

Electronic mode of control to obtain increased torque and improved power factor from an asynchronous machine

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that of Fig. 2, but, for this case,

$$\left. \begin{aligned} C_0 &= wt \epsilon_{33}^T (1 - k_{33}^2) / l \\ v &= v_b^D = (1 / \rho s_{33}^D)^{\frac{1}{2}} \\ Z_0 &= \rho wt v_b^D \\ \phi &= (1 / 2M) \operatorname{cosec} (l / \omega 2v_b^D) \\ X_1 &= Z_0 M^2 \sin (l / \omega v_b^D) \end{aligned} \right\} \dots \dots \dots (3)$$

where

$$M = g_{33} / (\omega Z_0 s_{33}^D)$$

and l , w and t are defined in Fig. 2.

Side-electroded length-expander bar: The new equivalent circuit for the side-electroded length-expander bar is shown in Fig. 3. Here

$$\left. \begin{aligned} C_0 &= lw \epsilon_{33}^T (1 - k_{31}^2) / t \\ v &= v_b^E = (1 / \rho s_{11}^E)^{\frac{1}{2}} \\ Z_0 &= \rho wt v_b^E \\ B_1 &= (1 / Z_0) M^2 \sin (l / \omega v_b^E) \\ J &= |B_2| \end{aligned} \right\} \dots \dots \dots (4)$$

where

$$B_2 = -(2 / Z_0) M \sin (l / \omega 2v_b^E)$$

$$M = wd_{31} / S_{11}^E$$

and l , w and t are defined in Fig. 3.

The circuit of Fig. 3 contains an ideal 'admittance inverter', with an inverter parameter $J = |B_2|$. This is a 2-port network with the property that, when an admittance Y is connected to one port, the input admittance at the other port is J^2 / Y . The inverter in Fig. 3 is defined by its general-circuit-parameter matrix:

$$\begin{pmatrix} A & B \\ C & D \end{pmatrix} = \begin{pmatrix} 0 & j/B_2 \\ jB_2 & 0 \end{pmatrix} \dots \dots \dots (5)$$

(Note that this inverter behaves like a section of transmission line of characteristic admittance J with a phase shift of $+90^\circ$ if B_2 is negative and -90° if B_2 is positive.)

By checking the open-circuit impedance parameters for the new thickness-expander plate circuit (Fig. 2) against those for the circuit of Fig. 1, an exact equivalence may be demonstrated between these two circuits. Similar analyses show that the new circuits for the other two transducer types also correspond exactly to their counterparts in Reference 1, the validity of which is well established.

To demonstrate the simplification afforded by use of the new circuits, consider the effect of open-circuiting the electrical port in Fig. 1. The result is an impedance

$$(j\omega C_0)^{-1} - (j\omega C_0)^{-1} = 0$$

across the transformer primary. This short circuit, reflected through the transformer, connects together points P and P'. The result may be shown to be the lumped-element T equivalent circuit of an acoustic transmission line with no electrical loading. By contrast, this result is immediately apparent from the new circuit for the thickness-expander plate (Fig. 2).

It is possible to interpret the three elements on the secondary side of the transformer in Mason's circuit (Fig. 1) as an acoustic transmission line² (by again recognising that they form the T equivalent circuit of the line); however, note that the acoustic forces V_1 and V_2 are not developed across this transmission line alone, but are developed partly across the line terminals and partly across the secondary of the transformer. In the new circuits, on the other hand, the acoustic forces appear directly across the transmission-line terminals. This is physically more reasonable, and permits a clear distinction to be drawn between the lumped-element

electrical behaviour and the wave acoustic behaviour of the transducer.

In each of the new circuits, it is easy to see the effect on the acoustic transmission line of an arbitrary impedance connected to the electrical port. This is not the case in the corresponding circuits of References 1 and 2.

Circuits similar to those presented here may be derived for transducers excited by nonuniform field distributions or with nonuniform piezoelectric properties. In particular, for certain electrical interconnections, a single simple circuit resembling those in Figs. 2 and 3 may be drawn to represent an array of transducers of alternating polarities. Such circuits will be discussed in an article currently in preparation.

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27th May 1970

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ELECTRONIC MODE OF CONTROL TO OBTAIN INCREASED TORQUE AND IMPROVED POWER FACTOR FROM AN ASYNCHRONOUS MACHINE

Indexing terms: Asynchronous machines, Power-factor correction, Torque

It is indicated that, by changing the electronic switching mode of the rotor current of an induction machine, it is possible to operate the machine at improved (capacitive) power factors and increased torque, or conversely at lower effective current and capacitive power factors at rated torque.

Introduction: The control of wound-rotor slip-ring induction machines by insertion of an external impedance in the rotor circuit has been applied in the past. To a certain extent these methods have had electronic or static counterparts for a long time. The advent of compact and reliable semiconductor switching elements has stimulated the interest in all types of electronic control schemes anew. Electronic control of the rotor current of induction machines by employing thyratrons or thyristors to switch in a delayed ignition angle mode has been proposed and investigated, for instance, by Erlicki *et al.*¹ and Shepherd *et al.*² These methods employ an antiparallel arrangement of controlled rectifiers utilising natural commutation as had been proposed long ago by Lenz.³ The purpose of this letter is to suggest a new mode of control without the drawbacks of the above-mentioned schemes, and to present some experimental results.

Theoretical considerations: Consider a multiphase induction machine to be controlled in its rotor. When the angle of ignition of a pair of antiparallel controlled rectifiers is delayed, it is found that the power factor of the circuit decreases with the decrease in current. This is brought about by the natural commutation employed by this method. To compare the conventional method of electronic control and the method to be proposed, an extremely simplified model for the controlled induction machine will be assumed. This should serve only to illustrate the essentials.

Assume that the stator resistance and leakage reactance of the considered machine to be so low that air-gap flux of essentially constant magnitude, rotating with uniform velocity, is established in the machine. This results in a sinusoidally

induced e.m.f. e_r^x , as shown in Fig. 1. Consider the x th phase:

$$e_r^x = \sqrt{2} E_r \sin \omega_r t = s \hat{E}_{r0} \sin \omega_r t \quad (1)$$

where ω_r is the rotor angular frequency, and s is the electromagnetic slip. (Refer to Fig. 1a.) Conventionally the con-

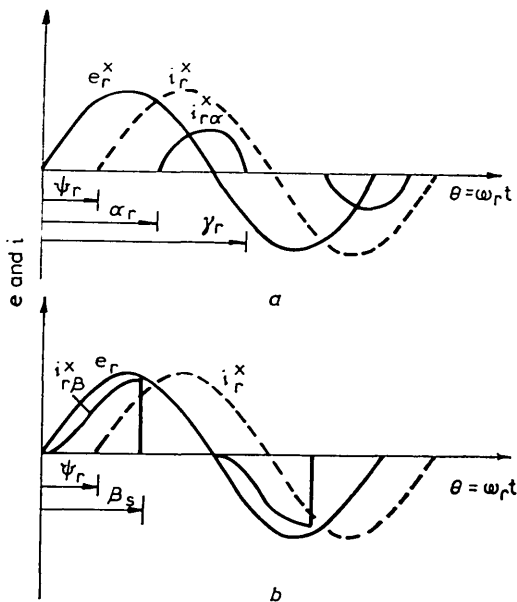


Fig. 1 Definitions and idealised waveforms
a Current-ignition control or α_r control
b Current-extinction control or β_r control

duction is delayed over an angle α_r , where α_r is subject to the constraint

$$\alpha_r > \psi_r$$

for symmetrical operation, and ψ_r is the steady-state uncontrolled phase angle between induced voltage e_r^x and rotor phase current i_r^x . The extinction angle γ_r is subject to

$$\gamma_r < \pi + \psi_r$$

resulting in the current flow indicated in Fig. 1a.

Now power-factor improvements for passive inductive circuits have been suggested recently by Emanuel-Eigeles *et al.*,⁴ and will now be applied to the present instance of a rotating machine. It is proposed that the current be started

at the condition of zero induced rotor voltage, and extinguished by forced commutation at angle β_r , where periodicity results in the constraint

$$\beta_r < \pi \quad (\text{see Fig. 1b})$$

An approximate expression for the electromagnetic torque of the machine may now be taken to be

$$T_e = \frac{mp}{2\omega_s} \hat{E}_{r0} \hat{I}_{r1} \cos \psi_1 \quad (2)$$

with the rotor current expressed as the series

$$i_r^x(t) = \hat{I}_r \sum_{n=1}^{\infty} [\sqrt{(a_n^2 + b_n^2)} \{\sin(n\omega_r t + \psi_n)\}] \quad (3)$$

where

$$\hat{I}_r = s \hat{E}_{r0} (R_r^2 + \omega_r^2 L_{r\sigma}^2)^{-1/2} \quad (4)$$

and R_r and $L_{r\sigma}$ are the equivalent rotor resistance and leakage inductance, respectively, where the m -phase, $2p$ -pole machine has a peak value of induced rotor voltage \hat{E}_{r0} at standstill, and ψ_1 is the phase angle between the fundamental current component and the induced voltage. The usual Fourier coefficients of n th order are given by a_n and b_n . Evaluation of the coefficients results in the torque for the two cases being

$$T_{e\alpha} = \frac{mp \hat{E}_{r0}^2}{2\pi\omega_s \{(R_r/s)^2 + (\omega_s L_{r\sigma})^2\}^{1/2}} \times \left[\left\{ (\gamma_r - \alpha_r) \cos \psi_r - \frac{\sin(2\gamma_r - \psi_r) - \sin(2\alpha_r - \psi_r)}{2} \right\} + \frac{2 \sin(\alpha_r - \psi_r)}{\operatorname{cosec}^2 \psi_r} \{(\cot \psi_r \sin \gamma_r + \cos \gamma_r) \times \exp - (\gamma_r - \alpha_r) \cot \psi_r - (\cot \psi_r \sin \alpha_r - \cos \alpha_r)\} \right] \quad (5)$$

$$T_{e\beta} = \frac{mp \hat{E}_{r0}^2}{2\pi\omega_s \{(R_r/s)^2 + (\omega_s L_{r\sigma})^2\}^{1/2}} \times \left[\left\{ \beta_r \cos \psi_r - \frac{\sin(2\beta_r - \psi_r) + \sin \psi_r}{2} \right\} - \frac{2 \sin \psi_r}{\operatorname{cosec}^2 \psi_r} \{(\cot \psi_r \sin \beta_r + \cos \beta_r) \exp - \beta_r \cot \psi_r - 1\} \right] \quad (6)$$

Calculation of the torque/speed characteristics from eqns. 2-6, with the control angles α_r , β_r as parameters for a 2-pole 2-phase machine represented by the basic parameters $E_{r0} = 39$ V, $R_r = 0.995 \Omega/\text{phase}$ and $L_{r\sigma} = 14.4$ mH/phase, results in the relations shown in Figs. 2a and b for the two cases, respectively. These Figures indicate the interesting fact that without changing the machine parameters, but merely the switching mode, it becomes possible to obtain torques much in excess of the normal maximum torque attainable with the machine. On the other hand, the same torque should be obtainable at a much lower machine current, since the power factor has been increased.

Regarding the calculations for the current extinction or β_r control, it must be remarked that Fig. 2b reflects pessimistic values for the basic system parameters cited above. In order to approach the true experimental arrangement more closely, the basic values were adapted in order to take into account resistive and reactive effects of the forced commutation switch inserted into the rotor of the experimental machines. The values were $E_{r0} = 39$ V, $R_r = 2.04 \Omega/\text{phase}$ and $L_{r\sigma} = 17.5$ mH/phase.

Experimental verification: Although it is not to be expected that the simple model used would adequately describe the asynchronous machine with a periodically operated electronic switch in the rotor, preliminary experiments have indicated that the same type of effect may be observed on a practical machine. A pair of antiparallel thyristors for α_r control and

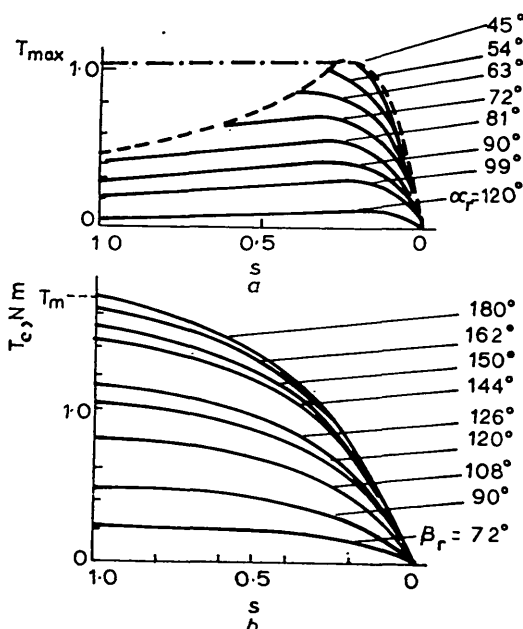


Fig. 2 Examples of torque/speed characteristics for the electronic rotor-control systems
a α_r control
b β_r control

Table 1 COMPARISON BETWEEN OPERATING MODES FOR CITED INDUCTION MACHINE

Speed rev/min	Stator voltage		Stator current		Rotor current		Power factor	
	V_s	V_s'	I_s	I_s'	I_r	I_r'	(PF)	(PF)'
0	110	110	4.20	1.22	9.80	3.3	0.306	0.748
300	110	110	4.14	1.24	9.70	3.35	0.319	0.752
2000	110	110	3.55	1.8	8.35	4.65	0.534	0.883

$V_s, I_s, I_r, (PF)$, refer to normal uncontrolled operation
 $V_s', I_s', I_r', (PF)'$ refer to operation with the proposed current-extinction control

a forced commutation switch for β_r control for each rotor phase of a 2-phase Westinghouse generalised machine operated as an induction motor with a wound rotor were used. The effects resulting in a difference between the experimental system and the simplified model presented are:

- (i) iron losses and magnetising current
- (ii) nonzero stator resistance and leakage reactance
- (iii) tooth and winding harmonics
- (iv) finite voltage drop across the naturally or forced commutation switch in the rotor
- (v) impossibility to reduce the rotor current instantaneously to zero.

The experimental ratio

$$\left(\frac{T_m}{T_{max}}\right)_{exp} = 2.2 \dots \dots \dots (7)$$

while theoretically

$$\left(\frac{T_m}{T_{max}}\right)_{theor} = 1.7 \dots \dots \dots (8)$$

The improvement in power factor is clearly indicated by the experimental results in Tables 1 and 2, while the reduction in

Table 2 COMPARISON BETWEEN TYPICAL OPERATING CHARACTERISTICS OF CURRENT-IGNITION CONTROL (α_r CONTROL) AND CURRENT-EXTINCTION CONTROL (β_r CONTROL)

Speed = 1400 rev/min $T_e = 0.5$ p.u. = $0.25 T_{max}$				
	Stator voltage	Stator current	Rotor current	Power factor
α_r control	V 109.75	A 2.390	A 4.90	0.288 ind
β_r control	V 110.00	A 0.835	A 2.15	0.560 cap

stator and rotor currents to obtain the same torque may also be noted. Table 2 indicates how the current-ignition control and current-extinction control results in inductive and capacitive power factors, respectively, proving how the

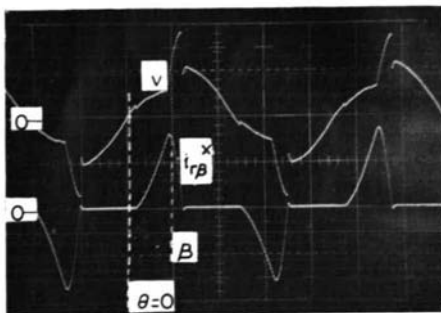


Fig. 3 Observed waveforms on β_r control system
 5 m s/division horizontal
 Top trace: input voltage to electronic power switch
 Bottom trace: rotor current
 Taken at 300 rev/min, and 0.5 p.u. torque (see Table 2)

β_r control enables the operation of an induction machine at a capacitive power factor.

Fig. 3 gives an indication of the rotor current. The top trace records the input voltage to the rotor-power electronic switch, which has not been discussed in this letter. This trace gives an indication of the rotor zero-voltage condition as shown. It may be seen that, owing to electronic-control considerations, the rotor current was not started at $\theta = 0$, but some time later. The time necessary to commutate the rotor current to zero is also evident from Fig. 3.

Conclusion: It has been indicated theoretically and experimentally that, by employing electronic current-extinction control in the rotor circuit of an induction machine, it is possible to achieve much lower currents at rated torques, or conversely much higher torques at rated currents. This method of control results in capacitive power factors for the induction-machine electronic-control system.

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27th May 1970

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EXCESS LEAKAGE-CURRENT NOISE IN JUNCTION FIELD-EFFECT TRANSISTORS

Indexing terms: Field-effect transistors, Leakage current, Noise

Measurements are reported of the excess gate leakage current I_G in several n channel f.e.t.s, showing that I_G varies exponentially with the inverse square root of the bias voltage between drain and gate. I_G shows full shot noise, together with a component of the form $I_{gn}^2 \propto I_G^\beta / f^\alpha$, where values of 1.4 and 1.6 have been found for α and β , respectively.

Fowler¹ has given measurements of the gate leakage current of junction field-effect transistors, showing that the leakage current under certain operating conditions can be considerably larger than that measured with zero drain current. The excess leakage current was found to be proportional to drain current I_D , and increased very rapidly