A HYBRID 18-PULSE RECTIFICATION SCHEME FOR DIODE FRONT END VARIABLE FREQUENCY DRIVES

Mahesh Swamy
Member, IEEE
Yaskawa Electric America
Waukegan, IL 60085, USA
mahesh_swamy@yaskawa.com

Tsuneo J. Kume
Fellow, IEEE
Yaskawa Electric Corporation
Kokura, Fukuoka, 803-8530, Japan
tjkume@yaskawa.co.jp

Noriyuki Takada
Yaskawa Electric America
Waukegan, IL 60085, USA
nory_takada@yaskawa.com

Abstract - Diode rectifier with large DC bus capacitors, used in the front ends of Variable Frequency Drives (VFDs), draw discontinuous current from the power system resulting in current distortion and hence voltage distortion. Typically, the power system can handle current distortion without showing signs of voltage distortion. However, when the majority of the load on a distribution feeder is made up of VFDs, current distortion becomes an important issue. Multi-pulse techniques to reduce input harmonics are popular because they do not interfere with the existing power system either from higher conducted EMI when active techniques are used or from possible resonance, when capacitor based filters are employed.

In this paper, a new 18-pulse topology is proposed that has two six-pulse rectifiers powered via a phase-shifting isolation transformer, while the third six-pulse rectifier is fed directly from the AC source via a matching-impedance. This idea relies on harmonic current cancellation strategy rather than the flux cancellation method and results in lower overall harmonics. It is also seen to be smaller in size and weight, and lower in cost compared to an isolation transformer. Experimental results are given to validate the concept.

Index Terms- Eighteen Pulse Rectification, Harmonics, Multi-pulse Rectifiers, Power Quality.

I. INTRODUCTION

Harmonic distortion concerns are serious when the power ratings of the VFD load increases. Large power VFDs are gaining in popularity due to their low cost and impressive reliability. In many large power installations, current harmonic distortion levels achievable using 12-pulse techniques are insufficient to meet the levels recommended in IEEE 519 (1992) [1]. In view of this, lately, the 18-pulse VFD systems are being proposed to achieve much superior harmonic performance compared to the 12-pulse systems.

In general, the characteristic harmonics generated by a semiconductor rectifier is given by:

$$h = kq \pm 1 \qquad (1)$$

where, h is the order of harmonics; k is any integer, and q is the pulse number of the rectifier (six for a 6-pulse rectifier). The per unit value of the characteristic harmonics present in the theoretical current waveform (assumes a rectangular current waveform spanning 120 electrical degrees) at the input of the semiconductor rectifier is given by $1/h$. In practice, the observed per unit value of the harmonics is much greater than $1/h$. From these observations, it is clear that increasing the pulse number from 6 to either 12 or 18 will significantly reduce the amplitude of low order harmonics and hence the total current harmonic distortion.

For 18-pulse operation, there is a need for three sets of 3-phase AC supply that are phase shifted with respect to each other by 20 electrical degrees. Traditionally, this is achieved using a four winding isolation transformer that has one set of primary windings and three sets of secondary windings. One set of secondary winding is in phase with the primary winding, while the other two sets are phase shifted by +20 electrical degrees and –20 electrical degrees with the primary. This arrangement yields three phase-shifted supplies that allow 18-pulse operation as shown in Fig. 1. The main disadvantage of the scheme shown in Fig 1 is that the isolation transformer is large in size, heavy and costly.
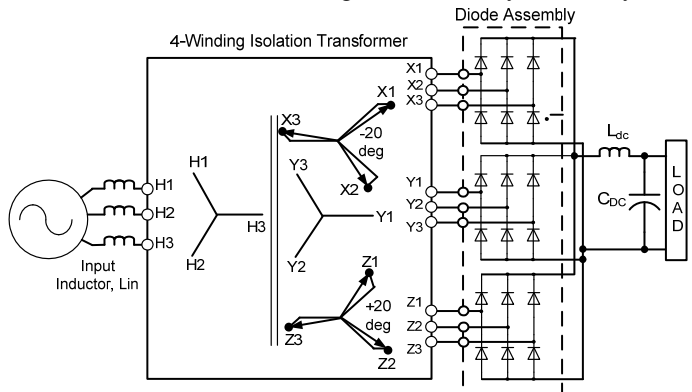

Fig. 1. Schematic representation of a traditional 18-pulse converter circuit.

Instead of using $\pm$ 20 degree phase-shifted outputs from isolation transformer for 18-pulse operation, a nine-phase supply can be used, where each phase lags the other by 40 electrical degrees. Autotransformers have been developed that implement the above idea [2], and are widely used. Due to the nature of autotransformers, the size, weight, and cost can be reduced compared to the conventional technique shown in Fig. 1 [3], [4]. However, there are some issues that need to be addressed while utilizing autotransformers.

A popular 18-pulse autotransformer configuration is shown in Fig 2 [5]. In the configuration of Fig. 2, the delta-connected tertiary winding is included in the wye fork. This construction is popularly known in the industry as the windmill topology. Initially the delta tertiary winding was present in each phase and the size of the transformer was still
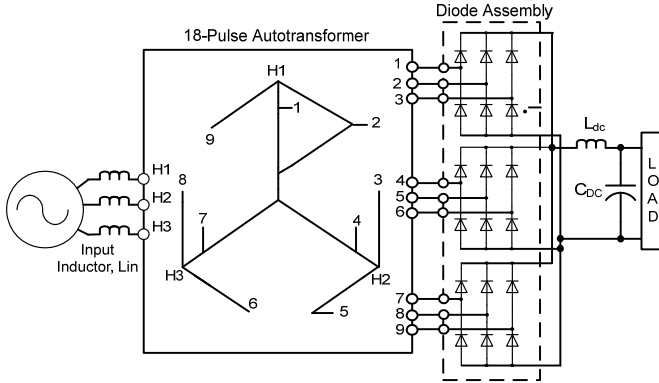
Fig. 2. Schematic of a typical windmill construction of the 18-pulse autotransformer configuration [5].

big. The effective kVA rating was about 60% of the load kVA rating. By intelligently removing the windmill structure from two of the three phases, it was shown that the performance remained equally good, while the effective rating was successfully reduced to 55% of the load kVA rating.

In all the 18-pulse autotransformer methods, the change of current from one conducting diode pair to the other is quick. Though the rms current rating may not exceed the current rating of the diode, attention should be given to the di/dt of the current through the diodes. One solution is to use additional inductors in between the autotransformer and the input rectifier to lower the di/dt. This makes the overall scheme bulky and expensive. The rectangular current through the windings also increases losses, prompting the need to use fans to keep the size of the transformer small.

II. ISSUES WITH EXISTING AUTOTRANSFORMER BASED TOPOLOGIES

Some important shortcomings of the autotransformer based topologies are as follows:

a. Autotransformer based topologies require significant input impedance (shown as L_{IN} in Fig. 2) to smooth the current and reduce the overall input current distortion,

b. Autotransformer techniques utilize complex winding structures, either of the stub-type or the polygon type. These transformers are labor intensive to manufacture and result in poor core utilization,

c. Because of complicated winding structure and the fact that partial turns are not practically feasible to build, the rounding off error influences the final performance, and so large input impedance is used to mask this problem,

d. Higher than normal di/dt stress in rectifier diodes and increased winding loss due to rectangular current waveform should be considered for high reliability systems.

e. Autotransformer topologies that convert a 3-phase system to a 9-phase output create an aberration in the dc bus ripple content of a VFD. When one or two of nine output phases has a bad rectifier due to high di/dt, the increase in dc bus

ripple is hardly noticeable and this reduces the chance for detection of failure. The power flow is now shared by existing rectifiers that can eventually fail.

Given the above important drawbacks, it is clear that there exists a need for a better 18-pulse topology that does not suffer from the same disadvantages pointed out here.

III. PROPOSED TOPOLOGY

The proposed topology is derived from the structure shown in Fig. 1. One disadvantage of the four winding arrangement of Fig. 1 is its size and cost. On re-examining this circuit, it can be noted that one set of windings does not have any phase-shift with respect to the primary windings. This is important because it allows one six-pulse rectifier circuit to be connected directly to the AC source via a balancing inductor to match the inductance in front of the other two sets of six-pulse rectifier circuits to achieve 18-pulse operation.

The above mentioned concept was successfully exploited to derive the hybrid 12-pulse structure, mentioned in reference [6]. In [6], a standard delta-wye transformer rated at half power is utilized to achieve the necessary 30-deg phase shift, and the other path is fed directly into the 2nd rectifier via a matching inductor to achieve optimal sharing of current. It was shown in [6] that the concept of current cancellation achieves better harmonic performance compared to a traditional isolation transformer based 12-pulse scheme that is based on flux cancellation concept.

Similar to the idea in [6], the proposed 18-pulse scheme has two six-pulse rectifiers powered via a phase-shifting isolation transformer, while the third six-pulse rectifier is fed directly from the AC source via a matching-impedance. Such an 18-pulse arrangement will be referred to in this paper as hybrid 18-pulse configuration and is shown in Fig. 3. The phase-shifting zigzag transformer feeding two of the three six-pulse rectifiers is sized to handle 2/3 the rated load power. Similarly, the matching inductor is sized to carry only 1/3 the rated input load current. This arrangement results in the overall size of the transformer and matching inductor combination to be smaller and less expensive than the four winding arrangement of Fig. 1. As an example, a 50hp conventional 18-pulse transformer without input inductor L_{IN} was recently quoted to be 42"(H)x36"(W)x24"(D) with an estimated weight of 880-lbs, while the hybrid 18-pulse structure (without L_{IN}) to handle the same load was quoted to be 24"(H)x26"(W)x14"(D) with an estimated weight of 550-lbs. The proposed topology requires a matching inductor to perform comparably, which will add 12-lbs, for a total weight of 562-lbs. Both these structures are naturally cooled.

The phase shift in the transformer shown in Fig. 3 is achieved by winding extra teaser windings on appropriate limbs of a transformer. The teaser windings are marked "T" with subscript denoting the phase that they are wound on. For example, T_{H21} denotes a teaser winding that is wound on the H2 winding of the primary side of the isolation

transformer and is used in the first set of secondary winding to yield a phase shift of +20 degrees.

A. Theoretical Estimate of Input Current Distortion

In order to estimate the total current harmonic distortion in the input line current I_{in}, the following assumptions are made:
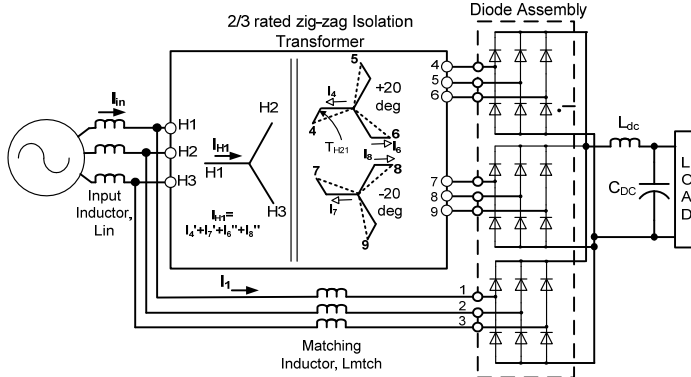

Fig. 3. Schematic of proposed hybrid 18-pulse topology.

a. The inductance in the dc bus is large enough to assume that the dc bus current has no ripple,

b. The current in the secondary windings of the transformer are rectangular due to the assumption in a. above,

c. The harmonic analysis is carried out at rated current, I_{DC},

d. The windings of the transformer and the matching inductance are well balanced such that the load current is equally shared among the three different paths. In other words, the height of the rectangular current pattern in each of the parallel paths is $I_{DC}/3$,

e. The primary winding marked as H1, H2, and H3, have N1 turns, the long part of the secondary fork has N2 turns and the short part of the secondary fork has N3 turns. The vector combination of N2 and N3 should yield N1 to result in a 1:1 transformation ratio.

In order to determine the contribution of secondary currents in the primary side input current, it is important to determine the turns-ratio among N1, N2, and N3. This is achieved using the expanded vector diagram shown in Fig. 4.

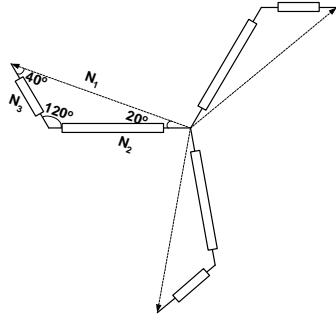**Fig. 4.** Vector representation of one of the secondary windings of Fig. 3.

From Fig. 4, the turns-ratio in terms of N_1 is computed as:

$$\frac{N_1}{\sin(120)} = \frac{N_2}{\sin(40)} = \frac{N_3}{\sin(20)}$$

$$N_2 = 0.7422 N_1$$

$$N_3 = 0.3949 N_1 \qquad (2)$$

From equation (2) and Fig. 3, the components that make up the primary current, I_{H1} are derived next.

$$I_{H1} = I_4' + I_7' + I_8'' + I_6''$$

$$I_4' = 0.7422 \cdot \frac{I_{dc}}{3} \angle 20^\circ$$

$$I_7' = 0.7422 \cdot \frac{I_{dc}}{3} \angle -20^\circ \qquad (3)$$

$$I_6'' = 0.3949 \cdot \frac{I_{dc}}{3} \angle -100^\circ$$

$$I_8'' = 0.3949 \cdot \frac{I_{dc}}{3} \angle -140^\circ$$

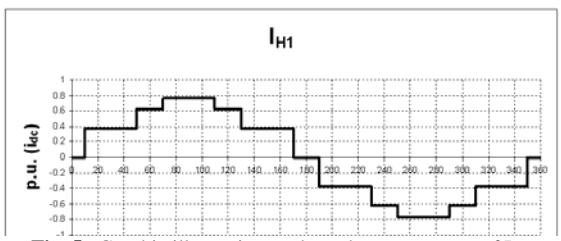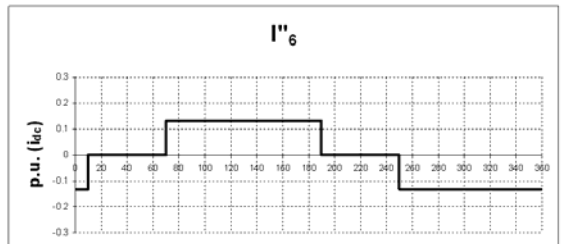The current wave shape of I_{H1} and the relative positions of the components of I_{H1} are shown in Fig. 5.
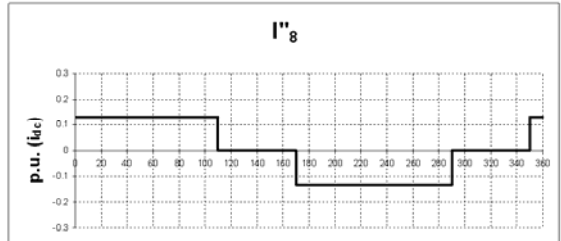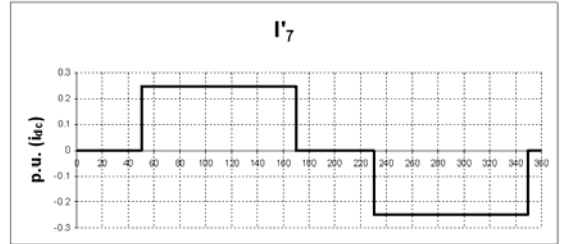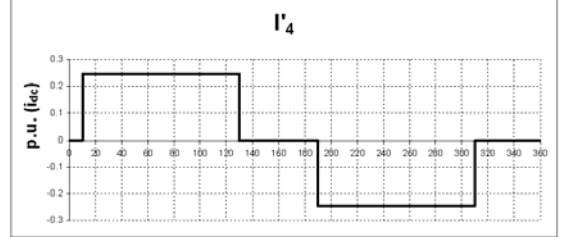

Fig. 5. Graphic illustration to show the components of I_{H1}.

The input current drawn from the ac system is a combination of the input current into the transformer and the current I_1 flowing through the matching inductor, L_{mtch}, into the third diode bridge. A graphic illustration of the components of I_{IN} is shown in Fig. 6. The waveform corresponding to I_{IN} in Fig. 6 will be used to mathematically estimate the total input current harmonic distortion.
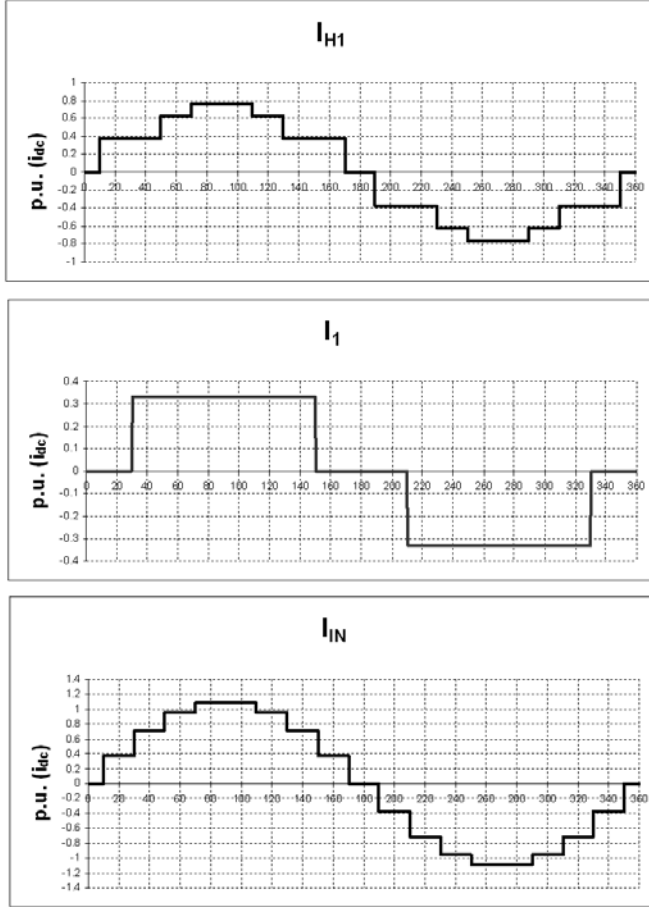

Fig. 6. Graphic illustration to show the components of I_{IN}.

The staircase waveform shown in Fig. 6 for I_{IN} is analyzed using Fourier series (Equation 4) to estimate the theoretical value of the total input current harmonic distortion.

To facilitate this process, certain features of I_{IN} are presented here. On inspecting the waveform shown in Figure 8, the following observations are made:

a. The waveform of I_{IN} does not have any dc component. This would mean that the dc component representation in the Fourier series is zero.

b. I_{IN} exhibits mirrored symmetry about the π axis. This would mean that I_{IN} is an odd function and not an even function. This property will render the even coefficient in the Fourier series to be zero.

c. From observations mentioned in items a through c, only the odd coefficient b_n needs to be computed.

Coefficient b_n is computed next.

$$b_n = \frac{2}{\pi} \cdot \int_0^\pi I_{in} \cdot \sin(n\theta) \cdot d\theta$$

$$= \frac{2 \cdot I_{DC}}{\pi} \cdot \left\{ \int_{\pi/18}^{\pi/6} 0.379 \cdot \sin(n\theta) \cdot d\theta + \int_{\pi/6}^{5\pi/18} 0.712 \cdot \sin(n\theta) \cdot d\theta \right.$$

$$+ \int_{5\pi/18}^{7\pi/18} 0.959 \cdot \sin(n\theta) \cdot d\theta + \int_{7\pi/18}^{11\pi/18} 1.091 \cdot \sin(n\theta) \cdot d\theta$$

$$+ \int_{11\pi/18}^{13\pi/18} 0.959 \cdot \sin(n\theta) \cdot d\theta + \int_{13\pi/18}^{5\pi/6} 0.712 \cdot \sin(n\theta) \cdot d\theta$$

$$\left. + \int_{5\pi/6}^{17\pi/18} 0.379 \cdot \sin(n\theta) \cdot d\theta \right\}$$

$$b_n = \frac{2 \cdot I_{DC}}{\pi} \cdot \left\{ \begin{array}{l} \dfrac{0.666}{n} \cdot \cos(n\pi/6) + \dfrac{0.494}{n} \cdot \cos(5n\pi/18) \\[2mm] + \dfrac{0.758}{n} \cdot \cos(n\pi/18) + \dfrac{0.264}{n} \cdot \cos(7n\pi/18) \end{array} \right\} \quad (4)$$

From equation (4), the following observations can be made:

a. The total current harmonic distortion is computed to be **8.8%** for the assumed staircase waveform.

b. There does not exist any triplen harmonics.

c. The first set of lowest order harmonics encountered is the 17th and the 19th.

d. The 17th is observed to be of negative sequence, while the 19th is seen to be of positive sequence.

B. Theoretical VA rating of the Proposed Transformer

From the waveform for the primary side input current of the proposed transformer configuration shown in Fig. 5, the VA rating of the transformer is computed next.

$$I_{H1} = \sqrt{\frac{1}{\pi} \cdot \left\{ \int_0^\pi (i'_4 + i'_7 + i''_6 + i''_8)^2 \cdot d\theta \right\}}$$

$$= \sqrt{\frac{4 \cdot I_{DC}^2}{18} \cdot \left\{ 0.626^2 + 0.758^2 + (2 \cdot 0.379^2) \right\}}$$

$$I_{H1} = 0.5278 \cdot I_{DC} \quad (5)$$

The VA rating of the proposed phase-shifting isolation transformer is computed next by multiplying the rms value of the input current with the rms value of the applied line-neutral voltage (V_{LN}) and further multiplying the result with 3 to account for all the windings.

$$VA_{xfmr} = 3 \cdot V_{LN} \cdot I_{H1} = 3 \cdot V_{LN} \cdot 0.5278 \cdot I_{DC}$$

$$= 1.5834 \cdot V_{LN} \cdot I_{DC}$$

$$P_{out} = \frac{3 \cdot \sqrt{3} \cdot \sqrt{2} \cdot V_{LN}}{\pi} \cdot I_{DC}$$

$$\frac{VA_{xfmr}}{P_{out}} = \frac{1.5834 \cdot \pi}{3 \cdot \sqrt{6}} = 0.677 \quad (6)$$

From equation (6), it is seen that the proposed transformer processes about 2/3rd the rated output power.

This conclusion matches the physical reasoning because the power processed by the non-phase shifted section that consists of the matching inductance L_{mtch}, is about 1/3rd the rated output power.

C. Key Differences between Autotransformer method and Proposed method

It is important to point out the salient differences between the autotransformer method of Fig. 2 and the proposed method shown in Fig. 3. They are as follows:

a. In autotransformer methods, the harmonic cancellation is primarily due to the reflected current waveform on to the branch carrying the main phase current. Due to non-ideal coupling and leakage effects, the cancellation is not complete and so the harmonic performance is not close to theoretical levels. In the proposed method, the harmonic cancellation is not dependent on the idealness of the magnetic coupling. It is achieved by current combination at the input junction point between I_1 and I_{H1} in Fig. 3.

b. Due to the non-ideal coupling, most autotransformer methods need to use large values of input inductor L_{IN} to comply with harmonic standards prevalent in the Industry, particularly IEEE 519-1992. The typical value used in such cases range from 0.05pu to 0.075pu. This adds cost and increases the overall size. Input inductors are associated with voltage drop and need to be compensated elsewhere. Due to the different way in which harmonic cancellation is achieved in the proposed method, the reliance on external inductor to achieve good performance is minimal. Typical values needed for the proposed method range from 0.015pu to 0.02pu to meet 5% THID requirements and no external inductor when 8% or higher THID is required to be met. This results in a more compact, less expensive, and a more efficient system.

c. Autotransformer methods rely heavily on stub windings. Stub windings are difficult to be practically implemented resulting in wastage of copper and poor utilization of the core. Manufacturing time is also longer due to the many complicated interconnections that need to be performed. In contrast, the proposed method relies on standard isolation transformer technique and is easy to manufacture with very few windings. Core and copper utilization is better, resulting in a more efficient and less expensive product.

d. In autotransformer methods, the transition of current from one diode pair to the next is very quick. In contrast, in the proposed method, the transition is observed to be much slower and is associated with a lower di/dt. This reduces the stress to the rectifier diodes, lowers power loss, and improves system reliability.

IV. EXPERIMENTAL VERIFICATION OF PROPOSED TOPOLOGY

The test results discussed in this section pertain to the use of the proposed hybrid 18-pulse converter to reduce the current harmonics at the input of a variable frequency drive with diode front end. The hybrid 18-pulse system tested was rated at 100hp. The VFD used was rated at 460V, 100hp. The value of L_{in} used is 0.11mH, which corresponds to 0.02pu on a 460V, 124A base. The relevant waveforms are shown in Figs. 7a and b. Harmonic data is given in Table 1.
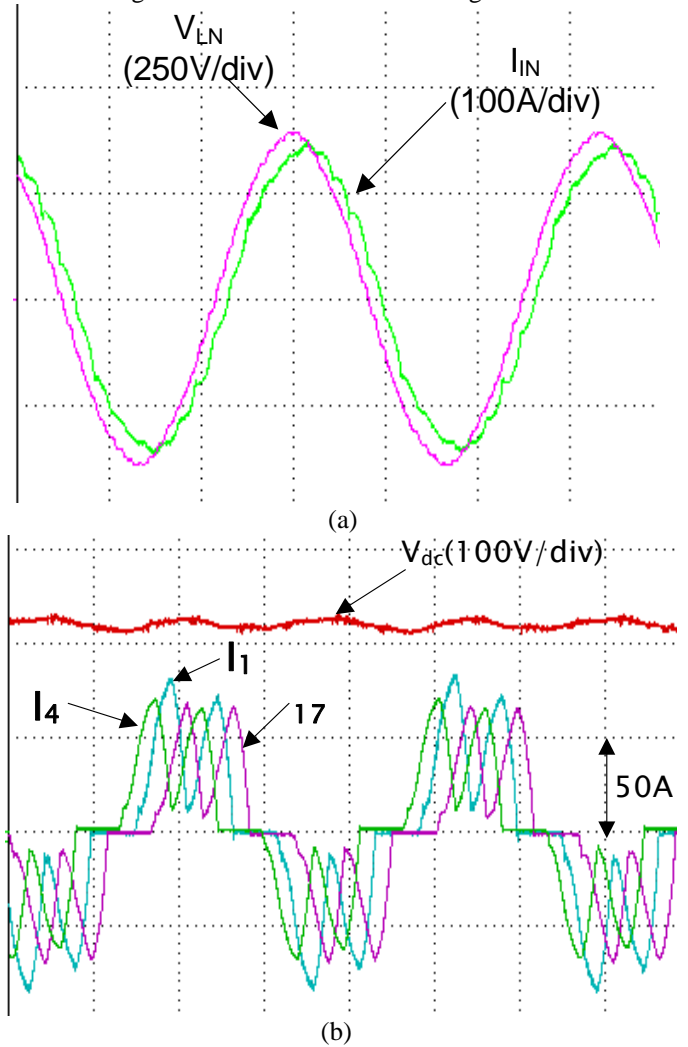

Fig. 7. (a) Line-neutral voltage and line current, (b) DC bus voltage and line current components for a 100hp, hybrid 18-pulse system with 0.02pu of input inductor, L_{IN}.

Relevant performance indices recorded at rated operating condition are as follows:

Input Power/phase:	24.9kW
I_4, I_7, I_1:	37.1A, 35.4A, 43.1A
True Power Factor:	0.96
Average DC bus voltage at no-load=	669.0V
Average DC bus voltage at rated load=	620.3V

Harm. Comp.	RMS	Fund	5th	7th	11th	13th	17th	19th	THD (%)
I_{in} (A)	95.8	95.7	2.1	0.6	0.4	0.7	2.3	1.5	4.5
V_{LN} (V)	270.4	270.3	3.4	2.0	0.8	0.3	0.6	0.3	1.7

Form the results presented in Fig. 7 and Table 1, the following important inferences can be made:

1. The total input current harmonic distortion of less than 5% at rated load condition is achievable using the proposed hybrid 18-pulse topology.
2. The true power factor of 0.96 is close to the displacement power factor of 0.99, which is indicative of low harmonic distortion.
3. The dc bus voltage reduces from 669V under no-load to 620V at rated load, indicating a drop of 7.3%, which is due to the impedance of the input transformer and 0.02pu input line inductor.
4. The current balance among the three different current paths is good, indicating that good matching is achievable. The current path through the L_{mtch} shows higher values of peak current compared to the other two paths and this can be improved to further reduce the total input current distortion. Current balance here refers to more of the wave shape rather than the rms value. Under light load condition, the currents can be discontinuous but still the balance could be good to yield low total current harmonic distortion.
5. Current cancellation technique is shown to be more effective than flux cancellation since the impedance through the direct path can be altered to improve balance and hence cancellation magnitude.

Tests were conducted with no input inductor to show that even in the absence of input inductor, L_{IN}, the performance is good due to good matching possible using the proposed technique. Results for the case with NO input inductor, L_{IN} are shown in Figs. 8a and b and the data is given in Table II.
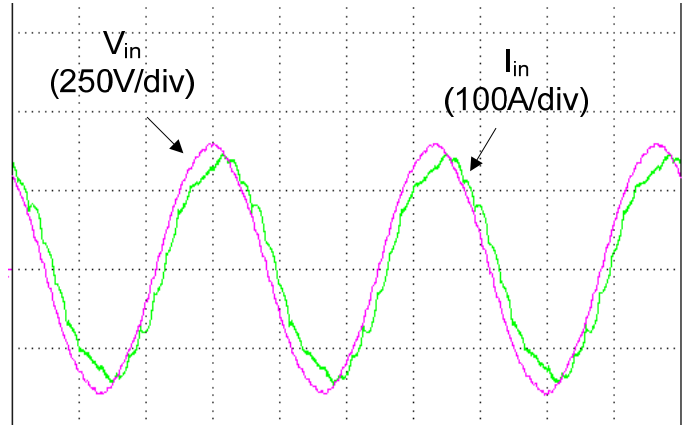
Relevant performance indices recorded at rated operating condition are as follows:

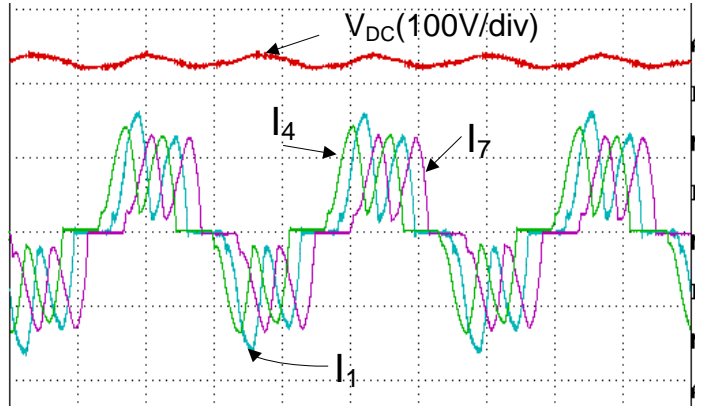|	
Input Power/phase:	24.53kW
I_4, I_7, I_1:	36.9A, 34.4A, 40.4A
True Power Factor:	0.96
Average DC bus voltage at no-load=	669.0V
Average DC bus voltage at rated load=	629.7V

Harm. Comp.	RMS	Fund	5th	7th	11th	13th	17th	19th	THD (%)
I_{in} (A)	93.8	93.6	3.1	0.9	0.8	0.8	2.6	1.6	6.3
V_{LN} (V)	272.0	272.0	3.4	1.8	0.7	0.2	0.8	0.3	1.8

(a)

(b)

Fig. 8. (a) Line-neutral voltage and line current, (b) DC bus voltage and line current components for a 100hp, hybrid 18-pulse system with NO L_{IN}.

Form the results presented in Fig. 7 and Table 1, the following important inferences can be made:

1. The total input current harmonic distortion of less than 8% at rated load condition is achievable using the proposed hybrid 18-pulse topology, without L_{IN}.
2. The true power factor of 0.96 is close to the displacement power factor of 0.99, which is indicative of low harmonic distortion.
3. The dc bus voltage reduces from 669V under no-load to 629.7V at rated load, indicating a drop of 5.9%, which is primarily due to the impedance of the input transformer.
4. The current balance among the three different current paths is good, indicating that good matching is achievable. Again, the current path through the L_{mtch} shows higher values of peak current compared to the other two paths and this can be improved to further reduce the total input current distortion.
5. The advantage of current cancellation technique over flux cancellation is reinforced from this test.

V. CONCLUSIONS

From the discussions thus far, the following important conclusions can be drawn:

a. The theoretically estimated harmonic distortion based on equation (4), which assumes no input inductor is 8.8%, while the measured distortion with no input inductor is 6.3%. The reason for this difference is the system impedance, which is not zero. Finite system impedance makes the input current smooth without any sharp edges. Also, the dc link inductor is finite and the dc bus current has 360Hz ripple and is not perfectly smooth. These non ideal characteristics present in a 3-phase AC system, in fact, aids in reducing the overall harmonic distortion compared to theoretical estimates.

b. The proposed topology is easy to manufacture, less complicated, and has superior performance. It has good copper and core utilization and is smaller in size and weight – For example, a 50hp zigzag transformer weighs 550lbs compared to 880lbs for a traditional transformer.

c. The hybrid 18-pulse topology proposed here achieves low harmonic distortion since the idea proposed here is rooted in current cancellation rather than flux cancellation as in isolation transformer or as in autotransformer techniques.

d. Due to ease of manufacturability and good core and copper utilization, the cost of the proposed technique is expected to be much lower than that of either the isolation transformer shown in Fig. 1 or the autotransformer technique shown in Fig. 2.

e. Existing autotransformer techniques need significant value of input inductance to improve the total current harmonic distortion. The topology of Fig. 2 needs L_{IN} of 0.075pu to achieve good performance, compared to only 0.02pu for the proposed hybrid 18-pulse topology to achieve similar performance.

REFERENCES

[1] *"IEEE Recommended Practices and Requirements for Harmonic Control in Electrical Power Systems"*, IEEE Std. 519-1992.

[2] Derek A. Paice, *"Power Electronic Converter Harmonics – Multipulse Methods for Clean Power"*, IEEE Press, 1996; ISBN 0-7803-1137-X.

[3] G. R. Kamath, D. Benson, and R. Wood, "A Novel Autotransformer based 18-pulse Rectifier Circuit", in *Applied Power Electronics Conference and Exposition"*, 2002.

[4] Mahesh Swamy, "Eighteen Pulse Rectification Scheme For Use with Variable Frequency Drives", USPTO Application: 11/650,811, 5 January 2007.

[5] Derek A. Paice, "Simplified Wye Connected 3-Phase to 9-Phase Auto-Transformer", U.S. Patent 6,525,951 B1, Feb. 25, 2003.

[6] Tsuneo Kume, "Multi-Pulse Rectifier Circuit," Japanese Patent P3591548, September 3, 2004.