Simulation of a 10 MW Series Loaded Resonant Converter

Peter Bartal

July 31, 2013

1 Introduction

The present document summarizes the results of the simulation study of a SLR with a rated power of 10 MW.

The documentation comes with simulation models built in PLECS.

2 Simplifying Assumptions

The following simplifying assumptions have been made:

- ideal semiconductors
- ideal transformer
- lossless resonant tank components

3 Calculations

It is assumed that the three-phase generator has an output line voltage of $V_{gen} = 690\,\mathrm{V}$ (AC RMS). The uncontrolled diode rectifier fed from this generator will have as output voltage:

$$V_{gen_{DC}} = \frac{3\sqrt{2}}{\pi} V_{gen} = 1.35 V_{gen} \approx 930\,\mathrm{V}\ \text{(DC)} \tag{1}$$

Ideally, this voltage will be pure, ripple-free DC (as will be assumed subsequently) and feeds a full-bridge inverter with IGBTs. The inverter will produce a high frequency square-wave voltage ($V_{in_{sq}}$) to drive the resonant tank. The switching frequency is chosen to be $f_s = 1\,\mathrm{kHz}$ To simplify the analysis, only the fundamental of the square wave voltage $V_{in_{sq}}$ will be accounted for. The peak value of the fundamental of the square-wave will be:

$$\hat{V}_{in_1} = \frac{4}{\pi} V_{gen_{DC}} \approx 1185\,\mathrm{V}\ \text{(AC peak)} \tag{2}$$

The RMS value of the fundamental can be calculated as:

$$V_{in_1} = \frac{\hat{V}_{in_1}}{\sqrt{2}} = \frac{2\sqrt{2}}{\pi} V_{gen_{DC}} \approx 840\,\mathrm{V}\ \text{(AC RMS)} \tag{3}$$

At resonance the LC-tank has zero impedance, which results (ideally) in zero voltage drop. However, the voltage across the inductor and the capacitor are far from zero, although their sum is (they are shifted by 180°, but have the same magnitude). It is also known for a fact that this voltage stress is maximal at resonance and can reach prohibitive values. To reduce the stress on the insulation, the converter will need to operate at frequencies below (or above) resonance. The rated operating point has to be chosen at such switching frequency. The difference between the switching frequency (f_s) and the resonant frequency (f_0) will have an influence on the quality factor of the resonant circuit (Q). For a larger difference in frequency Q will be smaller, due to the smaller bandwidth of the resonant tank and vice-versa. The departure from resonance also implies that there will be a non-zero voltage drop across the resonant tank, which will now consume (or produce) reactive power that needs to be kept within bounds. The design procedure laid out in the following will take this fact into account.

As a simplification, throughout this document it will be assumed that only the fundamental plays a noteworthy role in the energy transmission. In reality, the harmonics may be neglected only if the quality factor is sufficiently large, implying a narrow bandwidth of the resonant tank.

The resonant tank will operate in inductive, that is sub-resonant mode. The voltage across the tank will be calculated such as to produce a $\Delta = 5\,\%$ decrease in magnitude of the fundamental of the transformer primary voltage compared to resonance conditions: $V_R = (1 - \Delta) \times V_{in_1} = 0.95 \times V_{in_1} = 798\,\mathrm{V}$. The voltage vector of the resonant tank is perpendicular on the voltage of the transormer primary, thus:

$$V_{tank} = \sqrt{V_{in_1}^2 - V_R^2} = V_{in_1}\sqrt{1 - (1 - \Delta)^2} \tag{4}$$

With the output power known ($P_{out} = 10\,\mathrm{MW}$), the equivalent load resistance, as perceived by the resonant tank can be determined:

$$R_e = \frac{V_R^2}{P_{out}} = \frac{(1 - \Delta)^2 V_{in_1}^2}{P_{out}} \tag{5}$$

accordingly, the total reactance of the tank can be calculated as:

$$X_{tank} = \frac{V_{tank}}{I_{in_1}} = \frac{V_{in_1}^2 (1 - \Delta)\sqrt{1 - (1 - \Delta)^2}}{P_{out}} = \frac{V_{in_1}^2 (1 - \Delta)\sqrt{\Delta (2 - \Delta)}}{P_{out}} \tag{6}$$

Once the switching frequency and the resonant frequency are chosen, it is possible to set up a system of two equations to find the values of L and C:

$$\begin{cases} \omega_s L - \frac{1}{\omega_s C} = X_{tank} \\ \omega_0 L - \frac{1}{\omega_0 C} = 0 \end{cases} \tag{7}$$

For the current application, with $f_s = 1000\,\mathrm{Hz}$ and $f_0 = 1050\,\mathrm{Hz}$, the following values for the resonant tank components were found: $L = 137.97\,\mathrm{\mu H}$ and $C = 179.26\,\mathrm{\mu F}$.

The transformer winding ratio can be found after calculating the transformer secondary voltage, if the voltage at the output has to be $V_{MVDC} = 70\,\mathrm{kV}$:

$$n = \frac{V_{R_2}}{V_R} = \frac{\frac{2\sqrt{2}}{\pi} V_{MVDC}}{V_R} \approx \frac{0.9 V_{MVDC}}{V_R} \approx 79 \tag{8}$$

In case a resistive load is used, smoothing capacitor may be needed at the output of the converter. Its value will be chosen such as to obtain 1 % voltage ripple (peak-to-peak). The general formula for the voltage ripple is:

$$\Delta V = \frac{\Delta Q}{C_F} \tag{9}$$

The change in electrical charge of the capacitor (ΔQ) is the product of the charging (discharging) current and the charging (discharging) interval:

$$\Delta Q = \int_{\frac{\pi}{4}}^{\frac{3\pi}{4}} i(\omega t) d(\omega t) - \frac{\pi}{2} I_{DC} = \frac{4 - \pi}{2} I_{DC} \tag{10}$$

Since a charging takes 1/4 of the resonance frequency, the capacitance is:

$$C_F = \frac{\Delta Q}{\Delta V} = \frac{(4 - \pi)\sqrt{\dfrac{P_{out}}{R}}}{8 f_0 \Delta V} \approx 2.192 \times 10^{-5}\,\mathrm{F} \tag{11}$$

If the smoothing capacitor is to be split in two series capacitors (i.e. to be able to ground the centerpoint of the DC output), the value of each capacitor will be twice the one just calculated: $C_{F_s} = 4.384 \times 10^{-5}\,\mathrm{F}$.

The simulation was performed with the load modelled as the Thévenin equivalent of a voltage source, since the substation boosting the MVDC to HVDC will work with voltage control. In this case the capacitors at the output of the converter represent the capacitance of the MVDC cable.

The frequency can be expressed as a function of the load current. The starting point is the equation of the voltage drop across the resonant tank:

$$I = \frac{V_{tank}}{Z_{tank}} \tag{12}$$

After replacing V_{tank} and Z_{tank}:

$$I = \frac{\omega_s C \sqrt{V_{in_1}^2 - V_R^2}}{\omega^2 LC - 1} \tag{13}$$

Assuming that the input and the output voltage are constant, the expression also represents the relationship between frequency and power:

$$f_{1,2} = \left| \frac{C\sqrt{V_{in_1}^2 - V_R^2} \pm \sqrt{C^2\left(V_{in_1}^2 - V_R^2\right) + 4LCI^2}}{4\pi LCI} \right| \tag{14}$$

The two solutions are for sub- $(-)$ and super-resonant $(+)$ operation, respectively. The formula is suited for a simple feedforward control.

4 State Space Model of the System

The system is nonlinear, yet there is the possibility of linearizing it, which largely simplifies the control design procedure. In order to achieve this goal, the rasonant tank will be represented in $d-q$ coordinates. Again, only the fundamental will be taken into account. By superposition, effect of the harmonics may be modelled.

Starting from the equation:

$$V_{in_1}^2 = V_R^2 + V_{tank}^2 \tag{15}$$

and by denoting the angle between V_R and the d axis to be θ_R, the following decomposition can be found:

$$\begin{cases} V_{in_d} = V_R \cos\theta_R + V_{tank}\cos\left(\frac{\pi}{2} + \theta_R\right) \\ V_{in_q} = V_R \sin\theta_R + V_{tank}\sin\left(\frac{\pi}{2} + \theta_R\right) \end{cases} \tag{16}$$

or:

$$\begin{cases} V_{in_d} = V_{R_d} + I_q\left(\omega_s L - \frac{1}{\omega_s C}\right) \\ V_{in_q} = V_{R_q} - I_d\left(\omega_s L - \frac{1}{\omega_s C}\right) \end{cases} \tag{17}$$

Taking advantage of the fact that the angle θ_R can be chosen arbitrarily, some components, namely V_{R_q} and I_q can be eliminated. (The angle between them will be zero as long as the load can be considered purely resistive.) To achieve this, $\theta_R = 0$:

$$\begin{cases} V_{in_d} = V_{R_d} \\ V_{in_q} = -I_d\left(\omega_s L - \frac{1}{\omega_s C}\right) \end{cases} \tag{18}$$

These relationships can be replaced in the equations describing the system. The general state-space model may now be written as:

$$\begin{cases} \dot{\mathbf{x}} = \mathbf{A}\mathbf{x} + \mathbf{B}\mathbf{u} \\ \mathbf{y} = \mathbf{C}\mathbf{x} \end{cases} \tag{19}$$

For the studied system, this translates to:

$$\mathbf{x} = \begin{bmatrix} i_d & i_q & v_{C_d} & v_{C_q} & v_0 & i_0 \end{bmatrix}^T \tag{20}$$

$$\mathbf{A} = \begin{bmatrix} 0 & \omega_s & -\dfrac{1}{L} & 0 & 0 & 0 \\ -\omega_s & 0 & 0 & -\dfrac{1}{L} & -\dfrac{4}{\pi L}X_{tank} & \\ \dfrac{1}{C} & 0 & 0 & \omega_s & 0 & 0 \\ 0 & \dfrac{1}{C} & -\omega_s & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & -\dfrac{1}{L_0} \\ 0 & 0 & 0 & 0 & \dfrac{1}{C_0} & 0 \end{bmatrix} \tag{21}$$

$$\mathbf{u} = \begin{bmatrix} v_{in_d} & v_{in_q} \end{bmatrix}^T \tag{22}$$

$$\mathbf{B} = \begin{bmatrix} \dfrac{2}{\pi L_0} & 0 \\ 0 & 0 \\ 0 & 0 \\ 0 & 0 \\ 0 & 0 \\ 0 & -\dfrac{1}{C_0} \end{bmatrix} \tag{23}$$

where L_0 and C_0 are the inductance and the capacitance of the output filter. The capacitance of the underwater cable can also be lumped into C_0.

$$\mathbf{y} = \begin{bmatrix} v_0 & i_0 \end{bmatrix}^T \tag{24}$$

$$\mathbf{C} = \begin{bmatrix} 0 & 0 & 0 & 0 & 1 & 1 \end{bmatrix} \tag{25}$$

The derived state space model has the limitation of being valid only for a given frequency, therefore, it can be successfully used only for phase-shift control at fixed frequency.

5 Simulation Model

The schematic of the simulated circuit can be seen in Fig. 1.

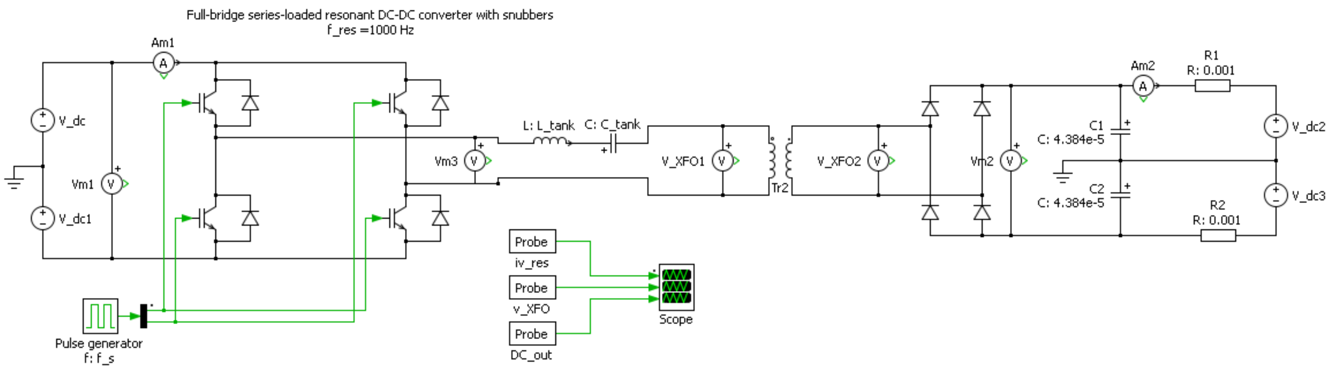

Figure 1: Schematic of the simulated SLR

6 Simulation Results

The results of simulation for steady-state are presented in the subsections below.

The no-load voltage of the converter is $V_{MVDC_0} = 77.5\,\mathrm{kV}$ (independently on the switching frequency).

Under short-circuit conditions the current reaches a peak value of $\hat{I}_{sc} = 1\,\mathrm{kA}$ (when the switching frequency is the one used for rated load).

6.1 SLR in Inductive Mode

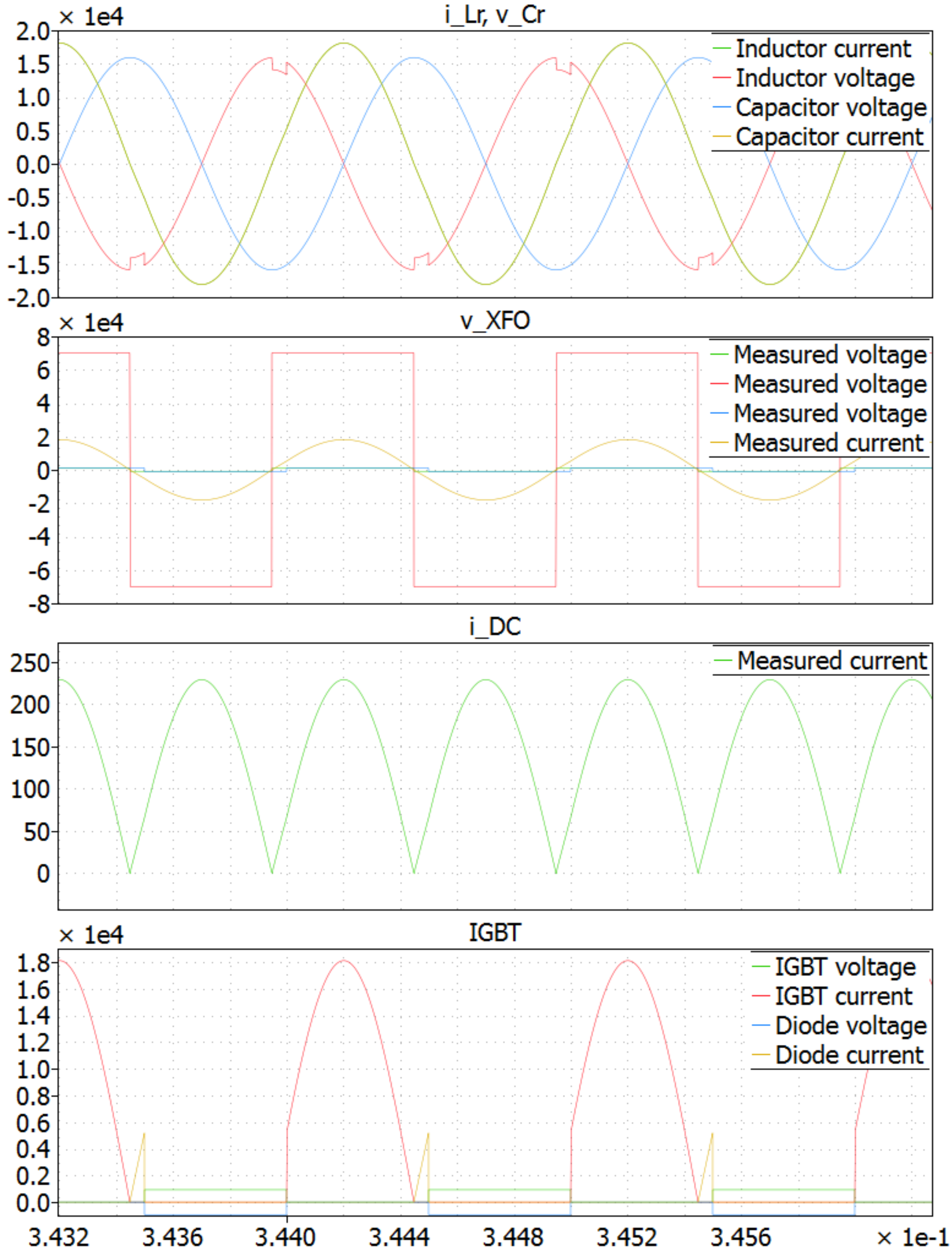

Figure 2: SLR resonant tank and output current waveforms (at 100 % load, inductive)

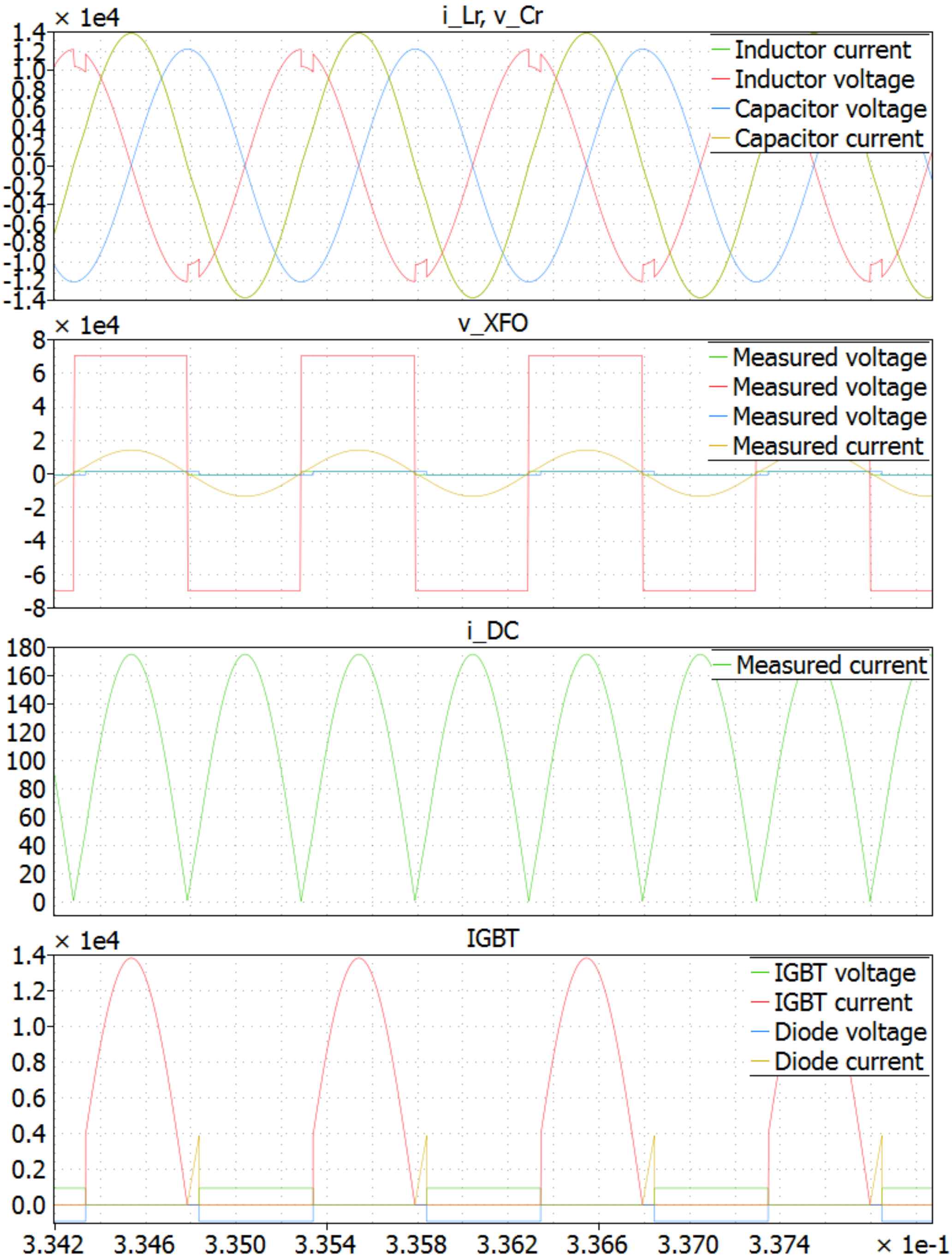

Figure 3: SLR resonant tank and output current waveforms (at 75 % load, inductive)

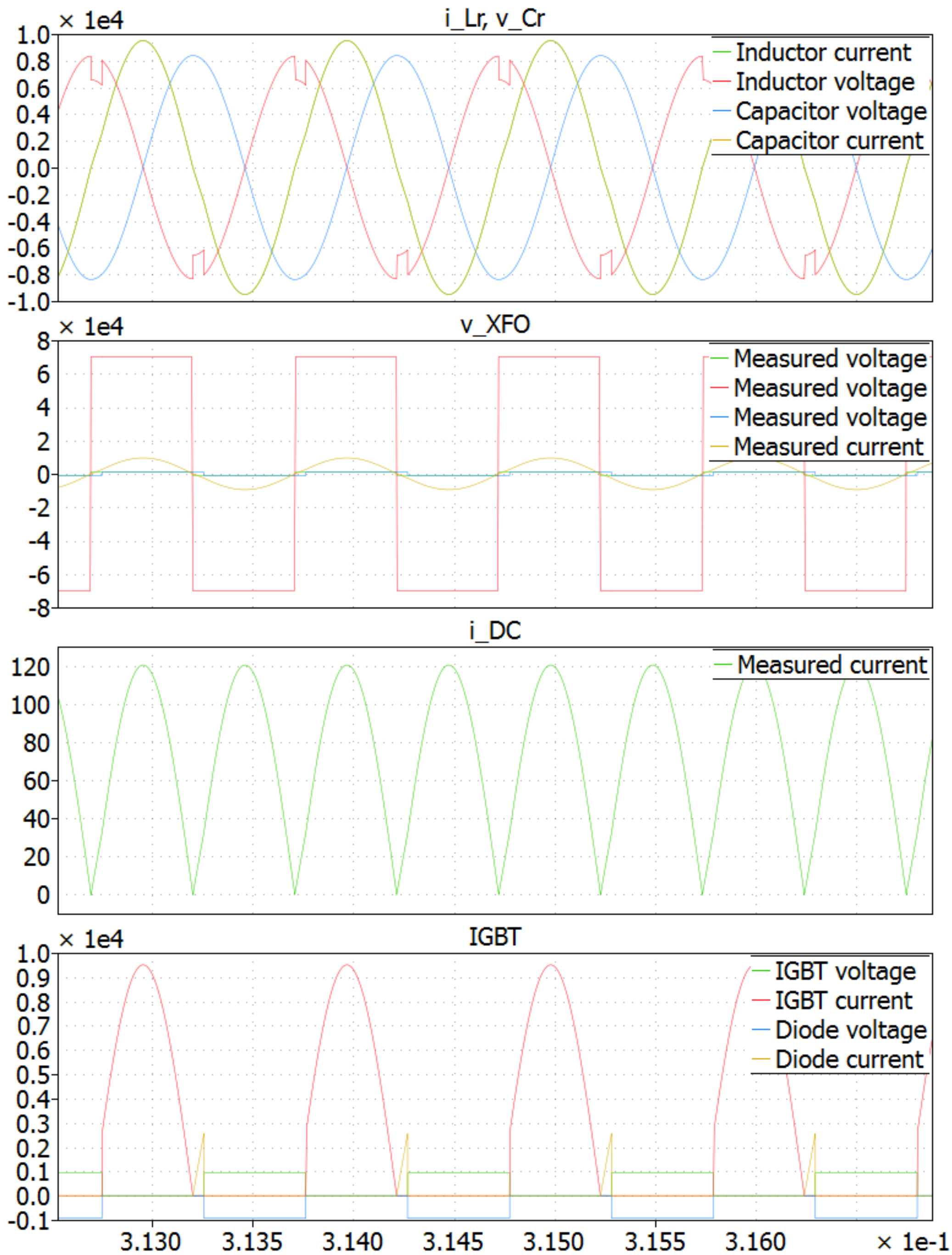

Figure 4: SLR resonant tank and output current waveforms (at 50 % load, inductive)

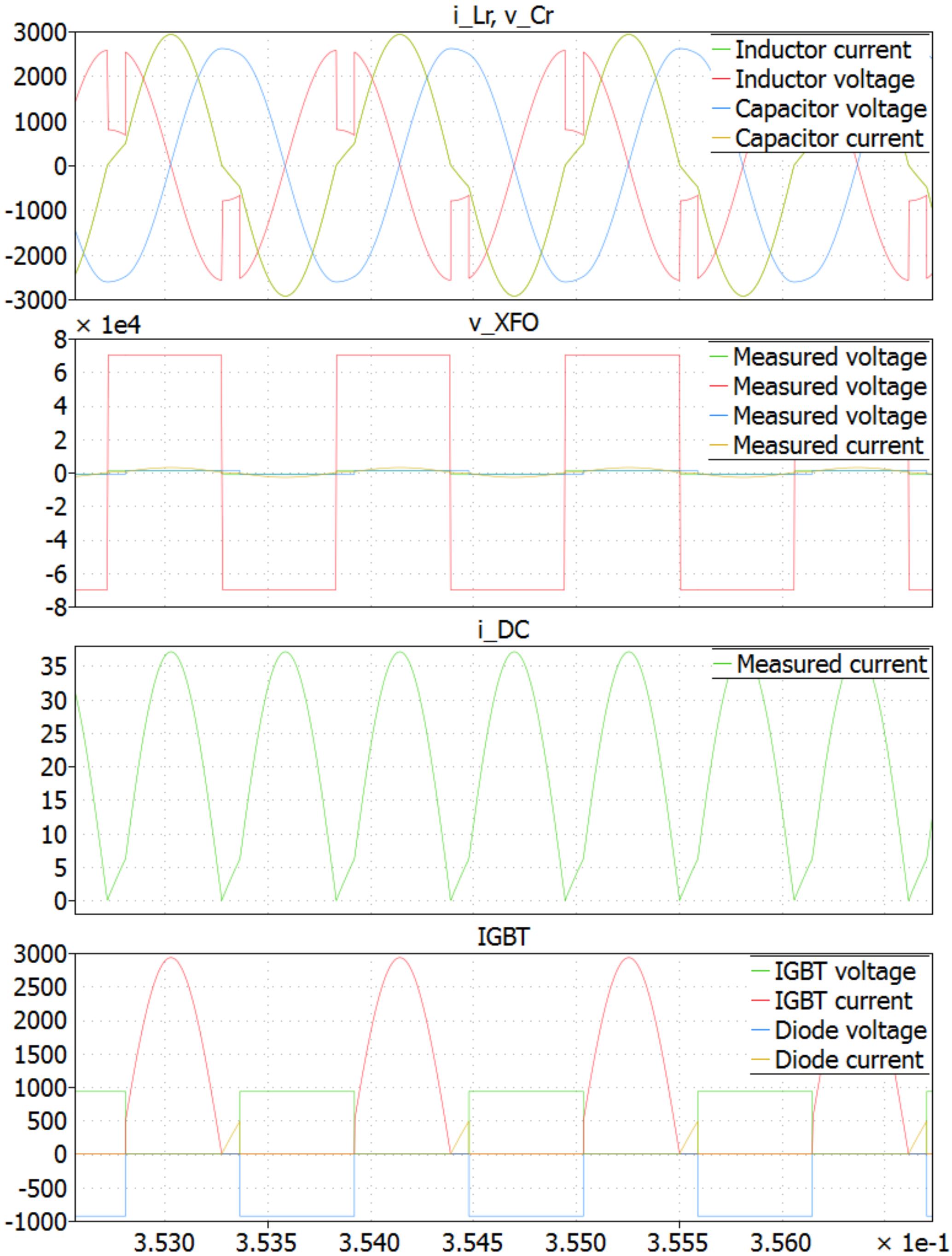

Figure 5: SLR resonant tank and output current waveforms (at 10 % load, inductive)

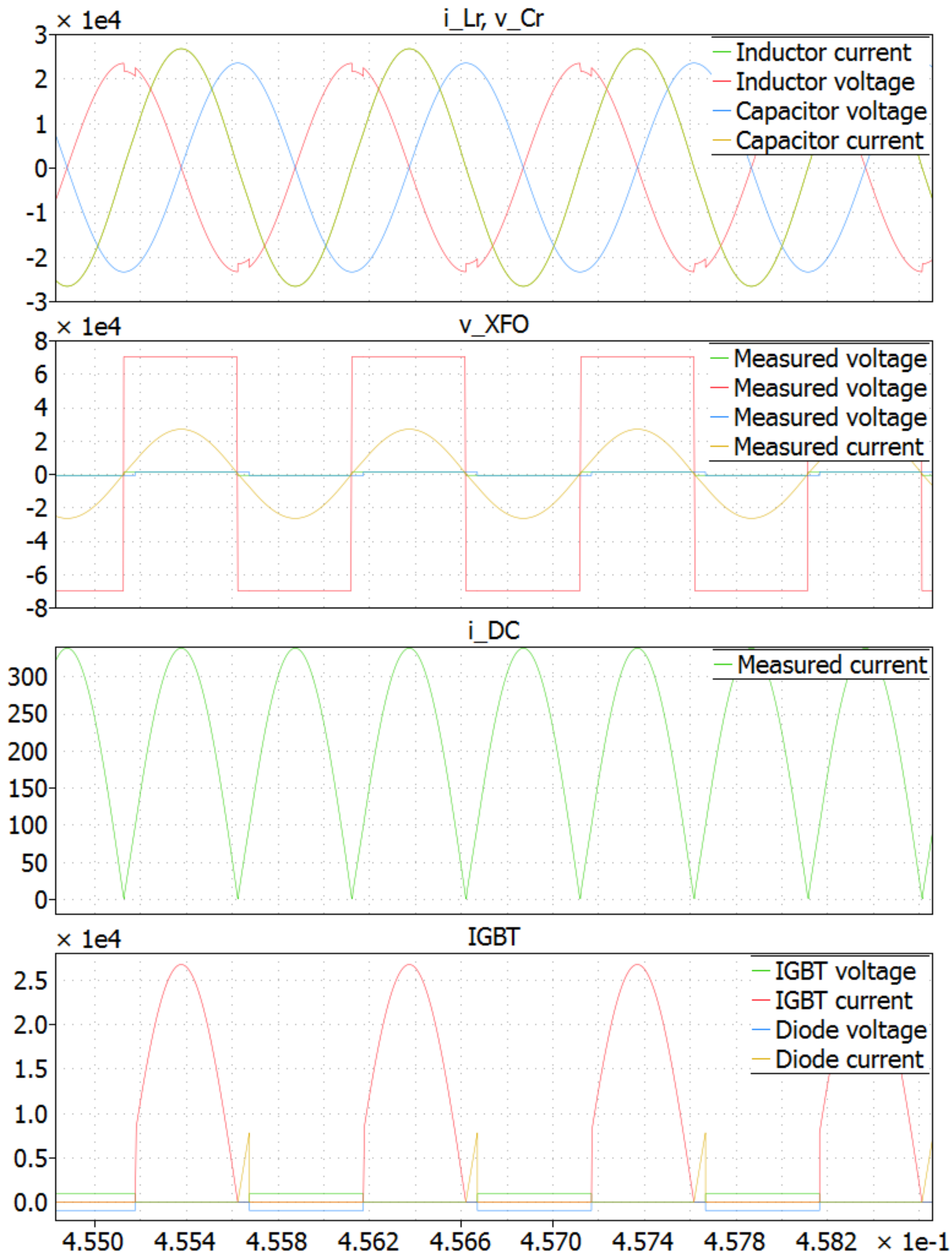

Figure 6: SLR resonant tank and output current waveforms (at 150 % load, inductive)

6.2 SLR in Capacitive Mode

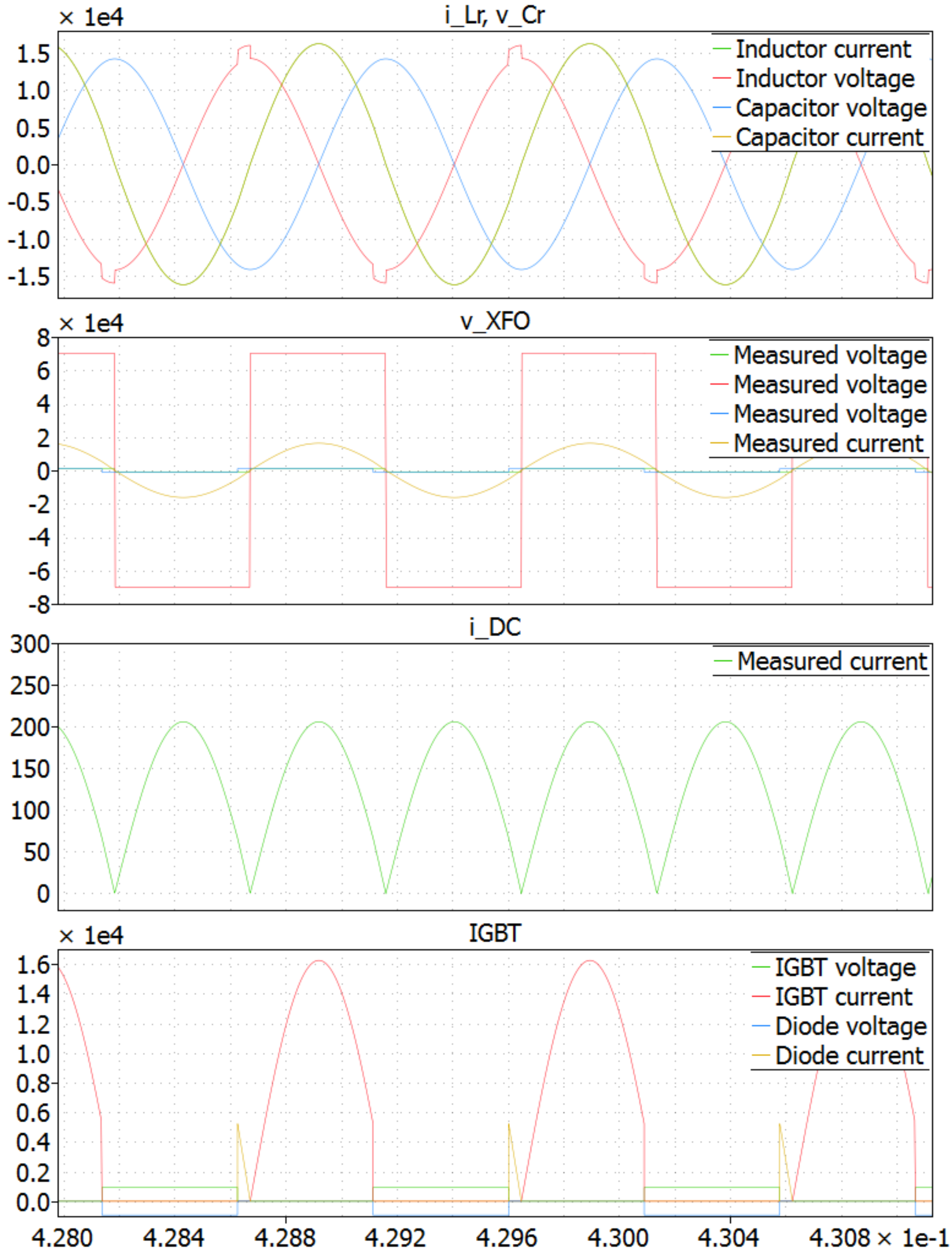

Figure 7: SLR resonant tank and output current waveforms (at 100 % load, capacitive)

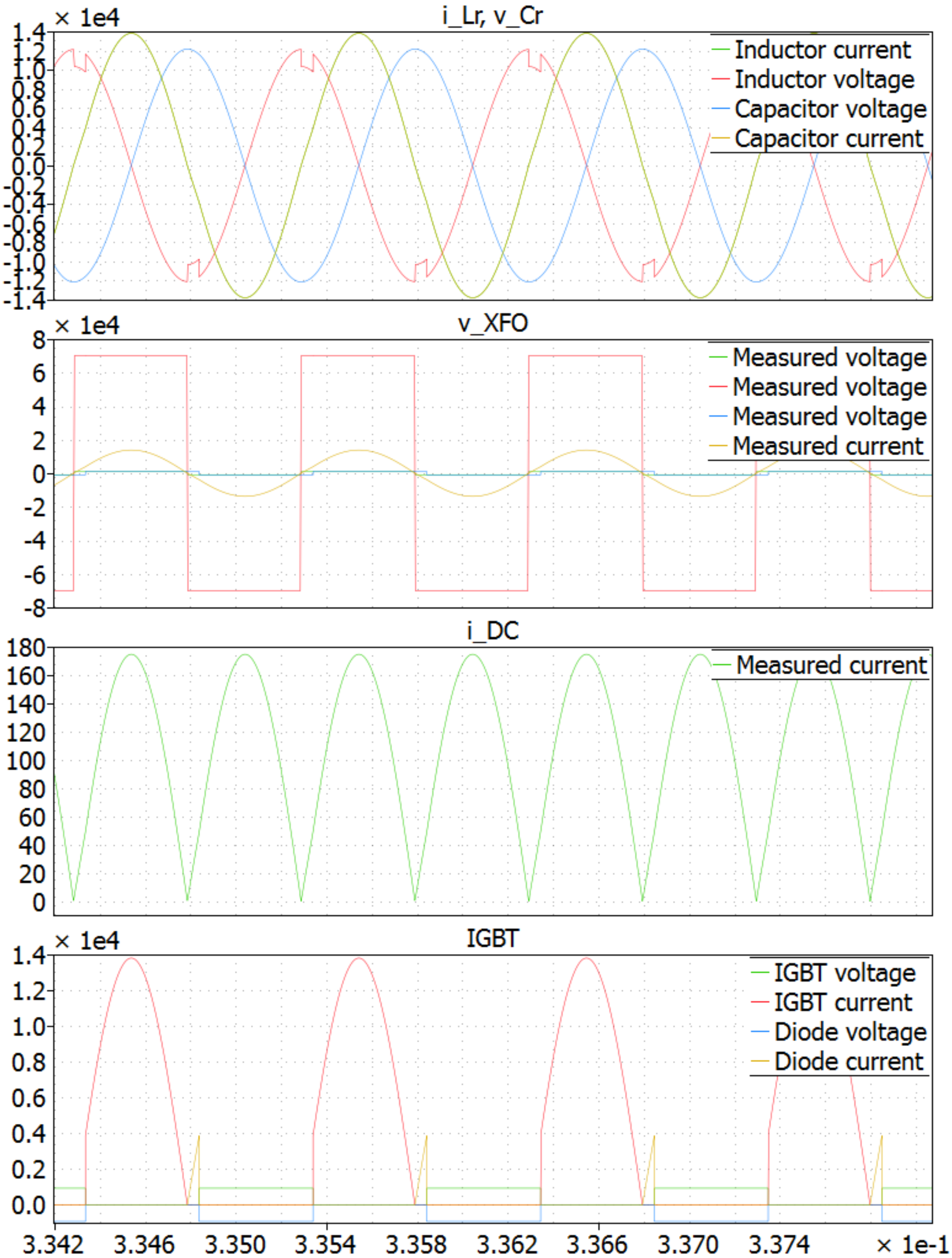

Figure 8: SLR resonant tank and output current waveforms (at 75 % load, capacitive)

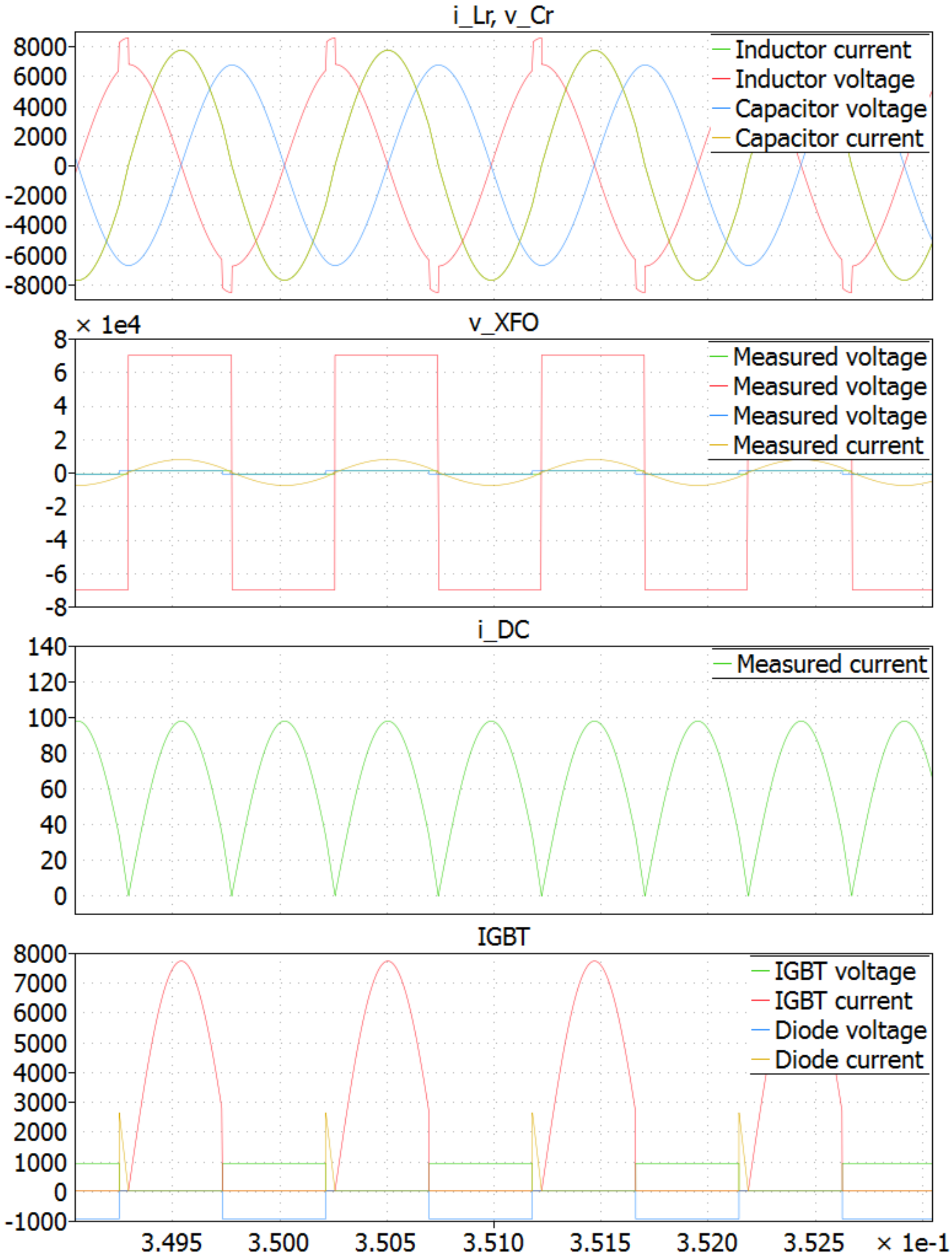

Figure 9: SLR resonant tank and output current waveforms (at 50 % load, capacitive)

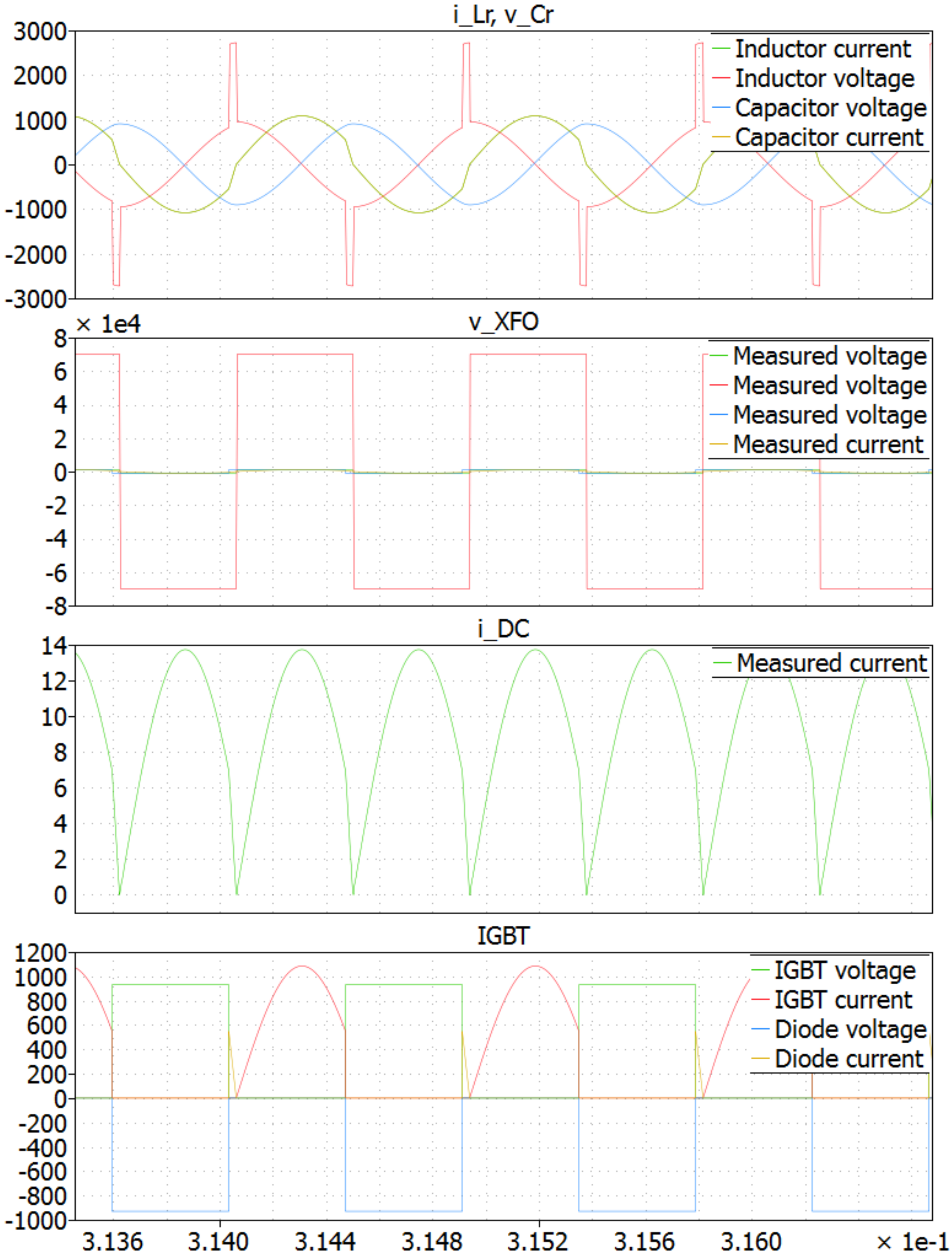

Figure 10: SLR resonant tank and output current waveforms (at 10 % load, capacitive)

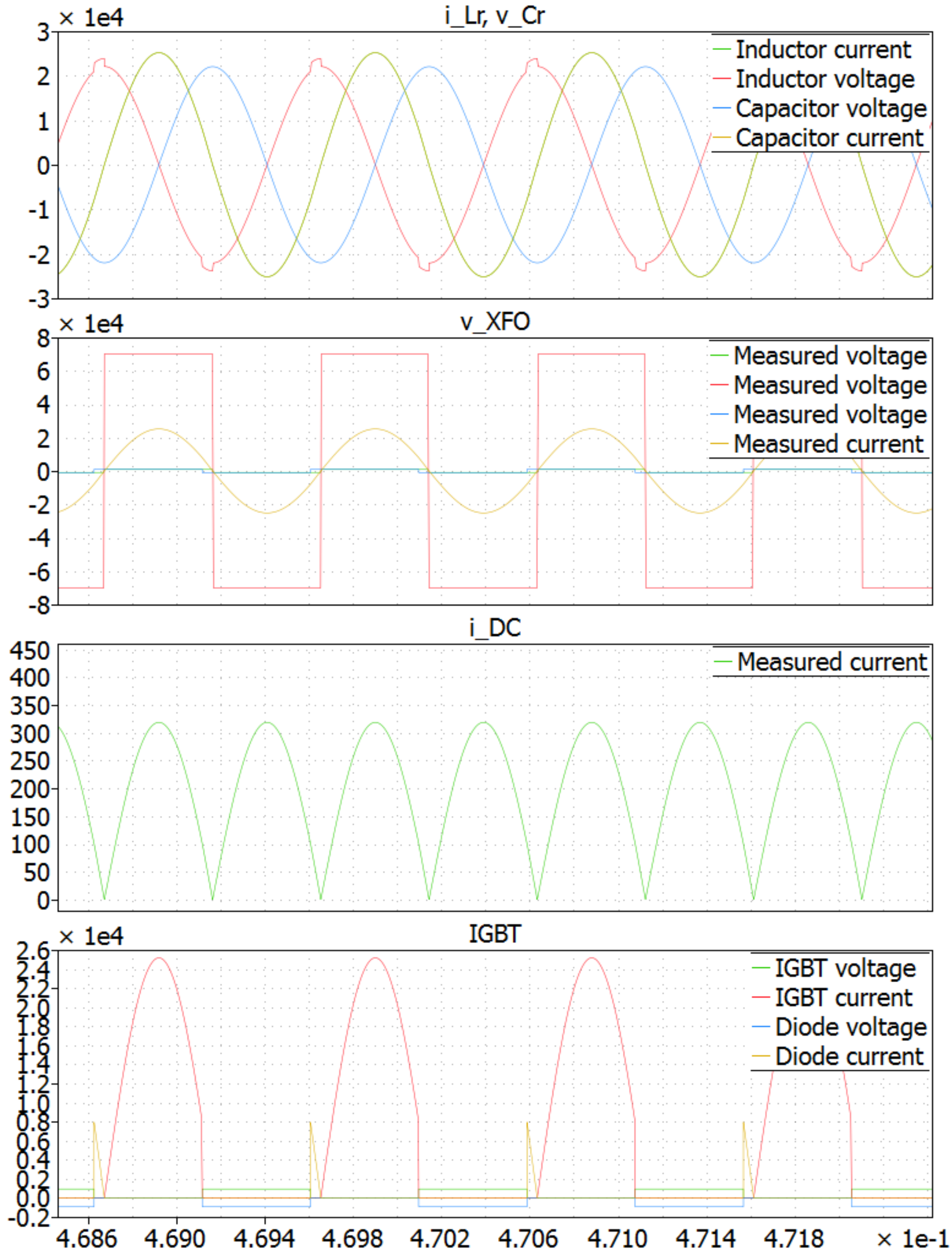

Figure 11: SLR resonant tank and output current waveforms (at 150 % load, capacitive)

7 Files

The following files contain simulation models and scripts:

- `resonant_tank_design.m` – Matlab script to design the resonant tank, as described previously
- `FullBridgeSLR_Q20_newres.plecs` – Plecs model of the resonant converter, with voltage source to model the MVDC grid
- `FullBridgeSLR_Q20_phsh.plecs` – Plecs model of the resonant converter, with phase shift control
- `FullBridgeSLR_simple.plecs` – Plecs model of the resonant converter, without XFO
- `FullBridgeSLR_smallsignal.plecs` – Plecs model of the resonant converter, set up for small signal analysis

The `old` directories contain previous attempts to simulate the converter.
In all cases, the parameters of the circuit can be modified by going to:
`Simulation -> Simulation parameters... -> Initialization`.