# A Novel Method to Determine the RC Interconnect Circuit Outputs

Patricia Renault, Pirouz Bazargan-Sabet
LIP6 - University of Paris 6, 4 Place Jussieu, 75005 Paris, France
patricia.renault@lip6.fr, pirouz.bazargan-sabet@lip6.fr

**Abstract— With the process evolutions, interconnect reduction is an important problem in the verification of integrated circuit. In this paper, we present an exact method to compute the analytic time-domain response for a RC circuit including coupling capacitors under step input. Accuracy and efficiency of the method is shown for various examples.**

## I. Introduction

With deep submicron processes, some new parasitic elements appear in the interconnect of the circuit. These parasitic elements are mainly due to the increased number of metal level, thinner metal width, increased wire height versus width radio and smaller wire spacing. Typically, interconnect are modelled by resistance and capacitance. Figure 1 gives an example of gates connected through a RC-circuit. Some coupling (or crosstalk) capacitance can

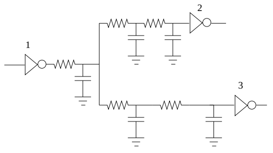


Fig. 1. Example of RC-circuit

be added between two neighbouring wires (fig. 2). The set of RC elements used to model the wire is called RC network. Theses networks can have two significant impacts on circuit: timing or functional failure. When the output of the gate 1 makes a transition a current is injected through the coupling capacitance on the interconnection located between the gates 4 and 5. If the gate 4 is making a transition, the injected current can accelerate or decelerate the transition causing timing failure. If the gate 4 is in a steady state, the injected current can generate a noise, which can produce a functional failure if it is not rapidly absorbed by the emitter 4.

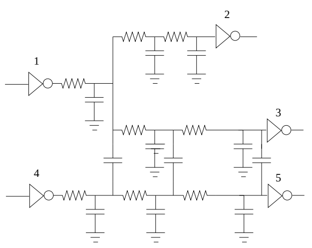


Fig. 2. Example of RC-circuit with coupling capacitance

Nowadays, these parasitic elements can not be neglected and must be taken into account in the design methodologies, design and verification tools. [2] and [10] propose some methodologies to reduce the impact of parasitic elements. New place and route tools are developped in order to take into account the crosstalk phenomenon: [11] proposes a new global timing placement, [12] presents an algorithm to solve crosstalk violation elimination problem and in [4] a technique is presented considering the river routing crosstalk constraints. Motivated by the circumstance, [13] exposes a method to capture crosstalk-induced noisy waveform for static timing analysis. Several methods have been proposed to evaluate the crosstalk noise due to RC-circuit with crosstalk coupling capacitance [3] [8] [9]. These methods are quite simple but they can suffer from a lake of accuracy.

In real-size circuit, the RC-circuits can comprise an important number of parasitic elements (near 1000 or 10000 resistances and ground capacitors) and have a complex topology with crosstalk coupling capacitors. Thus, the parasitic RC-circuits can not be directly taken into account in verification tools. We propose to reduce the complexity of the RC-circuit by reducing the RC-circuits. The reduction consists in modelling the original RC-circuit with a small number of elements while maintaining the circuit's characteristics. This has been an area of considerable research over the last few years [1][7][14][6][15][16]. The method proposed by W. C. Elmore [1] in 1948 and modified by J. Rubinstein [7] proposes to estimate the 50% propagation delay for RC-tree without coupling capacitance. This method can not easily expressed the propagation delay for RC-network with coupling capacitance. Recently, the AWE method [14] and its derivative, the first three moments method [6], give an approximated expression of a RC-network output by matching the first $n$ (or three) moments of the impulse response. The experience shows that these methods can generate some important error with coupling capacitance. In addition, it is well-known that AWE can be unstable. The PVL [15] and PRIMA [16] methods propose, to match the first n moments of the RC-networks with different mathematical techniques. Due to the complexity of these methods and to the difficulty of finding the number of moments which give a satisfying accuracy/computation time ratio, it is really difficult to use these methods.

This paper describes an original approach for determining the exact analytic waveform of the RC-circuit outputs and is organised as follow. Section 2 gives a mathematical formulation of the problem. In section 3, our method is developed, based on the time-domain equations. Section 4 shows some results. In section 5, we discuss the computation time according to the accuracy of the method. Finally, some conclusions and future works are offered in section 6.

## II. Problem Formulation

Let's consider a RC-circuit composed by $m+1$ nodes numbered of 0 to $m$. Some nodes, called input nodes, represent the initial point of a wire. These nodes, connected to the output of gates, have a known voltage. Some nodes, called output nodes, represent the final point of a wire. They are connected to the input of a gate. Except the input and output nodes, there are internal nodes, which are not connected to gates.

Let's consider a node $i$ connected to the node $l$ through a resistor $R_{il}$ and to the node $k$ through a capacitor $C_{ik}$. $C_{i0}$ is the ground capacitor (see figure 3).

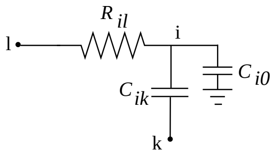


Fig. 3. Node $i$ connected to several nodes

Let's note $x_i(t)$ the expression of the voltage of i in time domain. Then

$$\frac{1}{R_{il}}(x_i - x_l) + C_{i0}x_i' + C_{ik}(x_i' - x_k') = 0 \qquad (1)$$

The general form of this equation can be written as:

$$G_i.x_i + C_i.x_i' = \sum_{l=0}^{m} G_{il}.x_l + \sum_{k=0}^{m} C_{ik}.x_k' \qquad (2)$$

where

- $G_i$ is the total conductance of node $i$: $G_i = \sum_{l=0}^{m} \frac{1}{R_{il}}$
- $C_i$ is the total capacitance of node $i$: $C_i = \sum_{k=0}^{m} C_{ik}$
- $G_{il}$ is the conductance between node $i$ and $l$: $G_{il} = \frac{1}{R_{il}}$. When the nodes $i$ and $l$ are not connected by a resistor, $G_{il} = 0$ (particularly $G_{ii} = 0$).
- $C_{ik}$ is the capacitance between node $i$ and $k$. When the nodes $i$ and $k$ are not connected by a capacitor, $C_{ik} = 0$ (particularly $C_{ii} = 0$)

We note 0 the ground voltage. 1 to $n$ are the internal nodes and the output nodes. The input nodes are indexed from $n+1$ to $m$.

With the equation 2, a RC-circuit composed by $m+1$ nodes is characterised, in the time-domain, by a system of $n$ equations ($S_t$).

$$S_t \begin{cases} G_1.x_1(t) + C_1.x_1'(t) &= \sum_{j=0}^{m} G_{1j}.x_j(t) + \sum_{j=0}^{m} C_{1j}.x_j'(t) \\ \vdots & \\ G_i.x_i(t) + C_i.x_i'(t) &= \sum_{j=0}^{m} G_{ij}x_j(t) + \sum_{j=0}^{m} C_{ij}.x_j'(t) \\ \vdots & \\ G_n.x_n(t) + C_n.x_n'(t) &= \sum_{j=0}^{m} G_{nj}.x_j(t) + \sum_{j=0}^{m} C_{nj}.x_j'(t) \end{cases} \qquad (3)$$

Note that for each $i, j$ $C_{ij} = C_{ji}$ and $G_{ij} = G_{ji}$.

It is well known that the resolution of this system gives, for each node, the following solution

$$x_i(t) = \sum_{k=1}^{n} a_{ik}.e^{-h_k.t} + a_{i0} \qquad (4)$$

where $a_{i0}$ is the voltage when $t \to +\infty$ of the node $i$, $a_{ik}$ a coefficient in Volt and $h_k$ a frequency.

The aim of this work is to propose a method to determine the coefficients $a_{ik}$ and the frequencies $h_k$.

## III. Determining the Analytic Waveform

The method consists in resolving the system $S_t$. Note that no approximation is used to solve this system.

First, we will consider the simple case of a RC-circuit without coupling capacitance. We determine the frequencies $h_i$ and then the coefficients $a_{ik}$ for each node. After, we will extend our method to real RC-circuits with coupling capacitance.

### A. Determining the frequencies without coupling capacitances

To determine the $n$ frequencies we try to find $n$ equations. A first equation is given by the value at time 0 of the node i

$$a_{i0} + a_{i1} + \cdots + a_{in} = x_i(0) \qquad (5)$$

so

$$a_{i1} + a_{i2} + \cdots + a_{in} = x_i(0) - a_{i0} \qquad (6)$$

in writing $x_{i0} = x_i(0) - a_{i0}$

$$a_{i1} + a_{i2} + \cdots + a_{in} = x_{i0} \qquad (7)$$

where each $a_{ij}$ are unknown and $x_{i0}$ is known.

However, we know not only the initial value of node $i$ but also the initial values of all the nodes of the circuit. Thus, we can determine the initial value of the first derivative of the node $i$, $x_i'(0)$ :

$$\begin{array}{rcl} G_i.x_i(t) + C_i.x_i'(t) & = & \sum\limits_{j=0}^{m} G_{ij}.x_j(t) \\ x_i'(0) & = & \frac{1}{C_i}(\sum\limits_{j=0}^{m} G_{ij}.x_j(0) - G_i.x_i(0)) \end{array} \tag{8}$$

On the other hand, from the expression of the node $i$, $x_i(t)$, we obtain

$$x_i'(t) = -\sum_{k=1}^{n} a_{ik} h_k.e^{-t.h_k} \tag{9}$$

which gives: $x_i'(0) = -\sum\limits_{k=1}^{n} a_{ik} h_k$.

We note $x_{i1}$, $x_i'(0)$. We obtain a second equation

$$a_{i1}.h_1 + a_{i2}.h_2 + \cdots + a_{in}.h_n = -x_{i1} \tag{10}$$

By the same way, we obtain a set of equations where $a_{i1}, \cdots, a_{in}$ and $h_1, \cdots, h_n$ are the variables.

We take the $n+1$ firsts equations

$$S_{der} \left\{ \begin{array}{lllll} \sum\limits_{j=1}^{n} h_j^0 a_{ij} & = & (-1)^0 x_{i0} & = & (-1)^0 (x_i(0) - a_{i0}) \\ \sum\limits_{j=1}^{n} h_j^1 a_{ij} & = & (-1)^1 x_{i1} & = & (-1)^1 x_i^{(1)}(0) \\ \vdots \\ \sum\limits_{j=1}^{n} h_j^p a_{ij} & = & (-1)^p x_{ip} & = & (-1)^p x_i^{(p)}(0) \\ \vdots \\ \sum\limits_{j=1}^{n} h_j^n a_{ij} & = & (-1)^n x_{in} & = & (-1)^n x_i^{(n)}(0) \end{array} \right. \tag{11}$$

Unfortunately, $S_{der}$ is a non-linear system. We try to modify this system to make it linear by eliminating the variable $a_{ij}$.

First, $a_{i1}$ is eliminated by subtracting the line $p$ and the line $p-1$. We obtain the $n$ equations

$$\left\{ \begin{array}{lll} \sum\limits_{j=2}^{n} h_j^0 (h_j - h_1) a_{ij} & = & (-1)^1 (x_{i1} + h_1 x_{i0}) \\ \vdots \\ \sum\limits_{j=2}^{n} h_j^{p-1} (h_j - h_1) a_{ij} & = & (-1)^p (x_{ip} + h_1 x_{ip-1}) \\ \vdots \\ \sum\limits_{j=2}^{n} h_j^{n-1} (h_j - h_1) a_{ij} & = & (-1)^n (x_{in} + h_1 x_{in-1}) \end{array} \right. \tag{12}$$

Then, in the same way, we proceed to the elimination of the other variables $a_{ij}$. After $n-1$ iterations, we obtain the two following equations

$$\left\{ \begin{array}{ll} a_{in} \prod\limits_{j=1}^{n-1} (h_n - h_j) = & (-1)^{n-1} (x_{in-1} + (h_1 + \cdots + h_{n-1}) x_{in-2} \\ & +(h_1 h_2 + \cdots + h_{n-2} h_{n-1}) x_{in-3} \\ & + \cdots + h_{n-1} \cdots h_1 x_{i0}) \\ a_{in} h_n \prod\limits_{j=1}^{n-1} (h_n - h_j) = & (-1)^n (x_{in} + (h_1 + \cdots + h_{n-1}) x_{in-1} \\ & +(h_1 h_2 + \cdots + h_{n-2} h_{n-1}) x_{in-2} \\ & + \cdots + h_{n-1} \cdots h_1 x_{i1}) \end{array} \right. \tag{13}$$

Finally, the last iteration gives

$$\begin{array}{l} 0 = x_{in} + (h_1 + h_2 + \cdots + h_n) x_{in-1} \\ +(h_1 h_2 + \cdots + h_1 h_n + h_2 h_3 + \cdots + h_2 h_n + \cdots) x_{in-2} \\ + \cdots + (h_1 h_2 \cdots h_n) x_{i0} \end{array} \tag{14}$$

It can be written as

$$\begin{array}{l} 0 = x_{in} + \sum\limits_{i=1}^{n} h_i.x_{in-1} + \sum\limits_{i=1}^{n} \sum\limits_{j=i+1}^{n} h_i h_j.x_{in-2} \\ + \cdots + \prod\limits_{i=1}^{n} h_i.x_{i0} \end{array} \tag{15}$$

We note

$$\left\{ \begin{array}{ll} \sigma_{n-1} & = \sum\limits_{i=1}^{n} h_i \\ \sigma_{n-2} & = \sum\limits_{i=1}^{n} \sum\limits_{j=i+1}^{n} h_i h_j \\ \vdots \\ \sigma_0 & = \prod\limits_{i=1}^{n} h_i \end{array} \right. \tag{16}$$

The equation 15 can be written as

$$\sigma_{n-1} x_{in-1} + \sigma_{n-2} x_{in-2} + \cdots + \sigma_0 x_{i0} = -x_{in} \tag{17}$$

Applying this equation on each node, we obtain $n$ equations with $n$ unknowns $(\sigma_i)$.

$$\left\{ \begin{array}{ccccccc} x_{10}\sigma_0 & + & \cdots & + & x_{1n-1}\sigma_{n-1} & = & -x_{1n} \\ \vdots \\ x_{n0}\sigma_0 & + & \cdots & + & x_{nn-1}\sigma_{n-1} & = & -x_{nn} \end{array} \right. \tag{18}$$

This linear system can be written as

$$\left( \begin{array}{ccc} x_{10} & \cdots & x_{1n-1} \\ \vdots & \vdots & \vdots \\ x_{n0} & \cdots & x_{nn-1} \end{array} \right) . \left( \begin{array}{c} \sigma_0 \\ \vdots \\ \sigma_{n-1} \end{array} \right) = - \left( \begin{array}{c} x_{1n} \\ \vdots \\ x_{nn} \end{array} \right) \tag{19}$$

where the $x_{ij}$ are known. $\sigma_i$ are obtained from the resolution of this system. Then, we determine the frequencies $h_i$ from $\sigma_i$.

Let's consider the polynomial $P_n$, such as

$$P_n(h) = \prod_{i=1}^{n} (h - h_i) \tag{20}$$

It can be written as

$$\begin{aligned}P_n(h) &= h^n + (-1)\sum_{i=1}^{n} h_i.h^{n-1}\\ &+(-1)^2\sum_{i=1}^{n}\sum_{j=i+1}^{n} h_ih_j.h^{n-2}\\ &\vdots\\ &+(-1)^n\prod_{i=1}^{n} h_i\\ &= h^n - \sigma_{n-1}h^{n-1} + \sigma_{n-2}h^{n-2} + \cdots + (-1)^n\sigma_0\end{aligned} \tag{21}$$

Consequently, $h_i$ can be ontained from the roots of the polynomial $P_n(h)$. We propose to use a Newton-Raphson method to find the first root of $P_n$. Then, the subsequent roots are determined iteratively by deflating $P_n$.

### *B. Determining the coefficients*

Now, all the coefficients $h_i$ are known. With the first equation of the system 13 we can determine $a_{in}$

$$\begin{aligned}a_{in}\prod_{j=1}^{n-1}(h_n-h_j) = (-1)^{n-1}(&x_{in-1}+(h_1+\cdots+h_{n-1})x_{in-2}\\ &+(h_1h_2+\cdots+h_{n-2}h_{n-1})x_{in-3}\\ &+\cdots\\ &+h_{n-1}\cdots h_1x_{i0})\end{aligned} \tag{22}$$

The system $S_{der}$ is symetrical according to $h_1, \cdots, h_n$. So for each $k \in \{1, \cdots, n\}$, we have the following equation and we can determine each coefficient $a_{ik}$.

$$\begin{aligned}a_{ik}\prod_{\substack{j=1\\ j\neq k}}^{n}(h_k-h_j) = (-1)^{n-1}(&x_{in-1}+\sum_{\substack{j=1\\ j\neq k}}^{n} h_j.x_{in-2}\\ &+\sum_{\substack{j=1\\ j\neq k}}^{n}\sum_{\substack{l=j\\ l\neq k}}^{n} h_jh_l.x_{in-3}\\ &+\cdots\\ &+\prod_{\substack{j=1\\ j\neq k}}^{n} h_jx_{i0})\end{aligned} \tag{23}$$

### *C. General case with coupling capacitance*

To resume, the proposed method is composed by four steps. We determine:

- the initial values of the derivatives of each node
- the coefficients $\sigma_0, \cdots, \sigma_{n-1}$
- the frequencies $h_i$ in using root-finding method of $P_n(h)$
- the coefficients $a_{ij}$

When the RC-circuit has some coupling capacitance, determine the derivative of the node is complex. For a node $i$ with coupling capacitance, we have:

$$G_i.x_i(t) + C_i.x_i'(t) = \sum_{j=0}^{m} G_{ij}.x_j(t) + \sum_{j=0}^{m} C_{ij}.x_j'(t) \tag{24}$$

It is not possible to compute easily $x_i'(0)$ because it depends on $x_j'(0)$.

In this section, we propose a method to transform the system with coupling capacitance in an equivalent system without coupling capacitance.

The initial system $S_t$ can be written in using matrix such as

$$G.X + C.X' = 0 \tag{25}$$

where $G$ is the conductance matrix and $C$ the capacitance matrix. We note

$$G = \begin{pmatrix} G_1 & -G_{12} & \cdots & -G_{1n} & -G_{1n+1} & \cdots & -G_{1m}\\ -G_{21} & G_2 & \cdots & -G_{2n} & -G_{2n+1} & \cdots & -G_{2m}\\ \vdots & \vdots & \ddots & \vdots & \vdots & \ddots & \vdots\\ -G_{n1} & -G_{n2} & \cdots & G_n & -G_{nn+1} & \cdots & -G_{nm}\\ 0 & 0 & \cdots & 0 & 0 & \cdots & 0\\ \vdots & \vdots & \ddots & \vdots & \vdots & \ddots & \vdots\\ 0 & 0 & \cdots & 0 & 0 & \cdots & 0 \end{pmatrix} \tag{26}$$

and

$$C = \begin{pmatrix} C_1 & -C_{12} & \cdots & -C_{1n} & -C_{1n+1} & \cdots & -C_{1m}\\ -C_{21} & C_2 & \cdots & -C_{2n} & -C_{2n+1} & \cdots & -C_{2m}\\ \vdots & \vdots & \ddots & \vdots & \vdots & \ddots & \vdots\\ -C_{n1} & -C_{n2} & \cdots & C_n & -C_{nn+1} & \cdots & -C_{nm}\\ 0 & 0 & \cdots & 0 & 0 & \cdots & 0\\ \vdots & \vdots & \ddots & \vdots & \vdots & \ddots & \vdots\\ 0 & 0 & \cdots & 0 & 0 & \cdots & 0 \end{pmatrix} \tag{27}$$

Without coupling capacitance $C$ is diagonal but with coupling capacitance the matrix is not diagonal. So, we propose to apply some addition and subtraction between line such as the matrix $C$ becomes diagonal. The operations made on $C$ have to be done on $G$. So, we obtain a new matrix equation:

$$G_{new}.X + C_{new}.X' = 0 \tag{28}$$

with

$$G_{new} = \begin{pmatrix} G_1' & -G_{12}' & \cdots & -G_{1n}' & -G_{1n+1}' & \cdots & -G_{1m}'\\ -G_{21}' & G_2' & \cdots & -G_{2n}' & -G_{2n+1}' & \cdots & -G_{2m}'\\ \vdots & \vdots & \ddots & \vdots & \vdots & \ddots & \vdots\\ -G_{n1}' & -G_{n2}' & \cdots & G_n' & -G_{n+1}' & \cdots & -G_{3m}'\\ 0 & 0 & \cdots & 0 & 0 & \cdots & 0\\ \vdots & \vdots & \ddots & \vdots & \vdots & \ddots & \vdots\\ 0 & 0 & \cdots & 0 & 0 & \cdots & 0 \end{pmatrix} \tag{29}$$

and

$$C_{new} = \begin{pmatrix} C'_1 & 0 & \cdots & 0 & 0 & \cdots & 0 \\ 0 & C'_2 & \cdots & 0 & 0 & \cdots & 0 \\ \vdots & \vdots & \ddots & \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \cdots & C'_n & 0 & \cdots & 0 \\ 0 & 0 & \cdots & 0 & 0 & \cdots & 0 \\ \vdots & \vdots & \ddots & \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \cdots & 0 & \cdots & \cdots & 0 \end{pmatrix} \quad (30)$$

which can be written as

$$S_{eq} \begin{cases} G'_1.x_1(t) + C'_1.x'_1(t) & = \sum_{j=0}^{m} G'_{1j}.x_j(t) \\ \vdots & \\ G'_i.x_i(t) + C'_i.x'_i(t) & = \sum_{j=0}^{m} G'_{ij}.x_j(t) \\ \vdots & \\ G'_n.x_n(t) + C'_n.x'_n(t) & = \sum_{j=0}^{m} G'_{nj}.x_j(t) \end{cases} \quad (31)$$

With this equivalent system, there is no coupling capacitance. It is now possible to determine $x'_i(0)$ according to $x_i(0)$ and then to apply the method proposed in this section.

## IV. Results

A prototype tool that implements all the concepts described in this article has been developed. It is composed by five steps: determine the initial values and the matrix equation, compute an equivalent capacitance matrix without coupling capacitance, find the polynomial roots and calculate the coefficients.

This prototype has been tested (on a 1 GHz PC computer with 1 GByte RAM) over two kind of RC-circuits: networks with coupling capacitance with two outputs (see fig. 4) or three outputs (see fig. 5) and a variable number of internal nodes ranging from 50 to 500.

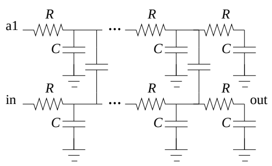


Fig. 4. Network with coupling capacitance and two ouputs

We expose here the number of bits used to represent the floats and the execution time to compute the frequencies and the coefficients according to the networks.

### A. Number Of Bits Used To Represent Floats

The experience has shown that the use of traditional float (64 bits) is not appropriate because the range of

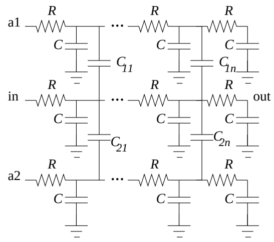


Fig. 5. Network with coupling capacitance and three outputs

the polynomial coefficient is very important. We have used the GNU Multiple Precision Arithmetic Library [5] to represent the floats with a variable number of bits. Figure 6 gives the number of bits used to represent floats for different number of internal node.

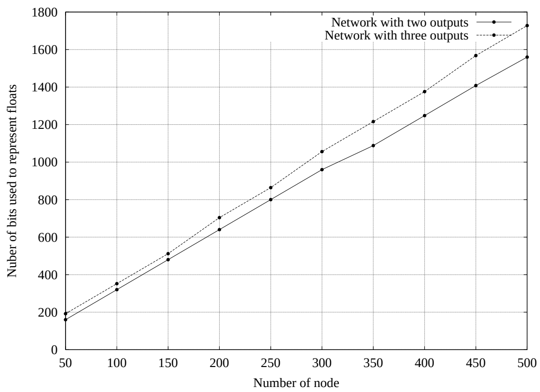


Fig. 6. Number of bits used to represents floats

We can see that the number of bits used to represent the floats is globally linear with the number of internal node.

### B. Execution Time

The figure 7 shows the execution time obtained to determine the frequencies $h_j$. For example, 50 seconds are needed to compute the frequencies of a network with three outputs and 500 internal nodes.

Figure 8 shows the execution time obtained to compute the coefficients $a_{ij}$. 25s are needed to compute the coefficients of a network with three outputs.

## V. Discussion on Computation Cost

Here, we make a brief discussion on the computation cost of the proposed method. We agree that the computation cost of the proposed method is not small compared to the main methods [1][7][6][15][16]. But, the proposed method computes without approximation the exact waveform of the output according to the input and the main

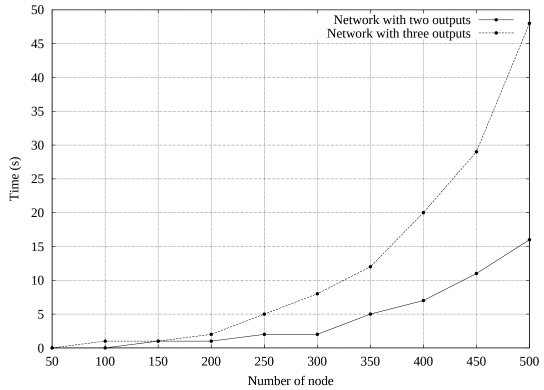


Fig. 7. Execution time for determining the frequencies

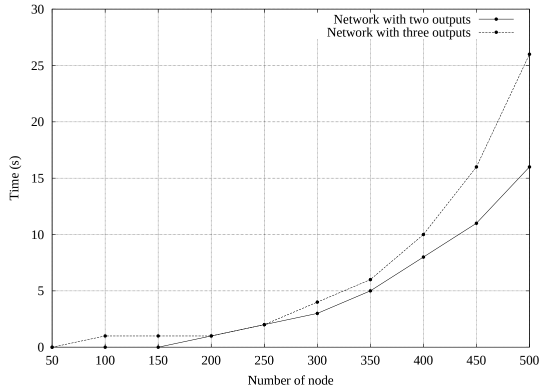


Fig. 8. Execution time for determining the coefficients

methods make some approximations and can generate an error about 15%.

## VI. CONCLUSION

Signal integrity is becoming a major issue in the verification process of high performance designs. Coupling capacitance and RC-circuit of interconnect are one of the factors that may cause timing and functional failure in the circuit.

In this paper, we have presented a method for determining the analytic waveform of a RC-circuit output without approximation, based on representation in the time-domain. The precision is exact but the computation time is important compared to an electrical simulation. On the other hand, the electrical simulation gives numerical points whereas this method gives numerical functions.

The proposed method can be improved following two directions. First, we can develop techniques to accelerate the method in decreasing the number of internal node. We can analyse the RC-circuit and merge the nodes which have the same constant time or we can split a RC-circuit in several RC-circuits. Thus, the analytic waveform of each output is calculated more quickly. The second improvement concerns the method. For very deep submicron processes the inductance of the interconnect can have a significant impact on the circuit. Thus, the method has to be modified to take into consideration the inductance of the interconnect.

## References

[1] W. C. Elmore, "The transient response of damped linear networks with particular regard to wideband amplifiers", Journal of applied physics, vol. 19, January 1948

[2] R. Saleh and D. Overhauser and S. Taylor, "Full-Chip of UDSM Designs", ICCAD, IEEE/ACM, pp. 453-460, 1998

[3] T. Xue and E.S. Kuh and D. Wang, "Post Global Routing Crosstalk Risk Estimation and Reduction", ICCAD, IEEE/ACM, pp. 302-309, 1996

[4] H. Zhou and D.F. Wong, "An Optimal Algorithm for River Routing with Crosstalk Contraints", ICCAD, IEEE/ACM, pp. 310-315, 1996

[5] http://swox.com.gmp, "GNU Multiple Precision Arithmetic Library"

[6] B. Tutuianu and F. Dartu and L. Pileggi, "An Explicit RC-Circuit Delay Approximation Based on the First Three Moments of the Impluse Response", DAC, IEEE/ACM, 1996

[7] Jorge Rubinstein and Paul Penfiels and mark A. Horowitz, "Signal delay in RC Tree Networks", IEEE transactions on computer-aided design, vol. cad-2, July, 1983

[8] A. B. Kahng and S. Muddu and D. Vidhani, "Noise model for multiple segmented coupled RC interconnects", Proceedings of the International Symposium on Quality Design, 2001

[9] Paul B. Morton and Wayne Dai, "An Efficient Sequential Quadratic Programming Formulation of Optimal Wire Spacing for Crosstalk Noise Avoidance Routing", Proceedings of the ISPD, 1999

[10] Murat Becer and Ravi Vaidyanathan and Chanhee Oh and Rajendran Panda, "Signal Integrity Management in an SoC Physical Design Flow", Proceedings of the ISPD, 2003

[11] Andrew B. Kahng and Stefanus Mantik and Igor L. Markov, "Min-Max Placement For Large-Scale Timing Optimization", Proceedings of the ISPD, 2002

[12] Hua Xiang and Kai-Yuan Chao and Martin D.F. Wong, "An ECO Algorithm for Eliminating Crosstalk Violations", Proceedings of the ISPD, 2004

[13] Masanori Hashimoto and Yuji Yamada and Hidetoshi Onodera, "Capturing Crosstalk-Induced Waveform for Accurate Static Timing Analysis", Proceedings of the ISPD, 2004

[14] Lawrence T. Pillage and Ronald A. Rohrer, "Asymptotic Waveform Evaluation for Timing Analysis", IEEE transactions on computer-aided design, vol. 9, april,1990

[15] R. W. Freund and P. Feldmann, "Reduced-Order Modeling of Large Passive Linear Circuits by Means of the SyPVL Algorithm", Proceedings of the ICCAD, 1996

[16] Altan Odabasioglu, Mustafa Celik, Lawrence T. Pileggi, "PRIMA: Passive Reduced-Order Interconnect Macromodeling Algorithm", IEEE Transactions on Computer-Aided Design of Integrated Circuits and Systems, Vol. 17, N. 8, August 1998