

Saturable Reactor for Power Flow Control in Electric Transmission Systems: Modeling and System Impact Study

A Dissertation Presented for the
Doctor of Philosophy
Degree
The University of Tennessee, Knoxville

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May 2015

DEDICATION

This dissertation is dedicated to my wife, Deanne Young, and to my entire family for their continued support throughout my educational endeavors.

ACKNOWLEDGEMENTS

I would like to express my sincere gratitude to Dr. Yilu Liu for her continuous guidance and support as my advisor during my PhD studies at the University of Tennessee.

I am thankful to the Oak Ridge National Laboratory (ORNL) for providing the opportunity to participate in innovative research and for its financial support throughout my PhD study. My sincere thanks to Mr. Tom King of ORNL for being supportive of my educational mission and to Dr. Aleksandar Dimitrovski of ORNL for his mentorship throughout the project. My thanks also go out to the ORNL Power and Energy Systems group for their encouragement and support.

My heartfelt thanks go to Dr. Kevin Tomsovic, Dr. Lee Riediner and Dr. Aleksander Dimitrovski for their time and effort in serving as members of my dissertation committee. I would also like to thank all the professors and friends of CURENT for their encouragement throughout my research.

I would like to acknowledge the U.S. Department of Energy (DOE) Advanced Research Projects Agency – Energy (ARPA-E) for providing financial support.

ABSTRACT

A Saturable Reactor for Power Flow Control (SRPFC) is a novel application of a concept well known to electronics and power electronics engineers that provides continuous modulation of line reactance by controlling the magnetization in a ferromagnetic core. The novelty of the SRPFC is the target application, which is power flow control in meshed electric power networks. Modulation of the line impedance occurs by altering the current in a DC winding to control the magnetization of the ferromagnetic core, thereby varying the reactance of the AC winding. The deployment of power system equipment, like SRPFC, requires careful planning and study to determine if the equipment will meet design objectives and how the equipment, and associated power system, respond to power system transients. This dissertation focuses on the development of SRPFC models for simulation at time resolutions compatible with many power system transient studies. The SRPFC models developed within the course of this effort are simulated within simple power system models under nominal and line to ground fault conditions to observe effectiveness. The simulations also provide a basis for comparison between the methods and the hardware test results of a prototype device. The simulation results demonstrate the effectiveness of all but one of the modelling methods to represent the SRPFC from the perspective of the AC power system.

Fault conditions can drive the SRPFC outside of the control point in sub-cycle time and could affect distance protection along the controlled line. This dissertation includes a system impact study that quantifies these effects through transient simulations of a SRPFC model, scaled to transmission levels based on full-scale design parameters, under fault conditions in an industry accepted power system model. Impacts on protection are observed through playback of the resulting waveforms to a distance relay via a relay test set. The results demonstrate that the interaction of the SRPFC has minimal effect on distance protection reach, even when measurements are acquired by coupling capacitor voltage transformers (CCVTs).

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CHAPTER I: INTRODUCTION AND GENERAL INFORMATION

Background

The objective of electric power systems is to generate and deliver electric energy while meeting strict technical requirements in a reliable, secure, and economically efficient way. Due to the continuously growing demand in both quantity and quality of electric energy, existing power grids are under stress in terms of capacity and capability. The U.S. Energy Information Administration reports that electricity demand in the U.S. has doubled in the past thirty years and expectations are for 30% additional growth in the next three decades [1]. At the same time, variable renewable sources, power markets, inefficiencies, and accelerated aging of power grid components are making the situation ever more challenging. The system expansion needed to address these issues is costly. For instance, the median cost for adding new transmission capacity to accommodate new wind generation is about \$300/kW [2], which exceeds the cost of the most expensive Flexible AC Transmission Systems (FACTS). However, only 50% of the overall capacity of more than 300,000 miles of transmission lines in the United States is ever utilized at any given time [3]. While some of this under-utilization is due to redundancy requirements, clearly, there is a problem of power flow control. The lack of adequate power routing methods effectively results with the grid having 100% redundancy. Improved power flow control could alleviate congestion issues by taking advantage of assets already in place, thereby delaying expansion of transmission networks.

Challenges

Achieving adequate and optimal power flow control has been an arduous task for power engineers to address. The challenges stem from the complex nature of the power grid and its operation. Electric energy is difficult to store in large quantities, so generation of electricity must occur synonymously with consumption. The load demands vary continuously at timeframes ranging from seconds to seasons. The addition of renewable sources like wind and solar cause some of generation also to vary on a continual basis. The complexity of the grid itself, for instance, the highly interconnected topology and wide geographic distribution of the

transmission and distribution systems, makes it even more complicated to apply any control process. Presently, means for power flow control include tap-changing/phase shifting transformers, switched shunt-capacitors/inductors, synchronous condensers, and FACTS. These devices are either coarse in functionality or relatively expensive to install and operate, which limits the feasibility of power flow control in the contemporary power grid. Recent efforts to develop a power flow controller based on saturable reactor technology suggest that it is possible to relieve congestion with high reliability and at practical costs [4]. The technology may offer an economically viable migration path for achieving improvements in power flow control until FACTS-based devices achieve the cost and reliability readiness required for widespread commercial acceptance. The Saturable Reactor for Power Flow Control (SRPFC) is effectively a series reactor that is continuously variable within its design limits.

Power Flow Control Scenario

The concept of, and need for, power flow control is illustrated by example using the simple 3-bus system in Figure 1a. The example contains three branches of equal impedance per unit distance [5]. The topology of the system is such that the branch directly connecting the generator at Bus 1 to the load at Bus 3 is the path of shortest distance, thus it is the path of least impedance. This path of least impedance, line₁₋₃, conducts higher current than the parallel path, line₁₋₂ and line₂₋₃, causing the branch to surpass its thermal rating while the parallel path operates at sub capacity. This operating scenario is undesirable as it can hinder consumer access to lower cost generation. A 2009 study on transmission congestion by the U.S DOE identifies areas where congestion creates significant costs and alludes to its effects on reliability [6]. A solution to this congestion problem is application of a series line reactor to the overloaded line as in Figure 1b. The additional reactance balances the impedance paths in the network and power transfer occurs equally along both paths to the load. With the line reactor remaining in service, consider a 30MW load at Bus 2 as illustrated in Figure 1c. The resulting power flows are suboptimal as line₁₋₂ is overloaded. Simply taking the additional line reactance out of service actually shifts the overload condition to another branch. However, an optimal system condition is possible by varying the value of the line reactor as in Figure 1d.

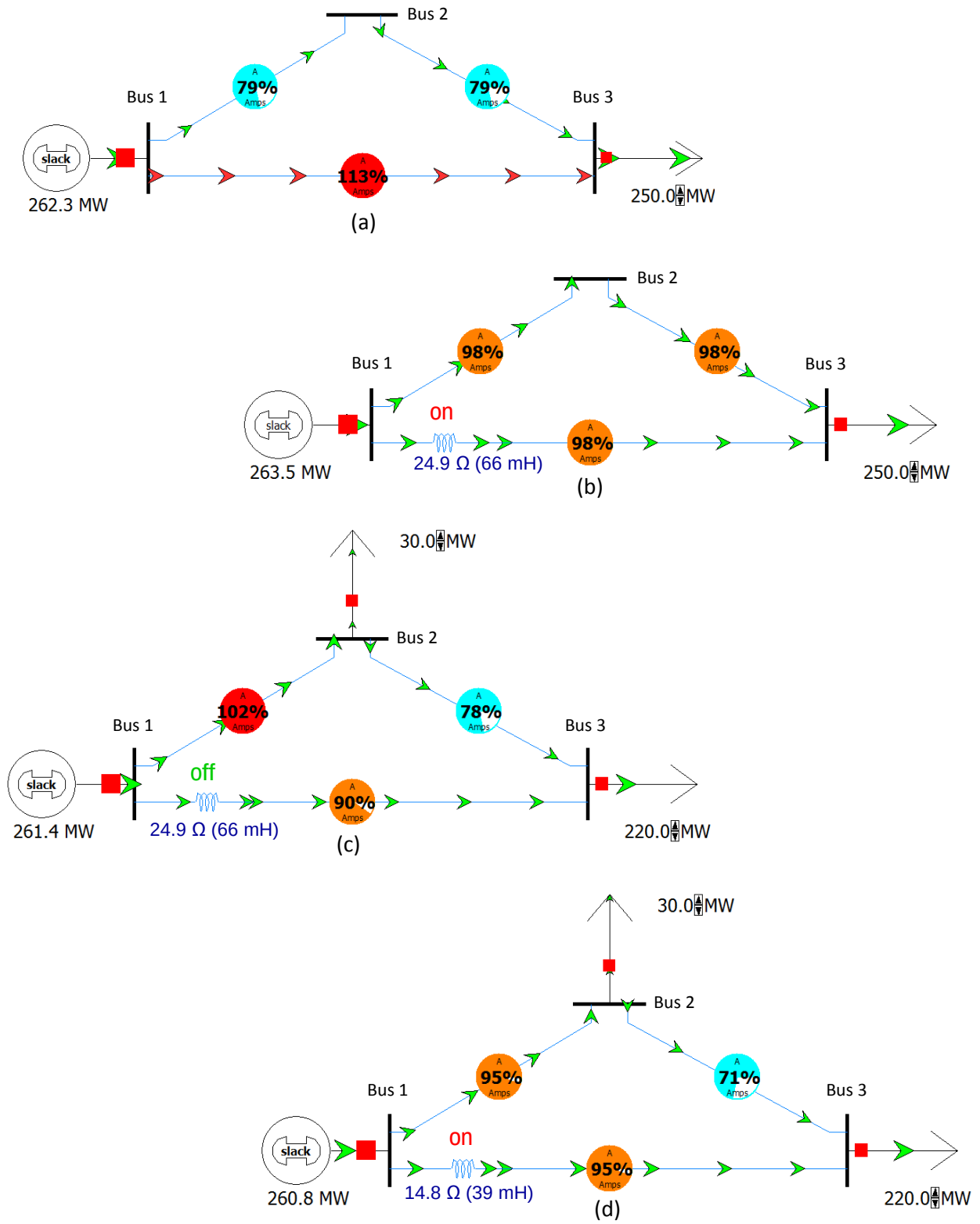


Figure 1 Impact of Line Reactance on Simple 3-Bus Power System

While series line reactors are successful in mitigating line congestion for all of the scenarios in Figure 1, the use of discrete reactors is granular and may not provide the optimal solution in all cases. A more comprehensive method is to apply a reactor that is continuously variable to achieve a power flow that is optimal for the system over a broad range of operating conditions.

Technology Overview

A publication originally published by the U.S. Navy [7] defines a saturable reactor as a device that controls the AC reactance of a coil by controlling the effective permeability of the magnetic core on which the coil is wound. In its most simple form, a saturable reactor is a ferromagnetic core with windings that connect a DC circuit, and a second set of windings that connect an AC circuit as illustrated in Figure 2a. A class of saturable reactor is the magnetic amplifier, which is identical in terms of magnetic operation but typically includes additional circuit components that modify the performance or behavior of the device (e.g. the magnetic amplifier in Figure 2b is a saturable reactor with a rectifier on the AC circuit). The terms saturable reactor and magnetic amplifier are often used interchangeably to describe the same apparatus without regard to the specific application.

In terms of operation, application of a DC bias results in current through the DC winding to establish a flux in the core. Current in the AC winding establishes a flux in a direction opposite or the same as the DC flux depending on the direction of the AC half cycle. The interaction of the fluxes sets the state of the magnetic core and determines the permeability and the reactance of the AC winding. There must be sufficient ampere-turns by the DC winding that balance the ampere-turns of the AC winding in order to control core saturation. Figure 3 illustrates the amplification principle of the saturable reactor. The waveform “INPUT 1” is applied without DC control of the core, which causes the core to operate in the linear region. Thus, the output current is small. Saturation of the core is achieved by application of sufficient DC bias, as in the case of “INPUT 2”, resulting in maximum output current. The design in Figure 2 has a significant limitation in that it contributes to the controlled line reactance for only one-half cycle of electric current unless the DC bias is high enough to cause complete saturation of the core. In the magnetic amplifier, the rectifier causes current flow in a uniform direction that

allows the DC circuit to force saturation with a smaller amount of effort (e.g. no countering flux from the AC winding).

The single-cycle limitation is overcome with modifications to the core design, e.g. the use of two cores as in Figure 4. With this configuration, each core assumes a state of operation for a half cycle. The cores then switch roles in the next half cycle. A significant design issue of a saturable reactor is the induced voltage from the AC flux on to the DC windings, or back emf. The back emf can retard controllability of the device and possibly damage equipment on the DC circuit if magnitudes are of significant levels. To mitigate this issue, the winding configurations are designed such that the series combination of the back emf for the two DC windings is zero. Other mitigation steps are possible depending on the configuration of the saturable reactor.

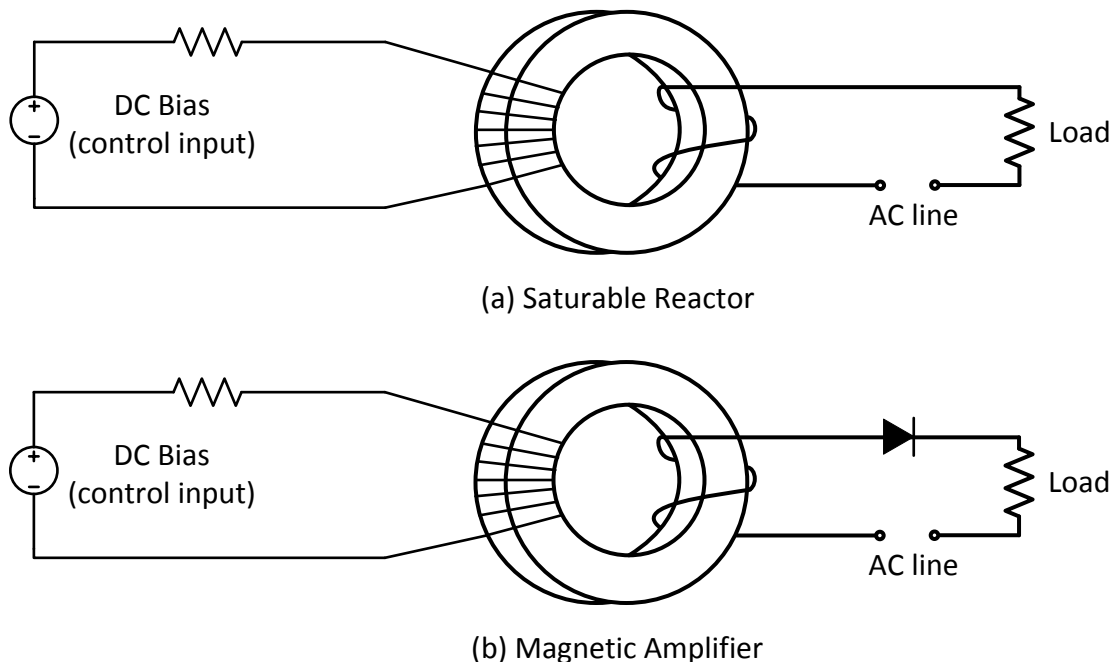


Figure 2 Basic Saturable Reactor and Magnetic Amplifier Control Circuits

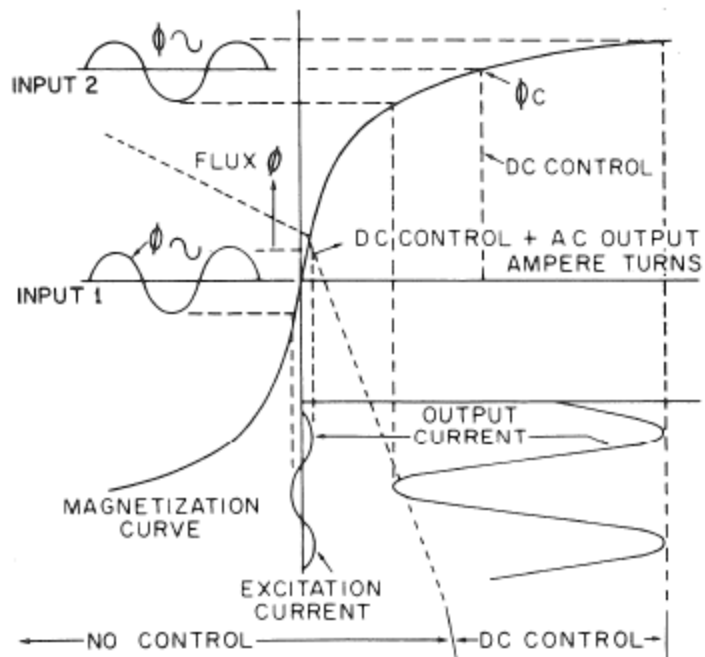


Figure 3 Amplification in a Simple Saturable Reactor [8]

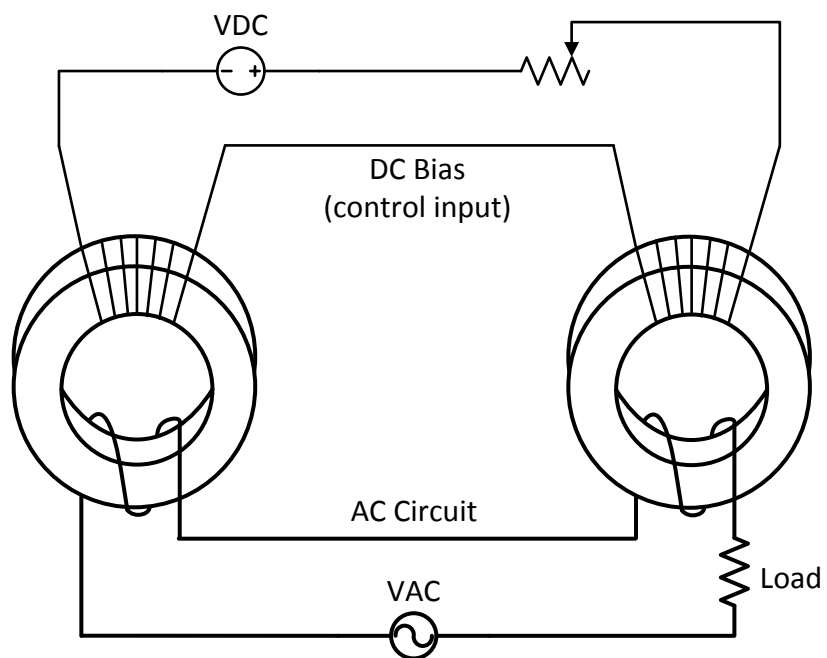


Figure 4 Two-Core Saturable Reactor

Historical Perspective

Saturable reactors and magnetic amplifiers have existed since the late 1800s with applications in motor control, radio, and electronics. Literature provides brief historical narratives on the development and application of both technologies [7] [9] [8]. The literature agrees that initial development of the technology was in the United States, but Germany enhanced the technology during World War II by introducing material improvements and new applications. The enhancements made the technology useful in military applications including guidance systems and master gun stabilizers. In addition to military use, there were applications in power flow. These applications were typically for control of end-use loads like commercial lighting. The authors in [7] and [8] make mention of civilian applications in wartime Germany that includes power flow control on electric power lines up to 50 MVA; however, a more detailed account of this application could not be found during literature review.

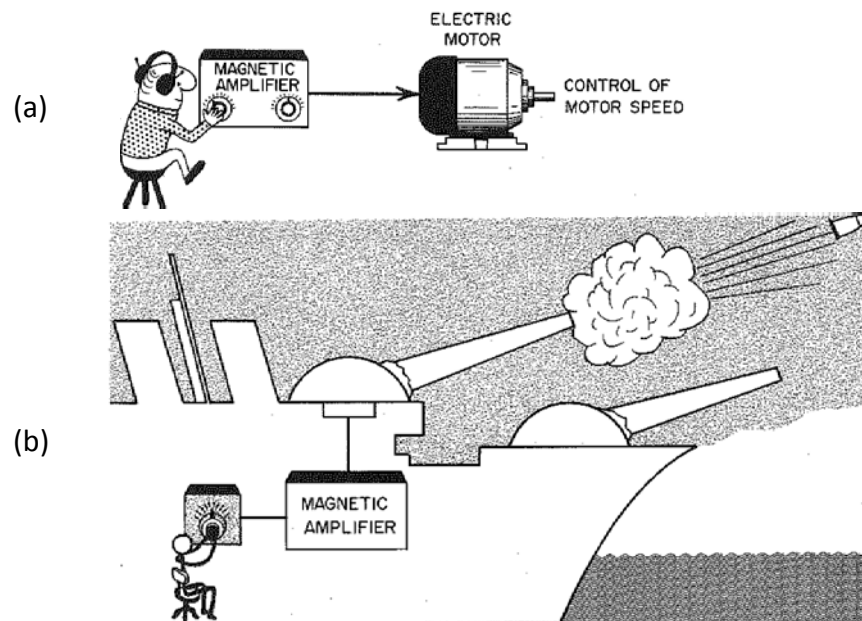


Figure 5 Motor Control (a) and Naval Applications (b) during Early 1900s [9]

Recent Application in Power Systems

Over the past decade, the development of saturable reactors for application in power systems has focused on fault current limiters. However, there have been efforts to utilize the technology at voltages less than 1kV for magnetically controlled lighting ballasts [10]. Literature review did not reveal efforts to develop saturable reactors for power flow control in high voltage power systems, so a brief summary of efforts to develop Saturable-core Fault Current Limiters (SCFCLs) is included to gain some familiarity with existing technology. SCFCLs operate in a constant state of saturation during nominal system operation by setting the DC control current sufficiently high. Fault current creates a flux that opposes the DC bias flux and forces the core into the linear region during the peaks of the fault current that effectively limits the fault current. At cessation of the fault, the core returns to saturation region and the SCFCL presents minimum reactance to the power system.

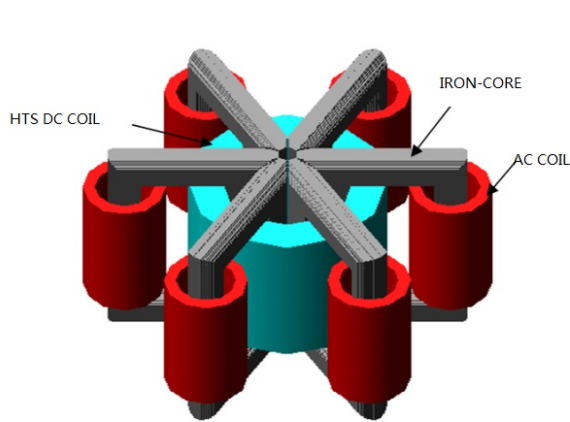
Since 2007, four SCFCLs have been deployed into commercial power systems with purpose to demonstrate feasibility and performance. In 2008, the Innopower Superconductor Company, Ltd., (Innopower) deployed a 35kV/90MVA SCFCL, Figure 6a, at the Puji substation of the Southern China Power Grid in Yunnan, China [11]. Figure 6c illustrates the configuration of the magnetic core, and reveals that each electrical phase uses two rectangular cores with the outer legs containing the AC windings. All of the cores together form a complete three-phase system with a common DC winding that encompasses all of the inner legs. The DC winding is superconducting to achieve high ampacity for meeting the ampere-turns design requirement using a low voltage DC supply. According to the manufacturer, the SCFCL can limit the magnitude of fault currents down to 50% of their expected levels. Innopower has since installed a 220kV, 300MVA SCFCL at the Shigezhuang substation in Tianjin, China. The installation shown in Figure 6b was completed in 2012 and maintains a core configuration according to Figure 6d, which similar to the 35kV unit [12]. Innopower announced in 2013 the intent to test a 500kV single-phase SCFCL sometime in 2014 [13]. All of four these SCFCLs utilize separate magnetic cores as shown in Figure 4.



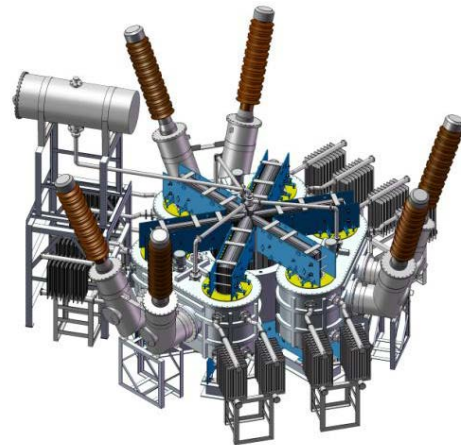
(a) 35kV SCFCL Installation



(b) 220kV SCFCL Installation



(c) 35kV Core Configuration



(d) 220kV Core Configuration

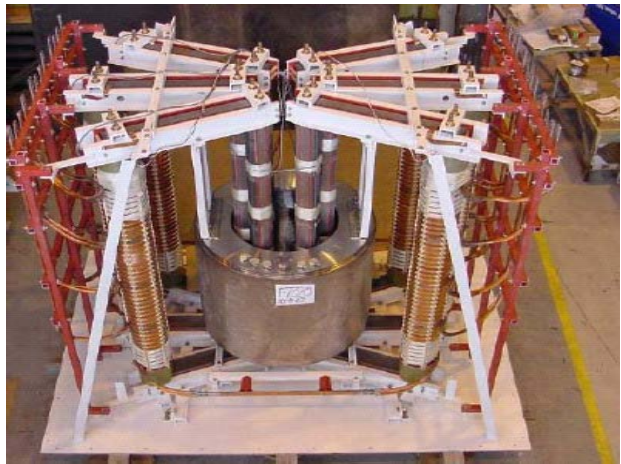
Figure 6 Innopower SCFCL Demonstrations [12]

In 2009, Zenergy Power Inc., (Zenergy) installed a 15kV, 30MVA SCFCL at Southern California's Shandin substation located in San Bernardino, California. The installation, Figure 7a, was performed at Southern California Edison's "Circuit of the Future", a 12kV feeder that serves as a test platform to incorporate new technologies and methods that increase reliability and safety while controlling customer costs [14]. The configuration of the cores in Figure 7b is similar to the Innopower design, but subtle design differences are likely in place to achieve specific performance requirements. Zenergy's SCFCL assets were sold to Applied Superconductor Ltd. in 2012 and no additional installations have been reported.

GridOn installed two of 11kV SCFCLs in the United Kingdom. Installation of a 10MVA unit at a UK Power Networks substation, shown in Figure 8, was completed in Newhaven, East Sussex in 2013 [15]. A second unit, rated at 30MVA, was installed at a Western Power Distribution substation located in Birmingham, United Kingdom in 2014 [16]. The core configuration of the GridOn FCL design is unknown as there is limited information on the design specifics in literature.



(a) Commercial Installation



(b) Core Configuration

Figure 7 15kV SCFCL by Zenergy [17]



Figure 8 11kV GridOn SCFCL Demonstration [18]

The waveforms in Figure 9 demonstrate the effectiveness of the SCFCL and are the responses from a fault test of an SFCL designed and tested by Zenergy [17]. The potential fault current, illustrated by the black curve, is 15kA with an initial peak of 30kA. Prior to the fault, the terminal voltage is very small. At onset of the fault, the terminal voltage increases significantly as the fault current, illustrated by the red curve, is limited by approximately 30%. In addition to demonstrating the effectiveness of the SFCL, the waveforms demonstrate the dynamic nature of SCFCLs in how their reactance changes throughout the AC current cycle. Under fault conditions, there is noticeable distortion at the peaks of the terminal voltage waveform. The distortion is caused by the transition of the core with opposing fluxes from the linear to the saturation region as the instantaneous AC current approaches the zero crossing. At this point of the current cycle, the instantaneous AC current approaches nominal values and the AC ampere-turns do not counter the DC ampere-turns enough to maintain operation in the linear region. However, as the instantaneous AC current approaches peak values, the AC ampere-turns sufficiently counter the DC ampere-turns and the device assumes operation in the linear region. Thus, there is a dip at the voltage peaks coinciding with the sub-cycle saturation. The remaining voltage waveform represents operation in the linear region.

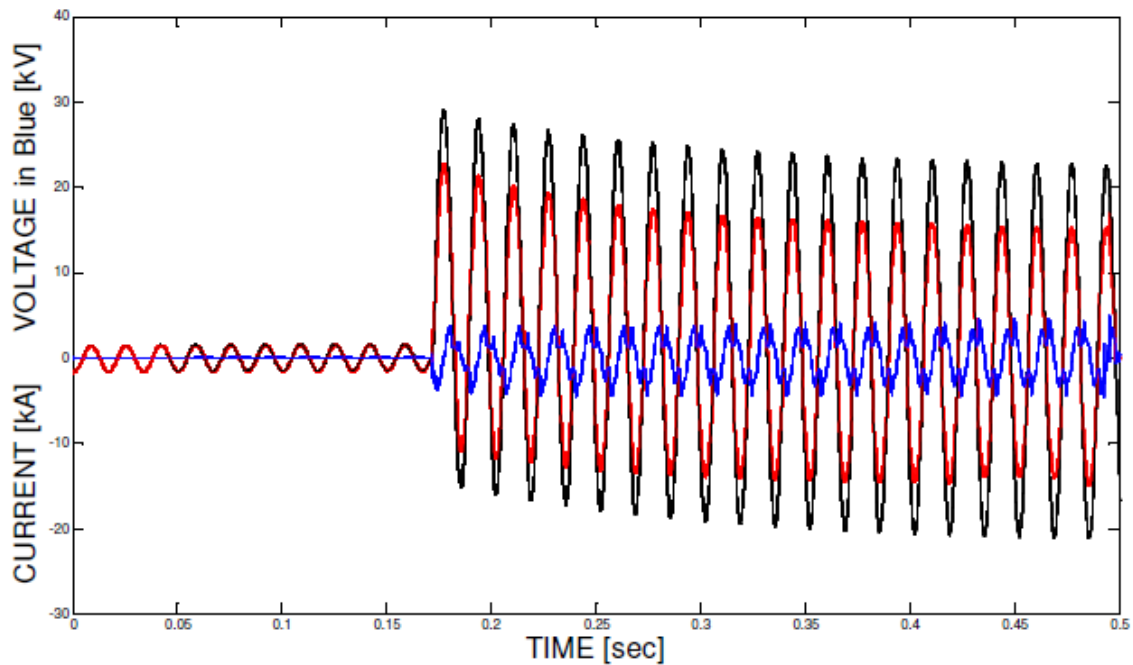


Figure 9 Zenergy SCFCL Response to 30kA_{peak} Fault [17]

SRPFC Development

ORNL is leading an effort to develop SRPFC technology for use in commercial power systems. The effort is sponsored by the Advanced Research Program Agency-Energy (ARPA-E), a program within the U.S. Department of Energy (DOE) [19]. An objective of the project under the ARPA-E is to deploy a full-scale demonstration into a commercial power system. Target specifications for the full-scale device are 115kV, 300MVA (1.5kA). The three-phase system is to be composed of three single-phase SRPFC devices. In 2013, the scaled prototype of an initial design was tested at ORNL.

ORNL Prototype

Figure 10 illustrates the single-phase design and operation of the SRPFC prototype. The prototype consists of a three-legged core with rectangular cross section and air gaps located on the center leg. The core material, TRAN-CORE® H-1 CARLITE® by AK Steel (AKH1) [20], is a material with relatively low core losses that is of common use in power system transformers. The DC magnetization curve for AKH1 steel is included in APPENDIX A. The DC windings are located on the outer legs and are connected as a common circuit. The AC windings are located on the center leg. Single-phase parameters for the prototype are summarized in Table 1. The motivation behind this design is to provide a path for the DC flux that is unimpeded by air gaps. The configuration makes it possible to achieve the DC flux necessary to achieve core saturation using a DC power supply of feasible size. Otherwise, it would require a significant DC power to achieve DC flux required to overcome the relatively large reluctance. A three-phase system is achieved by connecting three single-phase devices.

Operation of the SRPFC is dependent on the AC and DC flux components that determine the state of the core according its nonlinear magnetic characteristic. Adjusting the DC bias moves the operating point along the B-H curve and sets the inductive reactance on the AC side, thereby modulating the flow of power on the line within design limits. The saturation and linear regions of the B-H curve determine these limits. Both DC windings are connected in electrical series and wound in a common direction so that in a given half cycle the AC and DC fluxes oppose in one side of the core and add in the other. Figure 10 illustrates operation over a half

cycle, where the opposing flux components in the left leg force that side of the core into linear operation unless the DC bias is sufficient to maintain a control point at other locations on the B-H curve. The addition of the AC and DC flux components in the right leg place it into saturation, thus minimizing its inductive reactance. When operating in saturation, the air gap reluctance in the center leg dominates the core and confines the DC flux mostly to the outer ring of the core. The AC flux sees parallel reluctance paths as it splits outwardly from the center leg, so distribution of the AC flux is dependent on the state of each portion of the core at any instant of time. The outside legs exchange roles in the next half cycle. Thus, the left and right sides of the core operate in different states throughout a large portion of an AC current cycle.

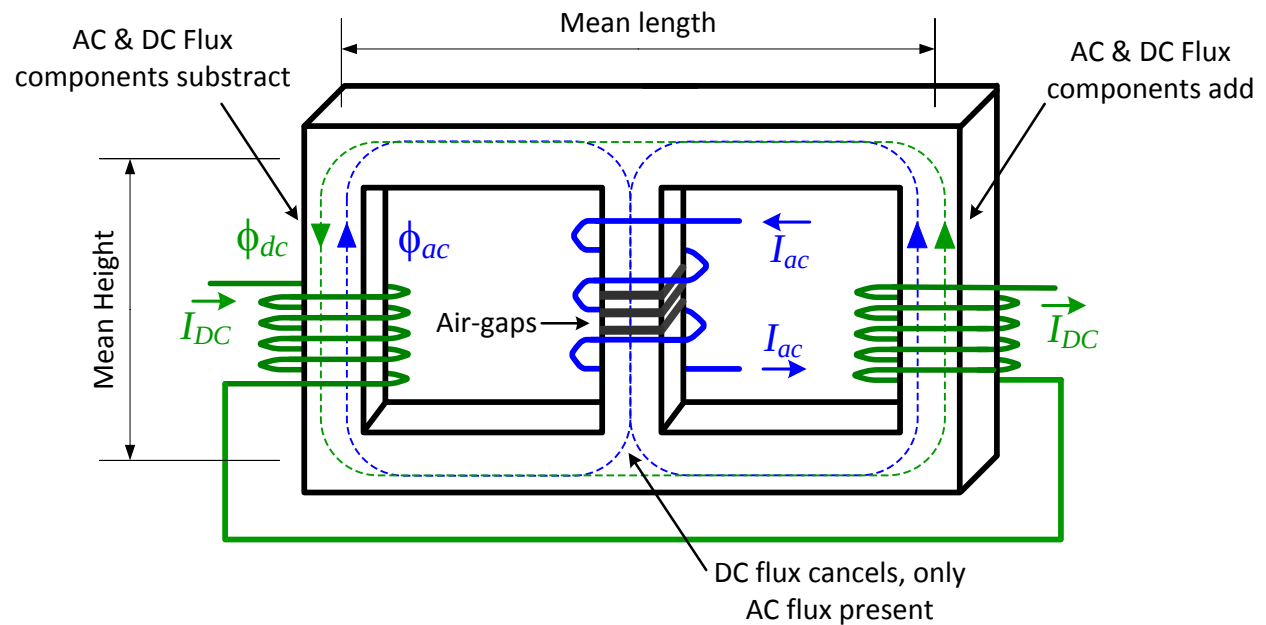


Figure 10 Configuration and Operation of SRPFC Prototype

Under nominal conditions, the DC bias controls the inductive reactance and the AC current changes in response to that reactance. However, high AC current magnitudes, like those that occur under fault conditions, create enough AC ampere-turns to overcome the DC flux and force the portion of the core with opposing AC and DC flux components into saturation. As is the case with capacitors used in series-compensated lines, this sub-cycle change in the

inductive reactance may have adverse effects on the power system, including line distance protection [21].

Table 1 Approximate Parameters for SRPFC Prototype

Parameter	Value
Rated voltage	277V
Rated current	200A
Rated VA	55kVA
AC turns	12
DC turns (each side)	24
Cross section	$0.1295 \times 0.1295 \text{m}^2$
Mean height	0.521m
Mean length	0.515m
Total air gap length	2mm

Hardware Tests

The SRPFC prototype underwent tests at ORNL to evaluate performance. The equipment in Figure 11 includes the SRPFC and the DC biasing source in their respective enclosures. Single-phase tests were performed by applying constant AC current while increasing the DC bias. The apparent reactance of the SRPFC was calculated from measurements of the line current and terminal voltage acquired at each DC current step. The test results in Figure 12 reveal a range of apparent reactance of approximately 0.7Ω at $I_{DC} = 0\text{A}$ and down to 0.13Ω at $I_{DC} = 200\text{A}$ for AC currents of 60A_{RMS} and 120A_{RMS} . These values correspond to 1.9mH and 0.33mH of inductance by the AC winding and indicate that the SRPFC can provide a maximum apparent reactance of more than five times its saturation, or minimum, value. The prototype transitions completely into saturation for DC currents beyond $I_{DC} = 150\text{A}$.



Figure 11 277V/480V SRPFC Prototype at ORNL

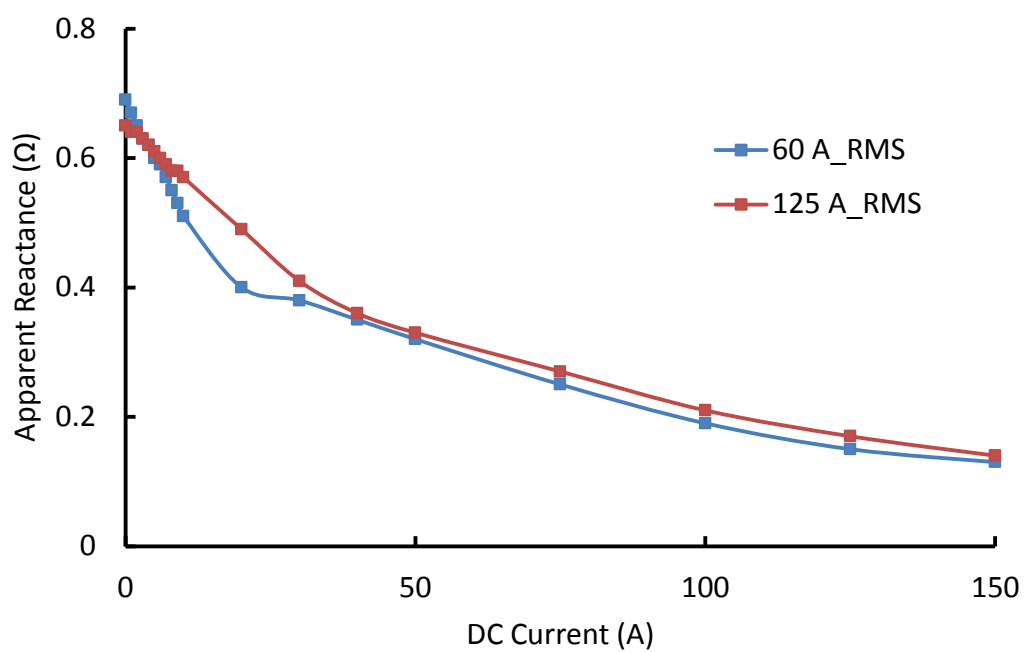


Figure 12 Single-Phase Reactance Curve from Test Measurements

Motivation for Modeling

Deployment of power system equipment requires careful planning and study to determine if the equipment will meet design objectives and how the equipment and power system respond to power system transients. Power system transients are phenomena that follow switching events and system disturbances that cause excessive currents and voltages on a system [22]. Power system transients are classified by timeframe and include the phenomena identified in Figure 13 that occur within microseconds up to within several seconds. These excessive currents and voltages stress the equipment and can cause acceleration of equipment lifecycle, equipment failures, and miss-operation of system protection. Thus, it is necessary to study the transients associated with the power system and the equipment to be deployed in order to assess impacts.

The power system transient related-phenomena highlighted in [23] include impulse transients, overvoltages, faults, transient recovery voltage (TRV), protection systems, and power flow controllers. Impulse transients are a sudden spike in voltage or current that often occurs, for example, from lightning strikes and can result in voltage, or current, many times higher than nominal but of short duration (e.g. several microseconds) [24]. Overvoltage is classified as an increase in RMS voltage magnitudes of at least 10% beyond nominal [24]. Overvoltage is less severe than impulse transients in terms of magnitude, but have durations of cycles, seconds, or minutes. Faults are events where there is a sudden interruption in the system (e.g. line to ground faults) that can be disruptive to equipment and lead to thermal or voltage related breakdowns. According to the *IEEE¹ Guide for Application of TRV for AC High-Voltage Circuit Breakers* [25], TRV is the voltage that appears across a pole of a circuit breaker after interruption. The waveform shape of the TRV depends on the nature of the circuit, whether it is resistive, capacitive, or inductive. Protection schemes detect faults and isolate equipment for purposes of human safety, equipment safety, and to minimize sustained outages. Analysis of transients and the protection scheme is necessary to verify that the protection is coordinated correctly and that any distortions in the voltage and current waveforms do not cause modern

¹ Institute of Electrical and Electronics Engineers

relays to miss-operate [26]. Studies of power system transients require application-specific models with appropriate time resolution and device physics to capture the response.

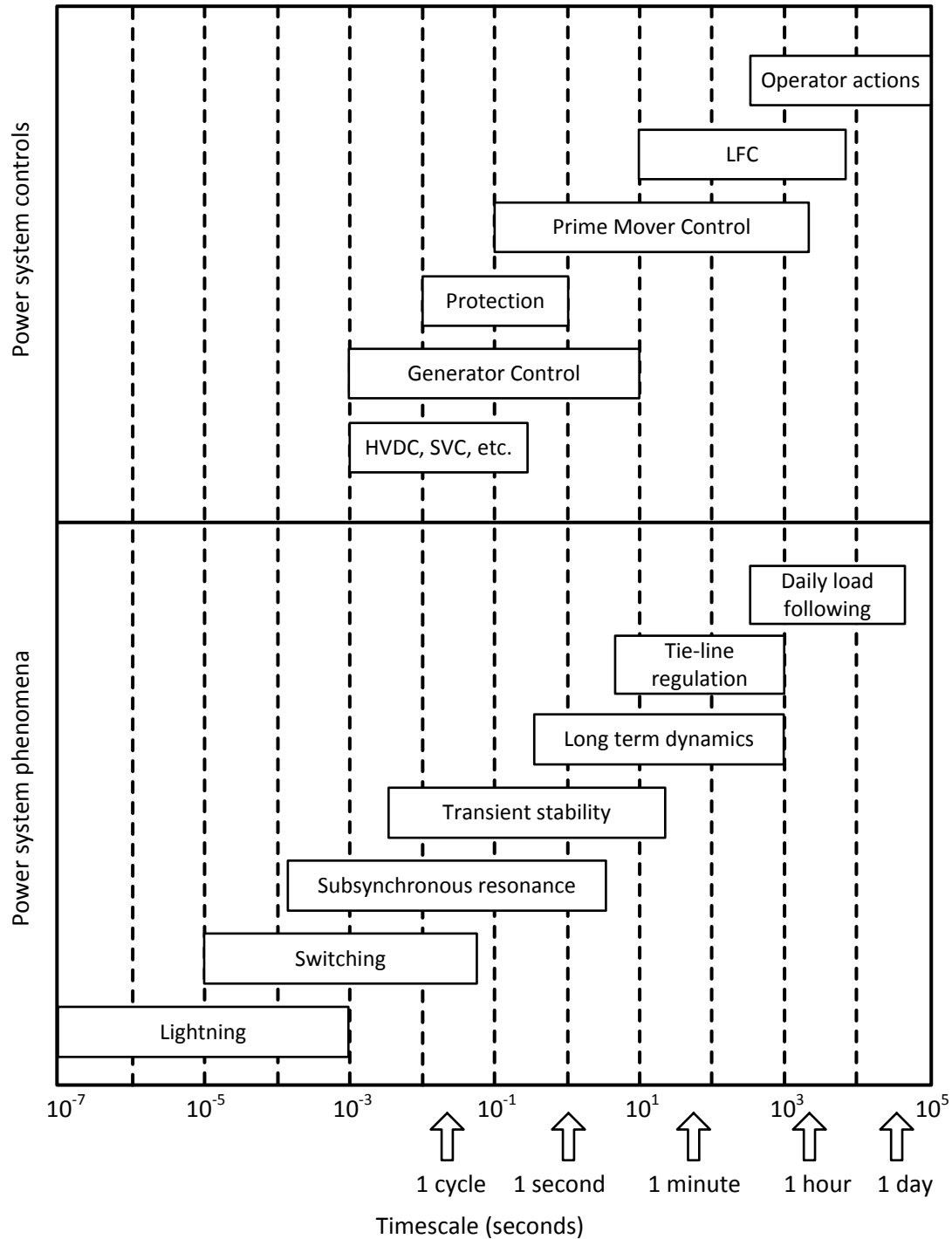


Figure 13 Timeframes of Power System Phenomena [22]

It is understood that SCFCLs change their reactance dynamically, and in sub-cycle time, when faults occur [27] [28]. However, it is unclear how an SRPFC will respond under fault conditions, what stresses it may experience, and what additional stress it may put on the power system. Therefore, this dissertation is an effort to develop an SRPFC model that represents the dynamics of the SRPFC as observed from the AC power system under fault conditions. The model, based on the 480V/277V prototype, will be scaled up to transmission levels and applied to a test power system to observe how the voltages and currents on the line react with the SRPFC in service. The data generated by simulation of the model will then be applied to a commercial distance relay to observe the performance of the protection scheme. Three modeling methods are developed based on finite element analysis and the hardware tests results.

Functional Requirements

The concept of modeling is to focus on behaviors that are significant to the device, or element, of study from the specified phenomena point of view [29]. In other words, model development should take into account the nature of the device under the conditions it is expected to operate under. Therefore, studies of power system transients require equipment models and power system models appropriate to represent behavior for the transient of interest. An adequate model is one that has high time resolution and represents the physics of the equipment under the specific transient event. Based on these criteria, the functional requirements for the SRPFC model are described as follows:

- Dynamic impedance under variable AC conditions and static DC current
- Voltage drop across the SRPFC terminals
- Adjustable DC current (though static during simulation)
- Dynamic impedance under fault conditions

The functional description of the transient model in Figure 14 describes the model in terms of inputs and outputs. The model inputs are instantaneous AC input current and DC control current, the outputs are AC current and voltage drop at the terminals of the device. Essentially,

the model should operate like a variable inductor that is a function of the instantaneous AC current and a static DC current.

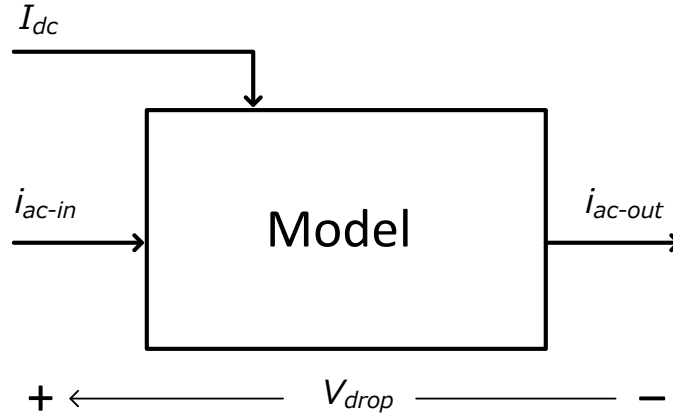


Figure 14 Functional Description of SRPFC Transient Model

The key to developing a model that is optimal – one that accurately represents the SRPFC and can be solved in sufficient simulation time – is to find a balance between the simplicity of the model and the best representation of real parameters of a given element [29]. Since the objective of the modeling effort is to achieve a model for assessing impacts to the power system under transient conditions, phenomena within the SRPFC are only necessary to gain initial understanding and some degree of confidence that the model is operating according to the intended physics. With these objectives in mind, effects and interactions that occur within the device that do not directly affect the transient behavior of the power system can be neglected in power system simulations. Additionally, the model can neglect behaviors that have minimal impact on the power system transient of interest. These behaviors the effects of magnetic hysteresis and eddy currents, voltages induced across the DC control winding by the AC winding, and time-varying DC current.

Hysteresis in magnetic materials occurs when the B and H relationship is different for increasing and decreasing values of B due to irreversible processes that result in heat loss [30]. This phenomenon typically occurs in iron and steel at varying degrees depending on the

composition and quality of these materials. Eddy currents are circular electric currents induced within conductors by a changing magnetic field in the conductor, due to Faraday's law of induction [31]. Previous efforts to understand transformer core behavior under transient conditions has shown that hysteresis and eddy currents are loss components that do not significantly impact the response [32]. According to [33], the predominate effect in transformer cores is saturation, which causes a response not encountered in linear systems. One SCFCL developer observed that physical SCFCLs do not “display” hysteretic behavior [34].

The induced voltage from the AC winding back to the DC winding is called back emf (Electromotive Force). The back emf results when the AC flux components in the outer legs are not equal, resulting in nonzero net AC flux within the DC circuit. The back emf indirectly affects the power system side of the SRPFC by perturbing the control set point or damaging the DC source. The amount of back emf is not only dependent on the state of the SRPFC, but also on the characteristics of the DC control circuit. Since methods are being implemented in the design of the SRPFC to mitigate the back emf, it is neglected.

Operation under varying DC bias current is possible, but the model must adequately represent the time constant associated with the DC circuit. This is due to the earlier point that the ampere-turns of the AC and DC windings must establish a balance point in order to make the device controllable. The AC circuit will conduct large currents with a minimal number of AC turns to achieve the desired inductance range. Therefore, the time constant on the AC winding is sufficiently small with negligible effects on the AC waveform. On the DC circuit it is practical to use a DC supply of reasonable size and cost, so the ampere-turns are achieved by increasing the number of turns. The increased number of turns increases the inductance of the DC winding and increases the time constant. This time constant will effect control and is necessary if performing dynamic control, e.g. for the application of damping electromechanical oscillations.

An SRPFC model meeting the stated functional requirements will serve as a starting point for conducting studies of power system transients and will provide users with the capability to simulate SRPFC applications within their respective power systems.

Methods to Model Saturable Reactors

A variety of modeling techniques have been applied to saturable reactors. Most techniques focus on applications in electronics or power applications at low voltage ($< 500\text{V}$). However, the magnetic interactions described by the models are relevant to all voltage classes. This section summarizes methods uncovered through literature search.

One of the earliest methods dates to the late 1950s. The method models the saturable reactor using block diagrams and transfer functions derived from equivalent circuits [35]. While the approach is useful for steady-state analysis, the basis for the method is half cycle averaging that limits the time resolution of the model. Another method proposed in the early 1960s uses equivalent electric circuits and traditional equations that describe RMS transformer voltage and current values. The method models the nonlinear magnetic behavior by tracking the change in B as H changes. A dynamic resistance, which is the core component of the model, is formed by taking the ratio of the voltage to the current. The use of RMS parameters bring into question the use of the method for transient analysis, although the method is demonstrated to provide sub-cycle solutions.

Magnetic equivalent circuits using electric circuit components to represent flux paths and windings are suggested in [36]. Traditionally, a resistance-reluctance approach is used which models the reluctance paths of a magnetic core with electric resistances [37]. However, this method models an energy storage element using a loss component, which is controversial amongst some technical experts [38] [39]. Replacing resistors with inductors mitigates the energy storage controversy, but results in a model topology that bears little resemblance to the physical application (e.g., the model represents device physics but the topology of the magnetic circuit does not visually represent the geometry of the device) [40]. The use of capacitors mitigates the energy storage controversy and results in an electric circuit configuration that is similar to the geometric construction of the magnetic device. [41]. While effective at modeling the internal interactions and design specific features of saturable reactors, magnetic equivalent circuits can become complex if there are several flux paths to consider and require an interface to link the magnetic and electric domains together into a single system. However, the method

can represent nonlinear core characteristics and support time resolutions sufficient for transient analysis.

Recent methods to model SCFCLs derive closed-form magnetic circuit equations using two ideal magnetic cores [42] [34] [43]. Each ideal core simplifies to a single magnetic path and a single magnetomotive force composed of AC and DC components. The closed-form equations are solved numerically with nonlinear reluctances related to the B-H or permeability curve. The model can be simulated at time resolutions sufficient for transient analysis. The SCFCL models developed by this approach represent physical designs that are of two-core construction, thus the simplification is a direct representation of an actual device. However, the two-core method may not be appropriate for all core designs.

Finite element analysis (FEA) is the most comprehensive modeling method in terms of capturing the electromagnetic interactions of saturable reactors. FEA is not limited to specific core designs and FEA applications can operate in tandem with transient simulators that solve the FEA model every time step in the simulation of an electric circuit. The disadvantage of FEA is simulation speed. While it is possible to optimize models and take advantage of other simplifications, like symmetry, it may not be possible to obtain solutions at speeds that are reasonable for performing studies that require repetitive simulations. Co-simulation may also not be feasible with large power system models. This relegates FEA and the co-simulation process to an effective first step in understanding device physics and as a validation step for developed transient models.

Considering the drawbacks of FEA simulation speed, it is possible to construct two-dimensional data matrix consisting of data derived from FEA simulations or measurements acquired during testing or field operation. The method uses interpolation of a two-dimensional data matrix that consists of DC control current, AC instantaneous current, and inductance or flux linkage values. The method was successfully used to model electromagnetic regulator in an electronic ballast [10] [44] [45].

The methods developed in the 1950s and 1960s were geared towards analog simulation or early computers that required the use of the simplifications that limit time resolution. However, the remaining methods can be applied using modern computers and software applications to get time resolution sufficient to simulate transients in power system studies. This dissertation focuses on the development of a SRPFC models that are applied using modern software tools and that can be simulated at time resolutions corresponding with most transients.

Dissertation Outline

The structure of this dissertation is as follows: Chapter II presents an effort to model the SRPFC using the two-core approach. The approach attempts to implement the three-legged design by using equivalent parameters derived by FEA. Chapter III describes the development of an FEA-derived model. The model incorporates a two-dimensional data matrix consisting of FEA-derived values that is applied to a commercial power system transients simulator. Chapter IV implements a magnetic circuit approach using capacitors to represent the core paths and gyrators that interface the magnetic circuit to DC and AC electric circuits. Chapter V presents a power system study applying the FEA-derived model in an industry-accepted power system model to determine impacts of the SRPFC on line distance protection during balanced and unbalanced faults. Finally, relevant conclusions and future work are summarized in the Chapter VI.

CHAPTER II: TWO-CORE MODEL

An approach to modeling the SRPFC is to apply two magnetically decoupled cores. The use of this technique was inspired by the analysis of SCFCLs that are physically constructed using separate magnetic cores [42] [34] [43]. The method represents the theoretical operation of the SRPFC as a function of the AC and DC currents. The advantage of the two-core approach is that the SRPFC can be represented by closed-form equations that are applied without difficulty in transient analysis programs for power systems analysis. However, there are inherent inconsistencies when modeling device with a three-legged core as a device with two separate homogenous cores. These inconsistencies are addressed by equivalent parameters determined by Finite Element Analysis (FEA).

Ultimately, the two-core model falls short on accuracy when compared to hardware test data. While unsuccessful at replicating measurements to desired accuracy, the two-core model provides a fundamental understanding of device physics from which to initiate modeling efforts. Hence, this chapter presents the theory operation, analysis, and application of the two-core method to model the SRPFC prototype.

Theory of Approach

The two-core model is constructed by splitting the three-legged core into two identical cores as illustrated in Figure 15. The two cores are identical in geometry and material composition. Another requirement for the cores is that they are homogenous (e.g. no air gaps or any other material transitions). The DC windings retain their configuration with respect to the three-legged core. For instance, the DC winding on the left leg will become the DC winding on core#1, but the number of turns and winding direction will not change. The DC winding on the right leg is applied identically to core#2. The AC winding on each of the two cores is identical, though wound in opposing directions to reduce back emf, and has the same number of turns as the middle leg AC winding on the three-legged core device. The full number of AC turns provides the flux linkage necessary to represent the each core when operating in the linear and saturation regions.

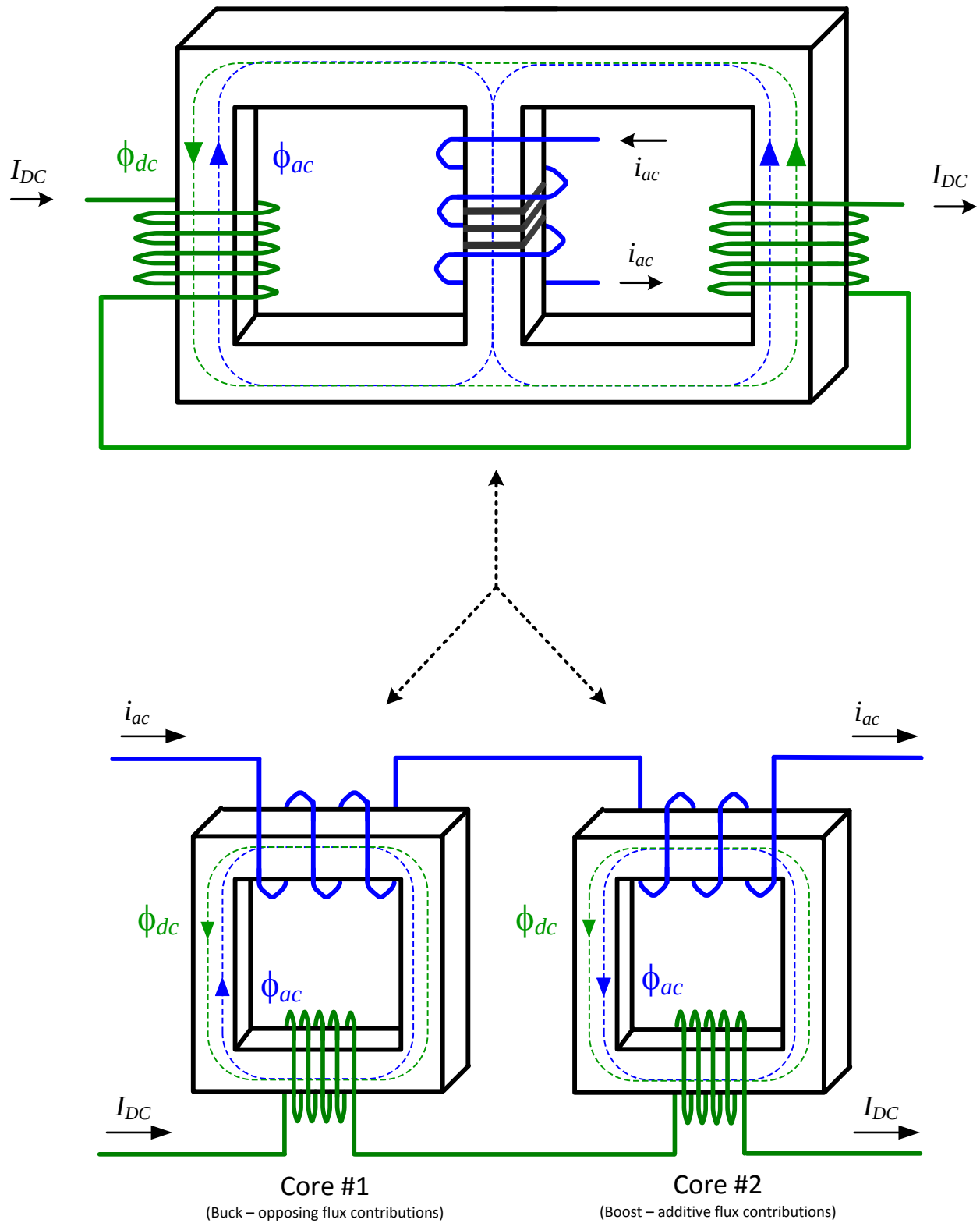


Figure 15 Three-Legged Core Split into Two Magnetically Decoupled Cores

Each core contains a uniform flux that is a combination of the AC and DC flux components. During each half electric cycle, the AC and DC fluxes counter each other in one core (buck core) and sum in the other core (boost core). The counter fluxes in the buck core cause operation in the linear region of the B-H curve, resulting in increased AC line impedance. Whereas, the sum of the AC and DC fluxes cause the boost core to operate in the saturation region, resulting in AC line impedance that is proportional to the permeability of free space. Either flux component can be significantly large enough such that the sum and/or difference of the AC and DC components result in core saturation. This is the case when the SRPFC is operating in saturation under nominal conditions, when DC current is high, or during most of a fault current cycle, when instantaneous AC current is high. Working together throughout each half cycle, the buck and boost cores provide impedance characteristics that transition proportionally to the B-H curve. With the flux interactions modeled this way, the model attempts to capture the behavior of the cores as they transition between linear and saturation regions. The electric circuit representation of the two-core model in Figure 16 shows that the cores are coupled electrically with common current in the AC windings and common current in the DC windings. However, the two-core model assumes that both cores are magnetically decoupled as there are no direct magnetic interactions between them.

Magnetic Equations

The mathematics describing the flux interactions and impedance characteristics of the two-core model begin with Ampere's law, which defines the relationship between current and magnetic field intensity.

$$\oint H \cdot dl = Ni \quad (2.1)$$

Where: H = magnetic field intensity {A·turns/m}

N = number of turns {turns}

i = current {A}

dl = differential length {m}

Evaluating the contour integral over a fixed length leads to the expression of magnetic field intensity as functions of effective core length and mmf in (2.2).

$$H = \frac{Ni}{l} \quad (2.2)$$

Where: l = mean magnetic length {m}

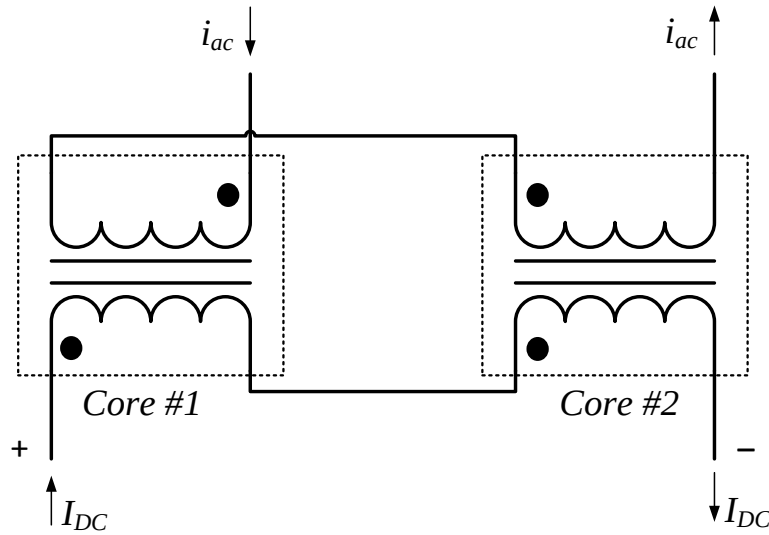


Figure 16: Electric Circuit Representation of Two-Core Model

A relationship between the flux intensity and the AC and DC parameters of each core, (2.3) and (2.4), is found by expanding (2.2) to represent the mmf in terms of AC and DC components.

$$H_{core\#1} = \frac{N_{DC}I_{DC} - n_{ac}i_{ac}}{l} \quad (2.3)$$

$$H_{core\#2} = \frac{N_{DC}I_{DC} + n_{ac}i_{ac}}{l} \quad (2.4)$$

Where $H_{core\#1}$ and $H_{core\#2}$ are the magnetic field intensities that represent each core, N_{DC} and n_{ac} are the number of turns of the AC and DC windings, I_{DC} and i_{ac} are the AC and DC currents. The expressions in (2.3) and (2.4) describe the two-core model mathematically according to the interaction of the AC and DC components. During the positive half AC cycle, core#1 flux intensity is proportional to the difference of the AC and DC mmf components as described in (2.3). The countering mmfs lead to a reduced total, or effective, mmf in core#1. Thus, core#1 operates in buck mode. Core#2 operates in boost mode during this positive half cycle, as the total mmf is the sum of the AC and DC mmf components as in (2.4). The cores exchange roles during the negative half cycle. This operating scenario assumes that the DC current remains positive as changing the sign of the DC current will reverse the operating mode of each core. The nonlinearity of the core becomes a factor upon determining B from H, or vice versa. The total magnetic field intensity, B, of each core relates to magnetic field density through (2.5).

$$B = \mu H \quad (2.5)$$

Where: B = magnetic field density {Wb}

μ = permeability {H/m}

The permeability, μ , is determined from the B-H curve, which represents the nonlinear magnetic characteristics of the core material. For modeling purposes, the nonlinear characteristic of the magnetic system may be represented by an equivalent B-H curve, which combines the B-H characteristics of multiple materials (e.g. iron and air) together into a single, or lumped, characteristic. Since the two-core model is restricted mathematically to homogeneous cores, an equivalent B-H curve is necessary to account for the core air gaps and other non-ideal core characteristics. These equivalent B-H characteristics are determined by FEA.

Once B is determined, it is possible to calculate the inductance. The total magnetic flux of each core is calculated by applying (2.6). Since the DC flux merely influences the magnetic bias in the magnetic core to establish the operating mode, the flux linking the magnetic circuit to the AC coil determines the inductance. The flux linkage is determined by application of (2.7).

$$\phi = \oint B dA = BA \quad (2.6)$$

$$\lambda = n_{ac}\phi \quad (2.7)$$

Where: ϕ = magnetic flux {T}

λ = flux linkage {T·turns}

A = cross sectional area {m²}

To express flux linkage in terms of the AC and DC currents, the definitions of B in (2.5) and flux in (2.6) are applied to the flux linkage expression (2.7). The resulting flux linkage, (2.8), is a function of the permeability and magnetic field intensity. Application of the magnetic field intensity definitions in (2.3) and (2.4) into the expression of flux linkage in (2.8) leads to a definition of flux linkage in terms of the AC and DC currents for a single core (2.9).

$$\lambda = n_{ac}A\mu H \quad (2.8)$$

$$\lambda = \frac{n_{ac}A}{l} \cdot (N_{DC}I_{DC} \pm n_{ac}i_{ac}) \cdot \mu \quad (2.9)$$

The differential inductance is the result of combining the flux linkages and application of the partial derivative with respect to the AC current as follows:

$$L_d = \frac{d\lambda}{di_{ac}} = n_{ac}A \frac{d(\mu H)}{di_{ac}} \quad (2.10)$$

This method requires expression of the flux linkage for each core as a function of AC current [43] [28] [27]. The series-connection of the two AC windings enables summation of the flux linkages to form a total flux linkage (2.11), where the negative sign of the first term denotes the relative direction of the AC winding of core#1 to core#2. This result assumes that the DC current is constant, so the process of determining the total flux linkage must be repeated when the DC current changes. Expansion of (2.11) by application of the core flux linkages from (2.9)

leads to an expression for the total flux linkage of the two-core model in terms of the AC and DC components (2.12).

$$\lambda_{total}(i_{ac}) = -\lambda_{core\#1}(i_{ac}) + \lambda_{core\#2}(i_{ac}) \quad (2.11)$$

$$\begin{aligned} \lambda_{total}(i_{ac}) = & -\frac{n_{ac}A}{l} \cdot (N_{DC}I_{DC} - n_{ac}i_{ac}) \cdot \mu_{core\#1} \\ & + \frac{n_{ac}A}{l} \cdot (N_{DC}I_{DC} + n_{ac}i_{ac}) \cdot \mu_{core\#2} \end{aligned} \quad (2.12)$$

Where $\lambda_{core\#1}$ and $\lambda_{core\#2}$ are the flux linkage of each core AC winding and λ_{total} is the total flux linkage and $\mu_{core\#1}$ and $\mu_{core\#2}$ are the permeability of the cores. Both $\mu_{core\#1}$ and $\mu_{core\#2}$ are also functions of the AC current.

The partial derivative of flux linkage is applied with respect to the instantaneous AC current as in (2.10). Since both μ and H are functions of the instantaneous AC current, the derivative is applied using the product rule. Equation (2.13) express the differential inductance in terms the differential permeability and design-specific constants. The full derivative produces four additional terms related to the DC current. Two terms are constants that reduce to zero upon application of the derivative. The two remaining terms cancel, as they are equal and opposite in sign. The differential permeability is defined in (2.14).

$$L_d = \frac{n_{ac}^2 A}{l} \cdot \left[\left(\mu_{core\#1} + i_{ac} \cdot \frac{d\mu_{core\#1}}{di_{ac}} \right) + \left(\mu_{core\#2} + i_{ac} \cdot \frac{d\mu_{core\#2}}{di_{ac}} \right) \right] \quad (2.13)$$

$$\mu_d = \mu + i_{ac} \cdot \frac{d\mu}{di_{ac}} = \frac{dB}{dH} \quad (2.14)$$

Where: L_d = differential inductance {H}

μ_d = differential permeability {H/m}

The total differential inductance can be expressed as the combination of the differential inductance for each core as follows:

$$L_{d_total} = L_{d_core\#1} + L_{d_core\#2} \quad (2.15)$$

$$L_{d_core\#1} = \frac{n_{ac}^2 A}{l} \cdot \mu_{d_core\#1} \quad (2.16)$$

$$L_{d_core\#2} = \frac{n_{ac}^2 A}{l} \cdot \mu_{d_core\#2} \quad (2.17)$$

Where $L_{d_core\#1}$ and $L_{d_core\#2}$ are the differential inductance for each core and $\mu_{d_core\#1}$ and $\mu_{d_core\#2}$ are the differential permeability of each core. A circuit representation of the differential inductance using two variable inductor elements is illustrated in Figure 17.

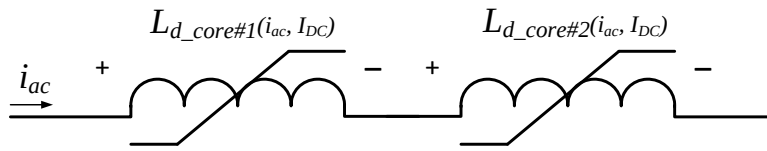


Figure 17: Series Combination of Differential Inductances

Plots of a generic B-H curve and its permeability curve are illustrated in Figure 18 as an example to show that highly nonlinear nature of the permeability curve, especially at values of magnetic field intensity close to origin. The differential permeability defines the permeability at a point on the B-H curve for small-signal analysis and is equal to the slope of a line tangent to the magnetization curve at any given value of H. This method leads to relatively small change of inductance in response to small changes in current and provides the most accurate representation of inductance characteristics for small-signal analysis. Permeability is generally expressed as the slope of a line starting at the origin of the B-H curve and intersecting any given point defined by H. This method is sufficient when considering operation in the linear region,

but underestimates the permeability within in the knee-point and saturation regions of the B-H curve. A more detailed discussion regarding differential permeability and differential inductance is provided in APPENDIX B, which derives a result proportional to (2.14). For the purpose of describing the two-core mode, the differential permeability is expressed in (2.14) as the change in B with respect to H.

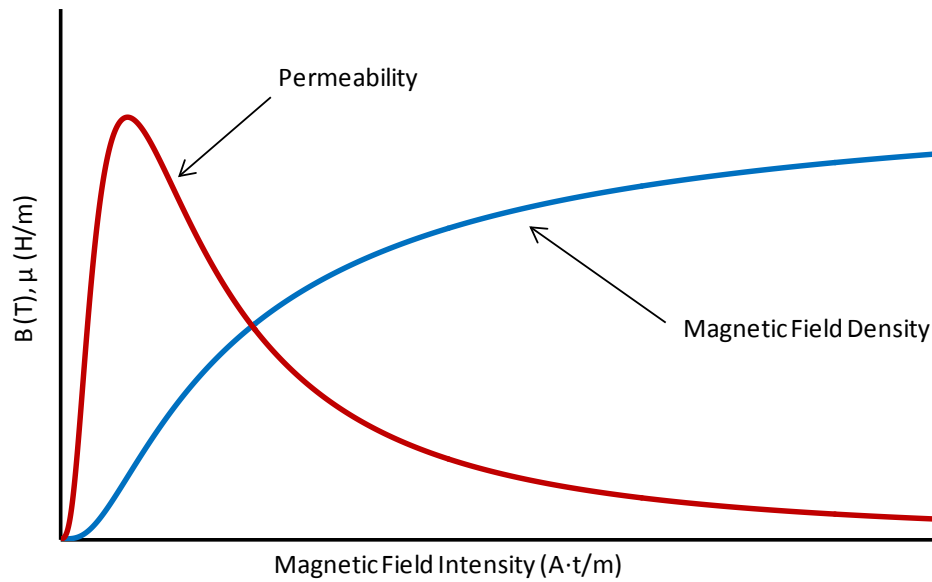


Figure 18: Magnetization and Permeability Curves

Summation of Flux Linkage Curves

The combination of the flux linkages makes it possible to model the SRPFC using equivalent flux linkage curves determined by FEA. Therefore, it may be possible to represent various core geometries using the two-core approach. This is an advantage over an equivalent B-H curve, because it is possible to model behaviors that the two-core model may not simulate inherently. Magnetic flux can travel multiple directions, across varying reluctance paths, and distribute itself non-homogenously throughout a core in practical devices, so implementation of an equivalent B-H may not produce adequate results in the two-core model. Since the magnetic flux through the AC winding determines inductance, an equivalent flux linkage as a function of

AC current may provide the most viable characteristic to apply to the two-core model. The equivalent flux linkage is determined by FEA. Consequently, the flux linkage and B-H curves are proportional, so the shapes of both curves are almost identical.

The example in Figure 19 demonstrates the summation of flux linkages from the boost core and the buck core as defined by (2.17) for $I_{DC} \gg 0$ and $I_{DC} = 0$. In this example, the flux linkage of each core is defined by a two-segment² curve within the instantaneous AC current range $-i_{max} < i_{ac} < i_{max}$. Summation of the flux linkages results in a total flux linkage defined over the instantaneous AC current range. For $I_{DC} \gg 0$, the flux linkage curves are offset from the origin by the DC flux and the flux linkage curve for core#1 is inverted due to the relative direction the AC windings. The combination of these curves yields a total flux linkage curve with slope values that are twice that of each core. As I_{DC} approaches zero, the flux linkage curves move towards each other until at $I_{DC} = 0$ they completely overlap. The differential inductance curves in Figure 20 are the partial derivative of the total flux linkage curves with respect to the instantaneous AC current. The differential inductance curves in Figure 20 are rigid due to the use of two-segment curves, but they demonstrate how the shape of the flux linkage curves impact the behavior of the model. For $I_{DC} \gg 0$, the differential inductance curve has two peaks separated by a range of instantaneous AC current. The two-core model operates in saturation for DC currents much greater than zero as long as the peak instantaneous AC currents are within the two peaks. At $I_{DC} = 0$, there is a single peak centered at zero DC current which is twice the magnitude of the peaks in $I_{DC} \gg 0$ case. In this state, the two-core model operates in the linear region for portions of the instantaneous AC current waveform within the range that define the width of the peaks. As DC current increases from zero, the flux linkage curves shift away from the origin and the single peaks begin to separate until two smaller peaks are formed. Further increase of DC current separates the peaks as described by the $I_{DC} \gg 0$ case.

² Mathematically, the two-segment curve is discontinuous at the intersection of the linear and saturation regions. In application, the transition region is continuous as the knee-point occurs over a range of instantaneous AC current. The discontinuities are neglected for the two-segment and three-segment curves presented in this dissertation.

The B-H curves of ferromagnetic materials generally have an inflexion point near the origin. The presence of this inflexion point affects the differential inductance characteristic of the two-core model. To demonstrate the impact, a three-segment flux linkage curve is applied to the two-core model where the first segment represents the inflexion point. The resulting composite flux linkage curve in Figure 21 contains an additional segment in the second and third quadrants that break the high slope segment. The differential inductance curves in Figure 22 illustrate how the additional segment alters the shape of the differential inductance peaks. In this example, the inflexion point affects the differential inductance curve as it causes a noticeable dip in the middle of each peak. The point of the example is to show that subtle undulations in the flux linkage curve become apparent upon taking the derivative.

Recent efforts to construct saturable-core fault current limiters, [42] [34] [27], depend on the inductance characteristics for $I_{DC} \gg 0$ described in Figure 20 and Figure 22. The DC current is selected such that the nominal AC current operates within the “valley” portion of the curve between the differential inductance peaks. Under fault conditions, the peaks of the instantaneous AC current result in device operation in the peaks of the differential inductance. Figure 23a is an example that shows how instantaneous nominal and fault currents set the operational state of the saturable-core FCL with a two-segment flux linkage curve. The sudden transition from low to high inductance effectively limits the fault current. Thus, an ideal saturable-core FCL application will have minimal impact on the system during nominal operation; but must react in fractions of a cycle to limit fault currents. If the instantaneous AC current magnitude increases too much during a fault, the peaks of the instantaneous AC current waveform go beyond the differential inductance peaks and the fault current limiter assumes reverse saturation until the instantaneous AC current changes direction and falls back into the high inductance region. Ideally, a saturable-core FCL will have a differential inductance curve with peaks that are steep in order to maximize the fault impedance. The saturable-core FCL must be biased high enough to maintain saturation under nominal current, but provide linear operation during fault currents.

Unlike the SCFCL, SRPFC operates within the entire range of DC current. This type of operation contrasts the application of a saturable-core FCL, which operates at a single DC bias current. This makes operation more complex than a fault current limiter as the differential inductance characteristic changes for every value of DC current. Figure 23b describes operation of a power flow controller with a two-segment flux linkage curve as the DC current is modulated. The inductance peaks merge together as the DC current decreases. The nominal current will see the changing inductance until the DC current increases to a point where the AC waveform operates entirely within the “valley” region between the peaks. In order to accommodate inductance change gradually over the entire AC waveform, it is desirable to have a flux linkage curve with a smooth transition from the linear to the saturation region, e.g. differential inductance curves that gradually ascend and descend. A smooth transition permits the SRPFC to vary the differential inductance while minimizing harmonics. It is obvious from the example that a two-segment curve is not sufficient for designing an SRPFC as it results in differential inductance peaks that transition abruptly between linear and saturation regions. Ultimately, the smoothness of the transition depends on the core design, core material, the DC ampere-turns, and the AC ampere-turns. Therefore, it is the design objective, whether SCFCL or SRPFC, that determines the shape of the differential inductance curve.

Equivalent Parameters

The two-core model is an ideal representation of the SRPFC. However, the prototype is a three-leg core device with an air gap on the center leg. These characteristics of the prototype are incompatible with the two-core model, which assumes that each core is homogenous (e.g. magnetic circuit is represented by a single reluctance). However, it may be possible to apply equivalent parameters to the two-core model that have the core characteristics of the three-legged core embedded into them. The objective of this effort is to develop an equivalent flux linkage profile as a function of the magnetic field intensity, H , to apply to the two-core model.

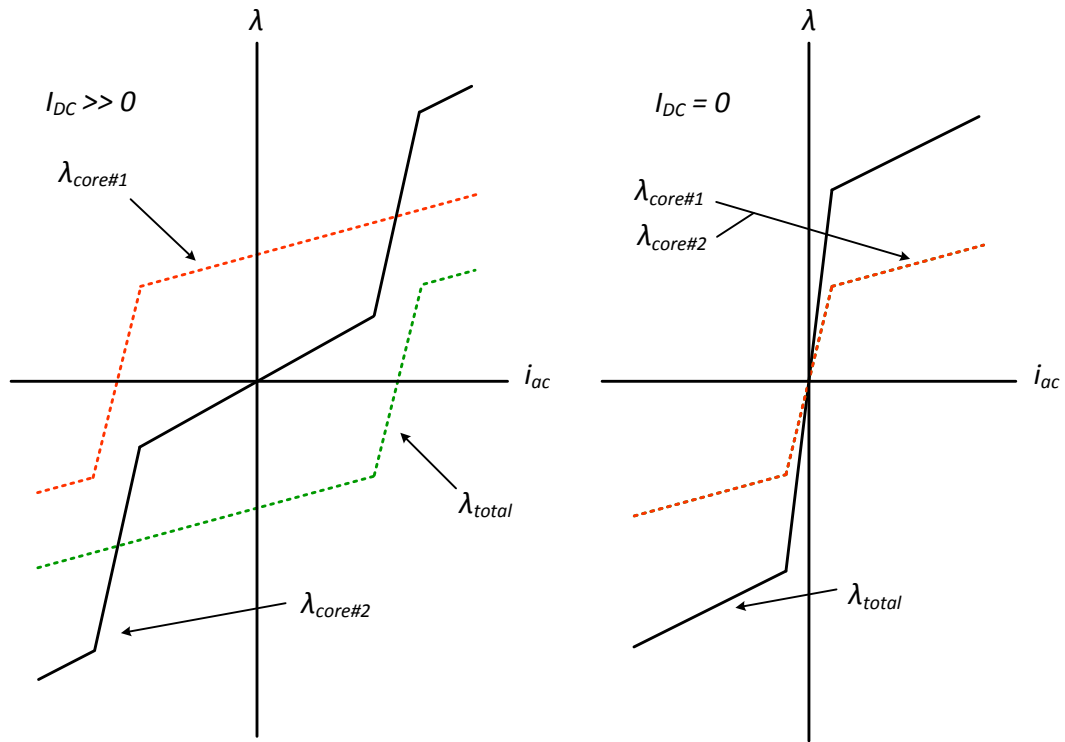


Figure 19 Two-Segment Flux Linkage Curves

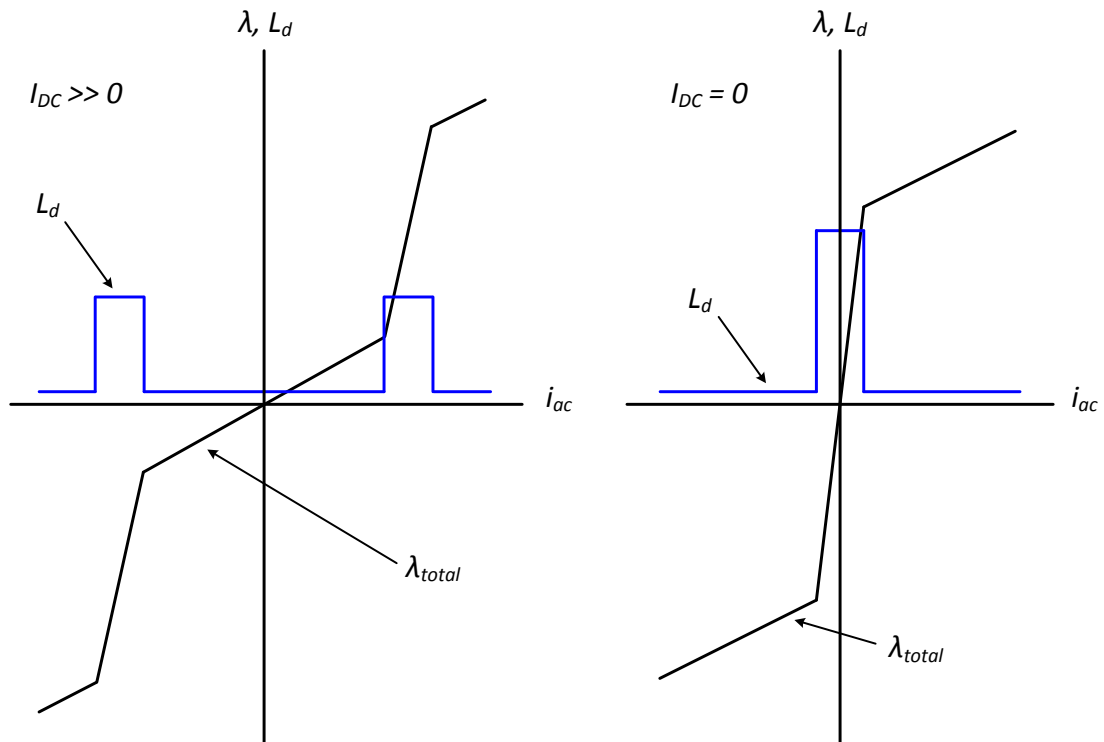


Figure 20 Differential Inductance Curves from Two-Segment Curves

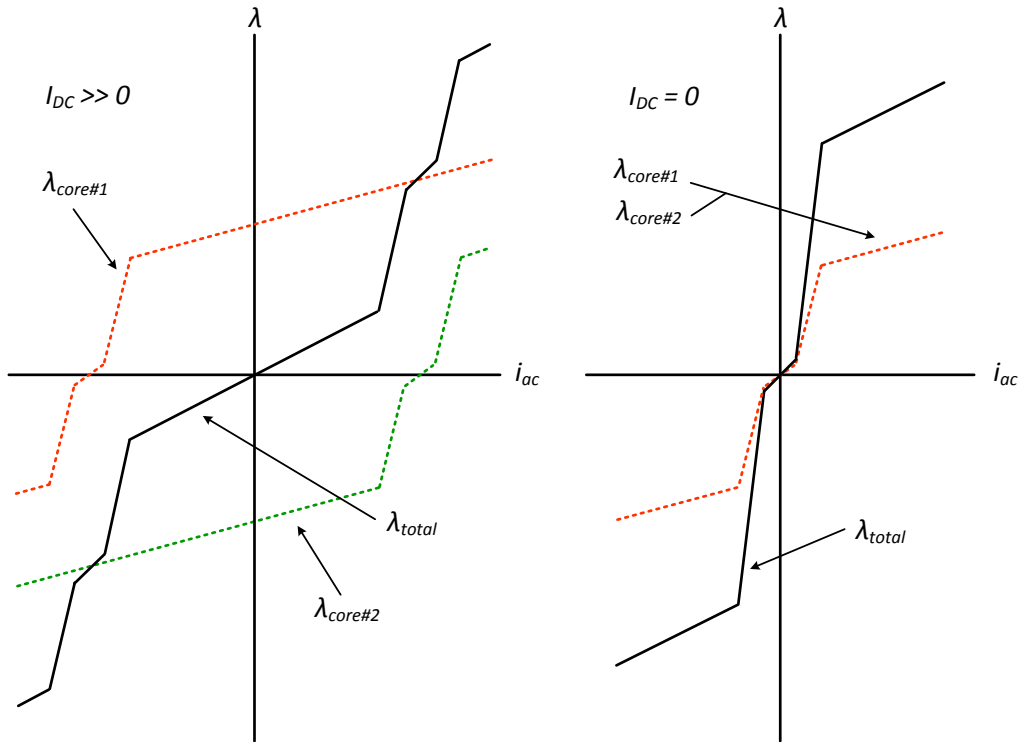


Figure 21 Three-Segment Flux Linkage Curves

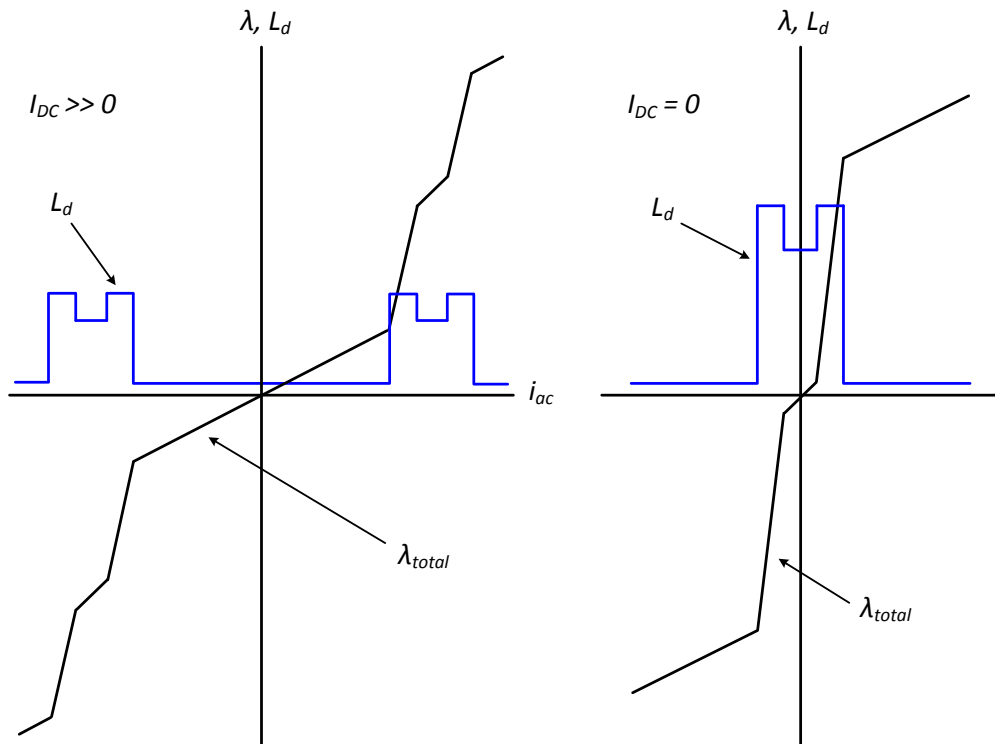
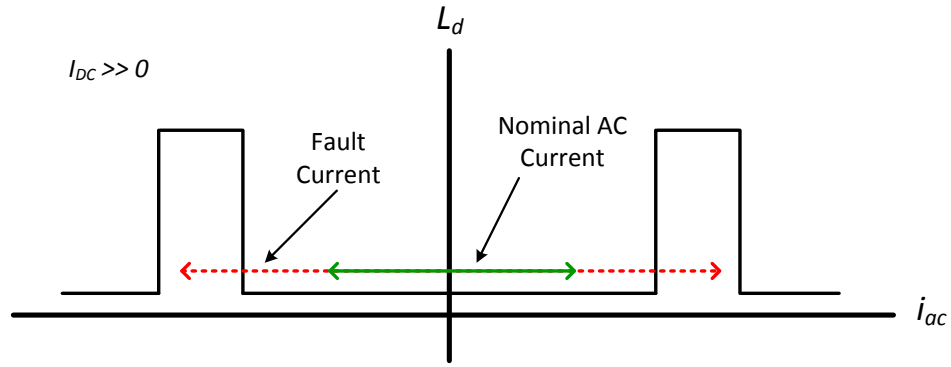
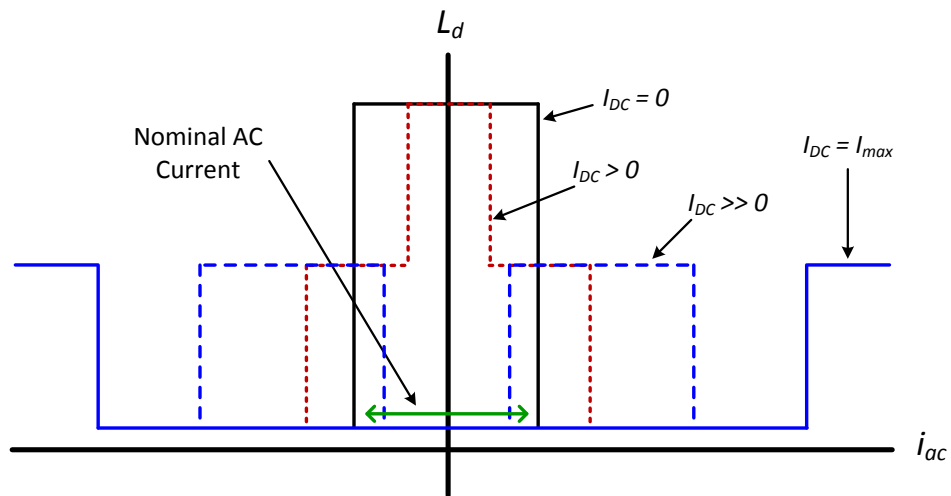


Figure 22 Differential Inductance Curves from Three-Segment Curves



a) Example -- Saturable-Core FCL



b) Example -- MAPFC

Figure 23 Two-Segment Examples of Saturable-Core FCL and SRPFC

The FEA contains the detailed modeling of the nonhomogeneous geometric and material characteristics of the core, thereby imparting these characteristics to the flux linkage profiles. The two-core model calculates the total H of each ideal core based on AC and DC parameters; then applies it to the equivalent parameters from FEA to estimate equivalent flux linkage for the given operating condition.

Finite Element Modeling and Analysis

Finite element models were constructed in an effort to develop equivalent parameters for the two-core model. The objective of the FEA is to acquire equivalent flux linkage at the AC terminal

as a function of H . The curves were generated by opening the DC circuit and applying a current ramp to the AC terminal. The resulting flux linkage curve for each model has a strong resemblance to the B-H curve with each model producing proportional results. Details of the FEA modeling effort and the resulting equivalent flux linkage curves are provided in APPENDIX C. The three models constructed are the three-legged core, split-core, and single-core. FEA provides equivalent flux linkage curves in all three cases. The effort determined that the three FEA models produced equivalent flux linkage curves that were proportional to each other. The resulting equivalent flux linkage curve is applied to the two-core model later in this chapter.

Effective DC Current

The two-core model and the FEA split-core model fail to capture the behavior of three-legged core model in terms of the DC flux as (2.3) and (2.4) assume that both flux components see the same reluctance paths. The AC leg of the three-legged core is a high reluctance path to the DC flux since one or more air gaps exist there, so the majority of the DC flux is confined to the outer ring of the core as illustrated in Figure 24. However, the DC flux in the FEA split-core model has no option but to traverse the air gap. Thus, the FEA split-core model requires more DC current to achieve the same level of DC flux. While the two-core model does not possess an air gap, it exists by virtue of the parameters embedded within the equivalent flux linkage curve. To model the low reactance DC flux path of the three-legged core, an effective DC current is proposed. The method attempts to scale the DC current so that it causes a DC flux value correlating to the low reluctance path of the core's outer ring.

The method begins by considering the two ideal cores illustrated in Figure 25. The cores have identical DC winding turns, mean magnetic length, and cross-sectional area. No AC windings are present. The only difference between the cores is that one core has an air gap and the other is gapless. The method depends on two observations from the cores in Figure 25. First, the reluctances of the cores are not equal, as the core with air gap is a higher reluctance to the DC flux. Second, the higher reluctance of the core with air gap will require more DC current to achieve the same amount of flux than the gapless core. Considering these observations, the method equates fluxes of the cores in terms of their respective DC currents and determines

these DC currents using a numerical analysis. The method provides a correction curve that relates the effective DC current to an applied DC current. The numerical analysis makes use of two homogenous cores that are characterized by the equivalent magnetization curves in Figure 26. The curves are the result of FEA for the gapless and air gap split-core models without AC windings and reveal that the presence of the air gap decreases the slope of the linear region, corresponding to low magnetic field density, while preserving the slope of the saturation region. This result is consistent with previous observations [46].

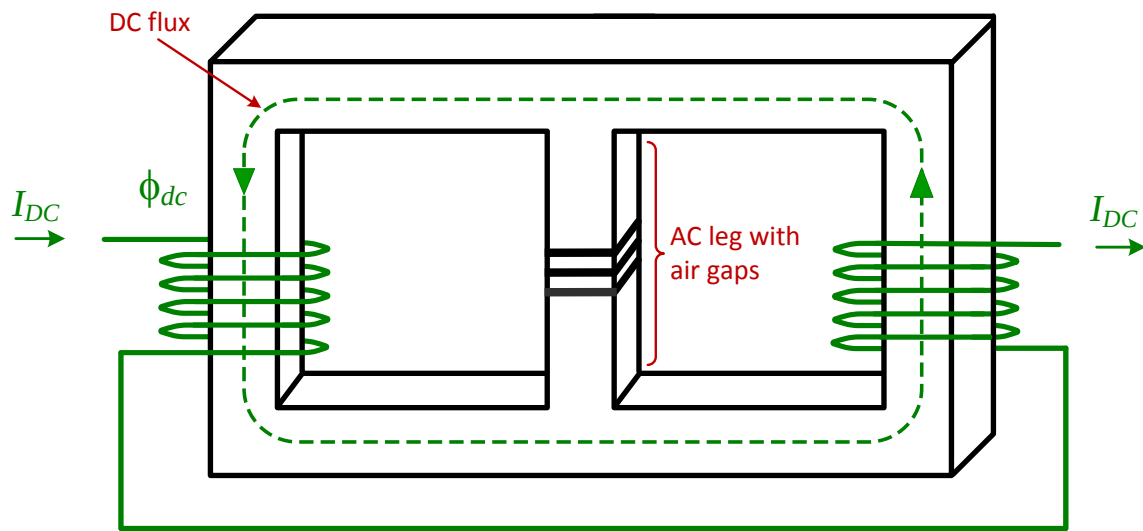


Figure 24 Diagram of Three-legged Core showing the DC Flux Path

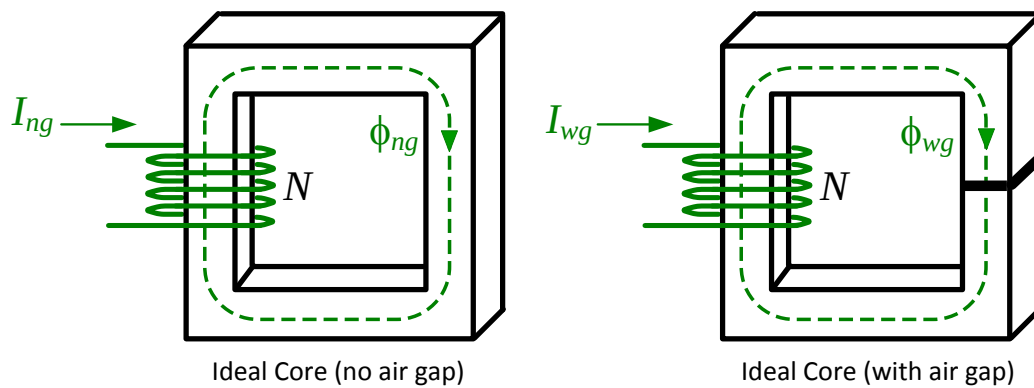


Figure 25 Core Models used to Determine Effective DC Current

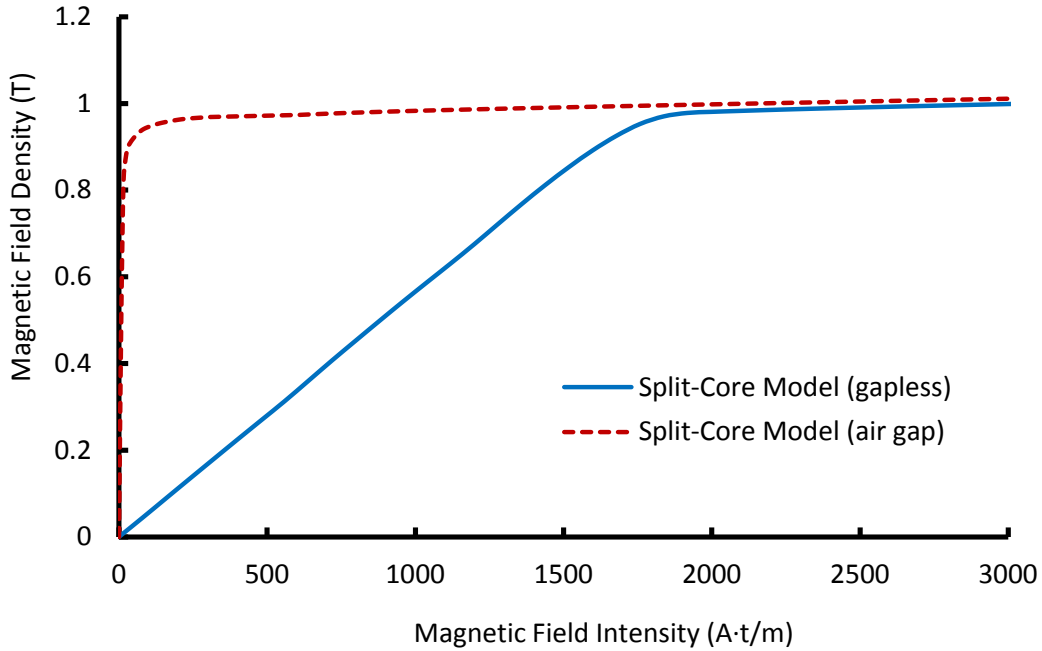


Figure 26 Magnetization Curves used for Calculation of Effective DC Current

The analysis equates the flux in both cores as in (2.18). The use of two homogenous cores allows the mmf of each core to be represented by a single term; thus, each core is represented by a single magnetic field intensity as in (2.19) and (2.20). The flowchart in Figure 27 describes the numerical processes and was applied using MATLAB® (APPENDIX E). First, an identical flux value is applied to both cores. For this analysis, the applied flux values are in a range from zero to some maximum value. Next, the magnetic field density is determined for each core and used to estimate the magnetic field intensity for each core by interpolation of the equivalent B-H curves. The DC currents for each core are determined using from the estimated magnetic field intensities according to (2.19) and (2.20). The ratio of the estimated DC currents yields a correction factor that is applied to (2.21) to determine an effective DC current. The resulting correction curve in Figure 28 describes a nonlinear relationship between the applied DC current and the effective DC current. For small applied values of DC current, there is significant increase in the effective DC current. This increase is synonymous with the dominance of the air gap reluctance in the single-core when operating in the linear region. Therefore, a large effective DC current relative to the applied current is necessary to achieve an equivalent flux in the

homogenous core. For very low values of applied DC current up to approximately 50 amps, there is little change in the effective DC current. In this region, the gapless core is fully saturated while the air gap core operates in the linear region. As the applied DC current value increases beyond 50 amps, both cores are in saturation and the relationship between both cores takes on a linear profile.

$$\phi_{ng}(H_{ng}) = \phi_{wg}(H_{wg}) \quad (2.18)$$

$$I_{ng} = \frac{H_{ng}N_{DC}}{l} \quad (2.19)$$

$$I_{wg} = \frac{H_{wg}N_{DC}}{l} \quad (2.20)$$

$$I_{eff} = \frac{I_{wg}}{I_{ng}} \cdot I_{applied} \quad (2.21)$$

Where: H_{ng} = magnetic field intensity of gapless core
 H_{wg} = magnetic field intensity of core with air gap
 ϕ_{ng} = magnetic field intensity of gapless core
 ϕ_{wg} = magnetic field intensity of core with air gap
 I_{ng} = DC winding current for gapless core
 I_{wg} = DC winding current for core with air gap
 l = mean magnetic path of each core
 $I_{applied}$ = DC current applied to two-core model
 I_{wg} = effective DC current for two-core model

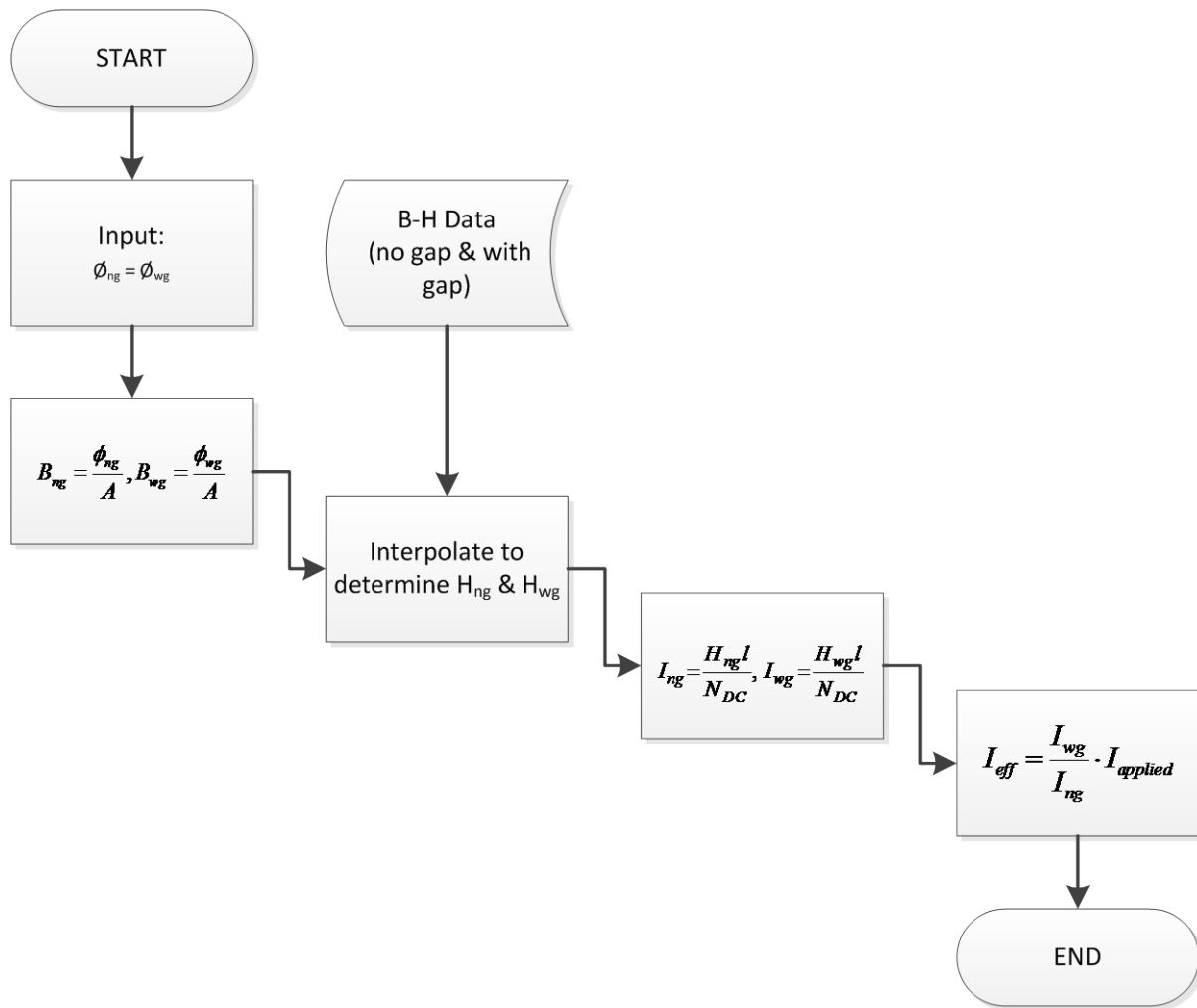


Figure 27 Flowchart for Calculation of Effective DC Current

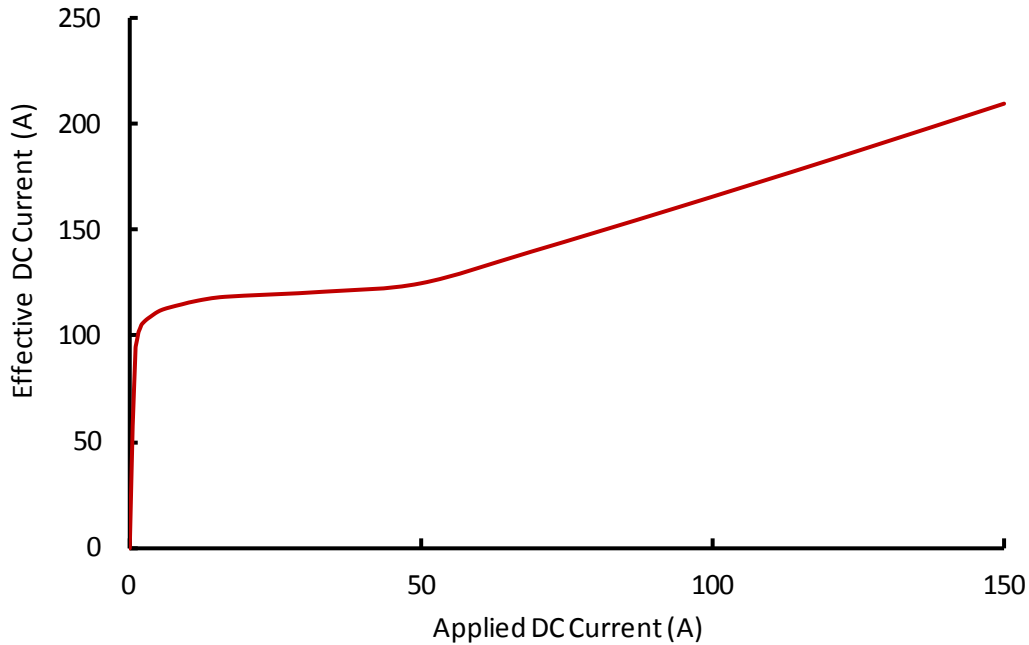


Figure 28 DC Current Correction Curve

Simulation in MATLAB®

Simulations of the two-core model in MATLAB® (APPENDIX E) were performed using the equivalent flux linkage curves to estimate the differential inductance behavior and the apparent reactance profiles. Differential inductance simulations were performed using the numerical process summarized by the flowchart in Figure 29. Inputs to the process include the equivalent flux linkage curve, a range of instantaneous AC currents, and a single DC current value. The first step is calculation of the magnetic field intensity for each core from the AC and DC currents. Second, the flux linkage is estimated by interpolating the equivalent flux linkage curve using the H calculated for each core. The third step is to combine the flux linkages as in (2.11). The last step is to determine differential inductance by taking the difference of the flux linkage with respect to instantaneous AC current.

The numerical process used to determine the apparent reactance profile is summarized by the flowchart in Figure 30. The process applies instantaneous AC current defined over one electrical cycle at a frequency of 60Hz as defined by the first step. No harmonics are present in the AC

current waveform. Second, the magnetic field intensities are calculated for each core. The third step is interpolation of the flux linkage of each AC winding using the calculated H values. The fourth step is the calculation of the instantaneous voltage from the time derivative of the total AC flux linkage according to (2.22).

$$v_t = L_d \cdot \frac{\partial i_{ac}}{\partial t} = \frac{\partial \lambda_{tot}}{\partial t} \quad (2.22)$$

The RMS voltage and current are calculated over the electrical cycle. The apparent reactance is the result of dividing the RMS voltage by the RMS current. The process in Figure 30 is repeated for all DC current values in the operating range.

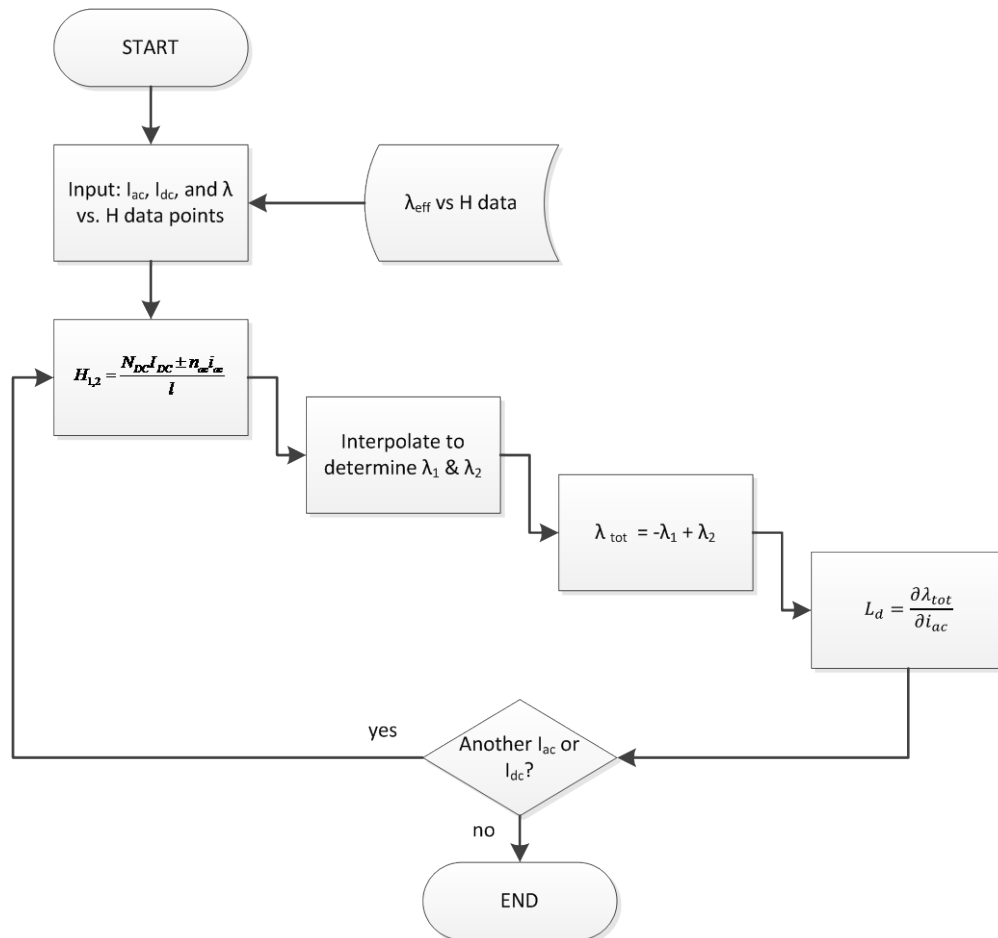


Figure 29 Numerical Procedure for Calculating Differential Inductance using Two-Core Model

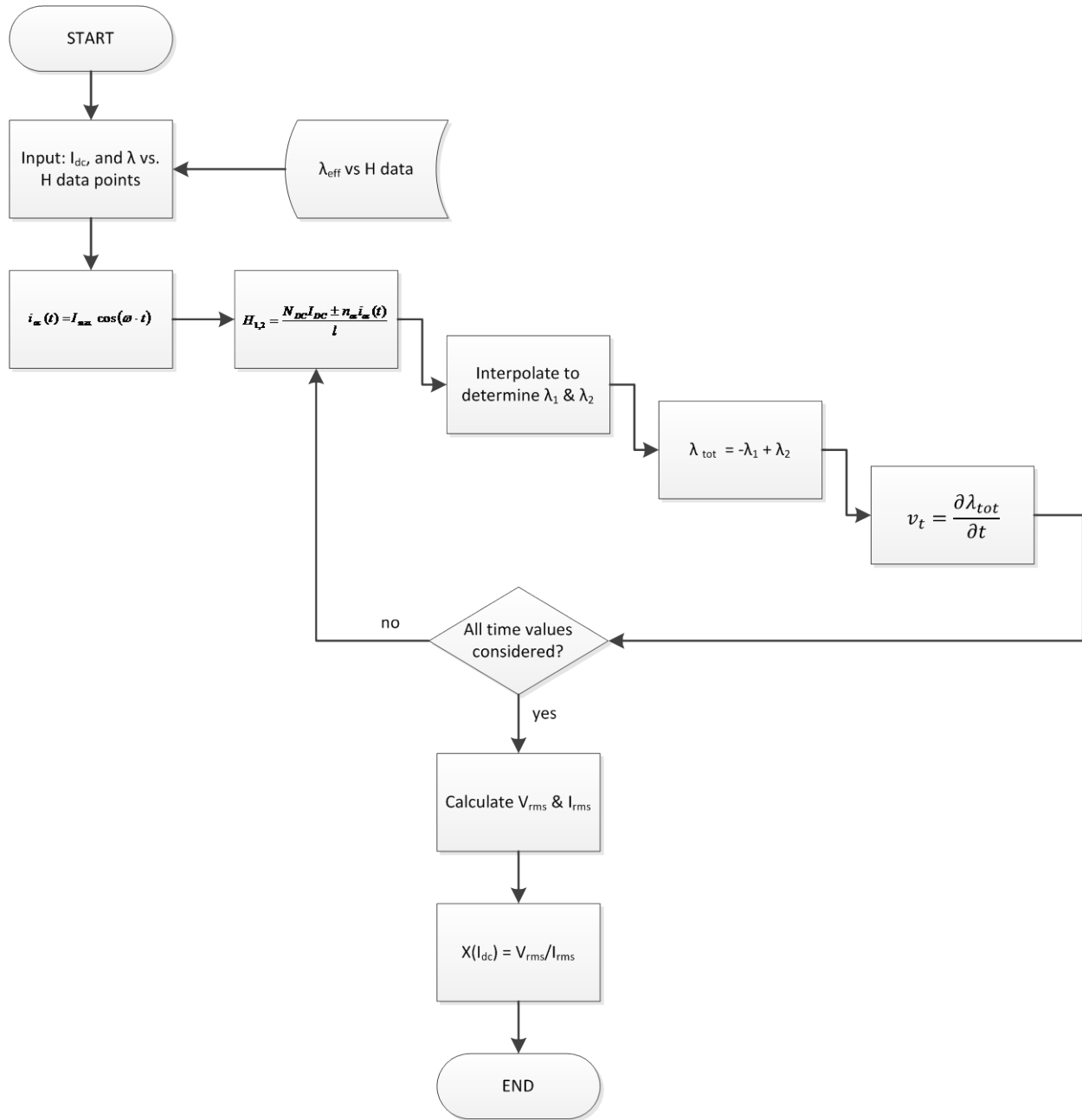


Figure 30 Numerical Procedure for Calculating Apparent Reactance using Two-Core Model

The block diagram in Figure 31 represents the final form of the simulation. The simulation includes conversion of the applied DC current to effective DC current. Inputs to the simulation include applied DC current, AC current defined over a time range, and the flux linkage profile with respect to magnetic field intensity. Each simulation was performed using a constant DC current and the flux linkage profile generated by FEA of the single-core model. The AC current was applied over one 60Hz cycle.

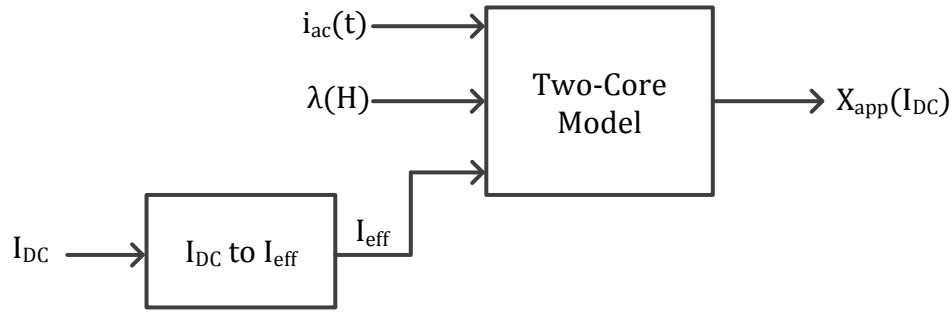


Figure 31 Block Diagram of Two-Core Model Simulation

Summary of Results

Simulations of the two-core model were performed using MATLAB® and the equivalent flux linkage profiles from FEA. The results include apparent reactance curves, or DC sweep, and differential inductance curves, or AC sweep. Results from the two-core model using single-core equivalent parameters are also presented and include application of the effective DC current and comparison to hardware tests.

Two-Core using Split-Core Equivalent Parameters

The differential inductance curves in Figure 32 and Figure 33 are from simulation of the two-core model using equivalent flux linkage curves. The results in Figure 32 were obtained using the gapless split-core parameters and the results in Figure 33 were obtained using parameters of the split-core with air gap. Both sets of curves include the differential inductance, core flux linkages, and total flux linkage as functions of instantaneous AC current. The curves are normalized in order to observe the relative differences in the curves based on the

implementation of the air gap and to observe the interaction of the flux linkages with respect to the differential inductance curves. Both sets of curves demonstrate the behavior of the two-core model for the operating range of DC current. The AC current is constant, $125A_{RMS}$, in both cases.

The differential inductance curves from simulation of the gapless split-core, Figure 32, are presented as an example to show operation typical of SCFCLs. The curves reveal that operation in the linear region is limited to a narrow range of instantaneous AC current. This behavior is reflective of the equivalent flux linkage of the gapless split-core in Figure 114, which has a large slope in the linear region. In addition to the width of the peaks, the magnitude of the peaks is approximately two orders of magnitude higher than the curves from the model with air gap. The high magnitude is consistent with the high slope since it is proportional to the derivative of the flux linkage with respect to the instantaneous AC current. When $I_{DC} \gg 0$, the location of the peaks with respect to the AC current indicates that the model operates in the saturation region for the entire AC current cycle. As I_{DC} approaches zero, the peaks combine and form a single peak of double the magnitude at $I_{DC} = 0$. The differential inductance curves in Figure 33 were determined from simulation of the split-core model with air gap. The curves indicate that the model operates in the linear region for a wide range of instantaneous AC current. In addition to the increased range, the magnitude of the peaks is two orders less than the gapless core. The result is reflective of the equivalent flux linkage curve for the split-core model with air gap in Figure 114, which has a gradual slope. At $I_{DC} = 0$, the normalized peak is 1 unit high and its width includes the entire instantaneous AC range (-175A to 175A). In this operating condition, the model operates at maximum differential inductance over the entire AC cycle. As I_{DC} increases, the peaks begin to separate and the differential inductance changes throughout portions of the AC current cycle. At $I_{DC} \gg 0$, the peaks separate enough that the model operates in saturation for the entire AC current cycle.

Comparison of the differential curves reveals significant differences between the models in terms of width and magnitude of the differential inductance peaks, and emphasizes the impact of air gap in the design. Consequently, the gapless core produces characteristics optimal for an

FCL application; where the DC is applied for constant operation in saturation and the peaks of the fault current drive the device into the saturation region. In contrast, the core with air gap is optimal for power flow applications as the change in the inductance profile is gradual and mostly uniform across the nominal AC current range.

The apparent reactance curves in Figure 34 and Figure 35 echo the behavior of the differential inductance curves. The curves for the gapless core in Figure 34 have apparent reactance that changes significantly for I_{DC} close to 0, which reflects full separation of the differential inductance peaks. The apparent reactance then gradually decreases as the differential inductance peaks are displaced throughout different portions of the AC current cycle. This trend continues until saturation is achieved when the peaks are fully displaced outside of the AC current cycle. Curves for the core with air gap in Figure 35 have similar behavior, but produce a more gradual change in apparent reactance with magnitudes more than two orders of magnitude smaller. For I_{DC} near 0, the values are constant as the width of the differential inductance peaks is large enough not to effect operation within the AC current cycle. The two peaks separate in three stages. First, the unity peak is reduced in width and portions of the AC current cycle result in operation at full and $\frac{1}{2}$ the unity differential inductance at I_{DC} near zero. Second, the peaks separate as $I_{DC} > 0$ and the model operates in saturation for instantaneous AC current close to zero and at $\frac{1}{2}$ peak differential inductance for instantaneous AC current close to the peaks of the current waveform. Third, the peaks are displaced by the DC current enough so that the entire AC current cycle results in saturation of the model.

The apparent reactance curves in both cases are significantly different in terms of their profile, but suggest similar behavior. The saturation point in both cases changes with respect to the RMS current as smaller RMS currents occupy a smaller instantaneous AC current range, which requires smaller DC current to displace the differential inductance peaks outside the range of the AC current cycle. The change of apparent reactance over the DC current range is reflective of the interactions of the differential inductance peaks as they are displaced.

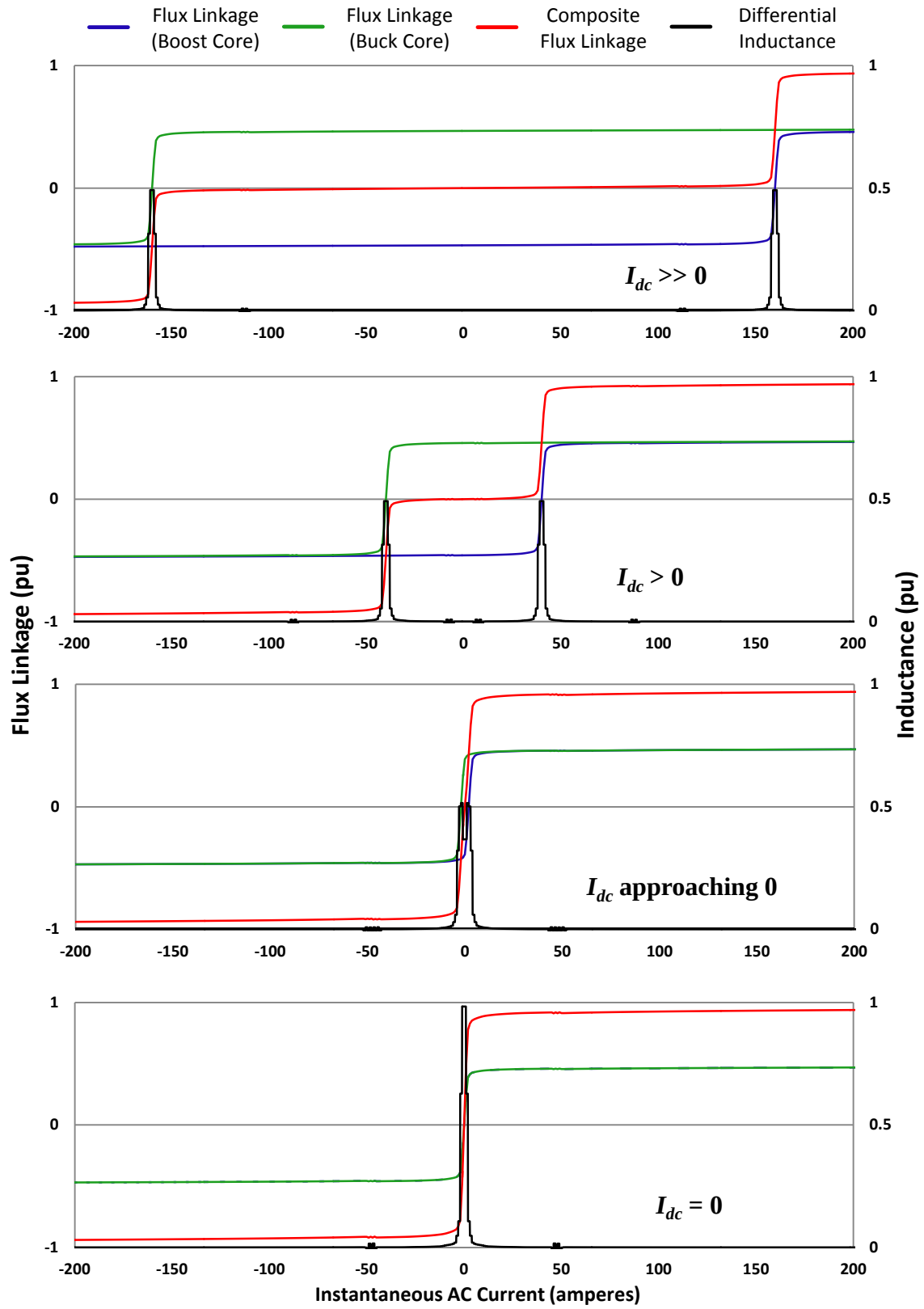


Figure 32 Two-Core Model using Equivalent Parameters without Air Gap

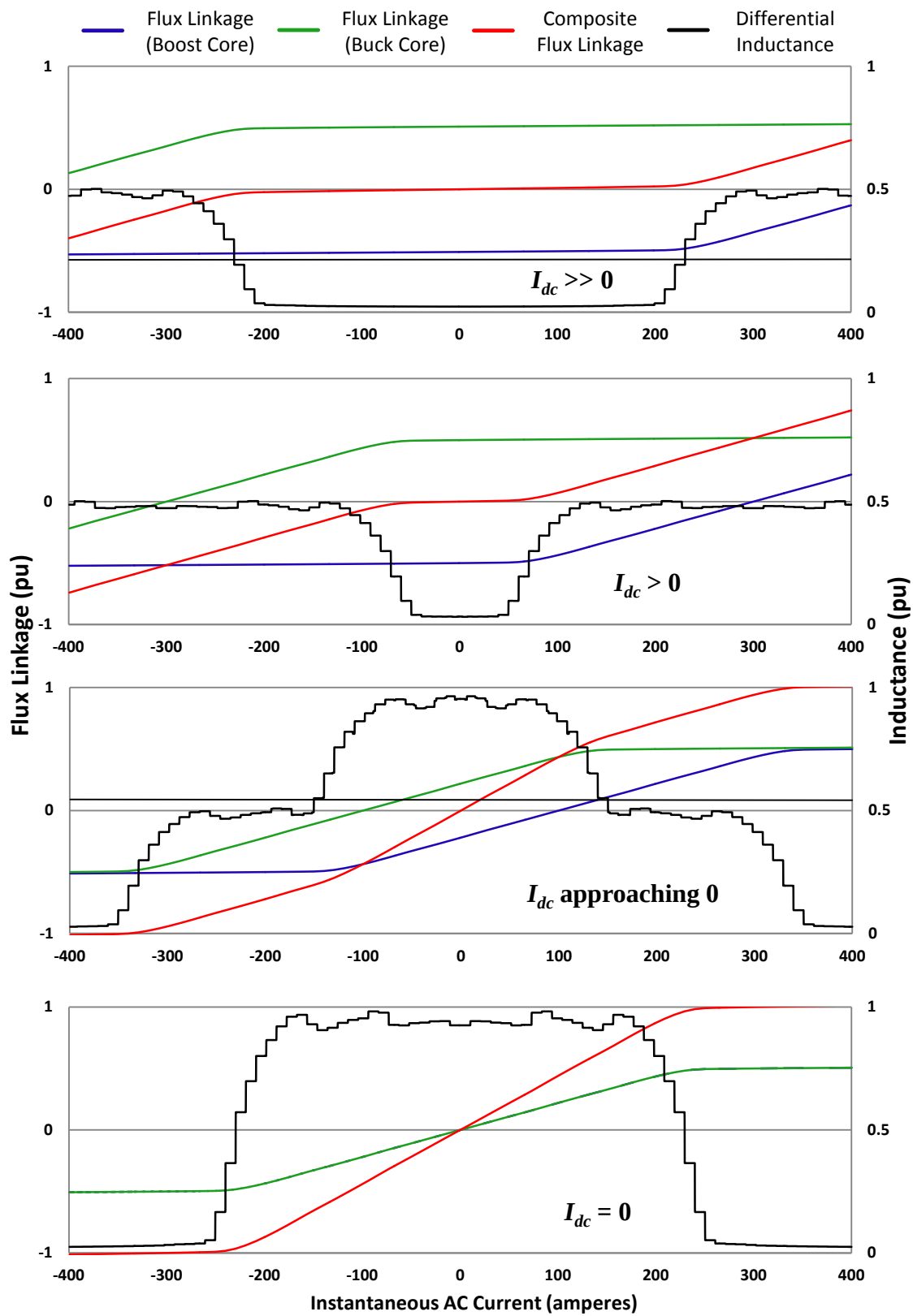


Figure 33 Two-Core Model using Equivalent Parameters with Air Gap

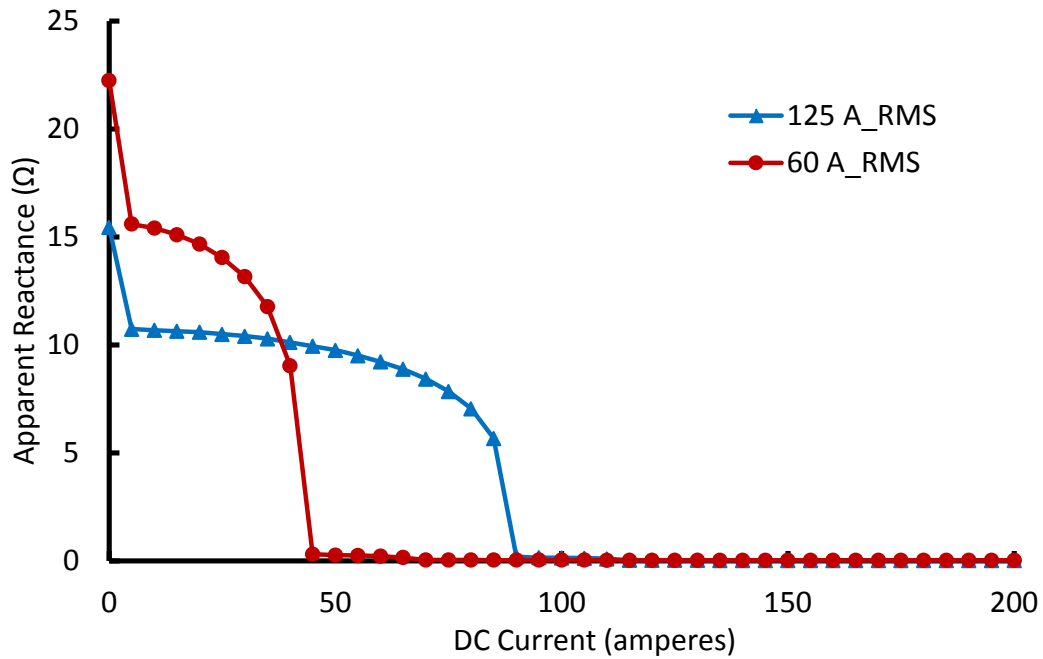


Figure 34 Reactance of Two-Core Model using Equivalent Parameters without Air Gap

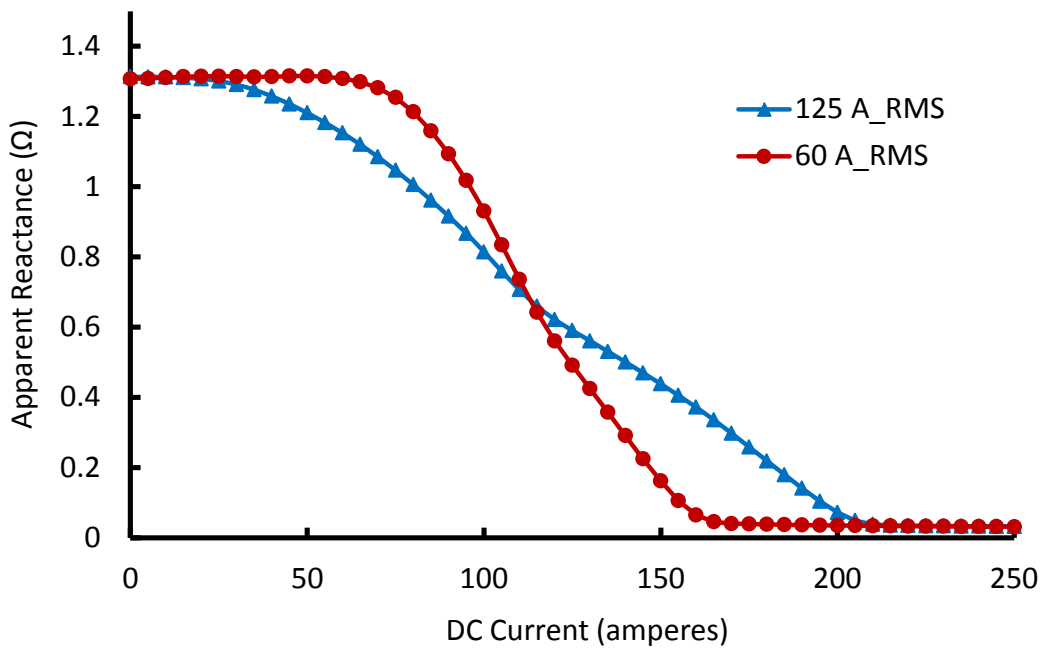


Figure 35 Reactance of Two-Core Model using Equivalent Parameters with Air Gap

Two-Core using Single-Core Equivalent Parameters

The applied reactance curves in Figure 36 are the result of FEA and MATLAB® simulation of the two-core model. The curves from FEA are the result of performing the DC sweep in the ANSYS® MAXWELL. The curves from the two-core model represent simulations in MATLAB® with application of the split-core equivalent parameters. The two-core results are identical to those in Figure 35 in every way except in magnitude, which is a factor of two smaller due to the reduction of flux linkage. There is correlation between the FEA and two-core model results, but there are two distinct differences. First, the magnitude of the apparent reactance from the two-core model at low values of DC is slightly lower than the FEA. Second, there is a small difference in the saturation levels. The differences between the results is possibly due to precision as the FEA analyzes the geometric structure of the core while the two-core model is based on average parameters.

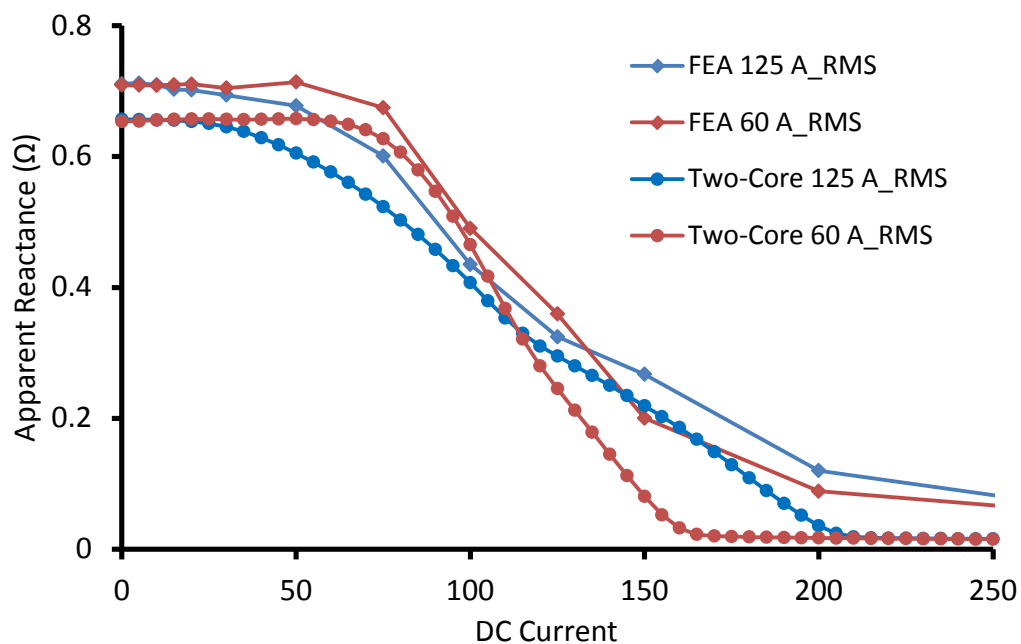


Figure 36 DC Sweep Results for FEA Split-Core and the Two-Core Model

Results from the two-core model are compared to the hardware test data in Figure 37. The curves are significantly different and appear to inflect in opposing directions. The only similarity between the curves is the apparent reactance at $I_{DC} = 0$. The hardware test curves decrease significantly for low values of I_{DC} before gradually descending towards saturation. In contrast, the two-core model curves hold constant for low values of I_{DC} followed by a rapid decent towards saturation. Since hardware test data is only available for I_{DC} up to 150A, it is unclear if saturation is achieved at the same point as the two-core model. The opposing behavior of the two sets of curves suggests that significant DC flux exists at small DC current in the two-core model. This result reveals that the split-core model in FEA and two-core model do not adequately represent the DC flux interactions that exist in the three-legged core.

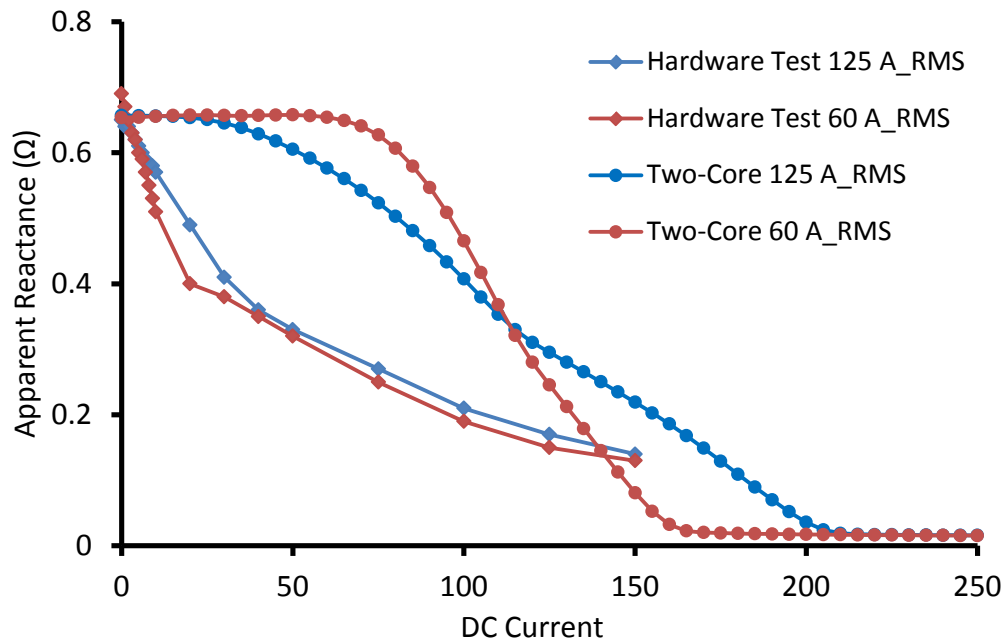


Figure 37 DC Sweep Curves from Hardware Tests and the Two-Core Model

Application of an effective DC current using the relationship in Figure 28 improves the behavior two-core model. Results from the two-core model with effective DC current are compared to the hardware tests in Figure 38. Curves from the two-core model now follow the behavior of the hardware tests with sharp transition for low values of DC current, followed by a gradual

transition to the saturation region. There are still differences between the two results. Most notable is the drop of apparent reactance for low I_{DC} is more significant with two-core model. The two-core model also achieves deeper saturation. Another observation is the divergence between the 125A_{RMS} and 60A_{RMS} two-core model curves after the initial drop and before the saturation point.

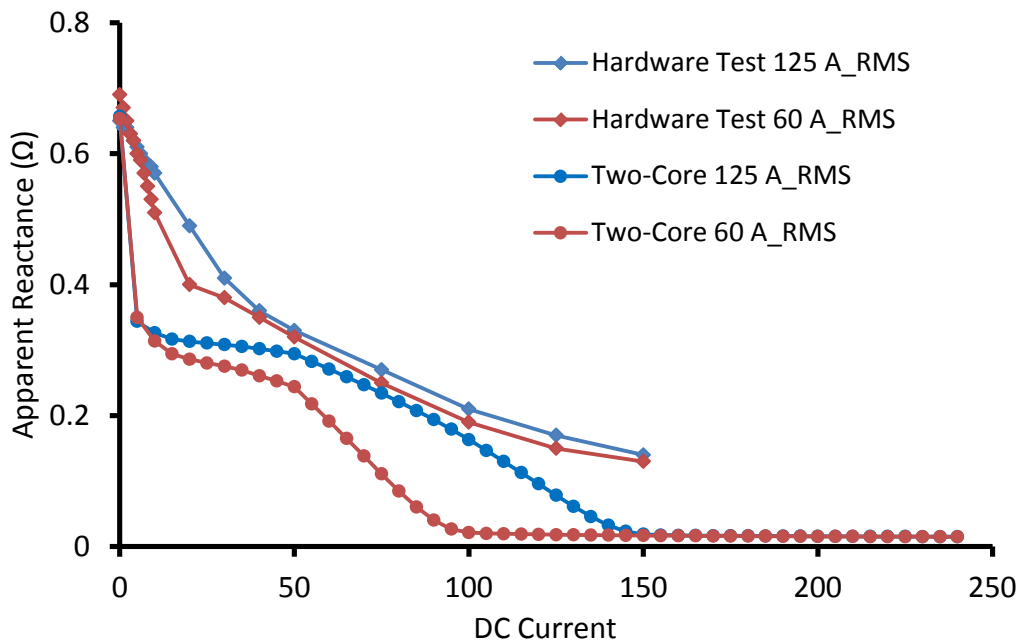


Figure 38 Reactance Curves from Tests and Corrected Two-Core Model

The two-core model is a simple method for modeling the behavior of the SRPFC. It provides insight into basic operating principles and is relatively simple to apply in simulation. Its simplicity is due to an assumption of ideal core characteristics. These ideal characteristics facilitate uniform distribution of B and H throughout each core, which allow expression of H directly in terms of AC and DC currents. However, practical designs for power flow control do not lend themselves to such assumptions. Multiple flux paths and air gaps are examples of elements that complicate modeling and require corrections to compensate for their impacts. The application of two equivalent parameters led to improved results with the two-core model. The first correction was the application of equivalent AC flux linkage derived from FEA. The

resulting flux linkage lumps the non-ideal core properties and represents these properties indirectly upon application to the two-core model. Subsequent simulation of the two-core model produced apparent reactance values similar to the hardware tests for zero DC current, but correlated poorly for non-zero values of I_{DC} . The second correction was application of an effective DC current to improve the DC flux, which produced apparent reactance curves following the general trend of the hardware tests. While a vast improvement, divergence of the curves for mid-range DC currents did not match the hardware tests. This implies that the two-core model does not adequately represent the SRPFC.

In the end, the final version of two-core model was not able to match the measurements with desired accuracy. Additional corrections could provide further improvement to the two-core model, but would likely increase its complexity. For these reasons, it was decided to suspend further development of the two-core model and focus on methods more appropriate for modeling devices with non-ideal magnetic properties.

CHAPTER III: MODEL DERIVED BY FINITE ELEMENT ANALYSIS

The two-core example described in Chapter II assumes that the cores are identical and homogenous, with a model represented by closed form equations. However, it was determined that the three-legged core, cannot be expressed in closed form equations due to multiple non-linear reluctance paths. Thus, the problem requires a numerical approach. One option is to build an electromagnetic circuit model that represents all branches; then use a robust solver to calculate the flux flows [38] [40]. Another method is to use FEA to determine the flux linkages of the device under specified DC biases. This Chapter describes the latter option since finite element analysis is a more fundamental approach when there is limited experience on how the physical device behaves and operates. Design engineers commonly use FEA in design of electromagnetic devices due to its ability to provide detailed analysis of complex geometries that have non-linear magnetic characteristics. FEA is also applicable to modeling both steady-state and transient behavior [47].

A drawback of FEA for transient simulations is the time required to resolve all of the mesh calculations for each time step, which can lead to significant time durations to achieve solutions. Additionally, FEA does not lend itself to transient circuit simulations unless interfaced to a transient circuit simulator. To address both of these challenges, this chapter focuses on the development of a transient model of the SRPFC based on FEA and uses the subsequent results to construct a transient model in PSCAD™ [48]. The FEA provides the differential reactance profile of the SRPFC over a specified range of DC and instantaneous AC operating currents. The FEA results form a two-dimensional data matrix that a transient circuit simulator interpolates to simulate SRPFC operation under nominal and transient conditions. A similar method has been used to model the behavior of magnetically-controlled ballasts with encouraging results [44].

Transient circuit simulations are performed in PSCAD™ using a data matrix and compared to transient circuit co-simulation with FEA to verify that the data matrix method preserves the behavior of the SRPFC. The results show strong correlation between the data matrix approach and co-simulation with FEA. The data matrix approach achieves significant increase in

simulation speed over co-simulation. Additionally, steady-state simulations are compared to hardware test results obtained at ORNL.

Finite Element Analysis

Model Description

The finite element model (FEM) represents a single phase of the SRPFC prototype tested at ORNL. This design has a three-legged core with DC windings on the outer legs and an AC winding on the middle leg. A strategically placed air gap located in the middle leg achieves the specified output inductance when the SRPFC operates in the linear region. The construction of the core consists of stacked laminations of TRAN-COR[®] H grain oriented steel with a 2mm air gap on the center leg. The DC windings are located on the outer legs and consist of 24 turns each; they are connected in series for a total of 48 turns. Two series connected AC windings are evenly distributed on both sides the air gap. The model was constructed in three-dimensions in order to adequately capture significant core features and magnetic paths. The FEM was constructed and simulated using ANSYS[®] Maxwell. A caption of the three-dimensional FEA model is shown in Figure 39.

FEA Simulation and Results

Simulation of the FEM was performed by applying a fixed current on the DC windings and a current ramp on AC windings. The current ramp was of fixed rate to create a monotonically changing flux within the center leg of the core. The ramp initiates at 0A and terminates at 4kA, which is sufficient to represent operation of the device under nominal and fault conditions (for peak fault currents up to 20 times the nominal rating). The FEA calculates the flux linkage of the AC windings for each simulation step resulting in a flux linkage curve that is a function of the ramped current. This process was repeated by incrementing the DC current and reapplying the current ramp on the AC windings. The resulting set of flux linkage curves, shown in Figure 40, spans the specified DC current range. For initial DC current values, the flux linkage curves resemble the B-H curve in shape. As the DC current increases, there is a noticeable decrease in the slope of the linear region. As the DC current approaches the maximum, the initial portion of

the flux linkage curve forms two linear regions prior to the knee-point. All of the flux linkage curves converge to a common slope beyond the knee-point. An enlarged view of the flux linkage curves, shown in Figure 41, provides greater detail of SRPFC operation leading into the knee-point. These curves show that the SRPFC is always in linear operation under nominal conditions, and that adjustment of the DC current simply alters the shape of the linear region. This behavior is different than the two-core model flux linkage described in Figure 33, where the model could achieve full saturation. The behavior in Figure 41 suggests that another flux path exists that limits the device from achieving full saturation. The flux linkage curves form a three-dimensional surface as shown in Figure 42.

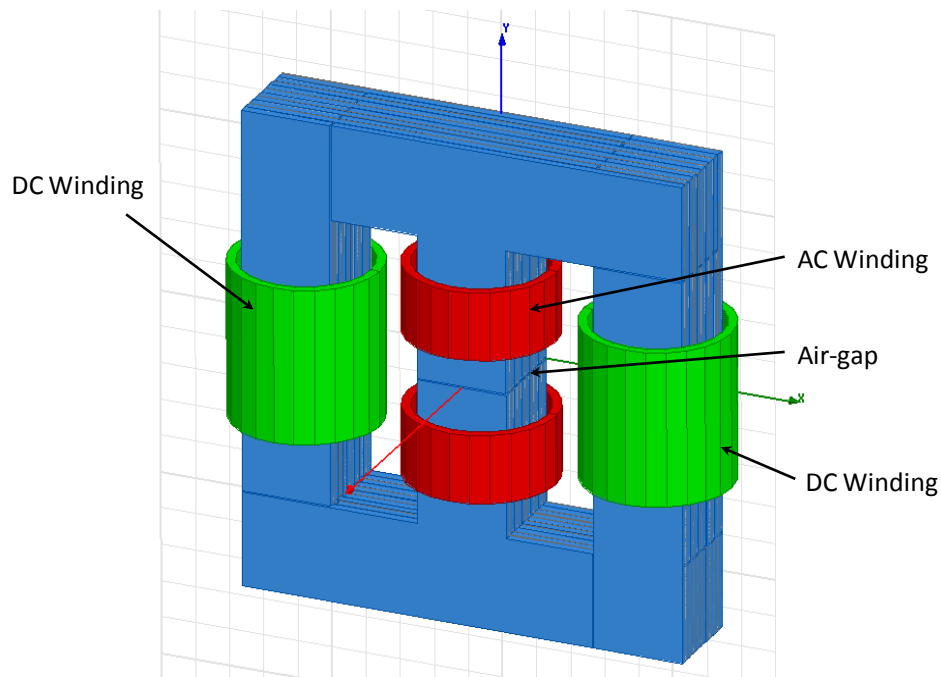


Figure 39 Caption of Three Dimensional FEA Model

The differential inductances are determined by taking the difference of the flux linkage curves with respect to the instantaneous current values as in (2.10). The resulting differential inductances, shown in Figure 43, are functions of the DC current and instantaneous AC current and reflect the intended operation of the SRPFC under nominal and transient operating conditions. The differential inductances are highest within the nominal AC current range for

lower DC currents, then decrease within the nominal AC current range as the DC current increases. In addition to nominal operation, the flux linkage curves represent operation of the SRPFC for instantaneous AC currents beyond the nominal range. Once the instantaneous AC current exceeds nominal levels, the SRPFC will operate in saturation mode, resulting in the minimum differential inductance. Therefore, fault currents will dynamically change the operating state of the SRPFC within a half cycle. Under fault conditions the SRPFC will tend to have high inductance near the zero crossings, then transition to the saturation region as the fault current ascends to its peak. Since the high impedance occurs near the zero crossing, the SRPFC may not significantly affect the fault current, but has the potential to cause high voltage spikes corresponding to these points at the terminal of the device.

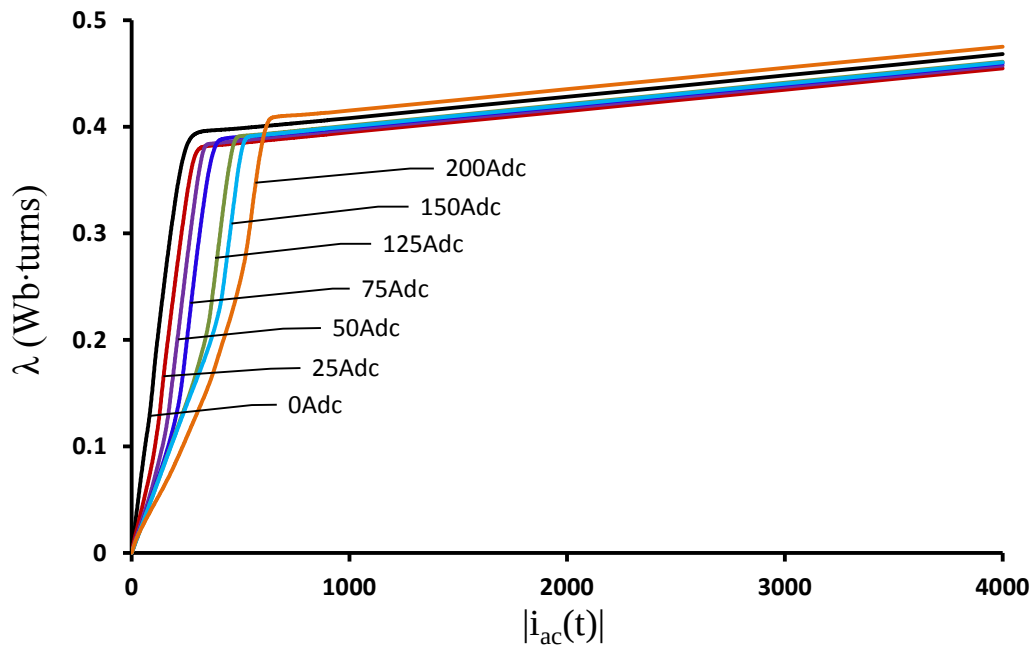


Figure 40 Flux Linkages Curves from FEA

The differential inductances in Figure 43 are defined for positive values of AC current, as they are assumed to be symmetrical for negative AC currents. Since each flux linkage curve requires approximately 3 days of simulation time, only positive AC current values were simulated to in

order to generate enough curves to adequately address the DC current range within a realistic project timeline.

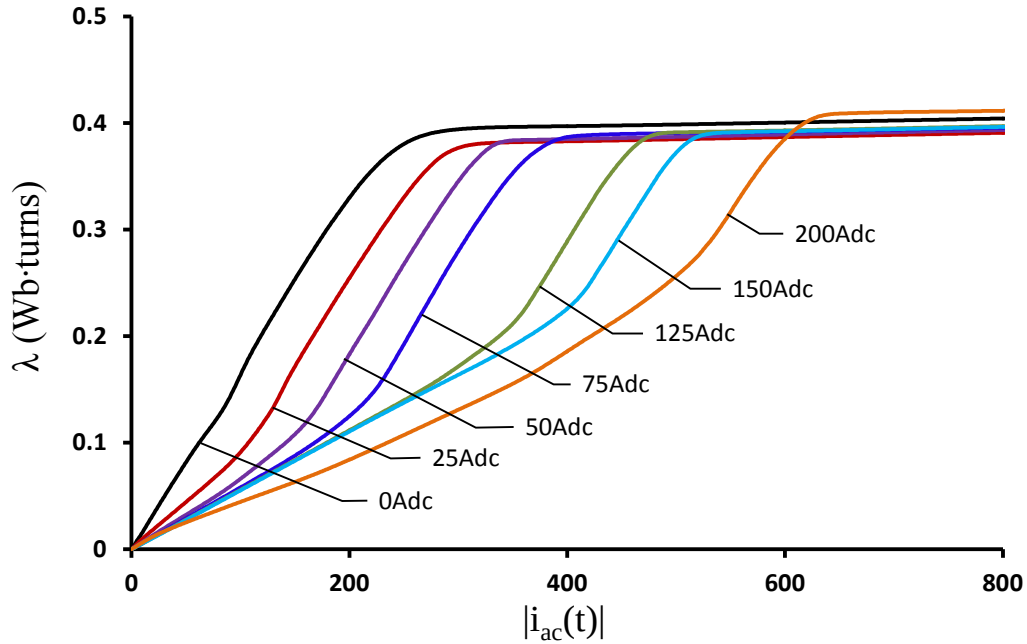


Figure 41 Flux Linkage Curves from FEA (shown close to the knee-point)

Application in Transient Simulation

The FEA results form a two-dimensional matrix of data values that describe the differential inductance as a function of the AC and DC currents. During transient circuit simulations, approximations of the differential inductance values from the matrix require bilinear interpolation based on input values of AC and DC current. Bilinear interpolation not only approximates differential inductance values along a defined curve, but it also interpolates between curves of two DC levels. Therefore, model accuracy is dependent on the size of the DC current increments. For the model defined within this study, the seven differential curves, corresponding to seven DC currents, defined in Figure 43 are sufficient to represent the SRPFC over the specified range of DC current (0 to 200A_{DC}).

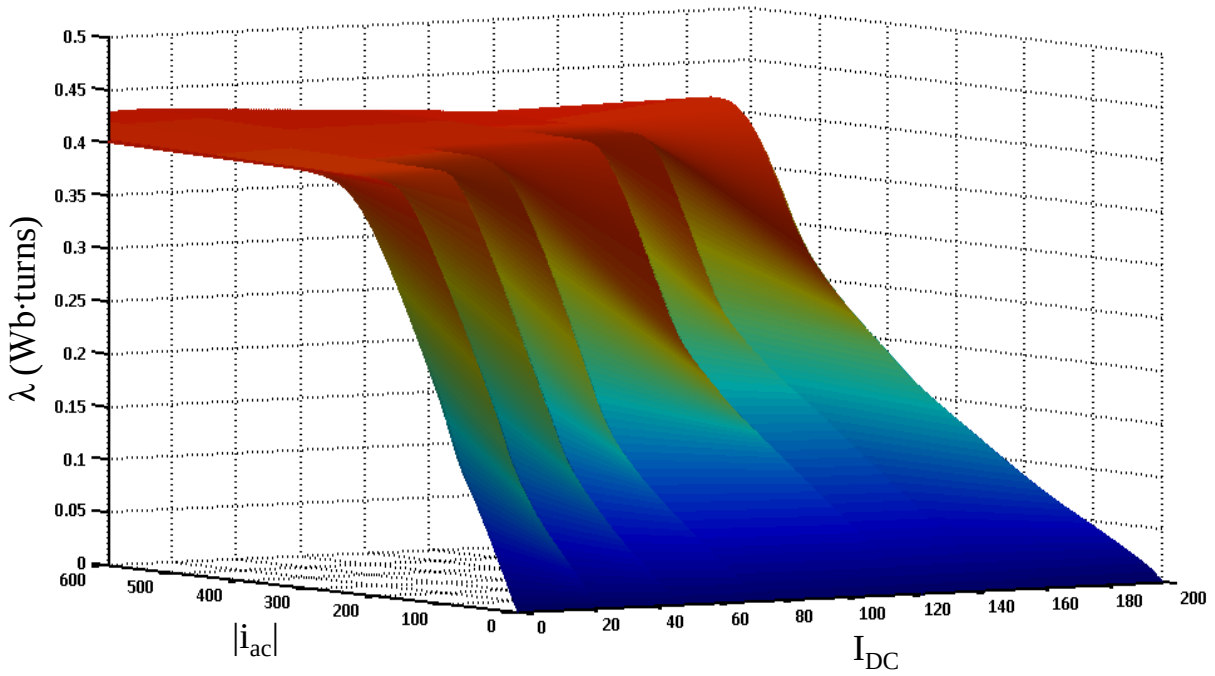


Figure 42 Surface Plot of Flux Linkage as Functions of AC and DC Current

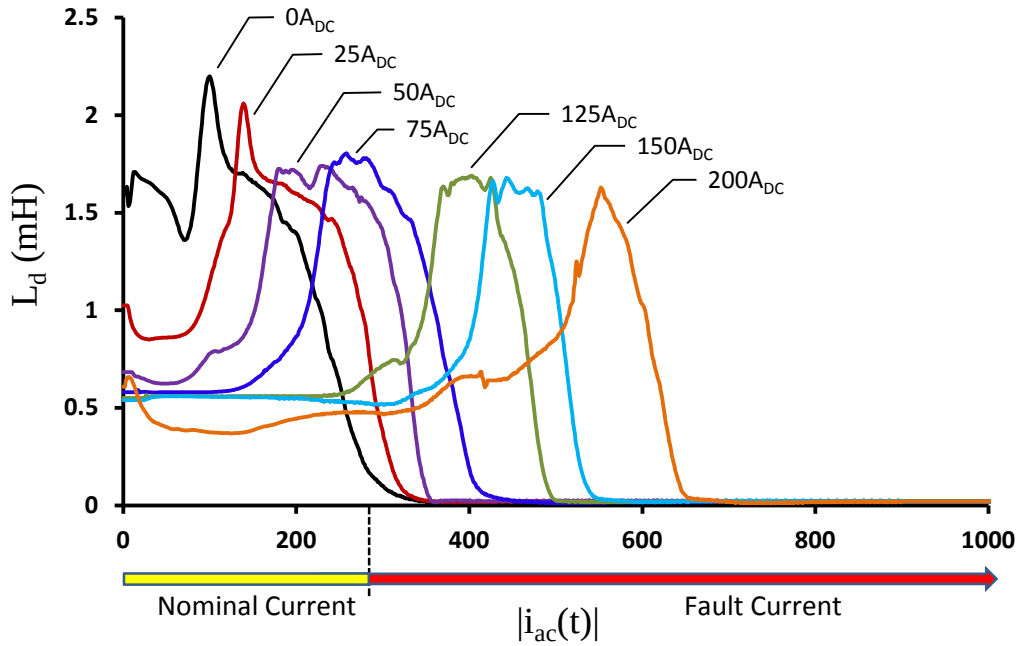


Figure 43 Differential Inductance Curves Derived from FEA Results

Application of the FEA-derived model in transient circuit simulation is describe by the block diagram shown in Figure 44. The user-input is the DC current, which is fed directly to the model. The AC current is an instantaneous value provided by the transient simulator for every time step. The model uses the absolute value of the instantaneous AC current for interpolation as the data matrix is defined only for positive values. The absolute values are a valid input since the differential inductance curves are an even function of the instantaneous AC current. Another way to represent the model would be to define the data matrix for positive and negative values of AC current, then feed the instantaneous AC current directly into the model.

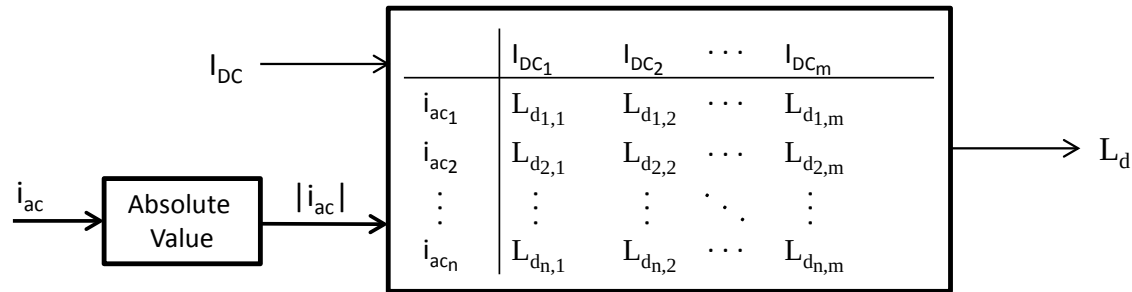


Figure 44 Block Diagram Describing the Application of FEM Results into a Transient Circuit Simulator

Application to PSCAD™

The data matrix is applied to PSCAD™ as shown in Figure 45. PSCAD™ supports a look-up table element that is compatible with the data matrix containing the SRPFC differential inductances. The look-up table element requires a specific data format within a text file (.txt). To conform to format requirements, a MATLAB® script (APPENDIX E) was written to convert the data accordingly. The look-up table element drives a variable inductor element that is connected in series with the transmission line. Figure 45 shows application of the data matrix in PSCAD™, where the SRPFC is simulated under nominal current conditions.

Fault Simulations

The SRPFC was simulated under fault conditions within the simple power system model shown in Figure 46. The single-phase system is driven by a 277V_{LN} ideal voltage source with resistive

source impedance. The value of the resistive impedance was chosen to achieve a symmetrical fault current of 4kA_{peak} . A resistive load provides prefault current conditions, which was set to achieve almost full rated current. Table 2 describes the fault simulation parameters. The PSCAD™ fault simulation, shown in Figure 47, uses the data matrix approach to model the SRPFC. Simulations were performed with time steps of $100\mu\text{s}$ for six electric cycles at 60Hz (one cycle of nominal conditions followed by five cycles of applied fault).

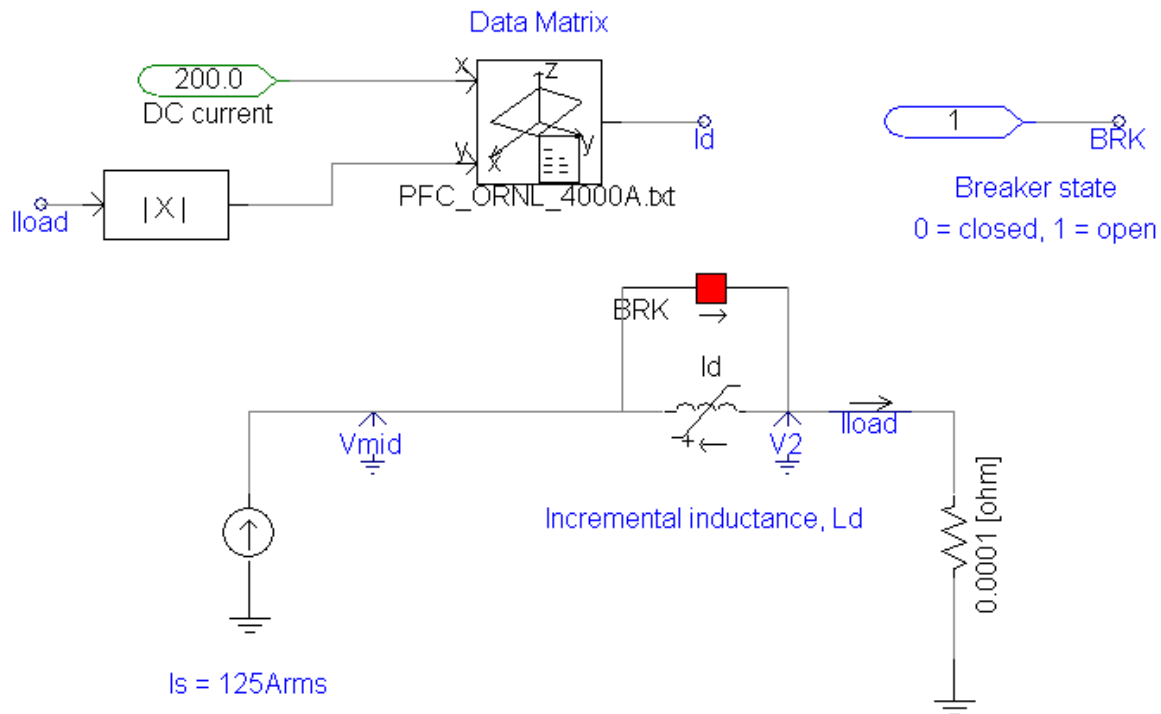


Figure 45 Nominal Simulation Circuit in PSCAD™

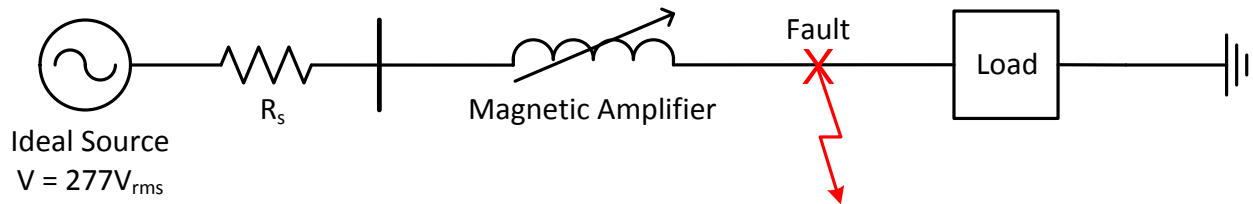


Figure 46 One-Line Diagram of Single-phase Fault Simulations

Table 2 Parameters of Fault Simulation

Parameter	Description	Value
R_s	Surge resistance	97.95m Ω
L_s	Surge inductance	negligible
R_{Load}	Load resistance	2 Ω
R_{fault}	Fault resistance	0.1m Ω
t_{fault}	Fault initiate time	16.67ms (1 cycle)
Δt	Time step	100 μ s
t_{end}	Duration of simulation	100ms (6 cycles)

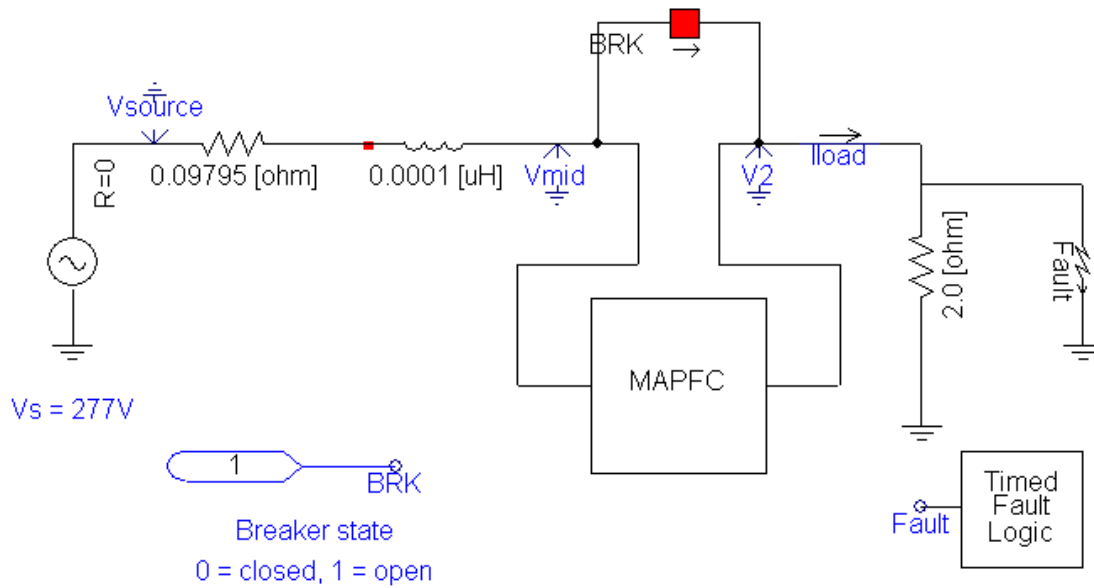


Figure 47 Fault Current Simulation Circuit in PSCAD™

Co-Simulation of FEM and Transient Circuit

The ANSYS® software suite includes a transient circuit simulator, ANSYS® Simplorer, that can interface the FEA in ANSYS® Maxwell. This co-simulation ability enables the FEA of a device to be simulated within the context of a power system. During each time step, ANSYS® Simplorer communicates with ANSYS® Maxwell to solve the FEA and import the results to solve the electric circuit. The nominal and fault simulation models are shown in Figure 48 and Figure 49. These models are replicas of the system described by Figure 46 and Table 2.

The co-simulation method is comprehensive in terms supporting device detail when there is no electric equivalent circuit available. However, this capability comes at the cost of simulation time. In this case, it required approximately one week to complete six cycles of simulation. In contrast, the PSCAD™ model took seconds to solve. The $100\mu s$ time step was applied in all cases.

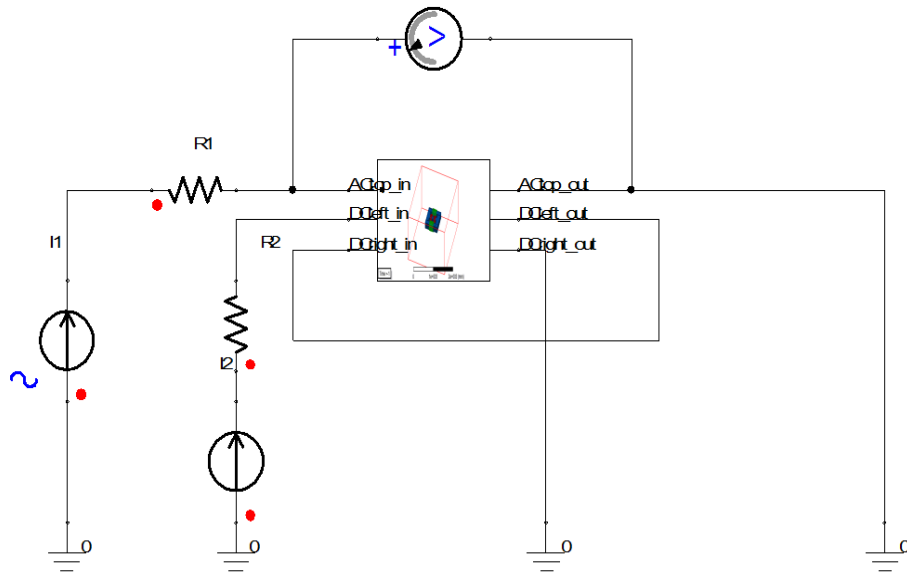


Figure 48 ANSYS® Simplorer Simulation for Nominal Conditions

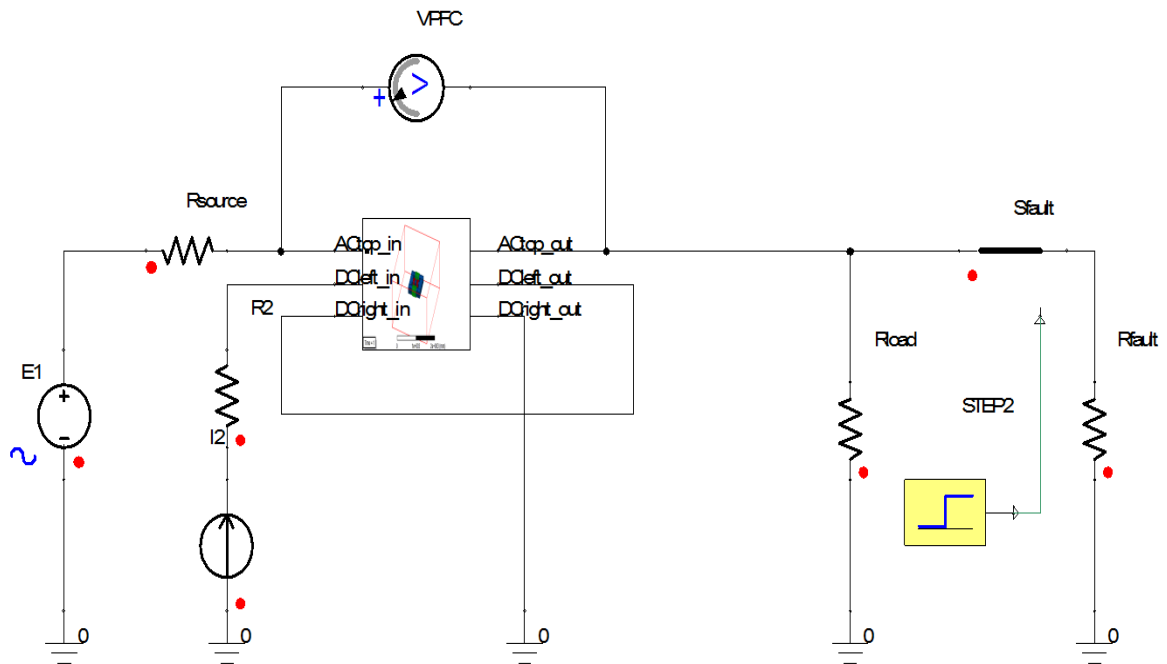


Figure 49 Co-Simulation using ANSYS® Maxwell and ANSYS®Simplorer

Simulation Results

Nominal Conditions

The data matrix approach using PSCAD™ and the co-simulation method were simulated under nominal conditions. For these simulations, $125A_{RMS}$ was applied to the AC windings using an ideal current source for several electrical cycles at 60Hz. The apparent reactance was estimated by taking the ratio of the RMS voltage across the terminals of the SRPFC to the applied RMS current. The process was repeated for values of DC current within the 0-200A_{DC} range. Figure 50 compares the results of both simulations to hardware of the ORNL prototype. The simulations produced nearly identical results, thus providing a degree of confidence that the data matrix approach maintains integrity of the FEA solution with minimal loss of resolution. Both simulations correlate well with the test results, especially at the lower range of DC current.

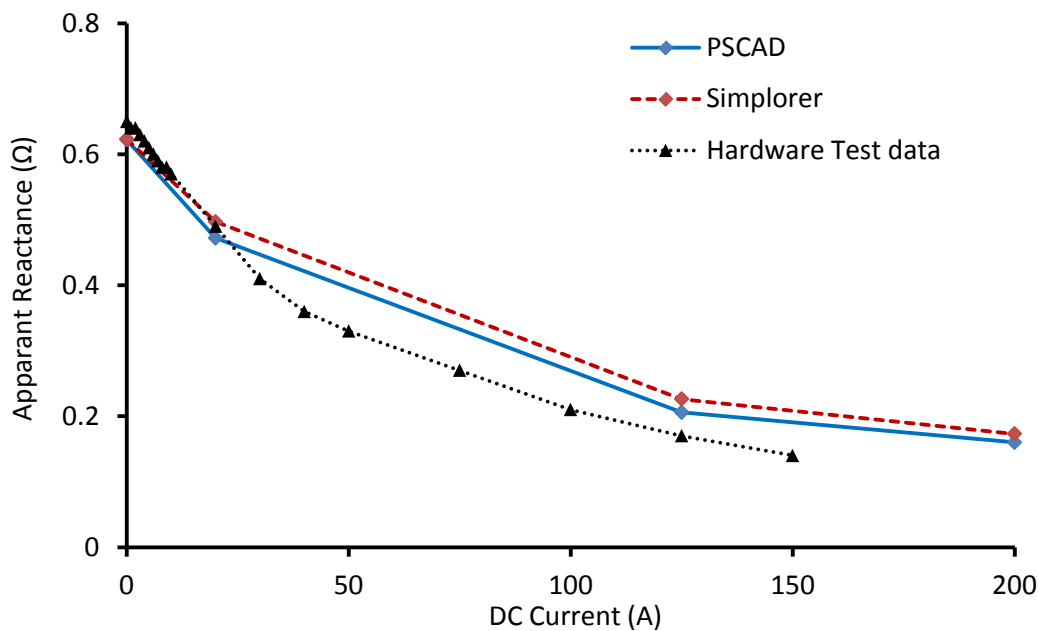


Figure 50 Apparent Reactance Curves Acquired from Hardware Tests, PSCAD™, and ANSYS® Simplorer

Figure 51 shows waveforms of nominal state simulations over the course of three cycles at DC currents of 0, 20, 125, and 200A_{DC}. The voltage represents the voltage drop across the terminals of the SRPFC. In all four cases, the data matrix approach using PSCAD™ is nearly identical with the co-simulation (ANSYS® Simplorer). There are subtle variances that occur where ANSYS® Simplorer appears to experience numerical stability issues. These numerical instabilities occur where there the differential inductance changes significantly with respect to the instantaneous AC current.

Fault Conditions

The results of fault simulations shown in Figure 52, Figure 53, and Figure 54 represent DC bias currents of 0, 75, and 200A_{DC}. The co-simulation and the data matrix approach are an identical match in the 0A_{DC} case. For 75A_{DC} and 200A_{DC}, there are small deviations at the current peaks and zero crossings of the voltage. Since the faults were applied at the SRPFC, at zero distance from the device in terms of line length, the fault impedance is purely inductive and the voltage is 90 degrees out of phase with respect to the current.

The fault currents in all three cases are impacted by the changing inductance of the SRPFC near the zero crossings as the lower instantaneous AC current values cause the device to transition into its linear region of operation. The impact of the SRPFC on the fault current is magnified by the relative weakness of the power system, which consists of an ideal generator, the SRPFC, and a load. In an interconnected system, which will likely resemble an infinite bus, the SRPFC will have less impact on the fault current.

Adapting the Model for Transmission Studies

Scaling from 277V to 66.4kV

The data matrix approach reflects the characteristics of the 277V, 200A prototype device, but full-scale application of the SRPFC is expected to occur at 115kV (66.4kV line-to-neutral), 1.5kA. The specifications of the prototype and requirements of the full-scale device are described in Table 3. The transient model of the SRPFC is not dependent on voltage rating since insulation design is a topic for another study. However, the model is dependent on current rating as it

affects the number of turns, the flux, and the overall size of the core. The apparent reactance requirements for the full-scale device are 5.1Ω at $0A_{DC}$ and 1.4Ω at $200A_{DC}$; these values represent the fundamental frequency component. The rated fault current is 20 times the nominal current rating ($30kA_{RMS}$), which scales down to the 4kA level achieved in the prototype simulations.

Table 3 Requirements for Prototype and Full-Scale SRPFC

Parameter	Description	Prototype	Full-Scale
X_{max}	Maximum apparent reactance	0.64Ω	5.1Ω
X_{min}	Minimum apparent reactance	0.16Ω	1.4Ω
I_{rated}	Rated AC current	200A	1.5kA
V_{rated}	Rated AC voltage (line-to-neutral)	277V	66.4kV
$I_{DC\ max}$	DC bias current for minimum reactance	200A	200A
I_{Fault}	Fault rating	4kA	30kA

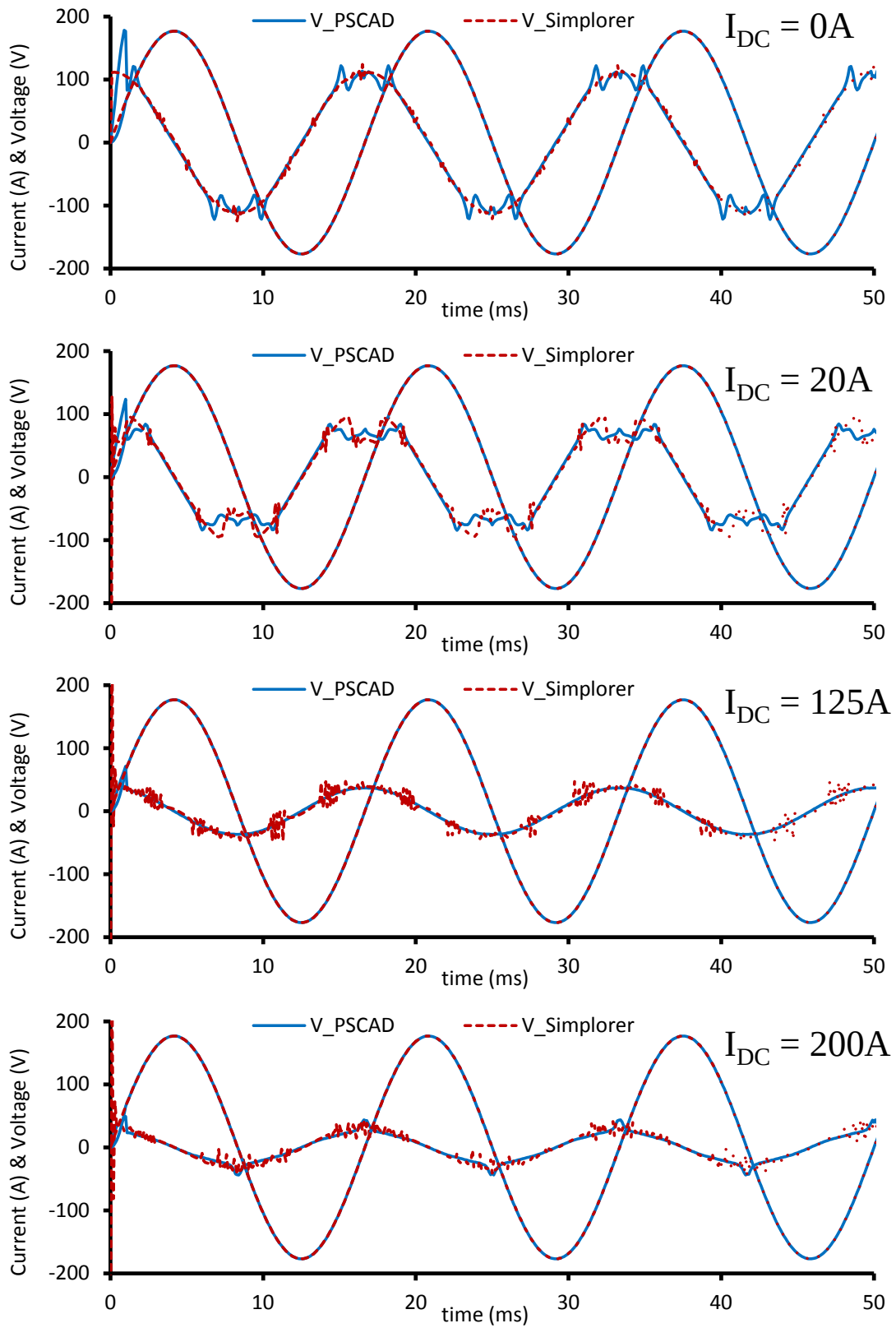


Figure 51 Nominal State Simulations Results from PSCAD™ and ANSYS® Simplorer

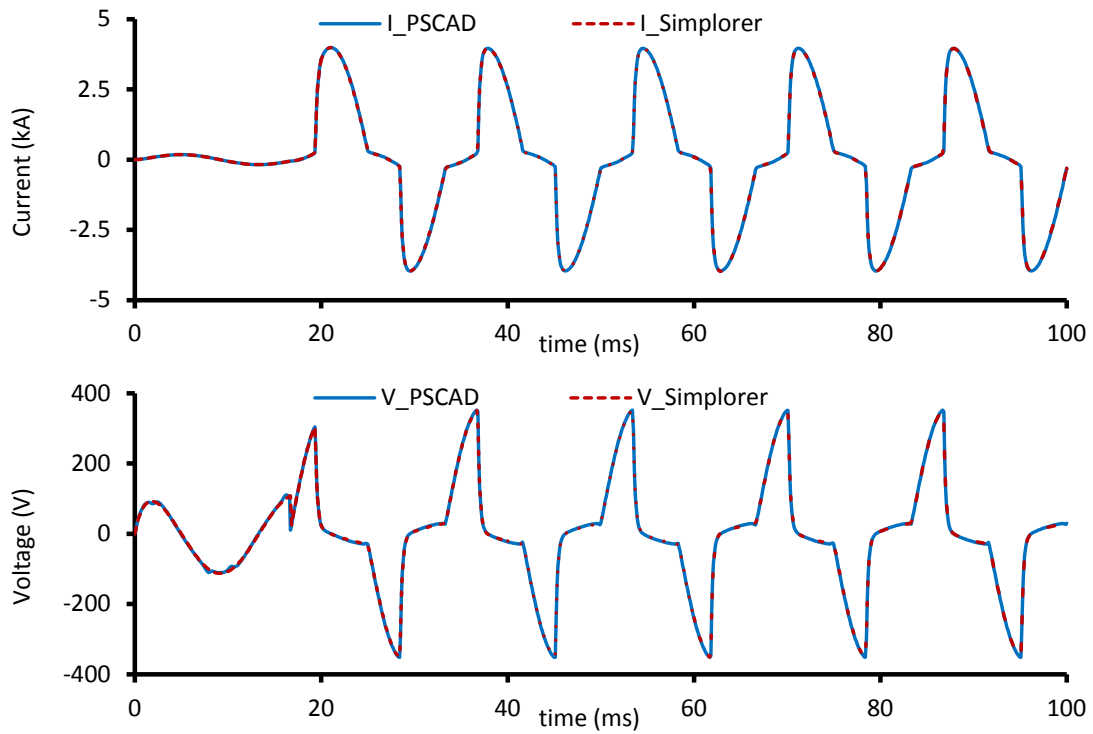


Figure 52 Comparison of PSCAD™ and ANSYS® Simplorer Fault Simulations for $I_{DC} = 0A$

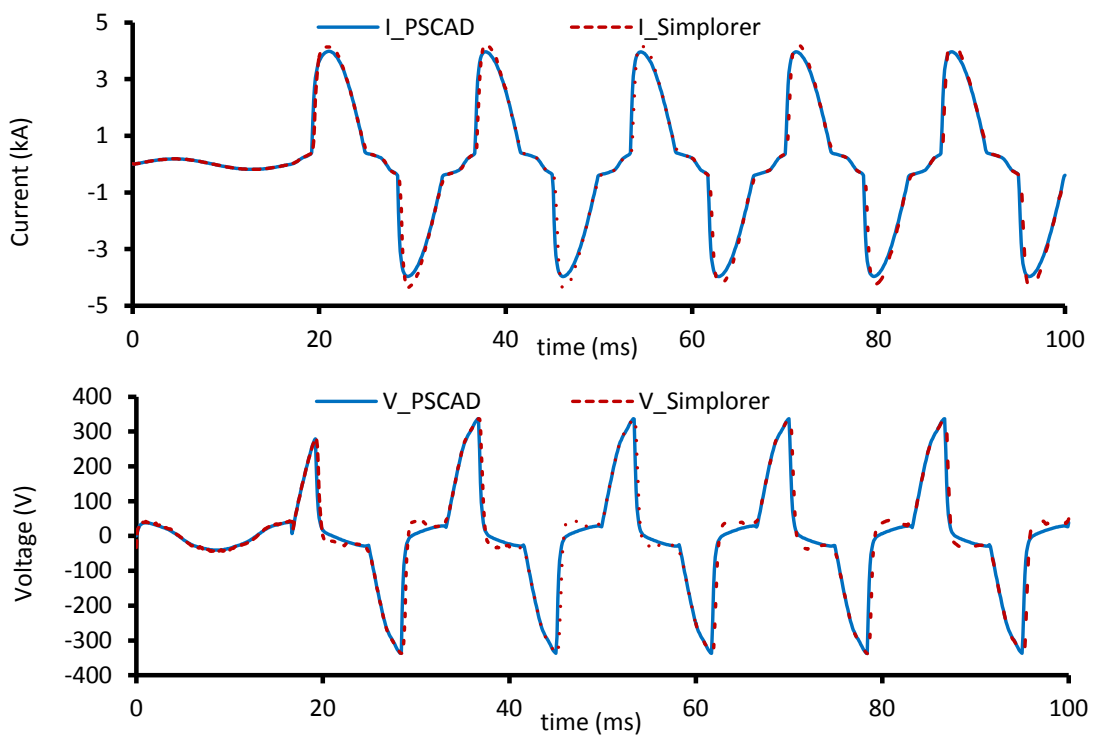


Figure 53 Comparison of PSCAD™ and ANSYS® Simplorer Fault Simulations for $I_{DC} = 75A$

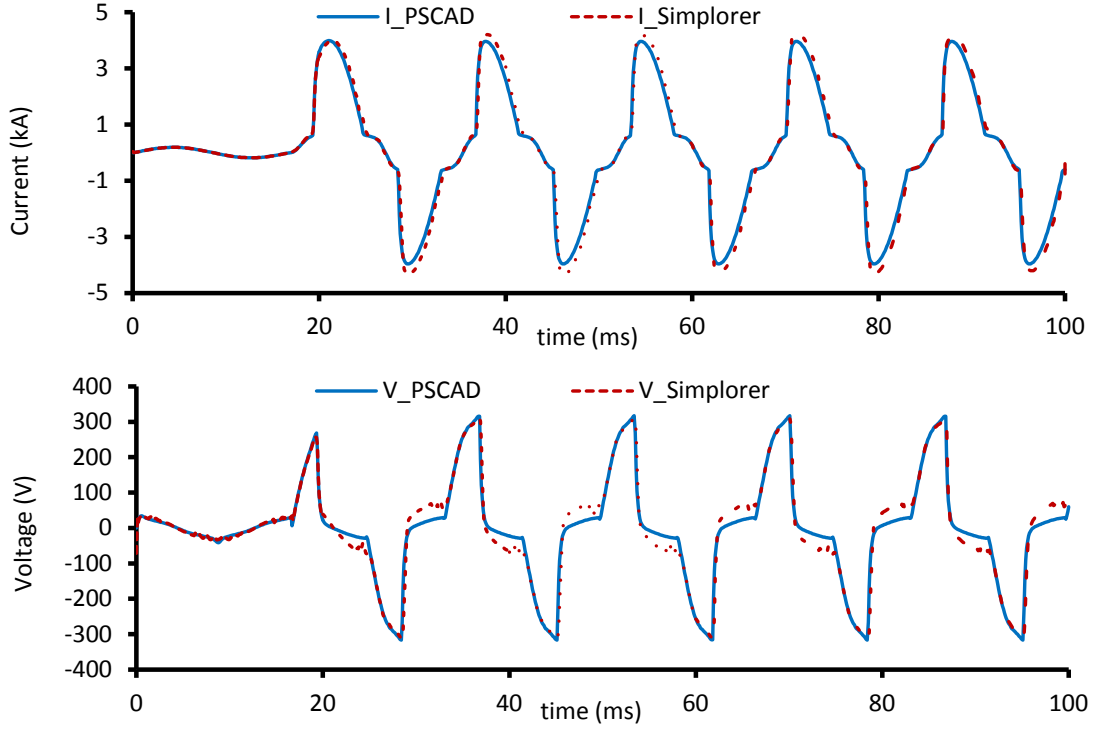


Figure 54 Comparison of PSCAD™ and ANSYS® Simplorer Fault Simulations for $I_{DC} = 200A$

To realize a full-scale model of the SRPFC, the prototype model was scaled using the requirements listed in Table 3. This was accomplished through scaling the differential inductance curves in Figure 43 using a two-step process. First, the instantaneous current axis was scaled by the ratio of the current ratings as described in (4.1).

$$current\ scaler = \frac{i_{rated(full-scale)}}{i_{rated(prototype)}} \quad (4.1)$$

Secondly, the magnitudes of the differential inductances were scaled approximately by the ratio of the maximum apparent reactance of the full-scale device to that of the prototype as described in (4.2).

$$L_d\ scaler = \frac{X_{max(full-scale)}}{X_{max(prototype)}} \quad (4.2)$$

The scaling factors were applied to the differential inductance curves in Figure 43. The full-scale differential inductance curves are shown in Figure 55. Application of the full-scale curves to the data matrix followed by simulation in PSCAD™ nets the apparent reactance curves in Figure 56. The apparent reactance curve matches the prototype in terms of shape and meets the magnitude requirements for the full-scale model. The table within Figure 56 shows that the $X_{app} = 5.11\Omega$ at $0A_{DC}$ and $X_{app} = 1.43\Omega$ at $200A_{DC}$.

Application of Full-Scale Model to PSCAD™

The requirements listed in Table 3 describe a single-phase device. Three-phase application requires deployment of three single-phase devices. Figure 57 describes three-phase application of the full-scale model in PSCAD™, where each phase is equipped with a series-connected variable inductor that is driven by the data matrix.

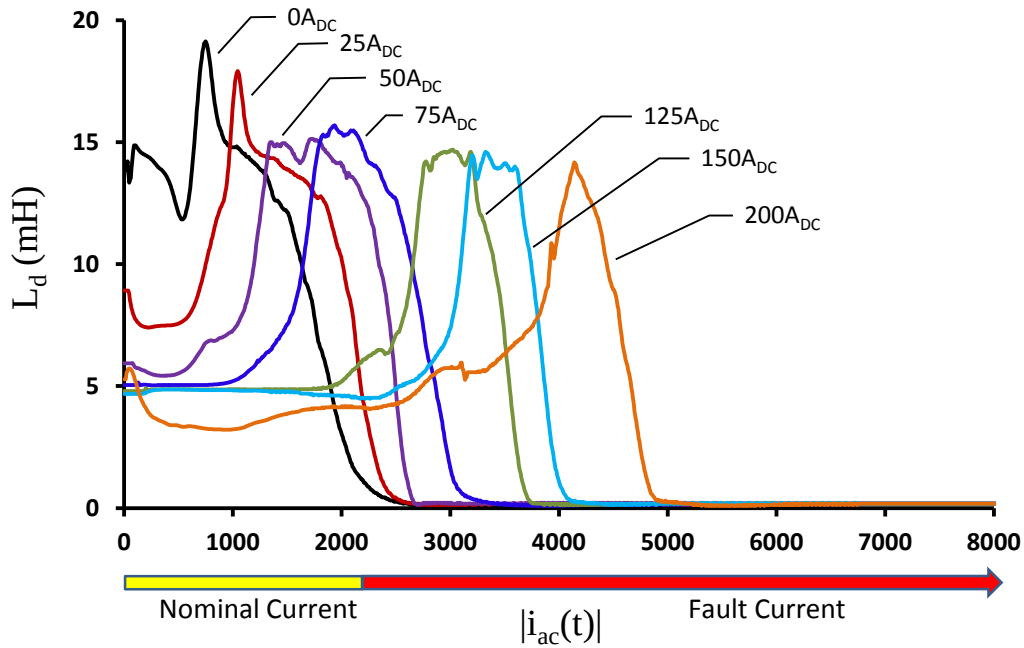


Figure 55 Differential Inductance Curves for Full-Scale Model

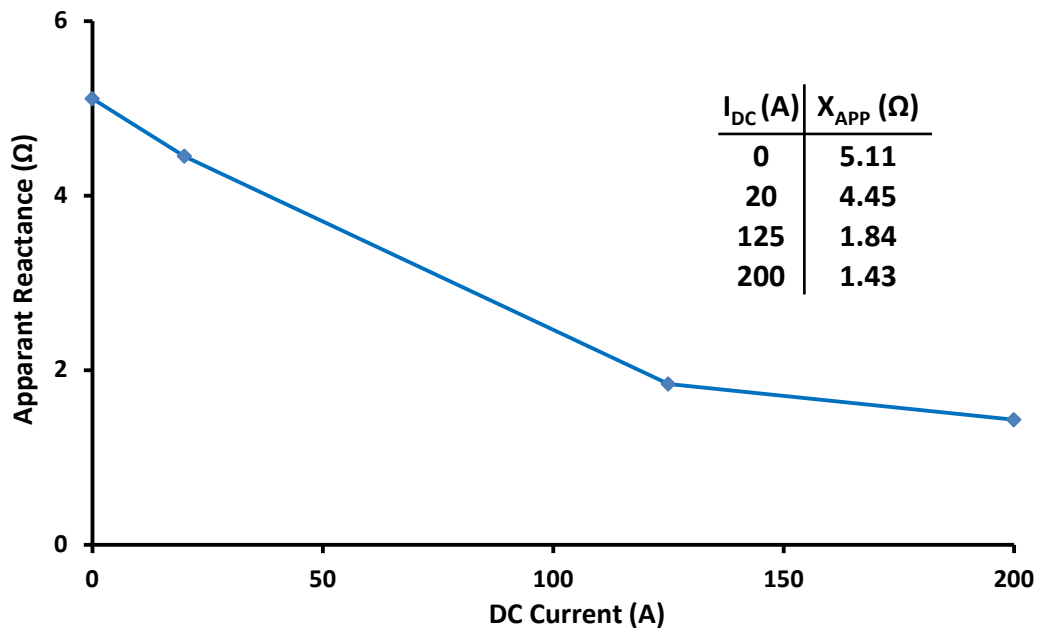


Figure 56 Apparent Reactance Curve for the Full-Scale Model

Summary of Results

The FEA-derived modeling approach appears to be an adequate method of representing the SRPFC in transient circuit simulations. The FEA results were implemented in the transient simulation tool, PSCAD™, via a two-dimensional data matrix that is interpolated by the transient simulator using bilinear interpolation. Simulations of the single-phase prototype produced results with strong correlation to a co-simulation method where the FEM was solved during every time step. Results were obtained and compared for nominal and fault conditions. In addition, the simulation of nominal conditions had good correlation to the test results obtained at ORNL. The data matrix approach produced significant speed-up in terms of simulation time over co-simulation with FEA; the former was solved in seconds while the latter required days. Once an adequate model of the single-phase prototype was achieved, it was scaled to transmission level requirements and implemented in PSCAD™. Expansion of the model to three-phases required applying a single-phase model to each phase of a transmission line.

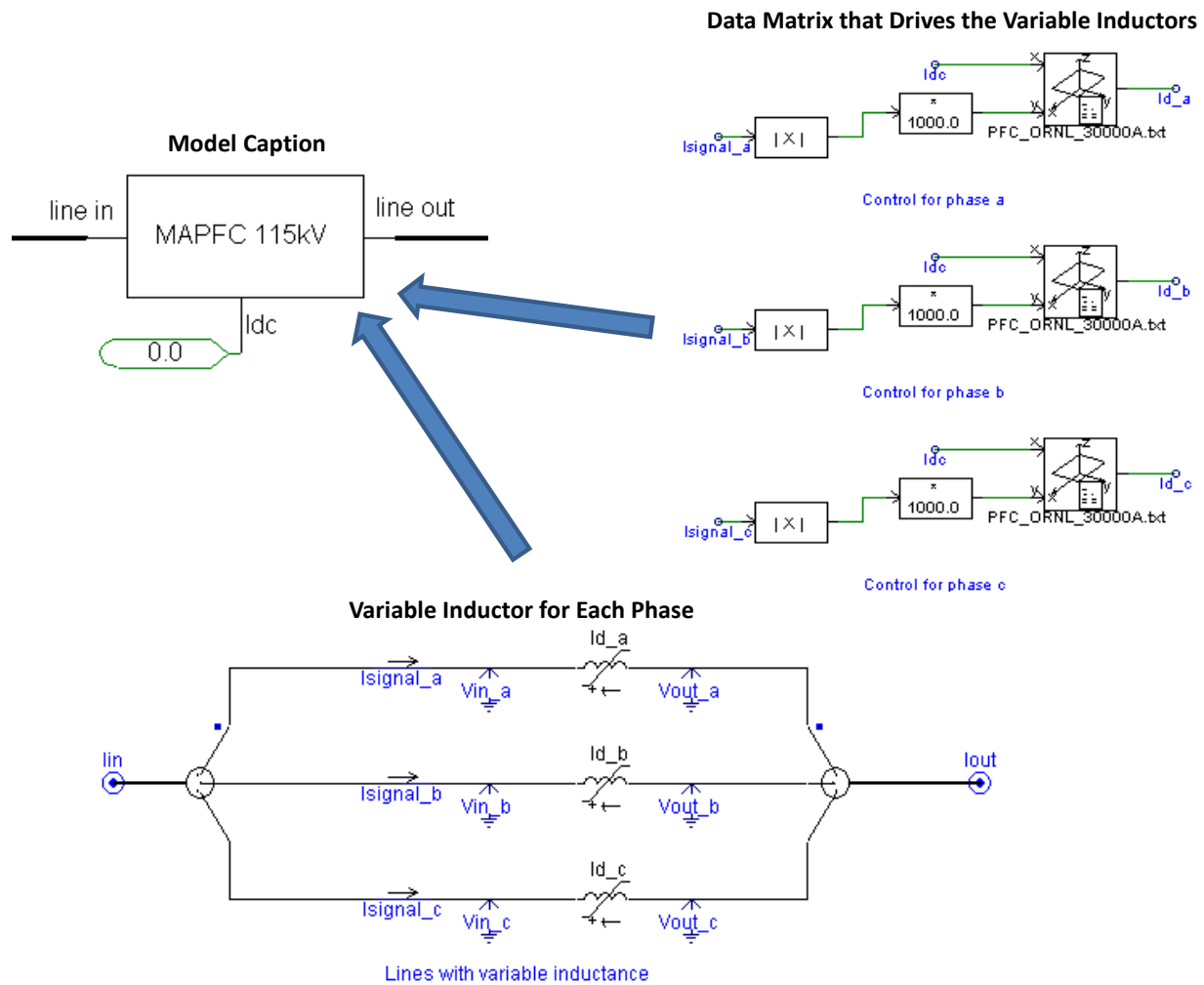


Figure 57 Screen Captions of Full-Scale Model Applied in PSCAD™

CHAPTER IV: GYRATOR-CAPACITOR MODEL

A conventional approach for modeling magnetic circuits is the reluctance-resistance analogy where lumped equivalent circuits are formed using resistors that represent flux paths and voltage sources that represent magnetomotive force [37]. The method is effective in modeling magnetic behavior, but its use is typically isolated to analysis of magnetic circuits with known winding currents. The coupling of magnetic circuits to electric circuits for tandem simulation requires that the system solve for the magnetic flux quantities and electric currents simultaneously. There must also be an energy transducing method that accommodates the transfer of energy between the two domains. Previous work suggests that it is possible to couple a reluctance-resistance network with an electric circuit using the principle of electromagnetic induction [49], but literature review findings reveal minimal use of this method. The gyrator-capacitor approach, first proposed by Rudolph Buntentbach in 1969 [38] [50], solves this coupling issue by introducing capacitors to achieve consistent flow variables and application of the gyrator as the interface between the two domains.

This chapter describes the gyrator-capacitor approach and applies the method in SPICE (Simulation Program with Integrated Circuit Emphasis) [51] to the SRPFC. The model is coupled to a DC control circuit and an AC power system that are solved in tandem with the magnetic circuit. Modifications to the resulting model then applied then to account for manufacturing imperfections and unintended flux paths that make the model operate more realistically according to observations from hardware testing. Simulations of the model are compared to the FEA-derived and FEA co-simulation models that are described in Chapter III. These simulations are performed under nominal and fault conditions. Results of the gyrator-capacitor model are also compared with existing hardware test data.

Theory of Approach

The reluctance-resistive approach models the magnetic behavior of a device or system by using a direct analogy to electric circuits composed of voltage sources and resistors. The equivalent circuits in Figure 58 illustrate the analogy between the magnetic and electric domains, where

resistors are used to represent the reluctance of magnetic pathways as described in (4.1) and (4.2).

$$R = \frac{l}{\sigma A} \quad (4.1)$$

$$\mathcal{R} = \frac{l}{\mu A} \quad (4.2)$$

Where: R = Electric resistance $\{\Omega\}$
 $\mathcal{R}$ = Magnetic reluctance $\{H^{-1}\}$
 l = Mean length $\{m\}$
 A = Cross-sectional area $\{m^2\}$
 σ = Electrical conductivity $\{S/m\}$
 μ = Permeability $\{H/m\}$

In electric circuits, voltage is the force variable that interacts with electric resistance to develop current, the flow variable, in accordance with Ohm's law as described in (4.3). Likewise, magnetomotive force, or *mmf*, is the effort, or force, variable that interacts with the reluctance to develop magnetic flux. This relationship is described in (4.4). The magnetic flux is the flow variable in the reluctance-resistance model that directly correlates to electric current in an electric circuit.

$$I = \frac{V}{R} \quad (4.3)$$

$$\phi = \frac{\mathcal{F}}{\mathcal{R}} \quad (4.4)$$

Where: I = Current $\{A\}$
 V = Voltage $\{V\}$

R = Resistance $\{\Omega\}$

ϕ = Flux $\{\text{Wb}\}$

$\mathcal{R}$ = Reluctance $\{\text{H}^{-1}\}$

$\mathcal{F}$ = Magnetomotive force $\{\text{A}\cdot\text{turns}\}$

Synthesis of the equivalent reluctance circuit from the electromagnetic device is straightforward. Figure 58a shows a simple electromagnetic device with a winding, core material of known type and geometry, and current flowing through the winding. The winding has N number of turns. The magnetic circuit, shown in Figure 58b, is constructed by application of an mmf source that is equal to $N\cdot i$. The reluctance is determined by applying (4.2) with knowledge of the permeability, mean magnetic length, and cross-sectional area of the core. This reluctance-resistance model is applied as an electrical circuit for simulation as shown in Figure 58c.

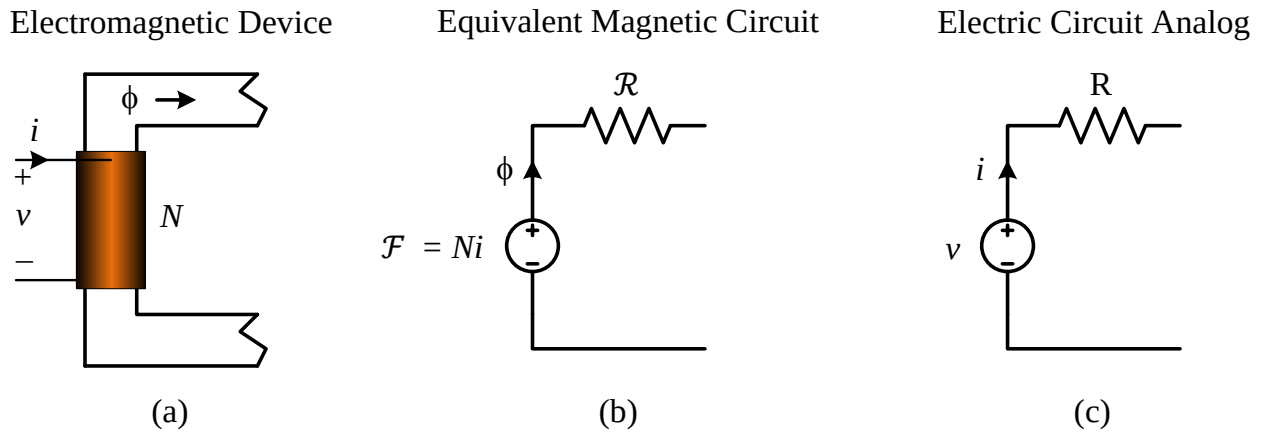


Figure 58 Reluctance-Resistance Approach to Magnetic Circuit Modeling

Table 4 describes the force and flow variables of the electric and magnetic domains as summarized in previous work [52]. A significant difference between the two domains is evident when taking the product of the force and flow variables. In the electric domain, the product of voltage and current is power, but in the magnetic domain, the product of mmf and flux is energy. Since power and energy are not compatible quantities, the two domains cannot couple

directly. The two domains have incompatible products because their flow variables are not consistent. For instance, the flow variable for the electric domain is current, which by definition is the rate of change of electric charge with respect to time expressed in coulombs per second. The flow variable for the magnetic domain is volt·seconds, which is not a time rate of change. Therefore, the flow variables represent different quantities. In short, the reluctance-resistance approach does not permit energy transfer directly between the two domains due to the inconsistent flow variables. The inability to couple the two domains requires that the electric current through the magnetic windings be a known quantity in order to solve the magnetic circuit, or the electric current can be an unknown if the core flux components are a known quantity. The reluctance-resistance model cannot provide a solution if the current through the windings and the flux components are unknown; this would be the case if a voltage source is connected to the winding and the current and flux components are dependent on the impedance of the electromagnet system.

In addition to inconsistent flow variables, authors have stated that it is confusing to represent the inductive nature of magnetic circuits that store energy with components like resistors that dissipate energy [39] [53]. Replacing resistances with inductors fixes both of these issues [40] [36] [53], but results in circuit topologies that bear little resemblance to the components that they model [40].

These issues are resolved by replacing resistors (reluctance) with capacitors (permeance) in the magnetic circuit. The use of capacitors make the flow variables consistent for linking the domains, represent the reactive nature of magnetic devices as oppose dissipating energy, and maintains topological resemblance of the actual device or system.

Table 4 Force and Flow Variables in Electric/Magnetic Domains

Domain	Force Variable	Flow Variable	Force · Flow
Electric	Voltage {V}	Current {A} $\equiv$ {C/s}	Power {V·A} $\equiv$ {W}
Magnetic	<i>mmf</i> {A·turns}	Flux {Wb} $\equiv$ {V·s}	Energy {V·A·s} $\equiv$ {J}

Permeance-Capacitance

Achieving a magnetic flow variable that is consistent with the electric domain starts by expressing the magnetic flux in terms of reluctance and mmf by substituting the definition of reluctance in (4.2) into (4.4).

$$\phi = \frac{1}{\mathcal{R}} \cdot \mathcal{F} = \frac{\mu A}{l} \cdot \mathcal{F} \quad (4.5)$$

As previously stated, electric current is the flow variable in the electric domain and is a time rate of change. As magnetic flux is not a time rate of change, it is possible to achieve a flow variable consistent with the electric domain by taking the time derivative of (4.5). The resulting flow variable is a time rate of change of flux as expressed in (4.6). The new constant formed by (4.6) is permeance, which is equal to the inverse of reluctance as described in (4.7).

$$\frac{\partial \phi}{\partial t} = \frac{\mu A}{l} \cdot \frac{\partial \mathcal{F}}{\partial t} \quad (4.6)$$

$$P = \frac{\mu A}{l} = \frac{1}{\mathcal{R}} \quad (4.7)$$

Where: $\frac{\partial \phi}{\partial t}$ = Time rate of change of flux {V·s/s} or {V}

$\frac{\partial \mathcal{F}}{\partial t}$ = Time rate of change of mmf {A·turns/s}

P = Permeance {H}

The relationships in (4.6) and (4.7) are significant because together they resemble the equation of current through a capacitor. The flux rate in (4.8) is obtained by expressing parameters of (4.6) in terms of P and $\dot{\phi}$. For comparison, (4.9) describes an electric current through a capacitor. There is distinct duality between the two they describe the product of a constant parameter and a time rate of change.

$$\dot{\phi} = P \cdot \frac{\partial \mathcal{F}}{\partial t} \quad (4.8)$$

$$i = C \cdot \frac{\partial v}{\partial t} \quad (4.9)$$

Where: $\dot{\phi}$ = Time rate of change of flux {V·s/s} or {V}

i = Electric current {A}

C = Capacitance {farads}

Another point of significance in (4.8) and (4.9) is that the force variables for both domains remain consistent and have not changed from the reluctance-resistance approach. Therefore, the permeance-capacitance approach utilizes magnetomotive force, $\mathcal{F}$, as an analog to electric voltage. The permeance-capacitance approach is illustrated in Figure 59.

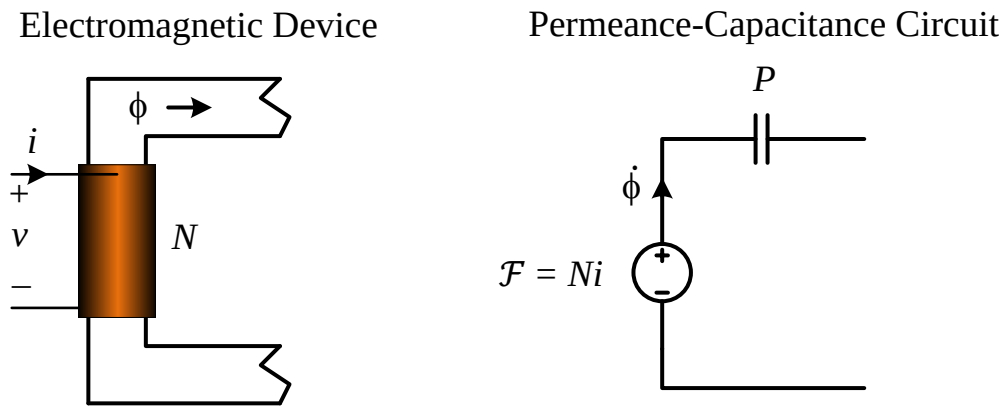


Figure 59 Permeance-Capacitance Approach to Magnetic Circuit Modeling

The force and flow variables of the three modeling approaches are compared in Table 5. The comparison reveals that the permeance-capacitance approach gives a force-flow product that is more consistent with the electric domain, but proportionally unequal to the electric circuit by the number of winding turns. This issue is resolved by considering the number of turns upon

transfer of energy between domains. With matching force-flow products, it is possible to transfer energy between the domains. However, the transfer of energy will require an interface that performs the transfer and accounts for the proportionality issue.

Table 5 Comparison of Force and Flow Variables

Domain	Force Variable	Flow Variable	Force · Flow
Electric	Voltage {V}	Current {A} $\equiv$ {C/s}	Power {V·A} $\equiv$ {W}
Magnetic (Reluctance-Resistance)	mmf {A·turns}	Flux {Wb} $\equiv$ {V·s}	Energy {V·A·s} $\equiv$ {J}
Magnetic (Permeance-Capacitance)	mmf {A·turns}	Flux rate {Wb/s} $\equiv$ {V}	Power {V·A·turns} $\equiv$ {W}

The Tellegen Gyrator

Consistency between flow variables makes it possible to transfer energy between the electric and magnetic domains, but an interface is required to accommodate the exchange. The gyrator is the link that completes the gyrator-capacitor model. Bernard Tellegen of the Philips Physics Laboratory first introduced the gyrator in 1948 [54]. The gyrator is an ideal, passive, two-port electrical circuit element that supports exchange of force and flow between two systems [55]. The circuit schematic of the gyrator is shown in Figure 60.

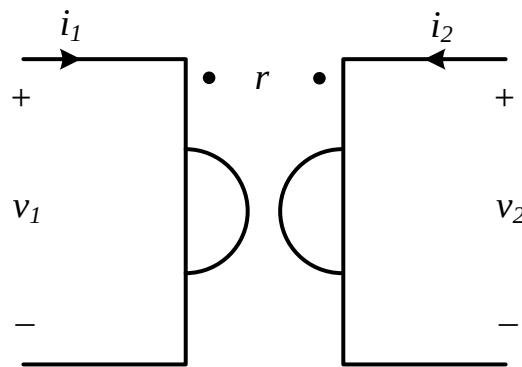


Figure 60 Circuit Schematic of a Tellegen Gyrator

The gyrator operates similarly to an ideal transformer in that the energy of both sides is proportional. However, a transformer relates the primary voltage to that of the secondary while the gyrator relates the primary voltage to the secondary current. Thus, an ideal transformer is formed by connecting two gyrators back-to-back. A constant of proportionality equalizes the energy on both sides. The relationship between the inputs and outputs of the gyrator are expressed in (4.10). The gyrator is lossless, in accordance with (4.11), so no energy dissipates during in the transformation process. Figure 61 illustrates application of the gyrator to link the electric and magnetic domains.

$$\begin{bmatrix} v_1 \\ v_2 \end{bmatrix} = \begin{bmatrix} 0 & -r \\ r & 0 \end{bmatrix} \begin{bmatrix} i_1 \\ i_2 \end{bmatrix} \quad (4.10)$$

$$v_1 i_1 + v_2 i_2 = 0 \quad (4.11)$$

Where:

- v_1 = Input voltage
- v_2 = Output voltage
- i_1 = Input current
- i_2 = Output current
- r = Constant of proportionality

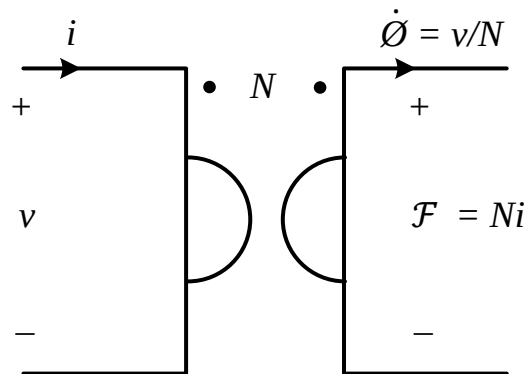


Figure 61 Gyrator-Capacitor Model

Application of the gyrator and a permeance-capacitance element achieves a gyrator-capacitor model that couples electric and magnetic circuits [52] [50] [38] [56]. Synthesis of the electromagnetic device in Figure 59 results in a gyrator-capacitor model as shown in Figure 62. The model operates in accordance with the system of equations described in (4.12). The constant of proportionality, N , is equal to the number of winding turns and cancels with the turns in the magnetic flow variable to make the units of both flow variables in $V \cdot A$.

$$\begin{bmatrix} v \\ \mathcal{F} \end{bmatrix} = \begin{bmatrix} 0 & N \\ N & 0 \end{bmatrix} \begin{bmatrix} i \\ \dot{\phi} \end{bmatrix} \quad (4.12)$$

Where: v = Terminal voltage {V}
 i = Terminal current {A}
 $\mathcal{F}$ = mmf {A·turns}
 $\dot{\phi}$ = Flux rate {Wb/s} or {V·A·turns}
 N = Number of turns

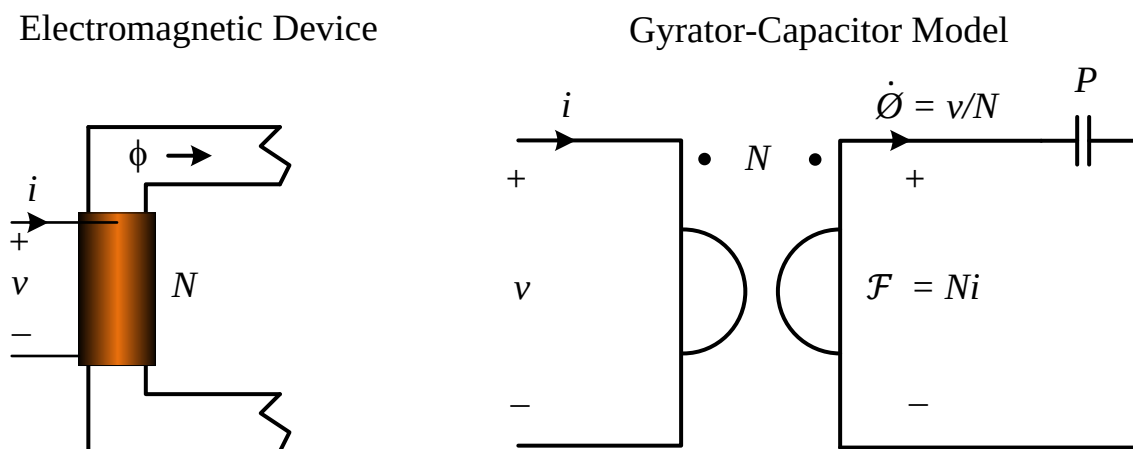


Figure 62 Synthesis of the Gyrator-Capacitor Model

Nonlinear Permeance

It is possible to model cores of nonlinear permeability with application of variable capacitors, or variable permeances, as shown in Figure 63. The choice of method in application of a variable permeance is dependent on the ability of a specific method to simulate the B-H characteristics accurately within the capabilities of the modeling/simulation tool. The method described here, and used to model the SRPFC, interpolates a table of B-H curve data points. Mathematical approaches are also available [56] [40] [57] [58], but do not adequately represent the B-H characteristics for some materials.

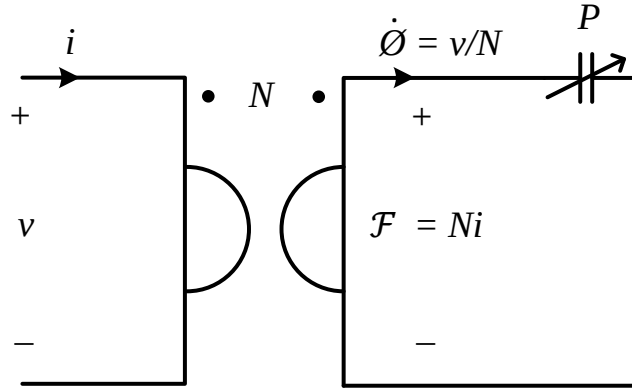


Figure 63 Gyrator-Capacitor Model with Variable Permeance

Variable permeance was modeled in SPICE using a method that calculates an equivalent permeance using either an equation or a data table [59] [41]. The method uses an ideal transformer model [60] to obtain an equivalent permeance, P_{eq} , by reflecting a unity permeance, P , on the secondary side over to the primary using a scaling factor, k_n , as illustrated in Figure 64. The ideal transformer consists of a dependent voltage source on the primary side and a dependent current source on the secondary side. The use of unity permeance on the secondary side makes the scaling factor inversely proportional to the equivalent permeance. The equivalent permeance is proportional to the differential permeability calculated from the B-H curve during each time step. This method was successful in previous work to implement a

mathematical curve fit to a B-H characteristic [41], but required modification in order to implement a table of data values [61].

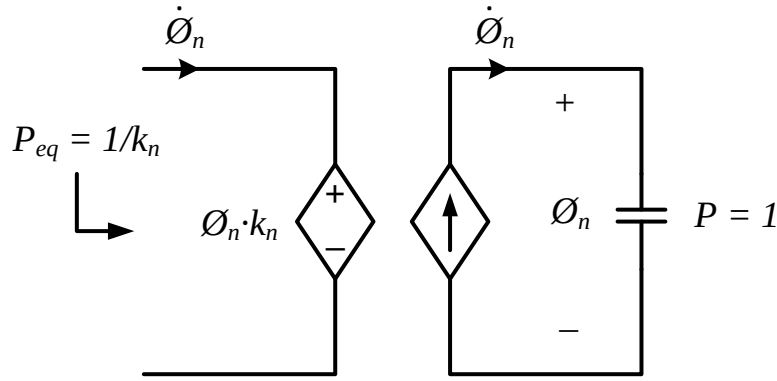
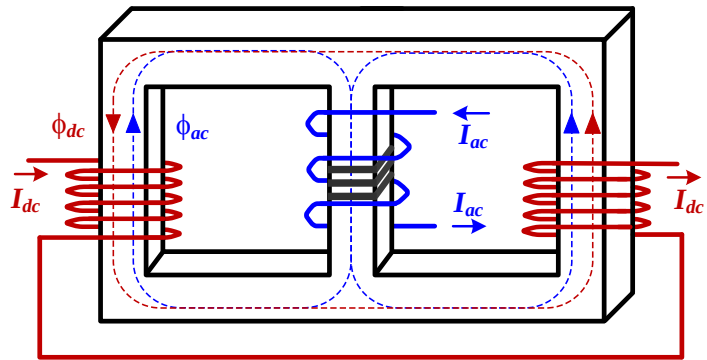


Figure 64 Variable Permeance for Application in SPICE

Application to SRPFC

The gyrator-capacitor approach was applied to the SRPFC using a commercial version of SPICE [62]. The magnetic circuit is represented by three variable permeances to model each leg of the core, a static permeance to model the air gap, and three gyrators to model the two DC windings and the single AC winding. The topology of the gyrator-capacitor model bears strong resemblance to the design configuration of the SRPFC prototype as shown in Figure 65. The core of the center leg is lumped into a single element with mean length equal to the length of the center leg minus the air gaps. The outer legs have mean lengths of approximately half the circumference of the rectangle. The geometry, material characteristics, and winding configurations are consistent with the specifications described in CHAPTER I. The magnetic circuit is coupled to a single-phase AC circuit consisting of a voltage source and a resistive load with parameters identical to the single-phase system described in CHAPTER III. The magnetic circuit is also coupled to a DC electric circuit that represents the DC winding and is driven by an ideal current source to control the DC bias.



Three-leg prototype



G-C Model

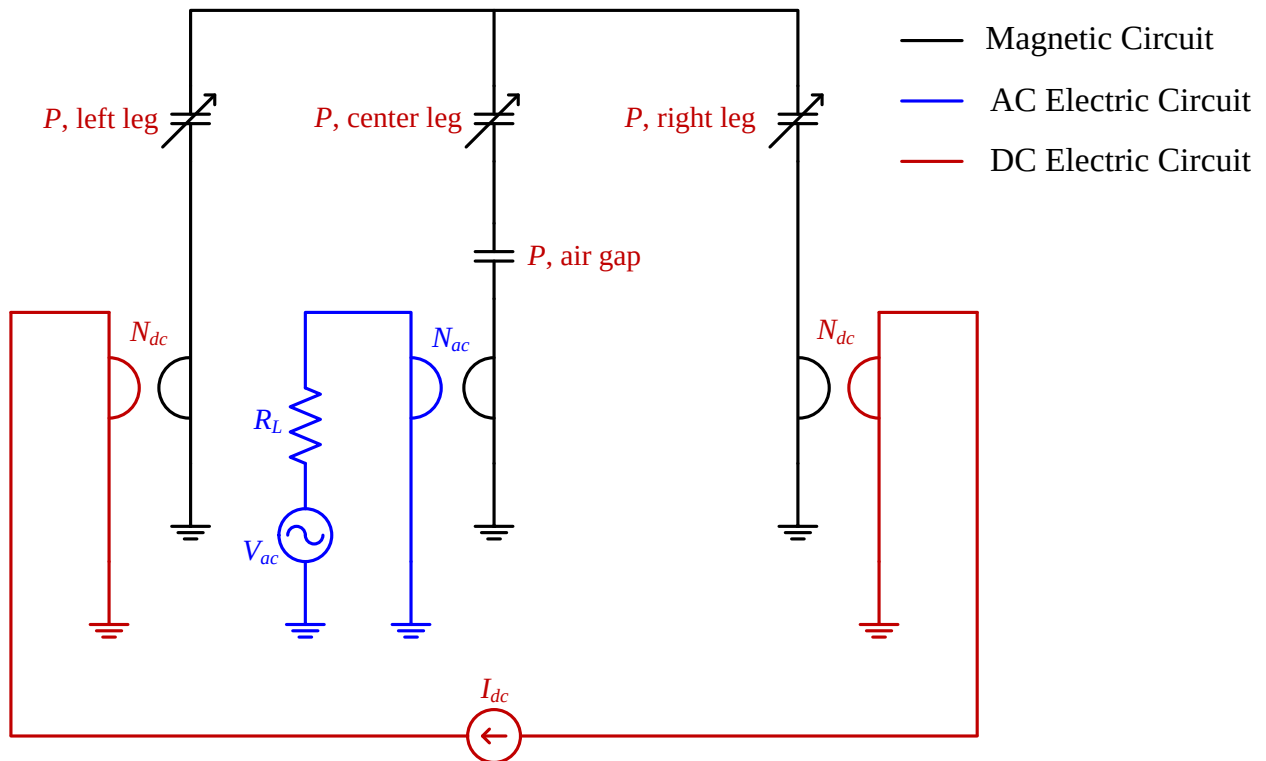


Figure 65 Gyrator-Capacitor Model of SRPFC

Application of the magnetic model in SPICE is shown in Figure 66. SPICE does not contain a gyrator library component, so each gyrator is constructed using dependent voltage sources [40] [41]. The nonlinear permeances are contained within hierarchical blocks to improve the appearance of the model. Each nonlinear permeance block utilizes a SPICE compatible look up table of the AKH1 Steel B-H curve generated using MATLAB®. The AC and DC circuit models in SPICE are shown in Figure 67. The AC circuit is implemented with a load resistor that sets the current under nominal conditions and a surge resistor that determines the fault current. The load resistor has a by-pass switch to simulate line-ground faults. The DC circuit is driven by an ideal current source. The gyrator-capacitor can solve for the back emf of the SRPFC, which is the voltage induced by the AC flux in the center leg onto the DC winding. The back emf is equivalent to the difference between the flux rate of change of the outer legs as follows:

$$E_{EMF} = (\dot{\phi}_{right} - \dot{\phi}_{left}) \cdot N_{DC} \quad (4.13)$$

Where: E_{EMF} = Back emf {V}
 $\dot{\phi}_{right}$ = Flux rate in right leg {V}
 $\dot{\phi}_{left}$ = Flux rate in left leg {V}
 N_{DC} = Number of DC turns

The back emf does not have a direct impact on the AC system, but it is significant to the design of the DC power supply.

The variable permeance blocks are constructed according to Figure 68. The flux calculation circuit applies the input flux rate to a unity capacitor, or permeance, circuit resulting in a voltage drop across the capacitor that is equal to the flux as follows:

$$mmf = \frac{1}{P} \int \dot{\phi} \cdot dt = \frac{\phi}{P} \quad (4.14)$$

The dependent voltage source in Figure 68a converts this flux to magnetic field density, B, by dividing it by the cross sectional area. The resulting B is passed to the interpolation element in

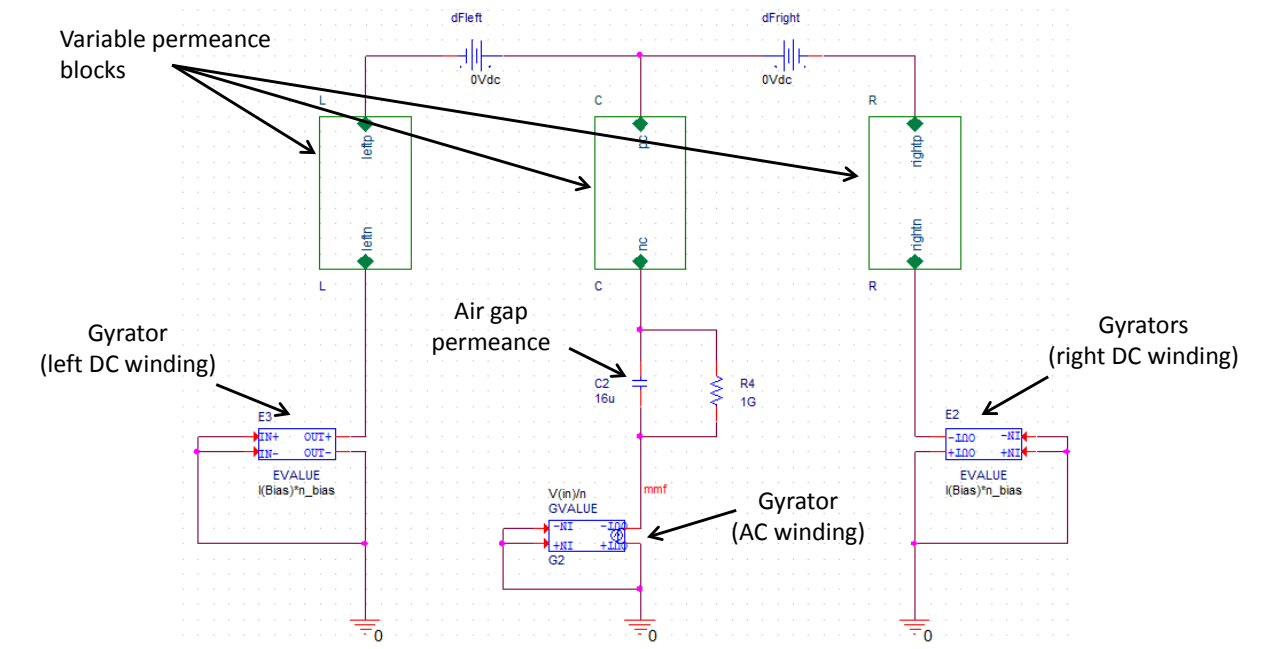


Figure 66 Gyrator-Capacitor Model of SRPFC in SPICE

There are components in the SPICE models, notably resistors and DC voltage sources, that improve convergence or take measurements. Prior to performing transient simulation, the SPICE solver must calculate a DC solution. Since capacitors are open circuits to DC, resistors are placed in parallel with each capacitor element to provide a conducting path. The value of these resistors is large, $10^9 \Omega$, so they do not impact the accuracy of the simulation. In contrast, small series resistances, $10^{-3} \Omega$, are connected between series connected voltage sources as SPICE will produce errors they are connected directly. The 0V DC voltage sources are a SPICE technique for obtaining current measurements that bias the dependent sources. SPICE provides current values only when there is potential. In this case, the DC sources are set to zero to avoid affecting the accuracy of the solution. These extraneous circuit elements are SPICE-specific tricks to enhance the probability of convergence and to observe solutions; they do not represent characteristics of the SRPFC.

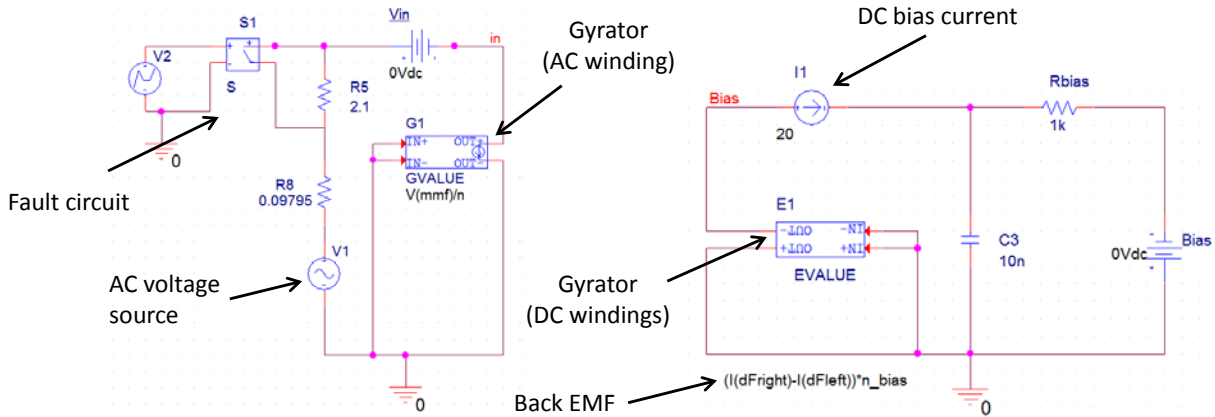


Figure 67 AC and DC Electric Circuits in SPICE

Manufacturing Effects

Modifications to the basic model in Figure 65 are necessary to account for issues in precision with the cutting and stacking of core laminates. In the 480V prototype, the 2mm air gap in the middle leg approached the threshold of the manufacturer's cutting instrument in terms of precision. Thus, there are errors in the air gap size that could affect SRPFC operation. Additionally, the three-legged core was constructed by stacking sheets of rectangular lamination. A parasitic air gap is formed at locations where the ends of the rectangular laminates join. The effect of these parasitic gaps along the entire portion of a leg results in a net, or equivalent, air gap. These equivalent air gaps are represented by static capacitors on the outer legs, P_{tol} , as shown in Figure 69. Precision of the middle leg air gap is adjusted by the static capacitor already in place. The value of these permeances was determined by trial and error to achieve an acceptable match to data from hardware testing and the FEA-derived model.

Outer Air Gap Flux

Hardware testing and FEA simulation at ORNL revealed that the reactance of the SRPFC in saturation mode was a higher level than the B-H characteristics of AKH1 steel indicate it should be. It was determined by observing test results and FEA simulation that a portion of the flux in each leg leaps across the outer air gaps as the core approaches saturation, as illustrated by

Figure 70. The flux takes the path of highest permeance. As the core reaches saturation, the permeance of the core is low compared to air and flux jumps the air gap. Contrastingly, core permeance is highest when the core is operating in the linear region and the flux remains within the core. The outer air gaps are accounted for by the addition of a permeance path between each outer leg and the middle leg as shown in Figure 71. Since the equivalent length and cross sectional area of these paths is unknown, their values are found through trial and error until the saturation point adequately matches the hardware test data and FEA-derived model. The final SPICE implementation of the magnetic circuit in Figure 72 implements the equivalent air gaps due to manufacturing precision and the outer air gaps.

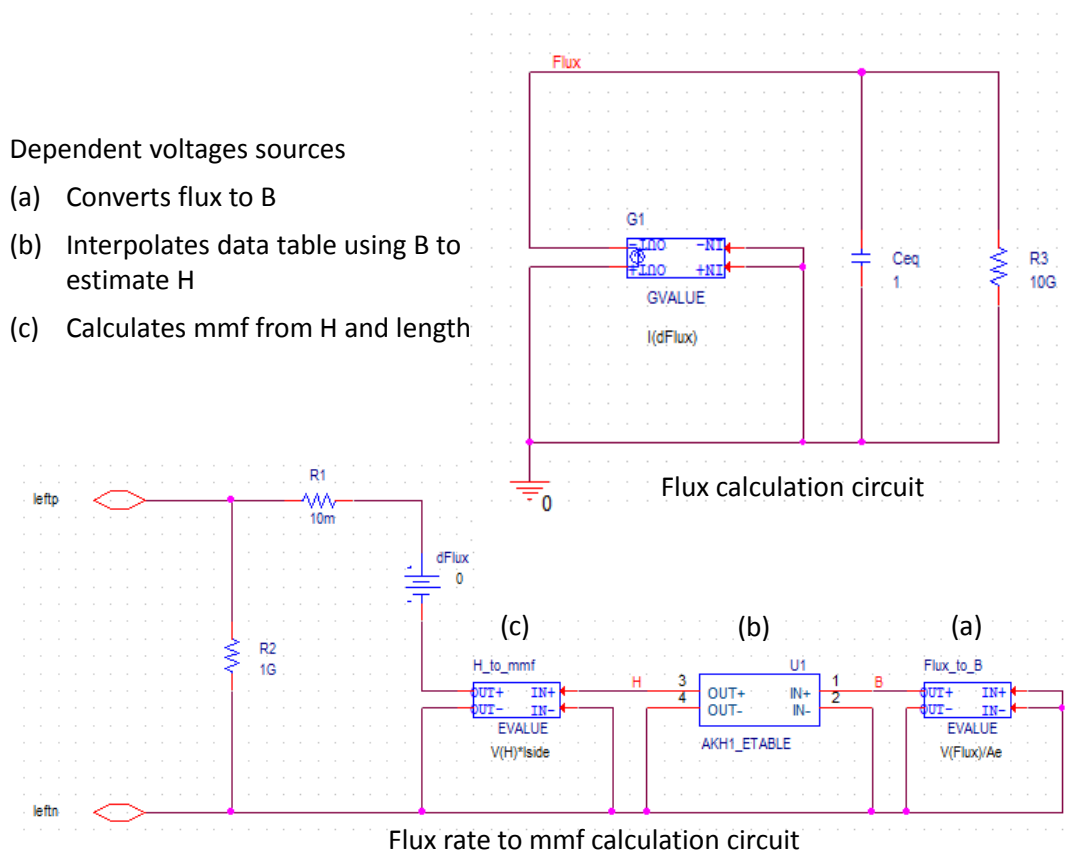


Figure 68 Variable Permeance Block in SPICE

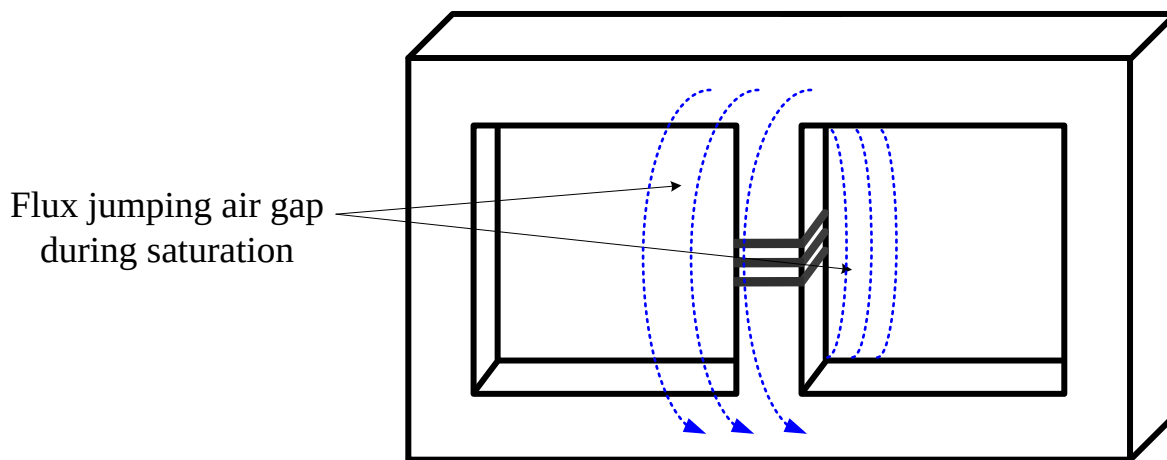
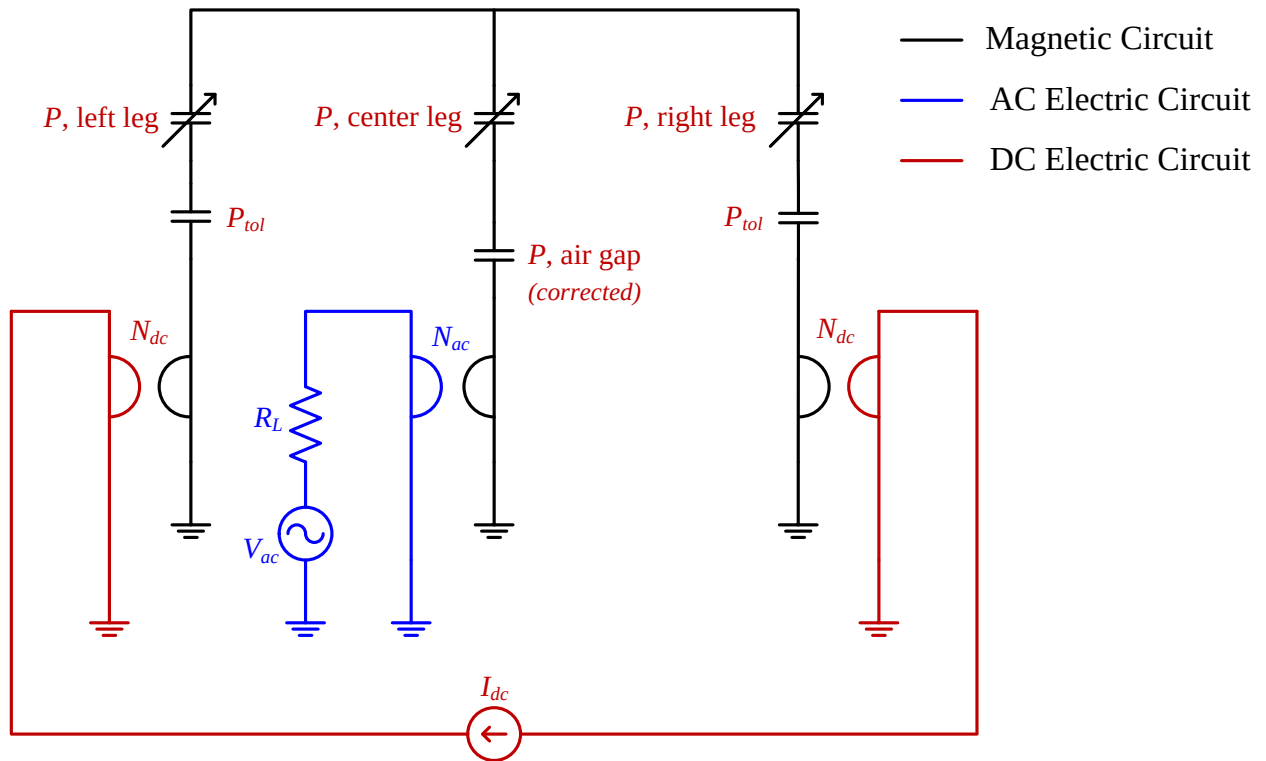


Figure 70 Illustration of Flux Jumping the Outer Air Gaps

Summary of Results

The gyrator-capacitor model was simulated in SPICE for the initial case, with modifications to account for manufacturing effects, and with modifications to include outer air gap flux paths. Each modification of the model has identification according to Table 6. The simulations were performed using AC circuit parameters like those in Table 2 from CHAPTER III. The SPICE solver utilizes variable time-steps that auto adjusts to improve convergence. The maximum time step size was set to $100\mu\text{s}$ in an effort to maintain consistency with the other modeling and simulation methods. Results of these simulations are compared to the apparent reactance curves obtained from hardware testing and the FEA-derived model. Permeance values of the equivalent air gaps were determined by performing simulations and altering the parameters until an adequate match of the apparent reactance curves was achieved. The results are compared to waveforms obtained from Simplorer (FEA-transient co-simulation) and PSCAD™ (FEA-derived model).

Table 6 Versions of the Gyrator-Capacitor Model

Name	Modification
Mod-0	Initial case with no modifications
Mod-1	Equivalent air gap permeances for manufacturing effects
Mod-2	Manufacturing effects and outer air gap permeances

Apparent Reactance

The apparent reactance curves in Figure 73 were obtained with an AC current of 125A_{RMS} from hardware test results, the FEA-derived model, and the gyrator-capacitor model. The curve from the initial gyrator-capacitor model, Mod-0, has strong correlation to the hardware tests and FEA-derived model in terms of behavior, but has a low value at 0A_{DC} and saturates at a level much lower than the other curves. The addition of equivalent capacitances that account for manufacturing precision are effective at increasing the 0A_{DC} point, shown by the Mod-1 curve, but have no effect on the saturation level. The addition of the outer air gap permeances has a

significant affect the saturation level and aligns the knee points of the curves with respect to the DC current. All of the modifications combined produce an apparent inductance curve, Mod-2, with adequate correlation to the hardware tests and the FEA-derived model in terms of scale and behavior.

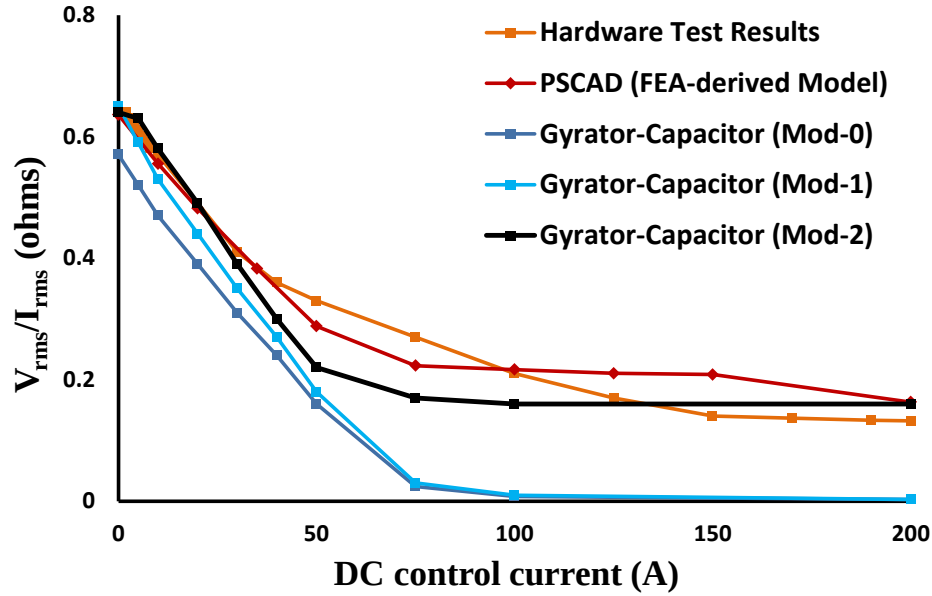


Figure 73 Apparent Reactance Curves

Nominal Conditions

The simulation results in Figure 74 through Figure 78 are the result of simulations of the Mod-1 gyrator-capacitor model. The waveforms in Figure 74 are the result of the gyrator-capacitor model with AC current of $125A_{RMS}$ and a DC bias current of $0A_{DC}$. The terminal voltage and current are purely sinusoidal and the outer leg fluxes are equal with no DC offset. The middle leg flux is equal to the sum of the flux in the outer legs. When the DC current is $20A_{DC}$, as in Figure 75, the outer flux components have a DC offset that is equal and opposite. The flux, current, and voltage waveforms appear sinusoidal as the core is just beginning its transition to the knee point of the B-H curve. This operation causes a minute reduction of terminal voltage magnitude and slight deformation of the current waveform near the zero current crossings. At

50A_{DC}, Figure 76, the core is operating within the knee point of the curve as portions of the flux in the outer leg transition from the linear to the saturation region within portions of the half cycle. The flux in the middle leg is no longer a pure sinusoid and the AC current shows significant deformation at the zero crossings. The most obvious change is the reduction of magnitude and harmonics in the terminal voltage. These harmonics are synonymous with the derivative of the current waveform as the deformations at the voltage peaks are consistent with the current zero crossings. This relationship between the voltage and current shows the inductive nature of the SRPFC and correlates with results from Simplorer and PSCAD™ simulations. Simulation of the model at 100A_{DC}, shown in Figure 77, is a state where the SRPFC initiates transition from the knee point to the saturation region. The flux in the outer legs transition to the linear region only at the peaks, resulting in a middle leg flux that is reduced from the previous simulations, but has significant harmonic content. The harmonics are observable in the terminal voltage, which has been reduced to a peak value of less than 50V. The deformation of the current waveform is indicative of the changing line impedance. The waveforms for the 200A_{DC} simulation in Figure 78 indicate operation in deep saturation region as the outer leg fluxes are almost entirely DC. The resulting terminal voltage is nearly zero at this point and the current is purely sinusoidal.

The terminal voltage waveforms from Mod-1 simulations are compared to results from Simplorer and PSCAD™ in Figure 79. The current waveforms, not shown, are identical in each operating state and purely sinusoidal. At 0A_{DC}, there is strong correlation between the three modeling methods, but the Mod-1 voltage waveform does not contain the harmonic effects from numerical instabilities that occur in the models that depend on FEA. The 20A_{DC} Mod-1 waveform is nearly identical to the 0A_{DC} curve. As previously stated, there is minute reduction in the voltage magnitude as the SRPFC begins its transition from the linear region to the knee point of the B-H curve. The Mod-1 curve does not contain the harmonics observed at the peaks of the Simplorer/PSCAD™ waveforms. The likely reason for this result is that the modifications in the Mod-1 model have moved the transition point to higher DC current values. The 125A_{DC} and 200A_{DC} operating points reveal further disparities between the Mod-1 model and the

Simplorer/PSCAD™ results. At these points, the Mod-1 model saturates at a level much lower than the Simplorer/PSCAD™ simulations as observed by the lack of terminal voltage.

Simulation results of the Mod-2 model in Figure 80 show significant improvement over the Mod-1 results. At $0A_{DC}$, the results are the same as Mod-1. However, the $20A_{DC}$ results have now have the same magnitude and contain the harmonic content as observed in the Simplorer/PSCAD™ results. The $125A_{DC}$ and $200A_{DC}$ results correlate strongly with the Simplorer/PSCAD™ results as the outer air gaps dominate the saturation point.

Fault Conditions

The fault simulation results in Figure 81 through Figure 84 were produced by applying a line-ground fault for 5 cycles following 1 cycle of nominal operation. The results for both Mod-1 and Mod-2 models show strong agreement with the Simplorer/PSCAD™ results. This agreement is observable in the waveforms obtained from operation at $0A_{DC}$ in Figure 81 and Figure 83. At $200A_{DC}$, all methods strongly correlate with Mod-1 and Mod-2. The Mod-1 voltage waveform has notching at its peaks not present in the Simplorer/PSCAD™ results, but the Mod-2 result is nearly identical at the peaks.

Back EMF

The waveforms in Figure 85 and Figure 86 represent the back emf under nominal conditions over a range of DC bias currents. SRPFC operation in linear and saturation produces balanced AC flux within the center leg, so back emf is nearly zero. However, operation in the knee-point results in net AC flux in the center leg, causing back emf. The maximum peak back emf estimated by the gyrator capacitor model is approximately 180V, which occurs for DC bias currents slightly greater than zero. The FEA of the three-legged core model produced comparable results, as the peak of the largest waveforms is approximately 200V. There is a small difference in the level of magnitude for each dc bias, which was expected since the FEA model saturates at a slightly higher level than the gyrator-capacitor model.

For fault conditions, the back emf waveforms from the gyrator-capacitor model shown in Figure 87 reveal peak voltages approaching nearly 500V. As expected, the back emf is higher under

fault conditions because the outer leg fluxes become unbalanced. In addition, large fault current magnitudes result in high flux rates. The back emf waveforms from FEA, simulated using a $200\mu\text{s}$ timestep, are shown in Figure 88. The results are comparable to those from the gyrator-capacitor model in terms of behavior, but reveal numerical issues with the FEA at the waveform peaks. The gyrator-capacitor model is able to achieve solutions that are both stable and identical for maximum time steps ranging from $200\mu\text{s}$ down to $1\mu\text{s}$.

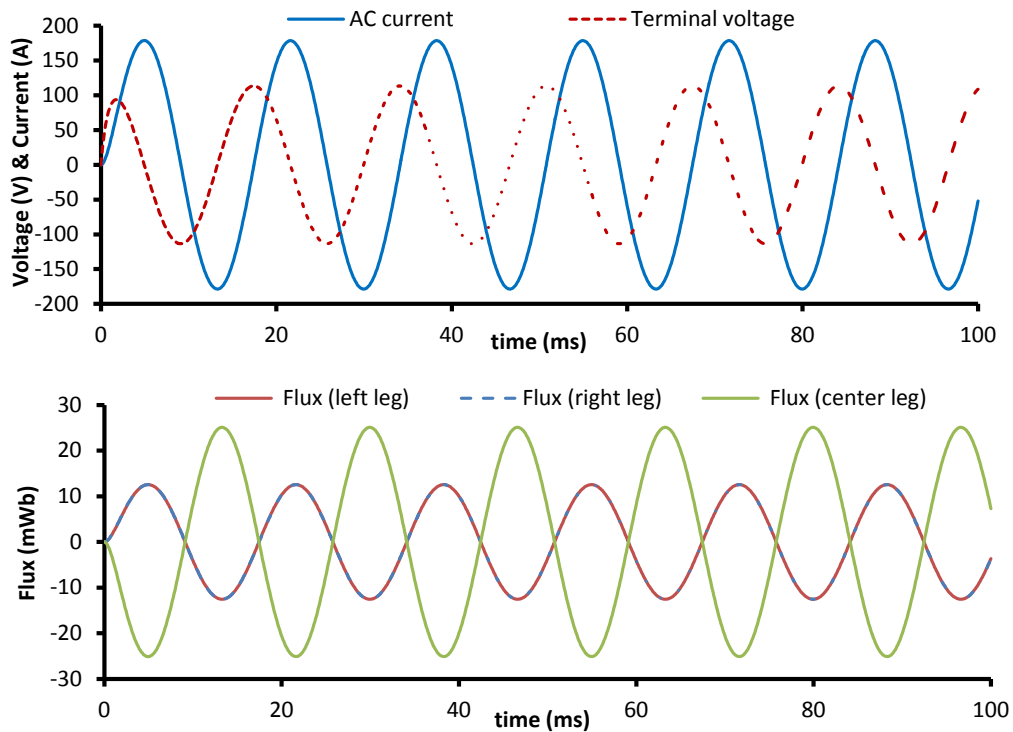


Figure 74 Mod-1 Gyrator-Capacitor Model: $I_{DC} = 0\text{A}$, $I_{AC} = 125\text{A}$

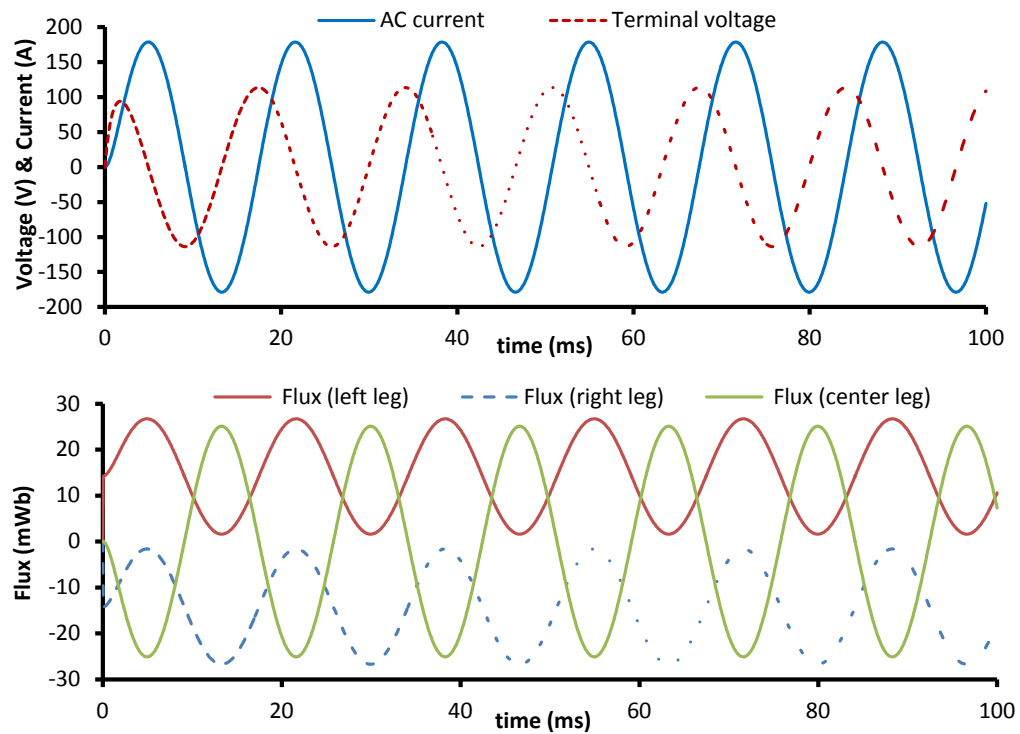


Figure 75 Mod-1 Gyrator-Capacitor Model: $I_{DC} = 20A$, $I_{AC} = 125A$

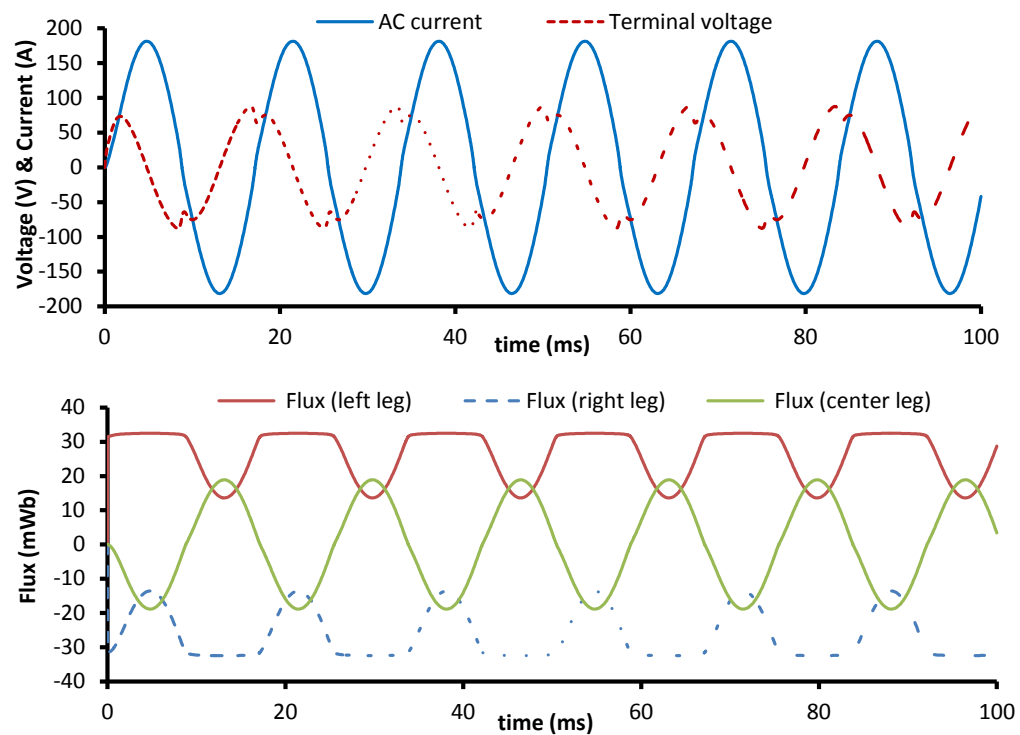


Figure 76 Mod-1 Gyrator-Capacitor Model: $I_{DC} = 50A$, $I_{AC} = 125A$

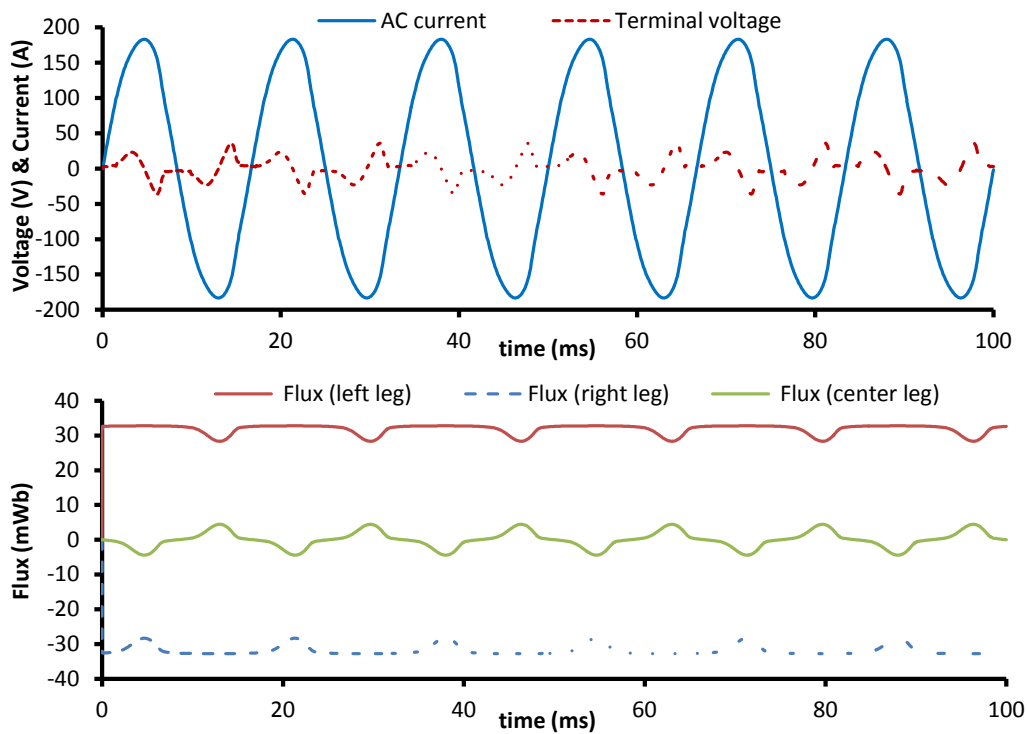


Figure 77 Mod-1 Gyrator-Capacitor Model: $I_{DC} = 100A$, $I_{AC} = 125A$

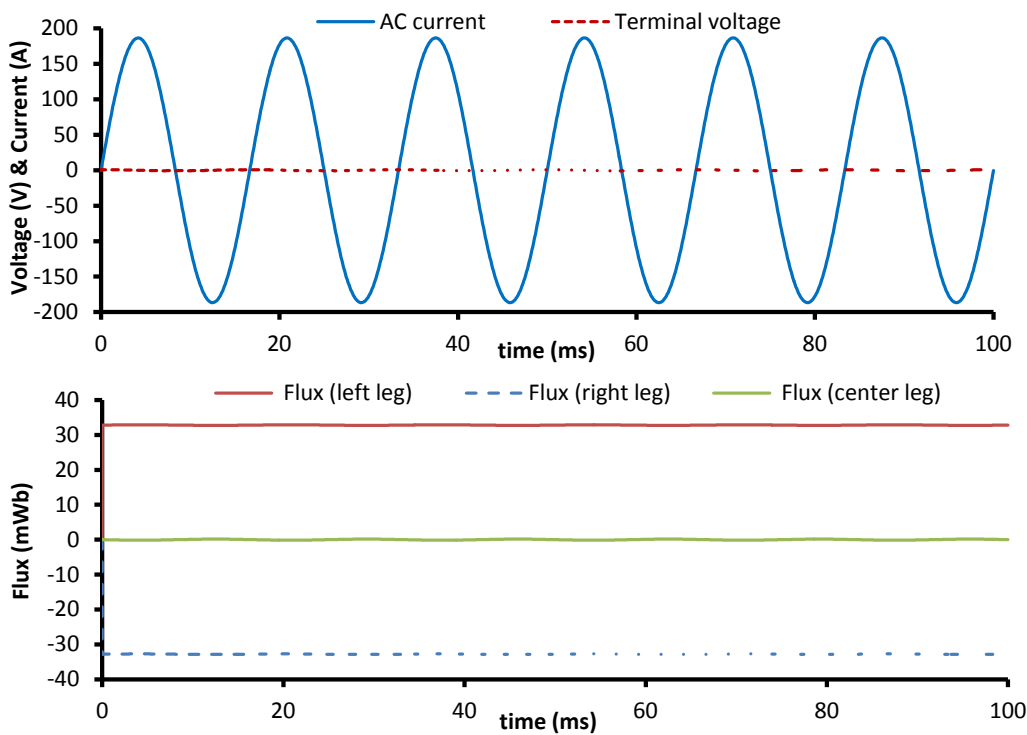


Figure 78 Mod-1 Gyrator-Capacitor Model: $I_{DC} = 200A$, $I_{AC} = 125A$

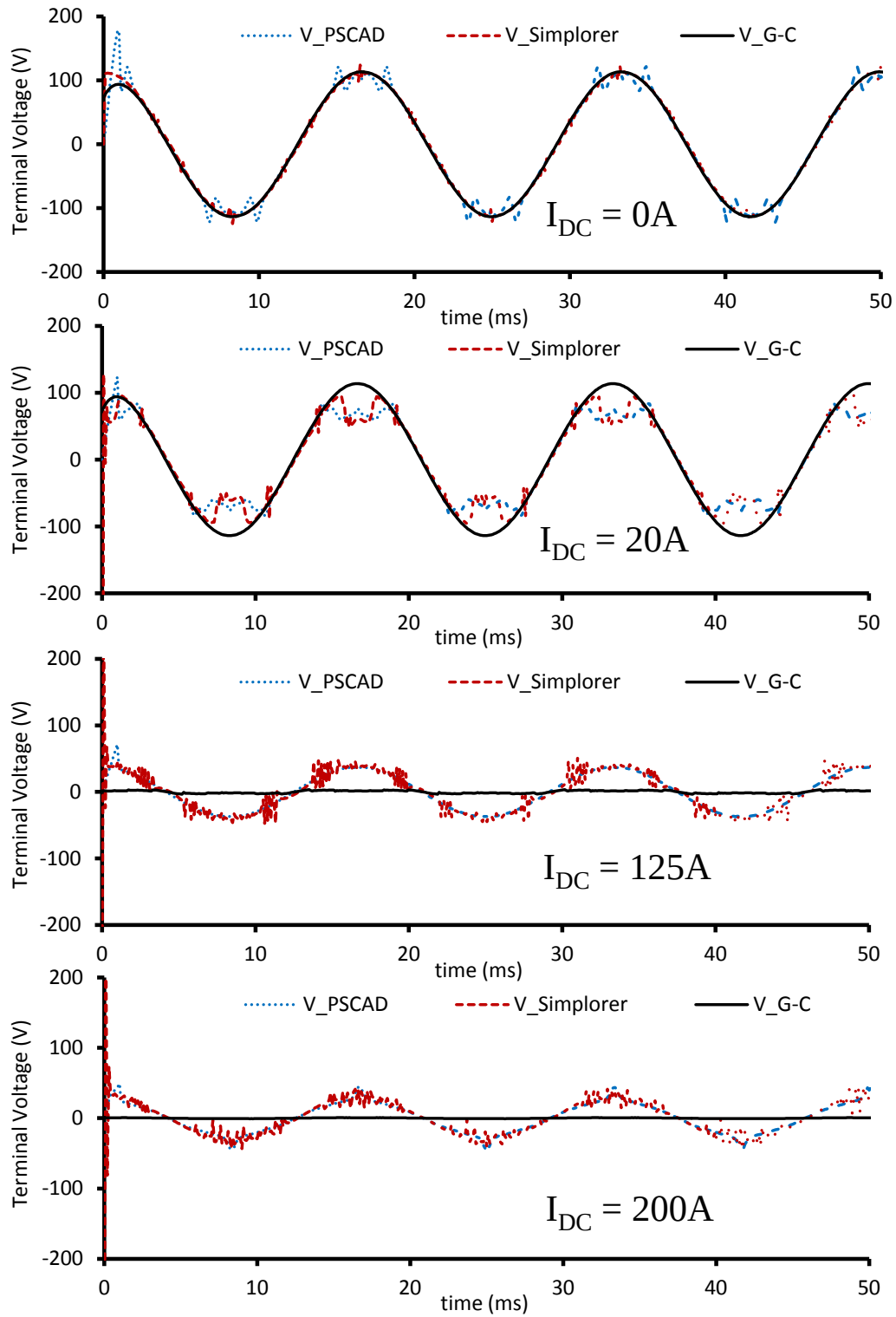


Figure 79 Mod-1 Gyrator-Capacitor Model, $I_{AC} = 125A$

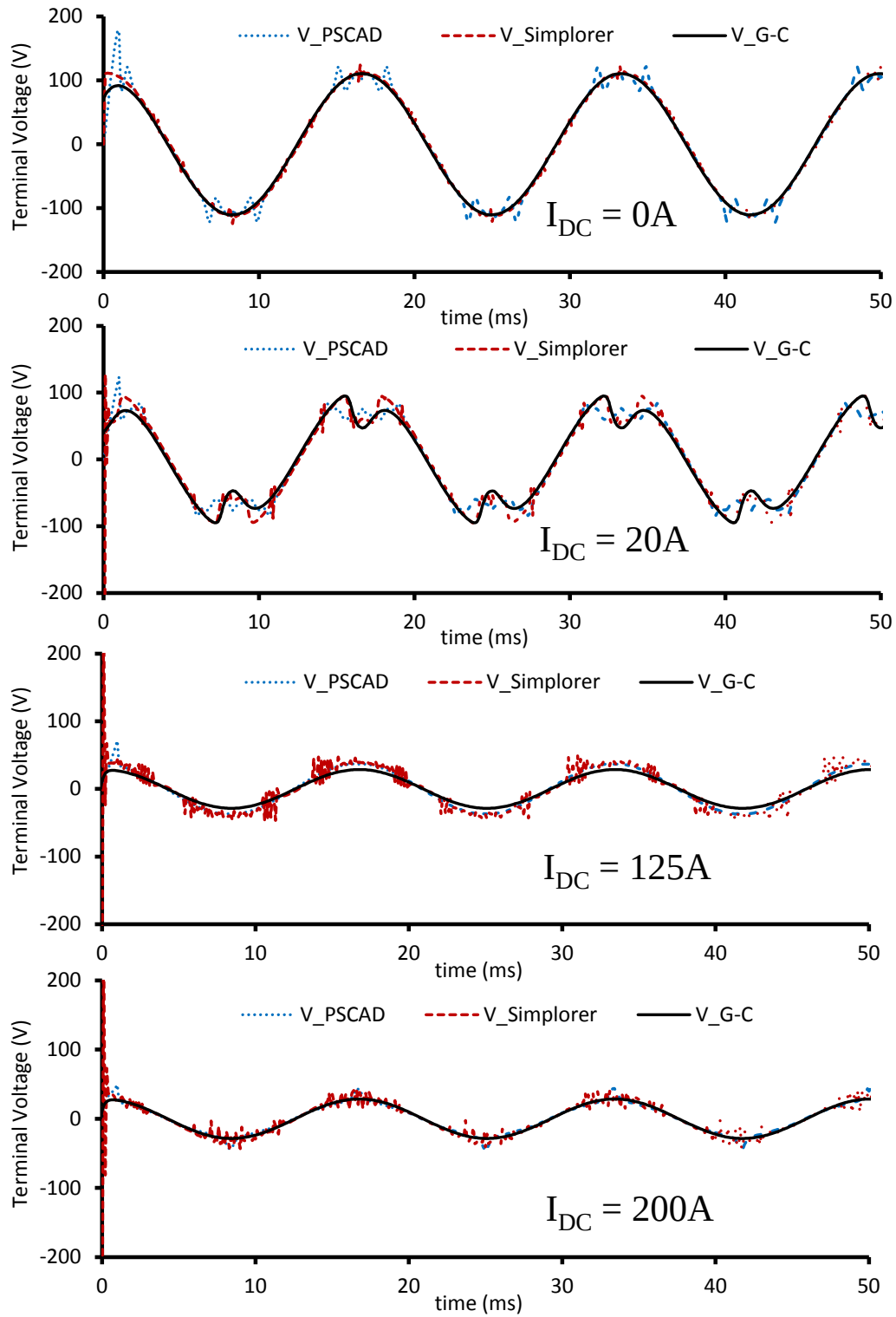


Figure 80 Mod-2 Gyrator-Capacitor Model, $I_{AC} = 125A$

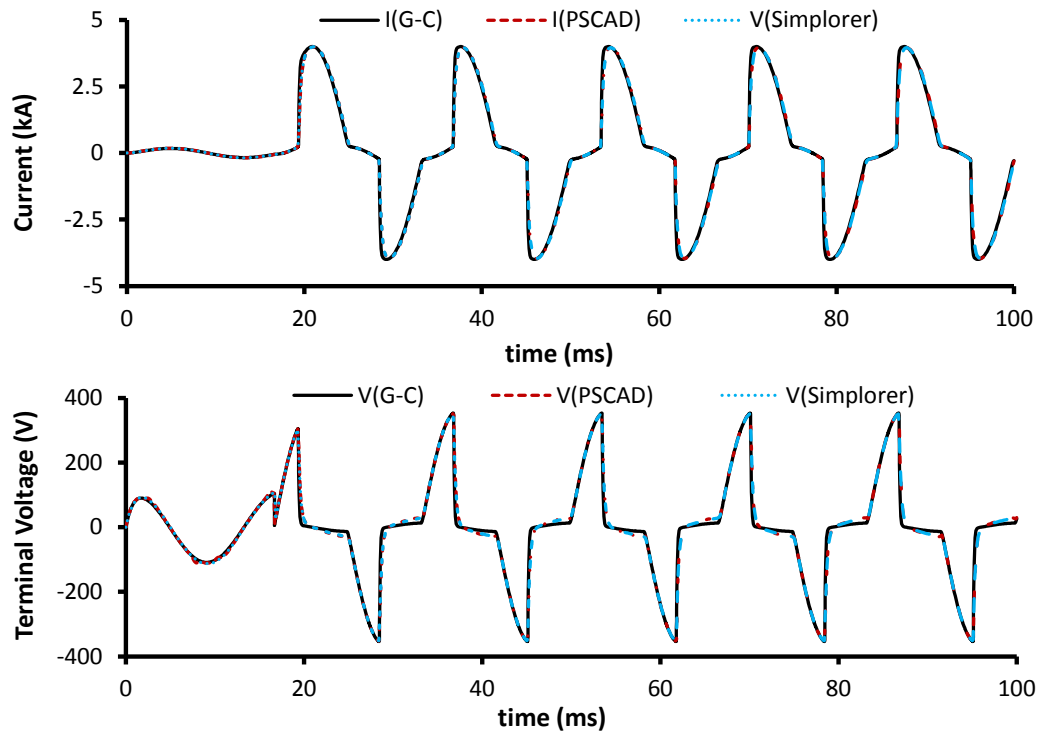


Figure 81 Mod-1 Gyrator-Capacitor Model Fault Simulations for $I_{DC} = 0A$

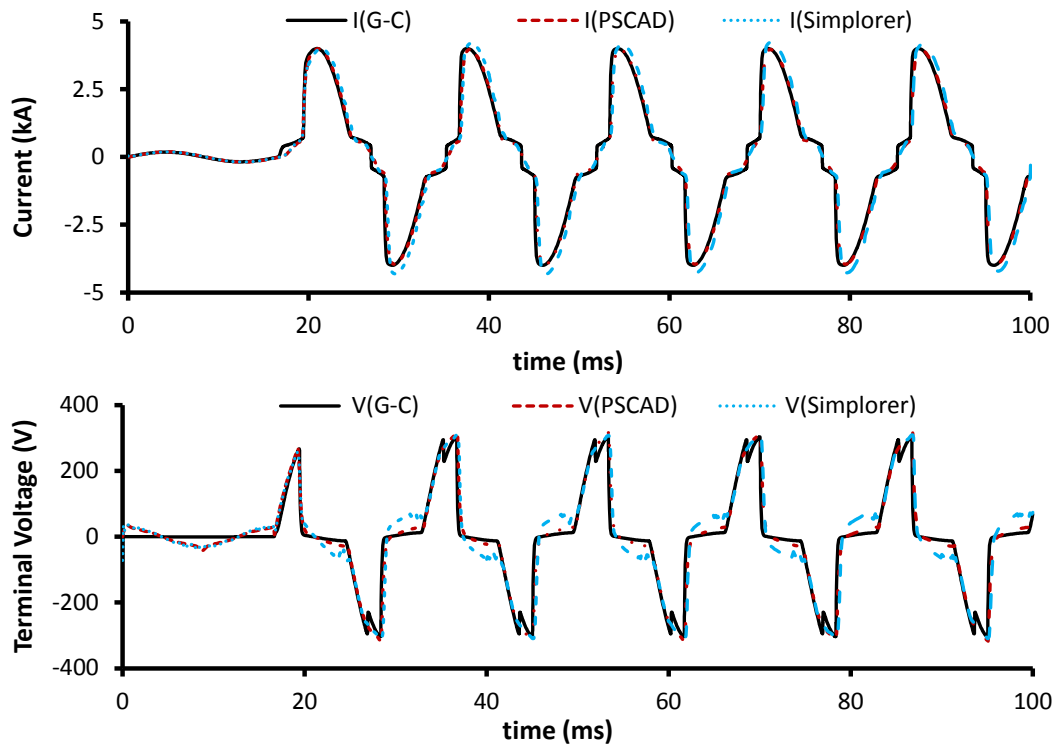


Figure 82 Mod-1 Gyrator-Capacitor Model Fault Simulations for $I_{DC} = 200A$

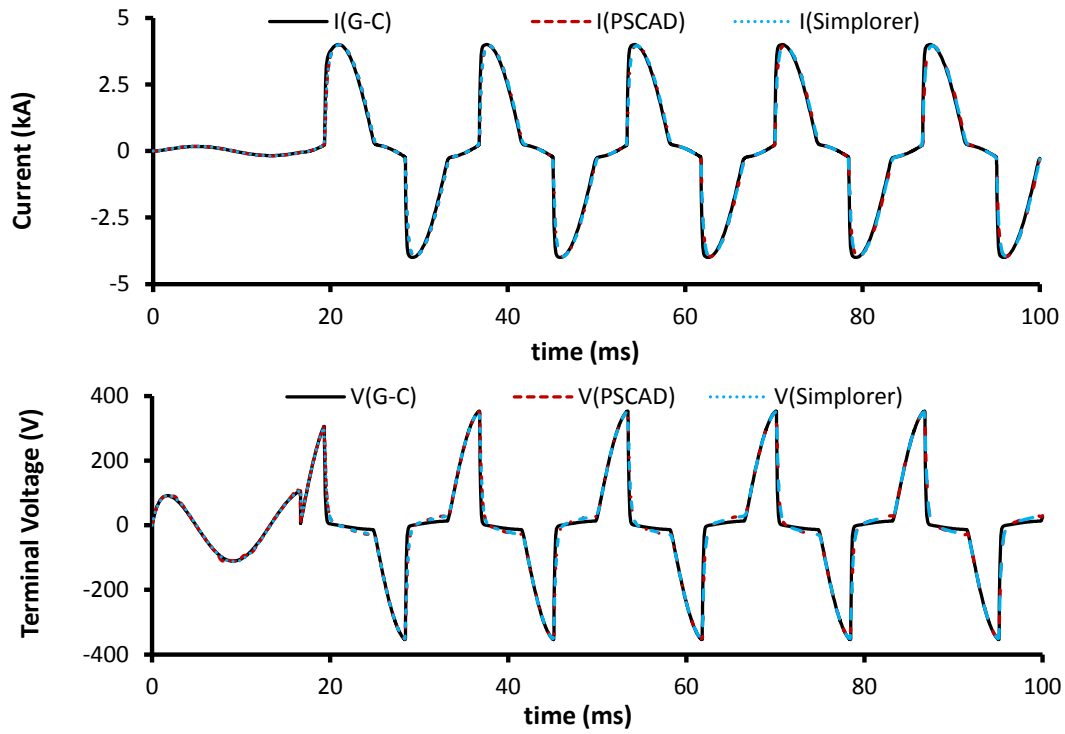


Figure 83 Mod-2 Gyrator-Capacitor Model Fault Simulations for $I_{DC} = 0A$

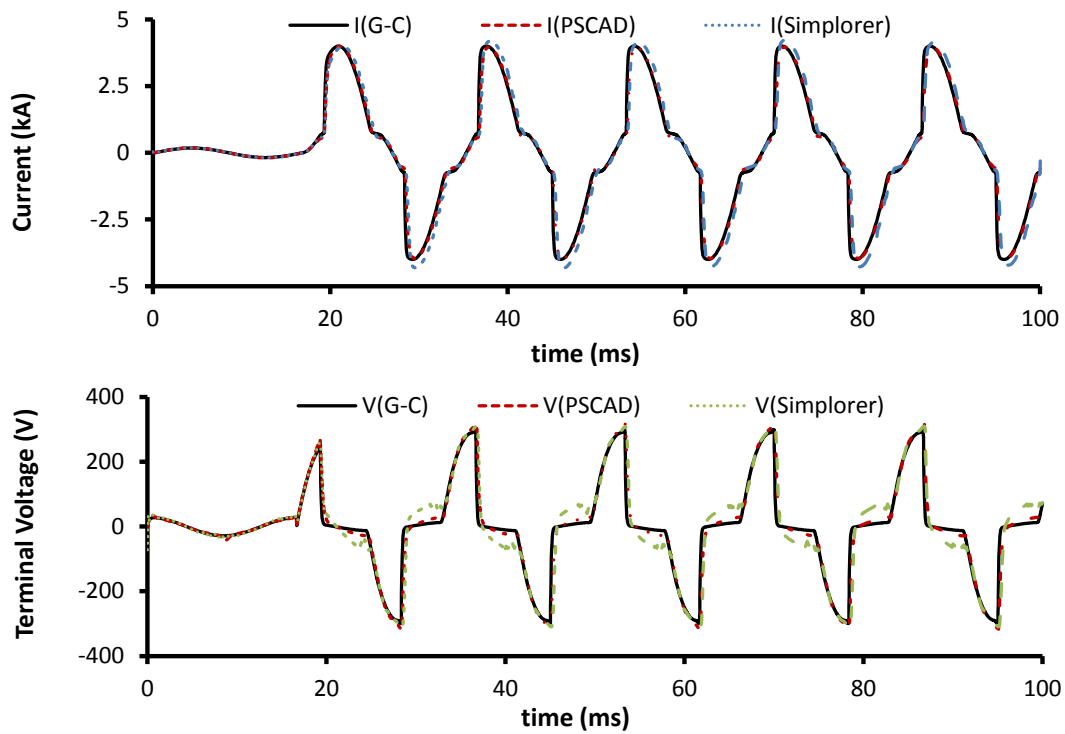


Figure 84 Mod-2 Gyrator-Capacitor Model Fault Simulations for $I_{DC} = 200A$

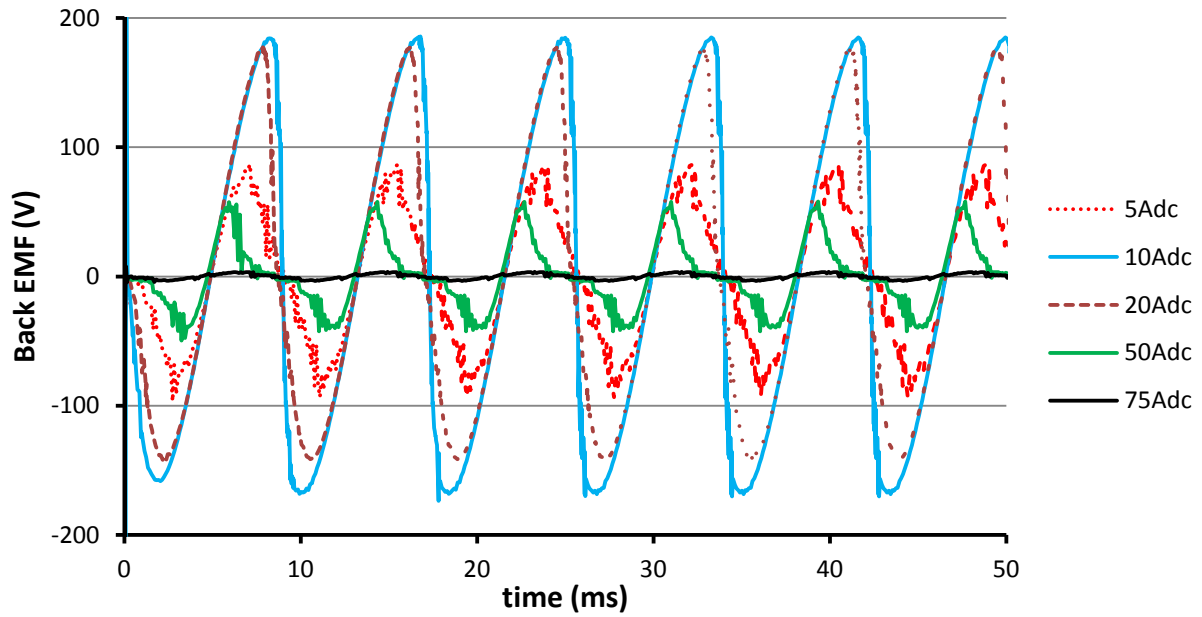


Figure 85 Back EMF of Mod-2 Gyrator-Capacitor Model, $I_{AC} = 125A$

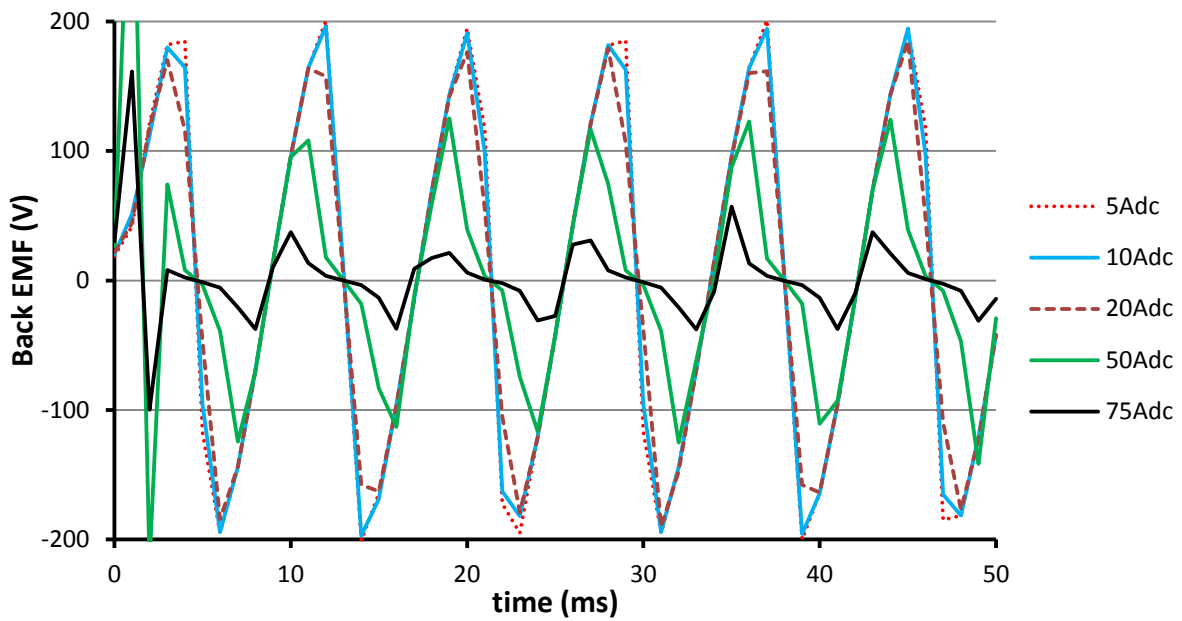


Figure 86 Back EMF from FEA, $I_{AC} = 125A$

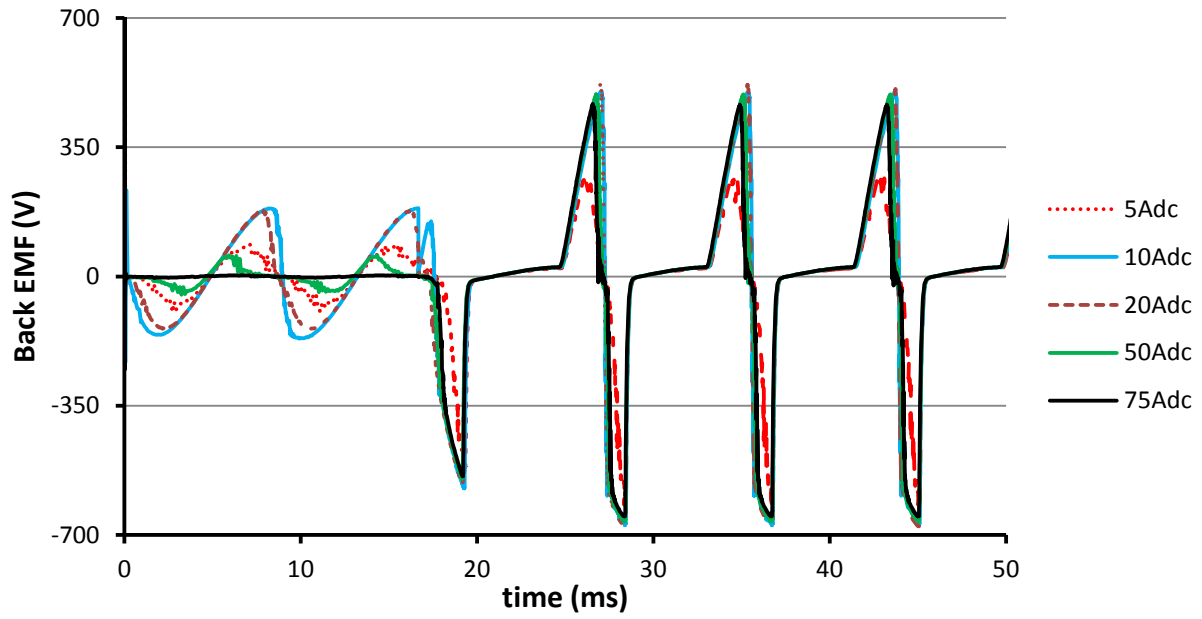


Figure 87 Back EMF of Mod-2 Gyrator-Capacitor Model under Fault

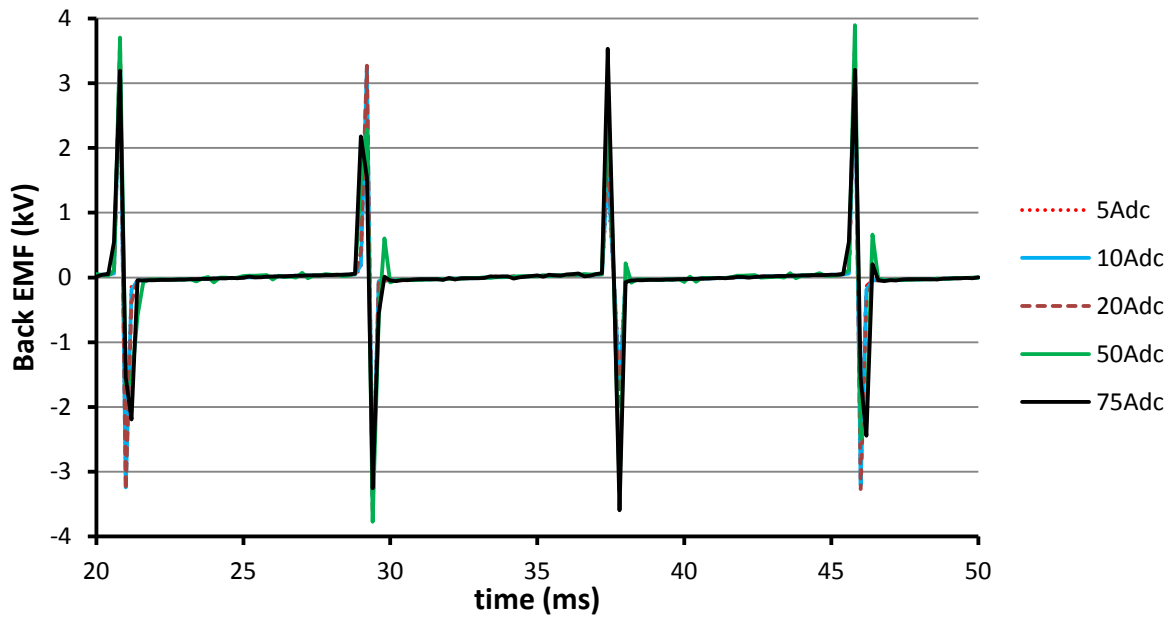


Figure 88 Back EMF from FEA under Fault

CHAPTER V: IMPACT STUDY ON DISTANCE PROTECTION

The behavior of the FEA-derived and G-C models constructed in Chapter III and Chapter IV describe how the SRPFC changes its operating state dynamically, and within each half cycle, as fault current drives the AC flux in and out of saturation. This behavior could make the line appear more inductive than the line actually is, causing the relay to under-reach its forward zone settings. If the amount of under-reach is significant, then the relay may fail to trip at required speed. Additionally, since the impedance changes on a sub-cycle basis, the relay's line impedance calculation may produce results that do not adequately reflect the actual impact of the SRPFC. The work presented here will address both of these issues.

This chapter evaluates the impacts of the SRPFC on line distance protection by application of the FEA-derived model to power system simulation under fault conditions. The resulting voltage and current waveforms are applied to a commercial line distance relay using a relay test set to observe impacts.

Distance Relaying

A distance relay calculates the impedance of a transmission line through measurement of the three-phase voltages and currents at one end of the line. If the line impedance changes such that it falls within a predetermined trip characteristic, then the relay detects that a fault exists on the line and executes a trip command to the appropriate breaker(s) that remove the line from service. Typically, the impedance of transmission lines are considered constant as they are of fixed length, so faults along a line effectively reduce the impedance seen by the relay. Distance relays use this principle to determine the location of faults in order to execute the appropriate protection action. The presence of dynamic impedances in the network can negatively affect distance protection, as the assumption of constant line impedance no longer holds true.

Application Practice

Prevailing industry practice is to minimize the number line outages by tripping only those breakers closest to the fault while keeping adjacent lines in service. This task is accomplished by

placing distance relays at both ends of each line and coordinating their operation by establishing zones of protection [21] [63]. There are typically three zones of protection for each relay as shown for Relay A in Figure 89. The primary zone reaches out 80% to 90% of the line distance and operates instantaneously upon detection of a fault. A second zone reaches the entire length of the line plus approximately 20% of the next line (120% total). The purpose of zone 2 is to provide backup for zone 1 of Relays A and B. To avoid conflicts with overlapping zones, zone 2 is delayed to allow zone 1 of the primary and the adjacent relays to respond first. An optional zone 3 provides additional backup protection by reaching out beyond zone 2 in the forward direction or behind the relay in the reverse direction. Delay settings for zone 3 are significantly larger than for zone 2.

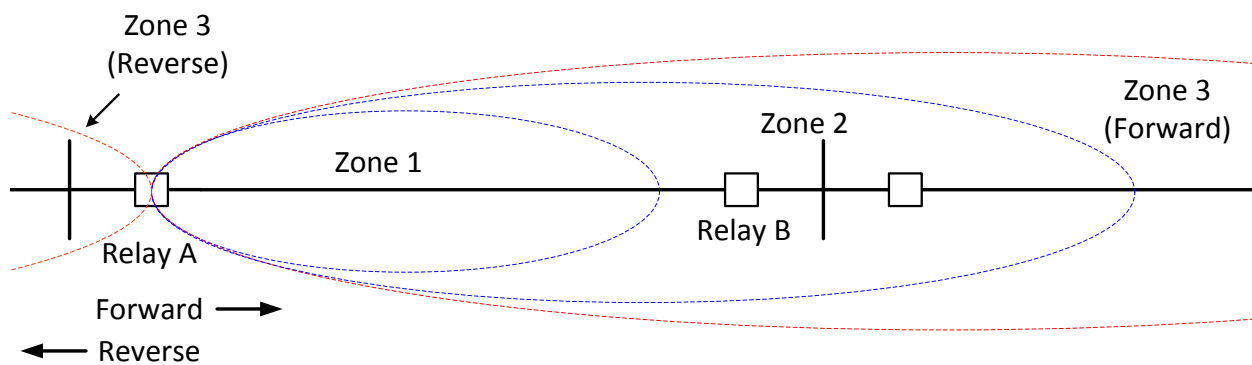


Figure 89 Zones of Protection for Transmission Line Distance Relays

The R-X diagram in Figure 90 describes a relay “mho” characteristic for a distance relay with two forward zones and a reverse zone 3. The mho characteristic is typically a circle defined by resistance (horizontal axis) and apparent reactance (vertical axis). The center of the circle is offset from the origin to make the distance relay directional. The impedance observed by the relay moves through the R-X plane as events occur. The movement of the impedance into one of the mho circles arms one or more of the distance elements within the relay. When the impedance falls within the zone 1 circle the relay trips instantaneously; while movement of the impedance within zones 2 or 3 (outside of the zone 1 circle) arm distance elements that result

in a trip only if the impedance stays within these zones for the duration of the time delay setting.

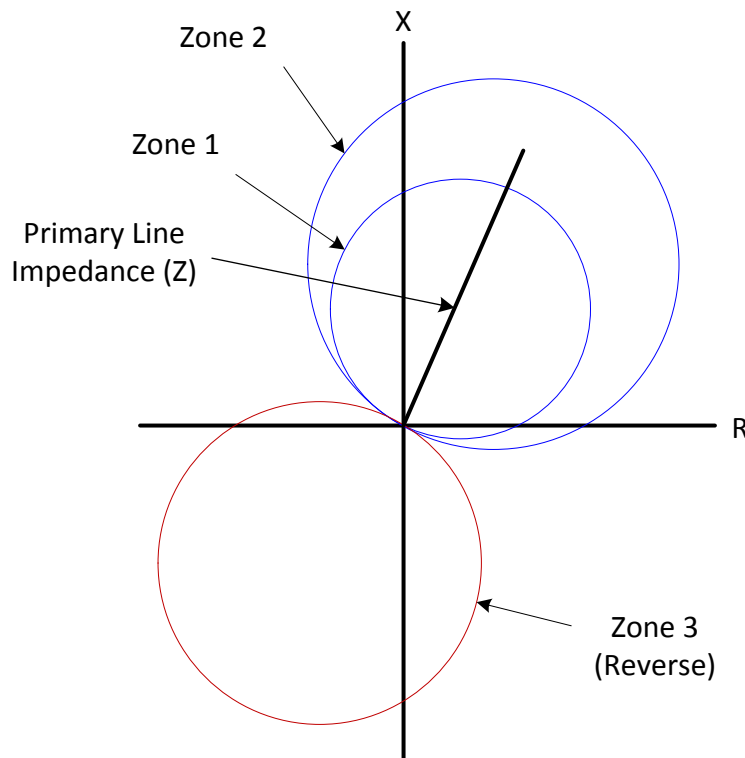


Figure 90 Zones of Protection on an R-X Diagram (mho Characteristic)

Dynamic Impedance

Modern distance relays use digital processes to perform the functions of their electromechanical predecessors and make calculations of line impedance that are based on fundamental frequency components. These processes include filtering the measurements to remove much of the harmonic content and to isolate the fundamental components, calculation of the line impedance, and comparison of the impedance to the trip settings. The performance of these algorithms can be adversely affected if the dynamic impedance significantly distorts the voltage and/or current waveforms [64] [26]. Additionally, these algorithms may not accurately represent the actual impedance if significant portions of the dynamic impedance occur outside of the fundamental frequency. Series compensation devices, like series

capacitors, are examples of equipment that introduce nonlinear variation and harmonics in transmission lines [65]. The *IEEE Guide for Protective Relay Applications to Transmission Lines* (IEEE Std C37.113-1999) suggests that protection engineers must pay careful attention to consider all of the implications when series compensation devices are in use [66]. In addition to the relay, measurement equipment may also be affected. For instance, coupling capacitor voltage transformers (CCVTs) can interact with dynamic impedances to cause spurious voltage measurements at the relay [67]. Electromechanical relays are also susceptible to issues related to the dynamic impedance, but are not as vulnerable as microprocessor relays to waveform distortion or CCVT-related issues.

Impacts of Dynamic Impedance Devices

Series-connected devices that introduce dynamic impedance on transmission lines are commonplace in the electric power industry and are known to affect the operation of line distance relays [68]. These devices typically provide reactive power support, but there also exist devices that regulate power flow.

Summary of Devices

A review of following series connected devices provides insight on distance protection issues caused by dynamic impedance.

- Fixed Series Capacitors
- Flexible AC Transmission Systems
- Fault Current Limiters
- Distributed Power Flow Controllers

Fixed series capacitors are commonly used to boost system voltage at strategic locations within a transmission network. Lines are inductive by nature; and long lines can produce voltage drops in excess of voltage stability requirements. In these situations, FSCs are often a cost-effective mitigating solution. However, switching an FSC into service causes the effective impedance of the line to change [21]. The presence of protection devices (metal oxide varistors and spark gaps) that protect the FSC device under fault conditions further exasperate the problem [69].

Spark gaps introduce variable resistance and metal oxide varistors introduce varying and nonlinear resistance. Additionally, FSC can also cause transients on a power system and create RLC resonance that results in harmonics near the fundamental frequency that can lead to undesirable effects on instrument transformers and/or impedance calculations within the relay. A representation of a typical FSC application with protection components is shown in Figure 91.

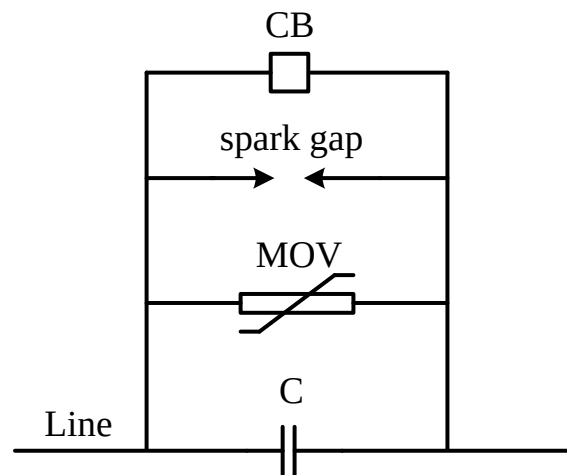


Figure 91 Fixed Series Capacitor with Protection Components

Consider an FSC located at the source end of the line as shown in Figure 92. The line impedance will assume the nominal line characteristic from the relay up to the location of the FSC. From location of the FSC into the fault, the impedance observed by the relay will be lower than the nominal line characteristic as the capacitor has increased the line voltage. Therefore, a fault located at the end of the line in zone 2 could cause the relay to detect the fault in zone 1, causing a false zone 1 trip. Another phenomenon, voltage reversal, causes the relay to lose its directionality. Voltage reversal occurs when the negative reactance of the FSC is higher than the reactance of the line up to the fault location. This condition causes the relay to detect the fault location to be behind the relay. An example of voltage reversal, shown in Figure 93, shows how the line voltage at the fault is the same level as behind the relay. In this case, the relay would respond by arming a reverse zone 3 rather than executing an instantaneous zone 1 trip. The tendency of FSC to cause over-reach and loss of directionality, combined with the nonlinear

effects of the protection devices, make distance protection with FSC a challenging task for protection engineers.

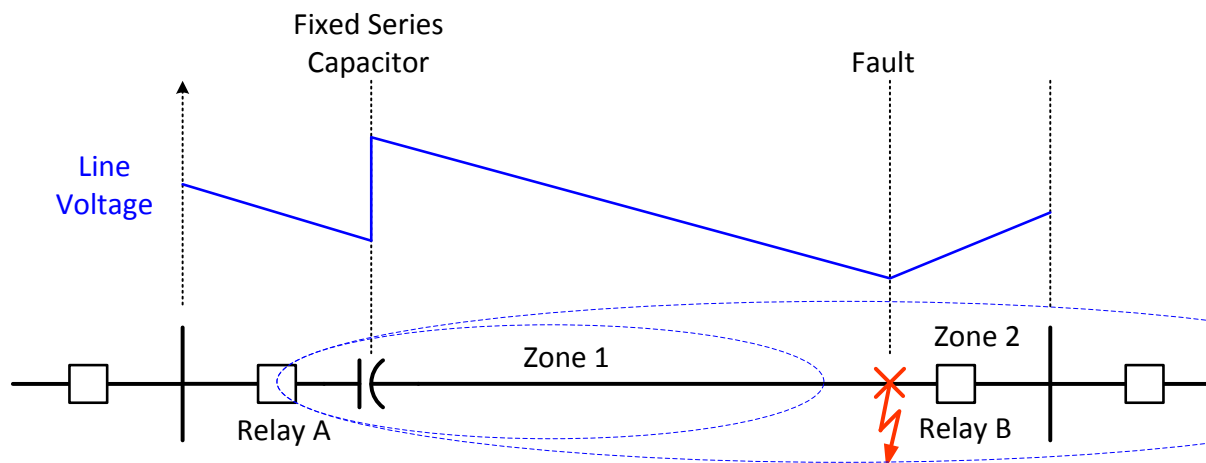


Figure 92 False Zone 1 Trip due to Relative Location of FSC and Phase-Ground Fault

FACTS devices can perform both power flow control and reactive compensation. These devices allow modulation of AC line conditions by using power electronics and advanced control algorithms. The primary cause of issues related to FACTS on distance protection are the changing voltage and current injections that affect the impedance estimate of the relay [70] [71] [72]. When the estimated impedance is compared to the set impedance characteristic, the location of the fault is not correctly determined by the relay, which results in miss-operation. Miss-operations can include over-reaching/under-reaching, delayed trips, improper phasing, and errors in fault location estimation. The active nature and possible number of operational states of FACTS devices contribute to the complexity of the distance protection issues. CCVTs present another concern as these devices are tuned to the line and are known to be adversely affected by certain transients [73].

Fault current limiters (FCLs) quickly increase the line impedance in response to faults to effectively reduce the magnitude of fault currents. Types of FCLs include resistive-type, inductive-type, saturable-core FCL, and power electronics applications [74]. Previous research

suggests that the sub-cycle impedance change can affect distance protection [75] [76]. A saturable-core FCL operates on the same physics principles as the SRPFC, but is designed to increase the line reactance at the peak portions of the fault current. Its design and method of operation make the saturable-core FCL ideal for limiting fault currents as opposed to power flow control. The device operates in saturation under nominal system conditions and allows the peak values of the fault current to drive it into the linear region. A saturable-core reactor was installed at the Puji substation in Kunming, China [11]. One literature source suggests an adaptive relay scheme to switch the relay group settings when the saturable-core reactor is in service in order to account for its impedance characteristic on the line [77]. The relay settings groups in this case are based on the line itself without an FCL in service, and the effective impedance of the saturable-core reactor during nominal and during fault conditions.

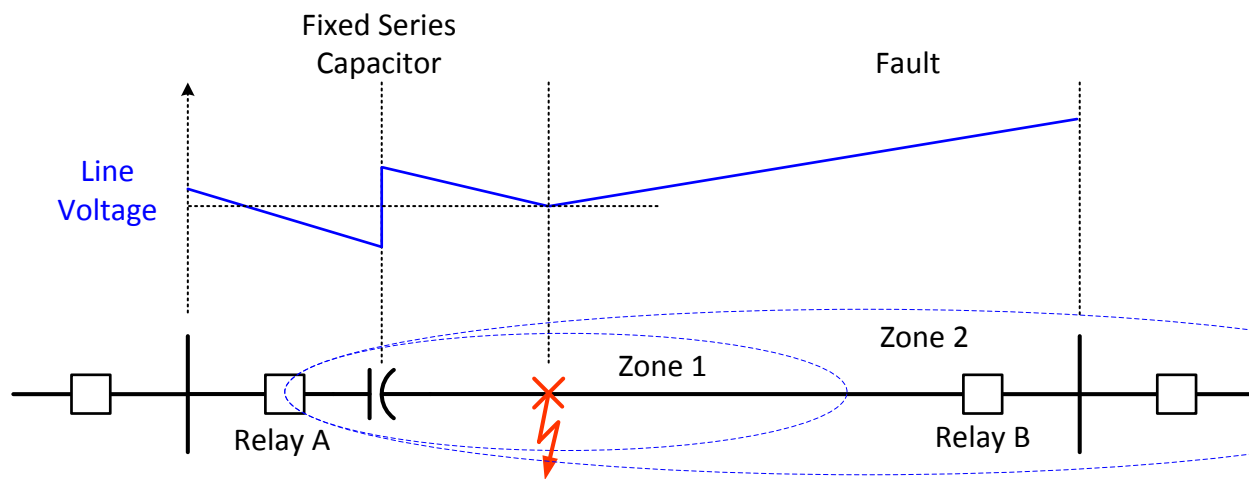


Figure 93 Voltage Reversal Causing False Directional Characteristic

Smart Wire Grid[®] is a distributed power flow technology that uses small power electronics devices mounted directly to the line at transmission towers [78]. A large number of these distributed devices, which are coordinated through communications, are required to provide significant impedance change on a transmission line. Currently, there is minimal information on how such devices will effect distance protection, but there are claims that the technology

avoids protection issues through a by-pass scheme that occurs before distance protection elements become armed [79].

Recommended Mitigations in Literature

FSC, and in many operational cases FACTS devices, add line capacitance that effectively reduces the impedance of a transmission line. Thus, distance protection will tend to over-reach as the line length will appear smaller with respect to the set impedance characteristic of the relay. The significance of the over-reach is dependent on several factors, including the type of series compensation device and its operating state. The *IEEE Guide for Protective Relay Applications to Transmission Lines* (IEEE Std C37.113-1999) provides suggestions on how to address series compensation and transmission line power flow devices [66]. For series compensation elements, the IEEE guide suggests that the distance relay record a memory voltage so that it can detect voltage reversal. Additionally, it suggests that compensating for over-reach in the relay settings is too simplistic and only accounts for the fundamental component. To account for all harmonics, the IEEE guide suggests that transient tests of protection systems with detailed models of the transmission line plus adjacent lines be performed in order to evaluate the protection system for a given installation. While these suggestions are focused more towards FSC applications, they cover all series compensation devices. For series reactors, the IEEE guide suggests that an adaptive relaying system be implemented to switch the relay settings when the reactor is by-passed and when it is in service.

Mitigation strategies throughout literature typically focus on two methods: communication-based distance protection and adaptive relaying. The communications-based relaying schemes suggested in [80] [65] [81] [82] can be utilized to perform phase comparison, which is immune to some of the protection issues associated with changing line impedance. Phase comparison relies on the difference between the current phase angles at the ends of a line to determine if the fault occurred within the line or outside of the line. Additionally, pilot-relaying schemes including Directional Comparison Blocking (DCB), Permissive Over-reaching Transfer Trip (POTT), and Direction Comparison Unblocking (DCUB) can be used to coordinate distance relays to work around series compensation and FACTS-based issues. While communications-based

distance schemes can provide mitigation, they introduce an additional reliability component in terms of communications channels. Adaptive relay schemes are suggested in [83] [84] [85] [86] [87] to change the relay protective boundaries in-situ. The various techniques proposed, some of them novel, automatically change relay group settings or relay parameters based on measured values and/or control inputs. Methodologies based on artificial neural networks, synchrophasors, and other methods are proposed.

In summary, series compensation and power flow elements can cause problems with distance protection that require careful consideration in each application. Since each type of device presents unique issues, and because all power systems are unique, careful testing and analysis is required in each instance to ensure proper operation of the protection system.

Potential Impacts of the SRPFC

The SRPFC introduces an inductive reactance into a transmission line that is variable, similar to FACTS devices that operate in an inductive mode but with a different impedance profile and operation speed. Thus, the primary impact on distance protection is an under-reaching effect, which is contrary to that of FSC. SRPFC adds reactance to the line, thereby making the line impedance appear larger to the distance relay as shown in Figure 94. It is possible that a fault in zone 1 with the MPAFC in service can cause the relay to determine that the fault occurred further down the line than the actual location, in zone 2 for instance. Such an occurrence would result in failure of the instantaneous zone 1 to trip and cause the fault to remain in the circuit until zone 2 times out or the adjacent relay responds. The significance of the under-reaching impact is dependent on the amount of impedance change observed by the relay during fault conditions and potential interactions of SRPFC operation with the transmission system and the measurement devices. The significance of this under-reach is unknown and requires power systems transient studies with an SRPFC model to understand the extent of the impact on distance protection.

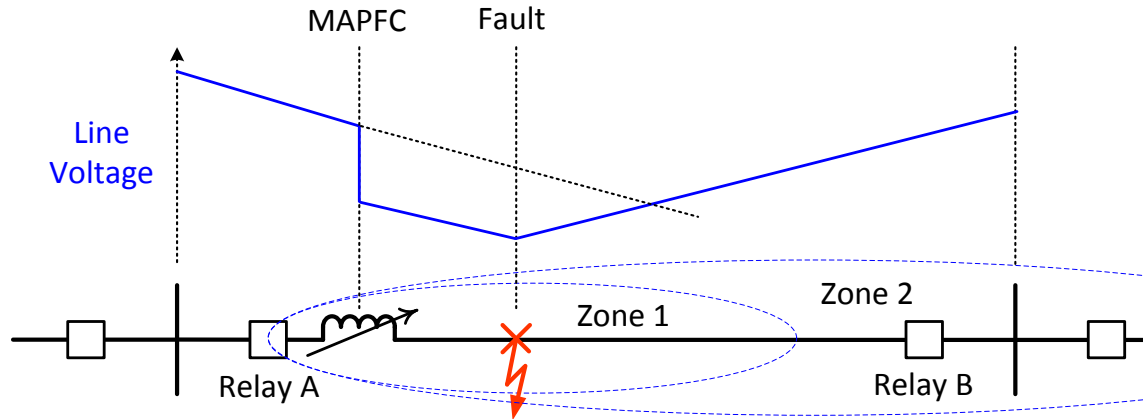


Figure 94 Failure to Trip by Zone 1 with SRPFC

Power System Study

Application of an SRPFC transient model in power system simulation is performed to observe how the device/system interactions affect distance protection. The SRPFC transient model, derived by FEA, is described in Chapter III. The Power Systems Relaying Committee (PSRC) system model for testing transmission line protection [88] is used for this study as it provides a faithful representation of power system transients and is an industry accepted tool for assessing distance protection schemes and relay performance in transmission systems. A specific configuration of the PSRC test system is applied that consists of two generators joined by two series-connected transmission lines. The system configuration avoids parallel transmission lines and tapped lines. However, the system does represent realistic generator response and the ability to supply fault current from both directions on a line. The one-line diagrams in Figure 95 describe the full PSRC test model and the configuration used to perform this study. An ideal source (S1) connected to Bus 1 and a synchronous machine (S2) connected through a delta/wye transformer on Bus 4 represent the two generators on the system. The controls of the synchronous machine were configured to reduce electromechanical oscillations under nominal operation, yet provide sufficient fault current under fault conditions. Potential transformers are located on the line-side and bus-side of the saturable reactor, which is located on the ZL1 side of Bus 1.

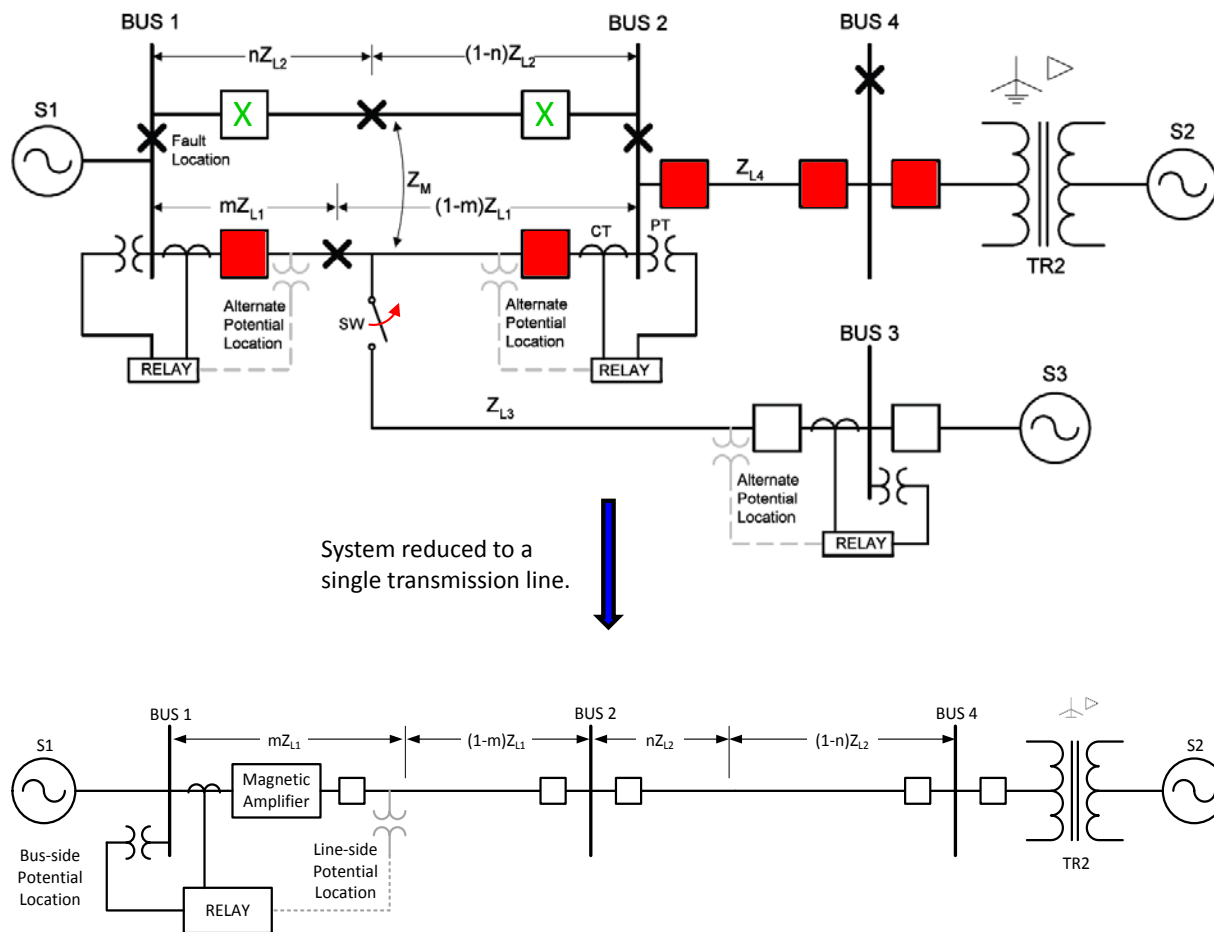


Figure 95 PSRC Power System Model for Testing Transmission Line Protection

Simulations for zone 1 testing were performed by applying faults along line Z_{L1} . To determine the zone 1 reach point, simulations were performed at 1% distance increments while approaching the zone 1 setting. The same method was used to test zone 2 reach along line Z_{L2} . In addition to the reach points, simulations were performed at 0%, 33%, and 66% of Z_{L1} to observe zone 1 operation with faults in close proximity to the saturable reactor. Potential measurements were recorded on both sides of the saturable reactor (bus-side and line-side) to observe differences in location of the measurement. Table 7 summarizes the simulation cases. To capture impacts over the operating range of the saturable reactor, simulations were performed at the minimum and maximum DC operating currents.

Table 7 Simulation Cases for SRPFC Simulation in the PSRC Power System Model

Fault Locations	Fault Types	Potential Locations	Saturable reactor Bias Current
Z_{L1} 0%	A-g	Bus-side	0A
Z_{L1} 33%	AB	Line-side	200A
Z_{L1} 66%	AB-g		
Z_{L1} (Z1 reach point)	ABC		
Z_{L2} (Z2 reach point)			

Fault Scenarios

There are 10 types of faults to consider when performing fault analysis. Six are permutations of the four fault types listed below:

- Line-ground: A-g, B-g, C-g
- Line-Line: AB, BC, AC
- Line-Line-ground: AB-g, BC-g, AC-g
- Three-phase: ABC

The assessment of the distance protection for this study is performed by considering one instance of each fault type (A-g, AB, AB-g, and ABC), so all of the remaining permutations are neglected. This approach is valid for two reasons: First, the SRPFC transient model is constructed from three independent, or de-coupled, single-phase models. Second, the PSRC test system is symmetric across its three phases, with exception of tower configuration. Since three identical SRPFC transient models compose the three-phase system, each fault permutation, within each fault type, should produce identical results on the PSRC test system.

The location of the potential measurements will effect relay operation. The line-side location does not see the full effect of the dynamic impedance for forward faults. However, the bus-side location will see all of the dynamics. The impact of measurement location on the voltage waveforms is shown in Figure 96 and Figure 97 for ABC-g close-in faults on line ZL1. The bus-

side measurement detects the dynamic inductance of the SRPFC, which is evident by significant voltage spikes that occur as the fault current approaches the zero crossings and the SRPFC transitions between the linear and saturation regions of operation. However, the line-side measurement observes only the line and fault impedance beyond the SRPFC. In both cases, this SRPFC behavior causes slight changes in the current that are noticeable at the peaks of the current waveforms. Close-in faults represent the worst case in terms of how the SRPFC dominates the shape of the voltage and current waveforms. As the distance from the relay to the fault increases (e.g. for faults that occur further down the line), the impacts of the SRPFC become less significant as the line impedance dominates. This behavior is observed in Figure 98, Figure 99, and Figure 100, where ABC faults are applied at lines distance of 33% and 87% of the total line length. At a distance of 33%, there is still noticeable distortion, but the voltage waveforms are nearly sinusoidal as the line impedance begins to dominate the total impedance seen by the relay. At 87% of the line distance, a point that resulted in zone 1 trip, there is little distortion in the voltage waveform. In both of these cases, the current waveforms appear purely sinusoidal.

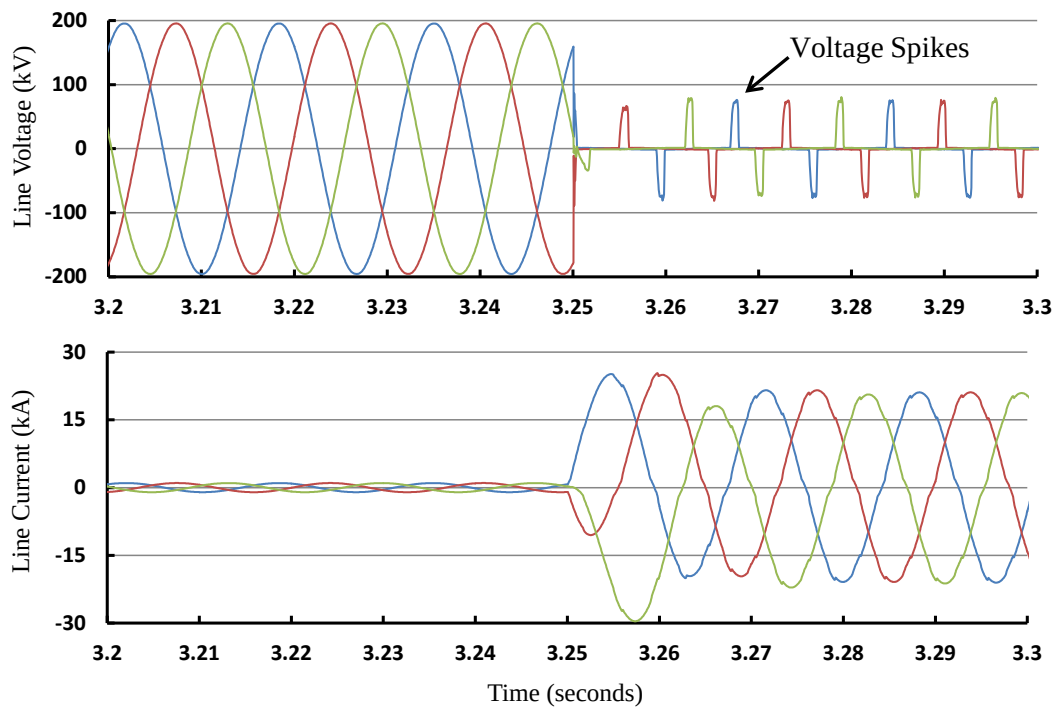


Figure 96 ABC Close-in Fault with Bus-side Measurements

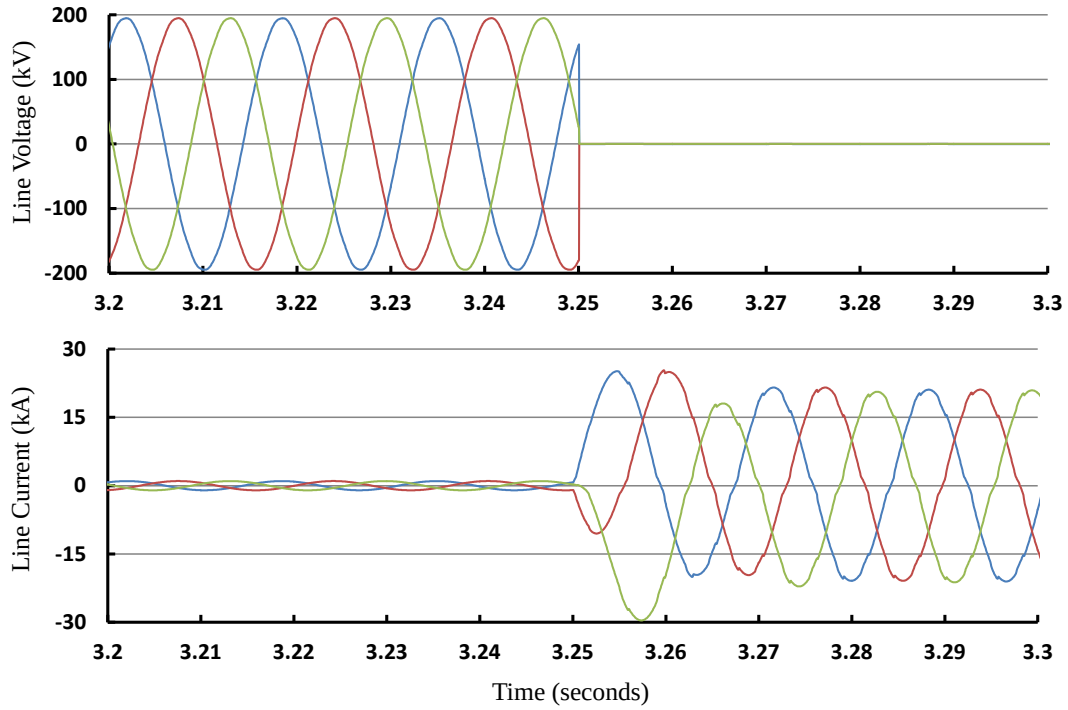


Figure 97 ABC Close-in Fault with Line-side Measurements

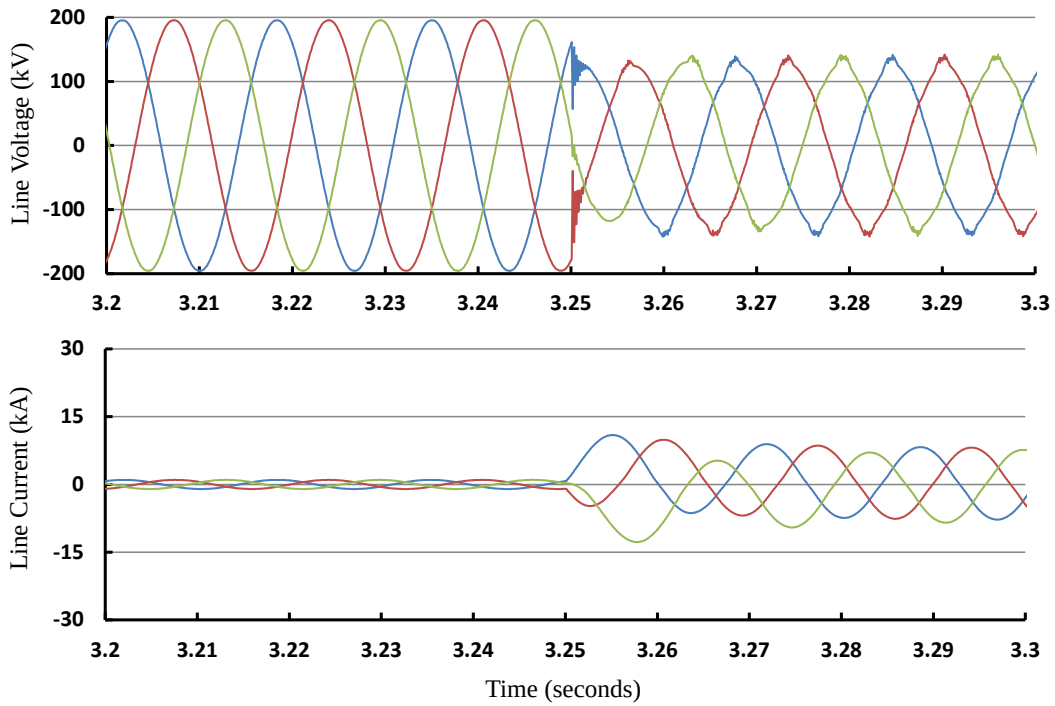


Figure 98 ABC Fault at 33% Line Length with Bus-side Measurements

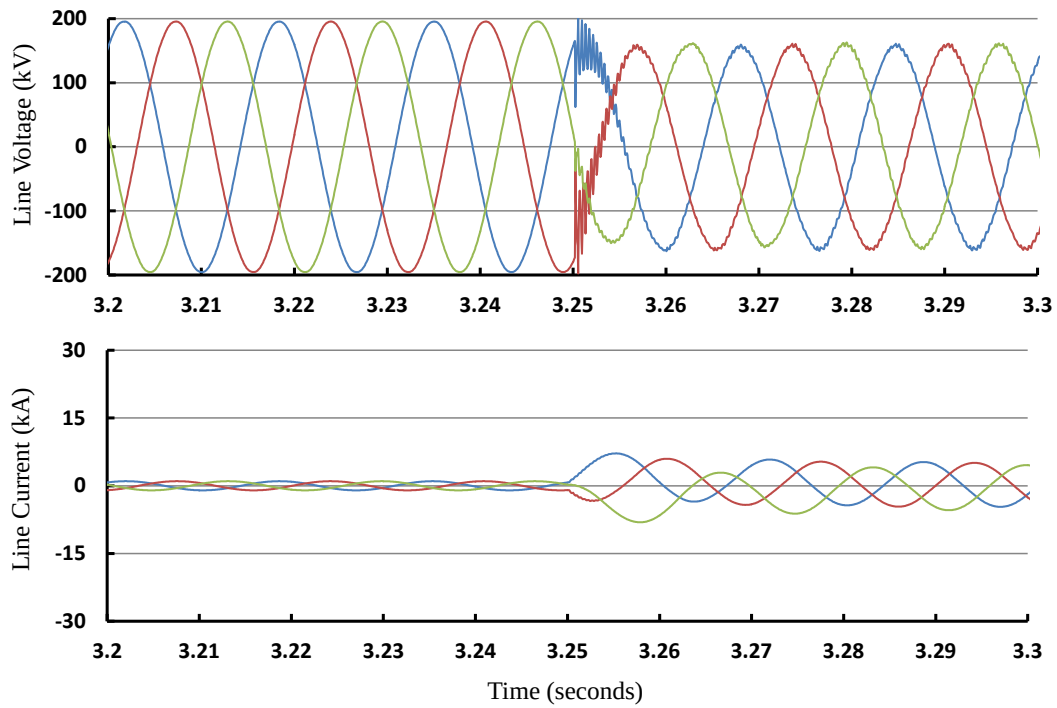


Figure 99 ABC Fault at 66% Line Length with Bus-side Measurements

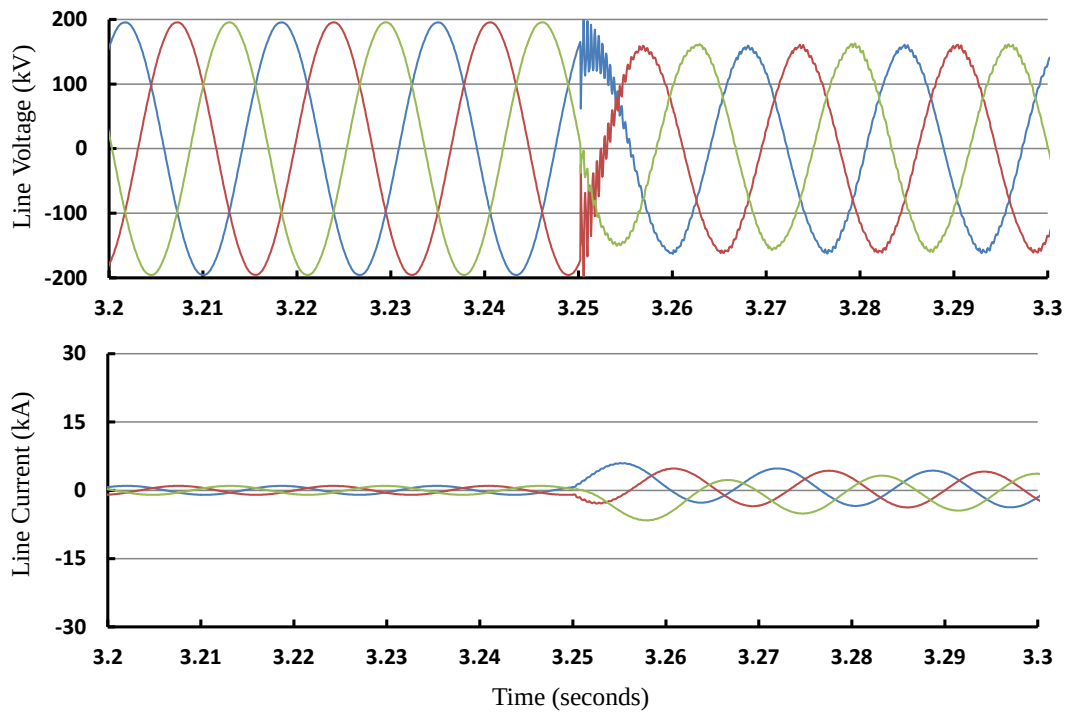


Figure 100 ABC Fault at 87% Line Length with Bus-side Measurements

Simulation Parameters

Simulations were performed using PSCAD™ [48]. In all cases, the duration of prefault simulation is 3.5 seconds that is followed by 0.5 seconds of fault condition. The total simulation time is 4 seconds. The 3.5 seconds of prefault allows the test relay adequate time to initialize under nominal power system conditions upon playback. Simulation time steps were typically set to 50 μ s. However, reduced times steps of 25 μ s and 10 μ s were required for cases when a line segment was less than 10% of the total line length. The smaller time steps enable PSCAD™ to solve the transmission line equations when the length of any line is significantly short. Simulation results were recorded in COMTRADE [89] format for relay testing.

CCVT Model

A CCVT³ is a capacitive voltage divider with a step down transformer, burden impedance, and a ferroresonant filter circuit that is used to provide secondary voltages to protection relays. The equivalent circuit model in Figure 101 describes the arrangement of CCVT circuit elements. CCVTs are a prevalent source of potential measurements in high voltage transmission systems above 138kV, as they are economically advantageous when compared to potential transformers for high voltage levels [67]. A drawback of CCVTs is the interaction of device capacitance with the inductance of the transmission system, which can lead to resonant circuits and RC time constants that cause misrepresentation of the secondary voltage. For this reason, it is necessary to include a CCVT model into the distance protection study in order to observe potential interactions with the SRPFC and its impact on distance protection.

The transient effects of CCVT operation are well known and the subject of previous research efforts [90] [67]. The most significant issue is the ringdown effect of the secondary voltage at the onset of fault conditions that is caused by the internal energy storage components of the device. The severity of the ringdown depends on multiple factors, including CCVT parameters

³ A capacitive voltage transformer (CVT) is effectively a CCVT without power line carrier accessories, so both terms are often used interchangeably.

and point on wave when a fault occurs [90]. The ringdown effect tends to cause over-reach by distance relays as depression of the 60Hz component make the observed line impedance appear smaller than the actual impedance to the fault [73] [91] [92]. Additionally, the presence of capacitance, inductance, and nonlinear elements in CCVTs reduce the fidelity of the secondary voltage under transient conditions that can also lead to problems with distance protection [93]. Microprocessor relays provide means to mitigate these issues, but impacts require careful understanding in order to properly tune CCVT parameters and determine relay settings.

The CCVT model described in Figure 101 exists in the PSCAD™ component library. In order to produce a faithful representation of CCVTs prevalent in industry, the model was tuned using the parameters of a Trench® CCVT [94]. Refer to Appendix D for the CCVT parameters applied in this study.

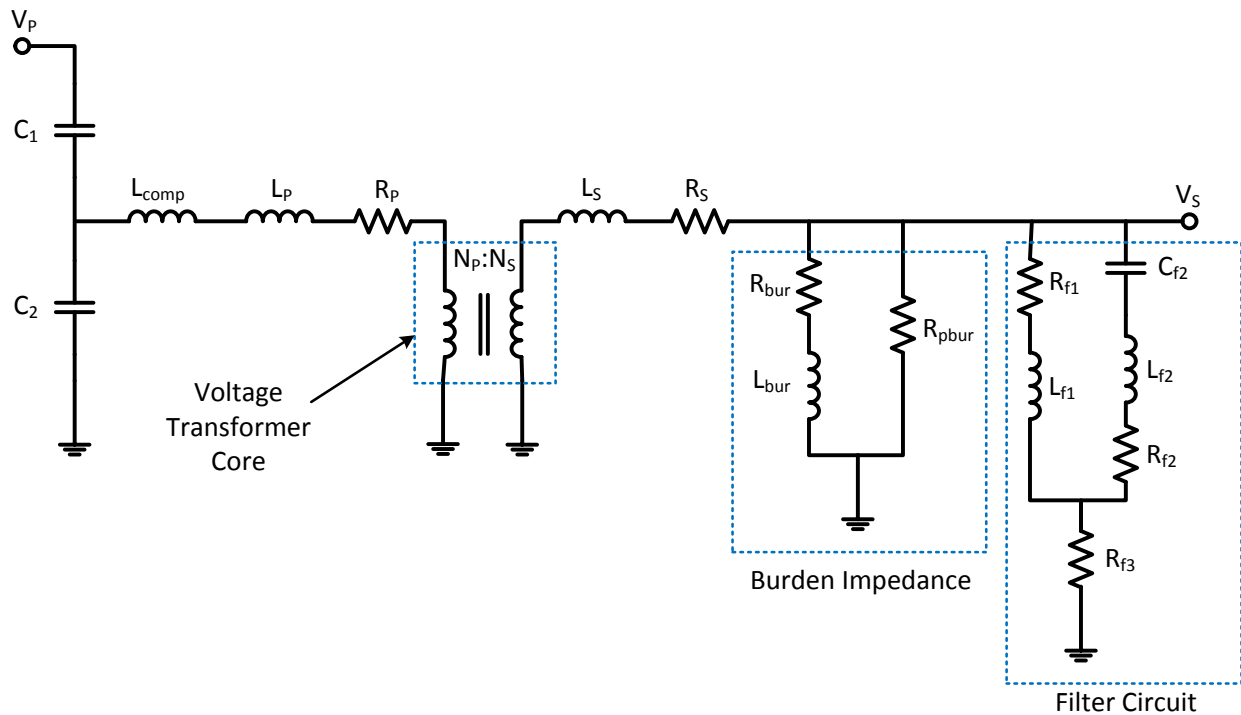


Figure 101 CCVT Circuit Model in PSCAD™

The voltage waveforms in Figure 102 and Figure 103 are the results of PSCAD™ simulations with close-in ABC faults. Close-in faults should result in little to no impedance between the relay and the fault location, as the fault location occurs very close to the relay, so the voltage observed by the relay is effectively zero. The voltage waveforms in Figure 102 represent a fault simulation with the SRPFC out of service. These waveforms possess noticeable ringdown immediately following fault inception due to the internal charging components of the CCVT. The frequency of the ringdown is much slower than the line frequency – approximately 3Hz in this case.

Figure 103 compares phase-C to neutral voltages with the SRPFC in and out of service. The CCVT captures the distortion of the voltage waveform as the voltage spikes travel along the 3Hz ringdown wave. In contrast, the voltage waveform from direct measurement (no CCVTs) of the voltage with SRPFC in service contains the waveform distortion from the SRPFC, but it does not contain the ringdown effect.

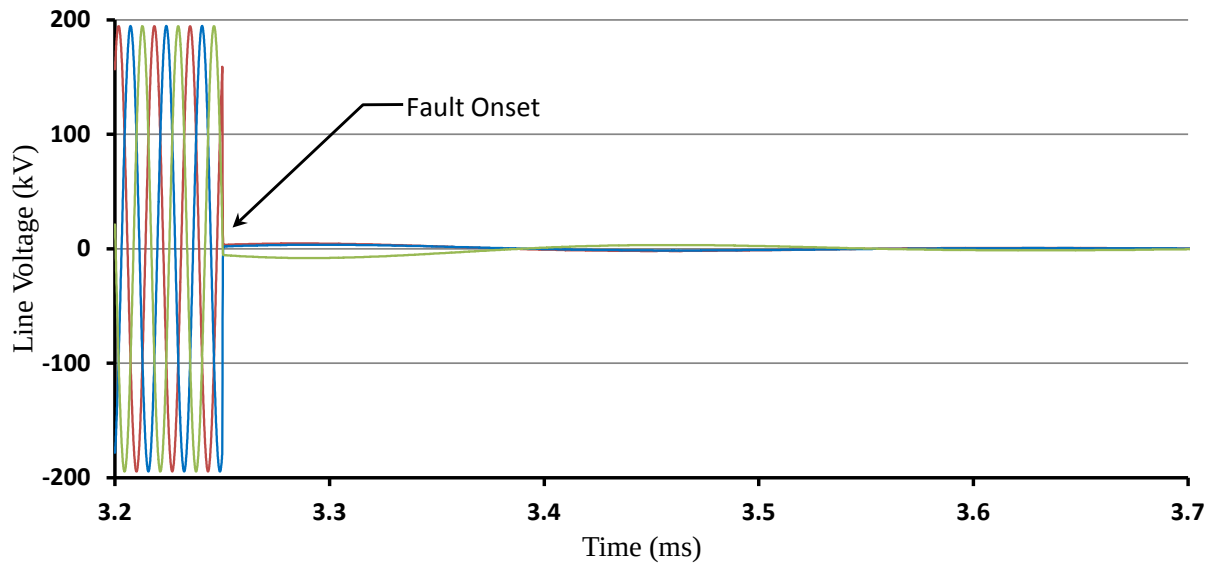
In the case of an ABC close-in fault, the relay's reach will likely not be affected by the ringdown as the 60Hz component is naturally forced to zero and the relay's impedance algorithm will not see the 3Hz component. However, for faults further down the line, the ringdown can suppress the 60Hz component and cause the relay to calculate less impedance to the fault than actually exists. The result of this error is a potential for the relay to over-reach its designated zones.

To observe potential impacts caused by the ringdown effect and the distortion originated by the SRPFC, the CCVT model is applied to all three phases of the PSRC test system.

Relay Configuration and Testing

To analyze potential impacts that a saturable reactor might pose on distance protection, the recordings of the transient simulations were played back into a SEL-421 line distance relay using a DOBLE® relay test set (Figure 104). The relay test set was configured to playback secondary PT and CT waveforms at proper scale. The current signals were directly measured in the simulations and required scaling to secondary CT levels of 1-5A by the relay test set. Direct voltage measurements (230 kV nominal) were scaled to 115 V (66.4V line-to-neutral). Waveforms from the CCVT models required no scaling. The 3-phase voltages (115V nominal)

and currents (1-5A nominal) generated by the relay test set were applied to the transducer inputs of the SEL-421 relay. Relay settings for the phase-distance and ground-distance elements are listed in Table 8.



Magnified to observe ringdown

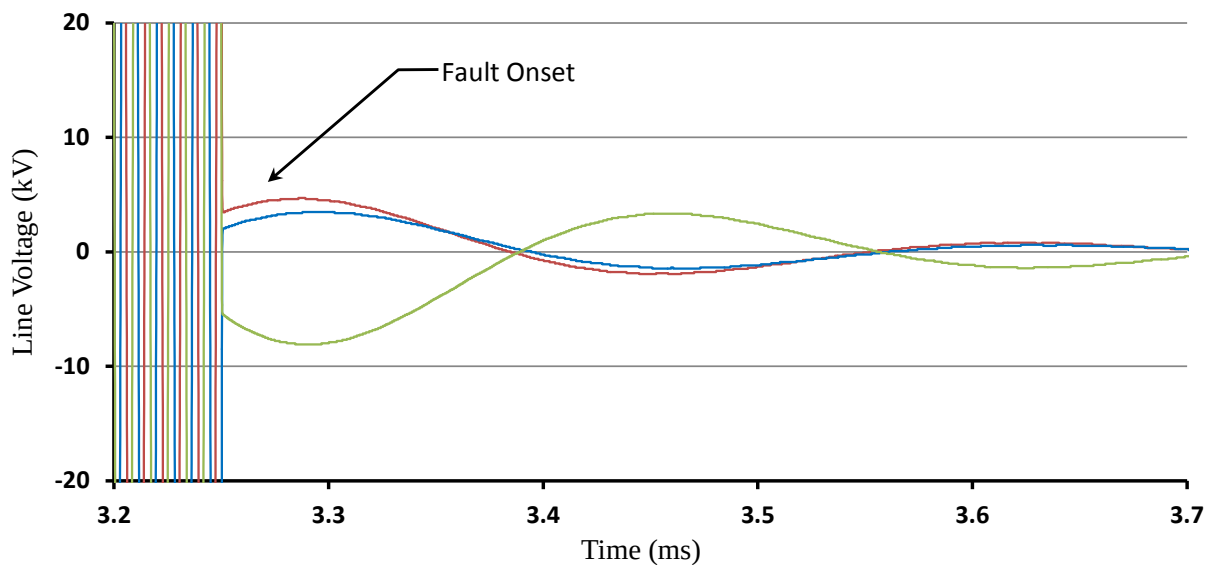
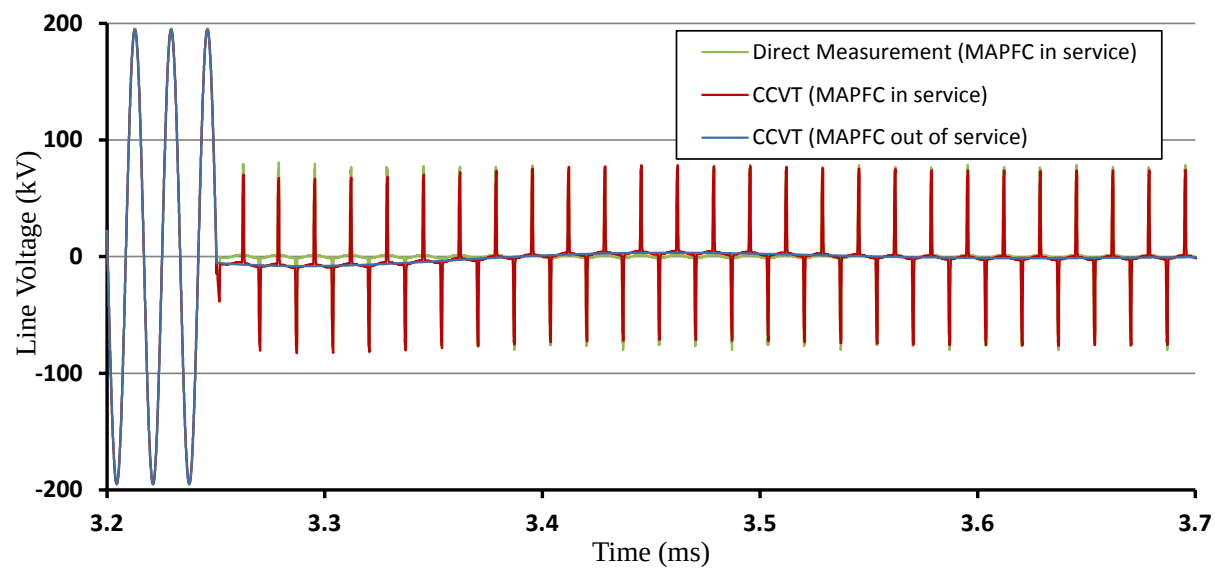


Figure 102 CCVT Line-Neutral Voltages during ABC Fault



Magnified to observe ringdown

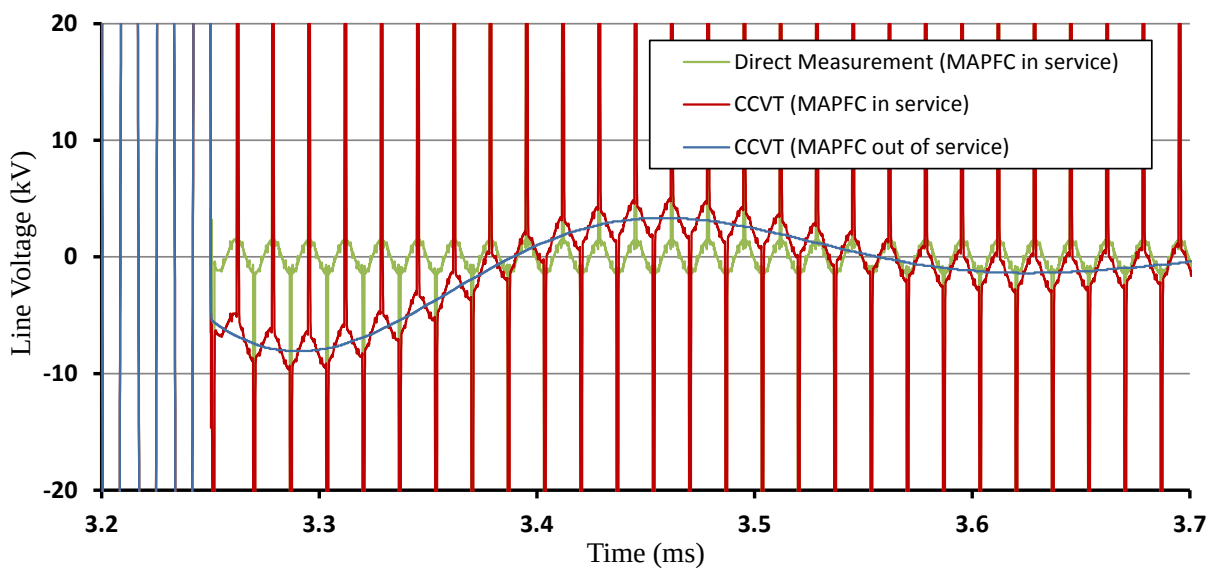


Figure 103 Phase C Line-Neutral Voltages during ABC Fault

The flow diagram in Figure 105 describes the testing procedure. The first step is to select the simulation parameters in PSCAD™, this includes selection of fault type and location. After the parameters are set, the simulation is executed and the resulting voltages and currents at the relay location are recorded in COMTRADE format. The next step is playback of the recorded currents and voltages into the relay transducer terminals. To accomplish this task, a COMTRADE file is uploaded into the DOBLE® and scaling factors are set as previously described. The SEL-421 relay, which is configured to the settings listed in Table 8, observes the 3-phase voltages and currents from the DOBLE® as they would an actual transmission line. The reaction of the relay is logged. For determination of zone reach, the fault location is incremented by 1% of the total line length until the relay fails to trip on the target zone.

Table 8 Distance Relay Settings

Setting	Value
Zone 1	90%
Zone 1 delay	Instantaneous
Zone 2	120%
Zone 2 delay	20 cycles (60 Hz)
Zone 3	N/A
Zone 3 delay	N/A
PT ratio	2000:1
CT ratio	400:1

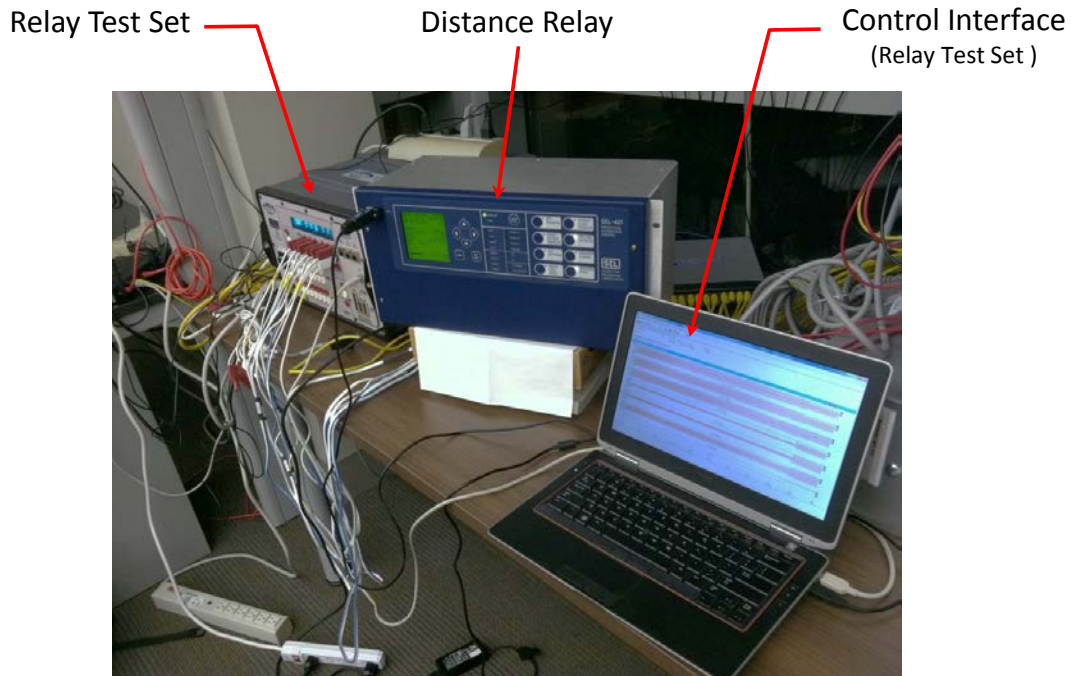


Figure 104 Relay Testing Set-up

Baseline Tests

Baseline tests were performed with the saturable reactor out of service using direct voltage measurements and CCVTs to compare the zone 1 and zone 2 reach limits to the relay settings. These tests will serve as benchmark for evaluating the protection when the saturable reactor is in service and for evaluating the impact of CCVTs.

The baseline test results for zone 1 are summarized in Table 9. The actual values represent the locations of simulated faults along the line length that resulted in zone 1 trip, but a 1% increase in the fault distance beyond this zone 1 trip point resulted in zone 2 trip. The estimated values are determined by the relays' fault location estimate, in percent of total line length, and indicate the fault location perceived by the relay. The baseline results show that zone 1 has approximately 3% under-reach in all but the A-g case. Since the SRPFC is out of service, this under-reach is inherent to the protection scheme. The application of CCVTs appears to have little or no impact the zone 1 reach points.

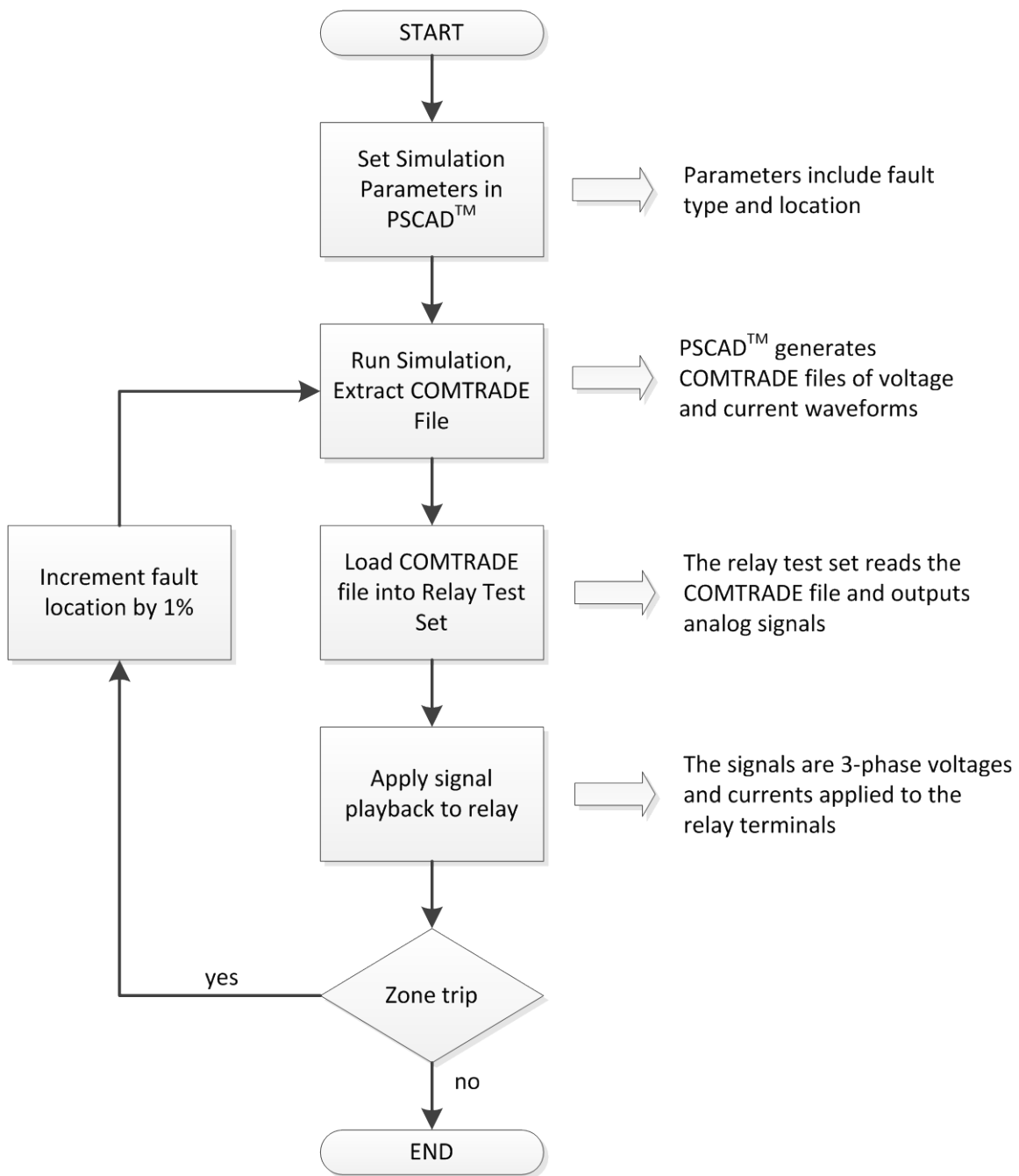


Figure 105 Relay Testing Flow Diagram

Table 9 Baseline Test Results for Zone 1

Fault	Zone 1			
	Actual (%)		Estimated (%)	
	Direct Measurement	CCVT	Direct Measurement	CCVT
ABC	87	88	90.28	90.74
A-g	90	90	90.36	89.81
AB	87	87	90.16	89.72
AB-g	87	87	90.15	89.67

Baseline test results for zone 2 are summarized in Table 10. They show, similarly to zone 1, an inherent tendency for the relay to under-reach. In all fault cases but A-g, the observed under-reach was approximately 5% of the relay setting. For A-g faults, the amount of under-reach was approximately 2%. The application of CCVTs appears to have a small impact, as it appears to increase the reach point by 1% in all four fault cases.

Table 10 Baseline Test Results for Zone 2

Fault	Zone 2			
	Actual (%)		Estimated (%)	
	Direct Measurement	CCVT	Direct Measurement	CCVT
ABC	115	116	118.94	119.19
A-g	118	119	118.93	119.57
AB	115	116	119.12	119.4
AB-g	115	116	119.12	119.57

The likely causes of the inherent under-reach include errors in the calculation of the relay settings and/or loss of precision between transient simulations, playback, and measurement by the relay. While the exact causes of the inherent under-reach are not quantified here, the baseline tests provide a reference point, or benchmark, from which to compare subsequent tests with SRPFC in service.

SRPFC in Service

The zone 1 reach tests for DC operating currents of 0A and 200A are summarized in Table 11. The far right column displays the amount of under-reach relative to the baseline. The line-side values are the observed zone 1 trip limits when the potential is measured on the line-side of the SRPFC; where the relay should experience minimal impact from the dynamic impedance. The bus-side measurements are located behind the SRPFC, so the relay should see the full impact of the dynamic impedance. Both DC operating states of the saturable reactor cause the relay to under-reach. Operating at zero DC current causes the highest under-reach with levels of 4% to 7% relative to the baseline. The 200A DC operating case improves the under-reach by a few percent in all fault cases but ABC-g. As expected, the line-side measurements are identical to the baseline.

Table 11 Zone 1 Reach Results with SRPFC in Service using Direct Measurements

Fault	I _{DC} = 0A				% Zone 1 change (BS Actual - LS Actual)
	Line-side (LS)		Bus-side (BS)		
	Actual (%)	Estimated (%)	Actual (%)	Estimated (%)	
ABC	87	90.25	83	90.11	-4
A-g	90	90.29	85	90.91	-5
AB	87	90.14	80	90.2	-7
AB-g	87	90.17	80	90.91	-7
Fault	I _{DC} = 200A				% Zone 1 change (BS Actual - LS Actual)
	Line-side (LS)		Bus-side (BS)		
	Actual (%)	Estimated (%)	Actual (%)	Estimated (%)	
ABC	87	90.27	83	90.11	-4
A-g	90	90.33	87	89.48	-3
AB	87	90.18	83	89.69	-4
AB-g	87	90.16	83	89.7	-4

Zone 2 reach tests are summarized in Table 12. They are much like the zone 1 results in terms of their behavior, but the level of under-reach is slightly larger. In the case of zero DC operating current, the level of under-reach is 7% to 9% of the baseline. At 200A DC operating current, the under-reach improves to 3% to 4% of the baseline. Again, the line-side measurements are identical to the baseline.

Faults applied on ZL1 at 0%, 33%, and 66% of the line length did not produce significant results as all instances resulted in instantaneous trip of zone 1.

Table 12 Zone 2 Reach Results with SRPFC in Service using Direct Measurements

Fault	I _{DC} = 0A				% Zone 2 change (BS Actual - LS Actual)
	Line-side (LS)		Bus-side (BS)		
	Actual (%)	Estimated (%)	Actual (%)	Estimated (%)	
ABC	115	118.94	107	119.07	-8
A-g	118	118.83	111	118.35	-7
AB	115	119.1	106	118.69	-9
AB-g	115	119.08	107	119.34	-8
Fault	I _{DC} = 200A				% Zone 2 change (BS Actual - LS Actual)
	Line-side (LS)		Bus-side (BS)		
	Actual (%)	Estimated (%)	Actual (%)	Estimated (%)	
ABC	115	118.94	112	119.47	-3
A-g	118	118.89	116	118.92	-2
AB	115	119.11	112	119.41	-3
AB-g	115	119.11	112	119.41	-3

SRPFC in Service with CCVTs

Test results for zone 1 reach with the application of CCVTs are shown in Table 13. The results are nearly identical to those from direct measurement. At the 200A operating point, the CCVT reduces the amount of under-reach by 1% in all but the AB-g case. The zone 2 results, shown in Table 14, have the same trend where the CCVTs may reduce the under-reach by approximately

1%. In Table 13 and Table 14, the 200A operating point results are compared to those of the 0A operating point as line-side measurements with CCVTs at the 200A operating point were not simulated. However, the trend with the direct measurements and those using CCVTs for 0A operating point show that line-side measurements match the baseline. Therefore, it is concluded that the CCVT measurements for the 200A operating point will be identical to those of the 0A operation point. Faults applied on ZL1 at 0%, 33%, and 66% of the line length did not produce significant results as all instances resulted in instantaneous trip of Zone 1.

Table 13 Zone 1 Reach Results with SRPFC in Service using CCVT Measurements

Fault	I _{DC} = 0A				% Zone 1 change (BS Actual - LS Actual)
	Line-side (LS)		Bus-side (BS)		
	Actual (%)	Estimated (%)	Actual (%)	Estimated (%)	
ABC	88	90.74	81	90.16	-7
A-g	90	89.81	85	90.54	-5
AB	87	89.72	80	89.72	-7
AB-g	87	89.67	80	89.33	-7
Fault	I _{DC} = 0A		I _{DC} = 200A		% Zone 1 change (BS Actual - LS Actual)
	Line-side (LS)		Bus-side (BS)		
	Actual (%)	Estimated (%)	Actual (%)	Estimated (%)	
ABC	88	90.74	84	90.55	-4
A-g	90	89.81	88	89.97	-2
AB	87	89.72	84	90.25	-3
AB-g	87	89.67	83	89.24	-4

Table 14 Zone 2 Reach Results with SRPFC in Service using CCVT Measurements

Fault	I _{DC} = 0A				% Zone 2 change (BS Actual - LS Actual)
	Line-side (LS)		Bus-side (BS)		
	Actual (%)	Estimated (%)	Actual (%)	Estimated (%)	
ABC	116	119.19	109	119.39	-7
A-g	119	119.57	112	118.77	-7
AB	116	119.4	107	119.18	-9
AB-g	116	119.57	107	118.79	-9
Fault	I _{DC} = 0A		I _{DC} = 200A		% Zone 2 change (BS Actual - LS Actual)
	Line-side (LS)		Bus-side (BS)		
	Actual (%)	Estimated (%)	Actual (%)	Estimated (%)	
ABC	115	118.94	113	119.7	-2
A-g	118	118.89	117	118.86	-1
AB	115	119.11	113	119.41	-2
AB-g	115	119.11	113	119.85	-2

Summary and Recommendations

An impact study was conducted to determine the impacts of the SRPFC on line distance protection. The study included the following activities:

- Application of the FEA-derived device model, developed in Chapter III, to the PSRC system model for testing line distance protection. Simulations of four fault types at locations in the system were performed using PSCAD™.
- Playback of the simulation waveforms through a relay test set to a SEL-421 line distance relay.
- Determinations of zone reach by observing relay response to fault location.
- The testing of zone reach using direct voltage measurements and CCVTs.

The SRPFC causes the relay to under-reach zones 1 & 2 when the potentials are measured on the bus-side on the device. The amount of under-reach is approximately 2% to 9% of the observed baseline reach depending on fault type and the amount DC bias current applied to the SRPFC.

The DC operating state had an observable, yet marginal, impact on relay reach. The 0A operating resulted in the largest over-reach, which is consistent with the higher impedance of the SRPFC when operating in the linear region. The 200A operating point reduced the amount of under-reach by a few percent, as the saturation impedance still increases the overall impedance of the line relative to SRPFC out of service. It is observed in Figure 55 that increasing the DC current reduces the incremental inductance over the defined instantaneous AC current range. As fault current approaches the zero crossing, it causes the SRPFC to assume the incremental inductance set by the DC operating point within the nominal range. This change of impedance over a relatively small portion of the AC current cycle is enough to change the impedance of the line as observed by the relay.

CCVTs made little difference in the outcome. The baseline tests revealed that the CCVT ringdown did not affect the reach. Further tests with SRPFC in service revealed a trend of the CCVTs to reduce the under-reach by 1% in most cases.

Unlike the use of capacitors in series-compensated lines that create a set of very challenging problems for distance protection, the impact of the SRPFC is limited and manifested in a small decrease in sensitivity of the scheme. The distance relay under-reach demonstrated in this study can easily be accommodated when setting the protective relays and pose no significant problem to protection. Potential measurements taken at the line-side mitigate the under-reach problem for both zones, but there could be an impact on reverse zone 3 or zone 2 from an adjacent relay located behind the SRPFC. In addition, having the highest under-reach at zero DC current supports the assumption that the transient reflection on the DC circuit can be neglected since the SRPFC will have protection that shorts the DC winding under transient conditions. Thus, the DC current will be forced to zero under transient conditions.

CHAPTER VI: OVERALL CONCLUSIONS AND FUTURE WORK

Conclusions and Contributions

This dissertation describes the development of models that represent the behavior of an SRPFC for simulation of power system transients. The SRPFC is essentially a variable reactor connected in series with a transmission line for the purpose of power flow control. Under transient conditions, like faults, the high current in the AC winding of the device results in high flux that forces the device into its saturation mode, or mode of least reactance. The effects of the dynamic reactance are quantified through the development of device models and simulation in an industry-accepted power system model. The key contributions of this dissertation are summarized as follows:

- Three methods are proposed to model the SRPFC for the analysis of power system transients in electric power systems. These models make it possible to assess the feasibility of SRPFC and to assess impacts on protection, fault currents, and other transient-related events. The modeling methods are unique in terms of their application for power flow control in power system equipment.
- Evaluation of the models by intercomparison of simulations in benchmark test systems is performed to gain confidence and gauge performance of the models. The models are validated using data available from hardware tests.
- An initial impact study using a SRPFC model is completed to observe potential effects on line distance protection and to quantify the extent its impact. A line distance protection scheme using a commercial relay is evaluated. The significance of the study is its uniqueness, as there are no known impact studies of this kind for an SRPFC. Reliability is of foremost concern to transmission utilities which necessitate an understanding of the impacts to protection schemes when considering the deployment of SRPFC devices. Therefore, this study will serve as an initial point for further studies on impacts to line distance protection.

The models developed during the course of this dissertation attempt to achieve a common objective, to model the SRPFC from perspective of an AC power system, but utilize different methods in the achievement of this objective. Two-core, FEA-derived, and gyrator-capacitor methods are attempted. The two-core model represents the AC and DC flux interactions using a single ideal core over a given half cycle. Application of the second core provides full-cycle representation of the device. In order to represent the three-legged core device as a two-core device, equivalent flux linkage curves derived from FEA and an effective DC current derived using a gapless core were applied to the model. While application of two-core model into commercial power system transients simulators is relatively simple, the accuracy of the model is not sufficient when compared to hardware test results.

The FEA-derived model is not constrained by core configuration as FEA breaks down any geometry into small pieces for analysis. Since FEA is hampered by relatively slow simulation speeds, the FEA-derived model uses a data table consisting of flux linkage, DC current, and instantaneous AC current recorded from a several FEA simulations. The resulting two-dimensional data table is interpolated at each time step in order to represent the SRPFC. The FEA-derived model is relatively simple to implement in most commercial software tools and provides sufficient accuracy and resolution for analysis in AC power systems. When compared with co-simulation (transient simulations coupled to the FEA model) the single-phase results showed strong agreement under nominal and line-ground fault conditions. The apparent reactance curve also showed strong agreement when compared to hardware test results. However, the method does not represent internal behavior of the SRPFC, e.g. flux quantities and back emf.

The gyrator-capacitor model provides high time resolution and represents the internal behavior of the SRPFC. The capacitor adequately models flux paths in the core and nonlinear capacitors amply model the nonlinear core characteristic. Gyrators provide an interface to couple the electric circuit representing the magnetics to the electric circuit representing the power system, making it possible to simulate the magnetic and electric circuit models simultaneously. Simulations of the single-phase model under nominal and line-ground fault conditions achieved

results that matched closely with the FEA-derived and co-simulation results when modifications to the model are included to account for stray flux paths and manufacturing tolerances. The apparent reactance curve correlated well with hardware test data. The method solves relatively quickly with high time resolution. A significant advantage of the method is its ability to represent the flux values in each leg of the core and the back emf on the DC winding. The disadvantage of this method is the difficulty in applying it to commercial power system transient simulation tools.

A power system transients study was performed using the FEA-derived model to assess affects that the SRPFC may have on line distance protection. To accommodate the study, the model was scaled to transmission voltage and current levels. Three individual models were used to represent a three-phase system. The model was applied to the PSRC line distance protection test system and faults were applied along segments of the line. A commercial relay configured to protect the line was tested using playback of the simulations through a relay test set. The results demonstrated minor under-reach of zone 1 and zone 2 with the SRPFC in service. The amount of under-reach ranged from 2% to 9% depending on the location and type of fault. CCVTs are known to impact measurements as they form an RLC circuit with the power system. To observe the combined effects of the SRPFC and CCVTs, an industry-accepted CCVT model was included in the transient analysis. The results demonstrated a tendency for the CCVTs alone to cause the protection scheme to over-reach. However, the amount of over-reach was relatively small as the combined effect of the SRPFC and the CCVTs resulted in a net under-reach of the protection system, though the amount of under-reach was only slightly higher without the CCVTs. Since the impact of the SRPFC is minor, protection engineers may choose to neglect the under-reach effects upon designing protection schemes. However, engineers may also choose to compensate for the under-reach when configuring zone 1 and zone2. Since the undesirable miss operation in line distance protection is zone 1 over-reach, engineers will likely neglect the effects for zone 1 and include compensation for zone 2 only.

Future Work

Potential work to follow the efforts presented in this dissertation include the following topics:

Expansion of Protection Study

The impact study on line distance protection utilized a configuration of PSRC test system designed to evaluate protection on two series-connected transmission lines. This configuration provided the most straightforward system in terms of establishing relay settings and understanding behavior of the system and the relay under fault conditions. Additional studies will observe the effects of the SRPFC on more complex configurations of the PSRC test system. Specific configurations to consider are parallel transmission paths and tapped transmission lines. A parallel path will change the voltage distribution along the adjacent line if a fault occurs on that line [63]. Tapped lines change the line impedance characteristic when they are placed in and out of service [63]. System protection has to perform amply in both situations, even with the added effects of the SRPFC that create additional complexity. The impacts of SRPFC on protection in both conditions would provide engineers with knowledge they can apply to improve their protection schemes when SRPFCs are to be operated in the system.

Transient Recovery Voltage

According to the *IEEE Guide for the Application of Transient Voltage Recovery Voltage for AC High-Voltage Circuit Breakers, IEEE Standard C37.011-2011*, transient recovery voltage (TRV) is the voltage that appears across the terminals of a pole of a circuit breaker after interruption [25]. The nature of TRV depends on the circuit being interrupted, whether it is resistive, capacitive, or inductive. TRV is a transient condition that can lead to restrike within a circuit breaker. This condition is potentially damaging to circuit breakers as sustained arcs will eat way at the contacts and cause protection to remove greater portions of the network from service than intended. Since TRV is dependent on circuit type, the inductance introduced to the line by the SRPFC may influence TRV. In addition, the changing nature of the inductance characteristic under fault conditions could also affect TRV.

Internal Effects under Fault Conditions

While the focus of this dissertation was to model SRPFC behavior from the perspective of the AC power system, the internal voltage is an important design consideration. Under fault conditions, the large AC current could induce high voltage on the DC control circuit. The induced voltage can disrupt the control set point of the SRPFC, or if high enough, damage the DC controller. FEA will likely be the most comprehensive method for studying these effects, but the gyrator-capacitor approach has shown promise that it can produce results. To model the actual effects of the gyrator-capacitor model, the constant current source in the DC control circuit will need to be replaced by a voltage source and equivalent impedance. The DC circuit impedance will likely impact the level of back emf and will represent the actual implementation. However, the use of a voltage source and equivalent impedance may cause convergence issues as it provides another unknown parameter in the model.

Application of Model Methods to Other Designs

The SRPFC prototype highlighted in this dissertation is an initial prototype developed by ORNL with collaboration by project partners. While the prototype demonstrated technical feasibility, it did present challenges in performance and controllability that eventually prompted alterations of the initial design. The modeling and analysis effort in this dissertation could be extended to new designs to understand system impacts and provide feedback to designers. As the models and the design modifications converge, potential users of the technology will be equipped with tools to evaluate feasibility and potential impacts within their respective systems.

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APPENDICES

APPENDIX A: MAGNETIZATION CURVE FOR AKH1 STEEL

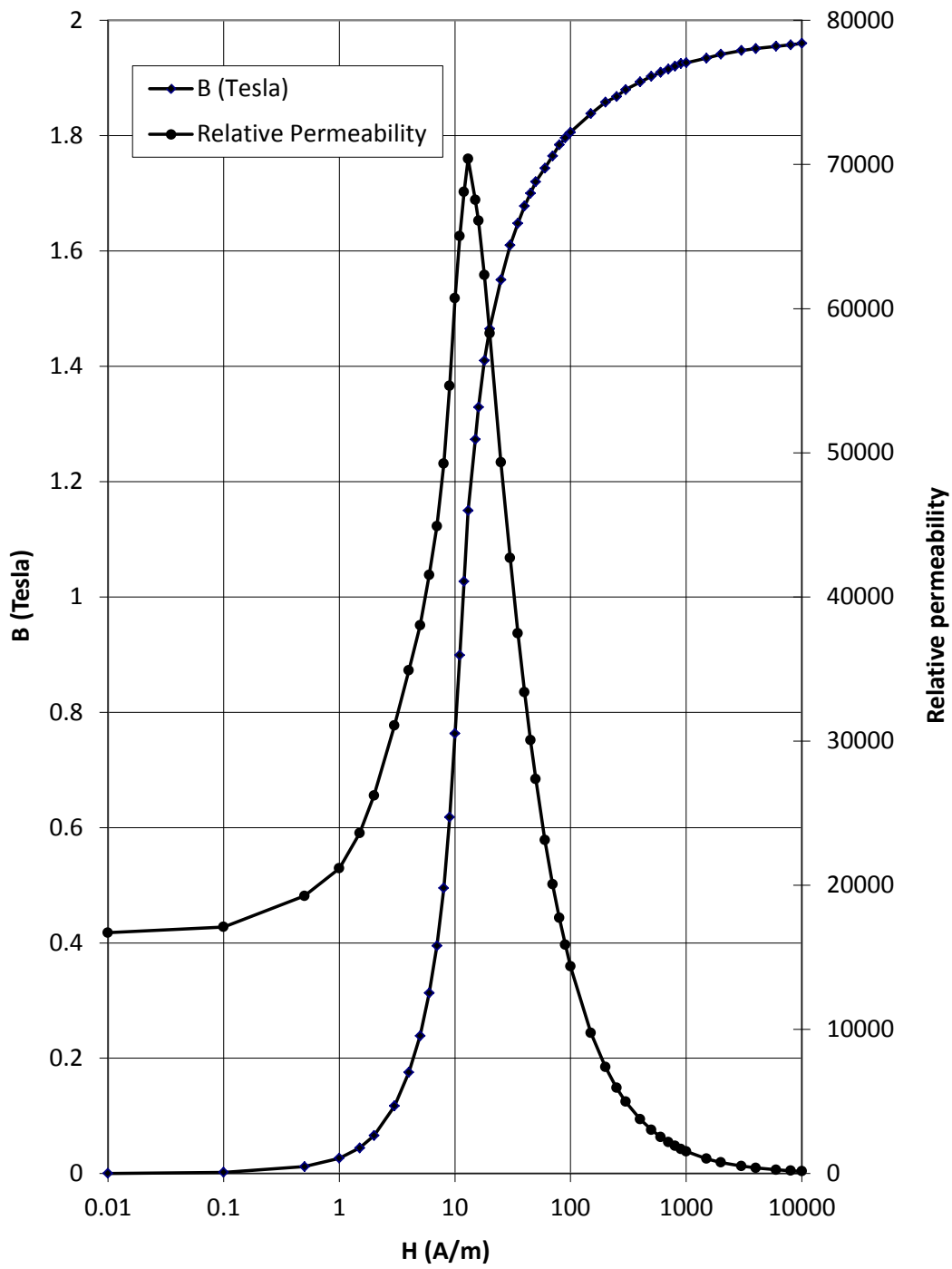


Figure 106 DC Magnetization Curve for AKH1 Steel

APPENDIX B: DIFFERENTIAL AND APPARENT INDUCTANCE

The use of inductance as a constant value is common in the analysis of electric and magnetic circuits. This statement holds true for analysis of both AC circuits and DC transient circuits, where currents change with respect to time. The assumption of constant inductance is a valid one if the magnetic behavior of a device is linear. Device properties and operational constraints are elements that can confine inductance to a single value, thus they are the basis for such assumptions. For example, air-core inductors have constant inductance proportional to the permeability of free space. Another example is operation of a transformer within its linear region. However, there are situations where constant inductance is not adequate. Transient analysis of electric machines and transformers can serve as example, as inrush currents often push these devices into a non-linear mode of operation. In these situations, analysis requires a variable inductance.

This section reviews the concept of differential and apparent inductance. The review defines these inductances and provides an example pertaining to the two-core model discussed in CHAPTER II. Discussions of these terms in literature refer to the B-H curve and the λ -i curve, indicating that differential inductance is proportional to incremental permeability [95] [96]. In similar fashion, apparent inductance is proportional to differential permeability. This review will focus the λ -i curve, since flux linkage is the basis for modeling the SRPFC within the context of this work.

Historical Perspective

During the industrial revolution, scientists were able to define many of the engineering units that are in use today. This includes units that are now common in electricity and magnetism. Transactions published by the British Association for the Advancement of Science⁴ in 1895 address definitions of inductance [95]. The discussions focus on inconsistencies between three principles that define inductance (energy, emf, and total induction). In turn, the authors propose corrections and alternatives to resolve these inconsistencies. One alternative was the

⁴ The British Science Association (formally known as the British Association for the Advancement of Science) was formed in 1831 [98].

recognition of two types of inductance. The first is described as the slope of a secant line starting at the origin to any point along the flux linkage curve (A.1). This inductance was defined as the “total inductance”. The second is described as the slope of a line tangent to any point on the flux linkage curve (A.2). This inductance was defined as the “differential inductance”. The differential inductance is proportional to the slope of a line tangent to any point on the B-H curve. In this case, the slope becomes the differential permeability. Both inductances are expressed in units of henries.

Recent literature supports both definitions of inductance [97] [96] [47]. However, the terminology is not always consistent. Terms for apparent inductance include large-signal inductance, secant inductance, and total inductance [98]. Terms for differential inductance include small-signal and incremental inductance. Incremental and differential permeability are also used interchangeably. However, there is a distinction between the two terms as incremental permeability has a special meaning in magnetics [30] [99]. This distinction is noted by the following definitions:

Incremental Permeability is the ratio of a cyclic change in the magnetic induction to the corresponding cyclic change in magnetizing force when the mean induction differs from zero [30].

Small B-H “incremental” loops are obtained when small perturbations in H are applied around a fixed operating point. The slope of these small loops is the incremental permeability [99].

While difficult to visualize, these definitions do not describe the slope of a line tangent to any point on the B-H curve. However, a point is made that incremental and differential permeability is the same when hysteresis is nonexistent [99]. This is likely the assumption made when the terms are used interchangeably [96] [47].

For consistency, this work refers to the slope of the secant line as the apparent inductance and the slope of the tangent line as the differential inductance. The relationships of apparent and differential inductance to the flux linkage curve are shown in Figure 107.

$$L_a = \frac{\lambda}{i} \quad (\text{A.1})$$

$$L_d = \frac{d\lambda}{di} \quad (\text{A.2})$$

Where: L_d = differential inductance {H}

L_a = apparent inductance {H}

λ = flux linkage {Wb·turns}

i = current {A}

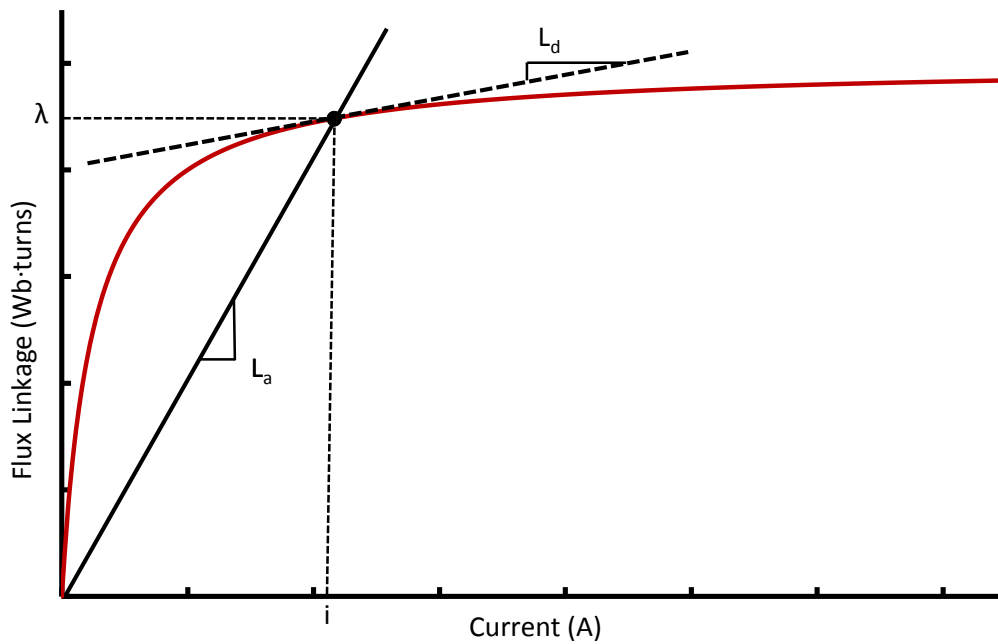


Figure 107 Flux Linkage Curve with Secant and Tangent Lines

Derivation

The relationship between L_a and L_d can be expressed explicitly through evaluation of inductor voltage (A.3), where both inductances are functions of the applied current [47]. Expansion of

(A.2) by applying the chain rule leads to an expression that relates the rate of change of current with respect to time multiplied by some scalar value at a given current (A.4). This scalar is the differential inductance expressed in terms of the apparent inductance (A.5). This result shows that L_d is dependent on the rate of change of L_a with respect to current. Absent magnetic saturation, the L_d equals L [96].

$$V = \frac{d\lambda}{dt} = \frac{d(L_a \cdot i)}{dt} \quad (\text{A.3})$$

$$\frac{\partial \lambda}{\partial t} = \left[L_a + i \cdot \frac{dL_a}{di} \right] \cdot \frac{di}{dt} \quad (\text{A.4})$$

$$L_d = L_a + i \cdot \frac{dL_a}{di} \quad (\text{A.5})$$

Relation to the λ -i Curve

The relationship of L_a and L_d to the λ -i curve is shown in Figure 107. This λ -i curve was calculated using the Frolich equation, and is presented here only for purpose of example [100]. The apparent inductance is straightforward as it is the slope of a line from the origin to any point (i, λ) on the curve. Two secant lines are shown in Figure 108. Their slopes are defined by discrete equations (A.6) and (A.7).

$$L_{a_k} = \frac{\lambda_k}{i_k} \quad (\text{A.6})$$

$$L_{a_{k-1}} = \frac{\lambda_{k-1}}{i_{k-1}} \quad (\text{A.7})$$

The apparent inductance is constant for small values of current as the curve is linear. However, as current increases and the λ -i curve approaches saturation, the apparent inductance decreases. As increasing current moves the flux linkage through the knee-point of the curve, changes in L_a become smaller. L_d behaves similarly to L_a , but in fashion that is more dramatic

as it falls to the saturation value at lower currents. The behavior of L_a and L_d for the example λ - i curve are shown in Figure 109.

The differential inductance is straightforward when articulating it in terms of the slope of a tangent line located at any point (i, λ) . Numerically, differential inductances can be calculated using (A.8) [57] as long as the steps are sufficiently small. The parameters in (A.8) are labeled in Figure 108.

$$L_{d_k} = \frac{\lambda_k - \lambda_{k-1}}{i_k - i_{k-1}} \quad (\text{A.8})$$

Another way to define the differential inductance is in terms of apparent inductance (A.5). This definition makes it possible to calculate L_d at a point on the flux linkage curve using L_a at some point (A.6) and L_a from the previous point (A.7). Figure 108 describes the parameters. To represent a line tangent to a point, the distance between current steps should be small. Note that the distance between points in Figure 108 is large to demonstrate the concept.

Equation (A.9) expresses (A.5) as a discrete equation and states that L_d at point k is the L_a from the previous point ($k-1$) plus the change in L_a from $k-1$ to k .

$$L_{d_k} = L_{a_{k-1}} + i_k \cdot \frac{L_{a_k} - L_{a_{k-1}}}{i_k - i_{k-1}} \quad (\text{A.9})$$

Substitution of the (A.6) and (A.7) followed by algebraic expansion reduces (A.9) to (A.10). This result matches the previous definition of differential inductance (A.8).

$$L_{d_k} = \frac{\lambda_{k-1}}{i_{k-1}} + i_k \cdot \frac{\left(\frac{\lambda_k}{i_k} - \frac{\lambda_{k-1}}{i_{k-1}}\right)}{i_k - i_{k-1}} = \frac{\lambda_k - \lambda_{k-1}}{i_k - i_{k-1}} \quad (\text{A.10})$$

Equation (A.9) can also be solved for apparent inductance. The proof requires substitution of (A.7) and (A.8) as shown in (A.11). Solving (A.11) for L_{a_k} matches the definition of apparent inductance given in (A.6).

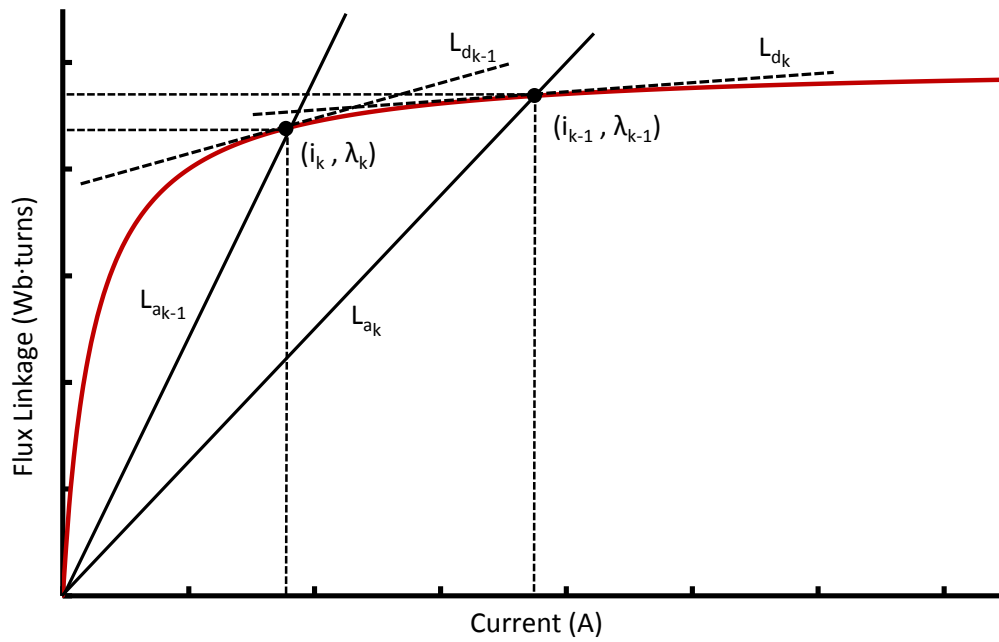


Figure 108 Apparent and Differential Inductance at Two Points on Flux Linkage Curve

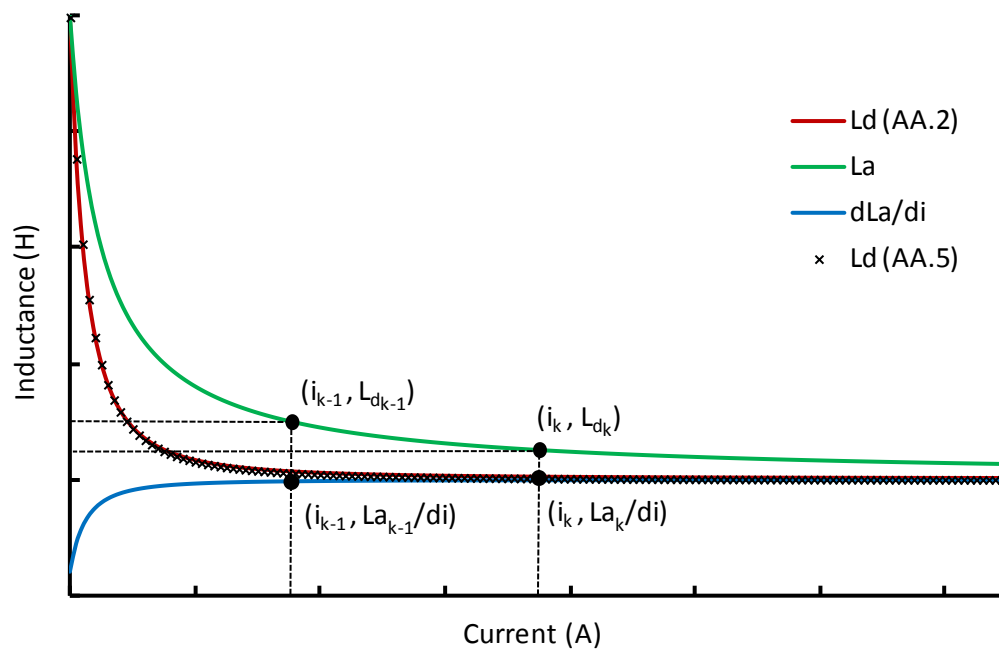


Figure 109 Apparent Inductance, Differential Inductance, and Derivative of Apparent Inductance

$$\frac{\lambda_k - \lambda_{k-1}}{i_k - i_{k-1}} = \frac{\lambda_{k-1}}{i_{k-1}} + i_k \cdot \frac{\left(L_{a_k} - \frac{\lambda_{k-1}}{i_{k-1}}\right)}{i_k - i_{k-1}} \quad (\text{A.11})$$

It is also possible to estimate L_a from L_d using continuous equations by multiplying both sides of (A.2) by di and applying integration as in (A.12). Application of the integral provides the flux linkage in terms of current and differential inductance (A.13). Expansion of the flux linkage terms using the relationship in (A.1) yields the apparent inductance in (A.14).

$$\int d\lambda = \int L_d \cdot di \quad (\text{A.12})$$

$$\lambda = \int L_d \cdot di \quad (\text{A.13})$$

$$L_a = \frac{1}{i} \int_0^i L_d \cdot di \quad (\text{A.14})$$

Example using SRPFC Model

Applying the definitions of apparent and differential inductance to SRPFC model data validates the relationships between both types of inductance. FEA of the three-legged core was used to generate the model data. The data set contains flux linkage values calculated at the AC windings for applied currents. In this case, the current was ramped linearly with respect to time from zero to a maximum value. Inductances were calculated by applying (A.1), (A.2), (A.5), and (A.14) to the flux linkage data set using MATLAB®. The results are shown in Figure 110.

There are key observations from Figure 110. First, both inductances are nonlinear with respect to the applied current. This nonlinearity is basis for operation of the SRPFC and the reason for the use differential inductance in the transient models. Second, changes in differential inductance are larger than the changes in apparent inductance. This behavior is as expected

and follows the definitions given in this APPENDIX. As the slope of a line tangent to the λ - i curve, the differential inductance will change more aggressively as the tangent line moves from point to point in accordance with the applied current. In contrast, the slope of the secant line experiences changes that are less drastic since the line is pinned to the origin. Thus, the apparent inductance appears as an “averaged” version of the differential inductance. Lastly, all definitions for apparent inductance yield the same result. The same holds true for the differential inductance.

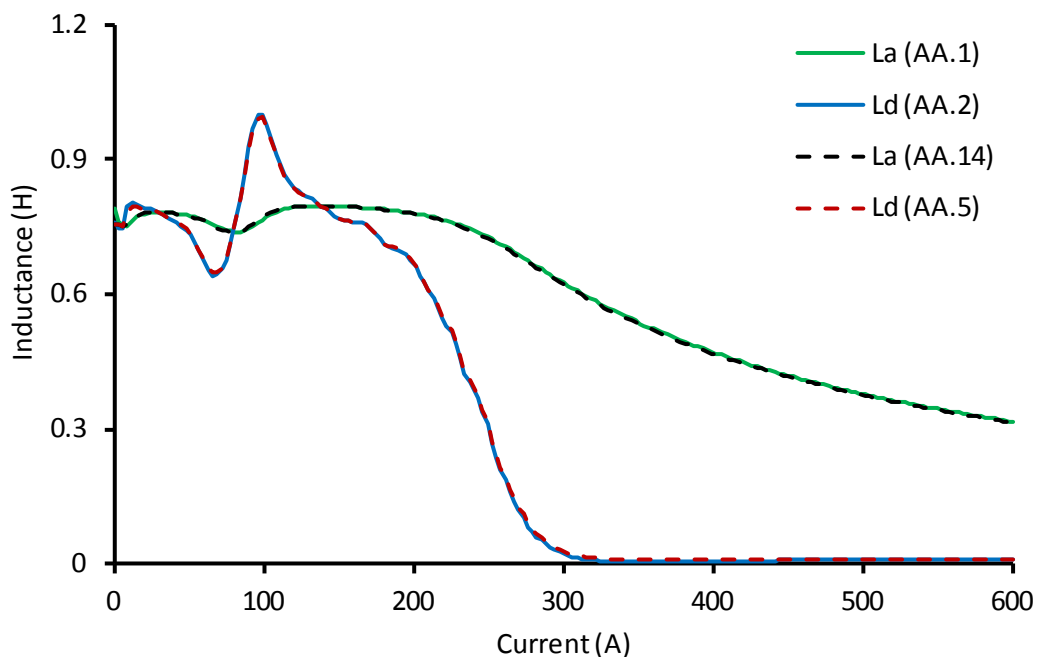


Figure 110 Apparent and Differential Inductance Applied to SRPFC Model Data

Summary

An assumption of constant inductance is not always an adequate one for analysis of electric circuit transients. Devices that exhibit nonlinear magnetic behavior can change inductance dynamically with respect to current. Such is the case for the SRPFC, which has a magnetic behavior that is highly nonlinear.

Differential inductance is a quantity defined as the slope of a line tangent at any point on the λ - i curve. It is small-signal value, meaning that differential inductance is valid for relatively small changes in current. Thus, differential inductance is suited for use in transient analysis where small time increments generally lead to small changes in current. Apparent inductance is a large-scale quantity common to AC circuit analysis that uses RMS values. Since it is the slope of a secant line from the origin to any point on the λ - i curve, the apparent inductance is similar to an average value over a range of current values. For linear magnetic behavior near the origin of the B-H curve, the differential and apparent inductances are equal.

Transient simulations of nonlinear magnetic devices require differential inductance for adequate representation of the inductance at a given current value, assuming that current is varying with respect to time. The differential inductance provides the resolution required to adequately model extrema. This allows for more realistic analysis of peak voltages and other parameters that are a concern under transient conditions. In these cases, apparent inductance will produce marginalized results that could lead to underestimation of peak values.

APPENDIX C: EQUIVALENT FLUX LINKAGE CURVES

The FEA model of the three-legged core may be insufficient for deriving equivalent parameters because all three legs are physically connected and share common flux paths. The proposed approach splits the three-leg core vertically through the center of the AC leg into two separate cores. Two versions of this approach are used to derive equivalent parameters. The first version maintains use of both cores, but provides flux linkage curves two times higher than actual. The second version remedies this issue by using only one of the cores. Analysis of these models by FEA entails calculation of the AC flux linkage for one of the AC windings while applying a current ramp to the AC terminal. Since only the AC flux linkage contributes to the reactance, the DC windings are open circuited for the FEA. The applied current ramp is linear with respect to time and positive for all values. The FEA model contains the multiple reluctance paths and air-gaps, so the resulting flux linkage curve contains their net effects. The development of these models and the equivalent parameters generated are described in this section.

The FEA split-core model in Figure 111 is the three-legged core model split vertically down the middle of the AC leg. Both cores are identical in terms of material properties and number of turns for the AC and DC windings. However, a transformation is required to maintain reluctance equivalence between the three-legged and split cores that result in non-uniform cross-sectional area. The air gap of the AC leg is retained in the transformation. The non-uniformity and presence of an air gap make the core non-homogenous. To the degree that the split-core model represents the three-legged core model, it will provide equivalent parameters that impart the nonhomogeneous nature of the three-legged core to the two-core model.

The transformation of the three-legged core to the split-core model is illustrated in Figure 112 and expressed in (C.1), (C.2), and (C.3). When no DC bias is applied, flux in the middle leg will split equally between the two outer legs. Thus, the outer legs are parallel paths with equivalent reluctance equal to $\mathcal{R}'_a/2$. In the split-core model, the parallel path no longer exists and the equivalent reluctance is twice as high, $\mathcal{R}'_a$. Thus, the two models are not equivalent. A method to correct this issue is to reduce the cross-sectional area of the AC leg by a factor of two. This effectively increases the reluctance of the AC leg by a factor of two, $\mathcal{R}'_b$, and compensates for

the increase in the reluctance from loss of an outer leg. Application of mmf to the AC limb of both cores in Figure 112 yields expressions (C.1) and (C.2). These expressions lead to identical reluctances that validate the transformation. The reduction of area in the split core results in a slight change in the equivalent length, denoted by l'_a . However, the change is significantly small and is neglected as expressed in (C.3).

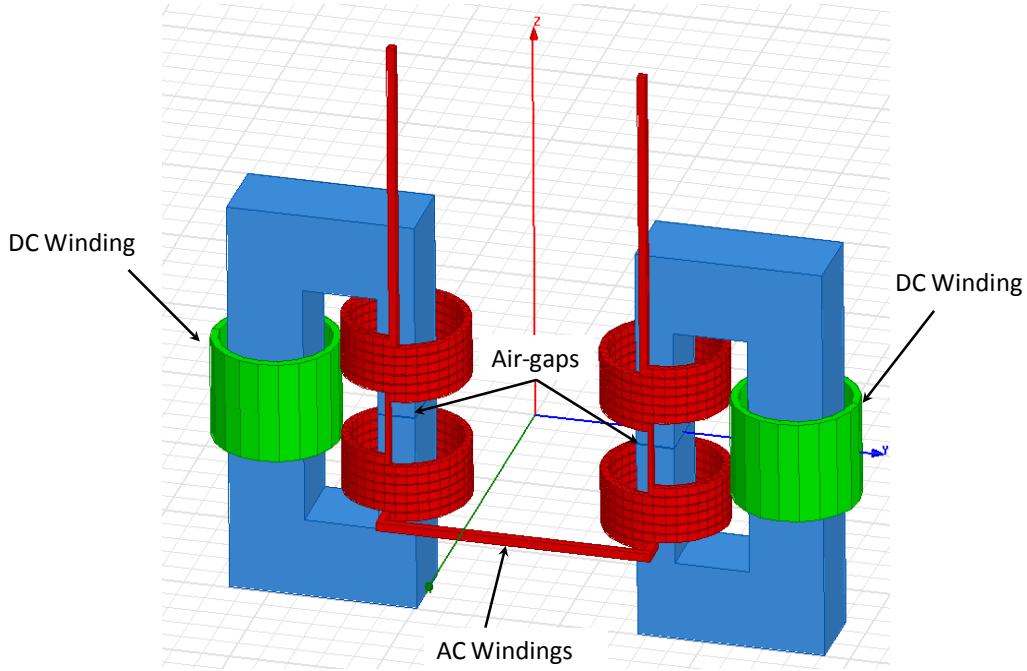


Figure 111 Split-Core Model for FEA

$$NI = \phi_1(\mathcal{R}_a + \mathcal{R}'_b) = \phi_1 \left(\frac{l_a}{\mu A} + \frac{l_b}{\frac{1}{2}\mu A} \right) = \frac{\phi_1}{\mu A} \left(\frac{l_a}{2} + l_b \right) \quad (\text{C.1})$$

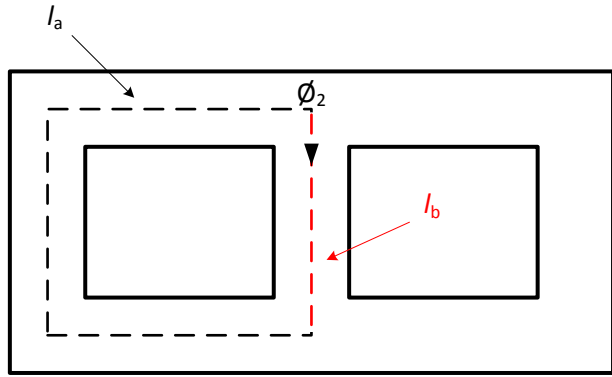
$$\mathcal{F} = NI = \phi_2 \left(\frac{\mathcal{R}'_a}{2} + \mathcal{R}_b \right) = \phi_2 \left(\frac{l'_a}{2\mu A} + \frac{l_b}{\mu A} \right) = \frac{\phi_2}{\mu A} \left(\frac{l_a}{2} + l_b \right) \quad (\text{C.2})$$

$$l'_a \approx l_a \quad (\text{C.3})$$

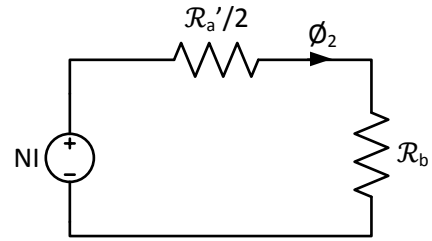
Where: $\mathcal{F}$ = magnetomotive force {A·turns}

$\mathcal{R}$ = reluctance {H⁻¹}

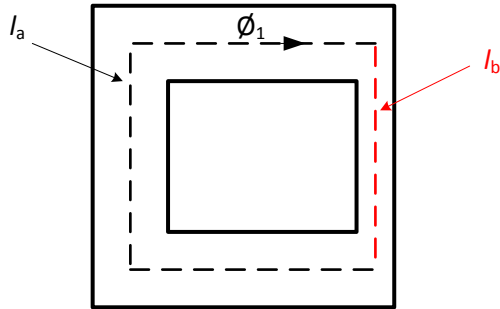
ϕ_1 and ϕ_2 represent the flux in the split-core and three-legged core models, $\mathcal{R}_a$ and $\mathcal{R}_b$ are the reluctances of the split-core model, $\mathcal{R}'_a$ and $\mathcal{R}'_b$ are the reluctances of the three-legged core model, l_a and l_b are the mean lengths, and l'_a is a corrected length for the split-core model.



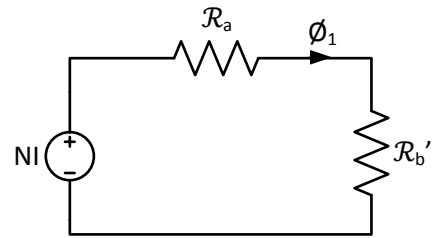
Three-leg Core (Area_b = Area_a)



Magnetic Circuit



Split Core (Area_b = Area_a/2)



Magnetic Circuit

Figure 112 Transformation of Three-Leg Core to Split-Core

While division of the three-legged core accounts for the flow of AC flux, application of the AC windings to each core results in twice the number of turns than the three-leg core as observed from the AC terminals. The additional AC turns result in flux linkage values that are double the expected quantity. This issue is addressed by removing one of the cores and using a single-core

model, illustrated in Figure 113, to derive the equivalent parameters. This has the same effect as dividing the flux linkage by a factor of two. Reducing the size of the model has an added advantage of improving the time required for FEA to achieve a solution.

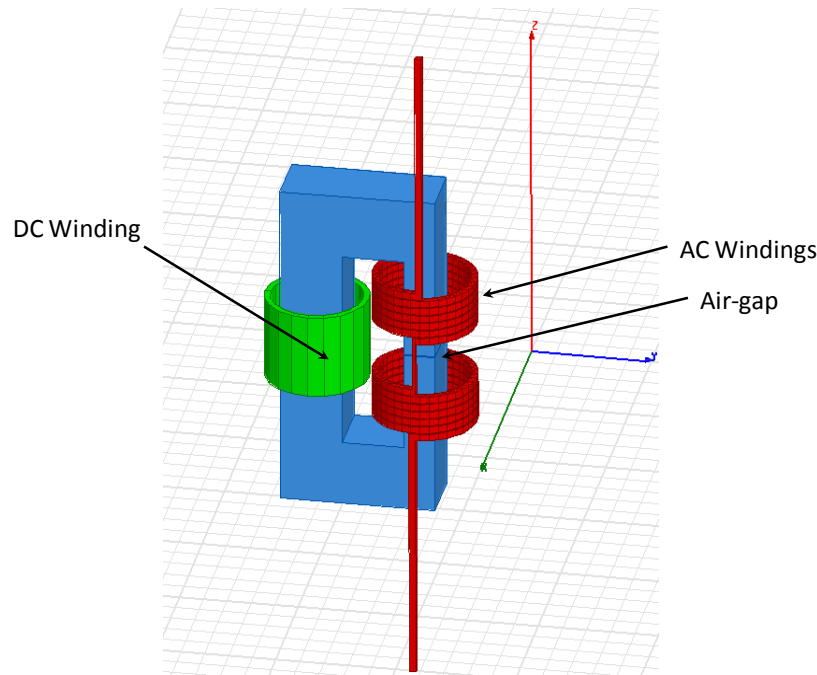


Figure 113 Single-Core Model for FEA

The flux linkage curves in Figure 114 were generated by FEA of the split-core model. One flux linkage curve is the result of simulating the split-core model without an air gap, which has high permeability, or slope, in the linear region. The second curve includes the air gap. The impact of the air gap is significant as it reduces the permeability for the linear region of the curve, which is the indented purpose of including an air gap in the design. The absence of the air gap provides a high reactance linear region that is desirable in saturable-core FCLs.

Upon analysis of the split-core model, the resulting apparent reactance values were twice the magnitude of hardware test data for zero DC current. After some reflection on the construction of the split-core model, it was apparent that the flux linkages were double the anticipated values. The increase in flux linkage values is due to the construction of the split core model and

application of the AC windings to each core. The number of AC turns applied to each core equals the number of turns on the middle leg of the three-legged core. This is necessary to represent the mmf; however, the total number of windings as seen by the AC current is doubled. Calculation of the flux linkage in FEA takes into account the correct mmf, but includes both AC windings. The flux linkage for the single-core model, which has a single AC winding, is expressed in (C.4). The flux linkage for the split-core model, which has two AC windings, results in a flux linkage double that of the single-core as expressed in (C.5).

$$\lambda_{single-core} = n_{ac}\phi \quad (C.4)$$

$$\lambda_{split-core} = 2n_{ac}\phi \quad (C.5)$$

FEA of a single-core model, constructed by removing one of the cores from the split-core model, was performed to validate the hypothesis. The curves in Figure 115 are results of the FEA that compare the split-core, single-core, and three-legged models. The result validates the hypothesis that the second set of AC windings is the cause of the issue as the single-core curve two times smaller than the split-core curve in terms of magnitude. The result also reveals that the three-legged core and single-core are equivalent when the DC terminals are open-circuited. However, the result does not account for interactions between AC and DC flux components, which are managed by the framework of the two-core model.

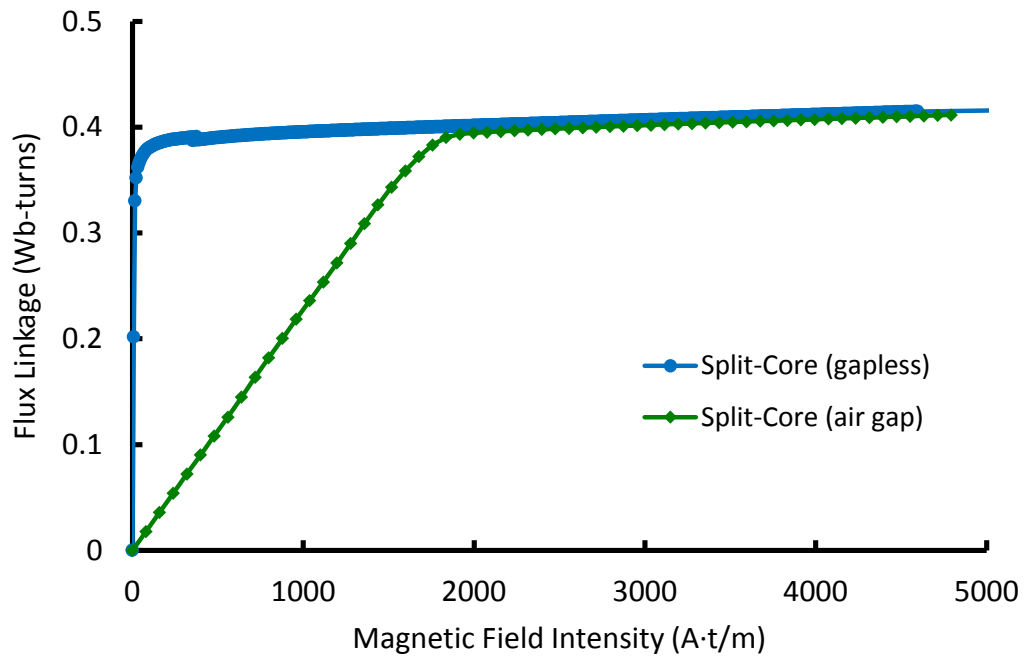


Figure 114 Equivalent Flux Linkages from Split-Core Model

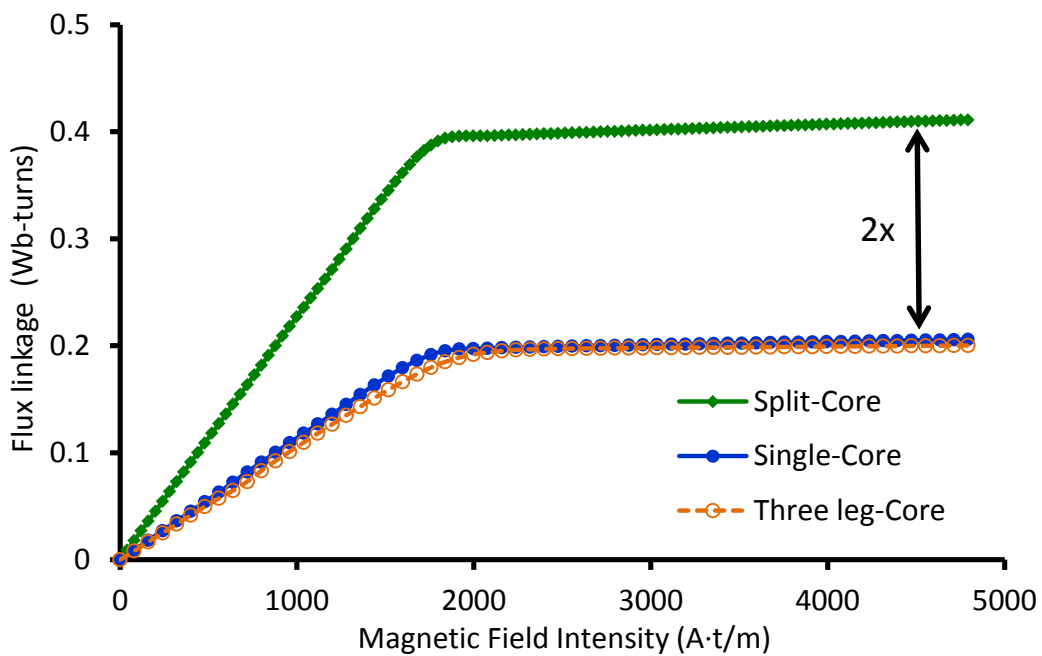


Figure 115 Equivalent Flux Linkage Curves from Two Models

APPENDIX D: SIMULATION PARAMETERS

This section includes parameters for the Trench[®] CCVT Model.

CCVT Model Parameters

Table 15 CCVT Model Parameters

PARAMETER	VALUE
C_1	1280 pF
C_2	263970 pF
L_{COMP}	20.954 mH
L_P	0.514 mH
R_P	0.0547 Ω
L_S	0.43 mH
R_S	0.18 Ω
R_{bur}	40 Ω
L_{bur}	4.66 H
R_{pbur}	1.3 Ω
R_{f1}	1M Ω
L_{f1}	1 μ H
R_{f2}	1 M Ω
L_{f2}	1 μ H
C_{f2}	0.15 μ H
R_{f3}	1 $\mu\Omega$
N_P/N_S	93.1
V_P	132.8 kV
V_S	66.4 V
Frequency	60 Hz

Saturation parameters for the voltage transformer were represented by the i-B curve per PSCAD™ requirements. The i-B formed by application of the following calculation to the B-H characteristic:

$$i = \frac{H \cdot l}{turns} \quad (D.1)$$

Where H is the magnetic field intensity, l is the mean length of the test magnetic core, and $turns$ is the number of turns in the winding.

PSCAD™ requires that the i-B curve be represented by power series of positive terms [101] [102], which it uses to determine the incremental permeability of the core for a given state of operation. The following power series was applied for the Trench® CCVT.

$$i = 0.000407B^{13} + 0.21688 \quad (D.2)$$

Equation (D.2) was determined by applying a two-term power series curve fit to the i-B data using MATLAB®.

PSRC Test System

Application of the PSRC test system in PSCAD™ is shown in Figure 116 and Figure 117. The system shows parallel transmission lines, but the top transmission line was removed from service for the studies presented in this dissertation.

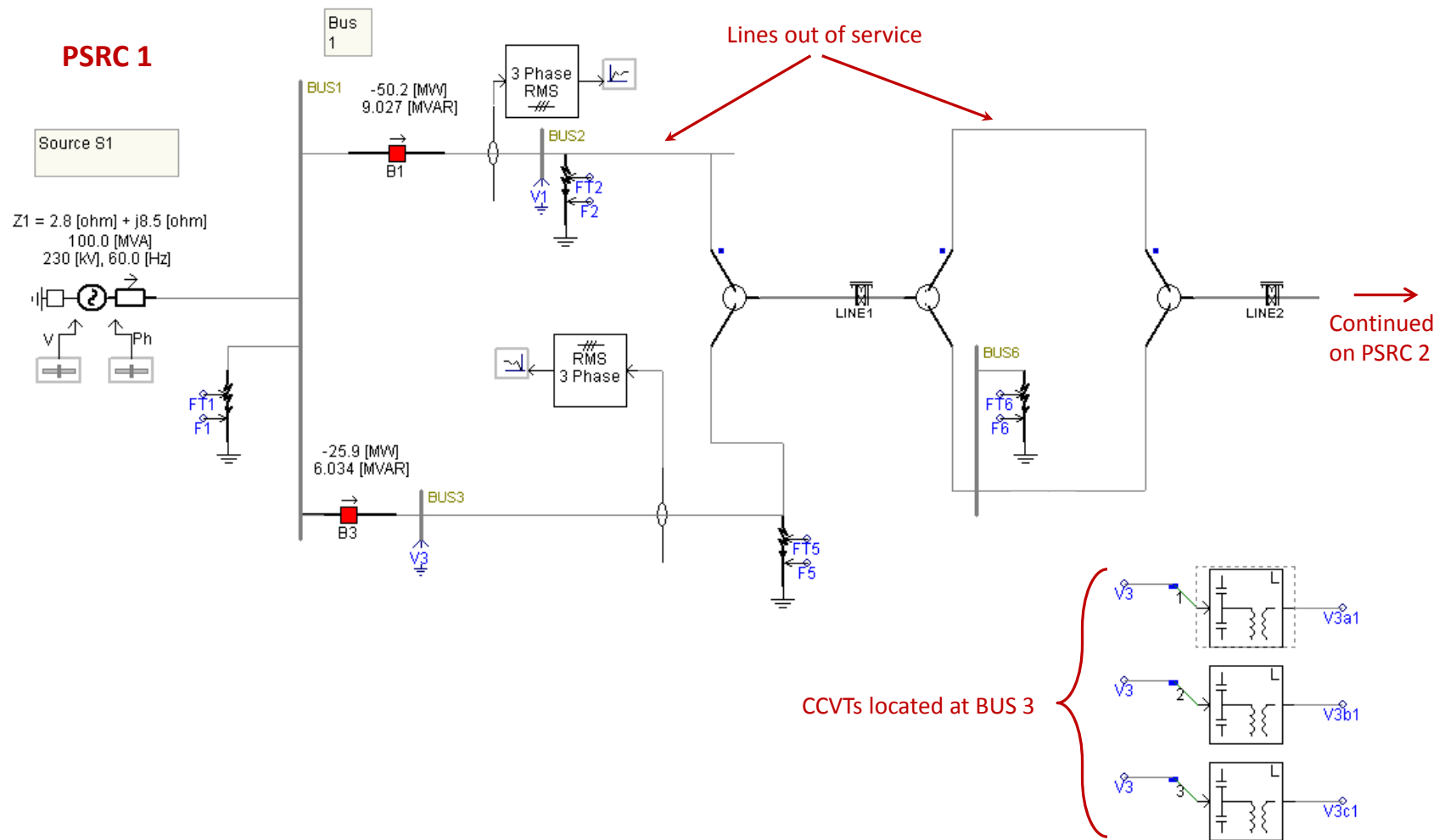


Figure 116 PSRC Distance Protection Test System (PSRC 1)

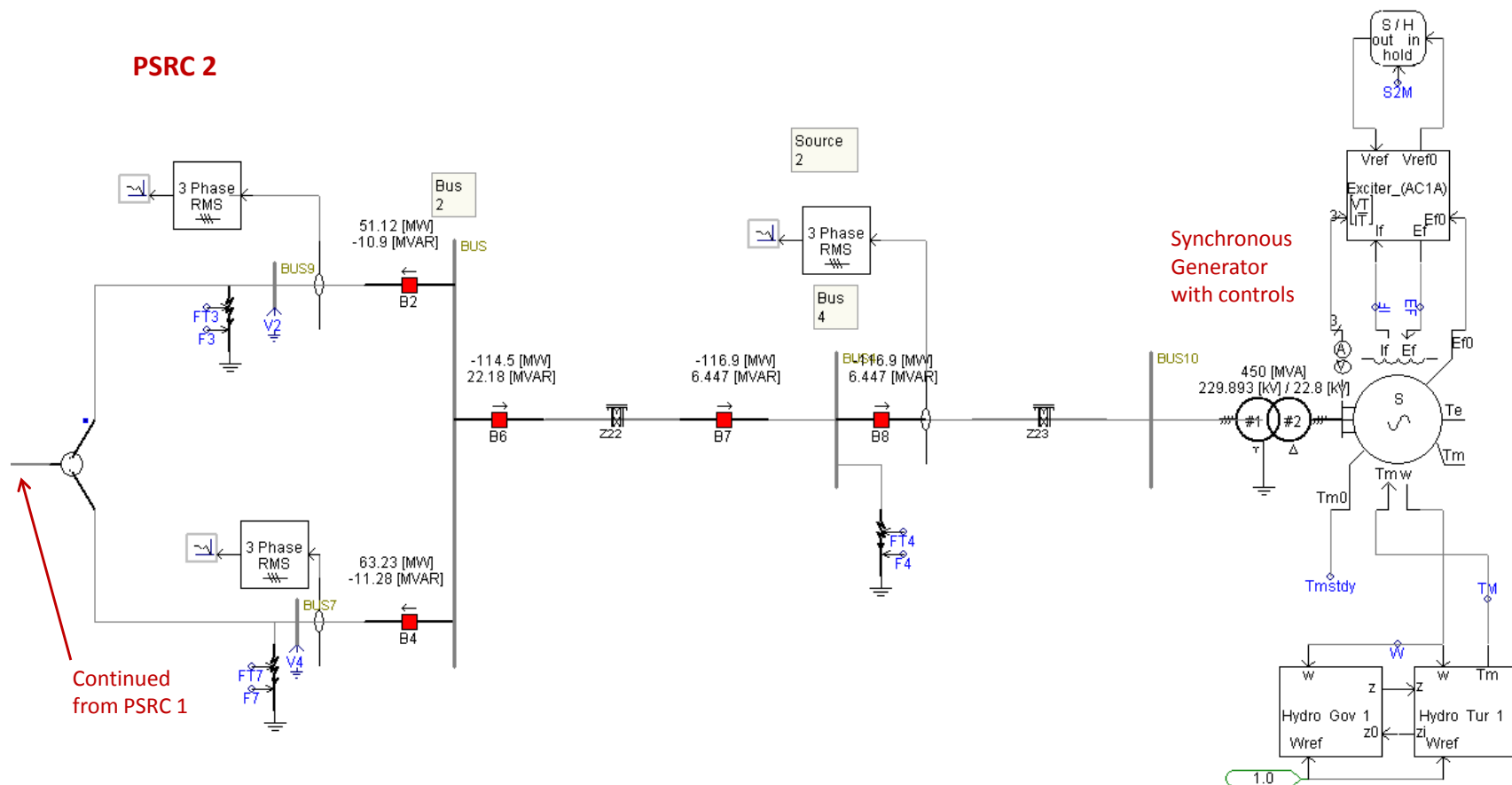


Figure 117 PSRC Distance Protection Test System (PSRC 2)

APPENDIX E MATLAB® SCRIPT

Effective DC Current

```
%IDC CORRECTION to EFFECTIVE DC CURRENT
%
%Calculates the relationship of the applied DC current to
%an effective DC current. The method considers an ideal core and a
%core with air gap. Each core is given effective parameters from FEA
%The method equates the fluxes and determines the currents required by
%designating flux values.
clear;
clc;

%Define variables
l = 1.5667;    %core mean length
la = 2e-3;    %airgap mean length
ldc = l;      %mean DC core length
A = .1295*.1295; %cross sectional area
N = 24;       %number of DC turns
muo = pi*4e-7; %permeability of free space

%load B-H curve (effective core = core + airgap)
load B_H_Case4.dat;
He = B_H_Case4(:,1);
Be = B_H_Case4(:,2);
mue = diff(Be)./diff(He);

%load B-H curve(effective core = gapless core);
load B_H_Case4_nogap.dat
Hng = B_H_Case4_nogap(:,1);
Bng = B_H_Case4_nogap(:,2);
mung = diff(Bng)./diff(Hng);

%Perform transformation and calculate currents
step = 1e-4;
phi = 0:step:(max(Be)*A);
B = phi./A;
for i = 1:length(phi)
    H1(i) = interp1(Be, He, B(i));
    H2(i) = interp1(Bng, Hng, B(i));
end
```

```

leff = H1.*(I/N);
lapp = H2.*(Idc/N);

%Put applied and effective currents in an array
%and output a csv file
I(:,1) = lapp;
I(:,2) = leff;
csvwrite('c',I);
%plot for verification
plot(lapp,leff)
xlabel('lapp')
ylabel('leff')
grid on;

```

DC Sweep Simulation

```

%TWO-CORE MODEL
%AC SWEEP DIFFERENTIAL INDUCTANCE SIMULATION
%
CLEAR;
CLC;
CLF;

%SIMULATION SPECIFIC PARAMETERS
LDMAX = 0.41; %MAX LD AT IDC=0 FOR SPECIFIC CASE (NORMALIZING FACTOR)
%0.405-GAPLESS
%0.0037-AIRGAP
CFLMAX = 0.85;
%0.7945-AIRGAP
%0.815-GAPLESS

%PFC PARAMETERS
NDC   = 24;
NAC   = 12;
LGTH  = 1.5667;
AREA  = 0.1295 * 0.1295;

%READ IN MEASUREMENTS
FILENAME = 'FLUXLINKAGE_SPLITCORE_GAPLESS.CSV';
%FLUXLINKAGE_SPLITCORE_GAPLESS.CSV
%FLUXLINKAGE_SPLITCORE_AIRGAP.CSV

```

```

M = CSVREAD(FILENAME, 0, 0);

%CREATE A THIRD QUADRANT B-H CURVE FROM THE DATA
LAMBDA MEASURED = M(:, 2);
H MEASURED = M(:, 1);

%CREATE A THIRD QUADRANT B-H CURVE FROM THE DATA
H3RD = - H MEASURED;
LAMBDA3RD = - LAMBDA MEASURED;

%OPERATING STATE
PEAK = 400; DELTA = 0.1;
IAC = -PEAK:DELTA:PEAK;
IDC = 80;

%2-CORE H CALCULATIONS
H1 = (NDC*IDC - NAC.*IAC)./LGTH;
H2 = (NDC*IDC + NAC.*IAC)./LGTH;

%INTERPOLATE LAMBDA-H CURVE FROM H VALUES
FOR I = 1:LENGTH(IAC);
    IF H1(I) < 0
        LAMBDA1(I) = INTERP1(H3RD, LAMBDA3RD, H1(I));
    ELSE
        LAMBDA1(I) = INTERP1(H MEASURED, LAMBDA MEASURED, H1(I));
    END
    IF H2(I) < 0
        LAMBDA2(I) = INTERP1(H3RD, LAMBDA3RD, H2(I));
    ELSE
        LAMBDA2(I) = INTERP1(H MEASURED, LAMBDA MEASURED, H2(I));
    END
END

%CALCULATE FLUX LINKAGES AT AC WINDINGS
LAMBDA12 = -LAMBDA1 + LAMBDA2;

%DIFFERENTIAL INDUCTANCE IS THE DERIVATIVE WRT IAC
INDUCTANCE = DIFF(LAMBDA12)./DIFF(IAC);
INDUCTANCE(LENGTH(INDUCTANCE)+1) = INDUCTANCE(LENGTH(INDUCTANCE));

% PLOT(IAC(1:LENGTH(IAC)-1), INDUCTANCE)

```

```

% PLOT(H1, LAMBDA1, 'B', H2, LAMBDA2, 'R')
% PLOT(IAC, -LAMBDA1, 'B', IAC, LAMBDA2, 'R', IAC, LAMBDA12, 'K')

PLOT(IAC, INDUCTANCE/LDMAX, IAC, LAMBDA12/CFLMAX,...
    'R', IAC, -LAMBDA1/CFLMAX, 'G', IAC, LAMBDA2/CFLMAX, 'B')
AXIS([-PEAK PEAK -1 1])
GRID ON;

M = [IAC' INDUCTANCE'/LDMAX LAMBDA12'/CFLMAX...
    -LAMBDA1'/CFLMAX LAMBDA2'/CFLMAX];
FILENAME2 = 'RESULTS NORM IDC80_IACSWEEP_SPLITCORE_GAPLESS.CSV';
%RESULTS NORM IDC0_IACSWEEP_SPLITCORE_AIRGAP.CSV
%RESULTS NORM IDC50_IACSWEEP_SPLITCORE_AIRGAP.CSV
%RESULTS NORM IDC150_IACSWEEP_SPLITCORE_AIRGAP.CSV
%RESULTS NORM IDC0_IACSWEEP_SPLITCORE_GAPLESS.CSV
%RESULTS NORM IDC1_IACSWEEP_SPLITCORE_GAPLESS.CSV
%RESULTS NORM IDC20_IACSWEEP_SPLITCORE_GAPLESS.CSV
%RESULTS NORM IDC80_IACSWEEP_SPLITCORE_GAPLESS.CSV
CSVWRITE(FILENAME2, M)

```

DC Sweep simulation

```

%CALCULATES A REACTANCE VS. IDC CURVE FOR A SINGLE IRMS CURRENT
%
%THE USER-DEFINED FUNCTION XOUTRMS IS USED TO PRODUCE XRMS AT EACH IDC
%VALUE. [XRMS] = XOUTRMSNUMERIC(IRMS, IDC)
%
CLC;
CLF;
CLEAR;

IRMS = [125];    %AC OPERATING CONDITIONS
%125, 60
IAPP = 0:5:250;  %APPLIED DC CURRENT (BY-PASS CONVERSION BY COMMENTING
    %THE CODE BELOW AND SETTING IAPP=IDC.

%CONVERT IAPP TO I EFF USING INTERPOLATION
CONVERTFILE = 'IDC_TO_IEFF.CSV';
M = CSVREAD(CONVERTFILE, 0, 0);
IIN = M(:, 1);  %READ IN APPLIED CURRENTS
IOUT = M(:, 2); %READ IN EFFECTIVE CURRENTS

```

```

FOR K = 1:LENGTH(IAPP)
    IDC(K) = INTERP1(IIN, IOUT, IAPP(K));
END

%PERFORM DC SWEEP
FOR K = 1:LENGTH(IRMS)
    FOR I = 1:LENGTH(IDC)
        [ XRMS(K, I)] = XOUTRMS_2COREFEA(IRMS(K), IDC(I));
    END
END

PLOT(IAPP, XRMS)

TITLE('PFC REACTANCE (VRMS/IRMS) VS. IDC')
XLABEL('IDC (A)')
YLABEL('REACTANCE')
GRID ON;

M = [IAPP' XRMS'];
FILENAME2 = 'RESULTSDCSWEEPIAC125_SINGLECORE_AIRGAP_IEFF.CSV';
%RESULTSDCSWEEPIAC60_SINGLECORE_AIRGAP_IEFF.CSV
%RESULTSDCSWEEPIAC125_SINGLECORE_AIRGAP_IEFF.CSV
%RESULTSDCSWEEPIAC125_SPLITCORE_AIRGAP.CSV
%RESULTSDCSWEEPIAC60_SPLITCORE_AIRGAP.CSV
%RESULTSDCSWEEPIAC30_SPLITCORE_AIRGAP.CSV
%RESULTSDCSWEEPIAC125_SINGLECORE_AIRGAP.CSV
%RESULTSDCSWEEPIAC30_SINGLECORE_AIRGAP.CSV
%RESULTSDCSWEEPIAC60_SINGLECORE_AIRGAP.CSV
%RESULTSDCSWEEPIAC125_SPLITCORE_GAPLESS.CSV
%RESULTSDCSWEEPIAC60_SPLITCORE_GAPLESS.CSV
%RESULTSDCSWEEPIAC30_SPLITCORE_GAPLESS.CSV
CSVWRITE(FILENAME2, M);

FUNCTION [ XLRMS] = XOUTRMS_2COREFEA( IRMS, IDC)
%CALCULATES RMS REACTANCE GIVEN A CONSTANT MAGNITUDE SINUSOIDAL CURRENT AND
%A SINGLE DC CURRENT OPERATING VALUE. THE METHOD CALCULATES THE REACTANCE
%OVER ONE CYCLE OF SINUSOIDAL CURRENT (DELTAT = 1US) AND A SIX SEGMENT B-H
%CURVE. INDUCTANCE IS CALCULATED FROM THE DERIVATIVE OF THE FLUX LINKAGE
%VS. IAC. REACTANCE IS 2*PI*F AT F = 60HZ.

%PFC PARAMETERS
NDC    = 24;

```

```

NAC = 12;
LGTH = 1.5667;
AREA = 0.1295 * 0.1295;

%READ IN MEASUREMENTS
FILENAME = 'FLUXLINKAGE_SINGLECORE_AIRGAP.CSV';
%FLUXLINKAGE_SPLITCORE_GAPLESS.CSV
%FLUXLINKAGE_SPLITCORE_AIRGAP.CSV
%FLUXLINKAGE_SINGLECORE_AIRGAP.CSV
M = CSVREAD(FILENAME, 0, 0);

LAMBDA MEASURED = M(:, 2); % (NAC * AREA) .*
H MEASURED = M(:, 1);

%CREATE A THIRD QUADRANT B-H CURVE FROM THE DATA
H3RD = - H MEASURED;
LAMBDA3RD = - LAMBDA MEASURED;

%AC CURRENT
T = 0:0.00001:0.01667;

IPEAK = IRMS*SQRT(2);
%%%%%%%%[1 2 3 4 5 6 7
HAR = [100, 0, 0, 0, 0, 0, 0]; % [86, 4, 2, 0, 0, 0, 8]
HARSUM = SUM(HAR);

IAC1 = HAR(1)*0.01*IPEAK*SIN(2*1*60*PI.*T);
IAC2 = HAR(2)*0.01*IPEAK*SIN(2*2*60*PI.*T);
IAC3 = HAR(3)*0.01*IPEAK*SIN(2*3*60*PI.*T);
IAC4 = HAR(4)*0.01*IPEAK*SIN(2*4*60*PI.*T);
IAC5 = HAR(5)*0.01*IPEAK*SIN(2*5*60*PI.*T);
IAC6 = HAR(6)*0.01*IPEAK*SIN(2*6*60*PI.*T);
IAC7 = HAR(7)*0.01*IPEAK*SIN(2*7*60*PI.*T);

IAC = IAC1 + IAC2 + IAC3 + IAC4 + IAC5 + IAC6 + IAC7;

%IAC = IRMS * SQRT(2) .* SIN(377.*T);

%2-CORE H CALCULATIONS
H1 = (NDC*IDC - NAC.*IAC)./LGTH;
H2 = (NDC*IDC + NAC.*IAC)./LGTH;

```

```

FOR I = 1:LENGTH(IAC);
%   IF IAC(I) == 0 %IAC = 0 PRODUCES NAN AT DERIVATIVE.
%       IAC(I) = 0.00001;
%   END
    IF H1(I) < 0
        LAMBDA1(I) = INTERP1(H3RD, LAMBDA3RD, H1(I));
    ELSE
        LAMBDA1(I) = INTERP1(HMEASURED, LAMBDA MEASURED, H1(I));
    END
    IF H2(I) < 0
        LAMBDA2(I) = INTERP1(H3RD, LAMBDA3RD, H2(I));
    ELSE
        LAMBDA2(I) = INTERP1(HMEASURED, LAMBDA MEASURED, H2(I));
    END
END

%CALCULATE FLUX LINKAGES AT AC WINDINGS
LAMBDA12 = -LAMBDA1 + LAMBDA2;

%CALCULATE VRMS AND IRMS
IRMS = SQRT(MEAN(IAC.^2));
VDROP = DIFF(LAMBDA12)./DIFF(T);
VRMS = SQRT(MEAN(VDROP.^2));
XLRMS = VRMS/IRMS;
END

```

B-H Curve for SPICE ETable

```

%SPICE TABLE FOR B-H
%AKH1 STEEL
%
%OUTPUTS B-H IN PSPICE TABLE FORMAT IN A .TXT FILE

CLEAR;
CLC;

%####READ IN BH CURVE###
BHDATAFILE = ['C:\USERS\MYOUNG7\DOCUMENTS\DISSERTATION SONY\'...
'DISSERATION MATLAB CODE\AKTRANCOR_AKH1.CSV'];
BHDATA = CSVREAD(BHDATAFILE);

```

```

HDATA = BHDATA(:,1)';
BDATA = BHDATA(:,2)';

%CREATE 3RD QUADRANT VALUES (BH CURVE IS SYMMETRICAL)
BDATA2 = -FLIPLR(BDATA);
HDATA2 = -FLIPLR(HDATA);
%COMBINE B AND H SUBMATRICES
BCOMB = [BDATA2,0,BDATA];
HCOMB = [HDATA2,0,HDATA];

%PLOTS TO VALIDATE THE CURVE
PLOT(HCOMB,BCOMB) %BH CURVE
% %DIFFERENTIAL PERMEABILITY
% PLOT(HCOMB(1:LENGTH(HCOMB)-1),DIFF(BCOMB)./DIFF(HCOMB),'-X')
% AXIS([10E-3 10E3 0 .16])
% GRID ON;

% %###POPULATE PSPICE TABLE IN TXT FILE FORMAT: +(X,Y)###
OUTNAME = ['C:\USERS\MYOUNG7\DOCUMENTS\DISSERTATION SONY\'...
'DISSERATION MATLAB CODE\PSPICE_TABLE_ORNL_PROTOTYPE.TXT'];
FILEID = FOPEN(OUTNAME,'W');
FOR I = 1:LENGTH(BCOMB);
    FPRINTF(FILEID,'%S %12.8F %S %12.8F %S\R\n',...
        '+' ,BCOMB(I),',', HCOMB(I), ');
END
FCLOSE(FILEID);

```

VITA

Marcus Young joined the University of Tennessee-Knoxville in January 2010 for a Doctorate of Philosophy Degree in Electrical Engineering. Marcus completed Bachelor of Science (2005) and Master of Science (2005) degrees in Electrical Engineering from University of Tennessee-Knoxville. Prior to 2005, he was a research assistant in applied superconductivity for energy delivery systems at the Oak Ridge National Laboratory. In 2005, he joined the Electric Power Research Institute and became a senior project engineer/scientist in EPRI's Power Delivery and Utilization Sector. In 2010, Marcus joined ORNL's Power and Energy Systems group. His current research interests include power system modeling, substation automation, and wide-area measurements. Marcus is a licensed Professional Engineer in the State of Tennessee.