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# FLUX SENSING AND CONTROL OF STATIC AC DRIVES BY THE USE OF FLUX COILS

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## Abstract

Application of air gap flux sensing coils to the control of static AC induction motor drives is discussed. Optimum flux coil configurations are developed for the most frequently used slot combinations. Practical results are presented which verify the validity of the design approach. Application of flux signals to the control of practical static AC drives is included with emphasis placed on state-of-the-art techniques.

## Introduction

Many manufacturing processes require the use of an adjustable speed rotating power source. Typical examples of such systems include pulp and paper drives, chemical processes, machine tool drives, pump drives and dynamometers. Although these applications may vary widely in horsepower rating, speed range, environment and speed of response, they are all traditionally served by the same motor drive, namely the Ward-Leonard DC drive. With the development of the thyristor in 1958, advances in solid state technology have had a significant effect in the motor drive industry. The phase controlled rectifier bridge has to a large extent replaced the DC generator armature supply in the Ward-Leonard arrangement. This AC/DC converter is but one of a wide variety of power converters which have recently been developed. In particular AC/DC/AC and direct AC/AC converters have made possible AC motor solutions to the adjustable speed problem. By adjusting the stator frequency by means of a frequency converter, the speed of an induction motor can be changed while maintaining the high efficiency inherent in low slip operation. These new static AC drives are being increasingly employed in applications where their inherent features are useful, for example in high horsepower situations, where very precise or very high speeds are required or in adverse environments.

Recently, attention has focused on control algorithms required to maximize the speed of response for a given application. Use of flux as a feedback signal has been found to overcome many of the drawbacks encountered with more conventional feedback variables such as rotor slip frequency, stator current or terminal voltage. Although Hall probes are sometimes used as flux sensors, the unfavorable environment generally discourages their use in favor of voltage search coils connected to high quality electronic integrators. Since flux sensing as a means of control is relatively recent, many of the practical problems involved in securing an accurate flux signal are not well known. This paper presents a detailed discussion of the concept of flux sensing by search coils and the problems involved in its implementation. In particular, the effects of slot harmonics and saturation are discussed. A design scheme is presented which attempts to minimize these effects. Implementation of flux signals in several practical motor control schemes is described.

## Approach to Flux Sensing

A consistent theme in the design of any AC machine is the layout of sinusoidally distributed armature windings in order to realize a sinusoidal distribution of air gap flux. However this situation can only be approached in practice. The most important limiting factor is the presence of discrete stator and rotor slots which causes variations in the air gap permeance. Slot permeances combine with the winding MMF's to produce flux harmonics in the air gap. These harmonics tend to degrade the quality of flux measurement concerned only with the fundamental component. If P denotes the number of pole pairs, S the number of stator slots and R the number of rotor slots, it is well known[1] that the stator slots set up fields having S-P and S+P pole pairs rotating forward and backward respectively at synchronous speed. The rotor slots set up fields having R+P and R-P pole pairs. The rotor poles rotate at a speed $1 \pm (1-s)R/P$ relative to the stator, where s is the rotor slip. Being a function of rotor speed, the negatively rotating rotor slot permeance harmonic introduces a particularly annoying effect since the frequency of oscillations passes through zero when $s = 1-P/R$. In many small machines, rotor slot harmonics are removed by skewing. Unfortunately, this approach is feasible only for cast rotor machines, generally below 100 HP. Skewing also introduces an extra "skew" leakage reactance, which is normally undesirable in high horsepower AC drives.

Slot harmonics are not the only source of measurement error. Because the stator slots are discrete, the ideal sinusoidal distribution of armature conductors can only be approximated. Additional odd-ordered rotating MMF harmonics are introduced in addition to the fundamental component. Most predominant of these "phase belt" harmonics are the negatively rotating 5th and positively rotating 7th harmonics. Finally, operation at high flux levels causes magnetic saturation. Third harmonics and multiples thereof begin to appear in the air gap flux wave. All of these effects tend to induce voltages in search coils inserted into the gap and thus produce errors in the measurement of air gap flux. Since stator slot harmonics rotate at synchronous speed they will induce voltages with the same frequency as the fundamental. Fortunately, when the teeth are not highly saturated these voltages are proportional to the fundamental component of MMF so that they can simply be considered as part of the effective magnetizing reactance. Of the remaining harmonics, most severe are the R+P and R-P rotor slot harmonics and the third harmonic flux wave produced by magnetic saturation and it is desirable to minimize their effect by interconnection of search coils.

When possible, it is best to locate the flux coils during manufacture. In this case the flux coils can simply be incorporated into the stator coil assemblies. The coil and winding span will then be identical. Unfortunately, such an approach is generally not possible for most applications. Although the theory to follow is directed primarily to flux coil design for existing machines the work can be readily modified to accomodate flux coil arrangements which follow conventional winding patterns.

When flux coils are inserted in an existing machine the most straightforward approach is probably to locate concentric coils symmetrically around the magnetic axis to be measured. Depending upon the winding pitch a magnetic axis can be located either over a stator tooth or stator slot. However, it is assumed initially that four concentric windings are arranged symmetrically around a stator tooth as shown in Fig. 1. It can be shown that the phase shift in the voltage induced in the coils between successive slots is 360 P/S° for the P field, 360·3P/S°for the third harmonic (3P) field, 360° for the S field and 360 R/S° for the R field.[1] From this result it can be determined that the phase angle between successive stator slot pitches is 360 (R-P)/S for the R-P field and 360(R+P)/S for the R+P field. The total voltage produced in all three coils is proportional to the turns A,B,C, and D.

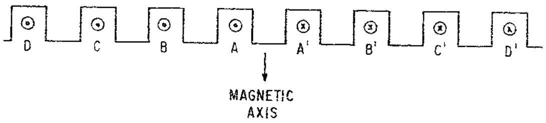

Fig. 1 Four Concentric Coils Centered Over a Stator Tooth.

For the P, S+P and S-P fields the voltage induced should be as large as possible. Hence,

$$D \sin (7\pi P/S) + C \sin (5\pi P/S) + B \sin (3\pi P/S) + A \sin (\pi P/S) = K \quad (1)$$

For the R-P field the voltage measurement should be zero, whence

$$D \sin [7\pi(R-P)/S] + C \sin [5\pi(R-P)/S] + B \sin [3\pi(R-P)/S] + A \sin [\pi(R-P)/S] = 0 \quad (2)$$

Similarly for the R+P field

$$D \sin [7\pi(R+P)/S] + C \sin [5\pi(R+P)/S + B \sin [3\pi(R+P)/S] + A \sin [\pi(R+P)/S] = 0 \quad (3)$$

and the 3P field

$$D \sin (21\pi P/S) + C \sin (15\pi P/S) + B \sin (9\pi P/S) + A \sin (3\pi P/S) = 0 \quad (4)$$

In general, these four equations define five unknowns namely A, B, C, D and K so that an infinite number of solutions are possible. However, Eqs. 2-4 can be written as

$$(D/A) \sin [7\pi(R-P)/S] + (C/A) \sin [5\pi(R-P)/S] + (B/A) \sin [3\pi(R-P)/S] = -\sin [\pi(R-P)/S] \quad (5)$$

$$(D/A) \sin [7\pi(R+P)/S] + (C/A) \sin [5\pi(R+P)/S] + (B/A) \sin [3\pi(R+P)/S] = -\sin [\pi(R+P)/S] \quad (6)$$

$$(D/A) \sin (21\pi P/S) + (C/A) \sin (15\pi P/S) + (B/A) \sin (9\pi P/S) = -\sin (3\pi P/S) \quad (7)$$

Equations 5-7 can now be solved for the ratios D/A, C/A and B/A. Substituting these results in Eq. 1 will yield the relative gain K/A for the optimum winding configuration. The ratio K/A must clearly be nonzero for a solution to be valid.

When the coils are located concentrically around a stator slot rather than a tooth the phase shift from the magnetic axis to Coil A the innermost coil is 360 P/S° rather than 180 P/S°. The phase shift between successive slots is the same as before. The fundamental voltage induced in the four coils can now be written

$$D \sin (8\pi P/S) + C \sin (6\pi P/S) + B \sin (4\pi P/S) + A \sin (2\pi P/S) = K \quad (8)$$

The equations for the R-P, R+P and 3P fields are phase shifted similarly.

Table 1 shows the optimum weighting factors for a number of popular slot combinations. Note that the voltages induced in the coils concentric with a slot are larger than those concentric with a stator tooth. In computing winding weighting factors care should be taken in solving Eqs. 5-7 since a solution does not exist in some cases, indicating that fewer than four coils are required to obtain a solution. One example of such behavior is the 4 pole machine with 36 stator and 44 rotor slots given in Table 1. It can be noted that D/A = 0 so that the D coil is essentially not required.

| Concentric Coils Centered Over A Tooth | | | | | | |
|---|---|---|---|---|---|---|
| P | S | R | B/A | C/A | D/A | K/A |
| 4 | 36 | 44 | -0.717 | 0.434 | 0.0 | 0.140 |
| 4 | 72 | 58 | -0.704 | 0.328 | -0.080 | 0.003 |
| 6 | 72 | 86 | -0.721 | 0.345 | -0.092 | -0.009 |
| 8 | 72 | 86 | -0.742 | 0.371 | -0.113 | -0.019 |
| Concentric Coils Centered Over A Slot | | | | | | |
| 4 | 36 | 44 | -1.000 | 1.532 | 0.0 | 1.026 |
| 4 | 72 | 48 | -1.272 | 0.837 | -0.272 | -0.018 |
| 6 | 72 | 86 | -1.348 | 0.906 | -0.331 | -0.061 |
| 8 | 72 | 86 | -1.438 | 1.000 | -0.438 | -0.147 |

Table 1

Optimum Concentric Coils for Typical Slot Combinations

Figure 2 shows a complete flux coil layout for the machine with 36 stator and 44 rotor teeth. It is assumed that the armature winding pitch is arranged such that the flux axes of the three stator phase windings are aligned along stator teeth. Note that two coil sets are used to obtain a two phase measure of air gap flux. Because the number of slots per pole is 9, one set of flux coils is located over a tooth (q-axis) and the other over a slot (d-axis). Since the voltage induced in the coils centered around the tooth are smaller, the gain of these coils have been increased by the ratio 1.026/0.140 = 7.33. Each coil is assumed to have an equal number of turns. The required fractional number of turns for optimum cancellation of harmonics is obtained by proper summing resistors on the flux integrators.

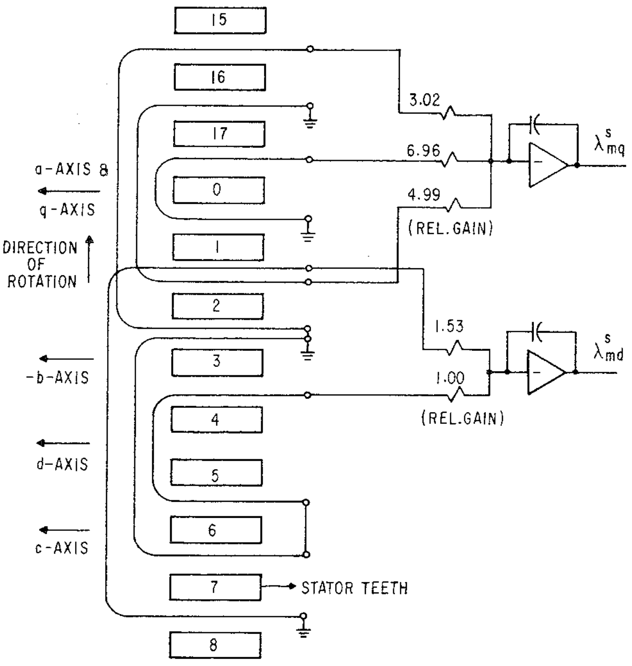

Fig. 2 Search Coil Arrangement Using Three Concentric Coils per Axis.

Examination of Table 1 indicates that the useful voltage K attained by the coil connections are much smaller than in individual coil voltages A, B, C and D. In effect, a large number of turns per coil or high integrator gain must be used to develop the required flux signal thereby increasing the sensitivity to small errors in the measuring apparatus. Since the coils must not link leakage flux components, the end turns of each coil must run closely along the ends of the stator stack. As the span of a flux coil increases, problems associated with mechanical integrity arise since abrasion by contact with the rotor or physical abuse become distinct possibilities. These problems can be reduced to a minimum by the coil arrangement shown in Fig. 3 in which each coil spans only one stator tooth. Also, note that since each coil is identical, the arrangement is easier to manufacture. In fact, since all coils are identical they can be installed without regard to location of magnetic axes and the proper axes located afterwards.

When the magnetic axis is centered over a stator tooth as in Fig. 3 the voltages induced in the coils are, for the fundamental

$$(D-C)\ \sin\ (7\pi P/S) + (C-B)\ \sin\ (5\pi P/S) + (B-A)\ \sin\ (3\pi P/S) + A\ \sin\ (\pi P/S) = K \tag{9}$$

for the R-P field

$$(D-C)\ \sin\ [7\pi(R-P)/S] + (C-B)\ \sin\ [5\pi(R-P)/S] + (B-A)\ \sin\ [3\pi(R-P)/S] + A\ \sin\ [\pi(R-P)/S] = 0 \tag{10}$$

for the R+P field

$$(D-C)\ \sin\ [7\pi(R+P)/S] + (C-B)\ \sin\ [5\pi(R+P)/S] + (B-A)\ \sin\ [3\pi(R+P)/S] + A\ \sin\ [\pi(R+P)/S] = 0 \tag{11}$$

and for the 3P field

$$(D-C)\ \sin\ (21\pi P/S) + (C-B)\ \sin\ (15\pi P/S) + (B-A)\ \sin\ (9\pi P/S) + A\ \sin\ (3\pi P/S) = 0 \tag{12}$$

Equations 10-12 can again be solved for D/A, C/A and B/A. When substituted into Eq. 9 the resulting fundamental component of voltage K/A is obtained.

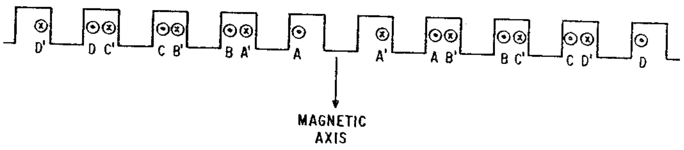

Fig. 3 Eight Tooth Pitch Coils Centered Over a Stator Tooth.

When the coils are centered around a slot the solution for the fundamental component is

$$(D-C)\ \sin\ (8\pi P/S) + (C-B)\ \sin\ (6\pi P/S) + (B-A)\ \sin\ (4\pi P/S) + A\ \sin\ (2\pi P/S) = K \tag{13}$$

The equations for the R-P, R+P, and 3P fields are similarly phase shifted. Table 2 shows the optimum coil configurations for the same slot combinations as Table 1. It is apparent that the fundamental component of voltage K is considerably larger portion of the voltage induced per coil compared to the concentric coil arrangement. Figure 4 shows the complete winding diagram for the four pole machine of Fig. 2. In this case the gains of the d-axis (slot-centered axis) have been increased to equalize the amplitude of the flux signals in the two axes. The required gain this time is only 0.425/0.232 = 1.83. Note that the three center coils serve a dual purpose since they effectively measure both d- and q-axis components. Hence, only nine coils are required for the complete flux measuring scheme.

| Tooth Pitch Coils Centered Over A Tooth | | | | | | |
|---|---|---|---|---|---|---|
| P | S | R | B/A | C/A | D/A | K/A |
| 4 | 36 | 44 | 0.064 | 0.468 | 0.0 | -0.425 |
| 4 | 72 | 58 | -5.158 | 4.792 | -3.426 | 0.407 |
| 6 | 72 | 86 | 2.837 | -2.212 | 2.271 | -0.782 |
| 8 | 72 | 86 | 0.848 | 0.348 | 0.848 | -0.543 |
| **Tooth Pitch Coils Centered Over A Slot** | | | | | | |
| 4 | 36 | 44 | -0.653 | 0.653 | 0.0 | -0.232 |
| 4 | 72 | 58 | 1.311 | -1.311 | 1.960 | -0.430 |
| 6 | 72 | 86 | -0.263 | 0.210 | 0.314 | -0.251 |
| 8 | 72 | 86 | -0.452 | 0.452 | 0.096 | -0.254 |

Table 2

Optimum Tooth Pitch Coils for Typical Slot Combinations

## Practical Considerations and Implementation

The installation of flux coils as outlined in the previous section fortunately does not usually present a severe hardship. Of course, relatively thin copper wire should be used, yet with sufficient strength to avoid breakage. Wire size approximately #26 AWG (15.9 mil Dia.) is a typical choice. In most cases, induction machines below several hundred horsepower contain random wound stator coils topped with a teflon strip (slot

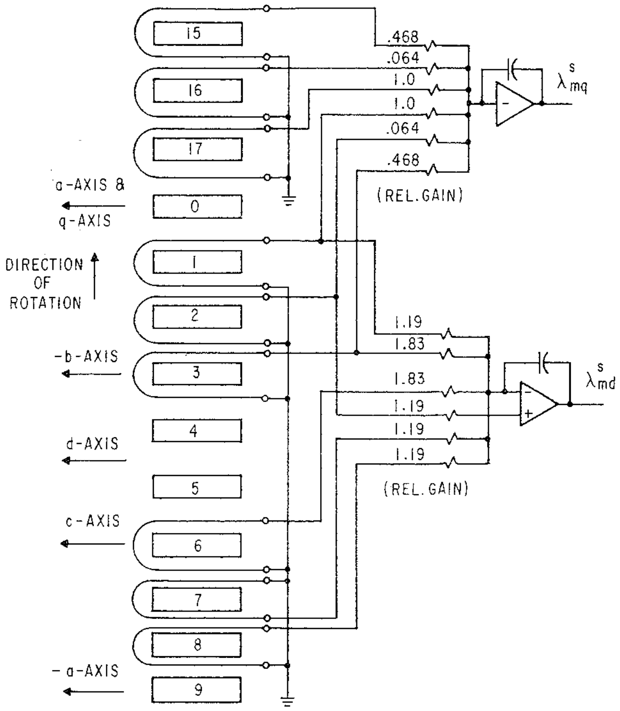

Fig. 4 Search Coil Arrangement Using Six Tooth Pitch Coils per Axis.

wedge). The core and armature assembly is baked with a polyester varnish to improve insulation qualities. This varnish can often be softened with a heat lamp or similar device thereby permitting removal of the teflon strip. The coils can then easily be inserted and the slots refilled with insulation material. Larger random wound and most form wound machines (500 HP and up) present a somewhat more difficult problem. In this case the slot wedge exists in the form of a mica-mat strip (top stick) which cannot be easily removed. The machine often is vacuum impregnated with epoxy insulation which softens at a much higher temperature than polyester varnish. Fortunately, as the rating of the machine increases the region above the top stick becomes larger so that the flux coils can generally be simply inserted above the top stick. If necessary, a small groove can be filed in the top stick without appreciably affecting mechanical strength.

For simplicity all coils should be wound with the same number of turns. Proper weighting factors are most readily obtained by adjusting the gain at the input of the integrator. In all but the smallest machines only several turns are sufficient to induce a sufficiently large voltage in the coils. Approximately ten turns is desirable for a four pole 25 HP motor. Since rating is proportional to air gap surface area, only one turn is needed above 250 HP. When flux signals are used in a feedback control system, the location of the two phase d- and q-axes must be carefully located. The magnetic axis of the d-axis can be located by energizing the phase a winding with a small direct current. The rotor is removed during the test. The point of flux reversal is then identified by moving a compass along the stator surface. By counting forward in the direction of rotation by S/6P tooth pitches the orthogonal q-axis can be located.

Because of the thermally non-stable environment, chopper stabilized operational amplifiers must be used in order to maintain accuracy in the flux measurement over a wide frequency range. A number of high quality, low noise, low drift amplifiers are commercially available for this task. Figure 5 shows wave forms obtained from the optimum search coil arrangement for the four pole machine containing 36 and 44 rotor teeth (Fig. 4). Figure 5(a) shows the voltage induced one individual search coil. Note the presence of the third harmonic due to saturation as well as the slot ripple harmonics. Figure 5(b) is a waveform of the weighted summation of voltages before and after integration. Before integration only a small residual amount of slot harmonics remain. After integration a sine wave of very low harmonic content is clearly evident.

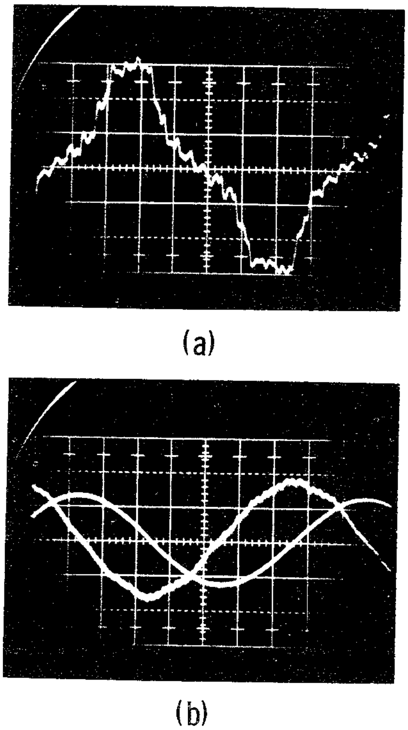

Fig. 5 Experimental Results Using Tooth Pitch Arrangement of Fig. 4. (a) Voltage Produced by One Search Coil. (b) Weighted Summation of Voltages Before and After Integration.

## Application to Static AC Drives

Although flux sensing has been used for decades as a measurement technique, the application to closed-loop control of static AC drives is much more recent. A nagging problem with these new drives is their poor open-loop damping. In fact, it has been established that in some cases dynamic (small signal)[2] or even static (steady-state) instability[3] can occur. Such systems are routinely compensated with feedback in order to improve damping. In particular, the rotor slip frequency and stator current amplitude are typically used as set point variables in conjunction with either voltage source[4] or current source types of inverters.[5]

Application of the current-slip control principle to a current source inverter drive is shown in Fig. 6. In the implementation shown, DC link current $i_d$ and angular slip frequency $\omega_{s\ell}$ are adjusted as functions of the torque command signal $T_e^*$. It is clear that the motor parameters must be accurately known in order to develop the desired motor torque with the proper motor flux condition. Unfortunately, motor parameters change widely with temperature, frequency and current amplitude so that these set point variables do not remain optimum over the entire range of operating conditions. Some deterioration in performance is particularly noticeable at low speed, during reversing operations, or during braking.

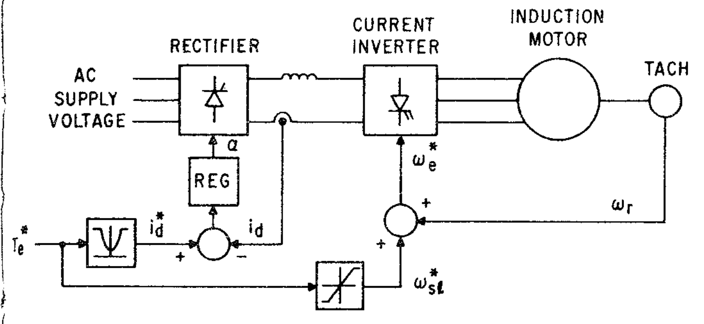

Fig. 6 Torque Controller for Current Source Inverter AC Drive Employing Current Amplitude and Slip Frequency Regulation.

Figure 7 shows a computer trace of the start-up transient for a typical 25 HP system. It is apparent that although the motor current and slip frequency are established very rapidly, the desired motor torque is not produced until the proper motor flux is developed in the air gap. It has been verified that speed of response cannot be improved by simply adjusting control gains and an entirely different control algorithm is required.

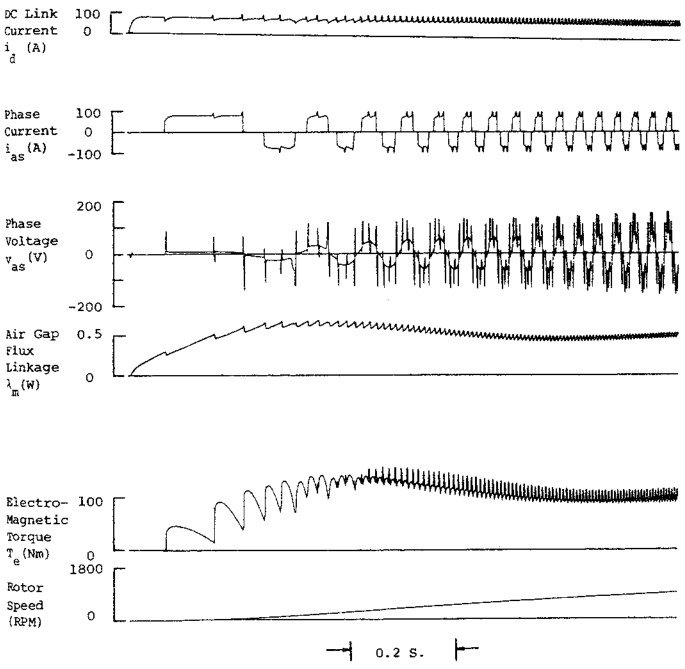

Fig. 7 Acceleration of Drive from a Stalled Condition with Inertia Load Using Control System of Fig. 6.

In order to develop useful power output two mechanisms must be present in any electrical machine. First, the machine must be magnetized as does any magnetic device. Second, a current component must be provided which sets up an MMF normal to the magnetizing flux. The special interaction of these two field components results in electromagnetic torque. In the case of a DC machine, these two components flow independently since the field winding (magnetizing current) and the armature winding (torque producing current) are electrically isolated. In the case of an induction machine (and to a lesser extent a synchronous machine) the case is quite different. Both the magnetizing and torque producing components must enter the machine through the same terminals (stator winding). Rapid control of an induction machine implies rapid changes in the power component of current in order to produce torque. However, at the same time the magnetizing component must be maintained at a constant amplitude in order to avoid saturating or underexciting the machine. The problems involved in "sorting out" these two components with the converter control system, in essence, lead to the sluggish response illustrated in Fig. 7. In order to alleviate these problems it has been found useful to sense the actual instantaneous flux distribution in the machine by inserting flux coils in the air gap of the machine. The two components of the current can then be sorted out relative to the measured flux vector and the system dynamic response markedly improved.

A so-called "decoupling" of the magnetizing from the torque producing component of rotor current can be accomplished by the method of field orientation.[6] The technique can be viewed as an exercise in coordinate transformation, a concept popular among electrical machine theorists.[7] In this approach the flux linkage corresponding to the air gap flux is visualized as a rotating vector. An axes transformation is employed in which the stator current and flux linkage variables are transformed from a stationary reference frame to a frame which rotates with the flux linkage vector. The q-axis of the rotating axes is aligned so that it is identically equal to the magnetizing component of stator current. Since torque is equal to the vector cross product of stator current and air gap flux, the stator current corresponding to the orthogonal rotating d-axis is the torque producing component. Figure 8 shows a practical extension of this scheme to a current source inverter induction motor drive. It should be mentioned that other methods have recently been developed which utilize the same control variables but do not employ the field orientation principle.[8,9] Again only the inner torque regulator is shown. Note that the torque is now regulated by means of the inverter frequency. Flux linkages are controlled by means of the phase-controlled DC bridge. As an added bonus it is important to note that a tachometer is not required. Figure 9 shows a computer trace of the start-up condition for a prototype of the same 25 HP system as Fig. 7. Much improved performance is clearly evident. It can be noted that after a short transient during which time the reference frame seeks to "lock-on" to the flux linkage vector, the machine accelerates very smoothly.

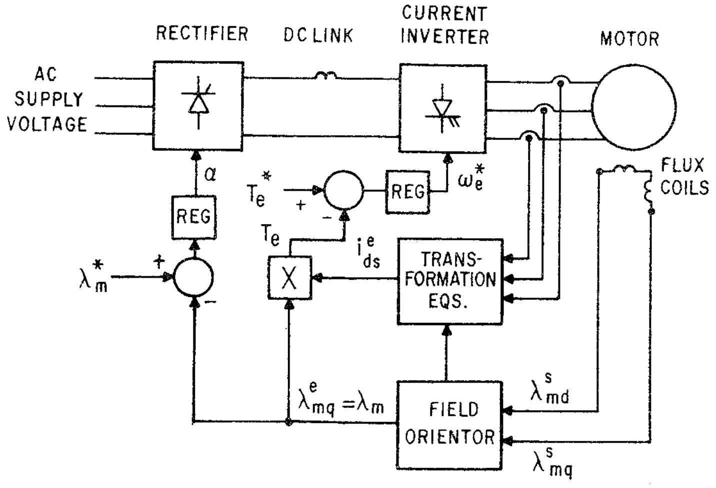

Fig. 8 Torque Controller for Current Source Inverter AC Drive Employing Field Orientation.

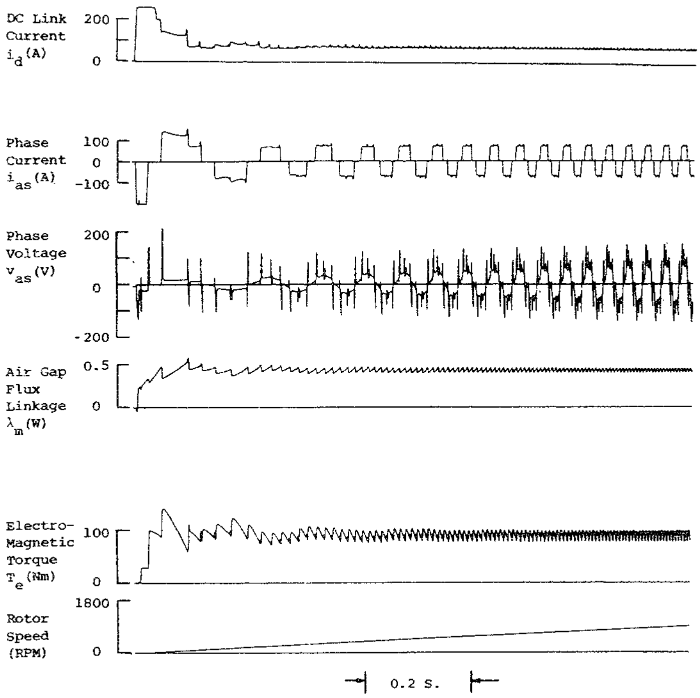

Fig. 9 Acceleration of Drive from a Stalled Condition with Inertia Load Using Control System of Fig. 8.

In Fig. 10 the step response of the two types of control systems are compared. Two time constants are evident in the case of the current-slip control scheme. The shorter time constant is approximately 20 mS indicating a moderately fast response. A second, much longer time constant is apparent associated with the magnetizing component of stator current. It is clear that the system response with field oriented control is much faster. An initial response on the order of 5 mS or less indicated. It appears that the basic response limitation to this type of control is the switching rate (commutating ability) of the inverter.

## Conclusion

Although flux coils are often used to monitor motor conditions the need for a high quality AC signal has appeared only recently. This paper has identified the problems involved in flux measurement and has outlined a design procedure to minimize their effects. Present-day applications have not required high dynamic response from an AC drive over a wide speed range. However, new applications are arising which exceed the response readily obtained from such a drive using established control techniques. To satisfy this need the effectiveness of flux signals in control of such static AC drives has been demonstrated.

## Acknowledgement

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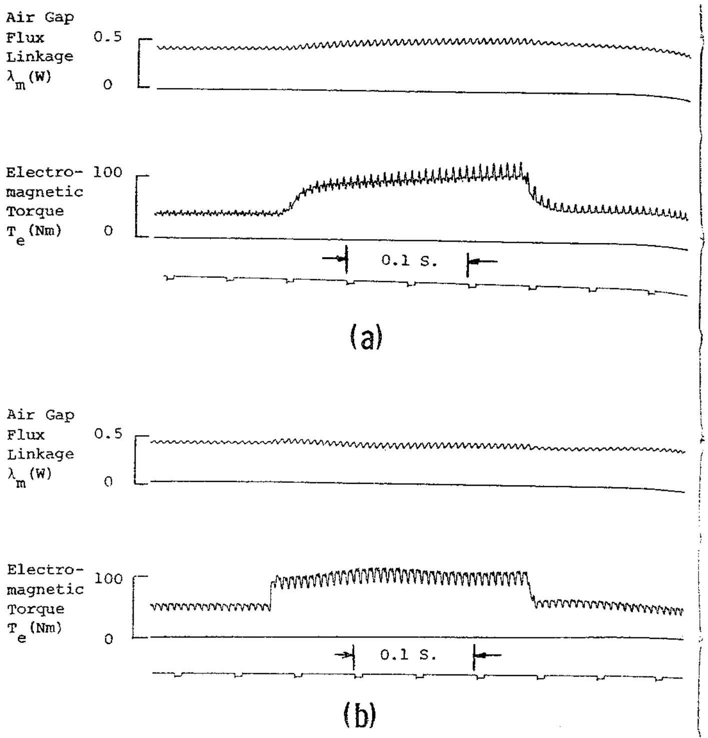

Fig. 10 Step Response of (a) Current Amplitude and Slip Frequency Control, (b) Field Oriented Control. Step Change in Torque Command from 0.5 to 1.0 per Unit at 30 Hz Line Frequency.

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