Use of the IGBT Module in the Active Region to Design a High Current Active Filter

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Use of the IGBT Module in the Active Region to Design a High Current Active Filter

A thesis submitted in partial fulfillment of the requirements for the degree of Master of Sciences in Electrical Engineering

by

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Abstract

Particle accelerators require high-precision magnetic fields on the order or 100ppm or less. This implies that the precision of the associated electrical current in the electromagnet that generates these fields should be smaller than 100ppm. However, conventional switching power supplies cannot offer this precision due to the frequency limitation of the switches. This research considers the use of power electronics devices operating in a linear as an alternative solution to meet the requirements of particle accelerator electromagnets.

This thesis presents the study of an insulated-gate bipolar transistor (IGBT) driver using a new control method that linearizes the IGBT’s collector-emitter voltage ($V_{ce}$) gain. This allows the use of the IGBT in its linear region. This enables the implementation of an active filter as part of the superconducting magnet test facilities within the Superconducting Magnet Division (SMD) at Brookhaven National Laboratory (BNL). The IGBT driver and active filter incorporate a dual feedback loop topology to control IGBT modules and achieves improved capability in achieving precise control of electromagnet currents. This approach allows the use of commercially available IGBT devices that are normally optimized to work as switches.

The design, simulation, prototyping, and test results of an IGBT driver using the proposed dual-loop feedback approach is performed using MITSUBISHI CM1000-24H IGBTs, LabVIEW software and National Instruments Compact RIO hardware. Experimental results are presented that confirm the effectiveness of the proposed method.
Acknowledgments

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Dedication

To my Lord Jesus Christ, who has been with me throughout this journey.

To my parents in heaven, Jorge and Rosario.

To the women in my life, my wife Erika, my daughters Chris and Lauren, my sisters Nelly and Susana, and my nieces Sara and Sandra.
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Chapter 1

Superconducting Magnet Test Facility

In this section a brief introduction to superconductivity, superconducting and conventional magnets and the description of a Superconducting Magnet Test Facility is presented.

1.1 Overview

At the end of the 19th century, scientists in Europe liquified Helium and were able to get temperatures of about 4° Kelvin (K), or -269° Celsius (C). They found that the resistance of some materials vanishes at a temperature approaching absolute zero. At these temperatures, materials show zero resistance and is referred to as the phenomenon of superconductivity. Recent discoveries have shown superconductivity of new materials with a temperature hundreds of degrees K above zero provided the opportunity to obtain high magnetic fields using small magnets with moderate sized power supplies.

1.2 Superconducting Magnet Division at Brookhaven National Laboratory

The Superconducting Magnet Division (SMD) of Brookhaven National Laboratory (BNL) researches and develops new technologies to improve the manufacturing and testing of superconducting magnets and devices. Its capabilities are:

- Magnetic structure design.
- Superconducting magnet fabrication.
- Vertical and horizontal cryogenic testing.
- Magnetic field measurements.
1.3 Conventional Magnets

Conventional magnets are built using windings surrounded by a ferromagnetic core. When an electrical current run through the winding, a magnetic field is generated. To controls the intensity of the magnetic field, the current and/or the number of windings is changed. The maximum magnetic flux density limit for conventional magnets is about 1.6 Tesla. To generate these magnetic fields, conventional magnets need current of several thousand amperes. Cooling down the cables carrying these large currents requires wide diameter cables with liquid cooling. This increases costs while resulting in larger overall size and increased complexity.

1.4 Superconducting Magnets

Superconducting magnets are made of windings of superconducting cables cooled with liquid helium. This reduces the temperature to 4° K such that the cable’s resistance is zero. Consequently, the cable can carry much larger currents and thereby achieve magnetic fields higher than 1.6 Tesla. Most superconducting magnet coils are made of a niobium-titanium (Nb-Ti) alloy embedded in a copper matrix.

After the magnet is energized, it achieves a “persistent mode” of operation. Because it does not show resistance, there is no electrical energy loss in the magnetic field. If the power supply is removed and the magnet is shorted, the magnet windings will continue to conduct the electrical current without reduction.

The operation of the superconducting magnet is as follows: a persistent mode switch is located across the terminals of the magnet, and the switch either short-circuits the magnet terminals or places a piece of superconducting cable warmed with a heater to bring it to a resistive state. At the resistive state, the current going through the cable generates a voltage that charge the magnet. When the magnet reaches the desired field, the heater is turned off and the switch is short-circuited
again. The resistance of the magnet is so small that its time constant L/R is very long. It can operate for days, even months, with a constant field. The maximum field strength is limited by the point where the magnet loses its superconducting state.

1.5 Power Supplies

Power supplies are used in the SMD for powering the electromagnets. AC-DC conversion is done using silicon-controlled rectifiers (SCRs). A diagram of a generic power supply is shown in Figure 1.5-1.

![Figure 1.5-1: Diagram of a generic power supply.]

Most of the power supplies in accelerator facilities use 12-pulse rectifiers. Twelve pulses are achieved by connecting in series the output of the rectifiers of two six-pole rectifiers. These rectifiers are powered by a phase-shifting transformer, with two secondary windings with a phase shift of 30 degrees.

Increasing the number of pulses tends to decrease voltage ripple and results in reduced line current harmonic distortion. Using complex and advanced transformer winding configurations can get 24 and 30 pulses however this becomes less practical and more expensive. For example, a 24-pulse rectifier requires several transformers and more intensive maintenance requirements. The collider accelerator in BNL employs one of them as part of the booster accelerator and magnet power supply system.
1.6 Vertical Test Cryostat Test Facility

At SMD, the hardware where the magnets are tested at low temperatures is referred as the test cryostat. The test cryostat is a vertical inner vessel inside of a dewar, (i.e., a vacuum flask) with a 4.5° K heat shield between them. The vessel has two plates; the one that covers the top is called the “top plate,” and the second located inside the vessel is called the “lambda plate.” Two areas are produced by the Lambda plate; the top is kept at 4.5° K and the bottom at 1.9° K.

In addition to the gas connections to control the temperature and pressure, there are several electrical connections for power and measurement. All are properly suitable to hold the transition from room temperature and preserve it at a very low temperature inside of the cryostat. This is called the header system. Figure 1.6-2 show an AutoCAD model of a Top Plate and Header System.
1.7 Superconducting Magnet Test Facility

The superconducting magnet test facility consists of the power supplies, filter, energy extraction, control systems, free-wheel diodes, and the cryostat. Two or three power supplies can be connected in parallel to get up to 30 kA. Very large, water-cooled cables are used to connect the power supplies. An extra filter is used to decrease the current ripple, and a fast energy extraction (EE) system is used in each power supply for quench protection. The magnet or device under test (DUT) is attached to the lambda plate and inserted into the cryostat. A data acquisition system uses a digital LabView control is operated from the control room. Figure 1.7-1 shows a Diagram of a Superconducting Magnet Test Facility.
Figure 1.7-1: Simplified diagram of a superconducting magnet test facility [6].
Chapter 2

Quench Phenomenon

2.1 Quench Phenomenon

A quench phenomenon usually happens in the highest field area of the electromagnet when a small part of the superconducting material enters the normal state. The superconducting cable recovers its resistance, and because a high current is going through it, that area of the magnet overheats [13]. By conduction, surrounding areas are warmed and recover their normal state. This process continues until all the energy storage in the magnet is fully discharged.

The quench is generated when:

- There is rapid increasing in the current density and field.
- The rate of change of the magnetic field becomes too high.

The quench event happens so fast that the voltage generated in the area may be high in kilovolts, but current is stopped as soon as voltage is about 100mV. If insulation is not adequate, then arcing will occur between winding layers. During the event, the magnet is exposed to high voltage, high temperature, and high forces, any of which can damage the magnet. One way to protect the magnet from catastrophic failure in the localized area where the quench originated is to warm the whole magnet, so the energy is dissipated throughout all the magnets. For this reason, heaters are installed in the superconducting magnets.

2.2 Superconducting Cables

Resistivity of superconducting materials at room temperature are in three orders of magnitude higher than copper and aluminum. For this reason, superconducting wires are enclosed in a copper tubing. The combination of copper tubing with superconducting material is called filament. The filament size in the superconducting wire used in the RHIC and SSC dipole magnet
design is 6 μm [12]. The copper tubing gives way to transport electricity and heat at room temperatures is called stabilizer. At low temperatures, this is the heat dissipation capacity of the cable. The amount of copper in the cable is usually more than the amount of superconductor. The copper ratio to the superconductor ratio of the cable changes according to the location of the cable on the magnet.

The cables are submerged in helium, which dissipates the heat and keeps the temperature at 4.2° K. Rutherford type of cable is wide and made of several wires (strands) twisted together in a spiral shape. To counteract skin effects and the crowding of electrical current towards the conductor surface, Rutherford cable is used in most large-scale production of accelerator magnets [12].

Figure 2.2-1: Rutherford cable [14].
2.3 Stability

The stability against temperature disturbances depends on the cooling and heat dissipation capacity of the cable. If the area that enters in a normal state or the temperature disturbance is small, then the heat dissipation capacity of the cable and/or the heat flow into the helium neglect the disturbance and keep the cable stable. The heat flow into the helium is neglected because the surface is small.

2.4 Quench Detection

Simplest way to detect a quench event is to measure voltage difference between two halves of the magnet coil and compare it with the threshold values. Another way to detect a quench is measuring the propagation velocity of the quench using pickup coils as voltage tap replacements. The distance between the pickup coils is known and times of the arrival of the quench front can be detected [12]. Figure 2.4-1 shows a typical example where the voltage rises linearly up to 25° K when the normal zone expands between two voltage taps or coils. The voltage then decays, and the velocity in this case is about 100m/s.

![Figure 2.4-1: Measurement of the voltage drops during a quench as a function of time [14].](image-url)
2.5 Magnet Protection

The basic procedure to protect a magnet from a quench event is to disconnect the power supply and dissipate a large fraction of the energy storage in the magnet using one of the methods shown in Figure 2.5-1.

![Figure 2.5-1: Quench protection methods.]

- **Self-protection (a)**– Once the quench is detected, the power supply is turned off, the switch is turned on, and the energy is dissipated in the coil.

- **Bypass elements (BP) (b)**– Once the quench is detected, the power supply is turned off, and the energy is forced to cross through bypass elements like resistors, single diodes, back-to-back diodes, or more complex protection schemes [11].

- **Energy extraction (EE)(c)** – Once the quench is detected, the power supply is turned off, and a switch is opened forcing the current to go through an external resistance where the energy is dissipated. This is the method used at BNL.

- **Active heating (quench heaters, CLIQ)(d)** – Once the quench is detected, the power supply is turned off, and the heater is turned on to warm the magnet. The magnet goes to normal state and the energy is discharged through all the magnet area instead of one small, localized spot that would otherwise cause permanent damage.
Chapter 3
Closed loop Driver to linearize the $V_{ce}$ response

3.1 Overview

Because of equipment aging associated with existing installed superconducting magnets, there are increasing demands for maintenance and repairs. To overcome the limitations and reliability concerns of existing equipment, this research seeks to develop an efficient, precise and nearly maintenance free active filter that can be fabricated from a minimal number of commercially available high-current IGBT modules.

3.2 Research Objective

The development of an efficient precise and almost maintenance free active filter based in few units of high current IGBTs that replace the capacitors banks. It should include leading edge technology to achieve superior characteristics compare to existing equipment.

3.3 Limitations of Prior Art

Conventional topologies using pulsed-width modulate switched-mode convertors require switching frequencies about near 200 kHz to get high precision dc currents needed for superconducting magnet applications. Presently, there is not devices capable for high current levels needed for particle collider magnet applications that can operate with switching times needed for 200 kHz operation[2]. The switching frequency suggested by the device manufacturers depends on the maximum heat dissipation of the IGBT. It is found that state-of-the-art IGBTs can hard-switch above 100 kHz if the current is reduced from its rate value[16]. However, particle collider superconducting magnets applications require full rated current. This will not be achieved if switching conditions limit its safe operating area to a corresponding lower switching frequency.
3.4 Proposed Solution Topology

This research considers the implementation of an IGBT driver that linearizes the IGBT $V_{ce}$ response so that a high current active filter can be implemented using commercially available IGBT modules working in the active region. Since the IGBT $V_{ce}$ is linearized, the resulting power supply can be controlled in a manner with very low ripple and precise current control.

The operation of IGBTs in the active region is unusual because most commercial IGBTs are optimized to work in the saturation region. The active region of IGBTs is characterized by small changes in the voltage gate-emitter ($V_{ge}$) produces abrupt changes of the voltage collector-emitter ($V_{ce}$), which produces similar changes of the collector current. A technique to use IGBTs in the active region is applying a $V_{ce}$ feedback loop to the IGBT gate. This voltage feedback circuit implemented at the driver keeps the static gain quasi-constant and helps with the stabilization of the IGBT dynamics.

This chapter studies the performance of the IGBT with its driver configuration in open-loop and then develop for closed-loop operation using $V_{ce}$ as feedback. Experimental data collected in open loop shows that the response $V_{ce} = f(V_{ge})$ is logarithmic. Design a IGBT controller for this type of response is challenging and often requires the use of feed-forward methods. When a feedback loop using the $V_{ce}$ is implemented in the IGBT driver, the response $V_{ce} = f(V_{ge})$ becomes nearly linear will be referred to as quasi-linearization of the $V_{ce}$. Afterwards, an outer-loop control design for a linear response in $V_{ce}$ is straight forward. In addition, this chapter shows the response $V_{ce} = f(V_{ge})$ in open and closed loop from experimental data and then the concept of an active filter to control the current for a high current magnet (RL circuit) is shown.
3.5 Advantages of using IGBTs in their active region.

Following are advantages of using IGBTs in their active region:

- They offer high-input impedance of the gate and the ability to control the IGBT from the gate drive.
- Modern IGBTs can be used in their active region in areas close to the limits of their safe operation area up to 10 us.
- In series connections of IGBTs, balancing voltage sharing is necessary, and having an IGBT with is collector-emitter voltage stable is beneficial to reliable operation.
- A single IGBT module can replace a bank of dozens of bipolar junction transistors.
- There is an increased bandwidth in controlling magnet current.

3.6 The IGBT Model

The device selected for the active filter and the fast-high current switch is the MITSUBISHI CM1000-24H module. To study the performance of the IGBT in the active filter, the model of the IGBT is needed. The model shown in Fig 3.6-1 has been considered where parasitic inductances are neglected. In the model:

- $C_{ge}$ and $C_{ce}$ vary with the IGBT collector-emitter voltage [9].
- $C_{ge}$ and $C_{gc}$ vary with the IGBT current [9].

![Figure 3.6-1: Linear equivalent IGBT model [10].](image)
3.7 Behavior of IGBT

- When IGBT is transitioning to a conducting on-state (ON), $V_{ge}$ is maximum, $V_{ce}$ is being reduced. The gate circuit formed by series resistance $R_{gin}$ and capacitance $C_{ge}$ results in band width limitations gate-emitter voltage $V_{ge}$. As a result, any signal through the gate is delayed by capacitance $C_{ge}$ (Miller Effect).
- During the change to the off-state (OFF), as the $V_{ce}$ rises, $C_{ge}$ is low and the circuit formed by $V_{gin}$ and $C_{ge}$ shows a much more rapid response and allows $V_{ce}$ to transition quickly. The turn-off characteristics are shown in Fig. 3.7-1.

Figure 3.7-1: Typical switching off waveforms for an IGBT (at 25° C) [3].

3.8 Design of a Closed-Loop IGBT Driver of High-Power IGBTs

To control the collector-emitter voltage of the IGBT, a feedback technique is implemented to linearize its behavior. The analysis starts from the equivalent circuit shown in Figure 3.8-1.
In this Figure, $V_{ge}$ is the voltage applied to the gate-emitter of the IGBT. $K_a$ is the effective gain of the IGBT when operating in the circuit for specified supply voltage.

**Inner-Loop Quasi-Linearization**

![Figure 3.8-2: Inner-Loop quasi-linearization.](image)

At operating condition of the magnet IGBT controlled supply. Using incremental voltage changes at the nominal operating conditions of $\text{Supply} = 20\text{V}$, then it is found that in open loop

$$K_a \triangleq \frac{\Delta V_{ce}}{\Delta V_{ge}} \approx \frac{\partial V_{ce}}{\partial V_{ge}} \left( 88.9 \frac{V}{V} \right)$$

(3.8-1)

$$\frac{\partial f(x)}{\partial V_{ge}} \approx K_a$$

(3.8-2)

![Figure 3.8-3: Open-loop $K_a$ relationship.](image)

Using feedback, the relationship can be made more robust using the following configuration shown in Fig. 3.8-4 with incremental values with respect to nominal conditions $K_a = \bar{K} + \Delta K_a$:

![Figure 3.8-4: $K_a=\bar{K}_a+\Delta K_a$ relationship.](image)
In using feedback, the linearization of $\Delta K_a$ represents the variability of $K_a$ due to the changes in operating conditions. The resulting closed-loop relationship becomes

$$\frac{V_{ce}}{V_{ce}^*} = \frac{K_p(\bar{K}_a \Delta K_a)}{1 + K_p(\bar{K}_a \Delta K_a)} = T_{pk}(\Delta K_a) \quad (3.8-3)$$

The sensitivity of $T_{pk}$ with respect to $\Delta K_a$ can be found to be

$$\sum_{\Delta K_a} = \frac{\partial T_{pk}}{\partial K_a} \frac{\Delta K_a}{T_{pk}} \quad (3.8-4)$$

Therefore, the sensitivity to $\Delta K_a$ can be reduced compared to the open-loop condition.

Using the experimentally determined value of $\bar{K}_a = 88.9$, then

$$T_{pk} = \frac{K_p \cdot 88.9}{1 + 88.9 K_p} \quad (3.8-5)$$

For example, with $K_p = 1.375$, then $T_{pk} = 0.9919$ and therefore provides robust tracking of $V_{ce}$.

- **Outer-loop magnet current control**

The magnet current will be controlled by $V_{ce}$ with feedback of magnet current. For the linearized system, using Kirchhoff’s voltage law gives

$$R_{in} + L \frac{din}{dt} = V_{ce} \quad (3.8-6)$$

In Laplace Domain,

$$(R + Ls)Im(s) = V_{ce}(s) \quad (3.8-7)$$

Or as a transfer function,

$$\frac{V_{ce}}{Im} = \frac{1}{Ls + R} \quad (3.8-8)$$
However, there are known time delays and quantization errors associated with National Instrument Compact Rio hardware as well as other sensor and software interfaces. These effects will be accounted for with a Padé approximation where $T_s$ is the digital time delay:

$$D(s) = \frac{-T_s s + 1}{T_s s^2 + 1}$$ (3.9-9)

The controller is selected to be a proportional-integral type. This provides robust tracking performance. The Root Locus Function can be found from the closed-loop magnet current system shown in Figure 3.8-6

For selecting $K_p$, first assume $K_i=0$ and the Root Locus Function is found from the characteristic equation:
\[ \Delta(s) = 1 + \left( \frac{K_p s + K_i}{s} \right) \left( \frac{1}{s + R \frac{L}{L}} \right) \left( \frac{-T_s s + 1}{T_s s + 1} \right) = 0 \] (3.8-10)

The root locus is plotted using MATLAB. This shows a large range of stable value with poles on the negative real axis. A value of \( K_p = 1 \) is selected since this provides a pole at \(-331 \text{ r/s} \) and \(-9.9 \times 10^4 \text{ r/s} \) which provides rapid decay of the transients.

Figure 3.8-7: Root Locus for selecting \( K_p \), assuming \( K_i = 0 \).

The integral coefficient is determined through a new root locus with \( K_p \) fixed. The revised root locus with respect to \( K_i \) is first formulated by:

\[ s(s + \frac{R}{L}) \left( \frac{T_s}{2} s + 1 \right) \left( \frac{K_i}{L} \right) \left( \frac{-T_s}{2} s + 1 \right) + K_i \left( \frac{1}{L} \right) \left( \frac{-T_s}{2} s + 1 \right) = 0 \] (3.8-11)

Which is put in standard form:

\[ 1 + K_i \frac{\frac{1}{L} \left( \frac{T_s}{2} s + 1 \right)}{\frac{T_s}{2} s^3 + \left( \frac{R}{L} + \frac{K_i T_s}{2L} \right) s^2 + \left( \frac{R}{L} \right) s + \left( \frac{K_i}{L} s^1 \frac{K_i}{L} \right)} = 0 \] (3.8-12)
The resulting root locus will have unstable values for $K_i$ depending on $T_s$. To avoid unstable conditions with an unknown $T_s$, a relatively small value of $K_i = 0.001$ is selected.

To study the behavior of the IGBT using the *PSIM Capture Tool* and the IGBT’s data sheet, a model of the Mitsubishi CM1000-24H IGBT was inserted in the following PSIM circuit. In the simulation, the reference signal is compared with the collector-emitter voltage, which is the feedback term to control loop. The collector-emitter voltage tracks the reference. Figure 3.8-9 shows the IGBT with a feedback configuration.
3.9 Simulation

Figure 3.9-1 shows a Psim simulation of the above circuit. The IGBT is a Psim model, which is good approximation the model shown in Fig 3.6-1. The IGBT’s data was taken from its data sheet.
In the simulation, using trial and error method, the PI controller was configured with 

\[ K_p = 1.00 \text{ and } K_i = -0.001 \]

Figure 3.9-2 shows the waveforms obtained from the simulation, \( V_{CE} \) tracks exactly \( V_{ref} \) and achieves the objective of the IGBT closed loop driver. It is also observed how \( I_{ce} \) changes according \( V_{ce} \).

![Waveform Diagram](image)

**Figure 3.9-2: Psim simulation response waves of IGBT-magnet closed-loop driver.**

The PSIM results shows how the \( V_{ce} \) tracks very closely the reference. In addition, Figure 3.9-2 shows the tracking behavior of the collector current (which is the magnet current).
Chapter 4

Prototype Construction and Experimental Results

A prototype system was fabricated to linearize the $V_{ce}$ gain of Mitsubishi CM1000-24H IGBT and test that the concept of developing an active filter for a current power supply using a quasi-linearized IGBT with a closed-loop gate driver is feasible.

Since the IGBT response with changes of $V_{ce}$ to variation of $V_{ge}$, the objective of these experiments is to linearize the response $V_{ce}=f(V_{ge})$ when the current is at a constant nominal operating condition. First, the $V_{ce}=f(V_{ge})$ is plotted from experimental data with the following procedure:

1. The $V_{ce}=f(V_{ge})$ is plotted from experimental data of the IGBT driver load (open-circuit) in open-loop gate drive when the magnet current is constant.

2. The $V_{ce}=f(V_{ge-ref})$ is plotted from experimental data of the IGBT- driver-Load circuit in closed-loop with constant magnet current. In this test, the feedback loop is implemented in the driver using the $V_{ce}$ as feedback.

3. The proposed active filter using two PIs nested, the inner is $V_{ce}$ PI, the outer loop is collector current PI.

4.1 Prototype Experiments

• Linearization Process:

1. Show the $V_{ce}=f(V_{ge})$ characteristics when the current is constant plot of the Mitsubishi CM1000-24H IGBT in open-loop.
2. Show the $V_{ce} = f \ (V_{ge-ref})$ when the current is constant plot of Mitsubishi CM1000-24H IGBT in closed-loop using $V_{ce}$ as feedback.

3. Confirm that the active filter using an IGBT with linearized ($V_{ce}$) properly characterizes a quasi-linear device response.

- **Software tools:**
  - LabVIEW. – Provides a controller graphic programming software to build PI Controllers using its FPGA Section.
  - MATLAB Tool: System Identification Toolbox accept numeric arrays of the input and outputs signals and returns an estimated of the transfer function
  - MATLAB Tool: PID Turner Toolbox accept transfer function of the plant and returns $K_p$, $K_i$ and $K_d$ constants of a PID Controller

- **Hardware and equipment:**
  - Compact Rio (cRio) systems from National Instruments: Provides a development chassis, input and output modules that can be programmed in three platforms, host computer, real time processor and FPGA. It is associated with LabView product software.
  - High current resistor and inductance, 20VDC @ 150 Amps power supply, measuring equipment.

- **Procedure:**
  - Perform test and save the required signals in the scope memory.
  - Export the data to an Excel file and insert the data to MATLAB as numerical array.
  - Import the numerical array to the System Identification Toolbox, derive the transfer function.
• Insert the transfer function in PID Turner Toolbox and confirm the $K_p$ and $K_i$ match the root locus analytical results.

• The $K_p$ and $K_i$ are then used as initial point of the controller design and are revised with adjustments if needed using trial and error techniques.

**Work Bench Set Up**

![Figure 4.1.1: Photograph of prototype system used in laboratory experiments.](image)

The experiments were conducted in Superconducting Magnet Division of BNL. Figure 4.1-1 shows the prototype circuit and experimental set up:

1. cRio Chassis
2. Laptop
3. Power Supply
4. IGBT air cooled heat sink
5. AFG
6. Inductance
7. Power resistor
8. DCCT head
9. Private network hobs
PI Controllers were implemented using LabView, Figure 4.1-2 shows the Human Machine Interface, it shows 2 PI controllers, its controls, settings and bypasses.

The programing part of the PI Controllers are shown in Figure 4.1-3. This includes the FPGA programing with the pin icons representing the hardware inputs and outputs of the circuit.
This design features three inputs and one output. The inputs are the reference voltage, feedback of the collector current and $V_{CE}$. The output is the signal going to the gate driver.

**4.2 Experimental measurement of the IGBT $V_{ce} = f(V_{ge})$ at $I_c = K$ in open loop**

<table>
<thead>
<tr>
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<td>$V_{ce}[\text{V}]$</td>
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<tr>
<td>7.035</td>
<td>16.91</td>
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<td>7.2501</td>
<td>0.869</td>
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<table>
<thead>
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<td>$V_{ce}[\text{V}]$</td>
</tr>
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<td>7.452</td>
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</tr>
<tr>
<td>7.6</td>
<td>1.25</td>
</tr>
</tbody>
</table>

Figure 4.2-1: (Left) Circuit used to get $V_{ce} = V_{ge}$.

Figure 4.2-2: (Right) Collected data.

Figure 4.2-3: Plot of the IGBT $V_{ce} = f(V_{ge})$ at constant current in open loop.

Observations:
• A small change in $V_{ge}$ produce an exponential change in $V_{ce}$ response.

• The range of $V_{ge}$ is less than 0.40V. Two implications of 0.40V in this parameter are:
  
  ▪ The maximum amplitude of the signal is 0.40V: if the amplitude of the reference signal is greater than 0.40V, then it must be reduced to avoid loss of definition.
  
  ▪ Noise amplitude usually is greater than 0.40V: Noise might drive the IGBT to saturation and the collector current will be the maximum, generating a hazardous situation that might damage the IGBT.

Design a IGBT controller for this type of response is not straight forward, often requires the use of supplemental feedforwards controls or other modifications.

4.3 Experimental measurement of the IGBT $V_{ce} = f(V_{ge})$ at $I_c = K$ in close loop

Table 4.3.1: Experimental data for $V_{ge}$, $I_c$ and $V_{ce}$ when voltage source is 20V.

<table>
<thead>
<tr>
<th>$V_{ge}$ [V]</th>
<th>$I_c$ [A]</th>
<th>$V_{ce}$ [V]</th>
<th>$V_{ge}$ [V]</th>
<th>$I_c$ [A]</th>
<th>$V_{ce}$ [V]</th>
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<tr>
<td>6</td>
<td>0</td>
<td>20</td>
<td>7.35</td>
<td>15</td>
<td>9.96</td>
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<td>6.72</td>
<td>1</td>
<td>19.95</td>
<td>7.353</td>
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<td>9.39</td>
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<td>6.84</td>
<td>2</td>
<td>19.32</td>
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<td>17</td>
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<td>6.94</td>
<td>3</td>
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<td>8</td>
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<td>10.73</td>
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<td>1.068</td>
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The process of programing the $V_{ce}$ inner-loop PI controller start with requesting a $V_{ce}$ value from the reference followed by entering $K_p$ and $K_i$ constants in the PI controller and comparing the readings of $V_{ce}$ with the reference. Table 4.3-1 show the associate the values of $V_{ge}$, $I_c$ and
$V_{ce}$. Once the PI controller is programmed, then collect results with the voltage values of the source and the reference to obtain the data for $V_{ce} = f(V_{ce-ref})$, as shown in table 4.3-2. Figure 4.3-1 shows the circuit used to obtain $V_{ce} = f(V_{ce-ref})$ with driver in closed loop.

![Figure 4.3-1: Shows the circuit used to get $V_{ce} = f(V_{ge})$ driver in closed-loop.](image)

The circuit used to semi-linearize $V_{ce}$ response is in Figure 4.3-1, it has a PI controller to control $V_{CE}$ and the feedback from $V_{CE}$, to its output at the gate driver.

Procedure:

- Adjust the $V_{ce}$ PI using the procedure of point 4.1
- Record the results under Table 4.3.1. Select the $V_{ce-ref}$, change the Source voltage to obtain the correct table value.
- Select a new $V_{ce-ref}$ and repeat above step making sure the current remains constant.
- Repeat the above steps for each current value in the table.
• Table 4.3-2: Experimental data collected.

<table>
<thead>
<tr>
<th>Vge-ref (V)</th>
<th>Vce (V)</th>
<th>Vge-ref (V)</th>
<th>Vce (V)</th>
<th>Vge-ref (V)</th>
<th>Vce (V)</th>
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<tr>
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<td>3.75</td>
<td>-15.70</td>
<td>3.75</td>
<td>-16.11</td>
<td>4.03</td>
</tr>
</tbody>
</table>

The results shown in Figure 4.3-2 demonstrates that the $V_{ce}$ PI controller can obtain the desire quasi-linearized $V_{ce}$. This confirms one of the significant purposes of the research.

Observation:
A 15V change in $V_{ce-ref}$ produce a nearly linear change in $V_{ce}$ response.

The range of $V_{ce-ref}$ is about 15V, it is about 30 times larger than the driver in open loop.

This design approach of the IGBT controller for achieving a quasi-linear response is straightforward and is an improvement over previous design methods.

For the prototype system, the outer-loop PI controller coefficient were obtained as

$$K_p = -1.375 \text{ and } K_i = 0.001$$

### 4.4 Active Filter for a current Power Supply using IGBT with Linearize the $V_{ce}$

The objective of this section is to evaluate the concept of the active filter with $V_{ce}$ semi-linearized feedback meets the superconducting magnet requirements.

Figure 4.4-1: Circuit of the concept of active filter for a high current power supply.

The configuration shown in Figure 4.4-1 is the same used to generate the $V_{ce}$ quasi-linearized response except for the addition of the outer PI control loop with the objective of controlling the current. For this reason, the feedback comes from a sensor that measures the magnet.
current. The reference comes from the frequency generator $V_{ce-ref}$, and will be referred to as the outer-loop current PI controller. Scaling of the Current Pi Controller is given as:

- DCCT feedback: $10V=200A$
- Reference ($V_{ce-ref}$): $10V=200A$

After completing the design process for the PI controller coefficients, the active filter will be evaluated by requesting various reference waveforms and comparing with the actual measured response.

4.5 Responses of the Active Filter for waveforms

The performance of the Active filter was tested by inserting in the reference square, sine, ramp and noise signals. In the output measurements of the magnet load current, it was observed how well the filter reproduce the reference waveforms. The distortion in the shape of the waveform and any noise or oscillation effects in the signal were the main factors to estimating the system bandwidth. Finally, the behavior of the filter was analyzed when either one or both loops bypassed.

The following setting were used for the experimental test specifications.

- The scope screen displaying four traces:
  - Channel 1_Yellow trace showing the $V_{ge}$ signal.
  - Channel 2_Blue trace showing $V_{ce}$ signal.
  - Channel 3_Violet trace showing the Current reference signal.
  - Channel 4_Green Trace showing the Output current or Collector current.

- PI Settings:
  - Current PI (outer)
    - $K_p = -1$
- $K_i = -0.001$

Voltage collector-emitter (inner)

- $K_p = -1.375$
- $K_i = 0.001$

- Reference signal for all waveforms
  - Amplitude: 3.8 Vpp
  - Offset: 2.871 V

**Square Wave Reference Signal**

![Square Wave Reference Signal](image)

Figure 4.5-1: Response of prototype circuit for square wave reference.

In Figure 4.5-1 A square wave is requested and the output is its accurate reproduction. The rising and failing edges show the presence of the large inductance in the load. The output is acceptable for the superconducting magnet application.
Sine Wave Reference Signal

![Sine Wave Reference Signal](image)

Figure 4.5-2: Response of prototype circuit for sine wave reference.

In Figure 4.5-2, a sine wave is requested, the output is its accurate reproduction. The output is acceptable for the superconducting magnet application. $V_{ce}$ show a non-symmetric noise. This effect may problematic and further analysis to explain this effect is given in the following section.

Ramp Wave Reference Signal

![Ramp Wave Reference Signal](image)

Figure 4.5-3: Response of circuit for ramp wave reference.

In Figure 4.5-3, a ramp wave is requested with the output as an accurate reproduction. Overall, this is an acceptable result for the magnet application.
Noise Signal with DC offset

In Figure 4.5-4 a 3.8Vpp noise from 1Hz to 100kHz on top a 2.80 VDC signal is requested, the corresponding output is a pure 2.80VDC and the noise is eliminated. This confirms the effectiveness of the proposed two-loop active filter. Also noted is that $V_{ce}$ does not show any oscillation.

Sine Wave Reference Signal at Multiple Frequencies

The prototype circuit was evaluated at 10Hz, 20 Hz, and 25 Hz. The results are shown in the following sequence of Figures.

Figure 4.5-5: Prototype circuit output for 10 Hz sinusoidal reference.
In Figures 4.5-5 to 4.5-7, an assessment was made to determine an upper bound of the usable sinusoidal frequency range of the prototype filter. Based on these results, the conclusions are:

1) Usable frequency for sine wave is 15 Hz; the bandwidth for triangular wave is 7 Hz; and for rectangular wave it is 10 Hz.

2) A delay is evident with increasing frequency and the phase angle lag also increases with the frequency.

3) Small noise in bottom part of the output increases with frequency.
To further evaluate the performance of the prototype circuit, the behavior with either one or both loops bypassed is considered.

Figure 4.5-8: Behavior of the circuit with both loops working normally.

Figure 4.5-8 shows the baseline output both loops operating normally. Figure 4.6-9 shows the behavior of the circuit when the outer loop has been removed. The current response is noisy in the bottom part with 180° phase.

Figure 4.5-9: Behavior of the circuit with outer loop bypassed.
Figure 4.5-10 shows the response of the circuit when the inner loop is being bypassed. As expected the circuit is no longer functional.

Figure 4.5-10: Behavior of the circuit with inner loop bypassed.

Figure 4.5-11 shows the response of the circuit when both inner and outer loop are bypassed, the resulting in the reference signal going directly to the gate of the IGBT. The IGBT goes to saturation in the bottom and in the top of the signal.

Figure 4.5-1: Behavior of the circuit with both loops bypassed.
Chapter 5

Conclusions

The concept of developing an active filter for a current power supply using a quasi-linearized IGBT with closed-loop gate driver has been explained from a theoretical basis and confirmed with laboratory experiments of a prototype circuit. The following are conclusions drawn from the analysis and experimental results for a superconducting magnet power supply and current controller:

- When the IGBT driver is in open loop:
  - A small change in $V_{ge}$ produce an exponential change in $V_{ce}$ response.
  - The range of $V_{ge}$ is less than 0.40V
  - Design a IGBT controller for this type of response is difficult and sensitive to small variations in circuit parameters.

- When the IGBT driver is in closed loop with $V_{ce}$ feedback:
  - A small change in the $V_{se}$ produce a quasi-linear change in $V_{ce}$ response.
  - The range of $V_{ge}$ is increased to approximately 15V and is almost 30 times larger than the driver in open loop.

- Design a IGBT controller for the proposed quasi-linearized IGBT is procedurally straightforward.

- When an IGBT $V_{ce}$ is quasi-linearized, the current loop PI controller coefficients can be designed first for $K_p$ by setting $K_i = 0$. The integral term is then selected with the $K_p$ value at its initial value.

- The usable frequency for sinusoidal reference input is 20Hz, for triangular waveform it is 7 Hz and for rectangular waveform is 10Hz
The $V_{ce}$ signal shows a non-symmetric noise as noted in Figures 4.5-8 to 11. In order to explain this phenomena, detailed simulations were developed to provide a more detailed analysis of the behavior of the circuit. The asymmetric noise was found to be due to diode conduction that is induced by digital quantization (voltage reversal across the magnet inductance). That is, the analog-to-digital converter used to measure $V_{ce}$ will cause a step change in voltage at the gate. Since the IGBT is in its active range, this step change attempts an abrupt change in the collector current. Due to the large inductance of the superconducting magnet coil, this change in current causes a large induced $L \frac{dt}{dt}$ voltage causing the anti-parallel fly-back diode to become momentarily forward biased. When this occurs, $V_{ce}$ is pull up towards the supply voltage until the induced voltage is relieved. The Simulink model is shown in Figure 5.1. This effect can be minimized and effectively eliminated by using an analog-to-digital converter with higher resolution.

Figure 5.1: Simulink file used to reproduce the noise in the prototype circuit.
A negative point of using IGBTs in the active region is the concern for power dissipation. The problem is that when an IGBT is operating in the active range, between fully off and fully on-state, the total voltage of the source is shared among the $V_{ce}$ of the IGBT and the load, while the current is the same for both. For high loads (small resistances), most of the power is dissipated for the IGBT making the circuit inefficient. For this reason, using switching topologies where the IGBTs work from open to saturation and have a higher efficiency is may result in this type of power converter to be used in other applications. For superconducting magnets, there are options for IGBTs operating in the active region since cryogenic cooling is available.

To evaluate the instantaneous power of the IGBT, analytical simulations were performed which addresses concerns about how much power is being dissipated in the IGBT:
• Outer Loop Power Dissipation

In the case of the prototype circuit, the instantaneous peak power was 60 W. For a full power superconducting magnet power supply, the IGBT power dissipation would need to be considered and supplemental thermal management may need to be added.
References


