MODEL ALGORITHMIC CONTROL : AN APPLICATION TO DC MOTOR SPEED CONTROL

by

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ABSTRACT

The growth of digital technology in the last few years has represented a challenge to automatic control research workers.The increasing use of digital minicomputers has considerably helped the industrial implementation of algorithmic types of control.For this type of control to be practical and efficient,it is imperative for the controller to have access to fast digital computing facilities with fast memory access and substential information storage capacity.

In this thesis a new method of digital process control which is applied to a number of industrial processes ranging from power plants to glass furnaces,is described.This new control method is called Model Algorithmic Control or equivalently Model Predictive Heuristic Control.A mathematical framework for the analysis of Model Algorithmic Control is developed and the operations of the main components of the control structure are described.

The Model Algorithmic Control strategy relies on three principles:

i.the plant is represented by its impuls response which will be used on-line by the computer for prediction,

ii. the behavior of the closed-loop system is prescribed by means of reference trajectory initiated on the actual output,

iii. the control variables are computed in a heuristic way. Future inputs are computed in such a way that when applied to the internal predictive model, it induces outputs as close as possible to the desired reference trajectory.

In this thesis, the Model Algorithmic Control is applied to dc motor speed control. Control algorithm is evaluated by a Z-80 based microcomputer, and the output of microcomputer is applied to a fully controlled thyristor converter unit as a speed reference input through a digital to analog converter. The actual speed of the motor is inputed to the microcomputer by means of a tachogenerator through an analog to digital converter.

The application of Model Algorithmic Control to dc motor speed control is devided in two parts. In the first part of the application, the set speed is constant, the algorithm causes the speed of the motor to reach that constant speed with a time constant similar to the reference trajectory. The time responses of motor output speed are observed for different time constants. In the second part, the set speed is time varying. In this part the set speed is changed as a linear ramp, the tracking behavior of the dc motor output speed is observed.

In chapter one, the methods for the speed control of a separately excited direct current motor are described and compared with each others. Consequently, the equations determining the dynamical behavior of the motor and the transfer function of the motor are derived.

In chapter two,solid state dc motor drives are described and the operation of a three phase full converter is explained in general terms.The features of the thyristor unit used in the realization of the control is then presented.

In chapter three,a mathematical framework for the analysis of Model Algorithmic Control is developed and the operations of the main components of the control structure are described.The equations to calculate the optimum inputs are derived.

In chapter four,the model which will be used in the algorithm is introduced.The discrete impuls response of the system is found and the overall set up for closed-loop control is given.Additionally the software developed is presented and explained.

In chapter five,the stability analysis of the Model Algorithmic Controlis developed.The robustness of the system to the parameter changes is discussed.Finally the effects of the constraints on input sequence to system stability are discussed,and the experimental results are presented.

ÖZETÇE

Geçmiş yıllarda sayısal teknolojideki önemli gelişmeler otomatik denetim araştırmacılarına yeni ufuklar açmıştır. Mikroişlemcilerin gittikçe artan kullanılım alanları, algoritma tipi denetim yöntemlerinin endüstride uygulanmalarına olanak sağlamıştır.Bu tip bir denetim yönteminin başarılı ve verimli olabilmesi için denetimci mikroişlemcinin hızlı sayısal hesaplama ve hafızayı en etkili ve hızlı bir şekilde kullanabilme yeteneğine sahip olması önkoşuldur.

Bu tez çalışmasında, güç üretici dizgelerden cam işlemcili-gine varana kadar pek çok endüstri uygulamalarında başarı ile denenmiş yeni bir denetim yöntemi tartışılmıştır. Bu yeni yönteme Model Algoritmasal Denetim veya Model Tahminle Deneyimsel Denetim adı verilmiştir. Model Algoritmasal Denetim yönteminin analizi için teori formule edilmiş, denetim yönteminin temel elemanlarının işlevleri açıklanmıştır.

Model Algoritmasal Denetim yöntemi üç temel prensibe dayandırılmıştır:

i. dizge, mikroişlemci tarafından dizge çıkışının tahmininde kullanılmak üzere, dürtü tepkesi ile tanımlanmıştır,

ii. kapalı döngü dizgenin davranışı, dizge çıkışı ile başlatılan bir referans izlence ile tanımlanır,

iii. denetim değişkeni deneyimsel bir yolla hesaplanmaktadır. Gelecekteki dizge girdileri, dizge çıkışını referans izlenceyi çok yakından izlemeye zorlayacak şekilde hesaplanmaktadır.

Bu tezde Model Algoritmasal Denetim yöntemi bir doğru akım motorunun hız denetimine uygulanmıştır. Denetim algoritması Z-80 tabanlı bir mikroişlemci tarafından uygulanmakta, mikroişlemci çıkışı bir sayısal analog çevirici ile tam denetimli bir tristör doğrultucunun hız referans girişine uygulanmaktadır. Motorun ansal hızı ise mikroişlemci tarafından bir analog sayısal çevirici yardımı ile takojeneratör çıkışının okunması ile elde edilir.

Model Algoritmasal Denetim bu uygulaması iki böyümde gerçekleştirilmiştir. Birinci bölümde ulaşılacak hız sabittir ve algoritma motor hızını bu sabit hıza, referans izlencenin zaman sabiti ile aynı olacak şekilde, ulaştıracak biçimde çalışır. Motor hızının zaman tepkesi değişik izlence zaman sabitleri için gözlemlenir. İkinci bölümde ise ulaşılacak hız zamanla değişkendir. Bu uygulamada ulaşılacak hız bir yokuş olarak alınmış ve motor hızının bu yokuş işlevini takip edişi gözlemlenmiştir.

Tezin birinci bölümünde alan uyarımlı doğru akım motorunun hız denetim yöntemleri tartışılmış, motorun dinamik davranışı incelenmiş, ve geçiş işlevi çıkarılmıştır.

İkinci bölümde katı hal doğru akım motor sürücüleri tanıtılmış ve genel olarak üç evreli tam denetimli doğrultucunun çalışması açıklanmış, ayrıca tezin uygulanmasında kullanılan tristör ünitesi tanıtılmıştır.

Üçüncü bölümde ise Model Algoritmasal Denetim yönteminin analizi için gerekli matematik çalışma yapılmış, temel denetim elemanlarının işlevlerine açıklık getirilmiş ve optimum dizge girişlerinin hesaplanması tartışılmıştır.

Dördüncü bölümde, doğru akım motorunun matematiksel modeli elde edilmiş, bu model kesikli dürtü tepkesinin bulunmasında kullanılmıştır.

Son olarak, dizgenin kararlılığı ve parametrelerdeki sapmalara karşı duyarlılığı tartışılmıştır. Ayrıca dizge girişindeki kısıtlamaların kararlılığa etkileri açıklanmıştır.

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LIST OF SYMBOLS

B	Friction coefficient
e_a	Motor armature voltage
e_g	Motor generated voltage
i_a	Motor current
J	Moment of inertia
K_a ,K_t	Dc machine constants
K_s	Speed controller gain
K_t	Tachogenerator gain
L_a	Motor armature circuit inductance
N	Averaga motor speed
R_a	Motor armature circuit resistance
T	Torque
T_L	Load torque
ω	Angular frequency
ϕ	Flux

1.DIRECT CURRENT MOTOR CHARACTERISTICS

The direct current (dc) motor is one of the first machines devised to convert electrical power into mechanical power.Many dc machines were built in the 1880's when dc was the principal form of electric power generation.With the advent of 50 Hz alternating current (ac) as the electric power standard,and invention of the induction motor with its lower manufacturing costs,the dc machine became less important.In recent years,the use of dc machines has become almost exclusively associated with applications where the unique characteristics of the dc motor (e.g.high starting torque for traction motor applications,high speed variation both above and below the rated speed) justify its cost.

The flexibility of the dc motor for speed control,its overload capability,and its reliability make it the dominant means of providing a controllable source of mechanical rotating power in industry.The steady-state characteristics of the motor determine how the motor can be controlled;the dynamic characteristics determine how the control system must be designed to obtain the required response.

A. Speed Control of a Separately Excited DC Motor

The equivalent circuit of a separately excited dc motor shown in Fig. 1.1 is equal to a resistance, R_a (total armature circuit resistance) in series with a parallel conbination of an inductance, L_a (total armature circuit inductance), and another resistance R_1. The resistence R_1 will not be involved in calculations. When the motor rotates the armature coil move in the stator magnetic field. The induced e.m.f. appears across the armature terminals as internally generated voltage, e_g. Therefore the equivalent electric circuit of the motor is the impedance connected in series to a voltage source e_g.

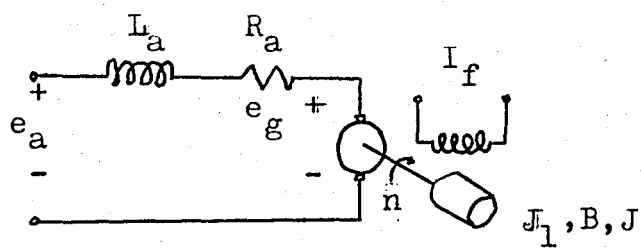


Fig. 1.1. Equivalent circuit of a sep.exc. dc motor.

The physical explanation for this model is that R_1 represents the losses in the magnetic circuit. The resistance R_1 is usually larger than R_a (typically about 5-10 times), and since the effect of R_1 on the motor operation is insignificant, it is possible to ignore this resistance in most practical application.

The steady state operation of the motor is governed by three equations. The total armature circuit voltage is

$$e_a = e_g + i_a R_a \qquad (1.1)$$

The generated armature voltage is given by

$$e_g = K_a \phi_f N \qquad (1.2)$$

And, the internal torque is

$$T = K_t \phi_f I_a \qquad (1.3)$$

The torque constant K_t ,and armature voltage constant K_a are equal in a consistent set of units.The solution of above three equations for speed gives

$$N = \frac{e_a - R_a i_a}{K_a \phi_f} \qquad (1.4)$$

Equation (1.4) shows that the speed of the motor can be controlled in three ways by changing;

i. the armature voltage e_a ,which is nearly proportional to speed

ii. the magnetik field ϕ_f ,which is inversly proportional to speed

iii. the armature circuit resistance R_a ,which is proportional to the speed drop.

The control of speed by the armature circuit voltage e_a abbreviated to armature voltage control,is the most desirable because the magnetic field can remain at full amplitude and full torque can be developed.Field control is accomplished by reducing the field current and is used to extend the speed above the value for full armature voltage.In this case the available torque decreases.Armature circuit resistance control is not practical except in very small motors because of the power dissipation.

The characteristics for a full range of speed are shown in Fig. 1.2.Armature voltage control raises the speed N_1 ; field control raises it additionally to speed N_2. The available torque at rated armature current is maximum to speed N_1 and drops with the field to speed N_2. The horse power increases linearly to speed N_1 and remains constant to speed N_2.

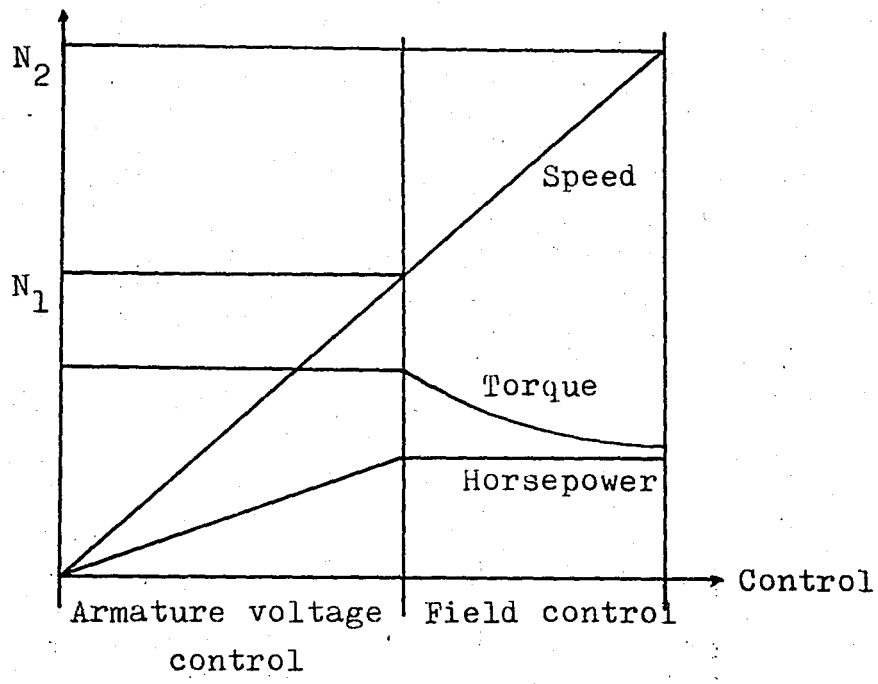


Fig. 1.2 Speed control range of sep.exc. motor under armature voltage and field control

B.Dynamics of a Separately Excited DC Motor

The dynamic behavior of the motor is governed by its energy storage properties. Energy is stored in two places: the magnetic fields in the magnetic circuit and by mechanical velocity in the inertia of the armature. Of course,energy is also stored in the inertia of the load and in magnetic fields of the electrical sources and must be included in tne consideration of the dynamics.

Consider a separately excited dc motor with armature voltage control as shown in Fig. 1.1. The voltage loop equation is

$$e_a = e_g + R_a i_a + L_a \frac{di_a}{dt} \tag{1.5}$$

where

$$e_g = K_a \phi_f n \tag{1.6}$$

The torque balance equation is

$$T_e = T_L + Bn + J \frac{dn}{dt} \tag{1.7}$$

where

$$T_e = K_a \phi_f i_a \tag{1.8}$$

In the Laplace domain, equations (1.5) through (1.8) can be written as

$$E_a(s) = E_g(s) + R_a I_a(s) + L_a s I_a(s) \tag{1.9}$$

$$E_g(s) = K_a \phi_f N(s) \tag{1.10}$$

$$T_e(s) = T_L(s) + BN(s) + JsN(s) \tag{1.11}$$

$$T_e(s) = K_a \phi_f I_a(s) \tag{1.12}$$

From equation (1.9)

$$I_a(s) = \frac{E_a(s) - E_g(s)}{R_a + sL_a} = \frac{(E_a(s) - E_g(s))/R_a}{1 + T_a s} \tag{1.13}$$

where $T_a = L_a/R_a$ is electrical time constant of the motor armature circuit.

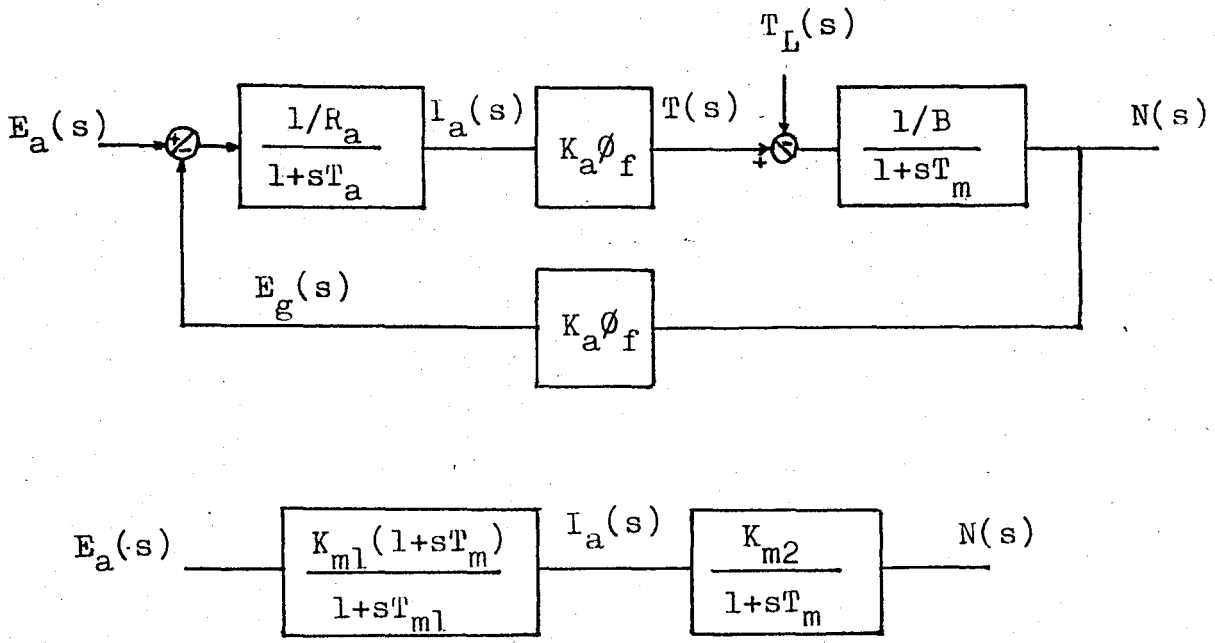


Fig. 1.3 Development of motor transfer function
(a). Functional block diagram
(b). Simplified transfer function

From equation (1,11)

$$N(s) = \frac{(T_e(s)-T_L(s))/B}{1 + T_m s} \qquad (1.14)$$

where $T_m = J/B$ is mechanical time constant of the motor.

These relationships are shown in block diagram in Fig. 1.3.a The feedback loop present in the form of the back e.m.f.This back e.m.f. provides the moderate speed regulation inherent in the separately excited dc motor.

Neglecting the load torque, the relation between the speed and armature voltage is found to be

$$\frac{N(s)}{E_a(s)} = \frac{K_a \phi_f}{(K_a\phi_f)^2 + R_a B(1+sT_a)(1+sT_m)} \qquad (1.15)$$

Electrical time constant T_a is always less than the mechanical time constant T_m, then T_a can be neglected and the expression simplifies to

$$\frac{N(s)}{E_a(s)} = \frac{K_a\phi_f}{(K_a\phi_f)^2 + R_a B + sR_a BT_m} = \frac{K_m}{1 + sT_{ml}} \qquad (1.16)$$

where

$$T_{ml} = \frac{R_a B}{(K_a\phi_f)^2 + R_a B} T_m \qquad (1.17)$$

$$K_m = \frac{K_a\phi_f}{(K_a\phi_f)^2 + R_a B} \qquad (1,18)$$

Thus the motor can be represented, for the purpose of analysing it for voltage control, as a first order system as in Fig. 1.4

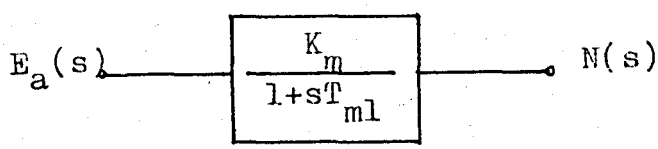


Fig. 1.4 Simplified transfer function as a block diagram

C.Closed Loop Speed Control

If a tachogenerator is attached to the motor shaft,a speed signal which is proportional to the actual one can be fed back and the error $E_n(s)$ used to control the armature voltage.This closed loop control scheme is shown in Fig.1.5.

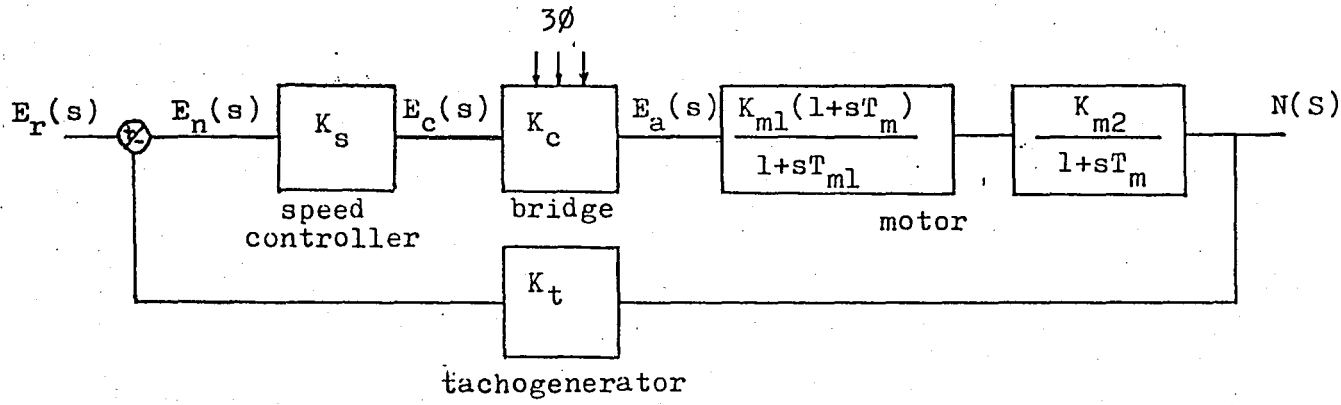


Fig. 1.5 Closed-loop Speed Control

From the block diagram shown in Fig. 1.5,the relation between the actual speed and the error is found to be

$$\frac{N(s)}{E_r(s)} = \frac{K_1}{1 + sT_1} \tag{1.19}$$

where

$$K_1 = \frac{K_s K_c K_{m1} K_{m2}}{K_s K_c K_{m1} K_{m2} K_t + 1} \tag{1.20}$$

$$T_1 = \frac{T_{m1}}{K_s K_c K_{m1} K_{m2} K_t + 1} \tag{1.21}$$

$$K_{m1} = \frac{B}{(K_a \phi_f)^2 + R_a B} \quad , \quad K_{m2} = \frac{K_a \phi_f}{B} \tag{1.22}$$

Since $K_s K_c K_{m1} K_{m2} K_t \gg 1$

$$K_1 = \frac{1}{K_t} \tag{1.23}$$

$$T_1 = \frac{T_{m1}}{K_s K_c K_{m1} K_{m2} K_t} \tag{1.24}$$

Also from the block diagram above

$$\frac{I_a(s)}{E_r(s)} = \frac{N(s)}{E_r(s)} \frac{I_a(s)}{N(s)} = \frac{K_I}{K_{m2}} \frac{(1+T_m s)}{(1+T_1 s)} \tag{1.25}$$

The current response to a step change in input $E_r(s)$ is

$$I_a(s) = \frac{K_1 E_r}{K_{m2} s} \frac{(1+T_m s)}{(1+T_1 s)} \tag{1.26}$$

We can write the above equation in time domain as

$$I_a(t) = \frac{E_r K_1}{K_{m2}} \left(1 + \frac{(T_m - T_1)}{T_1} \exp(-t/T_1) \right) \tag{1.27}$$

Since $T_m \gg T_1$, T_1 can be neglected. Normalizing the current with respect to the steady state $I_a(\infty)$

$$\frac{I_a(t)}{I_a(\infty)} = 1 + \frac{T_m}{T_1} \exp(-t/T_1) \qquad (1.28)$$

Therefore it can be concluded that a step input voltage results in a large sudden change in armature current which decays slowly. This transient overcurrent is undesirable from the standpoint of converter ratings and protection.

II.SOLID STATE DC MOTOR DRIVES

Direct current (dc) motor drives are extensively used in industry all over the world. The outstanding advantages of dc drives such as ease of control,precise and continuous control of speed over a wide range and speed of response,will ensure their popularity in the future.

Thyristors (silicon-controlled rectifiers,SCRs) have revolutionized the art of speed control of drives. In the speed of variable-speed drives,a dc motor controlled by a thyristor converter is still a popular choice. In recent years,attempts have been made to develop ac drives. However,ac drives still do not compete seriously with dc drives when variable speed is required.Although ac motors are smaller in size,and require minimal maintenance, control of their speed is relatively complex and expensive.Thyristor controlled dc drives range in rating from fractional horsepower to several thousand horsepower.

The thyristor controlled dc drives have the following advantages:

i.the thyristor-power unit eleminates the electrical time lag of the field and the armature.Time response is therefore limited only by dc motor commutating ability and the inertia of the motor,

ii. the operation is simple and reliable,

iii. minimal maintenance is required,

iv. operating efficiency is high, above 95%, because of the relatively low losses in SCRs,

v. small size, packaging flexibility, lower initial cost, and lower installation and operating costs.

The thyristor drives have the following disadvantages:

i. the high ripple content of the converter output adds to motor heating and commutation problems. The addition of a reactor in the armature circuit may be required to smooth out the ripple current,

ii. under certain operating conditions, the power factor in the ac supply is low,

iii. the overload capability is low,

iv. distortion of the ac supply voltage and telephone interface may be produced due to the switching action of SCRs,

v. regeneration requires a complex converter control.

A. Phase-controlled DC Drives

In phase-controlled dc drives an ac to dc phase-controlled converter is used to control the dc drive motor. The converter changes the ac input voltage to a controllable dc output voltage. In such converters thyristor commutation is easily achived by a process referred to as naturel commutation. When an incoming thyristor is turned on, it immediately reverse biases the outgoing thyristor and turns it off. No additional circuitry is required for the commutation process. Phase-controlled converters are

therefore simple and less expensive and are extensively used in industries.

Phase-controlled converters can be classified as semi-converters and full-converters. Semi-converters are one-quadrant converters, that is, they have one polarity of voltage and current at the dc terminals. Full-converters are two-quadrant converters in which voltage polarity can reverse, but current remains unidirectional because of the unidirectional thyristors.

B. Three Phase Full-Converter Operation

Large-horsepower dc drives take power from three phase sources. In such drives, the drive motor is controlled by three phase phase-controlled converters. The ripple frequency of the motor terminal voltage is higher than any other converters. Consequently filtering requirements for smoothing out the motor current are less. The motor current is mostly continuous, and therefore the motor performance is better than other converters.

The three phase full-converter is shown in Fig. 2.1. It uses

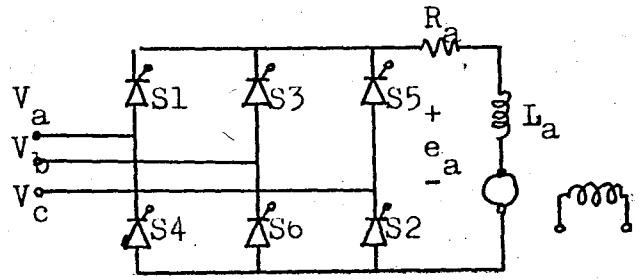


Fig. 2.1 Three phase full-converter

six thyristors and no diodes except for freewheeling operation. The thyristors are fired in sequence every 60° and the ripple is

always six pulses per cycle. The full wave converter produces less ripple than the three phase semi-converter,and is primarily used where regeneration is required. If the polarity of the motor armature voltaga is reversed with a reversing connector or by reversing the field current,then the three phase full-converter can be operated as an inverter to transfer power from the motor to the line and finally to accelerate the motor to its required reverse speed. The voltage and current waveforms of full-converter are shown in Fig. 2.2.

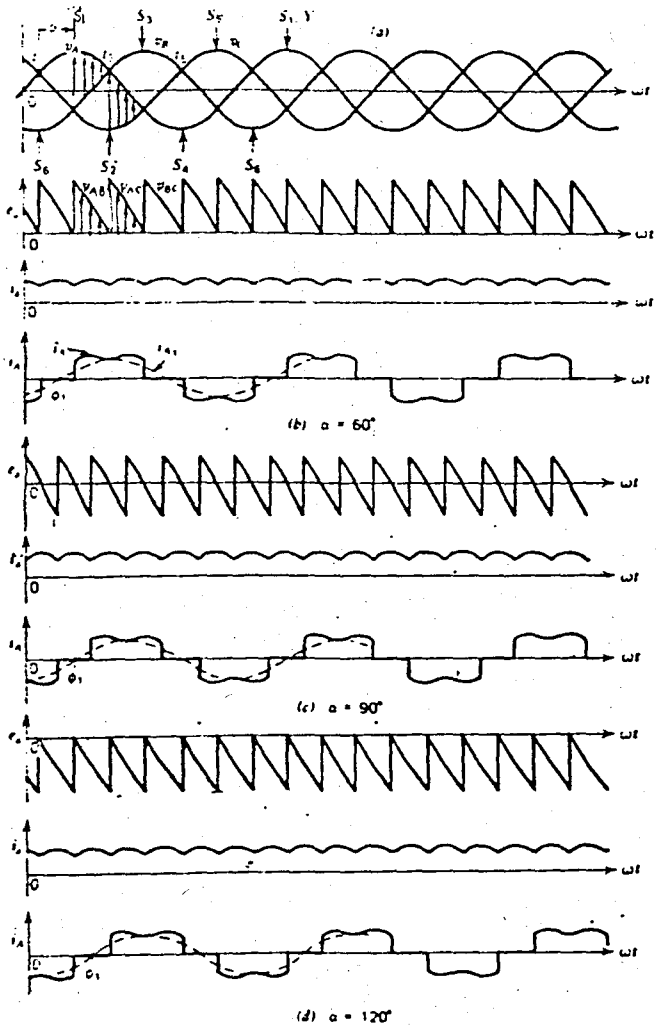


Fig.2.2 Three phase full-converter waveforms at different firing angles for continuous motor current.

At $\omega t = \pi/6+\alpha$, S1 turns on. Prior to this instant S6 was turned on therefore, during the interval $(\pi/6+\alpha)<\omega t<(\pi/6+\alpha+\pi/3)$, thyristor S1 and S6 conduct, and the motor terminals are connected to phase V_a and V_b, making $e_a = V_{ab}$. At $\omega t=\pi/6+\alpha+\pi/3$, thyristor S2 is fired, and immediately thyristor S6 is reverse biased and turns off. (Naturel or line commutation) The current from S6 is transferred to S2, and therefore the motor terminals are connected to phase V_a through S1 and phase V_c through S2 making $e_a = V_{ac}$. This process repeats after every 60° whenever a thyristor is fired.

The motor terminal voltage can become negative, as shown in Fig. 2.2.d, for the triggering angle $\alpha > 90^\circ$. This is the inversion mode of operation of the converter.

The average motor terminal voltage can be calculated as follows: let the supply voltage be

$$V_a = \sqrt{2}\, V \sin\omega t$$

$$V_b = \sqrt{2}\, V \sin(\omega t-2\pi/3)$$

$$V_c = \sqrt{2}\, V \sin(\omega t-4\pi/3) \qquad (2.1)$$

The average motor terminal voltage is

$$E_a(\alpha) = \frac{3}{\pi}\int_{\pi/6+\alpha}^{\pi/6+\alpha+\pi/3} (V_a - V_b)\, d(\omega t) \qquad (2.2)$$

Solving the above integral one gets

$$E_a(\alpha) = \frac{3\sqrt{6}}{\pi} V \cos\alpha \qquad (2.3)$$

The relation between E_a and α for continuous motor current is shown in Fig. 2.3.

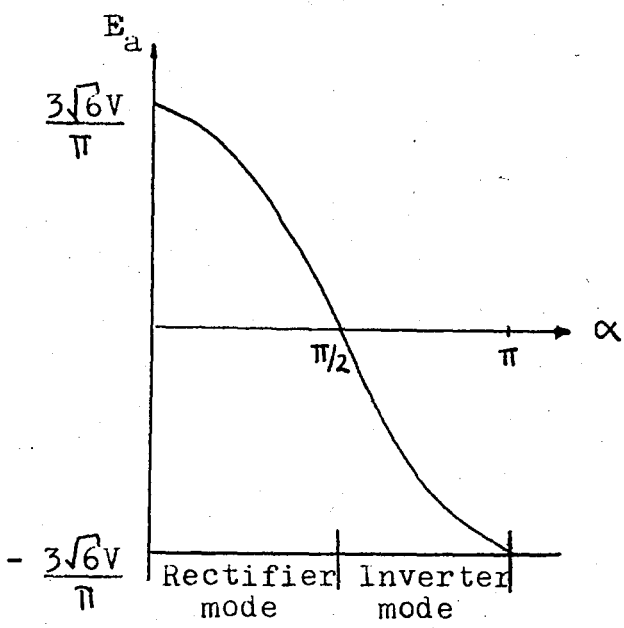


Fig. 2.3 Voltage-firing angle characteristic of full wave converter.

C.The Thyristor Unit Used in the Thesis

In the thesis the thyristor unit CA6000 is used for the experimental work. CA 6000 is an educational three phase six pulse bridge connected,nonreversible thyristor converter which can be used for laboratory exercises or practical training.

CA 6000 is based on a modern thyristor converter "Contravas VDB 380.30 ",developed and manufactured for industrial applications by Contravas Antriebstechnic AG,Zurich (Switzerland) and adapted for use as a laboratory training aid. This converter in addition to all classic functional units,like speed and current controllers,firing pulse control circuits,thyristor bridge,also contains a number of auxiliary protection and monitoring functions,

which are necessary for industrial applications.

Technical specifications of Thyristor Unit CA 6000 :

Ambient temperature range	0-45°C
Rated output dc current (continuously at 45)	30 A
Input current mains side	26 A ac
Max. power dissipation	150 W
Mains frequency (selectable)	50 or 60 Hz
Mains supply voltage	3x220 or 380 V ac
Field supply dc voltage	0.9xmains supply voltage
Field current (max.)	4 A
Stabilized reference voltage	± 15 V
Speed control range (with tacho feedback)	0 - 100 % of the rated speed

The internal circuitry of the thyristor unit can be divided into two basic parts :

i. power circuitry

ii. control, protection and monitoring circuits

The power circuits basically consist of a main thyristor controlled rectifier for armature supply and an auxiliary rectifier for field excitation supply. In thyristor unit CA 6000 the main thyristor rectifier consist of a fully controlled three phase bridge connected thyristor rectifier.

The auxiliary rectifier supplies the current for the field excitation of the dc motor. In most cases, dc motors are being operated with constant field and only a simple low power single phase diode bridge like in CA 6000 is used to supply the field current.

Closed-loop automatic control circuits are used for both controlling the speed and the current of the motor. The actual speed of the motor is measured by a dc tachogenerator. This tachogenerator is a low power dc generator with permanent magnet field excitation,designed so that its output voltage is a very accurate linear function of its rotor's angular velocity. The value of actual speed and direction of rotation are thus represented by the magnitude and polarity,respectively,of the tachogenerator's output voltage.

At the input of the speed controller,the tachogenerator output (actual speed) is compared to another voltage, called speed reference, which represents the desired motor speed.The speed reference voltage can be set by means of a potentiometer inside the unit, or may come from another source. In this thesis this reference voltage is generated by the microcomputer,applying the model algorithmic control algorithm,through a digital to analog converter.

The speed reference can be fed to the speed controller's input directly or via a linear ramp. The linear ramp prevents sudden changes of the speed reference. Linear ramp limits the rate of change of speed reference to a certain ,pre-setable, maximal value. The difference between the speed reference and actual speed voltages in the speed controller is thus a measure of the speed error,which has to be corrected by the control algorithm. In most drive applications, this correction has to be made as fast as possible. To determine the necessary change of the value of the armature current and to actuate this change

the following control scheme is used : Speed controller, the input signal of which is the speed error, determines the value of the current necessary to correct (or maintain) the speed, that is,the output voltage of the speed controller is the measure of the desired armature current,the current reference. A current controller then compares the signal representing the actual value of the armature current with the current reference and determines the angle at which the thyristors in the main rectifier are fired,as that the desired value of the armature current will be achieved (and maintained) ,irrispectively of the motor speed,within a short time.

The task of generating firing pulses at time instants, determined by the current controller and distributing them in the right sequence to the thyristors is performed by the firing pulse unit. The firing pulse generating circuitry is shown in Fig. 2.4. This figure actually shows only generation of firing pulses for the thyristor pair S1 and S4. Identical circuits are used for generating firing pulses for the two remaining thyristor pairs. The signal waveforms of firing pulse circuitry is shown in Fig.2.5.

The transformer T_r is used for synchronization,on the secondary part of the T_r we obtain the synchronization voltage U_{RT}. This synchronization is phase delayed by 60° in the RC filter (R_f, C_f). The zero crossing of this delayed voltage is thus coincide with the points of natural commutations of the thyristors S1 and S4. These zero crossings are detected by the zero crossing detector. Zero-crossing signal drives the saw-tooth waveform generator of which output frequency is twice the line frequency

and synchronized with the points of natural commutation of thyristors.

This saw-tooth voltage is then in the comperator compared to the control voltage v_V which is proportional to the firing angle. The control voltage changes from approximately 8 V in the beginning to approximately 1 V at the end, so the firing angle varies from approximately 30° to 150°.

The complete circuit for the three phase firing circuit is shown as a block diagram in Fig. 2.6.

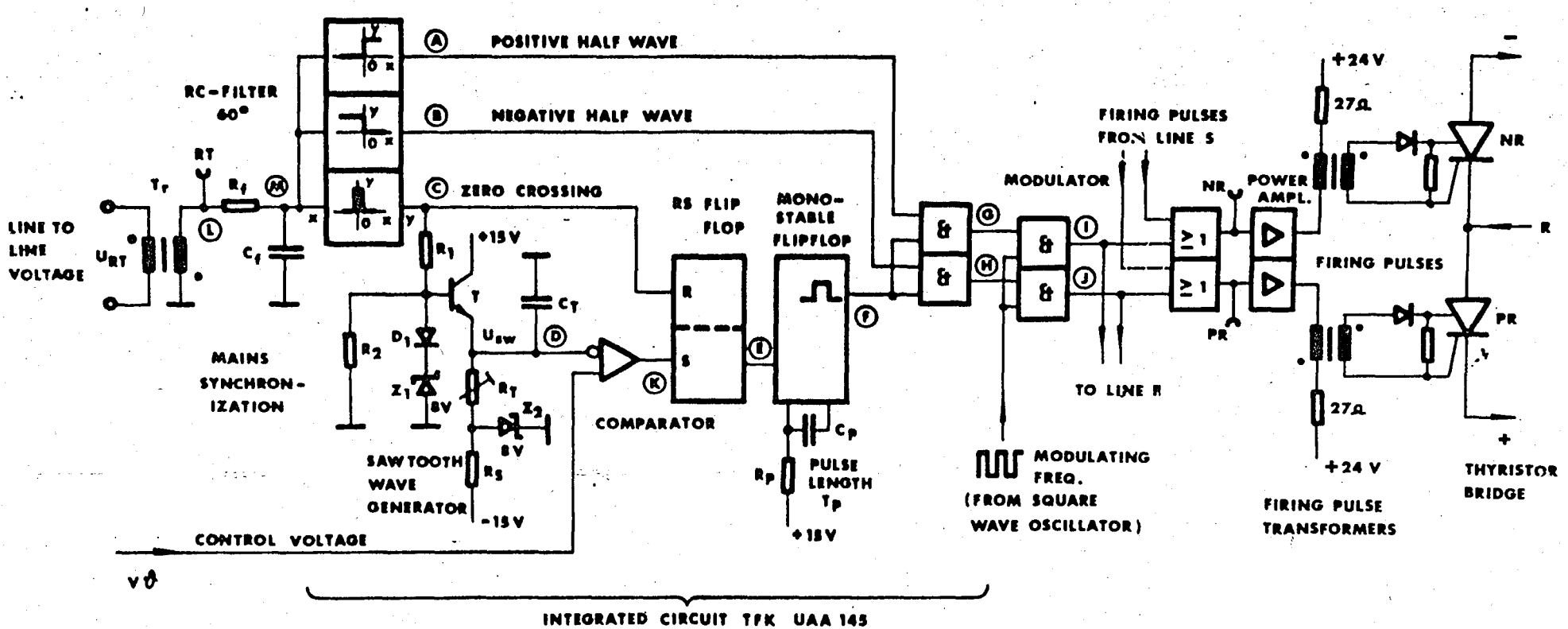


Fig. 2.4 Firing pulse generating circuitry.

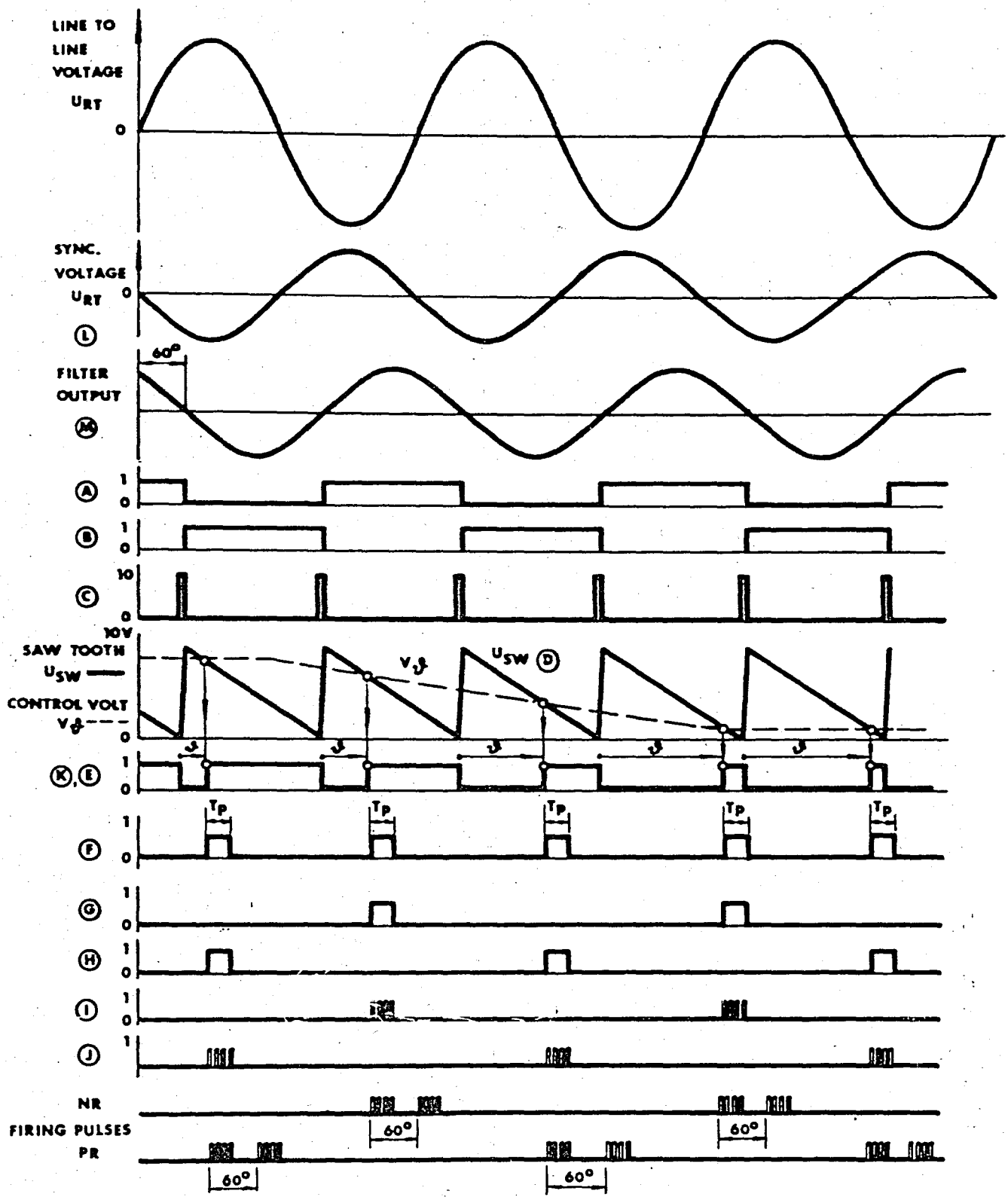


Fig. 2.5 Signal waveforms in the firing pulse unit

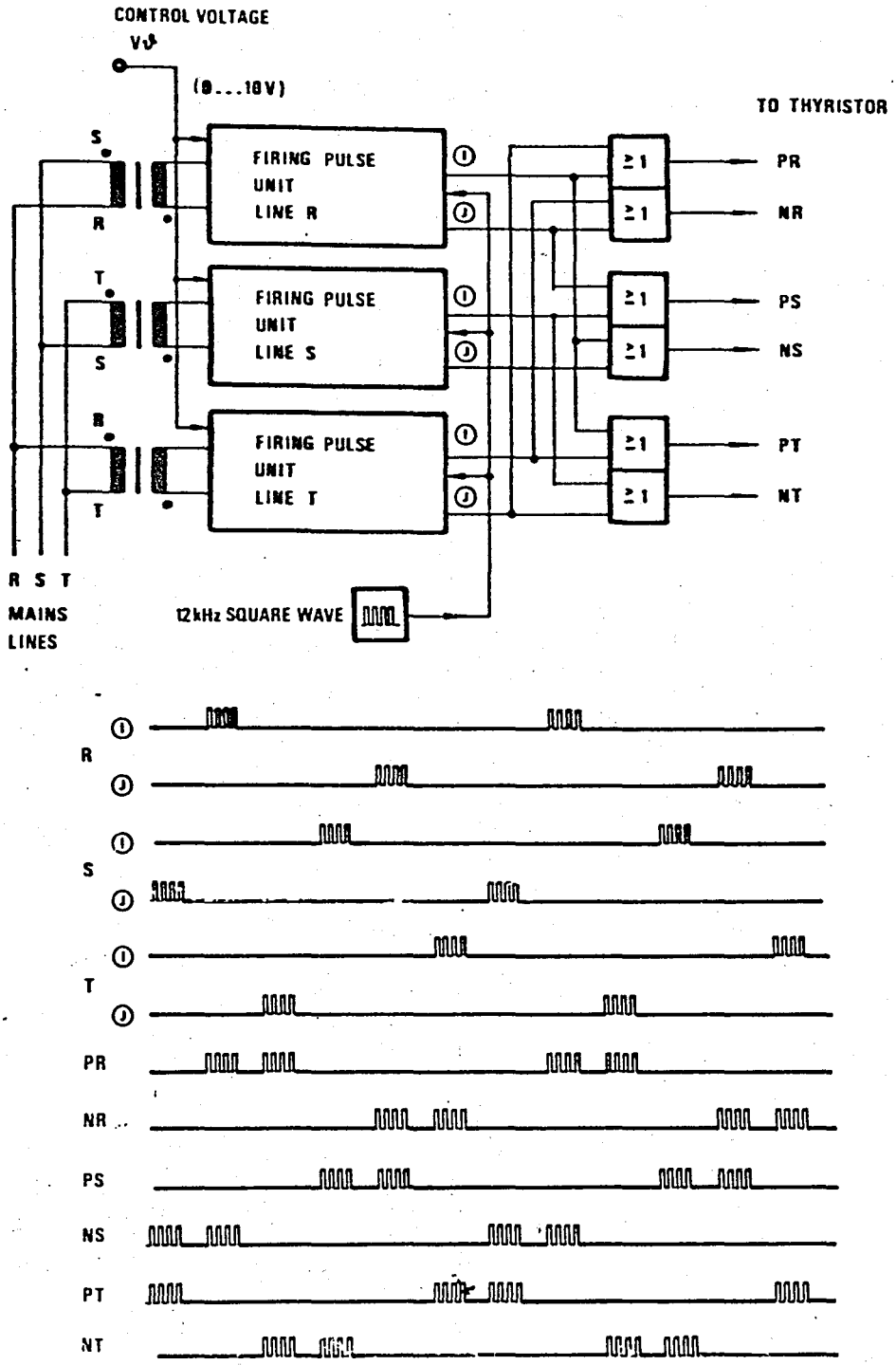


Fig. 2.6 Three phase firing unit

The speed controller part can be operated as proportional, proportional-integrating,or proportional-integrating-derivating

control by means of selectable external connections. Current control part can be operated also proportional and proportional-integrating control. In the thesis, the speed controller is achieved as proportional control.

III. MODEL ALGORITHMIC CONTROL

The increasing use of digital microcomputers has considerably helped the industrial implementation of algorithmic-adaptive types of control. Fast digital computing facilities with fast memory access and substential information storage capacity made the algorithmic-adaptive types of control to be practical and efficient.

The success of model algorithmic control scheme operating on a process is due to the particular representation of the process that it uses, namely impulse response representation from input to output. For most multivariable industrial processes, parametric models ,such as state-space models based on physical laws are difficult to obtain. It is known that parametric models can give results with large error if the order of the model does not agree with the order of the plant. Moreover in an industrial environment,perturbations affect the plant structure more often than the measurable variables. This requires a constant checking and updating of the model parameters. The impulse response representation is convenient, since in most industrial multivariable processes,the identification of impulse response is relatively simple. The error due to truncations and approximations

of impulse responses can be looked at as a mismatch between the plant and its model. This mismatch is,partially,responsible for the discrepancy between the process output and its predicted value. The closed loop model algorithmic control, where this discrepancy enters the computation of the plant input,displays a particularly high degree of robustness against model -plant mismatch.

The model algorithmic control strategy relies on three principles :

i. The process to be controlled is represented by its impulse response which constitute the internal model. This model is used on-line for prediction,its inputs and outputs are updated according to the actual state of the process. Though it could be identified on-line, this model is most of the time computed off-line.

ii. The strategy is fixed by means of a reference trajectory which defines the closed-loop behavior of the plant. This trajectory is initiated on the actual output of the process and tends to the desired set-point.

iii. Controls are not computed by a one-shot operator or controller but through a procedure which is heuristic in the general case. Future inputs are computed in such a way that, when applied to the fast time internal predictive model, they induce outputs as close as possible to the desired reference trajectory.

A. Representation and Prediction

The system is represented by its impulse response, the iden-

tification of which can be done both on-line and off-line. However in most cases, the off-line identification is accurate enough for the purpose of control and one can avoid the cost and complexity of an on-line identification procedure. This is due to the particular redundancy of the impulse response representation which allows a considerable enhancement of the robustness of the control scheme against identification errors and parameter perturbations.

The representation of the system is as follows :

$$y(t+1)= \underline{h}^T\underline{x}(t)= h_o x(t) + h_1 x(t-1) +.....+h_N x(t-N) \qquad (3.1)$$

where $y(t+1)$ is the plant output at time $t+1$, $h^T \in R^{N+1}$ denotes the plant impulse response. $X(t-j) \in \Omega \subset R$ for $j= 0,.....,N$, is the input at time $t-j$ to the plant. Ω is the constrained set of the input.

The model (3.1) is also called the actual model to emphasize that h^T represents the actual process. But since such a perfect knowledge of the plant impulse response is usually not possible, one has to use an approximation $\hat{h}^T$ to h^T. The model corresponding to this latter impulse response is than

$$y_m(t+1) = \underline{\hat{h}}^T \underline{x}(t) \ . \qquad (3.2)$$

The above model together with the past history of the plant output denoted by $Y(t) = \{ y(\tau) \quad \tau \leq t \}$ is used to predict the future value of the output.

B. The Reference Trajectory

The purpose of the control is to lead the output $y(t)$ along a desired, and generally smooth, path to an ultimate set point c. Such a path is called a reference trajectory. From the last sampled value $y(t_o)$ a trajectory $y_r(t_o+k)$ is initialized which reaches c according to some criterion (e.g. no overshoot, fixed time response). The desired values of the future output can be obtained from stored data or computed by a recursive equation. One of the simplest (first order) reference trajectory is

$$y_r(t_o+k) = \alpha^k y_r(t_o) + (1-\alpha^k)c \qquad |\alpha| < 1$$

$$y_r(t_o) = y(t_o) \tag{3.3}$$

The control algorithm has to find a set of future control variables such that the future outputs of the internal model will be as close as possible to the reference trajectory.

The reference trajectory can be chosen to be of higher order and the set point can be made time variant.

C. The Optimality Criterion and the Optimal Control Strategy

The optimality criterion should reflect the previously mentioned purpose of following the reference path to the desired set point c. This can be done by defining the optimum control strategy as the one which minimizes over a certain horizon in the future, the deviation of the predicted outputs from the reference path.

Formally, at each instant t_o , the optimum set of T future inputs $\{x^*(t_o), x^*(t_o+1), \ldots\ldots, x^*(t_o+T-1)\}$ are such that the predicted T outputs $\{y_p(t_o+1), \ldots\ldots, y_p(t_o+T)\}$ are as close as possible, in the sense of a weighted Euclidean norm, to the reference trajectory. Therefore the function to minimize is

$$J_T = \sum_{k=1}^{T} \left(y_p(t+k) - y_r(t+k) \right)^2 w_k \qquad (3.4)$$

where w_k is a nonnegative weighting factor.

The interval $\{t_o , t_o+T\}$ is called the horizon of control evaluation, and sometimes the horizon of prediction since at time t_o one has to predict a set of T outputs $y_p(t+k)$, $(k=1,2,\ldots,T)$. Ones the set $X^*(t_o)$ of the T optimum inputs is determined, it is possible to wait up to T periods before observing the process output $y(t)$, and reinitializing $y_r(t)$, predicting y_ps and computing the next set $X^*(t_o+T)$. This means that all the elements of the optimum control set $X^*(t_o)$ have been actually applied to the plant.

In this thesis, the horizon of control evaluation is selected as a one sampling period. This is called one-step prediction. The reference trajectory for one-step prediction is continuously initialized on the observed plant outputs $y(t_o)$, $y(t_o+1), \ldots, y(t_o+T-1)$ that is

$$y_r(t_o+j) = \alpha\, y(t_o+j-1) + (1-\alpha)c \ ,$$

$$\forall_{j=1,\ldots\ldots,T} \qquad (3.5)$$

Also the optimality criterion requires that

$$J_1 = (y_r(t+1) - y_p(t+1))^2 w_1 \tag{3.6}$$

$$y_r(t+1) = y_p(t+1) \tag{3.7}$$

The optimal control strategy is then, as shown schematically in the Fig. 3.1, realized as follows : At time t_o, the plant output is predicted from the reference trajectory, then the optimum input is calculated to lead the process to follow the reference trajectory as closely as possible.

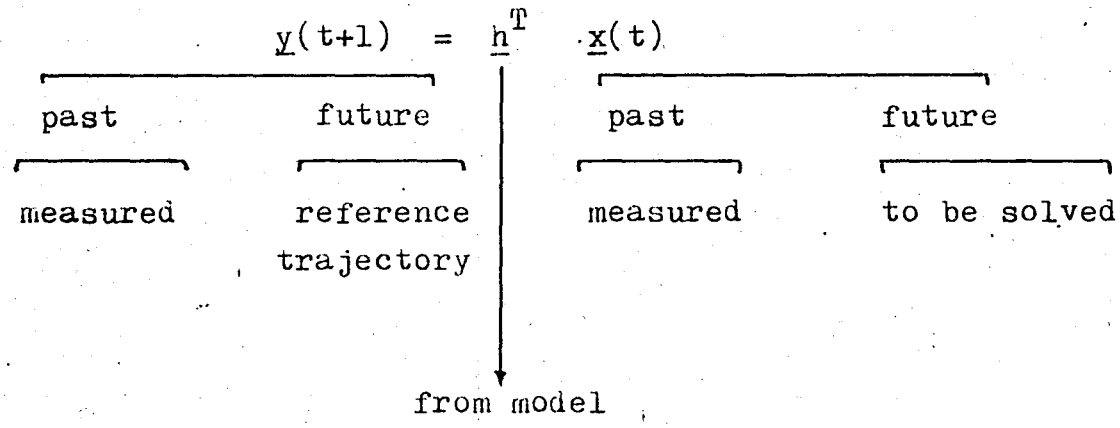


Fig. 3.1 Optimal control strategy

D.Open-Loop and Closed-Loop Prediction

The simplest one step (T=1) open-loop prediction scheme that one can imagine is

$$y_p(t+1) = y_m(t+1) = \hat{\underline{h}}^T \underline{x}(t), \tag{3.8}$$

that is, the model of the plant is used to predict the output one step ahead. The main inconvenience of such a prediction is that the output $y(t)$ of the controlled plant would not converge to the set point c when there is a mismatch between the model and the plant.

From equations (3.7) and (3.5)

$$y_p(t+1) = \alpha\, y(t) + (1-\alpha)c\ , \tag{3.9}$$

and writing the above equation in the expanded form, we get

$$\hat{h}_o x(t)+\hat{h}_1 x(t-1)+.....,+\hat{h}_N x(t-N) - \alpha(\ h_o x(t-1)+h_1 x(t-2)+.....+$$

$$h_N x(t-N-1)\ ) = (1-\alpha)c \tag{3.10}$$

$$\hat{h}_o x(t)+(\hat{h}_1-\alpha\, h_o)x(t-1) + (\hat{h}_2-\alpha\, h_1)x(t-2) +.....+$$

$$(\hat{h}_N-\alpha\, h_{N-1})x(t-N) -\alpha\, h_N x(t-N-1) = (1-\alpha)c \tag{3.11}$$

If the system is stable, the sequence converges to a value $\bar{x}$, then

$$(\hat{h}_o+ \hat{h}_1-\alpha\, h_o+ \hat{h}_2-\alpha h_1+.....+\hat{h}_N-\alpha\, h_{N-1}-\alpha\, h_N)\bar{x} = (1-\alpha)c \tag{3.12}$$

Therefore

$$\bar{x} = \frac{(1-\alpha)c}{\sum_{i=0}^{N} h_i - \alpha \sum_{i=0}^{N} h_i} \quad . \tag{3.12}$$

The output for this input is

$$y = h^T \bar{x} = \frac{(1-\alpha)c \sum_{i=0}^{N} h_i}{\sum_{i=0}^{N} \hat{h}_i - \alpha \sum_{i=0}^{N} h_i} \quad . \tag{3.13}$$

The difference between the set point and the output is

$$y-c = \left(\frac{(1-\alpha)c \sum_{i=0}^{N} h_i}{\sum_{i=0}^{N} \hat{h}_i - \alpha \sum_{i=0}^{N} h_i} - 1 \right) c \tag{3.14}$$

Therefore the bias term is found to be

$$\frac{\sum_{i=0}^{N} h_i - \sum_{i=0}^{N} \hat{h}_i}{\sum_{i=0}^{N} \hat{h}_i - \alpha \sum_{i=0}^{N} h_i} c \tag{3.15}$$

To alleviate the bias problem, the closed-loop prediction is applied. The closed-loop prediction is as follows :

$$y_p(t+1) = y_m(t+1) + (y(t) - y_m(t)) \tag{3.16}$$

where $y(t) - y_m(t) = (\underline{h} - \hat{\underline{h}})\, \underline{x}(t-1)$ represents a correction term similar to the innovation term of Kalman filter or the correction term of an observer, assuring the final convergence of the plant output to the set point.

Rewriting the equation (3.9)

$$y_p(t+1) = y_r(t+1) = \alpha\, y(t) + (1-\alpha)c \ , \qquad (3.17)$$

and inserting this equation into the equation (3.16), we get

$$(1-\alpha)\,(c-\ y(t)\ ) = y_m(t+1) - y_m(t) \ . \qquad (3.18)$$

Assuming that the system is stable, as $t \to \infty$ it can be shown that

$$(1-\alpha)\,(c-\ y(t)\ ) = 0$$

then

$$\lim_{t\to\infty} y(t) = 0 \qquad (3.19)$$

The optimum sequence of inputs $x^*(t)$ in the sense of minimizing J_1, for a closed-loop system, is generated by an auto-regressive equatio, which results from equation (3.18), where $y_p(t+1)$ and $y_r(t+1)$ are expressed in terms of inputs

$$(1-\alpha)\Big(c-\sum_{j=0}^{N} h_j x(t-j-1)\ \Big) = \sum_{j=0}^{N} \hat{h}_j x^*(t-j) - \sum_{j=0}^{N} \hat{h}_j x^*(t-j-1) \qquad (3.20)$$

$$(1-\alpha)(c-\sum_{j=0}^{N} h_j x(t-j-1)\ ) = \hat{h}_o x^*(t)+\sum_{j=1}^{N} \hat{h}_j x^*(t-j) -\sum_{j=0}^{N} \hat{h}_j x^*(t-j-1) \tag{3.21}$$

Therefore the optimum input is

$$\hat{h}_o x^*(t) =\sum_{j=0}^{N} \hat{h}_j x^*(t-j-1) -\sum_{j=1}^{N} \hat{h}_j x^*(t-j) + (1-\alpha)(c-\sum_{j=0}^{N} h_j x(t-j-1)\ ) \tag{3.22}$$

The summation in the last term in the equation (3.22) can be written as $y(t)$ because this summation is the actual speed obtained by measurements at every sampling instants.Also we can write the summations in the above equation in opened form and making some manipulations we get

$$x^*(t) = (\frac{\hat{h}_o-\hat{h}_1}{\hat{h}_o})x^*(t-1) + (\frac{\hat{h}_1-\hat{h}_2}{\hat{h}_o})x^*(t-2) +\cdots\cdots+ (\frac{\hat{h}_N-\hat{h}_{N-1}}{\hat{h}_o})x^*(t-N-1) - \frac{\hat{h}_N}{\hat{h}_o}x^*(t-N) + \frac{(1-\alpha)}{\hat{h}_o}(c-y(t)) \tag{3.23}$$

E.Tracking Type Model Algorithmic Control

In many applications, it is desired for the output of the system under the control to track a certain trajectory. The tracing behavior of the output does not complicates the theory of model algorithmic control. The only thing that we should have to do is to make the set point c time varying.

Time varying set speed can be achieved in different ways. For example, the desired trajectory might be the output of another

process. If at each sampling interval the computer reads this external process output as a set speed c, than the output of controlled process follows the time varying set speed, the output of the external process. The rate of change of external process output, of course, must be slower than the time constant of the software trajectory. If the time constant of software trajectory is lower than the external process, there will be always error between the desired trajectory and the process output, because the software trajectory will not have enough time to follow the external process output.

In the thesis, the time varying set speed C(t) is in the form of a linear ramp, the slope of which is made equal to desired acceleration. The maximum acceleration of the motor output can be calculated from the equation (1.7) by neglecting the load torque

$$T_e = Bn + J \frac{dn}{dt} = K_a \phi_f i_a \qquad (3.24)$$

For maximum speed and maximum generated torque, maximum acceleration $\frac{dn}{dt}$ is calculated. If the maximum speed is devided by the maximum acceleration, the time that the ramp reaches the maximum speed is found.

For the parameters of the system that the experimental investigations were carried on, the maximum acceleration that the system is capable of was found to be 11.27 rad/sec^2.

IV. MICROCOMPUTER BASED CONTROL

In this chapter the design of a microprocessor based control of a dc motor fed by a three phase full-converter is described The system is realized by Z-80 based microcomputer. The microcomputer manipulates the software of model algorithmic control and communicates with the thyristor unit and the tachogenerator through a D/Aconverter and a A/D converterrespectively.

A. The Internal Model of the System

The purpose of the internal model as mentioned in chapter three is to have a flexible representation of the controlled system stored in the microcomputer memory, which can be updated as the system changes and which can be used at any instant to predict the future behavior of the system under different control inputs.

The internal model of the system is generally obtained via off-line identification, using either a physical model structure when one is evailable or an input-output type representation model such as the impulse response model.

The simplified transfer function of the dc motor is

$$\frac{N(s)}{E_a(s)} = \frac{K_m}{1 + sT_{ml}} \tag{4.1}$$

where

$$K_m = \frac{K_a\phi_f}{(K_a\phi_f)^2 + R_aB} \tag{4.2}$$

$$T_{ml} = \frac{R_aB}{(K_a\phi_f)^2 + R_aB}\ T_m \tag{4.3}$$

In order to be able to identify the dc motor used in the thesis, we have to find the motor parameters $K_a\phi_f$, B, J, R_a and T_m. These parameters are found with the help of thyristor unit, so the transfer function obtained at the end of experiments represents both the thyristor unit and the dc motor.

The armature resistance R_a is measured when the motor is hot by using an ohm-meter. This measurement is repeater for different angular positions and the average value is calculated. After this process R_a is found to be 4.8 ohms.

The back e.m.f. constant $K_a\phi_f$ is obtained by running the machine as a generator at the rated field current and calculating the ratio of the generated voltage to speed. $K_a\phi_f$ is found to be 1.361 V/rad/sec.

The mechanical damping coefficient B is found with the help of equation below

$$J\frac{d\omega}{dt} + B\omega + T_L = T_e = K_a\phi_f I_a \ . \tag{4.4}$$

If the machine rotates at a constant speed and no load is applied the equation (4.4) reduces to

$$B\omega = K_a \phi_f I_a \qquad (4.5)$$

Measuring the I_a at rated speed, and repeating this a few times B is found to be 0.006498 rad/V.sec .

The equation (4.4) is also utilized in finding the total motor armature and tachogenerator inertia J. The speed reference potentiometer of the thyristor unit is preset to the value thet the motor rotates at 1400 rpm. The start switch is depressed and by means of a storage oscilloscope the acceleration recording in Fig. 4.1 is obtained. Since no load is applied the equation (4.4) becomes

$$J\frac{d\omega}{dt} + B\omega = K_a \phi_f I_a \qquad (4.6)$$

and the armature current is measured to be 0.85 A. Because of the internal dynamics of the thyristor unit, the acceleration recording consists of two parts with different time constants. Only the interval from 0 to 0.5 sec. part is considered for the calculation of J.

The slope in the interval 0 to 0.3 sec. is about $\frac{d\omega}{dt} = \frac{5.75}{1.5}$

then

$$J\frac{5.75}{1.5} + 0.0065\frac{1150\ 2\pi}{60} = 1.361\ 0.85$$

Solving the above equation the total inertia is found to be

$J = 0.098 \quad kg\text{-}m^2$.

The transfer function of the system is then

$$\frac{N(s)}{E_a(s)} = \frac{0.722}{1 + 0.249s} \tag{4.7}$$

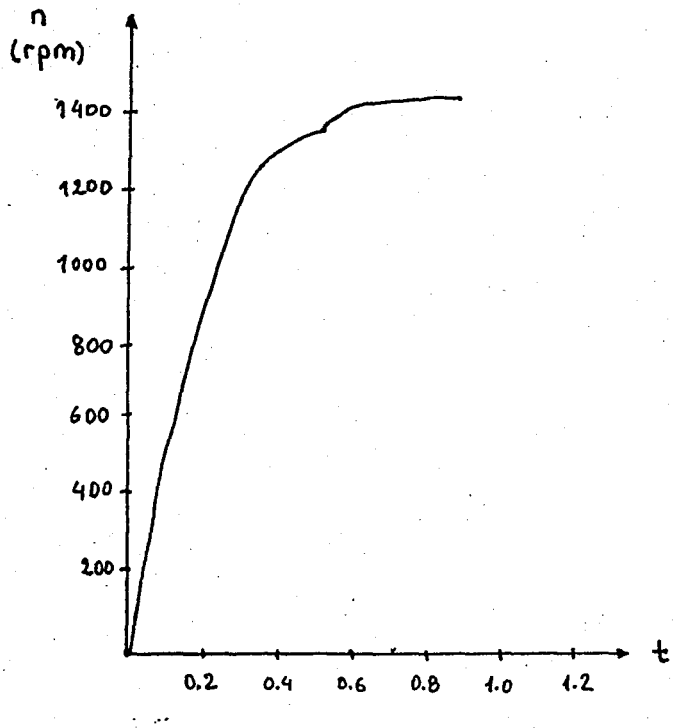


Fig. 4.1 Time responce of dc motor driven by thyristor unit..

The next step is to find the discrete impulse response of the dc motor.The control input $e_a(t)$ is assumed to be piecewise constant,then the system is as in Fig. 4.2.

$$G(s) = \frac{N(s)}{E_a(s)} = \frac{1-e^{-sT}}{s} \quad \frac{b}{s+a} \tag{4.8}$$

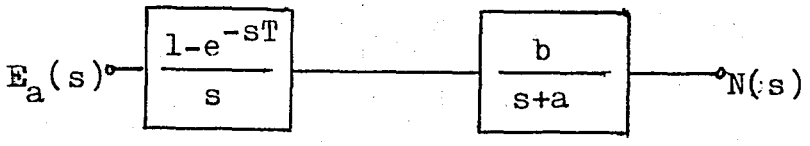


Fig. 4.2 System with Zero Order Holder

In the Z domain the equation (4.8) is

$$G(z) = \mathcal{Z}\left\{ \frac{b}{a} (1 - e^{-sT})\left(\frac{1}{s} - \frac{1}{s+a}\right)\right\} . \qquad (4.9)$$

Our system representation was $y(t+1) = \underline{h}^T x(t)$ so the output in the Z domain is multiplied by z then the transfer function in Z domain is

$$H(z) = z\, G(z) \qquad (4.10)$$

then

$$H(z) = \frac{b}{a} (z-1) \left(\frac{z}{z-1} - \frac{z}{z-e^{-aT}} \right) \qquad (4.11)$$

$$H(z) = \frac{b}{a} \left(1 - e^{-aT} \right) \frac{z}{z-e^{-aT}} \qquad (4.12)$$

The discrete impulse response is found by taking inverse Z transformation of the above equation. It is found to be

$$h(kT) = \frac{b}{a} \left(1 - e^{-aT} \right) e^{-akT} \quad \text{for} \quad k=0,1,\ldots,N \qquad (4.13)$$

where $a = 2.893$ and $b = 4.001$.

N, which is the number of the impulses to approximate the dc motor,is selected such that the h_N can be considered as zero. The sampling period T is selected as 8 msec. This time interval limits the computer time to calculate the control inputs at every sampling period. For the system used in the thesis N is selected as 145, because time limit lets us only that amount of data to manipulate.

As it is mentioned in chapter three, for model algorithmic control to be practical and efficient, the controller must have fast digital computing facilities. Our system is controlled by a Z-80 based microcomputer with clock frequency 2.5 MHz. This clock frequency never let us to be able to manipulate 145 multiplications and the same number of summations. This limitation forces us to group the coefficients in equation (3.25), without disturbing the exponential decay of them as much as possible. This approximation is shown schematically in Fig. 4.3.

Grouping the coefficients in equation (3.25) considerably reduces the computation time of the control inputs. Instead of 145 multiplications and 145 summations the computer does 145 summations and 13 multiplications. The mutliplication program consumes much more time than the summation, so reducing the number of multiplications saves quite an important amount of computation time.

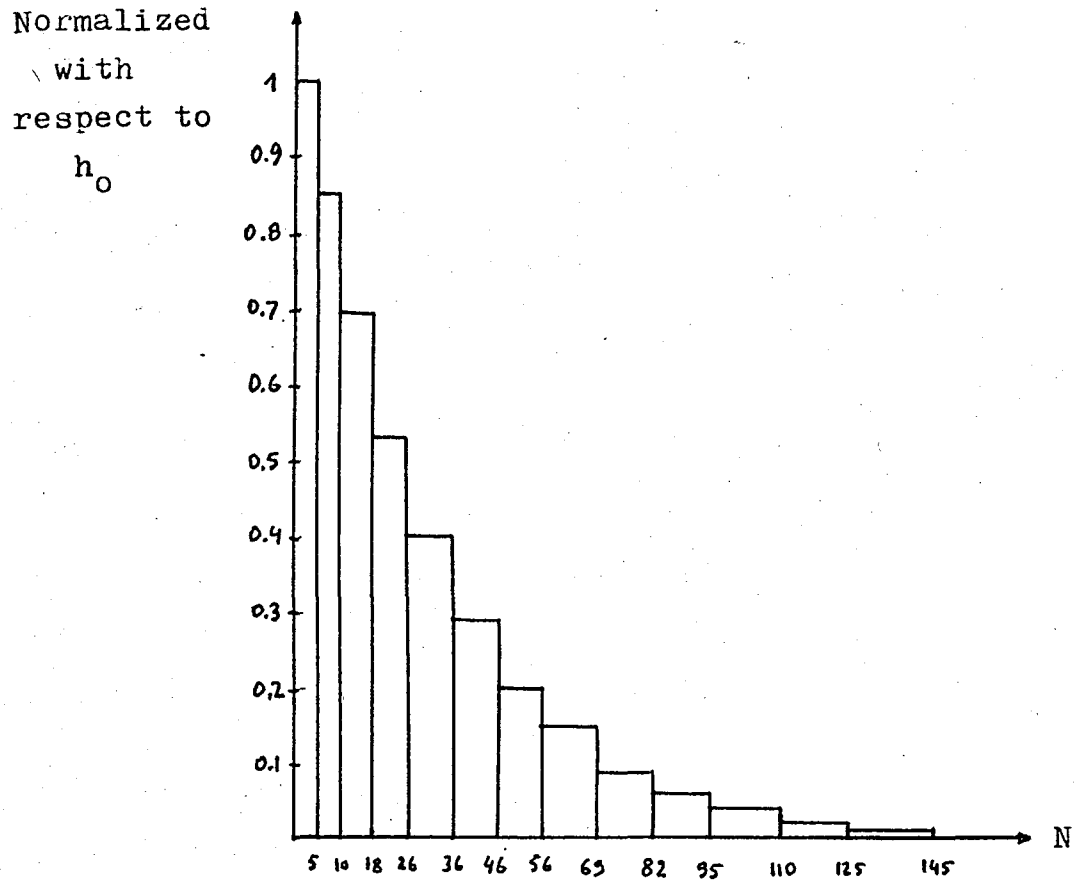


Fig. 4.3 The grouping of coefficients in control equation

The overall model algorithmic control system is shown as a block diagram in Fig. 4.4.

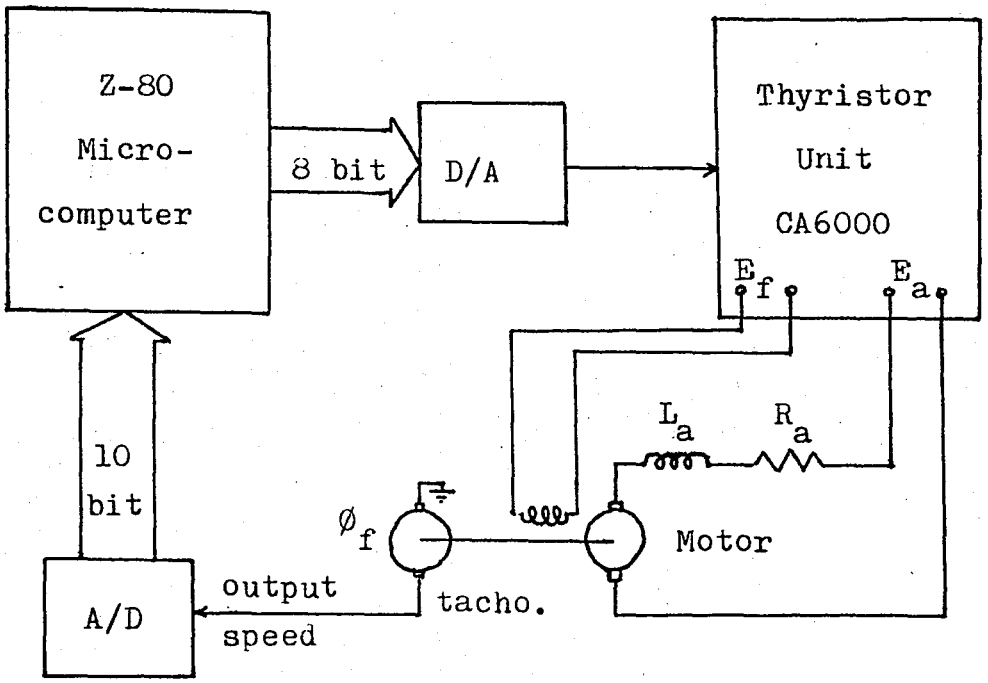


Fig. 4.4 Dc motor speed control system

B. Software

The optimum control input $x^*(t)$ was found in the previous chapter in equation (3.23). The coefficients in this equation are precalculated by solving the equation (4.13), discrete impulse response, and located in the memory of the microcomputer. The software of model algorithmic control for constant set speed in the memory solves the equation (3.23) for $x^*(t)$. The flow-

chart of the program developed is given in Fig. 4.5 and the software listing in Appendix A. Since software takes 8msec calculation time which is equal to the sampling period, an external timer is not included in the hardware of the system.

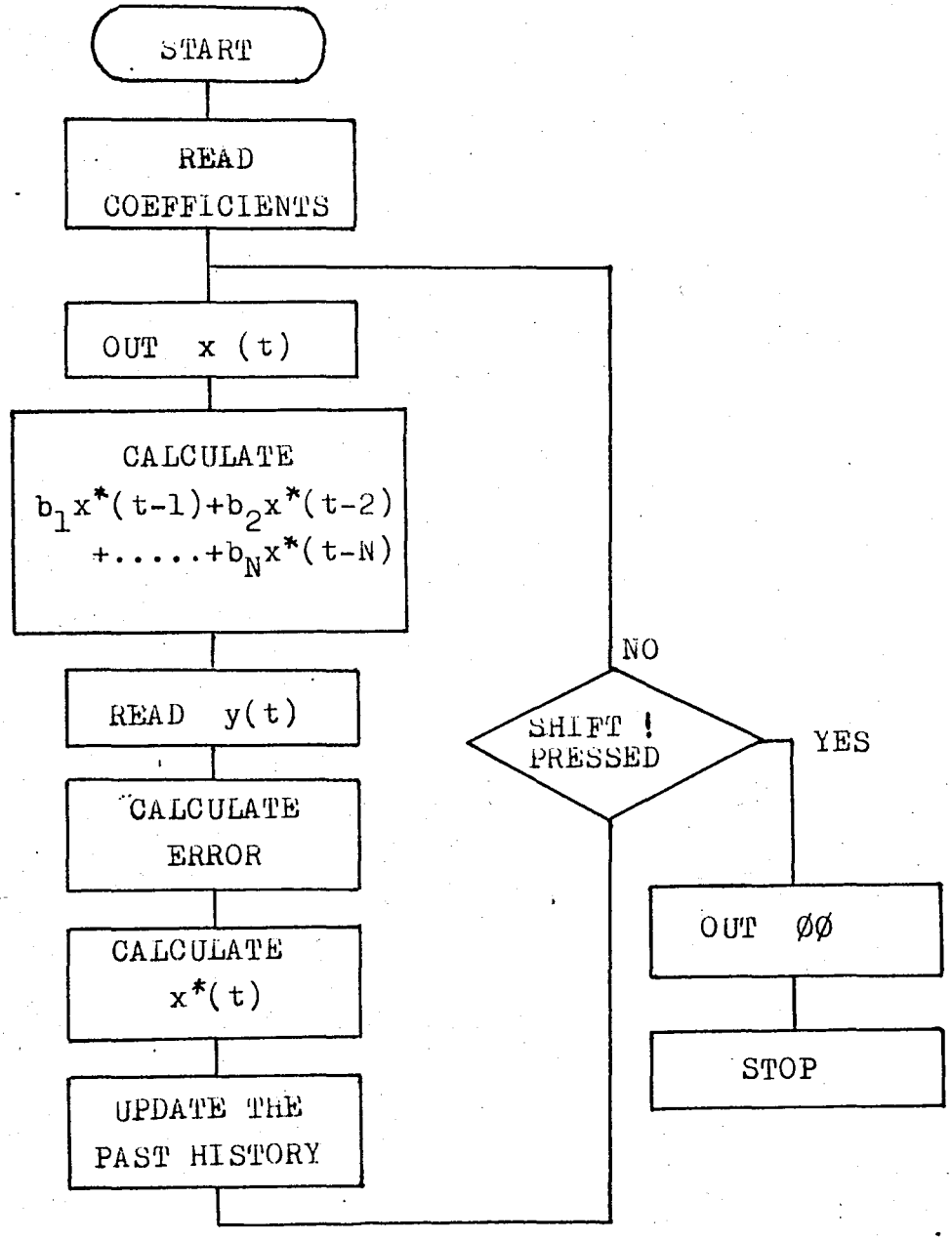


Fig. 4.5 Flow-chart

Also the software listing for model algorithmic control for time varying set speed is given in Appendix B.

V. STABILITY AND EXPERIMENTAL RESULTS

A. Stability

The optimum sequence of inputs $x^*(t)$, obtained such that the outputs minimizes J_1, for a closed-loop system , was generated in chapter 3 by an auto-regressive equation, where $y_p(t+1)$ and $y_r(t+1)$ are expressed in terms of inputs.

$$\hat{h}_o x^*(t) = \sum_{j=0}^{N} \hat{h}_j x^*(t-j-1) - \sum_{j=1}^{N} \hat{h}_j x^*(t-j) + (1-\alpha)(c - y(t)) \qquad (5.1)$$

If the sequence $x(t)$ tends to an equilibrium value $\bar{x}$, that is,if the corresponding auto-regressive model is stable,then the output $y(t)$ tends to an equilibrium value $y(\infty)$ which equals to set speed in the case of closed-loop prediction. But differs from set speed for open-loop prediction, because, as it is explained in chapter 3,section D, the open-loop prediction produces a bias term which deviates $y(\infty)$ from set speed.

The closed-loop one-step prediction control equation was :

$$(1-\alpha)(c - y(t)) = y_m(t+1) - y_m(t). \qquad (5.2)$$

Rewriting the above equation in terms of impulse responses one gets

$$(1-\alpha)(c-\sum_{j=o}^{N} h_j x(t-j-1)) = \sum_{j=o}^{N} \hat{h}_j x^*(t-j) - \sum_{j=o}^{N} \hat{h}_j x^*(t-j-1) \quad (5.3)$$

In order to be able to analyze the stability of the system , we have to examine the response of the optimally controlled system to the set point c, that is, the relationship between y(t) and c(t). This relationship is found in the Z domain as follows. If the equation (5.3) is transformed to the Z domain, the equation becomes

$$(1-\alpha)(C(z) - z^{-1} H(z)X(z)) = \hat{H}(z)X(z) - z^{-1} \hat{H}(z)X(z) \quad (5.4)$$

$$(1-\alpha)C(z) = (z^{-1}(1-\alpha)H(z) + (1-z^{-1})\hat{H}(z))X(z) \quad (5.5)$$

$$\frac{X(z)}{C(z)} = \frac{(1 -\alpha)}{z^{-1}(1-\alpha)H(z) + (1- z^{-1})\hat{H}(z)} \quad (5.6)$$

Since

$$Y(z) = z^{-1} H(z) X(z) , \quad (5.7)$$

then

$$\frac{Y(z)}{C(z)} = \frac{z^{-1}H(z)(1-\alpha)}{z^{-1}(1-\alpha)H(z) + (1- z^{-1})\hat{H}(z)} \quad (5.8)$$

The closed-loop transfer function obteined above can be represented as a block diagram as it is shown in Fig. 5.1.

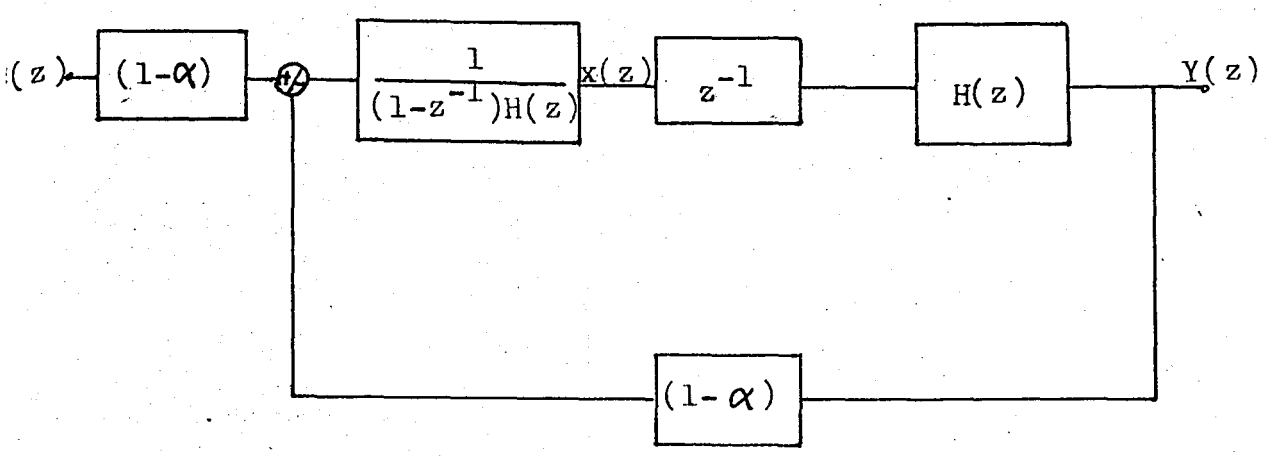


Fig. 5.1 Closed-loop prediction

If the identification of the system impulse response $\underline{h}^T$ can be considered as perfect ,that is, $H(z) = \hat{H}(z)$, the closed loop transfer function becomes

$$\frac{X(z)}{C(z)} = \frac{(1-\alpha)}{(1-\alpha z^{-1})H(z)} \tag{5.9}$$

and

$$\frac{Y(z)}{C(z)} = \frac{(1-\alpha)z^{-1}H(z)}{(1-\alpha z^{-1})H(z)} \tag{5.10}$$

The boundedness condition for the input sequence $x^*(t)$ is identical to the stability of the auto-regressive model, that is, the polynomial in the denominator of the equation (5.8) which is rewritten in equation (5.11) must have all its roots within the unit circle.

$$z^{-1}(1-\alpha)H(z) + (1-z^{-1})\hat{H}(z) = 0 \qquad (5.11)$$

Under the consideration of perfect identification the transfer function of $Y(z)$ with respect to $C(z)$ is the first order and identical to the reference trajectory. Since the $|\alpha|$ is always less than 1, the system is always stable.

We have to consider the case of imperfect identification $H(z) \neq \hat{H}(z)$. From the transfer functions (5.6), (5.8) it becomes clear that for the output $y(t)$ to be convergent and the optimum input sequence to be bounded, it is necessary and sufficient that the denominator of equation (5.8) has all its roots within the unit circle.

In general the exact plant transfer function $H(z)$ is not known and therefore one can not evaluate the exact expression of the polynomial in equation (5.11). However, if the identification of $H(z)$ is fairly good, Then it is expected that the roots of $z^{-1}(1-\alpha)H(z) + (1-z^{-1})\hat{H}(z)$ be close to the roots of $(1-\alpha z^{-1})H(z)$. Therefore, the stability of the identified model $\hat{H}(z)$ implies the stability of the controlled system. But when the identification error becomes large so that the discrepency between $H(z)$ and $\hat{H}(z)$ becomes significant, then to determine the stability of the system one should recourse to robustness analysis which involves determining the set of plant polynomials $H(z)$ for which the above characteristic polynomials have stable roots.

1. Robustness of the System

The control is said to be robust if the plant output $y(t)$

converges to its ultimate desired value c, under a range of plant and behavior changes caused , for instance, by failure of some of its components or by large parameter changes,etc.

With regard to the closed loop prediction control, the robustness problem is posed in terms of the stability of the input sequence $x(t)$ and the plant output $y(t)$. This is a consequence of the fact that the stability of the closed loop system guarantees the unbiased convergence of the output to the set point. Therefore, as long as tne stuctural changes of the plant,which amount to altering the impulse pesponse $\underline{h}^T$ from its nominal (identified) model $\underline{\hat{h}}^T$, are such that the characteristic polynomial (5.11) has all its roots within the unit circle, the control is robust.

Given the model $\underline{\hat{h}}^T$, it is great interest to find a range for the (N+1) dimentional vector $\underline{h}^T$ such that (5.11) is stable. The delimitation of such a subspace of R^{N+1} requires complex search algorithms, and, unrealistic amounts of computation. But fortunately, in most all practical problems,valuable additional information on the physical properties of the plant is aveilable, the appropriate use of this information reduces substantially the size and dimention of the search space. For instance, if the robustness against identification errors or small perturbations is considered, then

$$h_i = \hat{h}_i + \xi_i \qquad i=o,.....,N \qquad (5.12)$$

where $|\xi_i|$ are small with respect to $\hat{h}_i$ and their upper bounds can be estimated from the particular identification scheme in operation and also the stability can be tested.

As a simple example, let us consider the case where the mismatch between $\underline{h}^T$ and $\hat{\underline{h}}^T$ is a pure gain, $\underline{h}^T = q\hat{\underline{h}}^T$. As it is explained, the impulse response is represented by a finite number of elements, then a certain magnitude value of the impulse is accepted as zero. Because of the gain error, the impulse response representing the system will decay this certain value faster than the actual system. The maximum possible mismatch gain is usually larger than unity. Than slowing down the reference trajectory (increasing the α) improves the robustness of the system.

2.Amplitude Constraint on Input

As it is in this application, the input sequence $x(t)$ is usually not free of constraints. In general, there are both amplitude and rate constraints imposed by technical and cost considerations.

In this thesis the speed reference input of the thyristor unit is constrained by 10V as a result of 8 bit digital to analog converters output capability which is 10V. The speed reference input of the thyristor unit is 0 to ±15V, but the maximum output of the digital to analog converter is 10V which is inverted by an operational amplifier. If the system is forced to reach its maximum rated speed (1400 rpm) from zero speed, in the first few sampling intervals the error term $\frac{(1-\alpha)}{h_0}(c-y(t))$ is maximum. At these intervals the algorithm in the computer will calculate maximum speed references to be able to reach the set speed as fast as possible, however, no matter how big the input is

calculated,the applied maximum speed reference input is -10V. This causes the system reach the set speed slower.

The above discussion, amplitude constraint does not effect the system stability, but limits the selectable range of α . The meaning of this limitation in α , in practical application is that the rate of reference trajectory is also limited. The output of 10 bit analog to digital converter is 782 in decimal system (3ØE in hex.) when the motor rotating at 1400 rpm (146.6). The resolution of A/D converter is then 1.79 rpm/bit, or 0.18747 rad/sec/bit The FF in the input of D/A converter produces 10V at the output. From the equation (3.25) the error term is

$$\frac{(1-\alpha)}{h_o}\,(c-y(t)). \qquad (5.13)$$

Let us suppose that the motor is not rotating and no past history is available in the memory. In the first sampling period while calculating $x_1(o)$ only the above equation is manipulated. The The calculation will be as follows

$$0.18747\,\frac{(1-\alpha)}{h_o}\,(782-0) = x\,(o) = 255. \qquad (5.14)$$

Since all calculations are done in rad/sec, the speed of the motor is multiplied by 0.18747 to get the speed in rad/sec. Equation is equaled to 255 to be able to obtain maximum speed reference output from the D/A converter.

From the equation (5.14) minimum α is found to be 0.96 . Therefore

$$0.96 \leqslant \alpha < 1 . \qquad (5.15)$$

B.Experimental Results

The experiments are done for two different time constants of software reference trajectory,which are the maximum allowable α and the minimum one. Also to observe the system response to parameter variations, an induction motor is coupled to the dc motor to change the moment of inertia of the system.

Fig. 5.2 is the time response of the motor speed for α =0.967 with the induction motor coupled and without it. In Fig. 5.3 the same process is repeated for α=0.99.

As it can be seen from these occillograms, the change in one of the system parameters, namely the moment of inertia does not effect the performance of the system.

If the output of the system at time (t+1) is exactly equal to the predicted value, the relation between the time constant and the value of α can be obtained as below.

The reference trajectory was

$$y_r(t+1) = \alpha y_r(t) + (1-\alpha)c , \qquad (5.16)$$

and from equation (1.16) the system equation is

$$\dot{y}_m(t) = -\frac{1}{t_m} y_m(t) + \frac{K_m}{t_m} e_a(t) \qquad (5.17)$$

the discrete equivalent of above equation is

$$y_m(t+1) = e^{-\frac{T}{T_m}} y_m(t) + K_m(1 - e^{-\frac{T}{T_m}}) e_a(t) . \qquad (5.18)$$

Thus

$$\alpha = e^{-\frac{T}{T_m}} , \qquad (5.19)$$

therefore, for $\alpha = 0.99$, $T_m = 0.7959$ sec. and for $\alpha = 0.967$, $T_m = 0.2384$ sec.

Although the time constant of the step response is not expected to be equal to that determined by the value of α due to continual updating, it is seen that the time constants of the responses are very close to those values.

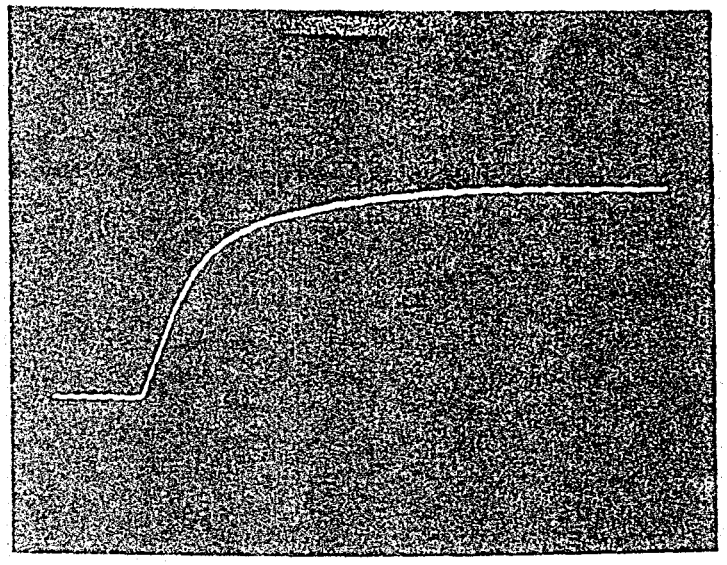
The oscillograms of Fig. 5.4 show the tracking behavior of the system. For these tests, the set speed is changed continuously for 3 seconds and then it is kept constant and finally for the last 3 seconds c is decreased in a linear fashion. In other words, a time optimal set speed change is applied. As discussed in chapter three, the maximum acceleration that the system is capable of is 11.27 rad/sec^2. Starting with an initial speed of 600 rpm, the set speed was therefore changed by 0.86 rpm per sampling interval until a final set speed of 923 rpm was reached. The same procedure was followed for deceleration. It is seen that from the oscilograms of Fig. 5.4 that this reference trajectory is followed very closely. The response for $\alpha = 0.99$ are not shown because for this value

of α ,the tracking behavior is not very good due to the fact that a change of 0.86 rpm in the set speed can only be followed with a time constant of 0.967. If slower time constants are applied,the tracking error accumulates at the end.

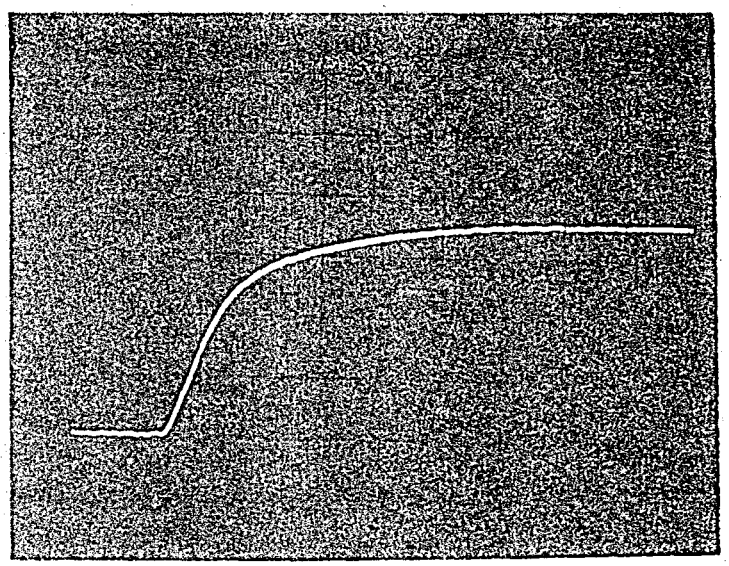
Fig. 5.4.b shows that the tracking performance does not change appreciably due to the change in J. The slight difference in Fig.5.4.a and b are due to the fact that the time of acceleration, the time during which set speed was kept constant and the time of deceleration were decided upon by a simple observation of the CRT spot in the horizontal axis.

Fig. 5.5.a and b show the responses of the system under loaded conditions. A comparison of these with those of Fig.5.2.a and b show that the loading does not effect the performance especially for the time varying set speed case.

56



(a)



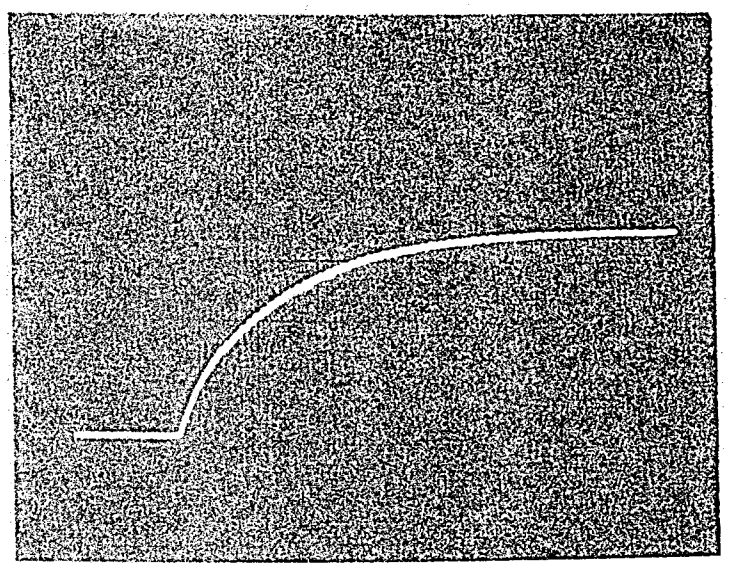
(b)

Fig. 5.2 The time response of the motor speed

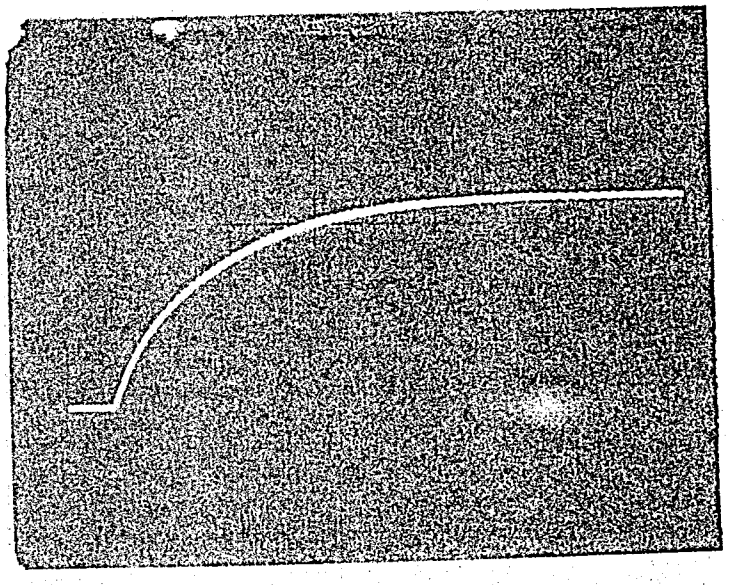
α =0.967 1 div.=0.2 sec = 200 rpm

(a) without induction motor

(b) with induction motor.



(a)



(b)

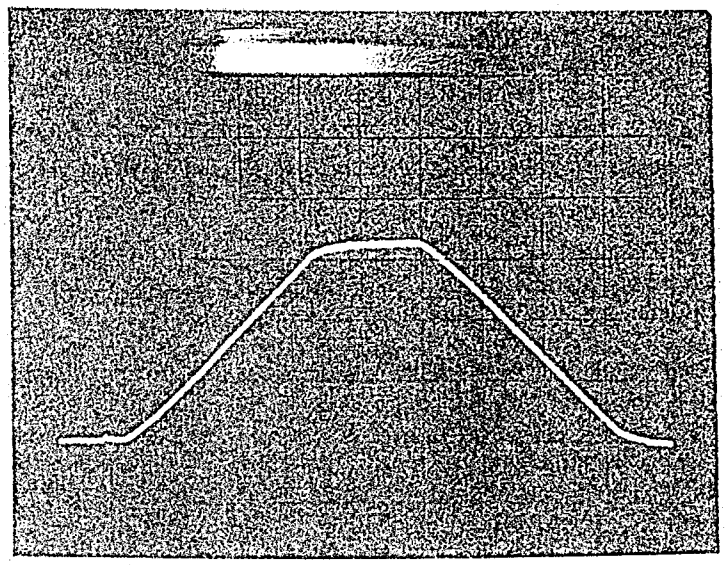
Fig. 5.3 The time response of the motor speed

α =0.99 1 div.= 0.5 sec = 200rpm

(a) without induction motor

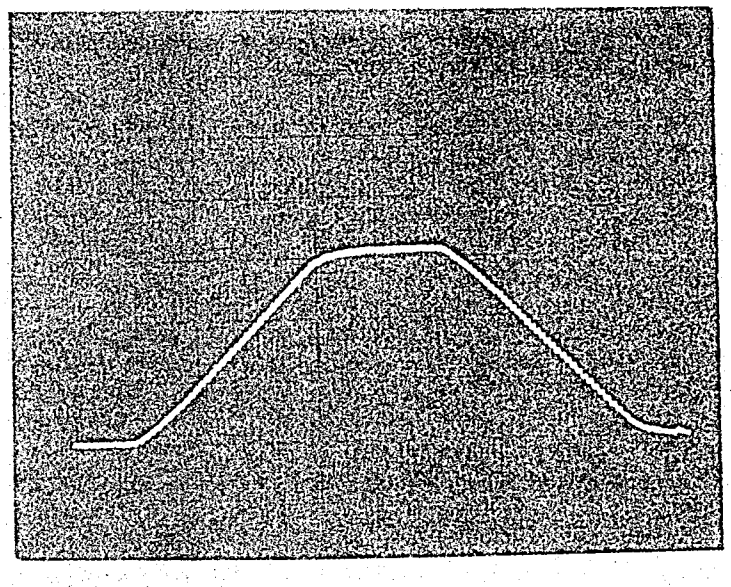
(b) with induction motor.

600 rpm



(a)

600 rpm



(b)

Fig.5.4 The tracking behavior of the motor speed

α =0.967 1 div.= 1 sec = 100 rpm

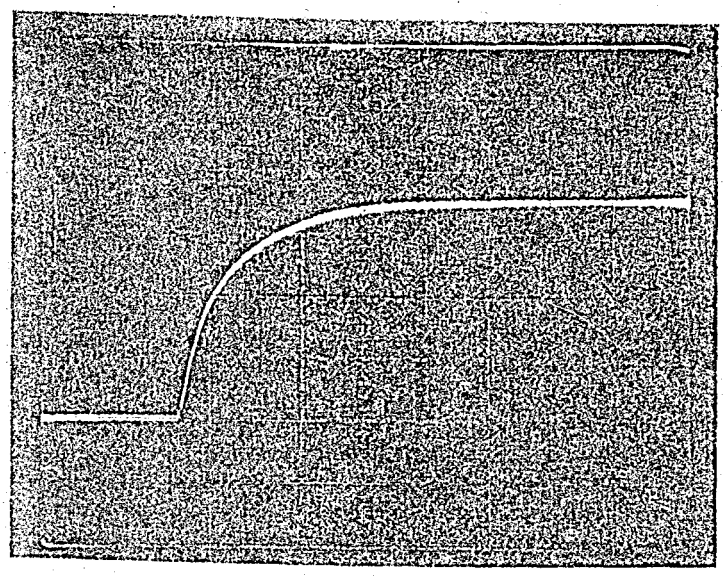
5 sec. acceleration, 2 sec. max. speed and

5 sec. deceleration

(a) without induction motor

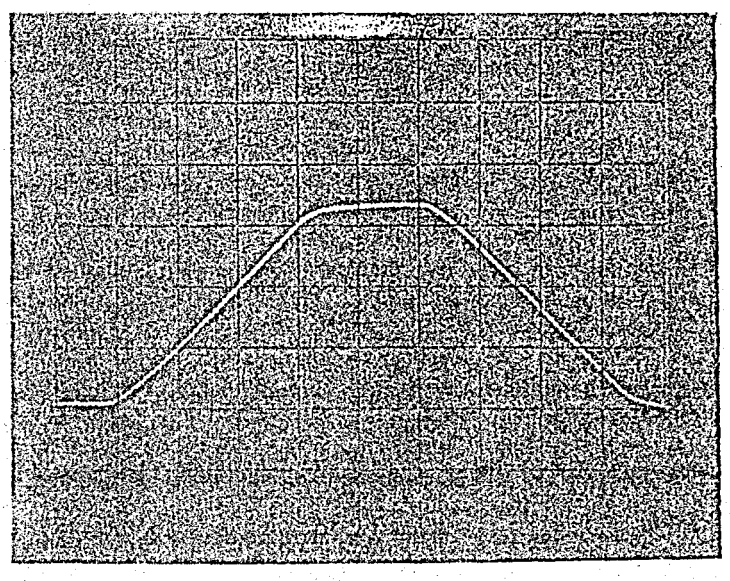
(b) with induction motor.

700 rpm



(a)

600 rpm



(b)

Fig. 5.5 The behavior of the system under load

(a) α=0.967 1 div.= 0.2 sec.= 200 rpm

(b) α=0.967 1 div.= 1 sec.= 100 rpm.

VI. CONCLUSIONS AND SUGGESTIONS FOR FURTHER WORK

In the thesis, the speed control of a seperately excited dc motor is achieved by using Model Algorithmic Control. The starting point of this theory is that the process to be controlled should be relatively slow and the controller must have fast digital computing and memory facilities.

The frequency of microcomputer used in the implementation of the theory is 2.5 MHz, than the controller can be considered as a relatively slow one. As it is explained, to overcome this disadvantage, the coefficients in the control equation is grouped as shown in Fig. 4.3

The coefficients are approximated twice, first because of truncation errors and than because of the error caused by grouping them. In spite of these approximations, it is shown that the steady state error can be considered as zero. Also it is shown that the system is robust, and therefore Model Algorithmic Control is a good candidate for cases in which robust performance is required.

The D/A converter used in the hardware is an 8-bit one and the inverted output of which is -10 V. The speed reference input of the thyristor unit is -15 V. This brings an important constraint in the optimal input sequence. This constraint does

not effect the system stability and robustness, but limits the value of α , and the time constant of the process output.This is especially important for tracking case, because the tracking performance of the Model Algorithmic Control is better for small software time constants. If a proper D/A converter could be available, the region of selectable α would have been wider, and the system would have been operated with smaller time constants.

The performance obtained by the work described in the thesis is very promising, warranting a more through theoretical analysis of the Model Algorithmic Control theory as applied to the dc motor speed control. The theorical analysis presented in this thesis neglects the effects of current control loop. The constraint imposed by the set maximum value of the current and the effects of the P controller of the current loop should be investigated to gain more insight to the design of a Model Algorithmic Controller.

The performance obtained suggest that the application of Model Algorithmic Control to manipulator control could alleviate the problems that arise in such a system due to high degree of coupling between the links and due to variations in the pay load. The Lagrange-Euler approach to formulate the robot arm dynamics result in a set of highly coupled differential equations making its control difficult. However for the application of Model Algorithmic Control theory only an impulse response representation is required and this can be obtained rather easily.The variations in the impulse response due to different arm positions should not effect the arm trajectory as demostrated in the thesis.

APPENDIX A

SOFTWARE OF MODEL ALGORITHMIC CONTROL FOR CONSTANT SET SPEED

```
21ØØ42          LD  HL,42ØØH
363C            LD  (HL),3CH    ;Coefficients of
23              INC HL           control equation
3633            LD  (HL),33H
23              INC HL
3629            LD  (HL),29H
23              INC HL
362Ø            LD  (HL),2ØH
23              INC HL
3618            LD  (HL),18H
23              INC HL
3611            LD  (HL),11H
23              INC HL
36ØC            LD  (HL),ØCH
23              INC HL
36Ø9            LD  (HL),Ø9H
23              INC HL
36Ø6            LD  (HL),Ø6H
23              INC HL
36Ø4            LD  (HL),Ø4H
23              INC HL
36Ø3            LD  (HL),Ø3H
23              INC HL
36Ø2            LD  (HL),Ø2H
```

```
23                      INC   HL
3601                    LD    (HL),01H
23                      INC   HL
363C                    LD    (HL),3CH      ; α

D300            MAIN:   OUT   (00),A        ;Apply speed ref.
210040                  LD    HL,4000H       input of thyristor
0605                    LD    B,05           unit
97                      SUB   A             ;Calculation of
57                      LD    D,A            b1x*(t-1)+.....+
86              L1:     ADD   A,(HL)         bNx*(t-N)
D23A30                  JP    NC,M1
14                      INC   D
23              M1:     INC   HL
10F8                    DJNZ  L1
210042                  LD    HL,4200H
4E                      LD    C,(HL)
5F                      LD    E,A
CD3232                  CALL  MULT
226232                  LD    (TEMP1),HL

210540                  LD    HL,4005H
0605                    LD    B,05H
97                      SUB   A
57                      LD    D,A
86              L2:     ADD   A,(HL)
D25B30                  JP    NC,M2
14                      INC   D
23              M2:     INC   HL
10F8                    DJNZ  L2
210142                  LD    HL,4201H
4E                      LD    C,(HL)
5F                      LD    E,A
CD3232                  CALL  MULT
226432                  LD    (TEMP2),HL

210A42                  LD    HL,420AH
0608                    LD    B,08H
97                      SUB   A
57                      LD    D,A
86              L3:     ADD   A,(HL)
```

```
D2753Ø          JP    NC,M3
14              INC   D
23        M3:   INC   HL
1ØF8            DJNZ  L3
21Ø242          LD    HL,42Ø2H
4E              LD    C,(HL)
5F              LD    E,A
CD3232          CALL  MULT
226632          LD    (TEMP3),HL

211240          LD    HL,4Ø12H
Ø6Ø8            LD    B,Ø8
97              SUB   A
57              LD    D,A
86        L4:   ADD   A,(HL)
D28F3Ø          JP    NC,M4
14              INC   D
23        M4:   INC   HL
1ØF8            DJNZ  L4
21Ø342          LD    HL,42Ø3H
4E              LD    C,(HL)
5F              LD    E,A
CD3232          CALL  MULT
226832          LD    (TEMP4),HL

211A4Ø          LD    HL,4Ø1AH
Ø6ØA            LD    B,ØAH
97              SUB   A
57              LD    D,A
86        L5:   ADD   A,(HL)
D2A93Ø          JP    NC,M5
14              INC   D
23        M5:   INC   HL
1ØF8            DJNZ  L5
21Ø442          LD    HL,42Ø4H
4E              LD    C,(HL)
5F              LD    E,A
CD3232          CALL  MULT
226A32          LD    (TEMP5),HL
```

```
21244Ø              LD    HL,4Ø24H
Ø6ØA                LD    B,ØAH
97                  SUB   A
57                  LD    D,A
86          L6:     ADD   A,(HL)
D2C33Ø              JP    NC,M6
14                  INC   D
23          M6:     INC   HL
1ØF8                DJNZ  L6
21Ø542              LD    HL,42Ø5H
4E                  LD    C,(HL)
5F                  LD    E,A
CD3232              CALL  MULT
226C32              LD    (TEMP6),HL

212E4Ø              LD    HL,4Ø2EH
Ø6ØA                LD    B,ØAH
97                  SUB   A
57                  LD    D,A
86          L7:     ADD   A,(HL)
D2C53Ø              JP    NC,M7
14                  INC   D
23          M7:     INC   HL
1ØF8                DJNZ  L7
21Ø642              LD    HL,42Ø6H
4E                  LD    C,(HL)
5F                  LD    E,A
CD3232              CALL  MULT
226E32              LD    (TEMP7),HL

21384Ø              LD    HL,4Ø38H
Ø6ØD                LD    B,ØDH
97                  SUB   A
57                  LD    D,A
86          L8:     ADD   A,(HL)
D2C73Ø              JP    NC,M8
14                  INC   D
23          M8:     INC   HL
1ØF8                DJNZ  L8
21Ø742              LD    HL,42Ø7H
```

```
4E                      LD      C,(HL)
5F                      LD      E,A
CD3232                  CALL    MULT
227Ø32                  LD      (TEMP8),HL

21454Ø                  LD      HL,4Ø45H
Ø6ØD                    LD      B,ØDH
97                      SUB     A
57                      LD      D,A
86              L9:     ADD     A,(HL)
D2C93Ø                  JP      NC,M9
14                      INC     D
23              M9:     INC     HL
1ØF8                    DJNZ    L9
21Ø842                  LD      HL,42Ø8H
4E                      LD      C,(HL)
5F                      LD      E,A
CD3232                  CALL    MULT
227232                  LD      (TEMP9)

21524Ø                  LD      HL,4Ø52H
Ø6ØD                    LD      B,ØDH
97                      SUB     A
57                      LD      D,A
86              LA:     ADD     A,(HL)
D2CB3Ø                  JP      NC,MA
14                      INC     D
23              MA:     INC     HL
1ØF8                    DJNZ    LA
21Ø942                  LD      HL,42Ø9H
4E                      LD      C,(HL)
5F                      LD      E,A
CD3232                  CALL    MULT
227432                  LD      (TEMPA),HL

215F4Ø                  LD      HL,4Ø5FH
Ø6ØF                    LD      B,ØFH
97                      SUB     A
57                      LD      D,A
86              LB:     ADD     A,(HL)
```

```
D2CD3Ø          JP    NC,MB
14              INC   D
23        MB:   INC   HL
1ØF8            DJNZ  LB
21ØA42          LD    HL,42ØAH
4E              LD    C,(HL)
5F              LD    E,A
CD3232          CALL  MULT
227632          LD    (TEMPB),HL

216E4Ø          LD    HL,4Ø6EH
Ø6ØF            LD    B,ØFH
97              SUB   A
57              LD    D,A
86        LC:   ADD   A,(HL)
D2CF3Ø          JP    NC,MC
14              INC   D
23        MC:   INC   HL
1ØF8            DJNZ  LC
21ØB42          LD    HL,42ØBH
4E              LD    C,(HL)
5F              LD    E,A
CD3232          CALL  MULT
227832          LD    (TEMPC),HL

217C4Ø          LD    HL,4Ø7CH
Ø6ØF            LD    B,ØFH
97              SUB   A
57              LD    D,A
86        LD:   ADD   A,(HL)
D2d13Ø          JP    NC,MD
14              INC   D
23        MD:   INC   HL
1ØF8            DJNZ  LD
21ØC42          LD    HL,42ØCH
4E              LD    C,(HL)
5F              LD    E,A
CD3232          CALL  MULT

227A32          LD    (TEMPD),HL
```

```
21ØØØØ          LD    HL,ØØØØ
ED5B6232        LD    DE,(TEMP1)
19              ADD   HL,DE
ED5B6432        LD    DE,(TEMP2)
19              ADD   HL,DE
ED5B6632        LD    DE,(TEMP3)
19              ADD   HL,DE
ED5B6832        LD    DE,(TEMP4)
19              ADD   HL,DE
ED5B6A32        LD    DE,(TEMP5)
19              ADD   HL,DE
ED5B6C32        LD    DE,(TEMP6)
19              ADD   HL,DE
ED5B6E32        LD    DE,(TEMP7)
19              ADD   HL,DE
ED5B7Ø32        LD    DE,(TEMP8)
19              ADD   HL,DE
ED5B7232        LD    DE,(TEMP9)
19              ADD   HL,DE
ED5B7432        LD    DE,(TEMPA)
19              ADD   HL,DE
ED5B7632        LD    DE,(TEMPB)
19              ADD   HL,DE
ED5B7832        LD    DE,(TEMPC)
19              ADD   HL,DE
ED5B7A32        LD    DE,(TEMPD)
19              ADD   HL,DE
226Ø32          LD    (TEMP),HL
DBØØ            IN    A,(ØØ)      ;Read the actual
5F              LD    E,A          speed
DBØ1            IN    A,(Ø1)
E6Ø3            AND   Ø3H
57              LD    D,A
2A5E32          LD    HL,(SETSPD)
37              SCF
3F              CCF
ED52            SBC   HL,DE
EB              EX    DE,HL
21ØD42          LD    HL,42ØDH    ;Calculate the
4E              LD    C,(HL)       error term
```

```
CD4C32                  CALL   MULT1
ED5B6Ø32                LD     DE,(TEMP)
19                      ADD    HL,DE          ;Calculate x*(t)
FC                      LD     A,H
32FF3F                  LD     (3FFFH),A
Ø192ØØ                  LD     BC,ØØ92H       ;Update the past
219Ø4Ø                  LD     HL,4Ø9ØH        history
11914Ø                  LD     DE,4Ø91H
ED5B                    LDDR
CD1BØØ                  CALL   GETKY
FE21                    CP     21H            ;SHIFT + ! pressed ?
CA1D32                  JP     Z,MODIFY       ;Yes
FE49                    CP     49H            ;I pressed ?
CA2632                  JP     Z,INCRS        ;Yes
FE44                    CP     44H            ;D pressed ?
CA2632                  JP     Z,DCRS         ;Yes
C32C32                  JP     IN
97          MODIFY:     SUB    A              ;Clear the speed
DCØØ                    OUT    (ØØ),A          reference input
Ø691                    LD     B,91H           of thyristor unit
21FF3F                  LD     HL,3FFFH
77          LR:         LD     (HL),A
23                      INC    HL
1ØFC                    DJNZ   LR
C3ØØ6Ø                  JP     BRK
21ØE03      INCRS:      LD     HL,Ø3ØEH       ;Increase the set
225E32                  LD     (SETSPD),HL     speed
C32C32                  JP     IN
2187Ø1      DCRS:       LD     HL,Ø187H       ;Decrease the set
225E32                  LD     (SETSPD),HL     speed
3AFF3F      IN:         LD     A,(3FFFH)
C33139      MULT:       JP     MAIN
21ØØØØ                  LD     HL,ØØØØ        ;6 bit multiplier
Ø6Ø6                    LD     B,Ø6H           with zero initial
CBØ9        LC2:        RRC    C               value
D23D32                  JP     NC,LC1
19                      ADD    HL,DE
CB13        LC1:        RL     E
```

```
CB12                    RL D
1ØF4                    DJNZ  LC2
Ø6Ø3                    LD    B,Ø3H
CB3C        LC3:        SRL   H
CB1D                    RR    L
1ØFA                    DJNZ  LC3
C9                      RET
21ØØØØ      MULT1:      LD    HL,ØØØØ     ;7 bit multiplier
Ø6Ø7                    LD    B,Ø7H        with zero initial
CBØ9        LB2:        RRC   C            value
D25732                  JP    NC,LB1
19                      ADD   HL,DE
CB13        LB1:        RL    E
CB12                    RL    D
1ØF4                    DJNZ  LB2
C9                      RET

                SETSPD: DEFW Ø3ØEH
                TEMP  : DEFW ØØØØ
                TEMP1 : DEFW ØØØØ
                TEMP2 :DEFW  ØØØØ
                TEMP3 : DEFW ØØØØ
                TEMP4 : DEFW ØØØØ
                TEMP5 :DEFW ØØØØ
                TEMP6 : DEFW ØØØØ
                TEMP7 : DEFW ØØØØ
                TEMP8 : DEFW ØØØØ
                TEMP9 : DEFW ØØØØ
                TEMPA : DEFW ØØØØ
                TEMPB : DEFW ØØØØ
                TEMPC : DEFW ØØØØ
                TEMPD : DEFW ØØØØ
                BRK   :EQU   6ØØØH
                GETKY :EQU   ØØ1BH
                       END
```

APPENDIX B

SOFTWARE OF MODEL ALGORITHMIC CONTROL FOR TIME VARYING SET SPEED

The difference between the software of constant set speed and time varying set speed is in the progpam sections labelled by INCRS and DCRS in Appendix A. If we change the below program with the programs labelled INCRS and DCRS in the software of constant set speed, the system runs as a tracking type model algorithmic control.

```
.                       .
.                       .
.                       .
C3ØØ6Ø                  JP    BRK

211ØØØ          INCRS:  LD    HL,ØØ1ØH   ;The increase at
ED5BBE32                LD    DE,(SPD)    the set speed
19                      ADD   HL,DE
22BE32                  LD    (SPD),HL
37                      SCF
3F                      CCF
11CØ61                  LD    DE,61CØH
ED52                    SBC   HL,DE
```

```
F23332          JP    P,PR      ;If reached the max.
C33A32          JP    NOR        set speed ?
EB        PR:   EX    DE,HL
22BE32          LD    (SPD),HL
C34932          JP    LL
2ABE32    NOR:  LD    HL,(SPD)
CB65            BIT   4,L
CA4532          JP    Z,XX
C34932          JP    LL
112ØØØ    XX:   LD    DE,ØØ2ØH
19              ADD   HL,DE
Ø6Ø5      LL:   LD    B,Ø5H
37        YY:   SCF
3F              CCF
CB1C            RR    H
CB1D            RR    L
1ØF8            DJNZ  YY
22BC32          LD    (SETSPD),HL
C38A32          JP    IN

111ØØØ    DCRS: LD    DE,ØØ1ØH  ;The decrease at the
2ABE32          LD    HL,(SPD)   set speed
37              SCF
3F              CCF
EB52            SBC   HL,DE
22BE32          LD    (SPD),HL
37              SCF
3F              CCF
11ØØØØ          LD    DE,ØØØØ
ED52            SBC   HL,DE
FA7332          JP    M,NN      ;If reached zero
C37A32          JP    OK         set speed ?
EB        NN:   EX    DE,HL
22BE32          LD    (SPD),HL
C37D32          JP    OO
2ABE32    OK:   LD    HL,(SPD)
Ø6Ø5      OO:   LD    B,Ø5H
37        MM:   SCF
3F              CCF
```

```
CB1C                    RR      H
CB1D                    RR      L
10F8                    DJNZ    MM
22BC32                  LD      (SETSPD),HL
3AFF3F          IN:     LD      A,(3FFFH)
C33130                  JP      MAIN

210000          MULT:   LD      HL,0000
   .                    .       .
   .                    .       .
   .                    .       .
```

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