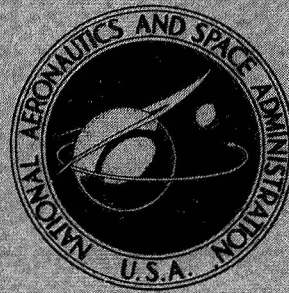


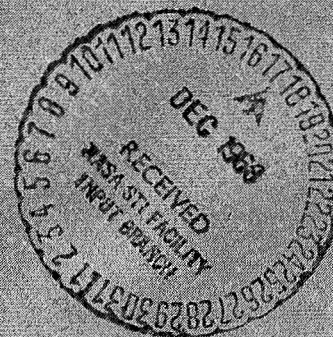
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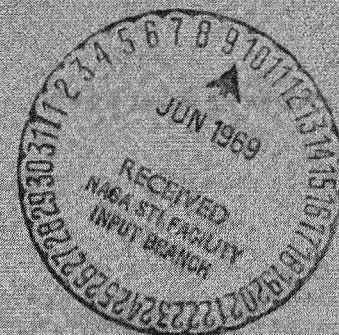


**STUDY AND DEVELOPMENT OF HIGH
EFFICIENCY INDUCTION MOTOR
AMPLIFIER SYSTEM**

by Sherwin Greenblatt and
John J. Wawzonek

Prepared by

BOSE CORPORATION
Natick, Massachusetts
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STUDY AND DEVELOPMENT OF HIGH EFFICIENCY
INDUCTION MOTOR AMPLIFIER SYSTEM

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ABSTRACT

This report examines several areas important to the design of variable speed induction motor systems with variable voltage, variable frequency drive. These systems are composed of the induction motor, the DC-AC inverter, and an overall control scheme.

The AC induction motor is analyzed to determine its inherent capabilities in these systems. Also, the motor is analyzed to determine its impedance and torque characteristics with non-sinusoidal excitation. It is shown that the design of the induction motor should minimize rotor resistance in order to maximize efficiency and to reduce spurious torques due to non-sinusoidal excitation.

The various DC-AC inverters are reduced to three basic forms:

- 1) The base band inverter. (That is, an inverter whose switching frequency is equal to the output frequency.)
- 2) The cycloconverter, and
- 3) The pulse modulator.

These inverters are analyzed to determine inherent performance capabilities and limitations with various applicable modulation schemes. Design considerations of power processing stages are discussed including an analysis of some thyristor forced commutation schemes. Specific performance requirements for inverters are discussed in terms of motor requirements.

The various means of controlling the motor parameters (e.g. open loop and closed loop systems for control of torque, flux, speed, etc.) are discussed. The trade-offs between complexity and performance capability are presented.

The integration of the components into a system is discussed and recommendations are given for future work.

FORWARD

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1. INTRODUCTION

The combination of the AC induction motor with a solid state inverter drive represents an attractive solution to the problem of design of an efficient and reliable servo system. The AC induction motor is inherently simple and reliable: with one moving part and friction at bearing surfaces only, it is particularly suited to the requirements of space applications.

The DC-AC inverter portion of the variable speed induction motor (VSIM) system must be a source of controlled variable frequency-variable voltage excitation to exploit the full performance and efficiency capability of the motor. While AC induction motor technology has advanced to the point where large motors of suitable design are available, the design of solid state inverters to drive induction motors of even moderate power (10 KVA) represents a very recent advance in the state-of-the-art. The detailed theory of operation of these units is not yet well understood. As might be expected, it is in the inverter and particularly in the power output stage where most of the design effort must be expended. Therefore, the crux of the VSIM drive problem lies in the successful realization of the inverter power output stage.

Intimately associated with the design of the power processing stages of the inverter are three areas: 1) The nature of the load requirements the motor presents to the inverter; 2) the manner in which control of power is achieved in the inverter; and 3) the control scheme that couples the inverter to the motor. These three aspects, in addition to the basic design considerations of the power processing stages, are the subject of this report.

Basic to the understanding of the manner in which the power to the motor must be controlled (i. e., the choice of voltage or current and the choice of frequency) is a complete understanding of the drive the motor requires to achieve full utilization of its performance capability. Sections 2.1 and 2.2 present the basic characteristics of the AC induction motor and apply them to an analysis of the motor performance in the VSIM system. It is shown that the minimization of the rotor resistance is essential for maximum efficiency and performance. This is followed by a discussion and analysis of the problems of coupling the inverter and motor. Section 2.4 provides a characterization of the motor impedance as a function of the motor speed. The generation of spurious torques with non-sinusoidal excitation is discussed in Section 2.5. It is shown that the minimization of rotor resistance also minimizes spurious torques.

The discussion and analysis of Chapter 2 shows that for highest performance and complete flexibility of operation (four quadrant torque-speed capability) the drive must possess the following characteristics:

1) Sufficient voltage capability to fully flux the motor at the highest stator drive frequency. This insures that maximum torque for optimal dynamic response will be available under all operating conditions.

2) Sufficient peak current capability to realize the maximum available torque under all motor operating conditions and in the presence of harmonic currents.

3) A capability for regenerating as well as supplying AC power. This feature provides optimal response during braking or when driving active loads.

4) A capability for providing controlled drive down to zero frequency. During motor reversing speed changes or when driving an active load at low speeds, DC drive is required to realize optimal motor torque capability.

5) Low harmonic content in the AC drive waveform, particularly at low drive frequencies, is required to minimize harmonic and pulsating torques. At high frequencies, low harmonic and spurious frequency content is required to minimize excess core losses and large stator currents at frequencies other than the drive frequency.

6) The drive must be responsive to variations in the load. For the highest level of performance, the drive to the motor must be adjusted by feedback means to compensate for load changes and to provide optimal transient response.

The study of inverters in Chapter 3 proceeds by reducing the many inverters described in the literature to the fewest simple forms and analyzing their inherent capabilities and limitations. Particular attention is given to various forms of the 1) base band system (switching is at the output frequency); 2) the square wave cycloconverter; and 3) the pulse modulator. The design problems center on those of the output power stage. The necessity for DC output requires that the power stage output switches be connected by a direct path to the terminals and precludes the use of transformer coupled output stages.

The discussion of Section 3.3 centers on control concepts (i. e., the

manner in which switching times and switching frequency are chosen in order to control the flow of power). The results indicate that to insure a small harmonic content at low frequencies, a sine wave approximation method, using samples at a relatively high frequency, should be employed.

Section 3.4 discusses the choice of the output stage switch and the choice of the switch configuration. At the power levels of the systems under study (greater than 2 KW) and the supply voltage levels considered (200 to 500 volts), the use of thyristor switching elements is mandatory. All the inverter systems considered in this study require forced commutation of at least some of the thyristors. The study of switching losses in forced commutation output stages indicates that the McMurray circuit, with a theoretical commutation efficiency of nearly 100% is an attractive choice. Experimental studies performed outside the scope of this contract indicate that this configuration can, in fact, commute at high efficiency and, with some modification, can be made to operate with high reliability. It has been found, however, that instantaneous current control, or at least instantaneous current limiting, is necessary to prevent commutation failure.

The completion of the VSIM system requires some form of overall system control to couple the inverter to the motor. The various parameter control schemes are discussed in Chapter 4. All of the methods of voltage, current, flux, torque and speed control, when properly implemented, will operate the motor. The more sophisticated methods will yield high performance. Several factors that govern the choice of parameter control techniques are evident: 1) Instantaneous current control or limiting must be employed to protect the inverter switching devices under dynamic load and fault conditions; 2) For optimal flux control, either flux feedback or some method of flux drive compensation with load variations must be employed; 3) The method of slip control employed must be studied for the particular motor, inverter, and mission requirements to optimize the performance parameters.

In general it is found that the crucial problems lie in minimization of power loss in the inverter. Recommendations are given for experimental investigation of simple versions of three inverters to determine the optimum system. In particular, the square wave cycloconverter, the pulse modulator, and the series inverter should be investigated.

2. INDUCTION MOTOR CHARACTERISTICS

2.1 BASIC TORQUE AND POWER RELATIONS

To proceed with an outline of the basic relations governing the motor performance, consider the per phase equivalent circuit shown in Figure 2.1. Power flow to the right through terminals a - a' represents power flowing from the stator of the motor into the air gap. We shall denote this power by the symbol P_g . In terms of the parameters of Figure 2.1, the real part of this power can be expressed per phase as

$$P_g = E_1^2 \frac{\frac{r_2}{s}}{\left(\frac{r_2}{s}\right)^2 + \left(L_2 \omega_s\right)^2} . \quad (2.1)$$

The internal mechanical torque, T , of any induction motor is related to P_g by the equation

$$T\omega_s = P_g . \quad (2.2)$$

Hence, the torque, per phase, can be expressed

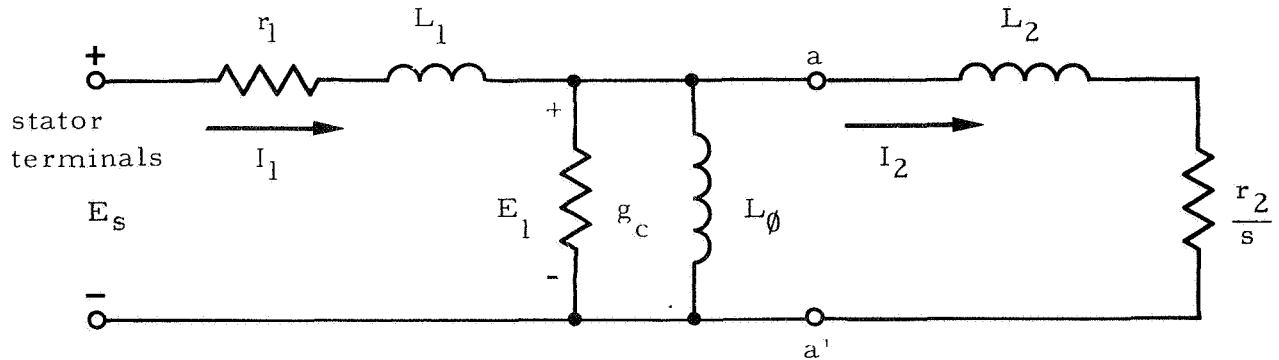
$$T = \frac{E_1^2}{\omega_s} \frac{\frac{r_2}{s}}{\left(\frac{r_2}{s}\right)^2 + \left(L_2 \omega_s\right)^2} . \quad (2.3)$$

We now define the slip frequency

$$\omega_r = \omega_s - \omega_m \quad (2.4)$$

where ω_s is the line frequency and ω_m the motor speed. The quantity ω_r is also the electrical frequency seen in the rotor. The rotor electrical time constant is defined as

$$\tau = \frac{L_2}{r_2} . \quad (2.5)$$



r_1 = Resistance of Stator Winding

r_2 = Resistance of Rotor Winding as Seen From the Stator

g_c = Conductance to Account for Core Losses. (In general this will be a function of the terminal voltage and frequency.)

L_1 = Leakage Inductance of Stator Winding

L_2 = Leakage Inductance of Rotor Winding as Seen From the Stator

L_ϕ = Magnetizing Inductance

s = Slip = $(\omega_s - \omega_m) / \omega_s$ where ω_s = Line Frequency
and ω_m = Mechanical Angular Velocity of Rotor

E_s = RMS Value of Stator Terminal Voltage

E_1 = RMS Value of Back EMF

FIGURE 2.1

PER PHASE EQUIVALENT CIRCUIT FOR INDUCTION MOTOR
(Valid for two pole motor.)

Thus, we can write Eqn. 2.3 in the form

$$T = \frac{E_1^2}{L_2 \omega_s^2} \frac{\tau \omega_r}{1 + \tau^2 \omega_r^2} \quad (2.6)$$

If we now hold $\frac{E_1}{\omega_s}$ constant and vary ω_r we find that the maximum torque for a given flux in the motor is given by

$$T_{\max} = \frac{E_1^2}{2L_2 \omega_s^2} \quad (2.7)$$

and it occurs at a rotor electrical frequency given by

$$\omega_{r_{\max T}} = \frac{1}{\tau} = \frac{r_2}{L_2} \quad (2.8)$$

which corresponds to a slip

$$s_{\max T} = \frac{r_2}{L_2 \omega_s} \quad .$$

Eqns. 2.6 and 2.7 can be used to define a single normalized torque characteristic that is applicable to all induction motors. This general relation is

$$\frac{T}{T_{\max}} = 2 \frac{\tau \omega_r}{1 + \tau^2 \omega_r^2} \quad (2.9)$$

where τ again is the rotor time constant, L_2/r_2 , and ω_r is the electrical frequency seen in the rotor. This relation is plotted in Figures 2.2 and 2.3. Figure 2.3, a plot of $\frac{T}{T_{\max}}$ versus $\tau \omega_r$, will be particularly useful later when we discuss torque generation for different drive frequencies in order to understand the spurious torques produced by non-sinusoidal drives.

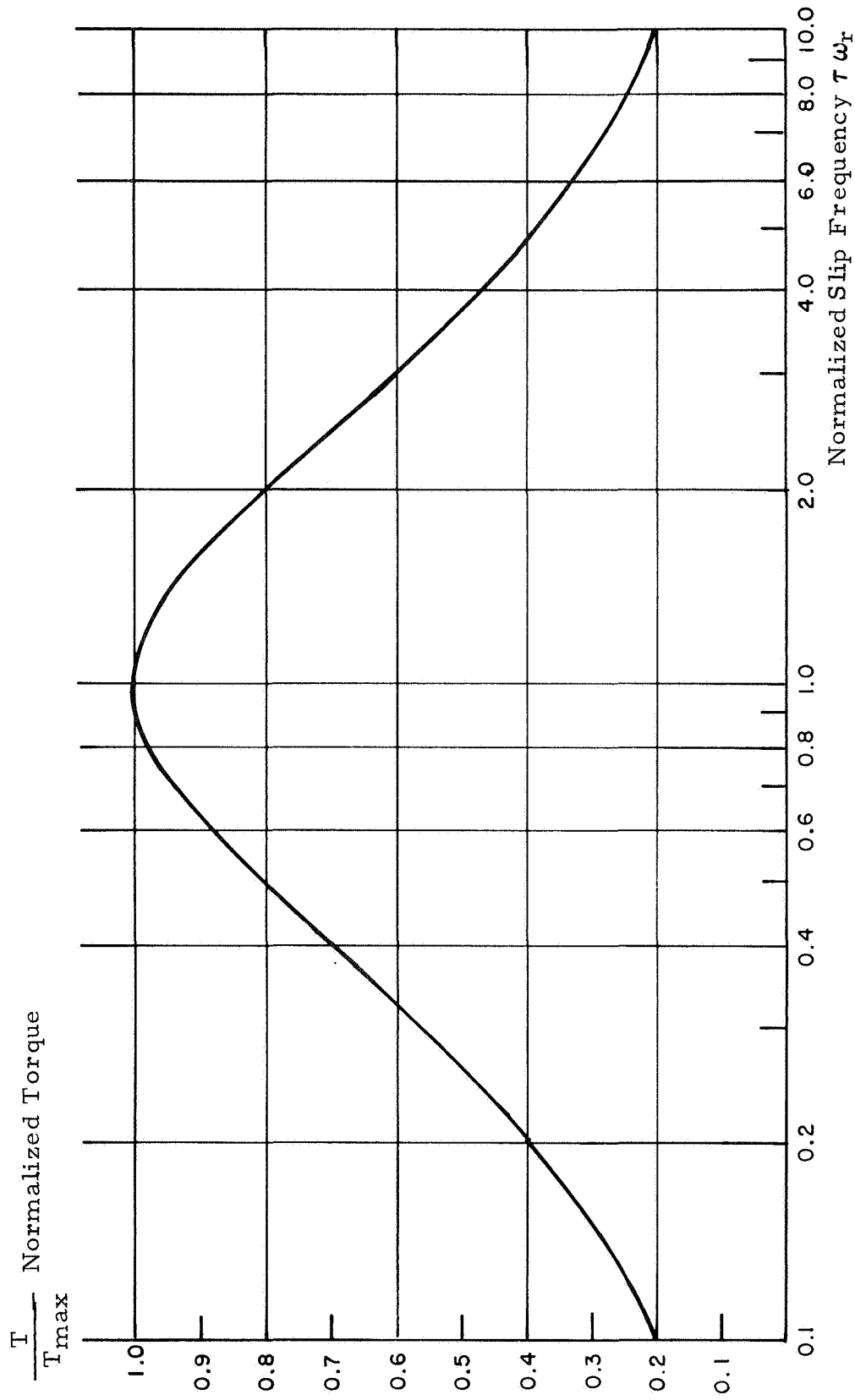


FIGURE 2.2

NORMALIZED INDUCTION MOTOR TORQUE-FREQUENCY CHARACTERISTIC

Logarithmic Plot

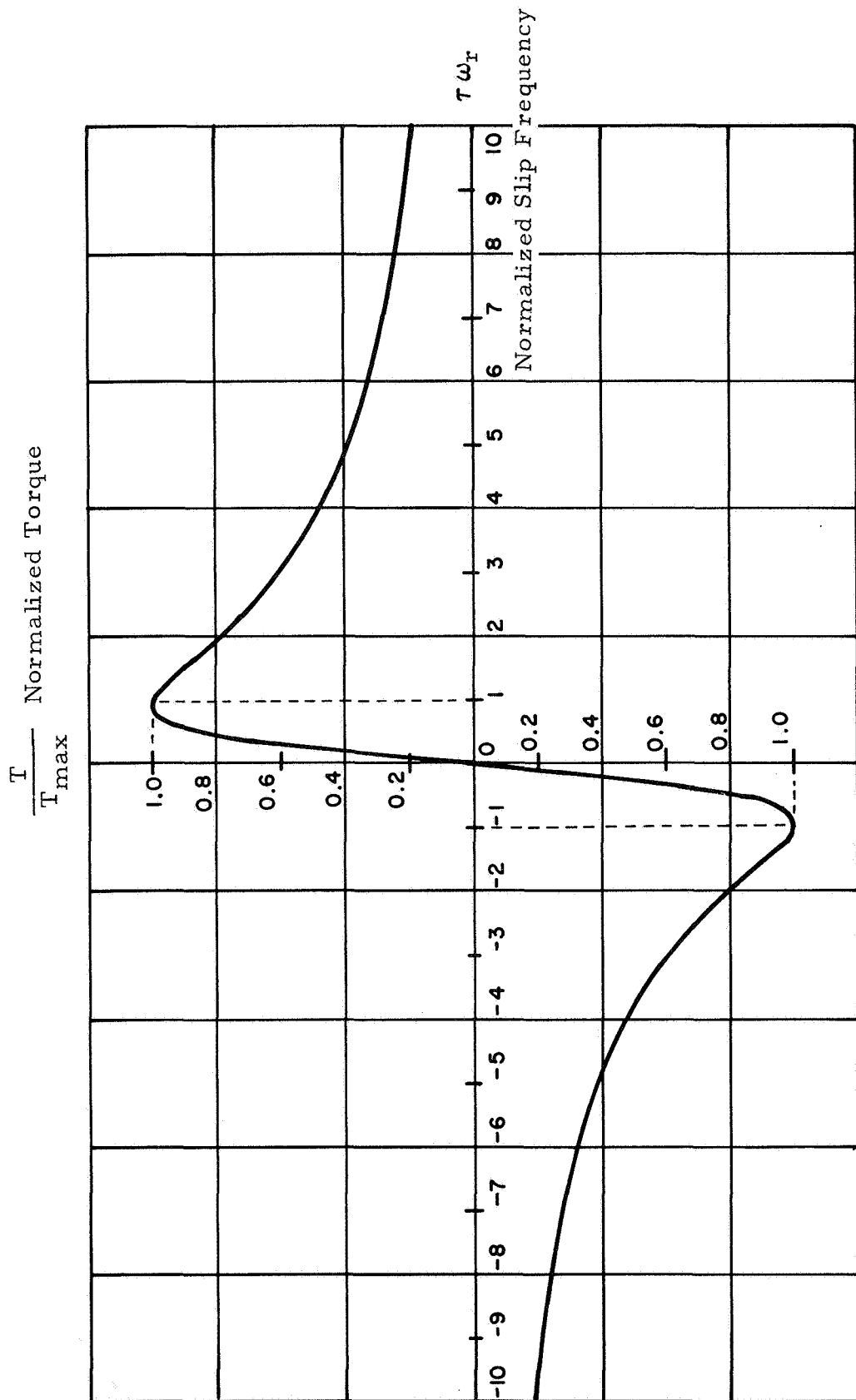


FIGURE 2.3

NORMALIZED INDUCTION MOTOR TORQUE-FREQUENCY CHARACTERISTIC

Linear Plot

The significance of Eqns. 2.7 and 2.9 is that they show that the maximum torque is independent of the rotor resistance r_z and the rotor frequency ω_r .

Also we see that maximum torque is produced at a rotor speed given by the relationship

$$\omega_m = \omega_s - \frac{1}{\tau} . \quad (2.10)$$

This equation shows that maximum torque can be obtained for any desired rotor speed ω_m including zero by proper adjustment of the drive frequency ω_s . In other words, the curve representing Eqn. 2.9 can be made to translate along the frequency axis so that its maximum lies at any desired rotor speed. Thus, we see also that if the motor is operated at constant flux,

i. e. $\frac{E_1}{\omega_s} = \text{constant}$, then we can obtain constant torque if $\omega_r = \omega_s - \omega_m$ is

maintained constant. The graphs of Figure 2.4 show several members of the family of torque speed curves for indicated values of $\frac{\omega_s}{\omega_{so}}$, where ω_{so}

is the maximum available line frequency. In these graphs, the reverse torque region that is useful for braking or reversing is also indicated. It is apparent that any torque, T , satisfying the relation $-T_{\max} \leq T \leq T_{\max}$ can be obtained at any desired speed by properly controlling ω_s .

Returning to the power relations, we note that the power transmitted to the rotor is given by Eqn. 2.2

$$P_g = T\omega_s . \quad (2.2)$$

The internal mechanical power is then

$$P_m = T\omega_m . \quad (2.11)$$

Thus, the power dissipated in the rotor is

$$P_r = T (\omega_s - \omega_m) = T\omega_r . \quad (2.12)$$

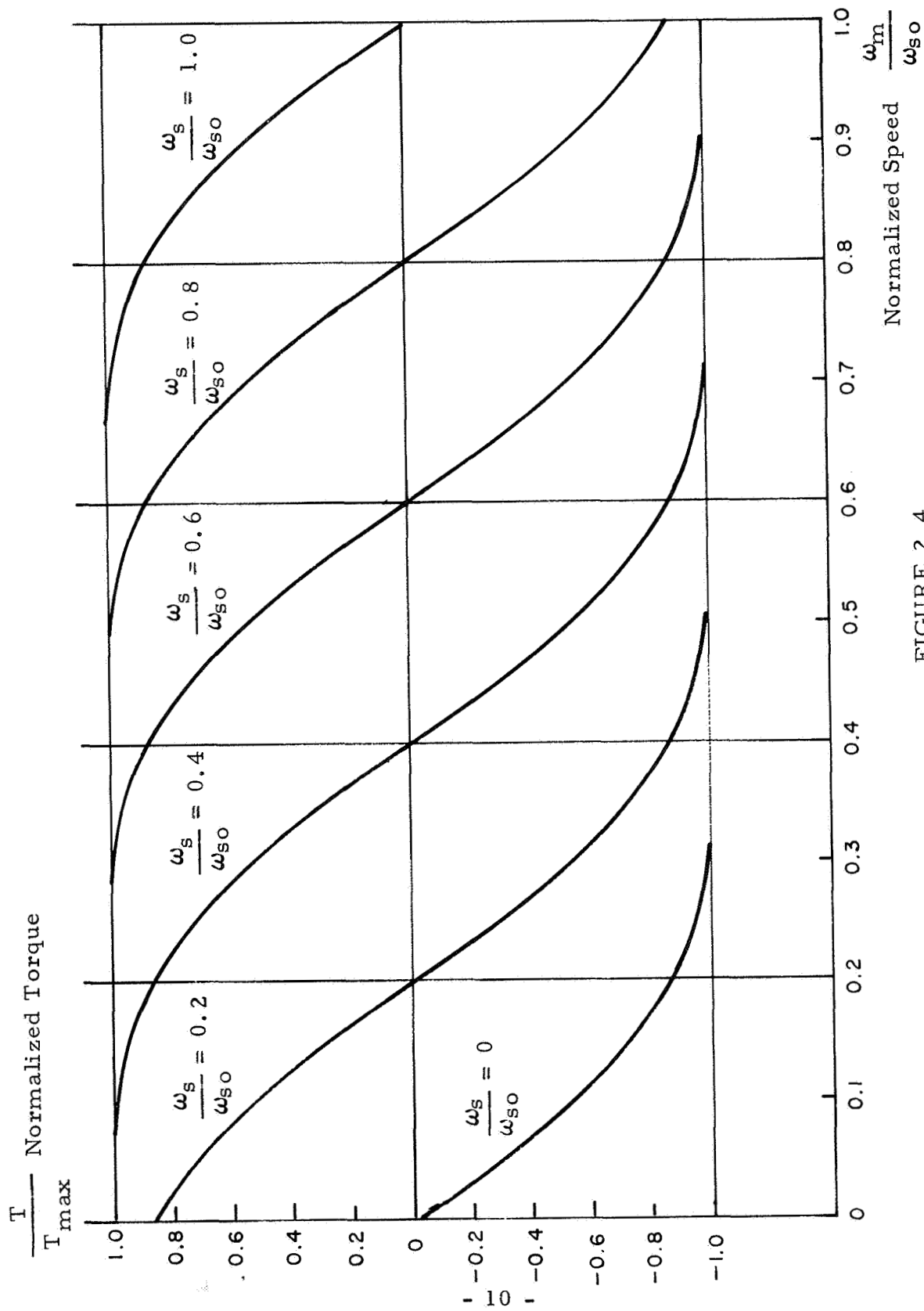


FIGURE 2.4

FAMILY OF NORMALIZED TORQUE-SPEED CURVES
 ($\tau = 0.008$, $\omega_{so} = \text{Maximum available line frequency} \approx 377 \text{ radians/second}$)

The rotor efficiency, η_r , is the ratio of power output, P_m , to the power transmitted across the gap, P_g , and is given by

$$\eta_r = \frac{P_m}{P_g} = \frac{\omega_m}{\omega_s} . \quad (2.13)$$

We note that η_r is just the ratio of rotor speed to excitation frequency.

The total real power transmitted across the gap has been given as $P_g = T\omega_s$. This power must be equal to that absorbed in the resistor $\frac{r_2}{s}$. Thus,

$$P_g = I_2^2 \frac{r_2}{s} = I_2^2 r_2 \left(1 + \frac{\omega_m}{\omega_r} \right) \quad (2.14)$$

$$P_g = T(\omega_r + \omega_m) . \quad (2.15)$$

We can, therefore, split the resistor $\frac{r_2}{s}$ into two parts,

$$\frac{r_2}{s} = r_2 + r_2 \frac{\omega_m}{\omega_r} \quad (2.16)$$

and identify the rotor dissipation as the power dissipated in r_2 , and the power converted into mechanical energy as that dissipated in $r_2 \frac{\omega_m}{\omega_r}$. This is expressed as

$$P_r = T\omega_r = I_2^2 r_2 \quad (2.17)$$

$$P_m = T\omega_m = I_2^2 r_2 \frac{\omega_m}{\omega_r} . \quad (2.18)$$

2.2 PERFORMANCE CHARACTERISTICS OF THE VSIM SYSTEM

Available Internal Torque

Let us assume that the maximum allowable drive frequency is ω_{s0} (this is an arbitrary limitation and is only loosely set by the motor design).

Figure 2.5 is a plot of $\frac{T}{T_{\max}}$ versus $\frac{\omega_m}{\omega_{s0}}$ under two conditions of opera-

tion. The bottom curve shows the typical servo motor with constant line frequency. The upper two curves show the motor with variable line frequency and two different rotor time constants. In the region where these curves are flat ω_s is adjusted to operate the motor at the peak of the torque characteristic. However, at the peak $\omega_r = \omega_s - \omega_m = 1/\tau$.

Thus, with a specified maximum ω_s , the large values of ω_m do not permit satisfying this relation and the available torque drops off. We see, however, that significantly more torque is available in the variable frequency system, and that best performance is achieved with the largest rotor time constant. (Note that this really means reducing the resistance, r_2 , and not increasing L_2 since doing so would reduce $T_{\max}$.)

Available Internal Mechanical Power

By multiplying the curves of Figure 2.5 by $\frac{\omega_m}{\omega_{s0}}$, we obtain the corresponding curves for the available internal mechanical power as shown in Figure 2.6. Thus, optimum rotor design along with frequency control makes it possible to extract appreciably more power from a given size motor at a given speed.

Rotor Dissipation

It is interesting to compare the rotor losses for the two conditions of operation. The graphs of Figure 2.7 show the normalized rotor losses for the two modes of operation at the maximum system capacities indicated by the curves of Figures 2.5 and 2.6. The reduction of the rotor losses at

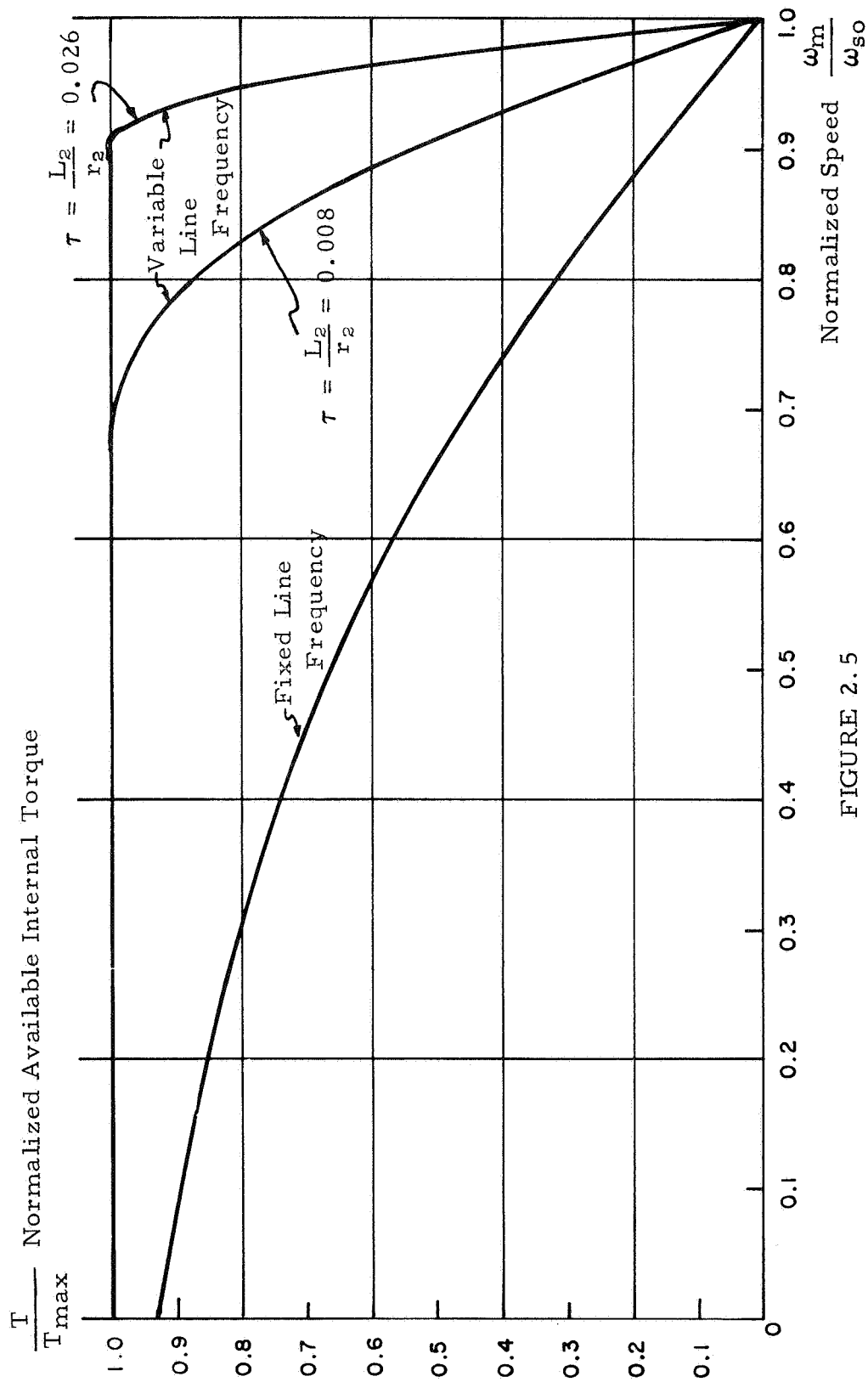


FIGURE 2.5

AVAILABLE INTERNAL TORQUE VS. $\frac{\omega_m}{\omega_{so}}$
 (ω_{so} = Maximum available line frequency.)

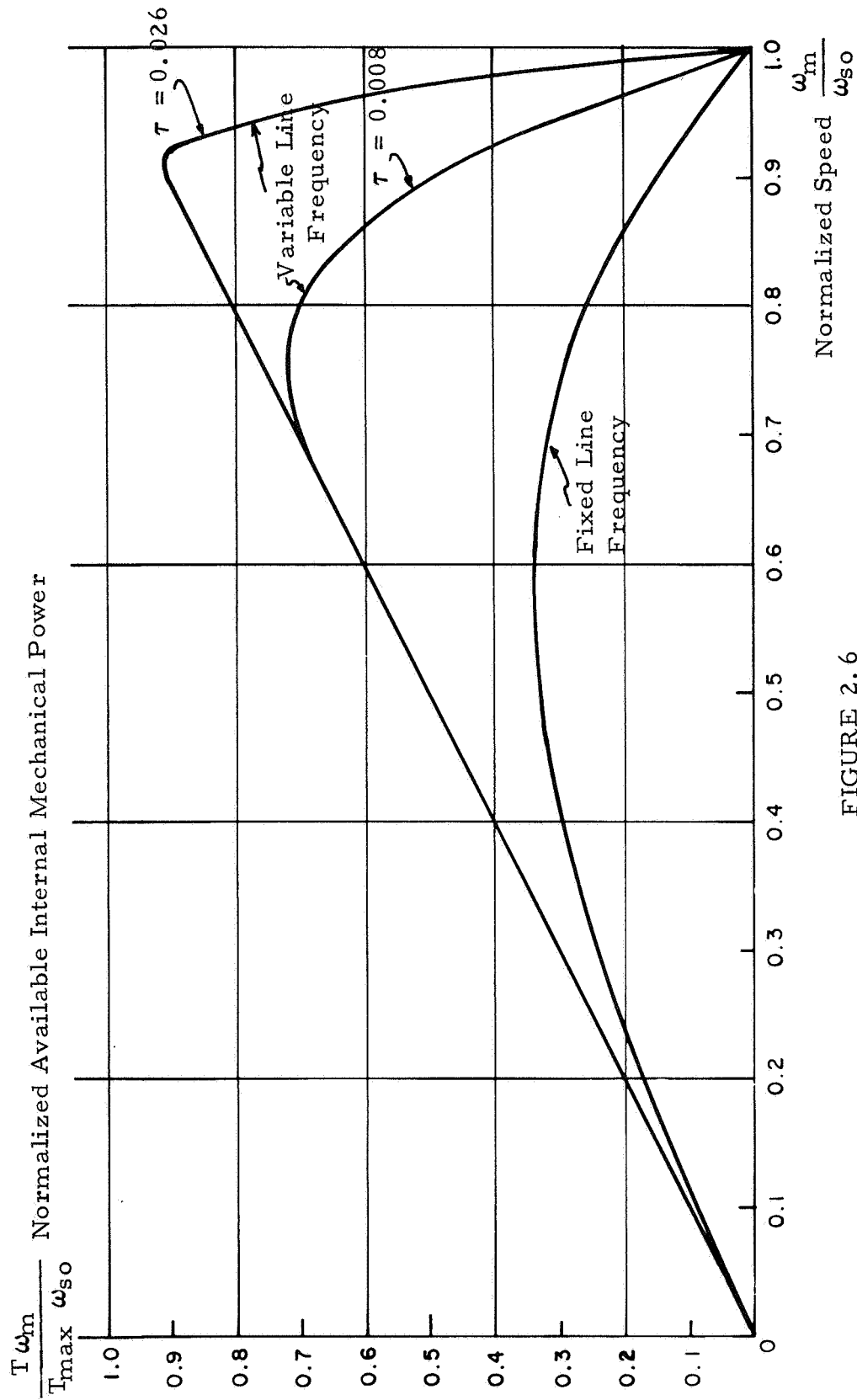


FIGURE 2.6

AVAILABLE INTERNAL MECHANICAL POWER VS. $\frac{\omega_m}{\omega_{so}}$

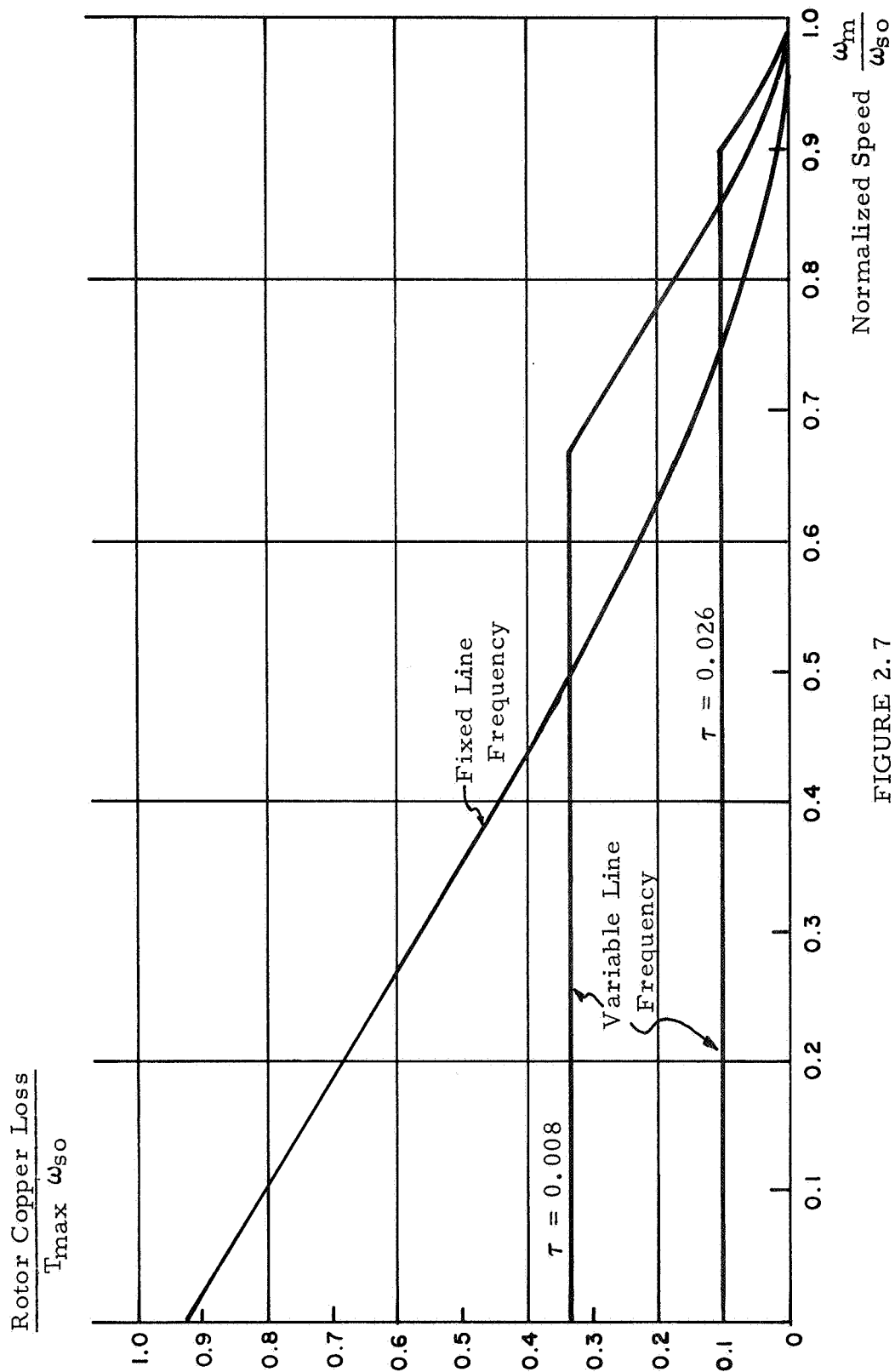


FIGURE 2.7

ROTOR COPPER LOSS VS. $\frac{\omega_m}{\omega_{so}}$

(Output torque conditions as shown in Figures 2.5 and 2.6.)

starting with variable frequency operation is significant not only with respect to efficiency but with respect to the primary limitation of the motor which is the heat generated in the rotor. The curves of Figure 2.7 indicate that large reductions in rotor losses at low speeds are possible with variable frequency operation. In studying the curves of Figure 2.7 it must be remembered that the torque and power delivered by the variable frequency system are greater at all speeds than that delivered by the conventional system. Thus, the fact that rotor losses in the variable frequency system at high speeds are greater than that of the typical servomotor is only a reflection of the fact that the proposed system is delivering greater power at these speeds. In order to illustrate this, the rotor losses have been replotted in Figure 2.8 for the condition that both systems are delivering the identical torque, and hence power, at all speeds. The magnitude of the reduction in rotor losses at all speeds is apparent from the curves.

We may also look at the rotor losses from the standpoint of rotor efficiency as a function of the motor speed. The rotor efficiency, η_r , given by Eqn. 2.13 is plotted in Figure 2.9 for the system operating conditions of Figures 2.5 and 2.6. It is apparent from Figure 2.9 that variable frequency operation significantly enhances the rotor efficiency in the low speed region where it is most needed.

Rotor Acceleration

The curves of Figure 2.10 show the times of acceleration of an inertial load to a given speed for the two modes of operation, under the assumption that the same T_{max} holds in each case. From these curves we can see that it is possible to obtain as much as a factor of two reduction in the time necessary to reach 0.95 of synchronous speed. Also, we can take advantage of the reduced rotor losses to increase the available torque for a given size motor and possibly gain still another factor of three or four in acceleration time.

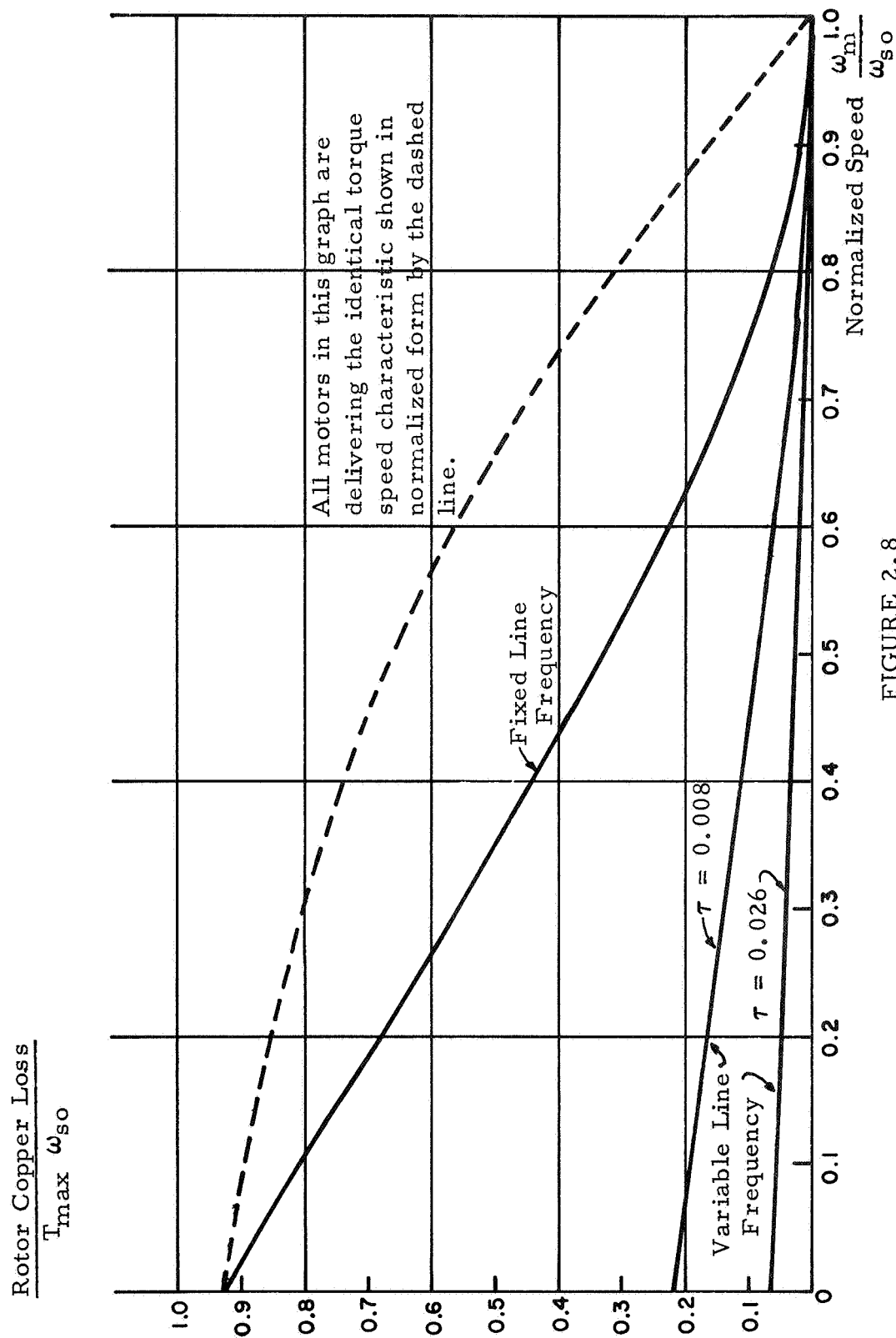


FIGURE 2.8

ROTOR LOSSES UNDER IDENTICAL TORQUE SPEED CONDITIONS

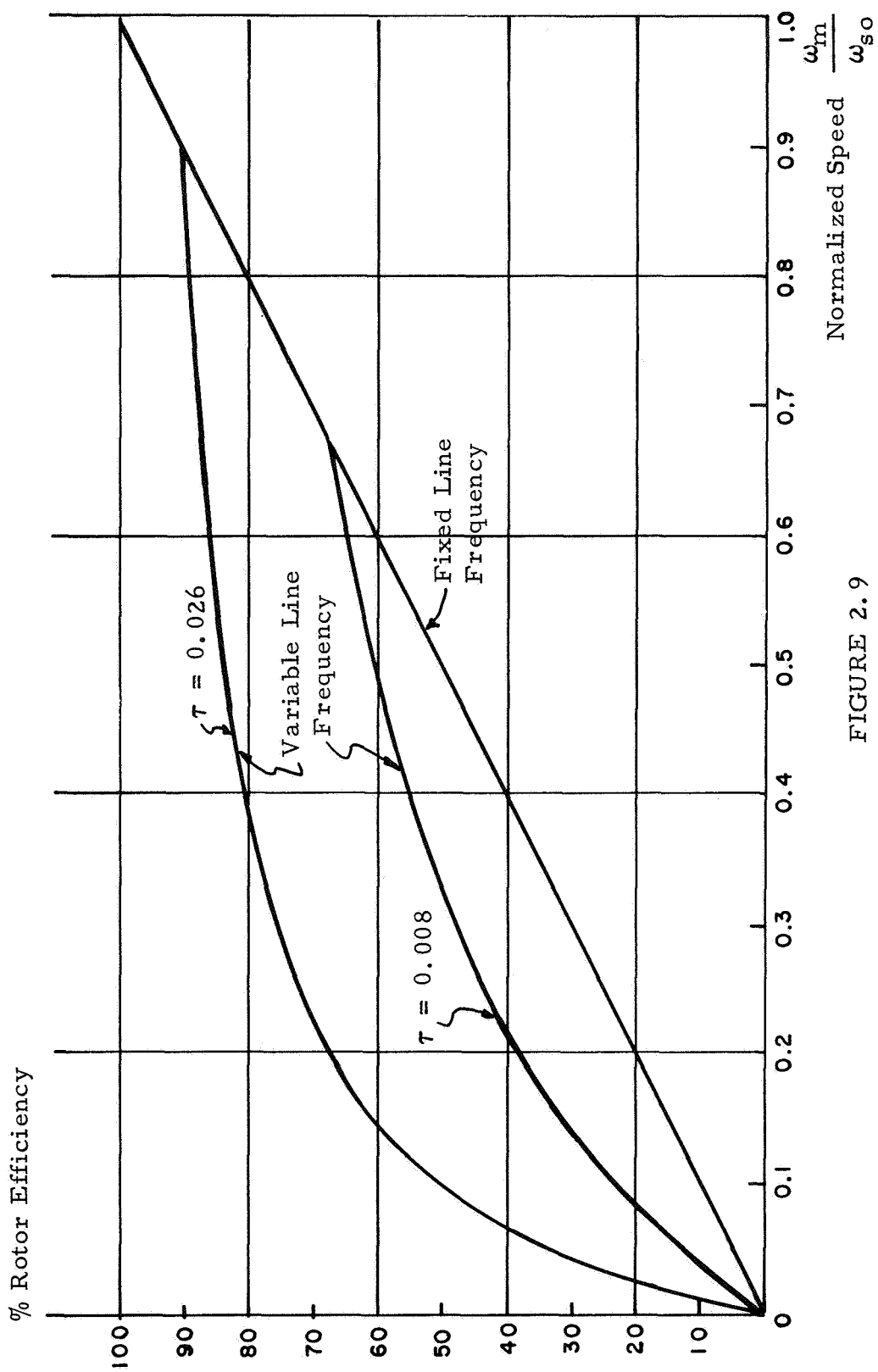


FIGURE 2.9

ROTOR EFFICIENCY FOR TORQUE SPEED CONDITIONS OF FIGURE 2.5

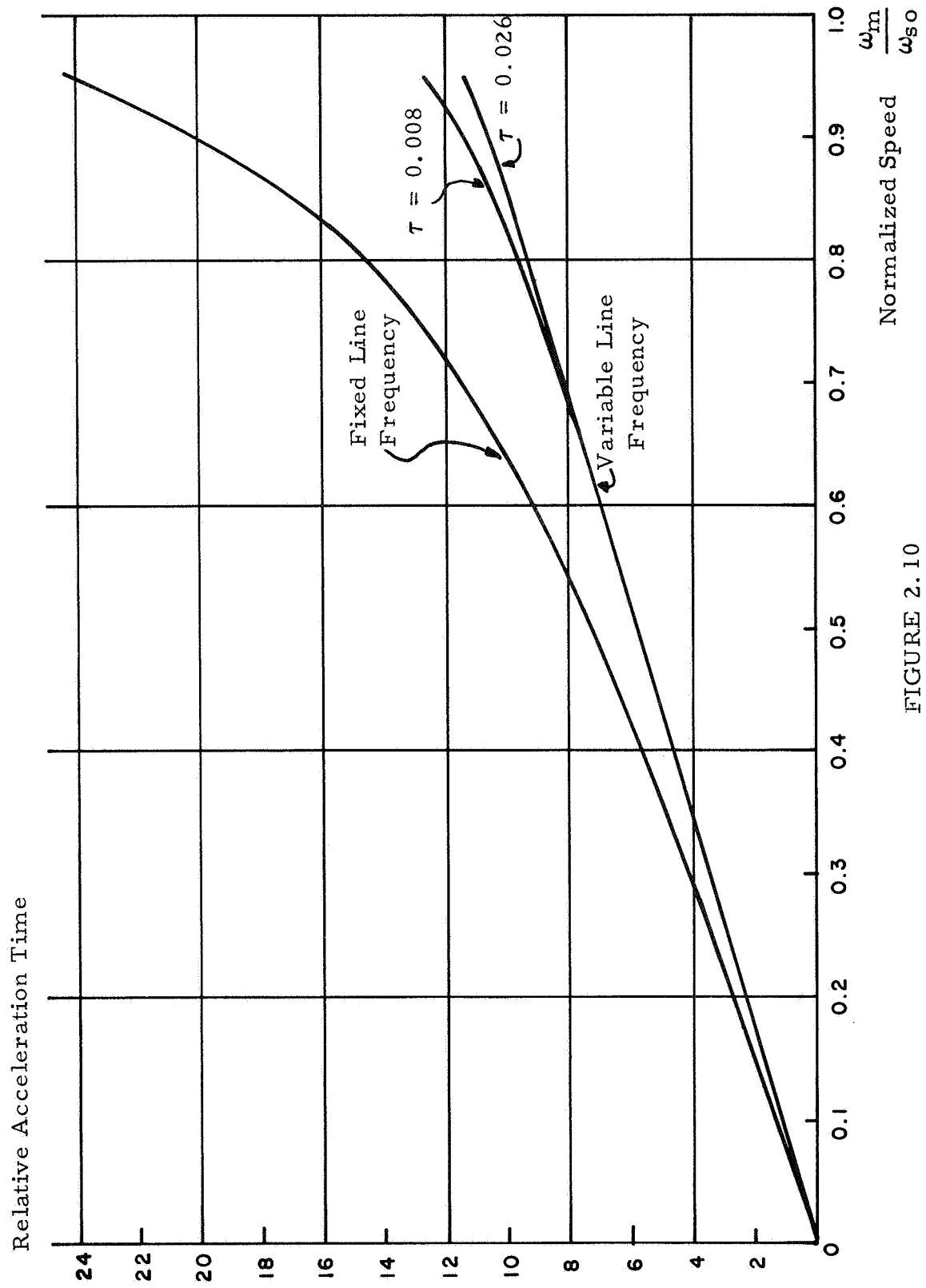


FIGURE 2.10

RELATIVE ACCELERATION TIME FOR INERTIAL LOAD

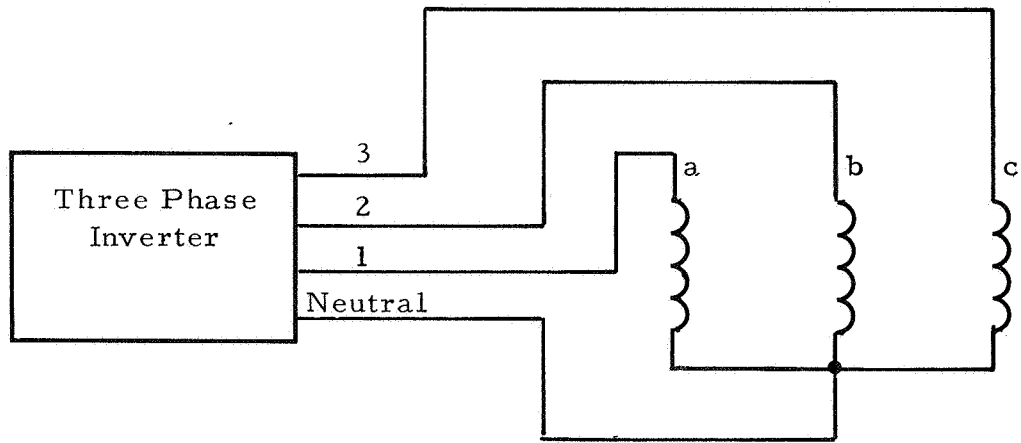
2.3 MOTOR VOLTAGES IN THE THREE PHASE CONNECTION

We shall illustrate in this section the various possible connections of a three phase motor to a three phase inverter, along with the various voltage waveforms and spectra for a square wave drive signal.

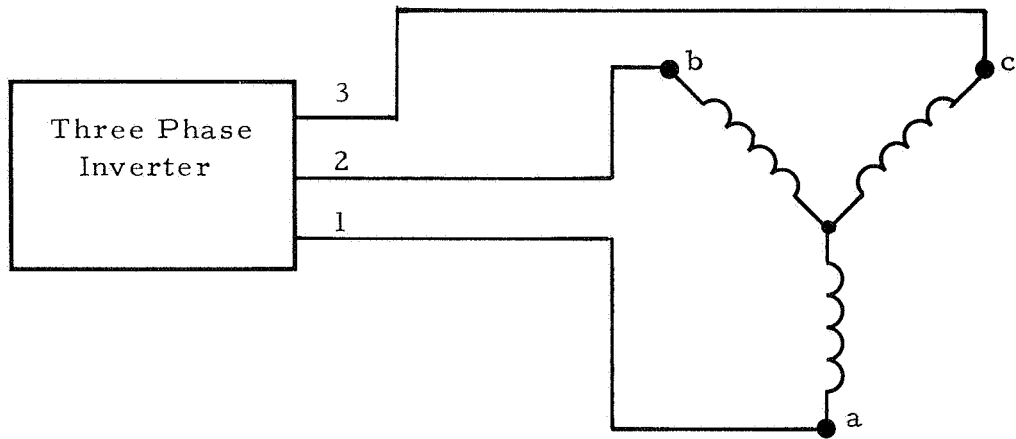
The three possible connections are illustrated in Figure 2.11. We shall refer to these as (a) star connection with neutral; (b) star connection without neutral; and (c) delta connection. The line to neutral output waveforms of the square wave three phase inverter are shown in Figure 2.12. Figure 2.13 shows line to line waveforms for the delta connection and for the star connection without neutral. The relevant waveforms for the star connection with neutral are the same as the line to neutral waveforms for the inverter.

Figure 2.14 shows the spectra corresponding to the various waveforms. In each case, the amplitude of the harmonic is referred to the amplitude of the fundamental. It should be noted that, despite the different waveforms, the spectra are identical except for the harmonics of order $n = 3k + 3$ where $k = 0, 1, 2, \dots$. The three phase connection results in harmonics of the order $3k + 3$ which are in phase when measured line to line. Therefore, with the neutral not connected, these components are balanced out and no current flows. If the neutral is connected, large currents will flow since the magnetic fields in the motor cancel and the motor impedance becomes nearly that of the stator resistance.

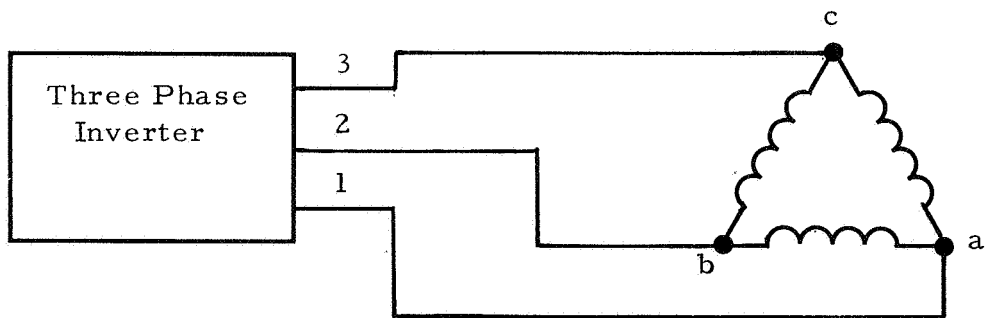
In general, the magnetic fields in the motor corresponding to a harmonic of order n the inverter output behave as follows. For those harmonics of order $n = 3k + 1$, where $k = 0, 1, 2, \dots$ the magnetic flux density rotates in the same direction. The phase velocity of this flux density is directly proportional to the order of the harmonic. For those harmonics of order $n = 3k + 2$, the magnetic flux density rotates in the opposite direction from those of order $n = 3k + 1$. These are referred to as counter-rotating fields. For those harmonics of order $n = 3k + 3$, the magnetic flux density does not rotate, leading to the effects described above.



a) Star Connection with Neutral



b) Star Connection without Neutral



c) Delta Connection

FIGURE 2.11

MOTOR CONNECTIONS TO INVERTER

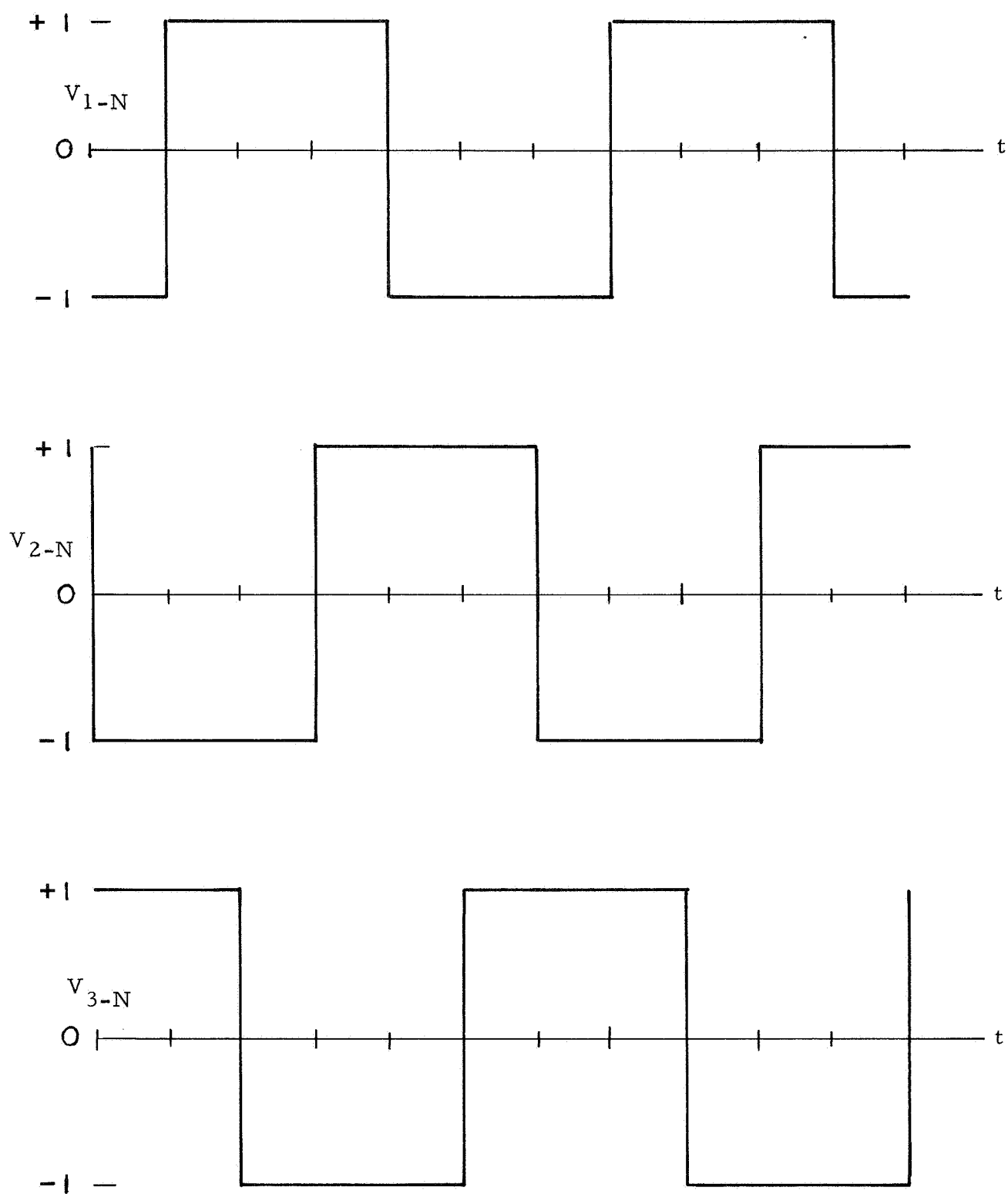


FIGURE 2.12

LINE TO NEUTRAL OUTPUT WAVEFORMS OF SQUARE WAVE THREE
PHASE INVERTER

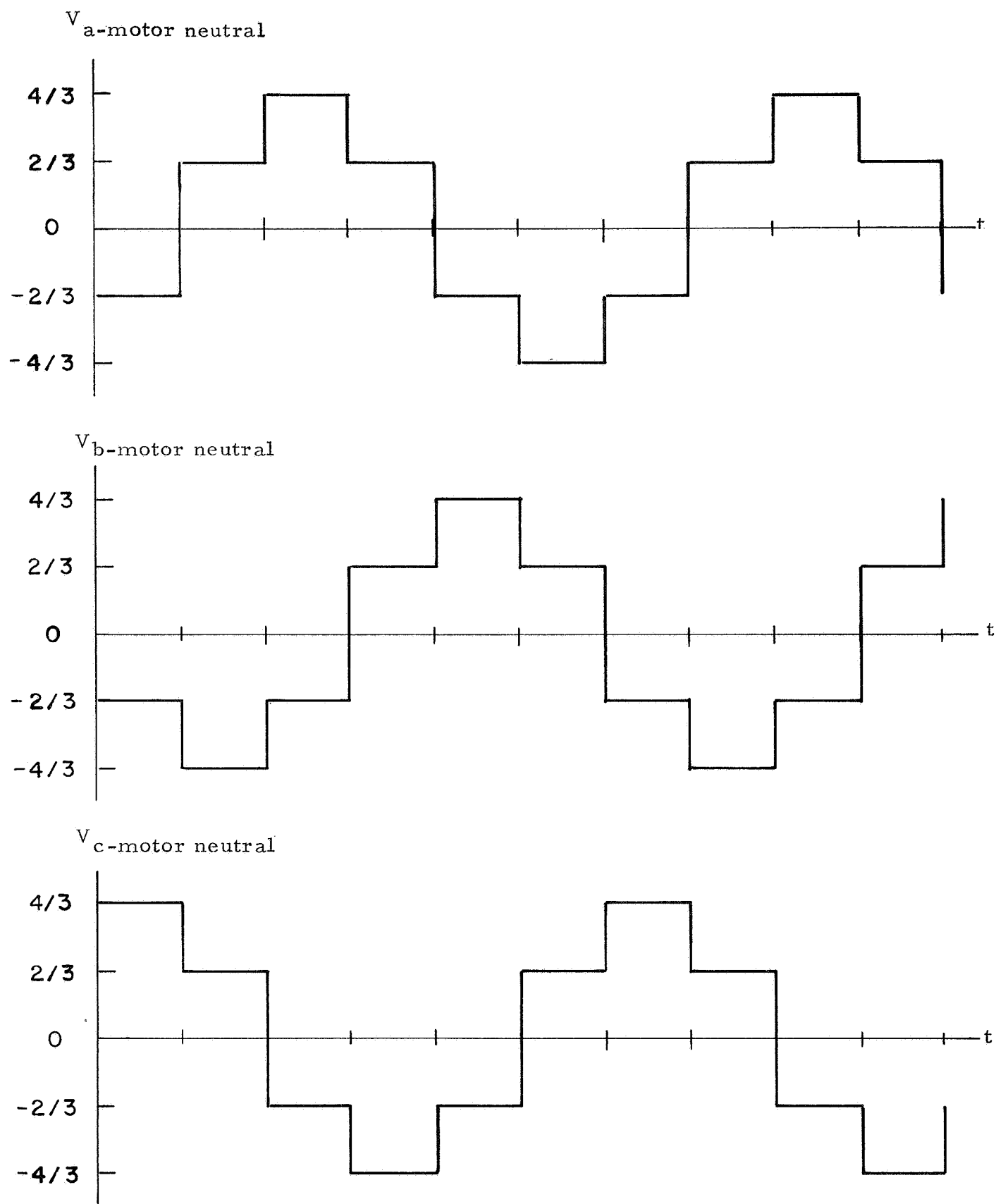


FIGURE 2.13a

LINE TO MOTOR NEUTRAL WAVEFORMS:
STAR CONNECTION WITHOUT NEUTRAL

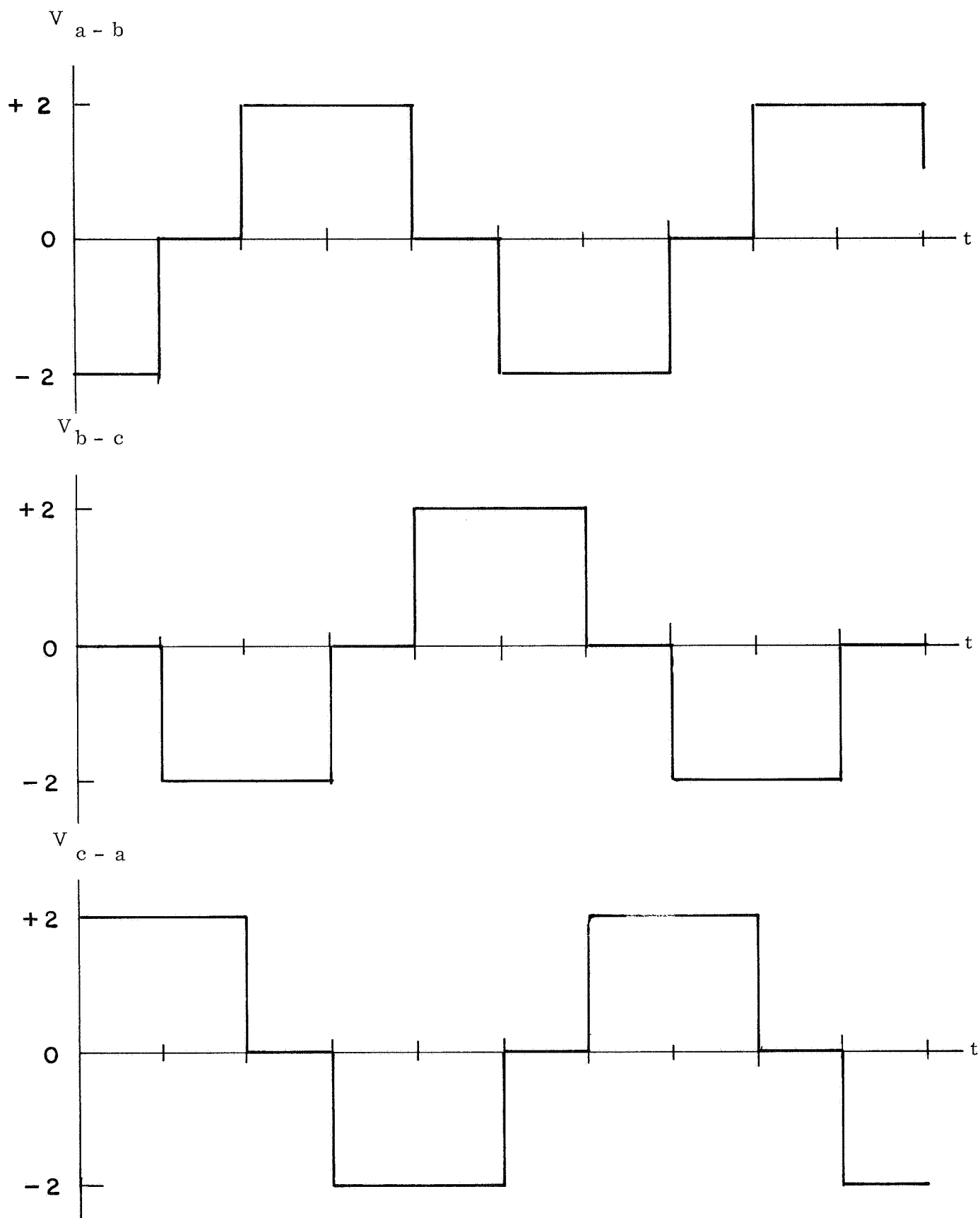


FIGURE 2.13b

LINE TO LINE WAVEFORMS: DELTA CONNECTION

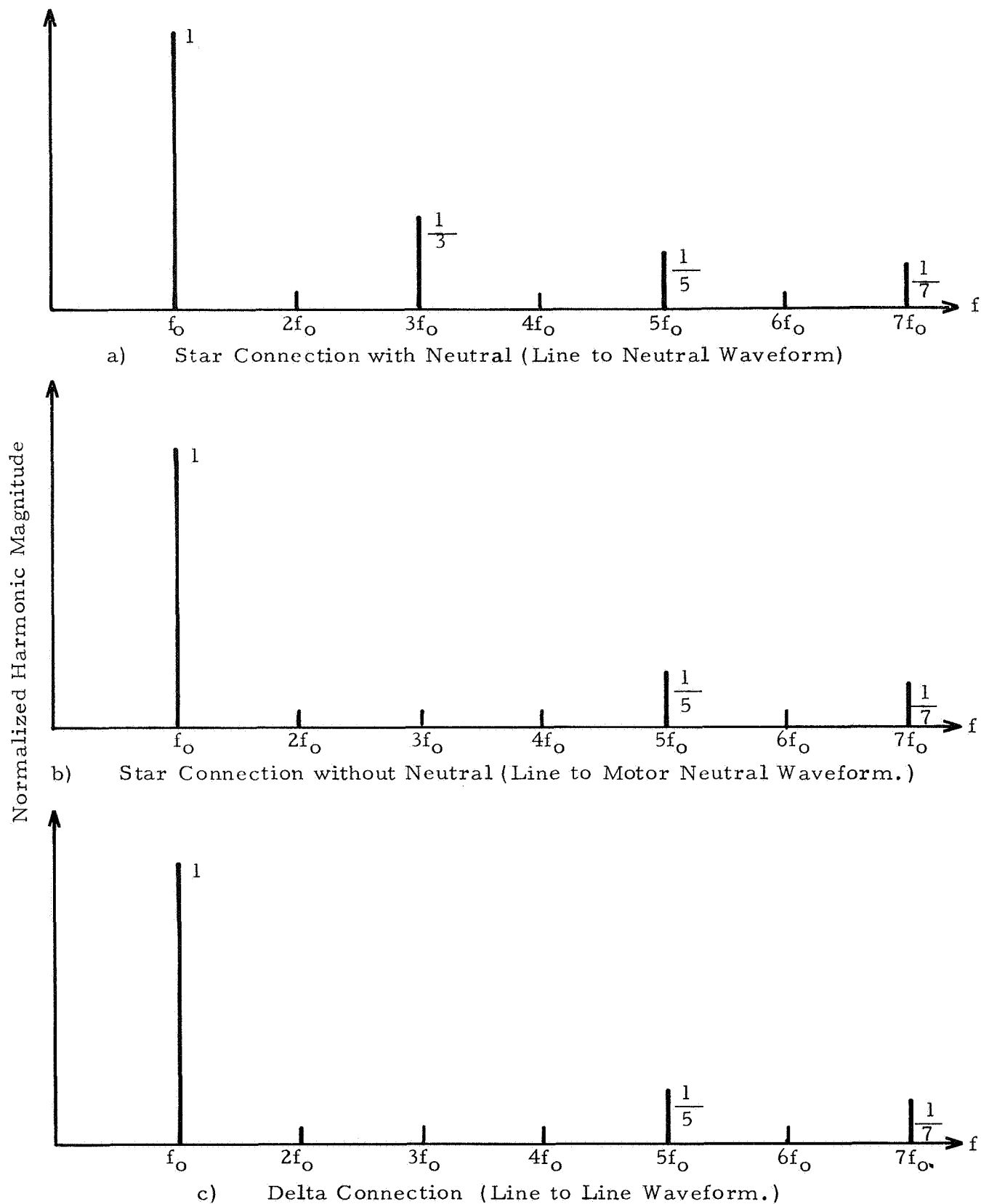


FIGURE 2.14

COMPONENT AMPLITUDES RELATIVE TO FUNDAMENTAL IN
FREQUENCY SPECTRA FOR WAVEFORMS OF FIGURES 2.12 AND 2.13

Spurious modulation frequencies will not rotate synchronously with the fundamental but may be handled by techniques similar to those used to handle harmonics.

2.4 INDUCTION MOTOR IMPEDANCE CHARACTERISTICS

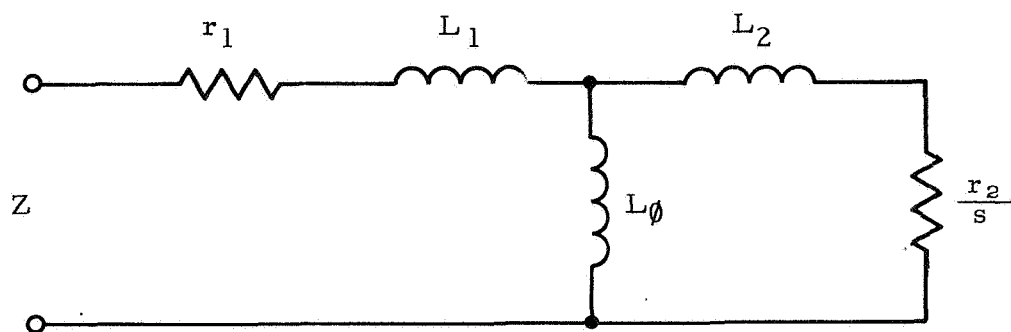
The nature of the load impedance the motor presents to the inverter has implications for the inverter design, particularly in regard to the peak voltages and currents that the inverter must supply for a given state of the motor. We need to characterize the motor impedance for all drive frequencies and for all speeds, thus including all the modes of operation: motoring, dissipative braking and regenerative braking.

The following information is desired from our analysis:

- 1) Limits on the magnitude of driving point impedance for all states of the motor and of the drive.
- 2) Limits on the angle of driving point impedance for all states of the motor and of the drive.
- 3) The existence of anomalies in the impedance (such as a condition where the harmonic frequencies see impedances lower than the fundamental).
- 4) The change in impedance characteristics with different motor designs.

The equivalent circuit, per phase, of the induction motor was presented in Figure 2.1 and is reproduced in Figure 2.15 for reference, with the exception of the conductance representing magnetic losses, which is neglected.

The meaningful presentation of the impedance information must take into consideration that the inductive elements have an impedance that varies directly with frequency whereas the resistance r_2/s varies both with the drive frequency and with motor speed. Accordingly, we assume that at a given instant of time the state of the motor is characterized by a speed ω_m . Since ω_m cannot change instantaneously (due to rotor inertia)



r_1 = Resistance of Stator Winding

r_2 = Resistance of Rotor Winding As Seen from the Stator

L_1 = Leakage Inductance of Stator Winding

L_2 = Leakage Inductance of Rotor Winding As Seen from the Stator

L_ϕ = Magnetizing Inductance

s = Slip = $\frac{\omega_s - \omega_m}{\omega_s}$

Where ω_s = Line Frequency

ω_m = Mechanical Angular Velocity of Rotor

FIGURE 2.15

PER PHASE EQUIVALENT CIRCUIT OF INDUCTION MOTOR
(Model valid for two pole motor.)

this allows us to present the driving point impedance as a function of line frequency, ω_s , with ω_m as a parameter.

The motor chosen for analysis has the following parameters:

$$r_1 = 0.3\Omega$$

$$L_1 = 0.14 \text{ mH}$$

$$r_2 = 0.14\Omega$$

$$L_2 = 0.6 \text{ mH}$$

$$L_\phi = 3.8 \text{ mH}$$

The motor has six poles and is designed for three phase operation at 220 volts at a line frequency of 60 Hz. Synchronous operating speed is 125.6 radians/second.

The results of the calculations are presented in Figures 2.16 and 2.17. An examination of the figures shows two limiting cases. For $\omega_s < 1.0$, the reactances are small and the impedance is mainly that of the stator resistance. For $\omega_s > 300$, the reactances dominate and the impedance is mainly that of a pure inductance.

For the intermediate frequencies, the impedance is a function of the operating mode of the machine. That is, the machine may be either motoring ($\omega_m < \omega_s$) or braking ($\omega_s < \omega_m$). The braking may involve regeneration of power if ω_s is high enough such that $\frac{r_2}{s}$ dominates the impedance. All the interesting behavior takes place at $\omega_m \approx \omega_s$, that is, relatively near synchronous operating speed.

For the case of very slow speed ($\omega_m = 1$), the anomalous behavior of $\frac{r_2}{s}$ takes place at a line frequency so low that the magnetizing reactance almost completely shunts $\frac{r_2}{s}$. Thus the impedance departs only slightly

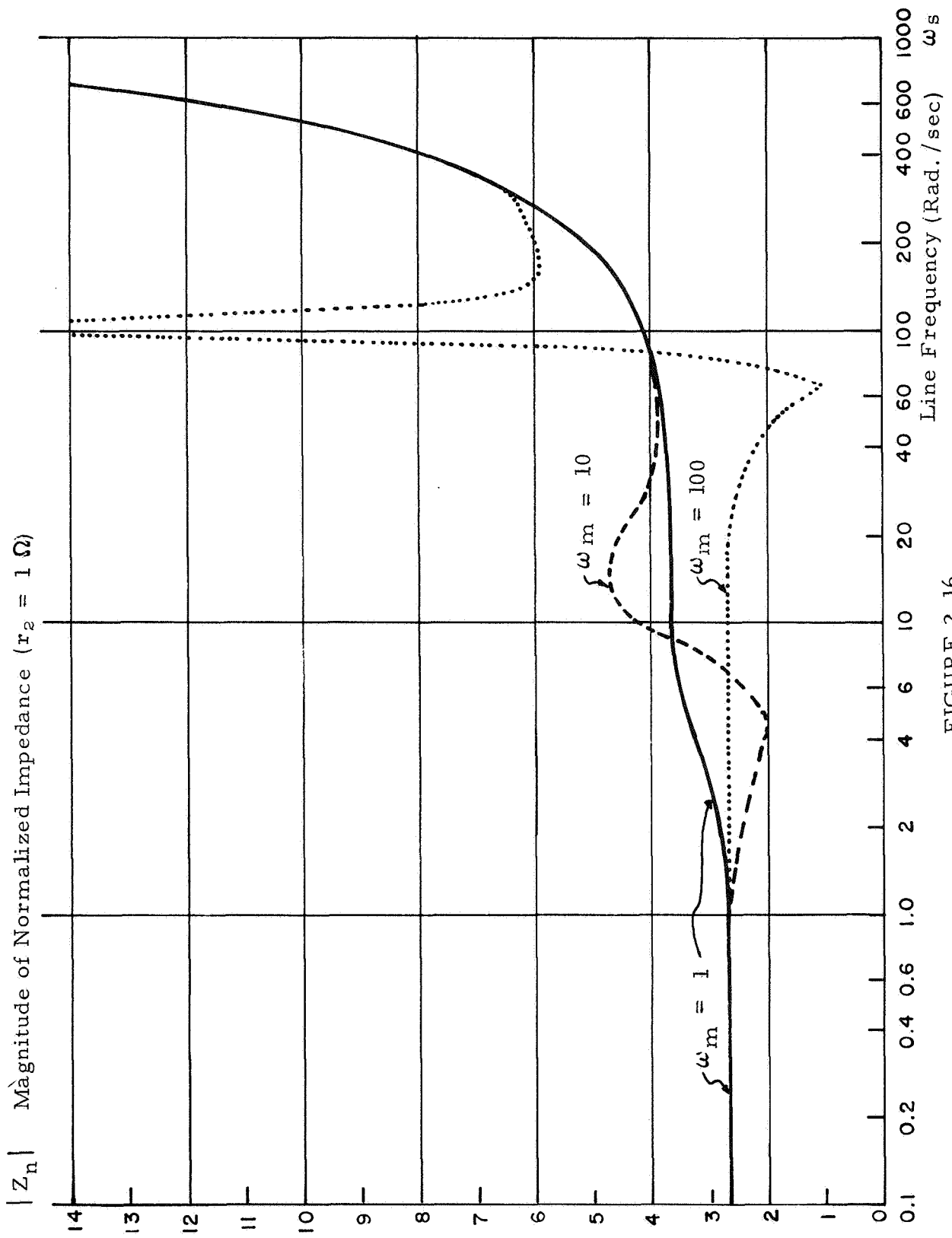


FIGURE 2.16

MAGNITUDE OF NORMALIZED MOTOR IMPEDANCE vs ω_s WITH ω_m AS A PARAMETER

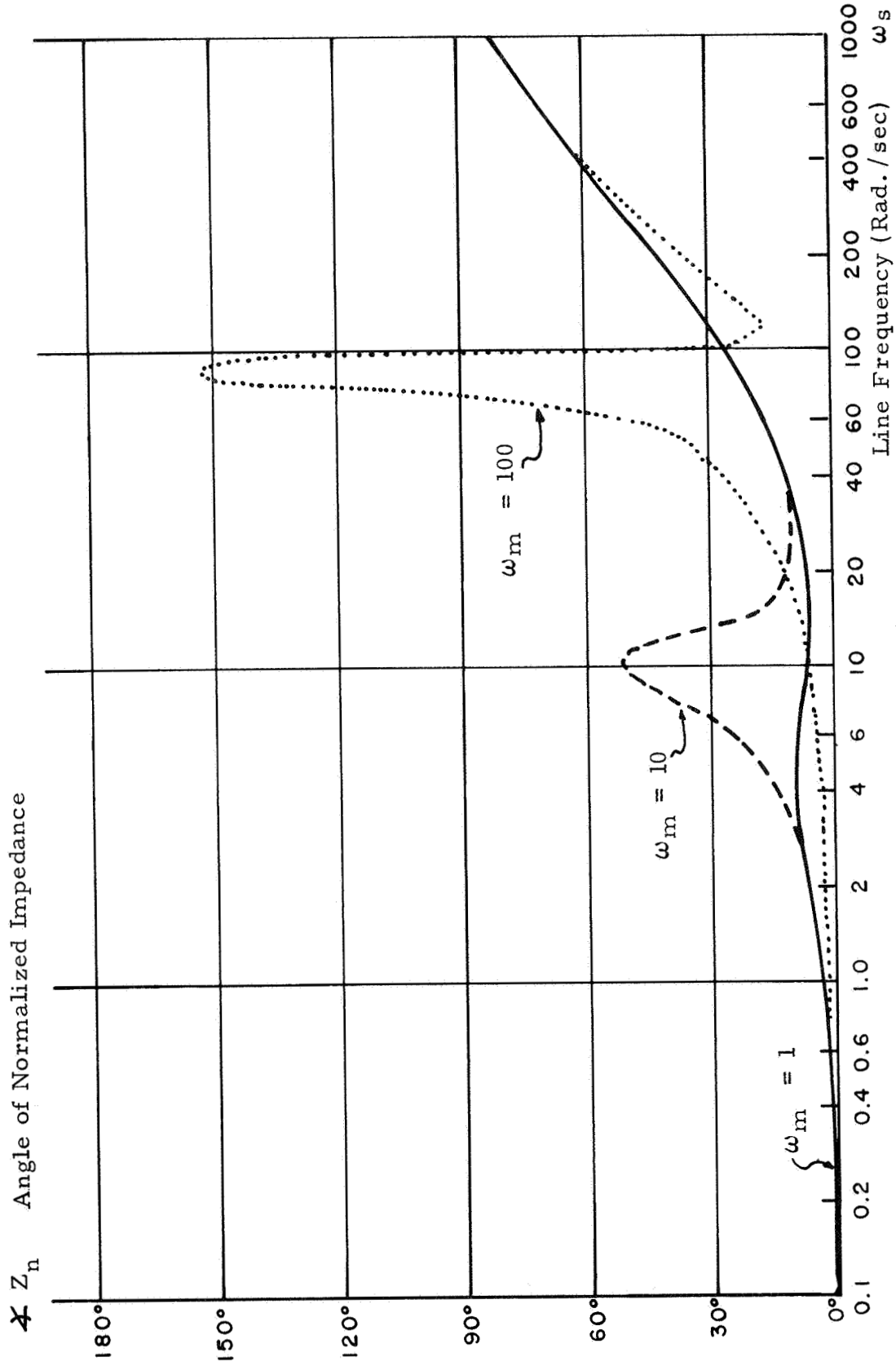


FIGURE 2.17

ANGLE OF NORMALIZED MOTOR IMPEDANCE vs ω_s WITH ω_m AS A PARAMETER

from that of a conventional R-L circuit.

For $\omega_m = 100$, however, the reactance of L_0 is quite large for $\omega_s \approx 100$, and r_2/s dominates the impedance. The three modes of operation of the motor are clearly evident. For $\omega_s < 60$, r_2/s is negative but small in magnitude. This corresponds to braking action although the real part of the impedance at the motor terminals is positive. At $65 < \omega_s < 100$, the motor impedance has a negative real part and power is regenerated. Figure 2.16 shows the change from braking to regeneration as a discontinuity in the derivative of the magnitude at $\omega_s \approx 65$. In Figure 2.17, the angle of the impedance exceeds 90° for $\omega_s > 70$ again indicating regeneration. For $\omega_s > 100$, motoring again takes place.

It is evident that under certain conditions, the impedance can be lower for a harmonic of the drive frequency than for the fundamental. For example, with $\omega_r = 100$, $\omega_s \approx 20$ (rapid braking) the impedance at approximately $5\omega_s$ will go through a minimum.

Many of the effects of change of motor design such as changing r_1 or L_1 are self-evident. We note that reducing the parameter $\frac{r_2}{L_2}$ (as desired for servo applications) makes the anomalous behavior of r_2/s evident only when ω_m is more nearly equal to ω_s , thus sharpening the variations in the curves. The magnitude of the variations remains unchanged since at $\omega_r = \omega_m$, r_2/s approaches infinity.

From this data, we may conclude the following:

- 1) The magnitude of the driving point impedance varies over a range of at least 20:1.
- 2) The angle of the driving point impedance varies from 0° to a value approaching 180° .
- 3) The impedance is a strong function of the state (speed) of the motor.

4) Anomalies such as low impedance at harmonic frequencies are evident.

Although this presentation characterizes the motor impedance completely, the specification of voltage and current requirements on the inverter depends on other parameters also. These include:

- 1) The particular motor chosen.
- 2) The mission requirements (i. e. required torque versus speed).
- 3) The control scheme chosen.

The last item needs some explanation. For a given torque and speed demand, the control scheme can choose to adjust either the drive frequency or the voltage or both. For example, the slip frequency might be set equal to a constant (constant slip operation) thus eliminating operation very near synchronism.

The final results will depend on evaluation of the total system including the motor, inverter and control scheme.

2.5 MOTOR TORQUES WITH NON-SINUSOIDAL EXCITATION

Many of the factors that most directly increase the cost and complexity of the inverter are those that demand minimization of harmonic frequencies in the output spectrum. Accordingly, it is necessary to know the penalty paid in performance for harmonic components in the inverter output. In this section, we examine the conditions under which these harmonics can produce spurious torques (both steady and pulsating) in the motor.

The Field Approach To Torque Generation In The Induction Motor

Torque is generated in the induction motor by the interaction of a magnetic field and a current according to the equation

$$d\bar{T} = \bar{R} \times i d\bar{s} \times \bar{B} \quad (2.19)$$

where $d\bar{T}$ is a vector element of torque, $\bar{R}$ is the moment arm vector, $id\bar{s}$ is a current element vector, and $\bar{B}$ is the magnetic field vector.

In the analysis of the AC induction motor, we can consider the total magnetic flux in the air gap as due to the terminal voltage E_s (neglecting drops due to the stator leakage reactance and the stator resistance). The flux density is radially directed in the air gap and (for a three phase motor) has the time and spatial dependence of a traveling wave. Thus, for a sinusoidal excitation at a frequency ω_s , the flux density in the air gap for a two pole motor is given by

$$\bar{B}_s = \bar{i}_r B_s \cos (\omega_s t - \psi) \quad (2.20)$$

where $\bar{i}_r$ is the radial unit vector, and ψ is the angular position in stator coordinates. For harmonics of the drive frequency, ω_s , the spatial distribution of flux density in the air gap is the same but the phase velocity is different. For example, the fifth harmonic in a square wave applied to a three phase motor would rotate in a direction opposite to that of the fundamental. This component of the field would be expressed

$$\bar{B}_{s5} = \bar{i}_r B_{s5} \cos (-5\omega_s t - \psi). \quad (2.21)$$

In general, the n th harmonic of the fundamental will produce a field

$$\bar{B}_{s_n} = \bar{i}_r B_{s_n} \cos (\pm n\omega_s t - \psi). \quad (2.22)$$

We note that in the usual three phase connection we can indicate the behavior of the field by a convention involving n . In particular, for harmonics of order $n = 3k + 1$, $k = 0, 1, 2, \dots$, the positive sign is taken in Eqn. 2.22 and indicates a positive direction of field rotation (or phase velocity). For harmonics of order $n = 3k + 2$, the negative sign is taken in Eqn. 2.22 and indicates a negative rotation or a counter-rotating field. Harmonics of order $n = 3k + 3$ do not rotate and, in fact, the resultant field component is zero.

Each component of the flux density induces a voltage in the rotor

resulting in a rotor current that depends on the rotor impedance. The frequency of the current in the rotor is $\omega_r = n\omega_s - \omega_m$ where n is the order of the harmonic and ω_m is the mechanical speed of the rotor. The current induced in the rotor interacts with the flux density in the gap according to Eqn. 2.19 to produce torque. Since the rotor impedance is inductive, the current in the rotor will lag in time behind the exciting flux density. This lag keeps the rotor magnetic field component from exactly aligning itself with the flux density, thus allowing torque to be generated continuously.

Each harmonic of the motor drive signal will lead to a component of flux density and to a component of rotor current. Any component of flux density can interact with any component of rotor current to produce torque.

In general it can be shown that the magnitude of torque due to interaction of the n th component of flux density with the n th component of rotor current gives a torque

$$T_{n,n} = \frac{|E_{s_n}|^2}{n^2 \omega_s^2 L_2} \frac{\tau [n\omega_s - \omega_m]}{[1 + \tau^2 (n\omega_s - \omega_m)^2]} \quad (2.23)$$

where $|E_{s_n}|$ is the magnitude of the n th component of the drive voltage (neglecting drops due to the stator resistance, r_1 , and the stator leakage reactance L_1), ω_s is the fundamental drive frequency, r_2 is the rotor resistance and τ is the rotor electrical time constant, L_2/r_2 . The torques produced by the interaction of the n th component of flux density with the corresponding (or n th) component of rotor current are steady torques and are here referred to as harmonic torques.

In general the n th component of flux density will also interact with the m th component of rotor current and the m th component of flux density will interact with the n th component of rotor current to produce a sinusoidally time varying torque. We will refer to these torques as pulsating torques. The frequency of the time variation is $(n - m) \omega_s$. The magnitude of the pulsating torque of frequency $(n - m) \omega_s$ is given by

(2.24)

$$|T_{m,n}| = \frac{|E_{s_n}| |E_{s_m}|}{nm\omega_s^2 L_2} \frac{\left[(n\omega_s - \omega_m) + (m\omega_s - \omega_m) \right] \tau}{\left[1 + \tau^2 (n\omega_s - \omega_m)^2 \right]^{\frac{1}{2}} \left[1 + \tau^2 (m\omega_s - \omega_m)^2 \right]^{\frac{1}{2}}}$$

where $|E_{s_n}|$, $|E_{s_m}|$ are the magnitudes of the n th and m th components of the stator voltage (again neglecting the drops due to stator resistance and stator leakage reactance).

In general, we need to determine the magnitude of the spurious torques in question (both harmonic and pulsating) relative to the magnitude of the fundamental component of torque. These spurious torques can become significant when the motor is operating very near synchronous speed where the fundamental component of torque is low.

Evaluation Of The Magnitude Of Spurious Torques

We will proceed to show that for constant flux the magnitude of the harmonic torque depends only on the frequency seen in the rotor. Low speed operation emphasizes harmonic torques because the harmonics of ω_s do not create rotor frequencies sufficiently different from the fundamental rotor frequency. Also, we will examine the magnitude of the fluxes expected for the various harmonics under various conditions. Finally, we shall examine the magnitude of pulsating torques as a first step in determining the effect of these torques on shaft position for various loads on the motor.

Harmonic Torques

The magnitude of the harmonic torque is given by Eqn. 2.23, reproduced here.

$$T_{n,n} = \frac{|E_{s_n}|^2}{n^2 \omega_s^2 L_2} \frac{\tau [n\omega_s - \omega_m]}{\left[1 + \tau^2 (n\omega_s - \omega_m)^2 \right]} \quad (2.23)$$

For the purposes of this discussion, we will write Eqn. 2.23 in the form

$$T_{n,n} = T_{\max} F(\beta_n) \quad (2.25)$$

where

$$T_{\max} = \frac{|E_{s_n}|^2}{2(n\omega_s)^2 L_2} \quad (2.26)$$

and

$$F(\beta_n) = \frac{2\beta_n}{1 + \beta_n^2} . \quad (2.27)$$

The quantity β_n is defined as

$$\beta_n = \tau \omega_{r_n} = \frac{L_2}{r_2} (n\omega_s - \omega_m) . \quad (2.28)$$

The maximum value of $F(\beta_n)$ is 1.0 which occurs at $\beta_n = 1$ corresponding to $\omega_{r_n} = 1/\tau$. Thus, $T_{\max}$ is the maximum torque the motor can produce for a given flux. (The maximum flux is, of course, limited by saturation of the iron in the motor.) Figure 2.18 is a plot of $F(\beta_n)$ versus β_n . This is exactly the same curve as that of Figure 2.3. Note that normal motor operation occurs for $\beta_n < 1$. Breakover torque is at $\beta_n = 1$. The effect of harmonic torques can easily be assessed from this graph.

If the fundamental drive frequency causes an operating point in the slip limited operating region of the motor, $|\beta_1| < 1$ then harmonic torques can be minimized if the harmonics fall such that $|\beta_n| \gg 1$. This requires that $\omega_{r_n} = |n\omega_s - \omega_m| \gg \frac{1}{\tau}$ where n is an integer that is neither 0 nor ± 1 . This condition is more easily satisfied if τ is large as is illustrated by Figure 2.19. The larger value of τ narrows the range of ω_{r_n} over which large torques are possible. Additionally, we can require that the harmonic frequencies fall at large values of ω_{r_n} , i. e. such that $|\omega_{r_n}| \gg \frac{1}{\tau}$. Since normal motoring operation results in $\omega_{r_1} < \frac{1}{\tau}$, when ω_m is small several harmonics may result in $\omega_{r_n} \approx \frac{1}{\tau}$.

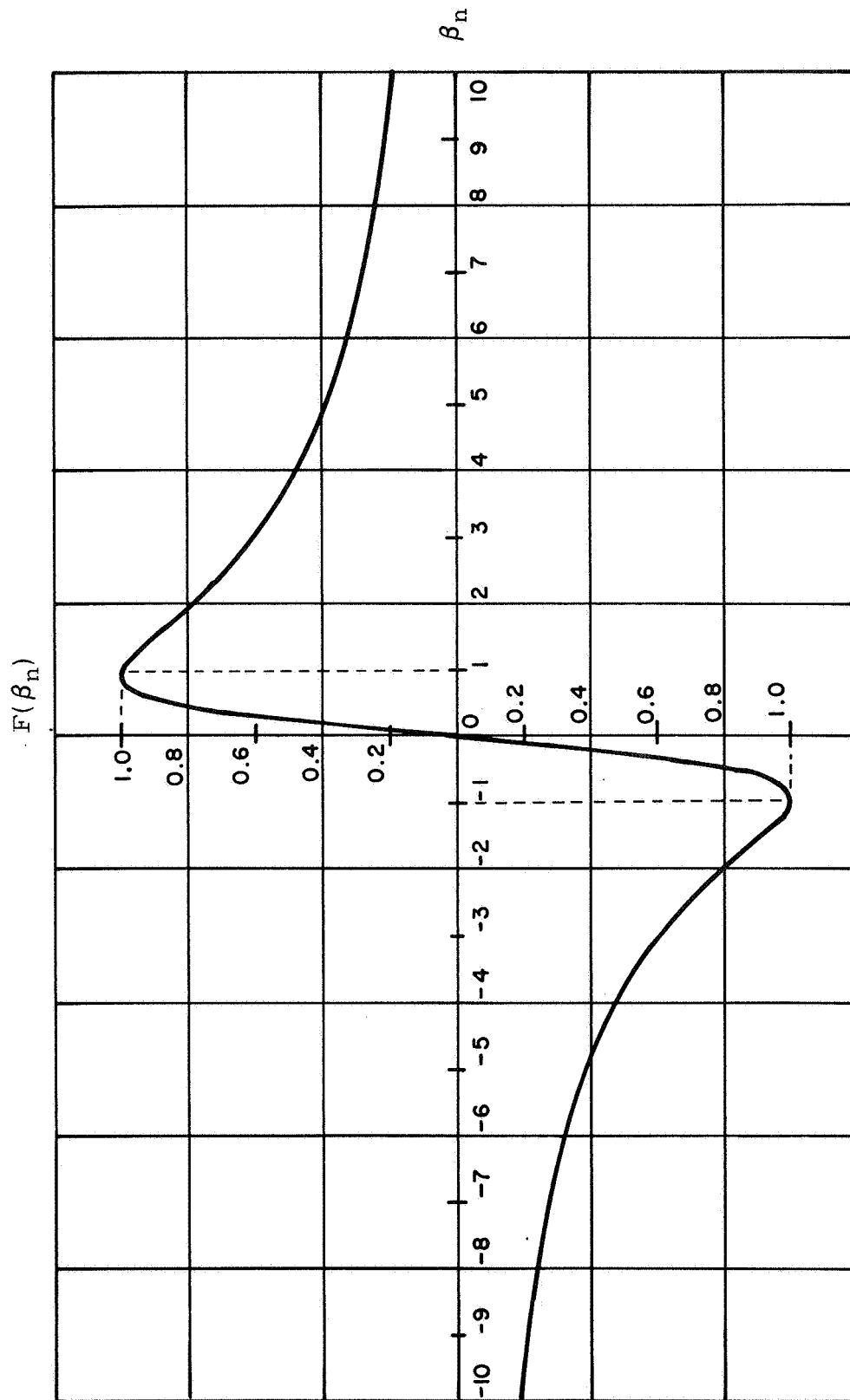


FIGURE 2.18

GRAPH OF $F(\beta_n)$ vs β_n (NORMALIZED TORQUE - FREQUENCY CURVE)

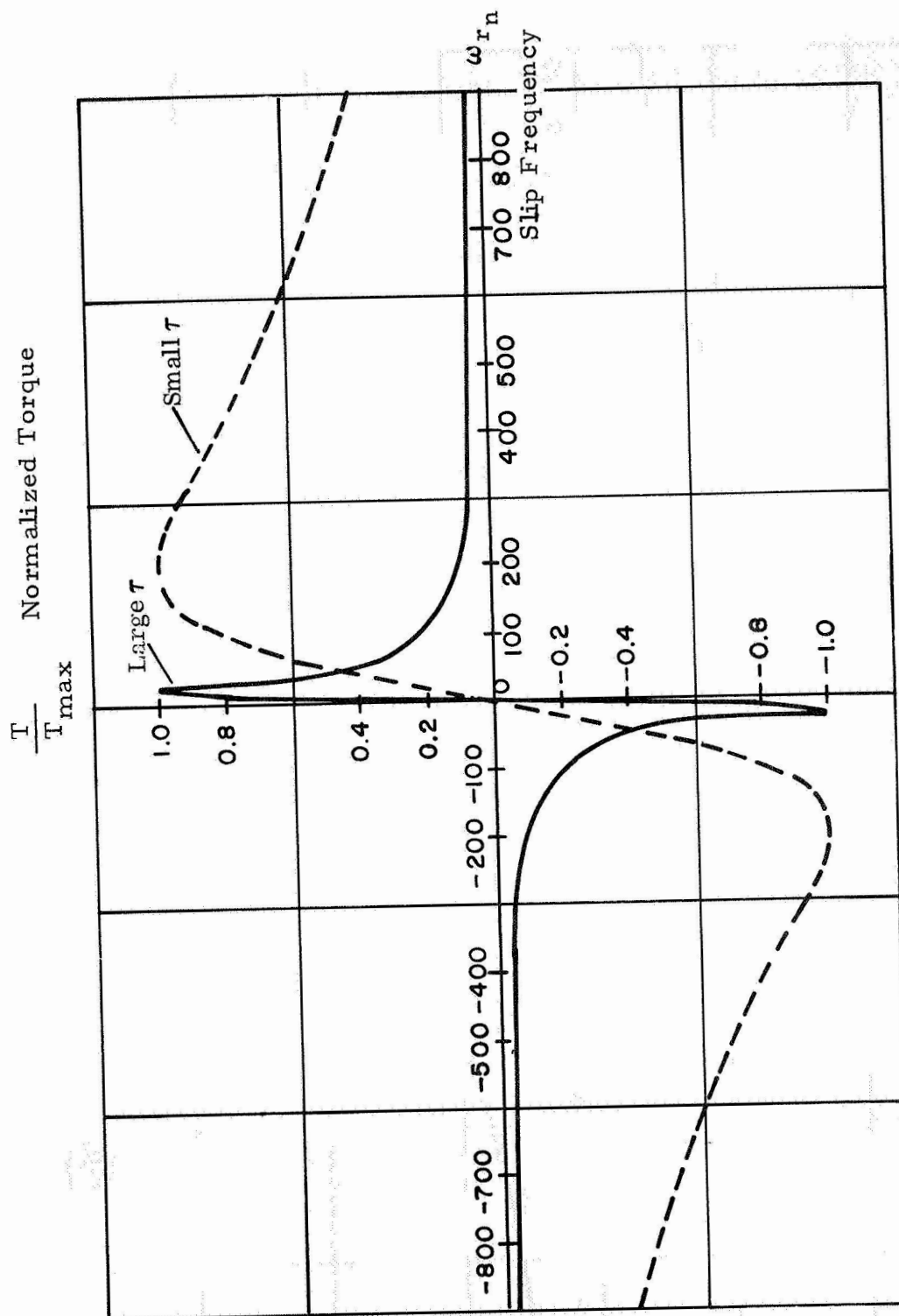


FIGURE 2.19
NORMALIZED TORQUE FREQUENCY CURVE FOR TWO VALUES OF ROTOR
TIME CONSTANT

Figures 2.20 and 2.21 illustrate this point. Both are plots of $\frac{T_{n,n}}{T_{\max}}$ for the same rotor ($\tau = 0.1$) but for different values of ω_m . In Figure 2.21 with $\omega_m = 1$ and a fundamental torque 20% of maximum torque, the fifth harmonic (counter rotating) torque is about 100% of maximum torque (for the same motor flux). Figure 2.21 shows the same case for $\omega_m = 100$. Here $|-5\omega_s - \omega_m| \gg \frac{1}{\tau}$ thus minimizing the harmonic torque.

However, in general the flux in the motor is less at harmonic frequencies. For a square wave inverter, the magnitude of E_{s_n} is proportional to $1/n$. Thus $|E_{s_n}|^2/(n\omega_s)^2$ is proportional to $1/n^4$. At sufficiently low frequencies however, the stator resistance dominates the motor impedance causing $L\theta$ (see Figure 2.16) to differentiate the input voltage. Thus, E_{s_n} would be proportional to n times the actual terminal voltage and $|E_{s_n}|^2/(n\omega_s)^2$ would be proportional to $1/n^2$. Using our previous example where $F(\beta_{-5}) \approx 1.0$, $F(\beta_1) = 0.2$, $1/n^2 = 1/25$, we find that the fifth harmonic torque is approximately $1/5$ the fundamental torque.

The worst case would occur as ω_{r1} , the fundamental slip frequency, passed through zero (this would correspond to decelerating the motor by frequency variation) thus causing the fundamental torque to approach zero. This case could be avoided by controlled slip operation.

Also, some pulse modulating schemes have higher harmonic content than the square wave and thus would be undesirable for low frequency operation.

Pulsating Torques

Eqn. 2.24 is an expression for the magnitude of the m, n interaction torque. If we let $\beta_n = \tau(n\omega_s - \omega_m)$, $\alpha_m = \tau(m\omega_s - \omega_m)$ then Eqn. 2.24 can be written

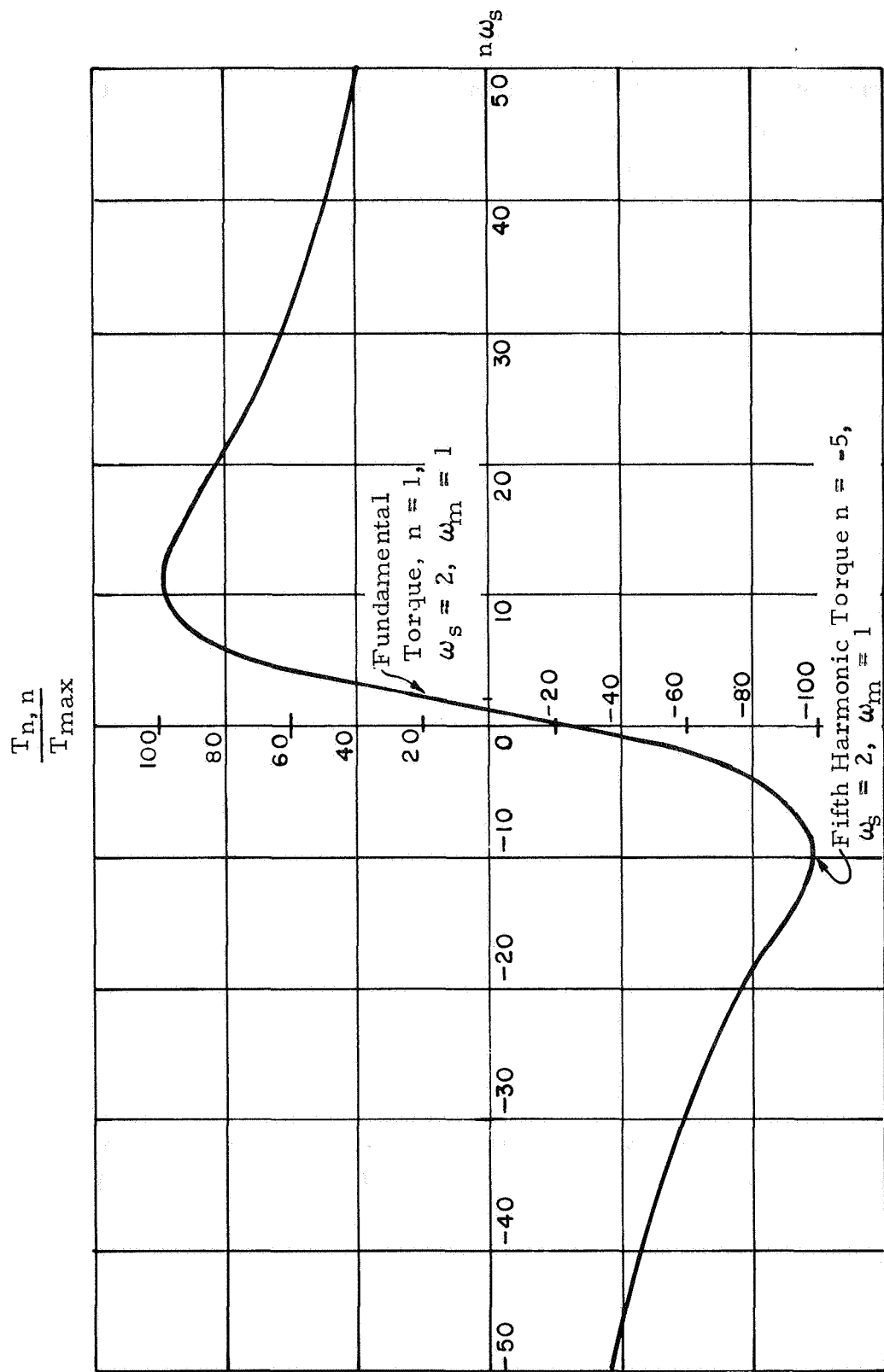


FIGURE 2.20

PLOT OF $\frac{T_{n,n}}{T_{max}}$ VS $n\omega_s$ AT SLOW MOTOR SPEED SHOWING FUNDAMENTAL TORQUE AND
TORQUE DUE TO FIFTH HARMONIC

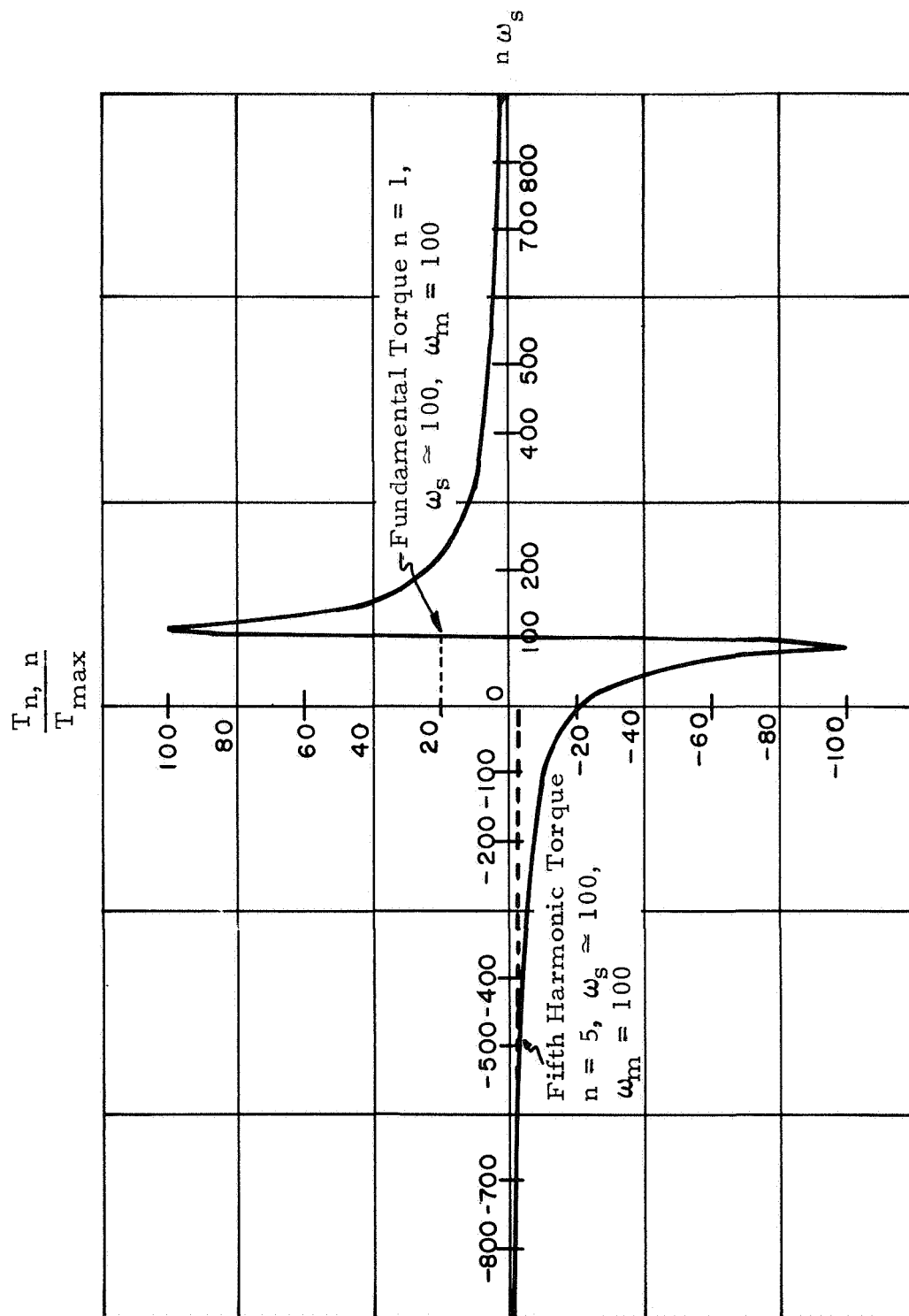


FIGURE 2.21

PLOT OF $\frac{T_{n,n}}{T_{max}}$ VS $n \omega_s$ AT HIGH MOTOR SPEED SHOWING FUNDAMENTAL TORQUE AND TORQUE DUE TO FIFTH HARMONIC

$$|T_{m, n}| = \frac{|E_{S_m}| |E_{S_n}|}{nm \omega_s^2 L_2} \frac{(\alpha_m + \beta_n)}{[1 + \alpha_m^2]^{\frac{1}{2}} [1 + \beta_n^2]^{\frac{1}{2}}} = T'_{\max} F(\alpha_m, \beta_n) \quad (2.29)$$

where

$$|T'_{\max}| = \frac{|E_{S_n}| |E_{S_m}|}{nm \omega_s^2 L_2} \quad (2.30)$$

and

$$F(\alpha_m, \beta_n) = \frac{(\alpha_m + \beta_n)}{[1 + \alpha_m^2]^{\frac{1}{2}} [1 + \beta_n^2]^{\frac{1}{2}}} \quad (2.31)$$

The quantity $F(\alpha_m, \beta_n)$ has a maximum value of 1 at $\alpha_m = \frac{1}{\beta_n}$. Thus, $T'_{\max}$ represents the maximum magnitude of pulsating torque.

If the relationship $\alpha_m = \frac{1}{\beta_n}$ can be satisfied, then a worst case condition for the generation of pulsating torque exists. This relationship requires that

$$(n\omega_s - \omega_m)(m\omega_s - \omega_m) = \frac{1}{\tau^2} \quad (2.32)$$

The solution for ω_s is

$$\omega_s = \frac{1}{2} \frac{(n+m)}{nm} \omega_m \pm \frac{1}{2 nm \tau} \left[\left(\tau \omega_m (n-m) \right)^2 + 4 nm \right]^{\frac{1}{2}} \quad (2.33)$$

If n and m have different signs, the solution does not always exist for small ω_m .

Since the magnitude of the pulsating torque is proportional to the product of the two flux components, then the largest pulsating torque will occur for the largest harmonic component of lowest frequency interacting with the fundamental component of flux. Therefore, for our evaluation of the magnitude of pulsating torque, we will choose $n = 1$, and relate the magnitude of the $T_{m,1}$ component of continuous torque. This relationship is given by

$$\frac{|T_{m,1}|}{T_{1,1}} = \frac{|E_{sm}|}{|E_{s1}|_m} \frac{(\alpha_m + \beta_1)(1 + \beta_1^2)^{\frac{1}{2}}}{2\beta_1(1 + \alpha_m^2)^{\frac{1}{2}}} \quad (2.34)$$

where

$$\alpha_m = \tau(m\omega_s - \omega_m) \quad (2.35)$$

and

$$\beta_1 = \tau(\omega_s - \omega_m). \quad (2.36)$$

The worst case condition for Eqn. 2.34 with respect to the variation of α_m (assuming a solution exists for which $\alpha_m = \frac{1}{\beta_n}$) is

$$\frac{|T_{m,1}|}{T_{1,1}} = \frac{|E_{sm}|}{|E_{s1}|_m} \frac{(1 + \beta_1^2)^{\frac{1}{2}}}{2\beta_1} \quad (2.37)$$

We can use Eqn. 2.34 to evaluate the relative magnitude of pulsating torque for the example used in the previous section on harmonic torque. For this example, the worst case solution does not exist, i.e.

there is no value of ω_s for which $\alpha_m = \frac{1}{\beta_n}$. With $m = -5$, $\tau = 0.1$,

$\omega_m = 1$ and $\omega_s = 2$ we find that

$$\frac{|T_{-5,1}|}{T_{1,1}} = \frac{|E_{s-5}|}{|E_{s1}|} \times 0.69. \quad (2.38)$$

For a square wave, $\frac{|E_{s-5}|}{|E_{s1}|} = 1/5$ and if we allow for the filtering effect

of the stator resistance and the magnetizing reactance, we find that effectively $\frac{|E_{s-5}|}{|E_{s1}|} = 1$. Thus,

$$\frac{|T_{-5,1}|}{T_{1,1}} \approx 0.69. \quad (2.39)$$

Although the condition $\alpha_m = \frac{1}{\beta_1}$ represents a worst case, the function $F(\alpha_m, \beta_n)$ has only a relatively small effect on the ratio $\frac{|T_{m,1}|}{T_{1,1}}$.

The absolute worst case occurs when $T_{1,1}$ goes to zero (i e. when the fundamental drive frequency is the synchronous frequency) just as with harmonic torque.

Accurate evaluation of the problem requires solving the dynamic motion problem for a particular shaft load (including rotor inertia, stray losses, load inertia and load friction) to determine the actual effect of the pulsating torques on shaft motion.

3. INVERTER SYSTEMS FOR VARIABLE SPEED INDUCTION MOTOR DRIVES

The design of a particular piece of power processing equipment is composed of a cascade of one or more basic power converters; that is, a cascade of circuits that perform one of the four electrical conversion functions:

- 1) AC-AC (frequency changing, amplitude scaling or regulation)
- 2) AC-DC (rectification)
- 3) DC-DC (voltage regulation, DC transformation)
- 4) DC-AC (inversion)

A survey of the available literature indicates that the study of variable voltage, variable frequency inverter systems, operating from an unconditioned DC source and suitable for use in variable speed induction motor (VSIM) drive applications, may be divided into three basic power conversion approaches shown in Table 3.1. For each approach, there are a multitude of different power converters which may be used to accomplish each step of the conversion. However, there are two fundamental considerations in the selection and evaluation of each converter; these are the selection of an output stage configuration and the selection of a control concept (or modulation system) to accomplish the desired conversion. Because the power switching device constraints limit the overall power output capability of a converter system, it is profitable to examine how the characteristics of the output stage circuitry and the load meet the device constraints, divorced from any particular system. On the other hand, the control concept often determines the performance of a power converter with respect to transient and steady state response and stability. Thus, it is also profitable to examine the characteristics of basic modulation systems.

In practical systems, it is the interaction between constraints of the modulation system and limitations of the output stage that determine the overall performance of the power converters. The output stage and modulation system must be "matched" to each other to achieve optimum results.

Table 3.1

APPROACHES --- UNCONDITIONED DC TO CONDITIONED AC		
METHOD	EXAMPLE	
Raw DC → Conditioned DC → Modulated AC → Filtered AC Output Raw DC → High Frequency AC → Modulated AC → Filtered AC Output Raw DC → Modulated AC → Filtered AC Output	Cascaded DC Regulator - Inverter Cascaded High Frequency Inverter - Cycloconverter Pulse Width Modulator	

3.1 MOTOR CONSTRAINTS ON THE INVERTER

The study of the induction motor in Chapter 2 indicates that the motor imposes certain constraints and requirements on the power supplied by the inverter system. These may be summarized in the five categories of

- 1) Voltage of fundamental component
- 2) Current of fundamental component
- 3) Frequency of fundamental component
- 4) Harmonic structure
- 5) Power - Thermal

The interrelationships and tradeoffs between the motor and the inverter may be couched in terms of these categories as parameters.

The first three categories represent direct constraints by the motor. For a given motor and a given set of motor output conditions (torque and speed), the interrelationships between voltage, current and excitation frequency are prescribed and only one of these three variables may be independent. The maximum flux limitation of the motor sets end points on the equations expressing these interrelations. Efficiency considerations of the motor and the inverter may be used as one criterion in selecting an optimum operating point from among those permissible.

The output voltage harmonic structure allowable is dictated by the construction of the motor and tolerable excess motor losses due to harmonic currents. Since the input impedance of the motor may be smaller at a harmonic frequency than at the fundamental, harmonic currents may be significant when compared to the fundamental and, hence, the peak current supplied by the inverter may be significantly larger than the peak of the fundamental current. (See Section 2.4) Since the fundamental current is constrained by the load requirements, large harmonic currents may necessitate an inverter with an oversized peak current rating.

Torques produced by harmonics in the drive voltage are generally not considered useful or desirable in high performance VSIM drive applications, hence optimal motor performance is achieved with pure sine-wave drive. The burden of producing pure sinewaves is generally excessive, leading to increased dissipation, cost and complexity in the inverter. Thus, the harmonic structure tolerable must be a compromise between the motor factors of harmonic losses, currents and spurious torques (both harmonic and pulsating) and the inverter factors of peak current rating, losses, cost and complexity. To optimize this compromise, these tradeoffs must be studied for a particular mission requirement.

To meet a specific mission requirement, a maximum voltage and current will be required, but the two may not be required simultaneously. Thus, the maximum power required by the motor is a design criteria for the inverter. Thermal design of a static inverter can account for a significant amount of the size and weight of the unit. Even though an inverter is electrically capable of producing maximum voltage and current simultaneously, a decrease in total power losses obtained by not demanding maximum power results in a smaller inverter package.

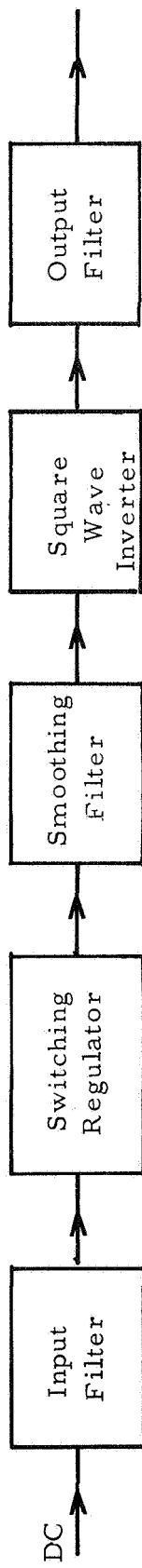
3.2 BLOCK DIAGRAMS

Table 3.1 tabulates the basic power conversion approaches to realizing the DC to conditioned AC required in VSIM applications. Figure 3.1 illustrates block diagram mechanizations of the conversion approaches in Table 3.1. The power converter (or converters) in each of the block diagrams consist of output stages combining switching elements, inductors, transformers and capacitors. The on-off times of the switches are controlled in a manner which yields the desired output with high efficiency. This section will examine the characteristics of the blocks of Figure 3.1 and present some of the basic output stage realizations for each of the approaches.

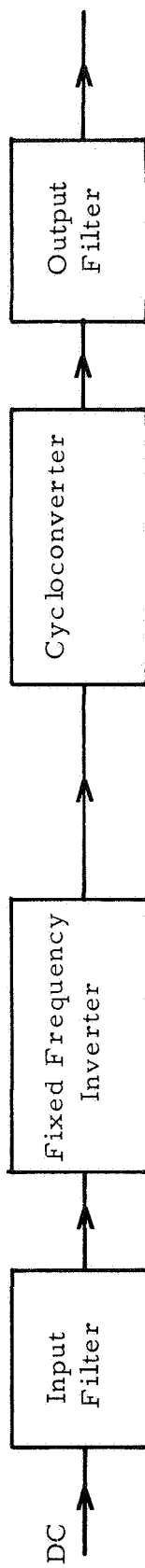
Base Band System

A base band system is defined as a system in which the inverter switches operate at the output frequency. The on-off times of the switches are not used for amplitude control, but rather a separate DC-DC converter/regulator preceding the inverter controls the main inverter supply voltage

Base Band System



Cycloconverter



Pulse Modulator

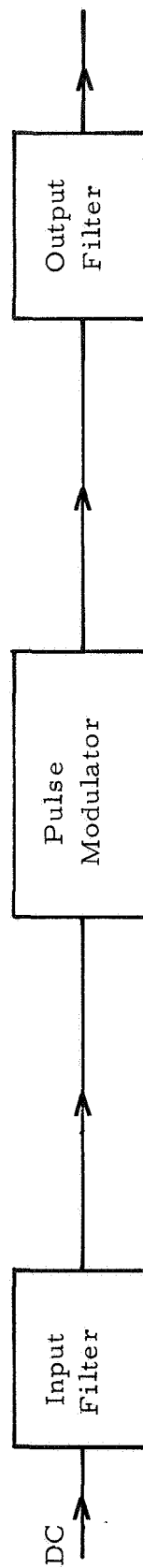


FIGURE 3.1

BLOCK DIAGRAMS OF THE THREE BASIC INVERTER SYSTEMS

and, hence, the peak output voltage. A filter is necessary between the converter/regulator and the inverter to provide filtering of the regulator ripple components. This filtering minimizes the generation of spurious frequency components in the output due to "beating" effects between the regulator and converter frequencies and prevents interactions between the control loops of the two units.

Figure 3.2 illustrates a simple realization of the baseband system using unilateral switching elements (e.g. transistors or force-commutated thyristors) that are "idealized" as perfect switches and diodes. The switching regulator is composed of a bidirectional DC "chopper" circuit whose duty cycle is modulated to yield the desired V_1 . The nature of this circuit is such that DC energy may flow in either direction. The inverter is composed of a "half bridge" connection of switches for each phase of the motor. For three phase operation, the motor winding would be connected in a "star" or "delta" configuration to the outputs of the inverter. The inverter switches are operated so as to produce a square wave output of peak-peak amplitude V_1 at the desired stator drive frequency as illustrated in Figure 3.3. The timing of the square wave transistors is adjusted to yield the proper phasing.

As shown in Section 2.3, this output waveform is rich in harmonic content at frequencies near the fundamental frequency, regardless of the connection of the motor windings; hence, at low frequencies spurious torques may become troublesome.

Additive Base Band System

A variation on the basic base band system employs a multiplicity of transformer coupled parallel inverters of the type shown in Figure 3.4 to generate a multistep waveform. Six of the inverters of Figure 3.4 may be combined in series to form a three phase system as shown in Figure 3.5. With the relative phase of the inverters as shown, the six waveforms of Figure 3.6a are generated. By proper connection of

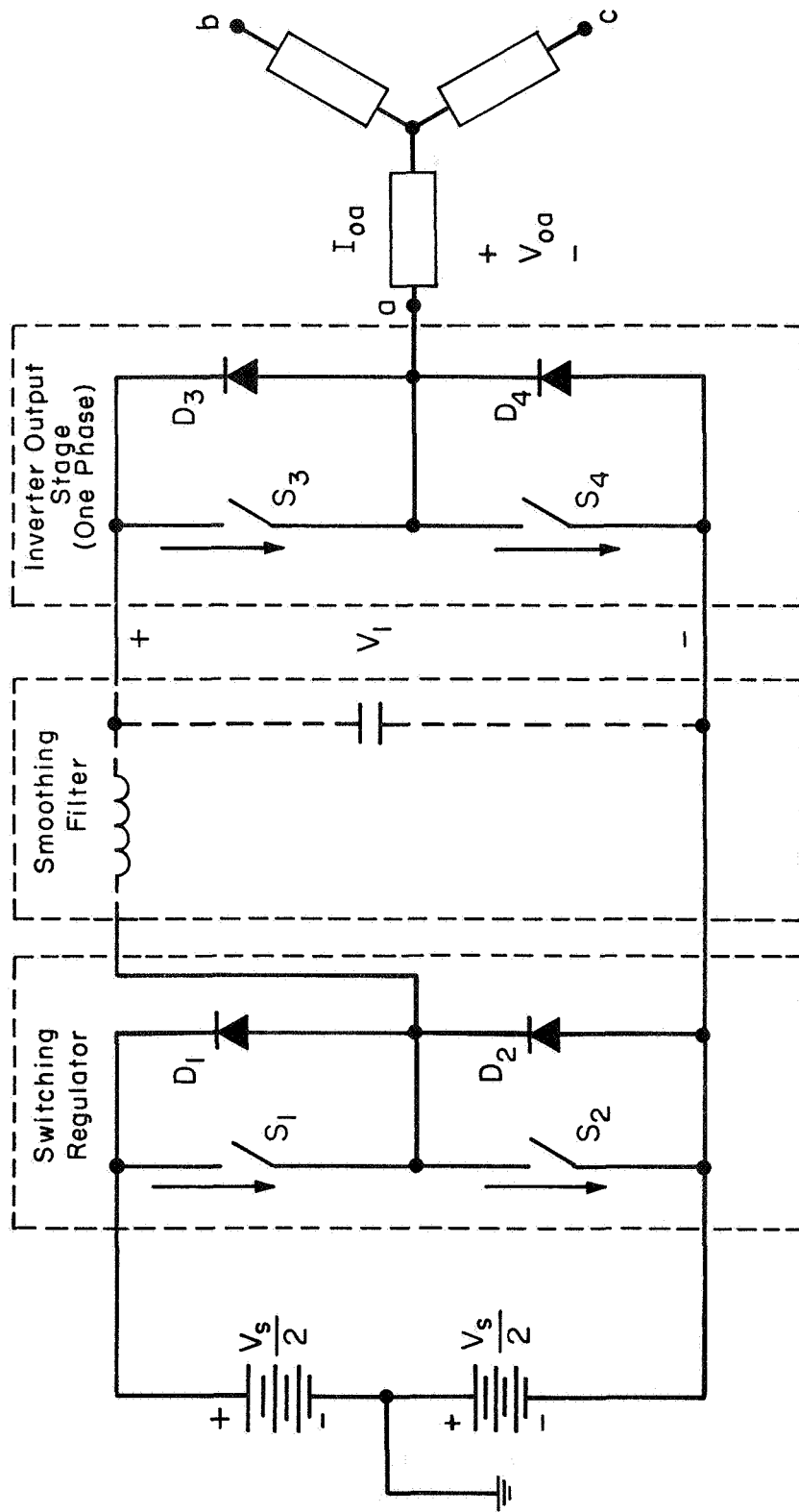


FIGURE 3. 2

SIMPLE REALIZATION OF BASE BAND SYSTEM

(Only one phase of a three phase output stage is shown. Arrows indicate direction of current flow in the switching devices.)

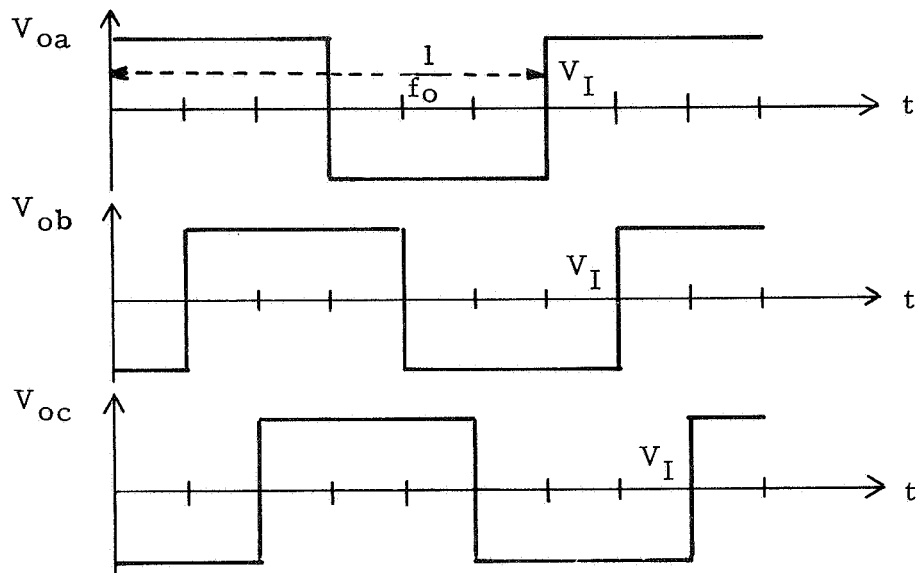


FIGURE 3.3

OUTPUT WAVEFORMS OF A SIMPLE THREE PHASE BASE BAND SYSTEM

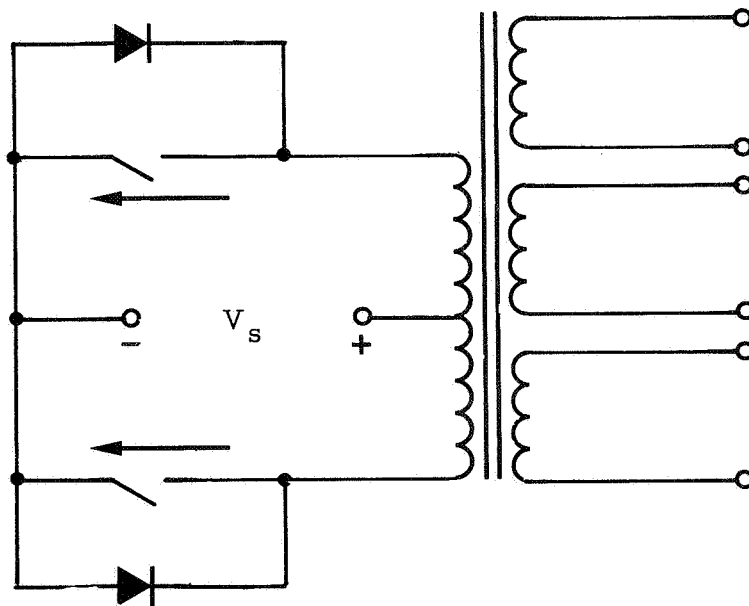


FIGURE 3.4

BASIC PARALLEL SQUARE WAVE INVERTER
(Arrows indicate direction of current flow in
switching devices.)

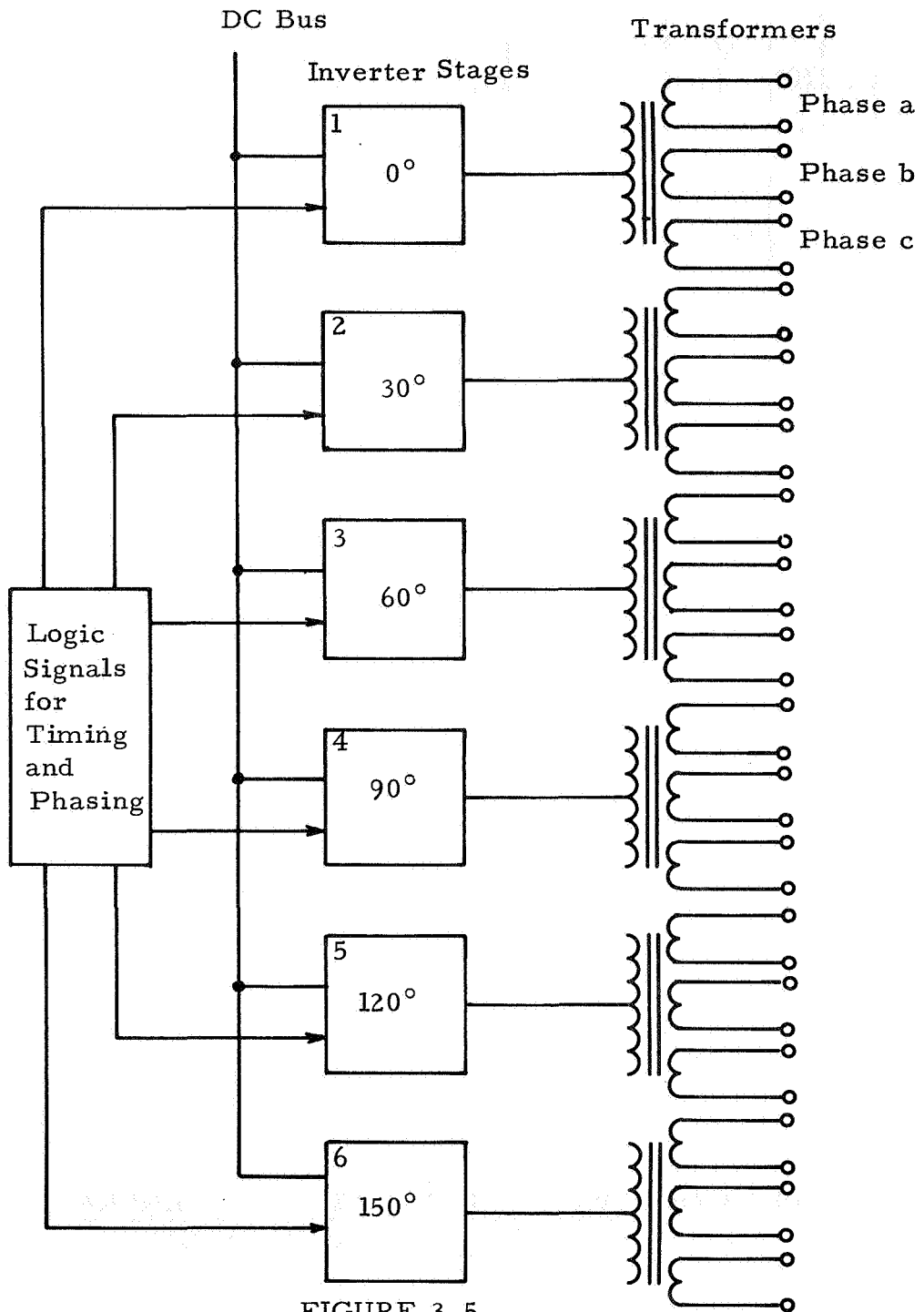


FIGURE 3.5

REALIZATION OF ADDITIVE BASE BAND SYSTEM

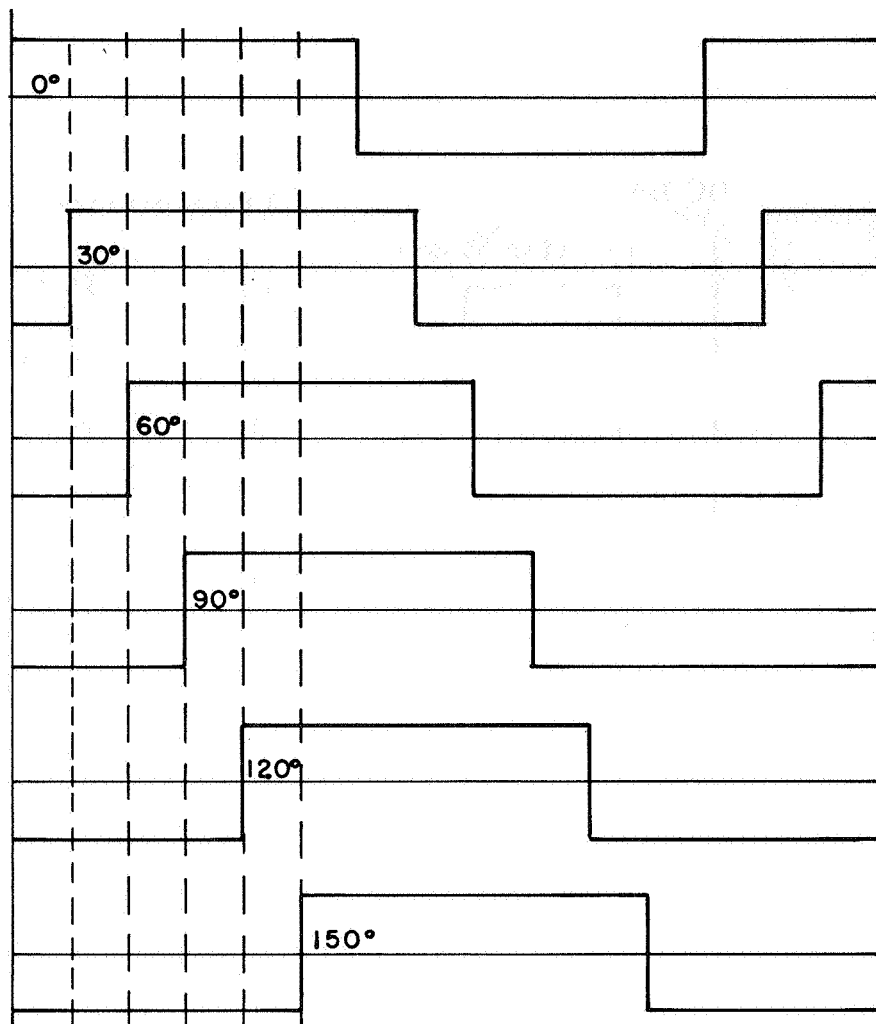


FIGURE 3.6a - INDIVIDUAL INVERTER WAVEFORMS IN ADDITIVE BASE BAND SYSTEM

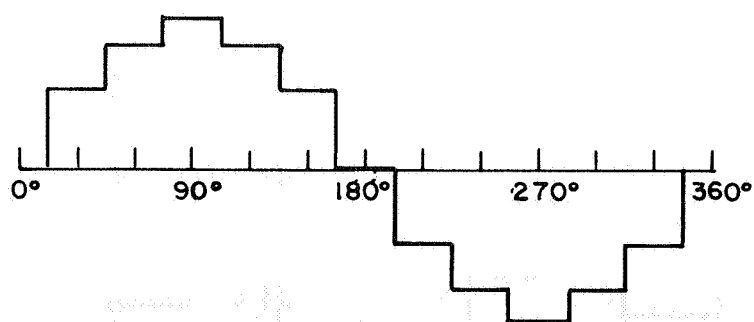


FIGURE 3.6b - TWELVE STEP SINE WAVE GENERATED BY ADDITIVE BASE BAND SYSTEM

the transformer secondaries, a three phase set of twelve step per cycle waveforms illustrated in Figure 3.6b may be generated with peak amplitude proportional to the DC supply voltage, V_I . For this example, employing six inverters, all harmonics of the fundamental up to the tenth may be cancelled by turns ratio adjustment.¹

Since the inverters are transformer coupled, operation down to DC is not possible. To drive an induction motor with an arbitrary load, special control techniques must be employed near zero frequency if this technique is to be employed.

Cycloconverter System

The cycloconverter employs a two-step conversion of DC power. The second entry in Table 3.1 shows the sequence of power conversion. The DC input is converted to a relatively high frequency AC signal. A modulator operating synchronously with the high frequency AC, known as a cycloconverter, produces an AC component at a lower frequency. By properly controlling the modulator, the frequency, amplitude and phase of the output signal may be controlled. A simple realization of the cycloconverter system is shown in Figure 3.7. A high frequency square wave is generated by a parallel inverter. This square wave is coupled to the cycloconverter by transformer T. The operation of the cycloconverter may be understood by reference to the waveforms of Figure 3.8. Assume that the inverter operates at frequency f_1 and the output current of the "a" phase is positive into a highly inductive load. Further, assume switch 1 (Figure 3.7) is closed. The cycloconverter is now transferring power from the source to the load. At $T_1/2$, the inverter switches and V_I changes sign. Because the load is inductive, current continues to flow and switch 1 remains closed. Power is now being transferred from the load inductance through the inverter and back to the DC source. At some time during this half cycle of inverter operation, switch 2 is closed and switch 1 opens. Power is again being delivered from the source to the load, but through switch 2 instead of 1. At time T_1 , the inverter again switches, during this half cycle, the sequence of power flow is identical to the previous half cycle, but the roles of switches 1 and 2 are interchanged. By proper control of the cycloconverter switching, the average voltage at the output may be controlled. Low frequency modulation of the cycloconverter switching allows a low frequency waveform to be constructed

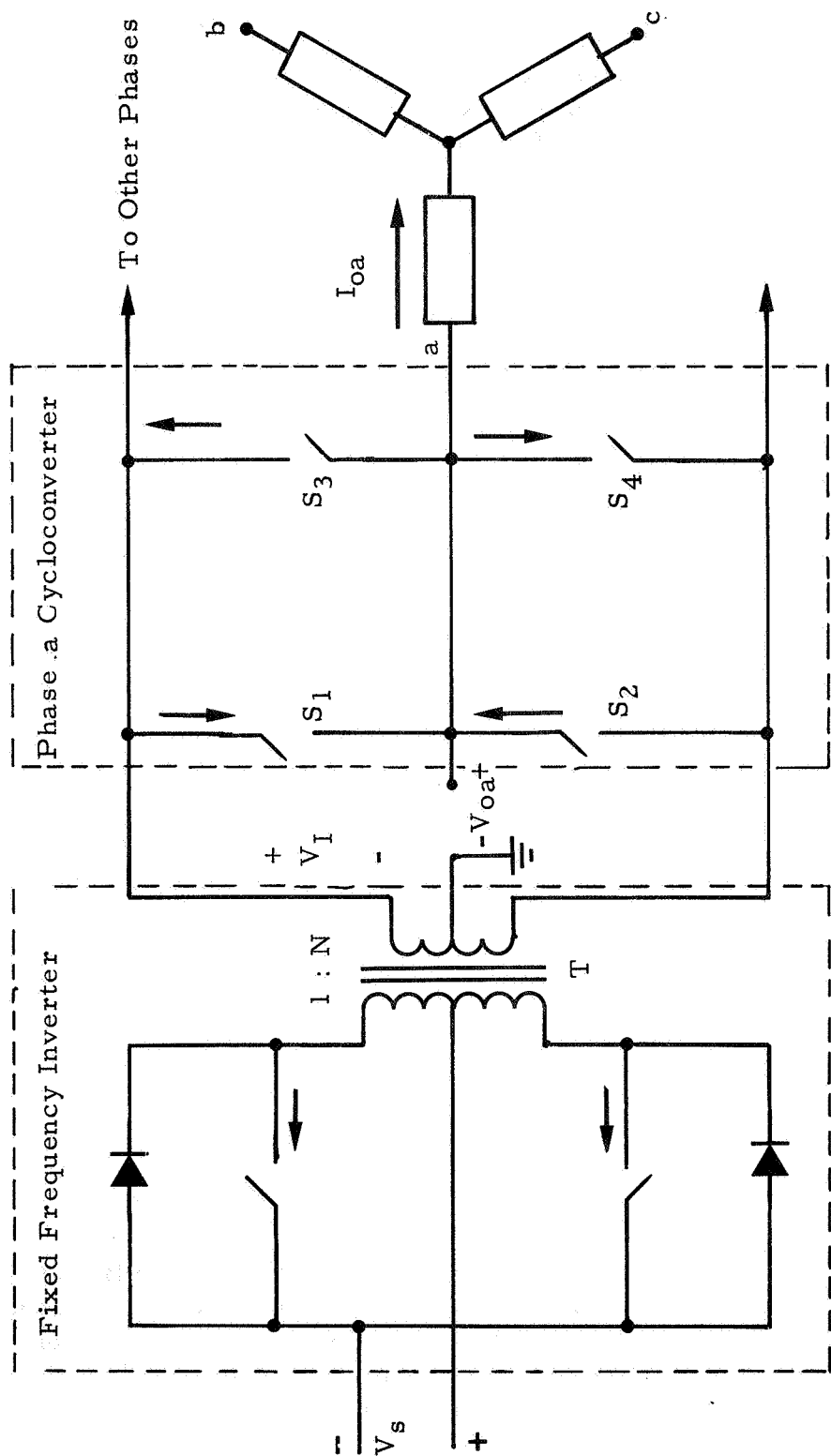


FIGURE 3.7

SIMPLE REALIZATION OF THE CYCLOCONVERTER SYSTEM
 (Only one phase of a three phase output stage is shown. Arrows indicate direction of current flow in the switching devices.)

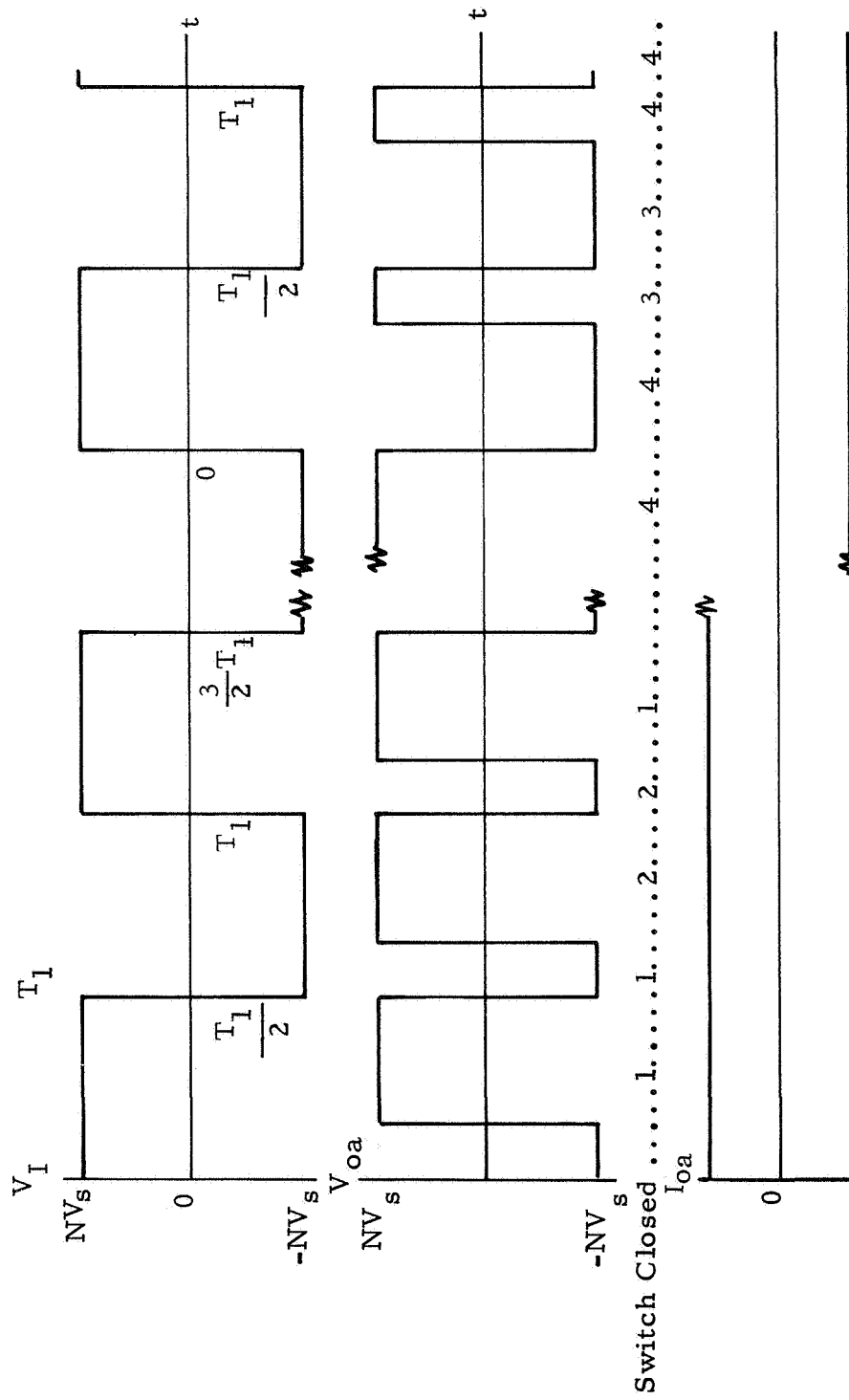


FIGURE 3.8

CYCLOCONVERTER OPERATION WITH INDUCTIVE LOAD ($T_1 = 1/f_1$)

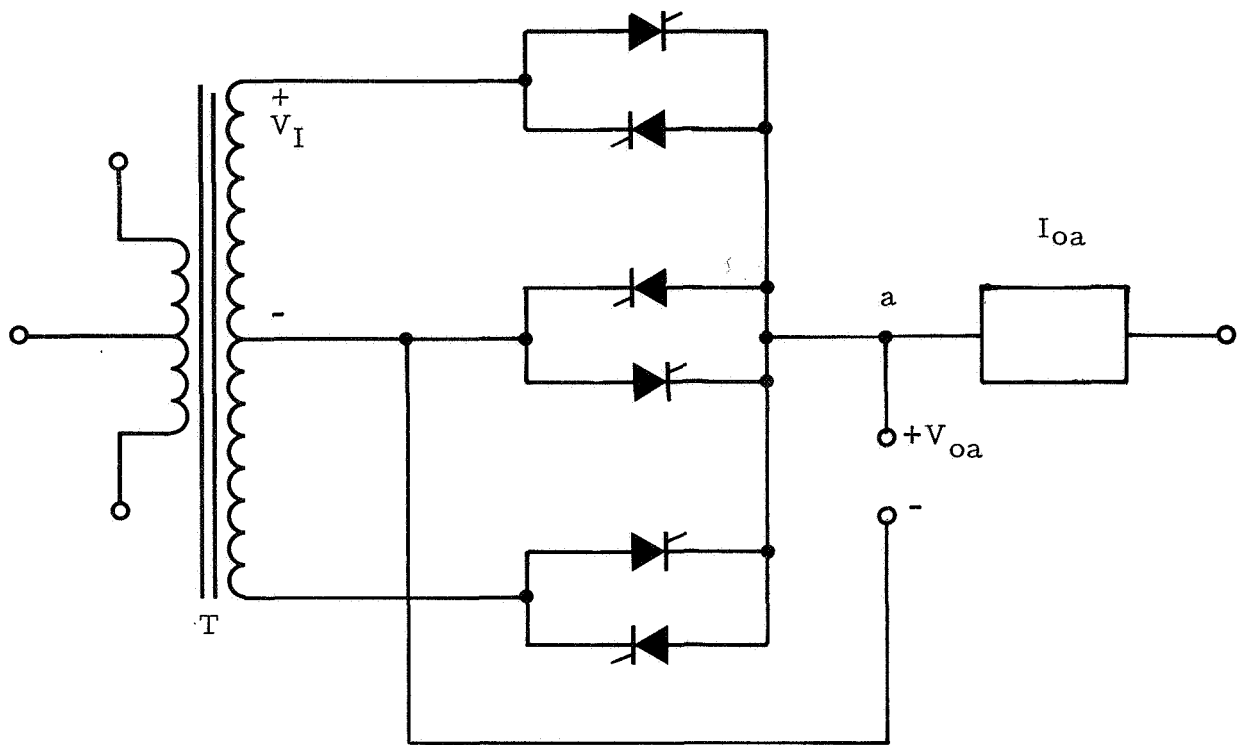
from samples of the high frequency inverter output. When the output current I_{oa} reverses, switches 3 and 4 assume the role previously played by switches 1 and 2, but operation is otherwise identical. As Figure 3.8 shows, the modulation generated is identical to fixed frequency pulse width modulation at twice the inverter frequency. Duty cycle modulation techniques employed in pulse width systems may be used to generate the transition timing signal with only slight modifications.

This basic cycloconverter scheme may be extended to multistate cycloconverters, that is, cycloconverters operating from a square wave source that allow multiple output voltages. Figure 3.9 shows the extension of one phase to three state and four state systems using thyristor switches. In the three state system shown, the output may be $V_{oa} = \pm V_I$ and 0. In the four state system, the outputs may be $V_{oa} = \pm V_I$ and $V_{oa} = \pm KV_I$ where K is some fraction of V_I less than 1 determined by the point at which transformer T is tapped. Proper operation of the switches in the multistate system can produce the same low frequency output as the two-state system, but the amplitude of the high frequency ripple components which must be filtered will be considerably reduced.

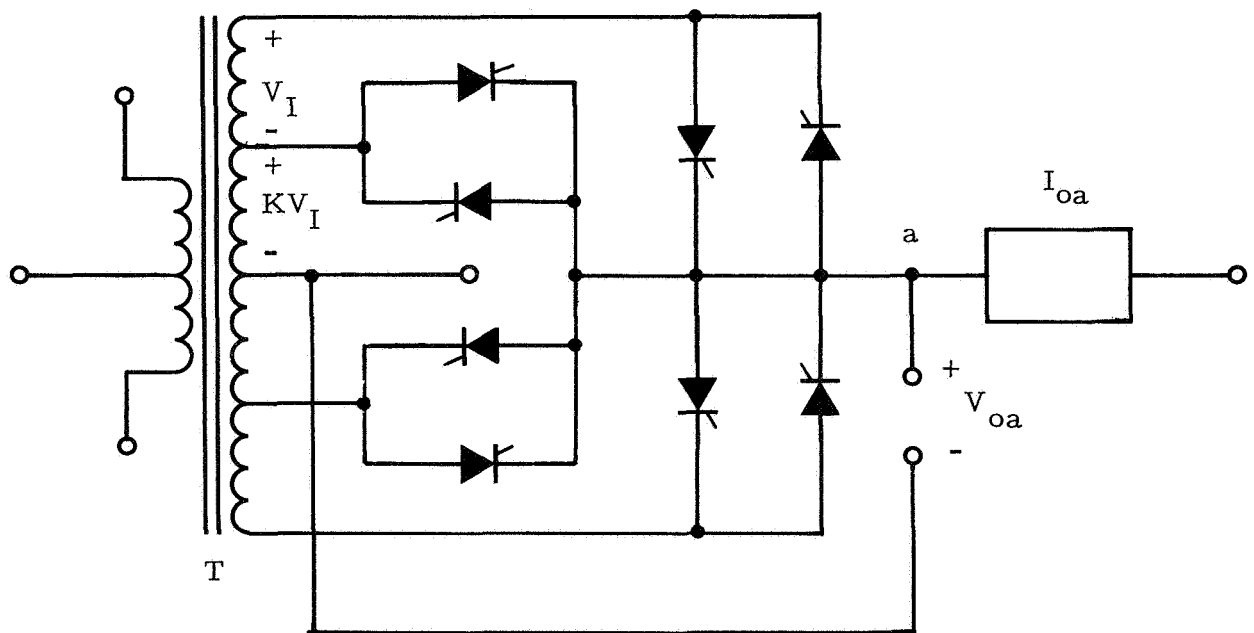
The same switching principles may be applied to cycloconverters having high frequency sine wave instead of square wave sources. Cycloconverters use a relatively large number of active switching elements when compared to other DC-AC conversion methods, but the fact that only the two inverter switches need be force commutated makes this scheme attractive with respect to efficiency considerations.

Pulse Modulator System

The final type of inverter system to be considered is the pulse modulator, the third approach tabulated in Table 3.1. This system converts DC directly to the desired AC output superimposed on a high frequency carrier. The simplest realization of this type of system is shown in Figure 3.10a. One phase of a three phase system is drawn. The switches, in a half bridge configuration, are operated alternately and the diodes provide a free-wheeling path for inductive loads. By proper control of the duty cycle of the switches in each phase, a three phase set of signals of controlled amplitude, frequency and phase may be generated directly from the DC source. The half bridge configuration shown in Figure 3.10a has an inherent bidirectional power flow



a) Three State Cycloconverter

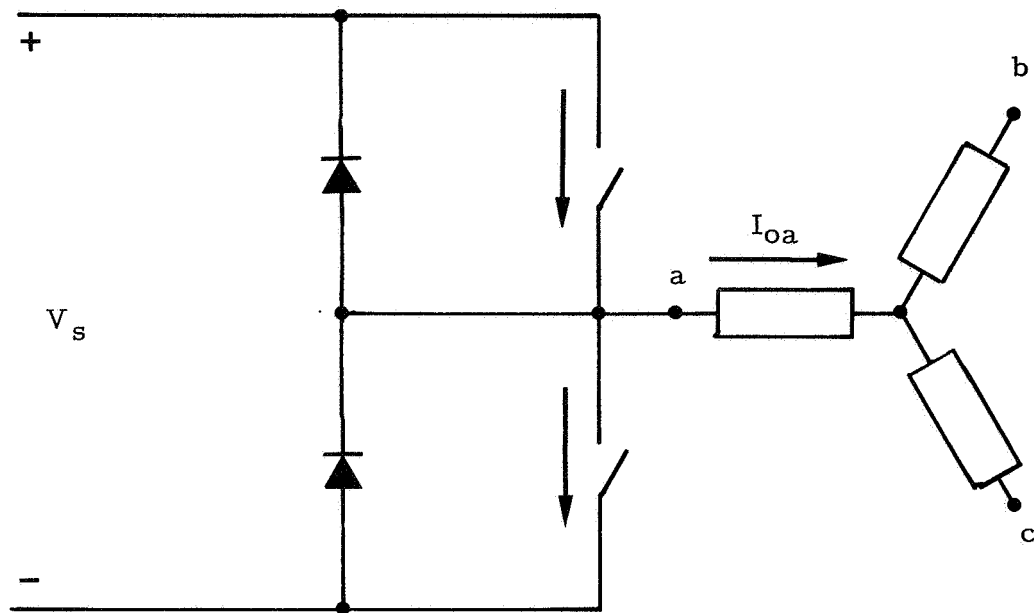


b) Four State Cycloconverter

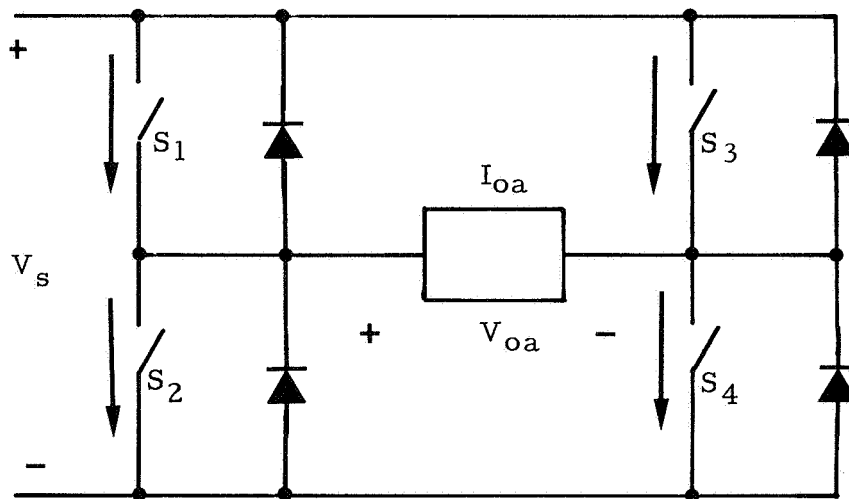
FIGURE 3.9

MULTI-STATE CYCLOCONVERTERS

(Only one phase of a three phase output stage is shown.)



a) Simple Realization of the Pulse Modulator System



b) Full Bridge Connection

FIGURE 3.10

TWO CONFIGURATIONS OF THE PULSE MODULATOR

(Only one phase of a three phase output stage is shown. Arrows indicate direction of current flow in the switching devices.)

capability. Thus, power may be returned to the source during deceleration.

Another pulse modulator configuration that may be employed is the "full bridge" connection; Figure 3.10b illustrates one phase of this output stage. Each phase of the load is separated and connected between the legs of the full bridge circuit. This connection may provide three output voltages. With switches 1 and 4 closed $V_{Oa} = +V_s$. With switches 2 and 3 closed, $V_{Oa} = -V_s$. With either switches 1 and 3 or 2 and 4 closed, $V_{Oa} = 0$. Thus, three-state operation independent of the other phases is possible with a full bridge circuit.

The three phase half bridge configuration can be realized with only six switching devices. This is a smaller number than is possible with any of the other methods outlined here, and accounts for the popularity of this configuration in pulse modulators. The full bridge, on the other hand, is a more versatile connection, allowing for three state operation, which can be made to generate less harmonics than a comparable two state system. Non-rotating harmonics (see Section 2.3) can cause troublesome circulating currents with a full bridge because of the ungrounded neutral.

AC pulse modulators can be operated in two basic modes, synchronous and asynchronous. In the synchronous mode, the switching frequency is made an integral multiple of the output frequency and is phase locked to it. Thus, the number of high frequency pulses per output cycle is constant regardless of output frequency. This approach produces modulation frequencies at well defined output frequencies harmonically related to the fundamental. In the asynchronous mode, the switching and output frequencies are unrelated and the modulation frequencies can only be defined as "bands" of frequencies related to the output and carrier frequencies and their various sums and differences.

Summary of Fundamental System Parameters

Certain properties of power handling stages are fundamental to the stages themselves and relatively independent of the control concept (or modulation system) used to generate the desired output. These properties are listed in Table 3.2 and tabulated for each of the systems discussed in this section.

Table 3.2

TABULATION OF SYSTEM PARAMETERS

	Base Band System 30 Fig. 3.2	Additive Base Band 30 Fig. 3.5	Cycloconverter Two State 30 Fig. 3.7	Cycloconverter Three State 30 Fig. 3.9a	Pulse Modulator Two State 30 Fig. 3.10	Pulse Modulator Three State 30 Fig. 3.10
1) Number of Power Handling Active Devices (S) and Power Diodes (D)	8S + 8D	(2 + 2K)D + (2 + 2K)S K is the number of inverters used.	14S + 2D	20S + 2D 26S + 2D	6S + 2D	8S + 8D 12S + 12D
2) Number of Devices in Series with Output at any Instant	1S or 1D in Input Stage. 2, Either S or D, in Output.	Effectively Same as Simple Baseband	1S or 1D in Input Stage. 2S in Output Stage.	Same as Two State Cycloconverter	2, Either S or D, in Output.	2, S or D
3) Number of Devices to be Force Commutated.	2 at Frequency f_1 6 at Frequency f_o (f_1 is the Regulator Switching Frequency.)	2 at f_1 2K at f_o	2 at Frequency f_1 6 at Frequency f_o (f_1 is the Regulator Switching Frequency)	Same as Two State Cycloconverter	6 at Frequency f_1 (f_1 is the Switching Frequency.)	4 at f_1 and 4 at f_o 6 at f_1 and 6 at f_o
4) Number of Forced Commutations Per Output Cycle.	$2\frac{f_1}{f_o} + 6$	$2\frac{f_1}{f_o} + 2K$	$2\frac{f_1}{f_o} + 6$	Same as Two State Cycloconverter	$6\frac{f_1}{f_o}$	$4\frac{f_1}{f_o} + 4$ $6\frac{f_1}{f_o}$
5) Nature of Filter Requirements	Input Filter: Blocks f_1 for RFI Protection of Supply. Smoothing Filter: Smooths Rectified Signal and Isolates f_1 From Output. Output Filter: Blocks f > f_o max only.	Input - Same as Simple Baseband Output - Less Harmonics than Simple System. Depends on K and Harmonic Cancellation Scheme Used.	Input Filter: Blocks f_1 for RFI Protection of Supply. Output Filter: Blocks f_1 for RFI Protection and Attenuates All Undesirable Harmonics.	Input - Same as Two State Cycloconverter. Output filter Sees Progressively Less Harmonics.	Input Filter: Blocks f_1 for RFI Protection of Supply. Output Filter: Blocks f_1 for RFI Protection and Attenuates $f > f_o$ max.	Input - Same as Two State Pulse Modulator. Output - $\frac{2}{3}$ the Harmonics of Two State.

6) Spectrum of Output Signal. Components Clustered About:	$f_o, 3f_o, 5f_o$	Same as Simple Baseband, but Lower Amplitudes. Harmonic Cancellation may be Used to Null Out Low Order Harmonics.	$f_o, f_1, 2f_1, 3f_1, \dots$ $f_1 > f_o \max$	Same as Two State Cycloconverter	a) Synchronous System ($f_1 = nf_o$), $f_o, 3f_o, 5f_o, \dots$ b) Asynchronous System (f_1 unrelated to f_o), $f_o, f_1, 2f_1, 3f_1, \dots$ $f_1 > f_o \max$	$f, g, i, j, 2f, \dots$
7) Peak Voltage and Current Requirements on Switching Devices.	Input Devices: $V_s \max, I_o \max$ Output Devices: $2V_o \max, I_o \max$	Input Devices: $V_s \max, I_o \sum_{k=1}^K N_k$ Output Devices: $2V_o, N_{\max} I_o$ $\frac{K}{\sum N_k}, N_{\max} I_o$ $k=1$ (V_o is Slightly Greater Than Simple Baseband.)	Input Devices: $2 V_s \max, N I_o \max$ Output Devices: $2N V_s \max, I_o \max$	Added Devices see Lower Voltages Than Two State Cycloconverter $I_o \max$.	$V_s \max, I_o \max$	$V_s \max, I_o \max$
8) Tolerance of Supply Voltage Variation (V_b = Breakdown Voltage of S or D)	$V_s \max < V_b$ $V_s \min > V_o \max$ Instantaneous Peak Value)	$V_s \max < V_b$ $V_s \min > V_o \max$ K N (N is Transformer Turns Ratio.)	$V_s \max < \frac{V_b}{N}$ (Input Devices) or $V_s \max < \frac{V_b}{2N}$ (Output Devices) $V_s \min > \frac{V_o \max}{N}$ (Instantaneous Peak Value)	$V_s \max < \frac{V_b}{N}$ for C.T.(S) $V_s \max < \frac{V_b}{K+N}$ for Added (S)	Same as Baseband System	Same as Two State Pulse Modulator.
9) Provision for Isolated DC Output	None	None	Yes	Yes	None	None
10) Provision for Isolated AC Output	None (May Be Provided with Addition of Output Transformer).	Yes	Yes	Yes	None (May Be Provided with Addition of Output Transformer).	None (May Be Provided with Addition of Output Transformer).

The following are the parameters which are attributable to the output stages of an inverter system that are basic to the performance evaluation of inverter systems and are tabulated in Table 3.2.

- 1) The number of power handling active devices and power diodes. (This affects size, weight, cost and reliability of the systems.)
- 2) The number of devices in series with the output at any instant. (This directly limits the efficiency of the system.)
- 3) The number of devices to be force commutated per inverter operating cycle. (This determines the size and complexity of the commutation circuits.)
- 4) The number of forced commutations per output cycle. (Since most energy loss during switching is in forcing commutation, this directly limits the efficiency of the system for a given forced commutation scheme.)
- 5) Nature of filter requirements. (This includes input line filters, interstage filters and output filters. The complexity and weight of the various filters are a function of both system requirements, e.g., the tolerable level of RFI fed back to the power line, and of the spectrum of the signals that the inverter generates.)
- 6) The spectrum of the output signal. (This influences motor losses and, at low speeds, the torque characteristics, in addition to affecting output filter design. If $f_{o \text{ max}}$ is the maximum output frequency required from the inverter and f_o is the frequency being called for at a particular time, then components in the spectrum at $f_o < f < f_{o \text{ max}}$ are particularly undesirable since they can cause spurious motor torques, contribute to dissipated energy, and cannot easily be filtered. We refer to them as "in band" harmonics. Also, since the harmonics appear to the motor as drive signals at high slip speed, the current drawn at these harmonics may be large. This excess current can add significantly to the current that must be commutated.)
- 7) Voltage and current requirements on switching devices. (This can determine if the system is realizable under the restriction of present device technology and directly affects the cost of the system.)

8) Tolerance of supply voltage variation. (This constrains the degree of compatibility between the inverter system and the characteristics of the power supply.)

9, 10) Provision for isolation of output from input. (This may be a system requirement.)

3.3 CONTROL CONCEPTS FOR SWITCHING INVERTER SYSTEMS

The normal practice in the design of power processing systems is to select an output stage which is capable of performing the desired power conversion and then to mate a control concept to the output stage. The selection of a control concept (or modulation system) to be applied to the output stage of an inverter system involves an inter-relationship of many complex factors. The output stage itself places certain constraints on the modulation, such as minimum and maximum "on" time or minimum duty cycles. Other factors are less precise but, nevertheless, important. The complexity or practical realizability of a control concept or the cost of a particular embodiment must govern the choice. Several control concepts may exist which, in conjunction with a suitable output stage, will yield inverter systems which meet the basic specifications for a given application. The engineer is then faced with a choice of "how good" to make the system: is added control system complexity and cost worthwhile to achieve lower harmonic content or tighter parameter control or greater bandwidth? These questions are usually answered by engineering judgement rather than analysis.

When a control concept is applied to a specific output stage, such as those described in Section 3.4, a number of questions about the combined system become pertinent. The questions can be grouped in the following interdependent factors:

- 1) Device and element utilization: Practical switching devices have maximum voltage and current ratings and magnetic components have volt-second or current limitations. In a given stage configuration, the control concept employed may affect these ratings for a given power output.

- 2) Dynamic losses: During turn-on or turn-off times of practical devices, conditions exist where energy is being dissipated by the switching devices or in the output stage circuitry. The power dissipated depends on the energy dissipated per operation of the switch and the rate at which the switches are operated. Control concepts which minimize the switching rate or the energy dissipated per switch operation (this can be done by judicious selection of the switching instant in certain output stages) will minimize dynamic losses.

3) Parameter control under steady state and transient conditions: Control of the output parameter of an inverter under various operating conditions may be affected by the selection of control concept. Under steady-state conditions, the system must provide smooth, stepless control of the average output voltage (or current) with respect to variations in amplitude and frequency of the output signal and must provide regulation against source voltage variations and load impedance variations. Under transient conditions, control must be maintained in a stable manner, with good transient recovery evident. In addition, under transient conditions control of the average output parameters (current or voltage) may be inadequate to guarantee safe operation of the output stage. A control concept may have to employ instantaneous control or limiting of an output variable (in particular output current) as well as average parameter control to achieve satisfactory operation under all operating conditions.

4) Harmonic Structure: Since only the fundamental component of inverter voltage and current contribute to the desired output, all other frequencies must be considered spurious components. Spurious voltages generate additional core losses in a motor while spurious currents generate additional stator and rotor copper losses and undesirable output torques.

All power conversion systems employing switching techniques generate spurious frequency components. The control concept determines the structure of these components while the output stage limitations govern the choice of the switching frequency and, hence, the frequencies of the spurious components.

Waveforms Generated By Various Control Concepts

Variable frequency AC suitable for VSIM drive applications, is synthesized from a DC source in two ways. In one method, a predetermined waveform is generated with transitions fixed in time relative to the period of the AC output. These systems use voltage variable clocks and counting circuits to generate switch timing information. The second method of AC waveform synthesis is to approximate a reference AC sine wave with a modulated high frequency signal. This method uses waveform generators and/or feedback sensors and

comparator circuits to generate the switch timing information.

The following discussion and examples indicate the nature of waveforms encountered using various control concepts. The discussion is in terms of a single phase output, but extension to multiple phases is straightforward.

1) Square Wave Drive: This drive, produced by the base-band system (See Section 3.2), represents the simplest predetermined waveform method of AC waveform generation. The spurious frequencies generated are odd harmonics of the fundamental and fall off as $1/N$ times the amplitude of the fundamental (N is the harmonic number of the spurious components).

2) Synchronous Pulse Frequency: This is a predetermined system which combines amplitude and frequency control in an inverse relationship; i.e., the volt second product of the waveform remains constant. This is accomplished by keeping the number of pulses per output cycle fixed.

Figure 3.11 shows an example of a six voltage pulse per cycle waveform at two different output frequencies. The height and width of each of the pulses is identical; the close spacing of the pulses at the high frequency implies that there is a larger fundamental component than for the low frequency case. Figure 3.11 shows equally spaced pulses through the cycle; however, any spacing that can be mechanically generated may be employed. For an N pulse per cycle synchronous pulse frequency system, where the width of each pulse is T_p , the maximum operating frequency is $f_{\max} = 1/(NT_p)$

and the output waveform at $f_{\max}$ is a square wave.

Pulse frequency modulation may be generated by the bridge circuit of Figure 3.10b. One half of the bridge is used to generate the pulses while the other side is switched on each half cycle to control the pulse polarity. Synchronous pulse frequency modulation is thus a three state modulation.

3) Sliding Wave (2 Pulse Synchronous): Another control concept suitable for applications with three state output stages is the "sliding wave" or two pulses per cycle synchronous modulation. Figure 3.12 illustrates how this modulation is constructed from two

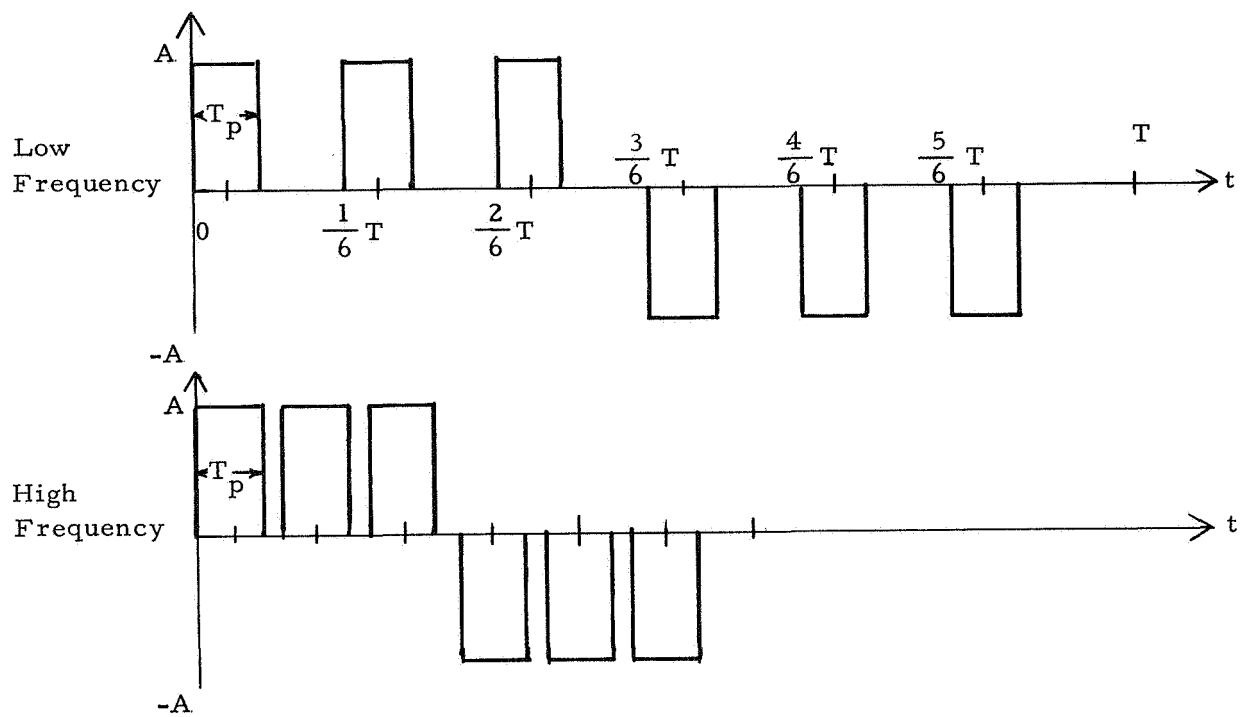


FIGURE 3.11

WAVEFORMS GENERATED BY SYNCHRONOUS PULSE
FREQUENCY CONTROL

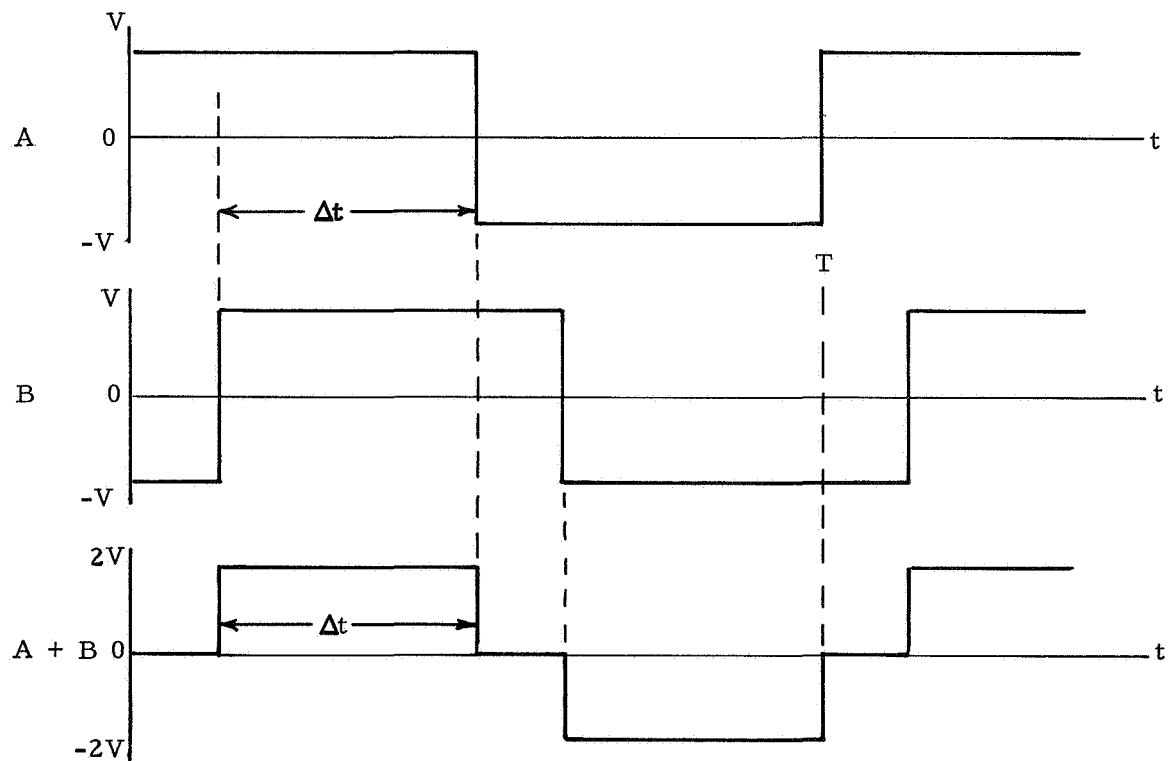


FIGURE 3.12

WAVEFORMS GENERATED BY "SLIDING WAVE" CONTROL

square wave inverters connected in series. The full bridge circuit (Figure 3.10b) can generate these waveforms. The inverters operate at the same frequency, but the phase of one of the inverters is shifted with respect to the other. The sum is a waveform with the same fundamental frequency as the component square waves, but the amplitude of the fundamental is dependent upon the relative phase between the two inverters. Figure 3.13 is a plot of the amplitude of the fundamental and the principle harmonics in relation to the source voltage versus the normalized pulse width parameter β , defined as

$$\beta = 2 \frac{\Delta t}{T} . \text{ At } \Delta t = T/2, \text{ both inverters are in phase and square}$$

wave drive results. As Δt approaches zero, the waveform approaches narrow pulses.² In Figure 3.14, the fifth and seventh harmonic components are plotted in relation to the fundamental. Since the fundamental drive component must be varied with drive frequency, this plot is a more meaningful way to assess the relative effect of the harmonic components. At very low motor drive frequencies, the fundamental amplitude must be made small. This corresponds to small values of β (see Figure 3.13). Figure 3.14 shows that for small β the magnitude of the harmonics approaches that of the fundamental, suggesting the possibility that the spurious torques described in Section 2.5 may become troublesome.

4) Pulse Width Modulation: The previously described methods were techniques for predetermined generation of AC. Pulse width modulation is a technique for approximating an AC reference sine wave with a modulated high frequency signal. Two forms of pulse width are of interest here, single edge and double edge modulation. To generate single edge modulation, a sawtooth waveform at the desired modulation frequency is subtracted from the reference sine wave, as shown in Figure 3.15. The output stage is turned on for a positive difference sensed by a threshold detector. The resultant rectangular waveform contains a major frequency component at the reference frequency and spurious frequencies as modulation components about the modulating frequency and its harmonics. Single edge modulation of the type shown is suitable for use with the square wave cycloconverter output stage if the sawtooth is synchronized with the high frequency inverter.

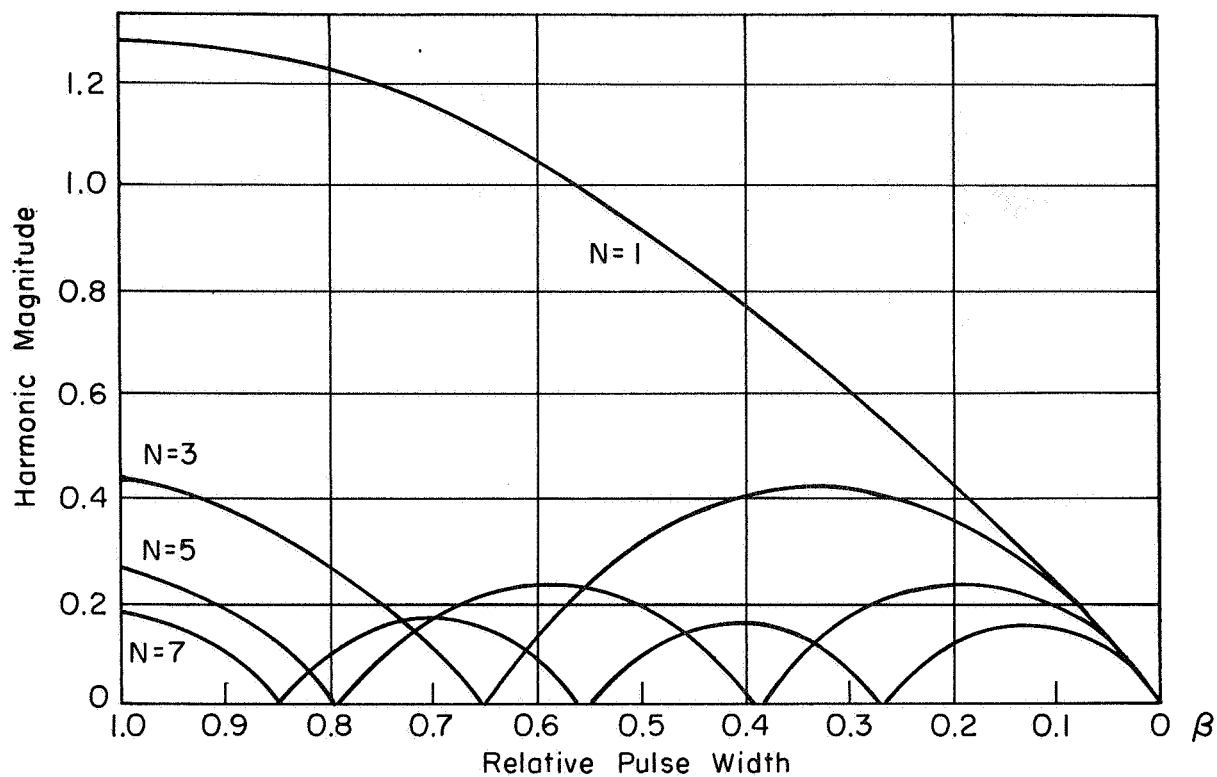


FIGURE 3.13

SPECTRUM OF "SLIDING WAVE" WAVEFORM RELATIVE TO SOURCE

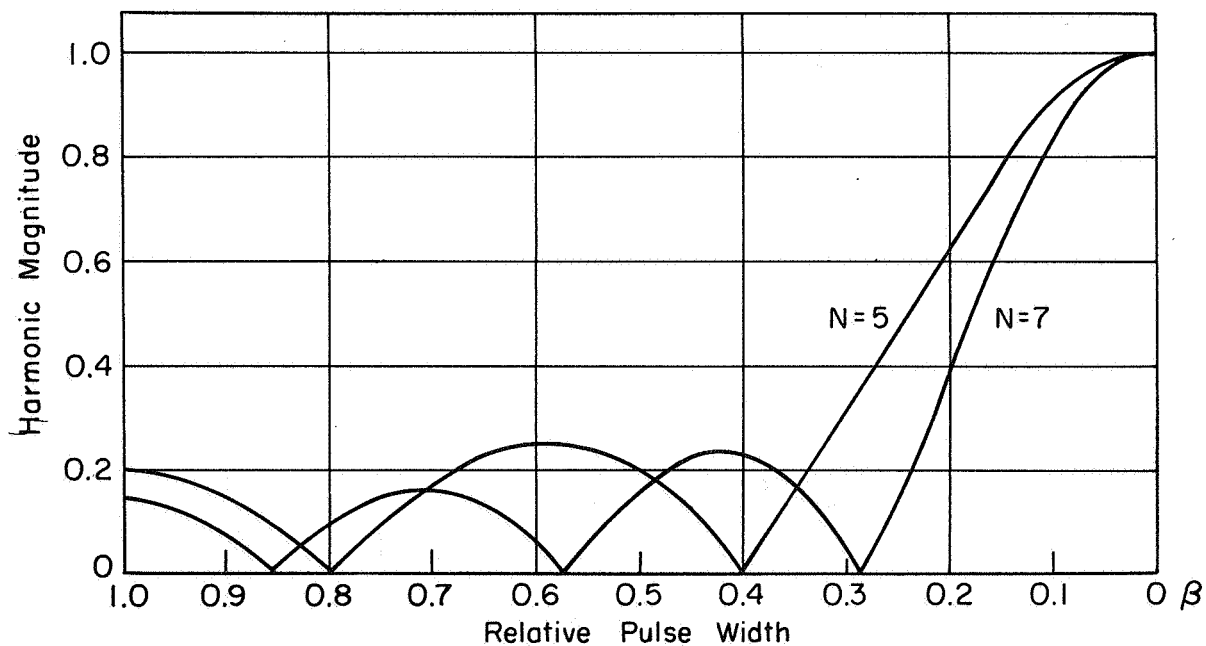


FIGURE 3.14

SPECTRUM OF "SLIDING WAVE" WAVEFORM RELATIVE TO
FUNDAMENTAL COMPONENT OF FIGURE 3.13

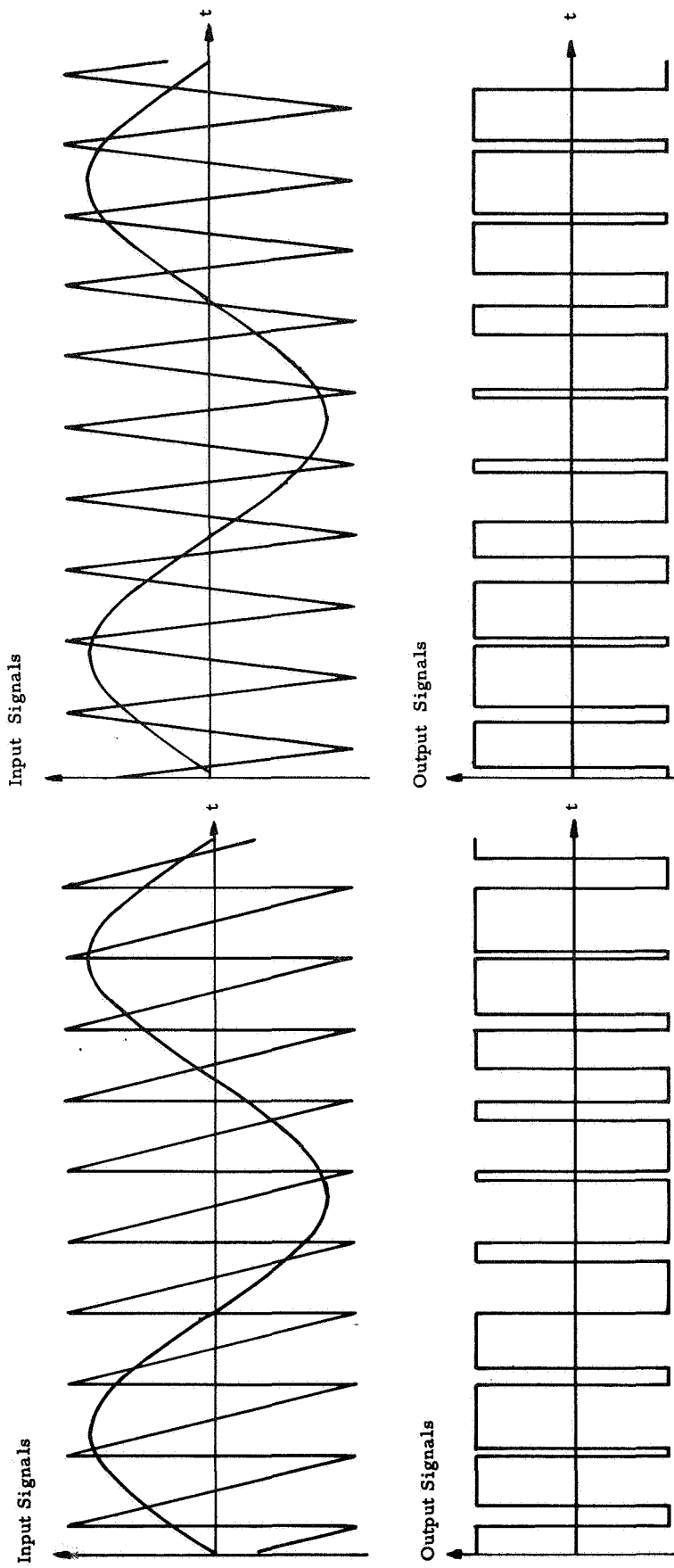


FIGURE 3.15 - WAVEFORMS ENCOUNTERED IN SINGLE EDGE
PULSE WIDTH MODULATION

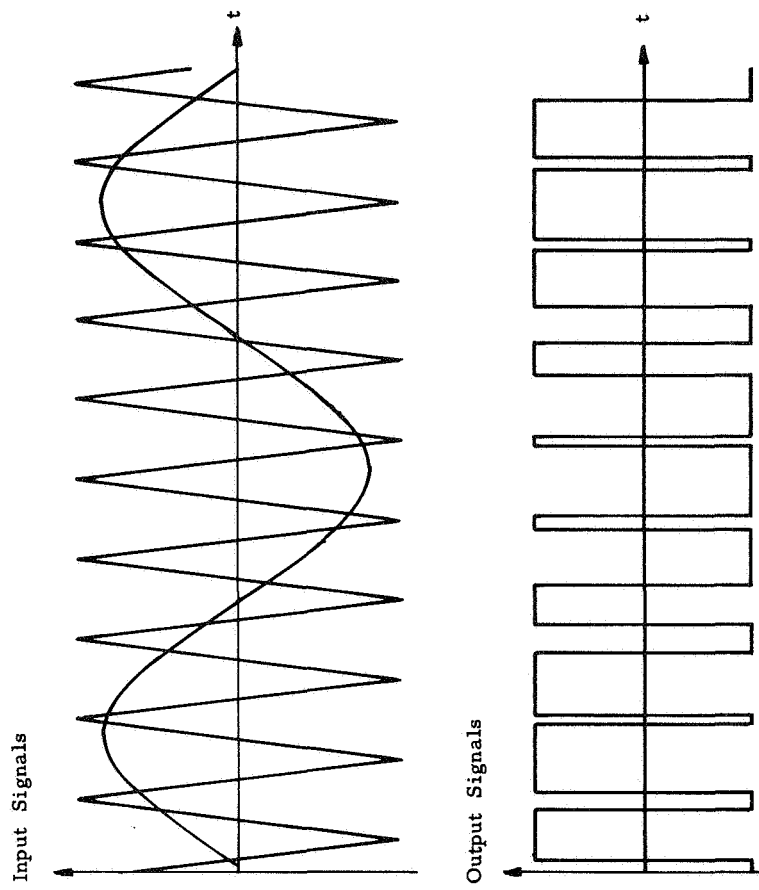


FIGURE 3.16 - WAVEFORMS ENCOUNTERED IN DOUBLE EDGE
PULSE WIDTH MODULATION

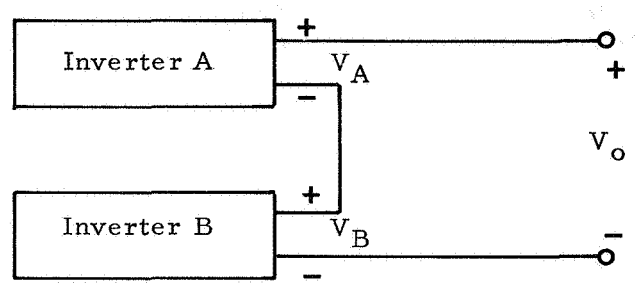
The method of generating double edge pulse width modulation is similar to that for single edge only a triangular waveform is used rather than a sawtooth. This case is illustrated in Figure 3.16.

5) Synchronous Pulse Width - "Subharmonic Method":

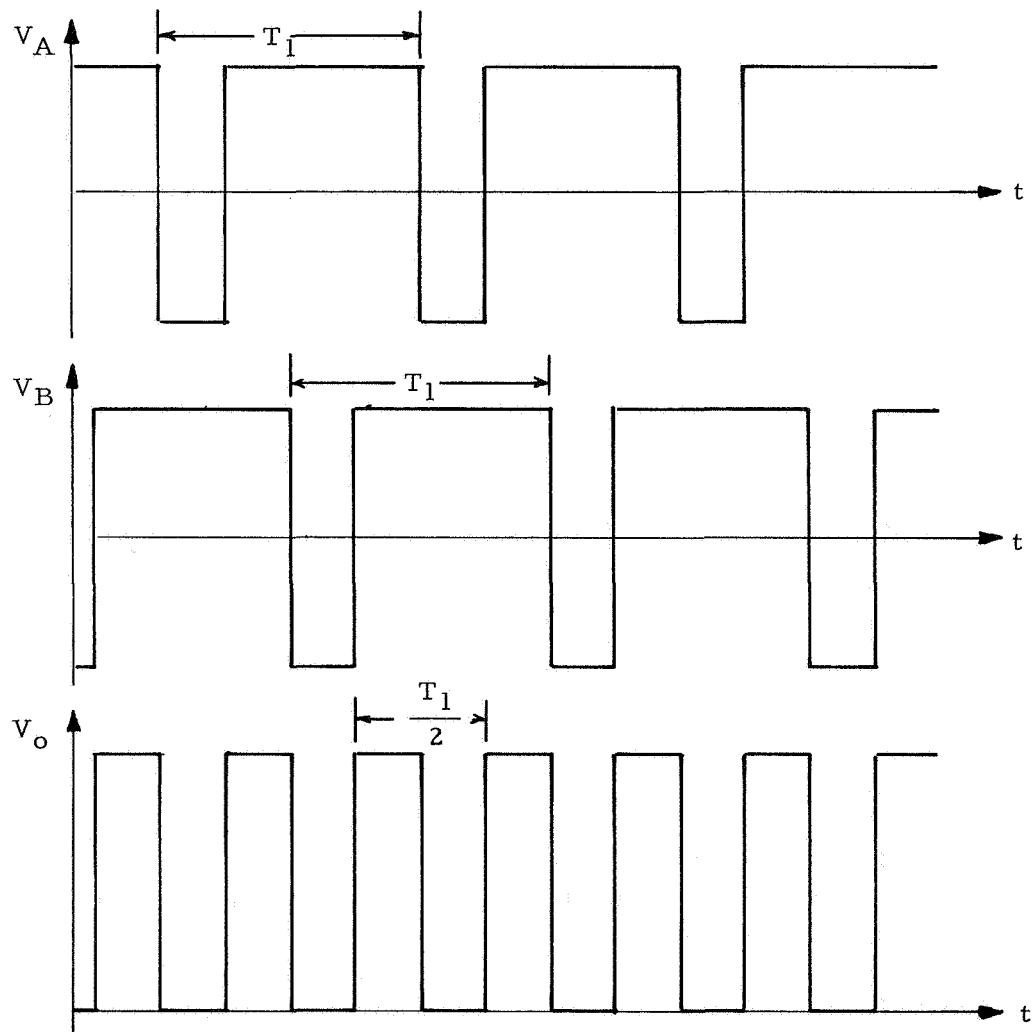
If the modulation generating waveforms in the pulse width system are synchronized to the desired AC signal; that is, if the frequency of the modulation signal is an integral multiple of the reference signal and phase locked to the reference, a form of pulse width modulation can be generated in multiple inverters which has the property of cancelling out the principle spurious frequency components of the modulation frequency. It has been reported that if a single triangular voltage is used to generate modulation signals from a three phase set of reference signals, the principle spurious component at the modulation frequency, which can be much larger than the output, will be non-rotating and, therefore will have no effect on a motor with ungrounded neutral. The remaining spurious components are smaller than the desired output signal.³

The subharmonic method can be extended to inverters connected in series to cancel complete groups of spurious components. How this may be accomplished is illustrated in Figure 3.17 for the case of two inverters in series (the full bridge connection may be used also) and a constant modulation. The modulation of inverter B is phase shifted by 180° with respect to the modulation of inverter A at the modulation frequency, f_1 . The resultant sum contains significant frequency components at $2f_1$ and integral multiples of $2f_1$ but no components at f_1 . The method may be extended to k series inverters, in which case, the significant spurious components would occur at kf_1 and integral multiples of kf_1 .

6) Current Control: The previous control concepts have generated a modulation based on control of the output voltage. In some motor drive applications, it is desirable to control the output current directly, rather than the voltage. A feedback loop can be employed to "servo" the voltage reference in a voltage control inverter to achieve output current control (see Section 4.1); or a control concept that performs this task inherently may be used as shown in Figure 3.18. The output voltage is applied to a load which has inductive characteristics at the switching frequency. The output current is sensed and compared



a) Series Connection of Inverters



b) Waveforms in Series Connection

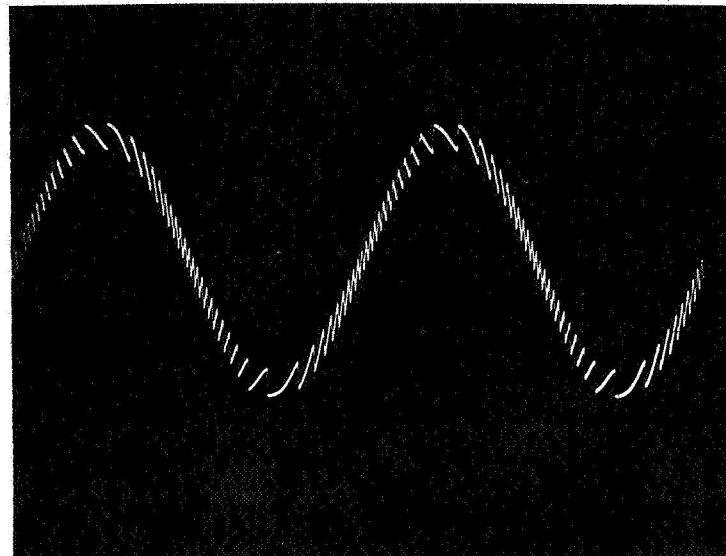
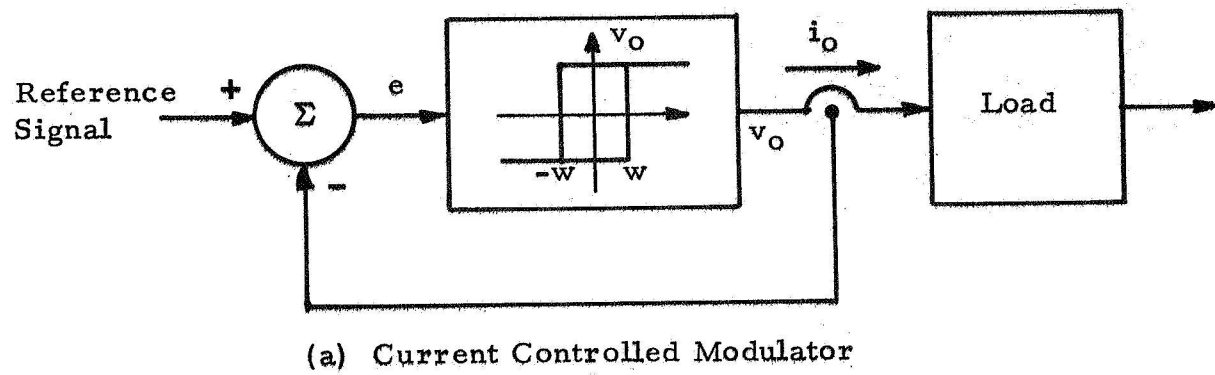
FIGURE 3.17

SPURIOUS FREQUENCY CANCELLATION WITH TWO INVERTERS IN SERIES

to a current reference signal. The instantaneous error signal is fed to a hysteretic switch, which causes the output stage to switch in a manner that will reduce the error magnitude when it exceeds a predetermined threshold. The resultant output has an average current equal to the reference current and a high frequency current ripple of constant peak-to-peak amplitude about the average current as Figure 3.18b shows. This control concept is of great interest because it controls both the instantaneous as well as average output current.

SUMMARY

In studying various control concepts used in power processors employing switching techniques, it becomes evident that there are two nearly independent parameters which must be specified in generating a switching waveform. One is the duty cycle of the output; that is, the ratio of "on" time to the total switching period. The allowable duty cycle at any instant is constrained by the load requirements and the source voltage; it must be the same regardless of the control concept employed. On the other hand, the switching frequency is only loosely constrained by the load through a spurious frequency constraint; any spurious frequency components must cause an insignificant degradation in performance. Since practical switching elements and magnetic components have limitations, the frequency is constrained between certain minima and maxima for a given duty cycle and output stage. Between these two constraints, the control concept is free to set the frequency. The principle differences in the control concepts presented here is the way in which the frequency is set. The way in which this constraint is set for the systems described and the use of this constraint as well as a summary of significant parameters due to the control concept alone are summarized in Table 3.3.



(b) Current Waveform Produced by Current Control Modulator

FIGURE 3.18

BLOCK DIAGRAM AND WAVEFORMS OF CURRENT CONTROL SYSTEM

Table 3.3

COMPARISON OF CONTROL CONCEPTS DESCRIBED						
	Square Wave Drive	Synchronous Pulse Frequency Voltage Control	Sliding Wave Voltage Control	Asynchronous Pulse Width Voltage Control	Synchronous Pulse Width Voltage Control	Current Control Two State
1) Parameter Control Average Current	Can be Obtained with Feedback	None	Can be Obtained with Feedback	Can be Obtained with Feedback	Can be Obtained with Feedback	Yes
Instantaneous Current	None	None	None	None	None	Yes
2) Harmonic Structure (Components Clustered About)	$f_o, 3f_o, \dots$ Falls Off as $\frac{1}{N}$	$f_o, 3f_o, \dots$	$f_o, 3f_o, \dots$	$f_o, f_1, 2f_1, \dots$	$f_o, kf_1, 2kf_1$ (Where k Depends on the Number of Inverters per Phase)	$f_o, f_1, 2f_1, \dots$
3) Variation of Harmonic Structure vs. Fundamental Frequency.	Scales with f_o	Scales with f_o , Increases for Decreasing f_o	Scales with f_o	None	Scales with f_o	None
4) Switching Frequency Constraint Set by	f_o	Integral Multiple of f_o	f_o	Clock	Integral Multiple of f_o	Rate of Change of Instantaneous Output Current
5) Use of Switching Frequency Constraint	Generates Output Frequency	Generate Constant Volts/Hz	Generates Output Frequency	Sets Frequency of Modulation Components	Cancel Spurious Frequencies	Maintain Constant Output Current Ripple

3.4 BASIC DESIGN CONSIDERATIONS OF POWER PROCESSING STAGES

The stages in the inverter system that handle the power (as opposed to the logic and control stages) directly limit the performance effectiveness in the areas of

- 1) efficiency
- 2) size
- 3) weight
- 4) cost

Additionally the performance capability is determined in the areas of

- 1) DC output
- 2) isolation
- 3) maximum frequency of operation
- 4) ability to regenerate power.

The basic system approach requires conversion of the incoming DC power to variable voltage variable frequency AC power. Thus, the power handling stages must accomplish

- 1) DC - AC inversion
- 2) AC frequency control
- 3) voltage control (either at the AC or DC level)

Figure 3.1 shows the breakdown of the basic systems we are considering (base band, cycloconverter, pulse modulator) into the component stages that accomplish the three functions listed above. In the design of each of these stages we must make two basic choices:

- 1) The configuration of the power switches, and 2) the type of power switch to be used. In addition we must, of course, provide any

auxiliary circuitry necessary for the operation of the switches (e.g. commutation circuitry, di/dt limiting, etc.).

Choice of Switch Configuration

The choice of configuration depends strongly on mission requirements since the performance capability of the inverter depends strongly on the configuration of the power processing stages. These requirements have the following implications:

- 1) DC Output Capability: This may be necessary depending on the motor drive requirements and the method of controlling the motor. (e.g. limited or controlled slip operation may preclude the necessity of DC drive capability). DC capability precludes the use of a transformer unless some modulation takes place after the transformer, as is the case with the cycloconverter.
- 2) Isolation Capability: This requires the use of a power transformer somewhere in the power processing chain, and may be a mission requirement. The weight of the transformer will be significant if the transformer operates at the output frequency. Also, modulation after the transformer is necessary to provide DC capability. Both these difficulties are overcome in the cycloconverter.
- 3) Power Regeneration Capability: This is usually a necessity in servo applications if any reasonable efficiency is to be achieved. Since the cycloconverter requires bilateral switches it is inherently capable of handling regenerated power. All the other power processing stages require that all the switches be made bilateral. This can usually be accomplished simply by paralleling each switch with a diode or providing parallel opposed unilateral switches.

There are really only two basic configurations in use in power processing stages. These are 1) the half bridge, and 2) the push-pull connection. The half bridge is the basic building block for the full bridge and the three phase bridge. The push-pull connection is also sometimes referred to as the "center-tapped load" connection, and generally requires a transformer.

Figure 3.19 shows the basic forms of each of these two circuits. Diodes have been included in order to make the switches bilateral. Thus, both configurations are capable of handling regenerated power. The main differences in capabilities of the two circuits are accounted for by the presence or absence of a transformer. We may summarize them as follows:

Half bridge Connection

will handle DC power

provides no isolation

no transformer required

Push Pull Connection

will not handle DC power

provides isolation

transformer required

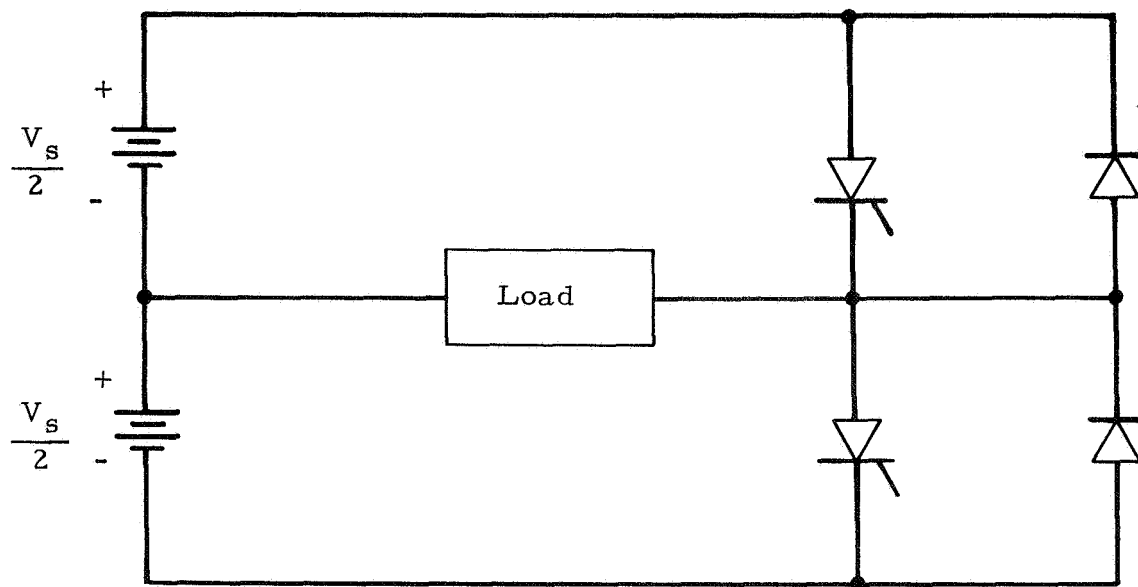
Choice of Power Switch

Currently, the choice of power switch is limited to the transistor and the thyristor. The following are the major factors in the choice.

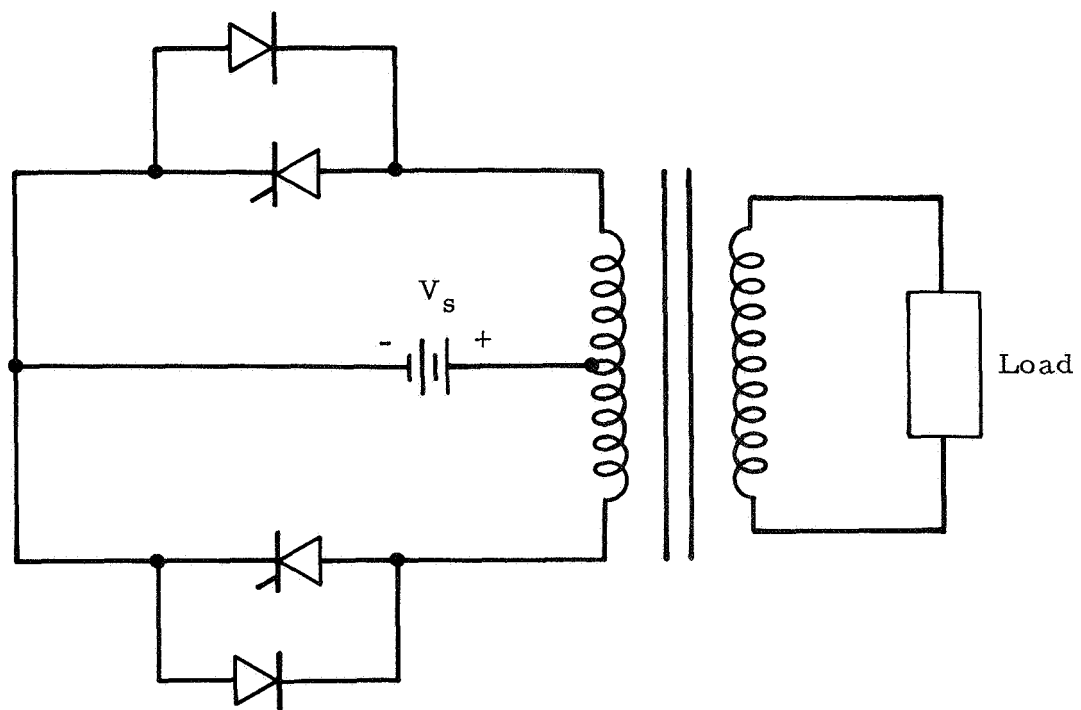
- 1) Maximum current
- 2) Maximum voltage
- 3) Speed (switching time)
- 4) Ease of turn off
- 5) Switching losses

We shall discuss in turn each of these factors.

1) Maximum Current and 2) Maximum Voltage: These two are inter-related since they determine the total power handling capability. Currently, thyristors are available up to about 55 amperes at 1000 volts with a guaranteed maximum turn-off time of 20 micro seconds or less. Units are available to handle currents in excess of 500 amperes but with longer and generally unspecified turn-off times. Transistors are available up to 60 amperes but only to about 300 volts. Higher voltage units (400 volts) are restricted to about 5 amperes.



a) Basic Half Bridge Connection



b) Basic Push-Pull Connection

FIGURE 3.19

BASIC SWITCH CONFIGURATIONS

3) Switching Time - Thyristors generally turn on virtually instantaneously, but turn off slowly. Most of the dynamic losses in thyristor power processing stages occur because of the need to transfer large amounts of energy during the long commutation interval. Turn-off time increases with power handling capability (see Table 3.4) but for the better units is on the order of 10 to 20 micro-seconds. Transistors are generally faster (with perhaps a 2 micro second or less turn-off time) but are not directly comparable since they are not available in the same power ranges as thyristors.

4) Ease of Turn-off - Thyristors have inherently no means for turn-off and thus special commutation circuits must be provided or they must be operated off an AC line (as in the cycloconverter) to provide natural commutation. Transistors are conventionally turned off by reverse biasing the base.

5) Switching Losses - Thyristor turn-on losses are low because of the very fast turn-on. Turn-off losses in the device itself are low since the current goes to zero in a time much less than the turn-off time. Switching losses in thyristor power handling stages occur mainly in the transfer of energy required to force commutation. Transistor losses during switching are appreciably higher since both voltage and current are supported in the device during switching. Nevertheless, the total losses may be less than for force commutated thyristor circuits. Generally the problem of losses during switching is a very complex one. A further exploration of the problem is included in the section on thyristor commutation below.

In general, for the medium to high power levels under consideration only thyristors have adequate voltage and current ratings.

Thyristor Survey

A survey has been made of power thyristors suitable for operation in inverter circuits. The results of the survey, are presented here.

The following parameters are of interest in forced commutation inverters (listed in approximate order of importance):

- a) Maximum RMS forward current ($I_{\text{RMS MAX}}$).

- b) Forward and reverse blocking voltages (PFV, PRV).
- c) Turn-off time (t_{off}).
- d) Maximum allowable dv/dt in the off state.
- e) Maximum allowable di/dt during turn-on.
- f) Forward voltage drop (V_f).
- g) Junction to case thermal resistance.
- h) Maximum operating junction temperature.

The maximum forward current and the peak forward voltage limit directly the average power the device can switch. The turn off time, t_{off} , besides limiting the maximum switching frequency, affects the amount of energy that must be supplied by the commutating circuits during turn-off and thus affects weight and efficiency. Generally dv/dt and di/dt protection can be designed into the circuit as long as the device restrictions are not too severe and are known. V_f limits the static efficiency of the inverter.

Devices are currently available with $I_{\text{RMS MAX}}$ to at least 470 amps and PFV in excess of 1,000 volts. However, the high power devices tend to have longer turn-off times. Since failure of the device to meet its t_{off} rating can result in failure to commute, leading to possible device failure, a guarantee of maximum turn-off time under specified conditions is essential. These factors, combined with the speed and efficiency limitations of slow thyristors, make it appear that the t_{off} maximum rating is at present the determining factor of high frequency performance of thyristor inverters.

The preliminary survey sought to tabulate those thyristors with $I_{\text{RMS MAX}} \geq 5$ amps and for which the manufacturer guaranteed a maximum turn off time less than 40μ seconds. These thyristors and their pertinent characteristics are tabulated in Table 3.4. Figure 3.20 is a plot of $t_{\text{off max}}$ versus the current at which $t_{\text{off max}}$ is specified and versus $I_{\text{RMS MAX}}$.

TABLE 3.4

SURVEY OF POWER THYRISTORS WITH SPECIFIED TURN OFF TIME

Type and Number	Manufacturer	$I_{RMS\ MAX}$ (amps)	t_{off} @ I ($\mu s.$) (amps)	PfV (volts)
40555	R. C. A.	5	6 @ 2	600
C12C	General Electric	6	14 @ 15	300
SCR F 58A	Semicon	7	6 @ 5	800
SCR F 580	Semicon	7	12 @ 5	800
2N3940	Texas Inst.	7 (DC)	8 @ 5	500
2N3272	Fairchild	8 (DC)	22 @ 5	400
2N4198	Motorola	8	25 @ 5	600
MCR 1907-6	Motorola	25	12 @ 10	400
TI 3042	Texas Inst.	25	10 @ 25	500
SCR F 168A	Semicon	35	6 @ 10	800
C141D	General Electric	35	10 @ 10	400
NLC40E	National Electric	35	12 @ 10	500
SCR F 1680	Semicon	35	12 @ 10	800
40216	R. C. A.	35	40 @ 18	600
2191S	Westinghouse	55	15 @ 50	800
2192Z	Westinghouse	55	20 @ 50	1,000
C154E	General Electric	110	10 @ 50	500
NL-C154-E	National Electric	110	10 @ 50	500
2541S	Westinghouse	110	15 @ 50	800

Type & Number	Manufacturer	$I_{RMS\ MAX}$ (amps)	t_{off} @ I ($\mu s.$) (amps)	PFV (volts)
C155E	General Electric	110	20 @ 50	500
NL-C155-E	National Electric	110	20 @ 50	500
2542Z	Westinghouse	110	20 @ 50	1,000
2181S	Westinghouse	200	20 @ 50	800
2182Z	Westinghouse	200	30 @ 50	1,000
C185E	General Electric	235	20 @ 250	500
2201S	Westinghouse	275	30 @ 100	800
2231S	Westinghouse	475	30 @ 100	800

Note: Only the highest voltage devices in a given line are listed.

With regard to achieving the maximum utilization of current thyristors, the following considerations are pertinent:

1) Many SCR's are available with $I_{RMS\ MAX}$ well in excess of those shown in Figure 3.20 and tabulated in Table 3.4. Although these are basically slow devices, manufacturers are generally willing to specially select these devices for $t_{off\ max}$ at some increase in cost.

2) Most of the thyristors with $t_{off\ max}$ already specified can be selected to more stringent specifications by the manufacturer at increased cost.

3) The maximum turn off time increases fairly slowly (in at least some units) with the current in the device just prior to commutation. Thus, it may be possible to use these devices advantageously at peak currents well in excess of their rated $I_{RMS\ MAX}$. This depends on the manufacturer's willingness to specify the $t_{off\ max}$ at these currents.

4) The maximum turn off time, is a strong (increasing) function of junction temperature. It is possible that guaranteeing a junction temperature below the maximum operating temperature would result in considerably improved performance.

Inverter Output Stages Using Thyristors: Minimization of Losses in Commutation.

The output stages in the literature that employ thyristors can be divided into two groups: those that employ line commutation (the cycloconverter) or resonant commutation, and those that use forced commutation. Self commutation schemes achieve high reliability by guaranteeing thyristor turn-off, thus helping avoid a possibly destructive "on-on" condition in the output stage. However, the cycloconverter requires a high frequency inverter before the output stage to generate AC from the DC line. The resonant commutation schemes do not inherently provide for handling regenerated power, nor do they provide the

fast response of the forced commutation schemes. Both commutation schemes rely on a guaranteed maximum turn-off time of the thyristor for avoiding the "on-on" condition.

The study which is outlined below, is part of an attempt to analyze carefully the fundamental limitations of various schemes described in the literature for forced commutation of thyristors. The goal is to meaningfully characterize the circuits in terms of energy loss as a result of the forced commutation and to determine the minimum losses for worst case load (an infinite inductor) and for optimum choice of component values and operating conditions, and also to determine the practicality of employing this optimum circuit in terms of required component sizes and device limitations. This analysis is not available in the literature, although Humphrey made a general analysis and comparison of these schemes. He did not, however, optimize the circuit performance prior to making his comparisons.⁴

Minimization of Trapped Energy

Energy loss during switching forced commutation inverters is primarily dependent on the energy required to force the commutation of the thyristors. Although in principle it requires no energy to commute a thyristor (since the voltage or current is reduced to zero instantaneously), in practice commutation is achieved by energy transfer between inductors and capacitors in order to momentarily reverse bias the thyristor.

This energy transfer may occur in such a way that the energy is ultimately delivered to the load or is used to reinitialize a capacitor for a succeeding commutation. In some cases, however, the energy is "trapped" in an inductor and it must be dissipated or special transformers must be provided to return this energy to the supply. Since the size and weight of these added components are proportional to the amount of energy they must handle, the approach taken here is to optimize the performance of various basis commutation circuits by minimizing the trapped energy.

The basic inverter output stage chosen for analysis is the half bridge configuration. This basic configuration, less commutating circuitry, is shown in Figure 3.19a.

Comparison of Three Commutation Schemes

Three forced commutation circuits have been analyzed. The first, which we may call the "transformer commutated inverter" is shown in Figure 3.21 the second, the "transformer commutated inverter with capacitor help", is shown in Figure 3.22. The means for charging the commutation capacitors C and C' and the trigger circuits are not shown. Figure 3.23 gives an example of a means for recovering trapped energy. The third, the McMurry circuit is shown in Figure 3.24.

The analysis performed is the worst case analysis with the load assumed to be an infinite inductor so that the output current is fixed. In addition, we assume that the thyristor stops conducting as soon as it is reversed biased and that we can ignore reverse current flow. We require that the voltage across the thyristor being commutated remain negative for at least $t_{off\ max}$ after the start of commutation. Also, we can, without loss of generality, assume a turns ratio of 1:1 since the transformer action conserves energy. (Variation of the turns ratio N allows a trade off between capacitor size and capacitor voltage. The reflected capacitor voltage is Nv_C , the reflected capacitor size is C/N^2 .)

The trapped energy minimization was carried out in detail. The results are given in Table 3.5. Table 3.6 gives typical numerical examples. In the case of the McMurry circuit, which is not discussed in the previous report,⁵ it can be shown that under the conditions assumed, the trapped energy is zero. In this case, the values presented in Tables 3.5 and 3.6 are derived from the results of Bedford and Hoft and represent the condition of minimization of energy transferred.⁶ This minimization is carried out under the same conditions used for the first two circuits, namely ideal components (i. e. infinite Q resonant circuits) and a fixed load current. It appears that the McMurry circuit is an attractive solution to the problem of switching losses during commutation. However, it remains for more careful theoretical work plus experimental verification to determine if the analysis is valid, and the extent to which second order effects (inductor resistance, diode forward voltage, etc.) will increase losses.

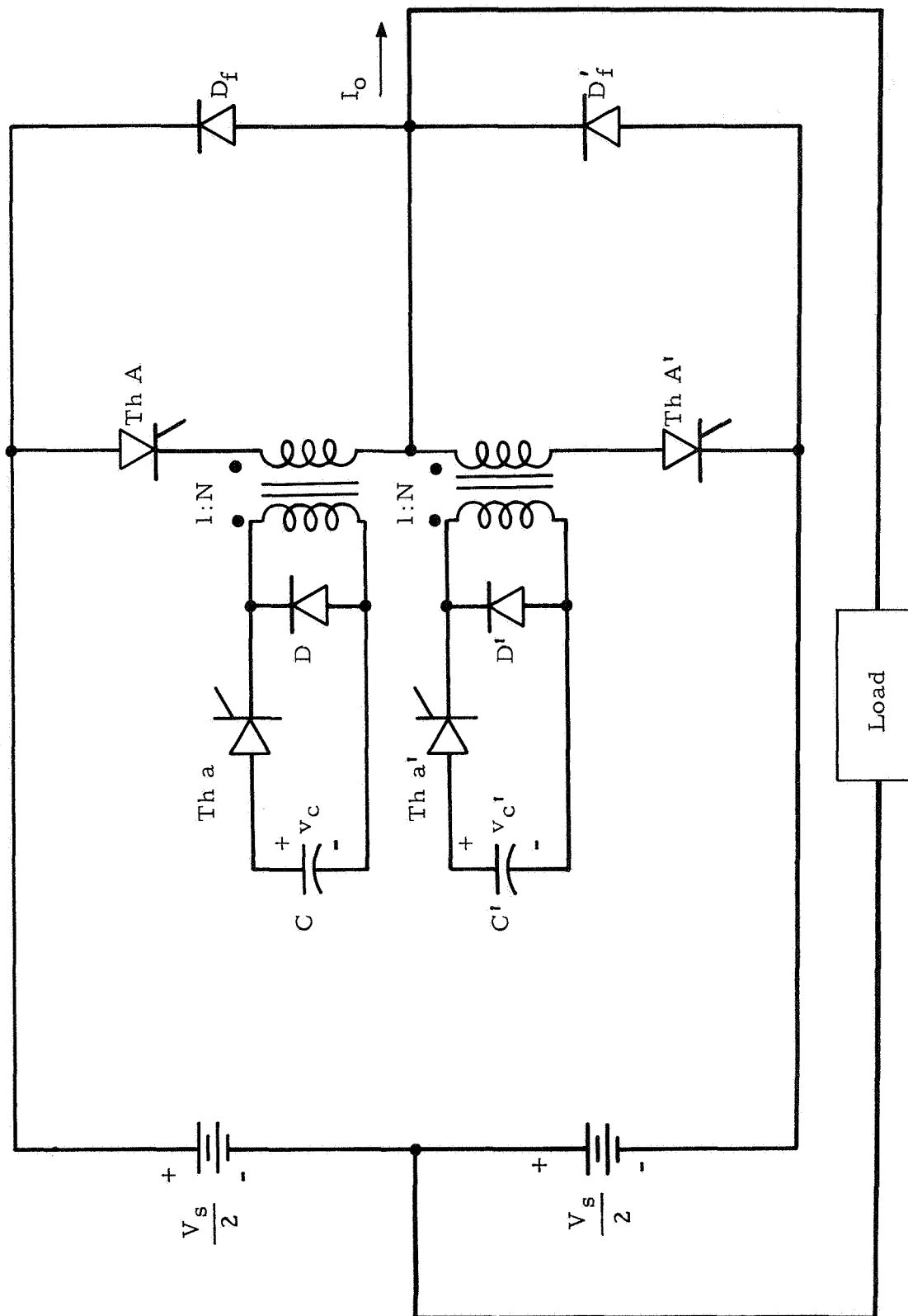


FIGURE 3.21
TRANSFORMER COMMUTATED INVERTER

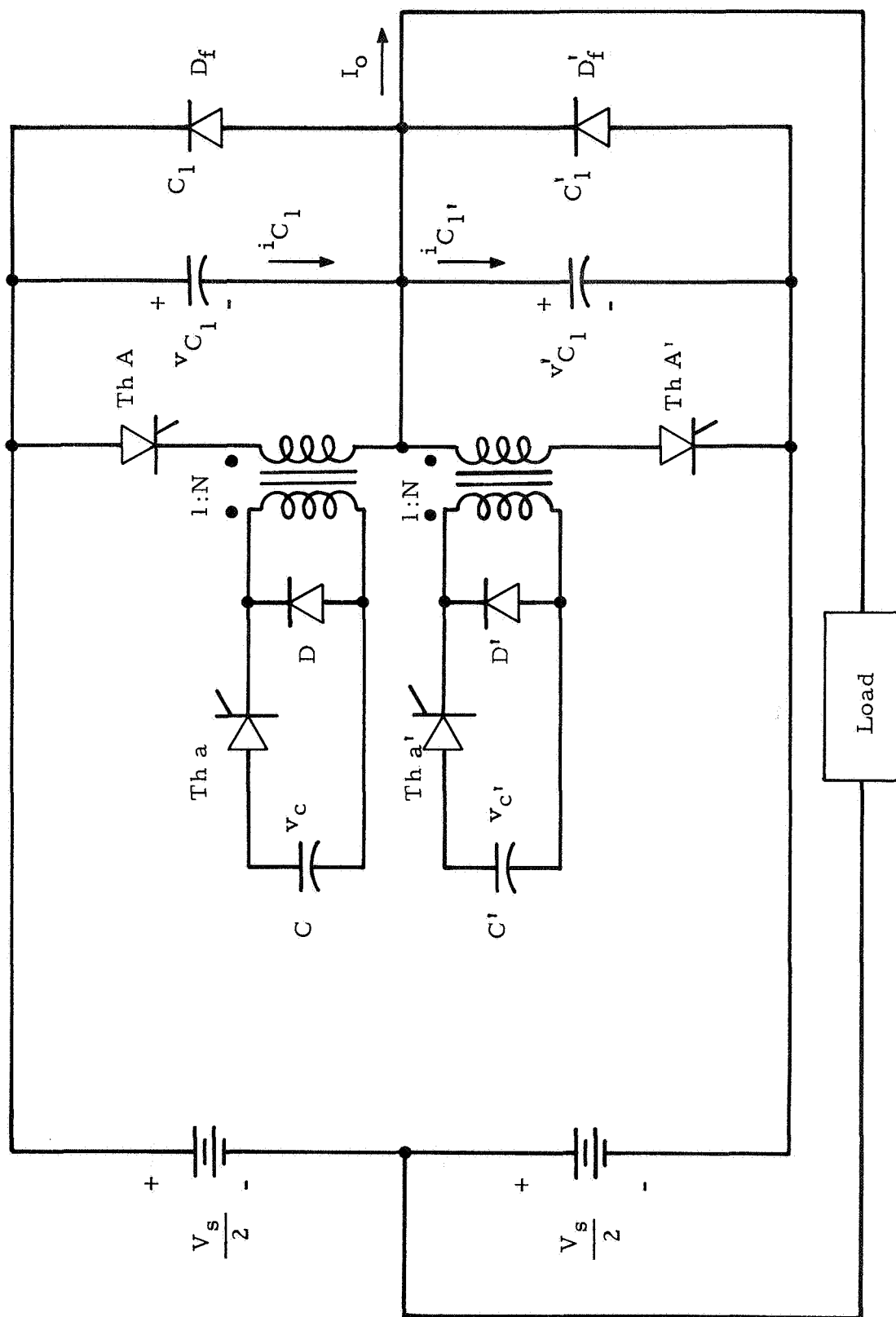


FIGURE 3.22

TRANSFORMER COMMUTATED INVERTER
WITH CAPACITOR HELP

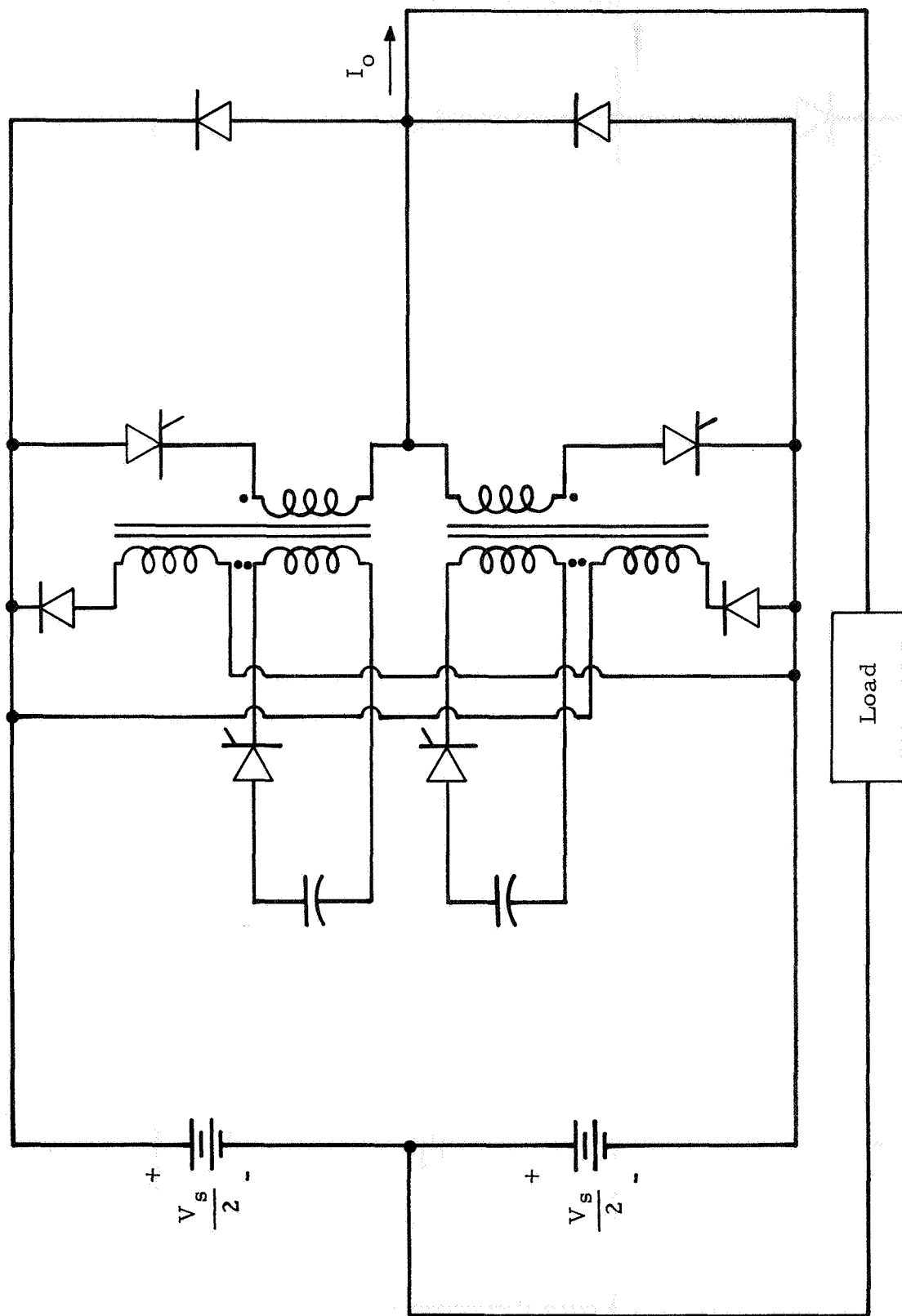


FIGURE 3.23

INVERTER OF FIGURE 3.22 WITH MEANS FOR
RECOVERING TRAPPED ENERGY

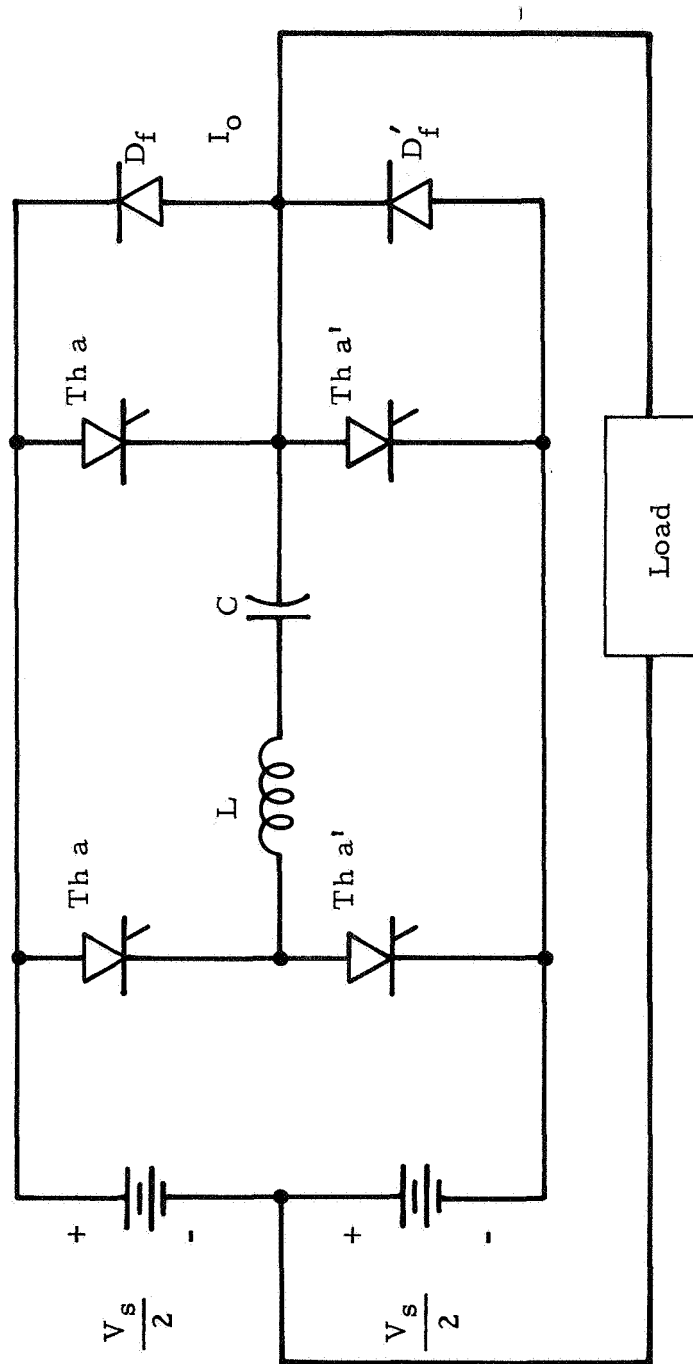


FIGURE 3:24

McMURRAY CIRCUIT

Table 3.5

TABULATION OF OPTIMUM VALUES		
Notation:		
V_s	=	Supply Voltage
I_o	=	Maximum Output Current
τ	=	Maximum Turn-Off Time of Thyristor
K	=	$\omega\tau$ where ω is the Resonant Frequency of the Commutating Circuit
E	=	Trapped Energy
C	=	Commutating Capacitor
C_l	=	"Helping Capacitor"
L_m	=	Magnetizing Inductance of Commutating Transformer
L	=	Commutating Inductor
V_c	=	Voltage on Commutating Capacitor Just Prior to Commutation
E_c	=	Energy Stored in Capacitor C
Q_c	=	Charge Stored in Capacitor C

Table 3.5 (Continued)

Transformer Commutated	Transformer Commutated	McMurry Circuit
$K \approx 0.5$	$K \approx 0.7$	-----
-----	$m \approx 0.2$	-----
$E \approx 3.85 V_s I_o \tau$	$E \approx 3.8 V_s I_o \tau$	$E \approx 0$
$C \approx \frac{2 I_o \tau}{V_s}$	$C \approx \frac{7.44 I_o \tau}{V_s}$	$C \approx \frac{2.35 I_o \tau}{V_s}$
-----	$C_1 \approx 0.2 C_{opt}$	-----
$L_m \approx \frac{2 V_s \tau}{I_o}$	$L_m \approx \frac{0.27 V_s \tau}{I_o}$	$L \approx \frac{2.35 V_s \tau}{I_o}$
$V_c \approx 1.69 V_s$	$V_c \approx 0.6 V_s$	$*V_{c \min} \approx V_s$
$E_c \approx 2.86 V_s I_o \tau$	$E_c \approx 1.34 V_s I_o \tau$	$*E_{c \min} \approx 1.18 V_s I_o \tau$
$Q_c \approx 3.38 I_o \tau$	$Q_c \approx 4.46 I_o \tau$	$*Q_c \approx 2.35 I_o \tau$

* These are given as minimum values since over charging is possible under some conditions.

Table 3.6

NUMERICAL EVALUATIONS FOR		
$V_s = 300 \text{ Volts}$ $I_o = 40 \text{ Amps}$ $\tau = 20 \mu\text{seconds}$		
Transformer Commutated	Transformer Commutated With Capacitor Help	McMurry Circuit
$E \approx 0.93 \text{ joules}$	$E \approx 0.91 \text{ joules}$	$E \approx 0.0 \text{ joules}$
$C \approx 5.32 \mu\text{farads}$	$C \approx 19.8 \mu\text{farads}$	$C \approx 6.25 \mu\text{farads}$
$L_m \approx 300 \mu\text{henries}$	$L_m \approx 40 \mu\text{henries}$	$L_m \approx 350 \mu\text{henries}$
$V_c \approx 500 \text{ volts}$	$V_c \approx 180 \text{ volts}$	$V_{c \text{ min}} \approx 300 \text{ volts}$
$E_c \approx 0.67 \text{ joules}$	$E_c \approx 0.32 \text{ joules}$	$E_{c \text{ min}} \approx 0.28 \text{ joules}$
$Q_c \approx 2.7 \times 10^{-3} \text{ coulombs}$	$Q_c \approx 3.6 \times 10^{-3} \text{ coulombs}$	$Q_{c \text{ min}} \approx 1.8 \times 10^{-3} \text{ coulombs}$

4. PARAMETER CONTROL

When a complete inverter system is mated with an induction motor, control and feedback circuitry must be added to complete the inverter-motor system. The parameters for which control might be provided are those at the interface between the motor and inverter (voltage, current, motor flux, frequency) and those at the output of the motor (torque, speed, shaft position). The interface parameters must be controlled to effect proper motor control and to prevent damage to the inverter in case of a line or motor fault or of an abnormal load condition which reflects an abnormal load impedance to the inverter. The output parameters must be controlled to meet the mission requirements of the system. This section will present representative examples of the techniques employed in realizing the necessary control functions.

4.1 VOLTAGE CONTROL

The inverter system provides control of both the external AC output voltage to meet the motor demands and internal control of the instantaneous voltages applied to the output stage switches for protection against breakdown. Internal voltage control is dependent upon the details of circuit design and will not be discussed here. For the purpose of examining external control, a basic inverter may be modeled as a sine wave source whose frequency output is proportional to one control input, S_f , and whose amplitude is proportional to another control input, S_v , and the supply voltage, V_s , (this is an unregulated inverter) as diagrammed in Figure 4.1.

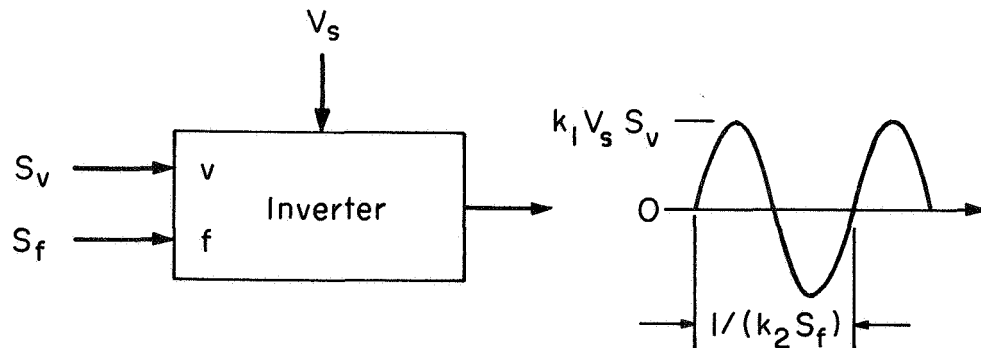


FIGURE 4.1

MODEL FOR OPEN LOOP CONTROLLED INVERTER

Thus, ignoring spurious frequency components

$$v_o(t) = k_1 S_v V_s \sin (2\pi k_2 S_f t) \quad (4.1)$$

where k_1 and k_2 are scaling constants. For a multi-phase system, additional outputs not shown in the figure provide the appropriate signals.

This inverter provides open loop control of the output amplitude. If the supply voltage regulation is poor, the output amplitude will vary. Amplitude control feedback may be employed to stabilize the output against supply variations, as shown in Figure 4.2.

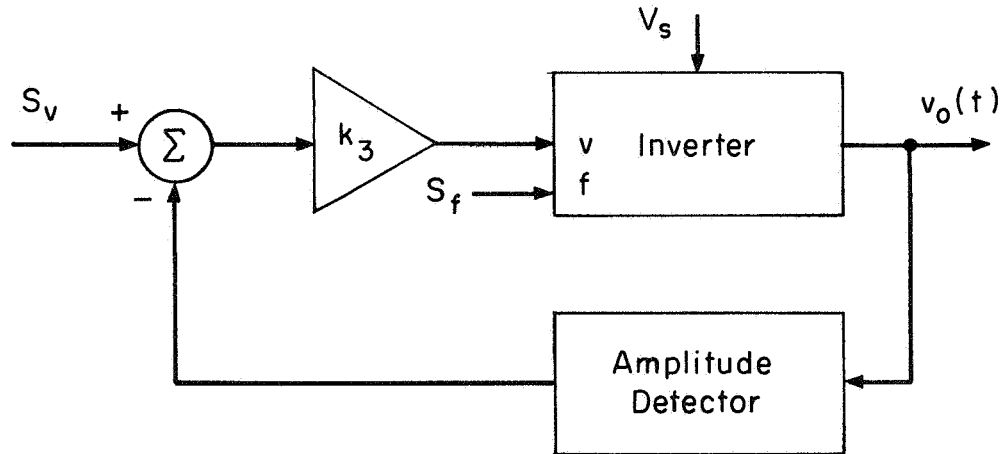


FIGURE 4.2

CLOSED LOOP AMPLITUDE CONTROL

With this modified system

$$v_o(t) = \frac{S_v}{\left(1 + \frac{1}{k_1 k_3 V_s}\right)} \sin 2\pi k_2 S_f t. \quad (4.2)$$

If $k_1 k_3 V_s \gg 1$ then the output amplitude will be substantially independent of supply variation.

The method of Figure 4.2 requires an amplitude detector circuit to provide the feedback necessary for control. This necessarily makes operation of the control loop slow compared to the period of the output sine wave. If a greater bandwidth is required, a form of instantaneous control of the output voltage must be employed: a scheme which matches the output signal to the reference signal on a sample by sample basis, thereby eliminating the effects of power source variations. This, of course, restricts the choice of inverter systems. The Two-State Modulation System, shown in Figure 4.3, is an example of an inverter system which accomplishes this task.

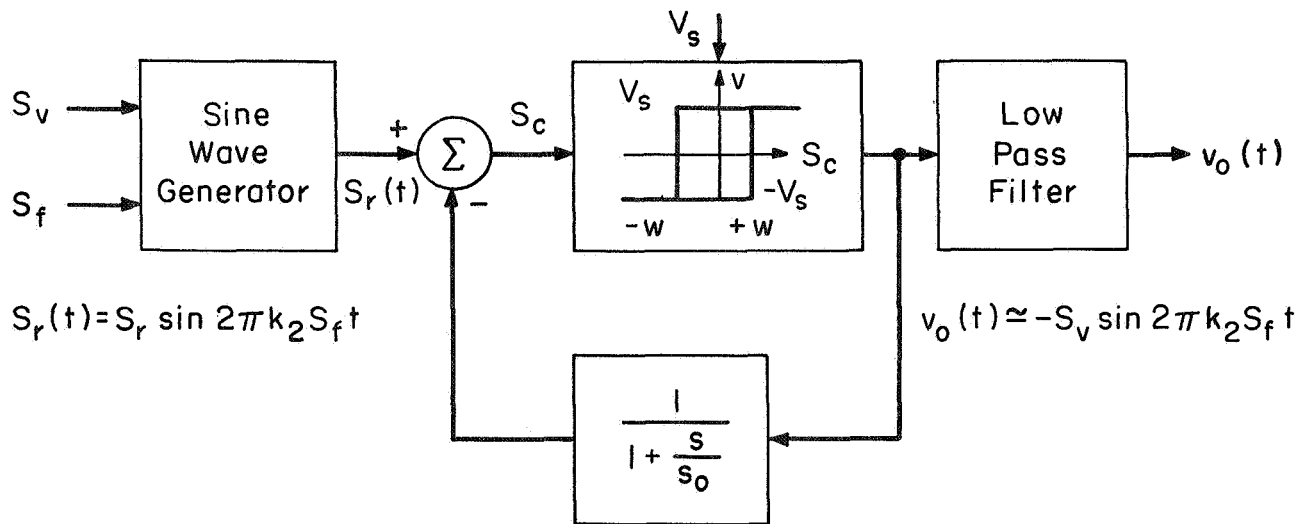


FIGURE 4.3

INSTANTANEOUS VOLTAGE CONTROL WITH THE TWO-STATE MODULATOR

The output switching waveform is passed through a low pass filter with a break frequency S_a which may be large when compared to the output frequency $\omega_f = 2\pi k_2 S_f$. This filtered waveform is compared to a reference sine wave and the resultant error is used to actuate a hysteretic switch, thereby generating the modulation. The low pass filtered output $v_o(t)$ has been shown to closely approximate the reference sine wave, $S_r(t)$, if the system parameters are chosen properly, even in the presence of transients in the source voltage, V_s .

4.2 CURRENT CONTROL

It is often desirable to control the current output of an inverter system instead of, or in addition to, the voltage output. As in voltage control, either the average or the instantaneous current may be controlled. In Figure 4.4a, a controlled voltage inverter system similar to that of Figures 4.1 to 4.3 is adapted to current control by the use of a current feedback loop which servos the voltage amplitude control input. Operation is similar to that of the controlled voltage inverter of Figure 4.2 and the limitation on transient response time due to amplitude detector response is the same. The controlled current inverter may be modelled as a single block, as shown in Figure 4.4b. (A controlled current inverter may be made controlled voltage by the addition of the voltage servo loop of Figure 4.2.)

For some systems, continuous control of the output current may not be necessary, but rather current limiting may be adequate. With a controlled voltage inverter, this may be accomplished in the manner shown in Figure 4.5.

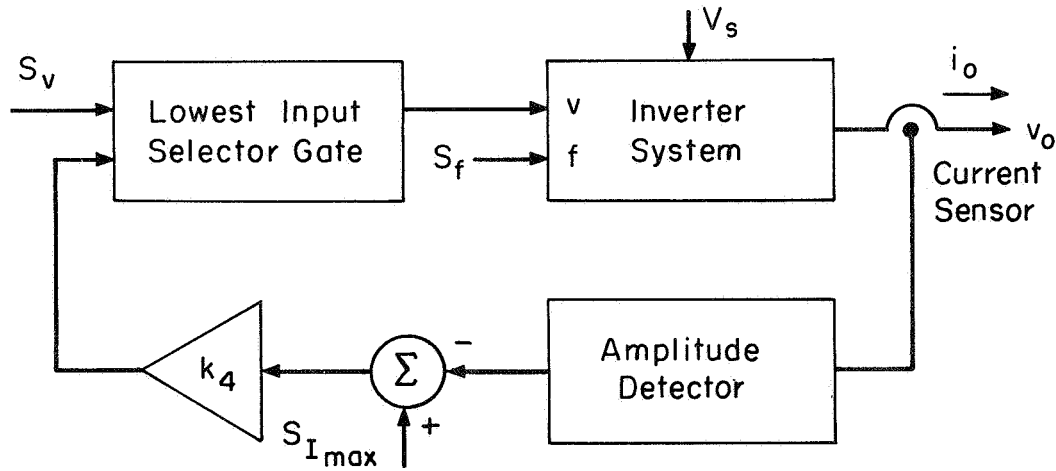


FIGURE 4.5

CONTROLLED VOLTAGE INVERTER WITH LINEAR CURRENT LIMITING

The amplitude of AC voltage output, v_o , will be proportional to the voltage control signal, S_v , unless the output current, i_o , exceeds the reference value of $S_{I_{max}}$.

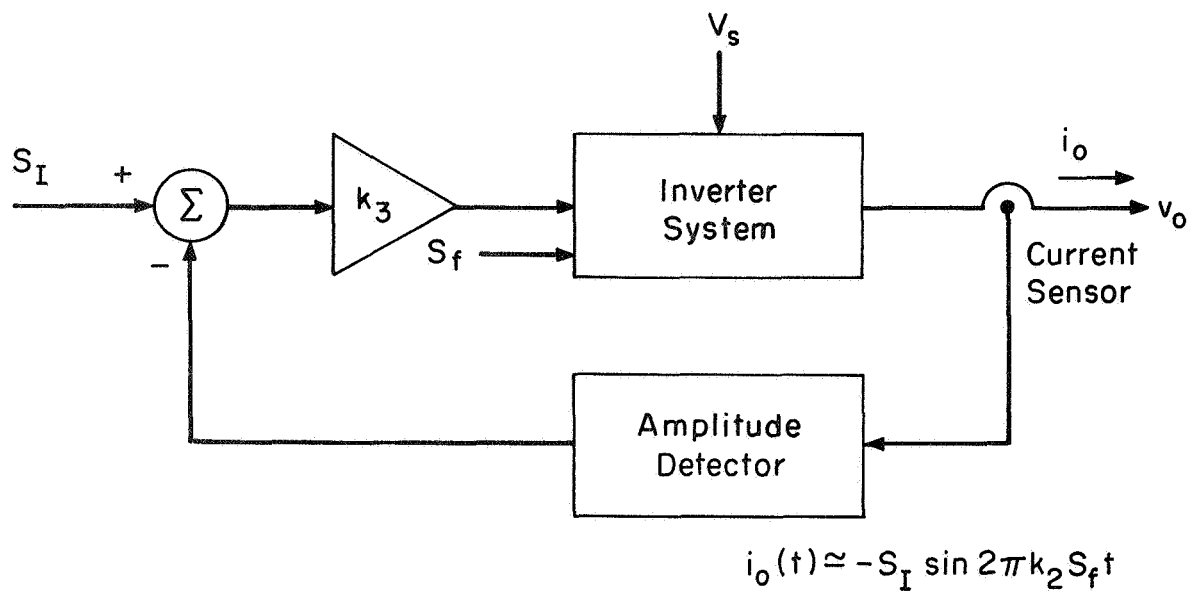


FIGURE 4. 4a

CONTROLLED VOLTAGE INVERTER WITH LINEAR
CURRENT FEEDBACK

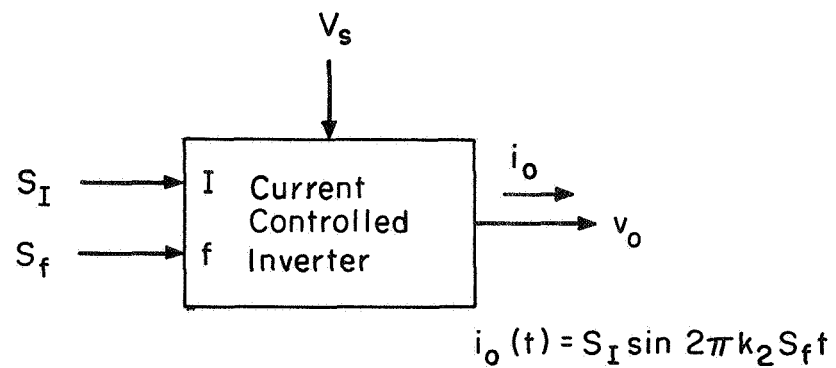


FIGURE 4. 4b

MODEL OF CONTROLLED CURRENT INVERTER SYSTEM

The current limit loop of Figure 4.5 and current control scheme of Figure 4.4 operate relatively slowly under transient conditions as previously explained. These control schemes will not be adequate to protect solid state inverter components from damage due to line faults. Fusing or other protection means may be provided, or the inverter system may be designed using an instantaneous current control method. One such method is a Two-State current control scheme block diagrammed in Figure 4.6. If the load is inductive at the switching frequencies, the instantaneous current will be linearly rising or falling depending on the state of the hysteretic switch. A sensed replica of the current is compared to the reference sine wave S_L and the difference signal between the two, S_e , used to actuate the hysteretic switch. The resultant modulation produces an output current whose average value follows the reference sine wave and whose current ripple has a constant peak to peak amplitude related to the width of the hysteresis. If the reference sine wave is passed through an instantaneous signal amplitude limiter, then the output current will be limited to an average value corresponding to the limiter threshold, $S_{s\max}$, in Figure 4.6. Further, since the current ripple is controlled, the instantaneous peak current (ignoring the effects of delays or circuit operation) is limited. The output devices in the inverter are thus protected on an instantaneous basis against over-current for all static and dynamic load conditions.

Instantaneous current limiting rather than continuous control may be provided by a feedback technique similar to Two-State current control. This technique, shown in Figure 4.7, allows any control concept to be used under normal operating conditions. If the output current exceeds a predetermined threshold set by the current threshold block, Two-State current control operation overrides the normal control and limits the average and peak current to the threshold values. When the output current falls below the maximum, normal operation resumes. Since the modulator and current limiter operate independently in frequency, suitable interlocks must be provided in the hysteretic switch, if necessary, to prevent the switch from changing state before the output stage commutating circuits have fully recovered.

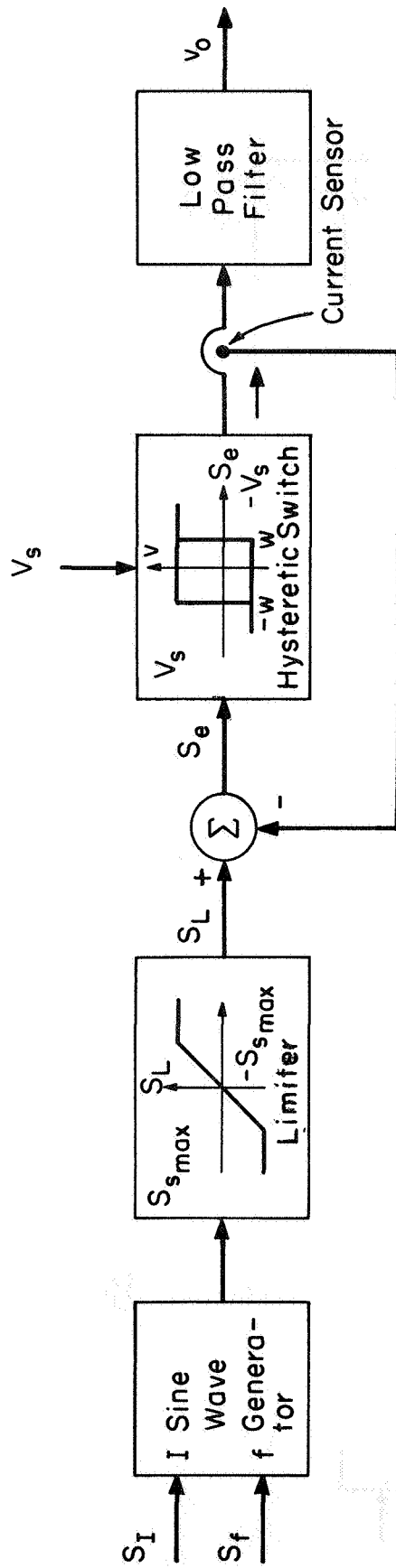


FIGURE 4.6

CONTROLLED CURRENT INVERTER EMPLOYING TWO-STATE CURRENT CONTROL TECHNIQUES

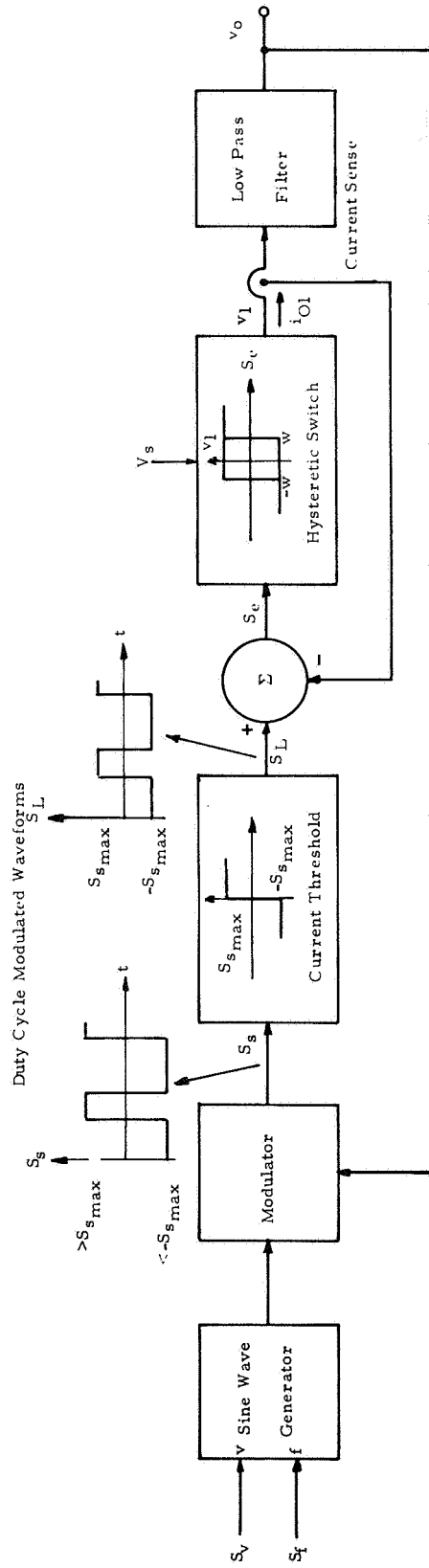


FIGURE 4.7

CURRENT LIMITING EMPLOYING TWO-STATE CURRENT CONTROL

4.3 FLUX CONTROL

All of the servo techniques discussed under voltage and current control may be modified to control the flux of the induction machine inverter load by the addition of a flux sensor. As an example, Figure 4.8 shows a method of direct flux control using a controlled voltage inverter. The output of the flux sensor is conditioned by an amplifier and the amplifier signal is fed to an amplitude detector. The amplitude signal is compared to a limited flux reference signal S_L . The difference is used to servo the voltage control input of the inverter to minimize the flux error. Limiting the flux reference signal limits the maximum flux, thereby preventing saturation of the motor.

If the induction motor to be used with the system is well characterized, direct flux control may be unnecessary. By examination of the equations governing induction motor operation, it can be shown that it is possible to infer the state of flux of the motor by sensing of the electrical variables. To illustrate this point, refer to the model of the induction motor in Figure 2.1, a portion of which is repeated in Figure 4.9.

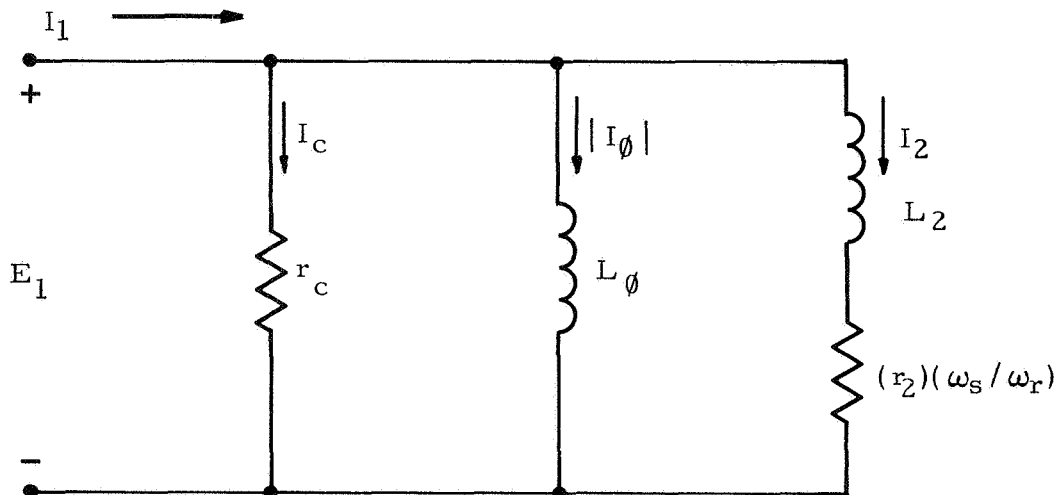


FIGURE 4.9 - INDUCTION MOTOR MODEL

$$\left[\text{Note: } |I_\phi| = |E_1| / (L_\phi \omega_s), \phi_m = E_1 / (N \omega_s) \right]$$

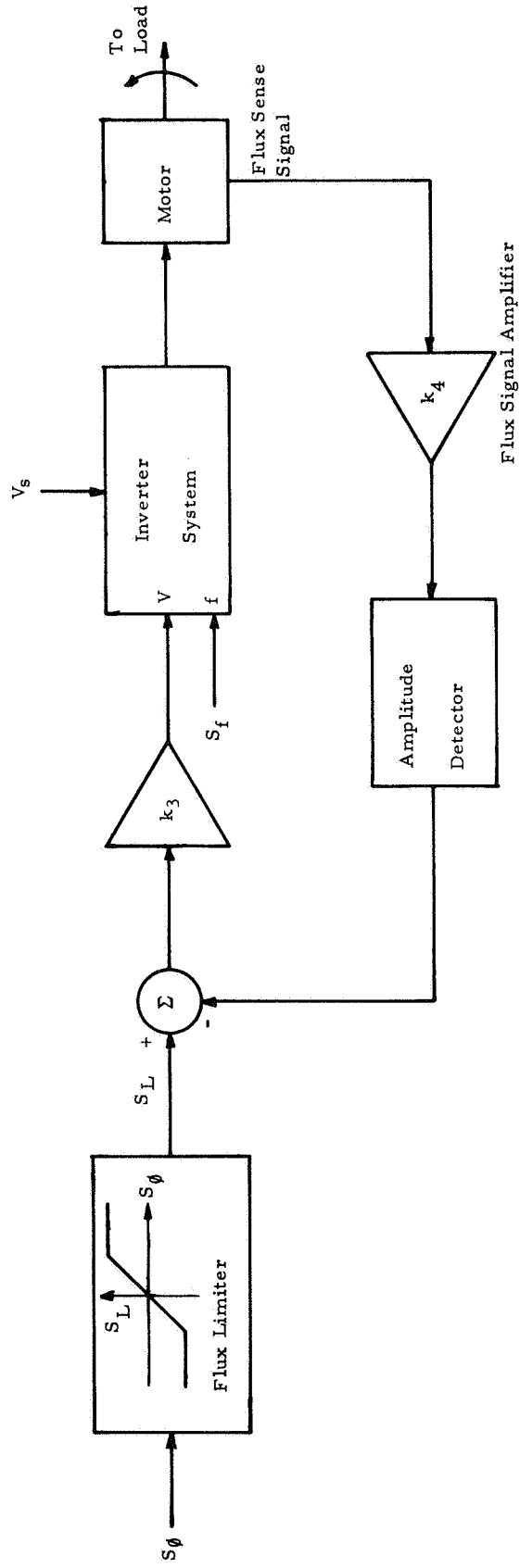


FIGURE 4.8

DIRECT FLUX CONTROL WITH A CONTROLLED VOLTAGE INVERTER

The stator current, I_1 , can be resolved into three components:

1) Core losses, modeled as a non-linear resistor whose resistance is a function of the excitation voltage, E_1 , and excitation frequency, ω_s , resulting in current, I_C .

2) Magnetizing current, I_ϕ , modeled as an inductor, L_ϕ . The magnetizing current is assumed to be directly proportional to the flux, ϕ_m , in the motor.

3) Rotor current, I_2 , which contributes to both motor output power and motor losses.

Ignoring core losses, Figure 4.9 shows that

$$I_1 \approx I_\phi + I_2 \quad (4.3)$$

If the flux is to be held constant, I_ϕ must be held constant. Since every term in the impedance expression for the magnetizing and motor legs of the model contains ω_s , the total current, I_ϕ , may be expressed in terms of I_ϕ independently of ω_s :

$$|I_1| = |I_\phi| \left| \frac{j(L_2 + L_\phi) + \frac{r_2}{\omega_r}}{jL_2 + \frac{r_2}{\omega_r}} \right| \quad (4.4)$$

L_2 , L_ϕ and r_2 are constants of the motor. The parameter, ω_r , is the slip frequency of the induction machine. As discussed in Section 2.1, ω_r is an important control parameter in a controlled slip drive system. In Figure 4.10, $\left(\frac{I_1}{I_\phi}\right)$, given by Eqn. 4.2 is plotted versus ω_r for a typical induction motor.

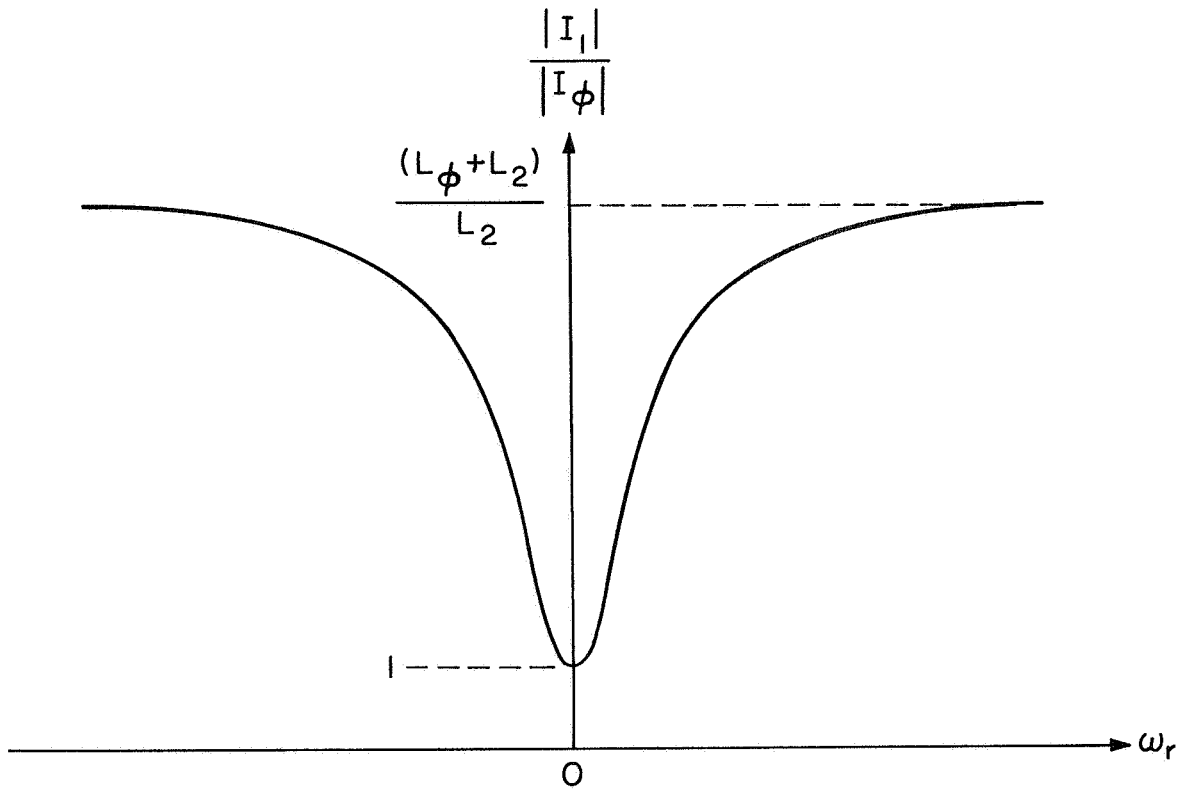


FIGURE 4.10

STATOR CURRENT vs. SLIP FREQUENCY FOR CONSTANT FLUX

By inspection of the figure, it can be seen that there is a unique correspondence between I_ϕ (hence ϕ_m) and I_1 for every value of ω_r . If I_1 is modulated with ω_r according to Figure 4.10, then the flux will be kept constant. Figure 4.11 illustrates a system which accomplishes this task. Further compensation dependent on the drive frequency, ω_s , and the desired flux level, I_ϕ , can be added to compensate for core loss effects. Alternatively, the flux level can be fixed to the desired flux at maximum stator excitation frequency. As the core loss decreases at lower frequencies, the motor will be slightly overfluxed, but the overall core losses will still be less than those at the highest frequency.

4.4 TORQUE CONTROL

The torque equation of the induction motor, Eqn. 2.6, repeated here for convenience, indicates the manner in which an inverter system might be operated to control torque.

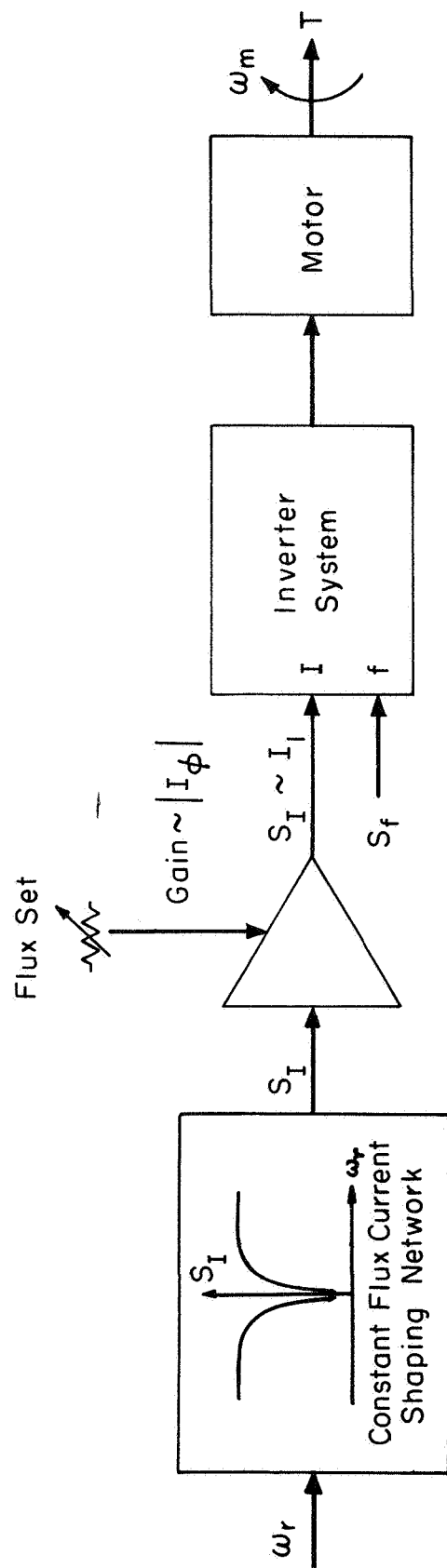


FIGURE 4.11

INDIRECT FLUX CONTROL BY CURRENT SHAPING

$$T = \frac{E_1^2}{\omega_s} \cdot \frac{1}{L_2} \cdot \frac{\tau \omega_r}{1 + \tau^2 \omega_r^2} . \quad (2.6)$$

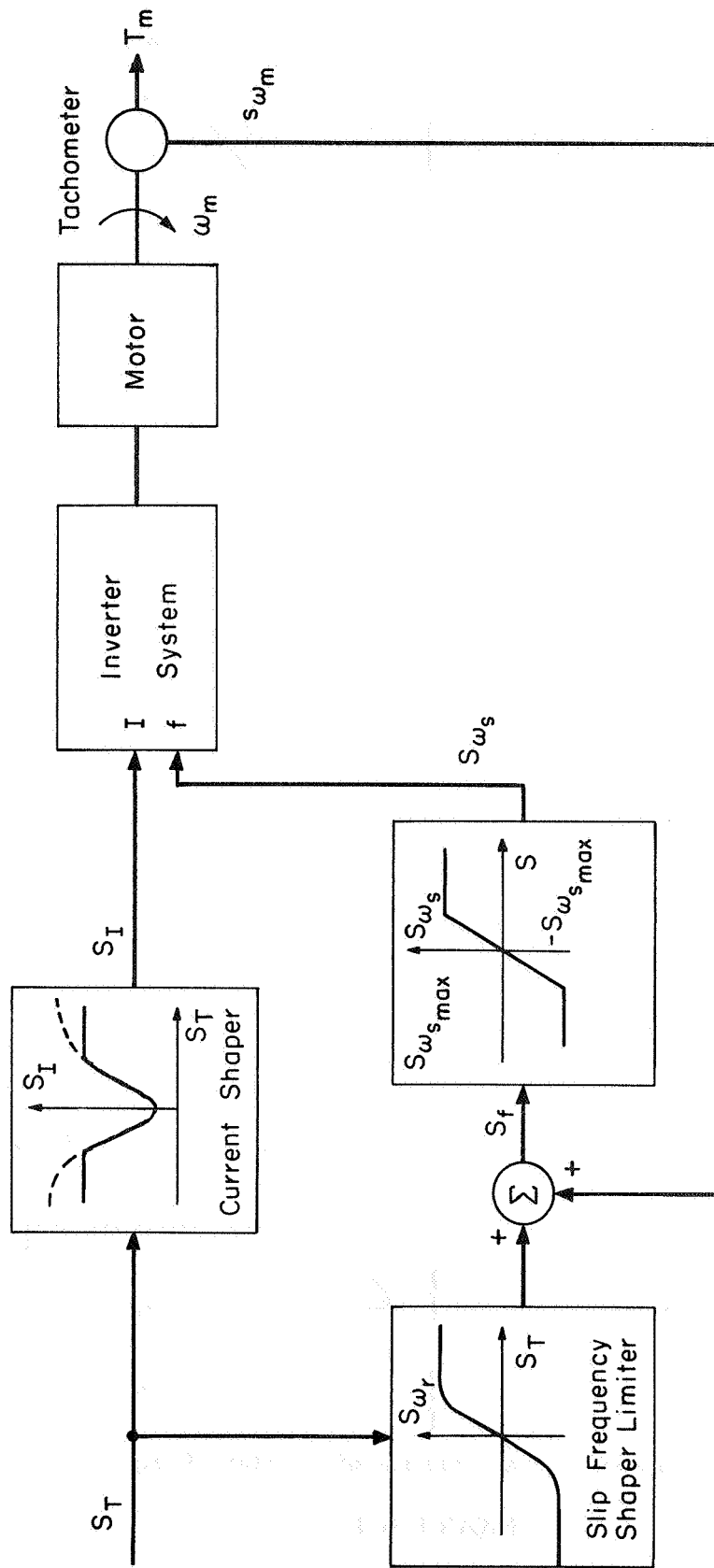
If $\frac{E_1}{\omega_s} \approx \phi_m$ is held constant, then the torque is dependent only on ω_r , the difference frequency between the stator applied electrical frequency and the rotor mechanical frequency. Eqn. 2.6 has been plotted in Figure 2.3. As is evident, if

$$-\frac{1}{\tau} < \omega_r < \frac{1}{\tau} , \quad (4.5)$$

the entire range of possible torque is available. Figure 4.12a shows a system which implements Eqn. 2.6 to control the torque. The slip frequency shaper-limiter restricts operation to the region specified by Eqn. 4.5. The stator frequency limiter prevents overspeed operation and the current shaper provide the constant flux required independent of torque. Figure 4.12b shows the idealized static transfer function of the system and Figure 4.12c shows the corresponding torque-speed output characteristics with the input S_T as a parameter.

4.5 SPEED CONTROL

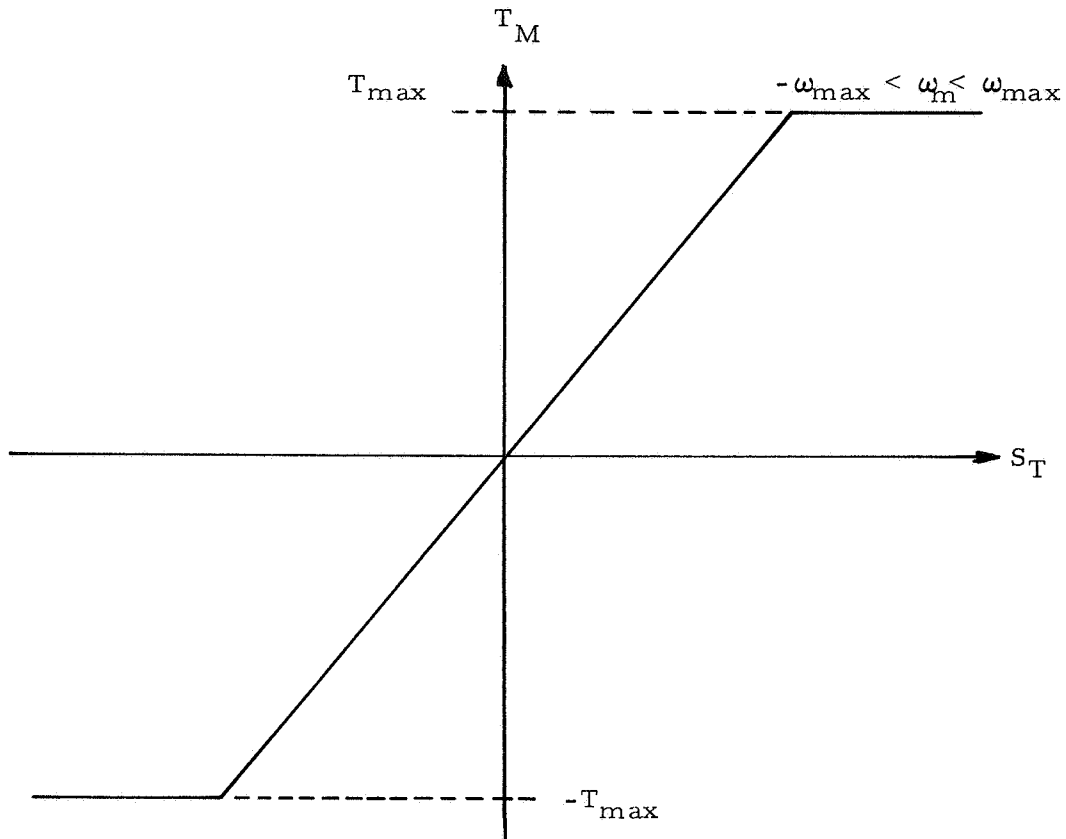
The usual application of variable speed induction motor drive systems is for control of the motor shaft speed in response to a speed control signal in the face of varying load torques and load inertia. Depending on the degree of control required; i.e. permissible speed error and dynamic response, increasingly refined systems of control may be employed. It is significant that all of these systems employ substantially the same inverter system and the same motor. For systems of appreciable power level, these two items dominate in the size, weight and cost of the system. The systems presented here differ only in the configuration of the circuitry used to generate the control signals for the inverter system, a relatively minor part of the total system.



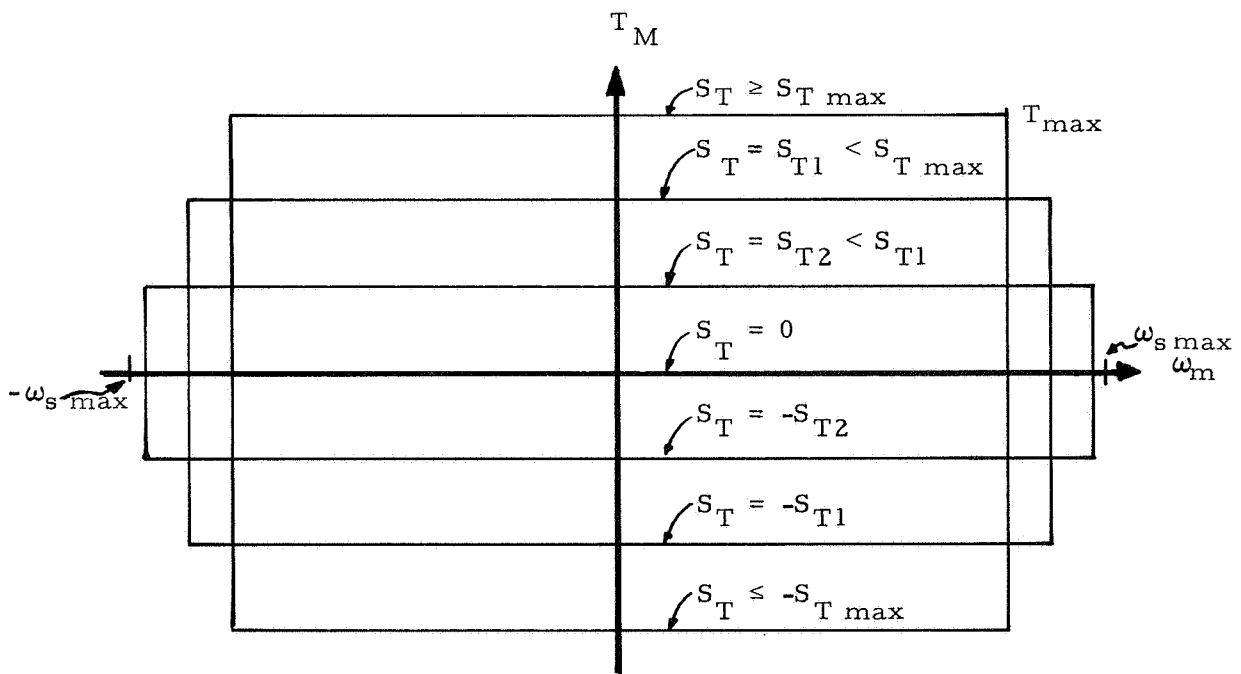
a) Block Diagram of Constant Torque System

FIGURE 4.12

BLOCK DIAGRAM AND CHARACTERISTICS OF CONSTANT TORQUE SYSTEM



b) Transfer Characteristics of Constant Torque System



c) Torque-Speed Characteristics of Constant Torque System

FIGURE 4.12 (cont.)

Open Loop Speed Control

The simplest drive to be considered is an open loop system shown in Figure 4.13. This system provides voltage amplitude control to the motor with a constant volts/Hz dependence with drive frequency. At low frequencies, where the stator resistance is dominant, constant voltage drive is provided. To prevent the motor from operating at beyond breakover torque during acceleration, a rate of change of speed limiter is included to limit acceleration of speed signal to a range less than the acceleration of the load. The static motor load torque must also be less than the breakover torque of the motor. If the motor load can be expressed as

$$T_L + J_L \frac{d\omega_m}{dt} \quad (4.6)$$

then these restrictions may be written as

$$T_{L \max} \leq T_{\max} \quad (4.7)$$

$$\frac{dS\omega_m}{dt} \leq \frac{T_{\max} - T_{L \max}}{J_{L \max}} \quad (4.8)$$

where T_L is the load torque

$T_{L \max}$ is the maximum T_L

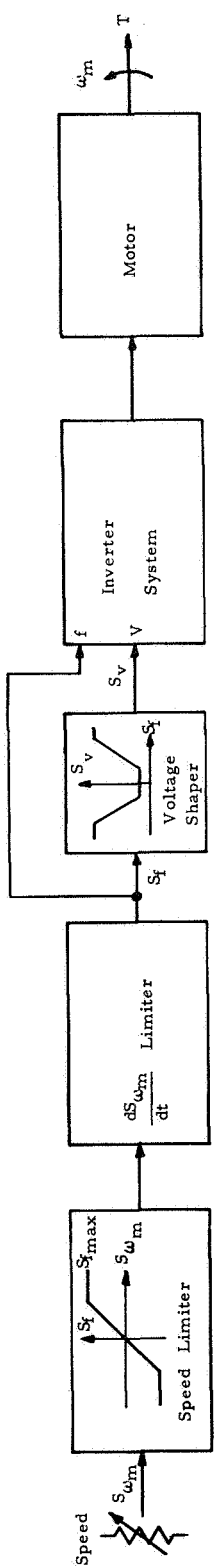
$T_{\max}$ is the breakover motor torque

J_L is the load moment of inertia

$J_{L \max}$ is the maximum J_L

$S\omega_m$ and ω_m as defined in Figure 4.13.

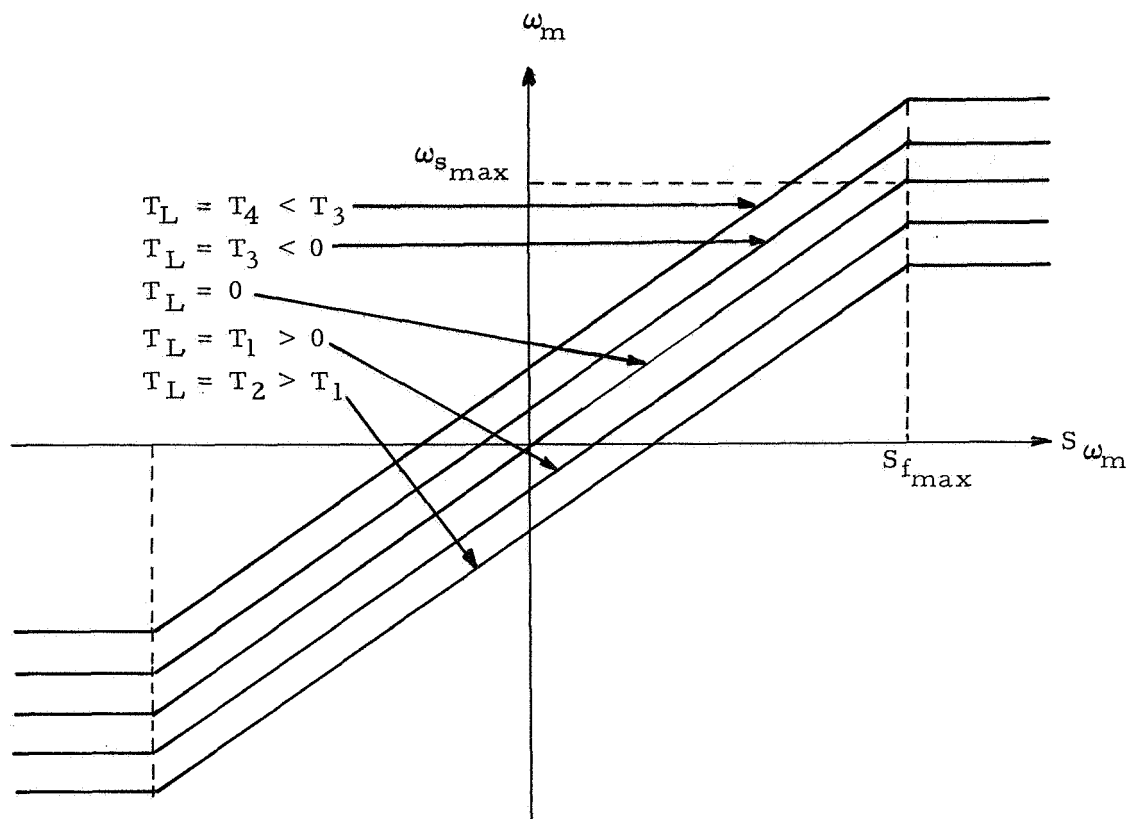
The open loop speed control system has several disadvantages.



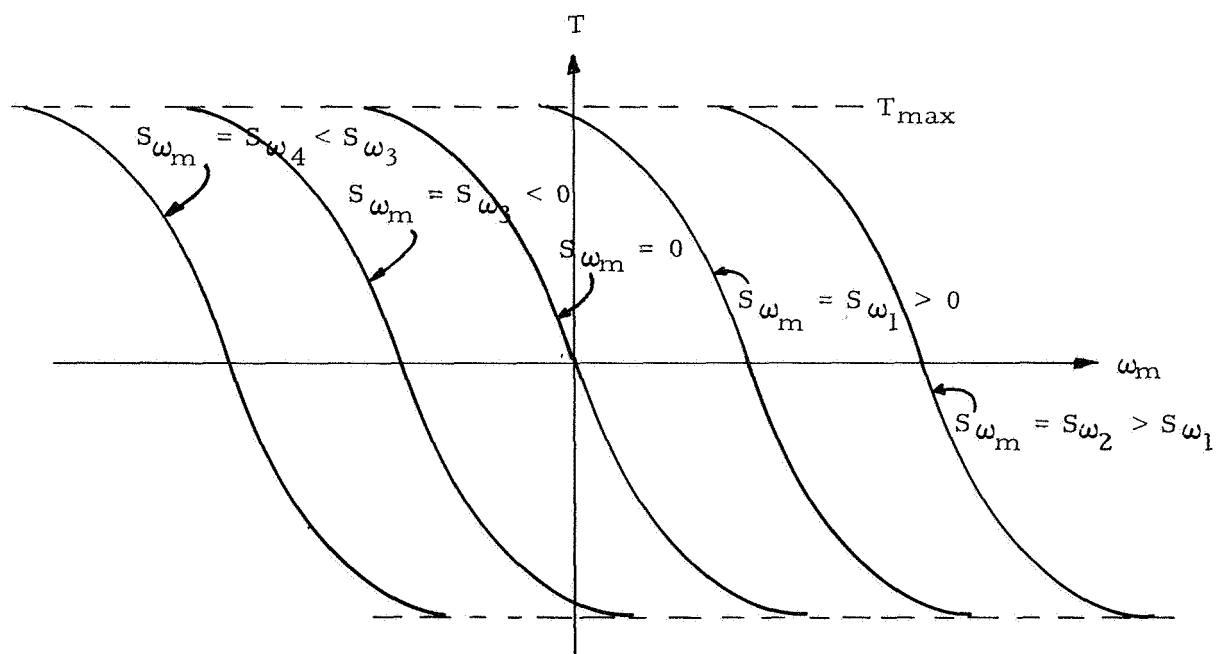
a) Block Diagram of Open Loop Speed Control System

FIGURE 4.13

BLOCK DIAGRAM AND CHARACTERISTICS OF OPEN LOOP SPEED CONTROL SYSTEM



b) Transfer Characteristics of Open Loop Speed Control System



Torque-Speed Characteristics of Open Loop Speed Control System

FIGURE 4.13 (cont.)

BLOCK DIAGRAM AND CHARACTERISTICS OF OPEN LOOP SPEED CONTROL SYSTEM

1) Speed regulation is limited to that inherent in the induction motor. The slip frequency of the breakover torque bounds the speed regulation band. In a well designed motor, the breakover slip frequency might be 5 to 10% of the motor design speed. The speed regulation band remains a fixed percentage of the motor design speed regardless of the actual shaft speed; hence, at low speeds percentage speed regulation is poor.

2) The response to an input step change in speed is limited to the response obtained with the worst case load. If the load applied to the motor varies, this response is far from optimal.

3) At any given speed, as the motor torque increases, the current into the motor increases. The magnitude of applied voltage is held fixed by the system; as the stator current increases, voltage drop in the stator resistance decreases the voltage available to flux the motor, hence the peak torque available drops. If the motor voltage is adjusted for maximum flux at no load, at full load the breakover torque can drop 5 to 15% below the motor's actual capability.

4) If the breakover torque of the motor is exceeded due to a static or dynamic momentary overload, no means is provided for the system to recover speed control. If the load is passive, the shaft speed will drop to nearly zero and the system will continue to operate in a very inefficient mode until manually reset.

In spite of these drawbacks, the simplicity of the open loop system has led to its widespread use in large scale industrial applications.

If a higher level of performance than that provided by the open loop system is desired, a system which infers the slip frequency of the motor from measurement of the electrical parameters may be employed. The block diagram for this system is presented in Figure 4.14. Operation is based on Eqn. 4.4, graphed in Figure 4.10. For each value of slip frequency, there is a corresponding stator current. If the motor flux is voltage controlled, then the stator current may be used as a measure of the slip frequency at which the motor is being operated. The slip frequency detector is the inverse of Figure 4.10. The sign detector

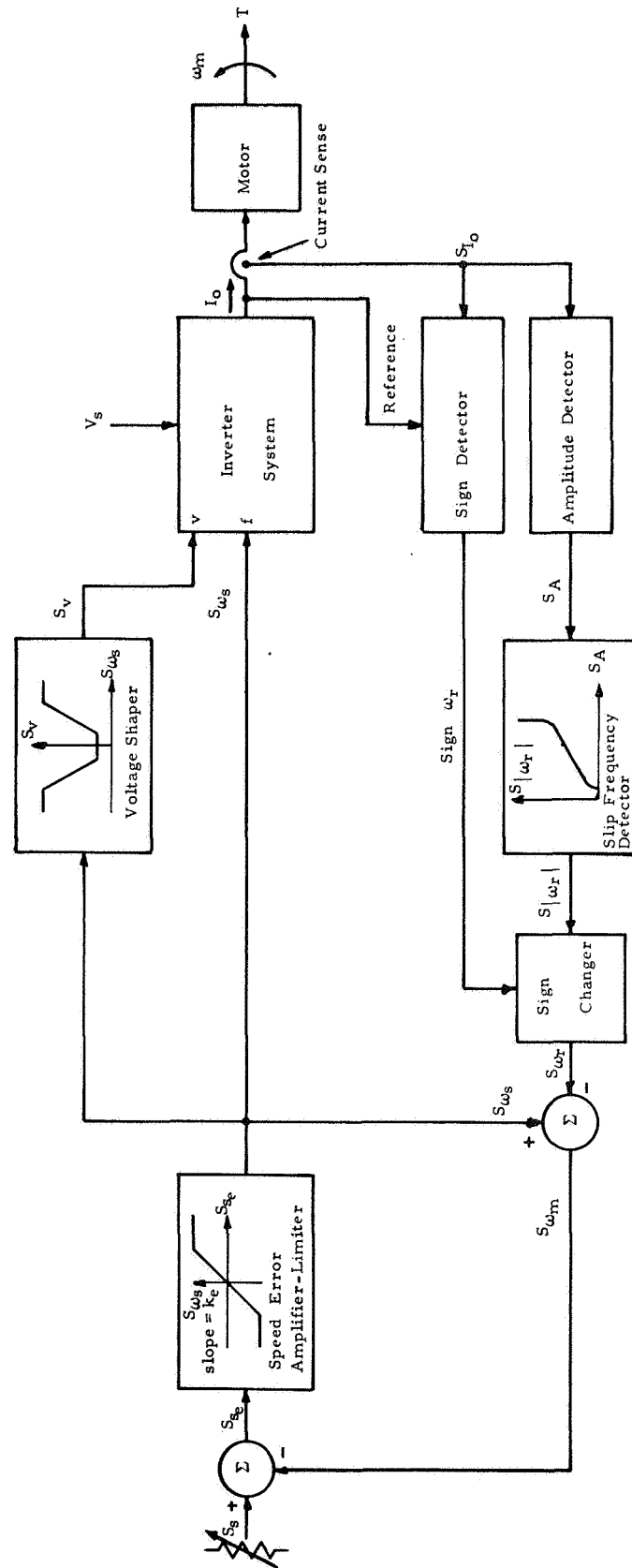


FIGURE 4.14

INFERRED SLIP FREQUENCY SPEED CONTROL SYSTEM

and changer is necessary to supply the slip frequency sign information. This is done by comparison of the phase of the output voltage and current. The resultant signal is an inferred replica directly proportional to the actual slip frequency. The inferred slip frequency signal S_{ω_r} is subtracted from the stator frequency signal S_{ω_s} resulting in a signal S_{ω_m} directly proportional to the actual mechanical rotation frequency ω_m . S_{ω_m} is compared to the speed reference signal S_s and the resultant speed error signal amplified and limited. This amplified speed error signal, S_{ω_s} , is used as the stator drive frequency which tends to minimize the speed error. Statically, the relation between the reference speed signal S_s and rotor speed ω_m may be expressed as

$$\omega_m \approx S_s - \frac{\omega_r}{k_e}, \text{ if } k_e \gg 1. \quad (4.9)$$

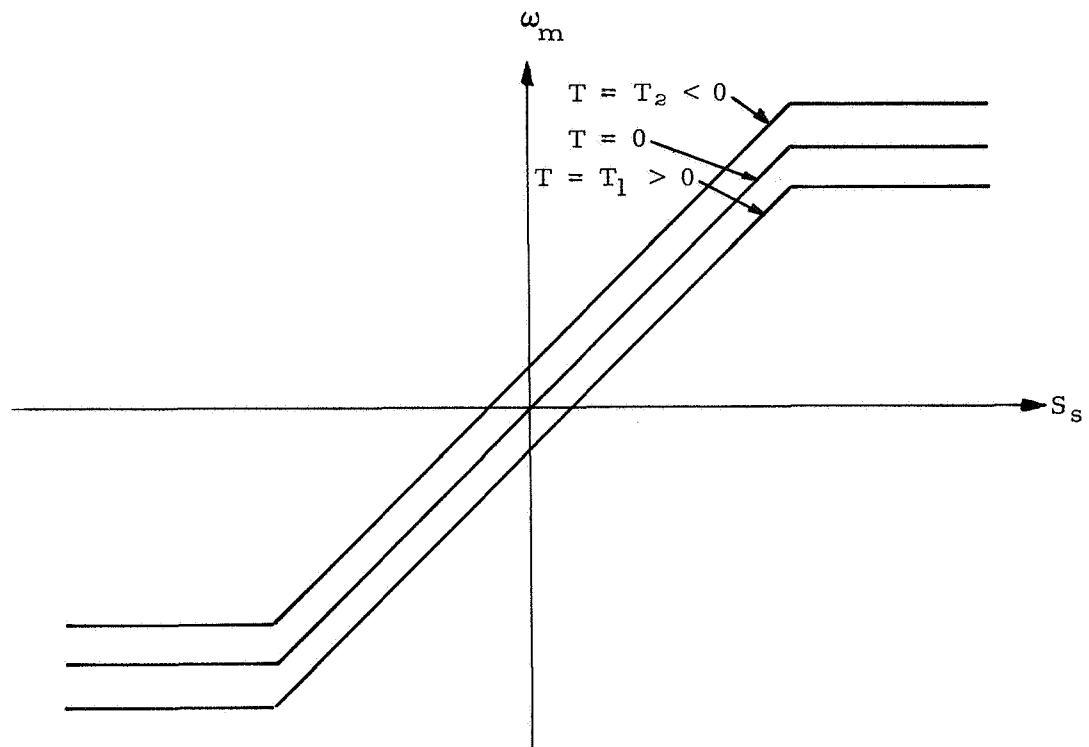
If the gain k_e of the error amplifier is large, then the speed error of the motor will be small (within the limitations of this model).³ Figure 4.15 shows the transfer and output characteristics of the inferred slip system. The transfer characteristics in Figure 4.15 are more closely spaced than those of the open loop system, Figure 4.13b, and in the torque-speed curves in Figure 4.15b there is much less variation of frequency with torque than in Figure 4.13c.

The inferred slip system presented here improves the speed regulation characteristics of the open loop system, but still shows the three other disadvantages of open loop operation:

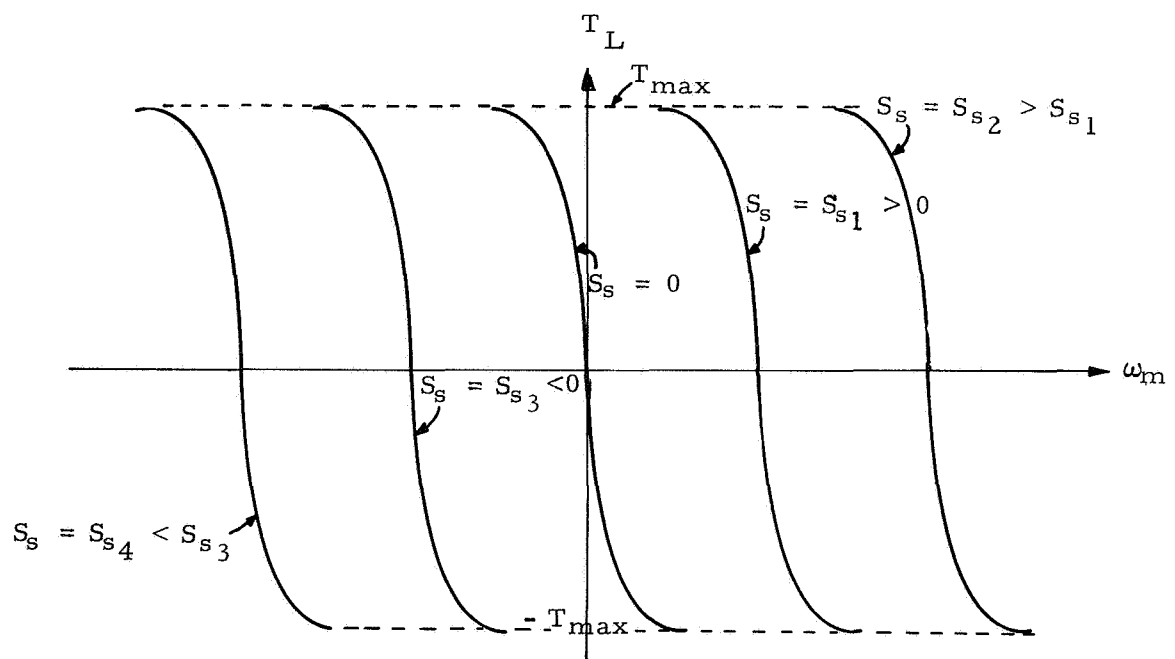
- 1) Poor transient response.
- 2) Lowered breakover torque.
- 3) Unstable operation if breakover torque is exceeded.

Speed Regulation by Feedback Means

Several parameters of the induction motor may be employed



a) Transfer Characteristics of System



b) Torque Speed Characteristics of System

FIGURE 4.15

CHARACTERISTICS OF INFERRED SLIP SPEED CONTROL SYSTEM

to effect speed regulation of the motor. The key to understanding these is the machine torque-slip frequency characteristic given by Eqn. 2.6. For a given shaft speed ω_{m_1} and torque output T_1 , this equation is plotted for two different sets of parameters in Figure 4.16. In figure 4.16a, the motor flux is held constant at its maximum value and the slip frequency (or equivalently, stator frequency if the mechanical speed is fixed) is adjusted to yield the required torque T_1 at ω_{m_1} . In Figure 4.16b, the slip is held constant at the breakover value $\frac{1}{\tau}$ and the flux in the motor is adjusted to yield T_1 . Obviously, there are many combinations of slip frequencies and flux that lie between the two cases presented above which will satisfy the T_1 , ω_1 load requirements. Figure 4.16c, a plot of the full flux torque versus ω_s for a given value of $\omega_m = \omega_{m_1}$ graphically shows the range of possible slip frequencies which meet the load requirement. At each value of permissible slip frequency, the flux $\frac{E_1}{\omega_s}$ may be adjusted according to Eqn. 2.6, or the corresponding plot, Figure 2.3 to achieve torque T_1 . Figure 4.16c shows that for a given flux there are two possible values of slip frequency, one on each side of the peak of the torque curve, which will yield the desired load torque and speed. Figure 4.10, however, indicates that the operating point with the larger slip frequency requires more stator current than the other to generate the same power output. The increased stator current corresponds to internal losses for the same output power, hence the high slip frequency operating point is not used. The permissible range of slip frequencies is restricted to

$$\omega_r \text{ min} \leq \omega_r \leq \omega_r \text{ max} = \frac{1}{\tau} \quad (4.10)$$

as shown in Figure 4.16c.

The examples to be discussed here employ constant flux, variable slip frequency to control the motor speed (Figure 4.16a). However, other systems employing constant slip frequency, variable flux or combinations of the two methods have been reported.

The disadvantages of the open loop and inferred slip systems may be overcome and stable, high performance speed regulation

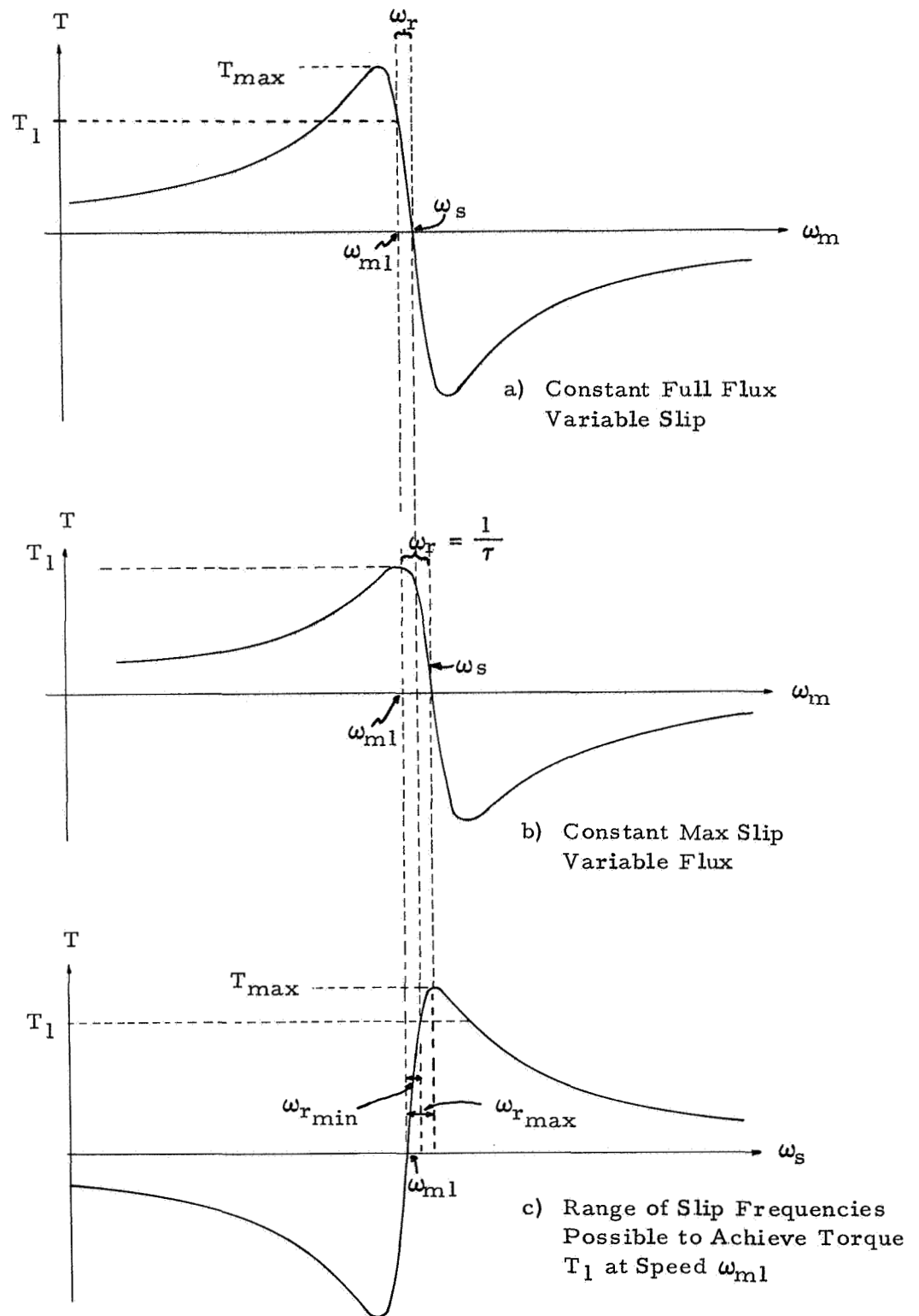


FIGURE 4.16

METHODS OF ACHIEVING A GIVEN TORQUE AND SPEED

realized if feedback around the motor is employed to control the actual slip frequency at which the motor operates. An example of a controlled slip speed regulation system is shown in Figure 4.17. The system diagrammed is based on the constant torque system of Figure 4.12. The torque of the motor is servoed to control the motor speed by using the speed error signal, S_T as the input to the torque generator. For very tight speed regulation, a speed error amplifier might be inserted at the input to the constant torque portion of the system. Figure 4.18 illustrates the characteristics of the system of Figure 4.17. Comparison with Figures 4.13 and 4.15 illustrate the superior speed regulation slip control allows. In addition, there are no unstable regions, regardless of the transient encountered, and transient inputs of disturbances are corrected with the maximum torque that can be supplied by the motor.

In highly critical applications, even the small speed error encountered in controlled slip operation may be intolerable. For these applications, the mechanical speed must be maintained in precise synchronism with a reference AC signal. For these applications, a phase lock system may be used to synchronize the shaft with the reference signal. The phase lock system, also like the controlled slip system, is based on the constant torque system of Figure 4.12. In Figure 4.19, the block diagram of a phase control system is shown. The constant torque portion of this system is modelled as a single block containing all of the parts diagrammed in Figure 4.12. Input is the torque signal, S_T , which is servoed to control the shaft position. If the system is out of synchronism, the rate of change of phase between the shaft position and the reference signal is large and the phase signal $S_{\theta e}$ oscillates at a relatively high frequency between $S_{\theta \max}$ and $-S_{\theta \max}$. The low pass filter attenuates these oscillations and $S_{\theta f}$ is very small. S_s is a signal proportional to the frequency of the reference signal S_{ref} . S_s is compared to the actual motor speed S_{ω_m} and the difference between the signals used to control the torque of the motor system to minimize the error. This behavior is identical to that of the controlled slip speed regulation system; the motor accelerates in a well controlled manner to a speed near synchronism with the reference signal. As synchronism is approached, the frequency

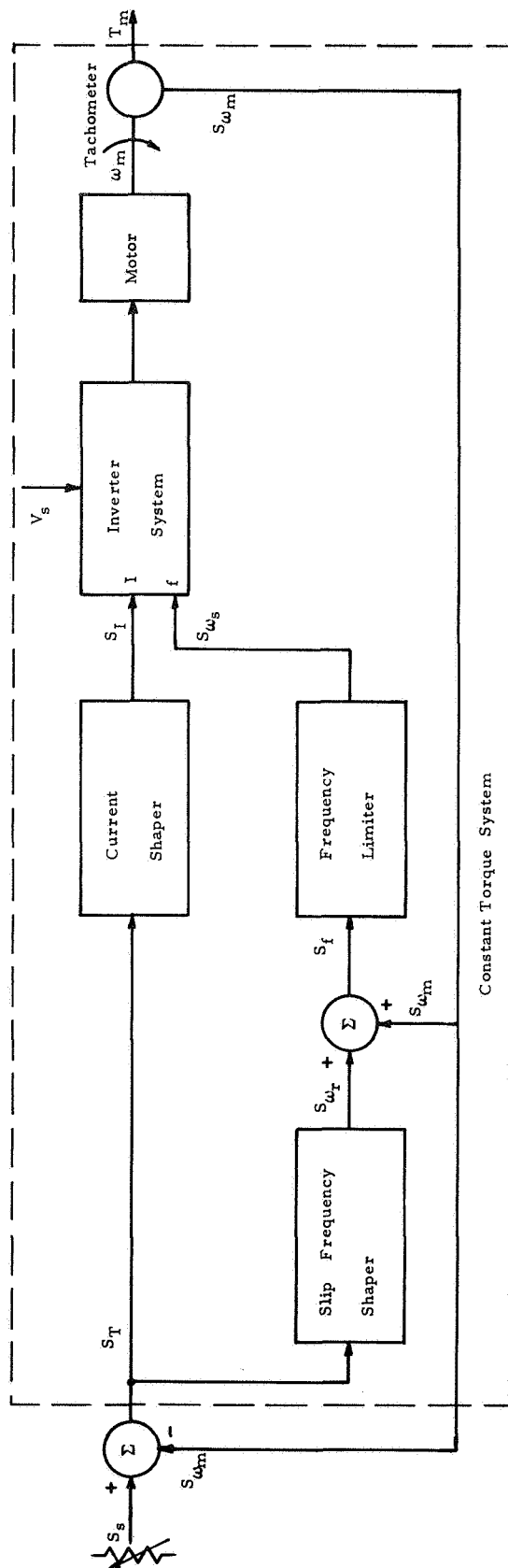
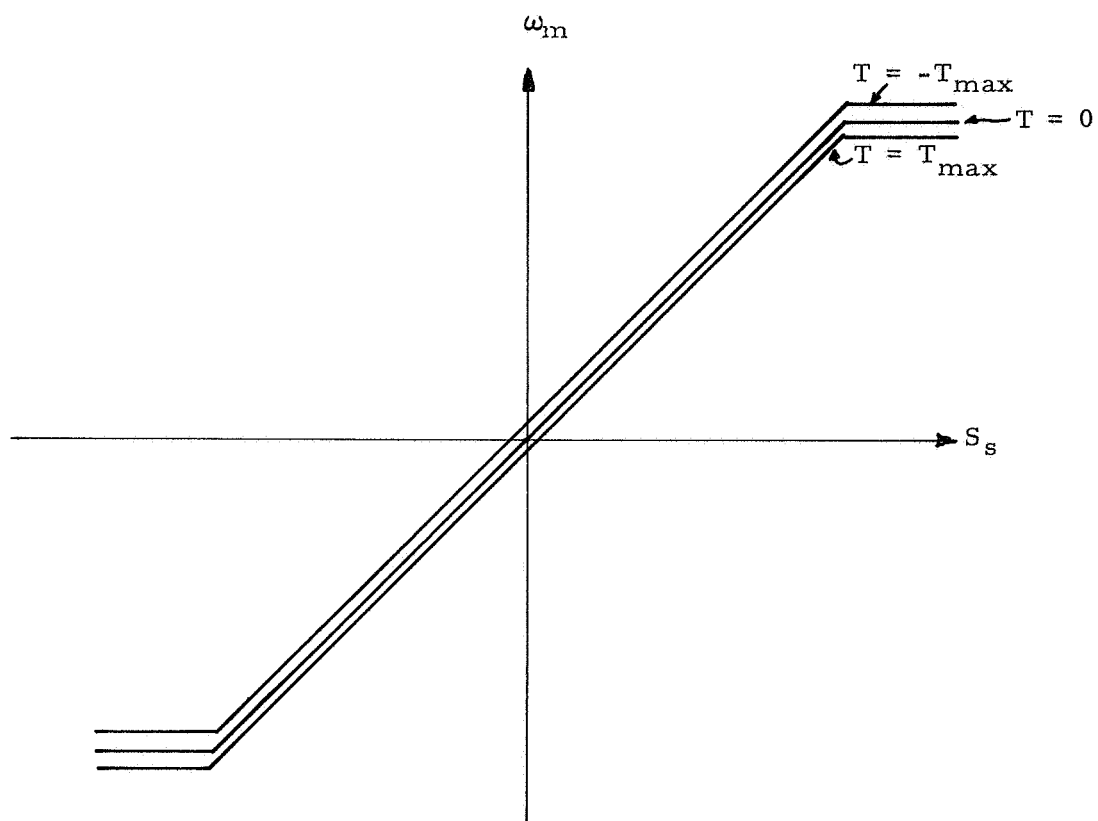
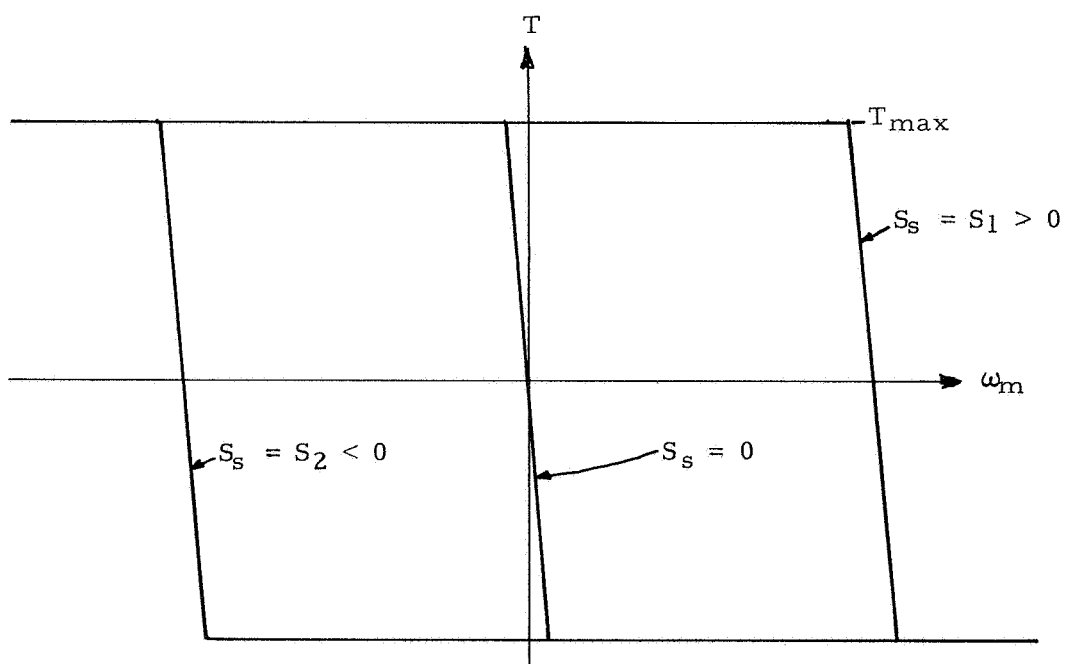


FIGURE 4.17

CONTROLLED SLIP SPEED REGULATION SYSTEM



a) Transfer Characteristics of System



b) Torque Speed Characteristics of System

FIGURE 4.18

CHARACTERISTICS OF CONTROLLED SLIP SPEED REGULATION SYSTEM

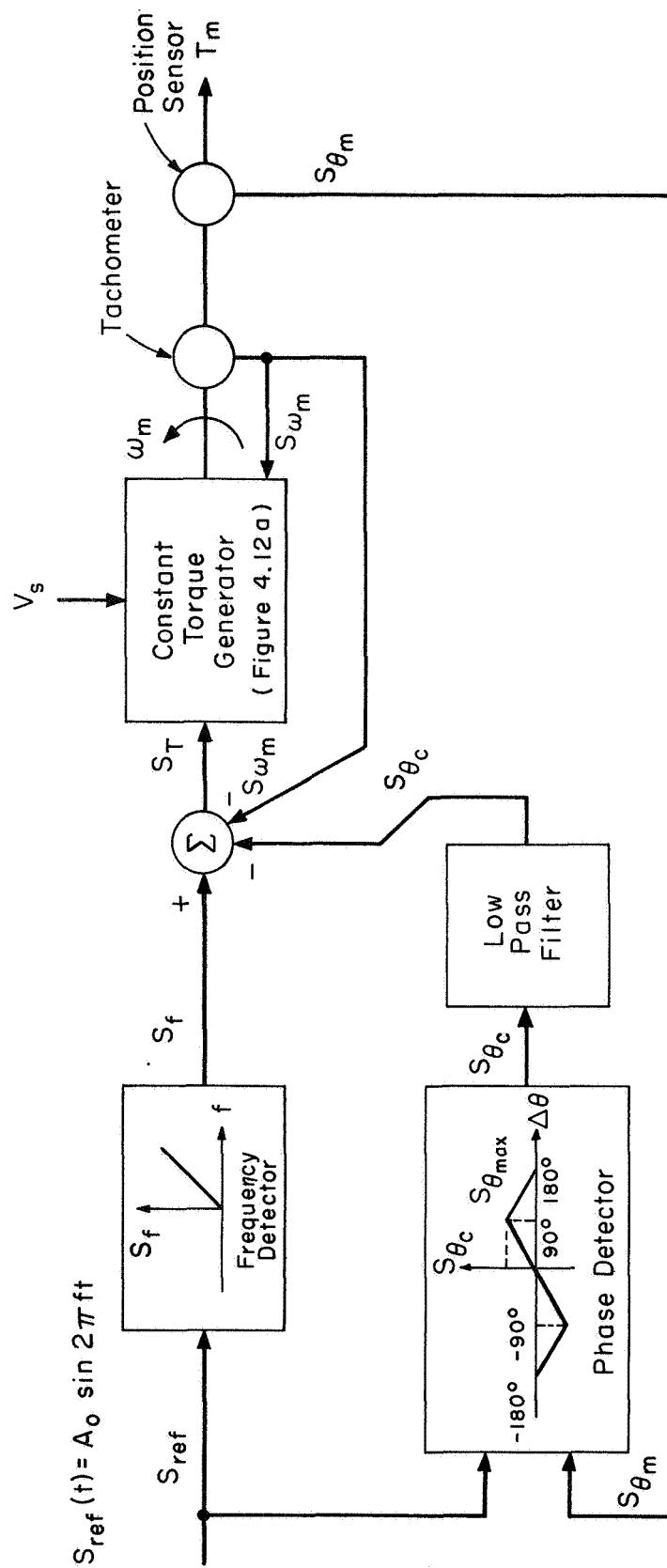
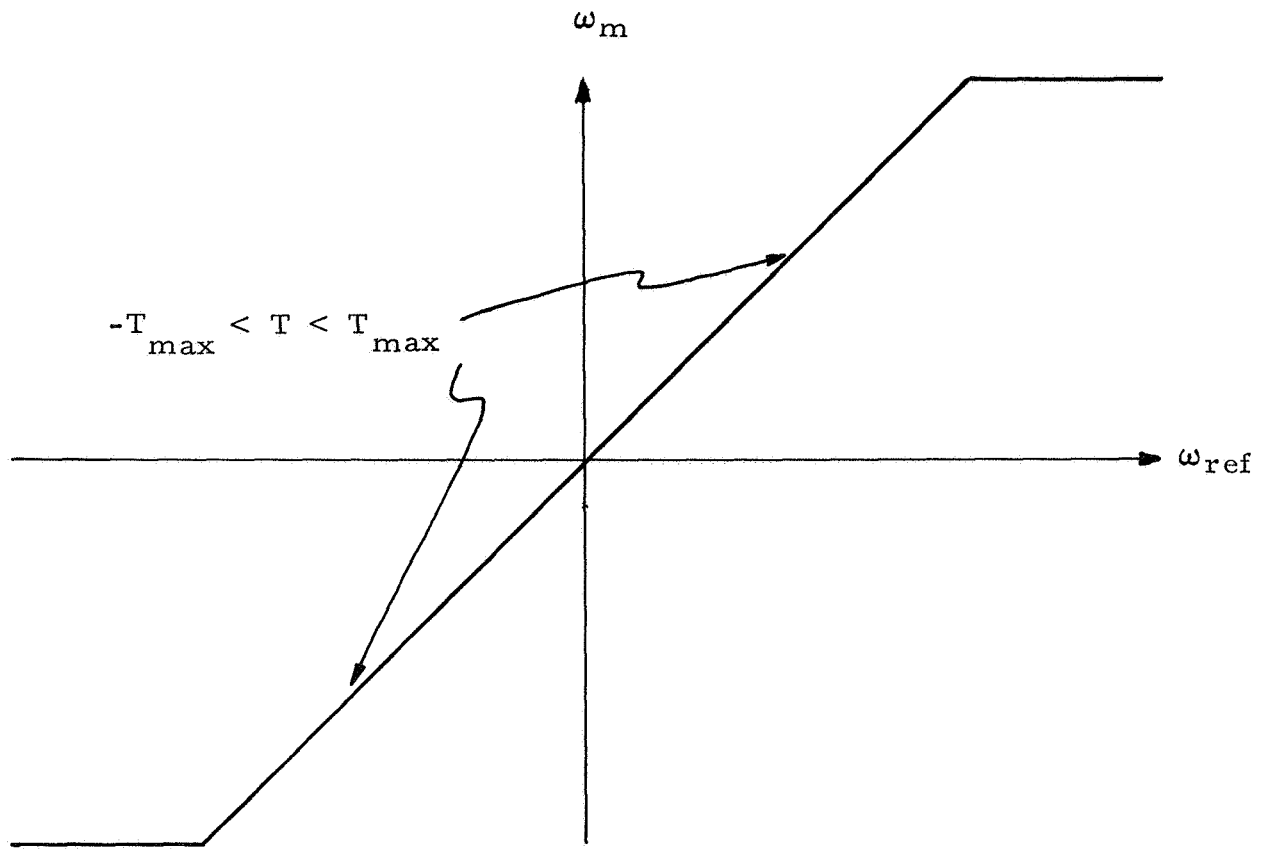


FIGURE 4.19

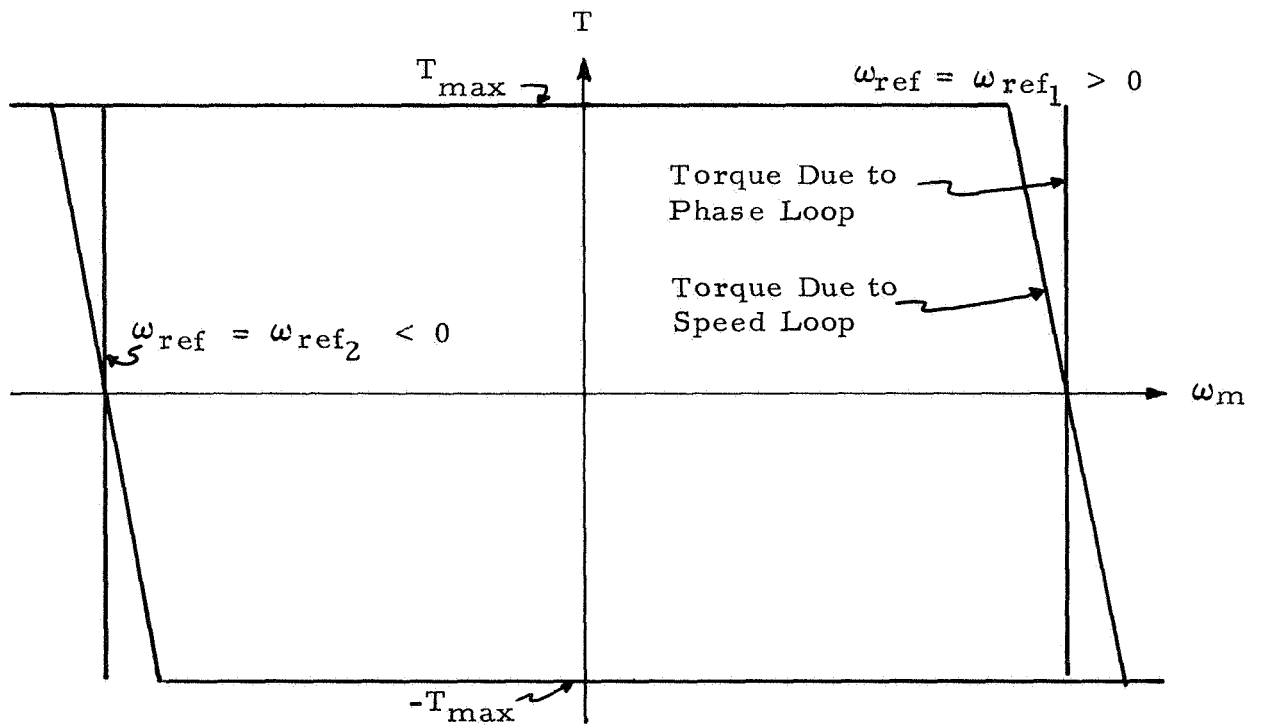
PHASE CONTROLLED INDUCTION MOTOR SYSTEM

of oscillation of $S_{\theta e}$ becomes small and $S_{\theta f}$ adds a pulsating component of torque to the speed error signal. If the frequency of pulsation is comparable to the mechanical time constant of the load, the load will follow the torque pulsations of alternately accelerating and decelerating in speed. If during an acceleration pulsation the motor speed becomes exactly equal to the reference frequency, the motor will lock into synchronism with the reference and the torque signal necessary to maintain synchronism will be supplied by the phase detector. When synchronism is achieved, the speed reference signal S_s will exactly equal the shaft speed signal S_{ω_m} and the difference between the two will be zero. Thus the speed control loop used for accelerating to near synchronism is effectively removed from the operation when synchronism is attained. If the motor decelerates due to an increase in load, the rotor position will begin to lag further behind the reference signal. This will generate a large phase error signal $S_{\theta e}$ which will accelerate the motor through an increase in torque to counteract the deceleration torque. If $S_{\theta \max}$ is large enough to drive the system to full torque, then exact synchronism will be maintained within the full torque capabilities of the motor. The phase error between the shaft position and reference signal will always be within $\pm 90^\circ$.

Figure 4.20 shows the performance characteristics of the phase lock system. Linearity in speed and regulation with respect to torque variations is virtually perfect provided the breakover torque is not exceeded. It is interesting to note that the mode of operation and the characteristics obtained are similar, but superior to, those of a synchronous-induction motor, often employed where precise speed control is necessary. The control system required to realize phase lock operation with an induction motor is more complex than that necessary to operate a synchronous motor, but the phase lock-induction motor system is more efficient, uses a lighter weight motor, and develops greater maximum torque for both initial acceleration and in synchronous operation than a comparable system employing a synchronous motor.



a) Transfer Characteristics of System



b) Torque Speed Characteristics of System

FIGURE 4.20

CHARACTERISTICS OF PHASE CONTROLLED SPEED REGULATION
SYSTEM

5. SUMMARY AND RECOMMENDATIONS

This report has examined several areas important to VSIM drive systems. The current state-of-the-art in these systems, as assessed from published literature has been studied and the significant factors and limitations, as well as representative examples of various approaches, have been presented. Previously unpublished analytical extensions of theory pertinent to operation of induction motors over a wide speed range, along with new approaches to the realization of inverters systems and methods of control of the motor-inverter system are included.

It is evident, from the study of induction motor characteristics in Chapter 2 that a high level of performance is attainable from an optimally designed induction machine. It was shown that minimization of the rotor resistance would maximize efficiency and also minimize the generation of spurious torque due to non-sinusoidal drives.

However, the high theoretical level of performance of the VSIM system can only be realized if an electrical drive capable of properly exciting the motor under steady state and dynamic conditions is provided. The discussion and analysis of Chapter 2 show that for highest performance and complete flexibility of operation (four quadrant torque-speed capability) the drive must possess the following characteristics:

- 1) Sufficient voltage capability to fully flux the motor at the highest stator drive frequency. This insures that maximum torque for optimal dynamic response will be available under all operating conditions.
- 2) Sufficient peak current capability to realize the maximum available torque under all motor operating conditions and in the presence of harmonic currents.
- 3) A capability for regenerating as well as supplying AC power. This feature provides optimal response during braking or when driving active loads.

4) A capability for providing controlled drive down to zero frequency. During motor reversing speed changes or when driving an active load at low speeds, DC drive is required to realize optimal motor torque capability.

5) Low harmonic content in the AC drive waveform, particularly at low drive frequencies is required to minimize harmonic and pulsating torques. At high frequencies, low harmonic and spurious frequency content is required to minimize excess core losses and large stator currents at frequencies other than the drive frequency.

6) The drive must be responsive to variations in the load. For the highest level of performance, the drive to the motor must be adjusted by feedback means to compensate for load changes and to provide optimal transient response.

The study of inverter systems presented in Chapter 3 indicates the nature of the problems involved in static inverter designs which can meet the induction motor drive requirements. The design problems center on those of the output power stage. Simple systems produce waveforms rich in harmonic and spurious frequencies. The necessity for DC output requires that the power stage output switches be connected by a direct path to the terminals and precludes the use of transformer coupled output stages. This constraint limits the number of usable output stages.

The presentation of the waveforms generated by various control concepts (Section 3.2) indicates that to insure a small harmonic content at low frequencies, a sine wave approximation method, using samples at a relatively high frequency, should be employed. If a pulse modulator or square wave cycloconverter is used, the predetermined waveform methods of control may be easily simulated if desired.

At the power levels of the systems under study ($>2\text{KW}$) and the supply voltage levels considered (200 - 500 volts) the use of thyristor switching elements is mandatory with the current state-of-the-art. All of the inverter systems considered in this study require forced commutation of at least some of the thyristors. The study of basic design considerations of the output stage (Section 3.4) indicates that the McMurray circuit, with a theoretical commutating efficiency of near 100%, is an attractive choice. Experimental studies performed outside of the scope of this contract indicate that this configuration can in fact commute at high

efficiency and, with some modification, can be made to operate with high reliability. It has been found, however, that instantaneous current control or at least instantaneous current limiting is necessary to prevent commutation failures.

The investigation of parameter control techniques presented in Chapter 4 indicates that the form of control employed is very much dependent on the particular mission requirements. All of the methods of voltage, current, flux, torque and speed control, when properly implemented, will operate the motor. The more sophisticated methods will yield a higher level of performance. Several factors that govern the choice of parameter control techniques are evident:

- 1) Instantaneous current control or limiting must be employed to protect the inverter switching devices under dynamic load and fault conditions.
- 2) For optimal flux control, either flux feedback or some method of flux drive compensation with load variations must be employed.
- 3) The method of slip control employed (constant slip frequency, constant flux or a combination of both) must be studied for the particular motor, inverter and mission requirements to optimize the performance parameters.

Recommendations

The crux of the VSIM drive problem lies in the successful realization of the inverter output power stage. Motor technology has advanced to the point where very large motors (> 1000 KVA) are not uncommon. The theory of motor operation and the art of motor design are both well advanced. While the combination of motor parameters optimal for VSIM applications are not those that are conventionally designed into motors, the design trade-offs are apparently understood in detail sufficient to allow optimization of these parameters by established techniques. On the other hand, the design of solid state inverters to drive induction motors of even moderate power (10 KVA) represents a very recent advance in the state-of-the-art. The detailed theory of operation of these units is not yet well understood and design is largely an art. As might be

expected under these circumstances, it is in the inverter, and more particularly in the inverter power output stage where most of the design effort must be expended. Therefore, the following experimental design and evaluation program is recommended.

The principle design parameter (in addition to well controlled, reliable operation) associated with the output stage is power loss. This is also one of the most difficult parameters to theoretically evaluate, since in a well designed output stage, the losses are due to effects that are often second order with respect to operation. Therefore, it is recommended that an experimental study of several output stages, combined with suitable control concepts, be undertaken to evaluate these stages for a particular mission requirement. To meet the motor requirements for high performance, the more versatile output stages should be investigated. It would appear that both the pulse modulator and square wave cycloconverter have merit in this application. With both of these approaches, the simpler schemes such as square wave drive and "sliding wave" drive may be simulated to assess their effect upon the overall system.

This study had been centered on and guided by the current literature on VSIM drive systems and inverter designs. Other basic approaches to the inverter problem exist, but have not yet been explored, at least not in the published literature. In particular, the series inverter⁷ holds the possibility of competing favorably with the two high performance systems recommended for further study. It would be appropriate to investigate the series inverter output stage along with others until all can be compared.

When the optimal output stage has been selected, control concepts suitable to that output stage may be selected and evaluated. The choice of control concept must be restricted to one which provides instantaneous current control or current limiting to insure reliable operation under all conditions.

When a complete inverter system has been realized, it may be mated to an induction motor (not necessarily one optimized for VSIM use) and parameter control concepts investigated. The optimum control concepts should be selected which meet a particular (but flexible) set of operating specifications. For example, a system which has unrestricted four quadrant torque-speed capability would be an interesting and fruitful one to investigate.

With an optimal inverter and control system, the induction motor may then be redesigned to possess the desirable characteristics which will have become evident in the parameter control study. At this time, the tradeoffs between desirable electrical characteristics and motor physical constraints (e.g. cooling) can be profitably studied.

Finally, the tradeoff and design data gathered during the study and development of the initial VSIM drive will allow the entire system to be redesigned as a whole to obtain the optimal overall system. The redesigned system will serve as a check on the validity of the design tradeoffs formulated during the initial study. To study the effect of scaling on the initial design conclusions, the redesigned system could also be scaled in power at the same time.

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NEW TECHNOLOGY APPENDIX

After a diligent review of the work performed under this contract, no new innovation, discovery, improvement or invention was made.

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