A Power Electronic Transformer for Three Phase PWM AC/AC Drive with Loss Less Commutation and Common-mode Voltage Suppression

Abstract—**This paper presents a novel topology for the generation of adjustable frequency and magnitude PWM three phase ac from a balanced three phase ac source with a high frequency ac link. The proposed power electronic transformer system provides single stage power conversion with bidirectional power flow capability. This topology along with the proposed control has the following advantages : 1) Input power factor correction, 2) Common-mode voltage suppression at the load end, 3) High quality output voltage waveform (comparable to conventional space vector PWM modulated two level inverter) and 4) Minimization of output voltage loss, common-mode voltage switching and distortion of the load current waveform due to leakage inductance commutation. A loss less source based commutation technique for the leakage energy has been proposed. This results in soft switching of the output side converter. The entire topology along with the proposed control scheme has been analyzed and simulated. The simulation results verify the operation and advantages of the proposed power electronic transformer.**

I. Introduction

Three phase ac/ac power conversion with a high frequency ac link has potential applications in generation of electric power from renewable energy sources [1]. One of the most important area of application is wind-power systems. A transformer is used to match the voltage level and to provide galvanic isolation. A high frequency transformer (HFT) has much lower size and volume in comparison with its line frequency counterpart. Replacement of conventional transformer with HFT implies increase in power density and reduction in the cost of copper and iron.

Single phase ac to constant frequency controllable amplitude ac with a high frequency ac link is described in [2] and [3]. For the same topology, a control strategy based on phase modulated converter is proposed in [4]. This results in soft switching when the output voltage and current are in the same quadrant. Single and three phase ac/ac converters with HFT based on flyback or push-pull topologies with multiple power conversion stages are proposed in [5], [6] and [7].

This paper focuses on topologies that provide single stage power conversion with bidirectional power flow capability without using any storage element. Matrix converter is used for direct three phase ac/ac power conversion. In literature [8]–[12] there exist three different matrix converter-based approaches for single stage three-phase ac/ac power conversion with a high frequency ac link.

The first method is based on the indirect modulation of the matrix converter. The input or primary side converter chops the virtual dc-link into a high frequency square waveform in an average sense [8], [9]. The load or secondary side converter synthesizes adjustable frequency and magnitude three-phase ac from the high frequency ac available at the transformer secondary. In the second approach, the input three-phase ac is chopped at a high frequency and applied to the primary of three HFTs. This can be done either by a full-bridge [10] [13] or by a push-pull [11] type of configuration in the primary side converter. A matrix converter is used in the secondary side in order to generate the required output voltage from the high frequency three-phase ac.

The windings of a transformer have leakage inductances. The output load, generally a three phase ac machine, is inductive. In both of the above-mentioned approaches, any change in the switching state of the output converter requires commutation of the leakage energy. Commutation of leakage inductances with clamp or snubber circuit is lossy. Commutation also results in the loss of the output voltage and distortion in the output load current [14]. It also causes common-mode voltage switching. Topologies described in [9]–[11] provides common-mode voltage elimination assuming there is no leakage inductance.

In [15] (third group), a topology has been proposed to minimize the number of switching transitions between the secondary winding of the transformer and the output load over a period in which the output voltage is synthesized along with a loss less source based commutation method of the leakage energy. This commutation method results in soft-switching (ZCS) of all the switches in the load side converter. In a matrix converter synchronously rotating vectors can be used in order to get zero common-mode voltage [16]. In this converter these vectors can also be used. But at any instant of time only three vectors are available rotating in particular direction, for modulation. This reduces the quality of the output voltage waveform.

This paper presents a topology (Fig. 1) that belongs to the third group, where six synchronously rotating voltage vectors are available for modulation (similar to a two level inverter). Modulation with synchronously rotating vectors in this topology results in high quality output voltage waveform. Also the loss less source based commutation process is simpler compared to [15]. In this topology, availability of more voltage (in comparison with [15]) for leakage energy commutation implies lower commutation time, higher frequency of operation and high power density. During source

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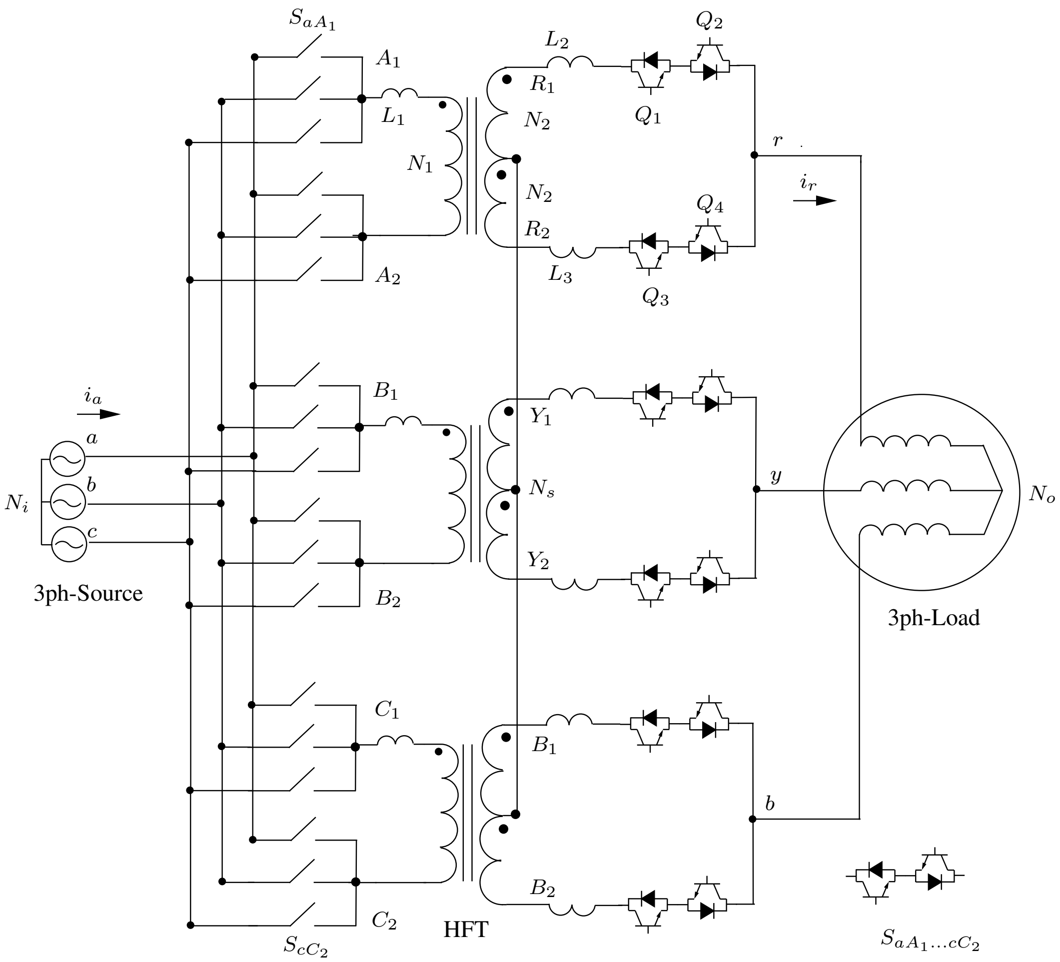

Fig. 1. Circuit diagram of the proposed topology

based commutation a zero vector gets applied to the load. In this converter commutation is done when modulation of the output voltage vector requires an application of zero vector. This results in less amount of output voltage loss during commutation. The advantages of the proposed converter are:

1) Voltage matching, isolation and high power density
2) Common-mode voltage suppression
3) A simple loss less commutation of leakage energy
4) Minimization of output voltage loss
5) High quality output voltage waveform
6) Single-stage power conversion (no storage element) with bidirectional power flow capability
7) Input power factor correction
8) Soft-switching

II. Analysis

The operation of the proposed power electronic system is described using a set of signals shown in Fig. 2. T_s is the period over which the average output voltages are synthesized. When S_1 is high, counter clockwise (CCW) vectors are used for modulation. Clockwise (CW) vectors are used for the rest of the time. The operation of this converter can be divided into two parts, modulation and commutation.

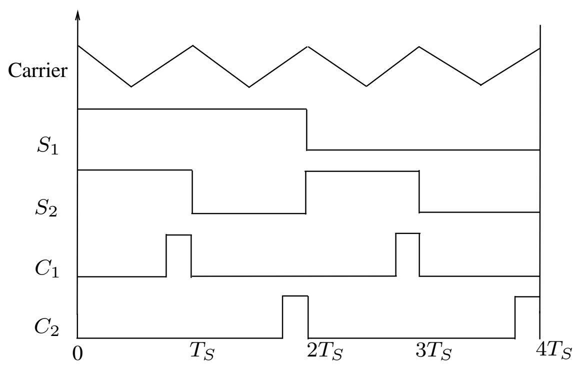

Fig. 2. Control Signals

A. Modulation

When S_2 is high power is transferred through the upper half of the secondary winding of each of the transformers (Q_1 and Q_2 is on, Fig. 1). Here the modulation of the output voltages with CCW vectors is explained in detail (S_1 is high). The three phase balanced voltages are given in (1). (2) gives the average output line to neutral voltages generated at the load end. (3) defines the instantaneous output voltage vector.

The reference average output voltage vector is given in (4). Using transformer relationship (5) and definitions (6), (7), (8), (9) it is possible to get (10). During this state ($S_1 = 1$ and $S_2 = 1$) $\mathbf{V_{s_1}} = \mathbf{V_o}$ and $\mathbf{V_o} = \frac{N_2}{N_1}\left(\mathbf{V_{A_1B_1C_1}} - \mathbf{V_{A_2B_2C_2}}\right)$. In Fig. 3 $\mathbf{V_{A_1B_1C_1}} = \mathbf{u_1} = \frac{3}{2}V_i e^{j\omega_i t}$ when switches S_{aA_1}, S_{bB_1} and S_{cC_1} are on. When switches S_{bA_2}, S_{cB_2} and S_{aC_2} are on, $-\mathbf{V_{A_2B_2C_2}} = \mathbf{w_3} = \frac{3}{2}V_i e^{(j\omega_i t + \frac{\pi}{3})}$. Combination of $\mathbf{u_{1,2,3}}$ with $\mathbf{w_{1,2,3}}$ produces six active voltage vectors $\mathbf{V_{1,2..,6}}$ (Fig. 3) which are available for modulation in this state. Here, modulation is similar to that of a two level voltage source inverter. For example, if at a particular instant of time $\mathbf{V_{ref}}$ is located in a sector formed by vectors $\mathbf{V_1}$ and $\mathbf{V_2}$, $\mathbf{V_{ref}}$ is synthesized on an average using these two vectors, (11), (12), (13). Where d_1 is the fraction of time for which $\mathbf{V_1}$ is applied, m is equal to $\left(\frac{N_1}{N_2}\right)\frac{V_o}{\sqrt{3}V_i}$ and $\alpha = \omega_o t + \phi - (\omega_i t - \frac{\pi}{6})$. The zero vector is obtained by the simultaneous use of $\mathbf{u_1}$ and $\mathbf{w_1}$. A mid point clamped carrier is used and the sequence in which the vectors are applied is 0120-0210. It is possible to show that application of these vectors results in zero common-mode voltage ($V_{cm} = \frac{1}{3}\left(V_{rN_o} + V_{yN_o} + V_{bN_o}\right)$).

$$
\begin{aligned}
v_{aN_i} &= V_i \cos \omega_i t \\
v_{bN_i} &= V_i \cos\left(\omega_i t - \frac{2\pi}{3}\right) \\
v_{cN_i} &= V_i \cos\left(\omega_i t + \frac{2\pi}{3}\right)
\end{aligned} \tag{1}
$$

$$
\begin{aligned}
\overline{v_{rN_o}} &= V_o \cos\left(\omega_o t + \phi\right) \\
\overline{v_{yN_o}} &= V_o \cos\left(\omega_o t - \frac{2\pi}{3} + \phi\right) \\
\overline{v_{bN_o}} &= V_o \cos\left(\omega_o t + \frac{2\pi}{3} + \phi\right)
\end{aligned} \tag{2}
$$

$$\mathbf{V_o} = v_{rN_o} + v_{yN_o} e^{j\frac{2\pi}{3}} + v_{bN_o} e^{-j\frac{2\pi}{3}} \tag{3}$$

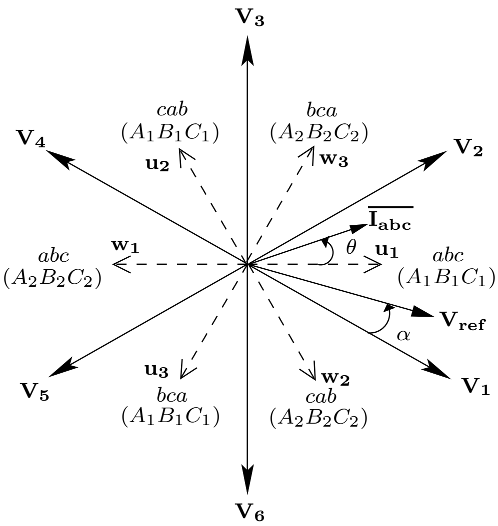

Fig. 3. Modulation with anti clockwise vectors

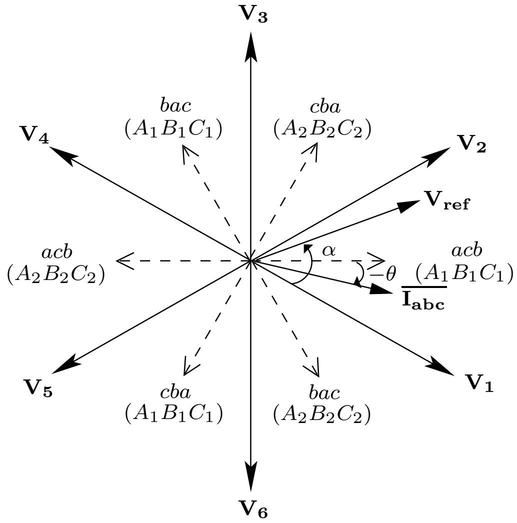

Fig. 4. Modulation with clockwise vectors

$$
\begin{aligned}
\mathbf{V_{ref}} &= \overline{v_{rN_o}} + \overline{v_{yN_o}} e^{j\frac{2\pi}{3}} + \overline{v_{bN_o}} e^{-j\frac{2\pi}{3}} \\
&= \frac{3}{2} V_o e^{(j\omega_o t + \phi)}
\end{aligned} \tag{4}
$$

$$
\begin{aligned}
v_{R_1N_s} &= \frac{N_2}{N_1}\left(v_{A_1N_i} - v_{A_2N_i}\right) \\
v_{Y_1N_s} &= \frac{N_2}{N_1}\left(v_{B_1N_i} - v_{B_2N_i}\right) \\
v_{B_1N_s} &= \frac{N_2}{N_1}\left(v_{C_1N_i} - v_{C_2N_i}\right)
\end{aligned} \tag{5}
$$

$$\mathbf{V_{s_1}} = v_{R_1N_s} + v_{Y_1N_s} e^{j\frac{2\pi}{3}} + v_{B_1N_s} e^{-j\frac{2\pi}{3}} \tag{6}$$

$$\mathbf{V_{s_2}} = v_{R_2N_s} + v_{Y_2N_s} e^{j\frac{2\pi}{3}} + v_{B_2N_s} e^{-j\frac{2\pi}{3}} \tag{7}$$

$$\mathbf{V_{A_1B_1C_1}} = v_{A_1N_i} + v_{B_1N_i} e^{j\frac{2\pi}{3}} + v_{C_1N_i} e^{-j\frac{2\pi}{3}} \tag{8}$$

$$\mathbf{V_{A_2B_2C_2}} = v_{A_2N_i} + v_{B_2N_i} e^{j\frac{2\pi}{3}} + v_{C_2N_i} e^{-j\frac{2\pi}{3}} \tag{9}$$

$$\mathbf{V_{S_1}} = \frac{N_2}{N_1}\left(\mathbf{V_{A_1B_1C_1}} - \mathbf{V_{A_2B_2C_2}}\right) \tag{10}$$

$$\mathbf{V_{ref}} = \frac{N_2}{N_1}\left(\mathbf{V_1} d_1 + \mathbf{V_2} d_2\right) \tag{11}$$

$$m e^{j\left(\omega_o t + \phi - \left[\omega_i t - \frac{\pi}{6}\right]\right)} = d_1 + d_2 e^{j\frac{\pi}{3}} \tag{12}$$

$$
\begin{aligned}
d_1 &= m \frac{\sin\left(\frac{\pi}{3} - \alpha\right)}{\sin\frac{\pi}{3}} \\
d_2 &= m \frac{\sin \alpha}{\sin\frac{\pi}{3}}
\end{aligned} \tag{13}
$$

TABLE I
$i_o > 0$

	Q_1	Q_2	Q_3	Q_4	D_1	D_2	D_3	D_4	V_p
P	1	X	0	0	0	1	0	0	X
C_1	1	0	1	0	0	1	0	1	-Ve
N	0	0	1	X	0	0	0	1	X
C_2	1	0	1	0	0	1	0	1	+Ve

TABLE II
$i_o < 0$

	Q_1	Q_2	Q_3	Q_4	D_1	D_2	D_3	D_4	V_p
P	X	1	0	0	1	0	0	0	X
C_1	0	1	0	1	1	0	1	0	+Ve
N	0	0	X	1	0	0	1	0	X
C_2	0	1	0	1	1	0	1	0	-Ve

Assuming the load power factor angle to be θ, the output load currents are given in (14). When vector $\mathbf{V_1}$ is applied $i_a = \frac{N_2}{N_1}(i_r - i_y)$, as a is connected to A_1 and B_2. Similarly during the application of $\mathbf{V_2}$, $i_a = \frac{N_2}{N_1}(i_r - i_b)$. The average input currents during this state are given in (15). Using definition (16), the average input current vector during this state is given by (17).

In the following cycle, when S_2 is low, power is transferred through the lower half of the secondary windings. In this state $\mathbf{V_{s_2}} = \mathbf{V_o}$ and $\mathbf{V_o} = -\frac{N_2}{N_1}(\mathbf{V_{A_1B_1C_1}} - \mathbf{V_{A_2B_2C_2}})$. Vectors $\mathbf{V_{1,2..,6}}$ (Fig. 4) are used to generate $-\mathbf{V_{ref}}$. This results in flux balance. In this state, the average input current vector is given (18). The average $\overline{I_{abc}}$ over one complete cycle of S_1 is $\frac{3}{2}I_o m e^{j\omega_i t}$ i.e. in phase with the input voltage vector. This results in input power factor correction.

$$\begin{aligned} i_r &= I_o\cos(\omega_o t + \phi + \theta) \\ i_y &= I_o\cos\left(\omega_o t - \frac{2\pi}{3} + \phi + \theta\right) \\ i_b &= I_o\cos\left(\omega_o t + \frac{2\pi}{3} + \phi + \theta\right) \end{aligned} \tag{14}$$

$$\begin{aligned} \overline{i_a} &= \frac{N_2}{N_1}\left[(i_r - i_y)\,d_1 + (i_r - i_b)\,d_2\right] \\ \overline{i_b} &= \frac{N_2}{N_1}\left[(i_y - i_b)\,d_1 + (i_y - i_r)\,d_2\right] \\ \overline{i_c} &= \frac{N_2}{N_1}\left[(i_b - i_r)\,d_1 + (i_b - i_y)\,d_2\right] \end{aligned} \tag{15}$$

$$\overline{\mathbf{I_{abc}}} = \overline{i_a} + \overline{i_b}e^{j\frac{2\pi}{3}} + \overline{i_c}e^{-j\frac{2\pi}{3}} \tag{16}$$

$$\overline{\mathbf{I_{abc}}} = \frac{3}{2}I_o m e^{(j\omega_i t + \theta)} \tag{17}$$

$$\overline{\mathbf{I_{abc}}} = \frac{3}{2}I_o m e^{(j\omega_i t - \theta)} \tag{18}$$

B. Commutation

Commutation is required at each transition of the signal S_2. Commutation refers to the following processes 1) reversal of the current in the primary leakage inductance (L_1) 2) exchange of the load current between the leakage inductances (L_2 and L_3) of the two halves of the secondary winding. Commutation is done on a per phase basis by applying a proper voltage at the transformer primary and controlling the individual igbts in the secondary side converter. Depending on the direction of the load current and two possible transitions of S_2, four cases are possible. The switching scheme for $Q_{1,2,3,4}$ and the direction of the applied primary voltage is given in Table I and II for all of these four cases. Here, the case when load current is positive and S_2 is making a transition from high to low is described in detail. Fig. 5(a), depicts the circuit just before C_1 goes high. The commutation stage is given in Fig. 5(b). According to Table I a negative voltage is required to be applied at the transformer primary. Say at this instant of time, it is found that V_{ab} is maximum in magnitude and negative in sign. Application of this voltage and turning on of Q_3 (at zero current) forward biases diode D_4. In this analysis magnetizing currents are neglected. According to transformer relationships (19) and (20) are valid. I_r is the value of i_r at this instant of time. By KCL at point r we get (21). (22) and (23) are obtained by applying KVL to the primary and secondary windings. Solution of these equations leads to (24). This gives the rate of change of the leakage inductance current i_3. The commutation time is maximum when the output load current is at its peak (I_o) and the available maximum line to line voltage is at its minimum ($V_i\sqrt{3}\cos\frac{\pi}{6}$). The time period for which C_1 and C_2 is high is chosen to be equal to T_{com}, (25). When i_3 reaches I_r, i_2 and i_1 becomes zero and $-I_r$ respectively and the commutation process comes to a natural end. When C_1 goes low Q_1 is turned off and Q_4 is turned on at zero current (ZCS). Fig. 5(c) shows the circuit configuration just after the commutation process is over.

$$\frac{e_1}{N_1} = \frac{e_2}{N_2} = \frac{e_3}{N_2} \tag{19}$$

$$i_1N_1 - i_2N_2 + i_3N_3 = 0 \tag{20}$$

$$I_r = i_2 + i_3 \tag{21}$$

$$v_{ab} = L_1\frac{d}{dt}i_1 + e_1 \tag{22}$$

$$e_2 + e_3 = L_2\frac{d}{dt}i_2 - L_3\frac{d}{dt}i_3 \tag{23}$$

$$\frac{d}{dt}i_3 = -\frac{v_{ab}\left(\frac{N_2}{N_1}\right)}{\left(\frac{L_2+L_3}{2}\right) + 2L_1\left(\frac{N_2}{N_1}\right)^2} \tag{24}$$

$$T_{com} = \left[\frac{\left(\frac{L_2+L_3}{2}\right) + 2L_1\left(\frac{N_2}{N_1}\right)^2}{V_i\sqrt{3}\cos\frac{\pi}{6}\left(\frac{N_2}{N_1}\right)}\right]I_o \tag{25}$$

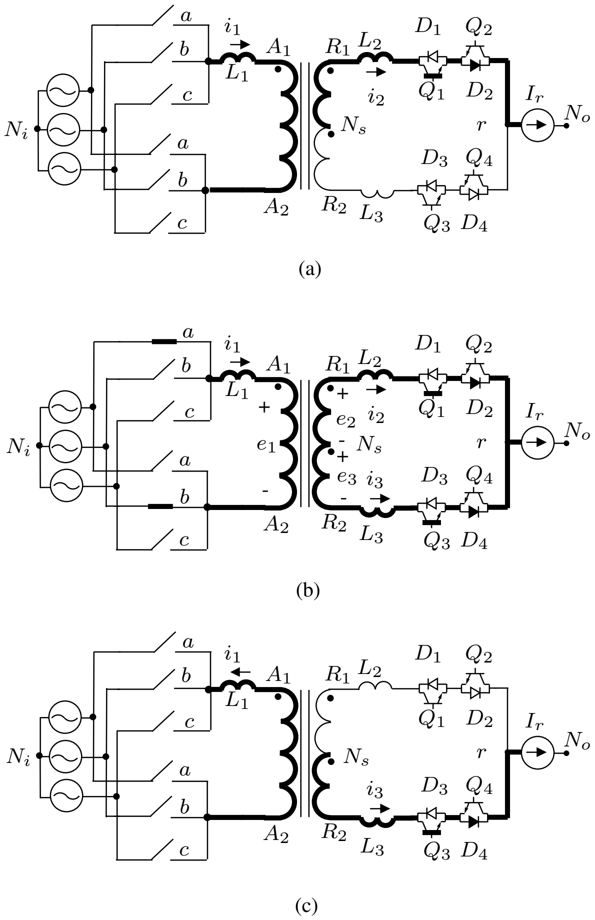

Fig. 5. Commutation

TABLE III
PARAMETERS

L_{load}	$10mH$
R_{load}	2.5Ω
$L_{1,2,3}$	$40\mu H$
$R_{1,2,3}$	0.2Ω
L_m	$15\ mH$
$\frac{N_2}{N_1}$	1
V_i	500V
$f_s = \frac{1}{T_s}$	5kHz
ω_i	$2\pi 60$
ω_o	$2\pi 60$
m	$\frac{0.4}{\sqrt{3}}$

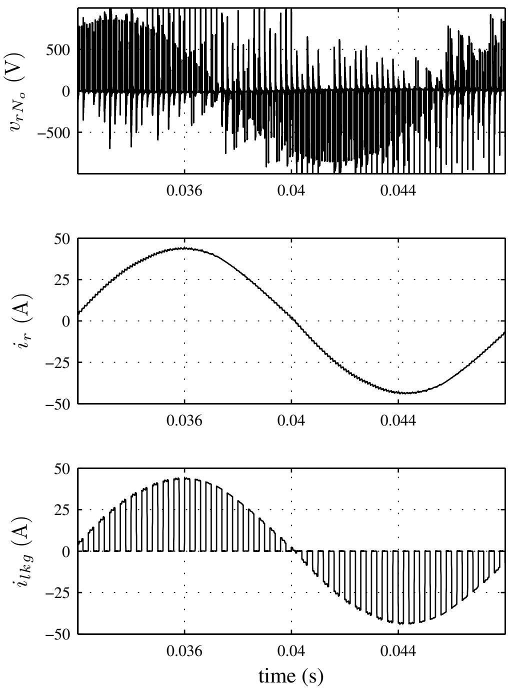

Fig. 6. Simulation result: a) output voltage b) output current c) current through the upper half of secondary winding

III. SIMULATION

The proposed topology along with the suggested control has been simulated in MATLAB/Simulink. The parameters for the simulation are given in Table III. Fig. 6 presents simulated output voltage, load current and current through the upper half of the secondary winding in phase r. The peak of the output load current is slightly lower than its analytically predicted value (Analytical $I_r = 44.21A$, Simulated $I_r = 43.5A$). This is due to voltage loss during commutation. Fig. 7 shows the filtered input current in phase a along with the corresponding input line to neutral voltage. This confirms input power factor correction. Simulated waveforms for one cycle of the signal S_2 is given in Fig. 8. The current in the upper half of the phase r winding (waveform i_{lkg}) linearly changes to its desired value during C_1 and C_2 as predicted by (24). These slopes match with analytical predictions. The magnetising current of the transformer in phase r is shown as the waveform i_m. It verifies flux balance over one cycle of S_2. The last waveform is the common-mode voltage. It is zero except during commutation (when C_1 and C_2 are high).

IV. CONCLUSION

In this paper a new power electronic transformer for direct three-phase ac/ac conversion has been proposed. Here, the power flow is bidirectional. A control technique has been developed which ensures loss less commutation along with high quality output voltage generation (comparable to space vector moudulated two level inverter) and common-mode voltage suppression at load end. Actually as commutation applies a zero vector to the load and commutation happens when zero vector is need to be applied due to modulation, the voltage loss due to commutation is low.The circuit has been analysed and its operation for leakage energy commutation and output voltage generation has been described. The following advantages of the proposed topology have been confirmed through the presented simulation results:

1) Common-mode voltage suppression.
2) Minimum number of switching transitions between the load and the transformer secondary windings. This

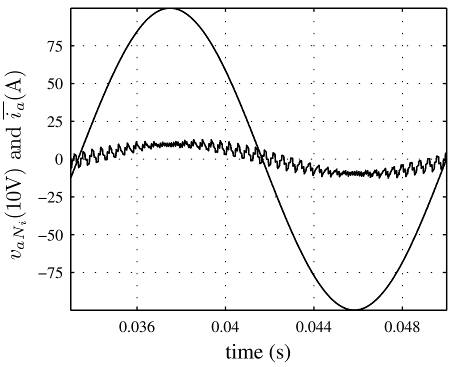

Fig. 7. Simulation results: Input voltage and filtered input current

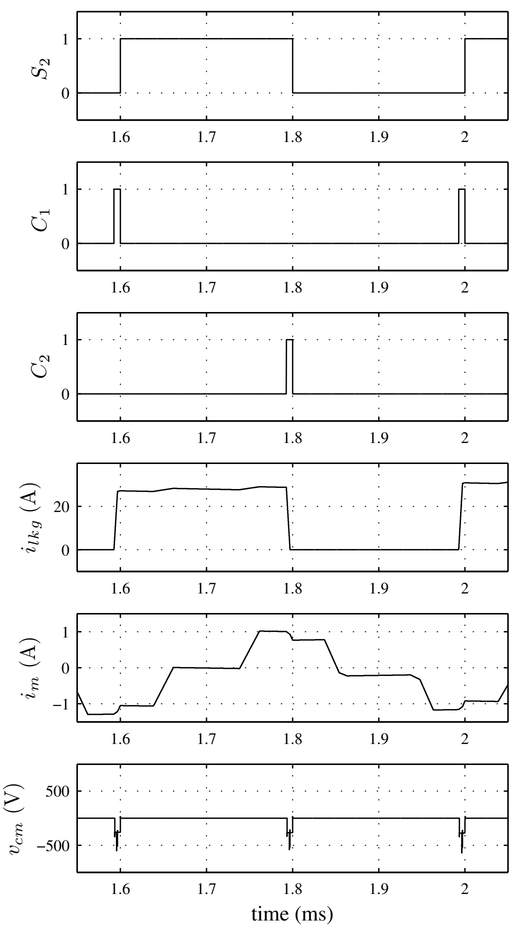

Fig. 8. Simulation results: Commutation

implies lower distortion and loss in the output voltage.

3) Loss less commutation of the leakage energy
4) Zero current switching (ZCS) in all secondary side switches
5) Input power factor correction

This topology is a promising and comprehensive solution for direct ac/ac three phase motor drive application with high frequency ac link. However, since the modulation for output voltage generation is done in the input side, the circuit has more number of switches in the input side (usually high voltage) which may be not be desirable.

REFERENCES

[1] R. K. Gupta, G. F. Castelino, K. K. Mohapatra, and N. Mohan, "A novel integrated three-phase, switched multi-winding power electronic transformer converter for wind power generation system," in *Proc. Industrial Electronics Society, 2009. IECON'09. The 35th Annual Conference of the IEEE*, Porto, Portugal, Nov. 2009.

[2] M. Kang, P. Enjeti, and I. Pitel, "Analysis and design of electronic transformers for electric power distribution system," in *Industry Applications Conference, 1997. Thirty-Second IAS Annual Meeting, IAS '97., Conference Record of the 1997 IEEE*, vol. 2, Oct 1997, pp. 1689–1694 vol.2.

[3] H. Krishnaswami and V. Ramanarayanan, "Control of high-frequency ac link electronic transformer," *Electric Power Applications, IEE Proceedings -*, vol. 152, no. 3, pp. 509–516, 6 May 2005.

[4] D. Chen and J. Liu, "The uni-polarity phase-shifted controlled voltage mode ac-ac converters with high frequency ac link," *Power Electronics, IEEE Transactions on*, vol. 21, no. 4, pp. 899–905, July 2006.

[5] D. Tang and L. Li, "Analysis and simulation of push-pull three level ac/ac converter with high frequency link," in *Industrial Electronics and Applications, 2009. ICIEA 2009. 4th IEEE Conference on*, May 2009, pp. 3366–3371.

[6] D. Chen, L. Li, J. Liu, S. Lin, and C. Song, "Novel current mode ac/ac converters with high frequency ac link," in *Power Electronics Specialists Conference, 2005. PESC '05. IEEE 36th*, June 2005, pp. 39–44.

[7] M. Manjrekar, R. Kieferndorf, and G. Venkataramanan, "Power electronic transformers for utility applications," in *Industry Applications Conference, 2000. Conference Record of the 2000 IEEE*, vol. 4, Oct 2000, pp. 2496–2502 vol.4.

[8] H. Cha and P. Enjeti, "A three-phase ac/ac high-frequency link matrix converter for vscf applications," in *Power Electronics Specialist Conference, 2003. PESC '03. 2003 IEEE 34th Annual*, vol. 4, June 2003, pp. 1971–1976 vol.4.

[9] K. Basu, A. Umarikar, K. Mohapatra, and N. Mohan, "High-frequency transformer-link three-level inverter drive with common-mode voltage elimination," in *Power Electronics Specialists Conference, 2008. PESC 2008. IEEE*, June 2008, pp. 4413–4418.

[10] K. Mohapatra and N. Mohan, "Matrix converter fed open-ended power electronic transformer for power system application," in *Power and Energy Society General Meeting - Conversion and Delivery of Electrical Energy in the 21st Century, 2008 IEEE*, July 2008, pp. 1–6.

[11] R. K. Gupta, K. K. Mohapatra, and N. Mohan, "A novel three-phase, switched multi-winding power electronic transformer," in *Proc. IEEE Energy Conversion Congress and Exposition (ECCE) 2009*, San Jose, CA, Sep. 2009, pp. 2696 –2703.

[12] K. Basu, R. K. Gupta, S. Nath, G. F. Castelino, K. K. Mohapatra, and N. Mohan, "Research in matrix-converter based three-phase power-electronic transformers," in *Power Electronics Conference (IPEC), 2010 International*, jun. 2010, pp. 2799 –2803.

[13] R. Gupta, K. Mohapatra, and N. Mohan, "Novel topologies of power electronic transformers with reduced switch-count," in *Grand Challenges in Modeling and Simulation (GCMS'09)*, July 2009.

[14] S. Nath, K. Mohapatra, and N. Mohan, "Output voltage regulation in matrix converter fed power electronic transformer for power systems application in electric ship," in *Electric Ship Technologies Symposium, 2009. ESTS 2009. IEEE*, april 2009, pp. 203 –206.

[15] K. Basu, A. Somani, K. Mohapatra, and N. Mohan, "Three phase ac/ac power electronic transformer based pwm ac drive with loss less commutation of leakage energy," in *SPEEDAM 2010, accepted for presentation*, June 2010.

[16] R. Gupta, K. Mohapatra, A. Somani, and N. Mohan, "Direct-matrix-converter based drive for a three-phase open-end-winding ac machine with advanced features," *Industrial Electronics, IEEE Transactions on*, vol. PP, no. 99, pp. 1 –1, 2010.