New Reciprocal Circuit Model for Lossy Waveguide Structures Based on the Orthogonality of the Eigenmodes

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***Abstract*— In this contribution, we present a new consistent equivalent transmission line model to describe the propagation along lossy hybrid waveguide structures. All existing consistent transmission line models are based on the assumption that the power propagated by the modes considered in the waveguide is the same as the power propagated in the model. In a lossy reciprocal waveguide, this leads to a nonreciprocal transmission line model because the modes are not power orthogonal. We start from the Lorentz orthogonality condition to construct a reciprocal transmission line model, even for lossy waveguides. For multiconductor waveguides, we will discuss what we call RI- and RV-models, in analogy with the existing PI- and PV-models. We will also present a generalisation of these RI- and RV-models to general waveguide structures. The theory is illustrated with a comparison of an RI- and PI-model for a lossy thick microstrip structure.**

I. Introduction

The representation of the fundamental modes of a nonhybrid multiconductor waveguide, such as the coaxial line, by a set of transmission lines is a direct result of Maxwell's equations because these modes are purely TEM. This representation is also natural and unambiguous for the fundamental modes of lossless hybrid, i.e., inhomogeneous, multiconductor waveguides in the quasi-TEM limit [1], [2]. At higher frequencies, or for higher order modes, the transmission line models are no longer unambiguous because currents and voltages can no longer be defined in a unique way due to the presence of nonnegligible longitudinal field components. In the past this problem has been tackled by different approaches.

In [3], a consistent transmission line model for a single mode in a lossy hybrid waveguide interconnection was constructed. This model was based on the power equivalence between the model and the waveguide. A model for coupled multiconductor waveguides was introduced in [4]. However, in [5], [6] it was shown that this model is correct only in the quasi-TEM limit because of the special choice of the partial powers (i.e., power per conductor and per mode) propagated in the structure. In [7], it was shown that the model of [4] does not yield a reciprocal model, even for lossless structures. In [5], [6], a consistent definition of the partial powers resulted in a new circuit model that is now reciprocal in the lossless case but not in the lossy one. This model was further generalised to anisotropic waveguides in [8]. The above is only a brief review of the extensive literature on the subject, for more references see [6].

When constructing a circuit model for lossless multiconductor waveguide structures, one generally takes only the fundamental modes into account. It is assumed that higher order modes are either below cut-off, or that they are not excited at the generator and load of the waveguide. Consequently, only the fundamental modes will propagate power and it is justified to demand that the power propagated by the fundamental modes in the waveguide is the same as the one propagated in the circuit model. For lossy structures, the higher order modes, even if they do not propagate, will dissipate power due to the losses. This means that the power demand looses its justification in the lossy case. As already mentioned above, it was shown in [6] that, in the lossy case, the power demand results in a nonreciprocal transmission line model. This results from the fact that the modes are not power orthogonal in lossy waveguides. This absence of reciprocity and the Lorentz orthogonality of the modes have led us to the construction of a new circuit model based on a reciprocity demand. This model is reciprocal for lossy structures and reduces to the classical power model for lossless structures.

Neither the power, nor the reciprocity demand, fully determines all the parameters in the transmission line model. One has to include extra information on how the hybrid waveguide is connected to and excited by the generator and the load. This has led to power-current (PI) models and power-voltage (PV) models [3], [5], [6]. When the reciprocity demand is used, we call the corresponding models the RI- and RV-models. In the RI-model, the currents on the lines are taken to be equal to the currents on the conductors of the waveguide. In the RV-model, the voltages on the lines are taken to be equal to line integrals of the electric field along well-chosen paths in the cross-section of the waveguide. These RI- and RV-models are only suitable for multiconductor waveguides and they are only approximations of reality. We will present a more general model that takes the three-dimensional electromagnetic interactions at the generator and at the load into account. The RI- and RV-models are special cases of this more general model. The drawback of this general model is that it is not independent of the load and generator and, as such, is more difficult to implement in a general purpose circuit simulator.

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We will construct the transmission line model for lossy waveguide structures consisting of reciprocal material that is anisotropic in the cross-section. These waveguides are of course reciprocal, but also bi-directional [9]. This means that corresponding modes propagating in opposite directions have the same complex wave number. Reciprocal waveguides that are not bi-directional cannot be represented by a classical transmission line model because such a model is inherently bi-directional. In [8], a more general transmission line model for nonbi-directional chiral waveguides was discussed based on the power demand.

II. Modes in a Waveguide Structure

Consider a lossy anisotropic waveguide structure that is invariant in the z-direction. An electromagnetic field that propagates along this structure can be decomposed in contributions from the different eigenmodes of the structure. For a general open waveguide, the set of eigenmodes consists of a discrete and a continuous part. In the representation of the final circuit model, we will take only C discrete modes into account. We will come back to the choice of C later on. Hence, we write an arbitrary field $\mathbf{E}(x,y,z), \mathbf{H}(x,y,z)$ which propagates in the positive z-direction as

$$\begin{aligned} \mathbf{E}(x,y,z) &= \sum_{f=1}^{C} \mathbf{e}_f(x,y,z) K_{w,f} \\ \mathbf{H}(x,y,z) &= \sum_{f=1}^{C} \mathbf{h}_f(x,y,z) K_{w,f} \end{aligned} \tag{1}$$

$\mathbf{e}_f(x,y,z)$ and $\mathbf{h}_f(x,y,z)$ $(f = 1,2,\ldots,C)$ are the properly normalised field patterns of the eigenmodes. $K_{w,f}$ is an excitation coefficient that determines the contribution of mode f to the total field $\mathbf{E}(x,y,z)$, $\mathbf{H}(x,y,z)$.

The z-dependence of the fields $\mathbf{e}_f(x,y,z)$, $\mathbf{h}_f(x,y,z)$ of mode f is characterised by the complex wave number $\beta_{w,f}$, such that

$$\begin{aligned} \mathbf{e}_f(x,y,z) &= \mathbf{E}_f(x,y)\exp(-j\beta_{w,f}z) \\ \mathbf{h}_f(x,y,z) &= \mathbf{H}_f(x,y)\exp(-j\beta_{w,f}z). \end{aligned} \tag{2}$$

Substitution of (2) in (1) yields

$$\begin{aligned} \mathbf{E}(x,y,z) &= \mathbf{E}^T(x,y)\exp(-j\mathbf{beta}_w z)\mathbf{K}_w \\ \mathbf{H}(x,y,z) &= \mathbf{H}^T(x,y)\exp(-j\mathbf{beta}_w z)\mathbf{K}_w \end{aligned} \tag{3}$$

where $\mathbf{beta}_w$ is a diagonal matrix with the C complex wave numbers on its diagonal. In (3), $K_{w,f}$, $\mathbf{E}_f(x,y)$ and $\mathbf{H}_f(x,y)$ are collected in the column vectors $\mathbf{K}_w$, $\mathbf{E}(x,y)$ and $\mathbf{H}(x,y)$ respectively. T is the transposition operator.

It is assumed that the material in the waveguide consists of lossy reciprocal anisotropic material characterised by **eps** and **mu** tensors of the following form:

$$\begin{aligned} \mathbf{eps}(x,y) &= \begin{pmatrix} \mathbf{eps}_{\rm tr}(x,y) & 0 \\ 0 & \varepsilon_z(x,y) \end{pmatrix} \\ \mathbf{mu}(x,y) &= \begin{pmatrix} \mathbf{mu}_{\rm tr}(x,y) & 0 \\ 0 & \mu_z(x,y) \end{pmatrix} \end{aligned} \tag{4}$$

with $\mathbf{eps}_{\rm tr} = (\mathbf{eps}_{\rm tr})^T$ and $\mathbf{mu}_{\rm tr} = (\mathbf{mu}_{\rm tr})^T$. $\mathbf{eps}_{\rm tr}$, $\mathbf{mu}_{\rm tr}$, ε_z and μ_z can take arbitrary complex values. With these assumptions, the waveguide is bi-directional [9]. In Appendix A it is shown that the following normalisation relation holds when we have materials of type (4):

$$\frac{1}{2}\int\int_S \{\mathbf{E}_f(x,y)\times\mathbf{H}_g(x,y)\}.\mathbf{u}_z dS = \delta_{fg} \qquad f,g = 1,2,\ldots,C \tag{5}$$

δ_{fg} is the Kronecker symbol and S is the cross-section of the waveguide. Equation (5) normalises the modal field patterns.

III. Modes in a Set of Coupled Transmission Lines

Suppose we have a set of C coupled transmission lines, with propagation along the z-direction. In the frequency domain, this set of transmission lines is described by the following system of equations:

$$\begin{aligned} \frac{d\,\mathbf{V}(z)}{dz} + \mathbf{Z}\,\mathbf{I}(z) &= 0 \\ \frac{d\,\mathbf{I}(z)}{dz} + \mathbf{Y}\,\mathbf{V}(z) &= 0. \end{aligned} \tag{6}$$

$\mathbf{I}(z)$ and $\mathbf{V}(z)$ are column matrices with elements $I_j(z)$ and $V_j(z)$ $(j = 1,2,\ldots,C)$, representing respectively the current and voltage on transmission line j. $\mathbf{Z}$, and $\mathbf{Y}$ are the circuit impedance and circuit admittance matrices respectively. These matrices are usually rewritten in terms of their real and imaginary parts as follows:

$$\begin{aligned} \mathbf{Z} &= j\,\omega\,\mathbf{L} + \mathbf{R} \\ \mathbf{Y} &= j\,\omega\,\mathbf{C} + \mathbf{G}. \end{aligned} \tag{7}$$

The circuit quantities $\mathbf{Z}$, $\mathbf{Y}$, $\mathbf{C}$, $\mathbf{G}$, $\mathbf{L}$ and $\mathbf{R}$ are frequency dependent. In a lossless situation for low frequencies in the quasi-TEM limit $\mathbf{C}$ and $\mathbf{L}$ take their classical static frequency independend values as capacitance and inductance matrix ([1] and [2]). We assume that $\mathbf{Z}$ and $\mathbf{Y}$ are nonsingular matrices.

The general solution of the differential equations (6) is given by

$$\begin{aligned} \mathbf{I}(z) &= \mathbf{I}_c\exp(-j\mathbf{beta}_c z)\mathbf{K}_c \\ \mathbf{V}(z) &= \mathbf{V}_c\exp(-j\mathbf{beta}_c z)\mathbf{K}_c \end{aligned} \tag{8}$$

where $\mathbf{beta}_c$ is a diagonal matrix with diagonal elements given by the square root of minus the eigenvalues of the matrix $\mathbf{YZ}$. The columns of the matrix $\mathbf{I}_c(\mathbf{V}_c)$ are the eigenvectors of $\mathbf{YZ}(\mathbf{ZY})$. In (8), the current $\mathbf{I}(z)$ and the voltage $\mathbf{V}(z)$ are expanded in the contributions of the C different eigenmodes of the set of transmission lines. For this reason we call $\mathbf{I}_c(\mathbf{V}_c)$ the circuit-mode current (voltage) matrix and $\mathbf{beta}_c$ the complex wave number matrix. $K_{c,f}(f = 1,2,\ldots,C)$ is an excitation coefficient that gives the contribution of mode f to the current $\mathbf{I}(z)$ and to the voltage $\mathbf{V}(z)$. In (8), we only took modes propagating in the positive z-direction into account. The complex wave number matrix for modes propagating in the negative z-direction is $-\mathbf{beta}_c$ because a set of transmission lines described by (6) is always bi-directional. From (6) and (8), the relation between $\mathbf{V}_c$ and $\mathbf{I}_c$ is the following:

$$\begin{aligned}\mathbf{V}_c &= j(\mathbf{Y})^{-1}\mathbf{I}_c\ \mathbf{beta}_c\\ \mathbf{I}_c &= j(\mathbf{Z})^{-1}\mathbf{V}_c\ \mathbf{beta}_c.\end{aligned} \tag{9}$$

Now we assume that the set of transmission lines is reciprocal, in other words we assume that $\mathbf{Y}$ and $\mathbf{Z}$ are symmetric. Reciprocity states that the transmission lines satisfy the Tellegen circuit law or the Kirchoff laws. This means that the following normality relation holds between the eigenmodes:

$$\frac{1}{2}\mathbf{V}_c^T\mathbf{I}_c = \mathbf{U}. \tag{10}$$

We have used this orthogonality property to normalise the eigenmodes such that $\mathbf{U}$ is the C by C unit matrix. (10) can be found by applying the law of Tellegen or, directly from (9) by demanding $\mathbf{Y}$ and $\mathbf{Z}$ to be symmetric.

Using (9) and (10), one can construct the following important relations:

$$\begin{aligned}\mathbf{Y} &= \frac{j}{2}\mathbf{I}_c\mathbf{beta}_c\mathbf{I}_c^T\\ \mathbf{Z} &= 2j(\mathbf{I}_c^T)^{-1}\mathbf{beta}_c(\mathbf{I}_c)^{-1}.\end{aligned} \tag{11}$$

These relations immediately show that $\mathbf{Y}$ and $\mathbf{Z}$ are symmetric.

Another important quantity describing a set of coupled transmission lines is the characteristic impedance matrix $\mathbf{Z}_{\text{char}}$. $\mathbf{Z}_{\text{char}}$ gives the relation between a voltage and a current wave $\mathbf{V}(z)$ and $\mathbf{I}(z)$ which propagate in the positive direction. $\mathbf{Z}_{\text{char}}$ is the input impedance matrix of the set of transmission lines. It follows that $\mathbf{Z}_{\text{char}}$ is given by:

$$\mathbf{Z}_{\text{char}} = 2(\mathbf{I}_c^T)^{-1}(\mathbf{I}_c)^{-1}. \tag{12}$$

Equation (12) shows that $\mathbf{Z}_{\text{char}}$ is also symmetrical.

IV. Identification of the Waveguide and the Set of Transmission Lines

In this section, we construct a circuit model for the C modes propagating in the waveguide of Section II. In other words, we want to determine the circuit parameters $\mathbf{Y}$ and $\mathbf{Z}$ in (6) in such a way that the propagation of the C modes in the waveguide is an "as good as possible" representation of the wave propagation in the set of transmission lines. The meaning of "as good as possible" will become clear from the sequel.

First, we demand the complex wave numbers of the modes in the waveguide to be identical to the complex wave numbers of the modes in the set of transmission lines

$$\mathbf{beta}_c = \mathbf{beta}_w. \tag{13}$$

Secondly, we demand the normalised modes in the set of transmission lines to be excited by the same amount as the normalised modes in the waveguide structure. In other words, we demand that:

$$\mathbf{K}_c = \mathbf{K}_w. \tag{14}$$

In a lossless waveguide, (14) combined with the normalisations (5) and (10), imply that the power propagated by the C modes in the waveguide is the same as the power propagated in the set of transmission lines.

Equations (13) and (14) allow us to omit the subscripts "c" and "w" in $\mathbf{beta}_c$, $\mathbf{beta}_w$, $\mathbf{K}_c$ and $\mathbf{K}_w$. The conditions (13) and (14) do not fully specify $\mathbf{Y}$ and $\mathbf{Z}$, since there are still C^2 degrees of freedom left. In (11) we still have to determine the circuit-mode current matrix $\mathbf{I}_c$ (or equivalently $\mathbf{V}_c$). The determination of $\mathbf{I}_c$ depends on how the waveguide is excited or, more precisely, upon the interconnection of the waveguide with external circuits or other waveguides. We will discuss various cases in the next section. Equation (12) shows that determining $\mathbf{I}_c$ is equivalent to determining the "impedance level" of the set of transmission lines.

The circuit model developed in [5], [6], [8] demands the power propagated by C modes in the lossy waveguide and the set of coupled transmission lines to be the same. In that case, and when we take $\mathbf{K}_c = \mathbf{K}_w$, it is easy to show that the reciprocity relation (10) should be replaced by

$$\frac{1}{2}\mathbf{V}_c^T\mathbf{I}_c^* = \mathbf{P} \tag{15}$$

where $\mathbf{P}$ is the modal cross-power matrix defined by

$$P_{fg} = \frac{1}{2}\int\int_S \{\mathbf{E}_f(x,y)\times\mathbf{H}_g^*(x,y)\}.\mathbf{u}_z dS \qquad f,g = 1,2,\ldots,C. \tag{16}$$

The matrix $\mathbf{P}$ is *not* a diagonal matrix because, in a lossy waveguide, the eigenmodes are *not* power orthogonal. If (15) replaces (10) then (11) should be replaced by

$$\begin{aligned}\mathbf{Y} &= \frac{j}{2}\mathbf{I}_c\mathbf{beta}(\mathbf{P}^T)^{-1}\mathbf{I}_c^{T*}\\ \mathbf{Z} &= 2j(\mathbf{I}_c^{T*})^{-1}\mathbf{P}^T\mathbf{beta}(\mathbf{I}_c)^{-1}.\end{aligned} \tag{17}$$

In [6] it is shown that these relations imply that $\mathbf{Y}$ and $\mathbf{Z}$ are *not* symmetric or, that the set of transmission lines is *not* reciprocal. We conclude that, if one uses a circuit model based on the power demand and not on the reciprocity demand, a reciprocal lossy waveguide is represented by a nonreciprocal set of transmission lines. When the structure is lossless, there is no difference between a model based on the reciprocity demand and a model based on the power demand. We remark that, in (17), there are again C^2 degrees of freedom left because $\mathbf{I}_c$ is still unknown.

V. Determination of the Circuit-Mode Current Matrix

In this section we will determine $\mathbf{I}_c$ in various cases. We will start with the junction of two general waveguides. Then, we will look at the excitation of a waveguide by a lumped element generator. Finally, we will construct the RI- and RV-models for multiconductor waveguides and we will show that they are special cases of the excitation by a lumped element generator.

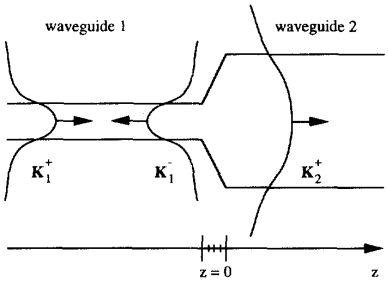

Fig. 1. Geometry of the junction of two waveguides.

A. Interconnection of Two Waveguides

Consider the situation of Fig. 1 where two general waveguide structures are interconnected. The junction is not necessary abrupt but can extend over a finite length in the z-direction. We take C modes in both waveguides into account. We want to represent the structure of Fig. 1 by the junction of two sets of C coupled transmission lines. In the model, the junction is represented by an abrupt junction at $z = 0$. The finite length junction between the real waveguides will be replaced by a lumped element circuit as discussed below. We assume that we know the impedance level $\mathbf{Z}_{\mathrm{char},1}$ or the circuit-mode current matrix $\mathbf{I}_{c,1}$ of the transmission line model of the left waveguide. Now we will determine $\mathbf{Z}_{\mathrm{char},2}$ or $\mathbf{I}_{c,2}$ of the transmission line model of the right waveguide.

Suppose C incoming modes in the left waveguide, with excitation coefficients $\mathbf{K}_1^+$, incident on the junction. They will excite transmitted modes, with excitation coefficients $\mathbf{K}_2^+$, propagating in the positive z-direction in the right waveguide and reflected modes, with excitation coefficients $\mathbf{K}_1^-$, propagating in the negative z-direction in the left waveguide. From a three dimensional electromagnetic full-wave analysis of the junction, we know by what amount incoming modes in the left waveguide excite transmitted modes in the right waveguide and reflected modes in the left waveguide. In other words, we know the reflection and transmission matrices $\mathbf{S}_{11}$ and $\mathbf{S}_{21}$ which describe the relation between $\mathbf{K}_1^-, \mathbf{K}_2^+$, and $\mathbf{K}_1^+$

$$\mathbf{K}_2^+ = \mathbf{S}_{21}\mathbf{K}_1^+ \qquad \mathbf{K}_1^- = \mathbf{S}_{11}\mathbf{K}_1^+. \tag{18}$$

In an analogous way, we write the relation between $\mathbf{K}_1^-, \mathbf{K}_2^+$ and $\mathbf{K}_2^-$ when there are incoming waves in the right waveguide

$$\mathbf{K}_1^- = \mathbf{S}_{12}\mathbf{K}_2^- \qquad \mathbf{K}_2^+ = \mathbf{S}_{22}\mathbf{K}_2^-. \tag{19}$$

We assume that the junction is reciprocal, i.e., that $\mathbf{S}_{12} = \mathbf{S}_{21}$. This reciprocity demand places some restrictions on the type of materials used in the junction. The matrices $\mathbf{S}_{ij}(i,j = 1,2)$ take into account all the effects inside the junction, including wave propagating effects, when the junction is not abrupt. The matrices $\mathbf{S}_{ij}(i,j = 1,2)$ build up the scattering matrix of the junction.

Fig. 2 shows the equivalent circuit of the structure of Fig. 1. Between the two sets of transmission lines, we introduce a series impedance matrix $\mathbf{Z}_0$ and a parallel admittance matrix $\mathbf{Y}_0$. These matrices take into account the excitation of higher order modes, losses or propagation effects inside the junction.

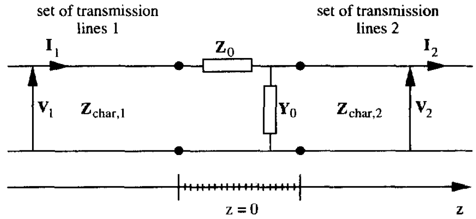

Fig. 2. Circuit model for the structure of Fig. 1 consisting of two coupled sets of transmission lines connected by a $(\mathbf{Z}_0, \mathbf{Y}_0)$-circuit.

The current and voltage waves in both transmission lines due to an excitation $\mathbf{K}_1^+$ can be written as

$$\begin{aligned}
\mathbf{I}_1(z) &= \mathbf{I}_{c,1}\exp(-j\mathbf{beta}_1 z)\mathbf{K}_1^+ - \mathbf{I}_{c,1}\exp(j\mathbf{beta}_1 z)\mathbf{K}_1^- \\
\mathbf{V}_1(z) &= 2(\mathbf{I}_{c,1}^T)^{-1}\exp(-j\mathbf{beta}_1 z)\mathbf{K}_1^+ \\
&\quad + 2(\mathbf{I}_{c,1}^T)^{-1}\exp(j\mathbf{beta}_1 z)\mathbf{K}_1^- \\
\mathbf{I}_2(z) &= \mathbf{I}_{c,2}\exp(-j\mathbf{beta}_2 z)\mathbf{K}_2^+ \\
\mathbf{V}_2(z) &= 2(\mathbf{I}_{c,2}^T)^{-1}\exp(-j\mathbf{beta}_2 z)\mathbf{K}_2^+
\end{aligned} \tag{20}$$

and those due to an excitation $\mathbf{K}_2^-$ as

$$\begin{aligned}
\mathbf{I}_1(z) &= -\mathbf{I}_{c,1}\exp(j\mathbf{beta}_1 z)\mathbf{K}_1^- \\
\mathbf{V}_1(z) &= 2(\mathbf{I}_{c,1}^T)^{-1}\exp(j\mathbf{beta}_1 z)\mathbf{K}_1^- \\
\mathbf{I}_2(z) &= \mathbf{I}_{c,2}\exp(-j\mathbf{beta}_2 z)\mathbf{K}_2^+ - \mathbf{I}_{c,2}\exp(j\mathbf{beta}_2 z)\mathbf{K}_2^- \\
\mathbf{V}_2(z) &= 2(\mathbf{I}_{c,2}^T)^{-1}\exp(-j\mathbf{beta}_2 z)\mathbf{K}_2^+ \\
&\quad + 2(\mathbf{I}_{c,2}^T)^{-1}\exp(j\mathbf{beta}_2 z)\mathbf{K}_2^-.
\end{aligned} \tag{21}$$

In (20) and (21) we used the relation (10). The circuit formed by $\mathbf{Z}_0$ and $\mathbf{Y}_0$ imposes the following relations at $z = 0$

$$\begin{aligned}
\mathbf{I}_1(0) &= \mathbf{I}_2(0) + \mathbf{Y}_0\mathbf{V}_2(0) \\
\mathbf{V}_1(0) - \mathbf{Z}_0\mathbf{I}_1(0) &= \mathbf{V}_2(0).
\end{aligned} \tag{22}$$

If (20)-(22) hold for arbitrary excitations $\mathbf{K}_1^+$ and $\mathbf{K}_2^-$, then one obtains the following expressions for $\mathbf{I}_{c,2}, \mathbf{Z}_0$ and $\mathbf{Y}_0$

$$\begin{aligned}
\mathbf{I}_{c,2} &= \frac{1}{2}\mathbf{I}_{c,1}[(\mathbf{U} - \mathbf{S}_{11})(\mathbf{S}_{12})^{-1}(\mathbf{U} + \mathbf{S}_{22}) + \mathbf{S}_{12}] \\
\mathbf{Y}_0 &= \frac{1}{2}[\mathbf{I}_{c,1}(\mathbf{U} - \mathbf{S}_{11})(\mathbf{S}_{12})^{-1} - \mathbf{I}_{c,2}]\mathbf{I}_{c,2}^T \\
\mathbf{Z}_0 &= 2[(\mathbf{I}_{c,1}^T)^{-1}(\mathbf{U} + \mathbf{S}_{11}) - (\mathbf{I}_{c,2}^T)^{-1}\mathbf{S}_{12}] \\
&\quad \times (\mathbf{U} - \mathbf{S}_{11})^{-1}(\mathbf{I}_{c,1})^{-1}.
\end{aligned} \tag{23}$$

Once $\mathbf{I}_{c,2}$ is known, one can determine $\mathbf{Z}_{\mathrm{char},2}$ using (12). Equation (23) indicates that one can only omit $\mathbf{Z}_0$ and $\mathbf{Y}_0$ when there are special relations between $\mathbf{S}_{11}, \mathbf{S}_{22}$ and $\mathbf{S}_{12}$. For $C = 1$, these relations are $\mathbf{S}_{11} = -\mathbf{S}_{22}$ and $(S_{11})^2 + (S_{12})^2 = 1$. In the lossless case this relation means that the sum of the power in the transmitted and reflected mode is equal to the power in the incoming mode.

The via-hole is an important special junction between two single line multiconductor waveguides. In practice, such a via-hole is characterised by its capacitance C_0 [10] and inductance L_0 [11]. In a circuit model, these are represented by a series impedance $Z_0 = j\omega L_0$ and a parallel admittance $Y_0 = j\omega C_0$, in accordance with the model of Fig. 2.

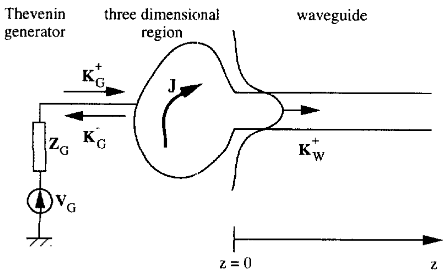

Fig. 3. Geometry of the interconnection of a waveguide with a lumped element generator.

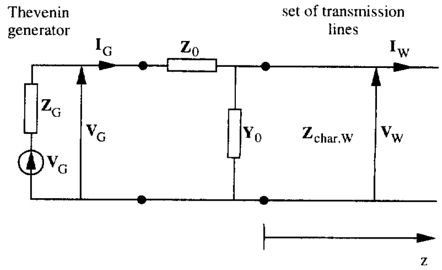

Fig. 4. Circuit model for the structure of Fig. 3 consisting of a coupled Thevenin generator and a set of transmission lines connected by a $(\mathbf{Z}_0, \mathbf{Y}_0)$-circuit.

When the junction is abrupt, it is possible to express the matrices $\mathbf{S}_{ij}(i,j = 1,2)$ as a function of the projection of the modes of one waveguide onto the modes of the other waveguide. This is done by imposing the continuity of the transverse field components at the junction as shown in Appendix B.

Instead of inserting a two-element circuit, consisting of $\mathbf{Z}_0$ and $\mathbf{Y}_0$, between the two waveguides, one can insert a π or T-circuit with three unknown elements. In this case, one can determine these three unknown elements and the characteristic impedance of the left waveguide can be chosen arbitrarily.

B. Excitation by a Lumped Element Generator

Fig. 3 shows a general configuration of a waveguide excited by a coupled Thevenin generator. This Thevenin generator can be a circuit model for the three-dimensional electromagnetic excitation of the waveguide or a real lumped element source as is often the case for a multiconductor waveguide. Again, we will assume C modes in the waveguide and hence, we will assume a C-dimensional coupled Thevenin generator, characterised by an impedance matrix $\mathbf{Z}_G$. For a multiconductor waveguide, C is generally taken equal to the number of conductors (exclusive the ground conductor). In this case the C modes correspond to the dominant modes in the waveguide. In the lossless situation, these C modes are the fundamental modes and the other modes are assumed to be below cut-off.

From elementary circuit theory, it is known that the voltage vector $\mathbf{V}_G$ across the coupled Thevenin generator and the current vector $\mathbf{I}_G$ through the coupled Thevenin generator can be decomposed into an incoming and a reflected part

$$\begin{aligned}\mathbf{I}_G &= \mathbf{I}_{c,G}\mathbf{K}_G^+ - \mathbf{I}_{c,G}\mathbf{K}_G^- \\ \mathbf{V}_G &= \mathbf{V}_{c,G}\mathbf{K}_G^+ + \mathbf{V}_{c,G}\mathbf{K}_G^-\end{aligned} \tag{24}$$

with

$$\frac{1}{2}\mathbf{V}_{c,G}^T\mathbf{I}_{c,G} = \mathbf{U} \qquad \mathbf{V}_{c,G} = \mathbf{Z}_G\mathbf{I}_{c,G}. \tag{25}$$

$\mathbf{K}_G^+$ and $\mathbf{K}_G^-$ are the excitation coefficients of the incoming and reflected modes of the generator respectively. Remark the resemblance between (24) and the first two equations of (20).

We can now proceed as in the previous case. Suppose that we know the relation between the amplitudes $\mathbf{K}_G^-, \mathbf{K}_W^+$ of the modes in the waveguide and an excitation $\mathbf{K}_G^+$. These relations follow from a three dimensional electromagnetic analysis of the generator and are represented by

$$\mathbf{K}_W^+ = \mathbf{S}_{21}\mathbf{K}_G^+ \qquad \mathbf{K}_G^- = \mathbf{S}_{11}\mathbf{K}_G^+. \tag{26}$$

The relation between $\mathbf{K}_G^-, \mathbf{K}_W^+$ and an excitation $\mathbf{K}_W^-$ is given by

$$\mathbf{K}_G^- = \mathbf{S}_{12}\mathbf{K}_W^- \qquad \mathbf{K}_W^+ = \mathbf{S}_{22}\mathbf{K}_W^-. \tag{27}$$

We assume a reciprocal interconnection with $\mathbf{S}_{21} = \mathbf{S}_{12}$. We represent the structure of Fig. 3 by the circuit model of Fig. 4 where again a coupled series impedance $\mathbf{Z}_0$ and parallel admittance $\mathbf{Y}_0$ is introduced. The circuit-mode current matrix $\mathbf{I}_{c,W}$, $\mathbf{Z}_0$ and $\mathbf{Y}_0$ are determined in the same way as in (23) with the subscripts "1" and "2" replaced respectively by "G" and "W"

$$\begin{aligned}\mathbf{I}_{c,W} &= \frac{1}{2}\mathbf{I}_{c,G}[(\mathbf{U} - \mathbf{S}_{11})(\mathbf{S}_{12})^{-1}(\mathbf{U} + \mathbf{S}_{22}) + \mathbf{S}_{12}] \\ \mathbf{Y}_0 &= \frac{1}{2}[\mathbf{I}_{c,G}(\mathbf{U} - \mathbf{S}_{11})(\mathbf{S}_{12})^{-1} - \mathbf{I}_{c,W}]\mathbf{I}_{c,W}^T \\ \mathbf{Z}_0 &= 2[(\mathbf{I}_{c,G}^T)^{-1}(\mathbf{U} + \mathbf{S}_{11}) - (\mathbf{I}_{c,W}^T)^{-1}\mathbf{S}_{12}] \\ &\quad \times (\mathbf{U} - \mathbf{S}_{11})^{-1}(\mathbf{I}_{c,G})^{-1}.\end{aligned} \tag{28}$$

C. RI- and RV-Models for Multiconductor Waveguides

In the models based on the power demand [3], [5], [6], [8], the matrix $\mathbf{I}_c$ for multiconductor waveguides was determined using the so-called PI or PV formulation. These models can also be introduced in combination with the reciprocity demand. In such a case, we suppose that we have a multiconductor waveguide with C conductors and that we take the C dominant modes into account.

In the RI-model, the currents on the conductors in the waveguide should be the same as the currents on the transmission lines in the equivalent transmission line model. We define a waveguide-mode current matrix $\mathbf{I}_w$ as

$$I_{w,jf} = \oint_{c_j} \mathbf{H}_f(x,y).dl \qquad j,f = 1,2,\ldots,C. \tag{29}$$

c_j is the boundary curve around conductor j. Element $I_{w,jf}$ is the current on conductor j for mode f. In the RI-model the $\mathbf{I}_c$ matrix of the model is defined by

$$\mathbf{I}_c = \mathbf{I}_w. \tag{30}$$

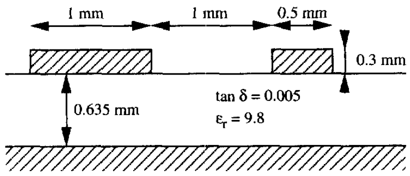

Fig. 5. Geometry of an asymmetrical coupled microstrip line with a lossy substrate.

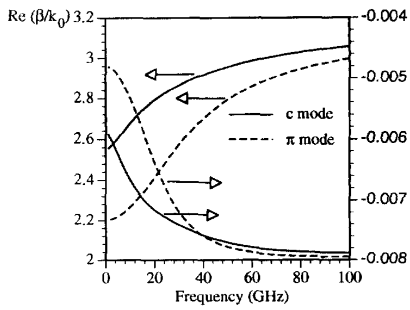

Fig. 6. Complex normalised wave numbers of the c- and p-mode for the structure of Fig. 5.

Contrary to (28), this definition has the advantage to be independent from the generator. It does not, however, take the detailed three-dimensional electromagnetic interaction with a generator into account. It is easy to see that (30) is a special case of the previous, more general model (28). Inverting the relations (28), using (30) and $\mathbf{Z}_0 = \mathbf{Y}_0 = 0$, yields the $\mathbf{S}_{11}, \mathbf{S}_{12}$, and $\mathbf{S}_{22}$ matrices associated with the RI-model.

The RV-model associates the voltages on the multiconductor lines with line integrals of the electric field along suitable paths $l_j(j = 1, 2, \ldots, C)$ between the ground plane and conductors in the waveguide. In analogy with (29), a waveguide-mode voltage matrix V_w is defined

$$V_{w,jf} = \int_{l_j} \mathbf{E}_f(x, y).d\mathbf{l} \qquad j, f = 1, 2, \ldots, C. \tag{31}$$

If we demand that $\mathbf{V}_c = \mathbf{V}_w$ and use (10), then $\mathbf{I}_c$ is given by

$$\mathbf{I}_c = 2(\mathbf{V}_w^T)^{-1}. \tag{32}$$

The difference between the RI- and RV-model (or equivalently PI- and PV-model) dissapears when $\omega \to 0$ because, in both models, the equivalent currents and voltages become the true currents and voltages [6].

For a discussion about when to use models based on current, i.e., RI- or PI-models, and when to use models based on voltage, i.e., RV- or PV-models, refer to [6].

VI. Integral Representations for C, L, R, and G Matrices

It is possible to express $\mathbf{Y}$ and $\mathbf{Z}$, and consequently, $\mathbf{C}, \mathbf{L}, \mathbf{R}$, and $\mathbf{G}$, as integrals of the modal field distributions over the cross-section of the structure. We start from the Maxwell curl equations

$$\begin{aligned}
&\nabla_{\text{tr}} \times \mathbf{E}_{\text{tr}} = -j\omega\mu_z\mathbf{H}_z\mathbf{u}_z\\
&\nabla_{\text{tr}} \times E_z\mathbf{u}_z - j\beta\mathbf{u}_z \times \mathbf{E}_{\text{tr}} = -j\omega\mathbf{mu}_{\text{tr}}\mathbf{H}_{\text{tr}}\\
&\nabla_{\text{tr}} \times \mathbf{H}_{\text{tr}} = j\omega\varepsilon_z E_z\mathbf{u}_z\\
&\nabla_{\text{tr}} \times H_z\mathbf{u}_z - j\beta\mathbf{u}_z \times \mathbf{H}_{\text{tr}} = j\omega\mathbf{eps}_{\text{tr}}\mathbf{E}_{\text{tr}}
\end{aligned} \tag{33}$$

where we have split the equations into transverse and longitudinal parts. Upon eliminating $\mathbf{E}_{\text{tr},f}$ in (5), with the aid of the second equation of (33) one obtains

$$\frac{1}{2}\iint_S \{j\omega(\mathbf{mu}_{\text{tr}}\mathbf{H}_{\text{tr},f}).\mathbf{H}_{\text{tr},g} + (\nabla_{\text{tr}} \times E_{z,f}\mathbf{u}_z).\mathbf{H}_{\text{tr},g}\}dS = j\beta_f\delta_{fg}. \tag{34}$$

Using Gauss' theorem, Sommerfeld's radiation condition and the third equation of (33), leads to the following expression:

$$\frac{\omega}{2}\iint_S \{(\mathbf{mu}_{\text{tr}}\mathbf{H}_{tr,f}).\mathbf{H}_{\text{tr},g} + \varepsilon_z E_{z,f}E_{z,g}\}dS = \beta_f\delta_{fg}. \tag{35}$$

Using the first and last equation of (33) a dual expression can be found

$$\frac{\omega}{2}\iint_S \{(\mathbf{eps}_{\text{tr}}\mathbf{E}_{\text{tr},f}).\mathbf{E}_{\text{tr},g} + \mu_z H_{z,f}H_{z,g}\}dS = \beta_f\delta_{fg}. \tag{36}$$

Upon substituting the integral expressions (35) or (36) for β_f in (11), one obtains the integral expressions for $\mathbf{Y}$ and $\mathbf{Z}$. Remark that the matrix $\mathbf{I}_c$, appearing in (11), is defined in function of the fields for the RI- and RV-models, with (29)–(32).

VII. Example

In this example, we will examine the difference between the PI- and RI-model for the two dominant modes in an asymmetric coupled lossy microstrip line shown in Fig. 5. The substrate of the line has a thickness of 0.635 mm and consists of lossy material with relative dielectric constant $\varepsilon_r = 9.8$ and loss tangent $\tan\delta = 0.005$. The strips have a thickness of 0.3 mm and respective widths of 1 mm and 0.5 mm. The strips are perfectly conducting and the distance between the strips is 1.0 mm.

This structure was analysed by a rigorous full-wave eigenmode analysis that takes both the losses and the thickness of the strips into account without approximations [12], [13]. With this analysis, we determine the modal complex wave numbers and the integrals of modal field components over the cross-section of the waveguide.

Fig. 6 shows the normalised complex wave numbers of the two dominant modes in the structure as a function of frequency. For a symmetric configuration, the c-mode corresponds to the even mode and the π-mode to the odd mode. As a result of the fact that the fields are pulled inside the substrate, the attenuation increases when the frequency increases.

Figs. 7 and 8 show, respectively, the elements of the capacitance $\mathbf{C}$ and inductance $\mathbf{L}$ matrix as a function of frequency for the PI- and RI-model. The scale of the figure shows no difference between the results of the PI- or RI-model. This is also the case between C_{12} and C_{21}, and between L_{12} and

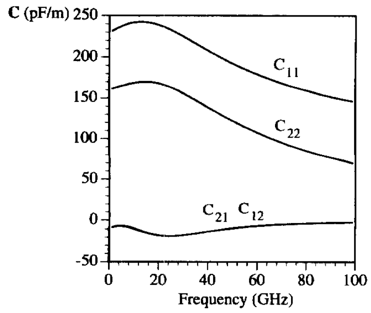

Fig. 7. Elements of the capacitance matrix **C** for the PI- and RI-model for the structure of Fig. 5.

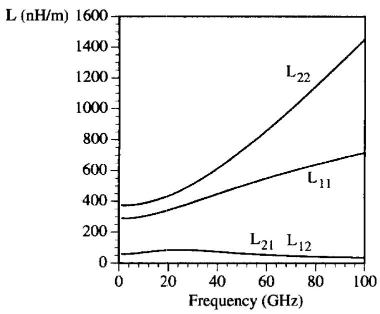

Fig. 8. Elements of the inductance matrix **L** for the PI- and RI-model for the structure of Fig. 5.

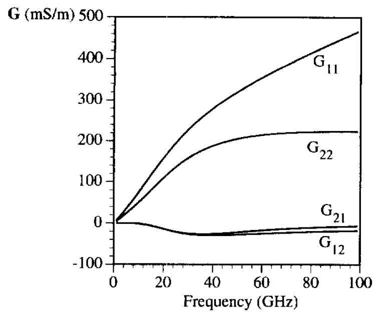

Fig. 9. Elements of the conductance matrix **G** for the PI-model for the structure of Fig. 5.

L_{21}. To get an idea of the differences between both models, we give the **C** and **L** matrices at 50 GHz for the PI-model

$$\mathbf{C} = \begin{pmatrix} 192.24 & -9.222 \\ -9.219 & 122.07 \end{pmatrix} \mathrm{pF/m}$$

$$\mathbf{L} = \begin{pmatrix} 499.50 & 62.032 \\ 61.992 & 730.27 \end{pmatrix} \mathrm{nH/m} \qquad (37)$$

and for the RI-model

$$\mathbf{C} = \begin{pmatrix} 192.30 & -9.262 \\ -9.262 & 122.04 \end{pmatrix} \mathrm{pF/m}$$

$$\mathbf{L} = \begin{pmatrix} 499.36 & 62.175 \\ 62.175 & 730.51 \end{pmatrix} \mathrm{nH/m}. \qquad (38)$$

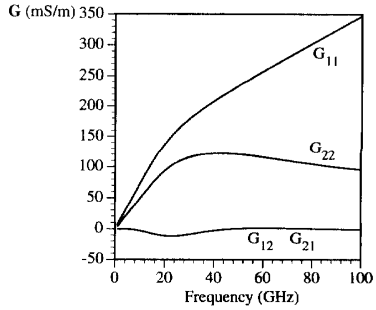

Fig. 10. Elements of the conductance matrix**G** for the RI-model for the structure of Fig. 5.

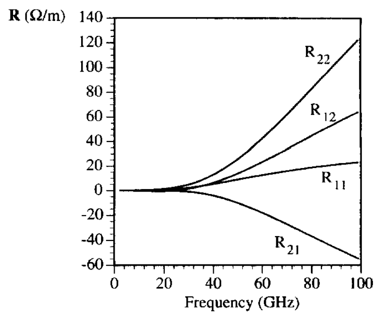

Fig. 11. Elements of the resistance matrix **R** for the PI-model for the structure of Fig. 5.

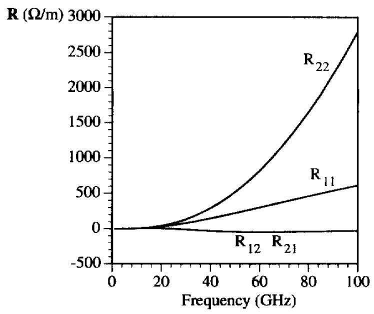

Fig. 12. Elements of the resistance matrix **R** for the RI-model for the structure of Fig. 5.

Figs. 9 and 10 show the conductance matrix **G** for the PI- and RI-model respectively. The nonreciprocity in the PI-model for the **G**-matrix is clearly visible in Fig. 9. Note the quite substantial differences between G_{11} and G_{22} in both models for the higher frequencies. Finally, Figs. 11 and 12 show the resistance matrix **R** for the PI- and RI-model respectively. First, we note that the losses inside the substrate also result in an **R**-matrix. However, the elements of **R** remain small when compared to $\omega\mathbf{L}$. Note also that the PI-model results in a strong nonreciprocal **R**-matrix. Remark further that the values of **G** are larger in the RI-model and that the values of **R** are larger in the PI-model. Finally, it must be said that the current hypothesis in both the RI- and PI-model becomes questionable at higher frequencies because the currents flowing inside the lossy substrate become more significant.

VIII. Conclusion

We have proposed a new consistent reciprocal transmission line model for hybrid anisotropic lossy waveguides. This model is based on reciprocity requirements instead of power requirements. We also studied the determination of the impedance level of the transmission lines in various configurations. We have generalised the classical voltage or current related models for multiconductor waveguides to more general models for general waveguide structures. The difference between a power related and reciprocity related transmission line model was illustrated in an example.

Appendix A

In this appendix we will give a short proof of the orthogonality relation (5). If ($\mathbf{E}_a$, $\mathbf{H}_a$) and ($\mathbf{E}_b$, $\mathbf{H}_b$) are solutions of the Maxwell equations for different sources then the following reciprocity relation [14] holds for reciprocal materials (i.e., $\mathbf{eps}^T = \mathbf{eps}$ and $\mathbf{mu}^T = \mathbf{mu}$) :

$$\oint_{S^{\text{tot}}} \{\mathbf{E}_a \times \mathbf{H}_b - \mathbf{E}_b \times \mathbf{H}_a\}.\mathbf{u}_n dS = 0 \qquad \text{(A.1)}$$

where S^{tot} is a closed surface bounding a source free region and $\mathbf{u_n}$ is the unit normal on the surface. Now, we apply this relation to a section of a waveguide of length L, i.e., $0 < z < L$, for two modes propagating in the positive z-direction. Upon inserting

$$\begin{aligned} \mathbf{E}_a &= \mathbf{E}_f(x,y)\exp(-j\beta_f z) \qquad \mathbf{H}_a = \mathbf{H}_f(x,y)\exp(-j\beta_f z) \\ \mathbf{E}_b &= \mathbf{E}_g(x,y)\exp(-j\beta_g z) \qquad \mathbf{H}_b = \mathbf{H}_g(x,y)\exp(-j\beta_g z) \end{aligned} \qquad \text{(A.2)}$$

in (A.1), one obtains

$$\exp[-j(\beta_f+\beta_g)L] \iint_{S(z=L)} \{\mathbf{E}_f \times \mathbf{H}_g - \mathbf{E}_g \times \mathbf{H}_f\}.\mathbf{u}_z dS - \iint_{S(z=0)} \{\mathbf{E}_f \times \mathbf{H}_g - \mathbf{E}_g \times \mathbf{H}_f\}.\mathbf{u}_z dS = 0 \qquad \text{(A.3)}$$

where $S(z = 0)$ and $S(z = L)$ are, respectively, the cross-sectional planes of the waveguide at the positions $z = 0$ and $z = L$. The contribution of the surface at infinity is zero. Because the modal field patterns $(\mathbf{E}_f, \mathbf{H}_f)$ and $(\mathbf{E}_g, \mathbf{H}_g)$ are z-independent, (A.3) yields

$$\iint_S \{\mathbf{E}_f \times \mathbf{H}_g - \mathbf{E}_g \times \mathbf{H}_f\}.\mathbf{u}_z dS = 0 \qquad \text{(A.4)}$$

where S is an arbitrary cross-section of the waveguide. Next, we change the direction of mode g in the waveguide. When the material parameters are of the form (4) the field corresponding to this mode is given by

$$\mathbf{E}_b = \mathbf{E}_g(x,y)\exp(j\beta_g z) \quad \mathbf{H}_b = -\mathbf{H}_g(x,y)\exp(j\beta_g z). \qquad \text{(A.5)}$$

This can be checked easily with (33). When using (A.5), (A.4) should be replaced by, provided $\beta_f \neq \beta_g$

$$\iint_S \{-\mathbf{E}_f \times \mathbf{H}_g - \mathbf{E}_g \times \mathbf{H}_f\}.\mathbf{u}_z dS = 0. \qquad \text{(A.6)}$$

Combining (A.4) and (A.6) results into the orthogonality relation (5). We emphasise that (5) is only valid for waveguides with material parameters of the form (4). More general materials such as chiral materials are either not reciprocal, not bi-directional or do not satisfy (A.5). Exceptions are waveguides with special symmetries in their cross-section [9].

Appendix B

In this appendix we express the scattering matrices $\mathbf{S}_{ij}$ $(i,j = 1,2)$ with respect to the modal field profiles, for an abrupt junction of two waveguides. Assume that a mode $f(f = 1,2,\ldots,C)$ in waveguide 1 is incident on the junction at $z = 0$. Now we express the continuity of the transverse field components at the junction between this incident mode and the reflected and transmitted modes in both waveguides

$$\begin{aligned} &\mathbf{E}_{\text{tr},1,f}(x,y)K_{1,f}^+ + \sum_g \mathbf{E}_{\text{tr},1,g}(x,y)K_{1,g}^- = \sum_g \mathbf{E}_{\text{tr},2,g}(x,y)K_{2,g}^+ \\ &\mathbf{H}_{\text{tr},1,f}(x,y)K_{1,f}^+ - \sum_g \mathbf{H}_{\text{tr},1,g}(x,y)K_{1,g}^- = \sum_g \mathbf{H}_{\text{tr},2,g}(x,y)K_{2,g}^+. \end{aligned} \qquad \text{(B.1)}$$

The summations run over the full set of modes (discrete and continuous spectrum) in both waveguides. If the first (second) equation of (B.1) is vector multiplied by $\mathbf{H}_{\text{tr},1,h}$ ($\mathbf{E}_{\text{tr},1,h}$) and integrated over the cross-section one finds, using the normalisation relations (5) and the definitions (18)

$$\begin{aligned} \delta_{hf}K_{1,f}^+ + S_{11,hf}K_{1,f}^+ &= \sum_{g=1}^{C} A_{hg}S_{12,gf}K_{1,f}^+ \\ \delta_{hf}K_{1,f}^+ - S_{11,hf}K_{1,f}^+ &= \sum_{g=1}^{C} B_{hg}S_{12,gf}K_{1,f}^+. \end{aligned} \qquad \text{(B.2)}$$

We have limited the number of modes to C and made the assumption that the other modes can be neglected, which is of course an approximation. The matrices $\mathbf{A}$ and $\mathbf{B}$ in (B.2) contain the projections of the modes in one waveguide onto the modes in the other waveguide

$$\begin{aligned} A_{hg} &= \frac{1}{2}\iint_S \{\ \mathbf{E}_{2,g}(x,y) \times \mathbf{H}_{1,h}(x,y)\ \}.\mathbf{u}_z dS \\ B_{hg} &= \frac{1}{2}\iint_S \{\ \mathbf{E}_{1,h}(x,y) \times \mathbf{H}_{2,g}(x,y)\ \}.\mathbf{u}_z dS. \end{aligned} \qquad \text{(B.3)}$$

If (B.2) has to be valid for all $K_{1,f}^+$ one finds in matrix notation

$$\begin{aligned} (\mathbf{U} + \mathbf{S}_{11}) &= \mathbf{A}\ \mathbf{S}_{12} \\ (\mathbf{U} - \mathbf{S}_{11}) &= \mathbf{B}\ \mathbf{S}_{12}. \end{aligned} \qquad \text{(B.4)}$$

Vector multiplication of the first equation of (B.1) by $\mathbf{H}_{\text{tr},2,h}$ and integration over the cross-section shows by comparison with the first equation of (B.4) that $\mathbf{A} = (\mathbf{B}^T)^{-1}$. Finally from

(B.4) we can determine $\mathbf{S}_{11}$ and $\mathbf{S}_{12}$

$$\begin{aligned}\mathbf{S}_{12} &= 2[(\mathbf{A}^T)^{-1} + \mathbf{A}]\\ \mathbf{S}_{11} &= [\mathbf{A} - (\mathbf{A}^T)^{-1}][(\mathbf{A}^T)^{-1} + \mathbf{A}].\end{aligned} \quad \text{(B.5)}$$

By considering modes incident on the junction from the right waveguide one shows that $\mathbf{S}_{21} = \mathbf{S}_{12}$ and $\mathbf{S}_{22} = [(\mathbf{A}^T)^{-1} - \mathbf{A}][(\mathbf{A}^T)^{-1} + \mathbf{A}]$.

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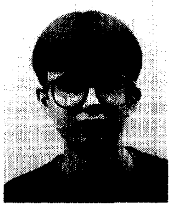

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