Solid-State Transformers for Interfacing Solar Panels to the Power Grid: An Optimum Design Methodology of a High Frequency Transformer for dc-dc Converter Applications

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Solid-State Transformers for Interfacing Solar Panels to the Power Grid
An Optimum Design Methodology of a High Frequency Transformer for dc-dc Converter Applications

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Abstract

Nowadays the use of power electronic interfaces to integrate distributed generation with the power grid is becoming relevant due to the increased penetration of renewable energy sources like solar, and the continued interest to move to a smarter and more robust electric grid. Those interfaces, which also provide a voltage step-up or step-down function, are of particular interest because renewable energy sources do not always have voltages compatible with the connecting grid. Among them, the so-called “power electronic transformer” or “solid-state transformer” (SST) is the focus of significant research. Advantages such as bidirectional power flow, improved system control, reduced size, and premium power quality at the output terminals, increase the interest of the SST for future electric grids. The SST consists mainly of two components: a high-frequency transformer (made out of advanced magnetic materials) and power converters (employing efficient power semiconductor devices like those based on silicon carbide (SiC)) to enable operation at frequencies higher than the grid frequency. This paper presents an optimum design method that can be employed to build a high-frequency transformer for a SST intended to interface a renewable energy source (e.g., a photovoltaic system) to the electric grid. Core material, geometry, and size will be analyzed in order to provide an optimum balance between cost, efficiency, thermal management, and size. Special consideration will also be given to the selection of the winding conductors given the skin effect associated with operation at high frequencies.

Keywords—transformer design, solid-state transformer, dual-active bridge.

I. INTRODUCTION

The solid-state transformer (SST), a proposed replacement for the conventional 50/60Hz line-frequency transformer (LF-XFMR), will utilize high-frequency switching converter topologies, such as the dual active bridge, to achieve many of the necessary functionalities demanded in future smart grids. Such functionalities include bidirectional power flow, intelligent system control, reduced size, and premium power quality. A converter is a power electronic interface that can perform either dc-dc, ac-dc or dc-ac power conversion using power semiconductor, such as silicon carbide (SiC). Such a converter, the dual active bridge (DAB), is utilized in the SST to perform dc-dc conversion and acts as the front-end interface for solar panels and battery storage units. The process by which this is accomplished begins with a dc voltage input (solar/battery), which is inverted into a high-frequency ac voltage, stepped up or down by the high-frequency transformer, then rectified back to a dc voltage as illustrated in Fig. 1 (X. She, R. Burgos, G. Wang, F. Wang, A.Q. Huang 2012). Though the various power SST topologies are not the subject of this research, it should be noted that based on the application and design, various types of conversions can be achieved such as ac-dc, dc-dc, or dc-ac with a number of different intermittent stages that offer the output of high or low voltage dc links. In the case of photovoltaic (PV) applications, where the output of a PV system is dc, the standard dc-dc DAB-based SST topology would be employed to convert the power delivered by the PV array to transmission or distribution voltage levels before being interfaced with the grid by a grid-tied inverter (S. Falcones, M. Xiaolin, R. Ayyanar 2010).

At the heart of the DAB is the HF-XFMR. A transformer is a voltage conversion component that steps up/steps down an ac signal from one voltage level to another by magnetically coupling two sets
of conductive windings using a magnetic core. Transformers are also utilized in applications requiring galvanic isolation and impedance matching between a source and a load. A key characteristic of a transformer is its dependency on the ac signal’s frequency, which holds an inverse relationship with the transformer’s size. In other words, higher frequency operation leads to transformers that are reduced in size (W.G. Hurley, W.H. Wolflle 2013).

These characteristics make the SST concept interesting for those applications where space is a limitation, but where they are still required to interface renewables in an efficient manner with the power grid. However, many design constraints must be taken into account when designing a transformer to operate at high frequencies. One such limitation is the availability of advanced magnetic materials that can handle higher power densities while maintaining minimal losses. At higher frequencies, heat losses are increased due to hysteresis (delayed realignment of magnetic dipoles within the core) and the development of eddy currents in the core. The smaller size and higher power densities will require more sophisticated thermal management strategies, some of which could include different core geometries and winding schemes. The conductive winding cross-sectional area must also be taken into consideration when high-frequency currents flow given the skin effect and its adverse effects.

Recent research in HF-XFMR design has presented techniques to engineers that will allow them to not only maximize efficiency with reduced size, but also utilize the leakage inductance of the equivalent circuit model in lieu of the standard external inductor implemented in the DAB topology. Researchers (K.D. Hoang, J. Wang 2012), (B. Cougo, J.W. Kolar 2012), and (P.A. Janse Van Rensburg, H.C. Ferreira 2004) describe the sizing of the leakage inductance of the transformer as being predominantly dependent on the geometry of the magnetic core and the winding scheme.

This paper presents an optimum transformer design method that can be employed to build a high-frequency transformer for a SST intended to interface a renewable energy source (e.g., a photovoltaic system) to the electric grid. Core material, geometry, and size will be analyzed in order to provide an optimum balance between cost, efficiency, thermal management, and size. Special consideration will also be given to the selection of the winding conductors given the skin effect associated with operation at high frequencies. To this end, this paper is organized as follows: relevant design considerations are given in section II, the proposed transformer design methodology is presented in section III, and a case study is provided in section IV. Finally, project conclusions and future work are addressed in section V.

II. DESIGN CONSIDERATIONS

Though the HF-XFMR will be an essential component of the power systems of the future and offers many benefits, several obstacles accompany the design process. This is due to the special considerations involved with the soft magnetic materials used to build the transformer, the increased copper winding resistances due to the skin effect, power loss calculation adjustments based on the non-sinusoidal input currents from the power electronic converters, and the thermal management issues surrounding the systems reduced size (T. Filchev, J. Clare, P. Wheeler, R. Richardson 2009). All of this must be taken into account while also keeping cost and efficiency in mind.

Magnetic materials come in two main varieties -- soft and hard magnetic materials. Hard magnetic materials are distinguished by their larger hysteresis loop under higher magnetic field intensities. This results in the need for a strong reverse magnetic field (low permeability) in order to change the flux density inside of the material. Soft magnetic materials do not require such strong magnetic fields (high permeability) to become fully magnetized or demagnetized, a
property that is employed in many power electronics today (W.G. Hurley, W.H. Wolfle 2013). Amorphous, ferrite, and nanocrystalline are soft magnetic materials presenting other desirable core characteristics (e.g., size reduction) that make them suitable to realize the HF-XFMR. Several of their properties are presented in Table I.

<table>
<thead>
<tr>
<th>Model</th>
<th>Amorphous</th>
<th>Ferrite</th>
<th>Nanocrystalline</th>
</tr>
</thead>
<tbody>
<tr>
<td>Metglas 2605</td>
<td>Epcos N87</td>
<td>Vitroperm 500F</td>
<td></td>
</tr>
<tr>
<td>Permeability ($\mu_i$)</td>
<td>10000 -150000</td>
<td>2200</td>
<td>15000</td>
</tr>
<tr>
<td>$B_{\text{max}}$ (T)</td>
<td>1.56</td>
<td>0.49</td>
<td>1.2</td>
</tr>
<tr>
<td>Curie Temp. (°C)</td>
<td>399</td>
<td>210</td>
<td>600</td>
</tr>
<tr>
<td>$K_c$</td>
<td>0.053</td>
<td>16.9</td>
<td>2.3</td>
</tr>
<tr>
<td>$\alpha$</td>
<td>1.81</td>
<td>1.25</td>
<td>1.32</td>
</tr>
<tr>
<td>$\beta$</td>
<td>1.74</td>
<td>2.35</td>
<td>2.10</td>
</tr>
</tbody>
</table>

Ferrite materials provide generally less expensive alternatives to newer materials such as amorphous and nanocrystalline; however they exhibit higher core losses per unit volume and are only suitable for temperatures up to 210 °C. Nanocrystalline is a premium material and is a suitable choice for extreme environments with relatively low losses per unit volume. However, nanocrystalline is more expensive than other materials and is still only manufactured in limited shapes and quantities. Amorphous presents the best balance between cost, losses per unit volume, operating temperature, and availability.

### A. Core Losses

The analysis of the different soft materials requires evaluating the core losses due to eddy currents and hysteresis loops. If the magnetic core is conductive, in accordance with Faraday’s law, eddy currents will be induced inside of it to oppose the change in the core’s magnetic field. The intrinsic resistivity of the core and these currents will create losses in the form of heat dissipation.

Fortunately, prior research has been conducted to aid in the design of a HF-XFMR in order to meet application, size, and efficiency requirements. When considering the core losses, traditionally measured in loss per unit volume (W/m3), the Steinmetz equation has been the standard model:

$$P_{fe} = K_c f^\alpha B_{\text{max}}^\beta$$  \hspace{1cm} (1)

where $K_c$, $\alpha$, and $\beta$ are constants typically provided by the manufacturer, or can be extrapolated from losses per unit volume graphs accompanying the core of choice. The maximum flux density at which the core is designed to operate is $B_{\text{max}}$, and $f$ is the excitation frequency. It can be seen that the core losses are proportional to the frequency and flux density while the size of the core is inversely proportional to frequency. There is a tradeoff between efficiency and size that the designer will need to consider.

One disadvantage of the Steinmetz equation is that it only considers sinusoidal inputs. For all SST applications, because of the discrete switching in the converters, square waveforms are generated and fed to the HF-XFMR. For more precise loss calculations, the improved General Steinmetz Equation (iGSE) must be used to accurately predict the core losses when factoring in non-sinusoidal excitation (W.G. Hurley, W.H. Wolfle 2013):

$$P_v = \frac{1}{T} \int_0^T k_i \left| \frac{dB(t)}{dt} \right|^\alpha \left| \Delta B \right|^{\beta - \alpha} dt$$  \hspace{1cm} (2)

Simplyfying (2):

$$P_v = k_i | \Delta B |^{\beta - \alpha} \left| \frac{dB(t)}{dt} \right|^\alpha$$  \hspace{1cm} (3)

where $\Delta B$ is the peak-to-peak flux density and $k_i$ is expressed as:

$$k_i = \frac{K_c}{2^{\beta - 1} \pi^{\alpha - 1} \int_0^{\pi/2} |\cos(\theta)|^\alpha d\theta}$$  \hspace{1cm} (4)

An accurate substitution for $k_i$, assuming square wave excitation, is:

$$k_i = \frac{K_c}{2^{\beta - 1} \pi^{\alpha - 1} \left(1.1044 + \frac{6.8244}{2.354} \right)}$$  \hspace{1cm} (5)

Finally, taking the piecewise behavior of the square waveform into account, the iGSE can be rewritten using a piecewise linear model as a function of the duty cycle of the signal:

$$P_v = k_i | \Delta B |^{\beta - \alpha} \frac{1}{T} | \Delta B |^{\alpha} \left[ (DT)^{1-\alpha} + [(1-D)T]^{1-\alpha} \right]$$  \hspace{1cm} (6)

These sets of equations will be used to estimate core losses during the initial design of the transformer. Subsequent design practices will be explored such that the user of the proposed methodology will be able to define system parameters to factor in when designing a.
HF-XFMR.

B. Winding Losses

Winding losses occur in the windings due to the intrinsic resistivity of the conductor coupled with the skin effect, a phenomenon associated with higher frequencies, and proximity effect. When calculating the losses due to these effects, it is difficult to keep them separate from one another. For simplicity, they will be referred to as eddy current losses and are largely results of the skin effect ($\epsilon$) (J.P. Vandelac, P.D. Ziogas 1988).

$$\epsilon = \frac{6.662}{\sqrt{J}}$$ (7)

A simple model of the copper losses ($P_{cu}$) in transformers using single conductor wire can be derived by applying Ohm’s law and by considering the geometry of the transformer and windings:

$$P_{cu} = \rho_w \sum_{i=1}^{n} \frac{N_i MLT (J_i A_w)^2}{A_w}$$ (8)

where $\rho_w$ is the resistivity of the winding conductor, $MLT$ is the mean length per turn, $N_i$ the number of turns, $J_i$ the effective current density, and $A_w$ is the cross sectional area of the conductor with skin effect. These parameters are all a product of the core geometry and provide for a comprehensive copper loss equation (W. Hurley, W.H. Wolfle, J.G. Breslin 1998).

This process is complicated when the designer chooses to utilize litz wire (multistrand) as the conductor. Litz wire is often the best choice because of the skin effect on single conductor windings and the large currents seen in higher power transformers. In the process of selecting litz wire, it is important to establish the cross-sectional area of the conductor as a function of the optimum current density before choosing a strand count and gauge.

$$J = K_i \sqrt{\frac{\Delta T}{2 k_u}} \left(\frac{1}{\sqrt[8]{A_p}}\right)$$ (9)

$$K_i = 48.2 \times 10^3$$ (10)

Dividing the current in the primary and secondary by the optimum current density yields the total cross-sectional area of the windings needed to facilitate the current on both sides of the HF-XFMR. The appropriate wire gauge, number of strands in the litz wire, and total equivalent resistance of each set of windings can be extrapolated from these two parameters and the resistivity of the copper.

C. Optimum Flux Density

When considering transformer core sizing, the area product ($A_p$) is a useful dimensional tool, both in general sizing and as a key design component. The area product is the product of the cross section area and the window area of the core. Reference (W.G. Hurley, W.H. Wolfle 2013) presents a method to derive an optimum area product by considering $B_o$, the optimum flux density and $J_o$, the corresponding current density:

$$A_p = \left[\frac{\sqrt{2} \sum VA}{K_f B_o k_f K_t \sqrt{k_u \Delta T}}\right]^{8/7}$$ (11)

where $\sum VA$ is the volt-ampere rating of the windings (typically twice that of the power rating of the transformer), $K_v$ is the waveform factor approximation (4.44 for sinusoidal and 4 for square), $k_f$ is the core-stacking factor, $K_t$ is a constant, $k_u$ is the window area utilization factor, and $\Delta T$ is the transformer temperature rise.

From this, an expression for the optimum flux density as a function of this area product and other key system parameters can be developed by making substitutions for the copper and core losses in governing transformer power equations. Extracting the optimum flux density can then be completed and is expressed as:

$$B_o = \frac{\left[\frac{h_c k_u \Delta T}{2^3 \rho_w k_w k_u 12 [k_c K_t]} \right]^{2/3} \left[rac{K_v k_f k_u}{12} \right]^{1/3}}{\frac{1}{5} \sum VA}$$ (12)

The flux density is limited by the saturation flux density of the chosen core material. This must be taken into account when selecting the magnetic components of the transformer and could be a factor in the overall size and cost of the device. This equation should be utilized if the efficiency is of critical importance in the design.

D. Specific Leakage Inductance

As mentioned before, efficiency may not always be the highest priority, but the design of the transformer such that the inductance required for maximum power transfer in the DAB topology could be fully integrated as the transformer leakage inductance. In this case, the designer should be
afforded the flexibility to simply choose an operating flux density, keeping in mind the saturation flux density of the core material selected. To minimize size and maximize power density while adhering to the limitations of the core material, best practice is to select an operating flux density at least 10% lower than the saturation flux density (G. Ortiz, J. Biela, J.W. Kolar 2010). An expression for the leakage inductance of a standard shell-type transformer core construction with both sets of windings wound around the center leg is:

\[ L_\phi = \frac{N_p^2 \mu MLT h}{3w} \]  

(13)

where \( L_\phi \) is the leakage inductance, \( N_p \) is the number of turns in the primary, \( \mu \) is the permeability of free space (a constant), \( MLT \) is the mean length per turn of the windings, \( h \) is the window height, and \( w \) is the window width. The \( MLT \) is a function of the core geometry:

\[ MLT = 2(a + d) + 0.8w(2 + \pi) \]  

(14)

where \( a \) and \( d \) are the lengths of the two sides of the transformer leg which the windings will be wrapped around and \( w \) is the window width. The E-type transformer core construction is the simplest in terms of leakage inductance control and thermal management because windings are wrapped around a center leg with two outer legs that aid in collecting, what would otherwise be stray magnetic field lines (B. Cougo, J.W. Kolar 2012).

III. PROPOSED METHODOLOGY

The methodology presented is optimized for designs requiring specific leakage inductances to meet DAB specifications. (W.G. Hurley, W.H. Wolfe, J.G. Breslin 1998), (J.P. Vandelac, P.D. Ziogas 1988), and (W. Hurley, W.H. Wolfe, J.G. Breslin 1998) outline effective means to minimize core and copper losses in order to maximize system efficiency while reducing size. Researchers (K.D. Hoang, J. Wang 2012), (B. Cougo, J.W. Kolar 2012), (P.A. Janse Van Rensburg, H.C. Ferreira 2004), and (G. Ortiz, J. Biela, J.W. Kolar 2010) present methods to achieve specific transformer equivalent circuit parameters. The proposed method is holistic in the sense that it provides insight from these different studies to provide the designer flexibility.

Regardless of application, an essential set of system parameters is first defined. This includes the power rating, desired efficiency, terminal voltages, planned temperature rise, switching frequency, duty cycle, the turns ratio, and desired leakage inductance. Following this, a core material is selected and its material parameters are documented. If efficiency is desired, equations (11) and (12) may be used to derive the optimum flux density and area product of the core. If size reduction or a specific leakage inductance is required, an operating flux density may be chosen, keeping in mind that it be at least 10% less than the saturation flux density of the material. Next, a core is selected along with its dimensions. From this, equation (14) is used to obtain the \( MLT \). For applications requiring a specific leakage inductance, equation (13) may be solved for \( N_p \). \( N_s \) can be shown then by dividing \( N_p \) by the turns ratio (\( N \)).

At this point, two checks can be done to verify design feasibility. Applying Faraday’s Law, the flux density of the selected core can be derived to ensure that the core will not saturate. If the leakage inductance is a critical design concern, the number of turns on the primary and secondary can be rounded down to the nearest whole number, to prevent the inductance from becoming larger than intended, and plugged into equation (13).

Once these design checks are confirmed, winding parameters can then be obtained. The skin effect and optimum current density can be calculated using equations (7), (9), and (10). The skin effect is the principle determinant of the required wire gauge, though a smaller gauge may be used. A larger than required gauge will result in an unnecessary increase in conductor volume. The current density yields the total necessary cross-sectional area of the windings which, when divided by the selected gauge of wires area, gives the total number of strands needed in the litz wire.
Another design check before continuing to the loss calculations is recommended. It may be possible that the selected conductors, with the number of turns, may not actually fit in the window area of the transformer core. Performing this simple check will ensure the transformer can be realized physically.

Finally, using the methods defined in section II.A, the total losses, both in the core and windings, can be calculated. Once all losses are compiled, the efficiency can be derived. If for any reason the transformer does not meet efficiency requirements, the process must reiterate from block two. If losses are too high, lowering the operating flux density or selecting a different core material will yield better results.

IV. CASE STUDY

For future SST projects to work, the SSEE lab requested a transformer for use in a three level full-bridge based DAB. Preliminary system specifications were first outlined, as shown in Table II. This transformer will be designed to meet a specific leakage inductance requirement, with an efficiency of 95% using an amorphous core in a shell-type transformer core construction for thermal management purposes.

With these system parameters in place, the core parameters are then compiled, along with the turns ratio necessary, to achieve the input and output voltages in Table III.

### Table II. System parameters

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Power Rating (P)</td>
<td>5kW</td>
</tr>
<tr>
<td>Efficiency (η)</td>
<td>98%</td>
</tr>
<tr>
<td>Primary Voltage (V_p)</td>
<td>600V</td>
</tr>
<tr>
<td>Secondary Voltage (V_s)</td>
<td>60V</td>
</tr>
<tr>
<td>Temperature Rise (ΔT)</td>
<td>70 °C</td>
</tr>
<tr>
<td>Switching Frequency (f)</td>
<td>20kHz</td>
</tr>
<tr>
<td>Duty Cycle (D)</td>
<td>50%</td>
</tr>
<tr>
<td>Leakage Inductance (L_ϕ)</td>
<td>405 μH</td>
</tr>
</tbody>
</table>

### Table III. Core material parameters

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Turns Ratio (N)</td>
<td>10</td>
</tr>
<tr>
<td>K_c</td>
<td>0.053</td>
</tr>
<tr>
<td>α</td>
<td>1.81</td>
</tr>
<tr>
<td>β</td>
<td>1.74</td>
</tr>
<tr>
<td>Saturation Flux Density (B_MAX)</td>
<td>1.56 T</td>
</tr>
<tr>
<td>Operating Flux Density (B_o)</td>
<td>0.5 T</td>
</tr>
</tbody>
</table>

### Table IV. AMCC-50 core dimensions

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Leg Width (a)</td>
<td>0.032m</td>
</tr>
<tr>
<td>Leg Depth (d)</td>
<td>0.025m</td>
</tr>
<tr>
<td>Window Height (h)</td>
<td>0.07m</td>
</tr>
<tr>
<td>Window Width (w)</td>
<td>0.02m</td>
</tr>
<tr>
<td>MLT</td>
<td>0.196m</td>
</tr>
<tr>
<td>Number of primary turns (N_P)</td>
<td>37.5 ~ 30-</td>
</tr>
<tr>
<td>Number of secondary turns (N_S)</td>
<td>3.75 ~ 3</td>
</tr>
</tbody>
</table>
Using equation (11) to find an optimum area product using the operating flux density yields 23cm$^4$. At this time it became apparent that a transformer core this small would not accommodate the windings and a larger core was selected. Adjustments were made later to accommodate this size increase. The AMCC-50 C-Core was selected and four of them will be combined to create the shell-type geometry with a total area product of 112cm$^4$. Equation (14) was used to derive the MLT while equation (13) was used to solve for the number of turns in the primary. The dimensions, along with the resulting MLT and numbers of turns, are listed in Table IV.

Equation (7) gives a skin effect radius of 0.0468cm, which leads to an effective cross-sectional area maximum of 6.9x10$^{-3}$ cm$^2$. The current density is found to be 250.1 A/cm$^2$. The litz wire selected was 36AWG/259 strands with a cross sectional area of 127x10$^{-6}$ cm$^2$. A single wire of this type will accommodate the current in the primary, but because the secondary will conduct a current 10x larger than the primary, these windings will require 10 of these wires be used in parallel. Using the MLT, number of turns, and the resistivity (13610 μΩ/cm) of the conductor, the primary and secondary equivalent resistances were found to be 26.27Ω and 2.627Ω respectively.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Core Losses ($P_{fe}$)</td>
<td>235.4 W</td>
</tr>
<tr>
<td>Copper Losses ($P_{cu}$)</td>
<td>2.19 W</td>
</tr>
<tr>
<td>Total Losses ($P_{TOT}$)</td>
<td>237.6 W</td>
</tr>
<tr>
<td>Efficiency ($\eta$)</td>
<td>95.5%</td>
</tr>
</tbody>
</table>

Performing the window area check shows that the windings will fit the window area with this litz wire and its required number of turns.

A final loss check is performed using the methods outlined in Sections II.A and II.B. See Table V for results.

V. CONCLUSIONS AND FUTURE WORK

This research developed and demonstrated the feasibility of the proposed HF-XFMR design methodology by presenting a case study. It is shown that there is a trade-off between size, efficiency, and cost when considering transformer design. For most applications where size and efficiency are critical components, materials like amorphous and nanocrystalline are desirable. However, in cases where cost is a factor, amorphous and ferrite are more suitable. Amorphous presents the best balance of cost, losses per unit volume, maximum temperature, and availability.

As shown, a transformer design aimed at satisfying specific leakage inductance requirements puts constraints on core size and winding turns. This oftentimes negates the calculation of an optimum flux density and area product. It is best, in these cases, to choose a flux density that matches with a specific area product such that the dimensions of the core allow for the design to accommodate the desired leakage inductance. Because of this, the process becomes reiterative in order to maximize efficiency and minimize size while meeting these requirements.

The transformer designed in this case study will be constructed and used for future SST project research in the SSEES lab at the University of Arkansas. Future work will include constructing and testing this transformer and implementing it in a 3 level full bridge based DAB topology. An analysis of thermal management options may follow.

ACKNOWLEDGMENT

Kenny George would like to thank the Arkansas Science and Technology Authority for the opportunity and financial support to conduct this summer research project. Kenny would also like to thank Dr. Juan Balda, Andres Escobar, and Roderick Garcia for their continued support and guidance in his academic pursuits.

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