CONTROL OF TRANSFORMERLESS UNINTERRUPTED POWER SUPPLY FOR UNITY POWER FACTOR AND SINUSOIDAL INDUCTOR CURRENTS

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ABSTRACT

This paper discusses the design of controller for transformerless uninterruptible power supply (UPS), with simultaneous ac and dc loading. The objective of the design is to maintain good regulation of the ac and dc voltages and sinusoidal inductor currents simultaneously. The volt-second balance across the output filter inductor is used to maintain the current in sinusoidal form which is the unique contribution of this paper. Complete analytical design solution is provided for both rectifier and inverter simultaneously. Implementation of the controller requires simple analog circuits. The proposed solution is validated through simulation and experimentation.

Keywords*: Ac And Dc Load, Carrier Based Modulation, Half Bridge Boost Rectifier, Transformerless UPS.*

1. INTRODUCTION

Efficiency is one of the important concerns in power electronics equipments. Number of switches and the switching frequency are the major contributing factors to power loss. Boost type converter type topology is best suited for providing better voltage regulation and reduced harmonic distortion. Several works have dealt with modeling, analysis and control of half bridge Boost rectifier[1-5,16-18]. Various configurations of boost converter topologies are presented in [1]. Among these, single phase half bridge voltage doubler boost converter has been a best choice for low cost, massive high power and variable power factor applications [2-4]. The configuration of the half bridge converter topology is expected to have higher efficiency over any converter topologies for UPS applications. It provides voltage doubling and allows transformerless power transfer. The circuit is extensively used in single phase ac line conditioner and frequency changer applications. Many works have been reported in literature in the direction of control of UPS [11-15]

The circuit diagram of the half bridge boost converter also known as transformerless UPS is shown in Figure 1.
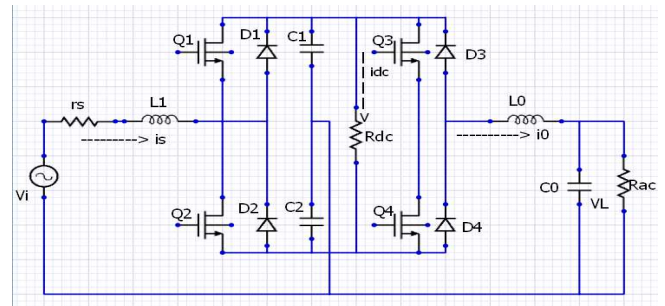

Figure.1.Circuit Diagram Of The Transformerless UPS

There are five challenges associated with the transformerless UPS control namely i) to regulate the dc bus voltage ii) to regulate the output ac voltage iii) To maintain input power factor at unity iv) to hold DC bus voltage equally in the two capacitors and v)to hold input inductor and output ac filter currents sinusoidal .

Detailed state space analysis and analytical results related to half bridge boost rectifier efficiency and voltage imbalances are reported in [3].The influence of circuit parameters variation and switching strategy on the boost rectifier's

performance and controller design are reported in [2].

A nonlinear carrier based control is proposed in [5] for high power factor boost rectifiers. The technique eliminates input voltage sensing, error amplifier in the current shaping loop and multiplier-divider circuitry. Design of a digital programmable control IC for common neutral half bridge ac-dc-ac converter is presented in [6]. The paper also discusses the inner current loop and the outer voltage loop for ac regulation. A simple analog circuit with complete analytical design solutions is presented for the rectifier in [4]. The paper also illustrates the function of carrier based modulation as an equivalent current loop. An inverse transfer function based zero phase shift notch filter control is proposed for SPWM inverter in [7] where the dc dynamics is kept constant. A four- leg type converter which operates as a rectifier, a battery charger/discharger and an inverter with necessary analysis and design of control schemes is presented in [8].

The analytical equations of the UPS show that ac and dc dynamics are decoupled for voltage regulation. Hence the ac and dc voltage regulations can be dealt separately. However the ac and dc dynamics are coupled for the current dynamics. Hence the design for sinusoidal input and output filter inductor currents should be dealt in the coupled form. This feature is the unique contribution of the proposed paper when compared to the existing literature. In this paper the complete control scheme is presented for regulating the ac and dc bus voltages. The method used for maintaining the unity power factor at the input is extended to maintain the output inductor currents in sinusoidal form. The control functions for both the ac and dc voltage regulations are derived from volt-second balance equations across the input and output filter inductors respectively. This results in simple analog control functions for both the rectifier and inverter which can be easily implemented.

2. CIRCUIT CONFIGURATION AND MATHEMATICAL MODELLING

The circuit diagram of the transformer less UPS is shown in Figure 1.Vi is the input voltage, L1 is the input inductor, rs is the equivalent series resistance of the inductor, C1 and C2 are the dc link capacitors, Rdc is the dc load resistor, L0 and C0 are the inductance and capacitance of the output filter and Rac is the ac load resistor. Switches Q1 and Q2 correspond to the rectifier portion and Q3

and Q4 correspond to the inverter portion of the UPS.

The equivalent circuit diagrams for various switch positions are shown in Figure 2. Operation of the circuit can be explained in four modes as given by Table 1 depending on the ON and OFF conditions of Q_1 & Q_2 and Q_3 & Q4 as indicated by S_1 and S_2.
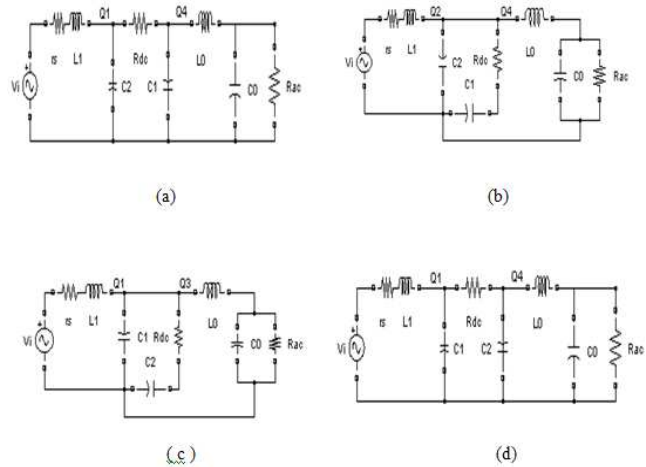

Figure 2. Equivalent circuit diagrams for ON and OFF conditions of the switches (a) Q_2 and Q_3 ON (b) Q_2 and Q_4 ON (c) Q_1 and Q_3 ON (d) Q_1 and Q_4 ON

Table 1: ON-OFF condition of the switches

Q_1	Q_2	S_1	Q_3	Q_4	S_2
ON	OFF	1	ON	OFF	1
OFF	ON	0	OFF	ON	0

The dynamic equations in terms of S_1 and S_2 can be given as follows.

$$L_1 \frac{di_s}{dt} = v_i - r_s i_s + S_1 v_{c2} - (1 - S_1) v_{c1} \quad (1a)$$

$$C_1 \frac{dv_{C1}}{dt} = -i_{dc} - i_0 S_2 + i_s (1 - S_1) \quad (1b)$$

$$C_2 \frac{dv_{C2}}{dt} = -i_{dc} - i_0 (1 - S_2) - i_s S_1 \quad (1c)$$

$$L_0 \frac{di_0}{dt} = S_2 v_{C1} - (1 - S_2) v_{C2} - r_d i_0 - v_l \quad (1d)$$

$$C_0 \frac{dv_l}{dt} = i_0 - i_l \quad (1e)$$

where S_1 and S_2 to be substituted from Table 1.

Upon replacing S_1 and S_2 by d_1 and d_2 where d_1 is the duty ratio of Q_2 and d_2 is the duty ratio of Q_3, the average model is obtained as in equations (2a-2e)

$$L_1 \frac{di_s}{dt} = v_i - r_s i_s + d_1 v_{C2} - (1-d_1)v_{C1} \qquad (2a)$$

$$C_1 \frac{dv_{C1}}{dt} = -i_{dc} - i_0 d_2 + i_s(1-d_1) \qquad (2b)$$

$$C_1 \frac{dv_{c2}}{dt} = -i_{dc} - i_0(1-d_2) - i_s d_1 \qquad (2c)$$

$$L_0 \frac{di_0}{dt} = d_2 v_{C1} - (1-d_2)v_{C2} - r_d i_0 - v_l \quad (2d)$$

$$C_0 \frac{dv_l}{dt} = i_0 - i_l \qquad (2e)$$

2.1 Small Signal Model

For small perturbations in the variables around their nominal values, a linearised small signal model is formed so as to facilitate the controller design.

Let , $d_1 = D_1 + \hat{d}_1$, $d_2 = D_2 + \hat{d}_2$,

$v_{C1} = V_{C1} + \hat{v}_{C1}$, $v_{c2} = V_{c2} + \hat{v}_{c2}$,

$i_s = I_s + \hat{i}_s$, $i_0 = I_0 + \hat{i}_0$, $v_l = V_l + v_l$

Where $D_1, D_2, V_{c1}, V_{c2}, I_s, V_0$ are the nominal values of the variables, $\hat{d}_1, \hat{d}_2, V_{c1}, V_{c2}, \hat{i}_s, \hat{i}_0,$ $\hat{v}_l$ are the small perturbations in the corresponding variables and $d_1, d_2, v_{c1}, v_{c2}, i_s, i_0, v_0$ are the respective instantaneous values.

$$L_1 \frac{d\hat{i}_s}{dt} = \hat{v}_i - r_s \hat{i}_s + \hat{d}_1 V_{C2} + D_1 \hat{v}_{C2} - (1-D_1)\hat{v}_{C1} + \hat{d}_1 V_{C1} \quad (3a)$$

$$C_1 \frac{d\hat{v}_{C1}}{dt} = -\hat{i}_{dc} - i_0 D_2 - I_0 \hat{d}_2 + (1-D_1)\hat{i}_s - I_s \hat{d}_1 \quad (3b)$$

$$C_2 \frac{d\hat{v}_{C2}}{dt} = -\hat{i}_{dc} - D_1 \hat{i}_s - I_s \hat{d}_1 - \hat{i}_0(1-D_2) + I_0 \hat{d}_2 \quad (3c)$$

$$L_0 \frac{d\hat{i}_0}{dt} = D_2 \hat{v}_{C1} + V_{C1} \hat{d}_2 - (1-D_2)\hat{v}_{C2} + V_{C2}\hat{d}_2 - r_s \hat{i}_0 - \hat{v}_l \quad (3d)$$

$$C_0 \frac{d\hat{v}_l}{dt} = \hat{i}_0 - \frac{\hat{v}_l}{R_{ac}} \qquad (3e)$$

Equations (3a) to (3e) describe the small signal model and equation (4) gives the Matrix form

$$\begin{bmatrix} \frac{d\hat{i}_s}{dt} \\ \frac{d\hat{v}_{c1}}{dt} \\ \frac{d\hat{v}_{c2}}{dt} \\ \frac{d\hat{i}_0}{dt} \\ \frac{d\hat{v}_l}{dt} \end{bmatrix} = \begin{bmatrix} -\frac{r_s}{L_1} & -\frac{1-D_1}{L_1} & \frac{D_1}{L_1} & 0 & 0 \\ \frac{1-D_1}{C_1} & -\frac{1}{R_1} & -\frac{1}{R_1} & \frac{D_2}{C_1} & 0 \\ -\frac{D_1}{C_2} & -\frac{1}{R_1} & -\frac{1}{R_1} & -\frac{1-D_2}{C_2} & 0 \\ 0 & \frac{D_2}{L_0} & \frac{1-D_2}{L_0} & -\frac{r_d}{L_0} & -\frac{1}{L_0} \\ 0 & 0 & 0 & \frac{1}{C_0} & -\frac{1}{R_{ac}C_0} \end{bmatrix} \begin{bmatrix} \hat{i}_s \\ \hat{v}_{c1} \\ \hat{v}_{c2} \\ \hat{i}_0 \\ \hat{v}_l \end{bmatrix} + \begin{bmatrix} \frac{1}{L_1} & 0 & \frac{V_{C1}+V_{C2}}{L_1} & 0 \\ 0 & -\frac{1}{C_1} & -\frac{I_s}{C_1} & -\frac{I_0}{C_1} \\ 0 & -\frac{1}{C_2} & -\frac{I_s}{C_2} & \frac{I_0}{C_2} \\ 0 & 0 & 0 & \frac{V_{C1}+V_{C2}}{L_0} \\ 0 & 0 & 0 & 0 \end{bmatrix} \begin{bmatrix} \hat{v}_l \\ \hat{i}_{dc} \\ \hat{d}_1 \\ \hat{d}_2 \end{bmatrix}$$

3. PROPOSED CONTROL SCHEME FOR SINUSOIDAL INDUCTOR CURRENTS

Consider the transformerless UPS circuit shown in Figure 1. Assuming similar switching frequency for both converter and inverter circuits, let T be the switching period. Assuming a constant and high switching frequency f (1/T) the duty cycle d_1 and d_2 can be determined by the volt-second balance across the inductors L_1 and L_0 respectively as given by

$$v_i - i_s r_s = v_{c1} - d_1(v_{c1} + v_{c2}) \qquad (5a)$$

$$v_l + i_0 r_d = d_2(v_{c1} + v_{c2}) - v_{c2} \qquad (5b)$$

As per control objective, the inductor current i_s should be made proportional to the input voltage v_i for unity power factor and the output inductor current should be sinusoidal with its frequency equal to the output voltage but need not be in phase with it.

Then $i_s = K_1 (V_{mi} \sin \omega_i t)$ (6a)

and $i_0 = K_2 V_{m0} \sin(\omega_0 t + \phi_0)$ (6b)

Where K_1, K_2 and ϕ_0 are constants.

Let us consider three integral based control loops one each for regulating v_{C1}, v_{C2} and v_l respectively as shown in figure 3. Then v_{c1}, v_{C2} and v_l of equations (5a) and (5b) can be replaced by corresponding controller outputs v_{m1}, v_{m2} and v_{m3} respectively for each switching cycle in steady state.

Equations (5a) and (5b) can be rewritten as given by 7a and 7b with v_i replaced in terms of i_s

$$i_s\left(\frac{1}{K_1} - r_s\right) = v_{m1} - d_1(v_{m1} + v_{m2}) \qquad (7a)$$

$$-r_d i_0 = v_{m3} - d_2(v_{m1} + v_{m2}) + v_{m2} \qquad (7b)$$

The variables d_1 and d_2 can be solved from the equations (7a) and (7b)

$$d_1 = \frac{v_{m1} - \left(\dfrac{V_i \sin \omega_i t}{K_1}\right)\left(\dfrac{1}{K_1} - r_s\right)}{v_{m1} + v_{m2}} \quad (8a)$$

$$d_2 = \frac{r_d K_2 v_0 \sin(\omega_0 t + \varphi_0) + v_{m3} + v_{m2}}{v_{m1} + v_{m2}} \quad (8b)$$

Equation 8b shows that the dc loop control commands Vm1 and Vm2 influence the duty cycle of the inverter switches for maintaining sinusoidal inductor currents.

Solving of (8a) & (8b) requires an estimate of K1, K2 and ϕ$_0$. An alternate solution is to generate the switching pulse using a carrier based approach [4, 9]. Let V_{c01} and V_{c02} represent the right hand side of equations 7a & 7b with d_1 and d_2 replaced by t/T. Then

$$V_{CO1}(t) = V_{m1} - (V_{m1} + V_{m2})\, t/T \; ; \; 0 < t < T \quad (9a)$$

$$V_{CO2}(t) = +V_{m3} + V_{m2} - (V_{m1} + V_{m2}) t/T; \; 0 < t < T \quad (9b)$$

It can be easily deduced from 9a and 9b that, $V_{CO1}(t)$ is equal to V_{m1} in the beginning of any interval T (t=0) and will be equal to $-V_{m2}$ at the end of the interval. $V_{CO1}(t)$ equals $i_s\left(\dfrac{1}{K_1} - r_s\right)$ at the instant t = D_1T as in equation (7a). Thus switching instant of inverter can be determined by comparing the input current $i_s\left(\dfrac{1}{K_1} - r_s\right)$ with $V_{CO1}(t)$ using a comparator as shown in Figure 3. Similarly, $V_{CO2}(t)$ is equal to $-V_{m2} - V_{m3}$ in the beginning of any interval T and will be equal to $V_{m1} - V_{m3}$ at the end of the interval. $V_{CO2}(t)$ equals $r_d i_0$ at end of D_2T as shown in (7b). Thus the switching instant of inverter can be determined by comparing the ac filter inductor current $-r_d i_0$ with $V_{CO2}(t)$ with the help of a comparator as shown in Figure 3.

This completes the design of control scheme for sinusoidal inductor currents. The voltage regulation loops are discussed in the following sections.
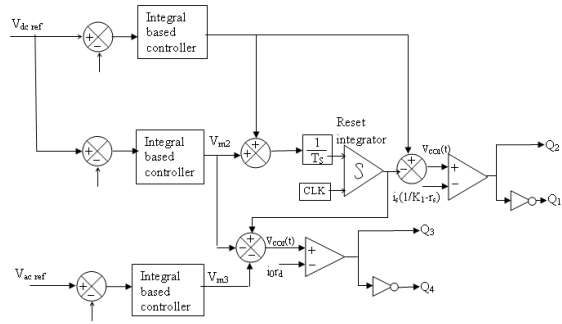

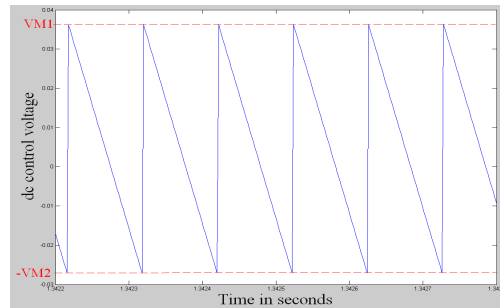*Figure3. Block diagram of the control scheme*
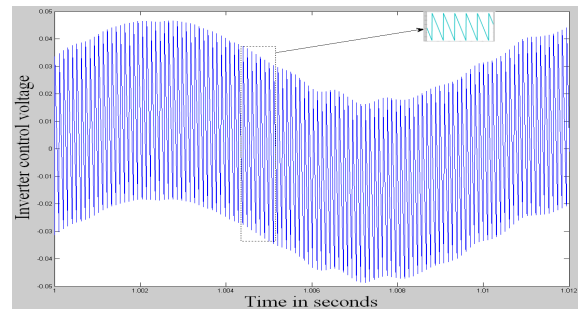

Figure. 4. Control voltage Vco₁(t)

Figure. 5. Control voltage Vco₂(t)

4. CONTROLLER DESIGN OF AC/DC VOLTAGE REGULATION

Considering the symmetry of the circuit the following simplifications can be made $V_{c1} = V_{c2} = V_c$, $\hat{v}_{C1} = -\hat{v}_{C2} = \hat{v}_C$, $D_1 = 0.5$, $D_2 = 0.5$. The simplified transfer functions are given as follows,

$$\frac{\hat{v}_c(s)}{\hat{d}_1(s)} = -\frac{I_s}{C_1}\left(\frac{s + \left(\dfrac{r_s}{L_1} - \dfrac{V_C}{I_S L_1}\right)}{s^2 + \dfrac{r_s}{L_1} + \dfrac{0.5}{C_1 L_1}}\right) \quad (10a)$$

$$\frac{\hat{v}_c(s)}{\hat{d}_2(s)} = -\frac{I_0 L_1\left(s+\dfrac{r_s}{L_1}\right)}{s^2 + \dfrac{r_s}{L_1}s + \dfrac{0.5}{C_1 L_1}} \tag{10b}$$

$$\frac{\hat{v}_0(s)}{\hat{d}_2(s)} = \frac{2V_c}{\left(s^2 + \left(\dfrac{R_{ac}r_d C_0 + L_0}{L_0 C_0 R_L}\right)s + \left(\dfrac{R_{ac}+r_d}{L_0 C_0 R_L}\right)\right)L_0 C_0} \tag{11}$$

In the steady state, the effects of $\hat{d}_1 \,\&\, \hat{d}_2$ on $\hat{v}_c$ and $\hat{v}_0$ are tabulated in table 2.

Table 2: Variation of $\hat{v}_0(s) \,\&\, \hat{v}_c(s)$ with $\hat{d}_1 \,\&\, \hat{d}_2$

	$\hat{v}_c$	$\hat{v}_0$
$\hat{d}_1$	$\dfrac{I_s r_s}{0.5} - \dfrac{V_C}{I_s}$	0
$\hat{d}_2$	$\dfrac{I_0 r_s}{0.5} L_1 C_1$	$\dfrac{2V_C R_{ac}}{R_{ac} + r_d}$

It can be observed from table 2 that the effect of $\hat{d}_2$ on $\hat{v}_c$ is insignificant when compared to the effect of $\hat{d}_1$ on $\hat{v}_c$ due to the product of r_s, L_1 and C_1. This results in two decoupled transfer functions as given by equations (10a) and (11).

Simple Proportional-integral control schemes can be employed for both the converter as well as the inverter. The transfer function of the controller is given by the equation (12). The proportional gain and the integral gain of the controller are selected through root locus technique.

$$C(s) = \frac{K_P s + K_I}{s} \tag{12}$$

5. SIMULATION AND EXPERIMENTAL RESULTS

The UPS is designed with parameters shown in table 3. The operational amplifier implementation of control circuit is shown in figure 6. The complete hardware setup is shown in figure

Figure 8 shows the input voltage and current waveforms in steady state. The waveforms

are perfectly sinusoidal with unity power factor. It is observed that the same (power factor being unity and the sinusoidal nature) is maintained over a wide range of operating conditions.

Table3:Parameters and Components of the Converter

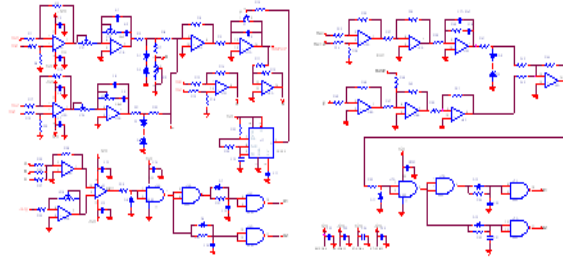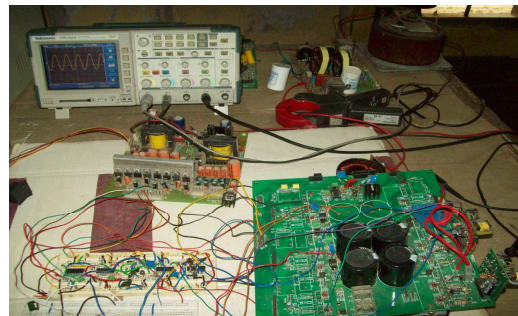| f_{SW}	10KHz
f_i	50Hz
L1	1mH
rs	0.2 Ω
C1	2200 µF
C2	2200 µF
Rdc	95 Ω
L_0	1mH
C_0	210 µF
R_{ac}	80 Ω

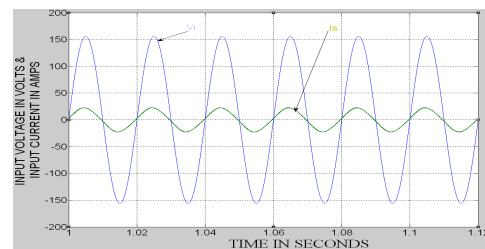*Figure.6.Circuit diagram of the controller*

Figure .7 Hardware setup

Figure 8.Input voltage (V_i) & input current (i_s)

The rectifier output transient and steady state voltage across the two dc link capacitors is

shown in figure 9a.The voltages settle to any desired value   over a wide range of input voltage, reference voltage and load current. It is possible to further reduce the magnitude of ripple in steady state
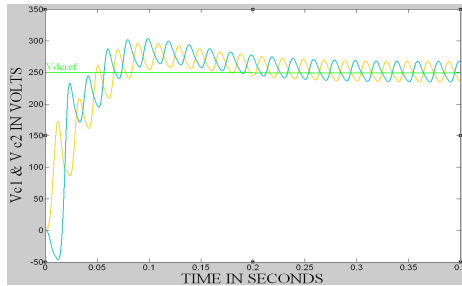

Figure.9a. V_{C1} & V_{C2}

Rectifier response for a step change of 200 volt (from 200 volt to 400 volt) in the dc reference is shown in figure 10b. It can be observed from Figure 9b that Vc1 and Vc2 have reached the set voltage with a small overshoot and undershoot.
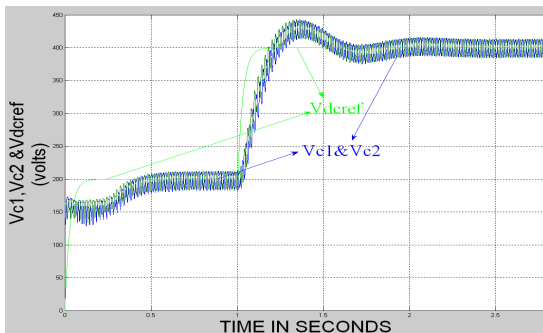

Figure.9b. V_{C1} & V_{C2} for a step change in Vdc reference

The output voltage and the output filter current in steady state are shown in figure 10. The Voltage and the current are sinusoidal. However the harmonic distortion in the output inductor current is higher than the input .
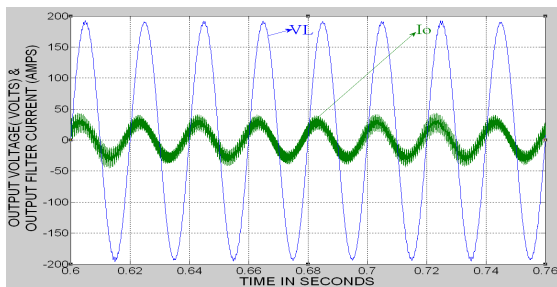

Figure.10.Output voltage (V_l) & output filter Current(i_0)

Experimental results are shown in figures 11a and 11b. Experimental results match with the results obtained through simulation. Figures 11a-11c show the input voltage and input current waveforms for varying converter parameters, input voltage and reference. Input current is maintained sinusoidal and the power factor is near unity. Rectifier output is shown in figures 12a and 12b. It can be observed that rectifier has very low ripple. The rectifier output reaches any set value.
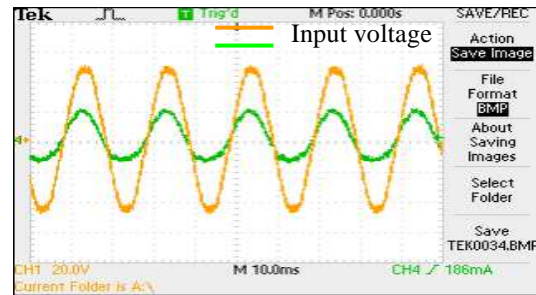
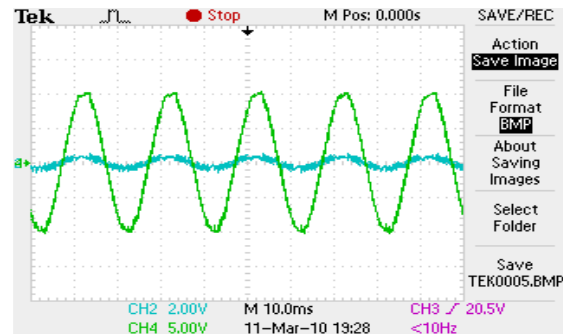

Figure.11a.Input voltage (V_i) & input current (i_s)

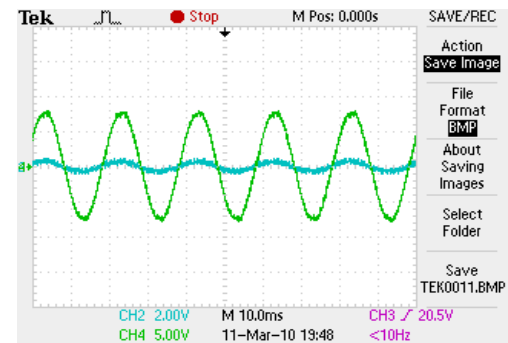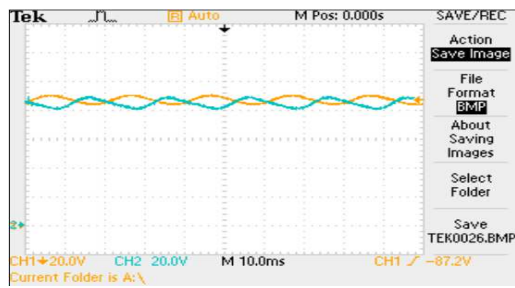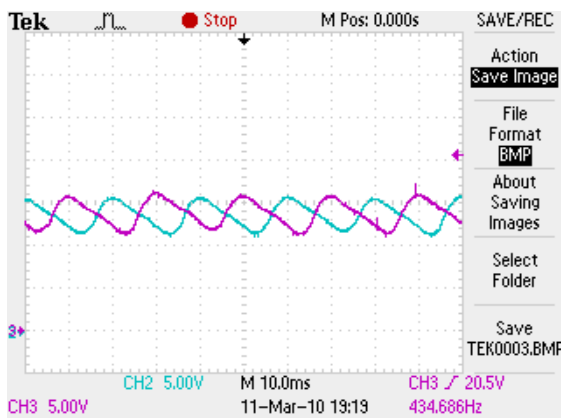*Figure 11 b. Input voltage (V_i) & input current (i_s)*

Figure 11 c. Input voltage (V_i) & input current (i_s)

Fig.12a .Rectifier output (voltage across V_{C1} and V_{C2})

Figuer.12b. Rectifier output (voltage across V_{C1} and V_{C2})

6. CONCLUSION

The dynamic equations describing the functioning of the UPS are presented .The control schemes are derived from volt-second balance of the input and output filter inductors. The regulation of ac and dc voltages is presented using small signal models. The transient and steady state performances are simulated for wide range of line and load variations. The switching pulses are generated with simple comparators and integrators. The performance of the controlled UPS is studied in simulation as well as experimentation. The method proposed in this paper has ensured sinusoidal input and output inductor currents and proper regulation of the ac and dc voltages.

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