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SOLAR HOME SYSTEMS

Manual for the design and modification of Solar Home System components

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Preface

This manual was produced by ECN in the framework of the World Bank project: Mitigating Global Climate Change through the Development of a Quality Process Infrastructure for Renewable Energy (QuaP-PV). This manual forms the basis of a training course for engineers who will be involved in the design and modification of solar home system equipment. The current text is a modified and extended version of a manual used in a similar training exercise provided by ECN at Hefei University in China in March and April 1999, in the framework of the China Renewable Energy Project of the World Bank.

The authors would like to thank Enno Heijndermans of the World Bank for funding the work and for providing guidance in formulating the manual. We would like to thank Jeroen van Twisk, Jan Pierik and Jaap Eikelboom of ECN for their detailed comments which have improved the text.

CONTENTS

Abbreviations and special terms

ADC	Analogue Digital Conversion
Ballast	The high frequency inverter, used for driving a fluorescent lamp
CFL	Compact Fluorescent lamp
Crest factor	(Peak voltage) / (RMS voltage)
Duty Cycle	100% * (‘on’-time / ‘off’-time)
EEPROM	Electric Erasable Read Only Memory
Electrical efficiency	100% * (Output power) / (input power)
ESR	Electro Static Resistance
FL	Fluorescent Lamp
FLI	Fluorescent Lamp Inverter
HVD	High Voltage Disconnect, the instant in time when the module is disconnected from the system in order to stop battery charging
HVR	High Voltage Reconnect, the instant in time when the module is reconnected to the system
LVD	Low Voltage Disconnect, the instant in time when the load is disconnected from the system because of a low battery
LVR	Low Voltage Reconnect, the instant in time when the load is reconnected to the system. Energy use is enabled
MOV	Metal-Oxide Varistor
NOCT	Nominal operating cell temperature
opamp	Operational Amplifier
PCB	Printed Circuit board
PWM	Pulse-Width Modulation
SHS	Solar Home System
SOC	State Of Charge of the battery

1. INTRODUCTION

1.1 Background

Solar photovoltaics is one of the most cost effective means to provide small amounts of electricity in areas without a grid. Especially when people live in scattered homes, the costs of alternatives to provide electricity are usually prohibitively high. Solar home systems (SHS) are small systems designed to meet the electricity demand of a single household. A Solar home system always consists of one or more photovoltaic (PV) modules, a battery, and a load consisting of lights, and one or more sockets for radio, television or other appliances. A battery charge regulator is usually added to control charging and discharging of the battery.

1.2 Objectives

This manual has two major objectives:

- To provide the necessary background material for the modification of existing designs of Solar home system components, especially the battery charge regulator, inverters for fluorescent lamps and DC/AC inverters;
- To provide background material for completely new designs of Solar home system components.

The target audience for this manual is primarily engineers in companies producing solar home system components. Engineers in solar energy research institutes would also benefit from the information provided.

1.3 Scope of work

The solar PV module is by far the most reliable component of a solar home system. Relatively few problems are related to the PV-module. We therefore limit ourselves to the designs of the rest of the components, the so-called Balance of System (BOS) components. PV-systems will be discussed but only for systems with a power level appropriate to household demands. Batteries will also be discussed, but only those aspects which are important for the BOS-components.

Solar Home System project in Lebak, Indonesia

Because building practices depend very much on local customs and sometimes on regulations, this manual does not extensively discuss the support structure for the modules, but focuses on the electrical and electronic components. The emphasis is on the battery charge regulator because it is the single most critical component in a solar home system. Lights are the only appliances that are discussed in this manual, because lighting is included in most solar home systems, and the lights are usually provided with the system. Some of the larger solar home systems can also be equipped with a DC/AC inverter to allow the use of (larger) AC appliances.

In this manual we have limited ourselves to user experiences relevant for design, technical and non-technical aspects of solar home systems. Financial aspects will be mentioned in relation to design and system sizing. Institutional settings, ownership issues, financing schemes and after sales arrangements will not be discussed, although these are also very critical aspects in the success of dissemination of solar home systems.

Extensive use (a number of figures and many quotations) has been made of the following five publications:

- Anil Cabraal, Mac Cosgrove-Davies, and Loretta Schaeffer, “Best Practices for Photovoltaic Household Electrification Programs, Lessons form Experiences in Selected Countries”, The World Bank, Technical Paper No. 324, 1996.
- Nieuwenhout, P.J.N.M. van de Rijt, E.J. Wiggelinkhhuizen, “Rural Lighting Services, A Comparison of Lamps for Domestic Lighting in Developing Countries”, Netherlands Energy Research Foundation ECN, July 1998.
- Hasna J. Kahn, Rezwan Khan, and Salma Khatun, “Optimal Design of DC Luminaires for Solar PV Applications (Final Draft)”, prepared for The World Bank, November 20th, 1998.
- Gerken, and D. Welsch, “A Pulse-Width Modulated High Reliability Charge Controller for Small Photovoltaic Systems”, Sandia, Report SAND97-0329, February 1997.
- Guy Séguier and Francis Labrique, “Power Electronic Converters, DC-AC Conversion”, Springer Verlag, 1989.
- “Battery guide for small stand alone photovoltaic systems”, IEA PVPS Task III, by Bo Andersson, Catella Generics and Staffan Ulvonas.

2. EXPERIENCE WITH SOLAR HOME SYSTEMS

Lectures on day 1

2.1 Introduction

Early in 1999, about one million solar home systems were in use in the world, and this number is rapidly growing. This is a strong indication that this technology provides desired services to rural households in non-electrified areas. However, technical and non-technical problems often arise, which can hamper the further wide-scale application of solar home systems in rural electrification. From time to time, experiences in different countries are documented or summarised. Engineers who are involved in the design of solar home systems and their components can learn from these experiences. Those system designs that take into account the demands of the users will be the most successful in the market place.

This chapter describes some of the findings that are most relevant for design engineers of solar home systems. This includes a number of non-technical aspects that are nonetheless important to guide design activities.

A Solar Home system can be defined as: a small, autonomous PV-system that consists of one or more solar modules, a battery and several 12Vdc appliances.

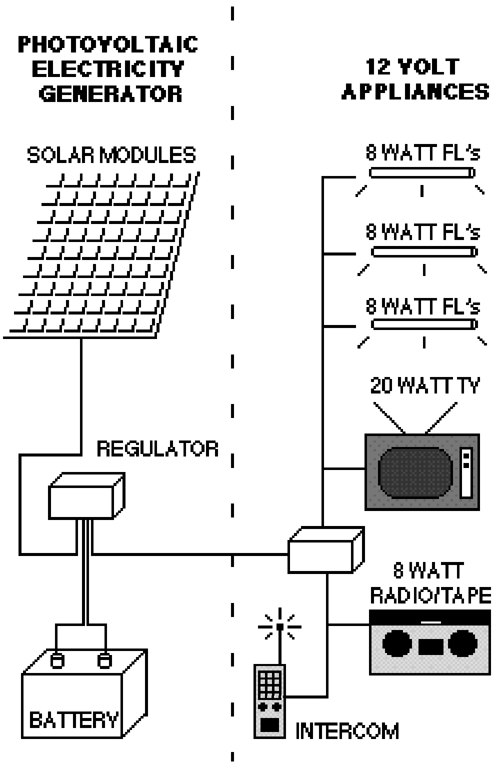

Figure 2.1 Basic set-up of a Solar Home System.

During daylight the battery is charged. The stored energy can be used for generating light and running a television or radio during the evening. When the battery is fully charged, the regulator disconnects the module in order to prevent the battery from becomming damaged. This moment is called High Voltage Disconnect (HVD). Below a certain level of discharge the battery can also be damaged. Therefore the regulator disconnects the load before the battery is completely drained. This moment is called Low Voltage Disconnect (LVD).

2.2 Non-technical aspects relevant for system sizing and design

2.2.1 Costs

In the investment cost of a solar home system, the PV module takes the major share of about 65% of the initial cost. Batteries take about 13% and the battery charge controller about 5%. Reducing the costs of the modules is clearly very important to make solar home systems more affordable to the rural poor.

Different components in a solar home system have different lifetimes. Modules can last for 20 years or more, but batteries usually last for about three years. In some countries this reduces to only two years. The total costs of the different components over a life cycle of 20 years, present a different picture than for the initial costs. The PV-module, which contributes 65% of the initial investment cost takes up only 33% of the life cycle costs[1], while the share of batteries increases from 13% to 46%, which is substantially larger than the life-cycle cost of the modules. Over a twenty year period the costs of the battery charge controller remains unchanged at 5% (see figure 2.2). To reduce the 20-year costs of solar home systems, it is therefore of utmost importance to increase the lifetime of the batteries used. Higher quality battery charge regulators are thus an essential requirement. In table 2.1 some typical figures for cost and lifetime of the different component of a solar home system are given. These figures have been used to calculate the values presented in figure 2.2.

Table 2.1 Typical costs and expected lifetime figures for a 40 Wp SHS.

Component	Cost [USD]	Lifetime [yr]
Module (40 Wp)	240	20
Battery	50	3
Controller	20	10
Light fixtures (2 pcs)	20	10
Fluorescent tubes (2 pcs)	2	1
Mounting materials, cables	40	20
Total	372	

[1] Calculated is the undiscounted life-cycle cost over a period of 20 years.

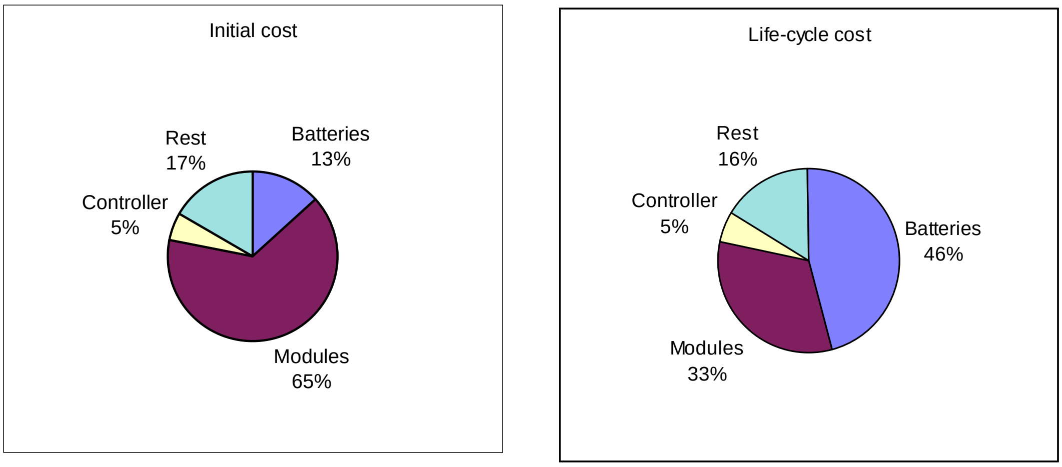

Figure 2.2 Breakdown of investment cost of solar home system components. The initial investments are shown on the left and the shares in the life-cycle cost on the right.

2.2.2 Ownership

There is a wide range of institutional and financial arrangements in solar home system projects. Consequently, ownership varies from complete ownership of the users in commercial distribution systems to state ownership in some government projects. In between are the credit schemes where the users will eventually own the systems after the loan has been completely repaid.

Ownership transpires to be an important factor in the attitude of people towards their PV-systems. A sense of ownership is a strong incentive for maintaining the systems. For this reason, projects where the government provides the systems free, usually show much higher failure rates than commercial schemes.

If known in advance, design engineers can take into account ownership of solar home systems. Especially in the case of publicly owned systems (either government or utility) it is necessary to ensure only a minimum amount of maintenance, and to expect that sometimes there will be no maintenance at all.

Often, aspects of ownership, quality of the products and after sales service are related to the financial success of the dissemination activity. Lower quality products and the absence of after-sales services generally leads to high default rates in financing schemes.

2.2.3 After sales services

Users of solar home systems must be able to get their equipment repaired in local workshops. The costs of replacement parts and labour costs can be covered by a guarantee arrangement, a fee-for-service arrangement or the user has to bear all the costs. In all cases, cost savings can be achieved when high quality products are used. Maintenance and repair cost are substantially higher in the case of low quality products. When analysed over the lifetime of a solar home system it is usually worth investing in improved quality. This is especially the case for the battery charge regulator.

2.2.4 Feed-back from field experience

In most countries, distribution of solar home systems starts with PV-modules in the range of 35 to 50 Wattpeak. After some time growth stagnates because the majority of people in rural areas cannot afford the full costs. But more people appear to be able to afford the costs of smaller systems of 10 to 20 Watt. This illustrates the need to listen carefully to the requirements of the users.

There are two main channels of information. Through the workshops involved in the after sales service one learns how often certain components fail. Ideally, all problems and complaints should be registered. Another channel is the distribution network, which provides information regarding the demand for certain (sizes) of systems. The larger the range of products, the more market information can be obtained in this way.

Very little information is available about the actual use of solar home systems in practice. There are some scattered monitoring activities. Where available, they provide very useful data about the sizing of the systems and relative sizing of the components, for example.

2.2.5 Environmental issues

Batteries constitute a major concern in the environmental effects of large-scale implementation of solar home systems. Batteries can be recycled, but the recycling process can be environmental harmful. Designers of charge regulators can contribute to the extension of the lifetime of a battery.

2.3 Technical aspects

2.3.1 System losses

Before the system sizing can begin, an estimate is needed for the system losses. When the amount of energy that the user needs is known, the size of the module can be calculated. Taking into account all these factors, the battery size can be chosen.

The first step is to define the different factors that contribute to the systems' energy-loss. All the available energy starts at the module, so we start with the loss-factors there.

PV-module output losses:

- *Orientation is not optimum*. Mostly the module is mounted in a fixed position. For every location on earth there is one direction and tilt angle that results in the highest annual electricity generated, or for the highest amount generated during the darkest month, whichever of the two is required. However, this is not critical. When the direction is within about 20 degrees of the optimum direction and the tilt angle within 10 degrees of the optimum angle, the electricity generated is less than 5% of the optimum.

- *Shading of the module.* During part of the day the module is often shaded by a tree or a building. Compared to a module in an open site, this means energy-loss. Furthermore, trees grow. So after a couple of years a tree could start shading a part of the module.
- *Dust on the module.* Modules need to be as clean as possible. Dust builds up on the surface of the module especially in the dry season. Therefore, never install a module with an inclination angle of less than about 15 degrees, to allow the rain to clean the panel. This dust causes energy losses which can be as high as 5-10% even in areas with frequent periods of rain.
- *Temperature effect on the module.* As described in section 2.4.1, the temperature effect on the module cannot be neglected. The higher the temperature, the lower the power output of the module. Modules are tested at a standard temperature of 25 ^{o}C. When lit by sunlight in tropical areas, the temperature can easily reach 70 ^{o}C. The power at the maximum power point of crystalline silicon cells decreases by about 0.4 to 0.5 % per ^{o}C of temperature increase. Taking a typical figure for the temperature of 60 ^{o}C, results in a reduction of power output by about 16%. Amorphous sicilicon modules have a lower temperature coefficient of about 0.2 to 0.25 % per ^{o}C of temperature increase. For the same temperature this results in only half the output reduction: 8% at 60 ^{o}C.
- *Nameplate mismatch.* Some manufacturers state an output power on the nameplate, which can be 10% higher than the actual output power. This has to be taken into account.

Other losses:

- *Cable losses.* When electrical energy is being transported via cables, energy loss is unavoidable. Section 2.4.7. explains more about wire-sizing. By selecting a sufficiently large wire size, the losses can be reduced to less than 5%.
- *Semiconductor energy loss.* (Components are described in section 4.3.2). Both the MOSFETs (metal-oxide semiconductor field-effect transistor) as the blocking diodes convert a certain amount of energy into heat. These components are always included within a charge regulator. On a daily base they can use about 10Wh. (Module mosfet during the day, load-mosfet during the night).
- *Charge regulator energy consumption.* The charge regulator continuously draws a small current of about 5 to 25 mA. With a quiescent current of 5mA (1.44Wh a day) in a 150Wh system losses will be 1%.
- *Chemical/electrical energy conversion losses inside the battery.* Conversion inside the battery takes energy. This energy loss also depends on the age of the battery. The electrical efficiency of a new battery can be 90%. During its lifetime it could reduce to 75%. Due to corrosion and increase in internal resistance in the battery, the capacity will be reduced to nearly zero, while the electrical efficiency will stay at 75% (for example).
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An estimate can be made from the total system losses.

Module output:

	Range	Typ.value
Orientation is not optimum	*5-10%*	5%
Shading of the module	*0-xx%*	*0%*
(assuming that the system is installed properly)		
Dust on the module	*5-10%*	*5%*
Temperature effect on the module	*0-20%*	*16% (at 60^{o}C)*
Nameplate mismatch	*5-15%*	*10%*

Other losses:

Cable losses	*5-10%*	*5%*
Semiconductor energy loss	*5-8%*	*7%*

Charge regulator energy consumption	*1-5%*	*2%*
Chemical/electrical energy conversion losses inside the battery	*10-20%*	*15%*
Total system efficiency		*50%*
Energy loss		*50%*

2.3.2 Sizing of the PV-module

The optimum size of a solar home system is directly related to its costs, household electricity requirements and their willingness to pay. Generally, people want more electricity, but there is always a trade off between what people want and what they are actually willing to pay. Unrealistic expectations should be avoided. A 10 Watt-peak module, which is expected to run a refrigerator through a 150 VA inverter, is certainly going to disappoint the owner. It is the responsibility of design engineers to make realistic calculations of the number of hours that the lights and other appliances can be operated with a certain module size.

What module Wattage is required for a solar home system? This question can be answered after taking the following three steps:

a) Determine the average daily electricity demand of the household;
b) Calculate the system losses (see previous paragraph);
c) Calculate the module Wattage.

Household surveys can provide information about the current demand for energy services of the households that intend to switch to solar home systems. However, an extra demand can be created through access to electricity, for example, through seeing advertising on television. Therefore, the determination of the average daily electricity demand of the household is often based on a number of assumptions. For example: two 5 Watt PL lights operated for 3 hours per day require in total 30 Wh per day. For two hours of watching television (with a 20 W Black and White television set) the total energy requirement amounts to 70 Wh per day.

Capacities of solar PV modules are given in Watt-peak. This allows you to calculate the electricity generated under different levels of sunshine. On a clear day, when the sun is high in the sky, the power of the sunlight on a flat plane facing the sun is approximately 1000 Watt for each square metre. To standardise the capacity of solar PV modules, the capacities are always given at an illumination of exactly 1000 Watt per square metre. 1 Watt-peak of PV-cells generate 1Watt of electric power under the standard test conditions of 1000 Watt per square metre and a temperature of 25 ^{o}C.

How much sunlight is available? Meteorological tables show the solar insolation, usually in kWh/m^2/day. This differs from day to day and shows a seasonal variation over the year. It is safe to design the system based on the average daily insolation in the month with the lowest insolation. In most areas near the equator this will be in the order of 4 kWh/m^2/day.

One Watt-peak of PV-module capacity is defined as 1 Wh of electricity generated per hour with 1 kWh/m2 of insolation. Therefore, with an insolation of 4 kWh/m^2/day, 4 Wh of electricity will be generated. However, this describes the ideal situation, when there are no losses. As shown in the example of the previous paragraph, energy losses can amount to 50 %, which implies a system efficiency of 50%. In this example, every Watt-peak of module capacity generates 2 Wh of electricity per day during the darkest month of the year. The amount of Wh electricity available to the user, per Watt-peak of PV-module capacity, is calculated by multiplying the average daily insolation by the system efficiency. And the other way around, if one knows the required daily amount of electricity, and wants to calculate the amount of Watt-peak required, one divides the electricity demand by the product of the average daily insolation and the system efficiency. For example, with a daily requirement of 70 Wh per day and insolation of 4 kWh/m^2/day and a system efficiency of 50%, the required module capacity amounts to 70/(4*0.5) = 35 Wp[2].

2.3.3 Sizing of components

As soon as the module capacity is known, the most important remaining free parametre is the capacity of the battery. This depends on the required autonomy, that is the number of days storage capacity. Using the numerical values of the previous paragraph, the household requires 70 Wh per day. Since this is delivered at 12 V, this is equivalent to 70/12 = 5.83 Ah per day. With an autonomy of 3 days, the battery needs to be able to provide 3 times 5.83 Ah = 17.5 Ah. However, due to the low voltage disconnect, one does not use the complete battery capacity, but only the top 40% (the actual value depends critically on the setting of the low voltage disconnect). With a maximum depth of discharge of 40%, the required capacity of the battery is 17.5 Ah/0.4 = 43.7 Ah. A practical value for the capacity close to this value, for example 45 Ah, should be chosen.

It would seem, at first sight, that a large battery capacity and a long autonomy period would be a good choice to be sure that there is sufficient energy stored to overcome periods with less sunshine. However, if the system is operated in such a way that low voltage disconnect happens almost daily, an autonomy which is much larger than one day is useless, because the battery will be seldom in a high state of charge (SOC). In this case, the larger the battery capacity, the lower the average SOC.

In the initial stages of PV-development in Sri Lanka, the battery charge regulator was left out, and smaller modules were used with 32 cells instead of the more usual 36 cells. With the lower number of cells, the maximum voltage attained when the module is connected to the battery is not high enough to cause gassing in the battery. Therefore, there was no need for a high voltage disconnect, one of the key functions of a charge regulator. However, when there is no charge regulator there is no low voltage disconnect to protect the battery against too low a state of charge. In practice, the benefits of using 36 cell modules outweighed the additional cost of a regulator.

2.3.4 Modularity

Small PV systems have a larger market potential than larger systems due to their lower costs. For example in Swaziland, users who want to extend their system, usually buy a new completely independent system. In Kenya people often start by buying a television set, followed by a battery, and only later a small module of 10 Wattpeak. Later one or more additional

[2] 70 W_eh/day divided by 0.5*4 kW_sh/m^2/day = 35 W_eh / kW_sh/m2 = 35 Wp, where the subscripts ‘e’ and ‘s’ are used to mark the difference between electric power and solar power

modules are bought. In the design of the systems, especially the smaller solar home systems it is useful to allow for modular expansion of the module capacity and the load.

2.4 Solar Home System components[3]

2.4.1 Modules

PV-promotion in Botswana

Solar PV modules are the most reliable component of a solar home system. Standards have been formulated (IEC 1215), and modules can be certified. For the certification, tests have to be passed regarding: visual inspection, performance at Standard Test Conditions (STC), measurement of temperature coefficient, measurement of nominal operating cell temperature (NOCT), performance at low radiance, outdoor exposure, thermal cycling, humidity freeze, damp heat, and robustness of termination.

In the design it should be noted that manufacturers have been known to supply modules with peak wattage about 10% lower than the nameplate capacity.

In addition, the temperature effect on modules can be critical in some areas. In full sun, the module temperature can increase to 70^{o}C. Normally a quality module has a temperature coefficient of about –2.5mV/ ^{o}C/cell. At 70^{o}C a 36-cell module should be able to charge the battery sufficiently. Because a protection diode is connected in series with the module in most systems, the voltage drop across this diode should also be taken into account. (Figure 4.1, Schottkey). Figure 2.3, the I/V-curve of a single cell at different temperatures shows how the output voltage reduces with the temperature. This relationship is almost linear.

[3] This section 2.4 draws heavily (with many citations) on “Best Practices for Photovoltaic Household Electrification Programs, Lessons from Experiences in Selected Countries”, Anil Cabraal et al., ASTAE, The World Bank.

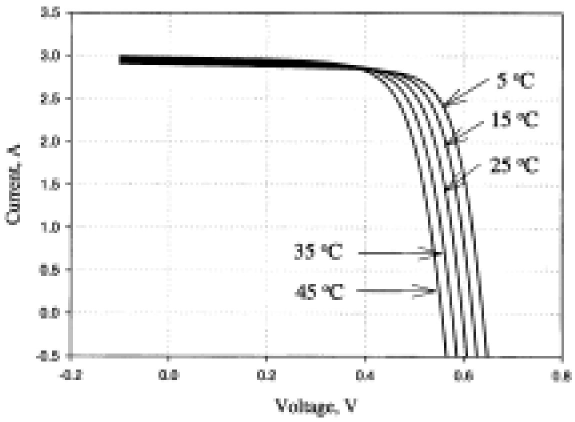

Figure 2.3 Temperature effect on a solar cell

The temperature coefficient increases with a decrease in module quality. At 25°C (STC) a 36-cell module based on cells from figure 2.3 would have an open circuit voltage of 21.96V. At 70°C this will be 17.91V. [(70-25) * 2.5mV/ °C/cell * 36 cells = 4.05V lower]

In India, the temperature coefficient of some (locally manufactured) modules was so high, that it was not possible in some situations to charge the battery up to HVD point. When the open circuit voltage Voc of a single cell drops to 0.4V at 70°C, the total module will give only 14.4V. When the blocking diode is also considered, the actual maximum charging voltage is 14.0V.Because the battery is never fully charged, this contributes to a decrease in the lifetime of the battery.

To illustrate the temperature effect on the module, figure 2.4 shows the temperature coefficients of a 36-cell module in %. In the upper left picture the temperature coefficient of the open circuit voltage is –3.26% for every 10 °C.

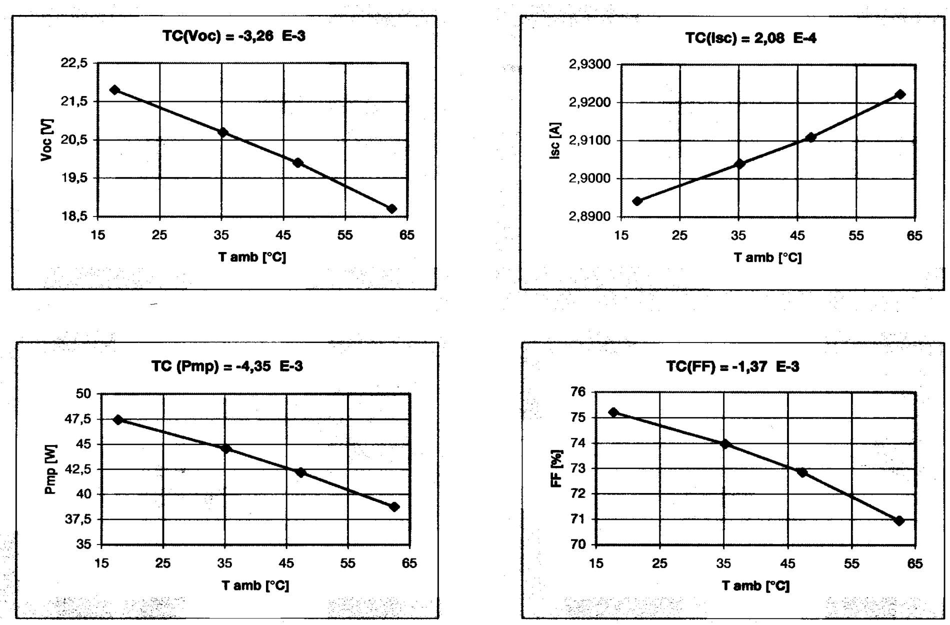

Figure 2.4 Temperature Coefficients of a Multicrystalline PV module, measured at ECN in 1999

As this diagram shows, the voltage reduces with temperature increase, while the current increases slightly. When the product of voltage- and current is examined, the voltage-coefficient wins. Hence the module power in the maximum power-point decreases by -4.35% / 10^{o}C.

2.4.2 Module Support Structure

The support structure for the PV-module(s) should be corrosion resistant (galvanised or stainless steel or aluminium) and electrolytically compatible with materials used in the module frame, fasteners, nuts and bolts. The design of the support structure should allow for proper orientation of the module, tilt and expansion of the system's capacity. Roof mounting may be preferable to ground or pole mounting since it is less costly, and requires less wiring. The module support should be firmly attached to the roof beams and not loosely attached to the roof tiles. The module should not be placed directly on the roof but 10-50 cm above the surface itself, to allow cooler and therefore more efficient operating conditions. If the module is mounted on a pole, the pole should be set firmly in the ground and secured with guy wires to increase rigidity. Pole mounted modules should be accessible for cleaning but high enough above the ground to discourage tampering.

2.4.3 Battery Charge Regulator

The charge and load controller prevents system overload or overcharging. For safe and reliable operation, the controller design should include:
- A low-voltage disconnect (LVD);
- A high voltage disconnect (HVD), which should be temperature-compensated if wide variations in battery temperature are expected. Temperature compensation is especially important if sealed lead-acid batteries are used;
- System safeguards to protect against reverse polarity connections and lightning-induced surges or over-voltage transients;
- A case or cover that shuts out insects, moisture and extremes of temperature.

To enhance the maintainability and usability of the solar system, the controller should:
- Indicate the battery charge level with a simple LED display or inexpensive analogue metre. Three indicators are recommended: green for a fully charged battery, yellow for a low charge level (pending disconnect), and red for a 'dead' or discharged battery;
- Be capable of supporting added modules to increase the system's capacity;
- Be capable of supporting more and bigger terminal strips so that additional circuits and larger wire sizes can be added as needed (this is necessary to ensure that new appliances are properly installed);
- Have a fail-safe mechanism to shut down the system in the case of an emergency and to allow the user to restart the unit.

Additional design considerations are:
- Low quiescent current (own consumption);
- A sturdy design to withstand the shocks and vibrations of transport;
- A sufficiently high lifetime, preferably longer than 5 years;
- Simple visual information on the casing should make the manual (almost) obsolete.

The charge controller could be equipped with a boost charging function to increasing the lifetime of the battery. Once every month or so, the battery is temporarily allowed to pass the high voltage disconnect setting. The resulting gassing will lead to cleaning of the battery plates and reduces stratification of the battery electrolyte.

Another optional feature in the design of an advanced battery charge regulator is pulse-width modulation (PWM). To charge the battery fully, a constant voltage algorithm is applied when the battery is almost full. This can be achieved with pulse width modulation.

New controller designs always have to be tested in the laboratory and in the field before being used in large-scale dissemination projects.

2.4.4 Battery

The most commonly used battery in solar home systems is a lead-acid battery of the type used in automobiles, sized to operate for around three days. Automotive batteries are often used because they are relatively inexpensive and available locally. Ideally, solar home systems should use deep-cycle lead-acid batteries that have thicker plates and more electrolyte reserves than automotive batteries and allow for deep discharge without seriously reducing the life of or damaging the battery. In a well-designed solar home system, such batteries can last for more than five years. However, deep-cycle batteries are not usually made locally in developing countries and high duties often increase the price of importing such batteries. Solar batteries can have a longer time between production and use than locally produced car batteries. Due to the longer storage time the quality advantage over car batteries is lost. Other disadvantages are the substantially higher costs and the difficulties in finding replacements. A very large number of solar home systems need to be distributed to make production of special PV batteries viable.

2.4.4.1 The lead-acid battery

Positive electrode design

Positive electrodes are designed to suit different applications. *Pasted grid plates* are primarily used for automotive applications like the SLI (starting, lighting, ignition) battery used in cars. The power capability of this battery type is very high and the deep discharge capability poor. Batteries with positive pasted plates are used in many applications, except for SLI applications, because of their low price.

In larger Solar Home Systems (100Wp or more) a truck battery can be used. They have thicker plates than a car battery, almost of the same thickness as special solar batteries. This will extend the battery life in a PV system significantly compared to a car battery.

Tubular plates are often used for traction batteries, i.e. for electric industrial or road vehicles. The main features are a fairly high specific energy per volume and a good capability for deep discharge. The charging time is average, 5 - 10 hours. In tubular plates, where a lead spine is surrounded by a highly porous, plastic tube, the active mass is located between the lead spine and the tube. The high current capability of this type of electrode is nevertheless limited, because the dimension of the tubes can not be reduced. The normal tube diameter is 8 mm (discharge time 3 - 10 h), which can be reduced to 6 mm for specific higher power applications (discharge time 1 - 3 h).

Rod plates are used in batteries for lighter traction as well as for some stationary applications. Rod plates consist of vertically arranged rods. The active mass surrounds the rods, and is completely enveloped in a pocket. The lead rods correspond largely to the spines of the tubular plate. This construction leads to a high utilisation of the active material and also to a high current capability.

2.4.4.2 What type of battery should be chosen?

For a typical small PV system the initial investment cost has to be kept low and the car batteries, truck batteries, solar batteries can be recommended in this order. In practice of course the local availability of batteries will also be a decisive factor. Therefore car or truck batteries are the best option in some developing countries where no other batteries are available.

2.4.4.3 Capacity will increase slightly in the beginning of operation

A new battery will not reach its full capacity during the first discharge cycle. In the standards, it is described that up to 10 charge/discharge conditioning cycles are allowed before performing the first capacity verification test. Normally, a battery should be fully charged and discharged a few times to activate the electrode materials. This is often difficult to achieve when PV panels are used for charging, as their charging capability is limited. This might delay the activation. The full capacity might not be reached during the first 50 cycles. This effect should not be overestimated. It will be in the range of 10-20% of the nominal capacity.

2.4.4.4 Temperature effect on capacity

The nominal capacity is normally measured at 20 ^{o}C battery temperature and down to a certain fixed cut off voltage of the battery. In cold climates the usable capacity may be significantly reduced, as low temperatures will slow down the chemical reactions in the battery. This will result in a useable capacity at for example minus 10 ^{o}C battery temperature of only 60% of the nominal value at 20 ^{o}C. The capacity is still there if the battery is heated to 20 ^{o}C but at low temperature one can not utilise the full amount. When possible the battery should therefore be placed indoors or otherwise sheltered from low temperatures by insulation or perhaps even placed in the ground if any other heat sources are not available. Seasonal storage containers with phase change materials with water as the main storage component have been shown to work well.

The opposite effect on capacity in warm climates is not of the same order of magnitude. In this case the battery should be placed in a way to avoid high temperatures. Already 10 ^{o}C temperature increase above 20 ^{o}C will double the corrosion velocity of the electrodes. In India some street lanterns had a black coloured battery-box, while the whole system was placed in full sunlight. Of course this is not a good solution.

2.4.4.5 Charging a battery

A lead-acid battery can generally be charged at any rate that does not produce excessive gassing, overcharging or high temperatures. In the laboratory, constant current charging is often used. Constant voltage charging is preferred in many stationary installations. This is especially true for sealed or valve regulated (VR) batteries. However, for deep cycling applications, constant voltage charging is not recommended due to the fact that the charging time is much longer than acceptable, several days or weeks. If the battery is charged at too high a voltage, which will shorten charging time, the corrosion is enhanced and the battery lifetime will suffer.

In a PV system the energy source is not regular and special charging considerations have to be made.

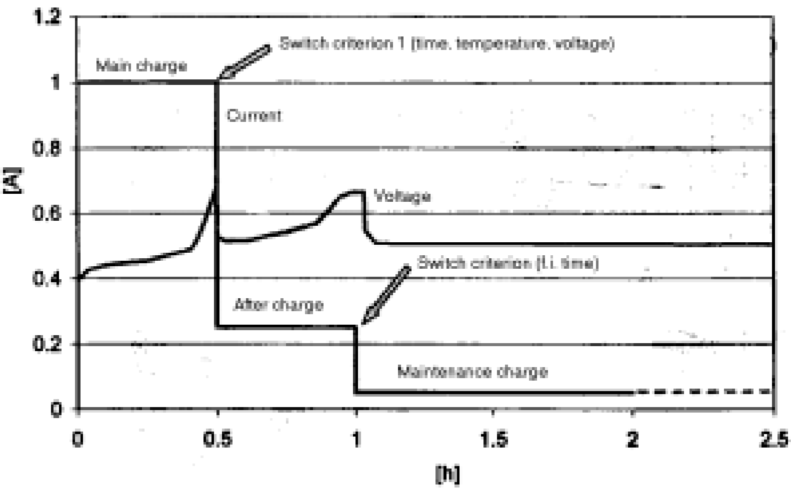

Figure 2.5 Different charging states while charging a battery

Different steps in the battery charging procedure:

The steps used to charge an open or vented lead acid battery are named:

main charge, used for charging the battery up to a level when gassing starts and the voltage rises. (Voltage limit 2.39 V at 25 ^{o}C; 2.33 V at 40 ^{o}C).
top-up charge, to reach the 100 % state of charge from a level of 90 - 95 %. (Retain the voltage limit by decreasing the current).
equalisation charge, used for equalising the capacity of the individual cells in a multi-cell battery. This is an important issue for improving life, but requires a special controller mode to create this in a system charged by PV panels. (Increase the voltage to 2.5 - 2.6 V/cell for a short time, 0.5 - 1 h, at regular intervals, once a week).
maintenance charge, used for maintaining the full capacity in a battery that is already fully charged but not frequently used for some period. (Approx. 2.20 - 2.25 V/cell or a current value equal to the capacity value divided by 100 (C/ 100)).
The battery is not very sensitive to the abuse created by the main charge, except for the temperature rise. It is preferable not to charge a very warm battery (>50 ^{o}C) if there is a possibility to cool the battery first.

When controlling the charging process according to the voltage of the whole battery, it is understood that the individual cells of the battery have the same voltage. If not, some cells may not be fully charged (undercharged). It is therefore important to check the voltage of each unit regularly. Undercharging may result in sulphatation of some cells of a battery. In this case an equalising charge may restore the capacity of the undercharged cells. On the other hand, one shorted battery cell will result in extensive overcharge of the other cells, which shortens the life of the whole battery or even creates a potential safety hazard.

Further explanation of charging principles can be found in section 4.3.4.

2.4.5 Battery Mounting

Batteries are sometimes left exposed on the ground and accessible to children. The potential dangers (burns from battery acids, shorts, and explosions) highlight the need for a well-designed battery enclosure to maximise safety and minimise maintenance. Such enclosures are being introduced in Indonesia and the Pacific Islands. Made of injection- molded plastic or Fiberglas, the enclosure contains the battery, battery charge regulator, charge indicators and switches. The electronic elements are isolated from the battery, and the battery enclosure has vents to disperse gases and can have channels to divert any acid overflow. There is no exposed wiring and the battery can be checked and filled easily.

2.4.6 Lamps, ballasts and fixtures

The principle reason most householders acquire a solar home system is that it provides brighter, safer, cleaner and more convenient lighting than kerosene lamps. Field observations show that additional lights increase satisfaction with and acceptance of solar home systems. In many cases, householders have subsequently added additional lights to their system while maintaining the overall level of energy consumption.

Efficient fluorescent lights (CFL or tube lights) are usually preferable to incandescent lights because of their much higher efficiency. However, fluorescent lamps require well-designed ballasts to ensure that the tubes will operate for more than 5000 hours and will not interfere with radio or television reception. In some programmes, the fluorescent ballasts have been the most problematic component of the PV-system.

Figure 2.6 Assembly of a CFL lamp in Botswana

Low-watt (1-2 W) incandescent lights may be preferable in cases where low-level area lighting is required (orientation lighting). Fixtures with reflectors are recommended to increase the effectiveness of the lights. Fixtures that use diffusers must be sealed against insects, since the "useful light" output and efficiency of the fixtures with diffusers can be drastically reduced by a build-up of dirt and insects inside.

2.4.7 Wiring, Switches and Outlets

Solar home systems should have high quality switches and outlets (preferably rated for DC operation). In Sri Lanka, standard AC surface-mounted wall switches were used along with two-prong AC wall outlets. If the lights and appliances draw little current, these AC switches and outlets are satisfactory substitutes for DC-rated components.

Undersized wiring is sometimes used in solar home systems, particularly when additional light fixtures are installed. This practice leads to energy losses and unacceptable voltage drops and should be strongly discouraged. All wiring should be stranded copper wire, preferably sized to keep voltage drops to less than 5 percent between battery and load.

Wire sizing tables like Table 2.2 can also be found on the Internet. 5% energy loss should be acceptable. When much lower losses are demanded, the cost of the wiring becomes too high for a Solar Home System. Table 2.2 shows such a calculation method. At a current of 4A the 12Vdc system handles 48Watt. Estimated cable length (Module--------battery--------lights/radio) will be 20 metres, 60 feet. Going to the right from 4A, the first distance higher than 60 feet mentioned is 84 feet. This corresponds to a gauge of wire of #8 that should be used.

Table 2.2 Cable gauge in 12Vdc systems

Amp in wire	Watt at 12Vdc	#14	#12	#10	#8	#6	#4	#2	1/0	2/0	3/0
1	12	84	131	206	337	532					
2	24	42	66	103	168	266	432	675			
4	48	18	33	52	84	133	216	337	543	685	
6	72	14	22	33	56	89	114	225	360	450	570
8	96	10	16	27	42	66	108	168	272	338	427
10	120	8.5	13	22	33	53	84	135	218	270	342
15	180	6	8.5	13	22	35	56	90	144	180	228
20	240		6.6	10	16	27	42	67	108	135	171
25	300			8.5	13	22	33	54	86	108	137
30	360			6.6	11	18	28	45	72	90	114
40	480				8	13	21	33	54	67	85

Twisted wire or spring-clip wire connections do not provide a good electrical connection and should not be used. Soldered or crimped connections or screwed terminal blocks are the best choices. Soldered wire connections require a non-corrosive, rather than an acid-flux solder, especially in regions where corrosion can be a serious problem. The wiring should be neatly and securely attached to the walls, either on the surface, in conduits, or buried in the walls.

Solar home systems should have a distribution panel that allows users to connect additional loads simply and safely, utilising the circuit protection and LVD features of the battery charge regulator. A distribution panel must always be used if additional circuits are to be installed. Direct battery connections must not be made.

2.4.8 Inverters

In larger solar home systems, DC/AC inverters can be used, where customers require AC electricity. It is not very common that users of systems smaller than 100 Wp require inverters. The cost of an inverter and associated energy losses recommend against its use in small systems. Several types of inverters are available: square wave (the least expensive and least efficient), modified square wave, and pure sine wave (the most expensive and most efficient).

3. LIGHTING IN SOLAR HOME SYSTEMS

This chapter provides an introduction to the application of lighting in solar home systems. It consists of two sections. In the first (3.1) the different types of lighting are discussed. With the help of a sample of actual measurements by ECN, some relevant characteristics of lighting are illustrated. In section 3.2 the operation of ballasts for fluorescent lights are analysed. This includes a number of possible design modifications.

3.1 Design considerations for lights[4]

Lighting is a basic necessity with a very high priority in households all over the world. Many different means are available to provide these lighting needs. When electricity first arrives in a village, lighting is by far the most common application. Electric lamps have some major advantages over their non-electric counterparts. These advantages are: increased lighting levels, higher quality of the lighting, greater ease of use, lower cost per unit of light output, no heat production and very low fire hazard.

In conventional rural electrification with grid extension, few incentives exist to apply energy efficient lighting. Consumers lack the knowledge of alternatives and the usually low electricity tariffs do not provide a strong incentive to consider an optimum choice of lighting source.

However, when solar photovoltaics provide the electricity for rural lighting, the situation is completely different. The high electricity cost per unit stimulates careful consideration in choosing the right type of lamp.

3.1.1 Types of lighting

There are four categories of lighting relevant for solar PV systems for households. These are:
- general lighting (illumination of a whole room);
- localised lighting (illumination of part of a room only);
- local or task lighting (e.g. illumination of a table);
- orientation lighting (only sufficient to recognise shapes

In relatively small rooms in which solar PV-systems are applied in developing countries, the distinction between localised and local lighting becomes insignificant. Both the luminaire and the lamp determine the possible use of the lighting.

General lighting

Lights that are intended to be attached high up in a room are categorised as general lighting. Most general lighting luminaires use fluorescent lamps and range in luminous flux[5] from about

[4] This paragraph 3.1 is mainly derived from a recent ECN report: “Rural Lighting Services, A Comparison of Lamps for Domestic Lighting in Developing Countries”
[5] Luminous flux is the technical term for the total light output of a lamp in all directions as perceived by the human eye. The measurement unit is the lumen [lm], which takes into account that the eye is more sensitive to some colours than to others.

150 to 500 lumen. For the user it is relevant to obtain the highest lumen output per Watt of energy consumption [lumen per Watt].

Figure 3.1: DC general / localised light from Goldstar

Localised and local lighting

In localised and local lighting, only a part of the room is lit. This can be accomplished with the smaller fluorescent tube types, or even better with compact fluorescent tubes. A reflector is usually included to concentrate the light to the task area, and to prevent glare.

Orientation lighting

Designers of solar home systems often overlook the fact that many people desire small orientation lights during the whole night. Very low lighting levels are sufficient to recognise the shapes of furniture. For users of PV systems, the power consumption of an orientation light is the most relevant factor. Small, 1-Watt incandescent lights are very suitable as a cost effective orientation light. Lately, the development of high efficiency LED lamps has made significant progress. For both localised and orientation lighting this can be a very suitable alternative for the future, because of its high efficiency and long lifetime.

Figure 3.2: Cluster LED lamp from Steca

3.1.2 Luminous flux

The Total amount of light emitted by a source, the so-called luminous flux, is one of the most important characteristics of a lamp. Luminous flux of lights in a test sample of 35 different lights measured by ECN ranged from about 1 to 1000 lumen for the electric lights and from about 10 to 2000 lumen for the non-electric lamps[6]. Figure 3.3 presents an overview of the luminous flux measurements of all the lights included in the test sample. As expected, the higher the power of the lamp, the higher the light output.

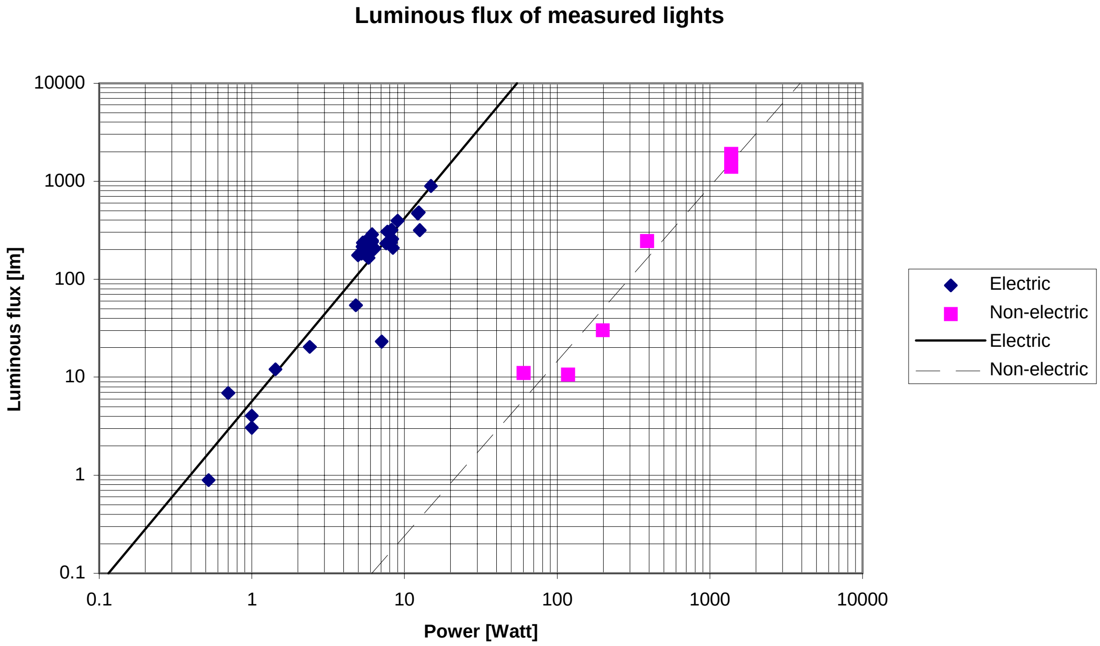

Figure 3.3 Luminous flux [lumen] and energy consumption [Watt] for 35 lights measured by ECN.

Electric lights are far more efficient than non-electric lights. For the household lights in the test sample it can be concluded that at the same level of luminous flux, the energy consumption of a non-electric light is about 65 times higher than for an electric light. There was no overlap in power between electric and non-electric lights in this sample: the highest power electric light is a 15 Watt compact fluorescent lamp, while the lowest power non-electric light is small candle with an energy consumption of 60 Watt[7]. If there had been an overlap in the power range, an electric lamp would have a lumen output of about 2000 times higher than the non-electric lamps at the same Wattage level.

Lights with a lumen output less than about 20 lumen are not meant for general lighting, but usually function as orientation lights only. Most users prefer to have one or more orientation lights in their homes.

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[6] "Rural Lighting Services, A Comparison of Lamps for Domestic Lighting in Developing Countries"

[7] This is a comparison on the basis of energy input into the light source. When comparing primary energy inputs, the equivalent energy consumption of the 15 W PLET lamp will be in the order of about 50 Watt, the exact value depending on the fuel mix.

3.1.3 Luminous efficacy

Conversion of fuels or electricity in visible light takes place in a number of different processes and under different conditions, resulting in a wide range of luminous efficacy[8]. Figure 3.4 shows the luminous efficacy in lumen per Watt of input power for all lights in the test sample.

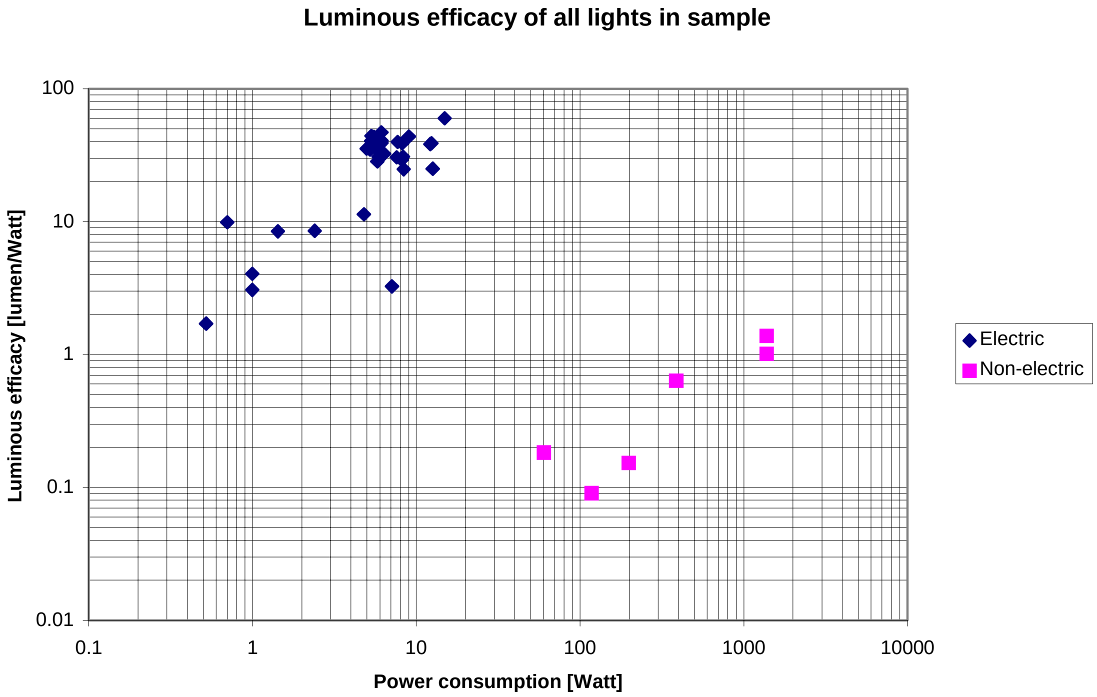


Figure 3.4 Luminous efficacy [Lumen/Watt] of all lights included in the test sample.

As could already be concluded from figure 3.3, electric lights are much more efficient than non-electric lights. Another interesting conclusion is that for both types of lights, the range in luminous efficacy levels is very wide. Electric lamps have luminous efficacy levels that range over a factor of 35, from 1.7 to 60 lumen per Watt. The non-electric lights in the sample range from about 0.1 to 1.4 lumen per Watt. These wide ranges result from the different technologies used and from the wide range in power of the selected lamps.

Careful design of fluorescent light ballasts and luminaires is required to prevent unnecessary low light output from lights in solar home systems.

3.1.4 Illuminance

The amount of light that falls on a surface is called the illuminance[9]. Illuminance of a surface area by a light source is one of the characteristics of a luminaire in which the users are especially interested. For reading, a minimum illuminance of about 50 lux is required. For general lighting 20-25 lux is sufficient, and for orientation lighting 10 lux or even lower can be

[8] Luminous efficacy is the technical term for the efficiency of a light. It is obtained by dividing the luminous flux by the power consumption. The unit is lumen per Watt [lm/W].
[9] The measurement unit is lux or lumen per square metre.

sufficient. In figure 3.5 illuminance levels are shown for light emitted in downward direction (relevant for ceiling mounted, general lighting).

Figure 3.5 shows the illuminance levels of the lights in the sample for which luminous intensity measurements are available in nadir direction (downwards). These are typically ceiling-mounted lights. Note that some of the lamps, for example the solar lanterns, do not radiate downwards. The vertical axis of figure 3.5 shows the illuminous efficacy, i.e. the illuminance level per Watt power consumption [lux/W].

For designers of lights for solar home systems, the following can be concluded. Especially the fluorescent tubes with reflector appear to be optimal in combining a high illuminance level with a high luminous efficacy. Adding a reflector to the fluorescent tube luminaire produces an approximate threefold increase in the efficiency, even though the luminous flux of lights with a reflector is about 10-20% lower due to reflection losses. In the low power range the cluster LED is relatively very efficient, but its illuminance is very low. This was one of the first prototypes of this new type of lighting and it will be improved greatly within near future.

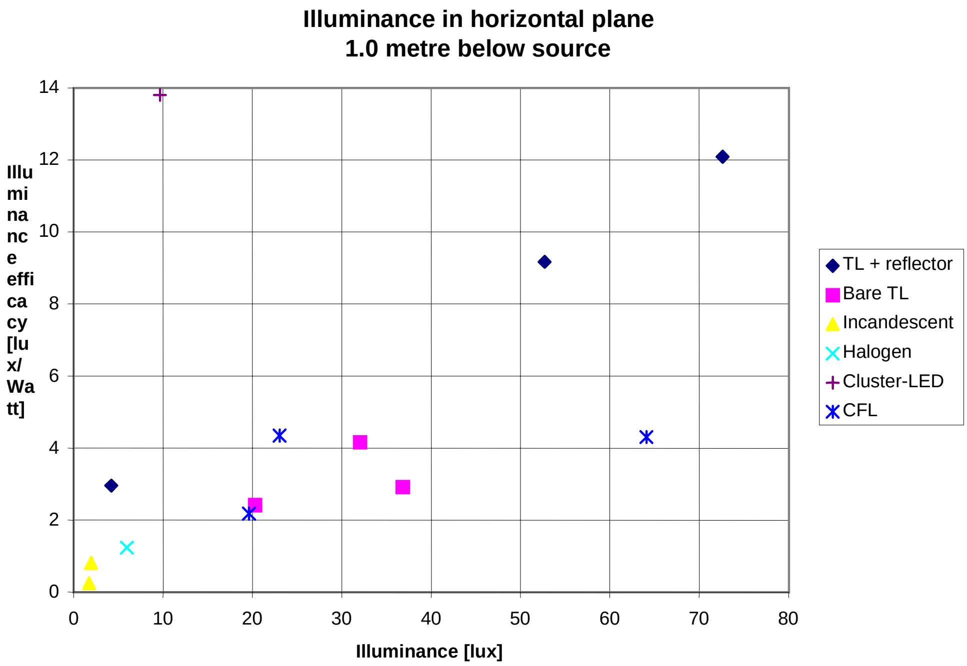


Figure 3.5 Illuminance [lux] and illuminous efficacy [lux/Watt] in a horizontal plane 1.0 metre below the light source.

3.1.5 Measurements of luminous flux

Usually, the luminous flux of electric lamps is measured in an integrating sphere (Ulbricht sphere, figure 3.6). An alternative approach is to use a goniophotometre in which the luminous intensity is measured in different directions and integrated to obtain the luminous flux of the lamp. Both methods require expensive equipment operated by specialists.

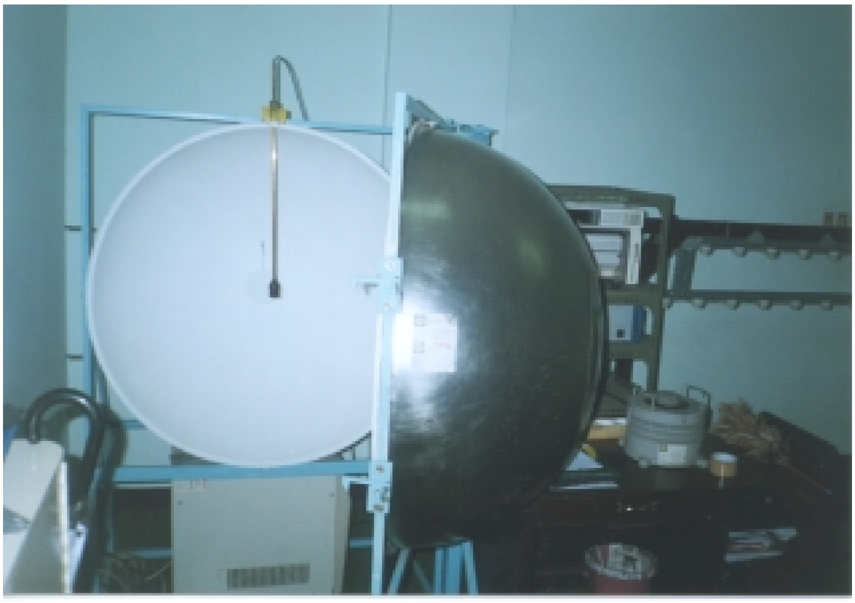


Figure 3.6 Ulbricht sphere for the measurement of luminous flux

Changes in the design of fluorescent light inverters, as will be discussed in section 3.2, will change the luminous flux and luminous efficacy of the light. Usually, organisations involved in modifying the equipment will not be the same as those testing equipment for certification. When modifications are required, the luminous flux of the light is usually already known. This is an advantage for institutes involved in modification, since it is much easier to measure changes in luminous flux than to measure its absolute value. In the rest of this paragraph it is assumed that the luminous efficacy of a light is known.

The following procedures lead to the determination of a luminous flux change, starting with a light with a known luminous flux:

- Measure the illuminance of the light in one direction [in lux] before modifications (= E_b), using a lux-metre (see next paragraph);
- Conduct modifications in the design;
- Measure the illuminance of the light in the same direction after modifications (=E_a).

The resulting luminous flux after modifications (Lu_a) can be calculated from the luminous flux before modifications (Lu_b) using the following formula:

$$Lu_a = (E_a / E_b) * Lu_b$$

To check if the design modifications produce an increase in efficiency the luminous efficacy before and after the modification are compared. It has improved if:

$$Lu_a/P_a > Lu_b/P_b$$

With Pa the DC input power after, and Pb before modification.

The illuminance of the light [in lux] is measured with a lux-metre. The light is mounted in a fixed (reproducible) position and distance from the lux-metre. This distance between lamp and lux metre is not critical since it involves a relative measurement. But it is important that this distance is exactly the same in both measurements. The light should be operated at nominal voltage, and current and voltage should be measured simultaneously. Stray-light is prevented from influencing the measurement, using a closed box of wood or other sturdy material of about 0.5 × 0.5 × 1.0 metre. The lamp is mounted at about 1 metre distance from the lux metre. Because the light output of fluorescent lamps is temperature dependent, the lamp should be allowed to stabilise for about 20 minutes before a lux measurement is conducted. During this time the box should be left open to prevent heating of the box. The box is closed during the lux measurements. Also, the ambient temperature of the two measurements should not differ by more than about 3 ^{0}C.

3.2 Fluorescent lamp inverters (FLI)

Some of the sections of 3.2 make use of the report “Optimal design of DC Luminaires for Solar PV application” - 1998, by Hasna J.Khan, Rezwan Khan and Salma Khatun, World Bank.

3.2.1 Introduction

Small local companies making fluorescent lamp inverters do not always have enough highly educated technicians available. Young graduates in electronics prefer to work for large organisations in the aircraft or computer industry, instead of working in rural areas. Because of this, many local manufacturers work with old circuit diagrams, sometimes based on poor designs, mainly caused by the lack of knowledge and information. Companies know the complaints from the field: early blackening, and a luminous flux, which degrades very quickly. It would be very beneficial for small producers of solar home system equipment to obtain technical support from a central research institute or university. These could provide technical advice about improving certain existing designs.

One of the weak components in the SHS is the fluorescent light. In many cases the tube starts blackening, or totally stops working very soon after installation. Another common problem in the field, is fatal damage of luminaires during transport to rural areas.

The most common causes of DC lamp failures in rural areas can be attributed to the following factors:
1- Variation of operating conditions compared to the rated design conditions,
2- Inefficient design stemming from insufficient technical information on DC-lamps,
3- Unavailability of reliable components for use in lamp construction,
4- Misuse of lamps originating from a lack of instructions for the user,
5- Insufficient quality control during manufacture of the lamps.

Not all of these issues are equally important for a superior field performance. However, these are some of the possible causes of problems that are associated with the design of DC-lamps.

3.2.2 Working principle of a fluorescent light

A fluorescent lamp (FL) is a high efficiency light source where electrical energy is converted into light with low losses of energy as heat.

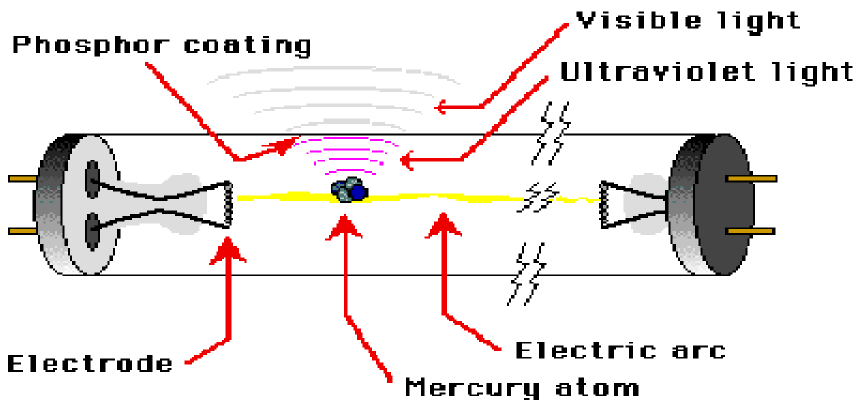


Figure 3.7 Schematic diagram of a fluorescent tube light

A schematic diagram of a fluorescent lamp tube construction is given in figure 3.7. The glass tube coated with fluorescent paint (Phosphor coating) is filled with mercury vapour. Tungsten filament connections are provided at the ends of
the tube, which are the sources of an arc discharge. Mercury can dissolve the filament material (by forming amalgam) and shorten the life of the filament. To prevent a direct contact with mercury vapour, these filaments are coated with an oxide layer.

A fluorescent lamp tube directly converts the electrical energy to electromagnetic radiation (visible light). When an electric current passes through the mercury vapour, the molecules are first ionised by absorbing kinetic energy. When an ionised atom recombines with an electron, and the electron drops back to its original energy level, the mercury atom radiates ultraviolet light. Ultraviolet light is not visible by human eyes. So, a fluorescent material is painted on the inner surface of the tube that converts the ultraviolet radiation into visible light.

AC voltage is needed, but at what frequency? Grid connected fluorescent lamps work on 50Hz. Ballasts for Solar Home Systems work on 20kHz or higher. Within SHS the power supply is a 12Vdc battery. The electronics inside the lamp fixture have to be as small as possible. A transformer is needed, but the size of the transformer can be smaller when the frequency is higher. At higher frequencies the efficiency of the FL is higher. At 20kHz this efficiency is 15-20% higher that at a frequency of 50Hz. On the other hand, when the frequency is too high, the lumen output will decrease, the internal resistance of the tube will be higher, so the voltage at the terminals will also rise.

With the onset of ionisation, the resistance of the tube drops dramatically and there is an inrush of current through the circuit. This non linear dependence of resistance of the tube on the current flow makes the operation of the fluorescent tube a unique component to the ballast circuit.

There are two techniques of initiating ionisation inside the tube. They are named as ‘hot start’ and ‘cold start’. In case of hot start, the filaments are initially heated by passing a strong current through the filaments only. The filaments have two functions: these are the electrodes for providing the electrons and the filaments for preheating. Heating of the filaments increases the emission of electrons from the filaments, which makes the ionisation easier. During a cold start the applied voltage across the tube terminal has to be higher than with a hot start. From the user point of view, the hot start requires some time (only a few seconds) to start up the lamp, but the cold start technique can start the lamp within a few milliseconds effectively making it almost instantaneous. Section 3.2.8 explains more about preheating.

3.2.3 Blackening of the fluorescent lamp

Degradation of fluorescent tubes starts when the ends become black. End blackening of the tube is the predominant mode of premature failure of fluorescent lamps. When a tube light is switched ON, it does not remain ON all the time. During each half cycle, it remains ON as long as the voltage across the tube is large enough to keep the gas inside the tube ionised. When the AC voltage passes through its zero crossing, the light goes OFF and becomes ON once again when the voltage magnitude rises in the next half cycle. When the light is ON, a number of things happen inside the tube light. Firstly, the gas molecules become ionised and start moving towards the ends due to the effect of the electric field. Hence, the ionised molecules become physically separated as the positively charged ions and the electrons move in opposite directions towards the negative and positive potential ends of the tube respectively. When the voltage wave goes through the zero crossing, these separated ions start their reverse journey to recombine and, if given enough time, almost all of them will reach the original energy level. Under such a circumstance, a significantly large voltage will be required across the lamp tube terminal to start it in the next half cycle. It is frequently observed in the current and voltage wave shapes of the tube lights operating at 50/60Hz that the current starts to flow after zero crossing when a large ionising voltage is build up across the tube light. On the other hand, if the frequency is high, there is not enough time for the electrons in the mercury atoms to fall back in their original energy level. Magnitude of current flowing through the tube light represents the actual number of molecules/atoms ionised. Lower magnitude of current means fewer ionised molecules and the voltage level required to restart the lamp after each zero crossing becomes larger. When the frequency is too high, the efficiency goes down and the terminal voltage will rise dramatically because of a too high internal resistance of the tube. Less atoms can be ionised, because there was no full recombination. *Higher terminal voltage causes the ions to bombard the filament with a higher kinetic energy and the oxide coating on the filament gets damaged at a faster rate.*

Four general causes for premature failure of DC-lamps can be defined:

1) *Damage of the filaments due to overvoltage across the tube:* ***End blackening***

High voltage at the tube terminal due to low DC input voltage:

Lower input DC voltage results in a lower magnitude of current in the tube light. Because its internal resistance is now higher (less molecules are ionised), the tube light terminal voltage is automatically raised. Hence, the oxide coating is damaged earlier than expected and mercury deposits on the filament causing corrosion by amalgamation. Fluid amalgam is sprayed from the filament by the effect of the AC potentials. **The sprayed amalgam shows up in the form of blackening of the tube ends.** Blackening of the ends has a two-fold effect.

(a) The amount of mercury vapour left inside the tube to form the ions is reduced as some of the mercury forms amalgam with the filament material.

(b) Shortage of the mercury vapour reduces the current flow with a consequent increase of the terminal voltage. This increase in terminal voltage enhances the rate of damage to the filament. Therefore, once blackening of the tube starts, the process of blackening continues even with the correct voltage.

High voltage at the tube terminal due to high DC input voltage:

This is a secondary cause of high voltage appearance at the tube terminal. With increasing input DC voltage, spikes may appear in the voltage across the lamp tube, because the transformer becomes saturated due to an incorrect number of turns in the primary winding, particularly if the number of windings is low. More of the transformer calculations can be found in section 3.2.10.

Why blackening frequently occurs at one end of a tube

Blackening takes place more quickly on one end of the tube. Because an asymmetric voltage wave shape is generated by the ballast circuit and high voltage spikes are generated during one of the half cycles. Usually, positive ions (mercury molecules) have a higher momentum compared to the electrons and damage is done on the negative potential terminal. In the case of a symmetric wave shape, the damage due to over voltage is shared equally by both the filaments. If the wave shape is asymmetric, one of the filaments is damaged more than the other and the lifetime of the lamp is drastically reduced.

2) *Contamination of the gas inside the tube*

Mercury vapour inside the tube may be contaminated if oxygen remains inside the tube due to some fault in the manufacturing process or due to leakage. In both cases, oxygen in the air reacts with mercury to form oxide and the life of the tube light is drastically shortened. Contamination of the mercury will require a higher voltage to ionise the gas inside the tube and this higher voltage will gradually damage the filament as explained before. This is not a major factor in lamp failure, although poor quality gas mixture is known to produce less light and reduce the tube life.

3) *Damage of the filament due to excessive current flow while preheating*

The filament of the tube light is designed to preheat the gas inside the tube for a short while, only at the time of starting. Excessive current flow or a longer period of preheating can result in filament overheating. The filament is damaged, when the filament and its oxide layer are burned much earlier than the expected lifetime of the tube.

4) *Damage of the filaments due to high current spikes.*

In some situations the filament can be damaged even without an excessive current flow, if the current wave-shape contains large spikes. Current spikes of short duration on the secondary side of the transformer cannot be detected in the input current consumed by the ballast circuit. But these short current-spikes mainly flow through the preheating capacitor (section 3.2.8), so through the filaments. Hence excessive heating will occur at the filaments. Under such a condition, a given lamp current will produce more heat and less illumination. Hence, it is important to design the ballast circuit to avoide high current spikes.

General recommendations for an optimised design for DC-lamps:

- The generated wave shape should be symmetrical in both half cycles.
- There should be no voltage spikes on the output wave-shape at the tube terminal. (low Crest-factor[10]).
- The ballast should normally be designed to operate the lamp tube at its rated power, otherwise blackening will occur very quickly in case of over-design, or the ballast is overloaded for an under designed ballast unit.

Possible remedies against blackening of the tubes

It is clear from the above discussion that rated current should be maintained to avoid blackening of the tube lights. This can be achieved:

- By keeping the input DC voltage constant, or
- By keeping the voltage across the lamp constant.

Using proper electronic circuitry it is possible to keep the input DC voltage constant even if the DC battery voltage is reduced (zener diode or voltage stabiliser). If the voltage control is to be applied at the tube light terminals, a variable frequency output voltage is needed. Normally, there is an impedance connected in series with the tube light to control the current through the lamp when the light is ON. If the series impedance is an inductor (usually the leakage inductance of the output transformer), the frequency should be reduced to reduce the voltage drop across the inductance. If the series impedance is a capacitance, frequency should be increased to get a similar effect. Changing the frequency is not possible with the conventional self-oscillating circuits. It is possible to obtain a variable frequency output if the oscillator circuit uses a Voltage Control Oscillator (VCO). A good electronic ballast circuit design ensures the long life of a tube light.

3.2.4 Lamp inverters with one or two transistors

As previously stated, a symmetric wave shape on the lamp terminals contributes to a longer lifetime. Some of the inverters tested by ECN were based on a one-transistor design. Two of the resulting waveforms are shown in figures 3.8 and 3.9. The upper diagram shows an inverter with a substantially asymmetric crest factor of almost 4.5 (!). At the start of each cycle, a very high voltage peak is supplied at its terminals. This causes early blackening on one side of the tube. Because the electronics are not built symmetrically, the tube will not light up during the negative half of the wave. There is current flowing, but the voltage was not high enough to start the ionising process. The lamp is OFF during this half cycle.

[10] The crest factor is defined as the ratio between the peak and the root mean square (RMS) value of the voltage or the current.

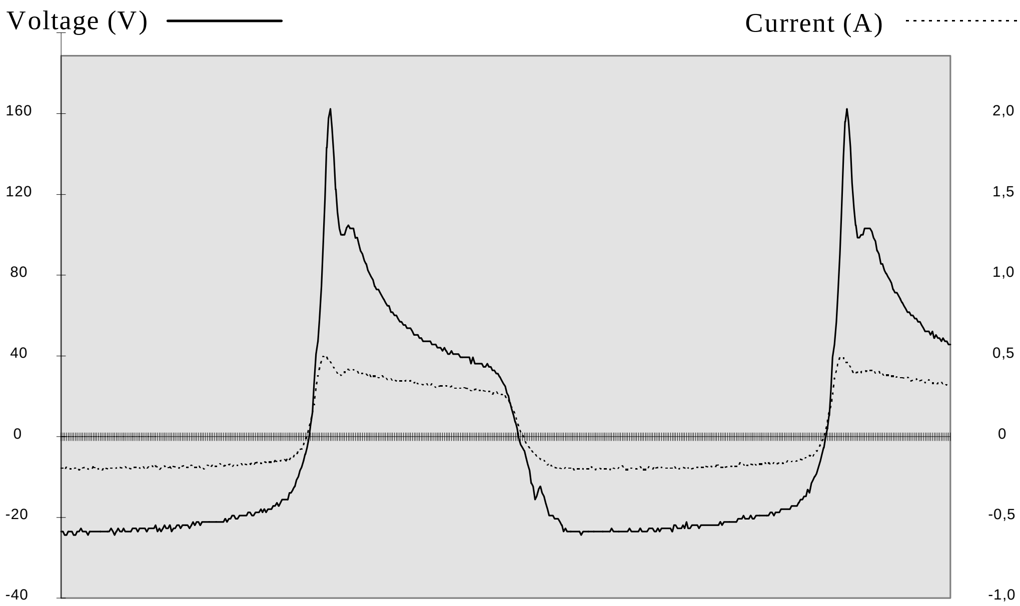


Figure 3.8 Waveform of a one-transistor DC-FLI

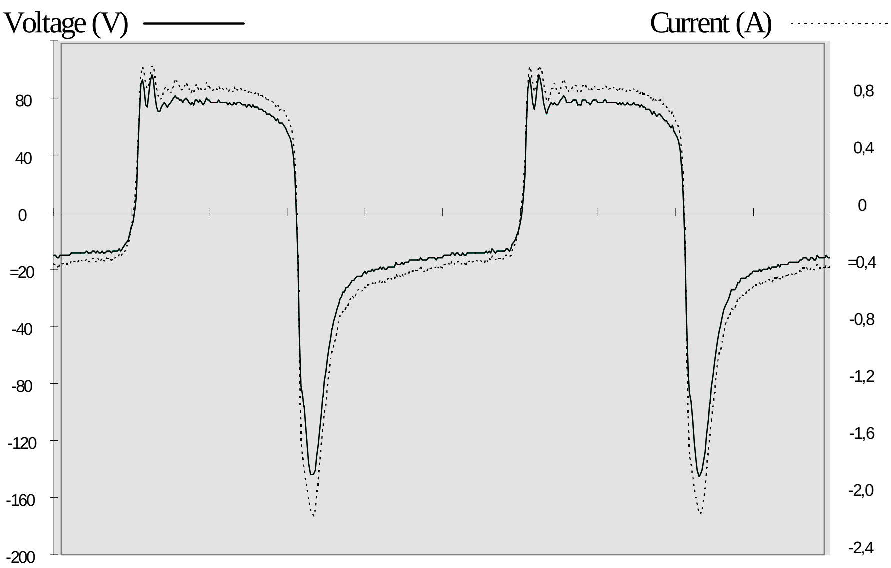


Figure 3.9 Waveform of a one-transistor DC-FLI

The most common design of HF-inverters uses a balanced LRC-circle to drive two transistors in a sine wave. Without a tube the output is a sine wave with an amplitude of about 600V. As soon as the tube (with its unique resistance characteristic) is added, it is a problem to maintain that 'ideal' sine wave (figure 3.10). The inverter below gives an example of a two-transistor inverter, which creates a more or less symmetrical waveform.

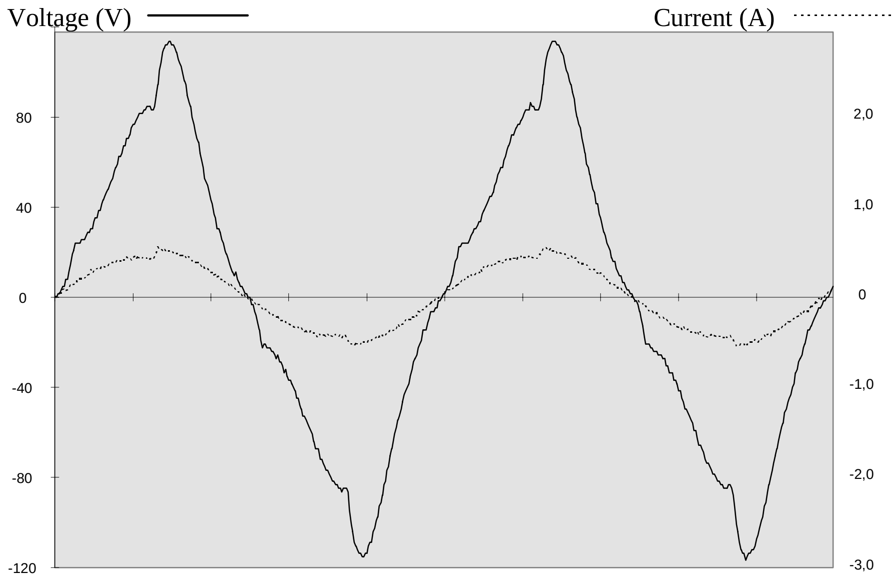


Figure 3.10 Waveform of a two-transistor DC-FLI

A remarkable factor is that the basic shape of the wave does not matter. There is no difference in lifetime with a square or a sine wave inverter, as long as the basic demands are regarded: low crest factor, symmetrical wave shape, etc.

During testing, the waveform should be checked in detail on the zero-crossings on the primary side, on the collectors of the transistors. If these zero-crossings have 'overlap', there will be unnecessary energy loss and even voltage and current peaks. During the overlap time the two transistors are both partially conducting. A short circuit is created via the transformer. Figure 3.11 illustrates this situation. The moment that a transistor starts its function at its half cycle, the other transistor has to be totally closed.

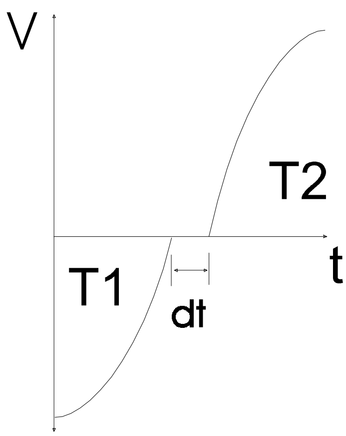


Figure 3.10 Well-trimmed transistor take-over: Δt>=0.

On the other hand, if the time Δt is too long, the luminous efficacy will reduce unnecessarily. Because of the LC-circuit at the secondary side of the transformer, the effect of this is difficult to detect in the tubes terminal voltage wave.

3.2.5 Different transistor set-ups

Firstly, two basic principles will be explained. Subsequently four different samples are discussed. These four types of ballast represent four degrees of performance and they were all manufactured in different developing countries.

In general two types of ballasts can be identified: the Chopper circuit and the Push-Pull configuration. The cheapest one is the Chopper design. The basic diagram is shown in figure 3.12.

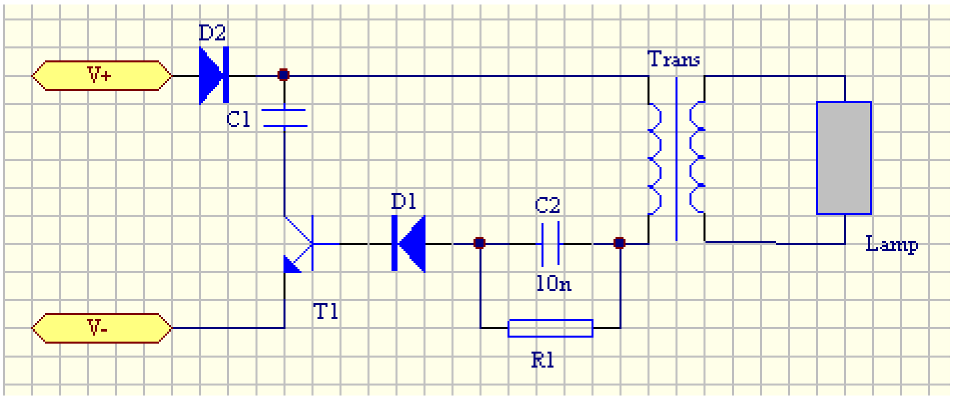


Figure 3.12 The Chopper design

When the system is switched on, a DC-current will charge capacitor C2 via the primary winding of the transformer. As soon as the anode of the capacitor is higher than 1.4V (Vbe of the transistor + voltage drop D1) the transistor starts conducting. C2 discharges via T1. When it is discharged the transistor closes again, so C2 can be charged. The charging is ‘chopped off’ when it reaches a certain level. This creates an AC-current in the primary side of the transformer during the ‘chop’. The result is a non-symmetric output signal, only functioning in one half a cycle. Most of the time there is some switching signal during the other half cycle, but this is not enough to ionise the lamp. During this half cycle the lamp will be off. The RC-time of R1/C2 and the inductance of the primary winding define the frequency of this system.

Diode D2 is the polarity protection at the DC-input. When T1 is conducting, the DC input is in short circuit. That means a lot of energy loss and a lot of heat. C1 is needed in order to avoid the DC short circuit. It only passes on AC-current, so the discharge energy from C2.

The second set-up is the Push-Pull configuration, shown in figure 3.13

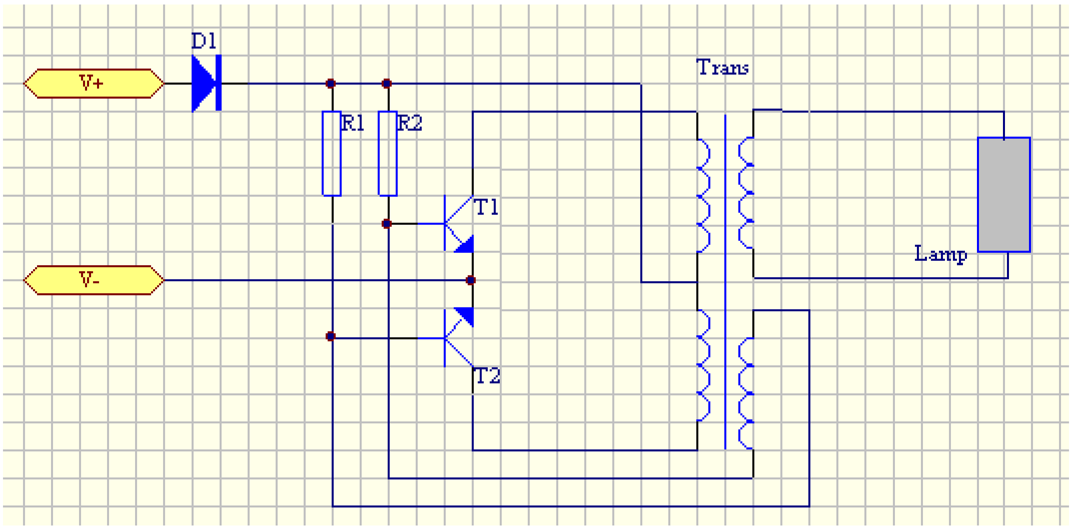


Figure 3.13 The Push-Pull configuration

Again, D1 is the polarity protection. Now the primary windings create a symmetric magnetic field in the core of the transformer. One half cycle T1 is making the field, the other half cycle T2 reverses this field. The feed-back winding is very important. This winding forces T2 to close when T1 is conducting and vice versa. So it actually is responsible for the oscillation. Without this, the two transistors would both be conducting, forming a short circuit via the primary windings.

Transistors are never *exactly* the same. When this system is switched on, the fastest transistor will conduct first. This creates a voltage difference at the feedback winding, which closes the slower transistor until the field in the core is totally built up. At that moment there is no voltage difference (which forces the slower transistor to close) at the feedback winding, so the slower transistor takes it over by reversing the field, while the faster transistor is now forced to close by the feedback winding.

The result is a symmetrical waveshape at the lamp terminal. During both half cycles there is enough voltage and current to start the ionisation process.
Four different samples:

Set-up 1: In this circuit no attention was paid to optimising of the circuit. The lamp had a very early stage of blackening and the efficiency is too low.

Set-up 2: This lamp was designed with a diffuser for area lighting, with little attention to circuit performance optimisation. Hence the circuit design was not efficient.

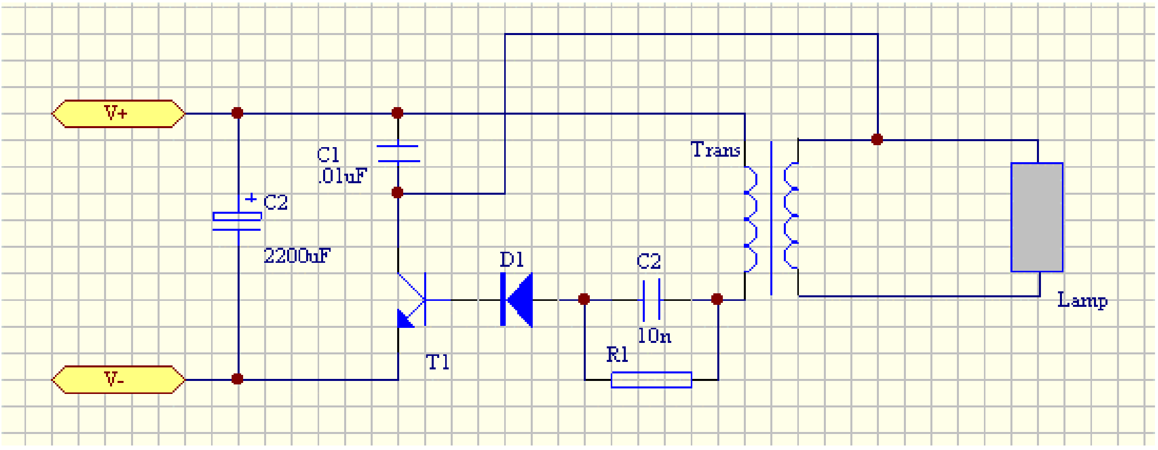
Set-up 3: The performance of this set-up was expected to be better than set 2. This design qualified other standards and the circuit performance was optimised by the manufacturer.

Set-up 4: This lamp was the best design in terms of efficiency (luminous efficacy) and ballast circuit performance. This design was known to meet stringent quality standards.

Set-up 1:
Rated operating Condition:

Input DC voltage : 12V
Output power : 8W

Diagram of the ballast circuit:



Main weaknesses in the circuit of design 1 are:

A. No prevention for voltage and current peaks at the lamp terminal
B. Thus producing a bad crest factor
C. Very asymmetric waveshape
D. The transistor needs to be cooled, because it produces heat.

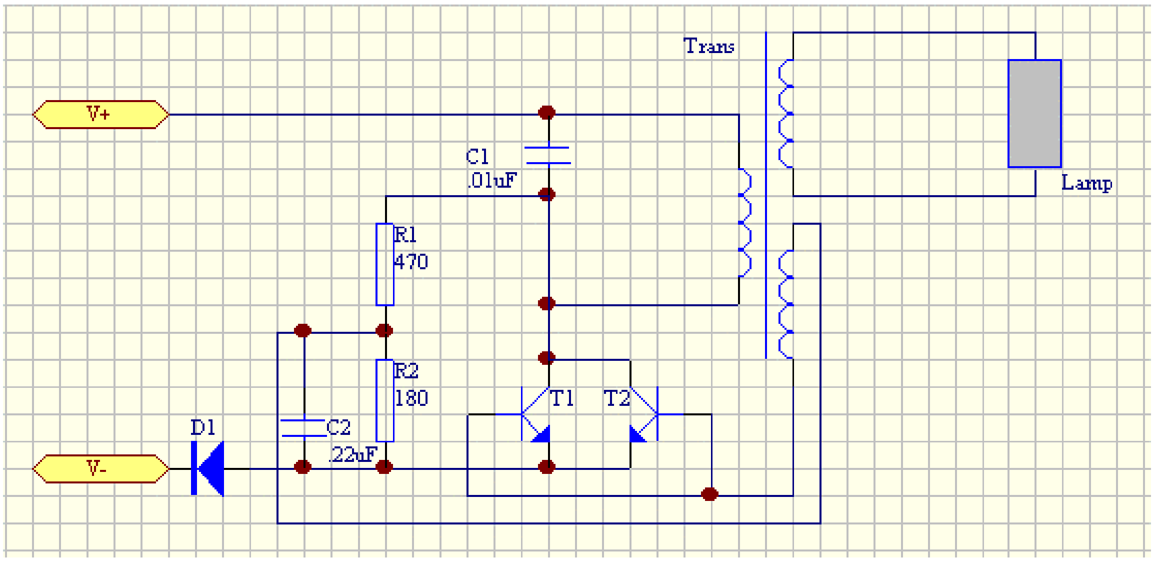
Dangerous point:

- No DC Polarity protection. There is an electrolytic capacitor (elco) added at the input (2200uF) which is polarity sensitive. This elco can be damaged after wrong connection, but it is not necessary in the first place. It is connected in parallel with the battery, so it functions like 'a very small battery next to a battery'. Its value is so high, that it is not a sufficient filter for high frequencies.

Design 2 is related to this design, so no further comments.

Set-up 2:
Rated operating Condition:
Input DC voltage : 12V
Output power : 8W

Diagram of the ballast circuit:



The principle of this design is about the same as design 1. The main differences are:

- Polarity protection on DC input
- Frequency generation at the collector in stead of at the base (.22uF / 180Ω).
- The oscillation (470Ω) thus needs to be activated.
- Two transistors parallel.
- Primary capacitor (.01uF)

Main weaknesses of this chopper design 2 are:

A. Two transistors have been used in parallel, apparently to share the load current. But it is not advisable to use parallel operation of the Bipolar Junction Transistors (BJTs) since they may not share the load equally. Due to the parallel connection of the transistors in this circuit, one of the transistors carries the full load and the temperature of the transistor in open-air conditions can reach 100 ^{o}C, which is extremely high. This may cause early damage of the transistors due to overheating and at the same time it indicates excessive power loss.

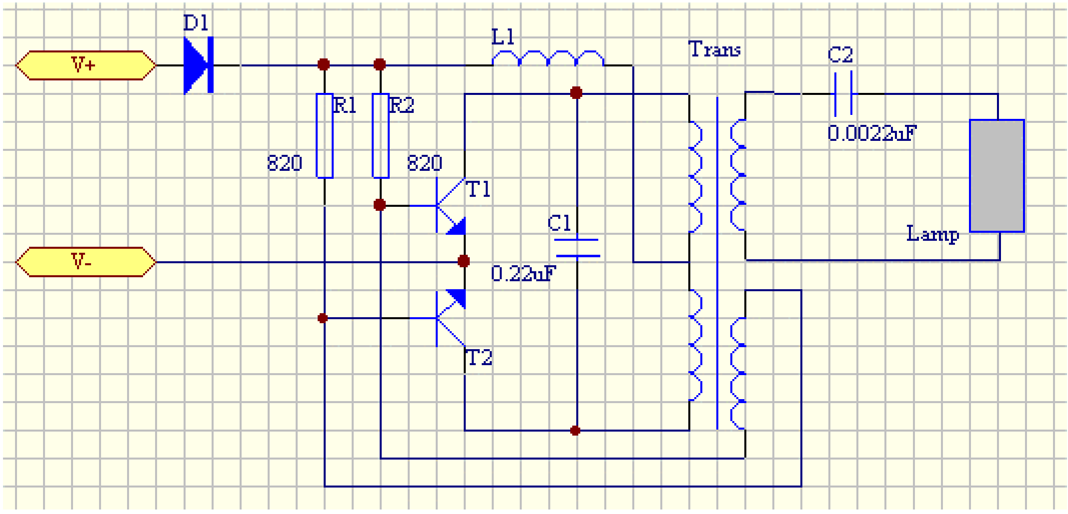
B. There is a high power 470 ohm resistance connected from the collector to the base of the transistors. This resistance is responsible for the initialisation of oscillation. As it is connected from the collector to the base, the resistance is effectively dissipating four times that required with a good design. An easy way to improve this is to replace this resistance by a higher value of resistance (around I K ohm) from the 12V supply to the base and a power rating of just 0.25W should be enough.

C. As far as the blackening of the tube light ends are concerned, it has been observed that the wave shapes become very distorted with voltage spikes at the output when the input DC voltage is reduced. During the preliminary tests, with 7V DC input, one end of the tube became black after only 25 hours of running.

Set-up 3:
Rated Operating Conditions:

Input DC voltage : 12V
Output power : 8W

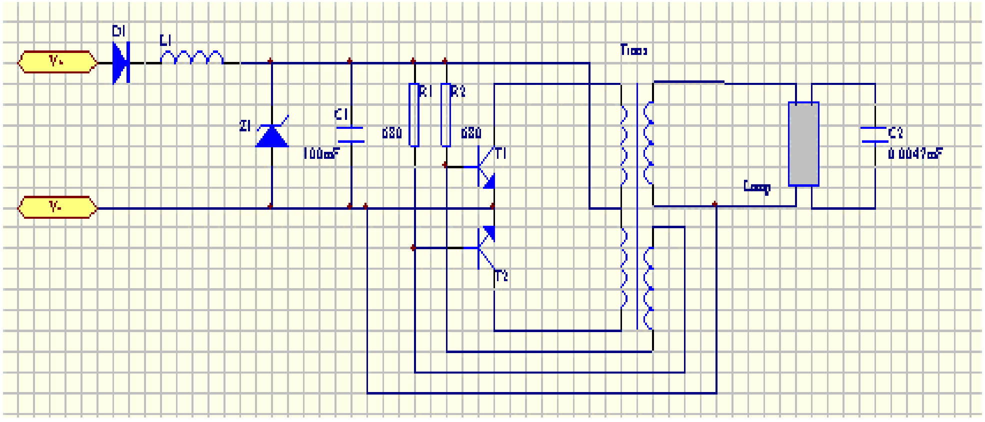
Diagram of the ballast circuit:



This Push-pull configured design was found to be a well designed circuit with very small loss in the transistors which was evident from the low temperature, which was nearly equal to the room temperature. The circuit used two transistors firing for each half cycle producing a symmetric wave shape. Control over the wave shape was much better and the wave shape distortion was much less with reduced voltage. Lumen per watt of the lamp was found to be quite high which reflects the small amount of loss occurring in the circuit. After 30 hours of running at 7V DC input, no apparent blackening of the tube fight was observed. However, in later tests, blackening was noticed after 137 hours of operation under this severe condition.

Set-up 4:

Rated operating condition:
Input DC voltage : 12 V
Rated power : 8 W



This proved to be the most efficient lamp design under consideration.

- On the input a constant voltage is maintained (Zener Z1).
- Symmetrical wave shape
- Preheating capacitor
- Low distortion

1. Uniform performance over the acceptable battery voltage range. Battery voltage in a solar home system varies from 14V to 11V depending on the depth of charging and discharging. Hence an ideal DC lamp should give uniform output over this voltage range. The usual DC lamps available worldwide give higher power output at higher battery voltage. To keep the power output uniform, an additional circuit may be needed which would marginally add to the cost of the DC lamps.

2. The frequency range over which most of the DC lamps operate is 20-30 kHz. This frequency range is chosen to give an optimum size of the ballast and to keep the switching loss of the transistors within acceptable limits. A higher frequency range would reduce the size of the ballast but increases the transistor switching loss. The leakage inductance (L) of the transformer has an impedance (ωL) which is proportional to the frequency. Any capacitor (C) used in the circuit has an impedance ($1/\omega C$) inversely proportional to the frequency. Hence changing any L or C alters the frequency and the overall impedance of the circuit, so correspondingly modifies the power input to the lamp.

3. Effect of low battery voltage: Voltage at the lamp terminals can be low if the battery voltage becomes low or the line loss is too high. Low voltage at the lamp terminals reduces current flow in the lamp. As the impedance of the lamp is highly non-linear and inversely related to the actual current flow through the lamp, its impedance increases with the decreasing current flow causing high voltage spikes in the voltage wave shape. These spikes cause high-energy ions to bombard the filaments and may cause irreversible damage resulting in blackening of the tube ends.

4. Effect of the voltage and current wave shapes: presence of high voltage spikes in the voltage waveform is the main reason for the blackening of the tube ends. The actual shape of the voltage wave (containing no spikes) does not affect the speed of the blackening process. High current spikes do not cause blackening if the spikes are not present in the voltage wave shape. It often happens that the current and voltage waves

contain DC offsets, especially those circuits without push-pull circuit configuration. These DC offsets are observed in the unequal peak values of the wave during two half cycles but cause no problem if spikes are not generated.

3.2.6 Polarity protection

Normally, DC-lighting is mounted on a wall or fixed to the ceiling for better access to the user. If people who buy and install new lights themselves do not have a technical background, they could easily make a mistake in the connection. This is one of the main reasons why polarity protection for DC-lighting is necessary.

Usually an input diode is used for polarity protection. These components have a voltage drop of about 0.6V. For a 12V DC-lamp with a rated power of 8W, this means 0.4W energy loss.

It is also possible to use a ‘POLY-SWITCH’. In normal use these have a voltage drop of about 0.4V, resulting in efficiency improvement of 3,3% towards the 12Vdc input. A poly-switch is current sensitive. When the current increases, the poly-switch resistance increases dramatically, so that the current is limited to a minimum. If the polarity of the ballast is wrong and there is no protection integrated, the reverse current becomes so large, that components are damaged. In the poly-switch, the current is reduced to a safe level, so that no components will be over-stressed. The lamp itself, however, will not be lit, because none of the transistors are in conduction, so no AC voltage is generated.

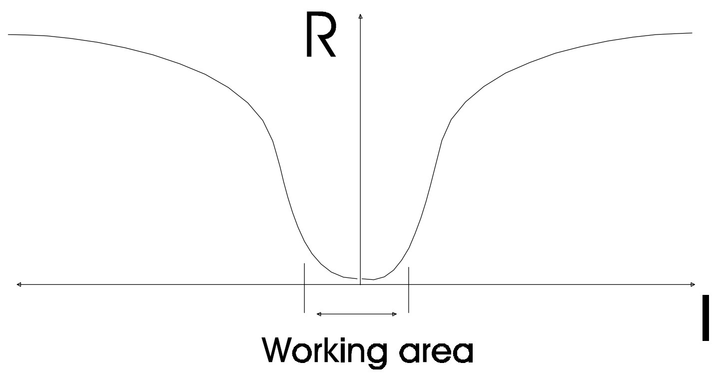


Figure 3.14 Characteristic of a polyswitch

A third option for polarity protection is the use of a MOSFET:

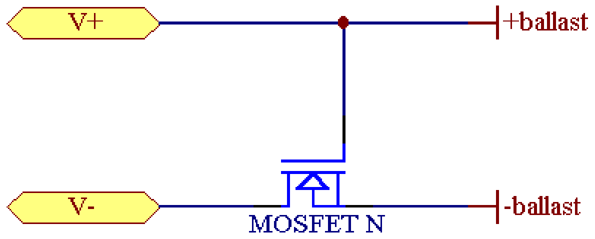


Figure 3.15 The use of a Mosfet for polarity protection

A cooled power-Mosfet is unnecessary in this case, because the current stays at about 1A. for an 8 to 10W fluorescent lamp. Depending on the internal conducting resistance, the voltage drop can be reduced to approximately 0.1 or 0.2 V. However, the price of a Mosfet is relatively high compared to that of a diode.

3.2.7 FET versus Transistor

Until now, most fluorescent lamp inverters (FLI) use transistors. Transistors were invented a long time ago and were commonly used when the first fluorescent lamp inverters were designed. Power MOSFETs were relatively expensive. Nowadays FETs have become cheaper, and when used in FLI they have certain advantages. For example:

- Less voltage drop, so less energy loss.
- Internal freewalk diode, so less voltage peaks when switching from OFF to ON.

Using FETs also has an important disadvantage:

- Because the input current of a FET is negligible, an oscillation with only a few components can not be established. An independent steering for the FETs has to be created.

When using MOSFETs, the steering needs to be different. You cannot just simply exchange transistors by FETs. Because no direct oscillation can be created, a signal-wave oscillator will have to be added. A simple square-wave is enough. If the frequency and the transformer-inductance are well trimmed, the output can be a good sine wave. Because of the freewalk diodes in the MOSFETs, less voltage and current peaks will occur. If there are, the same measures can be taken as for transistor-inverters. (See former paragraphs.) Also, a good input filter is needed between the main system and the high frequency part, because of the EMC effects on the system wires. The main system (without filter and other precautions) is shown in figure 3.16.

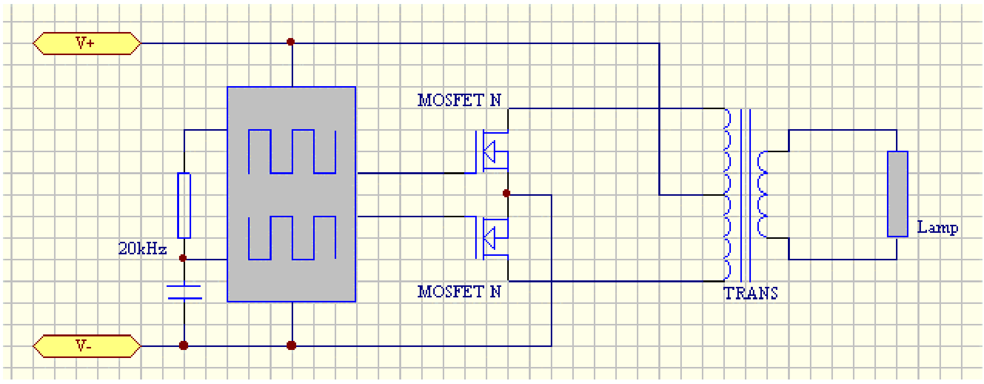


Figure 3.16 Example MOSFET-use in FLI

The Hefei University of Technology (China) created an identical inverter with such a perfect sine wave on the lamp (Crest Factor about 1.57) (!) that no voltage or current spikes could be detected. The disadvantage of their design was that they used a relatively expensive PWM IC for the steering part. It might be interesting to do some research on a FET controlled ballast with an analogue oscillation circuit.

3.2.8 Effect of filament preheating

Preheating the filaments helps to ionise the gas inside the tube and bring the lamp to its full glow more easily with less applied voltage. It is important to understand the actual mechanism of starting a DC lamp to ensure longer life of the lamp. When a lamp is switched ON, the lamp starts to glow immediately (without any noticeable time delay). It does not come to full glow instantly, but suddenly increases 4-5 seconds after the lamp is switched ON. During this period, the gas inside the tube is not completely ionised and the voltage across the tube remains high. The shorter the time required to come to full glow, the better it is for the lamp. Preheating minimises this start-up time and many of the manufacturers of DC lamp ballast circuits prefer filament preheating to ensure longer life. Figure 3.17 shows the configuration for filament preheating from the secondary side of the transformer. Four pins are connected to the tube-end for this configuration.

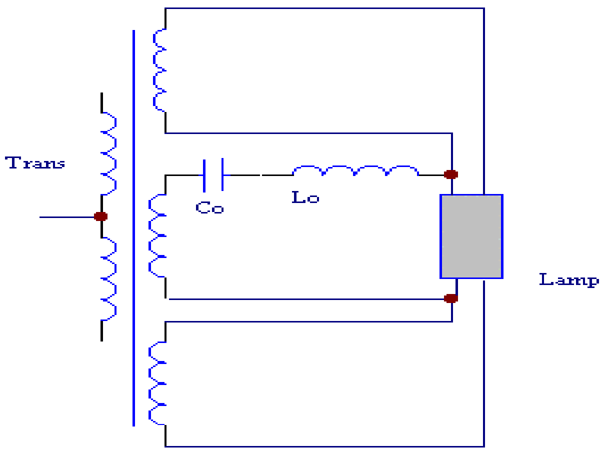


Figure 3.17 Configuration for Filament Preheating

Preheating of the filaments is simple and inexpensive. Two additional windings in the transformers can be provided for the filament supplies and can be connected at the ends of the tube. Although, not financially expensive, it will drain power all the time and the overall efficiency of the lamp will be reduced. However, the loss of energy is compensated by the longer life of the lamps. Hence, this trade off depends on the cost of energy compared to the cost of the lamp itself. In Solar Home Systems the cost of energy is very high, leading to many designs with cold start.

Another option for preheating, which does not make straight use of an extra winding at the transformer, is connecting a capacitor between the (not connected) pins of both filaments as shown in figure 3.19. The starting sequence for this type of FLI can be divided into three phases (figure 3.18):

- A phase where the voltage is being increased. The gas inside the tube is not yet ionised, so its internal resistance is very high. The voltage over the tube is being built up to 500 or 600 V. Because of the high internal resistance of the tube, this AC-current will flow through the parallel-capacitor. In this phase the current results in the heating of the filaments.
- The ionisation phase. When the voltage is high enough, the first ionisation will take place. The internal resistance of the tube will drop drastically. From now on the main current will flow through the tube.
- The steady state of the ballast and lamp. During this period the voltage will stay at approximately 60 V. During the ‘on’-period of the lamp, the current will flow through the tube, because of its low resistance. During the zero-crossings dI/dt is high and the lamp is off, so high resist. During this very short time current will flow through the capacitor. Because this is only a very short period, it will not contribute to effective heating as the filaments are still hot from the last cycle.

When the filament is being pre-heated, its temperature rises and electrons can leave the filament more easily. Free electrons gather around the filament, until the voltage is high enough to ionise the gas. Due to the higher number of free electrons, this ionisation can start at a lower voltage.

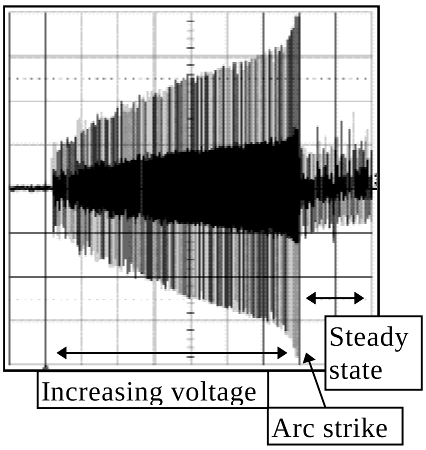


Figure 3.18 Starting sequence of a FLI

The important thing is that the overall efficiency is not reduced. Only one extra capacitor is required, so it is a very cheap and energy-saving method of preheating. The parallel capacitor is often used, in FLI of all price-classes.

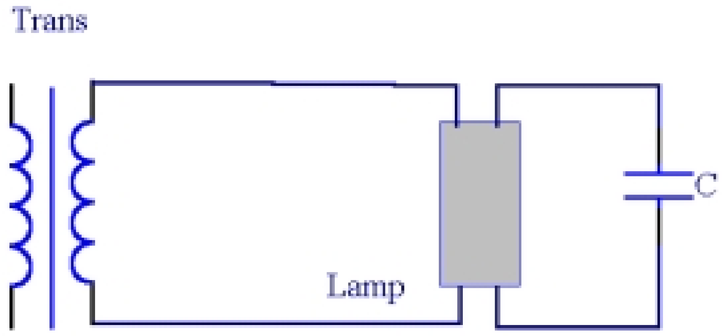


Figure 3.19 Filament Preheating with parallel capacitor.

3.2.9 Critical components and component quality

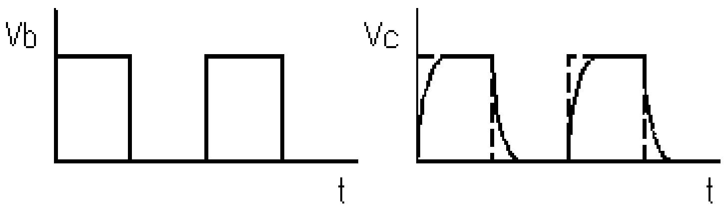
Following the basic alternative ballast circuit designs given in the previous section, this section discusses the role of the individual components in the overall performance and efficiency of the ballast.

3.2.9.1 The transistor(s)

The transistor is one of the critical components to determine the frequency of operation of the ballast circuit. A good choice of transistors is very important in the ballast circuit, because most critical components of the ballast are frequency dependent. The frequency of operation should be chosen considering the frequency response of the transistor. For an efficient operation of the transistor, the operating frequency should be chosen to be less than l/hFE (β) times the ultimate frequency limit fT , where hFE is the frequency response as a function of the gain factor (β). Mathematically speaking the operating frequency, f, should be set according to:

$$f < fT / hFE$$

If the operating frequency is higher than acceptable, the junction capacitance becomes prominent which makes switching the transistor ON and OFF slower. When the operating frequency is higher than the frequency mentioned above, junction capacitance makes the transition switching rate slower resulting in extra power loss inside the transistor. This situation is shown in the figure 3.20 (a) and (b). The base voltage VB has a square wave shape in (a), however, the switching function causes the collector voltage Vc, to be distorted with power loss, as shown in (b).



(a) (b)

Figure 3.20 Power loss during transistor switching

For the efficient operation of a transistor, it should either operate in the saturation or in the cut off region. Power dissipation inside the transistor is the product of transistor voltage drop and the transistor current. When the transistor is switched ON, it goes to saturation and the voltage across the transistor becomes very low (of the order of 0. 2V) although the current may be high. The product of the voltage and current is low due to low saturation voltage. When the transistor is switched OFF during the other half cycle, it must be able to handle the full voltage (open circuit) without breaking through. Hence the *product* of current and voltage is still very near to zero. Such an operation produces very little power loss and the transistor keeps cool, ensuring longer life and less possibility of failure.

3.2.9.2 The transformer

The transformer is the most critical component in the design of the electronic ballast circuit. DC lamps are normally operated in the frequency range of 20-30kHz. Because normal iron cores loose energy in this frequency range and ferrite cores are used. The voltage induced in a coil is given by Faraday's law of electromagnetic induction mathematically given by $Nd\phi/dt$, where N is the number of windings, ϕ is the flux inside the core and t is the time. If the operating frequency is high, $d\phi/dt$ is high. This means smaller amounts of flux can generate the required voltage and the actual size of the core can be reduced correspondingly. As discussed earlier, there are a number of physical constraints such as the switching loss and the price of the transistors which limit the upper frequency limit.

The transformer plays two very important roles in the successful operation of the lamp:
(a) Starting condition: generating high voltage during starting
(b) Running condition: producing the right value of inductance to prevent excess current to the lamp tube. It is very useful to experiment with the transformer, to verify the effects.

Ad.(a) Starting condition: Ideally, the ratio in the number of windings of the transformer should be such that the voltage generated at the lamp tube terminal is high enough to initiate ionisation of the gas. During this time, resistance of the gas is very high and the current through the tube is low which makes the inductive voltage drop Ldi/dt in the inductance small. Hence, the full voltage generated in the coil appears across the tube and the gas is ionised.
Ad.(b) Running condition. As the gas ionisation process increases, current flow also increases simultaneously and the voltage drop across the inductance increases, reducing the voltage across the lamp. When the lamp comes to full glow, the voltage drop across the leakage inductance should have a value so that correct voltage appears across the tube.

In most cases, it is difficult to satisfy both the starting and running conditions. In circuits without a capacitor across the primary side of the transformer (connected to the collector of the transistor(s)), current starts flowing in the primary coil when the transistor is ON. After half cycle, the transistor is suddenly switched OFF which tends to chop the current OFF instantly. This mean a very high value of di/dt which correspondingly induces very high voltage at the lamp ends helping the lamp to start even when there is a much lower voltage at the primary side of the transformer. Although this design may be suitable for starting the lamps, it has a number of drawbacks.

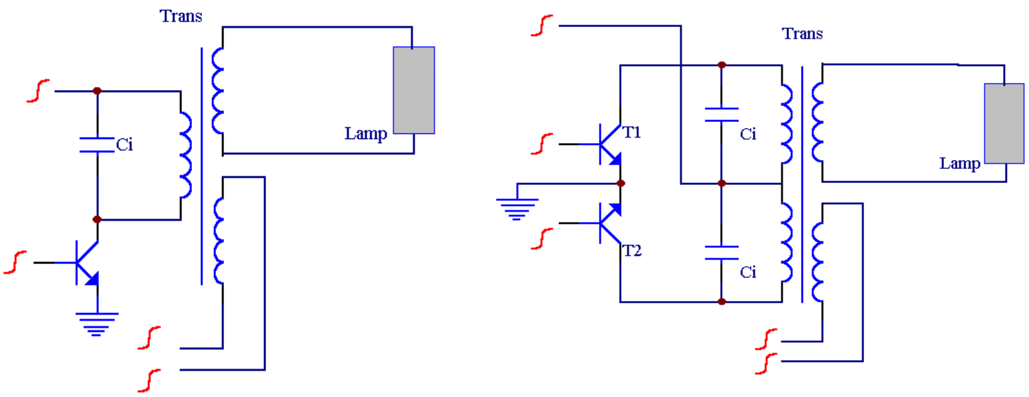
(a) Firstly, it makes the lamp glow at a much lower voltage, and the lamp may never attain its full glow.

(b) Secondly, even if the lamp comes to its full glow, it may take longer time. During the period of this partial glow, the ionisation of the gas in the tube remains incomplete. The impedance of the lamp remains much higher than that of the normal operation and high voltage remains across the tube. Hence, blackening of the tube ends take place much earlier.

One way to avoid such a damaging phenomenon is to prevent low voltage starting of the lamps: a lamp should start only when the input DC voltage is high enough to generate full glow. The solution lies with the addition of the capacitor in the circuit across the transformer on the primary side.

3.2.9.3 The capacitor and inductor on the primary side of the transformer

The capacitor across the primary winding of the transformer is an important component for ensuring the long life of the tube light and preventing end blackening. The two basic circuit configurations using capacitors across the primary winding are shown in figure 3.21.

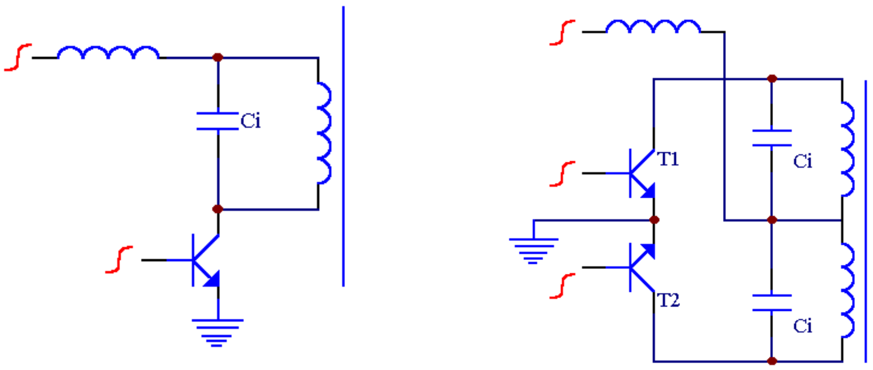


(a) Chopper circuit (b) Push-pull circuit

Figure 3.21 Capacitor across primary winding of transformer

When there is no capacitor across the primary side of the transformer, switching the transistor OFF from the ON state would generate high voltage spikes. As discussed earlier, this would initiate ionisation inside the tube prematurely even when the battery voltage is insufficient to produce sustainable full glow of the lamp. To eliminate this problem, a capacitor is connected in parallel with the primary windings of the transformer. When the transistor is switched OFF from the ON state, the current through the inductance is not interrupted, as there is a path for current flow through the capacitor. This reduces the voltage spikes and it is possible to design the transformer and choose the value of the capacitor in such a way that the lamp will be ON only when the DC battery voltage is at its right level.

Addition of such a capacitor would however cause current spikes through the transistor whenever the transistor is switched ON from the OFF position. This will increase power loss in the transistor causing over heating and reduce the overall efficiency. In extreme cases, the transistor may even get damaged. A small inductance in series would reduce any such current spike. The position of the inductance in the circuit is shown in figure 3.22.



(a) Chopper circuit (b) Push-pull circuit

Figure 3.22 Inductor in series with primary side of the transformer

3.2.9.4 The capacitor and inductor in the secondary side of the transformer (in series with the tube)

Role of the series capacitor during running: When a capacitor is added in parallel with the primary winding of the transformer as discussed earlier, the voltage spikes generated due to abrupt switching OFF of the transistors are suppressed. This causes difficulty in starting the lamp in the absence of the high voltage spikes. So, the transformer windings ratio is increased to have enough voltage at the output to initiate ionisation. However, such an increased voltage causes excessive current flow through the tube light under steady state (running) conditions and the lamp may get damaged. For this reason, a capacitor is added in series with the lamp to limit current in the circuit as shown in figure 3.23.

Role of the series capacitor during starting: It is interesting to observe that this series capacitor helps significantly in improving the overall performance of the lamp. It has already been mentioned that the lamp is a highly non-linear device with its resistance varying with applied voltage by a large margin. Every half cycle, the current through the lamp goes through a zero crossing and the lamp goes OFF momentarily. It will not start to glow again in the next half cycle until a significant voltage is reached. During this OFF period, the resistance of the tube remains very high. Once the capacitor is charged from one of the half cycles, it adds to the voltage of the next cycle until the lamp glows. This means the lamp always sees a higher voltage which is the sum of the capacitor voltage from the previous cycle and the voltage from the transformer. So the lamp starts much earlier each half cycle when compared to the situation when there is no series capacitor.

Role of the inductor in series with capacitor: Addition of the capacitor in series with the lamp has a disadvantage. When the lamp starts conduction after zero crossing in each half cycle, the resistance of the lamp suddenly decreases causing the capacitor to discharge very rapidly resulting in high current spikes. Sustained current spikes of high magnitude would generate more heat at the filaments and the overall efficiency of the lamp is reduced. The filaments may get damaged in the long run. However, the adverse effect of the current spikes are much less in comparison with the voltage spikes. Current spikes through the lamps can be suppressed by adding a small inductor.

The final form of the secondary side of the transformer circuit is shown in figure 3.23, which shows the configuration with a capacitor and an inductor in series with the tube.

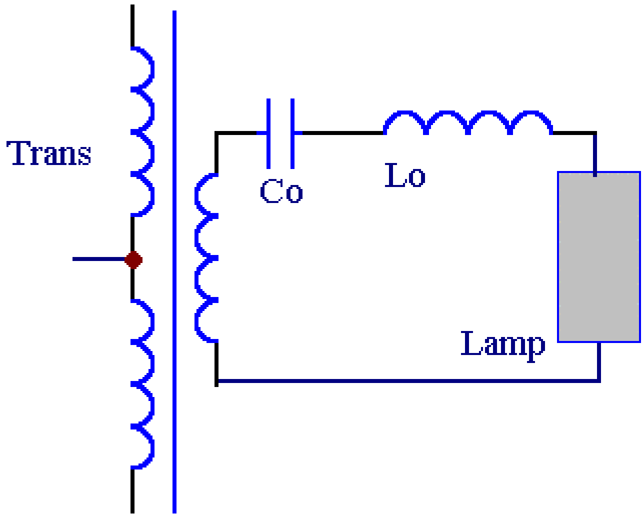


Figure 3.23 Capacitor and Inductor in Series with Tube Lamp

3.2.9.5 Tolerances of components

This chapter illustrates some specific types of variation in the performance of DC lamps from external causes, such as the quality of electronics components and spare parts used in manufacturing and maintenance. The results reflect on the significance of quality control and its absence in the markets of the developing countries. The objective is:

(1) to obtain information about the variation of input current consumption levels, which will affect the overall performance of the lamps,
(2) to demonstrate the variation of the actual from the rated values for inductance, resistance and transformer coil typically used in the DC lamps.

3.2.9.6 Performance of DC lamp ballast with different tubes

Fluorescent tubes used in DC lamps are typically purchased from the open market where various brands are available at different prices. Selection of the tube is strongly influenced by the price and past experience of the buyer. With an upcoming market for solar home system applications, the user is usually not experienced and the price and availability will guide the choice. A brief test shows the variation of input current consumption of the ballast in response to variation of the tube (using the available tubes from the market). Within the same brand of tube, once the tubes are replaced, current consumption is found to vary up to 5%, which is an indicator of the variation in tube impedance.

In some designs the luminous efficacy can vary up to 30% only by replacing the tube.

3.2.9.7 Current measurement of same ballast with different components

During the process of manufacturing lamps in the developing countries, availability of spare parts is a major constraint. Often, a replacement or equivalent component is used which is only nearly the same in performance. Using two different size capacitors of 0.22 and 0.47 μF, a combination of three resistors of 330, 470 and 680 Ω has been investigated. Current in Ampere has been measured for different tubes connected to ballast 2. It is evident that changing capacitor from 0.22 to 0.47 μF with identical resistors can increase the current by 50%. The rated current for this lamp being 0.8 Amps, the combination of 0.22 μF with 470 Ω is found to produce the best design conditions.

At the stage of optimising a ballast, sometimes component values have to be chosen, based on 'educated guesses' to see what happens. In practice, some components have to be chosen in an experimental way. A pre-calculated value can give a good starting point for the experiment, but will seldom work immediately, while the component quality is of high importance.

Quality control of the component is often the most critical issue being overlooked if the user is unaware of this variation. Percentage deviation up to 22% from the rated value has also been observed. Therefore, strict adherence to a reliable supply of components needs to be ensured for maintaining the desired performance of the ballast.

The leakage current and ESR (equivalent serial resistance) are very important factors of a capacitor used in FLI. They can have a large influence in the balance in the oscillation of the inverter. Furthermore they influence losses in the inverter.

Table 3.1 gives an overview of the quality of different types of capacitors.

Table 3.1: Some types of capacitors

Type	Cap. Reach	Vmax	accuracy	Temp. stability	ESR	Leakage current	Remarks
Ceramic	10pF..1uF	50..30000	--	--		-	• small • cheap
Mylar	.001uF..50uF	50..600	+	-		+	• cheap • good
Polystyrene	10pF..2.7uF	100..600	++	+	~0.05Ω	++	• high quality • filters
Polycarbon	100pF..30uF	50..800	++	++	~0.35Ω	+	• high quality • small
Polypropylene	100pF..50uF	100..800	++	+	~0.06Ω	++	• high quality • low absorption
Elco	0.1uF..1.6F	3..600	--	---		---	• short life • polarity • power supply

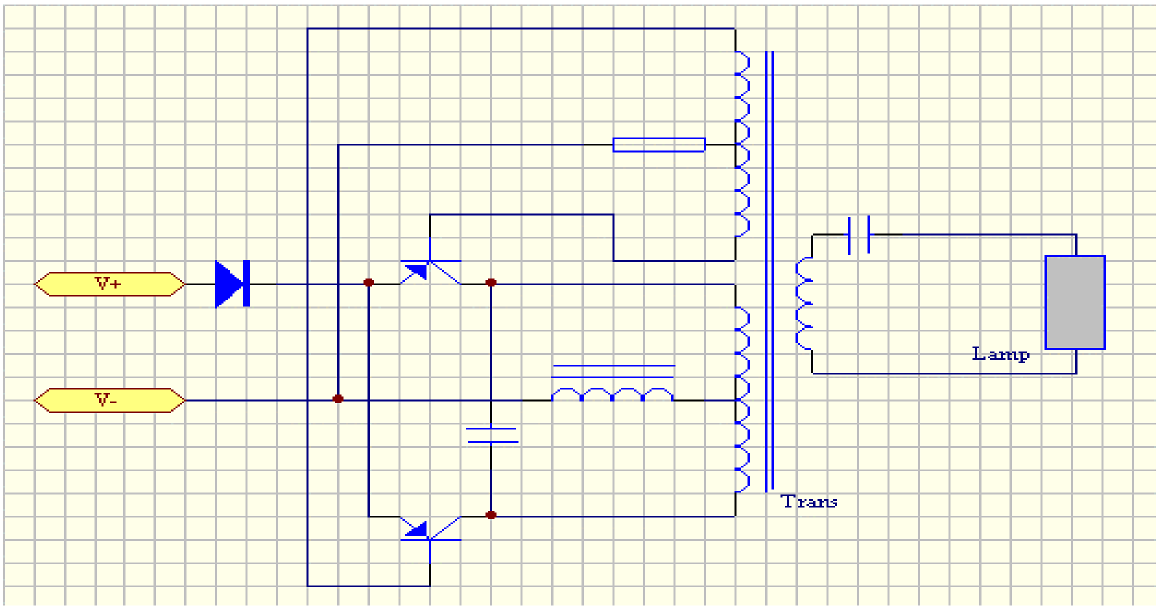
3.2.10 Inductance calculation of the transformer coil

The transformer is the most critical component of the fluorescent lamp inverter. Not only because of its important function, but also because it can be very fragile. *If a Solar Home System is being transported to remote areas, the inverter can be exposed to heavy shocks. The coil wire on the secondary (tube) side is quite thin. If this coil wire is dimensioned too thin, there can be a high percentage of failures in the field because of this under-dimensioning.* Hence the coil should be mounted very securely on the PCB.

Next, the coil dimensioning is discussed. For example: how to dimension the coil of a common emitter inverter with a tapped feedback winding.

Rated operating conditions:

Input DC voltage :12 V
Output power : 8W



First calculate the primary coil:

average current through the coil wire: 8W / 12V = 0,66 A

surface of the primary coil-wire = 1mm2/3A * 0.66 A * 0.5 (=Duty Cycle) = 0.11 mm2

Now the diameter (∅ prim.) = $\sqrt{(4 * 0.11)/\pi}$ = 0.376 mm

For the secondary coil:

Iout = 8W / 300V (working voltage) = 0.0266 A

surface of the secondary coil wire = 1/3 * 0.0266 * 0.5 = 0.0044 mm2

$\varnothing$ (sec.) = $\sqrt{(4 * 0.0044)/\pi}$ = 0.075 mm

As can be seen, this coil wire is very fragile, so we use 0.2 mm coil wire in any other further calculations.

The feedback coil:

Coil current = Ib = Ic / β = 0.66 / 30 (depending on the type of transistor) = 0.0222 A.

surface = 1/3 * 0.0222 * 0.5 = 0.0037 mm2

$\varnothing$(fb) = $\sqrt{(4 * 0.0037)/\pi}$ = 0.068 mm

Also this coil wire is too fragile, so we will have to take 0.2 mm wire.

The number of coil windings:

$$N = \frac{Vp * 10^8}{4 * f * Bsat * A}$$

with: Vp = input voltage - Vce = 12.0 - 0.3
Bsat = Saturation flux from the ferrite coil base = 2500 gauss
A = Surface of the coil wire

Primary coil: N = (11.7 * 108) / (4 * 20000 Hz * 2500 * 0.376) = 15.55 = 16 windings

Secondary coil: $N(sec) = N(prim) * \frac{Vsec.}{Vp}$ = 16 * 300 / 11.7 = 410.25 = 411 windings

A rounded figure of 415 windings is taken.

We took wire with a larger diameter, so we will have to add a series capacitor in order to maintain a safe FL-current.

Feedback coil: This output voltage will vary between 3.5V and 12V. The lower limit of 3.5V is used in the calculation:

$N(fb) = N(prim) * \frac{Vsec.}{Vp}$ = 16 * 3.5 / 11.7 = 410.25 = 4.7 = 5 windings

---

In practice, experimenting with the number of coil windings and diameter of the coil wire is required. The ferrite can get hot, implying substantial energy losses. In that case the diameter of the coils' ferrite should be enlarged or the total number of windings should be reduced. This type of calculation is a good starting point.

The ferrite core of the transformer can become saturated when the magnetic field intensity is too high. For example, consider figure 3.24. This figure is another possible set-up for a fluorescent lamp inverter. Its principle of working is a standard push-pull configuration. When Q1 goes from the 'off' to the 'on' position, the feedback (forward bias) is formed via L2 and L1. When Q2 is switched 'on' it creates a reversed bias. Due to this bias, Q1 is forced to go off when Q2 is on.

In practice, this does not work as planned, because the magnetic field of the core works against this. Figure 3.25 shows the magnetic field intensity in the core during one cycle. It has a hysteresis character. The cycle starts at point x. When we move to point d, transistor Q1 starts charging the core until it is almost in its negative saturation in point a. Because of the feedback, Q1 turns off at this point, after which B-H loop continues and goes to point b. From this point on the flux changes very quickly, through which it can easily can increase outside the saturation area. After the core almost reaches its positive saturation point, the loop starts again.

When the transformer is not dimensioned correctly, the magnetic field built up inside the core can increase outside its saturation area, which results in a temperature rise of the core. This saturation results in a clipped signal of the output voltage without the lamp. At ECN one type of DC-light was tested, which actually caught fire after 200 hours.

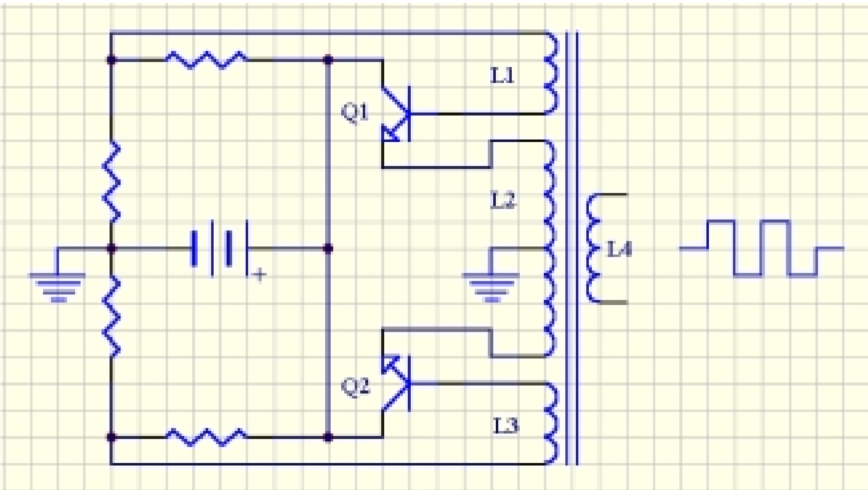


Figure 3.24: FLI-set-up for demonstrating flux saturation

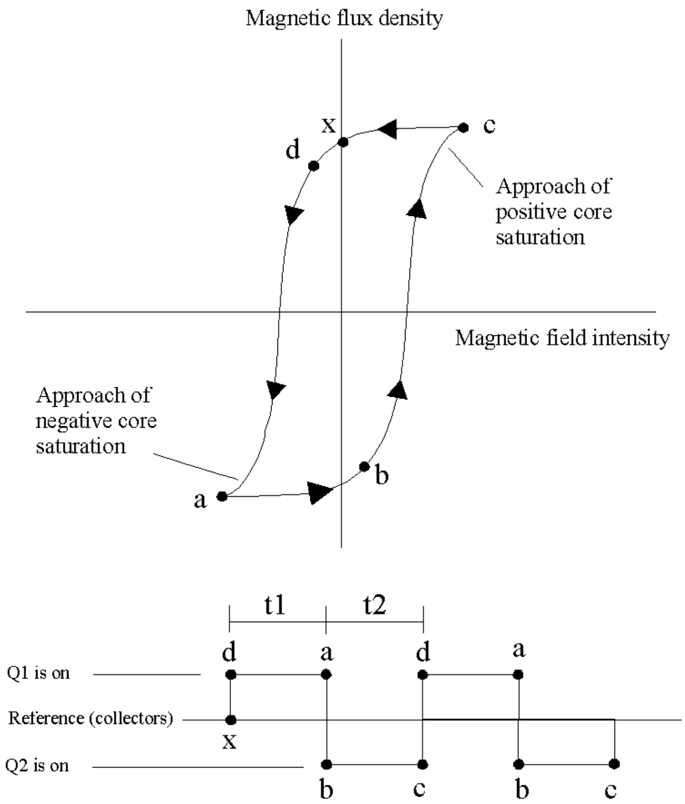


Figure 3.25 Core saturation area

3.2.11 EMC & PCB-design

The FLI is a device, which operates on middle to high frequencies. There is a risk of disturbing other equipment and it is very important that the electromagnetic emission of the FLI is reduced to the minimum. A fluorescent lamp should neither disturb other devices nor be disturbed by other devices. The emission can be divided into two factors:

- emission via the DC-input lines back to the system
- radio emission via the air

The transmission via the input DC-lines can be reduced by dimensioning the input impedance (input filter) correctly. This part of EMC has to do with discrete components. It is very important for the system that DC-lightning has no effects on the rest of the system. Radios can be disturbed by high frequencies on the system wires. Another important consequence of these frequencies is a rapid deterioration of the battery and a decrease of storage capacity, caused by the so-called ‘Skin-effect’. At 20 to 30kHz the energy can not be chemically stored anymore. The energy offered to the battery will collect on the surface of the plates and will find its way to the negative plate via the surface of the electrolyte (figure 3.26). This can be compared to VHF signals on transmitter PCBs. Open spaces on such a PCB should be filled with earth-sheets, to avoid VHF signals to interfere via the air. With the electrolyte inside batteries this effect can occur already at 10 to 20kHz. On the longer term this can be damaging for the battery. The intensity of this Skin effect depends on the amplitude of the AC signal component, superposed on the DC signal.

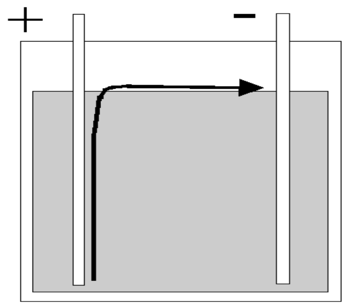
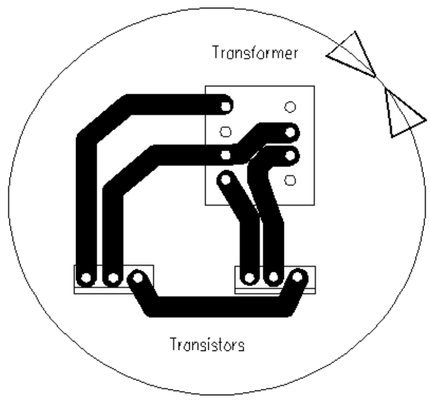


Figure 3.26 Skin effect in a lead-acid battery

The second type of EMC is influenced by the PCB is designed. A high ohm resistor together with a coil is a perfect antenna for transmitting and receiving signals. When the PCB is being designed, this should be taken into account. A different placing of components can have a tremendous effect on the EMC-properties. Also the shape of PCB tracks can result in a decrease of electrical efficiency. Placing the inverter in a metal box can also reduce the emission to the outer world. But now the shape of the tracks becomes even more important, as the following example shows.

At ECN we tried to design an inverter with an efficiency of about 80%. After some time the engineer succeeded in creating an efficiency of 80.4%. However, as soon as we placed the PCB back in the metal housing, its efficiency immediately dropped by about 6% (figure 3.27). The reason was probably, that the PCB tracks from transistor to transformer formed almost a circle. Together with the metal housing this circle formed a magnetic induction.

PCB:



PCB in Housing:

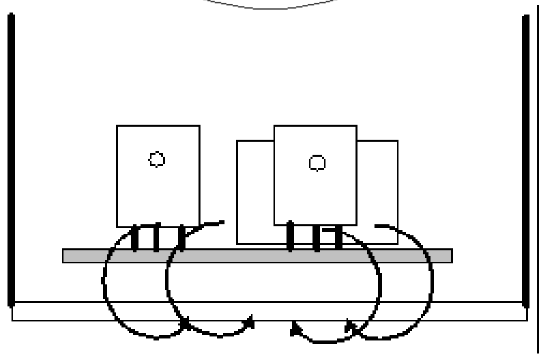


Figure 3.27 Decrease in efficiency caused by PCB design

The only solution is to place those tracks and components as close as possible to each other. The circuit itself is a circle, which can not be changed. As little room as possible should be left between the tracks from and to the transformer.

4. CHARGE REGULATORS

Lectures on day 2

Some of these sections make use of the report “A Pulse-Width Modulated, High Reliability Charge Controller for SHS” - 1997, by Kenneth Gerken of the Morningstar corporation.

4.1 Introduction

As PV cell costs continue to fall, the battery in a stand-alone PV-system becomes an increasingly large part of the system costs. Battery life now has the greatest impact on the economic viability of a small PV systems. The controller must manage a rapid, yet safe, recharge under a very diverse range of system conditions.

The charge controller in small stand-alone PV-systems is the primary driver of system reliability and battery life. An advanced controller will affect the system performance more than any other component, and an improved controller can potentially reduce 20-year system costs by 20% to 40% because the battery needs to be replaced less often.

There are a number of reasons why small PV controllers have not advanced with the rest of the PV industry, which include the following:

- Controllers only account for about 5% of the total PV system cost;
- Most small controllers are only simple on/off charge regulators, and most of these are hand-built in local, protected markets.
- PV applications and marketing channels are both highly diverse, so it is difficult to supply ‘standard’ small PV controllers.
- Since Solar Home Systems need to be as cheap as possible, more advanced and expensive charge regulators are not common. The challenge in the near future is to make sophisticated charge regulators viable for low cost applications.

In a PV-system a battery is the largest life cycle cost component in the system. There is a growing awareness of the impact that battery life has on the system economics. Also, as the volume of PV system installations increases, there is a growing emphasis on the high costs to repair or replace failed controllers. As PV becomes more widely accepted and the technology matures, users will demand more value and quality in their PV systems.

4.2 Objectives and scope of work

The purpose of this chapter is to give ideas and sugestions for developing and improving Solar Home System battery charge regulators. This chapter is divided in two parts:

1) design ideas and
2) modifications for improving the overall performance

Ad 1) On the question of design ideas, we first take a look at the power management. There are several ways to build up the power system. This has nothing to do (yet) with the methode of controlling the system. Pros and cons will be discussed.

Controlling the system is a secondary consideration. A basic controller can be created, using only a limited amount of electronics: the standard ON/OFF controller.

Several features can be added to the system: PWM, over-current protection, state off charge indication to the user, etc.. These can be added with analogue electronics.

Eventually if the amount of added analogue electronics becomes larger, a digital controller can be considered. With these digital controllers the number of components can be reduced, but the total costs are very much dependant on the type of features required.

Ad 2) Because many small companies work without much outside contacts, existing charge regulators are often based on old technology. Old circuit diagrams were never reviewed. If these diagrams were examined now, a great deal could be improved. In this section of this manual possible points of modification are discussed, in order to improve efficiency and reliability.

4.3 Design ideas

4.3.1 Series / shunt type regulators: Pros and cons

At the present time, many of the cheaper controllers use a shunt type of switching principle: the module short-circuited after the HVD set point is reached (Figure 4.1). The main reason for this is the general thought that less electronics is required than in the series type controllers. Furthermore, it is known that there is less voltage drop (=energy loss) while charging. The Schottkey is necessary, because it prevents the battery from discharging via the module during the night.

The disadvantage of this type of system is, that after HVD the module is in short circuit. The FET has to handle the full module current Isc. This causes much more heating inside the controller compared with the series-design. The cooling of the FET should be carefully evaluated. This is the main reason for an early breakdown of these types of controllers. If this type of system is used for PWM (Pulse Width Modulation, explained later on in section 4.3.4.), very high voltage and current peaks can occur in the system, which may damage some components.

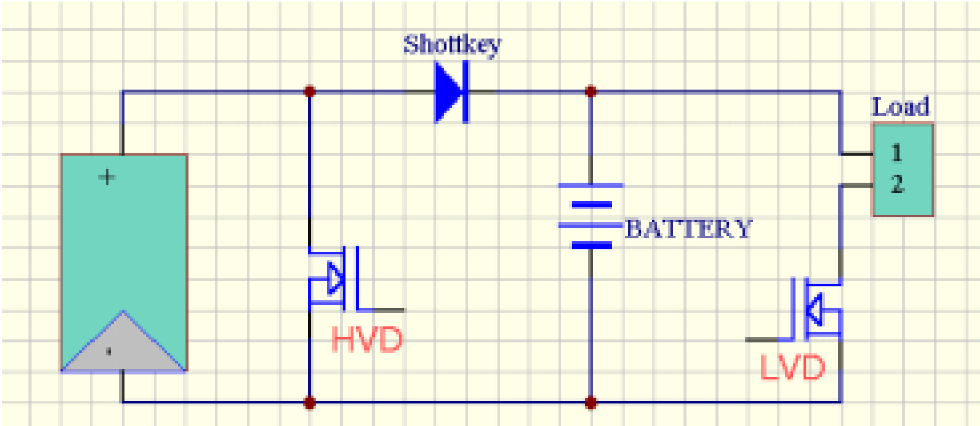


Figure 4.1 : System set-up in a shunt-type controller

A further problem is, that the World Bank states that the controller must have protection against reverse polarity connection for both the module and the battery. As can be seen in figure 4.1., this shunt design will definitely have problems with this requirement. A power FET always contains an internal freewalk diode, in the opposite direction of the switched current. If the module is connected with reverse polarity, it will immediately be in short circuit via this freewalk diode. In itself this is not a major problem, if the cooling is sufficient. But if the battery is connected with reverse polarity the FET and/or the Schottkey will burn immediately. There are two solutions to avoid this problem and meet the World Bank standards:

- add a fuse in the module line, just before or just after the Schottkey diode.
- add an extra Schottkey diode just before or just after the HVD-FET (in the FET shunt itself), in the direction of the module short circuit current.

The second solution adds an extra critical component to the controller, but the user / technician does not need to make repairs after making a wrong connection. The extra energy loss in this Schottkey is of no importance, Because this will be lost anyway. It is used when the module is disconnected. The first solution is the simplest one.

Some controllers however, are designed according to the series-principle (see figure 4.2):

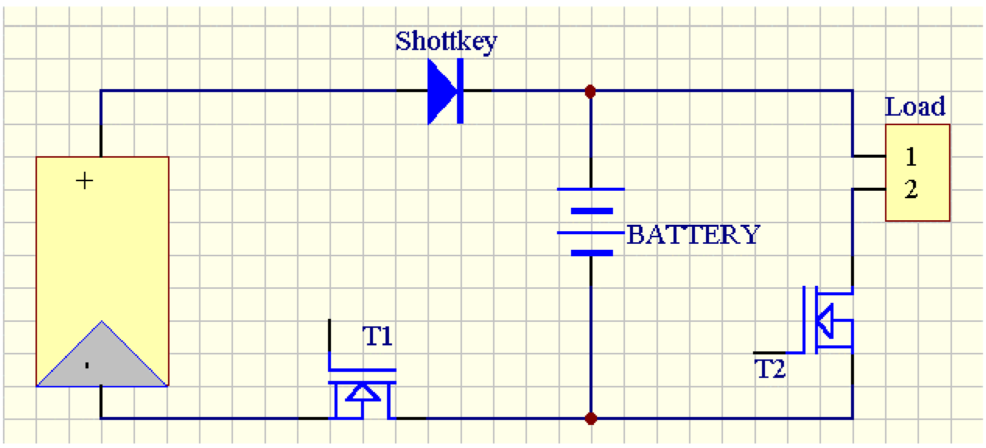


Figure 4.2: Serial, negative switching SHS controller, basic diagram.

Because of the internal freewalk diode in T1, the Schottkey is still necessary.

The main disadvantage is the extra energy loss in T1. You will therefore need to look over the data-sheets very carefully. The lower the internal conducting resistance, the lower the energy loss. This internal resistance differs considerably for different types and different manufacturers, and can vary between 0.3Ω and 0.01Ω.

Section 4.4.5 explains the effects of lightning induced transient pulses in both types of systems. The HVD mosfet is less stressed with a series design than with a shunt design.

Other advantages of a series configuration for solar home systems:

- the series FET has less voltage applied across it during all phases of operation when compared to the shunt FET (section 4.4.5.)
- less switching noise in a series design
- greater precision in battery charging, while the charging current is actively manageable
- pulse charging is possible

4.3.2 Positive switching / negative switching

It is known that there are significant problems in the field with Solar Home Systems. The customers and technicians sometimes do not know what the cause is, and sometimes they even bypass the controller. This only makes the problem worse, but they know of no other solutions. One of the causes is possibly unknowingly bypassing some controller functions. In most of the present controllers the 0V wires are switched. If the module and some load, or the controller itself is connected to 'ground', the FETs are in short circuit via this ground. Even the connections seem correct. Furthermore: in electronics it is *common* that the black wire can be connected to ground. For customers who work with electronics and do not know the system sufficiently, this can be very confusing and cause problems. For example: the module is connected to the roof. If this roof is made from metal, there is a chance the module is connected to earth (depending on the module construction). If the controller is now connected to ground, the HVD function does not work. The battery is overcharged every time. The damage can be so great, that the battery becomes defective within 3 months. Nevertheless, the system looks to be correct.

A second example: Almost all DC/AC-converters have a ground-connection at the box. When this device is connected to ground with also somehow the charge regulator, the total LVD function is bypassed. Even when all AC-load is switched off, the quiescent current of the converter is flowing from the battery all night.

The only way to make sure that this will not happen is to switch the +-wires from all devices. Now everything can be connected to ground without problems. Additionally, if the solar systems are interfered with, no damage is caused to the controller or the battery. Switching the positive lines contributes to the general reliability of the complete system.

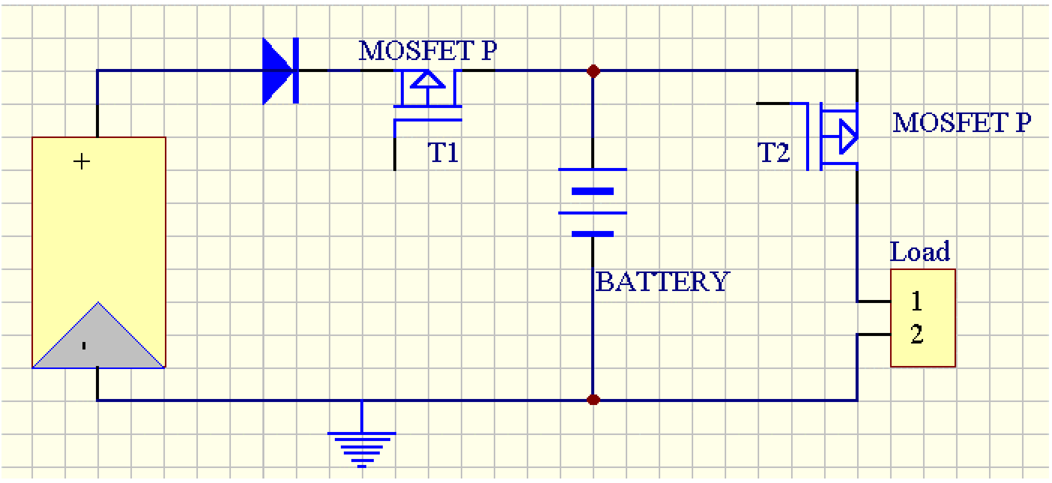


Figure 4.4 Positive switching SHS controller, basic diagram.

The aim of the charge regulator design is to keep the electronics as simple as possible.

When using the usual N-channel FETs for this positive switching, driving this component will be very difficult. The drain should be connected to the module, the source to the +battery. The gate voltage should be at least 10V higher than the source voltage in order to activate it (Vgs = +10V). Just before HVD the battery voltage will be about 14.5V. The gate voltage should be at least 24.5V. When the charging has to be stopped, Vgs must be 0V. This means that the gate voltage should be the same as the battery voltage! It will be relatively difficult to create this precision steering.

Another solution is the use of P-channel FETs with Vgs = -10V (Figure 4.4.). In the positive switching system the source voltage is the module voltage. If we make the gate voltage 0V (Vgs = -12V during charging) the FET is conducting. If the gate voltage is 22V (Voc from the module) Vgs=0V so it is blocking. In this type of FETs the drain-source current (Ids) is also negative, what is required. The steering principle for the 'load-FET' can stay the same, only exactly opposite towards a negative switching system. For the set-points we use Schmitt-triggers again. An example of the use of Schmitt-triggers can be found in the total diagram in appendix 1. The reference voltage can be taken towards the ground. **Only the module FET needs a voltage doubler, for a disconnected module terminal has the open circuit voltage (Voc). For a proper blocking of the module current, the gate voltage should be Voc=22V. at least.**

Another solution is to add a DC/DC-converter to the module, so that the input voltage is fixed. Then the voltage doubler for the module FET is not needed.

Only when a current limiter is desired, the reference voltage should be taken towards the positive system voltage. More details on this can be found in section 4.4.4. The only difference is, that now all reference voltages are trimmed towards the +Voltage of the battery (Appendix 1). The reference is taken from the positive voltage instead of the ground (Z3). Because of this, the Schmitt-triggers have to function in the opposite way.

The total costs for positive switching controllers will be a little higher, but the configuration is safer and more logical to the people in the field.

4.3.3 HVD-Switching in the battery-line: pros and cons

In serial disconnect controllers the module is totally disconnected from the system at HVD. In parallel controllers the module is brought in short circuit at HVD. The practical problem for the latter type of regulators is, that the cooling for the FETs has to be high. In almost all cheap parallel-controllers the cooling is insufficient. This results in an early breakdown, because the FET has to handle the full module short circuit current Isc.

A different approach is to place the module FET in the battery shunt. A FET always contains a reverse polarity freewalk diode that can handle the same current as the FET itself, which is really convenient for this method of use.

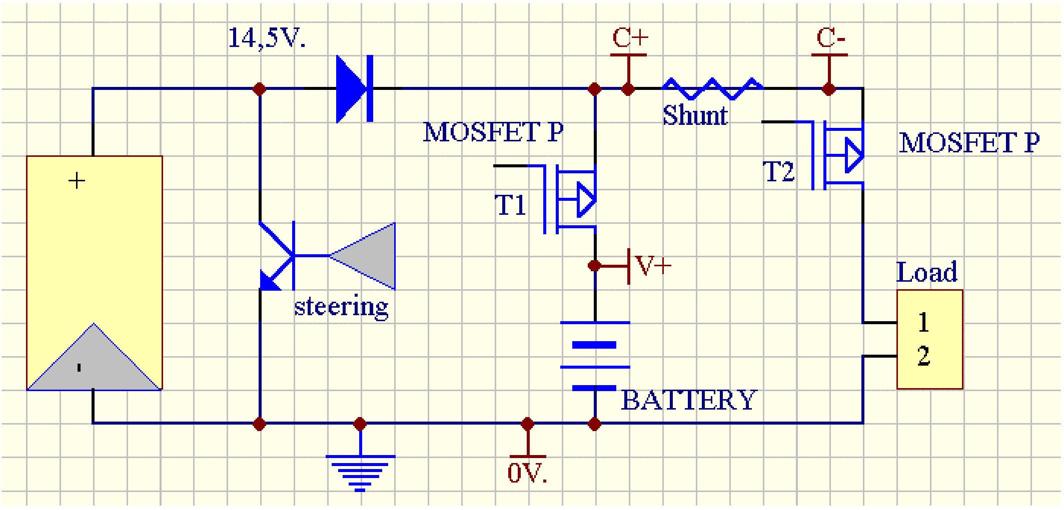


Figure 4.5 Basic system for a positive switching Charge regulator.

The line to the battery is now divided into two directions. One way is always open via the diode. That is the direction for the load current. The FET can block the opposite direction. That is the direction for the charging current. In the case of HVD the FET only blocks this charging current. The battery is still able to give energy to the load. The big difference is, that the module is also still connected to the load. The load can use the possible energy from the module after HVD. For example: the battery is fully charged at 16.00 in the afternoon. The module is totally disconnected in the standard type of systems. If the user wants to listen to the radio at that time, the energy is taken from the battery until the battery level is below the module reconnect voltage. In the mean time the possible energy from the module is wasted in stead of used by that radio. If the FET is placed in the battery shunt the energy from the module can be directly used by the load, while the battery is still in HVD-mode. This has a double benefit. The possible module energy is not wasted, and energy is not taken from the battery, but saved for later use.
There are some disadvantages to this system:

- Energy stored in the battery has to flow through an extra diode before if gets to the user. (It has to run twice through the HVD-FET.). This means a slight loss of energy. This principle is only worthwhile if it is certain that the extra module energy retrieved after HVD will be used during the day. The system should be slightly over dimensioned (the module should be a little too large for the battery), and energy should be used during daylight. Systems in hospitals and schools could be based on this principle.
- When the ‘Charging-FET’ is blocked and charging current is stopped, the systems’ voltage would increase to the open circuit voltage of the module. This could be damaging for user

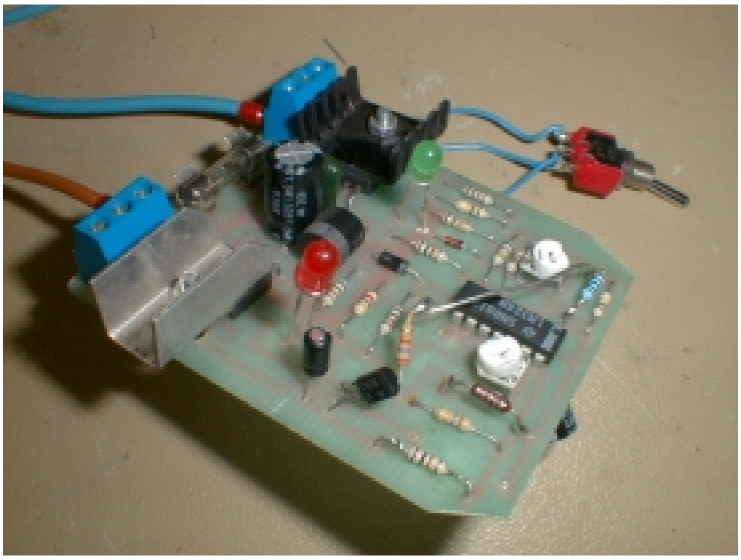
devices connected. At the moment that the module is disconnected from the battery, a dump load should be connected to the system. This load has to be sufficiently large that the module voltage (and so the system voltage) does not exceed the maximum of 14.5V. According to our experience, Solar home systems of 150Wp need to loose about 3 to 4A in order to reach that voltage. The more loads are connected, the less current needs to flow to the dump load. This is about the same effect as a DC/DC-converter on the module itself, only such a DC/DC-converter can also *increase* the module voltage when it is lower than a certain output voltage. This device can also be used for this purpose.

An important technical aspect of this system: the electronics' power supply should be taken straight from the battery (Figure 4.5: 'V+'), because if you take the V+ from the total system ('C+'), there will be problems with the set-points:

Without current flow the set-points can be correctly trimmed. As soon as load is connected, there is a certain voltage drop over T1. This FET is now in the battery shunt. The system voltage measured by the controller is not the actual battery voltage. This depends on the current flowing through T1. If the battery is discharged at a rate of 3A the LVD set-point will drop to 10,1V. If charging at a rate of 3A the HVD set-point will increase to 16,2V.

4.3.4 On-Off- charging / PWM- charging

A 1992 PV battery survey noted that most PV controllers use a simple 'on/off' regulation mechanism (see photograph below). These controllers interrupt charging early in the charging cycle when a voltage regulation set-point is reached. To prevent instability, a hysteresis of 1 volt is generally used to reconnect the PV array, and this causes the battery voltage to drift down for a period of time before charging can start again. A report by the US research institute Sandia in 1994, found that batteries charged with the standard set-points will typically average between 55% and 60% state-of-charge (SOC) for a period of years. This causes stratification of the electrolyte and sulfation of the battery plates, thus increasing internal resistance, which further reduces charge efficiency.



Low cost 'on/off'-charge regulator

There is another algorithm to charge batteries: Pulse Width Modulation. During the last part of the charging cycle, where the on-off regulation should disconnect the module, the module is switched on and off with a certain frequency. Normally the voltage of the car battery goes down when the charging device is removed. By connecting and disconnecting the charging device with a duty-cycle, charging can go on much longer and the battery will reach a much higher state of charge. The duty cycle is the ratio between the 'on'-time and the 'off'-time of the charging current.

The potential benefits of charging a PV battery with controlled pulses include:

- compact lead sulphate can be broken up to improve battery capacity and charge acceptance
- charge efficiency can be improved and effects of ageing can be reduced
- operating life of the battery can be increased
- higher voltage pulses can punch through resistive coatings between the grid and active material on the plates
- the opportunity for a gas bubbles to form can be reduced
- down pulses further improve charge efficiency and reduce gassing

One especially exciting area of pulsing may offer particular benefits for PV systems. The PWM charge pulses may be short enough in duration to reduce the ability for gas bubbles to form. Further, it seems that a down pulse makes gassing even less likely since it apparently helps to break up the precursor to a gas bubble.

With a PWM algorithm, there are constant voltage charging values for lead-acid batteries which provide for rapid charging and a safe regulation of voltage. These are between 2.35 v/cell and 2.40 v/cell where the battery enters the gassing stage. For effective PV charging, the proper constant voltage must be held within a narrow window.

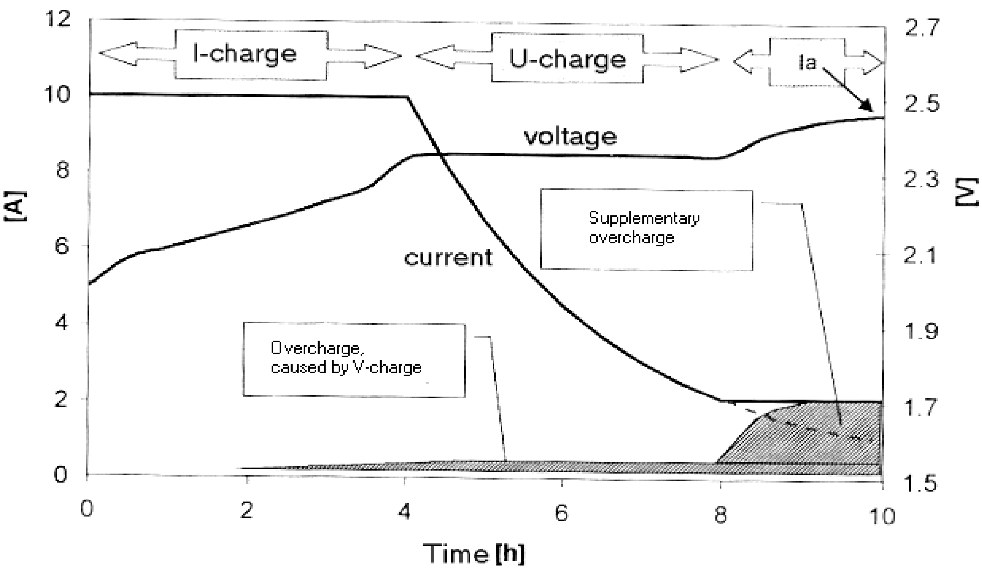


Figure 4.6 Charging current acceptance from a lead acid battery. The voltage and current are shown on the vertical axis.

Figure 4.6 above shows the charge characteristic from a battery, when it is charged with a constant maximum current (from point 0 until 4 hours). When each cell reaches its maximum voltage of about 2.30V, the battery is being charged with a constant voltage. When the battery reaches its (almost) full state of charge, its accepted charging current will decrease to 0.2*Imax. This is the point at which maintenance charging starts (after 8 hours) and from then on the cells will be more overcharged.

These three stages of charging are called trickle charging. Where normal on/off regulators stop the whole charging process when the cell voltage is at its maximum (figure 4.6, 4 hours), trickle charging continues until the battery reaches its full state of charge (100%). When the cell-voltage just reaches it maximum value, the actual state of charge is only 60%. Hence with standard on/off-regulators the battery is never fully charged.

Figure 4.6 applies charging without PWM, but with drip loading in order to compensate for the self-discharge of the battery (maintenance charging). This does not take place in PWM charge regulators, because these controllers keep on charging until the duty cycle is zero. If there is self-discharge of the battery, a low duty cycle will again be initiated.

A standard on/off controller will stop charging at the HVD point. Without regulation the voltage will increase further. Due to this high voltage, gassing will occur, which can be dangerous under normal operating conditions. The voltage should be kept at 14.5 V: the ‘constant voltage algorithm’.

In order to charge the battery with maximum power, the average charging current should be decreased in such a way, that the system voltage stays at 14.5 V. This can be done with Pulse Width Modulation (PWM). The duty cycle has to be decreased, so that the average charging current pulls up the (open circuit) battery voltage to 14.5 V. While going to its ‘fully charge’ state, the (virtual) open circuit voltage will steadily reach 14.5 V. The difference of the HVD voltage and this open circuit voltage will steer the duty cycle (D.C.):

D.C. (%) = A*(HVD - Vbat (open))

This system should only be triggered at the moment that the battery, charged with D.C.=100%, approaches the HVD set-point.

From another point of view: when the charging voltage approaches 14.5V, the duty cycle has to be reduced. Figure 4.7 shows a possible block diagram for this type of PWM.

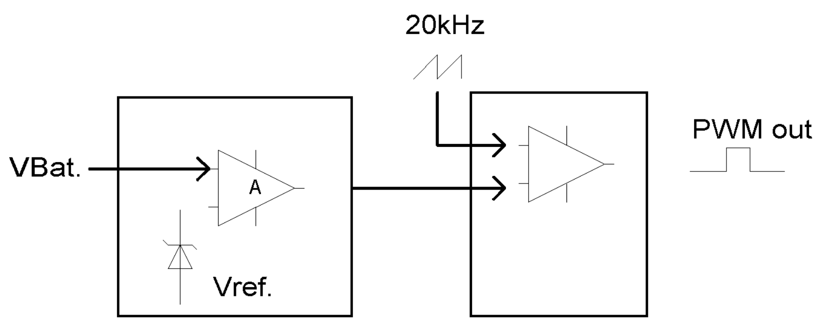


Figure 4.7 Block diagram for constant voltage PWM algorithm

Morningstar's 'Prostar' was one of the first digital controllers to use this algorithm. The SunSaver-6 (see photograph below) is Prostars analogue brother. Both controllers are known as very efficient charge controllers, which extend the lifetime of the battery to the maximum. Unfortunately, less attention is paid to the safety precautions. (section 4.4.2, 4.4.4.).

Morningstars Sun-Saver charge regulator

The PWM controllers tested by ECN had basic frequencies ranging from 8Hz to 300Hz. The basic frequency should not be too high (not higher than 10kHz for prevention of the battery 'Skin'-effect, section 3.2.11), and constant over the charging range. Battery charging is a chemical process, which needs some time to start up. Before it is effective. If the charging current is switched off too soon, the actual energy storage has not yet started. The Morningstar Sunsaver uses 300Hz. The Sunsaver-algorithm turned out to be one of the most effective charge regulators, but it is not clear if it will be even more effective if the basic frequency is brought down further.

The result for The Morningstar Sunsaver-6 is shown in figure 4.8 below.

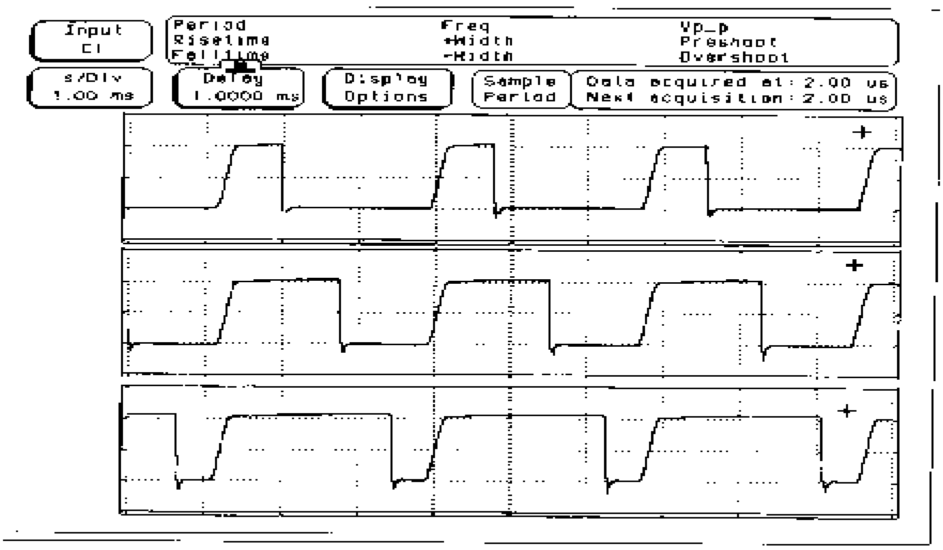


Figure 4.8 SunSaver-6 PWM results: voltage against time for three different duty cycles

As can be seen, on the cut off of the current, a down pulse is generated which can approach -10A. Morningstar has investigated the effect of this down pulse. They state that this characteristic has a beneficial impact on the lifetime of the battery:

"In a PV system, a down pulse (i.e. reverse or discharge pulse) will occur when charging in PWM with a load. Research indicates that there are good prospects for using pulse charging combined with down pulses to extend battery life in PV systems. The following benefits may be possible:

- the down pulse creates new activation sites, so with subsequent pulses there are more sites available to accept charge
- pulses break through corrosion to prevent capacity loss
- sulfation is reduced
- heating and gassing of the battery during charging are reduced "

4.3.5 Boost Charging

All batteries require some overcharging now and then to reverse sulfation and reach full charge. The finishing charge is of critical importance to enable recharging to 100% SOC, while avoiding excessive heating and gassing.

Gassing in the battery, which is initiated by some small overcharge, cleans the internal plates. If this is established once every three weeks, this will reduce the deterioration of the battery and increase its deep recharge capacity.

The *normal* HVD set-point should be:

- 14.1 V (2.35 V/cell) for Sealed batteries at 25 ^{o}C
- 14.4 V (2.40 V/cell) for Flooded batteries at 25 ^{o}C

While boost charging these HVD set-points are temporarily increased to about 14.5V and 14.8V at 25 ^{o}C.

Some Solar Home System controllers incorporate the function of this 'Boost Charging'. The biggest difference between the various types of controllers is the timing when the boost charging takes place.

- Some controllers only boost charge the battery when it is connected for the first time, so just after installation. This costs more in electronic circuits, while the boost function is not very effective.
- Others will apply the boost charge function every time the LVD set-point is reached. In this way, the extra sulfation is removed shortly after it is formed on the plates. However, overcharging too frequently is also not recommended, because this will shorten the battery lifetime. The result of this type of boost timing is that the deterioration of the battery will accelerate with age. Charging and discharging of a battery will be faster with the age of the battery. The older the battery, the more often it will reach its LVD and HVD points. The boost charging voltage will be applied to the system more frequently. In the end it is only boost charging.

- Hence the optimum situation is to make boost charge possible about once every three weeks.

The purpose of boost charging is generally accepted, but is difficult to achieve with only a few analogue components. Therefore, boost charging after every time the system fals below its LVD set-point seems a practical alternative. It can be achieved with relatively few extra components.

The next section explains more about microprocessor controlled charge regulators. Since the total control of the system is provided by software in these types of controllers the ‘ideal boost’ is easy to add, as extra components are required.

This function of boost charging is essential in areas where the temperature is subject to large fluctuations. This boost charging is especially recommended in tropical countries and desert areas. However, there are doubts whether this high voltage is often achieved. In practice people use as much as possible, so probably the average state of charge will be relatively low and HVD will not often be reached. In addition to this the module quality is important. As described in section 2.4.1, in the Indian desert the temperature effect on the (locally manufactured) module is such, that it cannot generate 14.5V in the first place. So the HVD point is never reached. In these cases a question can be asked about the usefulness of the boost charge function. However, in ideal systems it is essential in certain geographical areas.

4.3.6 Special function IC’s

As solar energy becomes increasingly accessible for commercial use, more and more special regulator ICs are being developed. Various manufacturers design special function circuits to include many types of different features. Only a minimum of external components needs to be added in order to create a fully functioning charge regulator. The main set-up of such a charge regulator is illustrated in figure 4.9 / 4.10.

The disadvantage of these chips is that its features are fixed. It is not possible to add extra features or change characteristics. In most special function ICs for this purpose a low-voltage disconnect for the load is not included, so external load-control is necessary. Many professional manufacturers therefore prefer a custom made design rather than a standard solution which is also available for private technicians.

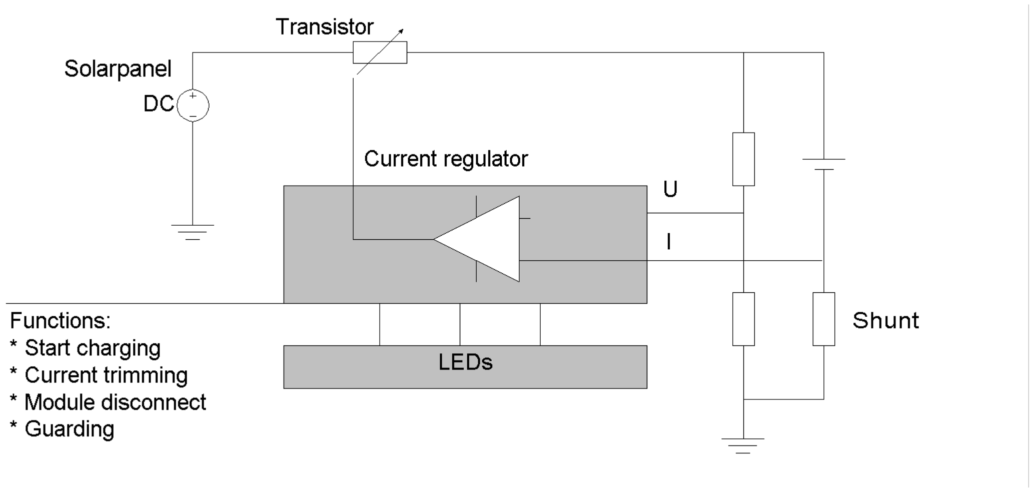


Figure 4.9 Charge regulator based on integrated circuit

Other special function ICs have a PWM algorithm integrated. The switching FET should be added externally.

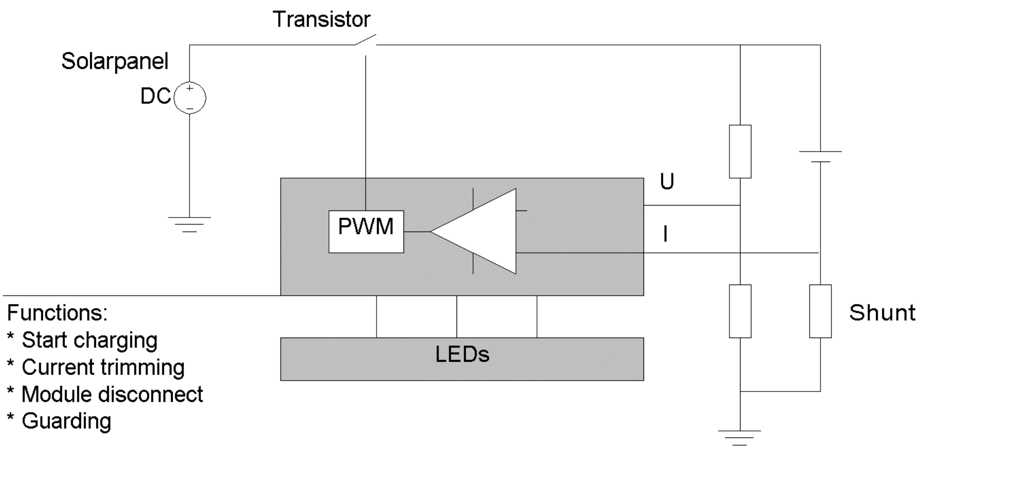


Figure 4.10 Charge regulator based in integrated circuit

Table 4.1 describes some of the special function ICs available for charge regulators for lead-acid batteries.

Table 4.1 Special function Ics for charge regulators

Type	**Manufacturer**	**Disconnect algorithm**	**Special**
MAX 1648	Maxim		Analogue programmable
UC3909	Unitrode	None	PWM Temperature compensation
BQ2031	Benchmarq	U,Δ2V,I,t	PWM Temperature compensation
ICS1735	Galaxy Power	None	Pulse charge and conditioning
MTA11200	Microchip	-ΔU, dT/dt	External current steering uP-based SOC-monitoring
LT1510	Linear Tech.	None: external switch needed	PWM at 200kHz internal end-amp
BST9303	Becker System		LED status-display
BST9705	BST	Softcool algorithm	LCD capacity display

4.3.7 Digital or analogue set-up

Most charge regulators are designed with analogue electronics. This is partially an inheritance from the slow development of charge regulators in the past. In recent years much has changed to improve microprocessor technology. One-chip processors with internal A/D-conversion, EEPROM, PWM-generator, etc. have become available in all price-classes. This brings the number of components of the system down to a minimum. Only a few analogue inputs and steering outputs are then required in the regulating part of the circuit.

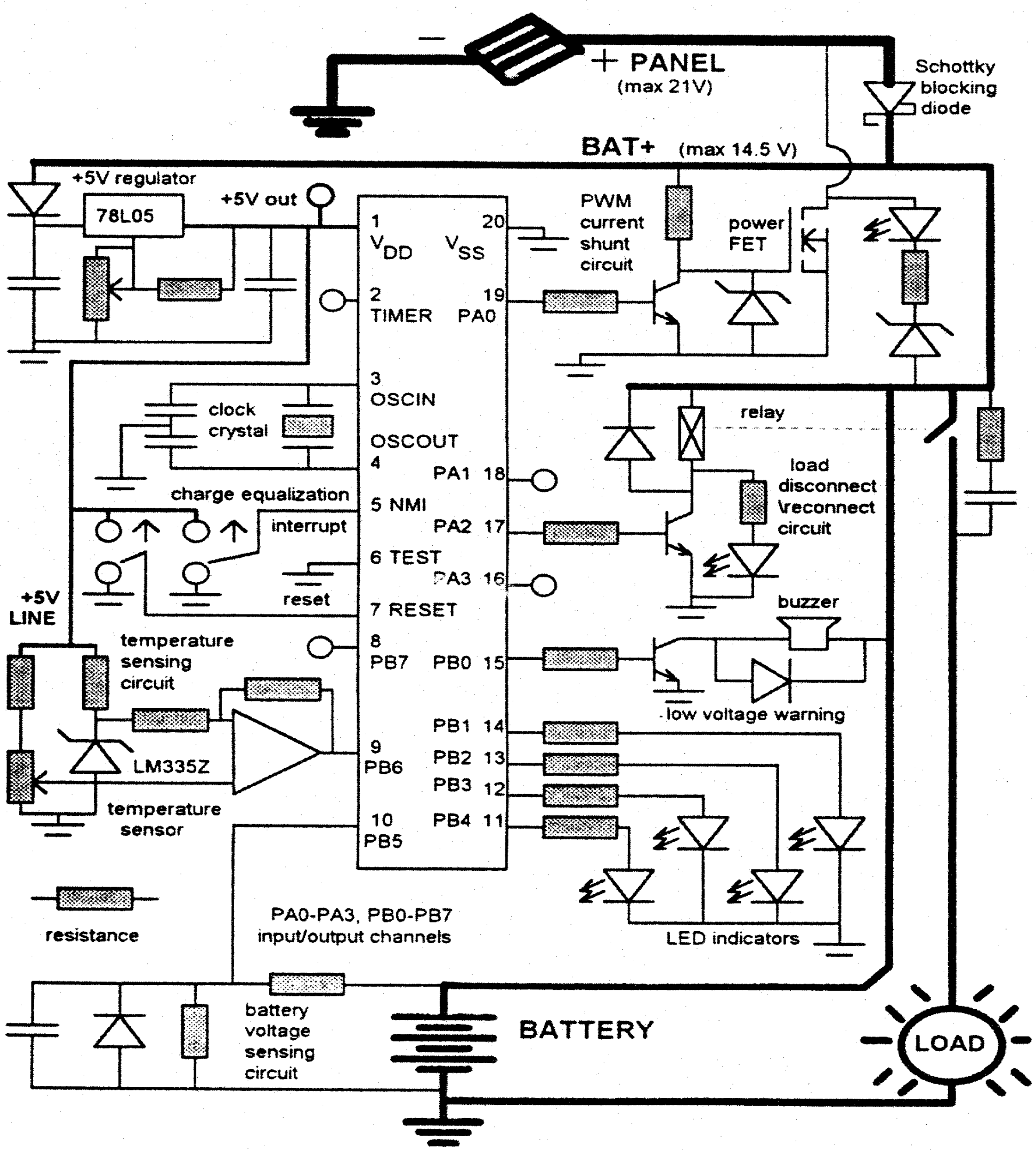


Figure 4.11: Example of a microprocessor based charge regulator

The great advantage of a processor-controlled regulator is, that extra functions (or changes in actual functions) can be edited very easily. As can be seen in figure 4.11., the hardware for the temperature compensation and for the state of charge indication is reduced to a minimum.

A standard processor without extra functions (ADC, PWM) costs about US$ 2.50 for large production quantities. Since there is no ADC on board, this should be created externally. This ADC can be made in various ways, so attention has to be paid to the ADC design and the total price.

The price-difference with ‘special-function’ processors is still considerable (US$ 10.-), which makes external ADC worthwhile in a digital design. If such a charge regulator is being developed, the engineer should be aware of and consider other options. For every extra option onboard a microprocessor the price rises disproportionately. In the near future, single-chip microprocessors will become cheaper and more advanced.

Sometimes it is worthwhile to base the design on analogue electronics. This depends on which functions are required by the manufacturer. For the design of a standard on/off controller, analogue electronics is relatively much cheaper and effective. Even a PWM-function as described in section 4.3.4 can be realised with analogue components, and still compete in price with a digital version.

Since the improvement of charge regulators in a recent development, there will be a growing interest in processor based charge regulators. Microprocessors have a high calculation capability. This can be used to determine the state of charge with complicated algorithms. Research is being performed on the use of ampere-hour balances for calculating the exact state of charge of the battery.

4.4 Modification points

4.4.1 Quiescent current reduction

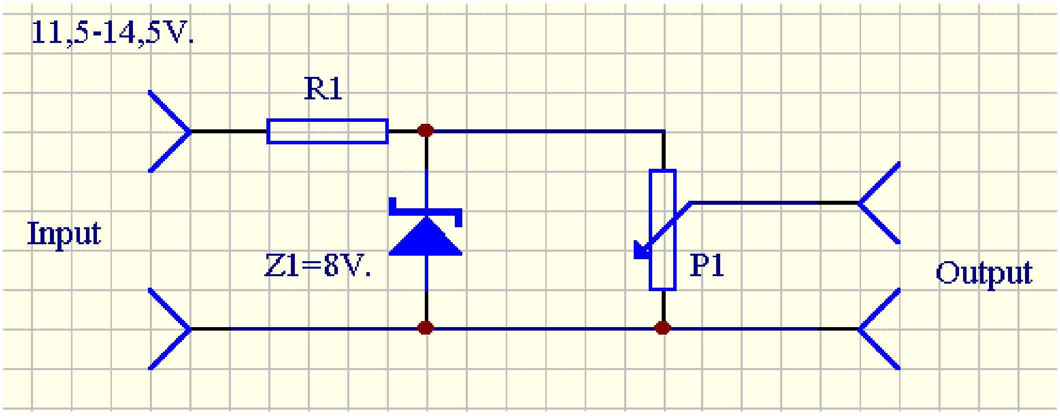
In small PV-systems energy saving is a major issue, not only for the user, but also for the designer of the system. All unnecessary energy loss is a waste. ‘Efficiency’ is a keyword in the balance of system (BOS) components. Therefore, charge regulator designs should be continuously revised with a view to decreasing its quiescent current. Every mA reduction can be spent by the user, and makes the system more efficient. A simple calculation:

> If the system has a life expectation of 15 years it means that 1 mA quiescent current reduction adds: 0.001A * 12V * 15year * 365days * 24hour = 1577 Watt-hour to the total energy used. Every 5 mA quiescent current reduction results in about 1% more energy available to the user.

The majority of the quiescent current of a basic analogue on/off charge regulator goes to following parts:

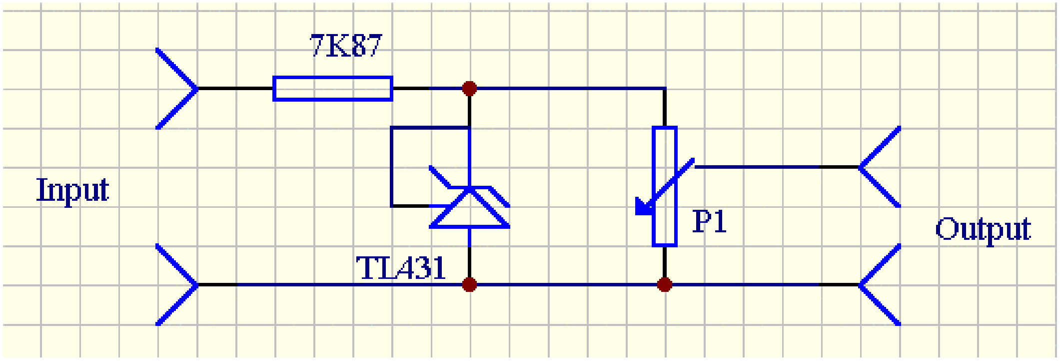
1. reference voltage
2. power supply for the main electronics
3. superfluous components (section 4.4.3.)
4. voltage dividers

Ad.1: Comparing the battery voltage with a certain reference voltage normally creates set-point control. For creating this reference voltage, different components can be used. In most diagrams you can find a zener diode:



There is a disadvantage with using a zener diode in this way. Its quiescent current is almost always more than it needs to be. The voltage drop over R1 at a battery voltage of 14.5V will be 6.5V. With a reference current of, lets say, 0.03 mA, its value must be 22kΩ for optimum efficiency. At a battery voltage of 11,5V this should be 12KΩ. Which resistor should you choose? If you chose 22kΩ and the battery voltage is 11.5 V, the reference is ‘too weak’ and its voltage will drop! If you take 12kΩ and the battery voltage is 14.5V, a lot of energy loss will take place in R1 and Z1 in order to maintain the 8V reference. In the end the resistor ends up in between these two values.

A better solution is a reference IC. Do not use the standard 78X.. series for analogue references, because these are made for power supplies. Their quiescent current is too large for this purpose. The ICs TL431 or REF25Z are more suitable, because they are designed to maintain the reference voltage at a fixed level with minimum energy loss over the input voltage range. This IC can be a simple substitute for the zener diode:



This reference voltage will now be about 2.50V, so the other resistors need to be recalculated. However, the quiescent current is significantly decreased!!

Ad2: If digital electronics is used in the controller, a +5V power supply is needed, which means selecting the 78X05 power supply. There is however a good substitute for this IC: the LP2905. This is also a 5V power supply, but has less quiescent current. In general: try to avoid digital flip-flops or CMOS ports in an analogue-based controller. Do not use them, unless it is absolutely necessary. Of course, in microprocessor based controllers a 5V powersupply is a must.

Ad 4: All electronics make use of voltage dividers, for example, for measuring the battery voltage, trimming the reference voltages on the opamp, etc. If the charge regulator is dimensioned properly, try to double all resistor values, or at least some voltage dividers.

4.4.2 Safety precautions: reverse polarity protection

Most stand alone SHS are being used in remote areas, far away from a service centre. This requires all BOS-components to be very reliable. If a component becomes defective, it will take a great deal of effort, time and money to replace it. Users should also have a reason to be confident in the relatively expensive solar system. In the 'ideal' solar system no BOS-components can be damaged by any mistake, not even on accidental wrong connection. The main issue is to design the controller in such a way that it can withstand all conceivable mistreatment.

This safety starts with the system itself. If all cables to and from the controller are equipped with plugs that can only connect in one way, the possibility of reverse polarity is already excluded. Of course both installation technicians in the field and users, sometimes try to 'repair' their system themselves if it does not meet their requirements. That is why a controller for remote areas should be able to withstand all types of wrong connections.

The World Bank has drafted a number of requirements regarding polarity protection for both the battery and the module.

The electronics itself has to be protected. If a 78X.. power supply is used (digital controllers), the downstream electronics is protected, but sensitive components before the IC are not. The safest way is to use a simple diode such as 1N4148 from the systems' voltage to the electronics. In such way the electronics are unharmed.

Protection of the system against wrong polarisation:

This implies that the module and battery as well as the user should not be endangered by a wrong connection. If the module is wrongly connected, no current should be able to flow, otherwise sparks may occur while disconnecting. Figure 4.1 and figure 4.2 show the series and the shunt controller. The Schottkey diode in the shunt controller prevents the battery from discharging during the night. If the modules' polarisation is opposite, it is in full short circuit via the internal safety diode. The Schottkey cannot be placed before the FET (from the module point of view) because then at HVD the battery is brought into short circuit. If the battery is connected wrongly, it will be in short circuit via the two diodes. As previously stated, an extra diode or a fuse should be added into this system.

In the series-controller the Schottkey and the safety diode are directed exactly opposite. This system is safe for wrong polarisation of both the module and the battery. No current will flow, so it is safe for both the system and the user.

If the module is switched in the battery line (Figure 4.5) and the battery is connected with reverse polarity, the electronics is not powered up. The module FET will not be activated, and the safety diode is blocking. Whenever the module is connected wrongly, the Schottkey is blocking. The system is now safe.

4.4.3 Reduction of the number of components

With all the foregoing in mind, the charge controller as illustrated in appendix 2 (the upper picture) is analysed. It has a quiescent current of 25mA. Four comparators monitor each one set-point (HVD, HVR, LVD, and LVR). Every comparator has its own reference voltage. In pairs they steer a digital flip flop, which steer the FETs directly. This means that the steering voltage is only 5V. When this voltage is compared with the FET graphics (appendix 3) 'Transfer characteristics' , it shows that there is only about 17A Ids (pulsed) possible. At Vgs=10V the FET allows 110 A Ids pulsed. This means that the internal resistance at Vgs=5V as well as the energy loss in the FET are much higher than necessary.

Under the 'actual diagram' a possible alternative set-up is pictured. By using the opamps in a different way, most of the original electronics becomes redundant. The FET gs-voltage can now be 12V, so less heat is produced inside the controller because of the lower internal resistance. The quiescent current is now 5mA, which can be further reduced by replacing the zener diode by the reference IC, as shown in appendix 4.

Some controllers use more components than needed for their proper function. Always look for a possibility to combine functions. All references can be led down from one voltage reference.

4.4.4 Safety precautions: overload protection / short circuit protection

For the standard analogue on/off controllers concerned, the cheapest way is switching with two opamps in 'Schmitt-trigger' form. When using a 14 pin opamp housing such as LM324, there are two opamps used. With some inexpensive components added, an electronic fuse can be created to maximise the benefit from the available electronics. Appendix 5 shows a possible design of such an overload protection for a negative switching system. The current limit is adjustable from 0 to 10 amperes. If that limit is exceeded, the load will be shut down for one minute. That gives the electronics time to cool down, and the user some time to switch off some load. After this interval of minute the load is automatically switched on again. This method of overload protection is much more user friendly than a fuse. In remote areas it is sometimes difficult to buy fuses. With this type of protection, replacement of fuses is no longer necessary. How does the electronics work?

Measure the load current: if this is more than 80% of the maximum FET current Ids (this is a generally accepted standard; components should be capable of handling 125% of the maximum system ratings) or a certain other current value, the load should be disconnected. The electronic fuse in a negative switching system gives a good indication on how it functions. Then combine the E-fuse with a positive switching system. The principle remains the same, but the electronics functions exactly in the opposite way.

In the negative switching system:

Appendix 5 shows the diagram of the current limiter in the negative switching system. A shunt resistor is added to measure the load-current. The voltage accross this resistor is compared with a reference (IC1d). Normally the output from pin 14 as well as from pin 8 is 0V. C4 is discharged via T3 / R16, because T1 is conducting. If the load-current exceeds the limit, the output from pin 14 is 12V. Pin 8 is still at 0V, so pin 10 will be about 4V. The IC1c output from pin 14 will flip to 12V. Now T1as well as the load, is shut off via T4. No load current is flowing, so the output from pin 14 will now return to 0V now. Pin 10 will have 8V. The output from pin 8 also slowly charges the capacitor C4. It takes about a minute to reach 8V. This is the point when the output from pin 8 turns to 0V again. T1 becomes active, C4 is being discharged, and the system is back in its original state. If there is still an overcurrent, this principle repeats itself. However, the time interval will only be the discharge time of the capacitor C4, which is a fraction of a second. C3 and R15 must be added, in order to suppress on/off switching peaks. Otherwise the current limiter will respond every time a lamp is switched on. When the controller is connected for the first time, then the input from pin 12 responds slower than the reference pin 13 due to C3. Therefore the current limiter stays off after the first installation. The system is now balanced correctly.

In the positive switching system:

The normal positive switching controller works in exactly the same way, but referenced from the systems +-voltage. Appendix 6 shows the same diagram, but now for the positive switching regulator.

As indicated earlier, the electronics power supply should be taken straight from the battery. This is in order to keep the set-points stable. The reference voltage for the Schmitt-triggers will have to come from the +-bat. Appendix 6 shows a prototype diagram with the current-limiter. IC1d is the over-current detector, IC1c is a 1-minute timer. As soon as FET T1 is conducting again, the timing capacitor C4 has to be discharged. So this discharging is depending on the opposite state of the T1 output. If T1 is high, there should be no discharging, if T1 is low C4 has to be discharged. That is why we need two transistors here. In the negative-switching regulator we also used two transistors, so there are no components added.

4.4.5 Lightning surge induction

Most standards in the world state that lightning surge induction protection inside the charge regulator is needed. This is an essential point, because in some areas lightning occurs very frequently. Temporary overvoltages can have many causes, but in solar home systems lightning induction in the cable from the module is the most important.

Because there is less system-level protection, small PV-systems are more sensitive for voltage transients than larger PV-systems, where the price of the electronics is of less importance.

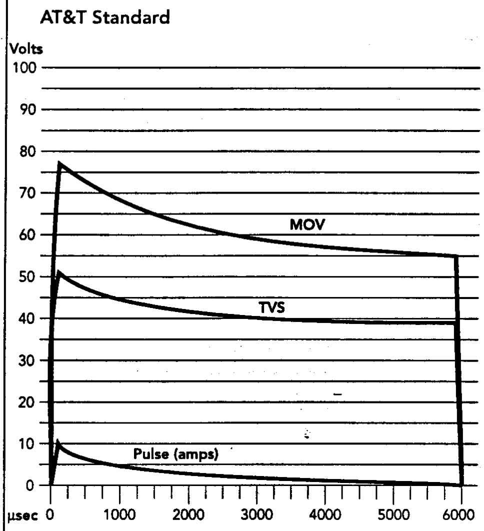
The Morningstar Corporation formed a project around the effects of transient induction on small Solar Systems. They also investigated two types of protection devices: the MOV (metal-oxide varistor) and the TVS (transient voltage surpressor). Most small charge regulators use MOV’s, because TVSs are much more expensive. Both devices work by shorting the transient current, so they are connected in parallel with the module input.

There is a standard IEC pulse-definition used in order to simulate the transients representative of field condition. Figure 4.12

Figure 4.12 shows that the resulting voltage peak, caused by the transient current pulse, is higher with a MOV than with a TVS. These TVSs are mainly used in more expensive systems.

Transients have many causes other than lightning, and it is very difficult to design for a 15-year operating life when considering how much exposure a PV controller will have to transient pulses.

In a reliable analysis of an 8/20 pulse (the IEC801-5 level 1 standard for transient pulse testing) the varistor (MOV) would clamp the FET to 60 V. However, in a series design the resulting voltage across the FET would only be 47.6 V (Figure 4.13)

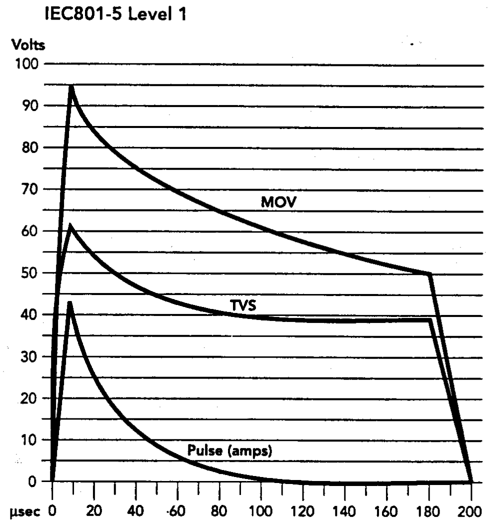


Amps:	Rise	Fall	Peak	Repetitions
	10µsec	1000 µsec	10 amps	3000

Device:	Size	Part Number	DC Standoff Voltage	Max Clamping Voltage
TVS	1500W	Z 6039	31.6V	51
MOV	15mm	Z 15L390	31.0V	77

Time µsec	Pulse amps	TVS volts	MOV volts
0	0.00	0	0
2	2.00	40	60
4	4.00	43	66
6	6.00	46	72
8	8.00	49	74
10	10.00	51	77
200	8.50	50	76
600	6.50	48	73
1000	5.00	45	68
2000	2.50	41	62
3000	1.20	40	58
4000	0.60	39	57
5000	0.30	39	56
5900	0.10	39	55
6000	0.00	0	0

10/1000 Pulse



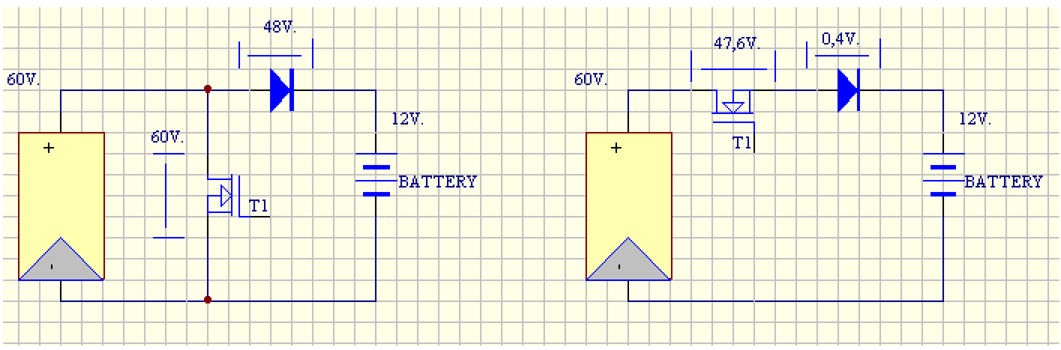
Amps:	Rise	Fall	Peak	Repetitions
	8µsec	20 µsec	43 amps	5

Device:	Size	Part Number	DC Standoff Voltage	Max Clamping Voltage
TVS	1500W	Z 6039	31.6V	61
MOV	15mm	Z 15L390	31.0V	95

Time µsec	Pulse amps	TVS volts	MOV volts
0	0.00	0	0
2	10.00	46	77
4	21.00	52	80
6	32.00	57	86
8	43.00	61	95
12	34.00	59	87
16	27.00	57	82
20	22.00	53	80
40	11.00	47	78
60	5.50	43	70
80	2.50	40	63
100	1.20	39	60
120	0.60	39	57
140	0.30	39	54
160	0.15	39	52
180	0.07	39	50
200	0.00	0	0

8/20 Pulse

Figure 4.12 The IEC standard test-pulse for testing the Transient Overvoltage Protection



shunt regulator

series regulator

Figure 4.13 Transient voltage drops in solar home systems

As can be seen in figure 4.13, the series FET is between 60V transient and the 12V battery. This produces only 47.6V, which is well within the (normally) 60V rating.

The shunt design, however, places the FET directly between the positive leg and the ground. This produces the full 60V across the FET. In addition, the Schottkey becomes a second component to be stressed by the transient. This means that the series design offers a greater degree of protection.

During a transient pulse, the system-voltage stays at 12V, and the working point of the module drops back to zero in its power curve, which means that the module itself and the cables absorb the energy of the transient pulse. This can never fully be the case because the response time of the Schottkey diode always is almost as long as the 8/20 pulse. While the Schottkey is not fully open the situation of figure 4.13 will be the case. As soon as the diode is conducting, a slight dip in the working point of the module may occur, but it will not drop totally to zero. By that time the transient pulse is already over.

4.4.6 Oscillation problems without battery

A solar home system is designed for charging batteries. This battery is the ‘buffer’ for the systems’ voltage. A module without any load gives an open circuit voltage of 21 V. When the battery is disconnected, the systems voltage will immediately go to its HVD action, and shuts off the module. The electronics does not have any power left to keep the FET conducting, because there is no battery connected. When the module is in open circuit again, the voltage rises to 21V. This repeats itself. The LEDs blink on the front panel. The speed of this oscillation depends on the total capacity of the controller. In itself this sequence is harmless. However, it can cause damage in shunt regulators.

In shunt-controllers, the module is brought into short circuit at HVD. This means that there is always a maximum current flowing at HVD. Because of the oscillation, the average gs-voltage is about 5V. This means that the internal resistance of the FET is higher, and it has to handle the full module current. This causes more heat than in a series-regulator, where no current flows during oscillation.

The oscillation problem is relevant for all tested controllers and both controller types. That is typical for a working controller: disconnect the module at HVD. In shunt-controllers this can cause overheating.

‘Oscillating’ is perhaps not the right word. It is more a ‘logical on/off switching’ which is normal for any controller. It is recommended that the oscillation speed is not too high, for example 0,5Hz. This can be achieved by adding a 100uF/50V capacitor straight over the power supply of the comparators. In this way the blinking disconnect-LEDs indicate that there is no battery connected. For shunt-type regulators, the biggest problem is the cooling for the FET. The stronger hardware has sufficient cooling. A small peace of metal surface is not enough. In cheaper types of charge regulators this is the main reason for failure.

A more expensive solution is a ‘battery detection circuit’. If there is no battery connected, this circuit changes the Module-FET into a ‘dummy-load’ and disables a module disconnect. The comparator LM324 can handle Vcc=30V max, so that is no problem. This may be an ideal solution, but is costly in terms of extra components.

Better solutions are possible, but ‘We need to keep the design as inexpensive as possible’.

5. DC/AC CONVERSION

Large parts of this section of the manual notes were taken from the following publications: "Power Electronic Converters: DC-AC conversion", Guy Séguier Francis Labrique, Springer Verlag Berlin Heidelberg 1993, "Power Electronics: Converters, applications and design". Ned Mohan, Tore M. Undeland, William P. Robbins, John Wiley & Sons, 1995, “200Watt Power inverter” by ELV GmbH, 1993, “A DC/AC Isolated Battery Inverter using HIP4082”, application note Intersil, May 1996

5.1 Scope of work

The use of DC/AC inverters in small solar home systems is rapidly growing. In some areas it is unthinkable to sell a SHS without one. Hence it is a worthwhile exercise to consider the advantages and disadvantages of using these devices and also for what purpose they can be used.

Firstly, the most common applications of DC/AC-conversion can be listed:

- **Television**. Many people in rural areas built up savings in order to buy a colour television, sometimes with a satellite dish.
- **Lighting**. In some rural areas standard 230Vac fluorescent lamps are used instead of the special 12Vdc fluorescent lamps because they are widely available.
- **Fan**. In tropical areas a fan is often desired. This is a luxury item, which usually bought only after a television set is obtained. This device consumes a lot of energy, so it can only be incorporated in larger systems.
- **Refrigerators.** The demand for refrigerators is growing, especially in areas where people have already worked with solar energy for some time. These devices also consume a lot of energy, so the dimensioning of the system has to kept in mind.

In many smaller devices (satellite receivers, radios, etc.) 230Vac input is transformed to 12Vdc. In televisions this dc-voltage is transformed afterwards to the high kVoltage to drive the electron beam to the screen. The use of a DC/AC-converter for these devices is theoretically rather useless. DC/AC-converters have an efficiency of approximately 85%. Downwards AC/DC-transformation always has energy-losses also, in the order of 90% efficiency. In total it means an unnecessary energy-loss of 100%-(90%*85%) = 23%.

At the present time there are many types of television sets, satellite receivers and fluorescent lamps operating at 12Vdc. Solar energy is relatively expensive, so devices that are used in combination with a SHS should be selected carefully.

On the other hand there are some important, very practical reasons why these 12Vdc devices can not (yet) be sold with a SHS in rural areas:

- Televisions: 12Vdc televisions are sometimes far too expensive, compared to the 230Vac versions. The market for 12Vdc televisions, *outside the Solar Home System market*, is very limited and mainly based on Western prices. It will take quite some time before the price for this type of device *and its availability* can compete with the 230Vac television.
- Lighting: The availability of fluorescent lamps in some areas is a problem. Travelling Nomads in the North of China prefer a DC/AC-converter with a 230Vac fluorescent lamp because 230Vac FL are available in almost every village they cross. From the aspect of energy consumption, this is not a good solution. In areas where 12Vdc FLI *can* be delivered, however, it must be recommended to use these. Because it is an important factor in getting the most out of a solar system.

- Refrigerators: At the present research is being performed in the field of 12Vdc refrigerators. It will still take some time for efficient units to enter the market at a reasonable price. Present 12Vdc designs use far too much energy for a small SHS. A 230Vac refrigerator with DC/AC converter is mainly used, but only in combination with larger PV-systems.

It is important to note that DC/AC-converters in combination with very small systems (up to 50 or 100 Wp) are not of much use (though sometimes practical). 230Vac equipment in general uses more energy than necessary. When the energy-loss in the DC/AC-converter is added to this unnecessary energy loss, there is not much left for other applications.

Taking this into consideration we can derive suitable basic structures for a DC/AC-converter that can be used in Solar Home Systems. This is very important, because there are many different ways to design them. Many different topologies are described in the literature. Much information is available about power electronics and DC/AC-conversion for a wide series of applications, unlike information about charge regulators and FLI for 12Vdc. On the other hand, the technology of DC/AC-converters is much more complicated than the technology of the other BOS-components.

In this project we focus on Solar Home Systems, so the examples we have chosen in this chapter are especially suitable for that purpose. Normally, the power limitation for DC/AC-converters in SHS is set on 500Watt maximum (see photograph below). Set-ups within this power range are different from set-ups designed for 5000Watts, for example. The basic power-topology may be the same, but safety-regulations and technical specifications are different. For example: it is hard to find a full sinewave inverter under 500Watt for a price that is suitable for customers in developing countries.

The common DC/AC-converters sold within the SHS market not only have to be as cheap as possible but also very reliable. In this chapter we will focus on topologies, set-ups and examples that can be found within this scope.

250-500 Watt DC/AC inverter from LUE.

5.2 Types of Inverter: Principles

We present the principle of the various inverter types for a single-phase output, by using the four "switch" (K1, K2, K'1 and K'2) configuration shown in Fig. 5.1. The supply placed on the DC side is assumed to be perfect, as are the switches (zero forward voltage drop, zero leakage current, instantaneous commutation).

i and u are used to denote the input current and voltage, i' and u', the output current and voltage,

i_{K1}, and v_{K1},, the current and voltage across K1,

T, the operating cycle of the inverter and its output values.

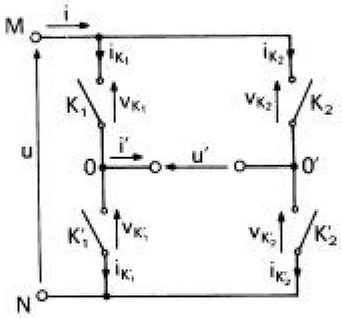


Figure 5.1 Basic diagram of a full bridge

The inverter is assumed to be in "full-wave" control I, i.e.:

- for $0 < t <$ T/2, K1 and K'2 are on and K2 and K'1 off.

u' = u $\quad i = i'$

i_{K1}, = i', $\quad v_{K1} = 0$

- for T/2 $< t < T$, K2 and K'1 are on, K1 and K'2 are off:

u' = -u, $\quad$ i = -i'

$i_{K1} = 0$, $\quad v_{K1} = u$.

If the variations in i_{K1} and v_{K1} are followed, the currents in the other switches and the voltages across them can be directly deduced:

$i_{K'2} = i_{K1}$, $\qquad v_{K'2} = v_{K1}$,

$i_{K2} = i_{K'1}$ $\qquad$ with $i_{K'1}(t + T/2) = i_{K1}(t)$

$v_{K2} = v_{K'1}$ $\qquad$ with $v_{K'1}(t + T/2) = v_{K1}(t)$.

Voltage-Source Inverters

A voltage-source inverter is an inverter *fed by a DC-voltage supply,* i.e. by a supply with negligible internal impedance; its voltage u is not affected by the variations in current i flowing through it. The DC supply *imposes the voltage* at the input - and thus at the output - of the inverter.

If u is constant and equal to U, whatever the value of *i*,

for $0 < t < T/2$, $u' = +U$

for $T/2 < t < T$, $u' = -U$.

The output *current i'* and thus the input current i depend on the load placed on the AC side. This load can be of any type, provided that it is not another voltage source (capacitor or AC emf) connected directly across the output terminals.

Switch K, must be made up of a fully controlled (turn-on and turn-off) switch TC_1. It must connect terminal 0 to terminal M at instant t = 0 and separate these two terminals at instant t = T/2, whatever the AC load may be. For i_{K1} to flow via K_1, if the latter is momentarily negative, an antiparallel- connected diode D_1 has to he added to TC_1.

The basic diagram of the autonomous voltage inverter can thus be proposed, as in Fig. 5.2.

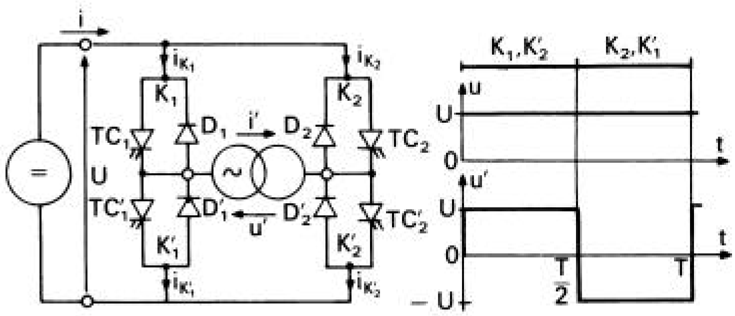


Figure 5.2

To show the influence of the load on the operation required of the semiconductor devices, we will consider the *load to be perfect,* i.e. that it takes a sinusoidal current. The cases of the inductive and capacitance loads are considered.

Figure 5.3 provides the waveforms of *u', i', i,* v_{K1}, and i_{K1} when the *load is inductive* (current i' lags voltage u' fundamental component).

For switch K_1,

from t = 0 to $t = t_1$, diode D_1 enables i', which is then negative, to flow;

from $t = t_1$ to t = T/2, TC_1 enables i' to flow;

at instant t = T/2, turning-off TC_1 leads to i' being transferred from TC_1 to D'_1.

Three remarks can be made about this operational mode:

-For t = t 1, TC_1 begins conducting when the current is zero; no overcurrent is produced in this device and no overvoltage across it.

-Controlled devices TC_1 and TC_2 turn-on drive signals must be applied throughout period [0, T/2] (and that of TC_2 and TC'_1 throughout period [T/2, T]), since the value of t_1 depends on the load. Although TC_1 is controlled at turn-on from t = 0, it remains non-conducting until $t = t_1$, when it starts to conduct.

-A short period of time must be left between the turn-on of TC'_1 (or TC'_2) and the turn-off of TC_1 (or TC_2) and vice-versa, in order to avoid short-circuiting the DC voltage supply. On account of the diodes, there is no risk of the AC current load circuit being opened.

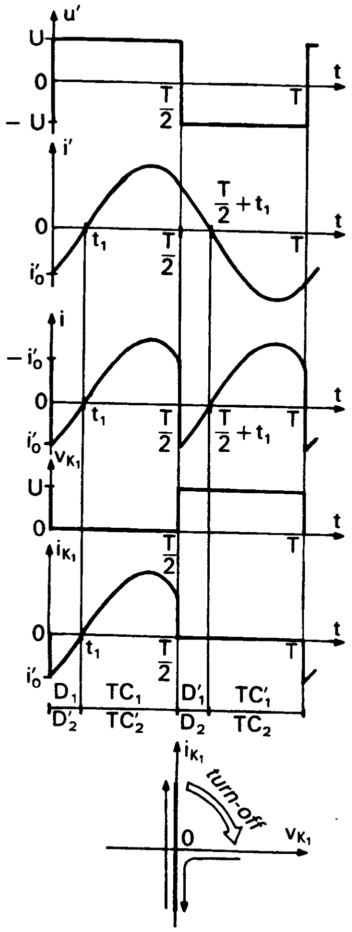


Fig. 5.3 inductive load waveforms

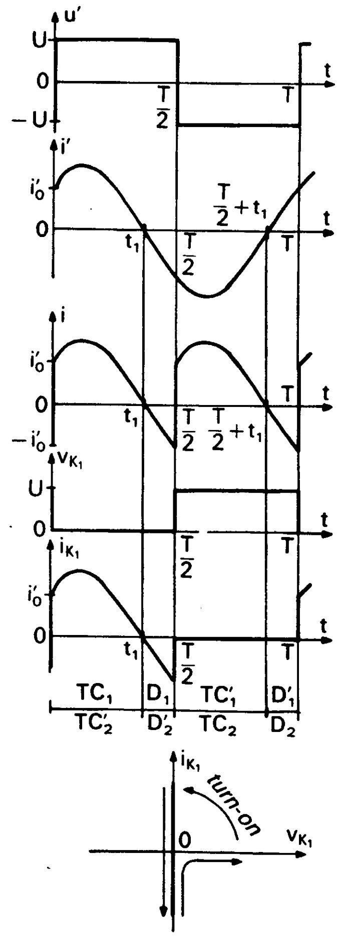


Fig 5.4 capacitive load waveforms

Figure 5.4 shows the waveforms of u', i', i, v_{K1}, and i_{K1} when the inverter feeds a capacitive load (current i' leads voltage u' fundamental component).

At instant t = 0, firing TC_1 means that current i' is transferred from diode D'_1 to TC_1; i' flows via the latter until t = t_1.

For t = t_1, current i' falls to zero and becomes negative, TC_1 turns off naturally and D_1 becomes conducting. D_1 conducts until instant t = T/2 when it turns off as a result of the turn-on of TC'_1.

As in the case of a flow across an inductive load, there is one commutation without overcurrent nor overvoltage. The other commutation is ensured by the turn-on of a controlled device. The ability for turn-off is not used in this case.

If the load can be either inductive or capacitive, the semiconductor switches must be controllable at both turn-on and turn-off.

5.3 DC-Voltage Supply Imperfection. Input Filter

In the case of inverters, the effects giving the most problems arise from *the inductance in the path where the current must be suddenly reversed:*

- DC supply inductance in a voltage-source inverter,
- AC load inductance in a current-source inverter.

Voltage spikes

In order to take into account the non zero value of inductance Ls of an imperfect voltage supply u, this supply is replaced by a perfect voltage source Us serially connected with this inductance Ls (Fig. 5.5a).

u = Us - Ls di/dt.

- *When feeding an inductive load*, during each commutation, current i must go from |i'o| to -|i'o| (see Fig. 5.3); this rapid decrease in i leads to a voltage peak Δu.

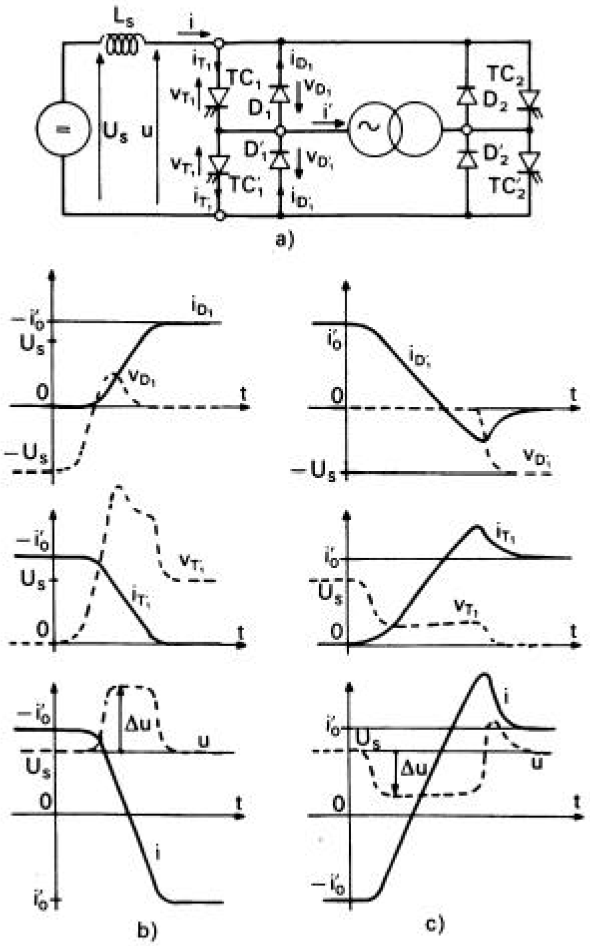


Fig. 5.5

Figure 5.5b indicates the waveforms

of v_{D1}, and i_{D1} (or of $v_{D'2}$ and $i_{D'2}$)

of $v_{T'1}$ and $i_{T'1}$ (or of v_{T2} and i_{T2}),

of i, equal to $i_{T'1}$ - i_{D1}, since $i_{T'1}$ equals i_{T2}

of u, equal to Us - Ls di/dt.

during the TC'_1-D_1 (or TC_2-D'_2) commutation, which occurs when, for t = 0, the turn-off of TC'_1 (and TC_2) is controlled.

Since v_{T1}, equals u + v_{D1}, the peak voltage Δu appears in the voltage across TC'_1; moreover, it can be increased by the forward voltage peak of diode D_1.

- *When feeding a capacitive load*, during each commutation, current i goes from -|i'o| to |i'o| (see Fig. 5.4); this leads to a voltage through Δu.

During the D'_1-TC_1 commutation controlled by the turn-on of TC_1. The reverse current peak in diode D'_1 can be found in the waveform of i_{T1} and, multiplied by 2, in that of i; it causes a voltage peak.

The input filter

In order to "correct" the DC-voltage supply, a capacitor C is connected in parallel across the inverter input (Fig. 5.6). This eliminates the sharp variations in u during commutations and reduces this voltage ripple during the period between commutations.

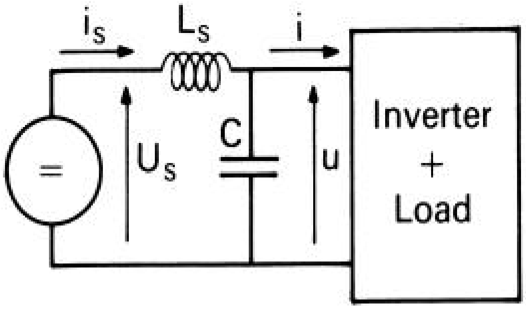


Fig. 5.6

Capacitor C, together with inductor Ls - whose value can be increased by adding an external inductance to the DC supply - form a *low-pass filter:*

• for voltages, in the supply to inverter direction,

• for currents, in the inverter to supply direction.

This enables a voltage u which is practically constant to be obtained at the inverter input, despite variations in i and even if voltage U_S is affected by a ripple (e.g. in the case of voltage supplied by a rectifier). It also enables current i, taken from the supply to be kept practically constant, while current i at the inverter input shows an important ripple. The cost of the filter decreases as the inverter operating frequency increases.

Battery Supply

A battery is a current-reversible DC voltage supply. It is thus *particularly well-adapted to supplying a voltage-source inverter* and can ensure the functional reversibility of the latter. For operation in the inverter mode, the mean value of i is positive and the battery discharges. During recovery, the mean value of i is negative and the battery charges up. A capacitor with a very low inductance is usually connected in parallel across the battery as the latter opposes very fast variations in the current.

An input filter can be used (see Fig. 5.6) to reduce the current ripple supplied or received by the battery.

Choosing the Inverter

For those inverters, intended to supply an AC load normally of *highly variable impedance* from a battery or a rectifier, *a voltage-source inverter* should be chosen. Such a choice is made for the following reasons:

1. The battery is particularly well adapted to supplying a voltage-source inverter. If necessary, an L-C filter can reduce the ripple of the current delivered by the battery. Such a filter enables a rectifier to supply a voltage-source inverter.

2 The operation of a voltage-source inverter is highly affected by the imperfections in the DC supply and hardly at all by those of the AC load. The reverse applies to the current-source inverter. For the uses being considered here, the DC supply can be selected but not the AC load.

3 Neither of the inverters directly gives a sinusoidal output voltage and an output filter must be used. In the case of the voltage-source inverter, the voltage to be filtered is imposed by the DC supply and the exact waveform to be filtered is known. With the current-source inverter, the voltage to be filtered depends on the load.

5.4 Single-phase Inverters Topologies

If only two "switches" are used obtain an AC voltage from a DC voltage, a mid-point is necessary, on either the AC output or DC input side.

This corresponds to:

• *the single-phase inverter with centre-tapped primary output transformer (often* called a push-pull inverter),

• *the single-phase inverter with a capacitive divider at the input* (often called a half-bridge inverter).

If the relative width of the square-wave pulses forming the output voltage half-cycles has to be varied, four "switches" are necessary. This is

• *the single-phase bridge* (or full bridge) *inverter.*

5.4.1 Single-Phase Inverter with Centre-Tapped Transformer

Figure 5.7 provides the configuration for the notations used for the single-phase voltage-source inverter with two "switches" and a centre-tapped primary transformer. The DC voltage supply has a terminal connected to the midpoint 0. Its other terminal is connected either to A by the conduction of TC_1 or D_1, or to B by the conduction of TC'_1 or D'_1.

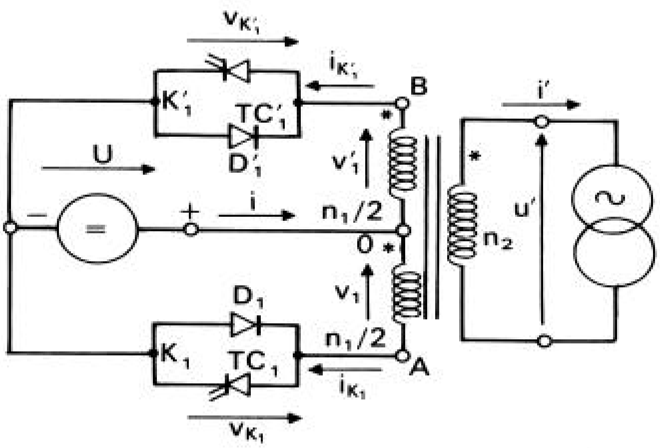


Fig. 5.7

Principle

If the transformer is assumed to be perfect, i.e. if the resistances and leakage inductances of its windings are ignored, the voltages are in relation to the number of turns:

$$v_1 = v'_1, \quad u' = v_1\, 2n_2 / n_1$$

If the magnetising m.m.f. is ignored, the primary and secondary currents are linked by the compensation in m.m.f.:

$$n_1 /2 \ (i_{K1}, - i_{K'1}) = n_2\, i'.$$

Commutations Required

-In the case of an inductive load, the commutations are forced. For $t = T/2$, when TC_1 is on, its conduction must be interrupted. This results in a voltage equal to + 2U across it; the current $(2n_2/n_1)i'$ is transferred from TC_1 to D_1.

-In the case of a capacitive load, the commutations are natural. For t = 0, when D'_1 is on, TC_1 is fired. This results in a voltage equal to - 2U across diode D'_1 which turns off. Current $(2n_2/n_1)i'$ is transferred from D'_1 to TC_1.

-In both cases, during each half-cycle, there is a free natural commutation between the semiconductor devices of a same switch - from D_1 to TC_1 or vice versa, or from D'_1 to TC'_1 or vice versa. This occurs when i' reverts its polarity.

Doubling the Voltage

The presence of a centre-tapped transformer means that the reverse voltage across the diodes and the forward voltage across the controlled devices are equal to twice the DC supply voltage:

$- v_{Dmax} = v_{Tmax} = 2U.$

5.4.2 Single-Phase Half-Bridge Inverter

The single-phase half-bridge inverter uses two "switches" which are bi-directional in current and a centre-tapped voltage supply. The mid-point can usually be obtained by using two capacitors C_1 and C'_1 with the same capacitance C. If the latter is sufficient, two voltages which are virtually constant and equal to U/2 are obtained. Figure 5.8 shows the configuration and the notations used.

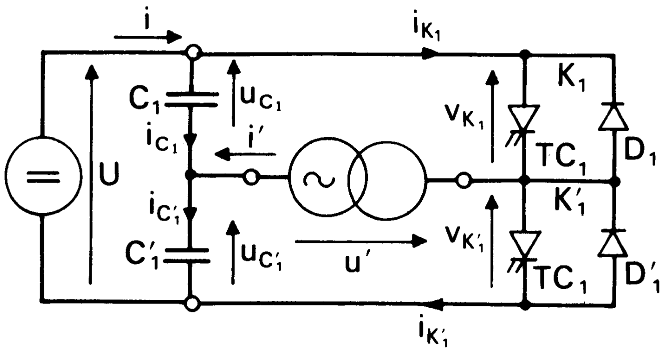


Fig 5.8 Single phase half-bridge configuration

Principle

As the sum of the voltages u_{C1}, + uc'_1 across the two capacitors is equal to U and as this voltage is assumed to be constant,

$u_{C1} + u_{C1'} = U$

gives

$C\, du_{C1}/dt, = - C\, du_{C1'}/dt$

$i_{C1} = - i_{C1'}$

The charging (or discharging) current of C_1 is equal to the discharging (or charging) current of C'_1.since

$i' = i_{C'1} - i_{C1},$

$i' = 2\, i_{C'1} = - 2\, i_{C1}$

Current i', which reaches the midpoint of the capacitive divider has twice the value of the current in one of the capacitors.

Commutations Required

The commutations required are the same as those for the push-pull inverter:

• forced commutations TC'_1-D_1 and TC_1-D'_1 when supplying an inductive load

• natural commutations D'_1-TC_1 and D_1-TC'_1 when supplying a capacitive load.

Doubling the Current

Owing to the capacitive divider, the current in the on-switch is twice the current in the DC supply.

I_{K1} or $i_{K'1} = 2i$

5.4.3 Single-Phase Full-Bridge Inverter

This is the full-bridge structure that was used in the previous chapter to present the various inverters. The configuration of the voltage-source inverter is shown again in Fig. 5.9.

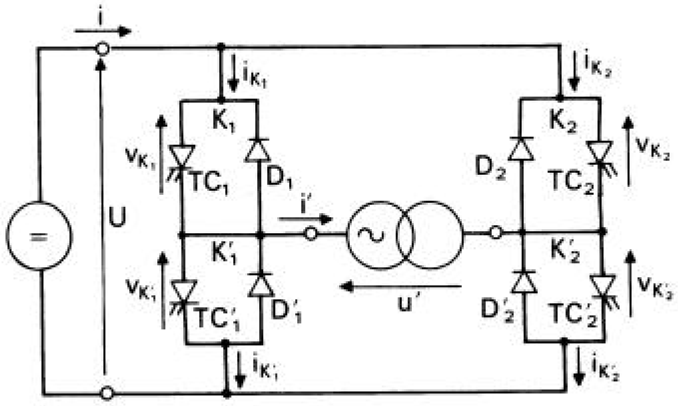


Fig. 5.9 Full-bridge structure

In the previous chapter, it was assumed that switches K, and K'2 were on during the first half-cycle of voltage u', making the latter equal to + U; switches K2 and K'1 were on during the other half-cycle, making u' equal to - U.

This control mode uses only a part of the possibilities of the full-bridge voltage-source inverter, because the latter enables not only the voltage frequency but also its value to be varied.

Principle

The control signals of K_1 and K'_1 must be complementary in order to avoid short-circuiting the voltage U supply and opening the current i' circuit. Similarly, for the other half-bridge, the controls of K_2 and K'_2 must be complementary. However the controls of both half-bridges need not be simultaneous; there may be a phase-shift between the two.

If ω is once again used to denote the angular frequency of the fundamental component voltage u', angle β can be used to characterise the phase shift between the control signals:

K_1 is on for $0 < \omega t < \pi$; K'_2 for $\beta < \omega t < \pi + \beta$

K'_1 is on for $\pi < \omega t < 2\pi$; K_2 for $\pi + \beta < \omega t < 2\pi + \beta$.

Commutations Required

The types of commutation required no longer depend solely on the nature of the AC load (positive or negative φ) but also on angle β.

Switch K_1 lets current i flow from $\omega t = 0$ to $\omega t = \pi$; this current reverses its polarity during this period. If i' is negative for $\omega t = 0$, D_1 conducts first of all and TC_1 then begins conducting when i' becomes positive; for $\omega t = \pi$, TC_1 must turn off current I' and perform a forced commutation.

If i' is positive for $\omega t = 0$, TC_1 takes the current which then goes via D_1 and is naturally transferred to TC'_1 when the latter is turned on at instant $t = \pi/\omega$.

5.5 Galvanic isolation

Often galvanic isolation is required between the high-voltage AC output and the DC input voltage. This can be achieved by either a low-frequency transformer at the ouput side of the inverter or by a high-frequency transformer before the inverting output bridge.

The main advantage of a high-frequency transformer is its reduced size and hence cost and weight compared to the rather bulky and heavy 50Hz transformers. However, the creation of a HF waveform and subsequent rectification requires additional circuitry and increases the complexity of the inverter. However the DC voltage often needs to be stepped up before the final inverting at the output bridge. In this case the HF transformer can provide both isolation and voltage boost. A low frequency transformer at the output of the inverter has the additional benefit of serving as a filter for disturbances at the load side which effectively protects the semi-conductor devices from the output bridge.

In the following section some issues are discussed regarding the design and construction of HF transformers for switched power supplies.

Isolation Transformer Representation

A high-frequency transformer is required to provide electrical isolation. Neglecting the losses in the transformer of Fig. 5.10b, an approximate equivalent circuit for a two- winding transformer is redrawn in Fig. 5.10c, where N_1: N_2 is the transformer winding turns ratio, L_m is the magnetising inductance referred to the primary side, and L_{t1} and L_{t2} are the leakage inductances. In the ideal transformer, $v_1/v_2 = N_1/N_2$ and $N_1 i_1 = N_2 i_2$.

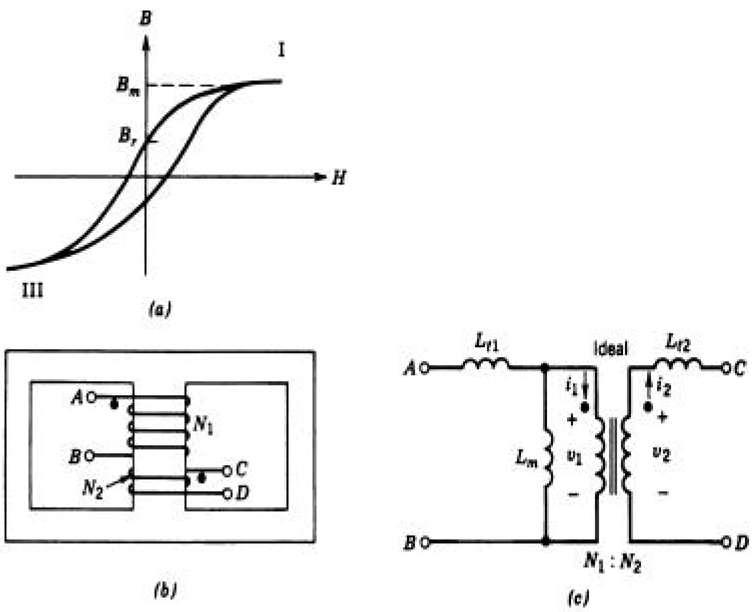


Fig. 5.10 Transformer representations: (a) typical B-H loop transformer core,
(b) two-winding transformer, (c) equivalent circuit

In a switch-mode converter, it is desirable to minimise the leakage inductances L_{t1} and L_{t2} by providing a tight magnetic coupling between the two windings. The energy associated with the leakage inductances has to be absorbed by the switching elements and their snubber circuits, thus clearly indicating a need to minimise the leakage inductances. Similarly, in a switch-mode dc-dc converter, it is desirable to make the magnetising inductance Lm in Fig. 5.10c as high as possible to minimise the magnetising current i. which flows through the switches and thus increases their current ratings.

Power transformers that are small in weight and size and have low power losses are the most desirable. The motivation for using high switching frequencies is to reduce the size of the power transformer and the filter components. If this benefit is to be realised, the power loss in the transformer core should remain low even at high frequencies.

Ferrite materials such as 3C8 are commonly used to build transformer cores. Similar to Fig. 5.10a, Fig. 5.11a shows a typical *B-H* loop for such a material, where the maximum flux density B_m beyond which the saturation occurs is in a range of 0.2-0.4 W_b/m^2 and the remnant flux density B_r is in a range of 0.1- 0.2 W_b/m^2. In Fig. 5.11b, the core loss per unit weight for several switching frequencies is plotted as a function of (ΔB_{max}) , where (ΔB_{max}) is the peak swing in the flux density around its average value during each cycle of the switching frequency f_s.

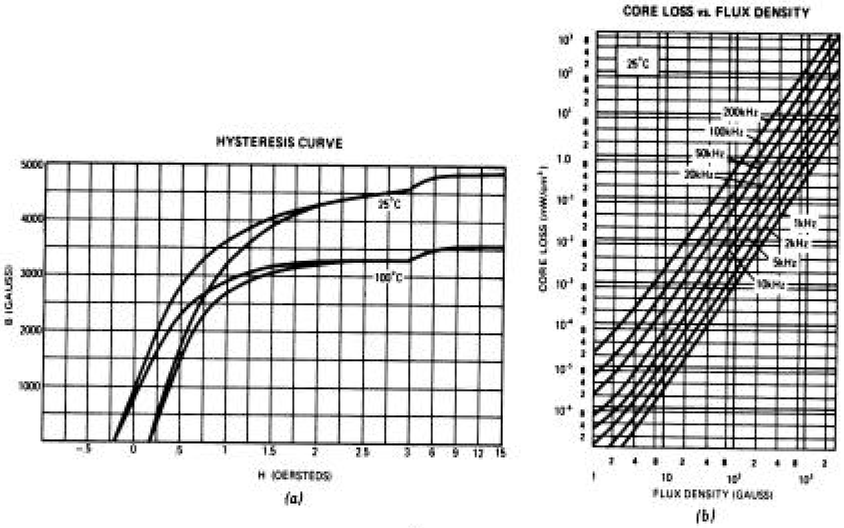


Fig. 5.11 3C8 Ferrite characteristics curve (a) B-H loop
(b) core losses curves (courtesy of ferroxcube division of Ampereex Electronic Corp.)

In the full-bridge converter with a bi-directional core excitation (example see Fig. 5.12)

$$(\Delta B_{max}) < B_m$$

Based on the foregoing discussion, the following conclusions can be reached regarding the desired core properties:

1. A large value of maximum flux density B_m allows (ΔB_{max}) to be large and results in a smaller core size.

2. At switching frequencies below 100 kHz, for example, (ΔB_{max}) is limited by B_m. Therefore, a higher switching frequency results in a smaller core area. However, at switching frequencies above 100 kHz, a smaller value of (ΔB_{max}) is chosen to limit the core losses.

3. In a converter topology where the core is excited in only one direction, (ΔB_{max}) is limited by B_m - B_r. Therefore, it is important to use a core with a low remnant flux density B_r in such a topology unless a complex-core resetting mechanism is used. In practice, a small air gap is introduced in the core that linearises the core characteristic and significantly lowers B_r.

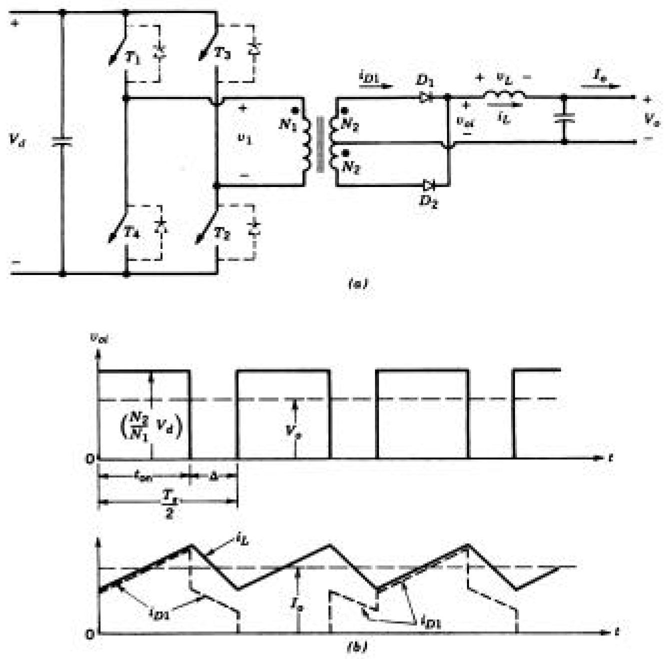


Fig. 5.12 HF transformer with bi-directional core excitation (using full-bridge topology)

In the converters with bi-directional core excitation topologies, the presence of an air gap prevents core saturation under start-up and transient conditions. But this does not prevent core saturation if there is a volt-second imbalance during the two half-cycles of operation (a volt-second imbalance implies that a dc voltage component is applied to the transformer core). In a practical implementation, there are several causes of such a volt-second imbalance, such as unequal conduction voltage drops and unequal switching times of the switches. The preferable way to avoid core saturation due to these practical limitations is to monitor switch currents, as is done in the current. Use of an appropriate control integrated circuit (IC) also eliminates saturation under start-up and transient conditions. The other way to prevent core saturation due to voltage imbalance is to use a blocking capacitor in series with the primary winding of the half-bridge and the full-bridge inverters. The blocking capacitor should he chosen appropriately so that it is not too large as to be ineffective under transient conditions and not too small to cause a large ac voltage drop across it under steady-state operating condition. In the push-pull converters, the current-mode control is used to prevent the switch currents from becoming unequal.

5.6 Comparison of the Required Components and Characteristics

The two types of two-switch voltage-source inverters (push-pull and half- bridge) require only two controlled semiconductor devices and two diodes, whereas the full-bridge inverter requires four controlled devices and four diodes. This advantage of the two former inverters is, however, reduced by the doubling of the voltage or the current:

-for a given voltage U of the DC supply,

for the push-pull inverter, $v_{Tmax} = - v_{Dmax} = 2$ U,

for the other two, $v_{Tmax} = -v_{Dmax} =$ U.

-for a given current i of the DC supply, the current in switch K_1, when it is on, has the value

$i_{K1} = |i|$ for the push-pull inverter,

$i_{K1} = 2\,|i|$ for the half-bridge inverter,

$i_{K1} = |i|$ (or $|i'|$) for the full-bridge inverter.

The two-switch inverters require half as many semiconductor devices but the voltage or current rating of the latter is doubled.

The push-pull inverter requires a transformer and the half-bridge inverter requires a capacitive divider, whereas the full-bridge inverter requires neither. These disadvantages of the two-switch inverters are less important that it would initially appear.

Virtually all inverters require a transformer to ensure galvanic isolation and to obtain the desired output voltage amplitude. In the case of the push-pull inverter, the transformer is inherent in the operating principle. However, the fact that it has a centre-tapped primary and that each half-primary is only used during one half-cycle means that the primary power rating is multiplied by square root (2).

Virtually all inverters require an input filter comprising an inductor and a capacitor. The capacitors of the half-bridge inverter can also act as a filter-capacitor. Instead of one capacitor of capacitance C submitted to voltage U, tow capacitors (each with a capacitance 2C and submitted to a voltage U/2) are used.

The main difference is that the full-bridge inverter (more expensive overall) enables the output voltage to be varied and thus, more precisely to be regulated. The full -bridge inverter requires the conduction of two series connected switches; the corresponding voltage drop and losses lead to a highly reduced efficiency when the voltage supply U has a relatively low value.

Single-phase inverters with an output voltage consisting of only one square-wave pulse per half-cycle are used for low-power equipment, which needs to deliver a relatively stable AC voltage. A low-pass filter must be connected between the inverter output itself and the AC load.

When there is low level of power use two-switch configurations wherever possible:

-push-pull inverter when the DC supply voltage has a very low value (e.g. 6 or 12 V battery); doubling the voltage is not a problem in this case.

-half-bridge inverter when U is about 100 V (e.g. AC voltage mains seen via a transformer and a rectifier); doubling the current raises relatively few problems and producing the capacitive divider is not too costly.

Only when the variations in input voltage U are important or when the output voltage must be correctly stabilised by the inverter itself does the full-bridge converter need to be used. In such cases, the mean operating point is placed at β equal to $\pi/3$. For this value, the harmonic ratio of voltage u' is reduced and, most importantly, there is no third harmonic. This makes it easier to produce the output filter and increase the efficiency of the latter.

5.7 Square-wave versus PWM switching

5.7.1 Square-wave switching principle

In the square-wave switching scheme, each switch of an inverter leg of a full-bridge inverter is on for one half-cycle (180º) of the desired output frequency. This results in an output voltage waveform as shown in Fig. 4.1a. From Fourier analysis, the peak values of the fundamental-frequency and harmonic components in the inverter output waveform can be obtained for a given input Vd as

$$(V_{A0})_1 = 4/\pi\ V_{DC}/2d$$

and

$$(V_{A0})_h = (V_{A0})_1/h$$

where the harmonic order h takes on only odd values, as shown in Fig. 5.13b.

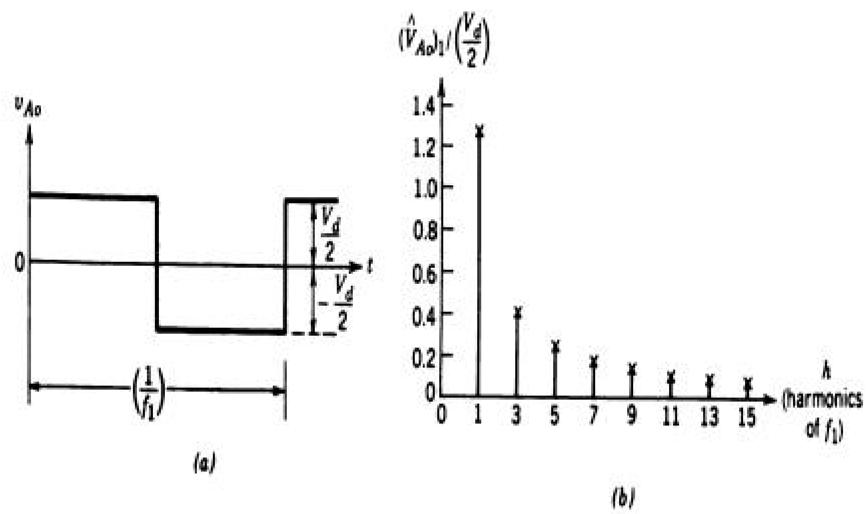


Fig. 5.13 Square-wave switching

One of the advantages of the square-wave operation is that each inverter switch changes its state only twice per cycle, which is important at very high power levels where the solid-state switches generally have slower turn-on and turn-off speeds. One of the serious disadvantages of square-wave switching is that the inverter is not capable of regulating the output voltage magnitude. Therefore, the dc-input voltage Vdc to the inverter must be adjusted in order to control the magnitude of the inverter output voltage.

5.7.2 PWM switching scheme

Increasing use is now made of pulse-width modulation (PWM) to obtain a rough estimate of a sinusoidal voltage by forming each of the output voltage half-cycles of a succession of suitable sized square-wave pulses. The advantage of the full-bridge inverter, which enables the output voltage value to be varied, disappears since the other types can now also guarantee this variation.

However, the two-switch inverters only provide two voltage levels (u' = + *U* or - U), whereas the full-bridge inverter gives three voltage levels (u' = + U or - U or zero).

The full-bridge inverter obviously makes it easier to approximate the required sinusoid. This allows a more effective use of components and an easier filtering. As soon as the power being used is important and the importance attached to performance (especially in reducing the overall weight of the equipment) become paramount, the full-bridge inverter will be used.

In order to produce a sinusoidal output voltage waveform at a desired frequency in inverter circuits, a sinusoidal control signal at the desired frequency is compared with a triangular waveform, as shown in Fig. 5.14a. The frequency of the triangular waveform establishes the inverter switching frequency and is generally kept constant along with its amplitude V_{tri}.

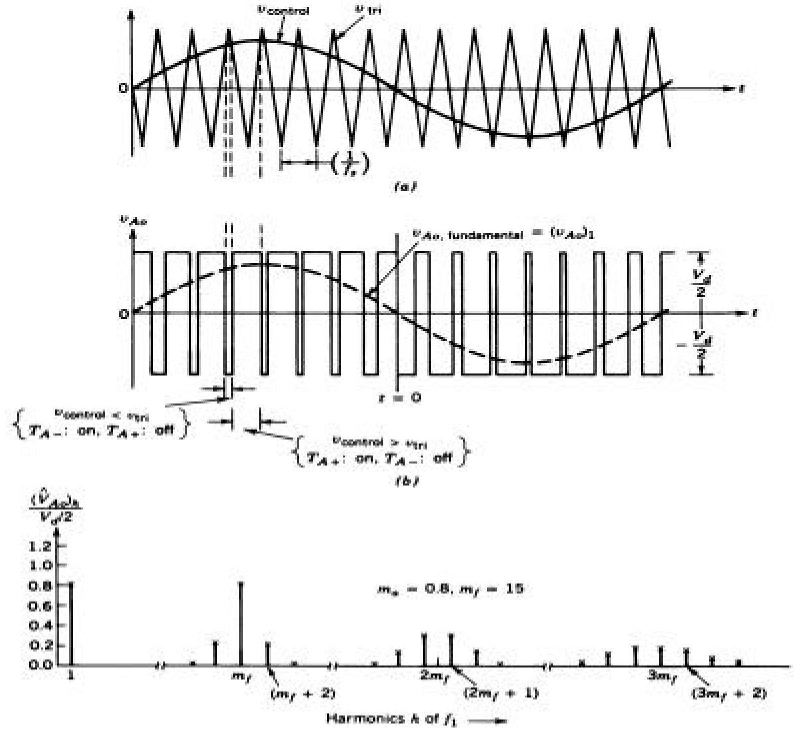


Fig. 5.14 Pulse-width modulation

The triangular waveform v_{tri} in Fig. 5.14a is at a switching frequency f_s, which establishes the frequency with which the inverter switches are switched (f_s is also called the carrier frequency). The control signal $v_{control}$ is used to modulate the switch duty ratio and has a frequency f_1, which is the desired fundamental frequency of the inverter voltage output (f_1 is also called the modulating frequency). It also recognises that the inverter output voltage will not be a perfect sine wave and will contain voltage components at harmonic frequencies of f_1. The amplitude modulation ratio m_a, is defined as

$$m_a = V_{control}/V_{tri}$$

where $V_{control}$ is the peak amplitude of the control signal. The amplitude V_{tri} of the triangular signal is generally kept constant. The frequency modulation ratio mf is defined as

$$m_f = f_s/f_1$$

Since the two switches are never off simultaneously, the output voltage v_{Ao} fluctuates between two values (½Vd and -½Vd). Voltage v_{Ao} and its fundamental frequency component (dashed curve) shown in Fig. 4.2b, which are drawn for $m_f = 15$ and $m_a = 0.8$.

The harmonic spectrum of v_{Ao} is shown in Fig. 5.14c, where the normalised harmonic voltages $(V_{A0})/½V_d$ having significant amplitudes are plotted. This plot (for $m_a < 1.0$) shows three important features of importance

1. The peak amplitude of the fundamental-frequency component $(V_{A0})_1$ is m_a times $½V_d$.
2. The harmonics in the inverter output voltage waveform appear as sidebands, centred around the switching frequency and its multiples, that is, around harmonics m_f, $2m_f$, $3m_f$, and so on. This general pattern holds true for all values of m_a in the range 0- 1.

3. The harmonic m_f should be an odd integer. Choosing m_f as an odd integer results in the fact that only odd harmonics are present and the even harmonics disappear from the waveform of v_{Ao}.

5.8 Semi-conductor switches and driver circuits

5.8.1 Semiconductor-switches

Which types of switch can be used in each case?

In voltage-source inverters (VSI's) the controlled semiconductor devices must have diodes connected across in anti-parallel; they do not have to withstand reverse voltage.

-*If the load is always inductive*, transistors with "long" control signals are used for low and medium power equipment; operation as a dual thyristor is particularly well-adapted to this operational mode. Care must be taken to supply the gate until the moment when the forward current builds up.

-*If the load is always capacitive*, classical thyristors are suitable, since there is natural commutation operation. Asymmetrical thyristors can be used.

-Controlled turn-on/turn-off devices must be used in cases where *the load can be of any nature.* Power transistors for low- and medium-power applications can be used.

METAL-OXIDE SEMICONDUCTOR FIELD EFFECT TRANSISTORS (MOSFET)

The circuit symbol of an n-channel MOSFET is shown in Fig. 5.15a. It is a voltage- controlled device, as is indicated by the i-v characteristics shown in Fig. 5.15b. The device is fully on and approximates a closed switch when the gate- source voltage is below the threshold value, *VGs(u,).* The idealised characteristics of the device operating as a switch are shown in Fig. 5.15c.

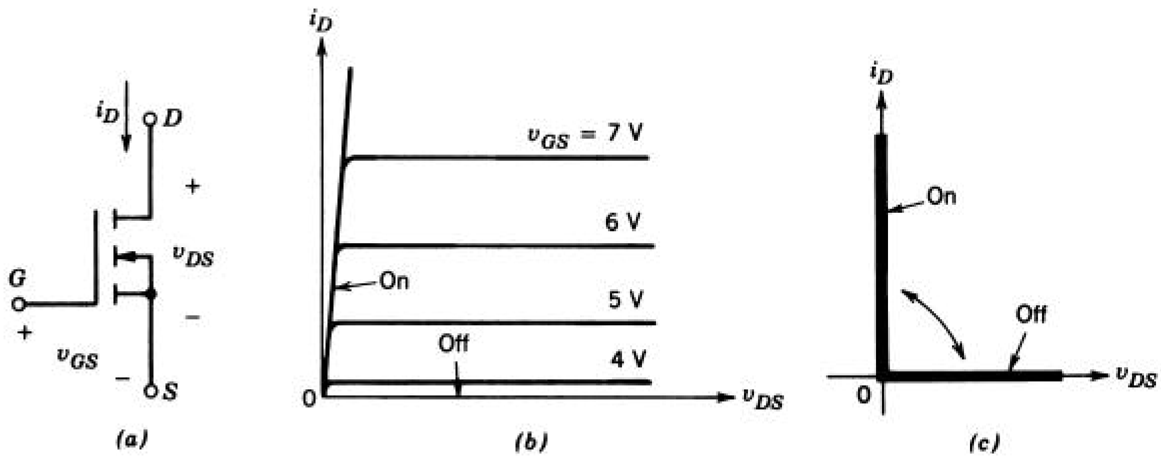


Fig 5.15 MOSFET characteristics

MOSFETs require the continuous application of a gate-source voltage of appropriate magnitude in order to be in the on state. No gate current flows except during the transitions from on to off or vice versa when the gate capacitance is being charged or discharged. The switching times are very short, being in the range of a few tens of nanoseconds to a few hundred nanoseconds depending on the device type.

The on-state resistance $r_{DS(on)}$ between the drain and source of the MOSFET increases rapidly with the device blocking voltage rating. Because of this, only devices with small voltage ratings with low on-state resistance and hence small conduction losses are available. However, because of their fast switching speed, the switching losses can be small. From a total power loss standpoint, 300-400-Y MOSFETs compete with bipolar transistors only if the switching frequency is in excess of 30-100 kHz. However, no definite statement can he made about the crossover frequency because it depends on the operating voltages, with low voltages favouring the MOSFET.

MOSFETs are available in voltage ratings in excess of 1000 V but with small current ratings and with up to 100 A at small voltage ratings. The maximum gate-source voltage is ±20 V, although MOSFETs are available, which can be controlled by 5V signals. Because their on-state resistance has a positive temperature coefficient, MOSFETs are easily paralleled. This causes the device conducting the higher current to heat up and thus forces it to share its current equally with the other MOSFETs in parallel.

INSULATED GATE BIPOLAR TRANSISTORS (IGBT)

The circuit symbol for an IGBT is shown in Fig. 5.16a and its i-v characteristics are shown in Fig. 5.16b. The IGBTs have some of the advantages of the MOSFET, the BJT, and the GTO combined. Similar to the MOSFET, the IGBT has a high impedance gate, which requires only a small amount of energy to switch the device. Like the BJT, the IGBT has a small on-state voltage even in devices with large blocking voltage ratings (for example, V_{on} is 2-3 V in a 1000V device). Similar to the GTO, IGBTs can be designed to block negative voltages, as their idealised switch characteristics shown in Fig. 5.16c indicate.

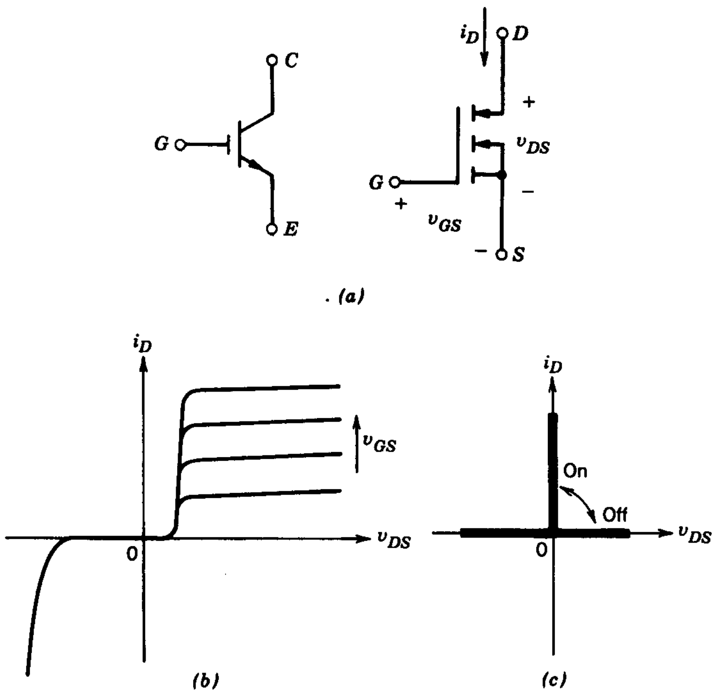


Fig. 5.16 IGBT characteristics

Insulated gate bipolar transistors have turn-on and turn-off times of the order of 1 µs and are available in module ratings as large as 1700 V and 1200 A. Voltage ratings of up to 2- 3 kV are projected.

5.8.2 Gate and base drive circuits

PRELIMINARY DESIGN CONSIDERATIONS

The primary function of a drive circuit is to switch a power semiconductor device from the off state to the on state and vice versa. In most situations the designer seeks a low cost drive circuit that minimises the turn-on and turn-off times so that the power device spends little time in traversing the active region where the instantaneous power dissipation is large. In the on state the drive circuit must provide adequate drive power (e.g., base current to a BJT or gate-source voltage to a MOSFET) to keep the power switch in the on state where the conduction losses are low. Very often the drive circuit must provide reverse bias to the power switch control terminals to minimise turn-off times and to ensure that the device remains in the off-state and is not triggered on by stray transient signals generated by the switching of other power devices.

The signal processing and control circuits that generate the logic-level control signals used to turn the power switch on and off are not considered part of the drive circuit. The drive circuit is the interface between the control circuit and the power switch. The drive circuit amplifies the control signals to levels required to drive the power switch and provides electrical isolation when required between the power switch and the logic-level signal processing control circuits. Often the drive circuit has significant power capabilities compared to the logic-level control signal processing circuits. For example, power BJTs have low values of beta, typically 5 - 10, so that the base current supplied by the drive circuit is often a significant fraction of the total load current.

The basic topology of the drive circuit is dictated by three functional considerations. Firstly, is the output signal, provided by the drive circuit, unipolar or bipolar? Unipolar signals lead to simpler drive circuits, but bipolar signals are needed for rapid turn-on and turn-off of the power switch. Secondly, can the drive signals be directly coupled to the power switch, or is electrical isolation required between the logic-level control circuits and the power device? Most electrically isolated drive circuits will require isolated de power supplies. Thirdly, is the output of the drive circuit connected in parallel with the power switch (the usual situation) or in series with the switch (cascode connection)?

Additional functionality may be required of the drive circuit, which will further influence the topological details of the circuit. Provisions may be included in the drive circuit design for protection of the power switch from overcurrents. Then communication between the drive circuit and the control circuit is needed. In bridge circuits, the drive circuit must often provide blanking times for the power switch. Incorporation of these types of functionality requires design inputs to both the drive circuit and the logic-level control circuit. Waveshaping of the drive circuit output may also he included to improve the power switch performance.

The specific details of component values to be used in a drive circuit will vary depending on the characteristics of the power switch being driven. For example, MOSFET drive circuits need only provide an initial large current as the device turns on and for the rest of the on-state time interval merely provide a large gate-source voltage at low current levels.

It is a good idea to consider how the drive circuit will be configured on a circuit board even at the earliest stages in the design process. The placement of components to minimise stray inductance and to minimise susceptibility to switching noise may affect the choice of topology for the drive circuit.

DC-coupled drive circuits

A simple MOSFET gate drive circuit with only one switch to control the gate current is shown in Fig. 5.17, where the output transistor of a comparator (e.g. LM31 1) controls the MOSFET. When the output transistor is off, the MOSFET is on and vice versa. When the comparator is on, it must sink a current V_{GG}/IR_1, and to avoid large losses in the drive circuit, R_1 should be large. This will slow down the MOSFET turn-on time. This means that the drive circuit is only suitable for low switching speed applications.

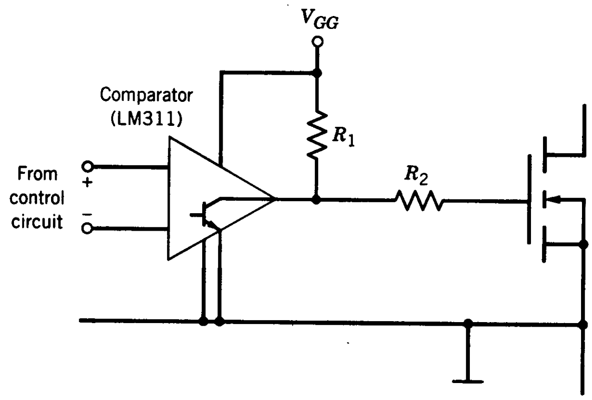


Fig. 5.17 Simple MOSFET gate drive circuit suitable for low-speed and low-switching frequency applications

The inadequacy of this circuit can be overcome by the MOSFET gate drive circuit shown in Fig. 5.18a where two switches are used in a totem-pole arrangement with the comparator (type 311) controlling the *npn*-*pnp* totem-pole stack. Here, to turn the MOSFET on, the output transistor of the comparator turns off, thus turning the *npn* BJT on which provides a positive gate voltage to the MOSFET. At the turn-off of the MOSFET: the gate is shorted to the source through R_G and the pnp transistor. Since no steady-state current flows through R_G in contrast to R_1 described in the previous paragraph, R_G can be chosen to be much smaller in value, which results in much faster turn-on and turn-off times. Very often, instead of using discrete components, similar performance can be obtained, as is shown in Fig. 5.18b by using buffer ICs such as CMOS 4049 or 4050 if a low gate current is needed or a DS0026 or UC 1707. These can source or sink currents in excess of 1 A.

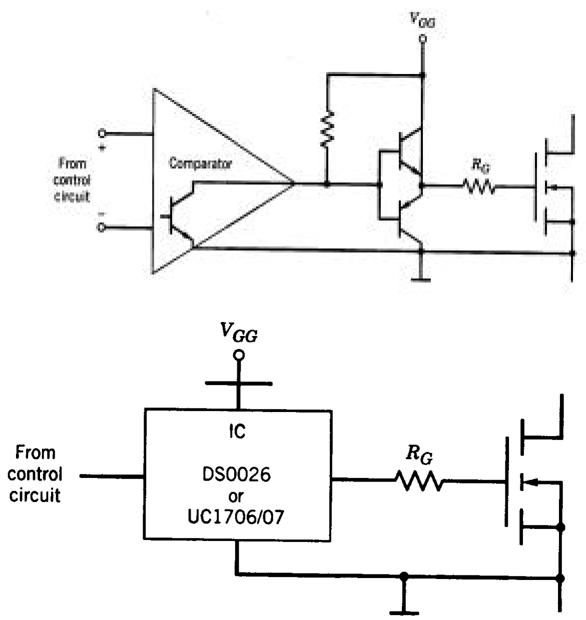


Fig. 5.18 A MOSFET gate drive circuit with a totem-pole configuration for faster turn-off times: (a) discrete totem-pole gate drive circuit; (b) integrated circuit totem-pole gate drive circuit.

DC-coupled drive circuits with bipolar output

In order to operate power semiconductor devices at high switching frequencies, drive circuits must he designed to turn-off the devices as rapidly as they turn on. The descriptions of the switching characteristics of BJTs, MOSFETs, IGBTs and other devices clearly illustrate the need for a reverse bias to he applied to the control terminals of the power switch in order to affect a rapid turn-off. Drive circuits with unipolar outputs are unable to provide the reverse bias required and are thus incapable of providing fast turn-off of power devices. In order to provide a reverse bias to the control terminals of the power device, the drive circuit must have a bipolar output (an output that can be either positive or negative). This in turn requires that the drive circuit is biased by a negative power supply as well as a positive power supply.

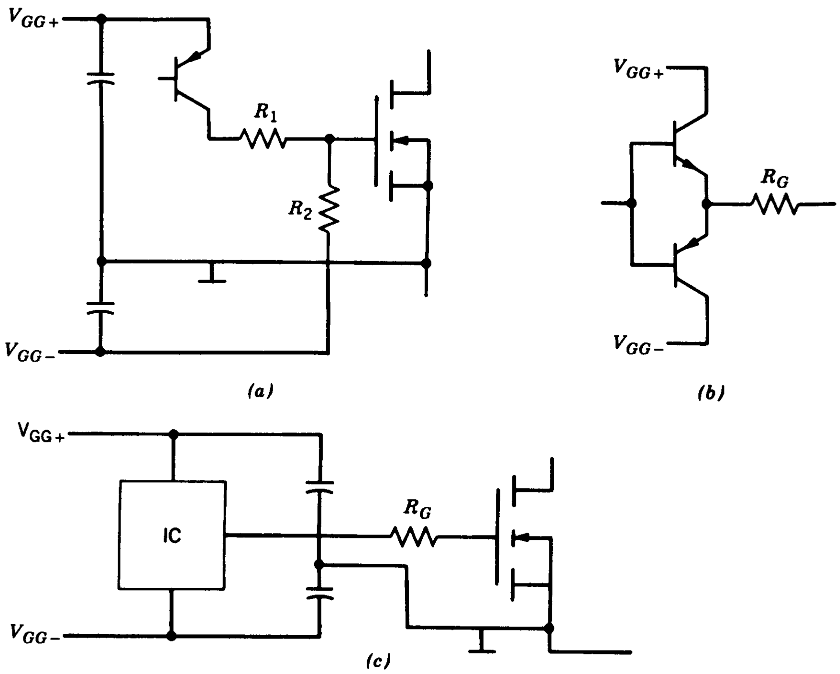


Fig. 5.19 Various gate-drive circuits using split DC power supplies for providing an n-channel MOSFET with positive gate-source voltages at turn-on and negative gate-drive at turn-off

A drive circuit for MOSFETs that provides positive gate voltages at turn on and negative gate voltages at turn off by means of a split power supply with respect to the MOSFET source is shown in Figs. 5.19a to 5.19c. If the control signal is supplied by the logic circuit, which is connected between V_{GG}, and the source of the MOSFET, then the reference input to the comparator should he shifted to be at the mid-potential between V_{GG}, and the MOSFET source using a pre-converter circuit.

ELECTRICALLY ISOLATED DRIVE CIRCUITS

Very often, there is a need for electrical isolation between the logic-level control signals and the drive circuits. The basic ways to provide electrical isolation are either by optocouplers, fiber optics or by transformers.

OPTOCOUPLER ISOLATED DRIVE CIRCUITS

The optocoupler consists of a light-emitting diode (LED), the output transistor, and a built-in Schmitt trigger. A positive signal from the control logic causes the LED to emit light that is focused on the optically sensitive base region of a phototransistor. In optocoupler-isolated drive circuits, the optocoupler itself is the interface between the output of the control circuit and the input of the isolated drive circuit. The input side of the optocoupler is directly coupled to the control circuit and the output side of the optocoupler is directly connected to the isolated drive circuit. The topology of the isolated drive circuit between the output of the optocoupler and the control terminal of the power switch can take many different forms.

Optocoupler-isolated drive circuits can also be used with power MOSFETs and IGBTs. The circuit shown in Fig. 5.20 uses a high common-anode noise immunity optocoupler (HPCL-4503) and a high-speed driver (IXLD4425) with a 3A output capability. The drive circuit uses a single-ended floating IS-V supply and provides a ±15V output voltage for high noise immunity and fast switching to drive the gate of a power MOSFET or IGBT. The integrated high-speed driver circuit connects the gate of the power device to the 15-V bus bar while it simultaneously connects the source to the negative side of the bias supply in order to turn the power device on. To turn the power device off, the drive circuit connects the gate to the negative side of the single-ended supply while it connects the source to the +15-V bus bar.

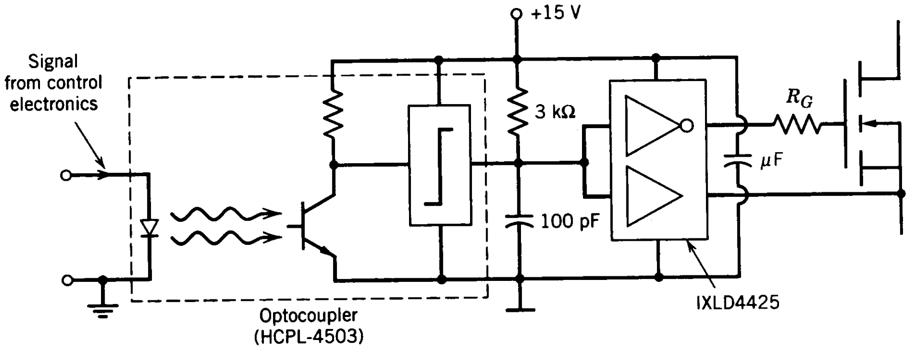


Fig. 5.20 Optocoupler isolation of base drive circuits

TRANSFORMER-ISOLATED DRIVE CIRCUITS PROVIDING BOTH SIGNAL AND POWER

The use of transformers for electrically isolating the drive circuit from the control circuit introduces a great deal of flexibility into the design of the drive circuit. The same transformer used to transfer the control signal from the control circuits to the isolated drive circuit can also be used to provide the isolated bias power and to replace a separate transformer for the isolated supplies.

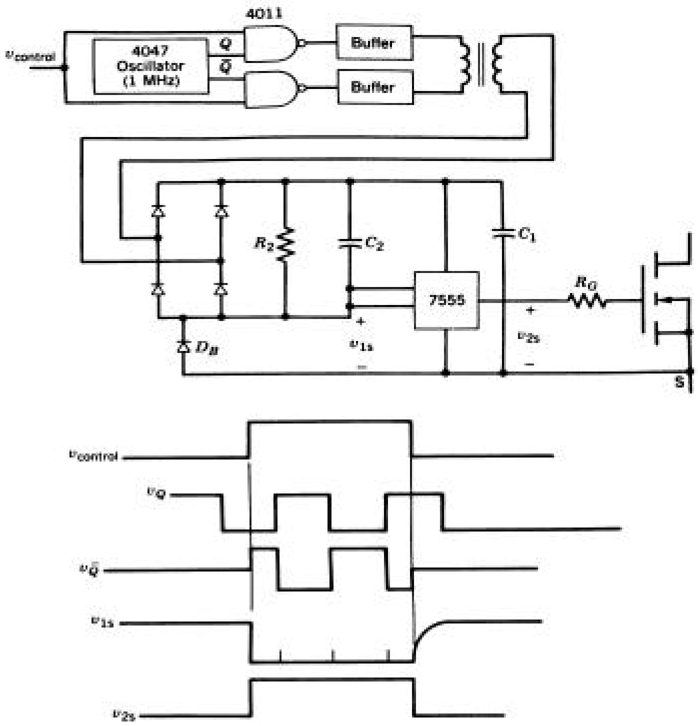


Fig. 5.21 Transformer-isolated MOSFET gate drive circuit using a high-frequency carriers so that the MOSFET can be held on for long periods. No auxiliary DC power supplies are needed since both the control signal and bias power come through the transformer.

If in a given application, the MOSFET to be controlled is to be on for a long time, the circuit shown in Fig. 5.21 can be used. In this circuit a high-frequency oscillator output modulates the control voltage before being applied to the buffer circuits. Now a high-frequency ac signal appears across the transformer primary when the control voltage is high. Thus charging the energy storage capacitance C_1 and the capacitance C_2 at the input to the 7555 IC, which is used here as a buffer and a Schmitt trigger because of its low power consumption. With the input to the 7555 low, it provides a positive voltage to the MOSFET gate, thus turning it on as is shown in Fig. 5.21. At turn off, the control voltage goes low and the voltage across the transformer primary goes to zero. Now C_2 discharges through R_2 and the input voltage to the 7555 goes high, which causes its output voltage to go low, thus turning the MOSFET off. The diode D_B is used to prevent the energy stored in the capacitance C_1 from discharging into the resistance R_2.

POWER DEVICE PROTECTION IN DRIVE CIRCUITS

OVERCURRENT PROTECTION

In some applications the potential may exist for currents to flow through a power device that exceed the capability of the device. If the device is not somehow protected against these overcurrents, it may be destroyed. Power devices cannot be protected against the overcurrents by fuses because they can not act fast enough. Overcurrents can be detected by measuring the device current and comparing it against a limit. At currents above this limit, a protection network in the drive circuit turns off the power device.

A cheaper and normally better way of providing overcurrent protection is to monitor the instantaneous output voltage of the device, for example, the collector-emitter on-state voltage of a BJT or the drain-source voltage of a MOSFET. Fig. 5.22a shows a simple circuit to provide overcurrent protection to a BJT based on this principle. The voltage during the on state at point C will be one forward-bias diode drop above V_{CEsat}. This voltage signal is one of the inputs to the overcurrent protection block that requires the control signal as another input. When the transistor is supposed to be on, if the voltage at point C with some delay is above some predetermined threshold, the overcurrent is detected, and the protection block causes the base drive to turn the BJT off. Depending on the design philosophy, the overall system may be shut down after such an overcurrent detection and may have to be manually reset. The overcurrent detection network can be combined with the anti-saturation network as is shown in the sub-circuit of Fig. 5.22b.

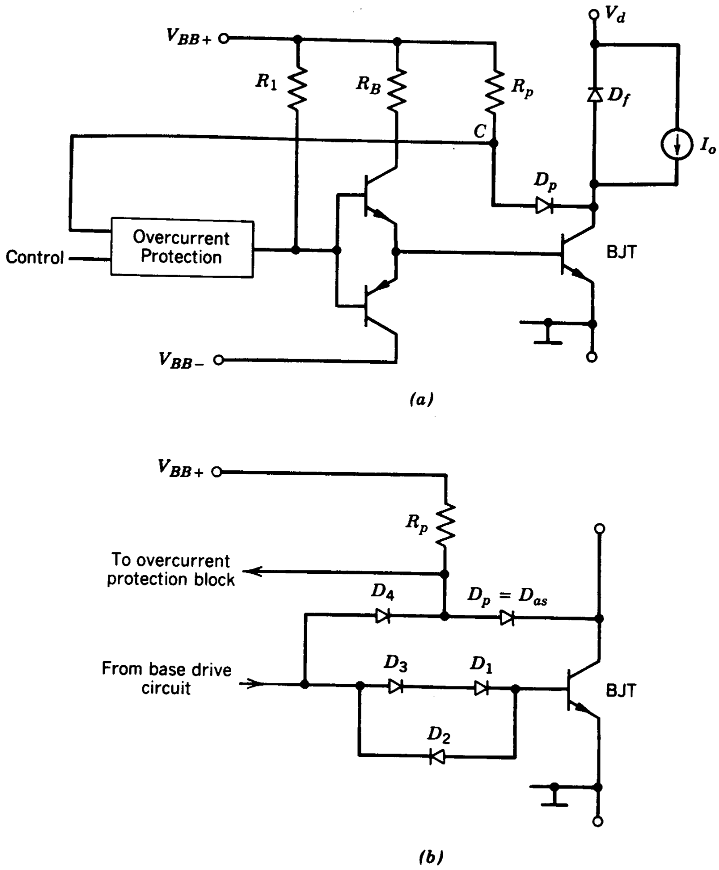


Fig 5.22 (a) overcurrent protection by measuring the instantaneous on-state collector-emitter voltage of the power transistors (b) this protection circuit can be used with an anti-saturation network.

The overcurrent protection should he combined with design measures that limit the maximum instantaneous current through the device. The overcurrent protection circuit must act within a few microseconds to turn off the MOSFET, otherwise it will be destroyed.

BLANKING TIMES FOR BRIDGE CIRCUITS

In the half-bridge and full-bridge circuits, where two transistors are connected in series in one converter leg, it is important to provide a blanking time. In this way the turn-on control input to one transistor is delayed with respect to the turn-off control input of the other transistor in the inverter leg. This blanking time should be chosen conservatively to be greater than the worst-case maximum storage time of the transistors being used to avoid cross conduction. Under normal operation, such a conservatively chosen blanking time will cause a dead time equal to the blanking time minus the actual delay time to occur in which both the transistors in the inverter leg are off. This dead time introduces an unwanted non-linearity in the converter transfer characteristic. This dead time can be minimised by the use of design enhancements to drive circuits, which minimise turn-on and turn-off delay times in power semiconductor devices being used as the power switches. These design enhancements include the use of anti-saturation diodes with BJTs, drive circuits with bipolar outputs, speed-up capacitors, and so forth.

This blanking time in the control inputs can be introduced by means of the circuit shown in Fig. 5.23a where the control signal is common to both BJTs of the converter leg. When the control signal is high, the upper transistor T, should be on and vice versa. The polarised RC network and the Schmitt trigger introduce a significant time delay in the turn-on of the BJT and almost no time delay in the turn-off of the transistor. The difference between these two time delays is the blanking time needed. The waveforms are shown in Fig. 5.23b where, when the bridge control input goes low, a significant time delay occurs in the control signal to turn on the bottom transistor T- and almost no time delay occurs in turning off the upper transistor T,. The blanking time and the dead time are also shown in Fig. 5.23b.

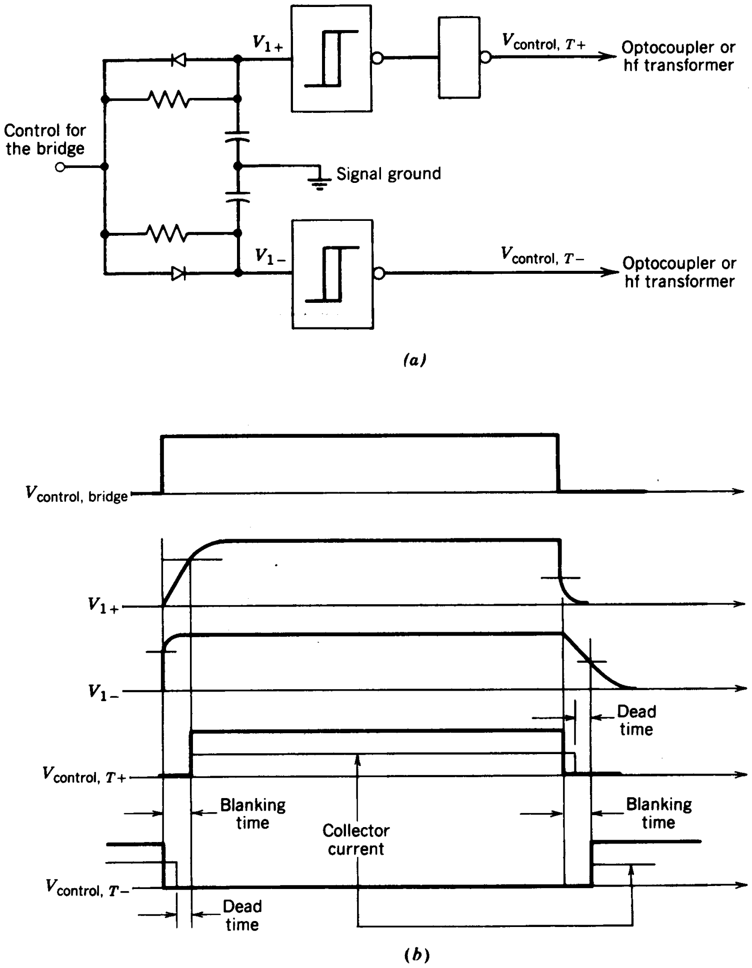


Fig. 5.23 (a) Circuit for providing blanking time to the base drives of BJTs in a bridge configuration so as to avoid cross conduction of BJTs. The dead times are the result of the BJT storage times, which are shown on (b) the collector current waveform.

CIRCUIT LAYOUT CONSIDERATIONS

MINIMISING STRAY INDUCTANCE IN DRIVE CIRCUITS

There are several practical considerations, in the design and fabrication of drive circuits, which are crucial to the successful operation of the circuits. The schematic shown in Fig. 5.24a serves as the focus of these considerations, and although it features a BJT the discussion based on this figure applies equally well to all power semiconductor devices. Firstly, the length of the conductor that connects the base drive circuit to the emitter of the power BJT should be as short as possible to minimise the stray inductance illustrated in Fig. 5.24b. Otherwise the turn-off will be slowed down and possibly unwanted oscillations may occur. Consider a positive base current i_B that turns the BJT on, which in turn causes the collector current i_C to increase rapidly. The stray inductance illustrated in Fig. 5.24b will induce a voltage that will tend to reduce the base current. If this then causes a reduction in the collector current, there will be a subsequent negative di_c/dt and the induced voltage will cause an increase in i_B. This then represents the start of unwanted oscillations.

In minimising the stray inductance, all power devices including BJTs, MOSFETs, thyristors, GT0s, IGBTs, and so forth, should be treated as four terminal devices having two control terminals and two power terminals (as is illustrated for the BJT in Fig. 5.24c). To facilitate the reduction of this stray inductance in high-power transistor modules, manufacturers provide a separate emitter terminal for the connection of the drive circuit as is shown in Fig. 5.24c. Such separate additional terminals are also found on thyristors, GTOs, and IGBTs.

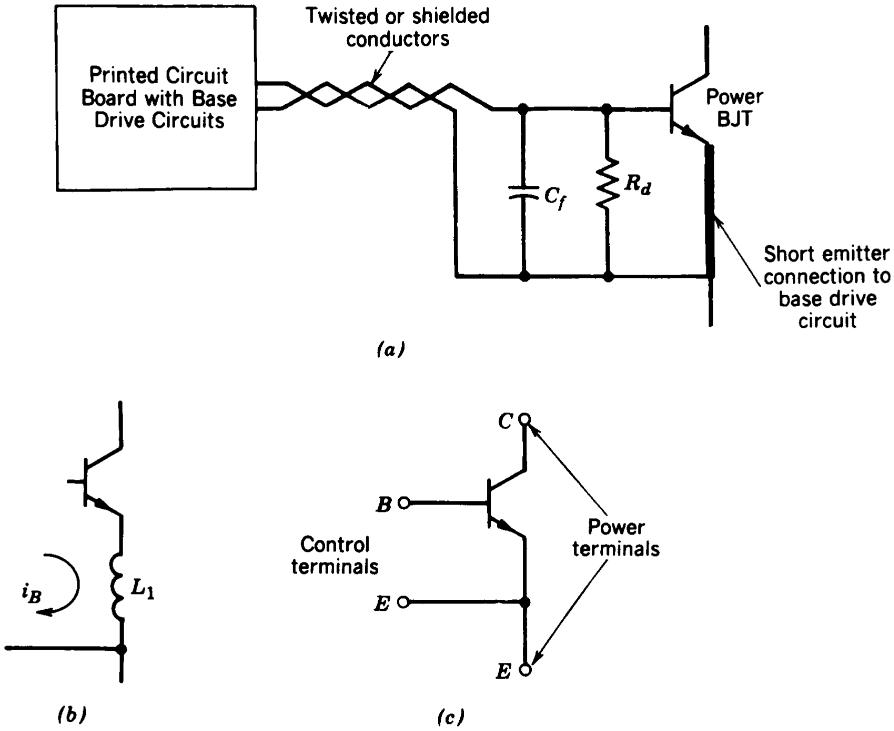


Fig. 5.24 (a) circuit lay-out and the interconnection considerations in connecting base drive circuits to power BJTs in order to minimise stray inductance (b) and other potential problems. Some BJTs have extra emitter connections as shown in (c) to help minimise such potential problems

SHIELDING AND PARTITIONING OF DRIVE CIRCUITS

Stray inductance must also be minimised in the high current power loop to which the output terminals of the power device are connected. If the stray inductance is not minimised, even with careful layout, it may be necessary to reduce further the overvoltages at turn-off by means of snubber circuits or control of the turn-off times. It should be kept in mind that one centimetre of unshielded lead has about 5 nH of series inductance. Thus the lengths of all unshielded leads should be kept to an absolute minimum.

In many designs, the basic drive circuit may be on a printed circuit board at some distance away from the power transistor, which is mounted on a heat sink. A twisted pair of wires or even a shielded cable where the shield is connected to the emitter terminal should be used to minimise the stray inductance and the inductive pick up of noise in the base drive circuit. A small filter capacitor C_f and damping resistor R_D can he added across the base and emitter terminals as shown in Fig. 5.24a to avoid oscillations and the problem of retriggering at the turn-off of the BJT.

If more than one base drive circuit is put on the same printed circuit board, they must he put on separately dedicated areas of the board with a minimum distance of at least 1 cm between the areas. This is especially important on double-sided or multiple-layer circuit boards. There must never be an intermixing of the printed wires of the different isolated base drive circuits on any area of the card.

5.9 Component temperature control and heat sinks

This section discusses the need to control the internal temperature of power electronic components and the factors to be considered in selecting passive components including resistors, capacitors, and heat sinks. Excessive internal temperatures are detrimental to all power electronic components, especially power semiconductor devices. A fundamental understanding of heat transfer is needed not only for the design and specification of heat sinks but also in the design of inductors and transformers where thermal considerations are a major part of the design.

5.9.1 Control of semiconductor device temperatures

The theoretical upper limit on the internal temperature of a semiconductor device is the so-called intrinsic temperature, T_i, which is the temperature at which the intrinsic carrier density in the most lightly doped region of the semiconductor device equals the majority carrier doping density in that region. However, the maximum internal temperatures specified on data sheets are much less than this limit. The power dissipation in power semiconductors normally increases with the internal temperature, and the losses become excessively high even at temperatures of 200°C. Device manufacturers will typically guarantee the maximum values of device parameters such as on-state conduction voltages, switching times and switching losses at a specified maximum temperature, which vary from one type of device to another and is often at 125°C.

In a design process, one of the design inputs is the worst-case junction temperature. A system intended to have, high reliability would be designed for a worst-case junction temperature in the semiconductor devices of 20-40°C below 125°C. Otherwise a value of 125°C is commonly used in the worst-case design input. Some power semiconductor devices and signal level transistors and ICs can operate at temperatures even slightly above 200°C. However their reliability (expected operating lifetime) is low, and the performance characteristics may be poor compared to operation at 125°C, for example. Moreover, the manufacturer will not guarantee the parameters above the maximum temperature specified on the data sheet.

In designing power electronic equipment, especially for high ambient temperatures, the thermal layout must be considered at an early stage. The heat sink size and weight, its location in the equipment cabinet and surrounding temperature should be considered at the beginning of the design process. It is important to be able to mount the heat sinks with their fins in a vertical position with ample room for natural convection of the air without a fan. The possibility of heating by the sun must be considered as part of a worst-case set of design inputs.

A bad thermal design will make the equipment much less reliable than intended. A rule-of-thumb to keep in mind is that the failure rate for semiconductor devices doubles for each 10-15°C temperature rise above 50°C. The choice of the correct (most economical or cheapest in production) heat sink is only a part of the thermal design processes for a power electronic system. At an early stage of the design, the designer should be free to consider a large or small heat sink that may be cooled by natural convection, or by a fan (ac motor fans controlled by a small power electronic inverter are much more reliable than a dc motor fan).

5.9.2 Heat transfer by conduction

THERMAL RESISTANCE

When a section of material such as is shown in Fig. 5.25 has a temperature difference across it, there is a net flow of energy from the higher temperature end to the lower temperature end. The energy flow per unit time, that is, power, is given by

$$P_{cond} = \lambda A \Delta t / d$$

where $\Delta T = T2 - T1$ in °C, A is cross-sectional area in m^2, d is the length in m, and λ is the thermal conductivity in W-m^{-1} °C^{-1}. For 90% pure aluminium, which is typically used for heat sinks, the thermal conductivity is 220 W-m^{-1} °C^{-1} . Values of λ for other materials can be found in the literature.

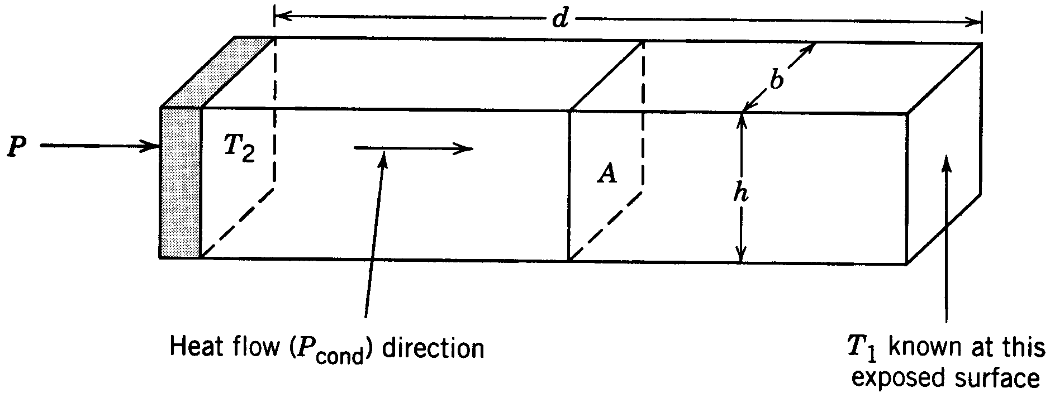


Fig. 5.25 An isolated rectangular rod that conducts P watts of heat energy per unit time

Example: A transistor module is mounted on an aluminium plate having dimensions h = 3 cm, b = 4 cm, and d = 2 mm (refer to Fig. 5.25). A temperature drop of 3°C is allowed from one surface of 3 * 4 cm^2 to the other.

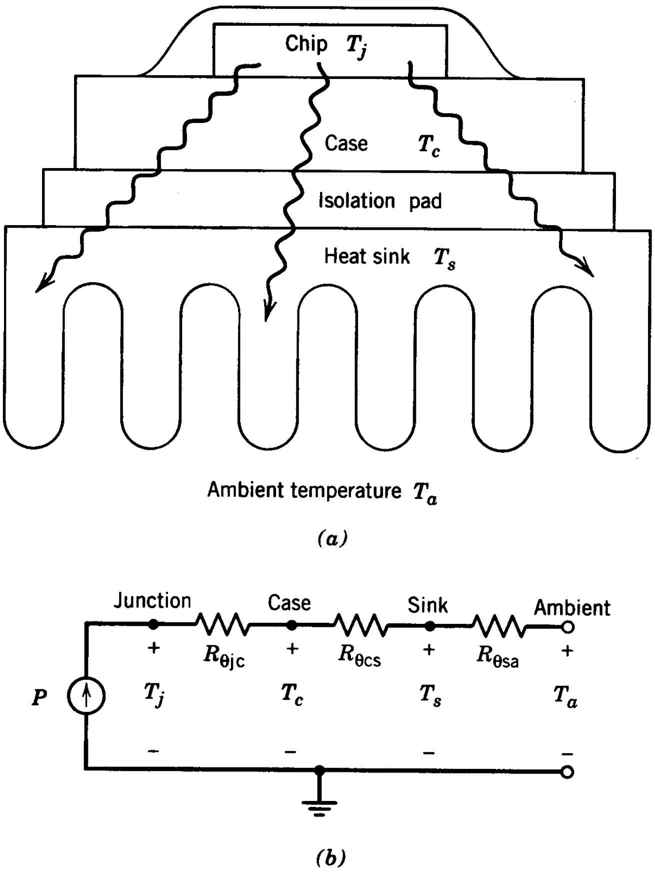


Fig. 5.26 Steady-state heat flow and thermal resistance in a multiple layer structure including a (a) heat sink and (b) equivalent circuit based on thermal resistors

Find the maximum power that can be generated in the module. Ignore any heat losses to the surrounding air.

$$P = \lambda A (T_2 - T_1) / d = (220)(0.03)(0.04)(3)/(0.002) = 396 \text{ W}$$

The thermal resistance is defined as

$$R_{\theta,cond} = \Delta T / P_{cond} \quad \text{or} \quad R_{\theta,cond} = d / \lambda A$$

The thermal resistance has units of degrees centigrade per watt.

Often the heat must flow through several different materials, each having different thermal conductivity and perhaps different areas and thickness. A multilayer example, which models the heat conduction path from a region in the silicon device to the ambient, is shown in Fig. 5.26. The total thermal resistance from the junction to the ambient (ja) is given by

$$R_{\theta,ja} = R_{\theta,jc} + R_{\theta,cs} + R_{\theta,sa}$$

Each contribution to the total thermal resistance is computed using the proper values of λ, A, and d. The resulting junction temperature, assuming a power dissipation of P_d, is

$$T_j = P_d (R_{\theta,jc} + R_{\theta,cs} + R_{\theta,sa}) + T_a.$$

in analogy with electric circuits. If there are parallel paths for heat flow, then the thermal resistances are combined in exactly the same manner as electrical resistors in parallel.

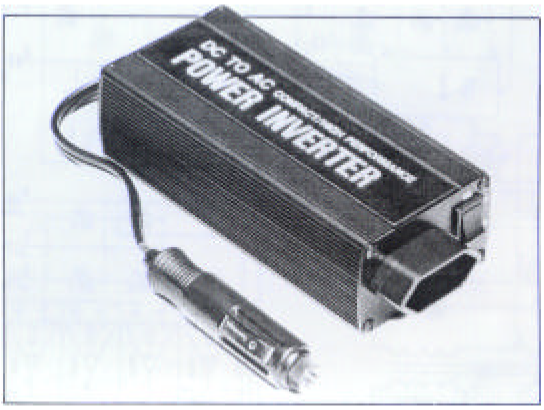
Manufacturers of power devices put great emphasis on keeping the thermal resistance as economically low as possible. This means keeping the length d, of all heat flow paths as short as possible, consistent with the requirements of breakdown voltage, mechanical ruggedness, and other requirements. It also means that the cross-sectional area A should be as large as possible consistent with other design requirements such as minimising parasitic capacitance. The package should be made of material with a high thermal conductivity.

5.10 Examples

Two designs will be discussed in this section:

- A HF full bridge square-wave-inverter with novel safety- and warning-functions
- A HF full-bridge inverter with a special function PWM-IC

5.10.1 150VA HF inverter



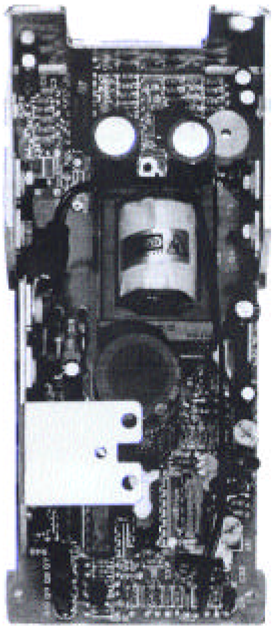


Figure 5.27 150Watt DC/AC converter from ELV GmbH

This inverter is small, light, has an efficiency of 90% and has many safety functions. Its block diagram is shown in figure 5.28. Appendix 9 shows the electronic circuit. IC1a and IC1b guard the battery voltage. Whenever the voltage falls below 11V the buzzer is activated. If the voltage falls below 10.5V the complete inverter is switched off. The chopper circuit, based on IC2 and transformer T1, form the first step to 230Vac. After transformer T1, 340Vdc is being generated. This is the peak-voltage of a 230Vac sinewave. The full-bridge Q12-Q15 creates the AC output-voltage. The steering for these FETs, IC3 and IC5, keep the effective output voltage Veff at 230V. IC4a to d form the security of this output-circuit. This security guards the temperature of the full-bridge, the peak- and continuous-current and the power-on delay of the output power.

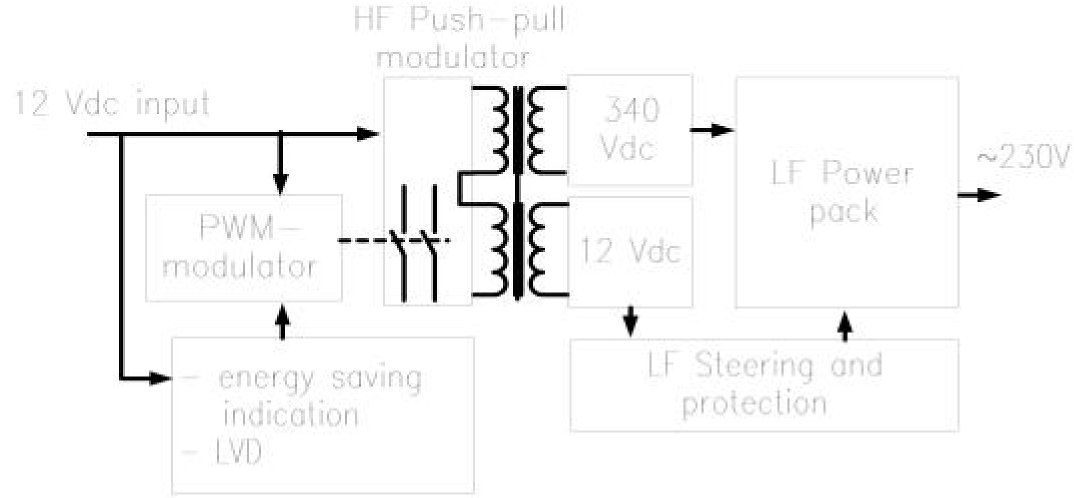


Figure 5.28 Block diagram of a HF-DC/AC-converter

As soon as the device is switched on, there will be an inrush current. The *charge regulator,* used in the SHS has to be able to accept this current. Furthermore, the wires to the charge regulator must be as short as possible, in order to minimise the energy losses.

IC2 generates the HF-steering at the primary side of the transformer. In this design this PWM-IC is 'abused' for the steering of Q1 to Q4, so that there is a 'blanking-time' generated in this part of the push-pull driver. At the frequency used (45kHz), the blanking-time become important. Otherwise there would be considerable energy-loss via T1 when two of the four FETs are not totally closed during the off cycle in this push-pull configuration.

At the secondary side of the transformer two windings are made. This creates 340Vdc for the full-bridge output circuit. The other winding creates the 12Vdc power supply for the steering of the secondary full-bridge. In this way the SHS input has total galvanic separation from the output circuit. At the primary side of the transformer, IC1a and IC1b form a low-voltage warning and security system. The disadvantage of a warning at 11V is, that when the usual charge regulator switches off at 11.5V there is no warning. The warning voltage within SHS should be set at approximately 11.8V. Whenever there is a low-voltage cut-off from IC1, IC2 is simply disabled, so that there is no AC generation at the primary side of the transformer.

At the secondary side of the transformer, the output full-bridge, Q7…Q10 are the drivers for the output FETs Q12…Q15. The gate-source voltage of Q13 and Q15 is floating towards the steering voltage of the full-bridge. In order to get it in conduction, you have to add D9/C24 and D8/C48. These capacitors are floating, just like Vgs of Q12 and Q14. Whenever Q12 and Q14 are conducting, C24 and C48 are connected to Vss of the secondary 12Vdc power-supply. They can now be charged upto 12V via the diode D9/D8, so that it can form a 'floating power-supply' for the FETs during the on-time.

One half of IC3 creates needle-pulses at a frequency of 100Hz. The other half of IC3 creates pulses with a duty-cycle dependent on the voltage at C26, which is the output DC-voltage of the transformer. In this way there is some sort of 'Pulse-Width' steering on the output, in order to stabilise the output voltage over the total output power range. IC5 is a flip-flop that creates the 50Hz steering from this 'PW' signal, in order to steer the FETs. This flip-flop configuration, steered by the PW-signal, also shuts down the full-bridge for a short period during the zero-crossings. This 'blanking-time' is steered in such a way that the effective output voltage is constant.

Finally, IC4 takes care of the safety function at the secondary side of the transformer. IC4d has two functions. Firstly it takes care of the power-on delay, based on the voltage on C21. IC4d disables IC5 during this delay, so that there is no output voltage during this time. The other comparators monitor other safety functions. As soon as it is necessary to switch off the output signal, these comparators discharge C21, so that IC4d can disable IC5 again. IC4a and IC4b check the peak-current and the continuous-current delivered to the load. C22 is larger than C23, so that peak-current-detection responds sooner than the continuous-current protection. And finally IC4c checks the FET temperature.

This is a very efficient DC/AC-converter with many safety features. In spite of the 150Watt design this is a good example for use in SHS. With some adaptations it could also be made suitable for larger power systems.

5.10.2 DC/AC Isolated Battery inverter using the integrated circuit HIP4082

This is a more advanced example of an inverter from Intersil. Appendix 10 shows the electronic circuit

The topology chosen uses the W4082 to provide a 60kHz, low voltage, square-wave to drive a small isolation power transformer. The output of this transformer is rectified, filtered, and the high voltage inverted again to produce the low frequency (55Hz) output waveform required. (55Hz was chosen as a compromise between the 50Hz and 60Hz power standards.) A variable duty-cycle quasi-square-wave output waveform was chosen over a sinusoidal waveform for simplicity and cost-effectiveness. The 60kHz frequency of the primary-side inverter minimises the cost and size of the transformer, while the square-wave output waveform minimises the size of the secondary side rectifier filter.

This DC-AC Inverter design presents one approach to regulating the output voltage, protecting the inverter from overcurrents and reducing output voltage when benign over-currents would tend to cause the system to overheat. A neon light is used to warn of high voltage when the unit is energised.

The primary inverter uses the HIP4082 to convert the 12V battery potential to approximately 160Vdc through transformer isolation. The second inverter converts the 160Vdc voltage into a quasi- square-wave representation of a sine-wave having a frequency of 55Hz with a peak voltage matching the high voltage DC-bus potential. A simple feed forward technique regulates the AC secondary voltage to 115Vac, while the battery varies over a range of approximately 11V to 15V. (Through component modification, 230Vac is possible.)

Primary Inverter Design

Input Filter

The primary-side inverter is comprised of a simple R-C input filter. Capacitors, C5 and C6 provide a stiff, sag-free source for the inverter-bridge comprised of Q1 to Q4 as shown in the schematic (see Appendix 10). To aid this process, resistor R37 and ceramic, non-inductive capacitor C7, parallel C5 and C6. Automotive applications can be fraught with voltage transients across the battery terminals. To prevent these transients from exceeding the Vcc voltage ratings of the HIP4082 and other ICs on the primary inverter section, R7, C4 and D3 clamp the Vcc voltage to 16V or less.

Primary Inverter Waveform Generation

To minimise the size of the secondary filter, a 50% duty cycle square-wave was chosen for primary excitation. With a nearly constant, low-ripple voltage, secondary filtering can be minimised and ripple nearly eliminated. An inexpensive ICM7555 timer was chosen. This timer, an improved 555 timer, reduces Vcc to ground cross conduction current spikes, thereby minimising bias current requirements.

The timer, U1, operates in the a-stable mode, accomplished by tying pins 2 and 6 of the timer together. The a-stable mode requires only one resistor R1, and one capacitor C3. A 50% duty-cycle square-wave is available at the "OUT" (pin 3) terminal of timer U1.

The timer 'out' pin drives the clock input, pin 3, of a CA4013 D-flip-flop connected as a divide-by-two circuit. To accomplish the divide-by-two function, the QNOT output of the flip-flop is fed back to its own datainput, D. The Q and QNOT outputs of the CA4013 provide an exact 50% duty-cycle square-wave at half the timer's output frequency and are applied to the ALI-BHI and the AHI-BLI gate control inputs of the HIP4082. The ICM7555 clock frequency was chosen to be 120kHz so that the primary inverter frequency would be 60kHz.

Choosing proper Dead-Time

The dead-time chosen for eliminating shoot-through currents in the QI-Q4 and Q2-Q3 MOSFET pairs is determined by the value of R2 connected between the DEL and Vss pins of the HIP4082. The 15K value chosen provides approximately 0.5s of dead-time, sufficient to avoid shoot-through when using RFP7ON06 MOSFETs.

Controlling di/dt and switching losses

Choice of gate resistor values for R3-R5, and R8 is based upon several factors. The gate resistors tailor the turn-on and turn-off rise times of the power MOSFETs and the commutation di/dt. The di/dt affects commutation losses and body diode recovery losses. As di/dt increases, recovery losses increase, but the commutation losses decrease. As di/dt decreases, recovery losses decrease, and commutation losses increase. Generally there is an ideal commutation di/dt which minimises the sum of these switching losses. Inductances which are in series with each power MOSFET, also control di/dt. Stray inductance between the filter capacitor and the positive and negative bus rails help reduce the switching di/dt. Gate-source capacitors help reduce the ringing at the inverter bridge terminals associated with the output choke employed to reduce EMI.

Transformer Specification

Current Product-to-Market needs often require having a magnetics supplier design the magnetic devices in your design. The electronics designer provides a detailed specification to the transformer supplier.

The specification should include the minimum frequency of operation, the maximum applied voltage and waveform, the continuous and overload current profiles and the operating ambient temperature. The required transformer regulation must also be specified. Transformer designers need to know this, in order to size the transformer wire and leakage inductance. The power handling capability and operating frequency influence the choice of core, size and geometry and ultimately the cost and size of the transformer.

It is important that the transformer designers know the transformer excitation waveform. This is important because the current waveshape dictates the form factor or the value of RMS (root-mean-square) current which will result for a given required average current. The size of the filter capacitor, the equivalent series impedance of the secondary, and the output voltage waveshape will determine the current waveform and form factor. The RMS current determines the power losses in the transformer and temperature rise. The leakage inductance of the transformer will be minimised if the ringing on the inverter bridge of the primary inverter is minimised.

In order to minimise transformer size and maximise winding fill, the primary was formed of 3 separate windings which were paralleled to supply the approximately $30A_{RMS}$ required at rated output power. The DC resistance of each primary winding is less than 1OmΩ. There are two equal, but separate, secondary power output windings. When series-connected, these secondary power output windings provide 230Vac load power. Series connection via soldered jumper wires allows for 230Vac operation, but this should not be attempted without changing the power MOSFETs and the voltage ratings of several capacitors. A 500V MOSFET with an Rds(ON) of 1.5Ω such as the IRF83OR would be a suitable device for 230Vac operation. Besides having to double the capacitor voltage ratings of C23, C27 and C13, the capacity of C23and C27 will have to be dropped by a factor of 4. This maintains the power dissipation in resistors, R34 and R38 to remain as they are for the 115Vac design.

A third secondary winding provides a low voltage control power to all of the secondary-side inverter logic and gate drivers. This voltage is nominally 20.5V (peak of the square-wave) when there is a nominal 13.6Vdc, applied to the battery input terminals. This winding must output at least 14V at the minimum battery voltage in order to keep the series regulator out of saturation. This winding carries less than 10OmA, so the winding gauge will be determined more for strength than for current carrying capacity.

Secondary-Side Inverter

The secondary-side inverter functions include the power MOSFETs Q6 to Q9, their associated gate resistors and capacitors, the snubber, the current-sensing resistor, the output choke, the indicator lamp and the filter. The inverter topology is a full-wave H-bridge and synthesises a pseudo sine-wave by alternately switching on Q6 and Q9 for positive half sine-waves and Q7 and QS for negative half sine-waves. Since the inverter requires the ability to regulate the RMS output voltage over a wide ranging DC battery input voltage, some means of varying the conduction period of the Q6-Q9 and Q7-Q8 pairs must be implemented.

The choice of square-wave output over sinewave output simplified the pulse-width-modulator (PWM) and minimised the MOSFET power dissipation. Varying the width of the positive and negative conduction periods inversely with the voltage level of the high voltage bus maintains the RMS value of the output wave-form relatively constant.

The left half-bridge includes MOSFETs Q6 and QS and the right half-bridge includes MOSFETs Q7 and Q9. The required phase-shift function is implemented by a simple control circuit. The technique can be expanded to create sinusoidal or other output waveform types with added complexity, of course.

The high voltage output waveform can exhibit a severe voltage transient, with the potential to upset the output voltage across the connected load and possibly to destroy the high voltage gate driver, HIP2500, or the secondary-side inverter MOSFETs. Therefore phase-to-phase and DC bus snubbers were added. Resistor, R34 and capacitor, C23, form the bus snubber and resistor, R38 and capacitor, C27, comprise the phase-to-phase snubber. Phase- to-phase or "AC" snubbers allow their capacitors to completely charge and discharge each cycle of the switching waveform and at high switching frequencies will dissipate a considerable amount of power. R38 and C27 were not used, but space for them was provided.

The energy required to charge a capacitor to a certain voltage and discharge it to its original voltage level is the product of the capacitance and the voltage attained across the capacitor during the charging cycle. The AC snubber dissipation is therefore: V^2bus * Csnubber * f_{PWM} , where f_{PWM}= 55HZ. The fact that the snubber power would have been quite substantial is one reason why a high frequency PWM modulation technique was avoided. The charge transferral in the DC bus snubber is almost negligible, because the capacitor voltage does not appreciably change or switch polarities like that of the AC snubber. The power rating of the serier resistor of the DC snubber can therefore be minimum (1/4W in this design).

A bifilar-wound choke was placed between the output banana jacks BJ3 and BJ4 and the high voltage inverter in order to reduce conducted EMI at the load. Capacitor C13 aids in this regard.

6. ADVISORY SERVICES BODY FOR IMPROVING BOS-COMPONENTS

6.1 Establishing an advisory services body

The normal method for companies to have their solar PV-products approved is to have them tested first. No internationally accepted standards yet exist, but some countries have (draft) national standards. The World Bank requires PV-products that are provided with their assistance to meet certain requirements. These have to be tested by an independent testing laboratory.

Some products, however, do not meet the requirements completely, mostly on one or two points. The result of the test-laboratory is a test report that describes the weak points of the product. Those problems will have to be solved in order to get the product officially accepted.

Larger or modern companies have the knowledge about electronics in house, so they will be able to improve their products themselves. Especially in the field of DC/AC-conversion, there will often be good technicians available within the company. There are many small companies however, which lack that knowledge. Some companies producing Solar Home System components work with older designs and do not have technicians who can further improve the product. They are dependent on external organisations which they can consul for technical advice. They will bring the test-report of the independent testing laboratory and ask how to improve the product.

An independent testing laboratory is not allowed to recommend improvements to the tested equipment. This should be done by an advisory services body which should be independent from the testing laboratory. The testing laboratory will only give a test report, but will not work directly on technical detailed improvements of the design.

An advisory services body can be a private company, a university department or a research and development organisation. It is important that the body is not directly related to any of the companies producing or importing solar PV-equipment in the country. Otherwise, conflicts of interest might arise.

6.2 Procedures of an advisory services body

The advisory services body can be approached by the PV-equipment manufacturer if products do not pass the tests of the independent testing laboratory. The report of the testing laboratory will provide some guidance on where to locate the problems and where to start further investigation.

Sometimes the manufacturer wishes to have their product optimised before having it tested by the testing laboratory. In that case the staff of the advisory services body need to run the tests completely. Often it will help to have a checklist, according to the local (national) requirements. Therefore this will be the starting point of the practical work in this training course.

After the complete tests have been conducted by either the testing laboratory or the advisory body, the actual design has to be analysed. When a certain factor has been improved and the circuit diagram has been changed, the product has to be tested completely. **One change for a particular problem can influence other settings in the device**. Especially fluorescent lamp inverters are very sensitive to this.

While making the changes in a design there is another aspect requiring attention:
The modifications in an existing design should not 'triple the costs of the product.' A company is usually reluctant to adopt a completely new design. The starting point is always to improve the existing design in some small details, so that the official requirements are met.

In practice it can occur that there are some very poor designs in production. Some of these designs are so poor, that they really cannot be improved by simple means. In that case you will have to develop a whole new set-up. In the end the manufacturer must take the main responsibility for this decision.

7. PRACTICAL PART OF THE TRAINING COURSE

In this part of the workshop an attempt will be made to test and modify one or more existing, local designs of fluorescent lamp inverters, charge regulators and DC/AC-converters. Most countries have several manufacturers so it should not be difficult to find samples for testing. It is recommended to obtain three samples from every type of a certain device in order to ensure that there are always spare devices in the event that one fails.

For all three subjects the following order will be maintained:

- First a checklist will be made, based on the local requirements and test procedures. These test procedures will be discussed, so that the execution of the tests is clear before starting. Examples of checklists are given in appendix 8.
- In this workshop the samples will first have to be fully tested. When a device has some characteristics that do not meet the requirements, these will have to be improved after completing all the tests. Never assume that a certain requirement will be met. **Always execute all tests**, even when a test seems unnecessary.
- An attempt will then be made to improve the weak spots. It can be very worthwhile just to experiment with some component values, in order to see what happens when some small changes are made. This increases the understanding in the set-up and the behaviour of the device.
- The aim is to have a short report on all three subjects at the end of the workshop, which describes the identified modifications and their effects on performance and safety of the device.

Each subject will take approximately one week to carry out.

7.1 Fluorescent lamp inverters

First Practical week

7.1.1 Measuring fluorescent lamp inverters

In this stage all functional tests can be carried out. When a calibrated digital oscilloscope is used, the electrical efficiency and the crest factor can be measured. A good electrical efficiency measurement can raise some problems. Input power can easily be measured. The difficulty is the output power measurement. Because true RMS metres are quite expensive, they are not always available in developing countries. A good solution is to store the file from the oscilloscope in an Excel-file, and multiply output voltage and -current from every measuring point. Excel can then calculate the output power to the lamp. Measuring with an external shunt resistor and adding this voltage to the scope is not a good solution (figure 7.1). This shunt resistor can add parasitic capacity or inductance to the secondary circuit. Especially when the resistor is metal wound. Furthermore, this resistor cannot be calibrated, and its signal has to be amplified so much, that the signal/noise ratio from the signal is too large. The result is not reliable. When measuring with the scope, a **calibrated current-probe** has to be used.

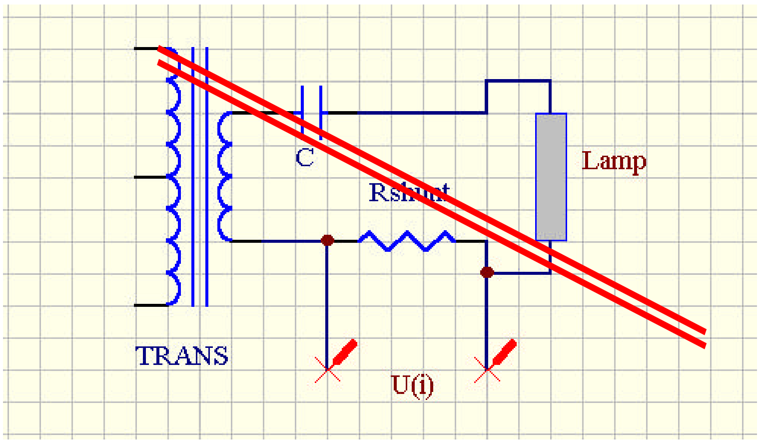


Figure 7.1 Wrong output-power measurement.

An important measurement, defined in most standards, is the luminous efficacy (Lumen/Watt$_{dc}$). Details and description of this measurement can be found in section 3.1.5. When some changes have been made in the design, a **relative** Lumen/Watt measurement will give the percentage increase in the luminous efficacy.

Within the time of this workshop it is not possible to make a long-term lifetime and switching durability test. One other possible test is the ability of the inverter to function without the lamp. This test normally takes one or two hours.

7.1.2 Modifications

Modifications are made with the knowledge already acquired and common sense. When the modifications are being tested, it often appears that the new design does not respond as expected. Inverters are balanced and sensitive devices. When one factor is changed, others will also be affected. In most cases some experimenting is required, in order to find the right balance and the right result.

In some cases the design is so poor, that optimisation of the circuit is not possible. For example, we can take a look at inverter set-up 1 in chapter 3.2.5. The basic set-up is clearly not ideal. The waveshape can never be made symmetrical, because of the use of only one transistor. The crestfactor might be reduced a little but so many components will have to be added to make the design **reliable under different circumstances** that a whole new set-up could be better.

7.2 Charge regulators

Second practical week

7.2.1 General

Within the last decade numerous different types of charge regulators have been developed. From very basic charge regulators, sometimes combined with a DC/AC-inverter or a solar lantern, to sophisticated digital devices with complex state of charge algorithms. This latter family of charge regulators will be manufacture by companies which have their own R&D department. When improvements are needed, they will rather improve it themselves than go to another institute and reveal all the details about their product. The companies that do need help, are those making the cheaper and simpler type of charge regulators. This type of device is mostly used in inexpensive Solar Home Systems for large-scale implementation in rural areas. Those manufacturers do not always have the knowledge themselves, so technical support can be very useful.

The keywords for the charge regulator are:

- safety
- reliability
- energy-saving

Solar Home Systems are often used under extreme conditions, in remote areas, or in fragile houses. Hence safety and reliability will be very important factors. New customers are not always used to living with electrical energy. Under these circumstances it is not surprising that problems can arise in the field. And when something *can* go wrong, it *will* go wrong sometimes. Because spare-parts are hard to get in remote areas, BOS-components have to be as reliable as possible.

Solar energy is relatively expensive, so unnecessary energy-loss should be reduced to a minimum. Once again, it can be stressed that over the lifetime of the system, a quiescent current reduction of 1 mA gives the customer 1.6 kWh of energy extra over the lifetime of the charge regulator. This has to be taken into account when modifications are made.

7.2.2 Measurements

Normally the features of the charge regulator are defined at first. Mostly the manufacturer does not send the complete documentation when he sends only a few test-samples. Before the testing starts it is best to know how the device will respond.

Normally in a laboratory, the battery and PV-module are simulated by power-supplies. Special attention has to be paid to the current limiters. They should not exceed 125% of the rated power of the charge regulator.

There are a few items of particular importance:

- Charge regulators can respond differently to power-supplies than they would do to a real battery and a real PV-module. Especially PWM-regulators could show a different behaviour. When some abnormal behaviour occurs, this has to be kept in mind.

- The overcurrent-protection test should be done **on one sample.** If the device fails, you still have the other samples to work with.

- **The most destructive tests have to be saved for last.** For this test the current limiter of the power-supply can be increased to 150% of the rated power of the charge regulator. Normally it should take action at 125%, so when the protection malfunctions there is a good chance that the regulator will not survive the test.

7.2.3 Modifications

When a charge regulator is functioning correctly according to the Standards, but only the quiescent current is too high, modifications need to be made for that aspect without influencing other functions.

However, a modification can sometimes be made for a whole part of the diagram. For example: a certain charge regulator has an ‘electronic fuse’ that makes use of many components. When another set-up of this ‘fuse’ can be created with far less components, it is worthwhile to try it out. Then you are making changes in the diagram with the objective of component reduction. The other features should, of course, not be influenced. These modifications might not always be necessary for passing the official tests. The manufacturer can be advised and decide whether the modification will be implemented or not.

7.3 DC/AC-converters

Third practical week

7.3.1 General

As described in section 5, there are many ways in which a DC/AC-converter can be arranged, not only in the way the power bridge is designed, but especially the steering principle. If it is an analogue design, improvements can be made easily. If the steering is based on a processor, using PWM techniques, it will not be possible to investigate the design within this one week of the workshop. The two reasons for this are:

- The processor technique itself is too complicated to be able to gain sufficient insight in its program, within the period of this training course.
- These types of devices are made by manufacturers, who do have good technicians available within the company. As indicated earlier, they are not likely to provide of their information, especially processor programs, to third parties.

This workshop focuses on the small, inexpensive inverters, most commonly used in the power range up to 500Watt.

7.3.2 Measurements at DC/AC-converters

When measuring at DC/AC-inverters, some important points have to be kept in mind. First of all:

1:

High Voltage alert !!

Whatever the measurement may be, use cables which are safe to touch! So no open endings or stripped wire endings. A mistake is easy made.

2: Never begin with the most destructive test. Just as for charge regulators, the short circuit test/ overload test at the output should be executed last.

3: Use short wires between the power source and the inverter which is being measured. Energy losses should be brought to a minimum. The connection wires have to be thick enough to handle 1.5 times the maximum rated current. It is preferable to use wires which are too thick rather than too thin. Not only because of the heat production, but also to ensure that a proper measurement is made. When the input voltage is being measured and the cables not suitable, voltage-drop may occur between the voltage metre and the inverter. When the efficiency is calculated, the input power (VA=Watt) will not be correctly measured.

7.3.3 Modifications

Testing a DC/AC-inverter can take place with a closed box. When modifications are being made, the box will have to be open. Before starting on any modifications, think very carefully to be sure that you have assessed all the implications. Think again very carefully before switching on the power. That also counts for FLI and Charge Regulators, but the consequences with DC/AC-inverters can be more severe.

The *modifications* to the electronics should be based on the same principle as set down for Fluorescent Lamps and Charge Regulators.

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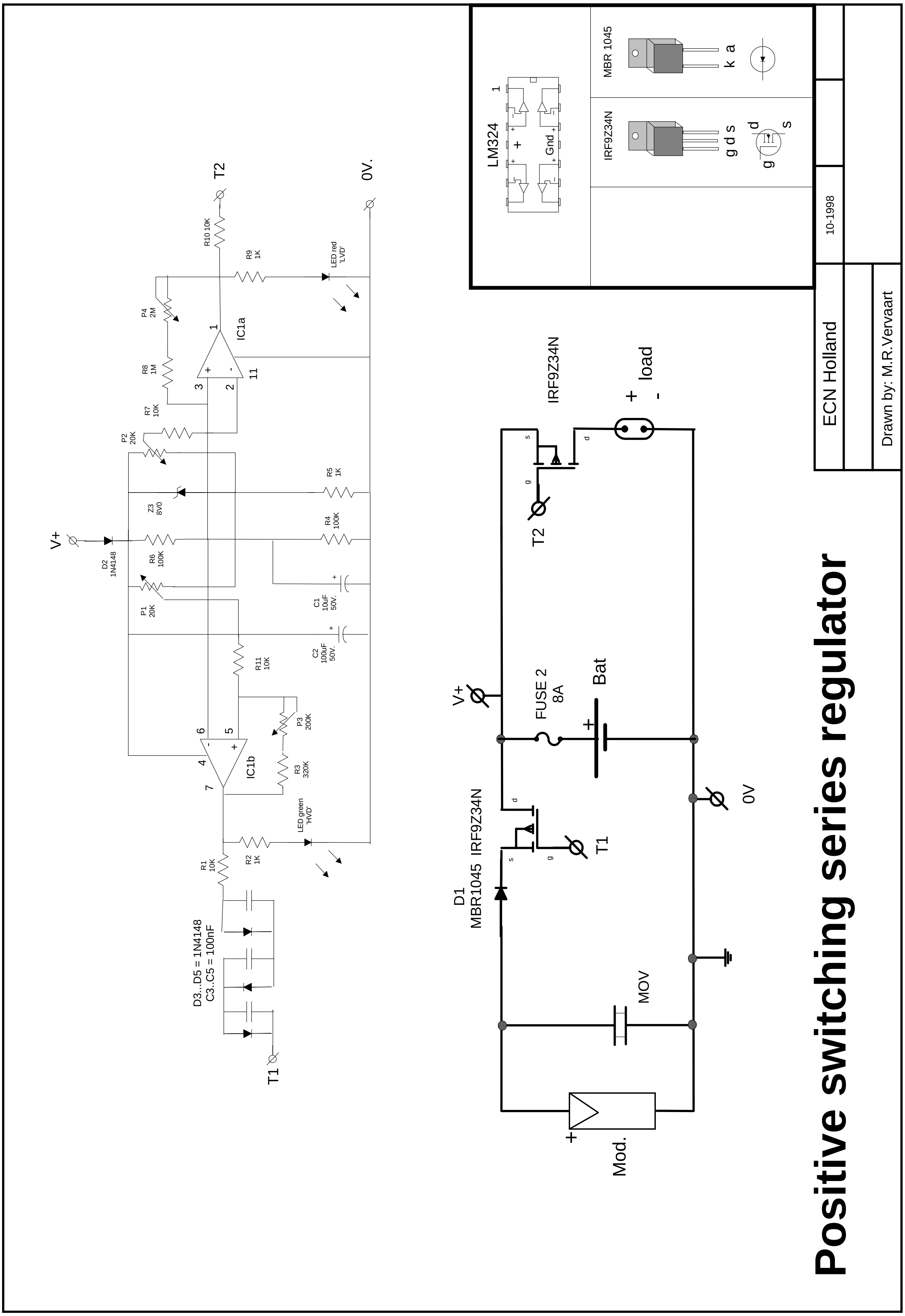
“Battery guide for small stand alone photovoltaic systems”, IEA PVPS Task III, by Bo Andersson, Catella Generics and Staffan Ulvonas

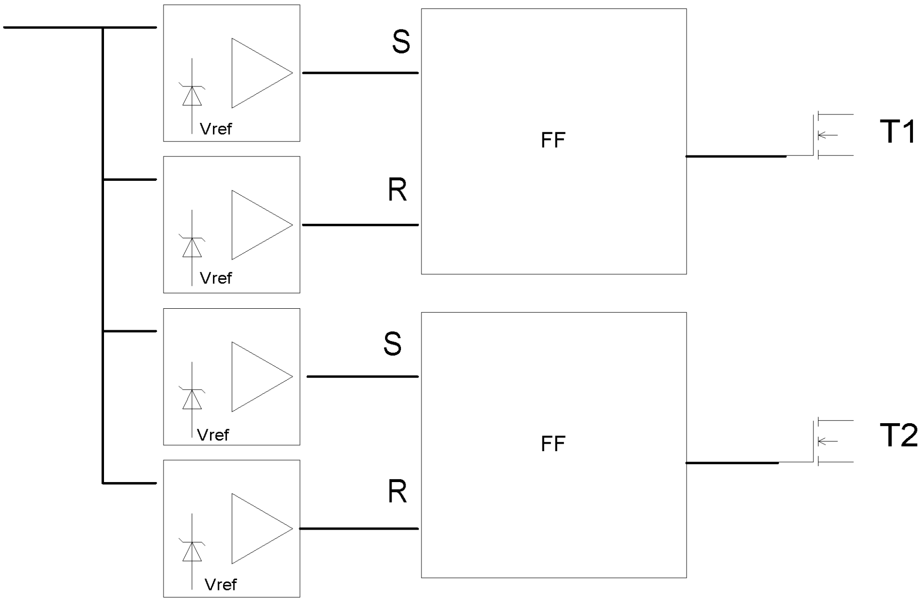
APPENDIX 1: POSITIVE SWITCHING CHARGE REGULATOR


Positive switching series regulator
ECN Holland
Drawn by: M.R.Vervaart
10-1998
LM324
1
+
Gnd
IRF9Z34N
g d s
d
g
s
MBR 1045
k a
T2
0V.
V+
T1
R10 10K
R9 1K
LED red 'LVD'
P4 2M
R8 1M
IC1a
1
3
2
11
R7 10K
P2 20K
R5 1K
Z3 8V0
R4 100K
D2 1N4148
R6 100K
P1 20K
C1 10uF 50V.
C2 100uF 50V.
R11 10K
P3 200K
6
5
4
7
IC1b
R3 320K
LED green 'HVD'
R1 10K
R2 1K
D3..D5 = 1N4148
C3..C5 = 100nF
IRF9Z34N
s
d
g
load
+
-
FUSE 2 8A
Bat
0V
D1 MBR1045
MOV
Mod.

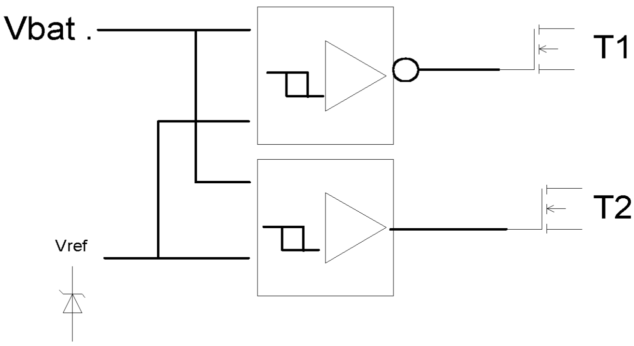
APPENDIX 2: BLOCK DIAGRAM MODIFIED CHARGE REGULATOR

Component reduction :

Actual diagram



Possible setup (exactly the same functions)



APPENDIX 3: DATA-SHEET IRFZ40

SGS-THOMSON MICROELECTRONICS

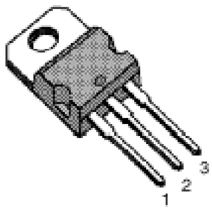
IRFZ40
IRFZ40FI

N - CHANNEL ENHANCEMENT MODE POWER MOS TRANSISTORS

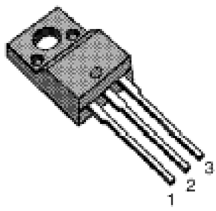
TYPE	V_{DSS}	$R_{DS(on)}$	I_D
IRFZ40	50 V	< 0.028 Ω	50 A
IRFZ40FI	50 V	< 0.028 Ω	27 A

- TYPICAL $R_{DS(on)}$ = 0.022 Ω
- AVALANCHE RUGGED TECHNOLOGY
- 100% AVALANCHE TESTED
- REPETITIVE AVALANCHE DATA AT 100°C
- LOW GATE CHARGE
- HIGH CURRENT CAPABILITY
- 175°C OPERATING TEMPERATURE

APPLICATIONS
- HIGH CURRENT, HIGH SPEED SWITCHING
- SOLENOID AND RELAY DRIVERS
- REGULATORS
- DC-DC & DC-AC CONVERTERS
- MOTOR CONTROL, AUDIO AMPLIFIERS
- AUTOMOTIVE ENVIRONMENT (INJECTION, ABS, AIR-BAG, LAMPDRIVERS, Etc.)

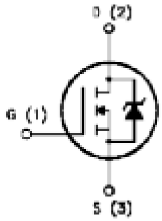


TO-220



ISOWATT220

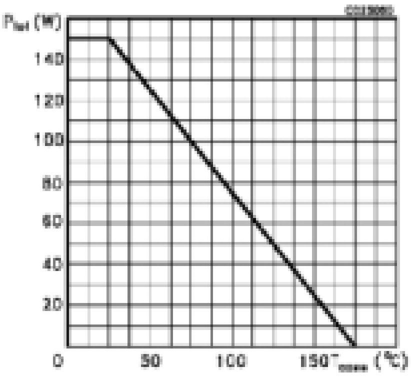
INTERNAL SCHEMATIC DIAGRAM

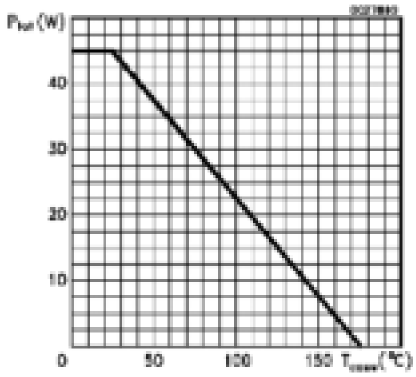


ABSOLUTE MAXIMUM RATINGS

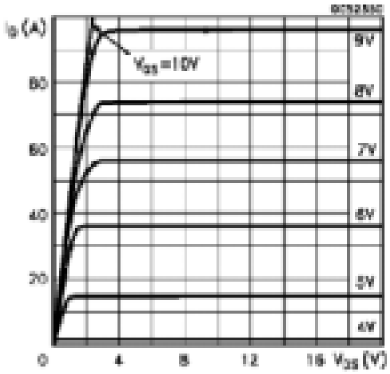
Symbol	Parameter	Value		Unit
		IRFZ40	IRFZ40FI	
V_{DS}	Drain-source Voltage (V_{GS} = 0)	50	50	V
V_{DGR}	Drain- gate Voltage (R_{GS} = 20 kΩ)	50	50	V
V_{GS}	Gate-source Voltage	± 20		V
I_D	Drain Current (cont.) at T_c = 25 °C	50	27	A
I_D	Drain Current (cont.) at T_c = 100 °C	35	19	A
I_{DM}(•)	Drain Current (pulsed)	200	200	A
P_{tot}	Total Dissipation at T_c = 25 °C	150	45	W
	Derating Factor	1	0.3	W/°C
V_{ISO}	Insulation Withstand Voltage (DC)	—	2000	V
T_{stg}	Storage Temperature	-65 to 175		°C
T_j	Max. Operating Junction Temperature	175		°C

(•) Pulse width limited by safe operating area

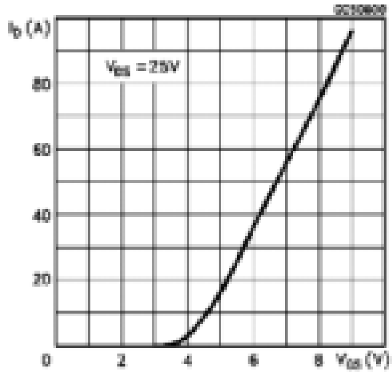




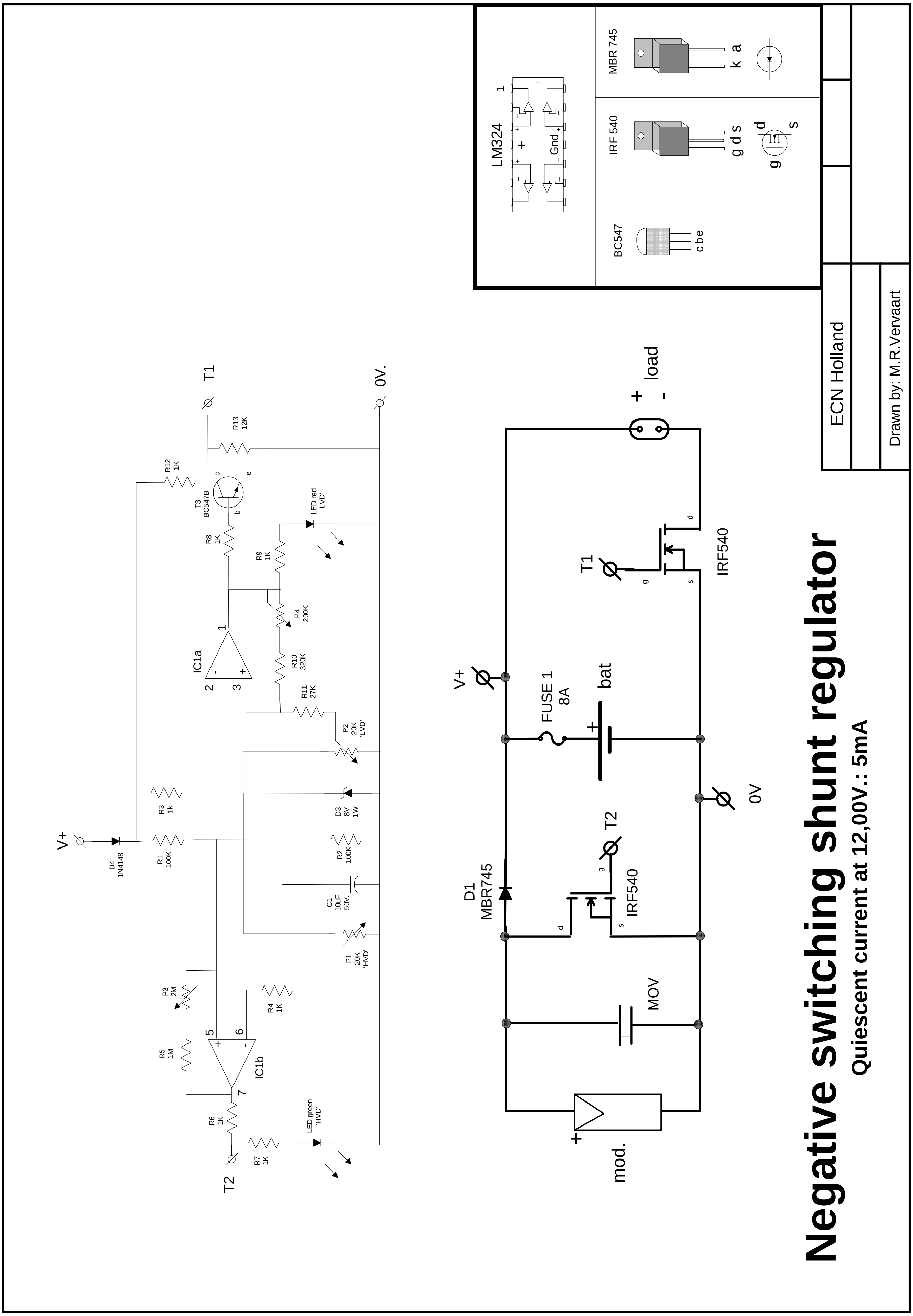
Output Characteristics



Transfer Characteristics

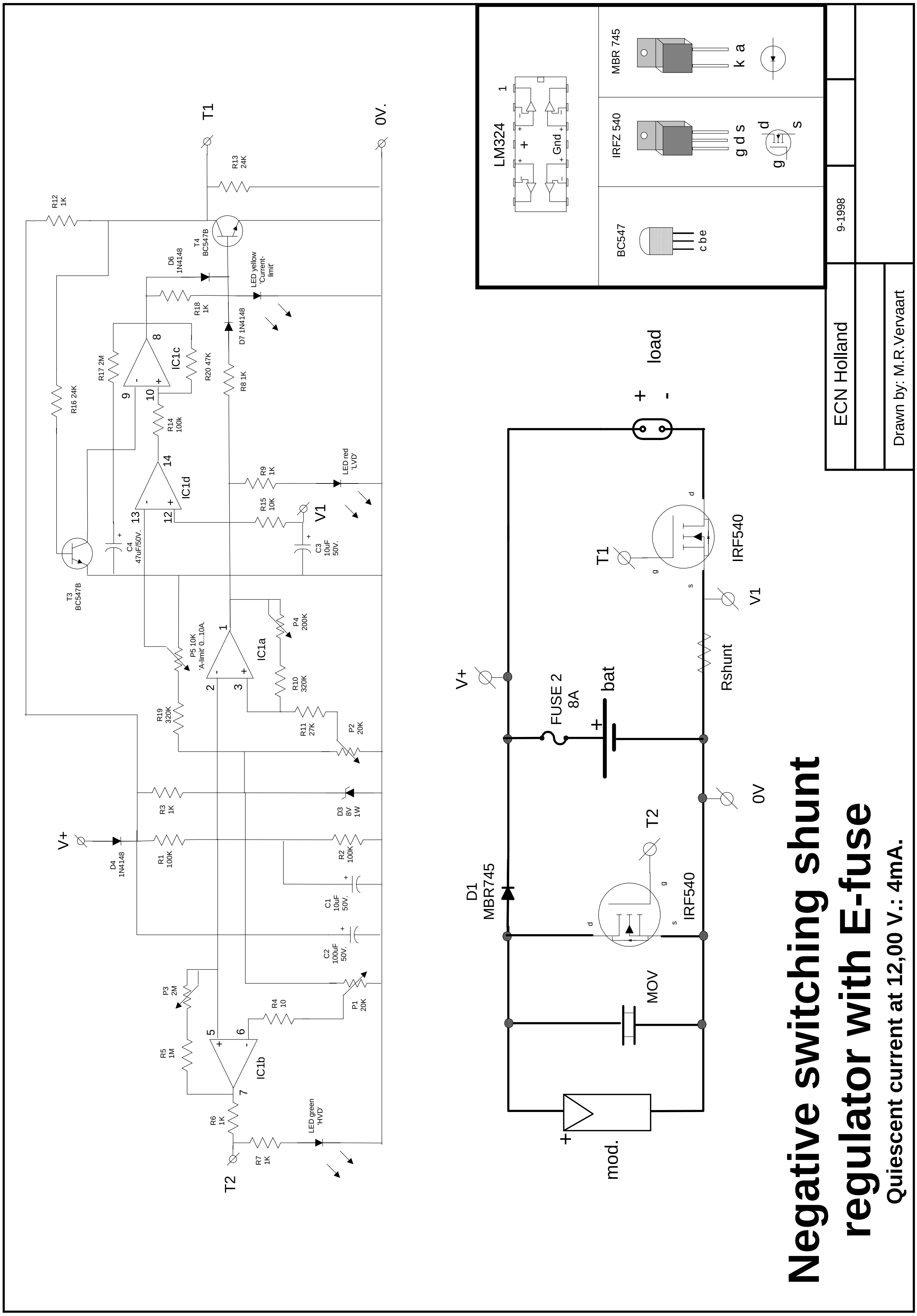


APPENDIX 4: NEGATIVE SWITCHING SHUNT REGULATOR

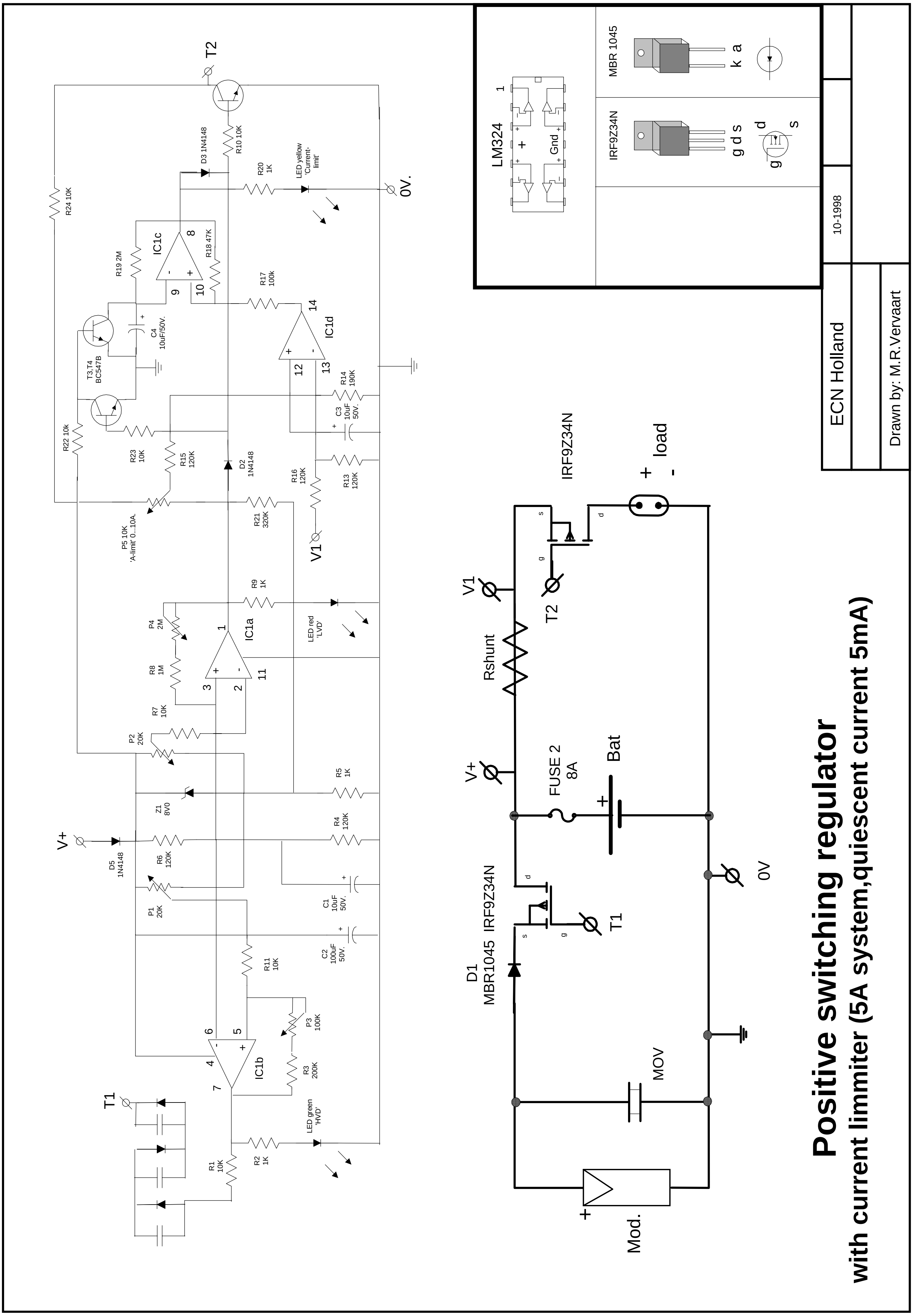


Negative switching shunt regulator
Quiescent current at 12,00V.: 5mA
ECN Holland
Drawn by: M.R.Vervaart
LM324
Gnd
BC547
c b e
IRF 540
g d s
MBR 745
k a
V+
D4 1N4148
R1 100K
R3 1k
R12 1K
R13 12K
T1
0V.
T3 BC547B
R8 1K
R9 1K
LED red 'LVD'
P4 200K
R10 320K
R11 27K
IC1a
P2 20K 'LVD'
D3 8V 1W
R2 100K
C1 10uF 50V.
P1 20K 'HVD'
R4 1K
P3 2M
R5 1M
IC1b
R6 1K
R7 1K
LED green 'HVD'
T2
load
IRF540
FUSE 1 8A
bat
0V
D1 MBR745
MOV
mod.

APPENDIX 5: NEGATIVE SWITCHING SHUNT REGULATOR WITH E-FUSE


Negative switching shunt regulator with E-fuse
Quiescent current at 12,00 V.: 4mA.
ECN Holland
Drawn by: M.R.Vervaart
9-1998
LM324
Gnd
BC547
c b e
IRFZ 540
g d s
MBR 745
k a
load
bat
FUSE 2
8A
Rshunt
MOV
mod.
D1
MBR745
IRF540
T1
T2
V+
V1
0V
0V.
IC1a
IC1b
IC1c
IC1d
T3
BC547B
T4
BC547B
D4
1N4148
D6
1N4148
D7 1N4148
D3
8V
1W
R1
100K
R2
100K
R3
1K
R4
10
R5
1M
R6
1K
R7
1K
R8 1K
R9
1K
R10
320K
R11
27K
R12
1K
R13
24K
R14
100k
R15
10K
R16 24K
R17 2M
R18
1K
R19
320K
R20 47K
P1
20K
P2
20K
P3
2M
P4
200K
P5 10K
'A-limit' 0...10A.
C1
10uF
50V.
C2
100uF
50V.
C3
10uF
50V.
C4
47uF/50V.
LED green
'HVD'
LED red
'LVD'
LED yellow
'Current-
limit'

APPENDIX 6: POSITIVE SWITCHING CHARGE REGULATOR WITH E-FUSE


Positive switching regulator
with current limmiter (5A system,quiescent current 5mA)
ECN Holland
10-1998
Drawn by: M.R.Vervaart
LM324
Gnd
IRF9Z34N
g d s
MBR 1045
k a
D1
MBR1045
IRF9Z34N
T1
T2
V1
V+
0V
Rshunt
FUSE 2
8A
Bat
load
MOV
Mod.
IC1a
IC1b
IC1c
IC1d
R1 10K
R2 1K
R3 200K
P3 100K
R11 10K
C2 100uF 50V.
P1 20K
C1 10uF 50V.
R6 120K
R4 120K
D5 1N4148
Z1 8V0
R5 1K
P2 20K
R7 10K
R8 1M
P4 2M
LED red "LVD"
LED green "HVD"
R9 1K
P5 10K 'A-limit' 0...10A.
R21 320K
R16 120K
R13 120K
R15 120K
R23 10K
R22 10k
T3,T4 BC547B
C4 10uF/50V.
R19 2M
R18 47K
R17 100k
R14 190K
C3 10uF 50V.
D2 1N4148
D3 1N4148
R10 10K
R20 1K
LED yellow 'Current-limit'
R24 10K
0V.

APPENDIX 7: INTRODUCTION TO LIGHT AND LIGHTING

Colour and wavelength

Visible light is electromagnetic radiation in wavelengths for which the human eye is sensitive, i.e. in the range of 400 to 800 nanometer[1]. Sensitivity is low for deep-blue coloured short wavelengths. It peaks in the green at about 555 nm, and is low in the deep red at around 800 nm. An approximation of the sensitivity curve of the human eye is presented in figure A1.1. It shows the relative sensitivity as a function of wavelength. The maximum of 100% is reached at a wavelength of 555 nm.

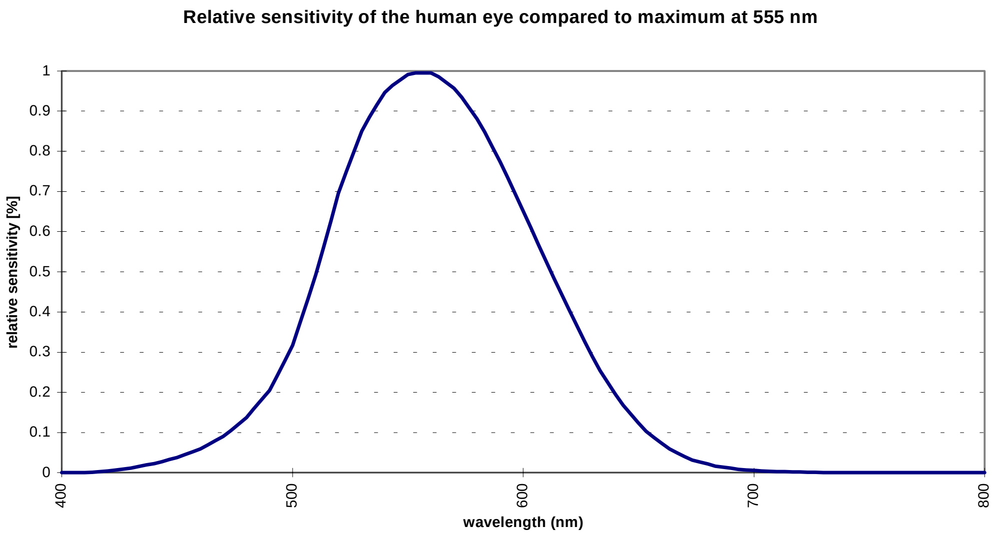


Figure A7.1 Normalised sensitivity curve of the human eye

Light output of a lamp: luminous flux

Since light is a form of energy, it is possible to express the light output of a lamp in units of power: Watt. However, this would give very little information about the visual impression of the light. To quantify the amount of visible light, the spectral intensity of the light (see figure A2.5)is multiplied with the eye sensitivity curve (see figure A1.1) and summed to obtain the luminous flux (Lu) in lumen (lm).

Efficiency of lamp or luminaire: Luminous efficacy

The efficiency with which a lamp produces light in the visible spectrum is defined as the luminous flux per unit of power: lumen per Watt.

Luminous intensity

The amount of light emitted in a certain direction is the luminous intensity (Li). It is defined as the luminous flux per steradian, with the unit candela (cd). Since there are 4 pi steradians in a

[1] One nanometer is one-billionth (10^{-9}) of a meter.

sphere, the relation between the average luminous intensity Li_{av} and the luminous flux of light source is as follows:

$$Lu = 4 * pi * Li_{av}$$

Illuminance

A measure of how well a surface is lit is presented by the illuminance. It is defined as the luminous flux per unit of surface area. The unit is lux. 1 lux = 1 lumen per square meter. The illuminance E at a point with a distance d from the light source can be expressed as a function of the angle a between the beam of light and the surface:

$$E = (Li/d^2)*sine(a)$$

Illuminous efficacy

Analogous to the luminous efficacy, the illuminous efficacy is a measure of the efficiency with which a light source illuminates a surface. It is presented in lux per Watt.

Correlated Colour Temperature

The colour appearances of the light source is expressed in the correlated colour temperature (CCT). When the CCT is below 3300 Kelvin, appearance of the light is reddish. This is called warm white. With a CCT higher than 5300 Kelvin the appearance of the light is bluish. An intermediate appearance occurs with a CCT between 3300 and 5300 Kelvin.

Colour rendering

The colour of an object appears to differ when illuminated by different light sources. Colour rendering is the degree to which the colours of surfaces illuminated by a given light source conform to those of the same surfaces under a reference[2]. It is presented as a percentage. Good colour rendering (80-90%) is achieved with incandescent lamps or fluorescent tubes with triphosphor coating. Moderate colour rendering (60-80%) is achieved with standard fluorescent tubes.

[2] Definition from: Rural Lighting, A guide for development workers, op. cit.

APPENDIX 8: EXAMPLE CHECKLIST

Test Report for Low Voltage DC Lights for Solar Home Systems

Project No.:	**Manufacturer:**
Client:	**Name of unit:**
	Type of Unit:
	Type Number :
	Serial number:
Tested by: M van Leeuwen	**Test Batch Number:**
	Date:
	Sheet **of**

Description of test

					Okay/Fail/ Value	Comments
1.General Inspection						
(C-G 8)	Visual Inspection					
:-G 9.1)	Funtional Test	DC range 10 Vdc -16Vdc				
		or Manufacturers Spec.				
2.4	Visual Inspection	Terminals, Safety Earth				
2.5	Visual Inspection	Creepage				
2.9	Visual Inspection	Accidental Contact to live parts				
2.Safety						
2.1 :-G 9.)	Insulation resistance	48hr @ 91%-95% RH, 20-30 Celsius	> 2 M Ohm			
2.2	Dielectric Strenght	50 Hz @ 500Vac	No Flashover			
2.3	Pulse Voltage	16Vdc + 3 x 16Vpeak,10ms				
2.7	Fault condition test					
Abnormal Conditions						
2.8.1	Lamp removal	1 Hour @ 16Vdc				
2.8.2	Lamp Failure	Filament resistor for 1 hr	(IEC 0081)			
2.8.3	Reverse Polarity	Blocking Diode				
3.Performance Tests						
3.1	Cold Start	12Vdc +/- 10% @ 10 up to 35 degrees Celsius				
	Operation	12Vdc +/- 10% @ 10 up to 50 degrees Celsius				
3.2	Open circuit lamp Voltage		@ 16Vdc	Vpk		
3.3	Lamp Current		@ 12Vdc	mA		
3.4	Inom.		@ 12Vdc	mA		
3.5	Max. Curent Electrode Supply.		@ 12Vdc	mA		
3.6	Lamp waveform @ I nom		@ 12Vdc			
3.7	see functional test					
			Vdc input		**A**	**Watt**
	Input power P @ 10Vdc		10			
3.8	Input power P @ 16Vdc		16			
3.9	Luminous Efficacy @ 2000 hr burnning period					
4.Endurance Test						
4.1	100 hr. burn-in Luminous Flux					
4.2	10,000 Switching Test	60 sec On,120 sec. Off				
5.EMC	See Separate test report					

Comments

Test Report for Low Voltage DC PV Charger Controllers for Solar Home Systems

Project No.:	**Manufacturer:**
Client:	**Name of unit:**
	Type of Unit:
	Type Number :
	Serial number:
	Number of samples:
Tested by:	**Test Batch Number:**
	Date:
	Sheet

Documentation — **Comments**

4.1.1, 4.1.2, 4.1.3

	Yes	No
* Installation, Operating and trouble shooting instructions		
* Waranty		
* Conditions of surroundings		
* Physical properties of the charge regulator		
* Electrical properties of the charge regulator		

6. Marking

* Maximum Module & Load Current		
* Nominal voltage		
* Polarity display		
* Labeling displays		
* Fuse values		

8.Electrical Parameters

	12V.	24V.
8.1 Nominal Voltage		

8.2 — **(PWM, Shunt or series)**

Working principle — Voltage control	
SOC algorithm	

	+	-
Switching polarity		

		Cenelec		Y	N
8.3 Switching thresholds	Imod=25%	HVD	2.30 V/cell		
		HVR	2.25 V/cell		
	Iload=25%	LVD	1.90 V/cell		
		LVR	2.10 V/cell		

8.5 Status Indicator Display	

Visual Inspection

9.	O.K.	Comments
Damage to the case, on terminals or displays		
Quality of case		
Quality of Electronic system		
Quality of wiring		

10.1 Function test

End of charge voltages

HVD	Imod. 6A	V
HVR		V

Iload:	0A	3A	4,0 A
LVD			
LVR			

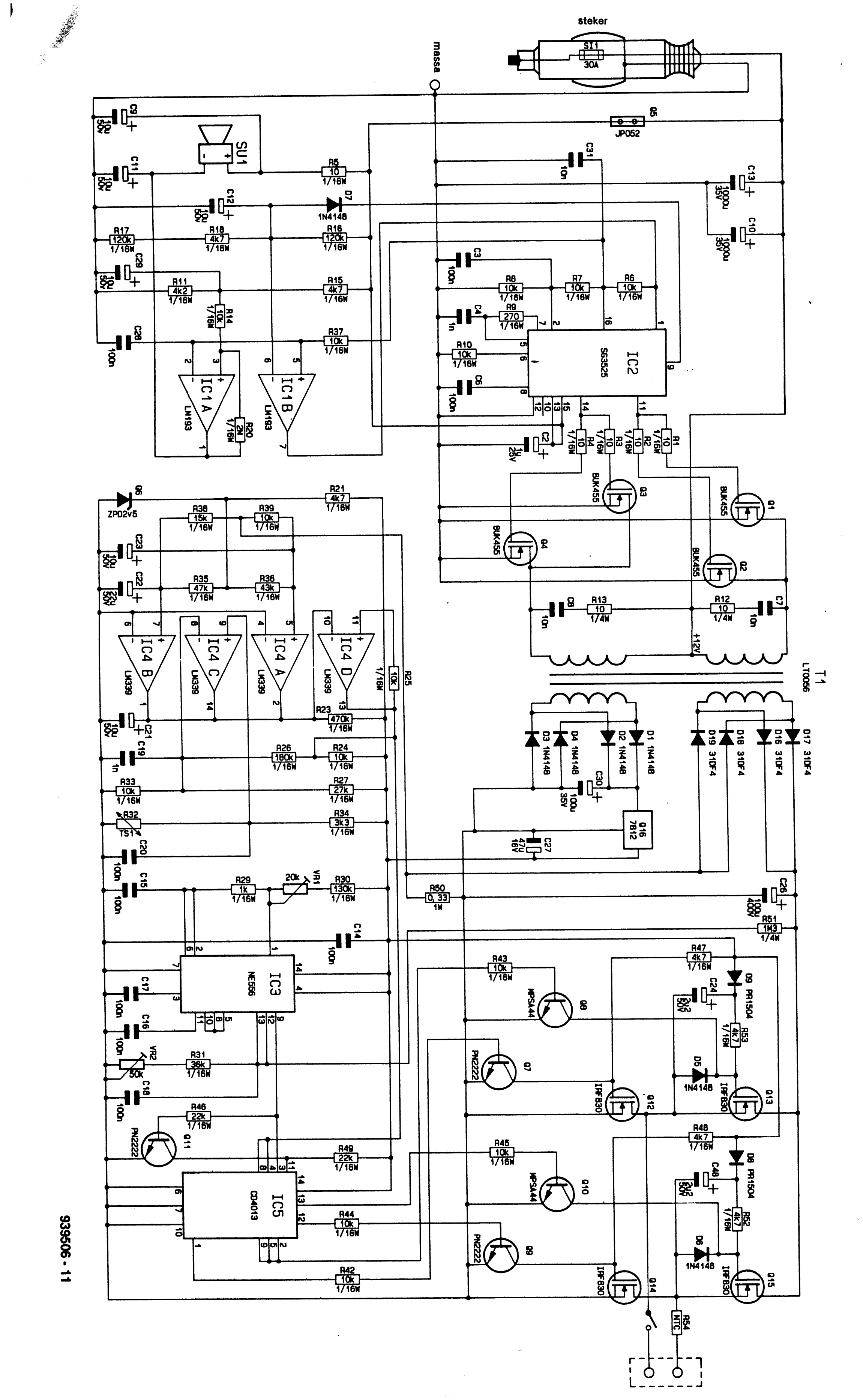
Gassing voltage	**(NP** = Not present)	np

Performance test — V at:

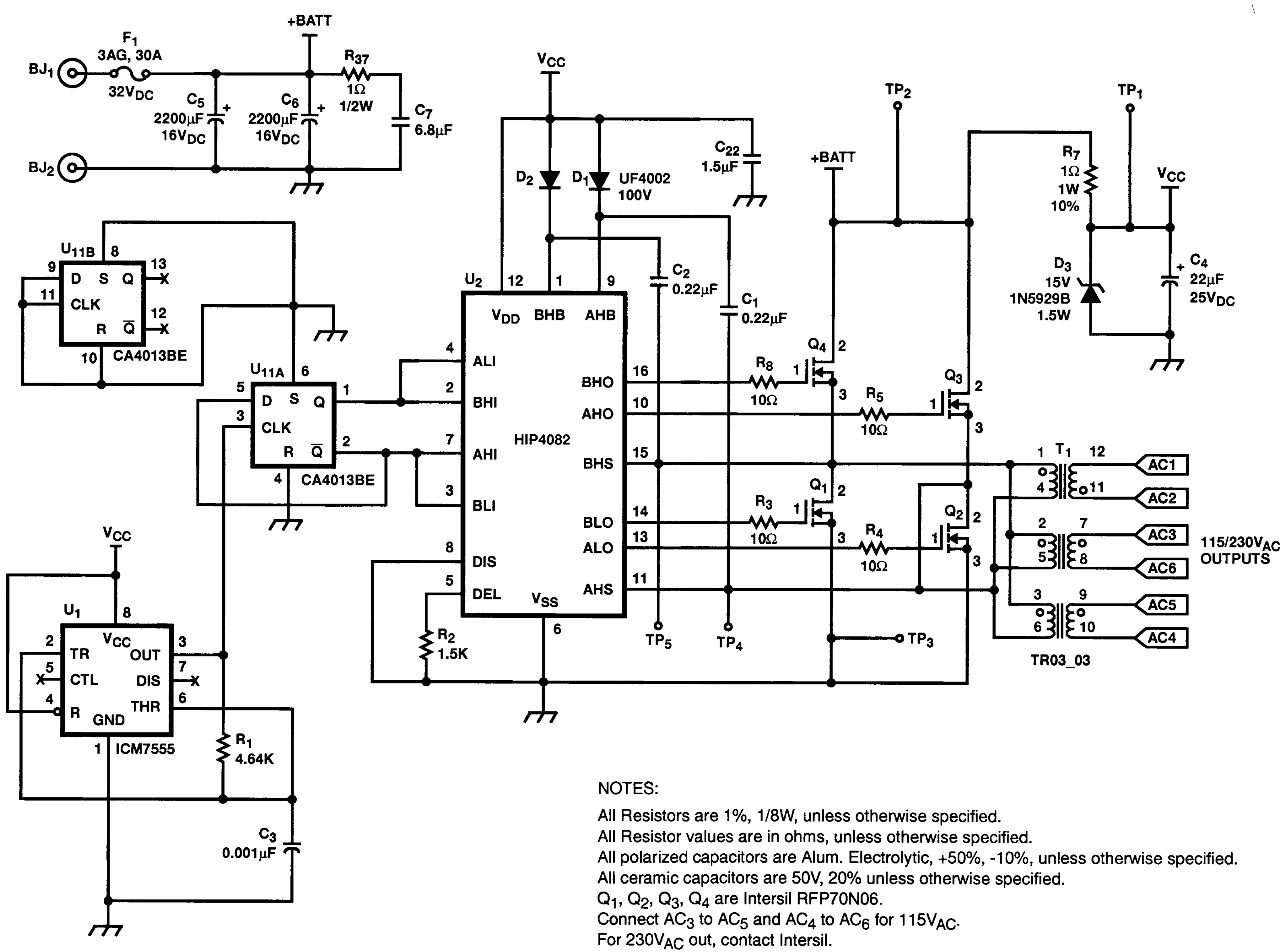
Quiescent current		mA.
Voltage drop Module-Battery	Imod. 6A	V.
Voltage drop Battery-Load	Iload 4.08 A	V.

	Y	N
Voltage drop compensation		
If yes, dV Battery terminal / Treshold controller < 100mV		

APPENDIX 9: 200VA HF FULL-BRIDGE INVERTER

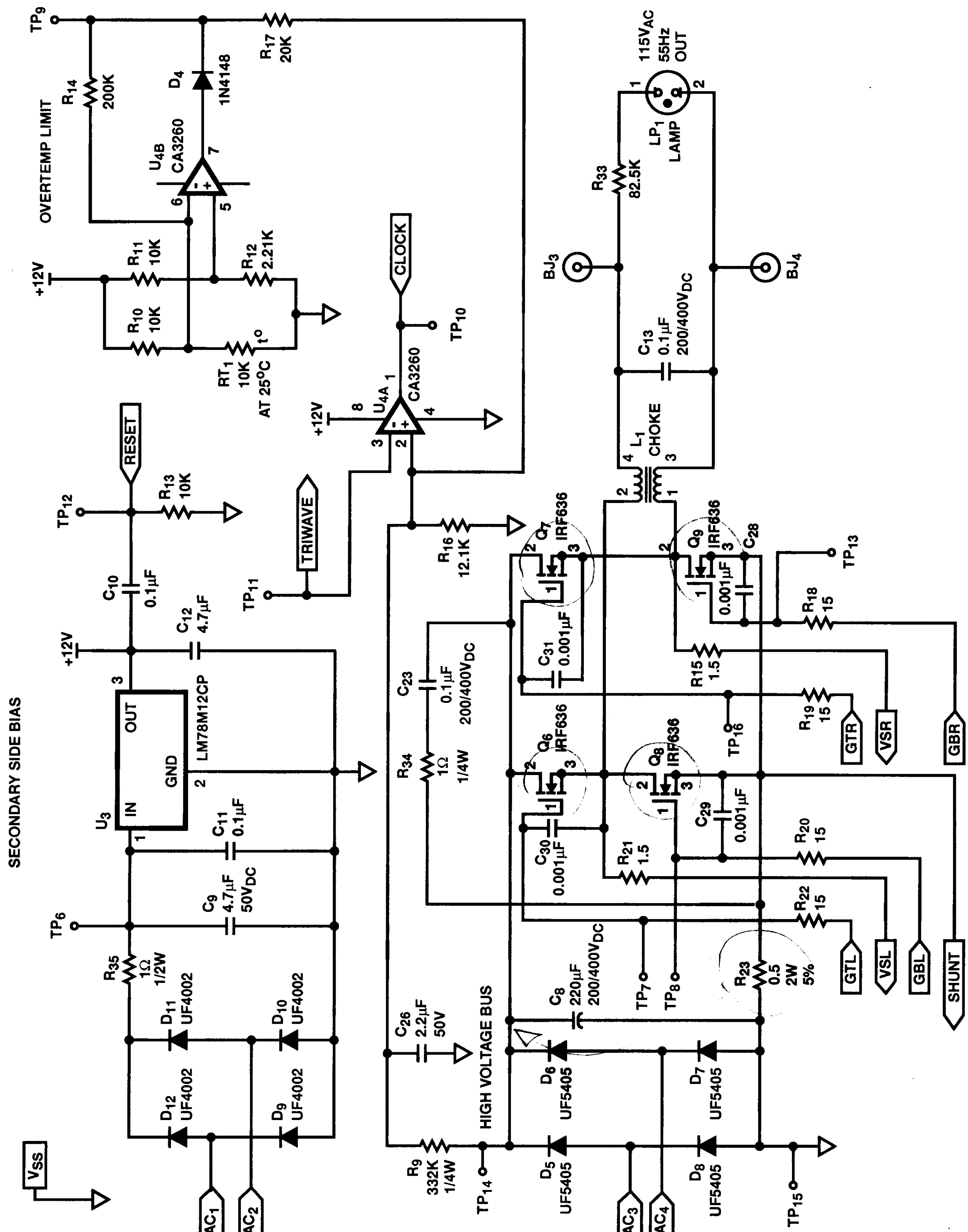

steker
massa
SU1
IC1A
IC1B
LM193
IC2
SG3525
IC4 A
IC4 B
IC4 C
IC4 D
LM339
IC3
NE556
IC5
CD4013
T1
LT0056
93506 - 11

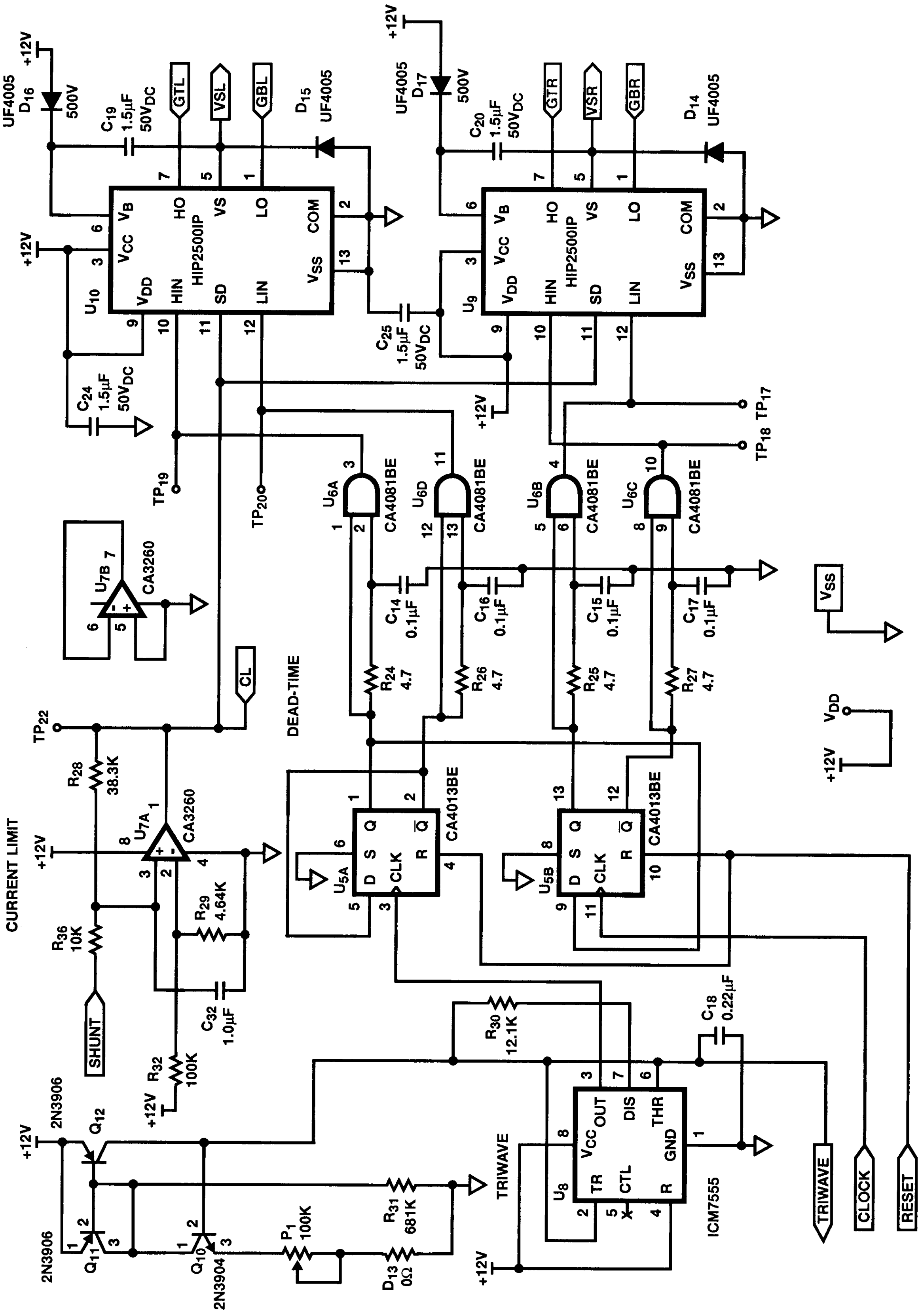
APPENDIX 10: HF FULL-BRIDGE INVERTER USING INTEGRATED CIRCUIT HIP4082



NOTES:

All Resistors are 1%, 1/8W, unless otherwise specified.
All Resistor values are in ohms, unless otherwise specified.
All polarized capacitors are Alum. Electrolytic, +50%, -10%, unless otherwise specified.
All ceramic capacitors are 50V, 20% unless otherwise specified.
Q_1, Q_2, Q_3, Q_4 are Intersil RFP70N06.
Connect AC_3 to AC_5 and AC_4 to AC_6 for $115V_{AC}$.
For $230V_{AC}$ out, contact Intersil.


SECONDARY SIDE BIAS
OVERTEMP LIMIT
HIGH VOLTAGE BUS
LM78M12CP
CA3260
1N4148
IRF636
UF5405
UF4002
CHOKE
LAMP
115VAC
55Hz
OUT
RESET
CLOCK
TRIWAVE
GTR
VSR
GBR
GTL
VSL
GBL
SHUNT
AC1
AC2
AC3
AC4
VSS


CURRENT LIMIT
DEAD-TIME
TRIWAVE
SHUNT
CL
CLOCK
RESET
GTL
VSL
GBL
GTR
VSR
GBR
HIP2500IP
CA4081BE
CA4013BE
CA3260
ICM7555
UF4005
2N3906
2N3904
+12V
500V