DESIGN OF A NOVEL CONTROLLER FOR A SMALL BATTERY CHARGING WIND TURBINE

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Small battery charging wind turbines are used to provide electric power in locations where grid connection would not be feasible. However, existing controllers for such small wind turbines are often inefficient and unreliable. This work therefore examines the application of the little used Cuk converter to charge efficiently a battery bank from a 2kW permanent magnet generator. This circuit was chosen because the buck-boost action allows battery charging over a wide range of windspeeds, and neither the input or output currents are discontinuous. A controlled resistive dump load both safely loads the wind turbine when the battery bank is unable to accept all the power available from the generator, and also provides a low grade source of heat.

A development of the basic circuit is to consider the use of self-regulating blades, which through safely limiting the maximum available generator power and controller input voltage, allows the significant simplification of the control electronics.

The final stage of the work is the development of a PC-based simulator for rapidly evaluating the effect of different turbine designs or control strategies under a range of actual recorded wind conditions. A test bed is described on which a motor, controlled by the PC, mimics the effect of the wind turbine driving an actual small generator.
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CHAPTER 1
INTRODUCTION

1 WINDPOWER
1.1 Development Of Windpower

Windpower has been used in England since at least the twelfth century for grinding grain, and more recently for land drainage, (Lipman, Musgrove & Pontin, 1982). The earliest windmills were wooden post mills which evolved into the smock mill, and finally to the brick or stone built tower mill, (Vince, 1987). Better control and reliability of steam power led to most windmills in the UK being made redundant by the mid 1920s, but more recent increases in the price of fuel, and also wider environmental concerns, have led to a new interest in the use of windpower as a renewable energy source for the generation of electricity.

In contrast to such grid connected wind turbines, Small Battery Charging Wind Generators (SBCWGs) are designed for use in applications where it is not feasible to have a mains connection, such as isolated dwellings or caravans. The design of these small wind turbines has improved considerably since early experiences (Piggott, 1983), but little work has been documented on the design of the electrical controller. This research project is based around the use of modern power electronics to overcome the limitations of the techniques which are currently used to control SBCWGs.
1.2 WIND ENERGY AND TURBINE DESIGN

The power due to the Kinetic Energy of a steady airstream of velocity $v$, acting over an area $A$, with a density $\rho$, is given by:

$$\text{Power} = 0.5\rho Av^3$$

For a wind turbine of swept area $A$, kinetic energy and momentum considerations mean that only 59.3% of the energy can be extracted, a figure which is known as the Betz criteria. The turbine power is thus given by:

$$\text{Turbine power} = 0.5\rho C_P Av^3$$

(where $C_P$ is the coefficient of power, a measure of the efficiency)

Since for ideal power matching the shaft speed ($\omega$) is proportional to the wind speed, the relationship between the maximum torque ($T_{\text{max}}$) and wind speed is given by:

$$T_{\text{max}} = k_1 v^2$$

Figure 1.1 shows the Power and Torque curves for an ideal wind turbine as a function of shaft speed.

The tip-speed ratio ($\lambda$) of a turbine gives the ratio of the blade tip-speed to the windspeed:

$$\lambda = \omega R / v$$

The optimum tip speed ratio (figure 1.2) gives the point of peak turbine efficiency. When operating at the optimum tip-speed ratio, $C_P$ will also be at a maximum.
Fig. 1.1 Optimum wind turbine power and torque as a function of shaft speed.

Fig. 1.2 Wind turbine $C_p$ characteristics.

Fig. 1.3 Drawing of a "standard" 910mm diameter 130W SBCWG
In the design of Small Battery Charging Wind Generators (SBCWGs), where there is no mains back-up supply, power availability is of greater importance than total power generated. This means that systems are designed to start generating power even in very low windspeeds (as little as 3 m/s). Most horizontal axis SBCWGs range in output between 20W and 10kW, (10m/s rating), with a typical 150W SBCWG shown in figure 1.3.

### 1.3 SBCWG SYSTEMS

#### 1.3.1 Context

SBCWG systems store energy in batteries, allowing energy to be stored for use during intervals of low or no wind, or at any time when the power demand exceeds that instantaneously available from the wind turbine. Here it is the ability to reliably provide power to appliances where a grid connection is not practical that is important, rather than the cost of the electricity generated, (Rapson & Grainger, 1989, Smith & Gwillim, 1995). Typical sites include boats, static and mobile caravans, remote homes, telemetry installations, disaster relief stations and farms. The power is used for a variety of loads, including lights, communication equipment, background heating, water pumps and refrigeration equipment. Battery to mains inverters are widely used, but the conversion process represents an additional energy loss, and so wherever possible it is best practice to use specialist battery-powered equipment.

Small Wind Turbine Generators (SWTGs) feeding local autonomous grids, supplying for instance small isolated communities, may comprise a significant part of the grid generation capacity, with additional power being provided by a diesel generator set, solar panels, or micro-hydro generators - a mix which is particularly important during prolonged times of low windspeed.
1.3.2 Design Of SBCWG Installations

A cutaway diagram of a typical 150W SBCWG (the "Rutland 150W Windcharger") is illustrated in figure 1.4, showing the simplicity and robustness of construction. Figure 1.5 shows the current output of this particular machine into a constant voltage battery, with the high windspeed (tail flap furling) power-limiting mechanism being described in chapter 4.2, (Falkner, 1995).

Figure 1.6 illustrates the simplest possible application with the SBCWG connected directly to the battery, and just a voltage regulator for battery protection, (Falkner & Fawkes, 1992).

Figure 1.7 illustrates a more complex application with a controller being used for both approximately matching the generator power demand to the windspeed, and for basic battery management through the use of a dump load resistor for dissipating surplus power. This system includes inputs from a diesel driven alternator and a solar panel, and a splitter diode arrangement for charging two battery banks.

1.3.3 Energy Storage

Battery Storage is by far the most popular method of decoupling the instantaneous load power demand from the instantaneous power available from the wind, since the electrical energy can be directly used as the power source for many pieces of equipment, and installation and maintenance are relatively simple. Vented Lead-Acid or Nickel Cadmium batteries are usual for large installations, although Valve Regulated Lead Acid (VRLA) batteries are becoming more common on smaller systems, (Chaurey, A. & Deambi, S., 1992.) See appendix 2 for further discussion on batteries used for remote energy applications.
Fig. 1.4 Cutaway view of a 910mm diameter 150W SBCWG, with furling tail for high wind speed protection

Fig. 1.5 Output characteristics of the 150W battery charging wind turbine shown above
Fig. 1.6 Simplest battery charging wind turbine system.

Fig. 1.7 Multi-source remote energy supply.
Less common methods of energy storage include:

- Hydrogen Storage, where electrical energy is used to electrolyse water, evolving hydrogen as one of the products, which can then be stored and used as fuel as needed.
- Water Heating, where water is heated through mechanical churning or electrical heating, and is then stored in well-insulated tanks for later use, (Twidell et al, 1989).
- Pumped Storage, where water is pumped to a high level reservoir, and then released to turn a hydro-electric generator as required.
- Flywheel Storage, where a flywheel is used for short-term mechanical energy storage. It is typically used in Wind/Diesel systems where it provides an energy reserve during lulls in the wind, reducing the frequency with which the Diesel generator must be started, (Infield, 1990).

1.4 EXISTING METHODS FOR CONTROLLING SBCWGS

1.4.1 Overview

For the generator to give maximum battery charge current over the operating windspeed range, the power drawn from the generator should be varied in proportion to the energy available from the wind. In the case of an un-controlled Permanent Magnet (PM) generator, the generator Open-Circuit Voltage (OCV) is proportional to the shaft speed, \( \omega \). Since the energy in the wind is proportional to the wind speed cubed, and assuming constant battery voltage and unity power factor, the optimum generator current is given by:

\[ I_g = k_2 \omega^3 \]

(For a resistive load, the optimum generator current is given by \( I_g = k_3 \omega^2 \).)
Currently, various electrical techniques are used for controlling the current, (Bolton, H.R. and Nicodemou, V.C., 1979), as shown in figure 1.8. While field control enables the power controller to be rated at a lower power level, the additional field winding and associated brush gear reduces both the reliability and maintenance intervals, and so a PM generator is usually favoured for SBCWGs. A 3-phase generator is desirable because of the better use of magnetic material, but this advantage must be offset against the additional down-lead and other components associated with an additional power cable.

1.4.2 Resistance Control

The electrical loading may be varied by using either a series multi-tapped resistor arrangement (figure 1.8.1), or a bank of parallel resistors (figure 1.8.2), either of which can be switched to vary the effective load resistance. Alternatively, a much wider variation in load resistance by the use of semiconductor switches for phase control is possible, allowing for much finer matching of the generator output to the load. Phase control can be implemented by either AC line-side triacs (figure 1.8.3), or a load-side thyristor-controlled bridge rectifier (figure 1.8.4). DC side connection of the resistance also allows simple DC chopper techniques to be used, using semiconductor switches operating under phase angle or Pulse Width Modulation (PWM) control. Another benefit of using resistors is that they can be used to provide background heating, and can provide a permanent load for the turbine if the battery is unable to accept sufficient charge to fully load the generator.

A simple and low-cost variation on this technique is to switch between full-wave and half-wave rectification (fig 1.8.5), a technique which is used successfully on the Rutland 20W SBCWG to reduce the output current when the generator thermostat trips, (Falkner & Fawkes, 1994).
Fig. 1.8  Electrical control of small wind turbines.

Fig. 1.8.1 Series resistor control

Fig. 1.8.2 Parallel resistor control

Fig. 1.8.3 Triac phase control

Fig. 1.8.4 Full bridge control

Fig. 1.8.5 Full/Half wave control

Fig. 1.8.6 Series reactance control

Fig. 1.8.7 Autotransformer control
1.4.3 Reactance Control

This is a similar technique to the switched resistor methods discussed above, only that inductors are used instead of resistors for loading the wind turbine generator. The Rutland 150W WTG uses a series choke to reduce the generator current and hence temperature when the generator thermostat trips, (similar to the operation of the switch shown in figure 1.8.6).

Since both the generator voltage (Vg) and inductor reactance (oL) are proportional to the shaft speed, the current is restricted to an approximately constant value of Vg/ oL.

1.4.4 Change Of Winding Configuration

Changing the winding configuration has been tried experimentally on a SBCWG (Campbell, R., 1991), in which the stator comprised two identical windings which were connected in series at low windspeeds, and in parallel for high windspeeds. However, in this case the gain in total power output over the operating windspeed range was found to be insufficient to justify the cost of additional winding changing circuitry.

1.4.5 Auto-Transformer Control

A multiple-tapped autotransformer can be used to approximately match the output voltage of the generator to that of the battery as the windspeed and hence generator voltage changes, (figure 1.8.7). The greater the number of transformer taps, the closer the power drawn will fit the turbine characteristics, and the smoother the transitions will be. German (1985) successfully used a thyristor switched five-tap auto-transformer on a Marlec 150W machine, with tap changing controlled by generator frequency. The use of an autotransformer provides a simple way of stepping-up the generator voltage to be
greater than the battery voltage when operating in low windspeed conditions. (This problem is discussed further by Smith, (1996)).

1.5 CURRENT COMMERCIAL SBCWG CONTROLLER DESIGN

Prior to starting the research into the design of the 2kW SBCWG controller, a re-design of a 500W (FM1800) Marlec wind turbine controller was undertaken, (appendix 7). This used the same basic electrical circuit as the original circuit (figure 1.9.1), but with the control electronics re-designed for greater reliability. This design work gives a practical insight into the difficulties of designing SBCWG control systems which are both efficient and reliable.

The generator on this 500W machine is a PM discoid type (Corbett, 1991), which is thermostatically protected by a choke connected in series with the generator output when the generator is too hot. The two autotransformer taps are changed by a relay, giving a two-stage approximation to the optimal power:windspeed characteristics, (figure 1.9.2). When the battery is fully charged, a dump load resistor is used to load the generator until the battery voltage has fallen to a lower voltage threshold, (figure 1.9.3).

This system has several shortcomings, all of which are overcome by the proposed 2kW SBCWG controller:

- Poor utilisation of the generator volume, since it is only a single-phase design.
- Non-ideal power:windspeed matching, particularly at the extremes of operation.
- Total loss of turbine electrical loading if just one phase connection should fail to open circuit.
Fig. 1.9 Commercial autotransformer-based WTG controller for battery charging.

Fig. 1.9.1 Electrical schematic.

Fig. 1.9.2 Close-up of tap-changing operation.

Fig. 1.9.3 Voltage control of charge: dump relay.
Once the battery is fully charged, the generator power is diverted to the dump load, and so the available capacity of the battery is gradually reduced as it is discharged by the appliances powered by the battery, (fig. 1.9.3). Charging only recommences when the battery has been sufficiently discharged for the voltage to fall below a set lower voltage limit.

- No user indication of charge current, battery voltage, or system status.

The remaining design weaknesses also lead to a sudden change in load, which imposes a sudden torque change on the generator and turbine. This is undesirable, since it increases wear on these components:

- If the battery is dis-connected, the generator Open Circuit Voltage (OCV) is both dangerous (up to 10 times nominal battery voltage), and can cause injury to the user and/or damage to the controller electronics.
- On poor batteries with a high internal impedance, the voltage falls as soon as the generator is connected to the dump load resistor, causing it to switch back to battery charging mode. This can set up an oscillation at 1 - 2 Hz.
- Thermostatic protection requires a bulky and costly protection choke to be mounted in the nacelle.

1.6 THE PROPOSED SMALL BATTERY CHARGING WIND GENERATOR SYSTEM

1.6.1 Outline Description

Figure 1.10 shows the block diagram of the basic system, with the wind turbine turning the generator rotor, the output of which is rectified to supply the variable voltage DC link. The generator power is then used for either charging batteries through the Cuk converter, or dissipated as heat by the controlled dump load. Figure 1.11 shows how in low windspeeds all the generator power is used for charging the batteries, but at higher windspeeds, if the battery is already well charged, some of the power can be diverted to
Fig. 1.10 2kW Battery charging wind turbine controller - block diagram

Fig. 1.11 The split of the available generator power between battery charging and the dump load

Fig. 1.12 2kW Battery charging wind turbine controller - high level schematic
the dump load. The control system was originally intended to be based on a 68HC11 microcontroller, but the later decision by the sponsor to use flexible blades meant that this was no longer essential, and so lower cost control solutions using discrete analogue and digital electronics are described. Figure 1.12 shows in more detail how the circuit was implemented.

1.6.2 Novel Aspects

The SBCWG controller is based around the relatively recent Cuk converter topology, which has to date been documented as being used almost exclusively in specialist aerospace and space applications. This topology was chosen because it allows buck:boost conversion using just one transistor, and has non-pulsating input and output currents. No other practical design uses this circuit over such a wide input:output range, the widest found being a span of just 46% on a fixed output voltage power supply, (Rensink et al, 1979). The SBCWG allows battery charging when the rectified WTG voltage is between 24V and 240V DC, and with a battery voltage that can vary between 20V (a just discharged 24V battery), and 165V DC (a fully-charged 110V battery). This gives a large voltage conversion ratio of 1:10 and 7:1 respectively, which (with a cost penalty) could be obtained with exactly the same component values.

This project is the first recorded occasion that the Cuk converter has been used in conjunction with a SBCWG. It has been used before in solar powered battery chargers, but in such an application there is an essentially constant source voltage, and there is not the problem of handling the excess power input which can occur in WTG installations in high windspeeds.

One of the practical disadvantages with the Cuk converter is its poor closed loop stability stemming from the three poles in its transfer function. This is because for a stable system it is necessary to have less than unity gain at 180° phase shift. This means
that depending on the positions of the poles, a highly damped system can result from the need to add an additional low frequency pole for stability. Hence open loop boundary control methods are explored, with a proportional magnitude boundary method finally being selected. The nominal rating of 2kW represents one of the highest power units built (although Cocconi & Cuk, 1983, describe a similar sized unit), and is believed to be the only unit using this type of boundary control.

Early SBCWGs used a fixed resistance dump load to absorb WTG power when the batteries are fully charged, but many of the disadvantages of this technique are overcome by using a PWM controlled transistor to vary the effective dump load resistance. This method also has the advantage of avoiding sudden torque changes due to the change between battery charging and the dump load. The dump load may be controlled by either a voltage servo to maintain a set battery voltage, or a current servo to maintain a fixed current.

Further work is outlined which would enable further "fine-tuning" of the power demand to compensate for other effects, such as changes in conductor resistance with temperature.

Part of this research is a PC-based Wind Generator Test Rig which simulates the effect on the generator power output and shaft speed in response to actual recorded wind patterns. An extension of this allows a computer to control a test rig which uses a motor to turn a generator, allowing a generator to work under simulated wind conditions. The software is Windows™ based, allowing fast and simple changes to be made, and hence the effects of changes in the SBCWG system to be rapidly evaluated.

This research project consists of two main phases;
- The design of a SBCWG system to improve both the power transfer and system reliability.
The design of a Wind Generator Test Rig (WGTR), which may be used for bench testing and evaluation of wind generator systems.

Chapters two and three describe in detail the design, construction and performance of the Cuk power and control circuits, chapter four the effect of using flexible blades, and chapter five the design of the wind generator test rig.
CHAPTER 2
SBCWG BATTERY CHARGER

2.1 INTRODUCTION
2.1.1 Basic Considerations

The purpose of the power controller is to match the WTG output to the battery in such a way that the battery is charged at the maximum rate possible at all windspeeds. The circuit which is designed to do this is considered in two sections - the power converter to efficiently match the varying generator output voltage to the battery voltage, and the control circuit to regulate the power to be a function of the windspeed.

2.1.2 Outline Of The Power Circuit

The battery bank specified for the 2kW SBCWG can have any nominal voltage between 24V and 120V. In practice, this means that the battery voltage can vary between 20V, (a discharged 24V battery) and 165V (an overcharged 110V battery). The DC link voltage supplied from the WTG varies between 0V and 240V as the wind speed changes, and so a wide-range step-up/step-down power converter is required. Since the link voltage is supplied by a three-phase generator, with DC-side capacitor filtering, it is assumed in the analysis to be essentially a steady DC voltage. An initial paper review suggested that the relatively little used Cuk converter would be best suited for this application, having the benefits of both single transistor control, and of low input and output current ripple.
The Cuk converter could be closed-loop controlled using either the wind generator (input) current, or battery charge (output) current as the controlled parameter. It was eventually decided to control the WTG current, which avoids having to account for the varying battery terminal voltage and converter efficiency when calculating the demand current.

2.1.3 Outline Of The Control Circuit

The control circuit is in two sections, one to match the WTG current to be a function of the windspeed, and the other to control the Cuk converter to draw this demanded current.

The control system to match the power drawn by the power converter to that available from the WTG should have characteristics based on those of figure 1.11. Above the maximum rated windspeed, it is assumed that a mechanical means of power limiting is used, (chapter 4).

The optimum power demand curve for the SBCWG WTG will not quite follow the ideal curve of figure 1.11 for several reasons, including:

- Internal generator resistance: As the generator current and temperature increases, the internal voltage drop and associated power losses will increase.
- Variation in the optimum $c_p$: The ideal curve assumes operation at a constant $c_p$, but in practice this is likely to vary with the windspeed.
- A current dependent voltage drop in the down leads.
- Possible deterioration of the PM generator performance over time.
- Possible deterioration of the turbine characteristics, for instance due to weathering or deposition of dirt.
To overcome these additional variations, it is suggested that a hill climbing algorithm is used to automatically modify this power demand by continuously seeking the maximum load. Such possibilities are discussed further in appendix 5.2.

There will be circumstances under which it is not possible for the battery to sink all the available WTG power, for instance if the battery is almost completely charged. This means that the WTG could be under-loaded, which could lead to permanent damage through exceeding the maximum design speed. To avoid this happening, a variable dump load resistance is used to ensure that the total WTG electrical load is always equal to the available WTG power, (chapter 3).

The battery charger is current controlled, with the charge current ultimately limited by the rated battery terminal voltage. It is recognised that this simple single stage voltage-limited charging does not make full use of the battery capacity, nor obtain maximum lifetime, but the irregular charge and discharge cycling of the battery makes more ideal battery charging or monitoring very difficult, (appendix 2).

It was originally intended to set the WTG demand current with a microcontroller (appendix 5), but the later decision to use flexible blades (chapter 4), led to this complexity not being needed, and so simpler discrete solutions are presented in 4.7.

Two generic types of control were considered for controlling the Cuk converter circuit; Proportional Integral Derivative (PID), and current boundary control. However, known difficulties with stabilising a PID control servo for controlling the Cuk converter favoured a current boundary control technique.

In windspeeds above the actual rated speed, the available WTG power could be greater than that which the Cuk converter can handle, and so it is assumed in this chapter that an ideal mechanical or aerodynamic means of power spilling is used, limiting the maximum
WTG power to exactly 2kW. Chapter 4 considers the effect of using the initial computer generated design of flexible blades, and the implications which it has for the design of the Cuk converter and control system.

2.2 SELECTION OF THE CHARGING CIRCUIT

2.2.1 Switched Mode Power Converters

High frequency operation of power conversion circuits means that reactive components can be much smaller than mains frequency equivalents, and a much faster transient response may be possible. The high frequency switching devices used are always run in saturation, keeping on-state conductive power losses to a minimum. However, a design trade-off is the increase in semiconductor switching losses with frequency.

The standard non-isolated or transformerless topologies (buck, boost, and buck-boost) are reviewed in appendix 3.

2.2.2 The Cuk Converter

This is a relatively new topology (figure 2.1), first reported by Slobodan Cuk in 1977, (Cuk and Middlebrook, 1977). It is a step-up, step-down converter, derived from cascading the boost and the buck converter. It is described as being optimum technology, since there are only five power components necessary to give an almost ideal DC transformer (Middlebrook, 1979), and the non-pulsating input and output currents give it an advantage over the buck-boost converter.

The steady state voltage transfer function is most easily found by considering the voltage across L1 during the on and off period. Since under steady state conditions the inductors are unable to sustain an average voltage, VL1 and VL2 will be zero, and hence;
Fig. 2.1 The Cuk converter

Fig. 2.2 The Cuk converter with transistor conducting

Fig. 2.3 The Cuk converter with diode conducting
Vi \( \frac{D}{(1-D)} \) (Where \( V_i \) = Input voltage, \( V_o \) = Output voltage, and \( D \) = fraction of the cycle for which the transistor is switched on).

Since the average inductor voltage is zero, the capacitor is charged to the sum of \( V_i \) and \( V_o \), and the output voltage is of opposite polarity to the input voltage.

When the transistor is on, the capacitor biases the diode off (figure 2.2), and currents \( I_1 \) and \( I_2 \) now flow through the transistor. When the transistor is off, the capacitor biases the diode off (figure 2.3), and so now currents \( I_1 \) and \( I_2 \) both flow through the diode. Currents \( I_1 \) and \( I_2 \) are therefore both continuous, with charge being transferred through the capacitor. This elegant circuit has clear penalties for component selection, since the capacitor sees the sum of the input and output voltage, and must be rated for a high ripple current. The semiconductors similarly have the sum of the input and output voltages applied when they are off, and the sum of the input and output currents applied when they are on. The ideal waveforms are shown in figure 2.4.

Since input and output currents are both non-pulsating, the size of any necessary additional filtering can be much reduced, and the topology means that the effect of any additional inductance at either input or output will actually be beneficial through assisting the reduction in current ripple. A further benefit is that the non-pulsating current waveforms will give rise to less electrical interference than topologies with rapidly changing currents. There are still pulsating currents present, but these are restricted to the central loop consisting of \( Q_1 \), \( C_1 \) and \( D_1 \), and are minimised by keeping the conductors short. This practice also keeps stray inductance to a minimum, reducing the peak voltage seen by the transistor and diode at switch off.

The usual implementation of the Cuk converter has one controlled transistor, and one diode, which is biased by the capacitor voltage to conduct alternately with the transistor.
Fig. 2.4 Ideal Cuk converter waveforms.

- **Ice**: Transistor Emitter Current
- **Ish**: Diode Current
- **Vce**: Transistor Collector Emitter Voltage
- **Vah**: Diode Anode-Cathode Voltage
- **I(1)**: Input Inductor Current
- **I(2)**: Output Inductor Current
- **V(1)**: Input Inductor Voltage
- **V(2)**: Output Inductor Voltage
- **Vcap**: Capacitor Voltage
- **Icap**: Capacitor Current
This is an important feature, since it means that the usual care with device overlap necessary in converters with two or more controlled devices is not necessary. The use of the capacitor for energy storage and transfer is unusual, but is claimed to be more efficient at energy transfer than the equivalent magnetic design, (Wu, C. 1982). Additionally, a better power transfer per unit mass and volume is achieved by this technique.

2.2.4 Extensions of the Cuk converter

The similarity of the waveforms across both chokes means that they can both be wound on the same core, with a reduction in the magnitude of current ripple in both chokes. This technique, known as "integrated magnetics", would benefit the production Cuk converter in this project through a reduction in the mass and cost of chokes, (Cuk, S. & Middlebrook, R. 1978, Cuk, S. 1980, Middlebrook, R. 1983 and Cuk, S. & Polivka, M. 1983).

Other variants of the Cuk converter are shown in figure A3.2, and include:

- The use of cascaded power stages (Middlebrook, R. 1988) or tapped inductors (Cuk, S. & Middlebrook, R. 1977) to allow higher input and output voltage ratios.
- A bi-directional version suitable for battery charging and discharging (Middlebrook et al 1978).
- A variant with a non-inverted (but pulsating) output, known as the SEPIC or Single Ended Primary Inductor Circuit, (Itoh, R. & Ishizaka, K., 1991).

The circuit was simulated using Spice (appendix 3), but the package used was disappointing in that it took too many cycles to arrive at the steady state condition, and so it was not possible to examine closely the normal operating waveforms. This meant that it could not be used for detailed examination of steady state waveforms, but it did confirm that the Cuk converter circuit worked as described in the literature.
2.3 POWER CIRCUIT DESIGN

2.3.1 Construction Of The Prototype SBCWG Controller

The experimental Cuk converter was designed to demonstrate the operation of the unit over the required power range when charging a nominal 24V battery.

The Cuk converter servo control circuit, gate drive and power circuit are shown in figure 2.5, with a full schematic of the 2kW SBCWG controller included in appendix 4. The heavy duty battery connection terminals can be seen in figure 2.6 on the rear (left) panel, with the fan blowing through the heatsink assembly behind the Printed Circuit Board, (PCB). The power mosfets (metal oxide semiconductor field effect transistors) and IGBTs (Insulated Gate Bipolar Transistors) are mounted on the edge of one of the heatsinks at the top left of the board, with their leads soldered direct to the PCB.

2.3.2 Component Stress Calculations

The voltage and current component stresses dictate both the choice of actual power circuit devices, and through their power loss, the overall circuit efficiency. The power loss contributions for each of the principle components are separately calculated in the spreadsheet (figure 2.7), from which the predicted efficiency is found. The peak collector-emitter voltage of the Insulated Gate Bipolar Transistors (IGBTs) is the only value which is in practice much larger than the basic formula indicates, and so careful practical trials with snubber networks were necessary to reduce this overshoot to a safe level.

The following expressions were used for calculating the component power losses:

\[
\text{Choke L1} = I_1
\]
Fig. 2.5 Circuit diagram of Cuk converter power and servo control circuit.

Fig. 2.6 Photograph of prototype 2kW SBCWG controller.
### Table 1: Current Waveforms

<table>
<thead>
<tr>
<th>Windspeed (m/s)</th>
<th>Vbus (Volts)</th>
<th>Max. Gen. Power (W)</th>
<th>Ib (Amps)</th>
<th>Il (Amps)</th>
<th>Ii (Amps)</th>
<th>K (Amps)</th>
<th>In (Amps)</th>
<th>Iw (Amps)</th>
<th>Im (Amps)</th>
<th>% Ripple Current</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
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<td>0.00</td>
<td>0.00</td>
</tr>
<tr>
<td>2</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
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<td>0.00</td>
<td>0.00</td>
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<td>0.00</td>
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<tr>
<td>4</td>
<td>0.00</td>
<td>0.00</td>
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<td>0.00</td>
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<td>0.00</td>
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<tr>
<td>6</td>
<td>0.00</td>
<td>0.00</td>
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<td>8</td>
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<td>0.00</td>
<td>0.00</td>
</tr>
<tr>
<td>10</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
</tr>
<tr>
<td>12</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
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<td>0.00</td>
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<tr>
<td>14</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
</tr>
<tr>
<td>16</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
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<tr>
<td>18</td>
<td>0.00</td>
<td>0.00</td>
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<td>20</td>
<td>0.00</td>
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<td>0.00</td>
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<td>0.00</td>
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<td>0.00</td>
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</tbody>
</table>

### Table 2: Period and Capacitor ripple voltage

<table>
<thead>
<tr>
<th>Windspeed (m/s)</th>
<th>Ton (ms)</th>
<th>Toff (ms)</th>
<th>Dck (%)</th>
<th>Period (ms)</th>
<th>Frequency (kHz)</th>
<th>Cap. Ripple (V)</th>
<th>Cap. % ripple (%)</th>
<th>Lout % ripple (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
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<tr>
<td>2</td>
<td>0.00</td>
<td>0.00</td>
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<td>0.00</td>
<td>0.00</td>
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<td>0.00</td>
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<td>0.00</td>
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<td>10</td>
<td>0.00</td>
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<td>0.00</td>
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<td>0.00</td>
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<tr>
<td>12</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
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<tr>
<td>14</td>
<td>0.00</td>
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<td>0.00</td>
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<tr>
<td>16</td>
<td>0.00</td>
<td>0.00</td>
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<td>18</td>
<td>0.00</td>
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<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
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<tr>
<td>20</td>
<td>0.00</td>
<td>0.00</td>
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<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
</tr>
</tbody>
</table>

### Table 3: Component power losses

<table>
<thead>
<tr>
<th>Windspeed (m/s)</th>
<th>Component Power Losses (Watts)</th>
<th>Actual Efficiency (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>Total</td>
<td>0.00</td>
</tr>
<tr>
<td>2</td>
<td>Total</td>
<td>0.00</td>
</tr>
<tr>
<td>4</td>
<td>Total</td>
<td>0.00</td>
</tr>
<tr>
<td>6</td>
<td>Total</td>
<td>0.00</td>
</tr>
<tr>
<td>8</td>
<td>Total</td>
<td>0.00</td>
</tr>
<tr>
<td>10</td>
<td>Total</td>
<td>0.00</td>
</tr>
<tr>
<td>12</td>
<td>Total</td>
<td>0.00</td>
</tr>
<tr>
<td>14</td>
<td>Total</td>
<td>0.00</td>
</tr>
<tr>
<td>16</td>
<td>Total</td>
<td>0.00</td>
</tr>
<tr>
<td>18</td>
<td>Total</td>
<td>0.00</td>
</tr>
<tr>
<td>20</td>
<td>Total</td>
<td>0.00</td>
</tr>
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</table>

Fig. 2.7 Cuk converter design spreadsheet
Resistive power loss = \( I_1^2 R_1 \)

**Choke L2**
- Current rating = \( I_2 \)
- Resistive power loss = \( I_2^2 R_2 \)

**Capacitor**
- Voltage = \( V_1 + V_0 \)
- Peak current = The peak current will be \( I_1 \) or \( I_2 \), depending on which is larger.
- Power Loss = \([I_2^2 D + I_1^2 (1-D)] x ESR\) 
  (Note that for clarity the Equivalent Series Resistance (ESR) of the capacitor is omitted from figures 2.1 - 2.3.)

**Diode**
- Reverse voltage = \( V_1 + V_0 \)
- Peak current = \( I_1 + I_2 \)
- Diode conduction losses = \((1-D) x (I_1 + I_2) x V_{AK}\)

**IGBT**
- Peak transistor voltage = \( V_{os}(V_1 + V_2)/100^9 \)
- Peak current = \( I_1 + I_2 \)
- Conduction loss** = \( D(I_1 + I_2)V_{CE}\)
- Switching loss = \( V_{pl}(I_1 + I_2) t_{sw}f / 6\)

*Where \( V_{os} \) is the magnitude of the overshoot voltage on switch-off as a percentage of the steady-state voltage.

**For comparison, conduction losses in a mosfet would be \( D(I_1+I_2)^2 r_{ds} \), where \( r_{ds} \) is the drain source resistance.

**IGBT Snubber**
- Power loss = \( C [(V_1 + V_2)0.01 V_{os}]^2 f \)
2.3.3 Predicted Circuit Performance

Figures 2.8.1 to 2.8.4 show the predicted Cuk converter efficiency, frequency, input current ripple, and capacitor voltage ripple. These are plotted as a function of DC link voltage, and assume maximum Cuk converter power demand for that windspeed. Two components dominate the power losses, and hence efficiency (figure 2.8.1) - the diode forward voltage drop and the output choke resistance. Little can be easily done to reduce the diode loss, but at greater cost a lower resistance choke could be specified.

The frequency range with the chosen control circuit (discussed in 2.5) is a compromise between high frequency operation at low voltages and low frequency operation at high voltages - a higher frequency at low link volts means a proportionately greater switching loss, but the low total power loss at low windspeeds means that the semiconductors are well able to withstand this. Figure 2.8.2 shows how the operating frequency falls with voltage, with the IGBT switching delay time helping to limit the frequency at lower voltages. This is also reflected in figure 2.8.3, which shows how the input ripple current increases with lower link volts. The capacitor ripple voltage (figure 2.8.4) increases with falling frequency.

2.3.4 Semiconductor Specification

At the high frequency at which the Cuk converter in this project is designed to operate, standard bipolar transistors are unsuitable, and so the choice is between mosfets and IGBTs. The spreadsheet shows that a switching junction voltage rating of 600V is just sufficient; since at this voltage, (for a given power rating) IGBTs have a slightly higher cost, but a lower on-state power dissipation. For the best cost:performance compromise, two IRGBC30U IGBTs were selected to operate in parallel, (with a
Fig. 2.8.1
Predicted Cuk converter efficiency against DC link voltage

Fig. 2.8.2
Predicted Cuk converter frequency against DC link voltage

Fig. 2.8.3
Predicted Cuk converter capacitor ripple voltage against DC link voltage

Fig. 2.8.4
Predicted Cuk converter input ripple current against DC link voltage
specification of $V_{ce} = 600V$, $I_{rms} = 23A$, $V_{ce \text{ sat.}} = 3.0V$, and $I_{peak} = 92A$.) Like mosfets, these devices have a positive temperature coefficient, which means that parallel devices automatically share current. A practical problem with these devices is that the TO-220 package is really too small for the voltage and current rating of the IGBT, and so instead the tab is used as the collector connection, via the heatsink. To ensure the transistors remain within their operating temperature range, a heatsink of $1.0^\circ C/W$ (naturally cooled rating) was necessary.

The choice of power diode is much more restricted, since only fast silicon diodes can operate at this frequency, voltage, and current. Two RURU80100 devices were used, which because they use a live tag which must not be connected to the live IGBT tab, had to be mounted on a separate heatsink. The two heatsinks are carefully arranged to allow for fan cooling, (figure A4.2).

2.3.5 Snubber Design

It was found that to reduce the voltage spike seen across the IGBT collector emitter junction at switch-off, it was necessary to place a snubber across it. During tests, it was found that a $22nF$ capacitor and a $10\Omega$ resistor were able to reduce the voltage overshoot to less than 25%, with a calculated maximum power loss of just $3.5W$. (The general selection of snubber components is considered in detail when considering the design of the dump load control circuit, section 3.5.1.)

2.3.6 Selection Of Capacitor And Inductors

For minimum resistive or “copper” loss, the gauge of copper wire used should be as large as practical.
The input choke inductance not only affects the input current ripple, but also the frequency of operation of the circuit. By iteration, it was chosen to be 10mH at 40mΩ. The output choke affects only the battery charge current ripple, and so must be large enough to reduce the output current ripple to an acceptable level. 10mH was also chosen as a reasonable cost-ripple compromise for this choke.

The capacitor must be able to work at both a high frequency and high voltage, and since the capacitor couples all the charge from the output to the input, it experiences a high ripple current. A specialist device is therefore needed, and again there is a trade-off between cost and voltage ripple. A 25μF 600V MKV type with a rated ripple current of 32A was eventually chosen. Further increasing the capacitance value reduces the capacitor voltage ripple - but also increases the cost.

2.3.7 Gate Drive Circuit

Since the Cuk power converter has an inverting function, the DC link voltage negative is connected to the battery positive, and so an isolated drive for the IGBTs is required. The need for a wide variation in duty cycles means that an opto-isolator was chosen for this application.

The transistor drive on the output of the steering latch is amplified by the high speed 2SC2500 transistor, which originally drove an opto-isolator. This isolated signal was then amplified by a discrete mosfet driver, (using the same circuit as the dump load mosfet drive in section 3.5.4). However, the switch-off delay (figure 2.9) was over 8μs, which under worse case conditions would cause the Cuk converter input current to rise over 50% above the set upper boundary level.

Using the new TLP250 dedicated integrated isolator and drive device, the switch-off delay was greatly reduced, and figure 2.10 shows how the maximum current overshoot
Fig. 2.9 Discrete gate driver switching delay (Opto-isolator input voltage and IGBT Icc)

Fig. 2.10 Integrated gate driver switching delay (TLP250 input voltage and IGBT Icc). (Icc = collector current)

Fig. 2.11 Comparator switch-off delay time. (Input to LM311 comparator and IGBT Icc)

Fig. 2.12 Comparator switch-on delay time. (Input to LM311 comparator and IGBT Icc)

Fig. 2.13 IGBT gate waveforms. (Vge = gate emitter voltage, Vce = collector emitter voltage)

Fig. 2.14 IGBT switch on waveforms.
was also reduced. The total transition time for the TLP250 is now just 2μs. (For comparison, figures 2.11 and 2.12 show the 1.4μs switch-off and 1.3μs switch-on delay of the drive amplifier stage alone with the old discrete opto-isolator and 1M311 comparator based circuit). Figure 2.13 shows the 170mA peak gate current at IGBT switch on, and 2.14 how the collector current has stopped within 220ns of the gate current pulse starting.

2.4 TESTING OF THE POWER CIRCUIT

2.4.1 Test Equipment

There was no 2kW 240V DC power supply available, and so an autotransformer based supply was assembled especially for the testing, (figure 2.15). This consisted of a 3-phase isolating and step-down transformer, connected to a 3-phase autotransformer, which was used to mimic the varying WTG output voltage. This gave a peak rectified open circuit voltage of c.300V DC. Since the ripple on a rectified 3-phase waveform is small, it does not matter that the variation with voltage is not the same as on a real WTG. (In practice, the DC link was found to have a higher voltage ripple than expected (figure 2.33), attributable to rectifier overlap caused by the large combined inductance of the two transformers used.) During tests, analogue ammeters were used for measuring both the link and battery voltages, a moving coil meter for the input current, and a Hall-effect meter for the battery current.

The battery bank comprises ten Alcad 220AH 2x1.2V nickel cadmium cells connected in series to give a nominal 24.0V stack (figure 2.16), with an 80A fuse link included for short circuit protection.
Fig. 2.15 Photograph of power transformers used for testing.

Fig. 2.16 Photograph of 24.0V 220AH Nickel-cadmium battery bank used for tests.
2.4.2 Circuit Waveforms

Figures 2.17 to 2.22 show the main Cuk converter waveforms when there is a 24V battery being charged from a link voltage of 100V, with 3A being drawn from the link. Figures 2.17 and 2.18 show how the WTG and battery currents have a very similar non-pulsating form, and figures 2.19 to 2.21 the switching action of the IGBTs and diodes. It can be seen that the waveform in figures 2.20 was captured at a much lower point on the link voltage than 2.19, and that the servo has compensated by increasing the pulse width. Figure 2.22 shows the capacitor being alternately charged and discharged, and figures 2.23 and 2.24 the transistor switching action, showing clearly that the switch on power loss is negligible compared to the switch off power loss. The IGBT collector emitter saturation voltage was not measured as exceeding 2.8V throughout the tests, which is in agreement with the datasheet value of Vce(sat) for Vge = 15V.

2.4.3 Results And Analysis Of Power Circuit Tests

The efficiency is lower than predicted, although it is clear from fig 2.25 that the efficiency does increase with current demand. There are several possible sources of additional power loss that were not included in the original predictions, which help account for the discrepancies:

- Copper loss of internal converter wires. This will be extremely small, since all wires were deliberately kept short and of broad gauge.

- Additional choke copper loss due to heating. The output choke did get warm during testing, but it is unlikely that it ever rose above 55°C, or a temperature rise of 30°C. The temperature coefficient of resistance of copper is +0.0043 %/K, and so the increase in resistance resulting from this increase in temperature is just 13%.
Figs 2.17 to 2.22 show important Cuk converter waveforms under the same operating conditions (Vin = 100V, Vbat = 24V, Iin = 3A).

Fig. 2.17 DC link voltage and Cuk converter input current.

Fig. 2.18 Battery voltage and charge current.

Fig. 2.19 IGBT collector emitter voltage and collector current.

Fig. 2.20 Diode anode-cathode voltage and IGBT collector current.

Fig. 2.21 IGBT Collector emitter voltage and diode current.

Fig. 2.22 IGBT Collector emitter voltage and capacitor current.
Fig. 2.23 IGBT switch-on loss. Vce and Icc (Vin = 200V & Vb = 25V, Icc = 2A)

Fig. 2.24 IGBT switch-off losses. Vce and Icc (Vin = 115V & Vb = 25V, Icc = 3.2A)

Fig. 2.25 Measured Cuk converter efficiency as a function of DC link voltage for different values of input current, (where power is proportional to the DC link voltage cubed).

Fig. 2.26 Boundary control operation. Current boundary levels as a function of DC link volts, (where power is proportional to the DC link voltage cubed)
Second order ripple effects. Component conductive losses will increase, since the poorer form factor due to the superimposed current ripple waveforms causes greater power loss across the near constant voltage diode and IGBT junctions.

Choke magnetising power. Since the choke will be magnetised in one direction only, the small flux excursion due to the current ripple will be small. The energy necessary for this is difficult to estimate without full core data, but is usually ignored in this type of application.

Power supply phase shift. The phase lag between the current and voltage ripple waveform means that there is a net power flow due to this ripple back to the supply. Referring to figure 2.32, this phase lag is measured as 130°, and the voltage and current ripple as 25V and 1A respectively. The net power flow back to the WTG of this ripple component is calculated as \( VI\cos\phi = 15 \text{ W} \).

Accuracy of the measurement equipment. The direct reading meters have a basic accuracy of ±1%, the current probe is similarly accurate to ±1%, but the oscilloscope can only be read to within c.±2%. However, these percentages are calculated for readings taken at full scale deflection, and so for the readings taken at levels near the bottom of the meter scale, the error will be considerably larger. Hence, particularly at lower power levels, the uncertainty of the results is greater than the simple sum of the quoted errors.

The low levels of power drawn from the WTG at low windspeeds means that although the efficiency may be low at low windspeeds, the actual power loss is also low. Any fixed losses will therefore have a disproportionate effect at such low windspeeds.
In practice, the varying DC link voltage made the operating frequency dither, and so it was not possible to obtain accurate frequency characteristics. However, it was observed that the frequency clearly decreased with increasing demand.

The measured performance of the Cuk converter is clearly not as good as predicted, but many reasons for this, particularly possible difficulties with instrumentation, are suggested to account for this. Further work should include an investigation to establish more closely where the additional losses, if indeed they are real, are actually occurring.

2.5 CONTROL CIRCUIT DESIGN

2.5.1 Selection Of Control Technique

The purpose of the servo control circuit is to maintain the current drawn from the WTG by the Cuk converter circuit to be close to the value set by the current demand. There are a large number of integrated circuits available for implementing Pulse Width Modulation (PWM) based Proportional, Integral, Derivative (PID) control, but as already mentioned, it would be difficult to stabilise the Cuk converter using this technique over the wide range of operating conditions. It was therefore decided to consider the current boundary control family of circuits, since these can work without any direct feedback. Boundary control works by allowing the controlled current to vary between set upper and lower current boundary levels, and so is sometimes referred to as hysteresis control. Usually the converter switching device is switched on when the current falls to the lower set boundary level, and off again when it hits the upper boundary level, as shown in figure 2.26. Various types were considered;

**Constant tolerance band**

The upper and lower current boundaries are a fixed current apart. The main disadvantage is that the percentage ripple gets high at low current levels, but it also stops working if the minimum current becomes negative.
Constant-off type
This overcomes the second problem above by always switching the device off for a fixed time. Both these techniques suffer from variable operating frequency.

Constant frequency method
This switches the device on at fixed intervals, switching it off when it reaches the top boundary. The disadvantage of this is the variable ripple magnitude.

Proportional tolerance band method
Since the demand is always positive, the lower current boundary will always be positive, overcoming the problem with the constant tolerance band type. The frequency with this method is still variable, being highest at low demand levels and high input voltages.

All methods suffer from either variable frequency or variable current ripple on the input and output of the converter. However, the spreadsheet simulation shows that by using the proportional boundary tolerance band method, the predicted operating frequency and current ripple under expected working conditions is acceptable. This is therefore the method that was selected.

2.5.2 Analysis Of The Proportional Boundary Control Technique

The following analysis gives greater detail of the Cuk converter input current waveform when under boundary control. This analysis is based on applying Lenz's law across the two inductors for both the on and off times. (The effective value of $L1$ will be slightly increased by the inductance of the generator, and so is included in the value of $L1$). Also included is the time delay between the input current reaching one of the boundaries and the current actually altering its direction of change, which limits the operating frequency and the servo current tracking performance. As the input voltage increases, the rate of
rise of current also increases, and so the maximum current excursion beyond the upper boundary also increases.

Figure 2.26 shows how the top (It) and bottom (II) set currents are set to be a fixed percentage above and below the demanded current level. Since the additional positive and negative current excursions due to switching delays will only rarely be same, the actual mean current level (Im) will move slightly from the demanded level, (Idem).

Using the nomenclature of figure 2.27.1, the following equations are used in the spreadsheet model in figure 2.7.1. Note that the equations for the actual current levels includes the extra term showing the effect of the on and off switching delays (dyn and df respectively) on both the current undershoot and overshoot.

- Set top current boundary;
  \[ \text{It} = \text{Idem} + \frac{\text{IP}. \text{Idem}}{2.100} \]
  where IP = demanded current window as a percentage of the demand current, Idem.

- Set bottom current boundary;
  \[ \text{II} = \text{Idem} - \frac{\text{IP}. \text{Idem}}{2.100} \]

- Maximum actual current boundary,
  \[ \text{Ix} = \text{Idem} + \frac{\text{IP}. \text{Idem} + \text{Vo}. \text{df}}{2.100} \]

- Minimum actual current boundary,
  \[ \text{In} = \text{Idem} - \frac{\text{IP}. \text{Idem} - \text{Vo}. \text{df}}{2.100} \]
Fig. 2.27.1 Nomenclature for the analysis of boundary control operation

- dyf = switch off delay time
- dyn = switch on delay time

- I = Set top current
- II = Set lower current
- IX = Actual maximum current
- In = Actual minimum current

Fig. 2.27.2 Control pulses for the boundary level control circuit

Fig. 2.27.3 Proportional boundary control circuit - current waveforms

Fig. 2.27.4 Fixed boundary control circuit - current waveforms
Actual current window

\[ I_w = I_x - I_n = \frac{I_{P \cdot Idem} + V_{d} \cdot dV_f + V_{o} \cdot dV_n}{100L_1} \]

Actual mean current

\[ I_m = \frac{I_n + I_x}{2} \]

From this analysis, the following ripple values can be calculated:

- Actual input (choke) current ripple (%): \[ = 100 \times \frac{I_w}{I_m} \]

- Actual capacitor voltage ripple (%): \[ = \frac{I_{L2} \cdot I_m}{C(V_i + V_o)} \]

The total on-time (t_on) and off-time (t_off) is given by:

- \[ t_{on} = \frac{L_1 \cdot (I_x - I_n)}{V_i} \]
- \[ t_{off} = \frac{L_1 \cdot (I_x - I_n)}{V_o} \]

### 2.6 Implementation Of The Proportional Boundary Control Circuit

Figure 2.27.1 shows how the current demand is used as a basis for setting upper and lower current boundaries. A comparator generates a short pulse when the measured Cuk converter current falls below the lower boundary (figure 2.27.2), setting the output steering latch to the Cuk converter IGBT "on" state. When the current exceeds the top boundary, another comparator generates a short pulse that resets the steering latch, switching the IGBTs off, and the cycle then repeats itself. Figure 2.27.3 shows with a constant input and output voltage, where the rate of change of current is constant, the
frequency increases as the current demand falls. Figure 2.27.4 is included to contrast the chosen proportional current boundary method with the rejected fixed boundary method.

The hardware (appendix 4.8) was implemented using operational amplifiers to set the proportional boundary levels, fast comparators to detect the cross-over points, and a directional latch to control the transistor state. The latch was a discrete level-triggered device, avoiding the possibility of loss of control that an edge-triggered type could give under noisy conditions.

Should the battery be removed, the Cuk converter will still attempt to maintain the same output current in to this open circuit, since it is controlled by a closed loop current servo. If the Cuk converter is not shutdown when this happens, the output voltage will be increased in an effort to maintain the current, and could become so high that the Cuk converter becomes permanently damaged. A circuit described in appendix 4.4 detects this condition, and safely shuts down the Cuk converter until the battery is reconnected.

2.7 TESTING OF THE CONTROL CIRCUIT

2.7.1 Basic Servo Operation

Figures 2.28 to 2.32 show very simply how the servo alters the mark:space ratio to charge the battery at different rates as the input current demand, input voltage and battery voltage are varied. Figure 2.33 shows the voltage and current ripple on the DC link, which is greater than expected, due mainly to the high inductance of the two series power transformers used.

Both the current tracking and transient response of the servo were tested:
Fig. 2.28 Control circuit testing: Vce and Ib (Vin = 60v, Vb = 12v, Ib = 5A).

Fig. 2.29 Control circuit testing: Vce and Ib (Vin = 60v, Vb = 24v, Ib = 5A).

Fig. 2.30 Control circuit testing: Vce and Ib (Vin = 30v, Vb = 24v, Ib = 5A).

Fig. 2.31 Control circuit testing: Vce and Ib (Vin = 120v, Vb = 12v, Ib = 2A).

Fig. 2.32 Control circuit testing: Vce and Icc (Vin = 240v, Vb = 24v, Iin = 6A)

Fig. 2.33 Tolerance band switching DC link current and IGBT Vce.
(Sinusoidal variation is mains ripple).
2.7.2 Current Tracking

For this test, the current demand was set on a potentiometer (figure 2.34) to a series of nominal current demand levels, and the DC link voltage then varied over the designed range for each of these current levels, (Figure 2.35). The rise in mean current as the voltage increases is attributable partly to the greater voltage across the input choke leading to a proportionate increase in the overshoot of the upper current boundary. Either increasing the input inductance, or reducing the time delays of the IGBT switching circuit, would improve this tracking performance.

2.7.3 Transient Performance

For this test, a storage oscilloscope monitored both the demand signal and actual input current when the switch is closed. When closed, the switch decreases the current demand by about 45%, and when opened, increases the demand by a similar amount. The response time (figure 2.36) of the actual current to undergo 2/3rds of the demanded change was always less than 300μs, which is taken as the characteristic response time, and was the same for both positive and negative transitions. There was no noticeable overshoot or instability, implying that the circuit is adequately damped.

It would also have been useful to initiate a step change in link voltage, but this was not practical, since there was no power supply available that could be controlled to give such a rapid change in output.

2.7.4 Operation With Different Battery Voltages

When charging a 24V battery, the controller never needs to work in the boost region, since the WTG will not be able to generate any power at windspeeds (and hence
Fig. 2.34 Servo test circuit.

Fig. 2.35 The current tracking of the Cuk converter, (marked currents are nominal settings at 25V).

Fig. 2.36 Negative transient servo response (Demand voltage and battery current)

Fig. 2.37 The cuk converter in boost mode, (charging a nominal 24V battery).
The main effect of raising the battery voltage is to proportionately reduce the battery charge current, since the WTG current will remain essentially the same.

Since in the testing of the Cuk converter using a 24V battery it was never used in the boost mode, a separate test was performed to verify operation in this mode. A 0-30V bench power supply was set at a constant 6.0A, and the output of the Cuk converter was connected to a 24V battery. The value of the Cuk converter "on" time (D), was monitored as the PSU voltage was varied from 0 to 30V, with the point of D = 0.5 occurring when the PSU voltage was just above the actual battery voltage, (figure 2.37). In this test, the analogue meters on the PSU were used for monitoring the input current, an analogue avometer for the battery current, and a digital multimeter for the battery voltage. This simple test confirms that the Cuk converter does function in the boost mode.

2.8 SUMMARY

This chapter has reviewed the practical work to demonstrate the feasibility of using the Cuk converter circuit under proportional boundary level control to charge a battery bank from a variable voltage WTG. It has confirmed that the circuit has features which make it attractive for this type of application, in particular the non-pulsating input and output currents, and the operation of the boundary control circuit. However, it has also highlighted further work which could be pursued to improve the performance of the circuit.

Attempting to design a switched mode circuit to work over such a wide operating range inevitably means that it is hard to maintain optimum circuit performance over the full range. Analogous to the two-step autotransformer circuit of the 500W WTG control circuit (figure 1.9), an additional means of changing the voltage transformation ratio could be used. Appendix 3 describes several methods for this, including the use of a
transformer, tapping one of the Cuk converter inductors, or using a cascaded circuit. The voltage transformation ratio could then be crudely set in perhaps two or more stages, and then the normal controller operation used over a smaller range for finer adjustment.

The choice of battery charger was a compromise between controlled frequency and variable speed, or controlled ripple and variable frequency. Once again, the wide operating range of the Cuk converter in this application led to a significant variation in operating frequency. Whilst the use of a high speed IGBT drive stage actually gave the control circuit a quite acceptable tracking performance, the adverse effect of the wide operating frequency on the efficiency of the Cuk converter itself could be reduced by modifying the boundary control circuit. Further work is also needed to consider the benefits of modifying the boundary control circuit in response to variations in either the battery voltage or the power demand: voltage relationship (due to some of the power being delivered to the dump load).

One of the most significant costs in the Cuk converter is that of the specialist coupling capacitor. Further increasing the capacitance value of this would reduce the ripple voltage, which would allow the Cuk converter to operate more efficiently over a wider operating range, but there would be a cost penalty for this.

Two factors dominate the power losses shown in the spreadsheet - the output diode and output choke losses. Little can be done to minimise the diode voltage drop, but at greater cost a lower resistance choke could be specified. However, the nature of the variation in flux in both of the inductors (and transformer, if used), means that the idea of integrated magnetics may be used. This allows the winding of both the input and output choke on the same core, meaning that the value of the output choke could then be significantly increased by the interaction of the flux waveform of the input choke winding. This would immediately mean that fewer turns would be necessary for the same inductance, and so the winding window could be utilised by using a greater wire gauge, both factors leading to a reduction in copper loss.
The Cuk power converter and battery charging circuit were shown by experiment to be capable of being used in this wide ranging voltage transformation ratio circuit, with input power of up to 1.5kW. However, it is recommended that further work as suggested in the summary is undertaken to further refine both the power converter and control circuits. The next chapter examines how the use of a dump load circuit can be used to safely load the WTG if the battery should be fully charged, or if a fault should prevent the Cuk converter loading the WTG at all.
CHAPTER 3
DUMP LOAD CIRCUIT

3.1 PURPOSE OF THE DUMP LOAD CIRCUIT

When the battery is unable to accept all the charge that the Cuk converter can source for a given windspeed, for instance if the battery is already fully charged, the WTG will be too lightly loaded for operation at maximum efficiency. In this condition, the WTG speed will be faster than that given by the ideal torque:windspeed relationship, and may even approach the "survival speed" of the WTG or maximum operating voltage of the controller.

By using an electrical dump load which can be varied to sink the difference between the available wind turbine power and the maximum power that the battery can accept, the WTG will always be optimally loaded. Not only does this help to avoid mechanical or electrical failure from overspeed, it also gives a useful source of resistive heating - and can even be used as a crude brake.

There will be a windspeed at which the available WTG blade power exceeds that which the dump load can dissipate, (fig 3.1) and so above this point, a mechanical means of power shedding must be relied upon, (section 4.1 - 4.2). An ideal power limiting mechanism which makes the power fall to zero above the maximum rated windspeed (as figure 1.11) is assumed in this section.
Fig. 3.1 Dump load power: shaft speed characteristics, (assuming generator voltage proportional to shaft speed).

Fig. 3.2 AC-side 3-phase thyristor controlled dump load.

Fig. 3.3 DC-side chopper controlled dump load
### 3.2 SELECTION OF DUMP LOAD CIRCUIT TOPOLOGY

The integrity of the dump load must be maintained under all conditions, since if it failed to open circuit, there could be a situation in which there is no load at all on the WTG. In order to reduce the risk of dump load disconnection through the failure of electronic components, the dump load should be connected to the WTG output with as few components as possible in the circuit between. Three resistors connected directly between WTG phases would therefore be ideal (figure 3.2), but this arrangement would mean that six thyristors would be necessary for power control.

A compromise between cost and security of resistor connection is a single resistor across the DC bus, with just the rectifier between the resistor and WTG downlead terminals (figure 3.3). Two types of transistor controlled operation are possible for varying the value of the load; linear or switched-mode. The main advantages of linear operation, where the transistor is controlled to drop a proportion of the link voltage, is that there are no fast changes of current or voltage that could lead to voltage overshoots across the transistor junction, and there is much less generation of electro-magnetic interference. However, in this application the switched-mode technique is selected because it is more efficient, since when conducting the transistor is always in saturation, and hence dissipates less power.

The efficiency of the dump load control circuit is important in applications where the power dissipated by the dump load resistor is used for space or water heating. In this circuit, the efficiency of the circuit is defined as the ratio of the power dissipation in the resistor to that of the whole dump load circuit (i.e. resistor plus control circuit losses).

The principle losses in the dump load control circuit comprise:

- Mosfet switching losses,
- Mosfet conduction loss,
Flywheel diode conduction loss,
Snubber loss.

(These losses exclude the negligible power consumed by the control circuit and the mosfet gate power consumption. In respect of the power losses, the fail-safe relay, with a coil rating of 6VA, is regarded as being part of the overall system protection rather than part of the dump load circuit, and so is not included in these calculations)

The spreadsheet shown in figure 3.4 was used as the basis for the power circuit design. Part 1 shows the bus voltage, generator power, dump power, dump current and conduction period of the dump load, and Part 2 uses this information to calculate the power losses of the mosfets, snubber resistor, and flyback diode. Figure 3.5 shows the calculated dump circuit efficiency as a function of link voltage.

3.3 BASIC DUMP LOAD CIRCUIT DESIGN

3.3.1 Specification Of Dump Load Resistor

The power rating of the resistor should be equal to the power output of the WTG when it is optimally loaded, with the wind blowing steadily at the maximum operating windspeed, (2,000W at 20m/s).

The resistance of 28.8Ω is found by considering the maximum voltage and power, (240V and 2,000W).

A low cost and mechanically robust dump load of this power and approximate resistance is available in a standard 240V "two-bar" electric fire, as shown in figure 3.6.
PART 1: DUMP POWER CALCULATIONS

<table>
<thead>
<tr>
<th>Windspeed (m/s)</th>
<th>Voltage Vbus (V)</th>
<th>Max Gen Power (W)</th>
<th>Max Dump Power (W)</th>
<th>% On-time</th>
<th>I Peak (Amps)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>0.00</td>
<td>0.0</td>
</tr>
<tr>
<td>2</td>
<td>24</td>
<td>2</td>
<td>2</td>
<td>0.10</td>
<td>0.9</td>
</tr>
<tr>
<td>4</td>
<td>48</td>
<td>16</td>
<td>82</td>
<td>0.19</td>
<td>1.7</td>
</tr>
<tr>
<td>6</td>
<td>72</td>
<td>54</td>
<td>185</td>
<td>0.29</td>
<td>2.6</td>
</tr>
<tr>
<td>8</td>
<td>96</td>
<td>128</td>
<td>329</td>
<td>0.39</td>
<td>3.4</td>
</tr>
<tr>
<td>10</td>
<td>120</td>
<td>250</td>
<td>514</td>
<td>0.49</td>
<td>4.3</td>
</tr>
<tr>
<td>12</td>
<td>144</td>
<td>432</td>
<td>741</td>
<td>0.58</td>
<td>5.1</td>
</tr>
<tr>
<td>14</td>
<td>168</td>
<td>668</td>
<td>1,006</td>
<td>0.68</td>
<td>6.0</td>
</tr>
<tr>
<td>16</td>
<td>192</td>
<td>1,024</td>
<td>1,317</td>
<td>0.78</td>
<td>6.9</td>
</tr>
<tr>
<td>18</td>
<td>216</td>
<td>1,458</td>
<td>1,658</td>
<td>0.88</td>
<td>7.7</td>
</tr>
<tr>
<td>20</td>
<td>240</td>
<td>2,000</td>
<td>2,067</td>
<td>0.97</td>
<td>8.6</td>
</tr>
<tr>
<td>22</td>
<td>264</td>
<td>2,882</td>
<td>2,489</td>
<td>1.0</td>
<td>8.8</td>
</tr>
<tr>
<td>24</td>
<td>288</td>
<td>3,456</td>
<td>2,992</td>
<td>1.0</td>
<td>9.4</td>
</tr>
<tr>
<td>26</td>
<td>312</td>
<td>4,384</td>
<td>3,477</td>
<td>1.0</td>
<td>9.8</td>
</tr>
</tbody>
</table>

Max Cont Dump Power: 0.00
Max Bus Volts at Pdmx: 0.00
Voltage/Windspeed Relationship: 0.00
Dump Load Resistance: 0.00

PART 2: COMPONENT POWER LOSS ANALYSIS

<table>
<thead>
<tr>
<th>Voltage Vbus (V)</th>
<th>R(on) (Ohms)</th>
<th>PQ(on) (Watts)</th>
<th>PQ(Sw) (Watts)</th>
<th>Psnubber (Watts)</th>
<th>Pdiode (Watts)</th>
<th>Ptotal (Watts)</th>
<th>Efficiency (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>88.1</td>
</tr>
<tr>
<td>24</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>0.00</td>
<td>94.5</td>
</tr>
<tr>
<td>48</td>
<td>0.00</td>
<td>0.10</td>
<td>0.01</td>
<td>0.01</td>
<td>0.01</td>
<td>0.01</td>
<td>94.5</td>
</tr>
<tr>
<td>72</td>
<td>0.00</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>94.5</td>
</tr>
<tr>
<td>96</td>
<td>0.00</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>94.5</td>
</tr>
<tr>
<td>120</td>
<td>0.00</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>94.5</td>
</tr>
<tr>
<td>144</td>
<td>0.00</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>94.5</td>
</tr>
<tr>
<td>168</td>
<td>0.00</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>94.5</td>
</tr>
<tr>
<td>192</td>
<td>0.00</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>94.5</td>
</tr>
<tr>
<td>216</td>
<td>0.00</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>94.5</td>
</tr>
<tr>
<td>240</td>
<td>0.00</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>94.5</td>
</tr>
<tr>
<td>264</td>
<td>0.00</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>94.5</td>
</tr>
<tr>
<td>288</td>
<td>0.00</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>94.5</td>
</tr>
<tr>
<td>312</td>
<td>0.00</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>0.10</td>
<td>94.5</td>
</tr>
</tbody>
</table>

Fig 3.4 Dump load circuit power loss spreadsheet

Fig 3.5 Dump load circuit performance
Fig. 3.6 Photograph of Fire-bar dump load.

![Diagram of Fire-bar dump load](image)

**Fig. 3.7** Spice simulation circuit.

![Spice simulation circuit diagram](image)

**Fig 3.8** Predicted mosfet and dump load voltage.

![Predicted mosfet and dump load voltage graph](image)
Above the rated windspeed, a mechanical means of power limitation is assumed. Below this windspeed, if the dump load resistor was permanently connected, then the WTG will enter the stall region.

In practice, it would be beneficial to specify the resistor to have a resistance lower than that calculated, giving a safety margin to allow for the mechanical WTG power limiting mechanism not cutting-in at precisely the maximum rated power point.

### 3.3.2 Selection Of Switching Device

A switching frequency of 25kHz was chosen, principally because it is just outside the human audible range - which is an important consideration in installations where the resistance may be used for domestic water or space heating. There is no benefit in raising the frequency further, since this increases switching losses. Consideration was given to running at a sub-audible frequency, allowing the use of slower switching devices. However, to avoid the DC link voltage ripple becoming too large, it would have meant an increase in the value of the link capacitor.

At the specified frequency, voltage, and current level, mosfets were considered to be the best cost-performance solution. The IRF 740 was eventually chosen, with a maximum drain source voltage rating of 400V, and an on-state resistance ($r_{ds}$) of 0.55 $\Omega$ at 25°C. Although having a continuous $I_{ds}$ rating of 10A, this is not achievable without the use of a very large heatsink, and so it was calculated to be both more power efficient and cheaper to use paralleled mosfets mounted on a smaller heatsink. The positive temperature coefficient of mosfet $r_{ds}$ makes current sharing simple, without the need for the emitter resistors necessary when paralleling bipolar transistors.

Hence the total conductive loss for $n$ devices, where $I$ is the total load current and $D$ the duty ratio of the waveform, is given by:
Power loss = \frac{Df^2L_{sn}}{n} (W)

The mosfet $r_{ds}$ is strongly effected by the device temperature, which is proportional to the device power generation, and hence $I^2$. In the spreadsheet calculation, $r_{ds}$ has been assumed to be twice the headline (25°C value), which reflects operation at more typical temperatures. Since the dump load mosfets dissipate less heat than the Cuk converter IGBTs for the same controlled WTG power, and the peak mosfet and IGBT power dissipation never occurs at the same time, the dump load mosfets can be mounted on the main IGBT heatsink without any heatsink rating penalty.

3.4 Spice Circuit Modelling

The dump load circuit was first modelled in Spice before building any hardware, so as to obtain an indication that the design was feasible. The circuit design for this is shown in figure 3.7, and sample outputs in figures 3.8 to 3.11. Figures 3.8 and 3.9 show the current and voltage waveforms associated with the resistive load, diode, and mosfet drain-source junction, and figures 3.10 and 3.11 show the current, voltage, and derived power dissipation of the mosfet Resistor-Capacitor (RC) snubber. (These plots are actually based on a snubber capacitor of 47nF, chosen to exaggerate the effect of the snubber). These waveforms are of similar form to those actually observed, but further refinement of the Spice model would be necessary to obtain a closer match.

3.5 DETAILED CIRCUIT DESIGN

3.5.1 Mosfet Snubbing

The current flowing in the stray inductance shown in the Spice simulation circuit causes a voltage surge when the mosfet is switched off, which is minimised by keeping the length of wire which causes this stray inductance to a minimum. The actual observed voltage appearing across the mosfet drain source junction shows a 55% voltage
Fig. 3.9 Predicted mosfet, dump load and diode currents.

Fig 3.10 Predicted snubber waveform.

Fig 3.11 Predicted snubber power dissipation.
overshoot (figure 3.12), which is too close to the device maximum drain-source voltage ($BV_{ds}$) rating at the maximum link voltage of 240V DC. However, the model gives a much more damped $V_{ds}$ voltage overshoot than this, attributable to some of the additional stray inductances being omitted from the Spice model.

To reduce this $Vds$ overshoot, a Resistor Capacitor (RC) snubber was placed across the drain-source junction, with the capacitor being selected to reduce the voltage overshoot. Figure 3.13 shows how the addition of this snubber reduces the overshoot to about 25%.

The RC snubber itself represents a power loss, since all the energy stored in the capacitor at each switch-off event (values for $Idd$, $Vds$ and $tsw$ are defined in figure 3.14), is discharged through the snubber resistor and parallel mosfet $r_{ds}$. This means that the capacitor should be chosen to be the minimum value which would adequately reduce the voltage overshoot. The $22\Omega$ resistance of the snubber resistor dominates the discharge path resistance, and so it can be considered that all the snubber energy is dissipated in this resistance. The value of this resistor was chosen such that it was small enough to allow high peak currents for fast capacitor charge and discharge times, but not so low that the time constant becomes so small that unwanted resonances might start.

The snubber charge ($Q$) is calculated from the $I.t$ area of the snubber current waveform shown in figure 3.15.

\[
Q = N\square \cdot I/D_t \cdot V_d/D
\]

\[
Q = 5 \times 1.0A \times 50\,\text{ns}
\]

\[
Q = 250\,\text{nC}
\]

(Where $N\square =$ Area of $I.t$ waveform in $\square$'s).
Volts Amps
120 0.6
80 0.4
40 0.2
50ns/div

Fig. 3.12 Mosfet Vds and Idd waveforms with no snubbing (link = 60V).
(Idd = drain current, Vds = drain source voltage)

Volts Amps
120 0.6
80 0.4
40 0.2
50ns/div

Fig. 3.13 Mosfet Vds and Idd waveforms with snubbing (link = 60V).

\[ E = Idd \times Vds \times tsw \]

Fig. 3.14 Calculation of Mosfet switching losses.

Volts Amps
120 1
80 0
40 -1
500ns/div

Fig. 3.16 Mosfet Vds and snubber current with no speed-up diode (link = 110V).

Volts Amps
120 1
80 0
40 -1
500ns/div

Fig. 3.17 Mosfet Vds and snubber current with speed-up diode (link = 110V).
This compares well with the theoretical charge on the capacitor when it charges to 100V:

\[ Q = C \cdot V \]
\[ Q = 2.2 \text{ nF} \times 100\text{V} \]
\[ Q = 220 \text{ nC} \]

This shows that the measured charge is well within the ±20% tolerance of the mosfet gate charge specification. Figure 3.16 shows the snubber discharge pulse to be of very similar form to, but of opposite sign from, the charge current waveform.

The power dissipated by the snubber network is given by:

\[ P_{\text{snub}} = C \cdot V_{pk}^2 \cdot f \quad (\text{W}). \]

(Where \( V_{pk} \) is the peak value of \( V_{ds} \), and \( f \) is the switching frequency - there are two snubbing events per switching cycle.)

Since the dump controller operates at 25kHz, and the capacitor was optimised at 2.2nF, the snubber power is just 0.055V^2 mW. This means that even at the peak transient drain-source voltage of 288V, the snubber power loss is just 5.0W.

The circuit was modified to see if the addition of a fast diode in parallel with the resistor would allow significantly faster capacitor charging, which could have further reduced the switch off voltage spike. However, it was found that the voltage and current oscillation was worse than without the diode, as shown in figure 3.17, and so this idea was not developed any further.
3.5.2 Flyback Diode

When the mosfets are switched off, the abrupt stopping of the load current leads to a voltage reversal across the dump load. To prevent this causing damage to the circuit, a flyback diode is fitted. To work fast enough to clamp the fast rising voltage transient, the device chosen is a BY229 fast recovery type. The current through this device is shown in figures 3.18 and 3.19 for a link voltage of 60V and 100V respectively. This current decay is described by the equation:

\[ I = I_0(1-e^{-t/\tau}) \quad (A) \]

The time constant (\( \tau \)) is measured from figure 3.18 as 1.5ms, and since \( \tau = L/R \) and the resistance of the firebar (R) is 31\( \Omega \), the total inductance is calculated as 45mH, dominated by the inductance of the fire bars. Neglecting the effect of reverse recovery time when D is large, the diode conduction loss is given by:

\[ P = \int_0^t V_{\text{ak}} I_0 (1-e^{-t/\tau}) \, dt \quad (W) \]

3.5.3 Mosfet Switching Losses

The mosfet switching waveforms were observed over a range of link voltages, with the peak load current varying in direct proportion to the voltage. During switch-on, the voltage falls before the current rises, and consequently the power loss is negligible (figure 3.20). However, at switch-off the voltage rises before the current has completely fallen, and so energy is lost at each switch-off. Switch-off waveforms are shown for link voltages from 60V to 210V in figures 3.21 to 3.23. The mosfet energy loss per device per switching transition is approximately calculated as, (where \( t_{\text{sw}} \) is the switching time);
Fig. 3.18 Flyback diode voltage and current waveforms (link = 60V).

Fig. 3.19 Flyback diode voltage and current waveforms (link = 100V).

Fig. 3.20 Mosfet switching waveforms (Vds and Idd of one device, link = 220V).

Fig. 3.21 Mosfet switch-off power losses (Vds and Idd of one device, link = 60V).

Fig. 3.22 Mosfet switch-off power losses (Vds and Idd of one device, link = 120V).

Fig. 3.23 Mosfet switch-off power losses (Vds and Idd of one device, link = 210V).
\[ E = 0.5 \ V_{pk} \ I_t \ t_{sw} \ (J) \]

This means that the total power loss for both devices is given by;

\[ P = V_{pk} \ t_{sw} \ f \ (W) \]

The increase in the rate of change of \( V_{ds} \) and \( I_{dd} \) as the link voltage increases means that the switching time only increases very slightly with link voltage, and so the power loss at switch-off is dominated by the \( V_{pk}.I \) product. Since in this circuit the current is directly proportional to the link voltage, the switching losses are proportional to the square of the link voltage.

Figure 3.24 shows how the switching “off” time varies with link voltage, from which is found the following empirical expression relating the switch off time to the link voltage:

\[ t_{sw} = (37 + 0.11V) \ ns \]

(For convenience, the switching time has been calculated as being from where the main slope of the current and voltage plots would intercept the \( I = V = 0 \) line, (figure 3.14)).

The total mosfet power loss per device is closely approximated by the sum of the conduction and switching losses:

- Conduction loss: \[ P = \frac{DV^2}{2R^2} \]

- Switching loss: \[ P = \frac{V_{pk}.V.t_{sw}f}{4R} \] (W)
Fig. 3.24 Mosfet switching time as a function of DC link voltage

Fig. 3.25 Circuit diagram of the dump load circuit
3.5.4 Gate Drive

The power for the main control electronics is derived from the 24V battery supply, but the Cuk power converter has an inverting function, and so the DC link voltage negative is connected to the battery positive. This means that, as with the Cuk converter IGBTs, an opto-isolator was needed for the gate drive, (figure 3.25).

The opto-isolator transistor output is buffered by the 7667 amplifier, specifically designed for mosfet driving, the power for which is provided by a floating +15V regulated supply. The 7667 totem pole output is current limited by the 22R output resistors, which gives a measured peak drive current of 350mA to each mosfet gate. This level of current (Ig) ensures that the mosfet is switched off quickly, hence reducing the switch off power loss.

The observed gate current and voltage waveforms show an oscillation is present, even when only one device is being driven, which is thought to be most likely due to stray inductance in the gate drive path. The energy necessary to switch each device on is found by examination of figure 3.26, with the relationship between the drain source voltage and the gate current shown in figure 3.27.

In a similar way to the calculation of the snubber charge in 3.5.1, by examination of figure 3.27, the total gate charge is approximately:

\[ Q_g = 3 \times 100\text{ns/div} \times 0.1\text{A/div} = 30\text{nC} \]
Fig. 3.26 Mosfet gate charge waveform. 
(Ig = gate current, Vgs = gate source voltage)

Fig. 3.27 Mosfet gate:drain waveform.

Fig. 3.28 Power-up/down mosfet/relay interlock circuit diagram.

Fig. 3.29 Operation of interlock circuit
The data sheet value for the average total gate capacitance is 1500pF ± 25%, and so the charge necessary to charge the gate to 15V is:

\[ Q = 1500 \text{pF} \times 15 \text{V} \]
\[ = 23 \text{nC} \]

The observed charge is therefore close to the range expected from the device data, and the total gate power is given by:

\[ P_g = 0.5 \times C \times V^2 \times f \quad \text{(W)} \]
\[ = 0.5 \times 1500 \text{pF} \times 15^2 \times 25 \text{kHz} \]
\[ = 4.2 \text{mW} \quad \text{(per device)} \]

This power loss is clearly negligible compared to the much greater conduction and switching losses of the drain source junction.

### 3.5.5 Fail-Safe Interlock Circuit

The loss of power to the dump load control circuit, for instance if the battery was removed, could leave the WTG completely unloaded. If this should occur, the normally closed relay in parallel with the mosfets is de-energised, and so the contacts short out the drain-source junction. This allows the WTG to be safely loaded by the dump load resistor, and slowly brought in to the stall region.

To avoid the arc that the relay contacts would draw when breaking a high voltage DC circuit, there is an interlock circuit which ensures that mosfets are switched on before the relay changes state. This action (figures 3.28 and 3.29), ensures that the relay only changes state when there is a low (just the mosfet on-state \( V_{ds} \)) voltage across the contacts.
3.6 Servo Design

This current mode servo (also shown in figure 3.25) is based around the industry standard 3524 PWM control IC. The dump current demand to the servo is set by the windspeed:current control circuit, and is set for a voltage of 250mV per amp of demanded current. The error amplifier in this IC is a transconductance type, with which it proved impossible to achieve both the necessary gain and stability. It was therefore decided to use an external summing amplifier, connected to the non-inverting input of the error amplifier within the 3524.

To determine the characteristics of the op-amp, it was necessary to characterise the output pulse width of the IC comparator as a function of input voltage. With a compensation resistor of 33k on the output of the 3524 error amplifier, the value of D from the PWM comparator was found to vary from 0-100% over the range of input voltage of 0.80V to 0.95V. The output of the external summing amplifier had therefore to be changed to have an output voltage swing to match this narrow input voltage range.

The first step was to include an attenuation network between the two amplifiers, made up of a 4k7Ω and 1k0Ω resistor. The value of Vo required for a 100% change in D is now:

\[
(4k7 + 1k0) \times 150mV = 855mV
\]

\[
1k0
\]

The voltage offset is obtained by adding a fixed negative current (of value Voff/Roff) to the summing amplifier input, and so the full equation for the servo error amplifier is now:

\[
I_{dem} + I_{dm} - I_{off} = I_o
\]
(Where \( I_{dem} \) = demanded current (always negative), \( I_{dm} \) = measured dump current, \( I_{off} \) = offset current and \( I_o \) = output current. \( V_{dem} \) etc. in the next equation are the related voltages).

These currents are set by the following voltages and resistances:

\[
\frac{V_{dem} + V_{dm} - V_{off}}{10k} = \frac{V_o}{10k + R_{off} + R_o}
\]

The servo is designed to regulate the output current such that a 10% change in actual current when at the maximum current demand will lead to a change in \( D \) of 100%. The value of the feedback resistor is therefore determined by considering the change in feedback voltage necessary to give this change in \( D \):

\[
V_{dm} = \frac{V_o}{10k + R_o}
\]

Since \( V_o = 0.855V \) and \( V_{dm} = 5V \times 0.1 \):

\[
R_o = \frac{(0.855 \times 10k)}{0.5} = 17.1k\Omega \quad (18k\Omega)
\]

If \( V_{off} \) is set to (-2.5V, adjustable by a potentiometer), then \( R_{off} \) is given by:

\[
R_{off} = \frac{(V_{off} \times R_o)}{V_o}
\]

\[
= \frac{2.5 \times 18k}{0.80V \times 5.7} = 10k\Omega
\]

---

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(Where Vo gives the output voltage necessary to reach the operating threshold of the comparator on the 3524 PWM IC).

3.7 SERVO TRACKING PERFORMANCE

The observed current tracking performance of the servo is shown for several values of nominal current demand as the voltage changes, (figure 3.30).

The turn-off delay of the opto-isolator driving the power mosfets from the PWM IC can be clearly seen in figures 3.31 and 3.32. This delay means that even with an output pulse width of just 2μs, the minimum on time that is possible is about 6μs. This is why when the nominal current demand is reduced to 0.20A, the dump current starts to operate in short bursts of current. The maximum repetition period of this is 10ms, which is due to the time constant of the current sensing feedback circuit. This leads to the mosfets being driven in sinusoidally modulated bursts, but the system remains stable, and so is able to function effectively at these low levels of current. However, the faster gate drive stage described in 2.37 would restrict this unsteady zone of operation to a much lower level of current.

3.8 TRANSIENT RESPONSE OF THE SERVO

The servo system has two changing inputs - the current demand and the link voltage. To obtain the transient response of the servo system, the dump current should be observed when both these parameters are independently and instantaneously changed, but as with the Cuk converter tests, it is only practical to test the step response from a change in demand current. The servo demand was changed by inserting an additional switched resistor into the demand circuit (figure 3.33), with the resulting step change in demand when the switch changes state giving the characteristic system response. The current was monitored using the filtered output of the LEM current module (figures 3.34 and
Fig. 3.30 Tracking of dump load servo, (marked currents are nominal currents)

Fig. 3.31 Mosfet switching delays (PWM IC output and Idd).

Fig. 3.32 Mosfet switching delays at low current demand (PWM IC output and Idd).
Fig. 3.33 Servo transient response test circuit.

Fig. 3.34 Transient performance of dump load servo, (+ve demand impulse).

Fig. 3.35 Transient performance of dump load servo, (-ve demand impulse).
3.35), and the time for the output current to make 67% of the transition was measured as c.3.3ms on both step increases and decreases in the current demand. There was no noticeable overshoot or instability, which implies that the servo response could probably be safely increased, although no practical requirement for this is foreseen.

3.9 SUMMARY

This chapter has described the use of a simple PWM controlled resistor circuit for loading the WTG when the battery is unable to absorb all the available power. The tests on the circuit show that it can easily dump the specified 2kW, and that the tracking and step response of the PWM control circuit is satisfactory.

This power and voltage limiting is an essential feature of any SBCWG, since both the speed (and hence output voltage) must be restricted to prevent damage to the WTG or the electronic controller. A potential problem with this type of PWM circuit is that if there is a long distance between the controller and the dump load, the self-inductance of the leads will increase the transient mosfet Vds spike at switch off. However, in practice it is expected that the dump load heat, which will most likely be used for space heating, could be placed within a few metres of the controller.

The dump load circuit was designed to absorb all the 2kW WTG power produced at a 20m/s windspeed. Above this windspeed, it is not sufficient to just electrically load the WTG to restrict its speed and hence output voltage, and so another means is needed. The next chapter describes the predicted performance of flexible (self-regulating) blades, and the effect that these would have on the design of the WTG controller.
CHAPTER 4

HIGH WINDSPEED PROTECTION
BY THE USE OF FLEXIBLE BLADES

4.1 HIGH WINDSPEED PROTECTION

In stand-alone off-grid installations, reliability is particularly important, since failure of the WTG would lead to total loss of power to that site. It is therefore critical to select a reliable method of limiting the WTG output power and speed during periods of high windspeed.

Several techniques, or combinations of techniques, are commonly used to ensure that WTGs can survive high windspeeds. These can be loosely classified as:

- rotation of the turbine out of the wind,
- varying the turbine aerodynamic efficiency,
- stall control or mechanical braking.

Some of these methods allow limited power generation at speeds in excess of the nominal rated speed. This gives greater average energy capture, particularly in areas of high average windspeed, although this is a secondary consideration to reliability.
This chapter examines the options for high windspeed control, the characteristics of the flexible blades which were selected, and the changes in the design of the controller that this allows. The basic controller design had assumed an abrupt limiting of power at the maximum rated windspeed, but the initial characteristics of the new blades show that this is impractical. Final changes to the control system must wait until after final blade tests have been completed.

4.2 HIGH WINDSPEED PROTECTION TECHNIQUES

This section lists the types of high windspeed protection considered:

Rotation out of the wind

- **Motor driven yaw control**: On larger machines a yaw axis motor is used for facing the turbine into the wind for maximum power capture up to rated speed. This same mechanism could also be conveniently used to yaw the turbine out of the wind in high windspeeds. The disadvantage is that the added complexity of a yawing motor, control wires, and servo control system make this unattractive for such a small WTG.

- **Tail flap furling**: The generator rotor is offset from the centre of the support pole such that at high windspeeds, the moment of the turbine about the pole axis is sufficient to overcome the restraining force on the hinged tail flap, (Ackerman, M., Musgrove, P. and Fawkes, J., 1989). The force on the tail flap may be provided by a spring, or more simply by gravity. However, experience of balancing smaller WTGs indicates that a passive furling mechanism on a machine of this size would be exceptionally difficult to correctly set up. This is mainly because as an upwind machine, it will be effected by the turbine wake or the aerodynamic shadow of the
support pole. Some models require manual resetting once furling through 90° has occurred, which is a severe disadvantage in unattended locations.

- **Tilt-back protection:** Some smaller machines, (such as the 1.4kW Australian Survivor windcharger) have a horizontally hinged nacelle. In high winds the force on the turbine makes the turbine progressively tilt backwards as it overcomes the opposing force of a spring.

**Varying the turbine aerodynamic efficiency**

- **Blade pitching:** The blades may be tilted to reduce their aerodynamic efficiency, and hence speed and power output, with the action usually achieved by an electrical or hydraulic actuator. This type of mechanism is most commonly used on simple water-pumping machines (Barlow, R., Crick, F. and Fraenkel, P., 1994), where the wind force on a drag plate parallel to the turbine mechanically pitches the blades.

- **Blade flap:** Tail edge flaps may be used to increase drag, slowing the turbine, which may be positioned either passively or by a controlled actuator. (Potential users have however indicated that blade flap control, blade pitching, and tilt-back protection would be perceived as being mechanically unreliable.)

- **Pneumatic spoiling:** Patterns of holes and airways can be included in the blades, which may be controlled by a valve to allow air through them, spoiling the aerodynamic efficiency. A study on a 500W machine, (Bannister & Fawkes, 1989) showed that a reduction of shaft speed of up to 30% is feasible, which was thought sufficient to protect the machine. This technique clearly has potential, but was thought to represent too great a technical risk.

- **Flexible blades:** The blades can be sufficiently flexible that they actually bend backwards in strong winds, thereby reducing their aerodynamic efficiency, and hence
output power. Alternatively, the blade tips may be weighted to progressively twist the blades by the centrifugal force as the shaft speed increases, also spoiling the aerodynamic efficiency.

Stall control or Mechanical Braking

- **Stall control**: This technique is based on the control system demanding a generator torque greater than that required for maximum power production. The rotational speed of the turbine is therefore reduced (starts to stall), and the output power of the wind turbine falls. If a resistive dump load is chosen, the heat dissipated may itself be utilised as useful energy for space or water heating. A limitation on very small WTGs is that the resistance of the generator severely limits the maximum possible current, with field tests on a Marlec 12V 72W WTG showing that even a short circuit on the generator output was insufficient to adequately brake the machine. On the significantly larger generator necessary in this project, the generator and downlead resistance are likely to be much lower, and so a larger electrical braking torque is possible. However, the control of the high currents which might be necessary make this method unattractive.

- **Mechanical brakes**: Hydraulic or electrically operated brakes may be used to stall the machine under high wind speed conditions, as well as during installation or maintenance. Frequent braking imposes stressful mechanical torque transients, and the brake lining will wear and eventually need replacement. However, machines of 2kW and below almost never have mechanical brakes, as the cost and mechanical complexity cannot be justified.

After carefully considering all the options, it was decided to opt for the simplicity of the flexible blade approach. This was made possible by the availability of expertise at Loughborough University, where the blades are now being designed under a three year CASE project at the CREST centre, (Feuchtwang J.B. and Infield D.G., 1995).
4.3 IMPLICATIONS OF USING FLEXIBLE BLADES

Flexible blades have been used successfully on several designs of wind turbines, with examples of manufacturers including Bergey, Proven and Carter. Their principal advantage is that in high windspeeds they offer a passive method of protecting the turbine, but they are also able to generate some power at these speeds. Limiting the power at the turbine, rather than further down the power transmission system, means that these other components do not have to be rated for higher powers.

The remainder of this chapter examines the key implications for using flexible blades on the WTG of the 2kW SBCWG controller. The most important implication is that the passive power regulation is considered much more reliable and robust than reliance on control from a separate electronic controller. This means that the dump load is no longer essential for ensuring that the WTG is safely loaded, but is still useful for providing heat.

4.4 DESIGN OF THE FLEXIBLE BLADES

The flexible blades are based on a novel type of glass fibre construction, where the direction of fibres in subsequent layers is carefully calculated to make the blade twist under the centripetal force in a carefully controlled manner. Figure 4.1 is a sketch of this type of blade, with the twisting mechanism indicated by arrows. Initial computer predictions of the performance of the designed blade are shown in figure 4.2.

The remarkable manner in which the power output is limited to 2,000W at speeds of up to 80m/s is best illustrated in figure 4.2, showing a change in power output of just 500W when the available (wind) power increases by 130kW (between 20 m/s and 75m/s). Although not shown here, the rate at which the WTG increases in speed falls
Fig. 4.1 How the blades twist under the centripetal force

Fig. 4.2 Predicted characteristics of the flexible blades (0 to 75m/s)

Fig. 4.3 Cuk converter current using flexible blades
dramatically with windspeed. This also limits the wind turbine voltage output, so protecting the electronics.

The graph also shows that this type of self-regulating turbine cannot perform as cleanly as the ideal power limiting mechanism assumed in the initial controller design - in which the power output was assumed to be abruptly limited at the maximum operating point. The penalty for this is that to restrict the maximum output power to just 2,000W, the wind turbine efficiency is deliberately reduced above just 8 m/s. The need for good low windspeed energy capture has therefore led to a wind turbine that is oversized above this relatively low windspeed. Alternative designs could trade off this low windspeed performance for a smaller and hence cheaper wind turbine, and then rely on a smaller reduction in aerodynamic efficiency at higher speeds to still enable the power output to be limited to just 2,000W.

4.5 CUK CONVERTER DESIGN CHANGES

It is assumed that the maximum power rating and input voltage of the Cuk converter with the flexible bladed WTG is the same as the fixed blade WTG. Since at all windspeeds up to the rated maximum, the available power (figure 4.3) is greater than that of the fixed bladed turbine, it means that for instance at 10m/s, the Cuk converter input current is now 8.3A, compared to just 2.1A at this windspeed with the earlier design. The reduction in speed and hence WTG voltage at higher windspeeds means that the current drawn from the WTG actually falls above around 20m/s.

These changes mean that some of the Cuk component specifications will need changing - in particular the input choke current rating will need raising. In addition, the setting of the separation of the upper and lower current thresholds on the boundary control circuit will need reviewing.
4.6 DUMP LOAD DESIGN CHANGES

The integrity of the dump load circuit originally dictated that it was placed close to the output of the WTG, effectively in parallel with the input of the Cuk converter. But since failure of the dump load circuit with a flexible-bladed turbine no longer compromises the mechanical or electrical reliability, this is no longer essential. It is therefore possible to move the dump load to the battery (figure 4.4), effectively allowing it to operate as a battery regulator. Unfortunately, a disadvantage of this is that the dump load efficiency is now the product of the Cuk converter efficiency and the dump load efficiency. However, two important practical benefits of this come from this new arrangement:

- The dump load, if properly sized, can serve as a regulator for additional battery charging sources, such as solar panels. This will reduce the amount of equipment and wiring in an installation, giving an attractive improvement in reliability, and also a reduction in total system cost.

- It is no longer necessary to calculate the split between battery charging and dump load current, reducing the number of microcontroller sub-routines needed.

Since the WTG blades are self-regulating, the fail-safe relay and associated drive and interlock circuits are also no longer needed.

4.7 NEW CONTROL SYSTEM

The cost premium of the flexible blades over the fixed blades means that costs must be reduced elsewhere in the design. Fortunately, the flexible bladed turbine means that a much simpler control system can be used:
Fig. 4.4 New system topology using the flexible blades

Fig. 4.5 WTG Power: shaft speed characteristics

Fig. 4.6 Characteristics of the dump load as a battery regulator
- No dump control transistor bypass relay is needed to prevent equipment damage at high windspeeds,
- The much simpler control system must now only control the WTG current to follow the characteristics shown in figure 4.5,
- No dump load/battery charging calculation is needed. Instead a simple constant voltage regulator servo is adequate, (figure 4.6),
- No action needs to be taken to avoid mechanical or electrical damage at windspeeds above the rated maximum.

Two low cost circuit designs are suggested for matching the WTG current to the windspeed, (figures 4.7):

**Analogue multiplier**

Since for an ideal WTG system the current demand is proportional to the DC link voltage squared, an analogue multiplier can be used to generate a squaring function, and the output inverted to match the negative input of the Cuk control servo. This circuit (figure 4.8) was built and performed well once the potentiometers had been adjusted for the correct gain and linearity, with an accuracy within ±1% of the ideal square law curve. However, in practice this type of circuit will drift with time and temperature. A key problem with this approach is that unlike fixed blades which follow much more closely the ideal current:windspeed square law, the power output from the flexible blades deviates much further from this simple law.

**Look-up table**

The look-up table technique stores the pre-determined current demand as a function of the link voltage in a memory chip, such as an eprom, (Smith, G.A., Undated). In this circuit, (figure 4.9) the link voltage is converted by an Analogue to Digital Converter (ADC) to give a digital eprom address, and a Digital to Analogue Converter (DAC) then converts the current demand data from that address to an analogue voltage, which sets the servo demand. It has several advantages over the analogue solution:
Fig 4.7 Possible Analogue WTG current control

Fig 4.8 Analogue multiplier circuit

Fig 4.9 Possible eprom based WTG current control circuit
Negligible temperature drift,
No complicated setting of potentiometers, (needed for setting up the analogue multiplier gain and linearity),
If the flexible blades have characteristics that deviate slightly from the ideal square law, then the data can be altered to suit,
The stored curve can have "notches" to avoid resonant frequencies of components such as the blades or mounting pole,
The curve can be easily modified to give no current demand at very low windspeeds, allowing the wind turbine to accelerate to the cut-in windspeed with no electrical load.

Since the microcontroller is no longer needed for the control of the SBCWG, it is possible to reduce the cost significantly by removing the microcontroller altogether. This is because the additional features that the microcontroller allows could originally only be justified because it was needed anyway for basic windspeed:power matching. Losing the microcontroller will mean the loss of the digital keypad and display for monitoring and data entry. However, the only parameter which the user will now have to set is the battery regulator voltage and the scaling factor for the current:windspeed relationship to cope with different battery voltages. Both of these can be altered in the field, or alternatively preset in the factory. An analogue battery voltmeter and WTG ammeter can be included to give the user this basic system information.

4.8 SUMMARY OF THE USE OF FLEXIBLE BLADES

The decision to use flexible blades part way through the project had significant implications for the design of the control unit. In particular, the self-regulating power output means that the major problems of excess power or too high a link voltage are eliminated. However, the additional cost of these blades means that it is a commercial
requirement to reduce the cost of the control system, and so it is proposed that the microcontroller and associated circuitry is removed. This is only possible because in addition to the above advantages, the dump load now operates independently as a simple constant voltage battery regulator, and so the microcontroller no longer undertakes any critical control or safety functions.

The modification of the controller to suit the computer generated characteristics of the WTG has necessarily been largely theoretical, since the final wind turbine design will not be available until after the controller design has been largely finished. However, the computer-based wind turbine simulator described in the next chapter allows the effect of different wind turbine designs to be simulated, and will allow development of wind turbine control equipment without needing to use a real wind turbine.
CHAPTER 5

WIND GENERATOR TEST RIG

5.1 THE NEED FOR A WIND GENERATOR TEST RIG

This equipment is an important part of the research project, allowing full testing of the SBCWG controller electronics and control strategy under controlled laboratory conditions, (Napoli et al, 1989). This overcomes the many practical problems that occur when using real test sites for system trials:

- **Predictability of wind conditions**: The unpredictability of windspeeds means that it is difficult to plan in advance when to do experiments.

- **Component Failure**: Subjecting a prototype SWTG to extreme conditions could lead to component failure.

- **Blade Design Changes**: The $c_p\lambda$ characteristics of blades must be physically altered to investigate the performance of different turbine characteristics, whereas using this model involves far less time and expense.

- **Monitoring**: It is expensive to build, properly equip and visit a site test hut.

- **Data Logging**: Data logging may not be practical for the short term variations that are of interest in optimising a control system, since it can be very hard to extract the required information of interest from the large amounts of data that will be logged.
5.2 THE WIND GENERATOR TEST RIG

The WGTR runs in two modes - Computer only, or a computer controlled test rig. Both systems allow the simulation of a defined wind turbine under real wind conditions - the difference being that in the computer only mode all calculations are performed within the computer, whereas in the test rig mode a generator is actually driven by a computer controlled motor. Electrical loading of the generator is controlled by the wind turbine control system, which is independent of the test rig computer.

5.3 PREPARING THE SIMULATION

Windows environment
The software runs under a windows environment, with data entered through a series of pull-down menus, as shown at the top of figure 5.4.

Wind data
The wind data is obtained from real wind patterns logged at a site in the Orkney Isles at two second intervals, with forty 18-minute samples available to choose from, (an example is shown in figure 5.1). This windspeed data may also be scaled, giving a further variety of wind patterns.

Turbine performance
The $C_p, \lambda$ curve is entered by using the mouse to adjust the five points on the graph shown in figure 5.2. A line-fitting routine joins the points, and displays the $C_p,\text{max}$ and $\lambda,\text{max}$ values. This graph is made up of over 1,000 points, which for speed of simulation are stored in a look-up table.
Fig. 5.1  Plot of stored windspeed against time

Fig. 5.2  Turbine $c_p$-$\lambda$ curve

Fig. 5.3  Entry/Display of turbine characteristics

100
Other turbine details
Other turbine details and operating limits are entered in the screen shown in figure 5.3.

Running
The torque:windspeed curves and relevant operating limits of the turbine are plotted, (all three limits are shown in figure 5.4, although usually only one or two would actually be used). In the case of the grid-connected example shown, the wind turbine starts from standstill and is gradually accelerated by the wind until it reaches synchronous speed. No generator torque is taken until it has reached the cut-in windspeed. The windspeed:torque point and corresponding turbine speed is updated every 400ms, (marked with a cross), and a lighter trail gives the past history of the turbine. In this example the turbine torque of a fixed speed grid-connected wind turbine has followed the maximum torque curve up to synchronous speed, where it has then remained, with the load torque changing as the generator torque changes.

A box at the top left-hand corner shows the cumulative power taken from the generator, and is a useful figure for comparing different designs of turbine or wind conditions. A fast mode allows the whole 18-minutes of windspeed data to be simulated in the model to give a fast cumulative kWh reading, which is very useful for rapidly assessing the effect of design changes.

5.4 COMPUTER ONLY SIMULATION

Blade torque
For variable speed operation, the blade tip:speed ratio \( \lambda \) in equation 5.1 will normally be at the optimum, \( \lambda_{\text{opt}} \). For wind speed \( v \) and blade radius \( R \), the shaft speed \( \omega \) can therefore be calculated from the relationship:

\[
\omega = \frac{v \lambda}{R} \quad \text{(Equation 5.1)}
\]
Fig. 5.4    Simulator Display
The value of $c_p$ corresponding to $\lambda$ is found from the look-up table describing the turbine characteristics. The blade torque ($T_b$) is then given by:

$$T_b = \rho \ c_p \ x \ R^2 \ v^3 / 2 \ \omega$$  \hspace{1cm} \text{(Equation 5.2)}$$

(where $\rho$ = density of air (1.23 kg/m$^3$))

In the constant speed mode, the shaft speed is almost constant, (mimicking the effect of a generator connected to a fixed frequency mains supply). The shaft speed is given by:

$$\omega = \omega_s \ (1 + s)$$ \hspace{1cm} \text{(Equation 5.3)}$$

Where $\omega_s$ is the synchronous speed, and $s$ is the value of slip, which will have a maximum value ($s_{\text{max}}$) of $\approx 0.03$ to $0.05$. The torque is then given by:

$$T = T_{\text{max}} \ (\omega_s - s)/\omega_s$$ \hspace{1cm} \text{(Equation 5.4)}$$

In the $c_p$ maximum/constant speed mode, equation 5.2 applies until $\omega_s$ is reached, thereafter equation 5.4 is used.

**Generator torque**

The speed demand is updated at intervals ($\Delta t$) of 0.1s. For the maximum power transfer condition, the generator load torque ($T_g$) should traverse the maximum $c_p$ curve:

$$T = \rho \ c_p \pi \ R^2 \ v^3 / 2 \ \omega$$ \hspace{1cm} \text{(Equation 5.5)}$$

Since on this curve $\lambda_{\text{max}} = \omega \ R / v$:

$$T_g = (\rho \ c_p \pi \ R^5 / 2 \ \lambda_{\text{max}}^3) \omega^2$$ \hspace{1cm} \text{(Equation 5.6)}$$

**Initial conditions**
At the start of a simulation there is no past history on which to calculate the starting speed of the turbine, and so in order to ensure that no energy is lost or gained over the cycle, the difference in stored rotating energy is added or subtracted from the kWh figure:

\[
\text{Net kWh} = \text{Final kWh} - 0.5J(\omega_0 - \omega_f)^2 \quad \text{(Equation 5.6)}
\]

(where \(\omega_0\) = initial shaft speed, \(\omega_f\) = final shaft speed, and \(J\) = angular moment of inertia of the turbine rotor and generator).

This model can be used for turbines of any reasonable size, and the actual generator torque is always assumed to be the optimum for that windspeed. However, when the software is used to drive a test rig, these factors are not always true, and so give rise to additional complications.

5.5 THE TEST RIG

Details of the hardware design (figure 5.5) of the WGTR for dynamic running are included in Appendix 8, and the basic system illustrated in figures 5.6 and 5.7. The software can monitor both the instantaneous torque taken by the generator load (\(T_g\)), and the instantaneous shaft speed (\(\omega\)). From these ‘real-world’ inputs, and the stored wind data, it calculates and outputs an up-dated control signal to the closed-loop controlled DC.

5.6 MODELLING A VERY LARGE TURBINE

If using the test rig to model a turbine much larger than that of the small (<9kW) generator fitted to the rig (figure 5.6), both the speed demand to the motor from the computer and the torque feedback can be appropriately scaled. Strictly, the torque
Fig. 5.5 Wind Generator Test Rig (configured for modelling a large generator (fig. 5.6))

Fig. 5.6 WGTR: Configuration for modelling a large generator

Fig. 5.7 WGTR: Configuration for modelling a small generator
The transducer is actually measuring the sum of $T_g$ (the load torque) and that of the accelerating torque, since:

$$T_{\text{meas}} = T_g + T_{\text{acc}}$$  \hspace{1cm} \text{(Equation 5.7)}

(Where $T_{\text{meas}} = \text{measured torque}$, $T_g = \text{generator torque}$, $T_{\text{acc}} = \text{accelerating torque}$.)

However, the inertia of such a large generator plus turbine will be so much larger than that of the test rig generator, that the large value of $J$ will mean that the accelerating torque is negligible compared to the generator torque, and so $T_{\text{meas}}$ is taken to be the same as $T_g$. The following analysis is also used in the computer only simulation already described.

The angular acceleration of the shaft is found from:

$$J \frac{\Delta \omega}{\Delta t} = T_b - T_g$$  \hspace{1cm} \text{(Equation 5.8)}

(where $T_g = \text{generator (load) torque}$)

This assumes that both the shaft is of infinite stiffness, and that there is no damping. Under steady-state conditions, $T_b = T_g$, and so the shaft speed is constant. However, as the generator load increases, the net load torque decreases (becoming negative), and so the shaft decelerates. Similarly, as $T_g$ decreases, the shaft will accelerate. The instantaneous speed change is calculated, with the modelled value of $J$ limiting the acceleration, and hence speed increment $\Delta \omega$, such that;

$$\omega_{\text{new}} = \omega_{\text{previous}} + \Delta \omega$$  \hspace{1cm} \text{(Equation 5.9)}

The motor drive maintains this speed profile, drawing as much current as is necessary to satisfy the speed demand. The inertia of the motor is low, and so will have a negligible effect on the speed response.
5.7 MODELLING OF A SMALL TURBINE

If the test rig is being used to control the actual generator on the rig (figure 5.7), then the accelerating torque of such a small generator will no longer be negligible, and so a different method of calculation must be used. The proposed solution is to overcome the torque measurement problem by using the computer to output a demanded blade torque to the drive, which also allows the inertia of the turbine and generator to be modelled within the computer. The torque transducer in this case measures the actual torque produced by the generator, which comprises the generator and accelerating torques:

\[ T_b = T_g + T_{ace} \]  \hspace{1cm} (Equation 5.11)

The torque demand is converted to a current demand to the drive, with a current servo using motor current feedback to overcome any drive losses, (figure 5.7). This closed-loop current servo is then trimmed using feedback from a shaft torque servo \( T_B \) in equation 5.11), which has the advantage of giving sufficient control with a much smaller servo gain, improving the system stability.

5.8 SUMMARY

This chapter has described the design of an easy-to-use Windows-based software package which allows the performance of a wind turbine of any size to be observed under different control or wind conditions. An extension of the work allows a real turbine generator of up to 7.5kW to be controlled by a motor controlled by the PC. This allows the laboratory testing of SBCWG controllers under a wide range of simulated wind conditions without the time and expense of real field testing.
CHAPTER 6

CONCLUSIONS AND FURTHER WORK

6.1 AIMS AND OBJECTIVES OF THE PROJECT

The main aim of this work was to design a reliable and efficient controller for charging batteries from a small (2kW) Wind Turbine Generator. For simplicity and reliability, a fixed blade turbine with a permanent magnet generator was originally specified, meaning that all control of the WTG must be done by the control unit. Whilst this work has shown that such an approach is possible, a further project is now under way to develop a design of flexible blades which could considerably simplify the control unit. In addition, an earlier report on this project to the sponsoring company has led to a separate project being started to design a brushless DC generator, shifting the cost and complexity from the control unit to the generator.

6.2 RELIABILITY

The performance of the controller contributes to the reliability of the whole system through both the reliability of the controller itself, and indirectly by the way in which it loads the WTG and controls the charge and discharge of the battery bank.
A critical success factor in a SBCWG design is the reliability of the control system, and so particular care was taken in the design of the electronics to ensure protection from common faults such as loss of battery, or excessive battery voltage. A particularly attractive feature of the Cuk converter is that any additional external inductance in connecting leads actually improves the circuit performance through greater current smoothing, rather than creating problems with switching voltage spikes which may be seen with other topologies.

The non-pulsating input and output currents mean that problems due to electrical interference are minimised. While stand-alone power generation equipment is not usually required to comply with existing regulations, this is clearly advantageous in reducing the possibility of interference to the communications systems which are frequently found on such supplies.

The WTG is protected against the danger of overspeed by ensuring that the total electrical load from the battery charger and dump load is always sufficient to match the available power. In the event of total loss of power to the controller, a de-energised relay connects the dump load directly across the WTG output, and so acts as a crude brake.

Above the maximum windspeed for power generation, an ideal means of mechanically limiting the generated power was assumed. However, the later decision to use self-regulating blades means the speed of the turbine is now reduced at speeds above the maximum, and so the generator can safely produce power in windspeeds of up to 75 m/s.

A practical problem on smaller 150W WTGs is that sudden changes of electrical load can ultimately lead to complete WTG failure, due to the actual generator assembly coming detached from the shaft. The smooth transition between the dump load and the
battery charging load mean that there should no longer be the possibility of such sudden torque shocks on the generator, and so this should not be a significant problem.

The simple voltage limiting battery charging cannot make maximum use of the battery charge capacity, but with temperature compensation, it would certainly be a significant improvement on current methods. Most of this benefit comes from the controller being able to split the WTG power between the dump load and the battery charger when the battery is almost or fully charged. This avoids the need in some existing circuits, such as that described in appendix 7, to effectively discharge the battery by diverting all the WTG power into the dump load until the battery voltage has fallen to a much lower level.

6.3 CONTROLLER EFFICIENCY

The advantage of the Cuk converter is that as a step-up and step-down voltage converter, batteries can be charged in low windspeeds when the WTG voltage is actually less than that of the battery bank. This enables greater power capture at low windspeeds.

The actual power converter used for experiments was not as efficient as hoped, but several reasons are given for this. One of the known significant contributors to the power loss is that of the output inductance resistance. By the use of the integrated magnetics techniques, the size of core needed for the inductance value can be greatly shrunk, so allowing a larger cross-section of copper conductor to be used for the same expense. Additionally, at extremes of operation where converter efficiency declines, a variation on the basic converter, such as using a transformer or tapped inductor, could be used to possibly decrease overall losses. Clearly, more work should be done in finding out exactly where the losses in the experimental test rig are occurring, both through experimentation and possibly developing a simulation using an improved version of Spice.
One of the drawbacks of the Cuk converter is the high ripple current rating of the capacitor, which is difficult to maintain at very low levels over the range of circuit operation. A ripple current rating leads to both a high cost for the capacitor, since this determines the lifetime of the capacitor, which is likely to be the least reliable device in the controller. Although the proportional magnitude boundary control circuit worked well, under some extremes the frequency of the Cuk converter became very low, imposing a large ripple current on the capacitor, and also increasing losses in other components. An investigation should also be extended in to reducing the boundary width to maintain a minimum frequency.

Several options for a control circuit to control the WTG current to be a function of windspeed were discussed, with a simple eprom-based look-up table being the most favoured option. However, further work should be done to explore the use of the hill-climbing technique to automatically compensate for variations in the WTG characteristics.

The use of flexible blades was shown to be very beneficial to the WTG system, allowing for a simpler controller design and safe operation at high windspeeds, which both protects the WTG system and gives additional power capture beyond the normal range of operation.

### 6.4 Wind Generator Test Rig

The development of the software for simulating wind turbines of different sizes and characteristics under real wind conditions, will be valuable not only for further work on the 2kW Wind Turbine featured in this research work, but also for other turbines. The extension of this software to allow the control of a real wind turbine generator, either the actual one to be used (where less than 7.5kW), or as a representation of a much
larger turbine, will be particularly useful for testing the behaviour of real wind turbine controllers.

6.5 FURTHER PROGRESS OF THE PROJECT TO DESIGN A 2kW SBCWG SYSTEM

An interim progress report on this controller project strongly endorsed the use of flexible blades. However, the company has now decided to also use a DC Brushless generator, allowing the controller to be rated for just a proportion of the power. This gives scope for reducing both the cost and the losses in the controller, although additional cable to control the rotor field strength are necessary. This will add cost to the brushgear and the generator design, but it is expected that the overall cost will be lower.

This decision is however a reflection as much on the skills of the Marlec design engineers as on the technical merits of the options, and it is anticipated that many aspects of this work will be incorporated into the final design.

6.6 SUMMARY

This work has demonstrated that the Cuk converter operating under boundary level control is suitable for controlling a battery charging permanent magnet (un-controlled) WTG. In addition, the use of a variable dump load allows optimum loading of the WTG under all load conditions, and gives space or water heating. The use of a flexible bladed Wind Turbine has been shown to reduce the complexity of the controller, and limit the power generated in high windspeeds.
The Wind Generator Test Rig project achieved its aims of allowing computer simulation of WTGs under real wind conditions, and with further work could have been used to mimic a real WTG for loading of the 2,000W WTG controller.
APPENDICES

APPENDIX 1  List of acronyms
APPENDIX 2  Selection of batteries for SBCWG installations
APPENDIX 3  The Cuk converter
APPENDIX 4  Circuit design and description
APPENDIX 5  Microcontroller control system
APPENDIX 6  Wind generator test rig
APPENDIX 7  Redesign of the 500W wind turbine controller
### APPENDIX 1

#### A1 LIST OF ACRONYMS

<table>
<thead>
<tr>
<th>Acronym</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>ADC</td>
<td>Analogue to Digital Converter</td>
</tr>
<tr>
<td>BUFFALO</td>
<td>Bit User Fast Friendly Assembler Language Operation</td>
</tr>
<tr>
<td>DAC</td>
<td>Digital to Analogue Converter</td>
</tr>
<tr>
<td>EPROM</td>
<td>Erasable Programmable Read Only Memory</td>
</tr>
<tr>
<td>EEPROM</td>
<td>Electronically Erasable Programmable Read Only Memory</td>
</tr>
<tr>
<td>HCl1</td>
<td>Motorola 68HC11 Microcontroller</td>
</tr>
<tr>
<td>IGBT</td>
<td>Insulated Gate Bipolar Transistor</td>
</tr>
<tr>
<td>LCD</td>
<td>Liquid Crystal Display</td>
</tr>
<tr>
<td>Mosfet</td>
<td>Metal Oxide Semiconductor Field Effect Transistor</td>
</tr>
<tr>
<td>Ni-Cad</td>
<td>Nickel Cadmium (Battery)</td>
</tr>
<tr>
<td>OCV</td>
<td>Open Circuit Voltage</td>
</tr>
<tr>
<td>PID</td>
<td>Proportional, Integral, Derivative (Control)</td>
</tr>
<tr>
<td>PM</td>
<td>Permanent Magnet</td>
</tr>
<tr>
<td>PRU</td>
<td>Port Replacement Unit</td>
</tr>
<tr>
<td>PWM</td>
<td>Pulse Width Modulation</td>
</tr>
<tr>
<td>RAM</td>
<td>Read Addressable Memory</td>
</tr>
<tr>
<td>RMS</td>
<td>Root Mean Square</td>
</tr>
<tr>
<td>ROM</td>
<td>Read Only Memory</td>
</tr>
<tr>
<td>SBCWG</td>
<td>Small Battery Charging Wind Generator</td>
</tr>
<tr>
<td>SG</td>
<td>Specific Gravity</td>
</tr>
<tr>
<td>SLI</td>
<td>Starting Lighting Ignition</td>
</tr>
<tr>
<td>SOC</td>
<td>State of Charge</td>
</tr>
<tr>
<td>SWTG</td>
<td>Small Wind Turbine Generator</td>
</tr>
<tr>
<td>VLA</td>
<td>Vented Lead Acid (Battery)</td>
</tr>
<tr>
<td>VRLA</td>
<td>Valve Regulated Lead Acid (Battery)</td>
</tr>
</tbody>
</table>
APPENDIX 2

BATTERIES FOR SBCWG INSTALLATIONS

A2.1 Types Of Batteries

Secondary storage batteries for use on SBCWG systems are usually one of the following types:

- Vented lead acid:
  - Stationary (Plante, tubular plate)
  - Traction

- Sealed lead acid
  - Absorbed electrolyte
  - Gelled electrolyte

- Vented nickel cadmium (ni-cad)

It is not unusual for the initial purchase cost of the batteries alone to exceed the cost of the remainder of the SBCWG system, and so correct choice of batteries is critical. Experience at Marlec (Fitches, P., 1993) has shown that most system problems are due to poor choice or maintenance of batteries. In particular, the deep discharge cycling that is characteristic of SBCWG installations means that the low cost Starting, Lighting, Ignition (SLI) lead acid batteries (commonly used in cars) should not be used, since they will rapidly fail through sulphation.
A2.2 Specification Of Batteries

Choice of batteries is principally dictated by the following factors:

- **Mechanical:** Physical dimensions, weight, vibration resistance.
- **Electro-chemical:** Volume and composition of emitted gas.
- **Voltage:** Nominal voltage, voltage regulation.
- **Current capability:** Peak discharge current, maximum charge rate.
- **Monitoring:** Ease of State of Charge (SOC) monitoring.
- **Cost:** Installation cost, maintenance costs.
- **Maintenance:** Maintenance intervals, battery lifetime.
- **Cycle type:** Number of cycles, depth of discharge, deep discharge survival capability, discharge rate, available capacity.
- **Temperature:** Maximum and minimum operating temperatures

In practice, almost all installations use either vented ni-cad or vented lead acid batteries, with ni-cads giving better performance, but at a higher cost. Typical characteristics of these common cells are included in figure A2.1.

A2.3 VENTED SECONDARY STORAGE CELLS

Advantages of Vented Ni-cad over Vented Lead Acid (VLA) batteries

- Slower self-discharge rate.
- Operation and storage over a wider temperature range, (Brambilla, M., Gregorio, R., Paglini, G., 1990).
- Virtually undamaged by being left in a deep discharge state.
<table>
<thead>
<tr>
<th>Parameter</th>
<th>Units</th>
<th>Vented Ni-cad</th>
<th>Vented Lead-Acid</th>
<th>VRLA</th>
</tr>
</thead>
<tbody>
<tr>
<td>Cell terminal voltage (at 25°C)</td>
<td>Volts</td>
<td>1.45</td>
<td>2.40</td>
<td>2.45</td>
</tr>
<tr>
<td>Cell temperature compensation voltage</td>
<td>mV/°C</td>
<td>-0.35</td>
<td>-4.0</td>
<td>-4.0</td>
</tr>
<tr>
<td>Minimum charge temperature</td>
<td>°C</td>
<td>-50</td>
<td>-20</td>
<td>-20</td>
</tr>
<tr>
<td>Maximum charge temperature</td>
<td>°C</td>
<td>+55</td>
<td>+50</td>
<td>+50</td>
</tr>
<tr>
<td>Equalisation Voltage</td>
<td>Volts</td>
<td>1.67</td>
<td>2.70</td>
<td>2.50</td>
</tr>
<tr>
<td>Lifetime</td>
<td>Years</td>
<td>10 - 20</td>
<td>2 - 4</td>
<td>&lt; 5</td>
</tr>
</tbody>
</table>

Fig. A.2.1  Typical characteristics of secondary cells

Fig. A.2.2  Vented ni-cad cell voltage against state of charge (at 0.1C)

Fig. A.2.3  Vented lead-acid cell voltage against state of charge (at 0.1C)
• Much less likelihood of plate failure (100 times less in tests on solar installations).
• Peak charge and discharge currents much greater.
• Much less damaged by overcharge.
• Less reserve capacity needed to avoid damaging deep discharge.
• Greater lifetime, much less effected by loss of plate material.
• More robust construction.
• More constant discharge voltage characteristic.
• Little electrolyte heating during charge or discharge, and so less cooling is required, and there is less risk of thermal runaway, (Berndt & Weissner, 1990).
• Current ripple on LA batteries should be limited to less than 0.1C A rms, (Nagai & Ozaki, 1990).

Disadvantages of Ni-cad compared to VLA batteries

• Difficulty of monitoring SOC, (since the SG and OCV remains fairly steady).
• Lower cell voltage means more cells are required for a given battery voltage.
• The cadmium in old batteries is toxic, and so must be carefully disposed of.

Ni-cads are clearly preferable for unattended sites, but the cheaper VLA type may be used if regularly maintained. Whatever type of vented battery is chosen, there is a compromise between the amount of stored charge and the loss of electrolyte through electrolysis during the final charge phase. Where maintenance visits are infrequent, it is common practice to considerably undercharge the battery (de-rating to as little as 50% of nominal), in order to reduce this electrolyte loss. Charging to a higher voltage in order to achieve this greater charge storage capability also reduces the effect of sulphation, but the gassing action causes active material to be lost from the plates, reducing battery life.
A2.4 SEALED LEAD ACID (VRLA) CELLS

The gelled type of VRLA (Valve Regulated Lead Acid) batteries are popular for leisure type applications because of their low maintenance requirement, robustness, and lack of emitted fumes. Their plates are held in a gelled electrolyte, making sulphation damage much less significant, and so are able to withstand deep discharge and heavy cycling conditions, (Harison & Bullough, 1990).

Dryout is the most significant cause of failure, due primarily to gassing and valve deterioration, and can be a serious problem in remote installations with only infrequent maintenance. VRLA batteries also have poorer heat transfer characteristics than vented types, and so more care must be taken with cooling, and ripple current should be kept low. Temperature compensation of the charge voltage is also important, (Gerner, Britmyer & Bornemann, 1990).

Small sealed ni-cad batteries are increasingly popular for applications with lower energy storage requirements, and consequently a range of battery management ICs intended for these applications has been developed. These allow the use of battery terminal voltage, temperature, charge time, and rate of change of voltage for SOC determination. Unfortunately, both the irregular charge current availability and the need for the battery bank to be connected to the load during charging, mean that only the terminal voltage method is appropriate in SBCWG installations.

A2.5 STATE OF CHARGE MONITORING

Being able to monitor the state of charge and total capacity of a battery bank is important both as a measure of how much energy is stored in the battery, and to prevent over-charging the battery. All methods of SOC measurement are limited by the settling time of the electrolyte, and so should be temperature compensated, (Armanta et al,
This settling time may be largely overcome by electrolyte agitation, (Lakeman, 1989), but the added complexity makes it impractical in remote SBCWG installations. Ideally, each individual cell in a battery bank should be monitored (Healy & Rhyne, 1990), but this too is impractical.

Many methods may be used for monitoring lead acid cells, since both their specific gravity and voltage varies significantly with state of charge, but ni-cads are harder to monitor, since they show less change in voltage, and almost no change in SG.

**Total discharge**: This involves the total discharge of a battery through a known load, where the capacity is a fraction of load current and time. Whilst accurate, it is usually unacceptable to totally discharge a battery in a remote SBCWG installation.

**Summed ampere-hour measurement**: This involves summing the total charge and discharge current from an initial known SOC, and should take into account the variation in capacity with charge and discharge currents, (Matsushima 1990). The cumulative errors in such systems make it unreliable for use in systems over long periods of time, or over irregular and varying regimes.

**Internal impedance**: A low impedance monitoring load is briefly applied to the battery, and the current and battery terminal voltage measured to calculate the internal impedance, (Barton & Mitchell, 1989). This method is the basis of BS and IEC standard tests.

**AC bridge technique**: This is superior to the DC load technique, and is becoming more widely accepted as a means of characterising batteries. However, the equipment is expensive, and unsuitable for use in the field.

**Open circuit voltage (OCV)**: The OCV is heavily affected by the diffusion time and stratification of the electrolyte, meaning that an error in the SOC measurement of up to...
±20% is possible, even after 24 hours of rest. Whilst this technique is prone to errors if the battery is being charged or discharged at high rates, the OCV is an easy parameter to monitor.

**Specific gravity (SG):** This is also affected by the electrolyte settling time, and in vented lead acid batteries there is a well-defined relationship between SG and OCV. The normal (fully charged) SG of a lead acid cell also varies, and so equipment must be calibrated for each different cell used. Ultrasound and fibre-optic methods allow for automatic sampling, but are impractical for use with standard batteries. The SG of nickel-cadmium cells does not vary in any significant way with SOC, and the electrolyte is in any case inaccessible in VRLA batteries, and so this technique is confined to vented lead acid batteries.

### A2.6 PROPOSED SBCWG BATTERY MANAGEMENT SYSTEM

Despite the limitations of accuracy with the technique, the OCV method is the easiest to implement, and is suitable for use with both vented and valve regulated lead acid batteries. However, since in practice the charge/discharge rate of the battery in SBCWG installations is much less than that which the battery is rated for, errors from the electrolyte settling time will only be small. A temperature sensing device could be used for temperature compensation of the chosen OCV, but since it can only be placed in the vicinity of the battery bank, rather than in the electrolyte itself, it will only give an approximation to the actual electrolyte temperature.

The controller should ideally ensure that the charge current does not exceed any of the following limits:

- Terminal voltage, (temperature compensated),
- Maximum permitted charge current,
- Maximum ambient temperature,
- Minimum ambient temperature.
Using the proposed microcontroller for controlling the battery charging regime, the battery lifetime and storage capacity could be maximised by including the following features:

- Stop charging under both very high and very low temperature conditions,
- Use temperature sensing as a crude over-charge sensing mechanism as a back-up to the OCV monitoring method,
- Set a maximum charge current limit,
- Temperature compensation of the charge voltage to reduce loss of electrolyte at high temperatures, and increase storage capacity at low temperatures,
- Keep the battery continuously trickle charged to maintain it near peak condition once charged,
- Incorporate equalisation charging as a further feature,
- Entry of parameters on digital keypad makes commissioning much easier, and eliminates the drift associated with potentiometer adjusters. Experience from SBCWG users shows that field adjustment of much existing equipment is so difficult that most system commissioners do not fine adjust the nominal battery settings at all, meaning that the batteries may be used far from their designed operating range.
APPENDIX 3

THE CUK CONVERTER

A3.1 NON-ISOLATED SMPS TOPOLOGIES

A3.1.1 Overview

The power converter for the 2kW SBCWG controller must have both a step-up and step-down action. This is so that at low windspeeds where the generator voltage is less than the battery voltage, the voltage can be boosted, and at high windspeeds, where the generator voltage is much greater, the voltage may be efficiently reduced. The two non-isolated converters that are considered, the buck-boost and the Cuk converter, are both derived from the basic buck and boost converters. The equations quoted all assume ideal components, and are true for steady state conditions only.

A3.1.2 Buck Converter

This is basically a step-down chopper, with the mean output voltage proportional to the fraction of time that the transistor is on, (D). The isolated version of this circuit is the forward converter.

While the switch is on, current flows from Vd to the load. When the transistor is off, the choke voltage reverses, causing the diode to conduct, allowing the choke to transfer the stored energy to the load. The relation between the input and output voltage under steady state conditions is:

\[ V_O = V_I \cdot D \]
Fig. A3.1 Basic SMPS Topologies

**Fig. A3.1.1 The Buck Converter**

**Fig. A3.1.2 The Boost Converter**

**Fig. A3.1.3 The Buck:Boost Converter**

**Fig. A3.1.4 The Cuk Converter**
(Where \( V_i \) = input voltage and \( V_o \) = output voltage).

The choke serves to smooth the output current, and the capacitor to reduce the voltage ripple. Without these filter components the output current and voltage would be discontinuous, and the output voltage would peak at \( V_i \). The chief disadvantages of this converter are the need for a floating transistor drive circuit, the large output filter necessary, the discontinuous input current, and the large inductive voltage spike that the transistor sees. Additionally, stray input inductance will cause a higher than predicted peak voltage to appear across the transistor collector-emitter junction.

**A3.1.3 Boost Converter**

The Boost converter is only capable of boosting (increasing) the input voltage. Operation is by the transistor allowing a current to flow through the choke for time \( D \). When the transistor is off, the choke back-emf biases the diode on, allowing energy to be transferred to the load. The relation between input and output voltage under steady state conditions is:

\[
\frac{V_o}{V_i} = \frac{1}{1-D}
\]

This circuit is superior to the buck converter in that the input current is non-pulsating, operation is beneficially affected by any additional input inductance, and there is no need for an isolated transistor drive circuit. The boost circuit is widely used for active power factor correction as a pre-regulator on AC supplied SMPS circuits.
A3.1.4 Buck-Boost Converter

The Buck-Boost converter is derived from a cascaded combination of the buck and boost converters, and is able to both step-up and step-down the input voltage. As in the buck converter, the input stage consists of a transistor and inductor working in a similar way. When the transistor is off, the diode allows current to be transferred to the load, but the output polarity is now opposite to the input. The relation between the input and output voltages under steady state conditions is:

$$\frac{V_o}{V_i} = \frac{D}{(1-D)}$$

For buck operation, $D < 0.5$, and for boost operation, $D > 0.5$.

The disadvantages of this converter are that it has a pulsating input current, needs an isolated transistor drive, is sensitive to any stray input inductance, and reverses the voltage polarity.

A3.2 VARIANTS ON THE CUK CONVERTER

The operation of the Cuk converter is described in detail in 2.2.2, but for completeness the circuit and the principal variants (A 3.2.1 - 7) are also shown here. Figures A3.3.1 - 3.3.4 show the circuit and results of the Spice simulation of the Cuk converter, described in 2.2.2.

Figure A3.2.4 Basic Cuk converter
The basic Cuk converter used in this research.
Figure A3.2.5 Cascaded Cuk converter
Allows for higher input:output voltage ratios without the problem of high peak currents due to low duty cycles in the standard configuration. Works by charging the capacitors in series, and then discharging them in parallel.

Figure A3.2.3 Tapped inductor Cuk converter
Connecting the transistor or diode to a tapped point of either of the inductors allows higher transformation ratios without the penalty of extreme duty cycles. However, both the input and output currents become pulsating.

Figure A3.2.4 Bi-directional Cuk converter
Useful if both charging and discharging a battery.

Figure A3.2.5 Transformer isolated Cuk converter
A 1:1 transformer isolates the input from the output. Changing the turns ratio can reduce the need to work at extreme duty cycles.

Figure A3.2.6 Cuk converter with integrated magnetics
The similarity of current waveforms means that both chokes, and the transformer if used, can be assembled on the same core, giving an effective increase in both the input and output inductances.

Figure A3.2.7 SEPIC converter
A variation on the Cuk converter where the output inductor and diode are interposed. The output polarity of the circuit is the same as the input, but the output current is discontinuous.
Fig. A3.2 Variations of the Cuk converter

Fig. A3.2.1 The basic Cuk converter

Fig. A3.2.2 The cascaded Cuk converter

Fig. A3.2.3 The tapped inductor Cuk converter topology

Fig. A3.2.4 The bi-directional Cuk converter

Fig. A3.2.5 The transformer coupled Cuk converter

Fig. A3.2.6 Transformer coupled Cuk converter using "integrated magnetics"

Fig. A3.2.7 The Sepic converter
Fig. A3.3 Spice simulation of the Cuk converter circuit

Fig. A3.3.1 The Cuk converter Spice simulation circuit

Fig. A3.3.2 Inductor and Capacitor currents

Fig. A3.3.3 Transistor, Diode and Capacitor currents

Fig. A3.3.4 IGBT drive waveforms
APPENDIX 4
CIRCUIT DESIGN AND DESCRIPTION

A4.1 CIRCUIT DIAGRAMS & CONSTRUCTION

Figure A4.1 shows the main controller circuit diagram, and figure A4.2 the mechanical layout of the principle components. The design of the microcontroller control board is shown in figures A4.3 - 4

A4.2 POWER SUPPLY

+15V Supply
This is supplied direct from the 24V battery supply, with an LM317 linear regulator being chosen for both good ripple rejection, and for a high maximum input voltage of 40V.

On-board Switched Mode Power Supply (SMPS)
In order to simply and efficiently generate the range of other voltages required for the control electronics, an on-board 25kHz SMPS supply is used. The specification is as follows:

Input voltage: 17 - 36V (24V nominal)

Output voltages:

<table>
<thead>
<tr>
<th>Supply</th>
<th>Description</th>
<th>Output voltage</th>
<th>Max. output current</th>
</tr>
</thead>
<tbody>
<tr>
<td>Supply 1</td>
<td>Current module supply</td>
<td>-15V</td>
<td>120mA</td>
</tr>
<tr>
<td>Supply 2</td>
<td>Microcontroller supply</td>
<td>+5V</td>
<td>100mA</td>
</tr>
<tr>
<td>Supply 3</td>
<td>Floating gate supply</td>
<td>+15V</td>
<td>50mA</td>
</tr>
</tbody>
</table>
Fig A4.1 Circuit diagram of the control board
Fig. A.4.2 Mechanical layout of the control unit

Fig. A.4.3 The microcontroller board
The SMPS is controlled by the 3524 PWM control IC, which is used in the push-pull configuration to drive the primary of a high frequency transformer. This IC is powered from the linear +15V regulator, and the transformer primary current is sourced from the 24V battery supply, reducing the load on the regulated supply. All outputs are smoothed using fixed linear regulators, reducing the amount of SMPS generated voltage ripple. The transformer design is critical for efficient operation over the wide range of possible battery voltages, since the servo necessarily controls the primary pulse width based on feedback from the output voltage of one supply only, (in this circuit the negative 15V supply was chosen). The implication is that for a high battery voltage, and a low current drain on this monitored supply, the pulse width will become very narrow, which would be a problem if the other power rails were heavily loaded at this time. The transformer turns ratios are therefore calculated to ensure that the output voltages are just adequate when the input voltage is at its lowest specified level. Should the input voltage fall below this minimum operating voltage, the output voltage will also fall, which could lead to failure of the control electronics, and so the SMPS is turned off when the input voltage falls below 17V. (At this voltage, the WTG will be automatically loaded by the dump resistor.)

**Fan regulator.** The fan is specified for operation between 12V and 28V DC, and so a discrete 26V series regulator is used to limit the voltage under high battery voltage conditions.

### A4.3 OVERVOLTAGE PROTECTION

When the battery supply voltage rises above the operating voltage of the 36V zener diode, the crowbar thyristor is triggered by the zener current, momentarily drawing a high current from the battery, rupturing the controller power supply fuse. Whilst this is inconvenient to replace, it is protecting only against the very rare occasion that the user connects more than a nominal 24V battery to the 24V supply input - a condition which...
can clearly only occur when the user is present, and so replacement is not a significant problem. The electronics are protected against short-term over-voltage by the upstream RC filter on the 15V regulator. Transient spike protection is also provided by further capacitors and a Voltage Dependent Resistor (VDR) on the battery side of the fuse, and the battery itself acts as a large capacitor to remove lower frequency transients.

**A4.4 NO BATTERY SENSING**

Since the Cuk converter is working under current mode control, the Loss of Battery (LOB) condition means that the output voltage will rapidly rise in an effort to sustain the set current. This voltage can be sufficient to give the user a dangerous electric shock, and possibly damage the power converter electronics.

To protect against this, a voltage comparator monitors the voltage on the battery terminals, and when it exceeds a set voltage, generates a Cuk converter disable signal. A 22μF capacitor across the output then slowly discharges from this set peak voltage, and when the voltage has fallen to a level set by the comparator hysteresis, the LOB signal is removed. The Cuk converter is then re-started, and the cycle continues until the battery is re-connected.

**A4.5 DUMP LOAD SWITCHING CIRCUIT**

This circuit ensures that the dump load relay never has to break a current that would lead to a high DC voltage appearing across the contacts. This is achieved by ensuring that it can only switch after the dump transistor in parallel with the contacts has itself first been switched fully on.

The switching thresholds for the dump transistor and relay are set by the voltage comparators, the transistor being switched on whenever the supply voltage lies between
15V and 17V, with the relay switching at 16V, (figure 3.29). This ensures that whenever the battery voltage is so low that the controller will not properly function, the dump load resistor will be safely connected to the WTG. External control inputs allow the microcontroller to close the dump load relay at any time, in which case the interlock sequencing is performed in software.

**A4.6 DUMP LOAD CONTROL CIRCUIT**

The Dump load power is controlled by the 3524 PWM control IC in single-ended mode. The output is isolated by an opto-isolator, which is needed because the source of the power mosfet is floating at a voltage above that of the control electronics 0V. A diode OR gate on the input to the opto-isolator enables the dump transistor to be controlled from either the microcontroller or the dump transistor/relay interlock circuit.

The output of the opto-isolator is buffered and amplified by one half of the 7667 mosfet driver IC, which drives the two parallel mosfets through 22Ω resistors. The mosfet snubbing components are set by experiment to give an acceptable voltage overshoot, and also to dissipate a minimum power. The high speed schottky fly-back diode across the load gives a path for the back emf generated when the transistor is switched off.

The current demand from the microcontroller controlled Digital to Analogue Converter (DAC) is given by a voltage of between 0V and -10V, and the current feedback is from a "Heme" isolated current module, which is smoothed to give an average value of current. An external op-amp is used to generate the error to the PWM comparator in the 3524, since the transconductance op-amp within the IC was found to be unsuitable. A null-offset voltage is fed in to the summing amplifier, over-coming the small voltage offset of the 3524 PWM comparator ramp waveform.
A4.7 CUK TRANSISTOR DRIVE

The Cuk transistor drive needs to be faster than that used for the dump transistor drive, and so an integrated TLP250 opto-isolator and amplifier is used.

A4.8 HYSTERESIS CONTROLLER

This circuit controls the DC to DC Cuk power converter by the hysteresis method. The input is a rectified 3-phase waveform from the WTG, and the output is to a large battery bank.

The hysteresis control method used here controls the Cuk power transistor switching pattern such that the input current is restricted to within a current window, (fig 2.27.1). This current window is derived from the mean (negative) reference current, and additional processing sets the two boundary limits.

The Upper boundary limit is set by op-amp 1, which has an adjustable gain of 1.0 - 1.2.

The Lower boundary limit is set by op-amp 2, which has an adjustable gain of 0.8 - 1.0.

The Set comparator 3 detects when the feedback current exceeds the upper boundary limit, giving an output pulse of logic 1 to set the steering latch. (A very small amount of hysteresis and filtering is used to avoid multiple switching transitions, which must be kept to less than the set current window magnitude.) The width of this reset pulse is automatically set to be sufficient to set the latch, since the pulse will be high until the latch has changed state and the current fallen below the upper current limit.

The Reset comparator 4 detects in a similar way when the feedback current falls below the lower current limit, again giving an output pulse of logic 1.
The Output steering latch is based on a level sensing RS flip-flop. Both inputs are normally logic 0, with short logic 1 pulses from the set and reset comparators changing the latch state. When the current reaches the upper current boundary limit, the reset signal is pulsed to logic 1. The latch output falls to 0, switching off the transistor, and the current now falls until the lower boundary is again reached, when the cycle repeats itself. The latch output is amplified by a fast bipolar transistor to drive the opto-isolator.

A4.9 MICROCONTROLLER CONTROL BOARD

The microcontroller development board (figures A4.3 and A4.4), is based around the Motorola 68HC11 (HC11) microcontroller. This board incorporates both the core functions which would be used on a production version, and also other and facilities which make development easier.

A4.9.1 HC11 Core Functions

4x 4 keypad 
This is scanned and read by the HC11, with switch de-bouncing performed in software.

Liquid Crystal Display (LCD) 
This 2 x 16 Line LCD and driver contains its own microprocessor and memory, which contains comprehensive character generation and scrolling facilities. It is controlled by sending strings of ASCII characters from the HC11, with all further handling and decoding being done within the module.

Digital to Analogue Converter (DAC) 
This DAC was chosen because it contained two low speed 8-bit DACs in one package; one for the dump current demand and one for the Cuk converter current demand. The
outputs are buffered, inverted and scaled by op-amps IC15, giving an output FSD of -15V. The signals are used for setting the current demands to the dump and cuk converter current servos.

Analogue to Digital Converter (ADC)
The 10-bit, 8-channel ADC was chosen to give sufficient resolution and number of channels for the system parameters that need to be accurately measured. This IC is used in the continuous sampling mode, with results accessed as required by the HC11. (The 8-bit ADC incorporated in the HC11 is of insufficient resolution, and so is used for monitoring less critical values.)

A4.9.2 DEVELOPMENT ONLY FUNCTIONS

Port Replacement Unit (PRU)
This part (IC4) is specifically made for the HC11, which gives access to the address bus of the HC11, but without the loss of functionality of the HC11 ports that working in this mode normally entails. This allows the use of additional peripheral ICs, which would not otherwise be possible to memory map in to the system.

Address decoding logic
ICs 5,6,7,13 & 14 are used for address decoding for the peripheral ICs in the system. On a production version with far fewer peripheral ICs, it is anticipated that there will be sufficient free output lines on the HC11 to allow much simpler address decoding, and so almost all of these ICs will be unnecessary.

Battery-backed ram
This ram (IC11) allows programmes to be stored under no power conditions, but without the inconvenience of using an eprom. Once the code is finalised, it will be stored in the 12k of rom incorporated in the HC11, and user set values held in the 128-bit eeprom (electronically erasable read only memory) also within the HC11.
Fig A4.4 Circuit diagram of the microcontroller board

UNIVERSITY OF LEICESTER
DEPARTMENT OF ENGINEERING

DATE 19/5/95
SNCF 3 OF 3 VC1

TITLE MCU CONTROL FOR
2KW WIND TURBINE

RS232
System reset
This IC (IC18) generates a controlled system reset signal, but in production units will be replaced by a cheaper method.

Real time clock/calendar
This IC (IC10) generates both date and time information, which could be useful for maintenance and fault logging requirements.

RS-232 serial communications IC
This IC (IC3) supports the serial link between the HC11 and the PC. It might be required in production units to allow system information to be downloaded to a portable PC for maintenance and analysis, and possibly for loading software changes.
APPENDIX 5

MICROCONTROLLER CONTROL SYSTEM

A5.1 DESCRIPTION OF MICROCONTROLLER ROUTINES

Total generator load power
The look-up table is addressed by the shaft speed \( (\omega) \), to obtain the look-up table power, \( P_{tb} \).

To this may be added the instantaneous value of the hill-climbing algorithm, \( P_{hc} \).
In this case the total generator load power, \( P_{dt} \) is the sum of the look-up table power and the hill climbing algorithm power, \( P_{hc} \):

\[ P_{dt} = P_{tb} + P_{hc} \]

Note that the value of \( P_{hc} \) may be negative, and so \( P_{dt} \) can actually be less than \( P_{dt} \).

Battery/dump load power split
The total generator load power demand above is split between the Cuk converter, \( P_{cd} \) and the dump load, \( P_{dd} \).

\[ P_{cd} + P_{dd} = P_{dt} \]

The split is determined principally by the maximum allowable battery voltage, such that all of the generator power is used for charging the battery \( P_{cd} \) until the battery charge voltage limit is reached, at which point the power is partially diverted to the dump load.
(Pdd), with the split adjusted to maintain the set maximum battery voltage. This maximum battery voltage is temperature compensated by the following equation relating the Maximum Battery Voltage ($V_{bmx}$), to the maximum battery voltage at 25°C ($V_{25}$), the battery temperature in °C ($T_b$), and the temperature compensation constant, ($k_c$):

$$V_{bmx} = V_{25} - k_c.(T_b - 25)$$

Note that the battery charge current is controlled by altering the value of the Cuk converter (input) demand power ($P_{cd}$), and so the actual battery charge current does not appear in the control system. The Cuk converter power demand may also be limited by conditions for maximum charge rate, maximum and minimum battery voltage, and maximum and minimum battery temperature.

Both battery and dump load power demands are converted into current demands by measurement of the DC bus voltage:

$$I_{cd} = \frac{P_{cd}}{V_{bus}}$$
and
$$I_{dr} = \frac{P_{dr}}{V_{bus}}$$

(Where $I_{cd} = $ Cuk converter input current demand, and $I_{dr} = $ dump load circuit current demand.)

### 5.2 USE OF THE HILL-CLIMBING ALGORITHM

Section 2.1.3 highlights some of the reasons why the actual power available from the WTG is likely to deviate from the ideal. These variations mean that the use of a hill climbing algorithm to automatically alter the current demand to seek this optimum is desirable. This technique works by slightly increasing or decreasing the current demand, and then observing the change in actual current. If for instance the WTG is slightly underloaded, then it will support a small increase in current demand, with only a small
decrease in speed. The current can be slowly increased until it is past the actual ideal (a in figure A5.1) - past which point the WTG will be overloaded and start to slow down (to b). This decrease in current is detected by the hill climbing control, and the current will then be decreased in a similar way (back to a).

For best system response it is suggested that the look-up table is still used for approximate current demand calculation, with the hill climbing just for fine adjustment. (This self-adjusting system that allows for variations in the system performance with time should be contrasted with the system discussed by Hadjimichael, 1992. This uses hill-climbing as a method of generating a set of stored "ideal" curves in a single "learning" session, but which crucially does not adapt to later changes in system.)

5.3 THE POWER SPLIT CALCULATION

The power split calculation ensures that the sum of the Cuk converter and dump load circuit power demand equals the total available from the WTG at that speed. If an accurate speed sensing signal is used, any small variation in generator volts/rpm can be overcome by using the look-up table to give a power demand, and then dividing this value by the measured generator voltage to give a compensated current demand. However, in this much simpler system where the generator voltage is used for indirectly sensing the shaft speed through DC link voltage measurement, this small error cannot be compensated for, and so the calculation could be in terms of current, not power.

5.4 USER INTERFACE INSTRUCTIONS

The user interface consists of the 4 x 4 keypad and alphanumeric display, which allows the viewing of all important system parameters and limits.
Increasing demand

Shaft speed

Maximum power point

Fig A5.1 Hill climbing - the effect of demanding too much current

<table>
<thead>
<tr>
<th>Location</th>
<th>Parameter</th>
<th>FSD value</th>
<th>Resolution</th>
</tr>
</thead>
<tbody>
<tr>
<td>8-bit DAC</td>
<td>Dump load demand current</td>
<td>10 Amps</td>
<td>39mA</td>
</tr>
<tr>
<td></td>
<td>Cuk demand current</td>
<td>10 Amps</td>
<td>39mA</td>
</tr>
<tr>
<td>10-bit ADC</td>
<td>2v5 reference</td>
<td></td>
<td></td>
</tr>
<tr>
<td>A10</td>
<td>Cuk converter current</td>
<td>12 Amps</td>
<td>12mA</td>
</tr>
<tr>
<td>A11</td>
<td>Dump load current</td>
<td>12 Amps</td>
<td>12mA</td>
</tr>
<tr>
<td>A12</td>
<td>Main battery voltage</td>
<td>210 Volts</td>
<td>205mV</td>
</tr>
<tr>
<td>A13</td>
<td>24v battery voltage</td>
<td>40 Volts</td>
<td>39mV</td>
</tr>
<tr>
<td>A14</td>
<td>Wind generator voltage</td>
<td>320 Volts</td>
<td>312mV</td>
</tr>
<tr>
<td>A15</td>
<td>Shaft speed</td>
<td>rpm</td>
<td>rpm</td>
</tr>
<tr>
<td>A16</td>
<td>Battery temperature</td>
<td>400 K</td>
<td>0.39 K</td>
</tr>
<tr>
<td>8-bit ADC</td>
<td>Thermostat</td>
<td>530V</td>
<td>2V</td>
</tr>
<tr>
<td>PE1</td>
<td>Phase 1 monitor</td>
<td>530V</td>
<td>2V</td>
</tr>
<tr>
<td>PE2</td>
<td>Phase 2 monitor</td>
<td>530V</td>
<td>2V</td>
</tr>
<tr>
<td>PE3</td>
<td>Phase 3 monitor</td>
<td>530V</td>
<td>2V</td>
</tr>
<tr>
<td>PE4</td>
<td>Dump load monitor</td>
<td>320V</td>
<td>1.25V</td>
</tr>
<tr>
<td>PE5</td>
<td>Battery current</td>
<td>90 Amps</td>
<td>3.6 amps</td>
</tr>
</tbody>
</table>

Fig. A5.2 Microprocessor analogue I/O signals
The menu is accessed using the scroll keys ← and →, with J being used as an enter key. CLR is used to escape to the previous level.

To display or enter values:
Scroll through the main menu until the desired set menu is reached. Press J.
Scroll through the sub menu until the desired parameter is reached. Press J.
Enter the new value on the keypad. Press J.
Press CLR at any time to return to the main menu.

In addition to the menu structure shown below, three further menus allow the entry and display of parameters shown in figure 5.2.

The Scaling Constants are for setting the ratio of actual measured voltages to the “real-world” values of current, voltage and speed which they represent.

- **Scaling Constants: 10-bit ADC**
  - F1c: Cuk converter input current
  - F1d: Dump current
  - FVb: Main battery voltage
  - FVab: Auxiliary (24V) battery voltage
  - FVbus: DC bus voltage
  - FWss: Shaft speed
  - FTbk: Battery temperature

- **Scaling Constants: 8-bit ADC**
  - FPh1: Phase 1 voltage
  - FPh2: Phase 2 voltage
  - FPh3: Phase 3 voltage
  - Flb: Battery charge current
Scaling Constants: DACs

Ficd  Cuk converter servo demand current
Fidd  Dump load servo demand current

Limits

Wmx    rpm  Maximum shaft speed
Wamx   rpm  Absolute maximum shaft speed
Pgmx   Watts Maximum generator power
Vbmx   Volts Maximum battery voltage
Ibmx   Amps Maximum battery charge current

System variables

Enable HCA  Flag  Use the hill climbing algorithm
Wc       rpm  Cut-in shaft speed
Ws       rpm  Shaft speed at end of initial ramp
Wp       rpm  Shaft speed at maximum power
PHmx     %    Maximum percentage deviation due to
          the HCA
Hstep    Amps HCA step size

PC interface

Run System Uploads system parameters to PC terminal

A5.5 MICROCONTROLLER USER INSTRUCTIONS

Stand-alone mode

This mode allows the system to be run without the PC terminal.

- Ensure mode switch is in "Non-Buffalo".
- Apply power, and press Reset (Sw.2)
**PC mode**

This mode allows the PC terminal to be used to alter the programme code. (User set values can be altered and retained in the stand-alone mode).

```bash
c\ cd hcl1\source .
pth .
a .
header.S19 files.
t .
load t .
F2
g 4004
```

Sets DOS path
Batch file assembles all code files, generating `header.S19` files.
Tells HCl1 to be in terminal mode.
Ensure mode switch is in "Buffalo" mode.
Press Reset on HCl1 board.
Sends `header.S19` file to HCl1
Starts programme at memory location 4004.

The data on the VDU is updated whenever any keypad key is pressed in the "PC Interface" mode.
APPENDIX 6

WIND GENERATOR TEST RIG HARDWARE

A6.1 DC MOTOR DRIVE

The DC motor drive unit (fig A6.1), is based around an "SSD" 512/32 9kW DC motor drive module. It is wired in the closed-loop speed control mode, with key system variables set by on-board potentiometers, (system stability, speed ramp rates, speed limits, current limit, and IR compensation). Field current is from an uncontrolled bridge rectifier, and armature current controlled by a half-controlled bridge rectifier, also sourced from the 2-phase 415V supply. The thyristors are protected by an external 32A 1,000V high speed semiconductor fuse. Speed feedback is from the DC tacho-generator mounted on the motor shaft.

Stop/Start control is by the two push-buttons "Start" and "Stop" mounted on the front panel, which for safety controls a 110V AC relay supplied from a 55-0-55V step-down transformer. One of the relay contacts shorts the "Start" button contacts after release, only being released when the Normally-Closed "Stop" button is pressed, breaking the supply to the relay coil. The other contact is used to enable the motor "Run" interlock circuit on the DC drive.

The speed demand is set by the double-pole toggle switch to either "Local" or "Remote". "Local" uses the 10-turn panel-mounted potentiometer for speed demand, and "Remote" uses an external voltage signal connected to the 9-pin socket. This may be sourced from either a remote 10k potentiometer, or any voltage signal of 0-10V DC.
Fig. A6.1 Wiring diagram of the motor drive for the Wind Generator Test Rig

POWER CUBICLE

FIELD SUPPLY

ARMATURE SUPPLY

32A-1000V HIGH SPEED

DC MOTOR DRIVE UNIT
RS 320-174 512/32

LOCAL SPEED DEMAND

SPEED DEMAND SELECT

REMOTE SPEED DEMAND

DC MOTOR

FELD TACHO-GEIGERCR

SHUNT HOUND
DC MOTOR
480V, 3000 rpm max
360V Excitation
250A, 22A Armature Cont. pos.

TACHO-GENERATOR
120V/300V/500V 22A Field

MAIN SUPPLY
The tacho signal is scaled on a small circuit board inside the motor drive cubicle to a maximum of 10V to suit the ADC card in the PC, (fig 6.2.1).

A6.2 INPUT/OUTPUT BOARD

This board (fig A6.2.2) interfaces the analogue conversion card to the motor drive and torque transducer amplifier. It consists of current and voltage limiting networks for both the torque level and shaft speed demand. Pin connections are shown in figure A6.2.3

A6.3 WGTR HARDWARE SPECIFICATIONS

- **Torque transducer**
  - EEL slip-ring type
  - 0 - 60 lb.ft
  - 2.172 mV/ lb.ft sensitivity
  - Connections: Bridge -, black, pin 1
  - Bridge +, red, pin 5
  - Transducer +, green, pin 3
  - Transducer -, yellow, pin 2

- **Torque transducer amplifier unit**
  - CLL meter type SGA1102
  - Amplifier designed and built at Leicester University
  - Excitation adjustable to 10V, (pin connections in fig A6.2.4)

- **Computer**
  - 486-Based IBM compatible
Fig. A6.2 Wind Generator Test Rig hardware

Fig. A6.2.1 Tacho output board

Fig. A6.2.2 Signal limiting board

Fig. A6.2.3 Drive - I/O box connector:
Pin allocation

Fig. A6.2.4 Torque transducer amplifier connections

Function Sixway 25Way
Rx 2 2
Tx 3 3
GND 5 7

1 +12V
11 A/D Ch.0 (Torque)
12 Ground
14 -12V
15 D/A Ch.1 (Dev Demand)
23 A/D Ch.2 (Tacho)
24 Ground

Blu 1 Bridge 'Neg.'
Yel 2 Transducer 'Neg.'
Grn 3 Transducer 'Pos.'
Red 4 Bridge 'Pos.'
5 Torque Out
6 0V
7 0V
8 0V
9 0V
10 0V
RTU Super 12-Bit converter card (2 x 12-Bit DAC, and 8 x 12-Bit ADC), now being replaced by an "in-house" converter card.

- **Tacho-generator**
  - Radio-Energie
  - 0.06V/rev/minute
  - 5000rpm maximum
  - 0.1A maximum current

- **Motor**
  - Mawdsley shunt wound DC motor
  - 440V 1/1060/1460/3000 rpm
  - 360V excitation
  - 8.5kW, 22A armature, continuous. (assumes fan cooling)

- **DC motor drive**
  - RS 320-764 32A d.c. drive (SSD)
  - Wired for 415V 2φ 50Hz supply.
  - Switch settings
    | Setting | Description |
    |---------|-------------|
    | 1 On, 2 On | (25 - 75V tacho) |
    | 3 Off | (Tacho) |
    | 4 Off | (Run control enable) |
    | 5 Off, 6 On, 7 Off | (12A max arm. current) |
    | 8 Off | (Ammeter scaling) |
Circuit function
The electrical schematic and outline description of the operation of this new controller is included in figure 1.9.1-3 and section 1.5 of chapter 1. This appendix gives further detail on the operation of the electronic control board.

Battery charge / Dump load control
The battery voltage is regulated by connecting the WTG to a dump load when the maximum safe battery voltage is reached. When the battery voltage has fallen to a lower voltage, the WTG is once again connected for charging the batteries. The battery voltage is sensed by IC1a, and triggers a 30-second pulse which is timed by the 555 monostable. This time delay is used to prevent the controller rapidly cycling between the "dump" and "charge" states if the battery is in very poor condition. IC1b acts as an OR gate to control the charge/dump signal to the charge/dump relay control IC, IC2a.

Autotransformer thermostatic protection
The WTG will also be loaded by the dump load if the autotransformer should become too hot, detected by a thermostat connected to Con3.

Autotransformer tap control
The WTG voltage is approximately matched to the battery voltage by a two step autotransformer. At low windspeeds, the lower tap is used, which boosts the voltage, and at higher windspeeds the upper tap is used, reducing the voltage. Tap changing is
controlled by IC3a, which changes to the high windspeed tap above a set WTG current, using a current transformer for sensing.

Battery disconnection
If the battery is disconnected, the rectifier output voltage rapidly rises, and could damage the electronics. IC2b therefore senses this abnormally high voltage, and switches the charge/dump relay to the (de-energised) dump position. This action not only safely loads the WTG, but also disconnects the electronics from the WTG, since it is being powered by neither the battery or the rectified WTG voltage. (In this condition, where the circuit is unpowered, the windspeed tap-changing relay will be in the de-energised or high-windspeed position, ensuring that WTG will not fail through over-current.) The circuit remains in this state until the battery is re-connected.

Operation with different battery voltages
The circuit is designed to operate with either a 12V or 24V battery supply, (with just a few changes of component values - shown as alternatives on the circuit diagram). However, if a higher battery voltage is required, then the battery voltage sense should be connected to the additional battery sense terminal, and the actual circuit powered from a 12V or 24V battery tap.
Fig. A7.1 Circuit Diagram for the new 500W (FM1800) SBCWG control board
REFERENCES


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