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Date: 12 MAR 98

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FROM: AFRL/DEOB-PA (6-6246)

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Judy L. Johnston
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Public Affairs Specialist

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Interaction Notes

Note 536

February 24, 1998

An efficient technique to calculate ideal junction scattering parameters in multiconductor transmission line networks

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Abstract

In multiconductor line network theory, ideal junctions represent the location where a cable may separate in different directions. Consequently, those junctions involve perfect connections between wires in each branch. But they may also involve short-circuits to the ground and open circuits. Because scattering parameters cannot be calculated with usual impedance or admittance matrix transformations, they have to be calculated directly. Widely inspired from an article written by A. K. Agrawal, H. M. Fowles, L. Scott and T. Simpson, in 1978, this paper presents an efficient technique to determine them. On typical examples, we show that this technique can be extended to any kind of ideal junctions. Moreover, in addition to the wide application field of this technique, the calculation time is really improved compared to *Besnier's Technique. Calculations performed with the CRIPTE code emphasized the improvement.*

Key words :

Scattering Parameters ; Multiconductor Transmission Line Networks ; Electromagnetic Topology ; Transmission Lines ; Electromagnetic Coupling ; Electromagnetic Interference

*This work has been made possible thanks to the Prospective Bureau of DGA and AFOSR.

1. Definition of the problem

Electromagnetic topology relies on the description of complex systems under the form of networks [1]. The calculation of the response of a network to an excitation is provided thanks to the famous BLT equation. The network equation is particularly well suited for ***multiconductor transmission line networks*** [2] where the BLT equation formalism finds numerous applications in EMC and EMI fields. A topological network is made of ***tubes***, supplying the propagation of the waves on the tubes, and ***junctions*** supplying the scattering of waves (figure 1-1). The concept of junction is essential in electromagnetic topology because it allows to take into account the bi-directed graphs associated to a given shielding level ([3], [4]). In particular, the notion of localized junctions can be extended to the notion of equivalent junctions. describing the entire scattering of waves on a subnetwork [5]. The extremities of the junctions connected to the tube wires are called "***ports***". Furthermore, for all tubes and junctions, we will suppose that there is a common signal reference that we call common "***ground***".

For cable networks, particular important junctions are the ones associated to the point where a given cable harness separates in several branches with different numbers of wires. In the network approach, each branch can be described as a tube. Such junctions are called "***ideal junctions***", according to the terminology used in the CRIPTE code [6]. Contrary to terminal junctions, those localized junctions do not involve any physical loads (figure 1-1). An ideal junction is a junction providing a perfect connection between the wires of the different tubes. The connection point between the wires is called a "***current node***" because Kirchoff's current law is applied on them.

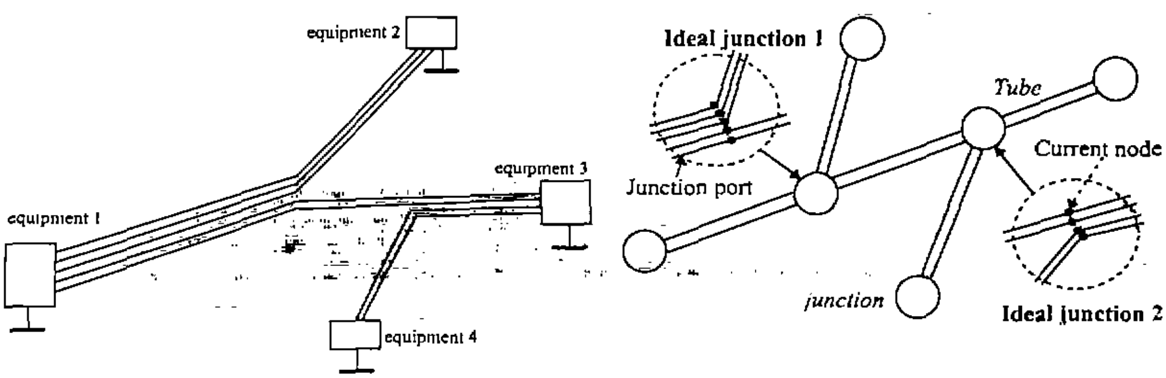


Fig. 1-1 : Example of a cable harness network and its associated topological network, involving 2 ideal junctions.

Compared to usual junctions, the particularity of such a junction is that it is not possible to calculate its scattering parameters by the usual impedance or admittance transformations [7]. Indeed, these matrices are not defined for such an ideal junction and the scattering matrix has to be calculated directly. Some examples are given in [7] for wire to wire connections between two tubes. In [8], a more general method is proposed to calculate the scattering parameters of an ideal junction. This technique requires a block analysis of the different characteristic impedance of the tubes connected to the junction. Particularly, the model accounts for multiple connections of wires at the same current node ("***fork***" current nodes). A fork current node is a current node involving at least 3 connections. For example, such a kind of connection happens in the general model of branched shielded cable harnesses in which the shield cannot be considered as the transmission line reference. In this case the connection of the equivalent wire associated to the shield can be described with a fork current node (figure 1-2).

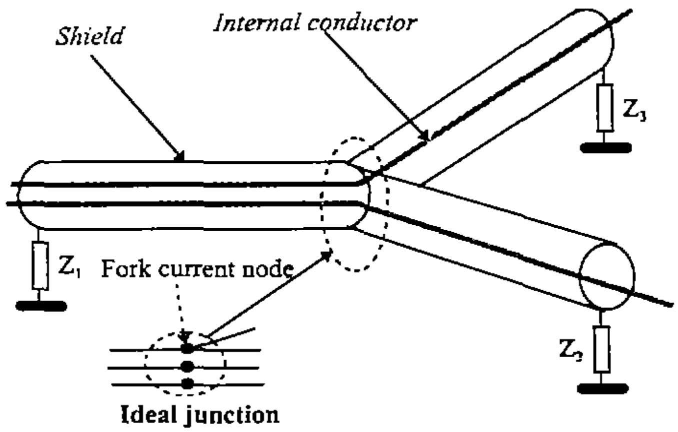


Fig. 1-2 : Modeling of the fork connection of a branched shielded harness.

Until this time, the technique described in [8] was the one applied in the CRIPTE code to calculate ideal junction scattering parameters. It has been used successfully in numerous electromagnetic topology applications. However, this technique still presented some drawbacks :

1 - The calculation time of the scattering parameters seemed to be very long for big junctions,
2 - The technique required much time and memory. Even if a stack allocation for the junction matrices had been recently introduced into the code, the number of matrices declared in the routine was still important,
3 - It was not possible to take into account ideal connections to the ground or ports remaining in open circuit.

In addition, a recent paper dealing with the optimization obtained when taking into account the sparse structure of the BLT matrix [9], emphasized the fact that the calculation of the junction scattering parameters was now more time requiring than the BLT resolution itself. The junctions used in the validation examples were essentially ideal junctions.

In an old paper written by Agrawal and al. on the analysis of simple transmission line networks [10], a very simple and efficient technique is described to determine the scattering parameters of ideal junctions. In this paper, those junctions are of great interest because, as the lines are lossless, the scattering parameters are not frequency dependent. Thus, in time domain, the reflection and transmission of waves they provide is easy to determine. In fact, such a technique to determine the scattering parameters seems to be really original and is not mentioned in usual circuit analysis references [11].

This paper deals with the analysis of the advantages of Agrawal's technique. First, we recall the equation providing the expression of the scattering parameters. Then, we show examples of the generalization of its application to ideal junctions involving short-circuits to the ground and open circuits. Finally, we demonstrate the time and memory improvements obtained on a big network, already described in [9].

2. Description of the method

2.1. Objective

In [10], the derivation of the scattering parameters of the ideal junctions is carried out via the determination of junction connection matrices C_I and C_V, respectively describing the Kirchoff's laws for currents and voltages. The whole demonstration of the calculation is based on the configuration of an ideal junction in which each wire in one given transmission line is connected to only one wire in another line. This way, the junctions considered in the article do not involve any fork current node, any short-circuit to the ground or any open circuit. However, the formulation the author gives is quite general. Here, we want to recall this formulation and to emphasize that it remains valid for more complex configurations of ideal junctions.

2.2. Possible definitions of scattering parameters

The scattering of signals at a junction is generally described by a scattering matrix ***relating incoming waves***, W^{+}, to ***outgoing waves***, W^{-}. Depending on the author, the terminology may change. The term wave is found in [7] and is equivalent to the "***combined voltage***" term used in [2]. Both terms are closely related to electromagnetic topology and multiconductor transmission line network theories.

$$W^{-} = S \cdot W^{+} \tag{1}$$

The waves are vectors defined as a combination of the voltage vector, made by all the individual voltages on each port of the junction, V, and the incoming current vector, made by the individual incoming currents on each port of the junction, I. I and V are related by a normalizing impedance matrix Z_0 in the following way :

$$W^{\pm} = V \pm Z_0 \cdot I \tag{2}$$

Thus, the value of the scattering parameters defined in (1) depends on the value of the normalizing impedance. For example, scattering parameters on 50 Ω are different from scattering parameters on 75 Ω. From a circuit point of view, the significance of scattering parameters referenced on Z_0 loads can be understood as transfer functions between the input voltage and the output voltage when all the ports are physically loaded by a Z_0 impedance network [9].

From a propagation and scattering point of view, it is important to realize that the scattering matrix gets the significance of a ***reflection coefficient***, only if the scattering parameters are referenced to the characteristic impedance matrix Z_c of all the transmission lines connected to the junction. It is only in this case that the incoming and outgoing waves defined in (1) are the same as the waves propagating on the connected lined. In [8], such scattering parameters are called "***topological***" scattering parameters because they are the one occurring in the BLT equation formulation.

From this point in this document, we will always consider that the reference impedance Z_0 is equal to Z_c and we will only consider topological scattering parameters.

In [10], the scattering parameters relate an ***input voltage*** vector, $V^{(in)}$, to a ***reflected voltage*** vector, $V^{(re)}$. These definitions are generally used in the theory of transmission lines [12] :

$$V^{(re)} = S \cdot V^{(in)} \tag{3}$$

Nevertheless, this definition does not modify the value of the scattering parameters. Indeed, in each point of the line, the voltage and the current vectors mentioned in (2) are given by the following combination of input and reflected voltage and current vectors :

$$V = V^{(in)} + V^{(re)} \tag{4}$$

and

$$I = I^{(in)} - I^{(re)} \tag{5}$$

The combination of (4) and (5) with (2) easily leads to :

$$V^{(in)} = \frac{V + Z_c \cdot I}{2} = \frac{W^{+}}{2} \tag{6}$$

and

$$V^{(re)} = \frac{V - Z_c \cdot I}{2} = \frac{W^-}{2} \qquad (7)$$

Because incoming and outgoing waves are related to input and reflected voltages with the same scalar coefficient, the definitions mentioned in (1) and (3) lead to the same value of the scattering matrix. A similar analysis can be made with the definitions of waves in microwave techniques where the factor "2" is replaced by a factor " $\sqrt{2}$ " for energetic considerations [12].

2.3. Determination of junction connection scattering matrices

2.3.1. Application of the Kirchoff's law

The Kirchoff's current law stipulates that the sum of all the currents at the ports of the junction must be equal to zero. Moreover, in the case of ideal junctions, we can say that the sum of all the currents at a given current node, "nc", is equal to zero. This can be summarized by :

$$\sum_j I_j^{nc} = 0 \qquad (8)$$

where I_j^{nc} represents the current on a port "j" connected to the current node "nc". It is easy to realize that (8) applies to any kind of fork current node. But it also applies for open circuit on a port "j". In this case, it is still possible to consider that we have a current node associated to the port "j" and the zero current on this port makes (8) become :

$$I_j^{nc} = 0 \qquad (9)$$

Let us call "N_c" the number of such (8) and (9) equations. If we put together all the "N_c" equations, it is possible to derive the following matrix equation, involving the current vector I :

$$C_I \cdot I = 0 \qquad (10)$$

The size of the matrix is N_cxN, where "N" is the total number of ports of the junction. Let us consider a current node, "nc", and a port "j". Then the definition of the matrix becomes :

- $C_I(nc,j) = 0$, if the port "j" is not connected to the current node "nc",
- $C_I(nc,j) = 1$, if the port "j" is connected to the current node "nc".

For a given current node, the number of non-zero terms is equal to the number of ports connected to the current node. In the particular case of an open circuit on the port "j", the line of C_I dealing with the associated current node "nc", contains only one non zero term, $C_I(nc,j)$.

2.3.2. Voltage relations

As we did for currents, it is possible to establish similar relations for voltages. For a given current node, all the voltages of the ports connected to "nc" are equal. If V_i^{nc} and V_j^{nc} are the voltages on two ports connected to the same current node, "nc", we can write :

$$V_i^{nc} - V_j^{nc} = 0 \qquad (11)$$

Such a relation is still valid in the particular case of a short-circuit to the ground. If "i" is such a port connected to the ground, (11) will be written in the following simplified form :

$$V_i^{nc} = 0 \qquad (12)$$

Let us call "N_v" the number of such (11) and (12) equation types. It is important to notice that :

$$N_c + N_v = N \qquad (13)$$

Consequently the sum of the number of the Kirchoff's current and voltage equations is equal to the number of ports of the junction.

Putting all the (11) and (12) equations together, we obtain the following matrix relation, counterpart of (10):

$$C_v \cdot V = 0 \qquad (14)$$

2.3.3. Derivation of the scattering parameters

From (2), it is easy to express the voltage and current vectors as a function of the incoming and outgoing waves. Thus, introducing the characteristic admittance, Y_c, as the inverse of the characteristic impedance Z_c of the tubes connected to the junction, we find an another way to demonstrate (4) and (5) :

$$Y_c = Z_c^{-1} \qquad (15)$$

$$V = \frac{W^+ + W^-}{2} = V^{(in)} + V^{(re)} \qquad (16)$$

and

$$I = Y_c . \frac{W^+ - W^-}{2} = I^{(in)} - I^{(re)} \qquad (17)$$

Combining (14) and (16) on one side, and (10) and (17) on the other side, it is possible to write the following two relations between incoming and reflected waves :

$$C_v \cdot W^+ = -C_v \cdot W^- \qquad (18)$$

and

$$C_I \cdot Y_c \cdot W^+ = C_I \cdot Y_c \cdot W^- \qquad (19)$$

(18) describes a linear system with "N_v" lines, whereas (19) describes a linear system with "N_c" lines. But both relate the N dimension W^+ and W^- wave vectors. Taking advantage of (13), (18) and (19) can be put together in a square NxN linear system :

$$\begin{bmatrix} -C_v \\ C_I \cdot Y_c \end{bmatrix} \cdot W^- = \begin{bmatrix} C_v \\ C_I \cdot Y_c \end{bmatrix} \cdot W^+ \qquad (20)$$

Comparing (1) with (20) the definition of the scattering matrix, S, naturally appears :

$$S = \begin{bmatrix} -C_v \\ C_I \cdot Y_c \end{bmatrix}^{-1} \cdot \begin{bmatrix} C_v \\ C_I \cdot Y_c \end{bmatrix} \qquad (20)$$

3. Examples

3.1. General ideas

In this section, we want to demonstrate the wide field of applications of (20). In particular, it is still valid for ideal junctions involving short-circuits to the ground and open circuits. The simple examples we have chosen here are limited to 3 ports but they do not remove any generality when (20) is applied on bigger junctions. To make the examples more self explanatory, we have chosen configurations with diagonal characteristic impedance matrix Z_c. The diagonal terms, Zc_i, are scalar and are the only non-zero terms of the matrix. This matrix corresponds to a configuration where each port "i" is connected to a single wire transmission line having a scalar characteristic impedance, Zc_i, or a characteristic admittance, Yc_i (figure 3-1). This means that the different wires connected to the junction are not coupled. Such a configuration can be obtained considering 3 shielded cables whose shields are connected to the common ground reference.

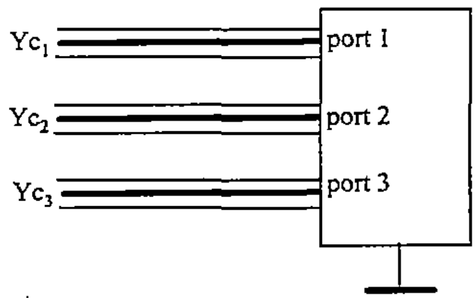
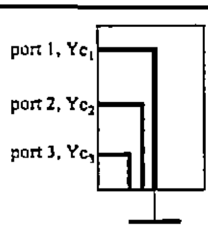


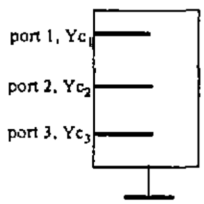
Figure 3-1 : Three port junction connected to three non coupled transmission lines

3.2. Short-circuit junction

	C_I	C_V	S
	$N_c = 0$	$N_v = 3$ $\begin{pmatrix}1&0&0\\0&1&0\\0&0&1\end{pmatrix}$	$S = \begin{pmatrix}-1&0&0\\0&-1&0\\0&0&-1\end{pmatrix}^{-1} \cdot \begin{pmatrix}1&0&0\\0&1&0\\0&0&1\end{pmatrix} = \begin{pmatrix}-1&0&0\\0&-1&0\\0&0&-1\end{pmatrix}$ (21)

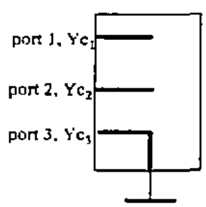
We find the well-known result that a short-circuit reflects entirely an incident wave with a phase opposition. The very important point to remember is that the matrix of a short-circuit is independent of the characteristic impedance of the transmission lines connected to its port.

3.3. Open circuit junction

	C_I	C_V	S
	$N_c = 3$ $\begin{pmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{pmatrix}$	$N_v = 0$	$S = \begin{pmatrix} Yc_1 & 0 & 0 \\ 0 & Yc_2 & 0 \\ 0 & 0 & Yc_3 \end{pmatrix}^{-1} \cdot \begin{pmatrix} Yc_1 & 0 & 0 \\ 0 & Yc_2 & 0 \\ 0 & 0 & Yc_3 \end{pmatrix} = \begin{pmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{pmatrix}$ (22)

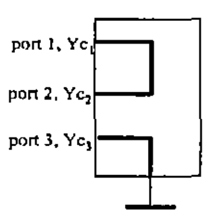
We find the well-known result that an open circuit reflects entirely an incident wave without any phase opposition. As for the short-circuit, the same remark on the non-dependency on the characteristic impedance can be made.

3.4. Combined short-circuit and open circuit junction

	C_I	C_V	S
	$N_c = 2$ $\begin{pmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \end{pmatrix}$	$N_v = 1$ $\begin{pmatrix} 0 & 0 & 1 \end{pmatrix}$	$S = \begin{pmatrix} 0 & 0 & -1 \\ Yc_1 & 0 & 0 \\ 0 & Yc_2 & 0 \end{pmatrix}^{-1} \cdot \begin{pmatrix} 0 & 0 & 1 \\ Yc_1 & 0 & 0 \\ 0 & Yc_2 & 0 \end{pmatrix} =$ $\begin{pmatrix} 0 & Zc_1 & 0 \\ 0 & 0 & Zc_2 \\ -1 & 0 & 0 \end{pmatrix}^{-1} \cdot \begin{pmatrix} 0 & 0 & 1 \\ Yc_1 & 0 & 0 \\ 0 & Yc_2 & 0 \end{pmatrix} = \begin{pmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & -1 \end{pmatrix}$ (23)

As in the previous cases of short-circuit and open circuit junctions, we notice that the scattering parameters are independent of the characteristic impedance.

3.5. Combined short-circuit and transmission junction

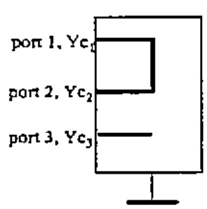
	C_I	C_V	S
	$N_c = 1$ $\begin{pmatrix} 1 & 1 & 0 \end{pmatrix}$	$N_v = 2$ $\begin{pmatrix} 1 & -1 & 0 \\ 0 & 0 & 1 \end{pmatrix}$	$S = \begin{pmatrix} -1 & 1 & 0 \\ 0 & 0 & -1 \\ Yc_1 & Yc_2 & 0 \end{pmatrix}^{-1} \cdot \begin{pmatrix} 1 & -1 & 0 \\ 0 & 0 & 1 \\ Yc_1 & Yc_2 & 0 \end{pmatrix} =$ $\begin{pmatrix} \frac{Yc_1 - Yc_2}{Yc_1 + Yc_2} & \frac{2 \cdot Yc_2}{Yc_1 + Yc_2} & 0 \\ \frac{2 \cdot Yc_1}{Yc_1 + Yc_2} & \frac{Yc_2 - Yc_1}{Yc_1 + Yc_2} & 0 \\ 0 & 0 & -1 \end{pmatrix}$ (24)

This example clearly demonstrates that a topological scattering matrix is not symmetric. This is related to the non symmetric property of the reflection coefficient. Moreover, the resultant matrix depends on the characteristic impedance matrix, except on the short-circuited port for which the scattering parameter remains equal to -1. The absence of symmetry and the dependence on the characteristic matrix are usual properties of scattering matrices sometimes forgotten by several users ([7], [12]). Nevertheless, if the characteristic impedance is the same on port 1 and port 2, the scattering matrix becomes :

$$S = \begin{pmatrix} 0 & 1 & 0 \\ 1 & 0 & 0 \\ 0 & 0 & -1 \end{pmatrix}$$

Because the wave does not see any mismatching between port 1 and port 2, it is entirely transmitted ($S_{12} = S_{21} = 1$).

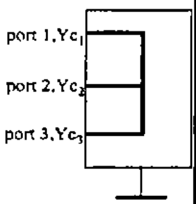
3.6. Combined open circuit and transmission junction

	C_I	C_V	S
	$N_c = 2$ $\begin{pmatrix} 0 & 0 & 1 \\ 1 & 1 & 0 \end{pmatrix}$	$N_v = 1$ $\begin{pmatrix} 1 & -1 & 0 \end{pmatrix}$	$S = \begin{pmatrix} -1 & 1 & 0 \\ 0 & 0 & Yc_3 \\ Yc_1 & Yc_2 & 0 \end{pmatrix}^{-1} \cdot \begin{pmatrix} 1 & -1 & 0 \\ 0 & 0 & Yc_3 \\ Yc_1 & Yc_2 & 0 \end{pmatrix} = \begin{pmatrix} \frac{Yc_1 - Yc_2}{Yc_1 + Yc_2} & \frac{2 \cdot Yc_2}{Yc_1 + Yc_2} & 0 \\ \frac{2 \cdot Yc_1}{Yc_1 + Yc_2} & \frac{Yc_2 - Yc_1}{Yc_1 + Yc_2} & 0 \\ 0 & 0 & 1 \end{pmatrix}$ (25)

As for the combination of a transmission and a short-circuit, the same comments apply in this case. Particularly, one will notice that the first 2x2 block is similar to the one of (24). In the particular case where the characteristic impedance is the same on port 1 and port 2 we obtain :

$$S = \begin{pmatrix} 0 & 1 & 0 \\ 1 & 0 & 0 \\ 0 & 0 & 1 \end{pmatrix}$$

3.7. Fork junction

	C_I	C_V	S
	$N_c = 1$ $(1 \;\; 1 \;\; 1)$	$N_v = 2$ $\begin{pmatrix} 1 & -1 & 0 \\ 1 & 0 & -1 \end{pmatrix}$	$S = \begin{pmatrix} -1 & 1 & 0 \\ -1 & 0 & 1 \\ Yc_1 & Yc_2 & Yc_3 \end{pmatrix}^{-1} \cdot \begin{pmatrix} 1 & -1 & 0 \\ 1 & 0 & -1 \\ Yc_1 & Yc_2 & Yc_3 \end{pmatrix} = \frac{1}{Yc_1 + Yc_2 + Yc_3} \cdot \begin{pmatrix} Yc_1 - Yc_2 - Yc_3 & 2 \cdot Yc_2 & 2 \cdot Yc_3 \\ 2 \cdot Yc_1 & Yc_2 - Yc_1 - Yc_3 & 2 \cdot Yc_3 \\ 2 \cdot Yc_1 & 2 \cdot Yc_2 & Yc_3 - Yc_1 - Yc_2 \end{pmatrix}$ (26)

Once again, the matrix is non-symmetric and depends on the characteristic impedance. Nevertheless, if the characteristic impedance of the three connected transmission lines is the same, we find :

$$S = \begin{pmatrix} 1/3 & 2/3 & 2/3 \\ 2/3 & 1/3 & 2/3 \\ 2/3 & 2/3 & 1/3 \end{pmatrix}$$

The matrix is now symmetric and independent of the characteristic impedance. One third of the signal is reflected on the input port ($S_{ii} = 1/3$), while the two other thirds are transmitted on the two other ports ($S_{ij} = 1/3$).

4. Improvements

4.1. Memory improvements

The technique described in this document has been implemented in the CRIPTE code. Compared to the old technique based on [8], the memory requirement has been significantly reduced.

In the code, all numerical implementations are made in FORTRAN 77. Recent memory optimizations, based on the used of a stack file allow an allocation of the tables close to the dynamic allocation provided in the FORTRAN 90. This way, it is possible to compare the memory requirement of the old method and the new method for a N port junction, provided that the sizes of all the tables are exactly matched to the number or ports.

The old method required :

- 8 double complex matrices NxN,
- 3 double complex vectors N,
- 3 integer matrices NxN.

Consequently the number of bytes required was equal to $8.N^2.16+3.N^2.2+3.N.16 \# 134.N^2$ for N large.

The new method requires :

- 3 double complex matrices NxN,
- 2 integer matrices N.

So the number of bytes required is equal to $3.N^2.16+2.N^2.2$ # $52.N^2$ for N large. The improvement is new close to 40 %. For instance, if the junction contains 60 ports as in the next example, the memory requirement will be :

- 485 kilo-bytes with the old method, and
- 187 kilo-bytes for the new method.

4.2. Calculation time improvements

To give an idea of the improvement obtained on the calculation time, we have chosen the same example as the fork one described in [9]. The network is made of 19 tubes organized as an assembling of junctions with three connected tubes (figure 4-1). Each tube is 1 meter long and contains 20 wires. The entire network contains 11 terminal junctions equal to 50 Ω loads applied on each wire and 9 ideal junctions. All the ideal junctions are similar and deal with the simple wire-to-wire connection on the three tubes connected to it. The internal topology of the 60 port junction is represented on figure 4-2.

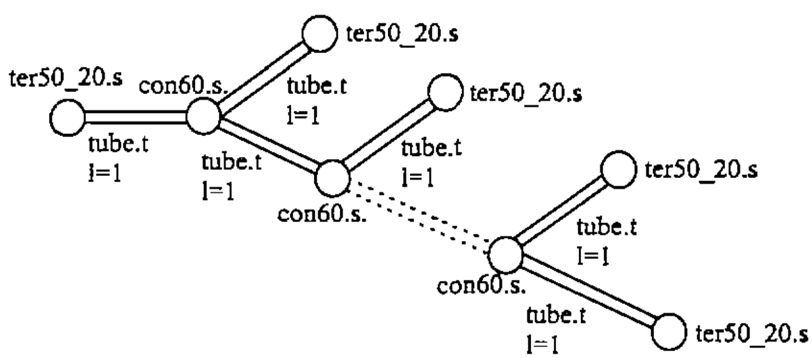


Fig. 4-1 : Fork network.]

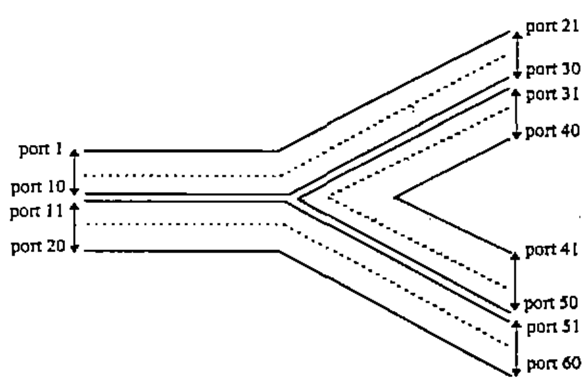


Fig. 4-2 : Internal topology of the internal junction.

As in [9], the calculation of the BLT equation on this network has been achieved when a 1 Volt localized generator is applied at the end of the first wire of the first tube. The calculation time for the different steps of the BLT equation construction and resolution has been checked, particularly, the time to calculate the scattering parameters of the junctions.

Table 4-1 shows the calculation time obtained with four different versions of the CRIPTE code :

- "***Old_non_opt***" : ideal junctions calculated according to [8] : no optimization of the compilation,
- "***New_non_opt***": ideal junctions calculated according to [10] : no optimization of the compilation,
- "***Old_opt***": ideal junctions calculated according to [8] : optimization of the compilation ("-O3" option),
- "***New_ opt***": ideal junctions calculated according to [10] : optimization of the compilation ("-O3" option).

The calculations have been carried out on the same SUN SPARK ONE work-station, with 64 Mbytes memory, as in [9]. Of course, the junction calculation times presented in table 4-1 do not apply only for ideal junctions. Nevertheless, despite the fact that the network contains only 9 ideal junctions over the 20 junctions, we notice that the calculation time is widely improved with the new technique. This means that the old ideal junction scattering parameter technique mainly imposed the calculation time on the junctions.

With no optimization of the compiler, the calculation time is improved by 8 %. With the "-O3" compiler optimization, the calculation time is reduced in both cases and the improvement is close to 10 %. For all those calculations, the BLT resolution times were respectively 30 s. and 8 s. for the compiler-non optimized and optimized calculations. In [9], it had been pointed out that, with the improvements of the BLT resolution, the calculation time on the junctions became the new limitation of the total calculation time. In the case of the optimized option of the compiler, this time becomes now comparable to the one required to calculate the scattering parameters of junctions.

Calculation Time	"Old_non_opt"	"New_non_opt"	"Old_opt"	"New_ opt"
Junctions	26 min. 06 s.	2 min. 11 s.	7 min. 10 s.	0 min. 40 s.
BLT resolution	0 min. 30 s.	0 min. 30 s.	0 min. 8 s.	0 min. 8 s.

Table 4-1 : Calculation time of the scattering parameters of all the junctions on figure 4-1 network

5. Conclusions

The calculation time of the scattering parameters of ideal junctions can be widely improved applying an efficient matrix technique described in [10]. In addition to [10], we have demonstrated that this technique could be applied without any modifications to the case of ideal junctions involving short-circuits to the ground and open circuits.

Compared to the technique described in [8], the new technique requires less memory and leads to a significant reduction of the computation time. This way, the calculation on junctions is not anymore the new limitation of the total calculation time that had been emphasized when taking into account the sparse structure of the BLT matrix [9]. With the same idea, in the next future, a similar effort should deal with the optimization of the calculation time of tube characteristics. Some recent papers describing analytical formula for the calculation of modes and propagation speeds on cable bundles already provide a pertinent answer to this problem [13].

Furthermore, especially in high frequency, it is now quite established that accounting for the non uniformity of cable bundles becomes a requirement [14]. The more straightforward technique deals with splitting the cable in different pieces, each having a particular impedance matrix and velocity matrix. The connection of the different pieces is performed thanks to ideal junctions. Depending on the non uniformity and the length of the cable, the number of ideal junctions may widely increase. The requirement of such junctions seems unavoidable. Even if special techniques like the interpolation of the velocities between two short sections of cables seem to be effective [15], the application of such techniques to cable bundles stillimposes non realistic hypothesis (uniform medium, structure of the characteristic matrices).

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