Self-Started Voltage-Source Series-Resonant Converter for High-Power Induction Heating and Melting Applications

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Abstract—An inverter configuration for high-power induction heating and melting applications is presented. The proposed inverter covers loads with quality factors up to 12, while featuring self-starting capabilities. This is achieved by properly distributing the compensated capacitor between the primary and the secondary of the matching transformer. The transient analysis of the configuration defines the maximum ratio of the primary and secondary capacitors, in order to assure self starting. On the other hand, the steady-state analysis defines the minimum ratio to limit the operating flux level of the matching transformer to a safe level. This paper identifies the sources of self-starting failures in the standard configuration and presents the transient and steady-state analyses toward a systematic procedure of components design for the proposed topology. Experimental results on a laboratory prototype to prove the theoretical considerations are also included.

Index Terms—Induction heating, self-starting inverter, series-resonant inverter.

I. INTRODUCTION

Static power converters are applied extensively in the area of induction heating and melting processes. In order to cover high-power applications, these converters use traditionally high-power low-turnoff-time asymmetrical switches, such as asymmetrical silicon-controlled rectifiers (ASCR’s), where self commutation would be an asset. The topologies based on a voltage-source series-resonant inverter present near unity input power factor, a wide range of load power control by adjusting the load frequency, and overall system simplicity [1]–[7].

However, voltage-source series-resonant inverters are not widely used in the industry, due to the starting problems associated with the transformer-coupled induction heating or melting load, specifically, a starting failure when self commutation is not achieved. This leads to a destructive short circuit across the dc source [8].

This paper identifies the sources of self-starting failures in the standard configuration. Moreover, a theoretical analysis is used to identify the region for safe starting. The analysis shows that this region is limited to loads with rated quality factors \( Q_0 \) lower than 7.

This work finally presents an SCR-based self-started voltage-source series-resonant inverter. This is achieved by properly distributing the compensated capacitor between the primary and secondary of the matching transformer. The transient and steady-state analyses are presented. It is shown in the paper that the proposed inverter configuration extends the load quality factor range up to 12. Experimental results on a laboratory prototype are used to verify the theoretical results.

II. VOLTAGE-SOURCE SERIES-RESONANT INVERTER DESCRIPTION

A. Power Topology

The power topology is depicted in Fig. 1. The inverter uses SCR’s \( Q_1, \cdots, Q_4 \) with antiparallel diodes \( D_1, \cdots, D_4 \) to feed back negative currents to the dc source. Only two of the SCR’s (either \( Q_1 \) and \( Q_3 \), or \( Q_2 \) and \( Q_4 \)) are on at a given time. In standard implementations, the primary capacitor \( C_p \) is very large \( (C_p \to \infty) \); therefore, for any analysis purpose, it can be considered a short circuit. A transformer is used to match the load voltage. The inductor \( L' \) represents the coil inductance and allows the operation of the converter in the series-resonant mode. Finally, the load is modeled by a resistor \( R_L \). Although the quality factor of the load \( (Q_0) \) can vary during the electroheat process, it can be considered constant during one inverter switching period.

B. Control Issues

Proper operation of the SCR’s should generate a square voltage waveform across the ac terminals of the inverter. Therefore, the power topology can be modeled by the equivalent circuits depicted in Fig. 2 as the circuit operates in
Fig. 2. Equivalent circuits of the voltage-source series-resonant inverter power topology. (a) Interval I—Q1 and Q3 are ON. (b) Interval II—Q2 and Q4 are ON.

Fig. 3. Power topology steady-state waveforms. (a) Inverter ac voltage \(v_p\). (b) Inverter ac current \(i_p\). (c) Magnetizing current \(i_m\). (d) Capacitor voltage \(v_C\). Simulated waveforms for \(V = 600\) V, \(P_o = 100\) kW, \(Q_s = 10\), \(x_m = 50\) p.u., \(f_s = 300\) Hz, and \(\omega_s = 0.75\omega_r\).

III. STARTING PROBLEMS OF THE VOLTAGE-SOURCE SERIES-RESONANT INVERTER

Although in steady state the self commutation of the power switches is assured by operating the inverter at switching continuous mode. The load power control is carried out by adjusting the voltage frequency, which is maintained between 0.5–1.0 p.u. (1 p.u. is the resonant frequency). This guarantees a load with an overall capacitive behavior, where the inverter output current leads the voltage and self commutation is always assured in steady state. Simulated steady-state waveforms are given in Fig. 3 for \(Q_s = 0.7\) and \(Q_s = 0.75\) p.u., \(\omega_s = 0.75\omega_r\), \(\omega_s = 0.75\omega_r\), 

Fig. 4. Inverter ac current \(i_p\) during starting for two starting load quality factors \((Q_s)\). (a) \(Q_s = 3.5\). (b) \(Q_s = 7\). Simulated waveforms for \(V = 600\) V, \(P_o = 100\) kW, \(f_s = 300\) Hz, and \(x_m = 30\) p.u.

frequencies ranging from 0.5 to 1.0 p.u., proper commutation during the starting may not be achieved. In fact, a successful starting depends upon three factors. These are as follows: 1) the value of the load quality factor at starting \((Q_s)\); 2) the starting frequency; and 3) the latching current of the SCR’s.

A. Commutation of the Switches

During transient (as the starting) and static operating conditions, for successful commutation of the inverter switches (for example, \(Q_1\) and \(Q_3\), or \(Q_2\) and \(Q_4\)), the output current of the inverter should go to zero and the antiparallel diodes \((D_1\) and \(D_3\), or \(D_2\) and \(D_4\)) must conduct for at least the minimum specified turnoff time for the switches. Otherwise, the triggering of the switches \((Q_2\) and \(Q_4\), or \(Q_1\) and \(Q_3\)) will cause a destructive short circuit across the dc source.

Fig. 4 shows the inverter starting current for two starting load quality factors \((Q_s)\) assuming that the SCR’s \(Q_2\) and \(Q_4\), or \(Q_1\) and \(Q_3\) will remain on and, if \(Q_2\) and \(Q_4\) are switched on, a destructive short circuit across the dc source is created.

In this paper, to identify the existence of a negative inverter current during the first half cycle, a relation between the starting load quality factor \(Q_s\), the rated load quality factor \(Q_o\), and the transformer impedance \(x_m\) is derived. Specifically, a commutation index \((M_c)\) will be defined.

By using the model given in Fig. 2(a), the starting inverter output current \((i_o)\) is found to be

\[
i_o = \frac{V}{L_m} t + V \sqrt{\frac{C_o}{L_o}} \exp\left\{\frac{-R_o}{2L_o} t\right\} \sin\left\{\frac{1}{\sqrt{L_o C_o}} t\right\} A \quad (1)
\]

where the rated load quality factor \(Q_o\) should satisfy

\[
Q_o = \frac{1}{R_o} \sqrt{\frac{L_o}{C_o}} \gg \frac{1}{2} \quad (2)
\]

and \(L_m\) is the transformer magnetizing inductance, \(V\) is the dc-bus voltage, \(C_o\), \(L_o\) are the resonant components, and \(R_o\) is used to define the starting quality factor. If \(\omega_r\) is the load...
angular resonant frequency \(\omega_r^2 = 1/C_o L_o\), the load quality factor at starting is

\[
Q_s = \frac{1}{R_s} \sqrt{\frac{L_o}{C_o}}.
\]

The expression for the inverter output current (1) can be written as

\[
i_o = \frac{V}{L_m} t + V \sqrt{\frac{C_o}{L_o}} \exp\left\{ -\frac{1}{2Q_s} \omega_r t \right\} \sin\{\omega_r t\} A.
\]

Choosing the input voltage \(V\), the rated load resistance \(R_o\), and the resonant frequency \(\omega_r\) of the load as the base values, (4) can be written in the following per-unit form:

\[
i_o = \frac{\omega_r t}{x_m} + \frac{1}{Q_o} \exp\left\{ -\frac{1}{2Q_s} \omega_r t \right\} \sin\{\omega_r t\} \text{ p.u.}
\]

where \(Q_o\) is the load quality factor at rated load, and \(x_m\) is the p.u. matching transformer magnetizing impedance. The boundary condition is defined when the inverter output current is equal to zero \((i_o = 0)\) at its first minimum \([\omega_r t_{\text{min}}\text{ in Fig. 4(b)}\). Fortunately, the first minimum of \(i_o\) occurs at \(\omega_r t_{\text{min}} = \pi\sqrt{2}\) which is approximately constant. This is valid for values of \(x_m\) ranging from 30 to 200 p.u. and for values of \(Q_o\) and \(Q_s\) ranging from 3 to 30. Therefore, using (5), the aforementioned boundary condition can be expressed as

\[
x_m \geq Q_o \exp\left\{ \frac{\sqrt{2}\pi}{2Q_s} \right\} \frac{\sqrt{2}\pi}{\sin\{\sqrt{2}\pi\}} \text{ p.u.}
\]

The commutation index is defined as

\[
M_c = \frac{x_m}{Q_o} \exp\left\{ -\frac{\sqrt{2}\pi}{2Q_s} \right\} \frac{\sin\{\sqrt{2}\pi\}}{\sqrt{2}\pi}.
\]

By inspecting the boundary condition (6) and the commutation index definition (7), it is found that for the inverter system to be self starting, \(M_c\) should be greater or equal to 1 under all starting conditions. For instance, the case depicted in Fig. 4(a) has a \(Q_o = 5\) and \(x_m = 50\) p.u., which yields \(M_c = 1,285 > 1\). The case depicted in Fig. 4(b) has a \(Q_o = 10\) and \(x_m = 50\) p.u.; hence, \(M_c = 0,702 < 1\). This confirms the lack of negative inverter output current in the case shown in Fig. 4(b).

In many applications, the starting load quality factor is 0.707 times the rated load quality factor. Using this simplification, Fig. 5 plots the minimum magnetizing inductance as a function of the rated load quality factor using (7). The line in Fig. 5 represents \(M_c = 1\); therefore, the area above the line (where \(M_c > 1\)) represents the region where a negative inverter output current during the starting cycle is assured. It can be seen that, for a magnetizing inductance \(x_m = 50\) p.u., the maximum rated load quality factor is approximately 7.

B. The Starting Frequency \(\omega_s\)

If the commutation index \(M_c\) is greater than 1, the inverter output current will be negative during the starting cycle. However, it is negative just for a limited amount of time \([t_{qs} \text{ in Fig. 4(a)}]\). Therefore, in order to avoid a destructive short circuit, only within this period should the next pair of SCR’s be switched on. Since, for practical values of \(Q_o\), \(x_m\), and \(Q_s\), the minimum of the inverter output current occurs at \(\omega_r t_{\text{min}} = \pi\sqrt{2}\) [Fig. 4(a)], the starting switching frequency \(\omega_s\) can safely be chosen to be \(\omega_s = \omega_r / \sqrt{2}\).

C. The Latching Current

SCR’s are switched on by applying to their gate a current pulse that features a given width \((t_p = 10\mu s)\). For the switch to stay on, the current through the valve must reach the latching current \((i_{\text{latch}} = 400\text{ mA})\). In this paper, to identify if the SCR’s current reaches the latching current during the starting, a latching index \(M_l\) will be defined.

By using (5) and assuming \(p\) parallel inverters feeding the same load, the current through an SCR at the end of the gate pulse \((i_{tp}\text{ p.u.})\) can be written in p.u. as

\[
i_{tp} = \frac{1}{pQ_o} \sin\{\omega_r t_p\} \text{ p.u.}
\]

where the magnetizing current and the exponential terms in (5) have been neglected. This is because \(t_p\) is very small. Since the base current is \(V/R_o\), the current through an SCR at the end of the gate pulse in amps is

\[
i_{tp} = \frac{\pi^2 P_o}{8 V pQ_o} \frac{1}{\sin\{\omega_r t_p\}} A.
\]

The boundary condition is defined when the current through an SCR at the end of the gate pulse is equal to the latching current. Therefore,

\[
i_{\text{latch}} \leq \frac{\pi^2 P_o}{8 V pQ_o} \frac{1}{\sin\{\omega_r t_p\}} A.
\]

The latching index is defined as

\[
M_l = \frac{\pi^2 P_o}{8 V pQ_o} \frac{1}{\sin\{\omega_r t_p\}}.
\]
Fig. 6. Equivalent circuits of the split compensated capacitor series-resonant voltage-source inverter power topology. (a) Interval I—Q₁ and Q₄ are ON. (b) Interval II—Q₂ and Q₄ are ON.

Fig. 7. Inverter ac current \( i_{ac} \) during starting for two split capacitor ratios \( x \). (a) \( x = 10 \), (b) \( x = 7 \). Simulated waveforms for \( V = 600 \) V, \( P = 100 \) kW, \( f = 500 \) Hz, \( Qₐ = 10 \), \( x = 0.05 \) p.u., and \( x_m = 50 \) p.u.

By inspecting the boundary condition (10) for the latching current and the latching index definition (11), it is found that \( M_{th} \) should be greater or equal to 1 under all starting conditions. For instance, the case depicted in Fig. 4(b) has a \( Qₐ = 10 \), \( P = 100 \) kW, \( V = 600 \) V, \( f = 500 \) Hz, which yields \( M_{th} = 1.615 > 1 \) and, thereby, the SCR’s will remain on after the first gate pulses. Equation (11) is used to find the maximum rated load quality factor \( Qₐ \) for a given set of conditions. As a result, load quality factors that are lower than 20 will always provide enough current through the SCR’s to ensure proper latching.

IV. PROPOSED SPLIT COMPENSATED SERIES-RESONANT VOLTAGE SERIES INVERTER

The previous section showed that load quality factors greater than 7 cannot be covered by the standard configuration \( (Cᵢ → ∞) \) and a magnetizing current of 2%. For \( Qₐ > 7 \), a commutation failure occurs during starting, due to a lack of a negative inverter ac current. This paper proposes to distribute the compensated capacitor between the primary and secondary of the matching transformer. This approach allows one to increase the maximum rated load quality factor \( Qₐ \) for a given set of conditions. As a result, load quality factors that are lower than 20 will always provide enough current through the SCR’s to ensure proper latching.

A. Transient Analysis

During the starting of the inverter, the SCR’s \( Q₁ \) and \( Q₃ \) are assumed to be switched on; therefore, the equivalent circuit shown in Fig. 6(a) can be used to analyze the starting inverter ac current. The angular resonant frequency \( \omega_r \), rated load quality factor \( Qₐ \), starting load quality factor \( Q₈ \), leakage inductance/resonant inductance \( x \), p.u. transformer magnetizing inductance \( x_m \), and ratio of the primary and secondary capacitors \( xₘ \) are defined as

\[
\omega_r = \frac{1}{\sqrt{(L_{th} + L_o)C_C/C₀}} \quad (12)
\]

\[
Qₐ = \frac{1}{R_C} \sqrt{\frac{L_{th} + L_o}{C_C/C₀}} \quad (13)
\]

\[
Q₈ = \frac{1}{R_C} \sqrt{\frac{L_{th} + L_o}{C_C/C₀}} \quad (14)
\]

\[
x = \frac{L_{th}}{L_C} \quad (15)
\]

\[
x_m = \frac{\omega_r L_m}{R_C} \quad (16)
\]

\[
x_i = \frac{C_i}{C₀} \quad (17)
\]

respectively. Using (12)–(17) and since \( 1 + xₘ \approx 1 \), the state variable model during the starting in p.u. is given by

\[
\frac{d}{dt} \begin{bmatrix}
U_{C_C} \\
U_{C₀} \\
U_{Cₘ} \\
U_{Cₘ} \\
U_{Cₘ}
\end{bmatrix} = \begin{bmatrix}
0 & \omega_r Qₐ x_i & 0 & 0 & 0 \\
-\frac{x_m + Qₐ xₘ}{x_m + Q₈ xₘ} & 0 & \frac{\omega_r Qₐ x_i}{x_m + Q₈ xₘ} & 0 & 0 \\
0 & -\frac{\omega_r Qₐ x_i}{x_m + Q₈ xₘ} & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & 0
\end{bmatrix} + \begin{bmatrix}
0 \\
x_m + Q₈ xₘ \\
x_m + Q₈ xₘ \\
x_m + Q₈ xₘ \\
x_m + Q₈ xₘ
\end{bmatrix} V. \quad (18)
\]

Fig. 7 shows the starting inverter ac current \( i_{ac} \) for two values of \( xₘ \) and the same conditions as Fig. 4(b). It can be seen that, for both values of \( xₘ \), the converter allows self starting. The period when the current is negative during the starting cycle is denoted by \( t_{qs} \) and plotted in Fig. 8 for different values of \( Qₐ \) as a function of \( xₘ \). The period \( t_{qs} \) becomes, in fact, the maximum available turnoff time at starting of the inverter. The most negative inverter ac current occurs at \( \omega_r t_{min} = \pi \sqrt{2} \) [Fig. 7(a)], which is approximately constant for a wide range of values of \( Qₐ \) and \( xₘ \). Therefore,
the safest starting switching frequency ($\omega_s$) becomes

$$\omega_s = 2\pi f_s = 2\pi \frac{1}{2f_{\text{min}}} = 2\pi \frac{\omega_r}{2\sqrt{2}\pi} = \frac{\omega_r}{\sqrt{2}} = 0.707\omega_r. \quad (19)$$

Under these conditions, the available turnoff time of the inverter at the starting becomes $t_{q0}/2$. Fig. 8 shows, indeed, that for a given $Q_o$, there is a maximum $x_i$ in order to achieve a minimum turnoff time. For instance, for $Q_o = 10$ (case shown in Fig. 7), a maximum $x_i = 12$ should be used.

### B. Steady-State Analysis

The output power of the inverter is effectively controlled in the frequency range of 0.5–1 p.u.; therefore, the analysis of the inverter is carried out in the continuous mode. Since the output voltage is defined in this mode (square waveform), a Fourier series method is used. The p.u. equivalent circuit of Fig. 6 for an $n$th harmonic is shown in Fig. 9, where

$$0.5 \text{ p.u.} < \omega = \frac{\omega_0}{\omega_r} < 1.0 \text{ p.u.} \quad (20)$$

is the operating frequency and $Q_{0n}$, $x_k$, and $x_l$ are defined by (13), (15), and (17), respectively. The expressions for voltages and currents are derived and later used in determining the steady-state flux level of the matching transformer. To do so, the following equivalent impedances are defined to simplify the expressions: $Z_{en}$ is the $n$th harmonic impedance at the output terminal of the inverter; $Z_{ym}$ is the $n$th harmonic series impedance in the primary of the transformer; $Z_{ym1}$ is the $n$th harmonic impedance looking into the magnetizing inductance of the transformer; $Z_{en}$ is the $n$th harmonic magnetizing impedance; and $Z_{en}$ is the $n$th harmonic series impedance in the secondary of the transformer. From Fig. 9, the respective expressions are

$$Z_{en} = Z_{ym} + Z_{ym1} = |Z_{en}|\max\theta_{en} \quad (21)$$

$$Z_{ym} = jQ_{0n}\left\{\frac{n\omega x_l}{1+x_l} - \frac{1}{n\omega + 1} + x_l\right\} = |Z_{ym}|\max\theta_{ym} \quad (22)$$

$$Z_{ym1} = z_{ym}z_{en} = |Z_{ym1}|\max\theta_{ym1} \quad (23)$$

$$Z_{en} = 1 + jQ_{0n}\left\{\frac{n\omega - 1}{1+x_l} - \frac{1}{n\omega + 1} + x_l\right\} = |Z_{en}|\max\theta_{en}. \quad (25)$$

The output voltage can be expressed in the form of a Fourier series as follows:

$$v_o(t) = \sum_{n=1,3,5,\ldots}^{\infty} \frac{4}{n\pi} \sin(n\omega_o t) \text{ p.u.} \quad (26)$$

By using the output voltage expression (26) and the total impedance (21), the inverter output ac current becomes

$$i_o(t) = \sum_{n=1,3,5,\ldots}^{\infty} \frac{4}{n\pi |Z_{en}|} \sin(n\omega_o t - \theta_{en}) \text{ p.u.} \quad (27)$$

The equation for the voltage across the magnetizing inductance is given by

$$v_m(t) = \sum_{n=1,3,5,\ldots}^{\infty} \frac{4}{n\pi |Z_{en}|} \sin(n\omega_o t + \theta_{en} - \theta_{on}) \text{ p.u.} \quad (28)$$

The magnetizing current of the inverter is given by the following equation:

$$i_m(t) = \sum_{n=1,3,5,\ldots}^{\infty} \frac{4}{n\pi |Z_{en}|} \sin\left(n\omega_o t + \theta_{en} - \theta_{on} - \frac{\pi}{2}\right). \quad (29)$$

The expression for the load current is given by

$$i_L(t) = \sum_{n=1,3,5,\ldots}^{\infty} \frac{4}{n\pi |Z_{en}|} \sin\left(n\omega_o t + \theta_{en} - \theta_{on}\right). \quad (30)$$

The voltage across the primary capacitor is given by the following equation:

$$v_{C_p}(t) = \sum_{n=1,3,5,\ldots}^{\infty} \frac{-4Q_{0n}}{n^2\pi\omega(x_l^2 + 1)} |Z_{en}| \cos\left(n\omega_o t - \theta_{en}\right) \text{ p.u.} \quad (31)$$

The voltage across the secondary capacitor is given by

$$v_{C_s}(t) = \sum_{n=1,3,5,\ldots}^{\infty} \frac{-4i(x_l)Q_{0n}}{n^2\pi\omega(x_l^2 + 1)} |Z_{en}| \frac{|Z_{en}|}{|Z_{en}|} \cdot \cos\left(n\omega_o t + \theta_{en} - \theta_{on}\right) \text{ p.u.} \quad (32)$$

Using (26)–(32), the time variation of the inverter voltages and currents for $Q_o = 10$, $\omega = 0.75$, and $x_l = 8.5$ are derived and shown in Fig. 10.

Computer simulation shows that the maximum flux level, which is proportional to the peak of the magnetizing current, increases as the ratio $x_l$ decreases. Thus, the rated flux level is defined for $C_i \rightarrow \infty$ and $\omega = 1.0$. In this case, the square voltage generated by the inverter is directly applied to the terminals of the matching transformer and a triangular...
magnetizing current [Fig. 3(c)] is thus generated. The peak value of the current \( (i_{\text{m, max}}) \) for \( C_i \rightarrow \infty \) and \( \omega = 1.0 \) is given by

\[
i_{\text{m, max}} = \frac{\pi}{2} \frac{V}{\omega L_m} = \frac{\pi}{2} \frac{V}{x_m R_o} \text{ A.} \tag{33}\]

Therefore, the rated flux density \( (B_o) \) can be expressed as

\[
B_o = \mu \frac{N}{T} i_{\text{m, max}} = \mu \frac{N}{T} \frac{V}{2 x_m R_o} \text{ T.} \tag{34}\]

where \( \mu \) is the permeability of the core, \( N \) is the number of turns, and \( L \) is the length of the core. In practical implementations, these parameters are chosen to obtain a rated flux density equal to a fraction of the maximum flux density of the material \( (B_{\text{max}}) \). For instance, \( B_o = 0.8 \cdot B_{\text{max}} \). Thus, the normalized flux density for an arbitrary operating condition

\[
B / B_{\text{max}} = 0.8 \frac{x_m i_{\text{m, max}}}{B_{\text{max}} \text{ p.u.}}. \tag{35}\]

where \( i_{\text{m, max}} \) is the peak magnetizing current in p.u. at a given operating condition. The flux density expression (35) is evaluated, using (29), and the results are plotted in Fig. 11. Fig. 11 shows, indeed, that for a given \( Q_o \), there is a minimum \( x_i \) in order to maintain the flux density below the maximum.

C. Operating Region

The transient and steady-state requirements of the inverter impose the opposite limits on the selection of the ratio \( x_i \): 1) for the self starting of the inverter \( C_i \rightarrow 0 \) (Fig. 8) and 2) for the minimum and load independent rating of the transformer \( C_i \rightarrow \infty \) (Fig. 11). Therefore, the ratio \( x_i \) should be selected in such a way that: 1) the self starting of the inverter is achieved and 2) the rated flux level of the transformer is not exceeded. Fig. 12 shows a region of operation for the inverter. For a given load \( Q_o \), the value of \( x_i \) should lie within the shaded region.

Fig. 12 has been built up using Fig. 8 (for the top limit) and Fig. 11 (for the bottom limit). Specifically, for a given \( Q_o \), the value of \( x_i \) can be read from Fig. 8 such that \( t_{qs}/2 = 0 \).
D. Available Turnoff Time for the Switch

Fig. 10(b) shows the available steady-state turnoff time \( (t_{q0}) \) for the switches for a given operating condition \( (\omega = 0.75) \). Let the output current of the inverter \( (i_o) \), given by (27), go to zero at \( \omega_o t = \epsilon \). Therefore,

\[
\sum_{n=1}^{\infty} \frac{4}{n\pi |Z_{en}|} \sin(n\epsilon - \theta_{en}) = 0. \quad (36)
\]

The value of \( \epsilon \) can be found using an iterative method. The available turnoff time can, therefore, be determined as follows:

\[
t_{q0} = \frac{\pi - \epsilon}{\omega_o} \text{ s.} \quad (37)
\]

Let \( 1 \text{ p.u. time} = 2\pi/\omega_p \), therefore, the p.u. turnoff time in steady state is given by

\[
t_{q0} = \frac{\pi - \epsilon}{2\pi\omega_p} \text{ p.u.} \quad (38)
\]

Fig. 13 shows the available turnoff time in steady state as a function of the operating frequency for different load conditions. It can be clearly seen that, in order to assure a minimum turnoff time, the maximum operating frequency...


must be lower than 1.0. In practice, the maximum operating frequency will be a function of the speed of the switches.

V. EXPERIMENTAL VERIFICATIONS

To verify the behavior and analysis of the inverter, a prototype inverter was built and tested in the laboratory. Fig. 14(a) shows the transient current waveforms of the inverter for \( x_1 = 22.5 \). This waveform shows that the inverter has the commutation failure at starting. Fig. 14(b) shows the transient current waveforms for \( x_1 = 8.5 \). It is evident from this figure that the split compensated series-resonant inverter has the self-starting capability.

Fig. 15 shows the steady-state experimental waveforms for the inverter for \( Q_o = 10 \), \( f_o = 8.4 \) kHz, \( f_s = 11.2 \) kHz, and \( x_1 = 8.5 \). The theoretical waveforms for the same conditions are shown in Fig. 10. Table I gives the p.u. comparisons of the experimental and theoretical results. A close agreement of the results, therefore, verifies the theoretical behavior of the inverter.

VI. CONCLUSION

A voltage-source series-resonant inverter for high-power induction heating and melting applications has been presented in this paper. The starting failure of the inverter has been identified and an improved converter configuration presented. Detailed transient and steady-state analyses have been given, and it has been shown that the converter has self-starting capability without exceeding the maximum flux level of the matching transformer. The proposed inverter has been found best suited for high-power melting loads with a quality factor of up to 12.

REFERENCES


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\[\begin{array}{|c|c|c|}
\hline
\text{V}_{ci} & 140 \text{ V} & 3.82 \text{ p.u.} \\hline
\text{V}_{co} & 45 \text{ V} & 0.41 \text{ p.u.} \\hline
i_o & 12.5 \text{ A} & 0.68 \text{ p.u.} \\hline
i_p & 40 \text{ A} & 0.727 \text{ p.u.} \\hline
\tau_{qo} & 26 \mu\text{s} & 0.279 \text{ p.u.} \\hline
\end{array}\]