A Model of Coplanar Waveguide Type Coupled Exponential Transmission Lines on semiconductor substrates

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Abstract

A new coupled exponential transmission lines (CETL) on semiconductor substrate are designed using an efficient simple method which is the transmission line theory combined with the CETL equivalent circuit. For this an analytical model is given to determine the Sij-parameters of the CETL versus frequency. The simulated results are compared to those of the coupled uniform transmission lines (CUTL). This comparison reveals that the use of an exponential shape of the lines and a semiconducting substrate contribute to obtaining a bandwidth of such coupler wider than that of the conventional ones formed by uniform transmission lines and dielectric substrates.

I-Introduction

For many years, considerable efforts have been devoted to the design and realisation of monolithic microwave integrated circuits (MMIC's). Once fabricated, monolithic circuits proved to be difficult to tune for optimum performance, wich is a major drawback. Hence, accurate theoretical models of MMIC components are required to predict the behaviour of these structures. Such characterization requires a mathematical rigorous solution for the field in the particular structures. Due to some applications in MIC, some studies of the structures, based on microstrip lines in inhomogeneous medium, have been done, by some authors, both in theoretical and experimental works [1-3]. According to their potential applications, many works have been devoted for the study of the nonuniform transmission lines (NUTL). Among such applications we cite: impedance-matching devices, pulse transformers, resonators and filters [4-8]. The principal and important aspect of the technologies based on NUTL structures is to reduce circuit size, to increase circuit bandwidth, to optimise the overall circuit performances [9] and also to solve possible problems of the wave-interaction between TL and nonlinear loads [10]. Among the commonly used particular types of the NUTL, there are the parabolic, cosinesquared, hyperbolic and the exponential transmission lines (ETL). Such models have been studied deeply both in experimental and theoretical analysis by many authors [10-16]. C. W. Hsue [16] examined the transient responses of the exponential line and then he extended the research to the nonlinear high speed backdriving in In-circuit test. From this study, C. W. Hsue concluded that the pulse signal distortion caused by ETLs could be used to compensate for the mismatched effects commonly associated with nonlinear loads.

Other useful NUTL, are the coupled nonuniform transmission lines (CNUTL) which also have important applications in microwaves, such as directional coupler and filters [17,18]

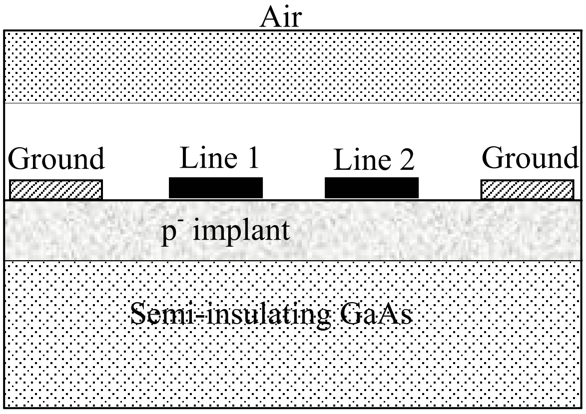
Numerous authors have contributed significantly to the analysis of NUTL by using different techniques. Most of these techniques treat the device as a cascade combination of multiple tiny uniform transmission line sections [19,20]. The inefficiency of this method is clear since the stepped-type of impedance profile is a poor approximation to the real continuously varied profile of a general NUTL. In his work, C. W. Hsue in 1991 [14] used the inverse of Laplace's transformation to present the frequency-domain scattering parameters of lossless ETLs.

Ke Lu [21] presents a new analytical technique including an ideal linear varied NUTL, (LNUTL). This technique is provided in order to replace the conventional one used by C. W. Hsue [15] and it consists in using cascaded ideal LNUTL sections to approximate the impedance profile of an arbitrary NUTL. This method has the merit but it needs much computer time and storage.

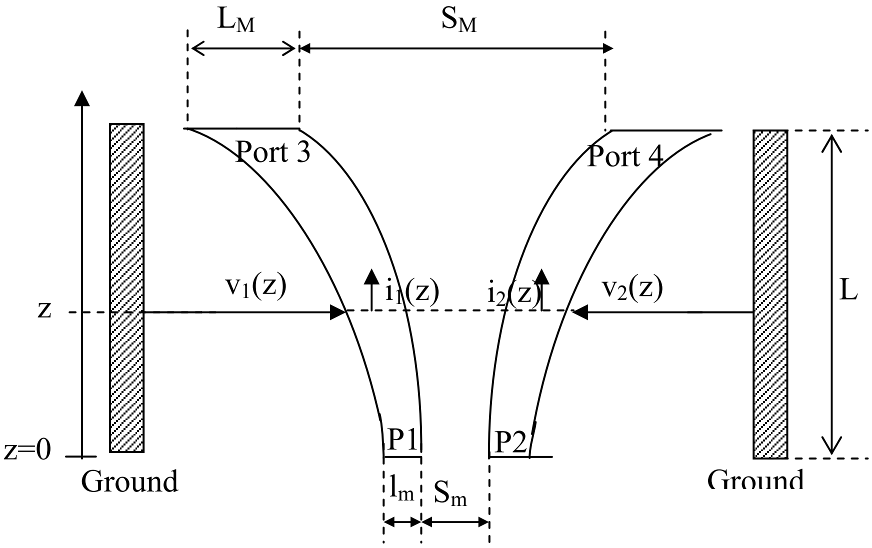
In this work, two new efficient and robust implementations are presented to study the coupled exponential transmission lines. The first one is that the structure is studied in an inhomogeneous medium based on semiconducting substrate. This structure can then give more performances than the one used by M. I. Sobhy [17] on dielectric substrate. The second aspect of novelty is that the technique which is used to characterize the structure consists in using the full-wave analysis based on the equivalent circuit model. This method has the merit not only of being efficient in obtaining accurate results but also in being simple to establish.

II-Analysis of coupled exponential transmission lines

Let us consider a schematic representation of the planar waveguide type exponential transmission lines. The cross section of the structure is depicted in fig 1a and the above view of the two lines and the two grounded electrodes are presented in fig. 1b.



(a)



(b)

Fig. 1- Schematic representation of the CETL :(a) cross-section, and (b) above view.

The device has two coupled exponential transmission lines placed between two grounded ohmic contacts on a thin layer of GaAs substrate. The equivalent circuit is obtained by combining the intrinsic equivalent circuit of the transversal cross section with the longitudinal part. This circuit is depicted in fig. 2.

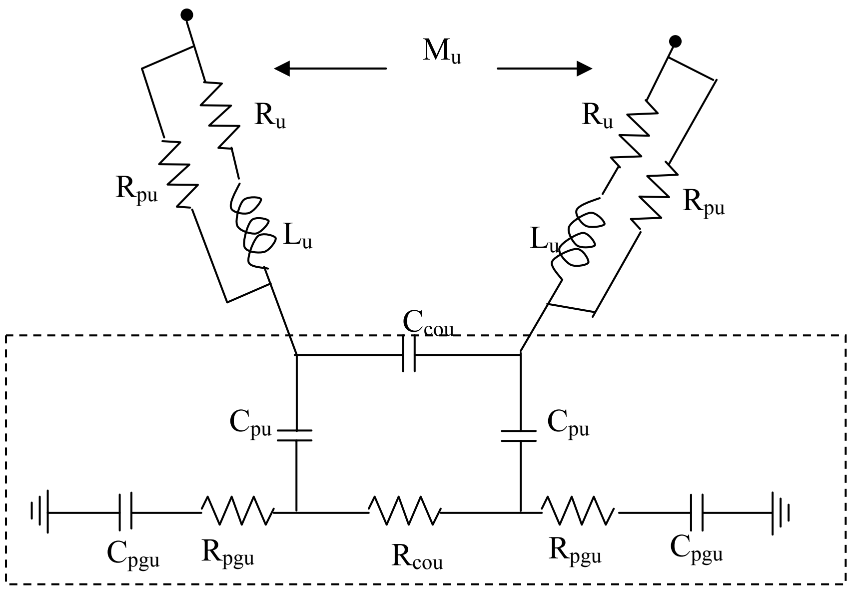


Fig. 2: Equivalent circuit per unit length of the CETL.

Let us note that the intrinsic part of the CETL is described by an admittance matrix Y_{int} as following:

$$[Y_{int}] = \begin{bmatrix} Y_{i11} & Y_{i12} \\ Y_{i21} & Y_{i22} \end{bmatrix}$$

With:

$$Y_{i11} = \left[Z_{co}[Z_p + \frac{Z_1.Z_{1p}}{Z_1 + Z_{1p}}] \right]^{-1} \quad (1)$$

and

$$Y_{i21} = \frac{1}{(Z_{co}+Z_p)(Z_1+Z_{1p})+Z_1Z_{1p}} \left[\frac{Z_1}{Z_1+Z_p} . \frac{Z_1Z_{co}(Z_1+Z_{1p})}{Z_1+Z_{1p}} - (Z_1+Z_{1p}).Z_p + Z_1Z_p \right] .Y_{i11}. \quad (2)$$

Where:

$$Z_{co} = \frac{1}{jC_{co}\omega}\ , \ Z_p = \frac{1}{jC_p\omega}\ , \ Z_{pg} = \frac{1}{jC_{pg}\omega},$$

and

$$Z_1 = R_{pg} + Z_{pg} \quad , \quad Z_{1p} = \frac{Z_1.Z_p}{Z_1+Z_p} + R_{co}$$

$$Y_{i12} = Y_{i21} \quad (3)$$

$$Y_{i11} = Y_{i22} \quad (4)$$

To take into account the actual structure and the real characteristics of the electrodes in the electromagnetic simulation, we apply the coupled transmission lines theory based on the schematic representations of the fig. 1b and fig. 2.

The proposed method is intensely developed in references [22] and [23] where the differential equations can be derived from the equivalent circuit of similar structures. We consider the variation of the inductances and the capacitances as follow:

$$C_{co} = C_{cou}\, e^{-\beta z} \quad (5)$$

$$C_{pg} = C_{pgu}\, e^{\beta z} \quad (6)$$

$$C_{p} = C_{pu}\, e^{\beta z} \quad (7)$$

$$L = L_{u}\, e^{\beta z} \quad (8)$$

$$M = M_{u}\, e^{-\beta z} \quad (9)$$

$$R_{p} = R_{pu}\, e^{-\lambda z}\, e^{-\beta z} \quad (10)$$

$$R_{pg} = R_{pgu}\, e^{\beta z} \quad (11)$$

$$R_{co} = R_{cou}\, e^{-\beta z} \quad (12)$$

$$R = R_{u}\, e^{-\beta z} \quad (13)$$

Where β is the coefficient of the exponential variation.

And the index u indicates that the element values are per unit length.

According to the equivalent circuit proposed in figure 3, it is possible to deduce the following relationships:

$$\frac{\partial i_1}{\partial z} = Y_{i11} v_1 + Y_{i12} v_2 \quad (14)$$

$$\frac{\partial i_2}{\partial z} = Y_{i21} v_1 + Y_{i22} v_2 \quad (15)$$

$$\frac{\partial v_1}{\partial z} = -Z_{11} i_1 - Z_{m} i_2 \quad (16)$$

$$\frac{\partial v_2}{\partial z} = -Z_{m} i_1 - Z_{22} i_2 \quad (17)$$

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where:

$$Z_{11} = Z_1 = \frac{(R_u + j\omega L_u)R_{pu}}{R_u + j\omega L_u + R_{pu}} \tag{18}$$

$$Z_{22} = Z_{11} \tag{19}$$

$$Zm = j\omega M \tag{20}$$

Moreover it is possible to write the equation system as follows:

$$\frac{\partial}{\partial z}\begin{bmatrix} v_1 \\ v_2 \\ i_1 \\ i_2 \end{bmatrix} = \begin{bmatrix} 0 & 0 & -Z_{11} & -Z_m \\ 0 & 0 & -Z_m & -Z_{22} \\ Y_{i11} & Y_{i12} & 0 & 0 \\ Y_{i21} & Y_{i22} & 0 & 0 \end{bmatrix}\begin{bmatrix} v_1 \\ v_2 \\ i_1 \\ i_2 \end{bmatrix} \tag{21}$$

The specific solutions are chosen as series of functions:

$$i_\alpha = \sum_i i_i e^{\gamma_i z} \tag{22}$$

$$v_\alpha = \sum_i v_i e^{\gamma_i z} \tag{23}$$

Where :

α = d, g

v_α and i_α are the eigenvectors.

v_i and i_i are the magnitude of the eigenvectors.

γ_i is the complex eigenvalue (propagation constant or wave vector) which is the solution of the characteristic equation :

$$\begin{aligned}&\gamma^4 + \gamma^2 \left[Z_m\left(Y_{i12} + Y_{i21}\right) - Z_{22}Y_{i22} - Z_{11}Y_{i11}\right] + Z_{11}Z_{22}Y_{i11}Y_{i22} \\ &+ Z_m^2 Y_{i12}Y_{i21} - Z_{11}Z_{22}Y_{i12}Y_{i21} - Y_{i11}Y_{i22}Z_m^2 = 0\end{aligned} \tag{24}$$

Let us note:

$A_1 = Z_{11}Y_{i11} - Z_mY_{i12}$

$A_2 = Z_{22}Y_{i22} - Z_mY_{i21}$

$B_1 = Z_mY_{i22} - Z_{11}Y_{i21}$

$B_2 = Z_mY_{i11} - Z_{22}Y_{i12}$

The equation (24) becomes.

$$\gamma^4 - \gamma^2 (A_1 + A_2) + A_1A_2 - B_1B_2 = 0 \qquad (25)$$

This equation (25) has four solutions ($\pm\gamma_e, \pm\gamma_o$) corresponding to the two possible propagation modes: the even mode and the odd mode.

If we consider:

$$A = \frac{A_1 + B_1}{2}$$
$$B^2 = B_1B_2$$

It is possible to obtain:

$$\gamma_e^{\,2} = A + K$$
$$\gamma_o^{\,2} = A - K$$

where :

$$K = \frac{A_1 - A_2}{2} + \sqrt{\frac{(A_1 - A_2)^2}{2} + B^2}$$

Consequently, according to the previous equations the line 1 and line 2 voltage-current can be deduced as follows:

$$v_1 = (a_1\exp(-\gamma_e z) - a_2\exp(\gamma_e z))\,\gamma_e K + (a_3\exp(-\gamma_o z) - a_4\exp(\gamma_o z))\,\gamma_e B_1 \qquad (26)$$

$$v_2 = (a_1\exp(-\gamma_e z) - a_2\exp(\gamma_e z))\,\gamma_e B_2 + (a_3\exp(-\gamma_o z) - a_4\exp(\gamma_o z))\,\gamma_e K \qquad (27)$$

$$i_1 = (a_1\exp(-\gamma_e z) - a_2\exp(\gamma_e z))(Y_{i11}K - Y_{i12}B_2) + (a_3\exp(-\gamma_o z) - a_4\exp(\gamma_o z))(Y_{i12}K + Y_{i11}B_1) \qquad (28)$$

$$i_2 = (a_1\exp(-\gamma_e z) + a_2\exp(\gamma_e z))(Y_{i22}B_2 - Y_{i21}K) + (a_3\exp(-\gamma_o z) + a_4\exp(\gamma_o z))(Y_{i21}B_1 + Y_{i22}K) \qquad (29)$$

The precision of analysis consists in getting the impedance matrix $[Z]_{ce}$ of the CETL.

Starting from figure 1(b) the principal boundary conditions are:

$v_1(0) = v_{p1}$, $v_1(L) = v_{p3}$, $v_2(0) = v_{p2}$, $v_2(L) = v_{p4}$

$i_1(0) = i_{p1}$, $i_1(L) = i_{p3}$, $i_2(0) = i_{p2}$, $i_2(L) = i_{p4}$

The combination of these boundary conditions with equations (26), (27), (28) and (29), leads to obtaining the constants a_1, a_2, a_3, a_4 as a function of v_{pi} and i_{pi}.

The indices pj in v_{pj} and i_{pj} indicate the port j of the structure.

Therefore, it is possible to obtain the impedance matrix $[Z]_{ce}$.

$$\begin{pmatrix} v_{p1} \\ v_{p2} \\ v_{p3} \\ v_{p4} \end{pmatrix} = \begin{pmatrix} Z_{11ce} & Z_{12ce} & Z_{13ce} & Z_{14ce} \\ Z_{21ce} & Z_{22ce} & Z_{23ce} & Z_{24ce} \\ Z_{31ce} & Z_{32ce} & Z_{33ce} & Z_{34ce} \\ Z_{41ce} & Z_{42ce} & Z_{43ce} & Z_{44ce} \end{pmatrix} \begin{pmatrix} i_{p1} \\ i_{p2} \\ i_{p3} \\ i_{p4} \end{pmatrix}$$

Finally, the matrix [S] is obtained as follows:

$$[S] = [[Z]_{ce} + Z_0\,[I]]^{-1}.[[Z]_{ce} - Z_0\,[I]]$$

Where Z_0 is the normalised impedance $Z_0 = 50\,\Omega$ and [I] is the unitary matrix.

III-Results and illustration

The structure studied is characterized by the measured electric parameters per unit length [2] given in table (I). For this, a simulated program is done and its numerical results are presented.

Table (I)

Element	Value per cm
L_u	6.9 nH
C_{pgu}	1.34 pF
C_{pu}	460 pF
R_{pu}	9750 Ω
R_{pgu}	1/9.3 Ω
C_{cou}	870 pF
R_u	186 Ω

The structure is illustrated by schematic representation fig. 1(a and b) and the equivalent circuit given in Fig. 2.

We try to compare the results concerning the directivity of the CETL with those of the CUTL.

In fig. 3 and 4, we mark the range of frequencies δF in which the coupler could be applied.

It should be noted from fig. 3 and 4 that the directivity of the CETL exists in a wider bandwidth than the one of the CUTL. In addition we can observe from these figures that, at high frequencies, the directivity of the coupler decreases due to the unequal phase velocity of even and odd modes [18].

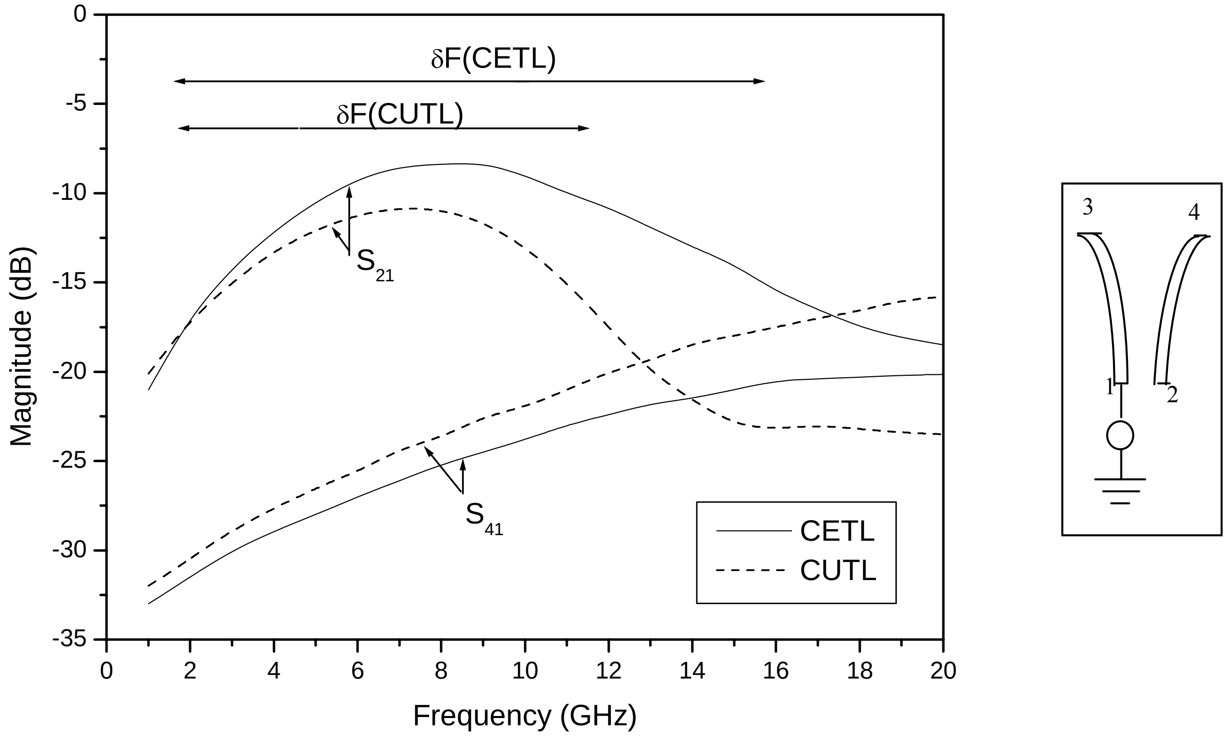

Fig. 3: Magnitude of the scattering parameters S_{21} and S_{41} of a CETL. (L = 12 mm, l_m = 0.4 mm, S_m = 0.04 mm, β = 50 m^{-1})

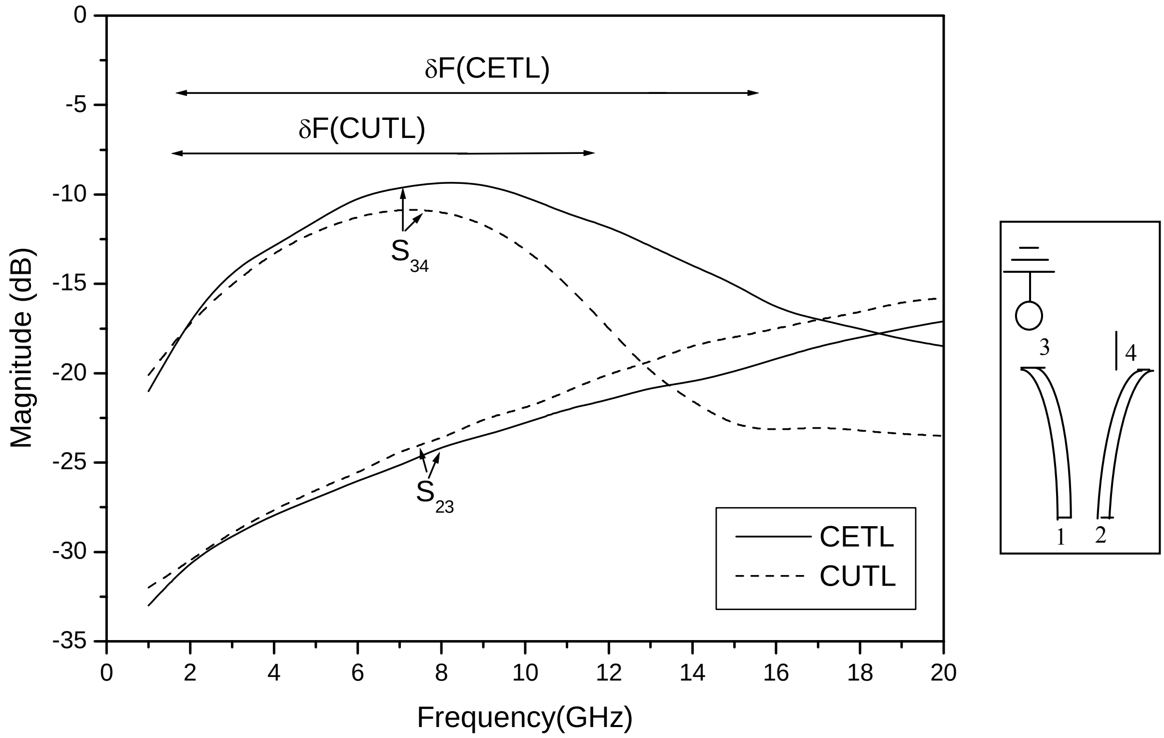

Fig. 4: Magnitude of the scattering parameters S_{34} and S_{23} of a CETL. (L = 12 mm, l_m = 0.4 mm, S_m = 0.04 mm, β = 50 m^{-1})

IV-Conclusion

The transmission lines theory (TLT) in quasi-TEM analysis is investigated to design the coupled exponential transmission lines (CETL) in an inhomogeneous medium with semiconducting substrate. According to an equivalent circuit, and taking into account the exponential variation of all the circuit elements, we determine the Sij parameters versus frequency. This method has the merit not only of being efficient in obtaining accurate results [22,23] but also in being simple to establish. Compared to the results obtained for the coupled uniform transmission lines (CUTL), the CETL shows a wider bandwidth response than that of the CUTL. It is also important to notice that the bandwidth is larger than that one corresponding to the structure which is optimized for the same coupling coefficient values, but with dielectric substrate [17]. In addition to the advantage given by the exponential shape of the lines, the use of the semiconductor substrate plays an important role in widening the frequency band of the coupler performances.

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