## AN ABSTRACT OF THE THESIS OF

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(1) René Spée

The advent of superior power semiconductor devices and converter topologies has renewed interest in ac drive systems. Although considerable research efforts have gone into improving power electronic converter devices and topologies, very little has been reported on the overall performance optimization of induction machine drive systems. The report of the research work presented in this thesis is an endeavor in that direction where enhancement in the system performance is achieved through optimization of the overall system.

The Brushless Doubly-Fed Machine (BDFM) can reduce the drive system cost and also retain the robustness of a cage rotor induction machine. Proof-of-concept prototypes have been used in the laboratory for investigation of BDFM operating modes. These prototypes, though providing insight into the operation of the BDFM were far from optimum. Thus, design procedures for optimizing the machine design needed to be developed. The optimized machine can then be integrated into an optimized system
by using a realistic, application-dependent converter selection scheme.
As mentioned earlier, recent developments in power semiconductor and converter technologies have led to a proliferation of circuit topologies and their modifications with sometimes contradictory performance claims. Consequently, for the non-specialist application engineer designing the BDFM system, this can often lead to uncertainty which potentially can result in non-optimum converter selection. An extensive and comprehensive literature review of presently available converter topologies is presented with a detailed comparative evaluation. Converter selection criteria, as applied to the BDFM, are also discussed in detail. These provide sufficient guidelines for selecting an optimum technology for a given application.

Stator design optimization, as discussed in this thesis uses a design parameter search algorithm and the BDFM steady state d-q model. Projected performance of an optimized stator BDFM design is compared with simulation results of present and past laboratory machines. An optimized design with good overall performance is presented. It is also shown that this optimization scheme lowers the power rating requirement of the converter.

Thus, for a given rotor structure and its parameters, the converter comparison and selection scheme for BDFM applications along with the optimization scheme for the stator design can lead to an overall optimized system with stator losses, reduced power converter rating and thus lower the initial investment and operational costs.
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by
Shibashis Bhowmik
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## APPROVED:

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# Assistant Professor of Electrical and Computer Engineering in charge of major 

## Redacted for Privacy

Head of Department of Electrical and Computer Engineering

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Date of thesis presentation September 10, 1992

Typed by Shibashis Bhowmik

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# SYSTEM OPTIMIZATION STUDIES RELATED TO <br> STATOR DESIGN AND AC/AC CONVERTER SELECTION FOR BRUSHLESS DOUBLY-FED MACHINES 

## 1. INTRODUCTION

### 1.1. Background

Recent developments in power semiconductor devices and converter topologies have led to increased research activities and applications of ac drive systems. A few notable applications of power electronic drive systems are adjustable speed drives (ASD), variable speed generation (VSG) schemes, high performance position controllers and commercial and residential appliances. Recently, new and more efficient converter topologies have been introduced, made possible by new devices and components as well as novel system control strategies. An emphasis is placed on the choice and usage of the optimum converter topology that provides the best system efficiency. At the same time, the topology should be capable of incorporating drive control schemes and maintain robustness and reliability. All of these features should be available at an economical and competitive price.

The most common ac ASD systems in industrial applications use a cage rotor induction motor (IM) with its stator excited by a variable-frequency, variable-voltage electronic power converter as shown in Fig. 1.1. Although the cage rotor is robust and cheap, the power converter is much more costly than the motor itself; since it has to process and provide for both output and losses. Moreover, for precise speed control an expensive speed feedback control loop is needed. A modification of this basic scheme
used in the industry, is obtained by connecting a wound rotor three phase IM in a doubly fed mode as shown in Fig. 1.2. Although this scheme provides for precise synchronous operation with a smaller, less expensive power converter, it losses the ruggedness of a cage rotor IM.


Figure 1.1 Conventional IM adjustable speed drive.


Figure 1.2 Wound rotor IM adjustable speed drive.

Conceptually, the Brushless Doubly Fed Machine (BDFM) is an equivalent of a wound rotor IM with both the power and control windings on the stator and a modified cage structure on the rotor. The schematic system configuration is shown in Fig. 1.3. The BDFM provides for the following advantages :
(i) robustness of a cage rotor IM and cost advantages over a wound rotor induction motor ;
(ii) reduction of the power converter rating; and
(iii) synchronous operation and thus open-loop speed control.

At Oregon State University prototype BDFM's have been designed with a 6 and 2-pole stator winding structure. Although providing insight into the operational characteristics, the prototype machines were not capable of achieving the desired performance. Most of these shortcomings are attributed to non-optimum stator and rotor designs, which result in poor efficiency and low torque production. This thesis discusses the method used to optimize the stator design for the 6 and 2pole machine using the steady state $\mathrm{d}-\mathrm{q}$ model developed by Li [9]. It also discusses the ac/ac converter topologies available and provides guidelines for selection of a converter topology for the BDFM in various applications. Research on the design of an optimized rotor will


Figure 1.3 Schemetic representation of the BDFM drive system. be reported in a separate thesis [12].

### 1.2. Literature Review

The need for a "single frame" cascade induction motor to reduce cost and improve performance over ordinary cascaded induction motors has long been established. During the beginning of the century, Hunt [1] first showed that "single frame" cascaded induction motors were feasible through ingenious motor design. Creedy [2] improved upon this basic configuration and proposed a 6 and 2-pole machine along with methods to design the rotor.

It was only again in the 1970's that Broadway [3] proposed a novel rotor structure now known as the "Broadway Rotor". It is based on the simple squirrel cage rotor structure with the associated simplicity and robustness. The rotor is modified to support the currents induced by the two stator windings, which are excited by voltages of different frequencies. However, in the synchronous mode of operation, where both the stator windings are excited, Broadway restricted his studies to fixed speed generation with dc excitation on the control winding.

Smith [4] was the first to investigate the synchronous behavior of the self-cascaded induction machine over a wide speed range, using a simplified steady-state equivalent circuit of two separate induction motors connected back to back. With the advent of power electronic converters, Kusko [5] and Shibata [6] developed slip-power recovery schemes for this type of machine. The equivalent circuits used for their analyses were essentially the same as those seen in the earlier papers.

A common feature of all the previous investigations is the underlying assumption that this type of machine can be represented by two magnetically separate, but electrically
connected wound rotor motors of different pole numbers, mounted on the same shaft. Moreover, most prior work has been limited to steady state analysis. Investigations on the BDFM at OSU have shown that the simplifying assumptions are appropriate for conceptual understanding of the BDFM operating principles, but are inadequate for detailed machine and drive system design [7].

Wallace et al [7] developed a detailed state variable machine design model to simulate machine performance for all operating conditions. The model separately represents each of the stator coils and nested rotor loops as well as the interaction among them. Electromagnetic field analysis carried out by Alexander [8] helped provide insight into the operation of the BDFM. Even though the detailed model and the electromagnetic field analysis are helpful tools, a reduced order dynamic model is needed to devise control strategies for BDFM drive systems and for overall system optimization. Li [9] developed the dynamic d-q model and a simplified steady state model for the 6 and 2-pole BDFM under investigation at OSU. His model is now being used for BDFM drive system development by Brassfield et al [10] and for BDFM parameter estimation by Ramchandran et al [11]. This thesis also utilizes the d -q model for stator optimization as discussed in Chapter 4.

### 1.3. Approach and Outline of Thesis Research

The areas of potential improvement of the BDFM system can be broadly divided into three categories :
(i) stator design for various pole structures, including winding distribution and slot space allocation to the two stator windings;
(ii) design of the die-castable rotor structure; and
(iii) optimal selection, design and control of the power electronic converter.

Only the first and the third areas as shown above are considered in this thesis. Rotor design studies are reported in a separate thesis [12].

The BDFM operating characteristics are described in Chapter 2, based on simulation and experimental results, with an emphasis on the active and reactive power flow structure in the 2 -pole winding. Chapter 3 discusses the available converter topologies and presents application based selection criteria. Chapter 4 discusses stator design optimization techniques for a 6 - and 2 -pole ASD. The use of the optimization program in designing a BDFM stator and its projected performance when compared to present and past laboratory prototypes is investigated in Chapter 5. Chapter 6 provides suggestions for future system optimization research activities.

## 2. OPERATIONAL CHARACTERISTICS OF THE BDFM

### 2.1. BDFM Operating Modes.

There are three different modes of BDFM operation as observed in the laboratory:
(i) singly-fed induction mode, where only one set of stator windings is excited and currents of a single frequency exist in the rotor,
(ii) doubly-fed asynchronous mode, where both the stators are excited, but currents of two different frequencies, as induced by both stators, exist in the rotor, and
(iii) doubly-fed synchronous mode, where currents of a single frequency exist in the rotor.

Of the three modes listed above and shown in Table 2.1, only the first and the third are practically important in BDFM operation. The doubly-fed asynchronous mode is of importance more for theoretical aspects than for any experimental or practical reasons, because of the instability and high losses associated with it. On the other hand, even though the doubly-fed synchronous mode of operation will be the desired mode for all BDFM studies in this thesis, the machine can operate in the singly-fed induction mode upon power converter failure. This is a useful feature for systems where continued rotor motion is of importance.

In the doubly-fed synchronous mode the speed of the rotor in terms of the frequencies on the power (6-pole) and the control (2-pole) side is given as follows :

$$
\begin{equation*}
N_{r}=\left\{\frac{f_{p} \pm f_{c}}{P_{p}+P_{c}}\right\} \times 60 \quad r / \mathrm{min} \tag{2.1}
\end{equation*}
$$

where $P_{p}, P_{c}$ are the number of poles on the power and control windings, respectively.

Table 2.1 Operational Modes of the BDFM in Steady State

| Mode of <br> Operation | Power winding <br> stator | Control winding <br> stator | Cage rotor |
| :---: | :---: | :---: | :---: |
| Induction <br> Mode | Excited with $f_{p}$ | Open or shorted | Only one <br> frequency $f_{r}$ |
| Asynchronous <br> Mode | Excited with $f_{p}$ | Excited with $f_{c}$ | Two frequencies <br> $f_{p}, f_{c}$ |
| Synchronous <br> Mode | Excited with $f_{p}$ | Excited with $f_{c}$ | Only one <br> frequency $f_{p p}=f_{r c}$ |

From the above it follows that with dc excitation on the 2-pole side and 60 Hz excitation on the 6 -pole side, the speed of the rotor is $900 \mathrm{r} / \mathrm{min}$. Machine operation with ac excitation on the 2-pole side of phase sequence opposite to that of the 6-pole leads to speeds below $900 \mathrm{r} / \mathrm{min}$ and is referred to as subnatural speed mode. Operation at speeds higher than $900 \mathrm{r} / \mathrm{min}$ is achieved with co-rotating 2-pole and 6-pole fields and is referred to as hypernatural speed mode of the BDFM.

### 2.2. 2-Pole Winding Power Requirements - Simulation Results

The active and reactive power flow associated with the 2-pole winding vary considerably over possible operating ranges. Depending on the application, operating conditions can involve any of the following constraints:
(i) maintaining constant p.f. on the 6-pole side with variations in load and
speed. This could be a desirable factor in many BDFM applications in the industry;
(ii) achieving synchronism with minimum 2-pole voltage excitation and/or minimum 2-pole kVA , which reduces the converter ratings and thus the overall system cost; and
(iii) maintaining constant current on the 6-pole side for different speeds of the BDFM.

The following sections show simulation results illustrating the active and reactive power flow structure of the BDFM as observed for these conditions in simulations. The simulations were conducted by the interactive program developed by Brassfield. For the simulations the 6-pole excitation is held constant at 230 V , while the 2-pole is excited


Figure 2.1 Dynamic simulation result illustrating 2-pole characteristics with constant 6-pole p.f. at 0.85 lagging and 6-pole excitation held constant at $230 \mathrm{~V}, 60 \mathrm{~Hz}$.
from a voltage source inverter. The load torque for each of the cases is held constant at either 7 Nm or 14 Nm , as specified in the plots.

### 2.2.1. Constant 6-Pole Power Factor

Initial studies with 14 Nm load, for a p.f. of 0.85 on the 6 -pole side, show real power flowing into the 2-pole winding from $900 \mathrm{r} / \mathrm{min}$ to about $577 \mathrm{r} / \mathrm{min}$ (corresponding to 2-pole excitation frequencies of 0 and 21.5 Hz respectively). Below the speed of $577 \mathrm{r} / \mathrm{min}$ the real power always flows out of the 2-pole winding. The reactive power flows into the 2-pole winding over the entire speed range. For the same load, but for a 6-pole p.f. of 0.9 it was found that real power only flows out of the 2-pole below 450 $\mathrm{r} / \mathrm{min}$; i.e. at a 2-pole frequency of 30 Hz . Again, the reactive power flows into the 2pole over the entire subnatural speed range. In order to achieve the high 6-pole p.f., the excitation requirements on the 2-pole side were significantly higher than those needed to barely maintain synchronism.

With decreasing torque or increasing 6-pole p.f., the change in the direction of the active power occurs at lower speeds i.e. at higher 2-pole frequencies due to change in the $I^{2} r$ losses in the machine. Direction of reactive power flow is not affected by load or p.f. while the magnitude varies substantially over the speed range, increasing with decrease in speed. For speeds around 900-1300 r/min, the dynamic model does not predict stable operation. In the hypernatural mode, the 6-pole p.f. requirements of 0.85 and 0.9 can only be achieved at speeds above $1450 \mathrm{r} / \mathrm{min}$. Negligible active power flows into the 2-pole winding between $1300-1400 \mathrm{r} / \mathrm{min}$. Both active and reactive power flow into the machine throughout the hypernatural speed range.


Figure 2.2 Dynamic simulation results of the 2-pole characteristics with minimum 2-pole excitation and 6-pole excitation held constant at $230 \mathrm{~V}, 60 \mathrm{~Hz}$.


Figure 2.3 Dynamic simulation results of 2-pole characteristics with minimum 2-pole kVA required for synchronism and 6-pole excitation maintained at $230 \mathrm{~V}, 60 \mathrm{~Hz}$.

### 2.2.2. Minimum 2-Pole Excitation (Voltage and/or KVA)

Simulations were also carried out such that the 2-pole was excited with the minimum voltage necessary to maintain synchronism. As mentioned earlier, the 6 -pole p.f. is expected to be low for this excitation condition. Thus, the magnetizing requirement of the machine is mostly met by the 6 -pole winding. For a load of 7 Nm , the direction of active power flow changes twice over the speed range of $75-900 \mathrm{r} / \mathrm{min}$ for the minimum voltage constraint as illustrated in Fig. 2.2. It can be seen that the direction of 2-pole active power flow is strongly influenced by the mechanical load.

Fig. 2.3 illustrates power flow for the constraint of minimum 2-pole kVA. As in the previous case, an increase in load causes 2-pole power flow reversal at lower speeds.

As in the case with constant 6-pole p.f., for speeds between $900 \mathrm{r} / \mathrm{min}$ and 1300 r/min stable operation cannot be achieved. Both the active and the reactive power flows into the machine over the entire hypernatural speed range. In both Figs. 2.2. and 2.3., the high reactive power requirement at hypernatural speeds is due to poor machine performance because of non-optimized machine parameters.

### 2.3. Laboratory Investigations

As stated earlier, extensive laboratory investigations have been conducted for prototype BDFM systems. Experiments for all BDFM operating modes show varying power requirements on the 2 -pole side. To give a clearer picture of the operating modes, 2 sets of BDFM experimental data are graphically presented here.

It should be mentioned that the operating conditions and constraints used for the


Figure 2.4 Plots of currents and voltages as observed in the laboratory while maintaining constant 6 -pole current at no-load.


Figure 2.5 Power flow in both winding while maintaining constant 6-pole current at no-load.


Figure 2.6 Plots of currents and voltages for minimum 2-pole excitation as observed in the laboratory.


Figure 2.7 Power flow in both winding for minimum 2-pole excitation.
experiments mentioned below are different from those used for the simulations in Sec.
2.2. Moreover, the tests are those of feasibility; so no constraints, such as the ones above were imposed on the operation. The result is evident in the data presented.

### 2.3.1. Constant 6-Pole Current

The machine was run at hypernatural speeds with the 6 -pole current maintained constant at no load. The 6 -pole winding was excited at 115 V rms from the power grid. The 2-pole was excited in the current mode to provide the best current balance. The only load on the machine during this experiment was that due to friction and windage.

### 2.3.2. Minimum 2-Pole Excitation

Again, the 6 -pole is excited with only 115 V rms and the 2-pole is excited by the converter operating in the current mode. The voltage on the 2 -pole side and the current on the 6 -pole side vary substantially over the speed range to maintain synchronism under this constraint. On the other hand, because the converter is in the current mode, the current on the 2-pole does not vary significantly. Since Fig. 2.7. shows only a small variation in 6 -pole input power over the speed range of the test, the large increase in 6-pole current is attributed to an increased reactive power requirement in the machine as speed was increased. When the 2-pole voltage was raised, causing an increase in the 2 -pole flux, the 6 -pole current dropped to a nearly constant value. This indicates the reactive power transfer capability between 6 and the 2 -pole windings. The high 2-pole active power at high speeds is due to the increased high-slip rotor losses, which are being supplied by the 2 -pole.

## 3. AC/AC CONVERTER TOPOLOGIES FOR BDFM APPLICATIONS

Since its inception, the BDFM research project has been utilizing a series resonant converter (SRC). While this converter topology has proven to be very flexible and appropriate during laboratory evaluation of prototype BDFM systems, the technology is relatively new with the associated reliability and availability constraints. This could lead to acceptance problems of BDFM systems. Thus, other options are evaluated which include widely available and accepted topologies as well as new technologies under development at university or industrial laboratories.

Recent developments in power semiconductor and converter technologies have led to a proliferation of circuit topologies with sometimes contradictory performance claims. This chapter presents a comprehensive review and a comparative evaluation of ac/ac converter topologies. This provides guidelines for selecting an optimum technology for a given application. While review and evaluation is done at a relatively high level, the comprehensive bibliography section will provide access to design details.

AC-AC converter topologies can be broadly classified into three categories depending upon the type of the intermediate power transfer link. All topologies have used one of the following (as shown in Fig.3.1): (i) dc link, (ii) ac link, or (iii) direct link. This method of classification is more methodical than categorization on the basis of topologies, primarily since recent research tends to use topological concepts belonging traditionally to a different converter scheme. An example of this phenomenon is the recent incorporation of Matrix converter theory for a PWM controlled rectifier [13] and VSI/CSI inverter [14].


Figure 3.1 Classification of AC/AC converter altematives based upon the type of intermediate power transfer link between the input and the output stage. $\checkmark$

### 3.1. DC Link

Examination of Fig. 3.1 shows that the maximum research effort to date had been with this link. Pulse Width Modulation (PWM), which is associated with hard switching, was first reported in the mid 1970's [15,16] while Resonant DC Link Technology, which features soft switching, appeared in 1986 [17]. These two broadly constitute the dc link domain of converters. Fig. 3.2 illustrates both hard and soft switched dc link VSI topologies. In the case of PWM, the link consists of a dc capacitor (for a VSI) or inductor (for a CSI). The Resonant DC Link includes a resonant circuit with a dc bias, such that, even though the voltage or current pulsate and become zero, they never cross zero. This illustrates the crossover between established topologies. Here the concept of resonance from ac-link conversion was applied to the dc link in a PWM topology.

Modifications of the original sinusoidal PWM include variation of switching frequency [18,19], non-sinusoidal harmonic injection and elimination schemes [20], model reference control [21], delta modulation [22-24], device dependent techniques [25,26] and control strategies based on matrix converter theory [13,14].

For higher power levels, in an effort to devise the most effective topology with the existing devices, mainly GTO-inverters incorporating PWM or six-step strategies are discussed in the literature. For instance, a current source GTO inverter with thyristor rectifier [25] is suitable for high power applications. Soft switching techniques with a modified and simplified snubber circuit (actually only one commutation circuit on the dc side) have been used in GTO based inverters for high frequency operations [26]. For


Figure 3.2 DC link ac/ac converter topology with PWM control of input and output stages. (Components in dashed box are required for soft switched resosnant operation.)
an effective, regenerative 4-quadrant PWM scheme, a simple chopper circuit can be used to control the power flow in the reverse direction without changing the switching mode of the main switches [27].

Increase in switching frequencies for PWM topologies, which is necessary to minimize the lower order harmonics, has been mainly supported by pioneering research and improvements in device switching speeds and SOA ratings. Modern power devices, power MOSFETs, IGBTs and Static Induction Transistors (SIT), promise to provide high power switching frequencies between 20 and 300 Khz [28-30]. Even though power devices have been improved considerably, their switching characteristics are still far from being ideal, which leads to very high switching losses in high frequency hard switched applications. Moreover, high dv/dt stresses associated with fast switching increase the possibility of insulation failures and electromagnetic interference (EMI).

With a power circuit similar to PWM topologies, with the addition of a resonant

LC tank at the dc bus (see Fig. 3.2.), resonant dc link (RDCL) converters minimize the switching losses, reduce harmonic problems and enable switching frequencies of 20 Khz and above [29]. At the point of common connection to the switches of the inverter stage, controlled oscillation of the dc bus enables commutation whenever the bus voltage (for VSI or parallel RDCL [17]) or bus current (for CSI or series RDCL [31]) reaches zero. Soft switching can also be achieved by Pole Commutated Inverters (PCI), which are typically characterized by one resonant commutation circuit per pole of the inverter to generate the desired soft switching conditions [32].

Recent enhancements for the RDCL include efficient control algorithms [33] and other modifications [32]. Most notably, the actively clamped version (ACRDCL) achieves a lower ratio of peak-to-average voltage and thus reduces the relatively high device voltage stress. This makes the ACRDCL one of the superior resonant dc link topologies available at present.

In an effort to eliminate the sub-harmonic spectra that is concomitant with discrete pulse modulation (DPM) systems, recent research has tried to incorporate the PWM system into the RDCL topology. The synchronized RDCL [34], the high frequency quasi-resonant dc voltage notching inverter (QRDCVNL) [35] and the ZVS RDCL-PWM inverter [36] are all topologies developed on the basic RDCL with PWM control features to eliminate the sub-harmonic energy spectra phenomenon. The quasiresonant dc link inverter (QRDCL) [37], yet another topology derived from the basic RDCL, addresses the high peak device stresses, which constitute the most important shortcoming of the basic topology.

The Series RDCL has the main disadvantage of having a current stress of around 2-2.5 pu on the devices and the resonant components. The use of a saturable core instead of the resonant inductor with a biasing current has been shown to reduce device current stress considerably [38].

Similarly, in an effort to reduce current stress and eliminate the sub-harmonic phenomenon associated with DPM control, research on the PCI has incorporated PWM control for the basic Resonant Pole Inverter (RPI). This resulted in the development of the Auxiliary Diode Pole Inverter (ADPI), Auxiliary Resonant Pole Inverter (ARPI) and Auxiliary Resonant Commutated Pole Inverter (ARCPI) [39,40]. A notable crossover of PCI technology is in the Matrix converter topology where the resonance associated with each pole of the converter can be used for high frequency, ZVS switching for the matrix converter [41]. It gives low device stresses, high spectral performance of the output voltage and power factor control of input quantities. The latest direct modification of the basic RPI circuit is the use of a non-linear resonant circuit implemented by a saturating inductor in the resonant pole. This reduces device current switching stress to around 1.5 pu [42].

### 3.2. AC Link

The need for reducing low order harmonics and minimizing switching losses at the same time has also stimulated research on resonant ac link converter topologies. These converters employ a high frequency resonant LC circuit in the power transfer path and are capable of providing four quadrant operation with low switching losses. Research in this area concerns mostly high power ranges. While these topologies have


Figure 3.3 Series resonant converter. (Note the requirement for input and output capacitors.)
been generating interest, the use of a higher number of switching devices and oversized resonant components have made them relatively uneconomical. Although the energy stored in the system is small, the resonant circuit has to handle the full-load power which leads to high VA ratings of the resonant components. Also, the control of bidirectional power flow and that of high frequency bus regulation is extremely complex. Thus, to date, resonant link converters have not been a viable competition for dc link topologies in the market place.

Research on the AC-link type converters has been dominated by the series resonant converter (SRC) topology, shown in Fig. 3.3. One of the most important advantageous features of this topology is that the zero crossing of the resonant link current allows natural commutation of the thyristors. Thus, unlike the new resonant parallel DC-link converters, which require gate turn-off devices, this topology uses relatively inexpensive and rugged devices for potentially very high power ratings. While
the SRC does not require commutation control, the necessity to back-bias the thyristors put an additional voltage stress of approximately 0.5 pu on the resonant capacitor. The power conversion process makes it possible to shape a current independently of the load impedance and voltage polarity if the primary power transfer and control mechanism is based on charge control [43]. Also, unlike dc link topologies, the SRC is suitable for severely unbalanced loads. The SRC has been shown to achieve desirable optimal power factor and power conversion ratio by modifying the modulation process [44]. Its possible use in high power variable speed generation, where the output total harmonic distortion needs to be minimized has been reported [45]. In an effort to reduce the number of thyristors in the original topology, a modified circuit with only half the number of thyristors was proposed [46]. However, this half bridge configuration reduces the line to line output voltage by a factor of $\sqrt{3}$ in comparison with the original SRC topology. The effects of soft switching as opposed to natural commutation can improve the performance of the series resonant converter [47]. At present, both device count as well as required component ratings make the SRC unattractive economically for many applications. However, it is anticipated that developments similar to active clamping in the RDCL will eventually be introduced and make ac link resonant converters more attractive and competitive.

The dual of the SRC, the parallel resonant converter, employs ZVS and is attractive for power supply applications where power transfer between different types of sources is essential for power distribution [48].

### 3.3. Direct Link

At the time of its appearance this was by far the best ac-ac conversion topology, performing conversion without energy storage devices. Direct Frequency Changer converter topologies have been known for some time. Though the Direct Frequency Changer topology in the form of the cycloconverter and the matrix converter can be found in industry, it suffers from the need for a bidirectional switch which would reduce


Figure 3.4 Matrix converter. (Note the requirement for bidirectional switches.)
the active device count [49]. The use of a large number of active devices and the complex control associated with it offsets the fact that it does not use any inductors or capacitors in the link. Frequency changer circuits are essentially arrays of bidirectional switching matrices, capable of sequentially connecting the voltages of a multiphase input source to a multiphase load, as illustrated in Fig. 3.4.

The cycloconverter is used in high power motor drives and in slip energy
recovery schemes. It can be shown that under certain conditions it behaves as an unrestricted frequency changer whose maximum frequency is not limited by the input frequency [50]. Cycloconverters can also be found in constant frequency applications where control over reactive power flow is of importance. Typical applications include correction of the power factor of drives and static conversion equipment, reduction of voltage flicker and voltage regulation of arc furnaces and voltage stabilization of transmission and distribution lines.

The matrix converter as proposed by Gyugyi and Pelly [50] was studied and perfected during the early part of the last decade by Venturini [51,52]. Realization techniques with existing unidirectional switches [53], and methods to devise adaptive switching transfer functions for original direct frequency changers [54] have been reported. The matrix converter is stated to be capable of operating at lagging, unity or even leading fundamental input power factor while supplying power to a lagging power factor load at a variable frequency [55]. A major disadvantage of the matrix topology is its limitation of the maximum output voltage to 0.866 of the input. Other shortcomings of the original proposals are: (i) assumption of ideal bidirectional switches (zero conduction loss, high bilateral reverse voltage blocking capability, zero turn-on and turnoff times); (ii) current commutation with the associated efficiency problems; and (iii) unsatisfactory short circuit protection.

Control strategies to address these problems have been reported in the literature. The ZVS of resonant pole inverters has been suggested for soft switched matrix converter at high frequency operation [40]. A commutation scheme which involves natural commutation for half the number of turn-offs and forced commutation for the

Table 3.1 $3 \phi-3 \phi$ AC/AC Converter Comparison

| Converter Topology | $\begin{aligned} & \text { Input } \\ & \text { Stage } \end{aligned}$ | $\begin{aligned} & \text { Output } \\ & \text { Stage } \end{aligned}$ | Device Count <br> Active (Passive) | Basic P Comp Co Link | issive <br> nent <br> t <br> Other | Number of Quadrents | Method of Switching | Device <br> Requirements | Output <br> Hermonic <br> Distortion | ${ }^{4}$ Max. Output Voltage | Control | Development Status | ${ }^{12}$ Approx: Device Voltage $1 \mathrm{pu}=$ $V_{\text {Hem }}$ | ${ }^{23}$ Approx. Device Current $1 \mathrm{pu}=$ Len |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
| Sinusoidal PWM VSI | passive active | PWM PWM | $\begin{aligned} & 6(12) \\ & 12(12) \end{aligned}$ | $\begin{aligned} & 1 \mathrm{C} \\ & 1 \mathrm{C} \end{aligned}$ | $\begin{aligned} & 0 \\ & 0 \end{aligned}$ | $\begin{aligned} & 2 \\ & 4 \end{aligned}$ | hard | mom-off | substantial | $\frac{\sqrt{3}}{2} \mathbf{v}_{4}$ | simple | $\begin{aligned} & \text { well } \\ & \text { established } \end{aligned}$ | 1.2 | 1 |
| Six-step VSI | passive active | $\begin{aligned} & \text { 6-step } \\ & \text { 6-step } \end{aligned}$ | $\begin{aligned} & 6(12) \\ & 12(12) \end{aligned}$ | $\begin{aligned} & 10 \\ & 10 \end{aligned}$ | $\begin{aligned} & 0 \\ & 0 \end{aligned}$ | $\begin{aligned} & 2 \\ & 4 \end{aligned}$ | hard | tum-off | maximum | $\frac{\sqrt{12}}{\pi} V_{d}$ | $\begin{aligned} & \text { very } \\ & \text { simple } \end{aligned}$ | $\begin{aligned} & \text { well } \\ & \text { established } \end{aligned}$ | 0.9 | 1 |
| Naturally Commutated Cycloconverter | single-tage (existence matrix) |  | 36 | 0 | 6 L | 4 | $\begin{gathered} \text { natural } \\ \text { commuts- } \\ \text { tion } \end{gathered}$ | SCR | very high | $\frac{3 \sqrt{3} v_{1}}{\pi}$ | complex | well established | 1.2 | 1 |
| RDCL Basic | passive active | DPM DPM | $\begin{aligned} & 6(12) \\ & 12(12) \end{aligned}$ | $\begin{aligned} & \text { 2C,1L } \\ & 2 \mathrm{C}, 1 \mathrm{~L} \end{aligned}$ | $\begin{aligned} & 0 \\ & 0 \end{aligned}$ | $\begin{aligned} & 2 \\ & 4 \end{aligned}$ | $\begin{aligned} & \text { soft } \\ & \text { ZVS } \end{aligned}$ | tum-off | subharmonics of $f_{\text {wat }}$ | $\frac{\sqrt{3}}{4}{ }^{2}$ | complex | emerging | 2.3 | 1 |
| RDCL <br> (Parallel) <br> Actively <br> Clamped | passive sctive | DPM <br> DPM | $\begin{gathered} 7(13) \\ 13(13) \end{gathered}$ | $\begin{aligned} & 2 \mathrm{C}, 1 \mathrm{~L} \\ & 2 \mathrm{C}, 1 \mathrm{~L} \end{aligned}$ | $0$ | $2$ | $\begin{aligned} & \text { soff } \\ & \text { zV } \end{aligned}$ | tum-off | subharmonics of $\mathrm{f}_{\mathrm{m}}$ | $S^{s}<\frac{\sqrt{3}}{4} V_{P}$ | complex | emerging | ${ }^{3} 1.2 \mathrm{k}$ | 1 |
| RDCL Basic Series | active | DPM | 13 (13) | 21,1C | 6 C | 4 | soft ZCS | SCR | subharmonics of $\boldsymbol{f}_{\boldsymbol{p}}$ | 'current link | $\begin{aligned} & \text { more } \\ & \text { complex } \end{aligned}$ | emerging | 1 | 2.3 |
| Auxiliary <br> Resonant <br> Commutated <br> Pole | PWM | PWM | 24 (24) | 2 C | $\begin{aligned} & 20 \mathrm{C} \\ & 12 \mathrm{~L} \end{aligned}$ | 4 | $\begin{gathered} \text { primary } \\ \text { switch-ZV } \\ \mathbf{S} \text { suxiliary } \\ \text { switch-ZC } \\ \text { S } \end{gathered}$ | tum-off | very low | $\frac{\sqrt{3}}{2} V_{4}$ | $\begin{gathered} \text { very } \\ \text { complex } \end{gathered}$ | emerging | 1.2 | 1 |
| $\begin{gathered} \text { Series } \\ \text { Resonent AC } \\ \text { Lint } \end{gathered}$ | DPM | DPM | 24 | LL,1C | 6 C | 4 | $\begin{aligned} & \text { soft } \\ & \text { ZCS } \end{aligned}$ | SCR | very low | "current link | $\begin{aligned} & \text { very } \\ & \text { complex } \end{aligned}$ | emerging | 1.5 | 2.5 |
| Matrix (Venturini) | $\begin{gathered} \text { singl } \\ \text { (existen } \end{gathered}$ | tage matrix) | unilateral <br> 18(18) or <br> bilateral 9 | 0 | 3 C | 4 | hard | biliteral turn-off (MCT) | substantial harmonics | $\frac{\sqrt{3}}{2} V_{1}$ | complex | potential | 2.2 | 1 |

[^0]remaining half had been proposed as "Semi-Natural Commutation" [53]. The short circuit protection scheme is an extension of the main control algorithm whereby it monitors the load current, abandoning PWM operation upon detection of overcurrent, and connecting each output phase to the input phase which will reduce the fault current to zero in the least time [53]. Modulation schemes other than PWM have also appeared in the literature for the realization of direct frequency changers in general, and matrix converters in particular. This includes optimal modulation schemes for frequency changers [54] and space vector modulation for matrix converters on both the input and output waveforms [60].

### 3.4. General Performance Evaluation and Selection

Table 3.1. [60] summarizes the performance and characteristics of some of the topologies discussed in Sect. 3.1. The table compares well established, hard switched topologies with emerging technologies. The information in the table assumes enough inductance of source and load to forego additional line inductors. Also, turn-off devices (transistors) are provided with separate reverse diodes. Converter and device voltages and currents are approximate and for relatively ideal circuit and switch conditions. Overall, the information is sufficient to form preliminary judgements on the merits and demerits of a particular topology based on component count, stresses, utilization, etc. Examination of the table illustrates, that the resonant dc link converters with reduced device ratings (active clamping) are at present the most likely competitors on the basis of performance.

Since cost, complexity, power range and reliability of any power electronic
converter are closely tied to the power devices used, the information in Table 3.1 needs to be correlated with the device characteristics listed in Table 3.2 [30,57]. Given the specifications required by an application, the tables can be used as tools in initially narrowing the possible converter choices. Subsequently, detailed information can be acquired from the references cited or by contacting possible suppliers. For resonant converter design, references [58] and [59] are recommended.

The following sections review some guidelines to consider when establishing application requirements and appropriate converter specifications.

Table 3.2 Turn-off Power Devices [30,57] (Peak Performance Consideration)

| Device | $\begin{aligned} & V_{\text {vex }} \\ & (V) \end{aligned}$ | $L_{m}$ $(\overline{\mathrm{A}})$ | $(\mu s)$ | $\begin{aligned} & \mathbf{L f}_{\text {an }} \\ & (\mathbf{K h z}) \end{aligned}$ | On State Voltage | Snubber <br> Resistance | $\begin{aligned} & \text { Gate } \\ & \text { Drive } \end{aligned}$ | Drive <br> Requirements |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
| BJT | 1200 | 300 | 15 | 5 | low | no | curreat | high |
| BJT | 550 | 480 | 5 | 5 | low | no | current | high |
| MOSFET | 1000 | 30 | 0.3 | 100 | high | no | voltage | low |
| IGBT | 1200 | 400 | 1 | 20 | low | no | voltage | low |
| STTh | 2000 | 600 | 2 | 10 | low | yes | current | high |
| ${ }^{2} \mathrm{MCT}$ | 3000 | 300 | 5 | 3 | low | no | voltage | low |
| GTO | 4500 | 3000 | 10 | 1 | low | yes | current | high |

${ }^{1}$ For hard awitched converters (typically a thermal rather than a device limit).
${ }^{2}$ Under dovelopment.

### 3.4.1. Power Rating and Flexibility

The required converter kVA rating is closely tied to the switching devices used.
Thus transistor based topologies are currently limited to a maximum output rating of approximately 250 kVA . Higher ratings can be achieved by using GTOs, at the expense of increased complexity and cost due to drive and snubber requirements. However,
recent developments in GTO technology also extend to the high power ranges, where GTO based technologies are replacing thyristor-based circuits in applications such as CSI drives which might find use in BDFM systems. It should be noted that the turn-off times of commercial GTOs may not yet be sufficient for some high frequency converters.

For high power requirements of the BDFM, as encountered in large adjustable speed drives, the voltage source PWM inverters preferably with snubber effected softswitching as in ARCPI or with some harmonics reduction scheme and matrix converters are likely to be the choice. The use of the matrix converter assumes the development of the MCT as a feasible commercial device. Moreover, because of its inherent limitation on the output voltage, the use of the matrix converter will be restricted to applications which do not demand BDFM operation at high slips. When the BDFM operates at speeds where the slip is resonably low, the excitation on the 2 -pole required to maintain synchronism is considerably lower than that at high slips. This was observed in the laboratory (refer Section 3.2.) and during simulations presented in Chapter 5. The use of the SRC in high power applications like variable speed generation systems has also been reported [45].

While any converter system can be provided with four-quadrant capabilities, the increased component count and complexity involved add to cost and can lead to reliability problems. For example, transistorized input stages as shown in Fig. 3.2 are exposed to voltage transients in the supply. Where bidirectionality of power flow is mandatory, as in most of the BDFM applications, the inherently four-quadrant
converters, such as the SRC, will find application niches.

### 3.4.2. Efficiency

While good efficiency is always desirable, it is truly a requirement at the high power levels. Soft switched converters are likely to be more efficient than their hard switched counterparts, though not as reliable. Thus, for relatively high power ranges present market trends still favor the hard-switched topologies. In the lower power ranges, RDCL topologies are superior to hard switched PWM converters, as the reduction in switching losses not only increases efficiency, but also allows for an increase in device rms current for a given thermal limit. If a true, bilateral MCT becomes available, matrix converter designs will likely achieve reasonable efficiencies, since only one device per phase is involved in the ac/ac conversion process.

In power ranges where a choice is possible, efficiency characteristics need to be examined in correlation with harmonic requirements as well as initial cost and other requirements. Since the overall BDFM system performance is sensitive harmonics, softswitched topologies have an edge over hard-switched ones. Due to reliability concerns, industrial end use for relatively higher power ranges still favors those PWM topologies which provide sufficient spectral performance. The ARCPI is an example of such a PWM topology which uses snubber circuits to provide soft-switching in an otherwise hard-switched topology. However, as resonant converters become available on a wider basis, and as more utilities provide efficiency incentives to industrial customers, the use of hard switched converters is likely to decrease.

### 3.4.3. Harmonic Considerations

Harmonic considerations of both output waveforms as well as input waveforms (proposed revisions to ANSI/IEEE 519) favor the soft switched, high frequency converters, which also have superior EMI characteristics. The high frequency converters minimize the size of required passive filtering components. Given the increase in cost for inductors and capacitors, cost considerations for harmonic mitigation also favor high frequency resonant converters, namely the RDCL in the lower power ranges and SRC and ARCPI for higher power applications.

### 3.4.4. Performance

BDFM systems targeted towards high performance applications require a converter with fast dynamic response and flexible control. In order to achieve the desired high control bandwidth, converter switching frequencies need to be relatively high. This favors the resonant topologies over hard switched PWM converters, as the high frequency is not associated with high switching loss. Added benefits of high frequency operation are improvement in power density and elimination of acoustic noise.

### 3.4.5. Reliability

Although hard switched converters are suspect of causing premature insulation failures due to high dv/dt stresses and can lead to thermal problems, overall converter reliability considerations at present favor established and proven technologies for any application in general. This is especially true in critical or conservative environments, such as process industries. However, as resonant converters are becoming more common, reliability problems associated with manufacturing are likely to be resolved.

Also, demonstration installations needs to be used to demonstrate reliable operation of new converter technologies for the BDFM systems.

### 3.4.6. Cost and Availability

In general, for industrial end-use applications, these issues again favor established technologies which have been cost-engineered and are available in quantity from a variety of manufacturers. Newer technologies, if available commercially, are often expensive due to engineering costs associated with still small production volumes. Also, some of these converters are only available on a sole-source basis from small companies.

For high performance applications, and for original equipment manufacturers (OEM), these constraints are not necessarily as restrictive as for industrial end-users. Thus, it is expected that new technologies, in tandem with the BDFM, will find their way to commercial acceptance through OEM applications such as heat pumps, compressors and air-conditioners.

### 3.5. Converter Selection Scheme for BDFM Applications

Though Figs. 2.1-2.3 are graphical presentations of the power structure for some particular operating conditions, they represent the overall active and reactive power flow trends in the 2-pole winding. It is very apparent from those figures that bidirectional capability of the converter is a mandatory criterion for BDFM applications in the subnatural speed zones. Although it is not compulsory to have a bidirectional converter for applications in the hypernatural speed ranges, it is indeed helpful for regeneration purposes if the machine happens to operate in the subnatural speed ranges.

Otherwise, the regenerated energy needs to be dissipated in external resistances. The possible BDFM applications and suggested converter choice is as follows:

- BDFM used for adjustable speed drive for industrial applications, such as fans, pumps and compressors. Section 3.4 suggests the use of PWM topologies with an emphasis on spectral performance, RDCL (Parallel and Series), SRC and Matrix converters (with the limitations as mentioned in Sec. 3.4.1) for this type of medium and low power industrial applications. Single phase BDFM ASDs for residential purposes could use the RDCL parallel converter to give a far better spectral and acoustic performance than that possible using a PWM converter.
- Variable speed generation systems for automotive, in-stream hydro or wind power applications. The PWM converter has been used in the studies for automotive applications of the BDFM. Studies show that the PWM converter is able to meet all the basic requirements for automotive generation systems. For higher power ac VSG schemes, such as in-stream hydro or wind power applications, SRC (series resonant converter) and RPI (resonant pole inverters) can be feasible choices. The high cost and the complexity of the SRC and RPI will likely be justified in the high power domains.

While hard switched and load commutated converters will continue to dominate the high power market for applications in general, in the foreseeable future, resonant soft switched converters are likely to make advances. Implementation of new soft switched technologies into the BDFM drive system, in the beginning only at the low power level,
is expected to find more widespread use in OEM and high performance market situations before expanding into industrial end-use applications. Of the resonant topologies, the clamped RDCL seems appropriate for lower power levels, while the resonant pole technology is applicable for higher power ranges. On the other hand, the SRC seems equally applicable in both high and low power ranges. Of the resonant link converters applied to BDFM systems, the SRC is likely to remain the dominant one, especially for its inherent bidirectionality and capability to operate into unbalanced loads. Successful matrix converter incorporation into the BDFM system or into any application in general still hinges on further development of the MCT.

## 4. STATOR OPTIMIZATION

While the prototype BDFM systems at OSU provided insight into the operating principles of this type of "single frame" induction machine, they were far from efficient and optimal. Thus, as part of the overall machine optimization, the stator design was investigated for this thesis. An optimization program with a direct search algorithm was developed for the purpose. Simulations conducted with the help of the optimization program provided stator configurations with improved performance characteristics over prototype BDFM systems in the laboratory.

It is necessary to reiterate the simplifications and assumptions made for deriving the original steady state model. This helps explain and clarify the excellent performance of some configurations generated by the program. The model inherently ignores any core-loss in the rotor and neglects higher order harmonics which are present in the machine. The results obtained from the program are very encouraging and should be evaluated on laboratory prototypes. Methods of designing the stator for different operating conditions and requirements are described in the following sections.

### 4.1. Development of the BDFM Optimization Model

The steady state model of the BDFM as derived from the d -q equivalent [9] was utilized in the stator optimization scheme. The optimization is motivated by the need for devising a mathematical model to generate optimized parameters for a given operating condition of the BDFM. This generalized problem can be treated as a constrained optimization problem [61] of the form :

$$
\begin{array}{cll}
\text { minimize } & f(x) & x \in \mathbf{R}^{n} \\
\text { subject to } & c_{i}(x)=0, & i \in E  \tag{4.1}\\
& c_{l}(x) \geq 0, & i \in I
\end{array}
$$

where $f(x)$ is the cost or the objective function, with the constraint functions $c_{i}(x)$, $i=1,2, \ldots . p . \quad E$ is the index set of equality constraints and $I$ is the set of inequality constraints in the problem; both sets being finite.

### 4.1.1. Configuring the BDFM for the General Optimization Problem

As mentioned in the previous chapters, a substantial part of the system cost is that of the electronic power converter. Overall system design involves a trade-off between device ratings of the converter and the overall system performance. It is necessary to minimize the current flowing through the converter, ie. the 2-pole current, which would help reduce the current rating of the power electronic devices in the converter and thus its cost. Moreover, the current in the rotor needs to be kept within limits to reduce the rotor $I^{2} r$ losses. It is also desirable to optimize the 6-pole current to minimize the overall $\mathrm{I}^{2} \mathrm{r}$ loss in the machine model. The optimized 2-pole and 6-pole current will reduce the investment made for the converter and provide stator parameters efficient enough to conform to the constraints laid down by the designer. Thus the modified constrained optimization problem appears as

$$
\begin{array}{cc}
\text { minimize } & \left|I_{6}\right| \\
\text { minimize } & \left|I_{2}\right|  \tag{4.2}\\
\text { minimize } & \left|I_{r}\right| \\
\text { subject to } & C_{1}(p) \leq 0, \quad i \in I
\end{array}
$$

where $I_{6}, I_{2}$ and $I_{r}$ are the d-q rms currents in the 6 -pole, 2 -pole and the rotor respectively. $C_{i}(p)$ is the set of constraints involving current carrying limitations in the windings, p.f. requirement on the 6 -pole side and a minimum overall efficiency.


Figure 4.1 6- and 2-pole BDFM Steady state equivalent circuit for synchronous operation [9].

The optimization as outlined in (4.2), with the three objective functions, is a relatively involved problem. Besides, it may be difficult to determine a set of parameters which would simultaneously satisfy all the constraints and optimize all three objective functions. So a single objective function was developed which reflects the effect of each of those in (4.2). This function uses the sum of the weighted squares of the currents. The currents were squared to account for the magnitudes of the currents in the objective function of (4.2). This is similar to a function of the sum of the
resistive losses in the machine model, but the weights can be adjusted to reflect other losses in different regions of the machine. Thus, the new optimization model for the BDFM is

$$
\begin{array}{ll}
\text { minimize } & \left\{\alpha I_{6}^{2} r_{6}+\beta I_{2}^{2} r_{2}+\gamma I_{r}^{2} r_{r}\right\} \\
\text { subject to } & \text { (i) } \cos \left\{\tan ^{-1}\left(\frac{I_{G \alpha}}{I_{G r}}\right)\right\} \geq p f_{\operatorname{mm}} \\
\text { (ii) }\left(\frac{\text { output }}{\text { output }+ \text { losses }}\right) \geq \text { eff } f_{\min }  \tag{4.3}\\
\text { (iii) } I_{6} \leq l i m_{6} \\
\text { (iv) } I_{2} \leq \lim _{2}
\end{array}
$$

where

| $\alpha, \beta, \gamma$ | weights for the various resistive losses, |
| :--- | :--- |
| $I_{6 i}$ | imaginary part of d-q steady state 6-pole current, |
| $I_{6 r}$ | real part of d-q steady state 6-pole current, |
| $p f_{\text {min }}$ | minimum 6-pole p.f., |
| effimin | minimum overall efficiency, |
| $l i m_{6}$ | maximum current carrying capability on the 6-pole side |
|  | set by the wire gauge in use and |
| $l i m_{2}$ | maximum current carrying capability on the 2-pole side |
|  | set by the wire gauge in use. |

The currents are calculated iteratively using the BDFM steady state model [49] as shown below and the equivalent circuit in Fig. 4.1 :

$$
\begin{equation*}
\dot{V}_{q 6}=\left(r_{6}+j X_{\sigma \sigma}\right) \dot{I}_{\phi \sigma}+j X_{m o} \dot{I}_{\phi r} \tag{4.4}
\end{equation*}
$$

$$
\begin{align*}
& \frac{\dot{V}_{q 2}}{s}=\left(\frac{r_{2}}{s}+j X_{a 2}\right) \dot{I}_{q 2}-j X_{m 2} \dot{I}_{q r} \quad \text { for } s \neq 0  \tag{4.5}\\
& \frac{\dot{V}_{q r}}{s_{1}}=j X_{m \sigma} \dot{\sigma}_{q 6}-j X_{m 2} \dot{I}_{q 2}+\left(\frac{r_{r}}{s_{1}}+j X_{r}\right) \dot{I}_{q r} \tag{4.6}
\end{align*}
$$

In equations (4.4) through (4.6)

$$
\begin{gathered}
s=\frac{\omega_{2}}{\omega_{6}}, s_{1}=\left(\frac{\omega_{6}-3 \omega_{r}}{\omega_{6}}\right)=\left(\frac{\omega_{2}+\omega_{r}}{\omega_{6}}\right), \\
X_{\infty}=\omega_{6} L_{\infty}, X_{s 2}=\omega_{6} L_{s 2}, X_{r}=\omega_{6} L_{r} \\
X_{m 6}=\omega_{6} M_{6}, X_{m 2}=\omega_{6} M_{2}, \omega_{6}=2 \pi f_{6}, \omega_{2}=2 \pi f_{2},
\end{gathered}
$$

with $L$ and $M$ representing the self and the mutual inductances in the model, respectively.

### 4.2. Algorithm of the Program

The diagrammatic representation of the program flowchart is shown in Fig. 4.2. The optimization model in (4.3) is a 6 -dimensional problem with the rotor parameters held constant. The parameters that need to be determined are the winding resistance and inductance for the 6- and 2-pole and the two stator to rotor mutual inductances. All these parameters are proportional to either the number of turns or the square of number of turns as shown below :


Figure 4.2 Simplified flowchart for the stator optimization program

$$
\begin{align*}
r_{p} & =r_{p, \operatorname{bax}} N_{p} \\
L_{p p} & =L_{\text {P, hax }} N_{p}^{2}  \tag{4.7}\\
M_{p} & =M_{p, \operatorname{cax}} N_{p}
\end{align*}
$$

where $p=6$ or $2, r_{p}$ is the resistance, $L_{s p}$ the stator inductance, $M_{p}$ the stator to rotor mutual inductance and $N_{p}$ the number of turns in the p-pole winding. The base values of the resistance and the inductances are calculated with a single turn on the 6 and the 2-pole windings. These base values are calculated by new computer programs or ones derived from the detailed BDFM model [63].

The search algorithm needs to sweep over all possible resistances that can be generated from the wire-gauge table. The self and mutual inductances will be calculated based on the winding structure specified. The rotor resistance and inductance can be determined depending on the rotor structure to be used and is maintained constant throughout the program execution. Thus the algorithm searches for parameters by varying the number of turns, the range being specified by the user, on both the 2 -pole and the 6 -pole windings. Even the wire-gauge size is varied with the help of a wiregauge look-up table on both windings. This changes the diameter and thus the resistance associated with the windings. To simulate realistic stator slot usage, machine parameters and eventual steady state calculations are carried out only for those number of turns and wire-gauge size combinations that satisfy the following slot-fill condition:

$$
\begin{equation*}
0.6 A<\left(N_{6} a_{6}+N_{2} a_{2}\right)<0.7 A \tag{4.8}
\end{equation*}
$$

where $N_{6}, N_{2}$ are the number of turns and $a_{6}, a_{2}$ are the cross-sectional area of the wires
used for the 6-pole and the 2 -pole respectively. $A$ is the cross-sectional area of the stator slot which needs to be distributed between the two windings.

Once condition (4.8) is satisfied, the steady state currents in the 6-pole, 2 -pole and the rotor are calculated iteratively by the Newton-Raphson method. The 6-pole excitation for the model on the 6 -pole side is held constant and that on the 2 -pole voltage is swept over a range of voltage around the corresponding constant $\mathrm{V} / \mathrm{Hz}$ operating point. The steady state model may not converge to a solution for all 2-pole voltages. Thus, to reduce program execution time, the steady state model convergence is checked for voltages over the specified range with an increment of 20 V . Whenever the program encounters a steady state convergence, say at $\mathrm{V}_{\text {2,con }}$, it sweeps the 2-pole voltage from $\left(\mathrm{V}_{2, \text { con }}-20\right) \mathrm{V}$ to the maximum voltage set by the user, using increments of 1 V .

In motoring operation, the load torque is usually specified and it can be shown that, although the 6 and 2 -pole input voltages are given, the angle, $\beta$, between the reference voltage $\mathbf{V}_{\mathbf{q} 6}$ and $\mathrm{V}_{\mathbf{q} 2}$ is an unknown function of both load torque and 2-pole excitation voltage. The solution of the steady state equations characterized by (4.4) through (4.6) thus requires that they be solved simultaneously with the torque equation. Before these equations can be solved, the complex equations are changed into real algebraic equations by the following relations :

$$
\begin{gather*}
\dot{V}_{Q 6}=V_{Q 6}+j V_{Q 4}=V_{q 6 \sigma}+j 0  \tag{4.9}\\
\dot{V}_{Q 2}=V_{Q 2 r}+j V_{Q 21}=V_{Q 2} \cos \beta-j V_{Q 2} \sin \beta \tag{4.10}
\end{gather*}
$$

where $V_{q 2}$ is assumed to be lagging $V_{q 6}$ by the angle $\beta$ and

$$
\begin{equation*}
\dot{I}=I_{r}+j I_{i} \tag{4.11}
\end{equation*}
$$

where subscripts $r$ and $i$ represent real and imaginary parts of the phasor quantities, respectively.

By substitution of (4.9) through (4.11) into the steady state equations of (4.4) through (4.6) along with the torque equation, and separating the real and the imaginary parts, the following set of non-linear algebraic equations are obtained [64] :

$$
\begin{align*}
& r_{6} I_{\text {eci }}+X_{o f} I_{\text {ger }}+X_{m o} I_{\text {gTr }}=0 \\
& \frac{r_{2}}{s} I_{q 2 r}-X_{\Delta 2} I_{q 2 i}+X_{m 2} I_{q \pi t}-\frac{1}{s} V_{q 2} \cos \beta=0 \\
& \frac{r_{2}}{s} I_{q 21}+X_{s 2} I_{q 2 \pi}-X_{m 2} I_{q r}+\frac{1}{s} V_{Q 2} \sin \beta=0  \tag{4.12}\\
& -X_{m 0} I_{q Q 1}+X_{m 2} I_{q 2 i}+\frac{r_{r}}{s_{1}} I_{q r T}-X_{r_{q r t}}=0 \\
& X_{m o l} I_{q \sigma r}-X_{m 2} I_{q 2 r}+\frac{r_{r}}{s_{1}} I_{q r t}+X I_{q r}=0
\end{align*}
$$

$$
\begin{aligned}
& +2 M_{2}\left(I_{q 2} I_{q r}-I_{q 2} I_{q H}\right)-T_{L}=0
\end{aligned}
$$

The procedure for solving these equations has been presented in [9] and [64]. It is reviewed here for the convenience of the reader. The above equations can be written as

$$
\begin{equation*}
F(Y)=0 \tag{4.13}
\end{equation*}
$$

where

$$
Y=\left[I_{q 6 r}, I_{q 6 i}, I_{q 2 r}, I_{q 2 i}, I_{q r r}, I_{q r}, \beta\right]^{t}
$$

is the vector of the unknown quantities and

$$
F=\left[f_{1}(Y), f_{2}(Y), f_{3}(Y), f_{4}(Y), f_{5}(Y), f_{6}(Y), f_{7}(Y)\right]
$$

is the function vector containing the seven scalar functions of the equations in (4.13).
Newton's algorithm for iterative approximation is adequate to obtain solutions to (4.13) as follows. At the $i^{\text {th }}$ iteration the unknown vector $Y^{(i)}$ and the previous approximation $Y^{(i-1)}$ and the difference $\Delta Y^{(i-1)}$ are related by

$$
Y^{(0)}=Y^{(t-1)}-\Delta Y^{(t-1)}
$$

where

$$
\Delta Y^{(i-1)}=\left[\frac{\partial F}{\partial Y}\right]^{-1} F^{(i-1)}
$$

The elements of the Jacobian

$$
\left[\frac{\partial F}{\partial Y}\right]=\left[\frac{\partial f_{m, n}}{\partial Y_{m, n}}\right] \text { for } m, n=1,2, \ldots 7
$$

are computed as follows.
It is noticed that four out of the seven equations in (4.13) are linear. Thus the Jacobian matrix entries for these equations are just the circuit parameters themselves.

The other terms of the matrix are given below without providing intermediate results :

$$
\begin{aligned}
& \frac{\partial f_{3}}{\partial \beta}=\frac{1}{s} V_{q 2} \sin \beta, \frac{\partial f_{4}}{\partial \beta}=\frac{1}{s} V_{q 2} \cos \beta, \frac{\partial f_{7}}{\partial I_{q \sigma r}}=-6 M_{6} I_{q r t} \\
& \frac{\partial f_{7}}{\partial I_{q \alpha i}}=6 M_{6} I_{q r r} \frac{\partial f_{7}}{\partial I_{q 2 r}}=-2 M_{2} I_{q r r}, \frac{\partial f_{7}}{\partial I_{q 2 i}}=2 M_{2} I_{q r} \\
& \frac{\partial f_{7}}{\partial I_{q T}}=6 M_{6} I_{q ब i}+2 M_{2} I_{q 2 r} \quad \frac{\partial f_{7}}{\partial I_{q r i}}=-6 M_{6} I_{q 6 r}-2 M_{2} I_{q 2 r}
\end{aligned}
$$

Table 4.1 Sample Output File

| Total <br> losses | 2-pole <br> voltage | 6-pole <br> no. of <br> turns | 6-pole <br> AWG <br> size | 2-pole <br> no. of <br> torss | 2-pole <br> AWG <br> size | 6-pole loss <br> (in watts) | 2-pole loss <br> (in watts) | rotor loes <br> (in watts) | efficieny in <br> \% | 6-pole <br> p.f. |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
| 148.5 | 158 | 29 | 13 | 24 | 13 | 30.4 | 23.006 | 95.153 | 76.0311 | 0.780 |
| 148.0 | 153 | 30 | 13 | 23 | 13 | 31.195 | 22.588 | 94.287 | 76.0911 | 0.783 |
| 148.2 | 187 | 30 | 13 | 28 | 14 | 31.426 | 23.208 | 93.652 | 76.0645 | 0.780 |
| 148.7 | 192 | 30 | 13 | 29 | 14 | 30.929 | 22.795 | 95.035 | 76.0065 | 0.787 |
| 148.0 | 193 | 30 | 13 | 29 | 14 | 31.195 | 22.587 | 94.287 | 76.0913 | 0.783 |
| 147.3 | 194 | 30 | 13 | 29 | 14 | 31.463 | 22.38 | 93.553 | 76.1741 | 0.780 |
| 148.1 | 149 | 31 | 13 | 22 | 13 | 32.342 | 22.161 | 93.648 | 76.0811 | 0.782 |
| 148.6 | 183 | 31 | 13 | 27 | 14 | 32.397 | 22.725 | 93.518 | 76.0212 | 0.782 |
| 148.4 | 189 | 31 | 13 | 28 | 14 | 32.126 | 22.114 | 94.172 | 76.0492 | 0.785 |
| 147.8 | 190 | 31 | 13 | 28 | 14 | 32.406 | 21.906 | 93.496 | 76.1233 | 0.782 |

To ensure faster convergence of the Newton-Raphson iteration method, an initialestimate for $\beta^{(0)}$ is used to solve the equivalent circuit of Fig. 4.1 and equations (4.4) through (4.6) for the initial values of $Y^{(0)}$. The simplified phasor equations derived for solving the initial values of $Y^{(0)}$ are as follows :

$$
\begin{align*}
& \dot{I}_{q 6}=\frac{\dot{V}_{q 6}-z_{m \sigma} \dot{I}_{q r}}{z_{\sigma}} \\
& \dot{I}_{q 2}=\frac{\left(\frac{\dot{V}_{q 2}}{s}\right)-z_{m 2} \dot{I}_{q r}}{z_{2}} \text { and }  \tag{4.14}\\
& \dot{I}_{q r}=\frac{\left(\frac{z_{m \sigma}}{z_{\sigma}}\right) \dot{V}_{q 6}+\left(\frac{z_{m 2}}{z_{2}}\right)\left(\frac{\dot{V}_{q 2}}{s}\right)}{\left(\frac{z_{m 6}^{2}}{z_{\sigma}}\right)+\left(\frac{z_{m 2}^{2}}{z_{2}}\right)-z_{r}}
\end{align*}
$$

where

$$
\begin{gathered}
z_{6}=r_{6}+j X_{\infty 0} \quad z_{2}=r_{2}+j X_{\infty 2} \\
z_{r}=r_{r}+j X_{r} \quad z_{m \sigma}=j X_{m 0} \quad z_{m 2}=j X_{m 2}
\end{gathered}
$$

Finally, suitable configurations of the BDFM which satisfy user specified 6-pole p.f. and overall machine efficiency requirements are written to an output file. The output file also contains information regarding the allocation of the $I^{2} r$ losses within the different parts of the machine. A sample output file is shown in table 4.1. Listings of the main optimization program are provided in Appendix A.

## 5. STATOR DESIGN STUDY

The stators for the BDFM laboratory prototypes thus far have been designed with the help of the detailed model $[7,63]$. While these prototypes provide considerable insight into the operational modes of the BDFM, rotor and stator design need to be improved considerably before commercially competitive performance is achieved. This is apparent from the simulation results in Chapter 2. Due to the complexity of the detailed model and its computational intensity, it is difficult and impractical to use in stator optimization activities. Thus, the design optimization scheme presented here, uses the d-q model. This chapter explains the use of the design algorithm outlined in Chapter 4. As explained previously, the method of optimization utilizes a search algorithm which scans for number of turns on the two stator windings, wire gauge sizes and 2-pole voltage.

A brief description of the laboratory prototype machine is provided and used as a comparative basis for the optimized stator BDFM. The use of the optimization program for designing a BDFM optimized over a speed range of $0-1800 \mathrm{r} / \mathrm{min}$ is described in this chapter. Simulation results illustrate the performance of the proposed configuration. These results are then compared with an existing laboratory prototype and with the configuration used by Li [9].

### 5.1. Background

The d-q model developed by Li [9] was actually based on a laboratory prototype common winding BDFM. Machines of this type use the same winding to create two equivalent stators with different pole numbers [1,3]. It can be shown that for different
pole numbers on the two equivalent windings they can be treated as two magnetically decoupled stator systems. Thus, for modeling purposes, the two windings are treated as isolated windings. Consequently, the BDFM d-q model can be used for both isolated and common winding design. The isolated winding design has the advantage of

Table 5.1 BDFM D-Q Parameters

| Machime | Number of tares |  | A.W.G. number |  | d-q parameters |  |  |  |  |  |  |  |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
|  | $\begin{gathered} 6- \\ \text { pole } \end{gathered}$ | $\begin{gathered} 2 . \\ \text { pole } \end{gathered}$ | 6. pole | $\begin{gathered} 2 \cdot \\ \text { pole } \end{gathered}$ | ${ }_{6}$ | $\mathrm{F}_{2}$ | $r_{\text {r }}$ | $\mathbf{x}$ | $\mathrm{X}_{\mathrm{z}}$ | $\mathrm{X}_{\mathbf{r}}$ | X ${ }_{\text {¢ }}$ | $\mathbf{X}_{\mathbf{2}}$ |
| Common winding | 60 | 60 | 19 | 19 | 0.81 | 0.81 | 327.5 $\mu$ | 30 | 237 | 0.016 | 0.334 | 1.623 |
| Lab. iwolated winding | 28 | 25 | 17 | 17 | 1.64 | 2.84 | $327.5 \mu$ | 34 | 228 | 0.016 | 0.405 | 1.8 |
| Optimized isolated winding | 33 | 20 | 14 | 12 | 0.97 | 0.71 | 327.5 $\mu$ | 47 | 146 | 0.016 | 0.478 | 1.44 |

allowing the use of different wire sizes for the two stator systems, thus providing an additional degree of freedom. Moreover, it was experimentally established in the laboratory that the isolated winding BDFM is superior in performance to the common winding version, which exhibits internal circulating currents. Thus, all design investigations here consider the isolated winding BDFM.

The common winding BDFM as described by Li [9] has a 4-nest, 6-loop nest rotor structure with a 4-bar cage. This rotor configuration is also used in the existing laboratory prototype and in the design optimization procedure. Due to the winding
structure of the common winding BDFM used by Li the 6 and the 2 -pole have the same winding resistance. The steady state $\mathrm{d}-\mathrm{q}$ parameters of this machine as listed in [9] along with those of the present isolated winding prototype and the stator optimized BDFM are tabulated in Table 5.1.

Recent investigations show that a rotor design without cage has superior low speed characteristics. While the results presented here assume a cage structure, design adaption to the cageless geometry are straightforward. The model representation of the existing laboratory prototype uses 28 single-layered turns on the 6 -pole and 25 singlelayered turns on the 2 -pole, both with A.W.G. 17. The wire gauge table used as a lookup table by the optimization program is provided in Table 5.2.

### 5.2. Design Derivation and Projected Performance

The main design criterion was to optimize the BDFM stator over the speed range of $0-1800 \mathrm{r} / \mathrm{min}$ for a torque load of 15 Nm . Thus, it was necessary for the design to incorporate a requirement that would enable the machine to achieve synchronism at the maximum speed of $1800 \mathrm{I} / \mathrm{min}$ with a load torque of 15 Nm and with a maximum of 230 V excitation on the 2-pole. In order to account for the additional losses not considered by the model, such as core-loss and windage and friction loss, a conservative upper bound of 200 V on the 2-pole side was enforced. This allows for additional control margin.

With these conditions implemented, the number of turns was swept from 20 to 40 on both windings. All designs that failed to support a load of 15 Nm at $1800 \mathrm{r} / \mathrm{min}$ were discarded by the program. It was found that configurations with lower number of
turns on the 2-pole (around 20) were suitable and met the requirements. The ones with higher 2-pole number of turns (around 25) needed excessively high 2-pole excitation.

The acceptable configurations, i.e. those with around 20 turns on the 2-pole and 33 turns on the 6-pole, were evaluated for efficiency and 6-pole p.f. It was found that at hypernatural speeds, the efficiencies provided by those configurations were almost

Table 5.2 A.W.G. Look-up Table

| Wire size A.W.G. | Current carrying capacity in Amps. | Diameter in mm . |
| :---: | :---: | :---: |
| 6 | 37.5 | 4.115 |
| 7 | 29.7 | 3.665 |
| 8 | 23.6 | 3.264 |
| 9 | 18.7 | 2.906 |
| 10 | 14.8 | 2.588 |
| 11 | 11.8 | 2.305 |
| 12 | 9.33 | 2.053 |
| 13 | 7.4 | 1.828 |
| 14 | 5.87 | 1.628 |
| 15 | 4.65 | 1.45 |
| 16 | 3.69 | 1.291 |
| 17 | 2.93 | 1.15 |
| 18 | 2.32 | 1.024 |
| 19 | 1.84 | 0.912 |
| 20 | 1.46 | 0.812 |
| 21 | 1.16 | 0.723 |
| 22 | 0.918 | 0.644 |
| 23 | 0.728 | 0.573 |
| 24 | 0.577 | 0.511 |
| 25 | 0.458 | 0.455 |

same. The decision to select A.W.G. 14 on the 6-pole and A.W.G. 12 on the 2-pole
was mostly based on the overall machine performance in subnatural speed ranges. It should be noted that for winding implementation the wire sizes may be too large. In this case, two wires of smaller size can be used in parallel for equivalent impedence (i.e. 2 parallel turns of A.W.G. 15 equal 1 turn of A.W.G. 12). This configuration need not be the best for all possible operating conditions, as mentioned in Chapter 2. It is optimized for supporting a load of 15 Nm at $1800 \mathrm{r} / \mathrm{min}$ with one of the lowest 2-pole excitations. The machine parameters are provided in Table 5.1.

Figs. 5.1 through 5.4 show the machine performance expected for this configuration over the speed range $0-1800 \mathrm{r} / \mathrm{min}$. The gaps in the all the plots are due non-convergence of the BDFM model at those speeds.


Figure 5.1 Plot of maximum efficiency and corresponding 2-pole excitation.


Figure 5.2 Variation of 6-pole p.f. at maximum efficient operating point.


Figure 5.3 Active and reactive power flow in 2-pole winding at maximum efficient operation.


Figure 5.4 Plot of maximum efficiencies of the different machines over a range of $0-1800 \mathrm{r} / \mathrm{min}$.


Figure 5.5 Variation of the 6-pole p.f. angle of the different machines operating under maximum efficiency conditions.


Figure 5.6 Plot of the 2-pole active power flow in the different machines under the maximum efficiency conditions.


Figure 5.7 Plot of 2-pole reactive power for the different machines under maximum efficiency conditions.

### 5.3. Comparative Evaluation

Figs. 5.5 through 5.8 compares the efficiency, 6 -pole power factor angle and the 2-pole active and reactive power over the entire speed range. It should be noted that all the simulations for this comparison were conducted with a load torque of 15 Nm . The simulation generated the most efficient steady state operating points for different speeds. These were used to plot and compare the performance of the different configurations.

As shown in Figs. 5.5 through 5.8, both the non-optimized machines fail to achieve synchronism at $1800 \mathrm{r} / \mathrm{min}$ with a 2 -pole excitation lower than 210 V . Fig. 5.5 shows that the efficiency of the optimized BDFM is distinctly higher than either of the two machines in both the low ( $0-400 \mathrm{r} / \mathrm{min}$ ) and high ( $1350-1800 \mathrm{r} / \mathrm{min}$ ) speed ranges. For all remaining speeds the efficiency of the optimized BDFM matches that of the common winding BDFM. It should be noted that the d-q model neglects the circulating currents in the common winding configuration, which in practice lead to distintly higher copper loss. Moreover, the optimized BDFM needs the lowest active and reactive power among the machines compared. Thus the power converter rating can be reduced, which in turn lowers system cost and potentially allows for competitive pricing.

## 6. CONCLUSIONS AND RECOMMENDATIONS

### 6.1. Conclusions on System Optimization Studies

Recent progress in power semiconductor and converter technologies has brought renewed interest in ac drive systems. Modifications to induction motor drive systems have been restricted to power converter improvements. Other proposed drives utilize high performance machines, such as Permanent Magnet motors, which are significantly more expensive than IM's. Conventional ac drives require a power converter of full rating, which is a major contributor to high system cost. An alternative drive system, such as the BDFM proposed at OSU, utilizes an induction-type machine with reduced converter requirements. As illustrated in this thesis, the goal of competitive performance can be met for decreased converter rating. This can eventually lead to a commercially competitive design.

Results reported in this thesis show that both the stator and the power converter can be optimized to achieve maximum system performance for specific applications. Optimization of the machine should also extend to the rotor, which significantly enhance BDFM performance. It is in the rotor where the maximum losses occur (both resistive and core). Rotor loss is estimated to be higher than the combined stator loss. This is primarily because of the high frequency currents present in the rotor bars.

As shown in chapter 5 , stator optimization has successfully been applied to reduce stator losses and reduce the required converter rating. Thus, the optimized design improves performance and helps reduce system cost. The reduced reactive power requirement enables the use of cheaper and readily available power converters for BDFM systems, at least for conventional applications.

### 6.2. Recommendations for Future Research

Based on the stator design optimization work reported in this thesis, it is felt that the following changes and enhancements are beneficial :
(i) Instead of a direct search algorithm, a faster optimization algorithm can be devised that would leave provisions for the future incorporation of rotor parameter optimization.
(ii) The optimization scheme should search for the best pole combination of the two stators. This can only be done with a generalized BDFM d-q model, which at present is not available.
(iii) The program should be able to modify the winding structure on the stator including their harmonic effects, unlike the present work, which only represents single and double layered windings.
(iv) The optimization model should simulate the effects of high frequency harmonics on the overall BDFM performance. This would help optimize the stator with respect to the rotor harmonics, which otherwise go unnoticed in this optimization scheme.

The search for the power converter which would suit the BDFM and meet various performance criteria was mainly based on the results reported in the literature. Future work should involve modeling the suitable converter topologies, interfaced with the BDFM d-q model, to simulate the device stresses and performance reliability of the converter for various operating constraints. Based on the results obtained from this simulation scheme, the best converter topology can be selected with more confidence.

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## APPENDIX

## Optim.c

## STATOR OPTIMIZATION PROGRAM --- DIRECT SEARCH ALGORITHM WRITTEN BY <br> Shibashis Bhowmik

The program uses single-layered winding on both the power and control sides. The pole number combination is fixed for the present program as 6 and 2 poles. The different subroutines with the ".c" extension and included in the main routine are used in the optimization scheme as mentioned below :
declare.c declaration of global variables
complex_math.c operational routines for comlex numbers
real_math.c operational routines for real matrices
circuit.c calculates initial guess for Newton Raphson iterations
function.c generates the function vector containing seven scalar functions of equation (4.12)
jacobian.c calculates the jacobian matrix
newrap.c updates the function vector and the jacobian matrix for each iteration and solves for steady state condition
iterations.c calls "newrap.c" and checks for limit on number of iterations possible; helps keep the numerical solutions within bounds and prevent running time blowing-up of the program
constraint.c checks whether the operational constraints are met by the suitable configurations, laid down by the designer
store.c store the output of the program into a file previously opened for the purpose
wiretable.c this is the wire-gauge table read in by the program

```
\#include <stdio.h>
\#include <math.h>
\#include <ermo.h>
\#define CLEAR_ERROR 25
\#include <declare.c>
\#include <complex_math.c>
\#include <real_math.c>
\#include <circuit.c>
\#include <function.c>
\#include <jacobian.c>
\#include <newrap.c>
\#include <iterations.c>
\#include <constraint.c>
```

```
#include <store.c>
#include <wiretable.c>
main0
|
FILE *fp1 ;
char filename1[20], filename2[20] ;
int i,j, p,q;
int np, nc;/* number of turns for the power and control windings */
double np_max, nc_max, np_strt, nc_strt ; % maximum number of turns */
double v2, v2_strt, v2_max, \2_b, v2_strt_new ;
double k_low, k_hgh, len, dia, cspan6, cspan2 ;
double A, ap, ac ;/* slot area and cross-sectional area of each wire */
double ph_ang, row, sum ;
double 16_base, 12_base, m6_base, m2_base ;
float r6_base, r2_base ;
float r6dq_base, r2dq_base ;
double cur[7] ; /* real and imaginery parts of current */
double data[30] ;
float pol[36][6], polt[6][36] ;
```

wiretable0 ;

```


```

    Open file containing base parameters, operating conditions and
    ```
    Open file containing base parameters, operating conditions and
    performance criteria (named startup) and an output file
    performance criteria (named startup) and an output file
    whose name is specified by the user
    whose name is specified by the user
#**************************************************************************/
#**************************************************************************/
    if ((fp1 = fopen("startup", "rt"))=NULL)
    if ((fp1 = fopen("startup", "rt"))=NULL)
    {
    {
        print("Error : file startup could not be openedN");
        print("Error : file startup could not be openedN");
        exit(1);
        exit(1);
    }
    }
        else
        else
        print("opened file startup.\n") ;
        print("opened file startup.\n") ;
    print("\n\\\Enter file name for writing stator configurations: ");
    print("\n\\\Enter file name for writing stator configurations: ");
    scanf("%ls", filename2) ;
    scanf("%ls", filename2) ;
    fp2 = fopen(filename2, "w+");
    fp2 = fopen(filename2, "w+");
    print("opened file %s.ln",filename2) ;
    print("opened file %s.ln",filename2) ;
    row = 17.24137931e-09;
    row = 17.24137931e-09;
/***********Read in file "startup"**************************/
/***********Read in file "startup"**************************/
    for (i=0; i<24; ++i)
    for (i=0; i<24; ++i)
    f fscanf(fp1,"%1f", &data[i]);
    f fscanf(fp1,"%1f", &data[i]);
        fscanf(fp1,"%*[^\n]") ;
        fscanf(fp1,"%*[^\n]") ;
    }
    }
/**********Read in wiretable**************************/
```

/**********Read in wiretable**************************/

```
```

len = data[0] ;
dia = data[1];
16_base = data[2] ;
12_base = data[3] ;
m6_base = data[4] ;
m2_base = data[5] ;
m= data[6] ;
lr = data[7];
v6 = data[8] ;
k_low = data[9];
k_hgh = data[10];
tor_1 = data[11];
speed = data[12];
pf6_min = data[13] ;
effi_min= data[14];
np_strt = data[15] ;
nc_strt = data[16] ;
np_max = data[17];
nc_max = data[18];
A = data[19] ;
cspan6 = data[20];
cspan2 = data[21];
count_max=data[22];
ph_ang = data[23];
(void)fclose(fp1) ;

```
/***********identify the winding structure using "rdpol.f"***********/
rdpol(pol,polt) ;
/**********calculate different slips, 2-pole voltage range************/
    \(\mathrm{f} 6=60.0\);
    fr \(=\) speed/60;
    \(\mathrm{f} 2=\mathrm{f} 6-\mathbf{4}^{*} \mathrm{fr}\);
    w6 = \(2^{*} \mathrm{pi}^{*}\) f6 ;
    s1 = (f6-3*fr)/f6 ;
    \(\mathbf{s}=\mathbf{f} 2 / \mathrm{f} 6\);
    v2_strt \(=\) (v6*labs(s)-k_low);
    if ( \(\mathbf{V}\) _strt \(<=0.0\) )
        v2_strt \(=1.0\);
    v2_max \(=\) (v6*labs(s)+k_hgh) ;
    if ( V 2 _max \(>=(\mathrm{v} 6-20)\) )
        v2_max = v6-20;
    if \((s<0.0)\)
        ph_ang += 180 ;
/*
    \(\mathrm{v} 2 \_\)inc \(=\left(-\mathrm{v} \mathbf{2}_{-}\right.\)strt \(+\mathrm{v} \mathbf{2}_{-}\)max \() / 5\);
/*********begin sweeping over number of turns, wire-gauge sizes and 2-pole voltage \({ }^{* * * * * * * * * * * * * * * * * * * * * * * * * * * * * * * * * * * * * * * / ~}\)
```

|******* 2-pole number of tums **********/
for (nc=nc_strt; nc<=nc_max; ++nc)
| ls2 = 12_base*nc*nc ;
m2 = m2_base*nc ;
/********* 6-pole number of turns *************/
for (np=np_strt; np<=np_max; ++np)
| ls6 = 16_base*np*np ;
m6 = m6_base*np ;
/********* wire-gauge on 2-pole *********/
for (q=0;q<20; ++q)
{ ac = pi*wire[q].d*wire[q].d**0.000001/4;
/********* wire-gauge on 6-pole *********/
for ( }\textrm{p}=0;\textrm{p}<20; ++p
{ ap = pi**ire[p].d* wire[p].d*0.000001/4 ;
/******* Check for slot-fill condition ***********/
if (((np*ap+nc*ac)>=0.6*A) \&\& ((np*ap+nc*ac)<=0.7*A))
1 r6_base = row*(2*len+4*(pi*dia*cspan6/360))/ap ;
12_base = row*(2*len+3*(pi*dia*cspan2/360))/ac ;
print("ap=%e\np=%dNp=%dn",ap,np,p);
print("ac=%elunc=%dtq=%d\n",ac,nc,q);
printf("area=%e\n",np*ap+nc*ac);
printf("r6_base=%e<br>2_base=%e\n",r6_base,r2_base);
gensr(\&r6_base,\&r2_base,pol,polt,\&r6dq_base,\&r2dq_base);
r6 = r6dq_base * np ;
r2 = r2dq_base * nc ;
printf("r6 = %ftr2 = %fn",r6,r2);
/******* use coarse 2-pole voltage increments ******/
for (v2_b=v2_strt; v2_b<=v2_max; v2_b += 20)
{ note =0;
iterations(v2_b, ph_ang, cur);
ermo = CLEAR_ERROR;/* error handling */
if (note=1)
{ if (20 > v2_b)
v2_strt_new = 1.0;
else
v2_strt_new = v2_b - 20.0;

```
/******** use fine 2-pole voltage increments ******/
                                    for ( \(\mathrm{v} 2=\mathrm{v} 2\) _strt_new; \(\mathrm{v} 2<=\mathrm{v} 2 \_\)max; \(\mathrm{v} 2+=1\) )
                    1 note \(=0\);
                    accept \(=0\);
                    iterations(v2,ph_ang, cur);
                        ermo = CLEAR_ERROR ;
                        if (note \(=1\) )
                        constraint(cur,p,q) ;

\section*{if (accept \(=1\) )}
store(v2,np,nc,p,q) ;
)
v2_b = v2_max + 1 ;
\}
)
)
\}
\}
    \}
)
(void)fclose(fp2) ;

\section*{declare.c}
```

typedef struct {
double re;
double im ;
| complex ;
typedef struct {
ints;
float cap;
float d;
} copper;
FILE *fp2;
int note, accept ;
double F[7], V[2], pi ;
double f6, f2, fr, w6, s1, s;
double v6, count_max ;
double r6, r2, rr, ls6, 1s2, Ir, m6, m2 ;
double tor_1, tor_e, speed, pf6_min, pf6, effi_min, effi1, effi2 ;
double Iq6, Iq2, Iqr ;
double xs6, xs2, xm6, xm2, xr ;
double Pr6loss, Pr2loss, Prrloss, optim, inpu12, inpu2q, ang6 ;
copper wire[30] ;

```

\section*{complex_math.c}
```

/************************************************************************
Library of functions for operation on complex numbers and matrices. The names of the functions explains the operation they program.
complex assign_values(real_part, imag_part)
double real_part;
double imag_part;
{
complex y;
y.re = real_part;
y.im = imag_part;
return(y);
}
complex cmplx_add(var1, var2)
complex var1;
complex var2;
{
complex var3;
var3.re = var1.re + var2.re ;
var3.im = var1.im + var2.im ;
return(var3);
}
complex cmplx_sub(var1, var2)
complex var1;
complex var2;
{
complex var3 ;
var3.re = var1.re - var2.re;
var3.im = var1.im - var2.im ;
return(var3) ;
}
complex cmplx_mulply(var1, var2)
complex varl ;
complex var2 ;
{
complex var3 ;
var3.re = var1.re * var2.re - var1.im * var2.im ;

```
```

    var3.im = varl.re * var2.im + var1.im * var2.re ;
    return(var3);
    }
complex multply_by_j(varl)
complex var1;
{
complex var2;
var2.re = -varl.im ;
var2.im = varl.re ;
return(var2);
}
complex real_cmplx_add(var1, var2)
double varl;
complex var2 ;
{
complex var3, var4;
var4.re = var1;
var4.im = 0;
var3 = cmplx_add(var2, var4);
return(var3);
}
complex real_cmplx_multply(var1, var2)
double varl ;
complex var2 ;
{
complex var3;
var3.re = varl * var2.re ;
var3.im = varl * var2.im ;
return(var3) ;
}
complex conjugate(var1)
complex varl ;
{
complex var2 ;
var2.re = var1.re ;
var2.im = -var1.im ;
return(var2) ;
}

```
```

complex cmplx_div(var1, var2)
complex var1, var2 ;
{
complex var3, var4, var5 ;
var4 = cmplx_mulply(var2, conjugate(var2)) ;
var5 = cmplx_multply(var1, conjugate(var2)) ;
var3 = real_cmplx_mulply(1/var4.re, var5) ;
return(var3) ;
}
complex real_cmplx_div(var1, var2)
double var1;
complex var2 ;
{
complex var3, var4;
var4.re = varl ;
var4.im =0;
var3 = cmplx_div(var4, var2) ;
retum(var3) ;
}
complex inv(var2)
complex var2 ;
{
complex var3, var4 ;
var4.re = 1;
var4.im = 0;
var3 = cmplx_div(var4, var2) ;
return(var3) ;
J

```

\section*{/*************************************************************************}

Complex matrix vector multiplication function
The integer arguments for the function are the number of rows and columns for the matrix.
```

void cmplx_matrix_vector_mulply(A,B,C, row_of_A, col_of_A)
int row_of_A, col_of_A;
complex A[[2], B[, CD ;
1
int i, j;
for (i=0 ; i< row_of_A; ++i)
l C[i].re = 0;
C[i].im = 0;
for ( }\textrm{j}=0;\textrm{j}< col_of_A; ++j
C[i] = cmplx_add(C[i],cmplx_mulply(A[i][j],B[j]);
}
}

```
/***************************************************************************

Complex matrix multiplication function
The integer arguments for the function are the number of rows and columns for the first matrix and the number of columns for the second matrix.
void cmplx_matrix_multply(A,B,C, row_of_A, col_of_A, col_of_B) int row_of_A, col_of_A, col_of_B ; complex \(\mathrm{A}[[7], \mathrm{B}[][7], \mathrm{C}[77]\);
\{
int \(\mathbf{i}, \mathbf{j}, \mathbf{k}\);
for ( \(\mathrm{i}=\mathbf{0}\); \(\mathrm{i}<\) row_of_A; ++i)
for ( \(\mathbf{j}=\mathbf{0} ; \mathbf{j}<\) col_of_B; +j )
\(1 \quad \mathrm{C}[\mathrm{i}][\mathrm{j}] \mathrm{re}=0\);
\(\mathrm{C}[1][j] . \mathrm{im}=0\);
for ( \(k=0 ; \mathbf{k}<\) col_of_A; ++k)
\(\mathbf{C}[i][j]=\) cmplx_add(C[i][j],cmplx_multply(A[i][k],B[k][j]));
\}
\}

\section*{real_math.c}
```

/**************************************************************************
Functions for operations on real matrices. The names of the functions explains the operation they program. "labs" determines the absolute value of a floating point number

```

```

double labs(varl)
double varl ;
{
double var2;
if (varl < 0.0)
var2 = -var1 ;
else
var2 = var1 ;
reurn(var2);
}
void transpose(A, B, row_of_A, col_of_A) /* B is the transposed matrix */
double A[[7], B[[7] ;
int row_of_A, col_of_A ;
{
int i,j;
for (i=0; i<row_of_A; ++i)
for ( }=0;\mp@code{j<col_of_A; ++j)
B[i][j] = A[j][i] ;
}
void vecadd(A, B, C, row_of_A)
int row_of_A;
double A[], }\textrm{B}[,\textrm{C}[\mathrm{ ;
{
int i;
for (i=0; i<row_of_A; ++i)
C[i] = A[i] + B[i] ;
}
void vecsub(A, B, C, row_of_A)
int row_of_A ;
double A[], B[, CD ;
I
int i;
for (i=0; i<row_of_A; ++i)

```
\[
\mathrm{C}[\mathrm{i}]=\mathrm{A}[\mathrm{i}]-\mathrm{B}[\mathrm{i}] ;
\]
```

|***************************************************************************
Real matrix vector multiplication function
The integer arguments for the function are the number of rows and columns for the matrix.

```
void mat_vec_multply(A, B, C, row_of_A, col_of_A)
int row_of_A, col_of_A ;
double \(\mathrm{A}[77], \mathrm{B}[\mathrm{C}, \mathrm{C}\);
1
int \(\mathbf{i}, \mathbf{j}\);
for ( \(\mathrm{i}=\mathbf{0}\); \(\mathrm{i}<\) row_of_A; + +i)
\(1 \quad \mathrm{C}[\mathrm{i}]=0\); for ( \(\mathbf{j}=\mathbf{0} ; \mathrm{j}<\) col_of_A; ++j) \(\mathrm{C}[\mathrm{i}]+=\mathrm{A}[\mathrm{i}][\mathrm{j}]\) * \(\mathrm{B}[\mathrm{j}]\);
\}
\}

\section*{/***************************************************************************}

Real matrix multiplication function
The integer arguments for the function are the number of rows and columns for the first matrix and the number of columns for the second matrix.
*************************************************************************/
void mat_multply(A, B, C, row_of_A, col_of_A, col_of_B) int row_of_A, col_of_A, col_of_B ;
double \(\mathrm{A}[77], \mathrm{B}[77], \mathrm{CD}[7]\);
1
int \(\mathbf{i}, \mathbf{j}, \mathrm{k}\);
for ( \(\mathrm{i}=0\); \(\mathrm{i}<\) row_of_A; ++i)
for ( \(\mathrm{j}=0 ; \mathrm{j}<\) col_of_B; +j )
\(1 \quad \mathrm{C}[\mathrm{i}][\mathrm{j}]=0\);
for ( \(k=0 ; k<\) col_of \(A ;++k)\)
\(\mathrm{C}[\mathrm{i}][\mathrm{j}]+=\mathrm{A}[\mathrm{i}][\mathrm{k}]\) * \(\mathrm{B}[\mathrm{k}][\mathrm{j}]\);
)
)

\section*{circuit.c}

```

Routine to determine the initial guess for the Newton-Raphson iteration method of solving equations for the steady state BDFM model. This involves an initial guess for the angle "beta"; the angle between the 6-pole and 2-pole voltage phasors.

```
```

void circuit(beta_in, I)
double beta_in ;
double II ;
1
int i ;
complex z6, z2, zr, zm6, zm2, c_s ;
complex param_a, param_b, param_c, param_d, param_e ;
complex param11, param12, param21, param22 ;
complex param[2][2], I62r[3], V_psr[2] ;
V _psr[0].re $=\mathrm{V}[0]$;
V_psr[0].im =0;
V_ps[[1].re $=\mathrm{V}[1]^{*} \cos \left(\mathrm{pi}^{*}\right.$ beta_in/180) ;
V_psr[1].im = -V[1]*sin(pi*beta_in/180) ;
/************Calculate reactances from the inductances calculated in********
******************* "optim.c". They are then transformed into complex********
**************** impedances using the complex function. ******************/
xs6 $=\mathbf{w 6}{ }^{*} 1 \mathrm{ls} 6$;
xs2 $=\mathrm{w} 6$ *ls2 ;
$\mathrm{xm6}=\mathrm{w} 6 * \mathrm{~m} 6$;
$\mathrm{xm} 2=\mathrm{w} 6^{*} \mathrm{~m} 2$;
$\mathbf{x r}=\mathbf{w} \mathbf{6}^{*} \mathbf{l r}$;
z6 = assign_values(r6, xs6) ;
z2 = assign_values(r2/s, xs2) ;
zr = assign_values(ri/s1, xr) ;
zm6= assign_values(0.0, xm6) ;
zm2 $=$ assign_values $(0.0,-\mathrm{xm} 2)$;
c_s= assign_values(s, 0.0 ) ;
/*********************** Common winding BDFM parameters ***************/
/*
16 $=0.806667$;
12 $=0.806667$;
$\mathrm{Ir}=327.5 \mathrm{e}-06$;
xs $6=30.00849$;
x $\mathrm{s} 2=237.0671$;
$\mathrm{xr}=0.015708$;
$\mathrm{xm6}=0.334382$;
$\mathrm{xm} 2=1.622758$;

```
```

    m6 = xm6/w6;
    m2 = xm2/w6;
    z6 = assign_values(r6,30.008490) ;
    z2 = assign_values(r2/s,237.067100) ;
    zr = assign_values(327.5e-06/s1,0.015708) ;
    zm6= assign_values(0.0, 0.334382) ;
    zm2= assign_values(0.0,-1.622758) ;
    c_s= assign_values(s, 0.0) ;
    */

```
/******* The model without the torque equation
    param_a \(=\) cmplx_multply(cmplx_div(zm6,z6),V_psr[0]) ;
    param_b = cmplx_multply(cmplx_div(zm2,z2),cmplx_div(V_psr[1],c_s)) ;
    param_c = cmplx_div(cmplx_multply(zm6,zm6),z6) ;
    param_d = cmplx_div(cmplx_multply(zm2,zm2),z2) ;
    param_e \(=\) cmplx_sub(cmplx_add(param_c,param_d),zr) ;
    162r[2] = cmplx_div(cmplx_add(param_a,param_b),param_e) ;
    I62r[1] = cmplx_div(cmplx_sub(cmplx_div(V_psr[1],c_s),cmplx_multply(zm2,I62r[2])),z2);
    I62r[0] = cmplx_div(cmplx_sub(V_psr[0],cmplx_multply(zm6,162r[2])),z6) ;
/****** Initial guess for the currents

    \(\mathrm{I}[0]=\mathrm{I} 62 \mathrm{r}[0]\).re ;
    \(\mathrm{I}[1]=\mathrm{I} 62 \mathrm{r}[0] . \mathrm{im}\);
    \(\mathrm{I}[2]=\mathrm{I} 62 \mathrm{r}[1] . \mathrm{re}\);
    \(\mathrm{I}[3]=162 \mathrm{r}[1] . \mathrm{im}\);
    \(\mathrm{I}[4]=\mathrm{I} 62 \mathrm{r}[2]\).re ;
    \(\mathrm{I}[5]=\mathrm{I} 62 \mathrm{r}[2] \mathrm{im}\);
    \(\mathrm{I}[6]=\) beta_in*pi/180 ;
)

\section*{iterations.c}
```

void iterations(volt, ini_ang, c)
double ini_ang, volt ;
double c[7] ;
{
int i, j, count;
double sum ;
double c1[7] ;
V[0]= 0.8660254038*v6;
V[1]= 0.8660254038*volt ;
circuit(ini_ang, c);
newrap(c,c1,7) ;
sum = 0.0;
for (i=0; i<7; ++i)
sum += labs(F[i]);
count =0;
while ((sum > 0.0001)\&\&(count <= count_max)\&\&(ermo!=ERANGE))
| for ( }=0;0;j<7;++j
c[j] = cl[j] ;
newrap(c,c1,7) ;
sum = 0.0;
for (i=0; i<7; ++i)
sum += labs(F[i]);
++count;
/* print("count = %\&N sum = %e\n",count,sum);
*/
}
for (j=0; j<7; + j)
c[j] = c1[j];
if ((coun\count_max) | (ermo == ERANGE))
| /*printf("Iteration limit exceededn"); */
note = 0;
}
else
note = 1;
]

```

\section*{newrap.c}

```

This routine updates the function vector and the jacobian matrix during each iteration. This also handles error codes generated by the compiler during run-time.

```
```

void newrap(X, X1, dim) /*dim is the order of the system*/
double X[, X1[ ;
int dim ;
{
function(X);
jacobian(X, J) ;
if (ermo != ERANGE) /*** if there is no error, so far, then... ***/
{
transpose(J, JT, dim, dim) ;
matinv(invJT, JT, \&dim) ;
transpose(invJT, invJ, dim, dim) ;
mat_vec_multply(invJ, F, invJF, dim, dim) ;
vecsub(X, invJF, X1, dim);
|
|

```

\section*{function.c}
```

void function(Ib)
double Ib [ ;
1
int i ;
$\mathrm{F}[0]=\mathrm{r} 6^{*} \mathrm{Ib}[0]-\mathrm{xs} 6^{*} \mathrm{Ib}[1]-\mathrm{xm} 6^{*} \mathrm{Ib}[5]-\mathrm{V}[0] ;$
$\mathrm{F}[1]=\mathrm{r} \mathrm{b}^{*} \mathrm{Ib}[1]+\mathrm{xs} 6^{*} \mathrm{Ib}[0]+\mathrm{xm} 6^{*} \mathrm{Ib}[4]$;
$\mathrm{F}[2]=\mathrm{r} 2^{*} \mathrm{Ib}[2] / \mathrm{s}-\mathrm{xs} 2^{*} \mathrm{Ib}[3]+\mathrm{xm} 2^{*} \mathrm{Ib}[5]-\mathrm{V}[1]^{*} \cos (\mathrm{lb}[6]) / \mathrm{s}$;
$\mathrm{F}[3]=\mathrm{r} 2^{*} \mathrm{Ib}[3] / \mathrm{s}+\mathrm{xs} 2^{*} \mathrm{Ib}[2]-\mathrm{xm} 2^{*} \mathrm{lb}[4]+\mathrm{V}[1]^{*} \sin (\mathrm{lb}[6]) / \mathrm{s}$;
$\mathrm{F}[4]=-\mathrm{xm} 6^{*} \mathrm{Ib}[1]+\mathrm{xm} 2^{*} \mathrm{Ib}[3]+(\mathrm{m} / \mathrm{s} 1)^{*} \mathrm{Ib}[4]-\mathrm{xr} \mathrm{r}^{*} \mathrm{Ib}[5]$;
$\mathrm{F}[5]=\mathrm{xm}^{*} \mathrm{Ib}[0]-\mathrm{xm} 2^{*} \mathrm{Ib}[2]+(\mathrm{rr} / \mathrm{s} 1)^{*} \mathrm{Ib}[5]+\mathrm{xr} \mathrm{r}^{*} \mathrm{Ib}[4]$;
$\mathrm{F}[6]=6^{*} \mathrm{~m} 6^{*}\left(\mathrm{Ib}[1]^{*} \mathrm{Ib}[4]-\mathrm{Ib}[0]^{*} \mathrm{Ib}[5]\right)+2^{*} \mathrm{~m} 2^{*}\left(\mathrm{Ib}[3]^{*} \mathrm{Ib}[4]-\mathrm{Ib}[2]^{*} \mathrm{lb}[5]\right)-\operatorname{tor} \_$;
\}

```

\section*{jacobian.c}
```

void jacobian(lb, par)
double Ib[], par[[7] ;
{
int i, j;
for (i=0; i<7; ++i)
for (j=0; j<7; ++j)
par[i][j] = 0.0;
par[0][0] = 16 ;
par[0][1] = -xs6 ;
par[0][5] = -xm6;
par[1][0] = xs6 ;
par[1][1] = 16;
par[1][4] = xm6 ;
par[2][2] = r2/s ;
par[2][3] = -xs2 ;
par[2][5] = xm2 ;
par[2][6] = (V[1]/s)*sin(Ib[6]) ;
par[3][2] = xs2;
par[3][3] = r2/s ;
par[3][4] = -xm2 ;
par[3][6] = (V[1]/s)*}\operatorname{cos(lb[6]);
par[4][1] = -xm6;
par[4][3] = xm2 ;
par[4][4] = r//s1 ;
par[4][5] = -xr ;
par[5][0] = xm6 ;
par[5][2] = -xm2 ;
par[5][4] = xr ;
par[5][5] = rr/s1 ;
par[6][0] = -6*m6*Ib[5] ;
par[6][1] = 6*m6*Ib[4];
par[6][2] = -2*m2*Ib[5] ;
par[6][3] = 2*m2*Ib[4] ;
par[6][4] = 6**m6*Ib[1] + 2*m2*Ib[3];
par[6][5] = -6*m6*Ib[0] - 2*m2*Ib[2];

```

\section*{constraint.c}
```

void constraint(I, type6, type2)
double ID ;
int type6, type2 ;
{
double input6, input2, output ;
int i;
Iq6 = sqrt(I[0]*I[0] + I[1]*I[1]);
Iq2 = sqt(I[2]*I[2] + I[3]*I[3]);
Iqr = sqrt(I[4]*I[4] + I[5]*I[5]);
output = tor_l* }\mp@subsup{2}{}{*}\mp@subsup{\textrm{p}}{}{*}\mp@subsup{}{}{*}\mathrm{ speed/60 ;
pf6 = cos(atan(I[1]/[0])) ;
input6 = 2*V[0]*Iq6*pf6 ;
if (I[2] >= 0)
{ if (I[3]>=0)
1/*print("Quadrant Nn");*/
input2 = 2*V[1]*Iq2*}\operatorname{cos}(\operatorname{atan(labs(I[3]/[[2]))+I[6]) ;
}
else if (I[3] < 0)
| /*prinf("Quadrant IVn") ;*/
input2 = 2*V[1]*Iq2*}\operatorname{cos}(\operatorname{atan}(labs(I[3]/[[2]))-I[6])
]
}
else if (1[2] < 0)
{ if (I[3]>=0)
l/*print("Quadrant INn");*/
input2 = 2*V[1]*Iq2*}\operatorname{cos}(pi+atan(labs(I[3]/[2]))-I[6])
}
else if (I[3] < 0)
| /*printf("Quadrant IIIn") ;*/
input2 = 2*V[1]*Iq2* cos(pi-atan(labs(I[3]//[2]))-I[6]);
J
}
Pr6loss = 2*Iq6*Iq6*r6;
Pr2loss=2*Iq2*Iq2*12;
Prrloss = 2*Iqr*'Iqr*}\pi\mathrm{ ;
optim = Pr6loss + Pr2loss + Prrloss ;
effi1 = output*100/(input6+input2);
effi2 =output*100/(output+optim);
print("effil = %7.4^teffi2 = %7.4^Nn",effi1, effi2);
printf("Iq6 = %e\Iq2 = %e\Iqr = %e\n", Iq6,Iq2,Iqr) ;
print("pf6 = %e\n", pf6) ;
print("effi2 = %e<br>", effi2);

```
    if (I[1] < 0.0)
    if (pf6 >= pf6_min)
    | if (Iq2 <= 2*1.224744871*wire[type2].cap)
        { if (Iq6 <= 2*1.224744871*wire[type6].cap)
        { if (effi2 >=effi_min)
                        accept =1;
            else
                        accept = 0;
            }
        }
    }
/* print("%fn",sqrt(3/2));
    print("%fN",2*1.224744871*wire[type2].cap);
    print("%An",2*sqrt(3/2)*wire[type6].cap);
    print("accept = %dvn", accept) ;
    return(accept);
    print("I[1] = %ANn", I[1]) ;
    accept = 1;
*/
}
```


## store.c

```
void store(p2v, n6, n2, type6, type2)
double p 2 v ;
int n6, n2, type6, type2 ;
(
fprintf(fp2,"\%10.4f \%3.0f \(\%\) d \(\%\) d \(\%\) d \(\%\) d \(\% 8.3 f \quad \% 8.3 f \quad \% 8.3 f \quad \% 7.4 f\)
\%6.4fn",optim,p2v,n6,wire[type6].s,n2,wire[type2].s,Pr6loss,Pr2loss,Priloss,effi1,pf6) ;
fflush(fp2) ;
)
```

```
/* the copper wire table */
void wiretable0
{
/*a.w.g. size*/
wire[0].s = 6;
wire[1].s = 7;
wire[2].s = 8;
wire[3].s = 9;
wire[4].s=10;
wire[5].s = 11;
wire[6].s=12;
wire[7].s = 13;
wire[8].s=14;
wire[9].s=15;
wire[10].s= 16;
wire[11].s= 17;
wire[12]. }=18\mathrm{ ;
wire[13] }s=19\mathrm{ ;
wire[14].s= 20;
wire[15].s= 21;
wire[16].s= 22;
wire[17].s= 23;
wire[18].s= 24;
wire[19].s= 25;
```

/* current capacity */
wire[0].cap $=37.5$;
wire[1].cap $=29.7$;
wire[2].cap $=23.6$;
wire[3].cap $=18.7$;
wire[4].cap $=14.8$;
wire[5].cap $=11.8$;
wire[6].cap $=9.33$;
wire[7].cap $=7.40$;
wire[8].cap $=5.87$;
wire[9].cap $=4.65$;
wire[10].cap $=3.69$;
wire[11].cap $=2.93$;
wire[12].cap $=2.32$;
wire[13].cap $=1.84$;
wire[14].cap $=1.46$;
wire[15].cap $=1.16$;
wire[16].cap = . 918 ;
wire[17].cap =. 728 ;
wire[18].cap $=.577$;
wire[19].cap $=.458$;

## /* diameter in mm */

wire[0].d $=4.115$;
wire[1].d $=3.665$;
wire[2].d $=3.264$;
wire[3].d $=2.906$;
wire[4].d $=2.588$;
wire[5].d $=2.305$;
wire[6].d $=2.053$;
wire[7].d $=1.828$;
wire[8].d $=1.628$;
wire[9].d $=1.450$;
wire[10].d $=1.291$;
wire[11].d = 1.150 ;
wire[12].d $=1.024$;
wire[13].d $=0.912$;
wire[14].d $=0.812$;
wire[15].d $=0.723$;
wire[16].d $=0.644$;
wire[17].d $=0.573$;
wire[18].d $=0.511$; wire[19]. $\mathrm{d}=0.455$;
$\mathrm{pi}=3.141592654$;

startup<br>(sample data file)

$0.075 \quad / *$ stack length */
0.21
1.155882e-04
9.690814e-04
3.840536e-05
$1.909421 \mathrm{e}-04$
$327.5 \mathrm{e}-06$
41.668e-06

230
50
50
45
850
0.9

87
20
20
40
40
200.823e-06

60
180
20
20
/* stack length */
/* machine diameter */
/* six-pole base inductance */
/* two-pole base inductance */
/* six-pole to rotor mutual base inductance */
/* two-pole to rotor mutual base inductance */
/* rotor resistance */
/* rotor inductance */
/* six-pole rms line-to-line voltage */
/* lower offset on two-pole rms line-to-line voltage */
/* upper offset on two-pole ms line-to-line voltage */
/* load torque */
/* speed */
/* minimum six-pole p.f. required */
/* minimum efficiency required */
/* starting number of turns for the six-pole */
/* starting number of turns for the two-pole */
/* maximum number of turns for the six-pole */
/* maximum number of turns for the two-pole */
/* stator slot cross-sectional area */
/* span of six-pole winding*/
/* span of two-pole winding */
/* maximum number of iterations allowed for convergance */
/* initial guess of angle between 2 -pole and 6 -pole voltages */


[^0]:    ${ }^{1}$ Device stresses are for ideal situations; circuit and device characteristics, snubber effects, and safety factors need to be considered.
    ${ }^{2} \mathrm{~V}_{\mathrm{H}=\mathrm{s}}=$ peak output line-to-line voltage.
    ${ }^{3} \mathrm{H}_{\text {pem }}=$ peak phase current.
    ${ }^{4} V_{\mathrm{d}}=\mathrm{dc}$ link voltage; $\mathrm{V}_{1}=$ peak input line-to-line voltage; $\mathrm{V}_{\mathrm{p}}=$ peak resonant de link voltage.
    ${ }^{3}$ The clamping factor, $k(1<k<2)$, can significantly reduce the device voltage, while only marginally affecting the output voltage.

    - Output voltage is limited by allowable component stresses.

