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CONTENTS

Improving the reliability of electric drives by using a.c. motors

P ATKINSON, BSc, ACGI, CEng, MIEE, MIERE and P R SAVAGE, BSc
Department of Engineering, The University of Reading

SYNOPSIS Ideally, electric drives for robotics applications should be robust and substantially maintenance-free. In general, owing to ease of control d.c. motors in position control applications have hitherto been preferred to a.c. motors in such applications. Apart from the commutatorless d.c. motors, which have some special problems of their own, d.c. motors have commutators and brushes which make them more prone to faults and more expensive to manufacture and maintain than a.c. induction motors. The simplicity of construction, low cost, and freedom from maintenance problems would appear to make the two-phase a.c. induction motor an ideal candidate for robotics applications. However until recently it was not as easy to control as the d.c. motor. Nowadays, the signal processing power of the microprocessor together with the convenience of power handling of the VMOS FET has removed this objection. Recent work by the authors has shown that it should be possible to marry micros to 400 Hz induction motors giving feedback positional servos which are both cheap and reliable. The paper describes the principles of the design of such systems and illustrates these principles with a practical example based on a 50 Hz motor.

1 INTRODUCTION

Pneumatic, hydraulic and electric drives have all been used for industrial robots. Each form of drive has robotic applications in which its particular advantages offer the most satisfactory economical and technical solution.

However in this paper it is the authors' objective to focus attention on electric drives which offer a number of distinct possibilities. In first generation robots with electric drives, two types of motor have been most popular - the stepping motor and the permanent magnet d.c. commutator motor. Stepping motors have been used for small low cost robots in which positional feedback is not considered essential. They suffer from various disadvantages, particularly when used on open loop. These include limited torque, speed and power, low efficiency particularly at low speeds, and possible loss of positional accuracy if the inertial or drag of the load changes or the pulse repetition frequency is incorrectly chosen. Accordingly robots which aim to achieve higher positional accuracy (e.g. the Unimation Puma, ASEA IR66 and the Lansing LIR10 (manufactured by Hitachi)) use permanent magnet commutator d.c. motors with positional feedback. D.C. permanent magnet commutator motors are highly efficient but themselves suffer from certain disadvantages. The main disadvantage is that the brushes must be replaced at frequent intervals (perhaps every 100 days for a motor operating efficiently with a high duty cycle). Furthermore with the lower cost ferrite permanent magnets, demagnetization due to accidental overheating can also cause failure. The use of rare-earth (samarium cobalt) permanent magnets (1) obviates this particular problem but increases the cost of the motor. Commutator motors also possess a high level of Coulomb friction and stiction which is especially high in printed armature motors (e.g. greater than 10% of maximum torque) due to their method of construction. Stiction is the cause of stick-slip motion in tracking servos which is a severe disadvantage in robotics applications. The sparking at commutators is also a cause of radio-frequency interference and a fire-hazard in combustible atmospheres.

Brushless d.c. motors do not yet appear to have found application in robot drives but recent developments (2) indicate that many of their problems are now being solved. When constructed with rare-earth permanent magnets they are highly reliable machines but very expensive and relatively difficult to control. Furthermore the torque-to-inertia ratios obtainable cannot be expected to match those which can be achieved using disc or cup armature d.c. machines.

Switched variable reluctance motors (3) are very robust but are moderately expensive and difficult to position control making them unattractive for robotics applications.

A.C. induction motors are by far the most widely used motors for constant speed applications (4,5). They are extremely robust requiring virtually no maintenance over the whole of their long working lives. They are also very cheap to manufacture. However they are much more difficult to speed or position control than d.c. motors. Both two and three phase motors have been used for speed control applications but the controls have primarily relied on voltage control based on the phase-control of thyristors (6). In the more demanding applications these have been found unsuitable and both voltage and frequency control have been used. These have resulted in very expensive controllers based on solid-state d.c. link inverter drives (5).

For robotics applications, position control is necessary and for this the two-phase induction motor is the most suitable form of a.c. drive.

A further advantage of the induction motor in this application is the very low level of stiction and Coulomb friction it exhibits. The motor has a stator which is a pair of windings (fitted into the laminated-iron stator structure) mutually at right-angles in space. The rotor can take one of several forms: squirrel cage, drag cup, or solid iron. Torque is produced by supplying the stator windings with a.c. voltages in phase quadrature at frequencies of normally 50 Hz, 60 Hz, or 400 Hz.

The use of rotary and linear induction motors, especially for position control applications, has been very much less in recent years than it was formerly. In the period between 1940 and about 1965, small two-phase rotary induction motors were extremely popular in position control applications, especially when incorporated in 'all a.c.' (i.e. carrier type) servomechanisms. The general control strategy (7) was to excite the reference winding from a constant frequency constant voltage supply and to apply quadrature control signals to the control winding. In these applications the amplifier was usually a thermionic valve device. Naturally this form of control was extremely inefficient, and relied heavily for its popularity on two features:

(i) a.c. supplies are usually available, and
(ii) a.c. amplifiers are drift free.

This second feature was, of course, a prime reason for the use of all-a.c. systems compared with systems using d.c. motors which require at least d.c. power amplification. The nominal disadvantage of d.c. power amplification is the associated amplifier drift which is bound to cause undesirable positional (or velocity) offsets.

The advent of low-cost high-power semi-conductor devices in the early 1960's had some marked effects on the situation. High-power d.c. amplifiers based on bipolar silicon planar transistors reduced the drift problem and the use of armature-controlled d.c. motors became so popular that a.c. motors practically disappeared in position control applications. Indeed even in many high power speed control applications d.c. machines under thyristor control became extremely popular.

However since two-phase induction motors are very much more reliable than d.c. motors, are potentially very much cheaper, have low stiction and Coulomb friction and are sparkless they would appear to be very good candidates for a new initiative, particularly for robotics applications.

This paper describes the problems encountered in using two-phase a.c. induction motors for positional servo applications and how they may be overcome using modern signal processing techniques and semiconductor power output devices.

2 CONTROL OF TWO-PHASE INDUCTION MOTORS

2.1 Historical Background

As explained in section 1, traditional control of the two-phase induction motor relied essentially on constant sinusoidal excitation of the reference winding with an amplitude-modulated, carrier suppressed signal in quadrature applied to the control winding. The control signals were usually obtained by class A tuned power amplification of the low power a.c. error signal. Not only were these methods of control highly inefficient in their own right, but they also led to the development of very inefficient motors. It is of course imperative for position control applications to design a motor to give a high torque at standstill and also to obtain a torque/speed characteristic with a negative slope. In the induction motor this may be achieved by using a rotor with high resistance. However in the traditionally controlled two-phase induction motor in which one winding is continuously excited, it is also necessary to design the motor to avoid 'single-phasing' (i.e. the gratuitous production of torque when the control winding is unexcited). This was normally achieved by increasing the rotor resistance to a value well beyond that required to achieve maximum torque at standstill. This resulted in a characteristic with very poor efficiency at maximum rated power output (e.g. 15% compared with about 50% for d.c. motors of the same power rating). Consequently two-phase induction servo motors were used for applications with less than 20 watts output, although 200 watt motors have been commercially available.

It would therefore appear that it was the traditional means of control that dictated the design of the old-fashioned types of two-phase induction motor, rendering them and their power controllers far less efficient than strictly necessary.

In a more recent development (8) it was necessary in the particular application to produce inputs for both windings from a d.c. source. It then became evident that in the interest of efficiency, it would be advisable to excite both windings from control signals in quadrature. In order to avoid class A power amplification, 'six-step' drives (Fig. 1a) in quadrature to both windings are advisable. Of course this involves the use of controlled d.c.-a.c. inverters and at the time the original work was performed (1970) it was necessary to use thyristors. There were commutation problems associated with these of some difficulty. Also the production of control sequences involved the use of considerable hard-wired logic.

Nowadays with improved power VMOS FET's commercially available together with the signal processing power of microcomputers, it is quite possible to develop a low cost, highly efficient controlled d.c.-a.c. inverter. The use of such a scheme will also allow new designs of two-phase induction motors to be developed with higher starting torque and efficiency at maximum power output.

2.2 Modern Methods of Control

There are essentially <u>three</u> principal techniques for controlling induction motors, viz:

(i) two-phase fixed frequency with amplitude control of one phase, the other phase being continuously driven on full amplitude;

(ii) two-phase fixed frequency control with amplitude control of both phases, reverse torque being achieved by means of phase reversal of one phase;

(iii) two- or three-phase variable frequency and variable voltage control.

Any of the above may be implemented using sinusoidal or switched voltage waveforms; however the use of switched voltages allows the use of highly efficient controllers and would be preferred unless other factors militate against its use.

Method (ii) has the advantage over method (i) that it gives less heating (i^2R) loss at low torque but at low speeds will give a considerably greater i^2R loss than a d.c. motor of equal rating. Method (iii) offers the advantage of low i^2R loss but is much more difficult to implement and appears to be best suited to speed control applications (4) than the position control requirements of robot servos. Hence although the motor heating of method (ii) is worse than method (iii), giving rise to a poorer torque/weight ratio, the electronic implementation is much simpler and accordingly is the method selected by the authors in their present development.

2.3 Types of Control Waveforms for the Two-Phase Induction Motor

When operating an induction motor from d.c. supplies it must first be decided which switching waveform is the most suitable for the particular application. The two most obvious contenders being the six-step waveform (Fig. 1a) and the pulse-width modulated (PWM) waveform (Fig. 1b). The main advantage of the six-step waveform is that the power devices only have to switch at line frequency, which is typically 50 Hz or at most 400 Hz in practice. The main advantage of the PWM technique is that harmonic elimination may be used to generate a more nearly sinusoidal output. In speed control applications this will be very advantageous because at low speeds it is necessary to drive at low supply frequencies; the presence of high harmonic content is then a cause of rough running because of the production of alternating torques due to the multiplicative action of induction motors. There are many kinds of PWM waveform which can be generated and used in any of the methods of control (i), (ii), or (iii) summarized in section 2.2. However the indiscriminate use of non-optimized PWM waveforms may cause difficulties when fast response is required. It may be demonstrated for example that certain kinds of non-optimum PWM waveforms actually cause reverse torques to be generated during demand for a transient torque. Alternatively if these difficulties are to be avoided the use of optimum PWM waveforms requires the use of complex, fast acting algorithms which are not simple to implement.

Due to the difficulties encountered in the use of PWM it was therefore decided to use the six-step waveform in this application. To control the two-phase motor it is necessary to drive its two stator windings with six-step waveforms in quadrature as illustrated in Fig. 2. The conduction angle ϕ must be varied from 0 to π radians as a function of the positional error signal. Unfortunately the static torque T_{gs} is not a linear function of conduction angle, but it may be shown that:

$$T_{gs} = k_1(1 - \cos\phi) - k_2(1 - \cos3\phi) \qquad (1)$$

where k_1 and k_2 are constants dependent on the supply voltage, the frequency, and the constants of the motor. In practice k_2 is small compared with k_1 (typically $k_2 = k_1/15$). However the characteristic is very flat for small values of ϕ, giving the motor an inherently 'soft start' in control applications. Since stiffness near the origin of the torque/command characteristic is essential for servo applications, it is vital to linearize this by incorporating a counter nonlinearity in the controller. In real-time microcomputer-controlled systems this may be most conveniently implemented in the form of a 'look-up table'.

3 CONTROL PHILOSOPHY

3.1 Overview

The implementation of a positional feedback control system based on the two-phase induction motor is complex in principle but need not be unduly expensive. In its most primitive form a suitable control scheme (Fig. 3) may be divided into:

(a) the two-phase induction motor and gear-box;
(b) positional transducer;
(c) differencing device;
(d) controller;
(e) nonlinear static torque compensator;
(f) waveform timers;
(g) d.c.-a.c. inverters.

The signal processing involved in (c), (d), (e) and (f) can be achieved by either analog or digital electronic circuitry. However an analog implementation becomes complicated and costly thus militating in favour of a digital solution. The digital implementation can be achieved in a number of ways, including hardwired logic, programmable logic array, and microprocessor. The microprocessor appears to offer the cheapest and most flexible solution, particularly if the waveform timing is executed externally by means of readily available programmable timer chips.

The choice of a microprocessor as the main signal processing element dictates the format of the incoming signals. The angular command must be produced in digital form and so must the feedback signal. The number of bits used to represent these signals must be fixed on the basis of the required resolution. This decision must also influence the form of positional transducer used. Again there is considerable choice here, the following being possibilities:

(i) absolute shaft encoder;
(ii) analog transducer with analog to digital converter (ADC);
(iii) incremental shaft encoder.

Although absolute shaft encoders are very expensive (about £1000 for 12-bit resolution over a single turn) they do perhaps offer the ideal technical solution. An analog transducer with an ADC, while possibly providing the cheapest solution, will not usually give the combination of high resolution and large mechanical range required in many servo applications. The incremental encoder fitted directly to the motor shaft rather than the load is a very popular solution in robotics applications and is capable of higher resolution. However, there is no measure of absolute position using this device, so that its associated pulse counter must be reset to an arbitrary reference position at switch-on. Furthermore positional inaccuracy must accrue wherever 'pulses' are lost or gained due to electrical interference so that it is advisable to reset to reference moderately often. Absolute accuracy will also of course be affected by inaccuracies in the gear-box and coupling misalignment.

In the pilot study described in this paper, the model servomechanism was constructed using a microprocessor as the main control element and a 12-bit absolute shaft encoder as the positional transducer to ensure high resolution and absolute knowledge of position. In robotics applications the incremental transducer might well be preferred in the interests of economy.

3.2 Controller Characteristics

The simplest form of control which could be used in this application is proportional control. However this form of control is very limited because it is not possible to tailor the dynamic response to meet anything but the most primitive specifications. In the robotics application it is generally necessary for the servo to track ramp-like command signals with minimal (hopefully zero) steady-state error and to nullify offsets due to constant load disturbances. In general it is possible to achieve these requirements by using integral (I) action combined with proportional (P) control to give stable operation. Unfortunately the 'P + I' control defined above places limitations on the form of dynamic response which can reasonably be achieved. The traditional method of improving the dynamic response is to use a derivative (D) term in addition. However this is very noise-enhancing in practice and rarely produces the expected degree of improvement because computed approximations to the derivative of positional error are inaccurate anyway. A vastly superior technique for gaining design flexibility is the use of a local velocity feedback loop based on a tachogenerator coupled directly to the motor shaft. This not only allows the dynamic response to be arbitrarily adjusted within wide limits but also regulates against the effect of dynamic load disturbances and ameliorates the effects of gearbox backlash, Coulomb friction and stiction (9).

Accordingly in the model servomechanism it was decided to use P + I control with a local velocity feedback loop based on a d.c. tachogenerator with an 8-bit ADC. To avoid the appearance of inband noise due to aliasing, it was necessary to include an anti-aliasing filter of suitable cut-off frequency within the local loop. The arrangement is outlined in Fig. 4.

The nonlinear static torque characteristic of the motor (equation 1) is not readily solved for ϕ in terms of T_{gs}, but values of T_{gs} for a set of values of ϕ can be computed off-line with ease for a given machine with known values of k_1 and k_2. Values of ϕ can be then be stored in a set of store locations, the addresses of which represent the required torque T_{gs}. In operation, the correct value of ϕ for a given value of T_f is rapidly obtained by 'looking up' the value of ϕ at an address dictated by T_f.

4 PRACTICAL IMPLEMENTATION OF MODEL SERVO-MECHANISM

4.1 General Arrangement

To demonstrate the use of a microprocessor in the control of a two-phase induction motor it was decided to use a currently available Evershed 3 W, 50 V, 50 Hz two-phase squirrel-cage induction motor modified in-house to carry a miniature d.c. tachogenerator for local feedback. The main positional feedback was implemented using a proprietary 12-bit absolute shaft encoder.

A Motorola 6809 8-bit microprocessor was chosen as the heart of the microcomputer controller. This device was chosen for the application because a development system complete with a monitor and a cross-assembler based on a PDP 11/44 running the UNIX operating system is generally available in the Electronics Laboratories at Reading University (10). Moreover its operating speed is quite adequate for this 50 Hz application. The main program which implements the 'P + I' algorithm and torque compensation is held in ROM with scratch pad RAM for storing the variables. The 12-bit command input is entered via a keypad and input to the computer through a peripheral interface adapter (PIA) in two bytes and the main positional feedback from the absolute shaft encoder is input in two bytes via a second PIA. The local velocity feedback signal is interfaced via an 8-bit ADC (ZN448). The anti-aliasing filter is a single lag 355 op-amp circuit with a time constant equal to 0.22 seconds (which gives a break frequency well below half the sampling frequency).

The software problems associated with generating the timing signals to drive the inverter are greatly simplified by the use of the Motorola 6840 timer chip (PTM). This chip in fact contains three separate 16-bit programmable timers which make it ideal for producing both the real-time interrupt clock and the control sequences for both inverter drives. Each timer appears to the microprocessor as if it were a memory location and as such can be programmed by loading the data at the appropriate address. When the data is loaded the counter output goes to logic high and the counter proceeds to count down to zero. When zero is reached the timer output goes low. It is thus possible to generate two pulse-width modulated waveforms from two counters which can then be used directly to control the actions of the inverter drives. The primary timing of the waveform edges at phase angles of 0 and $n\pi/2$ (where n is odd) must of course be separately timed. This is achieved by using the interrupt clock in the appropriate manner. The notional arrangement for the hardware is illustrated in Fig. 5.

4.2 The D.C.-A.C. Power Inverter

The implementation of an inverter can be achieved by a variety of current technologies. It is possible to use thyristors which are cheap and reliable but are difficult to commutate. It is also possible to use bipolar transistors but these have the major disadvantage that they require a heavy base current in the ON state. Accordingly it was decided to use power VMOS FET's which have very low drive requirements and can also be commutated directly by means of low-power gate switchings.

Various topologies were considered but the bridge circuit was selected because of the many advantages it offers. It reduces the component costs because a single-ended power supply can be used (allowing the V rating of the VMOS FET's to be halved); also free-wheeling of the inductive load is built into the operation due to the integral reverse source-drain diode, so reducing power supply ripple to negligible proportions and at the same time eliminating the need for RC snubbers. The actual circuitry used together with its isolation and drive circuitry is illustrated in Fig. 6.

It consists of four VMOS FET's T1, T2, T3, T4 connected in complementary pairs between the positive d.c. power line V_s and zero volts. The switching sequence is as follows. T1 is switched ON by a primary timing signal on the gate derived from the peripheral interface adapter (PIA); a very short time later T4 is switched ON from an auxiliary timing signal derived from the PIA and the PTM via a NAND gate and optical isolator. This ensures that the switching losses are confined to T4 and can be minimised by the use of well-defined high speed switching. It should be noted that it was found necessary to drive the VMOS FET's gates from the D469 Quad High Current Drivers in order to hold the gates as near zero volts as possible in the OFF condition. If the gates are driven by a relatively high impedance source (e.g. directly from the optical isolators) the existence of the drain-source and gate-source capacitances (typically 80pF and 400 pF respectively) may cause T2 to be switched ON unintentially while T1 is fully conducting, thus causing a short circuit across the supply and resultant malfunction.

When the positive current cycle is complete (i.e. when $\omega t = \pi$ for the control phase), T4 is gated OFF by the PTM via the logic circuitry and optical isolator. Similarly on the negative half cycle T3 is switched ON first at $\omega t = \pi$ and immediately after T2 is gated ON until $\omega t = \pi + \phi$ whereupon it is gated OFF.

It was found most convenient to operate the bridge with T1 switched ON from 0 to π and T2 switched ON from π to 2π. The in-built diodes thus allow the inductive load to freewheel on both the positive and negative half cycle after $\omega t = \phi$ and $t = \pi + \phi$ respectively.

It should be noted that in order to avoid exceeding the low source-gate maximum voltage on T1 and T3 (about 20 V), it was necessary to limit the gate swing to 12 V by connecting the zero volts terminal of the D469 to $(V_s - 12)$ volts. To provide simple interfaces to the PIA which provides logic signals at 0 and + 5 V and to prevent fault conditions in the inverter from damaging the microcomputer, optical isolators have been used.

4.3 Real Time Generation of Switching Signals

In order to generate the correct sequence of switching signals to operate the bridge inverters, the 6809 interrupt system is employed. This is driven by the external 100 Hz interrupt clock, which generates interrupts on both positive and negative clock edges by appropriately programming the PIA CA1 and CA2 inputs. The 6809 is equipped with microprograms to handle an interrupt request automatically.

Once the computer is interrupted, the program jumps directly to the starting address of the interrupt service routine. The interrupt inputs are derived from the PIA through CA1 and CA2 so as to enable identification of the source.

The phase counter to be updated can be determined by examining the state of bit 7 (the sign bit) of the control register of the PIA. A negative clock edge causes this to be set, so signalling that the control phase counter is to be altered. Transistor T1 or T2 for the control phase is switched on, followed by the selection of the appropriate lower transistor T4 or T2 via the PIA output, and the switching logic.

Finally the timer is loaded with the appropriate data which starts the timing sequence, a dummy read is performed on peripheral register A to clear the interrupt mask, and the microprocessor resumes the normal program flow.

4.4 The P + I Algorithm

The command position $\theta_i(t)$, output position $\theta_o(t)$ and velocity feedback voltage $v(t)$ are sampled on interrupt every 10 ms. When the interrupt routine is complete the computer determines a sampled error signal $V_e(k)$ at the kth sampling time given by

$$V_e(k) = K'_p(\theta_i(k) - \theta_o(k))$$

where K'_p is the number of bits per radian of shaft rotation. The computer then operates on this signal with a digital P + I algorithm to produce

$$V_c(k) = K\{V_e(k) + \frac{T_s}{T_i} \sum_{m=\infty}^{k} V_e(m)\}$$

where K is a proportional gain factor
T_i is the integral action time
T_s is the sampling interval (10 ms).

It should be noted that the integral term is computed using the Euler rectangular approximation which is quite accurate enough for this purpose.

Since the software is written effectively in machine code and the computer word length is only 8-bits, certain precautions have to be built into the program to ensure that the algorithm does not go awry. The coefficients K and T are stored as

16 bit words in two bytes to give sufficient resolution and all the multiplications are designed to produce 16-bit words. Of course the individual terms and the total $V_c(k)$ have to be checked for overflow after each program cycle. In the event of an overflow in either component, the component is set to its maximum positive or negative value, depending on the direction of the overflow. Should $V_c(k)$ overflow, then this is also set to its maximum positive or negative value.

The velocity feedback signal $Gv(k)$ is then subtracted from $V_c(k)$ and a check is made on possible overflow of the resultant 16-bit signal which is now proportional to the demanded value of torque. Again in the event of an overflow the demanded torque is set to the maximum positive or negative value.

It should be noted that the control action takes place one sampling period (i.e. 10 ms) after the output and output rate signals have been converted, so that a pure time delay of this amount is introduced into both control loops and this must be allowed for when designing the control parameters.

4.5 Control System Parameter Design

Having designed and tested the hardware and software and shown it all to be running satisfactorily in accordance with expectations it was then necessary to choose the various adjustable control system parameters (in essence G, K and T_i) such that the closed-loop behaviour of the system could be tested. It was decided therefore to design the system to a moderately tight specification as follows:

(i) zero offset in response to constant disturbances;
(ii) zero steady-state error in response to a constant velocity input;
(iii) step response with time to peak equal to (0.2 ± 0.05)s and overshoot $(20 \pm 5)\%$.

The first two of these specifications are satisfied by the use of the P + I action already designed into the compensating algorithm. To determine approximate values of the parameters required to ensure that the third specification is met we may use an approximate design method ignoring pure time delays and based on a small signal continuous linear equivalent model of the 'plant'. The plant may be imagined to consist of the linearisation algorithm, output interface, power d.c.-a.c. inverter, motor and gear-box. The two signal transducers and the ADC can be handled separately. A plausible transfer model of this plant takes the form:

$$\frac{\Theta_o(s)}{T_f} = \frac{K_p}{s(1 + sT_m)}$$

where $\Theta_o(s)$ is the Laplace transform of the output angular position and $T_f(s)$ is the Laplace transform of the continuous equivalent to the binary input to the linearisation algorithm. The values of K_p and T_m for the system used were found by applying a small step input to the input of the linearisation algorithm and by observing the output angular velocity of the open-loop plant on a storage oscilloscope. The measure of angular velocity used was the tachogenerator output.

Routine control system design (11) can then be used to determine suitable values of G, K, T_i.

4.6 Performance Verification

It was possible to demonstrate elimination of offsets due to constant load disturbances by making a manual displacement of the output shaft. The mark-space ratio of the 50 Hz actuating signal could then be observed gradually building up to a maximum with the associated increase in motor torque. At the time of writing, a ramp response test had not been completed.

The step response of the system was observed on a storage oscilloscope and it was found that the system response had a time to peak of approximately 0.2 s (c.f. (0.2 ± 0.05)s) which is within specification, and an overshoot of about 20% (c.f. $(20 \pm 5)\%$) which is also within specification.

5 RELIABILITY AND COST

It is not possible to quote quantitative figures for the reliability of the system described in this paper. The two-phase induction motor is extremely reliable so long as its power dissipation is kept within the specification. It is of course capable of handling very considerable transient overloads with impunity and can work in ambient temperatures of 125°C. Eventually (e.g. after at least 30000 hours) the bearings will wear out but these can be replaced easily by a mechanical technician. The life is thus limited only by the number of times this operation may be performed in an economical way. There are no brushes or commutator to wear and no magnets to be demagnetized. They are ideally suited to the worst of industrial environments.

The microprocessor and its support chips are generally thought to have a life in the order of 150 years. Power VMOS FET's are relatively new devices (about four years old) so that although they appear to be very rugged devices no significant in-service experience is yet available. The N-channel devices are better established than the P-channel devices and give high reliability so long as the gate drive circuitry has been properly designed. The inverter scheme is of course not intrinsically limited to the use of VMOS FET's; although they appear to the authors to be the best device for the application, bipolar power devices (for example Darlington transistors (12) or gate-turn off thyristors) could also be used.

The reliability of the gear-box and the transducers are not in dispute in this paper as these devices are necessary for any electric motor drive with feedback, and must be chosen carefully for any application where high reliability in the field is important.

All precision control systems are expensive and engineers must forever search for ways to reduce their cost without significantly reducing the technical performance or the reliability. D.C. machines are very expensive and not highly reliable. Induction motors are very much less expensive to produce and much more reliable. By controlling on both windings as described in this paper it is possible to improve the efficiencies considerably (typically from 15% to 40%) by correct design.

The electronic control system described in this paper is relatively inexpensive involving low-cost microprocessor chips for the low-power signal processing and VMOS FET's for the high power output. Power VMOS FET's are becoming relatively cheaper all the time and are very much cheaper to interface to computers than conventional bipolar transistor and gate turn-off thyristors.

The positional transducer used in the model servo is an expensive absolute shaft encoder. In robotics applications an incremental encoder is often preferred to reduce costs. It should be noted however that for precision position control, the incremental encoder is not a reliable method of monitoring absolute position, and it may well be better to consider the expensive but totally reliable alternative in situations where high accuracy over long cycle times is required.

6 CONCLUSION

Two-phase induction motors are demonstrably more reliable in servo applications than d.c. commutator motors. Their low efficiency when designed for application in traditional control schemes can largely be overcome by the use of systems which apply quadrature control signals of equal amplitude to both windings. Unfortunately this produces static torque characteristics which are highly nonlinear and requires significant amounts of signal processing. These difficulties can be overcome by using a microprocessor for the essential signal processing and linearisation. The microprocessor is especially useful in this application because traditional servo compensation can also be achieved within the same hardware using some simple additional software. The advantage of the scheme described is that it can be easily modified to handle motors designed to work at different frequencies and with widely different power ratings.

The paper summarises the hardware and software used successfully in a model servomechanism to demonstrate the use of these machines. The model utilizes complementary pairs of VMOS FET's in the d.c.-a.c. power inverter which interfaces the microprocessor to the motor and these appear to offer the simplest means of power control available for this purpose.

Although the model uses a 50 Hz motor, precisely the same hardware could be used to operate a 400 Hz motor with small software modifications.

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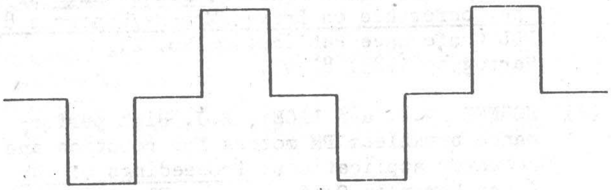

Fig 1a Typical six-step waveform

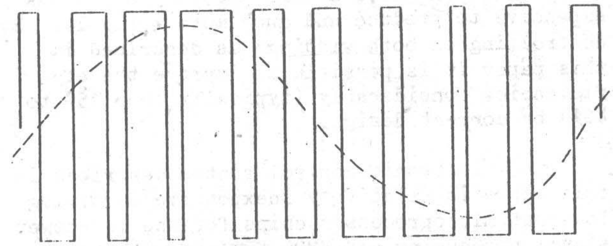

Fig 1b Typical PWM waveform

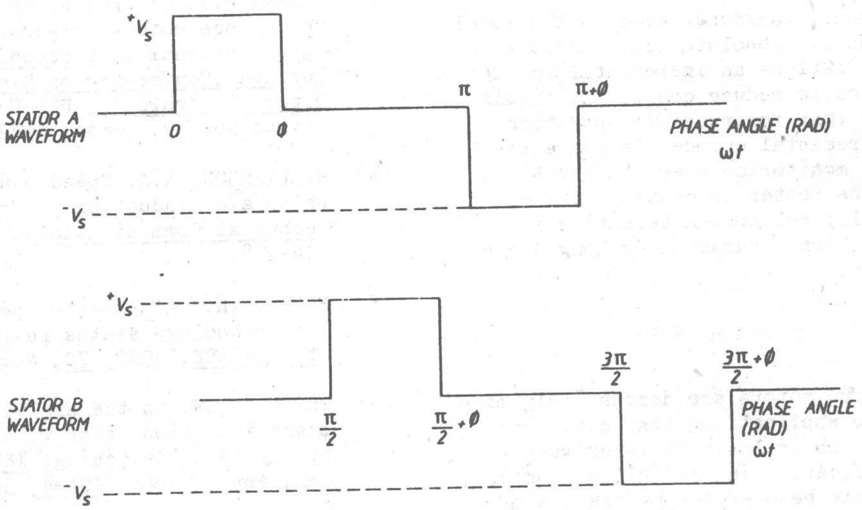

Fig 2 Required power waveforms (ϕ running from 0 to π)

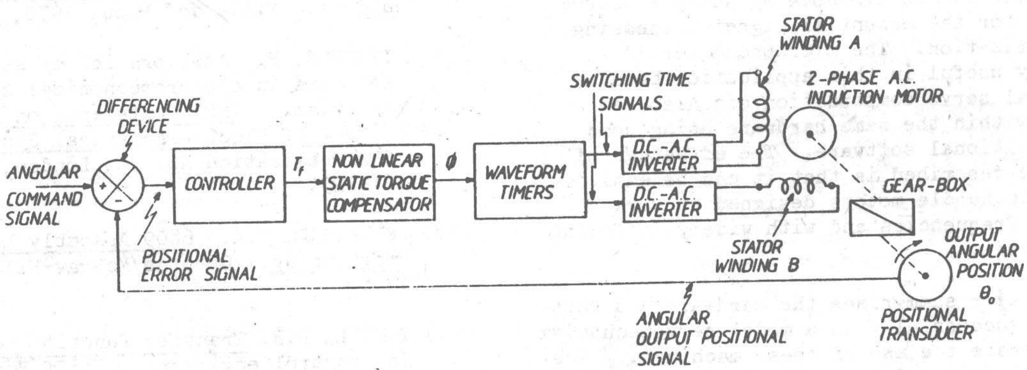

Fig 3 Notional position control system for two-phase a.c. induction motor

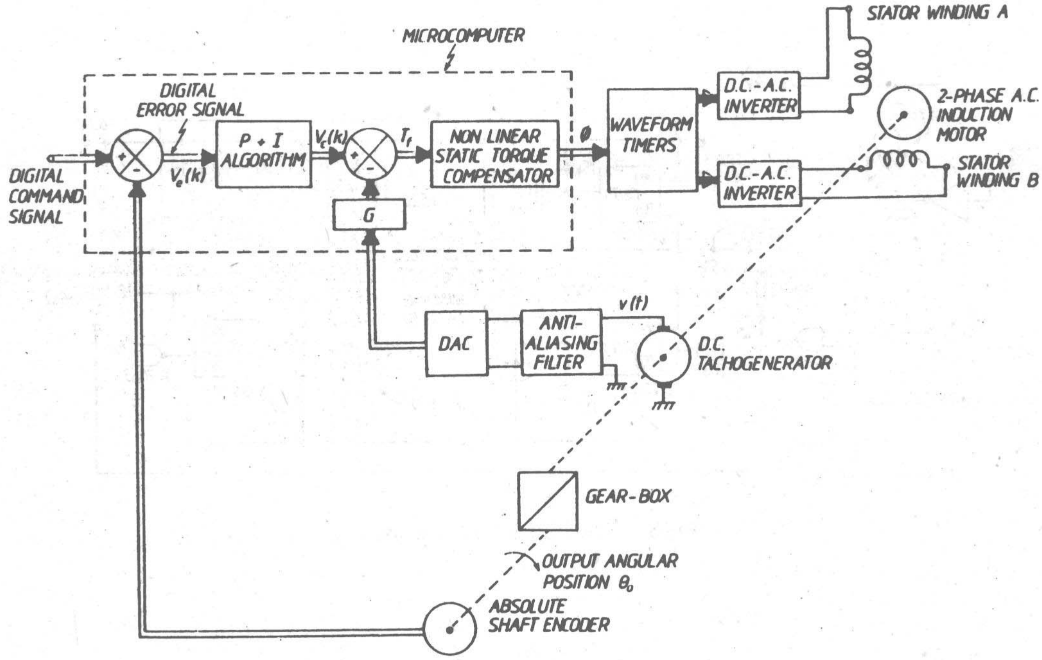

Fig 4 System diagram for induction motor servomechanism

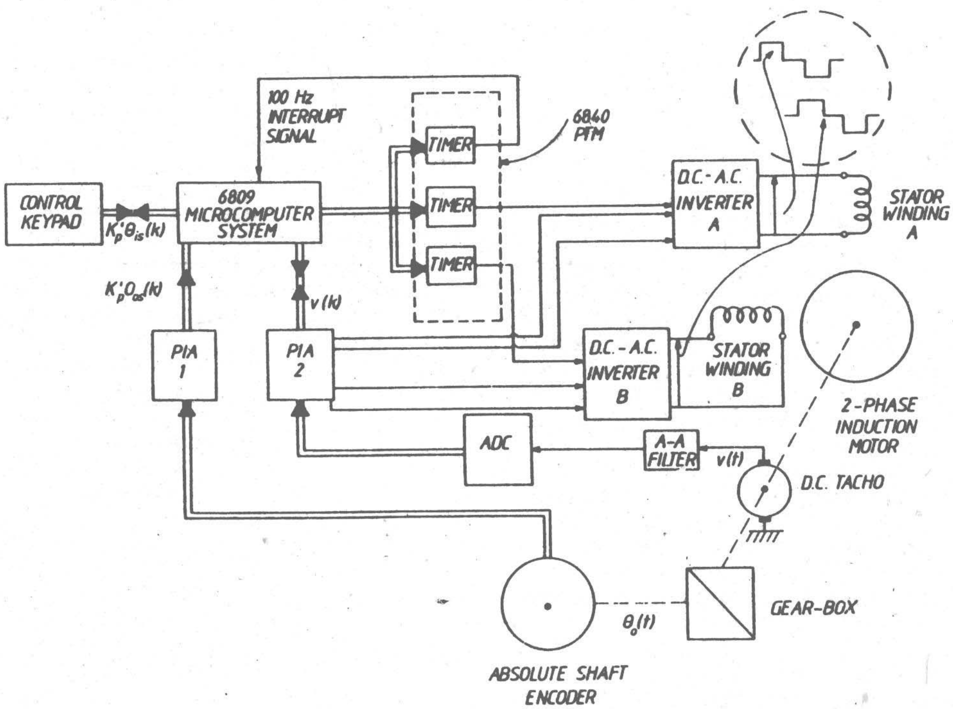

Fig 5 Notional block diagram for proposed control system

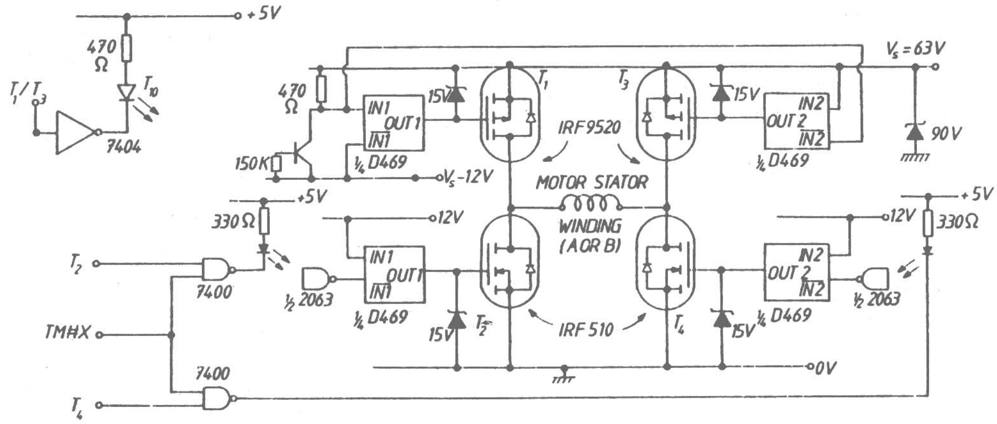

Fig 6 VMOS FET d.c.-a.c. inverter

The search for cost-effective robotic assembly

P B SCOTT, BSc, ACGI, MBCS and T M HUSBAND, PhD, CEng, FIProdE
Centre for Robotics and Automated Systems, Imperial College of Science and Technology, London SW7

SYNOPSIS Some of the theoretical bases of a new methodology for cost-effective robotic assembly are presented, together with a condensed example of how the methodology might be employed in practice.

1 INTRODUCTION

What appears to be required by industry, as an aid to cost-effective robotic assembly, is 'ready made experience'. At Imperial College we are currently synthesising a broad methodology for robotic assembly: partly by attempting to discover patterns and consensus within the wealth of existing, yet frequently insular, industrial and academic experience; and partly by seeking to theoretically uncover some of the underlying principles which govern robotic assembly, and then test these against observed reality.

One of the fundamental advantages of such an approach is that it allows the use of 'case lore' and 'rules of thumb' to guide the design process in a fashion which rapidly homes in on an optimal strategy. In this way, throughout the process, successively accurate 'ball-park' figures can be estimated and their consequences evaluated. Whereas such heuristics would naturally evolve in due course over the next decade, it is felt that the economic pressure for cost-effective robotic assembly demands that the 'natural' aggregation of case lore be accelerated by means of active research. In this way, the global stages of:

System Design
Analysis
Evaluation

can to an extent become instead:

PREDICTION
Analysis
Evaluation
Directed Detailed Design

In other words, analysis and evaluation are accomplished using predicted data, such as 'ball-park' figures. Only after an encouraging evaluation of a proposed robotic-assembly project need the detailed design of the flexible assembly system (FAS) be finalised. Even this latter stage is not performed unaided, because by then a broad skeleton of specifications has been determined towards which the detailed design can be aimed. Provided that the prediction stage is relatively straightforward and yet sufficiently accurate, the whole process through to evaluation is easy enough to rerun to be very responsive to changes. Similarly, comparatively little time will have been wasted should an alternative method of assembly to robots be suggested by the evaluation stage. This procedure is in stark contrast to the usual current approaches to design, analysis and evaluation of robotic-assembly projects (1,2,3). The serious optimisation problems inherent in such approaches have been expressed by the authors in earlier papers (4,5).

2 BASES OF THE NEW METHODOLOGY

2.1 Segregated cost rates

An important distinction made in the Imperial College methodology is between those costs incurred by the flexible, reusable portions of an FAS, and that expenditure (including design and labour costs) which is 'fixed' to a particular product design and is of no further benefit when that product is no longer being assembled. A further distinction is made between the costs of the robotic equipment and the remaining 'flexible' equipment. For example, a vibratory bowl-feeder, which might at first be thought of as 'fixed' equipment, is in fact reusable (though nonrobotic) because it is usually only the tooling of the feeder which requires replacement during a product design change. The vibratory bowl itself is as 'reusable' as a robot, and may still be present in the FAS a decade later. Yet such conventional (though 'flexible') automation tends to be substantially faster than purely robotic equipment, so a distinction between the two must be made (6).

Unit assembly cost thus comprises the expenditure (including such costs as design) associated respectively with robotic equipment, reusable nonrobotic equipment and equipment 'fixed' to a given product (and not used except when assembling batches of that product). In addition to these three costs, however, there is frequently a fourth 'hidden' overhead cost which may nevertheless be substantial. The major contributing factor to this overhead is often the inventory cost of work-in-progress (WIP). Conventional techniques of assembly, whether they involve dedicated automation or manual labour, tend to result in a significant value of components and partly-finished products simply waiting for assembly. Use of a number of integrated robots for assembly may result in more efficient processing of components, resulting in fewer parts being actually 'tied up in the system' at any given one time. As a result, the WIP inventory will cost less, so less of the company's capital will be tied up. The potential impact of robotics on such assembly overheads requires that such costs usually be considered separately from the three other costs previously mentioned.

The above distinctions can be employed in the form of Segregated Cost-Rates (SCR) as follows:

$$\text{unit cost} = t\,(r_{RO} + r_{RE} + r_{F} + r_{O})$$

where t is the cycle time taken to assemble a product

r_{RO} is the cost per unit time of the robotic equipment

r_{RE} is the (nonrobotic) reusable equipment cost rate

r_{F} is the cost rate of 'fixed' equipment

r_{O} is the cost rate of overheads such as inventory

It should be noted that the cost rate for fixed equipment (the 'fixity' cost rate) might well be calculated simply by dividing all the expenditure tied to a given product (such as special tooling, redesign, and so on) by the sum of all the periods the FAS was set up to assemble that product (that is, the total time that the 'fixed equipment' was actually of benefit). Calculation of robotic or reusable-equipment cost rates, however, would require account to be taken of the 'cost of capital' tied up in the equipment over a long lifetime of maybe a decade. Discounting procedures such as 'annual equivalent cost' would be appropriate for this adjustment.

2.2 SAM-times

There are immediate benefits simply from stating the SCR-formula. For example, it indicates the answer to the question, 'Is it better to employ a cheaper robot or a faster robot - which has most effect on unit cost?' When it is remembered that an FAS is always paced by the robot, not the parts-feeding equipment, then, clearly, it is better to use a robot which is twice as fast rather than half as expensive, because a halved cycle time is multiplied by the whole of the brackets, whereas a halved robot cost rate is just one of the terms within the brackets. This conclusion agrees with those of far more complex models (7). Yet, it is worth noting here that the 'maximum end-effector velocity' supplied by manufacturers' specifications may be a poor indicator of actual cycle time (8).

This is because a robot joint does not, of course, instantaneously reach its slew rate, but must instead accelerate to it, maintain it for a time, and then start to decelerate in time to stop at a desired location. The rate of acceleration and deceleration will be dependent on many factors, such as the mass of the robot arm, the power of the drive, and the control strategy, so it is unlikely to be the same for different models of robot. If a robot movement is a small one (as is common during assembly), then a given joint may not have reached its slew rate before it has to decelerate again. A robot with a 'snappier' response (though maybe a lower actual slew rate) might be able to perform the same movement quicker because it in fact reached a higher speed, more rapidly, than the potentially faster alternative. It may even temporarily have reached its slew rate before having to decelerate.

In order to reflect such robot 'snappiness', the authors employ the concept of a 'standard assembly movement' time (SAM-time). This corresponds to the empirically measured time required by a given robot to perform a series of movements designed to be representative of a typical robotic component-assembly cycle. The speeds and distances travelled during a SAM will vary for different models of robot, as speed settings and distances are specified in terms of the maximum available speed setting and the furthest reach of the robot arm respectively. Using SAM-times, a rough estimate can be easily obtained for the likely assembly time of a product:

assembly time = SAM-time x number of operations x correcting factor

where the correcting factor (which for a typical assembly equals unity) is dependent on such aspects as FAS layout and the nature of the assembly operations.

2.3 The principle of fixity

Returning to the implications of segregated cost-rates: the SCR-formula also suggests that as the number of <u>different</u> product designs to be assembled by an FAS increases, so the fixity expenditure (comprising equipment and costs 'fixed' to a given product) should be reduced, and the amount spent on equipment and services common to all batches should be increased to compensate. However, because it is quicker, and so cheaper for large runs, to assemble using dedicated automation, there is a balance to be maintained (illustrated in Figure 1) dependent upon how frequent the design changes are to be. [There is also the implication that the flexible portions of the FAS should employ as high a proportion of reusable yet nonrobotic equipment as possible, because (for instance for parts orienting) this will be faster than <u>purely</u> robotic approaches.]

The tradeoff between increased production rate and increased wastage when an FAS is reconfigured to assemble a new product design, is crystallised in the proposed 'Principle of Fixity', which states that:

> For optimal robotic assembly, the proportion of the total cost of an FAS which is due to expenditure 'fixed' to a particular product is directly related to the total time, spent throughout the life of the system, assembling that given product.