PASSIVE TECHNIQUES FOR REDUCING INPUT CURRENT HARMONICS

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1. Introduction

Events over the last several years have focused attention on certain types of loads on the electrical system that result in power quality problems for the user and utility alike. Equipment which has become common place in most facilities including computer power supplies, solid state lighting ballast, adjustable speed drives (ASDs), and un-interruptible power supplies (UPSs) are examples of non-linear loads. Adjustable speed drives are also known as Variable Frequency Drives (VFDs) and are used extensively in the HVAC systems and in numerous industrial applications to control the speed and torque of electric motors. The number of VFDs and their power rating has increased significantly in the past decade. Hence, their contribution to the total electrical load of a power system is significant and cannot be neglected.

Non-linear loads are loads in which the current waveform does not have a linear relationship with the voltage waveform. In other words, if the input voltage to the load is sinusoidal and the current is non-sinusoidal then such loads will be classified as non-linear loads because of the non-linear relationship between voltage and current. Non-linear loads generate voltage and current harmonics, which can have adverse effects on equipment that are used to deliver electrical energy. Examples of power delivery equipment include power system transformers, feeders, circuit breakers, etc. Power delivery equipment is subject to higher heating losses due to harmonic currents consumed by non-linear loads. Harmonics can have a detrimental effect on emergency or standby power generators, telephones and other sensitive electrical equipment.

When reactive power compensation in the form of passive power factor improving capacitors are used with non-linear loads, resonance conditions can occur that may result in even higher levels of harmonic voltage and current distortion thereby causing equipment failure, disruption of power service, and fire hazards in extreme conditions.

The electrical environment has absorbed most of these problems in the past. However, the problem has now reached a magnitude where Europe, the US, and other countries have proposed standards to engineer systems responsibly, considering the electrical environment. IEEE 519-1992 and EN61000-3-2 have evolved to become a common requirement cited when specifying equipment on newly engineered projects. Various harmonic filtering techniques have been developed to meet these specifications. The present IEEE 519-1992 document establishes acceptable levels of harmonics (voltage and current) that can be introduced into the incoming feeders by commercial and industrial users. Where there may have been little cooperation previously from manufacturers to meet such specifications, the adoption of IEEE 519-1992 and other similar world standards now attract the attention of everyone.

2. Why VFDs Generate Harmonics?

The current waveform at the input of a VFD is not continuous. It has multiple zero crossings in one electrical cycle as shown in Fig 1. All voltage source inverter based VFDs have an AC to DC rectifier unit with a large DC capacitor to smooth the voltage ripple. The DC bus capacitor draws charging current only when it is discharged in to the motor load. The charging current flows into the capacitor when the input rectifier is forward biased, which occurs when the instantaneous input voltage is higher than the DC voltage across the DC bus capacitor. The pulsed current drawn by the DC bus capacitor is rich in harmonics because it is discontinuous.

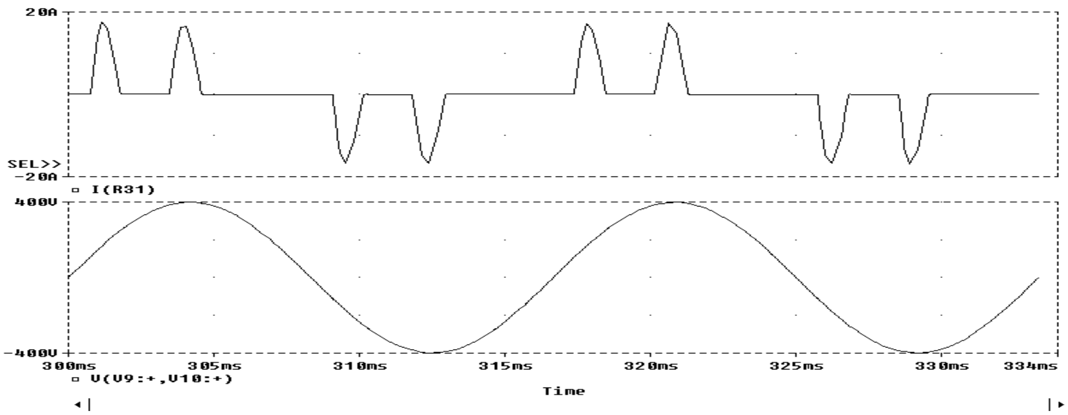

Fig. 1: Typical pulsed current waveform as seen at input of a VFD.

The voltage harmonics generated by VFDs are due to the flat-topping effect caused by a weak AC source charging the DC bus capacitor without any intervening impedance. The distorted voltage waveform gives rise to voltage harmonics, which is of more importance than current harmonics. The reason is simple. Voltage is shared by all loads and it affects all loads connected in an electrical system. Current distortion has a local effect and pertains to only that circuit that is feeding the non-linear load. Hence, connecting non-linear loads like VFDs to a weak AC system requires more careful consideration than otherwise.

The discontinuous, non-sinusoidal current waveform as shown in Fig.1 can be mathematically represented by sinusoidal patterns of different frequencies having a certain amplitude and phase relationship among each other. By adding these components, the original waveform can be reconstructed. The amplitude of the various sinusoidal components that need to be used to reconstruct a given non-sinusoidal waveform is expressed in terms of a mathematical expression called total harmonic distortion. The total harmonic current distortion (THD) is defined as: $THD_I = \frac{\sqrt{\sum_{n=2}^{n=\infty} I_n^{\ 2}}}{I_1}$; I_1 is the rms value of the fundamental component of current; and I_n is the rms value of the n^{th} harmonic component of current.

The reason for doing this is that it is easier to evaluate the heating effect caused by continuous sinusoidal waveforms of different frequencies and corresponding amplitudes than to estimate the heating effects caused by discontinuous non-sinusoidal waveforms.

The order of current harmonics produced by a semiconductor converter during normal operation is termed as Characteristic Harmonics. In a three-phase, six-pulse converter with no DC bus capacitor, the characteristic harmonics are non-triplen odd harmonics (e.g., 5th, 7th, 11th, etc.). In general, the characteristic harmonics generated by a semiconductor converter is given by:

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

h is the order of harmonics; k is any integer, and q is the pulse number of the semiconductor converter (six for a six-pulse converter). When operating a six-pulse rectifier-inverter system with a DC bus capacitor (Voltage Source Inverter or VSI), harmonics of orders other than those given by the above equation can be observed. Such harmonics are called non-characteristic harmonics. The reason for the existence of non-characteristic harmonics is due to imbalanced and/or distorted source voltage. In addition, equation (1) is based on a theoretical waveform that has a rectangular pattern made up of equal positive and negative halves, each occupying 120 electrical degrees. The pulsed discontinuous waveform observed commonly at the input of a VFD (Fig. 1) digresses greatly from the theoretical waveform. The non-characteristic harmonics are lower in magnitude but contribute to the overall harmonic distortion of the input current. The per unit value of the characteristic harmonics present in the theoretical current waveform at the input of the semiconductor converter is given by ***1/h*** where h is the order of the harmonics. In general, the observed per unit value of the harmonics is much greater than 1/h. The reasons are the same as mentioned above.

3. Harmonic Limit Calculations based on IEEE 519-1992

The IEEE 519-1992 relies strongly on the definition of the point of common coupling or PCC. The PCC from the power utility point of view will usually be the point where power comes into the establishment (i.e., point of metering). However, the IEEE 519-1992 document also suggests that, **"within an industrial plant, the point of common coupling (PCC) is the point between the non-linear load and other loads"** [1]. This suggestion is crucial since many plant managers and building supervisors feel that it is equally if not more important to keep the harmonic levels at or below acceptable guidelines within their facility. In view of the many recently reported problems associated with harmonics within industrial plants [2], it is important to recognize the need for mitigating harmonics at the point where the non-linear load is connected to the power system. This approach would minimize harmonic problems, thereby reducing costly downtime and improving the life of electrical equipment. If harmonic mitigation is accomplished for individual non-linear loads or a group of non-linear loads collectively, then the total harmonics at the point of the utility connection will, in most cases, meet or better the IEEE recommended guidelines. In view of this, it is becoming increasingly common for project engineers and consultants to require non-linear equipment suppliers to adopt the procedure outlined in IEEE 519-1992 to mitigate the harmonics to acceptable levels at the point of the offending equipment. For this to be interpreted equally by different suppliers, the intended PCC must be identified. If not defined clearly, many suppliers of non-linear loads would likely adopt the PCC to be at the utility metering point, which would not benefit the plant or the building but rather the utility.

Having established that it is beneficial to adopt the PCC to be the point where the non-linear load connects to the power system, the next step is to establish the short circuit ratio. Short circuit ratio calculations are key in establishing the allowable current harmonic distortion levels. For calculating the short circuit ratio, a determination of the available short circuit current at the input terminals of the non-linear load is required. If the short circuit value available at the low-voltage side of the utility transformer feeding the establishment (building) is known, and the cable impedance and other series impedances in the electrical circuit between the low-voltage side of the transformer and input to the non-linear load are known, then a calculation can be made for the available short circuit current at the non-linear load input. **In practice, it is common to assume that the short circuit current at the input to the non-linear load is the same as that at the low-voltage side of the utility transformer feeding the non-linear load**. The next step is to compute the fundamental value of the rated input current into the non-linear load. In case the non-linear load is a Variable Frequency Drive (VFD) operating an induction motor, the NEC amp rating for induction motor is used because NEC amps are fundamental amps that a motor draws when connected directly to the utility supply. An example is presented here to recap the above procedure.

A 100hp VFD/motor combination connected to a 480V system being fed from a 1500kVA, 3-Ph. transformer with an impedance of 4% is required to meet IEEE 519-1992 at the input terminals. The rated current of the transformer is: $1500*1000/(\sqrt{3}*480)$, which is calculated to be 1804A. In the absence of any more information, the short circuit current available at the low-voltage side of the utility transformer can be assumed not to exceed the rated current of the transformer divided by the per unit impedance of the transformer. For the above example, this is calculated to be: 45,105A. The NEC current rating for a 100hp, 460V induction motor is 124A. The short circuit ratio, which is defined as the ratio of the short circuit current at the PCC to the fundamental value of the non-linear current, is computed to be 364 (= 45,105A/124A). For the above calculations, it is assumed that the short circuit current at the VFD input is practically the same as that at the low-voltage side of the utility transformer. On referring to the IEEE 519-1992 Table 10.3 [1], the short circuit ratio falls in the 100-1000 category. For this ratio, the total demand distortion (TDD) at the point of VFD connection to the power system network is recommended to be 15% or less. For reference, Table 10.3 [1] is reproduced below.

Current Distortion Limits for General Distribution Systems

(120 V through 69,000 V)

	Maximum Harmonic Current Distortion in percent of I_L					
	Individual Harmonic Order (Odd Harmonics)					
I_{sc} / I_L	<11	11≤h≤17	17≤h≤23	23≤h≤35	35≤h	TDD
< 20*	4.0	2.0	1.5	0.6	0.3	**5.0**
20 < 50	7.0	3.5	2.5	1.0	0.5	**8.0**
50 < 100	10.0	4.5	4.0	1.5	0.7	**12.0**
100 < 1000	12.0	5.5	5.0	2.0	1.0	**15.0**
> 1000	15.0	7.0	6.0	2.5	1.4	**20.0**

Even harmonics are limited to 25% of the odd harmonic limits above.

* All power generation equipment is limited to these values of current distortion, regardless of actual I_{sc} / I_L **;** where I_{sc} is the maximum short circuit current at PCC and

I_L is the maximum demand load current (fundamental frequency) at PCC.

TDD is **Total Demand Distortion** and is defined as the harmonic current distortion in percent of maximum demand load current. The maximum demand current interval could be either a 15-minute or a 30-minute interval.

4. Harmonic Mitigating Techniques

Various techniques of improving the input current waveform are discussed below. The intent of all techniques is to make the input current more continuous to reduce the overall current harmonic distortion. The different techniques can be classified into three broad categories:

a. Passive techniques
b. Active techniques, and
c. Hybrid technique – combination of passive and active techniques

This document will concentrate only on the passive techniques. Each of the passive option will be briefly discussed with their relative advantages and disadvantages.

There are three different options in the passive configuration:

1. Addition of Inductive Impedance – line reactors and/or DC link chokes
2. Capacitor based harmonic filters – tuned as well as broad band type
3. Multi-pulse techniques (12-pulse, 18-pulse, etc)

4.1 Addition of Inductive Impedance

3-Phase Line Reactors

A line reactor makes the current waveform less discontinuous resulting in lower current harmonics. Since the reactor impedance increases with frequency, it offers larger impedance to the flow of higher order harmonic currents.

On knowing the input reactance value, the expected current harmonic distortion can be estimated. A table illustrating the expected input current harmonics for various amounts of input reactance is shown below:

Percent Harmonics vs Total Line Impedance

Harmonic	Total Input Impedance 3%	4%	5%	6%	7%	8%	9%	10%
5th	40	34	32	30	28	26	24	23
7th	16	13	12	11	10	9	8.3	7.5
11th	7.3	6.3	5.8	5.2	5	4.3	4.2	4
13th	4.9	4.2	3.9	3.6	3.3	3.15	3	2.8
17th	3	2.4	2.2	2.1	0.9	0.7	0.5	0.4
19th	2.2	2	0.8	0.7	0.4	0.3	0.25	0.2
%THID	44.13	37.31	34.96	32.65	30.35	28.04	25.92	24.68
True rms	1.09	1.07	1.06	1.05	1.05	1.04	1.03	1.03

Input reactance is determined by the series combination of impedance of the AC reactor, input transformer (building/plant incoming-feed transformer) and power cable. By adding all the inductive reactance upstream, the effective line impedance can be determined and the expected harmonic current distortion can be estimated from the above chart. The effective impedance value in % is based on the actual loading and is:

$$Z_{eff\ (pu)} = \frac{\sqrt{3} * 2 * \pi * f * L_T * I_{act\ (fnd.)} * I_{rated}}{V_{L-L}} * 100 \qquad (2)$$

$I_{act(fnd.)}$ is the fundamental value of the actual load current and $V_{L\text{-}L}$ is the line-line voltage. L_T is the total inductance of all reactance upstream and I_{rated} is the rated load current. The effective impedance of the transformer as seen from the non-linear load is:

$$Z_{eff,x-mer} = \frac{Z_{x-mer} * I_{act(fnd.)}}{I_r} \qquad (3)$$

$Z_{eff,x\text{-}mer}$ is the effective impedance of the transformer as viewed from the non-linear load end; $Z_{x\text{-}mer}$ is the nameplate impedance of the transformer; and I_r is the nameplate rated current of the transformer.

The reactor also electrically separates the DC bus voltage from the AC source so that the AC source is not clamped to the DC bus voltage during diode conduction. This feature reduces flat topping of the AC voltage waveform caused by many VFDs when operated with weak AC systems.

However, introducing AC inductance between the diode input terminal and the AC source causes voltage drop due to the series impedance offered by the inductor. The fundamental frequency voltage drop increases with load and results in lower available DC bus voltage. Hence, care should be taken not to add excessive amount of line inductance to prevent low voltage condition at the input terminals of the VFD.

AC inductor also causes overlap of conduction between outgoing diode and incoming diode in a three-phase diode rectifier system. The overlap phenomenon further contributes to the reduction of the average DC bus voltage. This reduction depends on the duration of the overlap in electrical degrees, which in turn depends on the value of the intervening inductance used and the current amplitude. The duration of overlap in electrical degrees is commonly represented by μ. In order to compute the effect quantitatively, a simple model can be assumed. Assume that the line comprises of inductance L in each phase. Let the DC load current be I_{dc} and the rms value of the line-line voltage at the input terminals of the rectifier (VFD) is $V_{L\text{-}L}$. It is assumed that the

DC load current does not change during the overlap interval. The current through the incoming diode at start is zero and by the end of the overlap interval, it is I_{dc}. Based on this assumption, the relationship between current and voltage can be expressed as:

$$v_{ab} = \sqrt{2} \cdot V_{L-L} \cdot \sin(\omega t) = 2 \cdot L \cdot (di / dt)$$
$$\sqrt{2} \cdot V_{L-L} \cdot \int_{(\pi/3)}^{(\pi/3)+\mu} \sin(\omega t)\, d(t) = 2 \cdot L \cdot \int_{0}^{I_{dc}} di \qquad (4)$$
$$I_{dc} = \frac{\sqrt{2} \cdot V_{L-L} \cdot (\cos(\pi/3) - \cos(\pi/3 + \mu))}{2\omega L} = \frac{\sqrt{2} \cdot V_{L-L} \cdot \sin(\pi/3 + \mu/2) \cdot \sin(\mu/2)}{\omega L}$$

For small values of overlap angle μ, $\sin(\mu/2)=\mu/2$ and $\sin(\pi/3+(\mu/2))=\sin(\pi/3)$. Rearranging the above equation yields:

$$\mu = \frac{2 \cdot \sqrt{2} \cdot \omega L \cdot I_{dc}}{V_{L-L} \cdot \sqrt{3}}; \qquad (5)$$

From the above expression, the following observations can be made:

1. If the inductance L in the form of either external inductance or leakage inductance of transformer or lead length is large, the overlap duration will be large.
2. If the load current, I_{dc} is large, the overlap duration is large.

The average output voltage will reduce due to the overlap angle as mentioned before. In order to compute the average output voltage with a certain overlap angle, the limits of integration have to be changed. This exercise yields the following:

$$V_O = \frac{3}{\pi} \int_{\mu+(\pi/3)}^{\mu+(2\pi/3)} \sqrt{2}\, V_{L-L} \sin(\omega t)\, d(\omega t)$$
$$V_O = \frac{3 \cdot \sqrt{2} \cdot V_{L-L} \cdot \cos(\mu)}{\pi} = \frac{3 \cdot \sqrt{2} \cdot \sqrt{3} \cdot V_{L-N} \cdot \cos(\mu)}{\pi} \qquad (6)$$

Thus, it can be seen that the overlap angle contributes to the reduction in the average value of the output DC bus voltage. Unfortunately, higher values of external inductive reactance, increases the overlap angle, which in turn reduces the average output voltage as seen from the above equation. In deriving the above expression, it is important to note that the line-line voltage V_{L-L} is the value at the input terminals of the rectifier and not that before the AC inductor. The fundamental voltage drop across the AC inductor should be subtracted from the actual value of the AC supply to arrive at the correct value of the voltage available at the input terminals of the rectifier.

DC Link Choke

Any inductor of adequate value placed in between the AC source and the DC bus capacitor of the VFD will help in making the input current waveform more continuous. Hence, a DC link choke, which is electrically present after the diode rectifier bridge and before the DC bus capacitor, can be used to reduce the input current harmonic distortion. The DC link choke appears to perform similar to the three-phase line inductance. However, on analyzing the behavior of the DC link choke, it can be seen that the DC link choke behaves similar to the input AC line inductor only from current distortion point of view but has a completely different influence on the average output voltage.

An important difference is that the output of a three-phase rectifier has significant DC component and so there is no voltage drop similar to that observed in three-phase AC line inductors. In addition, the DC link choke is after the diode rectifier block and will not contribute to the overlap phenomenon discussed earlier with regards to external AC input reactors. Hence, unlike the case with AC input reactors, there is no DC bus voltage reduction when DC link choke is used. The DC link choke increases the diode conduction duration. There is a critical DC link choke inductance value, which results in complete 60-degree conduction of a diode pair. Any value of DC link inductance beyond this critical value is of no further importance and thus introducing a very large DC link inductor will have marginal benefit. A larger inductance value will only help in reducing the DC bus current ripple but will be associated with an extra voltage drop due to the higher winding resistance. This can also result in somewhat higher power loss without altering the average output DC voltage significantly. The critical DC link choke that is needed to achieve complete 60-degree conduction is derived next.

It is assumed that the source is ideal with zero impedance. The forward voltage drop across the conducting diodes is also neglected. When there is no DC link current, the DC bus charges up to the peak of the input AC line. The average DC bus voltage remains at the peak of the input AC line-line voltage, neglecting voltage drop across the diode and assuming an ideal system. The critical inductance is that value that will result in the average DC bus voltage to drop from its peak value to the average 3-ph. rectified value, under rated load conditions. This occurs when each diode pair conducts for 60^{o} duration under rated load condition. The voltage difference from the peak line-line voltage under no-load condition to 3-ph. average rectified value under rated load condition appears across the DC link inductor. The inductance needed to achieve this is obtained as follows:

$$\begin{aligned} L_{cr}\cdot\frac{\Delta i}{\Delta t} &= V_m - V_{3-ph-avg} = V_m - \frac{3\cdot V_m}{\pi} \\ L_{cr} &= \frac{\pi-3}{\pi}\cdot V_m\cdot\frac{\Delta t}{\Delta i} = \frac{\pi-3}{\pi}\cdot V_m\cdot\frac{T/6}{I_{dc}} \end{aligned} \qquad (7)$$

L_{cr} is the critical value of the DC link choke and I_{dc} is the load current. The change in current Δi in equation (7) is the difference from no-load condition to rated load condition. Hence, Δi is the rated average DC link current that flows continuous for a 60-degree conduction interval. T is the period of the input AC supply. In equation (7), it should be pointed out that if continuous current conduction for 60-degree duration is desired at a lower value of DC load current, a DC link inductor of a large value is required.

From the expression for the critical DC link inductance, it is seen that the value depends on the load condition, frequency of the input AC supply and the peak value of the input AC line-line voltage, V_m. It is also important to note that the value of the critical DC link inductor for a 240V system for the same load is 1/4th the value for a 480V system.

4.2 Capacitor based Passive Filters

Passive filters consist of passive components like inductors, capacitors, and resistors arranged in a pre-determined fashion either to attenuate the flow of harmonic components through them or to shunt the harmonic component into them. Passive filters can be of many types. Some popular ones are: Series Passive filters, Shunt Passive filters, and Low-pass broadband Passive filters. Series and Shunt passive filters are effective only in a narrow proximity of their respective tuned resonant frequency. Low pass broadband passive filters have a broader bandwidth and attenuate a larger range of harmonics above the cutoff frequency.

Series Passive Filter

One way to mitigate harmonics generated by non-linear loads is to introduce a series passive filter (Fig. 2) in the incoming power line so that the filter offers high impedance to the flow of harmonics from the source to the non-linear load. Since the series passive filter is tuned to a particular frequency, it offers high impedance at only its tuned frequency. Depending on the physical property of L and C chosen, typically there exists a narrow band around the tuned frequency where the impedance remains high.

Series passive filters have been used more often in 1-ph. applications where it is effective in attenuating the 3rd-harmonic component. Such filters are generally designed to offer low impedance at the fundamental frequency. A major drawback of this approach is that the filter components have to be designed to handle the rated load current. Further, one filter section is not adequate to attenuate the entire harmonic spectrum present in the input current of a non-linear system. Multiple sections may be needed to achieve acceptable results, which makes them bulky and expensive.

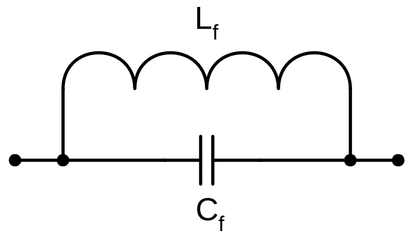

Fig. 2. Single-phase representation of a series filter configuration.

Shunt Passive Filter

The second and more common approach is to use a shunt passive filter, as shown in Fig. 3. The shunt passive filter is placed across the incoming line and is designed to offer very low impedance to current components corresponding to its tuned frequency. Another way of explaining the behavior of a shunt filter is to consider the energy flow from source to the non-linear load via the shunt filter. Energy at fundamental frequency flows into the shunt passive filter and the energy at the filter's tuned frequency flows out of the shunt filter since it offers lower impedance for flow of energy at its tuned frequency compared to the source. In other words, the harmonic component needed by the non-linear load is provided by the shunt filter rather than the AC source.

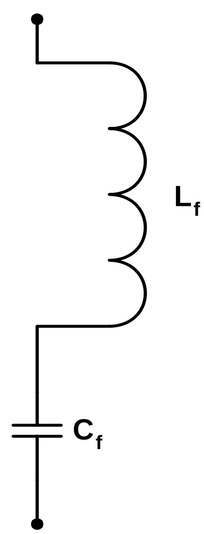

Fig. 3. Single-phase representation of a shunt-tuned filter configuration.

The fundamental frequency energy component flowing into the shunt filter is the reason for leading VARs and can cause over-voltage at the filter terminals. This can create problems with VFDs that are vulnerable to higher than normal voltage and under light-load condition can encounter over-voltage trips. Similar to the series tuned filter, the shunt-tuned filter is effective only at and around its tuned frequency and only one section of the filter alone is inadequate to provide for all the harmonic energy needed by a typical non-linear load (VFD). Multiple sections are needed, which makes them bulky and expensive.

The commonly used 3-ph. shunt filter sections comprise of individual sections tuned to the 5th, the 7th, and perhaps a high-pass section typically tuned near the 11th harmonic. Unfortunately, if care is not taken, the shunt filter will try to provide the harmonic energy needed by all non-linear loads connected across its terminals. In this process, it can be overloaded and be damaged if unprotected. In order to avoid import of harmonics, it is important to use AC line inductors in series with the shunt branches of tuned filters. The series impedance will impede the harmonic energy flow from other sources into the shunt tuned filter sections, as shown in Fig. 4. However, the addition of extra line inductance increases the size and cost of the filter section. The series inductance may also aggravate the over-voltage condition experienced by inverter drive systems at light-load conditions. The over voltage tolerance margin would be compromised and the VFD could be more vulnerable to fault out on over voltage resulting in nuisance trips. The probability of this occurrence increases under light-load conditions and especially if the installation happens to be close to a capacitor-switching utility substation.

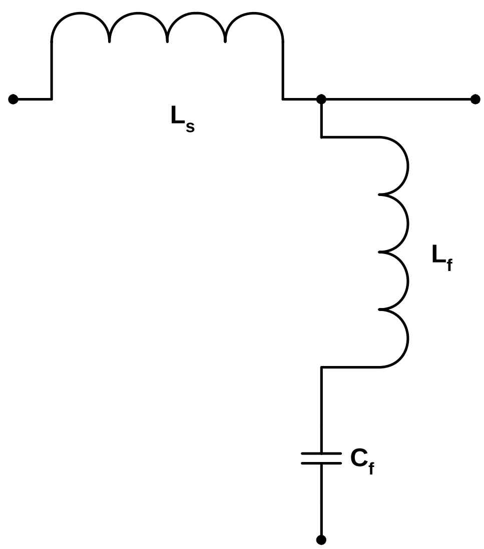

Fig. 4. Shunt tuned filter section with a series inductor to prevent import of harmonics.

Low Pass Broad Band Filter

The low-pass broadband filter is similar to the circuit configuration of Fig. 4. To improve the filtering performance, the inductor L_s is substituted by L_f and there is no series inductor with C_f. By removing L_f from the shunt path, the filter configuration changes from tuned type to broadband type. One advantage of the low pass, broadband harmonic filter is that unlike the shunt and series type filters, the broadband filter need not be configured in multiple stages or sections to offer wide spectrum filtering. In other words, one filter section achieves the performance close to the combined effect of a 5th, 7th, and a high-pass shunt tuned filter section. A typical broadband filter section is shown in Fig. 5. The series inductor L_f offers high impedance to limit import and export of harmonics from and to other non-linear loads on the system [2].

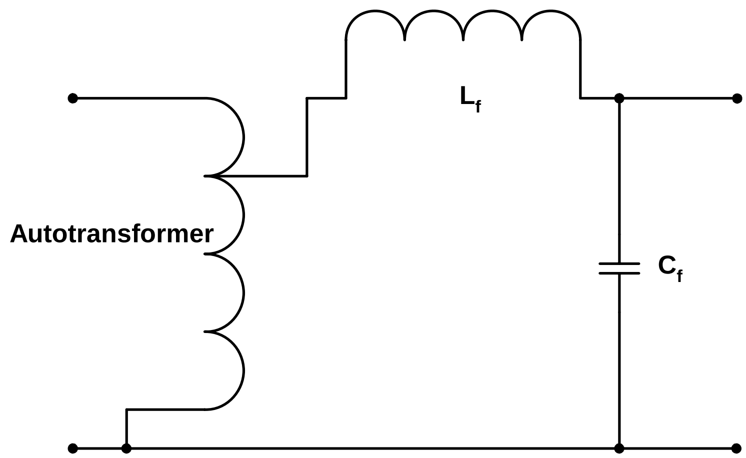

Fig. 5. Broadband filter section with autotransformer.

However, by removing L_f from the shunt branch and moving it to the series branch, aggravates the over-voltage problem experienced by VFDs. Autotransformers have been used in the past to address this problem. Since the over-voltage is a function of the load current, the correction offered by autotransformer works only at one operating point at best and is inadequate to handle wide range of operating conditions. In addition, the size and cost of the total filter configuration becomes high and less appealing. The leading VAR problem is not resolved and in fact has been found to interfere with power measurement and monitoring systems. These unfavorable features are serious enough to limit use of such filters for VFD applications.

4.3 Multi-pulse Techniques

As discussed in section 2, the characteristic harmonics generated by a semiconductor converter is a function of the pulse number for that converter. The higher the pulse number, the lower is the total harmonic distortion since the order of the characteristic harmonics shifts to a higher value. The pulse number is defined as the number of diode-pair conduction intervals that occur in one electrical cycle. In a 3-ph., six-diode bridge rectifier, the number of diode-pair conduction intervals is six and thus is known as a six-pulse rectifier. By using multiple six-pulse diode rectifiers in parallel and phase shifting the input voltage to each rectifier bridge by a suitable value, multi-pulse operation can be achieved.

4.3.1 12-Pulse Techniques

A 12-pulse rectifier operation can be achieved by using two six-pulse rectifiers in parallel with one rectifier fed from a power source that is phase shifted with respect to the other rectifier by 30 electrical degrees. The 12-pulse rectifier will have the lowest harmonic order of 11. In other words, the 5th, and the 7th harmonic orders are theoretically non-existent in a 12-pulse converter. Again, as mentioned in section 2, the amplitude of the characteristic harmonic is typically proportional to the inverse of the harmonic order. In other words, the amplitude of the 11th harmonic in a 12-pulse system will be 1/11 of the fundamental component and the amplitude of the 13th harmonic will be 1/13 of the fundamental component. There are many different ways of achieving the necessary phase-shift to realize 12-pulse operation. Some popular methods are:

a. Three winding isolation transformer;

b. Pseudo 12-pulse method;

c. Autotransformer method.

a. Three winding isolation transformer method

A three winding isolation transformer has three different sets of windings. One set of windings is typically called the primary, while the other two sets are called secondary windings. The primary windings can be connected in a delta or in a wye configuration. One set of secondary windings is connected in a delta while the other set is connected in a wye configuration. This arrangement automatically yields a 30-degree phase-shift between the two sets of secondary windings. A traditional 12-pulse arrangement using a three winding isolation transformer is shown in Fig. 6. The realization of 12-pulse operation in the circuit of Fig 6 is discussed next.

The current flowing out of the secondary windings, viewed independently, is similar to that observed in a six-pulse rectifier. However, since the voltages are phase-shifted by 30 electrical degrees, the currents are also phase-shifted by the same amount. In other words, if i_1 is the fundamental current through one set of secondary windings, and i_2 is the 30-degree phase-shifted current in the other set of secondary windings, then i_1 and i_2 can be expressed as follows:

$$i_1 = I_m \cdot \sin(\omega t); \qquad (8)$$
$$i_2 = I_m \cdot \sin(\omega t - \frac{\pi}{6});$$

The 5th harmonic component of the current in one of the secondary windings will be phase shifted with respect to its corresponding phase in the other set. However, it should be noted that the phase shift will get multiplied by the harmonic number as well. The 5th harmonic component in the two sets of windings can be represented as follows:

$$i_{5(1)} = I_{5m} \cdot \sin(5 \cdot \omega t) \qquad (9)$$
$$i_{5(2)} = I_{5m} \cdot \sin(5 \cdot \omega t - \frac{5 \cdot \pi}{6} - \frac{\pi}{6}) = -I_{5m} \cdot \sin(5 \cdot \omega t)$$

Similar expressions can be written for the 7th harmonic currents in each set of the secondary windings. From the above expressions, it can be said that the flux pattern formed by the 5th and 7th harmonic components by one set of secondary windings are theoretically equal and opposite to the 5th and 7th harmonic flux components produced by the second set of secondary windings. Consequently, there is no 5th and 7th harmonic component reflected on to the primary windings and so the 5th and 7th harmonic components do not theoretically exist in the input AC supply feeding the primary windings.

Based on the above explanation, it can be said that in a three-winding isolation transformer arrangement, magnetic flux coupling plays an important role in assuring the elimination of low order current harmonics. Any departure from the ideal scenario assumed above will yield sub-optimal flux cancellation and higher total current harmonic distortion. Leakage flux and the primary magnetizing flux create non-ideal conditions and are responsible for the existence of non-characteristic harmonics in the input current of a typical 12-pulse system. Minor winding imbalance between the two sets of secondary windings also contributes to sub-optimal performance.

Advantages

Some important advantages of the three winding isolation transformer configuration to achieve 12-pulse operation is listed below:

- 12-pulse operation yields low total current harmonic distortion,
- Three winding arrangement yields isolation from the input AC source, which has been seen to offer high impedance to conducted EMI.
- It offers in-built impedance due to leakage inductance of transformer. This smoothes the input current and helps further reduce the total current harmonic distortion.
- It is ideally suited for voltage level translation. If the input is at a high voltage (3.2kV or 4.16kV), and the drive system is

rated for 480V operation, this is ideal to step-down and to achieve the benefits of 12-pulse operation.

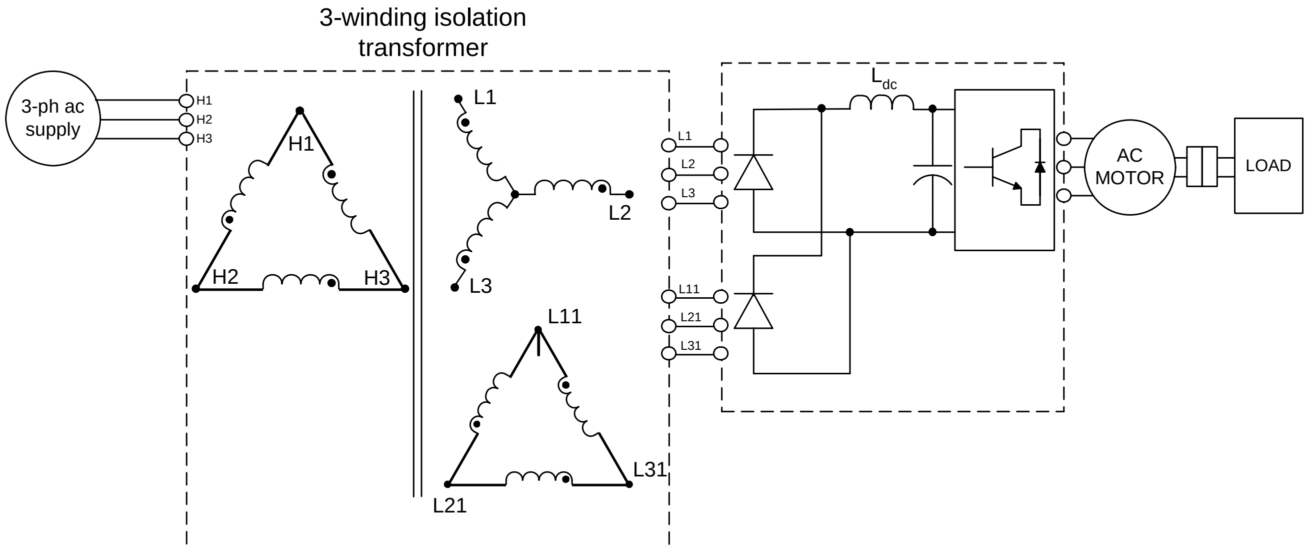

Fig. 6. Typical schematic of a 12-pulse configuration using a 3-winding isolation transformer. DC link choke improves performance.

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Disadvantages

In spite of its appeal, the three winding isolation transformer configuration has a few shortcomings listed below:

- The three winding transformer has to be rated for full power operation, which makes it bulky and expensive.
- Leakage inductance of the transformer will cause reduction in the DC bus voltage, which will require the use of taps in the primary windings to compensate for this drop. Addition of taps will increase cost.
- Due to minor winding mismatch, leakage flux, and non-trivial magnetizing current, the total current harmonic distortion can be higher than expected.
- This scheme requires the VFD to be equipped with two six-pulse rectifiers, which increases the cost of the VFD.

b. Pseudo 12-pulse method

One disadvantage of the three winding arrangement mentioned earlier is its size and cost. On re-examining the circuit of Fig 6, 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 directly connected to the AC source via some balancing inductance to match the inductance in front of the other six-pulse rectifier circuit to achieve 12-pulse operation.

The resulting scheme has one six-pulse rectifier powered via a phase-shifting isolation transformer, while the other six-pulse rectifier is fed directly from the AC source via matching impedance. Such a 12-pulse arrangement will be referred to in this document as pseudo 12-pulse configuration and is shown in Fig. 7. Another name for the same scheme is a hybrid 12-pulse configuration. The phase-shifting transformer feeding one of the two six-pulse rectifiers is sized to handle half the rated power. Similarly, the matching inductor is sized to carry only half the rated current. This arrangement results in the overall size of the transformer and matching inductor combination to be smaller and less expensive than the three winding arrangement.

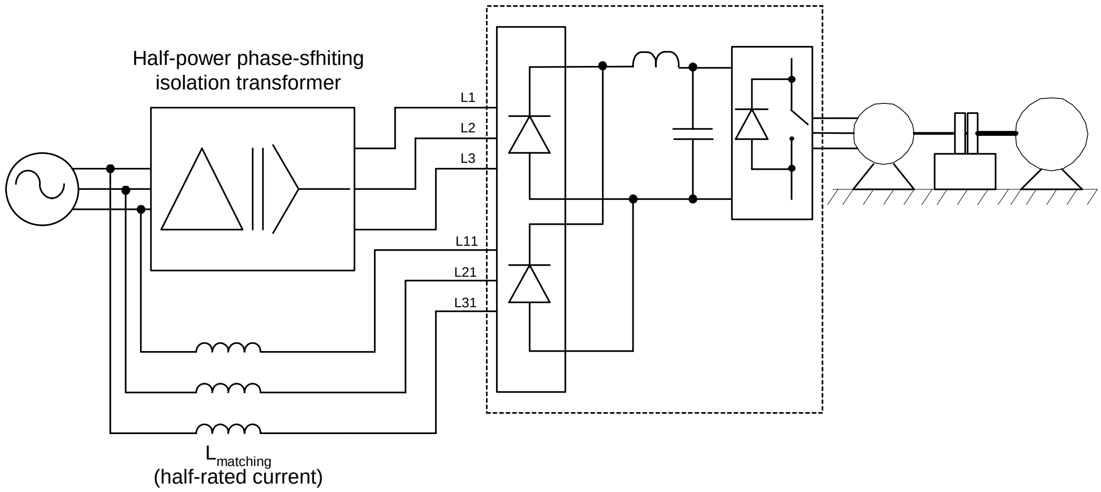

Fig. 7. Schematic of a pseudo 12-pulse arrangement.

Advantages

Some important advantages of the pseudo 12-pulse is listed below:

- Size and cost of the pseudo 12-pulse configuration is much less than the 3-winding arrangement.
- 12-pulse operation is achieved with low total current harmonic distortion.
- Unlike 3-winding method, in this method the current (instead of flux in the core) in the two bridges are combined at the source to cancel the low order harmonics. Leakage flux and winding mismatch problems do not occur.

Disadvantages

The pseudo 12-pulse method also has some important disadvantages that need to be pointed out.

- The impedance mismatch between the leakage inductance and the external matching inductance can never be accomplished for all operating conditions because the leakage inductance is a function of current through the transformer while the external inductance is in the form of self inductance, which is constant till its rated current value,
- In order to minimize the effect of mismatch, an input AC line inductor may need to be used sometimes to comply with the harmonic levels recommended in IEEE 519(1992).
- Use of extra inductance ahead of the transformer-inductor combination can cause extra voltage drop that cannot be compensated for.
- The arrangement of Fig 7 cannot be used where voltage level translation is needed.
- The advantage of high impedance to conducted EMI as offered by the three winding arrangement is lost on using the pseudo 12-pulse arrangement of Fig. 7.
- Similar to the three winding configuration, this method also requires the VFD to have two six-pulse rectifiers.

It should be pointed out that in spite of the shortcomings listed above, this method is gaining in popularity primarily because of size and cost advantage. The transformer leakage inductance and the external matching inductance are matched to perform at rated current so that low harmonic distortion is achieved at rated operating conditions. By specifying a maximum imbalance to be less than 5% at rated current operation, low current harmonic distortion is achievable.

c. Autotransformer method

The phase-shift necessary to achieve multi-pulse operation can also be achieved by using autotransformers. Autotransformers do not provide any isolation between the input and output but can be used to provide phase shift. Autotransformers are typically smaller compared to regular isolation transformers because they do not need to process the entire power. Majority of the load current passes directly from the primary to the secondary terminals and only a small amount of VA is necessary for the phase-shift processed by the autotransformer. This makes them small, inexpensive, and attractive for use in multi-pulse systems.

Though autotransformers are appealing for multi-pulse applications, they are not well suited for single VFD load. In all AC to DC rectification schemes, the diode-pair that has the highest voltage across the input terminals conducts in order to charge the DC bus. When parallel rectifiers are used as in multi-pulse techniques, it is important to maintain sharing of current among the multi-pulse rectifiers. If current sharing is compromised, then the amplitudes of lower order harmonics between the two rectifiers in a 12-pulse scheme will not cancel completely and this will result in poor harmonic performance. By electrically isolating one rectifier from the other either by using three-winding isolation transformer or by using half-power isolation transformer, in the two schemes discussed earlier, acceptable 12-pulse performance was possible. However, when autotransformers are employed, such isolation is lost and current from one set of phase-shifted windings can flow into the other set, thereby compromising the equal distribution of current between the phase-shifted sets of windings. One way to force the rectifiers to share correctly is to introduce an inter-phase transformer (IPT) in between the outputs of the two diode-rectifier units as shown in Fig. 8. A zero-sequence blocking transformer (ZSBT) in between the rectifier and one of the phase-shifted outputs of the autotransformer also helps in reducing non-characteristics triplen harmonics from flowing into the AC system. The autotransformer of Fig. 8 has phase-shifted outputs of $\pm 15^{o}$.

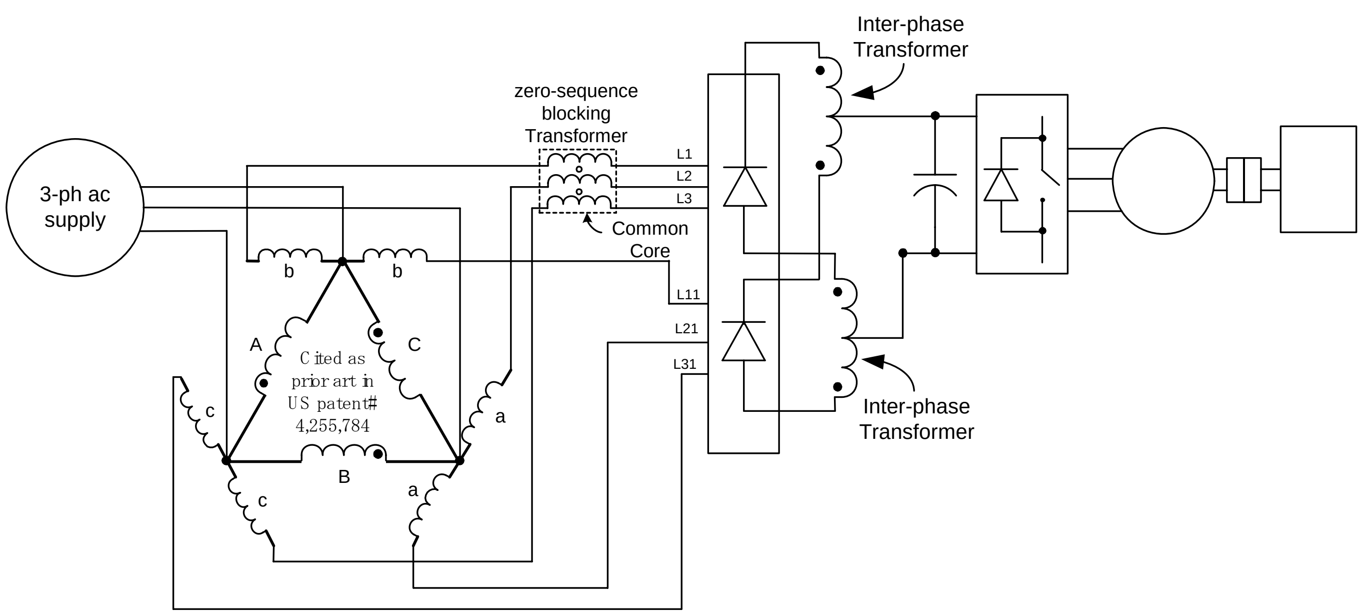

Fig. 8. Delta-fork autotransformer with ZSBT and IPT for 12-pulse applications.

The use of ZSBT and IPT makes the overall system bulky and expensive and the choice of autotransformer less appealing. In many cases, the VFD is not equipped with two rectifier units and so none of the 12-pulse schemes can be used. In such applications, if multiple VFDs are being employed and can be paired into approximately equal ratings then the delta-fork autotransformer shown in Fig. 8 can be effectively implemented. Instead of isolating the two diode rectifier units in one VFD, it is possible to use two different VFDs operating two independent loads of approximately equal rating and supplying them power from the phase-shifted outputs of the delta-fork transformer. This type of matched pair possibilities exist in a given system and is ideal for VFDs that do not have two independent six-pulse rectifier units. One such scheme of distributing the load between the phase-shifted outputs of a delta-fork autotransformer is shown in Fig. 9. This arrangement has been seen in the field to achieve low total current harmonic distortion even with load imbalance in the neighborhood of 20 to 25%.

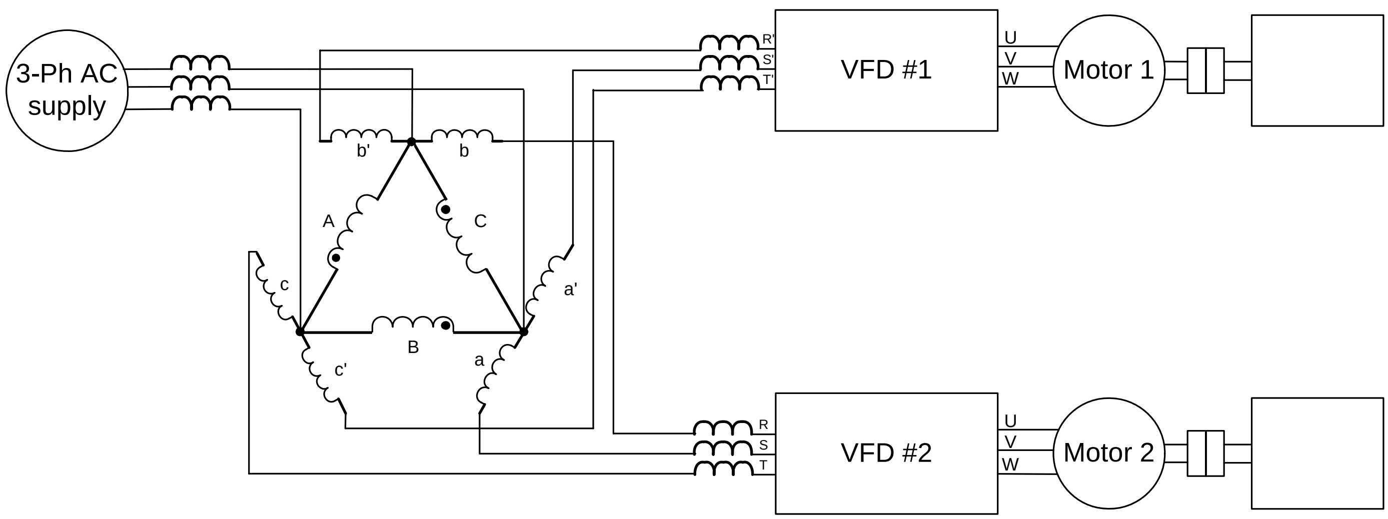

Fig. 9. Use of low cost autotransformer for 12-pulse operation in case of isolated and balanced loads.

Advantages

Some important advantages of the autotransformer connection shown in Fig 9 is listed below:

- VFDs do not need to have multiple rectifier units to achieve benefits of 12-pulse operation.
- Size and cost of autotransformer is less and unlike the circuit of Fig 8, there is no need for IPTs and ZSBTs.
- The 3-phase input AC reactor in front of each VFD helps in making the current more continuous. These may be replaced by DC link chokes.

Disadvantages

The circuit of Fig 9 has a few shortcomings and the reader should be aware of these. They are:

- Better harmonic performance is achieved if the loads are balanced. Since the loads are independent, many times it is not possible to guarantee balance and this may reduce the overall harmonic performance.
- Input AC line inductors or DC link chokes may be necessary to get better harmonic performance.
- VFDs need to be isolated to prevent cross current flow between the two sets of windings and to assure good sharing.

4.3.2. 18-Pulse Techniques

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. Use of large power VFDs increases the amplitude of low order harmonics that can impact the power system significantly. 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). In view of this, lately, the 18-pulse VFD systems are being proposed to achieve much superior harmonic performance compared to the traditional 12-pulse systems.

The 18-pulse systems have become economically feasible due to the recent advances in autotransformer techniques that help reduce the overall size and cost and achieve low total current harmonic distortion. As mentioned earlier, when employing autotransformers, care should be taken to force the different rectifier units to share the current properly. The 18-pulse configuration lends itself better in achieving this goal compared to the 12-pulse scheme. Some popular 18-pulse autotransformer techniques are discussed next.

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. 10.

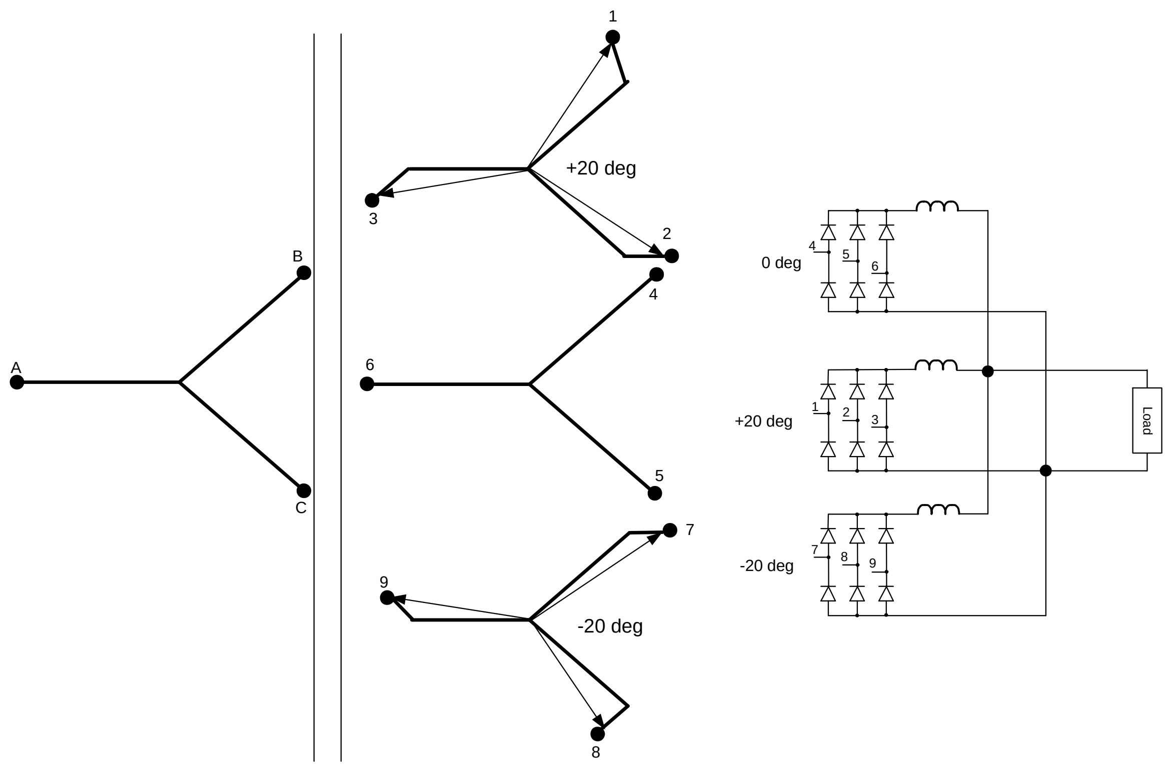

Fig. 10. Schematic representation of 18-pulse converter circuit fed from phase shifted isolation transformer.

The use of DC link choke as shown in Fig. 10 is optional. The leakage inductance of the transformer may be sufficient to smooth the input current and improve the overall current harmonic distortion levels.

The primary disadvantage of the scheme shown in Fig 10 is that the phase-shifting isolation transformer is bulky and expensive. A common disadvantage with all 18-pulse schemes is that all of them need three independent 3-phase rectifier units. Many VFD manufacturers do not provide this feature and the additional rectifier units needed may have to be provided external to the VFD.

Instead of using ± 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. Some patents that propose this scheme [3] are shown in Fig. 11.

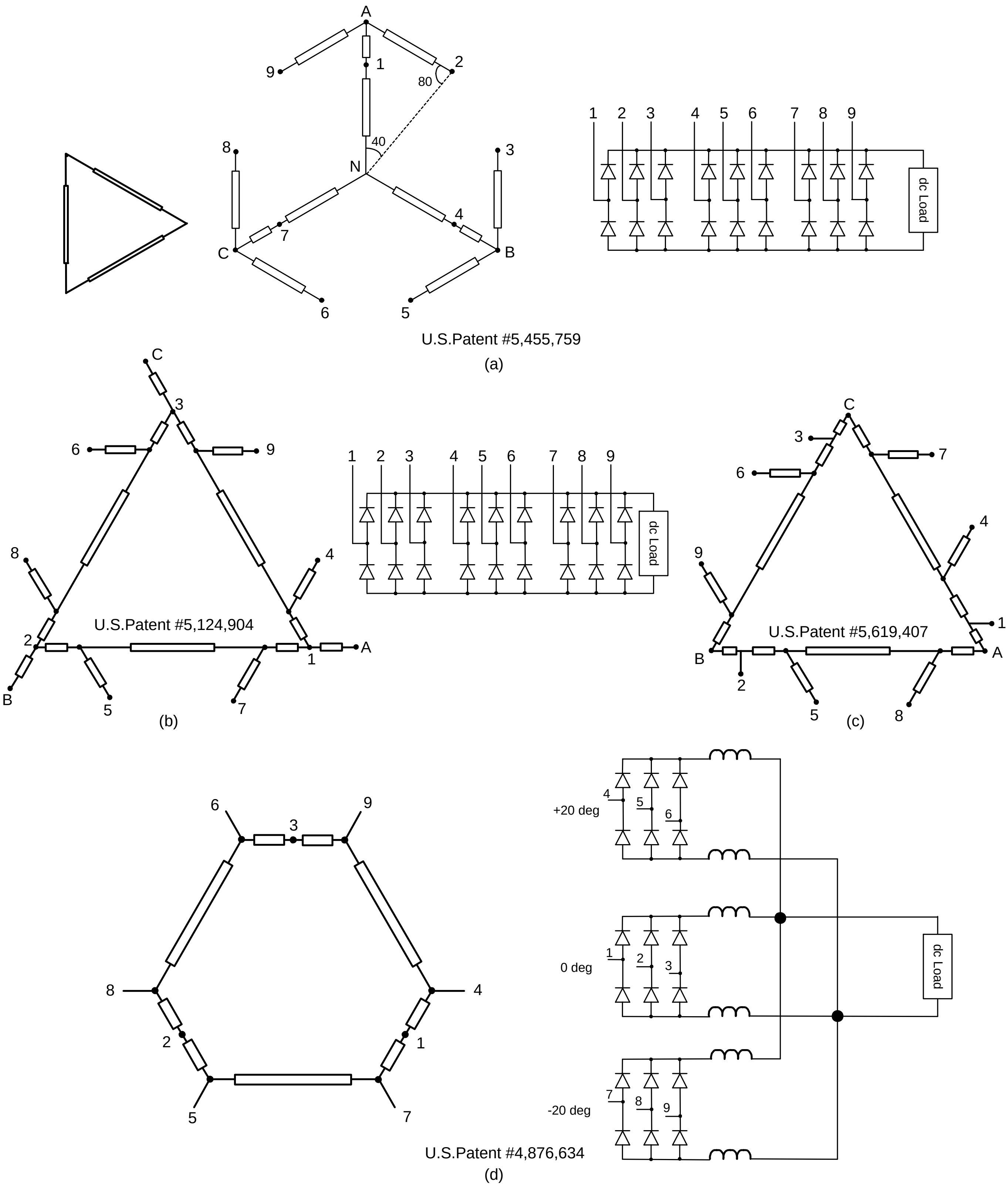

Fig. 11. Autotransformer methods of achieving 18-pulse operation.

Fig 11 (a) shows a nine-phase AC supply using wye-fork with a tertiary delta winding to circulate triplen harmonics. The size of the autotransformer is big and there is need for additional series impedance to smoothen the input AC currents. The rating of the transformer is about 70% of the rating of the load. If the series inductance is not used, then the output DC voltage is about 4.3% higher than that achieved when a standard six-pulse rectifier is used.

Fig 11(b) shows a nine-phase AC supply using delta-fork that does not require additional delta winding. In this configuration, the average DC output voltage is about 14% higher than that obtained using a standard six-pulse rectifier scheme. This can potentially stress the DC bus capacitors and the IGBTs in the inverter section of a VFD. In order to overcome this, additional teaser windings are used as shown. These windings not only add cost and increase the overall rating of the transformer, but also cause imbalance that results in higher than normal circulating currents in the delta windings, which need to be accommodated. The harmonic performance is good but the overall size is large with rated current flow through the teaser windings.

In order to overcome the 14% higher average DC bus voltage observed in the previous configuration, a modification of the configuration was proposed in the patent cited in Fig 11(c). The harmonic performance is equally good and the average DC bus voltage is equal to that observed in six-pulse rectifiers. Similar to the previous configuration, the stub winding currents are high and the teaser winding needs to carry rated load current making the overall transformer big in size and expensive to wind.

In autotransformer configurations using stub and/or teaser windings shown in Figs 11(a) through (c), the overall size and rating of the autotransformer is higher than the optimal value. Use of stub windings typically results in poor utilization of the core and involves more labor to wind the coils. Polygon type of autotransformer is better than stub type autotransformer from size and core utilization points of view. A polygon type autotransformer is shown in Fig 11(d). It should be pointed out that the configuration of Fig 11(d) needs the use of inter-phase transformers and input AC inductors to achieve low total current harmonic distortion. The reason is that the outputs are not equally spaced to achieve a nine-phase AC supply as in the previous configurations. The polygon autotransformer of Fig 11(d) provides +/-20° phase shifted outputs to achieve 18-pulse operation.

One of the most popular 18-pulse autotransformer configurations is that shown in Fig 12. This configuration is a modified version of the configuration shown in Fig 11(a) and was proposed by the same author. In the configuration of Fig 12, the delta-connected tertiary winding is included in the wye fork. This construction is called the windmill construction. Initially the windmill structure was present in each phase and the size of the transformer was still big. The kVA rating was about 60%. By intelligently removing the windmill structure from two of the three phases, it was shown that the performance remained equally good. By adopting the modified structure of Fig 12, the kVA rating of the autotransformer was reduced from 60% to 55%.

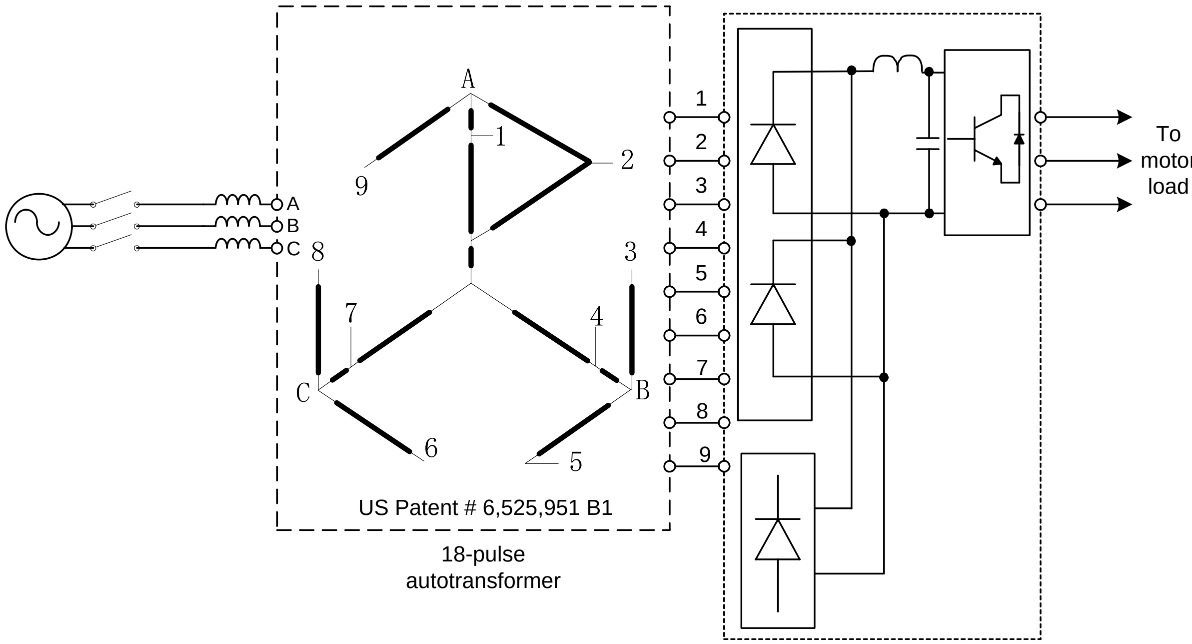

Fig. 12. Schematic of the modified windmill construction of the 18-pulse autotransformer configuration for used with VFDs.

In all the 18-pulse autotransformer methods, the change of current from one conducting diode pair to the other is quite sudden. 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. Since the use of autotransformer method of 18-pulse operation is recent, there is not much statistical data available to comment on the di/dt issue with diodes when used in conjunction with 18-pulse autotransformer techniques.

5. Conclusion

This document discusses the generation of current harmonics by non-linear loads and the IEEE-519-1992 standard to limit the quantity of these harmonics. A methodology of applying this standard to a practical industrial site has been discussed. Different harmonic mitigating techniques presently available in the industry have been highlighted. Multi-pulse techniques to achieve low total current harmonic distortion have been discussed. Relative advantage and disadvantage of the techniques presented have also been discussed. Based on the materials presented in this report, the following important conclusions can be drawn:

a. Passive techniques involving capacitors are associated with circulating current, leading power factor, and high DC bus voltage at light load condition and hence should be avoided as far as possible. Capacitor based filters are also associated with the possibility of causing network resonance and hence if installed, care should be taken to monitor and avoid any resonance conditions.

b. DC link choke is a better alternative than a 3-ph. AC line reactor for harmonic mitigation since it does not cause additional voltage drop. The DC bus voltage does not go below the standard 3-ph. average rectified value for DC link chokes greater than or equal to the critical inductance.

c. To handle transients and surges on the AC line, a combination of small value of AC inductance and DC link choke is preferred,

d. Multi-pulse techniques offer the best passive solution to handle harmonics. The 18-pulse autotransformer technique and the pseudo 12-pulse technique are attractive for medium power applications, while distributing the load on phase-shifted outputs of an autotransformer for small power application is an interesting alternative.

References

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

[2] Mahesh M. Swamy, Steven Rossiter, et. al, "Case Studies on Mitigating Harmonics in ASD Systems to meet IEEE 519-1992 Standards, **IEEE IAS Annual meeting, 1994, pp. 685-692.**

[3] "Power Electronic Converter Harmonics", Derek A. Paice, Paice and Associates, 1995. **ISBN 0-7803-1137-X**