Investigation of Active Antenna Arrays at 60 GHz

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Abstract—**There has been a significant effort to develop millimeter-wave active-array antennas for communications and radar applications [1]–[6], [31], [34], [35]. A dielectric waveguide is a promising medium for this application. However, the integration of active devices, transmission media, and antennas has been difficult to achieve. This paper presents the first successful demonstration of a phase locked array of millimeter wave grating surface emitters (MMWGSE). We discuss three aspects of MMWGSE: 1) The achievement of an optically steered millimeter wave grating surface emitter. 2) The demonstration of a frequency locked array of millimeter wave grating surface emitters. 3) Rigorous analytical studies of efficiently coupling power from a millimeter wave semiconductor device, to a waveguide which incorporates grating surface emitters. This work leads to a full monolithic array using pseudomorphic high electron mobility transistor (PHEMT)-devices.**

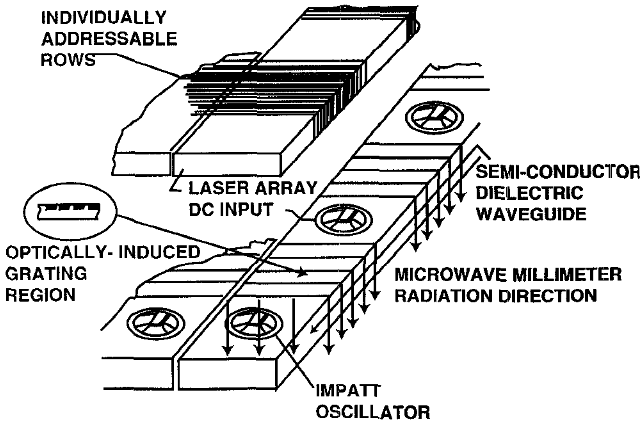

Fig. 1. Millimeter-wave surface emitter with optically controlled gratings.

I. Introduction

There is a need for the development of high efficiency high power solid state amplifiers that can replace conventional amplifier combining circuits in order to reduce the losses in the splitters/combiners guiding structures. This can be achieved by coupling the waves into and out of the amplifiers through radiating elements (space combining). This approach has led to an increased interest in the area of quasi-optical and spatial power amplification, since some of the added benefits of spatial power combining are: graceful degradation, and the realization of significant output power levels from modest size transistors. The use of lower power transistors is beneficial to the spatial combining scheme as they have lower cost and are, in general, more efficient than those with relatively higher power outputs. Spatial combining at X-band has been successfully demonstrated by others [7]–[10], for example:

1) A nine HEMT spatial amplifier at X-band. The amplifier was constructed by interconnecting many devices by microstrip lines. The maximum measured value of gain was found to be 5.6 dB, and the 3-dB bandwidth of 1 GHz was centered at 10.9 GHz.
2) A 4×4 array was constructed at 10 GHz. The measured results show an EIRP of 25 dBW, an output power greater than 4W, and a dc-rf conversion efficiency greater than 18%. Each element consists of a MMIC amplifier and a microstrip antenna coupled at its proximity.
3) A 4×4 spatial combining array with strongly coupled Gunn oscillators was demonstrated.
4) A 100 element heterojunction bipolar transistor (HBT) grid amplifier was demonstrated at 10 GHz. Self-biased HBT devices were used as the active devices. The grid had a 10 dB gain and a 1 GHz 3dB bandwidth. The maximum output power was 450 mW, and a 5% efficiency was reported.

The integration of active devices, transmission media, and antennas has been difficult to achieve at mm wave frequencies. Therefore, the task of millimeter-wave system design is hampered by the unavailability of high-power, compact, broadband millimeter-wave emitters. Available millimeter-wave generating devices, distribution elements, and radiating structures of integrated emitters are multi-planar and require complicated feed elements, which all radiate and degrade the quality of the desired space combined signal. One solution to this problem is the use of millimeter-wave grating surface emitters (MMWGSE), a concept that was introduced in 1982 and shown in Fig. 1.

Recently, we have demonstrated key aspects of this concept: 1) The optical steering ability of the millimeter wave grating surface emitter, up to 30°. 2) The frequency locking of the array of millimeter wave grating surface emitters (MMWGSE). Up to six IMPATT-based oscillators, operating at 62 GHz, have been mutually locked in a 2-D array utilizing top-metallized-image (TMI) guides with slot-type gratings. However, the far-field of this structure did not show a confined beam since no provision was made for phase adjustments between

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stages. 3) A rigorous analytical study of the coupling scheme from a millimeter wave semiconductor device to a waveguide incorporating grating surface emitters. Theory indicates that a power coupling efficiency to the fundamental waveguide mode of as high as 65% can be achieved.

This paper describes our efforts to obtain high power active antennas at millimeter waves. We lead the reader through a series of tasks, showing how the constraints of our goals influence design decisions and changes in our approach.

The investigation of IMPATT sources using both fixed and optically induced gratings as the antenna elements is included. We show how system constraints have caused the redirection of the investigation to PHEMT active elements. The latter became attractive because of the recent progress in the development of high efficiency power PHEMT amplifiers at 60 GHz. The change in active device causes a change in the wave guiding system resulting in a fixed grating in microstrip as the antenna element.

In this paper we not only list the many areas of technology that we have investigated, but also describe the research process leading up to our final methodology.

In Section II we discuss optically induced gratings and show how we have achieved greater than 30° steering ability through the application of a variable periodic optical source.

In Sections III and IV we describe the properties of the TMI guide and analyze the coupling from the probe to the guide respectively showing the feasibility of up to 65% coupling efficiency.

In Section V we show how the above concepts lead to the active array configuration. Experimental results of this array agree very well with our predicted results. However, the limitations of IMPATT sources have led us to consider the use of HEMT's and MMIC's which we discuss in Section VI. We have migrated from TMI guide to microstrip while preserving the approach of the previous works. Results were very encouraging and, once again, have agreed with theoretical predictions. This represents a milestone for an active antenna array utilizing PHEMT at 60 GHz.

II. Optically Induced Gratings in Silicon and Silicon on Sapphire (SOS) Dielectric Guides [3]

The propagation velocity of modes in the dielectric waveguides can be varied dynamically by either electrically or optically controlling the dielectric constant. At millimeter-wave frequencies, a promising technique for controlling the dielectric constant of materials is the use of free-carrier effects in the material.

We have extended our dielectric guide work, in which we use thin layers of a uniform plasma concentration for controlling the dielectric constant, to layers with periodically varying concentrations of plasma. Periodic plasma regions are obtained along the axis of propagation by using a linear array of injection lasers. The period is varied by exciting specified lasers of the array.

The first optically induced grating experiment was performed using a 1.4×1.4 mm dielectric guide. The mask used had a 0.2 mm grating stripe width and a 1.85-mm grating period. Teflon clamps were used to hold the emulsion mask in contact with the guide (with the emulsion side touching the guide). The emulsion masks served to create a periodic pattern of laser energy on the dielectric guide when illuminated by our high power laser diode array. Teflon and nonmetallic emulsion materials were selected to minimize interference with the millimeter waves. A 2D edge-emitting laser diode array with a power conversion efficiency of 40% was used as the optical source (Fig. 2). The output of the array was adjusted to about 150 W. This array was pulsed at a repetition rate of 100 Hz with a pulse width of 10 μs. An emulsion and a 2-D edge emitting laser diode array were used to induce a grating in a dielectric guide carrying a 60 GHz signal, and the far-field angle at which the peak in the pattern was observed, $\emptyset$, depends on the grating period, d. This angle increases with increasing period: $\emptyset = 90^\circ + \sin^{-1}[(\lambda_o/\lambda_g) - (\lambda_o/d)]$ where λ_o and λ_g are the free space and guide wavelength, respectively. The steering angle exceeded 30° from broadside as demonstrated by comparing the main lobe angle for five different masks. The measured and the theoretical values of these angles are illustrated in Fig. 3. Since we are using an incoherent laser source with a beam divergence of ±26°, the period of the induced grating is slightly greater than the period of the printed grating on the mask. Therefore the far field angle is greater than the expected angle by a few degrees. Fig. 4 depicts SOS optically induced gratings (OIG) steering at 60 GHz.

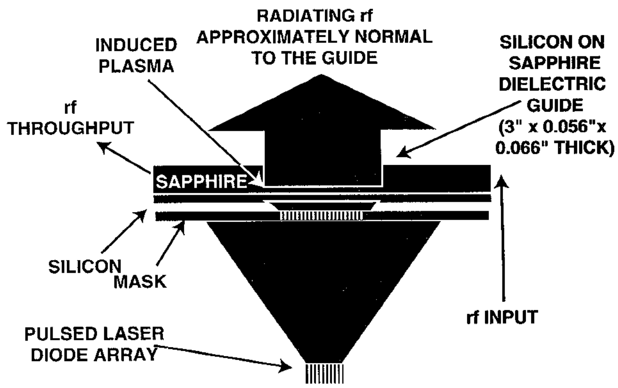


Fig. 2. Experimental setup for a millimeter-wave surface emitter with optically controlled gratings.

III. TMI Guide Modes [11]–[14]

In this work, we analyze a new coupling and waveguide structure designed to meet the needs for efficiently coupling power from a millimeter-wave semiconductor device, such as an IMPATT. The waveguide incorporates a grating surface emitter (IMPATT devices at 60 GHz were considered because of their high power capability).

The photograph in Fig. 5 identifies the critical parts of the millimeter-wave grating surface emitter (MMWGSE). An IMPATT is mounted at the bottom of a cylindrical cavity in the ground plane below the waveguide (see Fig. 6). The RF energy is coupled into the waveguide through a cylindrical post coming from the top of the IMPATT, passing through the dielectric, and contacting the top metal. Periodic slits in the top metal form a second-order grating which acts as the antenna.

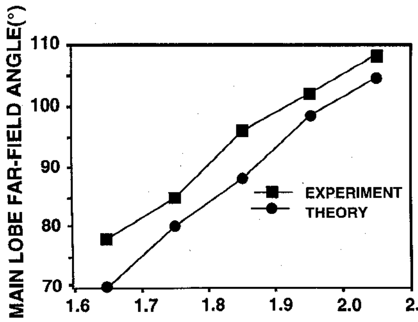


Fig. 3. The measured and theoretical main lobe angle for various grating periods (in silicon guide).

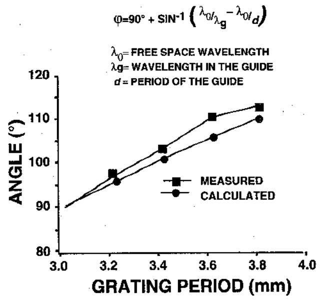


Fig. 4. Steering at 60 GHz (in SOS guide).

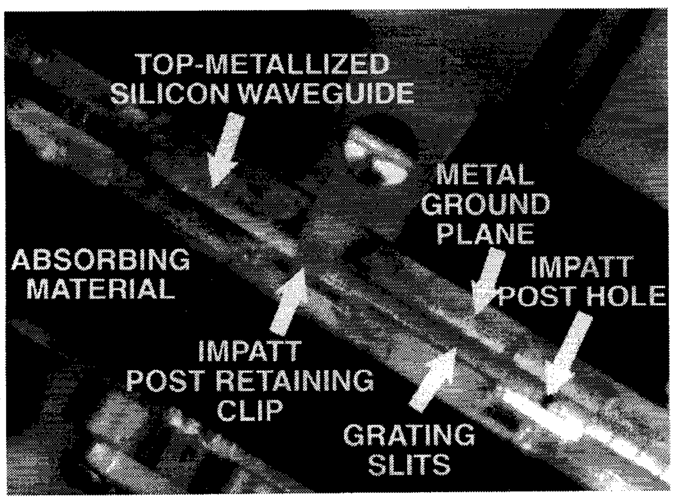


Fig. 5. Photograph of device showing the critical parts.

This structure introduces several beneficial features that maximize power coupling into the waveguide, facilitates mutual coupling of IMPATTS in the longitudinal direction for linear arrays, and provides energy for lateral coupling to adjacent waveguides through leaky waves for two-dimensional arrays. In the following paragraph we present a model for understanding the MMWGSE. In addition we calculate the power coupling into the waveguide and cite measurements in support of the model.

The waveguide consists of a top-metallized dielectric of height b and width $2a$ over a ground plane forming a top-metallized image guide (TMI-guide) which is strongly similar to an H-guide. We justify this claim by comparing the modes of the two waveguides. We numerically solve the vector wave equation for the TMI-guide and analytically solve the wave equation for the H-guide [15], [16]. Fig. 7 shows that the fundamental model of both guides has the same guided wavelength as a function of width. Furthermore, the wavelengths of the next order modes are within 5% of one another.

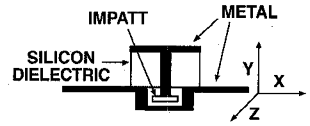


Fig. 6. Cross-section of waveguide, cavity, and post.

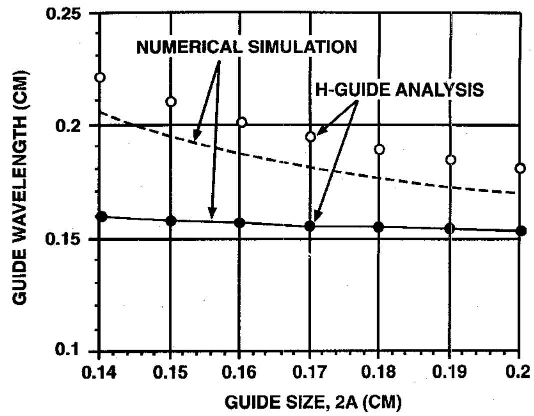


Fig. 7. Comparison of numerical solution for modes in the TMI-guide and analytic solutions for modes in an H-guide.

H-guides support two classes of modes bound to the dielectric which are denoted by PE_{mn} and PM_{mn}. The PE_{m0} modes, which have no variation in the y direction and therefore no z component, are equivalent to TE_{m0} modes and are the modes that are launched by a uniform source (with respect to y). If the source is also symmetrical in both placement and current flow (with respect to x), then only even PE_{m0} modes will be excited. Lateral leaky modes, which are present in finite H-guides (and therefore TMI-guides), reduce the power coupling from the post to the bound modes.

IV. Post Excitation of the Waveguide

Complete theoretical solutions for the coupling of RF power using a probe in a rectangular metal waveguide have been obtained [17]. Because of the difficulty in extending this earlier work to dielectric guides, we have used a different approach. Since the TMI-guide is an open guide with both continuous and discrete spectrums, we approximate the TMI-guide by an H-guide and then close the ends to form a rectangular waveguide as shown in Fig. 8. This configuration has several advantages: a) the closed waveguide makes the spectrum discrete, thereby simplifying the mathematics, b) the modes may be calculated analytically using the transverse-resonance method [18], and c) by letting the waveguide width, $2c$, go to infinity we recover the H-guide.

Several modes of this waveguide are shown in Fig. 9. The fundamental mode is bound to the dielectric and the higher-order modes belong to the rectangular waveguide. These higher order modes approximate the leaky modes of the H-

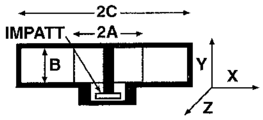
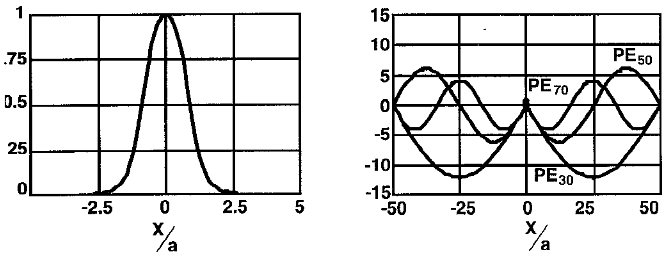


Fig. 8. Cross section of model used for calculating power coupling.



(a) (b)

Fig. 9. Modes of the dielectrically loaded rectangular waveguide ($c = 50a, a = 0.07$ cm, $f = 60$ GHz) normalized to unity at $x = 0$ (a) PE_{10} mode (b) PE_{30}, PE_{50}, and PE_{70} modes.

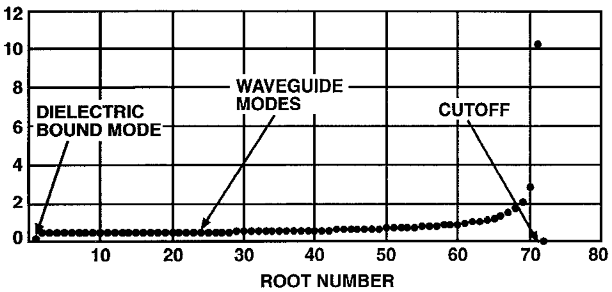


Fig. 10. Modes of the dielectrically loaded rectangular waveguide ($c = 250a, a = 0.07$ cm, $f = 60$ GHz).

guide (TMI-guide). As c is increased, the bound mode remains unchanged but the higher-order modes increase in number, thereby approximating the continuous part of the spectrum. Fig. 10 shows the mode spectrum for the even modes.

The calculation of coupling from the post to the waveguide modes is based on the assumption that the post thickness is negligible (z direction) and it can be treated as a rectangular strip. Although this is a serious assumption, we believe that it primarily affects an equivalent circuit for the post rather than the power coupling from the cavity to the waveguide.

Overlap integrals of the strip and the modes are computed. Since the overlap integral is proportional to the power coupled to a mode, the ratio of the overlap integral of the dielectric bound mode to those of all the propagating modes is the power coupling coefficient. Fig. 11 shows the results of these calculations for several values of post diameter, d, and rectangular waveguide width. We found that the power coupling coefficient varied linearly with respect to $\sqrt{a/c}$. When the calculations are extrapolated to $c \to \infty$, the power coupling coefficient for the H-guide (TMI-guide) is found.

Measurements were made in TMI-guides with and without a grating [19]. On structures without the grating we measured the power coupled to the guides by inserting a tapered section of TMI-guide into a standard V-band measurement fixture based on a rectangular waveguide. Record power was achieved: 427 mW at 62 GHz, as measured with a standard V-band metal

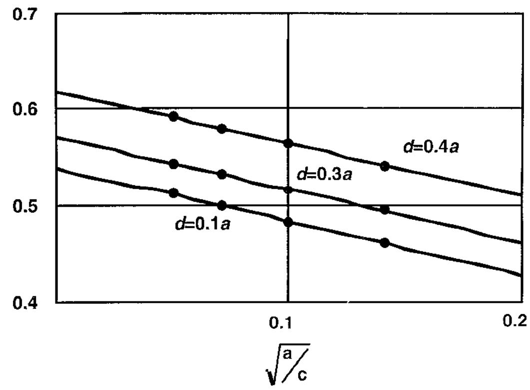
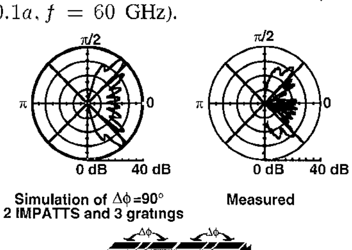


Fig. 11. The power coupling coefficient versus $\sqrt{a/c}$ ($a = 0.07$, $d = 0.4a, 0.3a, 0.1a$, $f = 60$ GHz).



(a) (b)

Fig. 12. Measured (a) and computed (b) far fields ($\Delta f = 90°$ and $a = 0.5\,\mathrm{cm}^{-1}$).

waveguide test bed. This is a factor of four higher than power from oscillators that were fabricated from either dielectric waveguide or image guide. We compared this power to the power achievable by the same IMPATT in a reduced height coaxial metal cavity and found the ratio to be 0.65, which closely corresponds to our calculation of 0.62 as shown in Fig. 11.

V. 2-D Active Antenna Utilizing IMPATT Devices [19]–[28]

Two IMPATT diodes have successfully and repeatedly been mutually locked to one another in one TMI guide with a 93% grating. Fig. 12 shows a comparison of measured and calculated far-field patterns. The calculated far field is based upon the theoretical near field pattern for a single propagating bound mode which varies as $\exp(-|\alpha z| + j(\Delta\phi/2)\,\mathrm{sign}(z))$ where $\Delta\phi$ is the excess phase introduced across the post coupling region and α is the attenuation rate in the waveguide due to the grating radiation and losses.

The far field has two symmetrical split beams. The symmetrical beams are due to the two grating antennas being situated on either side of the millimeter wave source and radiating away from the second bragg condition. The split beam is a consequence of phase delays introduced by the IMPATT/post structure as shown in Fig. 12. The split in each lobe is due to an estimated phase shift of about 90°. The agreement between the two patterns supports our contention that a single bound mode exists. The gain of this active antenna was measured to be 10 dB approximately.

A new set of measurements featuring three IMPATT diodes oscillating at the frequency of 61.7 GHz was performed. In these tests, all three IMPATT devices in the linear array

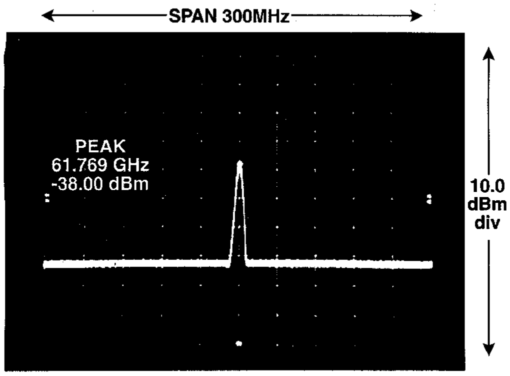


Fig. 13. Only the left oscillator of the three element linear array is biased.

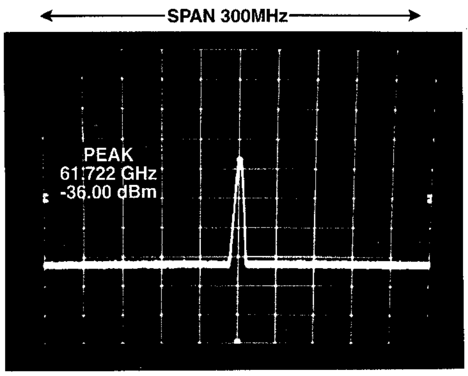


Fig. 14. The right and left oscillators of the three element linear array are locked to one another.

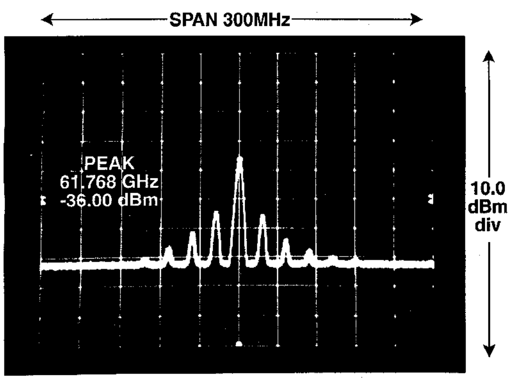


Fig. 15. The right and left oscillators are locked to one another and are beating with the center cell of the three element linear array.

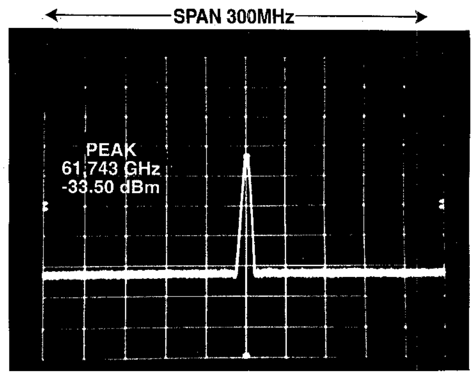


Fig. 16. All three elements of the linear array are biased and locked to one another.

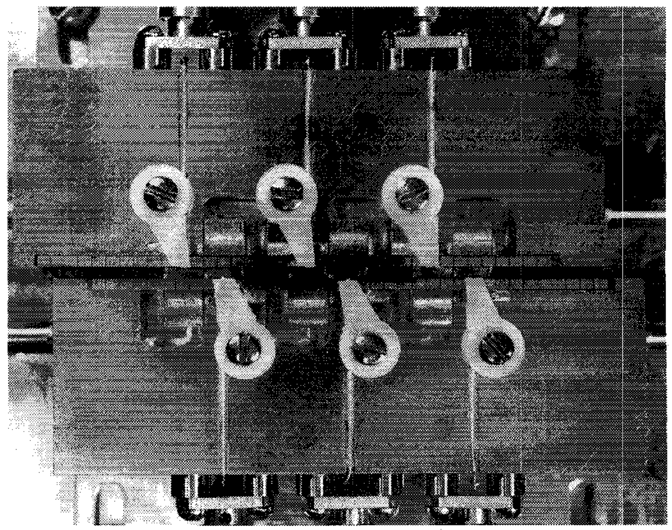


Fig. 17. 2-D array.

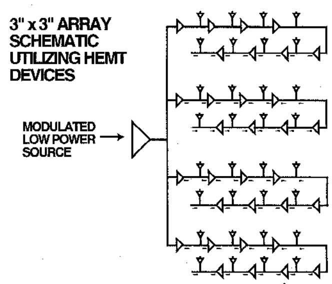


Fig. 18. New approach utilizing HEMT devices.

locked to one another. Fig. 13 depicts diode oscillation in the left circuit of the linear array. Fig. 14 depicts the right and left circuits locked to one another. Fig. 15 depicts the right and left circuits locked and the center beating with it (just before locking). Fig. 10 depicts the three diode MMWGSE, in operation (all oscillator circuits locked to one another). Each time an additional diode was biased, the power level emitted by the active, linear array and detected in the far-field increased (by an average of 2.25 dB with each additional diode). The far field of this structure did not show a confined beam, since no provision was made for phase adjustments.

Finally, we were successful in introducing two linear arrays, each consisting of three locked IMPATT oscillators, side by side (constituting a 2-D array having six locked IMPATT devices), Fig. 17. This accomplishment indicates that a large number of millimeter wave devices can be locked to one another through MTI guides without the utilization of isolators. The incremental phase between the radiating elements is adjusted, thus a far field can be achieved with a confined main beam. However, the IMPATT diode is a one port device, and its energy is coupled to the gratings on both sides of the IMPATT cavity. Hence, phase adjustments as well as the gratings' uniformity are very critical.

VI. A Linear Active Antenna Utilizing PHEMT—A First Step for Monolithic Array Configuration

As a consequence of: 1) the difficulty in providing a single main beam when using an IMPATT device between two sets of gratings (when the gratings are radiating off the Bragg condition), 2) the increased difficulty in processing IMPATT devices utilizing the MMIC approach, and 3) the availability of potential high efficient high power PHEMT's, we have investigated the possibility of replacing the IMPATT devices by PHEMT devices.

PHEMT's are making rapid progress in millimeter wave power amplifier applications, and very encouraging results are being reported for as high as 94 Hz [29], [30].

TRW [29] has reported a two stage amplifier with better than 370 mW output power, a compressed gain of 7 dB and greater than 11% power-added efficiency over the frequency range of 59.5–63.5 GHz. With low loss planar combiners, these modules produced an output power of 740 mW with a power gain of 11.68 dB.

VII. Microstrip Based Gratings

The design of wideband microstrip arrays is outlined here. A simple configuration of series or parallel arrays is used. By selecting the resonant frequency of each element according to the bandwidth requirements, a reasonably good bandwidth with acceptable VSWR can be achieved. Hence, the return loss can be made flat rather than selective at any specific frequency. A large number of elements and wide band match-terminations for traveling wave arrays need to be used. Elements are connected to the main feed at a distance of $n\lambda_g/2$ from the termination, and spaced $\lambda_g/4$ apart as each segment is a $\lambda_g/4$ matching transformer to the 50 ohms. A $\beta - k$ diagram can be used to derive the propagation characteristics of this periodic structure as shown by [33] and given by the following formula

$$\cos\beta_a d = \frac{\cos kd + j\dfrac{z_o}{2Z_{11}}\sin kd}{1 - j\dfrac{Z_o}{Z_{12}}\sin kd}$$
$$\beta_a = \beta + j\alpha$$
$$k_o = 2\pi\sqrt{\varepsilon_{\text{eff}}}/\lambda_o \tag{1}$$

where β_a is the complex propagation constant, Z_o and k are the feedline characteristic impedance and wavenumber respectively, Z_{11} and Z_{12} are the self and mutual impedances of the resonant element taking successive coupling into account, d is the interelement spacing, k_o is the effective wavenumber, and ε_{eff} is the effective dielectric constant.

The bandwidth of these radiating elements is limited due to the existence of stop bands in the $k - \beta$ diagram. The main beam direction can be calculated from the following equations

$$d\sin\theta + \sqrt{\varepsilon_{\text{eff}}}l = \lambda = \frac{c}{f} \tag{2}$$

where θ is the main beam angle, d is the interelement spacing in air, l is the length of the transmission line joining the successive elements, c is the speed of light in the dielectric, ε_r is the dielectric constant, and f is the frequency of operation [33]. The amplitude distribution along the array can be adjusted for low side lobe level performance simply, by changing the width of these radiating patch elements.

Simple calculations using $d = 1.78$ nm, $l = 0.1$ mn, $f = 60$ GHz, $\varepsilon_r = 13$ have indicated a main beam angle of 35 degrees off broadside (Fig. 19), which is in close agreement with the measurements, as seen in Fig. 20.

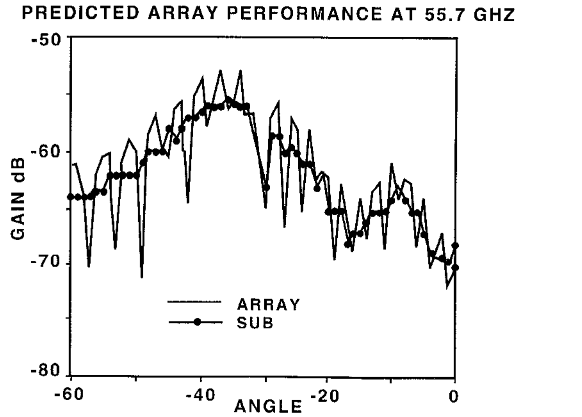


Fig. 19. Calculated subarray radiation pattern, and the predicted whole array pattern due to the contributions of such two subarrays.

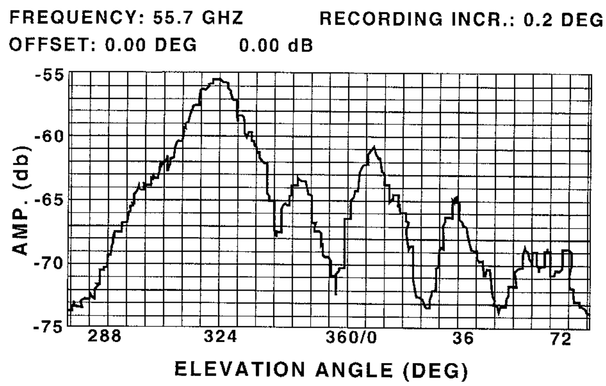


Fig. 20. Measured radiation pattern of a subarray at $f = 55.7$ GHz.

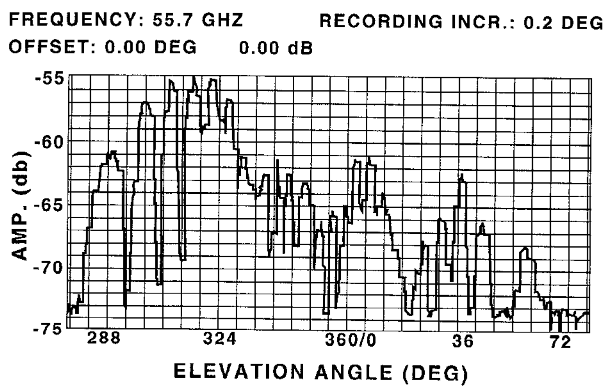


Fig. 21. The measured radiation pattern of two subarrays.

A combined PHEMT amplifier with a microstrip grating antenna comprises a "unit cell." Fig. 22 depicts the combination of two unit cells creating a one-dimensional (linear) antenna.

Experimental Results:

We have analyzed the periodic structure using a circuit ana lysis program (Touchstone), and an EM Simulator (Ensemble). The structure was modeled as successive wide transmission lines of 0.13 mm in width and 1.68 mm in length that are followed by narrow transmission lines 0.02 mm in width

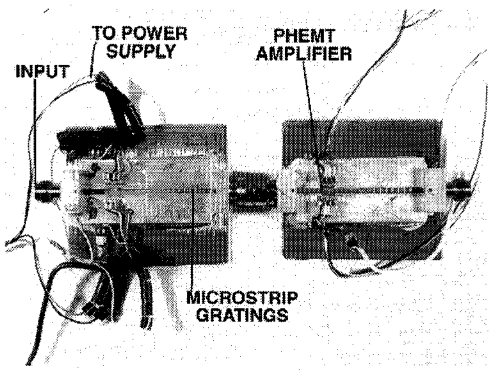


Fig. 22. Two unit cells creating a one-dimensional (linear) antenna.

and 0.1 mm in length. The structure consisted of 7 such narrow lines connected to 50-ohm terminations on both sides. The analysis showed a good VSWR performance over the 55–60 GHz frequency range (less than 2 : 1). The measured results agree closely with the predicted results. In addition, a transmission loss of 3 dB was measured over an overall transmission line of 10 mm in length (that includes all successive gap-discontinuities), which matches the predicted values closely.

We have used both (1), (2) and the EM simulation program ENSEMBLE to predict the radiation pattern of this series fed array. Equation (2) showed that a main beam should exit at an angle of 35 degrees off broadside. The radiation pattern as well as input-output return loss were also calculated. The pattern at 60 GHz, for example, has a beam at 35 degrees, which is very close to the value predicted by (2) and agrees very well with the measured results as seen in Fig. 20. The side lobe levels are relatively high since no effort was directed towards lowering their levels by controlling the amplitude or phase distribution of the whole array.

The antenna gain was measured relative to a standard feedhorn. The feedhorn was measured first in order to calibrate the system. Measurements of the microstrip grating antenna showed a 4 dB higher gain as compared to the horn antenna, indicating a relatively high 65% efficiency as calculated by $G/A = (4\pi/\lambda^2)\eta$.

Two such subarrays were connected in cascade to investigate spatial combining. The two subarrays were spaced 400 mils apart and the array factor of such an assembly is shown in Fig. 21. As expected the combined measured pattern showed the power combining from the two subarrays as the signal was increased by almost 2 dB, and the sidelobe levels and the number of nulls are related to the spacing between the two subarrays. Performance of this array can be significantly improved by eliminating the long section of interconnecting transmission lines between the two subarrays and by introducing some phase adjustments. Based on the input power of 5 mW, the MMIC amplifier power gain of 6 dB at 56.5 GHz, and a radiated efficiency of 65%, the estimated output power of a unit cell is 13 mW. The output power from the MMIC amplifier chip used in our experiments saturated above 5 mW of input power. The radiating structure has yielded a bandwidth of at least 1 GHz. This is the first demonstration of an active linear array antenna utilizing PHEMT at V-band. Furthermore, this design configuration lends itself to a monolithic implementation.

VIII. Conclusion

The three major components for a steerable millimeter wave active antenna were obtained: 1) an optically steered millimeter wave grating surface emitter with a 30° shift; 2) a phase locked active antenna with up to six IMPATT oscillators; and 3) an analysis of a new coupling structure and waveguide. Theory indicated a single bound mode with a coupling efficiency of as high as 65%. In addition, a linear active antenna utilizing two PHEMT amplifiers at 60 GHz and a microstrip configuration antenna were demonstrated. Furthermore, the design configured lends itself to a monolithic implementation.

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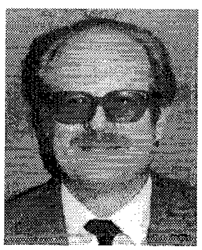

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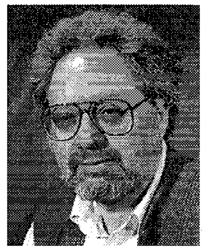
Dr. Rosen is the recipient of a 1972 and a 1982 RCA Laboratories Outstanding Achievement Award, the Microwave Prize at the 16th European Microwave Conference in 1986 for co-authoring the paper entitled "Optical Beam Control of Millimeter Wave Phased Array Antennas for Communications," a 1989 David Sarnoff Research Center Technical Achievement Award for conception and implementation of high-power optically activated semiconductor switching technology operating in the frequency range from DC to GHz, a 1989 IEEE Region One Award for significant contributions to microwave technology by the invention and development of microwave balloon angioplasty, and a 1990 David Sarnoff Research Center Technical Achievement Award for conceiving and proposing an optically controlled, surface emitting, millimeter wave steerable source. He is a Fellow of the IEEE. He has been a member of the IEEE MTT-S Technical Program Committee since 1979; he is MTT-S Technical Committee Chairman on Biological Effects and Medical Applications; a member of the IEEE MTT-S Technical Committee for Light-Wave Technology; Associate Editor of the IEEE JOURNAL OF LIGHT-WAVE TECHNOLOGY (JLT); and a member of the Editorial Board of Microwave and Optical Technology Letters and of Microwave and Guided Wave Letters. He has served on the Technical Committee for the IEEE International Conference on Microwaves in Medicine held in Belgrade, Yugoslavia, in April 1991, and on the Program Committee for Electro/International 1985 and 1991, in which he chaired the Medical Sessions. He is a Member-at-Large of the IEEE Health Care Engineering Policy Committee, and has served on the IEEE Educational Activities Board; and he holds the title of Professional Engineer in the Province of British Columbia, Canada. His biography has been selected for inclusion in Marquis Who's Who in the World.



Robert Amantea received the B.S.E.E. degree from City College of New York in 1965, the M.S. degree in electrophysics from Polytechnic Institute of New York in 1969, and a Ph.D. degree in electrical engineering from the University of California at Berkeley in 1977.

He joined the David Sarnoff Research Center in 1965 where he is a Member of the Technical Staff. During the interim he has received a David Sarnoff Fellowship to pursue the Ph.D. degree and two David Sarnoff Achievement Awards. He has presented or published approximately 50 papers in the fields of semiconductors, including bipolar transistors, MOS transistors, semiconductor lasers, and microwave devices. He is the holder of four patents. He specializes in the development of mathematical models for the understanding and optimization of semiconductor devices.

Paul J. Stabile (S'75–M'79–SM'88) received the B.E.E.E. degree summa cum laude from Manhattan College in 1979, the M.S.E.E. and Ph.D. degrees from Rutgers University in 1982 and 1992, repectively.

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Aly E. Fathy, (S'82–M'84–SM'92) photograph and biography not available at the time of publication.

Dean B. Gilbert, photograph and biography not available at the time of publication.

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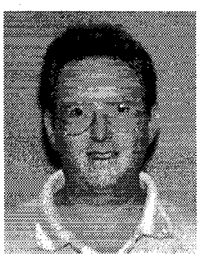
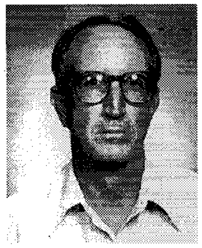
Walter M. Janton served in the U.S. Army as a Communications Technician from 1939 to 1945. In 1948 he graduated from the Baronian School of Horology, Philadelphia, PA. From 1953–1959 he was a Senior Research Technician at Bell Telephone Laboratories, Murray Hill, NJ in the Model Tube Laboratory. In 1959 he transferred to SFD Research Laboratory (Varian Associates). He joined RCA Laboratories in Princeton, NJ in 1972 as a Research Technician. He has worked on the electron tube injected beam transistor program and cathode-luminescent display devices. For the past ten years, he has been responsible for the mechanical design and fabrication on the millimeter-wave, p-i-n, and varactor diode projects, as well as laser arrays and their packaging. He has co-authored 30 technical papers and six patents.



Francis J. McGinty graduated from Pennco Technical Institute in 1978 and received the Associate degree in electronics from Bucks County College in 1980. From 1974 to 1980 he worked in Research and Development at Milton Roy Co. where he assisted in development of flow-control instrumentation. He received a commendation for his contribution to dialysate metering technology.

In 1980, he joined the Microwave Technology Division at David Sarnoff Research Center where he presently works as an AMTS. He has been involved in the development of microwave solid state power amplifiers and receivers and has contributed in the development of high definition TV and microwave video distribution systems. He is co-author of several publications.

Mr. McGinty has received five technical commendations including an individual Sarnoff Technical Achievement Award in 1994.



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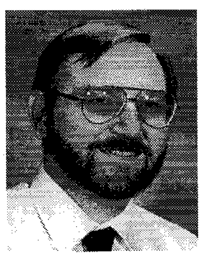
Dr. Butler was elected a Fellow of IEEE for contributions to semiconductor lasers and the theory of radiation characteristics of optical waveguides. He is a member of Sigma Xi, Tau Beta Pi, Eta Kappa Nu, and is a registered professional engineer in the State of Texas.



Gary A. Evans (S'69–M'75–SM'82–F'92) was born in Omak, WA. He received the B.S.E.E. degree from the University of Washington, Seattle, in 1970, and the M.S.E.E. and Ph.D. in electrical engineering and physics from the California Institute of Technology, Pasadena, in 1971 and 1975, respectively.

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Dr. Evans was elected a Fellow of the IEEE for "contributions to the development, fabrication, and understanding of semiconductor lasers." He is a licensed professional engineer, has served on numerous IEEE committees, is a past Chairman of the Princeton Lasers and Electro-Optic Society (LEOS), a past Chairman of the Santa Monica Bay Section of the IEEE, and is an Associated Editor of the IEEE JOURNAL OF QUANTUM ELECTRONICS.