|  |
| --- |
| Group F |
| AC Motor Variable Frequency Drive |
| Senior Design 1: Spring 2016 |

|  |
| --- |
| William Santos, Justin Barwick, Chris Guido, Merritt Robbins  4-28-2016 |

**Contents**

[Executive Summary 1](#_Toc449571835)

[Project Description 2](#_Toc449571836)

[Motivation and Goals 2](#_Toc449571837)

[Qualitative Statements on AC vs DC Motors 2](#_Toc449571838)

[Initial Objective Statements 3](#_Toc449571839)

[Efficiency 3](#_Toc449571840)

[Integrated Circuit and Motor Protection 3](#_Toc449571841)

[Safety 4](#_Toc449571842)

[Real-time Data Analytics 5](#_Toc449571843)

[Implementation of true field oriented control 5](#_Toc449571844)

[Design Specifications 5](#_Toc449571845)

[Drive Dynamic Performance 6](#_Toc449571846)

[Sensor Distribution and Real time Data Processing 7](#_Toc449571847)

[Efficiency Requirements Specifications 8](#_Toc449571848)

[Power System 9](#_Toc449571849)

[Safety Devices and Protection Circuitry 10](#_Toc449571850)

[Related Research 11](#_Toc449571851)

[Existing Similar Projects and Products 11](#_Toc449571852)

[Relevant Technologies 12](#_Toc449571853)

[Pulse Width Modulation 12](#_Toc449571854)

[Trench IGBTs and Power MOSFETs 13](#_Toc449571855)

[High Efficiency Synchronous Rectification 13](#_Toc449571856)

[Powerful and Highly – Integrated Microcontrollers 13](#_Toc449571857)

[Computer Aided Design 14](#_Toc449571858)

[The AC Induction Motor (Asynchronous Motor) 14](#_Toc449571859)

[Introduction to Rotating Machines 14](#_Toc449571860)

[The AC Induction Motor 15](#_Toc449571861)

[Synchronous vs Asynchronous Motors 17](#_Toc449571862)

[AC Induction Motor Selection Considerations 17](#_Toc449571863)

[Space - Vector Control 20](#_Toc449571864)

[Overview of Methods of Speed Control 21](#_Toc449571865)

[Shortcomings of Volts / Hertz Control (Open Loop Control) 21](#_Toc449571866)

[The Direct – Quadrature – Zero Transformation (Park Transformation) 22](#_Toc449571867)

[Space – Vector Pulse Width Modulation Integrated Circuit 22](#_Toc449571868)

[Space Vector Modulation 22](#_Toc449571869)

[Determining the Inverter State Vectors and Vector Notation 23](#_Toc449571870)

[Developing the Pulse Width Modulated Signal: 25](#_Toc449571871)

[Equating the PWM Signal 27](#_Toc449571872)

[Evaluating the Average Phase Voltage for each Sector 28](#_Toc449571873)

[Rectifier Topologies 29](#_Toc449571874)

[Half – Bridge Diode Rectifier 29](#_Toc449571875)

[Full – Bridge Diode Rectifier 30](#_Toc449571876)

[Silicon Controlled Rectifier (SCR) 31](#_Toc449571877)

[Synchronous Rectifier (Active Rectifier) 32](#_Toc449571878)

[Power Inverter 34](#_Toc449571879)

[Overview of Inverter Circuits 34](#_Toc449571880)

[Selecting a Power Switch Type for the Inverter 36](#_Toc449571881)

[Specific Design Considerations 37](#_Toc449571882)

[Selection of Protection Device Values 37](#_Toc449571883)

[Implementing Active Ground Management 38](#_Toc449571884)

[Comparisons of Temperature Sensing Methods 39](#_Toc449571885)

[Usage of a Digital Control Algorithm 39](#_Toc449571886)

[The Current Sensing Device 41](#_Toc449571887)

[Choice to Favor Well Reputed IC Manufacturers 41](#_Toc449571888)

[System Block Diagrams 42](#_Toc449571889)

[Overall System 42](#_Toc449571890)

[Power System 43](#_Toc449571891)

[Sensor Management and LCD 43](#_Toc449571892)

[Related Standards 44](#_Toc449571893)

[Applicable to Electrical Components 44](#_Toc449571894)

[Other Applicable Standards 45](#_Toc449571895)

[Realistic Design Constraints 47](#_Toc449571896)

[Economic and Time constraints 47](#_Toc449571897)

[Estimated Budget Breakdown 47](#_Toc449571898)

[Environmental, Social, and Political Constraints 47](#_Toc449571899)

[RoHS Compliance 47](#_Toc449571900)

[Attempting a Fully Lead – Free Design 48](#_Toc449571901)

[Ethical, Health, and Safety constraints 48](#_Toc449571902)

[Compliance with Standards Regarding Grounding and Isolation 48](#_Toc449571903)

[Consideration of Dynamic Breaking 49](#_Toc449571904)

[Regenerative Breaking 49](#_Toc449571905)

[Hardware and Software Design Process 50](#_Toc449571906)

[The A/C Induction Motor 50](#_Toc449571907)

[Motor Selection 50](#_Toc449571908)

[Mathematical Modeling of the A/C Induction Motor 51](#_Toc449571909)

[The Clark Transformation Applied 51](#_Toc449571910)

[Pseudo-stationary -system (Park’s Transformation Applied) 55](#_Toc449571911)

[Modeling in the Continuous Time Domain 56](#_Toc449571912)

[Synchronous Bridge Rectifier and DC Link 58](#_Toc449571913)

[Synchronous Bridge Schematic Design: 58](#_Toc449571914)

[Selection of Power Switches: 64](#_Toc449571915)

[Estimate of expected efficiency: 66](#_Toc449571916)

[Specification of DC Link Bulk Capacitors 67](#_Toc449571917)

[Test Plan: Synchronous Bridge 68](#_Toc449571918)

[Low Voltage Bus Regulation 69](#_Toc449571919)

[Specification of Input Transformer 70](#_Toc449571920)

[Rectifier Specification and Selection 70](#_Toc449571921)

[15V Regulator IC Selection 71](#_Toc449571922)

[15V Regulator Circuit Topology and Passive Component Selection 73](#_Toc449571923)

[3.3V Regulator IC Selection 76](#_Toc449571924)

[3.3V Regulator Circuit Topology and Passive Component Selection 77](#_Toc449571925)

[Low Voltage Rail Regulation Test Plan 78](#_Toc449571926)

[Power Inverter 81](#_Toc449571927)

[Inverter Schematic Design 82](#_Toc449571928)

[Selecting IGBTs for power switching. 85](#_Toc449571929)

[IGBT Driver IC Specification and Selection 88](#_Toc449571930)

[Output Filter Design 90](#_Toc449571931)

[Over-Current Protection 90](#_Toc449571932)

[Test Plan: Power Inverter 90](#_Toc449571933)

[Sensor System – Temperature, Voltage, and Current 91](#_Toc449571934)

[Temperature Sensors 93](#_Toc449571935)

[Mechanical Assembly 98](#_Toc449571936)

[Transmission/Drivetrain 98](#_Toc449571937)

[Microcontroller Selection 99](#_Toc449571938)

[Microcontroller Basics 99](#_Toc449571939)

[Microcontroller Selection 99](#_Toc449571940)

[The MSP430F5529 100](#_Toc449571941)

[The TMS320F28027F 101](#_Toc449571942)

[Microcontroller Inputs 102](#_Toc449571943)

[Sensor – Microcontroller – Display Interface 102](#_Toc449571944)

[Constraints on Input Signals 103](#_Toc449571945)

[Input Protection Circuitry Design 105](#_Toc449571946)

[Microcontroller Input Logic and Data management 107](#_Toc449571947)

[Analog – Digital Converter (ADC) 107](#_Toc449571948)

[Basics of Data Storage and Management 108](#_Toc449571949)

[LCD Display 111](#_Toc449571950)

[Project Prototype Construction and Coding 111](#_Toc449571951)

[Bill of Materials 111](#_Toc449571952)

[PCB Vendor and Assembly 113](#_Toc449571953)

[Final Algorithm Structure 113](#_Toc449571954)

[LCD Display Control 113](#_Toc449571955)

[Project Prototype Testing 124](#_Toc449571956)

[Software Test Environment 124](#_Toc449571957)

[Software Specific Testing 124](#_Toc449571958)

[Administrative Content 125](#_Toc449571959)

[Milestone Discussion 125](#_Toc449571960)

[Appendix A – Copyright Permissions i](#_Toc449571961)

[Appendix B – References ii](#_Toc449571962)

[Appendix D – Table of Figures iv](#_Toc449571963)

[Appendix E – Computer Simulation Screenshots ix](#_Toc449571964)

[MATLAB – Simulink Model of Induction Motor Dynamic response ix](#_Toc449571965)

[Appendix F – Circuit Schematics xiv](#_Toc449571966)

[Input Transformers, Synchronous Input Bridge Rectifier, and DC Link xiv](#_Toc449571967)

[Low Voltage DC Rectifier and Rail Regulation xvi](#_Toc449571968)

[Power Inverter xviii](#_Toc449571969)

[MSP430/Piccolo Interface and Protection Circuitry xix](#_Toc449571970)

[Thermocouple Interface xx](#_Toc449571971)

# Executive Summary

The three – phase power system is innately more efficient in power transfer when compared to its one or two – phase counterparts. A single phase system uses two conductors to create one AC power loop to transmit one unit of energy. In a balanced, three phase system[[1]](#footnote-2) three AC power loops are formed using three conductors transmitting three units of power. The advantage is clear. The three phase system has another powerful attribute which directly enable this project; three time varying vectors placed as position vectors in a complex plane each described with, where , the resultant vector will rotate with the frequency of the sinusoids (). Furthermore, if these sinusoids drive electromagnets placed as before, they generate a rotating magnetic field with the same rotational velocity. This is the core principle of all three – phase AC motors; the rotating magnetic field generated by a three – phase sinusoidal power source.

The three – phase alternating current motor is the workhorse of modern industry. Its combination of high conductor efficiency, extreme reliability, smooth operation, and lack of on – site waste disposal makes AC motors extremely popular for industrial drives of all types. Compared to DC motors for similar applications, AC motors tend to have all of the big wins in technology except one; speed and torque control. Speed and torque control in DC motors is of trivial complexity; just limit the current going into the rotor and the torque will follow. Speed and torque control of an AC motor is not nearly as simple. The three – phase AC motor uses the intrinsic properties of the three – phase balanced system to create a rotating magnetic field inside the stator. The speed at which this field rotates directly impacts the motors rotational speed. However, the speed of rotation is a function of the input sinusoidal signal’s frequency. This simple fact makes speed control of AC motors a complex problem. Solving it requires varying the frequency of the sinusoidal power signals operating the motor – that in and of itself is immensely more complex than simply controlling a DC current.

The solution is the Variable Frequency Drive (VFD); a very mature subject in electric machinery. VFDs perform the task of AC motor speed and torque control, they vary not only the frequency of input but the currents or voltages on each phase as well. The combination of varying the frequency and varying the amplitude of the power signal into the motor allows the control of speed and torque. Being that this system has two outputs (frequency and amplitude) and has one input (desired torque or desired speed) the control system which will operate it would seem initially complex. The problem of AC motor control is not easily solved intuitively as the time varying signals make mathematical modeling of the motor to generate desired motor signals for Proportional – Integral (PI) controllers to use is very difficult and computationally expensive to implement in a stationary reference frame.

The solution implemented by Group F is to employ Field Oriented Control (FOC), also known as Vector control to dynamically model and control an AC induction motor of our choosing. Field oriented vector control was first enabled by a publication by R.H. Park in 1929 which detailed a mathematical model which could be implemented with a series of trigonometric reference frame transformations which allowed the motor torque and flux “components” of the sinusoidal power signals to be separated from each other as orthogonal, DC vectors from the origin in two dimensions [1]. This publication allowed for the first implementations of field oriented control. Because the flux vector and torque vector are orthogonal, the controller can hold one vector constant (usually the flux) and vary the other (usually the torque) independently. This allows the PI controllers to have a direct way to control just the torque which our motor will output. This fact alone simplifies control enough where the only methods which existed before its inception couldn’t account for motor transient response at all and had to just de-rate the components enough to bear it.

The VFD as implemented by Group F is a highly flexible design. By inputting nameplate data from any induction motor within a range of power ratings, we can dynamically model and control it with our system. Attached to a motor which tolerates extremely high turn down, we can expect exceptional dynamic performance from our drive. The drive could also be easily adapted to work with a synchronous motor as well. And the algorithm to control a synchronous motor would be slightly simpler given the lack of slip. The induction motor is just so much more prevalent in low to medium power applications.

# Project Description

A brief overview of the motivations, initial objectives, and specifications for the variable frequency drive which will be designed and built by Group F to be presented in the fall 2016 senior design showcase.

## Motivation and Goals

It is apparent to all members of Group F that providing reliable motor control to consumers and industry will be a problem in electrical engineering for the foreseeable future. The design of AC motor drives was not covered in any formal class either required or offered by the university undergraduate catalog for electrical engineering. Given the clear future that AC motors will have in human industry and daily life, particularly given the growing sector of green energy, which all hinges on electric power; the group feels that this is a worthwhile project with clear educational merits in a subject not covered in our formal education.

This project is not intended to be groundbreaking technology or novel design, it is intended to enable the group to learn the current state of the technology as used by innumerable companies and individuals to dynamically control AC motors. The rise of electric vehicles in modern transportation all depend on reliable electric drivetrains that enable extremely high efficiencies and enormous dynamic range. These achievements are only possible using field oriented control algorithms developed throughout the last century.

### Qualitative Statements on AC vs DC Motors

AC and DC motors differ in their modes of operation and associated drive requirements.  The primary difference, of course, lies in the characteristic of the power they accept.  AC motors require AC voltage/current supplies to function while DC motors require a DC power supply.  In addition, there are universal motors that can operate on both DC and AC power. These universal motors will not be considered for the purpose of this project

DC motors in general are more complex in their function, requiring additional components to function as designed. DC motors contain a brushed internal mechanical commutation in order to reverse motor windings’ current in synchronism with rotation.  It generates torque directly from DC power supplied to the motor via internal commutation, stationary magnets, and rotating electromagnets.  The advantages of a DC motor lie in their extremely high torque, and straightforward control of motor speed by varying the armature winding’s current.   Disadvantages include higher maintenance costs, as the commutator and brushes need to be replaced from time to time, and losses due to commutation.

The speed of AC motors is more difficult to control than DC motors, as the motor’s torque is a function of not only input current but also electrical frequency, current state of the motor’s magnetic fields, and the slip speed of the rotor.  However, AC motors are better suited for most variable-speed applications because of their low maintenance cost and higher power efficiency (both due to the lack of brushes and commutators).   In addition, they generally come in a wider variety of power ratings and are less expensive than comparable DC motors of a given power due to their simplicity of construction and lack of permanent magnets.

## Initial Objective Statements

### Efficiency

Efficiency is always a necessary consideration in electrical power system design for modern solutions. This is for good reason, as loss of power corresponds to monetary loss for the end-user in the form of overhead costs associated with part replacement due to heat damage and cycling. Perhaps more importantly; any power that is lost by the system is heat that needs to be dissipated or else the components will easily overheat. This results in larger component packages, increased demand for heat sinking of components, and shorter system life overall because of elevated operating temperatures. These are all undesirable features of performance in any engineered system, and should be avoided. Therefore, the Variable Frequency Drive for this project is designed to be highly efficient, approaching or exceeding 90% overall at full load condition.

To implement an efficient power system there is an overwhelming need to exploit the conduction characteristics of modern power switching devices to achieve the efficiency desired. Using algorithms detailed in the Design sections of this document, the team designed a system which implements both an active front end in the form of a synchronous bridge rectifier and a high efficiency motor inverter using IGBTs to efficiently rectify upwards of 100A of current at 350VDC. Both of these design elements exploit specific conduction characteristics of the switch type chosen to maximize efficiency of the final control and drive system.

### Integrated Circuit and Motor Protection

The inputs to our microcontroller will be sensitive to high voltage transients. Unfortunately due to the reality of our motor being a large inductor, and the fact that our inverter will create high rates of change in the current apparent in our motor windings; back EMF will be a required consideration. Back EMF or back electromotive force are quick, high voltage transients generated at the terminals of an inductor when a high rate of change is made to the current. This voltage would be applied in the opposite direction to the change of current and without proper input protection, can easily get through our power circuit into the control circuit if proper considerations for isolating and shunting the transients are not taken. These considerations and design of the appropriate protection circuits and diodes is detailed in the Input Protection Circuitry Design section for the microcontroller and the appropriate sections for other subsystems (Inverter drivers, Rectifier drivers, and Sensor Drives accordingly).

### Safety

Safety of the end user is a major priority in the design of the Variable Frequency Drive. To ensure safety for the final user, a focus on isolation and proper grounding is employed. The DC link is physically and electrically isolated from all other circuitry. This requires properly insulating all portions of the bus, using over – specified capacitors and contacts all capable of handling the high ripple caused by high current switching in the inverter. At the high – voltage AC side there is equal risk for electrocution. All contacts and locations of high voltage need to be properly labeled, protected, and insulated from ground potential. The terminals of the high voltage capacitors on the DC link will also need parasitic resistors placed in parallel to discharge the capacitors in a timely manner when not in use. It would be highly unsafe to allow high voltages to remain in the circuit after any more than a minute or two after disconnection from mains power.

As an added layer of smart safety, during the power up sequence, the DC link transformer is not energized, it is isolated from the inputs through MOSFETs which are only turned “on” after a rigorous set of tests is performed on the system by the microcontroller and sensor network. This will greatly reduce the risk of accidental electrocution as well as add another layer of protection to the circuitry itself and the motor being driven.

One design element which contributes both to usability and safety is dynamic breaking. Whenever the desired speed is set well below the measured motor speed, the controller can attach a breaking load to the motor windings to quickly decelerate the rotor. This feature mainly serves to reduce required user control by self – stopping every time the motor is turned off. Dynamic breaking is extremely helpful for applications with no load conditions during motor start and stop, such as machining.

The motor will also require protection circuitry. This will mainly be focused on over – current and over – temperature protection. Over current protection will entail a combination of fuses, breakers, and current monitoring to ensure proper current levels are always present in the operation of the motor. The current monitoring on the motor phases is already an important feedback source for the controller and the measurement will be used for both the controller and over – current protection. Further protection is offered by fuses and breakers in the power path which trip at a set level of current, protecting the motor from short – circuit conditions. Temperature monitoring will be essential for assuring that the motor is operating in a safe range. If we start having issues with inductive heating and losses in the windings due to high frequency harmonics in the drive signal this will allow us to see it hopefully before it causes any permanent damage to the motor winding insulation by overheating.

### Real-time Data Analytics

Essential to the performance of our drive, diagnosing failure modes, and maintaining a safe motor drive system are real – time data analytics performed by our microcontroller. The data sampled from various peripheral sensors on temperature sensitive, self – heating devices throughout the system is essential for safe operation. Some of the measured sensor data is also displayed on the liquid crystal display (LCD) screen for user information during normal operation. The LCD output will have two modes; standard and diagnostic. In standard mode the LCD will display essential real – time statistics on motor and drive performance. Specifically, the display will be focused on motor speed, torque, and power. The microcontroller will still service interrupts as they arrive for any over – temperature condition and warn the user of the issue when necessary. In normal operation the screen should be low – clutter and easy to interpret. In diagnostic mode the LCD will display all measurements available to the controller (all temperatures, voltages, and currents).

When saved and analyzed over time, the data collected during operation of our drive will allow characterization of the motor drive’s behavior over time and under various loading conditions. All with temperature data as a plus. We can see how our drive is performing through a locus of multiple conditions by analyzing this data. This analysis enables the group to tune the system to optimize its overall performance for a variety of factors, clearly a good tool for design revision and one which is essential to normal operation. All of the data analytics is performed on a Texas Instruments® MSP430 series microcontroller.

### Implementation of true field oriented control

Because of the immediately evident merits of field oriented control (FOC) methodologies over other strategies for AC motor control, it is a main objective of this project to implement true FOC. The group wants to be able to vary the torque at any motor speed within our operating range and maintain the motor’s flux density constant throughout that process. Being able to manage the field inside the motor is the main advantage which FOC employs over more classical control algorithms. The design of FOC algorithms will be the most essential feature of this project and will be the hinging point for project success overall.

## Design Specifications

The following specifications define the variable frequency drive designed by Group F. To qualify as a successful, “working” prototype, all of these design specifications must be met.

### Drive Dynamic Performance

#### Qualitative Specifications:

The variable frequency drive designed by Group F must achieve the following characteristics:

* At low speed, chattering or vibration should be minimal if any.
* Allow operation at full torque from 1Hz up to synchronous speed if the motor can handle it.
* Be capable of over – speed operation of the motor.
* Be adaptable to any three – phase induction motor within the appropriate power range.
* The parameters to change the motor specification should all be available as nameplate values.
* Final prototype will operate on single phase, 240VAC mains power.
* Will have an operating power factor of greater than 0.95.
* Dynamic breaking ability to quickly stop motor rotation is required.
* Stall detection and protection is required
* Ramping turn on of the motor is required
* Maximum overshoot acceptable for step response is 2%
* Steady state error of 0% for a ramp response

#### Quantitative Specifications:

In order to achieve the dynamic performance required of the final variable frequency drive, it must adhere to the following required capabilities:

|  |  |  |
| --- | --- | --- |
| VFD Dynamic Performance Quantitative Specifications | | |
| **Parameter (sym)** | **Value(s)** | **Motivation** |
| **Min. output frequency**  **()** | 1Hz | Should be an easy specification to achieve as a frequency. But filtering will determine how sinusoidal it appears. |
| **Max. output frequency ()** | 1kHz | Efficiency design may dictate that we use a 5kHz PWM for our inverter. Instead of 20kHz |
| **Selectable over-speed limit** | 0%-50% | Improve performance in applications where it is necessary. Sensor feedback allows safe over rated operation. |
| **Turn on time for system** | 10sec | Allows time for the system to perform all necessary checks and display a series of informational outputs. |
| **Rise time of motor turn on ()** | 0.5s-10s | Required variability in motor rise time for ramped motor start operation. Allows slow turn on of motors under heavy load |
| **Accuracy of Speed control** | 1% | To be considered successful vector control, the speed control accuracy must be exceptional. |

Table 1 - Dynamic performance required of the Variable Frequency Drive

### Sensor Distribution and Real time Data Processing

#### Qualitative Specifications:

The sensor network on the variable frequency drive must be capable of the following real – time measurements:

* Real – time dedicated measurements of motor mechanical speed, motor power consumption, motor stator temperature, and motor phase current.
* Inverter power switch temperature, input rectifier temperature all need to be monitored
* DC link voltage, low voltage DC (LVDC) bus voltage, motor phase voltage, and input power voltage.
* Watchdog timer control (WTC) will be implemented on all sensors to ensure consistent operation and fault detection.

#### Quantitative Specifications:

In order to achieve the desired qualitative specifications, the sensors on the drive will be placed according to the following table:

|  |  |  |
| --- | --- | --- |
| Sensor Distribution Specifications | | |
| **Location and Type** | **Quantity** | **Details** |
| Motor - Rotor Speed | 1 | Rotary encoder attached to motor shaft for real time speed measurement |
| Motor - Stator Temp. | At least 3 | At very minimum, one temperature sensor per phase is required to adequately sense the stator temperature. |
| Inverter - Heatsink Temp. | 4 | One on each high side IGBT, and one on the heatsink center of mass. |
| Rectifier - Heatsink Temp. | 2 | One placed on the center of mass and one near a switch |
| Motor - Phase Voltages | 3 | Each phase shall be probed at the inverter connection, assuming short lead length. Isolated probes. |
| Motor - Phase Currents | 2 | Measure two, calculate the third. |
| DC Link – Voltage | 1 | Isolated probe near the inverter |
| AC Input – Voltage | 1 | Locate at the power transformer. |
| Low voltage DC bus - Voltage | 2 | Probe each bus near the regulator, the microcontroller can automatically sense its input voltage. |

Table 2 - Sensor distribution required on the Variable Frequency Drive.

To achieve the requirements defined by the qualitative specifications, the real – time data processing capabilities of the system must adhere to the qualitative specifications defined in Table 3 below.

|  |  |  |
| --- | --- | --- |
| Real Time Data Processing Specifications | | |
| **Data** | **Latency** | **Motivation** |
| Motor - Rotor Speed | 25µs | The motor rotor speed must be known for each pulse of the PWM, therefore half the fundamental period is specified. |
| Motor - Stator Temp. | At least 3 | At very minimum, one temperature sensor per phase is required to adequately sense the stator temperature. |
| Inverter - Heatsink Temp. | 4 | One on each high side IGBT, and one on the heatsink center of mass. |
| Rectifier - Heatsink Temp. | 2 | One placed on the center of mass and one near a switch |
| Motor - Phase Voltages | 3 | Each phase shall be probed at the inverter connection, assuming short lead length. Isolated probes. |
| Motor - Phase Currents | 2 | Measure two, calculate the third. |
| DC Link – Voltage | 1 | Isolated probe near the inverter |
| AC Input – Voltage | 1 | Locate at the power transformer. |
| Low voltage DC bus - Voltage | 2 | Probe each bus near the regulator, the microcontroller can automatically sense its input voltage. |

Table 3 - Quantitative specifications for real - time data processing.

### Efficiency Requirements Specifications

#### Qualitative Specifications:

The efficiency of the variable frequency drive will allow for the following characteristics of the final prototype:

* Heatsink only required for the input bridge rectifier and power inverter.
* All surface mount voltage regulators use PCB as heatsink
* DC link parasitic resistors will be disconnected during normal operation with depletion mode MOSFETs.
* Switching will be managed in final prototype such that minimal losses are incurred from power device switching
* Normal load heatsink temperatures should not exceed 80°C for device safety.
* Drive must maintain reasonable efficiency at low loads
* The drive will employ all synchronous type regulators for DC rails

#### Quantitative Specifications:

In order to enable the qualitative specifications outlined above, the following quantitative specifications for variable frequency drive efficiency must be adhered to:

|  |  |  |
| --- | --- | --- |
| VFD Efficiency Quantitative Specifications | | |
| **Subsystem** | **Efficiency** | **Justification** |
| **Synchronous Bridge** | 97%\* | For this low frequency application, high 90% efficiency should be easily doable. |
| **DC Link** | 99% | The ESR of the capacitors as well as the leakage resistors. |
| **Inverter** | 94% | Switching losses in the IGBTs is significant. This is as low as it can be and still be overall efficient of 90% at full load. |
| **Overall Efficiency ()** | 90%\* | This is a highly demanding specification. Will require careful selection and control of the power switches in the inverter specifically. |
| **Low Voltage DC Regulators** | 90% (all loads)\* | Due to modern switch mode synchronous regulator topologies, efficiency of greater than 90% over an extremely wide range of loads is expected. |
| **Input Transformers** | 98% at full load | Reasonable expectation with high gauge wire, proper core material, and winding specification. |

Table 4 - Efficiency of the power subsystems required on the Variable Frequency Drive.

\*Note: This figure does not include transformer losses.

### Power System

#### Qualitative Specifications:

The power system of the variable frequency drive must be designed in order to conform to the following specifications:

* Ability to drive a 5hp rated motor over its entire loading range reliably
* Provide adequate power for up to a 10% overrated operation of the motor
* Deliver adequate power to all subsystems
* Input rectifier will employ actively controller power switches to maximize efficiency
* Near unity power factor for line side
* DC link will be a nominal 340-360V
* Inverter will be capable of 20 kHz switching frequency if desired by the control algorithm.

#### Quantitative Specifications:

To adhere to the specifications made above, the final prototype submitted for the variable frequency drive must adhere to the following quantitative specifications:

|  |  |  |
| --- | --- | --- |
| Power System Quantitative Specifications | | |
| **Parameter** | **Value(s)** | **Details** |
| **Input Voltage ()** | 240V | 120V outlets in a standard US circuit cannot supply the required power for a 5hp motor at full power. |
| **Max Input Current ()** | 20A | With 90% efficiency and a 0.95PF we can expect to almost fully load the supply circuit at full power. |
| **Max Rated Motor Power ()** | 4kW | Initial spec, our final motor will be around 5hp, 3.92kW. This allows power to be taking from a standard 240V household plug. Theoretically we could have upwards of 5kW before overloading the outlet’s circuit. |
| **Over Power Limit** | 10% | Allows moderate over-power of the motor, while still maintaining safe operating range. The rated power of a motor can generally be exceeded by ten percent or so. |
| **Thermal Dissipation\*** | 450W | A bit over the 415W that 90% overall efficiency would predict. However this allows for over – power operation of the motor. |
| **Max output ripple voltage ()** | 5% |  |

Table 5 - Output power capabilities of the Variable Frequency Drive

### Safety Devices and Protection Circuitry

#### Qualitative Specifications:

The variable frequency drive must employ a safety and protection system which ensures the following specifications are met or exceeded by the final prototype:

* Processing of all sensor temperature data and operation critical functions have watchdog timer controls employed (WTC)
* Inverter output is tolerant of short circuit for at least 5ms
* Stall detection and rapid correction/power down of motor
* Fusing on all power connections interfacing with motor and mains
* Tests for short circuits are performed on all start-ups using voltage probes as continuity test probes during start up.
* Ground fault detection is required at all times
* Over temperature warning and auto – shutdown.
  + Both programmable for different motor ratings.
* Circuit breaker placed on AC input to allow a resettable over – current protection in series with the fuse.

#### Quantitative Specifications:

In order to meet the qualitative specifications above, the variable frequency drive must adhere to the following quantitative specifications for the specific elements of the protection system:

|  |  |  |
| --- | --- | --- |
| Safety Device and Protection System Quantitative Specifications | | |
| **Device/Method** | **Value(s)** | **Details** |
| **Input power breaker** | 20A | Protects the house building circuit to its rated level. 240V is generally a 20A circuit in the US. |
| **Input power fuse (mostly protects the HV transformer)** | 30A | We can see inrush currents of over 30A so the fuse needs to be selected appropriately to avoid false trigger |
| **LV transformer primary fuse** | 1A | Standard fuse, protects from large, sustained fault in transformer windings. |
| **Motor output fuse** | 150A | Protects from major faults only, prevents propagation of a major fault. |
| **Ground fault detector** | -- | Current sensing comparator on the ground path which will trigger if current is greater than 0.01mA for longer than a few microseconds |
| **Switch over – current** | -- | Most modern power switches have built in thermal limits which fold back output current if the junction temperate exceeds a set value. |
| **DC link parasitic resistors** | 5s | Choose resistors such as to discharge the DC link to less than 2V in 15 seconds. |
| **Motor current sensors** | 2 | Will be used to detect over – current on motor windings as well as control function. |

Table 6 – Safety and protection devices specifications.

# Related Research

Initially, the group had a very broad understanding of the theory of variable frequency motor control. The project was intended from initial proposal to be focused on topics that were not covered as part of our curriculum. As a result, Group F conducted a significant amount of research to gain an understanding of the theoretical operation of the control algorithm, sensor implementation, and power electronics design. The succeeding sections offer a summary of the research which was conducted to equip the group to tackle the implementation of a highly efficient variable frequency drive (VFD)

## Existing Similar Projects and Products

There are currently many products on the market which achieve highly efficient space vector control algorithms in small packages at reasonably affordable prices. The core of our project being a space-vector AC motor controller makes it highly similar to any motor controller on the market today targeted at high dynamic range, power output, efficiency, and compact size.

Beyond the motor controller itself, the rapidly growing number of Electric Vehicles in the market today all use AC induction motors as the heart of their drivetrain. This is the main inspiration for the project. The rapidly advancing market for EVs and their accelerating development clearly illustrates that the systems which enable EVs will be instrumental for any Electrical Engineer wishing to contribute to these fields.

Some specific examples of current electric drive packages supplied for use today are the motor and drive packages supplied by EVDrive. EVDrive provides high quality controllers with high rated, matched power components. They package these drives with specific motors with 150V – 600V battery pack voltages. Higher voltages correspond to higher power capability as would be intuitively expected [2].

Given that we are designing a motor driver which will be centered on a power system to drive a three – phase AC induction motor, power concerns are of paramount importance. A discussion of EVs cannot take place without mentioning the current forefront of EV technology; Tesla Motors (NASDAQ: TSLA). According to Tesla’s website, the Tesla Model S curb weight is 4,647.3 lbs. Its power is battery discharge limited to 315hp in the 70kWh battery version [3]. Equation 1 shows the power to weight ratio achieved by Tesla on their standard Model S sedan.

Equation 1 - The power to weight ratio of the Tesla Model S. Values from Tesla.com

## Relevant Technologies

Many technological advancements in manufacturing and methodologies of design have taken place which have a direct impact on the feasibility of the VFD as a project for a team of undergraduate students. The following sections offer a brief discussion of the most impactful advancements which enable this project.

### Pulse Width Modulation

Pulse width modulation allows the synthesis of sinusoidal signals with very high efficiency by operating the switches in full conduction mode to minimize losses. A pulse width modulated signal uses a constant frequency on the rising edge of a pulsed signal, and the time that the pulse stays on is specified as the pulse width. The key specification of a PWM signal are the high voltage amplitude, and the duty cycle. The duty cycle is defined as the ratio of the time the switch is on to the total time elapsed, as illustrated below:

Equation 2 - Definition of the Duty Cycle of a PWM.

By varying the duty cycle of an output the designed can synthesize a large number of output waveforms. For our purpose we want to make a quasi – sinusoidal output power signal to drive our motor and therefore we will use sinusoidal PWM. To create a sinusoidal PWM we will use the theories outlined in the Space Vector Modulation section of this document combined with the Power Inverter to enable the 3.729kW of specified maximum power to be safely delivered to the motor being driven.

### Trench IGBTs and Power MOSFETs

Modern advances in transistor technology have produced numerous exceptional power switching devices. MOSFETs have become ubiquitous in an enormous range of applications from CMOS to power switching, due to their high efficiency at low loads, and fast switching capabilities. IGBTs still hold one niche of application quire securely however; Motor drives. IGBTs are used extensively in motor drives simply because of their loss characteristics. IGBTs have a diode – like voltage drop under load where MOSFETs losses are characterized as resistive. This characteristic alone makes IGBTs exceptional for high power applications. An exponential output current with respect to voltage drop versus a linear model for MOSFETs means that at low current levels (through observation and estimation anywhere under 25A) MOSFETs have higher power transfer efficiency, and at high current levels (like the ones in a PWM motor drive) the IGBT is far more efficient.

Furthermore, the temperature coefficients of IGBTs are much lower than MOSFETs, boosting their efficiency under heavy loads even further. Because of the combination of high efficiency and low temperature coefficient makes the IGBT the ideal transistor for switching the high currents we will be dealing with on our inverter.

### High Efficiency Synchronous Rectification

One of the applications for which MOSFETs yield themselves particularly useful is their use as synchronous rectifying switches in an attempt to simulate a more ideal diode model in a full bridge rectifier. The power MOSFET is exceptionally well fitted to this application because of two main reasons. First, the MOSFET has the built in body diode intrinsic in its design. That diode will allow the MOSFET to conduct current from source to drain before the gate is driven. This voltage drop triggers sensing circuitry in the MOSFET drivers and turns the switch on to drastically reduce the power dissipation for the rest of the conduction cycle and then stays off to block the positive drain voltage which will follow from the other half of the input signal being rectified. The IGBT also has a built in diode in most packages, however the IGBT cannot conduct current from Emitter to Collector in the on mode, preventing them from being implemented in the same fashion as MOSFETs in this application. MOSFETs also possess resistive loss characteristics when in the non - saturated mode as described above and therefore perform well under low instantaneous power applications such as the rectifier.

### Powerful and Highly – Integrated Microcontrollers

The amount of logic that will be driving the functionality of this project makes the use of microcontrollers essential for success. In order to finely tune the output of our system to our specific design, we will need to be able to perform complex calculations in a very short amount of time. These computations will be dependent on multiple variables that we will need to consider. These variables include the input our system is receiving, the desired output that an ideal system should be outputting, and the actual output of our system. To provide the most optimal and precise control of our system, we will need to perform these calculations incredibly fast so that there is very little room for error. This is where a high power microcontroller unit will be useful.

In order to properly monitor our system, we will be implementing an LCD display for our project. The purpose of this display will be so that a user will be able to keep track of various conditions in the system, such as the temperature of various components or the speed of the motor itself. In order to drive this LCD display, we will require another microcontroller. However, controlling an LCD display is not a very power-intensive task. This means that using a high power microcontroller would more than likely be a waste of resources. Luckily, there is a myriad of types of microcontrollers for all sorts of applications, ranging for high power applications such as motor control or driving power line modems, to low power applications like monitoring a timer for a digital clock or driving an LCD display.

### Computer Aided Design

The usage of MATLAB, Simulink, KiCad, LTspice, and Microsoft Excel were all instrumental in the design of the variable frequency drive. Without these aids the calculation involved in iterative design would have been tedious at best. More importantly the real time calculations required to analyze and control the AC motor via space – vector control require significant design using appropriate IDEs and testing environments, this design is greatly enabled via computer simulations.

## The AC Induction Motor (Asynchronous Motor)

As the main inspiration for this project is the AC induction motor itself being such an amazing machine, the succeeding sections explore the AC induction motor conceptually. This information is essential to understanding the conceptual elements of AC motor functionality and the basis for how the VFD can control the output of an AC motor.

### Introduction to Rotating Machines

The basic principles common to both AC and DC machines is worth discussing for a development of relevant knowledge towards our design goals.  As we begin to analyze the AC induction motor, the following equation provides insight into the general goal of machine design:

Equation 3 - Induced Electromotive Force (Faraday – Lenz Law)

Where , the induced electromotive force, and λ is the magnetic flux linkage through a surface. This equation can be utilized to determine the voltages induced by time-varying fields per the work of Faraday, the direction (the negative sign) is given by Lenz’s Law.  It relates the induction of electromotive force (operates analogously to an induced potential) to the change in the flux resulting from any source. In a motor the change in flux is created by mechanical motion of the stator with respect to the rotor’s magnetic field. The flux linkage of the motor is defined as the following:

Equation 4 - Expression of Flux Linkage

Where N is the number of turns in the coil windings of our machine and φ is used to indicate the instantaneous value of a time-varying flux. Within electric rotating machines, voltages are yielded via windings or groups of coils by rotating these windings mechanically through a magnetic field, through mechanically rotating a magnetic field past the winding, or by designing the magnetic circuit such that reluctance of the motor varies with rotation of the rotor.  In utilizing any of these methods, the flux linking a particular coil is varied periodically, and a time-varying voltage is generated.

The term *armature winding* refers to these sets of coils.  Generally, this term is used to refer to a specific winding or set of windings on a rotating machine which carry ac currents.  In ac machines like synchronous and induction machines, the armature winding is usually on the stationary part of the motor, referred to as the stator, and so are usually referred to as stator windings.  In synchronous and dc machines there is usually a second winding or set of windings that carry dc current and are used to produce the main operating flux in the machine.  Such windings are referred to as the *field winding*.  On a dc machine these are found on the stator, while on a synchronous machine it is found on the rotor. In the case of the synchronous machine, current must be supplied to the field winding via a set of rotating mechanical contacts (slip rings).

Most stators and rotors of rotating machines are comprised of electrical steel, and the windings are installed in slots on these structures.  The use of a high-permeability material, such as electrical steel, maximizes the coupling effects between coils, hence increasing the magnetic energy density associated with the electromechanical interaction. The windings are usually comprised of enameled copper wire in the effort to reduce I2R heat losses and degradation of the coils.

### The AC Induction Motor

The AC induction motor is the brainchild of the renowned Nikola Tesla, and has a number of applications including industrial fans, pumps, compressor loads, and electric vehicles. Its application in electric vehicles is the primary motivation for this project. To summarize its operation, the AC induction motor, also known as the asynchronous motor, makes use of a rotating magnetic field in the stator (stationary motor casing) to produce the electric current in the rotor (rotating part of the motor) required to produce torque applied on the shaft of the motor. In essence, the rotating magnetic field “drags” the rotor around. The magnetic field of the stator winding yields this current via electromagnetic induction. Because of this principle in operation, the induction motor does not need mechanical commutation to produce torque, and does not require separate excitation schemes for a portion of or all of the energy transferred from stator to rotor. This is a requirement in universal, DC and large synchronous motors.

The cage of the induction motor can be either wound type or squirrel-cage type. In a wound type rotor, the rotor windings are connected to slip rings, which are in turn connected to external resistances. By changing these external resistances, the speed and torque of the wound type induction motor can be controlled. In addition, the rotor of the wound type rotor contains more winding turns, resulting in higher induced voltage and lower current than the squirrel-cage rotor. The squirrel-cage motor, on the other hand, exhibits long conductive bars that are usually made of aluminum or copper. These bars are set into grooves and connected at both ends by shorting rings to form a shape akin to a cage, hence its name.

By definition then, the AC induction motor is asynchronous, meaning that it does not run at a speed synchronized with the frequency of the AC voltage that is supplying it. This is due primarily to the fact that, in order for the currents in the rotor windings to be induced, the rotor has to slip with respect to the asynchronous frequency. Hence, this phenomenon is known as the slip speed, and it varies with the amount of mechanical load that the motor is driving.

The versatility of the AC induction motor provides the primary motivation for an electric vehicle system.  AC motors can be divided into two main categories: synchronous and asynchronous motors.  The AC induction motor is the most common form of asynchronous motor, meaning that rotor speed is not varied directly with the frequency of the phase voltages providing it power.  Instead, there is a slip speed associated with the motor.  This phenomenon, along with synchronous motors, will be discussed in this document.  Pioneered by Nikola Tesla in 1888, this ingenious motor design enjoys a number of applications in both fixed-speed and variable-speed services.  Home applications and appliances include clocks, power tools, disk drives, washing machines, audio turntables and fans.  In industry, these motors are used in pumps, blowers, conveyors, compressors, air conditioning units and electric vehicles. The selection of operation mode, that is, fixed-speed or variable-speed, is crucial to devising the proper control scheme for the drive system of the motor.  In the case of electric vehicles, variable speed control is a must.  The following image shows a model of Tesla’s first induction motor:



Figure 1 - A model of Tesla’s first induction motor. Tesla museum in Belgrade, Belgium

To summarize its operation, the AC induction motor, also known as the asynchronous motor, makes use of a rotating magnetic field in the stator (stationary motor casing) to produce the electric current in the rotor (rotating part of the motor) required to produce torque applied on the shaft of the motor.  In essence, the rotating magnetic field “drags” the rotor around. The magnetic field of the stator winding yields this current via electromagnetic induction.  Because of this principle in operation, the induction motor does not need mechanical commutation to produce torque, and does not require separate excitation schemes for a portion of or all of the energy transferred from stator to rotor.  This is a requirement in universal, DC and large synchronous motors.  In identifying a housing for the motor, the cage of the induction motor can be either wound type or squirrel-cage type.  In a wound type rotor, the rotor windings are connected to slip rings, which are in turn connected to external resistances.  By changing these external resistances, the speed and torque of the wound type induction motor can be controlled.  In addition, the rotor of the wound type rotor contains more winding turns, resulting in higher induced voltage and lower current than the squirrel-cage rotor.  The squirrel-cage motor, on the other hand, exhibits long conductive bars that are usually made of aluminum or copper.  These bars are set into grooves and connected at both ends by shorting rings to form a shape akin to a cage, hence its name.  In our design, we will be using a three-phase squirrel-cage motor.

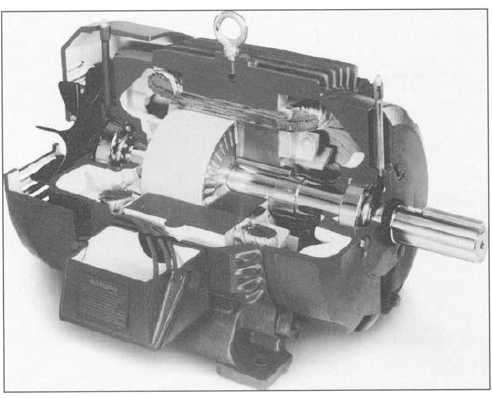


Figure 2 - Cutaway view of a three-phase squirrel-cage motor. Reprinted with permission from Rockwell Automation/Reliance Electric

### Synchronous vs Asynchronous Motors

In the world of three phase AC machines, there are two major groups. Synchronous motors, and asynchronous motors. Synchronous motors always rotate at their synchronous speed, in other words the rotor mechanical speed always equals the stator electrical speed. These motors use either permanent magnets in the rotor, or have a DC source drive the rotor windings to generate the rotor magnetic flux. Asynchronous motors, also referred to as induction motors, do not have this property; the rotor will never spin at the same mechanical speed as the stators electrical speed. This is due to the necessity of inducing the rotors current. Without a changing magnetic flux in the rotor there will be no induced current, thus requiring that the rotor moves with respect to the stators electrical frequency to create the varying magnetic field which induces the rotor current. [8]

### AC Induction Motor Selection Considerations

The selection of an appropriate induction motor with which to perform space vector control is of critical importance to this project. The power output, rated voltage, and operating frequency are of particular interest because these parameters will dictate the selection of components and the development of the initial electrical model to provide power to the MCU, Inverter and other stages that require low voltage power. A discussion on the parameters outlined in a motor’s nameplate will be provided so that parameter selection is clear.

The National Electrical Manufacturer’s association, or NEMA was established in 1926 to promote the standardization of electrical parts, apparatus, supplies, etc. Critical in ensuring that motors are interchangeable in systems is making sure that nameplate information is common among manufacturers. As such, this information enables engineers and installation/maintenance personnel alike to quickly understand and recognize exactly the type of motor one is dealing with. The National Electric Code, or NEC states that the motor nameplate must show the following parameters:

* Rated voltage or voltages
* Rated full-load amperage for each voltage level
* Frequency
* Phase
* Rated full-load speed
* Insulation class and rated ambient temperature
* Rated horsepower
* Time Rating
* Locked-rotor code letter
* Manufacturer’s name and address

A number of nameplate configurations also include data like bearing identification numbers, certification code, manufacturer serial number, and symbols/logos. Each motor nameplate parameter will be discussed so as to identify the importance of these parameters in motor selection. [4]

The rated voltage refers to the specific voltage level or combination of voltage levels required for optimal performance. In the consideration that voltage changes on a power distribution system occur due to changing load conditions within facility/utility supplies, motors are designed with a 10% tolerance for voltage above and below the rated nameplate value. For example, a motor with a rated nameplate voltage of 460V should be expected to operate successfully between 414V and 506V.

The rated full load amperage refers to the rated amperage when the motor is in full-load torque and horsepower mode. As the torque load on a motor increases, the amperage required to power the motor also increases. This value is determined by laboratory tests performed by the manufacturer and is usually rounded up slightly to be recorded as the nameplate value. The rounding allows for variations that can arise in manufacturing processes and normal voltage variations that might increase the full-load amperage of the motor. Full-load amperage is used to select correct wire size, motor starter, and overload protection devices necessary to protect the motor.

The rated full-load speed refers to the motor’s approximate speed at full-load conditions: when voltage and frequency are at rated values. A slightly lower number than the actual laboratory test result figures is stamped on the nameplate due to the fact that this value can change slightly in response to factors such as manufacturing tolerances, motor temperature, and voltage variations. On standard induction motors, the full-load speed is typically 96% to 99% of the no-load speed [4] .

A very important parameter that dictates motor life is the maximum temperature that occurs at the hottest portion of the motor. The temperature that occurs at that spot is a combination of motor design (temperature rise) and the ambient or surrounding temperature. The standard way of indicating these components is by showing the allowable maximum ambient or surrounding temperature, which is usually on the order of 40 degrees Celsius or 104 degrees Fahrenheit, and the class of insulation used in the design of the motor. Available motor classes are A, B, F, and H. The following table summarizes these temperature tolerance classes.

|  |  |  |  |  |
| --- | --- | --- | --- | --- |
| **Temperature Tolerance Classes of Induction Motors** | | | | |
| **Temperature Tolerance Class** | **Maximum Operation Temperature** | | **Allowable Temperature rise @1.0 service factor** | **Allowable Temperature rise @1.15 service factor** |
|  | *Degrees Celsius* | *Degrees Fahrenheit* | *Degrees Celsius* | *Degrees Celsius* |
| A | 105 | 221 | 60 | 70 |
| B | 130 | 266 | 80 | 90 |
| F | 155 | 311 | 105 | 115 |
| H | 180 | 356 | 125 | - |

Table 7- Allowable temperature fluctuations based on temperature tolerance class

Horsepower is the measure of how much work a motor can be expected to perform. This value is calculated from the motor’s full-load torque and full-load speed ratings. The following equation illustrates its formulation:

Equation – The mathematical/physical definition of horsepower

Standard motors are rated for continuous duty (24/7 operation) at their rated load and maximum ambient temperature. Specialized motors can be designed for “short-time” requirements where intermittent duty is all that is required. These motors can carry a short-time rating from five minutes to sixty minutes. The National Electrical Manufacturer’s association definition for short-time motors is as follows: “All short-time ratings are based upon corresponding short-load tests, which will commence only when the windings and other parts of the motor are within five degrees Celsius at the time of the test.” By utilizing short-time ratings, it becomes possible to reduce the size, weight and cost of the motor required for certain applications.

In addition to the common nameplate data on a motor, motor manufacturer’s might also include frame size, NEMA design letter, service factor, full – load capacity, and power factor.

Under the National Electrical Manufacturer’s Association’s system of standards for nameplate information, most motor dimensions are standardized and categorized by a frame size number and letter designation. In fractional horsepower motors the frame sizes are two digits and represent the shaft height of the motor from the bottom of the base in sixteenths of an inch. For example, a 56-frame motor would have a shaft height (“D” dimension) of 56/16 of an inch, or 3.5 inches [4].

The National Electrical Manufacturer’s Association design letter designation on a motor nameplate yields the specialized performance characteristics for a given motor application. For example, cranes and hoists that have to start with full loads imposed on the shaft of the motor may require motors with operating characteristics much different from what is required for pumps, blowers, and compressors with variables loads applied to the shaft. Motor performance characteristics can be altered by design changes in lamination, winding, rotor, or any combination of these three items. In most standard motors for general-purpose applications the specifications are over-met for design B motors. Design A and B motors are sometimes utilized in applications characterized by high breakdown, or pull-out torque requirements. Injection molding machines are an example of such an application. Design C motors are selected for applications that require high starting, or locked-rotor/inrush torque such as inclined conveyors [4].

The service factor (SF on the nameplate) is a measure of how much overload a motor can withstand when in normal operation within the correct voltage tolerances. It is readily advisable to avoid operation continuously above the rated load in the service factor area. Many motors might not provide the right amount of initial and breakdown torques, and erroneous sizing of the starter becomes a possibility.

Full-load efficiency is calculated as a percentage and illustrates how well the motor converts electrical power into mechanical power. In general, the larger the motor, the more efficient the motor. At one horsepower, three-phase motors can be rated up to 86.5 percent while a 300 horsepower motor can be as efficient as 95.8 percent.

The power factor presented on a motor nameplate is the ratio of motor load watts divided by volt-amps at the full-load condition. It is known that power factor is minimum in a no-load or low-load condition and that this value goes up as load is applied to the motor [4].

The following tables provide a comparison of nameplate motor parameters among possible motors selected for design: including the first fractional horsepower chosen for initial testing and motor selection research.

## Space - Vector Control

The advancement in computational power of microelectronic microprocessors and DSP’s has enabled the development of very precise digital vector control algorithms that significantly impact the reliability of the AC drive system. A common digital implementation method of vector controls is field oriented control, which allows for direct torque control. Direct torque control of the induction motor allows for increased ease in handling system limitations and also provides higher power conversion efficiency. The succeeding sections provide an overview of the research conducted on vector control methodologies.

### Overview of Methods of Speed Control

|  |  |  |  |
| --- | --- | --- | --- |
| **Comparison of the Three General Categories of Speed Control** | | | |
|  | **Direct Connection (on/off)** | **Volts / Hertz Control** | **Vector Control** |
| **Variable Speed** | NO | YES | YES |
| **Feedback Requirements** | NO | NO – Open loop designed to drive the desired steady state output values. | YES – Vector control requires feedback of motor phase voltage, current, and mechanical speed. |
| **Complexity of Design** | LOW – It’s just a switch | HIGH – Managing the transient behavior of the motor requires special considerations for protection circuitry to ensure reliability. | HIGH – The nature of a MIMO control system are innately complex in theory and practical implementation. |
| **Flexibility in Application** | HIGH - On/off control can work with any motor as long as the power supply meets motor specifications of voltage, frequency, and current. | LOW – Controller is designed with specific application and motor, cannot be easily adapted to a new motor or loading condition | HIGH – The feedback allows the controller to dynamically control the motor through all operating conditions. Allows adaptability to any three phase synchronous or asynchronous motor with appropriate power peripherals. |
| **Cost of Implementation** | VERY LOW – It’s just a switch. | HIGH – The components need to be massively over – rated because of uncontrolled transient behavior of the motor | MED – Dependent mainly on power components as the control system is merely a microcontroller and two PI controllers. No over – rated components needed as transient behavior is also controlled. |

Table 8 - Overview of the different methods of AC motor speed control.

### Shortcomings of Volts / Hertz Control (Open Loop Control)

Classical control algorithms for AC induction motors relied on the modeling of an AC motor at steady-state conditions, calculation of the required phase frequency and current to produce desired effects, and then applying those desired references to the stator windings using either PI control or a hysteresis comparator. There were many issues with this method including uncontrollable transient voltage and current, difficulty in using a sinusoidal reference for PI control, and a lack of general applicability in design; a controller must be specifically designed for asynchronous or synchronous motors. These drawbacks resulted in higher costs of construction, lower efficiency, and larger size [5].

### The Direct – Quadrature – Zero Transformation (Park Transformation)

Instrumental in the development of motor control systems was the vast simplification offered through application of the Direct – Quadrature – Zero (dq0) transformation, also referred to as the Park Transformation. The Park Transformation involves two axis projections and the spinning of the reference frame of the machine. With three inputs being sinusoidal varying voltages of a three-phase balanced AC machine; **a**, **b**, and **c**. The axis projections on their own result in three orthogonal vectors, **α**, **β**, and **Z**. Conveniently the **Z** vector always has magnitude equal to zero in a balanced three-phase system. This simplifies the three sinusoidal inputs of the system into two orthogonal DC values spinning about the **Z** in a stationary reference frame vector with frequency equal to synchronous speed. This transformation on its own is the referred to as the Clark Transformation or the αβ0 Transformation, and is useful in its own right, but further simplifying by spinning the reference frame about the **Z** vector at synchronous speed makes the old **α** and **β** vectors appear stationary in the reference frame, these new vectors are the **d** and **q** vectors we desire. [1]

The inverse Park transformation allows three time-varying vector quantities to be expressed as two orthogonal, non-time variant vectors in a new rotating reference frame, and the forward Park Transform does the opposite, allowing transformation freely between **d-q** quantities and **a-b-c** values.

In practice, this transformation will be essential in our control algorithm. It will allow us to model the motor itself in a far less mathematically intensive way, converting the phase voltages into DC values allows amazing simplification of the rest of the motor simulation calculations as it removes the frequency and phase between voltages, removing the trigonometric terms from future calculations. Illustrates a rough idea of the signal flow involved in the modeling of an AC induction motor. The inverse Park Transformation is highlighted, showing its role as the first calculation done in the signal flow, allowing all subsequent blocks to be designed with enormously greater simplicity than without it. The model is further refined in the Mathematical Modeling of the A/C Induction Motor section.

Once there is a complete motor model to simulate with, the control system itself will use the Park transformation in real time to calculate desired reference quantities for the PI controllers to follow. This will sacrifice complexity of the PWM algorithm in favor of a massively simplified control algorithm with greatly reduced stress on the PI controllers we will use in the implementation.

### Space – Vector Pulse Width Modulation Integrated Circuit

If we need an IC to do this discretely lets decide that here

## Space Vector Modulation

The purpose of space vector modulation is to represent a sampled voltage vector as a combination of inverter state vectors (and ultimately in terms of the ABC-system). The amplitudes of these inverter state vector components are then used to determine the duration for which each inverter is active per sample period *Ts*. The duration of each inverter is different and duration within in an inverter will change with time so that each inverter will generate a pulse width modulated (PWM) signal and PWM signal for each inverter are one hundred and twenty degrees out of phase. These PWM signals are placed across the stator at appropriate points with the purpose of rotating the rotor.

To start the sampling frequency (*Fs*) needs to be determined. To best represent an ideal sinusoidal signal, a high sampling frequency is demanded. For simplicity the sampling frequency will be two degrees higher than the fastest speed of the rotor to ensure that regardless of the rotor’s speed the sampling and error correction process will be at least sufficient.

### Determining the Inverter State Vectors and Vector Notation

Now that the sampling frequency is defined, the inverter state vectors need to be realized. Looking at the set of inverters in Figure 3 we can see that for each phase when an inverter in the top row is active, the inverter directly below it will be inactive. Using that fact we can define the inverter state vectors using only the top row of inverters.

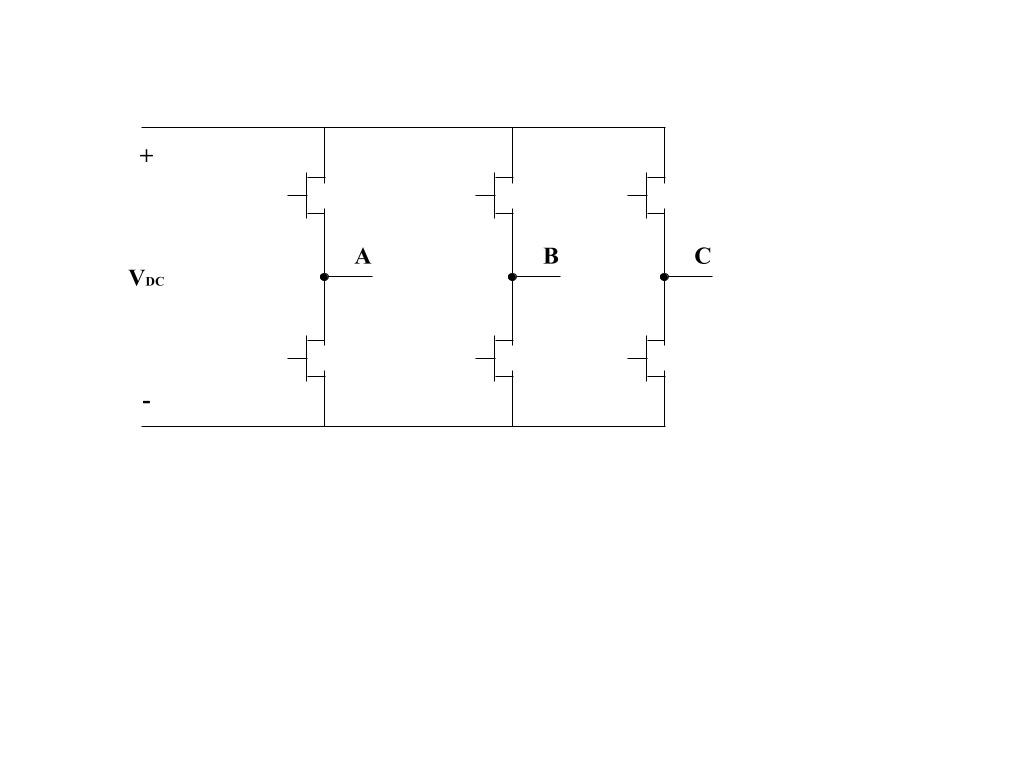


Figure 3 - Two State Inverter Set.

An inverter is either going to be on or off, meaning that when looking at all three inverters on the top row, there is a total of eight different states that the row of inverters can be in. The number 1 will represent when an inverter is on and the number 0 will represent when an inverter is off. Table 9shows the eight different states of the inverter system as well as the notion used for the rest of this document. Looking in Figure 3 it can be seen that the phases are displayed from left to right, A, B, and C but in the table the phases displayed in the opposite manner meaning that the on and off states of each inverter per phase will be represented in the opposite order: for the inverter state, 101, the top inverters for the A and C phases are active while the top inverter for the B phase is inactive (meaning the bottom inverter for phase B is active).

Now that these states are defined as a combination of the A, B, and C phases they can also be displayed as vectors in the ABC-system. Figure 4overlays the inverter state vectors onto the ABC coordinate system as well as the αβ-system since the sampled voltage will be expressed in the αβ-system. The space between each state vector is a sector. Starting from state vector V1 and proceeding counter clockwise the sectors are S1, S2, S3, S4, S5, andS6. It can be seen that any sampled voltage (Vs), whether in the ABC-system or αβ-system can be expressed as a combination of the inverter states (which exist in the ABC-system) and vice versa. The magnitude of the projection of the sampled voltage onto the inverter state vectors is used to determine the duration of each state when generating the PWM signal. This will be discussed later.

|  |  |  |  |
| --- | --- | --- | --- |
| **Inverter State** | **Phase C Activity** | **Phase B Activity** | **Phase A Activity** |
| 000 | Inactive | Inactive | Inactive |
| 001 | Inactive | Inactive | Active |
| 010 | Inactive | Active | Inactive |
| 011 | Inactive | Active | Active |
| 100 | Active | Inactive | Inactive |
| 101 | Active | Inactive | Active |
| 110 | Active | Active | Inactive |
| 111 | Active | Active | Active |

Table 9 - Inverter Activity States.

The dashed lines connecting each state vector is used to determine the maximum value of a sampled voltages. When an ideal sinusoidal waveform is plotted in terms of radians, one period of the waveform creates a circle in the plot (a variation of the unit circle). This concept is also applied to Figure 4. If the purpose of space vector modulation is to generate a signal that best mimics an ideal sinusoidal waveform, then the behavior of that waveform will be used as the limit of the amplitude of the sampled signal. From this concept we can see that the maximum value the sampled voltage can have (Vs,max) is shown in Equation 5.

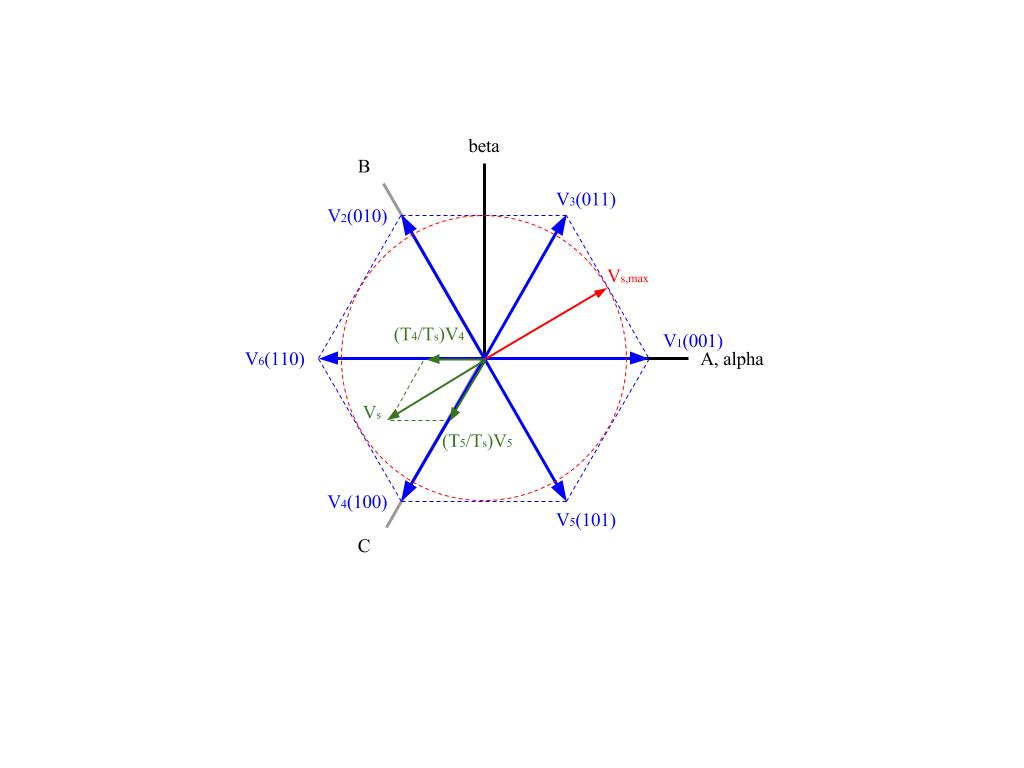


Figure 4 - Inverter State Vectors.

Equation 5 - Maximum Sampled Voltage.

In Equation 5the magnitude of an inverter state vector is replaced with half the DC bus voltage because the rectification of positive and negative polarity of the three phase sine waves will be separated and placed across the positive and negative rail of Figure 3respectively.

### Developing the Pulse Width Modulated Signal:

The ideal sinusoidal waveform will be approximated using pulse width modulation. The pulses will be generated using the inverters and the width of each pulse is determined by the active duration of each inverter. This duration, as mentioned before, is realized using the magnitudes of the projections of the sampled voltage onto its neighboring state vectors. First it must be apparent that the sampled voltage remains unchanged until another sample is taken meaning that all calculations and analyses are discrete. If the sampled voltage in Figure 4were to be expressed in terms of the adjacent state vectors, Equation 6 would show the relationship.

Equation 6 - Projection of the Sampled Voltage onto Neighboring Inverter State Voltages.

This relationship is useful but not complete since the sampled voltage is not expressed in terms of the zero state vector. This concept combined with the sampled voltage’s projections leads to Equation 7 which equates to sampled voltage vector to the time varying average of the inverter state vectors where the zero vector can be either *V0* or *V7* or both. For either case the zero state vector is expressed by *V0* in Equation 7. The durations of each state vector make up the sampled vector for the entire sample period meaning that the summation of each state vector duration must equal the sample period (Equation 8). Both Equation 7and Equation 8 can be used to find the durations of each inverter state used to generate the PWM signal. For convenience the applied sampled voltage will now exists in between state vectors *V1* and *V3* (Figure 5).

Equation 7 - Time Varying Average of the Inverter State Vectors.

Equation 8 - Subintervals of the Sample Interval.

Here we will utilize the alpha and beta axis to find the time durations. The projections of the sampled voltage, and each inverter state (including the zero vectors) onto the alpha axis can be seen in Equation 9. Equation 10shows the projections of each vector onto the beta axis.

Equation 9 - Projection onto the alpha axis.

Equation 10 - Projection onto the beta axis.

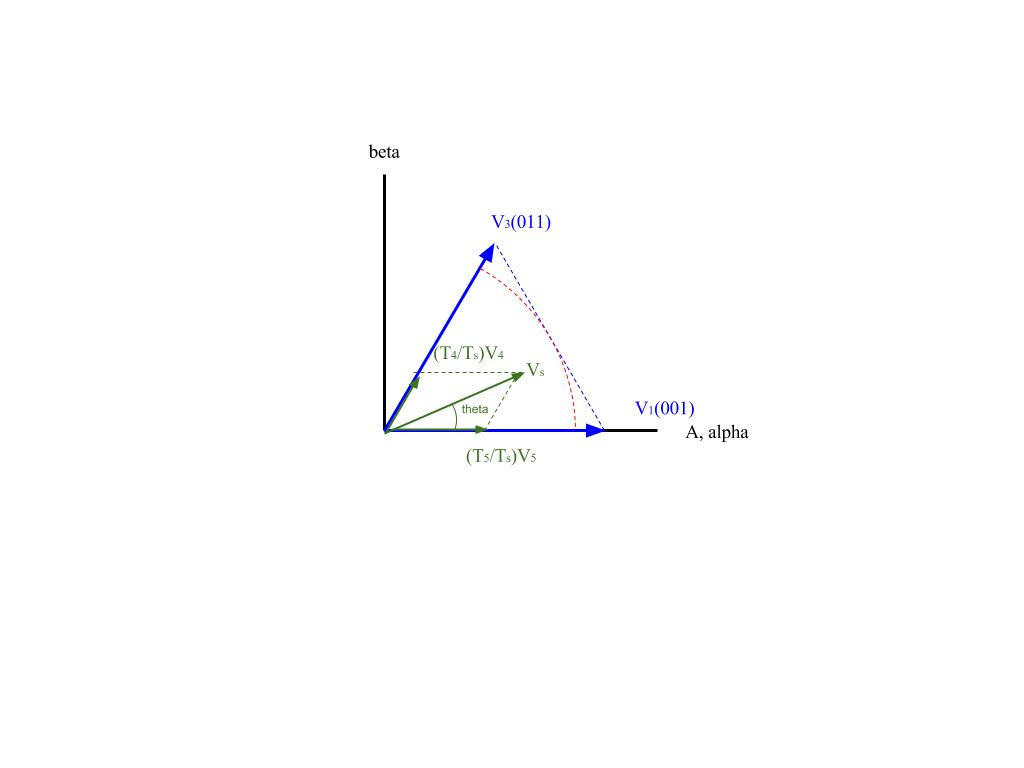


Figure 5 - Sampled Voltage Vector in Sector S1.

Using these equations we can solve for T3 and T1. It is important to note that the voltage Vs will be a measured value and the sampling period *Ts* will be known. V3 and V1 known for being the voltages across both inverters which is equal to VDC (as well as any other state vector). With these observations the equation to find the duration of state vector V1 is Equation 11 and the equation to find the duration of state vector V3 is Equation 12. The time duration for the zero vector can be found by subtracting these time durations from the sample period.

Equation 11 - Time Duration of Vector State V1.

Equation 12 - Time Duration of Vector State V3.

The purpose of the zero voltage state vectors are to keep the number of inverters switching between two time subintervals limited to one inverter. This process reduces the rippling effect produced when an inverter is suddenly switched on or off. Since the zero vectors do not cause the inverters to produce an output, these states will not only reduce the ripple but help transition from one sample to another since the voltage (and ultimately power) produced by the inverters will not suddenly change between samples provided that the zero voltage vectors are used on either end of the sampling period. This is accomplished by splitting the zero time duration into two halves and letting either the V0 state or V7 state occupy that duration (Figure 6).

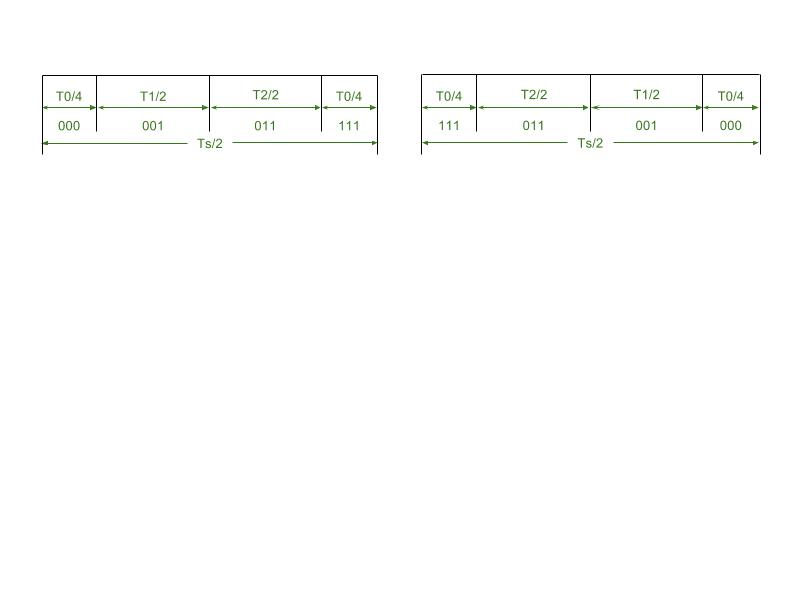


Figure 6 - Visualization of Time Subintervals

The sample period Ts1 and Ts2 have the same duration but the placement of each time subinterval is flipped around the halfway point of the sample period. This is done to maintain the one inverter switching limit. The transition between sample Ts1 and Ts2 is ideally nonexistent since the same inverter state will be active for the end of Ts1 and the beginning of Ts2 and the output of both states will be zero.

Each of the time durations in Figure 6 applied to their respective state vectors can be visualized in Figure 7where the activity of each state is expressed in terms of A, B, and C phase activity using the vector state notation in Table 9. The high notation refers the top (or positive) half of the inverters and the low notation refers to the bottom (or negative) half of the inverters. As mentioned before only one inverter per phase can be active at one time so when the high inverter is active the low inverter is inactive and vice versa. For each new sample voltage, the time durations are going to be different and the pulse width is going to vary simulating a sinusoidal signal.



Figure 7 - Visualization of the State Vector Activity in the ABC-system.

### Equating the PWM Signal

Using Figure 7an equation can be written for the average voltage per phase Equation 13. Each phase active in the high section can be expressed with a positive amplitude and each phase active in the low section can be expressed using a negative amplitude. The amplitude of each phase voltage is equal to the voltage across one inverter, which is equal to half of VDC.

Equation 13 - Equating the Pulses of Figure 7.

Do the opposite polarity of the two zero state voltages, the average voltage per phase is only dependent on the time duration of the neighboring state vectors. Conveniently these time durations were calculated in Equation 11and Equation 12. Substituting these equations into Equation 13 yields Equation 14. This equation redefines the average voltages in terms of theta.

Equation 14 - Average Phase Voltages Dependent on Theta.

This equation can be simplified more by substituting (VDC/2) for each phase voltage and by using trigonometric identities to get Equation 15. The angle theta can be replaced with “ωt” where ω is the sampling frequency of the sampled voltage provided that an initial condition is established where ωt is equal to zero (Figure 8).

Equation 15 - Average Phase Voltage in Sector 1.

### Evaluating the Average Phase Voltage for each Sector

The angle theta can be replaced with “ωt” where ω is the sampling frequency of the sampled voltage provided that an initial condition is established where ωt is equal to zero (Figure 8). The direction of the initial condition is point down so that the behavior A phase average voltage can be expressed with a sin function (allowing for the substitution of θ with ωt).

The first objective is to find the behavior of the A phase average voltage. From there the B and C phase average voltages can be expressed in terms of the A phase average voltage. The behavior of the A, B, and C phase average voltages in sector 1 will be used to simplify this process as well.

It can be seen that starting at the initial condition and continuing counter clockwise until the inverter state vector voltage V5, θ will vary from 30 degrees to 60 degrees (since each sector is 60 degrees). In this same window, ωt varies from 0 degrees to 30 degrees, meaning that in sector 5 the relationship between θ and ωt can be seen in Equation 16.

Equation 16 - Relationship between θ and ωt in Sector 5.

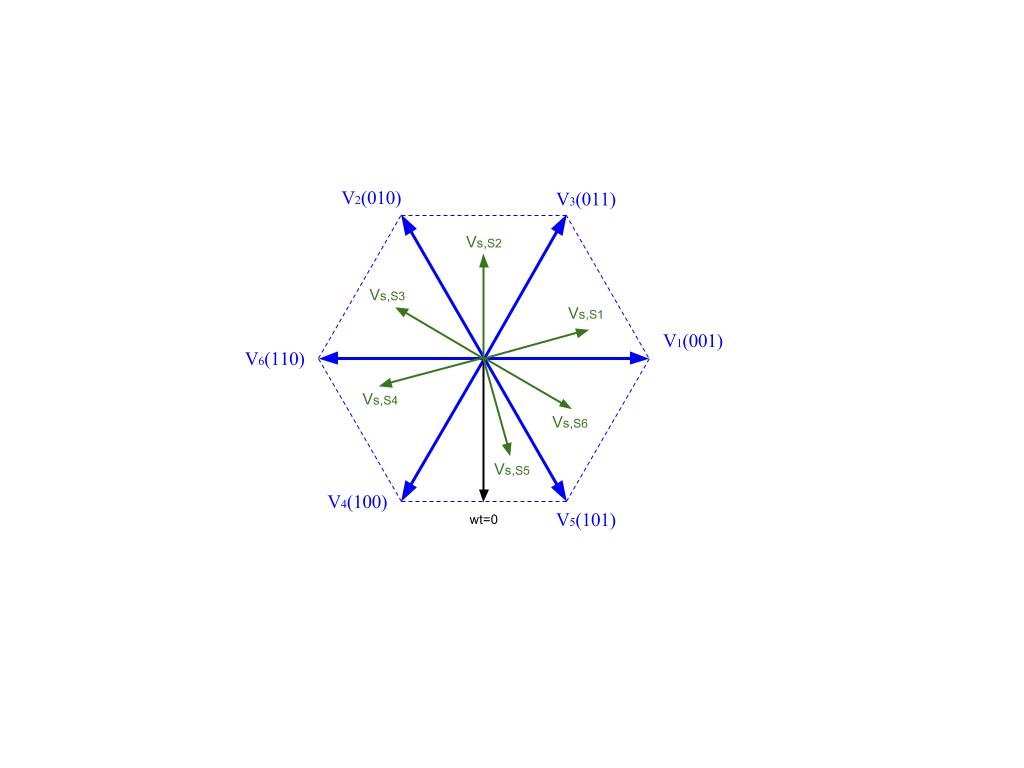


Figure 8 - Voltage Analysis per Sector.

The phase A voltage in sector 5 (Vs,S5) behaves just as the phase B voltage behaves in sector 1. The equation governing the average phase A voltage would be the average phase B voltage in Equation 15. Substitute Equation 16 into that equation and you get the behavior of the average phase A voltage in sector 5 (dependent on ωt).

Equation 17 - Average Phase A Voltage in Sector 5.

In sector 6, the phase A voltage (Vs,S6) is between two state vectors each with a notation stating that Vs,S6 would remain active throughout the sector. This means that the behavior of average phase A voltage in sector 6 can be expressed by the phase A equation in Equation 15 since the phase A voltage behaves in sector 1 the same as in sector 6. This yields:

Equation 18 - Average Phase A Voltage in Sector 6.

## Rectifier Topologies

In the first stage of our Variable Frequency Drive, power will be converted from alternating current into direct current through a rectifier. The rectifier, by definition, converts AC into DC. For the variable frequency drive to achieve the criteria outlined in the Design Specifications, we need a rectifier that has very low losses. In the interest of reliability, simplicity of diagnostics, and ease of repair, we want a rectifier that is reasonably simple to implement, hopefully without any digital control required for proper operation.

### Half – Bridge Diode Rectifier

The simplest and most easily designed rectifier topology commonly in use today is the half – bridge rectifier. The half bridge consists of only two components in its most basic form; a diode and a capacitor. The diode only allows current to pass one way through it, creating a series of “humps” which resemble the positive portions of the input sinusoid with a relatively constant DC offset of -Vγ where Vγ is the turn on voltage of the diode. This assumes a constant voltage drop diode model and in practice the voltage drop VDiode will be nonlinear with respect to forward current. A linear model can also be used to simulate the diode but still will not be able to truly model the diode’s forward current characteristics [6]. Figure 9 shows a basic topology for a half – bridge (half – wave) rectifier.

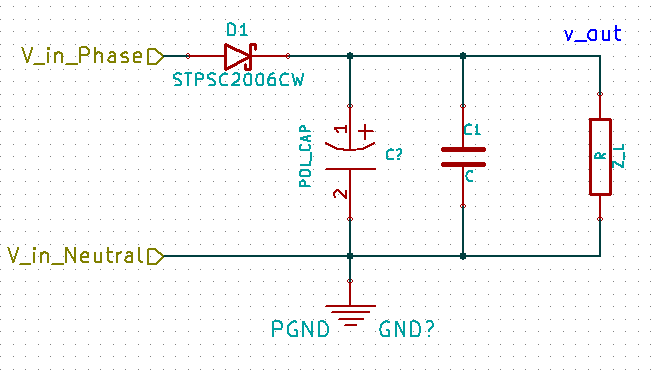


Figure 9 - Basic half - bridge rectifier topology modeled in KiCad

The diode current is inherently pulsating as it allows only the top half of the sinusoidal input to pass, and thus as it is technically DC by definition, it needs capacitive filtering at the very least to render it useful for any DC application. Therefore, the second of the two components is a capacitor. The capacitor stores energy between the “humps” of the diode current and smooths the ripple which was created by the diodes rectification.

### Full – Bridge Diode Rectifier

The full – bridge rectifier improves on the half – bridge rectifier by adding another two diodes to include the negative portion of the input AC power. This rectifier has the same characteristic “humps” on the output but has clear advantages in the ability to load both the positive and negative cycles of the AC source as well as supply more power to the load on the DC rail. A simple design for a full – bridge rectifier is shown in Figure 10.

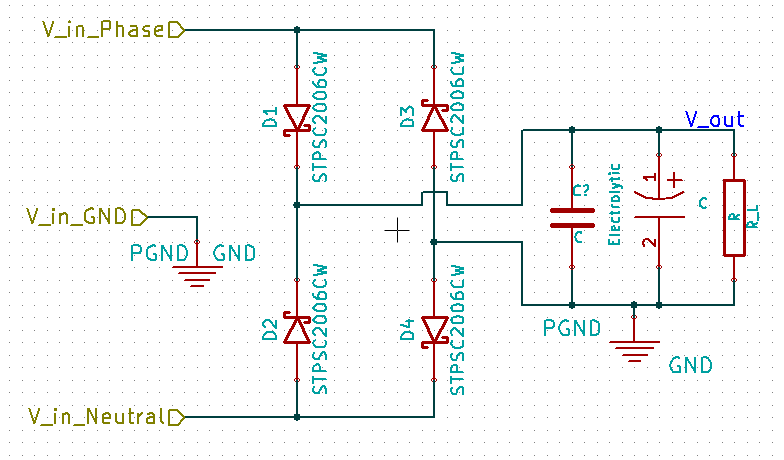


Figure 10 - Full - bridge rectifier topology modeled in KiCad

One consistent drawback to the aforementioned rectifier topology is their simple usage of diodes. Diodes, even very low forward voltage Schottky diodes have forward voltages of about 200mV at absolute minimum (and this is at very low current levels), and it rapidly increases with current. The STPSC2006CW 10A rated Schottky array has 1.7V forward voltage at 10A. Given that real power can be calculated, then, it follows that:

Equation 19 - Calculation of diode losses.

The current 10A was selected based on the ½ hp rating of our first prototype motor and the likely increase in power that will be made with a new motor, somewhere around 2.5 – 5hp. And average rectified power calculated with **Error! Reference source not found.**. This number is quite arbitrary at this point in design and is simply used as a rough estimate of a reasonable current we might expect in the final design. The choice of switching device for rectification is refined in the Synchronous Rectifier section within the Design Procedure which follows.

From Equation 19, conservatively we would be looking at 34 watts (two diodes always in conduction) lost in the rectifier diodes alone. This loss would be acceptable if efficiency was not a major priority for this project. Dissipating power in our rectifier is going to be unavoidable, but upwards of 40 watts is not ideal and better solutions should be considered.

### Silicon Controlled Rectifier (SCR)

One attempt at reducing the issue of diode loss and lack of control in the simpler diode topologies is the usage of Silicon Controlled Rectifiers (Thyristors). The benefits of using SCRs include significantly improved control of the rectifiers operation and reduced conduction losses because of lower forward voltage drop. Silicon controlled rectifiers operate similarly to diodes, but include a third pin; the gate. The gate must be forward biased for the SCR to allow current to flow, and once the gate is forward biased current will flow until the anode – cathode terminals are reverse biased again (it can’t be turned off by discharging the gate). The downside of SCR rectifiers is that they require moderately complex drive circuits and methods of isolating the drive signals for each SCR as they do not share common cathodes. Figure 11 shows a typical configuration for a Silicon Controlled Rectifier topology not including isolation transformers for driving the appropriate SCRs in a synchronized manner. The figure also excludes the drive signals themselves and focuses on the change from diodes to SCRs.

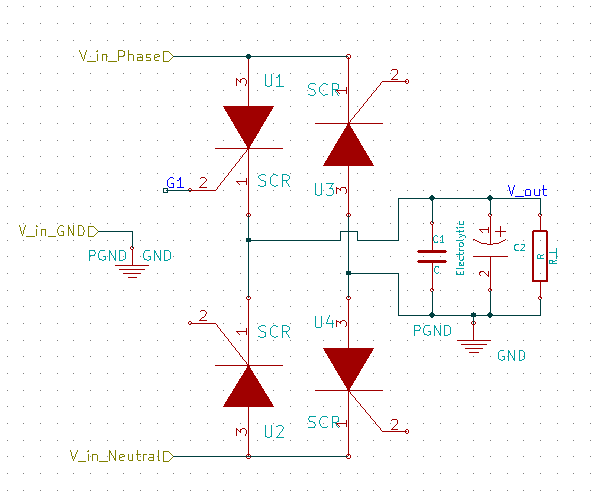


Figure 11 - The Silicon Controlled Rectifier Topology.

### Synchronous Rectifier (Active Rectifier)

MOSFETS have been growing in popularity as a switching device since their inception, and it is no surprise. Compared to other options (in this case diodes and SCRs) MOSFETS offer significantly better conduction characteristics for minimizing losses in the switching devices. For this reason, MOSFETS are often chosen for rectifier switches when efficiency and size are priorities for design. Rectifier which employ MOSFETS or other transistor elements as switching devices are termed active rectifiers. Active rectifiers require integrated circuits to drive the gates of the MOSFETS to control both the turn on and turn off of the switches. Because the MOSFET will conduct current back into the mains if allowed to stay on, the turn off is of major importance and one of the reasons why active rectifiers require the most complex driver circuit of the considered topologies. Because there are integrated circuits which are purpose built for this application, however, the active rectifier’s immensely superior efficiency is very appealing given that efficiency and size are both major design goals. Equation 20 below uses the Toshiba TK62N60X 600V Vds 61Amax rated MOSFET which has an on resistance Ron\_typical = 33mOhm at IDS = 40A.

Equation 20

Comparing this result to Equation 19 it is clear to see the MOSFET has massively superior loss characteristics when compared to a comparable Schottky diode previously considered. This is the main motivation for using a synchronous rectifier over a diode rectifier as either a half or full – bridge topology. Figure 12 shows a simple active rectifier topology using N channel MOSFETS. The drive signals at the bottom of the image will connect to a PMIC which will control the switching of each FET. This PMIC is detailed in **SECTION**.

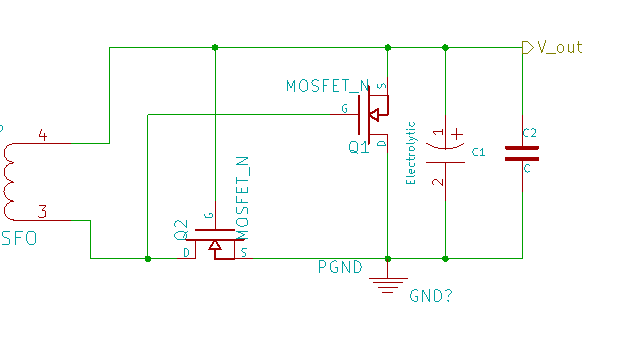


Figure 12 - The Active Rectifier using MOSFETS as the switching devices. Note the drive signals entering to control the switching of each FET. Designed in KiCad

The main drawback of the active rectifier is twofold, firstly, due to the issues of power flowing back into the lines and the requirements of MOSFET gate drive we will either a good custom algorithm for MOSFET switching or we will need to use a discrete IC to monitor the MOSFET and switch appropriately. Secondly because the complexity is higher and we will be using four MOSFETS instead of a packaged rectifier or building one out of Schottky diodes, the cost will be greater with an active rectifier than a passive one. This is acceptable, however, because the greater efficiency will save money in thermal management and help keep the overall efficiency of the drive in line with the design specification. For these reasons we will be using a synchronous rectifier topology for the high voltage power rectifier supplying the DC link. The starting topology will be an adaption of the full bridge topology switching MOSFETs for diodes.

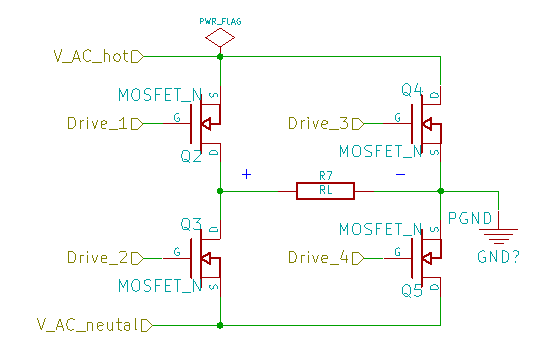


Figure 13 - The synchronous bridge rectifier; the starting point for our input power rectifier design for optimal efficiency.

## Power Inverter

The power inverter is the third core subsystem in the power supply design, following the transformer and synchronous rectifier. The inverter takes the 350V DC link and uses pulse width modulated power switching to synthesize a quasi-sinusoidal power signal to the motor which is variable both in frequency and in amplitude. To program a specific sinusoidal output, a space vector pulse width modulation algorithm will be implemented to optimize the loading of the DC link. By constantly sensing the DC link voltage and calculating the necessary pulse width at that measured voltage to generate the specified output

### Overview of Inverter Circuits

The basic topology for a single phase inverter is shown in Figure 14 below, it consists of four switching devices positioned very similarly to that of a bridge rectifier. And rightfully so, power inversion is the conversion of DC into AC, the opposite of rectification. To modulate a positive pulse onto the output, the IGBTs Q1 and Q3 both turn on, to modulate a neutral coasting period, leave both high and low side IGBTs off and allow the freewheeling diodes built in to the IGBTs to limit the motor voltage on the phases to that of the rails. The inverter allows the connection of either high or neutral lines to either node of the output voltage. Used with DC power this circuit is referred to as an H bridge and allows the direction control on a DC motor by controlling polarity on the output.

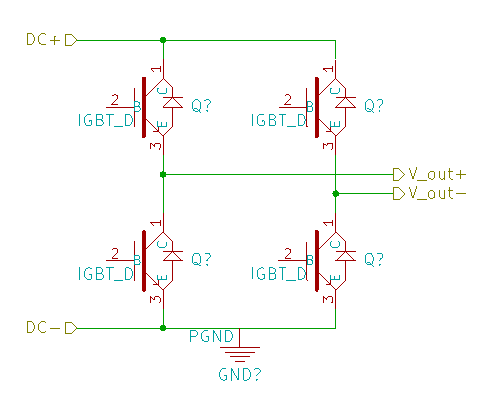


Figure 14 - The basic structure of a single phase inverter using IGBTs. Designed in KiCad

The inverter has two operating states. In state 1, Q1 and Q4 both on; creating a positive output pulse, and in state 2, Q2 and Q3 are on; providing a negative output pulse. These two states are mutually exclusive. When state 1 and state 2 are set in a constantly alternating pattern with set frequency of state changes, a square wave is generated at the output. This is the most basic form of inversion; transforming DC into AC. If the inverter operates as it did previously, but this time the high side switch is pulsed on and off during each state in a manner which modulates a low frequency signal onto the high frequency pulses, the inverter can be used to synthesize many function such as ramp, triangle, and sinusoidal outputs.

The switching algorithm which we will use to control our power inverter is described in the Space Vector Modulation section preceding. The inverter itself simply needs to take the commands issued from the microcontroller dictating the states of each IGBT, decouple them from noise, and apply the signals through an isolating driver to the high voltage IGBTS which make up the inverter. One notable consideration that the SVPWM algorithm will impart is that the high – side and low – side switching will average to the same frequency. This is in contrast to a direct sinusoidal PWM where the low – side switches only switch at the output frequency to switch the polarity of the output and the high side switches do all of the pulse width modulation to create the high or low cycle of the sinusoid desired.

### Selecting a Power Switch Type for the Inverter

At the heart of any inverter circuit is the power switch. The device which creates the pulses generated as the output of the space vector pulse width modulation used in our controller topology. These switches, for our design, will require greater than 400V rating and be easily capable of switching at approximately 20 kHz; a common frequency used in this type of power PWM for motor control. The main criteria that we will be considering are the switch’s size, on resistance, and simplicity of drive circuit topology.

Upon further research it becomes clear that there are ICs made to drive any type of switch we may choose at a very reasonable cost, with small footprint in a surface mount package. This observation renders the drive circuitry considerations less important, being more concerned with simply specifying an IC, and designing its supporting passive networks reduces design rigor immensely.

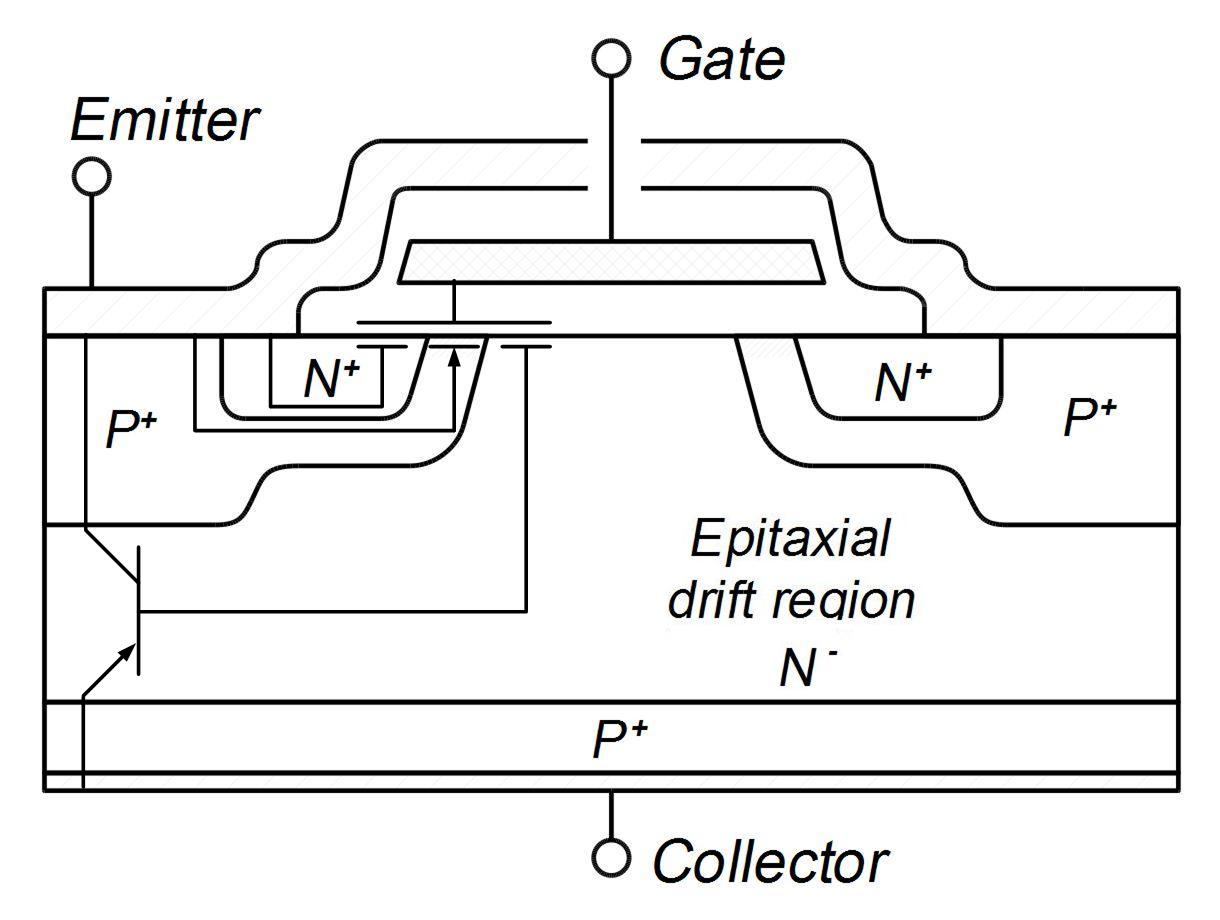


Figure 15 - Cross section of an IGBT, copied with permission requested from ElectronicDesign.com [7]. Permission request can be found in Appendix A – Copyright Permissions.

There are two current technologies we consider a strong candidate for our power switch. The power MOSFET and the Insulated Gate Bipolar Transistor (IGBT). It is clear from general research into current existing technologies that most motor drives use IGBTs. This raises a question of why they chose this device over a MOSFET and whether IGBTs will be the best choice for us.

|  |  |  |
| --- | --- | --- |
| MOSFET vs IGBT | | |
|  | MOSFET | IGBT |
| Charge Carriers Involved | Majority | Majority and Minority |
| Typical on Resistance | Low | Lower |
| Turn on Time (Ton­) | Similar | Similar |
| Turn off Time (Toff) | Shorter | Longer |
| Current Density (I / A) | Lower | Higher |
| Required Die Size | Higher | Lower |
| Noise Generation | Higher | Lower |

Table 10 - Qualitative comparison of MOSFET and IGBT [8]

As can be gathered from Table 10, the IGBT typically shows lower on resistance, and thus lower I2R power losses due to switch conduction. The MOSFET typically has significantly shorter turn off times on the switch, which makes the MOSFET conducive to high – frequency applications. Being that this drive will be ideally around 20 kHz, this feature of the MOSFET will likely be unnecessary. The fact that IGBTs employ both minority and majority carriers in their conduction bands allows them to have much higher current densities per volume of silicon, thus allowing for smaller package size, and therefore reduced cost. [7]

Because size, cost, and efficiency are all primary objectives of our motor drive, the Insulated Gate Bipolar Transistor seems to be the best option for us. It is no surprise as this is the current industry standard. The combination of small die size, low losses, and available literature make it the clear choice for the VFD Go-kart [8]. With ever advancing semiconductor technology and development of lower on resistance MOSFETS, it seems evident that the era of the IGBT may soon be ending and future iterations of this engineering solution will be more likely to choose MOSFETS over IGBTs.

## Specific Design Considerations

### Selection of Protection Device Values

The protection devices used on our system will prevent total system failure in the event of short circuit and more importantly protect against fire during a short circuit. Values were chosen based on estimates for the maximum current expected at given locations. The assumption of a maximum of 4000W in for the main power input and 38W in for low voltage rails were assumed. Nominal voltages were used as defined elsewhere in the document.

|  |  |  |  |
| --- | --- | --- | --- |
| **Fuse and Circuit Breaker Value Specifications** | | | |
| **Protection Location** | **Fuse or Breaker** | **Value** | **Justification** |
| Mains Input () | Breaker | 20A@250VAC | Safe loading level for a standard US 240V outlet |
| Power Transformer Secondary | Fuse | 20A@125VAC | Predicted maximum is 17A |
| LVDC Transformer Secondary | Fuse | 3A@30VAC | Predicted max current here is 2.5A |
| Motor Phases | Fuse | 20A@250VAC | Protect against total failure from short circuit |
| DC Link (both high and low) | Fuse | 15A@400VDC | Predicted max current is 11.5A |
| LV DC rail | Fuse | 5A@15VDC | Predicted max is around 3A. Could be higher or lower though. This fuse will be adjusted as necessary |

Table 11 - Specifications of circuit breaker and fuse values.

### Implementing Active Ground Management

To implement an active ground management system we simply use a current sensing resistor placed on the path to earth ground at the mains plug. A comparator with the trigger level being 0V will be used to sense any current flowing through this current sensing resistor. If there is current present, this indicates that somewhere in the system there is a short to ground (ground fault) and a logic high is sent to the microcontroller which raises the appropriate flag and interrupt to stop the power system.

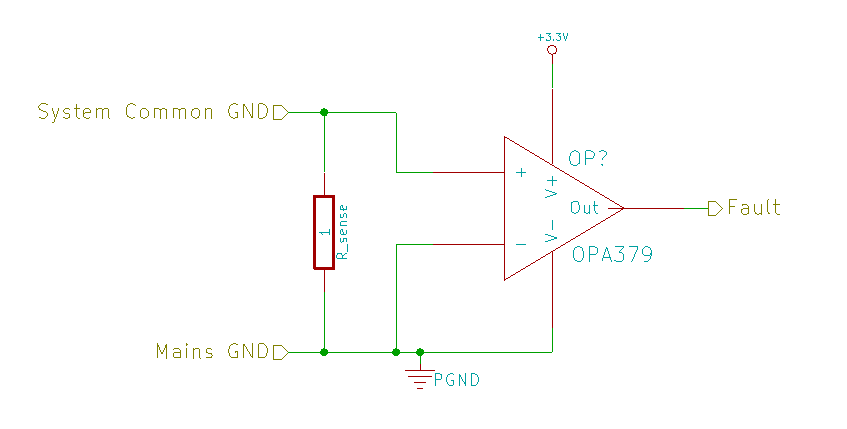


Figure 16 - The basic network used for ground fault detection. Designed in KiCad

### Comparisons of Temperature Sensing Methods

Discuss the different technologies available to sense temperature. RTDs, Thermistors (PTC and NTC), thermocouples. Choose the ones we will use to track temps of the ICs during test, and the ones inside the motor.

Temperature is defined as the energy level of matter which can be evidenced by some change in that matter. Temperature sensors come in a number of configurations and have one thing in common: they all measure temperature by sensing some change in a physical characteristic.

Resistive temperature devices, or RTDs, are electrical devices that measure temperature in a unique way among electrical devices. Rather than indicating temperature as a change in voltage, they take advantage of resistance. In general, RTDs are more linear than thermocouples. They increase in a positive direction, with resistance rising as temperature rises. On the other hand, the thermistor has an entirely different type of construction. RTDs tend to be more expensive because of their high accuracy, and therefore are not implemented in this design.

### Usage of a Digital Control Algorithm

Digital Signal Processing or DSP is of integral importance to our control design. The advantages that a digitally implemented control system offer are immense from both a simplicity of design perspective and an economic perspective of cost and space. Digital signal processing, data management, and control calculation all share the same benefits over their analog alternatives; they are cheaper, smaller, more reliable, and immensely more flexible to the specific application. Table 12 summarizes some of the design considerations made by the group when the decision was made between an entirely digital control system and an equivalent analog system.

|  |  |  |
| --- | --- | --- |
| **Digital Control vs. Analog Control** | | |
| Motor Control Design Consideration | Comments on Each Technologies’ Advantages/Disadvantages | |
| Digital Control | Analog Control |
| Cost of Final Controller Design | Low Cost, minimal passive components, chip under $20 | Can be hundreds of dollars in component costs only. |
| Size of Final Controller Design | Will fit on one or two four inch square PCBs due to small component count | Requires large, complex PCBs to accommodate the high component count. |
| Flexibility in Implementation of Final Controller | Controller can be easily customized for any motor within a range of power and voltage levels. | Due to physical calculation circuits, very limited deviation is allowed from the originally specified motor in design. |
| Complexity of Algorithm | Digital algorithm design can be difficult due to A/D conversion, differences between discrete and continuous time, etc. But allows for implementation of complex mathematical functions in a simple, small package. | Conceptual design can be simplified. However, difficulty of practical design is a major concern and therefore heavily limits the complexity of the control algorithm chosen. Also makes the physical layout extremely complicated due to powering high number of active devices. |
| Difficulty of Revision | Revision of design and control algorithm is straightforward. Simply re-flash the microcontroller memory and alter simple external networks if necessary. | Revision, even if small, will require rebuilding large portions of a highly complex active network. Rendering revision costly in both time and materials. |
| Efficiency | Due to the allowed complexity of control algorithm, digital control offers the opportunity for much greater efficiency overall. | Efficiency is nominal, with some losses in the control circuit itself. But much greater losses resulting from lack of criterion consideration required by the simple control algorithm. |

Table 12 - Comparison of Digital and Analog Control Algorithms

It is apparent through any considerations contained in Table 12 that digital control is vastly superior to analog alternatives for most applications of motor control. There may still exist specific solutions in control engineering for which analog control is best optimized, but for our VFD Go-kart, it is clear that digital is the only option worth attempting.

### The Current Sensing Device

The Hall Effect sensor is good for sensing current, can also use a resistive sensor with an extremely small, high current resistor placed in series with the motor phase inputs. These sensors would also require ability to isolate very high voltages from their measurement signals.

#### The Basics of Current Sensors:

Current sensors are either open-loop or closed-loop.  While open-loop current sensors measure AC and DC currents and provide electrical isolation between the circuit being measured and the output of the sensor that is, galvanic isolation takes place.  The primary current measured does not make electrical contact with the primary circuit.  Open-loop current sensors are generally more suited for battery-powered circuits and hence will not be utilized in this design.  Although less expensive, open-loop current sensors don’t provide the accuracy in measurement that the design requires.  [9]

#### The Advantage of Closed-Loop Current sensing:

Closed-loop sensors measure AC and DC currents, while also providing electrical isolation.  They offer fast response, high linearity, and low temperature drift.  The output current of the closed-loop sensor is relatively immune to electrical noise.  Sometimes called a “zero-flux” sensor, the closed-loop Hall Effect sensor feeds back an opposing current into a secondary coil, wound on the magnetic core so as to cancel or “zero” the flux produced in the magnetic core by the primary current.  They are the sensor of choice for this application because high accuracy is essential in this field oriented control scheme.  In general, they are comprised of a Hall generator mounted in the air gap of a magnetic core, a coil around the core, and a current amplifier.  The current carrying conductor placed through the aperture of the sensor produces a magnetic field that is proportionate to the current.  This field is concentrated by the core and sensed by the Hall generator.  The Hall generator is connected to the input of the current amplifier, which in turn drives the coil.  The current through the coil produces an opposing field, provided by the current through the aperture.  [9]

The output of this sensor is proportional to both the aperture current and the number of turns of the coil.  If the net primary current through the current transducer aperture is I, the number of turns in the compensation winding is N, and the current in the compensation winding is i, then at the zero flux condition I=Ni.  The output current of a closed-loop hall-effect sensor is converted to a voltage figure by connecting a resistor to the output of the sensor and ground.  Resistor value selection can result in scaling of the output.  [9]

### Choice to Favor Well Reputed IC Manufacturers

One consideration that can be a major issue for design is whether or not to choose to use the large, reputed IC manufacturers with higher priced parts but could also exhibit better reliability. The choice becomes one of the value of cost to your design. One of the more significant aspects of a manufacturer’s service to the designer, especially if the designer is an amateur, is their providing of datasheets which are not only adequate at describing the electrical characteristics of the part, but also adept in their description of the considerations that need to be made for design to ensure good functionality. Most of the top manufacturers provide exceptional datasheets for their products; they describe how the device works, give detailed descriptions of pin functionality and formulas to easily calculate component values. This information, as well as often superior graphical plots describing the components behavior, make the extra cost of the top branded ICs to be worth it at this point in the education of the group members. This is intended as a highly educational experience, where one can retain knowledge to reference on future designs. For these reasons the group feels it to be in our best interests to attempt to stick to manufacturers which provide a great datasheet for the product we buy whenever possible.

## System Block Diagrams

### Overall System

C:\Users\me361547\Downloads\Overall System.png

Figure 17 - Block diagram of the overall top, system - level perspective.

### Power System

C:\Users\me361547\Downloads\Power System.png

Figure 18 - Block diagram of the power system.

### Sensor Management and LCD

C:\Users\me361547\Downloads\Microcontroller Sensor Connection Scheme (1).png

Figure 19 - Block diagram of microcontroller sensor input.

# Related Standards

## Applicable to Electrical Components

There are various applicable standards for the components, PCB, and connections in our project. All widely used device packages have a standard which defines their dimensions and pin placement as well as the thermal characteristics of the package where applicable. An overview of the standards which apply to the components used in our design is provided in Table 13 on the next page.

Of the standards outlined in the table, most are standards for device package dimensions, these standards have little impact on the considerations of design for us as the package dimensions are specified on the datasheet so we don’t need to use a standard for our own design, the standards are important, however, because they are used by the manufacturers to define their packages. During PCB layout these standards will be useful because if a device is of standard dimensions, we can use a standard footprint for that package, saving time.

IEC 60269 and UL 248 are of special importance as they describe the requirements of fuses used for various purposes in low voltage systems like our VFD (<1kV). The standard focuses on the defining characteristic of time rupture characteristics. UL 248 includes specifications for fuses used on motor systems where short term over – current is expected, it will be important to make sure our fuses on the DC link are UL 248 rated to be sure that they will handle the inrush current properly.

|  |  |  |  |
| --- | --- | --- | --- |
| **Standards of Electronic Components Used** | | | |
| **Standard Name** | **Associated Organization** | **Applicable Components** | **Description** |
| IEC 60269 | International Electrical Council | Fuses | Provides technical requirements for low voltage fuses in many applications |
| UL 248 | Underwriters Laboratory | Fuses | Technical specifications for low voltage fuses including those for motor application. |
| Universal Serial Bus (USB) | USB Implementers Forum | Cables, connectors, and data protocol | Specifications for cable and connector construction as well as data protocol. |
| Small Outline Integrated Circuit (SOIC) | EIAJ and JEDEC | Surface mount IC packages | Specifies package dimensions for surface mount integrated circuits (note that the EIAJ and JEDC standards are different) |
| UL 94 | Underwriters Laboratory | FR – 4 | Standard for Safety of Flammability of Plastic Materials for Parts in Devices and Appliances testing |
| Dual In-line Package (DIP) | JEDEC Solid State Technology Association | Through – hole IC packages | Standard for package, lead spacing, and pitch of through – hole mounting packages. |
| TO - 220 | JEDEC | Through – hole power packages | Standard for package and lead spacing for heatsink mountable power packages. |
| TO – 247 | JEDEC | Through – hole power packages | Standard for package and lead spacing for heatsink mountable power packages. |
| Quad Flat-pack No-Lead (QFN) | JEDEC | Surface mount IC packages | Specifies package dimensions for surface mount integrated circuits |
| Micro Small Outline Package (MSOP) | JEDEC | Surface mount IC packages | Specifies package dimensions for surface mount integrated circuits |

Table 13 - Applicable standards for electrical components used on the VFD.

## Other Applicable Standards

Beyond the standards which define the components we will use, other standards have major impacts on the design of the variable frequency drive. These standards are summarized in Table 14 below.

|  |  |  |  |
| --- | --- | --- | --- |
| **Miscellaneous Applicable Standards** | | | |
| **Standard Name** | **Associated Organization** | **Impact on Design** | **Description** |
| ASME Y14.5 | ASME | All device dimension specifications | Provides standards of measurement and tolerances for dimensional specification documents. |
| IEEE 139-1988 | IEEE | Noise testing | Provides standards for the measurement of radio frequency emission from equipment. |
| IEEE C2-1997 | IEEE | Electrical safety | National electric safety code (NESC). Includes standards for grounding. |
| RoHS\* | European Union | Hazardous Material Content | Requirements set in 2006 for the reduction of hazardous materials in electronics |
| IEEE 1118.1-1990 | IEEE | IEEE Standard for Microcontroller System Serial Control Bus | We will be using a serial control bus in order for our microcontroller to connect and communicate with the LCD display module. |

Table 14 - Applicable miscellaneous standards.

\*Note: RoHS is not specifically a standard but a set of requirements laid out be a joint commission in the European Union which defines requirements for all production electronics sold after 2006.

The standard of most importance for us in the table above is the IEEE C2-1997 – the National Electric Safety Code (NESC). This standard defines procedures and design specifications to ensure personnel safety while using equipment which contains hazardous voltage levels. This standard impacts how the final project will be fused and grounded, assuring that all ground paths of power components are of a specific impedance and placement is a key to this standard.

The RoHS standard is, as mentioned in the table note, not technically a standard, however it functions like one. RoHS compliance is discussed in the Realistic Design Constraints section following. RoHS is included in this section because it has such a large impact on the electronics industry as a whole and therefore has a large impact on the parts which will comprise our system.

# Realistic Design Constraints

## Economic and Time constraints

The biggest constraint on the variable frequency drive project is time. Going from novel, divergent research on a topic not covered in course material is very arduous. Completing a design in the specified timeframe of one semester is extremely difficult given that learning curve. Time constraints apply mostly to the first semester of the project as Group F has opted to finish the design in the fall 2016 semester, giving two extra months of fabrication and testing time. It is understood by the group that much of the final design will come in the summer semester with prototyping and simulations that were not completed in the spring due to time constraints.

## Estimated Budget Breakdown

|  |  |
| --- | --- |
| **Estimated Budget** | |
| *Line Item Category* | *Cost* |
| 3 Phase, 240V AC induction motor | $500.00 |
| Mechanical assembly (case motor mount, etc.) | $400.00 |
| DC Link capacitors | $100.00 |
| Input transformers and other coils | $150.00 |
| Wire | $150.00 |
| Microelectronics (ICs, MCUs, LCD) | $300.00 |
| PCB Manufacturing | $100.00 |
| Testing Equipment (variac) | $50.00 |
| **TOTAL** | **$1750.00** |

Table 15 - Estimate of final project budget. Most items are predicted to be less expensive.

## Environmental, Social, and Political Constraints

There are various constraints which affect the variable frequency drive which come from political sources such as governing body regulations on hazardous materials (RoHS) as well as design trends which further the efforts in reducing lead specifically (Lead – free). These restrictions will be explored in the following sections.

### RoHS Compliance

The defining requirements of compliance to the Reduction of Hazardous Substances (RoHS) requirement imposed to all production electronics sold in the EU are shown in the following table:

|  |  |
| --- | --- |
| **Hazardous Material Levels in RoHS Compliant Devices** | |
| Lead (Pb) | < 1000 ppm |
| Mercury (Hg) | < 100 ppm |
| Cadmium (Cd) | < 100 ppm |
| Hexavalent Chromium (Cr VI) | < 1000 ppm |
| Polybrominated Diphenyl Ethers (PBDE) | < 1000 ppm |
| Bis(2-Ethylhexyl) phthalate (DEHP) | < 1000 ppm |
| Benzyl butyl phthalate (BBP) | < 1000 ppm |
| Dibutyl phthalate (DBP) | < 1000 ppm |
| Diisobutyl phthalate (DIBP) | < 1000 ppm |

Table 16 - Maximum allowable levels for hazardous materials in RoHS compliant devices. [10]

These requirements for RoHS compliance are rather lax and if a Lead – free design has already been chosen, then RoHS compliance is nearly guaranteed. To assure compliance we will be sure that all components selected are RoHS compliant themselves.

### Attempting a Fully Lead – Free Design

The removal of heavy metals from electronics components is an ongoing process today, and one which deserves much focus. Lead being the most prevalent heavy metal used in electronics components is of most concern to this project. In an effort to support the transition from lead based solders and coatings used in electronics Group F’s variable frequency drive will be built using entirely lead – free components and solder. This is a significant constraint for manufacturing because leaded solder is much easier to use and performs in many ways superior to lead – free solder. The vast majority of components offered today are offered as solely lead – free options or feature a lead – free variant, so sourcing components which conform to this constraint is not an issue.

In order to still achieve good quality in the manufacturing of our board we will rely heavily on a reflow solder process with surface mounting components. Solder paste – reflow soldering processes are the current standard for surface mount soldering, and work extremely well with lead – free solder compositions. Therefore it is imperative that most of our devices be surface mountable, to avoid the hassles involved with hand soldering lead – free. To help with any hand soldered components a separate tip set will be used on our iron, which will only be used for lead – free soldering.

To make the constraint less frustrating for the process of circuit prototyping with through – hole components when possible, leaded solder is permitted for initial circuit prototyping. The constraint of 100% lead free will only be imposed on the final version.

## Ethical, Health, and Safety constraints

### Compliance with Standards Regarding Grounding and Isolation

Due to the nature of private companies (profit), standards which are published by private groups are very rarely available free and to the public. For this reason, Group F does not currently possess a copy of IEEE C2-1997, the national electrical safety code. For assurance that we comply with this standard, inquiry will be made to IEEE about student access to standard C2-1997. This standard will specify the grounding requirements we must meet to declare compliance, and confidently state that our design is “safe” by IEEE standards.

### Consideration of Dynamic Breaking

Brining the motor to a stop quickly when it is turned off can be both a major convenience for the operator of the VFD, but also provides a method to stop the motor without mechanical means or requirement of a high load. The motor should be able to stop in an emergency situation to help prevent property damage or serious personal injury due to the spinning motor and load’s inertia. Dynamic breaking is a system where instead of allowing the motor to free wheel against a high impedance when it is not being driven by the inverter, the motor phases are loaded with a resistance and energy flows from the motor, slowing the motor and load quickly and reliably with no mechanical contact. The downside of dynamic breaking is that all of the energy is lost to heat if the motor is loaded with a high power resistive network.

### Regenerative Breaking

It is very possible to recover the power via regenerative breaking, like that which is used on the Tesla® Model S to boost its range significantly. For highly dynamic operation, regenerative breaking in an essential feature if maximizing efficiency and size is the goal. Therefore it would be a good thing to put up for consideration for our VFD design. Regenerative breaking probably would require some sort of energy storage like bulk supercapacitors, and a regulator to dump the power back into the DC bus.

# Hardware and Software Design Process

## The A/C Induction Motor

### Motor Selection

|  |  |  |  |
| --- | --- | --- | --- |
| Induction Motors for Consideration | | | |
| **Motor Designation** | Reliance Electric P56H5069G | Iron Horse  MTCP-005-3BD36 | GE  5K33GN2A |
| **Rated Voltage (V)** | 208-230  460-480 | 208/230  460 | 208-230/460 |
| **Rated full-load amperage (A)** | 1.1-1.2  2.2-3.2 | 12.6  6.3 | 1.4 |
| **Frequency (Hz)** | 60 | 60 | 60 |
| **Number of Phases** | 3 | 3 | 3 |
| **Rated full-load speed (rpm)** | 1725 | 3570 | 1800 |
| **Insulation class** | F | - | - |
| **Rated horsepower** | 0.5 | 5 | 1/4 |
| **Time Rating** | Continuous | Continuous | Continuous |
| **Locked-rotor code letter/Locked-Rotor Torque (lb\*ft)** | L | 16.19 |  |
| **Manufacturer’s name and address** | Reliance Electric Industrial Company | IronHorse Premium | GE |

Table 17 - A comparison of motors chosen for possible utilization in this design.

In this design the GE and Iron Horse Motor will be utilized for different life cycles of the prototype.  The GE motor is a motor recommended by Texas Instruments for use with the TMS320F2802 piccolo microcontroller.  More on this will be discussed in the hardware testing section of this documentation.  The IronHorse motor, on the other hand is intended for use in the final implementation of the Field Oriented Control system.

## Mathematical Modeling of the A/C Induction Motor

In order to gain greater understanding of the function of an AC induction motor, as well as make an attempt to predict the dynamic behavior of our AC motors, computer models will be created in an attempt to simulate the dynamic behavior of the AC induction motors we buy. If the dynamic models prove accurate the can be used as a means to generate the plant transfer functions to enable the design of the **dq** vector PI controller design process. Prominent software environments for mathematical simulation are MATLAB/Simulink, Mathcad, Python, C, etc. We will attempt to model the AC motor with various different modeling algorithms in Simulink and MATLAB due to the ease of use of the software and the prominence it has in industry today.

### The Clark Transformation Applied

When analyzing a three-phase AC induction motor, the electrical variables taken into consideration are the voltages applied across the windings for each phase and the currents passing through the windings of each phase. For convenience the three phases will be referred to as phases a, b, and c, making up the abc-system, where each phase is an axis and orthogonal to the other two. In this system, a-phase currents and voltages are denoted with a subscript a, b-phase currents and voltages are denoted with a subscript b, and c-phase currents and voltages are denoted with a subscript c. To fully analyze the current state of the induction motor and implement a correction within the abc-system, would take an immense amount of calculations. To reduce the time it would take to implement said change, the Clarke Transformation and the Park Transformation are utilized.

The Clarke transformation creates a two-phase system (the -system) the produces the same magnetomotive force (MMF) or magnetic potential as the three phase abc-system. The benefit of the Clarke transformation is to create a system that is less stressful to compute. It is important to note that the -system is nothing more than a fictitious system that would produce the same output as the original abc-system.

Since the MMF produced by the stator coils is dependent on the number of turns in each coil Ns (which would be the same) and the currents running through each coil, the MMF can be represented by a vector with a,b, and c component in the abc-system. Figure 3shows the graphical representation and Equation 21quantifies the projections of each current vector onto the MMF vector

Equation 21 - MMF of the abc-system

Through trigonometric identity manipulation and simplification the above equation can be simplified into a cosine component and a sine component. These two components can be used to represent the MMF in terms of an equivalent MMF produced by two windings that are orthogonal and ninety degrees to each other. These two windings will be the windings in the -system and their respective MMF equation can be seen in Equation 23*.*

Equation 22 - Simplified MMF equation

Equation 23 - MMF in the αβ0-system

The current is proportional to the summation of currents in the cosine component of Equation 22 and the current is equally proportional to the summation of currents that make of the sine component of the same equation. The relationship between these two systems can be shown in Equation 8 where the -system can be found from the measured abc-system. The matrix relating the two systems will be represented by the letter M for future convenience.

Equation 24 - Clarke Transformation of Currents Iabc

The zero-sequence component of the -system is usually the average of the components being transformed into the -system. Here we have set the zero-sequence current proportional to the average so that when the inverse of M is calculated, it will be proportional to the transpose of the M matrix. This is doable since the currents in the abc-system are balanced, meaning that the summation of , , and is equal to zero and that the zero-sequence component will not consume power.

The number of turns in the windings along the axis and axis () is different than the number of turns in the abc-system. This will be shown when comparing the power invariant or power variant cases of the entire system. For this project we will be using the power invariant case since the total power consumed will be compared to the total power provided to measure the power efficiency of the project. The power invariant case will be calculated to effects of the number of turns per winding in the -system.

For the power invariant case, the total current in the abc-system multiplied by the total voltage in the abc-system is equal to the total current in the -system multiplied by the total voltage in the -system (Equation 25).

Equation 25 - Power Invariant Equation

Using Equation 24 and Equation 25, the relationship between the total voltage in the abc-system and the total voltage in the -system is shown in Equation 26.

Or

Equation 26 - Clarke Transformation of Voltages Vabc

For simplicity we set the relationship between the current matrices and equal to the relationship between the voltage matrices and . This relationship yields the ratio between the number of turns in the windings of the abc-system and the -system.

Equation 27 - Ratio of number of Turns per winding between the αβ0- and abc- systems

Using this ratio and Equation 24and Equation 25the complete Clarke transformation is shown in Equation 28.

And

Equation 28 - Complete Clarke Transformation

If the project were to be analyzed using the time variant case, where the power consumed per component is conserved, the relationship between the currents and voltages of the abc-system and -system can be shown in Equation 29and the ratio of number of turns per winding can be seen inEquation 30. This shows that the ratio of turns is important in regards to the desired power invariance of the project.

Equation 29 - Power Variant Equations

Equation 30 - Ratio of Number of Turns per Winding for Power Variant Equations

The generalized machine description using the -system (Equation 31) was found doing a series of calculations and manipulations on to the generalized machine description using the abc-system. Even though the notation for the -system is still the same, the below equation considers both the stator values (subscript‘s’) and the rotor values (subscript ‘r’). This becomes apparent when the variables are expanded into their respective matrices.

The inductance matrix (as well as the speed matrix;) has a mutual inductance component that is dependent on the rotor angle at the instant of evaluation which demands constant evaluation of the system for every instance of time. To remove this dependency the -system is transformed into the dq0-system through Park’s transformation (with an intermediary pseudo-stationary -system which represent the rotor currents and voltages as a set of pseudo currents and voltages passing through and across a coil on the stator axes.

Equation 31 - General Machine Equation in the αβ0-system

Equation 32 - Voltage matrix in the αβ0-system Equation 33 - Current matrix in the αβ0-system

Equation 34 - The resistance matrix in the αβ0-system Equation 35 - The inductance matrix in the αβ0-system

Equation 36 - The components of the inductance matrix in the αβ0-system

Equation 37 - The speed matrix in the αβ0-system

Equation 38 - The components of the speed matrix in the αβ0-system

Pseudo-stationary -system (Park’s Transformation Applied)**:**

The same process used to determine the -system is used to determine the pseudo-stationary system. Here the MMF produces by the -system is also going to be produces by a pair of coils on both the -axis and -axis where one coil on each axis is the stator coil and the other is a stationary equivalent of the rotor coil. The relationship between the pseudo currents and the rotor current can be seen inEquation 39*.*

Equation 39 - The pseudo-stationary αβ0-system current

We also desire a power invariant transformation when entering the pseudo-stationary system. This condition will allow for the relationship between the motor voltage and the pseudo stationary motor voltage to be calculated (Equation 40)

Equation 40 - Power Invariance Condition

Equation 41 - The pseudo-stationary αβ0-system voltage

**Park Transformation:**

From here we can go into the park transformation. This transformation allows the motor to be controlled by a set of dc values rather than a set of sinusoidal functions. It is important to note that the motor is still controlled by sinusoidal values, but these values will be determined by the input dc values and the inverse transformation matrices. The MMF is also still conserved through transformation.

Through the conservation of the MMF the relationship between the -system and the dq0-system can be seen in Equation 42.

Equation 42 - dq0 Current Transformation

The benefit of this dq0 transformation is that the new machine equation is dependent on the angle between the MMF vector and the axis of the abc-system.

### Modeling in the Continuous Time Domain

A model was created using a design taken from the Antonino Trotta YouTube Channel [11]. In his video Antonio uses a simplified and efficient approach to modeling the dynamic response of an AC induction motor. Some similar videos exist publicly available and all show extremely similar results of simulation but employ unneeded complexity, such as inclusion of the zero – vector for all calculation. Inclusion of the zero – vector after the dq0 transformation is probably good practice as it can only be neglected in the balanced load case. However, because we are looking to simulate roughly the motors behavior, we can assume a balanced load condition. Presented in Figure 20 is an adapted version of Trotta’s simulation including parameters to test additional loads being applied over time, testing how the motor will respond to various static torque loads being applied over time as well as varying inertial loads is possible with this model. Detailed images of the simulation inputs, intermediate calculations, and outputs can be found in **Error! Reference source not found.**.

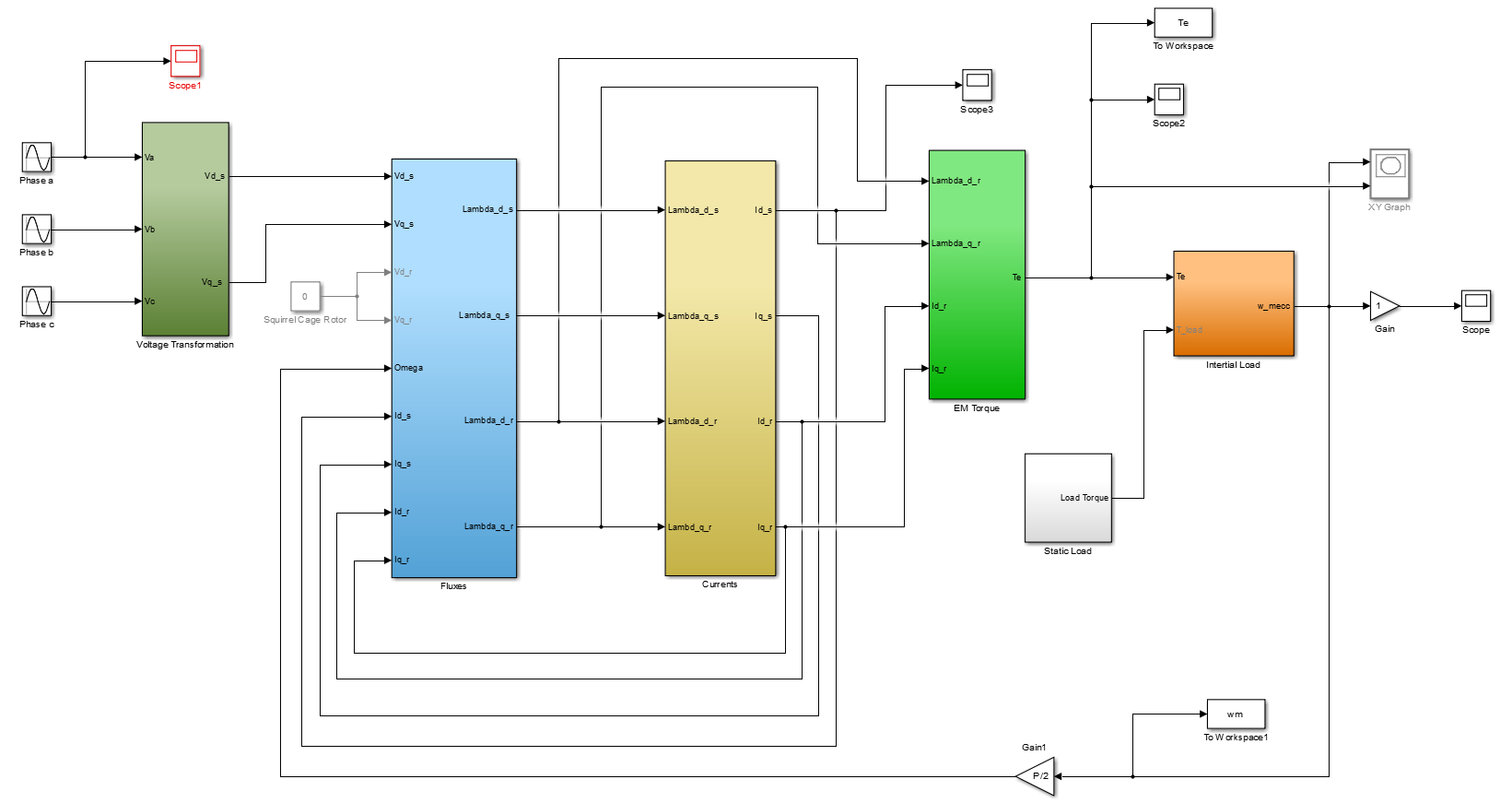


Figure 20 - Dynamic model high - level view of functionality. Taken from Antonio Trotta, 2015.

## Synchronous Bridge Rectifier and DC Link

This section details the design of the synchronous rectifier (active rectifier) we plan to use in the finished variable frequency drive. The decision to use a synchronous rectifier topology is covered in the **Error! Reference source not found.** Section under **Error! Reference source not found.**

### Synchronous Bridge Schematic Design:

The basic topology of a synchronous half – wave rectifier is shown in Figure 12 and this will be the starting point of our design. One immediate observation of the basic topology is that the gates of the MOSFETs are driven directly from the secondary side of the input transformer. This is acceptable only if the peak voltages on the secondary winding do not exceed the gate rating of the MOSFET. This is clearly not going to be a possibility in our application where we desire a 240/480VDC link. Given this immediate design constraint we must reconsider the drive of the MOSFETS to protect the gates from unacceptable over – voltage conditions.

One way of protecting the inputs is be implementing a crude voltage regulator based on a Zener diode. Figure 17 shows the basic topology of a linear voltage regulator we could use to drive the MOSFETs. The regulator regulates VO such that VO = VZ – VBE where VZ is the Zener diode breakdown voltage, VBE is the Base – Emitter voltage of the NPN on state. The resistor R6 limits the current which will flow into the base of the NPN, hence limiting the current of VO as well, this can be highly desirable for sensitive loads but for our application of MOSFET gate drive, we need high current capability as well as voltage regulation. One solution could be to simply reduce the value of R6 such that the base current was sufficient to drive the MOSFET in question, however, this will create significant power losses in the Zener as well as R6 when the base current is not needed and therefore flowing through the Zener instead. A better solution will be to implement capacitive energy storage with a series RC network on the base. This change is illustrated in Figure 21 and allows the base to draw excess current during the period at which it is needed (to provide the high current to charge the MOSFET gate) while also keeping the voltage regulated once the gate is charged without large power losses from the diode current.

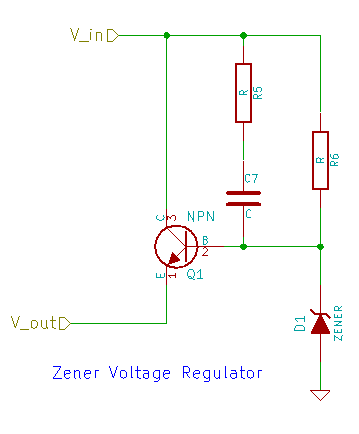


Figure 21 - The Zener voltage regulator with RC network energy storage. Designed in KiCad.

Applying the circuit in Figure 21 to the basic topology shown in Figure 12 yields a good starting point for the rectifier design. Figure 22 below shows this topology with generic components used for conceptual analysis. One thing to note in this layout which is generally based on the reference design from ElectronicDesign.com which was based on reference designs from Texas Instruments [12]. This design offers great improvement on the previous topology, however the issue of output voltage level still remains. Because we need output voltages which are vastly greater than the gate drive levels of the MOSFETs we need to further refine the design.

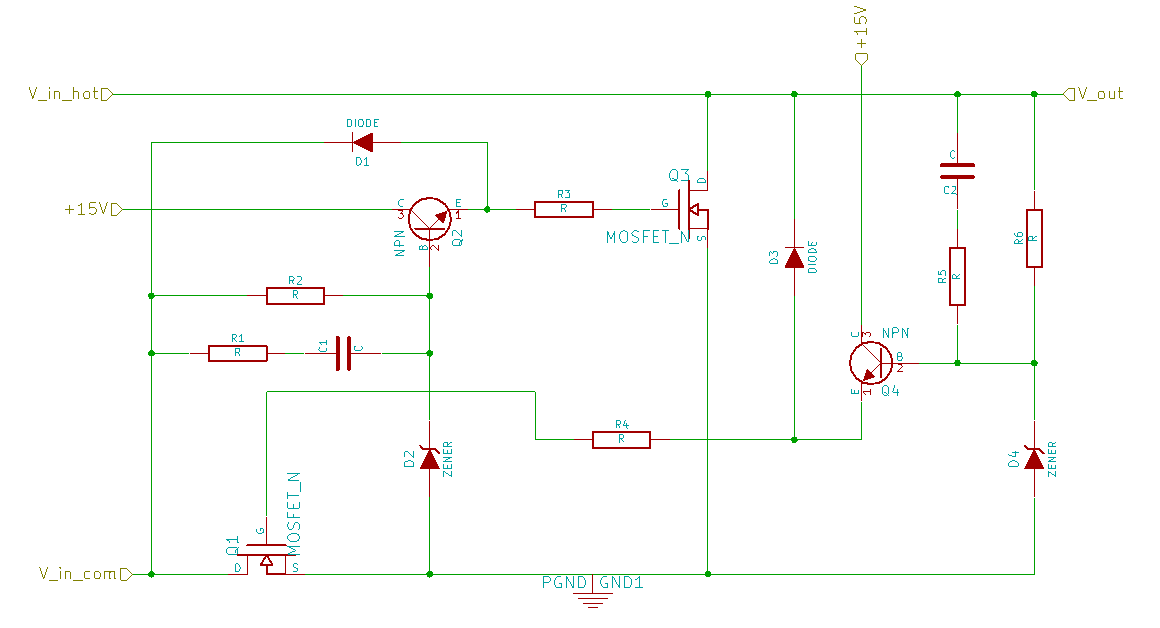


Figure 22 - The half - wave synchronous rectifier with regulated gate drivers.

One such way we can adapt the schematic to tolerate higher input and output voltages is to reconfigure the discharge diodes to be dumping back into the charging rail, in this case 15 volts. The previous version was convenient in that its output was the source for driving the MOSFETs and BJT control signals. In the 350V version this is not possible and therefore the diodes need to dump back into the control voltage bus. This could work, but would likely lead to unnecessary complexity of design and lacking safety. So other options should surely be favored.

The other option of design is the usage of integrated circuits. There exist many ICs which are designed to work in high frequency synchronous rectification applications where efficiency is a key concern. Usually this is added as the output rectifier to a switching mode regulator but in our case we can adapt the functionality of the IC to work as a full bridge for low frequency 60Hz mains input power. One such integrated circuit is the IR1167 synchronous rectifier controller. Some of the main characteristics which make this a solid choice for the application at hand are illustrated in the following table.

|  |  |  |
| --- | --- | --- |
| Characteristics of the IR1167BS | | |
| **Characteristic** | **Value** | **Importance** |
| Maximum gate drive current | Source 2A, Sink 7A | For a quick turn on and even more importantly a quick turn off to ensure no shorting of input power. Rise time of 125ns with a 10000pF gate capacitance. |
| Maximum VDS of controlled switch | 200V | Can handle mains voltage levels if we use one rectifier to handle positive side and one to handle negative side of DC link. |
| Gate clamp voltage | 14.5V | Ensures that the MOSFET is fully on, will account for even high voltage MOSFETs if we decide to change FET. |
| Quiescent Current | 1.8mA | This is important for the high side MOSFETs as they need a floating source to drive the IC while the high side is high. |

Table 18 - A summary of the characteristics which render the IR1167B an ideal MOSFET control IC for the synchronous rectifier section.

Implementing the IR1167 is quite simple in practice. Adding it to the topology in Figure 12 yields a topology shown in Figure 23. This model has a few important considerations built in. Firstly the integrated circuit has a built in minimum – on – time (MOT) pin which is used to prevent ringing from false triggering the turn off of the MOSFET immediately after it switches on. The pin does this by disabling the sensing of the drain – source voltage for a limited amount of time which can be programmed with a resistor to ground. The maximum value for this programmed delay is discussed in the datasheet to be about 3µs. This is not nearly long enough for our application at 60Hz frequency and very low RDS MOSFETs. Therefore a workaround must be devised. This workaround is implemented with an RC network between the gate drive and the source sense pin (one example is C9 R11 and R8). This RC network creates a decaying burst of current through R8 when the switch turns on, forcing the source sensing pin to read a higher value than is actually present on the source. This false reading will decay as the capacitor charges and will be unnecessary during the high conduction phase following because the RDS will create a large enough drop for the chip to not be false triggered. The time that this bias needs to exist is hard to predict but fortunately easily tunable with the series resistor to the capacitor. One would not make the other resistor (R8 in this case) too high, as this can distort measurements if the sensing pin’s bias current is high enough. Therefore adjustments should be made to R11 if needed. For initial design we can estimate that 10% of the total on or off time of the MOSFET may be distorted with ringing. A nominal calculation for the time that this charge will exist can be made with a simple RC time constant estimate and nominal values were selected accordingly. The design of this synchronous rectifier is based on the very similar design presented by Davide Giacomini and Luigi Chiné in their presentation paper. [13]

Further considerations needed to be made to allow the IR1167 to function with a floating ground. The chip requires that the VCC input be higher potential than ground by a minimum of about 12-20V in order to maintain the gate voltage at the specified 14.5V nominal. In order to achieve this a capacitor was added between VCC and GND pins of the high side chips. The capacitor is charged through a diode which is adequate to block the 160V or so that will be placed across it. Fortunately the capacitor does not have to be very large because it will only be maintaining the clamp voltage of the MOS capacitance as well as driving the low quiescent current of the IC (1.8mA). This will only need to be sustained for approximately 8.3ms (one half cycle at 60Hz frequency) and the calculation for the necessary size of this capacitor is shown below:

Equation 43 - Calculation of required bootstrap capacitor for high - side driver ICs.

The required minimum of 5.53µF should not be near the actual value of the capacitor. Because capacitors in SMD packages are extremely inexpensive and easily larger than the minimum with little to no added size or cost, it makes sense to grossly oversize the bootstrap capacitor to ensure that the IC gets plenty of energy to function for the half cycle without any chances of significant voltage drop of the capacitor. This was the motivation for choosing the value of 22µF; it is a common size and will easily supply all the power required. A specific measurement will be taken for each of the high side FET drivers to be sure the bootstrap capacitor is functioning properly.

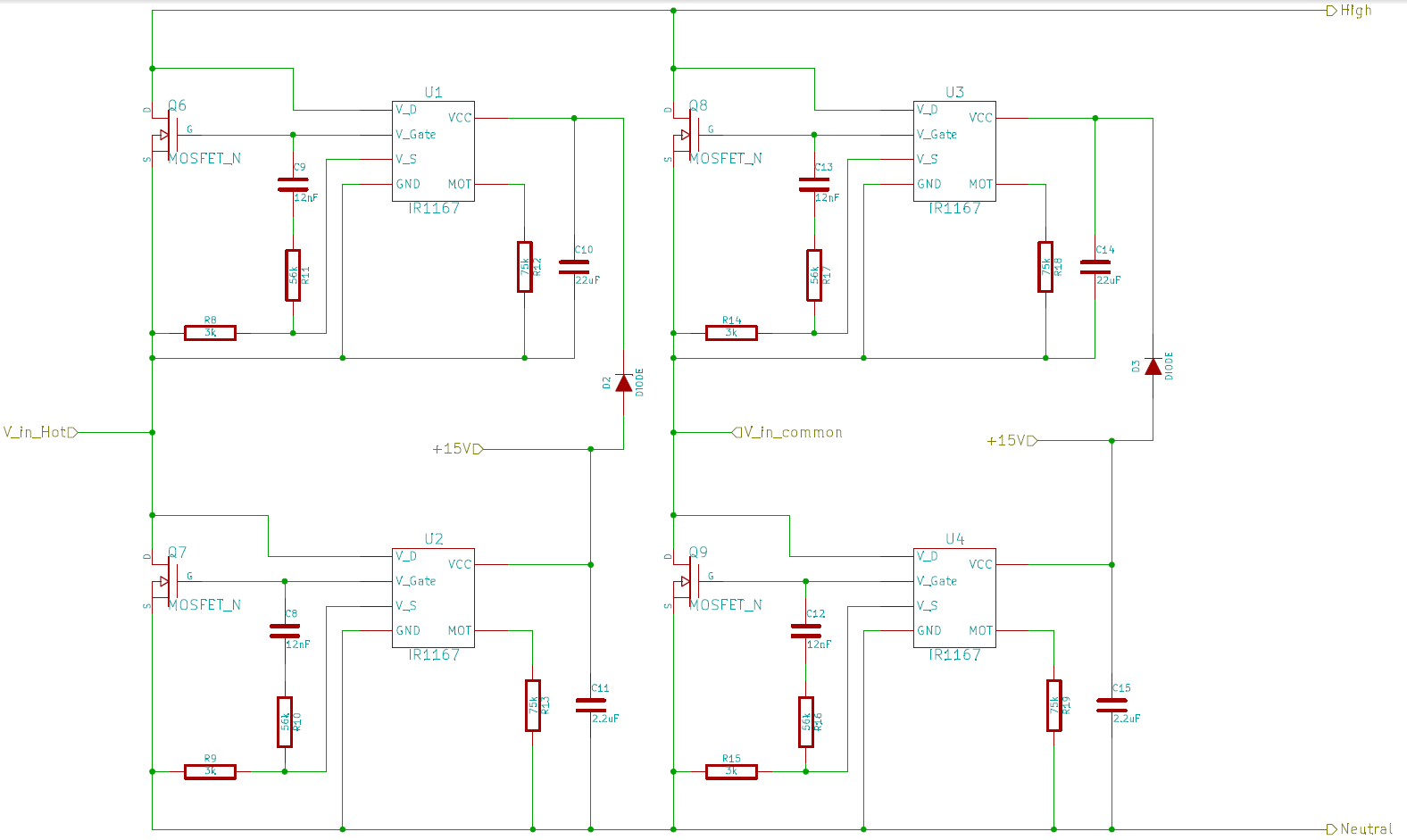


Figure 23 - One half of the synchronous rectifier using IR1167 ICs for MOSFET driving. Designed using KiCad

### Selection of Power Switches:

The first level of switch selection is type; should we use MOSFET, IGBT, or BJT. For this application we will consider mainly conduction loss and cost. The conduction losses of BJTs is characteristically lower than MOSFETs at high current levels. This is also true of IGBTs. However, at lower current levels the MOSFET’s low RON resistance has lower loss characteristics. These comparisons are summarized in Table 19. Upon consideration of the differences between the switching devices in question, as well as consideration of the choices made by various reference designs available publicly on the internet from multiple sources, it became clear that the power MOSFET was the standard of the industry for this task and would be the best choice for our rectifier simply due to the low conduction losses at the necessary current levels (not likely to exceed 20 amps at 240VRMS for any motor at or below 5hp)

|  |  |  |  |  |
| --- | --- | --- | --- | --- |
| **Comparison of Power Switch Types for the Synchronous Rectifier** | | | | |
|  | *Drive Type (Current or Voltage)* | *Conduction Loss Characteristics* | *Advantages* | *Disadvantages* |
| MOSFET | Voltage (gate capacitor charge) | Resistive if the gate voltage is held constant | Very low losses at lower current compared to either BJT or IGBT | RON is highly sensitive to device temperature and therefore require greater thermal care |
| BJT | Current (junction carrier injection) | Diode characteristics if the base current is maintained such that BJT is in on state | At high current levels, BJT exploit exponential characteristics of diode losses with current | With high current entering source, requires significant current driving the base to operate in the on state. |
| IGBT | Voltage (Gate capacitor charge) | Diode characteristics if the gate voltage is held high enough for the on state. | Similar to BJT, the IGBT excels at high current levels due to diode characteristics. | Higher switching losses than MOSFET due to minority carrier diffusion during turn off. |

Table 19 - Comparison of switch types for usage in the Synchronous Rectifier.

With the decision made to use a set of power MOSFETs for the power switches used in the synchronous rectifier, the next step in specification will be to compare various components which meet or exceed our design specifications. Given the specifications of the motor selected in Motor Selection section above, we plan to use a DC link voltage between 340 and 360 Volts. This will require that our power MOSFETs selected can withstand this voltage across their drain – source pins. We would also like the on – state resistance to be as low as possible, a reasonable initial specification is 50mOhm or lower typical at 25C. Close attention needs to be paid the temperature coefficient of this on resistance due to the fact that these are self – heating devices. Table 20 below shows the comparison between some of the MOSFETs which were considered for the role of the rectifier power switch. These MOSFETs all have very low RDS in the on state and all easily meet blocking voltage and current requirements for our application.

|  |  |  |  |  |
| --- | --- | --- | --- | --- |
| Comparison of Various Possible Power MOSFETs | | | | |
|  | **STY105NM50N** | **TK62N60X** | **TK100L60W** | **APT106N60B2C6** |
| **VDS Breakdown** | 500V | 600V | 600V | 600V |
| **RDS(on) typ.** | 19mOhm @ 52A | 33mOhm @ 21A | 15mOhm @ 50A | 35mOhm @ 53A |
| **Gate charge capacitance (pF)** | 9600pF | 6500pF | 15000pF | 8390pF |
| **Estimated temperature coefficient of RDS(on)** | 0.380mOhm per degree Celsius | 0.286mOhm per Degree Celsius | 0.104mOhm per Degree Celsius | 0.438mOhm per Degree Celsius |
| **Estimated RDS(on) at 100C** | 38.5mOhm | 55mOhm | 22.5mOhm | 63mOhm |
| **Device Package Options** | MAX247 | TO-247 | TO-3P(L) | TO-247-3 variant (T-MAX) and TO-264 |
| **Cost at 1 Qty.** | $20.27 | $10.24 | $31.20 | $17.44 |
| **Rated Drain Current** | 110A | 61.8A | 100A | 106A |

Table 20 - Comparisons between four different power MOSFETs which all possess desirable characteristics for our application and would all function adequately if chosen.

Upon considering the information presented in Table 20, it is immediately clear that when compared to the other available options, the TK100L60W is simply too expensive and has too much gate capacitance for the superior on resistance it has versus the others considered, therefore it is eliminated from this consideration. The APT106N60B2C6 has a very strong temperature coefficient when compared to the other remaining choices and also costs significantly more than the TK62N60X, therefore it too is eliminated. Between the STY105NM50N and the TK62N60x, the consideration is simple cost versus on resistance and gate capacitance. The difference of on resistance at 25C is significantly in favor of the STY105NM50N but when you consider its inferior temperature coefficient, it seems clear that despite its higher on resistance, the lower cost and lower gate capacitance option of the TK62N60X is the best option for maximizing cost versus efficiency.

### Estimate of expected efficiency:

It is a multifaceted problem to calculate the efficiency of the rectifier. The rectifier will only provide optimal efficiency at moderate – to high loads because losses intrinsic to the circuit will be swamped by conduction losses due to load current. Assuming this condition exists we can make a rough estimate of the final efficiency we expect to achieve merely by considering conduction losses we expect to see at a more nominal temperature than is used for standardized datasheet values (25 Celsius is usually used for RON figures).

Using information for the MOSFET selected in the above section; the TK62N60X, and assuming operation at a more realistic 100C junction temperature, we can calculate a rough estimate of the efficiency of our rectifier only accounting for the conduction losses of the MOSFETS, which will be the vast majority of the losses in the rectifier. The other losses such as switching losses, power used to drive the FETs and leakage in the circuit, are very small compared to the I2R losses of the conducting FETs. Beginning with an estimate of the power required from the rectifier, assuming 90% efficiency in the following high voltage inverter;

Equation 44 - Calculation of the necessary power rectified by the synchronous bridge.

Following the calculation above, we can then easily calculate the current which will flow through the MOSFETs. Assuming the drive operates at unity power factor;

Equation 45 - Calculation of the RMS current through the power FETs based on the power output and the nominal input voltage to the rectifier.

With the current known, and the FETs treated as a pure resistive load, the power loss is easily found as;

Equation 46 - Calculations of the amount of power lost in the FETs due to conduction of the necessary input RMS current.

And the efficiency naturally follows as calculated below:

Equation 47 - Calculation of overall rectifier efficiency based on the power lost in the FETs during conduction on required current at 120V RMS input voltage.

Taking into consideration other sources of power loss in the circuit, and assuming a less – than – idea operation of the FET, it seems safe to assume that our rectifier will be at least 95% efficient if it is operating normally. This is well within design specification and keeps power losses relatively low. Dissipating 65W in a heatsink is not very difficult, and will not require a very large heatsink at all.

### Specification of DC Link Bulk Capacitors

The DC link needs a large amount of bulk capacitance to be able to supply the high inrush currents of the induction motor. The bulk capacitance requirements were roughly estimated using the following series of calculations:

Equation 48 - Calculation of the peak operating current on the motor phases. All standard SI units for P, I, and V. This assumes PF=0.6, rated power 5hp, and motor efficiency of 90%

Equation 49 - Calculation of the peak inrush current expected for a motor with **XX** class operating characteristics and peak operating current above.

Important to note here is that the inrush current is 96A not 3\*96A because of the usage of space vector pulse width modulation. SVPWM guarantees that only one high side switch will be conducting at any time, thus guaranteeing that the instantaneous load on the DC link is that of only one phase. For this current condition, the ultra-high inrush current will only exist for the time it takes the synchronous rectifier circuit to catch up to the rail load, at worst the rectifier may take up to one that time could be anything up to approximately one quarter cycle, the time until the input transformer secondary can rise from neutral to a high or low which can be rectified. Therefore the formula used before to calculate decoupling capacitance for the bootstrap capacitors can be implemented:

Equation 50 - Estimate for required capacitance on the DC link to supply inrush transient with low voltage drop.

This value is somewhat of a fudge factor right now, because cost being a large player in the selection of these capacitors we need to be aware of the chance of using larger amounts of cheaper capacitors to achieve the necessary ripple current. If the ripple current hits 95A at peak then with four in parallel for each of the high and low voltage rails will mean each must be able to handle around 24A pulses at 20 kHz frequency. This specification proves costly to satisfy, however, simply because of the ripple current. The capacitor requirements will need to be researched further going forward in design to specify the necessary link capacitance to provide satisfactory energy storage for the output inverter.

### Test Plan: Synchronous Bridge

The first power up will be accomplished using a standard benchtop signal generator with built in protection circuitry. This will hopefully prevent any overload in our circuit and therefore provide safety for both the circuit and the test operators (us). The initial voltage input will be started at 20Vpp as this the standard maximum for the power supplies we have access to. The initial power up will be with a very light load, less than 500mW to accommodate the output limitations of the function generator. This load will be a static resistive load calculated as follows:

Equation 51 - Load resistance calculated for dissipation of 500mW at 30VDC output.

The initial test condition is purely to verify rectifier functionality and switching operation. At this point the oscilloscope will be connected to the input, the output, and the gates of both power switches on either the positive or negative side of the rectifier. Observations and screenshots will be taken of the aforementioned values and used as initial verification that the rectifier is operating properly. Special attention needs to be paid to any transient effects of the output smoothing inductor on the gate voltages, as these transients, if not shunted to ground, will reduce wear life or destroy the MOSFETs all together. Proper operation of the bootstrap circuit which drives the high side FETs will need to be verified at this point to ensure the drivers are getting plenty of power to operate normally. Measurements will also be taken for the VDS of the MOSFETs to verify both correct on – state resistance and proper switch drive current provided. If we do not observe strong on / off operation of the switches at this level, serious reselection of drive components needs to me made before we go any further as improper switching will result in large power dissipation and loss of efficiency, as well as the possibility for a short circuit of the input voltage resulting in fuse failure or worse later down the road.

Power testing of the rectifier prototype will begin with 120VRMS mains power as the source. A variable autotransformer will be used to slowly ramp the voltage from 0V through 120V mains voltage. Current limiting and sensing power resistors placed on the inputs to both protect the circuit from over – current and to allow easy measurement of input currents for testing purposes will be implemented. Both of these steps will hopefully prevent a rapid failure of components if issues with power dissipation arise. Two oscilloscope probes will be attached across this resistor to track input voltage and current. We will attach a load resistance to the output of the rectifier specified to allow approximately 50 watts output. The specification is calculated in the following equation:

Equation 52 - Load resistance calculated for dissipation of 50 watts on a 170VDC output.

The load resistor must be selected for high power dissipation and will require some sort of heatsink – likely just a large aluminum scrap at hand as this is just a temporary testing rig, no need to over – engineer it.

Beginning a slow ramp to the full 120VRMS of the mains, keep a close eye on the current levels, temperature of the switch heatsink, and the output power. Measurements should be taken across the power switches to verify proper switching operation once again, given the immensely increased load. Once the maximum input voltage is reached, measurements for power in and power out should be taken and efficiency calculated. This number will hopefully fall over 85% for this loading condition but will definitely not be within the optimal 95%+ range because of low loading conditions. Analysis of losses shall be conducted at this point to determine where the power is being dissipated, likely power losses in this power range will be the gate driving currents.

## Low Voltage Bus Regulation

A set of low voltage DC regulators will need to be implemented to drive the low voltage electronics; the microcontrollers, the sensor drivers, LCD display, gate drivers, and any other low voltage peripherals that become necessary during actual construction. To create these busses we could use the high voltage DC rails we will already be creating to drive the inverter circuit, however that would be inefficient and the input voltage of the first regulator would be very high. Much higher than general, widely available voltage regulators can handle. Beyond the efficiency issues, for safe operation it is imperative that the microcontroller and associated sensors are fully functioning before the high voltage DC link is energized at all. This implies that we require a separate power system in parallel with the high – voltage – high – power synchronous bride, DC link, and inverter. The devices we will be powering all fall onto two voltage rails described in the following table:

|  |  |  |
| --- | --- | --- |
| Required Low Voltage DC Busses | | |
| Bus | Components on bus | Estimated maximum load |
| +15V | MOSFET and IGBT drivers, These components can take significant current at high switching frequencies so we want to be sure this bus has plenty of power capability. | 3A |
| +3.3V | Microcontrollers, LCD Display, and sensor drivers. All low power items, the LCD, if it works on 3.3V will go here, it has not been selected. This bus is oversized for its application to accommodate future expansion. | 1A |

Table 21 - Brief overview of the required low voltage DC busses.

### Specification of Input Transformer

Considering the required busses described in Table 21, the method of approach is to use a low voltage step down transformer in parallel with the main power transformer. This low voltage transformer will be energized immediately upon the drive being connected to mains power, and will immediately supply power to a low voltage bridge rectifier and a cascade of two switch mode buck regulators with significant ripple filtering to yield relatively quiet DC busses for integrated circuit and sensor power. In order to assure that the first regulator always has sufficient input voltage but is never over driven the selection of transformer winding ratio is key. Fortunately because modern regulator ICs have relatively high maximum input voltages (up to 30V in many cases), we can easily specify a ratio of 16:1 for the primary – secondary turns ratio. The following two equations verify the viability of the selection of 16:1 turns ratio showing that the output of the transformer will have sufficient peak value even after a volt or so is lost in the rectifier.

Equation 53 - Output voltage of transformer with 120VRMS input.

Equation 54 - Output voltage of transformer with 115VRMS input.

The current specification of the windings is rather flexible and most wisely over designed for ease of assembly and added efficiency. An estimate for the amount of current needed in the primary to supply the necessary power can be computed as follows, using the approximation that power in will be approximately power out, this is only an approximation:

Equation 55 - Large overestimate of the primary RMS current needed to supply the low voltage bus.

And the secondary current rating will just be eight times the current determined above for the primary; 2.78ARMS. Both of these are so low that over specifying the winding is a matter of choosing the best price at anything well over this specification.

### Rectifier Specification and Selection

The low voltage DC bridge rectifier will need to be efficient as is always the case on this project, but because of the low power involved (merely up to about 4A­RMS rectified) the rectifier does not need to be synchronous, it just needs to be a specifically efficient Schottky bridge rectifier as described in the Full – Bridge Diode Rectifier section. The component selection was focused on making sure the current rating was adequate, and then simply prioritizing a good price for low forward voltage drop. Leakage is not a big issue here as it will contribute very little to any significant losses or cause issues elsewhere. Diode arrays were also considered for implementation but proved to provide inferior forward voltage characteristics. Based on these criteria the NXP PMEG3050EP Schottky diode was selected to serve as our bridge rectification diode for the low voltage bus. The PEMG3050EP provides a Vf­orward­ voltage drop of 315mV at 5A current rectified. Figure 24 shows the topology of the Schottky bridge rectifier to be used with the low voltage DC bus regulators described in the following sections.

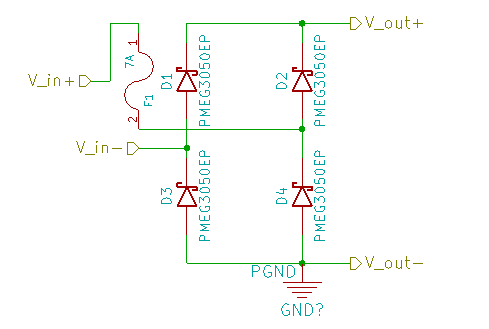


Figure 24 - The topology of the low voltage rectifier (approx. 21V). Component selection is justified above.

Approximating on the plots provided in the datasheet shows that at temperature of approximately 55°C and forward current of 4A we can expect a voltage drop of approximately 300mV, only slightly less than the 315mV 5A rating. At that rating we can expect power losses to be approximated by the following equation:

Equation 56 - Calculation of an estimate of the expected efficiency for the Schottky diode rectifier shown in Figure 13.

This number will be verified as part of the test plan, as it is the key statistic for the rectifier, being that the regulators it will be powering will both be designed for very high efficiency per design specifications.

### 15V Regulator IC Selection

The first buck regulator will drop the rectified unregulated DC (approximately 20-22V) into 15V. This 15V rail will power both its associated electronics and the 3.3V regulator for the low power electronics. Therefore the 15V regulator will need to be specified at approximately 150% the current level of the 3.3V regulator. This suggests that a good upper range of current specification would be three amps for the 15V regulator. Using parametric search functions prom large online retailers as well as direct catalogs from manufacturers (if necessary) selection of the appropriate IC mainly requires a list of parameters for refining a search.

|  |  |  |
| --- | --- | --- |
| Initial Parameters for IC Selection | | |
| **Parameter** | **Value(s)** | **Justification** |
| Output Current Maximum | 3A | Safe high – end estimate for the maximum current this regulator will see. Allows for future expansion without having to pick a new regulator. |
| Input Voltage Range | 18V < Vin < 25V | Safe range that includes all possible Vin predicted |
| Output Voltage | +15V or adjustable to it | Rail voltage desired |
| Efficiency | Greater than 90% over 1A-4A range | That range encompasses most loads expected. |
| Synchronous | YES | The only way we can expect to see the 90%+ efficiency we want. |
| Price | Less than 15 Dollars | Reasonable, seeing as we only need one. |

Table 22 - Base parameters for 15V IC parametric search and their justifications.

As discussed in the Choice to Favor Well Reputed IC Manufacturers section of Specific Design Considerations; for most of our integrated circuits, if possible, we would prefer to use products made by top manufacturers. Therefore, taking that into account, reviewing datasheets, and attempting to find a good balance of performance and price, the LT8640 was selected as the voltage regulator IC we will use for our final prototype. One thing that is essential to note with LT8640 is that it comes in a QFN package with solder pads underneath the package. This renders it very hard to prototype with, unless a great breakout board is found/made. Therefore the LT8640 will likely be our final regulator, not the initial. For the initial rough prototype to test the theory of operation we will likely use a 15V benchtop power supply to supply the rails in testing before we print power supply boards to implement small SMD packages.

|  |  |  |
| --- | --- | --- |
| Parameters of Interest for the LT8640 | | |
| **Parameter** | **Value(s)** | **Justification** |
| Output Current | 5A | Similar price to 4A variant, more efficient over larger current range. Over specified regulator allows for easy additions later if necessary. |
| Input Voltage | 3.4V < Vin < 42V | Easily handles any input voltage we will supply it in our application. |
| Loaded Efficiency | Greater than 95% efficient from 0.5A to 4.5A. | Surpasses our efficiency specification by 5% |
| Switching Frequency | Adjustable 300kHz – 3MHz | Allows tuning for optimal efficiency in our application. |
| Output Voltage | Adjustable 0.97V – 42V | Huge output voltage range, easily suits our needs. |
| Synchronous Rectifier | YES | As specified |
| Price | $9.30 | One of this chips down sides. Cost is quite high. Design notes will make implementation simple though. |

Table 23 - Justification for selection; Parameters of interest for the 15V regulator; the Linear Technology LT8640.

### 15V Regulator Circuit Topology and Passive Component Selection

One of the reasons which Linear Technology was chosen over their competitors was that their datasheets are exceptionally helpful. And the LT8640 datasheet is exceptionally helpful for ease of design. Starting with the provided reference topology for the IC, all that must then be done by the user is to specify a few component values using provided equations. Specifically we must determine the inductance of the inductor, the inductors current specification, and its saturation specification, set the output voltage feedback divider, set the frequency of switching, and design any necessary decoupling or additional features as needed.

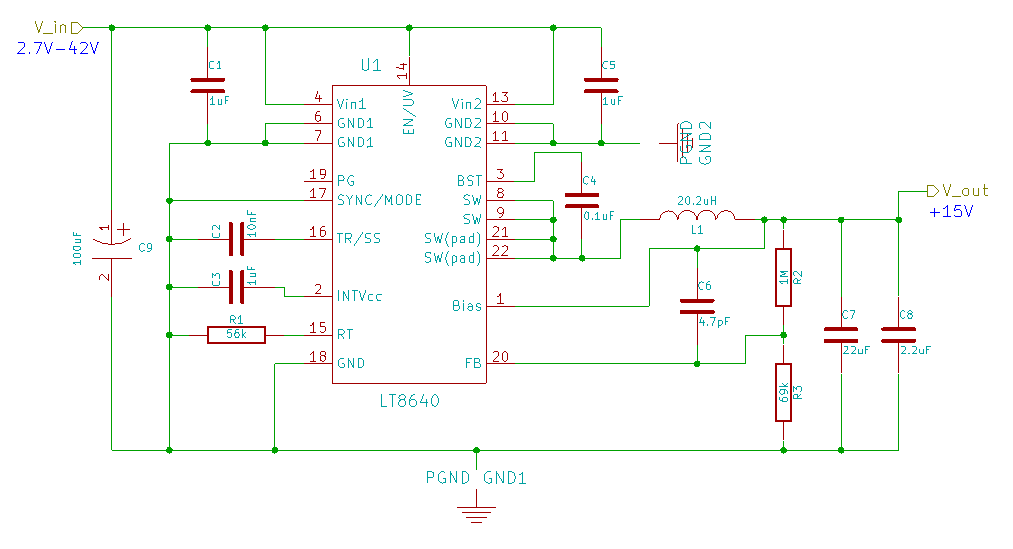


Figure 25 - The topology of the 15V synchronous regulator, design was made using provided reference design.

The first specification to specify is the switching frequency. Upon considering the graphs provided in the datasheet, for most efficiency operation point for the switching frequency is about 750 kHz. To program that value simply use the lookup table provided on the datasheet and use the corresponding resistor, for 750 kHz R = 56.1k. R1 in Figure 25 is the resistor used to program fsw­. The next specification of interest is the inductor value. The datasheet provides a formula recommended for selecting the value of the inductor as a function of output voltage and switch frequency. Equation 57 below shows this calculation.

Equation 57 - Calculation of inductor value using the formula supplied on the LT9640 datasheet.

Now that the inductor value is specified we can use Equation 58 provided on the datasheet to calculate the specification for the saturation current of the inductor core. It is also notable that a core material specified for high frequency switching should be selected to reduce heating.

Equation 58 - Calculation of the required inductor saturation current rating using the formula supplied in the LT9640 datasheet.

With the inductor fully specified attention must now be directed to the output voltage programming which is set with a feedback voltage divider. The regulator regulates the feedback (FB) pin to 0.97V, knowing this and knowing our desired output voltage of 15V we can specify the ratio of R2/R1 to achieve the desired output voltage. Applying voltage divider law as follows in Equation 59:

Equation 59 - Calculation of the feedback gain to achieve Vout = 15V. Assumes that the bias current into/out of the FB pin will be negligible.

To pick actual values the amount of current through the feedback divider must be considered, the LT9640 FB pin bias current is +/- 20nA max per datasheet specification. It is wise to make sure the divider current is many hundreds of times the bias current of the FB pin to render the FB current negligible in Equation 59. Also important to consider is that the current is as low as possible to optimize low load efficiency. Considering both of those goals, the values of R2 = 1MΩ AND R1 = 69.1kΩ. Given that 69.1k are very hard to find, 69k will be fine as reflecting in Figure 25.

Considering the behavior of the SYNC/MODE pin, the datasheet specified that for burst operation to be enabled for loads under 400mA, the SYNC/MODE pin should be tied to ground as shown. The PG pin is the power good indicator, it is pulled down through an open drain until the output voltage is within 8% of its final value, and will not be allowed high if there are any fault conditions. The PG pin will be useful during testing to know whether or not the IC thinks it is providing good power in the event of an issue.

The remaining capacitors are selected based on the datasheets recommendations. The specifications are outlined in the table below:

|  |  |
| --- | --- |
| **Specifications for Capacitor Values** | |
| **Capacitor number – name** | **Specification Statement** |
| C1,C5 – Input decoupling capacitors | Datasheet requires a value of at least 1μF very close to the pins. |
| C2 – Slew rate setting | Limits the slew rate, reference designs used 10nF but datasheet says the pin can be left floating. |
| C3 – INTVcc decoupling capacitor | Datasheet specifies a value of at least 1μF decoupled to power ground. Needs to be low ESR ceramic. |
| C6 – Phase leading feedback capacitor | Specified by datasheet to improve transient response. Typical values are given as 3.7pF – 22pF |
| C7,C8 – Output decoupling capacitors | Filter ripple on output current, reducing high frequency noise. This regulator is specifically designed for low noise applications. |
| C9 | Input bulk capacitor, likely will be a supercapacitor to ensure a good bulk capacitance if high load transients occur. Could be traded for a large electrolytic. |

Table 24 - Specification of 15V regulator capacitor values.

NOTE: Full schematics showing the interconnections of rectifier, 15V regulator, and 3.3V regulator can be found in

### 3.3V Regulator IC Selection

Similarly to the selection of the 15V regulator, there are numerous important characteristics to take into account when selecting the 3.3V regulator IC. The parameters used to narrow the selection of eligible 3.3V regulators are described in Table 25 below.

|  |  |  |
| --- | --- | --- |
| Initial Parameters for IC Selection | | |
| **Parameter** | **Value(s)** | **Justification** |
| Output Current Maximum | 1A | Safe high – end estimate for the maximum current this regulator will see. Allows for future expansion without having to pick a new regulator. |
| Input Voltage Range | 14V < Vin < 16V | Vin will always be a 15V nominal once the 15V regulator reaches regulation. Therefore this is the only value needed for input voltage specification. |
| Output Voltage | +3.3V or adjustable to it | Rail voltage desired |
| High Load Efficiency | Greater than 85% at 1A load | Extreme high end estimate of possible loading conditions |
| Low Load Efficiency | Greater than 85% at 10mA load | Extreme low end estimate of possible loading conditions |
| Synchronous | YES | The only way we can expect to see the 90%+ efficiency we want. |
| Price | Less than 10 Dollars | Reasonable, seeing as we only need one. |
| Package | SMD w/leads | Leads are easier to solder than pads underneath the IC. The 15V regulator has a pad WFN package which is not ideal. Avoid here if possible. |

Table 25 - Base parameters for 3.3V IC parametric search and their justifications.

Upon consideration of multiple ICs from multiple vendors, once again the combination of adequate specifications and exceptional datasheet contribute to the Linear Technology LTC3621 being selected as our 3.3V regulator. The regulator has many positive attributes, most significant of which are used in Table 26 to justify the selection of the LTC3621 as a highly satisfactory choice for the 3.3V rail regulation. One specific pro that this regulator has is its package; it comes in an 8-lead MSOP. These packages are much easier to work with by hand than QFN packages like the LT9640. It will be possible to buy a breakout board for prototyping this regulator from the start.

|  |  |  |
| --- | --- | --- |
| Parameters of Interest for the LTC3621 3.3V Regulator | | |
| **Parameter** | **Value(s)** | **Justification** |
| Output Current | 1A | At specification |
| Input Voltage | 2.7V < Vin < 17V | Includes our input of 15V |
| Output Voltage | 3.3V | Not variable output, but we don’t need it to be. 3.3V is a common output voltage and using a fixed output regulator eliminates two discrete components. |
| High Load Efficiency | 90% efficient at 1A load. | Good to have the efficiency at high loads if power requirements are above nominally expected. |
| Light Load Efficiency | 90% efficient at 3mA load with burst mode operation enabled. | There are good chances of very low load condition at times, during these times having high efficiency is still a design goal. |
| Price | $5.45 | Good price for a good regulator. Design is simple. |

Table 26 – Justification for selection; Parameters of interest for the 3.3V regulator; the Linear Technology LTC3621

### 3.3V Regulator Circuit Topology and Passive Component Selection

Unlike the LT9640, the switching frequency of the LT3621 is not programmable, and therefore not a consideration in our peripheral design. This simplifies design significantly as FB is tied directly to the output voltage. Like the LT9640, the LTC3721 has a PGOOD pin which indicates when the output rail is within about 6% of its set value, creating a signal which could be used in diagnostics to check if the regulator “thinks” it is putting out the correct voltage if we encounter an issue in prototyping.

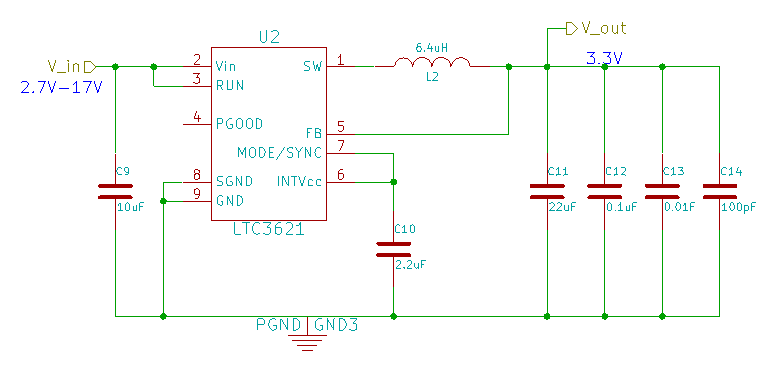


Figure 26 - The topology of the 3.3V synchronous regulator. Design was made using provided reference design.

The first specification for the LT3621 is the inductor value. Equation 60 was used to calculate the value of the inductor for the circuit in Figure 26. Where ΔVin(Max) is the maximum allowable ripple voltage. The datasheet suggests 40% of maximum output current so we will use 0.4A. We can also specify the inductor to be rated to carry 1ARMS current due to the max output load we have specified. Following this specification we can also say that the inductor saturation should be calculated as in Equation 58, yielding a saturation requirement of 1.2A. Using the equation below we can calculate the nominal value of the inductance needed and that will conclude our specification of the inductor.

Equation 60 - Calculation of the nominal inductor value, equations were given in the datasheet.

The datasheet discusses specifications of input decoupling capacitors and output smoothing capacitors, the specifications for the capacitors are described in Table 27 below and are all inexpensive simple parts to choose which are not worth taking space here for discussion.

|  |  |
| --- | --- |
| **Specifications for Capacitor Values** | |
| **Capacitor number – name** | **Specification Statement** |
| C9 – Input Bypass | Datasheet specifies 10uF on all of their reference designs. |
| C10 – INTVcc Bypass | Interval Vcc bypass, datasheet specifies that INTVcc pin be bypassed to ground with at least 1μF. Chose 2.2μF to de-rate a bit. |
| C11 – Output smoothing | Smooth output ripple: datasheet uses 22μF for all of its reference designs. Added 0.1, 0.01, 100pF to reduce noise at the microcontroller power input. |

Table 27 - Specifications of the external capacitor values for the LTC3721. Summarized from the datasheet.

It will not likely be necessary to add additional capacitive storage to the input due to the fact that large transients are not expected on the 3.3V rail, as well as the fact that the 15V regulator it is powered by has a very large bulk capacitance.

### Low Voltage Rail Regulation Test Plan

#### Input Transformer:

Measurements must be taken at the transformer inputs and outputs especially if we opt to wind the transformer ourselves, to ensure that the transformers output is within nominal value calculated by the turns ratio. If the transformer passes this initial test of safe operation the rest of its load testing will be satisfied by testing the low voltage rectifier in series with the transformer. We can say that all operation is normal if all of the bridge rectifier tests outlined below yield nominal results.

#### Bridge Rectifier:

The low voltage regulator will be simple compared to the high voltage rectifier because the power and voltage levels are much lower. The tests will essentially be to test the circuits individually to confirm proper individual operation. For each regulator circuit this will include efficiency measurements to better characterize our losses and tune components accordingly. With the rectifier we want to see the predicted output with appropriate no – load DC voltage approximated below:

Equation 61 - Estimate for the minimum expected no-load output voltage of the LVDC rectifier.

Equation 61 uses Vf specified in the datasheet for a 5A load, this was used in order to give a bottom end requirement to be considered functioning. The next test performed will be to test the full load condition. Using appropriate output resistance calculated using Ohm’s Law (V\_out / I\_load) to be approximately RL = 5Ω. Output voltage and ripple will be monitored and compared against datasheet specifications for the Schottky diodes, ripple current in the large capacitors will be measured and confirmed to be within datasheet specified values. Temperatures of the diodes will be monitored and confirmed to be within safe ranges over time under heavy load conditions.

#### 15V Regulator:

The 15V regulator testing requires confirmation of all nominally specified components in design. The testing party must supply approximately 20VDC input from DC power supply to the input pins with current limit set to about 10mA on the supply for initial power – up. Assuming there are no issues the current limit can be increased appropriately. Various loads will be purely resistive in nature. For the first check of functionality the output of the regulator will be measured and checked to be within about 5% of the specified value (.

Assuming that the regulator is functional in the initial test and output is within specification, then probes will be taken at various locations in the circuit to confirm good functionality. Table 28 lists the measurements taken for the 15V regulator and the purpose of each measurement to confirming proper functionality.

|  |  |  |
| --- | --- | --- |
| **Measurements Affirming Proper Function: 15V Regulator** | | |
| **Measurement (sym)** | **Predicted Value(s)** | **Details** |
| Switching frequency ( | 750 kHz | Value specified in design, easily measured with oscilloscope prove applied to the switch pin pad test point placed on the PCB. |
| Output Ripple ( | 0.1% | Taken at full load condition (5A). |
| Input Ripple ( | 0.1% | Taken at full load condition (5A). |
| INTVcc ( | 3.5V | Datasheet specification for Regulation. |
| Package Temperature ( | Les s than 85°C | Recommended safe operating limit for device package. |
| Output Slew Rate | See Note\* | Tested by switching in and out a resistive load. |
| Efficiency at various loading conditions (η) | >95% @ 1 - 2.5A | The optimal efficiency range will be found and attempts to tune it by changing inductors iteratively may be important for the initial prototype. Implies measuring power in and power out for a range of loading conditions and plotting the results for analysis. |
| Light Load Efficiency (ηlight­) | >95% @ 10mA | Predicted value based on datasheet plots. |
| Load Regulation |  |  |

Table 28 - Measurement specifications for the 15V regulator.

Note\*: Slew Rate was briefly mentioned on datasheet, but no information on the calculation of the TR/SS capacitor value for a desired slew rate was given.

If all of the criteria describe in Table 28 are satisfied then the 15V regulator can be deemed fully functional and implemented in sequence with the rectifier and transformer to test the three sections in series. Similar tests will be performed for load transient analysis and efficiency measurements of the whole 15V system. Documentation of testing will be made and included with the final report. This is true of all testing done on the project.

#### 3.3V Regulator:

Similarly to the 15V regulator tests, the 3.3V regulator testing centers around first confirming basic functionality of the regulator, then running it through a series of input, output, and efficiency measurements to confirm operation totally within specification per datasheet information and design choices. The 3.3V regulator will be supplied 15V with a DC power supply and a safe 10mA current limit for initial power – up. Current limit can be increased with observed safe operation. And again, the loading conditions will be purely resistive. Table 29 below specifies the measurements taken on the isolated 3.3V regulator system to confirm proper operation within design specifications.

|  |  |  |
| --- | --- | --- |
| **Measurements Affirming Proper Function: 15V Regulator** | | |
| **Measurement (sym)** | **Predicted Value(s)** | **Details** |
| Switching frequency ( | 1 MHz | Value specified in design, easily measured with oscilloscope prove applied to the switch pin on the package. |
| Output Ripple ( | 0.1% | Taken at full load condition (1A). |
| Input Ripple ( | 0.1% | Taken at full load condition (1A). |
| Package Temperature ( | Less than 85°C | Recommended safe operating limit for device package. |
| Output Slew Rate | -- | Tested by switching in and out a resistive load. Not a specified value in the datasheet but is interesting characteristic of the transient response. |
| Efficiency at various loading conditions (η) | >90% @ 0.2A-1A | The optimal efficiency range will be found and attempts to tune it by changing inductors iteratively may be important for the initial prototype. Implies measuring power in and power out for a range of loading conditions and plotting the results for analysis. |
| Light Load Efficiency (ηlight­) | >90% @ 4mA | Predicted value based on datasheet plots. |

Table 29 - Measurements specifications for the 3.3V regulator.

When the 3.3V regulator prototype achieves values within the specifications of Table 29 above, we can confidently say that the regulator works as specified by datasheet and design. The 3.3V regulator will then be connected to the output of the 15V regulator/transformer combination for the final series of tests.

#### Tests of Full Rail System:

Once individual subsystem operation is confirmed for the rail regulator system, the full system can be connected and tested for proper functionality. Mainly of concern here are power up behavior, load transient response, efficiency, input and output noise, and load regulation over the full range of supply capabilities. The Vin to 3.3V regulation should be around 85% efficient excluding losses in the input transformer. The input transformer will be designed to optimize efficiency but if possible we will buy a purpose – built transformer specified to maximize final design efficiency, this specification can only be made through iterative design and prototyping in the future.

## Power Inverter

The power inverter converts the DC link voltage into a pulse width modulated, quasi-sinusoidal, three phase output waveform to drive the attached AC induction motor that the system is controlling. The inverter receives commands for the switch states from the Piccolo controller, decouples them from switching noise, and drives the power switches accordingly. The inverter also assures that both IGBTs in a leg will never be on at the same time, shorting the DC rail. The inverter subsystem also contains the sensors for motor phase voltage.

### Inverter Schematic Design

The power inverter is the point where DC voltage on the high-voltage DC bus is converted into a pulsating AC voltage on the load. A basic topology for a 3-phase IGBT inverter circuit is shown in Figure 4. The six gate terminals left floating in Figure 4 will be attached to an IGBT driver IC with appropriate protection and limiting circuitry. Protection and limiting circuitry will be composed of diode configurations.

One very important note to make here is the reverse-parallel diode used in conjunction with each IGBT. This diode is essential to the function and reliability of the circuit. These diodes protect the IGBTs from any transients that will place a high voltage across the emitter – collector junction. IGBTs are not designed to handle current flowing into the emitter, and if this is allowed to happen, serious damage to the power switches can occur. As a result of this issue, most modern IGBTs come with a built-in diode in the aforementioned configuration. These diodes work well, however for safety and longevity, our design will include discrete fast-recovery diodes in parallel with the built-in diode. Figure 27 shows the initial design used for the three phase inverter portion. The IGBT’s are left generic to be specified in the proceeding section. Drive circuitry and integrated circuits are also not shown in Figure 27 to emphasize the simple design that will be implemented. An interesting consideration is that of whether or not we should be using the same IGBTs for the high side and low side switches. In some pulse width modulation algorithms the low side switches switch at a much lower frequency that the high side. In SVPWM, however, all switches switch at the same frequency, therefore this is not a concern.

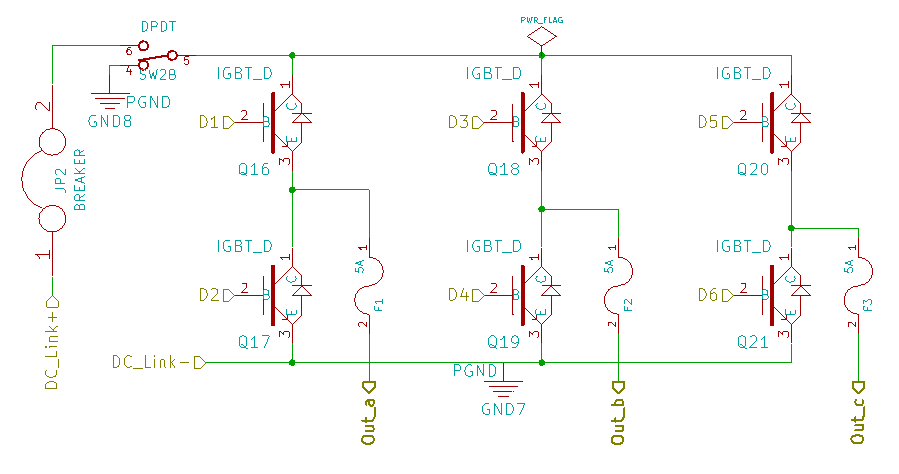


Figure 27 – 3 Phase Inverter using IGBTs with built – in antiparallel free-wheeling diodes.

Working from that basic design above, we need to specify the IGBT we will use as the switch, the IC which will drive the IGBT gates, consider the optional addition of an external discrete freewheeling diode, and design the bootstrap circuitry for the high side drivers. After those specifications are made we can attempt to make specifications as to the values needed in the output filter placed between the drive and motor to smooth switching noise. Protection circuitry to limit current flow should be considered for the switching branches.

Considerations must be made for driving the high side IGBT gates. Similar to the solution in the rectifier, bootstrap circuitry is implemented on the gates of Q1, Q3, and Q5 to allow the gate drive IC to operate nominally. Many gate driver ICs already have bootstrapping in mind for their application. This will be considered in the appropriate section after selection the power switch in the following section.

Working from information laid out in the section below, the FAN7190 was chosen as a decent first IGBT driver because of its low cost. The datasheet did not provide any reference designs to work with. Therefore design was based on reference designs for similar parts from TI. Figure 28 on following page shows the topology of the synchronous rectifier. The bootstrap capacitors were estimated both using Equation 43 as well as a tip from TI which recommends a capacitor at least ten times greater than the gate capacitance. 22μF was chosen because we will already be buying them for the synchronous rectifier and they are plenty large enough.

The diode specification is simple, requires the ability to pass substantial current of over 1A, as well as the ability to block 400V+. The losses in these diodes could prove to be a factor in our overall efficiency and if they do we may need to specify them for lower loss options such as Schottky.

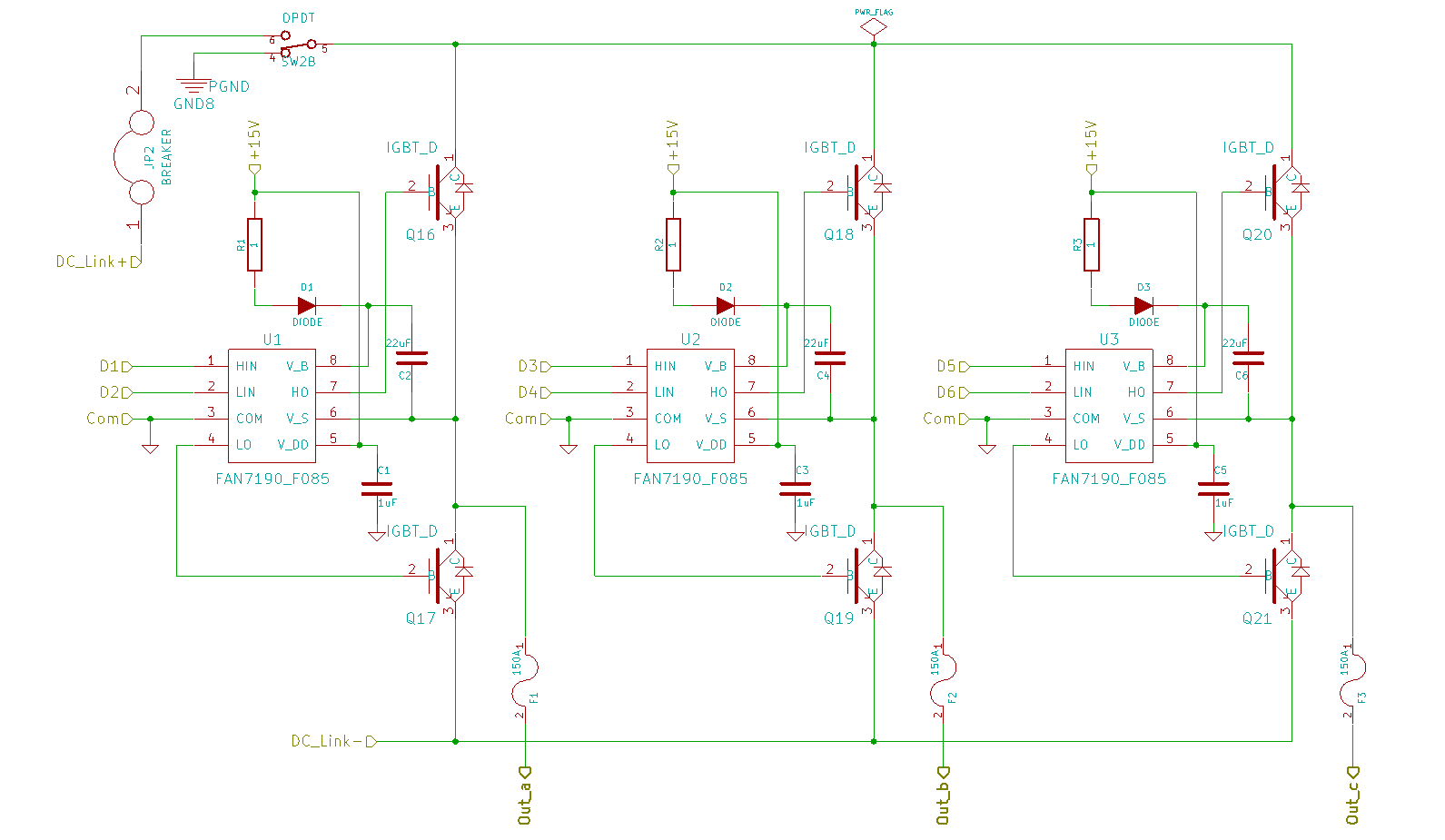


Figure 28 - The topology which will be used for the synchronous rectifier. Based on the FAN7190. Designed in KiCad

### Selecting IGBTs for power switching.

It is noted that some sources suggest that different switched be chosen for the high side and low side IGBTs. The reason being is that during sinusoidal PWM using classical control algorithms, the low-side switches only switch at the output sinusoid frequency (low) and the high – side switches switch at the PWM switching frequency (high). This means that the high – side switches often have much more focus on switching losses, while the low – side switches need to be specified to minimize conduction losses. In our application, however, because of space vector PWM, the high and low side switches will be switching at similar frequencies and therefore it will be fine to specify one switch with both fast switching characteristics and low conduction losses.

Initial specifications for the switch were made taking into account our application requirements, specifically the maximum inrush current calculated in Equation 49 - Calculation of the peak inrush current expected for a motor with **XX** class operating characteristics and peak operating current above. It will be wise to select an IGBT which can handle this level of current in normal operation for better lifetime of the IGBT, however as long as the IGBT can handle the 95A in pulses that should be fine, given as the ramping startup will use extremely narrow pulses at first (the first pulses will be by far the highest current). To ensure good control resolution especially for low amplitude starting signals, a controllable range of 1%-99% duty cycle is desirable. This means that at maximum the time to switch the IGBT on and off once must be less than 1% of a 50μs period. Therefore we specify that the on time and off time of the IGBT summed must be less than 500ns, or 1% of the possible pulse duration. If the desired pulse width is between zero and one percent, it will be rounded to zero, and if it is over 99 percent, it will be rounded to 100%.

Our inverter will need to be capable of producing output signal amplitude at minimum to drive our first motor. This means that the IGBTs will be blocking the DC link voltage which will be as high as 340VDC. This is not an exceptionally high specification and if the IGBT we select is capable of over 350V blocking that will be more than sufficient. Furthermore, the IGBT must be able to easily handle the inrush current of the motor. As calculated in Equation 49, the inrush current will not exceed 100A, so a specification of 100A current capability is safe. These and further specifications for the IGBT operating characteristics are specified and justified in the following table.

|  |  |  |
| --- | --- | --- |
| Initial Parameters for IGBT Selection | | |
| **Parameter** | **Value(s)** | **Justification** |
| Operating Current ( | Greater than 100A | This level will easily handle any predicted load for a 5hp motor. |
| Collector – emitter voltage | Less than 1.8V at specification. | Needs to be low for high efficiency, should stay low even under very high current |
| Type | Trench | For high frequency switching Trench are specifically designed to maximize efficiency |
| Package | Simple heatsink mountable power package | TO-220, TO-3P, TO-247, etc. All provide easy mounting and replacement in the event of switch failure. |
| Cost | Under $8 | We need six and will likely need replacements, so cost should be kept as low as possible. |
| Efficiency | >98% in conduction at 50A | Always have to be very efficient, this specification isn’t hard for IGBTs to achieve, they are innately strong in conduction losses. |

Table 30 - Initial search specifications for IGBT selection.

The following table compares IGBTs for suitability to our design.

|  |  |  |  |  |
| --- | --- | --- | --- | --- |
| **Comparison of Possible Inverter IGBTS** | | | | |
|  | **IGW30N65L5** | **FGA180N33AT** | **NGTB50N60FWG** | **FGA5065ADF** |
|  | 650V | 330V | 600V | 650V |
|  | 85A @ 25°C | 180A @ 25°C | 100A @ 25°C | 100A @ 25°C |
|  | 120A | 450A | 200A | 150A |
| **()** | 42ns | 107ns | 160ns | 52.4ns |
| **()** | 359ns | 288ns | 390ns | 73.6ns |
|  | 168nC | 169nC | 310nC | 72.2nC |
| **Gate Test Conditions** | 25°C, 400VCE, 30A, 15VGE, RG = 10Ω | 25°C, 200VCE, 40A, 15VGE, RG = 5Ω | 25°C, 400VCE, 50A, 15VGE, RG = 10Ω | 25°C, 400VCE, 50A, 15VGE, RG = 6Ω |
| **@ (?)Amp** | 1.05V @ 30A | 1.03V @ 40A |  | 1.7V @ 50A |
| **()**  **@25°C / @150°C** | 1820μJ / 2860 μJ | Not provided | 2300μJ / 3400μJ | 1659μJ / 2378μJ  **@ 175°C** |
| **Cost @ Distributor** | $3.83 @ Mouser | $4.33 @ Mouser | $6.31 @ Digi Key | $4.79 @ Mouser |

Table 31 - Comparison of eligible IGBTs for the inverter.

One thing that has become clear in component selection research is that predicting the switching losses of our IGBTs is hard unless that statistic is provided on the datasheet. Because the standard relation of does not include the diode recovery losses or the tail losses associated with the recombination of the minority carriers during the turn off cycle. The IGW30N65L5, FGA5065ADF, and NGTB50N60FWG datasheets all provide data on the switching losses of the devices at 25°C, the FGA180N33AT does not. This specification provides a measurement for the turn on and turn off energy at specified testing conditions. Using these values we can estimate the switching losses of the IGBTs we have in consideration to help judge an efficient operating frequency for the PWM as well as consider the options of lower switching energy IGBTs which allow high frequency switching at high efficiencies with 100A current capability in bursts, if there are any which do.

Using Equation 62 and an Excel® spreadsheet, we calculated the power loss of our various eligible switches. This data is presented in Table 32.

Equation 62 - Switching power losses.

|  |  |  |  |  |
| --- | --- | --- | --- | --- |
| **Comparison of Switching Losses in IGBTs** | | | | |
| **Device 🡪** | | **IGW30N…** | **NGTB50N…** | **FGA5065…** |
| **T = 25°C** |  | 36.4 (W) | 46.0 (W) | 33.2 (W) |
|  | 27.3 | 34.5 | 24.9 |
|  | 18.2 | 23.0 | 16.6 |
|  | 9.10 | 11.5 | 8.30 |
|  | 4.55 | 5.75 | 4.15 |
| **T = 150°C\*** |  | 57.2 | 68.0 | 47.6 |
|  | 42.9 | 51.0 | 35.7 |
|  | 28.6 | 34.0 | 23.8 |
|  | 14.3 | 17.0 | 11.9 |
|  | 7.15 | 8.50 | 5.95 |

Table 32 - Comparison of the switching losses of eligible IGBTs.

\*Note: 175°C for the FGA5065ADF

Because these losses are constant with respect to duty cycle they will be insignificant at high loads as the power output will be large, but at low loads the power lost in switching the IGBTs will be very significant. For other PWM applications this issue usually warrants pulse skipping modes where the switching frequency is effectively reduced while the load is very small, increasing efficiency immensely. The possibility of changing the switching algorithm to allow fewer switching cycles under light loads needs to be considered in the prototyping phase to see if we can use pulse – skipping modes to reduce the switching cycles. These considerations, as well as the final switch selection for the inverter will be made through prototyping and iterative design.

Considering the devices in Table 31, the conduction performance of the FGA180N33AT is extremely good, but its voltage rating is slightly under specification. This eliminates it from consideration assuming we are using a 340V DC bus. The remaining switches all have relatively similar loss figures from switching. As a result it seems that switching losses for these levels of current are significant across all devices, and therefore a lower PWM frequency from the initially desired 20 kHz is recommended to the rest of design team. A switching frequency below 10kHz, even as low as 5kHz will provide sufficient resolution to synthesize sinusoidal outputs of under a couple hundred hertz.

Because of the switches low cost this design will be one of iteration and testing. We will buy a variety of eligible IGBTs with various operating characteristics (note the switching times for the FGA5065) and see which yields the most efficient drive. Once a switch is selected the IC which drives it can be finalized but for initial testing it seems clear that having a high current specified drive is strongly desirable to handle any gate charge we may present it.

### IGBT Driver IC Specification and Selection

Any high power switching device requires a circuit to drive it on and off. In the case of the IGBT this means charging and discharging the gate capacitor. Similar to a MOSFET. As discussed in: Selecting a Power Switch Type for the Inverter,the IGBT is very similar to the MOSFET in its gate structure, both have a MOS capacitor which needs to be charged, then discharged at precise times, quickly. That is where the IGBT driver IC comes in, this IC takes a logic level signal from the MCU telling it which gates of which IGBTs to drive high and low, and uses internal power switching arrays to convert the logic level signal, to a power signal in the form of drive current on the gate(s) of the desired switches.

As discussed in the preceding section; the drive IC is selected based on the drive requirements of the final switch. Because the final switch will be selected through iterative testing under various conditions, we must specify a drive IC which can handle most switches we may choose.

Using the specifications in Table 30, the Infineon IGW30N65L5 TrenchstopTM IGBT. The device has typical values of 1.05VCE drop for a collector current of 30A and it is very stable over temperature ranges up to 150°C. The IGW30N65L5 key specifications are described in Table 33.

|  |  |  |
| --- | --- | --- |
| Selection Criteria for the IGBT Driver | | |
| **Parameter** | **Value(s)** | **Justification** |
| Max Collector – Emitter Voltage () | Greater than 350V | Nominal DC link voltage |
| Drive current (Source/sink) | Greater than 3A | Most gate resistor values on the IGBT datasheets were between 5 and 15 ohms at 15V drive, suggesting that they had initial current of 15/5=3A max at turn on. |
| Input Voltage | 15V | This is our specified rail for gate driving |
| Number of drivers | 2 | Allows for 3 drives to handle 6 IGBTs. One chip does one half – bridge. |
| Channel Type | Independent | We want to be able to control the IGBTs independently of each other using SVPWM. |
| Logic Level | 3V max | We use 3V logic on our microcontrollers so this would be optimal. |

Table 33 - Selection criteria for the IGBT driver IC.

Based on the parameters in Table 33, four eligible ICs were selected for consideration which all meet the specifications. The selection of IGBT drivers we will choose from is given in Table 34 below.

|  |  |  |  |  |
| --- | --- | --- | --- | --- |
| **Eligible Options for the IGBT Driver IC** | | | | |
|  | **IRS2186S** | **FAN7190M\_F085** | **UCC27714D** | **DGD2190** |
| **Drive Current Source** | 4A | 4.5A | 4A | 4.5A |
| **Drive Current Sink** | 4A | 4.5A | 4A | 4.5A |
| **Supply Voltage Range** | 10V-20V | 10V-22V | 10V-18V | 10V-20V |
| **Gate Voltage** | 5V-20V | 9.7V-22.3V | 0.3V-17.7V | -0.3V-19.7V |
|  | 600V | 600V | 600V | 600V |
| **Logic Level (low/high)** | 0.8V/2.5V | 1.2V/2.5V | 1.2V/2.7V | 0.8V/2.5V |
| **propagation delay** | 170ns / 170ns | 140ns/140ns | 90ns/90ns | 140ns/140ns |
| **Inverting Input?** | Both options | No | No | No |
| **Device Package** | 8 SOIC | 8 SOP | 14 SOIC | 8 SOIC |
| **Price @ Distributer** | $2.85 @ Mouser | $1.62 @ Digi key | $5.00 @ Digi key | $1.73 from Mouser |

Table 34 - IGBTs under consideration meeting specifications in Table 33.

The Final decision will be based both on the selection of the IGBT under test. For initial testing, however, any of these devices should work because their drive current ratings are all very high.

### Output Filter Design

The output filter of the power inverter soothes the carrier (switching) frequency harmonics and passes the low frequency modulated power signal into the motor. The filter is a simple second order low-pass LC network with sufficiently specified cutoff frequency and ripple current rating on the capacitor. The filter is best implemented as a Butterworth response to maximize the flatness in the passband. In order to achieve a Butterworth response the following criteria formulas can be applied to solve for L and C values based on load resistance:

Equation 63 - Calculation of L and C values for a Butterworth filter.

To adequately preserve the drive signal it would be wise to have the highest expected output frequency expected be well below the cutoff frequency. The frequency of PWM should be significantly higher than the cutoff frequency of the filter. Given that it would be unlikely to see an output frequency of over 200Hz, and that our switching will be at least 5 kHz, a cutoff frequency of between 750Hz and 1000Hz seems appropriate. The attenuation of the first harmonic of the PWM will be at least 16dB or 6.3x attenuation.

The final calculation of values will be done with motor parameters for the final motor selected. will be taken as the winding resistance of the stator windings. Once this measurement is taken on the motor we are using, the values will be specified for L and C. The filter will be implemented in the basic second order low pass LC filter topology.

### Over-Current Protection

Over current protection is mainly a consideration for protecting the power switches from overheating their junction and damaging the device. The motor windings can handle extremely high currents for short times. Modern power switching devices have a built in limit of their current based on junction temperature. If the current spikes, the junction temperature will rise almost immediately. A built in temperature sensing device in the IC can sense this rise and limit the conduction of the junction to protect it from self – destruction. Because these devices are built in, as long as we choose IGBTs which are rated for the transient currents that will occur during startup of the motor, they should be fine managing their own current limit if the situation should ever arise in testing.

### Test Plan: Power Inverter

The power inverter testing will begin with low power functionality confirmation. The inverter will be charged on the DC rails with a decoupled DC power supply and a three phase resistive load will be attached such that the DC source voltage will create a load amperage of below 5A. The Inverter will be driven with the SVPWM module set to synthesize a square wave output on the three phases. Appropriate current limiting will be used on the power supply initially to ensure safe start up. For this portion of the testing we will be observing the output waveforms of the inverter to confirm phase shift between the outputs and apparent operation of the switches. Then the switch rise and fall times will be measured and compared to datasheet values.

The switching on and off of the high and low side switches will be carefully observed for overlap. There cannot be any overlap between the switching of the high and low side (it will short the DC link). We will also observe the conduction characteristics of the switches by measuring the collector – emitter voltage drop during saturation. We really want to see symmetry here. All the phases need to see the same output impedance from the inverter. We will also probe the bootstrap capacitor input of the drive similarly to the rectifier test plan. We must ensure that the IC is getting fully sufficient voltage for proper operation at all times. If the functionality looks nominal; all currents, voltages, and switching operations are conducted without issue, we can continue to sinusoidal synthesis.

To synthesize a sinusoid we will use our SVPWM algorithm to create the switching commands for the IGBT drives. We will use a roughly designed LC low-pass filter on the output to smooth the switching ripple and observe the baseband signal underneath. We hope to see that the desired frequency is achieved as well as symmetrical amplitude on the positive and negative cycles of the waveform. The frequency of the output will be varied to simulate all possible output signal frequencies (1Hz – 1000Hz), we will observe for proper switching and synthesis. During this and the previous tests, special attention needs to be paid to the latency of the switching system, as the microcontroller, drive IC, and IGBT all have latencies which come into play with the output pulses. For consideration of the ADC sampling for the output phase voltage we need to be sure we sample it at a specific time during the PWM cycle, and this timing must be consistent. Therefore we must document and measure the total input to output latency on the switching system. This will be added as a constant offset in the code to assure proper measurement timing.

High power testing of the inverter will require the DC link to be operational as well as all accompanying sensor devices. Therefore full power testing will not take place until the full other power system supply elements (Synchronous rectifier, LV transformer and rectifier, LVDC Rail regulators, DC link) have fully functional prototypes which have been individually load tested to achieve their specified output rating.

## Sensor System – Temperature, Voltage, and Current

The sensor requirements, as specified in the appropriate Design Specifications section, are of three types; rotational speed, temperature, voltage, and current. Sensors are defined based on their function to the system being critical or auxiliary. The critical sensors are required for operation of the drive, auxiliary sensors are used for protection and start of protections only. Sensors are also specified based on accuracy. All critical sensors require 1% accuracy minimum, as a 1% accuracy in speed control is specified in the Design Specifications.

|  |  |  |  |  |
| --- | --- | --- | --- | --- |
| Sensor Type, Monitoring, and Protection Requirements | | | | |
| **Sensor(s)** | **Critical / Auxiliary** | **Multiplex?** | **MCU assignment** | **Accuracy** |
| Motor - Rotor Speed | Critical | NO | Piccolo | 1% |
| Motor - Stator Temp. | Auxiliary | YES | MSP430 | 5% |
| Inverter - Heatsink Temp. | Auxiliary | YES | MSP430 | 5% |
| Rectifier - Heatsink Temp. | Auxiliary | YES | MSP430 | 5% |
| Motor - Phase Voltages | Critical | NO | Piccolo | 1% |
| Motor - Phase Currents | Critical | NO | Piccolo | 1% |
| DC Link – Voltage | Auxiliary | YES | MSP430 | 5% |
| AC Input – Voltage | Auxiliary | YES | MSP430 | 5% |
| Low voltage DC bus - Voltage | Auxiliary | YES | MSP430 | 5% |

Table 35 - Overview of sensor inputs from the system per Design Specifications.

Given that the Piccolo controller will manage the space vector control algorithm and data processing, it will be in charge of all critical sensor data which is needed for vector control to function as intended. All other (auxiliary) sensor data processing will be handled by the MSP430 to better distribute the processing load between the processors.

All of the auxiliary measurements will be multiplexed at the MSP430 to save GPIO pins. The MSP430 will cycle through the MUX and probe each measurement in sequence. Because maintaining “real – time” is actually a quite generous sampling frequency allotment this can be done. The MSP430 will also control the LCD display so doing the sensor calculations which will be displayed on the LCD on the MSP430 makes sense.

For critical sensor data, each sensor had a dedicated GPIO pin fed into the Piccolo’s ADC which can be sampled at a high frequency for the space vector PWM control logic. Each sensor will need good accuracy, and the 12 bit ADC of the Piccolo will be instrumental in achieving the measurement accuracy we specify. The required accuracy of the critical sensors also plays a role in the sensor selection, as the transducer must be low latency to assure accurate timing of the measurement.

### Temperature Sensors

For data that is sampled from the motor appropriate transducers will be required.  A sensor to read data from the motor will interface with the encoder to provide motor mechanical speed data to the space vector control block of the variable frequency drive. As a part of the digital signal processing system there will need to be appropriate algorithms in place to perform anti-aliasing of the data (low pass filtering to band limit continuous time signals like motor rpm) appropriate sampling techniques, and analog-to-digital conversion.

The microcontrollers will be gathering various inputs from parts of the variable frequency drive. The microcontroller (likely MSP430) in charge of the low priority sensors will multiplex the sensor inputs to more efficiently use the microcontroller pins. This will sacrifice measurement speed immensely but with measurements of temperature which do not have very high rates of change with time that is acceptable. Despite the latency in multiplexing the signals, the microcontroller should still be able to sample the data at a rate which is completely acceptable for monitoring and safety purposes. In the case of the critical sensors (motor operational data of phase voltages and currents, the DC link voltage, and the motor mechanical speed), it is imperative that we not only dedicate microcontroller pins on the ADC solely to measuring these values but also that the microcontroller which is issuing the PWM control signal to the inverter is the one to sample it. That microcontroller will know when a pulse is sent and the current state of the power switches (assuming a rather constant turn on/off time for the switches). This will enable optimal timing for sampling motor phase voltage and current to best characterize the current state operations of the motor under control.

#### Thermocouple devices

Thermocouples are a common way of obtaining temperature measurements for a wide variety of industrial and general-purpose applications.  This project being a commercial application, is well suited to the advantages of using thermocouples to obtain the temperature of the motor during its operation.  Thermocouples are a popular temperature measurement device because of their relatively low price, wide temperature range, lack of required excitation, long-term stability, and proficiency with contact measurements.  Though robust, achieving extremely high accuracy with a thermocouple can be more difficult than a resistance temperature detector, but the low cost and versatility of this device make it desirable for use in this design.  In addition, as opposed to thermistors and RTDs, the use of thermocouples simplifies application circuitry because they require no excitation. Thermocouples do require a stable voltage reference and some form of cold junction compensation which needs to be performed to obtain proper measurements from the MSP430.  The TI ADS1118 is an ideal option for thermocouple measurement because of its integration of an internal voltage reference, multiplexer, and temperature sensor [14].

#### The 16-bit ADS1118

The ADS1118 is a precision, low power, 16-bit ADC that provides all of the necessary features required for temperature measurement of the motor during operation.  It contains a programmable gain amplifier, voltage reference, oscillator, and high-accuracy temperature sensor.  Its power supply range falls within 2-5.5V, hence it can be powered from out 3.3V analog supply rail.  Performing conversions at data rates up to 860 samples per second and with input ranges of +/-256mV to +/-6.144V allows for both large and small signals to be captured with pretty high resolution.  The input multiplexer allows for the measurement of two differential or four single-ended inputs.  In our case a two-channel thermocouple will be implemented [14].

#### The Seebeck Effect and its Importance to Thermocouples

A thermocouple consists of two wires made from differing conductors, usually alloys, which are soldered or welded together at one end.  The composition of these conductors varies widely, and is dependent on the required temperature range, accuracy, lifespan and environment that it is being measured in. All thermocouple types are common in their use of the thermoelectric or Seebeck effect.  In the case that a conductor experiences a temperature gradient from one end of the conductor to the other, a voltage potential develops.   This voltage appears because of the free electrons within the conductor that diffuse at different rates, depending on the temperature.  Electrons with more copious energy on the hot side of the conductor diffuse more rapidly than the lower energy electrons on the cold side of the thermocouple.  The resulting effect is that a buildup of charge occurs at one end of the conductor to create a voltage potential from the hot and cold ends.  The following figures illustrate the concept/structure of the thermocouple and the Seebeck Effect [14]:

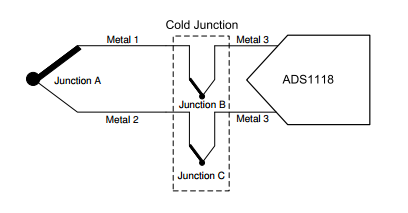


Figure 29 - Thermocouple junction diagram- Texas Instruments.

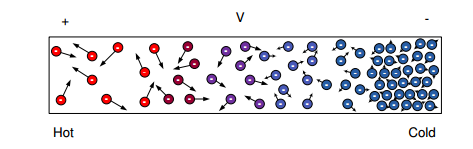


Figure 30 - Illustration of the Seebeck Effect - Texas Instruments.

Metals of different composition exhibit the Seebeck effect at varying levels of intensity.  When two different types of metals are paired together and joined at a certain point like junction A in the thermocouple junction diagram the differences in voltage potentials on the opposite end of the short, comprised of junctions B and C are proportional to the temperature gradient formed from either end of the conductors.  This effect serves to illustrate that the thermocouples do not actually measure an absolute temperature; they only measure the temperature difference between two points, known as the hot and cold junctions of the thermocouple.  Henceforth, in order to determine the temperature at either end of the thermocouple, the exact temperature of the opposite end must be known.  A true ice point reference is unnecessary in this design, as this would be impractical (we don't want to use an ice bath as part of this design).  Instead, the temperature of the junctions B and C of the thermocouple are continuously monitored and utilized as a point of reference to calculate the temperature at junction A at the other end of the device.  These endpoints of the device are referred to as junctions because they connect to a terminal block that transitions from the thermocouple alloys into the traces of a PCB, which are in most cases copper.  This movement back to copper is what creates the cold junctions B and C.  Junctions B and C can be treated as a single reference junction, provided that they are each held at the same temperature.  Once this reference temperature is known, the absolute temperature at junction A can be calculated.  Measuring that temperature at B and C then using it to calculate a second temperature at A is known as cold junction compensation [14].

#### Achieving accurate thermocouple measurements

It is important that the following conditions are met to achieve accurate thermocouple measurements.  First, the junctions B and C must be kept at identical temperatures.  This can be achieved by keeping these two junctions in very close proximity to each other and away from any sources of heat that may exist on the PCB.  Isothermal blocks are a good way of keeping the junctions at the same temperature.  A large mass of metal offers a very reliable way of achieving isothermal stabilization.  Second, the isothermal temperature of junctions B and C must be accurately obtained.  The closer that a temperature sensor like a diode, RTD or thermistor can be placed to the isothermal block, the better.  This stipulation applies to the ADS1118.  With a mere 500μW of power consumption, the effects of the ADS1118’s self-heating is negligible.  The ADS1118 already offers excellent uncalibrated precision, with 0.5°C in maximum error. The concern of air currents, however, must be taken into consideration.  Air currents can serve to diminish the accuracy of the cold junction compensation measurement.  In order to garner best performance, the cold junction should be kept in an enclosure so that air currents may be kept to a minimum near the cold junction.  In our case, a mechanical cover on the ADS1118 and connector block should do the trick.  In addition, the orientation of the PCB can impact the accuracy of the cold junction compensation.  This is due to the fact that if heat-generating elements are physically below the cold junction, inaccuracies can become significant as heat from those elements rises [14].

The schematic of Figure # lays out the connections for an independent, two-channel thermocouple system.  The circuit contains a low pass, anti-aliasing filter, mid-point bias, and open detection.  Although the digital filter of the ADS1118 strongly attenuates high-frequency components of noise, providing a first-order passive RC filter further improves this filtering performance.  The differential RC filter formed by the 500-ohm resistors and the one microfarad capacitor offers a cutoff frequency of about 320 Hz.  Additional filtering can be achieved by increasing the differential capacitor and resistance values, however for the purposes of this design it won’t be necessary to increase this differential.  In addition, by avoiding increasing the filter resistance beyond one kilo-ohm will ensure that the linearity and gain of the ADS1118 is maintained and not corrupted by the ADC input impedance.  Because of the high sampling rates supported by the ADS1118, post digital filtering with the MSP430 can alleviate the requirements of the analog filter, and also offers the flexibility to implement filter notches at 50Hz or 60Hz.  Two 0.1μF capacitors are also added to offer attenuation of high frequency, common mode noise components.

#### Circuit Component Considerations

Mismatches in the common-mode capacitors cause differential noise, an undesirable feature in this circuit implementation.  Therefore, the differential capacitor will be at least an order of magnitude larger than the common-mode capacitors.  In order to achieve electromagnetic interference immunity, it is noted that placing large capacitors in the signal path and supply are not very good at blocking out high noise frequency components.  Using small capacitors on the order of 10nF or smaller with low equivalent series resistance and low dielectric absorption in parallel with another higher capacitance capacitor on sensitive supply and signal paths can offer significant improvements to EMI immunity.  EMI is a concern for this project because of the large inductive load we are dealing with.  Therefore, additional EMI immunity can be achieved by incorporating a ferrite bead/common-mode choke on the inputs.  The ADC input protection circuit with clamping Schottky diodes can be implemented before the input filter to protect against ESD.

The two Mega-ohm resistors serve two purposes.  First, these components offer a common-mode bias near mid supply.  By connecting only one of the inputs to a common point, performance is decreased by converting common-mode noise into differential signal noise that is not strongly attenuated.  Secondly, these resistors offer a weak pull-up and pull-down for sensor open detection.  If a sensor is disconnected for whatever reason, the inputs to the ADC will extend to the analog supply and ground to yield a full-scale readout that indicates a sensor disconnection.  For unusually long thermocouples, these 1MΩ resistors might impact the measurement accuracy.  Increasing the resistance of serves to alleviate these effects.  On the other hand, two 1MΩ resistors connected as a voltage divider to one of the inputs of the ADS1118 can maintain the midpoint bias without affecting the measurement results.  This methodology, however, sacrifices open detection and a small amount of common-mode noise rejection [14].

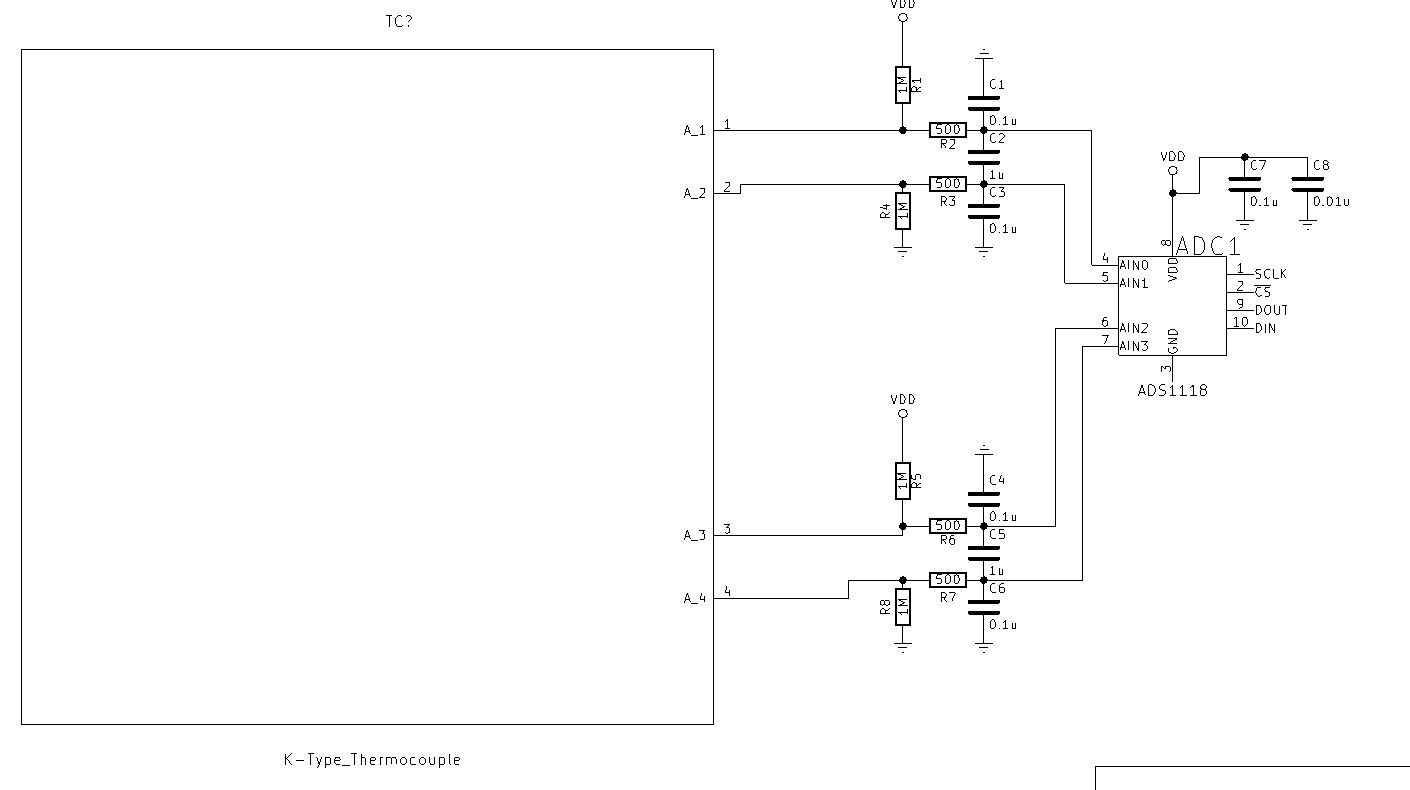


Figure 31 - Two-Channel Thermocouple system – Designed in KiCad

#### Software Discussion on Temperature Measurement

The actual calculation to achieve cold junction temperature compensation is relatively simple to achieve and can be done in several ways.  A typical way to interleave readings between the thermocouple inputs and the temperature sensor is to acquire the on-chip temperature result for each ADS voltage measured.  If the cold junction is in a sufficiently stable environment, the more periodic cold junction temperature measurements might be sufficient.  These operations that are performed will, in turn, yield two results for every thermocouple measurement and cold junction temperature cycle: the thermocouple voltage or Vtc, and the on-chip temperature or Tcjc.  In order to account for the cold junction, the temperature sensor inside the ADS1118 needs to first be converted to a voltage that is proportional to the thermocouple currently being used, to yield Vcjc. This process is, in general, accomplished by performing a reverse lookup on the table used for the thermocouple voltage-to-temperature conversion.  Adding the two voltages then yields the thermocouple-compensated voltage Vactual.  The following equation mathematically defines Vactual:

Equation 64 - Actual value from the reverse lookup table operation.

Then, Vactual is converted back to a temperature reading using the same lookup table as before and the actual temperature is achieved.  A block diagram illustrating this process is given below:

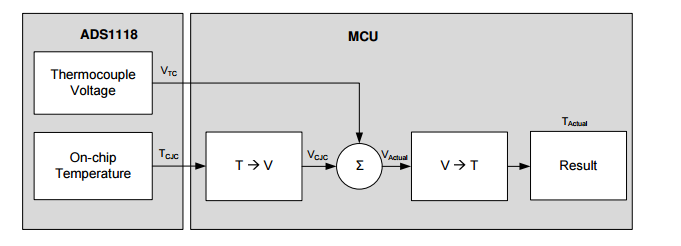


Figure 32 - Software flow block diagram- Reprinted with permission from Texas Instruments

#### Thermocouple Measurement Conversion

There are two ways to perform the conversion between thermocouple to voltage and voltage to temperature.  First, the coefficients of the lookup table can be programmed into the MSP430 from the high-order polynomial, and following, the calculation may be performed on each reading.  Although this method offers the smallest introduced error during the conversion, it is extremely processor-intensive and therefore is not practical for most applications.  The other and more common way to perform the conversion is through the use of a lookup table.  Thermocouple manufacturers will often times provide a lookup table with their respective thermocouple devices that offer excellent accuracy for linearization of a specific type of thermocouple.

## Mechanical Assembly

### Transmission/Drivetrain

The transmission converts power at one speed and torque, to (ideally) equivalent power at another speed and torque. Similar to the AC power transformer converting power of one voltage and current to another voltage and current, but power is unchanged as well (ideally). In reality losses in both analogous systems result in heat generation, wear on components, and loss of efficiency.

Because of the above limitations, the VFD Go-kart will be driven with as simple a transmission as we can build, as simplicity of design will result in less losses, higher reliability, lower cost, etc. Normally a designer of an internal combustion engine driven vehicle would not be able to make this bold assertion; the limitations of the internal combustion engine require a multi-speed transmission for practical use. However, as a result of the very core goal of this project; the design of a variable frequency drive (VFD), our motor will be able to have a greatly extended dynamic range relative to an internal combustion engine. This property allows the group to select a simple chain drive, where the only conversion will be the relationship of the tooth count on the sprockets. Equation **X** illustrates the relationship between torque output (Tout), input torque (Tin) and the ratio of teeth on the sprockets (N2/N1)

Equation 65- Output torque

## Microcontroller Selection

### Microcontroller Basics

In order to control the speed of the motor, we are going to need a computer that will be able to not only monitor the input that our system is receiving, as well as the output that our motor is producing, but also correct any error that is found between the input and the output of the motor. For an application as sensitive as motor control, we will need a computer that will not only compute on a fast clock, but also has access to pulse-width modulation channels, analog-to-digital converters, and easily accessible memory. For all of these different reasons, we have chosen to use a microcontroller to fulfill these different needs.

A microcontroller is a miniature computer on a single integrated circuit. The circuit includes a processor core, memory, and input and output peripherals that can be programmed to the user’s desire. Microcontrollers are primarily used in most embedded applications and systems, varying from home appliances to power tools to automotive control.

While microcontrollers tend to seem underwhelming by appearance, their practical and economic capabilities more than make up for their size. Their ability to provide low-power operations while managing basic input and output allows the entire computerization of many products and services used every day, and at very little cost. Since most microcontrollers only need to manage very basic tasks, they can perform their job efficiently with inexpensive hardware; though there are much more high-end microcontrollers for very precise and sensitive tasks.

### Microcontroller Selection

In choosing the microcontrollers we wish to use for our project, we must consider what tasks they will be performing. For example, we will be using a LCD display to show the user information of our system. This task is very ideal for a microcontroller, as its computational abilities will be quick enough to display and update data to the LCD screen at real-time.

Powering and controlling an LCD display is a task that is not typically memory intensive, so memory capacity, while important, is not of major circumstance. One very important consideration as well is that powering and updating an LCD display is not a very power-intensive function. For these reasons, we will be using the TI MSP430x5xx line of microcontrollers for this task, as it is specialized in low-power consumption embedded applications. Please see the MSP430F5529 section for more details.

Another thing we will be using a microcontroller for will be monitoring and controlling the motor using Pulse Width Modulation. This is obviously a much more sensitive undertaking than controlling an LCD display, so we are going to require a more powerful chip. This microcontroller is going to be taking in various analog inputs, so it will require accurate analog-to-digital converters. We are also going to need various pulse width modulation channels to control the various switches in the three-phase inverter. We will need much more accurate and precise data, so using a 32-bit architecture would be the most ideal. As we will be doing various computations with all of this data, we will require a very powerful processor. The TI C2000 line of microcontrollers is more than ideal for this task. Please see the TMS320F28027F section for more details.

### The MSP430F5529

The microcontroller that would ideally drive the LCD display would be the MSP430F5529. This microcontroller features a 16-bit RISC architecture, unified clock support, and four separate timers with several capture and compare registers per timer. It also features a clock frequency of up to 25 MHz, 128 kB of flash memory, and 10 kB of RAM. The pin diagram for the MSP430F5529 can be found in Figure 33.

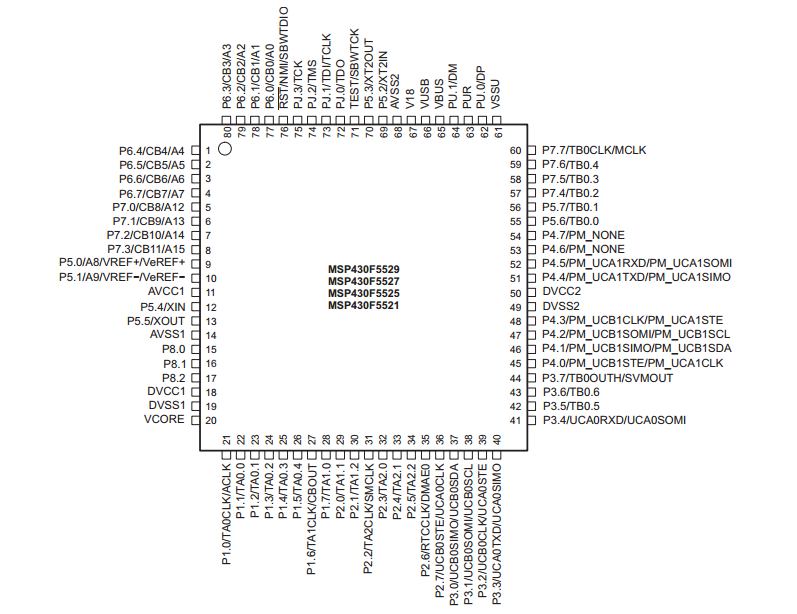


Figure 33 - The pin diagram of the MSP430F5529. This will be used as a reference for all KiCad schematics and footprint layouts.

In terms of power consumption, the MSP430F5529 features various power modes. The first one is the “shutdown mode.” This mode turns the current draw of the microcontroller to a mere 0.18 µA at 3.0 V. This power mode will only be functional while the system is turned completely shut down, so there will most likely not be any use of this power mode. The next power mode is the “off mode” power setting. This power setting is similar to the “shutdown mode” except for the larger current draw (1.1 µA at 3.0 V) and that there is full RAM retention. This will most likely be more useful if the system will be turned off, but will also have important data being handled by the microcontroller. Considering that the data that we will be handling will only be used during real-time when the system is running, the volatile properties of the RAM will not matter, meaning that this power mode most likely will also not be implemented.

The next power mode we will consider is the “standby mode,” however there are two different versions of this mode. The difference between the two only varies by the timers that are being used by the microcontroller. The less powerful of the two takes advantage of the low-power oscillator (or VLO) and the general-purpose counter. It also has a current draw of 1.4 µA at 3.0 V. These two peripherals are very efficient in terms of power consumption, but are also less accurate than a clock with a crystal. This is where the other power mode comes in, implementing the real-time clock with a 32 kHz crystal (or RTC.) This power mode uses a much more precise clock, but at the cost of a larger current draw than the other clock (1.9 µA at 2.2 V or 2.1 µA at 3.0 V.)

This power mode overall would be much more useful than the other previously covered power modes. The reason this is that in order to update the LCD display, we are going to be using interrupts frequently, which means we will be implementing a clock signal. We will need the clock peripherals available to create and handle the interrupts, which will include monitoring the value of various analog signals, quantizing those signals into digital values, storing those values into memory, and then updating the LCD display based on the value in specific memory addresses. However, since this power mode only allows the availability of a single clock system, and our system may require additional clocks, this mode may not be the most ideal to implement for our drive.

The last power mode that we will look at is the “active mode” power settings. This mode will be the one that we will take advantage of the most, as it activates all clock systems, the processor will be operating at its full potential, and any code that is stored in flash memory or the RAM will be executed. Obviously the last point will be the most important feature in this power mode, as all of the computations being done to the analog input signals will be executed in code that is stored in the memory of the system. That being said, the majority, if not all, of the functionality of the MSP430 and the LCD display will be performed in the MSP430’s “active mode.”

### The TMS320F28027F

The TMS320F2802xx (also referred to as “Piccolo”) family of microcontrollers would be aptly suited to handle the task of space – vector control and space vector pulse width modulation. While the MSP430 focused more on low-power functionality, the Piccolo microcontrollers are strongly suited in CPU intensive applications. It features a maximum 60 MHz clock (translating to a 16.67-ns cycle time,) a 32-bit TMS320C28x CPU, 64 KB of flash memory, and 12 KB of RAM. The F2802xF models specifically also feature TI’s InstaSPIN-FOC Technology, which will be helpful in interfacing with our system. The pin layout for the TMS320F28027F can be found in

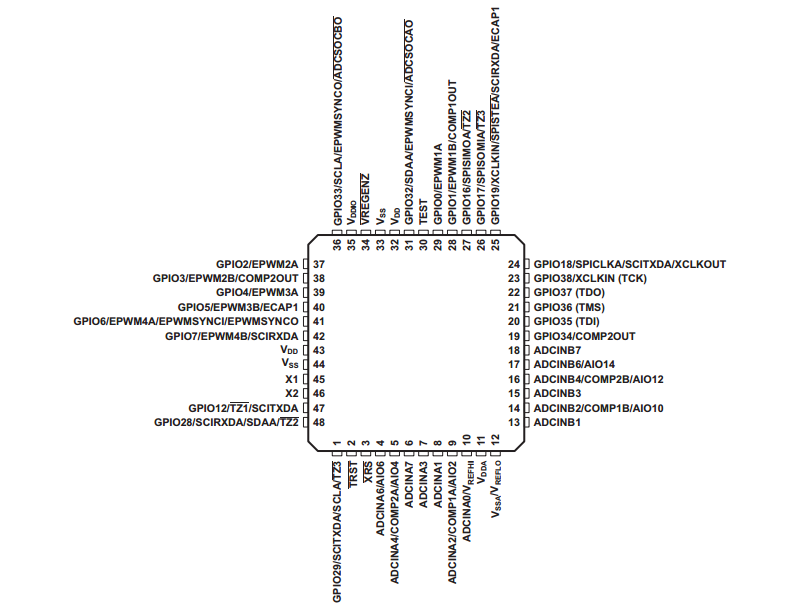
**

Figure 34 - The TMS320F28027F Pin diagram. This will be used for the KiCad schematic and PCB footprint.

The Piccolo microcontroller features various control peripherals that make it essential to controlling our system. It includes high-resolution PWM (HRPWM) module with four enhanced pulse width modulators (ePWM,) along with independent 16-bit timers for each ePWM module. This will make it excellent for controlling the individual stator coils. It also contains a seven-channel analog-to-digital converter with 12-bit resolution. This will aid in providing accurate feedback from the system, as well as receiving precise input from a user.

## Microcontroller Inputs

### Sensor – Microcontroller – Display Interface

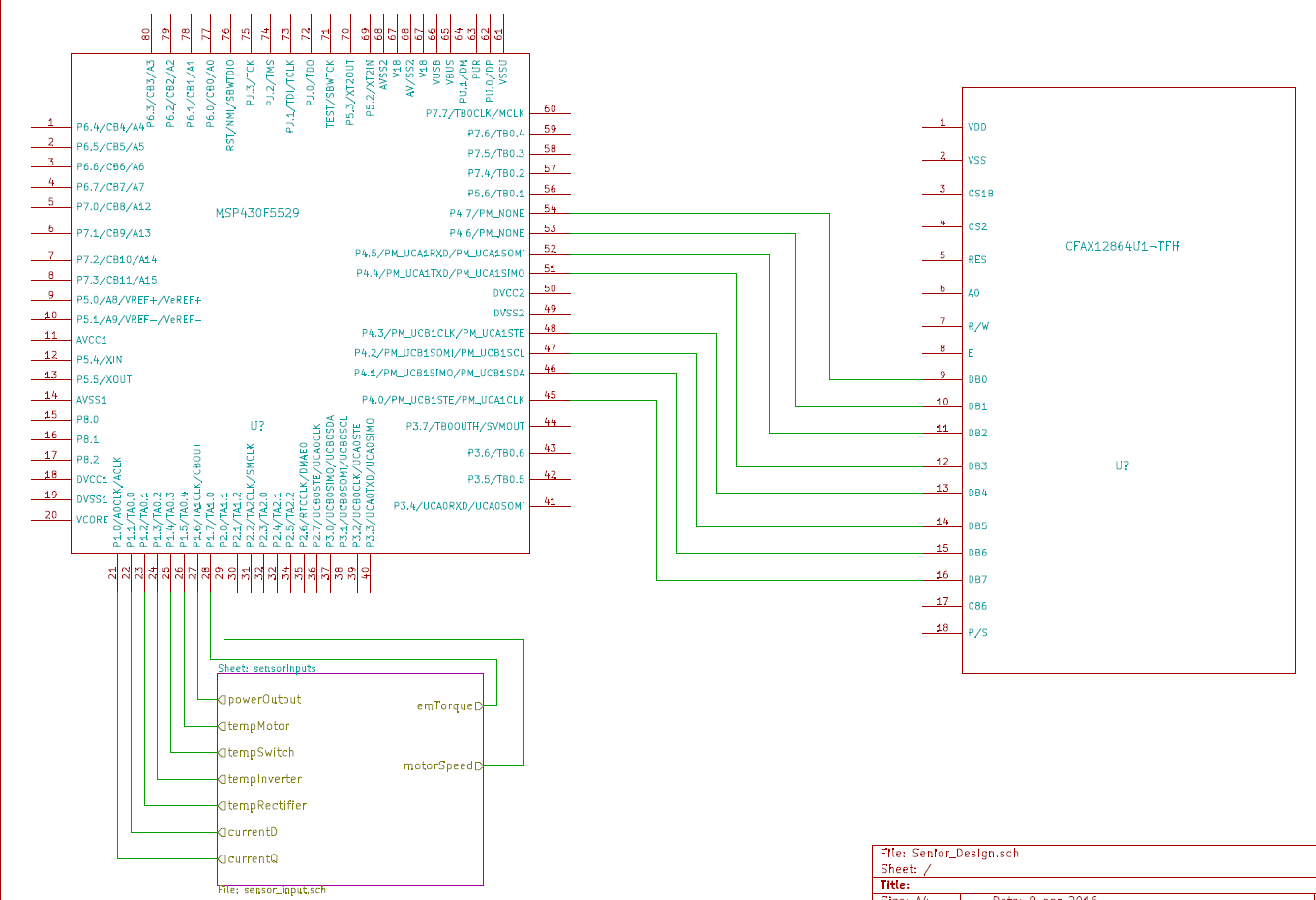


Figure 35 - The Interface required between sensor inputs, the MSP430, and the LCD display.

### Constraints on Input Signals

There is a range of inputs that microcontrollers are designed to handle. If they receive an input or attempt to generate an output that is outside of this range, then they could malfunction. The following table, which is taken from the microcontroller datasheet, describes the absolute minimum and maximum voltages, currents, and temperatures that the various peripherals of the TMS320F28027F are designed to handle:

|  |  |  |
| --- | --- | --- |
| **Absolute Maximum Ratings for TMS320F28027F** | | |
| Supply voltage range, VDDIO (I/O and Flash) | With respect to VSS | -0.3 V to 4.6 V |
| Supply voltage range, VDD | With respect to VSS | -0.3 V to 2.5 V |
| Analog voltage range, VDDA | With respect to VSSA | -0.3 V to 4.6 V |
| Input voltage range, VIN (3.3 V) | -- | -0.3 V to 4.6 V |
| Output voltage range, VO | -- | -0.3 V to 4.6 V |
| Input clamp current, IIK (VIN < 0 or VIN > VDDIO) | -- | ±20 mA |
| Output clamp current, IOK (VO < 0 or VO > VDDIO) | -- | ±20 mA |
| Junction temperature range, TJ | -- | -40°C to 150°C |
| Storage temperature range, Tstg | -- | -65°C to 150°C |

Table 36 - The TMS320F28027F Absolute Maximum Values.

The following table, which is taken from the microcontroller datasheet, describes the absolute minimum and maximum voltages, currents, and temperatures that the various peripherals of the MSP430F5529 are designed to handle:

|  |  |  |  |
| --- | --- | --- | --- |
| **Absolute Maximum Ratings for MSP430F5529** | | | |
|  | **Min** | **Max** | **Unit** |
| Voltage applied at VCC to VSS | -0.3 | 4.1 | V |
| Voltage applied to any pin (excluding VCORE, VBUS, V18) | -0.3 | VCC + 0.3 | V |
| Diode current at any device pin | -2 | 2 | mA |
| Maximum operating junction temperature, TJ |  | 95 | °C |
| Storage temperature range, Tstg | -55 | 150 | °C |

Table 37 - The MSP430F5529 Absolute Maximum Values.

The information that we will want to display will be shown on the CFAX12864U1-TFH LCD module selected for the MSP430. The following table, which is taken directly from the display module datasheet, describes the specific minimum and maximum inputs that the display is designed to handle:

|  |  |  |  |  |  |
| --- | --- | --- | --- | --- | --- |
| **Absolute Maximum Ratings for CFAX12864U1-TFH LCD Module** | | | | | |
| **Item** | **Symbol** | **Min** | **Typ** | **Max** | **Unit** |
| Operating Temperature | TOP | -20 | - | +70 | °C |
| Storage Temperature | TST | -30 | - | +80 | °C |
| Input Voltage | VI | VSS | - | VDD | V |
| Supply Voltage for Logic | VDD - VSS | 1.8 | - | 3.6 | V |
| Supply Voltage for LCD | VOUT - VSS | 6.0 | - | 14.2 | V |

Table 38 - The CFAX12864U1-TFH Absolute Maximum Values

The following table, which is taken directly from the CFAX12864U1-TFH datasheet, describes the specific electrical characteristics of the LCD module:

|  |  |  |  |  |  |  |
| --- | --- | --- | --- | --- | --- | --- |
| **Electrical Characteristics of CFAX12864U1-TFH LCD Module** | | | | | | |
| **Item** | **Symbol** | **Condition** | **Min** | **Typ** | **Max** | **Unit** |
| Supply Voltage for Logic | VDD - VSS | - | 3.0 | 3.3 | 3.6 | V |
| Supply Voltage for LCD | VDD - VOUT | Ta = -20°C  Ta = 25°C  Ta = 70°C | -  -  - | -  8.5  - | -  -  - | V  V  V |
| Input High Voltage | VIH | - | 0.8VDD | - | VDD | V |
| Input Low Voltage | VIL | - | VSS | - | 0.2VDD | V |
| Output High Voltage | VOH | - | 0.8VDD | - | VDD | V |
| Output Low Voltage | VOL | - | VSS | - | 0.2VDD | V |
| Supply Current | IDD | VDD | 0.18 | 0.18 | 0.18 | mA |

Table 39 - The CFAX12864U1-TFH Electrical Characteristics.

### Input Protection Circuitry Design

In general, an electronic system has inputs that are controlled by some end user.  These inputs are read by electronics and acted upon by using outputs.  Input sources include buttons, switches, sensors, relays, and communication devices to name a few.  In certain environments and situations, these input signals can pose a threat to the electronics reading them – especially if those electronics are designed without the thought of protection [15].

The consideration of over-voltage protection and overcurrent protection must be taken in order to adequately prevent damage to the inputs of the MCU.  The analog signals to be captured by the controller will be the phase currents and voltages of the motor.  The motor being a large inductive device, back-EMF spikes are of critical concern to the controllers.  The inputs must be protected at all times i.e. during power up as well as power down states in order to avoid damage to the ADC and GPIO inputs.  The damage can be catastrophic if design precautions are not taken: the ADC/GPIO pin can either be rendered dysfunctional or there will be a performance degradation to the extent that the expected system performance cannot be achieved.  The following methods will be discussed to be applied to the ADC inputs of the TMS320F28027F Piccolo Controller and MSP430F5529 which have the front end of their ADC blocks as the sample and hold input circuitry [16].

Input protection using clamps can be employed if the input is directly connected to the ADC input or the signal conditioning amplifiers of the ADC are operating at voltages greater than the ADC analog supply voltage.   According to Texas Instruments, Most ADC inputs have internal diodes which conduct when the input voltage goes beyond the supply voltage.  These diodes are not designed to carry a substantial amount of current for a longer amount of time.  The diode clamp structure emulates the ADC internal diode structure with external clamp diodes that are capable of higher continuous current protection.  The forward bias voltage characteristics of the clamping diode are important to consider, Schottky diodes are a preferred device due to their low forward-bias voltage characteristics and relatively fast switching speed. Reverse –bias leakage current is also a parameter to consider.  The effects of voltage dependent leakage currents are proportional to circuit impedances and can cause distortions.  In addition, leakage currents will also vary with temperature and must be evaluated over the intended temperature operating range.  Reverse bias capacitance is a voltage dependent junction capacitance that can cause distortion and it must be insignificant in comparison to the circuit component values. The datasheets for both the MSP430 and the piccolo microcontroller specifies that a maximum input voltage of 3.6V is allowed for the ADC, hence the forward bias voltage of the Schottky diodes must be chosen so that, given a high voltage transient in the system, the spike is clamped to within +/- 0.3 volts of the analog supply. The resistor in series with the ADC input serves to further limit the current absorbed by the ADC.  The following figure presents the KiCad layout of this basic input protection system [16]:

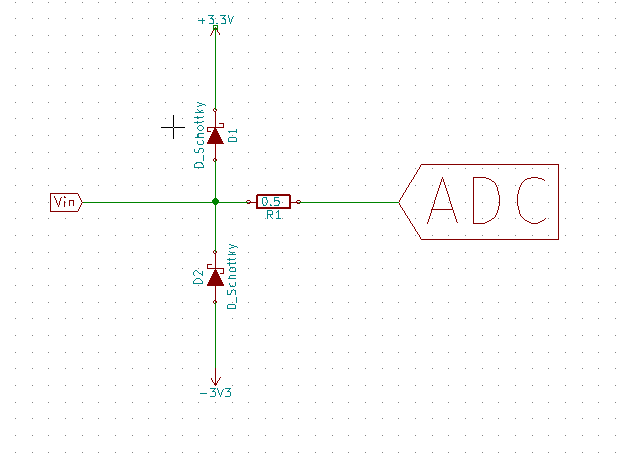


Figure 36-Input Protection Circuitry Modeled in KiCad

As long as the signal voltage does not exceed the analog supply voltage, the diode forward voltage drop is reverse biased and no current will be conducted into the supply rails from the signal.  This is important, as the integrity of the signal is maintained.  However, once the signal voltage exceeds the analog supply voltage the diode is then in forward bias mode and conducts current to the supply rails.  The power rail regulation for the +/- 3.3V supply is stiff and can sink current, therefore clamping at the supply voltage plus the forward bias voltage drop of the diode.

Equation 66-Clamping Voltage as a function of the analog supply and forward bias voltage

The following table presents a comparison of low forward bias voltage surface mount Schottky diodes:

|  |  |  |  |  |
| --- | --- | --- | --- | --- |
| Schottky Diodes Under Consideration | | | | |
| **Manufacturer/Part Number** | **Forward Bias Voltage** | **Reverse Voltage DC** | **Junction Capacitance** | **Recovery Time** |
| **Diodes Incorporated: DFLS130DITR** | 310 mV@1A | 30V | 76pF@10V, 1MHz | <=500ns |
| **STMicroelectronics:**  **BAT30F4** | 310 mV@10mA | 30V | 10pF@1V, 1MHz | <=500 ns |
| **TOSHIBA: CUS520** | 280 mV@10mA | 30V | 17pF@1MHz | <=500 ns |

Table 40-Comparison of Schottky Diodes to be used for input clamp

All three diodes in question have equal reverse voltage parameters and recovery times.  For the purposes of safety and reliability of the ADC: the Toshiba CU520 has the most desirable forward bias voltage as it will maintain the input signal below the maximum 3.6V.

By referring to the section on “constraints on input signals” we can obtain the maximum ratings of both the piccolo controller and MSP430.   The piccolo controller includes input clamping for current up to +/- 20 mA so over-current protection is already taken into account.  Over-voltage protection, however, is not provided internally by the controller.  The goal of input protection on the ADC is to guarantee that the ADC input voltage never exceeds the supply voltages of the converter, which in this case operate at a nominal 3.3 volts. Because the MSP430 and Piccolo Microcontroller have equal maximum ratings, the same input protection may be used on both modules.

## Microcontroller Input Logic and Data management

### Analog – Digital Converter (ADC)

In order to monitor all of the various conditions we need to keep track of to keep our system stable, or for the system to take input, we are going to require an analog-to-digital converter. An analog-to-digital converter (or ADC) is a device that takes a continuous signal and translates that signal into a quantized value that represents that strength, or amplitude, of that signal.

There are various factors and details we must understand about the ADC in order to effectively use one for our project. One of these concepts includes the resolution of the ADC. The resolution of the ADC represents the range of discrete values that the converter can produce when converting an analog signal. The higher the resolution the converter has, the more accurate the data we can receive, as we are able to reduce error with a larger set of values to use. The resolution of the ADC is generally stored in the form of a binary number, so the level of resolution is normally quantized as a power of two; the more bits that a binary number has available, the larger the range of discrete values.

One factor we must consider with the resolution of the ADC is quantization error, which is the rounding error produced when converting a continuous signal to a digital number. This is because a continuous signal does not have a rounded value, and the true signal amplitude gets cut off when translated into a binary number. This means that in order to reduce our quantization error, we want to have a high resolution to have more accurate data.

The sampling rate of an ADC is the rate at which the converter samples the amplitude of the signal. An analog signal is continuous, and this flow must be converted into digital values. In order to do this, the converter actually samples the signal at discrete time intervals. The rate at which the ADC samples the signal is known as the sampling rate.

We must be mindful that our ADC is capable of a high sampling rate, as a higher sampling rate results in more accurate data. Specifically, we want our converter to be able to sample the input signal at a frequency that is at least double the highest possible frequency the input signal is capable of. This is explained by the Shannon-Nyquist Sampling Theorem, which states that if a signal is sampled at twice the highest frequency of the signal, then the continuous signal can be successfully interpolated by the discrete values.

### Basics of Data Storage and Management

When programming directly with hardware, we must understand the concepts of memory and data management so that the software integrates correctly with the system. The main concepts that we must cover in order to fully understand what is happening in the microcontroller are registers and the stack pointer

Registers

Registers in a microcontroller are essentially volatile blocks of memory in the microcontroller that are used simply to hold values and variables. When dealing with mathematical operations, registers normally will have some sort of place in the equation, whether it be a variable being used in computations, or where the solution to the equation will be stored.

An example of how we can further clarify this idea would be variables in high level programming. Take the language C, for example. If we want to use any variables in C to hold values, we simply have to declare them. Depending on the size of the total memory of the computer, we could declare an incredibly large number of variables that we would want to use in our code. The number of variables is practically dynamic, with the developer being able to create any variable they wish to use whenever they need it.  
  
In a microcontroller, we can imagine registers as pre-declared variables that are meant to hold data we are working with. Instead of traditionally declaring variables as we need them and allocating the space accordingly, we already have a set amount of space automatically dedicated to storing variables. These registers will be used to hold and carry special values that we may need to store or send to the system. There are some temporary registers that are traditionally used for lower level programming, but we will not be using these. Instead, we will be accessing any special registers used to read values from certain peripherals, or sending information through peripherals to other parts of the system.

There are also other various registers that we will be using for the system to function. In the microcontrollers we are using, we will be using various timers, along with interrupts as well. In order to access those peripherals, we have to actually activate them in the code we are executing. We do this by declaring a specific variable. This variable is typically associated with a special register used on the hardware, and can vary by microcontroller. We then use a set of predetermined mask variables to set the specific bits in the register. The bits in these registers determine how the timer will run in the system.

This concept will also apply to interrupts as well. Since we are going to have various moments where we will be checking certain inputs, as well as sending data to multiple outputs, we will need to configure interrupts in the system in order for these services to be performed. In terms of how the software will be running the system, we will have to essentially “take breaks” in the middle of our code in order to service certain problems, such as updating the LCD display. These “interrupt” our current code process in order to take care of another problem. In order to configure these interrupts, we have to write specific values to the interrupt registers, just as we did with the timer.

We will most likely be using registers for one of two reasons: the first one being that we are dealing with data that will most likely be changing rapidly, most likely by mathematical operations being run in-code. The other reason would be that we will be configuring the interrupt and timer registers so that we can time our processes properly.

The Stack Pointer

We can imagine the structure of memory as a bunch of blocks stacked on top of each other. Some of these blocks have data in them, such as register values or memory addresses. However, there is a large part of this stack that has no data in many consecutive blocks. When one refers to the “stack” in memory, there are often referring to this large, empty part.

The reason why this empty space exists is so that we can reliably store information in memory and use it whenever we need it. One may think that we can simply use registers for this, but as mentioned before, general-purpose registers are volatile and change due to unforeseen conditions in code. However, in memory, data values are not changed unless explicitly told to. This makes writing to memory more reliable than storing all of our data in registers. Whenever we need more room, we can allocate space block-by-block in this part of memory.

However, we run into a problem when we just allocate memory for data storage. How do we know where the filled blocks of memory end, and where do the empty blocks of memory begin? This issue is resolved by the stack pointer. The stack pointer is an address stored in memory that points to the block of memory that is also the bottom of the stack. This means that the address that the stack pointer references is the last piece of memory being used in the stack. The following diagram below shows a basic structure of memory before the stack pointer is moved:

|  |  |  |
| --- | --- | --- |
| Stack Pointer | Address (in hex) | Data Blocks |
| SP==> |  |  |
|  | 0x0000 | Block 0 (not in use) |
|  | 0x0001 | Block 1 (not in use) |
|  | 0x0002 | Block 2 (not in use) |
|  | 0x0003 | Block 3 (not in use) |

As we can see, the stack pointer is not actually pointing to any data yet. This is because no space on the stack has been allocated for use yet. The next diagram shows an example of the same memory structure, but this time the stack pointer has been allocated to include block 0, block 1, and block 2 to be available:

|  |  |  |
| --- | --- | --- |
| Stack Pointer | Address (in hex) | Data Blocks |
|  |  |  |
|  | 0x0000 | Block 0 (in use) |
|  | 0x0001 | Block 1 (in use) |
| SP==> | 0x0002 | Block 2 (in use) |
|  | 0x0003 | Block 3 (not in use) |

Now we can use these three blocks of memory in our program. When we move the stack pointer to include more memory, we call this “decrementing the stack pointer.) We say this since we can imagine the stack pointer descending down the stack. To access these blocks of memory, we first reference the address of the stack pointer. This will bring us to the bottom of the stack. We then have to add the distance from the stack pointer to the memory address that we want. In the example above, if we wanted to access block 0, we would first reference the stack pointer, and then increment the reference by the size of two blocks of memory. If we wanted to deallocate this space once our program was finished running, we simply move the stack pointer back up to the top of the full memory block. We call this “incrementing the stack pointer.”

This technique allows us to physically allocate space for very important variables in our system. We would most likely implement this technique in storing variables that would be accessed by our peripherals. An example would be storing the values that will be shown on the LCD display. Since we want to have very accurate being displayed, we want this variables stored in a nonvolatile portion of memory.

## LCD Display

The LCD display is going to be used to display general information about the entire system. The following information we will be showing is including, but not limited to, the motor speed in rotations per minute, the electromagnetic torque in foot-pounds produced, and the power output by the system in Watts. These main variables are meant to give the standard user information on the system as it is operating.

A feature we will be including in the display is a diagnostic mode. The purpose of this mode is to show an advanced user additional information on the system that a normal user may not understand. This information includes, but is not limited to, the temperature of the motor in Celsius, the temperature of the power switches in Celsius, the temperature of the inverter in Celsius, the temperature of the rectifier circuit in Celsius, and the values of the D and Q current vectors of the motor in Amperes.

The specific display that has been selected is the CFAX12864U1-TFH by Crystalfontz. This part was selected for various reasons. To start, this display is a 128x64 parallel graphic LCD, meaning that there will be plenty of individual pixels to work with. This means that there will be plenty of freedom in terms of how we want to format the information we are displaying, including whether we want to show a lot of information in a smaller font, if we want to exhibit a few key measurements in a larger font, or even some sort of combination of the two. It also features a white backlight, meaning that it will be easier to view the screen even in low-light environments.

# Project Prototype Construction and Coding

## Bill of Materials



## PCB Vendor and Assembly

We will have our PCB made by Osh Park and any surface mount placement done by Quality Manufacturing Services locally. Osh Park is chosen because of their relatively short lead time and past record serving UCF senior design teams. Quality Manufacturing Services is chosen in the interests of keeping the assembly local. Locality of the company working with your board has some big plusses. A face to face conversation can go a long way.

## Final Algorithm Structure

### LCD Display Control

The following diagram represents the logic that will be implemented for driving the LCD display:

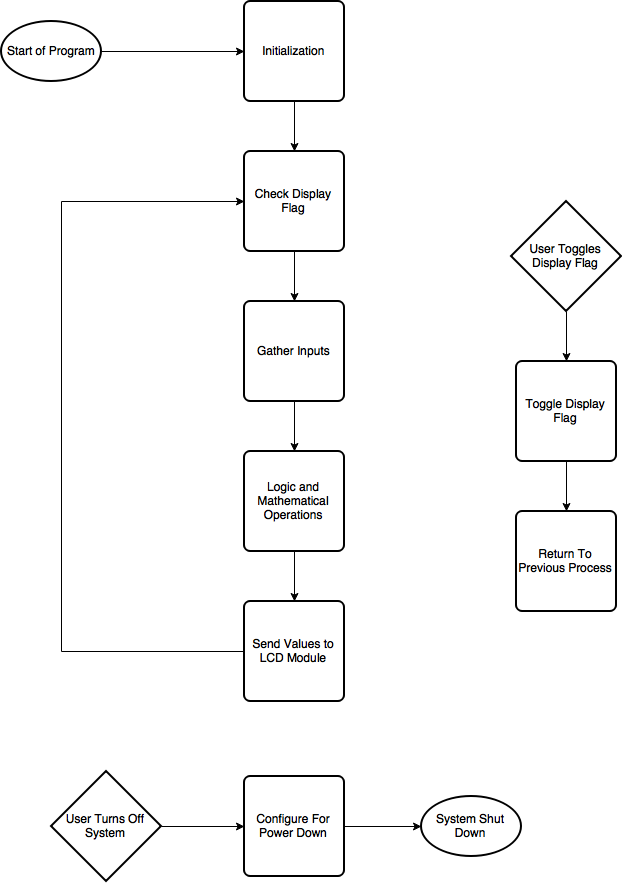
[](https://drive.draw.io/#G0B4PlVlS7guU8RnBLYWpqZWJ1Wnc)

Figure 37 - LCD Display Algorithm Flowchart.

#### Initialization

As soon as the system is turned on, the MSP430 will immediately begin its initialization phase. This block will set up any initial variables that will need to be declared. This includes configuring any clocks that will be used, any and all input and output pins that will be used, initializing the LCD display such that it is in the proper operating mode.

#### The LCD Character Array

This stage will also initialize the LCD character array. In order to display any characters on the LCD display, the characters will have to be translated into a grid of pixels. The dimensions of the grid will depend on the size of the character. Most likely, dimensions of each character will be 6x8 (this means a character will consist of a total width of 6 pixels and a height of 8 pixels.) This means that a character can be represented by a total of six hexadecimal values. The following diagram below represents an example of the number “3” broken down into a 6x8 pixel grid.

|  |  |  |  |  |  |
| --- | --- | --- | --- | --- | --- |
|  |  |  |  |  |  |
|  |  |  |  |  |  |
|  |  |  |  |  |  |
|  |  |  |  |  |  |
|  |  |  |  |  |  |
|  |  |  |  |  |  |
|  |  |  |  |  |  |
|  |  |  |  |  |  |

Figure 38 - Pixel Breakdown for “3”.

The way that we represent this as individual numbers is representing each column of pixels as a hexadecimal number. Each individual pixel is a single bit of the total eight bits in a hexadecimal number. If the pixel is dark, then the bit represents the pixel as a ‘0’ or the pixel being off. If the pixel is light, then the bit represents the pixel as a ‘1’ or the pixel being on. Using this logic, we can now translate the character above into six hexadecimal values, as shown below:

|  |  |  |  |  |  |
| --- | --- | --- | --- | --- | --- |
| 1 | 1 | 1 | 1 | 1 | 1 |
| 1 | 0 | 0 | 0 | 1 | 1 |
| 0 | 1 | 1 | 1 | 0 | 1 |
| 1 | 1 | 1 | 1 | 0 | 1 |
| 1 | 1 | 0 | 0 | 1 | 1 |
| 1 | 1 | 1 | 1 | 0 | 1 |
| 0 | 1 | 1 | 1 | 0 | 1 |
| 1 | 0 | 0 | 0 | 1 | 1 |

Figure 39 - Pixel Breakdown for “3” with Bit Assignment.

Since the numbers are representing columns, we have to read each column as number, from top to bottom. The next column then represents the next number, and so on and so forth. The following hexadecimal numbers are being used for this specific example:

|  |  |  |  |  |  |  |
| --- | --- | --- | --- | --- | --- | --- |
| **Binary** | 0b11011101 | 0b10111110 | 0b10110110 | 0b10110110 | 0b11001001 | 0b11111111 |
| **Hex** | 0xDD | 0xBE | 0xB6 | 0xB6 | 0xC9 | 0xFF |

Figure 40 - Hexadecimal Set for Representing “3”.

In order to display all of our information, we will need to have a representation of every letter, number, and some special characters and punctuation, so this process will be performed for every character that is used.

#### Check Display Flag

The LCD display will have two different variants of showing information on the system. There will be a "standard" mode and a "diagnostic" mode. The only difference between the two is that "diagnostic" mode displays the same information as "standard" mode, but also includes additional information on the system. The following table below describes the differences between the two modes:

|  |  |  |  |
| --- | --- | --- | --- |
| **Comparison of LCD Display Modes** | | | |
| **Information** | **Standard Mode** | **Diagnostic Mode** | **Units** |
| Motor Speed | Yes | Yes | RPM |
| Electromagnetic Torque | Yes | Yes | ft·lb |
| Power Output | Yes | Yes | W |
| Temperature of Motor | No | Yes | °C |
| Temperature of Power Switches | No | Yes | °C |
| Temperature of Inverter | No | Yes | °C |
| Temperature of Rectifier | No | Yes | °C |
| D Current Vector Value | No | Yes | A |
| Q Current Vector Value | No | Yes | A |

Table 41 - Comparing the Different Display Modes.

In order to determine which mode of operation the user wishes to view, the program will declare and keep track of a Boolean flag. If the flag is toggled to *false*, then the display will operate in the "standard" mode of operation. If the flag is toggled to *true*, then the display will operate in the “diagnostic” mode of operation. If the flag has been toggled since the last time that the system has entered this part of the code, then the LCD display will perform any additional initializations in this block. These initializations include writing the any additional variable names along with their units (if the display is switching from “standard” to “diagnostic,”) or clearing those variable names and units (if the display is switching from “diagnostic” to “standard.”)

#### Gather Inputs

At this stage of the program, the program will have determined its mode of operation. Based on the mode of operation, we will gather various inputs from the system. If the display is in "standard" mode, then we will only gather the inputs for the motor speed, the electromagnetic torque, and the power output. If the display is operating in “diagnostic” mode, then we will also gather the inputs for the temperature of the motor, the temperature of the power switches, the temperature of the inverter, the temperature of the rectifier, the D current vector value, and the Q current vector value, as well as the inputs gathered in the "standard" mode of operation.

#### Logic and Mathematical Operations

Once the microcontroller has gathered all of the necessary inputs, we will have to perform various calculations on each input in order to get the correct measurement out of the initial values. The operations performed will vary between each input.

#### Display Flag Toggled

This procedure will be an interrupt routine that will only occur when the user wishes to switch display modes. The user will ideally toggle the flag by pressing a button that will be attached to one of the MSP430 general purpose I/O pins. Whenever this button is pressed, the display flag will be toggled to the opposite of its current value. The program will then continue to execute whatever action it was in the middle of when the flag was toggled. When the program loop is finished executing, then the loop will begin again, but this time the display will operate in the other display mode.

#### Send Values to LCD Module

We now have all of the information that we wish to display on the LCD screen. However, we cannot simply send the value to the display itself. As covered in The LCD Character Array, we have to represent characters as hexadecimal values. In order to write to the LCD module, we will have to take each value, break it down into its individual characters, and write each value character-by-character to the screen.

For an example, if the motor was rotating at 156 RPM, we would first have to determine how much screen space we would need to allocate for the value. To do this, we will compare the value we are dealing with by powers of ten, with each increasing power adding an additional 6x8 pixel grid dedicated to the display.

Once we have determined the amount of space we will be using for the specific value, we then have to determine the individual digits in the value. This will allow us to pull the hexadecimal numbers necessary to display the specific character onto the LCD screen. All of the values we will be working with will not be too large, and any decimal values will most likely have only a few significant figures when rounding off, so determining individual digits will not require very complex mathematics.

Going off of the example, the following diagram shows the pixelated breakdown of each digit, along with the derivations of their hexadecimal values:

|  |  |  |  |  |  |
| --- | --- | --- | --- | --- | --- |
| 1 | 1 | 1 | 1 | 1 | 1 |
| 1 | 1 | 0 | 0 | 1 | 1 |
| 1 | 0 | 0 | 0 | 1 | 1 |
| 1 | 1 | 0 | 0 | 1 | 1 |
| 1 | 1 | 0 | 0 | 1 | 1 |
| 1 | 1 | 0 | 0 | 1 | 1 |
| 1 | 1 | 0 | 0 | 1 | 1 |
| 1 | 0 | 0 | 0 | 0 | 1 |

Figure 41 - Pixel Breakdown for “1” with Bit Assignment.

|  |  |  |  |  |  |  |
| --- | --- | --- | --- | --- | --- | --- |
| **Binary** | 0b11111111 | 0b11011110 | 0b10000000 | 0b10000000 | 0b11111110 | 0b11111111 |
| **Hex** | 0xFF | 0xDE | 0x80 | 0x80 | 0xFE | 0xFF |

Table 42 - Hexadecimal Set for Representing “1”.

|  |  |  |  |  |  |
| --- | --- | --- | --- | --- | --- |
| 1 | 1 | 1 | 1 | 1 | 1 |
| 0 | 0 | 0 | 0 | 0 | 1 |
| 0 | 1 | 1 | 1 | 1 | 1 |
| 0 | 1 | 1 | 1 | 1 | 1 |
| 0 | 0 | 0 | 0 | 0 | 1 |
| 1 | 1 | 1 | 1 | 0 | 1 |
| 1 | 1 | 1 | 1 | 0 | 1 |
| 0 | 0 | 0 | 0 | 0 | 1 |

Figure 42 - Pixel Breakdown for “5” with Bit Assignment.

|  |  |  |  |  |  |  |
| --- | --- | --- | --- | --- | --- | --- |
| **Binary** | 0b10000110 | 0b10110110 | 0b10110110 | 0b10110110 | 0b10110000 | 0b11111111 |
| **Hex** | 0x86 | 0xB6 | 0xB6 | 0xB6 | 0xB0 | 0xFF |

Table 43 - Hexadecimal Set for Representing “5”.

|  |  |  |  |  |  |
| --- | --- | --- | --- | --- | --- |
| 1 | 1 | 1 | 1 | 1 | 1 |
| 1 | 0 | 0 | 0 | 1 | 1 |
| 0 | 1 | 1 | 1 | 0 | 1 |
| 0 | 1 | 1 | 1 | 1 | 1 |
| 0 | 0 | 0 | 0 | 1 | 1 |
| 0 | 1 | 1 | 1 | 0 | 1 |
| 0 | 1 | 1 | 1 | 0 | 1 |
| 1 | 0 | 0 | 0 | 1 | 1 |

Figure 43 - Pixel Breakdown for “6” with Bit Assignment.

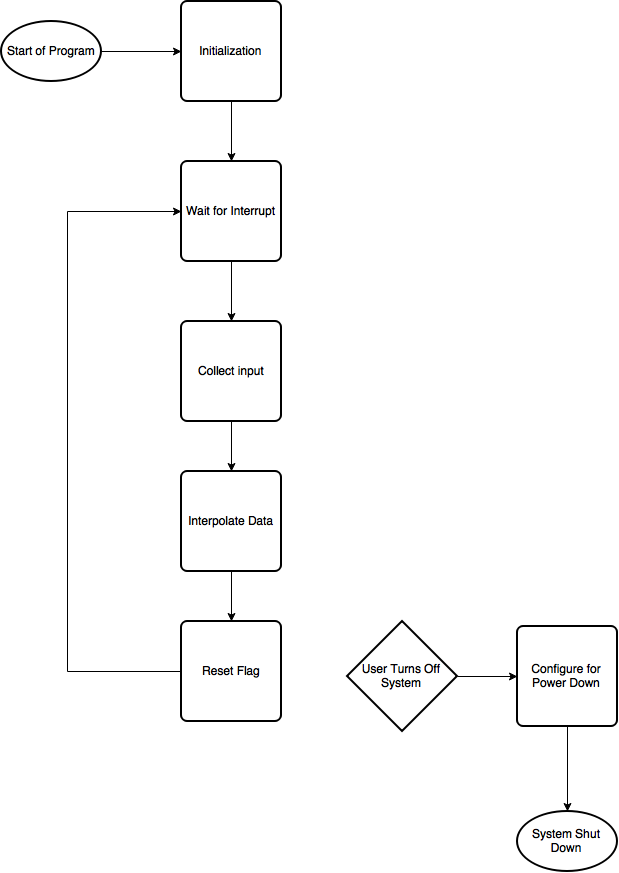
|  |  |  |  |  |  |  |
| --- | --- | --- | --- | --- | --- | --- |
| **Binary** | 0b11000001 | 0b10110110 | 0b10110110 | 0b10110110 | 0b11011001 | 0b11111111 |
| **Hex** | 0xC1 | 0xB6 | 0xB6 | 0xB6 | 0xD9 | 0xFF |

Table 44 - Hexadecimal Set for Representing “6”.

While this process may seem tedious, this work will already be done when the system is initialized at start-up. All that the system will be doing during this block of the algorithm will be identifying the individual digit, identifying the predetermined hexadecimal set that represents that digit, and then writing the values in the set to each column on the LCD display. This process will be repeated for each character that needs to be displayed.  
  
Once all of the necessary characters have been written to the display, the program will repeat back to the beginning of the “Check Display Flag” block and start the whole process over again. This process will continue until the system is shut off.

Configuring I/O Peripherals

The following diagram is a flowchart that will be serve as a base example of programming the input/output peripherals leading to the microcontroller:

[](https://drive.draw.io/#G0B4PlVlS7guU8QnNsRlJMcE42VXc)  
Figure 44 - Peripheral flow chart.

#### Initialization

The specifics of this process will vary based on the individual peripheral. However, the general theory will be the same for each one. The initialization process will involve configuring any general input/output pins on the microcontroller that are used to interface with the peripherals, as well as loading any special registers with initial values so that the peripheral behaves the way we want it to. This process may not be necessary for all devices in our system.

#### Wait For Interrupt

The microcontroller will require a method of knowing when to read a new value from its input/output pins. One method of performing this would be constantly reading in a new value from the pin. However this can provide a much more redundant process, as the value read from the peripheral will not change unless a new value is loaded into it. This means that we will be performing the same operations with the same information over and over again until a new value is loaded into the peripheral. To remove this redundancy, we will set an interrupt to go off whenever the peripheral has a new value loaded into it. This will cause the microcontroller to be put in a stand-by state until new calculations are ready to be made.

#### Collect Input

Once the interrupt has been set off, the microcontroller will receive the input from the peripheral that set off the interrupt flag. The microcontroller will now be able to perform any calculations necessary to continue our system process.

#### Interpolate Data

The microcontroller has now receive the input from the peripheral. The input now needs to be interpolated so that it will be in a format in relation to the system, rather than in the scope of the ADC. We do this normally by multiplying the input by some predetermined ratio to scale the value from the ADCs resolution to the values we expect to find in our system.

#### Reset Flag

Once we have performed all of the calculations necessary to translating the information we have received from our peripherals, we will move the data to whatever process it is necessary for, such as displaying information on the LCD module, or outputting a PWM signal to the inverters. We will then reset the interrupt flag so that the system will be ready to receive a new input whenever it is ready.

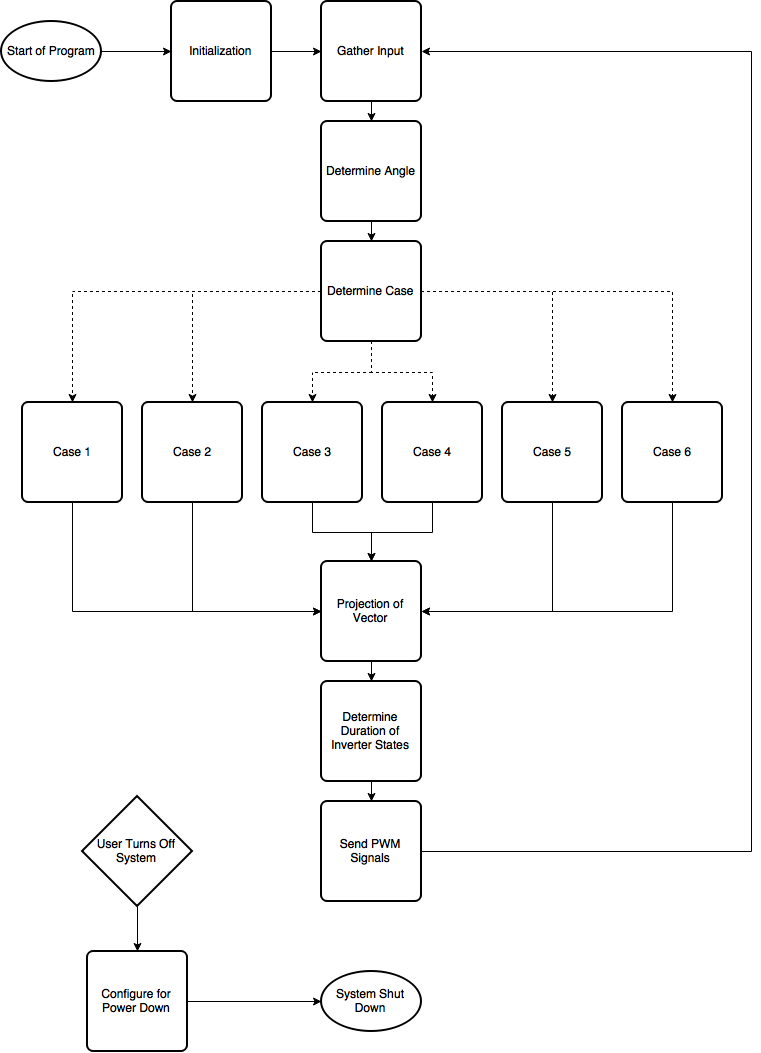
The process will now repeat until the user chooses to turn the system off.

#### Configure for Power Down

Once the user chooses to power down the system, there may be various peripherals that need to be programmed for shut-down. This may include writing values to registers and configuring timers, or even hard coding the peripheral itself to power down.

PWM Controller Algorithm

The following diagram is a flowchart demonstrating the logic of the PWM controller:

[[](https://drive.draw.io/#G0B4PlVlS7guU8WjdrQzhzZzJvV2s)](https://drive.draw.io/#G0B4PlVlS7guU8WjdrQzhzZzJvV2s)Figure 45 - PWM Logic Flowchart.

#### Initialization

The initialization of the PWM controller will involve configuring any pins that will be used in switching the individual inverters, as well as initializing any pins that will be used for collecting inputs from the system.

#### Gather Input

This process simply represents collecting any inputs necessary to drive the PWM. This may include many sample voltages from the system that will be averaged, or perhaps one sample that will be taken as the current value of the system. This may also include any encoder data that is used to describe the motor speed.

#### Determine Angle

Once all of the inputs have been gathered, we will have two sample voltages Vα and Vβ. We will use these voltages to represent a vector. In order to determine what inverter states we are going to use, we must first determine which sector that the vector is in. This is done by taking the inverse-tangent of our vector. We will then be able to determine which specific case we will be implementing in the current cycle.

#### Determine Case

There are 6 different individual cases that will determine the next inverter states that will be used to drive the motor. The following table represents the different possible outcomes:

|  |  |  |
| --- | --- | --- |
| **Cases for PWM Algorithm** | | |
| **Case (Sector)** | **Range of Angles** | **Inverter States Used** |
| 1 | Between 0° and 60° | State 1 (001) and State 3 (011) |
| 2 | Between 60° and 120° | State 3 (011) and State 2(010) |
| 3 | Between 120° and 180° | State 2 (010) and State 6 (110) |
| 4 | Between 180° and 240° | State 6 (110) and State 4 (100) |
| 5 | Between 240° and 300° | State 4 (100) and State 5(101) |
| 6 | Between 300° and 360° | State 5 (101) and State 1 (001) |

Table 45 - PWM Algorithm Cases.

#### Projection of Vector

Once we have determined which sector that our voltage vector is in, we must then project the vector onto the two neighboring inverter-state vectors. The magnitude of the projection on these inverter-state vectors will represent the duration of how long the system will be in that specific inverter-state.

#### Determine Duration of Inverter States

After the magnitudes of the inverter-state vector projections are found, we must determine the time durations in which the inverter states will be selected in sequence. This differs from the duration of the projections themselves, as we must toggle the inverters in such a way that the pulse of the PWM is properly represented in the system.

#### Send PWM Signals

Once the individual time durations of the sequence have be determined, we will transmit the individual signals to each inverter for specified amounts of time, which would be determined in the previous step. Once the signals have been generated and sent to the inverters, the microcontroller will then begin to gather new inputs and begin the process over again. This will continue to loop until the user powers down the system.

#### Configure for Power Down

This process includes reconfiguring any pins, registers, and peripherals that need to be reinitialized for the system to be properly powered down.

# Project Prototype Testing

## Software Test Environment

Initial software testing will begin with the CFAX12864U-TFH LCD module, and will be done using the MSP430F5529 microcontroller to drive the LCD, a desktop computer running 64-bit Windows 7 Ultimate edition, and running any software using Code Composer Studio. Additional testing involving the rest of the system will be performed on a portable laptop computer running Windows 10. The software being used for coding and debugging will still be Code Composer Studio.

## Software Specific Testing

LCD Module Testing

The initial testing of the LCD module will include making sure all of the registers are in working order. This will involve writing any values to such registers such that all of the pixels will blink. This will ensure that the individual pixels that we wish to use will be accessible in such a manner that will be useful for our system.

Any additional testing involving the LCD module alone will involve making sure that the character array being used will display properly on the display. Once we are sure that we can display any values that we wish to use, we will then attach additional peripherals to the system to see if we will be able to properly display any inputs that the system will receive.

I/O Peripheral Testing

Any testing involving the programmable peripherals that will be used for input or output will first involve making sure that we will be able to access any peripherals from the microcontroller pins. This will likely involve sending a simple test signal from the peripheral to the controller, and then having the microcontroller read the value. We will know what the input should be, so if the microcontroller returns a different value, then we will know that there are issues with the input peripheral.

PWM Controller Testing

Testing involving the PWM drive will initially begin as making sure that we are receiving the correct inputs from any peripherals to the Piccolo microcontroller. We will then provide basic test cases that will involve the various transformations involved in the space-vector modulation process so that we will test any mathematics that will be simulated on the microcontroller itself. We will monitor the output on a display to make sure that the process that we are modeling is done correctly.

# Administrative Content

## Milestone Discussion

The first milestone for this project was the decision to pursue the variable frequency drive (VFD) over all the other project designs. This project suited everyone’s desires to explore into a field of engineering that was not covered in any one particular class.

Following this was the decision to cater our design toward a moving vehicle. Tesla’s electric vehicle inspired us to develop a VFD that would appropriately control a go-kart type vehicle. This decision led to the addition of a battery pack and battery cell analysis and control. A regenerative breaking scheme was also an addition to the project.

Initially the VFD was going to be driven by the volts of frequency scheme so that the speed of the motor would be determined by a variable voltage produced from a potentiometer. This scheme was later changed due to the desire of a regenerative braking control scheme for the vehicle. To apply regenerative braking a vector oriented control scheme was needed.

Most research that talked about vector oriented controls discussed the Clarke transformation which converted the currently three phase system (where the rotor values are on a rotating plane) into a three phase system (one phase being a zero vector phase) with the possibility of expressing the rotor values as a stationary set of windings that depended on the angle of the rotor compared to its initial condition.

After the Clarke transformation was the Park transformation which expressed the Clarke transformation as two permanently stationary axes and a third zero vector axis. The Park transformation allowed for complete control of the motor with just two variables.

After the recognition of the usefulness and application of the Clarke and Park transforms the utilitarianism of space vector modulation (SVM) and its ability to project a desired value in the domain of the Clarke transformation onto the original three phase domain without performing any inverse calculations from the Clarke transformation.

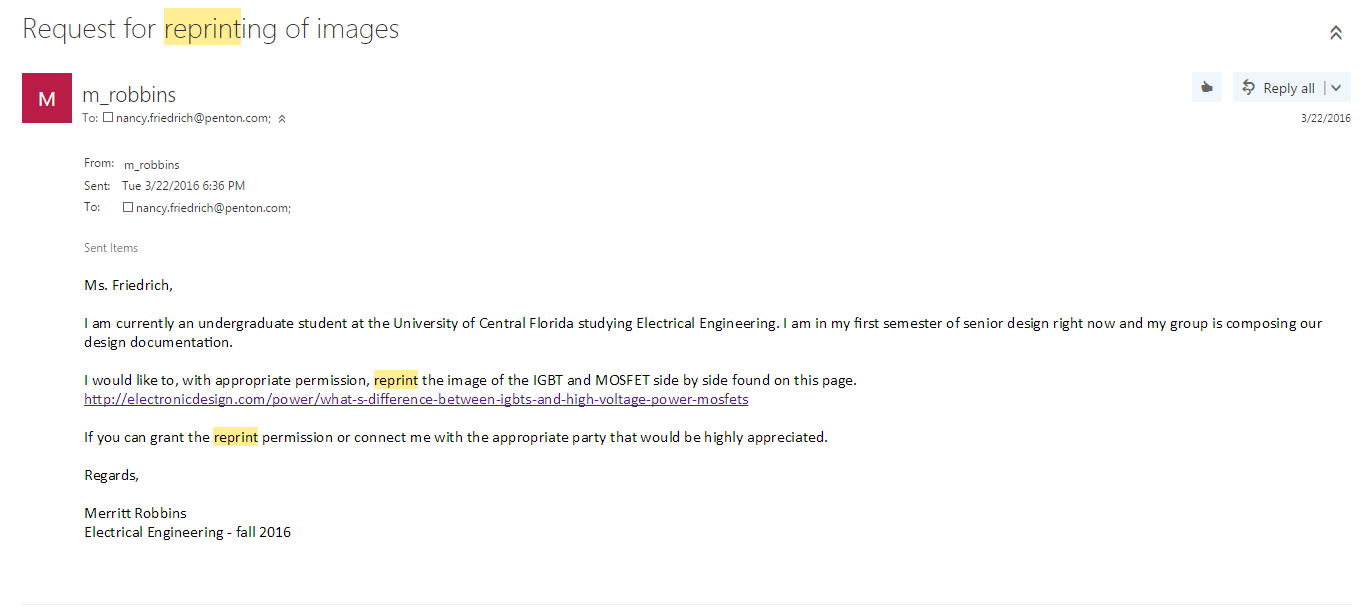
The output of SVM is a pulse width modulated (PWM) signal used to control the inverters. This use of PWM helped decide the microcontroller used to process the data and perform calculations. This controller is the Piccolo controller. This controller was fast and consumed low power and had more than enough channels to implement any method of PWM generation.

To display the speed of the rotor and the results from the Piccolo controller, another microcontroller and LCD module were selected to implement test sequences and show the current state of the motor.

About two thirds of the way through the semester it became apparent that a vehicle would be too excessive for a project since the battery management scheme would have been a senior design project all by itself. This decision led to the drop of the battery component of the project and the switch to the VFD being a wall plug-in that ran something stationary like a fan.

After revisiting the bridge rectification and the two sets of inverters (categorizing the inverters into positive and negative) the DC voltage that would have been generated by a diode full bridge rectifier would suffer significant voltage loss due to the nature of the diodes and would be significantly large. The full bridge rectifier was then replaced with a synchronous bridge rectifier that operated with less power consumption and was more efficient with high voltages. The inverter configuration was then switched so that the DC voltage required was only half the previous DC voltage. When applying a voltage to the negative inverter per phase, the polarity of the DC will change due to the added inverters.

# Appendix A – Copyright Permissions



# Appendix B – References

|  |  |
| --- | --- |
| [1] | R. H. Park, "Two-Reaction Theory of Synchronous Machines, Generalized Method of Analysis - Part I," *Transactions of the American Institute of Electrical Engineers,* vol. 48, no. 3, pp. 716-727, 1929. |
| [2] | EVDrive, "EVD Motor/Controller Packages," 2016. [Online]. Available: http://www.evdrive.com/products/evd-motor-controller/. [Accessed 26 March 2016]. |
| [3] | Tesla Motors, "Model S Specifications," 2016. [Online]. Available: https://www.teslamotors.com/support/model-s-specifications. [Accessed 2 April 2016]. |
| [4] | E. Cowern, "Understanding Induction Motor Nameplate Information," Baldor Electric Company, Fort Smith, 2004. |
| [5] | Texas Instruments Europe, "Field Oriented Control of 3-Phase AC-Motors," February 1998. [Online]. Available: http://www.ti.com/lit/an/bpra073/bpra073.pdf. [Accessed 20 March 2016]. |
| [6] | Neamen and D. A., Microelectronics Circuit Analysis and Design, New York: McGraw Hill, 2010. |
| [7] | S. L. Peter Wilson, "What's The Difference Between IGBTs And High-Coltage Power MOSFETS?," ElectronicDesign.com, 25 March 2014. [Online]. Available: http://electronicdesign.com/power/what-s-difference-between-igbts-and-high-voltage-power-mosfets. [Accessed 24 March 2016]. |
| [8] | ON Semiconductor, "ON Semiconductor's Motor Control IGBTs and Free - Wheeling Diodes," June 2012. [Online]. Available: http://www.onsemi.com/pub\_link/Collateral/AND9088-D.PDF. [Accessed 26 March 2016]. |
| [9] | M. Carolyn, "The Basics of Current Sensors," Digi-Key Electronics, Riverfalls, 2013. |
| [10] | RohsGuide, "Is Your Facility RoHS Compliant for 2016?," RohsGuide, 2016. [Online]. Available: http://www.rohsguide.com/. [Accessed 26 April 2016]. |
| [11] | A. Trotta, "Induction Machine (Asynchronous Motor) Dynamic Model," 26 June 2015. [Online]. Available: https://www.youtube.com/watch?v=4QZ1E7cvHvw. [Accessed 12 March 2016]. |
| [12] | A. Adamsky, "Build Your Own Low - Cost Driver for a Synchronous Rectifier," Electronic Design, 4 June 2010. [Online]. Available: http://electronicdesign.com/power/build-your-own-low-cost-driver-synchronous-rectifier. [Accessed 9 April 2016]. |
| [13] | D. Giacomini and L. Chiné, "A Novel High Efficient Approach to Input Bridges," 27-29 May 2008. [Online]. Available: http://www.infineon.com/dgdl/tp-080527.pdf?fileId=5546d462533600a40153573fb97c3e9f. [Accessed 14 April 2016]. |
| [14] | M. Beckman, "Precision Thermocouple Measurement with the ADS1118," Texas Instruments, Dallas, 2011. |
| [15] | S. Cubed, "Protecting Inputs in Digital Electronics," Digi-Key Electronics, River Falls, 2011. |
| [16] | S. More, "ADC Input Protection," Texas Instruments, Dallas, 2013. |
| [17] | "Bluetooth Core Specification," Bluetooth, 2016. [Online]. Available: https://www.bluetooth.com/specifications/bluetooth-core-specification. [Accessed 10 3 2016]. |
| [18] | Cadex - Battery University, "What's the Best Battery?," Coalescent Design, 1 11 2010. [Online]. Available: http://batteryuniversity.com/learn/article/whats\_the\_best\_battery. [Accessed 20 March 2016]. |
| [19] | A. Fitzgerald, C. Kingsley and S. D. Umans, Electric Machinery, New York: McGraw-Hill, 2003. |

# Appendix D – Table of Figures

[Figure 1 - A model of Tesla’s first induction motor. Tesla museum in Belgrade, Belgium 16](#_Toc449571972)

[Figure 2 - Cutaway view of a three-phase squirrel-cage motor. Reprinted with permission from Rockwell Automation/Reliance Electric 17](#_Toc449571973)

[Figure 3 - Two State Inverter Set. 23](#_Toc449571974)

[Figure 4 - Inverter State Vectors. 24](#_Toc449571975)

[Figure 5 - Sampled Voltage Vector in Sector S1. 26](#_Toc449571976)

[Figure 6 - Visualization of Time Subintervals 26](#_Toc449571977)

[Figure 7 - Visualization of the State Vector Activity in the ABC-system. 27](#_Toc449571978)

[Figure 8 - Voltage Analysis per Sector. 29](#_Toc449571979)

[Figure 9 - Basic half - bridge rectifier topology modeled in KiCad 30](#_Toc449571980)

[Figure 10 - Full - bridge rectifier topology modeled in KiCad 31](#_Toc449571981)

[Figure 11 - The Silicon Controlled Rectifier Topology. 32](#_Toc449571982)

[Figure 12 - The Active Rectifier using MOSFETS as the switching devices. Note the drive signals entering to control the switching of each FET. Designed in KiCad 33](#_Toc449571983)

[Figure 13 - The synchronous bridge rectifier; the starting point for our input power rectifier design for optimal efficiency. 34](#_Toc449571984)

[Figure 14 - The basic structure of a single phase inverter using IGBTs. Designed in KiCad 35](#_Toc449571985)

[Figure 15 - Cross section of an IGBT, copied with permission requested from ElectronicDesign.com [7]. Permission request can be found in Appendix A – Copyright Permissions. 36](#_Toc449571986)

[Figure 16 - The basic network used for ground fault detection. Designed in KiCad 38](#_Toc449571987)

[Figure 17 - Block diagram of the overall top, system - level perspective. 42](#_Toc449571988)

[Figure 18 - Block diagram of the power system. 43](#_Toc449571989)

[Figure 19 - Block diagram of microcontroller sensor input. 43](#_Toc449571990)

[Figure 20 - Dynamic model high - level view of functionality. Taken from Antonio Trotta, 2015. 57](#_Toc449571991)

[Figure 21 - The Zener voltage regulator with RC network energy storage. Designed in KiCad. 59](#_Toc449571992)

[Figure 22 - The half - wave synchronous rectifier with regulated gate drivers. 60](#_Toc449571993)

[Figure 23 - One half of the synchronous rectifier using IR1167 ICs for MOSFET driving. Designed using KiCad 63](#_Toc449571994)

[Figure 24 - The topology of the low voltage rectifier (approx. 21V). Component selection is justified above. 71](#_Toc449571995)

[Figure 25 - The topology of the 15V synchronous regulator, design was made using provided reference design. 74](#_Toc449571996)

[Figure 26 - The topology of the 3.3V synchronous regulator. Design was made using provided reference design. 78](#_Toc449571997)

[Figure 27 – 3 Phase Inverter using IGBTs with built – in antiparallel free-wheeling diodes. 83](#_Toc449571998)

[Figure 28 - The topology which will be used for the synchronous rectifier. Based on the FAN7190. Designed in KiCad 84](#_Toc449571999)

[Figure 29 - Thermocouple junction diagram- Texas Instruments. 94](#_Toc449572000)

[Figure 30 - Illustration of the Seebeck Effect - Texas Instruments. 94](#_Toc449572001)

[Figure 31 - Two-Channel Thermocouple system – Designed in KiCad 97](#_Toc449572002)

[Figure 32 - Software flow block diagram- Reprinted with permission from Texas Instruments 98](#_Toc449572003)

[Figure 33 - The pin diagram of the MSP430F5529. This will be used as a reference for all KiCad schematics and footprint layouts. 100](#_Toc449572004)

[Figure 34 - The TMS320F28027F Pin diagram. This will be used for the KiCad schematic and PCB footprint. 102](#_Toc449572005)

[Figure 35 - The Interface required between sensor inputs, the MSP430, and the LCD display. 103](#_Toc449572006)

[Figure 36-Input Protection Circuitry Modeled in KiCad 106](#_Toc449572007)

[Figure 37 - LCD Display Algorithm Flowchart. 113](#_Toc449572008)

[Figure 38 - Pixel Breakdown for “3”. 114](#_Toc449572009)

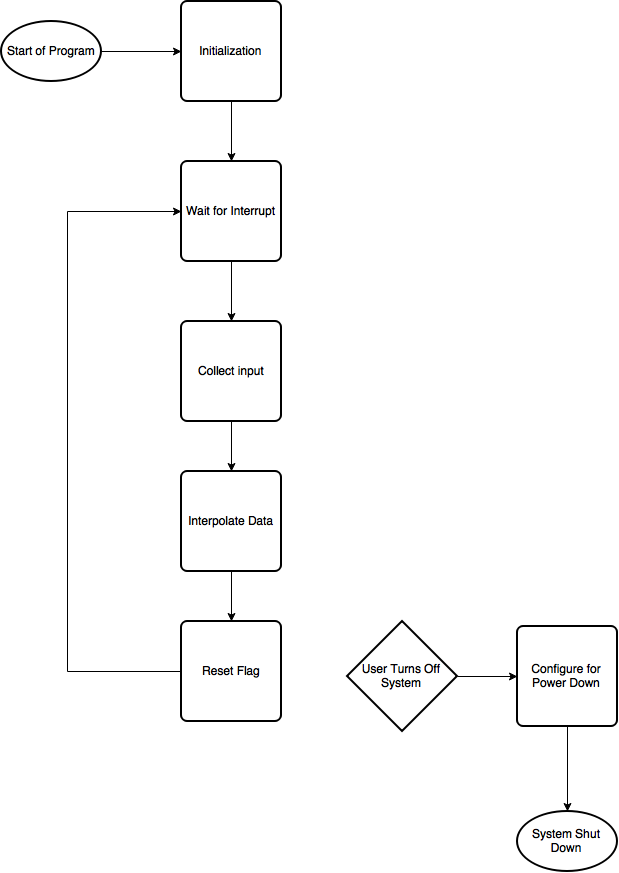
[Figure 39 - Pixel Breakdown for “3” with Bit Assignment. 115](#_Toc449572010)

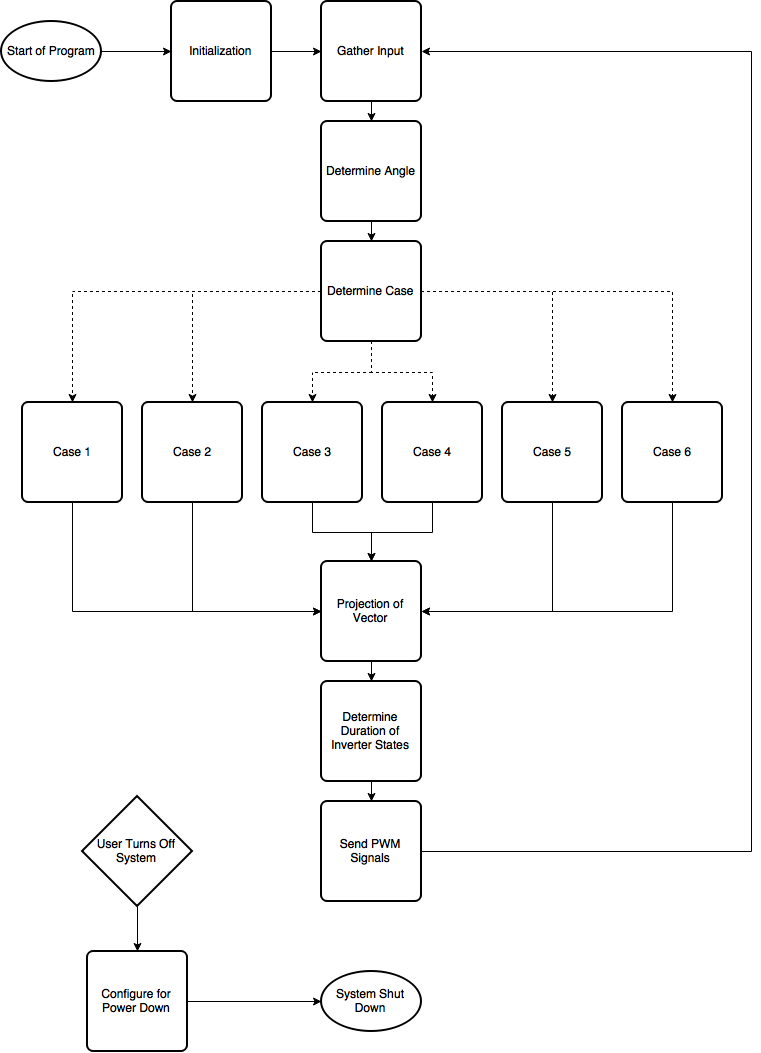
[Figure 40 - Hexadecimal Set for Representing “3”. 115](#_Toc449572011)

[Figure 41 - Pixel Breakdown for “1” with Bit Assignment. 117](#_Toc449572012)

[Figure 42 - Pixel Breakdown for “5” with Bit Assignment. 118](#_Toc449572013)

[Figure 43 - Pixel Breakdown for “6” with Bit Assignment. 118](#_Toc449572014)

[[](https://drive.draw.io/#G0B4PlVlS7guU8QnNsRlJMcE42VXc) Figure 44 - Peripheral flow chart. 120](#_Toc449572015)

[[](https://drive.draw.io/#G0B4PlVlS7guU8WjdrQzhzZzJvV2s) Figure 45 - PWM Logic Flowchart. 122](#_Toc449572016)

[Figure 46 - Input parameters used in the simulation of the AC induction motor. ix](#_Toc449572017)

[Figure 47 - The dq0 Transformation applied to the stator voltages Va, Vb, and Vc. The dark green block seen from the top level. ix](#_Toc449572018)

[Figure 48 - Calculation of rotor and stator flux vectors. The blue block when viewed from the top level. x](#_Toc449572019)

[Figure 49 - Calculation of rotor and stator currents. The yellow block when viewed from the top level. xi](#_Toc449572020)

[Figure 50 - Calculation of Electromagnetic torque from the motor. The light green block when viewed from the top level. xii](#_Toc449572021)

[Figure 51 - Calculation of motor mechanical speed using a varying load torque. The orange block when viewed from the top level. xii](#_Toc449572022)

[Figure 52 - Varying load torque generation. The white block when viewed from the top level. xiii](#_Toc449572023)

[Figure 53 - The input transformers. Connectors go to the rectifiers. xiv](#_Toc449572024)

[Figure 54 - Synchronous Bridge one bridge per DC link rail. xv](#_Toc449572025)

[Figure 55 - DC Link formation. Each block contains one bridge rectifier. xvi](#_Toc449572026)

[Figure 56 - Low voltage Schottky bridge. xvi](#_Toc449572027)

[Figure 57 - 15V DC regulator. xvii](#_Toc449572028)

[Figure 58 - 3.3V DC regulator. xvii](#_Toc449572029)

[Figure 59 - The power inverter. xviii](#_Toc449572030)

[Figure 60 - ADC clamping input. xix](#_Toc449572031)

[Figure 61 - MSP430 and LCD display interface. xix](#_Toc449572032)

[Figure 62 - Thermocouple sensing design. xx](#_Toc449572033)

# Appendix E – Computer Simulation Screenshots

## MATLAB – Simulink Model of Induction Motor Dynamic response

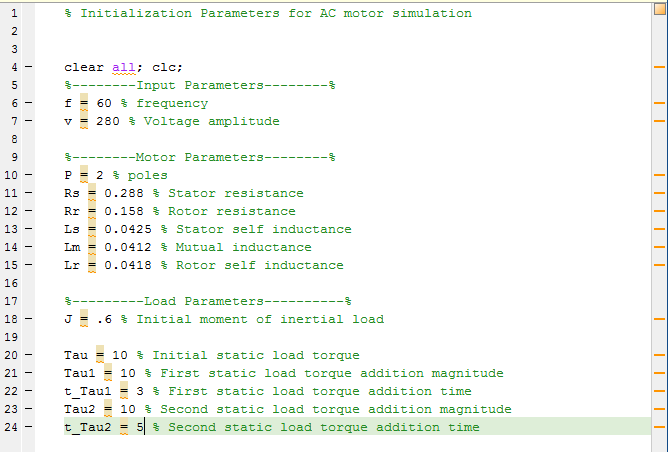


Figure 46 - Input parameters used in the simulation of the AC induction motor.

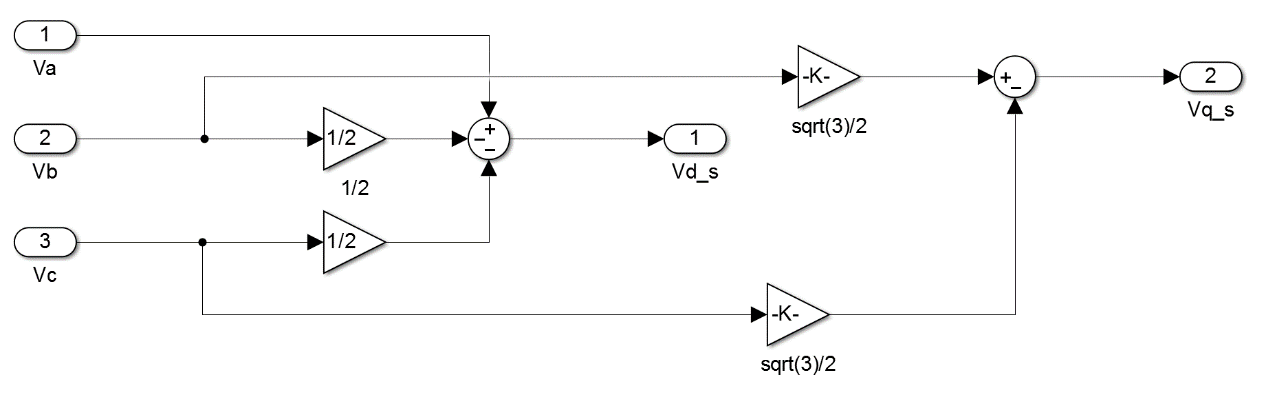


Figure 47 - The dq0 Transformation applied to the stator voltages Va, Vb, and Vc. The dark green block seen from the top level.

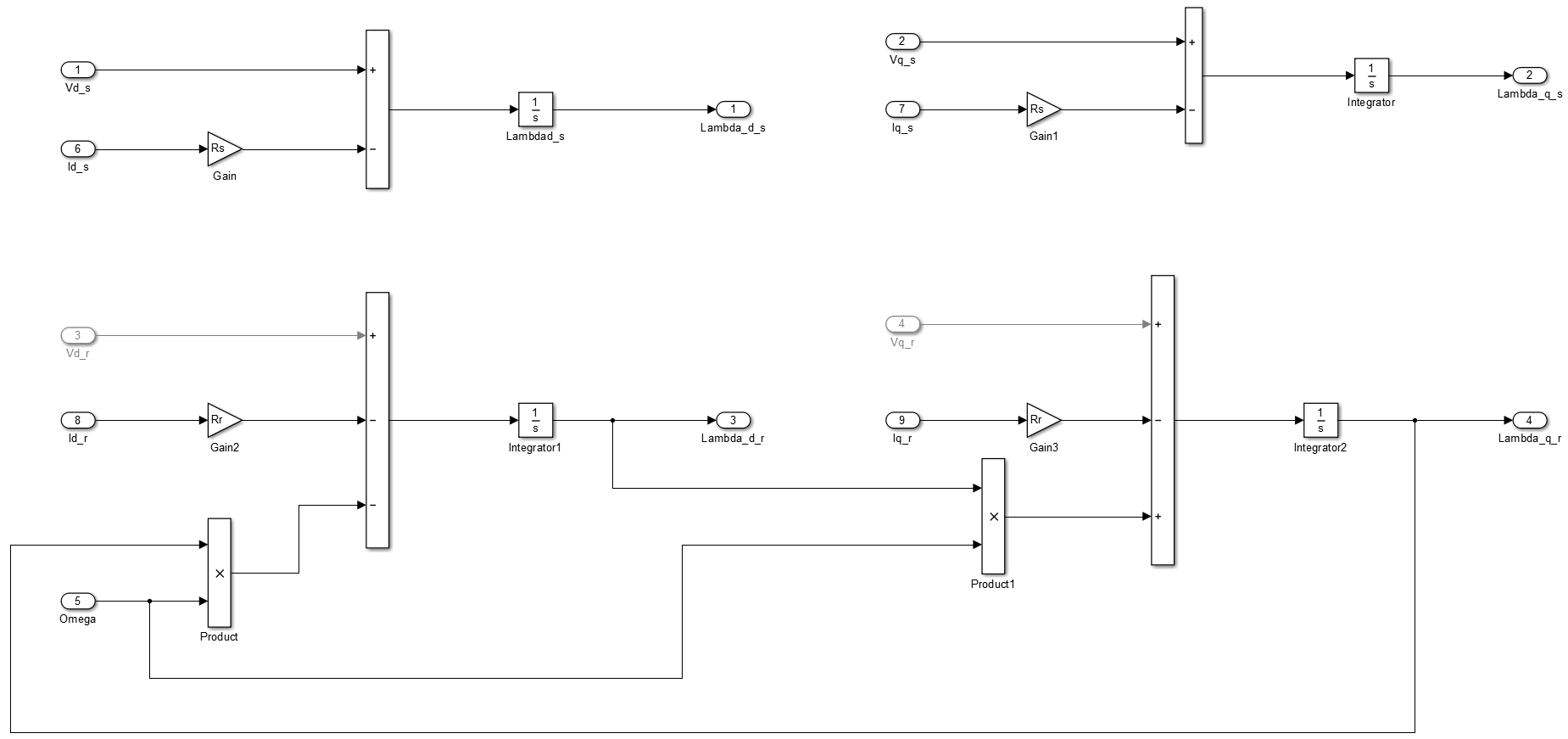


Figure 48 - Calculation of rotor and stator flux vectors. The blue block when viewed from the top level.

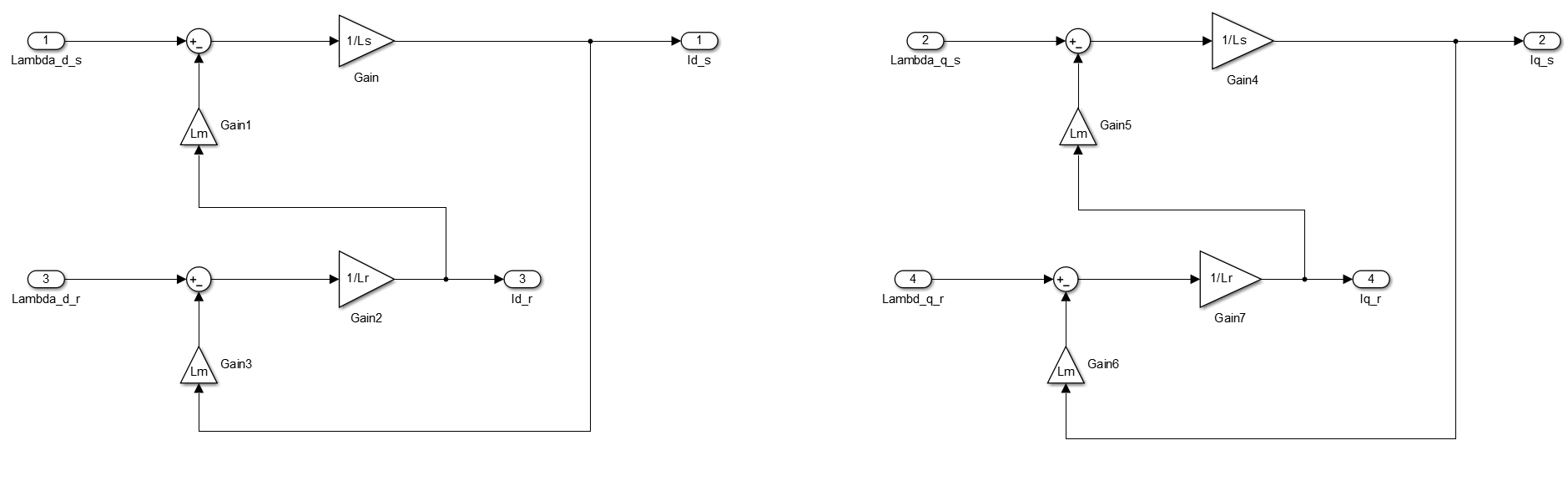


Figure 49 - Calculation of rotor and stator currents. The yellow block when viewed from the top level.

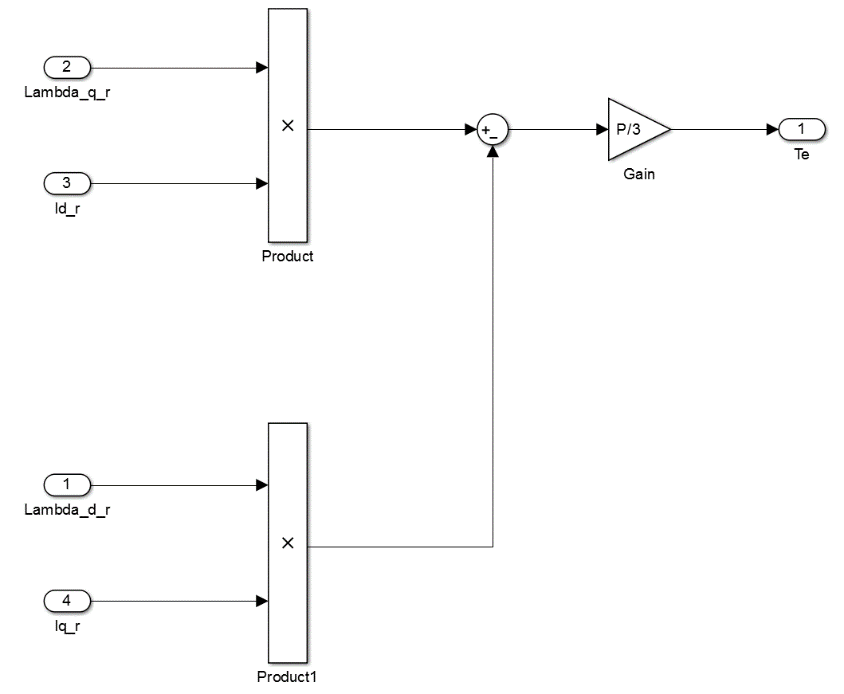


Figure 50 - Calculation of Electromagnetic torque from the motor. The light green block when viewed from the top level.

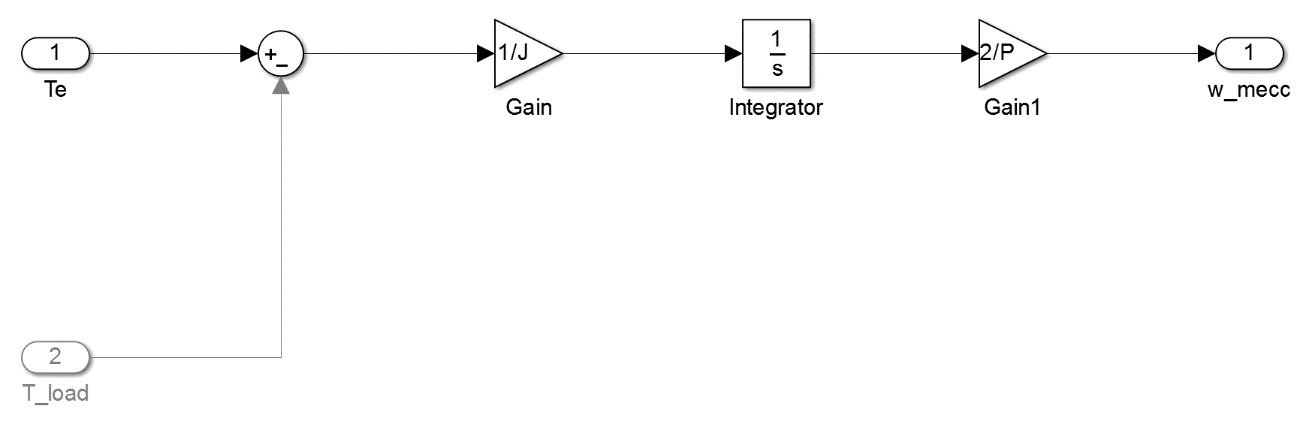


Figure 51 - Calculation of motor mechanical speed using a varying load torque. The orange block when viewed from the top level.

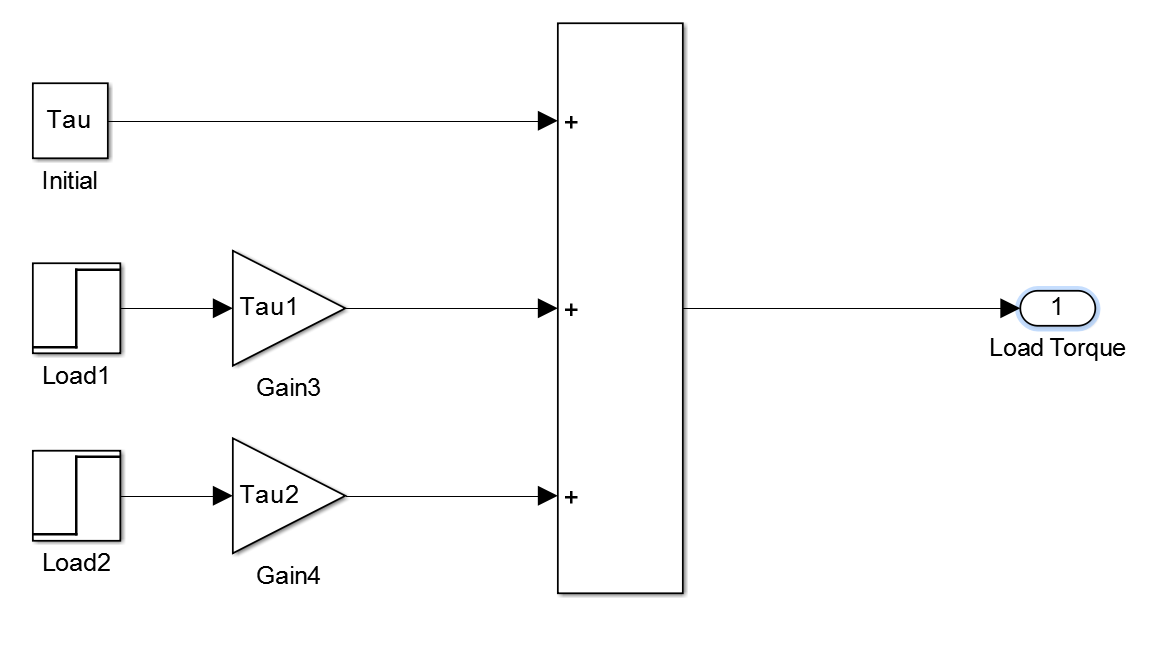


Figure 52 - Varying load torque generation. The white block when viewed from the top level.

# Appendix F – Circuit Schematics

## Input Transformers, Synchronous Input Bridge Rectifier, and DC Link

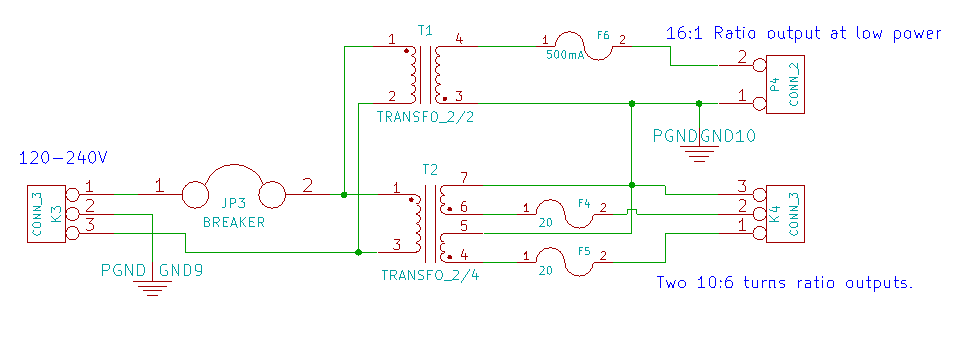


Figure 53 - The input transformers. Connectors go to the rectifiers.

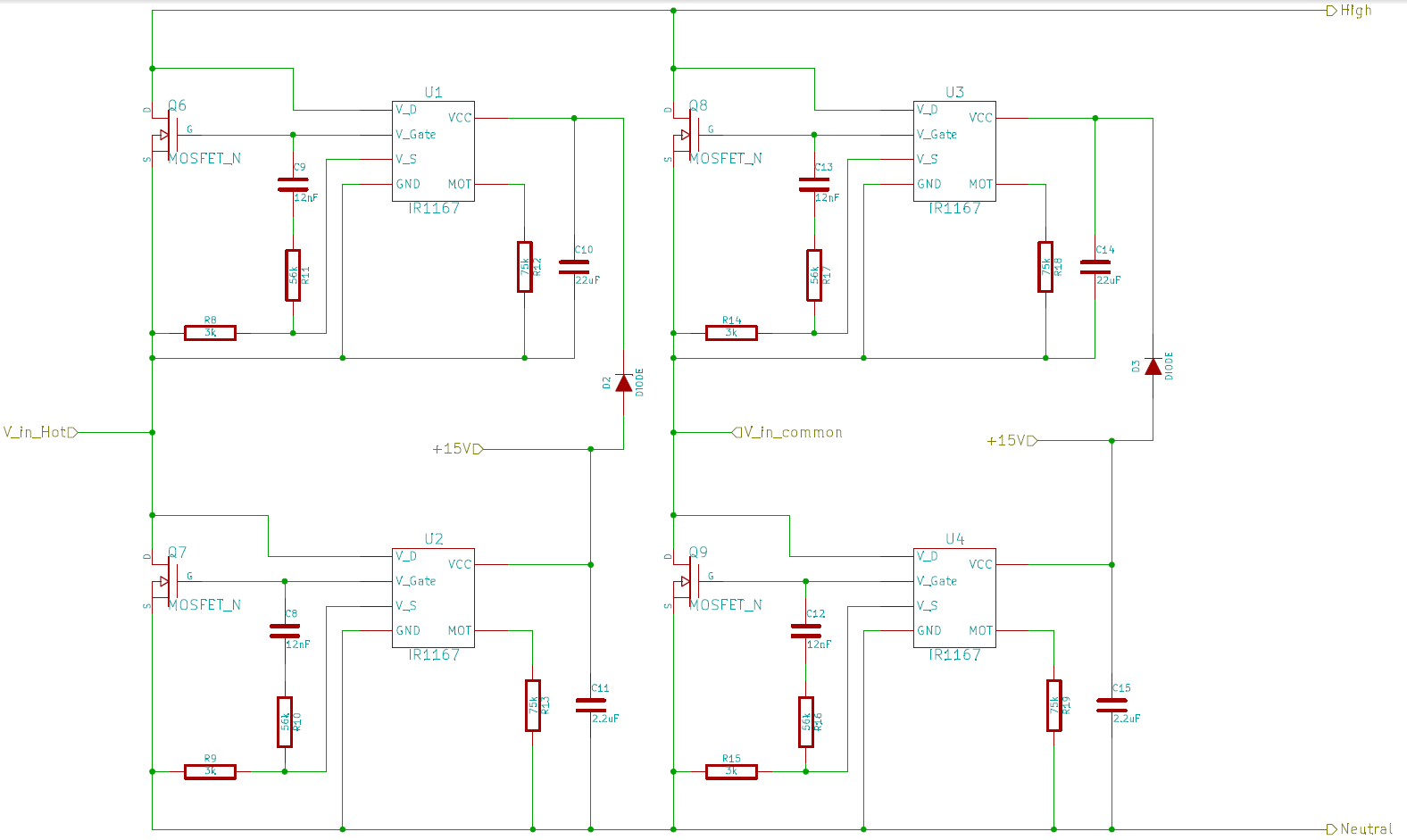


Figure 54 - Synchronous Bridge one bridge per DC link rail.

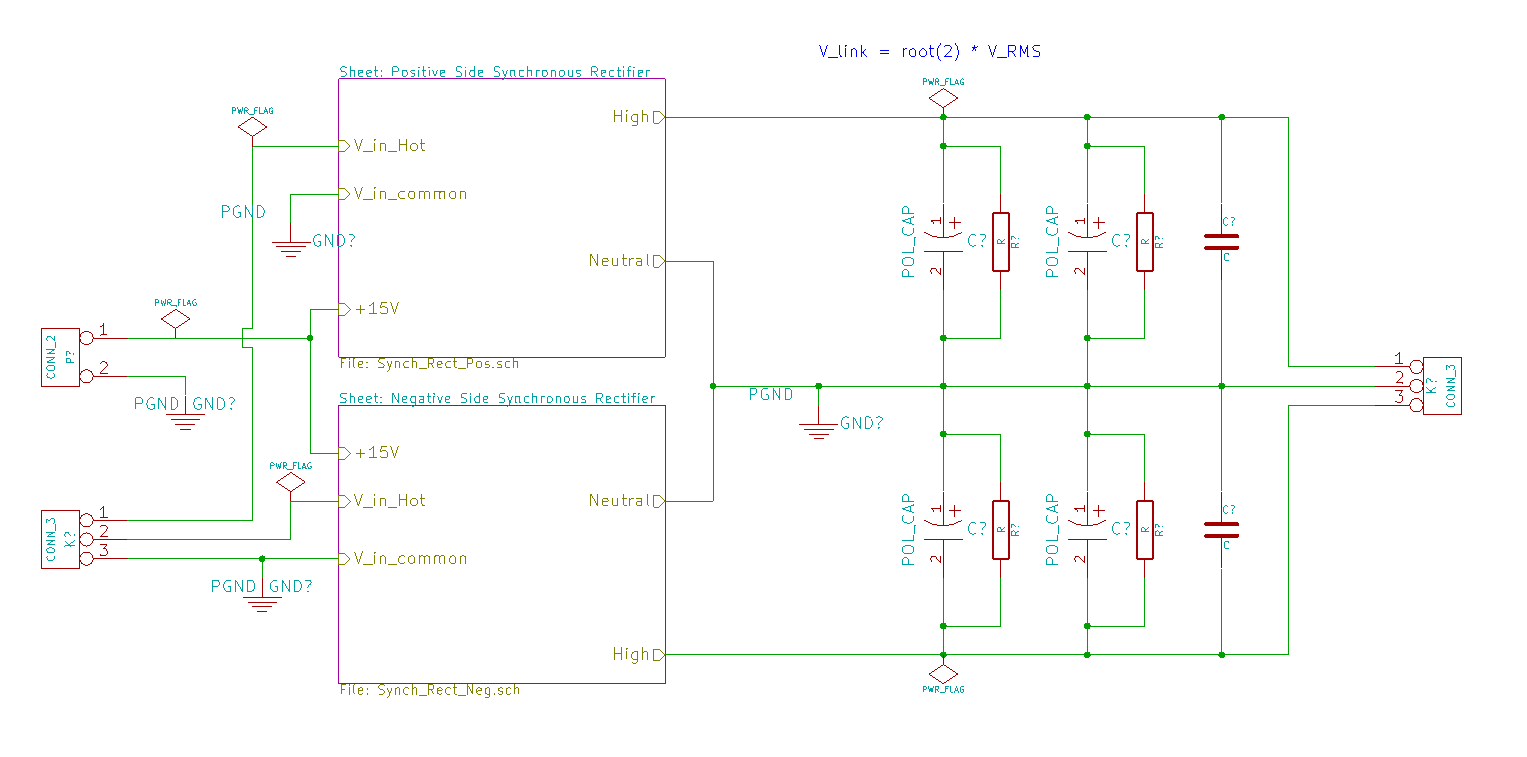


Figure 55 - DC Link formation. Each block contains one bridge rectifier.

## Low Voltage DC Rectifier and Rail Regulation

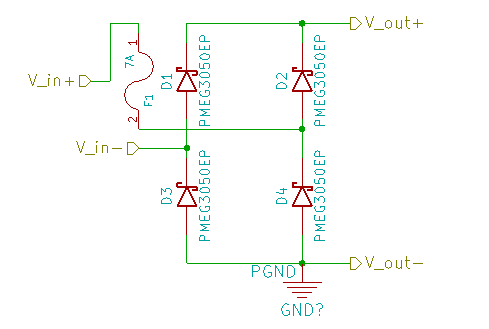


Figure 56 - Low voltage Schottky bridge.

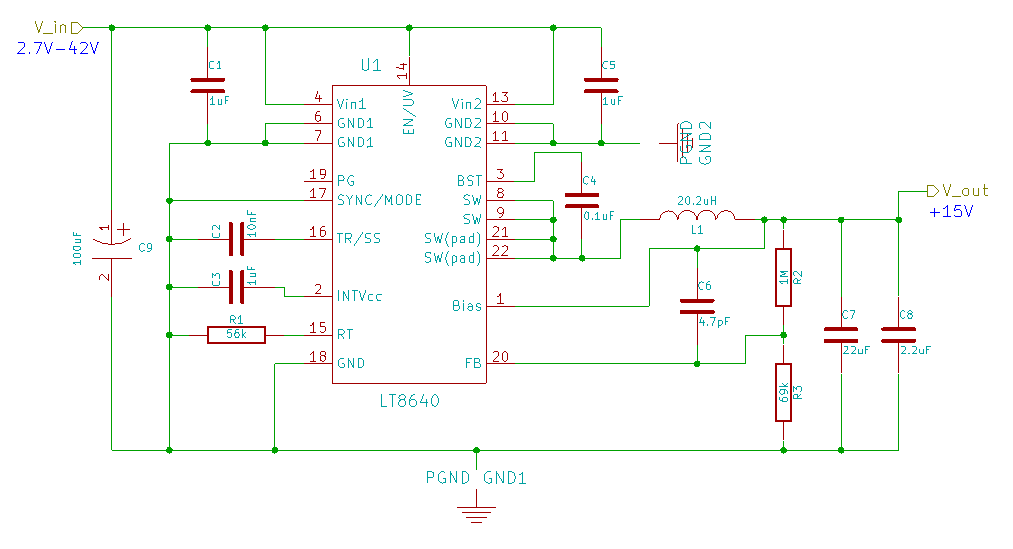


Figure 57 - 15V DC regulator.

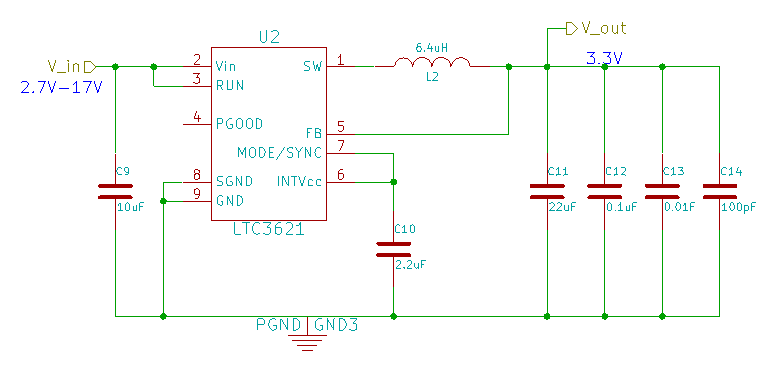


Figure 58 - 3.3V DC regulator.

## Power Inverter

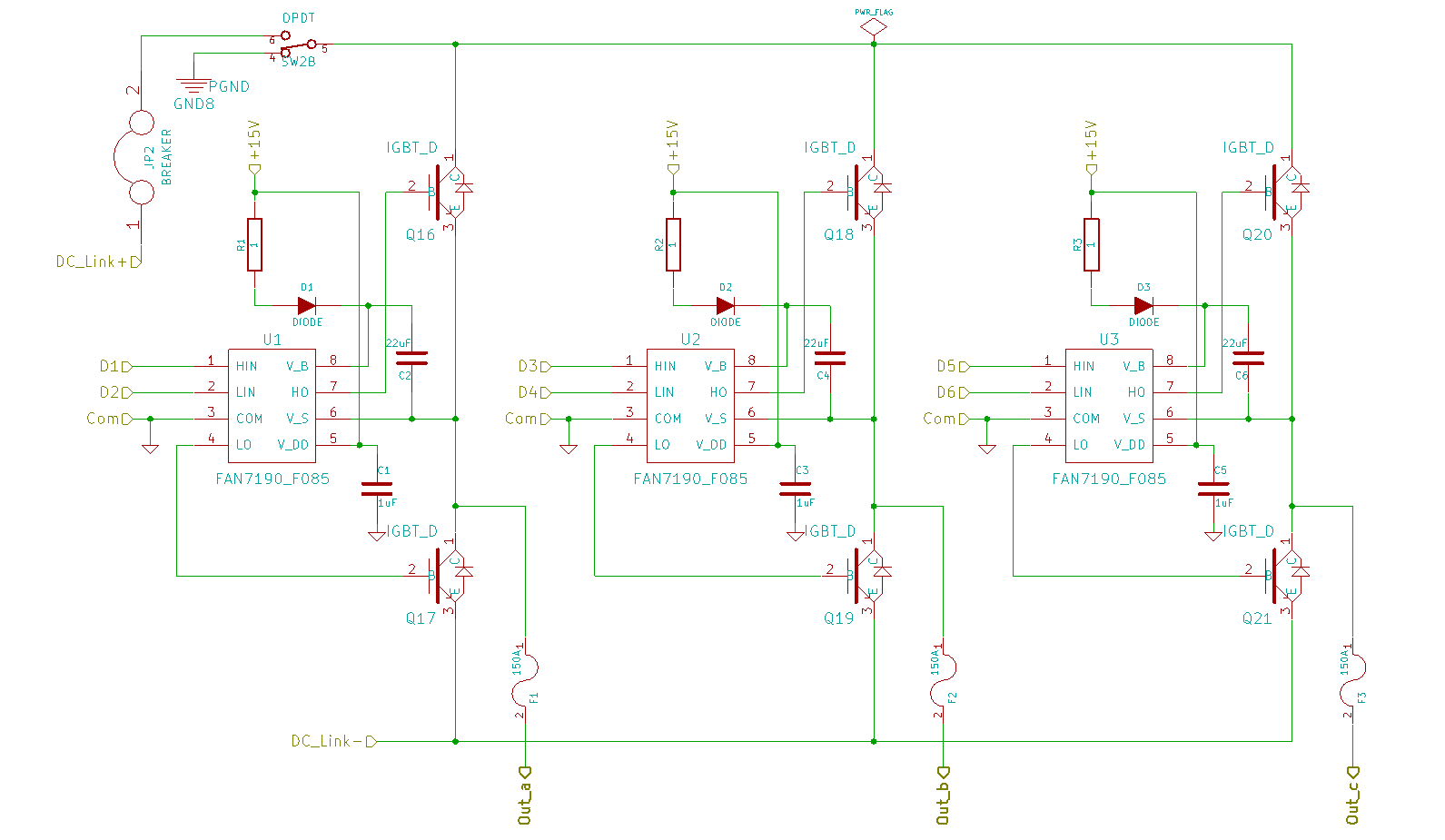


Figure 59 - The power inverter.

## MSP430/Piccolo Interface and Protection Circuitry

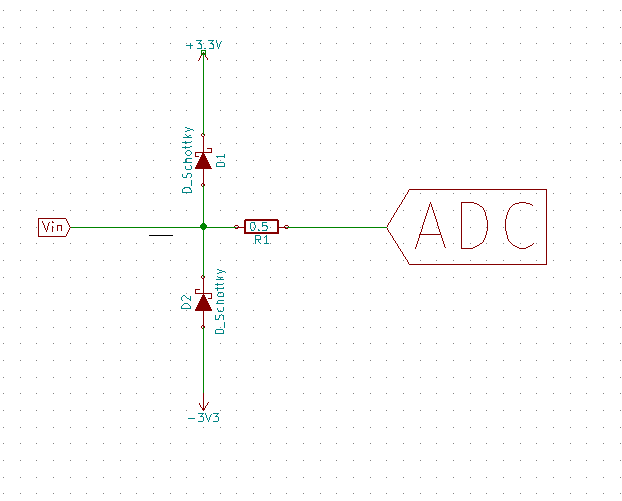


Figure 60 - ADC clamping input.

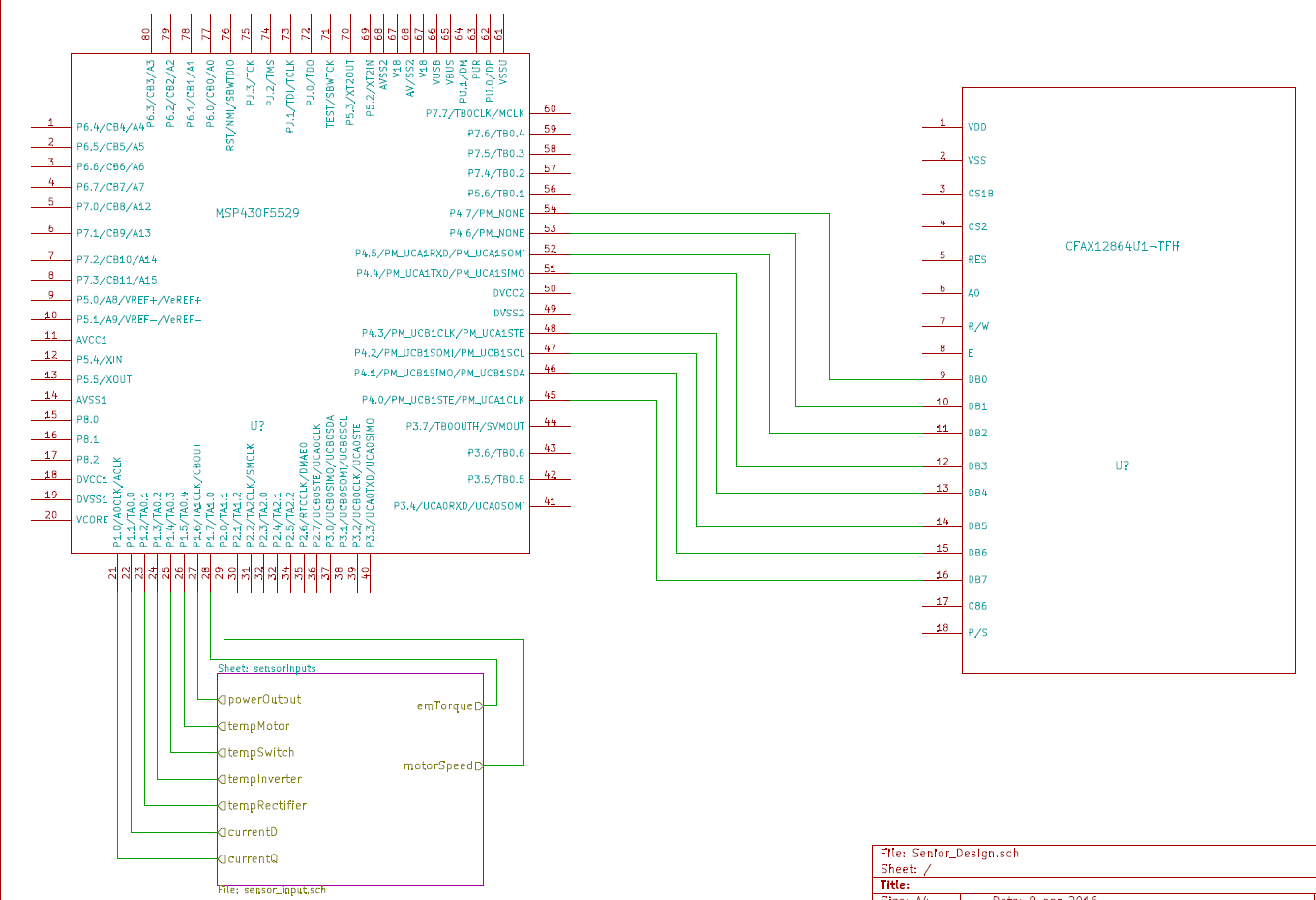


Figure 61 - MSP430 and LCD display interface.

## Thermocouple Interface

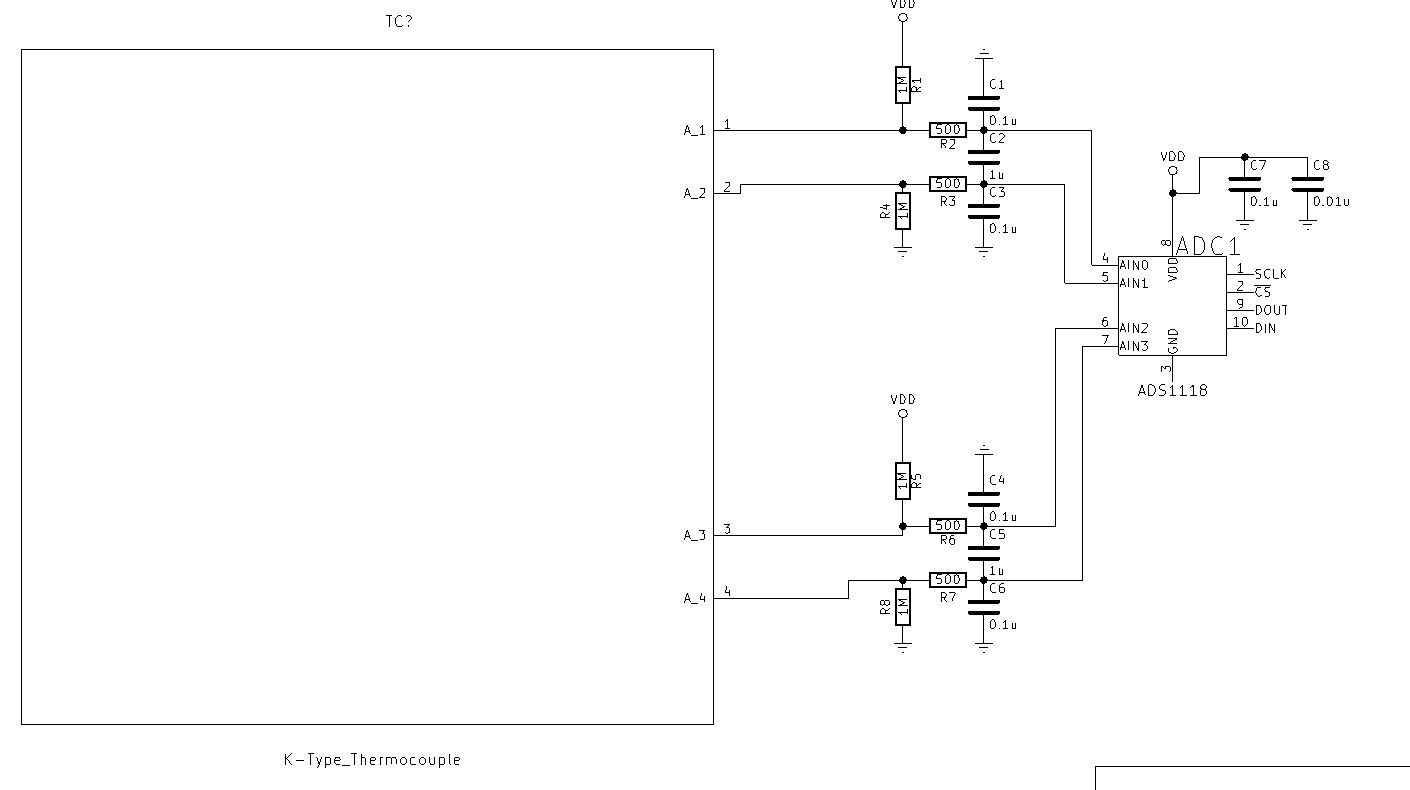


Figure 62 - Thermocouple sensing design.

1. All generators identical, 120° out of phase, purely sinusoidal voltage and current, symmetrical loading on all phases and symmetrical ground impedance. [↑](#footnote-ref-2)