DESIGN OF DUAL-OUTPUT ALTERNATORS WITH SWITCHED-MODE RECTIFICATION

by

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Abstract

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ABSTRACT

Electrification of many automotive functions and the desire to introduce many new features are dramatically increasing the electrical power requirements of vehicles. The increasing power demands are becoming very challenging within the context of the present 14V electrical system and have sparked investigation of a higher-voltage electrical system. The introduction of a 42V electrical system for future automobiles is therefore gaining widespread industry acceptance. The large number of electrical subsystems in today's 14 Volt vehicles make it extremely challenging for manufacturers to make a direct transition to a single 42 V electrical system, therefore dual-voltage (42V/14V) automotive electrical systems are attracting considerable interest. This push to introduce dual-voltage (42V/14V) automotive electrical systems necessitates power generation solutions that are capable of supplying power to both 14V and 42V electrical loads. A number of approaches for implementing dual-voltage electrical systems have been proposed, but most suffer from severe cost or performance limitations. This thesis explores the design of alternators incorporating dual-output switched-mode rectifiers. The approach enables the full load-matched power capability of the alternator machine to be achieved, with power delivered to the two outputs in any desired combination. Switched-mode rectifier topologies for this application are introduced. The design guidelines for alternators with switched mode rectifiers are established, and appropriate control laws derived. A prototype dual-output alternator incorporating a switched-mode rectifier is designed and built. Simulation and experimental results that demonstrate the feasibility and high performance of the approach are presented.

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Dedicated to
my wife Lami (Hepzibah) and my children Hadassah and Nasara
for their love
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1.1 Electrical System and Trends

In the 1950s automobiles used 6-Volt power supplies. The introduction of new features like radios, higher power headlamps, more powerful electric starting motors and the need for a higher ignition energy for the V8 engines all stretched the capability of the 6-Volt system. As a result the present 12-Volt electrical system\(^1\) was introduced in 1955.

Today the automobile industry is faced with a similar situation. Electrification of many automotive functions and the desire to introduce many new features are now dramatically increasing the electrical power requirements of vehicles. Functions including electrically powered pumps and valves, and amenities such as heated windscreens and seats, entertainment electronics and navigational aids are being installed in the car. A typical luxury class vehicle today draws between 1200W and 1500W of steady state power from the electrical system. The steady state electrical power needed in the year 2005 is estimated to be in the range of 2500W to 7000W. Figure 1.1 shows the projected trends in automotive electrical system power requirement.

---

\(^{1}\) One can denote the voltage of the electrical system either by the open circuit voltage of the battery (e.g. 12.6 ~ 12 V today) or by the alternator regulation voltage (e.g. 14.2 ~ 14 V today). For the remainder of this thesis, we will refer to the regulation voltage.
Figure 1.1  Projected trends in automotive electrical system (Cited in [1])

The increasing power demands are becoming very challenging within the context of the present 14V electrical system and have sparked investigation of a higher-voltage electrical system.[1, 4, 6]. Through the work of the MIT/Industry consortium on Advanced Automotive Electrical/Electronic Components and Systems and other organizations, introduction of a 42V electrical system for future automobiles is gaining widespread industry acceptance [4]. Introduction of a 42V bus provides important opportunities. First, it enables present loads that are mechanically powered by the engine to be electrified and placed under electronic control, with resulting improvements in performance and efficiency. Second, it enables the introduction of new high-powered features and functions.

The prospect of changing completely to a new voltage of 42 Volts is quite challenging for the Original Equipment Manufacturers (OEM) as this will mean
heavy investments in the redesign of their production plants. Furthermore, important low-voltage loads like incandescent lights and some low-power electronic modules that prefer low voltage operation appear unlikely to benefit from the introduction of the higher system voltage. It is because of this predicament that at least initially, both 42V and 14V networks are expected to coexist in vehicles. Such dual-voltage electrical systems provide the flexibility of accommodating both high-voltage loads and low-voltage loads.

1.2 Dual Voltage System Architectures

Various system architectures have been proposed for implementing dual-voltage systems[1, 2, 3, 6, 7, 8, 10, 11, 13, 14, 15, 16]. Architectures under consideration by the auto industry are:

1. Dc/dc Converter-based architecture,

2. Dual-Voltage Alternator-based architecture
1.2.1 Dc/dc Converter Based Architecture

In a dc/dc converter-based architecture (Figure 1.2), the alternator generates 42V and supplies the 42V bus. The 42V dc is further processed by a dc/dc converter to supply the 14V bus.

![Diagram showing dc/dc converter based architecture.]

**Figure 1.2** Dc/dc converter based architecture.

The dc/dc converter architecture is highly effective. The design of dc/dc converters is relatively well understood, and such converters can achieve very good performance. Furthermore, if a bi-directional converter is used, the 42V battery may be charged from the 14V battery if needed. Nevertheless, dc/dc converters for the application are extremely expensive by automotive standards, and are significantly larger than desired. As a result, this architecture is technically viable (and is likely to be used in the short term) but is probably not economically viable for long-term use.
1.2.2 Dual-Voltage Architecture

This architecture is further classified into two groups, namely:

- Dual-wound alternator architecture and
- Dual-rectified alternator architecture

1.2.2.1 Dual-Wound Alternator Architecture

Figure 1.3 illustrates the dual-wound alternator architecture. In this topology, the alternator has two separate sets of stator windings, each supplying an output via a rectifier.

![Dual-wound alternator architecture](image)

**Figure 1.3** Dual-wound alternator architecture.
In one possible implementation, the 42V bus is supplied by one winding via a diode bridge rectifier, while the 14V bus is supplied by the other winding via a phase-controlled rectification [3]. The two outputs are controlled by a combination of field control and phase control. The fact that field control is common to both outputs poses some difficulties in fully regulating both outputs, and in achieving good use of the alternator power capability under all operating conditions. Design and limitations of this approach are considered in [3].

1.2.2.2 Dual Rectified Alternator Architecture

The dual rectified alternator architecture (Figure 1.4) utilizes an alternator with a single winding and a two-output rectifier to supply the high and low voltage buses respectively. One implementation approach that has been explored in the past is illustrated in Figure 1.5 [2, 3, 5].

![Dual Rectified Alternator Architecture](image)

**Figure 1.4** Dual rectified alternator architecture.
In the implementation of Figure 1.5, a conventional diode bridge rectifier is used to supply the 42V bus. The 14V bus is supplied through a semi-controlled rectifier made of three additional thyristors in conjunction with the bottom diodes of the diode bridge. The two outputs can be regulated by a combination of field current and phase control.

This approach has some significant limitations. One limitation is that it is difficult to optimize the machine to be well utilized under different loading conditions. Furthermore, extremely large output filters are required due to the very large current ripple appearing at the outputs [3]. The frequency of this ripple is a small multiple of the low alternator electrical frequency and occurs as the thyristors chop current between the outputs.
1.3 ALTERNATOR WITH DUAL-OUTPUT SWITCHED-MODE RECTIFIER

As explored in this thesis, the limitations facing dual-output alternators with phase controlled rectifiers can be overcome by employing a dual output switched-mode rectifier (SMR) [2, 5]. Figure 1.6 illustrates one possible topology. In this topology, the lower diodes of the 42V-bridge rectifier are replaced by switches (MOSFETs) which are switched on and off at high frequency using pulse width modulation. Figure 1.7 shows alternative circuit topologies having similar characteristics that can be used for realizing such alternators.

![Dual-output switched-mode rectifier circuit](image)

**Figure 1.6** Dual-output switched-mode rectifier circuit.

The basic circuit operation of the switched-mode rectifier technique can be understood by considering the circuit topology of Figure 1.6.
At the beginning of a PWM cycle all MOSFETS \((Q_x, Q_y, Q_z)\) are switched on and remain on for a specified time. The Diodes \((D_x, D_y, D_z)\) are reversed biased (off) when the MOSFETs are on. At the turn off of the MOSFET a subset of the diodes became turned on (forward biased). After a specified diode conduction period, thyristors \((T_x, T_y, T_z)\) are fired and the machine phase currents are directed through a subset of thyristors for the remainder of the operating cycle. The machine current is therefore chopped back and forth between the two outputs at the high switching frequency of the PWM signal. The switching on and off of the switches at high frequency results in substantial output filter size reduction and improved control bandwidth for regulating the 14V bus. By controlling the field current, MOSFET duty ratio and thyristor duty ratio it is possible to regulate the two output voltages and at the same time meet the load matching condition of the alternator. These mentioned benefits of SMR cannot be obtained or achieved with conventional circuit implementations. Figure 1.7 shows alternative topologies that can be used to realize the SMR techniques.
Figure 1.7 (a), (b), (c) Alternative circuit topologies for the switched-mode rectifier.

1.4 Thesis Objectives and Contributions

The objective of this thesis is to fully investigate alternators with dual-output switched-mode rectification, to be used for the implementation of 42V/14V dual voltage electrical systems for the next generation automobile. This approach enables the full load-matched power capability to be achieved, with power delivered to the two outputs in any desired combination. The thesis will investigate the design of alternators incorporating dual-output switched-mode rectifiers. This will encompass the development of control laws and design rules.
rectifiers. This will encompass the development of control laws and design rules for dual-output alternators with switched-mode rectifiers. Furthermore, it includes the design, construction, simulation, and experimental evaluation of a prototype dual-output alternator.

1.5 Thesis Organization

This thesis consists of six chapters. Chapter 1 is the introduction. The development of control laws for the alternator/rectifier systems are presented in Chapter 2. Chapter 3 discusses the design and construction of the prototype alternator/SMR system. It also presents the experimental setup. Simulations and experiments and interpretation of the results are presented in Chapter 4. Chapter 5 discusses the economic considerations in the introduction of a higher voltage automotive electrical system and finally, Chapter 6 summarizes the findings of this research and presents possible suggestions for further work.
CHAPTER 2

CONTROL LAWS AND DESIGN TRADEOFF

2.1 LUNDELL ALTERNATOR

The Lundell alternator is a 3-phase synchronous machine. The field winding is wound on the rotor, and the regulator feeds the field current via slip rings. The simple circuit model of Figure 2.1 can be used to derive the electrical behavior of the alternator [9].

![Diagram of Lundell Alternator Model]

Figure 2.1 Simple Lundell Alternator Model
Based on the results of [9], the alternator power output $P_{out}$ can be calculated as

$$P_{out} = \frac{3V_x}{\pi} \sqrt{\frac{V_s^2 - \left(\frac{2V_x}{\pi}\right)^2}{\omega^2 L_s^2}}$$  \hspace{1cm} (2.1)$$

where $V_s$ is the output voltage, $V_x$ is the line-to-neutral back emf magnitude, $\omega$ is the alternator electrical frequency, and $L_s$ is the armature synchronous inductance.

Computing the back emf voltage as $V_s = k\omega i_f$ (where $k$ is the machine constant in V-s/(rad-A) and $i_f$ is the field current), this can be written as

$$P_{out} = \frac{3V_x}{\pi} \sqrt{\frac{(k\omega i_f)^2 - \left(\frac{2V_x}{\pi}\right)^2}{\omega^2 L_s^2}}$$  \hspace{1cm} (2.2)$$

To find the load matching condition where output power is maximized, (2.2) is differentiated with respect to $V_x$ and equated to zero:

$$\frac{\partial P_{out}}{\partial V_x} = \frac{3}{\pi} \sqrt{\frac{(k\omega i_f)^2 - \left(\frac{2V_x}{\pi}\right)^2}{\omega^2 L_s^2}} \frac{3V_x}{2\pi} \frac{1}{\omega^2 L_s^2} - \frac{3V_x}{2\pi} \frac{(k\omega i_f)^2 - \left(\frac{2V_x}{\pi}\right)^2}{\omega^2 L_s^2} \frac{8V_x}{\pi^2 \omega^2 L_s^2} = 0$$  \hspace{1cm} (2.3)$$
This can be solved for the load-matched operating voltage:

\[ V_X = \frac{\pi k \omega i_f}{2\sqrt{2}} \]  \hspace{1cm} (2.4)

Thus in the load-matched case, we find an average output current of

\[ \langle i_X \rangle = \frac{3k i_f}{\sqrt{2\pi L_s}} \]  \hspace{1cm} (2.5)

which results in a load-matched alternator output power of

\[ P_{out} = \frac{3k^2 \omega^2 i_f^2}{4\omega L_s} \]  \hspace{1cm} (2.6)

Thus, (2.4) specifies an output voltage for load-matching that results in output power (2.6). These results will be used to develop dual-output rectifiers that achieve load-matched operation.
2.2 Operation of the Switched-Mode Rectifier

The diagrams showing members of the proposed class of switched-mode rectifiers are shown in Figure 1.7 of Chapter 1. The basic circuit operation of the proposed class of dual-output switch-mode rectifiers has been briefly explained in Chapter 1, but is redeveloped here for clarity. We focus on the operation of the simplified circuit model of Figure 2.2. Figure 2.3 shows the pulse width modulation (PWM) signals applied to the gates of MOSFETS Q₁ and Q₂ of Figure 2.2. We note that the PWM switching period T is much shorter than an alternator electrical cycle, so that rectified alternator current \( i_x \) may be treated as a constant \( I_x \) over a switching period.

![Simplified model for the SMR](image)

**Figure 2.2** Simplified model for the SMR

As illustrated in Figure 2.3, each switching cycle can be broken up into three time segments: In the first segment, which we denote as being \( hrT \) in length, the diode \( D_3 \) conducts delivering the rectified alternator current \( i_x \) to the 42V bus. In the second time period, which lasts for \( h(1-r)T \), \( Q_2 \) is switched on. In this period diode \( D_2 \) and switch \( Q_2 \) conduct delivering the rectified alternator current \( i_x \) to the
14V bus. In the final time period, lasting \((1-h)T\), switch \(Q_t\) conducts, and the current \(i_x\) is shunted to ground bringing the voltage \(v_X\) to zero. (See waveforms "A" "D" and "E" of Figure 2.3).

Figure 2.3 Circuit waveforms over a switching cycle. (A): voltage \(v_X\), (B): at \(Q_t\) gate (C): at \(Q_s\) gate, (D): current into 14V bus and (E): current into 42V bus.
Equations for the average rectified bridge voltage $<v_x>$ and the average output currents $<i_{42}>$ and $<i_{14}>$ are derived from waveforms A, D and E in Figure 2.3:

\[
<v_x> = hr V_{42} \pm h(1-r) V_{14}
\]  \hfill (2.7)

\[
<i_{42}> = hr i_x
\]  \hfill (2.8)

\[
<i_{14}> = h(1-r) i_x
\]  \hfill (2.9)

where

$v_x$, $i_x$ are the instantaneous alternator voltage and current at the diode bridge output.

$v_{14}$, $i_{14}$ are the instantaneous voltage and current at the 14V bus.

$v_{42}$, $i_{42}$ are the instantaneous voltage and current of 42V bus and

"h" and "r" are switch time controls expressed as fractions.

2.2.1 Alternator Control Law

Here we consider a control law for the alternator and dual-output Switched-Mode Rectifier of Figure 2.2. Given desired (reference) output currents $i_{14}^*$ and $i_{42}^*$, the controls specify the control handles $h$, $r$, $i_f$ such that the desired output currents are achieved. In the event that the desired power exceeds the output capability of the alternator for a given operating point, the alternator should deliver the maximum power possible, with output currents in the desired proportion. Furthermore, the control laws should guarantee that maximum power capability of the alternator for a given operating point could be achieved.
From (2.6) the field current is expressed as

\[ i_f = \sqrt{\frac{4L_i P_{out}}{3k^2 \omega}} \quad (2.10) \]

we also have \( P_{out} = P_{14} + P_{42} = V_{14}i_{14}^* + V_{42}i_{42}^* \) where

\( i_{14}^*, i_{42}^* \) are the desired currents on the 14V and 42V buses respectively.

As described below, control laws meeting these requirements are given as follows:

\[ i_f = \sqrt{\frac{4L_i}{3k^2 \omega}}(V_{14}i_{14}^* + V_{42}i_{42}^*) \quad (2.11) \]

\[ r = \frac{i_{42}^*}{i_{14}^* + i_{42}^*} \quad (2.12) \]

\[ h = \frac{\pi k \omega i_f}{2\sqrt{2(rV_{42} + (1-r)V_{14})}} \quad (2.13) \]

where \( i_f, r, h \) are the control handles, \( i_{14}^*, i_{42}^* \) are the current commands (references), and

\( k \) is the machine constant in \( V \cdot s / rad \cdot A \)

\( L_s \) is the machine synchronous inductance in Henrys

\( \omega \) is the alternator electrical frequency in \( rad / sec \).

Equation (2.11) is based on the load-matched power capability derived in (2.6). It sets the field current \( i_f \) to that required for the desired output power (equal to \( V_{14}i_{14}^* + V_{42}i_{42}^* \)) up to the maximum permissible level. (2.12) picks the fraction "\( r \)"
such that the currents at the two outputs are in the desired proportion. As can be seen in Figure 2.3 "r" is the fraction of time \( i_r \) is delivered to the 42V bus as compared to the time it is delivered to either the 14V or 42V bus. Choosing "r" directly in this fashion guarantees that the output currents will remain in the desired proportion even if the total desired output power is not achievable. Finally, time fraction "\( h \)" in (2.13) is selected to maintain the load matched condition on the alternator (or as close to it as possible) such as that the full power capability of the alternator can be achieved under any operating condition. (2.13) follows directly from a combination of the required load-matched voltage \( v_x \) (2.4) and that imposed by the control (2.7).

2.3 Design Tradeoffs

In the design of dual voltage alternators with switched mode rectification, the interrelations of various parameters need to be addressed. The alternator machine parameters, (e.g., \( k, L_s, \) and \( R_s \)) depend on the machine geometry and how it is wound. Machine winding in part reflects a choice between many turns of small wire vs. fewer turns of large wire. Since the SMR control range depends on the parameters mentioned above, the alternator winding selection produces a tradeoff between the power capability at the two outputs vs speed and the ratings of the rectifier devices. It may be desirable to wind the alternator in such a way that maximum available load-matched power can be delivered to either output. However achieving this capability heavily impacts SMR device sizing. To illustrate
this tradeoff, Figure 2.4 shows graphs of device and output current ratings versus the normalized number of stator winding turns for a conventional 120 A, 14V alternator geometry. (The winding normalization is such that 1 represents the number of turns on a conventional 120 A, 14V alternator design.). Refer to Appendix A for the Matlab script used to calculate the data in Figure 2.5.

It can be seen that at idle speed, there is no remarkable change in the output current capability at the two buses with variation in number of stator winding turns. This is also true for the 42V current capability at cruising speed (up to a normalized number of winding turns of approximately 0.8.). There is a great variation, however, in the output current capability at the 14V bus at cruising speed and in the RMS current rating of the boost switch, as the number of stator winding turns varies. Achieving full load-matched power capability into the 14 V bus at cruising speed requires a greatly reduced number of winding turns (~ 0.3 normalized). This capability is achieved only at greatly increased current rating of devices Q₁ (shown) and Q₂ (which tracks I₁₄). This trend makes sense: if the alternator is to be able to direct full matched power to either output under all conditions, the machine must have a load matched voltage below 14 V under all conditions, and the SMR must operate over a wide boost range to direct load-matched power to either output. This results in a high rating of the SMR devices.

As full load-matched capability at cruising speed is not likely to be needed, an intermediate design that cannot direct full load-matched power to either bus under all conditions may be more desirable in view of rectifier component ratings.
For most practical designs, normalized winding turns in the range of $0.7 - 1$ may be desirable, as it enables full power delivery under idle-speed conditions to either bus, and full power delivery to the 42-V bus under any condition, while preserving low rectifier current rating. The graph of Figure 2.4 (or a similar one for an alternator geometry of choice) may be utilized to evaluate this tradeoff.

![Graph showing current tradeoff](image)

**Figure 2.4** Device and output current tradeoff.
CHAPTER 3
SMR CIRCUIT DESIGN AND EXPERIMENTAL SETUP

3.1 INTRODUCTION

This chapter discusses the design, construction, and experimental setup of a prototype dual-output alternator/SMR system. It addresses the selection of suitable components, including power MOSFETs, Schottky diodes, and output filter capacitors. Design of the converter heatsink and design and operation of the control circuitry is also addressed.

3.2 DESIGN CONSIDERATION

Figure 3.1 represents the actual circuit developed and used for testing. A detailed description of the operation of the SMR circuit is presented in section 2.2 of Chapter 2. In the circuit of Figure 3.1, the three alternator phases are each connected to the midpoint of a semi-bridge, with a MOSFET in the bottom position and a diode in the top position. These six elements combined perform the functions of $Q_1$ and $D_3$ in Figure 2.2, as well as the function of the diode bridge (schematically represented in Figure 2.2). Also connected to each alternator phase is the series combination of a diode and MOSFET. These elements serve the function of $D_2$ and $Q_2$ in Figure 2.2.
Figure 3.1  SMR circuit: partitioned into three segments.

For the purpose of explanation, the complete SMR system will be partitioned into three parts as illustrated in Figure 3.1. Part I constitutes the 42V SMR stage, Part II is the 14V SMR stage and Part III, the Pulse Width Modulation circuit.

3.2.1 42V SMR stage
This stage had been designed and built already as part of a single-output SMR-based alternator system [2], [5], [16], [17]. A detailed circuit diagram of the 42V SMR is seen in Figure 3.2.
Figure 3.2  42V SMR stage.
Chapter 3: SMR Circuit Design and Experimental Setup

The circuit is a 3-phase semi-bridge rectifier with N-channel MOSFETs (IXFN230N10) connected in the lower positions and diode modules (DSS2x41-01A), each containing a pair of parallel Schottky diodes, connected in the upper positions. High current integrated circuit (IC) FET drivers, UC 2710 and associated components are used to drive the MOSFETS. These ICs accept low-current high-speed digital inputs to activate the high current, 10V outputs required to turn on the MOSFETS. Two 3-terminal positive voltage regulators LM317H and LM340 are used to provide the +10V and +15V supplies to the electronic components. For the purpose of integration with the 14V SMR stage, the UC3823A PWM controller used for the single-output SMR-based system discussed in [16] has been removed. Instead another PWM controller, to be described later, is required for the dual-output SMR based system.

3.2.2 14V SMR Stage

The detailed schematic diagram of one phase of this stage is presented in Figure 3.3. For each of the three phases, an N-Channel MOSFET (IXFN230N10) and a diode module (DSS2x41-01A), each containing a pair of parallel Schottky diodes, are connected in series. Since the MOSFET switches are identical to the MOSFET switches in the 42V SMR stage, the same type of drivers, UC2710, are used to provide the gate voltage. Since the N-Channel MOSFETS on the 14V SMR boards are used for the high-side (or "upper") switching, the gate drive signal requires level shifting. This is achieved by the combination of a 15V single-output isolated
Figure 3.3  Single phase of 14V SMR stage
dc-dc converter module NDL1215, and the voltage regulator LM317HV to provide the floating +10V needed. They are shown in Figure 3.3 as U8 and U9 respectively.

3.3 THERMAL CONSIDERATIONS

A major design criterion for most of the power components in the SMR is limited maximum temperature. Power dissipated in semiconductor components and resistances is converted into heat and causes a rise in temperature. If the temperature of the junction of any semiconductor device exceeds a certain maximum level, the device may be damaged. To keep junction temperatures within limits, two major techniques are available: (1) Limit dissipation density by using large devices and if necessary, connecting them in parallel and (2), attach devices to a heat sink, to limit device case temperatures. Once the necessary heat sink performance requirement has been established, a suitable heat sink can be selected. Many factors such as performance, available space, mounting arrangements, and cost influence the choice of heat sink. Calculations for the appropriate heat sinks for the two stages are presented in the following subsections.
3.3.1 Specifications of Heat Sinks

Specifications of power components to be mounted on the heat sinks are given in the data sheets.

a. **MOSFET (3 sets):**

   Model: IXFN 230N10

   $$V_{DS} = 100 \text{ V}$$

   $$R_{ds\text{on}} = 0.006 \Omega \text{. @ } T_j = 25^\circ\text{C}$$

b. **SCHOTTKY DIODES (3 sets):** Each set represents a pair of diodes connected in parallel.

   Model: DSS 2 x 41-01A

   $$V_{R\text{RM}} = 100 \text{ V}$$

   $$V_F = 0.70 \text{ V}$$

3.3.1.1 42V SMR Stage

The currents through the power components are determined based on initial design calculations on the circuit in Figure 3.4 [16]. Figure 3.4 represents the parallel combination of all three phases in the three-phase system.

![Figure 3.4](image)

**Figure 3.4** Circuit Model used for determining maximum operating currents.
Chapter 3: SMR Circuit Design and Experimental Setup

The relationship between the current $i_x$ and the voltage $v_o$ depends on alternator model [16], and with the worst case condition $i_{d,\text{average}}$ equals 75A and $i_{q,\text{rms}}$ equals 95A[16]. These values are used to determine the size of the heat sink as follows:

(a).  Calculating Conduction and Switching Losses

For MOSFETs:

There are three parallel MOSFET modules. From data sheet this implies that at

$$T_j = 90 \, ^\circ\text{C}, \text{ normalized } R_{ds\,on} = (1.6 \times 0.006 )/ 3 = 3.2 \times 10^{-3} \, \Omega$$

$$P_{\text{conduction}} = i^2 R_{ds\,on} = (95A)^2 \times 3.2 \times 10^{-3} = 29.33 \, \text{W (approx. 30 W)}$$

The switching loss is considered to be fairly equal to the conduction loss. This assumption represents a rough estimation which is allowable because the devices are extremely oversized. The total conduction and switching losses is then:

$$P_{\text{total}} \approx 30 \times 2 = 60 \, \text{W}$$

For SCHOTTKY Diodes

$$P_{\text{conduction}} = (75A) \times 0.7 \, \text{V} = 52.5 \, \text{W}$$

For the package $R_{TH,JC}$

MOSFET : 0.26°C/W $\Rightarrow$ Net for 3 packages is 0.09°C/W

DIODE : 1.1°C/W each diode $\Rightarrow$ Net for 3 packages(6 diodes) is 0.18°C/W

![Diode schematic diagram](image_url)
\[ \Delta T_{CA} = (112.5 \text{ W}) \left( \frac{R_{TH,CA}}{\text{°C/W}} \right) \]

for \( \Delta T_{CA} \leq 40^\circ\text{C} \Rightarrow R_{TH,CA} \leq 0.36^\circ\text{C/W} \)

Therefore the specification of the heat sink required is as follows:

\[ R_{TH,CA} \leq 0.36^\circ\text{C/W} \]

Stacking 6 modules close together results in a size of 9" x 6" roughly needed.

To meet this requirement, IMI Marston model 96CN02500A200, size (40mm x 300mm x 250mm, \( R_{TH,CA} = 0.29^\circ\text{C/W} \)) was acquired.

This heat sink rating and \( \Delta T_{CA} \) are reasonable for bench operation. They cannot, however, be applied in an automotive situation where the underhood ambient temperature may reach 85°C or higher depending on vehicle speed and weather condition.

3.3.1.2 14V SMR Stage

The specification of the heat sink can be determined in a similar manner as in the previous section. In worst case situation, the maximum rms bus current is 126A. (from simulation at 6000 rpm alternator speed.)

(a). Calculating Conduction and Switching Losses

For MOSFETs:

We have three parallel MOSFETs, and from the data sheet

at \( T_j = 90 \text{ °C} \), normalized \( R_{ds,on} = (1.6 \times 0.006) / 3 = 3.2 \times 10^{-3} \text{ Ω} \)

\[ P_{\text{conduction}} = i^2 R_{ds,on} = (126\text{A})^2 \times 3.2 \times 10^{-3} = 50.50 \text{ W} \]
Because the switching frequency is the same as for the 42V stage, the same assumption about switching loss (equal to conduction loss) will be used.

\[ P_{\text{total}} \approx 50.50 \times 2 = 101 \, \text{W} \]

For SCHOTTKY Diodes

\[ P_{\text{conduction}} = 126A \times 0.7 \, \text{V} = 87.45 \, \text{W} \]

For the package \( R_{\text{TH,JC}} \)

MOSFET : 0.26°C/W \( \Rightarrow \) Net for 3 packages is 0.09°C/W

DIODE : 1.1°C/W \( \Rightarrow \) Net for 3 packages is 0.18°C/W

\[
\begin{align*}
\Delta T_{\text{CA}} &= (188.45 \, \text{W}) \times (R_{\text{TH,CA}} \, ^\circ\text{C}/\text{W}) \\
\text{Assuming} \quad \Delta T_{\text{CA}} &\leq 40 \, ^\circ\text{C} \quad \Rightarrow R_{\text{TH,CA}} \leq 0.21 \, ^\circ\text{C}/\text{W}
\end{align*}
\]

Therefore the specification of the heat sink required is as follows:

\[ R_{\text{THCA}} \leq 0.21^\circ\text{C}/\text{W} \]

The closest available is heat sink # 5429 Extrusion, size (31mm x 340mm x 260mm, \( R_{\text{THCA}} = 0.22^\circ\text{C}/\text{W} \)) and was acquired from Richardson Electronic Ltd. NY. Also as earlier mentioned, this rating is reasonable only for bench operation.
3.3.1.3 Output Filter Capacitor Calculation

While the operating frequency of this stage is quite high, the ripple current can be quite large. The circuit is proposed for use with a large battery load, which can be expected to have some tolerance for ripple current. Nevertheless, it has been decided to install filter capacitors to substantially limit ripple voltage. The output voltage, $V_{\text{out}}$ is 14 V. Assuming the worst case condition, the output current is assumed to be a square wave between 0 and 126 amperes. The a.c. current into the capacitor is then a zero-value square wave ±63A with rms value of 63A. A capacitor capable to handle this high current, from the ITW Pakton capstick capacitor selection guide for power conversion, is type CS4 having the specifications as indicated in Table 3.1.

<table>
<thead>
<tr>
<th>Capacitance ($\mu$F)</th>
<th>PF Code</th>
<th>DC Voltage (V)</th>
<th>ESR @ 500 KHz ($\Omega$)</th>
<th>RMS Current @ 500 KHz (A)</th>
</tr>
</thead>
<tbody>
<tr>
<td>10</td>
<td>106</td>
<td>100</td>
<td>0.003</td>
<td>15.30</td>
</tr>
</tbody>
</table>

Table 3.1 Specifications for Capstick Capacitor CS4.

The minimum number of capacitor needed is:

$$63/15.30 = 5$$

however, 9 Capstick capacitors are used (3 for each phase). These additional capacitors further reduce the ripple. In Figure 3.3, the capacitors in question are $C_{13}$, $C_{14}$ and $C_{15}$. In a similar manner of calculations, 6 Capstick (2 for each phase) are utilized at the output of the 42V SMR stage board. In Figure 3.2, the capacitors are $C_{1} \sim C_{6}$.
3.3.2  **PWM Controller**
A 16-Bit CMOS micro-controller, C167CR, is programmed\(^1\) to give the required pulse width modulating signals needed at the gates and also provide the flexibility of adjusting the duty ratios manually for setting the \(h\) and \(r\) values.

3.3.3  **Schematics and Board design.**
The schematic drawings and the printed circuit board design were accomplished using Protel SE 1999 software. The strategy in designing the PCB boards for the SMR system is to ensure that the high current paths have large copper surface area and also are as short as possible in length in order to mitigate \(i^2R\) loss. This also encouraged the utilization of the two sides of the board for the copper layout. See Figure 3.5 for the PCB layouts, and Figure 3.6 for a photograph of the resulting SMR.

\(^1\) PWM programmed by James Geraci.
Figure 3.5 Printed Circuit Boards: (a) Upper side; (b) Upper and Lower sides; (c) Top overlay
Figure 3.6 Photograph showing the SMR 42V and 14V boards.

3.4 EXPERIMENTAL LAYOUT

The setup for the experiment is based on a standard Lundell automotive alternator-(14 V, 65/130 A) that is driven by a computer controlled variable-speed drive (13.4 KW). For the purpose of this experiment, the internal alternator field regulator and the full bridge rectifier were disabled. An external constant current power supply is instead connected to the field winding to supply the needed field current. Electronic loads, each with electrolytic capacitor banks (40000 µF, 50VDC) form the constant-voltage loads to represent the batteries of a real system. The SMR outputs are connected to the electronic loads. The capacitor banks are connected to absorb any ripple at the outputs. Current probes are connected to measure the currents in the buses and are connected to the oscilloscope through
the current amplifier. A four-channel oscilloscope is connected for monitoring and measurements of currents, voltages, and gate signals. All the equipment used for the experiment are listed in Table 3.2 and the picture of the complete setup for the experiment is illustrated in Figures 3.7 and 3.8.

![Experimental setup](image)

**Figure 3.7**  Experimental setup.
Figure 3.8  Lundell alternator and Motor drive system.
<table>
<thead>
<tr>
<th>#</th>
<th>EQUIPMENT</th>
<th>MODEL and SPECIFICATION</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>Lundell Alternator</td>
<td>14 V, 65/130 A</td>
</tr>
<tr>
<td>2</td>
<td>Alternator Drive motor</td>
<td>PACTORQ: SC756A-001-01; 230V, 3ph 18.0hp(13.4KW) (Pacific Scientific)</td>
</tr>
<tr>
<td>3</td>
<td>Switched-mode rectifier</td>
<td>42V/14V SMR boards</td>
</tr>
<tr>
<td>4</td>
<td>System DC Power Supply</td>
<td>HP 6632A, (0 – 20V, 0 –5A, 100W)</td>
</tr>
<tr>
<td>5</td>
<td>DC Power Supply (2 units)</td>
<td>Tektronix PS 280</td>
</tr>
<tr>
<td>6</td>
<td>Electronic Loads</td>
<td>DYNALOAD Model DLVP 50 – 300 – 3000A</td>
</tr>
<tr>
<td>7</td>
<td>Electronic Loads</td>
<td>HP 6050, 3 channel 60/120A, 600W Load per channel.</td>
</tr>
<tr>
<td>8</td>
<td>Current Probe (2 units)</td>
<td>Tektronix A6303</td>
</tr>
<tr>
<td>9</td>
<td>3-Channel current amplifier.</td>
<td>Tektronix TM503</td>
</tr>
<tr>
<td>10</td>
<td>Four- Channel Digital Oscilloscope</td>
<td>Tektronix TDS 754D 500MHz, 2GS/S</td>
</tr>
</tbody>
</table>

Table 3.2  Equipment used in the SMR system experiment.
3.5 CONCLUSION

The experimental setup and prototype design described in this chapter have been found to be suitable for evaluating the proposed technology. The measurements and analysis of results are discussed in the next chapter.
CHAPTER 4

EXPERIMENTAL RESULTS AND INTERPRETATION

4.1 INTRODUCTION

This chapter presents experimental results for the prototype dual output alternator system described in the previous chapter. First the machine parameters used and calculations of control handles, \( I_s \), \( h \), and \( r \) used for the experiment are introduced. Measurement procedures are stated, and the experimental results are then presented and analyzed. It is worth mentioning that another alternator has been introduced in the later part of this Chapter. Details in section 4.3.1.1.

4.2 PARAMETERS AND CONTROL HANDLE CALCULATIONS

The parameters for the Lundell alternator used for the calculation of the control handles are given as below:

\[ R_s = 0 \text{ (The actual value is } \sim 33 \, m\Omega, \text{ but is neglected to enable the simplified equations to be used)} \]

\[ L_s = 135 \, \mu H \text{ (The actual value is closer to } 105\mu H, \text{ but increased to compensate for neglected stator resistance)} \]

\[ k = 0.004 \text{ v-s/(rad)A} \]

\( f \), the electrical frequency of the alternator's voltage is computed using:

\[
f = \frac{P}{2} \frac{rpm}{60} \tag{4.1}
\]

where \( P \) is the number of poles and \( rpm \) is the alternator's speed in revolutions per minute(rpm):
$f$ = 180 Hz, 600 Hz for idle (1800 rpm) and cruising (6000 rpm) speeds respectively.

$i_f$, the field current for this alternator (named Alternator-A), is set not to exceed 3.6 amps.

In order to determine the three control handles to be used in the experiment, the control equations developed earlier in chapter 2 are recalled here:

\[ i_f = \frac{4L_2}{3k^2\omega} (V'_{i_{14}} + V'_{i_{42}}) \]  
(4.2)

\[ r = \frac{i_{i_{14}}^*}{i_{i_{14}}^* + i_{i_{42}}^*} \]  
(4.3)

\[ h = \frac{\pi ko i_f}{2\sqrt{2(rV_{i_{42}} + (1-r)V_{i_{14}})}} \]  
(4.4)

We intend to display the capabilities of the experimental apparatus and the operation of the control laws by operating at many combinations of $i_{i_{14}}$ and $i_{i_{42}}$. It is convenient to gather and present data in the form of the interrelation between $i_{i_{14}}$ and $i_{i_{42}}$ when the alternator field current and speed are fixed. To select a matrix of test points, it is convenient to first select a range of values for $i_{i_{42}}^*$ to be achieved under the condition $i_{i_{14}}^* = 0$. We have selected the following points:

\[ i_{i_{42}}^* \bigg|_{i_{i_{14}}^* = 0} = (31A, 19A, 13A, 9A, 5A, 3A). \]

The corresponding values of $i_f$ are then calculated from equation (4.2) at idle speed.

The corresponding field currents are:

\[ i_f = (3.6A, 2.81A, 2.33A, 1.94A, 1.45A, 1.12A). \]
For each value of $i_f$, $r$ is varied from 0 to 1 with step of 0.1.

thus

$$r = (0, 0.1, 0.2, 0.3, 0.4, 0.5, 0.6, 0.7, 0.8, 0.9, 1).$$

These values of $r$, $i_f$ are used to determine the values of corresponding control handle $h$ from equation (4.4). The calculated values of the control handles $i_f$, $r$ and $h$ are presented in Table 4.1. The control handle $h$ represents a fraction of the rectifier's fundamental period $T$. It therefore is limited to values between 0 and 1.

However we see from Table 4.1 that some values of the control handle $h$ corresponding to entirely reasonable values of $r$ and $i_f$ are greater than 1. Considering the origin of equation 4.4, it is apparent that the condition $h > 1$ occurs, particularly for large values of $i_f$ and low values of $r$. A command of $h > 1$ reflects a requested power that exceeds the (idealized) alternator capability. $h$ is restricted to the range $(0, 1)$, which will result in the load-matched power if it is achievable, and the maximum power possible in the desired output combination if load-matching is not achievable.

It was also found through experimentation and simulation that, particularly at high field currents, a lower $h$ than predicted by the idealized equations results in the best output characteristics. This results from approximations developing the idealized equations (such as neglecting stator resistance and voltage harmonics). To correct for this, the originally-determined values of $h$ are multiplied by 0.8, referred here as the correction factor, which we find gives us very close to the desired performance. The adjusted values of $h$ are presented in column 5 of Table

51
4.1. The calculations of the positive widths \( t(h) \) and \( t(r) \) for the gate signals of the MOSFETS are derived from circuit waveforms reproduced in Figure 4.1(B), (C).

![Waveforms Image](image)

**Figure 4.1** Circuit waveforms over a switching cycle. (A): voltage \( v_x \), (B): signal applied to the gates of 42V SMR board MOSFETs and (C): applied to the gates of 14V SMR board.

From this Figure:

\[
\begin{align*}
    t(h) &= T \left( 1 - h_T \right) \\
    t(r) &= T \left( 1 - h \right) \\
    \text{where} \ T &= \frac{1}{f_{sw}}
\end{align*}
\]

The switching frequency, \( f_{sw} = 100 \text{ kHz} \) in the prototype.

The results of the above calculations are presented in column 6 and 7 of Table 4.1. Similar calculations were performed to determine the values of the control handles and gating signals at cruising speed (6000 rpm). These results are presented in Table 4.2. As will be described later, some measurements were carried out with a second alternator, designated as Alternator-B. \( i_f = 4.2 \text{A} \) is the full-scale field current of alternator-B. (See section 4.3.1.1). At the higher speed, the load-matching \( V_x \) is unobtainable for many operating points, including all points at or near full field.
current. Even multiplying $h$ by 0.8 did not bring the values within range for many cases. In those instances, $h$ was set to 1. A new column "h restricted" presents the parameters actually used in the test.

<table>
<thead>
<tr>
<th>$i_{cl}$ (A$^{-1}$)</th>
<th>$i_r$ (A)</th>
<th>$h$</th>
<th>$h \times 0.8$</th>
<th>$t(h)$ (usec)</th>
<th>$t(r)$ (usec)</th>
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<td>3.60</td>
<td>0.1</td>
<td>1.015</td>
<td>0.812</td>
<td>1.88</td>
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<td></td>
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<td>0.877</td>
<td>0.702</td>
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<td></td>
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<td>6.13</td>
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Table 4.1 Computed values of control handles used for experiments at idle machine speed (1800 rpm).
### Table 4.2
Computed values of control handles used for experiments at cruising machine speed (6000 rpm).
4.3 Measurements

4.3.1 Simulation measurements

Prior to the actual experimental measurements, simulations of the control laws and actual circuit were performed using Matlab and Pspice to determine the performance of the circuit under certain values of the machine parameters. The characteristics of the Alternator/SMR system were first investigated analytically with Matlab, using the control equations developed in chapter 2. The simulations were performed for idle speed (1800 rpm), and cruising speed (6000 rpm), using the machine parameters specified in section 4.2. In the Pspice simulations, two values of stator resistance and inductance were considered, corresponding to actual and approximate machine characteristics. Simulations were done both with and without approximate device drop values. Scripts used for the Matlab simulations can be found in appendix B.

4.3.1.1 Simulations Results

Results from Matlab simulations are illustrated in Figure 4.2a, b. Results obtained from varying the control handles $r$ and $h$ at $i_r = 3.6A$ for idle machine speed are illustrated in Figure 4.2(a), and Figure 4.2(b) for cruising speed. Table 4.3 and Figure 4.3 contain the results for the Pspice simulations performed with and without approximate device drop values. See appendix C for the Pspice model used for the simulation.
Figure 4.2  Simulated Alternator output over control range with parameters in section 4.2. Hatched regions represent achievable operating points (a): at 1800 rpm and (b): at 6000 rpm.
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<th>No device drops</th>
<th>device drops</th>
<th>device drops</th>
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<td>$R_s = 0, L_s = 135\mu H$</td>
<td>$R_s = 0, L_s = 135\mu H$</td>
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<td>$i_{42}$</td>
<td>$i_{14}$</td>
<td>$i_{42}$</td>
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<td>(A)</td>
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Table 4.3 Pspice simulation results with and without device drops as the value of resistance and inductance varies. All runs for $i_f = 3.6A$ and 1800 rpm. ($R_s = 0, \ 33\,m\Omega : L_s = 135\mu H, 105\mu H$).

![Figure 4.3](image)

**Figure 4.3** Pspice simulation results show circuit performance with and without device drops. All runs for $i_f = 3.6A$ and 1800 rpm, with $L_s = 135\mu H$ and $R_s = 0$. 

57
4.3.1.2 Discussion on simulation results

The hatched regions of Figures 4.3a, b represent achievable operating points. These show that it is possible to obtain any desired combination of bus currents, within machine capability, by selecting the appropriate control handles \( r, h \) and \( i_f \). In the Pspice simulations, the effects of stator resistance and device drops were investigated. Two values of stator resistance and inductance were considered, corresponding to actual and approximate machine characteristics. (For simplicity, the control equations assume zero stator resistance. The machine inductance is increased slightly to model the impedance associated with the neglected resistance.) The results indicate that the simplification used to develop the control equations is acceptable. The effect of device drops can be seen in the Pspice simulation results illustrated Figure 4.3. It can be seen that one effect of the device drops is to reduce the achievable current at the 14V output as compared to the theoretical calculations. The impact on the 42V output is much lower.

4.3.2 Experimental measurements

The first step was the switching on of power supplies and ensuring that the correct voltages are applied to the circuit boards. The next step was carefully setting the values of control handles \( r \) and \( h \) by manually adjusting the provided knobs on the PWM controller and verifying the correct gating signals are obtained. At this point the alternator drive is started and the speed set to the appropriate value (e.g., 1800 rpm for idle). The field current is then adjusted to the corresponding value. The currents in the buses, measured by the current probes, are read from the oscilloscope and entered in a spreadsheet. For a set of control handles, continuous monitoring of the alternator speed is done to ensure the drive speed-loop properly regulates and about three to five sets measurements of currents are performed in
order to verify that the results are consistent (e.g. to eliminate the influence of bias
drift in the current probes.)

Data was recorded without any attempt to achieve a thermal steady state at each
operating condition. Operating temperature is not expected to have a large effect
on data obtained in these experiments. The principal impact of temperature is on
winding resistance. The field winding is regulated to a fixed current, so the only
effect of field winding resistance variation is on the field voltage required to
achieve the regulated temperature. Variations in stator resistance will impact
circuit operation, but these effects will be small; stator currents are principally
determined by induced and switched voltages and by inductance, none of which
are strong functions of temperature.

4.3.2.1 Problems encountered

It is worth mentioning that after the alternator's idle-speed performance
characteristics were measured for all field currents in Table 4.1 the machine stator
winding was accidentally destroyed by overloading in an unrelated experiment.
This resulted in the use of a second alternator. Thus experimental measurements
for the idle and cruising speeds were performed on the newly installed alternator
which, will henceforth, for easy identification, be referred as alternator-B. The
former will be referred as alternator-A. Alternator-B has similar characteristics to
alternator-A except that its full-scale field current is approximately 4.2A instead of 3.6A.

4.3.2.2 Experimental Results

Representative waveforms of voltage $V_x$ (phase-to-ground for one phase) and the gating signals as recorded by the oscilloscope are shown in Figure 4.4. The values of the currents measured in the buses are presented in Table 4.4. As mentioned earlier, these are the values for measurements performed on alternator-A at idle speed (1800 rpm). The second sets of results are for the measurements on alternator-B at idle speed. These are presented in Table 4.5 while alternator-B’s performance at cruising speed of 6000 rpm is presented in Table 4.6.

![Oscilloscope waveforms](image)

**Figure 4.4** Oscilloscope: Voltage and gating signal waveforms
## Table 4.4  Experimental result: Bus currents of alternator-A at idle (1800 rpm) speed.
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<th>$x 0.8 h$</th>
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<th>$t(\text{r})$ (usec)</th>
<th>$\text{14V bus A/div}$</th>
<th>$\text{42V bus A/div}$</th>
<th>Oscilosc (mV)</th>
<th>Oscilosc (mV)</th>
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**Table 4.5** Experimental results: Bus currents of alternator-B at idle (1800 rpm) speed.
Table 4.5 (contd). Experimental results: Bus currents of alternator-B at idle (1800 rpm) speed.
### Table 4.6  Experimental results: Bus currents of alternator-B at cruising (6000 rpm) speed.
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Table 4.6(contd). Experimental results: Bus currents of alternator-B at cruising (6000 rpm) speed.
Figure 4.5 Experimental results: Alternator-A output currents (1800 rpm).
Figure 4.6 Experimental results: Alternator-B output currents (1800 rpm).
Figure 4.7 Experimental Results: Alternator-B output currents (6000 rpm).
4.3.2.3 Discussion of experimental results

Here we analyze the experimental results. Figure 4.5 (Alternator-A) and Figure 4.6 (Alternator-B) show experimental results for idle speed operation of the prototype system over the full output command range. Similar results are illustrated for cruising speed operation of Alternator-B in Figure 4.7. These results are consistent with the predicted performance of the system. They demonstrate the ability to achieve the desired system functionality.

Interestingly the results presented in Figure 4.6 and Figure 4.7 are for Alternator B, a different machine whose parameters may not be exactly the same with the original Alternator-A whose parameters were used in determining the control handles. Performance of Alternators A and B are very close except that full field current of Alternator-B is 4.2 A, rather than 3.6 A for Alternator-A. These results both show consistency with expectations, therefore indicating that the SMR technique is easily adaptable to different types of alternators without serious sensitivity to the machine parameters.

A careful observation of the graphs at idle speed shows that zero current at the 42V bus was not obtainable, especially at higher 14V bus currents. This is due to the drift in the regulation voltage of the 42V electronic load as zero current is approached, and does not reflect a limitation of the fundamental SMR technology. Another observation is the drop of output currents at the 42V bus under some conditions when the current to the 14V bus goes to zero. Normally one expects a higher current available to deliver to the 42V bus since at this instance all the switches are switched off. This may be attributed to the limitation of the model used to make the output predictions.
4.4 CONCLUSION

The experimental results obtained are very much consistent with the results obtained from simulations. These results demonstrate the feasibility and high performance of the proposed approach. Thus, the dual-output alternator overcomes the limitations prevailing with other designs as mentioned in chapter 1. It has been shown that to a reasonable approximation the SMR approach enables the full load-matched power capability of the alternator machine to be achieved at idle speed, with power delivered to the two outputs in any desired combination.
CHAPTER 5
ECONOMY AND POLICY CONSIDERATIONS.

5.1 INTRODUCTION

Most light vehicle owners are not aware of the predicament presented by the present 14-Volt automobile power supply. Many people assume that all the features needed in a vehicle are already provided, and that any additional features may be easily added. Why, then, is there serious discussion of a revolution in automotive electrical system? Why are 42-Volt and dual 42/14-Volt systems widely studied and discussed? The objective in this chapter is to discuss the issues pertaining to the demand for a dual/higher voltage automobile electrical system, and examine the driving forces, benefits, and concerns in the implementation of a higher voltage automotive electrical system.

5.2 PROBLEMS WITH 14-VOLT SYSTEM.

As briefly mentioned in Chapter 1, the present 14-Volt automobile electrical system was introduced in the early 1950's when the then 6-Volt system power capability was outstretched by new features introduced into the vehicle. While electrical power usage only increased at an annual rate of 2% until 1970, it has however increased at the rate of about 6% for the past 30 years [w4]. Projections of
the estimated electrical power requirement in automobile have been made by
various groups as illustrated in Figure 5.1. Due to the low annual rate of growth
prior to 1960, we will only look at a 50-year projections from 1960 to 2010. Graph
“A” in Figure 5.1 is the projection made by Delphi Automotive Systems. [w4]. It
depicts the generator peak power of an average passenger vehicle. It shows a peak
power of 500 watts in 1960 and an estimate of 5000 watts in the year 2010. Graph
“B” illustrates projected trends in automotive electrical power as cited in [I]. It also
shows an exponential increase in power from 1970 to 2005. Extrapolating the
graph to year 2010 gives us the power demand of about 4900 watts. Graph “C” is
a different projection from graphs “A” and “B”. It gives the picture of the electric
power consumption for different models of Renault vehicles from 1985 to 2005. As
illustrated in the figure, the power consumption for Renault 5 vehicle was 500-
watts in 1985, while the Renault Espace is anticipated to guzzle 5000 watts of
power by the year 2005. The wide range of consumption, all for model year 2001
Reneault vehicles, shows that the spread among vehicles in any model year is of
the order of the trend we try to present in this graph. The existence of this spread
is easy to understand, but it complicates the challenge to anyone trying to graph
the time trend; determining which model in one year corresponds to a given
model in another year. On the other hand, graph “D” shows the installed electrical
power in Volvo cars from 1960 to 1999 [19]. This also gives the picture of increase
in installed electrical power within the period of consideration.
In summary, it is appropriate to say that though these power projections made by different groups may vary, vehicles and it is apparent that there is a demand for a higher automotive electrical power over time. The power figures indicated are going beyond the capability of the present 14 Volts system.

**Figure 5.1.** Projected trends in automotive electrical system presented by various groups.

The vehicle manufacturer would desire to keep on adding new functionality, however the present automobile power supply is not adequate to supply all the add ons. At some point, any attempt to go on introducing additional features impairs the overall efficiency and is not economically reasonable. Significant
information being published or presented today strongly suggests that the growth in electrical power required for the future vehicles will be even greater than in the past. As a result, a proposal by the automotive community and MIT was set forth to raise the voltage level of the next generation automobile from the present 14V to 42V volts[6].

5.2.1 Why 42 Volts?
If increasing the voltage will be beneficial, one may ask, why settle for 42 volts and not higher? The limiting factor is the desire to maintain, for the next-generation automotive electrical system, an open wiring system, consistent with the present 14V system. To retain an open system, it is necessary to ensure that any high voltage adopted must be sufficiently low enough to guarantee personal safety of anyone who comes in contact with the system without the need for special protection. According to SAE standard J2232, [cited in 6], there is no need of protection against electric shock if the voltage, including periodic ripple does not exceed 65 volts dc. Hence 42 Volts was agreed upon for the next generation automobile electrical system.
5.3 **Driving Forces and Anticipated Benefits**

The driving forces for a higher voltage may vary by geographic market, however the major factors include:

- Higher electrical loads and new features
- Improving fuel economy
- Emissions reduction

5.3.1 *Higher electrical loads*

The power requirements for the future vehicles can be achieved by either increasing the current capacity or the voltage of electrical systems. Increasing the current will mean requiring thicker wiring which will add to the weight of the vehicle as well as difficulty in assembly. Higher current will also increase the cost of in semiconductor switches and other electrical components. A high-voltage system, by reducing current requirements at a given power, makes it possible to build some new electrically actuated features, and reduces the cost of others.

Also, the need for 42V system is being brought about by the need to improve fuel efficiency. An example is the conventional engine-powered hydraulic steering system which has a belt-driven pump that generates the hydraulic pressure. This pump runs continuously, placing a constant power drain on the engine. It is estimated that by converting to electric power steering system, in which power is applied only when the steering wheel is rotated, the fuel consumption is improved by 1 to 3%, depending on vehicle class [6]. Electric power steering is being adopted
today at 14V, but 42V applications will be more efficient. In some weight classes 42V will be required before electric power steering can be considered. Conversion of other engine-powered functionalities to electric-powered will greatly improve the fuel consumption. Table 5.1. shows some current functions to benefit from the 42-Volt system.[w4].

<table>
<thead>
<tr>
<th>Technology</th>
<th>Benefits</th>
</tr>
</thead>
<tbody>
<tr>
<td>Electric power steering</td>
<td>More power; improved fuel economy</td>
</tr>
<tr>
<td>Electric brakes</td>
<td>Redundant power supplies</td>
</tr>
<tr>
<td>Power windows, power seats, and power lift gates</td>
<td>Reduced size and mass of motors; more efficient operation</td>
</tr>
<tr>
<td>Heated catalytic converter</td>
<td>Lower emissions; quicker light-off time of converter</td>
</tr>
<tr>
<td>HVAC blower motors and cooling fans</td>
<td>Greater efficiency; smaller/lighter units; flexible packaging</td>
</tr>
<tr>
<td>Mobile multimedia</td>
<td>More power available for video, cellular phones, navigation systems, audio amplifiers, and fax</td>
</tr>
<tr>
<td>Water pumps</td>
<td>Improved efficiency; longer service life</td>
</tr>
<tr>
<td>Select engine management system components (such as EGR valves, ignition systems, and control actuators)</td>
<td>Reduced size and mass; increased performance</td>
</tr>
<tr>
<td>Fuel pumps</td>
<td>Reduced size and mass</td>
</tr>
<tr>
<td>Heated seats</td>
<td>Faster heating; more efficient operation; increased power</td>
</tr>
</tbody>
</table>

Table 5.1. Current Technologies to benefit from the 42-Volt system. [w4].

Another driving force for the 42V system is the need to introduce new high-tech automotive features that the present 14-Volt system could not easily support. These features include an integrated starter/generator, electronic brakes, steer by
wire and electromagnetic valve actuation. Not only do they offer comfort and convenience for the consumers, but they form the foundation for future automotive capabilities, such as adaptive cruise control and collision avoidance.

<table>
<thead>
<tr>
<th>Technology</th>
<th>Benefits</th>
</tr>
</thead>
<tbody>
<tr>
<td>Ride control systems</td>
<td>Improved ride, handling, and vehicle stability</td>
</tr>
<tr>
<td>Brake-by-wire</td>
<td>Improved vehicle parking and brake performance</td>
</tr>
<tr>
<td>Steer-by-wire</td>
<td>Enhanced performance; improved packaging; improved passive and active safety</td>
</tr>
<tr>
<td>Electromagnetic valve control</td>
<td>Lower emissions; optimum power; individual cylinder control; lower cost</td>
</tr>
<tr>
<td>Integrated starter-generator</td>
<td>Faster starts; quicker charging; design flexibility; low noise and vibration; improved fuel economy</td>
</tr>
</tbody>
</table>

Table 5.2. Future Technologies enabled by 42-Volt systems[w4].

5.3.2 Improved fuel economy/Emissions reduction

Concerns raised by environmentalists, Non-Government Organizations (NGO) and nations around the world requiring a cleaner environment through mitigation of carbon dioxide (CO₂) in atmosphere has also been a driving force for the 42V system. The transport sector is responsible for the rising share of CO₂ emission from fossil fuel use. Figure 5.3 shows that the sector emitted 816 million metric tons of CO₂ in 1990[18]. Road vehicle fuel consumption accounts for roughly 75% of this transport this emission as seen in Figure 5.3. As already mentioned in the preceding section, the new functionalities added offer valuable opportunities to increase engine efficiency and benefits in such important areas as fuel economy emissions, and passenger comfort. Any technology that will economically reduce CO₂ emissions will be highly embraced.
Figure 5.2. Carbon Emissions by Sector. [18].

Figure 5.3. Carbon Emissions from Transport Oil use. [18].
Chapter 5: Economy And Policy Considerations

5.4 CONCERNS TO 42V/14V SYSTEM ADOPTION

At present there is absence of a common or global standard. This makes the structure of the 42/14V electrical system very fluid, and it is not clear what will be the industry standard. This poses a challenging question for the OEMs and suppliers, whether the higher voltage system will be adopted through a dual-voltage/dual-battery architecture or through a single voltage/single battery system.

Another concern is the associated complexity that comes with the 42/14V system. It is clear that the basic cost of the new 42/14V electrical system supplying only today's vehicle electrical loads will be higher than that of a conventional 14V system. As a result the benefits arise only as new features, enabled by the new power system, are added. OEMs and suppliers are trying to avoid the initial risks and costs involved in adopting the new approach.

Another concern that needs to be seriously considered, despite the benefits of the 42-Volt system as earlier mentioned, is the competition that may be posed from improved 14 V systems over the next 10 years.

5.5 CONCLUSION

Future vehicles will be driven by the needs to reduce exhaust emissions, provide better fuel economy, improve safety and increase comfort and convenience. Most of these features can be obtained by electrical/electronic control of components that have historically been hydraulically or pneumatically controlled. This move to 42 Volts could open the markets for some automotive suppliers. Opportunities in
42 Volts are good for makers of batteries, voltage regulators, motor controllers, power semiconductors, converters and other types of load control devices. The automotive community must prepare for the introduction of dual/higher voltage vehicles by establishing an infrastructure to accommodate the needs of the system, including for example service tools, service parts and aftermarket sales groups. Education and training will need to be provided for the vehicle owner and service technicians.
CHAPTER 6
SUMMARY AND CONCLUSIONS

6.1 THESIS SUMMARY
The push to introduce dual-voltage (42V/14V) automotive electrical systems necessitates power generation solutions capable of supplying power to multiple outputs. A number of approaches for implementing dual-voltage electrical systems have been proposed, but most suffer from severe cost or performance limitations. This thesis explores the design of alternators incorporating dual-output switched-mode rectifiers for the implementation of 42V/14V dual voltage electrical systems for the next generation automobile. This approach enables the full load-matched power capability of an alternator machine to be achieved, with power delivered to the two outputs in any desired combination. The work presented here encompasses the development of control laws and design rules for dual-output alternators with switched-mode rectifiers. Finally, it includes the design, construction and experimental evaluation of a prototype dual-output alternator system.

The thesis is divided into six chapters. Chapter one presents the trends in automotive electrical system and shows the need for introduction of a higher voltage automotive electrical system. Discussions of the operation of the proposed system are also presented in chapter one. Chapter two presents the detailed operation and development of control laws for dual-output alternator/SMR systems and in addition, discusses the design tradeoffs. Chapter three presents the
design and construction of a prototype system and the experimental setup. Chapter four presents the simulations and experimental measurements. The final part of this chapter presents the interpretation of the results. Chapter five introduces the economic benefits of the higher electrical automotive system while the summary, conclusion and suggestions for further work are presented in chapter six.

6.2 Thesis Conclusions

The objectives of this thesis have been met. The first objective is the development of control laws for the proposed dual-output rectifier structure. Varying the control handles according to the developed equations, power can be delivered to the two outputs over a wide range of loading scenarios.

The second objective met in this thesis is the development of design guidelines for SMR-based dual-output alternators. It has been found that the system can be designed to deliver full load-matched power to either output under all conditions, but this imposes a heavy cost penalty on the switched-mode rectifier. Alternatively, the system can be designed such that full load-matched power is always deliverable to the 42V bus, load-matched power is deliverable to the 14V at idle, and increased (but less than load-matched) power is deliverable to the 14V bus at higher speeds. Such a design approach places much more modest requirements on the switched mode rectifier, and is likely to be much more closely
matched to actual load requirements. This thesis has quantified the tradeoffs between output power capability and rectifier design.

The final objective achieved is the development of a prototype 42V/14V dual-output alternator/SMR system. The prototype system has been utilized to validate the design and control laws proposed here, and demonstrated the feasibility of the proposed approach. A good match between theoretical, simulation and experimental results has been found.

6.3 RECOMMENDATIONS FOR FUTURE WORK.

The experimental measurements were performed with the desired values of control handles \( i_g, r \) and \( h \) being set manually. Future work should consider the development of a closed loop feedback control. The measurements were performed with only two machine speeds (1800 rpm and 6000 rpm). It will be worthwhile to establish the performance of the SMR technology under various machine speeds. The simplified model used here predicted more power than achieved in the actual machine. Future work should consider using an improved machine model to develop the control commands.

Finally, the prototype system developed here is suitable for bench-top testing, but is not designed for underhood automotive environmental conditions. Design and packaging of the switched mode rectifier for the underhood environment (and perhaps its integration into the alternator package) should be addressed in the future.
APPENDIX

MATLAB SCRIPT USED TO GENERATE FIGURE 2.5 IN CHAPTER 2

% DAWINDING1.M
% This file computes the maximum output currents and rms device currents in a
% Dual-Output Alternator. The dual-output alternator is of the dual-rectified type
% proposed in the Convergence 00 paper by Perreault, et. Al..
% (Device calculations are for the version with a diode bridge followed by a
% single switching stage.) The values are calculated for different winding
% numbers on the alternator (i.e. a few turns of large wire vs. many of small
% wire.)
% This allows a comparison of tradeoffs between output power capability on
% both busses and device ratings. The base values are those for a conventional 14 V
% alternator.
% Calculations from DJP's analysis 9/5/01.
% Written by D Perreault 9/6/01.

% Alternator parameters
Lbase = 135e-06; % base 1-n synchronous inductance value in
Henries.
kbase = 0.004;  % base machine (back emf) constant in V-s/rad-
A
ifield = 3.6;   % field current in amps (max field 3.6A)
w = (p/2) * wm_rpm*2*pi/60; % alternator electrical frequency in rad/sec
1800, cruise = 6000 % alternator mechanical speed in rpm idle =
p = 12;        % number of poles
V42 = 44;       % 42 volt bus voltage (includes diode drops)
V14 = 16;       % 14 volt bus voltage (includes diode drops)

% vectors to store data
I42_vec = []; % max 42 V current with I14 = 0
I14_vec = []; % max 14 V current with I42 = 0
Iq1rms_vec = []; % max rms shunt FET current at I14 = 0
Iq2rms_vec = []; % max rms "series" FET current at I42 = 0
n_vec = []; % normalized number of turns loop through
% winding range. n=1 is for a conventional
% alternator

for n = 0.25:0.05:1.5,
Ls = Lbase*n*n; % inductance scales with n^2

85
k = kbase*n;

n_vec [n_vec ; n] ;
% compute data for max current to 42 V bus, zero current to 14 V bus

h = pi*k*w*ifeild/ (2*sqr(2) *V42);
if h > 1
   h = 1
end;
Ih = (3/pi)*sqr(((k*k*w*w*ifeild*ifieild) - 4*h*h*V42*V42/(pi*pi))/ (w*w*Ls*Ls));
I42 = h*Ih;
Iq1rms = sqrt (1 - h) * Ih;

% compute data for max current to 14 V bus, zero current to 42 V bus

h = pi*k*w*ifeild/ (2*sqr(2) *V14);
if h > 1
   h = 1
end;
Ih = (3/pi)*sqr(((k*k*w*w*ifeild*ifieild) - 4*h*h*V14*V14/(pi*pi))/ (w*w*Ls*Ls));
I14 = h*Ih;
Iq2rms = sqrt (1 - h) * Ih;

% put data into vectors
I42_vec = [I42_vec ; real (I42) ];
I14_vec = [I14_vec ; real (I14) ];
Iq1rms_vec = [Iq1rms_vec ; real (Iq1rms) ];
Iq2rms_vec = [Iq2rms_vec ; real (Iq2rms) ];

end; % loop through number of turns

hold off
plot (n_vec , I14_vec , 'blue' )
hold on
plot (n_vec , I42_vec , 'red' )
plot (n_vec , Iq1rms_vec , 'green' )
plot(n_vec , Iq2rms_vec , 'black')
MATLAB Scripts Used to Generate Figures 4.2A.

% ALTERNATIVE CONTROL: filename: gimba511.m (2001)
% script to calculate power of dual output alternators vs. operating point(Rs = 0, k = 0.004, 1800 rpm)
% The constants in this script are based on Erica Salinas' alternator params

wm_rpm = 1800; % alternator mechanical speed in rpm
p = 12; % number of alternator poles
k = 0.004 % machine constant V-s/(rad)
Rsperk2 = 2062.5; % resistance coefficient = R/k~2
Lsperk2 = 6.5625; % inductance coefficient = L/k~2
%ifield = 3.6; % field current in amps
V42 = 42; % voltage of the 42V bus
V14 = 14; % voltage of the 14V bus

%Rs = Rsperk2*k^k % resistance(scales with square of machine const)
Ls = Lsperk2*k^k % inductance(scales with square of machine const), H
w = (p/2)*wm_rpm*2*pi/60; % alternator electrical freq. in rad/sec
d1_vec = 0; % vectors to store data
d2_vec = 0;
i42_vec = 0;
i14_vec = 0;
i42_vec = 0;
i14_vec = 0;
iifield = 0;
lx_vec = 0;
for i42 = 0:1:30,
    for i14 = 0:2:85,
c = i42 + i14*(29.3306)/(83.3061);
if c < 30
    i42 = i42;
i14 = i14;
else
    i42 = 0;
i14 = 0;
end;
M = sqrt((4*Ls*(i42*V42 + i14*V14))/(3*k*k*w)); % eqn. for field current in amps
if M > 3.6
    if field = 3.6; % setting the limit of field current to 3.6 amps max.
else
    if field = M;
end; % end of if (M)
r = i42/(i42 + i14);

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\[ h = \frac{(\pi k w \text{ifield})}{\sqrt{(8)(r V42 + (1-r)V14)}}; \]
\[ d1 = 1 - h; \]
\[ d2 = d1 + (hr); \]

\[ \% Vx = h^r V42 + h^{(1-r)} V14; \]
\[ Vx = (d2-d1)V42 + (1-d2)V14; \quad \% \text{average } Vx \]

\[ m1 = (k^*k^*w^*w^\text{ifield}^*\text{ifield}) - (4*Vx*Vx)/(\pi^*\pi); \]
\[ \% m2 = \sqrt{((w^*w^*Ls^*Ls^*m1) + (Rs^*Rs^*k^*k^*w^*w^\text{ifield}^*\text{ifield}))}; \]
\[ \text{if } m1 > 0 \]

\[ \% Ix = ((3/\pi)*m1)/((2*Vx*Rs/\pi) + m2); \quad \% \text{Ix with source resistance.} \]
\[ Ix = ((3/\pi)*\sqrt{(m1/(w^*w^*Ls^*Ls))}); \quad \% \text{Ix without source resistance.} \]
\[ \text{else } Ix = 0; \]
\[ \text{end; } \% \text{if} \]
\[ \text{if } Ix < 20.3061 \]
\[ I42 = Ix^*(d2-d1); \]
\[ I14 = Ix^*(1-d2); \]
\[ \text{else} \]
\[ I42 = 0; \]
\[ I14 = 0; \]
\[ \text{end; } \% \text{2nd if} \]

\[ d1\_vec = [d1\_vec ; d1]; \]
\[ d2\_vec = [d2\_vec ; d2]; \]
\[ i42\_vec = [i42\_vec ; i42]; \]
\[ i14\_vec = [i14\_vec ; i14]; \]
\[ I42\_vec = [I42\_vec ; I42]; \]
\[ I14\_vec = [I14\_vec ; I14]; \]
\[ \text{ifield\_vec = [ifield\_vec ; ifield];} \]
\[ \text{Ix\_vec = [Ix\_vec ; Ix];} \]
\[ X=I14\_vec; \]
\[ Y=I42\_vec; \]
\[ \% \text{RESULTS} = [i14,i42,ifield,Ix]; \]
\[ \text{Results} = [d1\_vec,d2\_vec,Ix\_vec,X,Y,Ix\_vec]; \]

\[ \text{end; } \% \text{for i14} \]
\[ \text{end; } \% \text{for i42} \]
\[ \text{plot(X, Y, 'x');} \]
\[ \% \text{axis([0 90 0 30])} \]
\[ \text{xlabel('I14 (A)');} \]
\[ \text{ylabel('I42 (A)');} \]
Output currents at \( L_s=11.667 \mu H; \) \( R_s=3.7 \text{m-ohm,} \) \( k=0.004, \) \( \omega_m=1800 \text{rpm}, \) field current = optimized)

\( d_1 = 0.005:0.99 \)

\( d_2 = d_1:0.005:0.99 \)

\( d_1,d_2 \) combinations for \( I_x < 84.31 \text{ amperes} \)
MATLAB Scripts used to generate Figure 4.2b.

```matlab
% ALTERNATIVE CONTROL. (Rs = 0, k=0.004, w=6000 rpm) filename: gimba514b.m (2001)
% script to calculate power of dual output alternators vs. operating point
% The constants in this script are based on Erica Salinas' alternator params

wm_rpm = 6000; % alternator mechanical speed in rpm
p = 12; % number of alternator poles
k = 0.004 % machine constant V-s/(rad)
Rsperk2 = 2062.5; % resistance coefficient = R/k^2
Lsperk2 = 6.5625; % inductance coefficient = L/k^2
ifield = 3.6; % field current in amps
V42 = 43; % voltage of the 42V bus + 1 volt diode
V14 = 15; % voltage of the 14V bus + 1 volt diode

Ls = 135/1000000
w = (p/2)*wm_rpm*2*pi/60; % alternator electrical freq. in rad/sec
d1_vec = 0; % vectors to store data
d2_vec = 0;
i42_vec = 0;
i14_vec = 0;
i42_vec = 0;
i14_vec = 0;
ifield_vec = 0;
Ix_vec = 0;
for i14 = 0:5:350,
    for i42 = 0:5:140,
        M = sqrt((4*Ls*(i42*V42 + i14*V14))/(3*k*k*w)); % eqn. for field current in amps
        if M > 3.6
            ifield = 3.6; % setting the limit of field current to 3.6 amps max.
        else
            ifield = M;
        end; % end of if (M)
        r = i42/(i42 + i14);
        h = (pi*k*w*ifield)/(sqrt(8)*(r*V42 + (1-r)*V14));
        if h < 1
            d1 = 1 - h;
            d2 = d1 + (h*r);
        else
            d1 = 0;
            d2 = r;
        end
    end
end
```
\[
\begin{align*}
\% V_x &= h^* r^* V_{42} + h^*(1-r)^* V_{14}; \\
V_x &= (d_2-d_1)^* V_{42} + (1-d_2)^* V_{14}; \quad \text{%average Vx}
\end{align*}
\]

\[
m_1 = (k^* k^* w^* w^* \text{ifield}^* \text{ifield}) - (4^* V_x^* V_x)/(\pi^* \pi);
\]
\[
\% m_2 = \sqrt{((w^* w^* L_s^* L_s)^* m_1) + (R_s R_s k^* k^* w^* w^* \text{ifield}^* \text{ifield})};
\]
if \( m_1 > 0 \)

\[
\% I_x = ((3/\pi)^* m_1)/((2^* V_x^* R_s/\pi) + m_2); \% I_x with source resistance.
\]
\[
I_x = ((3/\pi)^* \sqrt{m_1/(w^* w^* L_s^* L_s)}); \% I_x without source resistance.
\]
else
\[
I_x = ((3/\pi)^* \sqrt{m_1/(w^* w^* L_s^* L_s)}); \% I_x = 0;
\]
end; % if
if \( I_x < 130.4 \)
\[
I_{42} = I_x^*(d_2-d_1);
\]
\[
I_{14} = I_x^*(1-d_2);
\]
else
\[
I_{42} = 0;
\]
\[
I_{14} = 0;
\]
end; % 2nd if

\[
d_{1\_vec} = [d_1\_vec ; d_1];
\]
\[
d_{2\_vec} = [d_2\_vec ; d_2];
\]
\[
i_{42\_vec} = [i_{42}\_vec ; i_{42}];
\]
\[
i_{14\_vec} = [i_{14}\_vec ; i_{14}];
\]
\[
I_{42\_vec} = [I_{42}\_vec ; I_{42}];
\]
\[
I_{14\_vec} = [I_{14}\_vec ; I_{14}];
\]
\[
\text{ifield\_vec} = [\text{ifield\_vec} ; \text{ifield}];
\]
\[
I_x\_vec = [I_x\_vec ; I_x];
\]
\[
X=I_{14}\_vec;
\]
\[
Y=I_{42}\_vec;
\]
\[
\text{Results} = [i_{14}\_vec,i_{42}\_vec,d_{1\_vec},d_{2\_vec},\text{ifield\_vec},X,Y,I_x\_vec];
\]
end; % for i14
end; % for i42
plot(X, Y, 'x');
xlabel('I_{14} (A)');
ylabel('I_{42} (A)');
title('Output currents at L_s=135 \mu H; R_s=0 m-ohm, k=0.004, w_m=6000rpm, field current <= 3.6 amps')
text(50, 70, '\it\{i_{14} = 0:5:350\}')
text(50, 75, '\it\{i_{42} = 0:5:140\}')
text(50, 80, '\it(d_1,d_2 combinations for I_x < 130.4 amperes)')

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Appendix C

Figures C1 (a), (b) show the Pspice schematic diagrams for the SMR circuit simulations in order to observe its performance with and without approximate device drops. The three ac sources at the left in the figures represent the internal voltage of the alternator. The custom block immediately following, labeled RL_3PH, represents the resistive and (coupled) inductive impedance of the alternator. The N-channel MOSFET (IXFN230N10) and Schottky diode (DSS2x41-01A) are not available in the Pspice component library, thus an ideal diode, with a 0.4 Volt dc source and a 10 mΩ resistor are connected in series to give approximate characteristics of a non ideal Schottky diode (DSS2x41-01A). In a similar manner a switch (s1) in series with a resistance constitutes a non-ideal N-channel MOSFET (IXFN230N10). Its body diode is modeled as an ideal diode in series with a 1-volt dc source. For the simulations without approximate device drops, device and wire resistances were set to zero while all the dc voltages were dropped to 1 mV.
Figure C1: Pspice schematic diagrams for the SMR circuit simulations, (a) with device drops and (b) without drops
REFERENCES


References


ADDITIONAL SOURCES
In addition to the references cited, the following sites were consulted: (January 2003).

[w1] Emerging Opportunities in Automotive Power Electronics. (www.darnell.com/services/01-42v.stm)

