W-Band (75–110 GHz) Microstrip Components

KAI CHANG, SENIOR MEMBER, IEEE, DAN M. ENGLISH, RAGHBIR S. TAHIM, MEMBER, IEEE, ALBERT J. GROTE, THANG PHAM, CHENG SUN, MEMBER, IEEE, GEORGE M. HAYASHIBARA, PERCY C.H. YEN, AND WIESLAW S. PIOTROWSKI, MEMBER, IEEE

Abstract —**Various microstrip components, including mixers, IMPATT oscillators, Gunn oscillators, doublers, circulators, and IMPATT amplifiers, have been developed at *W*-band with state-of-the-art performance. The use of microstrip drastically reduces fabrication costs due to the less stringent machining tolerance. The design and performance of these components will be reported.**

I. INTRODUCTION

MILLIMETER-WAVE components have been reported and used in many system applications. Most approaches use waveguide, suspended stripline, and finline. Fabricating these components at *W*-band generally requires stringent tolerance and is expensive; microstrip can alleviate this problem because no critical machining is needed in its fabrication. The cost saving in large quantity production can be 5 to 10 times compared with waveguide. This paper reports many active and passive components fabricated in microstrip medium on Duroid substrate.

II. RAT-RACE BALANCED MIXERS

Microstrip rat-race mixers have been reported at *W*-band for very narrow-band operation [1]–[3]. The rat-race mixer (Fig. 1) consists of a ring-type power splitter, two mixer diodes, two RF chokes, and a low-pass filter. The LO input is split equally into two mixer diodes. The RF input is also split equally, but 180° out of phase at the mixer diodes. The LO and RF are mixed in these diodes which generate signals that are taken out through a low-pass filter. The RF choke provides the tuning mechanism and prevents the RF signal feeding into ground.

The dimensions of the ring are critical. It is 1.5 wavelength in circumference with four arms separated by 60° of angular rotation. Two input and output arms are spaced from one another. At the center frequency, the input power from arm A will split equally into arms B and D. Because of the length of the electric path, the phase relationship between arms B and D will be 180° out of phase. For the power input from arm C, the output will split equally in phase to arms B and D. The design of the ring requires that its impedance be equal to $\sqrt{2}$ times the characteristic impedance of each arm. For a 50-Ω system, this main ring impedance is equal to 70.7 Ω.

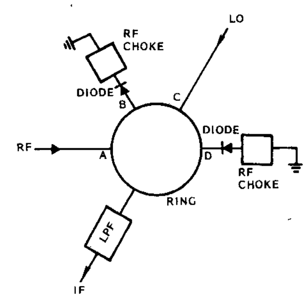

Fig. 1. Rat-race mixer and ring.

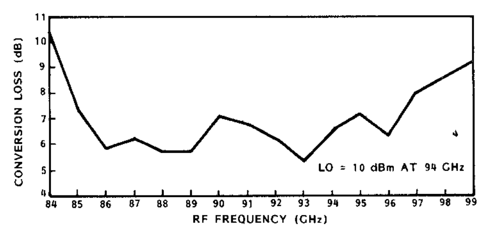

Fig. 2. Performance of *W*-band rat-race microstrip mixer.

A five-stage low-pass Chebyshev filter with 0.1-dB passband ripple is used in the mixer. The high-impedance line in the filter is 100 Ω and the low-impedance line is 20 Ω. The length of each section was optimized by the computer.

By carefully designing the ring size and the RF and IF matching, the bandwidth was improved considerably. A conversion loss of less than 7 dB was achieved for an instantaneous RF bandwidth of 9 GHz, as shown in Fig. 2. The circuit was built on 5-mil Duroid substrate. Broad-band finline-to-microstrip transitions were connected to the mixer for RF and LO power coupling during mixer testing. The transition has an insertion loss of less than 0.5 dB over a 35-GHz bandwidth. The mixer without the transitions is shown in Fig. 3.

Optimization was also carried out to lower the LO pump because a higher power LO drive is difficult to obtain at *W*-band; thus, it is always desirable to lower the LO drive to the mixer. With the external bias, the LO drive can be substantially lower while maintaining a reasonable conversion loss.

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K. Chang was with TRW Electronics and Defense. He is now with the Electrical Engineering Department, Texas A&M University, College Station, TX 77843.

D. M. English, R. S. Tahim, A. J. Grote, T. Pham, C. Sun, G. M. Hayashibara, P. C. H. Yen, and W. S. Piotrowski are with TRW Electronics and Defense, One Space Park, Redondo Beach, CA 90278.

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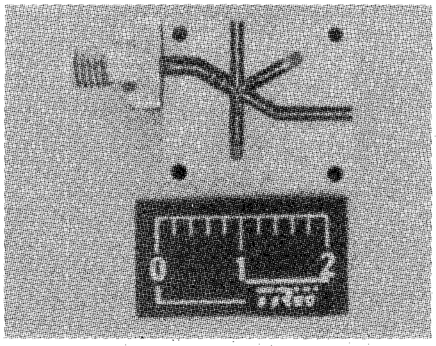


Fig. 3. W-band microstrip rat-race mixer.

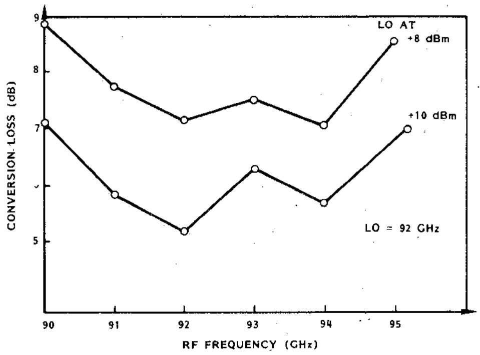


Fig. 4. Mixer performance without external bias.

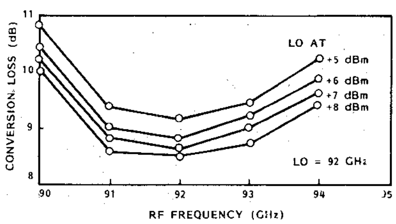


Fig. 5. Mixer performance with external bias.

A bias scheme has been devised to bias the mixer diodes. Fig. 4 shows the performance of the mixer without the external bias. The conversion loss increases very rapidly when the LO drive is below +8 dBm, which is out of scale and not shown in the figure. With the external bias, the LO drive can be as low as +5 dBm while still maintaining the conversion loss below 10 dB. The results are shown in Fig. 5.

III. IMPATT and GUNN Oscillators

Little work was reported on active microstrip components operating at W-band [4]–[7], and the output power was generally very low compared with their waveguide counterpart. By further optimizing the circuit reported in [4] and [7], better power outputs for IMPATT oscillators were achieved.

As shown in Fig. 6, the CW IMPATT oscillator consists of an IMPATT diode mounted on a 50-Ω microstrip line, coupled to a 50-Ω half-wavelength resonator with a high-impedance T-junction that transfers the RF power to a

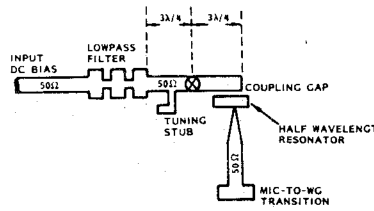


Fig. 6. Circuit schematic of IMPATT oscillator.

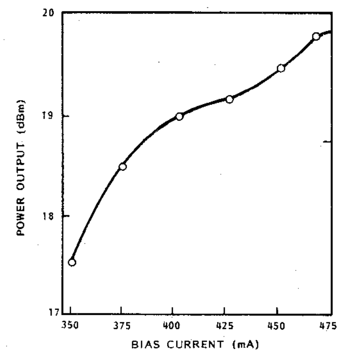


Fig. 7. CW IMPATT oscillator performance.

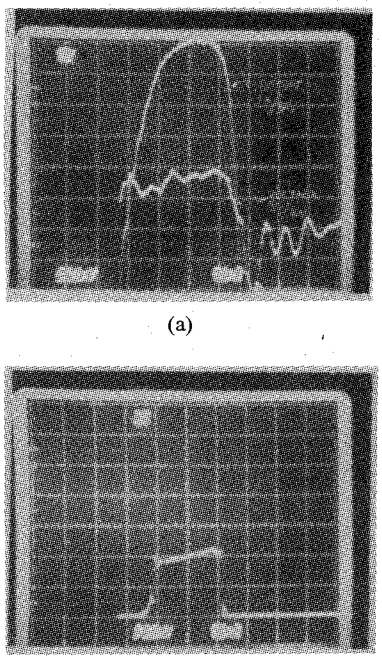
microstrip-to-waveguide transition. The circuit was built on Duroid substrate.

The IMPATT diode is mounted along the input 50-Ω line of three-quarter wavelengths at the desired frequency from the output end. DC bias is provided from the other end of the line. The connection from the bias side of the 50-Ω line to the $3\lambda/4$ resonator is completed by soldering a gold ribbon to each side of the circuit and the IMPATT diode. To suppress subharmonic oscillations, a quarter-wavelength stud at $f_0/2$ is placed in front of the IMPATT diode. The first low-impedance section of the low-pass filter in the dc bias line is located about a $3\lambda/4$ distance away from the diode center which, consequently, presents an open circuit to the diodes.

The gap between the $3\lambda/4$ and $\lambda/2$ resonators and the location of the T-junction determine the correct amount of output coupling as shown in Fig. 6. The T-junction tap point is most critical, as is also the impedance value. The high-impedance line is then tapered back up to 50 Ω over at least one wavelength. Fig. 7 shows the performance of this W-band oscillator. Over 90 mW at 100 GHz has been achieved, representing state-of-the-art performance using microstrip oscillators. At 94 GHz, over 100-mW output power was consistently achieved.

For pulsed operation, a peak output power of 5 W at 92.75 GHz was achieved with a 50-kHz pulse repetition rate and 100-ns pulsewidth using double-drift IMPATT diodes with a capacitance of 6 to 7 pF and a reverse breakdown voltage of 14. Fig. 8 shows the bias current/voltage and video output of the RF signal.

A similar circuit was used for Gunn oscillators. Typical power output is between +8 and +11 dBm. With a varactor incorporated, a tuning range of 800 MHz was achieved with over +8-dBm output power; the results are shown in Fig. 9.



(a)

(b)

Fig. 8. Performance of a *W*-band IMPATT oscillator pulsed operation. (a) Top trace for current and bottom trace for voltage. (b) RF output signal.

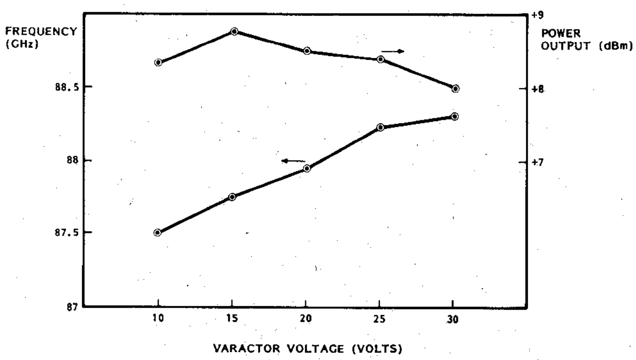


Fig. 9. Performance of *W*-band microstrip Gunn VCO.

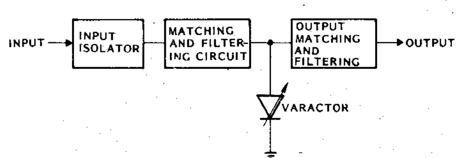


Fig. 10. Block diagram of microstrip multiplier.

IV. Frequency Doubler

The frequency multiplier is a viable approach to achieve low-noise, stable, high-frequency signals. All frequency multipliers reported at this frequency range use waveguide or suspended stripline [8]–[10]. A 46–92-GHz microstrip doubler is presented here with efficiency comparable to the suspended stripline multipliers.

Fig. 10 shows the block diagram of the doubler. It consists of a dc block, an input matching and filtering circuit, a varactor diode, and an output matching and

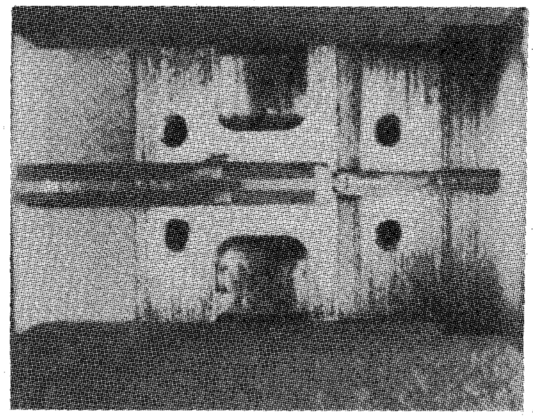


Fig. 11. Photograph of 46–92-GHz microstrip doubler.

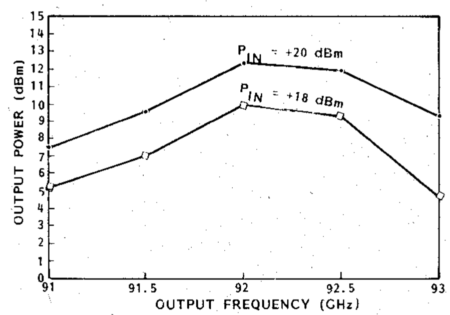


Fig. 12. Performance of 46–92-GHz microstrip doubler.

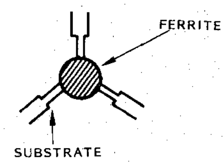


Fig. 13. Circulator circuit configuration.

filtering circuit. Multiplication is accomplished with a GaAs varactor diode, which has a cutoff frequency of about 500 GHz. The design began with the calculation of the diode impedance at input and output frequencies. Matching networks were then designed to match these impedance levels.

Fig. 11 shows the actual hardware; performance is summarized in Fig. 12. The conversion loss is about 8 to 9 dB over a 500-MHz output bandwidth.

V. Circulators

Circulators are important components for constructing oscillators and amplifiers using two terminal devices. At *W*-band, the ferrite disk becomes too small to support the lowest order mode; therefore, a relatively large ferrite supporting higher order modes is preferred to ensure producibility, repeatability of performance, and reasonable mechanical tolerance. The penalty for operating at higher order modes is the slightly higher insertion loss.

As a dielectric resonator, the junction ferrite supports the propagation of a mode, controlled by the $X_{1,m}$ root of the Bessel function $J'(x)$, and the ferrite radius is de-

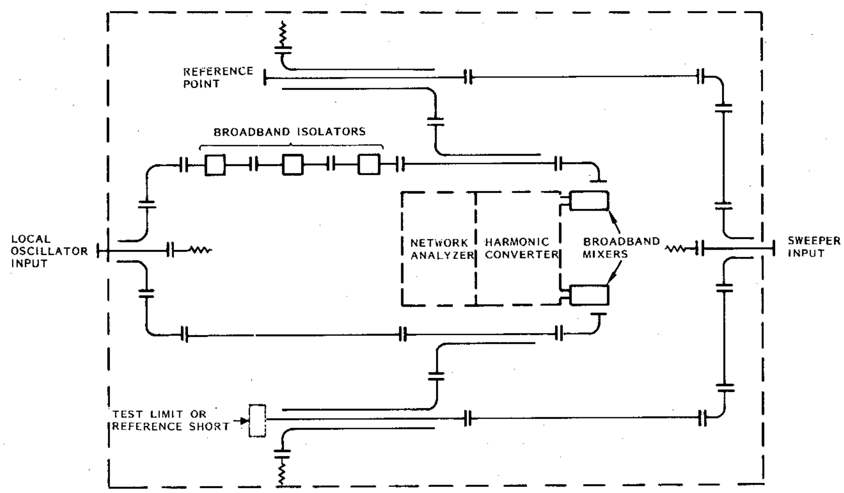


Fig. 14. TRW W-band network analyzer schematic.

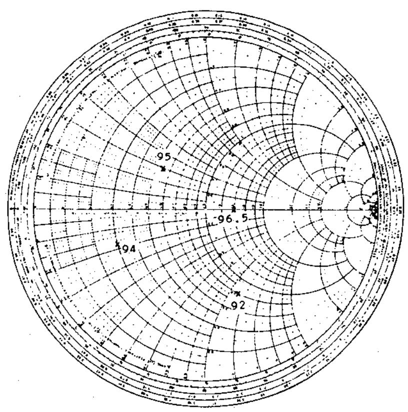


Fig. 15. Circulator impedance measurement.

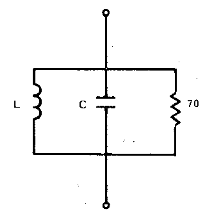


Fig. 16. Equivalent circuit of microstrip ferrite circulator.

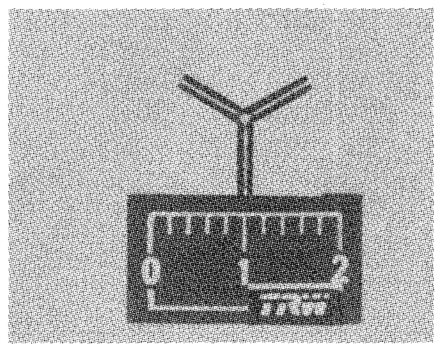


Fig. 17. Microstrip circulator with impedance quarter-wavelength transformer.

termined from

$$R = \frac{X_{1,m}}{\left[(2\pi/\lambda_0)^2 \epsilon_r\right]^{1/2}}$$

where λ_0 is the free-space wavelength at the design frequency and ϵ_r is the relative dielectric constant of the ferrite.

The calculated diameters for different modes are

fundamental mode:	$X_{1,m} = 1.8$, $2R = 0.021$ in
first higher order mode:	$X_{1,m} = 3.05$, $2R = 0.036$ in
second higher order mode:	$X_{1,m} = 3.83$, $2R = 0.044$ in.

The ferrite disk operating at the first higher order mode was selected with a diameter of 0.036 in. This large disk size has the advantage of less stringent mechanical tolerances and can thus be produced at lower cost. Fig. 13 shows the circuit configuration.

Another important analysis was to determine the material to be used for the ferrite disk. C-48 material was chosen because of its demonstrated performance in waveguide; Duroid 5880 substrate material was selected for its low loss and handling ease.

The matching circuit design was based on impedance measurements using the network analyzer (Fig. 14). The results of this impedance measurement for several frequencies are plotted on a Smith chart as shown in Fig. 15. It can be seen that the circuit resonance occurs at 96.5 GHz, which is slightly higher than the designed operating frequency of 94 GHz. The center frequency can be easily adjusted by using a puck with a 2-percent larger diameter. At the resonance, the ferrite puck becomes a pure resistance of 70 Ω. The equivalent circuit of the ferrite circulator can be as shown in Fig. 16 with a LC resonant circuit in parallel with a resistance. The circulator provides the highest isolation and lowest insertion loss at the resonance.

Based on this measurement, a matching circuit was designed to match the 50-Ω line impedance to the ferrite impedance. The circuit is shown in Fig. 17. The circulator has an insertion loss of less than 1.5 dB and an isolation more than 20 dB over a 2-GHz bandwidth (Fig. 18). No

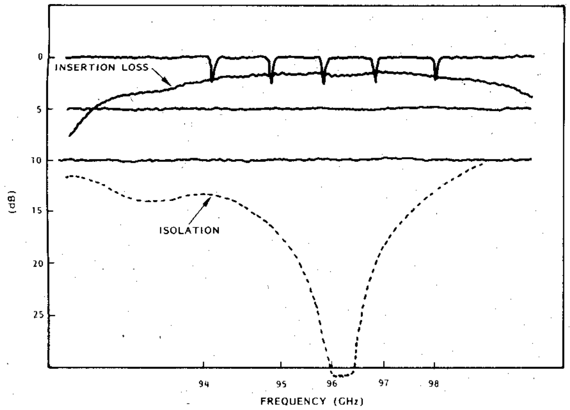


Fig. 18. Insertion and isolation measurement of microstrip circulator.

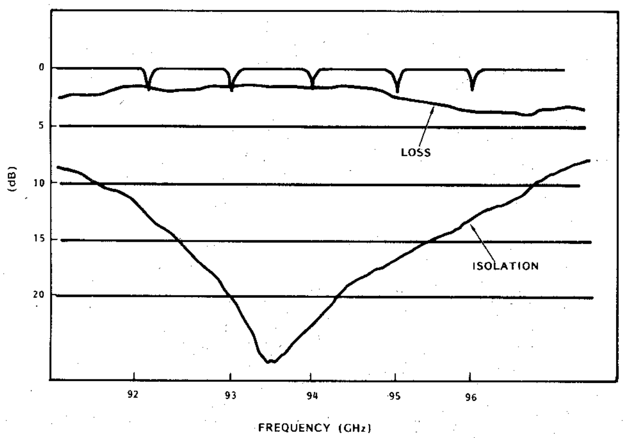


Fig. 19. Performance optimized at 94 GHz.

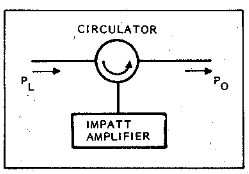


Fig. 20. Block diagram of injection-locked amplifier.

external tuning stubs were required in the present circuit to improve the matching.

The center frequency of the circulator can be adjusted by using slightly different sizes of ferrite disks. For example, an operating frequency of 94 GHz was achieved using a ferrite disk with a diameter of 0.0375 in. Fig. 19 shows the performance of this circulator.

VI. IMPATT Amplifier

The IMPATT oscillator can be connected to a microstrip circulator forming an injection-locked amplifier (Fig. 20). Since the source is to be injection-locked, the coupled-line microstrip circuit must be designed to allow for the effects of injection-locking on the real input impedance seen by the IMPATT diode. This means that the output coupling coefficient must be increased according to the following

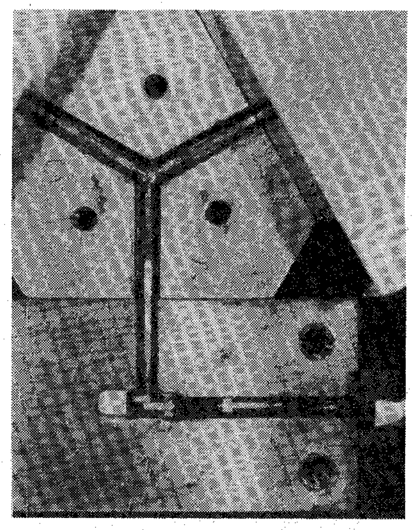


Fig. 21. *W*-band all-microstrip IMPATT amplifier.

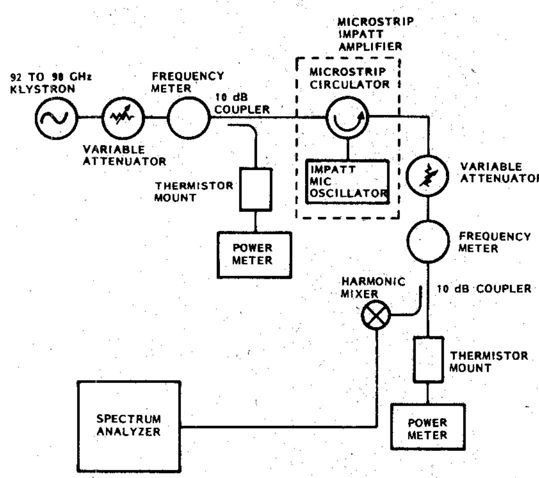


Fig. 22. Injection-locking measurement system.

equation:

$$B = \frac{Z_0}{Z_H}\left(\frac{2Q_0}{\pi}\right)\cos^2\theta$$

where

- B output coupling coefficient,
- Z_0 line impedance,
- Z_H tap point impedance,
- Q_0 unloaded quality factor,
- θ distance to tap point.

The equation indicates that the location of the tap point must be moved toward the input end of the half-wave resonator. This circuit modification allows the frequency to swing through a wide range and achieve wide locking bandwidth.

The oscillator was integrated with the microstrip circulator to form an amplifier. Fig. 21 shows the microstrip IMPATT amplifier. At the top is the microstrip circulator with the input and output ports coupled to the transitions. At the bottom is the microstrip IMPATT oscillator.

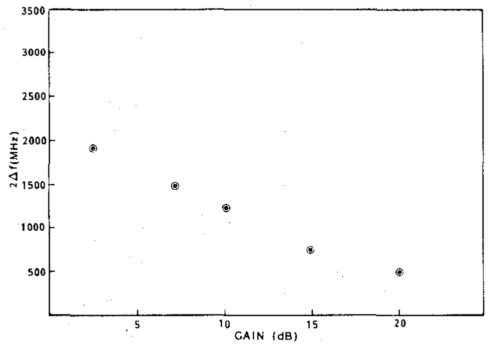
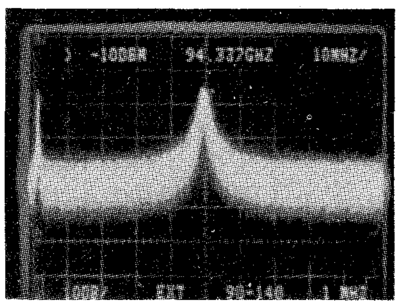
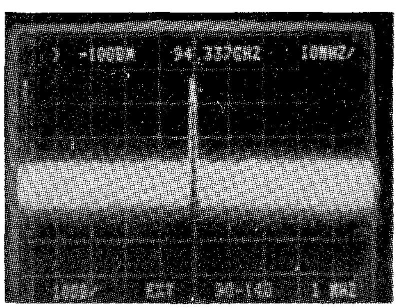


Fig. 23. Injection-locking bandwidth ($2\Delta f$) as a function of power gain (P_0/P_L).



(a)



(b)

Fig. 24. Spectrums of locked and unlocked IMPATT amplifier. (a) Free-running spectrum. (b) Injection-locked spectrum.

Using the measurement setup shown in Fig. 22, some preliminary injection locking data were obtained. The results are shown in Fig. 23. As the graph shows, a 500-MHz locking bandwidth was achieved at 20-dB locking gain and 1900 MHz at 3-dB gain. With these data and the following equation, the external Q of the circuit can be calculated [11]:

$$\frac{2\Delta f}{f_0} = \frac{2}{Q_{\mathrm{ext}}}\left(\frac{P_0}{P_L}\right)^{-1/2}$$

where $2\Delta f$ is the total locking bandwidth, and f_0 is the free-running frequency. P_0 is the free-running power output and P_L is the locking signal power. From Fig. 23, the external Q of approximately 37 was calculated. The high circuit Q is believed due to the built-in resonant stub.

The power variation was less than 2 dB for every value of input power and the maximum output power was +17 dBm. Fig. 24(a) shows the free-running spectrum of the IMPATT amplifier. After the application of the locking signal at an injection-locking gain of 20 dB, the spectrum exhibits excellent low-noise characteristics, as shown in Fig. 24(b).

VII. Conclusions

Various W-band microstrip components have been developed with state-of-the-art performance. These components can be fabricated at very low cost and can be used as building blocks for many systems.

Acknowledgment

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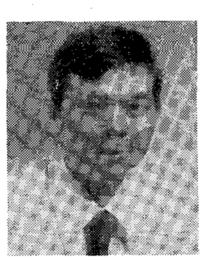


Kai Chang (S'75–M'76–SM'85) was born in Canton, China, on April 27, 1948. He received the B.S.E.E. degree from National Taiwan University, Taipei, Taiwan, the M.S. degree from State University of New York at Stony Brook, and the Ph.D. degree from the University of Michigan, Ann Arbor, in 1970, 1972, and 1976, respectively.

From 1972 to 1976, he worked for the Microwave Solid-State Circuits Group, Cooley Electronics Laboratory of the University of Michigan as a Research Assistant. From 1976 to 1978, he was employed by Shared Applications, Ann Arbor, where he worked in microwave circuits, microwave radar detectors, and microwave tubes. From 1978 to 1981, he worked for the Electron Dynamic Division, Hughes Aircraft Company, Torrance, CA, where he was involved in the research and development of millimeter-wave devices and circuits. This activity resulted in state-of-the-art IMPATT oscillator and power com-

biner performance at 94, 140, and 217 GHz. Other activities included silicon and gallium arsenide IMPATT diode design and computer simulation, Gunn oscillator development, and monopulse comparator and phase shifter development. In May 1981, he joined TRW Electronics and Defense, Redondo Beach, CA, as a Section Head in the Millimeter Wave Technology Department, developing state-of-the-art millimeter-wave integrated circuits and subsystems. In August 1985, he joined the Electrical Engineering Department of Texas A&M University as an Associate Professor.

Dan M. English is currently pursuing the B.S.E.E. degree at El Camino College, Torrance, CA.

From 1974–1979, he worked in the field of numerical machine tool technology as a Production Manager. In November 1979, he joined Hughes Aircraft Company, Electron Dynamics Division, where he was involved in the development of millimeter-wave solid-state CW and pulsed oscillators and stable amplifiers. In March 1982, he joined TRW Defense and Space Systems Sector, Redondo Beach, CA, as an Associate Engineer in the Millimeter Wave Technology Department. Currently, he is working with millimeter-wave integrated circuits.

Raghbir S. Tahim was born at Sarhali Jullundur, Punjab, India. He received the B.Sc. Eng. degree in electronics from Punjab Engineering College Chandigarh, India, in 1967 and the M.Sc. degree from the University of Bradford, England, in 1969 for his work on image canceling mixers.

Since 1970, he has engaged in research and development in microwave and millimeter-wave circuits and has contributed to the development of a wide range of microwave components and subsystems. From 1977 to 1981, he was with Hughes Aircraft Company, where he was responsible for the development of low-cost broad-band millimeter-wave mixers and frequency doublers. He also participated in the development of low-noise Gunn VCO's, step recovery, and varactor diode multiplier chains. Currently, he is with the Millimeter Wave Technology Department of TRW Electronics and Defense, Redondo Beach, CA, where he is engaged in the development of state-of-the-art millimeter-wave integrated circuits and subsystems.

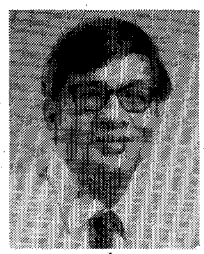
Mr. Tahim is the author and coauthor of over ten technical papers.

Albert J. Grote was born in Boston, MA, on September 18, 1928. He was educated at Northrup Institute.

He joined TRW in 1959 and presently is a Member of the Technical Staff. He has 30 years of experience in the design and development of frequency multipliers, oscillators, receivers, and transmitters in the VHF, UHF, and millimeter-wave frequency bands. His current interests are concerned with millimeter-wave integrated circuits.

Thang Pham has been an active participant in the development of microwave and millimeter wave circuits since 1980. He joined the Advanced Millimeter Wave System Group at TRW in 1982. Prior to that he was with the Millimeter Wave group at Hughes Aircraft Company. He has been involved with computer-aided circuit layouts and has contributed to the development of low-noise mixers and upconverters.

Mr. Thang Pham was born in Vietnam. Since his arrival in the United States, in 1975, he has been working actively towards the B.S. degree in electrical engineering.



Cheng Sun (S'63–M'65) received the B.S. degree in electrical engineering from National Taiwan University, Taipei, in 1958, and the M.S. and Ph.D. degrees from Cornell University, Ithaca, NY, in 1962 and 1965, respectively.

In 1964, he joined the RCA Corporation in development of various microwave solid-state sources and laser devices. From 1971 to 1980, he was with the Hughes Aircraft Company, Electron Dynamics Division, Torrance, CA, where he supervised a group engaged in the research and development of millimeter-wave circuits. Since 1980, he has been with TRW Defense and Space Systems Group, Redondo Beach, CA. He is now the manager in the Millimeter-Wave Technology Department, responsible for the development of a wide range of solid-state circuits and components in the EHF frequency range (20–300 GHz).

Dr. Sun is a member of Sigma Xi.

George M. Hayashibara was born in Long Beach, CA, on October 26, 1955. He received the B.S. degree from California State Polytechnic University, Pomona, in 1978. Currently, he is pursuing the M.S. degree in microwave systems engineering at the University of Southern California, Los Angeles.

From 1978 to 1981, he worked at Hughes Aircraft Company, Electron Dynamics Division, where he was involved in the development of millimeter-wave solid-state oscillators and power combiners. In August 1981, he joined TRW Defense and Space Systems Sector, Redondo Beach, CA, as a Member of the Technical Staff in the Millimeter Wave Technology Department. His current interests are concerned with millimeter-wave integrated circuits.

Percy C. H. Yen was born in Shanghai, China, in 1942, and received the B.S. degree from Cheng Kung University, Tainan, Taiwan, and the Ph.D. degree from the University of Pennsylvania, Philadelphia, both in electrical engineering, in 1964 and 1968, respectively. From 1969 to 1973, he had post-doctoral positions at Columbia, Rockefeller, and Columbia Universities with a primary interest in the area of RF systems and instrumentation.

His first industrial position was at Bunker-

Ramo Corporation, Westlake, CA, where he was engaged in the development of parametric amplifiers and ferrite devices. Later, he was employed by Transco Products, Inc., Venice, CA, working on switches and coplanar antennas, and by TRW, Redondo Beach, CA, working on RF transmitter and receiver design. In 1978, he worked at Ford Aerospace and Communication Corporation, Newport Beach, CA, working on the design of IMPATT oscillators and combiners. From 1979 to 1982, he was with Electron Dynamics Division, Hughes Aircraft Company, Torrance, CA, and involved in developing mixers, VCO's, dielectric guide components and integrated circuit, all in the millimeter-wave region. Since February 1982, he has been with TRW, Redondo Beach, CA, and engaged in the development of millimeter-wave integrated-circuit components and subsystems.

Wieslaw S. Piotrowski (M'63) was born in Zabno, Poland, on October 19, 1919. He received the B.S. degree in mechanical engineering and business and administration from Northeastern University, Boston, MA, in 1955 and 1957, respectively. He also received the B.S.E.E. degree from California State University, Los Angeles, CA, in 1960.

He has held a number of positions in the radar and communications industry before joining TRW Electronics and Defense Systems in 1972. During the past 13 years, he has been involved in the millimeter-wave ferrite components and ferrite materials technology.