A Quasi-Optical Multiplying Slot Array

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Abstract —**A quasi-optical frequency-multiplying slot array for high-power millimeter-wave applications has been demonstrated. Single slot conversion loss measured for *X*- and *Ka*-band models produced results that are similar to in-waveguide multipliers using the same type of diodes. An eight-element multiplying array exhibited better than −12-dB sidelobe level at 70 GHz. The simple planar structure makes millimeter IC fabrication of the slot array feasible.**

I. Introduction

THE FREQUENCY-multiplying slot array presented here is a slot antenna, a frequency multiplier, and a space combiner integrated into one component. This structure provides an alternative to a conventional solid-state source or an in-waveguide frequency multiplier, particularly for quasi-optical applications. It is difficult to obtain high-power solid-state sources at millimeter-wave frequencies. A conventional multiplier works under the limitation that the nonlinear element cannot handle more than a few hundred milliwatts. It is possible to power combine solid-state sources [1] or multipliers [2]. However, it is exceedingly difficult to combine more than several units due both to mechanical and electrical restrictions.

The proposed structure (Fig. 1) alleviates the difficulties described above. It has the following features: 1) only one source is required, 2) no splitting feed network is used, 3) it has a small size, 4) no post-fabrication tuning is necessary, 5) it uses built-in space power combining, and 6) it offers the possibility of planar integration.

Each slot in which a nonlinear element is installed is one-quarter free-space wavelength long at the fundamental input frequency and, hence, is resonant (one-half wavelength) at the second-harmonic frequency. These slots form a slot array for the second harmonic; thereby, the harmonic power is combined in free space. The slots are distributed along a single feeding transmission line in which the fundamental is guided. Since the slots are electrically small (one-quarter free-space wavelength long) at the fundamental frequency, a relatively small fraction of power couples to each slot. This is an important feature, unlike conventional multipliers in which the maximum fundamental power is coupled to the diodes. In the present structure, we would like to feed as many diodes as possible so that the total (combined) power is increased. From the antenna point of view, this makes the output beam narrower.

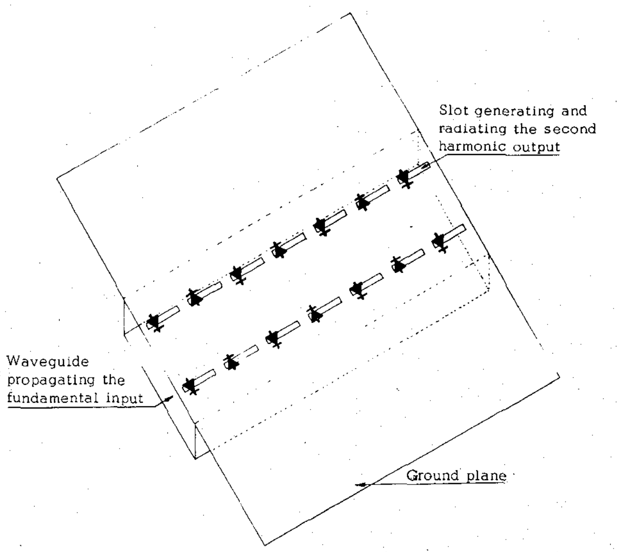

Fig. 1. A two-by-seven frequency-multiplying slot array fed by a waveguide.

II. Design

The fundamental input can be fed effectively to the slots in a variety of ways. Several types of transmission lines can be used for this purpose. A waveguide has been initially chosen as shown in Fig. 1 since it is best suited to demonstrate the operating principle and is one of the most popular transmission lines at millimeter-wave frequencies.

An important consideration in the design of frequency multipliers is that the output needs to be isolated from the input. The second harmonic generated at the diode is resonant in the slot and radiates but should not couple back to the transmission line feeding the fundamental. In the usual multiplier design, a low-pass filter [3] is used to prevent the second harmonic from leaking back into the transmission line feeding the fundamental frequency.

The theory of slots on a waveguide wall was derived by Stevenson [4] and later extended by several other authors [5], [6]. In the theoretical approach, the field distribution along a slot on a waveguide wall is given. The waveguide is then analyzed in three sections, namely, the section containing the slot, and the two waveguide sections that extend to infinity on each side of the slot. The coefficients of the modes propagating in both directions of the waveguide can then be solved by matching the boundary conditions.

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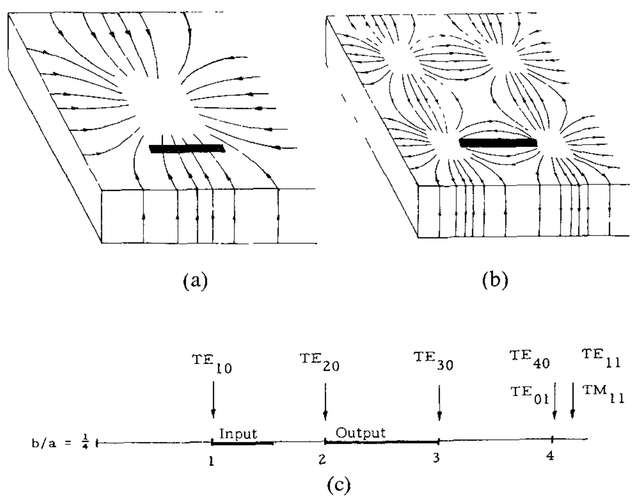

Fig. 2. (a) Surface currents on a waveguide wall for the TE_{10} mode. (b) Surface currents for the TE_{20} mode. The position of the slot is chosen not to couple to this mode. (c) Half-height waveguide cutoff for several modes with respect to the cutoff of the TE_{10} mode. Bold areas indicate feasible positions for the fundamental and the second harmonic.

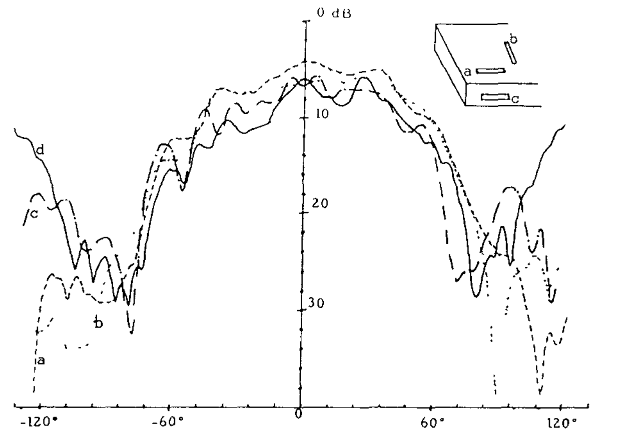

Fig 3. E-plane radiation pattern for a half-wavelength long slot at the second harmonic (8 GHz). In a, b, and c, the slot is backed by a half-height waveguide. In d, the slot radiates on both sides of the ground plane (no waveguide is involved).

From these solutions it is evident that a slot couples to a mode propagating in the waveguide if in any way it disturbs the current distribution on the waveguide wall for that particular mode. In other words, a slot couples to a mode if it intercepts the current distribution on the waveguide wall. This is a well-known technique that could give us information of whether or not a slot couples to the waveguide modes.

In this design, the waveguide feeding the slots has been chosen such that the least number of modes could propagate at the second harmonic while still allowing the fundamental mode to propagate. In this structure, the fundamental couples to the slot via the TE_{10} mode. The waveguide used is half the standard height of a conventional one such that at the second harmonic only the TE_{10} and the TE_{20} mode propagate. The positions of the slots are chosen such that they do not couple to the TE_{20} mode at the second harmonic as shown in Fig. 2(b), in which it is apparent that the slot does not disturb the current distribution for this mode. Fig. 2(c) shows the bandwidth of operation for such modes with respect to the cutoff of the TE_{10} mode for the fundamental input. The input frequency lies between 1 and 1.5 times the TE_{10} cutoff frequency. The second harmonic will be between two and three times the cutoff of the fundamental frequency, such that at the second-harmonic frequency only the TE_{10} and the TE_{20} modes can propagate. No coupling to the TE_{20} mode takes place due to the choice of slot location as described above. Some coupling of the second harmonic to the TE_{10} mode is inevitable. However, this coupling has been measured and was found to be 15 dB below the input power level at the second harmonic while the coupling from the fundamental mode to the slot is about 10 dB down. These values were measured by replacing the diode with a coaxial feed.

The choice of slot position for least coupling of the second harmonic back into the input waveguide has been demonstrated experimentally. Since measurement of power in an overmoded waveguide is difficult, this verification has been made using relative radiation intensities. Several slots and waveguide dimensions have been tried experimentally and the slot position of Fig. 2 proved to be the best. The distance of the slot from the waveguide wall is approximately 1/5 of the waveguide width. Fig. 3 shows radiation patterns, measured at 8 GHz, of three differently oriented half-wavelength slots in a ground plane that is backed by a half-height waveguide and fed by a coax at the second harmonic of the normal waveguide operating band. These radiation patterns are compared to the measured radiation patterns of a half-wavelength slot in a ground plane radiating on both sides of the ground plane. The ground-plane dimensions were the same in all cases. From the data it is evident that the radiation intensity for slot a is 3 dB higher than that of a slot radiating in both directions (Fig. 3(d)). This signifies that almost all the energy is radiated rather than coupled to the waveguide.

As mentioned in the Introduction, the power coupling from the waveguide to the slot at the fundamental frequency is relatively small so that many slots need to be excited in order to obtain high system efficiency. Matching of the nonlinear diode to the slot can be obtained by choosing an appropriate mounting location along the slot. Similar techniques have been used in a number of millimeter-wave slot antenna mixers [7], [8].

In an attempt to understand the variables involved in designing the waveguide slot-antenna multiplier, an equivalent circuit of one of the slots is shown in Fig. 4. The equivalent circuit consists of a transmission line which is coupled to the diode circuit by means of a directional coupler. The directional coupler is present because only a portion of the power in the waveguide couples to the slot. The output arm of the coupler is fed into a filter which passes the fundamental and rejects the second harmonic. The filter is required in the equivalent circuit since the position of the slot allows coupling to the dominant mode of the waveguide while preventing coupling of the second harmonic generated by the diode back into the waveguide. This filter is connected to the diode via a matching circuit at the input and the output is extracted from the diode via

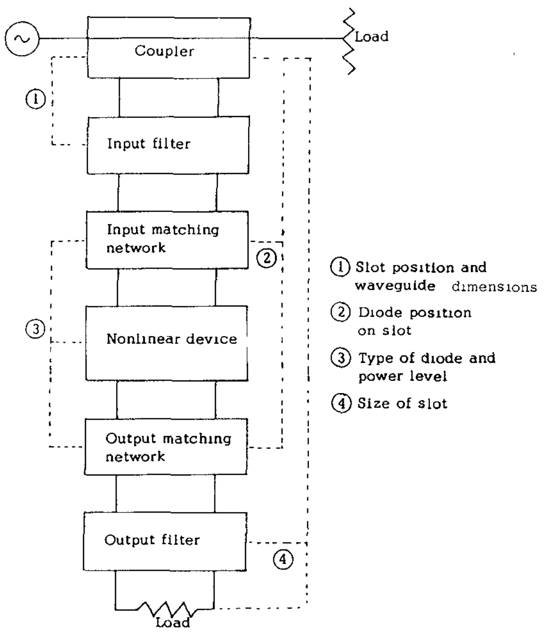

Fig. 4. Piecewise equivalent circuit of a slot multiplier indicating the four major variables effecting the performance of the multiplier.

a second matching circuit. These matching circuits represent the impedances obtainable by changing varactors or by moving the diode along the slot. Finally, the diode feeds into an output filter and a load resistor. The output filter passes the second harmonic and represents the slot resonance at the second harmonic where the slot is a half-wavelength long. The load impedance is the radiation impedance of the slot at the second harmonic and varies according to the slot dimensions.

From this equivalent circuit, one could then quantify the physical parameters that one could change to have an optimum system. More than one parameter is needed and each reflects one or more blocks of the equivalent circuit. Four major parameters and their effects are identified as follows.

1) The slot position and the waveguide dimensions will affect the coupling of the first harmonic from the waveguide to the slot. On the other hand, the coupling of the second harmonic back to the waveguide could be minimized by the appropriate choice of slot position and waveguide dimensions. A compromise needs to be reached at this point as to whether to minimize the amount of second-harmonic coupling back into the waveguide, or maximize the coupling of the fundamental waveguide mode to the slot. If one is interested in having high varactor efficiency, the filtering at the second harmonic is more important than the coupling of the fundamental mode to the diode. However, in a system design one needs to adjust the coupling so that the slot array efficiency will be high.

2) The position of the diode on the slot will affect both the input and the output match of the varactor. As the input match is altered, the degree of coupling will change since the coupling is a function of the impedance the slot presents to the waveguide at the fundamental frequency. The choice of diode position is one of the simplest tuning mechanisms for the slot multiplier.

3) The type of diode used is an important factor in the overall slot array performance. One needs to choose a type of diode whose properties match, as closely as possible, the other design restrictions. For instance, a diode with a high cutoff frequency is desirable, but diodes with high cutoff frequencies usually have a very small capacitance and this might cause matching problems for low-frequency circuits. So a diode with a lower cutoff and a high capacitance might be a better choice at low frequencies. The drive level of the varactor affects the fundamental and second-harmonic impedances of the diode and is also a major limitation for high multiplier efficiency. Since the varactors in this circuit are biased at zero volts, it is of great importance that the drive level of the varactor is chosen so as to improve the match to the remainder of the circuit.

4) The width of the slot helps to determine the level of rejection that the output filter will have at the input frequency and at the same time effects the radiation impedance presented to the varactor. Varying the length of the slot will also affect the coupling into and out of the waveguide since changes in the slot length alter the impedance at both the fundamental and at the second harmonic. One needs to consider that making the slot larger than a quarter-wavelength at the fundamental will increase the coupling of the fundamental to the slot but at the same time it will make the slot a better radiator at this frequency, thus reducing the performance of the output filter.

From the above discussion it should be apparent that the design of the slot multiplier is not straightforward and a significant amount of consideration and compromise is required. A particular characteristic of the slot multiplier design is that if the varactor is not well matched at the fundamental, the slot will present the equivalent of a short circuit to the waveguide. In such cases, the power is not reflected back to the generator but continues traveling down the waveguide past the remaining slots and finally being absorbed at the termination, just as if the slot did not exist. If the slot is kept to a quarter-wavelength at the fundamental, the coupling of the slot multiplier to the waveguide cannot be more than -10 dB. One could then assume that the coupling to each slot multiplier from the waveguide causes only a minor perturbation to the field in the guide. This implies that the TE_{10} mode for the fundamental is gradually attenuated as it propagates down the array and that no major reflections occur due to the slots.

III. Measurement of Conversion Loss for a Slot

The conversion loss in decibels or the conversion efficiency as a percentage of a multiplier is defined as the ratio of the output power to the input power. In the case of the slot doubler, the input power is the amount of power coupled to a single slot from the waveguide at the fundamental frequency, and the output power is the amount of the second harmonic which is radiated by this slot into free space.

Measuring the amount of fundamental power coupled to the slot doubler is not a trivial matter. Since the power coupled to the diode on the slot is at least 10 dB below the power propagating in the waveguide, it is difficult to

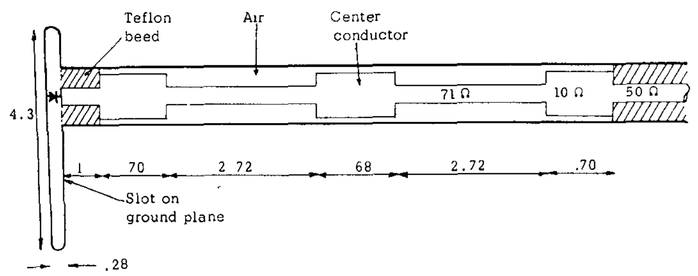

Fig. 5. Geometry of slot radiating at 34 GHz and coaxial bandpass filter at 20 GHz (all dimensions in millimeters).

evaluate the power coupled, from the measurement of the difference between the input and output power in the waveguide. This value of −10-dB maximum coupling has been determined using a coaxial feed across the waveguide slot and an infinitely variable impedance (stub tuner) attached to the coax and adjusted to obtain the maximum coupling. One cannot simply replace the diode by the coaxial feed and measure the coupled power because the impedance of the varactor diode is a function of the drive power. Thus, it is not practical to try and simulate the diode.

One way to evaluate the power coupled into the diode at the fundamental frequency is to inject the fundamental directly into the slot via a coaxial feed. In order to measure the conversion loss, a coaxial feed with a low-pass filter has been used to prevent the second-harmonic signal produced at the diode from leaking back into the fundamental-mode generator. This feed has been used to get an accurate measurement of the power coupled to the diode in the slot. A schematic diagram of such a feed to test diodes for a 34-GHz output doubler is shown in Fig. 5.

The next problem is to measure the amount of radiated power at the second harmonic. One way to measure the power output is to integrate over the radiation pattern of the slot at the second harmonic [9]. This method is very laborious and involves a potentially significant error since an absolute calibration of the radiation intensity is required. Also, measurement of the radiation pattern in three dimensions would be necessary since integrating over just the E- and H-plane patterns will result in a measurement error.

A better way of measuring the radiated power at the second harmonic is to calibrate a receiving antenna placed at a fixed distance from the slot at broadside. The calibration is made by replacing the slot multiplier with an identical slot feed radiating a known amount of power at the second-harmonic frequency. There are several ways in which the calibration slot could be excited. At microwave frequencies the slot could be excited by a coax, but at higher frequencies a waveguide taper would be more appropriate. A schematic depicting this method of calibration is shown in Fig. 6.

In all these calibration procedures, one needs to be sure that the power incident on the calibration slot is well matched. Stub tuners need to be used to optimize the match so that errors in the calibration will be kept to a minimum. One also needs to consider the losses of the

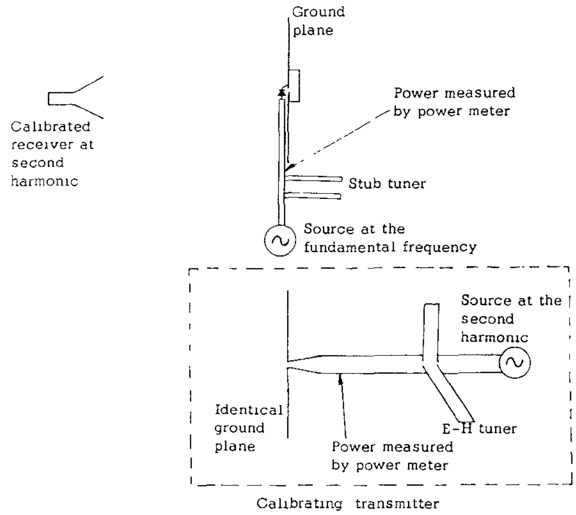

Fig. 6. Schematic for the measurement of conversion loss of the 17- to 34-GHz slot multiplier.

TABLE I
CONVERSION LOSS MEASUREMENTS FOR 8- AND 34-GHZ OUTPUT DOUBLERS USING VARIOUS DIODES

Type of Diode	Conversion loss (dB) for 8GHz Doubler	Conversion loss (dB) for 34GHz Doubler
N.E.C.		
V138(1)	2	-
ND4131	6	10
ND4141	7	11
Texas Instr. MD630	-	7
H.P.		
5082-2299	-	10
HSCH-5330	-	10
Metellics MSKM-717	-	10
MSKM-716	-	10

matching networks if the power is measured in front of these networks.

In this calibration, we assume that the radiation pattern of the calibration slot is identical to that of the doubling slot. For this assumption to hold, the dimensions of the ground plane in both cases need to be the same. One could argue that the radiation patterns are not identical since a nonlinear element is present on the doubling slot. However, we have found experimentally that the measured radiation patterns for both the doubling and the calibration slots are identical.

The conversion loss of several diodes have been measured using this feed system. Table I lists the various diodes tested as 8- and 34-GHz output doublers. For both the 8- and the 34-GHz experiments, the conversion loss obtained is comparable to in-waveguide doublers using the same type of diodes. This implies that the slot doubler

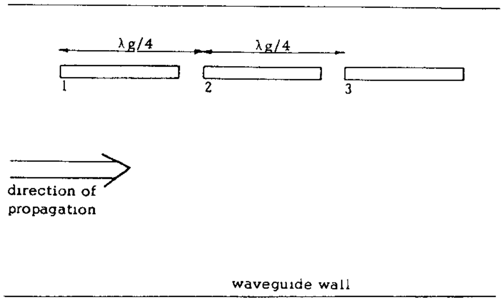

Fig. 7. Geometry of slots with respect to waveguide and direction of propagation.

circuit used in this design does not have any excessive loss. No conversion-loss measurements were made at higher frequencies since it is hard to feed a millimeter-wave signal via a coax. From the measurements made up to 34 GHz, it is evident that scaling up in frequency should not be detrimental and that the conversion efficiency will depend only on the type of diode used. The radiating third and fourth harmonics for the 8-GHz output doubler were measured to be 20 and 25 dB down from the second harmonic, respectively.

IV. Multiplying Slot Array

Since the multiplying slot array is a phased array at the second harmonic, the phase relationship of each antenna element at such a frequency needs to be taken into account in the design of the array. In our design, the slots are placed $\lambda_g/4$ apart where λ_g is the guide wavelength at the input frequency. Since the multiplying diode elements act as square-law devices, the slots are alternately out of phase by 180° at the second harmonic even though they are 90° ($\lambda_g/4$) out of phase at the fundamental.

This square-law behavior could be explained as follows. Let us assume that the second diodes are characterized by

$$i = AV^2$$

where i is the diode current, A is a constant, and V is the fundamental voltage across the diodes.

Referring to slots 1, 2, and 3 in Fig. 7, one could formulate the phase relationship of the second-harmonic currents in the slots. Since the slots are placed a quarter of a guide wavelength apart at the fundamental input voltage, one could represent the voltages of the slots with respect to the voltage of slot 1 as follows:

slot 1: $V\cos\omega t$.
slot 2: $V\cos(\omega t - \pi/2)$, quarter wavelength apart, 90° phase shift.
slot 3: $V\cos(\omega t - \pi)$, half wavelength apart, 180° phase shift.

One could then write the diode currents across the slots through the diode as follows:

slot 1: $i = \frac{1}{2}AV^2(1+\cos 2\omega t)$
slot 2: $i = \frac{1}{2}AV^2(1-\cos 2\omega t)$
slot 3: $i = \frac{1}{2}AV^2(1+\cos 2\omega t)$

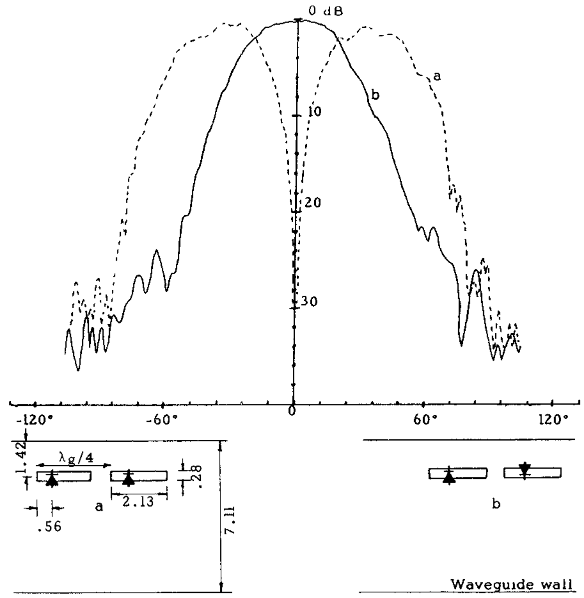

Fig. 8. E-plane radiation patterns measured at 70 GHz for two slots placed a quarter of a guided wavelength apart at the fundamental (35 GHz) (all dimensions in millimeters).

From these equations, it is evident that slot 1 and slot 3 are in phase at the second harmonic while slot 2 is 180° out of phase with slots 1 and 3. This phase difference between the three slots can be compensated for by changing the polarity of the diode on slot 2 with respect to the diodes in slots 1 and 3. Thus, all the slots will be in phase at the second harmonic, producing a single main beam at broadside.

This phase behavior can be explained best by Fig. 8, where two slots are placed $\lambda \hat{g}/4$ apart at the 35-GHz fundamental frequency. The pattern of Fig. 8(a) shows that the slots placed $\lambda_g/4$ apart are out of phase by 180° at the second-harmonic output, while the pattern of Fig. 8(b) shows that they are in phase due to the phase change caused by reversing the polarity of one of the diodes. This additional phasing effect, due to the square-law behavior of the diodes, makes it feasible to place the slots close to a half-wavelength apart at the radiating second-harmonic frequency, thereby obtaining a single beam. From the power combiner point of view, this phasing effect increases the density of the multiplying elements, resulting in a more compact design for a given number of elements.

Fig. 9 shows the normalized power patterns of an eight-element array multiplying from 35 to 70 GHz. The slots are placed $\lambda_g/4$ apart at the fundamental (corresponding to 0.59 of the free-space wavelength at the radiating second harmonic) in the E-plane direction. A picture of the array is shown in Fig. 10. This array produced sidelobes in the E-plane that are 12 dB below the main lobe. The sidelobe level in the E-plane depends on the illumination distribution of the array. In the array of Fig. 9, the power distribution decreases linearly along the array since it is fed from one side. A better way to feed the slot array is to power split the input waveguide right at the center of the array so that the slots in the middle will radiate the maximum

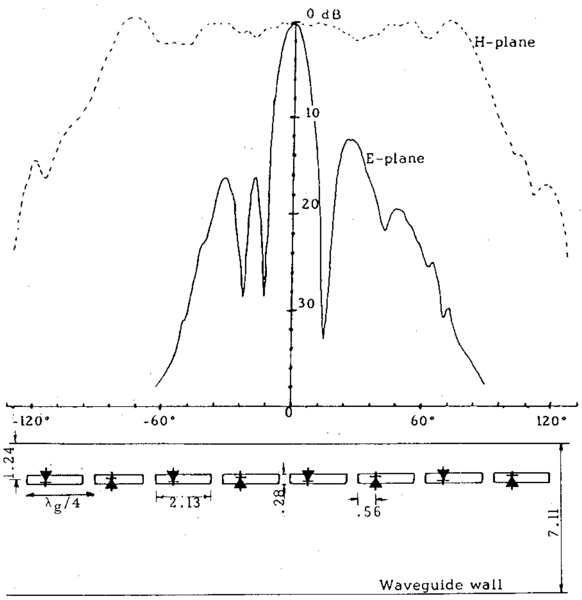

Fig. 9. Radiation pattern measured at 70 GHz for an eight-element multiplying slot array. This active array converts the 35-GHz waveguide fed input to 70-GHz radiated output (all dimensions in millimeters).

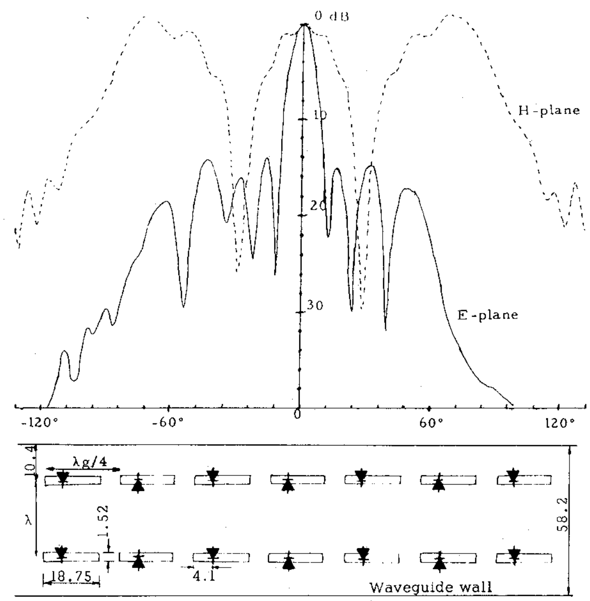

Fig. 11. Radiation pattern of a two-by-seven doubling slot array at 4.2-GHz fundamental frequency (all dimensions in millimeters).

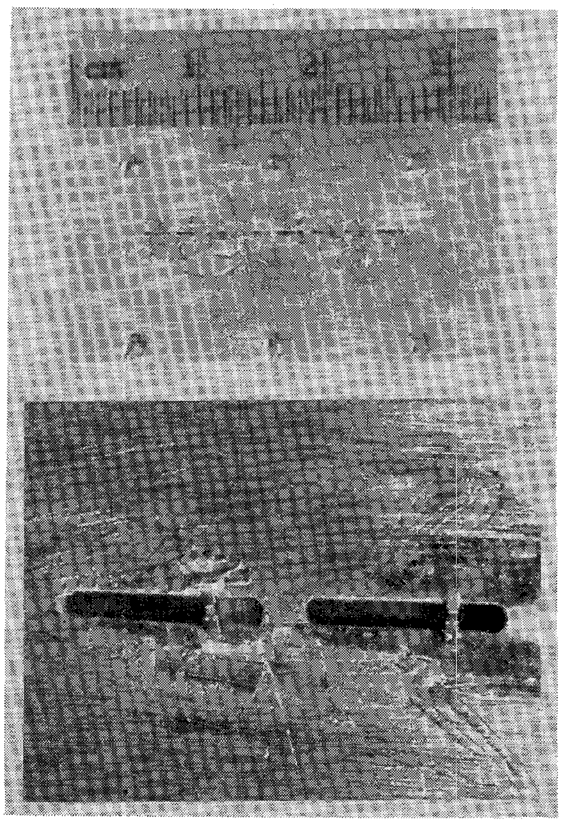

Fig. 10. Top picture shows a 70-GHz eight-element slot array. The bottom picture shows two slots magnified.

power and the illumination power will taper towards the ends of the array. The H-plane has a fan-shaped pattern which is only dependent on the ground-plane dimensions. If a pencil beam is desired, a cylindrical parabolic reflector can be used with the array as the feed.

In an attempt to determine the useful bandwidth of operation of the slot array, experiments have been performed at X-band where a powerful wide-band source is easily available for the fundamental. Fig. 11 shows the radiation pattern for a two-by-seven slot array with 8-GHz second-harmonic output. This radiation pattern indicates that the sidelobes are 14 dB down in the E-plane. It has been experimentally determined that the sidelobes are better than 11 dB down over a 20-percent bandwidth. The H-plane pattern for the two-by-seven slot array is maple leaf shaped. This occurs since the slots are placed one wavelength apart in the H-plane direction. Placing the slots closer together in this direction is difficult since the slot position is determined by the modes propagating in the waveguide.

V. Conclusions

A novel multiplying slot array with space-combining capability has been described. The feasibility of obtaining adequate single element conversion efficiency and the use of high density slot elements has been successfully demonstrated. The simplicity of construction of such a device makes it useful at millimeter-wave frequencies. The possibility of future monolithic design is very feasible due to the simplicity of the structure. Several beam shapes can be synthesized by using different diode polarities and reflectors. The use of planar transmission lines, such as microstrip, is also possible.

Acknowledgment

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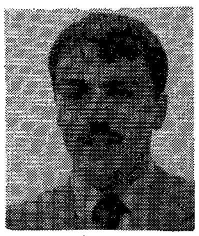

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