An Integrated Transition of Microstrip to Nonradiative Dielectric Waveguide for Microwave and Millimeter-Wave Circuits

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***Abstract*— Effective interconnects between two or more different waveguides are essential for using hybrid architecture of circuits as well as multipurpose instrumentation and measurements at microwave and millimeter-wave frequencies. In this paper, a new transition of microstrip line to nonradiative dielectric (NRD) guide is reported that makes it possible to design a class of compact NRD waveguide circuits directly integrated with planar-microstrip-based devices and components. A small aperture coupling theory is developed and effectively applied to model the proposed structure. It is found that theoretical results are in good agreement with experiments. The proposed transition promises to be instrumental in integrating planar microstrip devices and components with NRD guide circuits.**

I. Introduction

It HAS ALREADY been recognized that the nonradiative dielectric (NRD) guide [1]–[3] promises to be one of the most attractive building blocks for microwave, and in particular millimeter-wave, applications. Compared with other transmission lines, the NRD demonstrates its simplicity in mechanical assembling, ease of fabrication, and low-loss nature, offering potentially low-cost applications. Moreover, radiation at curved sections and discontinuities is nearly nonexistent. The NRD guide supports the design of various active and passive circuits for microwave and millimeter-wave systems or subsystems. It is particularly useful in applications involving a class of millimeter-wave circuits such as high-quality filters, antenna feeding networks, couplers, and high-quality oscillators. Nevertheless, the planar integrated circuits are still the mostly often used topology in microwave and millimeter-wave systems. Therefore, combining both circuit topologies together and taking their inherent advantages are the key issues for the further development of NRD guide circuits. To do so, an adequate transition integrating the two dissimilar structures is essential. So far, a few of transitions made from other transmission lines to NRD guide have been investigated, e.g., the transition from rectangular waveguide to NRD guide [4] and probe-type transition as well as a stripline to NRD transition [5]. However, none of them is suited for a compact integration of NRD guide with planar structure for circuit applications. In addition, design and fabrication of these transitions are rather difficult and expensive, pointing to no any advantages for the use of a hybrid geometry. As a matter of fact, these transitions were proposed mainly to be used in the NRD-related measurement systems.

To fill this void, a new type of transition proposed for the use of microstrip to NRD guide is presented in the paper that is basically derived from the concept of the aperture coupling [6]–[8]. It is known that the NRD guide is made of a low-loss dielectric material sandwiched in between two parallel metallic plates with a spacing smaller than half wavelength of free space. The proposed hybrid geometry consists of a microstrip line deposited on the top of a ground plane of the NRD guide, leading to a vertical coupling scheme. The transition is effectively achieved by a signal path running from the microstrip line via a slot in the ground plane down to the NRD or vice-versa. In fact, the ground plane of the microstrip line serves directly as one of the parallel plates of the NRD guide structure. With this compact transition, the planar circuits are obviously integrated with the NRD-based components in a single block for the best use of the complementary advantages of two different techniques.

In the present work, the coupling aperture between the two dissimilar guiding structures is studied with resort to the small aperture coupling theory [9], [10]. At first, a four-port network of the coupled microstrip-line and NRD guide is considered, the S-parameter matrix of which can be determined from a knowledge of the equivalent dipole moment of the aperture and the orthonormalized modal vectors of both guides. Next, one of the ports of each guide is open-circuited such that the transition of microstrip line to NRD is formed and modeled. In this way, the propagation properties of the new transition are easily obtained. Electrical performance of the proposed transition is examined. Theoretical and experimental results of a 20-GHz prototype demonstrate that the proposed transition offers wide bandwidth and high coupling efficiency.

II. Description of the Proposed Transition

Fig. 1(a) shows the geometry of the proposed transition that consists of a microstrip line deposited on the top of the NRD guide with a common ground plane. The microstrip line is orthogonal in space with respect to the dielectric strip of the NRD guide. The coupling is achieved through a rectangular slot etched on the ground plane. The coupling slot is oriented perpendicularly to the microstrip line. It is known that the fundamental mode of the microstrip line is a quasi-TEM

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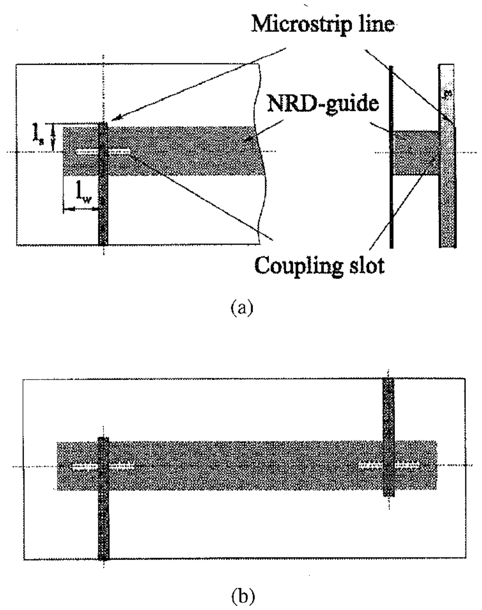

Fig. 1. (a) Geometry and (b) testing setup of the transition from microstrip line to NRD guide.

mode, while there are two nonradiating modes in the NRD guide that are usually refereed to as the LSE_{11} and LSM_{11} [1], respectively. The LSM_{11} mode is preferred for practical applications because it has the lowest propagation loss. The magnetic field of this mode is parallel to the air-dielectric interface of the NRD guide. Therefore, both magnetic fields of the quasi-TEM mode in the microstrip and the LSM_{11} mode in the NRD guide are well matched at the coupling aperture. The operating frequency range is expected to be dependent on the dimensions of the dielectric strip, which can be chosen by the way suggested in [2]. To study experimentally the electrical performance of the proposed transition, a test prototype is fabricated using two microstrip lines that are interconnected through two separated microstrip line-to-NRD transitions with the separation of a NRD guide as shown in Fig. 1(b). Obviously, the input and output remain in the form of microstrip lines that are easily connected to the network analyzer. The transitions can be located in either side of the NRD's ground planes. This is an additional advantage of the transition in its application of NRD-based components integrated with planar circuits. In this way, both ground planes of the NRD guide can be interfaced with and accommodate the planar circuits, leading a very compact integrated topology with a minimized parasitic interference between the two planar circuits blocks.

III. Modeling Technique Using the Equivalence Principle

The coupling problem and design of transition between two dissimilar guiding structures through an aperture have been studied by a number of modeling techniques [7], [11]–[13]. The used aperture may be in the form of a circular hole or a narrow rectangular slot depending on the guiding structures to be considered. Two basic problems in the transition design are geometrical conformity and impedance mismatch. The geometrical conformity in this case requires the use of a rectangular aperture considering an effective magnetic coupling scheme.

In the following, the small aperture coupling theory is used to deal with modeling of the proposed transition. The basic principle of the small aperture theory is to make an equivalence of the aperture with an electric dipole moment, which can be easily obtained by the standard technique. The electric dipole moment in question is proportional to the electrical field normal to the aperture plane, while its magnetic counterpart is related to the magnetic field tangential to the aperture plane by a characteristic matrix. To facilitate the modeling procedure, let us define the signal path running from the microstrip line (primary guide) to the NRD guide (coupling guide). In this transition, the microstrip line is symmetrically located at the center of the coupling aperture. The original coupling geometry consists of a complete four-port network of which two ports will subsequently be open-circuited. Fields inside both guides can be expressed as those radiated by the dipole moments excited by a source. The S-parameters of the network can be eventually extracted from the modal voltages of both the primary guide and coupling guide. Assuming that the network is excited at port 1 by the incident wave of $j1$ mode (j refers to the mode), the reflected waves of modal voltage at all the four ports become V_r^n with mode $r = i1$ on the primary guide and $r = i2$ on the coupling guide. The scattering parameters are then calculated from the modal voltages of both the primary guide and coupling guide [10], such that

$$\begin{aligned} S_{ij}^{11} &= \frac{V_{i1}^1}{V_{j1}^1} \\ &= \frac{j\omega}{2}\left[\frac{\mu}{\sqrt{Z_{i1}Z_{j1}}}([\hat{M}]\cdot h_{j1})\cdot h_{i1}^* \right. \\ &\quad \left. -\sqrt{Z_{i1}Z_{j1}}\,\varepsilon([\hat{P}]\cdot e_{j1})\cdot e_{i1}^*\right] \end{aligned} \tag{1a}$$

$$\begin{aligned} S_{ij}^{21} &= \frac{V_{i1}^2}{V_{j1}^1} \\ &= \delta_{ij} - \frac{j\omega}{2}\left[\frac{\mu}{\sqrt{Z_{i1}Z_{j1}}}([\hat{M}]\cdot h_{j1})\cdot h_{i1}^* \right. \\ &\quad \left. +\sqrt{Z_{i1}Z_{j1}}\,\varepsilon([\hat{P}]\cdot e_{j1})\cdot e_{i1}^*\right] \end{aligned} \tag{1b}$$

$$\begin{aligned} S_{ij}^{31} &= \frac{V_{i2}^3}{V_{j1}^1} \\ &= -\frac{j\omega}{2}\left[\frac{\mu}{\sqrt{Z_{i2}Z_{j1}}}([\hat{M}]\cdot h_{j1})\cdot h_{i2}^* \right. \\ &\quad \left. -\sqrt{Z_{i2}Z_{j1}}\,\varepsilon([\hat{P}]\cdot e_{j1})\cdot e_{i2}^*\right] \end{aligned} \tag{1c}$$

$$\begin{aligned} S_{ij}^{41} &= \frac{V_{i2}^4}{V_{j1}^1} \\ &= \frac{j\omega}{2}\left[\frac{\mu}{\sqrt{Z_{i2}Z_{j1}}}([\hat{M}]\cdot h_{j1})\cdot h_{i2}^* \right. \\ &\quad \left. +\sqrt{Z_{i2}Z_{j1}}\,\varepsilon([\hat{P}]\cdot e_{j1})\cdot e_{i2}^*\right] \end{aligned} \tag{1d}$$

in which $[\hat{M}]$ and $[\hat{P}]$ are the dyadic magnetic and electric polarizabilities of the aperture, respectively. h and e are the modal vectors satisfying the normalization condition. Z is the modal impedance of the designated guide. If $\delta_{ij} = 1$ the mode $i1$ in the primary guide becomes the same as $j1$, $\delta_{ij} = 0$, otherwise. The remaining S-parameters can be written in the similar way.

The electric and magnetic fields of both guides are formulated in terms of the modal field vector e, h, as well as the modal voltage V and current I, respectively, that is

$$\begin{aligned} E_\gamma &= V_\gamma e_\gamma; \\ H_\gamma &= I_\gamma h_\gamma \\ &= \frac{V_\gamma}{Z} h_\gamma \quad \gamma = 1, 2. \end{aligned} \tag{2}$$

It is known that the LSM_{11} modal of the NRD guide cannot be expressed in a simple form of the modal vector because of the composite dielectric-air geometry. To obtain a concise analytical formulation, an equivalent waveguide concept of NRD guide is introduced. This concept is based on the field profile of the LSM_{11}, which is similar to the TE_{01} mode bounded by a rectangular metallic enclosure. This equivalence is illustrated in Fig. 2. To do so, let us carefully look at the eigenvalue equation [2]

$$\begin{aligned} \beta_{mn} &= \sqrt{\varepsilon_r k_0^2 - \left(\frac{m\pi}{a}\right)^2 - q_n^2} \\ &= \sqrt{k_0^2 - \left(\frac{m\pi}{a}\right)^2 - p_n^2} \end{aligned} \tag{3}$$

in which p_n or q_n are the eigenvalues of the designated LSM_{mn} modes. They are obtained by solving a set of characteristic equations

$$q_n^2 + p_n^2 = (\varepsilon_r - 1)k_0^2$$
$$\begin{cases} q_n \cot \dfrac{q_n b}{2} = \varepsilon_r p_n & n = 0, 2, 4, \cdots \\ q_n \tan \dfrac{q_n b}{2} = -\varepsilon_r p_n & n = 1, 3, 5, \cdots. \end{cases} \tag{4}$$

In this way, the effective dielectric constant for a LSM mode can be calculated from

$$\begin{aligned} \beta_{mn} &= \sqrt{\varepsilon_{re} k_0^2 - \left(\frac{m\pi}{a}\right)^2} \quad m = 1, 2, \cdots \\ \varepsilon_{re} &= \varepsilon_r - \frac{q_n^2}{k_0^2} \\ &= 1 + \frac{p_n^2}{k_0^2} \quad n = 0, 1, 2, \cdots. \end{aligned} \tag{5}$$

This procedure of analysis suggests that the normalized modal vector of the LSM_{11} mode can be obtained in a similar way as the dominant TE_{10} mode of the rectangular waveguide with an appropriately efficient dielectric constant. To begin with, the modal functions for the NRD guide can be written in the same way. This is done by

$$\begin{aligned} e_1 &= \sqrt{\frac{2}{ab}}\, \hat{a}_x \sin \frac{\pi y}{a}\, e^{-j\beta_{11} z} \\ h_1 &= \sqrt{\frac{2}{ab}} \left[\hat{a}_y \sin \frac{\pi y}{a} + j\frac{\pi}{\beta_{11} a}\, \hat{a}_z \cos \frac{\pi y}{a}\right] e^{-j\beta_{11} z} \end{aligned} \tag{6}$$

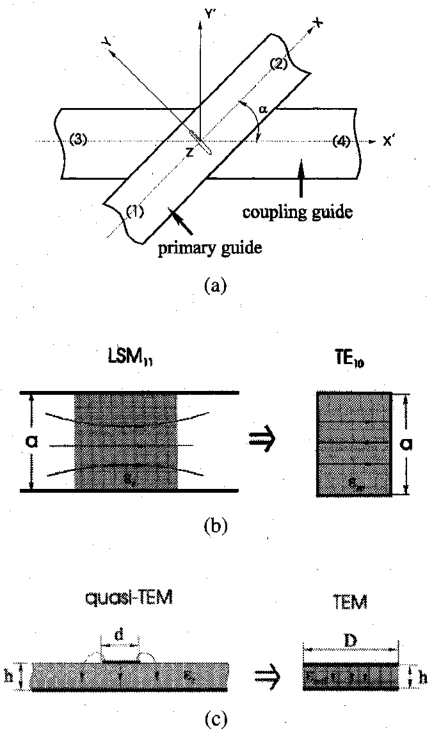

Fig. 2. Illustrations of (a) the coupling model of two waveguides passing across at an angle of α, (b) equivalent of NRD guide of LSM_{11} mode to rectangular metallic waveguide of TE_{01} mode, and (c) microstrip line replaced by its equivalent parallel plate guide.

in which β_{11} refers to the propagation constant of the LSM_{11} mode.

The normalized modal vectors for the microstrip line can also be obtained from its equivalence of a parallel plate waveguide model, which is well documented in [12], such that

$$\begin{aligned} e_2 &= \hat{a}_y' \sqrt{\frac{1}{Dh}}\, e^{-j\beta z'} \\ h_2 &= -\hat{a}_x' \sqrt{\frac{1}{Dh}}\, e^{-j\beta z'} \end{aligned} \tag{7}$$

in which D is the effective width of the parallel plate. Note that the length of the aperture slot does not affect the equivalent waveguide model since the coupling between the microstrip and the NRD guide is determined separately by the small aperture coupling theory. The modeling error comes basically from the intrinsic accuracy of the coupling theory. Generally speaking, smaller geometry of the coupling aperture yields better modeling accuracy.

Given that the electrical characteristics of the four-port network in question [letting α in Fig. 2(a) equal 90°] can be obtained from (1) and its subsequent equations, the S-parameter of the proposed transition that is concerned only with the two-port topology can be easily determined by letting two remaining ports remain open-circuited. Theoretical results will be presented and compared with experiments in the following sections.

IV. Electrical Performance of the Proposed Transition

With the above-described modeling technique, the electrical performance of the proposed transition can be easily obtained

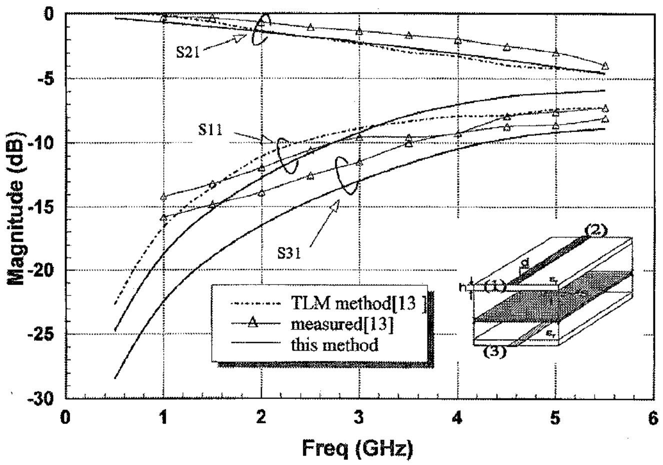

Fig. 3. A comparison for S-parameter of microstrip line-to-microstrip line transition via an aperture, in which $h = 0.254$ mm, $d = 0.246$ mm, and $\varepsilon_r = 2.2$, with the slot dimension of $l = 2.7$ mm and $s = 0.033$ mm.

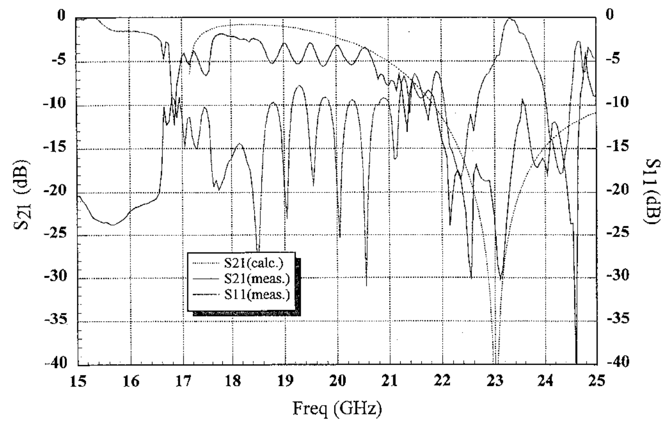

Fig. 4. Measured performance of two back-to-back transitions between microstrip lines, separated by 82 mm of NRD guide.

and optimized. Since there are the approximations made in the theoretical analysis, it is imperative to examine the numerical accuracy of the models of equivalence. Fig. 3 shows a comparison for a microstrip line-to-microstrip line coupling via an aperture [letting the angle α in Fig. 2(a) equal $0°$] in which the NRD guide is not involved. The results for the four-port S-parameters obtained from this technique agree relatively well with the experiments and those of the TLM method [13]. In addition, the equivalence principle implemented in the formulation is validated. Judging from its concise and analytical formulation, this technique is very efficient in terms of calculation time and memory storage.

In order to verify the applicability of the modeling technique for the proposed transition of microstrip line to NRD guide, a pair of transitions were fabricated and interconnected with a NRD guide having a length of 82 mm. The NRD guide was made of a rectangular dielectric strip (Rogers TMM®-3, $\varepsilon_r = 3.27$) and designed to operate around 20 GHz with $a = 6.1$ mm and $b = 6.5$ mm. The microstrip line was fabricated on a 60×98 mm^2 substrate (RT/duroid 5880, $\varepsilon_r = 2.3$) with a thickness of 20 mil and designed to have a line impedance of 50 Ω with a strip width of 1.53 mm. The coupling aperture on the ground plane is a narrow rectangular slot with 7.5×0.5 mm^2. The transmission and reflection coefficients of the complete signal path are plotted in Fig. 4 against a frequency that involves the two identical transitions, the input and output microstrip lines with the identical length, and the NRD guide. It is found that the theoretical and experimental results for S_{21} have a very good agreement, considering the fact that the modeling results are obtained only for the single transition without taking into account the losses of the two microstrip lines and NRD guide. The reflection S_{11} is generally better than -10 dB over the frequency band of interest. Some ripples in the frequency response were observed that may be attributed to our calibration problem. In this calculation, the interference between the two transitions is a complicated issue taht may degrade the electrical performance. It is simply ignored in the modeling. The best insertion loss is observed with less than 1.3 dB over 10% of the bandwidth and less than 4 dB over 40% at the center frequency of 20 GHz. It is observed experimentally that radiation of the open ended microstrip lines and coupling slots may contribute to the losses, although they cannot be separated in terms of the dielectric (negligible), ohmic, leakage, and radiation effects.

The modeling results in Fig. 5 indicate that the position of the open-ended microstrip line with respect to the aperture has a significant influence on the effective frequency bandwidth of the transition. Note that the open-end effect is ignored in the analysis and that the microstrip line is ideally supposed to be open-circuited at the specific location, as shown in the figure. The designated transition presents a cut-off frequency of 23 GHz with a complete stopband of $S_{21} = -30$ dB limited within 1.5% regardless of the variation of position of the open-ended microstrip line. Obviously, a better bandpass performance centered at different frequency is observed for the different length with the varied effective bandwidth. Based on this figure, our experimental prototype was fabricated and measured with $l_s = 1.4$ mm for this example. Nevertheless, such a design is believed to be far away from the complete optimization, which has to deal with a large number of tunable structural parameters. This issue, which is out of the scope of this paper, will not be discussed here. Contrary to the results of Fig. 5, the position of the open-end NRD affects strongly the location of the stopband of transition while the bandpass behavior is stationary at 18.7 GHz, as shown in Fig. 6. One important feature of changing the position of the NRD open-end is that the effective bandwidth of transition can be significantly improved by scarifying the better performance appearing at lower frequencies. Fig. 6 indicates that a flat transition bandwidth of 2 GHz with $l_w = 5.0$ mm can be readily obtained for a less than 2-dB insertion loss. Once again, these results suggest that an optimization procedure can be made in the design of the microstrip line-to-NRD transition.

Now, let us look at the electrical performance and paramedic effect of the proposed transition as long as the coupling aperture is concerned. Before getting into this analysis, one should be noticed that the use of the small aperture theory requires certain geometrical and electrical conditions imposed by the equivalent principle. These conditions have been documented in [10] and are not subject to the present discussion. It can be

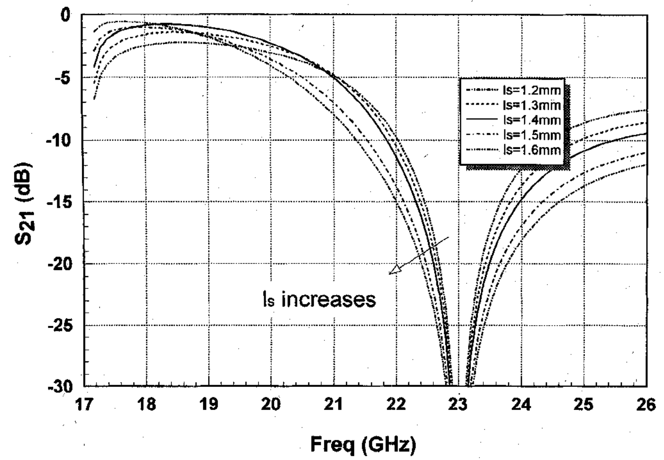

Fig. 5. Variation of the microstrip line open-ended position l_s with the position of an open-ended NRD fixed at $l_w = 5.6$ mm.

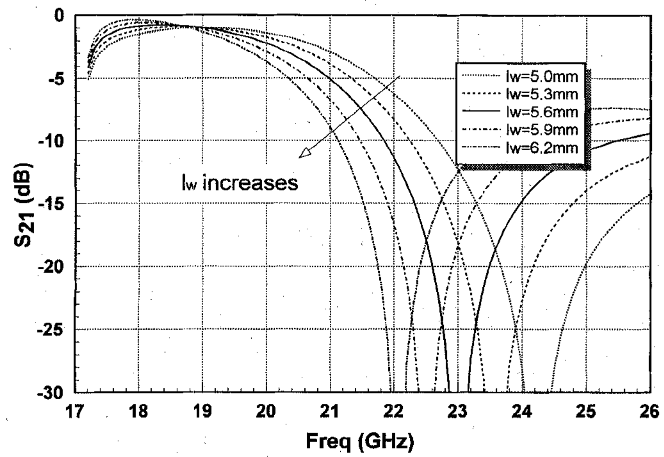

Fig. 6. Variation of the NRD open-ended position l_w with a microstrip line open-ended position fixed at $l_s = 1.4$ mm.

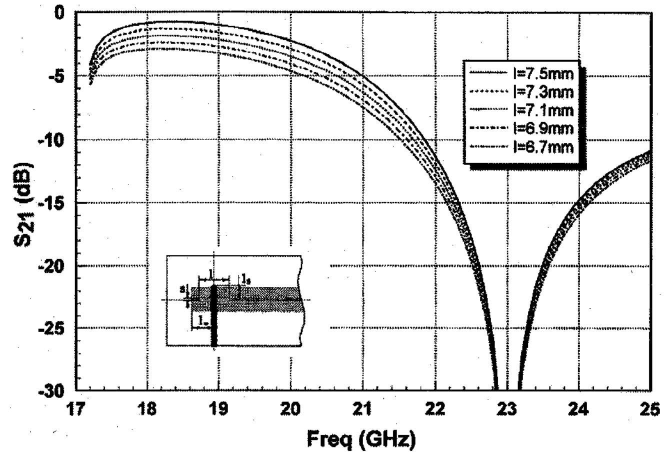

Fig. 7. Variation of the length l of an aperture having the width of $s = 0.5$ mm.

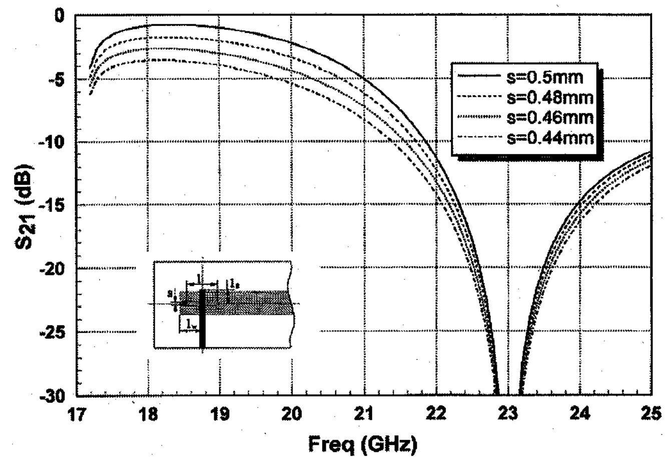

Fig. 8. Variation of the width s of an aperture having the length of $l = 7.5$ mm.

seen that the influence of the aperture length on the bandwidth performance for a fixed is described in Fig. 7, indicating that the better characteristics can be achieved by choosing larger length of the aperture. This implies that a lengthy aperture may enhance the magnetic coupling to some degree. Fig. 8 plots the frequency-responses of the transmission coefficient for different width of the aperture. As expected, a large width of the aperture yields better results. In any case, it is interesting to note that the central frequency of the stopband remains unchanged. This further suggests that the geometry of the aperture have virtually no influence on the stopband characteristics.

V. Conclusion

In this paper, a compact transition has been proposed to integrate the microstrip line with the NRD guides at microwave and millimeter-wave frequencies. The equivalence principle of the small aperture theory has been successfully applied in the modeling and analysis of the coupling aperture with the framework of the moment method. In addition, the equivalent waveguide techniques have also been developed and used to model the microstrip line and the NRD guide operating with the low-loss fundamental nonradiative mode. Therefore, an analytical method of microstrip and NRD guide has been formulated that yields an efficient algorithm. Theoretical and experimental results demonstrate a good agreement, thereby validating the proposed analytical approach. It is found that the open-ended position of both microstrip line and NRD guide with respect to the aperture has a significant but different impact on the bandwidth performance, while the geometrical dimensions of the aperture has no influence on the stopband characteristics. The results indicate that an optimization procedure can be made to achieve a wideband microstrip line-to-NRD transition by choosing its appropriate geometry for a particular frequency band. This study confirms that this type of transition is potentially suitable for low-cost applications of the hybrid circuits.

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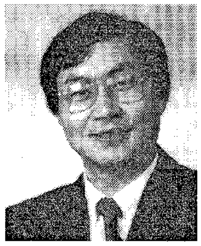

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