Abstract - This paper presents a Modified DC-DC boost converter for fuel cell vehicle application. In fuel cell vehicles, the output voltage of the fuel cell source is typically much lower than the voltage required by the DC bus and also this output voltage drops significantly as the output current increases. In order to match the output voltage of the fuel cell source to the DC bus voltage, a new DC-DC boost converter with a wide input range and high voltage gain was proposed to act as the required power interface, which reduces voltage stress across the power devices and operates with an acceptable conversion efficiency. Also a novel optimization technique for the closed-loop controller design of a boost type DC-DC converter based on the Particle Swarm Optimization (PSO) algorithm is proposed.

Key Words: Boost DC-DC converter, Fuel Cell Vehicles, High Voltage-Gain, Wide Input Range, Closed Loop, Particle Swarm Optimisation.

1. INTRODUCTION

In recent years, the usage of renewable energy systems such as fuel cell and photovoltaic systems are encouraged due to various environmental troubles caused by other fuels, such as climate change and global warming by increased emissions of carbon dioxide. With increasing attention to environmental problems, energy achieved from the fuel cell systems is focused on the low environmental effects and clean energy. Fuel cells are an effective alternative to replace fuels in emergency power systems and vehicles [1]-[3]. Vehicles powered by fuel cell sources may help to reduce transport's dependence on oil, and reduce polluting emissions. However, unlike batteries which have a fairly constant output voltage, the output voltage of fuel cells drops significantly with an increase of output current. Hence, a step-up DC-DC converter with a wide range of voltage-gain is essential to interface between the low voltage fuel-cell source and the high voltage DC bus of the motor drive inverter.

As shown in Fig.1, each energy source might require a specific DC-DC converter to be integrated into the high voltage (HV) DC link of the power-train. For bidirectional electric sources like super capacitors (SCs) and batteries, bidirectional DC-DC converters are essential to absorb the regenerative braking energy, which maximizes the overall efficiency of the system.

The proposed DC-DC converter with a high voltage gain is interfaced between low voltage (LV) fuel cell source and HV DC bus. When fuel cell vehicle accelerates, SC stacks supply power from DC bus by bi-directional converter (BDC). When fuel cell vehicle brakes or regenerates, energy is absorbed by SC stacks. When fuel cell vehicle runs smoothly, fuel cell source provides stable energy for inverter by proposed converter with corresponding voltage gain and charges SC stacks if needed. In theory, when the duty cycle approaches unity, the conventional boost converter can achieve a high voltage gain [4]. However, it is difficult to implement a high voltage gain (e.g. more than 6), due to the existence of parasitic elements (stray inductance, capacitance) and the extreme duty cycle required. In addition, the power semiconductors suffer from a high voltage stress - the DC bus voltage.

In order to obtain a DC-DC Boost converter with a high voltage gain and a low voltage stress, many different topologies have been proposed by researchers [5]-[8]. Compared with isolated converters, the cost and magnetic losses of non-isolated converters are lower. A high voltage-gain can be achieved by introducing a coupled inductor to topology e.g. [9]-[10], and the converter can maintain a low device voltage stress. The conventional quadratic DC-DC boost converter in [11] can obtain a high voltage-gain, but the voltage stress across the high side power semiconductors is as high as the output-voltage.

To solve this problem, the switched-capacitor configurations introduced in [12] and [13] are able to obtain a high voltage gain, but they cannot achieve flexible voltage regulation unless they are combined with other DC-DC converters [14]. A topology called the “switched-capacitor-based active-network” (SC-ANC) is presented in [15], the power switches may see a large voltage spike as a result of the leakage inductance of the circuit. The switched-capacitor...
circuit was studied in [16]: it achieves flexible voltage regulation by combining it with other DC-DC converters, however the difference in potential between the ground points of the input voltage source side and the load side is a high frequency pulse width modulated (PWM) voltage [17]-[18]. As a result, it may introduce issues associated with du/dt and these may limit its applications [19][20].

The design of the closed loop control parameters is framed as an optimization task such that the PSO algorithm can identify the controller parameters. An appropriate fitness function is then derived for the required objective and is used in the optimization process. The attributes of the large-signal model of the power converter, together with those of the optimization algorithm, provide excellent static and dynamic characteristics at all operating points.

2. PROPOSED SYSTEM

![Fig- 2. Configuration of the proposed system](image)

The configuration of the proposed system is shown in Fig.2. In order to boost the output voltage from the fuel cell source, a DC-DC boost converter is used which is having a wide input range, high voltage gain, acceptable conversion efficiency and reduces voltage stress across power devices. But the major drawback of this converter is the varying output voltage when the input from fuel cell source changes significantly. So a voltage source controller can design in order to achieve suitable static and dynamic performance. And also the closed loop controller parameters are designed by Particle Swarm Optimisation (PSO) technique. Thus the proposed system can act as the required power interface between the fuel cell source and DC bus.

2.1 Configuration of the proposed converter

To address the issues with regard to the conventional converters, a new non-isolated high ratio step-up dc-dc converter is proposed in this paper, which has the following features:

- It reduces the voltage stress across the power devices and has a common ground between the input and output sides.
- The two power switches turn on and off simultaneously. As a result, the control of the converter is simple, and power switches with low on-state resistance can be employed.
- The system operates with a high voltage gain and a wide input voltage range and does not use any extreme values for its duty cycle.

![Fig- 3. Proposed DC-DC boost converter topology](image)

The high voltage gain DC-DC Boost converter is shown in Fig.3. It comprises two active power switches (Q1 and Q2), five power diodes (D3-D7), two inductors (L1 and L2) and five capacitors (C1-C5). The fuel-cell source U_in and the inductor L1 are connected in series to charge capacitors C1 and C2 in parallel. Inductor L2 is another energy storage component which is used to realize a high voltage gain. The ladder type voltage multiplier (capacitors C3-C5 and diodes D5-D7) can improve the voltage gain further and reduces the voltage stress across the power semiconductors on the high voltage side.

2.2. Operating principle of the proposed converter

The gate signals of the two power switches (Q1, Q2) are identical - (Q1, Q2) are turned on and off simultaneously. Therefore, there are two switching states in each switching period, which are shown in Fig.4.

- **Switching state I**: As shown in Fig.4.(a), Q1 and Q2 turn on, L1 is charged by the DC source U_in i.e., (U_inL1−Q1), and L2 is charged by C1 and C2 in series i.e., (C1−L2Q2−C2Q1). Meanwhile, C3 is charged by C5 and C4 in series i.e., (C5−D5−C3−Q2C1).
- **Switching state II**: As shown in Fig.4.(b), Q1 and Q2 turn off, C1 and C2 are charged in parallel by the DC source and L1 i.e., (U_inL1−D1−C1, and U_inL1−C2−D2). At the same time, C5 is charged by the DC source, L3, and L2 in series i.e., (U_in−L1−D2−L3−C3−D7−C6). In addition, C4 and C5 are charged by the DC source, L1, L2, and C3 i.e., (U_in−L1−D3−L2−C4−D7−C6−C1), as well as through the load R. The output-voltage Uo is equal to the total voltages across C4 and C5.
The comparison of the proposed converter with other existing high voltage gain DC-DC Boost converters is shown in Table 1. It can be seen that the proposed converter achieves a high and wide voltage gain range by increasing the number of diodes by a small amount. Compared with the 3-level DC-DC converter, the proposed converter is more suitable for applications requiring a large step-up ratio.

<table>
<thead>
<tr>
<th>Topology</th>
<th>DC-DC Boost Converters</th>
<th>Proposed</th>
</tr>
</thead>
<tbody>
<tr>
<td>No.of switches</td>
<td>2</td>
<td>2</td>
</tr>
<tr>
<td>No.of diodes</td>
<td>2</td>
<td>4</td>
</tr>
<tr>
<td>No.of inductors</td>
<td>2</td>
<td>2</td>
</tr>
<tr>
<td>No.of capacitors</td>
<td>2</td>
<td>2</td>
</tr>
<tr>
<td>Voltage-gain</td>
<td>1/(1-d)²</td>
<td>1/(1-d)</td>
</tr>
<tr>
<td>Switch volt stress</td>
<td>dUo</td>
<td>2Uo/3</td>
</tr>
<tr>
<td>Diode volt stress</td>
<td>U0</td>
<td>2U0/3</td>
</tr>
<tr>
<td>Common ground</td>
<td>Yes</td>
<td>No</td>
</tr>
<tr>
<td>Efficiency</td>
<td>88%−93%</td>
<td>94%−96%</td>
</tr>
</tbody>
</table>

Considering the selection of the power switches, 3-level DC-DC converter will have its maximum device voltage stress (which is higher than half of the output voltage) when d = 0.5, whereas the maximum voltage stress across the power switches is less than half of the output voltage in the proposed converter. Considering the selection of the diodes, the maximum voltage stress across the diodes for the proposed converter is lower than that of the 3-level DC-DC converter. Although the Transformer-less DC-DC Converters has the advantage of the lower voltage stress, it does not have a common ground between the input and the output sides and this may cause additional du/dt issues.

### 3. STATE SPACE REPRESENTATION OF SYSTEM

It is assumed that the power semiconductors, inductors, and capacitors are analyzed for operation under ideal conditions. The average model and the small-signal model can be obtained by using the state-space averaging method. The capacitances are set such that $C_1 = C_2 = C_3 = C_4 = C_5 = C$ to simplify the analysis. The inductances are defined as $L_1$ and $L_2$, the load resistance is $R$, and $u_{di}(t)$, $u_{do}(t)$ and $d$ are the input variable, the output variable and the control variable, respectively. $u_{di}(t)$, $u_{do}(t)$, $u_{c1}(t)$, $u_{c2}(t)$, and $u_{c3}(t)$ are the state variables. According to Fig. 4(a), $C_1$, $C_2$, and $C_3$ are connected in series in the loop circuit when $Q_1$ and $Q_2$ turn on. It means the sum of voltages across $C_1$, $C_2$, and $C_2$ is $U_0$. There is an invalid state variable ($u_{c2}(t) + u_{c3}(t) = 0$, i.e. there are only two independent variables) in this loop circuit. By including the equivalent series resistance (e.g. $r_1 = r = 0.1\Omega$) in the same loop circuit, the coupling between $C_2$ and $C_3$ can be removed to avoid the invalid state variable. Similarly, as shown in Fig. 4(b), $C_1$ and $C_2$ are connected in parallel when $Q_1$ and $Q_2$ turn off, and this means the voltages across $C_1$ and $C_2$ should be equal, i.e. there is another invalid state variable. The coupling relationship between $C_1$ and $C_2$ can also be removed to avoid the invalid state variable ($u_{c1}(t) + u_{c2}(t) = 0$), by including the equivalent series resistance (e.g. $r_2 = r = 0.1\Omega$) in the loop circuits.

The state space model of system can be represented by following state and output equation respectively.

\[ \dot{x} = Ax + Bu \]
\[ y = Cx + Du \]

#### 3.1. Switching state 1

Switches $Q_1$ and $Q_2$ are turned ON:

\[ \frac{di_{L1}(t)}{dt} = \frac{1}{L_1}u_{in}(t) \tag{1} \]

$L_2$ is charged by $C_1$ and $C_2$ in series,

\[ \frac{di_{L2}(t)}{dt} = \frac{1}{L_2}u_{C1}(t) + \frac{1}{L_2}u_{C2}(t) \tag{2} \]

$C_2$, $C_3$, and $C_4$ are connected in series in loop circuit,

\[ u_{c2}(t) + u_{c3}(t) + u_{c4}(t) = 0 \tag{3} \]
So, equivalent series resistance, \( r_1 = r = 0.1 \Omega \) is included in loop circuit.

\[
\frac{di_{L2}(t)}{dt} = -\frac{r}{L_2} i_{L2}(t) + \frac{1}{L_2} u_{C1}(t) + \frac{1}{L_2} u_{C2}(t)
\]

(4)

Current through capacitor \( C_1 \) is,

\[
\frac{du_{C1}(t)}{dt} = -\frac{1}{C_1} i_{L2}(t)
\]

(5)

Current through capacitor \( C_2 \) is,

\[
\frac{du_{C2}(t)}{dt} = -\frac{1}{C_2} u_{C2}(t) - \frac{1}{C_r} u_{C3}(t) - \frac{1}{C_r} u_{C4}(t)
\]

(6)

Capacitor \( C_3 \) is charged by \( C_2 \) and \( C_4 \) in series,

\[
\frac{du_{C3}(t)}{dt} = \frac{1}{C_r} u_{C2}(t) - \frac{1}{C_r} u_{C3}(t) - \frac{1}{C_r} u_{C4}(t)
\]

(7)

Current through capacitor \( C_4 \) is,

\[
\frac{du_{C4}(t)}{dt} = -\frac{1}{C_r} u_{C2}(t) - \frac{1}{C_r} u_{C3}(t) - \frac{1}{C_r} u_{C5}(t)
\]

(8)

Current through capacitor \( C_5 \) is,

\[
\frac{du_{C5}(t)}{dt} = -\frac{1}{C_r} u_{C4}(t) + \frac{1}{C_r} u_{C5}(t)
\]

(9)

The output voltage across \( R \) is the sum of total voltage across \( C_4 \) and \( C_5 \),

\[
u_0(t) = u_{C4}(t) + u_{C5}(t)
\]

(10)

### 3.2. Switching state 2

Switches \( Q_1 \) and \( Q_2 \) are turned OFF:

Capacitors \( C_1 \) and \( C_2 \) are charged in parallel by DC source and \( L_1 \),

\[
\frac{di_{L1}(t)}{dt} = -\frac{1}{L_1} u_{C2}(t) + \frac{1}{L_1} U_{in}(t)
\]

(11)

Capacitors \( C_4 \) is charged by the DC source and \( L_1 \) and \( L_2 \) in series,

\[
\frac{di_{L2}(t)}{dt} = -\frac{1}{L_2} u_{C2}(t) - \frac{1}{L_2} u_{C4}(t)
\]

(12)

\( C_1 \), \( C_2 \) are connected in parallel when \( Q_1 \) and \( Q_2 \) are turned OFF,

\[
u_{C1}(t) + u_{C2}(t) = 0
\]

(13)

So, equivalent series resistance, \( r_1 = r = 0.1 \Omega \) is included in loop circuit.

Current through capacitor \( C_1 \) is,

\[
\frac{du_{C1}(t)}{dt} = -\frac{1}{C_r} u_{C1}(t) + \frac{1}{C_r} u_{C2}(t)
\]

(14)

Current through capacitor \( C_2 \) is,
The voltage relationship between the output and capacitor voltages can be found, in terms of the two switching states which are shown in Fig.4:

\[
\begin{align*}
U_{in} \times dT + (U_{in} - U_{C2}) \times (1 - d)T &= 0 \\
(U_{C1} + U_{C2}) \times dT + (U_{C2} - U_{C4}) \times (1 - d)T &= 0 \\
\end{align*}
\]  

(21)

The voltage relationship between the output and capacitor voltages can be found, in terms of the two switching states which are shown in Fig.4:

\[
\begin{align*}
U_{C1} &= U_{C9} \\
U_{C3} &= U_{C5} = U_{C2} + U_{C4} \\
U_{0} &= U_{C4} + U_{C5} \\
\end{align*}
\]

(22)

As a result, the output voltage \( U_0 \) can be obtained from (21) and (22) as follows:

\[
U_0 = \frac{3 + d}{(1 - d)} \times U_{in} = M \times U_{in}
\]

(23)

where \( M \) is the conversion ratio, i.e. the voltage gain. (23) shows that the proposed converter can theoretically obtain a high and wide voltage gain range.

5. PARAMETER DESIGN

5.1 Design of the power switches and diodes

The design of the power switches and diodes should refer to the most severe conditions that the semiconductor devices will operate in. Assuming that the maximum required voltage gain is 10 and the load power is 400W, the duty cycle \( d \) and the output current \( I_o \) can be obtained as follows:

\[
d = 0.3 \\
U_0 = 400V \\
I_0 = 1A
\]

(24)

It can be deduced that the maximum mean voltage stresses across \( Q_1 \) and \( Q_2 \) are 70V and 166V respectively, and the maximum mean current stresses on \( Q_1 \) and \( Q_2 \) are 16.5A and 6.2A respectively.

The inductances of \( L_1 \) and \( L_2 \) can be derived as (26):

\[
\begin{align*}
L_1 &= \frac{\Delta U_{C3} \times f_s}{(1 - d) \times \Delta U_{C1} \times f_s} \\
L_2 &= \frac{\Delta U_{C5} \times f_s}{(1 - d) \times \Delta U_{C2} \times f_s}
\end{align*}
\]

(26)

Assuming that the maximum acceptable voltage ripple across the capacitor is \( \Delta U_{C} \), the capacitances can be calculated as (27):

\[
C = \frac{dU_{C}}{dt} = \frac{dt}{du_{C}}
\]

(27)

where,

\[
du_{C} = \Delta U_{C} \text{ and } dt = \frac{d}{f_s}
\]

The capacitances of the five capacitors can be calculated as (28):

\[
\begin{align*}
C_1 &= \frac{2dU_{C}}{(1 - d) \times \Delta U_{C1} \times f_s} \\
C_2 &= \frac{(1 + d)U_{C} \times f_s}{(1 - d) \times \Delta U_{C2} \times f_s} \\
C_3 &= \frac{U_{C} \times f_s}{(1 + d)U_{C} \times f_s} \\
C_4 &= \frac{\Delta U_{C3} \times f_s}{U_{C} \times f_s} \\
C_5 &= \frac{\Delta U_{C5} \times f_s}{U_{C} \times f_s}
\end{align*}
\]

(28)

6. RESULTS AND DISCUSSION

The parameters of the experimental converter are listed in Table 2. An adjustable dc source with a range of \( U_{in} = 40V - 120V \) is used to emulate the fuel cell stack source. Hybrid power switches (MOSFETs, IRFP250N and IXTH88N30P) are employed in the low and the high voltage sides, respectively. DSEC60-03A diodes are used on the low voltage side and DPF60IM400HB diodes are used on the high voltage side.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Symbol</th>
<th>Specification</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input Voltage</td>
<td>( U_{in} )</td>
<td>20-80V</td>
</tr>
<tr>
<td>Output Voltage</td>
<td>( U_0 )</td>
<td>400V</td>
</tr>
<tr>
<td>Rated Power</td>
<td>( P )</td>
<td>300W</td>
</tr>
<tr>
<td>Switching frequency</td>
<td>( f_s )</td>
<td>20kHz</td>
</tr>
<tr>
<td>Inductor L1</td>
<td>( L_1 )</td>
<td>330\mu H</td>
</tr>
<tr>
<td>Inductor L2</td>
<td>( L_2 )</td>
<td>1mH</td>
</tr>
<tr>
<td>Electrolytic Capacitor C1/C2</td>
<td>C1/C2</td>
<td>540\mu F</td>
</tr>
<tr>
<td>Film Capacitor C3/C5</td>
<td>C3/C5</td>
<td>20\mu F</td>
</tr>
<tr>
<td>Film Capacitor C4</td>
<td>C4</td>
<td>40\mu F</td>
</tr>
<tr>
<td>Power Switch Q1</td>
<td>Q1</td>
<td>IRFP250N</td>
</tr>
<tr>
<td>Power Switch Q2</td>
<td>Q2</td>
<td>IXTH88N30P</td>
</tr>
<tr>
<td>Diode D3/D4</td>
<td>D3/D4</td>
<td>DSEC60-03A</td>
</tr>
<tr>
<td>Diode D5/D6/D7</td>
<td>D5/D6/D7</td>
<td>DPF60IM400HB</td>
</tr>
</tbody>
</table>

In addition, the switching frequency is 20 kHz, the inductors are \( L_1 = 330\mu H \) and \( L_2 = 1mH \) respectively (the inductances are increased to keep the current continuous), the electrolytic capacitances are \( C_1 = C_2 = 540\mu F \), and the film capacitances are \( C_3 = C_5 = 20\mu F \), \( C_4 = 40\mu F \). The input voltage \( U_{in} \) is variable from 40V to 80V, the reference output voltage
is 400V, and the load resistance is $R = 533\Omega$ (i.e. the rated power=300W).

6.1. Open Loop System

A pulse generator is used to give the required switching pulse shown in Fig.5, with a duty cycle of $d=0.3$. Simulation results of the open loop system is shown in Fig.6. When an input voltage of 60V DC is applied, an output voltage of 400V DC is obtained.

![Switching Pulse](image)

**Fig -5:** Switching Pulse

![Input & Output Voltage Waveform](image)

**Fig -6:** Input & Output Voltage Waveform

![Current through inductor L1 and L2](image)

**Fig -7:** Current through inductor $L_1$ and $L_2$

The inductor current $i_{L1}$ and $i_{L2}$ in the steady state are shown in Fig.7, when $U_{in} = 60\, V$, and $U_{o} = 400\, V$. From Fig.7, it is clear that $i_{L1}$ increases linearly when switch is closed. When switch is open, $i_{L1}$ decreases linearly. The average value of $i_{L1}$ is about 7A while the ripple rate is about 12.5%. Similarly, Fig.7 shows that the inductor current $i_{L2}$ has the same trend as $i_{L1}$ i.e., the average value of $i_{L2}$ is approximately 2.5A.

6.2. Closed Loop System

In the closed-loop control system, first, the actual output voltage $U_0$ is compared with its reference voltage using a comparator and the error voltage, so obtained, is processed by the PI controller. The output voltage of the PI controller is an analog signal which must be converted into a gating pulse for the MOSFET with an adjustable duty cycle. This task is performed by the modulator, which compares the PI controller output voltage with a ramp signal so that the output of the modulator is a gating pulse with its duty cycle varying in accordance with PI controller output voltage.

Closed loop results of input and output voltage of the system are shown in Fig.8. Input voltage is varied at a range of 20V, 30V and 40V and an output voltage of 400V DC is obtained.

![Closed loop input and output voltage waveform](image)

**Fig -8:** Closed loop input and output voltage waveform

6.3. Fuel Cell Stack Implementation

The fuel cell stack is the heart of a fuel cell power system. It generates electricity in the form of direct current (DC) from electro-chemical reactions that take place in the fuel cell. A single fuel cell produces less than 1 V, which is insufficient for most applications. Therefore, individual fuel cells are typically combined in series into a fuel cell stack. A typical fuel cell stack may consist of hundreds of fuel cells.

![Output voltage and output current waveform of fuel cell stack](image)

**Fig -9:** Output voltage and output current waveform of fuel cell stack

The output voltage and output current waveform of fuel cell stack with an output voltage of 60V is shown in Fig.9. Both voltage and current changes its value due to transient state. Voltage decreases initially and remains constant after 10s whereas current increases and remains constant after 10s.
The fuel cell stack parameter variations are shown in Fig.10. From 0 to 10s, flow rate starts increasing and remains constant until it reaches 50lpm. After 10s, flow rate increases up to 90lpm. The voltage and current required for the system are controlled by changing the flow rate of hydrogen and oxygen. As the flow rate is increased, there can be a reduction in utilisation of hydrogen from 100% to 42%. Whereas oxygen utilisation remains constant at 60%. Also, upto 10s, oxygen and hydrogen consumption gradually increases and remains constant. After 10s, its consumption decreases and remains constant. At time, t=0, due to transient state, stack efficiency first increases and then decreases and remains constant. With constant stack consumption, stack efficiency again decreases and remains constant after 10s.

Fig -10: Output voltage waveform of fuel cell stack

7. PARTICLE SWARM OPTIMISATION

The design of the closed loop control parameters framed as an optimization task such that the PSO algorithm can identify the controller parameters. An appropriate fitness function is then derived for the required objective and is used in the optimization process. The attributes of the large-signal model of the power converter, together with those of the optimization algorithm, provide excellent static and dynamic characteristics at all operating points.

Particle swarm optimization (PSO) is a population-based, robust, stochastic optimization technique developed in 1995 by Eberhart and Kennedy, and is inspired by social behavior of fish schooling or bird flocking. PSO is initialized with a number of random agents (particles) that constitute a swarm moving in the search space looking for optima by updating the particles positions during the generations. Each particle represents a candidate solution to the problem at hand. In a PSO system, particles change their positions by flying around in a multi-dimensional search space until a relatively unchanging position has been encountered, or until computational limitations are exceeded.

Boost type DC-DC converters are non-linear systems, and output voltage regulation in these converters using a traditionally derived feedback controller does not yield good dynamic responses at different operating points over the complete operating range. This novel optimization method can be designed to yield a robust, closed loop controller structure with stable static and dynamic characteristics for operating points over the whole operational range of the converter.

In the PSO algorithm shown in Fig.11., the population has n particles and each particle is an m-dimensional vector, where m is the number of optimized parameters. Incorporating the above modifications, the computational flow of PSO technique can be described in the following steps:

- Step 1: Initialize parameters.
- Step 2: Initialize population.
- Step 3: Evaluate fitness value.
- Step 4: Find particle best.
- Step 5: Find global best.
- Step 6: Update velocity.
- Step 7: Update position.
- Step 8: Evaluate.
- Step 9: Repeat steps 4 to 8 until a stopping criteria is met.

Fig -11:PSO Algorithm

7.1. PSO Results

In a PSO system, particles change their positions by flying around a multi-dimensional search space until a relatively unchanging position has been encountered. Fig.12. shows a swarm with n=100, initialised randomly in the entire search space. Swarm is called the cluster of moving particles and ‘n’ is the number of particles.
Evolution of the best values for controller parameters $K_p$ and $K_i$ are shown in Fig. 13. The entire particles moves along the entire search space until it has attained the best value for

$K_p$ and $K_i$. The best position of $K_p$ and $K_i$ is obtained as shown in Fig.14. and the controller constants obtained through standard PSO are: $K_p = 27$ and $K_i = 28$.

7.2. Overall System with Particle Swarm Optimisation

Configuration of overall DC-DC boost converter with PSO consists of input from the fuel cell source, PSO based PI controller and DC-DC boost converter and the output section. Output voltage of overall system is shown in Fig.15. where the input from fuel cell source of 20V to 60V is varied and an output voltage of 400V DC with better response is obtained.

8. PERFORMANCE COMPARISON OF PI CONTROLLER WITH PSO AND ZEIGLER NICHOL’S OPTIMISATION

Fig. 16. shows the step response of the system with Zeigler Nichol’s Optimisation. From the figure the values for $K=0.33$, $L=0.2$ and $T=1.6-L$. According to Zeigler Nichol’s (ZN) Optimisation,

$$T_i = \frac{L}{0.3} \quad (29)$$

$$K_p = \frac{0.9 \times T}{L} \quad (30)$$

$$K_i = \frac{K_p}{T_i} \quad (31)$$

From Eq.(29)-(31), the controller parameters can be calculated as, $K_p = 25.5$; $K_i = 63.6$.

In this paper more emphasis is given to improving the dynamic response of the boost converter by identifying the best values for the controller parameters. The dynamic parameters considered in this paper are rise time ($tr$), settling time ($ts$), peak time ($tp$), peak overshoot ($Mp$), and delay time ($td$).
9. CONCLUSION

A high voltage gain DC-DC Boost converter with a wide input range, continuous input current and common ground points between the input side and the load side has been proposed in this paper. It is suitable for the power interface between a fuel cell source and the DC bus for the motor drive in fuel cell vehicles. The design of a closed loop continuous conduction mode operation converter with PSO Optimization technique can identify the controller parameters. The attributes of the large-signal model of the power converter, together with those of the optimization algorithm, provide excellent static and dynamic characteristics at all operating points.

REFERENCES


Table -3: Performance Comparison

<table>
<thead>
<tr>
<th>Parameter</th>
<th>ZN-PI Controller</th>
<th>PSO-PI Controller</th>
</tr>
</thead>
<tbody>
<tr>
<td>Peak Overshoot (%)</td>
<td>0.7</td>
<td>0.45</td>
</tr>
<tr>
<td>Settling time(s)</td>
<td>0.028</td>
<td>0.025</td>
</tr>
<tr>
<td>Delay Time(s)</td>
<td>0.006</td>
<td>0.004</td>
</tr>
<tr>
<td>Rise Time(s)</td>
<td>0.008</td>
<td>0.006</td>
</tr>
</tbody>
</table>

Fig.17: Zeigler Nichol’s And PSO Comparison


