Control and Instrumentation for an Electric Farm Tractor

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CONTROL AND INSTRUMENTATION
FOR AN ELECTRIC FARM TRÁCTOR

BY
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Philippians 4:13
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Chapter I
INTRODUCTION

1.1 Project Origin

During the late 1970's the sharp rise in petroleum costs produced a flurry of activity to develop alternate energy sources. A primary goal was to produce efficient, cost-effective electric vehicles. Until recently no consideration had been given to producing electric farm vehicles. However, a shortage of diesel fuel for farmers in the early 1980's prompted a USDA-DOE grant to study the feasibility of electric farm tractors.

Research in this area has shown that a significant amount of farm work can be performed by electric vehicles.\textsuperscript{1,2,3} Tasks which are particularly suitable have characteristics of being regular routines performed on or near the farmstead which require stop-and-go operation. These tasks are particularly inefficient for internal combustion vehicles since they require only short, intermittent bursts of power.

Predictions of increasing petroleum prices make electric powered vehicles more attractive for farm use
since electricity can be generated by several sources. Other advantages include lower energy operating costs, less maintenance, longer vehicle life, quieter operation, easier starting, and elimination of noxious gases. Possibilities also exist for the electric farm tractor to serve as an auxiliary power source.

To further investigate the feasibility of electric farm vehicles, a project was undertaken to develop a battery-powered chore tractor. The philosophy of this project was that technology currently available would allow the use of standard components to assemble the vehicle. In the interest of time, a decision was made to build a tractor suitable for a majority of farm chores and to modify it accordingly as deficiencies appeared during testing.

A Versatile 160 tractor was chosen as the basis from which to start. This is a four-wheel drive tractor with an articulated frame. It is powered by an 85 hp diesel engine through a three-speed hydrostatic transmission. The unique design of this tractor provided a frame which could be easily modified to suit the particular needs of an electric tractor.

The result of this project, the Electric Choremaster, is shown in Fig. 1.1. Two electric motors
Fig. 1.1 Side view of Electric Choremaster.

Fig. 1.2 Electric Choremaster control panel.
are used: one to provide tractive power and the second to drive the Power Take Off (PTO) and hydraulic systems. The PTO/hydraulic (PTO/HYD) motor is located beneath the cab and the traction motor is located behind and beneath the batteries which are behind the cab. Table 1-1 contains a list of component specifications.

Table 1-1 Electric Choremaster Component Specifications

1. Overall Vehicle
   Weight: 12,000 lbs.
   Length: 150 in.
   Width: 80 in.
   Height: 109 in.
   Clearance: 16 in.
   Speed Range: 0-20 mph forward or reverse
   Transmission: Three range gearbox combined with a solid state controller.
   Braking: Electronic with a mechanical park
   Steering: Hydraulic with electronic actuation
   Tire Sizes: 13.6 x 24 in.

2. Traction Motor
   Power: 50 hp at 1500 rpm (1 hour rating)
   Size: 13 in. diameter, 475 lb.

3. PTO/Hydraulics Motor
   Hydraulics Power: 25 hp at 2500 rpm (30 min. rating)
   PTO Power: 22 hp at 625 rpm for 30 min. (same motor run through a 4/1 gear reducer with 10% efficiency loss)
   Size: 11 in. diameter
4. Battery

Capacity: 128 V, 340 A-h for 6 hour
discharge
Cell number: 64 cells in series, 2 volts
nominal
Type: Lead-acid
Cell size: 3 1/2 in. x 6 1/4 in. x 22 in.
Weight: 4000 lb. total

It is the purpose of this thesis to describe the control systems of the tractor and its instrumentation. A major portion of the control system is shown in Fig. 1.2. After the introductory material and literature review in Chapter I, Chapter II describes methods of motor control. Chapter III discusses auxiliary and instrumentation systems required to support and protect the drive circuitry as well as provide an adequate interface for the operator. An improved speed control system for the PTO/Hydraulic motor is analyzed in Chapter IV. Concluding remarks are contained in Chapter V.

1.2 Literature Review

There is a broad background of literature available on the topic of DC motor speed control. This section contains a representative sampling of current techniques and concludes with a discussion of electric
powered farm vehicles which have already been developed.

Joos and Barton have provided an overview of design considerations for DC motor drives. They note that in all but a few instances solid state controllers have taken over the task of motor control and allow DC motors to be driven from either DC or AC power sources. The basic control scheme normally involves two parallel loops for speed and current regulation. Normally the current loop remains inoperative until a maximum value of current is reached. At this point current control takes over to maintain a safe level. Generally speaking, an exact model of a DC machine is quite complex. Fortunately for most of the conditions of interest, a simplified model exists which does not take into account phenomena such as armature reaction, commutation, eddy currents, and brush drops, but still maintains acceptable accuracy. Another topic addressed is speed measurement. Although speed may be calculated from the armature terminal voltage, most high performance systems use tachometer generators. These signals are subject to noise and require adequate filtering. Use of proportional-plus-integral controllers can substantially reduce the effects of ripple and noise. This type of control is therefore
preferred over straight proportional control. Proper filtering and isolation of control signals are important design considerations.

Franklin developed the foundation for DC motor control by pulses. Equations are developed which predict current and torque as a function of motor speed. Included are discussions of pulse width control and current limit control. Several assumptions and approximations are made prior to the equation development. Speed is assumed to be constant during a given pulse condition. This is reasonable since chopping periods are generally quite short compared to motor time constants. Inductance is held constant even though inductance decreases somewhat with increasing current due to flux saturation. The nonlinear magnetization curve is approximated by a straight line. It is shown that the error introduced is small when compared to using step-by-step integration to find flux. Examples of current and torque vs. speed for both the shunt and series motors are given.

A continuation of the analysis of chopper fed DC motors has been provided by Damle and Dubey. Their approach divides motor performance equations into three intervals: duty, main thyristor is on;
commutation, commutation thyristor on and main thyristor off; and free-wheeling, both thyristors off. The equations are solved by numerical integration using fourth-order Runge-Kutta and modified Euler-Predictor-Corrector methods. An alternative approach is given using piecewise linear approximation. In contrast with Franklin, the nonlinear magnetization curve is approximated using linear interpolation between a set of coordinate points. By comparing theoretical predictions with experimental results for a DC series motor, the approach is verified as accurate. A digital computer program has also been written that can be used in many applications. It uses a state variable equation approach with suitable methods for reducing computation time.

Dubey and Shepherd predict the performance of a chopper-fed DC series motor using transient analysis. Instead of dividing the chopping cycle into intervals, a continuous approach is used to obtain average values rather than instantaneous values. Three different methods are presented to obtain the current and torque response to perturbations. As always, the magnetization curve presents a problem, but three approximations are given to circumvent this. A transfer
function can also be derived using these methods which can be used for stability studies in closed-loop systems. Both first and second order continuous models are derived with the second order model being more accurate. Experimental results are given to prove the accuracy of the methods which can be applied to both pulse rate control as well as current limit control systems.

A rather unique application of DC motor drives to electric road vehicles has been presented by Thompson. After acknowledging the superior torque characteristics at low speeds of chopped series motors, it was shown that at higher speeds field control can provide maximum power with high efficiency. By adding a transmission this advantage can be more fully utilized. Regenerative braking was also incorporated because of the increased safety and efficiency it can provide at road speeds. In order to operate the vehicle with field control only, an automatic transmission with a high stall torque converter was used in conjunction with a battery switching scheme. By running the motor at half battery voltage, the base speed of the motor is reduced to an acceptable level. This is the lowest speed the motor can run since field control is used. Increasing
speed couples the motor to the transmission through the torque converter. As soon as the vehicle has reached a minimum speed the controller automatically switches the battery to high voltage. This system was compared to a conventional armature chopping system used previously in the same vehicle with improvements noted in acceleration, maximum speed, and range. No comments were given, however, on the effect battery switching made on charging and battery life.

Bose and Steigerwald have presented work detailing a separately—excited DC motor drive using a thyristor armature chopper and a transistor field chopper. The system operates in three modes. Mode I is equivalent to series motor operation and is used for initial high acceleration. Both the armature and field current are chopped in phase to obtain maximum developed torque. Mode II is similar to shunt motor operation and is used below base speed of the motor. In this mode field current is maintained at rated value and armature current is controlled by the chopper. Mode III uses field weakening to increase motor speed. The armature chopper is bypassed and armature current is regulated by controlling field current. Both armature current and field current are subject to a temperature
overriding control. A stability analysis was performed on a hybrid computer.

A separately-excited motor drive system presented by Sen and MacDonald provides a good discussion of closed loop control. Using a separately-excited motor allows linear transfer functions for the motor to be developed. Two closed loops are used: an outer loop for speed control with an inner loop for current control. Additionally two types of controllers are discussed: proportional (P-type) and proportional plus integral (PI-type). It was found that excessive noise and ripple on the speed feedback signal necessitated use of a filter for satisfactory performance. If PI-type control was used a filter was not required. Using this type of control also reduced the steady-state error inherent in P-type control. This was graphically illustrated for torque disturbances. Current limiting was especially important during start-up. The normally large starting currents are eliminated thus protecting the controller thyristors.

This paper also gave experimental results showing the low order linear model is valid.

A second paper by Steigerwald describes the use of a fully transistorized controller. This system
uses a separately-excited DC motor and is also logically controlled by a microprocessor. During armature control the field current is maintained at a constant value, while in field control the armature chopper is bypassed. Four Darlington transistor pairs provide up to 400A for motoring and one pair is used to supply 200A of generating current during regenerative braking. Snubber circuits are essential to relieve turn-off and turn-on stress for the power transistors. In addition to this, an inductor is placed in series with the power transistors to limit inrush current. Also zener diode clamps are provided to limit voltage transients. Obviously transistor choppers require more overload protection than thyristor choppers. Significant interface electronics are needed to connect the microprocessor to the controller. A variety of armature chopper malfunction signals are capable of directly shutting down the controller without going through the microprocessor. Base drive to the power transistors, which is controlled by the processor, must provide -4 volts to insure transistor turn-off. A unique current control scheme implemented by discrete logic insures that transistors are gated only at the proper instant. A peak switching frequency of 2000 Hz is considerably
higher than thyristor choppers (400Hz).

Digital control using a finite difference equation algorithm is discussed by Jing-Ping and Marleau. The characteristic system differential equations are converted to discrete form using the $z$-transform. An algorithm for microprocessor control is then developed to produce a deadbeat response for a step, ramp, or parabolic input. It is interesting to note that the system does not exhibit good response for an input other than that for which it was designed. For cases in which the load torque can be measured directly, the digital computer can provide feed-forward action to decrease speed drop-off when the motors are subjected to load disturbances. This is most effective at high sampling rates. Often, however, load torque cannot be measured directly. Rather than feeding back only shaft velocity, as is commonly done, a combination of shaft speed, its derivative, and motor current was fed back to approximate the feedforward signal. It was found that this did not provide any improvement at the sampling interval investigated.

Plant, Jorna, and Chan present a method of microprocessor control for a DC motor using an AC supply. Use of an AC supply complicates the
controller since the SCRs may not be fired without considering the supply phase. System equations are handled using the state-space approach to minimize a cost function. Once the processor has minimized the cost function, a look-up table is consulted and the appropriate firing angle is chosen based on armature current and shaft speed. Considerable discussion of an appropriate choice for system eigenvalues to insure stability is included. It is suggested that this method has a faster algorithm since only four multiplications by a constant, five additions, and one logical decision are required.

A final topic which needs to be included in this section is a review of past electric farm vehicle development. Information in this area is sparse.

An electric tractor using fuel cells as the energy source has been developed by Allis Chalmers. Over 1000 fuel cells are connected in a series-parallel combination to produce 60 volts. A 20 hp electric motor operates through a resistance controller in a transmissionless drive. The tractor could develop 3000 ft-lb at the drawbar and easily pull a two-bottom plow.

Many electric garden tractors have been
developed. One such model, produced by the Farm Electrification Council and Lead Industries Association, had a 12 hp capability and was powered by six, 6 volt lead-acid batteries.\(^2\) Two 1 hp series-wound electric motors drive the vehicle, three 1.25 hp permanent magnet motors drive a mower attachment, and one 4 hp permanent magnet motor drives a snowblower attachment. A solid-state control was used. The tractor could operate about two hours and mow about 53,000 sq.ft. per battery charge.

A review of the available literature has shown an abundance of information on DC motor drive systems. Included are drives for series, armature controlled, field controlled, and separately-excited motors; discussions of SCR choppers, transistor choppers, and combinations of both; control and stability analysis of systems using transient analysis, P-type and PI-type feed back; and finally discrete microprocessor controlled systems. Concluding remarks have indicated that applications of this technology directed toward farm vehicles is practically nonexistent. It is hoped that this paper will provide a beginning in this area.
1.3 Drive System Components

In any battery powered electric vehicle three major components make up the drive system: the motor, controller and battery. These components are inter-related and one cannot be specified without giving consideration to the other two.

Drive system design begins with motor specification. Its purpose is to change the electrical energy in the battery to useful mechanical energy as efficiently as possible. Once the motor has been sized for power output and speed, motor type must be considered. AC induction motors are generally smaller and lighter than DC motors. They can also operate without brushes. However, they must have an inverter to convert battery current to alternating current. Series-wound DC motors can provide large amounts of torque at low speeds and for starting but have poor speed regulation with varying loads. Conversely, a shunt-wound DC motor provides better speed regulation but lower starting torque. A compound-wound motor lies between the two with corresponding advantages and disadvantages. Finally, the separately excited DC motor can be made to perform as either a shunt-wound or
series-wound motor. Its drawback lies in increased controller complexity. Because the Electric Choremaster is a low-speed vehicle requiring substantial torque, series-wound motors were selected. Series-wound motors are used almost universally in heavy-duty applications such as airport tow trucks and coal mining scoops.

Controllers must regulate the flow of energy from the battery to the motor such that the operator can easily control the speed of the motor. In addition they provide means of protecting battery, motor, and controller from electrical and thermal overloads. Controllers are built to operate with a specific type of motor, generally either AC induction, series-wound DC or separately excited DC. Shunt-wound motors see only very limited use in electric vehicles. Motor controllers are discussed in detail in Chapter II.

Batteries are the energy source, or fuel tank, for an electric vehicle. Batteries are specified according to voltage, and energy capacity in terms ampere-hours. Typically, lead-acid cells are used which nominally provide 2 volts each and are generally connected in series to provide the system voltage. These batteries are specially constructed to operate most efficiently when they are deep discharged before
being recharged. Batteries are the weak link in electric vehicles because of their comparatively low energy density (.086 MJ/Kg vs 43 MJ/Kg for diesel fuel). This puts a severe restriction on the amount of work which can be accomplished between charges and is the primary reason electric tractors are suitable only for chore work. Battery research indicates a near term efficiency improvement of 50%. Alternative battery cells, such as nickel-zinc, could provide a 100% increase in energy density within ten years.
Chapter II
MOTOR CONTROLLERS

2.1 The Silicon Controlled Rectifier

The advent of power semiconductors has revolutionized the world of motor control for battery-powered vehicles. One device in particular, the thyristor or silicon controlled rectifier (SCR), has found almost universal acceptance for controlling large amounts of current in DC motor drives. The thyristor is a three-terminal, three-junction, four-layer semiconductor of alternating p and n silicon layers whose structure and circuit symbol are shown in Fig. 2.1. Ideally, the thyristor acts as a switch. In the "off" state it offers a very high impedance between the anode and the cathode such that the device is essentially an open circuit. In the "on" state there is very low impedance between the anode and cathode such that the device is essentially a short circuit. Practically, however, when the device is "on" there is about a 1.5 volt drop from anode to cathode.

To understand thyristor operation the two-transistor model is shown in Fig. 2.2. The pnpn...
Fig. 2.1. Thyristor structure and circuit symbol.

Fig. 2.2. Two transistor model for a thyristor.
wafer can be considered as an npn and pnp transistor back to back sharing the two inside layers. Applying a positive input to the gate will forward bias the npn transistor turning it on. The collector of the npn transistor then provides the base drive for the pnp transistor and it turns on assuming the anode is positive with respect to the cathode. Both transistors saturate and the current is limited only by the external load.

Thyristor characteristics must be analyzed for three operating conditions. When the thyristor is reverse-biased (cathode positive with respect to anode), the outside junctions are reverse-biased, and no current flows. Application of a positive gate signal now will cause only a small anode leakage current to flow. Secondly, applying a forward bias to the anode and cathode will cause the two outer junctions to be forward-biased. However, the middle junction will be reverse-biased and no appreciable current will flow. The third operating condition, forward-biased and conducting, can be achieved by turning the thyristor on with any of four methods:

1. Light turn-on
2. Gate turn-on
3. Breakover voltage turn-on
4. dv/dt turn-on

All semiconductors are light sensitive and by directing a beam of light on the gate-cathode junction the thyristor can be turned on. This method is not applicable to electric vehicle motor control and will not be discussed further.

If a positive signal is applied to the gate while the anode is positive, both with respect to the cathode, the thyristor will turn on. The gate signal must supply sufficient current (up to 250 mA depending on thyristor size) over a sufficient period of time (a few microseconds). It takes a small amount of time for the thyristor to reach a conducting state and if the gate signal falls to zero before the latching current is reached, the thyristor will not remain in a conducting state. Power dissipation is significant in the thyristor during the turn-on period since there is still an appreciable voltage across the device as the current is rising.

Increasing the anode-cathode forward voltage causes the depletion region at the middle junction to widen. At the same time, the accelerating voltage for
minority carriers is increased causing these carriers to collide with fixed atoms releasing more minority carriers. Finally, avalanche break down occurs and the middle junction conducts. Now the anode current is limited only by the external load. The potential at which this occurs is called the breakover voltage (VBo). While this method is not destructive to the device, it is normally used only to turn on four layer diodes.

If the anode to cathode voltage increases rapidly (high dv/dt) transient currents can flow caused by the anode-to-gate and gate-to-cathode capacitances. This can cause the device to turn on but should be avoided.

Once the thyristor has been turned on, a way must be found to turn it off since removal of the gate signal alone will not accomplish this. By "off" it is meant that all conduction stops and reapplication of forward bias to the anode and cathode will not cause conduction. This turning off process is called commutation. There are three ways to commutate a thyristor:

1. Natural commutation
2. Forced commutation
3. Gate turn-off

When the current through a thyristor falls below a minimum value called the holding current the thyristor will turn off. To achieve this in a DC system a line switch must be opened, the thyristor by-passed, or the load impedance increased to reduce the current to less than the holding value.

Forced commutation involves reverse biasing the thyristor for a long enough period to cause the outer junctions in the device to become reverse biased and turn off the device. External circuitry is needed for forced commutation.

Specially built thyristors are available that can be turned off by applying a negative voltage to the base. This causes the effective holding current to increase until it is greater than the load current and the thyristor stops conducting. Only low power thyristors are available which can be turned off in this manner.

When using thyristors care should be taken to avoid exceeding maximum voltage, current, and power ratings. The peak forward voltage (PFV) is the limiting anode voltage above which the thyristor may be damaged.
Fortunately, the forward breakover voltage (VBo) is usually less than this so there is some inherent protection for the device. To increase the voltage rating thyristors may be built with greater wafer thickness. To do this, however, sacrifices fast turn-on times and increases forward voltage drops. To increase current handling capability the thyristor should be built with a large active crystal area and small thickness. Doing this compromises high voltage handling characteristics. Thus, a typical engineering trade-off faces the thyristor designer.

Thyristor power ratings can be divided into a number of power losses. The load current forward conduction loss is the current through the device multiplied by the voltage across the device. This is the major loss and large thyristors may be encased in massive heat sinks to dissipate the heat. A forward leakage power loss occurs when the thyristor is forward biased and nonconducting. As before, the power dissipated is equal to the voltage multiplied by the leakage current. During turn off high values of reverse current can flow. Since the voltage is increasing also at this time an appreciable loss occurs. Conversely, during turn-on an appreciable current can flow before
the voltage has dropped to its "on" value and a somewhat larger power loss can occur. Finally, there is a loss in the gate drive if the gate voltage is held positive instead of pulsed.

As a final note, thyristors are available with voltage, current, and power ratings which far exceed any requirement for an electric tractor. However, when very high voltages are required several thyristors can be placed in series to share the voltage drop. Also, several thyristors may be placed in parallel to handle extremely large amounts of current. When these techniques are employed additional protection circuitry is required.

2.2 Chopping Methods.

One way to control the voltage applied to a DC motor, known as chopping, is illustrated in Fig. 2.3a. When the switch, SW1, is opened and closed repetitively a voltage waveform as in Fig. 2.3b appears at the motor. The average voltage applied at the motor will be

\[ V_m = V_B \left[ \frac{t_{on}}{t_{on} + t_{off}} \right]. \]
Fig. 2.3. Illustration of PWM and PFM.
By varying $t_{on}$ and $t_{off}$ any voltage from 0 to $V_B$ may be applied to the motor. If $t_{on}$ is varied while the sum $t_{on} + t_{off}$ is held constant, a process known as pulse width modulation (PWM) is used to vary $V_m$. This is shown in Fig. 2.3c. Conversely, if $t_{on}$ is fixed and $t_{off}$ is varied the process is called pulse frequency modulation (PFM) and this is shown in Fig. 2.3d. Motor controllers generally apply a combination of PFM and PWM to achieve smooth motor operation throughout the applied voltage range.

The switch most commonly used in a voltage chopper is a thyristor. Many different types of choppers are available but all choppers may be categorized according to the type of commutation applied. There are four divisions:

1. Parallel-capacitor commutation
2. Parallel capacitor-inductor commutation
3. Series-capacitor commutation
4. Coupled-pulse commutation

Each type of commutation has advantages and disadvantages. A representative circuit of each type
will be presented and discussed.

Parallel-capacitor commutation can be further subdivided into two categories. The first type allows the capacitor to be charged for commutation through a path which doesn't contain the load. A representative chopper of this type is shown in Fig. 2.4. In this circuit the cycle begins when SCR1 is fired to initiate the load current. At the same time or a little later, SCR3 is fired which allows the capacitor to charge through \( L_1 \), SCR3, and SCR1 so that plate a is positive. Assuming negligible loss in \( L_1 \), the voltage across C reaches \( 2V_B \). When the capacitor is fully charged the current drops to 0 and SCR3 shuts off. To turn off SCR1, SCR2 is fired which places the capacitor across SCR1, reverse biasing it and turning it off. Characteristics of this circuit are:

1. It can operate as either a PWM or PFM chopper.
2. Minimum on time is \( t_{on} = \pi \sqrt{L_1 C} \). Minimum off time is load dependant. For light loads it could become excessively large.
3. Commutation voltage is not increased by load current \((D_1 \) prevents plate b of C from rising above \( V_B \)).
Fig. 2.4. Parallel-capacitor commutation, Type I.

Fig. 2.5. Parallel-capacitor commutation, Type II.
4. If SCR1 doesn't turn off when SCR2 fires, a second commutation can be reattempted by firing SCR3.
5. The rating of SCR1 is increased by the charging pulse for C.
6. A fault condition causing SCR2 and SCR3 to be fired simultaneously would short $V_B$ and destroy both devices.

Parallel-capacitor commutation of the second type allows the charging current for the capacitor to flow through the load. This type of commutation occurs in the circuit shown in Fig. 2.5. In this circuit SCR2 is fired first allowing the capacitor to charge through the load with plate b positive. When it is fully charged the current drops to 0 and SCR2 shuts off. When the main thyristor, SCR1, is fired the capacitor reverses its charge by resonating with $L_1$. The capacitor cannot discharge a second time since $D_2$ blocks current flow in this direction. When it is time to shut off SCR1, SCR2 is fired placing the capacitor across SCR1. The capacitor then charges up to $V_B$ with plate b positive and the cycle is ready to repeat.

The characteristics of this circuit are:
1. Both PWM and PFM are possible.
2. Minimum on and off times are the same as the previous circuit.
3. Commutation voltage is not boosted by load current.
4. If SCR1 fails to turn off the capacitor will discharge to 0 and SCR2 will turn off. No second chance at commutation is possible.
5. The rating of SCR1 is increased by the current required to charge C.
6. No fault condition is possible that would short-circuit the supply.

The second type of commutation, parallel inductor-capacitor, is illustrated in Fig. 2.6. In this circuit C is charged through $L_1$, $D_3$ and the load. When the load cycle begins with firing SCR1, C cannot discharge since $D_3$ is reverse biased. When SCR2 is fired C is charged oppositely by the resonant circuit C, SCR1, SCR2 and $L_1$. Finally, it discharges through $D_3$ and $L_1$ turning SCR1 off. $D_2$ limits the maximum discharge period. For this circuit the characteristics are:
Fig. 2.6. Parallel capacitor-inductor commutation.

Fig. 2.7. Series-capacitor commutation.
1. Both PWM and PFM are possible.
2. Minimum on and off times are $\sqrt{L_1 C}$.
3. The commutation voltage on C is increased proportional to load current due to transfer of energy from $L_1$ to C when SCR1 is off.
4. Only one commutation attempt is possible.
5. SCR1 rating increases due to resonant current of C.
6. There is no low impedance fault path.

Series-capacitor commutation is illustrated in Fig. 2.7. Operation begins when SCR1 is fired. This allows C to charge through $L_1$ and eventually turn SCR1 off. Then SCR2 is fired resetting C through $L_2$ and allowing SCR1 to be refired. This type of circuit has very limited use due to the capacitor reset time and is rarely used as a chopper.

A coupled-pulse commutation circuit is shown in Fig. 2.8. Circuit operation begins when SCR1 is fired. This allows $C_1$ to charge through $L_1$ to $V_B$. $D_2$ prevents $C_1$ from charging to a higher voltage. To turn SCR1 off SC2 is fired allowing $C_1$ to discharge through $L_1$. $L_1$, $L_2$ is an autotransformer so a voltage pulse is coupled from $L_2$ to $L_1$ turning SCR1
Fig. 2.8. Coupled-pulsed commutation.
off. C₂ allows SCR₁ to be fired before SCR₂ turns off. Firing SCR₁ allows C₂ to discharge through L₁ which couples a pulse to L₂ turning off SCR₂. The following characteristics apply to this circuit.

1. Both PFM and PWM are possible.
2. Minimum on-off times are $\frac{\pi \sqrt{L_1 C_1}}{1}$ and $\frac{\pi \sqrt{L_2 C_2}}{1}$ respectively.
3. Commutation voltage is fixed to $V_B$ by $D_1$ and $D_2$.
4. No second attempt at commutation is possible.
5. Both capacitor pulses flow through SCR₁ increasing its rating.
6. A fault path through SCR₁ and SCR₂ exists which could destroy both devices.¹⁹

Both series-capacitor and coupled-pulse commutation are unsuitable because high values of load current force prohibitively small values for L. Additionally, coupled-pulse commutation requires an auto-transformer increasing the cost. Generally speaking, parallel-capacitor commutation offers the most flexibility and is most often employed in electric vehicle controllers.
2.3 Commercial Motor Control Techniques

Motor controllers for electric vehicle drive systems are available from several companies. Each company has its own approach to motor control and it is the purpose of this section to compare and contrast the systems currently on the market.

The predominant form of control available utilizes SCR chopping to vary the speed of a series-wound motor. Three large companies produce products of this type: General Electric, Cableform, and Sevcon.

General Electric's controller, the EV-1, is available in a variety of sizes. It can accommodate battery voltages to 144 volts and average currents to 500 amperes with a peak current rating of 1100A. All components are mounted in a single unit which can operate in temperatures from -30C to +50C. It is recommended to mount the controller against the vehicle frame to aid in heat dissipation.

This controller is unique in that it uses parallel inductor-capacitor commutation. By using this method the commutating capacitor can be charged to
voltages exceeding battery voltage during all but the initial commutation cycle. General Electric uses three SCRs for commutation and does not allow the main SCR to go into full conduction. Instead, a bypass contactor is pulled in once 95% conduction is reached. A combination of PWM and PFM is used to assure smooth acceleration of the motor.

Safety features include a current sensor to limit motor current; pulse monitor trip which senses a shorted main SCR, or welded bypass contactor tips, and shuts down the controller; and thermal protection for the main SCR by means of a temperature sensor mounted in the heat sink.

Many additional features are also available. These include dual motor drives, dynamic or regenerative braking, and field weakening.

A second company producing standard SCR controllers for series-wound motors is Sevcon. Their Mark 7 system can handle battery voltages up to 170 volts and has continuous current ratings up to 330 amperes and peak current ratings up to 1000 amperes. Most models require mounting the traction panel to the vehicle frame for maximum heat sinking. A modular approach is used allowing the logic portion of the
controller to be mounted separate from the traction unit.

Parallel capacitor commutation is used to turn off the main SCR requiring use of two additional SCRs. As with the G.E. system, once the main SCR has reached 95% conduction a bypass contactor is pulled in to achieve full battery voltage at the motor terminals. This changeover is logic controlled to assure a smooth transition. During acceleration PFM is used until 85% of battery voltage is applied to the motor. From 85% to 95% conduction PWM is used.

Safety features include current limit, overtemperature cutback, and welded contactor detection. Additional protection to prevent engaging bypass circuitry under fault conditions is also standard.

Cableform is the third company which produces an SCR chopper control for series-wound motors. The Cableform system is unique in several areas. It utilizes a truly modular approach, breaking the controller into a power unit, motor unit, and logic unit. Commutation is by the parallel-capacitor method but uses only two SCRs. Additionally, the main SCR is allowed to reach full conduction which can preclude use of a bypass contactor. Battery voltages up to 140 volts
can be accommodated, while current ratings are 265 amperes continuously and 800 amperes peak. The controller can operate from -40°C to +50°C with no additional heat sinking. Standard circuitry includes thermal overload, current limit, and welded contactor or shorted main SCR protection.

Cableform controls were selected for use in the Electric Choremaster and a detailed description follows in the next section.

In recent years technology has made motor speed control possible using bipolar transistors. One company, Soleq, produces a control using this approach. This system uses a specially built motor which is separately controlled. For low-speed, high-torque performance the field current is at a maximum and the armature current is chopped. To produce high-speed, low-torque characteristics the field current is chopped and the armature current is constant. It is claimed that 3 to 4 times more power is developed at high speeds and that an overall efficiency of 87% is maintained.

This control system is still in the experimental phase and has been used only in highway vehicles. It appears to be a promising approach which optimizes the motor/controller/battery system for improved per-
performance.

Lucas Chloride EV Systems in England has developed two controllers which are used in commercial trucks. The larger of the two uses power transistors to control a 50 kW compound-wound motor. Maximum currents of 25 amperes for the field and 500 amperes for the armature can be supplied in this separately excited system. Battery voltage is in the range of 160-180 volts. The second controller is something of a hybrid in that it uses SCR chopping for the armature and transistor chopping for the field. This system, which operates from 150 to 250 volts, can supply peak currents of 350 amperes and 14 amperes to the armature and field respectively. Both systems can operate over a temperature range of -20°C to +40°C. Additionally, regenerative braking is a standard part of both systems.

Lucas Chloride has followed a systems approach and is involved in battery, DC-DC converter, and motor development in conjunction with the controllers mentioned above.

The five companies listed in this section are believed to be a fairly complete listing of reliable sources of motor controllers for large electric vehicles.
2.4 Electric Choremaster Motor Control

To perform all the functions of a chore tractor two electric motors were used. One motor propels the vehicle while the second powers the PTO and hydraulic (HYD) systems. As a result, two motor controllers are required. Each controller must interface its respective motor to the same power source independent of the other. The power source is a 128 V, 340 A-Hr battery; the traction motor is rated at 49 Hp for 1 hour and the PTO/HYD motor has a rating of 23 Hp for 1 hour. With these parameters in mind the specifications each controller has to meet are:

<table>
<thead>
<tr>
<th>CONTROLLER</th>
<th>Voltage</th>
<th>Ave. Current</th>
<th>Max. Motor Current</th>
<th>Temperature</th>
</tr>
</thead>
<tbody>
<tr>
<td>Traction</td>
<td>128 VDC</td>
<td>328 A</td>
<td>680 A</td>
<td>-34C to +43C</td>
</tr>
<tr>
<td>PTO/HYD</td>
<td>128 VDC</td>
<td>161 A</td>
<td>650 A</td>
<td></td>
</tr>
</tbody>
</table>

Additionally, the controllers must be capable of operating in environments typically encountered during farm work.
Cableform controllers which were purchased have the following specifications:

<table>
<thead>
<tr>
<th>CONTROLLER</th>
<th>Traction</th>
<th>PTO/HYD</th>
</tr>
</thead>
<tbody>
<tr>
<td>Voltage</td>
<td>84-140 VDC</td>
<td>84-140 VDC</td>
</tr>
<tr>
<td>Ave. Current</td>
<td>265 A</td>
<td>205 A</td>
</tr>
<tr>
<td>Max. motor current</td>
<td>800 A</td>
<td>800 A</td>
</tr>
<tr>
<td>Temperature</td>
<td>-40C to +50C</td>
<td></td>
</tr>
</tbody>
</table>

Comparison of the specifications indicates that the average current specification is not met by the traction controller. This is not a serious deficiency due to the nature of work performed by the vehicle. Chore routines rarely involve prolonged periods of heavy pulling but may include short periods where peak power is needed to the traction motor. The controller can supply these requirements. For the occasional instance when maximum power is demanded for extended periods the traction system was installed with a bypass contactor. Thus full power can be applied to the motor up to the capability of the battery.

Cableform uses a modular approach to controller design in that their MK10 system is divided into three
units. The logic unit contains the accelerator controls and all logic and oscillator circuitry required to operate and protect the high current components. The motor unit contains components necessary to condition motor response such as reversing contactors, plugging and flyback diodes, and transient suppression components. All SCRs and commutation components, as well as the shunt for monitoring battery current, are contained in the power unit.

A description of system operation follows. Please refer to Fig. A1.1 in Appendix A.

When the key switch is turned to the "ON" position B+ (128V) is applied to pin #8 on the motor unit. From pin #8 voltage is applied through the low current fuse to the suppression module SM-1, which contains a battery polarity sensing diode. B+ is applied to pin #4, one side of the forward and reverse contactors, the logic dropping resistor R1, and the power unit firing module dropping resistors R2 and R3. The motor unit B+ voltage at pin #4 is applied to logic unit pin #4, the power unit firing module yellow faston connection, and one side of the bypass contactor field coil.

From pin #3 on the motor unit, 9 to 14 volts is
applied to logic pin #3. From pin #2 on the motor unit, through R4, 7 to 14 volts is applied to the red faston on the power unit firing module.

When the system is energized as described above, and both the speed control logic safety trip relay and the system interlocks are closed, the vehicle is ready to run. Moving the Direction Switch forward and lightly pressing the Speed Control Unit (SCU) lever, closes the microswitch, MS1, and energizes the forward contactor field coil which closes the forward contactor and completes the motor circuit. Further pressing the SCU lever applies a variable voltage (0 to 5 volts) from the SCU potentiometer to the speed control logic, providing a variable pulsed output from logic pin #5 to the green faston on the power unit firing module. This applies gate voltages to the output and commutation SCRs, causing the motor to be switched to the battery at a rate of 250 to 320 hz.

Near the end of full travel on the SCU pedal the MS-2 microswitch is closed. This causes the bypass contactor field coil to be energized after the bypass timer runs out (2 sec. nom.). The contactor closes, short circuits the main SCR and simultaneously shuts off the logic oscillator. The vehicle continues to run in
bypass (full output) until the SCU pedal is released enough to open MS-2. At this time vehicle speed is again under SCU control.

The speed control logic and the power unit switched output is varied both by pulse width modulation (PWM) and pulse frequency modulation (PFM). Maximum motor current is controlled by the voltage fed from the power unit shunt to the speed control logic through pins #6 and #9.

Over-temperature protection for the power unit is provided by a thermistor located in the bottom power unit heat sink next to the main SCR. The cold thermistor resistance is approximately 70 ohms and increases with temperature. The resistance change is fed through the blue and white firing module fastons to pins #9 and #12 of the speed control logic to decrease the duty cycle if the power unit temperature rises above a safe level.

The safety trip relay prevents vehicle run away if an abnormality occurs such as a shorted output SCR or a welded bypass contactor. The speed control logic output is compared with the power unit output to determine if the system is operating correctly. If the logic sees an abnormality the safety trip relay opens.
This removes the B- return to the forward/reverse contactor coils, opens the contactors, and hence opens the motor circuit and brings the vehicle to a stop.

Diodes D1 and D2 in the motor unit are the braking diode and the flywheel diode. These diodes are protected by suppression modules SM2 and SM3 from high voltage spikes that would damage the diodes.

The braking diode is connected across the armature. Braking occurs when the motor is rotating forward, and the direction switch is reversed. With the motor field reversed in relation to the direction of the armature, currents circulate around the braking diode and the armature loop causing the anode to be 1.5 volts positive in relation to the cathode. This 1.5 volts is fed through wire #10 (off the AA buss) causing the logic to generate a narrow braking pulse, with variable frequency, which brakes the motor to a stop. When the motor stops, the 1.5 volts is no longer present, braking pulses stop, drive pulses resume, and the vehicle drives in the opposite direction.

The flywheel diode is across the armature and field. Its function is to damp the oscillatory ringing caused by pulsing the motor.

In the A1264SP01 motor unit the function of D1
and D2 changes with direction. In the forward direction D1 is the braking diode and D2 is the flywheel diode. In the reverse direction the opposite is true.

Circuitry in the speed control logic clamps the pulsed output when the direction switch is in the neutral position. Thus if the direction switch is changed from forward to reverse, under power, the clamp stops the pulsed output, and the contactors switch over without motor current present. This feature insures extended contactor life.

The system is shut off by returning the Key Switch to the "OFF" position.

The aforementioned system description is for the traction drive. The PTO/HYD motor drive is similar except that the motor runs in only one direction. Consequently, no reversing contactor, switch, or braking is necessary. This drive is shown in Fig. A1.2.

Commutation of the main SCR is accomplished in a unique manner using only one auxiliary SCR. Fig. 2.9 illustrates the parallel-capacitor commutation circuit. After the control card unlocks the gate, a pulse fires SCR2 which allows C to charge to battery voltage. When C is fully charged current starvation causes SCR2 to turn off. When the control card has sensed that C is
Fig. 2.9. Cableform controller commutation.
fully charged the gate to SCR1 is unlocked. Firing SCR1 places the load across the battery. In addition, a resonant circuit is formed around C, SCR1, L, and D. This allows the capacitor to reverse its charge but D blocks any additional oscillation. Charge reversal occurs in less than 1 millisecond. SCR1 remains in a conducting state until the control card fires SCR2. This places C directly across SCR1, momentarily reverse biases it, and turns it off. The capacitor continues to discharge through the battery and load until it charges fully in the opposite direction. The voltage across C at this point depends on the lead and motor inductance. Lack of current once again causes SCR2 to shut off.

Appendix A contains all schematic diagrams for the Electric Choremaster. Only minor changes were made to the basic Cableform control system. These changes involved additional circuitry, which was required to perform functions essential to a chore tractor. Chapters III and IV describe in detail the modifications that were made.

2.5 Electric Braking

Three forms of electric braking will be
discussed: plugging, dynamic, and regenerative. The Electric Choremaster is capable of plugging only.

Plugging is an industry-wide term which describes using the motor to apply a retarding torque to vehicle motion. This is accomplished by reversing the field of the motor while the vehicle is still in motion. The controller then applies braking pulses of short duration such that the vehicle comes to a smooth stop. After stopping, the vehicle accelerates normally in the reverse direction if the accelerator is not moved to the neutral position. Braking torque can be increased or decreased by varying the position of the accelerator. An internal resistor in the logic unit determines the severity of braking and this is currently set to a minimum value. Plugging in high gear is a very smooth operation. However, with the vehicle in low gear the drive train gear reduction is reduced to the point where plugging becomes quite stiff and jarring. Plugging is a feature which is common to almost all motor controllers and provides an acceptable way to stop a motor and reverse it without upsetting operator, vehicle, or load.

A more sophisticated form of braking, dynamic braking, converts the motor to a generator and dissipates the energy produced in a resistive load.
When the operator demands braking, the logic switches the motor armature in parallel with the field. The kinetic energy of the vehicle causes the motor to become a self-excited generator which is connected to a resistive load. Thus, the vehicle's kinetic energy is dissipated as heat in the load and forward motion is retarded.

Regenerative braking allows the kinetic energy of motion to charge the battery. The procedure is essentially the same as dynamic braking except that the energy is diverted to partially recharge the battery, instead of dumping energy in a resistive load. Higher efficiency can be obtained using this approach but the effects it has on battery performance and life are not known.

Regenerative and dynamic braking can produce very smooth braking response. However, they are only effective at substantial vehicle speeds. Since a show tractor operates at slow speeds, the marginal benefits of these techniques are far out-weighed by additional circuit complexity, maintenance, and cost.

Recent information obtained from Lucas-Chloride indicates that they have developed a system which brakes to zero ground speed. Obviously, little can be gained in
efficiency, but the application of a retarding torque when the vehicle is operating on inclined surfaces would be quite desirable.

2.6 DC-AC Drives

Recently, considerable effort has been expended in developing drives for electric vehicles powered by AC motors. This work has been concentrated on passenger vehicles intended primarily for highway use. Obviously, the requirements for these vehicles are radically different from those for an agricultural tractor. The high torque provided by the series-wound DC motor at low speed is the dominant factor in motor selection for farm chore work. However, before dismissing AC drives the relative advantages and disadvantages they offer with respect to DC drives should be investigated.

AC motors are advantageous since they are smaller, lighter, do not require brushes or a wound rotating armature, can operate at high speeds, and are cheaper than DC motors. For an electric tractor the first two items, size and weight, are not necessarily advantageous since the tractor is a large vehicle and by nature must be heavy to pull implements.
The ability to operate at higher speeds again is not clearly advantageous. Since the tractor is a slow moving vehicle, a motor operating most efficiently at high speeds would necessitate additional gear reduction. Brush arcing is no problem in any environment encountered by a farm tractor. The last item, cost, favors the AC motor. However, the cost advantage would have to be large to compensate for all other factors.

A major advantage of the DC motor is that it can be driven from the battery using relatively simple choppers as discussed previously in this chapter. In order to drive an AC motor the DC source must be converted to AC. Secondly, to vary the speed of an AC motor the frequency of the input must be varied. These items, coupled with the fact that suitable AC motors are usually three-phase machines, means that a complex inverter must be used to control the AC motor. Additionally, since all commercially available drives for electric vehicles are DC, the design philosophy of using off-the-shelf items for the electric tractor would be violated with an AC drive.

The factors listed above favor an AC drive system. However, current research efforts may develop AC drives suitable for an electric farm vehicle.
Therefore, future tractor development should not preclude this possibility.
Chapter III

INSTRUMENTATION AND AUXILIARY SYSTEMS

3.1 Motor Protection

As described in Chapter II, both the traction motor and the PTO/HYD motor are well protected from electrical overloads by the controller. However, two types of faults remain to be addressed: thermal overloads and motor overspeed.

Thermal protection is a necessary requirement for practically all motors because heat is usually the primary factor in insulation breakdown. The motors purchased for the Electric Choremaster are equipped with two thermostatic switches. One is a normally open switch which closes when additional cooling for the motor becomes necessary. The second is a normally closed switch which opens when the motor is too hot to be operating. Fig A 1.3 details the interfacing of these switches into the control system. The normally open switch switch in the traction motor is connected (via points z & zz) in series with a 12 volt squirrel cage blower on the right hand side of the drawing. Twelve volt blowers were chosen instead of 128 VDC
blowers because of availability and cost factors. As noted on the drawing, the PTO/HYD blower is connected in a similar fashion. Both blowers are physically located near their respective motors and are ducted to the motor inlet.

In the same figure, note that the normally closed switch in each motor is in series with pin #8 on the respective motor unit. Pin #8 provides 128V to the motor units which, among other duties, powers the contactors placing the motor in the power circuit. Thus, when a motor overheats and the normally closed switch opens, voltage is removed from that motor. When the motor has cooled sufficiently to close the switch, voltage can be resupplied to the motor by returning the controller accelerator to neutral to reset the system.

Motor overspeed is a potential problem with series-wound motors. In theory, the hyperbolic torque-speed curve of a series-wound DC motor shows that as torque approaches zero, speed approaches infinity. In practice this is not possible because of rotational resistance in the motor; however, very high speeds can be reached without adequate protection. If, for example, a differential u-joint breaks in the traction system, the torque load to the motor will be negligible. Under
a fault such as this the motor, and parts of the drive train connected to it, will overspeed and possibly damage part of the drive train and perhaps the motor. In order to prevent such occurrences the tachometer/overspeed circuit of Fig. A 1.4 was introduced into the system.

The circuit can be described as follows: magnetic reluctance sensors provide a frequency modulated input to the LM2907N-8 integrated circuits. This is accomplished by sensing cogs on a disc connected to the motor armature shafts. The LM2907N-8 is a frequency-to-voltage converter that provides a DC output proportional to the speed of the motors. The conversion factor is set by the 75 kilohm resistor, the .01 microfarad capacitor and the supply voltage to the chip. For this reason, an LM317 is used to provide a regulated voltage source. A speed of approximately 3000 rpm produces an output of 5.1 VDC on pin #7 of the LM2907N-8. The 1 microfarad capacitor is used to smooth the ripple of the output voltage. Two op-amps serve as comparators which are referenced at 5.1V by the zener diode. An output from either tachometer greater than 5.1V switches the respective comparator output high which turns on the associated transistor. When either
transistor turns on, the 12V relay pulls in causing the 120V relay to latch. Once the second relay has latched, the line between the key switch and controllers is interrupted and both controllers turn off. Resetting the key switch allows the controllers to operate once again. Without a latch built into the circuitry, the motors would oscillate about the overspeed set point after a fault had occurred.

3.2 Auxiliary Power System

There are a number of additional auxiliary circuits in any vehicle which are not directly associated with the vehicle drive system. Included in this category are systems such as vehicle lights and instrumentation. With the Electric Choremaster there are also unique systems such as the motor blowers and overspeed circuits discussed previously. It is most convenient to operate these circuits from a low voltage supply, typically 12 VDC. To accomplish this function a 12V automotive battery was installed on the vehicle. When the operator closes the key switch, a relay in series with the battery closes to supply voltage to the auxiliary circuits. All circuits other than the blower
circuits are protected by a two ampere fuse. The blowers are protected by a 10 ampere fuse. Typically, during normal operation, the motors rarely heat enough to require additional cooling. This is due to the "stop-and-go" nature of chore work. Since the blowers represent the only significant load on the 12V battery, it requires recharging only after several cycles of the main battery system. After nine months of operation no significant problems have been noted using this system other than the nuisance of recharging the battery periodically. A more ideal method would be to use the main batteries to power the low voltage circuits through the use of a DC-DC converter.

3.3 DC-DC Converters

The purpose of a DC-DC converter is to allow auxiliary circuitry to be supplied by the main energy source on the electric vehicle. As a result, auxiliary energy sources, as well as their charging and maintenance requirements, are eliminated. A converter should provide auxiliary power in adequate amounts, isolate the 12V circuits from the traction battery, and be protected from short circuits and overloads.\textsuperscript{21}
DC voltage conversion can be accomplished using chopper circuits such as those in Chapter II. However, a more common approach is to use a chopper to feed pulsed DC to the primary winding of a transformer. The transformer secondary is wound to produce the desired output voltage which can be full-wave rectified and filtered. Often a 12V battery is used as a part of the converter load for additional regulation and as a source of emergency power in the case of a converter failure.

Specifications, in addition to the features listed above, for a DC-DC converter to use on the Electric Choreomaster are listed below.

Input: 128 VDC
Output: 13.8 VDC (nominal)
Output Current: 15A continuous
Temp Range: -30 to +120 F
Supplemental cooling: none
Mountable in any position
Able to clean by spraying

DC-DC converters are available from at least three sources: Soleq, Sevcon, and General Electric. G.E. has redesigned their converter. At the present
time it appears that it will meet the above specifications and be reasonably priced. Unfortunately, production units will not be available until the summer of 1985. Addition of a converter to the Electric Choremaster will have to be delayed until then.

It should be mentioned that 12V auxiliary power may be obtained by tapping six cells of the main battery. This approach may be adequate for the case when auxiliary current draw is negligible compared to the main drive. For larger auxiliary loads the cells in the battery tap will discharge at a rate unequal to the rest of the battery. Upon recharging, these cells will not rise to voltages as high as the rest of the battery and will again discharge faster as the battery is used. The problem can become so severe that even an equalization charge cannot bring the voltage tap cells to par with the rest of the battery. Therefore, voltage taps are avoided in most vehicles.

3.4 Operator Controls

Any vehicle's control system is perceived as being only as good as the operator's ability to control the vehicle. For this reason, the operator-vehicle
interface must be well designed. It may be surprising to note that an electric vehicle is generally easier to operate because it has fewer controls. However, operation of vehicles with internal combustion engines is so deeply ingrained within most people that transition to electric vehicles can be difficult and awkward. Thus, it is desirable that electric controls mimic, to some extent, controls on internal combustion engine vehicles.

A great advantage of the electric vehicle is that to "start" the vehicle, the key switch is merely turned on. This simple procedure is all that is required regardless of temperature. Following this, the appropriate gear is selected without the need for clutching. The next step is to bring the PTO/HYD motor up to speed to operate the power steering. Usually a speed of 750-1200 rpm is sufficient. The improved control system discussed in Chapter IV makes setting this control a one-time operation. Finally, the traction motor control is moved in the desired direction of travel. Both the reversing switch and the accelerator for the traction motor are combined into one control. With the lever in the middle of the range, the motor has no voltage applied to it. As the lever is
moved in the direction of travel, the appropriate directional contactor in the traction motor unit is energized. Further motion in this direction allows the motor to accelerate. If desired, the lever can be moved in the opposite direction, through neutral, causing the motor to brake and reverse direction. This type of control imitates the hydrostatic drive found in some conventional tractors without the corresponding power loss associated with hydraulic systems.

Incidental to this discussion is the fact that engine monitoring instrumentation can normally be eliminated, since protection circuitry is automatic for electric motors. Thus, the operator is freed from constantly monitoring oil pressure, coolant temperature, transmission oil temperature, and any other functions associated with conventional vehicles. However, since the Electric Choremaster is an experimental vehicle, some instrumentation is required for research purposes.

3.5 Instrumentation

Most instruments available today are built to measure steady state DC or sinusoidally-varying AC parameters. Unfortunately, the currents and voltages
present in the Electric Choremaster are nonsinusoidally-varying waveforms. Therefore some consideration should be given to instrumentation.

For the general case, considering both AC and DC circuits, several guidelines for power measurements have been developed:

1. Average power and rms values of voltage and current have definite physical meaning in the case of general periodic waveforms and are therefore the most useful measurements.
2. Instantaneous power, rectified average measurements of voltage and currents, and peak values of voltage and current may be useful in special applications.
3. Power factor has been defined for nonsinusoidal waveforms, but its limited application is stressed.
4. For the purpose of analysis, Fourier series expansion of periodic waveforms plus knowledge of circuit and load impedances at one particular frequency (60 Hz typically) is the correct approach (assuming linear devices).
5. Use of reactive power measurements should be avoided for nonsinusoidal waveforms.²²
The choice for DC measurements is a permanent magnet, moving coil instrument. Errors can be introduced if AC components reach the movement producing eddy currents. The resulting torques produced are small and may be ignored except when very accurate measurements are being attempted. Thermocouple and electrodynamometer devices are unacceptable since they respond to AC as well as DC components. Instrumentation devices also should not be frequency dependent. For example, a current shunt should have little inductance associated with it in a low impedance circuit.

Refer once again to Fig. A 1.3 which includes details of the instrumentation on the Electric Choremaster. Instruments and components located in the tractor cab are noted by a small letter c contained within a circle.

The battery voltmeter is engaged when the key switch is on. This provides the operator with knowledge of the key switch position, the unloaded battery voltage, and the battery voltage under load.

Ammeters are provided for both the PTO/HYD motor drive and the traction motor drive. The shunts are located on the tractor control panel while the meter
movements are in the cab. By positioning the shunts as shown, only the battery current is measured. If the shunts were located between the motors and their respective motor units (i.e. between B+ on the controller motor unit and armature terminal A), the ammeters would read motor current. Motor current and battery current are not the same since a circulating motor current flows through the free-wheeling diode during the time that the main SCR is off. Although motor current is a parameter necessary to the design of appropriate controller circuitry, it was thought that battery current should be measured when evaluating vehicle performance. The ammeter in the traction drive has a capacity of 1000A while the PTO/HYD meter capacity is 500A.

A tachometer is necessary so that the operator can monitor motor speed. Fortunately, the tractor had one in place before it was converted so that only installation of magnetic reluctance sensors and discs on the motor drive shafts was necessary. By switching in the appropriate sensor, the operator can monitor the speed of either motor or vehicle ground speed. Additional tachometer circuitry was installed to prevent motor overspeed and improve speed control as
discussed in Section 3.1 and Chapter IV respectively.

A battery state of charge (SOC) device was installed to provide an electric analog for a conventional fuel gauge. The motor controller chopping action produces dips in voltage at the battery terminals each time the thyristors fire. The SOC integrates these dips over time to provide an indication of the battery state. In order for a fair degree of accuracy to be maintained the battery should be discharged along a standard discharge curve. Because of the wide variety of work this tractor performs, the rate of discharge of the batteries cannot be guaranteed. Consequently, this device has not been a reliable indicator of the battery's state of charge. Other devices are available which may be more accurate. Additional investigation in this direction is indicated.

The instruments discussed thus far are found on a majority of electric vehicles. Due to the experimental nature of the Electric Choremaster, a DC kilowatt-hour meter was installed. By measuring both battery voltage and current, and integrating their product over time, a measure of the energy supplied by the battery to the vehicle is produced. Using this device, side-by-side energy comparisons have been made
between the Electric Choremaster and a Versatile 160 performing identical tasks. The DC kilowatt-hour meter can be used to measure energy consumption of all drive train components from which component efficiency data may be obtained.
Chapter IV
IMPROVED SPEED CONTROL

4.1 The Problem

Initial laboratory testing indicated, and later on-farm testing verified, the initial speed control system was inadequate. The speed-torque characteristics of series-wound DC motors is shown in Fig. 4.1. Indicated speed and torque follow an inverse relationship which is approximately

\[ \omega = \frac{V_t}{(k_f T)^{1/2}} \]

where \( \omega \) = motor speed (rad/sec)
\( V_t \) = terminal voltage (volts)
\( k_f \) = motor constant
\( T \) = torque (N * m)

neglecting armature and field resistance. Any change in torque produces an inverse change in motor speed which can be quite pronounced depending on the portion of the curve in which the motor is operating.

Due to the nature of the tasks which the Electric Choremaster performs (i.e. stop-and-go, backward-and-forward), the operator is continually adjusting tractor motor speed such that speed changes due to torque disturbances often go unnoticed. However,
Fig. 4.1. Speed-torque characteristic of series-wound dc motors.
many operations are greatly simplified by setting the PTO/HYD motor at a fixed speed throughout the chore routine. A common occurrence is to set the PTO/HYD motor speed to maintain adequate hydraulic fluid flow for power steering, typically 1000 rpm. Any use of the hydraulics then places a torque disturbance on the motor and severely reduces motor speed. Since the hydraulic system has a priority valve for the power steering circuit, power steering is maintained but any additional hydraulic usage requires increasing the motor speed control. Following this, once the torque disturbance is removed, the motor overspeeds wasting battery energy by dumping hydraulic fluid through the pressure-relief valve and causing excessive wear on the drive train. This type of problem can be generalized to PTO operation. Since the PTO load is rarely constant, continual adjustment of the speed control would be necessary if a relatively constant speed is desired.

4.2 Proposed Solution

As a result of this problem, the original open-loop control system, as shown in Figure 4.2 was modified to form a closed-loop system to improve speed
Fig. 4.2. Open-loop control system for PTO/HYD drive.

Fig. 4.3. Closed-loop control system for PTO/HYD drive.
regulation as shown in Figure 4.3. The additional circuitry required to close the loop consists of only the summing network since the tachometer was in place as a requirement for motor protection. There was a difficulty in implementing the summing network since the voltage source for the network was 0-12 VDC making it impossible to produce a negative error signal, e(t). For this reason, a summing network was utilized with four op-amps which split the error signal into two parts. Positive error signals (indicating overspeed) are subtracted from the set speed, \( v_s \), and negative error signals (indicating underspeed) are added to the set speed. The summing network is shown in Figure 4.4.

U1a and U1b act as buffers to prevent any loading effects from influencing the set speed, \( v_s \), and the actual speed, \( v_a \), input signals. U2a, U2b, and U2c act as difference amplifiers. U2a performs \( v_a - v_s \) giving a nonzero output only if \( v_a > v_s \). U2c performs \( v_s - v_a \) giving a nonzero output only when \( v_s > v_a \). Only one of these op-amps will have a nonzero output at any instant in time. The op-amp U2b subtracts the overspeed error signal obtained from U2a, from \( v_s \). This gives a corrected signal at its output. U2d acts as an adder for the output of U2b and U2c. For the overspeed
All op-amps are LM224

All resistors, unless otherwise noted, 10kΩ

Fig. 4.4. Summing network for closed-loop PTO/HYD drive.
condition, \( v_a > v_s \), the output of U2c is zero so the output voltage \( v_o \) is \( v_s \) - overspeed error. For the underspeed case, \( v_a < v_s \), the output from U2a is zero, hence the output from U2b is \( v_s \). U2c performs the function \( v_s - v_a \) and applies this signal to the adder U2d such that its output equals \( v_s + \text{error signal} \).

The overall function this network implements is:

\[
v_o = v_s - (v_s - v_a) = 2v_s - v_a.
\]

The 470 ohm resistors connected to the output of each op amp are to provide enough loading to insure each op-amp output will be driven to zero for a zero-producing input. The zener diode is to prevent the input to the motor controller from exceeding 5.1 VDC. A switch (S1) was installed to permit bypassing the feedback network which enables the operator to evaluate closed-loop and open-loop performance of the system easily.

After implementing the closed-loop system two problems were noted. First, when the motor was run at speeds in the upper half of its operating range (>1500 rpm) and a torque disturbance was introduced, the system would oscillate. Further investigation revealed
significant noise was present on the feedback signal. This is not surprising since the signal originates at the rear of the tractor (on the controller panel), runs the full length of the tractor past the high-voltage choppers for the motor drives to the roof of the cab. In order to bypass this noise the asterisked capacitor and resistor were added to the non-inverting input of Ulb. The second problem was excessive steady state error (25-30%).

In order to optimize the system response it was necessary to develop transfer functions for each block of the system. A problem with this approach is that both the controller and the motor/pump combination are quite nonlinear so that only an approximate model over small portions of the dynamic range of the system can be generated.

4.3 Transfer Function Development: motor/pump

This section describes development of the transfer functions for the motor/pump combination, the motor controller, and the tachometer as denoted in Fig. 4.5.

Unfortunately, series-wound DC motors are not


\[ H_c(t) = \text{controller transfer function} \]

\[ H_m(t) = \text{motor/pump transfer function} \]

\[ H_r(t) = \text{tachometer transfer function} \]

**Fig. 4.5.** Transfer function block diagram for FTO/HYD drive.

**Fig. 4.6.** Circuitry used to determine step response of motor/pump.
linear devices. So it follows that combining the motor and hydraulic pump into one unit could be expected to produce a high-order nonlinear system. One common approach to handling nonlinear systems is to analyze small perturbations about stable operating points. This approach assumes the nonlinear system response can be modelled linearly if the operating range is restricted sufficiently. A method presented by Puchalka & Wozniak indicates an approximate second-order model with no zeros may be used for a series DC motor if the magnetization curve is approximated with a polynomial expression. Hydraulic pumps are often modelled as third order systems without zeros. It seems reasonable then to expect the motor/pump combination to result in a fifth-order transfer function.

Analyzing the motor and pump together is reasonable since the pump is physically attached to the motor and normal operation continuously couples the loading effect of the pump to the motor. Also, the most pronounced lack of speed regulation occurs during use of hydraulic systems with the tractor. The primary purpose, then of this development was to reduce speed variations produced by hydraulic loading on the motor. The effect of PTO system loading was not addressed
directly, however, this type of loading typically is less severe. Torque disturbances via the PTO are generally smooth in nature after initial start-up. Since the PTO directly affects the motor only and not the hydraulic pump, it was thought that analyzing the motor/pump combination with the end result to be reducing speed variation due to hydraulic torque disturbances will provide a worst-case analysis for system stability.

To adequately cover the operating range of the motor/pump it was decided to find the transfer function at four different operating speeds: 500 rpm, 1000 rpm, 1500 rpm, and 2000 rpm. In order to do this, the motor was isolated from the rest of the electrical system. A DC voltage was applied to maintain motor speed at one of the above operating points. Then a small voltage increase in the form of a step function was applied to the terminals to provide an increase in speed of approximately 200 rpm. The time response of the motor speed was plotted for this step input.

Fig 4.6 is an illustration of the circuitry used to obtain the motor/pump step response. Resistors R1 and R2 are adjusted such that when S1 is open the operating speed is maintained, and when S1 is closed the
motor/pump speed is increased approximately 200 rpm after all transient responses have died out. The power source was the battery from the Electric Choremaster. The tachometer function was obtained from the overspeed protection circuitry. Compared to the motor/pump, the response time of the tachometer was negligible so that its transfer function is merely a constant.

Figures 4.7 through 4.10 present the data collected at the four operating speeds. Several interesting observations can be made from a casual inspection of the data.

1. The curves are smooth, continuous, and well-behaved.
2. There appears to be no overshoot or oscillation. This suggests an underdamped or critically damped system.
3. The slope of the curves at t=0 is zero which suggests no effects from system zeros.

Even though the system step response is known, there is little known about the actual form of the transfer function other than what has already been discussed. One of the most important parameters of the
Fig. 4.7. Step response of motor/pump at 500 rpm.
Fig. 4.6. Step response of motor/pump at 1000 rpm.
Fig. 4.9. Step response of motor/pump at 1500 rpm.
transfer function concept is the order, or number of poles. One method for determining system order is by use of a Hankel matrix. The Hankel matrix is an arrangement of the discrete impulse response data of a system in the form:

\[
H_T = \begin{bmatrix}
    h_T(1) & h_T(2) & \cdots & h_T(n) \\
    h_T(2) & h_T(3) & \cdots & h_T(n+1) \\
    \vdots & \vdots & \ddots & \vdots \\
    h_T(n) & h_T(n+1) & \cdots & h_T(2n-1)
\end{bmatrix}
\]

where \( h_T(n) \) denotes the discrete pulse response at time \( n \). The argument, \( n \), is a multiple of sampling time, \( n=kT \), where \( T \) is the sampling interval and \( k \) is the multiplier. Since the Hankel matrix is the product of the system controllability and observability matrices, it has the property that its rank is equal to the order of the system. Notice, however, that no information is given concerning the location of the system poles.

In order to employ the Hankel matrix we must choose a sampling interval, convert the data from continuous form to discrete form, and find the discrete impulse response from the discrete step response.
Care should be exercised in choosing the sampling interval. To avoid masking higher frequency components, guidelines provided by the Nyquist sampling theorem should be followed,

\[ T < 1/2f_{\text{max}} \]

where \( T = \) sampling interval

\( f_{\text{max}} = \) max. freq. component.

As the data show in Figures 4.7 to 4.10, there appears to be very little or no oscillatory nature. Thus, we are not severely restricted in our choice of sampling intervals and it appears that \( T=.2 \) sec may be reasonable. Indeed, as tests are made for higher system orders the sampling interval is forced to become smaller. For example, to test for a tenth order system \( 2n-1=19 \) samples are required. For data ranging from 0 to 1 sec,

\[ T < \frac{1\text{ sec}}{19 \text{ samples}} = .05 \text{ sec/sample}. \]

To convert from continuous data to discrete data is fairly simple. After choosing the sampling
interval, the corresponding discrete value is obtained from the continuous value at the sampling instant. This value is maintained during the sampling interval until a new value is obtained at the next sampling instant. More error is introduced for longer sampling intervals which is further reason for keeping the sampling interval short.

Once the sampling interval has been chosen and the discrete data obtained, the final step is to convert from the discrete step response to the discrete impulse response. The discrete impulse response is easily found from the relationship

\[ h_T(k) = r(kT) - r(kT-T) \quad k = 1,2,3 \ldots. \]

where \( h_T(k) \) = discrete impulse response
\( r(kT) \) = discrete step response
\( T \) = sampling interval.

This equation shows that the impulse response at a given sampling instant is the difference between the step response at that instant and the step response at the previous sampling instant.

It is reasonable to assume that since one system is being analyzed at four different speeds, the Hankel
matrix need only be applied at one operating speed to determine the system order. This assumption is indicated by the data since the form of the response is similar at each operating speed. The data also seem to indicate the system order may be less than might be expected due to the smooth non-oscillatory response.

The approach employed to determine the rank of the Hankel matrix was a modified form of the Gauss-Jordan method for solving linear systems. To use the Gauss-Jordan method to find the rank of the Hankel matrix involves entering the values for the Hankel matrix into the nxn matrix for Gauss-Jordan and letting the method perform elementary row operations until a diagonal matrix is formed. The method is stopped at this point and the number of nonzero values on the diagonal is observed to ascertain the rank of the matrix. Note that the n+1 column does not affect any of the row operations. This method is most efficiently executed using a digital computer.

Initially, the rank of a small (3x3) Hankel matrix was investigated. When no zeros appeared on the diagonal a larger (5x5) matrix was formed and evaluated. Again no zeros were obtained on the diagonal and the matrix was enlarged. This phenomenon kept occurring
until finally a 10x10 Hankel matrix was formed. The values in this matrix were obtained from data collected in the 1000-1200 rpm operating range with a sampling interval $T = 0.04$ sec. The results were once more inconclusive as ten nonzero values were obtained:

\[
\begin{array}{cc}
18 & -6.42 \\
-22 & -2.34 \\
-1.78 & 4.13 \\
83.16 & 0.278 \\
4.76 & -6.97 \\
\end{array}
\]

At this point it appeared as though the method was failing since it was unlikely that the motor/pump system would be of order 10 or greater. It might be argued that the value 0.278 could be considered close enough to be regarded as zero. But the next larger value (-2.34) is less than an order of magnitude away which really doesn't provide a clear break as to when a value could be considered zero. Even if it were zero the Hankel matrix would be indicating a ninth order system which still seems unreasonably large.

There may be several reasons why the Hankel matrix failed to give a clear indication of system
order. It can be verified that this method works well for impulse responses from known linear systems of various order. But the system in question is decidedly nonlinear. The initial assumption was that by limiting the operating range the system could be made to look more linear. Perhaps if the range were limited even further the system could be made to look linear enough for this approach to work. Further restrictions are undesirable, however, since it would provide very little information over the entire range of operation. Another reason for failure of the Hankel matrix may be accuracy. The data were obtained from a storage oscilloscope and so for all practical purposes two-digit accuracy was all that could be obtained. Due to the number of computations required to find the rank of large matrices a significant amount of error may have been introduced. Regardless of why the Hankel matrix failed in this case, it would be most interesting to investigate further the effects that system nonlinearity and data accuracy make on this method.

After trying to determine system order by the approach used above, it was decided to investigate the system by forming a finite-difference equation. A single-input single-output system which can be
represented by the differential equation

\[ y^{(n)}(t) + a_{n-1} y^{(n-1)}(t) + \ldots + a_0 y(t) = \]
\[ b_m u^{(m)}(t) + \ldots + b_0 u(t), \]

where the superscript in parentheses denotes differentiation, can be represented in discrete form by

\[ y_T(k+n) + a_{Tn-1} y_T(k+n-1) + a_{Tn-2} y_T(k+n-2) + \ldots + a_{To} y_T(k) = \]
\[ b_{Tn-1} u_T(k+n-1) + b_{Tn-2} u_T(k+n-2) + \ldots + b_{To} u_T(k) \]

Whereas the nth order differential equation contains n derivatives of \( y \), the finite difference equation contains n past values of \( y \). This may be seen more easily by writing it in the form

\[ y_T(k) = -a_{Tn-1} y_T(k-1) - a_{Tn-2} y_T(k-2) - \ldots - a_{To} y_T(k-n) \]
\[ + b_{Tn-1} u_T(k-1) + b_{Tn-2} u_T(k-2) + \ldots + b_{To} u_T(k-n) \]

In this form all variables have been shifted n units in time. For time-invariant systems the coefficients

\[ \{a_{To}, a_{T1}, \ldots, a_{Tn-1}\} \]

and

\[ \{b_{To}, b_{T1}, \ldots, b_{Tn-1}\} \]
are constant but dependent on the sampling interval \( T \). Calculating a new value for \( y \) requires knowing \( n \) past values of \( y \) and \( n \) past values of the input \( u \).

If the discrete impulse response is known for a given system the finite-difference equation can be calculated. This may be done by solving a linear equation system of the form:

\[
\begin{bmatrix}
h_T(1) & h_T(2) & \ldots & h_T(n) \\
h_T(2) & h_T(3) & \ldots & h_T(n+1) \\
\vdots & \vdots & \ddots & \vdots \\
h_T(n) & h_T(n+1) & \ldots & h_T(2n-1)
\end{bmatrix}
\begin{bmatrix}
a_T(0) \\
a_T(1) \\
\vdots \\
a_T(n-1)
\end{bmatrix}
= \begin{bmatrix}
h_T(n+1) \\
h_T(n+2) \\
\vdots \\
h_T(2n)
\end{bmatrix}
\]

Solving this system by using an appropriate numerical method will yield all coefficients for terms containing the past values of \( y \) in the finite-difference equation. Notice that the left-hand matrix is the Hankel matrix.

Once the \( a_T \) coefficients have been found, the \( b_T \) coefficients may be solved for directly. At this point the nth-order finite-difference equation has been found giving a discrete approximation to the continuous system.

Just as the nth-order differential equation for
the continuous system can be represented in transfer function form, so also the discrete finite-difference equation can be represented by using the $z$ operator.

$$H_T(z) = \frac{b_{Tn-1}z^{n-1} + b_{Tn-2}z^{n-2} + \ldots + b_{To}}{z^n + a_{Tn-1}z^{n-1} + \ldots + a_{To}}$$

Equation 4-1

Notice that this representation is for the case of a proper system with no feed-forward term (i.e. the highest power of $z$ in the numerator is less than the highest power of $z$ in the denominator.

Using the discrete transfer function concept, many of the analysis techniques used in the continuous case may be applied. More importantly, the discrete poles of the system can be located and then converted to continuous poles using the relationship

$$p_i = \frac{\ln(p_{T1})}{T}$$

where $p_i$ = $i$th continuous pole

$p_{T1}$ = $i$th discrete pole.

Equation 4-2

In order to find the discrete poles, the denominator of
Equation 4-1 is set equal to zero and the n roots of the equation are found. Similarly, solving the numerator polynomial yields location of the system zeros.

Using the techniques discussed above, the 10th-order Hankel matrix found for the motor/pump operating at 1000-1200 rpm was used to find the corresponding finite-difference equation. The coefficients found are listed below.

\[ a_{T0} = -.01268 \quad b_{T9} = 18.00 \]
\[ a_{T1} = -.10125 \quad b_{T8} = 33.45 \]
\[ a_{T2} = .13591 \quad b_{T7} = 40.47 \]
\[ a_{T3} = -.14924 \quad b_{T6} = 46.02 \]
\[ a_{T4} = .03151 \quad b_{T5} = 46.96 \]
\[ a_{T5} = -.31423 \quad b_{T4} = 34.83 \]
\[ a_{T6} = .14542 \quad b_{T3} = 22.85 \]
\[ a_{T7} = -.14122 \quad b_{T2} = 13.87 \]
\[ a_{T8} = .37302 \quad b_{T1} = 8.806 \]
\[ a_{T9} = .19194 \quad b_{T0} = 6.737 \]

Once these coefficients had been found, the discrete system poles and zeros could be located using Equation 4-2. These were found to be
It is interesting to note that all pole locations lie within a unit circle about the origin. This is a necessary condition for discrete system stability. Since the motor/pump was stable, the pole locations seem to be consistent with the physical system. Another insight into the system can be gained by plotting pole and zero locations on the complex z-plane. This is done in Fig. 4.11. Particularly important to notice is that all complex-conjugate poles are nearly cancelled by a zero! Taking into consideration the amount of calculation required to obtain the pole-zero locations, as well as the degree of accuracy necessitated in collecting the data, it is reasonable to conclude that the complex poles are cancelled by the complex zeros. The result is that the motor/pump can be accurately modeled by a second-order
Fig. 4.11. Discrete pole-zero locations for motor/crump operating at 1000 rpm.

Fig. 4.12. Motor controller input and output levels.
system.

Once the system order has been determined, the next step is to find the location of the continuous poles. In deriving the second order finite-difference equation, choice of sampling interval, $T$, is important since only four sampling instants are needed. Through a trial and error method it was found that $T = .15$ second, resulted in the most accurate finite-difference equation of order two. For this sampling interval,

$$
\begin{align*}
 h_T(1) &= 102 \\
 h_T(2) &= 82 \\
 h_T(3) &= 32 \\
 h_T(4) &= 12
\end{align*}
$$

which leads to solving the matrix equation:

$$
\begin{bmatrix}
102 & 82 \\
82 & 32
\end{bmatrix}
\begin{bmatrix}
a_{T0} \\
a_{T1}
\end{bmatrix}
= 
\begin{bmatrix}
-32 \\
-12
\end{bmatrix}
$$

producing $a_{T0} = .012$ and $a_{T1} = -.40$.

These values produce the characteristic equation:
\[ z^2 - 0.40z + 0.012 = 0 \]

From which the discrete poles are found to be

\[ \begin{align*}
    p_{T1} &= 0.37 \\
    p_{T2} &= 0.03.
\end{align*} \]

Finally, Equation 4-2 can be used to find the continuous poles

\[ \begin{align*}
    p_1 &= \ln(0.37)/0.15 = -6.63 \\
    p_2 &= \ln(0.03)/0.15 = -23.4
\end{align*} \]

The form of the transfer function for the motor/pump operating at 1000-1200 rpm is

\[ H_m(s) = \frac{K}{(s+6.63)(s+23.4)} \]

Equation 4-3

since the data indicate no system zeros. Since a step input was used to obtain the data in Fig. 4.8, the output can be expressed as

\[ Y(s) = \frac{K}{(s+6.63)(s+23.4)} \times \frac{7.5}{s} \]
Using the final value theorem the value of $K = 4861$ can be found. So that the transfer function is now defined to be

$$H(s) = \frac{4861}{[(s+6.63)(s+233.4)]}$$

Equation 4-4.

for the motor/pump operating at 1000-1200 rpm.

Once the transfer function has been found for one operating range the same techniques may be employed to find the transfer function for the other operating speeds.

For the motor/pump operating at 2000-2200 rpm, the data from Fig. 4.10 were used. Once again, by trial and error, it was found that $T = .15$ sec produced the most accurate second-order finite-difference equation. The discrete poles were

$$P_{T1} = .37 \quad P_{T2} = .09$$

which resulted in continuous poles of

$$p_1 = -6.63 \quad \text{and} \quad p_2 = -16.05.$$
It is interesting to note the $p_1$ is the same at both 1000-1200 rpm and at 2000-2200 rpm. The resulting transfer function is

$$H_m(s) = \frac{2885}{(s+6.63)(s+16)}$$

Equation 4-5

However, when this transfer function was used to estimate the curve in Fig. 4.10, it was found to contain considerably more error than the transfer function in Equation 4-4. This was especially apparent during the first quarter second of the response curve. Since this is the period when $p_2$ is exerting its influence, and since $p_1$ was also obtained in Equation 4-4, the location of $p_2$ was suspected to be the source of the increased error. By letting $p_2$ slide down from 16 to 8, various transfer functions were obtained and analyzed for error with the results that if $p_2 = 10$ error was minimized. After relocating $p_2$, the final transfer function for 2000-2200 rpm was obtained,

$$H_m(s) = \frac{1803}{(s+6.63)(s+10)}$$

Equation 4-6.
The transfer function for the motor/pump operating at 1500-1700 rpm was obtained from the data in Fig. 4.9. As in the first two cases discussed, the most accurate sampling interval was found to be $T = 0.15$ sec. Using the same approach, the discrete poles were found to be complex conjugates which leads to continuous poles which are also complex.

$$p_{1,2} = -5.8 + 3.9j$$

The corresponding transfer function for these poles is

$$H_m(s) = \frac{1317}{s^2 + 11.6s + 48.9}$$

Equation 4-7.

However, this transfer function was found to be less than ideal. Its accuracy was not good and it produced an underdamped response which was not indicated by the data. It was also surprising that since this operating speed lies between the first two, that the transfer function did not lie somewhere "between" the first two (Equations 4-4 and 4-6).

With this insight in mind various transfer functions with the form
$$H_m(s) = \frac{K}{(s+6.63)(s+a)}$$

where \(a\) was varied over the interval \(20 > a > 15\), were derived. It was found that \(a = 16\) produced the most accurate result. This led to a final transfer function of

$$H_m(s) = \frac{2856}{(s+6.63)(s+16)}.$$  

Equation 4-8

Finally, the motor/pump transfer function at 500–700 rpm was derived. Following the same procedure as before with \(T = .15\) produced poles which were complex-conjugates. This type of transfer function was rejected as it was in the 1500–1700 rpm case. Reviewing the functions derived at the other operating speeds indicated that the second pole, \(p_2\) appeared to be moving negatively along the real axis as the operating speed was decreased. It seemed reasonable then that perhaps the motor/pump could be adequately modelled as a first-order system at this operating speed. Using this approach a transfer function of the form
was evaluated. Leaving the pole located at $p_1 = 6.63$, as it was at the three other operating speeds, produced a curve whose accuracy was marginal. Since this is a first-order model moving the pole to $1/T_c = 5.56$, which is the time required for the system to attain 63.2% of its final value, is suggested. Moving the pole to this location resulted in the transfer function

$$H_m(s) = 191/[s+5.56]$$

which produces a response curve displaying accuracy of the same degree as that obtained from the other transfer functions.

To summarize this section the four transfer functions for the motor/pump system are shown in Table 4-1.
Table 4-1

Transfer functions for the motor/pump system

<table>
<thead>
<tr>
<th>RPM Range</th>
<th>Transfer Function</th>
</tr>
</thead>
<tbody>
<tr>
<td>500-700 rpm</td>
<td>( H_m(s) = \frac{191}{s+5.56} )</td>
</tr>
<tr>
<td>1000-1200 rpm</td>
<td>( H_m(s) = \frac{4861}{(s+6.63)(s+23.4)} )</td>
</tr>
<tr>
<td>1500-1700 rpm</td>
<td>( H_m(s) = \frac{2856}{(s+6.63)(s+16)} )</td>
</tr>
<tr>
<td>2000-2200 rpm</td>
<td>( H_m(s) = \frac{1803}{(s+6.63)(s+10)} )</td>
</tr>
</tbody>
</table>

4.4 Transfer Function Development: controller

The motor controller, like the motor/pump, is nonlinear. This is where the similarity ends as the type of nonlinearity one system exhibits is quite different from the other. Whereas the motor/pump could be linearized by reducing its operating range, the controller remains nonlinear regardless of operating
range. Additionally, the controller response is dependent both on frequency and amplitude of the input signal. A block diagram of the controller with its input and output signals is shown in Figure 4.12.

The controller output is taken to be the average voltage seen at the motor terminals. The reason for this is that the motor responds to the average voltage seen at its terminals. By the method described in Chapter II the controller provides a chopped DC waveform to the motor. The average voltage at the motor terminals is directly proportional to the percent "on" time of the main SCR.

Under steady-state conditions the controller acts as a voltage amplifier. An input of zero volts produces an output of zero volts; an input of 5 volts produces an output of 130 volts or whatever the battery will deliver. Hence the DC gain is $K_c = 130/5 = 26$.

The DC gain $K_c$, represents an upper limit on what the output voltage of the controller can reach. If the input moves negatively, the output is obliged to follow it, regardless of frequency, if such a change would mean the output would exceed the upper limit dictated by $v_{max} = 26v_{in}$. For example if $v_{in} = 3V$, then $v_{max} = 78V$.

Assuming $v_o = 78$, if $v_{in}$ is suddenly reduced to $v_{in} = 2V$,
then \( v(t) = 26v_{in}(t) \) during this interval provided \( dv_{in}(t)/dt < 0 \). However, if \( v_{in} = 3V \), \( v_o = 20V \), and \( v_{in} \) is suddenly reduced to \( v_{in} = 2V \), it is not necessary for \( v_o(t) = 26v_{in}(t) \) during this interval since \( v_{max} = 52V \) is never reached by \( v_o \). In fact \( v_o \) will increase during this period.

Now consider the case where \( v_o(t) < v_{max} = 26v_{in}(t) \). During intervals such as these the controller output will increase at a constant rate with regard to time. The reason for this is that the controller does not allow full battery voltage to be supplied to the motor instantaneously. This is to prevent the application of unnecessary torques to the drive system of the tractor.

In order to measure the slope of the ramp, a 0-5 volt step input was applied to the controller and the DC average output voltage was measured as a function of time. To do this required the use once more of a storage oscilloscope and, additionally, a low pass filter.

There are two possible places to measure the output voltage of the controller. These are indicated in Fig. 4.13 along with the circuitry required to measure \( v_o \). In order for the controller to function properly it must be connected to an inductive load.
Fig. 4.13. Method used to find controller transfer function.
This required this test to be conducted without isolating the controller from the motor/pump. With the Cableform system the gate voltage, $v_1$, is high when the SCR is on and is low when the SCR is off. Since the controller logic alone determines this signal, it is a reliable indicator of the average DC voltage output of the controller (i.e. the motor/pump does not affect the signal). It was found experimentally that the voltage across the SCR, $v_2$, is an exactly inverted form of the gate voltage, $v_1$, and can provide an indication of the controller output voltage as well. In summation, $v_o$ can be found by two means:

$$v_o = 26v_1$$
$$v_o = 130 - v_2$$

Equation 4-9

Because the actual output voltage of the controller is a chopped form of the battery voltage, a low-pass filter is required to observe the average DC level. This function is performed by the 2nd op amp shown in Fig. 4.13. The circuit is a 2-pole Sallen and Key filter with a Butterworth response and a DC gain of 2. Proper voltage reduction is provided by the first op
amp to compensate for the filter gain and also reduce the voltage, \( v_2 \), to safe levels for the integrated circuits. The filter break frequency is set at 40 Hz and a roll-off of approximately 35 dB/decade was observed. The filter worked quite nicely for the data collection, removing nearly all trace of the 400 Hz chopping of the controller.

To observe the controller's ramp-like increase when the output voltage, \( v_0 \), is less than \( v_{\text{max}} \), a 0-5 VDC step was applied to the input and the output recorded on the storage oscilloscope. Fig 4.14 shows the results. Two interesting observations can be made. First is the step from 127V to full battery voltage occurring approximately 1.1 sec. after the step input. The controller operates in the switching mode until the SCR has reached 95% conduction. After 95% conduction the SCR is turned completely on. This is done to eliminate overly short "off" times which are damaging to the SCR.

A second observation is that instead of ramping at a constant slope, two slopes are indicated. A slope of approximately 76 V/sec occurs until \( v_0 = 44 \), then a second slope of approximately 160 V/sec begins. This phenomenon was entirely unsuspected and caused increased
Fig. 4.14. Step response of Cableform controller.
difficulty in dealing with this part of the system.

For the sake of completeness, a negative step input was applied to the controller (5 → 0 volts) and the output was observed to follow the input as discussed previously. As a first attempt to deal with this unique nonlinearity the describing function concept was investigated.\textsuperscript{24} This technique involves applying a sinusoid to the input and expanding the output into a Fourier series. However, to linearize the system only the fundamental component of the series is retained. The result is a function dependent on the input sinusoid's amplitude and frequency. While this approach at first sounds promising, an analytic solution is impossible because of the inability to find a limit when integrating to get the Fourier coefficients.

To illustrate this suppose a sine wave is applied to the input, $v_{in}=M\sin\omega nt$. During the first half cycle a ramp limit is described according to $v_{max}=26\sin\omega nt$. The input is negative during the second half cycle which forces $v_{max}=0$. When $v_{o}<v_{max}$ the output will ramp up until $v_{o}=v_{max}$. After this point is reached $v_{o}=26v_{in}$ until the input reaches $v_{in}=0$. During the second half-cycle $v_{in}<0$ so $v_{o}=0$ for this portion of the period. Fig. 4.15 shows $v_{o}$ and $v_{max}$ assuming, for simplicity,
\[ v_{\text{max}} = 26v_{\text{in}} = 26M\sin \omega t \]

**Fig. 4.15.** Controller output for a sinusoid input.

**Fig. 4.16.** Controller output as a function of frequency.
that the ramp has a constant slope. Notice that for
\( \omega t < \omega t_1 \), the output is ramping up at a fixed rate,
\[ v_0 = a(\omega t). \]
But for \( \omega t > \omega t_1 \) the output is following \( v_{\text{max}} \),
\[ v_0 = v_{\text{max}} = 26M \sin \omega_n t. \]

In order to find the Fourier coefficients for the describing function the following integrations must be performed:

\[ A_n = \frac{(2/T)}{I} v_0 \cos \omega t \, d(\omega t) \]
\[ B_n = \frac{(2/T)}{I} v_0 \sin \omega t \, d(\omega t) \]

Since \( v_0 \neq 0 \) for only \( 0 < \omega t < \pi/2 \) the integration for \( A_n \) becomes

\[ A_n = \int_{0}^{\omega t_1} [a(\omega t) \cos \omega t] \, d(\omega t) \]
\[ + \int_{\omega t_1}^{-\frac{\pi}{2}} [26M \sin \omega_n t \cos \omega t] \, d(\omega t) \]
Equation 4-10

In order to solve for \( A_n \), the integration limit, \( \omega t_1 \), must be found as a function of the input.

This time occurs when
\[ a(\omega t_1) = 26M \sin(\omega t_1) \]

Equation 4-11

This equation cannot be solved for \( t_1 \) analytically.
Thus a promising route for analysis grinds to a halt.

At this point it was thought that a digital computer model should be the next step. The model might provide useful insights into an alternative approach for obtaining a transfer function for the controller, as well as provide the necessary translation for the controller if the entire motor speed control system were to be modelled on a computer.

Appendix B contains the digital simulation for the motor controller. The program was written such that a variety of parameters could be selected by the user to determine their effect on the controller operation.
Inputs are limited to functions of time, \( u = f(t) \), for example \( u = \sin t \) or \( u = (a) \exp(-6t) \). Deterministic as well as random functions could be used. A provision is made to choose either a one-slope model, where the ramping rate is constant over the entire output range, or a two-slope model where the ramping rate can be one value over part of the output range and a second value for the rest of the range. When using the two-slope model a
slope changeover point (SLOCHA) must be specified in the output range. Finally, the initial input voltage must be specified, \( u(0) \). The model assumes that a steady-state output has been reached for \( u(0) \) before new values are calculated for succeeding time intervals. Using the digital model, with appropriate input values, resulted in a close approximation to the step input data in Fig. 4.14.

Once the digital model has been formed and verified accurate, a variety of inputs can be applied and the output analyzed. Consider using as an input a sine wave of the form \( u(t) = 5\sin \omega_n t \). A function of this form would describe an output limit function as \( v_{\text{max}} = 130\sin \omega_n t \). If \( \omega_n \) is small enough, the output could be expected to ramp as fast as the input is changing and the controller would act as an amplifier with a gain \( K_c = 26 \). However, for higher values of the output would not be able to increase as fast as the limit function, \( v_{\text{max}} \). In this case the output would ramp up until the limit function had passed its peak and had reached a value where \( v_o(t) = v_{\text{max}}(t) \). At this point the output will follow the limit function back to zero. Of course for the second half of the cycle, \( u(t) < 0 \), the output will be zero. This discussion is illustrated in
Fig. 4.16. It is interesting to note how magnitude and phase of the output change as a function of frequency. The magnitude remains constant until a frequency is reached where the controller cannot ramp up as fast as the input. At higher frequencies the maximum value of the output becomes less and less as the frequency increases. Table 4-2 contains the actual values of $v_o^{(\text{max})}$ as a function of frequency for the input $u(t)=5\sin \omega_n t$. For $\omega_n>1$ the magnitude appears to roll off with increasing frequency at a rate of $-10\text{dB/decade}$ for $1<\omega_n<10$ and at a rate of $-25\text{dB/decade}$ for $10<\omega_n<100$.

Table 4.2
Maximum Output Values and Phase Shift of the Controller for an Input $u(t)=5\sin \omega_n t$.

<table>
<thead>
<tr>
<th>$\omega_n$</th>
<th>$v_o^{(\text{max})}$</th>
<th>Gain(dB)</th>
<th>phase shift</th>
</tr>
</thead>
<tbody>
<tr>
<td>.1</td>
<td>130</td>
<td>28.3</td>
<td>0</td>
</tr>
<tr>
<td>.3</td>
<td>130</td>
<td>28.3</td>
<td>0</td>
</tr>
<tr>
<td>.5</td>
<td>130</td>
<td>28.3</td>
<td>0</td>
</tr>
<tr>
<td>1.0</td>
<td>130</td>
<td>28.3</td>
<td>0</td>
</tr>
<tr>
<td>2</td>
<td>113</td>
<td>27.1</td>
<td>27</td>
</tr>
<tr>
<td>2.5</td>
<td>94.6</td>
<td>25.5</td>
<td>40.5</td>
</tr>
</tbody>
</table>
As the frequency, $\omega_n$, increases, it is apparent that the maximum value of the output shifts in relation to the input maximum. At low frequencies the output peak occurs at the same time as the input peak, as shown in part (a) of Fig. 4.16. But as the frequency increases, the output peak lags the input peak (Fig. 4.16(b)&(c)). Finally, at high frequencies the output peak lags the input by almost 90° (Fig. 4.16(d)). Phase shift in this sense is listed in Table 4.2. This is quite analogous to the phase shift produced by a simple pole at $\omega_n=1$. But the analogy breaks down because the phase shift is reduced to zero at $\omega_n=\pi/2$ since both the input and the output reach zero at the same instant. Furthermore, for the negative half-cycle of the input, the output remains at zero. This nonlinear behavior will undoubtedly complicate the

<table>
<thead>
<tr>
<th></th>
<th>(a)</th>
<th>(b)</th>
<th>(c)</th>
<th>(d)</th>
</tr>
</thead>
<tbody>
<tr>
<td>3</td>
<td>78.7</td>
<td>23.9</td>
<td>49.5</td>
<td></td>
</tr>
<tr>
<td>5</td>
<td>41.8</td>
<td>18.4</td>
<td>67.5</td>
<td></td>
</tr>
<tr>
<td>10</td>
<td>22.2</td>
<td>12.9</td>
<td>76.5</td>
<td></td>
</tr>
<tr>
<td>30</td>
<td>7.77</td>
<td>3.8</td>
<td>87.5</td>
<td></td>
</tr>
<tr>
<td>50</td>
<td>4.65</td>
<td>-0.6</td>
<td>87.5</td>
<td></td>
</tr>
<tr>
<td>100</td>
<td>2.34</td>
<td>-6.6</td>
<td>87.5</td>
<td></td>
</tr>
</tbody>
</table>
actual response with harmonics of the fundamental components.

However, the data contained in Table 4-2 can be used as a worst-case analysis for the controller transfer function. The harmonics generated by the nonlinearity can largely be ignored from a systems viewpoint since the motor/pump will act as a low-pass filter and remove them from the system response. The magnitude data in Table 4-2 is the actual magnitude response of the controller and is therefore accurate. And in the case of the phase response, negative values for the input are not allowed in the system. Therefore, the $90^\circ$ phase shift occurring at the higher frequencies could be considered worst-case since it is always corrected to $0^\circ$ when the input reaches zero.

With these assumptions in hand, a Bode diagram can be constructed for the linear portion of the system and the response of the controller can be added in to give the overall system response as a function of frequency. This analysis was done and the results are shown in Figures 4.17 - 4.20.

The transfer function of the tachometer can be easily derived. A speed input of 3000 rpm to the tachometer provides an output of 5.6 volts. Hence,
Fig. 4.17. Bode diagram for PTO/HYD system at 500 rpm.
Fig. 4.18. Bode diagram for PTO/HYD system at 500 rpm.
Fig. 4.19. Bode diagram for PTO/HYD system 1500 rpm.
Fig. 4.20. Bode diagram for PTO/HYD system at 2000 rpm.
\[ H_T(s) = \frac{5.61}{3000} = 0.0019. \]

Equation 4-12

4.5 Stability Analysis and Optimization

Figures 4.17-4.20 illustrate the overall system response as well as the response due to only the linear portion. In the magnitude plots the upper curve is the system response and the lower curve is the response of the linearized portion. With the phase plots, the upper curve denotes the response of the linearized portion. Several observations are possible. The linear portion is unconditionally stable since its gain is always less than 0dB. This was verified experimentally. When the nonlinear portion of the system, the controller, is added in, the gain for \( \omega < 3 \) is greater than 0dB but the phase shift does not reach 180° until \( \omega > 10 \) which results in a stable system. In fact, the gain and phase margins are about 20dB and 100° respectively. From these observations, one approach to optimizing the system would be to increase the DC gain by 10dB to reduce the response time and steady state error.

A method proposed by Popov for determining the stability of a nonlinear system is applicable to this
system. The type of system under consideration with this method, as shown in Fig. 4.21, consists of a nonlinear portion, N, and a linear portion, G, with zero input. N is a time-invariant, memoryless nonlinearity characterized by a transfer function such that

\[ 0 < \xi \leq \frac{\phi(\xi)}{k} \leq k - \xi \]

which places an upper bound, k, and a lower bound, \( \xi > 0 \) on the gain. G is also time-invariant and nonanticipative. Its zero-input response is bounded and continuous for any initial state, and its impulse response tends to zero as \( t \) approaches infinity. If these conditions are met and there exists a real number \( q > 0 \) and a positive \( \delta \) such that

\[ \text{Re}[(1 + qj\omega)G(j\omega) + 1/k] \geq \delta > 0 \quad \text{for all} \quad \omega > 0 \]

then for any initial state the zero-input response is bounded and tends to zero as \( t \) approaches infinity. This relationship implies that a Nyquist plot of \( G(j\omega) \) must lie to the right of a line given by

\[ \text{Re} \, G(j\omega) = -(1/k) + \omega q \text{Im} \, G(j\omega). \]

The slope of this line is frequency dependent. Using a
Fig. 4.21. System considered in Popov's method.

Fig. 4.22. Nyquist plot of PTO/HYD system showing Popov line.
simple transformation, a plot which is not frequency
dependent can be made on a modified frequency plane.
The transformation is defined as

\[ G^*(j\omega) = \text{Re} \, G(j\omega) + j\omega \text{Im} G(j\omega) \]

and the equation of the line becomes

\[ \text{Re} G^*(j\omega) = -(1/k) + q \text{Im} G^*(j\omega) \]

which is frequency insensitive.

Using the transfer function obtained for the
motor/pump, a Nyquist plot can be made, the Popov line
determined, and the maximum value for k obtained. Fig.
4.22 illustrates the results. As the figure shows, the
Popov line can be drawn so that it intersects the real
axis arbitrarily close to the origin. The implication
is that the nonlinear gain, k, can be any value from
zero to infinity and the system will remain stable.
This result lends support to the previous transfer
function analysis which also indicated stability. It
must be remembered, however, that this result is valid
only for the zero input condition. Other methods must
be used for nonzero inputs.
Both methods presented, transfer function analysis and Popov, have shown that the system is stable. There is a difference in permissible gain allowed, however. Popov's upper limit of infinity describes a bang-bang system that switches the controller output to 130 volts for a negative speed error and 0 volts for a positive speed error. But the transfer function analysis indicates a gain margin of approximately 20dB which translates to an overall maximum controller gain of 260. Practically speaking this is also a bang-bang system since only a small speed error, about 12 rpm, would saturate the controller if it had a gain this large. From a physical viewpoint the methods are in agreement.

The preceding analysis provides a foundation from which to optimize the control loop. The present system exhibits a large steady state error for torque disturbances. This can be reduced by increasing the loop gain, however, a more satisfactory approach would be to introduce PI-type feedback. The transfer functions developed in this chapter allow the system to be modelled conveniently on a digital computer which would greatly facilitate optimization procedures.
Chapter V

THE FUTURE OF THE ELECTRIC CHOREMASTER TRACTOR

5.1 Summary

During the first week of June 1984, the Electric Choremaster became a reality. At that time the vehicle incorporated the systems discussed in Chapters I to III. Two SCR controllers were fitted to a traction motor and a PTO/HYD motor. The system utilized a direct current drive system to take advantage of the high torque capability of series-wound motors.

Several instruments were fitted to the vehicle to provide the measurements required of electric vehicles in general and also measurements unique to vehicles intended for research. System voltage and current are measured directly as well as motor and ground speed. A battery fuel gauge has been installed along with a DC kilowatt-hour meter for energy efficiency studies.

After laboratory testing and on-farm operation, it became apparent that a regulated speed control system was necessary for the PTO/HYD motor. Chapter IV described the circuitry installed. A stability study detailing analysis of nonlinear transfer functions was undertaken to insure a well behaved system.
5.2 Tractor Performance

After nine months of operation and testing, many features of the Electric Choremaster, both good and bad, have become apparent. After the vehicle had been tested in the laboratory, it was especially interesting to note how on-farm testing magnified the tractor's positive and negative characteristics.

Chore performance of the electric tractor is excellent. One reason for this is the articulated four-wheel-drive configuration. This vehicle moves in the desired direction without slipping and sliding like two-wheel-drive vehicles. There is an abundance of power available in all chore-type operations encountered. This is due in part to the large overload capability inherent in DC motors and also to the nature of the tasks performed. Another factor was the added vehicle weight imposed by the batteries which increased vehicle traction.

Performance of the Cableform controllers was also excellent. Acceleration is smooth and power from the motors is available on demand without lurching or jerking the tractor. An area in which the Electric
Choremaster is clearly superior to conventional tractors is creep speed. When a task requires extremely low speed, such as detaching the loader, the Choremaster can be made to creep as slowly as desired, contrasting sharply with conventional tractors requiring clutching to perform similarly. When full power is demanded, engagement of the bypass contactor is also smooth. Generally, the only way an operator can tell if the bypass is engaged is if he hears the engagement of the contacts. A final feature of the controller is the ability to reverse direction of travel simply by moving the traction accelerator. This virtually eliminates clutching and shifting from most tasks and is smoother and more efficient than hydrostatic drives.

Addition of the feedback speed control system of Chapter IV for the PTO/HYD motor has greatly aided loader operation. With this feature the operator can set the motor speed when loader work is necessary and not touch the control again until the task is completed. This system will also provide constant speed at the PTO which should greatly aid many chore activities. Cab instruments, with the exception of the state-of-charge meter, have performed well and allow the operator to monitor voltage, current, and speed.
Coasting is the most glaring deficiency. The only retarding force after the speed control has been backed off (but not reversed) is rolling resistance and gear train inertia. The vehicle coasts exceedingly well on firm surfaces and this is magnified on a declining surface. It is not only an unfamiliar response but a potentially dangerous one as well. To solve this problem two approaches are available: 1) regenerative braking and 2) plugging to produce a torque to retard motion while the vehicle is in neutral. Both of these approaches would require additional and perhaps complex circuitry.

Even though constant speed has been attained for the PTO/HYD system, the problem is still not completely solved. Ideally, during loader operation the motor should increase speed only when the loader is operated, but should remain at a low speed adequate for power steering needs when the loader is not in use. This would further optimize the efficiency of the tractor. Addition of this function would require extra control circuitry but should be fairly easy to realize.

Another deficiency is the state of charge meter or battery fuel gauge. The instrument obtained has its accuracy based on battery discharge following the rated
discharge curve of the battery. This cannot be guaranteed to any extent due to the varied nature of chores performed.

Finally, a deficiency inherent in any electric vehicle is the limited energy capacity of the battery. The Electric Choremaster was designed with the goal of performing 4-8 hours of chore work between charges. Once again, the varied nature of the chores determines whether or not the tractor can perform for this period of time before recharging becomes necessary. Battery technology is progressing, however, and batteries with 50% more capacity on a weight basis are currently available. Solution of this basic problem would insure the success of electric vehicles.

In conclusion, it has been shown that a control system for an electric farm vehicle can be developed with satisfactory performance using current technology. A unique method of nonlinear system analysis has been developed to allow optimization of closed loop speed regulation. Instrumentation has also been implemented with acceptable results in most areas.
5.3 Future Development

There are a number of areas which need to be addressed further. The unique nature of this project offers applied research which differs from much of what is currently being conducted. Design considerations for over-the-road vehicles, tow tractors, and mining vehicles decidedly differ from electric farm tractors. Another aspect is that chore work is quite varied while other electric vehicles are designed primarily for one task. In summation, producing an electric farm vehicle is not a simple procedure, but rather involves optimization of many subsystems to produce a tractor capable of acceptable performance in all areas.

Optimization of the PTO/HYD system is a primary area for further work. The entire system is impractical because separate control of PTO and hydraulics is impossible with the present system. Applying separate controllers and motors to each would be one solution. Once the hydraulics are independent, other means of control become possible such as hydraulic accumulators, pressure control, etc. Even with the current system, the feedback network can be optimized to reduce steady-state error and response time. Modifying the
system so that motor speed is a function of hydraulic requirements was mentioned in the previous section.

Since virtually all tractors have a governor, a feedback system to provide constant speed for the traction motor should be implemented. This function is not as critical for the chore tractor as it is for the field tractor due to the stop-and-go nature of chore routines. However, there are occasions when constant speed control enables the operator to perform a task more efficiently and with greater safety. Because the traction motor reverses rotation and is subject to braking pulses, constant speed control becomes somewhat more difficult to implement. The Cableform controller can be used to advantage, since when braking is detected, it overrides the normal acceleration ramps. Thus, during this period it would ignore any large error signals produced until all motion has stopped. Immediately following, the vehicle would accelerate normally in the opposite direction if the operator so desired. An additional aid is that the controller function is the same for both drive systems and its response has already been analyzed.

The problem of coasting was mentioned in the preceding section. Solution of this problem should be
given high priority since it directly affects safe vehicle operation. Developing a control that would apply a retarding torque of proper magnitude to the motor whenever vehicle speed is greater than the desired speed is the key to addressing this deficiency.

Accurately determining remaining battery capacity during operation of the Electric Choremaster is not currently possible. A detailed review of the literature and current techniques in this area is in order. Perhaps an alternative approach would be to directly measure energy into and out of the battery.

Advanced research and development is being done in the area of separately-excited motors and controllers. Indications are that better performance is possible over a larger range of operating speeds with increased efficiency. More precise control of the motor is obtained since field and armature function independently, but complexity of the controller is increased. Retaining SCR control of the large armature current but applying the field current through transistor choppers seems to be a viable approach. Development in this area would most likely require replacing the current motors and controllers in the Electric Choremaster.
A final area for development is the use of a microprocessor to supply all logic functions for the vehicle. Virtually every area of control and instrumentation could be regulated by a single processor with appropriate support components. A partial listing might include:

1. Control of motor choppers, including separately excited systems.
2. Constant speed regulation.
3. Braking application.
4. Optimal hydraulic motor control.
5. Optimal motor condition for greatest efficiency.
6. Battery state of charge measurement.
7. System energy usage and efficiency.
8. Data acquisition during vehicle testing.

Added performance, however, is acquired by added complexity. Careful consideration of the tractor's operating environment would be required before a microprocessor system could be installed.
5.4 A Future for Electric Farm Tractors

As discussed in Chapter I, the primary purpose of this project is to determine the feasibility of using electric tractors on the farm. Obviously the question cannot be answered completely after testing one research vehicle. Even so, some observations can be made.

From the view point of electric control, electric tractors can be made to operate as well or better than conventional tractors in most areas. Generated power is comparable; operator controls are better; speed control is comparable. Electric motors run more efficiently, are quieter, are capable of performance equalling gasoline or diesel engines, and are generally smaller. Batteries are the greatest drawback to electric farm vehicles. The energy capacity per unit weight is not comparable to conventional sources. As a result, electric tractors are capable of only extended light duty work or short periods of work with large power requirements. Thus, they are practical only for specialty tasks.

The Electric Choremaster configuration is based on a general purpose commercially available tractor. Converting it to electric power produced a specialty
vehicle and therefore its use is restricted. Farmers today cannot buy an expensive special purpose vehicle if they cannot use it enough to justify the purchase. It is much more cost effective to spend the money on a general-purpose vehicle which can be used continuously. Electric farm vehicles can find greatest acceptance when designed for special purposes that emphasize the advantages of electric motive power.
APPENDIX B

Contained in this appendix is a description of a digital simulation for the Cableform controller transfer function where the input is a time-varying bounded voltage, $0 \leq v_m < 5$. The output is an average voltage applied to the motor, $0 \leq v_o \leq 130$. Basic controller operation is as follows. If the output is less than $26 \times v_m$, it will ramp up at a steady rate until $v_o = 26 v_m$. If the output is greater than $26 \times v_m$, it will immediately drop to $v_o = 26 v_m$ and follow any further negative excursion. Additionally, the slope changes when $v_o = 45$, becoming greater for $v_o > 45$. This is illustrated in Fig. 4.14. The simulation allows the user to choose either a one-slope or two-slope model.

Inputs and outputs for the simulation are listed below.

1. STEPP = step size in time (seconds)
2. INTER = number of iterations
3. U = input function
4. MODEL model selection
MODEL = 1 for 1-slope model
MODEL = 2 for 2-slope model

5. SLOPE  slope of ramp (volts/second)
   SLOPE1 = 1 for 1-slope model
   SLOPE2 = 2 for 2-slope model

6. SLOCHA  slope change over point for 2-slope model

7. U initial input voltage (line 100)

OUTPUTS:

1. Y(KT) output voltage
2. KT time in seconds
3. U(KT) input voltage

A program listing, written in BASIC, implementing the simulation is shown below. The values shown are those used in Chapter IV to derive the controller transfer function. An output listing for this program is shown in Fig. B 1.1. The input in this example is u(t)=5sin(2\pi t), in line 160, and the values obtained were rounded off as shown. Notice that although negative inputs are not allowed physically the simulation will handle them appropriately. This allows
easy formulation of input functions.

This simulation could be easily used as a subroutine for the controller portion in calculating overall system response using state variable analysis on a digital computer.

Program listing for digital model of the Cableform Controller:

10 REM DIGITAL MODEL OF THE CABLEFORM CONTROLLER
20 REM INPUT VARIABLES
30 STEPP = .125
40 ITER =8
50 MODEL =2
60 SLOPE1 = 76.7
70 SLOPE2 = 158
80 SLOCHA = 44
90 REM ENTER INITIAL INPUT VALUE
100 U =0
110 PRINT "TIME", "U(KT)", "Y(KT)"
120 Y = 26 * U
130 PRINT 0,U,Y
140 FOR K=1 TO ITER
150 REM ENTER INPUT FUNCTION U=F(KT)
160 U = 5 * SIN (6.28 * K * STEPP)
170 RAMLIM = 26 * U
180 IF RAMLIM < 0 THEN 190 ELSE 200
190 RAMLIM = 0
200 IF Y <= RAMLIM THEN 210 ELSE 310
210 IF MODEL = 1 THEN 220 ELSE 260
220 Y = Y + (SLOPE1 * STEPP)
230 IF Y > = 130 THEN 240 ELSE 250
240 Y = 130
250 GOTO 350
260 IF Y < SLOCHA THEN 220 ELSE 270
270 Y = Y + (SLOPE2 * STEPP)
280 IF Y > 130 THEN 290 ELSE 250
290 Y = 130
300 GOTO 350
310 Y = RAMLIM
320 IF Y < 0 THEN 330 ELSE 340
330 Y = 0
340 GOTO 370
350 IF Y > RAMLIM THEN 360 ELSE 370
360 Y = RAMLIM
370 PRINT K * STEPP, U, Y
380 NEXT K
390 STOP
<table>
<thead>
<tr>
<th>TIME</th>
<th>U(KT)</th>
<th>Y(KT)</th>
</tr>
</thead>
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<td>0</td>
</tr>
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<td>.125</td>
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</tr>
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<tr>
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</tbody>
</table>

Fig. B 1.1 Output Listing for Simulation Program
BIBLIOGRAPHY


