Analysis of the Current-Fed Push-Pull Parallel Resonant Inverter Implemented with Unidirectional Switches

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Abstract— A modified Current-Fed Push-Pull Parallel Resonant Inverter (CFPPRI), that includes diodes in series with the switches, was analyzed and tested by simulation and experimentally. The modified power source is capable of producing a low THD output voltage within a relatively wide frequency range. A comparison between the modified topology and the basic CFPPRI, in terms of output signal quality and efficiency was carried out. It was found that the modified inverter is more efficient and that it produces an output signal that is less distorted as the running frequency deviates from resonance. Operation with a capacitive load in an off-resonance mode was demonstrated by driving a rotary piezoelectric motor by the modified inverter.

I. INTRODUCTION

The favorable drive signal for a number of capacitive loads is a sinusoidal waveform. For example, piezoelectric devices and in particular piezoelectric motors need to be driven by a high frequency sinusoidal waveform [1]. This is required since the optimal performance is obtained when the drive is of low harmonics contents. The Current-Fed Push-Pull Parallel Resonant Inverter (CFPPRI) was shown earlier [1] to be a useful sinusoidal voltage source suitable for driving loads that are sensitive to the harmonics contents of the drive voltage. However, the basic CFPPRI operation is optimal only when the switching frequency is equal to the resonant frequency. Any deviation from the optimal point will result in efficiency reduction and output signal distortion. Recently, a modified CFPPRI that includes diodes in series with the power MOSFETs was described [2]. This implementation follows the earlier realization that was common, before MOSFETs were available, when applying unidirectional devices such as Thyratrons [3]. One of the features of this topology variation is a lower THD of the output voltage when the switching frequency deviates from the resonant frequency. The objective of this study was to quantize the differences between the basic CFPPRI and the modified one, in terms of THD and efficiency.

II. THE CURRENT-FED PUSH-PULL PARALLEL RESONANT INVERTER (CFPPRI)

The basic CFPPRI topology (Fig. 1) includes two transistors (Q_1, Q_2) in a push-pull configuration that drive a parallel resonant network that is fed by a series inductor L_{in}. The load (capacitive in this example) is connected to a secondary winding that allows amplitude adjustment by the turns ratio ($n_2/2n_1$) as well as isolation. The resonant frequency (f_r) of the inverter will be [4]:

$$f_r = \frac{1}{2\pi\sqrt{L_m \Sigma C_r}}\sqrt{1+\frac{L_m}{n^2 L_{in}}} \qquad (1)$$

where L_m is the secondary inductance, $\Sigma C_r = (C_1 n^2 + C_2 + C_L)$ is the total capacitance reflected to the secondary, C_1 is the primary capacitance, C_2 is the secondary capacitance, C_L is the load capacitance and n is the turns ratio ($n_2/2n_1$).

When the gate voltage of Q_1 (V_{GS1}, Fig. 1) goes high and the gate voltage of Q_2 (V_{GS2}, Fig. 1) goes low, the drain voltage of Q1, (V_{DS1}, Fig. 1) is forced to zero and the tank voltage (V_{DS2} in this case) starts to rise in a sinusoidal shape. If this voltage reaches back to zero at the next switching instance, the transistors will operate under Zero Voltage Switching (ZVS) conditions. This situation corresponds to the case f_s=f_r (Fig. 2a). If the switching frequency is lower than the resonant frequency (Fig. 2b), the body diode of Q_2 will start conducting and the operation will include a boost period [4], which will cause distortion of the output voltage and increase conduction losses. If the switching frequency is higher than the resonant frequency, the operation will be under hard switching conditions (Fig. 2c). That is, the resonant capacitor, C_1, will be short circuited by the transistors at the beginning of each half cycle. An exact matching of the resonant to the switching frequency is not possible, in practice, due to load variability, cable connection etc.

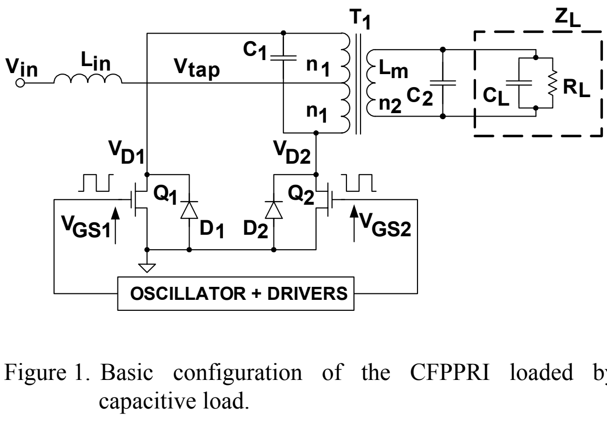

Figure 1. Basic configuration of the CFPPRI loaded by a capacitive load.

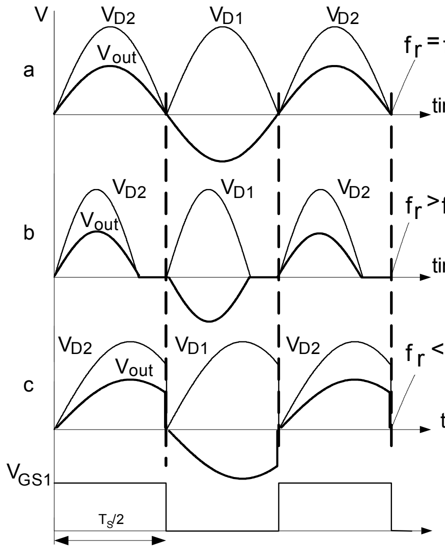

Figure 2. Gate voltage (V_{GS1}), drain voltages (V_D) and output voltage (V_{out}) when the gate-drive frequency matches the resonant frequency (a) and when it is lower (b) or higher (c) than the resonant frequency.

Consequently, in practical cases one would expect the output of the CFPPRI to be distorted due to off resonance operation. One possible solution to alleviate the problems associated with the frequencies mismatch is to automatically adjust the switching frequency to be equal to the resonant frequency [5]. Another possible solution to the problem is to add diode in series with the switches [2]. This option is examined in followings.

III. OPERATION AND ANALYSIS OF THE CFPPRI WITH UNIDIRECTIONAL SWITCHES (CFPPRI-US)

The modified CFPPRI-US includes two series diodes to the transistors (Fig. 3), making the power switches unidirectional. In the followings, the operation of the CFPPRI-US will be described by considering the first half cycle when V_{GS1} goes high and V_{GS2} goes low.

When the switching frequency is equal to the resonant frequency (Fig. 4a), the CFPPRI-US operates under ZVS, much like the operation of the basic CFPPRI. In the case that the resonant frequency is higher then the switching frequency (Fig. 4b) the tank voltage V_{S2}, will reach zero at the instant t_1, and D_4 will be reversed biased. This prevents the "boost" type operation typical of the basic CFPPRI. This period ends at t_2 when a new one-half cycle begins. When the resonant frequency is lower than the switching frequency (Fig. 4c), the tank voltage does not reach zero during the end of each half cycle. Consequently, at the first part of the half cycle (t_0-t_1) when Q_1 conducts, the tank voltage is negative but D_4 blocks the conduction of Q_2. This prevents the shorting of the resonant capacitance via the switches, as is the case in the basic CFPPRI. At instant t_1, the drain voltage becomes positive, D_4 start conducting and Q_2 will carry the output voltage.

It should be noted that except in case of f_s=f_r, the power switches are switched on while the voltage on them is not zero. That is, under hard switching conditions. However, the switching losses in this case are much lower compared to those in the off-resonant modes of the basic CFPPRI, where the resonant capacitance is short-circuited by the transistors. *Output Signal Quality*: Considering the operation of the CFPPRI-US as described above, and the fact that only one switch is conducting at each half cycle, it can be modeled by a parallel resonant network that is fed by squarewave current source (Fig. 5).

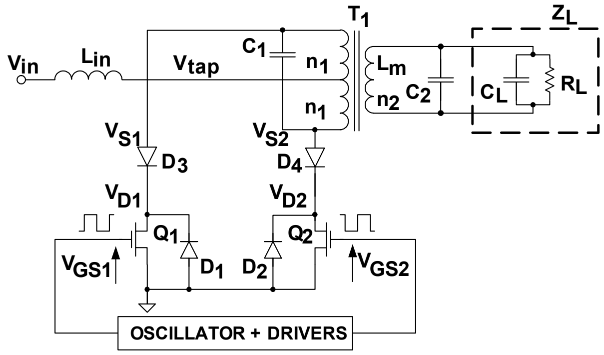

Figure 3. The CFPPRI with diodes in series to the power switches.

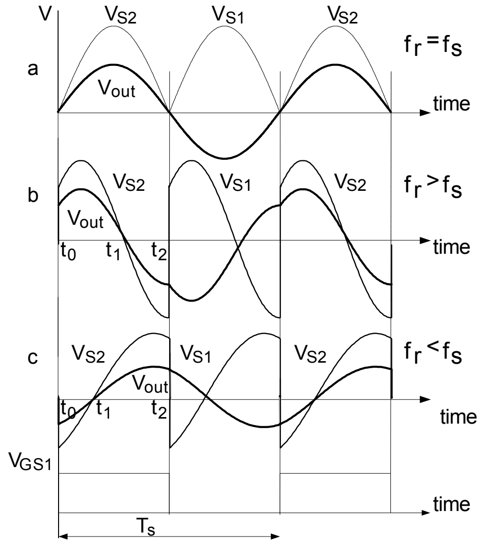

Figure 4. Gate voltage (V_{GS1}), switch voltages (V_S) and output voltage (V_{out}) when the gate-drive frequency matches the resonant frequency (a) and when it is lower (b) or higher (c) than the resonant frequency.

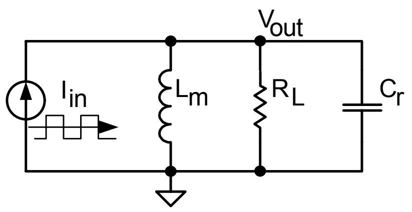

Figure 5. Simplified CFPPRI circuit.

This approximation assumes that L_{in} is large enough so that the current through it can be considered constant over one switching cycle (or that the ripple current of L_{in} is small). The equivalent current source can be expressed by its harmonic content as:

$$I(t)=\frac{4I_{DC}}{\pi}\sum_{k=0}^{\infty}\frac{1}{2k+1}\sin(2k+1)\omega_r t \tag{2}$$

where I_{DC} is the DC current of L_{in}, $\omega_r=2\pi f_r$ is the resonant frequency and k is the harmonic's index number.

The resonant network impedance is given by:

$$Z(s)=\frac{s\omega_r}{s^2+s\frac{\omega_r}{Q}+\omega_r{}^2} \tag{3}$$

where 's' is the Laplace operator, $Q=R_L/Z_r$ is the quality factor, RL is the load resistance and $Z_r=\sqrt{L_r/C_r}$ is the typical impedance of the resonant network.

Under steady state conditions $s=jk\omega_r$, and the normalized absolute value of the network impedance for a given harmonic is (normalization was done by considering Z_r as the base value):

$$|Z(k\Omega)|=\left|\frac{Z(jk\omega_r)}{Z_r}\right|=\frac{1}{\sqrt{\frac{1}{Q^2}+\left(k\Omega-\frac{1}{k\Omega}\right)^2}} \tag{4}$$

where Ω is the ratio between the resonant frequency and the running frequency ($\Omega=\omega/\omega_r$).

The output voltage of the k^{th} harmonic can be found by:

$$V_{out}(k)=I_k\cdot Z_k \tag{5}$$

and the rms value of the output voltage is:

$$V_{out\_rms}=\sqrt{\sum_{k=0}^{\infty}\frac{V_{out}(k)^2}{2}}\ . \tag{6}$$

In order to estimate the output signal, a numerical calculation (using Matlab) of the Total Harmonic Distortion of the output voltage was carried out by:

$$THD_{V_{out}}[\%]=\frac{\sqrt{\sum_{k=2}^{\infty}V_{out}(k)^2}}{V_{out}(k=1)} \tag{7}$$

Fig. 6 shows the output voltage THD of the CFPPRI-US as a function of the deviation from the resonant frequency. The results were compared to the output voltage THD of the basic CFPPRI.

It should be noted that the variation rate, dV_{out}/df_s , of the output voltage depends on the quality factor, Q, of the resonant network, as can be observed from Eqs. 3 and 4. However, it was found that the THD of the output signal THD is largely independent on Q. The latter is valid as long as s Q>1. This will also ensure proper operation of the inverter.

Efficiency: The equivalent circuit of the CFPPRI-US (Fig. 5) describes a current source that feeds a resonant network. In practice, the inverter is fed by a DC voltage source via a relatively large input inductance L_{in} (Fig. 3). The input power of the inverter can be expressed as:

$$P_{in}=I_{DC}V_{in} \tag{8}$$

where V_{in} is the average input voltage.

The output power is given by:

$$P_{out}=\frac{V_{out\_rms}{}^2}{R_L}\ . \tag{9}$$

The losses of the CFPPRI-US, ΔP, are considered to be mainly due to the conduction losses of the power switches (Q_1, Q_2, Fig. 3) and the series diodes (D_3, D_4, Fig. 3). Neglecting the input ripple current, the losses for one switching cycle are given by:

$$\Delta P=2\left[\left(I_{DC}\sqrt{\frac{1}{2}}\right)^2R_{on}+I_{DC}\frac{1}{2}V_{on}\right]=I_{DC}{}^2R_{on}+I_{DC}V_{on} \tag{10}$$

where R_{on} is the switches on resistance and V_{on} is the diodes forward voltage.

The efficiency of the CFPPRI-US is then expressed as :

$$\eta=\frac{P_{out}}{P_{out}+\Delta P}\ . \tag{11}$$

Fig. 7 shows the calculated efficiency of the modified inverter obtained by (10 and (11) as a function of the deviation from the resonant frequency. The results were compared with experimental measurements of the efficiency of the CFPPRI-US and to the basic CFPPRI topology.

RMS Output Voltage Gain: Combining (8), (9) and adding the losses obtained in (10), yields the power equilibrium of the inverter.

$$I_{DC}V_{in}=\frac{V_{out\_rms}{}^2}{R_L}+\Delta P\ . \tag{12}$$

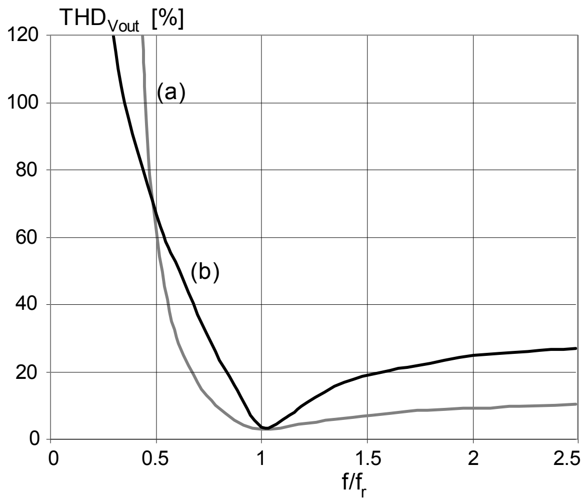

Figure 6. Total Harmonic Distortion of output voltage as a function of the deviation from the resonant frequency. (a) CFPPRI-US (numerical calculation). (b) The basic CFPPRI (SPICE simulation).

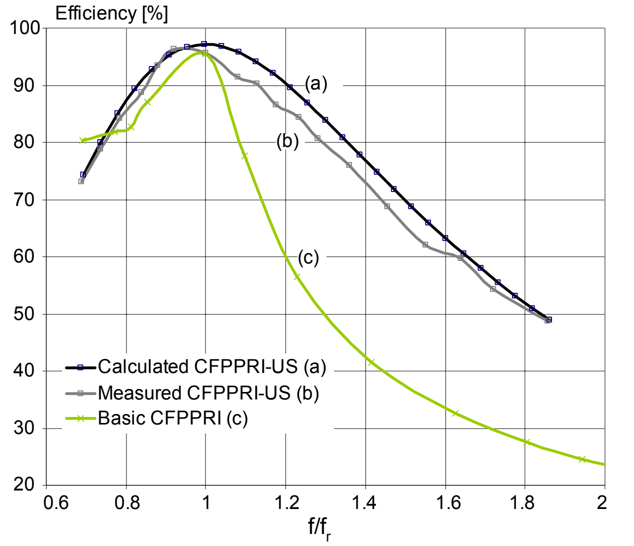

Figure 7. Efficency as a function of the deviation from resonant frequency. (a) Calculated efficiency of the CFPPRI-US. (b) Measured efficiency of the CFPPRI-US. (c) Measured efficiency of the basic CFPPRI topology.

Isolating (V_{out_rms}/V_{in}) yields the output to input voltage gain of the CFPPRI-US:

$$\frac{V_{out\_rms}}{V_{in}} = \frac{1}{\dfrac{V_{out\_rms}}{I_{DC}R_L} + \dfrac{\Delta P}{V_{out\_rms}}} . \qquad (13)$$

Eqs. 5, 6 and 13 imply that the value of the output voltage depends on the running frequency of the inverter. Namely, the output voltage increases as the switching frequency deviates from resonance. However, the operation range is limited since as the running frequency deviates from the resonant frequency, high losses are introduced (Fig. 7) and will consequently reduce the output to input voltage gain. Additional voltage gain reduction occurs at frequencies that are lower than half of the resonant frequency. Over this region, the output signal will be highly distorted with respect to the resonance.

Output Voltage Gain, first harmonic approximation: When working around resonance, the modified inverter produces an output signal that has low harmonic content (Fig. 6). This signal can be approximated to a sinusoidal waveform by setting k=1 in (6).

$$V_{out\_1rms} = \frac{V_{out}(k=1)}{\sqrt{2}} = \frac{I_1 \cdot Z_1}{\sqrt{2}} . \qquad (14)$$

Assuming no losses and inserting (2), (4) and (14) into (13), yields the sinusoidal output to input voltage gain of the CFPPRI-US:

$$\frac{V_{out\_1rms}}{V_{in}} = \sqrt{1 + Q^2\left(\Omega - \frac{1}{\Omega}\right)^2} . \qquad (15)$$

IV. Experimental Verification

A prototype CFPPRI-US, built, simulated and tested experimentally. The parameters of the experimental resonant unit were: Load capacitance: C_2 = 1nF. Transformer characteristics were: Core type: RM10; Magnetizing inductance (secondary side): L_m = 1.9mH; Turns ratio: $n_2{:}n_{11} = n_2{:}n_{12} = 8$. Typical waveforms of the experimental unit, driving a rotary piezoelectric motor is given in Fig. 8.

Fig. 9 shows the calculated output to input voltage gain as a function of the deviation from the resonant frequency (numerical calculations were carried out by Matlab): (a) RMS gain calculation by Eq. 13 and (b) First harmonic approximation gain (eq. 15). The results were compared to: (c) Ideal cycle-by-cycle SPICE simulation, (d) Cycle-by-cycle SPICE with losses and (e) Experimental measurements.

V. Discussion and Conclusions

The CFPPRI-US topology was shown to be able to drive capacitive loads with a drive signal that has low harmonic content within wide frequency range. Although ZVS is not obtained in the off-resonant modes, the power losses are reduced compared to the basic CFPPRI. In addition, the output voltage THD of the modified power source was found to be considerably smaller than the THD of the basic topology.

The output voltage gain has its minimum value at resonance and increases, as the running frequency deviates from the resonant frequency. This could be beneficial in applications where the voltage needs to be controlled such as dimming control. In practice however, operation is limited from half to twice the resonant frequency, due to losses

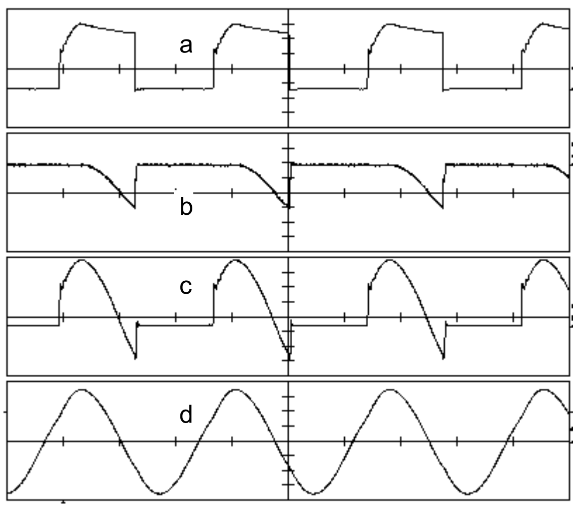

Figure 8. Experimental results. The inverter is loaded by a rotary piezoelectric motor (below resonance operation, f/f_r = 0.65). (a) V_{D1} (10V/div). (b) V_{D3} (20V/div). (c) $V_{D3}+V_{D1}$ (10V/div) and (d) output voltage (50V/div). Horizontal scale (5µs/div).

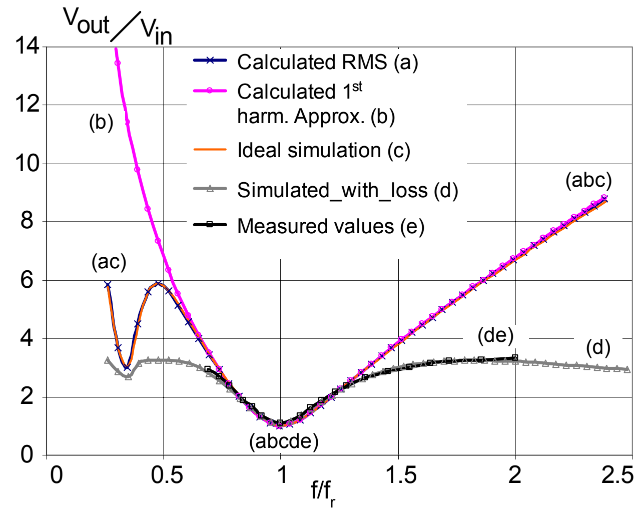

Figure 9. Measured, simulated and calculated output gain of the CFPPRI-US as a function of the deviation from the resonant frequency.

which cause reduction of the output voltage and consequently reduce the output to input voltage gain.

Another limitation to the output voltage gain was found to be the distortion of the output signal that increases as the switching frequency deviates form the resonance.

The validity of the analysis based on the first harmonic approximation was found to be limited to a narrow operation range, around the resonant frequency, since this approximation assumes a sinusoidal output voltage. At frequencies far from resonance, the output THD will increase and the assumption will no longer applies.

The output THD of the CFPPRI-US in the operation range of 0.7 of the resonance and above was found to be relatively small, in the region of 10% to 20%, probably sufficient in many applications.

The comparison between the CFPPRI and the CFPPRI-US reveals that the latter could be a better choice if the frequency of operation is expected to shift from the resonance frequency.

ACKNOWLEDGMENT

This research was supported by THE ISRAEL SCIENCE FOUNDATION (grant No. 113/02) and by the Paul Ivanier Center for Robotics and Production management.

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