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Efficacy of STC in Magnetically Coupled High Frequency Power Electronics Systems: Full-Bridge DC-DC Converter

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Abstract— Due to its simple gain selection and implementation procedures high-gain sliding mode control idea was introduced. Apart from achieving robustness features, the aim was to reduce product-design cycle time. To overcome its basic chattering issues, it was replaced by complex super twisting control (STC). Subsequently, the research in this field has been extensive, intense and persistent. The idea has been validated in several applications generating great hope in minds of industry experts. Its real success, however, would depend on its adoption and diffusion into industry domain, where an analysis of validating STC based products is due. In STC, selection of gains is based on worst-case values of disturbance and/or its derivative. Can such procedure ensure controller's reliable functioning? This article proposes that such simplistic procedure of gain selection may not work for power electronics controllers where, particularly, the actuation or controlled power transfer to load is through magnetically-coupled system. Using practical approach, this article elaborates that optimally designed transformer, driven by high-gain super twisting controller, invites problem of core saturation leading to higher switching losses, poor operating duty cycle of the system and there could be problem of reliability. It further details an alternate high-gain controller that generates superior efficacy.

Index Terms—Constant current (CC) load, full-bridge DC-DC converter (FBC), proportional plus integral (PI) control, second order sliding mode control (SOSMC), super twisting control (STC), transformer core saturation.

I. INTRODUCTION

GRADUALLY, for effective and efficient use, major part of energy demand would be handled through use of compatible power electronics systems [1]-[4]. Full-bridge DC-DC converter FBC (see Fig. 1) is one popular configuration being used for control of power flow to wide range high-power applications [4]-[8]. High-power FBC consists of several functional blocks, such as,

1. Power converter and topologies [8] used, therein
2. Transformer for isolation and voltage translation [9]
3. Sensors and actuators [10]
4. Embedded system [11]
5. Control concepts [11]-[22], etc.

Ideally, for effective utility of each block, they should be functionally compatible to each other; should not pose as any constraint to other. Then, say, for control of secondary side outputs (e.g., current I_a or voltage V_L), the system in Fig. 1 could be represented as secondary side controller (see Fig. 2a).

In industrial domains, the utility of PI control [11]-[15] is unmatched. Since its inception its application potential was recognized and the concept was soon integrated into industrial control system applications. Due to integral action, it could achieve zero steady state error in presence of uncertainty [13]. PI control is conservative because its two gains are chosen based on the desired response characteristics involving the complete dynamical system that includes the power controller. The gain selection is flexible to desired response characteristics.

The super twisting control (STC) [16]-[22] is an improved version of chattering-prone discontinuous sliding mode control function [23], the control is continuous. Though, it is perceived, at least, in theoretical domain, as a superior concept, its utility in industrial application is still to be demonstrated [24], [25]. From industry perspective, it is important to analyze the issue in details purely in practical domain.

Fig. 1: Typical power circuit full-bridge DC-DC converter

Like PI control, STC also needs setting of two control loop gains [21], [22]. But, the approach of gain selection is different, it is based on the magnitude of matched disturbances present in process or load. Neither the primary side of transformer TR (see Fig. 1) nor the inverter switching frequency f_s play any role in gain selection. Compared to PI, value of each gain is large. It makes the duty cycle d_{pwm} of PWM pulses sensitive to error. For magnetically coupled controlled actuation, fast rate of change in d_{pwm} , caused by disturbance or ripple content in STC function, could make adverse impact on nonlinear magnetic circuit. The prospect of DC bias in core would be more [26]-[28]. Presence of DC flux in nonlinear core would result increase in power loss in inverter of multiple origin. The worst could lead to core saturation where one pair of devices draw large magnetizing current and incur large power loss. There could be pre-matured thermal tripping of the controller and or

failure of power devices. The article analyzes with detailed practical demonstration that high gain fractional order STC function may be less effective for control application where power transformer is part of controlled energy transfer to the load. The article is structured as follows: Section II briefly discusses a typical FBC. It further elaborates the results of PI concept for control of current for a particular load profile. Section III develops one STC based SOSM controller for the same load profile, and further validates the concept with superior control response. Section IV elaborates the next step of innovation process i.e., complete product validation. It details that STC fails to clear a magnetically coupled product on several aspects such as on power loss, poor availability and reliability issues. Finally, Section V, details, with practical validation, an alternate high-gain controller for robust control of secondary side variables with better efficacy.

II. FB DC-DC CONVERTER AND PI CONTROL

The power circuit of FBC of Fig. 1 could be used for control of either current or voltage of load. Though, load characteristics compatible several soft-switched topologies exist. for clarity, hard-switched version would be discussed here. It is compatible to any transient load conditions. It consists of power inverter, transformer TR, secondary side fast rectifier and an inductor L_1 . Due to small conduction loss, metallized polypropylene capacitor is used in DC link for C_{DC} . Each IGBT turns-on at half the supply voltage V_{DC} , the turn-on loss of inverter is small. For ideal TR with zero leakage inductance l_{TR} , IGBTs turn-off at peak magnetizing current I_m , not dependent on load current. Turn-off loss of each IGBT and recovery and conduction losses of diodes D1-D4 would as well be small. The value of switching frequency f_s could be increased if SiC Mosfets are used in place of Q1-Q4, or, SiC diodes are used in place of D1-D4. Here, as representative load, constant current (CC) arc welding process would be considered. From control perspective of secondary variables, FBC could be represented as shown in Fig. 2a. When either pair of (Q1, Q4 or Q2, Q3) or equivalently Q5 turns on the dynamics of current I_a is expressed as,

$$\frac{d}{dt}(I_a) = \frac{1}{L_1}(V_{oc} - V_L) = \frac{1}{L_1}\left(\frac{V_{dc}}{n}d_{pwm} - V_L\right) \quad (1)$$

When Q1-Q4 or Q5 is turned off, D5/D6 or equivalently D8 conducts, corresponding current dynamics is,

$$\frac{d}{dt}(I_a) = \frac{1}{L_1}(-V_L) \quad (2)$$

V_{oc} is controlled input voltage to arc circuit, V_L is arc voltage. L_1 , of significant value, is pre-requisite to contain ripple for wide range arc loads and also for ensuring arc stability. Furthermore, arc is extinguished when short-circuiting takes place, but the energy stored in it assists re-ignition of arc. During switching on and off of Q5 the desired current I_a in both cases are shown in Fig. 2b. For a particular application, with fixed ripple current ΔI_a , the value of L_1 depends on f_s , because,

$$L_1 = \frac{1}{2f_s} \frac{V_L(V_{oc} - V_L)}{\Delta I_a V_{oc}} \quad (3)$$

At high f_s , small value of L_1 would ensure continuity of I_a for arc stability. To reduce associated high-frequency power loss,

load-compatible soft-switching topologies [29], [30] are mostly adopted.

Fig. 2: a) For secondary side control, ideal representation of full-bridge converter, and, b) Typical secondary side DC waveforms

A. PI Control for Constant Current Applications

For control of I_a , the control inputs V_{oc} in power domain and u_1 in control circuit are expressed as,

$$V_{oc} = k_1 u_1 = \frac{d_{pwm} V_{DC}}{n} = V_{oc}(\max) d_{pwm} \quad (4)$$

$$\text{where, } d_{pwm} = \frac{k_1}{V_{oc}(\max)} u_1 = k_2 u_1 \quad (5)$$

$V_{oc}(\max)$ is maximum value of V_{oc} , k_1 and k_2 are constant. V_{DC} is input voltage, n is turns ratio of TR, d_{pwm} is PWM duty cycle of inverter. PI control is mostly used in power electronics controllers. The control function for power actuation is,

$$V_{oc} = K_P e + K_I \int e dt; \quad e = I_{ref} - I_a \quad (6)$$

K_P and K_I are constant and e is current error. The control u_1 in PI i.e., u_{PI} could be expressed as,

$$u_{PI} = \frac{K_P}{k_1} e + \frac{K_I}{k_1} \int e dt = k_p e + k_i \int e dt \quad (7)$$

The closed loop transfer function $G_{CL}(s)$ could be written as,

$$G_{CL}(s) = \frac{\frac{1}{L_1}(K_P s + K_I)}{s^2 + \frac{1}{L_1}(R_L + K_P) + \frac{K_I}{L_1}} \quad (8)$$

$R_L = V_L/I_a$ represents the equivalent resistive load in V_L . The denominator $d(s)$ could be written as $d(s) = s^2 + 2\xi\omega_n s + \omega_n^2$ where L_1 and K_I decide the natural frequency ω_n and damping ratio ξ depends on controller gain K_P . In PI control, ω_n is related to frequency of rectified voltage V_{oc} i.e., $2f_s$, like [14],

$$\omega_n = \frac{2f_s}{5\xi}, \quad \tau_c = \frac{1}{\xi\omega_n} \quad \text{and} \quad \tau_s = 4\tau_c \quad (9)$$

Where τ_c is time constant of the system and τ_s is settling time. To improve response time τ_s , f_s is increased.

Considering the value of ξ as 0.8, the value of ω_n for V_{oc} at 40 kHz is 10 kHz. The value of K_P and K_I are [12],

$$K_P = \left(2\xi\omega_n - \frac{R_L}{L_1}\right) L_1 \quad \text{and} \quad K_I = L_1 \omega_n^2 \quad (10)$$

Considering load R_L of arc at 0.09Ω and with L_1 at 75 μH , to generate controlled V_{oc} , the value K_P is 1.2 and K_I is 7500.

B. Practical Implementation of PI Control

For practical validation, one 20 kHz, 400A FBC was developed. For sensing of I_a , a Hall effect-based transducer (output: 1A = 10mV) was used. The value of k_1 and k_2

respectively was 8 and 0.1. For practical implementation, in control circuit domain i.e., for u_{PI} the value of k_p and k_i were chosen as 0.2 and 1000 respectively. At nominal value of V_{DC} at 580V the value of $V_{oc}(\max)$ was 80V.

Waveforms of different dynamic parameters for making two welding joints under different welding conditions are shown in Fig. 3. In arc welding occasional shorting of electrode with weld pool is considered as disturbance. Here, arc or load disappears, but current continues to flow through the electrode. Arc needs to be regained when shorting gets removed. It is clear from Fig. 3a, that while welding at 100A, in spite of overshoot and undershoot during the disturbances, there was no problem in making a good joint. Similarly, as shown in Fig. 3b, there was no problem of arc stability when welding was performed at higher current level 200A.

Fig. 3: There was no problem in welding, but, overshoot and undershoot were noticed in PI control ($k_p = 0.2$ and $k_i = 1000$): a) at small current of 100A, and, b) when current was increased to 200A. [$C_{DC} = 55 \mu F$; $L_1 = 50 \mu H$]

III. STC BASED DESIGN AND CONCEPT VALIDATION

Gains in STC are decided based on the process disturbance, it needs the process or load to be clearly modelled. The arc is nonlinear. In stable welding arc zone, arc voltage V_L written is,

$$V_L = V_0 + E h_a + R_a I_a \quad (11)$$

E (V/m) is field intensity of arc, V_0 is anode and cathode drop together, h_a is arc length and R_a is arc resistance.

Considering error e (6) as sliding surface, the objectives of SOSMC are to meet following conditions,

$$e = I_{ref} - I_a = 0 \quad \text{and} \quad \dot{e} = 0 \quad (12)$$

The derivative of sliding surface e for control of CC arc is,

$$\begin{aligned} \dot{e} &= -\dot{I}_a = -\frac{1}{L_1} V_{oc} + \frac{1}{L_1} V_L \\ &= -\frac{1}{L_1} k_1 u_1 + \frac{1}{L_1} (V_0 + E h_a + R_a I_a) \\ &= -\frac{1}{L_1} k_1 u_1 + \frac{1}{L_1} \psi \end{aligned} \quad (13)$$

The sliding surface can be steered into sliding manifold using STC, with following control function in power domain,

$$V_{oc} = k_1 u_{STC} = \rho_0 |e|^{0.5} \text{sgn}(e) + \rho_1 \int \text{sgn}(e) d\tau \quad (14)$$

Where ρ_0 and ρ_1 are gains. The input u_{STC} in control circuit is

$$u_{STC} = \rho_{01} |e|^{0.5} \text{sgn}(e) + \rho_{11} \int \text{sgn}(e) d\tau \quad (15)$$

The closed loop system with STC (14) is,

$$\dot{e} = -\frac{\rho_0}{L_1} |e|^{0.5} \text{sgn}(e) - \frac{\rho_1}{L_1} \int \text{sgn}(e) d\tau + \frac{1}{L_1} \psi \quad (16)$$

Let, $Z = -\frac{\rho_1}{L_1} \int \text{sgn}(e) d\tau + \frac{1}{L_1} \psi$ then,

$$\dot{Z} = -\frac{\rho_1}{L_1} \text{sgn}(e) + \frac{1}{L_1} \dot{\psi} \quad (17)$$

$$\text{Where, } \psi = V_L \text{ and } \dot{\psi} = R_a \dot{I}_a \quad (18)$$

Suppose, the perturbation terms ψ and $\dot{\psi}$ are bounded by,

$$\delta_1 \sqrt{|e|} \geq \psi \text{ and } \delta_2 \geq \dot{\psi} \quad (19)$$

Compared to PI control, the approach for gain selection in STC is different. It is ignorant about the parameters of the inverter e.g., frequency f_s , etc. For calculation of gains ρ_0 and ρ_1 the information on maximum value of matched disturbances ψ and $\dot{\psi}$ are needed, they depend on process behavior. For CC process, at rated current of, say, 400A, the value of V_L is 36 V. Consider, conservatively the current error at 1.0 % of rated current i.e., at 4 A. and the value of R_a at 0.025Ω . The bounded values of ψ and $\dot{\psi}$ depend on the dynamics of the process, their values for different value of L_1 are listed in Table I. For a dynamical system where $\dot{\psi}$ is dominant, compared to the approach given in [22], [21] would yield smaller gains. To cater different transient conditions, respective set of gains are listed in Table I. It is clear that for design of controller for complete arc welding process [7], [17] the value of each gain would be large. To study the efficacy of STC, simple CC arc is considered where, even, $\dot{\psi}$ is ignored where gains are least (see Table I).

TABLE I
STC GAINS FOR DIFFERENT LEVELS OF DISTURBANCE LEVEL

	STC gains at different value of dI_a/dt or dI_{ref}/dt A/s					
	0	50000	50000	100000	300000	≥ 500000
δ_1	18	18	18	18	18	18
δ_2	0	1250	1250	2500	7500	≥ 12500
ρ_0	40	40	46	46	46	46
ρ_{01}	5	5	5.7	5.7	5.7	5.7
ρ_1	25k	104k	45.4k	92k	415k	$\geq 950k$
ρ_{11}	3125	13k	5690	11.5k	51.8k	$\geq 120k$

A. SOSMC Implementation

Here, to validate the STC function, simplest dynamical situation of arc load was considered, the gains were selected based on minimum disturbance. The controller was analyzed for CC arc welding process where the disturbance $\dot{\psi}$ was zero. As shown in Table I, for V_{oc} (14), the calculated gains were ρ_0 (=40) and ρ_1 (> 25000). To generate u_{STC} (15), gains were changed to ρ_{01} (=5.0) and ρ_{11} (= 3125). To validate high gain STC practically, hardware-based implementation was preferred for prompt corrective action where input (15) was simplified as,

$$\begin{aligned} u_{STC} &= \rho_{01} k_e e + \rho_{11} \int \text{sgn}(e) d\tau \\ &= \rho_{01} (e) e + \rho_{11} \int \text{sgn}(e) d\tau \end{aligned} \quad (20)$$

Where, $k_e = 1/\sqrt{e}$

The error specific gains are listed in Table II. The hardware circuit for u_{STC} [19] was based on analog switch (see Fig. 5).

TABLE II
ERROR SENSITIVE EFFECTIVE GAIN VALUE

Sliding surface e , V	k_e	Switch active	$\rho_{01}(e)$	$\rho_{01}(e)$
≥ 2.0 ($\geq 200A$)	0.7	All disabled	$0.7\rho_{01}$	3.5
1.0 (100A)	1.0	S1	$1.0\rho_{01}$	5.0
0.5 (50A)	1.4	S2	$1.4\rho_{01}$	7.0
0.20 (20A)	2.24	S3	$2.24\rho_{01}$	11.2
0.04 (4A)	5.0	S4	$5.0\rho_{01}$	25.0

Fig. 4: Hardware circuit for implementation of the super twisting control

B. Experimental Results

The STC controller needs to meet the desired performance in presence of wide range disturbances with multiple origin, like,

1. Sharp change in load or process disturbance
2. Sharp change in reference or process requirement
3. Increase in ripple content in error, particularly, at large current, caused by drooping nature of $L1$, and or,
4. Ripple content at source V_{DC} – due to availability of metallized polypropylene capacitors with high RMS current handling capability the value of DC link capacitance C_{DC} ($\mu F/A$) is drastically reduced.

Multiple welding experiments were conducted using STC functions. For concept validation, values of C_{DC} and $L1$ were varied. Initially, their values ($C_{DC} = 55\mu F$ and $L1 = 50\mu H$) were kept same like PI control of Fig. 3. Fig. 5 shows waveforms of electrical variables for two different welding conditions. Control of current while welding manually at 100A (see Fig. 5a) during sudden load change was robust, there was no overshoot or undershoot. The control performance was repeated when the load current was 200A (see Fig. 5b). When compared with PI control (see Fig. 3), performance of STC was definitely superior, but the ripple content in u_{STC} was more than u_{PI} . Control response was robust (see Fig. 6a & b) and ripple content in u_{STC} was reduced when $L1$ was increased to 75 μH . It was also noticed that the ripple content in u_{STC} further reduced when the value of C_{DC} was increased to 110 μF . Increase in value of C_{DC} and $L1$ would increase cost and reduce power density, might not be acceptable in price sensitive competitive world.

Fig. 5: STC could achieve robust control of current, but ripple in u_{STC} was large at, a) current 100 A, and, b) current 200A. [$L1 = 50\mu H$ and $C_{DC} = 55\mu F$]

Fig. 6: Current control was robust and ripple in u_{STC} was reduced when $L1$ was increased to 75 μH for welding at, a) 100 A, and, b) 200A. [$C_{DC} = 55\mu F$]

IV. PRODUCT VALIDATION OF STC BASED FBC

Often, even practical validation of a concept might not be a good indicator for its adoption and diffusion into (certain) industry or commercial domains; at least, for SOSMC, it did not happen that way. Apart from demonstrating superior control performance against process disturbances, efficacy of STC, as a superior control function, could be judged if the concept mandatorily meets or does not violate following criteria of a product deemed to be commercialized:

1. High degree of availability of the controller
2. Long service life i.e., the reliability estimate
3. Efficiency or power loss characteristics, and,
4. Cost competitiveness or
5. Favorable performance vs cost parameterization

While testing the controller for continuous duty applications in simple CC process, when compared with PI controller's performance, following observations were recorded:

1. STC controller's operating duty cycle (continuous run time in each ten-minute cycle) was much less. Due to frequent thermal shut down, its availability was poor.
2. The power losses in Q1-Q4 and D1-D4 were more
3. PWM controller entered cycle-by-cycle current mode shut-down frequently, and,
4. Prospect of IGBT failure increased, it would make impact on product reliability

To understand the origin of frequent thermal shut-down, separate heatsinks i.e., HS1 for Q1-Q4 and HS2 for D5-D6 were used. It was noticed that the thermal tripping was from HS1. It

meant that there was increase in power loss in inverter switches. Power loss in inverter could be categorized as follows:

1. Turn-on loss in Q1-Q4
2. Turn-off loss in Q1-Q4
3. Recovery loss in D1-D4, and,
4. Conduction loss in Q1-Q4 and D1-D4

The turn-on power loss of Q1-Q4 depends on current I_a reflected at primary of TR whereas their turn-off loss depends on the value of current at turn-off moment i.e., the peak magnetizing current I_m . The recovery loss of D1-D4 also depends on I_m , plus presence of I_{TR} adds more losses to them. It is important to study the impact of STC on functioning of magnetically coupled inverter.

Fig. 7: Relation between slope of control function and PWM duty cycle

A. Impact of Control Function on Transformer

For design of TR, N87 Grade EE8020 cores were used. The primary to secondary turns ratio ($n_p:n_s:n_s$) was 15:2:2. At V_{DC} of 580V, the value of peak operating flux density B_m for maximum value of d_{pwm} ($=0.8$) was 0.197 T. At 100 °C, the value of saturation flux density B_{sat} was 0.39 T. Among soft ferrites, linearity of B-H (H : magnetomotive force) curve of N87 Grade was found to be superior. To reduce the prospect of volt-second unbalance in core [28], each IGBT pair (Q1, Q4) or (Q2, Q3) was driven by single pulse transformer. For variable d_{pwm} , employing DC blocking capacitor in inverter would not be effective. Though the inverter was protected by cycle-by-cycle current limit, it was effective only at large current. The expression of peak magnetizing current I_m is,

$$I_m = \frac{l_{mag} B_m}{\mu_0 \mu_r n_p} \quad (21)$$

B_m is peak flux density, n_p is number of primary turns of TR, l_{mag} is mean length of core and μ_r is its relative permeability. B_m is directly linked to the control input, like,

$$B_m = \frac{n_s k_1}{4 A_C f_s n_p^2} (u_{SM} \text{ or } u_{PI}) \quad (22)$$

A_C is core area and n_s is number of secondary turns.

For a particular application, constant value of u_{STC} (its slope $m=0$ in a cycle) is desirable because it maintains B_m at constant value. The switch-on duration of (Q1, Q4) or (Q2, Q3) is same at, say, T1 (see Fig. 7) to ensure average flux in core at zero value. If m is other than zero in a cycle, d_{pwm} would be different in negative (say, T2) and positive (T3) half cycles (see Fig. 7, bottom). The flux variation $\Delta\Phi$ over one complete cycle is [26],

$$\Delta\Phi = \frac{V_{DC}(T_2 - T_3)}{n_p} = \frac{V_{DC}}{n_p} \frac{m}{(C-m)} \frac{1}{2f_s} \quad (23)$$

The rate of change of flux density ΔB_v caused by non-zero m is,

$$\Delta B_v = \frac{V_{DC}}{2 A_C n_p} \frac{m}{(C-m)} \quad (24)$$

C is slope of sawtooth waveform with peak at V_{st} (see Fig. 7) and m is slope of u_{PI} or u_{STC} . The time t_{sat} taken for the core to attain the core saturation is $t_{sat} = B_{sat}/\Delta B_v$. In this application, $B_{sat} \approx 2B_m$. For reliable operation of the system the condition $t_{sat} < \tau_s$ (9) needs to be satisfied. If, by design, m is kept comparable to C , swing in B_m in a cycle would be large where non-linearity in B-H loop could play adverse impact. Actually, m is considered small so that the value ΔB_v in one complete cycle is negligible.

By design, the operating point is kept in linear zone and I_m solely depends on d_{pwm} . Its effective swing depends on the operating point in the B-H loop. For persistent non-zero value of m the value of d_{pwm} keeps increasing (see Fig. 8a) to make the operating B-H loop shift towards one particular quadrant (see Fig. 8b) where the B-H curve for a particular half-cycle would be increasingly non-linear, leading to DC bias in core. The value of μ_r (21) would be less in that half cycle, could lead to different value of I_m in positive and negative half-cycles. Fig. 8b shows that with positive m , DC bias in core is introduced where H needed to maintain the flux swing is large i.e., the value of I_m is large. I_m does not contribute to power transfer but causes turn-off loss in Q1-Q4 and recovery losses in D1-D4, its minimum value is desired for reliable performance.

Fig. 8: a) Primary voltage of TR with increasing PWM duty cycle, and, b) it shifts the operating hysteresis loop and creates of DC bias in core

B. Influence of PI Control on Primary Current

In FBC the current at secondary DC side is controlled. But, the actuation for control execution takes place at primary of TR. Ideally, the controller should be designed so that the u_{PI} does not disturb TR in its functioning for voltage translation in isolated manner. It is possible if the average flux in a PWM cycle is negligible. It is decided by the slope m_{PI} of u_{PI} , like,

$$m_{PI} = k_p \frac{de}{dt} + k_i e \approx k_p \frac{de}{dt} = f \left(\frac{de}{dt} \right) \quad (25)$$

Where $de/dt = k_{hs} dI_a/dt$, k_{hs} is constant for current to voltage conversion in hall sensor (10mV/A). Consider transient disturbance at 100 kA/s i.e., de/dt of 1000 V/s, here, with k_p at 1.0 and k_i at 1000, the value of m_{PI} is 1000. Fig. 9a and 9b show waveforms of I_{pri} for two different applications. Uniform waveform of I_{pri} suggested that the secondary control did not disturb the magnetic circuit of TR. The average flux inside the

core was zero. The value of I_m was negligible because the peak value of I_{pri} measured during turn-off at welding current 100A was 14A, and for welding at 200A, it was 27A. Current I_{pri} and I_a were relatable and following relation was established:

$$I_{Q1,Q4} = I_{Q2,Q3} = \frac{n_s}{n_p} I_a \quad (26)$$

With V_{st} of 10V, for 20 kHz inverter the value of C would be 400000, and the value of ΔB_v would be at 0.25% of B_m .

Fig. 9: Magnetic saturation or DC bias in core was absent under PI control, where there was no ripple in u_{pi} while welding at: a) 100A, and, b) 200A. [L1: 50 μ H and C_{DC} : 55 μ F, k_p : 1.0 and k_i : 1000].

C. Influence of STC on I_{pri} when $C_{DC} = 55 \mu$ F and $L1=50\mu$ H

The slope m_{STC} of u_{STC} is complex and quite different from m_{PI} , it is difficult to define its trajectory. It is expressed as,

$$m_{STC} = \left(\frac{\rho_{01}}{2\sqrt{e}} \frac{de}{dt} + \rho_{11} \right) \text{sgn}(e) = f\left(e, \frac{de}{dt}, \text{sgn}(e)\right) \quad (27)$$

For same value of de/dt of 1000 V/s, the maximum value of m_{STC} is 28125. The corrective measure is 28-time faster than PI. It is sensitive near zero error; even ripple in I_a could modulate u_{STC} at fast rate at frequency $2f_s$. The prospect of core saturation is large because the value of ΔB_v is large, at 7.6% of B_m .

Fig. 10: High ripple content in high-gain u_{STC} coupled with nonlinearity in B-H curve of core resulted in DC bias and I_{pri} was more in one half-cycle, at, a) 100A, and, b) 200A.

Fig. 10 shows two sets of waveforms with same value of C_{DC} and $L1$ as used for PI control in Fig. 9. Even in steady state condition, due to large ripple in V_{DC} and, particularly, in I_a resulted large ripple in u_{STC} (see Figs. 10a and 10b). High ripple in u_{STC} resulted DC bias in core and eqn. (26) was violated.

D. Influence of STC on I_{pri} when $L1$ or C_{DC} was increased

In order to reduce the ripple in I_a as well as in u_{STC} , the value of $L1$ and C_{DC} were increased, one at a time. It is clear from Fig. 11 that, in steady state condition, the ripple content in I_a

and u_{STC} were only marginally reduced. Still, as shown in two welding conditions, I_{pri} in each half cycle in both cases were quite different (see Figs. 11a and b). The current in one pair of IGBTs was different from the other and (26) was not met.

Fig. 11: a) While welding at 100A, ripple in u_{STC} was marginally reduced when C_{DC} was increased to 110 μ F, but DC bias in core persisted leading to unequal current in IGBT pairs and, b) when $L1$ was increased to 75 μ H, similar results were achieved while welding at at 200A

Fig. 12: a) When a transient boost in I_{ref} was added to effect anti-sticking, STC resulted sharp change in d_{pwm} within a single PWM cycle, leading to DC bias in core and unequal current in IGBT pairs, and, b) Same situation persisted even when u_{STC} was getting stable after the transient condition, average flux in core per cycle was not zero. [L1: 75 μ H; C_{DC} : 110 μ F].

E. Influence of STC on I_{pri} during Disturbance

Here, to reduce ripple in V_{DC} , C_{DC} was increased to 110 μ F. Two experimental waveforms in Fig. 12 demonstrate the response of STC in presence of sharp transient disturbance for welding at 220A. Here, to create anti-sticking force during short-circuiting of electrode with weld pool ($V_L \approx 0$), transient reference equivalent to de/dt at 3000V/s was added. STC reacted sharply and changed the value of d_{pwm} almost cycle-by-cycle basis (see Fig. 12a). Large peak value of I_{pri} in one pair of IGBT at 60A (220 % of PI) signified that current limit was operational. It resulted increase in conduction and high-frequency losses in inverter. Fig. 12b shows the behavior of I_{pri} when u_{STC} was gradually stabilizing from transient disturbance. The DC bias in core persisted because the operating point was already in nonlinear zone of B-H loop.

V. MODIFIED CONTROL FUNCTION TO BOOST EFFICACY

It was clear that 20 kHz PI controller was reliable but it had overshoot and undershoot. To improve the control response the value of f_s is increased (9). Calculated values of different control parameters (see Table III) get better. Soft-switching topologies [29], [30] are employed to eliminate associated high-frequency losses. Power density of inverter also improves

because the size of TR, L1 and heat sink are reduced.

TABLE III
CAPABILITY OF HIGH FREQUENCY PI BASED FBC

Freq., kHz	20	50	100	150
L1, μH	65	25	14	9
τ_s , μs	500	200	100	70
k_p	0.13	0.112	0.128	0.125
k_i	940	1950	4375	6330
m_{PI}	130	224	640	1250

STC could resolve the control related issues. But, even small disturbance or, ripple in I_a changed u_{STC} instantly resulting increase in power loss in magnetically coupled controller. There were concerns on reliability and availability features. Ideally, to cater wide range need of the process the theoretical value of two gains would be extremely large (see parameters in red ink in Table I). To avoid DC bias in core, respective value of m_{STC} (27) would need unrealistically large value of C and f_s where PI performance would be great (see Table III).

It was noticed that the primary reasons for creation of DC bias in core and associated excess power loss in inverter were:

1. Extremely high gain near zero error (see Table II), and,
2. Existence of ripple in output, or ripple in error.

To make high-gain function reliable, L1 is increased to ensure ripple in e (12) is under acceptable limit, and for soft corrective measure around zero error, following alternate control u_{SMP} [7], [15] of u_{STC} (15) is proposed,

$$u_{SMP} = \rho_{01}e + \rho_{11} \int \text{sgn}(e)dt \quad (28)$$

Fig. 13: Current control was robust when alternate function u_{SMP} with $\rho_{10} = 5.0$ and $\rho_{11} = 3125$ for welding at, a) 100 A, and, b) 200A. [L1: 100 μH]

It has been practically validated in several process applications [7], [19], [31] that (28) strongly meets conditions of (12). Here, like m_{PI} , slope m_{SMP} of u_{SMP} depends on de/dt as,

$$m_{SMP} = \left(\rho_{01} \frac{de}{dt} + \rho_{11} \right) \text{sgn}(e) = f \left(\frac{de}{dt}, \text{sgn}(e) \right) \quad (29)$$

For same value of de/dt of 1000 V/s, the maximum value of m_{SMP} is around 8000. For C at 400000, compared to STC, the value of ΔB_v was small at 2.0% of B_m .

Fig. 13 shows control of current using new function (28) in similar applications (see Fig. 3, 5, 6), control responses were robust. Fig. 14 shows respective primary current I_{pri} . Though, same gains (ρ_{10} : 5.0, ρ_{11} : 3125) of u_{STC} were used, but DC bias in core was negligible and conditions of (26) were satisfied.

Fig. 14: DC bias in core was absent because the ripple in u_{SMP} was negligible, at: a) 100A, and, b) 200A. [L1: 100 μH].

VI. CONCLUSION

This article, for the first time, addressed the critical issues pertaining to the extremely slow pace of adoption and diffusion of widely researched STC concept. It detailed that efficacy of a new concept lied on their effective and efficient utilization to yield definite benefits. It further addressed that, apart from concept validation, commercialization of a product based on a novel idea needed to meet availability and reliability criteria. When the role of control was to directly modulate the duty cycle of PWM based converter, the control function that caused minimum stress to high-frequency power switching devices of a product would be preferred. It elaborated that the process-disturbance based gain selection procedure in STC might not be suitable for vast majority of products that used magnetically coupled power transfer mechanism. The high-gain STC could vary the duty cycle of PWM switches in each half-cycle, it was also highly sensitive to output ripple. Such fast change in duty cycle could inject DC bias in magnetic circuit of transformer that caused excess power loss in inverter. On the other hand, the choice of gains in PI control was more collective. It, not only involved the process dynamics, parameters of power controller were part of the design; smoothing inductance and switching frequency of inverter were part of gain selection.

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