frequency. We used a LM311 comparator for this purpose. Schematic diagram of this circuit is given in Fig. 5.10.

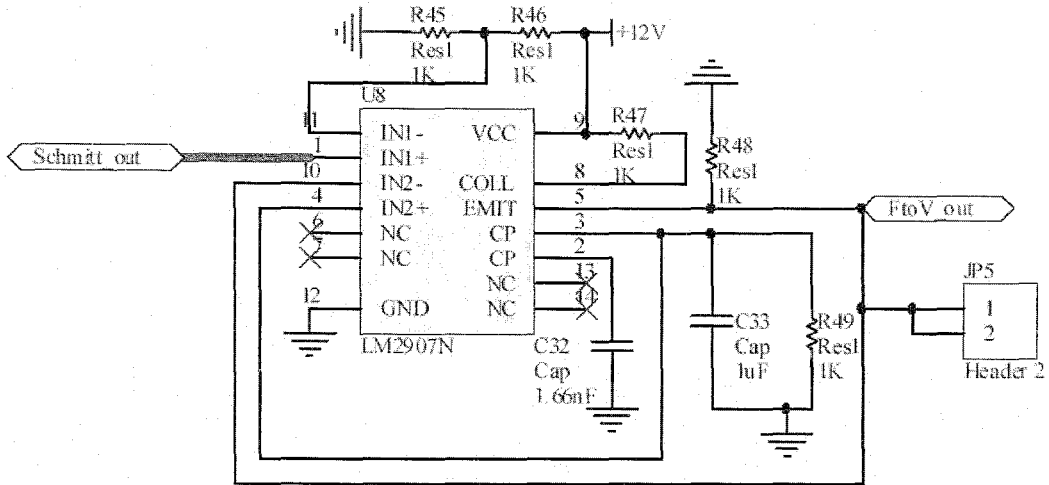


Figure 5.11. Schematic diagram of the frequency to voltage converter circuit.

The time it takes V_{out} to stabilize at a new frequency increases as the size of C_{32} increases. Three major factors that have effect on performance are ripple voltage, response time and linearity.

$$C_{32} = \frac{V_{3 \text{ fullscale}}}{R_{49} \times V_{cc} \times f_{\text{fullscale}}} \quad (5.6)$$

Selecting a ripple voltage:

$$C_{33} = \frac{V_{cc}}{2} \times \frac{C_{32}}{V_{\text{ripple}}} \times \left(1 - \frac{V_3}{R_{49} \times I_2} \right) \quad (5.7)$$

Here V_3 is the average voltage at pin 3.

$$V_3 = V_{cc} \times R_{49} \times C_{32} \times f \quad (5.8)$$

For our case $V_{cc} = 12V$, $C_{33} = 1\mu F$, $V_{3\text{max}} = 10V$, $I_{3\text{min}} = 150\mu A$, $I_2 = 200\mu A$, $C_{32} = 10nF$.

$$R_{49} \geq \frac{10V}{150\mu A} = 66667\Omega$$

We chose 100K for R_{49} .

When we are working at the lowest frequency of 70Hz the ripple voltage at the output is at its maximum and equals to:

$$V_{ripple} = \frac{12V}{2} \times \frac{10 \times 10^{-9} F}{1 \times 10^{-6} F} \times \left(1 - \frac{12V \times 70Hz \times 10 \times 10^{-9} F}{200\mu A} \right) = 57.48 \times 10^{-3} V_{pp}$$

When we are working at the highest frequency of 1000Hz the ripple voltage at the output is at its minimum and equals to:

$$V_{ripple} = \frac{12V}{2} \times \frac{10 \times 10^{-9} F}{1 \times 10^{-6} F} \times \left(1 - \frac{12V \times 1000Hz \times 10 \times 10^{-9} F}{200\mu A} \right) = 9.89 \times 10^{-3} V_{pp}$$

5.7 Peak Detector

We used a peak detector circuit to detect the peak value of the input signal. The peak level of the signal at the output of the instrumentation amplifier is converted into a DC signal. This DC signal can also be used to display the amplitude of the input signal. Schematic diagram of this circuit is shown in Fig 5.12.

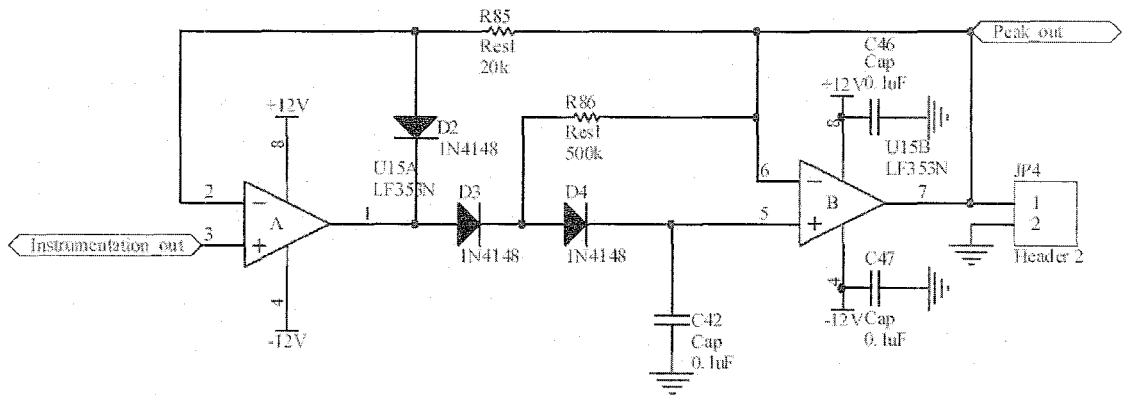


Figure 5.12. Peak detector circuit schematic.

5.8 Display

In the display unit we used a standard ICL7107 ADC to display the input signal frequency. Schematic diagram of the display unit is presented in Fig. 5.13.

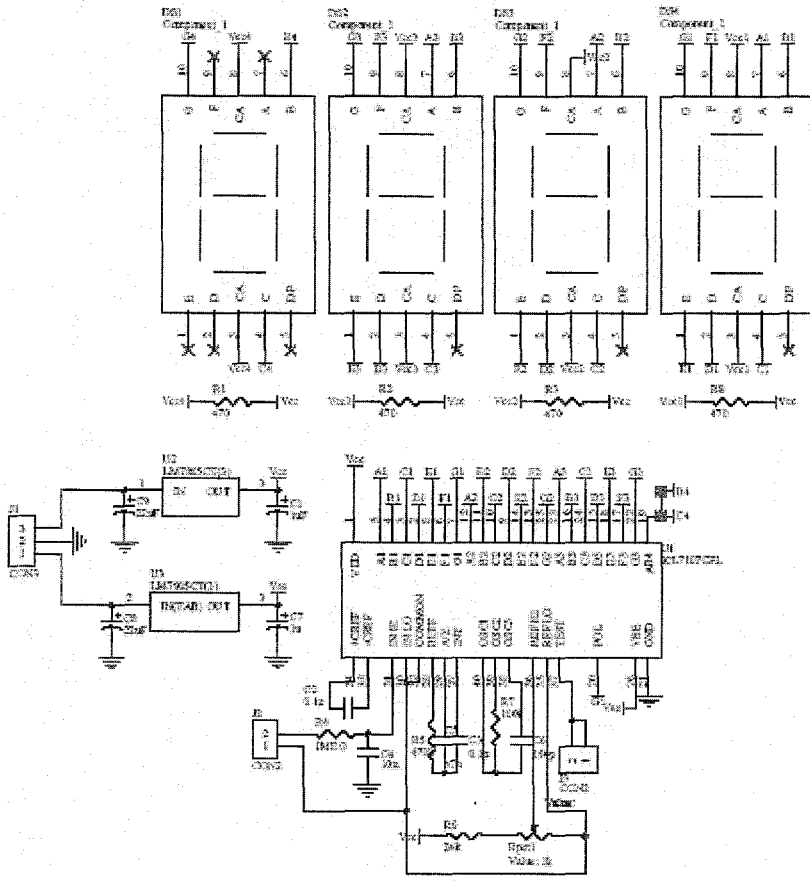


Figure 5.13. Schematic diagram of the display unit.

5.9 Reset Integrator I

The DC signal proportional to the input signal frequency is first inverted using an inverting amplifier then integrated using a reset integrator and converted into a saw tooth signal. This signal is compared with a reference signal to determine the position of the trigger signal. To produce a trigger pulse at the same position each time (corresponding to a standstill picture of the vocal folds) we used a resistive voltage divider with an adjustable potentiometer. To produce a variable position trigger pulse we used another integrator that has a longer time constant. Schematic diagram of the integrator circuits are shown in Fig 5.14 and Fig. 5.15.

Integrator calculations are as follows [33]:

Charging time is:

$$T_{ch} = \Delta t = \frac{C_{35}}{I} \times \Delta V \quad (5.9)$$

$$I = \frac{|V_{in}|}{R_{54}} \text{ and } \Delta V = V_T \quad (5.10)$$

$$f_0 = \frac{1}{T_{ch} + T_d} \quad (5.11)$$

$$f_0 = \frac{1}{\frac{R_{54} \times C_{35} \times V_T}{|V_{in}|} + T_d} \quad (5.12)$$

Here T_d is depended on the resetting circuit and it is small compared to T_{ch} but if we neglect it there can be errors. We choose the capacitor and resistor that minimizes the pulse on time. Further decrease in pulse time cannot be achieved because if it becomes smaller the integrating capacitor cannot be discharged.

5.10 Reference I (Resistive Voltage Divider)

Resistive voltage divider includes a potentiometer that allows adjustment of the reference voltage.

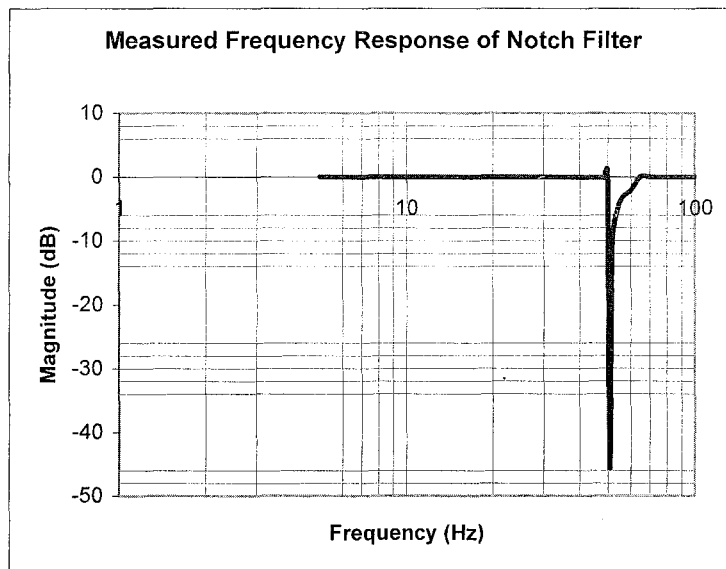


Figure 5.1 Measured frequency response of notch filter

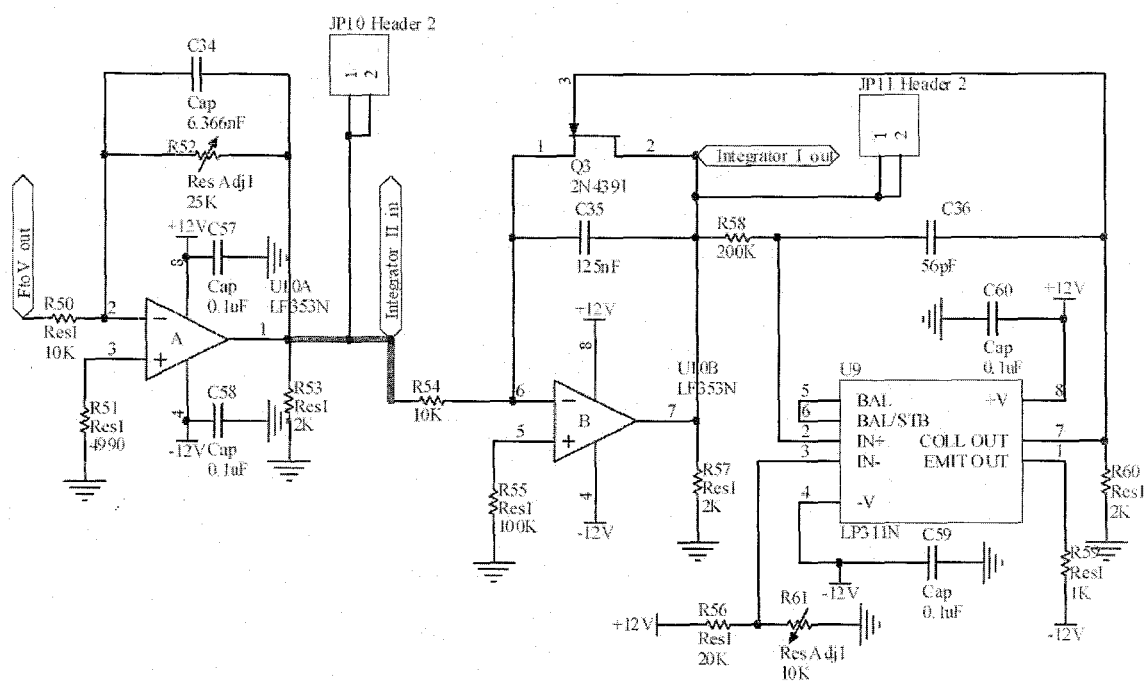


Figure 5.14. Reset integrator I circuit schematic including inverting amplifier at the input.

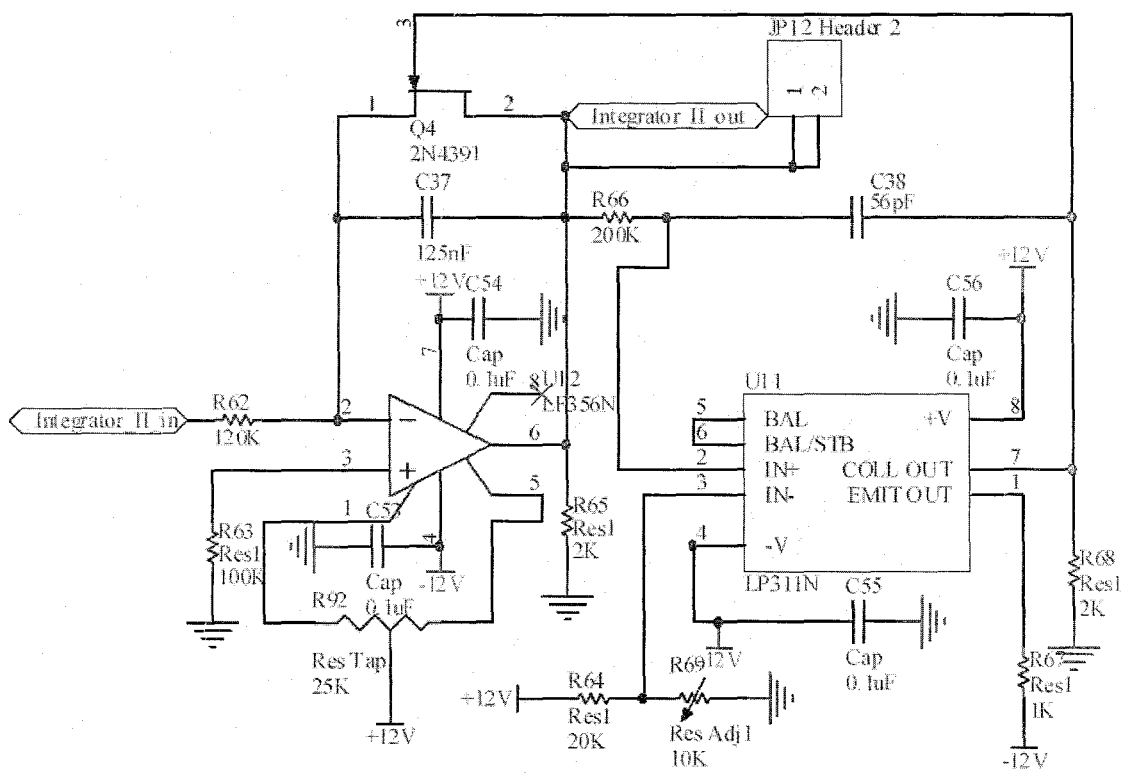


Figure 5.15. Reset integrator II schematic diagram.

5.11 Reference II (Integrator II)

Integrator II is the same as Integrator I but with a longer time constant. Because of the higher value resistance is bigger than the one in Integrator I, the saw tooth wave period is longer than Integrator I. The charging resistor is a potentiometer and increasing its value provides smaller time periods. Schematic diagram of this circuit is shown in Fig 5.15.

5.12 Comparator

The comparator compares the Integrator I output with the reference voltage. If the reference voltage is taken from a resistive voltage divider the trigger will occur at the same position every time, which means a standstill image. Adjusting the potentiometer allows to change triggering position. If, on the other hand, the reference voltage is taken from Integrator II output, the position of the trigger will change periodically and therefore different phases can be observed automatically and the vocal folds seems moving with a frequency determined by Integrator II. By adjusting the potentiometer of Integrator II the charging time and so the frequency of the motion can be adjusted. We used a LM311 comparator for this purpose. Schematic diagram of this circuit is shown in Fig 5.16.

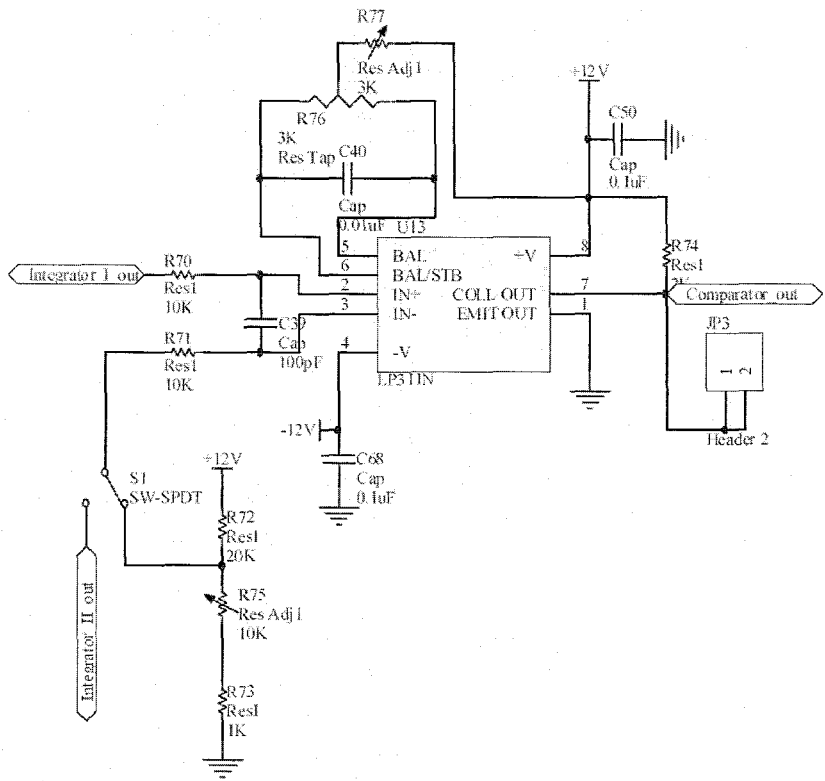


Figure 5.16. Comparator circuit schematic.

5.13 Differentiator and Isolated SCR Trigger

The pulse at the output of the comparator is differentiated and a pulse is formed. This pulse is transferred to SCR gate by an optic isolator. Schematic diagrams of this circuitry are shown in Fig 5.17 and Fig. 5.18.

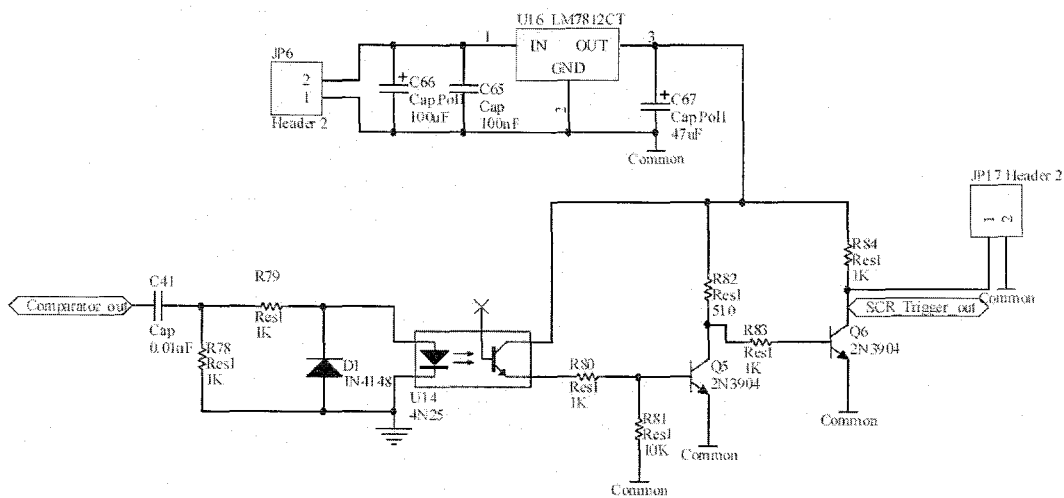


Figure 5.17. Isolated SCR trigger circuit schematic.

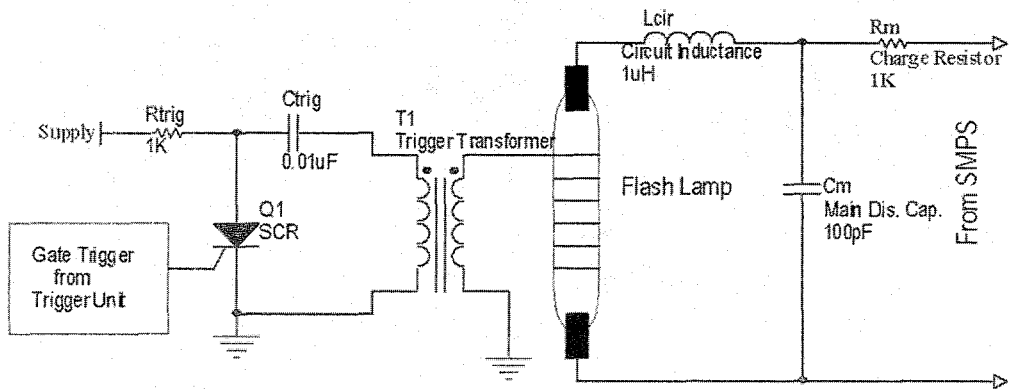


Figure 5.18. Flash lamp triggering circuit.

6. CONCLUSIONS

6.1 General Conclusions

In this work we designed and realized a power supply and triggering circuit for Video Laryngeal Stroboscopy VLS. The power supply is a Switched Mode Power Supply (SMPS) and uses a flyback topology, which is the topology used to generate high voltages in TVs, defibrillators and X-Ray units. Generally flyback power supplies are used below 150 or 200W. This power level can be increased through design optimization and through the choice of suitable devices of sufficient quality. The power supply has been tested at a power level of 250W. Since it operates in quasi-resonant mode, zero or low voltage switching is obtained that means minimization of MOSFET switching losses. Due to the frequency reduction at light loads, MOSFET switching losses are again minimized. In the absence of this property the switching frequency may reach to a value as high as 300kHz which approximately means a 6 times higher switching power loss in the MOSFET.

If desired, the power supply's mode can be changed from free-running to fixed frequency mode with a jumper setting. Moreover, the regulation characteristics can be changed from primary regulation to secondary regulation (via opto coupler feedback) with a jumper. The output voltage of the power supply can easily be adjusted by a potentiometer. This allows it to be used in other applications such as laser pumps, defibrillators, etc.

The power density of the power supply is about $3\text{W}/\text{inch}^3$, which is a good figure and reflects a state-of-the-art technology.

We also derived a method to predict the switching frequency of the SMPS. Using simplified equations we obtained transfer functions that describe the dynamic behavior of the unit.

The Triggering Unit (TU) is used to extract the fundamental frequency of vocal folds. Its input is an audio signal obtained from a laryngeal microphone. This signal is

amplified and filtered through a three-stage filter. Signal conditioning circuits are analog. Frequency is displayed using an ADC. The triggering signal at the TU output can take two forms. One is for the standstill images of vocal folds. The second is for obtaining mobile images of the vocal folds. A peak detector is also included in the TU and this can also be used to determine the peak amplitude of the speech after some calibration.

6.2 Recommendations of Future Work

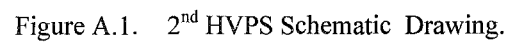
The following is a list of recommendations for future work:

1. The fan in the High Voltage Power Supply (HVPS) included to cool the power components at peak power levels consumes about 3 to 4 watts. It thus has an adverse effect on the efficiency. However, the supply mostly operates at lower power levels, at which the heat sinking used is sufficient, and there is no need for the fan. Using a temperature sensor like LM26 which is capable of driving fans directly and operating the fan only when needed, the efficiency at low power levels can be increased².
2. We placed a current-compensated choke and used its leakage inductance with an X type capacitor as a DM filter. However the leakage inductance is very low value and bigger X type capacitors are required. In order to overcome this problem we can use a separate DM inductor at the input.
3. Lower ESR capacitors can be used at the output.
4. In order to increase the efficiency and also the power level we should fight with leakage inductance that consumes a considerable power. Better coupling can be achieved by better winding arrangement as discussed in Chapter 3. Special winding techniques are available like interleaved winding.
5. Since the transformer was hand-wound, the coupling between the primary and the secondary is not very good. If winding is done with a winding machine better coupling can be achieved.

² This temperature sensor is not totally on or totally off, it can speed the fan as the temperature rises and slows down as it decreases.

6. The SMPS transfer functions can be derived by taking off time into account that will provide more accurate response characteristics.
7. In filter circuits and also in integrators the resistors should be of low tolerance and of high quality. Since there are more than 200 components on the triggering circuit, this results in an expensive circuit. Some of these triggering actions can be performed using a microcontroller.
8. The power supply is a source of EMI, the inductors and heat sinks behave like antennas. They should be screened properly and the power supply should be put into a shielded case. This case should be connected to earth to send CM currents back to its origins. Moreover, sensitive circuits like filter and integrator circuits should be away from the power supply. CM inductors can be used to attenuate CM currents at the amplifier inputs.

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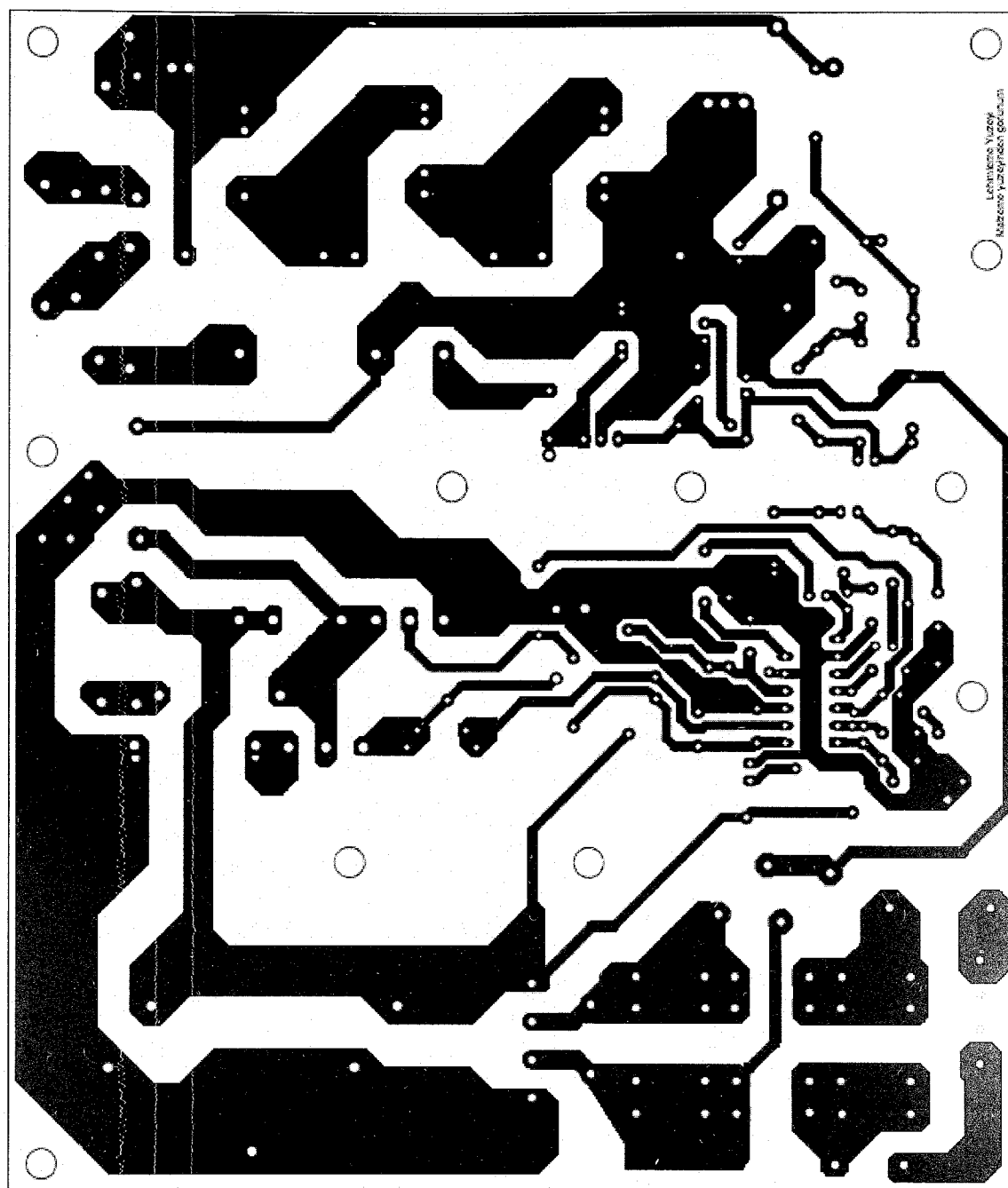


Figure A.3. 2nd HVPS PCB bottom layer.

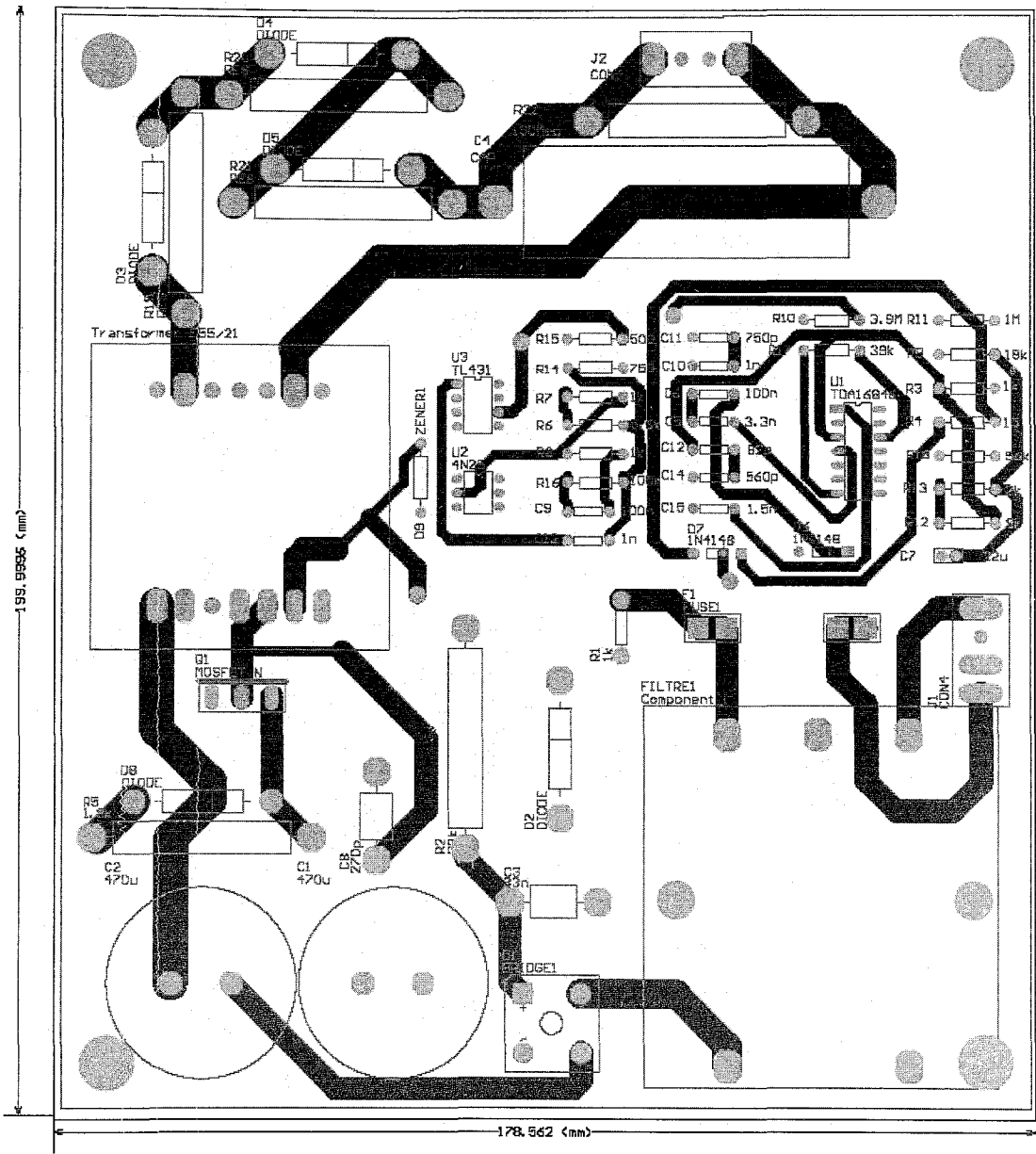


Figure A.4. 1st HVPS PCB bottom layer.

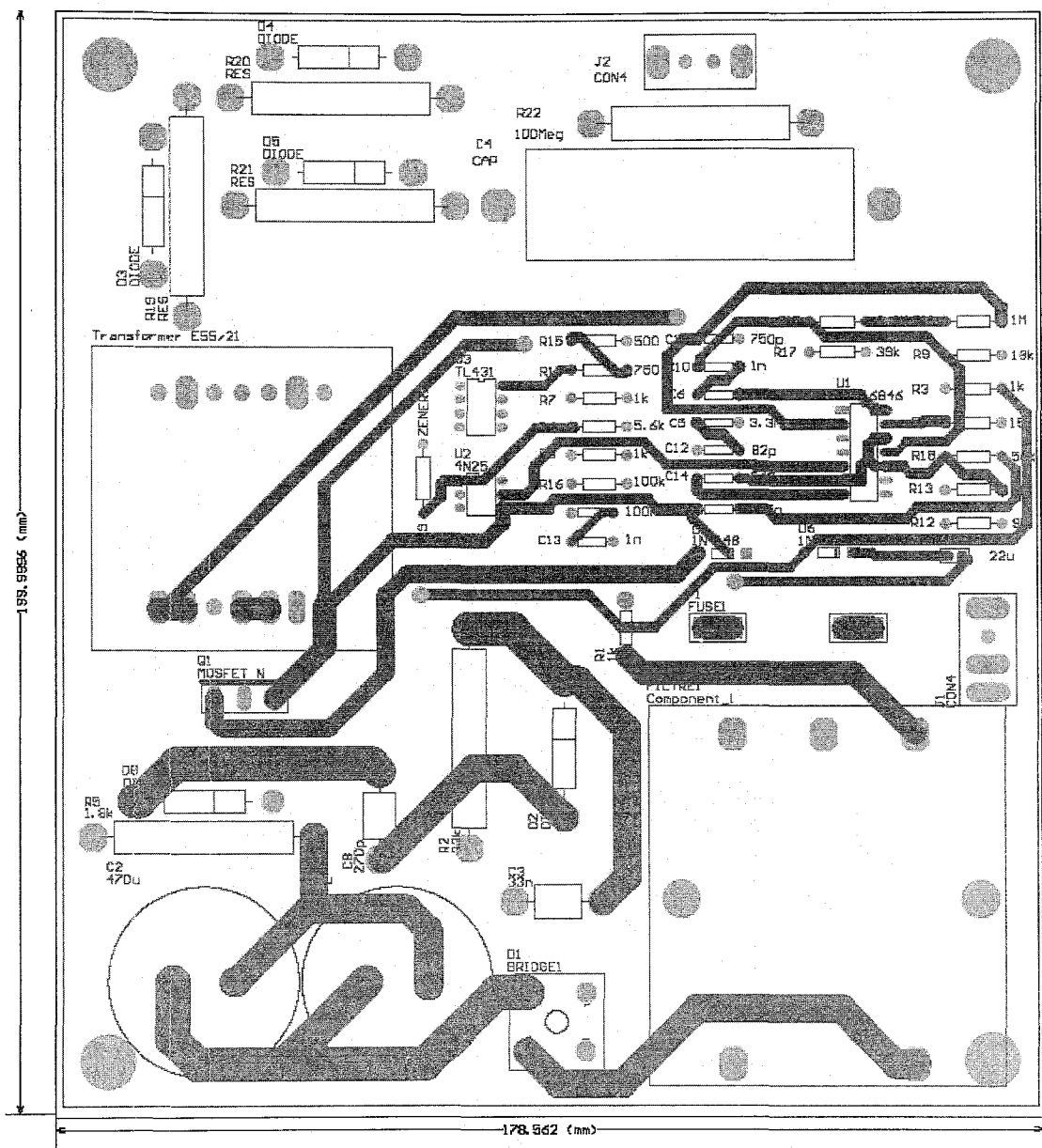


Figure A.5. 1st HVPS PCB top layer.

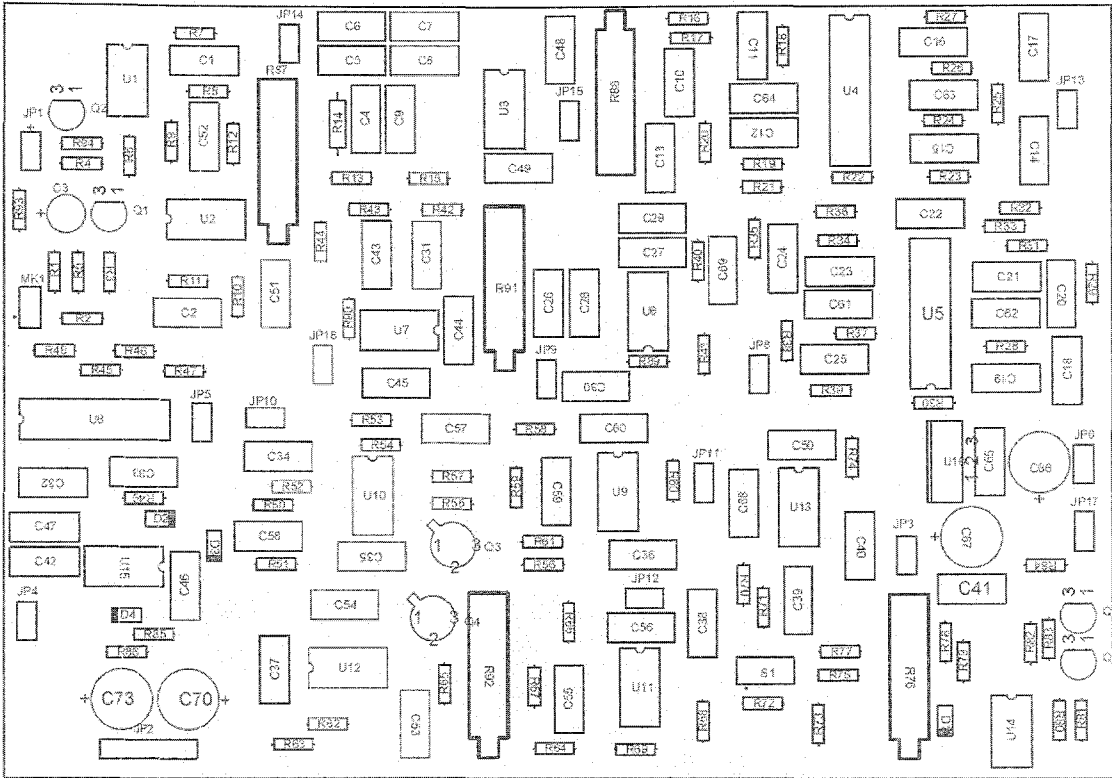


Figure A.6. TU PCB top overlay.

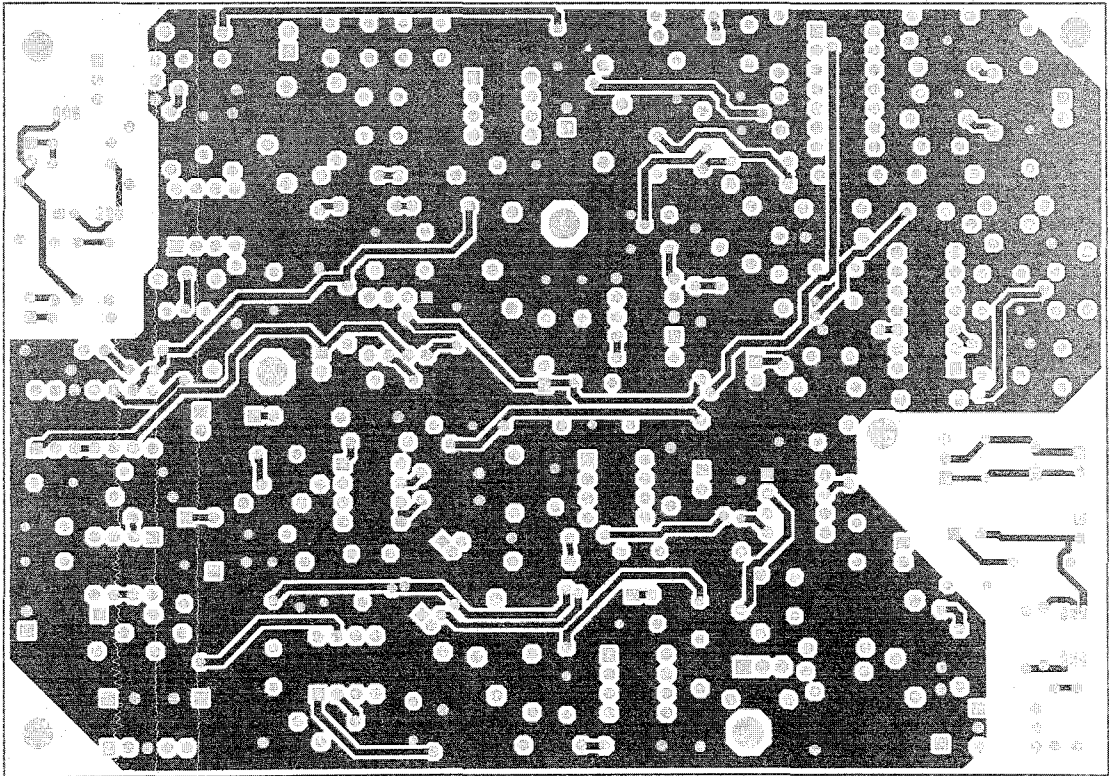


Figure A.7. TU PCB top layer.

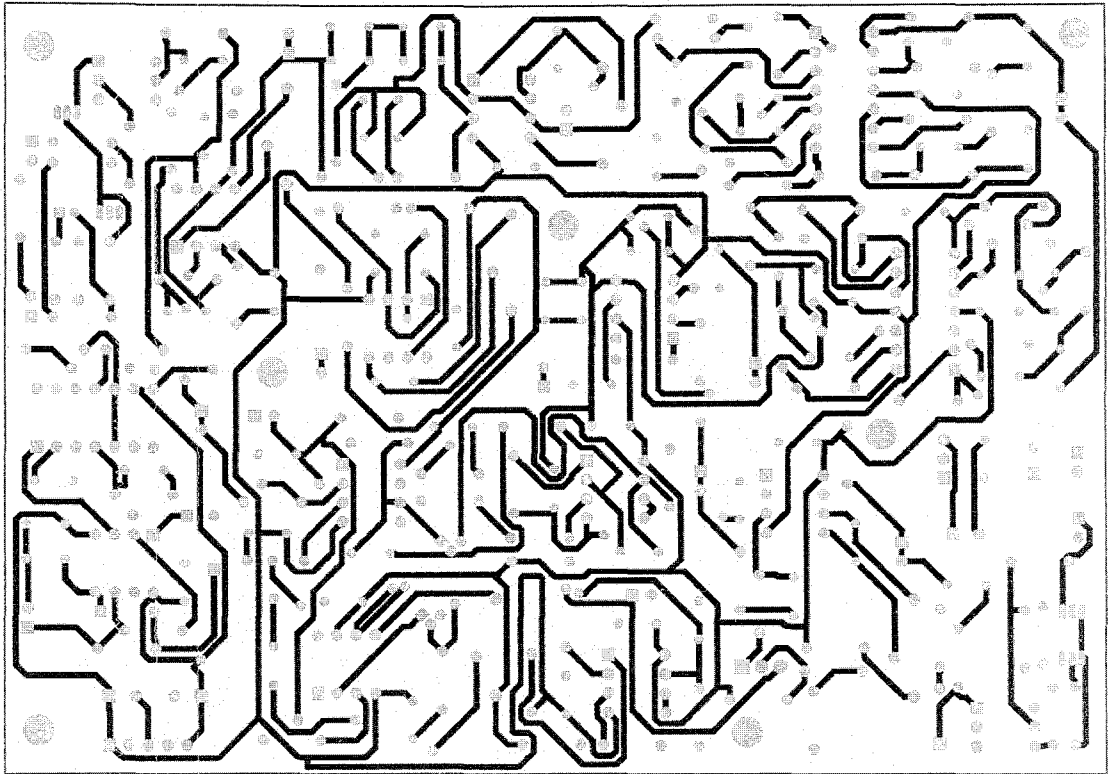


Figure A.8. TU PCB bottom layer.

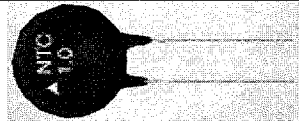
APPENDIX B: SOME SEMICONDUCTOR DATA

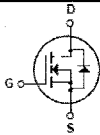
Output Rectifier Diode Specifications [34]		
MUR4100E		
Parameter	Value	Description
V_{RRM}	1000 V	Peak repetitive reverse voltage
$I_F(Av)$	4 A @ $T_a = 35^\circ C$	Average rectified forward current
T_j	-65 to $175^\circ C$	Maximum operating junction temperature
$R_{\theta ja}$	Depends on the mounting conditions ($\approx 51^\circ C/W$)	Junction to ambient thermal resistance
t_{rr}	$100 \times 10^{-9} s$	Maximum reverse recovery time

1N4148 Signal Diode (High Speed) [35]		
Parameter	Value	Description
V_{RRM}	100 V	Maximum repetitive reverse voltage
I_F	0,2 A	Average rectified forward voltage
$V_F @ I_F$	1 V @ 0,01 A	Forward voltage drop
$I_R @ V_{RRM}$	5 μ A @ VR = 75V	Reverse current
t_{rr}	4 ns	Reverse recovery time ($I_F = 10mA$, $V_R = 6V$ (60mA), $I_{rr} = 1mA$, $R_L = 100 \Omega$)

Bridge Rectifier [36]		
KBU8K		
Parameter	Value	Description
V_{RRM}	800 V	Maximum Repetitive Reverse Voltage
V_{RMS}	560 V	Maximum RMS Bridge Input Voltage
V_R	800 V	DC Reverse Voltage (Rated V_R)
$I_F(Av)$	8 A	Average Rectified Forward Current, @ $T_a = 50^\circ C$
I_{FSM}	300 A	Non-repetitive Peak Forward Surge Current
T_{sig}	-55 to $+150^\circ C$	Storage Temperature Range
T_j	-55 to $+150^\circ C$	Operating Junction Temperature
Thermal Characteristics		
P_D	6,9 W	Power Dissipation
$R_{\theta JA}$	18 $^\circ C/W$	Thermal Resistance, Junction to Ambient, per Leg
$R_{\theta JL}$	3 $^\circ C/W$	Thermal Resistance, Junction to Lead, per Leg
Electrical Characteristics		
V_F	1 V	Forward voltage, per bridge @ 8 A

Output Capacitor (Alcon Electronics)	
Parameter	Value
Capacitance	3μF
Rated Voltage	1000 VDC
ESR	2.5Ω

NTC Specifications							
							
R_{25} (Ω)	I_{max} (0...65°C)	No of R/T Characteristic	$B_{25/100}$ (K)	C_T 110V (μF)	C_T 110V (μF)	Parameters for $R(I)$	
						k	n
2,5	11	1203	2900	1000	4000	1,04	-1,32
Maximum power @ 25°C (P_{max})			5,1 W				
Resistance tolerance ($\Delta R_N/R_N$)			± 20 %				
Rated temperature (T_N)			25 °C				
B value tolerance ($\Delta B/B$)			± 3 %				
δ_{th} (approximately)			24mW/K (Dissipation factor in air)				
τ_c (thermal cooling time constant in air)			100 s (approximately)				
Heat capacity (C_{th})			2400mJ/K (approximately)				

MOSFET Specifications [21]		
Fairchild FQA13N80		
Parameter	Value	Description
V_{BVDSS}	800V	Maximum permissible drain to source voltage
$I_{d\ max}$	13A	Maximum permissible drain voltage
R_{dson}	0.75Ω	On state resistance
C_{oss}	$2.75 \times 10^{-10} F$	Output capacitance
T_r	$1.1 \times 10^{-7} s$	Turn-off fall time

Varistor Specifications		
Epcos S20 K300		
Parameter	Value	Description
V_{rms}	300V	Maximum RMS voltage
V_{DC}	385V	Maximum DC voltage
i_{max}	8000A (8/20 μs)	Maximum surge current
w_{max}	173J (2ms)	Maximum energy
P_{max}	1W	Maximum power consumption
Response Time	<25ns	-
Electric Strength	2.5kV _{rms}	-

Current Compensated Ring Choke Specifications		
Siemens B82723		
Parameter	Value	Description
L	2700 μ H	Inductance
I	4A	Rated current
n	2	Number of lines
R	60m Ω	Resistance
V_{rms}	250V	Rated voltage
L_{leak}	30-40 μ H	Leakage inductance

$$I_{DC} = \left(\frac{0.9 \times A_l}{K3} \right)^{\frac{1}{K4}} \text{ (A)}$$

Where A_l is in nH

Core Material Specifications [11]		
N27 Material		
Parameter	Value	Description
μ_i	2000 2000	Relative initial permeability
B_s	500mT	Saturation magnetization (B at 25 °C)
B_s	410mT	Saturation magnetization (B at 100 °C)
H_c	27 A/m	Coercive field strength (Hc at 25 °C)
H_c	19 A/m	Coercive field strength (Hc at 100 °C)
$T_c >$	220 °C	Curie temperature
f_{min}	25kHz	Optimum frequency range minimum frequency
f_{max}	150kHz	Optimum frequency range maximum frequency
$\tan \delta / \mu_i$		Relative loss factor of material at $H \rightarrow 0$
η_B	$1.5(10^{-6} / mT)$	Hysteresis material constant
α_f	$3(10^{-6} / K)$	Relative temperature coefficient of material
Density	4750 Kg/m ³	Material density
DC resistance	3Ωm	
$P_{v(f,B,T)}$	155 kW/m ³	Relative core losses

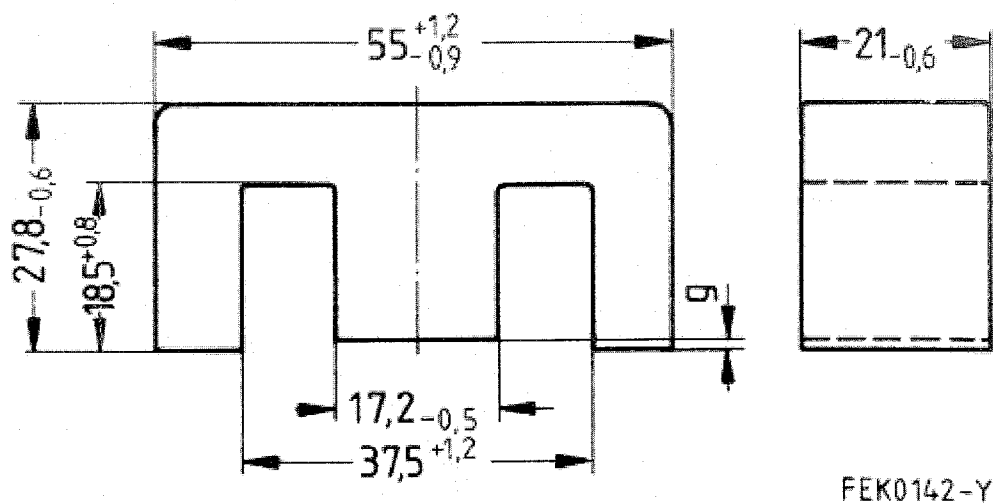


Figure C.1. E55/21 Ferrite core geometric dimensions

APPENDIX D: WIRE TABLE

Wire Table [14]							
Diam. Bare mm	Diam. Enamelled mm	Area mm ²	AWG no	Resistance at 20 °C Ω/km	Current at 3 A/mm ² mA	Length m/100g	Weight 100 g/km
0,04	0,05	0,0013	46	13700	3,8	8200	0,12
0,05	0,06	0,0020	44	8750	6	5400	0,18
0,06	0,07	0,0028	42	6070	9	3800	0,22
0,07	0,08	0,0039	41	4460	12	2800	0,35
0,08	0,09	0,0050	40	3420	15	2100	0,47
0,09	0,11	0,0064	39	2700	19	1700	0,59
0,10	0,12	0,0078	38	2190	24	1400	0,71
0,11	0,13	0,0095	37	1810	28	1100	0,91
0,12	0,14	0,011		1520	33	950	1,00
0,13	0,15	0,013	36	1300	40	820	1,21
0,14	0,16	0,015	35	1120	45	710	1,40
0,15	0,17	0,018		970	54	620	1,60
0,16	0,18	0,020	34	844	60	560	1,80
0,17	0,19	0,023		757	68	490	2,05
0,18	0,20	0,026	33	676	75	440	2,25
0,19	0,21	0,028		605	85	390	2,55
0,20	0,22	0,031	32	547	93	360	2,77
0,25	0,27	0,049	30	351	147	230	4,35
0,30	0,33	0,071	29	243	212	160	6,25
0,35	0,38	0,096	27	178	288	120	8,35
0,40	0,43	0,13	26	137	378	90	11,15
0,45	0,48	0,16	25	108	477	70	14,10
0,50	0,53	0,20	24	87,5	588	57	17,50
0,55	0,58	0,24		72,3	715	47	21,01
0,60	0,64	0,28		60,7	850	40	25,0
0,65	0,69	0,33	22	51,7	1,0A	34	29,4
0,70	0,74	0,39		44,6	1,16	29	34,5
0,75	0,79	0,44		38,9	1,32	25	40,0
0,80	0,84	0,50	20	34,1	1,51	22	45,5
0,85	0,89	0,57		30,2	1,70	20	50,0
0,90	0,94	0,64	19	26,9	1,91	18	55,5
0,95	0,99	0,71		24,3	2,12	16	62,5
1,00	1,05	0,78	18	21,9	2,36	14	71,5
1,10	1,15	0,95		18,1	2,85	12	83,5
1,20	1,25	1,1		15,2	3,38	10	100,0
1,30	1,35	1,3	16	13,0	3,97	8,5	118,0
1,40	1,45	1,5		11,2	4,60	7,5	140,0
1,50	1,56	1,8		9,70	5,30	6,4	155,0
1,60	1,66	2,0	14	8,54	6,0	5,5	179,0
1,70	1,76	2,3		7,57	6,7	5,0	200,0
1,80	1,86	2,6	13	6,76	7,6	4,5	225,0
1,90	1,96	2,8		6,05	8,5	4,0	250,0
2,00	2,06	3,1	12	5,47	9,40	3,5	285,5

APPENDIX E: PHOTOGRAPHS OF THE UNITS DESIGNED

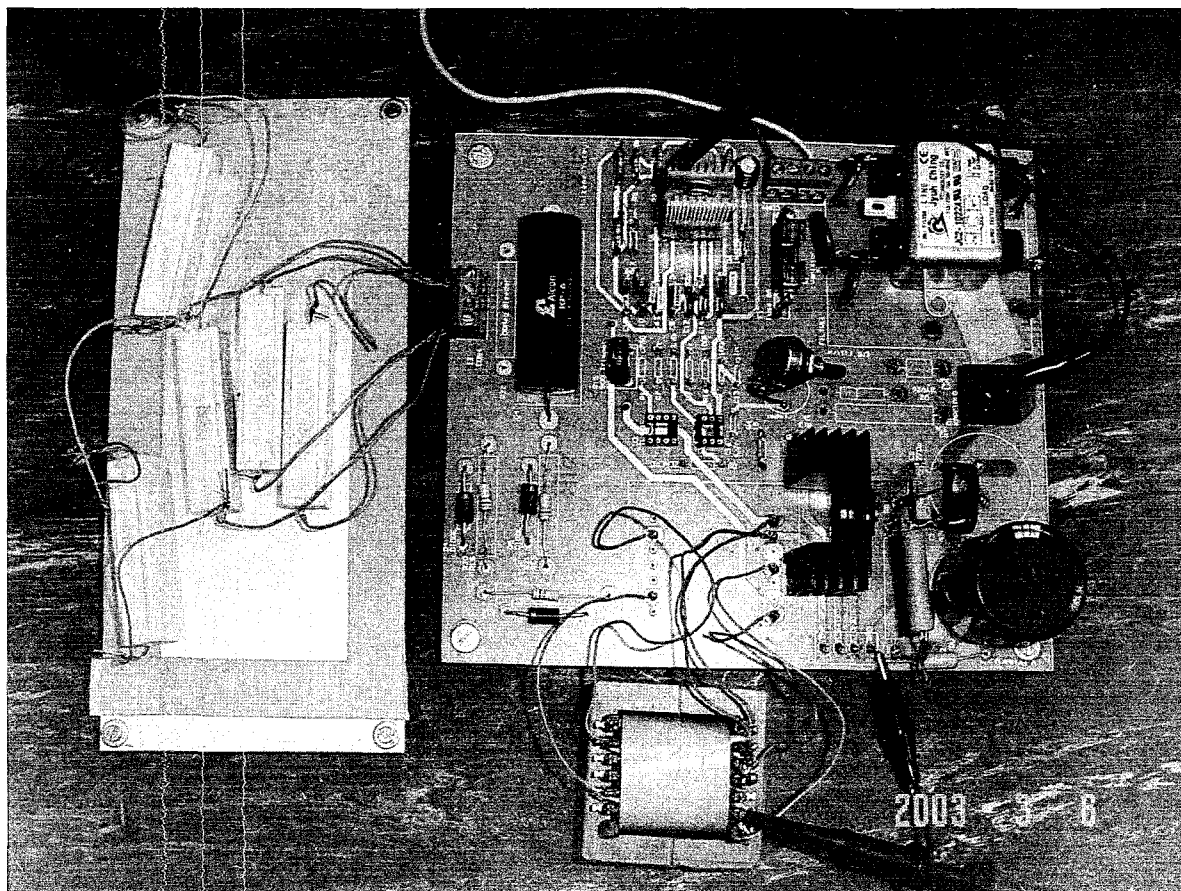


Figure E.1. 1st HVPS Unit PCB.

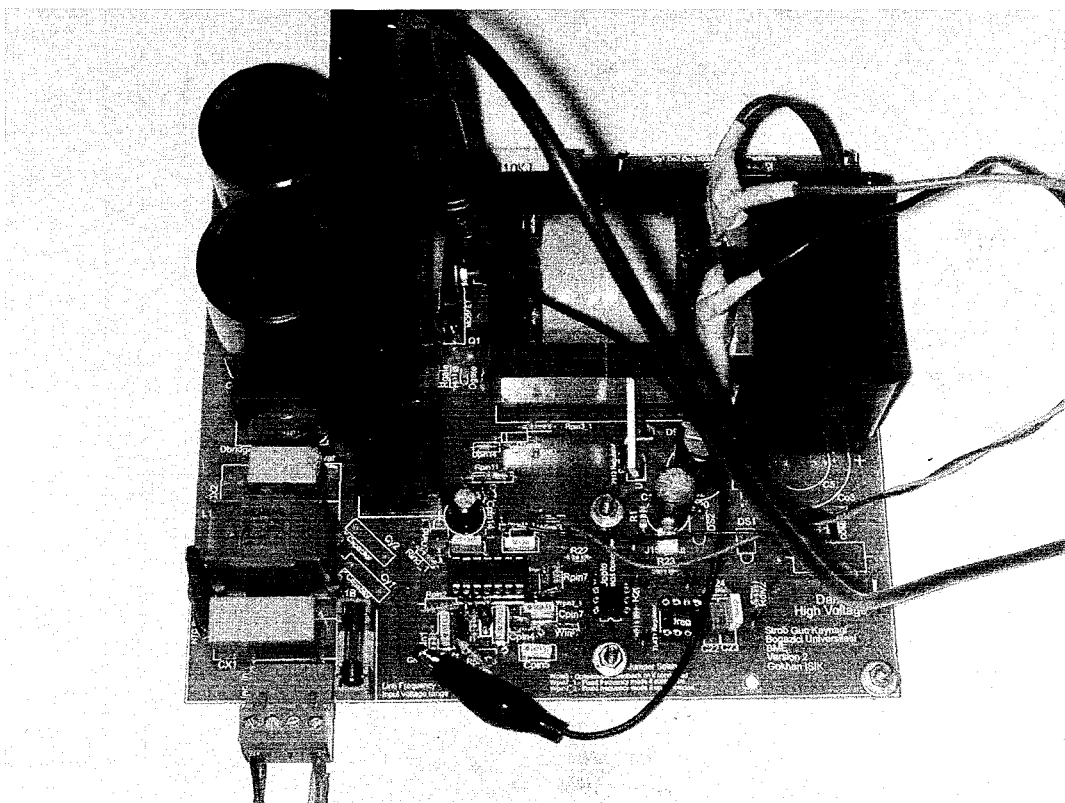


Figure E.2. 2nd HVPS Unit PCB.

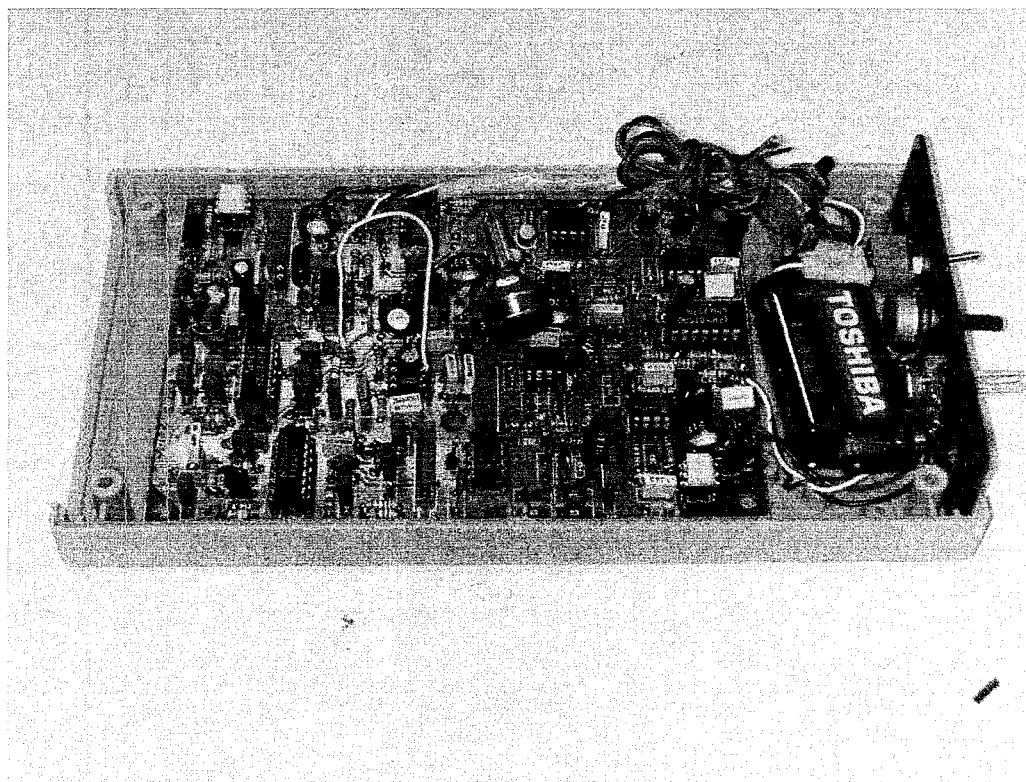


Figure E.3. TU PCB.

APPENDIX F: DRAIN VOLTAGE AND CURRENT WAVEFORMS AT DIFFERENT LOADS

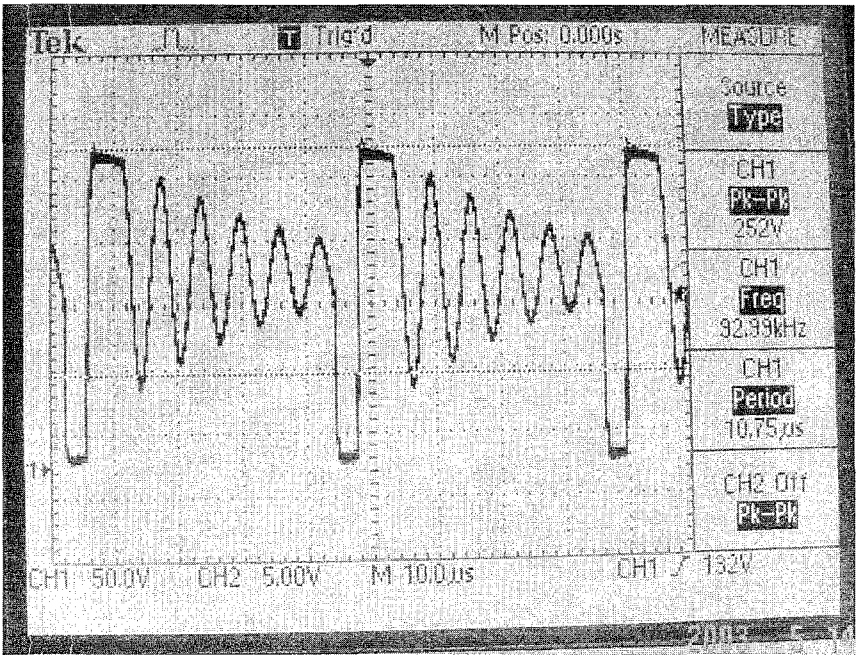


Figure F.1. Drain voltage at low power.

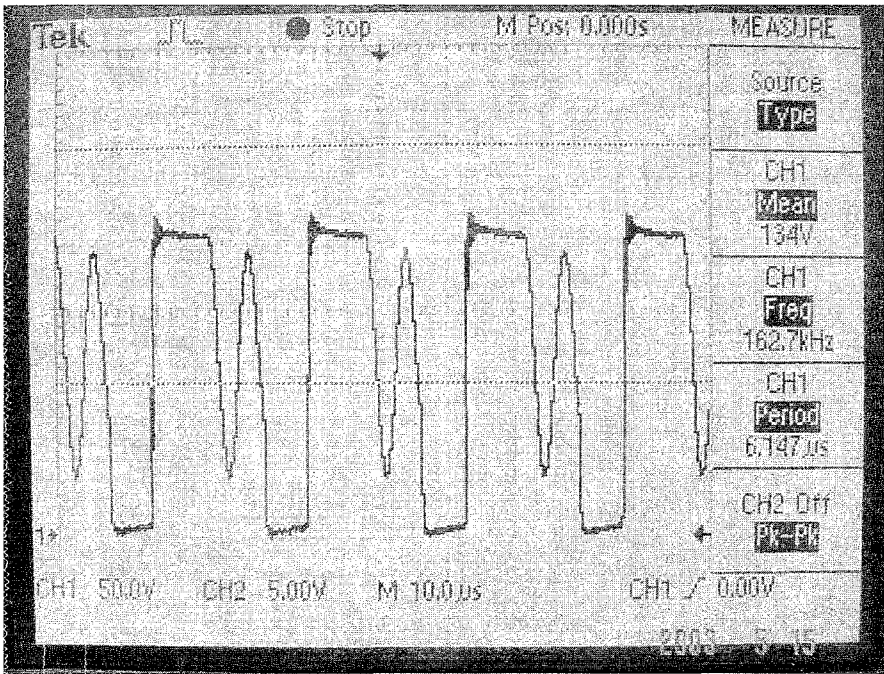


Figure F.2. Drain voltage at medium power.

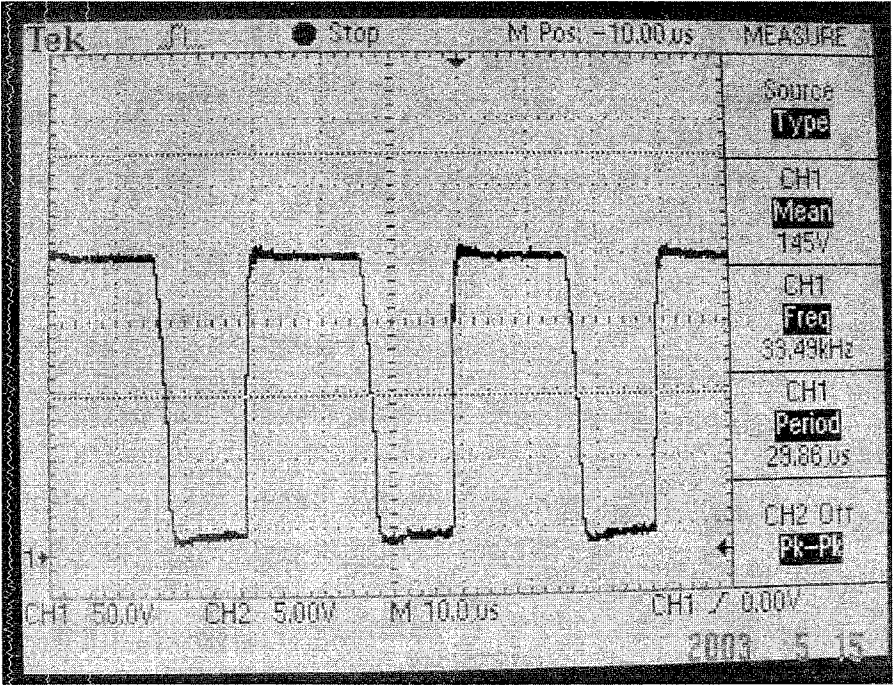


Figure F.3. Drain voltage at high power.

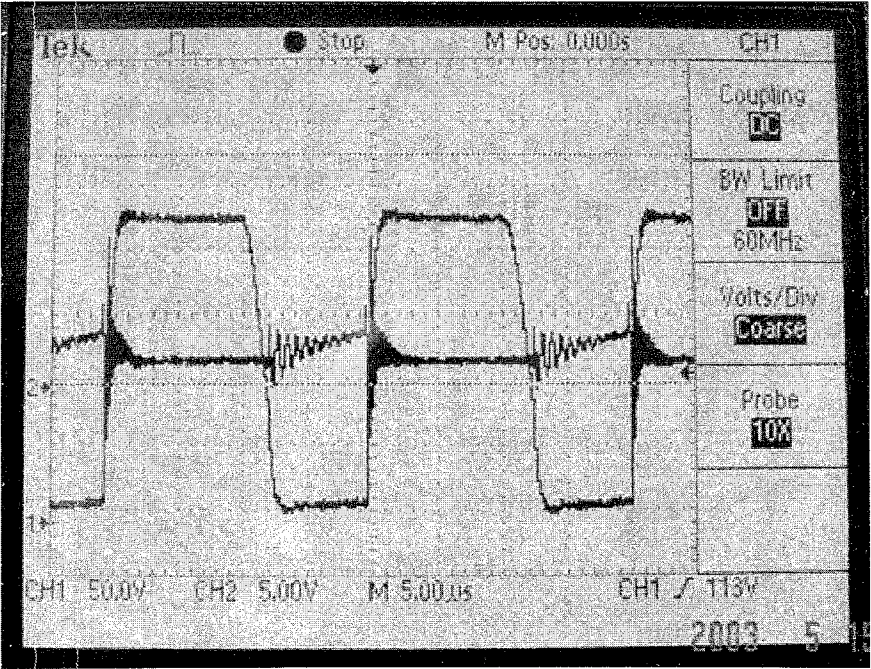


Figure F.4. Drain voltage and drain current at high power.

REFERENCES

1. Bless, D. M., M. Hirano, R.J. Feder, "Videostroboscopic Evaluation of the Larynx", *Ear, Nose and Throat Journal*, Vol. 66, pp. 48-58, July 1987.
2. Kitzing, P., "Stroboscopy – a Pertinent Laryngological Examination", *The Journal of Otolaryngology*, 14:3, pp.151-157, 1985.
3. Casiano, R.R., V. Zaveri, D.S. Lundy, "Efficacy of Videostroboscopy in the Diagnosis of Voice Disorders", *Otolaryngology Head Neck Surgery*, 107:95, 1992.
4. Weir, N., I. Bassett, "Outpatient Fiberoptic Nasolaryngoscopy and Videostroboscopy", *Journal of the Royal Society of Medicine*, Vol. 80. pp. 299-300. May 1987.
5. Yorulmaz İ., İ. Koçak, A. Demireller, Ç. Cuhruk, "Larinks Morfolojisinin ve Vokal Kord Vibratuar Hareketlerinin Değerlendirilmesinde Videolaringostroboskopi", *KBB Dergisi*, 2(1), pp. 65-70. 1993.
6. Tatlıpınar A., "Videolaringostroboskopi", Ankara Üniversitesi Kulak-Burun-Böğaz Hastalıkları Anabilim Dalı Seminer Arşivi, Nisan 1999.
7. Buckmire, R. A., "Stroboscopy", *eMedicine Journal*, Vol. 2, No. 9, September 2001.
8. Svec, J. "On Vibration Properties of Human Vocal Folds", PhD Thesis, University of Groningen, 2000.
9. Pressman, A.I., *Switching Power Supply Design*, McGraw-Hill, New York, 1992.
10. Kilgenstein, O., *Switched-mode Power Supplies in Practice*, John Wiley & Sons, New York, 1991.
11. EPCOS - Ferrite Magnetic Design Tool, Version 3.0., Epcos AG, 3/2003.
12. EPCOS Magnetic Materials Databook, Epcos AG, 2000.
13. Preller, P., AN-TDA 1684X - A Controller Family for Switch Mode Power Supplies Supporting Low Power Standby and Power Factor Correction (PFC), Application Note, Infineon Technologies AG.
14. *Electronic Components Catalogue*, Elfa, 2003.
15. Carter, T., "Switch Mode Power Supplies: An EMI Engineers Point of View", *Southcon/94 Conference Record*, 29-31 March 1994, pp. 295-300.
16. Pei, X., J. Xiong, Y. Kang and J. Chen, "Analysis and Suppression of Conducted EMI Emission in PWM Inverter", *Electric Machines and Drives Conference, IEMDC'03*, 1-4 June 2003, Vol. 3, pp. 1787 -1792, *IEEE International*, 2003.
17. *Rectifier Applications Handbook, Reference Manual and Design Guide*, HB214/D, On Semiconductor, 2001.
18. *Snap-in Capacitors, B43304, Datasheet*, Epcos AG, 2002.

19. *Application Bulletin AB-8, Selection of MOSFETs in Switch Mode DC-DC Converters*, Fairchild Semiconductor Corporation.
20. Barkhordarian, V., *Power MOSFET Basics*, International Rectifier.
21. *FQA13N80 Databook*, Fairchild Semiconductor.
22. Rivet, B., *Series Operation of Fast Rectifiers*, Application Note, ST SGS-THOMSON Microelectronics.
23. *TDA 16846/ TDA 16847, Controller for Switch Mode Power Supplies Supporting Low Power Standby and Power Factor Correction*, Data Sheet, Infineon Technologies AG, 2000.
24. *AN-ICE1QS01 A New Quasiresonant Controller for Switch Mode Power Supplies Supporting Low Power Standby and Power Factor Correction*, Application Note, Infineon Technologies AG, 2003.
25. Pirwitz, J., I. Zverev, *Considerations of Conducted EMI in SMPS, AN-EMI-01*. Application Note, Version 1.1, Infineon Technologies AG, 2003.
26. Zöllinger, H., *AN-SMPS-1683X-1 Coolset, TDA16831--34 for Off-Line Switch Mode Power Supplies*, Application Note, Infineon Technologies, 2000.
27. Ogata, K., *Modern Control Engineering*, Prentice Hall, New Jersey, 2002.
28. *Optimum Feedback Amplifier Design For Control Systems*, Technical Paper #3, Vencable Industries.
29. *TL431 Datasheet, Adjustable Precision Shunt Regulator*, Texas Instruments.
30. Sedra, A.S., Smith, K.C., *Microelectronic Circuits*, Oxford University Press, New York, 1998.
31. *High Q Notch Filter*, National Semiconductor Linear Brief, 1995.
32. Kitchin, C., L.A., Counts, *A Designer's Guide to Instrumentation Amplifiers*, Analog Devices.
33. Franco, S., *Design with Operational Amplifiers and Analog Integrated Circuits*, McGraw-Hill, New York, 2002.
34. *Mur4100E Power Diode Databook*, Motorola Semiconductors.
35. *1N4148 Databook*, Fairchild Semiconductor.
36. *KBU8K Bridge Rectifier Databook*, Fairchild Semiconductor Corporation, 2001.
37. *Switch Mode Power Supply Transformers, Databook*, ICE Components, Inc.

REFERENCES NOT CITED

- Adragna, C., *L6565 Quasi-Resonant Controller*, Application Note, ST Microelectronics, November 2002.
- Agrawal, J. P., *Power Electronic Systems, Theory and Design*, Prentice Hall, New Jersey, 2001.
- Basso, C., *AN1692, Spice Model Eases Feedback Loop Designs*, Application Note, Motorola Semiconductor.
- Basso, C., J. Turchi, *A Quasi-Resonant SPICE Model Eases Feedback Loop Designs*, Application Note, On Semiconductor.
- Chen, J., R. Erickson, D. Maksimovic, "Averaged Switch Modeling of Boundary Conduction Mode DC-to-DC Converters", *IECON'01: The 27th Annual Conference of the IEEE Industrial Electronics Society*.
- Chen, T.H., W. Lin, C.M. Liaw, "Dynamic Modeling and Controller Design of Flyback Converter", *IEEE Transactions on Aerospace and electronic Systems*, Vol. 35, No. 4, October 1999.
- Chung, H., S.Y.R. Hui, and W.H. Wang, "An Isolated ZVS/ZCS Flyback Converter Using Leakage Inductance of the Coupled Inductor", *IEEE Transactions on Industrial Electronics*, Vol. 45 No. 4, pp. 679-682, 1998.
- Control IC for Switched-Mode Power Supplies using MOS Transistors, TDA4605-2*, Siemens Semiconductor Group, 1994.
- Czarkowski, D., K. Kazimierczuk, "Linear Circuit Models of PWM Flyback and Buck/Boost Converters", *IEEE Transactions on Circuits and Systems – 1: Fundamental Theory and Applications*, Vol. 39, No. 8, August 1992.
- Dallago, E., G. Venchi, "Analytical and Experimental Approach to High-Frequency Transformer Simulation", *IEEE Transactions on Power Electronics*, Vol. 14, No.3, May 1999.
- Data Book Library 2003/2004, Technical Product Information, Epcos AG.
- Determining the Free Running Frequency for QR Systems*, Application Note, ON Semiconductor, 2002.
- Electromagnetic Spectrum*, University of Arkansas at Little Rock, Physics and Astronomy Department
URL: <http://lepus.physics.ualr.edu/~tahall/EXAM2/emspec.jpg>
- EMC Filters*, Data Book, Epcos AG, 2001.
- Erickson, R.W., *Fundamentals of Power Electronics*, Kluwer Academic Publishers, Massachusetts, 2000.
- Ferrite and Accessories*, Data Book, Epcos AG, 2001.
- Fisher, M.J., *Power Electronics*, Thomson Information / Publishing Group, Boston, 1991.
- Goldwasser, S., Notes on Troubleshooting and Repair of Small Switch Mode Power Supplies, Version 2.74, Sci. Electronics Repair FAQ.
URL: <http://www.repairfaq.org/>.

- Gus,E., J. Kayafas, H.N. Abrams, *Stopping Time the Photographs of Harold Edgerton*, 2000.
- How to Keep a Flyback Switch Mode Power Supply Stable with a Critical-Mode Controller*, Application Note, Motorola, Inc., 1999.
- IGBT Application Note for Camera Strobe*, Fairchild Semiconductor, July 2000.
- Illumination Fundamentals*, Lightning Research Center, Renselaer Polytechnic Institute, 2000.
- Irving, B.T., Y. Panov, M.M. Jovanovic, "Small Signal Model of Variable-Frequency Flyback Converter", (2003 IEEE).
- Irving, B.T. and M.M. Jovanovic, "Analysis and Design of Self-Oscillating Flyback Converter", IEEE 2002.
- Jacobsen, D., P. Katzman, *Benefits of Xenon Technology for Machine Vision Illumination*, Application Note, Perkin Elmer Optoelectronics, 2001.
- Kassakian, J., M.F. Schlecht, G.V. Verghese, *Principles of Power Electronics*, Addison-Wesley, 1991.
- L6565 Quasi-Resonant Controller*, Application Note, ST Microelectronics, 2002.
- Lempinen, J., T. Suntio, "Small-Signal Modeling for Design of Robust Variable-Frequency Flyback Battery Chargers for Portable Device Applications", IEEE 2001.
- Lempinen, J., T. Suntio, "Modeling and Analysis of a Self Oscillating Peak-Current Controlled Flyback Converter", *IECON'01: The 27th Annual Conference of the IEEE Industrial Electronics Society*, 2001.
- Lin, B.T., Y.S. Lee, "A Unified Approach to Modeling, Synthesizing, and Analyzing Quasi-Resonant Converters", *IEEE Transactions on Power Electronics*, Vol. 12, No. 6, November 1997.
- Middlebrook, R.D., "Small Signal Modeling of Pulse Width Modulated Switched-Mode Power Converters", *Proceedings of the IEEE*, Vol. 76, No. 4, April 1998.
- Nagel, A., R.W.D. Doncker, "Systematic Design of EMI-Filters for Power Converters", *Conference Record of the IEEE Industry Applications Conference*, 8-12 October 2000, Vol. 4, pp. 2523 -2525 , 2000.
- Panov, Y., M.M. Jovanovic, "Adaptive Off-Time Control for Variable Frequency, Soft Switched Flyback Converter at Light Loads", *IEEE Transactions on Power Electronics*, Vol. 17, No. 4, 2002.
- Preller, P., *SMPS with CoolMOS and TDA16846X*, Application Note, AN-SMPS-CoolMOS1, Infineon Technologies AG.
- Principles of Power Conversion*, Computer Products, Power Conversion Europe.
- Reis, F.S., J.C.M. de Lima, V.M. Canalli, J.A. Pomilio, J. Sebastian. And J. Uceda, "Matching Conducted EMI to International Standards", *IEEE 33rd Annual Power Electronics Specialists Conference, PESC '02*, 23-27 June 2002, Vol. 1, pp. 388 -393, 2002.
- Rutkowski, J., *Stroboscopes for Industry and Research*, Pergamon Press, 1966.
- Sercarz, A, J.A. Berke, G.S. Gerratt, B.R. Ming, Y. Natividad, "Videostroboscopy of Human Vocal Fold Paralysis", *Ann. Otol. Rhinol. Laryngology*, 101. pp. 567-577, 1992.

75W SMPS with TEA1507 Quasi Resonant Flyback Controller, Application Note, Philips Semiconductors, 2000.

Sokal, O.N. and R. Redl, "Control Algorithms and Circuit Designs for Optimal Flyback-Charging of an Energy-Storage Capacitor (e.g., for Flash Lamp or Defibrillator)", *IEEE Transactions on Power Electronics*, Vol. 12, No. 5, 1997.

Spiazzi, G., D. Tagliavia, S. Spampinato, "DC-DC Flyback Converters in the Critical Conduction Mode: a Re-examination", *IEEE Industry Applications Society Ann. Meet.*, (IAS), Rome, October 2000. pp. 2426-2431, 2000.

Spini, C., *25W Quasi-Resonant Flyback Converter for Set-Top Box Application using the L6565*, Application Note, ST Microelectronics, November 2001.

Strobe Lights and Design Guidelines, Useful Circuits, and Schematics
URL: <http://www.repairfaq.org/>.

Suntio, T., J. Lempinen, K. Hynynen, P. Silvetoinen, "Analysis and Small Signal Modeling of Self-Oscillating Converters with Applied Switching Delay", IEEE 2002.

Titze, I.R., *Principles of Voice Production*, Prentice Hall, 1994.

Weston D.A., *Electromagnetic Compatibility, Principles and Applications*, Marcel and Dekker, New York, 2001.

Xenon Flash Lamps, Technical Information, Hamamatsu Photonics K.K., Electron Tube Center, April 1999.

Yaakov, S.B. and D.A. Edry, "Average Models as Tools for Studying the Dynamics of Switch Mode DC-DC Converters".

Zhang, Y.F., L. Yang, and C.Q. Lee, "Spectral Analysis and EMI Comparison of PWM, Quasi-Resonant, and Resonant Converters", *Proceedings of the 37th Midwest Symposium on Circuits and Systems*, vol. 2, 3-5, pp. 1285 -1288, August 1993.