CHAPTER 2

PERFORMANCE ANALYSIS OF CONVENTIONAL INVERTER FED INDUCTION MOTOR AND FOUR SWITCH THREE PHASE INVERTER FED INDUCTION MOTOR

2.1 INTRODUCTION

The inverters are widely used semiconductor device with the variable speed AC motor applications. The SSTPI has been widely utilized for variable speed drives. These inverters have some drawbacks, which involve the switching losses of the six switches as well as the complexity of the control algorithms and interface circuits to generate six PWM logic signals due to high switching frequency. Hence application of FSTPI is viable alternative for conventional inverter fed induction motor as described by Takijawa et al (2000). The FSTPI is cheaper due to minimal number of interface circuits to supply logic signals for the switches and less chances of destroying the switches due to the lesser interaction among switches. The reduction of the power switches from six to four improves the cost effectiveness, the volume compactness and the reliability as described by Mohamed Azab et al (2001). This chapter presents the simulation and implementation of FSTPI fed induction motor. Wenyi Zhang et al (2006) analyzed and observed that the poor quality of current and voltage of a conventional inverter fed induction machine is due to the presence of
harmonics. Yu-Tzung et al (2007) proposed harmonic reduction techniques in pulse width modulation strategy with six power switches. This chapter deals with the simulation study of FSTPI fed induction motor. The experimental work is done using microcontroller and the results are obtained.

2.2 FSTPI-CONSTRUCTION AND WORKING

A standard three-phase voltage source inverter utilizes three legs with a pair of complementary power switches per phase. A reduced switch count voltage source inverter uses only two legs with four switches as shown in Figure 2.1.

![Figure 2.1 Basic Circuit of Four Switch Three Phase Induction Motor](image)

The circuit consists of two parts; first part is a front-end rectifier powered from single phase supply. The output DC voltage is smoothed through a two series connected capacitors called split capacitors. The second part of the power circuit is four-switch inverter. The maximum obtainable peak value of the line voltages equals $V_{dc}$. The inverter switches are
considered as ideal switches. The output voltages are defined by the gating signals of the two leg switches and by the two dc link voltages $V_{dc}$.

The output can be taken from the mid points of the two legs and from the midpoint of the split capacitors. In the analysis, the inverter switches are considered as ideal power switches and it is assumed that the conduction state of the power switches is associated with binary variables $q_1$ to $q_4$. The binary “1” will indicate a closed switch and “0” will indicate the open state. Pairs $q_1,q_3$ and $q_2,q_4$ are complementary and as a sequence it can be written as

\[ q_3 = 1 - q_1 \]  
(2.1) and
\[ q_4 = 1 - q_2 \]  
(2.2)

It is assumed that a stiff voltage is available across the two dc-link capacitors and

\[ V_{C1} = V_{C2} = \frac{E}{2} \]  
(2.3)

where $E$ corresponds to a stiff dc-link voltage.

The phase voltage equations of the motor is given by

\[ V_a = \frac{V_{dc}}{3} \left[ 4S_a - 2S_b - 1 \right] \]  
(2.4)

\[ V_b = \frac{V_{dc}}{3} \left[ -2S_a + 4S_b - 1 \right] \]  
(2.5)

\[ V_c = \frac{V_{dc}}{3} \left[ -2S_a - 2S_b + 2 \right] \]  
(2.6)
The above equations are the function of switching logic of the switches and the dc-link voltage. Where $V_a$, $V_b$, $V_c$ are inverter output voltages and $V_{dc}$ is the voltage across dc-link capacitors. $S_a$ and $S_b$ are the switching functions for each phase leg.

In matrix form, the above equations can be written as

$$
\begin{bmatrix}
V_a \\
V_b \\
V_c
\end{bmatrix} = \frac{V_{dc}}{3} \begin{bmatrix}
4 & -2 \\
-2 & 4 \\
-2 & -2
\end{bmatrix} \begin{bmatrix}
S_a \\
S_b
\end{bmatrix} + \frac{V_{dc}}{3} \begin{bmatrix}
-1 \\
-1 \\
2
\end{bmatrix}
$$

(2.7)

For a balanced capacitor voltages, the four switching combinations lead to four voltage reactors. Table 2.1 shows the different modes of operation and the corresponding output voltage vectors of the inverter.

Table 2.1 Inverter Modes of Operation

<table>
<thead>
<tr>
<th>Switching function</th>
<th>Switch ON</th>
<th>Output Voltage vector</th>
</tr>
</thead>
<tbody>
<tr>
<td>$S_a$</td>
<td>$S_b$</td>
<td>$V_a$</td>
</tr>
<tr>
<td>0</td>
<td>0</td>
<td>$q_3$</td>
</tr>
<tr>
<td>0</td>
<td>1</td>
<td>$q_2$</td>
</tr>
<tr>
<td>1</td>
<td>0</td>
<td>$q_1$</td>
</tr>
<tr>
<td>1</td>
<td>1</td>
<td>$q_1$</td>
</tr>
</tbody>
</table>

The modes of operation are explained by the following figures.
Figure 2.2 Circuit Model for the Switching State 0, 0 (S_a and S_b open state)

Figure 2.3 Circuit Model for Switching State 0, 1 (S_a open state and S_b closed state)

Figure 2.4 Circuit Model for Switching State 1, 0 (S_a closed state and S_b open state)
2.3 THE EFFECT OF HARMONICS

In the converter fed induction motor applications the waveforms are either pulse width modulated (PWM) or square waveform instead of sinusoidal. Such waveforms can be analyzed by Fourier series and are shown to have a fundamental or useful component and undesirable harmonic components. The harmonics generally have two undesirable effects:

- Harmonic heating
- Torque pulsation

2.3.1 Harmonic Heating

For simplicity, only the balanced three-phase, square-wave voltage supply is considered. For a balanced three-phase square-wave voltage supply, only the odd harmonics should be considered. Again, the third and its multiple (triplen) harmonics are in the same phase (co-phasal – often defined as zero sequence component) and therefore cannot cause any triplen harmonic current in a wye-or delta-connected machine winding with isolated neutral.
Considering only the non-triplen harmonics, the Fourier series of the phase voltages can be given as

\[ V_{as} = V_{lm}\sin\omega_{et} + V_{5m}\sin5\omega_{et} + V_{7m}\sin7\omega_{et} + \ldots \quad (2.8) \]

\[ V_{bs} = V_{lm}\sin\omega_{et} + V_{5m}\sin(\omega_{et}-2\pi/3) + V_{7m}\sin(\omega_{et}-2\pi/3) + \ldots \quad (2.9) \]

\[ V_{cs} = V_{lm}\sin\omega_{et} + V_{5m}\sin(\omega_{et}+2\pi/3) + V_{7m}\sin(\omega_{et}+2\pi/3) + \ldots \quad (2.10) \]

where \( V_{lm}, V_{5m} \) are the peak values. Equations (2.9) and (2.10) can be simplified as

\[ V_{bs} = V_{lm}\sin(\omega_{et}-2\pi/3) + V_{5m}\sin(\omega_{et}+2\pi/3) + V_{7m}\sin(\omega_{et}-2\pi/3) + \ldots \quad (2.11) \]

\[ V_{cs} = V_{lm}\sin(\omega_{et}+2\pi/3) + V_{5m}\sin(\omega_{et}-2\pi/3) + V_{7m}\sin(\omega_{et}+2\pi/3) + \ldots \quad (2.12) \]

For each harmonic component, the machine can be approximately represented by a constant-parameter and linear-equivalent circuit. The resultant ripple current can be solved by the superposition principle. The harmonic-equivalent circuit is shown in Figure 2.6 where the core loss resistor \( R_m \) has been omitted. In this Figure ‘n’ is the order of the harmonic and \( S_n \) is the slip at the 5th harmonic. Equations (2.8), (2.11), and (2.12) indicate that the 5th harmonic voltage component has negative phase sequence, whereas the 7th harmonic has positive phase sequence like the fundamental component. Obviously, the 5th harmonic will create air gap flux which will rotate in the backward direction, whereas the fundamental and the 7th harmonic will create flux that will rotate in the forward direction. The effect of higher harmonics can be derived in a similar manner. Since rotor speed is related to fundamental frequency only, the rotor appears practically stationary with respect to a fast-moving harmonic field, that is, \( S_n = 1.0 \).
Mathematically, the slip at the $n^{th}$ harmonic field can be given as

$$S_n = \frac{n\omega_e \pm \omega_r}{n\omega_e}$$  \hspace{1cm} (2.13)$$

where negative and positive signs relate to forward and backward rotating fields. Substituting $\omega_r / \omega_e = (1 - S)$ in Equation (2.13) and simplifying yields to

$$S_n = (n-1) + S_1/n$$  \hspace{1cm} (2.14)$$

$$S_n = (n+1) - S_1/n$$  \hspace{1cm} (2.15)$$

The above expressions are corresponding to forward and backward-rotating field respectively. In these equations, $S_1$ is the slip for fundamental frequency. For example, if $S_1$ varies from 0 to 1, $S_3$ and $S_5$ will vary in ranges from 1.2 to 1.0 and 0.857 to 1.0, respectively. It is to be noted that the harmonic currents are not influenced by the fundamental frequency operating condition and they are independent of the torque and speed of the machine.

The rms harmonic ripple current $I_h$ is

$$I_h = \sqrt{I_5^2 + I_7^2 + I_{11}^2 + I_{13}^2 + ...}$$

$$= \sqrt{\sum I_n^2}$$  \hspace{1cm} (2.16)$$
where \( I_5, I_7 \) are the rms harmonic components of the current. The total stator and rotor copper losses due to fundamental and ripple currents can be given as

\[
P_{ls} = 3(I_{sl}^2 + I_{h}^2)R \tag{2.17}
\]
\[
P_{lr} = 3(I_{rl}^2 + I_{h}^2)R_r \tag{2.18}
\]

where \( I_{sl} \) and \( I_{rl} \) are the fundamental stator and rotor currents respectively.

The core loss in the machine created by time-varying flux in the machine laminations consists of hysteresis and eddy current losses. This loss is contributed by the fundamental as well as harmonic components of the supply voltage. The per phase stator core loss at fundamental frequency can be given as

\[
P_{cs} = K_h \Psi_m^2 + K_e \phi_m^2 \tag{2.19}
\]

where \( \Psi_m = \) fundamental air gap flux linkage,
\( K_h = \) hysteresis loss coefficient and
\( K_e = \) eddy current loss coefficient.

The parameter \( K_h \) depends on the magnetic properties of iron and machine size and \( K_e \) depends on machine geometry size, lamination thickness and resistivity of iron. Under normal operating conditions, the fundamental rotor frequency \( S\omega_c \) is a small fraction of the stator frequency and the corresponding rotor core loss \( P_{cr} \) is given by

\[
P_{cr} = K_h S\omega_c \phi_m^2 + K_e (S\omega_c)^2 \phi_m^2 \tag{2.20}
\]

where \( S\omega_c \) has been substituted for \( \omega_c \). Obviously, the rotor core loss is small compared to that of the stator. Substituting \( \Psi_m \) value in equation (2.20), the total core loss is given as

\[
P_c = P_{cs} + P_{cr} = [K_h (1 + S/\omega_c) + K_e (1 + S^2)] \frac{V_m^2}{R_m} \tag{2.21}
\]
Therefore, the equivalent core loss resistance can be written as

$$R_m = \frac{1}{K_h \left[ 1 + S/\omega_c \right] + K_e (1 + S^2)} \quad (2.22)$$

Assuming that the core loss due to harmonic flux is governed by the same principles that determine the fundamental core losses, the coefficients $K_h$ and $K_e$ are the same at harmonic frequency. Since the harmonic slip $S_n \approx 1$, the equivalent core loss resistance $R_{mn}$, at harmonic frequency $n\omega_c$, can be given as

$$R_{mn} = \frac{1}{\left[ K_h/ n\omega_c + K_e \right]} \quad (2.23)$$

In a variable-frequency drive, harmonic core loss is normally smaller than harmonic copper loss. In addition to copper and core losses, there are stray losses in the machine. By definition, stray losses are those losses that occur in excess of copper loss, friction and windage loss. These losses are essentially due to hysteresis and eddy current losses induced by stator and rotor leakage fluxes.

### 2.3.2 Torque Pulsation

Pulsating torque is produced by air gap flux at one frequency interacting with rotor mmf at a different frequency. Considering only the fundamental frequency or any harmonic frequency alone, the angle $\delta$ remains fixed and therefore only unidirectional torque is produced. A harmonic component of air gap flux induces rotor current at the same frequency and therefore torque is developed in the same direction as that of the rotating flux. For example, the 7th harmonic frequency torque will add with the fundamental frequency torque, but the 5th harmonic frequency torque will oppose it.
The torque pulsation will occur when angle $\delta$ varies with time. The pulsating torques can be calculated by superimposing the flux and rotor current phasors of various frequencies. Mathematically, the $6^{th}$ harmonic torque expression can be written as

$$T_{e6} = K[\varphi_{1m}l_7\sin(\pi -6\omega_c t)+ \varphi_{m}l_7\sin(\delta-6\omega_c t)+$$
$$\varphi_{5m}l_7\sin(\delta-6\omega_c t)]$$  \hspace{1cm} (2.24)

$$= K[\varphi_{1m} (l_7-l_5) \sin 6\omega_c t+ l_n (\varphi_{m}+ \varphi_{5m})\cos 6\omega_c t]$$
$$- K\varphi_{1m}(l_7-l_5)\sin 6\omega_c t$$  \hspace{1cm} (2.25)

whereas the fundamental component of torque is given as

$$T_{e1} = K\varphi_{m} l_7\sin\delta$$  \hspace{1cm} (2.26)

where, $K$ is torque constant and $\delta$ is approximately taken as $\pi/2$ in equation (2.25). Since the harmonic flux components $\Psi_{7m}$ and $\Psi_{5m}$ are very small, the contribution of the second term can be neglected. The pulsating torque will tend to cause jitter in the machine speed, but the effect of high-frequency components will be smoothened due to rotor inertia. Assuming pure inertia load, the speed of $6^{th}$ harmonic torque can be given as

$$J\omega_m/dt = T_m\sin 6\omega_c t$$  \hspace{1cm} (2.27)

(or)

$$\omega_m = T_m\cos 6\omega_c t/ J6\omega_c$$  \hspace{1cm} (2.28)

where $J$ is rotor moment of inertia, $\omega_m$ is rotor speed and $T_m$ is peak value of the $6^{th}$ harmonic torque

Equation (2.28) indicates that at higher inertia and higher harmonic frequency, the speed pulsation will be highly attenuated. On the other hand, at low frequency, mechanical resonance may be induced, causing severe shaft
vibration, fatigue, wearing of gear teeth and instability of the feedback control
system.

2.4 FSTPI FED INDUCTION MOTOR

Figure 2.7 illustrates the circuit of FSTPI fed induction motor. The
FSTPI consists of a two-leg topology. The reduced structure inverter is fed by
a battery pack which is divided into two equal parts. The two phases of the
motor are fed by the legs of the inverter while the third one is connected to the
middle voltage point of the battery pack.

![Four Switch Three Phase Inverter Fed Induction Motor](image)

Figure 2.7 Four Switch Three Phase Inverter Fed Induction Motor

The analysis of the pulse width modulated waveform observed at the
output of an inverter is better understood introducing the concept of a
switching function. Consider the following switching function

\[
S(t) = \frac{\tau_0}{T_p} + \frac{2}{\pi} \sum_{n=1}^{\infty} \frac{1}{n} \sin \left( \frac{n \pi \tau_0}{T_p} \right) \cos(n \omega_p t)
\]  

(2.29)

where \(\tau_0\) is initial pulse width, \(T_p\) is sampling interval and \(\omega_p\) is carrier
frequency.
The switching function given by equation (2.29) is said to be pulse width modulated if $\tau_p^0$ is no longer fixed but varies linearly with a low frequency signal. Let the pulse width of the switching be expressed in terms of the modulating signal as

$$\tau_p(t) = \tau_p^0 [1 + m \cos(\omega_m t)]$$

(2.30)

where $m$ is the modulation index and $\omega_m$ is modulating frequency.

Substituting (2.30) into (2.29) and defining $\tau_p^0 = T_p/2$ the following expression is derived for the pulse width modulated switching function.

$$S(t) = \frac{1}{2} + \left(\frac{m}{2}\right) \cos(\omega_m t) + \frac{1}{\pi} \sum_{n=1}^{\infty} \frac{1}{n} \sin\left[\frac{n\pi}{2} + \frac{mn\pi}{2}\cos(\omega_m t)\right] \cos(n\omega_p t)$$

(2.31)

For the particular case where $m=1$ and $\omega_p = 6 \omega_m$, the general expression of the three phase output voltages of the SSTPI can be derived from the Equation (2.31),

$$v_i(t) = V_m \left\{ \frac{1}{2} \cos(\omega_m t - i \frac{2\pi}{3}) + \frac{2J_4(\frac{\pi}{7})}{\pi} \cos(2\omega_m t - i \frac{2\pi}{3}) - \frac{J_6(\frac{\pi}{7})}{\pi} \cos(3\omega_m t - i \frac{2\pi}{3}) - \frac{2J_2(\frac{\pi}{7})}{\pi} \cos(4\omega_m t - i \frac{2\pi}{3}) + \frac{2J_0(\frac{\pi}{7})}{\pi} \cos(5\omega_m t - i \frac{2\pi}{3}) + \frac{J_5(\frac{\pi}{7})}{\pi} \cos(6\omega_m t - i \frac{2\pi}{3}) - \frac{J_6(\frac{\pi}{7})}{\pi} \cos(7\omega_m t - i \frac{2\pi}{3}) \right\}$$

(2.32)

where $i=0$ indicates the first phase and $i=1$ indicates the second phase and so on. Also, $V_m$ is the amplitude and $J_k(x)$ is the Bessel function of order $k$.

From the equation (2.32) the SSTPI voltage vector can be expressed as

$$v_{SSTPI}^+(t) = \sqrt{3} \left\{ \frac{1}{2} \exp(j\omega_m t) + \frac{2J_4(\frac{\pi}{7})}{\pi} \exp(-j2\omega_m t) + \frac{2J_2(\frac{\pi}{7})}{\pi} \exp(j4\omega_m t) + \frac{2J_0(\frac{\pi}{7})}{\pi} \exp(-j5\omega_m t) + \frac{J_5(\frac{\pi}{7})}{\pi} \exp(j7\omega_m t) \right\}$$

(2.33)
In the case of the FSTPI,

\[ v_{FSTPI}^+(t) = \sqrt{3} \left\{ \frac{1}{2} \exp(jw_mt) + \frac{4J_s(\pi)}{\pi\sqrt{3}} \exp(-j2w_mt) + \frac{2J_s(\pi)}{\pi\sqrt{3}} \exp(j2w_mt) + \frac{14J_s(\pi)}{\pi} \exp(j3w_mt) + \frac{14J_s(\pi)}{\pi} \exp(-j3w_mt) + \frac{4J_s(\pi)}{\pi\sqrt{3}} \exp(j4w_mt) + \frac{2J_s(\pi)}{\pi\sqrt{3}} \exp(-j4w_mt) + \frac{2J_s(\pi)}{\pi} \exp(-j5w_mt) + \frac{2\sqrt{7}J_s(\pi)}{4\pi} \exp(-7w_mt) + \frac{2\sqrt{7}J_s(\pi)}{4\pi} \exp(j6w_mt) + \frac{J_s(\pi)}{\pi} \exp(j7w_mt) + \ldots \right\} \]

(2.34)

The Fourier analysis of the output waveforms of the SSTPI and FSTPI structures are presented in Equations (2.33) and (2.34). This analysis demonstrates that the FSTPI structure always presents the third harmonic while SSTPI does not have third harmonic. Further, the FSTPI has positive and negative sequences of second and fourth harmonics. The SSTPI presents only the negative sequence of the second harmonics and the positive sequence of the fourth harmonic.

2.5 SIMULATION RESULTS

2.5.1 Design procedure of simulation parameters

The FSTPI is fed by a battery pack. The batteries remain the most difficult element to model, hence an electric equivalent model of the battery pack has been considered. The prediction of dynamic behaviors of batteries is essential to estimate the state of charge (SoC).

The most common technique considered for the identification of battery equivalent electric circuits is the electrochemical impedance spectroscopy. It is a non-destructive method consisting of the superposition of the battery voltage to a low amplitude sinusoidal signal whose frequency is
changed in a given range. The internal impedance of the battery is obtained by the ratio of the voltage to the load current.

The identification of an equivalent electric circuit of battery is very difficult due to the complexity of the diffusion phenomenon. Thus, simple equivalent electric circuit representing batteries have been considered. The scheme considered is shown in Figure 2.8, where \( E_{\text{eq}} \) is the no-load emf, \( R_\Omega \) stands for electrolyte and connection resistances, \( R_{tc} \) is the resistance related to the charge transfer, \( C_{dl} \) is the equivalent capacitance in between the two electrodes and \( Z_w \) is the impedance characterizing the diffusion phenomenon. Such a scheme takes into account both static and dynamic phenomena.

Referring to Figure 2.8, the internal impedance of the battery is reduced to

\[
Z(p) = R_\Omega + \frac{R_{tc}}{1+R_{tc}C_{dl}p} + Z_w(p)
\]

(2.35)

Figure 2.8 Equivalent electric circuit of a battery

\( Z_w \) has been modeled by the following expression:

\[
Z_w(p) = \frac{k_2}{\sqrt{p}} \tanh \left( \frac{k_1}{k_2 \sqrt{p}} \right)
\]

(2.36)

where constants \( k_1 \) and \( k_2 \) are identified by spectroscopy.
The hyperbolic tangent can be expressed in terms of a series of functions which can be written as

$$\tanh(x) = \sum_{n=1}^{\infty} \frac{2x}{x^2 + \left(\frac{n\pi - \frac{\pi}{2}}{2}\right)^2}$$  \hspace{1cm} (2.37)

From the expression 2.36, \(x = \frac{k_1}{k_2} \sqrt{p}\) and \(n=1\) to \(\infty\).

Considering expression (2.37), the equation (2.36) is rewritten as

$$Z_{W}(p) = \sum_{n=1}^{\infty} \frac{1}{2k_1 \left(\frac{n\pi - \frac{\pi}{2}}{2}\right)^2 + k_1 k_2 p}$$  \hspace{1cm} (2.38)

which can expressed as

$$Z_{W}(p) = \sum_{n=1}^{\infty} \frac{1}{R_n + C_n p}$$  \hspace{1cm} (2.39)

where \(R_n\) is a resistance which is computed by \(\frac{8k_1}{(2n-1)^2 \pi^2}\) and \(C_n\) is a capacitance which is computed by \(\frac{k_1}{2k_2^2}\).

The THD presented has been calculated by

$$\text{THD} = \sqrt{T_{\text{hd}a}^2(p) + T_{\text{hd}b}^2(p)}$$

where \(T_{\text{hd}a}(p)\) and \(T_{\text{hd}b}(p)\) are the total harmonic distortion of \(\alpha\) and \(\beta\) voltage components which can be calculated by

$$T_{\text{hd}x}(p) = \frac{100}{a_1} \sum_{i=2}^{p} \sqrt{\left(\frac{a_i}{a_1}\right)^2 \frac{2}{i}}$$  \hspace{1cm} (2.40)

where \(a_1\) is the amplitude of the fundamental, \(a_i\) is the amplitude of \(i^{th}\) harmonic and \(p\) is the highest of harmonics taken into consideration.
The simulation has been done using matlab/simulink and the results are presented. This FSTPI fed drive system consists of a three phase diode bridge rectifier, a split capacitor, FSTPI and three phase squirrel cage induction motor. The parameters of the induction motor are given below.

Stator resistance, $r_s = 1.11 \Omega$, Rotor resistance, $r_r = 1.08 \Omega$

Stator inductance, $l_s = 0.006 \text{H}$, Rotor inductance, $l_r = 0.006 \text{H}$

Mutual inductance, $M = 0.2 \text{H}$

Poles, $P = 2$

Moment of inertia, $J = 0.02 \text{Kg m}^2$

Co-efficient of viscous friction, $F = 0.00575 \text{Nm/ (rad/sec)}$

The parameters of the simulation circuit are

Snubber resistance, $R_s = 500 \Omega$

Snubber capacitance, $C_s = 250\text{e-9}$

The complete simulink block diagram of the system is shown in Figure 2.9. This block diagram consists of two blocks. The block which consists switching pulses generate PWM pulses for four switches of FSTPI at the terminals Out 1 to Out 4. The other block consists FSTPI, induction motor and diode bridge rectifier. Voltage block is used to display the three phase voltage waveforms. The simulation circuit diagram of the drive system is shown in Figure 2.10. This circuit contains three phase input AC power supply, diode bridge rectifier, split capacitor, four MOSFET switches and three phase induction motor. The output line voltage waveform of FSTPI is
shown in Figure 2.11. The THD of FSTPI fed induction motor is found to be 5.00% as seen in Figure 2.12. The simulation circuit diagram of the drive system using SSTPI is shown in Figure 2.13. The output line voltage waveform of SSTPI is shown in Figure 2.14. The THD of SSTPI fed induction motor is found to be 13.96% as seen in Figure 2.15. As a result it is possible to reduce the THD by using FSTPI system.


Figure 2.9 Simulink block diagram of FSTPI fed induction motor
Figure 2.10 The simulation circuit of FSTPI fed induction motor

Figure 2.11 Three Phase Output Voltages
Figure 2.12 FFT Analysis (4 Switch Inverter)

Figure 2.13 The simulation circuit of SSTPI fed induction motor
2.6 EXPERIMENTAL STUDY

The experimental work of FSTPI fed induction motor is developed with 1KW laboratory model of three phase induction motor. ATMEL 89C2051 Microcontroller is used as a main device in this hardware because it changes operating frequency without changing hardware.
2.6.1 Hardware Implementation

In the hardware implementation, the voltage is stepped down by using shunt transformer. The AC voltage is rectified into DC and it is converted into variable frequency AC using an inverter. The pulses received by the MOSFETs are generated by Microcontroller. They are amplified by using driver circuits. The block diagram of hardware is shown in Figure 2.11. Port1 of ATMEL Microcontroller is used for determining the gate pulses. Timer is used for producing the delay required for $T_{on}$ and $T_{off}$ periods. The pulses from the Microcontroller are amplified using IR 2110. It amplifies the pulses to the level of 10V.

![Block Diagram of Hardware Implementation](image)

2.6.2 Pulse Generation Program

The pulse generation program using the 89C2051 ALP is as follows:

```
org 000h
Start :
    mov p3,#09h
    mov Tmod,#01h
    mov T10,#0ffh
    mov th0,#0a5h
    setb tr0
```

---

**Figure 2.16 Block Diagram of Hardware Implementation**
again:
  jnb tf0,again
  clr tf0
  clr tr0
  mov p3,#06h
  mov Tmod,#01h
  mov t10,#0ffh
  mov th0,#0a5h
  setb tr0

again 1:
  jnb tf0,again 1
  clr tf0
  clr tr0
  mov p3,#0Ah
  mov Tmod,#01h
  mov t10,#0ffh
  mov th0,#0a5h
  setb tr0

again 2:
  jnb tf0,again 2
  clr tf0
  clr tr0
  mov p3,#05h
  mov Tmod,#01h
  mov t10,#0ffh
  mov th0,#0a5h
  setb tr0

again 3:
  jnb tf0,again 3
  clr tf0
  clr tr0
  sjmp start
2.6.3 CONTROL CIRCUIT

The control circuit is shown in Figure 2.12. Various elements of the control circuit are explained in the following sections.

![Control Circuit Diagram](image)

Figure 2.17 Control Circuit

2.6.4 MOSFET – IRF 840

Features:

- 28A, 100 V
- \( r_{DS\ (ON)} = 0.0770 \)
- Single Pulse Avalanche Energy Rated
- Nanosecond Switching Speeds
- Linear Transfer Characteristics
- High Input Impedance.
2.7 EXPERIMENTAL RESULTS

Experimental setup is implemented by using a microcontroller which is equipped with power switch control. Port1 of ATMEL Microcontroller is used for generating the gate pulses. Timer is used for producing the delay required for \(T_{on}\) and \(T_{off}\) periods. The pulses from the Microcontroller are amplified using IR 2110.It amplifies the pulses to the level of 10V. An interface board was built to receive the gate drive signal, isolated them and connected to the four switches which were implemented using MOSFET. The output from FSTPI was connected to a three phase induction motor. The three phase induction motor has the follows parameters. Three phase, squirrel cage induction motor 400V, 1KW, \(R_s=7.29\Omega\), \(R_r=8.78\Omega\), \(L_s=3.23\)H, \(L_r=3.23\)H, \(P=2\), \(J_m=0.008\)Kgm\(^2\). The hardware module for the FSTPI fed induction motor is shown in Figure 2.17. Pulses applied to the gate of the MOSFET are shown in Figure 2.18. The waveform in Figure 2.19 demonstrates the output voltage of FSTPI system. The THD of experimental setup is shown in Figure 2.20. The similarity between simulation study and experimental study is revealed in Table 2.2.

### Table 2.2 Comparison between simulation study and experimental study

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value from simulation study</th>
<th>Value from experimental study</th>
</tr>
</thead>
<tbody>
<tr>
<td>Output Voltage</td>
<td>200V</td>
<td>200V</td>
</tr>
<tr>
<td>THD</td>
<td>5.00%</td>
<td>5.84%</td>
</tr>
</tbody>
</table>
Figure 2.19 Experimental Setup of FSTPI Fed Induction Motor System

Figure 2.20 Driving Pulses to the MOSFETs 1 and 3

Figure 2.21 Output of PWM Inverter
2.10 CONCLUSION

The comparison of FSTPI fed induction motor performance with conventional inverter fed induction motor has been discussed in detail in this chapter. From the simulation results, THD of FSTPI system is found to be 5.00% compared to 13.96% for the existing SSTPI system. This comparison shows that FSTPI fed induction motor has reduced THD due to reduced number of switches and reduced switching losses. The validity of the proposed method has been verified by experimental results. The experimental work has been developed using Microcontroller and the results are driving pulses to the switching device, output voltage and THD.