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Single –Stage Multi Input DC-Dc/AC Converter Using SPWM Control

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Abstract*: This paper proposes a new extendable single stage multi-input dc-dc/ac boost converter. This proposed structure comprises two current bidirectional ports in the converter central part to interface battery and load in order to provide reliability in the supply. This proposed topology consists of two set of parallel boost converters, which are actively controlled to produce two independent output voltage components. Choosing two pure dc or two dc biased sinusoidal values as the converter reference voltages, situations of the converter operating in two dc-dc and dc-ac modes are provided. The proposed converter utilizes minimum number of power switches and is able to step up the low-level input dc voltages into a high-level output dc or ac voltage with-out needing any output filter arrangement. In order obtained desired output voltage components for the converter by the resulting in autonomously charging and discharging of the battery to balance the power flow.
***Keywords*: boost converter, filters,**

I. INTRODUCTION

Energy sources like wind turbines and photovoltaic (PV) systems are intermittent and unpredictable; therefore, they are not highly reliable. In order to address this issue, renewable resources are either combined with each other or fuel cells (FCs) and energy storage systems (ESSs). Nowadays, the concept of multiple-input converters (MICs) has been proposed to accommodate several renewable energy resources. These converters provide simple circuit topologies, centralized control, high reliability, low manufacturing cost, and size. The systematic approaches of generating and synthesizing MICs have been introduced. In general, all various MICs are responsible to accept dc voltage sources at their input ports, while from the output point of view they can be broken into two categories: dc–dc MICs and dc–ac MICs.

For the first category, in three multi-input converters have been proposed in [1]- [2] based on the structure of the dc boost converter. The three-input dc–dc boost converter proposed by the author's benefits from simple unified structure and minimum numbers of power switches.

In this paper, a new extendable single-stage multi-input boost converter is proposed in [2]-[4] which can operate in both dc–dc and dc–ac modes. The proposed converter integrates two bidirectional ports for battery storage and output load and several unidirectional ports for different input dc sources. Although dc–ac MICs gain several advantages, all of them utilize at least two conversion stages and need an output filter to support ac loads. Accordingly, for all MICs, multiple power conversion stages may result in increasing the count of devices, system power losses, size, weight, and cost of hybrid systems. In this regard, single-stage topologies, which integrate performance of each stage of multistage power converters, are becoming more attractive. Although they may cause control complexity, they offer higher efficiency and reliability, and lower cost and size. , A single stage single-input z-source inverter with coupled inductor, which is able to step-up low-level input dc voltages into a high-level output ac voltage, has been proposed. If two dc sources are embedded in the x-shape impedance network of the z-source inverter. This circuit lacks enough degrees of freedom for the control system and does not provide acceptable current ripple for the input sources. Conventional single-stage boost inverter has been utilized by the authors as a series grid-connected single-stage PV inverter, which also plays the role of grid voltage compensator for an a.c load. Unfortunately, all single-stage inverters are not multi input and therefore not suitable for hybrid systems. As discussed previously, it can be said that for multisource systems, better performances can be accessed providing that a single stage multi-input converter is utilized. Such a converter becomes more attractive if it can operate in both dc–dc and dc–ac modes, without any changes in the converter structure. This can be proposed in [5]-[6] so multi-input converter is able to directly step up the low-level input dc voltages into a high-level output dc/ac voltage and does not need any output filter, while it only uses the minimum number of power switches. The load voltage is differentially obtained from the converter two output voltage components. For dc–dc mode operation, these voltages are regulated at two different dc values, while in dc–ac mode they are

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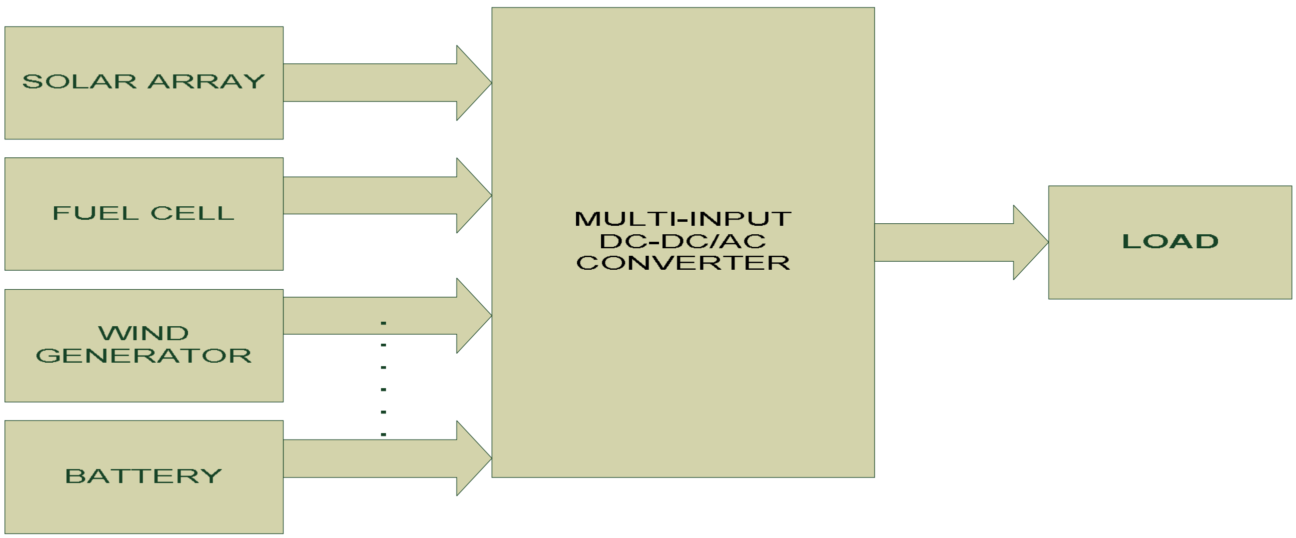
controlled to be two 180◦ out-of phase dc-biased sinusoidal voltages. Owing to the multi input structure of the converter, the integral state feedback strategy of multi-input multi
output(MIMO) control systems is applied to the converter small signal model to achieve its control laws. As a result, several input current regulator loops for the input dc sources and two output voltage control loops are designed for the proposed converter. Finally, the effectiveness of the proposed converter and its control performance are studied and verified by simulation and experimental results. In comparison with other multi-input double-stage dc–ac converters, the proposed converter applies only one power conversion stage and also is able to operate in both dc–dc and dc–ac modes without using any output filter. To be more precise, it utilizes less number of power switches and passive elements. Moreover, compared to a system that utilizes the same number of active and or passive elements, the proposed system has superiority over minimizing voltage ratings and sizes of the elements.

BLOCK DIAGRAM



The MICs proposed is essentially based on parallel connection at the output of a number of boost converters and buck–boost converters. Such MICs do not enjoy the advantage of reduced device and element counts. However, the multiple-input isolated full bridge boost converter and multiple-input half-bridge boost converter share the output rectifier through a multiple-winding transformer. A systematic approach for synthesizing MIC topologies was reported. The concept of the pulsating voltage-source cell (PVSC) and the pulsating current-source cell (PCSC) were introduced and these pulsating source cells (PSCs) were extracted from the six basic non isolated converters, including buck, boost, buck–boost, Cuk, Zeta, and SEPIC converters. The presented approach was to insert these PSCs into the six basic non isolated converters. The input sources of the generated non isolated MICs can transfer energy to the load individually or simultaneously. Using this approach, the resulting topologies do not provide isolation, and the topologies with time multiplexing control scheme are not considered.

In this paper, a new VSI is proposed, referred to as boost inverter, which naturally generates an output ac voltage lower or larger than the input dc voltage depending on the duty cycle. The proposed boost inverter achieves dc–ac conversion, as indicated in the figure, by connecting the load differentially across two dc–dc converters and modulating the dc–dc converter output voltages sinusoidally. This concept has been discussed; using the Cuk converter. The blocks A and B represent dc–dc converters. These converters produce a dc-biased sine wave output, so that each source only produces a unipolar voltage. The modulation of each converter is 180 out of phase with the other, which maximizes the voltage excursion across the load. The load is connected differentially across the converters. Thus, whereas a dc bias appears at each end of the load, with respect to ground, the differential dc voltage across the load is zero. The generating bipolar voltage at output is solved by a push–pull arrangement. Thus, the dc–dc converters need to be current bidirectional.

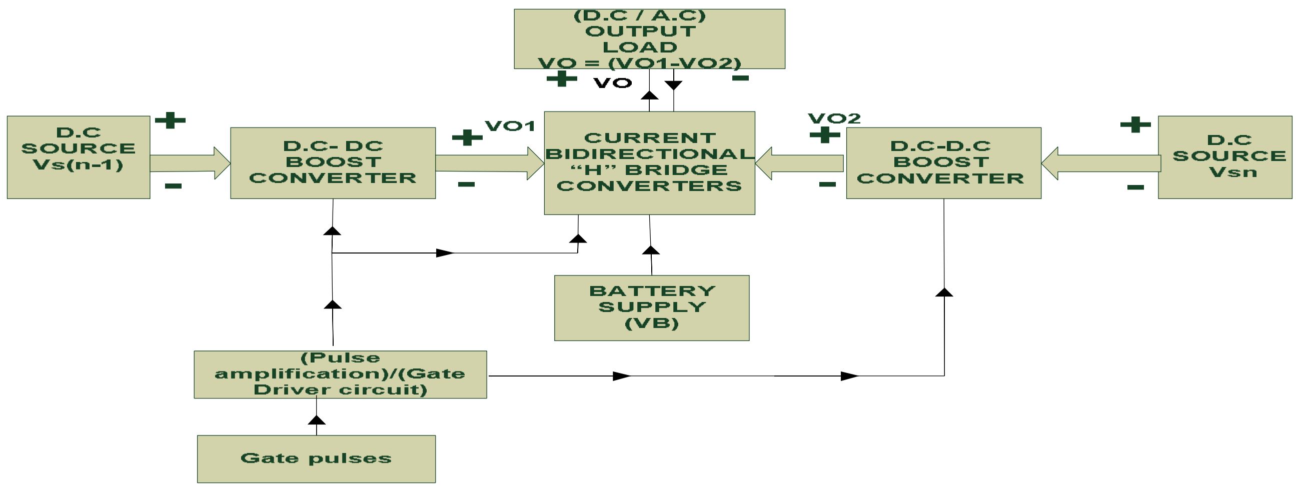
The voltage gain, for the boost inverter, can be derived as follows: assuming that the two converters are 180 out of phase the output voltage is given by

$$Vo = V1 - V2 = \frac{Vin}{1-D} - \frac{Vin}{D}$$

$$\frac{Vo}{Vin} = (2D - 1)/D(1 - D)$$

The gain characteristic of the boost inverter is shown in the figure interesting to note that the feature of zero output voltage is obtained for D=0.5. If the duty cycle is varied around this point, then there will be an a.c voltage at the output terminal For the purpose of optimizing the boost inverter dynamics, while ensuring correct operation in any working condition, a sliding mode controller is a more feasible approach. Sliding mode control has been presented as a good alternative to the control of switching power converters. The main advantage over the classical control schemes is its insusceptibility to plant parameter variations that leads to invariant dynamics and steady-state response in the ideal case. In this paper, a sliding mode controller for the boost inverter is proposed.

II. PROPOSED SYSTEM



For the proposed system, the single power switch of each unidirectional boost converter is controlled to regulate the dc power of its corresponding input source, while in the converter central part both upper and lower power switches of the bidirectional boost converters are complementary switched to produce their corresponding output reference voltages. Thus, n different duty ratios (d1 , . . . , and dn) are introduced to control the converter power switches S1, S1*, S2, S2*, S3 , . . . , and Sn, respectively. These duty ratios are the converter controlling variables that facilitate current and voltage regulation goals of the proposed system. For the proposed system, the input dc sources Vi3 , . . . , Vin should deliver n-2 ripple free dc currents, i.e., IL3 , . . . , ILn, respectively. Thus, the total generated dc powers at the both sides of the converter are expressed as follows:

$$Pi1 - \sum_{k-2}^{n/2} Vi(2k-1)\, IL(2k-1), \qquad Pi2 - \sum_{k-2}^{n/2} Vi2k\, IL2k$$

In this mode initially we assume capacitor C1 had fully charged through diodes D3 and D5 since switches S3 and S5 are not triggered with proper gate-pulses so the entire input voltage is charged the capacitor C1 similarly this type of operation also applicapable to capacitor C2 and switches S2 and S4 now the switches S1 and S3 are triggered so that inductors L3 and L5 charging takes place in the form of current with the direction for L3 is source voltage Vi3 ,inductor L3 and switch S3 and return to the source(Vi3) for L5 direction of current Source Vi5, inductor L5 and switch S5 and return back to the source (Vi5) in this condition the already charged voltage in the capacitor C1 will discharged to the load via the current bi-directional bridge converter in the proposed converter central part to interface load and battery. Now the switches S3 and S5 are off condition then the charged current in the inductor (L3) will discharging takes place as follows in this direction source(Vi3), inductor (L3) and through diode D3 to capacitor C1 and return back to the source(Vi3) similarly L5 also as follows source (Vi5), inductor(L5) and through diode D5 to capacitor C1 and return back to the supply (Vi5) same operation also valid for another parallel boost converters in the converter second part which charging and dis-charging of the capacitor C2 with help of battery and bi-directional h-bridge converter in the central part depending up on the load requirement . If two different dc values are chosen as the converter output reference voltages, then a pure dc voltage appears across the output load as follows:

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$$Vo1 - V1, Vo2 - V2, Vo - V1 - V2$$

For an output resistive load the consumed power can be expressed by the following equations:

$$Io = (V1 - V2)/(R)$$

$$Po = VoIo = \frac{(V1^2 - 2V1V2 - V2^2)}{R}$$

Now, we obtain the converter first- and second-side powers Po1 and Po2 delivered to the load as follows

$$Po1 = Vo1Io = (V1^2 - V1V2)/(R)$$

$$Po2 = -Vo2Io = -(V1^2 - V1V2)/(R)$$

As seen in (4), first side of the converter delivers the positive power amount P_{o1}, while the negative power amount of Po2 is absorbed by the converter second side. These powers are constant values that only depend on the converter both side voltages. By neglecting converter power losses, the battery's both sides powers are obtained as follows:

$$Pb1 - Po1 - Pi1 - \frac{V1^2 - V1V2}{R} - \sum_{k=2}^{n/2} Vi(2k-1)IL(2k-1)$$

$$Pb2 - Po2 - Pi2 - -\frac{V1V2 - V2^2}{R} - \sum_{k=2}^{n/2} Vi2k\ Il2k$$

The summation of *Pb*1 and *Pb*2 gives the battery total exchanged power as follows:

$$Pb = Pb1 + Pb2 = Po - \sum_{k=3}^{n} VikILk$$

As (6) shows, the battery source supplies or absorbs the power difference between the load power and the total generated input power. The battery current *Ib* and its both side inductor currents *IL*1 and *IL*2 are determined as follows:

$$IL1 = \frac{Pb1}{Vb} = \frac{Po1 - Po2}{Vb}$$

$$Ib = \frac{Po}{Vb} - \sum_{k=3}^{n}(Vik\ ILk)/(Vb)$$

$$IL2 = \frac{Pb2}{Vb} = \frac{Po1 - Po2}{Vb}$$

In this mode of operation the switches S3,S5 and S4,S6 are remains unaltered but the converter central part that is current bi-directional h- bridge converter switches S1,S1* and S2,S2* are triggered with uni-polar sinusoidal pwm pulses which are 180 degrees out of phase so S1 and S2* are complementary switches similarly S2 and S2* are complementary switches and the battery will generate two dc biased sinusoidal components these are 180 degrees out of phase with the help of switches S1,S1*and S2,S2* and inductors L1 and L2 suppose switch S1* is OFF condition then the switch S1 is ON so the inductor L1 will be charging the current through switch S1 and battery (Ib) it is dc biased sinusoidal value when the switch S1* is ON so that the charging current in the inductor (L1) will be discharged to the load at the same capacitor C1 also discharged to the load now the load is sinusoidally varied dc quantity and the capacitor C2 will charging takes place via the R load and the battery current (Ib) will return back to battery (Vb) through capacitor C2.

Now the switch S2* OFF condition then the switch S2 is ON so the inductor L2 will charging the current through switch S2 and battery (Ib) it is also a dc biased sinusoidal value but it is 180 degree out of phase nature when the switch S2* is ON so that the charging current in the inductor (L2) will be discharged to the load at the same capacitor C2 also discharged to the load now the load is sinusoidally varied dc quantity and this is 180 degrees out-of phase compared to the C1 discharged quantity in the previous case since both S1* and S2* triggered with the uni-polar sinusoidal pwm pulses which are 180 degrees out-of phase nature and the capacitor C1 will charging takes place via the load and the battery current (Ib) will return to the battery (Vb) through the capacitor C1 .

If the converter reference voltages are chosen as (8), then the proposed converter will operate in the dc–ac mode as follows.

$$Vo1(t) = Vdc + (Vm\sin(wt))/2$$

$$Vo2(t) = Vdc - (Vm\sin(wt))/2$$

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where their dc parts are the same as Vdc and the modulation of each sinusoidal part is 180 degrees out of phase with the other one. This concept results in generating a pure sinusoidal voltage across the load as follows:

$$Vo = Vo1(t) - Vo2(t) = Vm\sin(wt)$$

Instantaneous current and power of an output resistive load can be expressed by the following equations:

$$Io(t) = Im\sin(wt),\ Im = \frac{Vm}{R}$$

$$Po(t) = Vo(t)Io(t) = Po^{\sim}o + Po^{-} = -\frac{(VmIm)}{2}\cos(2wt) + [\frac{VmIm}{2}]$$

In equation (10), the average quantity (Vm *Im)/2 corresponds to the load average power ¯ Po , while the alternative term at the angular frequency of 2ω denotes the pulsation component of the load power. Now, the converter first- and second-sides instantaneous powers delivered to the load are obtained as follows:

$$Po1 = Vo1(t)Io(t) = Vdc\,Im\sin(wt) + \frac{VmIm}{2}\sin(wt)^2$$

$$= Vdc\,Im\sin(wt) - (Po^{-}\cos(2wt) + [\frac{Po^{-}}{2}]$$

$$Po2 = -Vo2(t)Io(t) = -Vdc\,Im\sin(wt) - \frac{Po^{-}}{2}\cos(2wt) + \left[\frac{Po^{-}}{2}\right]$$

As the average powers delivered to the load at the both sides of the converter are the same and equal to the half of the load average power for all power operation conditions. Also, the instantaneous powers associated with the converter capacitors $C1$ and $C2$ ($XC1 = 1/C1\omega$ and $XC2 = 1/C2\omega$) are obtained as follows.

$$Pc1 = (Vo1 * Ic1) = (Vdc + \frac{Vm}{2}\sin(2wt))\ \left(\frac{C1d}{dt}\right)\left(Vdc + \frac{Vm}{2}\sin(wt)\right)$$

$$= \frac{VdcVm}{2XC2}\cos(wt) + \frac{Vm^2}{8XC2}\sin(2wt)$$

$$Pc2 = (Vo2 * Ic2) = -\frac{VdcVm\cos(wt)}{2XC2} + \frac{Vm^2\sin(2wt)}{8XC2}$$

Neglecting the converter power losses and the stored energy in the converter inductors, the battery's both side powers are obtained as follows

$$Pb1 = Po1 + Pc1 - Pi1 = (Vdc * Im\sin(wt)) + (Vdc * Vm\cos(wt))/\ (2XC1)$$

$$-\frac{Po^{-}}{2}\cos(2wt) + \frac{Vm^2 * \sin(2wt)}{8XC1} + [\left(\frac{Po^{-}}{2}\right)\ \ \Sigma_{k=2}^{\frac{n}{2}}(Vi(2k\ \ 1) * IL(2k\ \ 1)]$$

$$Pb2 = Po2 + Pc2 - Pi2 = -(Vdc * Im\sin(wt)) - (Vdc * Vm\cos(wt))/\ (2XC2)$$

$$-\frac{Po}{2}\cos(2wt) + \frac{Vm^2 * \sin(2wt)}{8XC2} + [\left(\frac{Po^{-}}{2}\right) - \Sigma_{k=2}^{\frac{n}{2}}(Vi(2k) * IL(2k)]$$

Summing $Pb1$ and $Pb2$ and assuming $XC1 = XC2 = XC$ gives the total instantaneous battery power as follows:

$$Pb = \widetilde{Pb} + \overline{Pb} = -\overline{Po}\cos(2wt) + \frac{Vm^2 * \sin(2wt)}{4XC} + [\overline{Po} - Pi1 - Pi2]$$

Where $\overline{Pb}$ represents the battery average power, which is equal to the system power difference between the total generated dc power and the average load power. Besides, the power component ˜ PB is the summation of the pulsation component of the load power and the total reactive power associated with the converter capacitors. Now, neglecting the saved energy in the converter inductors, the battery current and it's both side inductors currents are obtained.

$$IL1 = \frac{pb1}{Vb} = \frac{Vdc * Im\sin(wt)}{Vb} + \frac{Vdc * Vm\cos(wt)}{2VbXC} - \frac{\overline{Po} * \cos(2wt)}{2Vb} + \frac{(Vm^2 * \sin(2wt))}{8VbXC} + [\frac{\overline{Po} - 2Pi1}{2Vb}]$$

$$IL2 = \frac{Pb2}{Vb} = -\frac{Vdc * Im\sin(wt)}{Vb} - \frac{Vdc * Vm\cos(wt)}{2VbXC} - \frac{\overline{Po} * \cos(2wt)}{2Vb} +$$

$$\frac{(Vm^2 * \sin(2wt))}{8VbXC} + [\frac{\overline{Po} - 2Pi2}{2Vb}]$$

$$Ib = \frac{Pb}{Vb} = -\frac{\overline{Po} * \cos(2wt)}{Vb} + (Vm^2 * \sin(2wt))\frac{1}{(4Vb * XC)} + [\frac{(\overline{Po} - Pi1 - Pi2)}{Vb}]$$

Two currents *IL*1 and *IL*2 contain two similar alternative terms with the angular frequencies of ω and 2ω and two different dc terms that depend on the total generated power by their corresponding input dc sources. Moreover, the battery current in (16) is a dc-biased sinusoidal waveform at the angular frequency of 2ω, which guarantees supplying the expected dc and ac power components of the battery.

III.PROPOSED CONVERTER DESIGN

The five-input structure of the proposed converter is shown in Fig. 5. As is seen in the figure, eight passive elements *C*1, *C*2 , *L*1, *L*2 , . . . , *L*5 , and *L*6 exist in the converter structure ($n = 6$). All these passive elements values can be determined in such a way that the desired current and voltage ripples are satisfied for the converter. It can be easily shown that in order to have the maximum current and voltage ripples of Δ*I*max and Δ*V*max for the system inductors and capacitors, respectively, they should be chosen bigger than the following minimum values:

$$Limin = \frac{Dmax * Vi}{\Delta Imax * fs}, \qquad Cmin = \frac{Dmax * Vm}{RL * \Delta Vmax * fs}$$

where *D*max is the converter predefined maximum duty ratio (comes from the converter stability consideration), *fs* is the converter switching frequency, *Vi* is the input dc voltage connected to the *i*th inductor, *RL* is the load resistance, and *Vm* is the peak value of the load sinusoidal voltage. Hence, for the input and output voltage ranges in the simulation section, Dmax = 0.5, fs = 20 kHz, ΔImax = 1 A for L3 to L6 , ΔImax = 7 A for L1 and L2 and ΔVmax = 10 V for C1 and C2 the following minimum values are obtained for the converter passive elements:

L1 min = L2 min = 0.339mH, L3 min...L6 min = 2.45mH

Cmin =35.35μF.

As is mentioned in the dc–ac operation mode, the proposed converter power switches experience the voltage stress of Vstr = Vdc+Vm/2. This voltage stress can be possibly minimized if the voltage Vdc is minimized. Therefore, assuming that all input dc voltages and the converter duty ratios take values in the
following ranges:

$$VL \le Vi \le VH$$

$$DL \le di \le DH,\ DL, DH \in [0, 1)$$

It can be shown that, if the lowest and highest duty ratios are set at zero and the predefined value Dmax, respectively, the voltages Vdc and Vstr are minimized, while a minimum acceptable value VL min is dictated to input dc voltages as

$$Vdc(min) = VH + Vm/2$$

$$Vstr\ (min) = VH + Vm$$

$$VL\ (min) = (1 - Dmax)(VH + Vm).$$

For example, if the minimum and maximum values of the converter duty ratios are designed to be *DL* = 0, *D*max = 0.5 and with *VH* = 100 V, and *Vm* = 280 V, using (16) gives the following voltage design considerations for the system:

*V*dc(min) = 240V; *V*str (min) = 380V; *VL* (min) = 76V.

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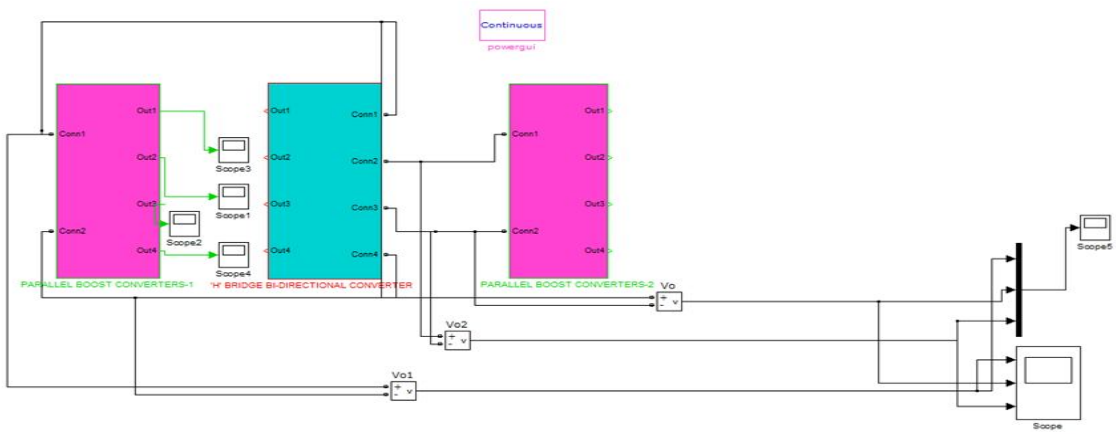
IV. SIMULATION CIRCUIT AND RESULTS

In this section, the introduced converter in Fig. 2 is simulated by MATLAB/PSCAD software to show its operation performances in both dc–dc and dc–ac modes. The simulation parameters for the converter circuit are listed in Table 3. In addition, a changeable resistive load is connected to the output port. In simulations, using the obtained input/output voltages in (24) and considering the voltage drops across passive and active elements, the voltages of input dc sources, battery storage, and the maximum and dc-bias values of the converter output voltage are set at following values:

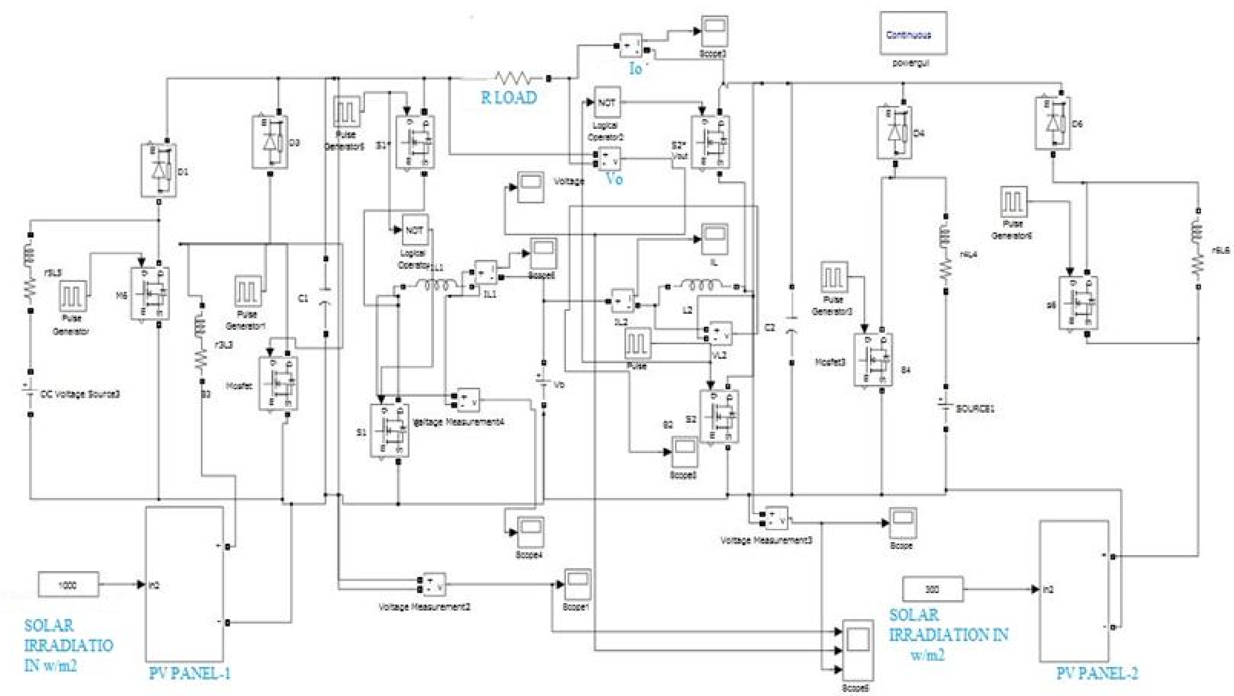
Vi3 = 80V, Vi4 = 95V, Vi5 = 90V, Vi6 = 85V

VB = 96V, Vm = 280V Vdc=230V, S3=45%, S5=60%

S4=45%, S6=40%, S1=25%, S1*=75%, S2=25%,S2*=75% and switching frequency (Fs)=20khz



A. *Simulation of proposed system*



In this mode, the load voltage is desired to be VO = 230 V. And for this matter, the converter output voltages VO1 and VO2 are regulated at the reference dc voltages of VO1ref = 366 V and VO2ref = 136 V, respectively. The reference currents of the input dc sources are considered to be variable, which will result in providing three different periods of simulation.

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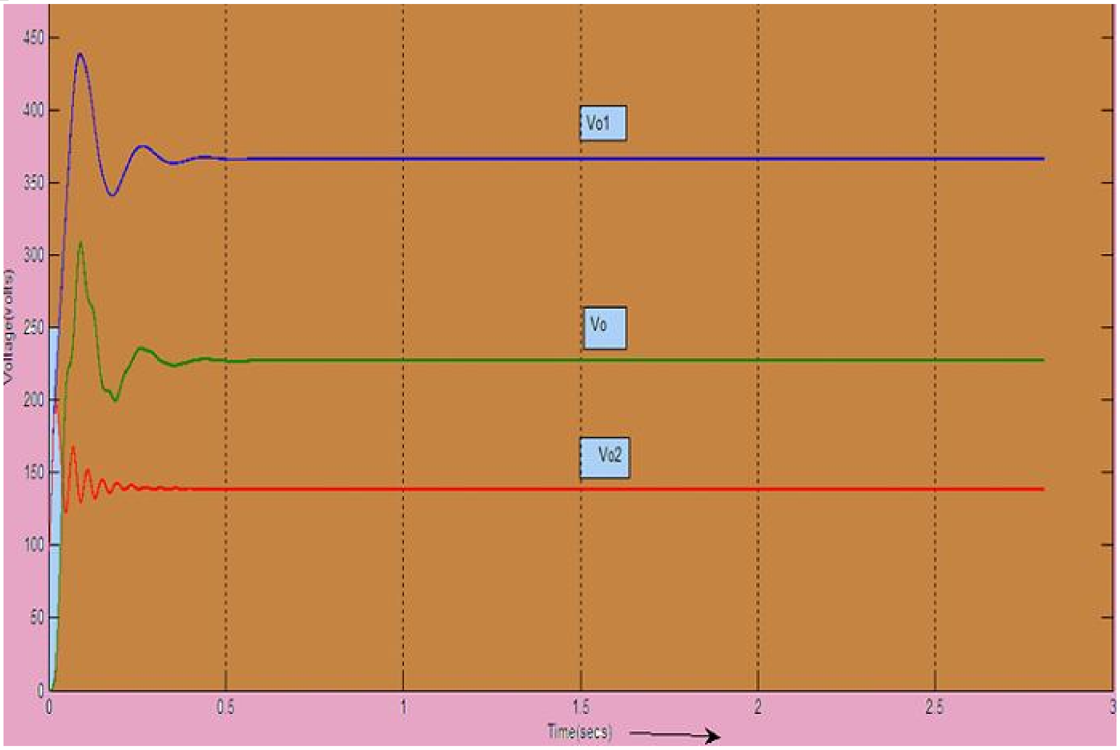
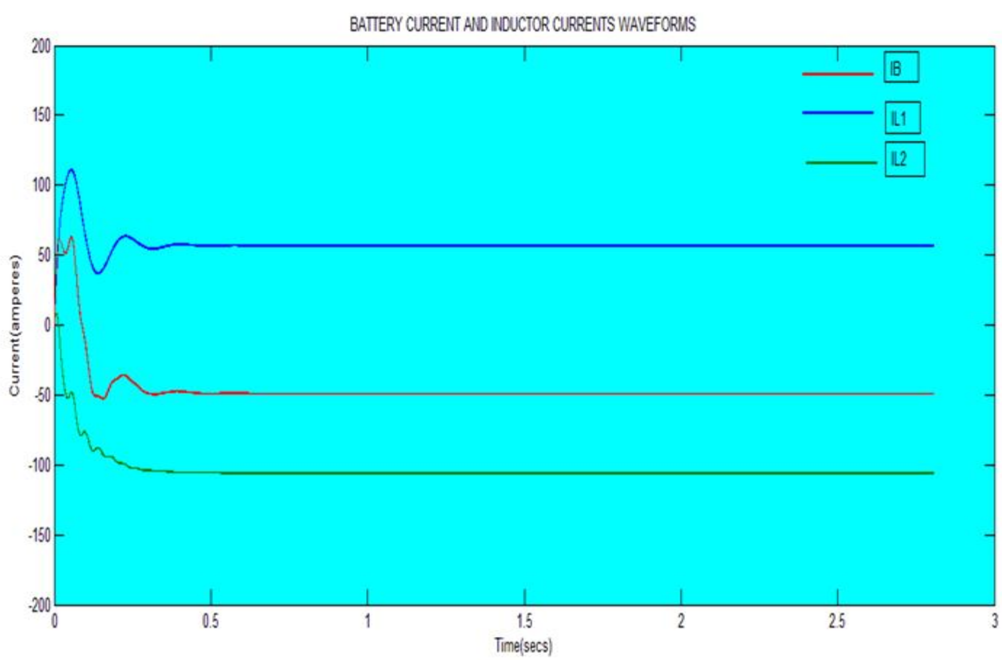


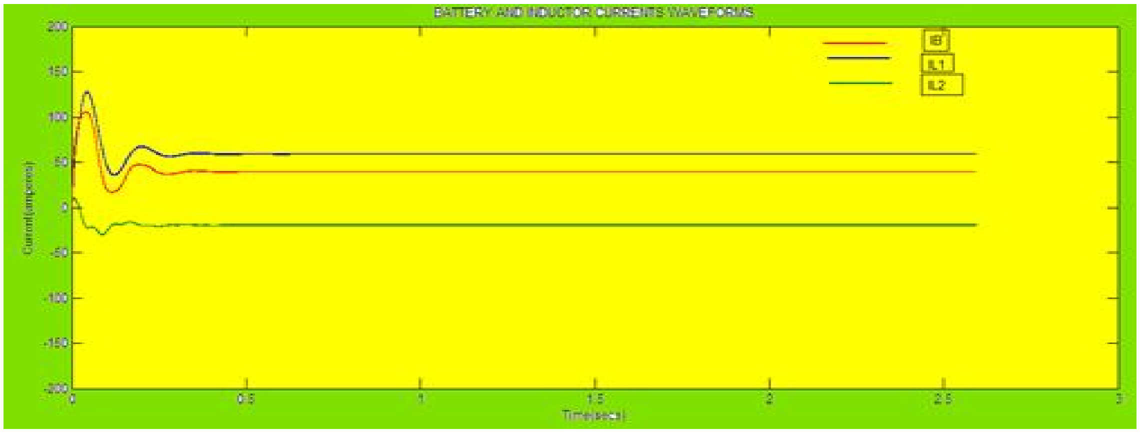
Output Voltage waveforms in dc-dc mode



Battery and inductor currents in dc-dc mode.

*IL*1 and *IL*2 take positive and negative values, respectively, and battery current (Ib) are negative this means battery is completely charging through discharged capacitor C2 , switch S2*.and inductor L2 and battery and again return to the capacitor C2 now the entire power deliver to the load by discharged capacitor C2 and in this case capacitor C1 charging takes place. And these values are given below.

IL1 = +58A, IL2 = -106A and Ib = -48A respectively.

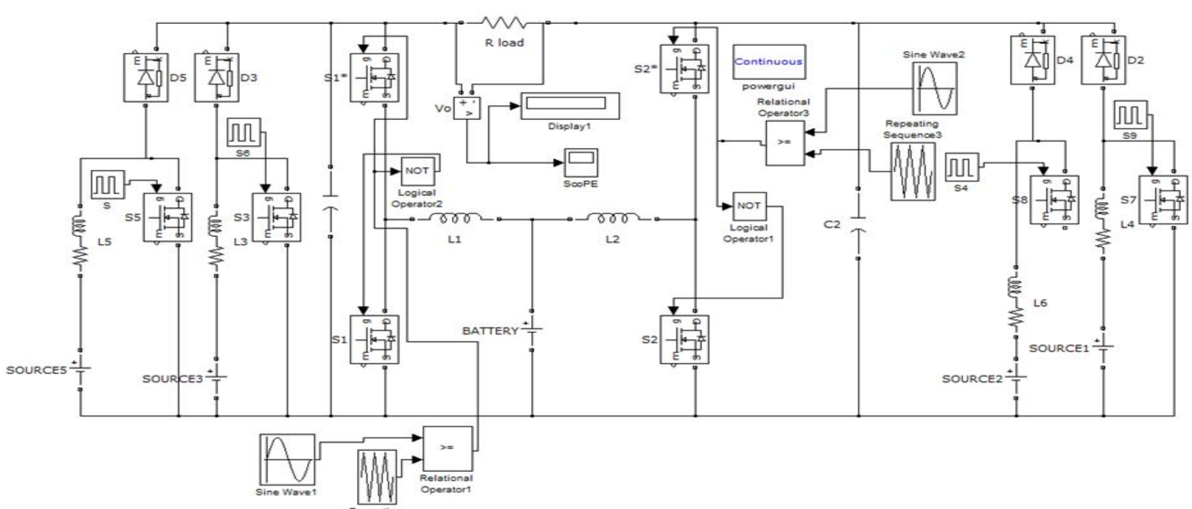


Battery and inductor currents in the dc-dc mode

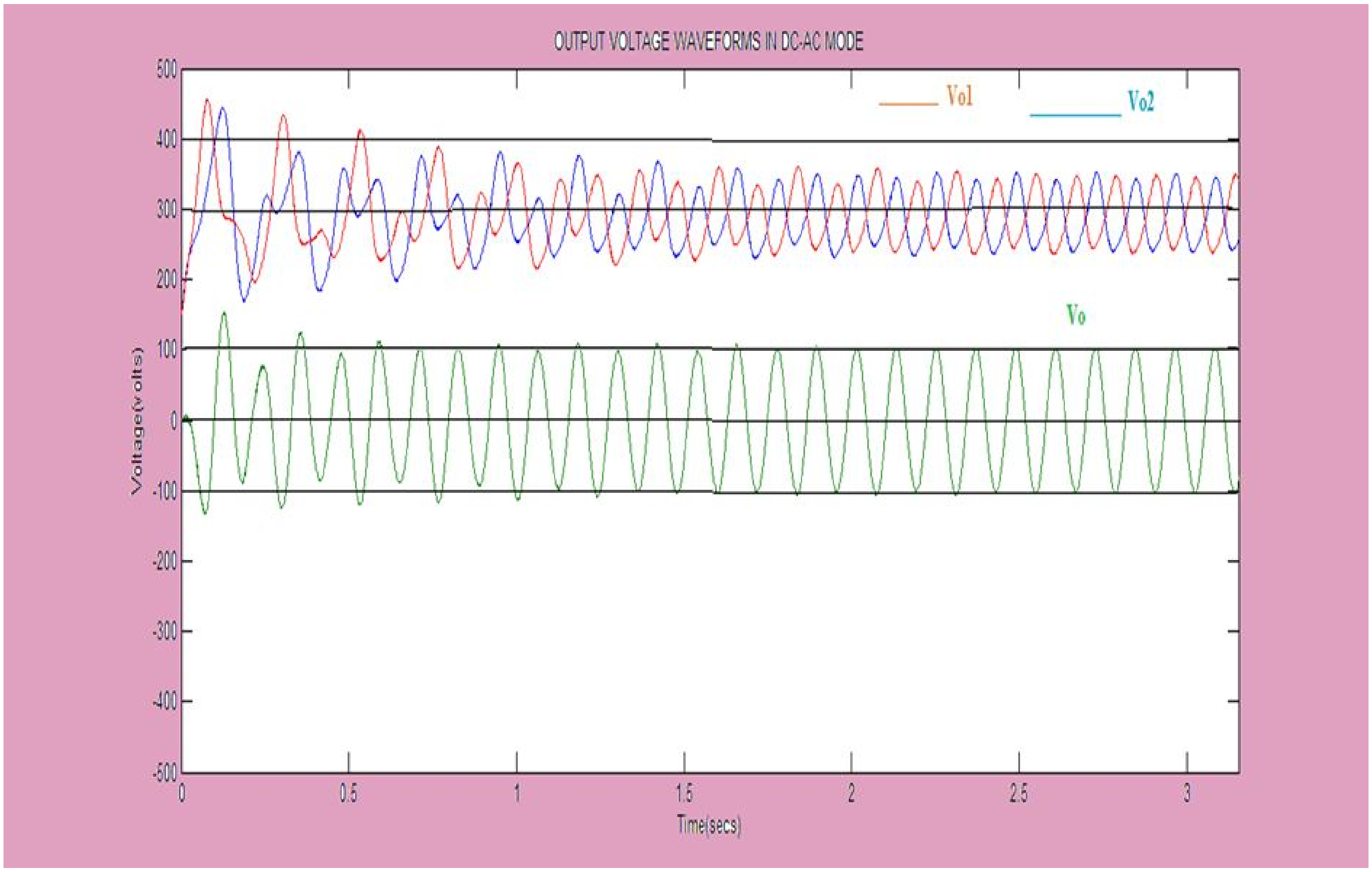
the inductor currents (IL1) and (IL2) positive and negative values and battery current (Ib) is positive value so this means now the battery is completely discharged through inductor L1, switch S1 and return back to battery so inductor L1 will be charging through switch S1 and battery current (Ib).In this case also the power delivered to the load by the capacitor C2 and capacitor C1 still charging. The values of inductor currents and battery current are given below.
IL1 = +60A, IL2 = -20A and Ib = +40A respectively.

Table.1.1 Simulation Results in dc-dc mode

S1 (%)	S2 (%)	S3 (%)	S4 (%)	S5 (%)	S6 (%)	S1* (%)	S2* (%)	VO1	VO2	VO
35	35	36	50	38	52	65	65	266V	166V	100V
35	35	36	45	38	47	65	65	266.3V	158V	108.3V
25	25	34	44.4	36	46.3	75	75	346.2V	146.2V	200V
25	25	36	35	38	33	75	75	350	140	210V
25	25	45	45	60	40	75	75	366	136	230V



Simulink model in dc-ac mode

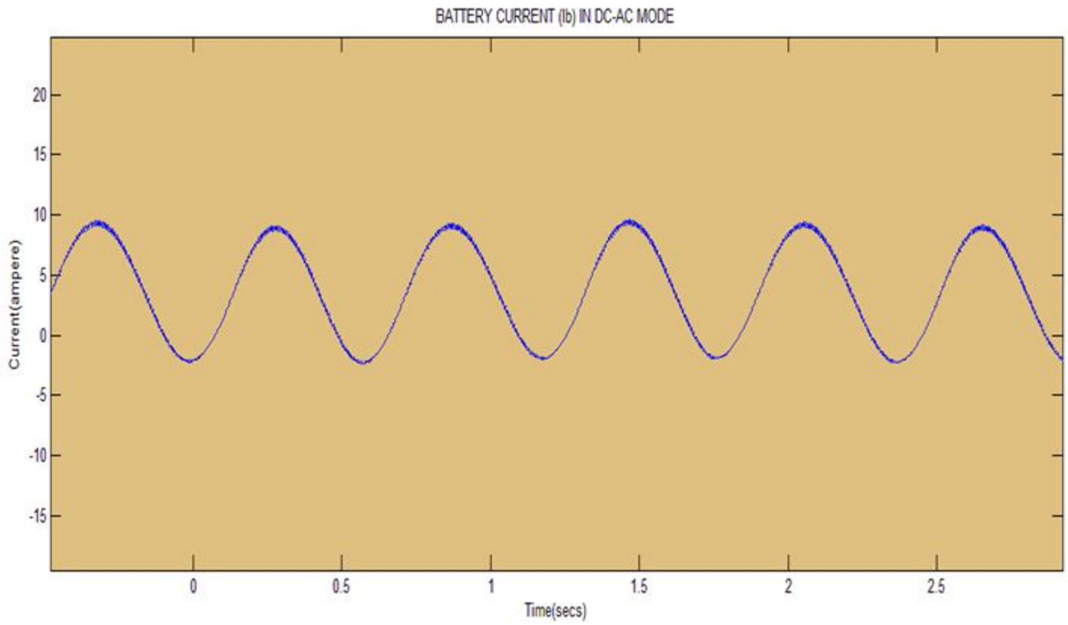


Sinusoidal Output Voltage waveform in dc-ac mode

Table.2 Smulation results in dc-ac mode

L1(mH)	L2(mH)	Vb(Volts)	Vo1(volts)	Vo2(volts)	Vo(volts)	T.H.D(%)
3.5	3.5	110	141	68.35	195	3.80
4.5	4.5	90	141	68.35	200	2.974
8.5	8.5	96	154.6	119.3	100	2.411
5	5	110	138.9	68.69	150	1.98
5	5	150	240	348	100	0.696

As is clear the load voltage is perfectly generated sinusoidal (THD =0.696%) and does not experience any voltage quality problems, i.e., voltage sags/swells, coming from the fact that the battery is immediately charged/discharged to support input sources in online producing the desired output voltages. With the intention of achieving similar results, the load power, the input dc sources *Vi*3 to *Vi*6 and their reference currents values are chosen to be as the same as those in the dc–dc mode. Therefore, three similar periods are provided for the system.



Battery current wave-form in dc-ac mode.

the sinusoidally varied battery current waveform this can be generated by the battery and 180 degrees phase-shift provided by unipolar sinusoidal pulse-width modulation switching scheme and this provided in the converter central part by the interface of current bi-directional H-bridge converter and battery storage system this can be automatically controlled by generation of 180 degree out of phase unipolar sinusoidal pulse width modulation switching pulses to the H-bridge converter in the central part.

V. CONCLUSION

The single stage multi-input boost converter which can work in both dc-dc and dc-ac modes has been proposed in this project it is extendable to accept more number of input dc sources and the proposed converter does-not need any output filter arrangement. The proposed converter utilizes minimum number of power switches, current ripple<5% in dc-dc mode and Total harmonic distortion is T.H.D<3% (0.696%) drastically reduced in dc-ac mode and so pure sinusoidal voltage wave form appeared across converter output with-out using any external filter arrangement. This can be possible by the battery generated dc biased sinusoidal voltages and autonomously charging/discharging of battery placed at the converter central part and also balance the power flow between source and load. Simulation results shows all the converter capabilities such as low-current ripples for input dc sources, autonomous battery charging/discharging, and producing high-quality dc or ac output voltages.

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