Microstrip-Fed Planar Frequency-Multiplying Space Combiner

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***Abstract*—A frequency-multiplying power combining array has been made of slots in the ground plane on a substrate with a microstrip feedline on the back side. The second harmonic generated by a diode in each slot is combined in free space. A design procedure and experimental results are presented.**

I. Introduction

GENERATION OF a coherent millimeter-wave signal with a reasonable strength becomes more difficult as the frequency of operation is increased. One of the methods to alleviate this problem is the use of a power combining technique [1]. However, it is increasingly difficult to combine the power outputs from more than a few devices. In many power combining schemes, very involved design procedures, complicated fabrication, and tricky postfabrication tuning are required. An alternative approach is the use of frequency multiplication [2]. In this approach, the output power level is substantially lower than the input signal level. In addition, the power level that can be handled by a diode-type frequency multiplier is quite limited. To alleviate many of the deficiencies described above, a quasi-optical frequency-multiplying space power combiner has been proposed and tested [3]. The proposed structure contains in a single component (1) a slot antenna array, (2) frequency multipliers, and (3) a power combiner. The structure has a simple configuration and requires no bias and no postproduction tuning.

In this paper, a new printed circuit configuration of the frequency-multiplying space power combiner is presented. In contrast to the work in [3], no waveguide is used for feeding. In the new structure, the array of slots on the substrate is fed by a microstrip line or a coplanar waveguide fabricated on the back side of the substrate. The new structure has other desirable features, such as (1) a planar structure suitable for monolithic integration, (2) large coupling, (3) controllable coupling, and (4) flexibility in the arrangement of slots on a planar circuit.

Fig. 1 shows a typical configuration of the planar multiplier/combiner. On the front side of the substrate, there is an array of slots and in each of these a diode is placed for frequency multiplication. A microstrip line feed on the back side is used in this example. The incident signal at the frequency f_0 is successively fed to these slots, which are typically a quarter wavelength long. The diode in each slot generates a second harmonic of $2f_0$. Then the slot behaves as a half wave slot dipole and an efficient radiation of a $2f_0$ signal takes place. The radiated powers from all the slots are combined in free space. This paper reports an experimental design of the structures scaled for a low-frequency operation at the X-band. This scaling allows us to use more precise characterizations than those attainable at millimeter-wave frequencies, so that useful design data can be collected more efficiently.

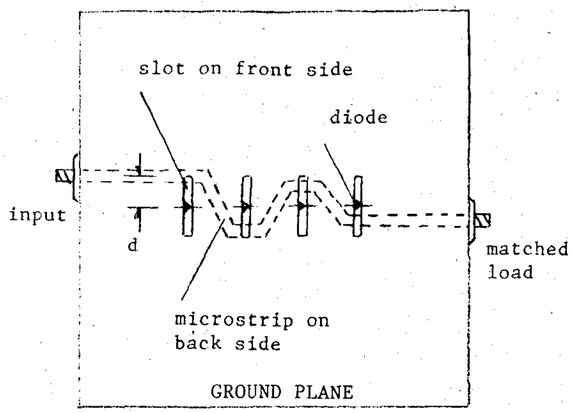

Fig. 1. Structure of multiplying slot array fed by microstrip.

II. Design Procedure

In order to design the proposed structure, it is necessary to know the characteristics of the microstrip-to-slot transition, the phase relationship between slots, and the conversion gain of each slot. Let us first study these characteristics in detail.

A. Magnitude Characteristics of Microstrip-to-Slot Transition

The transition characteristics have been studied based on the approximate equivalent circuit shown in Fig. 2 [4]. The magnitude characteristics of the coupling from the microstrip to the slot, the insertion loss of the slot, and the return loss of the slot with a 50-Ω load located at the

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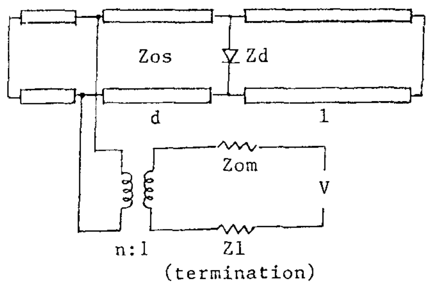

Fig. 2. Equivalent circuit for one slot ($l = \lambda_0 8$).

center of the slot are given as follows:

$$\text{Coupling} = 20\log|2nZ(\cos\beta d + \tan\beta d \sin\beta d) / ((Z_{0m} + Z_L + Z)(1 + jZ_{0s}\tan\beta d / Z_3))| \quad (1)$$

$$\text{Insertion loss} = 20\log|2Z_L(Z_{0m} + Z_L + Z)| \quad (2)$$

$$\text{Return loss} = 20\log|(Z_{0m} - Z - Z_L)/(Z_{0m} + Z + Z_L)| \quad (3)$$

where

- Z slot impedance seen from the microstrip,
- Z_{0m} characteristic impedance of the microstrip,
- Z_{0s} characteristic impedance of the slot line,
- Z_L termination impedance (50 Ω),
- Z_3 slot impedance seen from the center of the slot.

According to the calculated results, the smaller the distance from the center of the slot to the feeding position, the larger the coupling becomes. In order to examine the possibility of controlling the power coupled from the microstrip to the slot, we measured the transition characteristics. Measurement of microstrip-to-slot transition has been carried out with the conventional coaxial-to-slotline transition [5] as a probe for picking up the signal over the slot in place of the diode. From the measurement results, it is found that the control of the power coupled from the microstrip to the slot is possible by controlling the distance from the center of the slot to the feeding position. Also, it is seen that the equivalent circuit for the transition gives a prediction that agrees well with the measured data, if $n = 1$ is chosen, as long as the feeding position is not too close to the end of the slot. The results are shown in Fig. 3. The dotted lines show the calculated results under the assumption of $n = 1$ in Fig. 2.

B. Phase Relationship between Slots

The phase relationship between slots is determined by (1) the electrical length of the microstrip feed line between slots, (2) the phase characteristics of the transition, and (3) the square-law characteristics of the diode. Among these factors, the main contribution is the electrical length of the microstrip feedline between the slots. This quantity can be controlled by the physical length of the microstrip line between the slots.

The phase characteristics of the transition can be evaluated from the phase change of the voltage from the

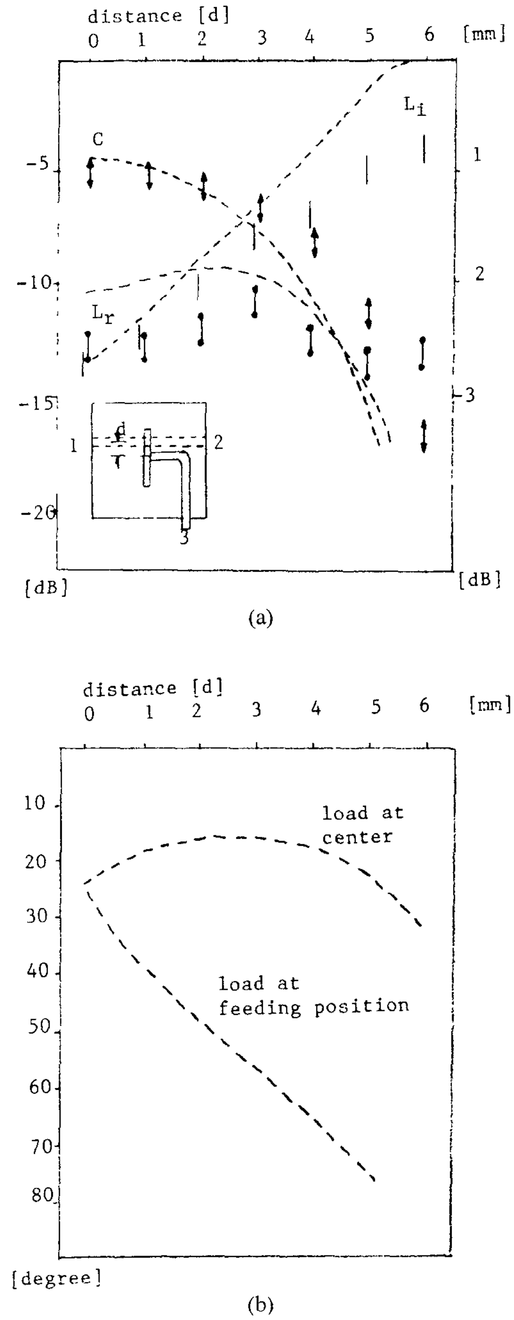

Fig. 3 Transition characteristics from microstrip to slot. (a) Coupling (C), insertion loss (L_i), and return loss (L_r). (b) Phase change of transition with load position.

transition point of the microstrip to the 50-Ω load on the slot by the following equation:

$$\text{Phase change} = \arg\left[2nZ(\cos\beta d + \tan\beta d \sin\beta d) / ((Z_{0m} + Z_L + Z)(1 + jZ_{0s}\tan\beta d / Z_3))\right]. \quad (4)$$

The results are plotted in Fig. 3(b) as a function of the position of the load. From these results, we find that the difference of the phase change to the feeding position can be neglected if the load is located at the center of the slot.

A diode located across the slot can be thought to pick up the voltage of the input frequency signal. The voltage on the slot forms a standing wave pattern given by

$$V(x) = \sin\beta(d + l_h - x)/\sin\beta(d + l_h). \tag{5}$$

Since this voltage has a maximum value at the feeding position, it is better to locate the diode at the feeding position for a better matching and a stronger coupling. As shown in Fig. 3(b), the phase change of voltage versus feeding position of each transition cannot be neglected in this case. Therefore, it is required that the difference of the phase change at each transition be considered for determination of the length of the microstrip line between the slots.

In the case of a two-element slot array, the length of the microstrip feed line between the slots is $\lambda_g/2$, where λ_g is the guide wavelength at the input frequency. However, the diodes should be placed with the same polarity at each slot for the broadside operation. This can be explained by the square-law behavior of the diode [3].

C. Conversion Gain of Each Slot

The conversion gain of a multiplier is defined as the ratio of the output power to the input power. In the present case, the input power is the amount of power coupled to a single slot from the microstrip at the fundamental frequency. The output power is the amount of the power at the second harmonic frequency, which is radiated by this slot into free space. Thus, we have to evaluate the amounts of input power and radiated power to find the conversion gain. The impedance of the mixer diode used in this paper is a function of the drive power. However, if the diode is driven sufficiently (more than 10 dBm), the impedance is known to be about 50 Ω. Thus, the power coupled into the diode at the fundamental frequency can be estimated by using C in Fig. 3(a). The direct measurement of the amount of the radiated output power is very difficult. Hence, we use an indirect method based on the fact that the power received at a fixed receiving antenna is proportional to the output power of a transmitting antenna. This relation is given by [6]

$$P_r = P_t + G_t + K \tag{6}$$

where

P_r receive power (dBm),
P_t transmit power (dBm),
G_t transmit antenna gain (dB),
K constant determined by the gain of receiving antenna, distance between the transmitting antenna and the receiving antenna, and the test frequency.

For the DUT (device under test), the transmit power can be expressed by

$$P_{t,d} = P_{i,d} + C_d + G_c \tag{7}$$

$P_{t,d}$ transmit power (dBm) from DUT,
$P_{i,d}$ incident power (dBm) to the microstrip feed line in DUT,
C_d power coupling factor (dB) to diode in DUT,
G_c conversion gain (dB) of DUT.

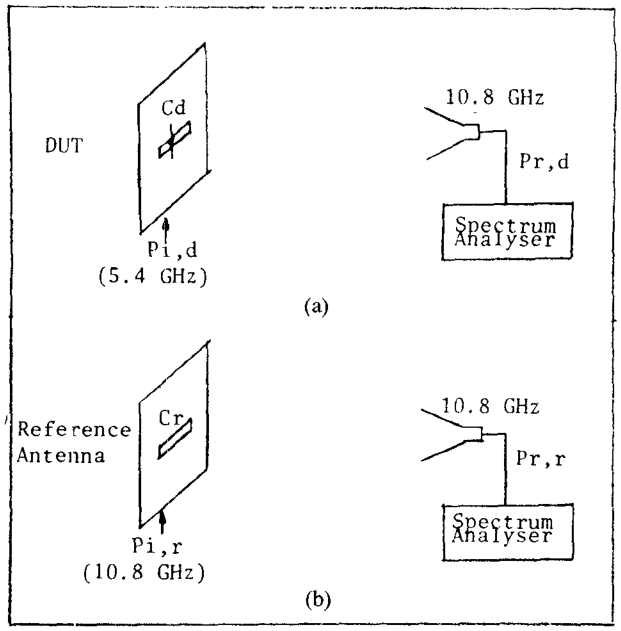

Fig. 4. Test setup for measuring the multiplication conversion gain. (a) DUT setup. (b) Reference setup.

Using (7), the power relation applied to the DUT setup shown in Fig. 4(a) is given by

$$P_{r,d} = (P_{i,d} + C_d + G_c) + G_{t,d} + K \tag{8}$$

where

$P_{r,d}$ receive power (dBm) in DUT setup,
$G_{t,d}$ antenna gain (dB) of DUT.

For the reference antenna, the transmit power can be expressed by

$$P_{t,r} = P_{i,r} + C_r \tag{9}$$

where

$P_{t,r}$ transmit power (dBm) of reference antenna,
$P_{i,r}$ incident power (dBm) to the microstrip feed line in reference antenna,
C_r power coupling factor (dB) to reference antenna.

Using (9), the power relation applied to reference setup shown in Fig. 4(b) is given by

$$P_{r,r} = (P_{i,r} + C_r) + G_{t,r} + K \tag{10}$$

where

$P_{r,r}$ receive power (dBm) in reference setup,
$G_{t,r}$ gain (dB) of reference antenna.

Using (8) and (10), the conversion gain can be calculated from

$$G_c = P_{r,d} - P_{r,r} - P_{i,d} + P_{i,r} - C_d + C_r - G_{t,d} + G_{t,r}. \tag{11}$$

D. Design Procedure

Let us consider the design procedure of the multiplying slot array. Coupling is defined by

$$\text{Coupling} = P_{\text{diode}} - P_{\text{input}} \quad [\text{dB}]. \tag{12}$$

For simplicity, we locate all the diodes at the center of each slot and design an equally excited, broadside multiplying slot array by applying (12) to the measured results in Fig. 3. First, from the power requirement of the diode in the first slot, we calculate the necessary coupling by (12); then we evaluate the location of the microstrip-to-slot transition from C in Fig. 3(a). We can then find out the amount of power left over on the microstrip line from L_t in Fig. 3(a). Next, we find the location of the transition of the second slot so that the power provided to the second diode is equal to that for the first diode. Then, we determine the length of the microstrip to the second slot to obtain an appropriate phase shift between slots. This process is repeated until all the slots are taken care of. Notice from Fig. 1 that the location of the transition moves inward as the feed signal progresses along the microstrip line. The decreasing power available from the feed line is compensated for by an increasing degree of coupling so that all the diodes are given an equal excitation power. It is possible to design various multiplying slot array antennas with a suitable arrangement of multiplying slots in the ground plane with the proper excitation powers and phase differences [6]. These design quantities can be obtained by controlling the locations of the transitions and the lengths of the microstrip between the slots.

III. Experimental Results

For frequency multiplication, nonlinear reactive elements such as varactor diodes are desirable for achieving good performance. However, only the Schottky-barrier diodes designed for X-band mixers have been available to the authors. The conversion gain, G_c, measured by the method described above for the structure with the ND5051(NEC) diode has been -6.2 dB. In this experiment, the power coupling factor to the reference antenna, C_r, can be estimated by

$$C_r = 10\log\left[(P_{in} - P_{ref} - P_{tra})/P_{in}\right] \tag{13}$$

where

- P_{in} incident power (mW) to microstrip feed line in reference antenna,
- P_{ref} reflected power (mW) at input port in reference antenna,
- P_{tra} transferred power (mW) to the matching load in reference antenna.

An estimated power coupling factor of -7 dB results. Note that the only difference in antenna structure between the DUT and the reference antenna is that there is a diode in the DUT while there is no diode in the reference antenna. This fact causes a difference in the gain of the two antennas. The difference of the two antenna gains can be estimated by measuring the reduction of the receiving power in the reference setup shown in Fig. 4(b) when a diode is located at the center of the slot of the reference antenna. About 10 dB reduction in the antenna gain due to the loading effect of the diode is measured.

For the other higher order harmonics generated by the diode, the slot can also be thought of as a slot dipole antenna. However it is observed that the receiving power of the third harmonic signal is lower than that of the second harmonic signal by more than 10 dB.

In order to design the multiplying slot array antenna, it is necessary to know the characteristics of the element radiation pattern of the single multiplying slot antenna. First, the effect of the ground plane size is evaluated. Fig. 5(a) shows the radiation patterns of a single slot antenna with a ground plane of different sizes. It is found that the length of one side of the ground plane should be larger than seven wavelengths at the radiating frequency (about 20 cm at 10.8 GHz). Otherwise, the element radiation pattern of the slot antenna has several peaks and valleys. This effect of the finite ground plane on the radiation by a slot can be explained theoretically by the uniform theory of diffraction [6]. Next, the effect of the location of the transition is tested. Fig. 5(b) shows the radiation patterns of a single multiplying slot antenna with the transition at different locations. We notice that the position of the feeding point has a negligible influence on the radiation pattern. This is due to the fact that for the second harmonic signal, the diode located at the center of the slot can be considered a source. Even though its output power is dependent on the feeding position, its position on the slot is not changed with the feeding position. Therefore, the radiation power of the multiplying slot antenna is changed with the feeding position but the radiation pattern for the front space of the slot seems to be insensitive to the position of the narrow microstrip feed line on the back side.

For the loading effect of the diode on the radiation pattern, the radiation patterns of the single slot antenna with the diode and without the diode are measured. By comparing Fig. 5(b) with Fig. 5(a), we notice that the diode located at the center of the slot reduces the number of undulations in the radiation pattern. However, the overall radiation pattern is not changed.

Based on the experimental transition characteristics and the elementary radiation pattern of the single slot, we designed and tested several uniformly excited broadside multiplying slot arrays. Figs. 5(c) and (d) show the radiation patterns of a two-element H-plane slot array antenna for different spacings between the slots. It is observed that the radiation pattern of an H-plane array with more closely spaced elements shifts more from the broadside direction. This is due to the increased internal mutual interference between the slots. The spacing between H-plane slots should be more than one wavelength at the second harmonic frequency to avoid interference. However, this requirement produces grating lobes in pairs for a broadside antenna. For the design of an array with closer spacing, it is necessary to measure the element radiation pattern including the effect of internal interference. Figs. 5(e) and (f) show the radiation patterns of a two-element E-plane slot array antenna with different degrees of coupling. It is observed that the radiation pattern of an array

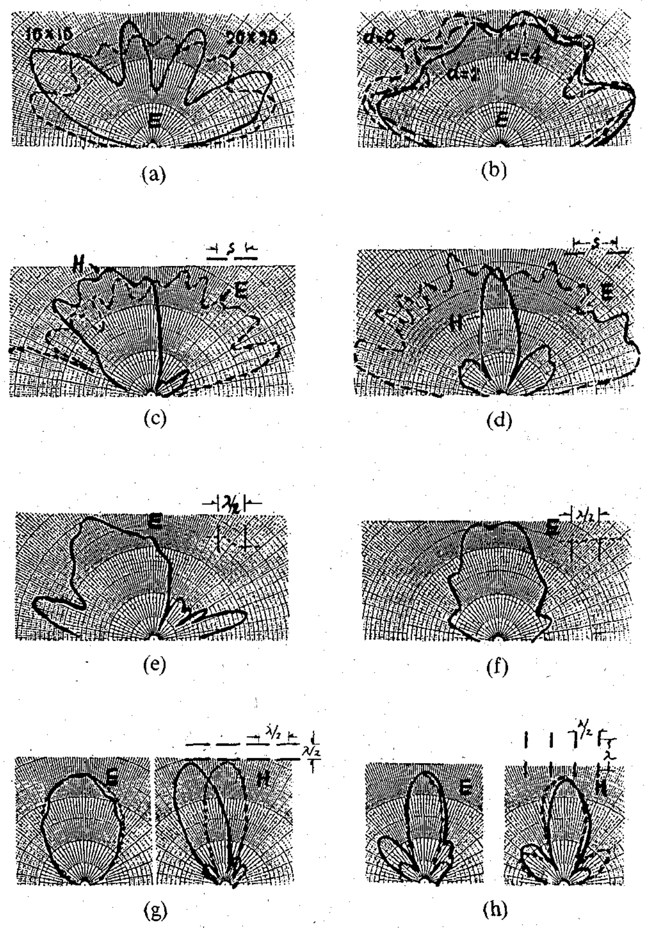

Fig. 5. Measured radiation patterns of various multiplying slot antennas. (The magnitude is plotted in linear scale.) (a) Radiation pattern of a single multiplying slot antenna for different ground plane sizes (without diode). (b) Radiation pattern of a single multiplying slot antenna for different feeding positions (with diode, 20×20). (c) Radiation pattern of a two-element H-plane slot array antenna (spacing $=\lambda/2$). (d) Radiation pattern of a two-element H-plane slot array antenna (spacing $=\lambda$). (e) Radiation pattern of a two-element E-plane slot array antenna for large coupling (-5 dB). (f) Radiation pattern of a two-element E-plane slot array antenna for small coupling (-10 dB). (g) Radiation pattern of a 2×4 H-plane slot array antenna. (h) Radiation pattern of a 2×4 E-plane slot array antenna. (The dotted lines indicate calculated radiation pattern.)

with more strongly excited slot elements shifts more from the broadside direction. This is because the reflected power at the transition is increased as the coupling increases and because the neglected phase shift at the microstrip-to-slot transition is increased as the difference of the feeding positions at the two slots increases.

Figs. 5(g) and (h) show the radiation patterns of the 2×4 H-plane and the 2×4 E-plane slot arrays with the calculated radiation patterns, respectively. The measured radiation patterns agree well with the theoretical results except for the H-plane radiation pattern of the 2×4 H-plane slot array. The reason for this is explained above. These arrays are excited with the incident power of 26 dBm at 5.4 GHz. The power fed to each diode is estimated to be 11 dBm. Fig. 6 shows a photograph of the 2×4 H-plane multiplying slot array antenna for which the radiation pattern is already plotted in Fig. 5(g).

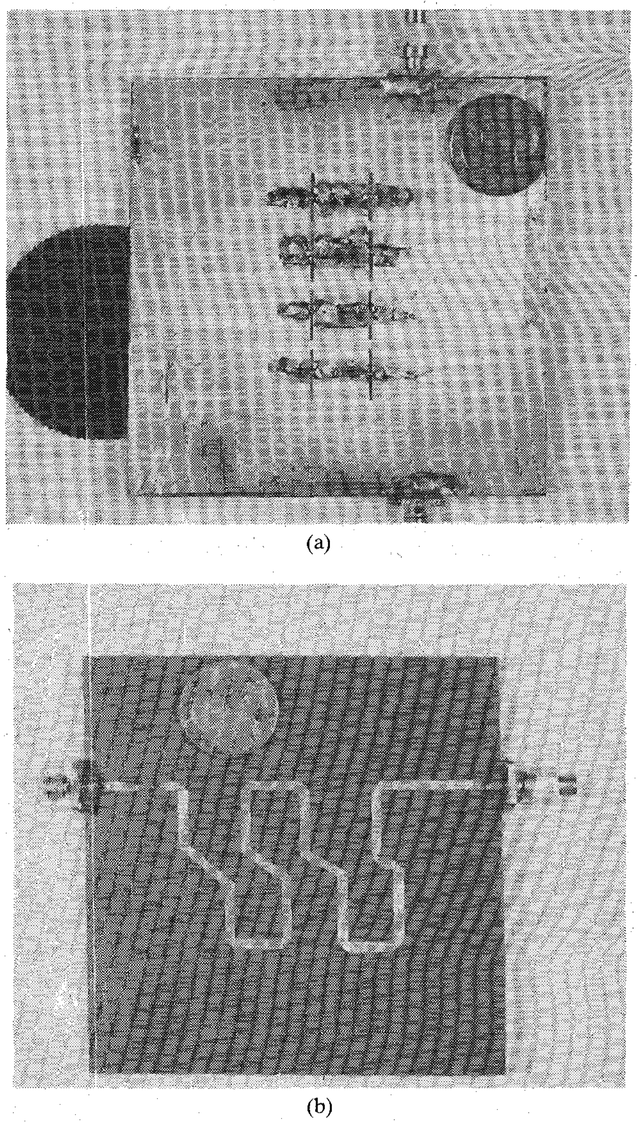

Fig. 6. Pictures of a 2×4 slot array antenna. (a) Slot array pattern. (b) Microstrip feed pattern.

IV. Conclusions

By using experimental data which characterize the transition from the microstrip to the slot, we designed a planar multiplier/combiner in the form of a slot array fed by a microstrip on the back side of the substrate. The results show the feasibility of the proposed structure. The new structure has desirable features, such as controllable power coupling and flexibility in array geometry on a planar substrate, as well as the features of a general multiplying slot array [3]. The present structure is suited for microwave and millimeter-wave integrated circuit applications.

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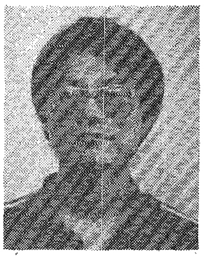

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