NTNU
Norwegian University of
Science and Technology

Power Supply for Down-hole Instrumentation and Actuators

Christian Eidsaune

Master of Science in Energy and Environment
Submission date: June 2009
Supervisor: Lars Einar Norum, ELKRAFT

Norwegian University of Science and Technology
Department of Electrical Power Engineering

Problem Description

Wireless communication with associated down-hole power generator, has long been a missing link for the ultimate smart well system. In this project we will address the technical challenges that must be met when the well does not have cables for power supply. In the proposed project we will focus on local electrical power generation and storage for a down-hole instrumentation unit.

The main objective of the project is to establish or develop reliable electrical power for a down-hole instrumentation unit, and develop the electronic for back up battery charging and voltage control.

A generator based on the induction principle, converting energy from the kinetic energy in the well stream to electric power is developed and patented.

The following main tasks will be initiated in the proposed project:

1. Design of power converters for voltage control and energy management system
2. Testing of the laboratory prototype electrical generator and high temperature electronics for storage and voltage control.

The project is initiated by industrial partners and a research project at SINTEF

Assignment given: 25. January 2009
Supervisor: Lars Einar Norum, ELKRAFT

PREFACE

This semester I have used much time to learn how to design SMPS', this together with reading and understanding how to implement processes for high temperature electronics. This has been done to research how an implementation of a high temperature instrumentation unit can be supplied with energy. This report comes with theoretical recommendations for how to build a high temperature converter, and I hope that you will have an interesting reading.

It has been an interesting semester, but it would have been more interesting if experimental testing of high temperature electronics have been done. The difficulties to apprehend large capacitors even though they are available, made the time disappear. What should have been done earlier was to order especially high temperature transistors, and testing their specifications. Then especially testing their voltage sharing capabilities, to check if it possible to connect the available high temperature transistors in series to achieve higher breakdown voltages.

I would like to thank all the people that have supported me throughout the last semester, and the time at NTNU.

Trondheim, June 29. 2009

Christian Eidsaune

SUMMARY

To create the ultimate wireless instrumentation unit for down-hole applications high temperature electronics with very high reliability is needed. It is possible to use ordinary bulk-CMOS devices at temperature up to 175 ^{0}C, but the lifetime at these temperatures is to low for a down-hole instrumentation unit. An alternative is to use s Silicon on Insulator process under the fabrication of the semiconductors. The SOI process is a fabrication process where there is buried a oxide layer in the silicon wafer, and thus allowing higher breakdown voltage and/or lower current leakage. The low current leakage allows the semiconductors to be used at higher junction temperature. SOI devices that are commercial available off-the-shelf as a expected lifetime for at least 5 years at 225 ^{0}C and thus much lower at junction temperatures below 200 ^{0}C.

The SOI technology can then be used together with hybrid circuits using ceramic substrate as a replacement for organic PCB and thick-film technology for the passive devices. A package like this gives a system with high reliability both toward high temperature operation and lifetime. The main limitation in the high temperature design is the availability off the larger capacitors; the limitation for high temperature stacked capacitors is 200 ^{0}C.

The converters designed are the standard step-up and step-down switch-mode power supplies. The converters are designed with current mode control; current mode control is used because of the advantage that comes with it. One off the advantages is the possibility to limit the inductor current; another advantage is the possibility to use constant current charging for the battery.

When designing the SOI devices for high temperature operation it is difficult to achieve high enough breakdown voltage. With this in mind, the high temperature converter is designed with series coupled transistors to achieve high enough breakdown voltage for high voltage operation. The transistors have always some small perturbations in their specifications, this has to be considered when connecting transistors in series. This perturbations in for example turn-off speed makes an uneven voltage sharing; this is solved by connecting suitable capacitors in parallel with the switches to maintain even voltage sharing.

TABLE OF CONTENTS

Appendices

Simulation files, MatLAB script and diagrams is attached in a zip file.

TABLE OF FIGURES

TABLE OF TABLES

ASIC – Application Specific Integrated Circuit

CCM – Continuous conduction mode

CMOS – Complementary MOS

DCM - Discontinuous conduction mode

IGBT – Insulated Gate Bipolar Transistor

MCM – Multi-chip module

MLC – Multi-layer ceramic

MOS – Metal Oxide Semiconductor

PWM – Pulse Width Modulation

SOAR – Safe operating area range

SOI – Silicon on Insulator

VLSI – Very-Large-Scale integration

1 INTRODUCTION

The oil and gas fields of today are approaching more inferior environments. The deeper the oil and gas fields is, the temperature and pressure rises drastically; as an example the Kristin gas field is that is at a depth of about 5000 meters, and has a temperature and pressure of 170 ^{0}C and 900bar(Statoil). In average the temperature rises with a geothermal gradient at 25-30 ^{0}C/km (Schlumberger); this shows the need of electronics that must withstand quite harsh environments as the fields get deeper.

A gas field like Kristin is going to produce for ten to twelve years; other fields again are delivering in up to 20years. This shows that the instrumentation equipment does not only have to cope with the high pressure and temperature, but they also have to have a very high reliability.

To get detailed knowledge of the oil field the operators want to know and control the individual flow of oil, water and gas from each source within the field; and at the same time monitoring temperature and pressure. One way to supply the instrumentation equipment in the well has been through a cable from topside, as the depth increases the cost of solving the problem by cable also increases. An alternative solution is therefore desired.

A stand-alone self supporting well system with wireless communication and high reliability is a missing link in the ultimate smart well system. This system has to have local electrical power generation and storage.

A prototype of a self-excited induction generator; that converts kinetic energy in the well stream to electric power, is developed and patented.

This thesis has looked at how a practical design can be implemented in a high temperature environment. The report is looking at the technical properties of the devices needed, and designing practical converters with their control circuitry.

FIGURE 2-1: HIGH TEMPERATURE INSTRUMENTATION (1)

2 HIGH TEMPERATURE ELECTRONICS

Traditionally silicon based integrated circuits have been used for temperatures up to 125 ^{0}C; Figure 2-1 shows the ranges for semiconductors for different applications. In the extreme conditions and harsh environments of oil & gas, aerospace and automotive applications, integrated circuits that can withstand higher temperatures are needed. Materials used today that can withstand these temperatures is with their belonging maximum junction temperatures CMOS/BiCMOS (<200 ^{0}C), epi-CMOS (<250 ^{0}C), SOI (<300 ^{0}C), GaAs, SiC and diamond.

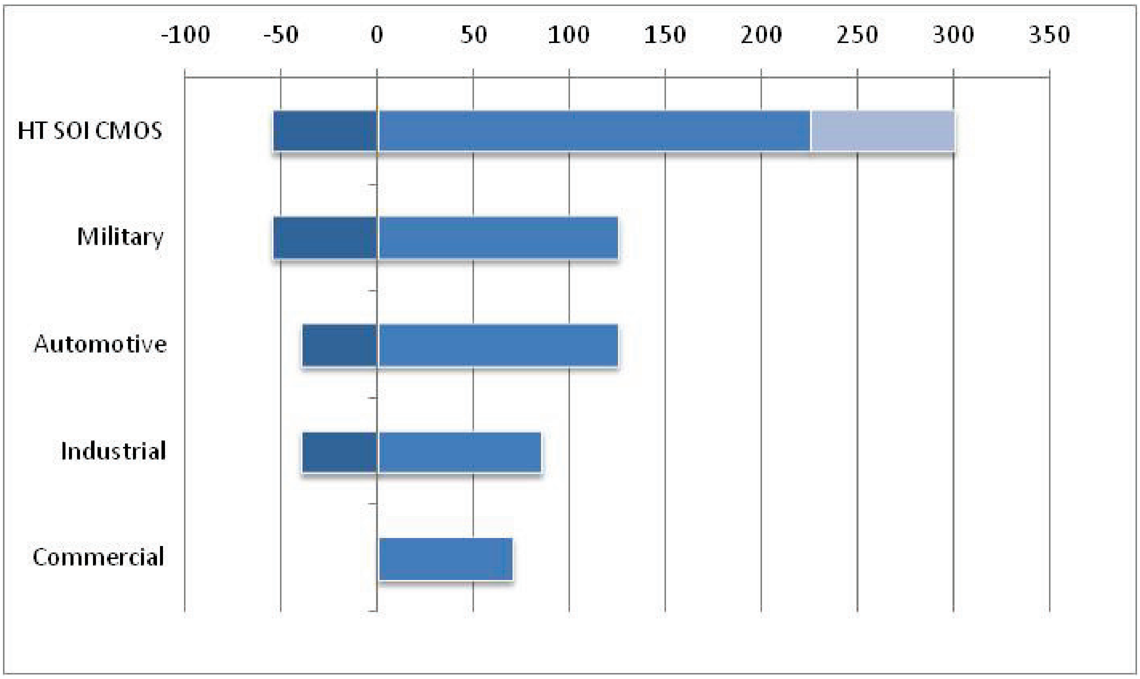


FIGURE 2-1 OPERATING TEMPERATURE RANGE FOR SEMICONDUCTORS

As for now temperatures above 175 ^{0}C are interesting for the application in hand. The CMOS/BiCMOS technology can withstand temperatures up to 200 ^{0}C but this is for a shorter lifetime. SINTEF with research partners has developed systems for monitoring and control of intelligent oil wells (1). This instrumentation equipment is supplied from topside, and is tested at temperatures at 200 ^{0}C, and are among others using BiCMOS HTASIC® modules. This shows the possibility to develop electronics for the desired environments.

For higher temperatures, reliability and power other solutions must be found, one interesting technology is the Silicon on Insulator technology. Silicon on Insulator has been researched for many years and is now a mature technology; this is one of the manufacturing strategies employed to allow the continuing miniaturization of VLSI devices.

For the last year off-the-shelf high temperature Silicon on Insulator semiconductor devices has been released, this mainly from CISSOID and Honeywell Aerospace. These components can operate at junction temperatures at 300 ^{0}C for a year with derated performance. This gives a boost in the reliability for the instrumentation units, and at the same time gives the possibility to create more powerful integrated circuits. The next step for a production process is the silicon carbide technology; this is not a process that is mature enough to be used for integrated circuits. Schottky diodes is one discrete semiconductor that is available with SiC technology, these are a good alternative to be used as rectifiers in the design.

2.1 SILICON ON INSULATOR

Silicon on Insulator is a variety of dielectric isolation method where an insulator material is “buried” in the silicon wafer. This gives several advantages compared to ordinary bulk-CMOS; some of these advantages are reduced parasitic device capacitance and increased breakdown voltage. The breakdown voltage increases with increased thickness of the buried insulator oxide.

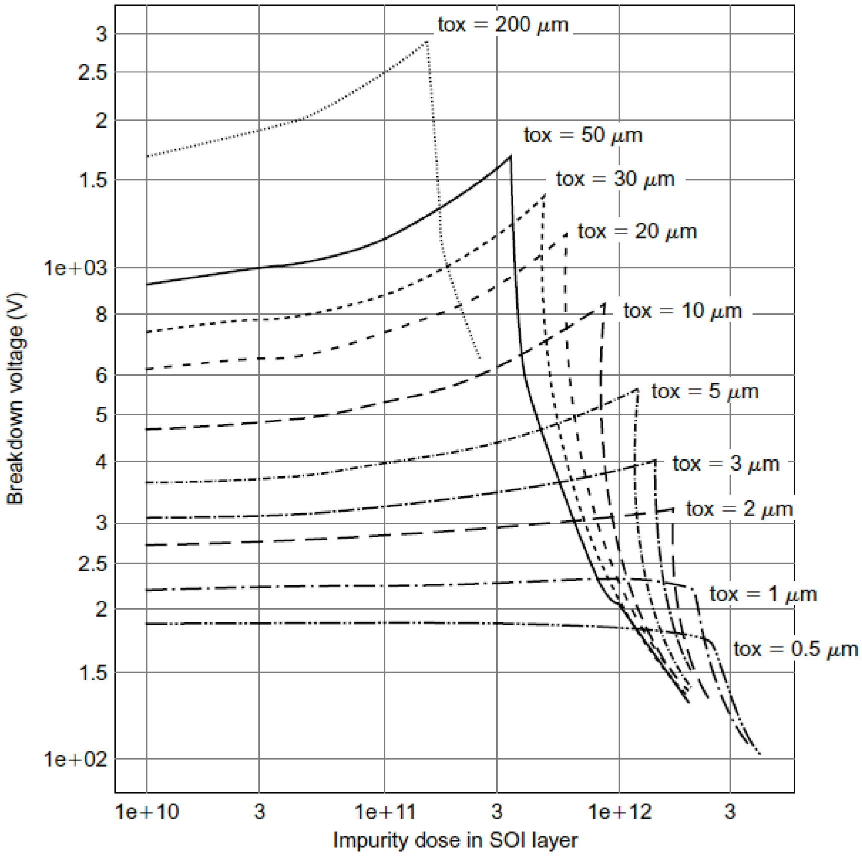


FIGURE 2-2: BREAKDOWN VOLTAGE VERSUS IMPURITY DOSE IN SOI LAYER WITH OXIDE LAYER THICKNESS (2)

The leakage current of SOI devices reduces as the SOI layer becomes thinner; the reduced leakage current gives the possibility to operate at high temperature, but it limits at the same time the maximum breakdown voltage. As long as the SOI layer thickness is kept high enough, the same CMOS fabrication process can be applied without modification. Leakage current versus SOI layer thickness is shown in Figure 2-3.

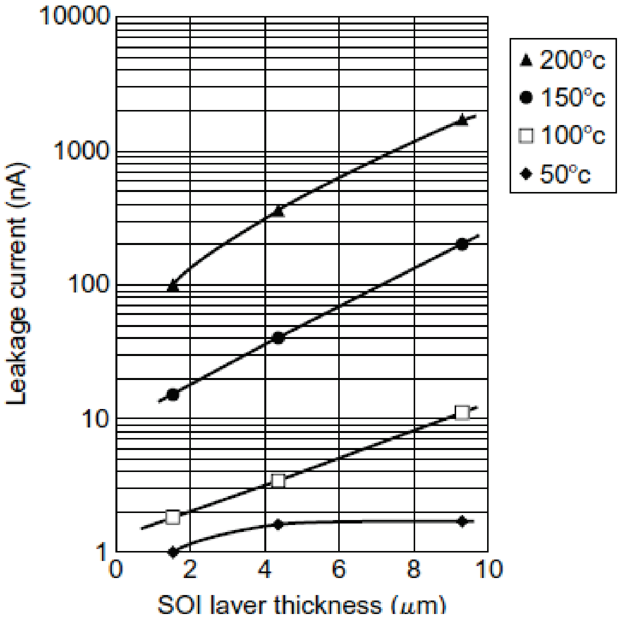


FIGURE 2-3: LEAKAGE CURRENT VERSUS SOI LAYER THICKNESS (2)

Using this technology together with VLSI, gives the possibility to design power integrated devices or at least ASICs with as much as possible of the circuit on one die. One of the biggest advantages with this approach is that the parasitic properties are minimized. A problem that might occur is to trench isolate the devices with the thin SOI layer that is needed for high temperature operation.

2.1.1 PRODUCTION PROCESS

X-FAB is one foundry group manufacturing silicon wafers for mixed-signal integrated circuits. They have four different SOI processes, with two different oxide thicknesses; 0.6µm and 1.0µm. An interesting process is the process with 1.0µm oxide thickness and a breakdown voltage at 90v designed for high temperature up to 225 ^{0}C.

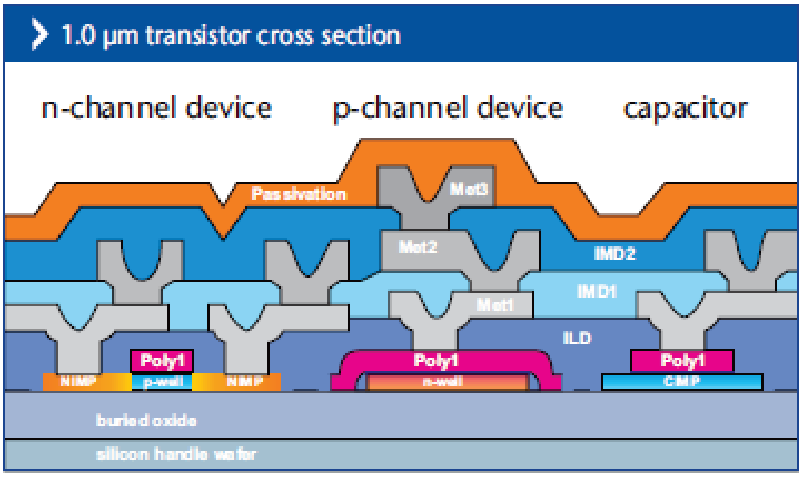


FIGURE 2-4: CROSS SECTION OF SOI TRANSISTOR USING 1.0µM SOI PROCESS (X-FAB)

2.1.2 METALLIZATION

The metallization is an essential part of the transistors, integrated circuits, multichip-modules, hybrids and printed circuit boards. In ordinary silicon integrated circuits and semiconductors the metallization is done by the use of aluminum-based alloys for the interconnection. The reliability of the aluminum-based alloys is limited by electromigration, stress-induced voiding and corrosion process. Thus by replacing the aluminum-based alloys with other materials with higher immunity towards electromigration, a higher lifetime for the device is found. The rate of electromigration is decided by several factors, where one of them is temperature. Thus replacing the aluminum alloy is necessary for increased lifetime for high temperature operation. There are several other metals that can be used for such purposes where higher reliability is needed, among these is gold, copper tungsten and many others. Not every metal is useful for such application even though they have better conductivity; one example for this is copper. With the use of copper as interconnection in integrated circuits effective diffusion barriers is needed to prevent the copper to diffuse into the silicon material.

One metallization method that is used by CISSOID is to use tungsten for interconnection for increased lifetime. Several other methods are found in (3).

2.2 HIGH TEMPERATURE COMPONENTS

2.2.1 INTEGRATED SEMICONDUCTOR

There are mainly two semiconductor companies that deliver off-the-shelf integrated semiconductors; both have an expected lifetime for at least five years at 225 ^{0}C for their devices. These are CISSOID and Honeywell, and their standard product family is shown in Table 2-1.

TABLE 2-1: HIGH TEMPERATURE STANDARD PRODUCTS

Company	Standard devices
CISSOID	AD converter Operational amplifiers Full bridge MOSFET driver Logic family MOSFET Oscillators and 555 timer Voltage reference Linear voltage regulators
Honeywell	Operational amplifiers Linear voltage regulators Analogue switch NFET Oscillator Analog multiplexer Static RAM 83C51 microcontroller

The transistors are the most interesting component in their portfolio, since the main part of a switch-mode power supply is the switch. These have to be able to switch quite high currents and withstand high voltages and at the same time have high reliability. Their characteristics from the datasheets is represented in Table 2-2

TABLE 2-2: TRANSISTOR CHARACTERISTICS (4)(5)

	CISSOID - NMOS80	Honeywell – HTNFET
Electrical characteristics		
Drain-source breakdown voltage [V]	80	55
On-state resistance [Ω]	0.9 (225⁰C)	0.4(25⁰C)
Gate leakage current	170pA	100nA
Dynamic characteristics		
Input capacitance [pF]	430	290
Output capacitance [pF]	60	87
Feedback capacitance [pF]	12	14
Switching characteristics		
Turn-on delay time [ns]	140	10
Rise time [ns]	330	20
Turn-off delay time [ns]	45	64
Fall time [ns]	100	20
Drain current [A]	3.4 (225⁰C)	Up to 1
Thermal characteristics		
Thermal resistance [⁰C/W]	5	

The prices for these devices are available at request at the companies respectively; at least CISSOID delivers up to 5 samples of each of their devices. CISSOID delivers a reference design (4) for a step-down regulator, the pricing for the components in the reference design was in April 2009 1115.03€.

2.2.2 PASSIVE DEVICES

The passive devices is components such as resistors, capacitors and inductors are a necessity also in high temperature applications, some of these provide a limitation factor in design of high-temperature electronics. Especially the unavailability of certain components such as high capacity capacitors is a high limitation factor. A common feature is that all the passive components must be derated for high temperature application.

The resistors are available for high temperature application, discrete leaded resistors have been available for numerous of years for long term high temperature operation up to 300 ^{0}C. Thick- and thin-film resistors either incorporated in hybrid circuits (as discussed later), or as discrete chip versions are also used for high temperature application. Especially in hybrid circuits thick-film resistors are a good choice because of the precision in resistance values.

When it comes to transformers and inductors, which is the main part of a switch-mode power supply, special considerations regarding windings and core material must be taken. The limitation for inductors and transformers is typically because of the wire insulation since the wire conductor and core can survive a higher temperature. For moderately high temperature ordinary organic insulation can be used, but for increased temperatures over 250 ^{0}C the insulation must be replaced by inorganic materials. Table 2-3 shows examples for materials used for conductors and insulation at different temperature ranges.

TABLE 2-3: MATERIALS FOR WIRES, CABLES AND CONNECTORS (3)

Temperature range [^{0}C]	Conductors	Insulation
Low, ≤200-250	Copper, aluminium	Organics: high performance “engineering” plastics (polyimide), fluorpolymers (Teflon®)
Moderate, 250-500	Aluminium, Ag-plated Cu, Ni-plated Cu, stainless-steel-clad Cu, Ag	Inorganics: fibreglass, glass, mica, asbestos, silica, magnesium oxide, aluminium oxide, and other ceramics
High, 500-1000	Au, Pt, or stainless steel	

The magnetic portion of inductors and transformers must also be considered since the core material decides the behaviour. One fundamental requirement is that the Curie temperature for the core material is at a much higher temperature than the operation temperature. The Curie temperature is the temperature where the ferromagnetic material loses its ferromagnetism. Most of the ferromagnetic materials used have a Curie temperature much higher than 200 ^{0}C, so the ferromagnetic materials should not be a limitation to the inductor design.

Inductors rated for operation at 200 ^{0}C is found off-the-shelf at some distributors, one of these are Datatronics Distrubution Inc. To get an inductor that is best suited for the application, inductors can easily be designed by the developer.

When it comes to the last passive device the one with the highest limitation is found. High capacity capacitors with high voltage rating are a missing link for high temperature applications. Capacitors for power-conversion circuitry must maintain a low ac loss and dc leakage at high temperature. Typically the high capacity capacitors are found in electrolytic types, the electrolyte inherence a very big limitation in temperature range. Therefore inorganic capacitors such as ceramic capacitors are required for high temperature operation; ceramic multilayer capacitors have been tested at 200 ^{0}C for thousands of hours (3). In hybrid circuits thick- and thin-film capacitors have good reliability. There is one drawback by using inorganic capacitors and that is their low volumetric efficiency compared to electrolytic capacitors.

In switch-mode power supplies "large" capacitors must be used to minimize the voltage ripple at the output of the converter. In order to achieve capacitors with high enough capacity and voltage rating the ceramic multilayer capacitors must be paralleled and coupled in series. AVX delivers stacked MLC capacitors intended for switch-mode power supplies that are rated for operation at up to 200 ^{0}C (6). Even though there are drawbacks by using ceramics capacitor there are benefits by very low ESR and dc leakage.

2.3 MULTICHIP MODULES AND HYBRIDS

Multichip modules, MCM, are a technique where the semiconductor dies or other modules are package so it is possible to use the chip as a single integrated circuit. By using the semiconductor dies, the limitations that are with the SOI process can be bypassed and it will be able to miniaturize the circuit. This gives lower parasitic properties compared to using PCB.

For temperatures above 175 ^{0}C organic PCB and standard surface mount technology passive components renders a limitation. A solution for this is to use a hybrid integrated circuit. A hybrid integrated circuit is a circuit using ordinary semiconductors on a ceramic substrate material using thick-film technology instead of an organic PCB with SMT passives.

2.4 ENERGY STORAGE

High temperature rechargeable batteries for high temperature operation are a missing, but battery technologies are under a large development focus in general. There are some providers that can provide non-rechargeable batteries for high temperature applications; among these are Energychem and SAFT.

The battery chemistry is the limitation for high temperature applications. High temperature battery can be designed with an electrolyte that is solid for normal room temperature, but are aqueous for higher temperatures. This is then allowing batteries to be operated at increased temperatures.

In this report it is assumed that an eventual high temperature rechargeable battery is of a lithium-ion type, and thus requires constant current charging. Typically open circuit cell voltages for lithium-ion based cells are 3.7V.

2.5 MECHANICAL ISSUES

2.5.1 SOLDERING

Ordinary unleaded soldering that is used for consumer electronics has a melting point at 217^0C; this temperature is to close to the operating temperature for high temperature electronics to render possible to use. Even though hybrid circuits are being used, the active components must be attached to the ceramic substrate and interconnections.

This can be solved by choosing soldering materials and adhesive materials that can be used at higher temperature by adding materials with higher melting point to the solder, such as silver. This type of soldering is called high melting point soldering and uses soldering materials with melting temperature around 300 ^{0}C.

2.5.2 PACKAGING

The system must be packed and hermetically sealed to provide protection from mechanical damage and the harsh environment. One thing that has a great impact on the lifetime for the circuit is moisture and various gases, therefore is the hermetically sealing imperative.

3 CONVERTER

The power supply must implement several different converter topologies. The converter has to charge a battery, supply external equipment, and at the same time make sure that everything can be supplied with enough energy.

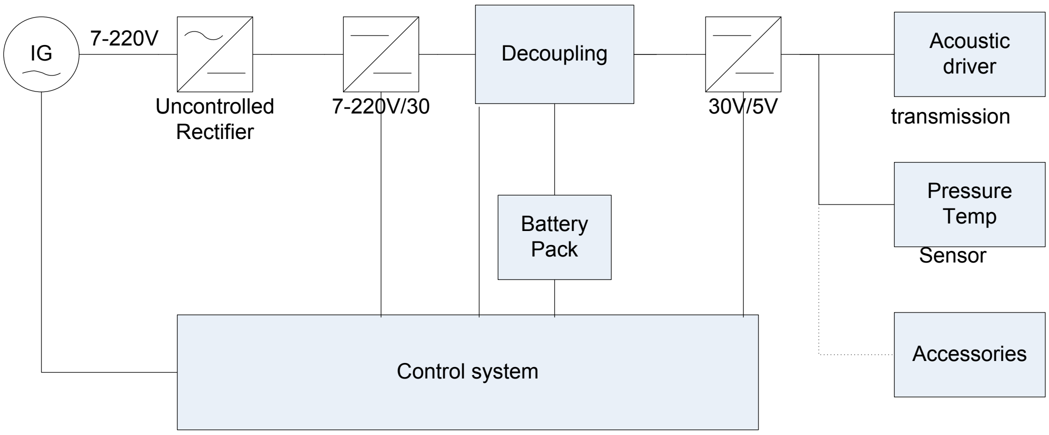


FIGURE 3-1: SYSTEM TOPOLOGY

The low side converter is the easiest converter to design. This is a low power step down converter whose biggest load is when transmitting data upstream. The biggest challenge lies in the high side converter; this converter copes with higher voltages and large variations in input voltage. The proposed characteristics are shown in Table 3-1.

TABLE 3-1: ASSUMED CONVERTER SPECIFICATIONS

Minimum generator voltage [V]	7
Maximum generator voltage [V]	200
Maximum output power [W]	10
Minimum output power [W]	3
Battery voltage [V]	25
Charging voltage [V]	30
Load voltage [V]	5

The energy source in the proposed well system is a self-excited induction generator that converts kinetic energy in the well stream to electric power.

3.1 AC/DC CONVERTER

3.1.1 UNCONTROLLED RECTIFIER

Since the induction generator is self-excited there is no need to be able to deliver power to the induction machine. This makes rectifying the alternating voltage from the induction generator easy by the use of an uncontrolled rectifier. The uncontrolled rectifier used is a three-phase, full-bridge rectifier as shown in Figure 3-2.

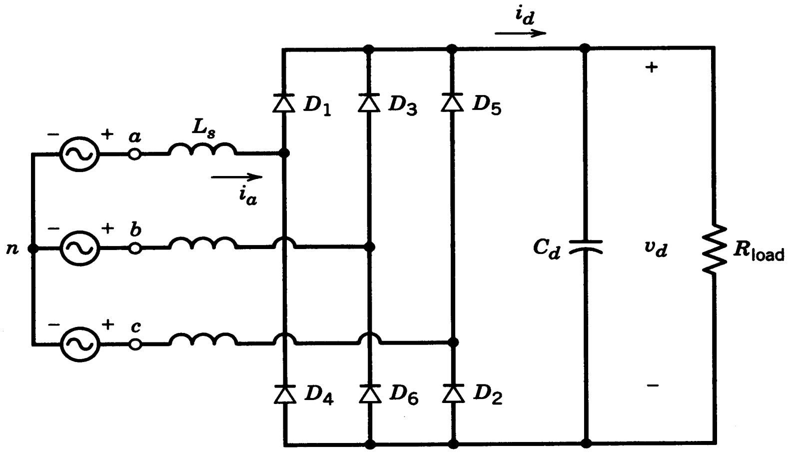


FIGURE 3-2: THREE PHASE FULL BRIDGE DIODE RECTIFIER (7)

The dc-link voltage is

$$V_d = 1.35 V_{LL} - \frac{3}{\pi} \omega L_s I_d \tag{3-1}$$

3.2 DC/DC CONVERTER

To be able to have numerous controlled voltage levels converted from the varying generator voltage, several converter topologies must be implemented. Characteristics of the converter:

- Step up and down the input voltage from the generator.
- Step down the voltage from battery link.
- Constant current constant voltage battery charging (Li-ion).
- Constant voltage to load.

3.2.1 STEP-DOWN CONVERTER

The step-down converter have a output voltage that are lower than the input voltage. The main type of step down converter is the buck converter; where the output voltage is given ideally by (4-1):

$$V_o = D \cdot V_d \tag{3-2}$$

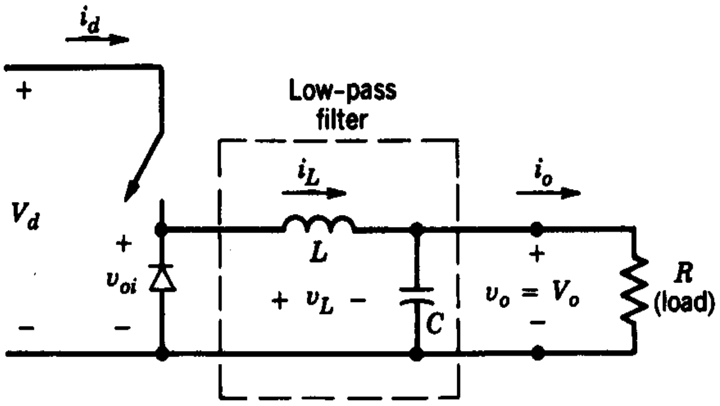


FIGURE 3-3: STEP-DOWN DC-DC CONVERTER (7)

The step-down converter, Figure 3-3, function is when the switch is operated; the inductor tries to withstand the change in current. The amount of energy stored in the inductor is decided by the switch duty ratio. This energy stored in the inductor is discharged when the switch opens; this creates an average output voltage lower than the input voltage. The output capacitor is needed to be sufficient to keep the output voltage ripple minimized.

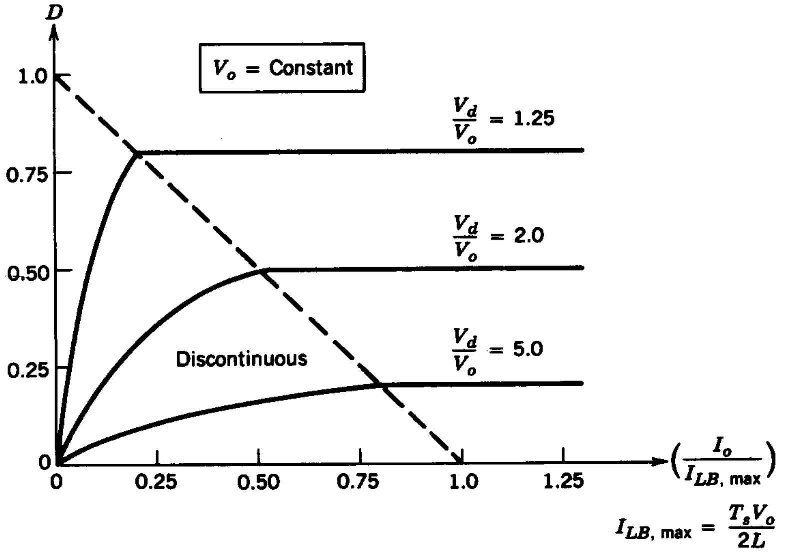


FIGURE 3-4: STEP-DOWN CONVERTER CHARACTERISTICS (7)

The step-down converter has two operating modes; discontinuous conduction mode (DCM) and continuous conduction mode (CCM). These modes tell whether the current through the inductor L flows continuously or discontinuously. For a step down converter it is desired that the current always flow continuously, to keep the peak current minimized and to avoid low voltage blackout. Figure 3-4 shows the border between DCM and CCM for a constant output voltage.

To be sure that the converter is operating in CCM the inductance of the inductor has to sufficient to be able to supply the output continuously also when the switch is closed. This boundary is given by:

$$I_{LB} = \frac{1}{2} i_{L,peak} = \frac{t_{on}}{2L}\left(V_d - V_o\right) = \frac{DT_s}{2L}\left(V_d - V_o\right) = I_{oB} \qquad (3\text{-}3)$$

The inductor current ripple is reduced by increasing the inductance.

A synchronous buck topology can easily be implemented with a slightly better efficiency. This can be done by replacing the diode with a transistor with a lower on-state resistance. Thus this topology has to be implemented in such a way that the switches are never on at the same time. Non-overlap logic is found slightly more efficient to use compared to implementing dead band to give the switches time to turn-off.

3.2.2 STEP-UP CONVERTER

For situations where the input voltage is lower than the desired output voltage, a step-up converter has to be implemented. The main configuration is shown in Figure 3-5 and is called boost converter, for the boost converter operating at CCM the output voltage is given in equation 3-4.

$$V_o = (1-D)^{-1} \cdot V_d \tag{3-4}$$

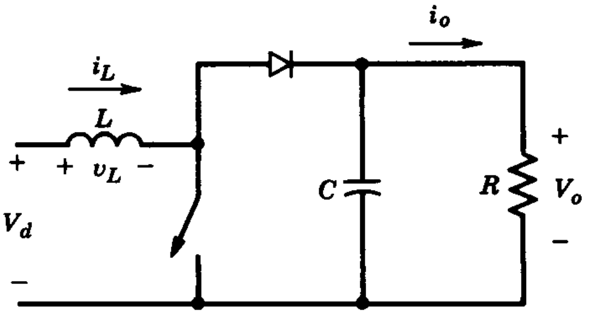


FIGURE 3-5: STEP-UP DC-DC CONVERTER (7)

The function of the boost converter is such that when the switch is turned on the diode gets reversed biased, and thus isolating the output stage so the inductor choke is supplied with energy. When the switch is turned off, the extra energy stored in the inductor flows to the output, and thus the output voltage gets a boost from the inductor voltage. As a temporary "storage unit" the capacitor is used to minimize the voltage ripple to approved limits.

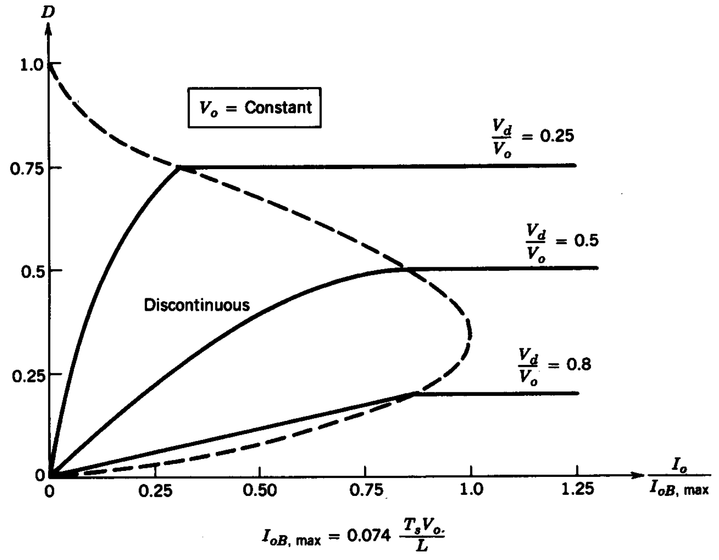


FIGURE 3-6: STEP-UP CONVERTER CHARACTERISTICS KEEPING OUTPUT VOLTAGE CONSTANT (7).

The boost converter also has two operating modes similar to the buck converter, discontinuous and continuous conduction mode. Compared to the buck converter it is desired to operate the boost converter in DCM for stability reasons; the reasons for this will come later. The boundary for operating at DCM is decided by equation 3-5.

$$I_{oB} = \frac{T_s V_o}{2L} D(1-D)^2 \tag{3-5}$$

3.2.3 NON-INVERTING BUCK-BOOST CONVERTER

A cascade connection of a step-down and a step-up converter gives the characteristics to deliver a constant voltage whether the input voltage is at higher or lower voltage level. This gives three different operation modes; buck, buck-boost and boost. The cascade connection of the converter can mainly be done by simply connecting the boost converter at the output of the buck converter or it is possible to use one inductor, as proposed in Figure 3-7 and thus removing the capacitance at the output of the "buck" part of the converter.

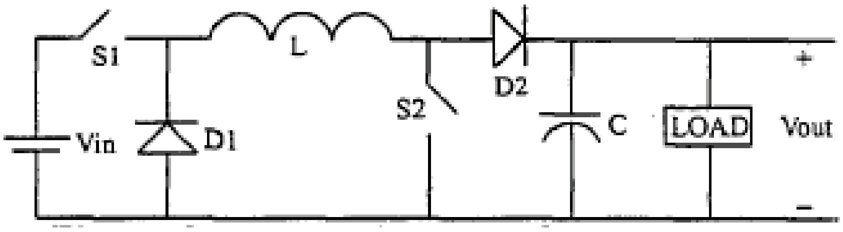


FIGURE 3-7: NON-INVERTING BUCK-BOOST (8)

This gives easier construction, but the inductor has to be large enough so CCM is possible for the buck converter, and then force the boost converter into CCM. This gives the need for a large choke in terms of both inductance and current rating.

3.2.4 MULTILEVEL CONVERTER

A multilevel converter is a converter often used in high powered converter in bridge connection to either convert from dc to ac or change the frequency. The multilevel converter shares the load over several switches; then often IGBTs with their anti-parallel diodes as switches. The switches are then coupled in either series or parallel, and they are often diode clamped. In these topologies snubbers are very essential because of the operation outside the safe operation area range.

The reason for the use of these topologies is that the lower rated switches can be used in a converter that delivers high voltage and power. This means that switches or transistors can be series coupled to withstand higher voltage, and paralleled to withstand higher current. In such cases several things must be considered, especially with regard to the perturbations in characteristics of each transistor. It seams like the main problems are leakage current and the turn-off time.

There is a need for such a solution in a down-hole application; since the voltage rating of the components available today has a lower breakdown voltage than there is need for. The assumed generator voltage can get up to about 200V or more, and it is not desired to have a large bulky transformer down-hole, and the transistors available have a breakdown voltage at 80V; so the need for such a solution speaks for itself.

To be sure that the voltage is shared evenly across the switches capacitors can be place in parallel with the switches. This helps the switch to share the dynamic change in voltage.

3.3 CONVERTER PROTECTION

When the converter is suddenly applied voltage the capacitor looks like a short circuit for the source, in the worst case scenario the input capacitor is completely discharged. This makes a very large current rush into the capacitor; to limit this inrush current the current charging the capacitor need to be controlled. This can be solved by the use of among others, a resistor, thermistor or MOSFET.

A resistor is a good current limiter, but it has a poor influence on the efficiency of the converter. The thermistor with a negative temperature coefficient is then a better current limiter; the NTC gets heated up by the current rushes through it and thus achieving a smaller resistance when it gets heated up. The latter solutions can not be controlled, thus an active inrush current control is desired. This can be done by mounting a MOSFET in the return path that is gradually decreasing its resistance, an example for this is found in reference (9) and is shown in Figure 3-8.

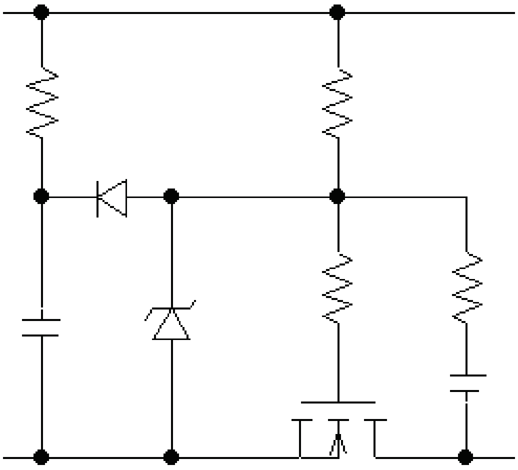


FIGURE 3-8: ACTIVE INRUSH CURRENT CONTROL (9)

Also the switches must be protected for transients when turning off and on the switches, especially when operating the devices outside the safe operation area region. For this snubbers can be implemented.

Soft-start can easily be implemented by a capacitor to gradually increasing the reference voltage, and using the current limitation on the control voltage as shown later to limit the start up inductor current.

4 CONTROL SYSTEM

The control system has to make sure that switch-mode power supply have good regulation to different variations. A SMPS should be designed to:

- Have good line regulation; output voltage remains the same due to variations in input voltage.
- Have good load regulation; output voltage remains the same due to load changes.
- Have good transient response
- Remain stable under all operating conditions.

4.1 DIRECT DUTY RATIO PWM CONTROL

The duty ratio is used to control the output voltage of SMPS, and the output voltage is in most cases compared to a reference voltage to adjust the duty ratio. In direct duty control the duty ratio is obtained by comparing the error signal, which is decided from the comparison of the output voltage and the reference voltage, with a fixed frequency sawtooth voltage. This approach is shown in Figure 4-1 borrowed from (7).

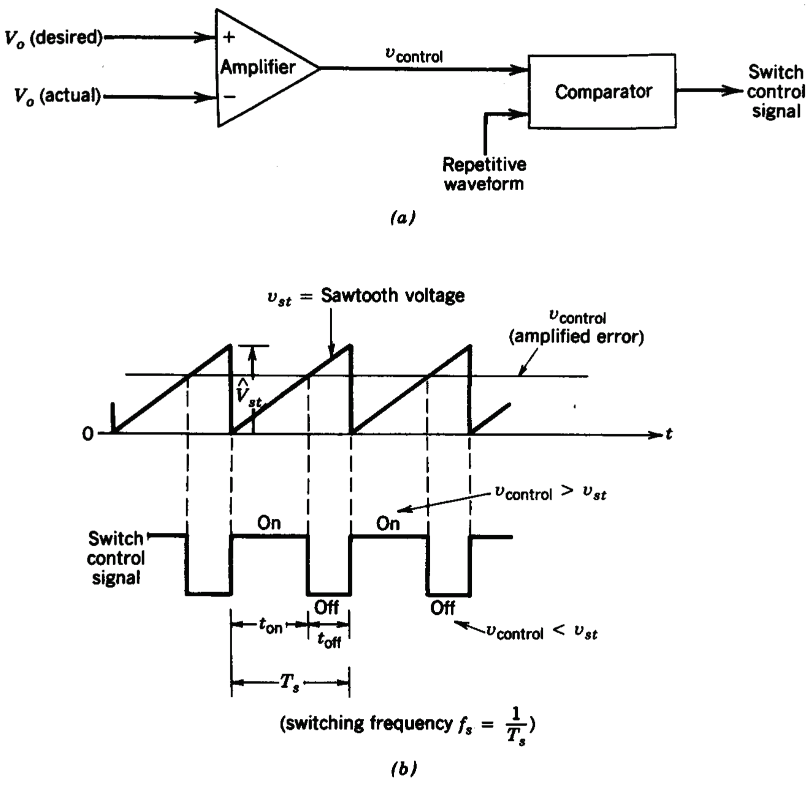


FIGURE 4-1: DIRECT DUTY RATIO PWM CONTROL

4.2 VOLTAGE FEED-FORWARD PWM CONTROL

The voltage feed-forward PWM control is a slight improvement of the previous control method. In direct duty PWM control the dynamic line regulation is poor since the output must react to variation before the duty cycle is changed. To get better line rejection the input voltage is fed forward to regulate the ramp, with that the peak of the sawtooth signal. This is only a good control method for CCM buck converter or DCM buck-boost converter.

4.3 CURRENT MODE CONTROL

In current mode control the PWM duty cycle is derived from the inductor current, thereby making a second internal feedback loop. The second loop then cancels out the double LC pole. This approach to controlling the voltage gives a supposedly simpler compensation, because of one less system pole, but the mathematical modelling is more challenging. The mathematical modelling is more challenging mainly because there are two feedback loops in the system; the voltage feedback loop, and a current feedback loop.

As the name tells this type of controller directly controls the inductor current in the converter, and thus setting the output voltage. The current can be set by the use of several methods; either by letting the control voltage set the average current or setting the peak current. The most used current mode method in SMPS is to set the peak inductor current for when the switch is turned off, and turned on again at the next constant-frequency switching period. If the converter is to be used in both CCM and DCM, a constant frequency switching is needed.

The main advantages of current-mode control are:

- Eliminates the effect of the inductor in the feedback loop.
- Limitation to the peak switch current.
- Precise load sharing between two paralleled converters.
- Voltage feed-forward property.

There are some properties with current mode control that have to be dealt with. These are loop instability when operating at duty cycles over 0.5, a voltage feed-forward property that is not perfect, and a sub-harmonic oscillation at half the switching frequency. These issues can be solved by slope compensating the control voltage or the set inductor peak. The slope compensating is done by use of a sawtooth waveform that has a down-slope of one half of the slope the inductor current when the switch is off.

4.4 FEEDBACK COMPENSATION

The feedback signal must be properly tailored so that the SMPS operates within the expected requirements. If the error signal is just amplified there will be a steady state error on the output voltage. It is not the dc analysis that is important; it is the small signal ac analysis that is the small perturbations in the output voltage. To get better transients the poles of the power plant is needed to be shifted with a proper regulator.

A system is bounded input bounded output stable (BIBO) only if all the system poles are in the left half plane; this is the case of all the converters.

4.4.1 TRANSFER FUNCTIONS

The transfer function of the different converters plays an important role regarding the design of the feedback compensation. It is after these that the cancelation poles and zeros of the error compensation are chosen. There are the transfer functions from duty cycle to output that is interesting when designing the control system.

Buck transfer function in CCM:

$$\begin{aligned} G_p(s) &= \frac{\tilde{v}_o(s)}{\tilde{d}(s)} = V_d \frac{\omega_0^2}{\omega_z} \frac{s+\omega_z}{s^2+2\xi\omega_0 s+\omega_0^2} \\ \omega_0 &= \frac{1}{\sqrt{LC}} \\ \xi &= \frac{1/CR+(r_c+r_L)/L}{2\omega_0} \\ \omega_z &= \frac{1}{r_c C} \end{aligned} \tag{4-1}$$

This shows that the buck has two left plane poles due to the denominator term and one well behaved zero that also is in the left hand plane thus giving a negative phase.

Boost transfer function in CCM:

$$
\begin{aligned}
&G_p(s) = \frac{\tilde{v}_o(s)}{\tilde{d}(s)} = V_d f(D) \frac{(1 + s/\omega_z)(1 - s/\omega_z)}{as^2 + bs + c} \\
&G_p(s) = \frac{\tilde{v}_o(s)}{\tilde{d}(s)} = V_d \frac{1}{(1-D)^2} \omega_0^2 \frac{(1 + s/\omega_{z1})(1 - s/\omega_{z2})}{s^2 + 2\xi\omega_0 s + \omega_0^2} \\
&\omega_0 = \frac{1}{\sqrt{\underline{L}C}} \\
&\xi = \frac{1/CR + (r_c + r_L)/\underline{L}}{2\omega_0} \\
&\omega_{z1} = \frac{1}{r_c C} \\
&\omega_{z2} = \frac{R(1-D)^2}{L} \\
&\underline{L} = L/(1-D)^2
\end{aligned}
\tag{4-2}
$$

The boost transfer function shows that it is dependent on the duty ratio itself; this makes the low frequency gain non-linear. The boost converter has two zeros and two poles, where one of the zeros is in the right hand plane and giving a positive phase contribution. The right hand zero can be explained by what happens when there is a sudden increase in duty cycle when the converter is operating in CCM. The increase in duty cycle makes the voltage drop, this makes the duty cycle larger again and instability might occur. This undesired action can be avoided if the converter operates in DCM; this helps the stability in such a way that no charge is stored in the inductor after each switching cycle.

The transfer functions of the converters can be easily transferred in to discrete transfer function by the use of the MATLAB function $c2d(G_s, T_s, \text{'}method\text{'})$, where H is the transfer function, T_s is the sampling frequency and 'method' is the method used to find a discrete transform of the transfer function.

The transfer functions of the pulse width modulator are also needed in the design of the feedback compensation. They differ for which feedback control is used; direct duty, voltage feed-forward or current mode control of the duty cycle.

The transfer function of the direct duty is a simple ramp decided by the peak sawtooth voltage Vs and is shown in equation 4-3.

$$G_m(s) = \frac{\tilde{d}(s)}{\tilde{v}_c(s)} = \frac{1}{\hat{V}_s} \tag{4-3}$$

In the voltage feed-forward control, the peak voltage is decided by the input voltage, and thus the equation is shown in 4-4.

$$G_m(s) = \frac{\tilde{d}(s)}{\tilde{v}_c(s)} = \frac{K}{V_d} \tag{4-4}$$
$$K = \text{scaling factor}$$

With current mode control it is easier to take a different approach to finding the transfer function. Since the inductor current is implemented in the feedback, and thus removing the inductor pole, it is easier to consider both the power stage and the modulator in one step.

CCM buck transfer function:

$$\begin{aligned} G_s(s) &= \frac{\tilde{v}_o(s)}{\tilde{v}_c(s)} = KR_{Load}\frac{\omega_p}{\omega_z}\frac{s+\omega_z}{s+\omega_p} \\ \omega_p &= \frac{1}{RC} \\ \omega_z &= \frac{1}{r_c C} \\ K &= \text{scaling factor} \end{aligned} \tag{4-5}$$

DCM Boost transfer function:

$$\begin{aligned}
&G_s(s)=\frac{\tilde{v}_o(s)}{\tilde{v}_c(s)}=K\cdot G_0^{-1}\omega_p\frac{(1+s/\omega_{z1})(1-s/\omega_{z2})}{s+\omega_p}\\
&G_0^{-1}=\frac{2}{1+\sqrt{1+2D^2TR/L}}\\
&\omega_p=\frac{1}{RC}\qquad\sqrt{\qquad\qquad\qquad}\\
&\omega_{z1}=\frac{1}{r_cC}\\
&\omega_{z2}=\frac{R_L(1-D)^2}{L}\\
&K=\text{scaling factor}
\end{aligned}\tag{4-6}$$

4.4.2 COMPENSATION TECHNIQUES

There are mainly three types of compensations that are used; type 1, 2 and 3. The type number depicts the number of poles provided by the compensation, thus the most comprehensive compensation technique is type 3 compensation. These compensation techniques can be used for both analog and digital feedback compensation.

4.4.3 ANALOG FEEDBACK COMPENSATION

The analog feedback compensation is done by the use of either trans-conductance amplifiers or operational amplifiers that have negative “feedback”. The trans-conductance amplifier does not have a direct feedback as it is with the operational amplifier. The gain of the amplifiers, and thus the error compensator, is decided by the input network and feedback network. The most common compensation types are given under in equations 4-7, 4-8 and 4-9.

- Type 3 with two zeros and three poles:

$$\begin{aligned}
G_c(s) &= K\frac{(s+\omega_{z1})(s+\omega_{z2})}{s(s+\omega_{p1})(s+\omega_{p2})} \\
K &= \frac{R_1+R_3}{R_1R_3C_1} \\
\omega_{z1} &= \frac{1}{R_2C_2} \\
\omega_{z2} &= \frac{1}{(R_1+R_3)C_3} \\
\omega_{p1} &= \frac{C_1+C_2}{R_2C_1C_2} \\
\omega_{p2} &= \frac{1}{R_3C_3}
\end{aligned} \tag{4-7}$$

- Type 2 with one zero and two poles:

$$\begin{aligned}
G_c(s) &= K\frac{s+\omega_z}{s(s+\omega_p)} \\
K &= \frac{1}{R_1C_1} \\
\omega_z &= \frac{1}{R_2C_2} \\
\omega_p &= \frac{C_1+C_2}{R_2C_1C_2}
\end{aligned} \tag{4-8}$$

- Type 1 with one pole at zero:

$$G_c(s) = \frac{1}{R_1C_2}\frac{1}{s} \tag{4-9}$$

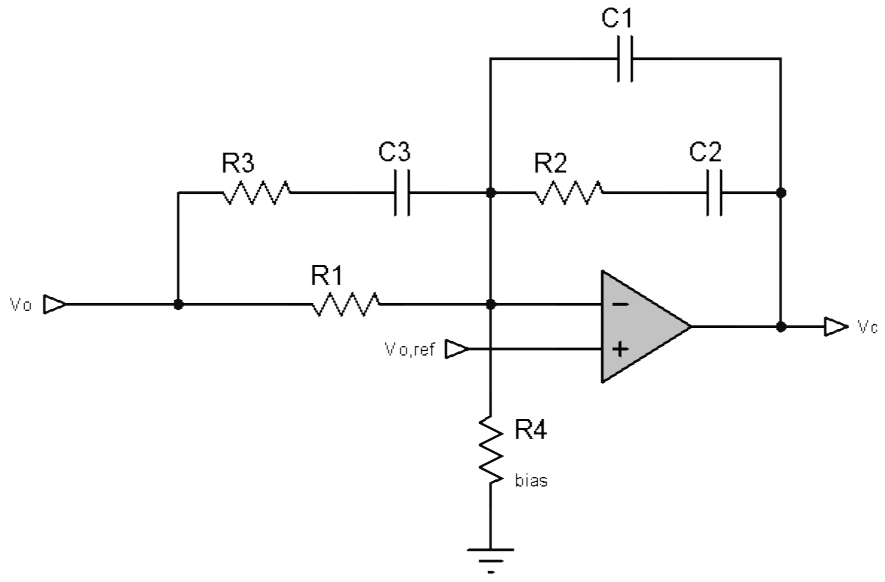


FIGURE 4-2: TYPE 3 CONTROLLER

Figure 4-2 show the implementation of type 3 controllers, the other types can be found by removing the components that are not a part of their transfer function. The bias resistor is there as a part of the voltage divider together with R1 to downsize the output voltage to a comparable voltage level. The voltage divider can either be placed like it is done here, or it can be placed before the controller since it does not play a part of the feedback anyway.

4.4.4 DIGITAL FEEDBACK COMPENSATION

Digital feedback compensation does not rely on any components as with the analog implementation. The drawback with analog feedback control is the parasitic influence in drift of components specifications. In digital control this is avoided, but the digital control has a much lower bandwidth so the high frequency perturbations can not be considered and thus has to be filtered out. One other aspect that is better with digital control is that it is possible to use more comprehensive control schemes.

The limitation in the bandwidth of the system can be seen if a buck converter is converted from a continuous model to a discrete model. If a sampling frequency of 100 kHz is used with the Tustin method in MATLAB, the bode-plot will become as in Figure 4-3. This shows that the discrete model does not consider what happens at higher frequencies than half the sampling frequency.

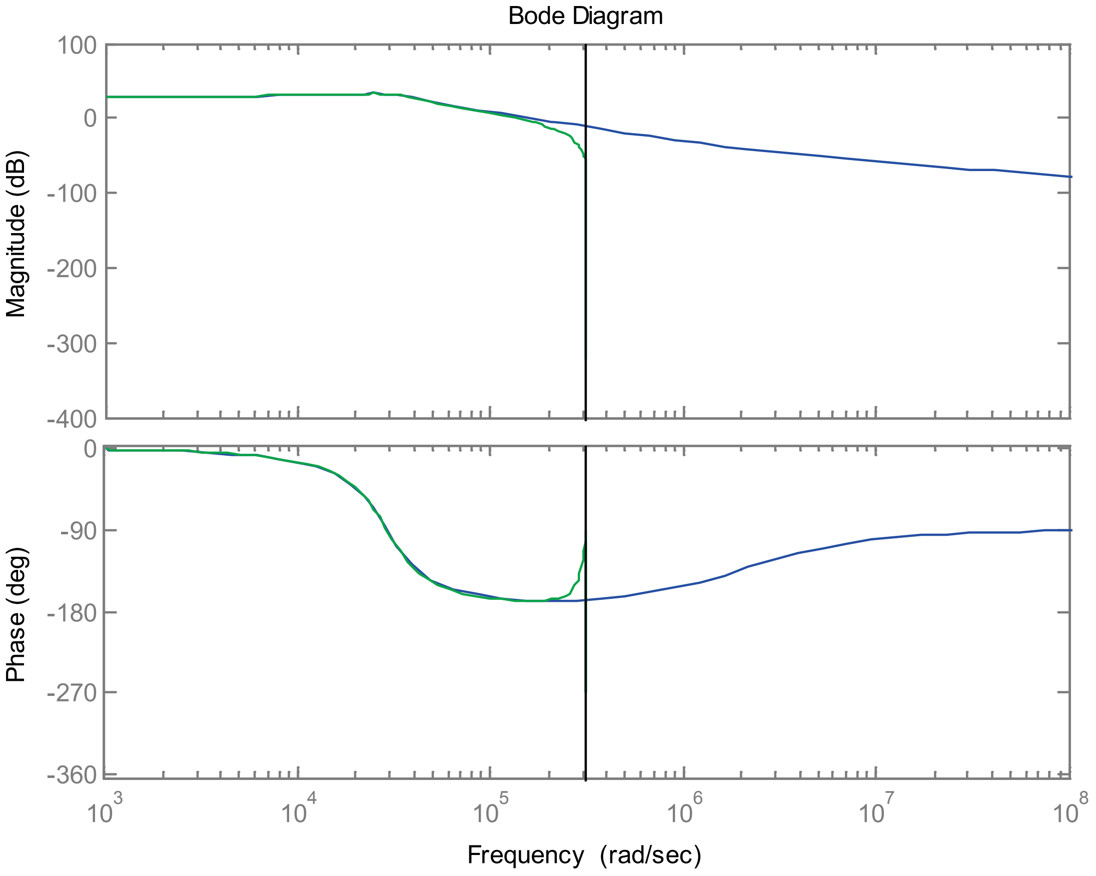


FIGURE 4-3: BODE DIAGRAM

An example for a type 3 digital controller is shown in equation 4-10.

- Type 3 with two zeros and three poles:

$$G_c(z) = \frac{b_0 + b_1 z^{-1} + b_2 z^{-2} + b_3 z^{-3}}{1 + a_1 z^{-1} + a_2 z^{-2} + a_3 z^{-3}} \qquad (4\text{-}10)$$

A comprehensive compensation like the type 3 controller might not be necessary to use since some of the zeros and poles might be at a higher frequency.

4.5 MONITORING

The system needs an “intelligent” monitoring system that can perform decision making for the system; some of these decisions are found underneath.

- Monitor battery voltage.
- Monitor generator voltage and rpm.
- Decide whether buck, boost or buck-boost is used for the generator voltage in hand.
- Make sure that the generator or battery can supply the power needed.
- Store the measurement from the instrumentation.
- Decide when to send the data up-hole, listen for incoming data.
- Etc.

To be able to monitor the system Honeywell’s HT83C51 μcontroller can be implemented, this is a pin equivalent μcontroller to the standard Intel 8XC51FC μcontroller.

5 CONVERTER DESIGN & SIMULATION

5.1 PRACTICAL DESIGN

For better control and the advantages it is chosen to implement a current mode control for both the buck and boost converters. Especially for the converters that are going to charge a battery it is important to implement current mode control to be able to limit the charging current. The practical design of the control system shown in Figure 5-1 is provided by the use of 555 timers and operational amplifiers.

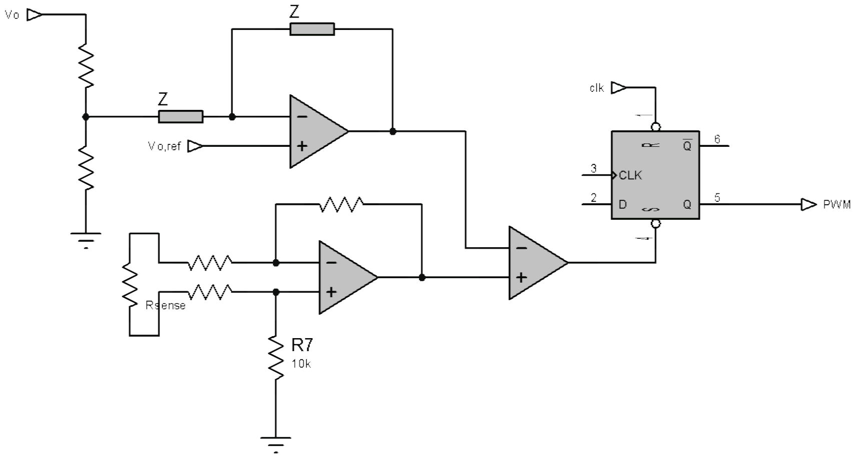


FIGURE 5-1: CURRENT MODE CONTROLLER

Since current mode is used for the control of the pulse width modulation and thus removing the inductor pole, only a type 2 controller is needed. The type 2 controller has one zero to move the output pole due to the output capacitor and the load. To be able to use the current mode compensation the inductor current must be measured, this is done by using a small resistor in series with the inductor and differencing the voltage measured over the resistor. This converts the inductor current to a voltage with reference to ground. A better method for measuring the voltage would have been by the use of a hall element, but those does not seem to be available for these temperatures. For a finished product the inductor current can be measured over one of the switches, and thus removing the series resistor.

The comparison and latching to make the pulse width modulated signal is done by the use of a 555 timer. The 555 timer is built up of two comparators, one SR flip-flop with inverting output, one discharge transistor and one inverter. For the PWM modulator only the comparators and flip-flop is used.

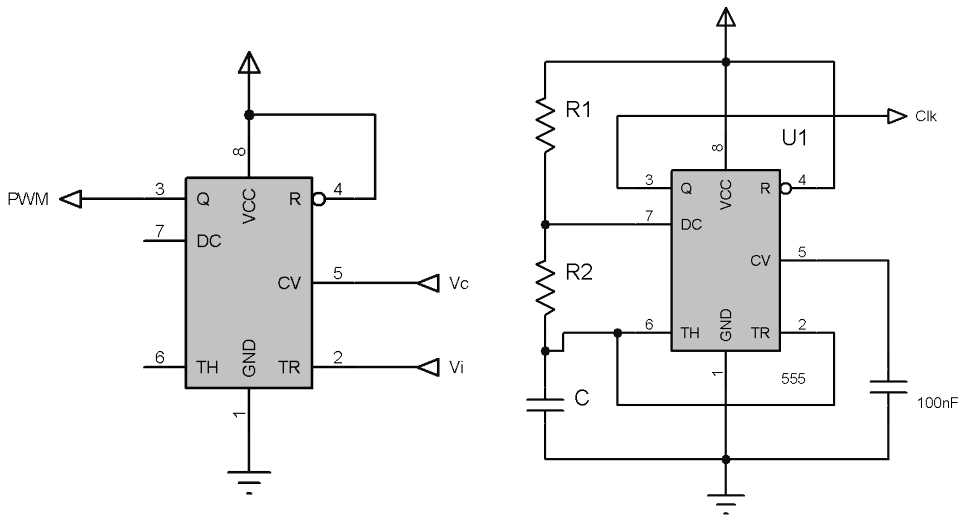


FIGURE 5-2: PULSE WIDTH MODULATOR & ASTABLE WITH 555 TIMER

The constant frequency clock is provided by an astable 555 timer shown in Figure 5-2, the frequency is decided by $f = 1.44 / \left[\left(R_1 + 2R_2 \right) C \right]$.

To get the same voltage at both drain and source for a MOSFET a voltage on the gate is needed that is 5V (for the CHT-NMOS80) higher than the source voltage. For a low side switch this is easily achieved because the source is connected to ground, but for a high side switch others methods must be used. This can be solved by the use of among others bootstrapped drivers or linear regulators that are coupled over gate source of a MOSFET. The method that is used for this application is by the use of a full bridge bootstrapped driver CHT-FBDR. The bootstrap is a capacitor that is charged with a charge pump, if the charge pump is place in parallel with the switch the capacitor voltage will be the same has the drain voltage.

5.2 HIGH SIDE STEP-DOWN AND STEP-UP CONVERTER

To the battery link a voltage at about 30v is desired, the input voltage from the generator is assumed to be rectified by a three phase diode bridge; it is assumed that the rectified generator voltage varies from 7-200V. The calculations for the converter design are done with a MATLAB script and are shown in appendix A. This script does also calculate the pole placements for a type 2 controller and is done by selecting a crossover frequency at least a third less than the switching frequency, and placing the poles and zeros at the different system poles and zeros. Other methods like the use of a K factor as found in (7) and thus choosing a phase margin may be better, but the approach found in (10) gives a good match. The circuit diagrams are found in appendix B

The step-down converter has to deal with the voltage input that is larger than the output voltage, and the converter is assumed to always operate in CCM to keep the current ripple minimized. Therefore the inductor has to be sufficient to maintain a constant current flowing also when the required power is at its maximum, and the duty cycle is high. The calculated specifications are represented in Table 5-1

TABLE 5-1: STEP-DOWN CONVERTER SPECIFICATIONS

Specifications	
Output voltage [V]	30
Maximum power output [W]	10
Minimum power output [W]	3
Maximum voltage[V]	200
Minimum voltage [V]	35
Switching frequency [kHz]	200
Minimum inductance [µH]	956
Minimum capacitance [µF]	0.3

The step-up converter has to deal with the last part, the voltage variation from 7-30V, for stability reasons it is desired that the step-down converter is operating in DCM. If the boost converter is going to deliver maximum power at the lowest input voltage components with quite high current rating must be implemented, a solution to this is to clamp the control voltage at a lower voltage and thus limiting the power delivered during low generator output.

Table 5-2 represents the calculated specifications.

TABLE 5-2: STEP-UP CONVERTER SPECIFICATIONS

Specifications	
Output voltage [V]	30
Maximum power output [W]	10
Minimum power output [W]	3
Maximum voltage[V]	25
Minimum voltage [V]	7
Switching frequency [kHz]	100
Minimum inductance [µH]	18.7
Minimum capacitance [µF]	8.5

As seen in the specifications of the converters they do not cover the entire range of input voltages from lowest to highest voltage. The reason for this is that the converter is going to be controlled like a buck-boost converter for the input range from 25 to 35V.

5.3 LOW SIDE STEP-DOWN CONVERTER

The low side converter is s stepping down the voltage from the battery link that is assumed to be near to constant; this is to supply the instrumentation and the wireless communication up- and down-hole. A small difference in input voltage is shown by the need for a voltage higher than the battery voltage for charging. So the dc-link for the low side converter needs to cope with a small voltage variation from 25-30V according as the power is delivered from the generator or the battery. The calculated specifications for the low side converter are represented in Table 5-3.

.

TABLE 5-3: STEP-DOWN CONVERTER SPECIFICATIONS

Specifications	
Output voltage [V]	5
Maximum power output [W]	10
Minimum power output [W]	3
Maximum voltage[V]	35
Minimum voltage [V]	25
Switching frequency [kHz]	100
Minimum inductance [µH]	53.5
Minimum capacitance [µF]	20

5.4 SIMULATION

The simulation of the systems is done in LTSpice; LTSpice is a free spice simulator that is provided by Linear Technology.

5.4.1 VOLTAGE SHARING

To share the voltage across series coupled switches, capacitors must be placed in parallel with the MOSFET to dynamically share the voltage over the switches. Figure 5-3 shows what happens without voltage sharing when two different switches are arbitrary chosen.

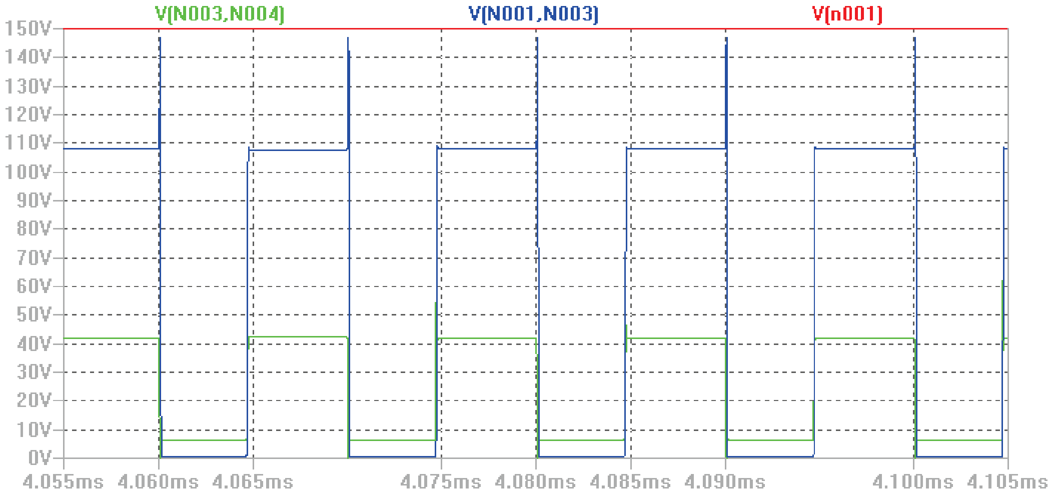


FIGURE 5-3: SERIES COUPLED SWITCHES WITH DIFFERENT CHARACTERISITICS WITHOUT VOLTAGE SHARING

If a capacitor is placed in parallel with each switch a completely different switching pattern will occur as shown in Figure 5-4, and the voltage is shared evenly.

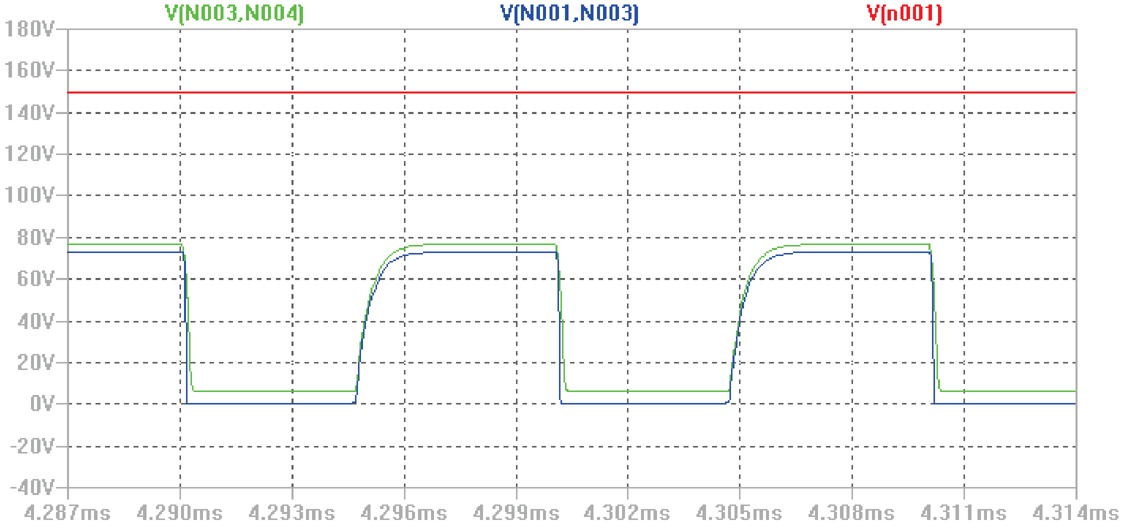


FIGURE 5-4: SERIES COUPLED SWICHES WITH DIFFERENT CHARACTERISTICS AND VOLTAGE SHARING

5.4.2 5V STEP-DOWN REGULATOR

The 5v regulator is the regulator that is going to deliver power to the instrumentation equipment and other electronic circuitry. Figure 5-5 shows the circuit diagram for the simulated 5 v regulator.

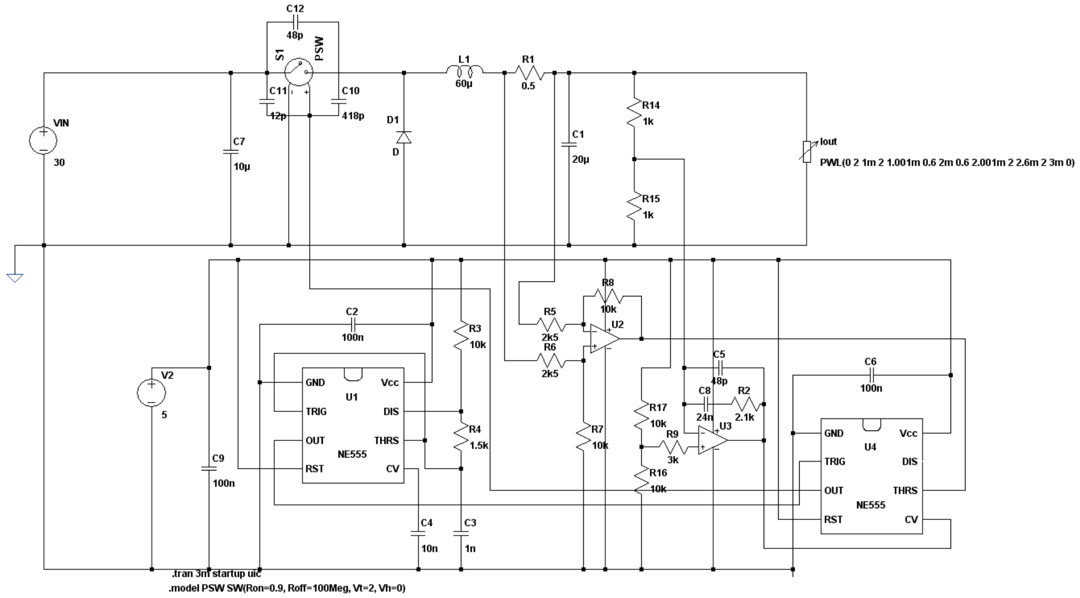


FIGURE 5-5: SIMULATION OF 5V REGULATOR

Figure 5-6 shows the startup of the 5v regulator and it shows that the regulator has good line rejection. The peak value of the output voltage can be limited by limiting the control voltage with a zener-diode and by ramping up the switching frequency.

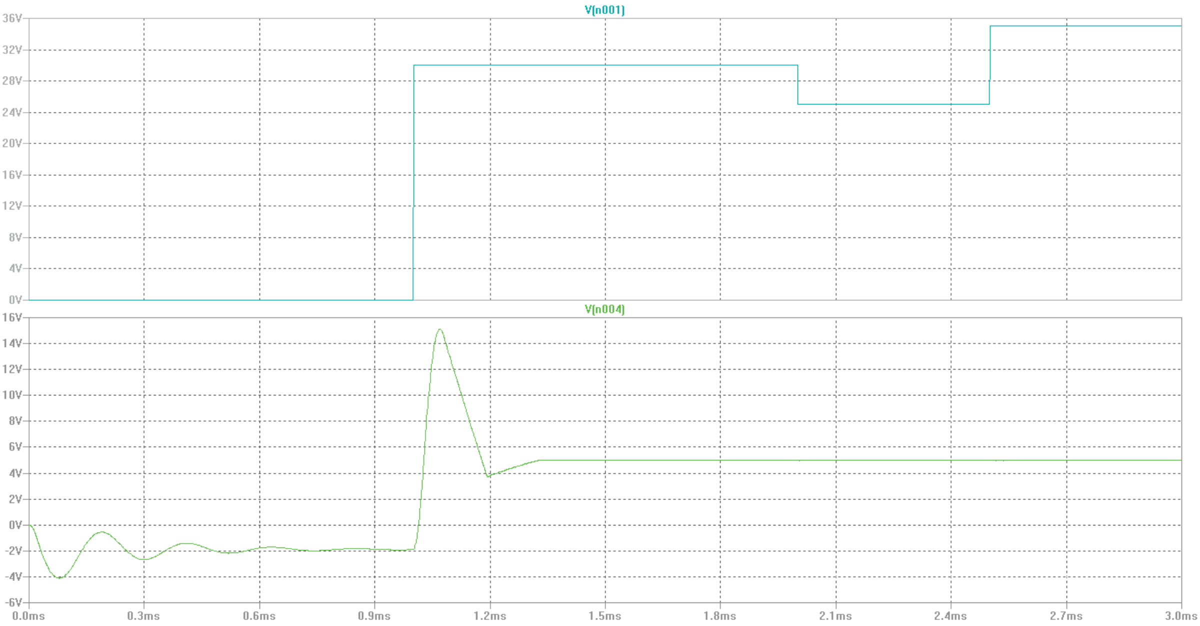


FIGURE 5-6: STARTUP OF 5V REGULATOR WITH CHANGE IN INPUT VOLTAGE

Figure 5-7 shows that the converter as a much better step response when the converter is helped by the use of soft-start techniques.

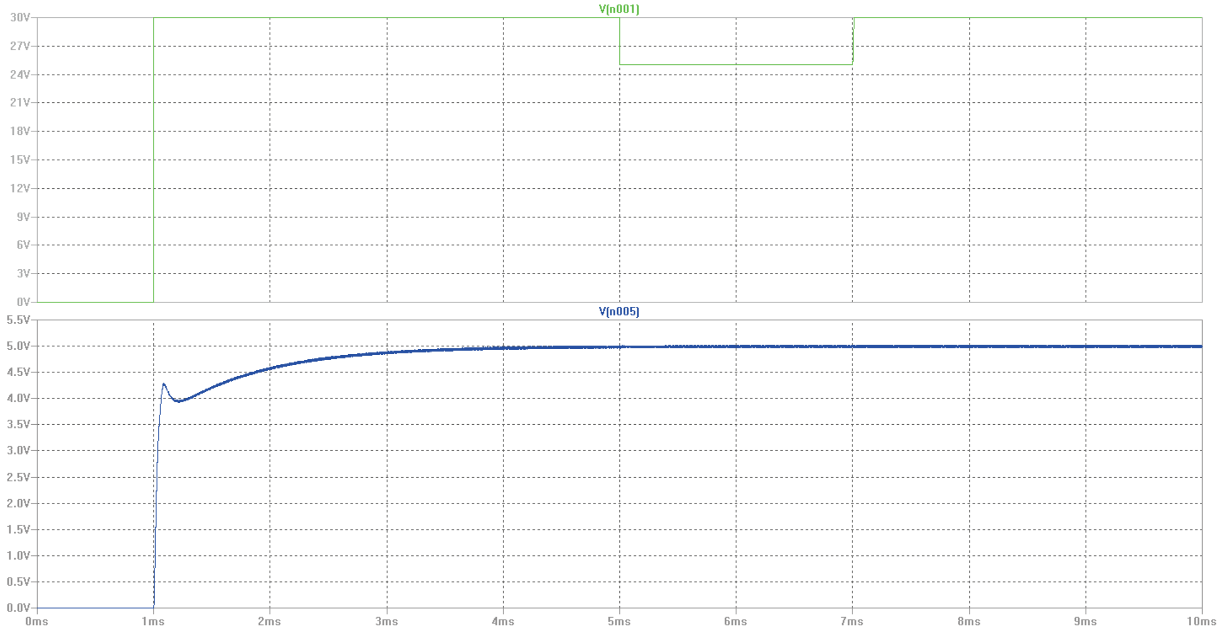


FIGURE 5-7: STEP-DOWN REGULATOR WITH SOFT-START

The step response for change in load is another specification that must be checked, Figure 5-8 shows that the step response is ok.

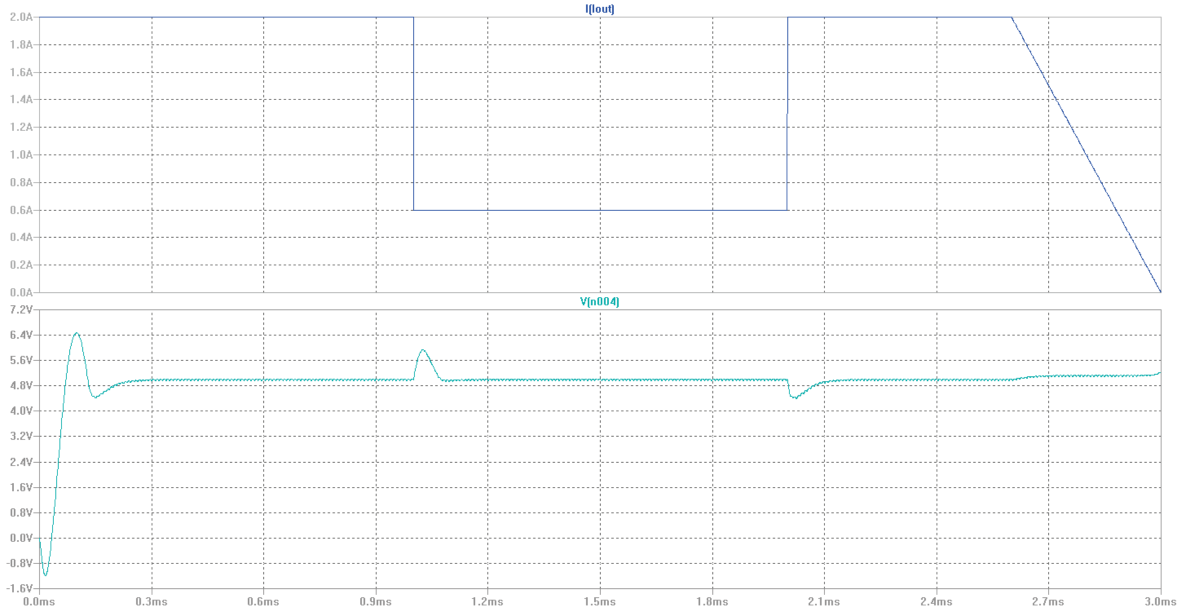
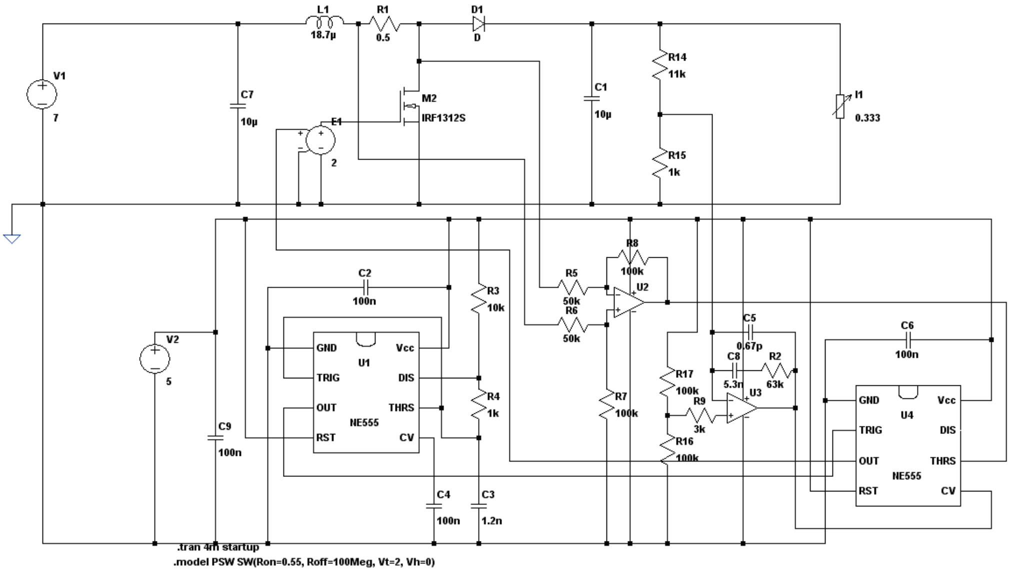


FIGURE 5-8: OUTPUT VOLTAGE FOR STEP IN LOAD CURRENT

5.4.3 30V STEP-UP REGULATOR



```
.tran 4m startup
.model PSW SW(Ron=0.55, Roff=100Meg, Vt=2, Vh=0)
```

FIGURE 5-9: 30V STEP UP REGULATOR

Figure 5-9 shows the simulation circuit diagram of the step-up converter, which is converting 7-25Vdc to 30Vdc. Also this converter has been simulated for step in input voltage and load current, Figure 5-10 and Figure 5-11 shows respectively the step response.

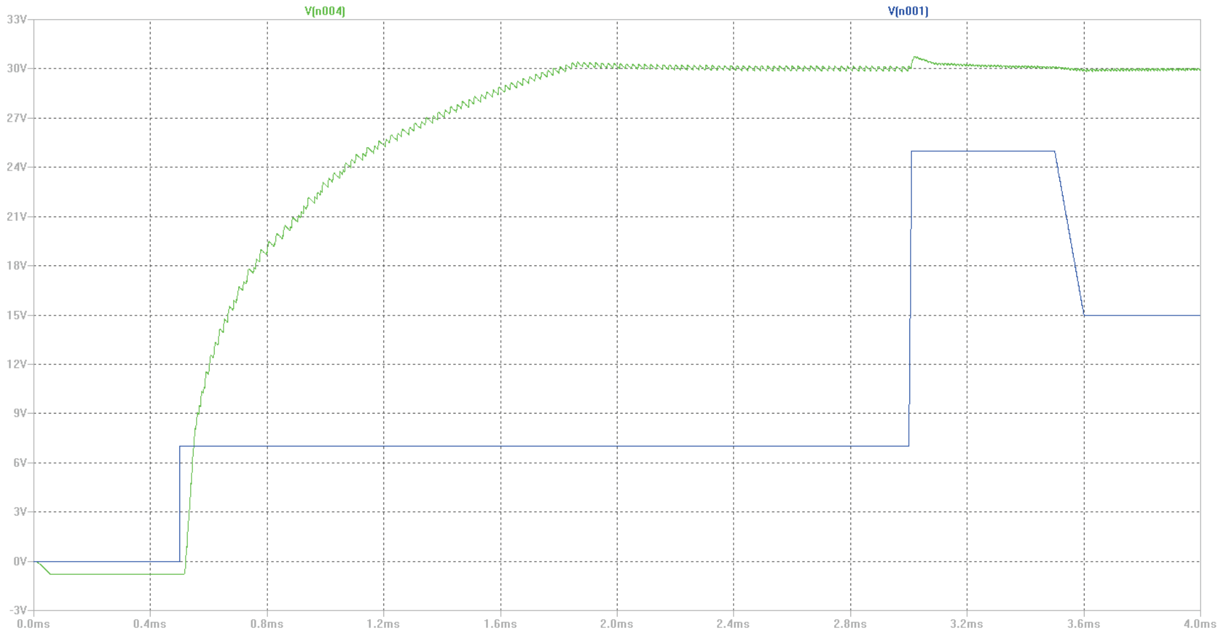


FIGURE 5-10: OUTPUT VOLTAGE DUE TO CHANGE IN INPUT VOLTAGE

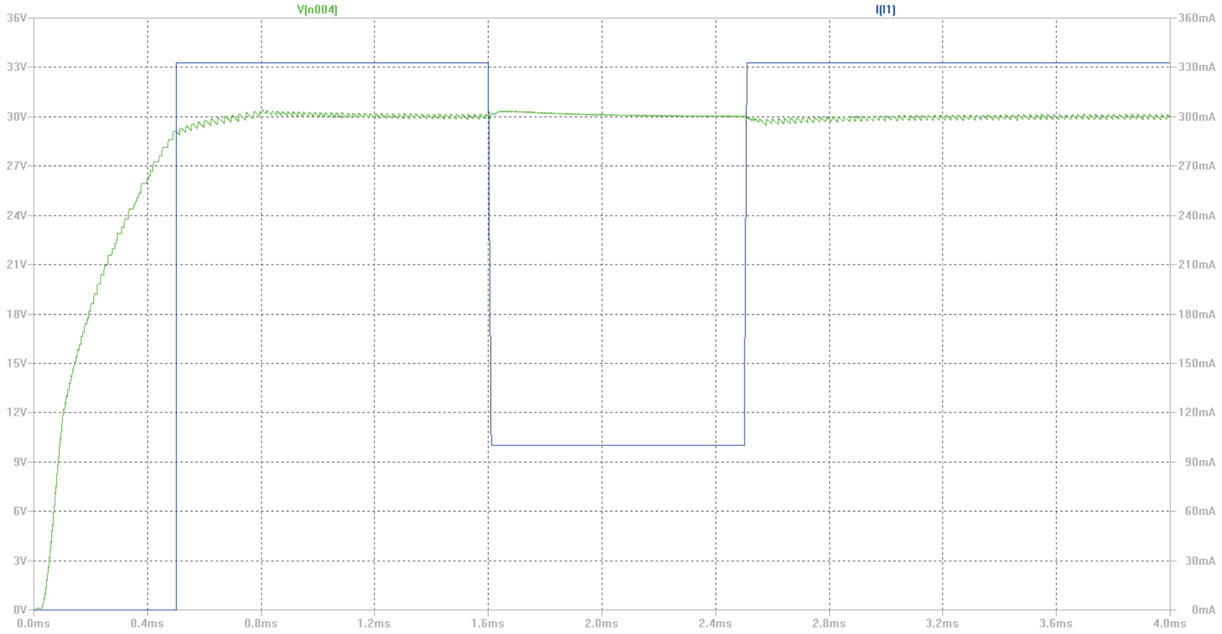


FIGURE 5-11: OUTPUT VOLTAGE DUE TO CHANGE IN OUTPUT CURRENT

This simulation has some problems with running at low effect at high input voltage; because of the PWM modulator used has a limitation in minimum duty cycle in the simulation. This might not happen in real life.

As read earlier the MOSFET CHT-NMOS80 has a current limitation at 3.4A for operation at 225^{0}C. This renders the possibility to deliver the highest output current for low input voltages impossible. A solution then is to design the control routine so high power output is not allowed during low input voltage. Another solution is to limit the control voltage and thus limiting the inductor current, then the output voltage from the boost converter will fall and thus letting the battery help to maintain operation. Figure 5-12 shows the inductor current when delivering maximum power at low input voltage.

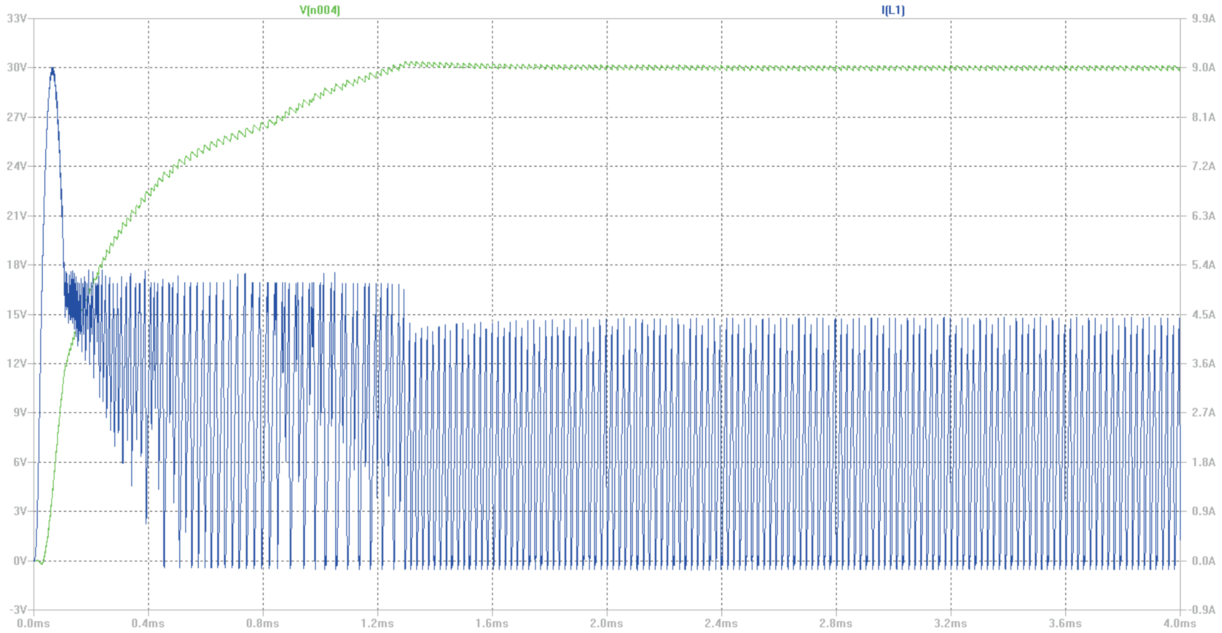


FIGURE 5-12: INDUCTOR CURRENT FOR HIGH POWER OUTPUT AT LOW INPUT VOLTAGE

5.4.4 30V STEP-DOWN REGULATOR

The last converter that has to cope with the largest input range is a step-down converter. To be able to operate this converter a relatively large inductor is needed, this renders the simulation difficult to run. The reason for this is that Spice originally was designed for simulation of integrated circuits, not switch-mode power supplies. The simulation files are attached in a zip-file, thus making it possible to test the simulation on a much more powerful computer. The prototype design is assumed working, since the other converters runs nicely. Figure 5-13 shows the simulation layout for the 30V step-down converter.

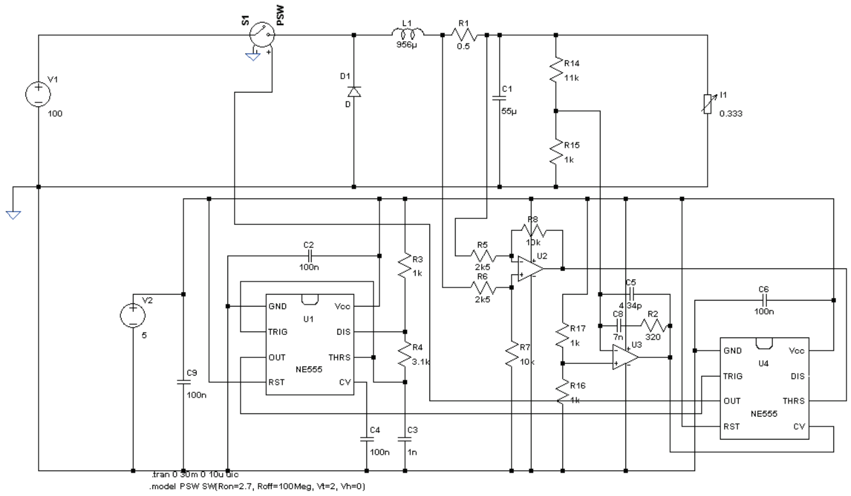


FIGURE 5-13: 30V STEP-DOWN CONVERTER

6 DISCUSSION

The Silicon on Insulator technology seems like a good choice for this kind of converters for high temperature operation. Silicon on Insulator technology has for the last years become a mature technology, and thus has good reliability and lifetime. A NASA report (11) evaluates CISSOIDS NMOS80. In this report is the SOI MOSFET device found to maintain good operation throughout the operation range from -195 ^{0}C to 225 ^{0}C with minimal changes in characteristics. The MOSFET device did not have any problems restart at the extreme temperature. This shows the potential to use this type of transistor for operation down-hole.

To maintain best possible reliability many different technologies must be implemented. SOI power ICs with a 250V, 0.5A, three phase, 1-chip inverter have been fabricated in a 5μm thick SOI layer and tested for increased temperatures up to 200^{0}C (2). For the high reliability in lifetime needed for a converter in down-hole applications, other solutions are found better. The control system is best found to be implemented by the use of analog control either as an application specific integrated circuit or multichip circuit module. The reason for this is that the advantages in minimizing parasitic properties when using either ASIC or MCM, and thus letting the μcontroller supervise the system instead.

The power circuitry is found most suitable to use a hybrid circuit solution with a ceramic substrate instead of using high temperature PCB. On a hybrid circuit the passive devices is implemented with thick-film technology, the inductor choke and output capacitors must still be implemented together with active components by the use of through-hole technology. The capacitors that seems most suitable for this application, is the AVX stacked multi layer ceramic capacitors. These are among others designed for use in SMPS in down-hole applications. The inductor can be implemented by the use of ordinary ferrite core, but with for example Teflon insulation.

The active components that are to be placed on the hybrid circuit, and thus not in the MCM, is the MOSFET switches and MOSFET driver. All of the CISSOID and Honeywell standard semiconductors can be ordered both in die and with casing and thus letting a quite versatile system be built.

A hermetically sealed packaging design must be used to pack the system; this to keep the humidity and gases from the harsh environment on the outside.

When the power supply has been produced each individual module has to be through a screening. This has to be done to avoid infant failures, a screening should at least apply temperature cycling, burn-in storage and free-fall drops.

There are quite large losses with devices available commercial off-the-shelf for high temperature operation today. As an example is the on-state drain-source resistance for the power MOSFET NMOS80; the resistance is almost 1Ω for operation at 225 ^{0}C. This gives an on-state loss given by the relation $P_{loss} = R_{ds} \cdot I_{ds}^2$, in switching operation the loss calculations gets different. An example for when this design exhibits large losses is when boost region is used. In the boost region must the buck switches stay closed for the entire operation; minimum three switches must be series coupled in the buck converter. This gives a series resistance at almost 3Ω when operating at 225 ^{0}C, by the high current drawn in boost operation a low efficiency is found. The reference dc-dc buck converter design supplied by CISSOID (4) has up to 85% efficiency, so the efficiency is at least not higher than 85% for this converter design.

7 CONCLUSION

For an intelligent instrumentation unit for down-hole high temperature applications high reliability electronics must be used. Ordinary bulk-CMOS circuits have a practical ambient operational temperature limit below 175 ^{0}C. This renders the need for a technology replacing the bulk-CMOS process. A successor is then the Silicon on Insulator technology, SOI is using the same fabrication as the bulk-CMOS. The difference lies in a buried oxide layer in the silicon wafer. The implementation of this oxide layer gives the possibility to make semiconductors with higher breakdown voltage and/or higher junction temperature.

The SOI technology is then used together with hybrid circuits using ceramic substrate as a replacement for organic PCB and thick-film technology for passive devices. This package gives a system with high reliability both toward high temperature operation and lifetime. When operating the SOI devices at temperatures of 175 ^{0}C, the lifetime is about 12-14 years, and 5 years for 225 ^{0}C.

The converters designed are the standard step-up and step-down switch-mode power supplies. The converters are designed with current mode control; current mode control is used because of the advantage that comes with it. Since an eventual battery probably needs a constant current charge, a current mode control is needed.

8 BIBLIOGRAPHY

1. **Fallet, Truls and Jensen, Mari Haugen.** High Temperature Instrumentation for Oil Applications. *SINTEF.* [Online] 16 January 2006. http://www.sintef.no/upload/IKT/9032/HiTEC%20final%20paper%20tfa.pdf.

2. **Nakagawa, Akio.** Power IC Technologies. [book auth.] Wai-Kai Chen. *The VLSI handbook.* Boca Raton, FL. : CRC PRESS, 2007.

3. **Kirschmann, Randall.** *High-Temperature Electronics: Edited by Randall Kirschmann.* New york, NY : IEEE Press, 1999. 0-7803-3477-9.

4. **CISSOID.** High Temperature Standard Products. *CISSOID Web site.* [Online] http://www.cissoid.com/high-temperature-electronics/ht-standard-products/high-temperature-standard-products.html.

5. **Honeywell Aerospace.** High Temperature Electronics. *Honeywell Aerospace web site.* [Online] 2004-2009. http://www.honeywell.com/sites/portal?smap=aerospace&page=High-Temp-Electonics3&theme=T5&catID=C82A27CF1-C0F1-76E9-6B52-2C477FB52FF7&id=H5E761CAC-F16E-40AF-B54E-3DFBA7F0A988&sel=1#Literature.

6. **AVX corporate.** SMPS Stacked MLC Capacitors - SMX Style. *AVX Online.* [Online] http://avx.com/prodinfo_productdetail.asp?I=1207&K=search&C=&word=smx&form=number&Number=.

7. **Mohan, Ned, Undeland, Tore M. and Robbins, William P.** *Power electronics: converters,apllications, and design.* Hoboken, N.J. : John Wiley & Sons, Inc., 2003.

8. **Weissbach, Robert S. and Torres, Kevin M.** *A NONINVERTING BUCK-BOOST CONVERTER WITH REDUCED.* [IEEE] Erie, PA : s.n., 2001. 0-7803-6748-0.

9. **Power-One.** Inrush current control: Application note. [Online] 5 March 2001. http://huqixin.blog.dianyuan.com/u/43/1152163498.pdf.

10. **Maniktala, Sanjaya.** *Switching Power Supplies A to Z.* Burlington : Elsevier, 2006.

11. **Patterson, Richard and Hammoud, Ahmad.** SOI N-Channel Field Effect Transistor, CHT-NMOS80, for Extreme Temperatures. *NASA technical reports server.* [Online] March

2009. http://ntrs.nasa.gov/archive/nasa/casi.ntrs.nasa.gov/20090014040_2009013116.pdf.

12. **Harris, Scott.** Power Electronics Technology. *Stacked-Ceramic Caps Brave High Temperatures.* [Online] August 1, 2007. [Cited: June 11, 2009.] http://powerelectronics.com/passive_components_packaging_interconnects/capacitors/708PET24.pdf.

13. **Whittington, H. W., Flynn, B. W. and Macpherson, D. E.** *Switched Mode Power Supplies: design and construction.* Taunton : Research Studies Press Ltd., 1997.

14. Considerations for series connection of IGBT and MOSFET switches. *DATAWEEK Electronics & communications technology.* [Online] Technews Publishing (Pty) Ltd, 28 May 2008. http://dataweek.co.za/news.aspx?pklNewsId=29028.

15. **Poley, Richard.** Control Theory Seminar Series: 4 modules. 2008-2009.

16. **Li, Jacob M.** Packaging Design & Manufacture of High Temperature Electronics... *Vectron International.* [Online] 2004. http://www.vectron.com/products/appnotes/HiTEC2004_TA1_Jacob_Li_BBL.pdf.

17. **AnaLog.** A Word About Solder. *AnaLog Services, Inc. home page.* [Online] 9 March 2007. http://www.logwell.com/tech/servtips/solder.html.

18. Wikipedia, the free encyclopedia. [Online] http://en.wikipedia.org/wiki/Main_Page.

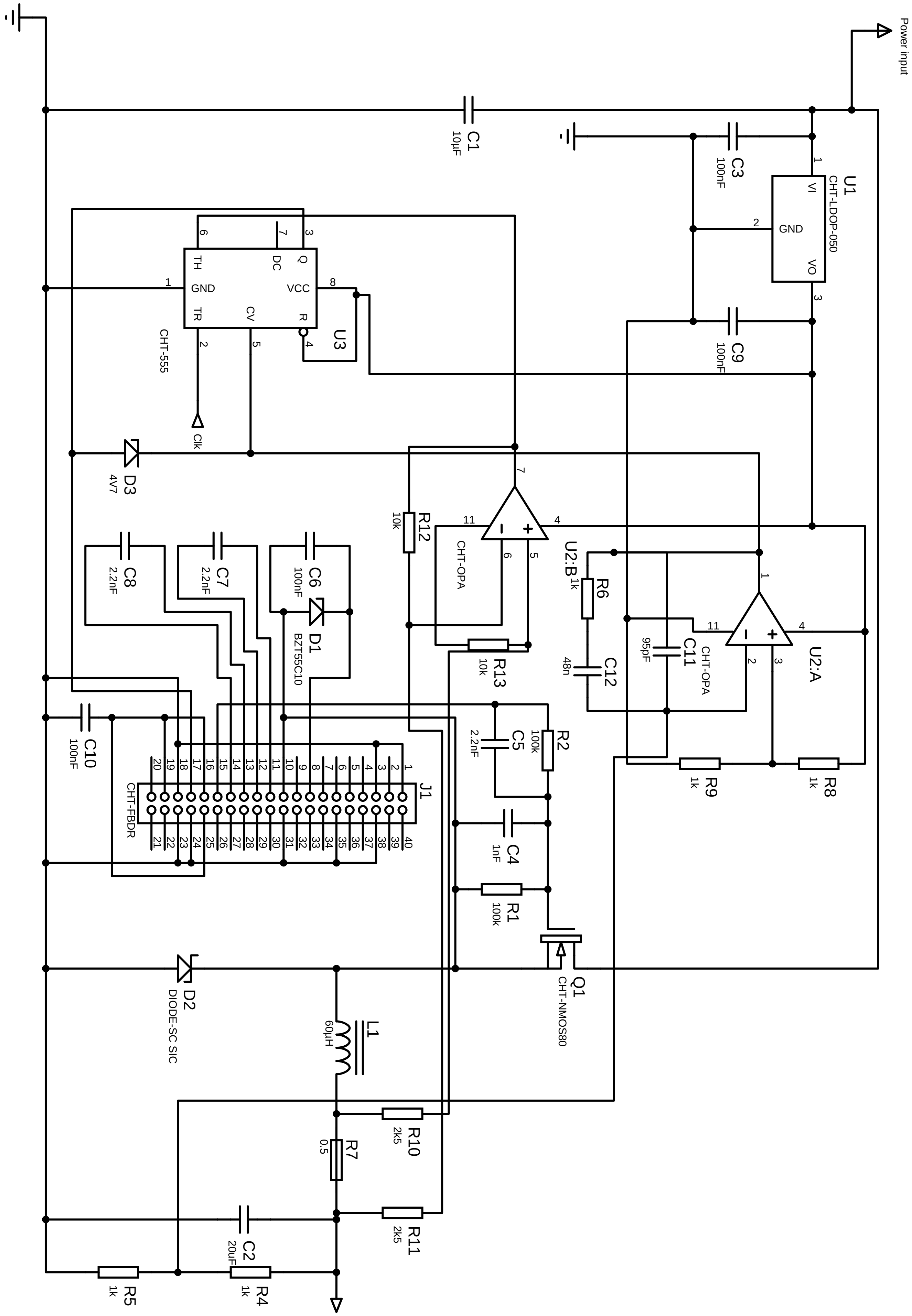
19. **Cristoloveanu, Sorin.** Silicon-on-Insulator Technology. [book auth.] Wai-Kai Chen. *The VLSI handbook.* Boca Raton, FL : CRC PRESS, 2007.

```
clear all
clc

disp('Converter Topology')
c=input('Choose convert topology(1=buck,2=boost): ');

if c==1
disp('You chose to design a buck converter')
Vmax=input('Maximum input voltage[V]: ');
Vmin=input('Minmum input voltage[V]: ');
Vo=input('Desired output voltage[V]: ');
Pmax=input('Maximum power output[W]: ');
Pmin=input('Minimum power output[W]: ');
fs=input('Desired switching frequency [kHz]: ');
fs=fs*1e3;

%Inductor calculations
ts=1/fs;
D=Vo/Vmax;
toff=(1-D)*ts;
Et=Vo*toff;
r=0.4;
LxIL=Et/r;
Io=Pmax/Vo;

L=LxIL/Io;
Irated=(1+r/2)*Io;

%Capacitor calulations
disp('The voltage ripple alowed is below 1% of the output voltage')
C=ts^2*(1-D)/(8*L*0.01);

% Controller calculations
R1=input('What is the feedback resistor value?[ohm]');
fcross=fs/3;
R=(Vo^2)/Pmax;
K=1;
G0=K*R;
%A=((1/R)+(1-0.5-D)/(L*fs))^-1;
%B=2;


fp=(2*pi*R*C)^-1;
fp0=fcross/G0;

L
C
R1
C2=(2*pi*R1*fp0)^-1
R2=(2*pi*C2*fp)^-1
C1=(5e-3*C/R2)

elseif c==2
disp('You chose to design a boost converter')
Vdmax=input('Maximum input voltage[V]: ');
```

```
Vdmin=input('Minmum input voltage[V]: ');
Vo=input('Desired output voltage[V]: ');
Pmax=input('Maximum power output[W]: ');
Pmin=input('Minimum power output[W]: ');
fs=input('Desired switching frequency [kHz]: ');
fs=fs*1e3;

cm=input('1=CCM, 2=DCM:')

%Inductor design
if cm==1
%CCM
D=(Vo-Vdmin)/Vo;
ts=1/fs;
ton=D*ts;
Et=ton*Vdmin;
LIL=Et/0.4;
IL=(Pmax/Vo)/(1-D);
L=(LIL/IL)/1e-6
Irated=1.2*IL

%Capacitor design
C=((D*ts)/(0.01*Vo^2/Pmax))/1e-6

%Controller design
m=1;
Rmap=2;
R=Vo^2/Pmax;
D=(Vo-Vdmin)/Vo;
R1=11e3
fesr=(2*pi*5e-3*C*1e-6)^-1

A=((2/R)+((1-0.5)*(1-D)^3)/(L*fs))^-1;
B=Rmap/(1-D);
Go=A/B;
fp=(2*pi*A*C*1e-6)^-1;

fcross=fs/24;
fp0=fcross/Go;
C2=(2*pi*R1*fp0)^-1
R2=(2*pi*C2*fp)^-1
C1=(2*pi*R2*fesr)^-1

elseif cm==2
% DCM
%Inductor design
ts=1/fs;
D=(Vo-Vdmin)/Vo;
Iomax=Pmax/Vo;
L=(ts*Vo*D*(1-D)^2)/(2*Iomax)
L=input('Set desired Inductance[H]: ')
%Capacitor design
C=((D*ts)/(0.01*Vo^2/Pmax))/1e-6

%Controller design
```

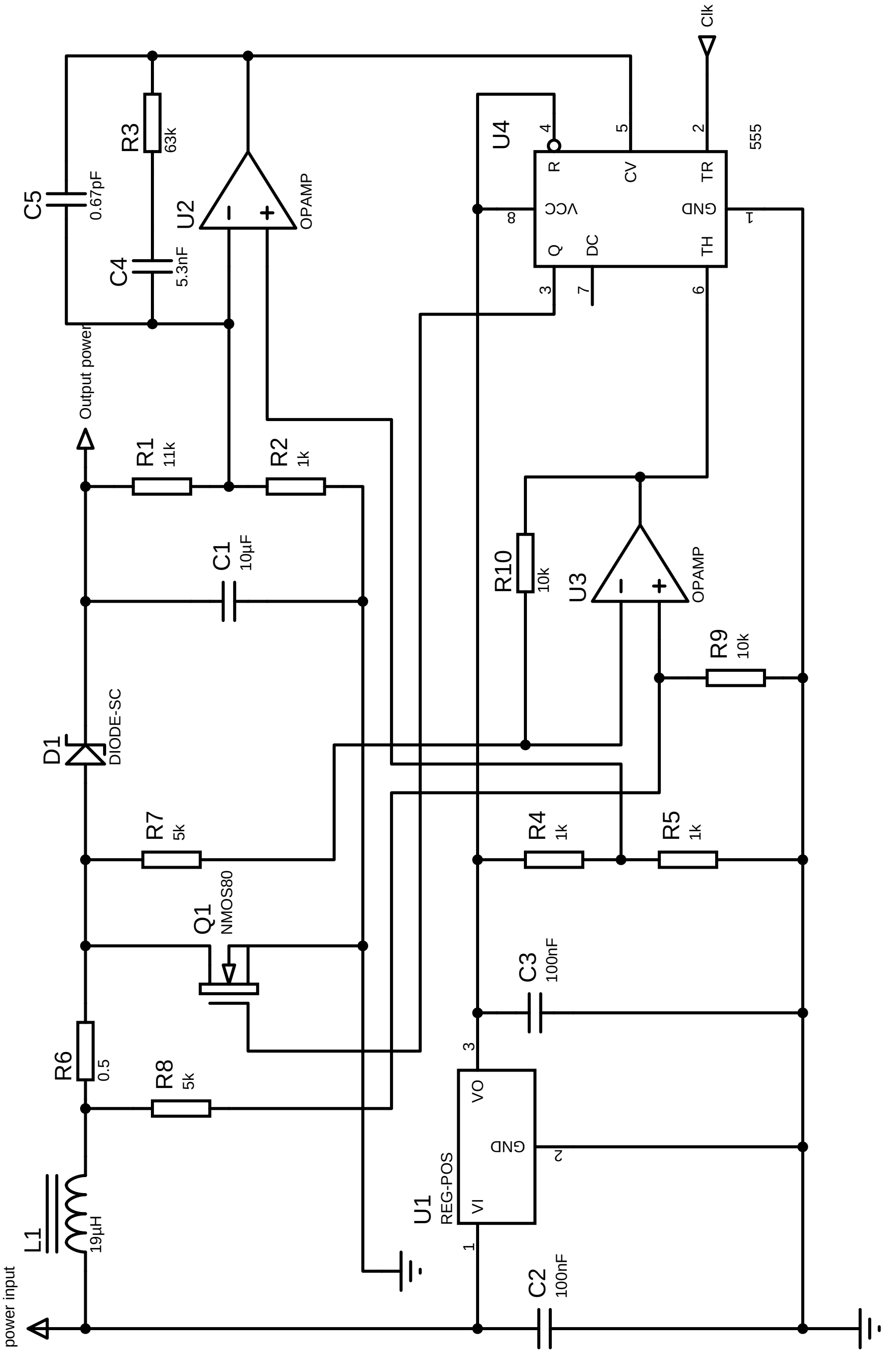
```
m=1;
Rmap=1.5;
R=Vo^2/Pmax;
D=(Vo-Vdmin)/Vo;
R1=11e3
fesr=(2*pi*5e-3*C*1e-6)^-1

A=((2/R)+((1-0.5)*(1-D)^3)/(L*fs))^-1;
B=Rmap/(1-D);
Go=A/B;
fp=(2*pi*A*C*1e-6)^-1;

fcross=fs/6;
fp0=fcross/Go;
C2=(2*pi*R1*fp0)^-1
R2=(2*pi*C2*fp)^-1
C1=(2*pi*R2*fesr)^-1

end




end
```

Appendix B

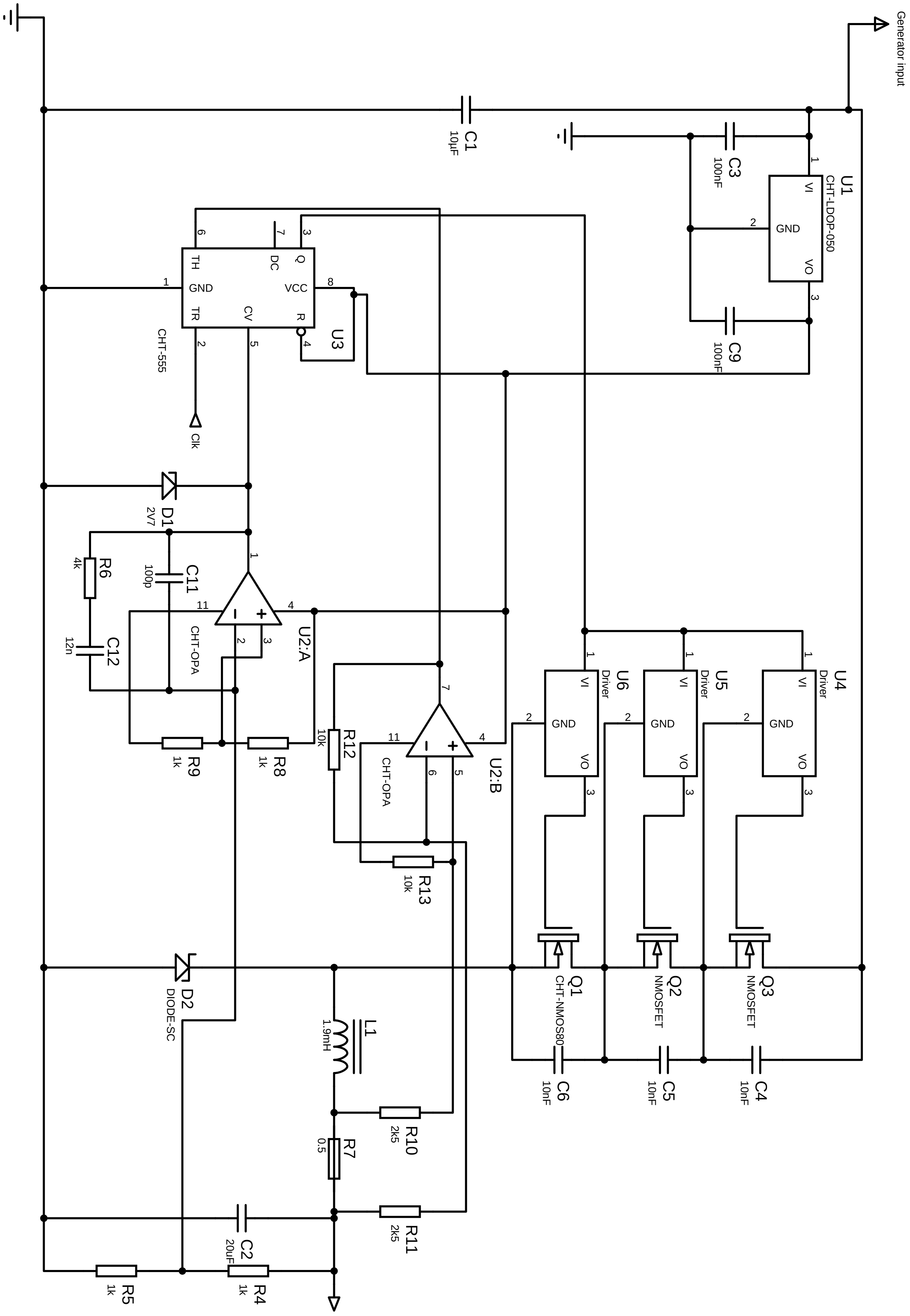


Power input
C1
10µF
C3
100nF
U1
CHT-LDOP-050
VI
GND
VO
1
2
3
C9
100nF
U3
CHT-555
TH
DC
Q
GND
VCC
TR
CV
R
6
7
3
1
8
2
5
4
Clk
D3
4V7
R12
10k
CHT-OPA
11
7
4
6
5
U2:B
R6
1k
C11
95pF
C12
48n
U2:A
CHT-OPA
1
11
4
2
3
R13
10k
C8
2.2nF
C7
2.2nF
C6
100nF
D1
BZT55C10
C5
2.2nF
R2
100k
R9
1k
R8
1k
C10
100nF
J1
CHT-FBDR
1
2
3
4
5
6
7
8
9
10
11
12
13
14
15
16
17
18
19
20
21
22
23
24
25
26
27
28
29
30
31
32
33
34
35
36
37
38
39
40
C4
1nF
R1
100k
Q1
CHT-NMOS80
D2
DIODE-SC SIC
L1
60µH
R10
2k5
R7
0.5
R11
2k5
C2
20uF
R5
1k
R4
1k

Appendix B


power input
L1
19µH
R6
0.5
R8
5k
Q1
NMOS80
R7
5k
D1
DIODE-SC
C1
10µF
R1
11k
R2
1k
Output power
C5
0.67pF
C4
5.3nF
R3
63k
U2
OPAMP
U4
555
R
CV
TR
VCC
GND
Q
DC
TH
4
5
2
8
1
3
7
6
Clk
R10
10k
U3
OPAMP
R9
10k
R4
1k
R5
1k
C3
100nF
U1
REG-POS
VO
GND
VI
3
2
1
C2
100nF

Appendix B


Generator input
C1
10µF
C3
100nF
C9
100nF
U1
CHT-LDOP-050
VI
GND
VO
1
2
3
U3
CHT-555
TH
DC
Q
GND
VCC
TR
CV
R
6
7
3
1
8
2
5
4
Clk
D1
2V7
D2
DIODE-SC
R6
4k
C11
100p
C12
12n
U2:A
CHT-OPA
11
4
1
2
3
R9
1k
R8
1k
R12
10k
U2:B
CHT-OPA
7
11
4
6
5
R13
10k
L1
1.9mH
R10
2k5
R7
0.5
R11
2k5
C2
20uF
R5
1k
R4
1k
U6
Driver
U5
Driver
U4
Driver
VI
GND
VO
Q1
CHT-NMOS80
Q2
NMOSFET
Q3
NMOSFET
C6
10nF
C5
10nF
C4
10nF