Printed Circuit Antennas with Integrated FET Detectors for Millimeter-Wave Quasi Optics

WILBERT CHEW AND HAROLD R. FETTERMAN, SENIOR MEMBER, IEEE

Abstract —**Planar twin dipole microstrip antennas with integrated FET detectors have been constructed and found to provide antenna patterns suitable for millimeter-wave quasi-optical applications. The circuits are suitable as individual elements of an imaging array. A 63 GHz heterodyne mixer using such a circuit produced a system noise temperature of 7900 K.**

I. Introduction

UTILIZATION of the millimeter-wave region demands the development of economical and practical techniques. Quasi-optical systems using unconfined beam propagation often become more practical than systems using very small waveguide components. Rapid improvements in FET's (field effect transistors) and printed circuit antennas are making them economical and effective components. FET's have demonstrated conversion gain and low noise as heterodyne mixers at lower frequencies [1]. FET detectors can be integrated with printed circuit antennas to take advantage of their qualities in millimeter-wave quasi-optical systems.

An attractive application of these circuits making use of the high resolution of millimeter waves is millimeter-wave imaging. Printed circuit antennas are the most economical way to provide enough antennas for an unscanned imaging array. In an imaging array, in contrast to the more conventional phased array, each array element operates independently and provides the signal for a separate picture element, or pixel. A large reflector or telescope can illuminate an entire quasi-optical array at once. Such an array can image a broad field of view, as optical systems can, without repositioning the telescope for each pixel. Each array element has its own feed antenna. The feed antenna need not have high gain, but should have low side lobes and a fairly symmetrical central lobe which can be efficiently covered by a conventional reflector.

A simple twin dipole antenna design inspired by previous millimeter-wave twin slot [2] and twin microstrip dipole [3], [4] designs can provide performance suitable for imaging and other quasi-optical systems. The antenna is printed on one side of the substrate, along with IF (intermediate frequency) and bias lines in a single patterning step (e.g. Fig. 1). The detector FET is mounted on the surface between the twin full-wave dipoles. The FET will normally be used as a heterodyne mixer for sensitive detection. In this study, they also operated as video detectors for easily testing the antenna patterns. The essential nonlinearity used for either video detection or heterodyne mixing in a FET is the nonlinear drain current versus gate voltage characteristic, the slope of which is the gate-voltage-dependent transconductance. The FET mixer/detector is biased near the turn-on point, where the nonlinearity is greatest, unlike a FET amplifier. In either case, the lines away from the antenna need carry only dc bias voltages and IF or video signals, avoiding the difficulties of providing a low-loss millimeter-wave transmission line.

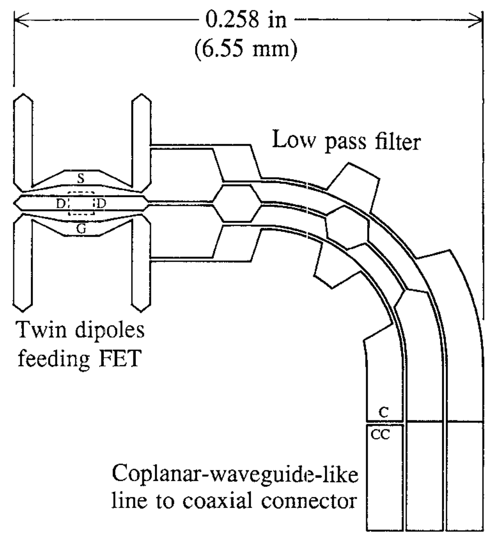


Fig. 1. Quasi-optical FET mixer/detector microstrip circuit for 0.011 in. (0.28 mm) thick PTFE/glass substrate. FET source, drain, and gate are wire-bonded to points S, D, and G. Gate bias chip capacitor is attached at point CC and wire-bonded to point C.

II. Design

The microstrip circuit of Fig. 1 was designed as a FET gate mixer on a 0.011 in. (0.28 mm) thick glass-filled PTFE (polytetrafluoroethylene) substrate. Both RF (radio frequency) signal and LO (local oscillator) inputs are sent

Manuscript received May 27, 1988; revised October 3, 1988. This work was supported by TRW and the University of California through the University of California MICRO program and by the Air Force Office of Scientific Research.
The authors are with the Department of Electrical Engineering, University of California at Los Angeles, Los Angeles, CA 90024.
IEEE Log Number 8825391.

0018-9480/89/0300-0593$01.00 ©1989 IEEE

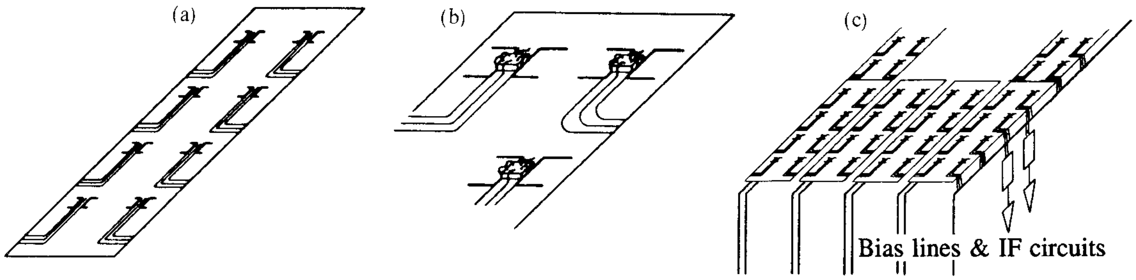


Fig. 2. Projected construction of a large two-dimensional imaging array using tested circuits. (a) Fabricate printed circuits (b) Cement FET's to substrate and wire-bond FET's to circuits. (c) Assemble substrates to form focal-plane array and make connections to IF circuits

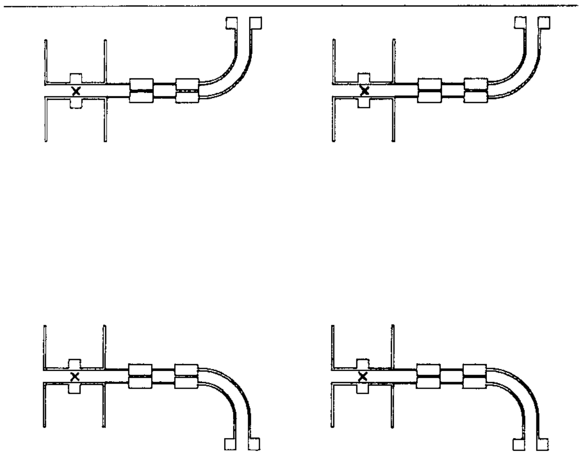


Fig 3. Four-element building block for large imaging array. Microstrip circuits for beam-lead Schottky diodes. Diodes are placed at points marked X. Lines indicating edges of substrate are 0.400 in. (10.16 mm) apart. Substrate is 0.009 in. (0.23 mm) thick fused quartz.

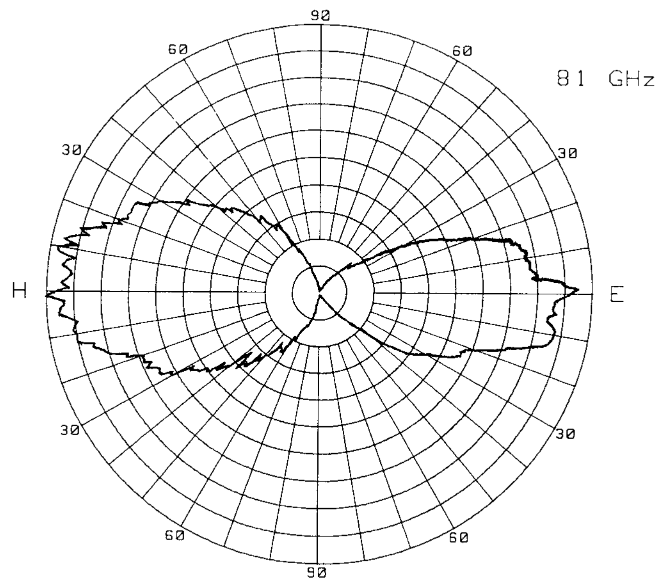


Fig. 4. Antenna pattern of twin microstrip dipoles with diodes. Video response at 81 GHz (linear voltage scale, arbitrary normalization) versus position angle. Left side: H-plane cut. Right side: E-plane cut.

through the antenna to the FET gate–source input. The FET is used to convert from the balanced dipole RF feed at its input to an unbalanced IF transmission line at the drain output. The IF transmission line is intended to work as coplanar waveguide backed by a lower ground plane [5], [6], with the lines used to bias the FET gate and source serving as the coplanar ground conductors and the line from the drain used as the center line of the coplanar waveguide. The IF/bias lines consist of cascaded quarter-wave sections for RF forming low-pass filters which present a low impedance for RF and LO at the FET output as desired for a gate mixer, but present high impedances to the dipoles to avoid disturbing the antenna performance. Effects of the IF at the FET input are minimized by operating the IF line in the coplanar waveguide mode and suppressing the balanced IF mode. The undesired balanced IF mode [6], in which the coplanar ground conductors take on opposing mode voltages instead of being at the same potential, would allow spurious low-frequency responses and IF noise. The IF ground conductor used to bias the FET gate is connected to the other ground conductors via a 100 pF chip capacitor.

Many ingenious and efficient quasi-optical mixers use too much area for a reasonably dense two-dimensional imaging array. The circuits here are compact enough to place detectors two free-space wavelengths apart in an arbitrary large array, as in Fig. 2. The practicality of such an array design was tested using beam-lead diode detectors in the circuits of Fig. 3. The diodes are similar to those described by Calviello *et al.* [7]. The measured video response versus angle for one of them at 81 GHz is shown in Fig. 4. Pattern cuts taken in the E plane and the H plane are shown on opposite sides of the same plot. The pattern is respectable despite the proximity of neighboring elements and the substrate edge.

Although diode mixers are presently better developed than FET mixers at millimeter-wave frequencies, FET mixers offer the possibility of conversion gain, at ever-increasing frequencies [8]. Also, FET's are three-terminal devices which allow some separation of input and output. For example, the diode circuits of Fig. 3 require baluns when operated as mixers with coaxial output; the FET circuit of Fig. 1 does not, because the FET separates the balanced dipole RF input from the unbalanced IF output.

III. Antenna Patterns with Integrated FET's

To study antenna problems without the problems caused by using FET's beyond their usual operating frequencies, we tested X-band (8–12 GHz) antenna models with pack-

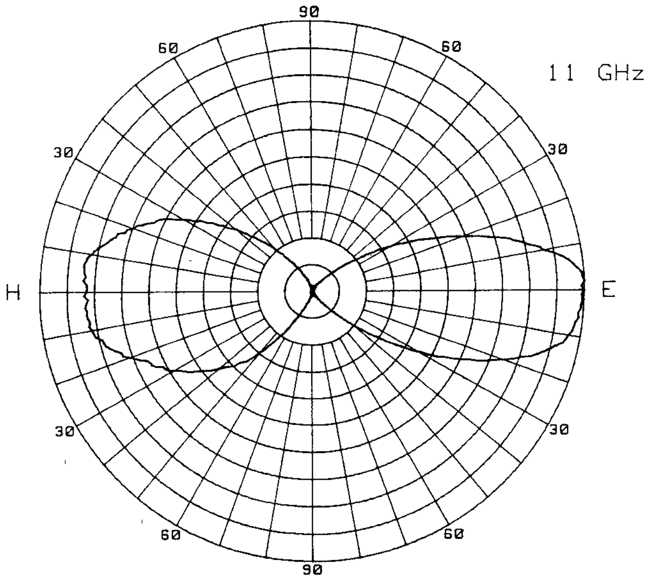


Fig. 5. Antenna pattern of *X*-band fused quartz superstrate antenna with FET. Video response at 11 GHz (linear voltage scale, arbitrary normalization) versus position angle. Left side: *H*-plane cut. Right side: *E*-plane cut.

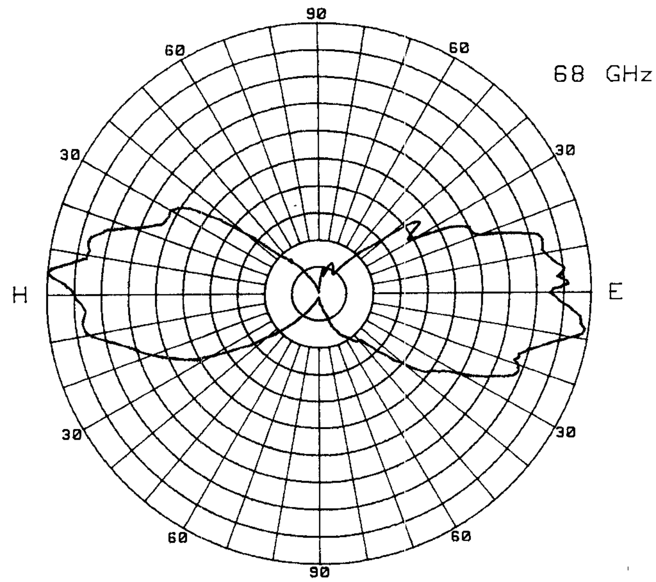


Fig. 7. Antenna pattern of millimeter-wave fused quartz microstrip antenna with FET. Video response at 68 GHz (linear voltage scale, arbitrary normalization) versus position angle. Left side: *H*-plane cut. Right side: *E*-plane cut.

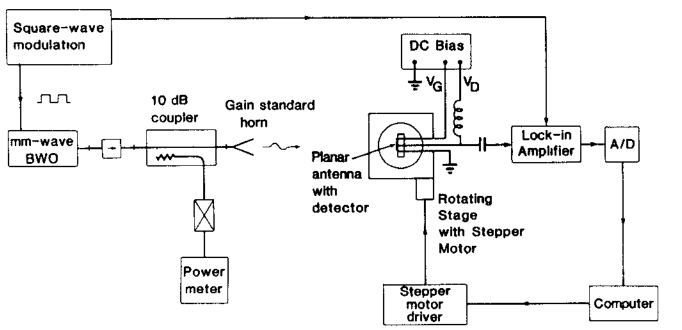


Fig. 6. Microcomputer-controlled bench-top antenna pattern measurement setup. Video response of the integrated detector is recorded as the antenna is rotated.

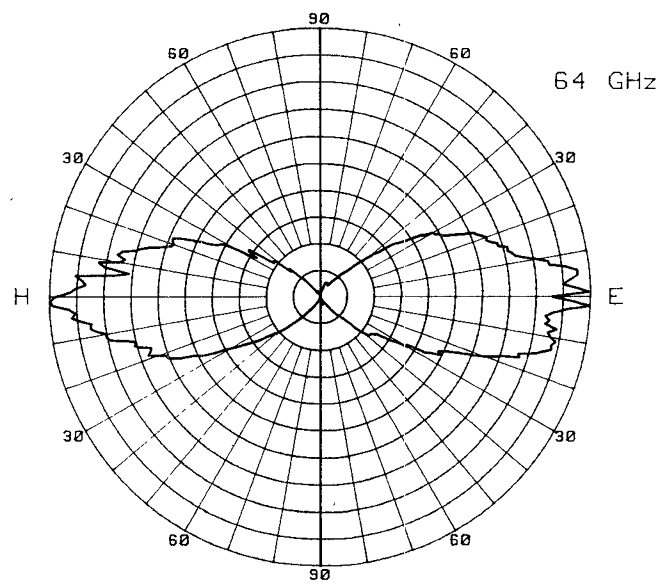


Fig. 8. Antenna pattern of millimeter-wave PTFE-glass microstrip antenna with FET. Video response at 64 GHz (linear voltage scale, arbitrary normalization) versus position angle. Left side: *H*-plane cut. Right side: *E*-plane cut.

aged FET's (NEC NE70083) before trying millimeter-wave circuits. *X*-band copper foil microstrip antennas on fused quartz and glass-filled PTFE (3M CuClad with permittivity of 2.3) substrates, and an inverted microstrip antenna on a fused quartz superstrate, were fabricated and tested.

The inverted microstrip antenna is a practical modification of a superstrate structure in which a microstrip antenna on a low-permittivity substrate is covered by a high-permittivity superstrate. Such a superstrate structure was suggested by Alexopoulos and Jackson [9] to prevent losses to substrate or surface modes. In the inverted microstrip structure, an air gap replaces the low-permittivity substrate. Air gap thickness was experimentally adjusted to 0.11 in. (2.8 mm) with a 0.050 in. (1.27 mm) thick fused quartz superstrate to obtain the antenna pattern of Fig. 5. The pattern is easily ruined if the air gap is too wide.

We measured antenna patterns using the apparatus of Fig. 6. On all three microstrip types, twin dipole antennas (with circuit patterns similar to Fig. 1) gave acceptable antenna patterns. All acceptable antenna patterns were essentially similar; the representative pattern of Fig. 5 came from the inverted microstrip structure. Video response from the FET is shown versus orientation (0° represents the direction perpendicular to the substrate surface).

Since all three *X*-band models gave acceptable patterns with FET's, we constructed 60 or 70 GHz versions of each. The inverted microstrip (superstrate) antenna proved difficult to fabricate without disturbing bond wires to the FET while fitting them within an air gap as narrow as desired for 70 GHz. The nominal FET chip thickness of 140 μm is an appreciable fraction of the desired gap of less than 400 μm. Using a wider air gap did not produce an acceptable antenna pattern. Monolithic construction could be more practical.

The other two structures gave better results. The video response antenna pattern at 68 GHz of a quartz microstrip circuit wire-bonded with an NEC NE71000 FET chip is shown in Fig. 7. The video response antenna pattern at 64 GHz of a similar PTFE-glass microstrip circuit is shown in Fig. 8. Both are suitable patterns for quasi-optical applications with conventional mirrors and lenses, having well-shaped main lobes directed perpendicular to the substrate surface, and small side lobes (on a linear scale).

IV. Mixer Results

The microstrip circuit of Fig. 1, which gave the antenna pattern of Fig. 8, was constructed on RT/Duroid 5880 (Rogers Corp., permittivity 2.2) using an NEC NE71000 FET (a widely used 0.3 μm gate length GaAs FET). The coplanar IF line was soldered to an SMA coaxial connector at the substrate edge. It was operated as a gate mixer with the gate dc bias near the turn-on point at -0.8 V and the drain bias at 0.7 V, both with respect to the grounded source.

The maximum conversion gain of a gate mixer and the LO power requirement for maximum gain can be estimated using Maas's [1], [8] simple modifications of the expressions of Pucel *et al.* [10]. Maximum conversion gain is approximately

$$G_c \approx \frac{g_1^2 R_L}{\omega_{\mathrm{RF}}^2 C^2 R_{\mathrm{in}}}$$

and the required local oscillator power for maximum gain is approximately

$$P_{\mathrm{LO}} \approx (\omega_{\mathrm{LO}} C V_{\mathrm{LO}})^2 R_{\mathrm{in}}/2$$

where g_1 is the magnitude of the fundamental component of the time-varying transconductance, ω_{RF} is the RF signal radian frequency, C is the time-averaged gate–source capacitance, R_L is the IF load impedance at the drain, R_{in} is the input resistance (including gate, source, and intrinsic resistances), ω_{LO} is the LO radian frequency, and V_{LO} is the LO voltage amplitude for maximum gain. Using $g_1 \approx G_{M\max}/4$, where $G_{M\max}$ is the maximum value of the dc transconductance, using V_{LO} as the voltage swing between the turn-on point and the point of maximum dc transconductance, and using typical values of $G_{M\max} \approx 68 \times 10^{-3}$ S, $C \approx 0.28 \times 10^{-12}$ F, $R_L \approx 55\ \Omega$, $R_{\mathrm{in}} \approx 5.1\ \Omega$, $V_{\mathrm{LO}} \approx 1$ V, and the values $\omega_{\mathrm{RF}} = 2\pi \times 63 \times 10^9$ Hz and $\omega_{\mathrm{LO}} = 2\pi \times 62 \times 10^9$ Hz, we obtain an estimated maximum conversion gain of 0.25 or a conversion loss of 6 dB, with an LO requirement of 30 mW. The orders of magnitude match the measured values, indicating that at these millimeter-wave frequencies, the basic microwave gate mixer mechanisms still apply and the quasi-optical input has reasonable efficiency.

To operate it as a mixer, LO and RF inputs were combined using a quasi-optical diplexer [11] in the quasi-optical noise measurement system of Fig. 9. This produced

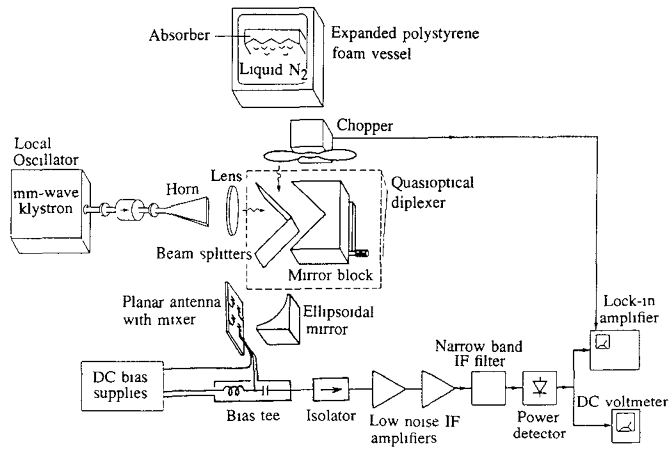


Fig. 9. Quasi-optical noise measurement system. LO (from klystron) and RF (from blackbody) are combined using a quasi-optical diplexer. System noise and system response to hot and cold blackbody radiation are measured using the mixer under test.

a directly measured system noise temperature of 7900 K with a conversion loss of 9.5 dB at an IF of 1.25 GHz and an LO of 35 mW at 62 GHz. This was a double sideband measurement made at room temperature (298 K). Since the circuit is most sensitive near 64 GHz and is much less sensitive to the lower sideband near 61 GHz, the single sideband noise figure should be nearly as low as this double sideband noise figure measurement, while the single sideband conversion loss should be almost half of this double sideband measurement. The system noise temperature includes any noise from the klystron LO, the losses from optics, diffraction, obscuration, and coupling inefficiencies, IF isolator loss, and the noise of the IF chain, in addition to the noise of the mixer circuit. Noise from the klystron LO is expected to be low, and is filtered by the quasi-optical interferometer diplexer [11]. The IF isolator and amplifier chain had a measured noise temperature of 113 K. The mixer conversion loss includes all losses and gains between the hot/cold blackbody radiator (chopper or absorber) and the input of the IF isolator. The integrated structure makes separate losses difficult to evaluate accurately without more detailed measurements. The LO power level was optimized for noise figure, and at a saturation level for conversion efficiency. Both conversion loss and noise were higher when LO power was increased to 50 mW. Though the mixer produced conversion loss rather than conversion gain, and the LO requirement is relatively high, our results show a promising noise figure. The noise measurement system constitutes a practical Dicke [12] radiometry system, so these measurements represent realistic system performance.

V. Conclusions

These results show that a FET mixer integrated with a simple printed circuit antenna can provide useful performance at 63 GHz. The circuits are suitable for imaging and other quasi-optical applications. The results were obtained using a popular commercial FET wire-bonded to the circuit. Further advances in performance may be expected with custom optics, refinements in circuit design, beam-lead devices or monolithic approaches, and relentless advances in available FET's.

ACKNOWLEDGMENT

The authors wish to thank J. A. Calviello of the AIL Division of Eaton Corporation for providing beam-lead diodes and W. W. Ho of Rockwell International Science Center for providing a low-noise millimeter-wave local oscillator for noise measurements.

REFERENCES

[1] S. A. Maas, *Microwave Mixers.* Dedham, MA: Artech House, 1986.

[2] A. R. Kerr, P. H. Siegel, and R. J. Mattauch, "A simple quasi-optical mixer for 100–120 GHz," in *1977 IEEE MTT-S Int. Microwave Symp. Dig.*, pp. 96–98.

[3] P. T. Parrish *et al.* "Printed dipole-Schottky diode millimeter wave antenna array," *SPIE Proc. Millimeter Wave Technology*, vol. 337, pp. 49–52, 1982.

[4] H. R. Fetterman *et al.*, "Printed dipole millimeter wave antenna for imaging array applications," *Electromagentics*, vol. 3, pp. 209–215, 1983.

[5] C. P. Wen, "Coplanar waveguide: A surface strip transmission line suitable for nonreciprocal gyromagnetic device applications," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-17, pp. 1087–1090, Dec. 1969.

[6] M. Riaziat, I. J. Feng, R. Majidi-Ahy, and B. A. Auld, "Single-mode operation of coplanar waveguides," *Electron. Lett.*, vol. 23, pp. 1281–1283, 19 Nov. 1987.

[7] J. A. Calviello, J. L. Wallace, and P. R. Bie, "High-performance GaAs beam-lead mixer diodes for millimetre and submillimetre applications," *Electron. Lett.*, vol. 15, pp. 509–510, 16 Aug. 1979.

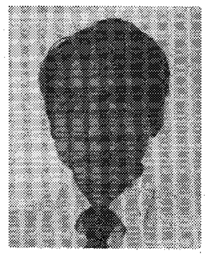
[8] S. A. Maas, "Design and performance of a 45-GHz HEMT mixer," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-34, pp. 799–803, July 1986.

[9] N. G. Alexopoulos and D. R. Jackson, "Fundamental superstrate (cover) effects on printed circuit antennas," *IEEE Trans. Antennas Propagat.*, vol. AP-32, pp. 807–816, Aug. 1984.

[10] R. A. Pucel, D. Masse, and R. Bera, "Performance of GaAs MESFET mixers at X band," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-24, pp. 351–360, June 1976.

[11] J. M. Payne and M. R. Wordeman, "Quasi-optical diplexer for millimeter wavelengths," *Rev. Sci. Instru.*, vol. 49, pp. 1741–1743, Dec. 1978.

[12] R. H. Dicke, "The measurement of thermal radiation at microwave frequencies," *Rev. Sci. Instr.*, vol. 17, pp. 268–275, July 1946.



Wilbert Chew received the B.S. degree in applied physics from the California Institute of Technology in 1978 and the M.S.E.E. degree in electrical engineering from the University of Washington in 1980.

From 1981 to 1983 he analyzed infrared imaging systems at Hughes Aircraft Company. He is currently a research assistant and Ph.D. student in electrical engineering at the University of California, Los Angeles, studying solid-state electronics and millimeter-wave systems and components.

Harold R. Fetterman (SM'81) received the Ph.D. degree from Cornell University in 1967.

He is currently a Professor in the Department of Electrical Engineering and Associate Dean of the School of Engineering and Applied Science at the University of California, Los Angeles. He joined UCLA after 14 years at the MIT Lincoln Laboratory, where he was Project Leader of the submillimeter/millimeter-wave detector and source programs. He successfully directed the development of heterodyne receivers and solid-state sources with applications in plasma diagnostics, remote sensing, and imaging radars. Since coming to UCLA, he has been active in the development of millimeter-wave FET devices and large-area arrays, and established the new UCLA High Frequency Electronics Center.