A CURRENT-FED RESONANT INVERTER FOR
HIGH-FREQUENCY
HIGH-POWER APPLICATIONS

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ABSTRACT

A new current-fed series resonant inverter suitable for high frequency high power applications is introduced and
analyzed. Steady-state characteristics of the inverter are obtained
using a normalized state-space model. Switch implementation,
advantages, limitations and design considerations of the inverter
are discussed.

1. INTRODUCTION

Resonant inverters have been used since the days of vacuum tubes in many industrial applications such as spacecraft
ac power supplies, induction heating and melting, ultrasonic cleaners and radio transmitters at low frequency
[1],[2]. Recent developments in semiconductor power devices such as thyristors, power transistors and power
MOSFETS have led to inverters with better performance than with vacuum tubes.

These inverters are switched power circuits that generate nearly sinusoidal output voltages and currents from a DC
supply. Typically switches connect either a constant voltage or current across a resonant load. Systematic methodologies for
deriving resonant inverter topologies are discussed in [1]. The major advantages of resonant inverters are the lower switching
losses and waveforms with reduced harmonics content as compared to square wave inverters.

The use of these inverters is mostly in high power applications. The thyristor has been used more than other
switching devices. There are two main reasons for this. First, the thyristor can handle high power when compared with other
switching devices. Secondly, since the thyristor turns off when its current goes to zero, the principle of natural commutation on
which many resonant inverters work turns off the thyristor without the need for an additional commutation circuit.

The analysis of these inverters has mostly focused on voltage-fed types. The use of duality may result in new circuits
with different practical characteristics [3],[4]. It is not always the case that the input characteristics of the inverter are those of a
voltage source; in some cases they are that of a current source. This paper analyzes a current-fed series load, thyristor inverter
obtained by the application of duality rules to a voltage-fed parallel inverter [2]. Closed form solutions using a normalized
second order model of the inverter are obtained. The switches stresses and the resonant components' ratings are obtained for
various controlling parameters such as the switching-to-resonant frequency ratio and the Q-factor of the resonant tank circuit.
These curves are used in designing the inverter for different values of input current, output voltage, power and frequency.

2. DUALITY CONSIDERATIONS

Duality has shown and it will continue to be a very useful link between power converters. It can yield new
converter topologies as well as significantly improve understanding of the relationship between converters and their
equivalent models [1],[5]. The type of switch and the modes of operation can differ for two dual converters. This can be
illustrated using the four resonant inverter topologies shown in Fig. 1. Topologies A and B are voltage-fed
parallel inverters with bidirectional current switches. Topologies C and D are their respective current-fed series inverters.
Topology A is introduced by Mapham [6], and its dual topology C is introduced by Kassakian [3],[4]. Topology B is introduced
by Kasturi and its dual topology D is introduced in this paper. The following comparisons give some insight into the
similarities and differences that dual topologies can exhibit.

(i) The switches in topologies A and B have to carry bidirectional currents, while the switches in topologies C and D
have to support bidirectional voltages. Thyristors can be used in topologies A and B with no additional commutation circuit
needed. In contrast, thyristors cannot be used in topology C without an additional commutation circuit, whereas they can be
used in topology D without any additional commutation circuit, and they can be turned off by forced commutation. Also, since a
thyristor is a bidirectional switch, only two switches are needed in topology D, as compared to four switches in topologies A, B
and C.

(ii) Topology A, which is a divided-inductor version of topology B, can only operate at frequencies less than the
resonant frequency, but it has better switch stresses (di/dt, dv/dt) and lower switching losses than topology B. Also, the
thyristors used in topology A are changed to transistors in its dual topology C. Since thyristors usually handle a larger
amount of power, then topology A has a power handling capability larger than that of topology C. Also, topology A uses
two inductors and one capacitor, while topology C uses two capacitors and one inductor. Capacitors can in general be lighter
and closer to ideal than inductors, so one might argue that topology C is better in that respect.

(iii) If a transformer is used to isolate the load, then in topologies A and B a transformer with larger magnetizing
inductance and small leakage is needed. On the other hand, a transformer with high leakage inductance is needed to isolate the
load in topologies C and D.

In conclusion the above comparison has demonstrated the power of duality in determining the relationships among dual
converters. Depending on the load characteristics, one has a choice between four inverters that are topologically the same but
different in characteristics.
3. STEADY STATE CHARACTERISTICS

Fig. 2 shows a half-bridge version of a the current-fed series inverter which has been shown earlier in Fig. 1D. The circuit is the dual of the voltage fed parallel inverter of Fig. 1B. Kassakian introduced a variation of this circuit [3], using a current-dependent switches closing scheme and split resonant capacitors. He stated that an independent scheme using the same circuit would be impractical due to high transient switch current. In this paper we investigate that scheme but using a single resonant capacitor.

The current source can be implemented by a voltage source in series with a large inductor. Switches S1 and S2 are operated at a frequency f_s and 50% duty cycle, generating a quasi-sinusoidal voltage across the resonant load. The power supplied to the load is controlled through variation of the switching frequency.

The state equations governing the circuit operation are given below:

\[
\begin{align*}
\frac{d i_L(t)}{dt} &= \left( \frac{-R}{L} \quad 1 \quad 0 \right) \left( \frac{i_L(t)}{i(t)} \right) + \left( 0 \right) I_s \\
\frac{d v_C(t)}{dt} &= -\frac{1}{C} \left( \frac{v_C(t)}{v(t)} \right) + \left( 1 \right) I_s
\end{align*}
\]  

\( (1) \)

The + and - signs in (1) correspond to the first half-cycle (S1 OFF) and the second half-cycle (S2 OFF) respectively. The following basis are used for normalizing the state equations in (1):

\[
\begin{align*}
I_{BASE} &= \text{input current} = I_s \\
Z_{BASE} &= \text{characteristic impedance} Z_0 = \sqrt{L/C} \\
t_{BASE} &= \text{switching period} = T_s = \frac{1}{f_s}
\end{align*}
\]  

(2) (3) (4)

The normalized state equations are then:

\[
\frac{dx(t)}{dt_{BN}} = 2\pi f_s \begin{pmatrix} -1/Q & 1 \\ -1 & 0 \end{pmatrix} x(t) \approx \pm 2\pi f_s \begin{pmatrix} 0 \\ 1 \end{pmatrix}
\]

\( (5) \)

where the state vector is given by:

\[
x(t) = \begin{pmatrix} x_1 \\ x_2 \end{pmatrix} = \begin{pmatrix} i_L(t)/I_s \\ v_C(t)/I_s Z_0 \end{pmatrix}
\]

\( (6) \)
and $f_0$ is the resonant frequency of the tank:

$$f_0 = \frac{1}{2\pi\sqrt{LC}}$$

and the $Q$ of the circuit is given by:

$$Q = \frac{2\pi}{R}$$

The characteristic equation associated with (5) gives a sinusoidal response when the $Q$ of the circuit is less than 0.5.

Solution of the state equations over a full cycle is:

$$X(t_N) = F(t_N)X(0) + G(t_N)\text{ for } 0 < t_N < 1/2$$

$$X(t_N) = F(t_N)X(0) - G(t_N)\text{ for } 1/2 < t_N < 1$$

(9)

where $t_N=0$ is the beginning of any cycle and $t_N=1$ is the end of it. The matrices $F(t_N)$ and $G(t_N)$ for a fixed $Q$ and $f_0/f_o$ are given by:

$$F(t_N) = \begin{pmatrix} a_0 & e^{\alpha t}e^{\phi t}
\end{pmatrix}$$

(10)

$$G(t_N) = \begin{pmatrix} 1 - \frac{a_0}{a_0}e^{\beta t}e^{\phi t}
\end{pmatrix}$$

(11)

where the parameters $a_0, a_0, \phi, \alpha$ and $\beta$ are given by:

$$a_0 = \sqrt{1 - \left(\frac{1}{2Q}\right)^2}$$

(12)

The cyclic steady state is characterized by:

$$X(1) = X(0) = X$$

(16)

Because of symmetry of the waveforms, the initial state can be written in terms of the state at half the cycle:

$$X(0) = -X(1/2)$$

(17)

Substituting (16) and (17) in (9) and solving for $X$ gives:

$$X = -(I + F(1/2))^{-1}G(1/2)$$

(18)

Since equations (10) and (11) are given for fixed values of $Q$ and $f_0/f_o$, they describe, together with (18), the steady state operation of the circuit for different values of $Q$ and $f_0/f_o$.

Fig. 3 shows typical steady state waveforms for relatively high values of $Q$. As $Q$ decreases, these waveforms begin to look more triangular than sinusoidal due to the effect of the load damping. Similarly, as the switching frequency approaches the resonant frequency, the amplitude of the waveforms increases up to very large values.

The normalized power $P_{N}$ delivered to the load, can be expressed in terms of normalized state variables as below:

$$P_{N}(Q,f_0/f_o) = \frac{1}{Q} \int x_1^2(t_N) \, dt_N$$

(19)
Numerical integration of equation (19) for various values of $f_s/f_o$ and Q is shown graphically in Fig. 4. It can be seen that power converted increases as Q increases and as the ratio $f_s/f_o$ approaches unity. High values of Q result in high energy conversion levels for switching frequencies close to the resonant frequency; however, the bandwidth available for a proper control gets reduced.

Another important characteristic of the circuit is the component ratings. Besides the stresses the resonant components must withstand, stresses on the switches must also be considered. Graphs showing the variation of the normalized inductor current $x_1$ and capacitor voltage $x_2$ peaks for different values of the switching-to-resonant-frequency ratio $f_s/f_o$ and Q are shown in Fig. 5 and 6. These peaks are very high when switching at frequencies near resonance and when Q is high. A close look at Figs. 4, 5 and 6 will reveal that for the region in which energy conversion can be adequately controlled, peak voltages and currents are not only relatively high, but also subject to large variations. The peak voltage on the switch is the same as the peak voltage in the capacitor; on the other hand, the switch current is a constant equal to 2 units of $I_p$.

The switches $S_1$ and $S_2$ operate alternately. Implementation of these switches and how they are driven depends on the range of operating frequencies. Since the use of thyristors offer the advantage of high power handling, it is convenient to establish the frequency boundaries in which it is possible to use them. In general, this is possible as long as the voltage across the switch at the time it is turned on is positive. The condition to be satisfied at $t_N = t/2$ is then:

$$
\sin \beta \left( 1 - \sin \alpha \left( \frac{\sin \phi}{\phi} + \frac{\sin \phi}{\omega_o} \right) \right) < 0 \tag{20}
$$

This condition shows that the circuit has to be switched below resonance. However, the upper limit of the frequency range depends on Q. As Q decreases this limit will also decrease.
5. REFERENCES.


Fig. 6. Normalized peaks of the capacitor voltage

4. CONCLUSIONS

A current-fed series inverter suitable for high-frequency and high-power applications has been introduced and analyzed. The steady state characteristics are obtained as a function of different controlling parameters such as the normalized load conductance Q and the switching-to-resonant-frequency ratio. Normalized curves for the output power and the component ratings have been obtained. The switch implementation and its limits have been determined.