Class-E/F Inverter Applied to plasma generator with adaptive Control

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***Abstract* – At the radio frequency the induction plasma generator is very attractive for several industrial applications especially in material processing, so an inductively coupled plasma generator is developed in this paper. A high-frequency power supply class-E/F inverter is required to generate the magnetic field at frequency ranging from hundreds of kilohertz to tens of megahertz offer high efficiencies at high power densities. A resonant inverter based on zero-voltage switching and an adaptive control via a backstepping based on Lyapunov theory is used as a power supply. This paper presents the circuit modelling, simulation and control considerations.**

Keywords – Plasma Generators, class-E/F Inverter, Buckstepping control, Tuning functions.

I. INTRODUCTION

With the development of power semiconductor devices, many new circuit techniques and control schemes, research about high frequency circuits using advanced power devices such as MOSFETs, IGBTs and so we have been performed for high power applications, it has made it possible to implement high frequency inverters for induction heating, dielectric heating, and plasma generation. These applications generally require power levels from watts to megawatts at a single frequency ranging from hundreds of kilohertz to tens of megahertz [1]. Soft switching techniques have been used in power converters to reduce switching losses and alleviate electromagnetic interference (EMI); it has been studied based on its topology [2]. The various resonant inverters which are class-D, class-E, class-F and class E/F etc. inverter using power devices such as MOSFETs and IGBTs offer reduced switching loss by effective means of soft-switching technique [4]. The class-E inverter is an example of a resonant inverter which allows obtaining, almost the sinusoidal current–voltage at the frequency above (several of KHz to several of MHz) [3]. The Class E inverter is a well-known resonant converter that can operate at these frequencies with very high efficiency and produce up to several kilowatts of power [1]–[4]. It is a single-ended or push–pull topology where a transistor is soft switched, and therefore, its switching losses are significantly reduced.

A. Comparison and Motivation for class-E/F inverter

Comparing class E to the two class-F tunings, several advantages and disadvantages are apparent. Class E has the advantage of being capable of strong switching operation even with a very simple circuit, whereas class F allows this only as a limiting case using a circuit with great complexity [4]. Whereas the class-E amplifier is limited only by the intrinsic switching speed of the active device, class-F amplifier tunings may find their switching speed dominated by the limited number of harmonics, which have been utilized in the waveforms. Additionally, class E has the advantage of incorporating the output capacitance of the transistor into the circuit topology. The simple class-F implementation will not work in the presence of large output capacitance since the harmonics that were intended to be open circuited at the transistor will instead be capacitive. Classes F and 1/F also have advantages. First, they present more desirable waveforms. It is desirable to have waveforms with low peak voltage and rms current, (it is clear that class-F and F^{-1}amplifiers can perform better in these respects) [4].

This paper presents an approach to adaptive control of class E/F resonant inverter via a backstepping tuning function control design. This design removes several obstacles from adaptive linear control. Since the design is based on a single Lyapunove function incorporating both the state of the error system and the update law, the proof of global uniform stability is direct and simple. Moreover, all the error states except for the parameter error converge to zero [9].

These issues motivate the search for more desirable strong-switching amplifier tunings, which would, ideally, have some of the best features of both the class-E and class-F tunings. If such a harmonic tuning strategy is to be found, it would ideally have the following features:

- *Incorporation of the transistor output capacitance into the tuned circuit, where this capacitance may be as large as possible.*
- *ZVS to eliminate discharge loss from this capacitor.*
- *Simple circuit implementation.*
- *Use of harmonic tuning to achieve improved waveforms for better performance.*

II. SYSTEM CONFIGURATION

Fig.1 shows the system configuration of class-E series-parallel (LCL) resonant inverter for inductively coupled plasma generator. It consists of one or more switches (MOSFET IRFP) connected with output parasitic capacitance and a freewheeling drain-source diode.

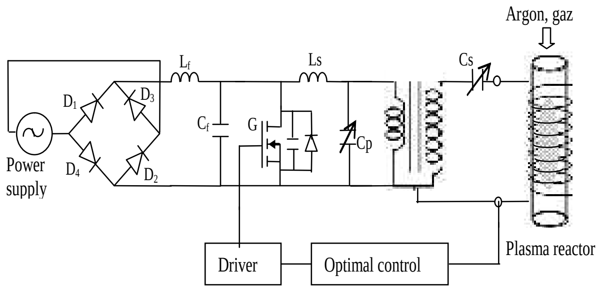

Fig.1. Class E inverter for inductively coupled plasma generator

The output resonant equivalent circuits were constructed by the output capacitor, matching transformer and plasma reactor. The load is modelled by the equivalent impedance which is varied during the heating process [7].

A. *Operation of a class-E resonant inverter*

The class-E resonant inverter is power topology specially suited for high frequency operation due to its low switching losses [8]. The key point of a class-E resonant inverter is the capacitance C voltage evolution after switch S is turned off. To minimize the switching losses, transistor S must be turned on while diode D is in conduction, thus providing to minimize the switching losses, transistor S must be turned on while diode D is in conduction, thus providing zero-voltage-switching. Capacitor C also operates as a turn-off snubber, further reducing the switching losses. The switch must operate with ZVS commutations. Failure to do so will result on capacitor discharging through the main switch thus increasing turn-on losses strongly. The voltage waveform applied to the resonant tank has strong harmonic content.

B. *Steady-State analysis topology*

The applied analysis method is based on a state-space description of the circuit and calculations of its properties using a dedicated program written in MATLAB.
The power switch of the inverter are turned on and off during each a constant interval T, the circuit variables namely, voltages v_c, v_{cs}, v_{cp} and inductor currents i_{Ls}, i_{Lr} are chosen as the state variables, such as:

$$\dot{x} = Ax + Bu\ , \quad y = Cx\ , \qquad (1)$$

where $x = [x_1, x_2, x_3, x_4]$ is the state vector.
$x_1 = v_{cp}$, $x_2 = v_{cp}$ $x_2 = v_{cs}$, $x_3 = i_{ls}$, $x_4 = i_{Lo}$.

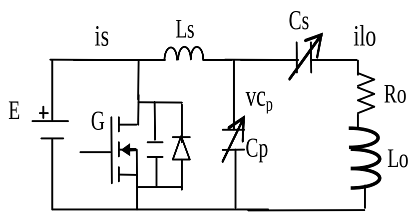

Fig.2. Class E inverter for inductively coupled plasma generator

Frequency characteristic of the resonant load section:
The load is a series-parallel resonant circuit; from the equivalent circuit of the load section, the load impedance in frequency domain is shown in equation 2.

$$Z_{Load} = j\mu L_s + \frac{1}{\dfrac{1}{R_{Lr} + j\mu L_r} + \dfrac{1}{R_c + \dfrac{1}{j\mu C}}}\ , \qquad (2)$$

Simulation of the inverter for deferent natural resonant frequencies of the load at 30, 45 and 50 Khz.

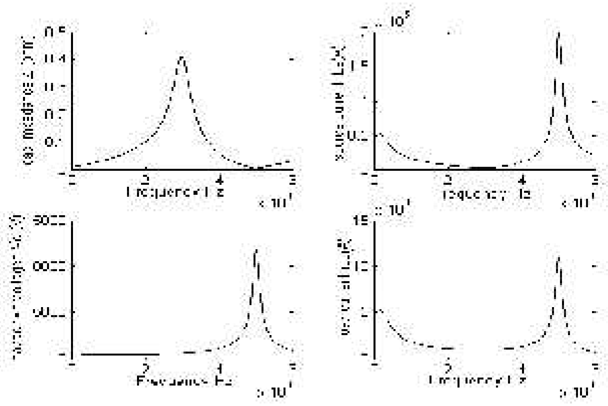

Fig.3. Amplitude of the load impedance at 30 KHz

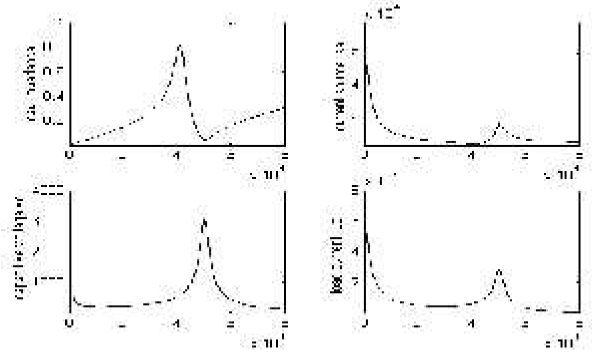

Fig.4. Amplitude of the load impedance at 50 KHz.

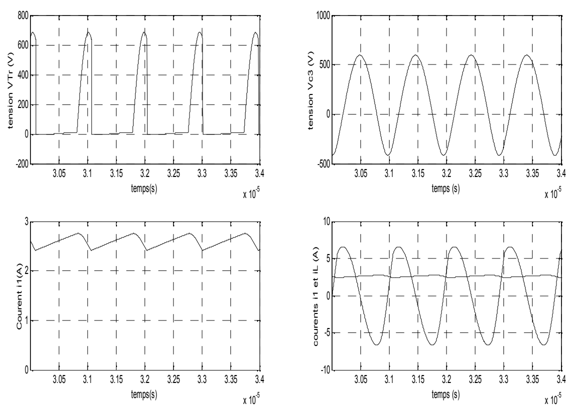

Fig.5. The waveforms of the switching voltage (a) switching current (b) and i_s and i_L currents of class-E inverter.

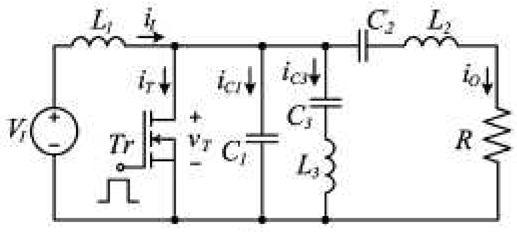

Fig.7. Circuit diagram of class E/F2 inverter.

III. CLASS-E/F$_2$ RESONANT INVERTER

C. *Analysis method*

The applied analysis method is based on a state-space description of the circuit and calculations of its properties using a dedicated program written in MATLAB.

The main points of this method for the Class E/F$_2$ inverter are presented below [4].

The following assumptions are used throughout the analysis

1) The transistor acts as a switch with a resistance of R_{Ton} for the ON interval, infinite resistance for the OFF interval, and zero switching times.

2) The shunt capacitance C_1 is linear including the transistor output capacitance.

3) All passive elements are linear, ideal, and time invariant.

The transistor is driven at frequency f and at any duty cycle D, where the duty cycle is defined as the switch ON time divided by the switching period T. The circuits in Fig. 3 can be described by a normalized set of state equations of the form:

$$dx/dt = Ax + Bu \tag{3}$$

where $x = [x_1, x_2, x_3, x_4, x_5, x_6]$ is the state vector.

$x_1 = v_{c1}, x_2 = v_{c2}, x_3 = v_{c3}, x_4 = i_{l1}, x_5 = i_{LR},$

$$A = \begin{bmatrix} -a_{11} & 0 & 0 & \frac{1}{C_1} & -\frac{1}{C_1} & -\frac{1}{C_1} \\ 0 & 0 & 0 & 0 & \frac{1}{C_2} & 0 \\ 0 & 0 & 0 & 0 & 0 & \frac{1}{C_3} \\ -\frac{1}{L_1} & 0 & 0 & 0 & 0 & 0 \\ \frac{1}{L_2} & -\frac{1}{L_2} & 0 & 0 & -\frac{R}{L_2} & 0 \\ \frac{1}{L_3} & 0 & -\frac{1}{L_3} & 0 & 0 & 0 \end{bmatrix} \tag{4}$$

$$B = \begin{bmatrix} 0 & 0 & 0 & \frac{1}{L_1} & 0 & 0 \end{bmatrix}^T, \tag{5}$$

Denoting the matrix A: for the ON interval $a_{11} = 1/C_1R_{Ton}$ and $a_{11} = 0$ for the OFF interval.

The steady-state solution is completed by adding the soft-switching conditions. The ZVS condition demands that the transistor voltage at the instant the transistor is switched on should be zero.

D. *Design procedure*

The design procedure is explained with an example circuit of the Class E/F inverter. The specifications of the example circuit are described in the table I.

TABBLE 1:
DESIGN PARAMETERS FOR THE CLASS-E AND CLASS-E/F$_2$ INVERTER

Description	Value
Operating frequency	f = 1 MHz
The peak transistor voltage	V_{Tm} = 450V
the rms transistor current	I_{Trrms} = 5A
Parameters L_1, L_2 and L_3	L_1 =270µH, L_2 =16.8 µH, L_3 =8 .3µH
Parameter C_1, C_2 and C_3	C_1=1.64 nF, C_2=1.99 nF, C_3=0.71nF
Resistance of the ON interval	R_{Ton} =0.182
The load resistance	R =20 .33 including R_{L2}
Input voltage	V_I =200V

The parameter L_1, L_2 and L_3 was selected to reduce losses and size of the Inverter.

VI. SIMUTATION RESULTS

The theoretical results have been verified experimentally using the Class E/F$_2$ inverter circuit in Fig. 6. The basic Class E circuit is extended by an input filter (L_3 and C_3) and parasitic series resistances (R_{L1}, R_{L2} and R_{L3}) of the inductors. They are used to calculate the output power and total efficiency of the inverter.

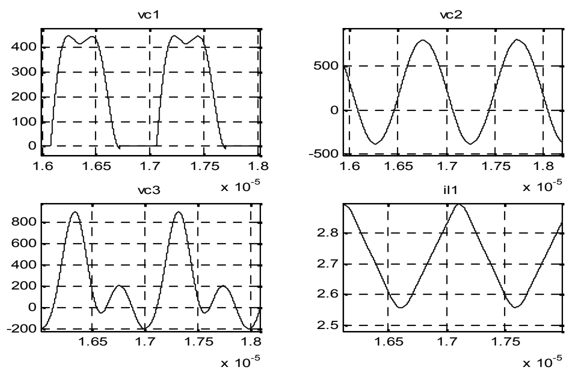

Fig.8. Capacitors voltage v_{c1}, v_{c2}, v_{c3} and i_{l1} current:

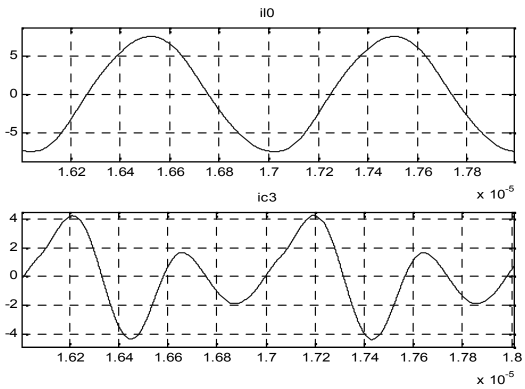

Fig.9. Load current i_{lo} and parallel capacitor current.

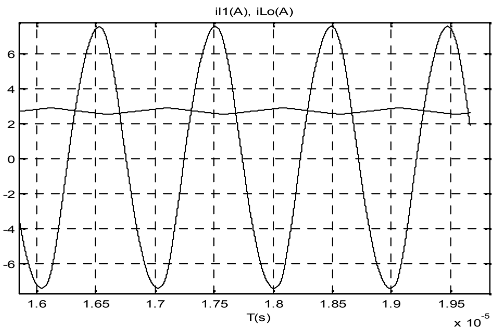

Fig.10. Input and output current i_{l1} and i_{lo}.

The analysis shows that the Class E/F inverter has the best performance, and therefore, the remainder of this paper will focus on this one.

V. ADAPTIVE BACKSTEPPING DESIGN

The control objective is to generate a feedback control input *u(t)* for the plant with unknown parameters —, such that all closed loop signals are bounded, and the plant output $y(t)=x_1(t)$ tracks a given bounded reference output $y_r(t)$ with pounded derivatives $\dot{y}_r(t)$, $\ddot{y}_r(t)$ [10], [11], [12].

Consider the plant of class E series resonant inverter:

$$\begin{aligned} \dot{x}_1 &= -a_1x_1 - a_2x_2 + b_1u \\ \dot{x}_2 &= -a_3x_1 \\ y &= cx_1 \end{aligned} \tag{6}$$

Where the specific parameters are: $a_1 = R/l$, $a_2 = 1/l$, $a_3 = 1/C$ and $b = 1/l$.

1. State estimation filters:

We start by representing the plant (6) in the observer canonical form:

$$\begin{aligned} \dot{x}_1 &= -a_1y + x_2 + bu \\ \dot{x}_2 &= -a_2a_3y \\ y &= cx_1 \end{aligned} \tag{7}$$

Or, in a more compact way, as

$$\dot{x} = \begin{bmatrix} 0 & 1 \\ 0 & 0 \end{bmatrix} x - y\begin{bmatrix} a_1 \\ a_2a_3 \end{bmatrix} + \begin{bmatrix} b \\ 0 \end{bmatrix} u \tag{8}$$

$$y = Cx \tag{9}$$

Where

$$A = \begin{bmatrix} 0 & 1 \\ 0 & 0 \end{bmatrix},\ a = \begin{bmatrix} a_1 \\ a_2a_3 \end{bmatrix}\ \ B = \begin{bmatrix} b \\ 0 \end{bmatrix} \text{ and } C = \begin{bmatrix} 1 & 0 \end{bmatrix}.$$

In this situation, we are able to express (7), (8) as

$$\dot{x} = Ax + f(y,u) \tag{10}$$

where

$$f(x,u)^T = \begin{bmatrix} u & -y & 0 \\ 0 & 0 & -y \end{bmatrix}$$

and the parameter vector $\text{—}^T = \begin{bmatrix} b & a \end{bmatrix}$

For state estimation we employ the filters

$$\dot{\P} = A_0\P + Ky \tag{11}$$

$$\dot{\Omega}^T = A_0\Omega^T + f(y,u)^T \tag{12}$$

Where the vector $K^T = \begin{bmatrix} K_1 & K_2 \end{bmatrix}$ is chosen so that the matrix:

$A_0 = A - KC^T = \begin{bmatrix} -K_1 & 1 \\ -K_2 & 0 \end{bmatrix}$ is Hurwitz, and hence P exists such that

$$PA_0 - A_0^TP = -I,\ P = P^T \tag{13}$$

With the help of the those filters our state estimate is

$$\hat{x} = \P + \Omega^{T\text{—}} \tag{14}$$

and the state estimation error

$$\pounds = x - \hat{x} \tag{15}$$

By exploiting the structure of $f(y,u)^T$ we denote the Ω filter by:

$$\Omega^T = \begin{bmatrix} \text{ª} & \text{¦} \end{bmatrix} \tag{16}$$

and show that to the special dependence of $f(y,u)^T$ of *u*, the equation for ª and ¦ are governed by:

$$\begin{cases} \dot{\text{ª}} = A_0\text{ª} + e_1u \\ \dot{\text{¦}} = A_0\text{¦} + e_2y \end{cases} \tag{17}$$

the vector $\P$ in (11) can be obtained from the filter (17) through the algebraic expression

$$\P = -A_0^3\text{¦} \tag{18}$$

What has been achieved thus far is a static relationship between the state x and the unknown parameter — :

$$x = \P + \Omega^{T\text{—}} + \pounds \tag{19}$$

In conclusion, from (19) and the expressions of KREISSELMEIR filters an equivalent expression for the virtual estimate $\hat{x}$ is

$$\hat{x} = -A_0^3\text{¦} - \sum_{i=1}^{3} a_iA_0^i\text{¦} + b_0\text{ª}$$

$$\hat{x} = B(A_0)\text{ª} - A(A_0)\text{¦} \tag{20}$$

Where $A(\cdot)$ and $B(\cdot)$ are matrix–valued polynomial functions.
For the plant (6) with relative degree 2, an adaptive backstepping control design procedure consists of two steps. It starts with its output y,

$$\dot{y} = x_2 - yC^T a \tag{21}$$

From the algebraic expressions (19) we have

$$x_2 = ¶_2 + \Omega_{(2)}^{T-} + £_2 \tag{22}$$

$$= ¶_2 + [ª_2 \quad ¦_2]^{T-} + £_2$$

$$= bª_2 + ¶_2 + [0 \quad ¦_2]^{T-} + £_2 \tag{23}$$

Substituting both (22) and (23) into (21), we obtain the following two important expressions for $\dot{y}$:

$$\dot{y} = ¶_2 + µ^{T-} + £_2 \tag{24}$$

$$= bª_2 + ¶_2 + \overline{µ}^{T-} + £_2 \tag{25}$$

and $\dot{ª}_2 = -K_2 ª_1$ (26)

Where the 'regressor' µ and the 'truncated regressor' ´ are defined as

$$µ = [ª_2, \quad ¦_2 - yC^T]^T, ´ = [0, \quad ¦_2 - yC^T]^T \tag{27}$$

All of these states are available for feedback. Our design task is to force the output y to asymptotically track the reference output y_r while keeping all the closed loop signals bounded. We employ the change of coordinates

$$z_1 = y - y_r \tag{28}$$

$$z_2 = ª_2 - ϭ\dot{y}_r - \sim_1 \tag{29}$$

Where ϭ is the estimate of $° = 1/b$, our goal is to regulate $z[z_1 \quad z_2]^T$ to zero.

Step 1: Let the tracking error $z_1 = y - y_r$ and introduce $z_2 = ª_2 - ϭ\dot{y}_r - \sim_1$, where $\sim_1$ a function to be designed is. Then, from (28) we get:

$$\dot{z}_1 = bª_2 + ¶_2 + \overline{µ}^{T-} \tag{30}$$

By substituting (29) into (30), we get

$$\dot{z}_1 = b\sim_1 + ¶_2 + \overline{µ}^{T-} + £_2 - b\tilde{ϭ}\dot{y}_r + bz_2 \tag{31}$$

Scaling the first stabilizing function $\sim_1$ as

$$\sim_1 = ϭ\overline{\sim}_1 \tag{32}$$

Choosing the design function $\overline{\sim}_1$ as:

$$\overline{\sim}_1 = -c_1 z_1 - d_1 z_1 - ¶_2 - \overline{µ}^{T-} \tag{33}$$

Results in the system

$$\dot{z}_1 = -c_1 z_1 - d_1 z_1 + ¶_2 + \overline{µ}^{T-} + £_2 - b(\dot{y}_r + \overline{\sim}_1)\tilde{ϭ} + bz_2 \tag{34}$$

and considering the first partial positive definite function:

$$V_1 = \frac{1}{2} z_1^2 + \frac{1}{2}\tilde{-}^2 \Gamma \tilde{-} + \frac{|b|}{2¥}\tilde{ϭ}^2 + \frac{1}{4d_1}£^T P£ \tag{35}$$

We examine the derivative of V_1 as:

$$V_1 = z_1\left[-c_1 z_1 - d_1 z_1 + £_2 + \left(µ - ϭ(\dot{y}_r + \overline{\sim}_1)C\right)\tilde{-} - b\left(\dot{y}_r + \overline{\sim}_1\right)\tilde{ϭ} + \hat{b}z_2\right]$$
$$-\tilde{-}^T \Gamma^{-1}\dot{\hat{-}} - \frac{|b|}{¥}\tilde{ϭ}\dot{ϭ} - \frac{1}{4d_1}£^T£ \tag{36}$$

$$= -c_1 z_1^2 + bz_1 z_2 - |b|\tilde{ϭ}\frac{1}{¥}\left[¥ \operatorname{sgn}(b)\left(\dot{y}_r + \overline{\sim}_1\right)z_1 + \dot{ϭ}\right] + \tilde{-}^T \Gamma^{-1}$$
$$\left[\Gamma\left(µ - ϭ(\dot{y}_r + \overline{\sim}_1)C\right)z_1 - \dot{\hat{-}}\right] - d_1 z_1^2 + z_1 £_2 - \frac{1}{4d_1}£^T£ \tag{37}$$

To eliminate the unknown indefinite $\tilde{-}$ and $\tilde{ϭ}$ terms in (37) we choose

$$\dot{ϭ} = -¥ \operatorname{sgn}(b)\left(\dot{y}_r + \overline{\sim}_1\right) \tag{38}$$

Were (38) is used as the actual update law for ϭ, and $\dot{\hat{-}} = \Gamma ²_1$, where

$$²_1 = \left(µ - ϭ\left(\dot{y}_r + \overline{\sim}_1\right)C\right)z_1 \tag{39}$$

Substituting (38) and (39) into (37) we obtain

$$\dot{V}_1 \le -c_1 z_1^2 + \hat{b}z_1 z_2 + \tilde{-}^T\left(²_1 - \Gamma^{-1}\dot{\hat{-}}\right) \tag{40}$$

Step 2: from (29) with the help of (26) we obtain

$$\dot{z}_2 = \dot{ª}_2 - ϭ\ddot{y}_r - \dot{ϭ}\dot{y}_r - \dot{\sim}_1\left(y, ¦, ϭ, -, y_r\right) \tag{41}$$

or

$$\dot{z}_2 = \sim_2 - {}_2 - \frac{\partial \sim_1}{\partial y}\left(µ^T\tilde{-} + £_2\right) - \frac{\partial \sim_1}{\partial \hat{-}}\dot{\hat{-}} \tag{42}$$

Since our system is augmented by the new state z_2, we augment the Lyapunov function (35) as

$$V_2 = V_1 + \frac{1}{2}z_2^2 + \frac{1}{4d_1}£^T P£ \tag{43}$$

The derivative of V_2 satisfies

$$V_2 \le -c_1 z_1^2 - c_2 z_2^2 - d_2\left(z_2\frac{\partial \sim_1}{\partial y} + \frac{1}{2d_2}£_2\right)^2$$
$$\le -c_1 z_1^2 - c_2 z_2^2 \tag{44}$$

And the control law

$$u = \sim_2 + {}^{\sigma}\ddot{y}_r \tag{45}$$

IV. SIMUTATION RESULTS

TABLE 2:
DESIGN PARAMETERS FOR THE OBSERVER BASED CONTROLLER

Description	Value
Liapunov gains	c_1=2e^{-5}, c_2=0,02
Observer gains	k_1=0,01 , k_2=2
Damping gain	d_1= 1e^{-5}, d_2=0,01
Adaptation gains	g_1=10, g_2=10

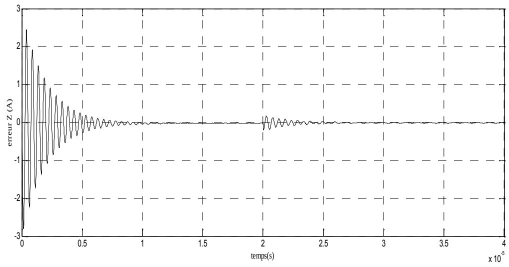

Fig.11. Steady state tracking error z_1 and z_2.

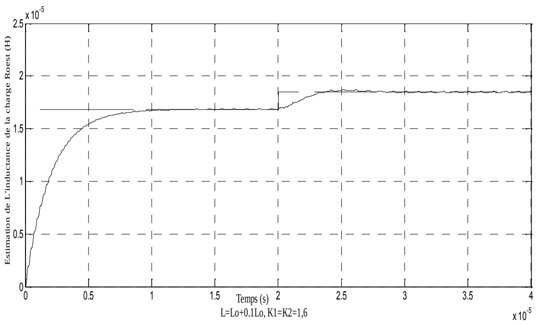

Fig.12. Estimation of parameters theta1

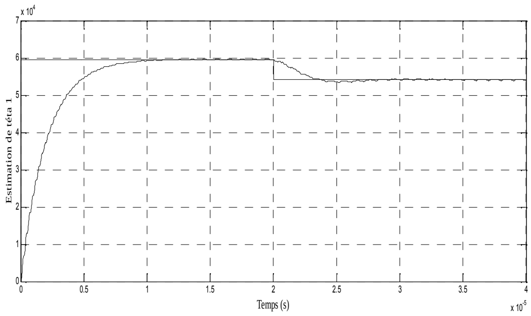

Fig.13. Estimation of parameters teta2 (RO)

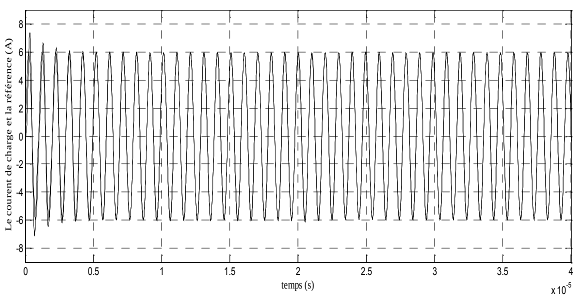

Fig.14. Estimation of the load current

VI. CONCLUSION

In this paper, the design of a MOSFET based class-E series-parallel resonant inverter power supply for an inductively plasma generator system has been presented. The variable load is highly inductive and requires a several KW active power at a frequency of several MHz Based on a detailed topology investigation, An LCC-resonant circuit supplied by a voltage source class-E and class-E/F_2 inverter are chosen. An analysis of the circuit and basic design rules are given. These E/F tunings allow strong-switching operation as in class E, but show a greater tolerance for transistor output capacitance and present waveforms approaching those of the more desirable class-F^{-1} the family exhibits a tradeoff between circuit complexity and performance. A control scheme allowing operation of the inverter with the MOSFET switching-losses is explained and simulation results verifying the operation of the control are shown.

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