

COMPARATIVE STUDY BETWEEN SLIDING MODE CONTROLLER AND PI CONTROLLER FOR AC/DC BRIDGELESS CONVERTER UNDER UNCERTAIN PARAMETERS

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ABSTRACT

In this paper, a comparative study between nonlinear sliding mode controller (SMC) and standard PI controller for stabilizing bridgeless AC/DC converter under uncertain parameters is presented. This type of converters is widely used in harvesting low energy systems as in wind turbine, piezoelectric transducers and heat exchange transducers. Designing robust controllers to enhance the efficiency and accuracy of these converters has become a promising track in control engineering field. The traditional bridge rectifier is widely used in the majority types of AC/DC conventional converters to gain the rectified DC voltage from the low AC input voltage source. However, these traditional converters are not effective for the low output voltage of renewable sources due to the voltage drops across rectifier's diodes. The proposed SMC-PI controller is used to enhance the stability and the response of these converters under uncertain parameters comparing with the standard PI controller. The proposed approach consists of both Boost and Buck-Boost converters with two controllers in order to maximize the useful output energy from the source. The graphical method has been used to obtain the limitation of the coefficients of the standard PI controller. A parameter space approach is used to find all robust stabilization PI coefficients and the stability regions. A comparative study using simulations in MATLAB is presented to ensure the effectiveness and robustness of the proposed SMC-PI controller under some external disturbances.

Keywords: AC/DC converters, Boost converter, Buck-Boost converter, Energy harvesting, Sliding mode Controller (SMC), Propotional-Integral controller PI.

1. INTRODUCTION

The incredible growth of the renewable energy sources that generate low AC output voltage (millivolt) has led to enormous research in developing converters to utilize these useful energy [1]. For instance, lots of devices which operate in stabilized DC voltage like sensors used in transport area and wireless communication sets. These devices need to be self-powering with about DC voltage (3 volt) [1], [2]. The increasing demand on such kind of energy sources encourage the researchers in this field to deeply search for effective type of converters that convert the low AC output voltage from these renewable energy sources immediately to a beneficial DC voltage without using the traditional bridge method [13]. The traditional power converters consist mainly from two sequential parts: the diode bridge rectifier and a standard Buck or Boost DC/DC converter. The main disadvantage of this old method is the voltage drop on diodes which leads to decreasing the efficiency of the converter and cut off the converter in low voltage range [1-4]. adding two DC/DC converters in parallel way have an advantage of converting the AC/DC immediately to skip using the traditional bridge [10]. In [1] and [4], Dwari, Dayal and Parsa published a promising method regarding direct AC/DC converter for effective low voltage energy harvesting. The operated converter which used has an efficiency of 63%. The researchers in [5] show double Boost converter for AC/DC power converter where the output DC bus is divided into two stages connected capacitors. Although the converter in [5] has merit, it has disadvantage in requiring large capacitors which leads to slow response in output voltage of the converters. Another architecture of bridgeless AC/DC converter is shown in [6] that demonstrate an efficiency about 71% at 54.5mW. Starting from the

suggested method in [1,4], the authors in this paper present an improvement to the converter and make comparison between the linear and nonlinear controller and determine the limitation of the linear controller coefficients by the graphical method.

The main contributions of this research are that:

- The standard PI controller is designed for the bridgeless converter which consists of the boost and buck-boost converter.
- The nonlinear controller Sliding Mode Control with PI mechanism is illustrated.
- Comparative study between the two controllers to demonstrate the response of the controller.
- Obtain the limitation of the stabilized PI controller coefficients.
- Lastly, checking the response of the controller under changing the load, changing the voltage source, and under tracking the output voltage the desired voltage.

In this paper, detailed analysis of the proposed converters is presented. Closed form relations between the output voltage and the duty cycle with parameters of the converter are derived. As a result, the performance of the proposed controller is tested using MATLAB simulation. This paper is structured as: Section II describes the process modes of the proposed converter. The derivation of the model for the Boost and the Buck-Boost converter is demonstrated in section III. Section IV presents the derivation of both PI and SMC-PI controllers. Section V illustrates the graphical method to obtain the coefficients under uncertain parameters. Section VI presents the results

of the suggested controller under load change, input change and voltage tracking. Section VI presents the conclusion.

In this study, a direct AC/DC converter is proposed in Fig. 1.

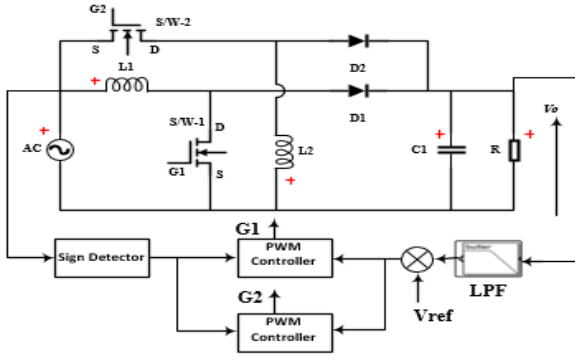


Fig. 1 Suggested AC-DC bridgeless converter

The converter includes Boost and Buck-Boost converter. Both half cycles of the renewable energy source are utilized by the converter. The capacitor is charged continually by the converters during both the positive half cycle and the negative half cycle [14].

2. CONVERTER OPERATION

There are four phases for the converter process [1, 4]. The first two phases 1 and 2, are concerned to the Boost converter which runs on the positive half cycle. The remained two phases 3 and 4 are related to the Buck-Boost converter which runs on the negative half cycle. It is noticeable that over the positive half cycle switch G_1 only operates. however, in the negative cycle the switch G_2 is operating and G_1 is off [11]. In this study, the Boost and the Buck-Boost converters are controlled with the same duty cycle to ensure stable output voltage. Fig. 2, 3, 4, 5 shows the process phases of the converter.

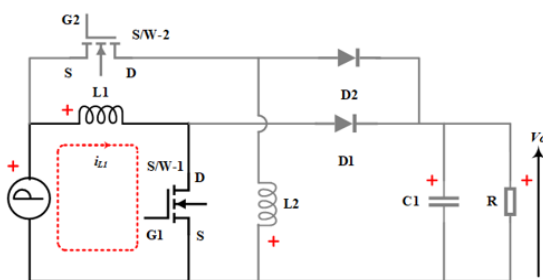


Fig. 2 Phase 1 of the converter

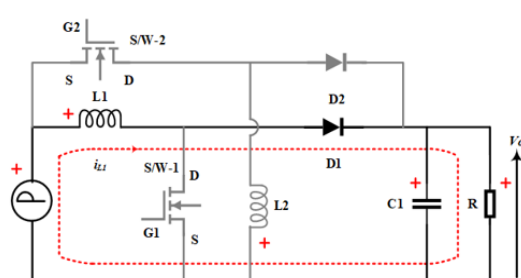


Fig. 3 Phase 2 of the converter

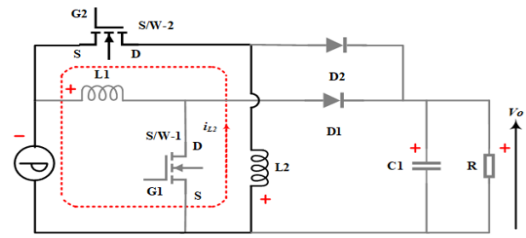


Fig. 4 Phase 3 of the converter

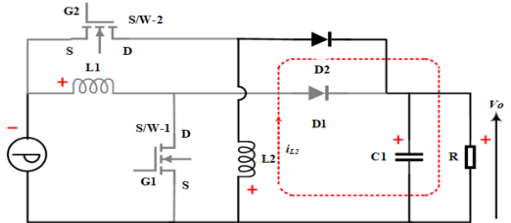


Fig. 5 Phase 4 of the converter

3. CONVERTER MATHEMATICAL MODEL

In this section, a short illustration of the mathematical modelling for the Boost and Buck-Boost converters is described.

3.1. Boost Converter

Generally, the Boost converter job is focused on rising the input voltage to higher voltage. In this section, the state space model and the transfer function are obtained by using the small signal model of the Boost converter. The model is derived in order to use it in the following PI and SMC-PI designing sections.

The two phases of the boost converter [ON, OFF] are described in Fig. 6 and Fig. 7. Besides that, the equations which describe the states of the voltage and currents are also derived.

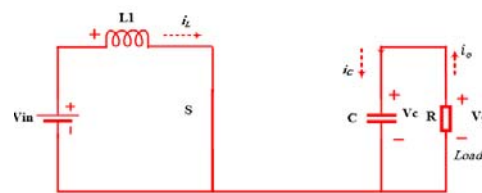


Fig. 6 Boost converter at ON state

$$V_{in} - L \frac{di_L}{dt} = 0 \quad (1)$$

$$i_c + i_o = 0 \quad (2)$$

$$C \frac{dv_c}{dt} + \frac{V_c}{R} = 0 \quad (3)$$

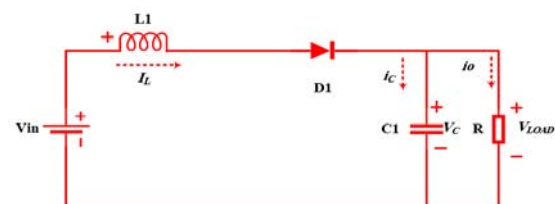


Fig. 7 Boost converter at OFF state

$$V_{in} - \frac{L di_L}{dt} - V_c = 0 \quad (4)$$

$$i_L = i_c + i_o \quad (5)$$

$$i_L = C \frac{dv_c}{dt} + \frac{V_c}{R} = 0 \quad (6)$$

Equation 7 describes the state space model of the Boost converter by the small signal model approach and the below hypothesis: [7].

$$d = D + \hat{d}, \quad v_{in} = V_{in} + \hat{v}_{in} \quad i_L = I_L + \hat{i}_L \quad v_c = V_c + \hat{v}_c$$

$$\begin{bmatrix} \hat{i}_L \\ \hat{v}_c \end{bmatrix} = \begin{bmatrix} 0 & -\frac{(1-D)}{L} \\ \frac{(1-D)}{C} & -\frac{1}{RC} \end{bmatrix} \begin{bmatrix} \hat{i}_L \\ \hat{v}_c \end{bmatrix} + \begin{bmatrix} \frac{V_o}{L} \\ -\frac{I_L}{C} \end{bmatrix} [\hat{d}] \quad (7)$$

3.2. Buck-Boost Converter

Generally, the Buck-Boost converter's job is to rise/down the input voltage depending on the duty cycle value. In this section, the state space model and the transfer function are obtained by using the small signal model of the Buck-Boost converter. This is derived in order to use it in the following PI and SMC-PI designing sections. The two phases of the Buck-Boost converter [ON, OFF] are described in Fig. 8 & 9. Beside that the equations which describe the states of the voltage and currents are also derived.

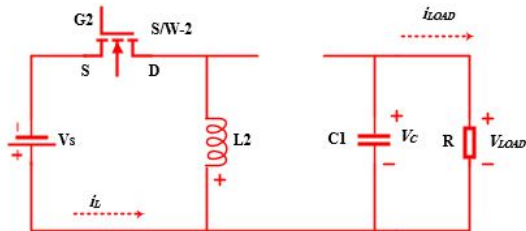


Fig. 8 Buck Boost converter at ON state

$$\frac{dv_c(t)}{dt} = -\frac{v_c(t)}{RC} \quad (8)$$

$$\frac{di_L(t)}{dt} = \frac{v_s(t)}{L} \quad (9)$$

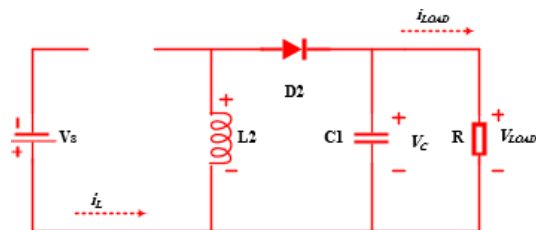


Fig. 9 Buck Boost converter at OFF state

$$\frac{di_L(t)}{dt} = \frac{v_c(t)}{L} = \frac{v_{load}(t)}{L} \quad (10)$$

$$\frac{dv_c(t)}{dt} = -\frac{v_c(t)}{R} - \frac{i_L(t)}{C} \quad (11)$$

Equation 12 describes the state space model of the Buck-Boost converter by the method of small signal model approach and the below hypothesis: [7].

$$d = D + \hat{d}, \quad v_{in} = V_{in} + \hat{v}_{in} \quad i_L = I_L + \hat{i}_L \quad v_c = V_c + \hat{v}_c$$

$$\begin{bmatrix} \hat{i}_L \\ \hat{v}_c \end{bmatrix} = \begin{bmatrix} 0 & \frac{1-D}{L} \\ -\frac{(1-D)}{C} & -\frac{1}{RC} \end{bmatrix} \begin{bmatrix} \hat{i}_L \\ \hat{v}_c \end{bmatrix} + \begin{bmatrix} -\frac{V_c}{L} \\ \frac{I_L}{C} \end{bmatrix} [\hat{d}] \quad (12)$$

4. PROCESS OF DESIGNING THE CONTROLLER

4.1. PI Controller

The principal control equation for the PI controller is specified in this arrange:

$$u(t) = k_p e(t) + k_i \int e(t)$$

where k_p , k_i are the proportional, integral respectively. The coefficients of the proposed PI controller are selected for the optimal values using MATLAB Simulink toolbox. The figure below describes the closed loop converter with PI controller in case of changing the load.

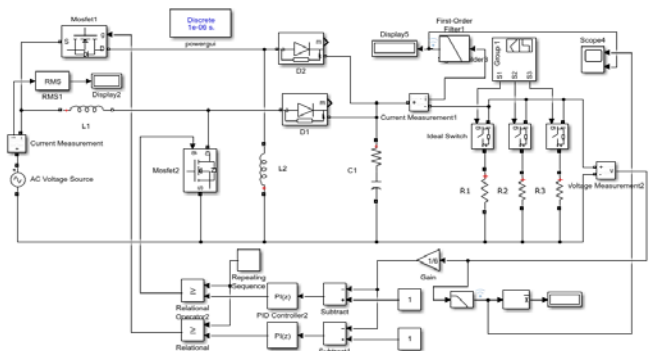


Fig. 10 Converter with different loads

The purpose of sawtooth generator is generating continuous triangle sequence with fixed frequency which allows to control the duty cycle. The comparator block function is used to compare between the PI signal and the sawtooth to set the required duty cycle that runs the mosfet. In addition, the switch block is used to guarantee that Boost and Buck-Boost converters run separately and independently. The 3 resistors are used to change the load in different times in order to check the robustness of the controller.

4.2. Sliding Mode Control (SMC)

Sliding mode control (SMC) is a nonlinear control technique that changes the behavior of the system by applying a discontinuous signal. This signal obliges the system to slide along a surface. The approach is implemented by using high-speed switching control law which obliges the trajectory of the system to move forward prespecified path in the state variable space called (sliding surface) and to stay in that surface thereafter. Before the system reaches the switching surface, there is a control

directed towards the switching surface which is called sliding mode.

Consider the system in the form:

$$\dot{x}(t) = f(x, t, u)$$

Where x is the state vector of the system, u is the control input and f is a function vector. If f is discontinuous on a surface $S(x)=0$ which called sliding surface in the sliding mode theory then:

$$f(x, t, u) = \begin{cases} f^+(x, t, u^+), & s > 0 \\ f^-(x, t, u^-), & s < 0 \end{cases} \quad (13)$$

In sliding control mode, the control purpose is to find a control input u such that the state vector x tracks a desired trajectory x^* in the presence of the model uncertainties and external disturbances. Then the sliding surface can be written as:

$$S(x) = x - x^* \quad (14)$$

Since the objective is to force the system states to the sliding surface, the accredited control plan must guarantee the system trajectory to move toward and stay on the sliding surface from any initial condition if the following condition meets:

$$S\dot{S} \leq \eta|S| \quad (15)$$

where η is the positive constant that guarantees the system trajectories hit the sliding surface in finite time, the required sliding mode controller achieving finite time convergences to the sliding surface is given by:

$$u = \begin{cases} 1 & \text{for } s > 0 \\ 0 & \text{for } s < 0 \end{cases}$$

Let $x_1 = i_l$ and $x_2 = v$ as the states of the boost converter and using the state equations given in equation (7) and (12) now the aim is to achieve a desired constant output voltage V^* . That is, in steady state the output voltage should be the desired voltage V^* . Thus,

$$x_2 = V^* \quad (16)$$

$$\dot{x}_2 = V^* = 0 \quad (17)$$

The state variable error, defined by the difference to the reference value, forms the sliding function:

$$S(x) = x_1 - x_1^* = 0 \quad (18)$$

This means that the control forces the system to slide on the sliding surface. The reference value x_1^* is derived internally to the controller from the output of the linear controller. In order to enforce sliding mode in the manifold $S=0$, the corresponding control signal for the ideal switch for the Boost and Buck-Boost converters as follows:

$$u = \frac{1}{2}(1 - \text{sign}(S)) \quad (19)$$

Since the aim is to guarantee that the state trajectory of the system is directed to the sliding surface $s=0$ and slides over it, this is achieved with a suitable design of control law using the reaching condition:

$$S\dot{S} \leq 0 \quad (20)$$

Since:

$$\dot{S}(x) = \dot{x}_1 - \dot{x}_1^*$$

The steady state values of state variables coincide with the corresponding reference values and they are constants, then

$\dot{x}_1^* = 0$, Replacing equation (18) into equation (20) and solving it, we get:

$$x_2 > V_{in}$$

This means that the sliding mode exists if the output voltage is higher than the source voltage. The closed loop control is necessary to maintain the output voltage when the input voltage has some variations. The analysis of the converter shows that the system dynamics can be divided into fast (current) and slow (voltage) motion. In this study two loop control an inner current control and outer voltage control loop are used. The voltage loop controller is a standard PI type. Since the motion rate of the current is much faster than that of the output voltage, a sliding mode controller is used in the inner current loop. The block diagram of the overall system is shown in Fig. 11 below.

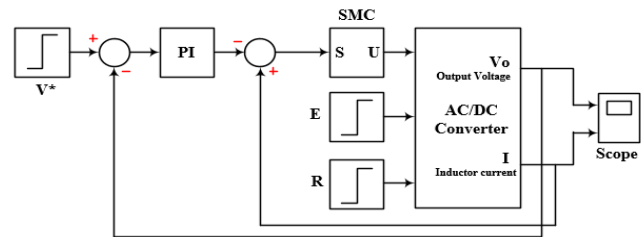


Fig. 11 Nonlinear controller Bloch diagram

The components R and E are used in Fig. 11 to represent the load changes and desired voltage respectively in different times in order to check the robustness of the controller in different cases. The below figure describes the sliding mode controller under changing the input signal.

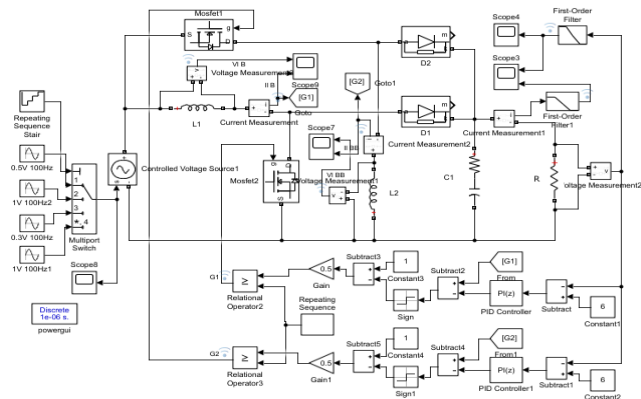


Fig. 12 Converter with SMC-PI Controller under input change

In the SMC-PI controller, the sign block is used to work as on/off switch to attain the required sliding mode in finite time and then multiply the result by 0.5 to secure the compatible control signal.

5. CONTROL OF BOOST CONVERTER WITH UNCERTAIN PAREMETER

In this section, we present the design steps of a robust controller for Linear Time Invariant (LTI) uncertain systems. The type of the uncertain parameters which we will consider is the parametric uncertainties such as resistance of the load and capacitance (R and C). A possible approach in finding the stability regions in the parameter space is the mapping of the stability boundary to the parameter space, this way is very useful if only two parameters are unclear. The main purpose of the graphical method is to find the limitation of PI coefficients that makes the system stable.

By using the pole spread method we can conclude that the system is stable for some poles and unstable for remain poles. We choose the two uncertain parameters (R, C)

Where: $\epsilon \in [3.5, 7.5]$, $C \in [1.5 \times 10^{-4}, 5 \times 10^{-4}]$

The open loop Transfer function for the boost converter with two uncertain parameters is as follows:

$$\frac{\hat{v}_o}{\hat{d}} = \frac{-\left(3.2R(5.73 \times 10^{12}s - 4.05 \times 10^{16})\right)}{(6.87 \times 10^{11}RCs^2 + 6.87 \times 10^{11}s + 8.1 \times 10^{11}R)} \quad (21)$$

The graphical method is a well-known method illustrated in Systems with Uncertain Physical Parameters book by J Ackermann in 1933. It used to determine the region of stability under range of uncertain parameter. Finally, I choose $k_d = 0$ because it's too small value so I neglected it. Actually, I used PI controller and make comparison with SMC-PI controller.

The transfer function of PID controller is as follow:

$$G(s) = \frac{K_d s^2 + K_p s + K_i}{s} \quad (22)$$

So, the closed loop characteristic equation is as follows:

$$P(s) = (3.436 \times 10^{12}CR - 3.20 \times 10^{14}k_d)s^3 + (2.26 \times 10^{18}k_d - 6.87 \times 10^{14}k_p + 3.43 \times 10^{12})s^2 + (1.41 \times 10^{16} - 6.87 \times 10^{14}k_i + 2.26 \times 10^{18}k_p)s + 6.483 \times 10^{17}Rk_i \quad (23)$$

The polynomial family should be belonging to interval class in order to apply Khartivov theorem for stability [15]. For that, the coefficients of the multilinear polynomial can be over bounded in order to convert it to interval one. This can be done by assuming that the coefficients of the multilinear polynomial are independent [16]. The characteristic polynomial in equation (23) is of order 3, hence testing only one polynomial will be enough to guarantee the stability by the sense of Khartivov theorem [16]

$$P^{+-}(s) = a_0^+ + a_1^- s + a_2^- s^2 + a_3^+ s^3 + a_5^+ s^4 \quad (24)$$

a_0^+ is the largest value that could be after constituting by the uncertainty range.

a_1^- is the smallest value that could be after constituting by the uncertainty range. Also a_2^-, a_3^+, a_5^+ is the same

$a_0^+ = 4.86 \times 10^{18}k_i$
$a_1^- = (1.41 \times 10^{16} - 6.87 \times 10^{14}k_i + 2.26 \times 10^{18}k_p)$
$a_2^- = (2.26 \times 10^{18}k_d - 6.87 \times 10^{14}k_p + 3.43 \times 10^{12})$
$a_3^+ = (1.28 \times 10^{10} - 3.20 \times 10^{14}k_d)$

By substitution the given parameters in equation (24) the required polynomial to test the stability is:

$$P^{+-}(s) = (1.28 \times 10^{10} - 3.20 \times 10^{14}k_d)s^3 + (2.26 \times 10^{18}k_d - 6.87 \times 10^{14}k_p + 3.43 \times 10^{12})s^2 + (1.41 \times 10^{16} - 6.87 \times 10^{14}k_i + 2.26 \times 10^{18}k_p)s + 4.86 \times 10^{18}k_i \quad (25)$$

Here, the parameter space approach is used in order to find the stability regions for both of the above polynomials [16].

For P^{+-} the Real and Imaginary parts are:

$$P^{+-}_{even} = -(2.26 \times 10^{18}k_d - 6.87 \times 10^{14}k_p + 3.43 \times 10^{12})w^2 + 4.86 \times 10^{18}k_i \quad (26)$$

$$P^{+-}_{odd} = -(1.28 \times 10^{10} - 3.20 \times 10^{14}k_d)w^3 + (1.41 \times 10^{16} - 6.87 \times 10^{14}k_i + 2.26 \times 10^{18}k_p)w \quad (27)$$

The above two equations are arranged in the matrix form as:

$$\begin{bmatrix} 4.86 \times 10^{18} & -2.26 \times 10^{18}w^2 \\ -6.87 \times 10^{14}w & 3.20 \times 10^{14}w^3 \\ (6.87 \times 10^{14}k_p - 3.43 \times 10^{12})w^2 \\ -(1.28 \times 10^{10})w^3 + (1.41 \times 10^{16} + 2.26 \times 10^{18}k_p)w \end{bmatrix} \begin{bmatrix} K_i \\ K_p \end{bmatrix} + \begin{bmatrix} 0 \\ 0 \\ 0 \\ 0 \end{bmatrix} = 0 \quad (28)$$

It is clear that the equation (28) is constructed of the form $Ax + b = 0$ and the determinant of matrix A should not equal to zero in order to find a solution for it. In this case, the determinant of the matrix A in equation (28) is:

$$\text{Det}(A) = \begin{bmatrix} 4.86 \times 10^{18} & -2.26 \times 10^{18}w^2 \\ -6.87 \times 10^{14}w & 3.20 \times 10^{14}w^3 \end{bmatrix} = 0 \quad (29)$$

The value of the parameter K_p has to be chosen within range to guarantee that the two lines are identical as follows

$$\begin{aligned} \frac{4.86 \times 10^{18}}{-6.87 \times 10^{14}w} &= \frac{-2.26 \times 10^{18}w^2}{3.20 \times 10^{14}w^3} \\ &= \frac{(6.87 \times 10^{14}k_p - 3.43 \times 10^{12})w^2}{-(1.28 \times 10^{10})w^3 + (1.41 \times 10^{16} + 2.26 \times 10^{18}k_p)w} \\ &= 0 \end{aligned} \quad (30)$$

Then,

$$w^2 = \frac{(1.1 \times 10^{37}k_p + 6.9 \times 10^{34})}{(6.5 \times 10^{28} - 4.72 \times 10^{29}k_p)} \quad (31)$$

Since the frequency should be positive, the value of k_p is chosen such that $k_p > -6.27 \times 10^{-3}$. Therefore, $k_p > -6.27 \times 10^{-3}$ is the condition to insure the identicality of the lines in equation (26) and equation (27). Let $k_p = 0.04$ then value of $w = 989$.

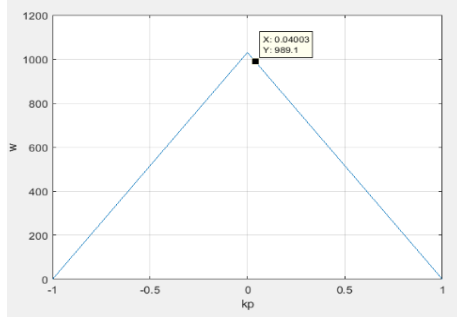


Fig. 13 Plot of K_p as function of w

By Substituting value of k_p in equation (26) and (27), then

$$P_{even}^{+-} = 4.86 \times 10^{18} k_i - (2.26 \times 10^{18} k_d - 2.40 \times 10^{13}) w^2 \quad (32)$$

$$P_{odd}^{+-} = (3.20 \times 10^{14} k_d - 1.28 \times 10^{10}) w^3 - (6.87 \times 10^{14} k_i - 1.04 \times 10^{17}) w \quad (33)$$

Fig. 14 demonstrates the stability regions of both k_i and k_d for P^{+-} where:

1. Real root boundary (RRB) at $w = 0$ is $K_i = 0$ as shown in equation (32)
2. Infinity root boundary (IRB) at $w = \infty$ at $K_d = 0$
3. Complex root boundary (CRB) at $w = 989$. The plot of the line equation below at $0 < w < \infty$

$$1.91 \times 10^{24} k_d + 4.18 \times 10^{18} k_i = -9.27 \times 10^{19} \quad (34)$$

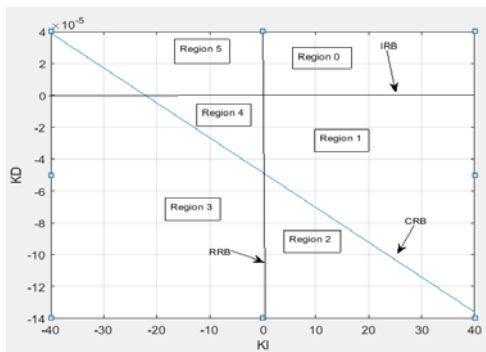


Fig. 14 $K_i - K_d$ plane with boundaries

The curves in Fig. 14 divide the plot into six different regions. The Boundary Crossing Theorem indicates that if one of these regions contains a stable polynomial, then that region must contain only stable polynomials. Conversely, if one of these regions contains an unstable polynomial, then that region must contain only unstable polynomials. Hence, by picking one polynomial for each region and testing its stability, by applying this step to the graph in Fig. 14, it's obvious that there is only one stable region.

For region 0 let $K_i = 5$, $K_d = 2 \times 10^{-5}$ then the close loop TF:

$$P(s) = 2.43 \times 10^{19} s^3 + 1.02 \times 10^{17} s^2 + 2.13 \times 10^{13} s + 6.47 \times 10^9$$

And the roots of the polynomial are:

$$\begin{aligned} & -1.02 \times 10^{-4} + 2.38 \times 10^{-4} i \\ & -1.02 \times 10^{-4} - 2.38 \times 10^{-4} i \\ & -0.00397 \end{aligned}$$

We note that region 0 has all poles at LHP, Take $K_i = 3$, $K_d = 2 \times 10^{-5}$, $K_p = 0.04$. We will neglect K_d because it's too small.

6. SIMULATION RESULTS

The converter is designed depending on Texas instrument reference [9] in CCM. η is the efficiency of power conversion. Normally, it is suitable to calculate the inductor peak-to-peak current of less than 40% of the average value of inductor current. Therefore, $\Delta I_{L(p-p)}$ current is supposed to be less than 40 % of $I_{L(DC)}$.

Table 1 Designing the parameters of the converter

Converter elements	part	Rate
Input Voltage	V_{in}	0.5 V
Output voltage	V_o	8 V
Output Current	I_{out}	1.5 A
Inductor	L_1, L_2	$2.37 \times 10^{-6} H$
Load resistance	R	5.5 Ω
Capacitor	C	$2.06 \times 10^{-4} F$
Frequency	F_{sw}	25 KHz

Table 2 Calculated Converter component values

$D = 1 - \frac{(V_{in} \times \eta)}{V_o}$	$D = 1 - \frac{(0.5 \times 0.90)}{6} = 92.5 \%$
$I_{L(DC)} = \frac{(V_o \times I_{out})}{V_{in} \times \eta}$	$I_{L(DC)} = \frac{8 \times 1.5}{0.5 \times 0.9} = 26.66 A$
$\Delta I_{L(p-p)} = \frac{30}{100} \times I_{L(DC)}$	$\Delta I_{L(p-p)} = \frac{30}{100} \times 26.66 = 8 A$
$I_{L(p)} = I_{L(DC)} + \frac{\Delta I_{L(p-p)}}{2}$	$I_{L(p)} = 26.66 + \frac{8}{2} = 30.66 A$
$\Delta I_{L(p-p)} = \frac{(V_{in} \times D)}{L \times F_{sw}}$	$L = \frac{V_{in} \times D}{\Delta I_{L(p-p)} \times F_{sw}} = 2.37 \mu H$
$V_{ripple} = I_{L(p)} \times R_{ESR}$	$V_{ripple} = 30.66 \times 0.003 = 9.19 \%$
$C_{out} = \frac{(D_{max} \times I_{out})}{F_{sw} \times V_{ripple}}$	$C_{out} = \frac{0.95 \times 1.5}{25000 \times 0.09198} = 2.06 \times 10^{-4} F$
$R_{load} = \frac{V_{out}}{I_{out}}$	$R_{load} = \frac{8}{1.5} = 5.5 \Omega$

Load variation: different loads have been applied (5.5 -20-5.5) ohm at different times (0.1, 0.2) seconds. Fig. 15, 16, 17 describe the response for each controller.

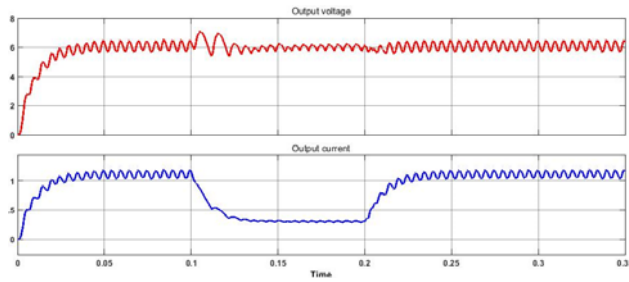


Fig. 15 Output response of PI controller for load change

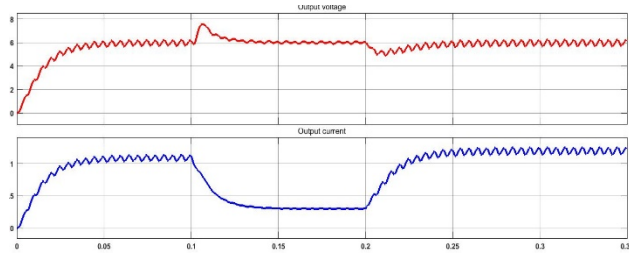


Fig. 16 Output response of SMC-PI controller for load change

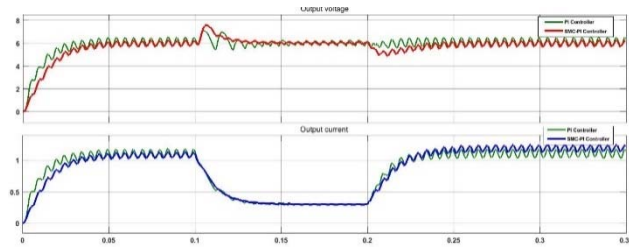


Fig. 17 Compare between the controllers with load variation

Its noted here that SMC has less ripple in the output voltage with less overshoot compared with PI controller. SMC has also less settling time. We notice above that SMC-PI has less chattering than PI contrary to custom and this happened because we used high order sliding mode (HOSM) algorithm where the system slides on the manifolds $\dot{S} = 0$ and $S = 0$. This is the best alternative to obtain continuous and "smooth enough" control signals. There is trade-off between overshoot and steady state error.

Input Variation: we implement a variable input with 0.5V to 0.8V at different times to test the stability of the output voltage under this disturbance. Fig. 18, 19, 20 illustrate the response in each controller with input variation.

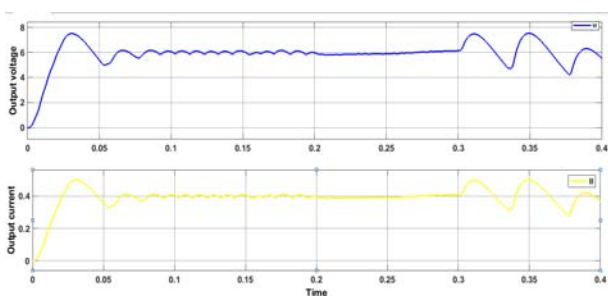


Fig. 18 Output response of PI controller for input change

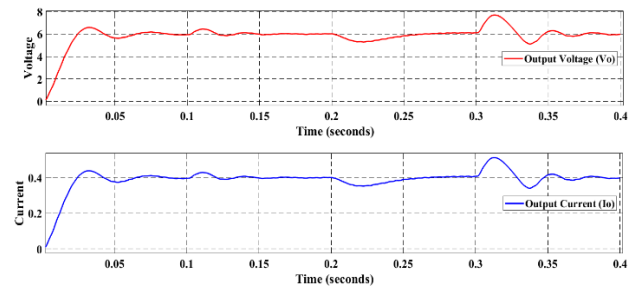


Fig. 19 Converter with SMC-PI Controller (Input variation)

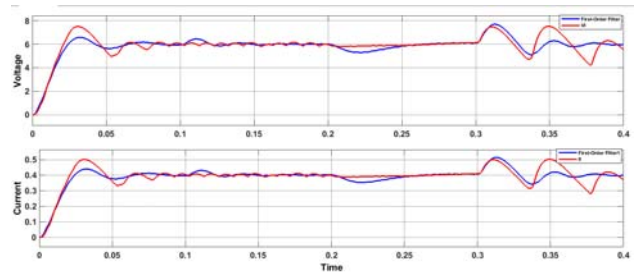


Fig. 20 Comparison between PI and SMC with Input variation

Its noted here that SMC has no ripple in the output voltage with less overshoot compared with PI controller. SMC has also less settling time.

Voltage Tracking: Variable desired output voltages (6V, 8V, 4V) are implemented to observe the response of system. Fig. 21, 22, 23 show the response in each controller for voltage tracking.

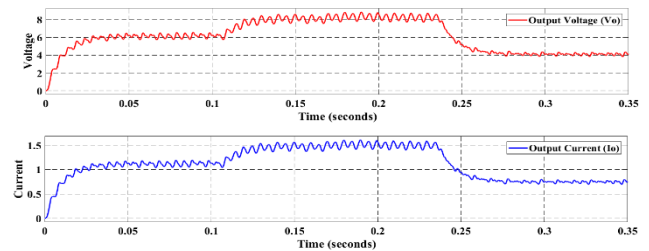


Fig. 21 Output response of PI controller for voltage tracking

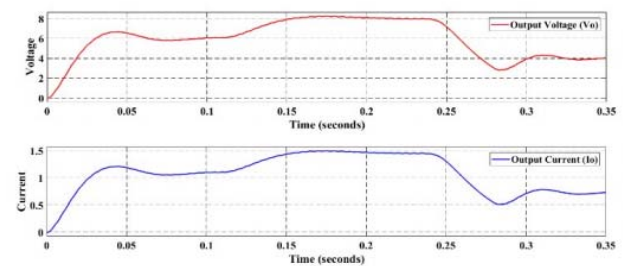


Fig. 22 Output response of SMC-PI controller for voltage tracking

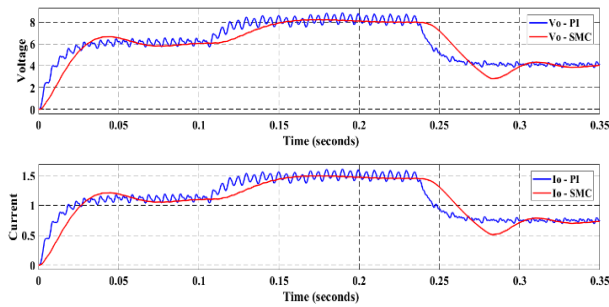


Fig. 23 Comparison between PI and SMC with Tracking voltage

The last figure shows that SMC-PI has no ripple and both of the two controllers have the same settling and rising time. The comparisons illustrate the positive behaviour of the proposed controller in stabilizing the output voltage within required voltage of value 6V under input voltage variations, load variation and tracking voltage.

Step response for the boost converter:

After selecting two random points in the same region 0, we take two points:

Point 1: $K_p = 0.04$, $K_i = 3$

Point 2: $K_p = 0.04$, $K_i = 7$

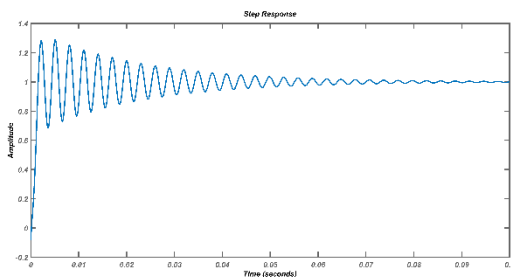


Fig. 24 Step response for point 1 in the region

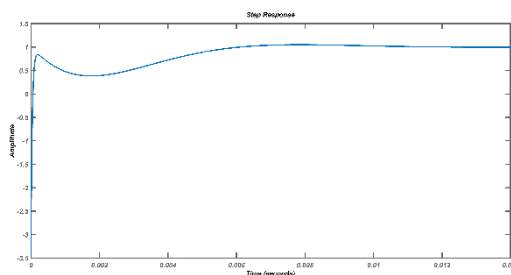


Fig. 25 Step response for point 2 in the region

Table 3 Comparison between two points in region 0

Criteria	Point 1	Point 2
Overshoot	28.78%	14.1 %
Steady state error	0.0026	0.0026
Settling time	0.0604	0.0056
Rising time	0.0012	9.454×10^{-5}

After we make a comparison between the two points in the same region, we conclude that at different points we have different responses. So, to optimize the optimal point which gives us the best performance in the response this relates to adaptive and optimal control courses which will lead to the optimal point. The value of K_p , K_i control the performance of the system as overshoot, settling time. Big Bang-Big Crunch method and genetic algorithm are ways used to optimise the range of the coefficients. Each point has different performance index.

The efficiency of the system:

The efficiency for the converter is tested with the exact parameters for both controllers as shown below.

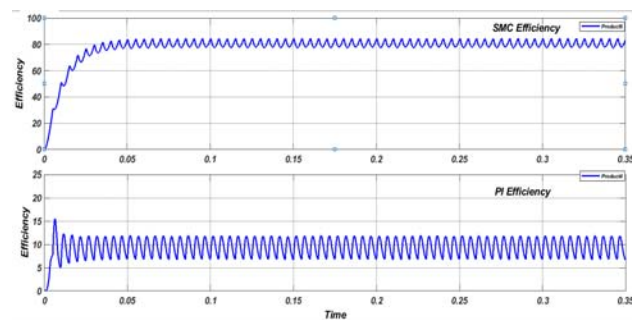


Fig. 26 Efficiency comparison between PI and PI-SMC

This figure above obviously shows that PI-SMC is better than PI controller at the same parameters value with 82 % for PI-SMC controller and about 15% for PI controller. It's incomparable to compare the two controllers.

7. CONCLUSION

The suggested SMC-PI direct AC/DC low voltage energy-harvesting converter overcome the traditional bridge rectification and attain more efficiency than the PI controller with efficiency about 82%. The suggested controller approach demands two separate controllers. Particular investigation of the immediate AC/DC power

converter is accomplished. Design instructions are offered for choosing values of the basic elements and the parameters of the controller for the converter. PI and SMC-PI controllers are tested and results demonstrate that SMC-PI has more efficiency and performance than PI. The specification of output voltage ripple, settling time, rising time and steady state error in SMC-PI is higher than in PI controller. The efficiency reaches about 82% for PI-SMC. In addition to that, the simulation results display the robustness of the propose controller under disturbances. As future studying, the writers will apply the proposed controllers to piezoelectric low voltage source applications as in order to approve it in real scenario. Moreover, the authors will apply another parametric uncertainty and change the type of the polynomial which will lead to complex solution to find the range of the solution. Moreover, applying recent optimization methods as Big Bang-Big Crunch method and genetic algorithm to find the range of the coefficients.

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