State Space Models of Diode and Bipolar Transistor for Simulation of Power Converters.

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Abstract - The perfecting of digital control and regulation loop of power converters cannot be done without a simulation phase. The aim of this paper is to show the behavioural approach of components through a model described in the state space by an inner model, in order to allow the identification of the parameters of the model from experimental curves.

I. INTRODUCTION

In the area of power electronic simulation, software are numerous and so are the methods to describe the components. There are two main methods for semiconductor device modelling : macro-circuit and micro-circuit. Among the macro-model approaches, complete diode and bipolar transistor models for electrical circuit simulators have been developed to the LEEI in Toulouse [1]. But one point has to be resolved : the possibility to determine the values of the parameters of the models directly from experimental curves.

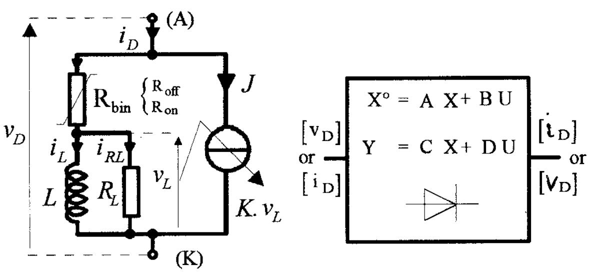
The purpose of this paper is to show an automatic approach of power electronic simulation, where electrical circuits are described in the state space by an inner model. Its performance is demonstrated using the Matlab-Simulink software. The automatic tools are used for the identification of the parameters of the component models.

II. DIODE MODELLING.

The converter designer has to know power components outer characteristics (delay time switching, recovery current etc...). In the case of the diode, we want to represent its imperfections which are predominant when designing a power converter. In particular, it is important to reproduce the transient behaviour at turn off. This has been achieved with an electrical equivalent circuit such as resistance, inductance, and a controlled current source (figure 1a). The inductance L and the resistance R_L characterise a 'current sensor' and the controlled current source J represents a two-state capacity whose value depends on the two-state resistance R_{bin}.

[1] The author is with B. Feuvrie, C. Bergmann and Z. Bestaoui

The diode inner model is represented figure 1b. The only state variable is the current in the inductance L. The state of R_{bin} depends on the control U and the state vector i_L.



a) Electrical equivalent circuit. b) Inner model

Figure 1 : Diode Modelling

II.1 DIODE_IU (INPUT CURRENT AND OUTPUT VOLTAGE)

In the case of a diode with an input current and an output voltage, we have :

$$X = [i_L] \qquad U = [i_D] \qquad Y = [v_D]$$

- <u>First inner model equation</u> :

From figure 1a, we can write :

$$i_D = i_L + i_{R_L} + J = i_L + L\frac{d\,i_L}{dt}\left(\frac{1}{R_L} + K\right) \qquad (1)$$

Whence we deduce :

$$\frac{d\,i_L}{dt} = \frac{-R_L}{L(1+K R_L)}\,i_L + \frac{R_L}{L(1+K R_L)}\,i_D \qquad (2)$$

As $i_D = U$ we can deduce from Eq. (2) :

$$A = \left[\frac{-R_L}{L(1+KR_L)}\right] \qquad B = \left[\frac{R_L}{L(1+KR_L)}\right]$$

- Second inner model equation :

From figure 1a, we deduce :

$$v_D = v_L + v_{bin} = v_L + R_{bin}\left(i_L + \frac{v_L}{R_L}\right)$$

$$\Rightarrow \quad v_D = L\frac{di_L}{dt}\left(1 + \frac{R_{bin}}{R_L}\right) + R_{bin}\, i_L \qquad (3)$$

Eq. (2) and (3) becomes :

$$v_D = R_L \frac{KR_{bin} - 1}{1 + KR_L} i_L + \frac{R_L + R_{bin}}{1 + KR_L} i_D \qquad (4)$$

As $v_D = Y$ and $i_D = U$, we can deduce from Eq. (4) :

$$C = \left[\frac{R_L(KR_{bin} - 1)}{1 + KR_L}\right] \qquad D = \left[\frac{R_L + R_{bin}}{1 + KR_L}\right]$$

- Rbin test : If $i_{bin} > 0$, then $R_{bin} = R_{min}$, else $R_{bin} = R_{max}$.

$$i_{bin} = i_D - J = i_D - KR_L\, i_{RL} = i_D - KR_L\left(i_{bin} - i_L\right)$$

$$\Rightarrow \quad i_{bin}(1 + KR_L) = i_D + KR_L\, i_L \qquad (5)$$

As $(1 + KR_L)$ is always positive, we can write :

If $(i_D + KR_L\, i_L) > 0$ then $R_{bin} = R_{min}$, else $R_{bin} = R_{max}$

$$R_{bin} = R_{min}\,\mathrm{sign}((U + KR_L X) > 0)) + R_{max}\,\mathrm{sign}((U + KR_L X) < 0)$$

II.2 DIODE_UI
(INPUT VOLTAGE AND OUTPUT CURRENT)

In the case of a diode with an intput voltage and an ouput current, we have :

$$X = [i_L] \qquad U = [v_D] \qquad Y = [i_D]$$

Arranging Eqs (2) (3), and(4), we demonstrate that :

$$A = \left[\frac{-R_{bin}R_L}{L(R_{bin} + R_L)}\right] \qquad B = \left[\frac{R_L}{L(R_{bin} + R_L)}\right]$$

$$C = \left[\frac{R_L(1 - KR_{bin})}{R_L + R_{bin}}\right] \qquad D = \left[\frac{1 + KR_L}{R_L + R_{bin}}\right]$$

- Rbin test : If $i_{bin} > 0$, then $R_{bin} = R_{min}$, else $R_{bin} = R_{max}$.

$$v_D = v_L + R_{bin}\, i_{bin} = R_L(i_{bin} - i_L) + R_{bin}\, i_{bin}$$

$$\Rightarrow \quad v_D = i_{bin}(R_L + R_{bin}) - R_L\, i_L \qquad (6)$$

Whence we deduce :

$$i_{bin}(R_L + R_{bin}) = v_D + R_L\, i_L \qquad (7)$$

As $(R_L + R_{bin})$ is always positive, we can write :

If $(v_D + R_L\, i_L) > 0$ then $R_{bin} = R_{min}$, else $R_{bin} = R_{max}$

$$R_{bin} = R_{min}\,\mathrm{sign}((U + R_L X) > 0) + R_{max}\,\mathrm{sign}((U + R_L X) < 0)$$

III. DIODE PARAMETERS IDENTIFICATION

We are confronted with 2 kinds of problems :

- The first one is linked to the experimental signal acquisition.

- The second one is due to the mathematical identification method.

In order to separate the 2 problems; we have applied the following method :

First Phase : Validation of the identification methods.

1) Arbitrarily, we choose the diode parameters. They are called reference parameters.

2) We simulate a test circuit.

3) From simulations of v_D and i_D, we applied mathematical methods for the identification of the diode parameters.

4) We compare the identified parameters with the reference parameters. If the error is very low, the mathematical identification method is validated.

These different points are summed up figure 2.

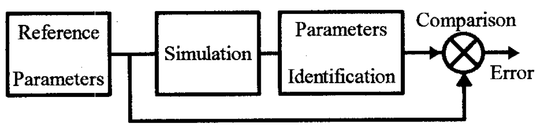


Figure 2 : Principle of the identification from simulations

Phase 2 : Identification from experimental signals.

6) From an experimental test rig, we recover the waveforms of i_D and v_D.

7) In using the mathematical methods validated in the first phase, we identify the experimental parameters of the diode.

8) A simulation is made with the identified parameters.

9) We compare simulation and experimental results. If both are similar, the identification problem is resolved. In the contrary case, the measure of v_D and i_D has to be improved.

These different points are summed up figure 3.

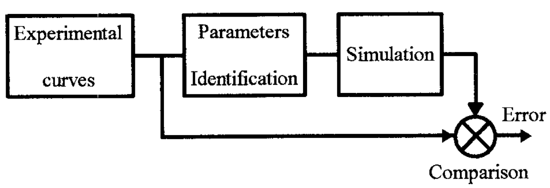


Figure 3 : Principle of parameters identification from experimental signal.

III.1. TEST CIRCUIT

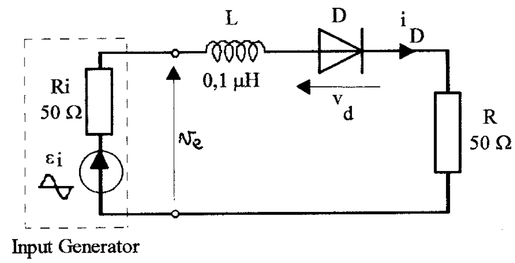
The test circuit is presented figure 4a. The input voltage $v_e(t)$ is a high frequency sinusoidal source voltage in series with a diode and a load resistance. The inductance L characterises the line.

From figure 4a, we write :

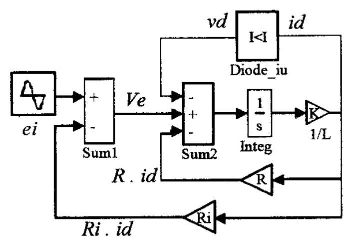
$$L \frac{d i_D}{dt} = v_e(t) - v_D(t) - R\, i_D(t) \qquad (8)$$

and : $$v_e(t) = e_i(t) - R_i\, i_D(t) \qquad (9)$$

The equivalent Simulink circuit is represented figure 4b. In this case a diode model with an input current and an output voltage has to be used to avoid an algebraic loop in the simulation circuit.



a) Electrical test circuit.



b) Equivalent Simulink circuit.

Figure 4 : Simulation Test Circuit

III.2. SIMULATION RESULTS

The parameters which have to be identified are :

- the gain K of the current source J and the resistance R_{on}.

- the resistance R_L.

As recommended in [1], $L = 10^{-8}$ H and $R_{off} = 10^5\ \Omega$.

We have shown figure 5 the simulated waveforms of the diode with the following reference parameters :

$$R_L = 0{,}05\ \Omega \qquad R_{on} = 2\ \Omega \qquad G = 1000$$

As soon as the input voltage is positive, the diode is on (R_{bin}=R_{on}). The diode current increases, reaches a maximum and decreases. The diode is off only when the charges have vanished (I_D=I_{RRM}). The diode recovery is soft. Current i_D tends to zero according to the following exponential law :

$$I_D = I_{RRM} \exp\left(\frac{-t}{L/R_L}\right) \qquad (10)$$

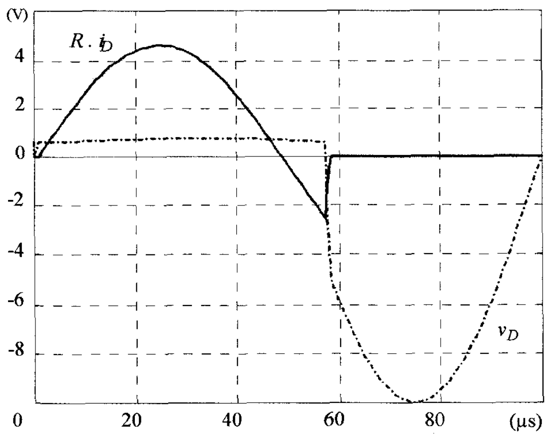


Figure 5 : Simulation curves

III.3. IDENTIFICATION METHODS

Nowadays, the identification of the physical components in different systems is a classical technic in control. We can find a lot of identification functions in the toolboxes of the software integrated Matlab. Their classification depends on the state nature (discrete or continue) and also on the description of the choice's model whose type can be :

- Black box (free parameters),

- Physical parameters system (linked parameters).

We always have to make the difference between the Single Input - Single Output models (SISO) and the Multiple Input - Multiple Output Models (MIMO) as well.

This paper describes the analysis based on the identification of a parameter SISO off line system. Two methods are used :

- ARX (Auto Regressive with eXternal input),

- PEM (Predictive Error Model).

ARX Method

Least square method is the principle of this function. In Matlab, the model is described as follows :

$$A(q)Y(t) = B(q)U(t) + v(t) \quad (11)$$

where : $A(q) = I + A_1 q^{-1} + \ldots + A_m q^{-m}$

$B(q) = B_0 + B_1 q^{-1} + \ldots + B_m q^{-m}$

A(q) is a discret equation, q^{-1} is the delay operation.

The least square estimation is described as follows :

$$\hat{\theta} = \left[\frac{1}{N}\sum_{t=1}^{N}\varphi(t)\varphi^T(t)\right]^{-1}\frac{1}{N}\sum_{t=1}^{N}\varphi(t)Y(t) \quad (12)$$

$\hat{\theta}$ contains the parameter vectors of A(q) and B(q) :

$$\hat{\theta}^T = \left[A_1 \ldots A_m\ B_1 \ldots B_m\right],$$

$\varphi(t)$ is made by the measurement and observation vectors U(t) and Y(t) which are the observation matrix in the linear regression representation.

N consists in the number of measures.

Experimentally, there are always noises in the measurement signals. In that case, y(t) can be described by the following equation :

$$Y(t) = \varphi(t)\,\theta_R + v(t) \quad (13)$$

θ_R represents the real value parameters

$v(t)$ is the introduced noise.

With this assumption, the estimated value $\hat{\theta}$ can be expressed by :

$$\hat{\theta} = \theta_R + \left[\frac{1}{N}\sum_{t=1}^{N}\varphi(t)\varphi(t)^T\right]^{-1}\frac{1}{N}\sum_{t=1}^{N}\varphi(t)v(t) \quad (14)$$

In order to have the estimation value of $\hat{\theta}$ closest to the real value θ_R, many conditions have to be satisfied. Particulary :

- $\sum_{t=1}^{N}\varphi(t)\,\varphi(t)^T$ must be a non singular matrix.
- v must be a white noise.

If these conditions are not fulfilled, we can say that there's an estimation error.

PEM Method

Like the ARX method, the PEM method is based on the least square method. The aim is the minimisation of a quadratic criterion with respect to prediction errors defined by :

$$J_{PEM} = \frac{1}{N}\sum_{t=1}^{N} \varepsilon(t,\theta)\,\varepsilon^T(t,\theta) \qquad (15)$$

where $\varepsilon(t)$ is the error between the real output process and the output estimate process. To refine the parameters, we wish to minimise the quadratic error.

The research of the minimum value is realised by a recursive algorithm with 10 iterations.

The initial conditions are obtained with the ARX method and then we observe the evolution of the system to obtain the convergence between the internal model calculated and the real model (figure 5).

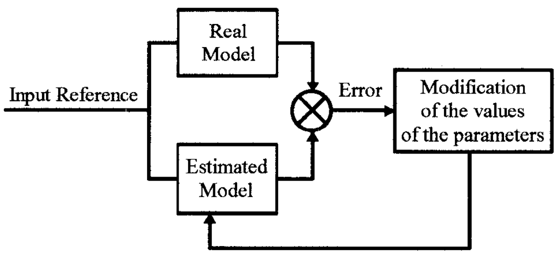


Figure 5 : Calculation Algorithm of the PEM method

The sampling period and the filter must be chosen to have signals allowing a good convergence.

III.4. RESULTS

The global model of the diode is non linear. We identify his parameters in two state : diode on ($R_{min} = R_{on}$) and diode off ($R_{min} = R_{off}$), but we cannot identify both states together.

We consider the part of signals where the diode is on. The results obtained are shown below :

Input reference parameters		Identified parameters	
Parameters value	State Matrix value	Parameters value	State Matrix value
$R_L = 0.05\ \Omega$	$A = -9.8039*10^4$	$R_L = 0.05\ \Omega$ (calculated from τ and L)	$A = -9.8039*10^4$
$L = 10^{-8}$ H	$B = 9.8039*10^4$		$B = 9.8039*10^4$
$R_{on} = 2\ \Omega$	C = 1.9598	$R_{on} = 1.9941\ \Omega$	C = 1.9540
G = 1000	D = 0.0402	$G = 9.999*10^2$	D = 0.0459

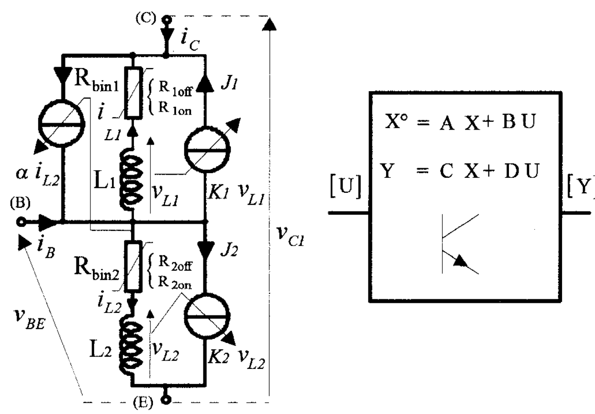
The parameters identified are very closed to the reference parameters. The error is 10^{-3}. The identification methods are validated.

For the moment, the identification of the diode parameters from experimental curves has not given any good results. This is due to the quality of the signals measured (too much noise). We hope to resolve this problem rapidly.

IV. BIPOLAR TRANSISTOR MODELLING.

A bipolar transistor model must characterise the delays in turn-on and turn-off switching. It has be done with the electrical circuit presented figure 6a. The transistor model is similar to the ebers-moll model except that the diodes are modelled as discussed in the previous section.

The transistor inner model is represented figure 6b. The state variables are the currents in the inductance L_1 and L_2, the state of R_{bin} depending on the control U and the state vector X.



a) Electrical equivalent circuit. b) Inner model

Figure 6 : Bipolar Transistor Modelling

IV.1. Equations of a Transistor IU

In the case of a transistor with an input current and an output voltage we have :

$$X = \begin{bmatrix} i_{L1} \\ i_{L2} \end{bmatrix} \qquad U = \begin{bmatrix} i_C \\ i_B \end{bmatrix} \qquad Y = \begin{bmatrix} v_{CE} \\ v_{BE} \end{bmatrix}$$

- <u>First inner model equation</u> :

From figure 6a, we can write :

$$i_C = \alpha\, i_{L_2} - i_{D_1} \quad \text{with} \quad i_{D_1} = i_{L_1} + K_1 L_1 \frac{d\, i_{L_1}}{dt}$$

$$i_{D_2} = i_C + i_B \quad \text{with} \quad i_{D_2} = i_{L_2} + K_2 L_2 \frac{d i_{L_2}}{dt}$$

So we deduce :

$$\frac{d i_{L_1}}{dt} = -\frac{1}{K_1 L_1} i_{L_1} + \frac{\alpha}{K_1 L_1} i_{L_2} - \frac{1}{K_1 L_1} i_C \qquad (16)$$

$$\frac{d i_{L_2}}{dt} = -\frac{1}{K_2 L_2} i_{L_2} + \frac{1}{K_2 L_2} i_C - \frac{1}{K_2 L_2} i_B \qquad (17)$$

So we deduce from Eqs.(16) and (17) :

$$A = \begin{bmatrix} \frac{-1}{K_1 L_1} & \frac{\alpha}{K_1 L_1} \\ 0 & \frac{-1}{K_2 L_2} \end{bmatrix} \qquad B = \begin{bmatrix} \frac{-1}{K_1 L_1} & 0 \\ \frac{1}{K_2 L_2} & \frac{1}{K_2 L_2} \end{bmatrix}$$

- <u>Second inner model equation</u> :

From figure 1a, we deduce :

$$v_{BE} = v_{D_2} = R_{bin2}\, i_{L_2} + L_2 \frac{di_{L_2}}{dt}$$

$$= \left(R_{bin2} - \frac{1}{K_2}\right) i_{L_2} + \frac{1}{K_2} i_C + \frac{1}{K_2} i_B \qquad (18)$$

$$v_{BC} = v_{D_1} = R_{bin1}\, i_{L_1} + L_1 \frac{di_{L_1}}{dt}$$

$$= \left(R_{bin1} - \frac{1}{K_1}\right) i_{L_1} + \frac{\alpha}{K_1} i_{L_2} - \frac{1}{K_1} i_C \qquad (19)$$

$$v_{CE} = v_{CB} + v_{BE} = -v_{BC} + v_{BE} \qquad (20)$$

We deduce from Eqs.(18) and (20) :

$$C = \begin{bmatrix} \frac{1}{K_1} - R_{bin1} & R_{bin2} - \frac{1}{K_2} - \frac{\alpha}{K_1} \\ 0 & R_{bin2} - \frac{1}{K_2} \end{bmatrix} \qquad D = \begin{bmatrix} \frac{1}{K_2} + \frac{1}{K_1} & \frac{1}{K_2} \\ \frac{1}{K_2} & \frac{1}{K_2} \end{bmatrix}$$

- <u>Rbin test</u> :

In the case of the bipolar transistor, $i_{bin1} = i_{L1}$ and $i_{bin2} = i_{L2}$.

So :

$$R_{bin1} = R_{min}\, \text{sign}\,(X(1) > 0) + R_{max}\, \text{sign}\,(X(1) < 0)$$

$$R_{bin2} = R_{min}\, \text{sign}\,(X(2) > 0) + R_{max}\, \text{sign}\,(X(2) < 0)$$

IV.2. Equations of a Bipolar Transistor UI

In the case of a transistor with an input voltage and an output current, we have :

$$X = \begin{bmatrix} i_{L_1} \\ i_{L_2} \end{bmatrix} \qquad U = \begin{bmatrix} v_{CE} \\ v_{BE} \end{bmatrix} \qquad Y = \begin{bmatrix} i_C \\ i_B \end{bmatrix}$$

Arranging Eqs (16) to (20), we demonstrate that :

$$A = \begin{bmatrix} \frac{-R_{bin1}}{L_1} & 0 \\ 0 & \frac{-R_{bin2}}{L_2} \end{bmatrix} \qquad B = \begin{bmatrix} -\frac{1}{L_1} & \frac{1}{L_1} \\ 0 & \frac{1}{L_2} \end{bmatrix}$$

$$C = \begin{bmatrix} K_1 R_{bin1} - 1 & \alpha \\ 1 - K_1 R_{bin1} & 1 - K_2 R_{bin2} - \alpha \end{bmatrix} \qquad D = \begin{bmatrix} K_1 & -K_1 \\ -K_1 & K_1 + K_2 \end{bmatrix}$$

The tests on R_{bin1} and R_{bin2} are the same as previously.

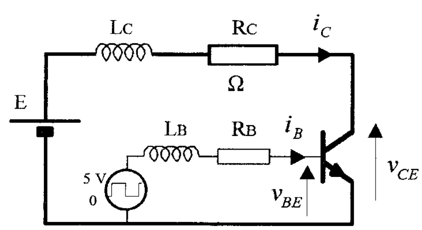
IV.3. Simulation Example

The electrical circuit of a switching transistor on a resistive load is represented figure 7a. The equations are :

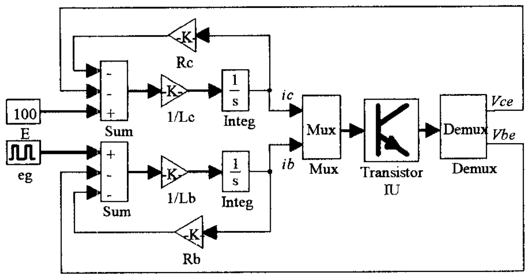
$$L_C \frac{di_C}{dt} = E - R_C I_C - V_{CE} \qquad (21)$$

$$L_B \frac{di_B}{dt} = e_g - R_B I_B - V_{BE} \qquad (22)$$

The equivalent Simulink circuit is shown figure 7b. It is deduced from Eqs. (21) and (22).



a) Electrical circuit



b) Equivalent Simulink circuit

Figure 7 : Switching of a bipolar transistor.

The corresponding simulations are shown figure 8.

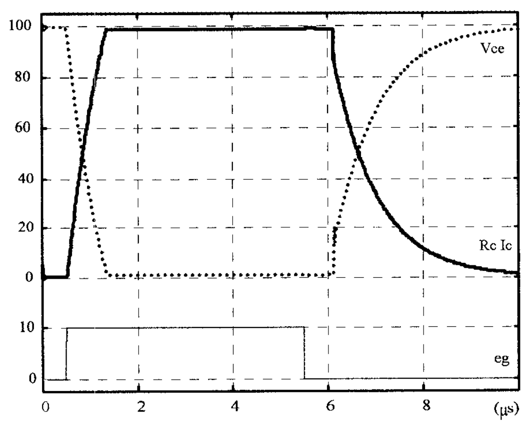


Figure 8 : Simulation of a bipolar transistor.

We can observe some delays at the turn on and turn off switching.

V. CONCLUSION

In this paper we have presented a new concept for the modelisation of power electronics components, in the goal to identify the parameters of the model directly from experimental curves.

The methodology used has been clearly highlighted. The identification of the diode parameters from simulation curves has given some very good results. Presently, we are trying to carry out the manipulation again with the experimental signals. But it is more complicated because of the noise in the signal.

In the future, we hope to identify the parameters of the bipolar transistor model from experimental curves and to develop new inner model of, MOS, IGBT etc....

The state space approach in the modelisation of power electronic circuits allows also the perfecting of regulation loops very easily. But in this case, the switches have to be simplily (2 state resistance).

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