Journal of Microwaves, Optoelectronics and Electromagnetic Applications, Vol. 22, No. 2, June 2023
DOI: http://dx.doi.org/10.1590/2179-10742023v22i2270080

A New Theoretical Approach of Studying Resonances in Single Finline Transitions

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***Abstract* - In this article an innovative method of studying and removing the resonances, inherently exhibit by some waveguide to microstrip transitions, is presented. By modeling an equivalent circuit, this new approach allows to obtain the constructive parameters of a finline to microstrip transition, only using the values of the resistance and capacitance components of the equivalent circuit. This procedure will allow small microwave design Companies to realize these transitions only implementing circuit analysis software, and not having to afford electromagnetic analysis software, which are very expensive and time-consuming. A full 3D electromagnetic analysis confirms that the simulation results are in excellent agreement with the results obtained by the new equations discussed in this work.**

Index Terms – Equivalent Circuit, Finline to Microstrip Transitions, Resonance, Spatial Power Combiners.

I. INTRODUCTION

In the last decades, developing technologies for terrestrial, satellite communication, RADAR applications has always been an environment in which TWTs (Travelling Wave Tubes) has been widely used for HPAs (High Power Amplifiers). Vacuum Tubes (VTs) can perform high power outputs wideband and with a center frequency up to hundreds of gigahertz [1-5]. However, they have several drawbacks caused by tubes technologies. In fact, they are heavy and with heat dissipation issues, a quite long warm up time and their dimensions are remarkable. Moreover, when power VTs are used in many systems, quite often the system suffers of single point failure. HPAs can work either in pulsed or continuous mode and tubes are nowadays the best option in case of pulsed mode for megawatt output powers in RADAR applications. Concerning SSPAs (Solid State Power Amplifiers) in recent years kilowatt – high frequency continuous wave have been developed along with new HPAs design strategies and architectures [6-15]. The most efficient and compact structure capable of delivering large amounts of power in the range of millimeter waves is certainly the one called Spatial Power Amplifier (SPA), or Spatial Power Combiner (SPC). SPAs are lighter, have graceful degradation and high reliability, also requiring low voltage supplies to work properly. SPAs have been developed using high power – high efficiency MMICs based on semiconductors such as GaN, GaAs and InP. Different finline to microstrip (FLuS) transitions [16] have been developed through the years with different topologies

ISSN 2179-1074

and EM field modes that can be pulled out from these structures using probes [17-19]. Basically, this type of transitions is based on Vivaldi antenna topologies [20, 21] and they are massively implemented in broadband applications, even in the THz domain and with very interesting harvesting capabilities [22-24]. When FLuS are used inside waveguides, because of the parallel – in air division, the losses introduced by these devices are very small and they are only dependent by the number of amplifiers involved [25]. One of the main issues to deal with in Spatial Power Combining structures is represented by electromagnetic resonances. They can be caused either by the presence of a mechanical cavity, or by the internal RF components of the SPA. One of these components is the aforementioned FluS. For specific applications, a FluS can be made by a single couple of ground and hot metallic lines and in this case, it is called “Single Finline”. Single Finlines are always resonating and an accurate model and method to remove this resonance, ensuring high performance, has not been developed yet.

In this paper, a project method based on a complete equivalent circuit is developed, presenting new design equations that links components to physical dimensions, in order to realize non resonant finline to microstrip transitions without referring to EM simulators, decreasing significantly time and costs needed to produce a device. A WR90 transition is simulated, showing the expected results with this new resonance removing technique.

II. RESONANCE

In Figure 1 is represented the exponential FLuS transition [26-28] considered in this paper. The substrate material is Al_2O_3, with a relative dielectric permittivity $\mathcal{E}_r = 9.9$. It is visible the equivalent cylindrical cavity which resonates [29-31] in the operative bandwidth once the single finline is placed inside a waveguide. The resonance has a very high Q (Quality Factor), so it can be represented as a resonant circuit cell coupled with the main transmission line. The resonant frequency can be evaluated with the following equation [32]:

$$f_{n,11} = \frac{c}{2\pi\sqrt{\epsilon_{eff}}}\left[\left(\frac{P_{n1}}{R_s}\right)^2 + \left(\frac{\pi}{b}\right)^2\right]^{\frac{1}{2}} \tag{1}$$

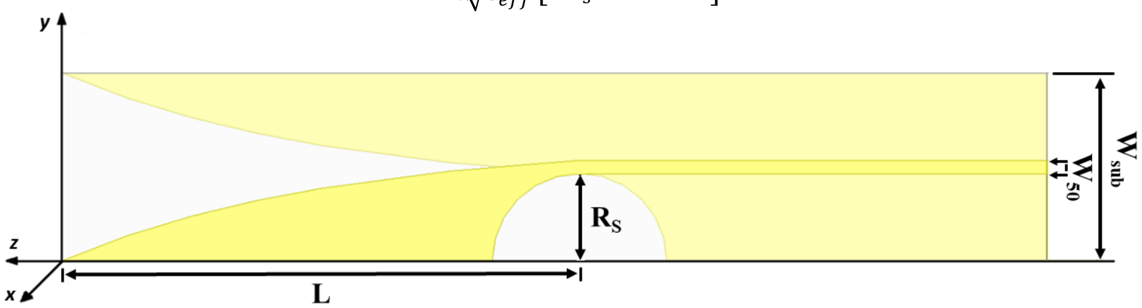

Fig. 1. Finline to microstrip transition, FLuS,

In Figure 1, L is the length of the exponential transition, while z is the coordinate across the waveguide. In equation (1), ϵ_{eff} needs to be experimentally evaluated and P_{n1} is the zero of the Bessel function of first species and order n. From resonating cavities theory [16], we can assume that decreasing or increasing cavities dimensions allows to, respectively, increase and decrease the resonance frequency (f_r). It is also possible, from literature [29, 30] to shift the resonance frequency by filling the dielectric surface, on the hot plate side, with a metal semicircle; this process is trial and error, so there is yet no method to remove the resonance. Another technique is to partially remove the dielectric in the resonating region, decreasing the dielectric constant; at the best author's knowledge, also this technique is not a methodical procedure.

A new methodical way proposed in this article consist in moving the resonance outside the waveguide operating bandwidth by varying the design parameter R_s. When a FLuS is designed, R_s has a fixed value ($\overline{R_s}$) which depends on the constructive parameters of both the waveguide and the microstrip, according to the following equation [26]:

$$\overline{R_s} = w_0\left(1 - w_f\right) - w_{50} \tag{2}$$

$$w_0 = \frac{w_{50} + w_{sub}}{2\left(1 - w_f\right)} \tag{3}$$

$$w_f = \frac{Z_{\mu s}}{Z_{wg}} \tag{4}$$

In this equation:

- w_{50} is the width of the microstrip hot plate.
- w_{sub} is the width of the dielectric, which is equal to the small side of the waveguide b.
- $Z_{\mu s}$ is the output impedance of the structure, which is the characteristic impedance of the microstrip.
- Z_{wg} is the characteristic impedance of the fundamental mode of the waveguide.

Using a WR90 waveguide, whose long side wall is $a = 22.86mm$ and the short side wall is $b = 11.16mm$, $Z_{\mu s} = 50\Omega$ and $Z_{wg} = 400\Omega$ with $w_{50} = h_{sub} = 0.254mm$, $w_0 = 6.52mm$ and $w_f = 0.125$, $L = 70mm$. $\overline{R_s}$, from equation (2), is $4.953mm$.

It is possible, then, to change R_s values: the more R_s decreases, the more the resonance frequency increases and viceversa. Increasing the radius is not suitable because the metallization design in this case results in some exotic design with important influences on the signal phase. Decreasing the radius dimension, instead, is possible modifying the design equations. The new equation proposed for the ground profile in this paper is here shown:

$$w_{GND}(z) = \frac{R_s + w_{50}}{1 - w_f}\left(exp\left(\frac{z}{L}\right) ln\left(w_f\right) - 1\right)\left(\frac{w_{sub} - R_s}{w_{50} + R_s}\right) \tag{5}$$

The resulting upper conductor profile consists now of two transitions: a first exponential transition and a second one, which in this paper is chosen linear.

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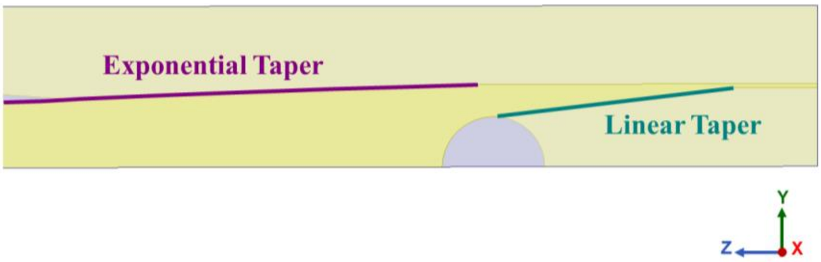

Fig. 2. Modified FLuS whit new shape, in order to cancel the resonance

This solution allows to keep the phase relation between a single finline and other multiple finline eventually used to increase the output power [6-14].

To propose a general method, when there is the need to increase the resonance frequency, as in the case of this work, the radius has to be decreased. Instead, when there is the need to decrease the resonance frequency, the filling gold semicircle can be applied [32]. Such frequency shifting can be represented by increasing or decreasing the value of the capacitance (C_r) coupled to the resonating RLC circuit, as shown in Figure 3. This equivalent circuit will be studied in the following section.

III. EQUIVALENT CIRCUIT

The equivalent circuit developed in this paper is shown in Figure 3. There are the two characteristic impedances, the fundamental mode waveguide impedance Z_{wg} and the microstrip impedance Z_{us}. The losses of these two propagative media are R_g for the waveguide and R_u for the microstrip. Since every FluS performs an impedance transformation, a transformer made by inductors can represent this phenomenon (in Figure 3), while the exponential transition of the conductor is represented by the tapered transmission line (in Figure 3). The resonating cavity and eventually the filling semicircle, are represented by the resonant RLC circuit with a parallel capacitor C_r. This whole resonating section is coupled to the main propagation line by a transformer (in Figure 3).

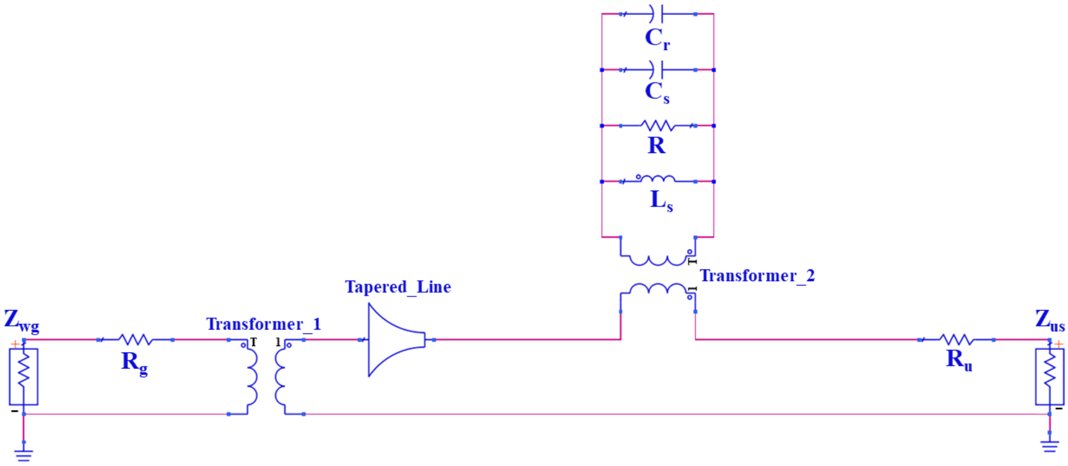

Fig. 3. Equivalent Circuit of a Single FLuS

To realize the required structure with a fast and effective method, the equivalent circuit components need to be related to the physical dimensions through new design equations. These equations are obtained as described in the following: for a discrete value of R_s without the filling semicircle and a discrete value of R_r without modifying R_s, the equivalent circuit is tuned to the resonance of the physical structure. Doing so, an experimental equation is derived through interpolation between, respectively, the equivalent circuit capacitors *Cs* - *Cr* and *Rs* - *Rr*.

The resistance is chosen $10M\Omega$ to keep high the Quality Factor (Q) of the RLC filter. The inductor is fixed, so that the whole behavior of the resonance is based on C_s and C_r.

In this paper, a WR90 is the waveguide on which the method is tested because WR 90 is widely used for radar and broadband systems. We can predict the resonance frequency using equation (1), so it is $11.465\ GHz$.

IV. NEW EQUATIONS

Several simulations can be performed by varying the values R_s and R_r ; for every R_s value that determines f_r, the RLC circuit (i.e. C_s) is tuned. In the same way, different values of C_r can be found for each R_r. Once this is done, the relation between C_s and R_s (or C_r and R_r) is derived through interpolation. L_s is chosen so that the lowest value of C_s expressed in picofarad matches the resonance with the lowest value of R_s expressed in mm, so that the $R_s - C_s$ relation has the same starting value. In this paper, $L_s = 50.405pH$. This has been done for several waveguides and can be seen that the $R_s - C_s$ relation is a second order function, meanwhile the $R_r - C_r$ is a first order rational relation for every waveguide analyzed.

Having the original physical structure $R_s = \overline{R_s} = 4.953mm$, the equivalent circuit is tuned to the same resonance frequency; thus, $C_s = 3.824\ pF$ and $C_r = 0\ pF$. The equivalent circuit behaviour is shown in Figure 4.

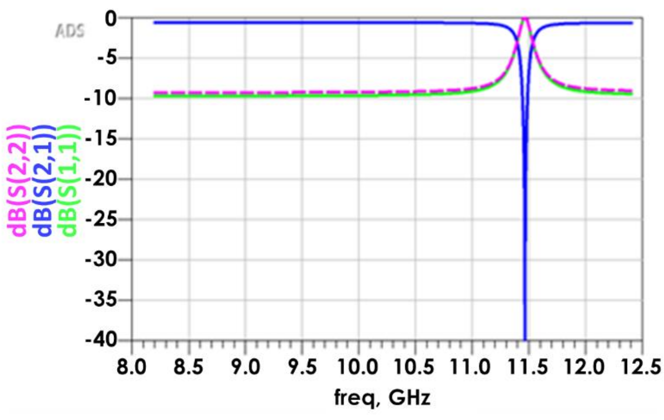

Fig. 4. *S Parameters of the Equivalent circuit, related to* $R_S = 4.953mm$

Decreasing or increasing f_r depends on the position of the resonance regarding the waveguide bandwidth: the bandwidth of the WR90 is $8.2GHz - 12.4\ GHz$ and f_r is close to the upper bandwidth

ISSN 2179-1074

limit, so we propose this new resonance removing method based on decreasing R_s dimension. The equation derived is the following, with R_s expressed in mm and C_s expressed in pF:

$$R_s = -1.118C_s^2 + 9.493C_s - 15.03 \tag{6}$$

The interpolation curve used to derive equation (6) is shown in the following figure:

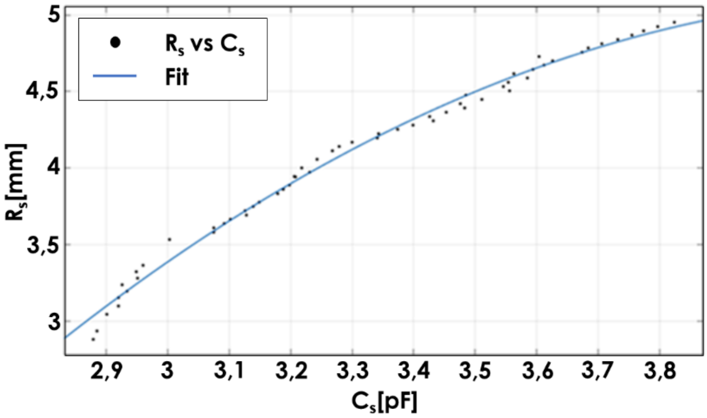

Fig. 5. *Interpolation Curve between C_s and R_s*

It is now necessary to move the resonance out of the operating bandwidth, so a Δf is chosen so that $f_r = 12.4GHz + \Delta f$, in this case, $\Delta f = 0.7GHz$. It is important to have a margin between f_r and $12.4GHz$ because a back-to-back structure is necessary to a full SPA design [6-13]; the back-to-back structure will be tested later in this paper. Therefore, $f_r = 12.4GHz + 0.7GHz = 13.1GHz$ and we derive $C_s = 2.92pF$. Using the equation (6) we observe that $R_s = 3.157mm$.

V. ELECTROMAGNETIC SIMULATIONS

To validate the effectiveness of the obtained quantities, it is necessary to design and simulate, using electromagnetic analysis software, a 3D CAD model of the finline transition. The software used for the electromagnetic characterization is Ansys HFSS.

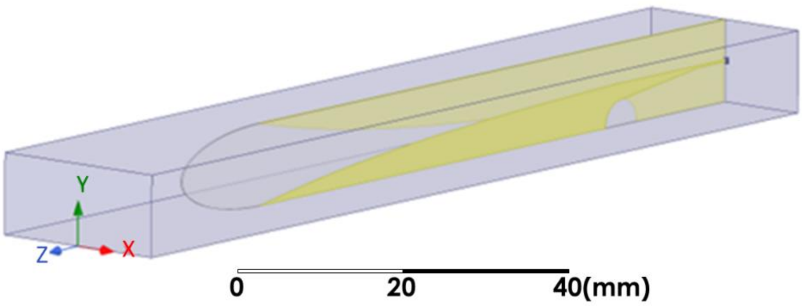

Fig. 6. Single FLuS in WR90 Waveguide for EM Simulations

Since this is a waveguide to microstrip transition, the set-up of the simulations consists of a waveguide section, in this case a WR 90 waveguide (*a* = 22.86 mm and *b* = 10.16 mm), inside which the FluS transition is placed. The finline is made by a ceramic substrate of Al_2O_3, with a thickness of

254 µm and two layers of gold placed on the upper and the bottom side of the substrate, which are the hot and the cold plate respectively.

As expected and as anticipated in the previous sections, the radius of the semicircle (R_s) has decreased, thus shifting the resonance towards higher frequencies. Since the bandwidth of the finline is greater than the bandwidth of the waveguide, it is sufficient to move the resonance outside the spectrum covered by the waveguide, in order to not have limitations in terms of bandwidth.

The design of this transition was carried out with this perspective and the simulations demonstrate the validity of this approach, as it is possible to see in the following figure:

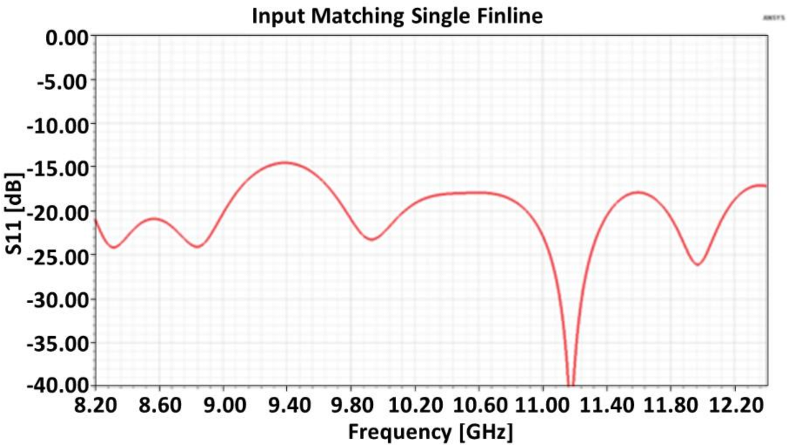

Fig. 7. Input Matching of the Single Non-Resonant Finline

According to the matching at the input port of the system, in this case the waveguide port, it is possible to notice how the natural resonant frequency of the single finline has been removed, by pushing it outside the WR90 bandwidth. The input matching is always or at most equal to 15dB on the entire WR90 bandwidth (8.2 – 12.4 GHz).

Concerning the losses of the structure, it is possible to estimate them by the scattering parameter S21, as given in the following figure:

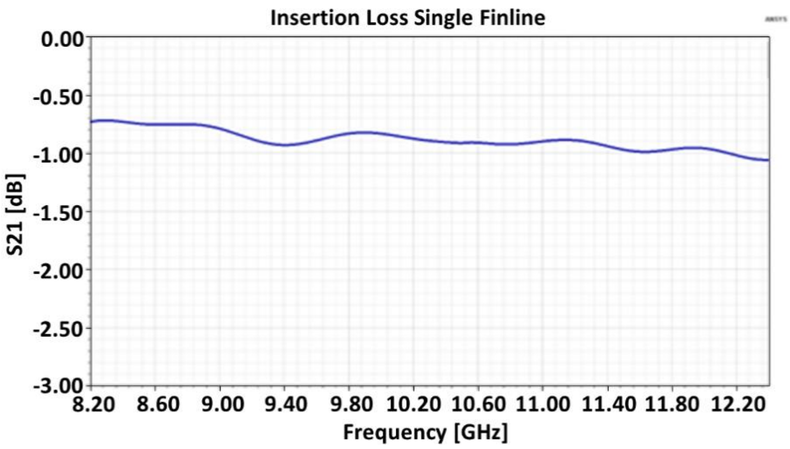

Fig. 8. Insertion Loss of the Single Non-Resonant Finline

As shown by the graph above, the trend of the insertion losses is very flat along the whole band. This flatness further confirms the quality of this transition, especially in power applications, where the gain flatness is a primary specification.

During the project of a power combination structure, such as a spatial power combiner, a fundamental step of the project is to validate the passive structure in which the power amplifiers will be inserted. In the specific case of the structure reported in this work, it is necessary to mirror the entire system, thus obtaining a back-to-back structure, shown in Figure 9:

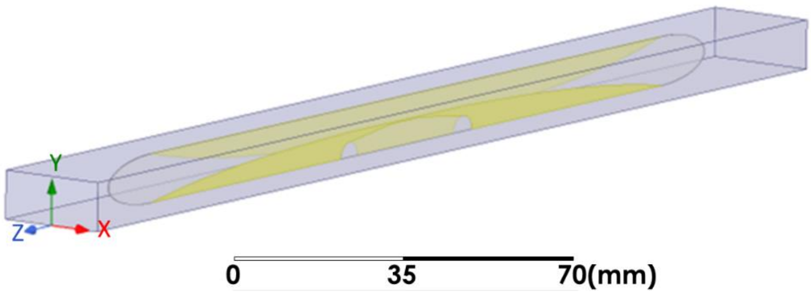

Fig. 9. Single Non-Resonant Finline in Back-to-Back Configuration

Inside a spatial combiner, the monolithic amplifier (MMIC) is placed at the center of the structure. Thus, after the finline has converted the traveling power in the waveguide into a power that propagates inside a microstrip, this power will arrive at the input port of the MMIC. The output power provided by the amplifiers will be sent through the microstrip and the output finline will launch this power inside the waveguide section, in order to reduce the transmission losses and save as much power as possible.

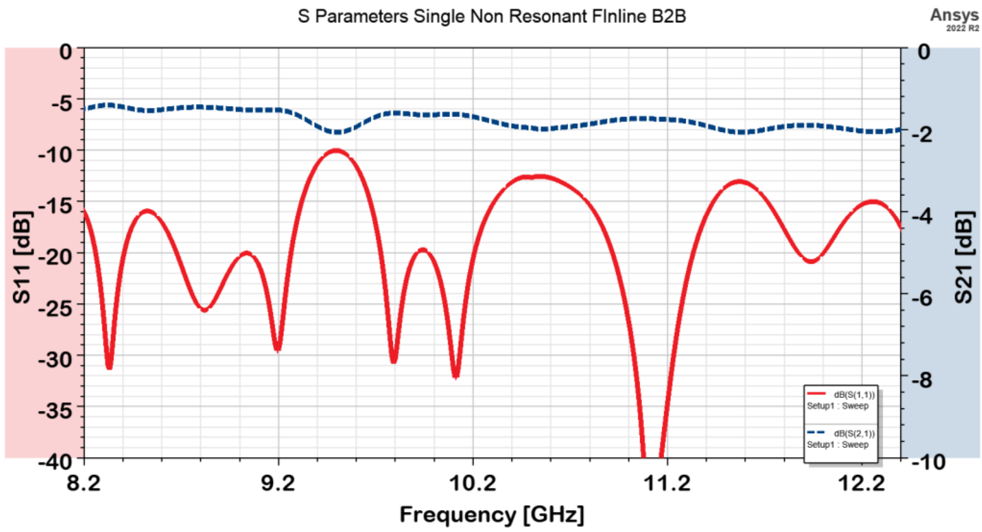

Fig. 10. S Parameters of the Single Non-Resonant Finline in Back-to-Back Configuration. Input Matching (Continuous Line) and Insertion Loss (Dashed Line)

The input matching of the B2B structure is always below the standard value of 10 dB and the insertion losses show an almost flat pattern across the entire bandwidth.

These results show how this innovative approach for resonance cancellation is completely valid and allows to use, in power combination applications for instance, structures such as the single finline. All this without the previous limitations in terms of bandwidth, due to the natural resonance of the FLuS transition. Thanks to the design method reported in this work, the resonance has been eliminated from the useful band.

Electromagnetic simulations confirm the validity of the method proposed in this article.

With our innovative design method presented in this work, it is therefore possible to calculate the construction parameters of the single finline without using electromagnetic simulation software, this last having a considerable cost. From a small business point of view, the application of this design method might be crucial for the growth of the company.

VI. CONCLUSIONS

Starting from a structure considerably known in literature, which is the single finline transition, an accurate equivalent circuit model was made. Since this waveguide to microstrip transition is an intrinsically resonant structure, it has been possible to study and validate an innovative method of moving this resonance outside the working bandwidth.

This process also made it possible to obtain the constructive parameters of the non-resonant finline transition, through the dimensioning of the equivalent circuit components. For a small company just born, this method allows to design these transitions using circuit simulation software exclusively, thus being able not to purchase electromagnetic analysis software, which are extremely expensive and time consuming and requiring dedicated workstations.

The EM simulations carried out in this article have shown how this new theoretical and methodical approach is completely valid and allows to realize waveguide to microstrip finline transitions, whose parameters perfectly fit in the design of microwave and millimeter-wave structures. In fact, a value always below 10 dB of input matching and a quite flat trend of the insertion loss have been achieved.

This innovative method for the removal of resonances through the study of the equivalent circuit will certainly be applied in many other high frequency structures in the coming years.

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