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Control of Saturation level in the magnetic core of a welding transformer by Hysteresis Controller (HC) and Proportional Integral (PI) Controller

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ABSTRACT

The objective of this paper is to analyse the performances of two controllers such as Hysteresis control (HC) and proportional integral (PI) control to control saturation level in the magnetic core of a welding transformer in a middle-frequency direct current (MFDC) resistance spot welding system(RSWS). It consists of an input converter, welding transformer, and a full-wave rectifier mounted at the transformer secondary. The unequal ohmic resistances of the two transformer's secondary circuits and the different characteristics of the diodes of output rectifier certainly lead to the magnetic core saturation which, consequently, causes the unwanted spikes in the transformer's primary current and over-current protection switch-off. The goal is to analyse the performance of both controllers in terms of transients, total harmonic distortion(THD) and variations in primary current and flux in the magnetic core of a welding transformer of highly nonlinear system of RSWS. The simulation study has been done in Matlab/Simulink environment and presented performance analysis. The responses shows that from the aforementioned aspects, proportional integral Controller is the better choice for controlling the saturation level in magnetic core of a welding transformer which is widely used in automobile industry welding system.

Keywords**:** Magnetic core, core saturation, Hysteresis, Proportional Integral control, welding transformer.s

I. INTRODUCTION

Resistance spot welding is one of the most widely used inexpensive and efficient material joining processes in the automotive industry. This work deals with the modeling, analysis and corresponding control design of the welding current source, which represents an electromagnetical subsystem of the entire welding system. However, the technical questions of welding itself are not a subject of this work.

Medium Frequency Direct Current (MFDC) Resistance Spot Welding System (RSWS) is extensively used in a large number of industries, such as automotive, nuclear power, home appliances, as well as civil infrastructure products. It has a great many particular positive features in industrial

applications [1]. The working process of the RSWS is a very complex one, which involves interactions among electromagnetic, thermal, mechanical, and metallurgical phenomena. Compared to Alternative Current (AC) resistance spot welding device, MFDC RSWS has a more complex structure because it needs to generate a higher frequency than that by its original power supply source. Generally speaking, the working frequency of the MFDC RSWS is about 1000Hz, while the frequency of the original power supply is about 50/60Hz. Some necessary transitions

from common electrical power to a low-voltage,

high-current and high-frequency electrical power supply

should be accomplished. The complex structure and working mechanism of the system may induce some problems [2-4], such as magnetic saturation and unwanted spikes in the currents. Klopcic [2-5] analyzed the special electromagnetic structure and proposed one method to deal with them. However, the work focused on the magnetic saturation and the mathematic model which the work used was developed using electromagnetic features. Thus the work concerned fewer about the variation of welding current. In this paper, after studying the structure and working principle of MFDC RSWS, different possible operating modes of the system is found. The modes can be described by how many diodes in two secondary coils of welding transformer are switched on. And then a new mathematical model is developed to precisely describe the dynamic behavior of the whole system.

When the current spikes are prevented actively, closed-loop control of the welding current and magnetic core flux density is required. Thus, the welding current and the magnetic core flux density must be measured. While the welding current is normally measured by the Rogowski coil [10], the magnetic core flux density can be measured by the Hall sensor or by a probe coil wound around the magnetic core. In the latter, the flux density value is obtained by analogue integration of the voltage induce in the probe coil [7]. Integration of the induced voltage can be unreliable due to the unknown integration constant in the form of remnant flux and drift in analogue electronic components. The drift can be kept under control by the use of closed-loop compensated analogue integrator [9].

An advanced, two hysteresis controllers based control of the RSWS, where current spikes are prevented actively by the closed-loop control of the welding current and flux density in the welding transformer's magnetic core, is presented in [9]. This solution requires measuring of the welding current, while instead of measured flux density only information about magnetization level in the magnetic core is required. Some methods tested on welding transformer's magnetic core, that can be applied for magnetization level detection are presented in [7], [8]. All these methods require Hall sensor or probe coils which make them less interesting for applications in industrial RSWS, due to the relatively high sensitivity on vibrations, mechanical stresses and high temperatures. In order to overcome these problems, PI controller is introduced. A dc-dc converter must provide a regulated dc output voltage under varying load and input voltage conditions. The converter component values are also changing with time, temperature, pressure, and so forth. Hence, the control of the output voltage should be performed in a closed-loop manner using principles of negative feedback. The most common closed-loop control method for PWM converter, namely, the current-mode control is presented schematically in below section. The current-mode control scheme is presented in section III. An additional inner control loop feeds back an inductor current signal, and this current signal, converted into its voltage analog, is compared to the control voltage. This modification of replacing the sawtooth waveform of the voltage-mode control scheme by a converter current signal significantly alters the dynamic behavior of the converter, which then takes on some characteristics of a current source. Among other control methods of converters, a hysteretic (or bang-bang) control is very simple for hardware implementation. However, the hysteretic control results in variable frequency operation of semiconductor switches. Generally, a constant switching frequency is preferred in power electronic circuits for easier elimination of electromagnetic interference and better utilization of magnetic components So the constant switching frequency gives better performance in the application of resistance spot welding system (RSWS). It uses the hysterisis controller. When it is used frequency cant be maintained. And the transformer saturation also happens due to the change in resistance of the RSWS.

In this paper, PI controller works well and giving better performance in terms of limiting flux density in order to limit the spikes in the primary current caused by the saturation to prevent the over current protection switch-off.

II. DYNAMIC MODEL OF THE RSWS

MFDC RSWS consists of an input rectifier, an H-bridge inverter, a welding transformer with a full-wave rectifier and corresponding load. A detailed schematic presentation of MFDC RSWS is shown in Fig. 1 [4]:

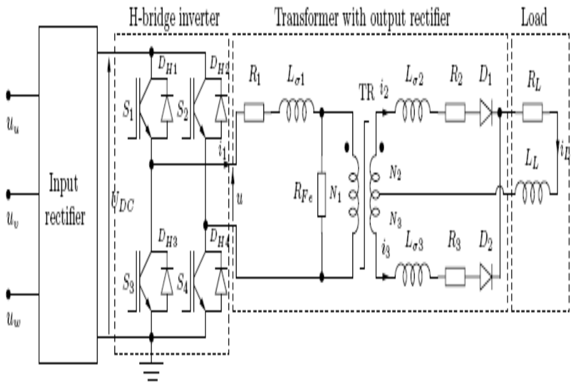


Fig.1.schematic representation of RSWS

The input rectifier is a three-phase full-wave rectifier,

which can change the common three-phase alternative

current (AC) voltage into a proper single-phase current. The output welding current is controlled by the voltage pulses generated through the pulse width modulation (PWM) controller to drive the H-bridge inverter. In above schematic presentation, the AC voltages u_u, u_v, u_w, which are provided from the common electric grid, are rectified and smoothed through the input rectifier in order to produce an approximate direct voltage U_{DC}. The square wave voltage u, which is the voltage in the transformer's primary coil, is generated by the H-bridge inverter which is composed of IGBT transistors S_1 to S_4 and corresponding diodes D_{H1} to D_{H4}.During working process, the PWM controller is applied to generate IGBT' switching patterns for required input voltages of the welding transformer. In other words, control of the MFDC RSWS is the control of the status of the IGBTs in real time. The welding transformer has one primary coil (denoted by subscript 1 in Fig. 1) and two secondary coils (denoted by subscripts 2 and 3 in Fig. 1). N_1, N_2 and N_3 are the number of turns, i_1, i_2 and i_3 are the currents in the coils, $L_{\sigma 1}$, $L_{\sigma 2}$ and $L_{\sigma 3}$ are the leakage inductances, while R_1, R_2 and R_3 are the ohm resistances of the corresponding transformer's coils. The welding transformer, which contains special nonlinear magnetizing features, is represented by *TR*. The iron core losses of *TR* are accounted for by the resistor R_{Fe}. The secondary coils of *TR* are connected to output rectifier diodes D_1 and D_2. The resistor and induction coil of the load are denoted by R_L and L_L.

The operation for the MFDC RSWS is to regulate the welding current i_L to a magnitude in between the predetermined upper bound I_{MAX} and lower bound I_{MIN} (a desired constant value is the best, but this is impossible to achieve, thus a proper bound is an alternative). At the same time, the magnetic flux density (*B*) of the transformer's iron core should be in between its upper and lower magnetic saturation bounds[4]: [$-B_M$, B_M]. This can be achieved by changing the input voltage for the welding transformer in three states: *U*, *-U*, and *0V*, through adjusting the patterns of IGBTs in the H-bridge inverter by PWM controller.

The welding current (*iL*) is the sum of the currents in the two secondary coils (Fig.1). A positive input voltage (*U*) can actuate the top secondary coil; while a negative input voltage (*-U*) can actuate the bottom secondary coil. Hence, both of *U* and *-U* can increase the load current. Only a zero input voltage (*0V*) can decrease the load current. However, *U* and *-U* can generate the opposite effect for variation of the magnetic flux density (*B*). For example, if when *U* increases the load current, but simultaneously *B* reaches the bound, *U* must be changed into *-U*, which can also increase the welding current, but *B* will increase toward the opposite direction, which can avoid the magnetic saturation.

When opposite input voltage is provided, the energy

which is stored by inductance coil in original circuit will decrease; while that in the other circuit will increase. And when the welding current should be decreased and a zero voltage is provided, the inductor coil will substitute the power source and a new back circuit will form. Thus in a certain period, both of the two diodes in the secondary coils are switched on at the same time because the inductor coils can suspend the transformation. And this phenomenon can appear when the pattern of input voltage changes between its three states (*U*,*-U* and *0V*) because of the same reason. Normally, it is impossible for the two diodes to be switched off at the same time, unless the welding process is over.

The dynamic model of the RSWS was built based on the schematic presentation, shown in Fig.1. In this work the model is built with the programme package Matlab/Simulink based on the following set of equations (1) – (9).

$$uH = R_1 i_1 + L\sigma_1(di_1/dt) + N_1(d\varphi/dt) \quad (1)$$

$$0 = R_2 i_2 + L\sigma_2(di_2/dt) + N_2(d\varphi/dt) + dip_1 + R_L i_L + L_L(d(i_2 + i_3)/dt) \quad (2)$$

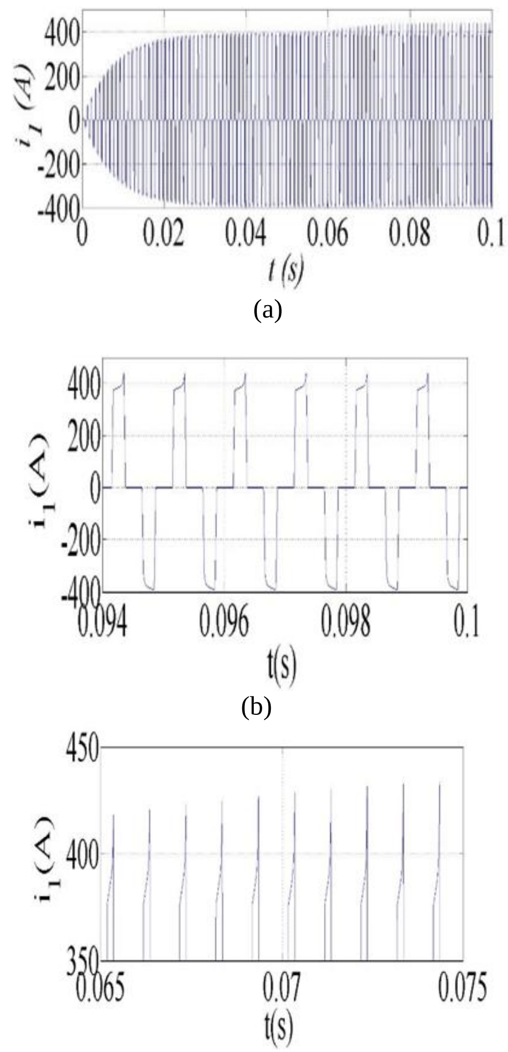
$$0 = R_3 i_3 + L\sigma_3(di_3/dt) - N_3(d\varphi/dt) + dip_2 + R_L i_L + L_L(d(i_2 + i_3)/dt) \quad (3)$$

$$N_1 i_p + N_2 i_2 - N_3 i_3 = H(B) l_{ic} + 2\delta B/\mu_0 \quad (4)$$

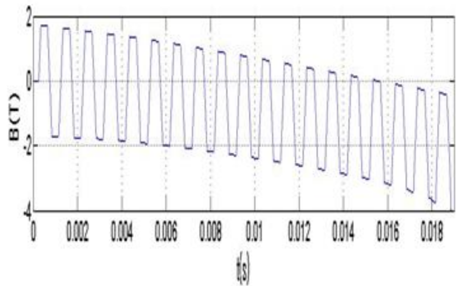
$$i_L = i_2 + i_3 \quad (5)$$

$$i1 = i_{Fe} + i_p \quad (6)$$

The results of simulations, obtained by the dynamic model of the RSWS, show that small difference in resistances R2, R3 and in characteristics of the rectifier diodes D1 and D2 can cause unbalanced time behavior of the magnetic core flux and the current spikes in the primary current i1, shown in Fig.2. The a) and b) graphs in Fig. 2 show the same variables in different time scales. The current spikes appear approximately after 0.06s (Fig.2(c)). After 0.07s the current spikes become high enough to cause the over-current protection switch off of the RSWS.



(c)



(d)

Fig.2. (a),(b)and (c) : Time behaviour Primary Current i_1 (d) FluxDensity

III. CONTROLLER DESIGN

The current spikes in transformer primary current are the direct consequence of transformer iron core saturation caused by the offset of flux density (Figs. 3 and 5). The basic idea on how to eliminate these current spikes is, therefore, the design of advanced control, which will closed-loop control both, saturation level in the transformer iron core and the welding current.

A dc-dc converter must provide a regulated dc output voltage under varying load and input voltage conditions. The converter component values are also changing with time, temperature, pressure, and so forth. Hence, the control of the output voltage should be performed in a closed-loop manner using principles of negative feedback. The most common closed-loop control method for PWM converters, namely, the current-mode control, are presented schematically in Fig.5.

I. *Hysteresis Controller* :

Reference currents are generated by DC to AC converters using a current control technique such as a hysteresis control. The hysteresis band is used to control load currents and determine switching signals for inverters gates, George & Agarwal (2007) Suitable stability, fast response, high accuracy, simple operation, inherent current peak limitation and load parameters variation independency make the hysteresis current control as one of the best current control methods of voltage source inverters. In this approach the current error, (difference between the reference and inverter currents) is controlled in hypothetical control band surrounding reference current.

When the load current exceeds the upper band, the comparator output activated so the output voltage is changed in such a way to decrease the load current and keep it between the bands and deactivated at lower limit. Switching frequency varies with respect to distance between upper and lower band. The other parameters like inverter-network inductance and DC link voltage affect significantly on the switching frequency. inverter can be controlled in unipolar or bipolar PWM method. In this approach the current error, (difference between the reference and inverter currents) is controlled in hypothetical control band surrounding reference current as shown in Figure 3.

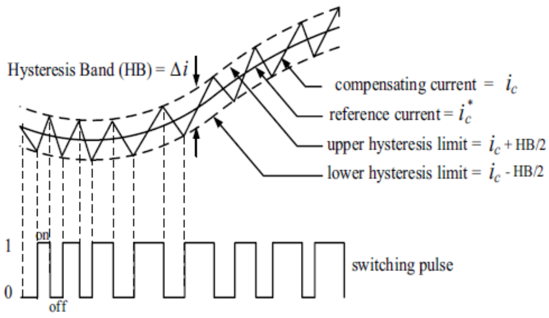


Fig.3 : Basic concept of Hysteresis Control

In hysteresis current control based on unipolar PWM, there are two upper bands and lower bands in order to change the slop of inverter output current based on their level voltages, +Vo, 0 and -Vo. The idea is to keep the current within the main area but the second upper and lower bands are to change the voltage level in order to increase or decrease the di,/dt of inverter output current.

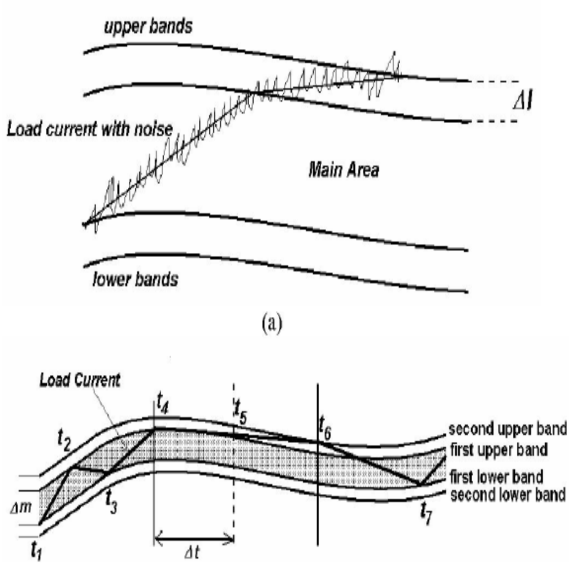


Fig.4 : Noisy Load Current with lower and upper bands.

ΔI cannot be very small as the noisy signal changes the switching time due to instantaneous comparison between the load and the reference currents and increases the switching losses and it cannot be big as the total harmonic distortion may be increased.In APF, load current has several different slopes within one cycle and to have a fast current tracking, the control algorithm in unipolar current control has been defined based on magnitude and time errors control as shown in Figure 3 (b). In this case, the second upper or lower band values can be big enough in order to remove the noise issue of the inverter output current but the second decision to change the level is based on time error. For example, when the load current exceeds the first upper band at t4, the output voltage of inverter is change from $+V_o$ to 0. The controller waits for Δt, if the inverter output current does not cross the second upper band within this period, thenthe controller changes the output voltage from zero to –Vo at t5. In this case, when the slope of reference current is close to the slop of inverter output current, then the time error control improves the quality of the APF and pushes the inverter current into the main area. It is proven that the current control based on unipolar PWM has a low switching losses or better performance compare to the other methods of control techniques. The load and compensated currents THD(total harmonics distortions) can be reduced sufficiently by using hysteresis current control based unipolar PWM.

(b) *PI Controller*

In control engineering, a PI Controller (proportional-integral controller) is a feedback controller which drives the plant to be controlled with a weighted sum of the error (difference between the output and desired set-point) and the integral of that value. PI controllers consist of a proportional gain that produces an output proportional to the input error and an integration to make the study state error zero for a step change in the input.

The controller output is given by

$$k_p\Delta + k_I \int \Delta \, dt \qquad (1)$$

where Δ is the error or deviation of actual measured value (PV) from the set-point (SP).

$$\Delta = SP - PV. \qquad (2)$$

A PI controller can be modelled easily in software such as Simulink using a "flow chart" box involving Laplace operators:

$$c = \frac{G(1+\tau s)}{\tau s} \qquad (3)$$

Where, $G = K_P$ = proportional gain and $G / \tau = K_I$ = integral gain.

Setting a value for G is often a tradeoff between decreasing overshoot and increasing settling time. The integral term in a PI controller causes the steady-state error to reduce to zero, which is not the case for proportional only control in general.

The current-mode control scheme is presented in Fig.1 An additional inner control loop feeds back an inductor current signal, and this current signal, converted into its voltage analog, is compared to the control voltage. This modification of replacing the sawtooth waveform of the voltage-mode control scheme by a converter current signal significantly alters the dynamic behavior of the converter, which then takes on some characteristics of a current source.

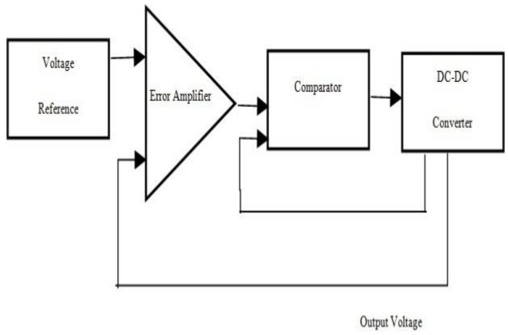


Fig. 5: Current mode control of PI control

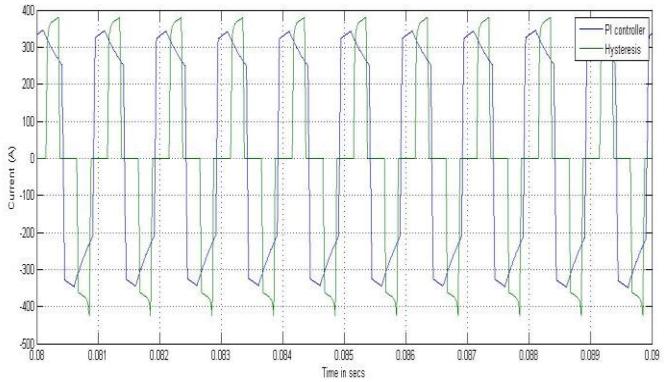
The current-mode control scheme is presented in Fig.5(b) An additional inner control loop feeds back an inductor current signal, and this current signal, converted into its voltage analog, is compared to the control voltage. This modification of replacing the sawtooth waveform of the voltage-mode control scheme by a converter current signal significantly alters the dynamic behavior of the converter, which then takes on some characteristics of a current source.

The output current in PWM converters is either equal to the average value of the output inductor current or is a product of an average inductor current and a function of the duty ratio. In practical implementations of the current-mode control, it is feasible to sense the peak inductor current instead of the average value. As the peak inductor current is equal to the peak switch current, the latter can be used in the inner loop, which often simplifies the current sensor. Note that the peak inductor (switch) current is proportional to the input voltage. Hence, the inner loop of the current-mode control naturally accomplishes the input voltage-feed forward technique. Among several current-mode control versions, the most popular is the constant-frequency one that requires a clock signal. Advantages of the current- mode control are the input voltage feed forward, the limit on the peak switch current, the equal current sharing in modular converters, and the reduction in the converter dynamic order. The main disadvantage of the current-mode control is its complicated hardware, which includes a need to compensate the control voltage by ramp signals (to avoid converter instability).Among other control methods of converters, a hysteretic (or bang-bang) control is very simple for hardware implementation.

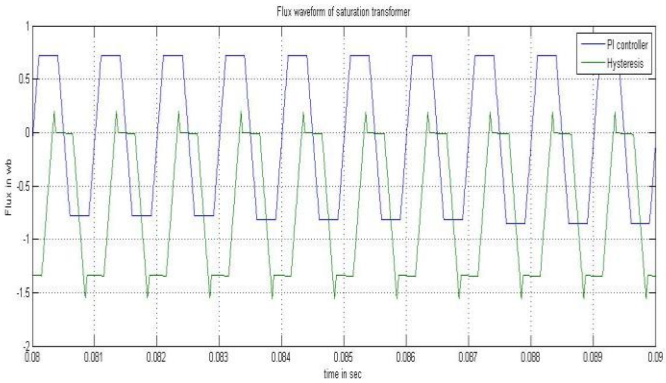
However the hysteresis control results in variable frequency operation of semiconductor switches. Generally a constant switching frequency in power electronic circuits for easier elimination of electromagnetic interference and better utilization of magnetic components.

IV RESULTS AND DISCUSSION

Simulation results of Hysteesis and PI controllers are presented here. primary current and fluxdensity of both controlleres are shown in fig.6.(a) and (b). In fig.6(a), primary current of a welding transformer can be seen spikes in a time scale of 0.08 to 0.09 sec. Since the hysteresis controller is not able to maintain the flux density with in the preset values (i.e., -1T to 1T) this spikes are not eliminated successfully. For the time scale, this spikes are completly eliminated by maintaining the flux density with in the preset values as -1T to 1T with PI controller can be seen in fig.6.(b). Total Harmonic Distortion (THD) variation can be seen in fig.7.(a) and (b). THD of HC is 60.81% and PI is 37.32%. PI gives better performance than the HC from the aforementioned aspects.

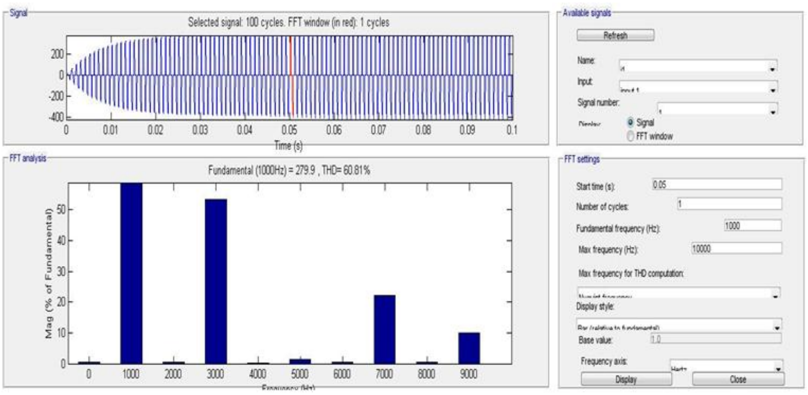


(a)

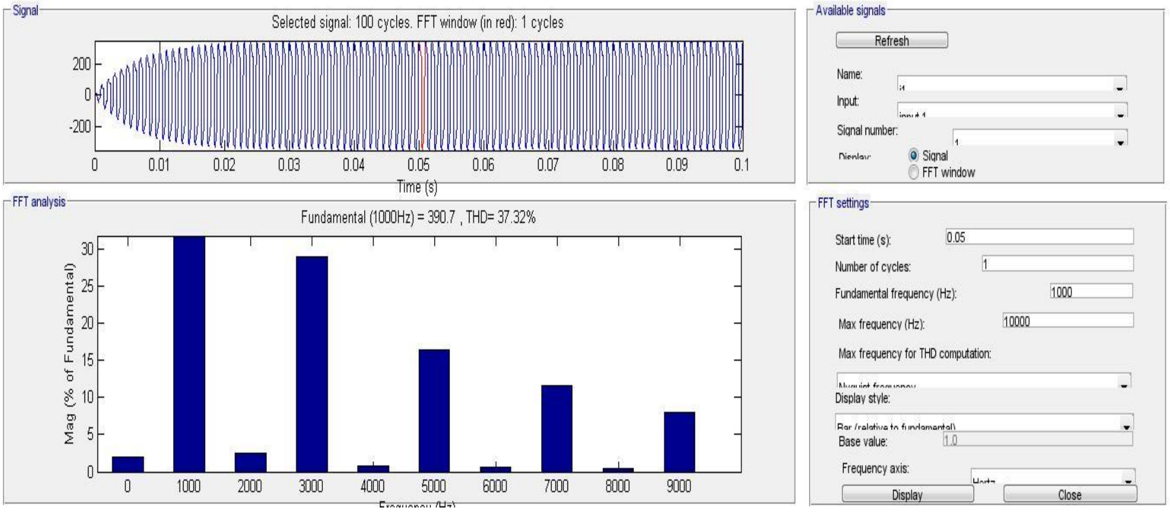


(b)

Fig.6. (a) and (b): combined Primary current and FluxDensity of HC and PI control



(a)



(b)

Fig.7. (a) and (b): THD of HC and PI Control

V CONCLUSION

In this paper, two controllers such as Hysteresis and PI are successfully designed. Based on the simulation results and the analysis, a conclusion has been made that PI control having less THD(37.32%) than Hysteresis control(60.81%). PI controller is capable of controlling the saturation level in the magnetic core of a welding transformer of nonlinear RSWS system

Flux Density can be maintained with in a preset values successfully in order to eliminate the spikes in the primary current of a welding transformer.

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