International Journal of Power Electronics and Drive System (IJPEDS)
Vol. 7, No. 2, June 2016, pp. 431~439
ISSN: 2088-8694 ❒ 431

Harmonics Reduction in a Current Source Fed Quasi-Resonant Inverter Based Induction Heater

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Article Info

Article history:

Received Dec 4, 2015
Revised Jan 18, 2015
Accepted Feb 2, 2016

Keyword:

IGBT
LPF
Parallel quasi-resonant inverter
PSIM
THD

ABSTRACT

This paper investigates an approach to reduce the injection of unwanted harmonics from the input side power source. The proposed work finds an application of a suitably designed low pass filter (LPF) for harmonics reduction. The Low Pass Filter is incorporated in between the input power source and high frequency current source fed single switch parallel quasi-resonant inverter of a domestic Induction Heater. The inverter has an Insulated Gate Bipolar Junction Transistor (IGBT) switch. Undesirable harmonics injection to the input power source makes voltage and current waveforms non-sinusoidal. Fast Fourier Transform (FFT) analysis has been applied to study the effect of harmonics in frequency domain. The entire system is realized in Power System Simulator (PSIM) environment. It is finally proved that the proposed filter reduces the total harmonic distortion (THD) and the distortion factor (DF) of the input current. It is also proved that the proposed LPF can make the input power factor closer to unity.

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1. INTRODUCTION

In recent years, Induction heating has become a very popular technique to generate controllable desired high temperature quickly for applications like domestic cooking, melting steel, brazing, hardening and also in some medical applications. It is a contactless pollution free heating process [1–5]. Any induction heater consists of a ‘Resonant Inverter’ as a source of very high frequency alternating current. Induction heaters use different inverter topologies like single switch quasi-resonant [6–8], the half bridge inverter and full bridge inverters according to different applications. Insulated Gate Bipolar Junction Transistor (IGBT) is used as the power semiconductor switch in this application which is very much suitable and effective [10].

A quasi-resonant inverter can be suitably used in domestic induction heater. It uses a Parallel resonant load that has the capability of current-magnification to produce more heat. Any quasi-resonant inverter can be fed either from a voltage source or a current source. At the input side, it may have any passive or active filter to provide less harmonic injection to the input source [9] from the output side.

A current source fed quasi-resonant inverter is suitable for variable work load conditions. Harmonics distortions occur due to harmonics injection at the input current due to high frequency switching. The circuit diagram of total DC-link high frequency current source fed quasi resonant inverter fitted induction heater system is given in Figure 1.

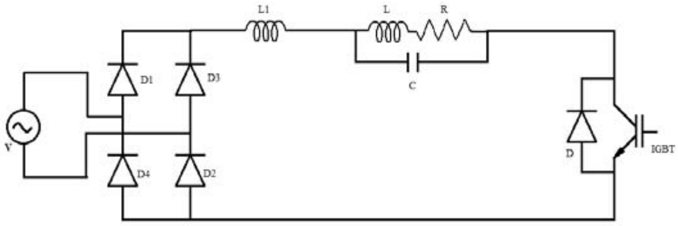

Figure 1. Circuit Diagram of DC-link high frequency current source fed quasi resonant inverter fitted induction heater system

2. OPERATIONAL PRINCIPLE OF PARALLEL QUASI-RESONANT INVERTER BASED INDUCTION HEATER

The Figure 2 as shown below represents simplified equivalent circuit of a single ended quasi-resonant inverter based induction heating system.

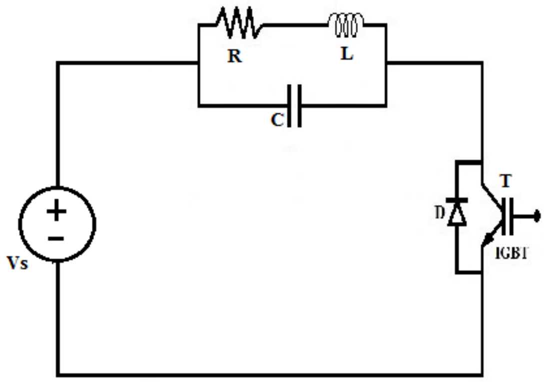

Figure 2. Simplified equivalent circuit of the quasi-resonant inverter based induction heater with load

The proposed quasi-resonant inverter based Induction Heater can operate under Zero voltage switching (ZVS) and Zero current switching (ZCS) conditions in two modes in every switching period. The equivalent circuits of each mode are shown in the Figure 3 and Figure 4 respectively.

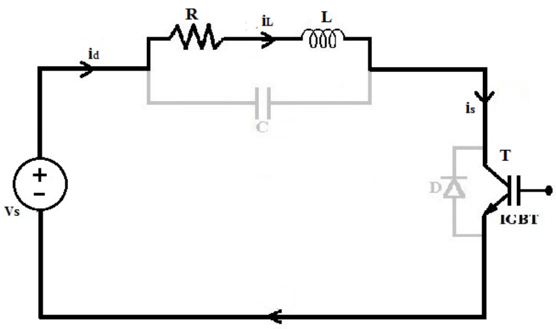

Figure 3. Simplified equivalent circuit of Mode-I while IGBT on

2.1. Mode-I Operation

In this mode the IGBT initiates conducting current immediately after its firing by the gate pulses. The voltage v_C across the capacitor C and across the R-L load is practically constant and equals to V_s. The load current i_L starts increasing exponentially and is equal to the switch current i_s. In this mode the inductor 'L' receives energy from the power source.

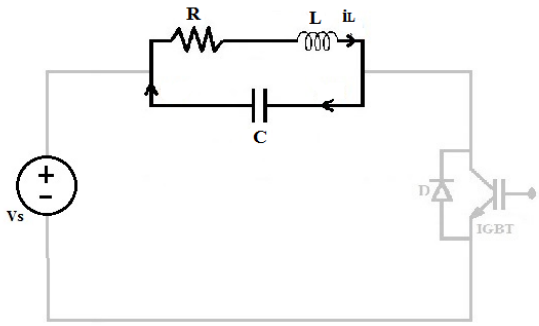

Figure 4. Simplified equivalent circuit of Mode-II while IGBT off

2.2. Mode-II Operation

In this mode, when the IGBT is turned-OFF, resonance in the parallel resonant circuit produces oscillation in the load current which exists till v_C reaches V_s. In this mode, the inductor 'L' of the load releases its stored energy and frequently exchanges it with the resonant capacitor 'C'.

Moreover, the proposed inverter system can perform under two different operating conditions to yield either ZVS condition or not which are described as follows.

2.3. Sub-Optimum Operation

In this mode, the capacitor voltage v_C becomes equal to the supply voltage V_s, when i_L is negative, so that to satisfy Kirchhoff's Voltage Law (KVL), the diode 'D' starts conducting, which decides the end of mode-II and the beginning of mode-I of the next switching cycle. The duration of the diode conduction is very short. At the time interval between the instant of diode conduction and the instant when i_L becomes zero, the IGBT must be triggered so that it can take over the diode current. Since, in this case, the IGBT is turned-ON before i_L reaches to zero, so switching loss occurs.

2.4. Optimum Operation

In this mode, the capacitor (C) voltage v_C reaches Vs and the inductor current i_L reaches zero at the same instant. The IGBT must be turned on at this instant to yield zero current switching (ZCS) condition and thus switching loss is negligible. The next cycle is initiated in this period. The two different conditions are graphically shown in the following Figure 5(a) and Figure 5(b) respectively.

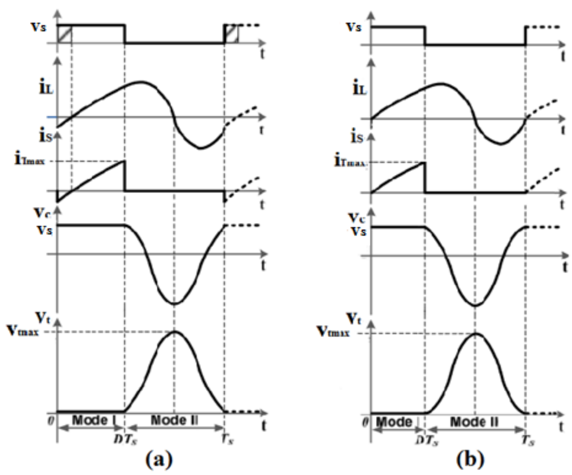

Figure 5. (a) Sub-optimum and (b) Optimum operations of quasi-resonant inverter

3. GENERAL EQUATIONS OF SUB–OPTIMUM AND OPTIMUM OPERATIONS

The following equations give the mathematical expressions of load current (i_L), the capacitor voltage (v_C) and the voltage across the IGBT (v_t) during the IGBT 'ON' period and 'OFF' period respectively for the sub-optimum operation of the inverter.

$$i_{Ln}(\omega_s t)=\frac{1}{2}\sqrt{1+\frac{1}{\alpha_n^2}}+\left(i_{Ln}(0)-\frac{1}{2}\sqrt{1+\frac{1}{\alpha_n^2}}\right)\exp\left(-2\frac{\alpha_n}{\omega_{sn}}\omega_s t\right) \quad \text{For } 0<\omega_s t\le 2\pi D \tag{1}$$

Where: $i_{Ln}(0)$ is the value of $i_{Ln}(\omega_s t)$ at $\omega_s t=0$, D = Duty cycle of the IGBT

and

$$i_{Ln}(\omega_s t)=\exp\left(-\frac{\alpha_n}{\omega_{sn}}(\omega_s t-2\pi D)\right)\left[\begin{array}{l} i_{Ln}(2\pi D)\cos\left(\dfrac{\omega_s t-2\pi D}{\omega_{sn}}\right)+ \\ \left(\sqrt{1+\alpha_n^2}-\alpha_n i_{Ln}(2\pi D)\sin\left(\dfrac{\omega_s t-2\pi D}{\omega_{sn}}\right)\right)\end{array}\right] \quad \text{For } 2\pi D<\omega_s t\le 2\pi \tag{2}$$

Where: $i_{Ln}(2\pi D)$ is the value of $i_{Ln}(\omega_s t)$ at $\omega_s t=2\pi D$

From Equation (2), $i_{Ln}(\omega_s t)$ has damped sinusoidal oscillation for $2\pi D < \omega_s t \le 2\pi$

$$v_{Cn}(\omega_s t)=1 \qquad \text{For } 0<\omega_s t\le 2\pi D \tag{3}$$

and

$$v_{Cn}(\omega_s t)=\exp\left(-\frac{\alpha_n}{\omega_{sn}}(\omega_s t-2\pi D)\right)\cdot\left[\begin{array}{l}\cos\left(\dfrac{\omega_s t-2\pi D}{\omega_{sn}}\right) \\ +\left\{\alpha_n-\dfrac{1}{2}\left(\alpha_n+\dfrac{1}{\alpha_n}\right)\left(1-\exp\left(-4\pi D\dfrac{\alpha_n}{\omega_{sn}}\right)\right)\right\}.\sin\left(\dfrac{\omega_s t-2\pi D}{\omega_{sn}}\right)\end{array}\right] \tag{4}$$

For $2\pi D < \omega_s t \le 2\pi$

and

$$v_t(\omega_s t)=V_s-v_C(\omega_s t) \qquad \text{For } 0<\omega_s t\le 2\pi \tag{5}$$

Where: i_{Ln}= normalized inductor current, v_{Cn}= normalized capacitor voltage, ω_{sn}= normalized switching frequency; Z_0= characteristic impedance of the inverter-load system; ω_0= resonant frequency of the parallel resonant system and ω_s= switching frequency of the IGBT.

The mathematical expressions of the above parameters are as follows.

$$i_{Ln}=\frac{i_L}{\dfrac{V_s}{Z_0}} \tag{6}$$

$$v_{Cn}=\frac{v_C}{V_s} \tag{7}$$

$$\alpha_n=\frac{R}{2\omega_0 L} \tag{8}$$

$$\omega_{sn}=\frac{\omega_s}{\omega_0} \tag{9}$$

$$Z_0=\sqrt{\frac{L}{C}} \tag{10}$$

$$\omega_0 = \sqrt{\frac{1}{LC} - \left(\frac{R^2}{4L^2}\right)} \tag{11}$$

The parameter α_n can be substituted by Q factor of the resonant load as follows.

$$Q = \frac{\sqrt{\frac{L}{C}}}{R} = \frac{1}{2}\sqrt{1 + \frac{1}{\alpha_n^2}} \tag{12}$$

From Equation (11), it is obvious, that $v_{Cn}(\omega_s t)$ has under damped oscillations for $2\pi D < \omega_s t \leq 2\pi$ and during the optimum operation of the proposed induction heating system, to achieve the turn-ON and turn-OFF and the following set of equations are to be satisfied.

$$v_t(2\pi) = 0; \frac{dv_t}{dt}(2\pi) = 0 \text{ and } i_{Ln}(2\pi) = 0 \tag{13}$$

So, from Equation (13), by KVL the capacitor voltage v_C reaches V_s and inductor current $i_L(t)$ reaches zero at the same instant and therefore, to obtain the optimum operating conditions, the load current $i_L(t)$, capacitor voltage $v_C(t)$ and the voltage $v_t(t)$ can be calculated using the equations (1)-(5) taking $i_{Ln}(0)=0$ and the relation among α_n, D and ω_{sn} can be expressed by the following equations respectively.

$$\left[1 - \exp\left(-4\pi D\frac{\alpha_n}{\omega_{sn}}\right)\right]\cos\left(\frac{2\pi(1-D)}{\omega_{sn}}\right) + \alpha_n\left[1 + \exp\left(-4\pi D\frac{\alpha_n}{\omega_{sn}}\right)\right]\sin\left(\frac{2\pi(1-D)}{\omega_{sn}}\right) = 0 \tag{14}$$

and

$$\exp\left(-2\pi(1-D)\frac{\alpha_n}{\omega_{sn}}\right)\left\{\begin{aligned}&\cos\left(\frac{2\pi(1-D)}{\omega_{sn}}\right)\\ &+\left[\alpha_n - \frac{1}{2}\left(\alpha_n + \frac{1}{\alpha_n}\right)\left[1 - \exp\left(-4\pi D\frac{\alpha_n}{\omega_{sn}}\right)\right]\right].\sin\left(\frac{2\pi(1-D)}{\omega_{sn}}\right)\end{aligned}\right\} = 1 \tag{15}$$

From the Equations (14) and (15) respectively, it is clear that the IGBT must be controlled with a variable duty cycle and variable switching frequency, which depend on the quality factor Q i.e. α_n that determined by the circuit components. Increase in α_n causes increase in switching frequency ω_s. For smaller values of Q, the voltage v_t cannot reach V_s in mode-II and soft turn-ON of the IGBT is not possible that causes high switching loss. So, higher value of Q should be chosen to get soft turn-ON.

4. TOTAL HARMONIC DISTORTION (THD)

It is a measure of distortion of a non-sinusoidal waveform from its sinusoidal fundamental component. This is an index to visualize the impact of harmonics and the level of distortion that they cause in a waveform. It is mathematically expressed by the following equation.

$$THD = \frac{\sqrt{\sum_{n=2,3,...}^{n} I_{nrms}^2}}{I_{1rms}} \tag{16}$$

Where: I_{Irms}= root mean squared value of the fundamental component and I_{nrms}= root mean squared value of the n-th harmonic component. High value of THD of any waveform indicates more distortion of it from its fundamental component.

5. DISTORTION FACTOR (DF)

A distortion factor indicates total amount of harmonics that remain in any waveform, after the waveform is subjected to 2nd order attenuation (i.e. divided by n^2), where, n= 1, 2, 3,…, n. It is given by the following expression.

$$DF = \sqrt{\sum_{n=2,3,....}^{n} \left(\frac{I_{nrms}}{n^2}\right)^2} \Big/ I_{1rms} \tag{17}$$

6. TOTAL HARMONIC DISTORTION (THD), DISTORTION FACTOR (DF) AND POWER FACTOR (PF) DETERMINATION OF THE PROPOSED INDUCTION HEATER WITHOUT FILTER

In Figure 6, the waveform of the input source current of the current source fed quasi-resonant Inverter without Filter is given using PSIM and the PSIM circuit diagram is shown in Figure 7. From this waveform the RMS value of the input source current is I_{irms} = 5.5544 A and the RMS value of the fundamental component is I_{1rms} = 3.904 A respectively. The Figure 8 also represents the FFT spectrum of this case. From the FFT spectrum the THD and the DF of the input current can be obtained which are as follows.

$$THD = \frac{\sqrt{\sum_{n=2,3,...}^{n} I_{nrms}^2}}{I_{1rms}} = \frac{\sqrt{\left(1.639^2+0.40038^2\right)}}{3.904} \times 100\% = 43.22\% \tag{18}$$

$$DF = \sqrt{\sum_{n=2,3,....}^{n} \left(\frac{I_{nrms}}{n^2}\right)^2} \Big/ I_{1rms} = \frac{\sqrt{\left(\frac{1.639}{3^2}\right)^2 + \left(\frac{0.40038}{5^2}\right)^2}}{3.904} \times 100\% = 4.67\% \tag{19}$$

From the PSIM diagram in Figure 7, the input power factor of the current source fed parallel quasi-resonant Inverter without Filter is 0.33183.

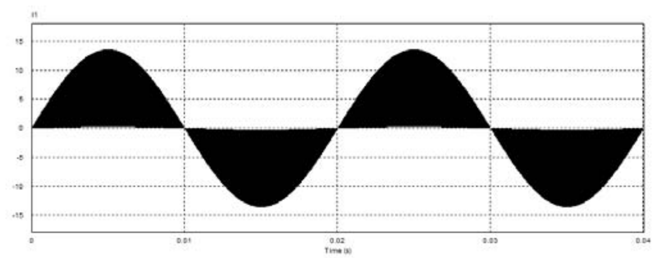

Figure 6. Input Current waveform of the current source fed quasi resonant inverter without filter

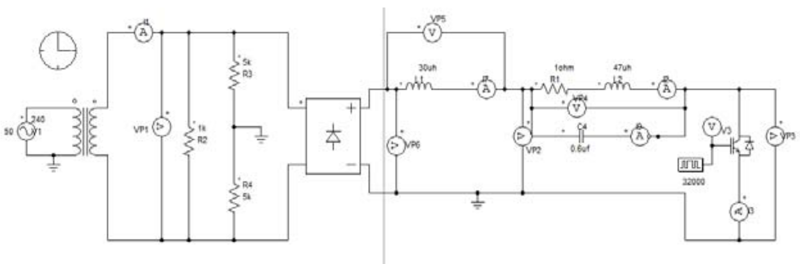

Figure 7. Simulated circuit of the current source fed quasi-resonant inverter without filter

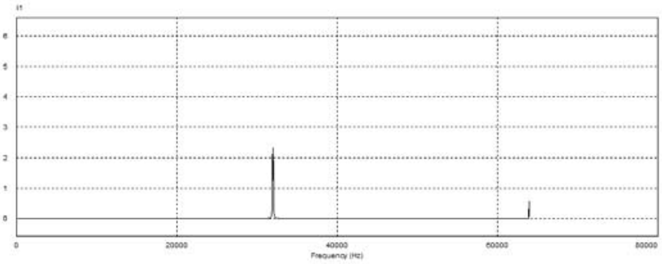

Figure 8. FFT spectrum of the input current of the current source fed quasi-resonant inverter without filter

7. TOTAL HARMONIC DISTORTION (THD), DISTORTION FACTOR (DF) AND POWER FACTOR (PF) DETERMINATION OF THE PROPOSED INDUCTION HEATER WITH FILTER

In Figure 9, the waveform of input source current of the current source fed quasi-resonant Inverter with Low Pass (LP) Filter is given using PSIM and the circuit diagram is shown in Figure 10. From this simulated waveform the RMS value of the input source current is obtained as I_{irms} = 7.2256 A and the RMS value of the fundamental component is I_{1rms} = 6.9179 A respectively. The Figure 11 also represents the corresponding FFT spectrum. From the FFT spectrum the THD and the DF values of the input current can be obtained which are as follows,

$$\mathrm{THD} = \frac{\sqrt{\sum_{n=2,3,...}^{n} I_{nrms}^{2}}}{I_{1rms}} = \frac{\sqrt{\left(0.89262^{2}\right)}}{6.9179} \times 100\% = 12.9\% \tag{20}$$

$$\mathrm{DF} = \sqrt{\sum_{n=2,3,....}^{n} \left(\frac{I_{nrms}}{n^{2}}\right)^{2}} \Bigg/ I_{1rms} = \frac{\sqrt{\left(\frac{0.89262}{3^{2}}\right)^{2}}}{6.9179} \times 100\% = 1.43\% \tag{21}$$

From the PSIM diagram as shown in Figure 10, the input power factor of the current source fed quasi- resonant Inverter without Filter is 0.80045.

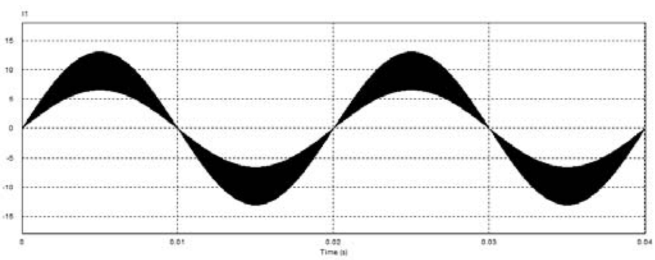

Figure 9. Input Current waveform of the current source fed quasi-resonant inverter with filter

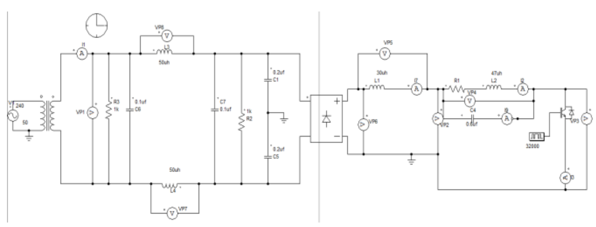

Figure 10. Simulated circuit of the current source fed quasi-resonant inverter with filter

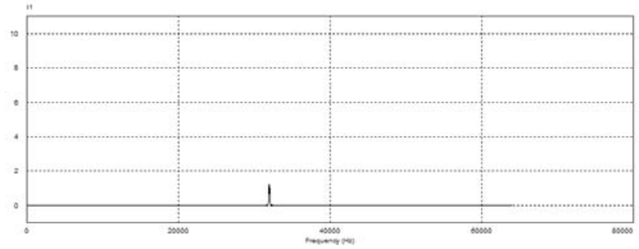

Figure 11. FFT spectrum of the input current of the current source fed quasi-resonant inverter with filter

8. SIMULATION AND RESULTS

From the PSIM simulations as shown in Figure 7 and Figure 10 of the proposed induction heating system, two different results are obtained. The Figure 6 and Figure 8 show the input current waveform and the FFT spectrum of the input current of the inverter without filter. From the FFT spectrum the 3rd and the 5th harmonics are found to be dominant making the input current non-sinusoidal with a relatively higher THD and DF values. Besides the input power factor is 0.33183, which is quite low. As such, to suppress the effects of these harmonics at an improved power factor an LC passive filter is used. The Figure 9 shows the input current of the inverter with filter and Figure 11 shows the FFT spectrum of the input current of the inverter with filter. The filter circuit used in this case is a passive LC Low Pass Filter (LP) whose parameters are suitably selected to get desired results. From the FFT spectrum, the harmonics are almost absent in the input current and thus the THD is relatively less and DF is also greatly reduced. Besides the input power factor is quite improved to a value of 0.80045 i.e. close to unity, which is the utility of the installed filter.

9. CONCLUSION

In the present work, the impact of the presence of harmonics in the input current is studied for a current source fed quasi-resonant inverter operating at high switching frequency with and without incorporating a passive filter. In the first case without filter, higher THD and DF values of the input current indicate the presence of high degree of noise level. The low input power factor indicates greater reactive power injection from the power frequency input source. In the second case, when the LC Low Pass Filter is incorporated, the THD and DF values of the input current reduce significantly to a low value and thus reduces the noise level of the input current after harmonics suppression satisfactorily and in the second case, the input power factor is also improved to a larger value resulting less reactive power injection from the input power source.

ACKNOWLEDGEMENTS

Authors are thankful to the UNIVERSITY GRANTS COMMISSION, Bahadurshah Zafar Marg, New Delhi, India for granting financial support under Major Research Project entitled “Simulation of high frequency mirror inverter for energy efficient induction heated cooking oven” and also grateful to the Under Secretary and Joint Secretary of UGC, India for their active co-operation.

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