Elimination of Discrete Position Sensor and Current Sensor in Switched Reluctance Motor Drives

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Abstract—**Shaft position sensing is essential in switched reluctance motors (SRM) in order to synchronize the phase excitation pulses to the rotor position. In order to implement closed-loop torque control, current sensors are also unavoidable. These sensors can constitute a substantial portion of the total system cost and tend to reduce system reliability.**

A low-cost position-sensing scheme using stator inductance measurement is presented in this paper. An analog electronic technique can be applied to measure the inductance of a nonconducting phase by using a linear frequency modulated (FM) converter. The output of the FM converter is decoded to get the shaft position signal. The control functions and converter switching signals are processed in a low-cost microcontroller.

The discrete current sensors for closed-loop operation have been eliminated by using MOS-gated power switches with integrated current sensing capability. The combination of the above two schemes resulted in a totally sensorless SR motor drive as described in the paper.

I. Introduction

The SWITCHED reluctance motor (SRM) drive technology has gone through steady and significant development over the last two decades [1]. The simplicity in both motor construction and power converter requirement, together with the availability of fast-acting high-performance semiconductor switches, have made SRM an attractive alternative in adjustable speed drives. The major emphasis of published research thus far has been on motor design and construction and on converter topology [2] and not much attention is given to the research on the control aspects of the motor. Successful and efficient four-quadrant SRM drives have been reported previously [3], [4] but with all these systems came the requirement of shaft position sensor and discrete current sensors. A few techniques of driving the SRM without the shaft position sensor have been reported previously, but these either added additional complexity to the drive circuitry [5], [6] or required significant real-time computation [7]. Adding complexity to the simple SRM converter configuration is undesirable. In addition, noise and A/D converter delays pose serious limitations around the commutating region in real-time computation of the inductance. This paper describes a simple way of measuring the inductance of the nonconducting phase directly, thereby giving the rotor position information without a shaft transducer. Furthermore, the need for discrete current sensing has been eliminated by using MOS-gated power switches (i.e., MOSFET's and IGBT's) with integrated current sensing leads.

In an SRM, torque is developed by the tendency of the rotor to seek the minimum reluctance position, which consequently is the complete alignment of the rotor. The phase inductance is maximum in this position. The rotor position information is, therefore, essential in an SRM drive to generate the commutation signals and typically mechanical devices like optointerrupters with a slotted disc have been used to determine the shaft position, whereas isolated current sensors like current transformers or Hall effect sensors have been used for current sensing. These sensors not only add complexity to the system but also come with associated costs.

The method of determining the rotor position by direct stator inductance measurement avoids complex circuitry and at the same time eliminates the mechanical mounting of devices for shaft sensing. This technique would be specially applicable where size and weight is a major constraint (e.g., in aerospace applications) and also where cost is a significant concern, such as in consumer products. The complete integration of our simple electronics can give an impetus to replacing conventional adjustable ac and dc drives by SRM drives.

This paper, starting with a brief discussion on basic principles of SRM, describes the inductance measuring technique and the method of generating the commutating signals based on that information. The application of integrated current sensing to SRM drives has also been described. Experimental verification of the developed techniques is also presented.

II. Basic Principles

The details of the principle of operation of the SR motor have been discussed in [1] and [2] and only salient points are summarized here. A schematic diagram of the SR motor with six stator poles and four rotor poles is shown in Fig. 1. The stator windings on diametrically opposite poles are connected in series to form one phase. Note that there are no windings or magnets on the rotor. When a stator phase is energized, the corresponding rotor pole pair is attracted toward the

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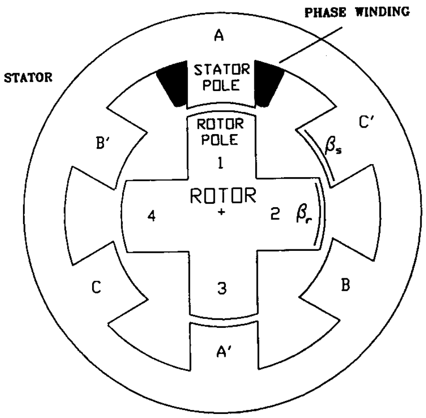


Fig. 1. Cross section of a 6/4 SR motor.

energized stator phase to minimize the reluctance of the magnetic path. Therefore, by energizing consecutive stator phases in succession it is possible to develop constant torque in either direction of rotation.

Several other combinations of stator and rotor poles are possible but a 4/2 or 2/2 configuration has the disadvantage that if the rotor and stator poles are aligned exactly it would be impossible to develop a starting torque. The configurations with higher numbers of stator/rotor poles have less torque ripple and do not have the problem of starting torque.

III. Torque Production

Torque is developed by the tendency of the magnetic circuit to adopt a configuration of minimum reluctance and is independent of the direction of current flow. Consequently, unidirectional currents are required and a simple configuration is sufficient as the drive circuitry. The torque in terms of coenergy W is [1],

$$T(\theta, i) = \frac{\partial W(\theta, i)}{\partial \theta} \tag{1}$$

where θ is the angle describing the rotor position and i is the current in the stator windings. Under the simplifying assumption of no magnetic nonlinearity the torque equation becomes

$$T(\theta, i) = \frac{1}{2} i^2 \frac{dL}{d\theta} \tag{2}$$

where L is the self-inductance of the stator phase at any value of θ.

The idealized inductance profile of an SR motor is shown in Fig. 2(a). For positive or motoring torque, phase current is switched such that it flows during the rising inductance period and, for braking phase current, must coincide with the decreasing inductance period as shown in Fig. 2(b). It should be noted that to maximize motoring torque the current in a phase should be switched on during the constant inductance region so that the current can build up before the period of increasing inductance starts. In addition, the current should be switched off before the end of the increasing inductance period to allow the current to decay fully so that no negative torque is produced.

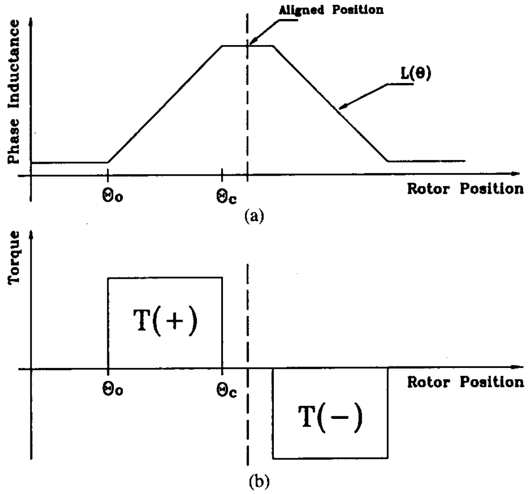


Fig. 2. (a) Idealized inductance profile for one motor phase; (b) idealized motoring torque $T(+)$ and braking torque $T(-)$ for constant phase current.

IV. Power Circuit

Each stator phase of an SR motor must be energized by an unidirectional current pulse while the rotor is appropriately positioned relative to the stator. A rapid response is necessary for positive and negative changes in the demanded current level. This function requires a two-quadrant power converter that is capable of applying equal positive and negative phase voltages to produce equal rates of current increase and decrease. There are several circuits that provide these requirements, and the conventional two-switch two-diode per phase configuration has been chosen as the converter for the sensorless drive. This configuration has the advantage over the others in that it does not need a bifilar winding or split power supply or even number of motor phases. The converter configuration for the totally sensorless SR motor drive is shown in Fig. 3. Current-sensing MOSFET's that have an additional lead for providing information about current flowing through the switch can be used for power devices. A high-voltage-integrated circuit (HVIC) driver chip IR-2110 manufactured by International Rectifier is used to drive the power switches. The HVIC eliminates the need for separate floating power supply for each of the upper switches. A bootstrap technique is used to provide a floating bias supply in IR-2110 in combination with level shifting a ground-referenced input signal [8]. IR-2110 is configured directly to drive a pair of power MOSFET's or MOSIGT's connected in half-bridge or other configurations.

V. Position Information from Inductance Variation

Several indirect rotor position sensing schemes have been proposed in the past [5]–[7] and almost all of them use the inductance variation information to detect the rotor position. This is because the phase inductance of the SR motor varies significantly between the aligned and unaligned rotor positions. The ratio of maximum to minimum inductance in SR motors is usually 3 or greater. In fact some SR motors have a maximum to minimum inductance ratio of 10 or more.

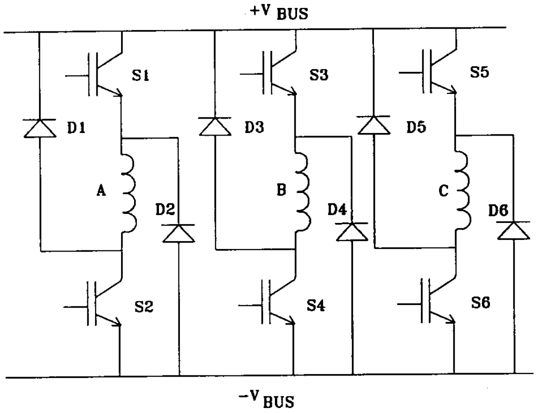


Fig. 3. Power converter circuit for a three-phase SR motor.

Acarnley et al. [9] first suggested monitoring the current waveforms to detect the rotor position in stepping and switched reluctance motors and applied the technique successfully to variable reluctance step motors. The fundamental idea behind this technique is that the rate of change of current in a phase depends on the incremental inductance, which in turn is rotor-position dependent. The appearance of a "back-EMF" term in the di/dt expression relating incremental inductance of an energized phase complicates the determination of rotor position using this method. Extensions to this technique are the calculation of real-time inductance or measuring the inductance of an unenergized phase with simplified calculations.

The general equation governing the flow of stator current in an SR motor can be written as

$$V = Ri + L\frac{di}{dt} + i\frac{dL}{dt} \tag{3}$$

where V is the dc bus voltage and L is the instantaneous inductance of a coil.

Solving for inductance L

$$L = \frac{(-B + AL_0)e^{-At} + B}{A} \tag{4}$$

where $A = (di/dt)/i$, $B = (V/i) - R$ and L_0 is the value of L during the previous sample.

Solving (4) would determine the rotor position indirectly since the inductance of an SR motor varies significantly with rotor position. However, the computation time for obtaining real-time inductance can be too long for high-speed or dynamic-response applications. Also, the algorithm would be quite vulnerable due to noise, A/D converter delays, etc. around the commutating region. Therefore, the real-time calculation of inductance of the SR motor for position estimation is not a good solution.

An alternative approach is to pulse an unenergized phase and calculate the inductance with a simplified algorithm. If phase voltage is applied for a short period of time of Δt, the current remains small, the rotor position does not change significantly and the inductance does not saturate. The voltage of (3) can then be approximated as

$$V = L(\theta)\frac{\Delta i}{\Delta t}.$$

Therefore

$$L(\theta) = V\frac{\Delta t}{\Delta i}$$

and

$$\theta = F^{-1}(L).$$

The pulsing can be done either by the main converter itself [6] or from an external electronic oscillator presented in this paper. The new concept developed at the Power Electronics Laboratory of Texas A&M University uses a frequency modulated (FM) converter that generates a voltage whose magnitude is proportional to the instantaneous inductance. This analog voltage signal contains the absolute shaft position information and can be processed to obtain the commutation instants.

VI. FREQUENCY MODULATED CONVERTER

The basic concept of the FM encoder technique is to generate a train of square wave signals whose frequency is proportional to the instantaneous inductance of a phase. Any analog circuitry that converts an inductance into frequency can be used for the purpose. The circuit proposed by Senani [10] for inductance to time-period conversion using an unity gain differential amplifier (UGDA) is a simple method that can be used to convert the phase inductance of an SRM to a proportional frequency signal. An example circuit schematic is given in Fig. 4 and the time period of this oscillator is given by

$$T = \frac{L}{(R + R_{\text{Loss}})}\left\{\ln\frac{1 + \alpha\dfrac{V_{01}}{V_{02}}}{1 - \alpha} + \ln\frac{1 + \alpha\dfrac{V_{02}}{V_{01}}}{1 - \alpha}\right\} \tag{5}$$

where $\alpha = R_2/(R_1 + R_2)$, R_{Loss} is the stray resistance of the phase coil and V_{01} and V_{02} are the positive and negative saturation levels, respectively.

Choosing the appropriate circuit parameters and assuming constant saturation voltages the time period, T of the output signal can be written as

$$T = K_1 L \Rightarrow f = \frac{1}{K_1 L}$$

where K_1 is a proportionality constant depending on circuit parameters.

The frequency-encoded signal of the phase inductance can then be decoded by directly supplying this signal to a microcontroller and counting the frequency using its time or using an F-V (frequency to voltage) converter to obtain a voltage proportional to the frequency. For our experiments a Teledyne F-V converter TSC9400CJ, which has a linear relationship between input and output, was used to generate the voltage waveform given by

$$V = K_2 f$$

where K_2 is also a proportionality constant.

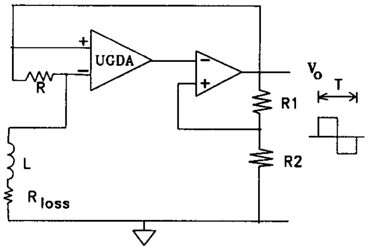


Fig. 4. Linear L/T converter.

Therefore

$$L = \frac{1}{KV}$$

where $K = K_1/K_2$.

Because phase inductance depends on the position of the shaft we, therefore, get the shaft position information in terms of the output voltage of the FM converter. An attractive feature of the FM converter is that it can operate from dc to $100k$ Hz. Therefore, the shaft position can be determined with high sensitivity.

VII. Phase Selection for Sensing Inductance

Two important problems have to be solved for implementing the inductance-sensing scheme. The first problem concerns the selection of a suitable phase that is not conducting to measure its inductance. The second, more critical, problem is how to connect a motor phase winding, which is in the power circuit, to the FM converter, which is in the control circuit.

The phase windings in the SR motor are independent and therefore the non-torque-producing stator phases at a given instant can be used for sensing inductance and detecting rotor position indirectly. A voltage profile that is inversely related to the phase inductance is obtained at the output of the FM converter. The output voltage profile of phase A obtained with a rotor speed of 750 r/min is shown in Fig. 5. Because only discrete rotor positions need be detected for successful commutation, sensing is accomplished by comparing the output of the FM converter to a threshold dc level. For example, while phase C is producing torque, phase B is sensed to detect when to switch phase A on. Similarly, when phase A is producing torque, phase C is sensed to detect when to switch phase B on and when phase B is conducting phase A is sensed to detect when to turn phase A on.

Phase excitation can be advanced or retarded by increasing or reducing the dc threshold level around the commutating region. The sensitivity of the FM converter determines the precision of advance angle control of the drive. The output voltage of the FM converter, which is inversely proportional to the phase inductance, varies from 1–4 V in our experiment. We are particularly interested in the output of the FM converter around the low-phase inductance region, which corresponds to voltage levels around 4 V, because of the high rotor position accuracy it offers. Harris and Lang [11] have shown that the effect of mutual inductance in the SR motor is negligible around the minimum inductance region. Because we are also interested in the performance of the FM encoder around that region, the mutual inductance will not pose a serious problem in the scheme.

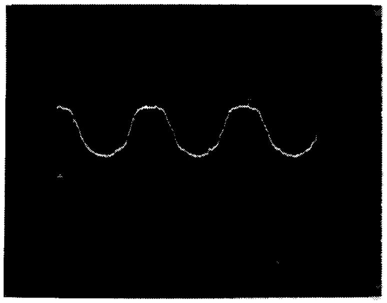


Fig. 5. $1/L$ profile of an SR motor, 5 ms/div, 1 V/div.

The problem of connecting the SR motor phase winding to the sensing circuitry is overcome by the use of photovoltaic BOSFET power switches that can connect analog signals from thermocouple level to 300 V peak ac or dc polarity. The BOSFET power IC switch [12] made by International Rectifier contains two power MOSFET's in inverse series connection for distortion-free control of bidirectional (ac) and dc signals. The BOSFET switch is controlled by a photovoltaic generator, which is energized by radiation from a dielectrically isolated light emitting diode (LED). This switch can provide the necessary voltage isolation as well as the switching speed required.

The block diagram of the sensor elimination scheme for the SR motor drive is shown in Fig. 6. Only one phase is shown for illustration and the other two phases have an identical control structure. The commutation instants are determined by a voltage comparator and the information is fed to the microcontroller.

A low-cost microcontroller is used to implement the new rotor position sensor-elimination scheme. The microcontroller is configured to have three inputs and six outputs. The three input signals correspond to the three comparator output signals. Among the six output signals, three provide driving signals to the power switches while the other three outputs are used to control the turn-on and turn-off of the BOSFET switches.

VIII. Starting Algorithm

At standstill, the microcontroller determines which phase should be excited to deliver torque in the desired direction by checking the inductance of all the three phases. Each phase is checked in sequence and the phase with the lest inductance is connected to the supply for positive torque production.

IX. Steady-State Test Results

Extensive tests were conducted to evaluate the performance of the new position sensor elimination scheme. Fig. 7 shows the gate pulses of phase A on the upper trace of the oscillograph and the current through phase A appears in the lower trace. A pulse-width modulation with unequal turn-on and turn-off times is employed to demonstrate the feasibility of using PWM with the new scheme. Fig. 8 shows the

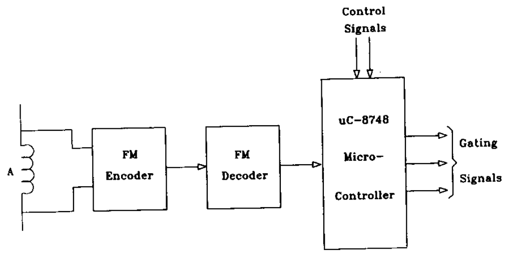


Fig. 6. Block diagram of the FM-encoded sensor-elimination scheme for the SR motor.

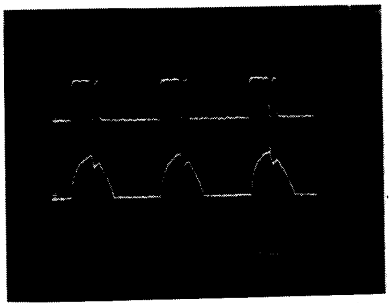


Fig. 7. Gate pulses and the current waveform of phase A at 1050 r/min. Upper trace: gating signal, 10 V/div, 5 ms/div. Lower trace: phase current, 1 A/div, 5 ms/div.

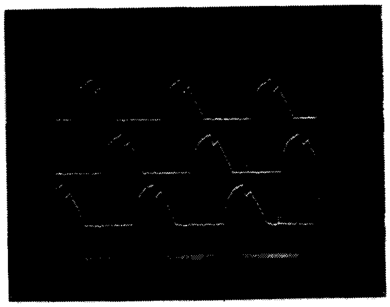


Fig. 8. Current waveforms of phases A, B, and C, 5 ms/div, 1 A/div at 1050 r/min.

oscillographs of the current waveforms of the three phases of the SR motor running at 850 r/min on no load. This result demonstrates that symmetrical and balanced three-phase currents are injected into the motor winding in synchronism with the rotor position with the new sensorless scheme.

The accuracy of the sensorless scheme was compared to the scheme that uses the traditional discrete position sensors by conducting two tests. In the first test the gating signals generated by the two schemes are compared. In Fig. 9 the top trace shows the gating signals of phase A obtained with the new sensorless scheme under normal positive torque producing mode of operation. The bottom trace shows the gating signals for the same phase obtained from the opto-interrupter and slotted-disk type of discrete position sensors. The correlation between the two schemes is excellent. The appearance of the notches in the top trace is due to the use of PWM on the actual converter, whereas the bottom trace shows the entire phase A conduction period from the discrete position sensor-gating logic circuit prior to the PWM stage. The important comparison here is the accuracy of phase conduction timing.

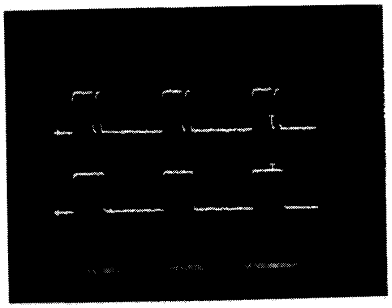


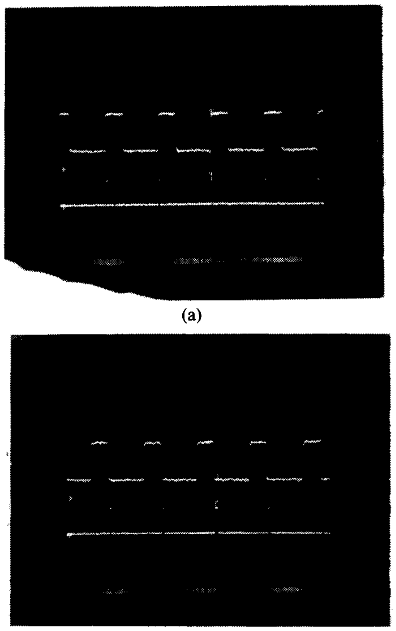
Fig. 9. Gating signals of phase A of the SR motor running at 1050 r/min. Upper trace: sensorless scheme, 10 V/div, 5 ms/div. Lower trace: with sensor, 10 V/div, 5 ms/div.

The second test directly determines the position-sensing accuracy by comparing the discrete position sensor output with the output of the threshold level comparators of the sensorless scheme. Fig. 11(a) shows the phase A position sensor output and the phase B comparator output while Fig. 11(b) shows the phase A position sensor output and phase C comparator output. In the new scheme the rising edge of the comparator initiates turn-on of one phase and turn-off of the other. Thus, sensor A signal should lie in between the outputs of the comparators of phases B and C, which is actually the case as shown in Fig. 10. The accuracy of the discrete position sensor scheme depends on the accuracy of the mounting of the opto-interrupters as well as that of the slotted disk. However, the accuracy of the new sensorless scheme depends on the accuracy of tuning the circuit.

X. COMPARISON WITH PULSED IMPEDANCE-SENSING TECHNIQUE

The foundation of both the FM encoder technique and pulsed impedance-sensing technique [6] is instantaneous phase inductance detection. However, in the FM technique an external circuit is used to measure the inductance, whereas in the latter the phase is pulsed from the main converter. One serious drawback of the pulsed impedance-sensing technique is that because the phases are pulsed from the main inverter dc bus the current takes a longer time to decay to negligible levels, and the next sensing pulse cannot be applied until the previous one has decayed completely. Therefore, the sensing pulses are at relatively low frequency. This limits the attainable shaft position sampling frequency and resolution. This waiting period is virtually nonexistent with the FM encoding technique as a bipolar low voltage analog circuit is used to pulse the phases two twice per cycle with much higher frequency.

In the impedance-sensing scheme, the inductance profile flattens in and around the aligned position due to saturation since the phases are pulsed to higher current levels from the high-voltage inverter bus. In the FM encoder technique the phase is pulsed by a signal level external circuitry and because bipolar voltage is applied it is not necessary for the current to decay completely. Hence, saturation problems are absent in this technique. The qualitative difference in $1/L$



(a)

(b)

Fig. 10. Opto-interrupter and comparator outputs: (a) Upper trace: phase A sensor output, 5 V/div, 5 ms/div. Lower trace: phase B comparator output, 2 V/div, 5 ms/div; (b) upper trace: phase A sensor output, 5 V/div, 5 ms/div. Lower trace: phase C comparator output, 2 V/div, 5 ms/div.

curves obtained using the impedance-sensing technique and FM encoding technique is shown in Fig. 11. The di/dt curve has an almost flat profile while, whereas the FM-encoded curve follows the inductance variation of the upper trace in an around the aligned position. In the method of comparing the inductance profile with a threshold level, proper commutation will take place as long as the $1/L$ curve follows the inductance profile. Therefore, saturation will limit the attainable phase advance angle in the impedance-sensing technique while more phase advance is possible with the FM encoder technique.

The main problem with the FM encoding technique is the need for isolation of the electronic circuit from the phase winding when it is connected to the main dc bus. However, this can be used to advantage in multiplexing the consecutive phases with a single FM encoder electronic circuit. In other schemes the number of sensing circuitry required is usually equal to the number of motor phases.

The tail current in the deenergized phase may limit the window of time available for impedance sensing on that phase when the RPM is very high. This problem is compounded in the three-phase SRM because the most suitable phase to probe is the last energized phase. For this reason this sensing scheme will work better with four-phase SRM's because in a four-phase motor the most suitable phase for inductance sensing is not adjacent to the torque-producing one. However, with a three-phase motor, low-speed applications will pose no problem because the window available for phase pulsing is sufficient after the residual current decay in the nonconducting phase.

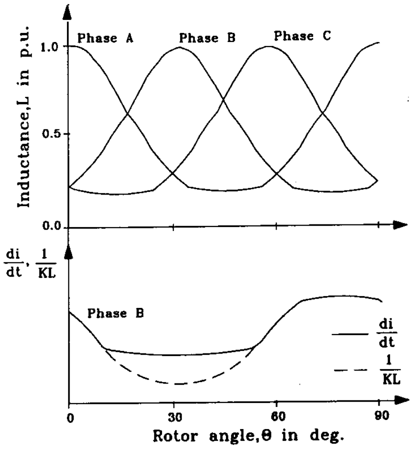


Fig. 11. Inductance profile and $1/L$ curves obtained using the impedance-sensing technique (di/dt) and the FM-encoder technique ($1/KL$).

The small number of components required and the simplicity of position measurement using this FM encoder technique suggests that a sensorless SRM drive using this technique would cost less than other possible SRM drives. The elimination of discrete position sensors reduces the costs of mechanical mountings. The cost can be further reduced by integrating the different components of the FM encoder into one single chip.

The response time of the FM converter dictates how efficiently speed transients can be handled. The shaft position can be monitored continuously by keeping the FM encoder always connected to a nonconducting phase and with a fast enough FM oscillator sudden changes in speed can be taken care of quite efficiently. However, the influence of high phase currents on the performance of the FM technique is an area of ongoing study.

XI. INTEGRATED CURRENT SENSING

Torque in an SR motor is controlled by closing the current loop. The back emf of the winding is small at low speeds and phase current is usually limited by "chopping" or "constant frequency PWM." The chopping method is applied to maintain the current constant and produce constant torque. At high speed motor back emf is itself large enough to restrict the rise of current. However, if the current increases excessively even at high speed the chopping mode of control can be restored.

The most popular method of measuring current in an SR motor is a simple resistive shunt located in series with one of the phase windings. The resistive shunts are inexpensive, but the power losses in these resistors become excessive especially when the ratings of the SR motor increase to the integral horsepower range. Isolated current sensors such as current transformers or Hall effect sensors can also be used for each motor phase for complete current feedback, but the associated costs often become prohibitive. An attractive alternative to discrete current sensors is the MOS-gated power switches that have integrated current-sensing leads [13]. In these new generation of MOS-gated power switches a few of

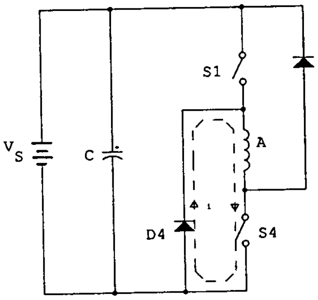


Fig. 12. Freewheeling path for phase current while chopping.

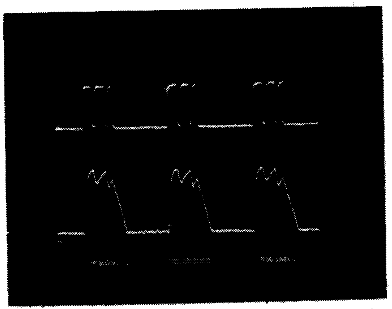


Fig. 13. Torque control with integrated current sensing and hysteresis band controller of one phase at 980 r/min. Upper trace: gating signal, 5 V/div, 5 ms/div. Lower trace: phase current, 2 A/div, 5 ms/div.

the identical device cells are segregated and a small, but fixed, fraction of the main device current is diverted to a separate sensing terminal. This small current signal can then be conveniently amplified and conditioned to provide useful information about the current flowing through the device. Power MOSFET's and IGBT's are now available from several device manufacturers with ratings up to several hundreds of volts and several tens of amps with current-sensing capability.

The integrated currents sensors impose their own constraints. The problem of level shifting for the upper switches of the converter becomes evident since the current-sensing terminal is not electrically isolated from the associated power device source (emitter) terminal in any available n-channel MOSFET's (IGBT's) provided with such sensing leads [14]. The level shifting for the upper switches is necessary in order to have all the feedback information with respect to the same reference. In addition, the information obtained from these sensors are not complete because sensing lead carries current only when the switch is conducting. These associated problems of integrated current sensors are avoided by using sensor switches only in the lower legs of the converter, whereas regular three-pin power MOSFET's are used for the upper switches.

In the conduction period of one phase, the phase current is chopped only by closing and opening the upper switch while the lower switch is kept closed. During the off period the current will freewheel through the path shown in Fig. 12 and continuous information of current through the phase winding is available. A hysteresis-type controller can be used to restrict the current within the desired band. The hysteresis controller generates on and off gating signals only for the upper switch, whereas the total conduction period for the phase is determined by the inductance (position)-sensing loop. The lower switch is kept on during the entire period of phase conduction. Fig. 13 shows the oscillograph of the gating signal and the resulting current flowing through phase A of the integrated sensorless scheme.

XII. CONCLUSION

An indirect position sensing scheme to eliminate discrete position sensors in an SRM drive has been presented in this paper. The scheme, taking advantage of the high degree of inductance variation in an SR motor phase winding, senses the phase inductance that directly gives the shaft position information. A frequency modulated (FM) converter composed of analog electronic components is used to measure the inductance of a nonconducting phase. The converter translates the position-sensing problem into a dc-level voltage-sensing problem. This information is fed to a low-cost microcontroller that processes the signals and generates the gating pulses.

The need for discrete current sensing for closed-loop operation has also been eliminated in this project by using MOS-gated power switches with integrated current sensing leads. Experimental results demonstrate the practicality of the circuit.

REFERENCES

[1] P. J. Lawrenson, J. M. Stephenson, P. T. Blenkinsop, J. Corda, and N. N. Fulton, "Variable speed reluctance motors," *Proc. Inst. Elec. Eng.*, vol. 127, pt. B, no. 4, pp. 253-265, July 1980.

[2] M. Ehsani, J. T. Bass, T. J. E. Miller, and R. L. Steigerwald, "Development of a unipolar converter for variable reluctance motor drives," *IEEE Trans. Industry Applications*, vol. IA-23, no. 3, pp. 545-553, May/June 1987.

[3] T. J. E. Miller, P. G. Bower, R. C. Becerra, and M. Ehsani, "Four-quadrant brushless reluctance motor drive," in *Proc. IEEE Conf. Power Electron.*, July 1988, pp. 273-276.

[4] B. K. Bose, T. J. E. Miller, P. M. Szczesny, and W. H. Bicknell, "Microcomputer control of switched reluctance motor," *IEEE Trans. Industry Applications*, vol. IA-22, no. 4, pp. 708-715, July/Aug. 1986.

[5] J. T. Bass, M. Ehsani, and T. J. E. Miller, "Robust torque control of switched reluctance motors without a shaft position sensor," *IEEE Trans. Ind. Electron.*, vol. IE-33, no. 3, pp. 212-216, Aug. 1986.

[6] S. R. MacMinn, P. M. Szczesny, W. J. Rzesos, and T. M. Jahns, "Application of sensor integration techniques to switched reluctance motor drives," in *IEEE-IAS Conf. Rec.*, Oct. 1988, pp. 584-588.

[7] A. Lumsdaine, J. H. Lang, and M. J. Balas, "A state observer for variable reluctance motors: Analysis and experiments," in *Proc. Incremental Motion Contr. Syst. Symp.*, June 1986, pp. 267-273.

[8] S. Young, "High-speed high-voltage IC driver for HEXFET or IGBT bridge circuits," *Int. Rectifier Application Notes*, vol. AN-978, 1988.

[9] P. P. Acarnley, R. J. Hill, and C. W. Hooper, "Detection of rotor position in stepping and switched reluctance motors by monitoring of current waveforms," *IEEE Trans. Ind. Electron.*, vol. IE-32, no. 3, pp. 215-222, Aug. 1985.

[10] R. Senani, "On linear inductance-time and related conversions using IC op amps," *IEEE Trans. Ind. Electron.*, vol. IE-34, no. 2, pp. 292-293, May 1987.

[11] W. D. Harris and J. H. Lang, "A simple motion estimator for variable-reluctance motors," in *IEEE-IAS Conf. Rec.*, Oct. 1988, pp. 281-286.

[12] *Microelectronic Relay Designers Manual*, International Rectifier, 1989.

[13] W. Schultz, "Lossless current sensing with SENSEFETs enhances motor drive design," *Power Conv. Intell. Motion*, pp. 30-34, Apr. 1986.

[14] T. Jahns, R. C. Becerra, and M. Ehsani, "Integrated current regulation for a brushless ECM drive," in *Proc. Appl. Power Electron. Conf.*, March 1989, pp. 81-90.

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