microwave JOURNAL

INTERNATIONAL EDITION □ VOL. 29, NO. 11 □ NOVEMBER 1986

- DRO coupling coefficients
- Hybrid ring directional coupler
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- Matched square waveguide corner

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MICROWAVE PACKAGING

horizon house

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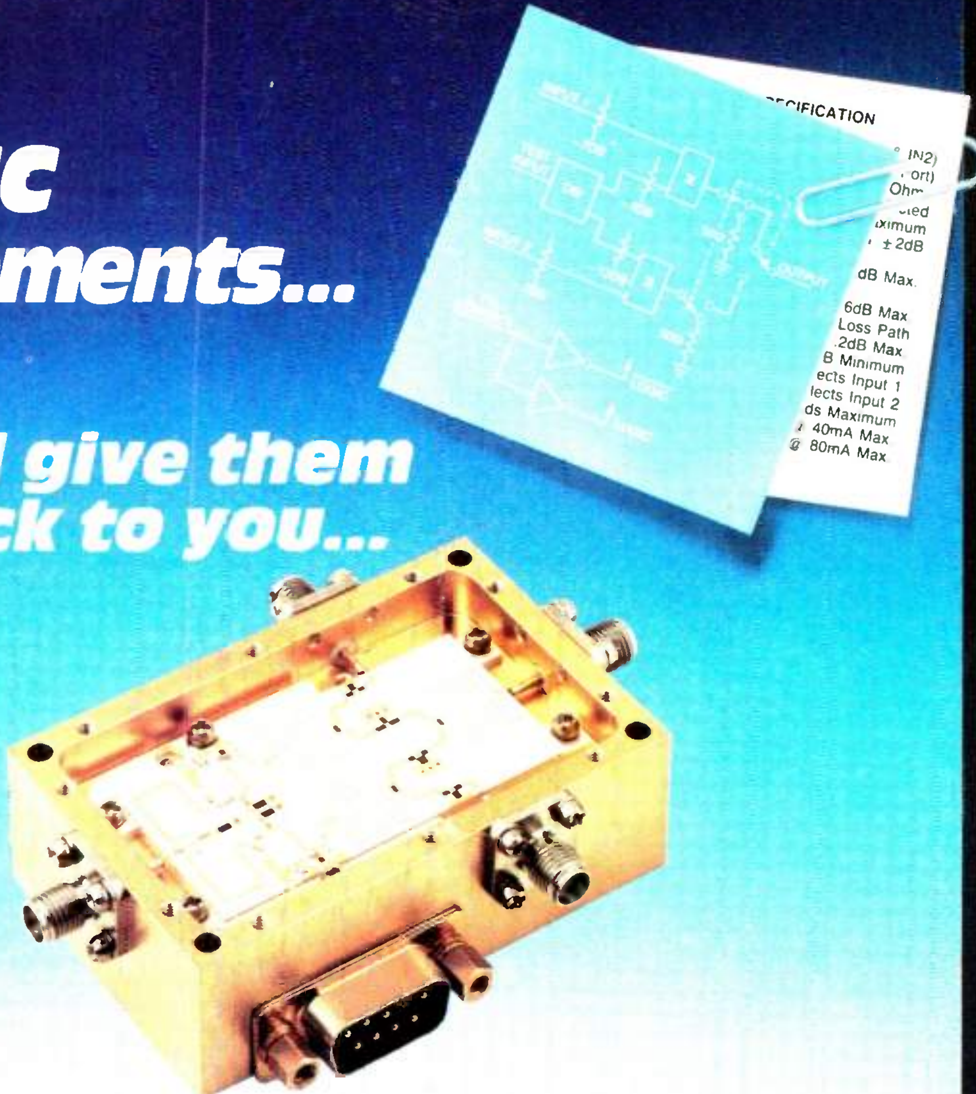
and we'll give them right back to you...



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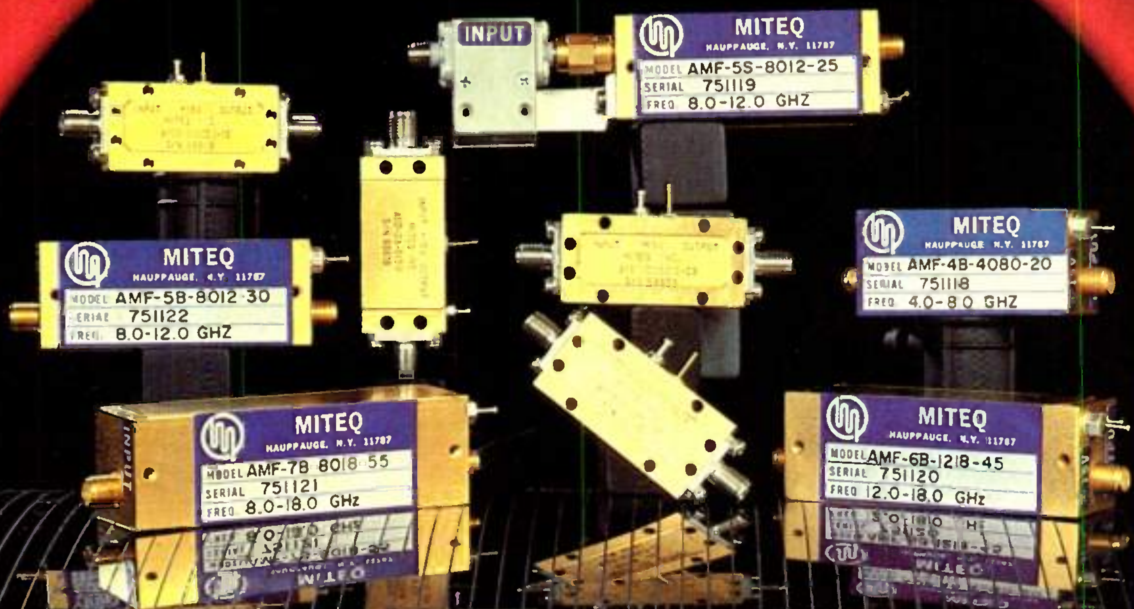

Once again, **MITEQ** is setting new standards of excellence in the performance of wideband low-noise amplifiers.

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AUD-2A-0150-13	.001-.5	30	.5	1.3	2:1	2:1	+10	25
AFD3-005010-09	.5-1.0	35	.5	0.9	2:1	2:1	+15	100
AFD3-010020-10	1.0-2.0	35	.5	1.0	2:1	2:1	+15	100
AFD3-020040-13	2.0-4.0	32	1.0	1.3	2:1	2:1	+15	100
AMF-4B-4080-20	4.0-8.0	35	1.5	2.0	2:1	2:1	+15	150
AMF-5S-8012-25	8.0-12.0	30	1.5	2.5	1.5:1	2:1	+15	225
AMF-7B-1218-45	12.0-18.0	30	1.5	4.5	2:1	2:1	+15	300
AMF-7B-8018-45	8.0-18.0	30	1.5	4.5	2:1	2:1	+15	300

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F82-2 REV B

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Max. 20dB Stop Frequency (MHz)	19	70	90	147	210	290	410	580	750	840	1000	1100	1340

Prices (ea.): P $9.95 (6-49), B $24.95 (1-49), N $27.95 (1-49), S $26.95 (1-49)

HIGH PASS Model *HP-		50	100	150	200	300	400	500	600	700	800	900	1000
Pass Band (MHz)	start, max.	41	90	133	185	290	395	500	600	700	780	910	1000
	end, min.	200	400	600	800	1200	1600	1600	1600	1800	2000	2100	2200
Min. 20dB Stop Frequency (MHz)		26	55	95	116	190	290	365	460	520	570	660	720

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*Prefix P for pins, B for BNC, N for Type N, S for SMA example: PLP-10.7

F82-2 REV 3

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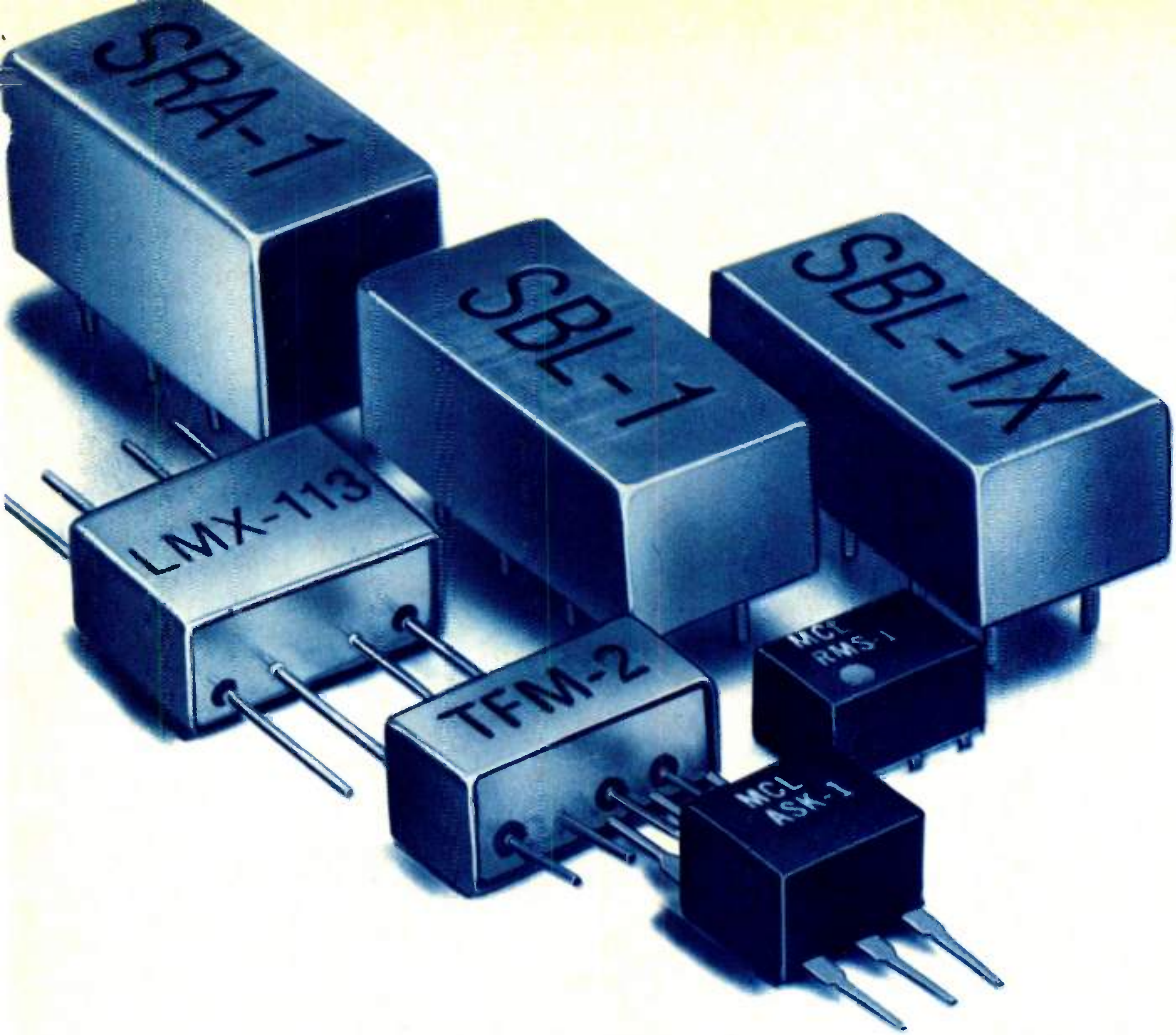

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TFM-2*	1-1000	6.0	40	11.95	(6-49)
SBL-1	1-500	5.5	45	4.50	(10-49)
				3.95	(100)
SBL-1X	10-1000	6.0	40	5.95	(10-49)
ASK-1	1-600	5.5	35	5.95	(10-49)
LMX-113*	5-1000	6.5	40	14.95	(6-24)
RMS-1	0.5-500	5.5	33	6.95	(10-49)

*units are not QPL listed
for testing and screening to MIL-M-28837/1A add /MIL to model no.
Also add $3 to unit price for SRA-1, $6 for all other models

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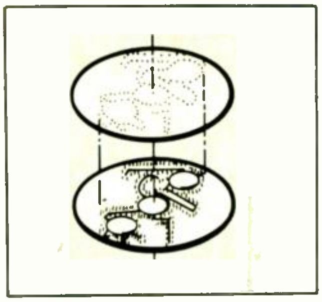

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ON THE COVER

Package design must keep pace with the expanding capabilities of microwave devices and integrated circuits to permit those capabilities to be fully exploited by system designers. Contributing Editor Howard Bierman reviews the state of microwave package design in his special report, beginning on p. 26.

*Euro-Global edition only

Press run for this issue is 53,260 copies.

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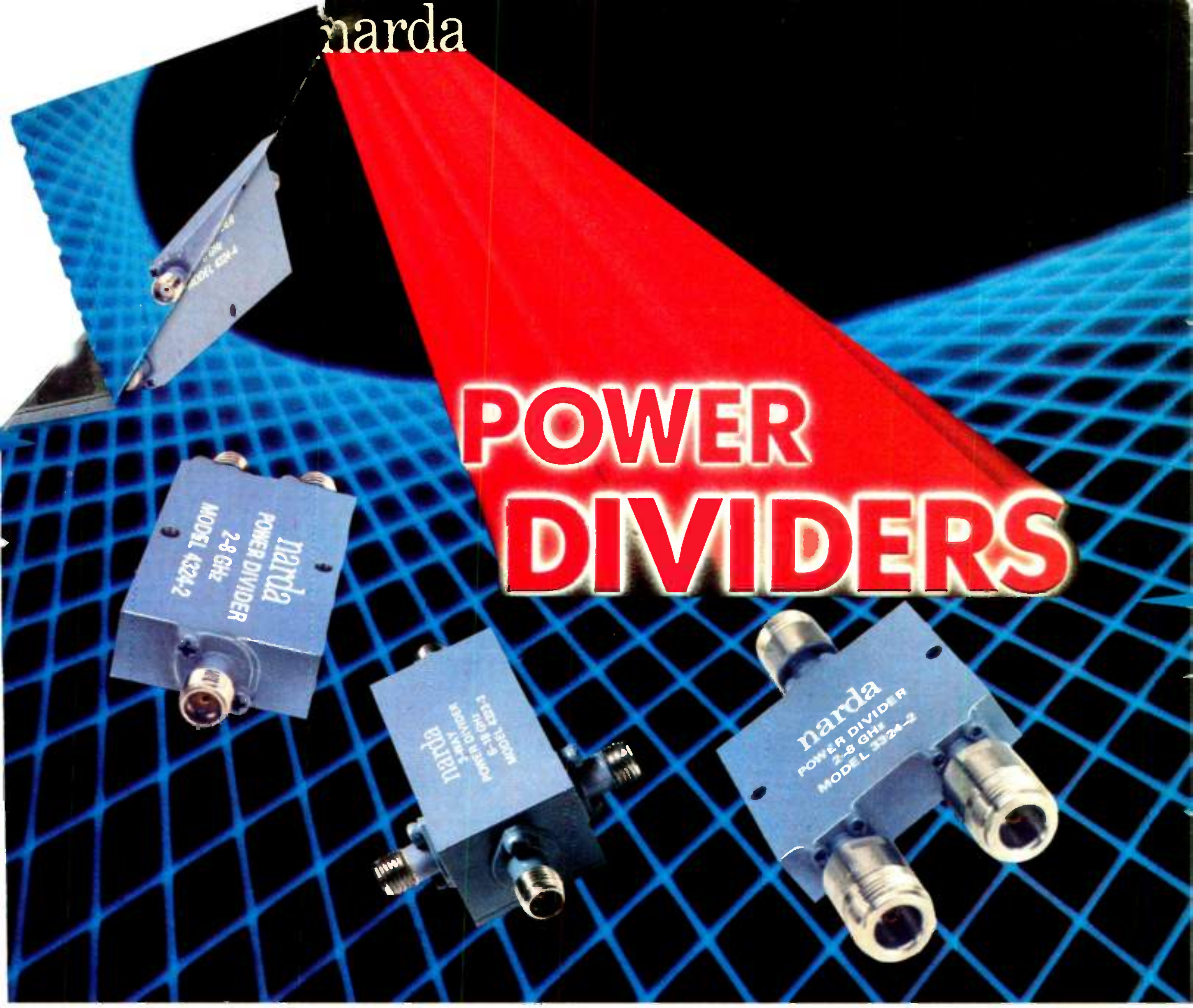

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Multi-Octave, Precision	7mm	2.0-18
Ultra Broadband, 36:1	SMA	0.5-18
High Power	TNC	2.0-18
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Coaxial	TYPE-N	0.215-11
Mini-Stripline 90°	SMA	0.125-18
Quadrature 3dB	SMA	2.0-18
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Coming Events

28th ARFTG Conference Dec. 4-5, 1986 Sponsor: IEEE Automatic RF Techniques Group. Topics: Theme is "Precision Microwave Measurements." Measurement methods or techniques for improving the precision of microwave measurements. Place: St. Petersburg Beach, FL. Contact: John Barr, Hewlett-Packard - 4US-Q, 1400 Fountaingrove Parkway, Santa Rosa, CA 95401 (707) 577-2350.

1986 IEEE International Electron Devices Meeting Dec. 7-10, 1986 Sponsor: IEEE Electron Devices Society. Topics: Modeling and simulation; quantum electronics and compound semiconductor devices; device technology; electron tubes; integrated circuits; detectors, sensors and displays; and solid-state devices. Place: Los Angeles, CA. Contact: Conference Manager Melissa M. Widerkehr, Courtesy Associates Inc., 655 15th St. NW, Suite 300, Washington, DC 20005 (202) 347-5900.

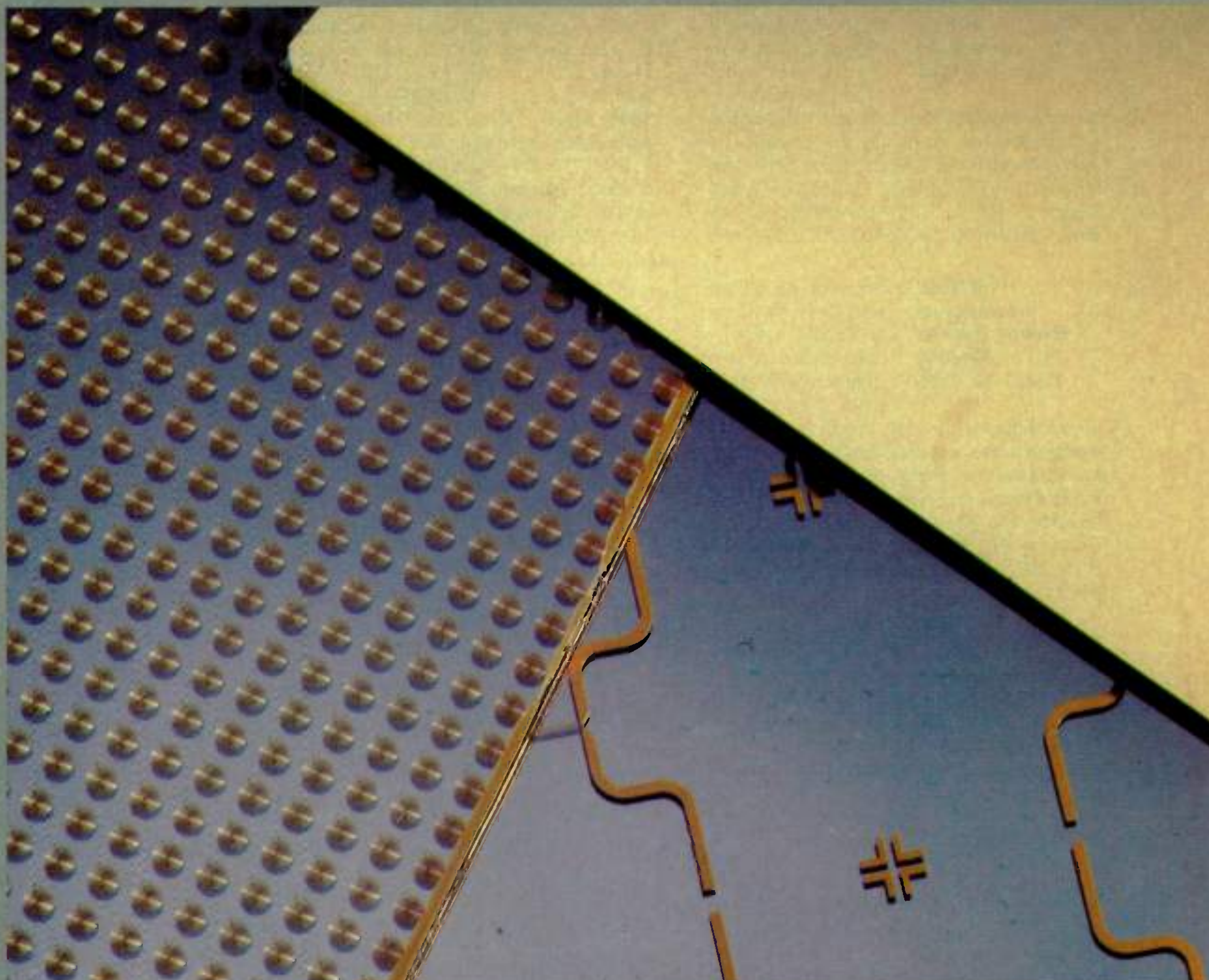
12th International Conference on Infrared and Millimeter Waves Dec. 14-18, 1987 **Call for papers.** Sponsor: Massachusetts Institute of Technology. Topics: Millimeter-wave sources, systems, detectors, devices, atmospheric physics, spectroscopy, mixers/imaging, guided propagation and ICs; plasma diagnostics and FIR materials; sub-mm-wave devices and detectors; lasers; FEL and gyrotrons. Send to: Kenneth J. Button, MIT, Box 72, MIT Branch, Cambridge, MA 02139-0901 (617) 253-5561. **Deadline: July 1, 1987** for 35-40-word abstracts. Place: Lake Buena Vista, FL. Contact: Button, see address above.

1987 IEEE International Solid-State Circuits Conference Feb. 25-27, 1987 Sponsor: IEEE Solid-State Circuits Council, IEEE New York Section and University of Pennsylvania. Topics: Advances in solid-state circuits: digital, analog, memory and signal processing; also, CAD and simulation, design verification, test generation, modeling, optoelectronics, etc. Place: New York City, NY. Contact: Lewis Winner, 301 Almeria Ave., Coral Gables, FL 33134 (305) 446-8193.

1987 IEEE Microwave and mm-Wave Monolithic Circuits Symposium June 8-9, 1987 **Call for papers.** Sponsor: IEEE Microwave Theory and Techniques Society, IEEE Electron Devices Society. Topics: Analog and related digital ICs; solid-state devices and circuits; fabrication technology and yield; radiation effects and reliability; packaging and testing; systems, subsystems and components; signal control and modulation; computer-aided design techniques; integrated optoelectronic circuits. Submit: 5 copies of 500-1,000 word summary with illustrations and clear explanation of contribution. Also, 5 copies of 30-50 word abstract with author address. *Do not submit same material to the MTT-S Symposium and this one.* **Deadline: December 12, 1986.** Send to: Derry Hornbuckle, c/o LRW Associates, 1218 Balfour Drive, Arnold, MD 21012 (707) 577-3658. Place: Las Vegas, NV. Contact: General Chairman Yalcin Ayasli, Hittite Microwave Corp., 21 Cabot Road, Woburn, MA 01801 (617) 933-7267.

1987 IEEE MTT-S International Microwave Symposium June 9-11, 1987 **Call for papers.** Sponsor: IEEE Microwave Theory and Techniques Society. Topics: New microwave theory and technique developments, from acoustics to optics. Biological effects and medical applications; CAD; solid-state devices and circuits; microwave systems; ferrite devices; GaAs monolithic cir-

Thin film coatings for advanced microwave circuitry?



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Coming Events

cuits; high power devices and systems; integrated optics, fiber optics and optical techniques; low noise techniques; microwave and mm-wave integrated circuits; microwave acoustics; communications systems; field and network theory; passive components; phased and active array techniques; sub-mm-wave techniques and devices; measurement theory and techniques; and manufacturing methods. Send: 15 copies of a 500-1000 word summary with supporting illustrations. Clearly explain contribution, originality and relative importance. Also, 10 copies of a 30-50 word abstract including author's name and affiliation, plus a separate sheet with address of author and a statement categorizing as full length, short or open forum. **Deadline: December 12, 1986.** Send to: Dr. R.S. Kagiwada, c/o LRW Associates, 1218 Balfour Drive, Arnold, MD. Place: Las Vegas, NV. Contact: Steven March, Symposium Steering Committee Chairman (714) 987-4715.

1987 IEEE AP-S International Symposium June 15-19, 1987

In conjunction with URSI Radio Science Meeting. **Call for papers.** Sponsor: IEEE AP-S and URSI. Topics: Adaptive, microstrip and reflector antennas, antenna measurements and metrology, antenna theory, environmental effects on waves, feeds and radiating elements, mm-waves, numerical methods, phased arrays, propagation, remote sensing, scattering and diffraction, wave-propagation theory; other topics will be considered. Send: Original and three copies of a summary, not to exceed four pages, single-spaced, including text, references, figures and photographs. Final paper instructions are available. **Deadline: January 2, 1987** for summaries. Send to: Charles W. Bostian, Technical Program Chairman, Dept. of Electrical Engineering, Virginia Polytechnic Institute and State University, Blacksburg, VA 24061 (703) 961-6834. Place: Blacksburg, VA. Contact: Bostian, see address above.

International Microwave Symposium/Brazil July 27-30, 1987

Call for papers. Sponsor: Brazilian Microwave Society in cooperation with the IEE, IEEE, IEEE MTT-S, IEEE AP-S, and IEEE ED-S. Topics: antennas and arrays, microwave radio propagation and radiometeorology, terrestrial and satellite communication systems, microwave active/passive devices and components, mm-wave components, circuits and systems, microwave hybrid and monolithic circuits, microwave techniques in radar, ECM, remote sensing and radio astronomy, microwave measurements, CAD/CAM, scientific, biological, medical and industrial, optical communications, and field and network theory. Submit: Original and three copies of six-page single-spaced paper (including figures, tables and references) with address of author(s) and telephone or telex number. Type on one side only. Author's name, affiliation and address should start two lines below title and text should start three lines below this, w th an abstract (max. 50 words), followed by the introduction. Number pages in pencil. Send to: Prof. Alvaro A. de Salles, 1987 International Microwave Symposium Committee, Rua Marques de Sao Vicente 225, Gavea Rio de Janeiro, Brazil 22453. **Deadline: December 31, 1986.** Place: Rio de Janeiro, Brazil. Contact: L. Vasquez, *M*I*L*A, 38760 Northwoods Drive, Wadsworth, IL 60083 (312) 249-1900 or (800) 367-7378.

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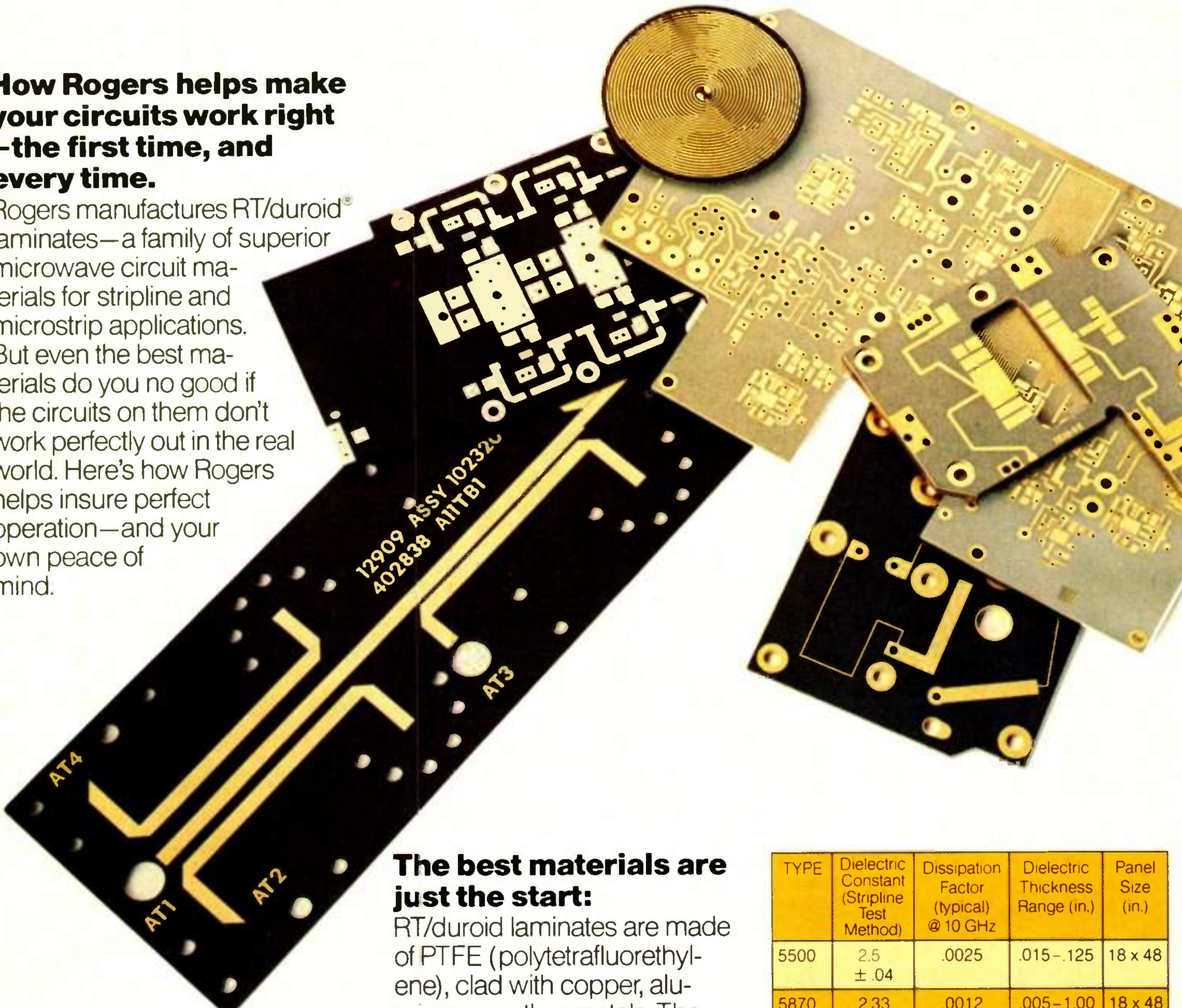

How Rogers helps make your circuits work right —the first time, and every time.

Rogers manufactures RT/duroid® laminates—a family of superior microwave circuit materials for stripline and microstrip applications. But even the best materials do you no good if the circuits on them don't work perfectly out in the real world. Here's how Rogers helps insure perfect operation—and your own peace of mind.

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5870	2.33 ± .02	.0012	.005–1.00	18 x 48
5880	2.20 ± .02	.0009	.005–1.00	18 x 48
6006	6.0 ± .20	.0012	.010–.100	20 x 20
6010.2	10.2 ± .25	.0015	.025–.100	20 x 20
6010.5	10.5 ± .25	.0015	.010–.100	20 x 20

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Planning for fail-safe operation:
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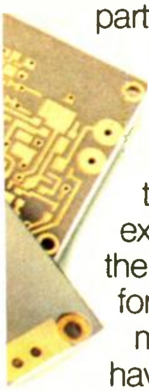
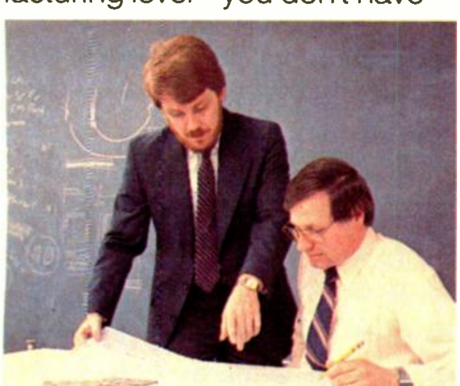


Rogers people are always available for on-site help at the manufacturing level—you don't have to wait for the next seminar. Nobody understands microwave materials like Rogers, so it stands to reason that Rogers is your best day-to-day source for the most accurate information and cost-efficient help.



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CIRCLE 15 FOR IMMEDIATE NEED
CIRCLE 16 FOR INFORMATION ONLY

Sum Up

Editor's Note: The November issue of the *Microwave Journal* concentrates on passive components and their integration into microwave systems. Some articles in this issue (identified by asterisks on their title pages) were invited for publication because of their appropriateness to this theme.

An Experimental Technique for Determining Coupling Between Dielectric Resonators

Jerry C. Brand describes an experimental method to determine the coupling coefficients between two dielectric resonators. S-parameter measurements are used to characterize a circuit model of the dielectric resonator. This model is correlated with data obtained from resonance frequency changes, caused by means of a sliding short, to determine the coupling coefficient. The author presents comparisons between calculated and measured coupling coefficients.

Stopband Filter: Resonance Frequency Shift due to an Uncoupled Length of Resonator

Dipak S. Kothari notes that the coupled length of a resonator is different from 1/4 wavelength at the resonance frequency. Since an additional resonator length lowers the resonance frequency, the coupled resonator should be 1/4 wavelength long at the higher frequency. The author introduces a calculation method for the higher resonance frequency and the length of the coupled resonators.

Sensitivity Analyses of 3 dB Branchline Couplers

A.F. Celliers investigated the effects of small changes in line lengths and impedance levels on the amplitude of the coupled ports and through ports. He presents data on the magnitude of performance degradation due to small line length and line width errors of a T-junction circuit, which he used in his computer model, at 6 GHz.

A Hybrid Ring Directional Coupler for Arbitrary Power Divisions

A.K. Agrawal and G.F. Mikucki state that the maximum power split ratio between two output ports of a conventional printed circuit hybrid ring coupler is limited by the highest impedance line that can be realized. They describe an advanced design that allows a higher power split ratio for the same impedance lines, thereby increasing the range of the power split ratio. The authors also present a theoretical analysis using the scattering matrix and experimental verification of their theoretical predictions, using a Ku-band stripline configuration.

Matched, Dual Mode Square Waveguide Corner

P.K. Park, R.L. Eisenhart and S.E. Bradshaw present a design method for matching a square waveguide right angle corner for both E and H plane (TE_{10} and TE_{01}) modes. Experimental results showed that SWRs of better than 1.05 for 1 GHz bandwidths could be achieved at X band with mode-to-mode isolation greater than 30 dB.

Simplified Expressions for the Calculation of the Impedance of a Shielded Slab Line

Stewart M. Perlow presents simple equations for the calculation of the characteristic impedance of strip lines and slab lines. These equations hold for any rectangular cross-section of the inner conductor, including the degenerate cases of infinitely thin horizontal or vertical strips. The method described may be used with hand-held calculators and for CAD programs since one set of equations covers all possible inner conductor shape ratios.

Howard Ellowitz

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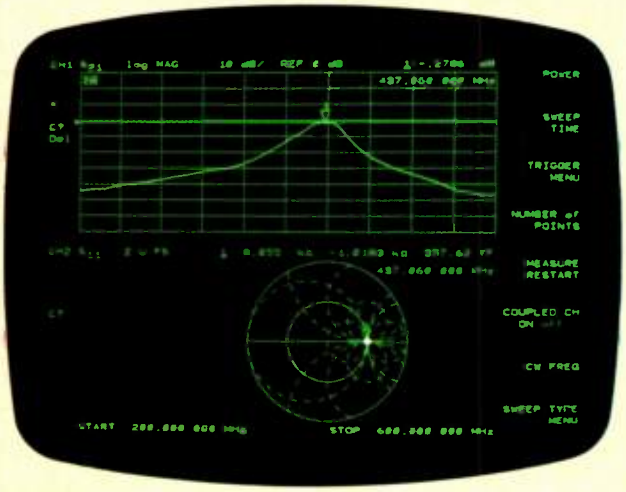

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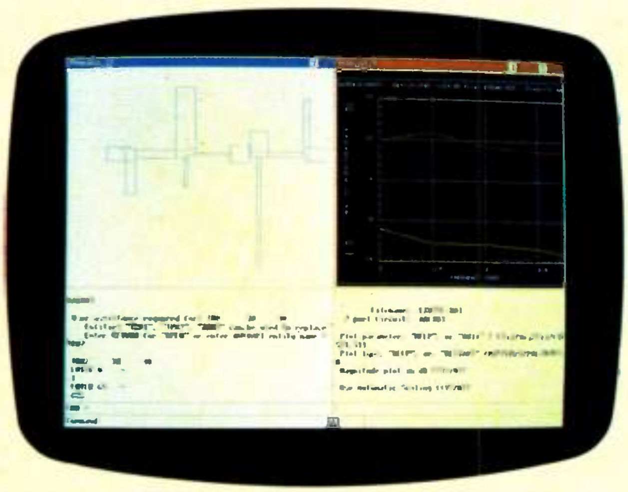

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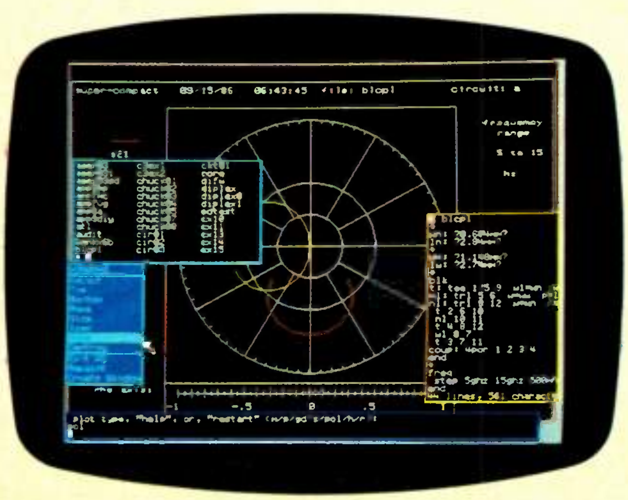

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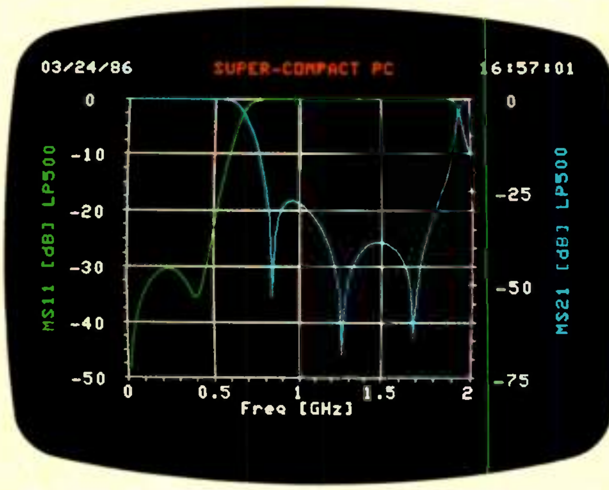

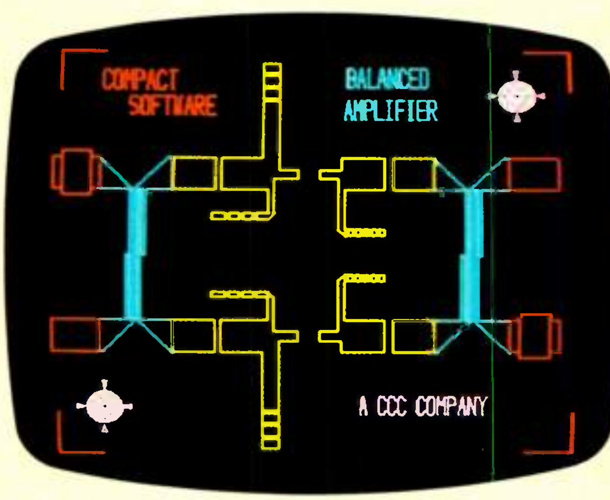

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Site: Atlanta, GA.
Date: Dec. 2-5, 1986.
Contact: Deidre Mercer, Dept. of Cont. Ed., Georgia Institute of Technology, Atlanta, GA 30332-0385 (404) 894-2547.

mm-Wave Engineering and Applications

Topics: Introduction, propagation characteristics of mm-waves, mm-wave sources (tubes and solid-state), antennas, transmission media, waveguide circuits and components, integrated mm-wave circuits and components, mm-wave systems and applications, mm-wave radar, mm-wave communication. Fee: $860.
Site: Washington, DC.
Date: Dec. 8-11, 1986.
Contact: Stod Cortelyou, Cont. Eng. Ed. Dept., George Washington Univ., Wash., DC 20052 (202) 676-8520 or (800) 424-9773.

Antenna Measurement Techniques

Topics: Antenna parameters for measurement, far-field range pattern measurements, anechoic chamber and compact range pattern measurements, near-field range pattern measurements, automation of antenna measurements, equipment specification for antenna measurements. Fee: $760.
Site: Atlanta, GA.
Date: Dec. 9-12, 1986.
Contact: Lynda S. Epstein, course registrar, Technology Service Corp., 962 Wayne Ave., Suite 600, Silver Spring, MD 20910 (301) 565-2970 or (800) 638-2628.

Infrared Detectors and Systems

Topics: Geometrical optics, radiometry, figures of merit for detectors, visible detectors, infrared detectors, focal plane arrays. Fee: $595.
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Contact: V. Amico, University of Central Florida (305) 275-2123.

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RF/Microwave Videotape Courses

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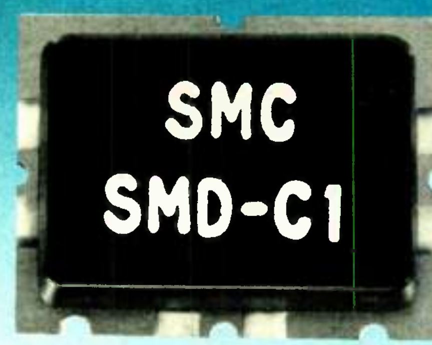

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SMD-2	20-1500	DC-1500	+7
SMD-2M	20-1500	DC-1500	+17
SMD-2H	20-1500	DC-1500	+23
SMD-3	20-2500	20-600	+7
SMD-3M	20-2500	20-600	+17
SMD-3H	20-2500	20-600	+23

MODEL	RF/LO MHz	IF MHz	LO POWER dBm	PRICE (100 pcs.)
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SMD-C1M	1-1000	DC-1000	+17	7.00
SMD-C1H	1-1000	DC-1000	+23	11.00
SMD-C2	20-1500	DC-1500	+7	7.00
SMD-C2M	20-1500	DC-1500	+17	8.00
SMD-C2H	20-1500	DC-1500	+23	12.00
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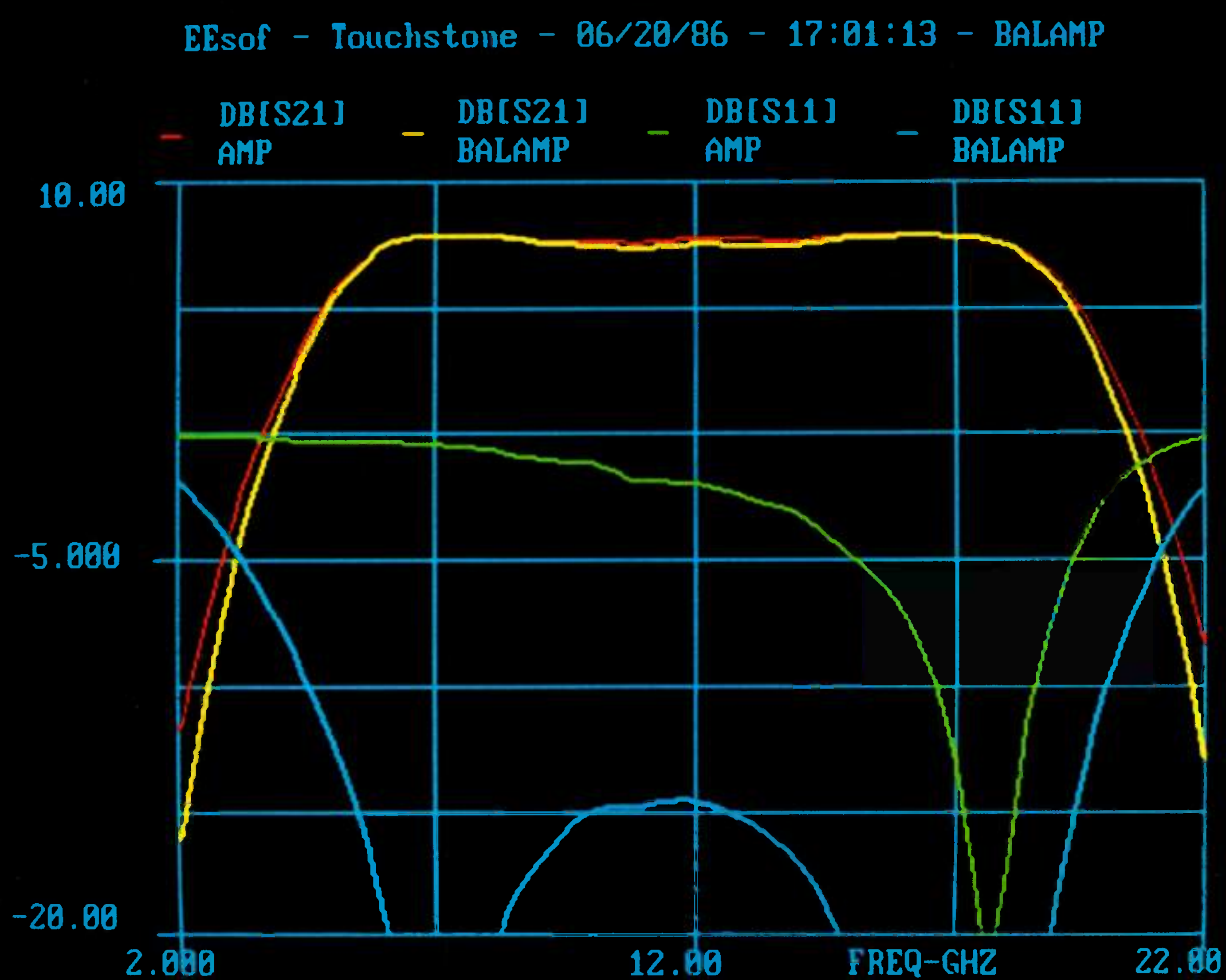

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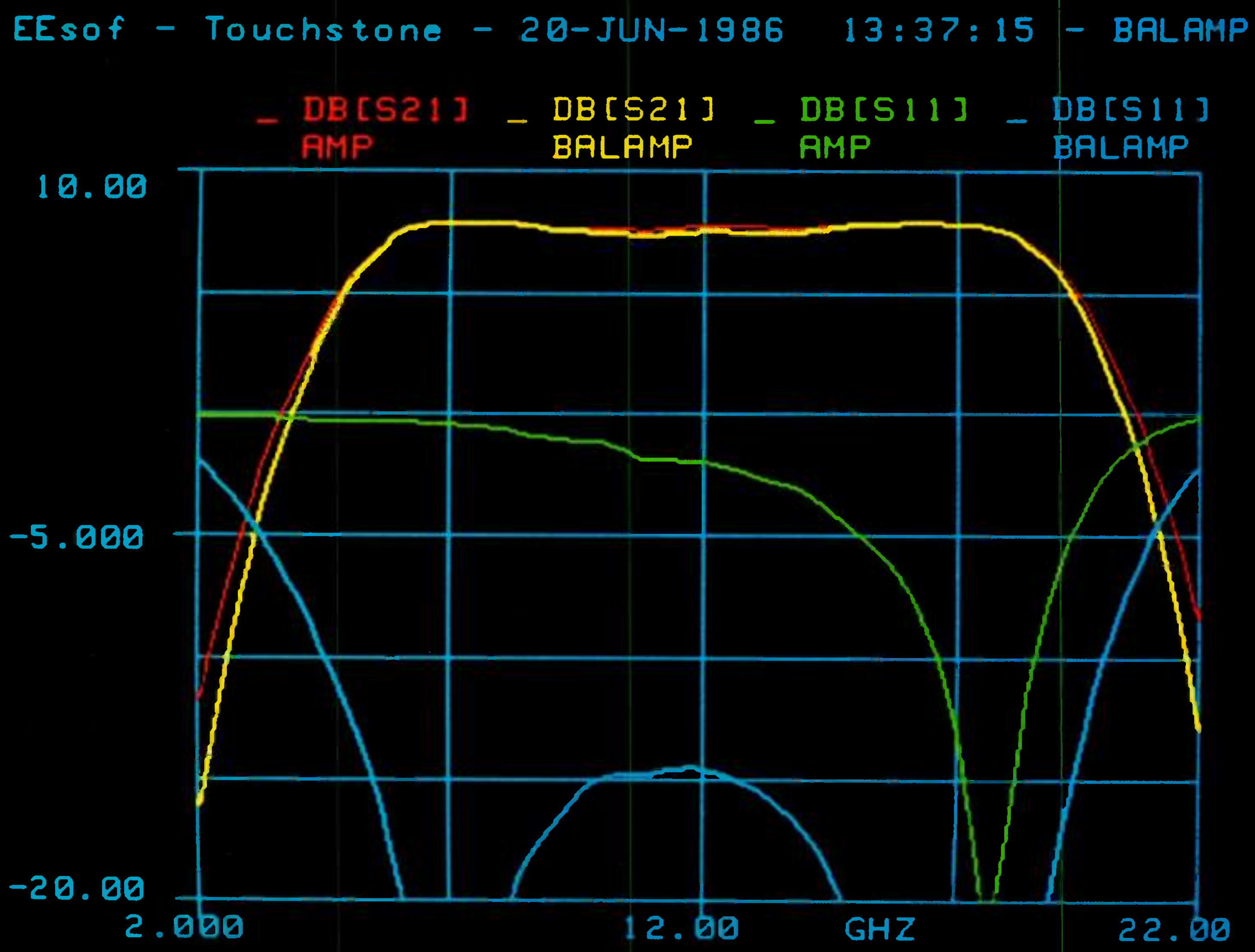

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Microwave Packages Are Meeting GaAs Challenges

Howard Bierman
Contributing Editor

Until recently, packaging approaches for microwave semiconductor chips and subsystems were considerably different than those employed for lower frequency consumer/computer devices. The goals for consumer/computer assembly packages were minimum size, weight and cost. Package requirements for microwave equipment, with greater emphasis on performance and reliability to meet the needs of its many military/space customers, included built-in characteristic impedance, propagation delay and signal line reflections.

Today's demands by computer and military/space systems designers to make use of affordable, high-frequency, high-speed GaAs circuits in their next-generation products places a heavy burden on packaging designers dealing with analog chips operating up to 20 GHz and digital ICs with rise times and gate delays less than 100 ps. At these speeds, signals routed along parallel line patterns are susceptible to crosstalk; reflections stemming from characteristic impedance mismatches can upset circuit performance; and an interconnection longer than 0.6 inch can inject a lengthier transmission line delay than the gate delay of the circuit chip. In addition to these stringent package demands, designers are expected to come up with solutions that are considerably more cost-effective than present schemes and that will lend themselves to a reasonable degree of production-line automation.

These points were reinforced at last month's GaAs IC Symposium at Grenelefe, FL by Jeffrey Frisco of Harris' Government Aerospace Division (Melbourne, FL). In a microwave application, Frisco pointed out, the package is generally part of the signal propagation medium, either shielded microstrip or waveguide, and is thus often an expensive custom assembly. Conventional silicon digital circuits, on the other hand, are traditionally identified with low-cost, high-production runs where unique and expensive packages are shunned. Similarly, Frisco added, while microwave assemblies are integrated using costly, labor-intensive hybrid packages or coaxial cables, digital building blocks take advantage of automated assembly techniques such as multilayer PC boards and wire-wrap arrangements.

With the current thrust by a growing number of GaAs chip makers to supply low-cost, high-performance analog and digital monolithic microwave integrated circuits (MMICs), it becomes obvious that low-cost packages and mass-production assembly techniques must be developed to take full advantage of the available semiconductor technology. A number of component and systems firms, including Harris, M/A-Com, TriQuint Semiconductor, Tektronix, GigaBit Logic and Vitesse Semiconductor, are offering commercially-available packages while others, such as GE and the Mayo Foundation, are developing clever packaging schemes.

Chip Carriers

Small, multiport microwave chip carriers (MCCs) have been developed at Harris Aerospace Systems Division to house high frequency, high speed GaAs chips. One design, shown in Figure 1, consists of two ceramic layers plus a metal base and cover. During the package design process, signal path length between the ceramic interface and the external pads was modeled as microstrip, the path length between ceramic layers was modeled as nonsymmetrical stripline, and the path length inside the package was modeled as covered microstrip; inductance and capacitance were included to model bond wires and edge capacitance. Signal lines and adjacent lines separated by ground conductors were analyzed and optimized for return loss, isolation and resonance. Based on the modeling analysis, MCCs were fabricated using co-fired aluminum oxide for the ceramic layers. The metal base and cover were braze-attached to the ceramic layers.

[Continued on page 28]

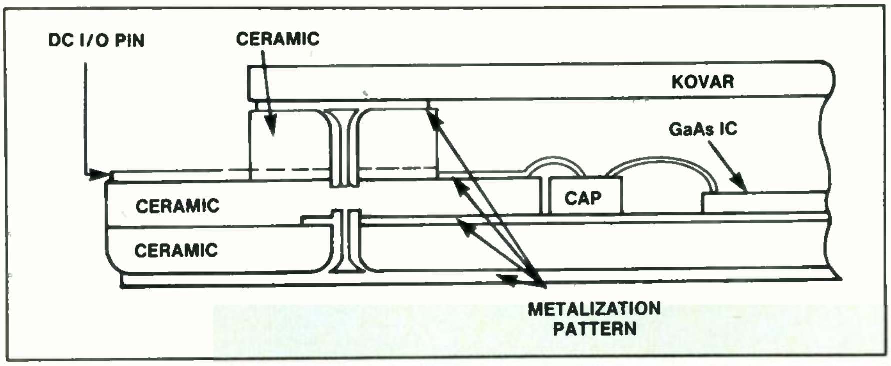
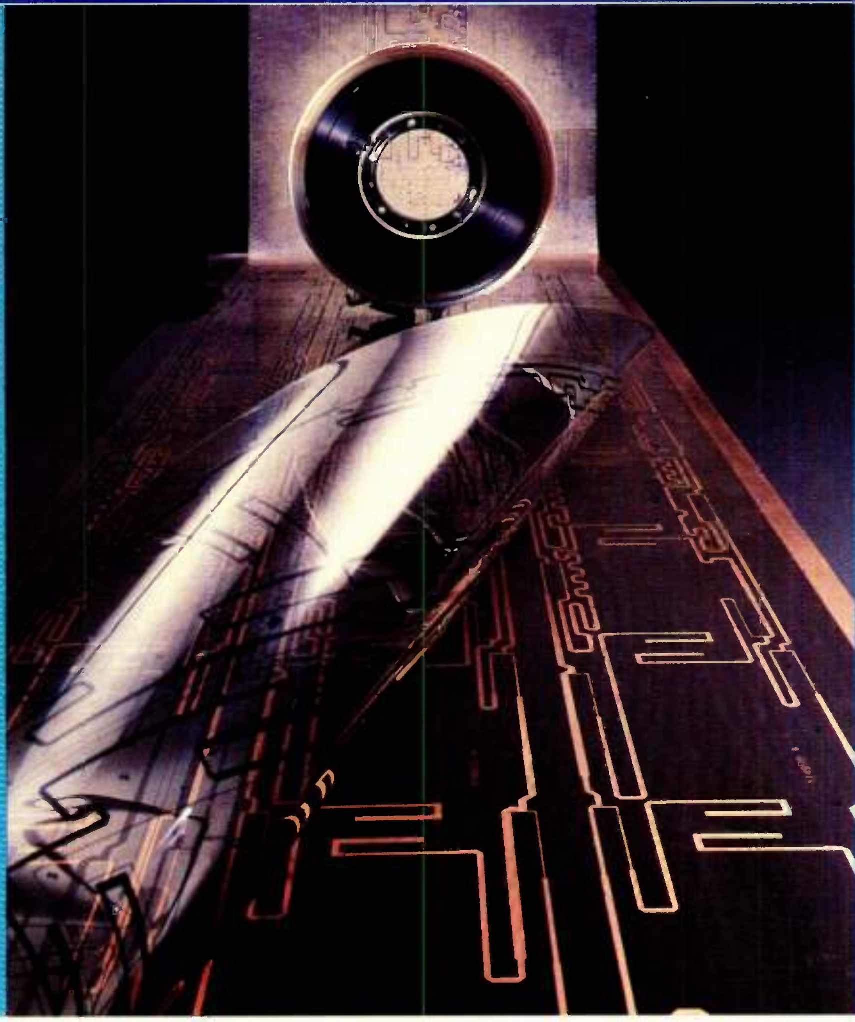


Fig. 1 Cross-sectional view of Harris' MCC showing the two ceramic layers with metal base and cover.



A "cavity-up" topology was selected to achieve a smooth microstrip transmission line launch to an interconnecting medium using ribbon or wire bonding. Sealed plated-through holes were used for grounding. Heat transfer was accomplished by means of a heat-conducting mounting structure attached to the base. The 0.35 inch square 24-pin high-speed MCC package offers hermeticity, ruggedness, high temperature stability and excellent thermal conductivity. Each input/output pad has adjacent ground pads to improve launching and also to increase signal path isolation; a 0.05 inch center-to-center spacing for signal paths on the outside of the chip carrier is consistent with a number of currently-used microwave subsystem interconnect schemes, including Harris' Waffleline. The Harris MCC, with metal-to-ceramic hermetic sealing and heat transfer capability to handle up to 2 W, has a demonstrated insertion loss of only 0.25 dB, return loss of 20 dB at all ports, and better than 30 dB isolation between adjacent signal paths through 20 GHz.

Ceramic-based packaging for the hermetic sealing of high power MMIC amplifier chips has been developed at M/A-Com's Advanced Semiconductor Operations (Lowell, MA). The 0.5 x 0.5 inch package includes planarized ground-signal-ground (G-S-G) RF transitions to allow direct cascading of packaged MMICs. Maximum use of subsystem space is achieved by permitting the control and bias signals to be applied to the packaged MMIC in the RF signal plane. The package's planar input/output transitions facilitate planar connections from adjacent microstrip or coplanar transmission lines of various substrate materials and thicknesses; the planarized RF transition minimizes the dependency on the quality and dimensions of the base ground connections to the RF subsystem.

The cutaway view of the package, Figure 2, shows the metal base, ceramic substrate, ceramic seal frame, Kovar lead frame and ceramic lid; screen-printed tungsten co-fired with the integrated aluminum structure is used for metalization. The MMIC die is mounted on a molybdenum or laminant carrier and tested prior to assembly into the package. A eutectic gold-tin alloy is used to attach the MMIC die to the carrier, and the carrier is then attached to the package using a tin-silver alloy. DC and control connections to the package are made with standard photo-etched lead frame brazed to the ceramic substrate, allowing a planar connection either to the motherboard or to circuitry above the MMIC in high density RF subsystem designs. The package's thermal impedance is minimized by the use of a high conductivity thermal via in the metal base; when used to house an MMIC power amplifier chip, the package permits a base operating temperature up to 60°C and FET junction temperatures as high as 150°C. RF modeling and performance test results, shown in Figure 3, indicate that insertion loss

[Continued on page 31]

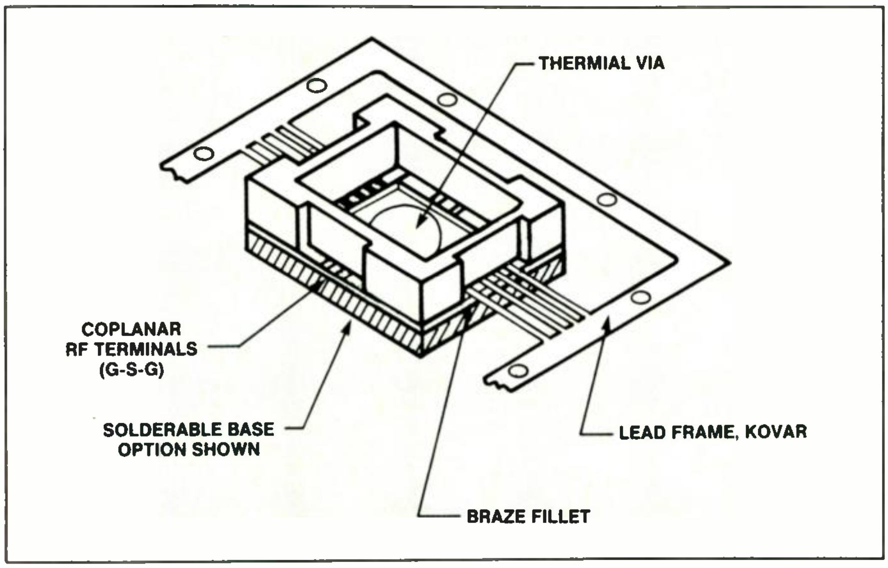


Fig. 2 M/A-Com's package includes planarized ground-signal-ground RF transitions.

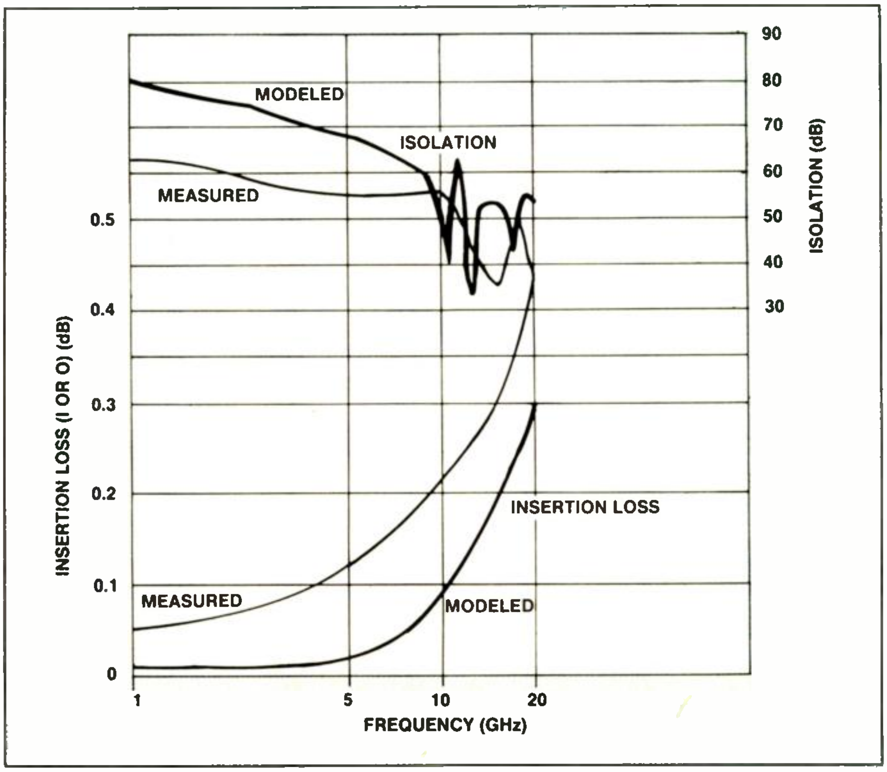


Fig. 3 Modeled and measured isolation and insertion loss for the M/A-Com MCC.

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At 8 GHz	-67	-64	-87	-72	-87	-80	-107	-101
At 18 GHz	-60	-54	-80	-60	-80	-68	-100	-97

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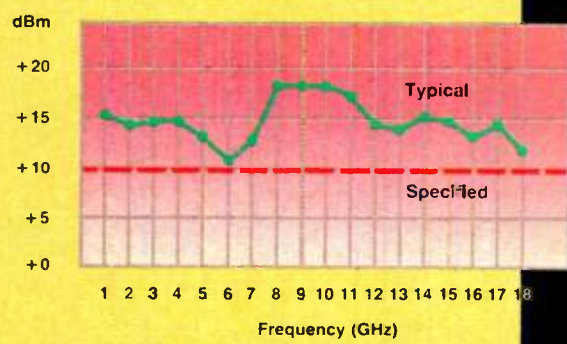

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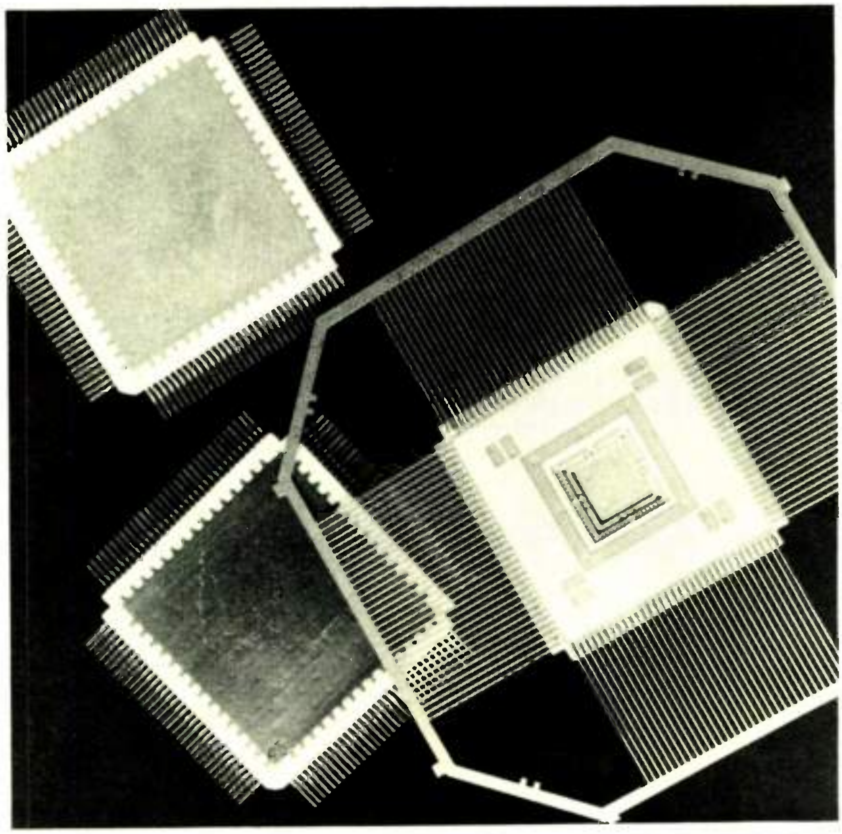


Fig. 5 TriQuint offers a 132-pin multilayer ceramic package in leaded and leadless formats.

chip capacitors. The 0.95 x 0.95 x 0.06 inch 132-pin package provides 64 controlled impedance signal lines implemented in a modified stripline structure. The measured propagation delay of the longest line on the 132-pin package is approximately 100 ps from the contact pad to the die bonding pad.

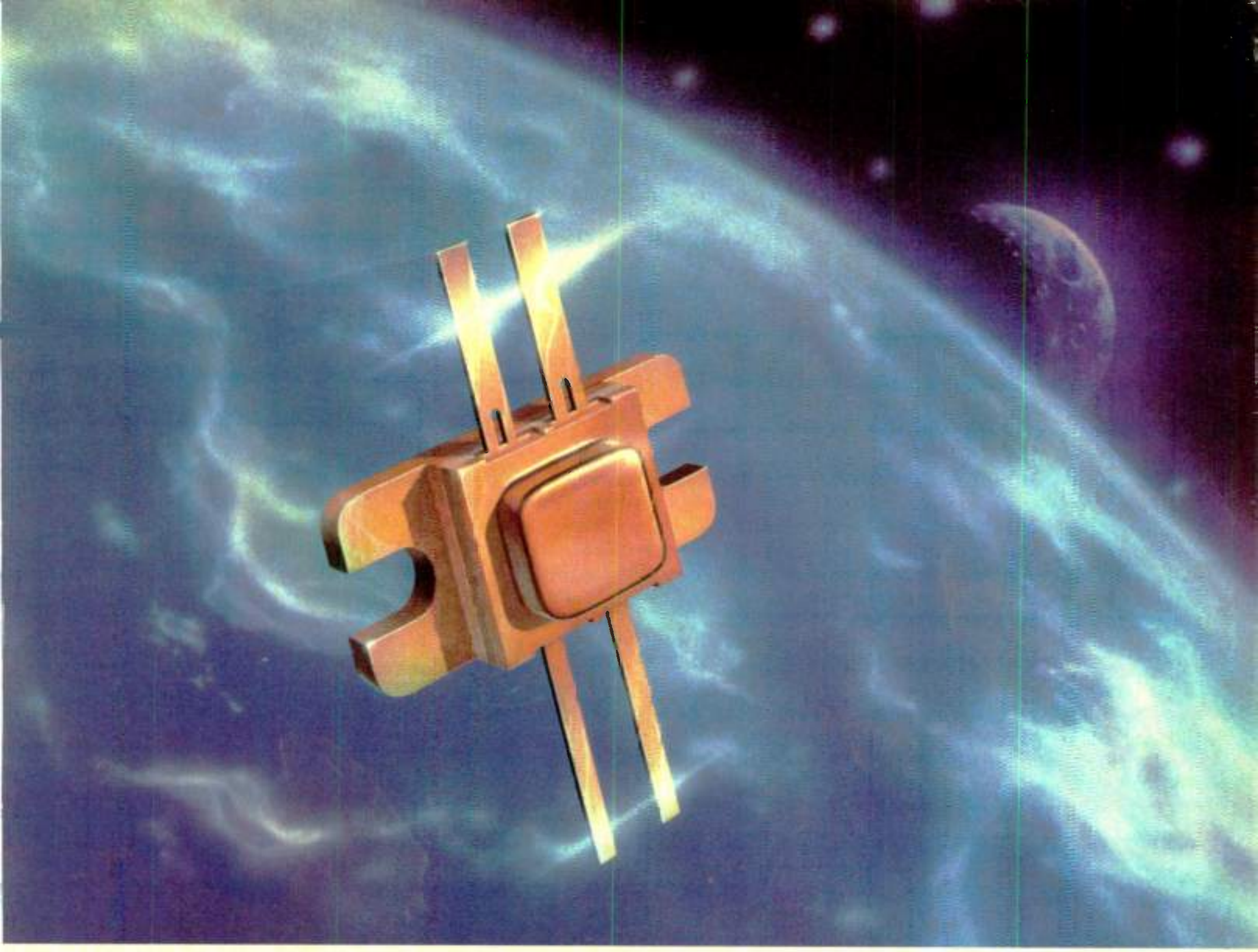
At the Mayo Clinic (Rochester, MN), chip carriers have been developed for a research project funded by DARPA/DSO for use with two types of GaAs ICs: low lead count digital devices operating at gigahertz rates and high lead count devices operating up to 200 MHz. Described by Barry Gilbert and Dan Schwab at last month's GaAs IC Symposium, the 500 mil square package designed for high frequency die provide 64 signal contacts and 24 power contacts for a total of 88 input/output spaced on 20 mil centers. To achieve power delivery with adequate decoupling devoid of inductive spikes, two voltage contacts and a ground are provided near each corner of the die cavity; the ground contact is wide enough to hold a dual parallel-plate capacitor. These corner contacts act as the power feed for the internal transistors on the die while eight contacts, two on each side of the die cavity, are available as power feeds for the chip output drivers.

Packages were fabricated using four different technologies: co-fired alumina, co-fired beryllia, copper/polyimide, and thin-film beryllia. All four were designed to have the base or back surface solder-attached to a circuit board or substrate in order to:

- Achieve strong mechanical support
- Provide a wide area ground return path contact to all signal I/O ports and
- Provide a large thermal heat transfer path.

An 8 GHz bandwidth was measured on the copper/polyimide package, a 4.5 GHz bandwidth was measured for the co-fired beryllia package, and a 3.0 GHz bandwidth was provided for the co-fired alumina assembly; no data was available

[Continued on page 34]



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	Configuration	Fo	Pout	Vcc	Pg	P.W.	D.C.	Tstg & T_j
SD1564	Common emitter	400-450 MHz	400 Watts	40 Vcc	7.5dB	1ms	10%	Tstg = −65 - +200°C T_j = +200°C
SD1565	Common base	400-450 MHz	500 Watts	40 Vcc	10.0dB	250μs	10%	Tstg = −65 - +200°C T_j = +200°C

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for the thin-film beryllia package.

The lower frequency package will support up to 192 signal contacts plus an additional 24 power/ground contacts for a total of 216 contacts. To keep the chip carrier size small, an input/output pad pitch of 10 mils was used to produce a 650 mil x 650 mil outline. The I/O connections for this mini-carrier to the circuit board are made using tape-automated bonding (TAB) or wire/ribbon bonds (Figure 6); the bonds can absorb the stresses induced by different coefficients of expansion. Onboard terminating resistors are included on the package with one end of each resistor connected to the voltage termination plane and the other end tied to an open pad; when the package is bonded to a printed circuit board or substrate, a second bond can be added to connect the resistor if required.

TO-Packages

Efforts to use low cost, plug-in TO-type transistor packages, such as the 0.3 inch outside diameter (OD) TO-5 and the 1.0 inch OD TO-3, also are being pursued at Harris. Single GaAs chips, several cascaded chips and entire hybrid substrates have been mounted in these packages with excellent results. For example, tests on a TO-8 package show better than 20 dB return loss through 15 GHz with isolation between adjacent ports greaer than 50 dB.

However, conventional TO-style package limitations for high frequency operation include inadequate RF grounding, seals that are not 50 ohm-matched, a right-angle transition between the pin extending up into the package and the chip, and cavity resonance. For operation in the Ku band, Harris selected a TO-8 package with an inner diameter of 0.625 inch having a cavity resonant frequency exceeding 17 GHz. First, the seals in the conventional TO-8 package were replaced with high quality glass seals so that the inner pin conductor, the glass dielectric and the outer metalization of the seal appear similar to a 50 ohm characteristic impedance transmission line.

The next problem tackled was the metal pin extending into the package for bonding to the chip, which appears highly inductive at microwave frequencies. A metal ridge, taller than the pin extension into the package, was formed to encircle the pin and create capacitance around the pin to match out the inductance. To provide adequate RF grounding, an additional cover was added to the TO-8 package with screw holes along its lip; this cover is placed on top of the standard lid and screwed to the ground surface of the next level of integration. This type of clamping scheme allows simple, low risk part replacement, a critical consideration in subsystem repair.

A TO-8 header with four microwave monolithic ICs providing 40 dB gain up to 16 GHz is shown in Figure 7; this assembly includes eight RF ports and two power pins. A major advantage of the TO-packages, according to Harris, is that they offer the minimum physical outline for a given chip size and the

[Continued on page 36]

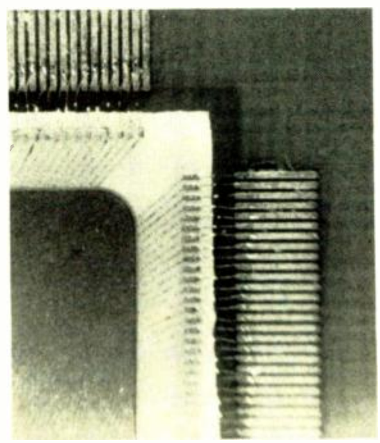

Fig. 6 Photo-micrograph of the attachment of 10 mil perimeter pads of Mayo Clinic's 216-pin microchip carrier.

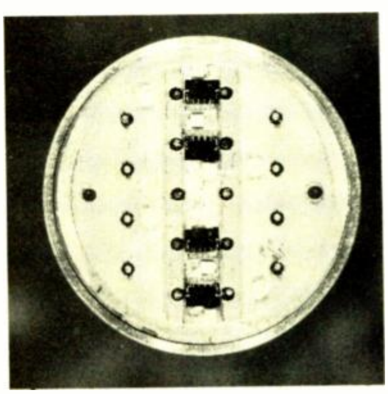

Fig. 7 Four MMICs are packed into a 0.625 inch inside diameter TO-8 header.

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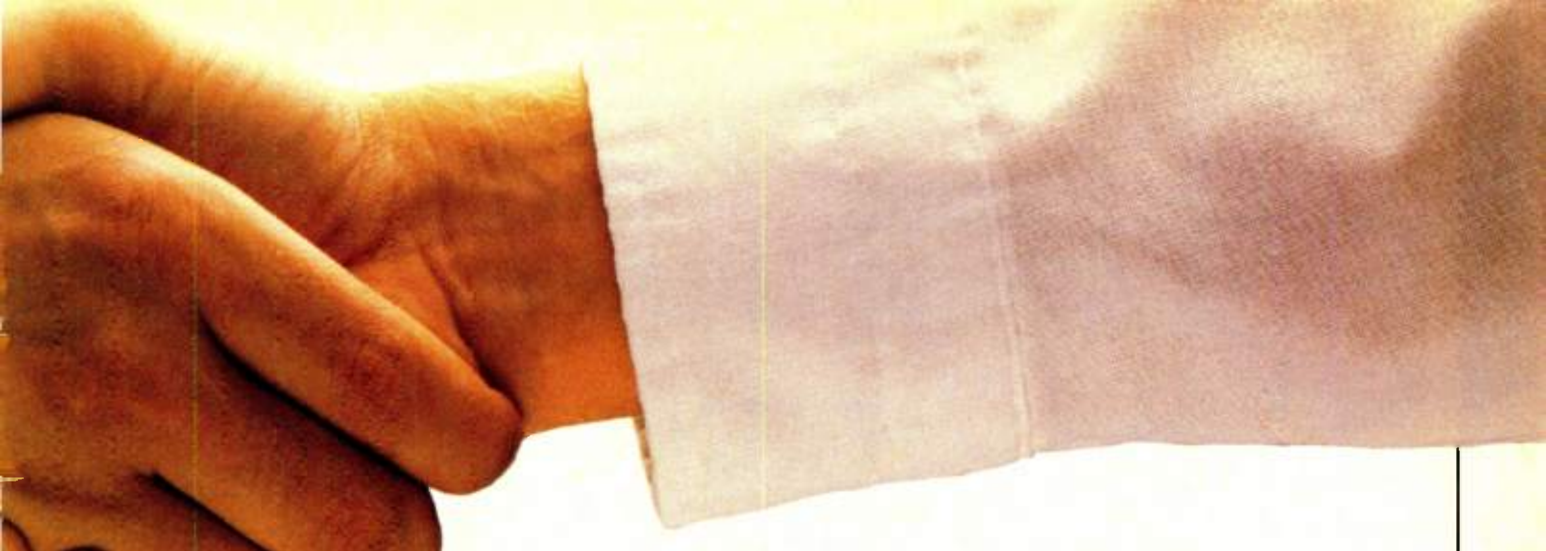

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Henry Eisenson
Sciteq Electronics Inc.
San Diego, CA

The frequency synthesis function permeates every sector of the electronics industry and now appears as a critical building block in major systems of all types.

General-purpose laboratory synthesizers cover wide bandwidths with good spectral purity; collectively, they define the state of the synthesizer art. That's partly because instrumentation specialists can distribute development costs across hundreds or even thousands of production units. Design efforts benefit from formalized production engineering to optimize producibility, economy and testability.

Similar advantages are seldom available to engineers with the task of satisfying an original equipment manufacturing (OEM) requirement; these programs involve much smaller production volumes. Nonrecurring engineering (NRE) or development charges can sometimes equal the cost of the hardware. The product may be tailored to the requirement as much as possible, but the designers usually are constrained by the expected production volume — they simply cannot start from scratch and still keep the result cost-effective.

In RF electronics, engineers use a set of conventions that divide the spectrum into baseband, VHF, UHF and microwave. Baseband synthesis has recently benefited from important evolution in digital techniques and devices, and VHF/UHF products have improved through hybridization with digital designs. Microwave synthesizer technology, however, has remained relatively static for years.

The typical OEM microwave synthesizer today suffers from lack of innovation. Despite improvements in technique and devices, most of today's OEM needs are met by products that were conceived nearly a decade ago. Except in some cost-no-object programs, it is the *requirement* that must adapt to the capabilities of available products rather than the other way around.

This is particularly true in the microwave domain because of the cost and financial risk involved in development work above 1 GHz. The business side of the synthesizer industry keeps its eye on the profit line and limits speculative development when an existing product can do the job in a satisfactory (if not exceptional) manner. Rapid growth of the microwave industry, in general, has helped shape that attitude because sales have steadily increased without a need for technological acceleration. There is no need to seek a solution when the market doesn't see a problem.

That "market" consists of system designers, who develop their block diagrams based upon accepted limitations of the frequency synthesis function and then work from that point. Because of the central location of the synthesizer in typical diagrams, the limitations of that function often define the limitations of the overall system. In fact, many system designers admit that frequency synthesis is a major limiting factor in their work.

Another issue that muddies the evolutionary stream is environment. Laboratory synthesizers reside on test benches in air-conditioned areas and are rarely subjected to shock and vibration. The laboratory unit can therefore be designed without consideration of the impact upon performance created by environmental stresses.

The OEM product is usually designed to operate over wide temperature ranges and while exposed to considerable shock and vibration, yet it often must be *more* reliable than its instrument counterpart.

There is obviously a paradox. Though exceptions exist, the OEM synthesizer must work well as an instrument, yet be derived from a much smaller development investment. The users of OEM synthesizer technologies do not recognize that paradox because they also are concerned with cost-effectiveness and are generally willing to finance only adaptation of existing designs to their specific requirements. They seldom seek, or are willing to support, something entirely new.

With evolving technologies, advanced radar provides longer ranges and better resolution; navigation becomes more precise and communication systems become more capable. With electronics in general becoming more sophisticated, it takes an astute observer to note that frequency synthesis — despite the fact that it is a core function that often defines the capability of the system — has been relatively static.

Various synthesizers proposed for a given requirement often are based upon similar techniques and, therefore, have very similar limitations. This helps convince system designers that they are all state-of-the-art designs. Buying decisions are influenced more by business issues than by technical performance, and attention shifts to the relative marketing strengths of the potential suppliers.

There is a light at the end of this particular tunnel. Better than any

[Continued on page 44]

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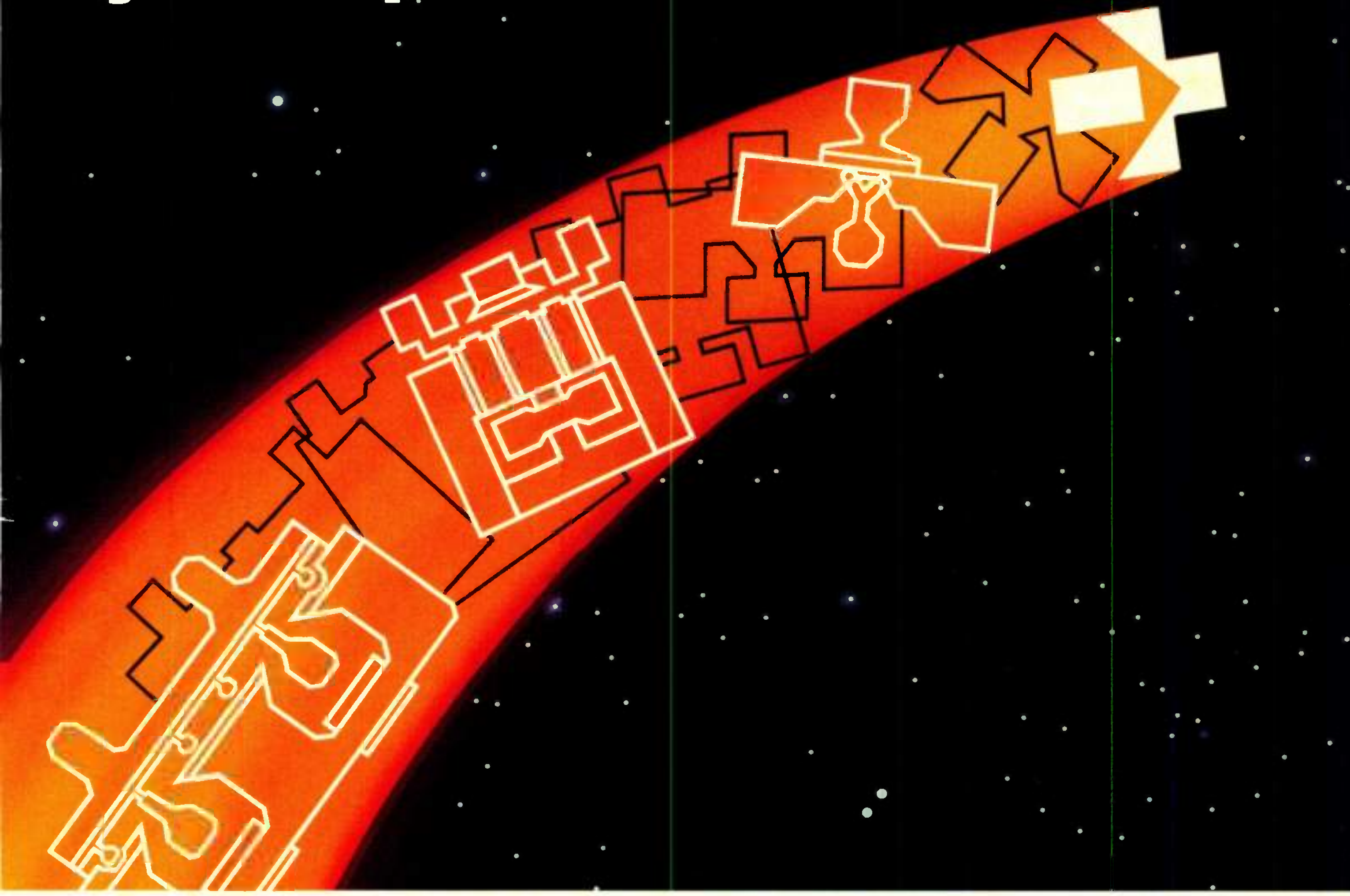

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Fig. A: *MPD H503 HEMT.* 1.8dB noise, 9.5db gain are unmatched at 18GHz. Best first stage for LNAs at 10–30GHz.

Fig. B: *DXL 3904.* 21dBm power FET with 5.5dB gain* for Ku band.

Fig. A

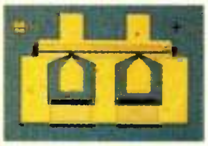

Fig. B

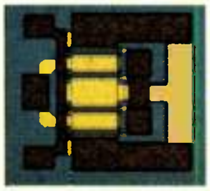

Fig. C

Fig. D

Fig. C: *DXL 2706.* Dual-gate for maximum gain from a single device. Noise 3.8dB, gain 13dB.*

Fig. D: *DXL 0408.* Effective amplification at frequencies to 60GHz. 9db gain, 3.1db noise.*

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For complete product details and the name of your Gould MPD representative, call us at (408) 943-9055. Or write: Device Marketing, Gould Inc., Microwave Products Division, 2580 Junction Ave., San Jose, CA 95134.

*All specifications measured at 18GHz.

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other, the electronics industry demonstrates the effect of competition upon business processes. In any relatively static field, if one company takes a market share greater than what its advertising budget warrants, the others take note. Internal R&D funding decisions usually are justified by a need to compete in the existing or foreseen market, and when competition looms, money loosens up.

If all the players in an industry work placidly in parallel, then who will act as the wild card that stimulates growth? Our system has an answer for this question as well: it's the entrepreneur with a better idea.

He'll grow until his competition sees itself losing market share and makes the necessary investments in new product technology. The entrepreneur enjoys his earned profits from his market penetration and rapidly becomes part of the establishment that has caught up with him. If the scenario continues to work (and it will), his products will remain static but profitable until the next entrepreneur catalyzes *his* growth. ■

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Litton

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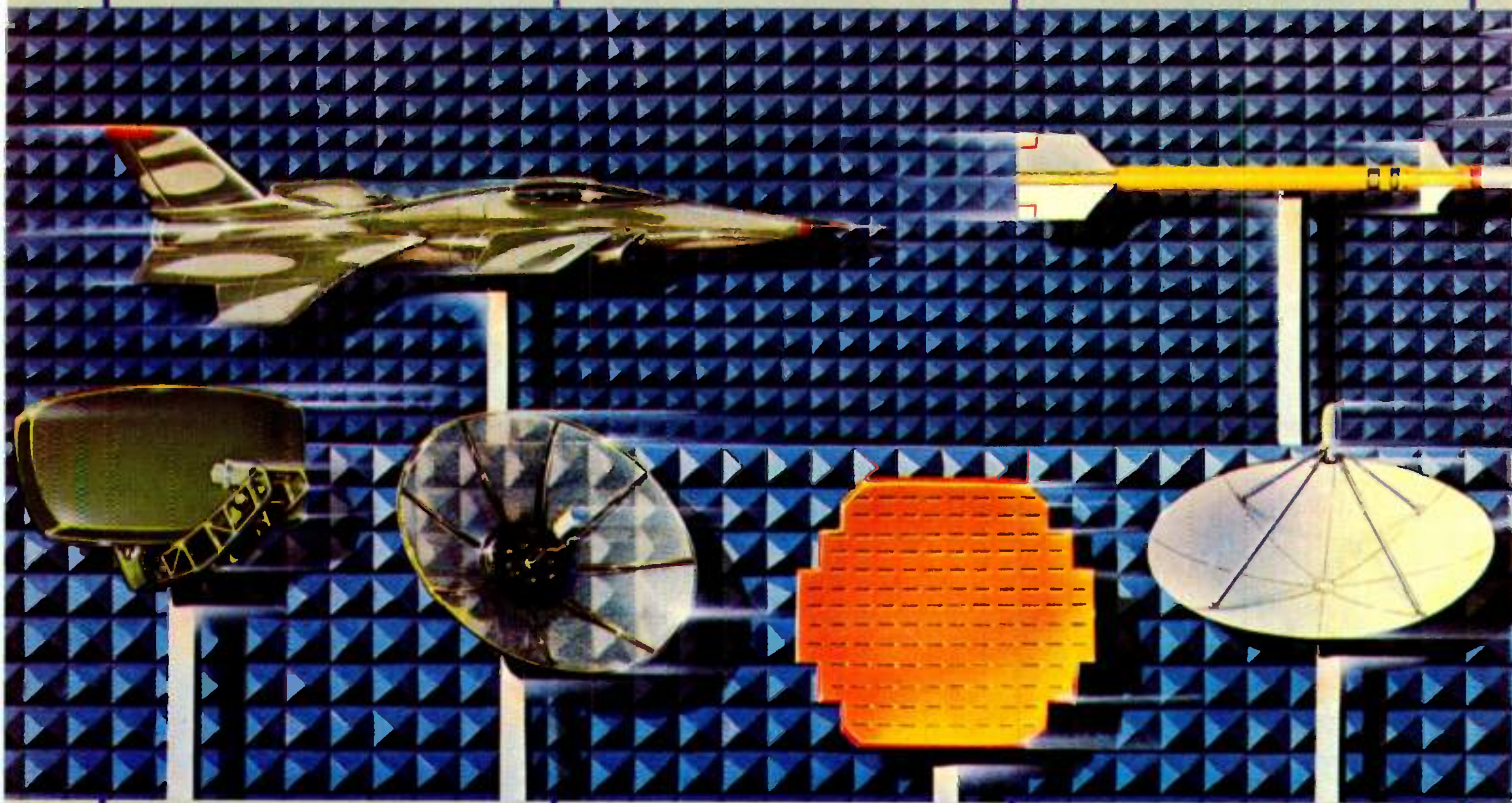

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BETTER CONTROL. Indoors, on a compact range, test operations are free from the whims of the weather. Let it rain, snow, let the wind blow, it won't affect your test measurements. Plus, the shielded construction of a compact range greatly reduces RFI/EMI interference.

INCREASED CONVENIENCE. Testing times can be reduced 50-95% with a compact range. And, because it can be located at design, production or maintenance facilities, transportation to and from the range is minimized.

SUPERIOR RESULTS. An indoor testing environment eliminates many variables that can affect test accuracy. Results are uniform and consistent on a compact range.

REDUCED COSTS. Unlike far field operations, compact ranges do not require extensive and expensive real estate commitments. On a compact range, your valuable staff can keep schedules and maintain productivity. And again, transportation—sometimes a significant expense—is lessened considerably with an on-site compact range.

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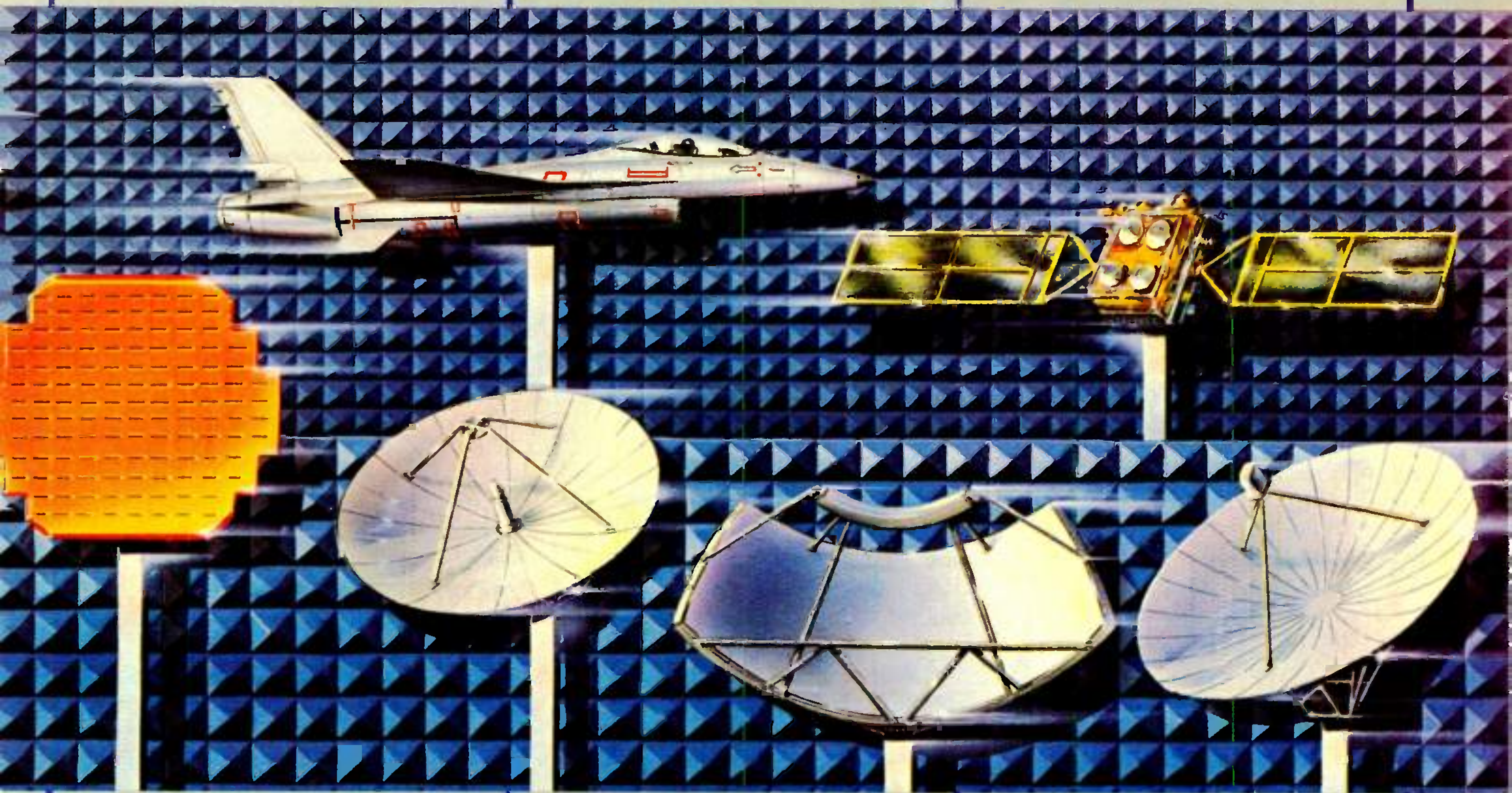

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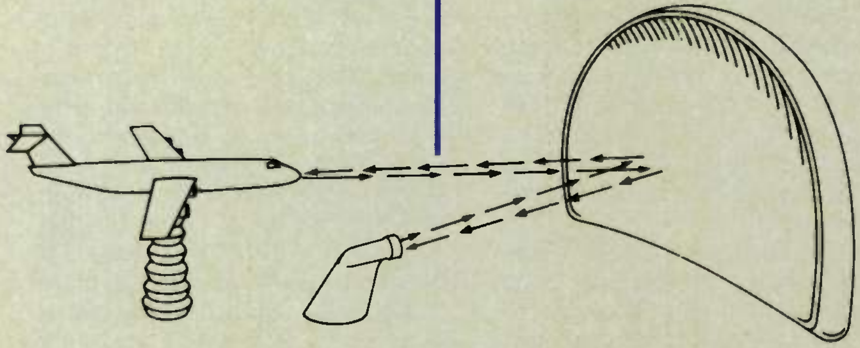

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The 1987 International Microwave Symposium: A Preview

Steven L. March
General Chairman

With over 500 conventions, meetings and symposia held there each year, Las Vegas can truly be labeled the "Convention Capital of the World." The Consumer Electronics Show and several of the IEEE Computer Society's meetings, with 70,000 to 100,000 attendees, use all of the city's 54,000 guest rooms; the exhibits occupy nearly all of the 1,000,000 square feet of space in the Las Vegas Convention Center. Probably no other city has the facilities that Las Vegas has to cater to the needs of the conventioneers and the exhibitors.

The IEEE is well known in Las Vegas; nearly everyone from room service personnel to taxi cab drivers to waitresses know that "IEEE" is the Institute for Electrical and Electronic Engineers. Some of the IEEE-sponsored or co-sponsored meetings that have been held in Las Vegas in recent years are:

- IEEE Nuclear and Space Radiation Conference, July 1982
- IEEE International Reliability Physics Symposium, April 1984
- 23rd IEEE Conference on Decision and Control, December 1984
- 22nd IEEE/ACM Design Automation Conference, June 1985
- IEEE International Reliability Physics Symposium, January 1986
- 23rd IEEE/ACM Design Automation Conference, June 1986.

Additionally, the 8th Annual Electrical Overstress/Electrostatic Discharge Symposium and the 13th International Symposium on Gallium Arsenide and Related Compounds were both held in September 1986 in Las Vegas. Expo Surface Mount Technology took place at the Tropicana Hotel at the beginning of October. The International Telemetry Conference was held last December at the Riviera Hotel, where it will be held again later this year.

In addition, many other professionals from abdominal surgeons to opticians to realtors have chosen to hold their annual conventions in Las Vegas. Other groups, such as the American Chemical Society, the American Society of Lubricating Engineers and the Instrument Society of America have opted to do likewise.

Forty years ago, just two resorts existed on what is now the famed Las Vegas "Strip." Today, more than three dozen luxurious hotels and casinos line both sides of the 3.5 mile stretch and, yearly, more than 12 million visitors stroll along this unique ribbon of real estate. The corner of Flamingo Road and Las Vegas Boulevard, boasting the presence of Bally's Grand Hotel (formerly the MGM Grand Hotel), Caesars Palace, the Barbary Coast (with the 2900-room Flamingo Hilton Hotel next door), and the Dunes on its four corners has replaced New York's Times Square as the busiest pedestrian intersection in the country.

In June 1987, this intersection will get even busier — nearly 7,000 registrants and exhibitors are expected to be in attendance during Microwave Week in Las Vegas. The week will begin with the sixth IEEE Microwave and Millimeter-Wave Monolithic Circuits Symposium on Monday and Tuesday, June 8 and 9. The 31st edition of the IEEE MTT-S International Microwave Symposium will take place June 9-11. The 29th ARFTG Conference and several workshops on June 12 and 13 will conclude the busy week of technical excellence.

The International Microwave Symposium's Steering Committee has arranged to nearly monopolize the corner of Las Vegas Boulevard and Flamingo Road with conference attendees. We have currently secured 2500 guest rooms at Bally's Grand Hotel, 1100 rooms at Caesars Palace, 500 rooms at the Flamingo Hilton and 400 rooms at the Dunes, with more coming. The room rates that have been contracted are lower than we have seen in many years and lower than we will again experience for many years. Rooms at Bally's Grand, the conference headquarters and the location where the technical sessions and the commercial exhibit will be held, will cost $75 per night. A rate of $65 per night has been secured at both Caesars Palace and the Flamingo Hilton. For those on a limited budget, such as government employees or those on a university per diem, rooms at the Dunes will cost only $50 per night. All rates are for run-of-the-house, single or double occupancy (suites excluded from these rates).

In addition, because McCarran International Airport is located within 10 minutes (and three miles) of "The Strip" hotels, transportation to the convention hotels will be inexpensive compared to transportation

[Continued on page 50]

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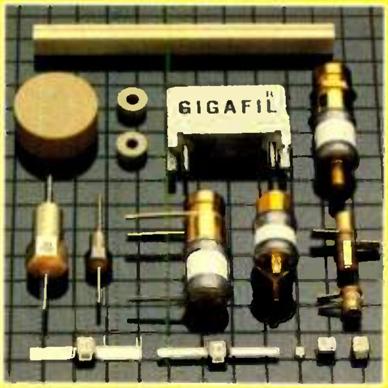

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[From page 48] MARCH

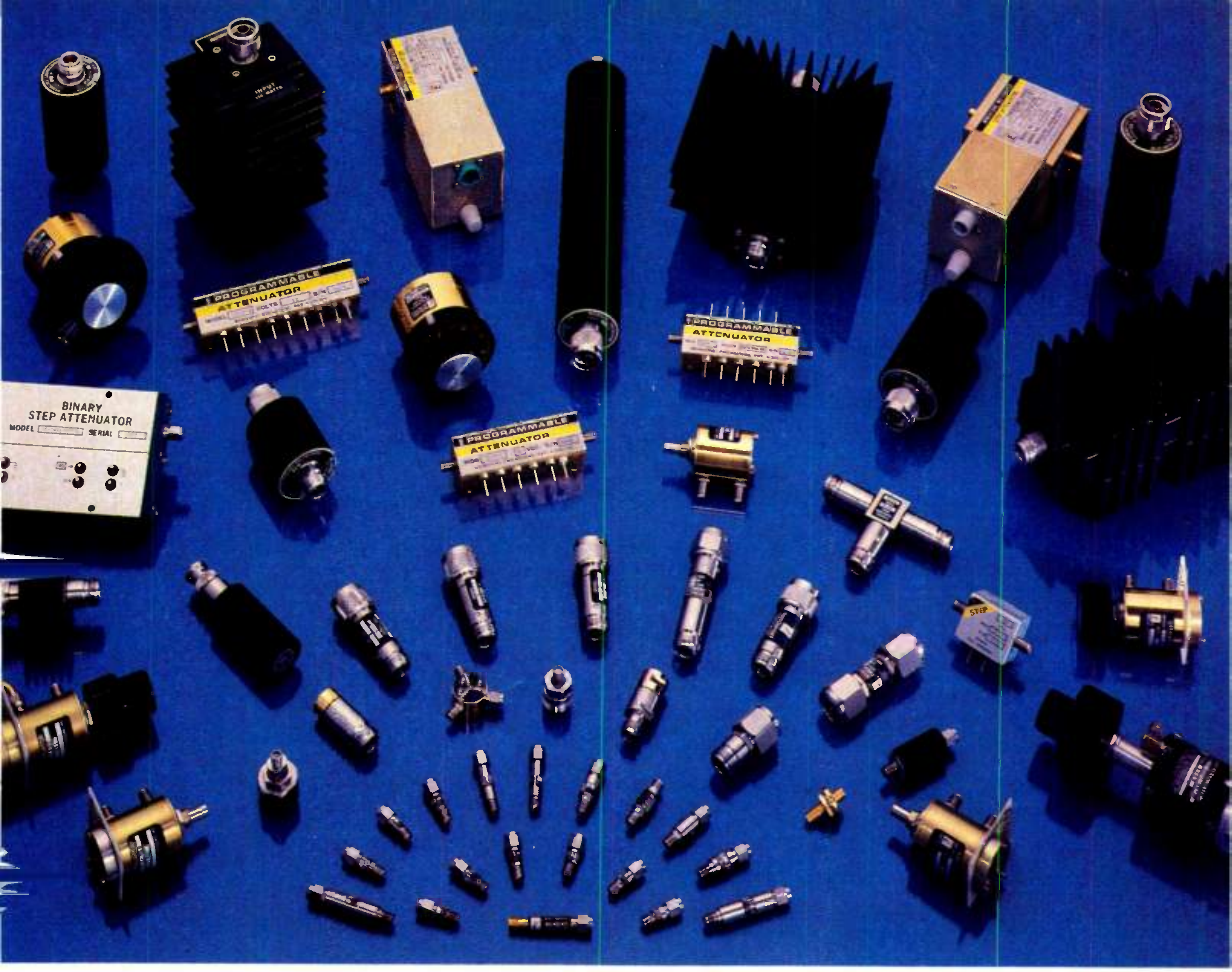
in other cities. A taxi cab ride should cost approximately $6.50. For most attendees, rental cars will not be needed in Las Vegas; this will help to keep expenses to a minimum.

The Steering Committee is arranging for Symposium registration fees to be payable by either VISA or MasterCard, making it easier for overseas attendees to pay their registration fees without resorting to arranging for drafts in US funds. The committee also is contemplating dividing the International Microwave Symposium Digest into a case-bound, two-volume set so that it will be easier to carry to technical sessions during the symposium. As has been the practice for the past two years, the symposium will again contract with United Parcel Service to provide, for a small fee, shipping of materials such as literature, magazines and registration materials collected by conference attendees.

The deadline for the submission of papers for possible inclusion at next year's IEEE MTT-S International Microwave Symposium is December 12, 1986. Submissions should be sent to Dr. Reynold S. Kagiwada, c/o LRW Associates, 1218 Balfour Drive, Arnold, MD 21012 in accordance with the instructions contained in the Call for Papers. For more information, contact Dr. Kagiwada at (213) 535-5515.

Recommendations for panel sessions and workshops are being actively solicited. If you have a proposal, please submit it to Dr. Tatsuo Itoh, The University of Texas at Austin, Department of Electrical Engineering, Austin, TX 78712, telephone (213) 471-1072. Recommendations for "focused sessions," begun this past June in Baltimore, also are welcome. The 100-plus member Technical Program Committee will meet on January 15, 1987 to select papers for presentation. These papers will form the nucleus around which the remainder of the Symposium will be formed.

By the time the 1987 edition of the International Microwave Symposium arrives next June, many of the Steering Committee members will have worked on this project for over four years. We are going to make it the best technical meeting that the MTT-S has ever had — the one to become the standard by which all future symposia will be judged. ■



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Model Number	Frequency (GHz)	Gain (Min.) (dB)	Gain Variation (Max.) (± dB)	Noise Figure (Max.) (dB)	VSWR (Max.) Input	VSWR (Max.) Output
100 MHz Bandwidth						
AFD3-005006-09	.5-.6	32	.25	0.9	1.5:1	1.5:1
AFD3-006007-09	.6-.7	32	.25	0.9	1.5:1	1.5:1
AFD3-007008-09	.7-.8	32	.25	0.9	1.5:1	1.5:1
AFD3-008009-09	.8-.9	34	.25	0.9	1.3:1	1.5:1
AFD3-009010-09	.9-1.0	34	.25	0.9	1.3:1	1.5:1
AFD3-010011-09	1.0-1.1	34	.25	0.9	1.3:1	1.5:1
AFD3-011012-09	1.1-1.2	34	.25	0.9	1.3:1	1.5:1
AFD3-012013-09	1.2-1.3	34	.25	0.9	1.3:1	1.5:1
AFD3-013014-09	1.3-1.4	34	.25	0.9	1.3:1	1.5:1
AFD3-014015-09	1.4-1.5	34	.25	0.9	1.3:1	1.5:1
AFD3-015016-09	1.5-1.6	34	.25	0.9	1.3:1	1.5:1
AFD3-016017-09	1.6-1.7	34	.25	0.9	1.3:1	1.5:1
AFD3-017018-09	1.7-1.8	34	.25	0.9	1.3:1	1.5:1
AFD3-018019-09	1.8-1.9	34	.25	0.9	1.3:1	1.5:1
AFD3-019020-09	1.9-2.0	34	.25	0.9	1.3:1	1.5:1
AFD3-020021-09	2.0-2.1	34	.25	0.9	1.3:1	1.5:1
AFD3-021022-09	2.1-2.2	34	.25	0.9	1.3:1	1.5:1
AFD3-022023-09	2.2-2.3	34	.25	0.9	1.3:1	1.5:1
200 MHz Bandwidth						
AFD3-005007-10	.5-.7	32	.35	1.0	1.5:1	1.5:1
AFD3-006008-10	.6-.8	32	.35	1.0	1.5:1	1.5:1
AFD3-008010-10	.8-1.0	34	.35	1.0	1.5:1	1.5:1
AFD3-010012-10	1.0-1.2	34	.35	1.0	1.3:1	1.5:1
AFD3-012014-10	1.2-1.4	34	.35	1.0	1.3:1	1.5:1
AFD3-014016-10	1.4-1.6	34	.35	1.0	1.3:1	1.5:1
AFD3-016018-10	1.6-1.8	34	.35	1.0	1.3:1	1.5:1
AFD3-018020-10	1.8-2.0	34	.35	1.0	1.3:1	1.5:1
AFD3-020022-10	2.0-2.2	34	.35	1.0	1.3:1	1.5:1
AFD3-022024-10	2.2-2.4	34	.35	1.0	1.3:1	1.5:1
AFD3-024026-10	2.4-2.6	34	.35	1.0	1.3:1	1.5:1
AFD3-027029-10	2.7-2.9	34	.35	1.0	1.3:1	1.5:1
AFD3-029031-10	2.9-3.1	30	.35	1.0	1.3:1	1.5:1
500 MHz Bandwidth						
AFD3-005010-10	.5-1.0	32	.5	1.0	2:1	2:1
AFD3-007012-10	.7-1.2	32	.5	1.0	1.5:1	1.5:1
AFD3-010015-10	1.0-1.5	35	.5	1.0	1.5:1	1.5:1
AFD3-015020-10	1.5-2.0	30	.5	1.0	1.5:1	1.5:1
AFD3-020025-13	2.0-2.5	30	.5	1.3	1.5:1	1.5:1
AFD3-025030-13	2.5-3.0	30	.5	1.3	1.5:1	1.5:1
AFD3-030035-13	3.0-3.5	30	.5	1.3	1.5:1	1.5:1
AFD3-037042-13	3.7-4.2	30	.5	1.3	1.5:1	1.5:1
700 MHz Bandwidth						
AFD3-017024-14	1.7-2.4	35	.5	1.4	1.5:1	1.5:1
1000 MHz Bandwidth						
AFD3-014024-14	1.4-2.4	30	.5	1.4	1.5:1	1.5:1
AFD3-010020-11	1.0-2.0	30	.5	1.1	1.5:1	1.5:1
1500 MHz Bandwidth						
AFD3-005020-13	.5-2.0	30	1.0	1.3	2:1	2:1
AFD3-015030-15	1.5-3.0	30	1.0	1.5	2:1	2:1
2000 MHz Bandwidth						
AFD3-020040-15	2-4	28	1.0	1.5	2:1	2:1

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News from Washington

APG-68/F-16 Radar Gets Initial VHSIC Units

Westinghouse Defense and Electronics Center has delivered the first three key functional parts of the APG-68/F-16 VHSIC programmable signal processor (VPSP). When installed, the units will constitute one of the first insertions of VHSIC technology into a production system. The VPSP will capitalize on Phase 1 VHSIC technology (1.25 micron) being developed to achieve benefits in processing performance, life cycle cost and reduction in size, weight and power.

The VHSIC insertion is part of a 28-month program awarded to Westinghouse by USAF's Aeronautical Systems Division for $13.8 M.

Development of C^2I for FAAD Initiated by Army

The Army Missile Command has initiated the development of a command, control and intelligence (C^2I) system as part of the Army's forward area air defense system (FAADS). FAADS is a tactical system designed to protect Army ground units from attacking enemy aircraft.

A major objective of the program is to automate key functions of command and control. Acquisition and tracking of incoming air threats, passing of friendly and unfriendly identification, and alerting and cueing of FAADS weapons are examples of functions to be automated.

The Missile Command has awarded a $58 M contract to TRW Defense Systems Group to develop the software and integrate a system to link the weapons, sensors and all the C^2 elements of FAADS. Data processing and display hardware and software will be GFE. TRW also will integrate the sensors and identification devices to enhance the effectiveness of weapons including the Stinger, Chaparral and Patriot Missile Systems and the Vulcan Air Defense System.

TRW subcontractors include Hughes Aircraft Co., the Command, Control and Communications Corp. and Ford Aerospace and Communications Corp.

Services to Merge ICNIA/INEWS, ICNIA Contract Restructured

The Air Force and the Navy have agreed to merge their ICNIA (integrated communications, navigation and IFF avionics) and INEWS (integrated EW system) programs, according to Assistant Secretary of Defense for C^3I Donald C. Latham.

In related action that illustrates the growing importance of electronics, the House Appropriation Committee is reportedly directing the Air Force to delay its ATF program to match the schedule of ICNIA/INEWS.

The Air Force, meanwhile, is restructuring its ICNIA contract by dropping ITT and retaining TRW as prime contractor.

Shared Aperture Program Advances

A Navy development program aimed at integrating communications, radar, IFF and electronic warfare antenna functions into a shared aperture is making progress. GE, Hughes, Raytheon, Sanders, TI, and Westinghouse are participating in the program.

Phase 1A (concept development) is expected to be completed in the first quarter of FY '87. Results are reported to be "very promising." Phase 1B (concept feasibility demonstration) awards are scheduled for the third quarter of FY '87. Phase 2 (advanced development), scheduled for FY '89, will include construction of flyable models. Phase 3 (preproduction) is scheduled for FY '96.

The Air Force has a number of complementary projects. The services are expected to combine their efforts in FY '87. ■

You Make the Choice

And M/A-COM will supply a customized Rotary Joint to meet your specific requirement in coaxial, waveguide or multi-channel configuration. Let us work for you in applying over 30 years of design and manufacturing experience to service any military or commercial requirement you may have now or in the future. Our experience covers all frequency ranges from DC to 50 GHz. M/A-COM Microwave Components, Inc. maintains a highly qualified staff of microwave specialists to assist you. Complete information is now available in a new comprehensive catalog. Please contact:

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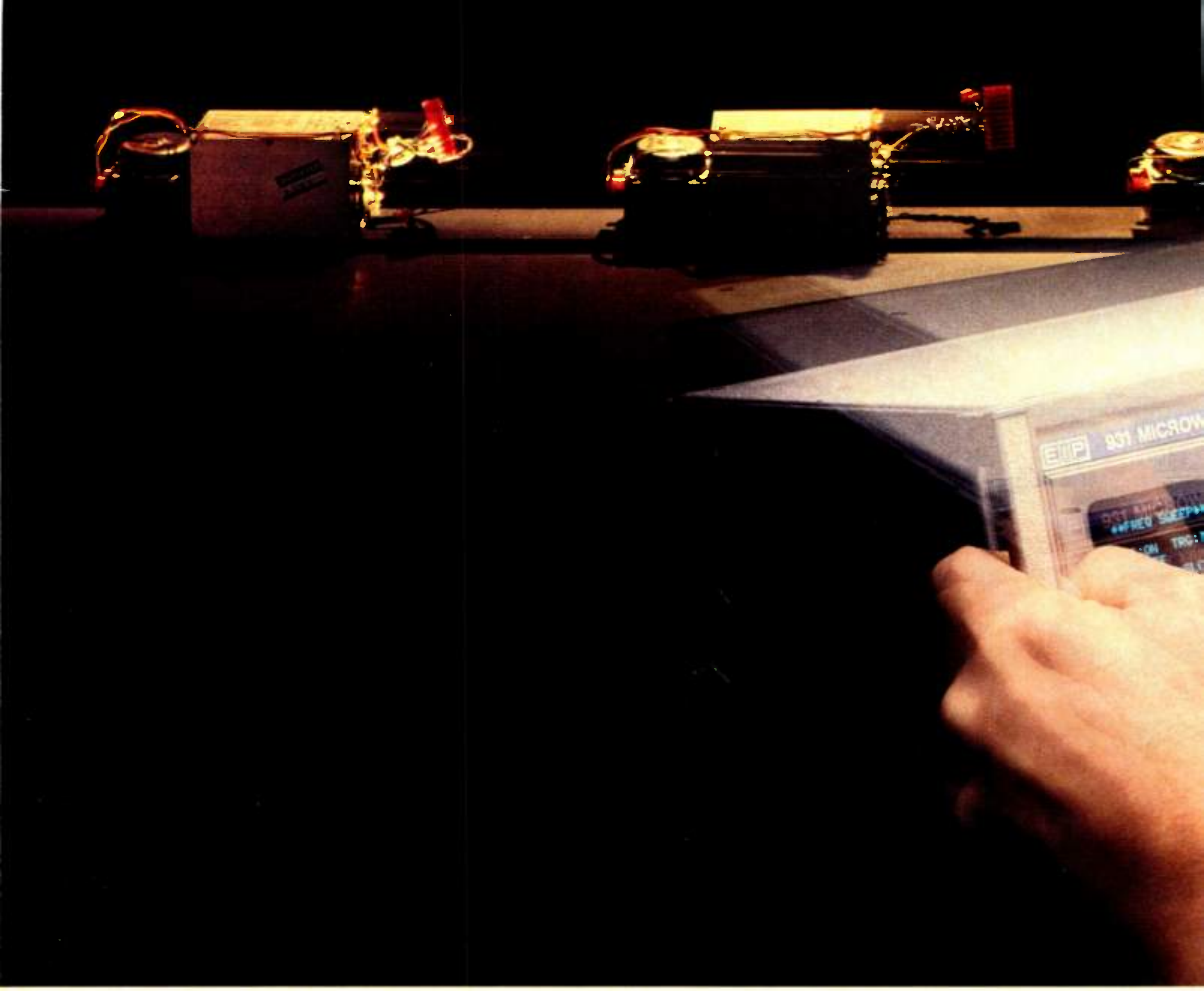
M/A-COM

CIRCLE 44 ON READER SERVICE CARD



Rapid Fire Testing

EIP's new Microwave Sweeper

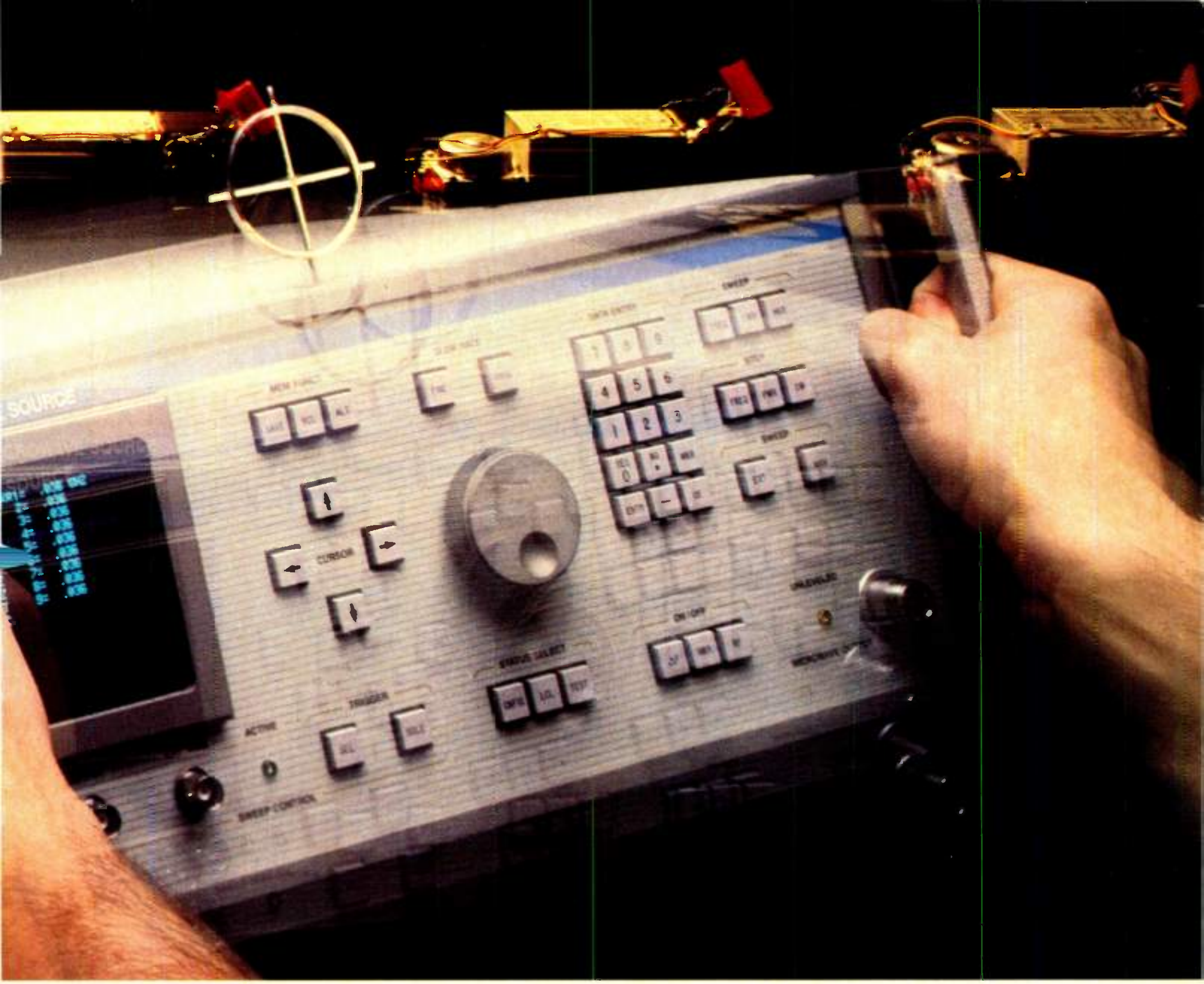
EIP's new family of microwave sweepers help you move product, fast! This unique new sweeper speeds the testing process with features tailored to demanding production requirements.

+13 dBm of Leveled Output Power over a Broad Frequency Range

The EIP 930 Series delivers from −70 dBm to +13 dBm of leveled output power, with power leveling as precise as 0.05 dB at the front panel, or at an external test point. Frequency range covers 10 MHz to 18.6 GHz, with 1 MHz resolution. An optional pulse modulator provides fast, high isolation pulse modulation.

Easy and Fast to Use.

Consider the time you'll save, and the testing bottlenecks you'll eliminate, with these features: Nine programmable test set-ups; automatic frequency and power steps, plus "on the bench" frequency and



power calibration. The 931's CRT display not only shows instrument status, but it can be remotely programmed as a terminal to interactively communicate with the operator. And for fully automated test systems, the 931 is tailored for remote programming over the IEEE bus.

Self Calibration—It's About Time

Power and frequency calibrations can at last be performed on the bench with only a power meter and a counter. And with only 19 internal adjustments, (compared with over 80 you're faced with in competitive units), a full instrument calibration can be completed in a fraction of the time you've had to spend in the past.

So to learn more about what is certain to become the new standard for performance and value, and to see what the 930 Series can do to speed up testing at your facility, line up those modules and call Product Manager Roger Shaw at (408) 946-5700 for a demonstration.

EIP Microwave, Inc.
2731 North First Street
San Jose, CA 95134
(408) 946-5700
TWX: 910-338-0155

The Tradition Breaker

EIP

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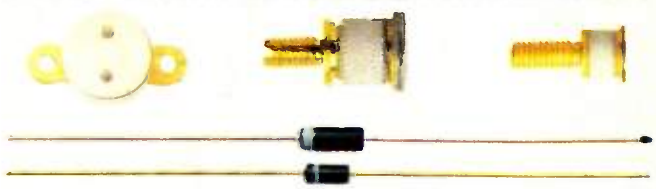
MORE COMPETITIVE

THOMSON SI PIN DIODES 30 V-3000 V / 1 MHz - 100 GHz
100 % AVAILABILITY 100 % CAPABILITY 100 % RELIABILITY



These diodes are 100 percent.

Especially, we have developed a new line of Si PIN diodes: these products provide low-series resistance; low-cost package; breakdown voltage from 30 to 3000 V; and fast switching time. They're designed to be used in phase shifters; very fast SPNT switches; and in RF/VHF/UHF devices that use frequency hopping. Thomson Si PIN diodes. The one hundred percenters.

We also offer you a complete line: we've got exactly what you want, exactly when you want it. Including:
- power generating diodes (Gunn and IMPATT),
- receiving diodes (Schottky and PIN),
- tuning and control diodes.

Thomson Hybrid and Microwave Components offers a broad line of microwave and optoelectronic components including: GaAs and Si semiconductors; Ferrite material and devices; Microwave Hybrid Circuits (GaAS amplifiers and oscillators etc...); Semiconductor lasers.

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Components Corporation
Special Products Division
301 Route 17 North
RUTHERFORD NEW-JERSEY 07070
Tel.: (201) 438.2300
Telex: 7109897 286 Rtfd THCSF.

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SEMICONDUCTEURS
HYBRID AND MICROWAVE COMPONENTS DIVISION

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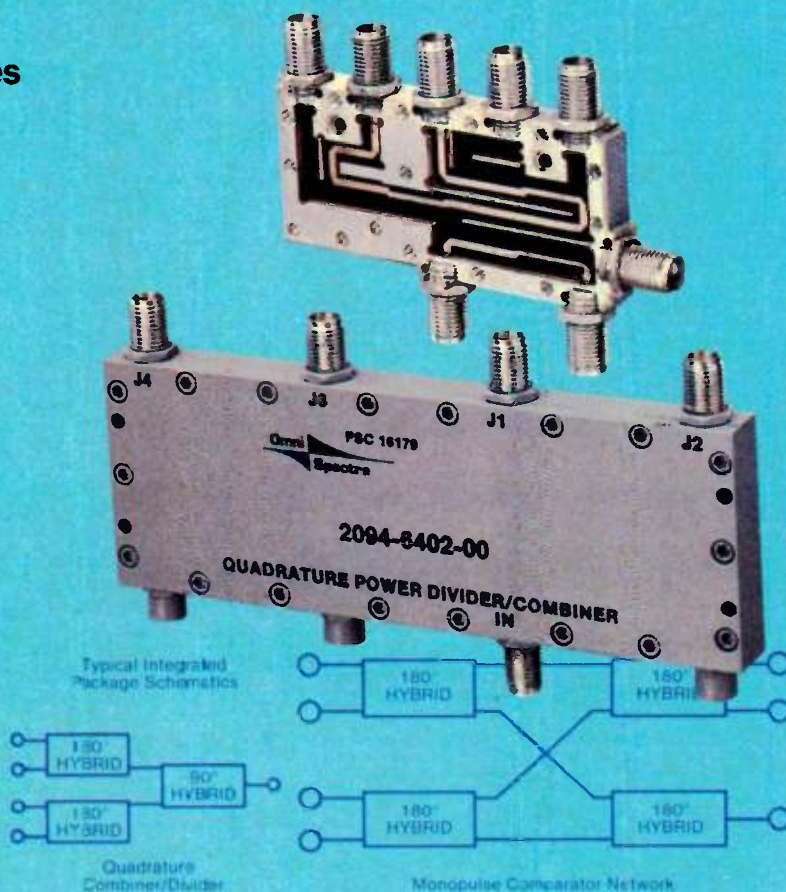
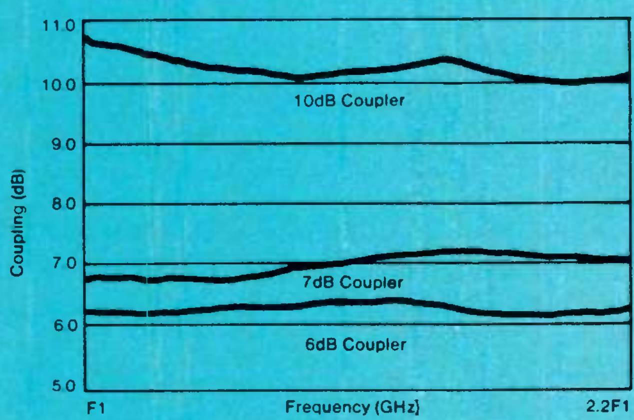
PROVEN PERFORMANCE . . .

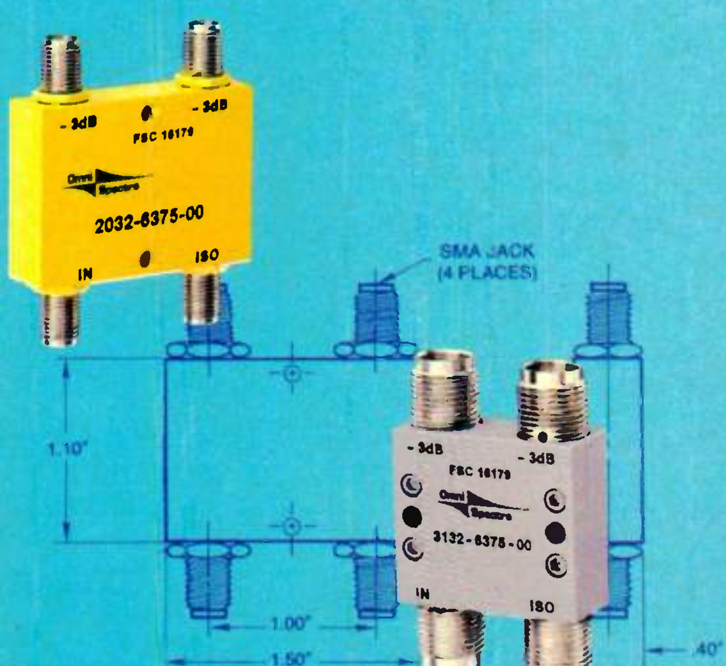
Multicouplers and Integrated Packages

- Octave and Multi Octave Bandwidths
- Stripline Construction
- OSM® (SMA) Connectors Approved to MIL-C-39012
- Multi-Function Packages
- Custom Designs Available





Typical Multicoupler Coupling/Flatness Data



High Power 90° Hybrid Couplers

- Small Package Size
- 100 Watts Average Power
- Low Insertion Loss
- OSM® (SMA), N or Mode Free TNC Connectors to 18.0 GHz Available
- Custom Designs Available

High Power 90° Hybrid Couplers

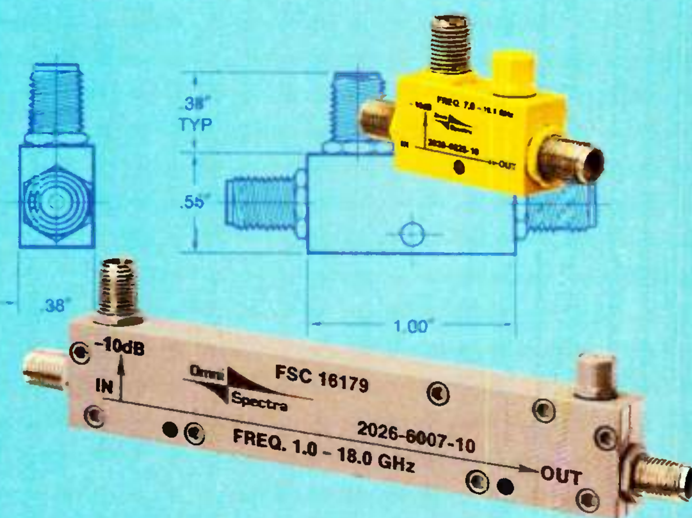
Specifications	2032-6375-00	3132-6375-00
Frequency	4.0-18.0 GHz	6.0-18.0 GHz
Amplitude Balance	±0.5 dB	±0.75 dB
Insertion Loss	0.80 dB	0.60 dB
Isolation	18 dB	15 dB
VSWR	1.45	1.50
Power In (Average)	100 W	100 W
Power In (Peak)	5000 W	5000 W

M/A-COM Omni Spectra, Inc.
21 Continental Boulevard
Merrimack, New Hampshire 03054
(603) 424-4111

Couplers, Hybrids and Custom Assemblies from M/A-COM Omni Spectra

Octave Band (Mini) and Broad Band Couplers

- .5 Through 18.0 GHz Frequency Ranges
- Smallest and Lightest Couplers Available
- High Directivity — Low VSWR
- Meets MIL-E-5400 and MIL-E-16400 Environments
- Custom Designs Available

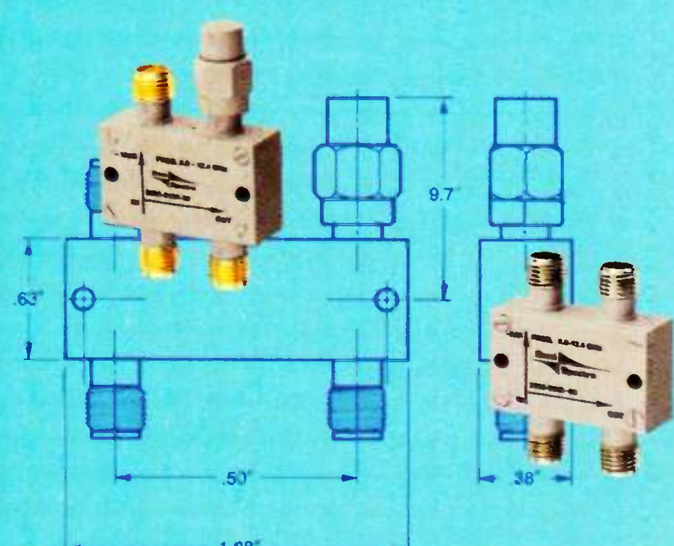


Octave Band (Mini) and Broad Band Couplers

Specifications	2020-6625-10	2026-6007-10
Frequency	7.0-18.0 GHz	1.0-18.0 GHz
Coupling (Incl. Sens.)	10 ± 1 dB	10 ± 1 dB
Freq. Sensitivity	±0.50 dB	±0.50 dB
Insertion Loss (Max.)	0.40 dB	1.20 dB
Directivity	20 dB	15 dB 1.0-12.4 GHz 12 dB 12.4-18.0 GHz
VSWR (Prim. Line)	1.30	1.40
VSWR (Sec. Line)	1.40	1.40
Power In (Average)	50 W	50 W
Power In (Peak)	3000 W	3000 W

Airline Hybrid and Directional Couplers

- Air Dielectric for Low Loss
- 1.0 to 18.0 GHz Frequency Ranges
- Octave Bandwidths
- MIL-C-15370/8 and MIL-C-15370/10 Equivalents
- Custom Designs Available



Airline Hybrids and Directional Couplers

Specifications	2035-6368-00	2023-6134-10
Frequency	8.0-12.4 GHz	8.0-12.4 GHz
Coupling (Incl. Sens.)	3 dB	10 ± 1 dB
Amplitude Balance	±.5 dB	—
Freq. Sensitivity	—	±0.75 dB
Insertion Loss (Max.)	0.35 dB	0.25 dB
Directivity	—	15 dB
Isolation	15 dB	—
VSWR	1.30	1.30
Power In (Average)	50 W	5 W
Power In (Peak)	3000 W	2000 W

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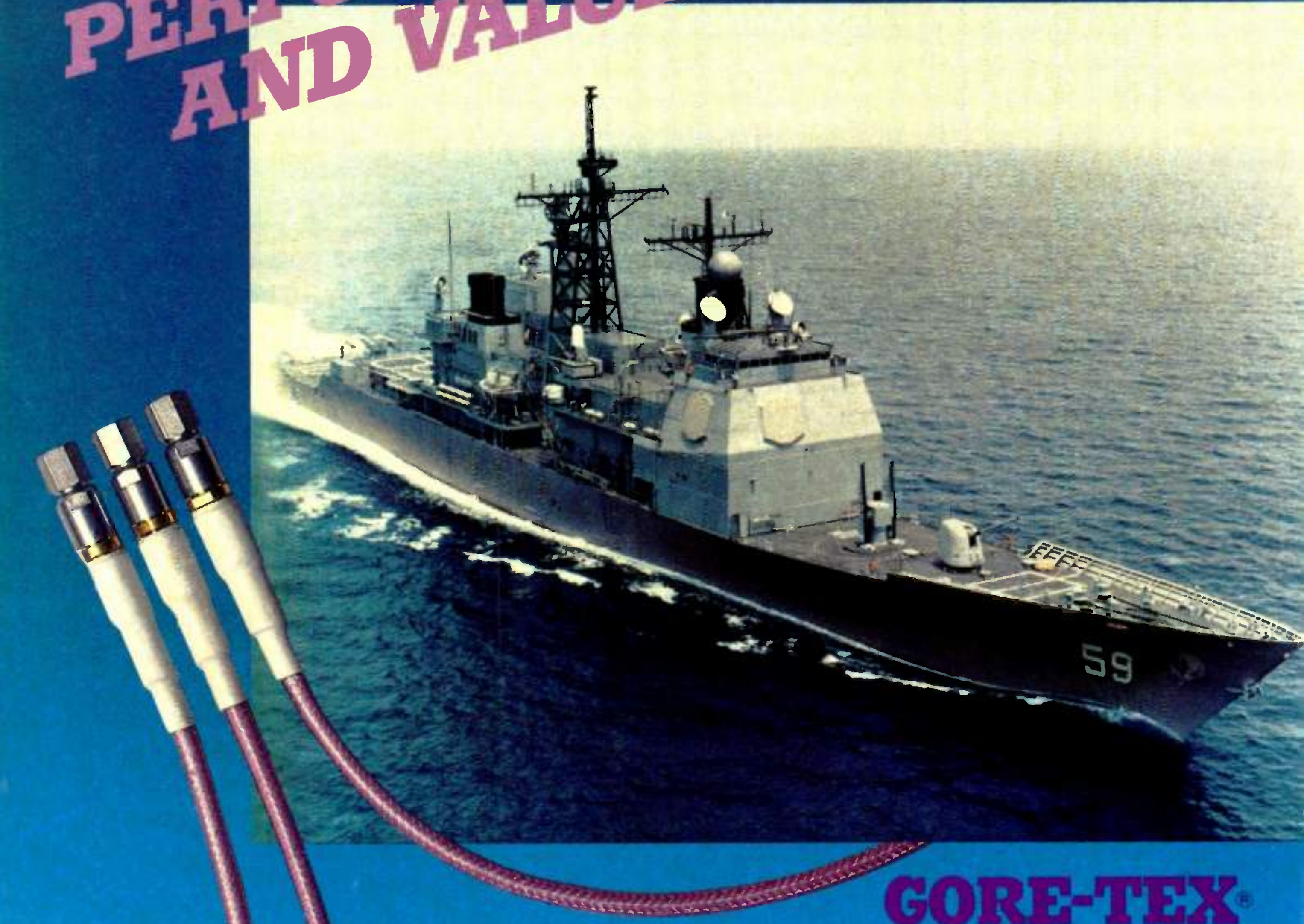
CIRCLE 48 FOR IMMEDIATE NEED



The Commercial Market

Stephen Shaw, *Correspondent*

FCC Releases Order to Simplify Equipment Authorization Requirements

The FCC has released an order simplifying the identification requirements for equipment covered under the authorization program in Part 2 of Title 47 of the Code of Federal Regulations.

The commission amended its rules to permit the marketing of multiple models of electrically identical equipment under the same FCC identifier without the need for additional filings.

"Experience with the equipment identification rules has shown that to require separate filings for new models of electrically identical equipment is unnecessary. Differences among such devices usually extend only to cabinet color or style and have no bearing on the devices' compliance with the commission's authorization process," the Order stated.

Court Reverses Pre-emption of Intrastate RCC's

The US Court of Appeals for the District of Columbia circuit has reversed an FCC order pre-empting state regulation of intrastate radio common carrier (RCC) services. The commission had pre-empted state regulation that restricted new entry of RCC services provided on FM subcarrier frequencies.

The court ruled that the commission lacked authority to pre-empt such restrictive legislation, stating that such a policy must be formulated by the US Congress, not the FCC. The Communications Act, the appeals court pointed out, reserves to the states authority to regulate intrastate wire or RCC services. The FCC's pre-emption would subject intrastate common carrier services to federal control solely on the basis that they utilize radio signals, a use expressly prohibited by Congress.

The appeal of the FCC's decision was filed by the California Public Utilities Commission.

AID Makes Available $50,000 for Third World Telecommunications Equipment Purchases

The US Agency for International Development (AID) has made available $50,000 in matching funds for the Center for Telecommunications Development, an arm of the International Telecommunications Union. The Center's purpose is to assist developing countries to plan and implement their basic telecommunications infrastructures. Manufacturers from developing countries, expected to be the source for much of the equipment, also are expected to benefit from the center's activities.

A number of US companies had reacted favorably to the center, but had withheld financial support until assurances from the US government were forthcoming that it would support the center. The $50,000 AID grant is intended to provide that assurance.

The AID monies must be matched by contributions from the private sector. The US Telecommunications Suppliers Association is leading the private sector initiative to raise funds for the effort.

Rural Electric Co-ops Form New Telecommunications Council

The nation's 1,100 rural utility cooperatives have agreed to form a new organization to provide a wide range of satellite-based telecommunications services for rural electric and telephone systems. The National Rural Telecommunications Cooperative (NRTC) was formed jointly by the National Rural Electric Cooperative Association and the National Rural Utilities Cooperative Finance Corp., which contributed $150,000 in seed capital to fund the new organization.

[cont.]

The Commercial Market

Co-ops [cont.]

NRTC will concentrate on developing four major telecommunications projects for the rural cooperatives. The first is a package of satellite television programming to be sold to consumers with backyard earth stations located primarily in, but not restricted to, rural areas. Second, NRTC is developing a pilot satellite data/video network to interconnect a number of cooperatives for data processing and training applications. Finally, the organization will be involved in setting standards and marketing satellite earth equipment for both members and non-members.

Intelsat Sells Transponders to Iran

The 112-nation International Telecommunications Satellite Organization (Intelsat) has approved the sale of two transponders to the Islamic Republic of Iran for domestic public television network use. The international satellite consortium also recently sold three additional transponders, for a total of six, to the Federal Republic of Germany for use in the West German public television network.

The first transponder to Iran is scheduled to begin service November 1 to broadcast television programming to more than 100 Iranian cities and villages.

Iran and West Germany join six other countries that have purchased transponders for domestic television use under Intelsat's Planned Domestic Service offering, first introduced in December of last year. Those countries are Gabon, Israel, Italy, Norway, the People's Republic of China and Turkey.

Fowler Places Ultraphone Digital Call

Mark Fowler, chairman of the FCC, recently placed the first all-digital radio telephone call as he initiated an FCC experiment using the new ultraphone technology.

The service, provided under a temporary FCC authorization, is provided by Mountain Bell, a US West company, using radio microwave channels and digital encoding techniques to substitute for traditional wireline telephone service.

The new wireless telephone service potentially could serve the approximately 500,000 US residents who are beyond conventional telephone service areas.

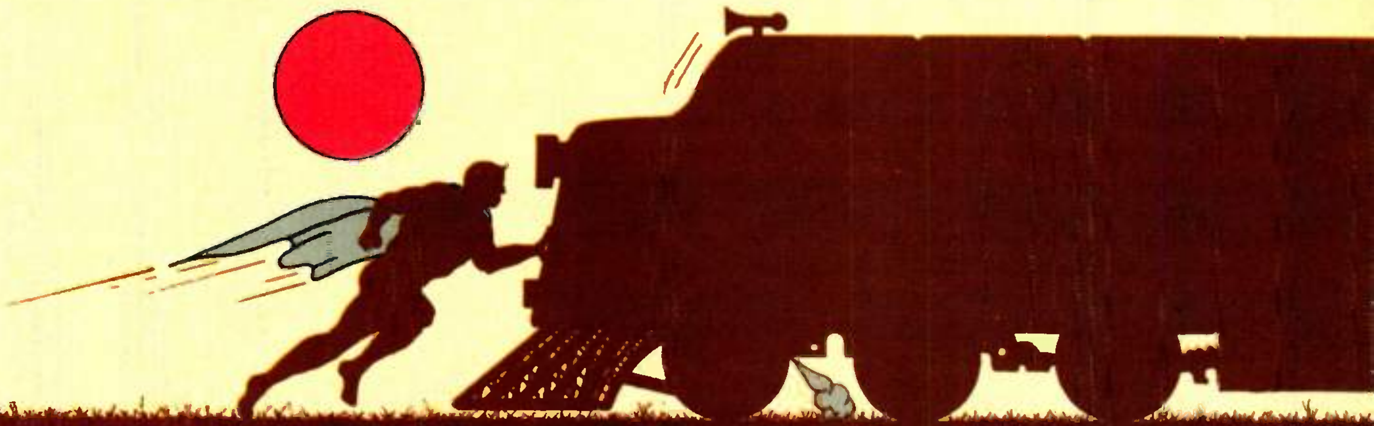
Commission Dismisses Proposal for Computer Radio Service

The FCC has dismissed a petition for rulemaking filed last year to create a new Public Digital Radio Service. As proposed, the service would create a radio alternative to the public switched telephone network for computer-to-computer communications. Donald L. Stoner, who filed the petition, requested exclusive assignment of the 52 to 54 MHz frequency band for the proposed service.

The commission received 45 comments on the proposal. More than half, mostly from personal computer users, were in favor of the proposed service. Nineteen comments from amateur radio operators argued that the proposal should be rejected.

The commission rejected the proposal, arguing that the adoption of a non-amateur digital radio service in the 52 to 54 MHz spectrum would conflict with international spectrum allocations as regulated by the International Telecommunications Union. The commission also said that re-allocation of the proposed spectrum would adversely affect existing amateur operators in the band.

The commission concluded, however, "This action should not discourage alternative proposals to establish a new computer hobbyist radio service." ■

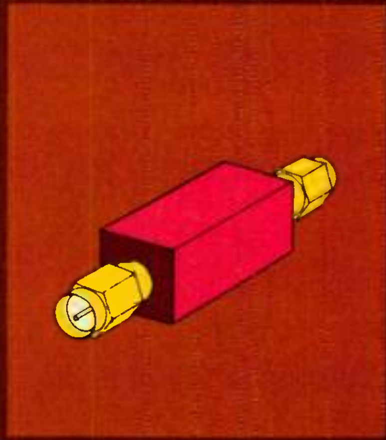


POWERFUL...

High Power Limiters

Most multi-octave limiters fail when faced with power bursts of more than a few microseconds. And they can't tolerate CW power except at very low levels. We have the ***only*** Broadband High Power Passive Limiter on the market that is designed to stand up to and survive peak powers in excess of 500 watts at 15 microseconds pulse widths as well as CW levels of 50 watts @ 85°C!

This high performance miniature device is the result of years of successful development and experience supplying microwave diodes and control devices to demanding military specifications. Call us for custom designs including connectorized, modular, or MIL STD-883-screened hermetic devices.



TYPICAL PERFORMANCE (Model 8108-CK)

Frequency Range	4 to 18 GHz	Peak Power	500 watts @ 15 μs PW
Insertion Loss	2.5 dB	CW Power	50 watts
VSWR	2.0:1	Leakage	50 mW CW
Recovery Time	5 μs		

Send for complete specifications

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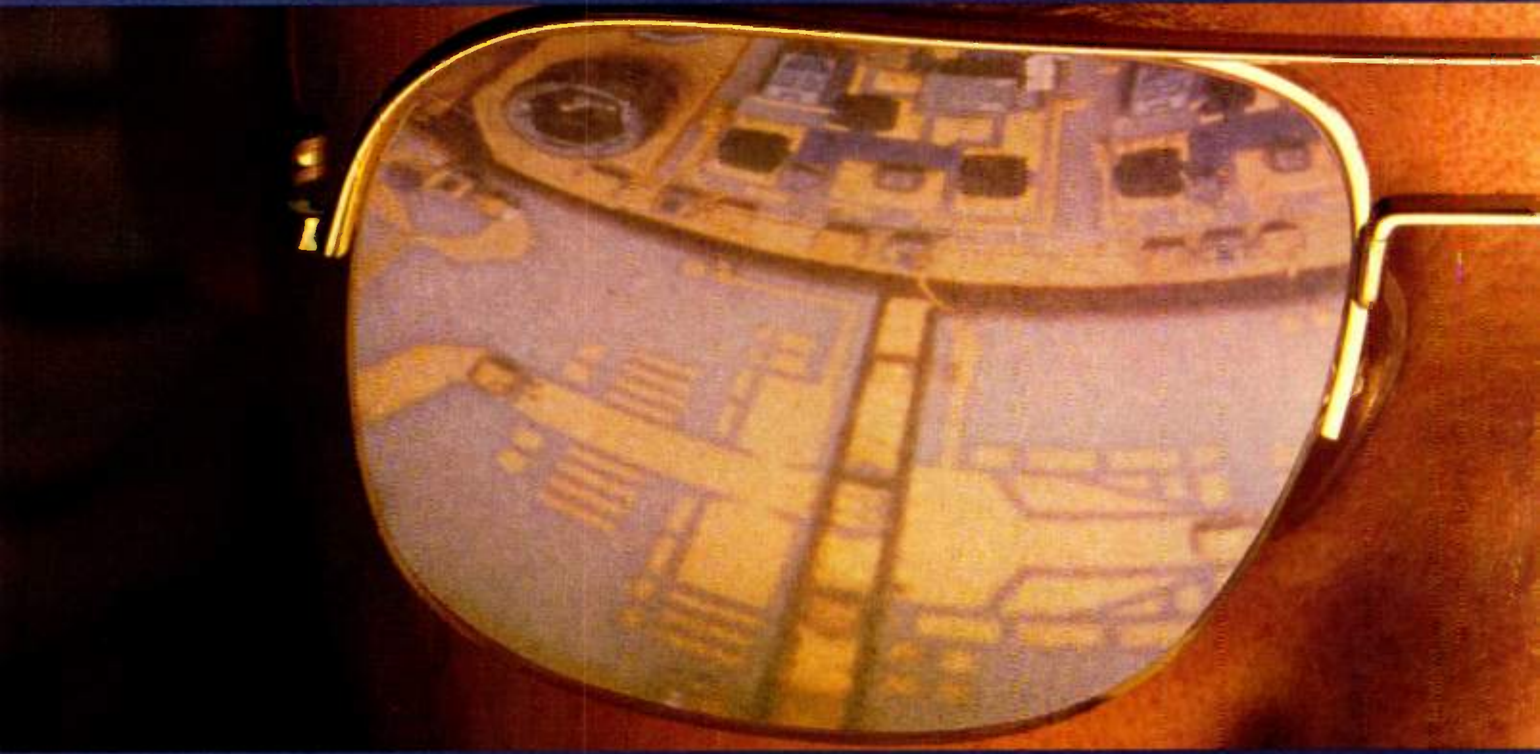
CIRCLE 40

EXTENDING THE



YOU DON'T HAVE TIME TO GET STUCK IN AN AMPLIFIER DESIGN

You've got a million things to think about, and a high power amplifier is just one of them. Get yourself too involved in details like that and pretty soon you're out of time completely.

On the other hand, if you're not looking closely enough, you can find yourself stuck with a solution to somebody else's problem instead of the unique amplifier your system needs.

CALL EPSCO IN EARLY

Because the microwave or RF amplifier typically defines the performance of the whole system, you can't afford to compromise on it.

The best thing to do is get Epsco working on the design as early as possible. That way, details like intermodulation, phase noise, size, and thermal environment can be resolved along with the basic requirements like frequency range, gain, and output power.

TYPICAL MICROWAVE POWER AMPLIFIER DESIGNS

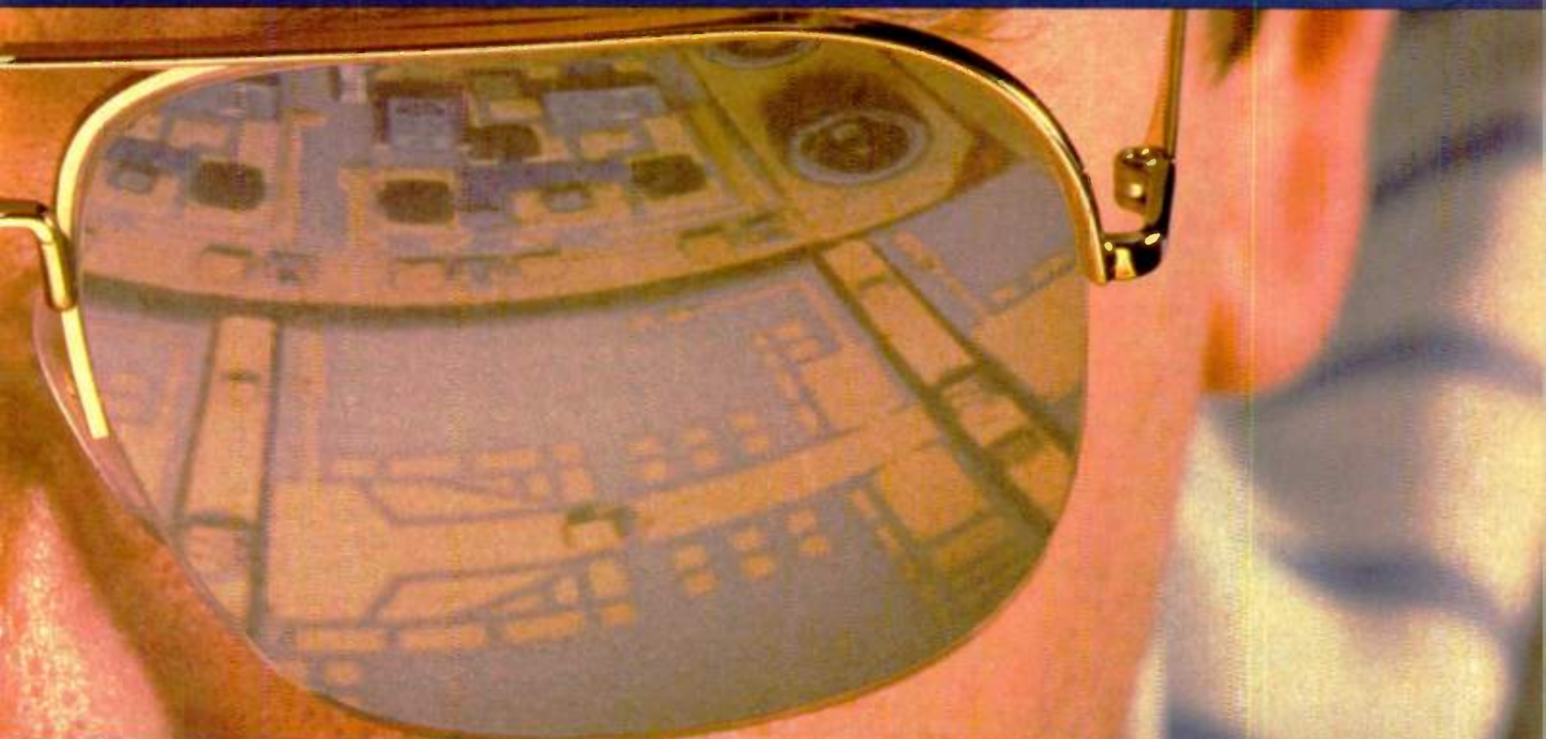
Frequency (GHz)	Output (dBm)	Gain (dB)
2.0 - 4.0	37	27
5.9 - 6.4	40	40
9.5 - 10.0	40	38
14.0 - 14.5	34	28

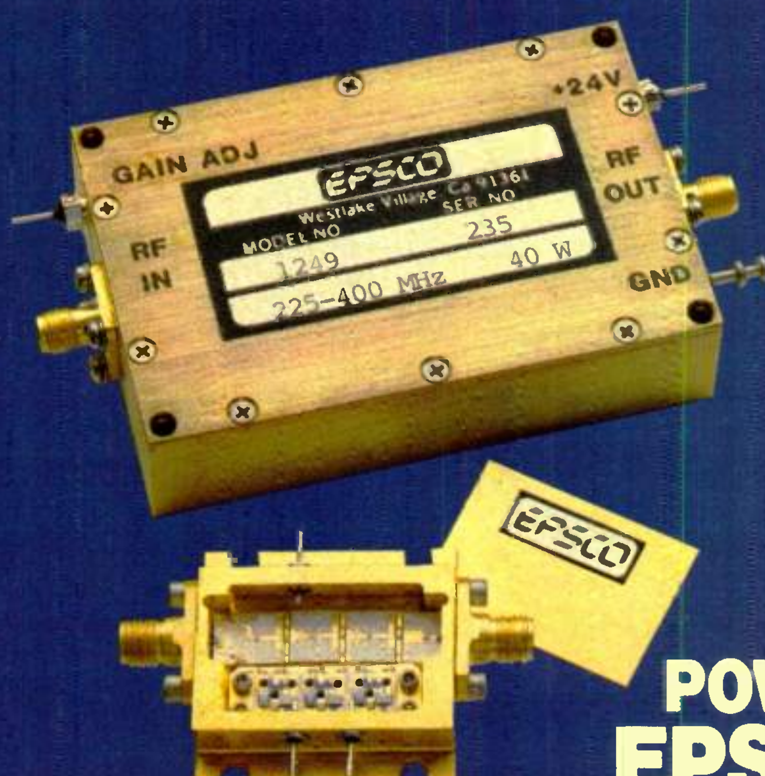
TYPICAL RF MODULE AND HPA DESIGNS

Frequency (MHz)	Output (dBm)	Gain (dB)
20 - 100	46	46
225 - 400	45	45
500 - 1000	43	43
20 - 100*	60	60
225 - 400*	53	53
100 - 500*	56	56

*HPA Designs

LIMITS OF POWER





HIGH POWER EPSCO

From 1MHz to 16GHz and output power levels to 10W in microwave and 10KW in RF, Epsco has amplifiers and designs that can be quickly tailored to your specifications. We have complete in-house capability in both microwave and RF technologies. And our experienced design teams have all the state-of-the-art tools needed to solve your most critical requirements (and keep you freed up for better things).

For more information on our RF capabilities, contact Epsco, RF Division, 31355 Agoura Road, Westlake Village, CA 91361. (818) 889-5200. TELEX 18-3378. For microwave applications, contact Epsco, 411 Providence Hwy, Westwood, MA 02090. (617) 329-1500. TELEX 95-1110.

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HIGH POWER EPSCO

CIRCLE 42 ON READER SERVICE CARD

Around the Circuit

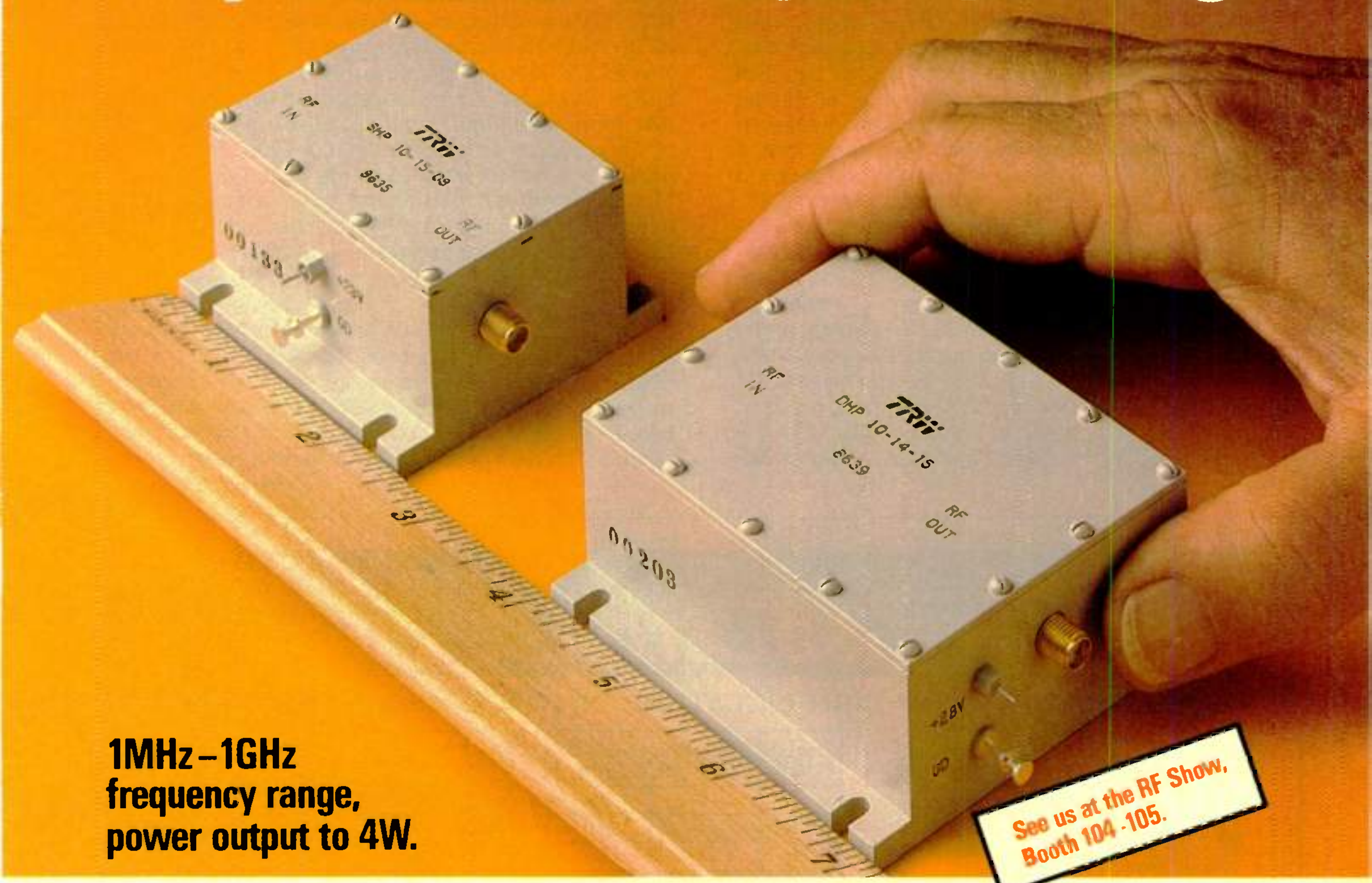
INDUSTRY NEWS

Coleman Microwave has added 3,300 square feet to its Edinburg, VA facility. The new space, to be used for aluminum dip brazing/metal finishing, meets MIL-B-7883 requirements and conforms to the MIL-I-45208 quality control system already in place at Coleman.... **Engelhard/Precision Microwave Circuits has broken ground for a 36,000-square-foot manufacturing and administrative building** in Milford, MA. The facility is expected to be completed by July 1987.... **Gould Inc., Microwave Products Division, has expanded and upgraded its GaAs device fabrication facility** in Santa Clara, CA. Gould has added a molecular beam epitaxy system and other equipment aimed at improving product reliability and quality. The company said that with the addition of the MBE system, it becomes the first commercial producer of high electron mobility transistors (HEMTs). In other news, Gould has changed the structure of its Information Systems Business. The Computer Systems Division will be headed by Patrick L. Rickard, president and GM; and the Imaging and Graphics Division will be headed by VP and GM Richard C. Baker, former Computer Systems Division VP of development. A new division, the Federal Systems Division, is comprised of the Systems Development Operation, which was transferred from the company's Defense Systems Business, and a former computer systems unit called the Federal Computer Systems Organization.... **Sanders Associates is being restructured following its recent acquisition by Lockheed Corp.** Included is a consolidation of the corporate office, Federal Systems Group and Component Products Group. Albert B. Wight will continue as president of Sanders Associates, which will retain its Nashua, NH headquarters, and former Sanders Chairman Jack L. Bowers will assist Roy A. Anderson, acting president of the newly formed Lockheed Electronics Group. The consolidation will result in the reduction of about 165 staff positions in the former Sanders corporate and Federal Systems organizations, according to Anderson. Those whose jobs are eliminated and who cannot be placed in other Sanders or Lockheed positions will be given help in finding new jobs, according to Sanders. The restructuring also includes the reassignment of CalComp, a former Sanders computer graphics company, and the Display Products Division to Lockheed's Information Systems Group.... **Sealectro Corp.**'s Watertown Operations will be transferred to the company's New Britain, CT facility in the first quarter of 1987. The move is designed to increase operating efficiency, expediting movement of components to the finished product stage.... **Landis Manufacturing Systems** has created a Military Electronics Division dedicated to close-tolerance machining of parts for manufacturers of electronic systems and components. LMS has over 60,000 square feet of manufacturing space and is a prime contractor to the DoD, meeting MIL-Spec 1-45208-A requirements.... **MCL/Inc.** has moved to larger facilities in Bolingbrook, IL and plans to employ an additional 50% personnel after the move.... **Spire Corp.** has opened a facility for ion beam enhanced deposition surface modification of materials. The IBED coatings are designed for applications including laser mirrors, optical coatings and magnetic storage media.... **Centel Corp.** and **M/A-Com Inc.** have agreed to Centel's purchase of M/A-Com Information Systems Inc., a systems integrator and value-added reseller of information systems. Closing on the transaction is expected by December 27, 1986.... **The International Telecommunication Union** has named **Horizon House-Microwave Inc.** publishing contractor for its monthly *Telecommunication Journal*. Horizon House is responsible for the promotion and printing of the *Journal*, as well as the canvassing for and contracting of its advertising. The ITU will retain control over editorial content. The *Journal* is published in English, French and Spanish and is received by members of national telecommunication administrations and industry officials. The ITU is the specialized United Nations agency for the planning, coordination, regulation and standardization of telecommunications worldwide. Horizon House also publishes the *Microwave Journal* and other technical magazines and books.... **Axiam Inc.** has purchased **Eaton Corp.'s Test System Division,** a supplier of discrete semiconductor test equipment. Axiam, which owns Mastech Inc., said that with this acquisition it becomes the world's largest supplier of this equipment line. The newly-acquired division will be called by its original name of LORLIN and will continue operations under its present GM, Richard Barnard.... **GAIN Electronics Corp.** plans to offer standard and custom GaAs and AlGaAs epitaxial wafers grown by molecular beam epitaxy for digital and microwave applications. Delivery to several customers is expected to start before the end of 1986.... **VERTEQ Process Systems Division** has received a US patent for the non-contact, labyrinth bowl seal developed for the company's Superclean 1600 rinser/dryer. The seal eliminates particulate generation, a major cause of wafer contamination, VERTEQ reports.... **Watkins-Johnson Ltd.** has received notification that it meets the requirements of Allied Quality Assurance Procedure 1, the most stringent quality control level applied to NATO contracts.... **KDI Electronics Inc.** has appointed a new board of directors member to meet DoD requirements, since parent company KDI Corp. has had a management change that includes a foreign interest. General Alton Slay (USAF, retired) has been appointed to the board and will act as proxy between KDI Corp. and KDI Electronics, which retains its full former staff.

[Continued on page 74]

New from TRW...

A broadband, medium-power amplifier in a compact package.



TRW RF DEVICES has developed two new lines of packaged linear amplifiers that feature . . .

- Heavy-duty machined housing
- EMI-RFI shielding
- SMA connectors
- Low noise
- High gain (14-36dB)

These new modules are ideal for applications in instrumentation, communications equipment and military systems. And they are lower priced than comparable products.

The SHP series package measures only 7.2 cubic inches; the DHP series package measures 9.8 cubic inches.

All products utilize a push-pull configuration and operate with bias voltages of 15V, 24V or 28V. They also are protected against reverse bias. They meet MIL-SPEC 883, Methods 1004 and 1009, for humidity and salt spray.

Prices (in quantities of 1-9) range from $240 to $275 for the nine models in the SHP series. The five models in the DHP series—which feature higher gains and power outputs at comparable frequencies—are priced from $380 to $455.

Look to TRW RF DEVICES for the latest developments in RF POWER TECHNOLOGY.

For data sheets, applications assistance or the location of your nearest stocking distributor, contact Don Murray, Sales Engineer:

RF Devices Division
TRW Electronic Components Group
14520 Aviation Blvd.
Lawndale, CA 90260, 213.536.0888

TRW

RF Devices Division
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CIRCLE 52 ON READER SERVICE CARD

FINANCIAL NEWS

Raytheon Co. reports sales of $1.7B for the third quarter, compared to $1.6B for the same period last year. Earnings were $101.5M ($1.31/share), compared to $95.3M ($1.17/share) for the corresponding quarter a year ago.... **Gould Inc.** says net revenues for the second quarter, ended June 30, were $321.3M, compared to $356M for the same quarter the previous year. Net earnings were $4.3M (10¢/share), compared to a net loss of $144.7M ($3.30/share) for the second quarter last year.... **AEL Industries Inc.** quotes revenues of $29.8M for the second quarter, ended August 29, compared to $24.5M for the same period a year ago. Net loss was $248K (5¢/share), compared to a net income of $237K (5¢/share) for the comparable quarter in 1985.... **California Microwave Inc.** says sales for the fiscal year ended June 30 were $104.4M, compared to $105.4M for fiscal 1985. Net income was $4.4M (55¢/share), compared to $4.6M (55¢/share) for the corresponding period last year.... **Radiation Systems Inc.** reports sales of $47.5M for the year ended June 30, up 42% from $33.5M for the previous year. Net earnings were $4.2M (87¢/share), up from $3.2M (70¢/share) for fiscal 1985.... **Andersen Group** quotes sales of $20.9M for the six months ended August 31, compared to $22M for the same period a year ago. The company incurred a loss of $391K, compared to a profit of $437K for the comparable six-month period last year.... **Herley Microwave Systems Inc.** says revenues for the year ended July 31 were $16.3M, up 69% from $9.6M last year. Net earnings were $2.1M (72¢/share), up from $1.2M (42¢/share) for the previous year. For the fourth quarter, revenues were $5.1M, up from $3.5M for the comparable period last year.... **North Hills Electronics Inc.** reports net sales of $1M for the second quarter, ended July 31, up 30% from $787K for the corresponding quarter in 1985. Net loss was $252K (10¢/share), compared to a loss of $198K (7¢/share) for the same period last year.... **Kevlin Microwave Corp.** says net sales were $828K for the first quarter, ended August 31, compared to $2.5M for the comparable quarter in 1985. Net loss was $137K (5¢/share), compared to a net income of $229K (9¢/share) for the same period last year.... **Pacific Monolithics Inc.** has completed a second round of financing, raising $3.5M to continue its planned growth and development over the next several years. The financing included repeat investments by Shaw Venture Partners, Vanguard Associates and Investment Advisers Inc., as well as several individual investors.... **Aydin Corp.** says it has resumed the stock repurchase program authorized by its board of directors February 4. When the program was discontinued late in February the company had purchased 184,400 shares of the 500,000 shares authorized for repurchase on the open market.... **Watkins-Johnson Co.**'s board of directors has declared a dividend distribution of one common share purchase right on each outstanding share of the

[Continued on page 76]

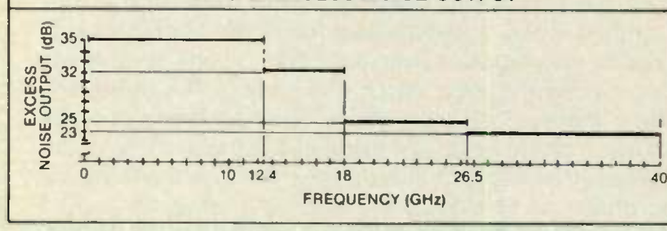
Wideband, mm-wave, high-level...
MSC's got a nose for noise.

MSC's Solid State Noise Source product line results from our advanced microwave power semiconductor technology. Hybrid IC construction yields mechanical integrity capable of withstanding the most severe environmental conditions with out sacrificing reliability and long-term stability.

Solid State Noise Sources are available with excess noise ratios (ENR) of 15.5 dB as direct replacements for gas discharge tube noise sources. Sources with higher level ENR allow noise to be injected into receiver front-ends via a directional coupler.

System Noise Sources are specifically designed for "BITE" applications for monitoring of parameter changes and system performance. ENR outputs of up to 35dB allow detection of receiver deterioration and developing faults.

TYPICAL EXCESS NOISE OUTPUT



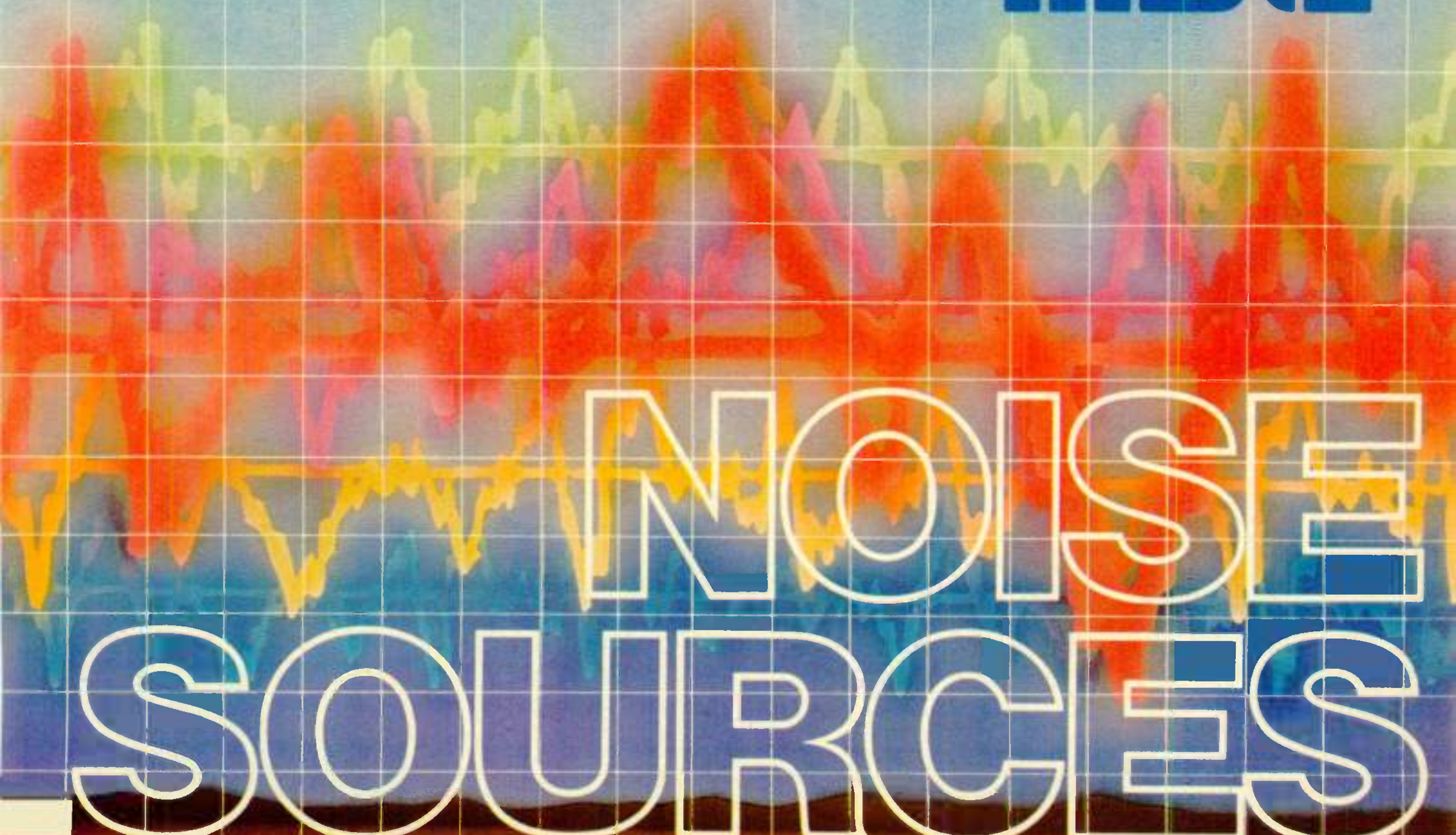
WIDEBAND COAXIAL SOURCES 10 TO 18000 (MHz)				
MODEL NUMBER	FREQUENCY RANGE	ENR	FLATNESS	POWER REQ. MAX.
MC 1000	10 - 1500	35.0dB	± 0.50dB	+ 28V, 10mA
MC 1100	10 - 1500	15.5dB	± 0.50dB	+ 28V, 10mA
MC 1040/1040P	10 - 4000	25.5dB/15.5dB	± 0.50dB	+ 28V, 15mA
MC 5112/5112P	1000 - 12400	25.5dB/15.5dB	± 0.50dB	+ 28V, 15mA
MC 5118/5118P	1000 - 18000	25.5dB/15.5dB	± 0.50dB	+ 28V, 15mA
MC 50018/50018P	10 - 18000	25.5dB/15.5dB	± 0.75dB	+ 28V, 15mA

WAVE GUIDE BAND SOURCES (GAS TUBE REPLACEMENTS)				
MC 51218W	12400 - 18000	15.0dB	± 0.50dB	+ 28V, 15mA
MC 51826W	18000 - 26500	25.0dB	± 2.00dB	+ 28V, 20mA
MC 52640W	26500 - 40000	23.0dB	± 3.00dB	+ 28V, 20mA

HERMETIC NOISE SOURCES				
MC 7090H	8325 - 9675	35.0dB	± 0.50dB	+ 28V, 15mA
MC 7146H	13500 - 15700	30.0dB	± 0.50dB	+ 28V, 15mA

Microwave Semiconductor Corp.
100 School House Rd., Somerset, N.J. 08873
A Siemens Company
(201) 469-3311 TWX(710)480-4730 TELEX833473

MSC



CIRCLE 93 ON READER SERVICE CARD

company's common stock to all shareowners of record as of October 20. The rights, which expire in 10 years, will trade with common shares until a person or group acquires at least 20% of the company's shares or announces an offer to acquire 30% or more of the outstanding stock, at which time the rights become exercisable, will separate from the common shares and will trade independently. When exercisable, the rights entitle a shareowner to purchase one new share of stock for each share held, at a price of $160 per share. If the company is involved in a merger, the rights will permit purchase of shares in the surviving entity at a 50% discount from the then-existing market price. The rights also provide protection against self-dealing transactions by a controlling shareowner.

CONTRACTS

Harris Corp. has received a $65M three-year contract from the Naval Air Systems Command for computerized automatic test systems. Harris will produce about 100 series 2000 automatic test systems for use on Naval aircraft including the F/A-18, AV-8B, A-4, A6A, LAMPS Mark III helicopter, EA6B, E-2C, F-14, A/D, S3 A/B and V-22 Osprey.... **Aydin Corp.** has been awarded a $57.4M US Army contract for the GRC-222 microwave radio program. The $57.4M includes $24.2M in probable production options. Aydin will supply 733 radios initially and another 500 for the probable production options. Aydin also has received a $7.8M contract to supply transmitter sets to the US Navy for missiles and a $4.7M award for airborne data acquisition systems to be delivered to a large US aircraft manufacturer.... **Loral Corp.** has been awarded a $41M USAF contract for additional units of the ALR-56C radar warning receiver for the F-15 fighter aircraft.... **Racal-Dana Instruments Inc.** has received a five-year USAF contract valued between $10M and $16M to supply 8,000 to 12,500 universal counters. The first delivery, worth more than $2M, is scheduled for early 1987.... **Honeywell** has been awarded an initial $12.9M USAF contract for design and delivery of the prototype GBU-15 part task trainer (PTT), which will train Tactical Air Command's weapon system officers in GBU-15 and AGM-130 weapon system use.... **American Satellite Co.** has awarded a long-term contract to Cable & Wireless Communications Inc. for fiber-optic transmission facilities. The minimum value of the contract is estimated at $12M over a 10-year period.... **Raytheon Service Co.** has been awarded an $11.6M USAF contract to provide base operations support at Richards Gebaur Air Force Base near Kansas City, MO. The contract, with options spanning three years, includes logistics, fire protection, transportation, property maintenance and data processing.... **American Electronics Laboratories** has concluded a successful airworthiness flight test after three years of R&D for an airborne seeker evaluation test system (ASETS), part of a $9M contract with the Air Force Armament Division's 3246 test wing.... **Electromagnetic Sciences Inc.** has received a $7.9M order from Norden Systems Inc. for the final develop-

[Continued on page 78]

SAW Questions and Andersens.

Dispersive devices:

QUESTION: I'm curious about dispersive devices. Where would I use them?

ANDERSEN: Our matched filters are widely used in digital communications and radar. With pulse expansion/compression you can increase dynamic range without increasing transmitter peak power or sacrificing resolution. They're also used in compressive receivers for real time spectrum analysis with 100% probability of intercept.

QUESTION: What range of bandwidth, dispersion and center frequencies are available?

ANDERSEN: SAW dispersive devices are used for wide bandwidth applications (up to 500 MHz) with dispersions up to 100 μs. IMCON dispersive devices have narrower bandwidths but provide dispersions up to 600 μs (IMCON's have been cascaded to produce dispersions of 10 ms). Center frequencies of Andersen dispersive devices range from 1 MHz to 750 MHz.

QUESTION: If I want to use dispersive devices in my system, what do I do?

ANDERSEN: We can help you specify the devices you need. Or we can supply the entire subsystem (compression/expansion module, compressive receiver, etc.). We've been supplying such systems for over 20 years. And with our in-house hybrid facility we can provide a compact, high performance unit tailored specifically for your system needs.

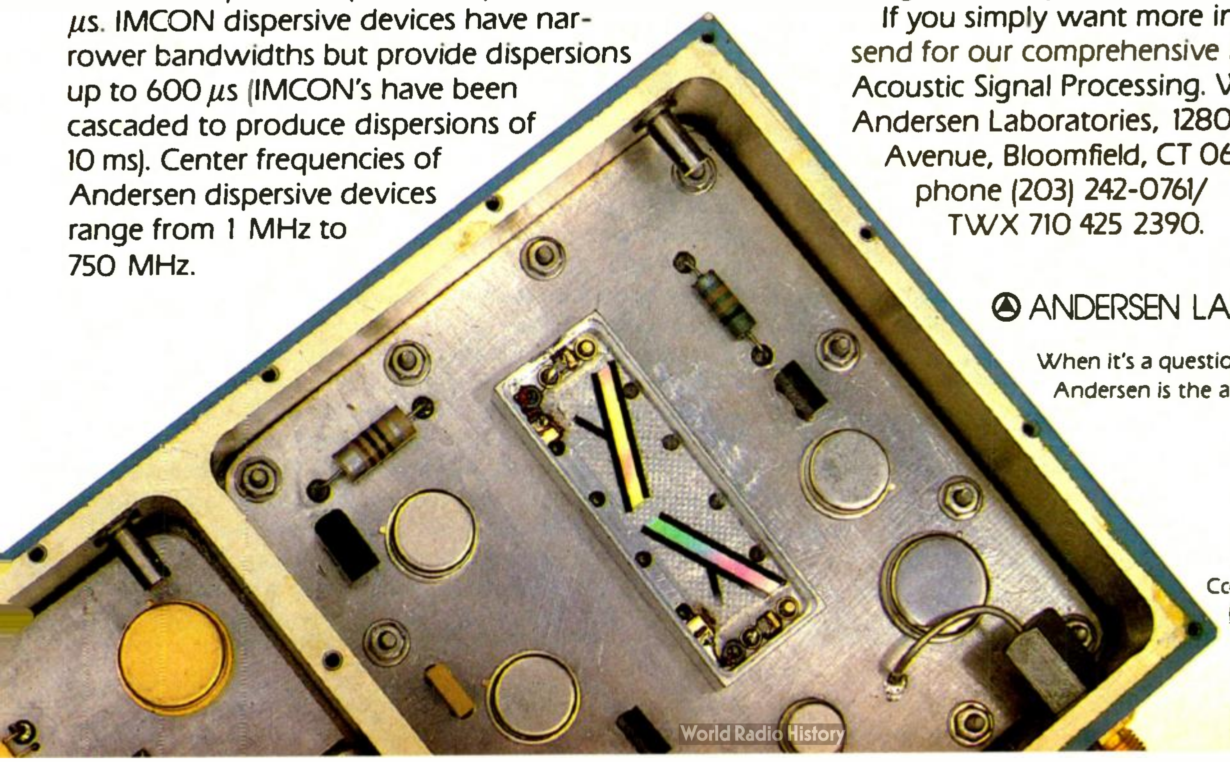
QUESTION: How do I get started?

ANDERSEN: Just give us a call. We'll do everything we can to meet your needs. With our recent major staff increase, we offer one of the largest groups of SAW designers of any U.S. company.

If you simply want more information, send for our comprehensive handbook on Acoustic Signal Processing. Write to Andersen Laboratories, 1280 Blue Hills Avenue, Bloomfield, CT 06002. Or phone (203) 242-0761/ TWX 710 425 2390.

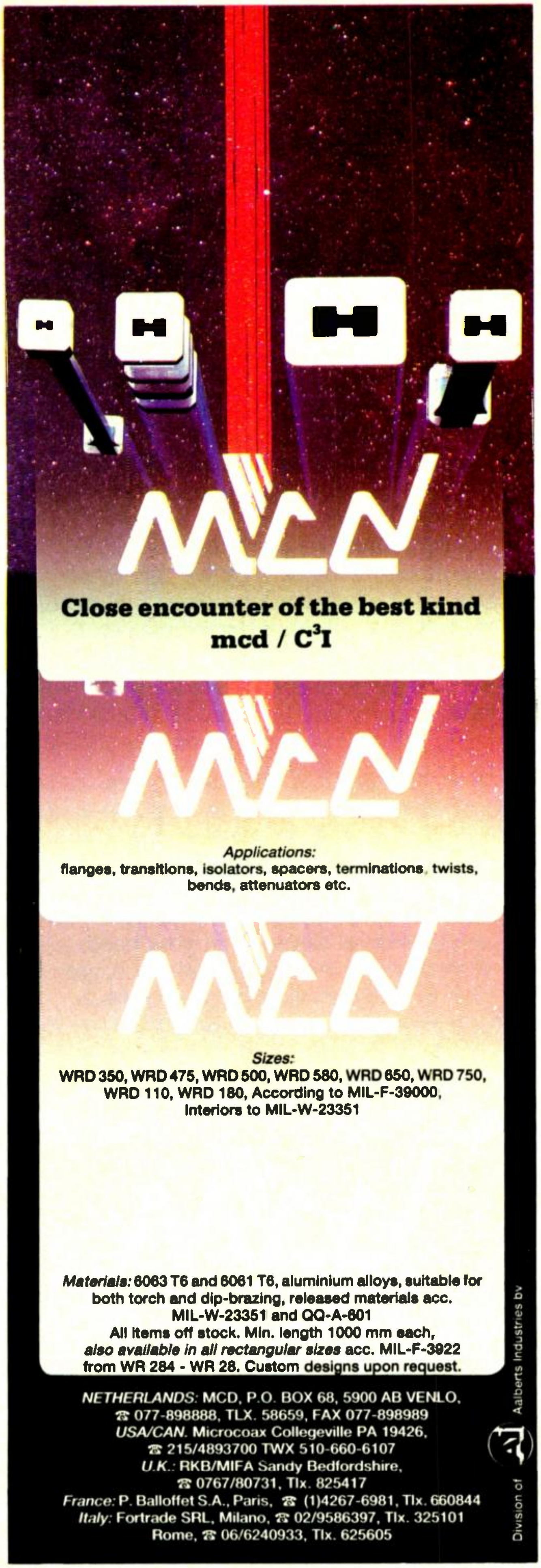
ANDERSEN LABORATORIES

When it's a question of SAW, Andersen is the answer.



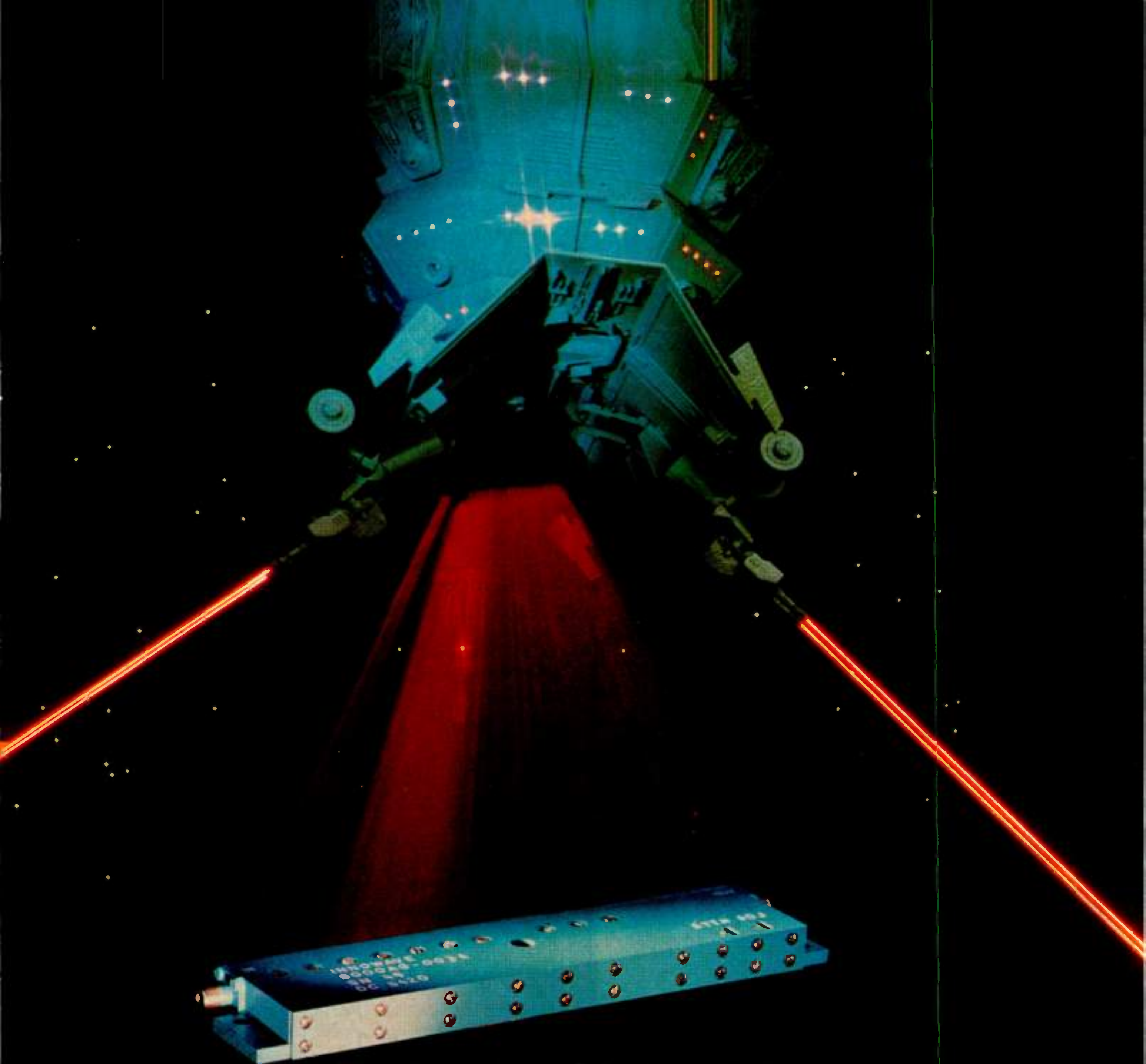
Compressive Receiver (actual size)

CIRCLE 56



ment and production of dual ferrite phase shifters. The phase shifters will equip antennas for a Norden airborne radar system being developed for the US Army and USAF Joint Surveillance Target Attack Radar (Joint-Stars) Program.... **Watkins-Johnson Co.** has received a $7.2M order from Hughes Aircraft Co. to deliver integrated hybrid microwave circuits, using Si and GaAs devices in monolithic form, for the AIM-120 Amraam air-to-air missile.... **Anaren Microwave Inc.** has received a $7.1M order from the US Navy for digital frequency discriminators. They will be used as spares for electronic defense systems on Navy aircraft.... **M/A-Com Microwave Power Devices Inc.** has been awarded a $7M US Navy contract for RF power calibration equipment. The contract calls for the development and production of a microprocessor-controlled power meter calibration system to be used on Navy equipment at various bases.... **Herley Microwave Systems Inc.** has received 12 contract awards totaling $5.3M. Included are a Rockwell Corp. contract for oscillators for the MILSTAR Program; a US Navy contract for amplifiers for the SPS-49 program; and a USAF contract for oscillators, triplexers, detectors and other products for the ALQ-119 and ALQ-131 Electronic Countermeasure Programs.... **Rockwell International Corp.**'s Collins Defense Communications has been awarded a $4.6M contract to produce its AN/ARC-182 UHF/VHF radios for the US Navy.... **Electrospace Systems Inc.** has received a $4.2M USAF contract for the design, development and fabrication of 34 satellite interface modules for AWACS aircraft. Installation of the production prototype is scheduled for early 1987.... Under an existing contract between Digital Video Systems (a Scientific-Atlanta Inc. division) and British Telecom, **Scientific-Atlanta** has received a $3M order to supply B-MAC satellite communications encryption products for Satellite Racing Development Ltd. In other news, Federal Express plans to use S-A's proprietary B-MAC transmission system to transmit video broadcasts from its central facility in Memphis, TN to its network of remote locations. The system is being provided by Private Satellite Network Inc., an integrator of satellite communication services for business.... **Comtech Inc. subsidiary Comtech Microwave Corp.** has been awarded a $1.6M US Army contract for the analog simple data interface, which provides a passive interface for voice or data communication over switched analog telephone wires and trunks.... **STAR Microwave** has received two USAF contracts to supply helix traveling-wave tubes for use in the AN/ALQ-94 ECM system and the AN/ALQ-119 ECM pods. STAR is to deliver 686 tubes over the next three years under this contract. In addition, STAR has been awarded USAF contracts totaling $1.4M, adding 201 traveling-wave tubes to STAR's planned production for the ALQ-94/137 electronic countermeasures system. STAR also has received a $368K contract from US Army LABCOM for R&D of low cost, wideband TWTs.... **Peninsula Engineering Group Inc.** has received orders totaling over $175K for the RF-2000 2 GHz solar powered microwave repeater. Included are orders from GTE of the Northwest, the US Forest Service and three major microwave manufacturers.... ANT Telecommunications

[Continued on page 80]



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CIRCLE 58 ON READER SERVICE CARD

Inc. has accepted a purchase order from Cornell University for a high power y-waveguide circulator. The 0.6 MW, 50 MHz circulator will be used by the university's Laboratory of Nuclear Studies for research.

PERSONNEL

Clifford C. Christ has been appointed president and GM of Gould Inc.'s NavCom Systems Division. He was most recently program manager for flight simulation and digital control systems at General Electric's Simulation and Control Systems Department. In addition, **David Simpson**, Gould vice chairman, will retire from the post, effective December 29. The position will not be filled.... **Allen F. Standley** has been appointed president of Allied-Signal's Amphenol Products unit. He succeeds **Phillip W. Arneson**, who has been made executive assistant in the Electronics and Instrumentation Sector. Standley was most recently VP and GM of Amphenol's Bendix Connector Operations.... **Terence W. Ede** has been named president of Sealectro's US Division. In this capacity, Ede also will be a member of the corporation's board of directors. He had been executive VP and GM.... **Raymond A. Foos** has been named president and COO of Brush Wellman Inc. He had been executive VP and president of the company's Beryllium and Specialty Materials Group.... **Robert J. Snyder** has been appointed president of Micro Chassis, a Keene Corp. division. He was previously VP, operations at Terminal Data Corp.... **Walter H. Palmer** has been elected VP, public and financial relations at Raytheon Co. He had been director of equal opportunity programs, public affairs and community relations. Palmer replaces **Richard R. Mau**, who left in June to be VP, communications for Rockwell International Corp. Also at Raytheon, **Gretchen S. Stephens** has been named production manager for the Electromagnetic Systems Division. She had been director of manufacturing management development at the corporate headquarters.... **Robert E. Sullivan** has joined Harris Corp. as senior VP, administration. He was previously senior VP, finance and administration at Harris Graphics Corp. In addition, **Leo Fong** has been promoted to manager of new business development at Harris Semiconductor Component Operations. He had been program manager.... **Eugene J. Ferrari** and **Roy W. Jacobus** have been elected VPs of the MITRE Corp.'s Bedford C³I Division. Ferrari had been technical director of strategic communications and Jacobus had been technical director of aerospace surveillance and defense.... **Stuart Wilson** has been appointed VP of manufacturing at RACOM Corp. He was previously manufacturing manager of microwave components at Litton's Applied Technology Division and, prior to that, process engineer at Watkins-Johnson Co.... **Christopher G. Conlin** has been promoted to GM of Murata Erie North America Inc.'s Georgia Division and **Tony R. Coalson** has been named marketing manager. Conlin had been national marketing manager and Coalson had been group product manager for ceramic capacitors.... **David J. Kuzmick** has been appointed product marketing specialist for the resistive elements product line at M/A-Com Omni Spectra Inc. He had been TVRO sales

[Continued on page 82]

Get away from it all.

If we can remote the antenna 5 km—with only 15 dB loss—think what our fiber optic systems mean for you!

Remoting antennas this distance always used to be just about impossible. Ortel makes it practical and affordable. Because we design and manufacture complete laser fiber optic links for low loss microwave transmission. For us, 5 km is nothing; at 20 km, the loss would be a trifling 15 dB more.

And antenna remoting barely starts the list. What could specs like this do for *your* application?

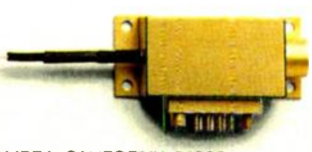
insertion loss	10 dB
link loss	1 dB/km
input/output VSWR	2.0
1 dB compression point	+15 dBm
3rd order intercept	+28 dBm
bandwidth	up to 6 GHZ
input impedance	50 ohms
equivalent input noise	-120 dBm/Hz

There's more. Transmission is three-ways secure: impervious to secret intercept, EMI and EMP. And installation's a snap. Fiber is exceptionally lightweight, and thin enough to snake through small apertures.

This is not "tomorrow's technology." It's here and now, today. Proven. In operation. But it is unarguably what you will have to have tomorrow to keep your designs competitive—or find yourself saddled with yesterday's solutions.

Call or write Ortel. Together we'll explore ways that optoelectronics can help solve the thorniest problems of your microwave transmission and signal processing applications—whatever those are. And Ortel, the example-setter in this field, will stick with you, step-by-step.

Call now. Tomorrow's almost come.



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CIRCLE 61 ON READER SERVICE CARD

[From page 80] **AROUND THE CIRCUIT**

engineer.... Precision Monolithics has promoted **Larry Brown** to director of product assurance and appointed **Doug Divine** to manager of reliability and quality engineering. Brown had been reliability and quality engineering manager and Divine had been director of product operations at Holt Integrated Circuits.... **Robert M. Hautzik** has been appointed director of operations at Ferretec Inc. Hautzik had been manager of the microwave component product line for Varian Associates' Microwave Equipment Division.... STAR Microwave has appointed **Malcolm J. Carruthers** project engineer. He was previously senior engineer at English Electric Valve, Tube Division.... **Harold A. Wheeler**, chairman Emeritus of and a consultant to Hazeltine Corp., has

Roy W. Jacobus

David J. Kuzmick

Gary Martin

Michael A. DeGiso

Richard Girard

been formally inducted into the National Academy of Engineering for "an outstanding lifetime of contributions to the analysis and fundamental understanding of radio, FM, television, microwave and antenna technologies".... **Barry Dunbridge** has been named assistant GM for technology at TRW's Electronics and Technology Division. He had been manager of TRW's 500-employee Microelectronics Center.... **Gary Martin** has been named national sales manager for Varian Associates Inc.'s Semiconductor Equipment Group. He was most recently marketing manager for the Thin-Film Technology Division. In addition, **William J. Nanney** has been appointed to the new post of plant manager at Varian's Eimac Division. Nanney had been operations manager for the Accu-Glass Division of Becton, Dickinson and Co.... **Michael A. DeGiso** has been promoted to sales manager of Alpha Industries' Advanced Technology Division. He was most recently field sales engineer for the company's New England sales office.... Microwave Networks Inc. has expanded its sales and marketing department with the addition of three regional sales managers: **Robert E. Davis** has been named sales manager, Midwest region, for microwave networks; **Becky G. Coyne** has been named sales manager-Southern region, for microwave networks; and **Robert J. Farrell** has been named sales manager, Western region, for microwave networks. Davis had been national sales manager for Light Communications Corp., Coyne had been national director of sales and marketing for Compucon, and Farrell had been Western sales manager for Gemlink, prior to its sale to Motorola.... **Victor Riess** has been promoted to marketing manager at AEG Corp. He had been a microwave device marketing engineer.... MAST Microwave has appointed **Dennis Flanders** manager of sales and marketing. He was most recently Western regional sales director for Applied Engineering Products.... **Richard Girard** has been named marketing manager for Augat Inc.'s Interconnection Components Division. He had held sales and marketing management positions for Texas Instruments, Connector Division.... **R.S. Pengally** has joined Tachonics Corp. as director of IC design. He had been at Plessey Research Caswell Ltd. in England, where he established the company's MMIC activity.

NEW MARKET ENTRY

Channel Microwave Corp., specializing in the design, manufacture and sale of passive microwave components, has been formed. The company's initial work will be in coaxial and waveguide ferrite devices. Edward L. Vadnais, who was founder of Junction Devices and served in the posts of president and CEO at that company, is president of Channel Microwave. Channel's principals have 70 years of combined technical experience. They include: Richard Roach, VP, operations; Charles Gasior, VP, engineering; and Nick Pena, VP, marketing. The address is 4630 Calle Quetzal, Camarillo, CA 93010, tel. (805) 482-7280.

REP APPOINTMENTS

TMR Associates of Groton, MA has been appointed New England area representative for **STAR Microwave**.... **Log Tech** has appointed **Tekelec Airtronic** sales representative in France; **Midoriya Electric Co.** sales representative in Japan and **March Microwave** sales representative in the UK for Log Tech's family of log amplifiers and threshold detectors.... **Ferranti Semiconductors** has appointed **Distributed Microtechnology Inc.** of Orange, CA as a distributor.... **Sage Laboratories Inc.** has appointed the following companies as sales representatives: **Hughes Associates Inc.** of Huntsville, AL covering Tennessee, North Carolina, South Carolina, Mississippi, Alabama, Georgia and Florida; **Northern Technical Sales** of Liverpool, NY covering Syracuse, Utica and Rochester, NY; and **Kaizer International Co.** of Huntington Beach, CA covering Taiwan.... **Channel Microwave Corp.** has named the following sales representatives: **MCA Repco Inc.** for Southern California; **AD COM** for Northern California; **Allis Associates-Northwest** for Oregon and Washington; **RF Resources** for Texas, Arkansas, Oklahoma and Louisiana; **EIR Co.** covering Iowa, Missouri and Illinois; and **Schibley Associates** covering Maryland, Virginia, West Virginia and Washington, DC. ■

Avantek Component Solutions... 0.5 to 18 GHz

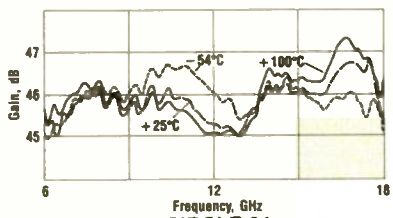
New Amplifiers Added to .5 to 18 GHz Family

Avantek has the range of amplifiers to provide the right solution for your system design needs. Our new super temperature compensation and gain control amplifiers can significantly improve your systems' performance. And, since all Avantek amplifiers are extremely compact, system size can be reduced while you're increasing performance.

Significant Improvement in System Dynamic Range

Avantek's new active temperature compensated amplifiers offer a 40% reduction in gain variation versus temperature over the best amplifiers previously available. The new AWT-18648 6 to 8 GHz active temperature compensation amplifier has a 2.5 dB peak-to-peak variation at 45 dB gain over the −54° to +100°C temperature range. Avantek's new active temperature compensation is available for all small-signal (to +20 dBm) AWT and AMT series 6 to 18, 8 to 18 and 12 to 18 GHz amplifiers.

Active Temperature Compensation



CIRCLE 64

Variable Gain Amplifiers Cover 2 to 18 GHz

Avantek's new AGT series variable gain amplifiers can improve your system's dynamic range. These standard amplifiers are available with 10 or 20 dB gain control range over the 2 to 8 and 2 to 18 GHz frequency bands, and offer extremely fast response time—full variation in less than 1 μsecond. These amplifiers feature a high impedance voltage control (0 to 10V, 3 to 5 mA typical), which simplifies gain control power requirements and permits rapid control changes. This gain variation may be used for temperature compensation.

Selected AGT-series Variable Gain Amplifiers

AGT-model	Frequency Range (GHz)	Gain at Zero Gain Control (dB.min.)	Gain Control (dB.min.)	Noise Figure (dB.max.)	P₁dB (dB.min.)
-8233	2–8	25	10	5.5	+15
-8235	2–8	45	20	5.5	+15
-18634	2–18	25	10	6.5	+15
-18636	2–18	40	20	6.5	+15

CIRCLE 65

Avantek Amplifiers Pack Big Gains in Small Packages

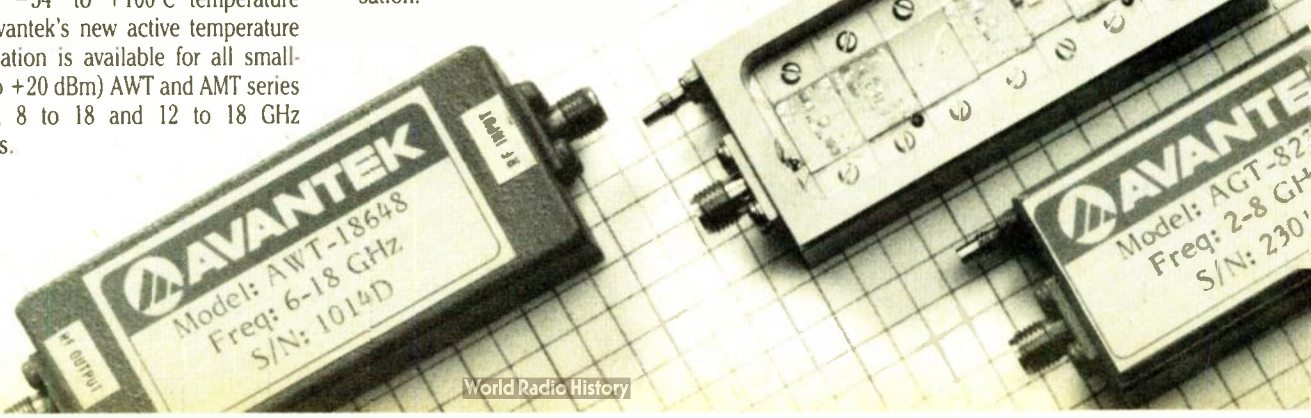
Avantek's Avanpak™ and Avanpak Plus™ low-noise amplifiers allow system designers to put gain, in small or large increments, where it's needed. They are economically priced and provide uncompromised performance in less than 20% of the volume and 25% of the weight of more conventional amplifiers. These amplifiers feature low-profile, miniature, SMA-connectored, hermetic packages that are stripline compatible. They're optimized for applications where low-noise, high-gain and dynamic range are critical.

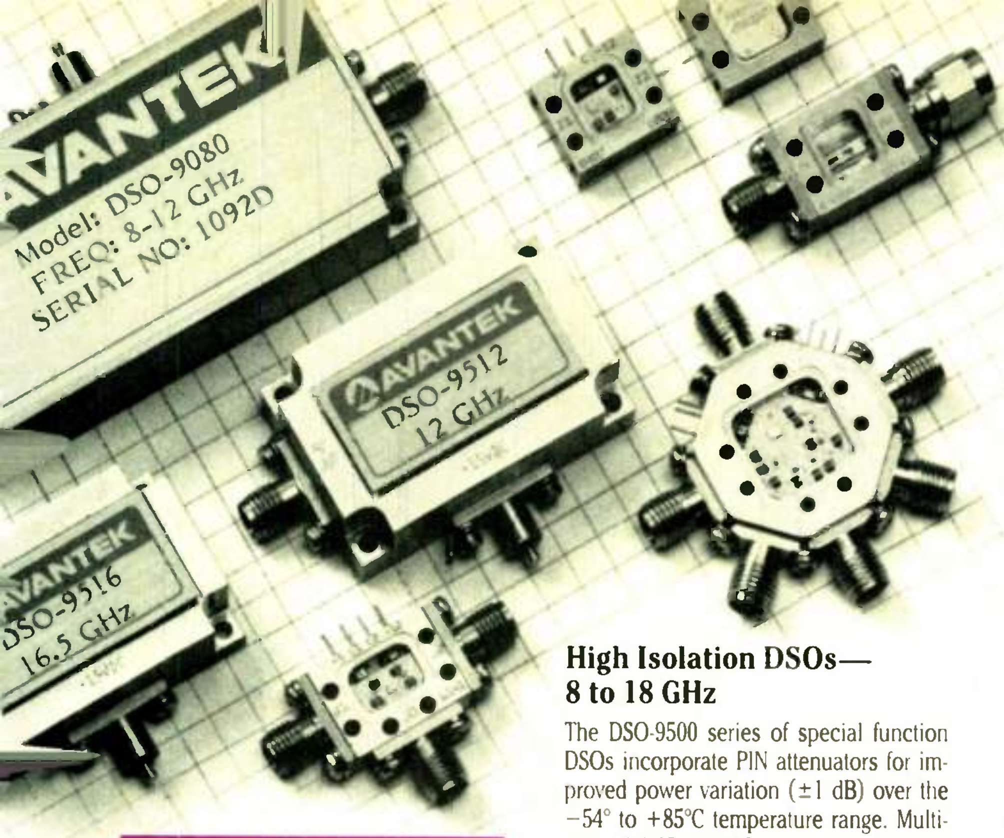
Selected Avanpak and Avanpak Plus Amplifiers

Model	Frequency Range (GHz)	Gain (dB.min.)	Noise Figure (dB.max.)	P_1dB (dBm.min.)
AFT-2031*	0.5–2	10.0	3.5	+11
AWT-2034-M**	0.5–2	47.0	3.5	+13
AFT-8231*	2–8	6.5	4.5	+12
AWT-8035-M**	2–8	41.0	4.0	+15
AFT-18631*	6–18	5.5	7.3	+10
AWT-18637-M**	6–18	44.0	6.5	+15

*Avanpak case **Avanpak Plus case
.22" H x .66"–.87" W x 1.5"–2.8" L

CIRCLE 66



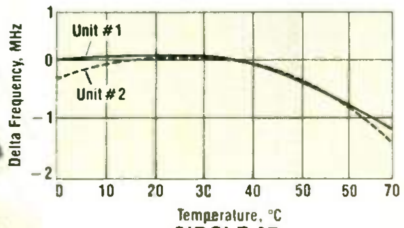


Oscillators Offer Temperature Compensation Solutions

Frequency Tracking DSOs—4 to 18 GHz

The new DSO-9000 series of special function dielectrically stabilized oscillators (DSOs) are available as matched pairs for applications where frequency tracking between units must be held to very tight tolerances. Typical frequency tracking specifications are better than ±1 MHz. Common applications include receive/transmit pairs, where one unit is a signal source for built-in test and the other unit is the LO in the receiver. Offset frequencies between paired oscillators can be specified at 70 or 160 MHz.

DSO-9000 Series Frequency Change Vs Temperature



CIRCLE 67

High Isolation DSOs—8 to 18 GHz

The DSO-9500 series of special function DSOs incorporate PIN attenuators for improved power variation (±1 dB) over the −54° to +85°C temperature range. Multi-stage MMIC amplifiers are used to increase power output while providing better than 60 dB isolation from the load. This series is designed for applications where load VSWRs are high and frequency pulling is critical.

Frequency Versus Load VSWR

Load VSWR (All Phases)	Delta Frequency (kHz)
1.1:1	0.25
1.67:1	2
3.0:1	5

CIRCLE 68

Control Component Solutions — Distributor Direct

2 to 18 GHz Miniature Thin-Film Switches

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AH Series 2 to 18 GHz Switches

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AHF 1802-0	SP5T	55	8

*10–90% rise or fall times. Driver delay typically 20 nsec

CIRCLE 69

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CIRCLE 70

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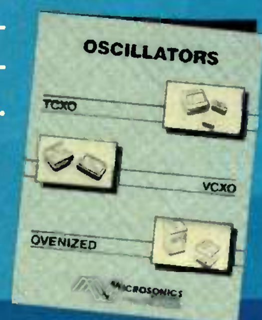

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An Improved Hybrid-Ring Directional Coupler for Higher Power Split Ratios*

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and
Gerald F. Mikucki
RCA Missile and Surface Radar Division
Moorestown, NJ

A directional coupler in the form of a hybrid ring particularly suited for printed circuits is described. The maximum power-split ratio between the two output ports of a printed-circuit conventional hybrid-ring coupler is limited by the highest impedance line that can be realized.[1] The hybrid-ring directional coupler described in this paper allows a larger power-split ratio for the same impedance lines and thereby increases the range of the power-split ratio that can be realized for printed circuits. A theoretical analysis was conducted using the scattering matrix, and experimental verification of the theoretical results was achieved in a stripline configuration at Ku band.

Introduction

The hybrid-ring directional coupler is an appealing choice for the basic power division element in beam-forming networks for printed circuit array antennas for two primary reasons: the output arms are isolated from each other, and the input impedance is matched when the other arms are terminated by matched impedances. Since the conventional T (or Y) junction power dividers do not possess these properties, directional couplers are preferable for antenna array feed systems where the isolation between the output arms of the power divider is essential to minimize mutual coupling between radiating elements.

The two-dimensional structure of stripline facilitates construction of the feeding network and antenna elements, such as dipoles on a single printed circuit board. At high frequencies, limited real estate makes hybrid-ring couplers preferable to branchline and parallel-line couplers; the hybrid-ring couplers have an inherent 90° phase difference between output ports. For an antenna array that is excited by an equi-phase, symmetrical, corporate feed network, the hybrid-ring directional coupler has a definite advantage over the parallel-line and branchline couplers because no phase-compensating element is necessary. The hybrid-ring coupler also has a broader bandwidth than the branchline coupler.[1]

The configuration of a conventional hybrid-ring directional coupler is shown in Figure 1. The characteristic admittances of the four arms are normalized to unity. The variable parameters Y_1 and Y_2 represent the characteristic admittances of two lines of the ring. They determine the degree of coupling of the output arms and the matching condition for the input arm. When the signal is fed into sum port 3, the output voltages in arms 1 and 2 are in phase, and their relative amplitudes are related by

$$\frac{b_1}{b_2} = \frac{Y_2}{Y_1} . \tag{1}$$

When the signal is fed into difference port 1, the output voltagess in arms 3 and 4 are 180° out of phase, and their relative amplitudes are related by

$$\frac{b_3}{b_4} = \frac{Y_2}{Y_1} . \tag{2}$$

In both cases, having the input arm perfectly matched requires that Y_1 and Y_2 satisfy the condition

$$Y_1^2 + Y_2^2 = 1 . \tag{3}$$

From Equations 1 and 2, it is evident that the output voltage ratio is directly proportional to the ratio of the characteristic admittances (or impedances) of the transmission lines forming the ring. The impedance values required for various power-split ratios are presented in Table 1. In microstrip and stripline circuits, the highest impedance line that can be realized limits the

*Invited paper.

maximum power-split ratio between the two output arms. The highest attainable impedance value for strip lines and microstrips is a function of the physical characteristics of the substrate (e.g. dielectric constant and thickness). The practical impedance limit for most substrates, using conventional etching techniques, is around 150 ohms. Referring to Table 1, this limits the maximum power-split ratio for a conventional hybrid-ring coupler to approximately 9 dB.

To obtain a higher power-split ratio between the output ports, the three-quarter wavelength line in the ring is split into three quarter-wavelength lines of characteristic admittances Y_1 and Y_2, as illustrated in Figure 2. This modified hybrid-ring directional coupler was analyzed using the scattering matrix for the four-port devices. Experimental results for a stripline coupler with a power-split ratio of 6.4 dB are presented in this paper. The modified hybrid ring is similar to the conventional hybrid ring in that: its input impedance is matched when the two adjacent arms are teminated by matched loads, and the voltages in the two output arms are either in phase or 180° out of phase, depending on the input arm chosen. The power-split ratio is adjusted by varying the impedances of the lines in the ring between the arms.

The modified hybrid-ring coupler differs from the conventional ring coupler in that the two output arms are not perfectly isolated at the center frequency, but it provides sufficient isolation over a frequency band to satisfy requirements for most applications.

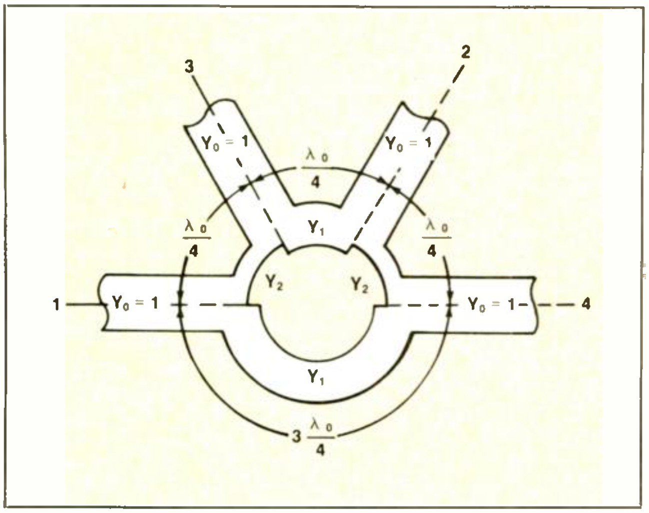

Fig. 1 Conventional hybrid-ring directional coupler.

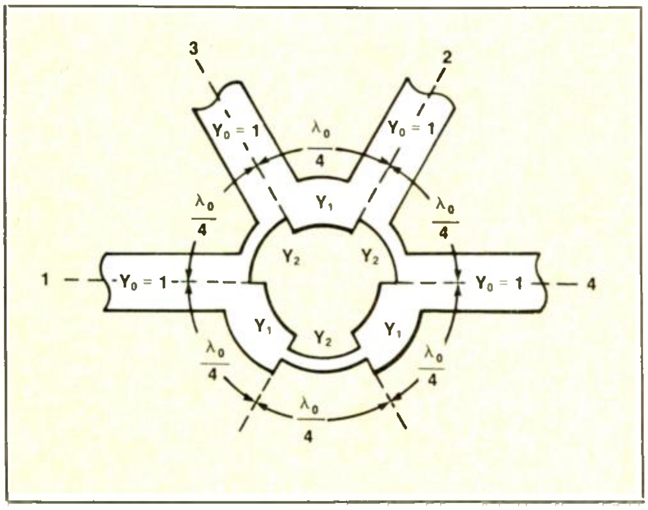

Fig. 2 Modified hybrid-ring directional coupler.

Analysis of the Hybrid-Ring Directional Coupler

The configuration of the modified hybrid-ring directional coupler is illustrated in Figure 2. The characteristic admittances of the four arms are equal and normalized to unity. The variable parameters are the two characteristic admittances Y_1 and Y_2 of the quarter-wave lines of the ring; these two admittances determine the power-split ratio between the two output arms and the impedance matching condition for the input arm.

The modified hybrid-ring coupler was analyzed by using the procedure for the analysis of symmetrical four-port networks[1,2] and by reducing the four-port network to a two-port network by taking advantage of the symmetry about the plane A-B. Then two in-phase waves of unit amplitude are applied to terminals 1 and 4 or to terminals 2 and 3, the current is zero at the plane A-B. As a result, the ring can be open-circuited at this plane, and only one half of the circuit needs to be analyzed. This condition is referred to as the even mode, and all parameters associated with this mode are denoted by subscript e (see Figure 3).

Similarly, when two opposite-phase waves of unit amplitude are applied to terminals 1 and 4 or to terminals 2 and 3, the voltage is zero at the plane A-B. As a result, the ring can be short-circuited at this plane, and only one half of the circuit needs to be analyzed.

[Continued on page 90]

CHARACTERISTIC IMPEDANCES OF THE LINES FOR TWO HYBRID-RING DIRECTIONAL COUPLERS

Power-Split Ratio (dB)	Conventional Hybrid Ring		Modified Hybrid Ring	
	Z_1	Z_2	Z_1	Z_2
0.0	70.7	70.7	70.7	70.7
1.0	75.1	67.0	72.3	66.9
2.0	80.4	63.9	74.5	63.8
3.0	86.5	61.3	77.3	61.1
4.0	93.7	59.1	80.8	59.0
5.0	102.0	57.4	84.9	57.1
6.0	111.6	55.9	89.8	55.6
7.0	122.6	54.8	95.6	54.4
8.0	135.2	53.8	102.3	53.5
9.0	149.5	53.1	109.9	52.7
10.0	165.8	52.4	118.6	52.1
11.0	184.3	51.9	128.4	51.6
12.0	205.2	51.6	139.4	51.2
13.0	228.9	51.2	151.9	50.9
14.0	255.5	51.0	165.9	50.7
15.0	285.6	50.8	181.6	50.5

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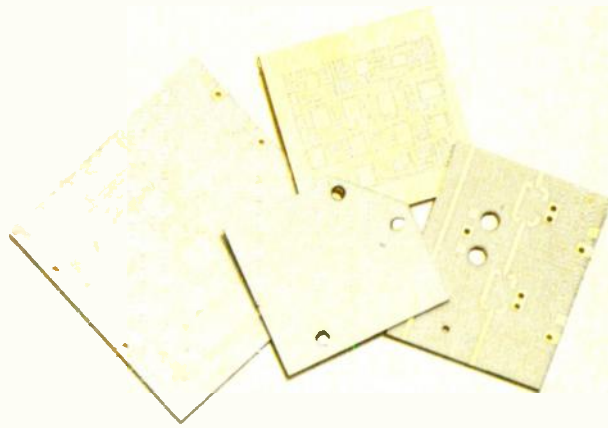

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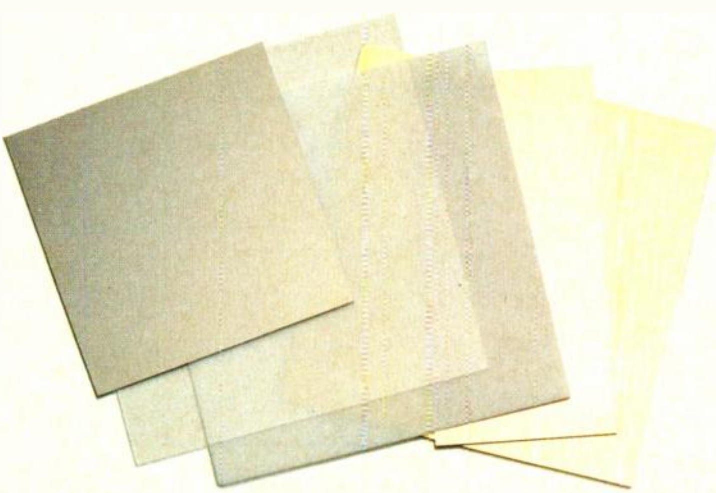

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This condition is referred to as the odd mode, and all parameters associated with this mode are identified by the subscript o (Figure 4). The equivalent circuits for these two modes are shown in Figures 3d and 4d, respectively. After deriving the scattering matrices for the even and odd modes, and superimposing the incident and reflected waves for these modes, the following resultant waves are obtained. For the first case, in which the incident waves are at arms 1 and 4, the resultant waves are given by:

$$a_1 = 2,\ a_4 = 0 \tag{4}$$

$$b_1 = \frac{2[1-(Y_2^2+Y_1^3/Y_2)^2+(Y_1^2-Y_1^4/Y_2^2)]}{(1+Y_2^2+Y_1^3/Y_2)^2+(Y_1-Y_1^2/Y_2)^2} \tag{5}$$

$$b_2 = \frac{-4Y_2(Y_1-Y_1^2/Y_2)}{(1+Y_2^2+Y_1^3/Y_2)^2+(Y_1-Y_1^2/Y_2)^2} \tag{6}$$

$$b_3 = \frac{-4jY_2(1+Y_2^2+Y_1^3/Y_2)}{(1+Y_2^2+Y_1^3/Y_2)^2+(Y_1-Y_1^2/Y_2)^2} \tag{7}$$

$$b_4 = \frac{4jY_1(Y_1/Y_2+Y_2^2+Y_1^3/Y_2)}{(1+Y_2^2+Y_1^3/Y_2)^2+(Y_1-Y_1^2/Y_2)^2}. \tag{8}$$

And for the second case, in which the incident waves are at arms 3 and 2, the resultant waves are given by:

$$a_3 = 2,\ a_2 = 0 \tag{9}$$

$$b_1 = \frac{-4jY_2(1+Y_2^2+Y_1^3/Y_2)}{(1+Y_2^2+Y_1^3/Y_2)^2+(Y_1-Y_1^2/Y_2)^2} \tag{10}$$

$$b_2 = \frac{-4jY_1[1+(Y_1/Y_2)(Y_2^2+Y_1^3/Y_2)]}{(1+Y_2^2+Y_1^3/Y_2)^2+(Y_1-Y_1^2/Y_2)^2} \tag{11}$$

$$b_3 = \frac{2[1-(Y_2^2+Y_1^3/Y_2)^2-(Y_1^2-Y_1^4/Y_2^2)]}{(1+Y_2^2+Y_1^3/Y_2)^2+(Y_1-Y_1^2/Y_2)^2} \tag{12}$$

$$b_4 = \frac{-4Y_2(Y_1-Y_1^2/Y_2)}{(1+Y_2^2+Y_1^3/Y_2)^2+(Y_1-Y_1^2/Y_2)^2} \tag{13}$$

The resultant waves for the two cases referred to as difference and sum modes are summarized in Figures 5a and 5b. The output voltage ratio between arms 3 and 4 for the difference mode in Figure 5a is given by:

$$\frac{b_3}{b_4} = -\frac{Y_2(1+Y_2^2+Y_1^3/Y_2)}{Y_1(Y_1/Y_2+Y_2^2+Y_1^3/Y_2)}, \tag{14}$$

[Continued on page 92]

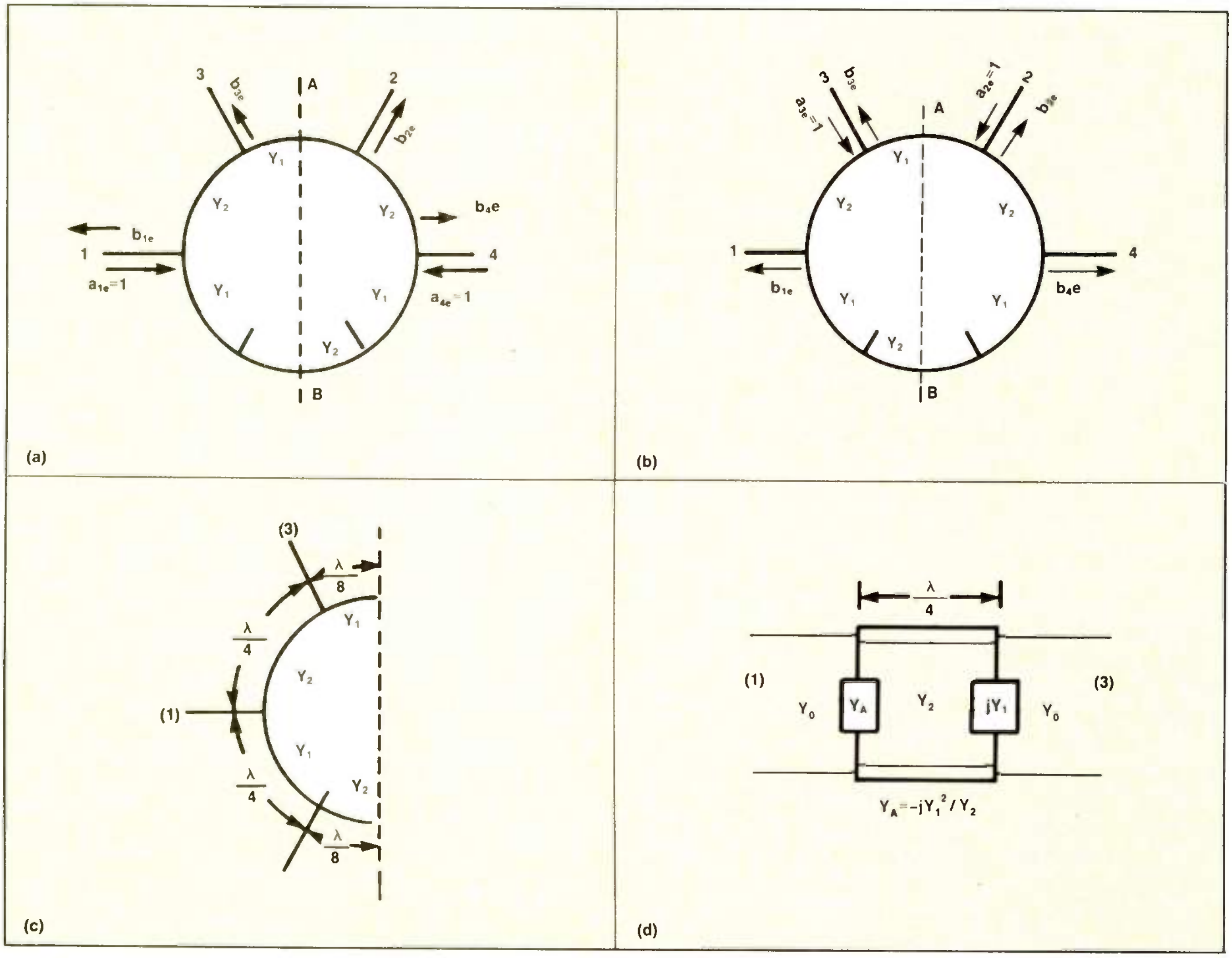

Fig. 3 Even mode: (a) incident waves at arms 1 and 4; (b) incident waves at arms 2 and 3; (c) open circuit at A-B; (d) equivalent circuit.

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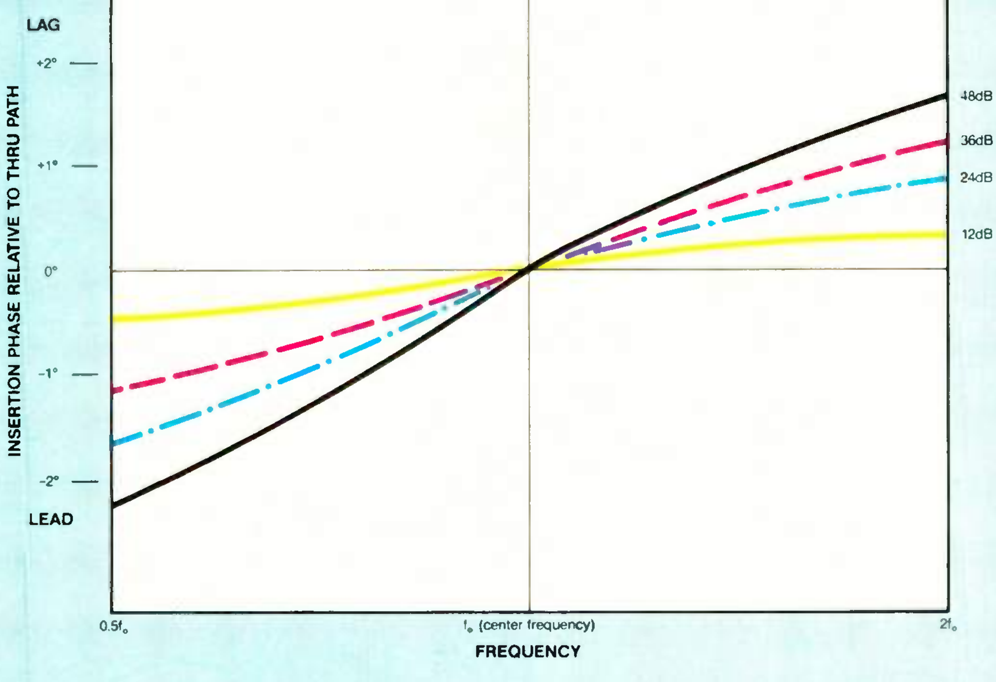

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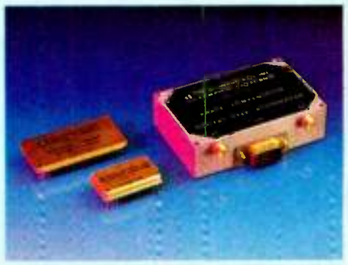

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and the output voltage ratio between arms 1 and 2 for this sum mode in Figure 5b is given by

$$\frac{b_1}{b_2} = \frac{Y_2(1 + Y_2^2 + Y_1^3/Y_2)}{Y_1[1 + Y_1/Y_2(Y_2^2 + Y_1^3/Y_2)]} \quad (15)$$

The condition that the input arm in both cases be matched requires that b_1 in Equation 5 and b_3 in Equation 12 be zero. A close look at Equations 5 and 12 indicates that a perfect match condition cannot be obtained simultaneously at both ports 1 and 3. A near-perfect match is achieved at ports 1 and 3 for the following condition:

$$Y_2^2 + \frac{Y_1^3}{Y_2} = 1 . \quad (16)$$

When Y_1 and Y_2 satisfy this condition, the resultant reflected waves in Equations 5 and 12 are given by

[Continued on page 94]

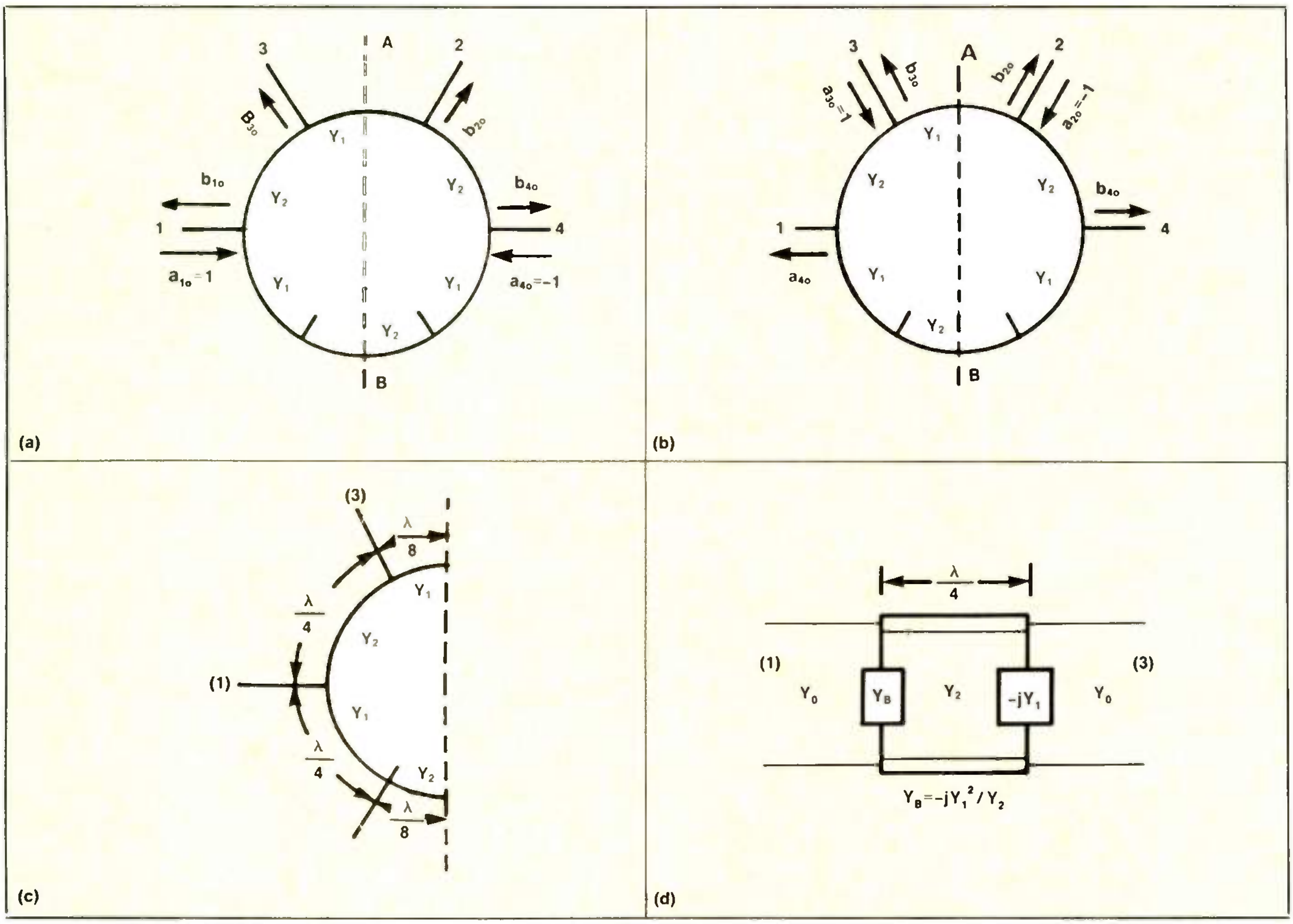

Fig. 4 Odd mode: (a) incident waves at arms 1 and 4; (b) incident waves at arms 2 and 3; (c) short circuit at A-B; (d) equivalent circuit.

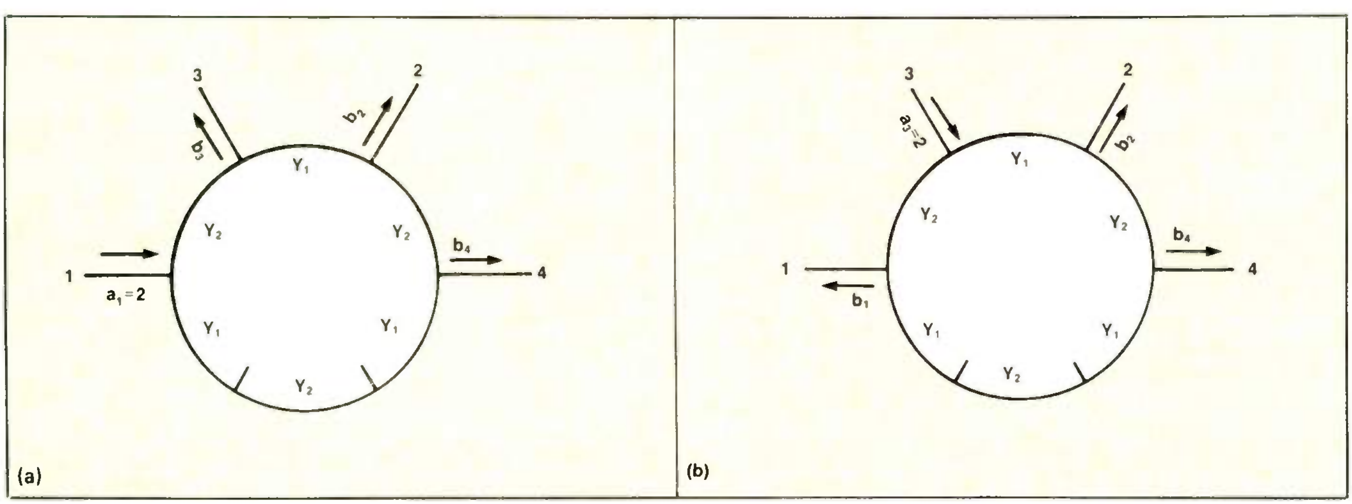

Fig. 5 Incident and reflected waves for the (a) difference and (b) sum modes.

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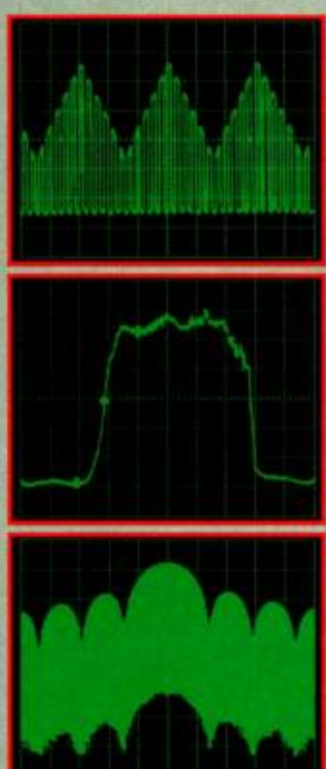

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$$b_1 = \frac{2(Y_1^2 - Y_1^4/Y_2^2)}{(1 + Y_2^2 + Y_1^3/Y_2)^2 + (Y_1 - Y_1^2/Y_2)^2} \quad (17)$$

(Difference mode)

$$b_3 = -\frac{2(Y_1^2 - Y_1^4/Y_2^2)}{(1 + Y_2^2 + Y_1^3/Y_2)^2 + (Y_1 - Y_1^2/Y_2)^2} \quad (18)$$

(Sum mode).

The output voltage ratio between arms 3 and 4 for the difference mode and arms 1 and 2 for the sum mode are given by

$$\frac{b_3}{b_4} = -\frac{2Y_2}{Y_1(1 + Y_1/Y_2)} \quad (19)$$

$$\frac{b_1}{b_2} = \frac{2Y_2}{Y_1(1 + Y_1/Y_2)}. \quad (20)$$

Equation 19 states that for an input voltage at arm 1, the output voltages at arms 3 and 4 are 180° out of phase. Equation 20 states that for an input voltage at arm 3, the output voltages at arms 1 and 2 are in phase. The characteristic impedance values Z_1 and Z_2 for various power-split ratios between the output ports are tabulated in Table 1 for comparison with the conventional hybrid-ring coupler.

The magnitude of the reflected wave in Equations 17 and 18, for values of Y_1 and Y_2 corresponding to the characteristic impedances in Table 1 for various power-split ratios, is very small; it results in an SWR of less than 1.1 for smaller power-split ratios and decreases to 1.05 for higher power-split ratios, establishing the useful range for this directional coupler. By symmetry, if the input waves are incident at ports 4 and 2, the reflected wave amplitudes will be the same as in Equations 17 and 18, respectively. Thus, the input impedance at any port is matched (SWR $<$ 1.1) when the other arms are terminated by matched impedances.

The characteristic impedance values Z_1 and Z_2 for various power-split ratios and the isolation values between output ports for the sum mode are plotted in Figures 6 and 7, respectively. The isolation between the output ports is about 20 dB for power-split ratios of less than -6 dB. As the power-split ratio approaches 0 dB (equal split, "rat-race" ring), the isolation approaches infinity.

Although the theoretical isolation between the output ports for the modified hybrid ring is on the order of 20 dB, as compared to the finite isolation for the conventional hybrid ring, the average difference between the measured results over a frequency band for the two cases is approximately 5 dB. A 20 dB isolation is adequate for most applications.

Comparing the impedance values for the modified hybrid ring and for the conventional ring (Table 1), the range of the realizable power-split ratio is significantly increased for the same realizable impedance values. For example, for a power-split ratio of 8 dB, the conventional ring requires a line of impedance of 135.2 ohms, while for a 12 dB power-split ratio, the modified ring requires a line of impedance of 139.4 ohms. Thus, the realizable range of power-split ratio for the modified hybrid-ring directional coupler is increased by approximately 4 dB.

The results obtained above are for a single frequency. To find the frequency dependence of this hybrid coupler, it is necessary to consider the variations in the lengths of the quarter-wave lines with the frequency in the equivalent circuits. The frequency characteristics of this hybrid-ring coupler were obtained by analyzing the circuit on MIDAS (a computer program for analyzing microwave circuits).[3] The power-split ratios, isolation between ports, phase difference between output ports and input SWR for a ring with power-split ratio of 6.4 dB are plotted in Figures 8 and 9 as a function of frequency (15 to 18 GHz), along with the experimental results discussed in the next section. From these figures, we note that the hybrid-ring coupler has a bandwidth of approximately 20 percent at Ku band.

Experimental Results

A stripline hybrid-ring directional coupler with a power-split ratio of 6.4 dB between the two output arms 1

[Continued on page 96]

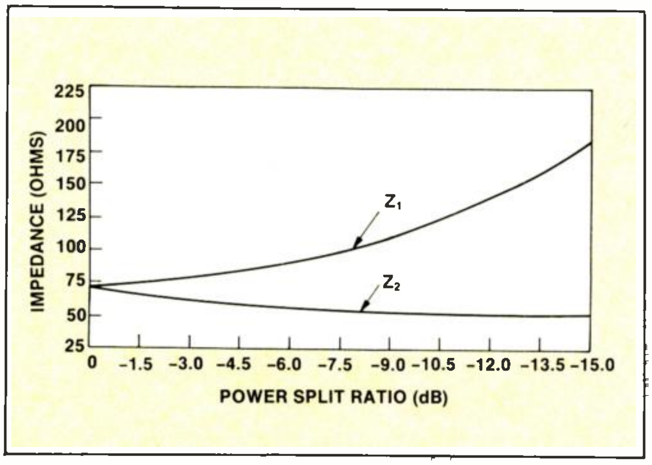

Fig. 6 Characteristic impedances of the transmission lines vs power-split ratios.

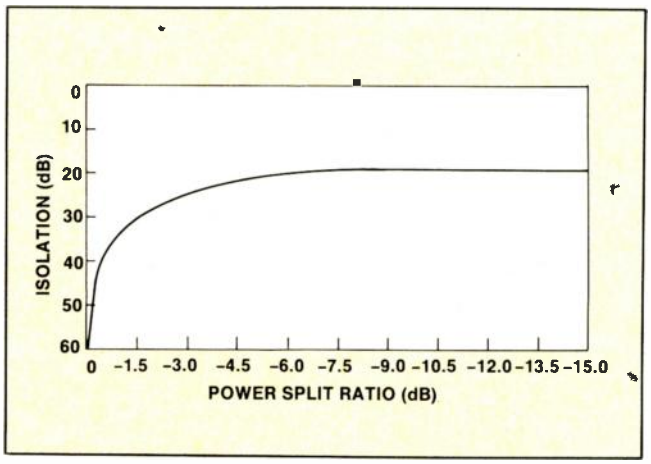

Fig. 7 Isolation between the output ports vs power-split ratios.

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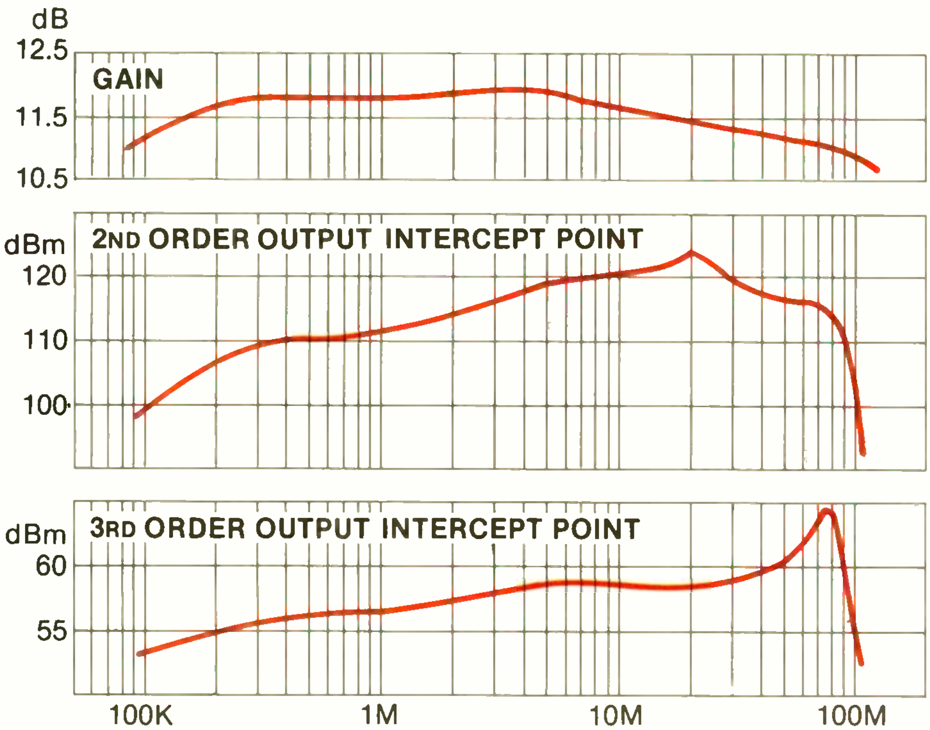

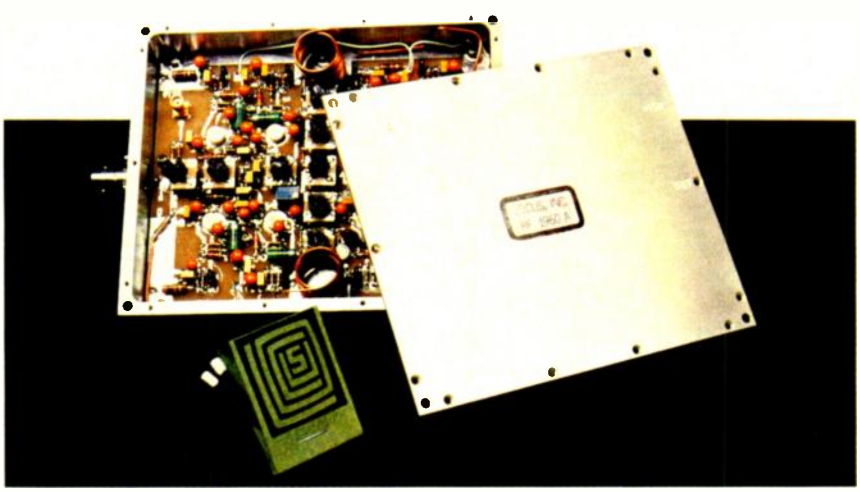

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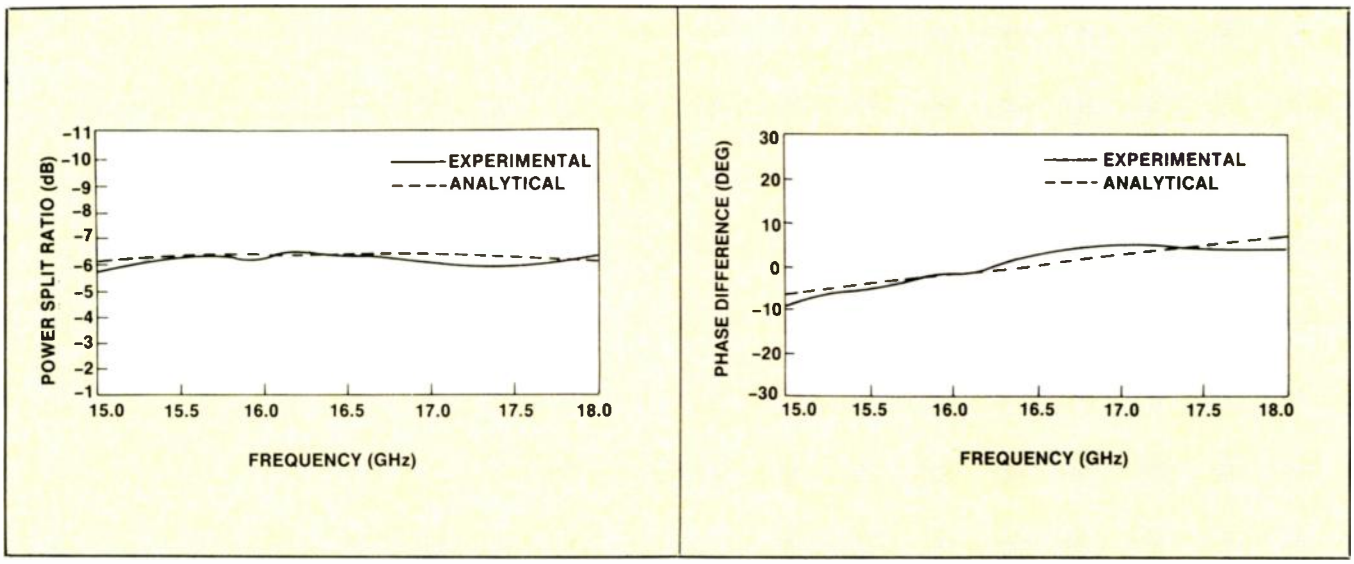

Fig. 8(a) Power-split ratio between ports 1 and 2 and (b) phase difference between ports 1 and 2 of the 6.4 dB hybrid-ring coupler.

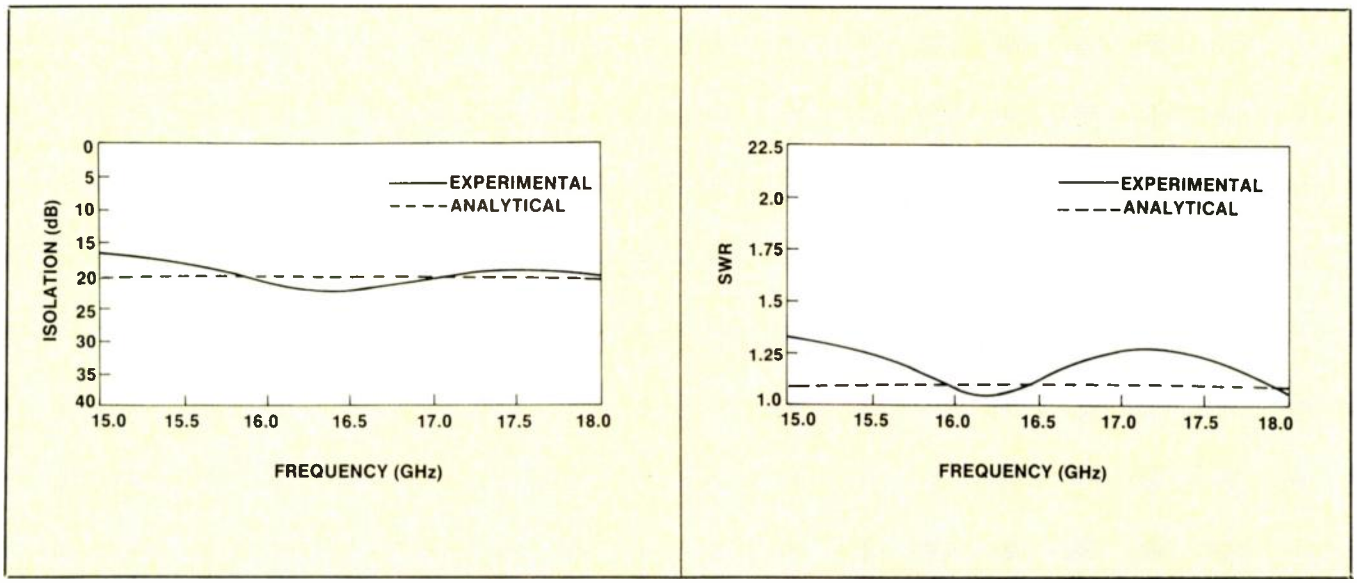

Fig. 9(a) Isolation between ports 1 and 2 and (b) input SWR at port 3 of the 6.4 dB hybrid-ring coupler.

and 2 (sum mode) at the center frequency of 16.5 GHz was designed and tested. The impedance values used for this ring are Z_1 = 92.0 ohms and Z_2 = 55.1 ohms. The measured power-split ratios, isolation, phase difference between the output ports and input SWR, along with the analytical results, are shown in Figures 8 and 9. A close agreement between the analytical and experimental results is observed over the frequency range of 15 to 18 GHz. The input port is fairly well matched, with an SWR of less than 1.5 over the frequency band.

The stripline hybrid rings were fabricated on a 50 mil duroid 5880 substrate. Duroid 5880 substrate has a dielectric constant of 2.2 and provides low circuit loss.

Conclusion

The analysis and design of a hybrid-ring directional coupler that provides a higher power-split ratio than the conventional hybrid-ring coupler is presented. The measured results for two couplers developed in stripline configuration at Ku band agree well with the analytical results. The hybrid-ring directional coupler has a bandwidth of approximately 20 percent. The input arm of the coupler is matched (SWR $<$ 1.5) over the frequency band. The power-split ratio can be adjusted by varying the characteristic impedances (Z_1 and Z_2) of the two lines forming the ring. Similar hybrid-ring directional couplers also can be realized in microstrip configuration.

Acknowledgment

The authors would like to acknowledge several helpful discussions with Farzin Lalezari of Ball Aerospace Corp. and to thank Vince Zvanya for his contribution in the fabrication of the hybrid rings. ■

[Continued on page 98]

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From 1973 to 1974, he was a research fellow at the Indian Institute of Technology, Kharagpur, India. From 1976 to 1982, he worked as a research scientist at Mission Research Corp. in Albuquerque and from 1982 to 1983 he worked as a senior research engineer at Dikewood Corp. in Albuquerque.

During 1982 to 1983, he also was an adjunct faculty member at the University of New Mexico, where he taught graduate courses on antennas. In 1983, he joined the Missile and Surface Radar Division of RCA Corp. in Moorestown, NJ as a principal member of the engineering staff. Since then, he has been involved in R&D work on phased array and microwave antennas.

Agrawal has authored over 25 journal articles and conference presentations.

He was vice chairman of the AP-S/MTT-S/EMC Albuquerque chapter during 1982 to 1983 and is currently chairman of the AP-S/MTT-S Philadelphia chapter.

Agrawal is a member of Tau Beta Pi.

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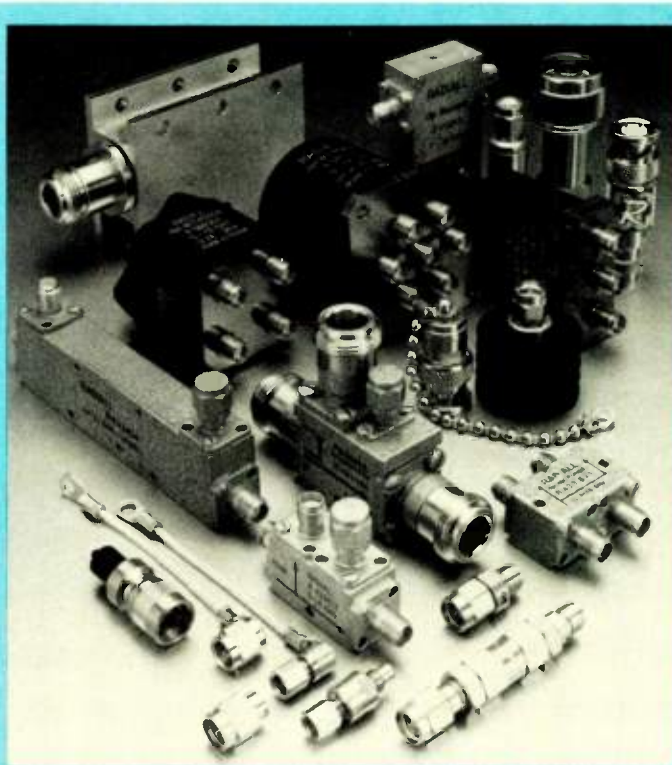
From 1983 to 1985, He was with Combat Surveillance and Target Acquisition Laboratory, Fort Monmouth, NJ, where he was involved in the development of electronic scan techniques for mm-wave antennas.

Since 1985, he has been with the Missile and Surface Radar Division of RCA Corp. in Moorestown, NJ, where he is actively involved in the design and development of microwave circuits and antennas.



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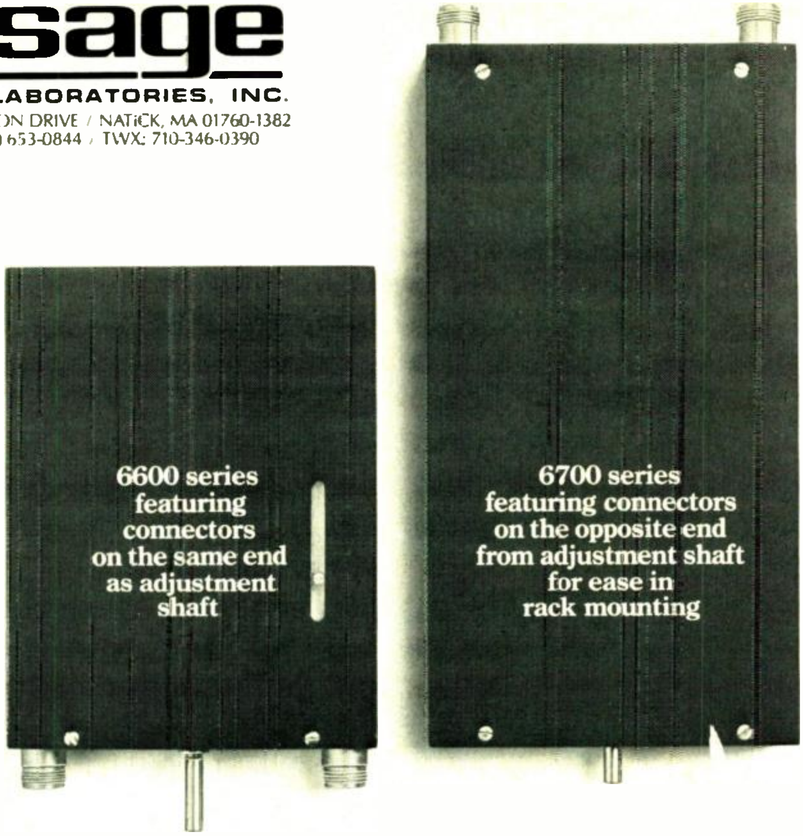

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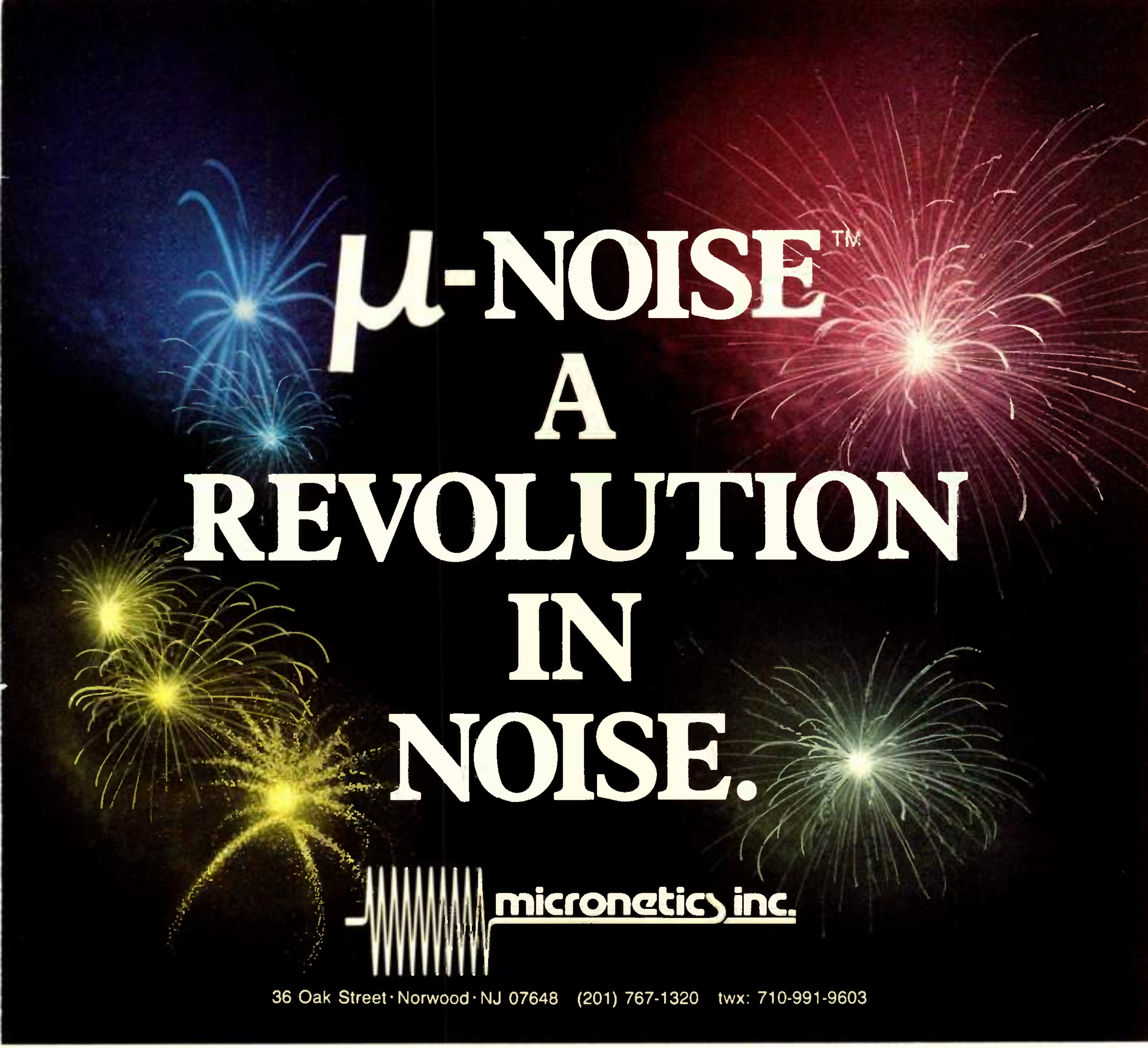
Frequency Range (GHz)	DC-.75	DC-2.0	DC-4.0	DC-8.0	DC-18.0
Minimum Phase-Shift (Deg./GHz)	900	290	190	95	40
Nominal Phase-Shift (Deg./GHz/Shaft-Turn)	14	5.7	5.7/3.8	1.9	1.5
Nominal Insertion Phase at Minimum Phase Setting (Deg./GHz)	1560	950	790/700	450	140
Maximum VSWR	1.20	1.35	1.50	2.00	1.60
Maximum Insertion Loss (dB)	1.0	0.5	0.5	1.0	1.5
Power: Average (Watts/$\sqrt{GHz}$)	100	350	350	350	150
Peak (Watts)	1000	2500	2500	2500	1000

CIRCLE 78 ON READER SERVICE CARD



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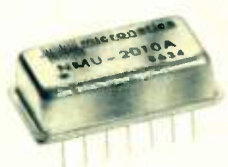
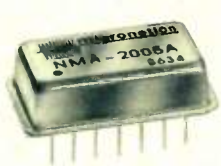


- Package:
  14 Pin DIP
- Output:
  5dB, 15dB, 30dB ENR
- Upper Freq. Limit:
  500MHz to 1.5GHz
- Pulse:
  TTL Input

CIRCLE 79

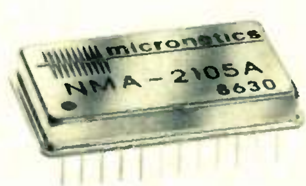
μ-NOISE
NMA 2000 Series



- Package:
  14 Pin DIP
- Output:
  10 millivolts RMS
- Upper Freq. Limit:
  100KHz to 500MHz

CIRCLE 80

μ-NOISE
NMA 2100 Series



- Package:
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- Output:
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- Upper Freq. Limit:
  100KHz to 500MHz

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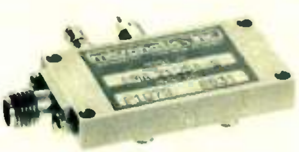
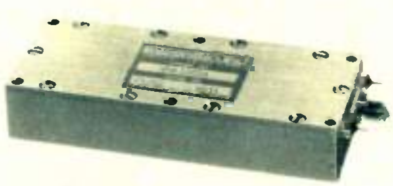


- Package: 1.5 x .8 x .25 inches
- Output: 1 volt RMS
- Upper Freq. Limit: 100KHz to 500MHz

CIRCLE 82

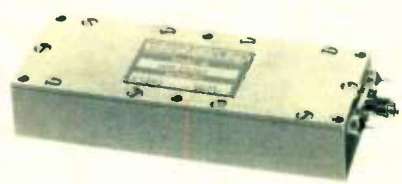
μ-NOISE NMA 2200 Series



- Package: 3.75 x 1.5 x .7 inches
- Output: + 13dBm
- Upper Freq. Limit: 100KHz to 300MHz
- Output VSWR: 1.2 to 1

CIRCLE 83

μ-NOISE NMA 2300 Series



- Package: 3.75 x 1.5 x .7 inches
- Output: 3 Watts Peak
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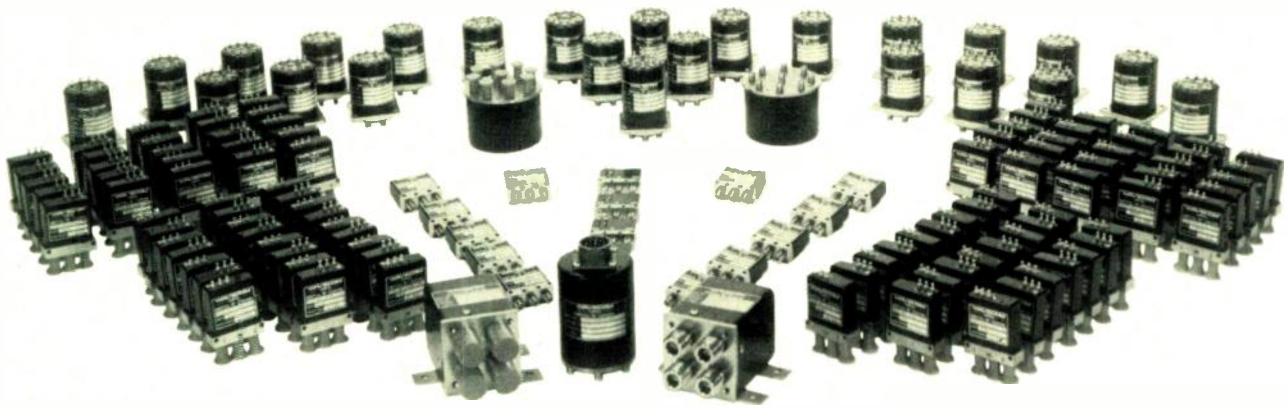

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Select the appropriate number or letter in each group, starting with Group 1. The 3rd digit in Group 1 indicates envelope size. Make one or more choices from Group 3; only one from each other group.

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SPECIFICATIONS (Typical)

- VSWR from 1.15:1 (3 GHz) to 1.5:1 (26 GHz).
- Isolation from 100 db (2 GHz) to 58 db (26 GHz).
- Insertion loss from 0.1 db (3 GHz) to 0.5 db (26 GHz).

GROUP 1 SWITCH SERIES

010 - SPST	063 - SP6T
020 - SP2T	064 - SP6T
022 - SP2T	070 - SP7T
023 - SP2T	080 - SP8T
024 - SP2T	090 - SP9T
030 - SP3T	100 - SP10T
032 - SP3T	110 - SP11T
033 - SP3T	120 - SP12T
040 - SP4T	130 - 2P2T
042 - SP4T	132 - 2P2T
043 - SP4T	140 - 2P6T
050 - SP5T	150 - 2-SP2T
052 - SP5T	160 - 4P4T*
053 - SP5T	170 - 4P8T*
060 - SP6T	180 - 8P8T*
062 - SP6T	190 - 8P10T*

031 - 101 - HIGH FREQUENCY
035 - 065 - HIGH FREQUENCY MINIATURIZED
7XXX - SPACE QUALIFIED
8XXX - CUSTOMER ASSIGNED NUMBER
9XXX - SPECIAL
916 - SP16T
936 - SP36T

GROUP 3 OPTIONS

1 - 50-OHM TERMINATION
2 - INDICATOR CIRCUITRY
3 - SUPPRESSION DIODES
4 - TTL LOGIC HIGH
5 - TTL LOGIC LOW
6 - MANUAL OVERRIDE
7 - SELF-DE-ENERGIZING CIRCUITRY
8 - HIGH POWER HANDLING
9 - OTHER (SPECIFY)
X - 75 OHM TERMINATION
Y - BCD TTL DECODER

GROUP 5 TERMINALS

1 - SOLDER
2 - POWER CONNECTOR/ FAST DISCONNECT (STANDARD)
3 - POWER CONNECTOR/ FAST DISCONNECT (SPECIFY TYPE)
4 - OTHER (SPECIFY)

GROUP 7 FREQUENCY

1 - DC - 3 GHz
2 - DC - 8 GHz
3 - DC - 12.4 GHz
4 - DC - 18 GHz
5 - DC - 18.5 GHz
6 - DC - 26.5 GHZ
7 - DC - OTHER (SPECIFY)
8 - DC - 19.5 GHz
9 - DC - 22 GHz
X - DC - 24 GHz
X1 - DC - 32 GHz*
X11 - DC - 40 GHz*

GROUP 9 MOUNTING

1 - BRACKET
2 - FLANGE
3 - OTHER (SPECIFY)

TYPICAL PART NUMBER†

022 - B 1234 - A 1 C - 6 C 0

GROUP 2 ACTUATING

A - FAILSAFE
B - LATCHING
C - LATCHING RESET
D - NORMALLY OPEN
E - OTHER (SPECIFY)
DA - NORMALLY OPEN FAILSAFE TO POSITION 1
F - MAKE BEFORE BREAK
G - MAKE BEFORE BREAK WITH SWITCHING TIME ADJUSTMENT

GROUP 4 RF CONNECTORS

A - SMA	E - TNC
B - BNC	F - SC
C - F	G - OTHER (SPECIFY)
D - N	H - P/C BOARD MOUNTING

GROUP 6 VOLTAGE

A - 6 VDC	F - 110 VAC
B - 12 VDC	G - OTHER (SPECIFY)
C - 24 VDC	H - 15 VDC
D - 28 VDC	I - 18 VDC
E - 48 VDC	J - 20 VDC

GROUP 8 POLARITY

A - COMMON PLUS
B - COMMON MINUS
C - NOT RELEVANT TO APPLICATION OR SWITCHES WITH LOGIC

* RF switching soon available.
† Sample part number identifies: SP2T switch, latching, 50-ohm termination, indicator circuitry, suppression diodes, TTL logic, SMA connectors, solder terminals, 24 vdc, DC - 26.5 GHz, polarity non-relevant.
‡ Patent pending (advanced design for 50-ohm terminations for all multiposition switches, to 26.5 GHz).

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Very High Power mm-Wave Components in Oversized Waveguides

M. Thumm,
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R. Wilhelm
Institut fur Plasmaforschung
Universitat Stuttgart
Stuttgart, FRG

Introduction

High power gyrotron tubes are used for plasma startup and electron cyclotron resonance heating (ECRH) for magnetic confinement fusion research and for plasma production in industrial processes, such as isotope separation. Millimeter-waves of 40 GHz frequency are required to obtain reactor relevant plasma densities at optimum magnetic confinement conditions. Commercial gyrotrons launch mixtures of circular electric TE_{0n} modes (mainly TE_{02} at frequencies between 28 and 70 GHz and, in the near future, TE_{03} or TE_{04} at frequencies between 100 and 140 GHz[1,2]) into an oversized circular waveguide (inner diameter = 63.5 mm). In particular, the TE_{01} mode is appropriate for transmission through long runs of straight circular waveguide because of its low ohmic wall losses.

However, TE_{0n} modes are unpolarized and produce a hollow conical radiation pattern with zero power along the waveguide axis. Efficient plasma heating or high power mm-wave radar techniques (space communication, high resolution radar) and next generation linear particle accelerators require an axisymmetric, narrow, pencil-like beam with well-defined polarization. The almost perfectly linearly polarized Gaussian HE_{11} hybrid mode, radiated from an open-ended, circumferentially corrugated, oversized circular waveguide, satisfies these conditions best.[3] This mode is ideal for quasi-optical propagation and launching by the use of focusing mirrors and polarization twist reflectors.[4,5] The HE_{11} mode can be generated from TE_{0n} gyrotron modes by the two multistep mode conversion sequences:

$$TE_{0n} \rightarrow TE_{01} \rightarrow TE_{11} \rightarrow HE_{11}$$

(references 5 and 6)

and

$$TE_{0n}\ TE_{01} \rightarrow TM_{11} \rightarrow HE_{11}$$

(reference 4).

The first approach, which uses the TE_{11} mode as polarized intermediate mode, has the advantage that the converters all can be made without bends, allowing an arbitrary choice and fast change of the polarization plane by simply rotating the TE_{01}-to-TE_{11} converter around its axis. However, the second method is more suitable for high power, 140 GHz transmission lines because efficient TE_{01}-to-TM_{11} transducers can be made half as long as the corresponding TE_{01}-to-TE_{11} mode converters, which must be made very long in order to suppress the undesired TE_{12} and TE_{21} modes.[7] Moreover, due to the relatively large number of coupling periods, the TE_{01}-to-TE_{11} transducer bandwidth is inherently narrower than that of the TE_{01}-to-TM_{11} transformer. The bandwidth of the corrugated TE_{11}-to-HE_{11} converter also is reduced, relative to that of the TM_{11}-to-HE_{11} transformer, by the need to start the corrugation depth at $\lambda/2$.[4,8,9]

This paper reports computations and measurements on mode converter systems of the first type at 28 and 70 GHz and of the second type at 140 GHz. The structure of wall perturbations in the rippled wall mode converters and the curvature distribution in the bent, smooth-walled TE_{01}-to-TM_{11} mode transducer were optimized by numerically solving the proper coupled-wave differential equations. Computer-aided optimization of circumferentially corrugated mode converters has been achieved with a scattering matrix code employing the modal field expansion technique: modular analysis concept (MAC).[10] All optimum mode transformers were fabricated by direct machining on a numerically controlled lathe and assembled from individual sections.

The mode purity in the transmission lines is conserved by using numerically optimized diameter tapers with non-linear profiles and corrugated gradual waveguide bends with tapered curvature and matched corrugation. Special mode-selective corrugated-wall mode filters (for asymmetric modes) with anisotropic surface reactance are used in order to protect the gyrotron from excessive reflections

This article is based on one contributed to the MIOP 1986 conference held in West Germany last June.

and to avoid arcing in the waveguide caused by trapped mode resonances. Mode content and reflected power are determined with a novel device that measures the spectrum of axial wavenumbers in oversized waveguide (k-spectrometer).[11] The mode beating structure in the different sections of the transmission lines is measured with stacked thermographic paper mode pattern analyzers (burn patterns). Absolute power calibration is performed with calorimetric loads containing an organic absorbing and cooling fluid with significantly larger attenuation length than water. Air-cooled loads with solid absorbers (fire clay) serve as convenient, reflection-free dummy loads for gyrotron and component test measurements.

High power operation of the complete transmission line has been successfully demonstrated at 28 and 70 GHz using pulsed gyrotrons (200 kW, 40 to 100 ms). A 70 GHz/1 MW-ECRH system (3 s, 5 transmission lines) and a 140 GHz/200 kW transmission line are in preparation.

ΣTE_{0n}-to-TE_{01} Mode Conversion

In both conversion methods, mode transducers with axisymmetric, periodic radius perturbations ($\Delta m=0$) convert TE_{0n} into $TE_{0,n-1}$ power, provided the converter radius is periodic with the beat wavelength

$$\lambda_B(n,n-1)$$

of the two modes.[7,13,14] Efficiency optimization is studied by numerically solving the corresponding coupled-mode differential equations for the complex forward-traveling wave amplitudes of all propagating TE_{0n} modes in the waveguide.[7] In the case of small waveguide distortions far from cutoff, reflections can be ignored. Ohmic attenuation is included in the analysis. The waveguide radius is approximately given by

$$a(z) = a_0 \left[1-\epsilon_0 \cos\left[(2\pi z/\lambda_W) \times (n,n-1)\right]\right]/(1-\epsilon_0). \qquad (1)$$

The optimum wavelength $\lambda_W(n,n-1)$ of the wall perturbations differs slightly from $\lambda_B(n,n-1)$, resulting in a reduction of the remaining TE_{0n} mode content.[15] The conversion efficiencies η_0 were further improved by superimposing small amounts of additional, phase matched, higher harmonic modes on the wall structure[15] and/or by inserting cylindrical phase shifters[7] in order to reduce the amplitudes of the undesired modes $TE_{0,n+1}$ and $TE_{0,n-2}$ at the output. In this case of symmetrical cosine-perturbations, the conversion efficiencies from $TE_{0,n-1}$ to TE_{0n} and from TE_{0n} to $TE_{0,n-1}$ are identical. Results of optimization calculations for several signal-step 200 kW TE_{0n}-to-$TE_{0,n-1}$ mode transducers (n=2 to 6) at 140 GHz (a_0=13.9 mm) are summarized in Table 1.

The numerical computations show that the number N of main coupling periods of the various mode transducers may be somewhat reduced without substantial reduction of the conversion effi-

[Continued on page 106]

GEOMETRICAL DATA, CALCULATED CONVERSION EFFICIENCIES (OHMIC ATTENUATION IS INCLUDED) AND BANDWIDTH FACTORS (AT 0.3% FREQUENCY DEVIATION) OF 140 GHz TE_{0n}-to-$TE_{0,n-1}$ MODE CONVERTERS (n=2-6, a_0 = 13.9 mm)

	$TE_{06} \rightarrow TE_{05}$	$TE_{05} \rightarrow TE_{04}$	$TE_{04} \rightarrow TE_{03}$	$TE_{03} \rightarrow TE_{02}$	$TE_{02} \rightarrow TE_{01}$
Converter length L	797 mm	862 mm	945 mm	1073 mm	1352 mm
Number of periods N	14	12	10	8	6
Perturbation ϵ_0	0.0063	0.0089	0.0137	0.0238	0.0535
Conv. efficiency η_0	0.986	0.986	0.988	0.995	0.997
Bandwidth $1/(1+\delta^2)$	0.992	0.994	0.996	0.998	0.999

ciency. In the case of TE_{02}-to-TE_{01} transformation, efficiencies of approximately 99 percent can be achieved with merely two geometrical periods (N=2).

The frequency sensitivity of rippled-wall mode transducers is discussed in reference 14. The conversion efficiency η is approximately given by

$$\eta = \eta_0/(1+\delta^2)$$

where δ=2N(Δf/f_0). Numerically reduced bandwidth factors $1/(1+\delta^2)$ for the various mode transformers at a typical gyrotron frequency deviation of 0.3 percent also are given in Table 1. TE_{03} or higher TE_{0n} gyrotron modes (e.g. TE_{04}) can be converted to the basic TE_{01} mode by cascading the corresponding single-step mode transducers, resulting in a total conversion efficiency of 99.2 percent or 98.0, respectively; ohmic losses are included (Cu-waveguide). Direct TE_{03}-to-TE_{01} conversion is impossible because

$$\lambda_B(3,1) \approx \lambda_B(3,4),^7$$

whereas a TE_{04}-to-TE_{01} conversion efficiency of 98.0 percent also was obtained by using a direct TE_{04}-to-TE_{01} mode transformer (N=46, ϵ=0.0173, L=1.95 m). However, the bandwidth of such a high N-number converter is inherently narrow ($1/(1+\delta^2)$=0.929 for $\Delta f/f_0$=0.3 percent).

The efficiencies of TE_{0n}-to-$TE_{0,n-1}$ transducers for 70 GHz (a_0=13.9 mm) and 28 GHz (a_0=31.7 mm) are

[Continued on page 108]

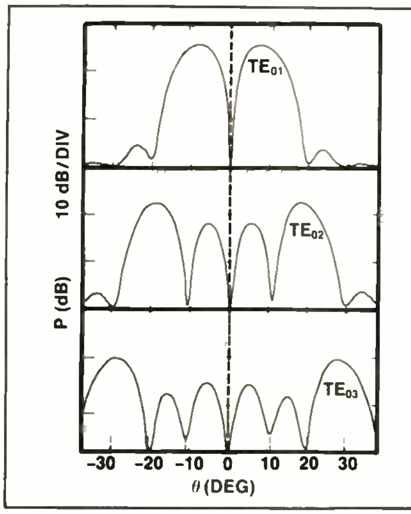

Fig. 1 Measured far-field radiation patterns (low-power measurements) of the TE_{02} and TE_{03} modes successively generated from a pure TE_{01} input mode at 70 GHz (inner waveguide radius a_0 = 13.9 mm). Aperture ϕ = 27.8 mm.

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The short, low torque, high performance coax rotary joints shown, as well as MDL's extensive line of waveguide rotary joints have set the standards of the industry. We developed low resistance, low noise contacts for these coax rotary joints based on our long experience in the design and manufacture of slip rings. These contacts have the proven advantage of long life. The center conductor design is covered by a U.S. patent. Moreover, these units meet or exceed MIL-E-5400 and MIL-E-16400 specifications. Call or write Microwave Development Laboratories, Inc. for units to meet your specific requirements, such as mounting flanges.

MODEL	CONNECTOR	FREQUENCY (GHz)	VSWR (MAX)	MAX INSERTION LOSS IN dB	MAX WOW	LENGTH	MAX STARTING TORQUE
180RS46	"SMA"	DC-18	1.25 DC-10 GHz 1.5 10-18 GHz	0.2 to 10 GHz 0.5 to 10-18 GHz	1.02	1.00"	.5 oz.-in.
120RK36	"N"	DC-12.4	1.15 DC-6 GHz 1.25 6-12.4 GHz	0.2 DC to 10 GHz 0.3 10 to 12.4 GHz	1.01	1.95"	.5 oz.-in.

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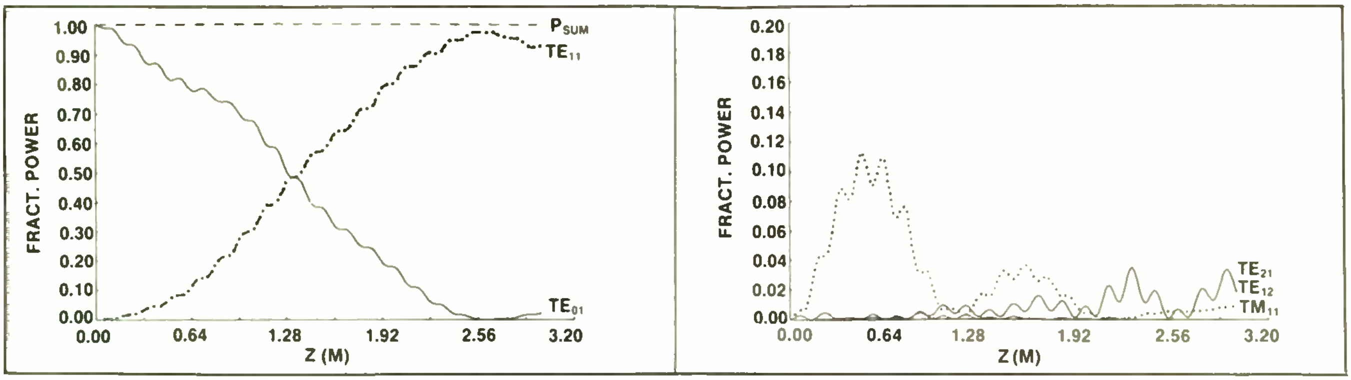

Fig. 2 Calculated fractional power in each mode along the optimum 70 GHz TE_{01}-to-TE_{11} converter (a_0 = 13.9 mm).

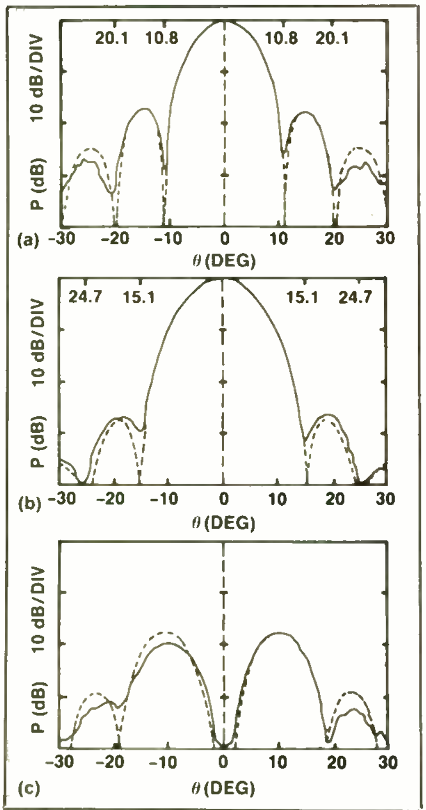

Fig. 3 Measured E-plane (a), H-plane (b) and 45° cross-polarization (c) far-field patterns of the TE_{11} mode generated by the optimum 70 GHz TE_{01}-to-TE_{11} converter (low power measurements). Aperture ϕ = 27.8 mm. The theoretical patterns are superimposed (dashed curves).

almost the same as for the 140 GHz converters because, for a given number N of geometrical periods, the ratio of the coupling coefficient $C_{n,n'}$ between TE_{0n} and $TE_{0n'}$ modes (for a smooth radius change) and the difference $\Delta\beta_{n,n'}=2\pi/\lambda_B(n,n')$ of the unperturbed wavenumbers remains constant.

The experimental conversion efficiencies of the 70 GHz mode converters, which were measured with the k-spectrometer and by determination of insertion losses through two back-to-back converters using frequency sweeping and phase shifting techniques, are in good agreement with the predicted values

$$\eta_{exp}(TE_{01}/TE_{02})=(99.5\pm0.5) \text{ percent}$$

and

$$\eta_{exp}(TE_{02}/TE_{03})=(99.0\pm0.7) \text{ percent}.$$

This also is valid for the 28 GHz mode transformers.[7] Figure 1 shows the measured far-field radiation patterns of the TE_{02} and TE_{03} modes successively produced from a pure TE_{01} input mode at 70 GHz. The agreement of these patterns with those that were computed is good and consistent with calculated efficiencies.

Phase-locked gyrotron mode mixtures can be converted to pure modes by using matched TE_{0n}-to-$TE_{0,n-1}$ transformer sections at specified axial positions of the longitudinal beat structure in the waveguide (phase matching with bellows).[7] The experimentally determined ΣTE_{0n} (mainly TE_{02})-to-TE_{01} conversion efficiency is (99±1) percent at 28 and 70 GHz (99.5 percent predicted).

TE_{01}-to-TE_{11} Mode Converter

Conversion of the circular symmetric TE_{01} mode to the almost linearly polarized TE_{11} mode is achieved by a non-axisymmetric mode transducer with constant radius and periodically perturbed curvature in one plane (Δm=1).[7,14] The waveguide radius coordinate in a rotatable converter (arbitrary choice and fast change of the polarization plane) with simple cosine-shaped perturbation is given by

$$a(z,\phi) = a_0\,[1+\epsilon_1 \times \cos(2\pi z/\lambda_B(01,11))\cos\phi], \quad (2)$$

where ϕ is the azimuthal angle with ϕ=0, π in the plane of the wiggles. The E-field plane of the TE_{11} mode is perpendicular to this plane. Six coupled modes were included in the theoretical analysis: TE_{01}, TE_{11}, TE_{12}, TE_{21}, TM_{11}, and TM_{21}. Ohmic attenuation and possible ellipticity coupling, introduced by the manufacturing procedure, are included in the coupling matrices. According to the multi-mode computations, the conversion efficiency η_0=81.6 percent (unwanted modes: 6.9 percent TE_{01}, 4.8 percent TE_{12} and 0.7 percent TE_{21}; ohmic attenuation in Cu-waveguide: 1.4 percent of an eight-period 70 GHz converter (a_0=13.9 mm, L=2.49 m, ϵ_1=0.0594) can be improved to 94.0 percent by a 1.6 percent increase of the perturbation period ($\lambda_W=1.016\lambda_B(01,11)$). This reduces the unwanted TE_{12}-mode level and causes a continuous rematching of the required phase difference between the TE_{01} and TE_{11} waves,[7] resulting in a reduction of the remaining TE_{01} mode content from 6.9 to 0.9 percent.

The decisive improvement of the efficiency up to η_0=97.4 percent (unwanted modes: 0.6 percent TE_{01}, 0.4 percent TE_{12} and 0.1 percent TE_{21}) is achieved by superimposing two additional, phase-matched, small perturbations, whose geometrical periods correspond to the beat wavelengths between the TE_{01} input mode and TE_{12} and the TE_{11} output mode and TE_{21}, respectively.[15] Such improved perturbation structures result in a minimization of the TE_{12} and TE_{21} mode levels at the converter output. In Figure 2 the calculated fractional power in each mode along the optimum 70 GHz TE_{01}-to-TE_{11} converter is plotted. Figure 3a shows measured far-field

[Continued on page 110]

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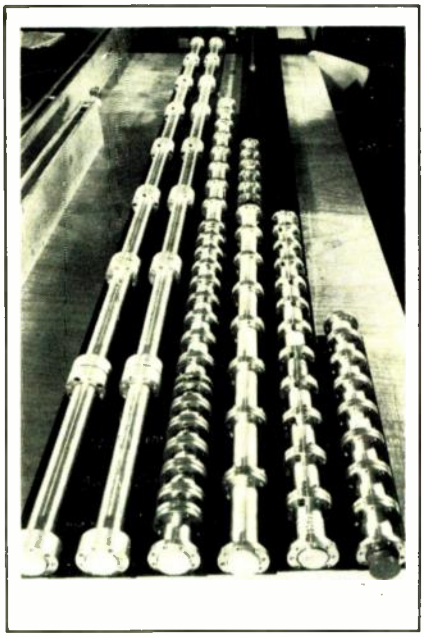


Fig. 4 Mode converter assemblies.

E and H planes as well as cross-polar (XP 45°) patterns of the TE_{11} mode, generated from a pure TE_{01} input mode. Figure 4 depicts a series of mode converter assemblies. The TE_{01}-to-TE_{11} mode converters are shown at the left of the photograph.

The agreement of the power distributions with theoretical ones (dashed curves) is good and consistent with the calculated efficiency. This far-field radiation pattern measurement, together with insertion loss measurements using a second converter (and re-conversion multimode calculations) and measurements with the k-spectrometer revealed a measured efficiency of $\eta_{0,exp}$=(97.2±0.5) percent, which is in good agreement with the predicted value. Good agreement between theoretical and experimental data exists also for the corresponding 28 GHz serpentine converter (a_0=20.0 mm).[7] The numerical optimization calculations showed that at the output of the advanced 140 GHz TE_{01}-to-TE_{11} converter (a_0=13.9 mm, length L=5.12 m !), the TE_{01} input mode is converted to the TE_{11} mode with an efficiency of η_0=95.0 percent (unwanted modes: 1.0 percent TE_{01}, 0.1 percent TE_{12} and 0.1 percent TE_{21}; ohmic attenuation: 3.8 percent).

TE_{11}-to-HE_{11} Mode Converter

Adiabatic TE_{11}-to-HE_{11} mode conversion is achieved in a straight corrugated waveguide section (a_0=20 mm at 28 GHz and a_0=13.9 mm at 70 GHz) in which the electrical depth of the annular slots gradually decreases from an initial value of almost $\lambda/2$ ($\lambda=2\pi/k$ is the free-space wavelength) to a final slot depth of approximately $\lambda/4$ (Figure 5). Unwanted mode conversion is associated primarily with the entirely cross-polarized EH_{12} mode.[8] Computer-aided optimization of converter length, shape of corrugations and slot depth profile is achieved with a scattering matrix formalism employing the modal field expansion technique: modular analysis concept (MAC).[10] At high k x a values, optimized non-linear profiles of mechanical slot depth are absolutely necessary for achieving a mode purity of almost 99 percent and cross-polar levels lower than –28 dB with converter lengths of 220 mm at 28 GHz and 370 mm at 70 GHz, respectively.[9] The calculated return losses are lower than –56 dB. The optimum slot depth profile is very close to that of constant coupling along the converter[4] and decreases most slowly at the converter input where the coupling constant for hopping to the TM_{11}-EH_{12} conversion branch is largest.[8] Low power insertion loss measurements at 28 and 70 GHz are in good agreement with the calculated efficiencies, and no return losses are detectable.[9] Figure 6 shows the intensity contours of the generated HE_{11} mode (70 GHz) radiated from an open-ended corrugated waveguide antenna (D=63.4 mm) and focused by an elliptical mirror (f=450 mm). The angle of incidence on the mirror was 64°. The HE_{11} mode produces

[Continued on page 112]

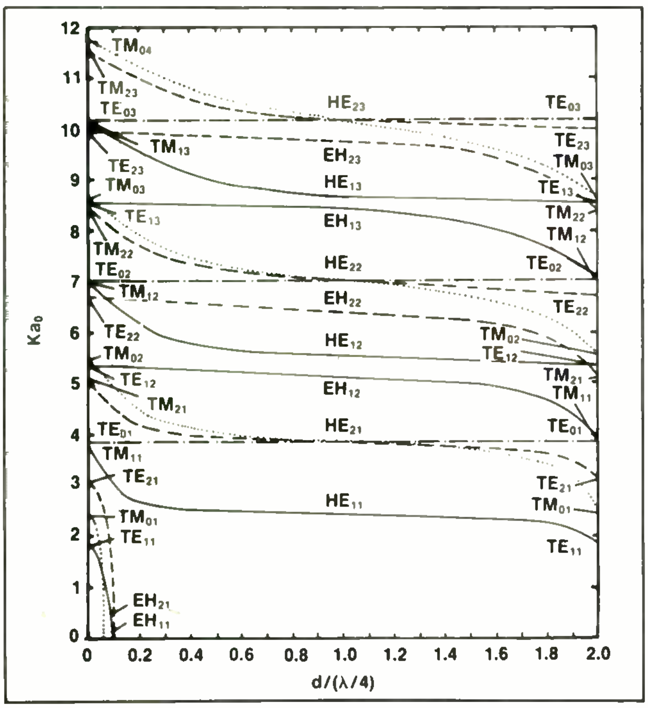


Fig. 5 Eigenvalues Ka_0 of the characteristic equation vs normalized electrical slot depth $d/(\lambda/4)$ for various modes in corrugated waveguide (a_0 = 13.9 mm, 70 GHz).

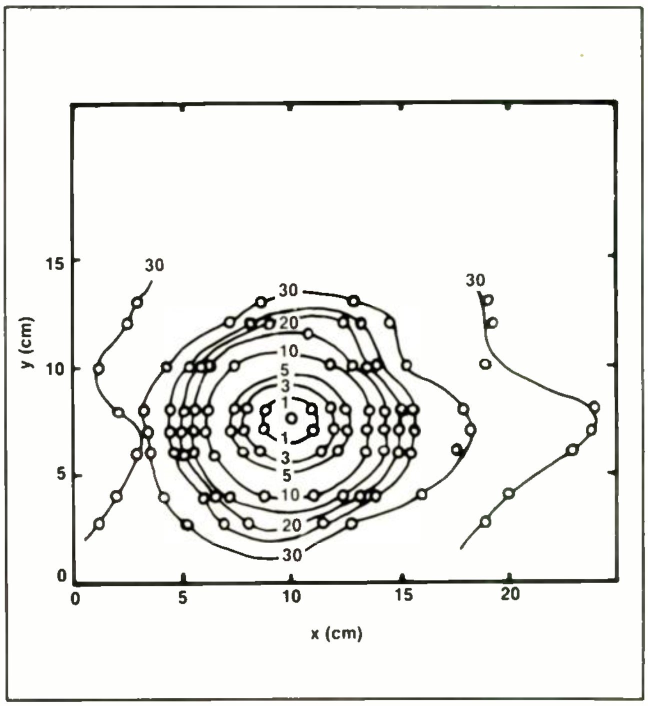
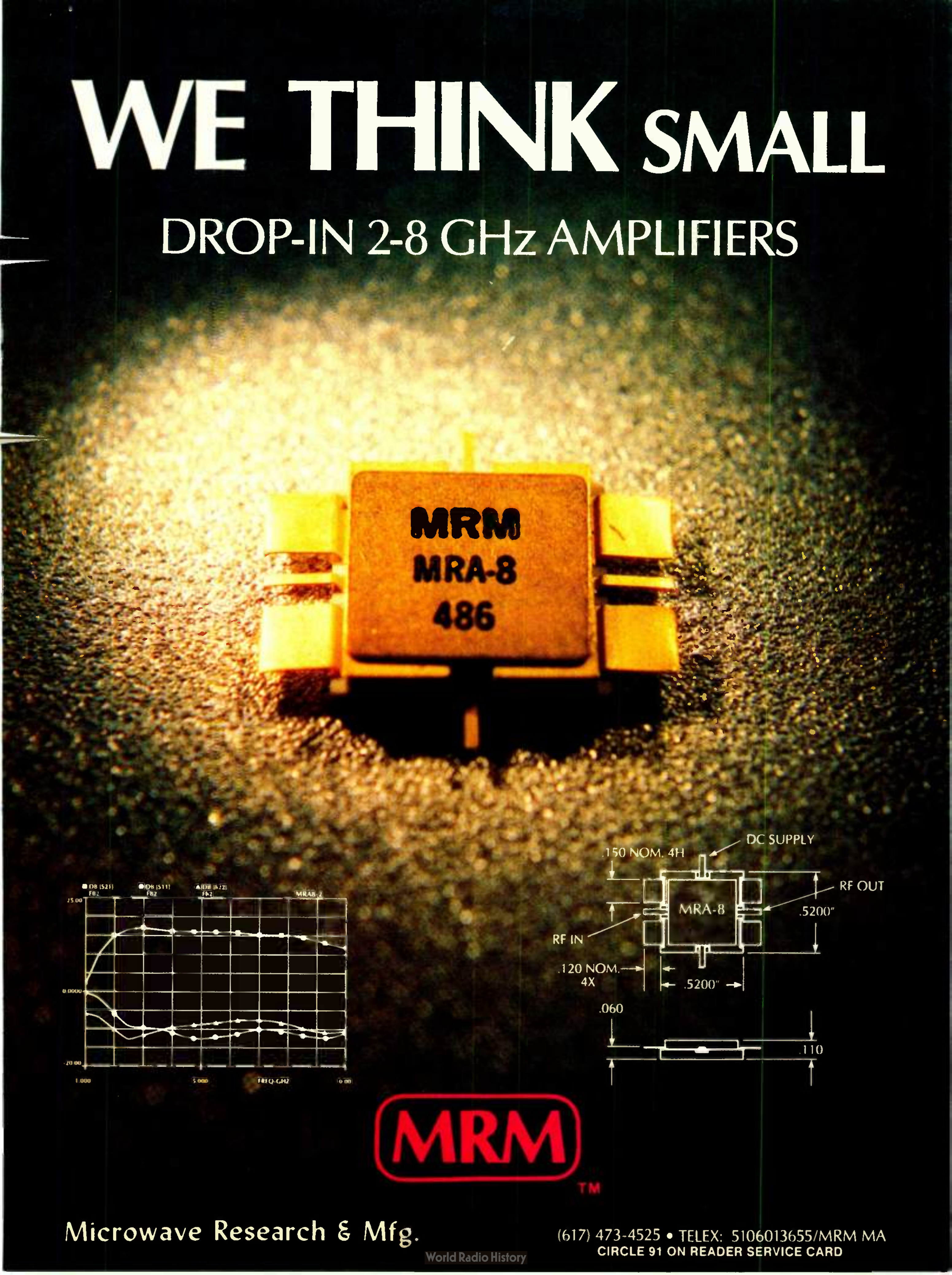


Fig. 6 Intensity contours of the generated HE_{11} mode (70 GHz, antenna aperture D = 63.4 mm) focused by an elliptical mirror (f = 450 mm). The E-vector is oriented along the x-axis.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High Power Mode Conversion Measurements

The mode conversion sequence TE_{02}-to-TE_{01}-to-TE_{11}-to-HE_{11} was measured at 28 and 70 GHz by thermographic mode-pattern analyzers inserted into the waveguide. Figure 7 shows the burn pattern measured at the ends of the various mode converters. It can be seen that the modes contain some small fractions of asymmetric spurious modes produced by the gyrotrons. At 70 GHz the near-field radiation pattern of the HE_{11} mode was measured by means of five pickup antennas at a distance of R=360 mm from an open-ended corrugated waveguide antenna (D=63.5 mm). Figure 8 shows the comparison of the measured HE_{11} mode H-plane pattern with a theoretical TEM_{00} mode power distribution; the agreement is very good.

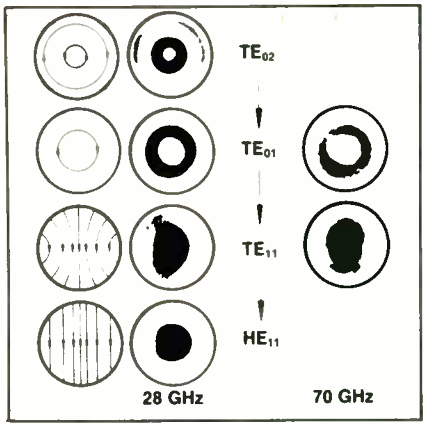


Fig. 7 Sequence of mode conversion TE_{02} to TE_{01} to TE_{11} to HE_{11} measured at high power (170 kW, 1 ms) at 28 and 70 GHz by thermographic mode pattern analyzers.

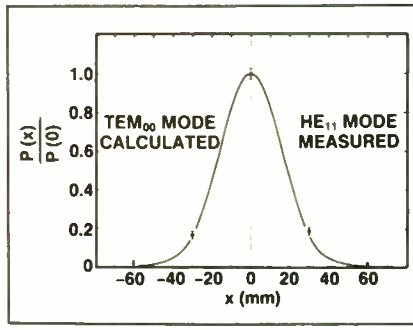


Fig. 8 High power measurement (170 kW) of the HE_{11} mode H-plane near-field pattern at 70 GHz and comparison with the theoretical TEM_{00} mode pattern (Gaussian mode). Aperture ϕ = 63.5 mm. Distance R = 360 mm.

TE_{01}-to-TM_{11} Mode Converter

In an alternative two-step mode conversion process, the HE_{11} mode is generated from TE_{01} using TM_{11} instead of TE_{11} as the intermediate polarized mode. First, a smooth circular waveguide is bent with an optimized curvature distribution at a proper angle to convert virtually all of the TE_{01} power to TM_{11}, which is polarized perpendicularly to the plane of bend. Power is continuously coupled from TE_{01} to TM_{11} along the converter because TE_{01} and TM_{11} are degenerate modes in smooth-walled circular waveguide (same propagation constants). A waveguide bend of angle

$$\theta_c = k_{01} \times \lambda / (2\sqrt{2} \times a_0) \qquad (3)$$

will convert all the power from one mode to the other[4]; k_{01} = 3.8317 is the first root of the Bessel-function $J_{0'} = -J_1$. The bend must be long enough to avoid unintentional mode conversion to unwanted modes, but ought to be as short as possible to minimize ohmic (wall) losses from the high loss TM_{11} mode. Bent 140 GHz TE_{01}-to-TM_{11} mode converters with a_0 = 13.9 mm were optimized by means of numerical integration of the coupled-mode equations for the six coupled modes: TE_{01}, TM_{11}, TE_{11}, TE_{12}, TE_{21} and TM_{21}.[16]

Ellipticity coupling is included in the coupling matrices. Due to the multi-mode coupling, the conversion angle θ_c depends on the curvature distribution and differs by a small amount from the value calculated with Equation 3. Lowest spurious mode levels, together with highest power transmission efficiency (shortest arc length: L_c=2.52 m), were achieved using sinusoidal curvature distribution instead of constant curvature as used in reference 4. The optimum 140 GHz TE_{01}-to-TM_{11} mode converter (computed conversion efficiency η_0=95.2 percent) was fabricated by bending commercially available C76 waveguide (a_0=13.9 mm) in a preassembled frame with a shape defining the curve. According to the multimode calculations, the influence of the measured ellipticity (≤0.02 mm) on the conversion efficiency turned out to be negligible. Figure 9 shows a photograph of the converter. The measured 140 GHz TM_{11}-mode pattern (E-plane) is plotted on a linear scale in Figure 10. This far-field pattern measurement, together with insertion loss measurements through two back-to-back mode converters, delivered an experimental conversion efficiency of (95.0±1.0) percent, in good agreement with the computed value.

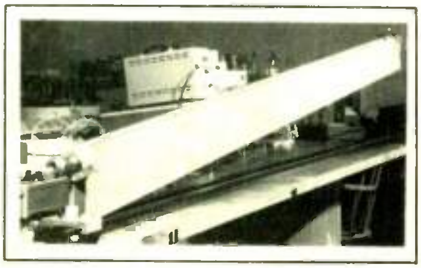


Fig. 9 The 140 GHz TE_{01}-to-TM_{11} mode converter (a_0 = 13.9 mm).

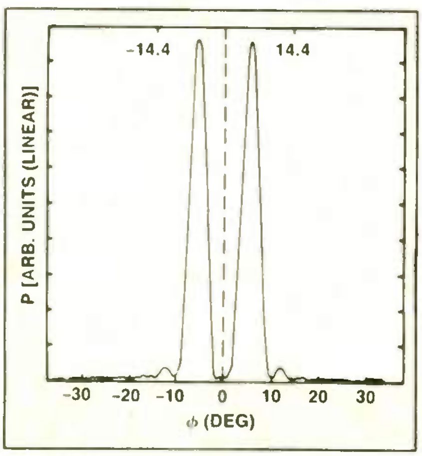


Fig. 10 Measured E-plane far-field pattern of the generated TM_{11} mode at 140 GHz. Aperture φ = 27.8 mm.

TM_{11}-to-HE_{11} Mode Converter

Adiabatic TM_{11}-to-HE_{11} conversion is obtained in a straight, circumferentially corrugated waveguide section by non-linearly tapering the corrugation depth from 0 to nominally λ/4 (see Figure 5). This transducer has to be sufficiently long (L=0.84 m for a_0=13.9 mm at 140 GHz; slot width = tooth width = λ/6) in order to suppress mode hopping to the TE_{11}-EH_{11} sur-

DATA OF DIAMETER TAPERS FROM D = 63.4 - 40.0 mm AT 28 GHz AND D = 63.4 - 27.8 mm AT 70 AND 140 GHz						
Progagating Mode	Inner Waveguide Surface	Main Coupled Modes	Frequency (GHz)	Length (mm)	Suppression of Spurious Modes (dB) calc.	meas.
TE_{01}	smooth	TE_{02}	28	310	40	(37±4)
TE_{01}	smooth	TE_{02}	70	705	40	(36±5)
TE_{01}	smooth	TE_{02}	140	1410	40	
TE_{02}	smooth	TE_{01}, TE_{03}	28	310	37	≥25
TE_{02}	smooth	TE_{01}, TE_{03}	70	705	37	≥25
TE_{03}	smooth	TE_{02}, TE_{04}	140	920	37	
TE_{11}	smooth	TM_{11}, TE_{12}	28	510	25	(24±2)
TE_{11}	smooth	TM_{11}, TE_{12}	70	1095	25	(23±2)
HE_{11}	corrugated	HE_{12}, EH_{12}	28	310	30	≥25
HE_{11}	corrugated	HE_{12}, EH_{12}	70	700	30	≥25
HE_{11}	corrugated	HE_{02}, EH_{12}	140	1400	30	

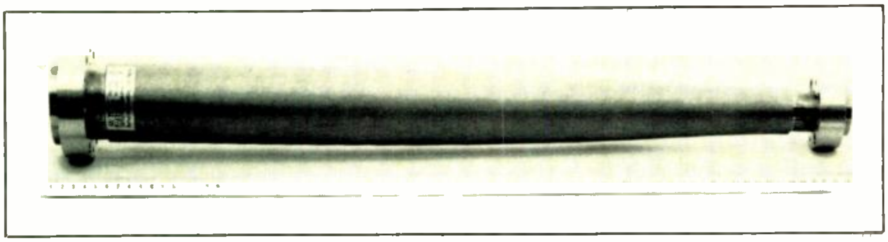


Fig. 11 Electroformed 70 GHz TE_{02}-waveguide diameter taper.

face mode branch, as well as the TE_{12}-EH_{12} mode branch.[4,8] As in the case of the TE_{11}-to-HE_{11} mode converter, computer-aided optimizations of converter length, shape of corrugations and slot depth profile along the converter also have been studied with the scattering matrix code (MAC). The theoretical conversion efficiency is around 98.5 percent. The slot depth increases most slowly at the converter input, where coupling to both unwanted mode branches is strongest.

Waveguide Diameter Taper

Tapered transitions are needed for the interconnection of oversized waveguides with different cross-sections. Waveguide diameter tapers should not generate spurious modes above a certain acceptable level. Very low unintentional mode conversion from the propagating mode to unwanted modes is obtained by employing relatively short nonlinear tapers with a gradual change of the cone angle. The numerically calculated optimum taper contours with approximate Chebyshev mode conversion response depend on the propagating mode of interest and are described by a Fourier series.[17] These synthesized contours are analyzed by numerical integration of the corresponding coupled-wave equations. The properties of several tapers developed for ECRH transmission lines are summarized in Table 2. The optimized tapers have been fabricated by electroforming methods. Figure 10 shows a photograph of the 70 GHz TE_{02}-taper.

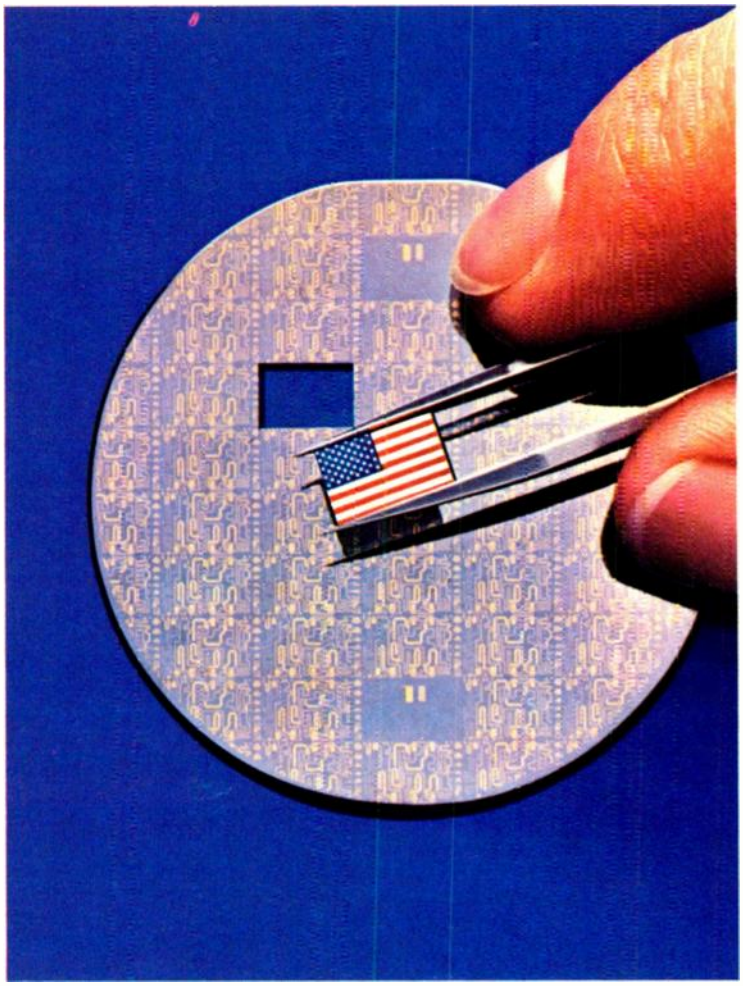
Corrugated Waveguide Bends

Unintentional mode conversion losses in circumferentially corrugated TE_{0n} bends scale with

$$a_0^3 f^2 / k_{0n}.$$

Therefore, bends can be made shorter if they propagate higher TE_{0n} modes; only gradual bends for TE_{0n} modes with $n \geq 3$ seem to be feasible at 140 GHz. In corrugated bends, a pure propagating mode

[Continued on page 116]



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DATA OF VARIOUS CORRUGATED WAVEGUIDE BENDS WITH INNER DIAMETER D=63.4 mm AT 28 GHz AND D=27.8 mm AT 70 AND 140 GHz						
Propag. Mode	Bend Angle	Frequency (GHz)	Arc Length (m)	Curvature Distrib.	Coupled Modes 1st order	2nd order
TE_{01}	90°	70	2.4	triangular	HE_{11}, EH_{12}	HE_{21}
TE_{02}	90°	28	3.3	sin	HE_{12}, EH_{12}, EH_{13}	HE_{22}, EH_{22}
TE_{02}	90°	70	1.74	sin	HE_{12}, EH_{12}, EH_{13}	HE_{22}, EH_{22}
TE_{03}	90°	140	3.5	sin	HE_{13}, EH_{13}, EH_{14}	HE_{23}, EH_{23}
HE_{11}	47°	70	1.5	sin^2	HE_{21}, TE_{01}	HE_{12}, EH_{12}, HE_{31}

can be made the dominant one by proper choice of the surface reactance. In the case of TE_{0n} bends, a matched corrugation removes the degeneracy of the mode pairs TE_{0n}/TM_{1n} (as in smooth-wall waveguides), and it prevents a degeneracy between the hybrid HE_{2n} and the working TE_{0n} mode at corrugation depths of about $\lambda/4$ (Figure 5).

The properties of the actual bends are optimized using numerical integration of the corresponding coupled-mode equations. Optimum electrical slot depths appear at values of $(0.2\text{-}0.4)\times\lambda/4$ for TE_{0n} bends and at $1\times\lambda/4$ for HE_{11} bends. Unwanted conversion to spurious modes is reduced by changing the curvature continuously instead of abruptly. The data of various corrugated waveguide bends with spurious mode level at the output of approximately -20 dB are given in Table 3. Figure 12 shows a photograph of the 70 GHz TE_{02}-mode bend (90°, electroformed corrugated Cu-waveguide). In the case of an E-plane bend propagating the HE_{11} mode, TM_{02} replaces coupling to TE_{01}.[4] However, optimum H-plane bends also can be used as E-plane bends because the curvature coupling constants $C(HE_{11}/TE_{01})$ and $C(HE_{11}/TM_{02})$ are identical. 140 GHz TE_{01} and HE_{11} bends with low mode conversion can be realized as quasi-optical mitre bends (D=63.4 mm). In this case, the experimentally observed spurious mode levels also are around -20 dB.

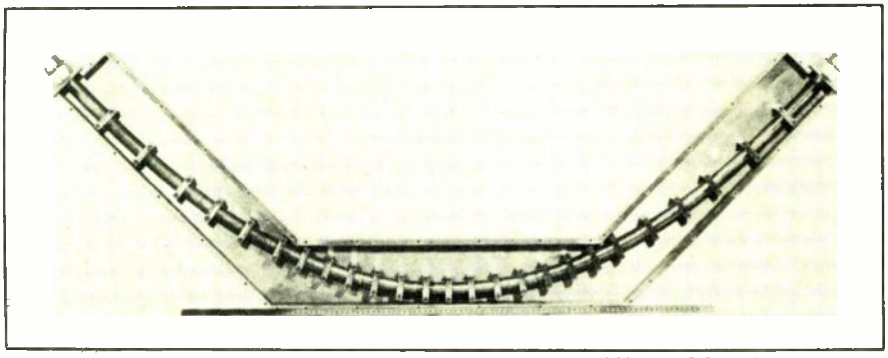

Fig. 12 The corrugated 70 GHz TE_{02}-mode bend (90°, a_0=13.9 mm).

MEASURED ATTENUATION OF VARIOUS MODES IN CORRUGATED WALL-MODE FILTERS					
Freq. (GHz)	Diam. (mm)	Attenuation (dB/m) TE_{11}	TE_{12}	TM_{11}	TE_{01}
28	63.4	8.5	4.0	11.0	≤0.03
70	27.8	7.0		8.5	≤0.04

Mode-Selective Corrugated Wall Filters

Full power gyrotron operation is affected by even small power reflections from down-tapers and from non-ideal loads. Moreover, unwanted asymmetric modes lead to trapped mode resonances between inhomogeneities in the waveguide. Therefore, selective suppression of these modes with mode filters is necessary to decouple the different waveguide sections and to avoid arcing. Optimum mode filtering results are achieved with circumferentially corrugated stainless steel filters with tapered slot depth. The filters convert non-circular modes into the corresponding EH-surface waves (step-by-step conversion, e.g.

$$TE_{13} \rightarrow TM_{12} \rightarrow TE_{12} \rightarrow TM_{11} \rightarrow TE_{11} \rightarrow EH_{11}).$$

These modes are highly damped by the special wall structure, while attenuation of the transmitted TE_{01} (or other TE_{0n}) mode is negligible.

Computer-aided minimization of input reflection and optimization of total length, shape of corrugations and slot depth variation (anisotropic surface reactance) is achieved with the scattering matrix code. Proper length and appropriate placement of phase-delay elements of straight smooth-walled waveguide between the different filter sections also is studied by the use of the MAC code. The attenuation of unwanted asymmetric modes in such optimized corrugated wall-mode filters is considerably higher than in resistive wall- or helix-mode filters. The measured attenuation of various modes in numerically optimized corrugated wall-mode filters is given in Table 4. Measured and theo-

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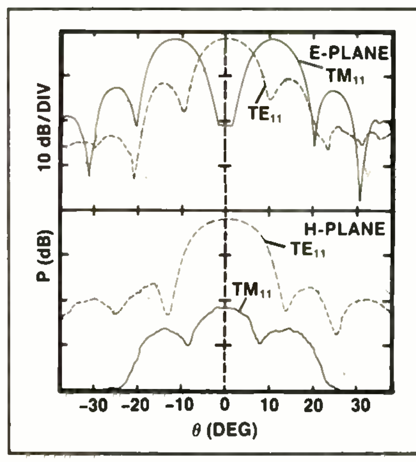

Fig. 13 TM_{11}-to-TE_{11} conversion and attenuation (dashed curve) of a TM_{11} wave (solid curve) at 70 GHz by a corrugated wall mode filter (L = 0.5 m). A second filter section attenuates TE_{11} (dotted) Aperture ϕ = 27.8 mm.

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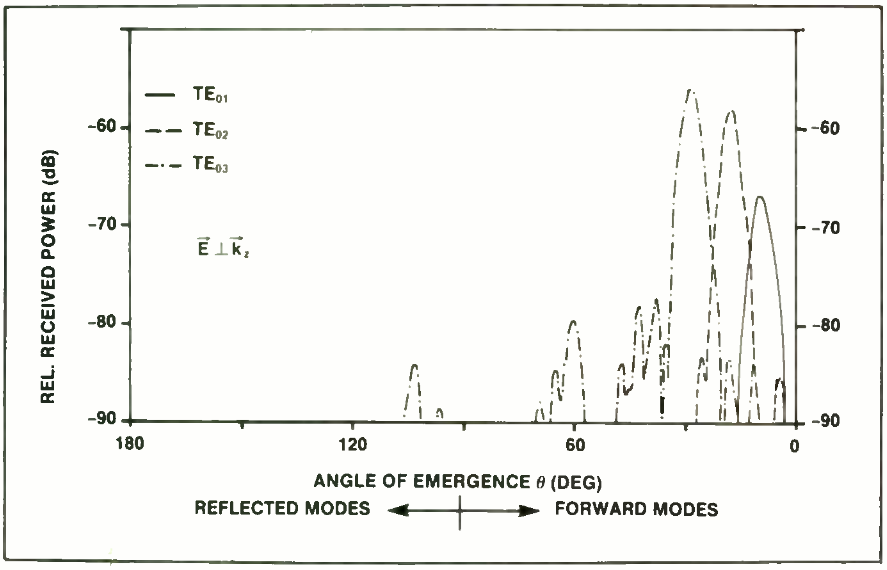

Fig. 14 Spectra of TE_{01}, TE_{02} and TE_{03} modes measured with the k-spectrometer at 70 GHz (low power measurements).

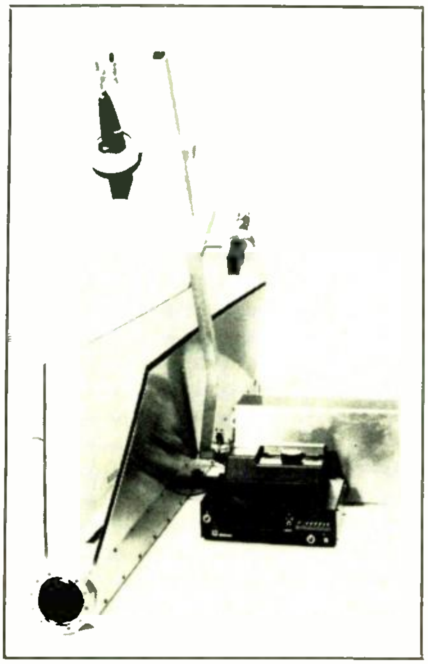

Fig. 15 A 70 GHz k-spectrometer (a_0 = 31.7 mm).

retical return losses are lower than −12 dB. The attenuation of the working mode TE_{01} is in all cases negligible. The mode conversion and attenuation performance of the 70 GHz filters is shown in Figure 13 (far-field measurements).

Wavenumber Spectrometer

This device allows mode analysis of mm-waves in oversized waveguides by measuring the spectrum of axial wavenumbers. In circular waveguides, the propagation constants of TM and TE modes are given by

$$k_z = \omega/c \times \sqrt{1 - (c\nu_{mn}/\omega a_0)^2}$$

with $J_m(_{mn})=0$ for TM_{mn} waves and $J'_m(\nu_{mn})=0$ for TE_{mn} modes, where $\nu=\omega/2\pi$ is the frequency of the wave, c is the velocity of light and J_m are the Bessel functions. The wavenumber spectrometer operates as a leaky-wave antenna in the wall of the oversized waveguide. If a mm-wave with wavenumber k_z propagates in the waveguide, a small amount of power is coupled out by the antenna structure into the free space. The resulting wave fronts of the radiated field are cones with an angle of emergence θ given by $\cos\theta=k_z/k$, where $k=\omega/c$ is the wavenumber in free space.

To determine the wavenumber and direction of a propagating mode, the direction of the main lobe of the radiated field has to be measured with respect to the waveguide axis. If the frequency is known a priori or measured, the Bessel-zero, which characterizes the mode, can then be calculated from the Equation given above. In a first experimental device (a_0= 13.9 mm), the leaky-wave antenna structure consists of a linear array of 230 holes at a distance of 2 mm in the waveguide wall. The diameter of the holes is tapered to achieve a Gaussian aperture distribution. This leads to a radiated beam with low sidelobes and plane phase fronts, thus allowing the measurement of the angle θ in the near field. The receiving antenna, an optimized non-linear horn with an angular resolution (FWHM) of 5° and 28.4 dB gain, can be swept around the leaky-wave structure on an arm with the axis perpendicular to the waveguide by means of a motor-drive.

To avoid stray radiation, the system is shielded with two Eccosorb-covered plates. Low power and high power tests were performed with TE and TM modes at 70 GHz. As an example, Figure 14 shows measured spectra of TE_{01}, TE_{02} and TE_{03} modes generated with rippled-wall

[Continued on page 120]

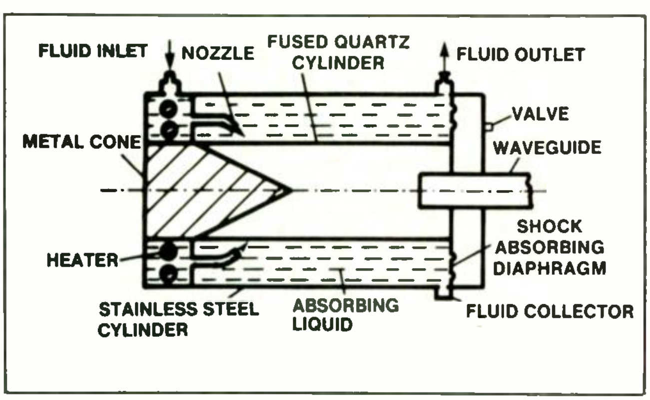

Fig. 16 Cross-sectional view.

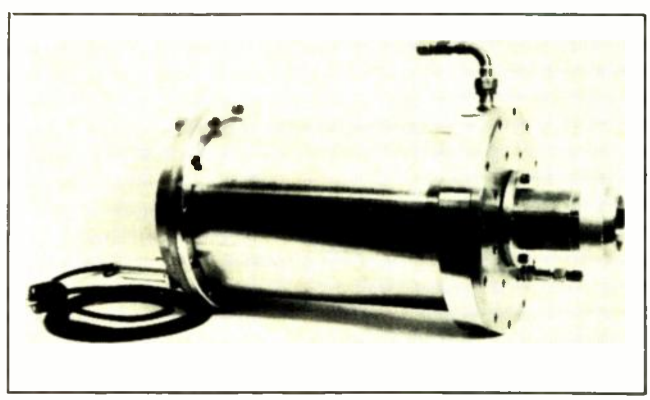

Fig. 17 The TE_{0n} mode calorimeter.

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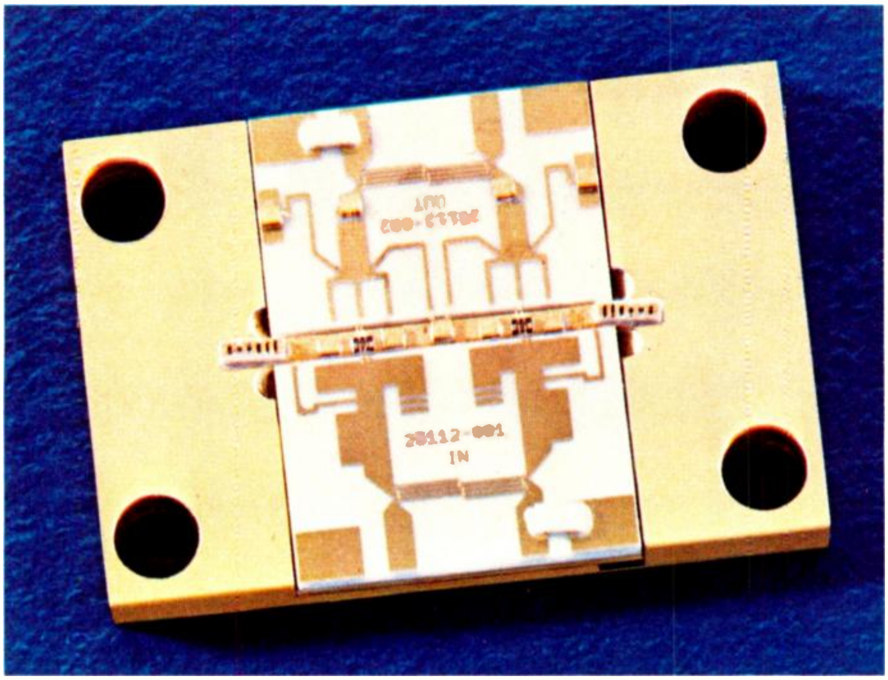

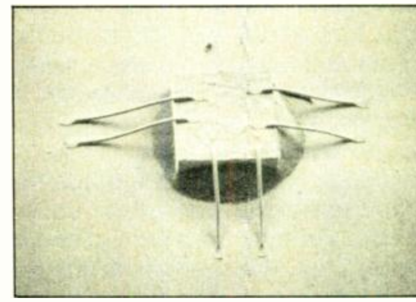

Fig. 1: SEM photomicrograph of an SLC bonded to a 1-in. square alumina substrate with six gold wires (45X).

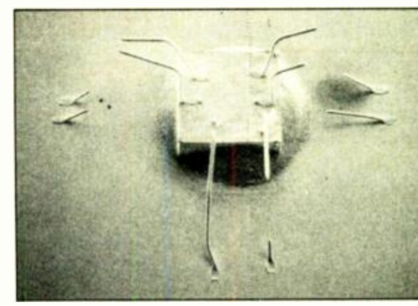

Fig. 2: SEM photomicrograph of the SLC after bond-pull strength testing.

The Effect of Termination on Single-Layer Capacitors Interconnected to MICs.

In microwave integrated circuit assemblies, passive components such as single-layer ceramic capacitors (SLCs) are generally wire-bonded to another circuit element. The reliability of such bonding depends not only on the integrity of the bond, but also on the reliability of the metallized terminations of the SLC.

SLC terminations usually consist of layers of thin or thick film metals. The base layer is chosen for its adhesion to both the ceramic body and the outer layer; the outer is usually pure gold. Although the terminations are generally plated, AVX PathGuard™ SLCs are metallized by sputtering.

Test Method

Tests to determine whether the method of termination affects overall bond stength were conducted by mounting six capacitors with either sputtered or plated metallizations on a 1-inch square alumina substrate. The terminations were titanium-tungsten-gold and varied in thickness between 50 and 150 microinches, while the gold on the substrate was kept at 150 μin. Each SLC was bonded six times with 0.7-mil diameter gold wire using a wedge bonder under consistent compression and stage and bonding tool temperature conditions.

Table 1:
Bond-Pull Strength in Grams (± δ)

Sputtered		Plated	
Type I	3.6 ± .5	Type A	2.3 ± .4
Type II	3.5 ± .5	Type B	2.1 ± .4
Type III	3.5 ± .5	Type C	2.2 ± .4
Type IV	3.1 ± .8	Type D	2.2 ± .5
		Type E	2.2 ± .7

δ: One Standard Deviation

Test Results

Average bond pull strength was derived from 36 data points for each thickness and metallization type. The results, shown in the Table, demonstrate that bond pull strength is significantly higher for SLCs with sputtered terminations—about 3.5 grams compared with 2.2 g—for metallizations between 60 and 120 μin. Above 150 μin. excessive termination thickness can reduce bond reliability In similar tests of gold ribbon weldability, the frequency of metallization lifting from the SLC body increased above 150 μin. for both types of termination.

Discussion

Sputtered metallizations show greater adhesion characteristics than do plated types, and they have a broader process window for welding applications. Data indicated that the thickness of the metallization is also an important factor, with the recommended value between 80 and 100 μin.

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mode converters. The spectra confirm the calculated low sidelobe level of the device and they demonstrate the high purity of these modes. During the high power experiments, the k-spectrometer was equipped with two horns, one serving as power monitor for the TE_{01} mode in forward direction, the other detecting reflected modes. Figure 15 shows a photograph of this device. The power signal was fairly reliable and was not influenced by the reflected modes. The backward signal was used to switch off the gyrotron in the case of strong reflections. Measurements with the k-spectrometer and with stacked burn pattern analyzers revealed some spurious TE_{13} mode content at the gyrotron output.

Matched High Power Loads

For high power gyrotron and waveguide component tests and for absolute calibration of the mm-wave power, two types of matched loads were built. These loads substitute for the customary water loads whose applications are limited due to the very short attenuation length for millimeter waves in water ($\approx$0.15 mm at 70 GHz). One type of load, with fire clay as absorbing medium (attenuation length at 70 GHz $\approx$ 15 mm), is used as a simple dummy load. In combination with the k-spectrometer, it forms a device for convenient and quick adjustment of operational parameters of gyrotrons with respect to good mode purity and maximum output power. These loads dissipate 200 kW/100 ms pulses at full gyrotron duty cycle (1 percent) with negligible reflections. Two versions of these loads were fabricated, one for the axisymmetric TE_{0n} modes and another for linearly polarized TE_{11} and HE_{11} modes. The other type of load, in which octanol, a monovalent alcohol, serves as a fluid absorbing and cooling medium (attenuation length at 70 GHz $\approx$ 10 mm), is used as a calorimetric load for exact high power measurement (low flow rate of $\approx$ 15 l/min. without any blistering even for 100 ms pulses). The principle of construction for both types of loads is demonstrated in Figure 16, wnich gives a cross-sectional view of the calorimetric load.

Figure 17 is a photograph of the load. The mm-wave power enters by an open-ended waveguide of 63.4 or 27.8 mm diameter, respectively. The power is reflected by a circular symmetric metallic cone into a cylinder that contains the absorbing medium. In the fire-clay load, the absorbing material is cooled by an air stream. In the calorimetric load, the liquid absorber is behind a fused-quartz cylinder. The fluid enters the load via a chamber where a heater of 1 kW power permits calibration of the load. The loads are slightly modified for measurements of linearly polarized modes (TE_{11} or HE_{11}). The metal cone is replaced by a tilted tile of fire clay in the dummy loads and by a tilted slab of an appropriate dielectric material, which separates the absorbing fluid from the waveguide interior, in the case of the calorimeter. Tile and slab are mounted at the Brewster's angle (fire clay: 65°, Al_2O_3: 72°, Teflon: 54°) where no reflections occur for matched linearly polarized radiation. ■

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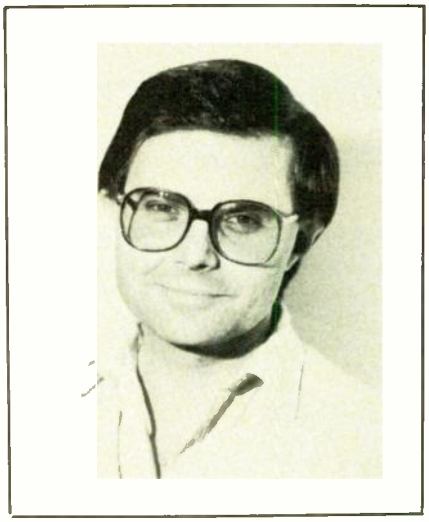
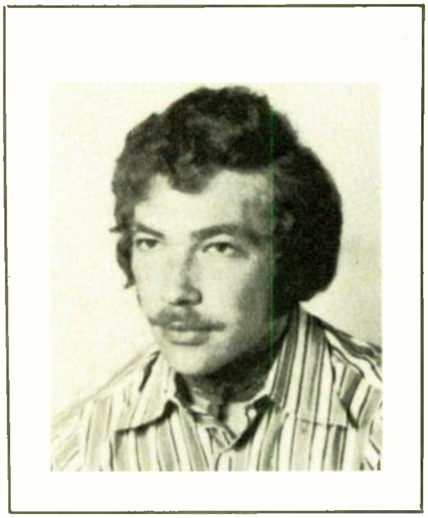


Manfred Thumm *was born in Magdeburg, Germany on August 5, 1943. He received the Dipl Phys and Dr rer nat degrees in physics from the University of Tubingen, Germany in 1972 and 1976, respectively. At Tubingen he was involved in the investigation of spin-dependent nuclear forces in inelastic neutron-nucleus scattering. From 1972 to 1975 he was a doctoral fellow of the "Studienstiftung des deutschen Volkes."*

In 1976 he joined the Institut fur Plasmaforschung (electrical engineering department) of the University of Stuttgart, Germany where he worked from 1976 to 1982 on confinement and RF heating of toroidal high-beta pinch plasmas for thermonuclear fusion research. Since 1982 his research activities have been mainly in the areas of development of components for transmission of very high power millimeter waves through overmoded waveguides and of antenna structures for RF-plasma heating with microwaves. Thumm is a member of the German Physical Society.



Volker Erckmann *was born in Brettheim, Germany on January 25, 1950. He studied physics at the University of Stuttgart, Germany, the University of Graz, Austria and the Technical University of Munich, Germany, where he received his Dipl Ing degree. In 1983 he got the Dr Ing degree at the University of Stuttgart. There, he worked on mag-*

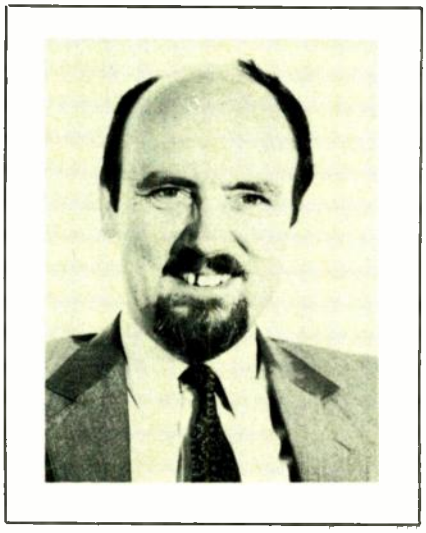
neto-acoustic heating and flux-conserving equilibria in a belt pinch.

In 1982 he joined the Max-Planck-Institut fur Plasmaphysik in Garching, near Munich, as a member of the stellarator group. He concentrated his research activities on electron cyclotron heating of stellarators in the framework of controlled thermonuclear fusion research.

Walter Kasparek *was born in Eitorf, Germany on July 22, 1952. He received the diploma in physics and the Dr Ing degree in electrical engineering from the University of Stuttgart in 1979 and 1984, respectively. There, at the Institut fur Plasmaforschung, he was involved in the development of infrared laser scattering techniques to determine ion temperatures in plasmas.*

Since 1984, he has worked in the field of electron cyclotron heating of fusion plasmas. His main activities are the development of measurement components and quasi-optical techniques for the transmission of high-power millimeter waves.

Helga Kumric' *was born in Opatija, Yugoslavia on October 1, 1959. She received her Dipl Ing degree in electrical engineering for development of microstrip lines at the University of Zagreb, Yugoslavia in 1982. Since 1984, she has been developing components for transmission of very high power millimeter waves through overmoded waveguides for RF-plasma heating with microwaves at the Institut fur Plasmaforschung of the University of Stuttgart.*



Gunther A. Muller *was born in Berlin, Germany on December 19, 1938. He received the Dipl Phys degree in 1966 from the University of Stuttgart, Germany. In 1967 he joined the Institut fur Plasmaforschung of the University of Stuttgart.*

His first research activities concerned experiments with magneto-hydrodynamic waves for diagnostics and heating of plasmas. In 1973 he received the Dr Ing degree. From 1974 to 1982 he worked on fusion plasma experiments such as theta pinch and belt pinch.

Since 1982 his main research activities have been in the field of plasma heating with microwaves. He is working on the development of microwave instrumentation and electrical systems for gyrotrons.

Paul-G. Schuller *obtained the Dipl Phys degree from the University of Karlsruhe, Germany, in 1964. He then joined the Institut fur Plasmaforschung of the University of Stuttgart, Germany, where he worked in the field of plasma magnetic wave propagation and its application to plasma heating. In 1974 he received the Dr Ing degree from the University of Stuttgart for his thesis on Alfven wave propagation in inhomogeneous magnetic fields.*

He is presently mainly engaged in investigations of plasma heating by high power microwaves and development of instrumentation for measurement of high microwave power.

Rolf Wilhelm *was born in Bremen, Germany on February 21, 1939. He received the Dipl Phys degree in 1963 at the Techn. University of Hannover, Germany and the Dr rer nat degree in 1968 at the Techn. University of Munich, both in the field of plasma physics. From 1965 to 1979 he joined the Max-Planck-Institut fur Plasmaphysik at Garching near Munich. Besides his thesis work there, he was mainly engaged in electrical and RF technical problems of plasma production and heating. After earning his degree, he became group leader and, since 1974, project leader of a larger fusion project, again dealing with problems of high power RF heating of plasmas.*

In 1979 he became a full professor at the Institut fur Plasmaforschung at the University of Stuttgart. There, he and a small research group started a program for generation, transmission and radiation of high power millimeter waves in cooperation with the IPP Garching (1980 until now).

Wilhelm is a member of the German Physical Society.

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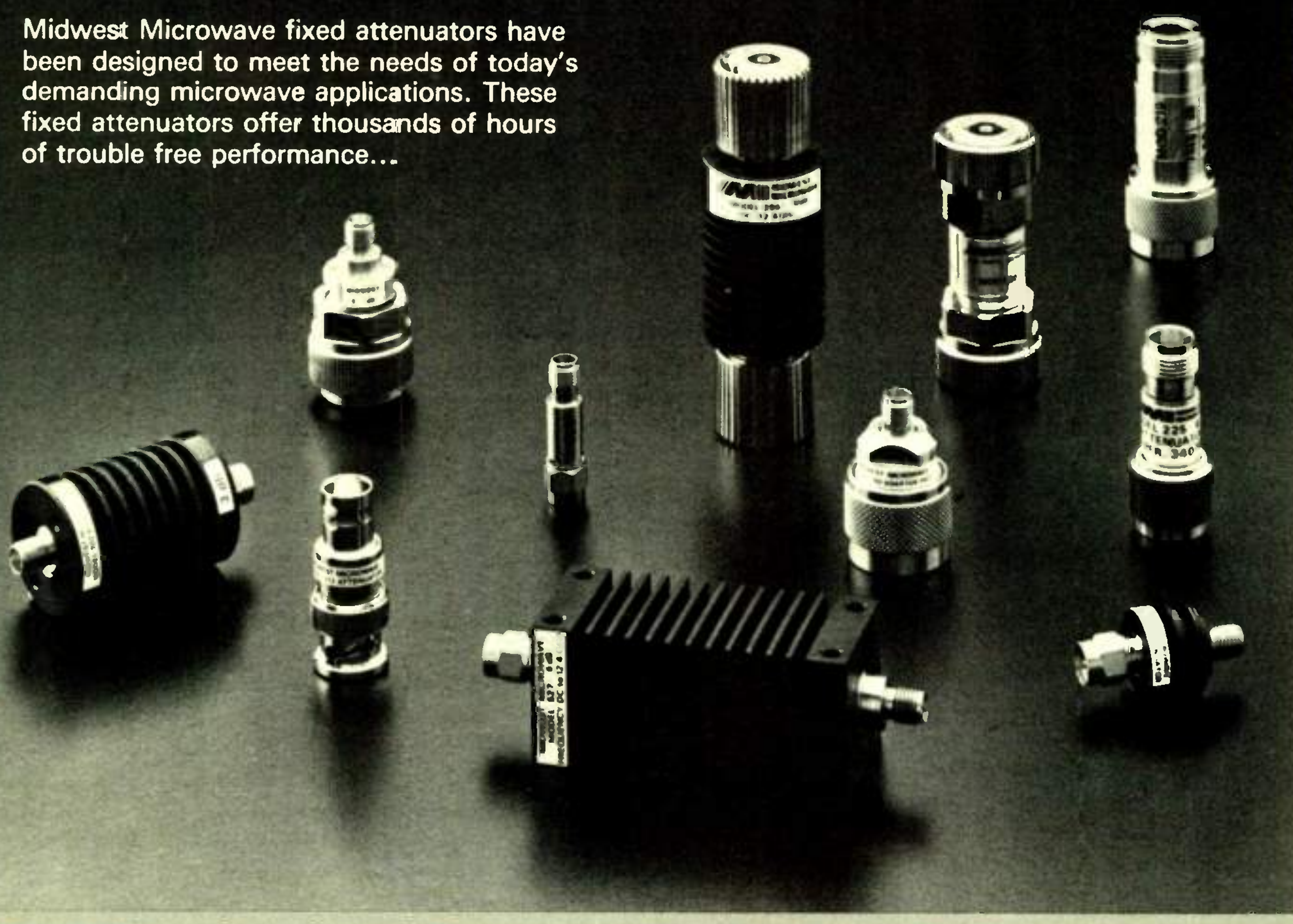

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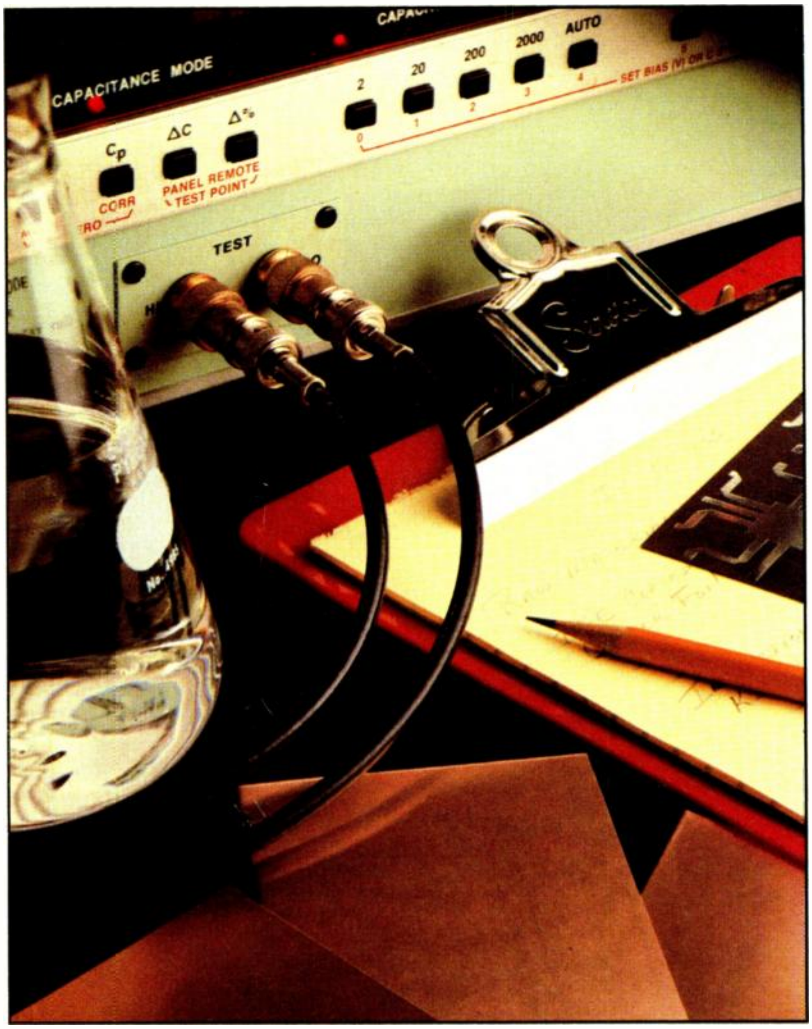

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Simplified Expressions for Calculating the Impedance of Shielded Slab Line of Any Shape Ratio

Stewart M. Perlow

RCA Microwave Technology Center
Princeton, NJ

Simple formulas are presented for calculating the characteristic impedance of strip line or slab line. The formulas are valid for any rectangular cross-section of the inner conductor, including the degenerate cases of infinitely thin horizontal or vertical strips. The method is especially useful for hand-held calculators due to the simplicity of calculation. It also is appropriate for CAD programs since one set of equations covers all possible inner conductor shape ratios.

Introduction

The shielded slab transmission line has been used for many years. The general structural form consists of a rectangular inner conductor sandwiched between two parallel plane outer conductors. The name stripline usually is applied to this type of line when the thickness of the rectangular inner conductor becomes small.

Bates calculated the impedance of the shielded slab line exactly, in terms of elliptic function.[1] The solution of these elliptic functions is not a trivial task, and this method is not particularly useful in CAD programs. Several authors have derived approximate expressions for these elliptic functions. The expressions are easy to evaluate and give excellent results but are restricted to certain conductor width and height ranges.[2,3,6] These restrictions do not affect most of the practical physical realizations of slab and strip lines. However, it is useful to have expressions or techniques available which have as few restrictions as possible for use in CAD type programs.

The technique presented here is unrestricted. The equations maintain their validity for any rectangular cross-section including the two degenerate cases of a thin, vertical strip or a thin, horizontal strip between two parallel ground planes. In addition, the evaluation of the expressions is so simple that it can be accomplished with a hand-held calculator. This technique consists of calculating the static capacitance per unit length for each surface of the rectangular inner conductor, as if the surfaces were infinitely thin strips, and correcting for the excess or overlapping fringing capacitances at the corners.

Calculation of the Characteristic Impedance

The characteristic impedance of a lossless uniform TEM mode transmission line is related to its static

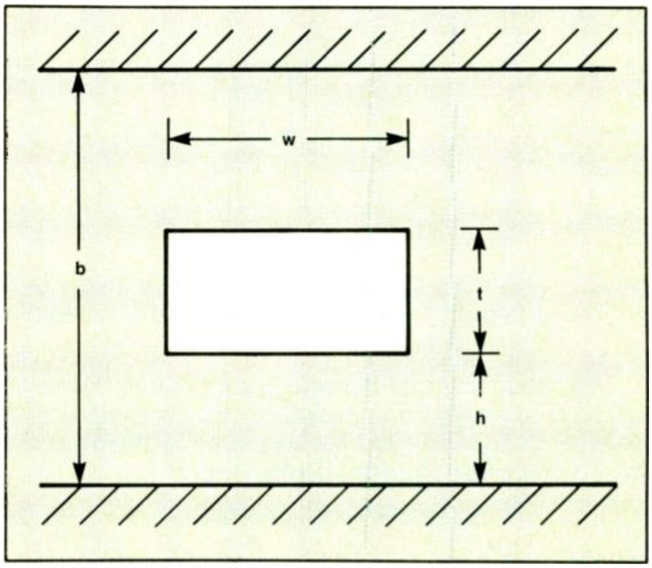

Fig. 1 Shielded slab line cross-section.

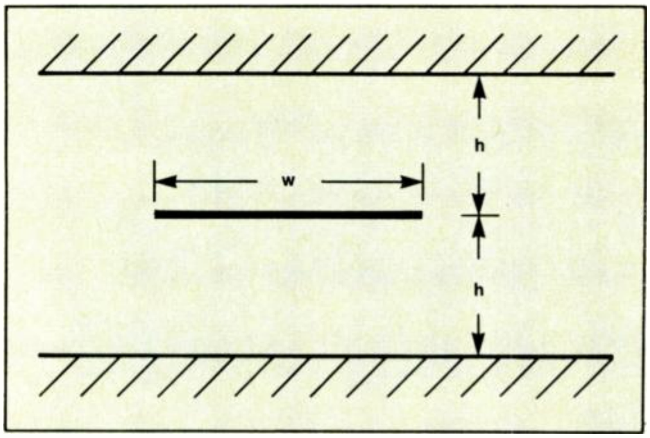

Fig. 2 Cross-section of slab line with infinitely thin horizontal center conductor.

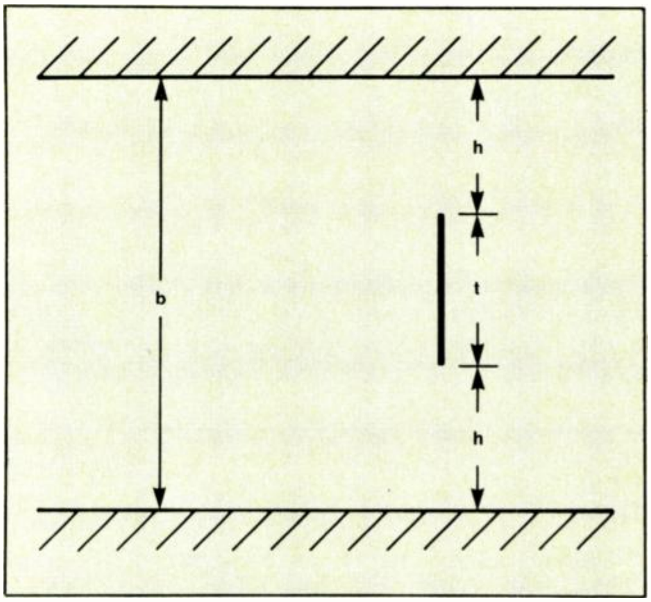

Fig. 3 Cross-section of slab line with infinitely thin vertical center conductor.

capacitance per unit length by

$$Z_0 = \frac{376.7}{\sqrt{\epsilon_r}\left(\dfrac{C}{\epsilon}\right)} \qquad (1)$$

where

ϵ_r is the relative dielectric constant of the medium and

C/ϵ is the ratio of the static capacitance per unit length to the permittivity of the dielectric medium.

The capacitance of the slab line shown in Figure 1, which is of width w and thickness t and is sandwiched between two parallel planes separated by the distance b, is composed of three capacitances,

$$C/\epsilon = CH/\epsilon + Cv/\epsilon - \Delta C/\epsilon \qquad (2)$$

where

Ch/ϵ is the capacitance of a thin horizontal strip of width w sandwiched midway between two parallel planes which are separated by the distance 2h;

Cv/ϵ is the capacitance of a thin vertical strip of height t sandwiched midway between two parallel planes which are separated by the distance b ($b = 2h + t$); and

$\Delta C/\epsilon$ is the overlapping fringing capacitance at the four corners of the slabline center conductor.

Determination of the Component Capacitances

Any of the many known characteristic impedance calculations for thin strips (t=0) can be used to obtain Ch/ϵ and Cv/ϵ. Hilberg presents solutions for both of these transmission line types in accurate, compact and easy-to-use forms.[4,5] The following relationships have been obtained using his solutions.

Thin Horizontal Strip

The normalized capacitance per unit length, Ch/ϵ for a thin horizontal strip as shown in Figure 2 is given as:

$$Ch/\epsilon = \frac{4}{.4413 + 2/\pi \, \ln[\coth(\pi/8 \; w/h)]}, \quad w/h < 1.1174$$

$$= 2\frac{w}{h} + 4(.4413), \quad w/h \geq 1.1174. \qquad (3)$$

Note that for $w/h > 1.1174$, Ch/ϵ has two components that are easily identifiable as the parallel-plate capacitance, $2w/h$, and four corner fringing capacitances, each of $2/\pi \; \ln 2 = .4413$ which is identical to Cohn's expression for wide thin strips.[6]

Thin Vertical Strip

The normalized capacitance of the thin vertical strip shown in Figure 3 of height h sandwiched midway between two parallel plates, separated a distance b, is given as:

$$Cv/\epsilon = \frac{4}{.4413 + 2/\pi \, \ln[\cot(\pi/4 \; t/b)]}, \quad t/b \leq 0.5,$$

$$= 4(.4413) + \frac{8}{\pi} \ln[\cot(\pi/4 \,(1-t/b))], \quad t/b > 0.5. \qquad (4)$$

Overlapping Fringing Capacitance

The overlapping fringing capacitance is calculated as:

$$\Delta C/\epsilon = 4\sqrt{(Cov(t/b)\,Cov(w/b))} \qquad (5)$$

where

$$Cov(x)/\epsilon = .3540\,(x)^{0.0915}, \quad 0 \leq x \leq 0.01;$$
$$= .4413\,(x)^{0.1412}, \quad 0.01 < x \leq 1, \qquad (6)$$
$$= .4413, \quad x > 1$$

[Continued on page 128]

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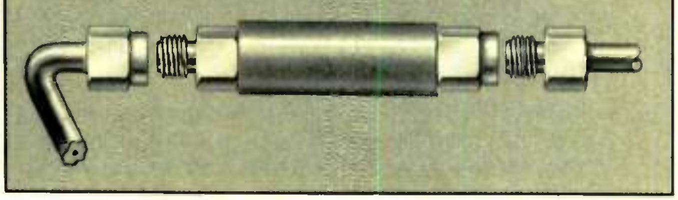

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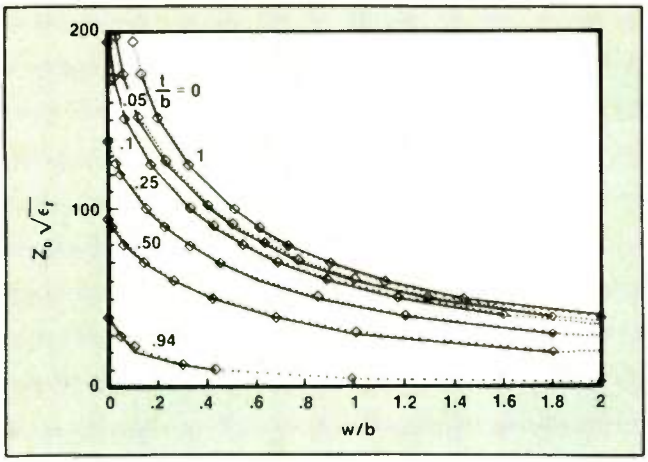

Fig. 4 Characteristic impedance of slab line as a function of the normalized width. Solid lines are exact solution by Bates. Dotted lines and markers are values using this technique.

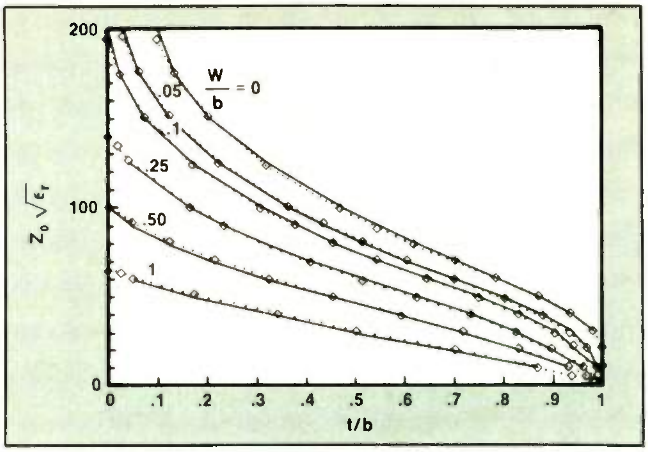

Fig. 5 Characteristic impedance of slab line as a function of the normalized thickness. Solid lines are exact solution by Bates. Dotted lines and markers are values using this technique.

and x is either t/b or w/b. Of course, t/b must be less than 1. If $\Delta C/\epsilon$ is greater than Ch/4, then it is equal to Ch/4, or if $\Delta C/\epsilon$ is greater then Cv/4, then it is equal to Cv/4.

Calculated Results

Figure 4 shows the characteristic impedance of slab line plotted as a function of the normalized width, w/b, with the normalized thickness, t/b, as a parameter. The solid curves are the results of Bates.[1]. The dotted lines are calculated using Equations 1 through 6. The markers indicate the calculated points.

Figure 5 shows the characteristic impedance as a function of t/b with w/b as a parameter. The results are in excellent agreement over the entire range of w/b and t/b.

Figure 6 shows the results of Bates, Wheeler, and Chang and this technique plotted together for a very narrow line (w/b = 0.05). It must be pointed out that this represents one of the worst cases for the results of Wheeler and Chang, very narrow thick lines. In most practical cases where the lines are wider than they are thick, their results also are in excellent agreement with

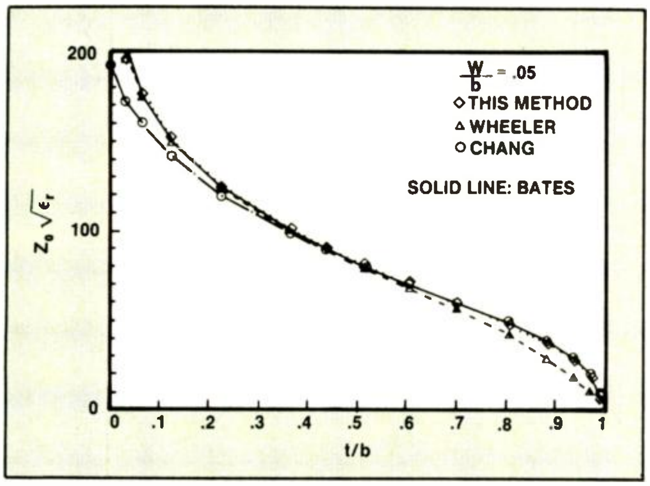

Fig. 6 Comparison of characteristic impedance using different methods of computation.

those of Bates. The solid line is again taken from Bates. The data points represented by the diamonds were calculated using this technique. Triangular data points were calculated using Wheeler's equations and the circular points using Chang's. It is interesting to note that Chang's results are too low for small values of normalized thickness (the derivation requires w/h>1), while Wheeler's tend to be too low for large values of t/b.

Derivations

The derivation of this technique is based upon the static capacitance components of the slab line. The capacitances from each face of the inner conductor are calculated from the thin line (t=0) cases are shown in Figure 7. The total capcitance of the line is simply

$$C = 4\,(Cv/4 + Ch/4 - \Delta C/4) \qquad (7)$$

where $\Delta C/4$ represents the overlapping fringing capacitance at the corners, which already has been in-

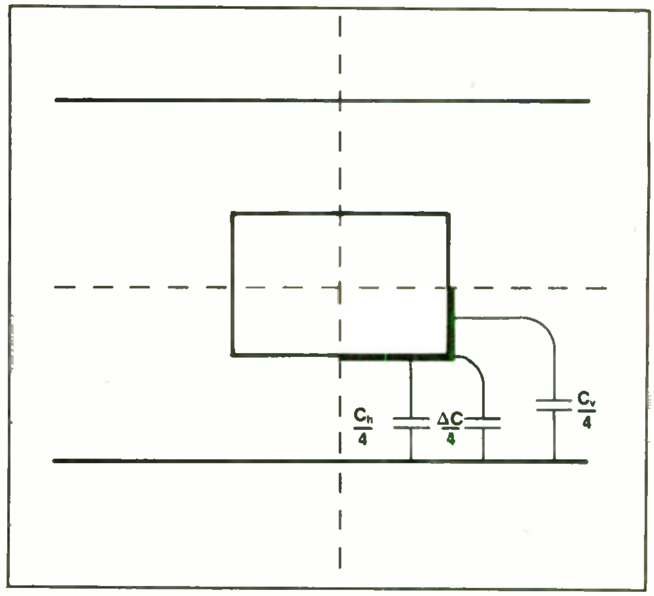

Fig. 7 Thin line capacitance components of thick center conductor.

cluded in both the horizontal and vertical thin line capacitances.

The derivation of the capacitances for thin vertical and thin horizontal strips can be obtained from reference 4.

The equations for the overlapping fringing capacitances were empirically derived on the basis of an inner conductor with a square cross-section. The impedance of a similar transmission line with a circular rather than square inner conductor is given by Frankel as[7]

$$Z_0 = \frac{60}{\sqrt{\epsilon_r}} \ln (4/\pi \; b/do) \qquad (8)$$

where do is the diameter of the circular inner conductor. The circular inner conductor can be equated to a square inner conductor[8] as

$$w = do/1.176 \qquad (9)$$

where w is both the width and thickness of the line (t/w = 1.0). Substituting Equation 9 into Equation 8 results in the equation for the impedance of a transmission line consisting of a square inner conductor between two parallel planes,

$$Z_0 = 60 \ln \left(1.0827 \; \frac{b}{w} \right). \qquad (10)$$

The normalized capacitance for the slab line containing a square inner conductor is

$$Csq/\epsilon = \frac{6.2788}{\ln (1.0827 \; b/w)} \qquad (11)$$

which is valid for $w/b < 0.5$.

The overlapping fringing capacitance at just one corner of the square inner conductor line is one quarter of the difference between the total capacitance and the sum of the capacitances of the infinitely thin horizontal and vertical lines.

$$\frac{Cov(x)}{\epsilon} = \frac{1}{4} \left(\frac{Csq}{\epsilon} - \frac{Cv}{\epsilon} - \frac{Ch}{\epsilon} \right). \qquad (12)$$

The value of $Cov(x)/\epsilon$ is plotted in Figure 8 for a wide range of x values. Equation 12 can be used to calculate Cov(x). However, Equation 6, which has been obtained from Figure 8, is in a form that makes the calculation

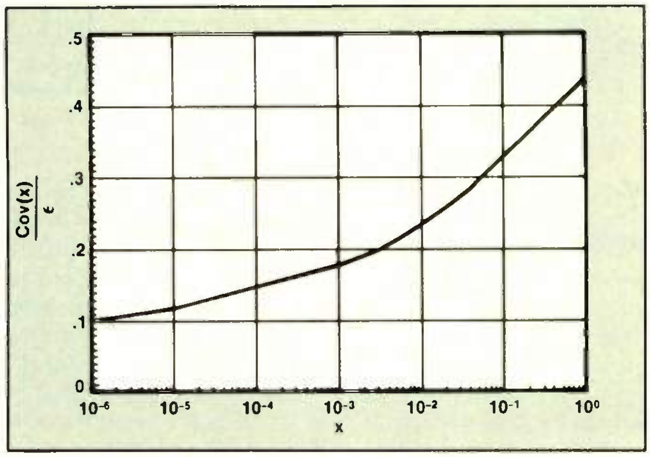

Fig. 8 Excess or overlapping fringing capacitance at one corner of a slab line with square center conductor.

much simpler to perform.

The total overlapping fringing capacitance at all four corners of a rectangular cross-section, $\Delta C/\epsilon$, is empirically derived by noting that $\Delta C/\epsilon$ must become zero as either w or t becomes zero, and should be 4Cov (w=t)/ϵ for a square cross-section. The simplest function that has these properties is the geometric mean of two square cross-sections, one of which is square with the dimension t and the other square with the dimension w. The resulting equation for the total overlapping fringing capacitance is given in Equation 5, where the factor 4 is used because all four corners are being considered. The excellent results obtained by using the geometric mean is its justification.

Conclusion

Simple formulas were presented for calculating the characteristic impedance of strip or slab line. The formulas are valid for any rectangular cross-section of the inner conductor, including the degenerate cases of infinitely thin horizontal or vertical strips. The method is especially useful for hand-held calculators, due to the simplicity of calculation, and for CAD programs, since one set of equations covers all possible inner conductor shape ratios. ■

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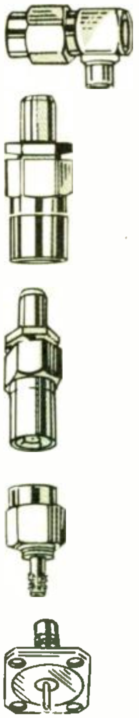

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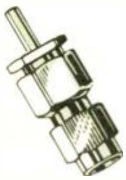

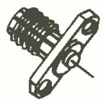

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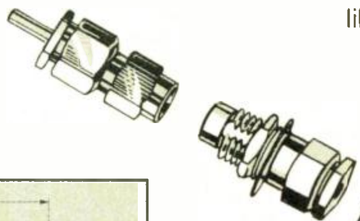

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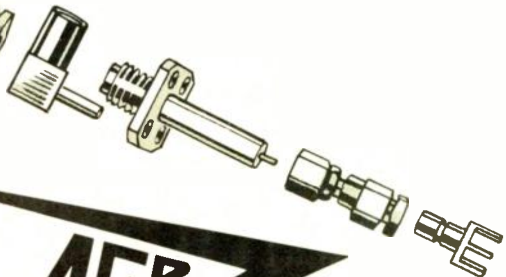

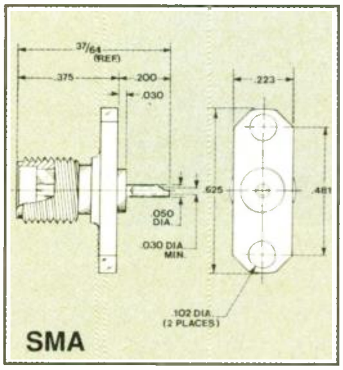

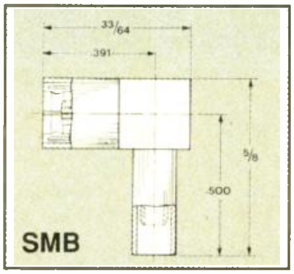

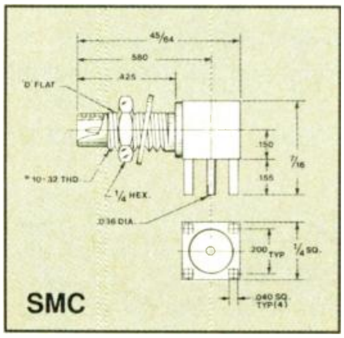

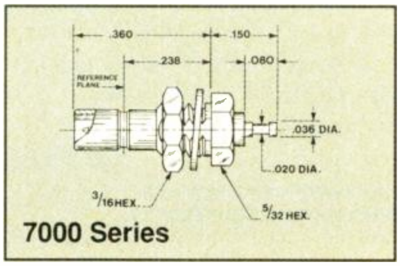

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High Performance MMIC Hermetic Packaging

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Introduction

GaAs MMIC packages in the future will have to meet stringent electrical and environmental requirements that must be addressed concomitantly with the requirements of subsystem level integration and affordability. A high performance ceramic based packaging for the hermetic housing of the GaAs MMIC devices has been developed.

An example of this packaging technology is shown in Figure 1. This is a 0.5" x 0.5" package designed for high power MMIC amplifier chips. The design of the controlled impedance I/O ports readily facilitates integration at the microwave subsystem level. The planarized G-S-G RF transitions permit direct cascading of the packaged MMICs. Computer modeling of the package ensured desired RF performance with an economically viable structure.

Insertion loss of each I/O port is less than 0.38 dB to 18 GHz with a complementary I/O port isolation in excess of 35 dB. I/O isolation below 10 GHz is above 50 dB. A low thermal impedance is presented to the MMIC device by means of a specially designed laminar chip mounting structure, which is located above a thermal via integrated into the package base.

The present packages can be mounted by solder reflow methods. This approach allows a high RF subassembly packaging density while providing a low thermal impedance to the MMIC device. Other base designs would facilitate flange or stud mounting. Solder sealing of the package lid ensures a hermetic environment for the enclosed MMIC chip. The planar RF transitions and the lead frame for DC connections permit the design of production fixturing for automated testing. Some of the more salient features of this package are highlighted in Figure 2.

Package Design and Construction

The package is designed to service the needs of both narrow- and wideband MMICs with their often inherent high power densities. The design of a wideband I/O port transition that is readily interconnected into typical microwave subassemblies makes direct cascading of MMIC functions practical and subsystem assembly economically viable. The inclusion of bypass features internal to the hermetic package simplifies the subsystem electrical design and ensures stable out-of-band operation.

The interior features of the package, which has been designed to house an MMIC power amplifier chip, are shown in Figure 3. Maximum utilization of subsystem space is accomplished by allowing the bias and control signals to be supplied to the packaged MMIC in the RF signal plane or from an elevated plane. The package thermal imped-

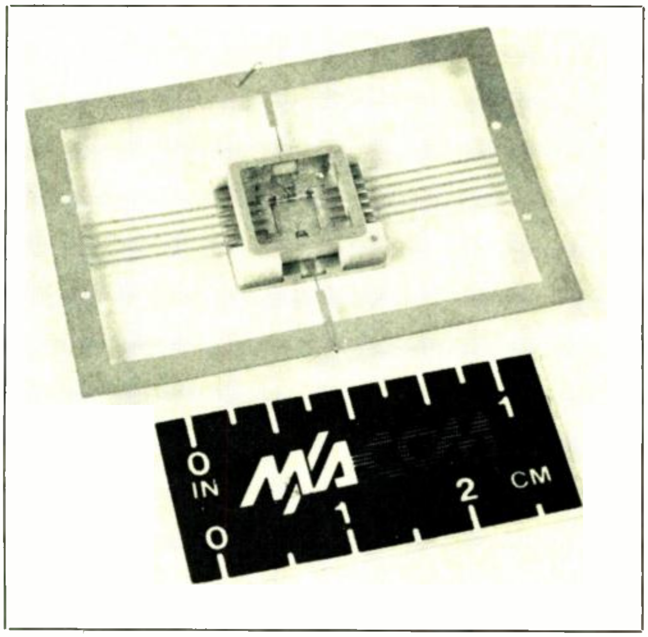

Fig. 1 Hermetic MMIC package for high power amplifier chips.

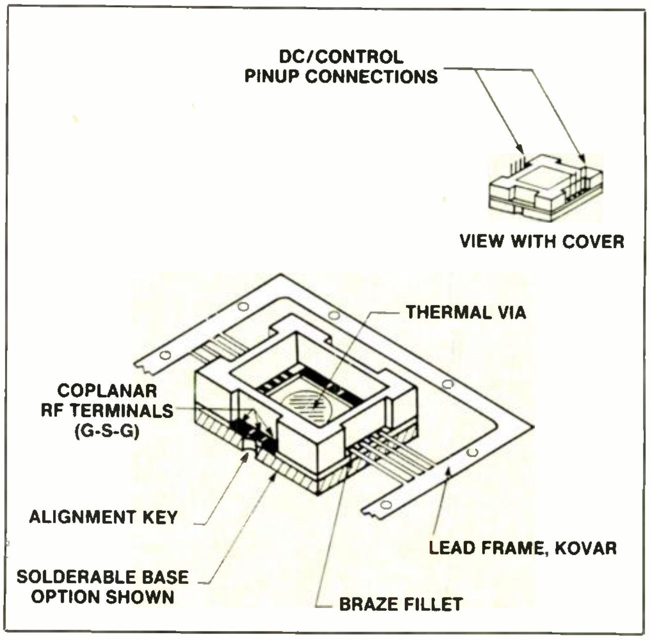

Fig. 2 MMIC package with highlighted features.

ance can be kept at a minimum when necessary through the use of a high conductivity thermal via in the metal base. Typically, the thermal impedance presented to an MMIC power amplifier chip would permit its operation at a package base temperature of +60°C with FET junction temperatures of less than +150°C. The normalized thermal impedance relative to the MMIC die attach surface is given as

$$\theta = 0.133/[W_g\ (\text{mm})\ A_c\ (\text{in}^2)],\ ^\circ\text{C/W}.$$

Thus, for an MMIC PA die with a thermal contact of 0.0053 in^2, the thermal impedance from the die attach surface to the base bottom surface is 2.2°C/W.

The package I/O transitions are designed to facilitate planar connections from adjacent microstrip or coplanar transmission line of various substrate materials and thickness and is illustrated in Figure 4. The optimum performance is achieved with 96 to 99.6 percent purity alumina substrates of 0.025" thickness. The planar RF transition removes the typical dependency on the dimensions and quality of the base ground connection to the RF subsystem. Repeatability and cost effectiveness of a subsequent subsystem assembly is enhanced. This wideband package I/O connection to the MMIC is facilitated by the use of the G-S-G planar configuration into the chip. The wideband package performance is achieved with the use of the planar I/O transitions as well as the design of the hermetic wall/lid sealing structure.

The exact features of the package were designed using an equivalent circuit that allows the physical features to be optimized for desired RF performance within the constraints of the packaging materials, processing technology and the ultimate cost objectives. The package equivalent circuit was implemented and analyzed using the computer program Super-Compact. The equivalent circuit electrical model used for this package development is shown in Figure 5.

The package wall height was selected to provide an adequate thermal impedance to facilitate the lid sealing operation. The location of the grounding via holes, the wall height and the I/O port dimensions significantly affect the I/O isolation function. The insertion loss, which consists mostly of metal losses due to the porous screen-printed structures with their inherent poor edge acuity, is less than 0.25 dB through X band per I/O port.

The package illustrated in Figures 1, 2 and 3 consists of a metal base, ceramic substrate, ceramic seal frame, kovar lead frames and ceramic lid. The ceramic substrate and seal frame are co-fired alumina structures. The metalization used is screen-printed tungsten co-fired with the integrated alumina structure. Following the CuAg braze attachment of the metal base and lead frames, high quality (50 μinches) nickel and gold electro-plating is deposited on all the metal surfaces. All the ceramic components are of 94 to 95 percent alumina fabricated with green tape technology. The

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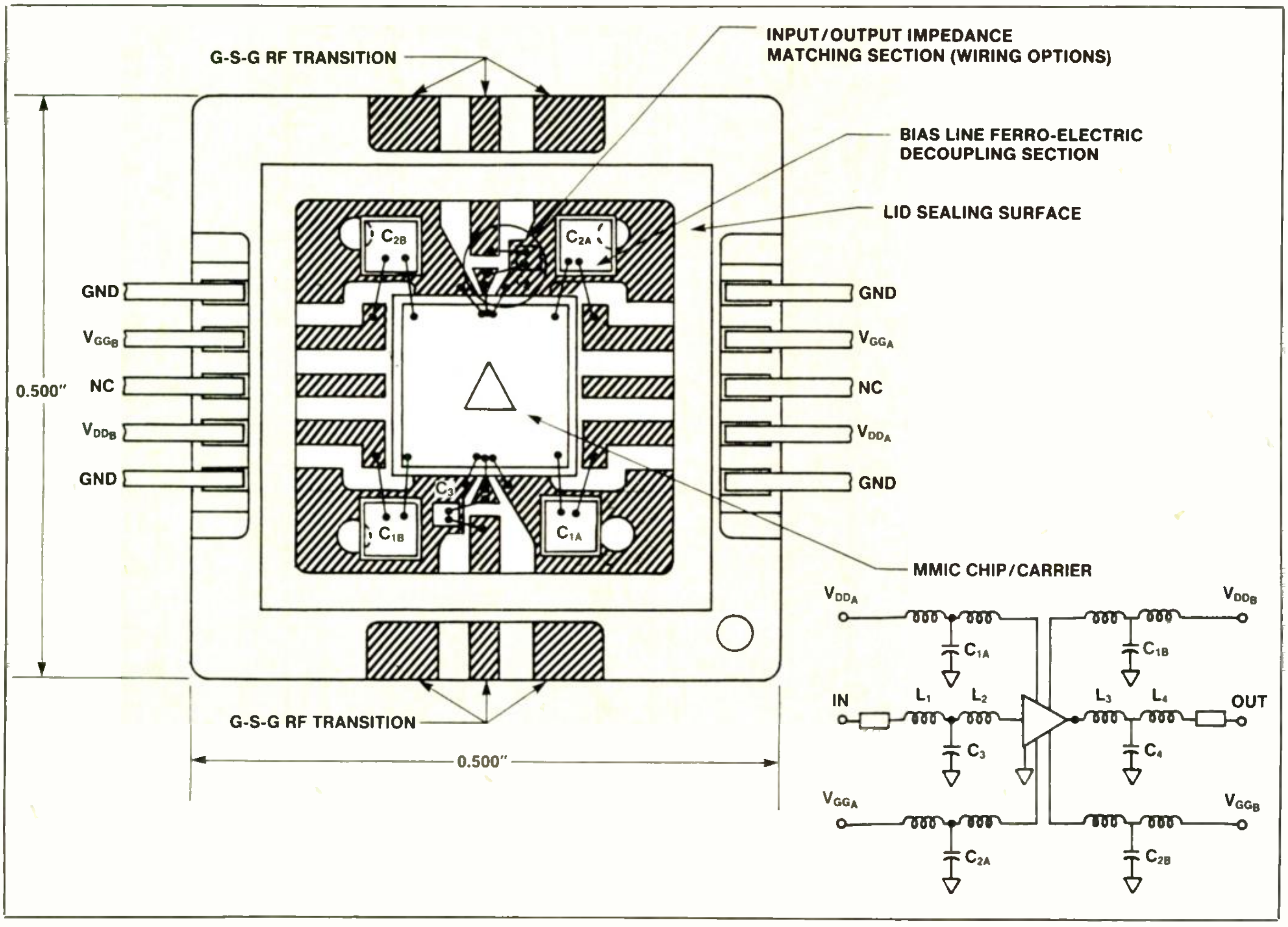

Fig. 3 MMIC power amplifier package assembly detail.

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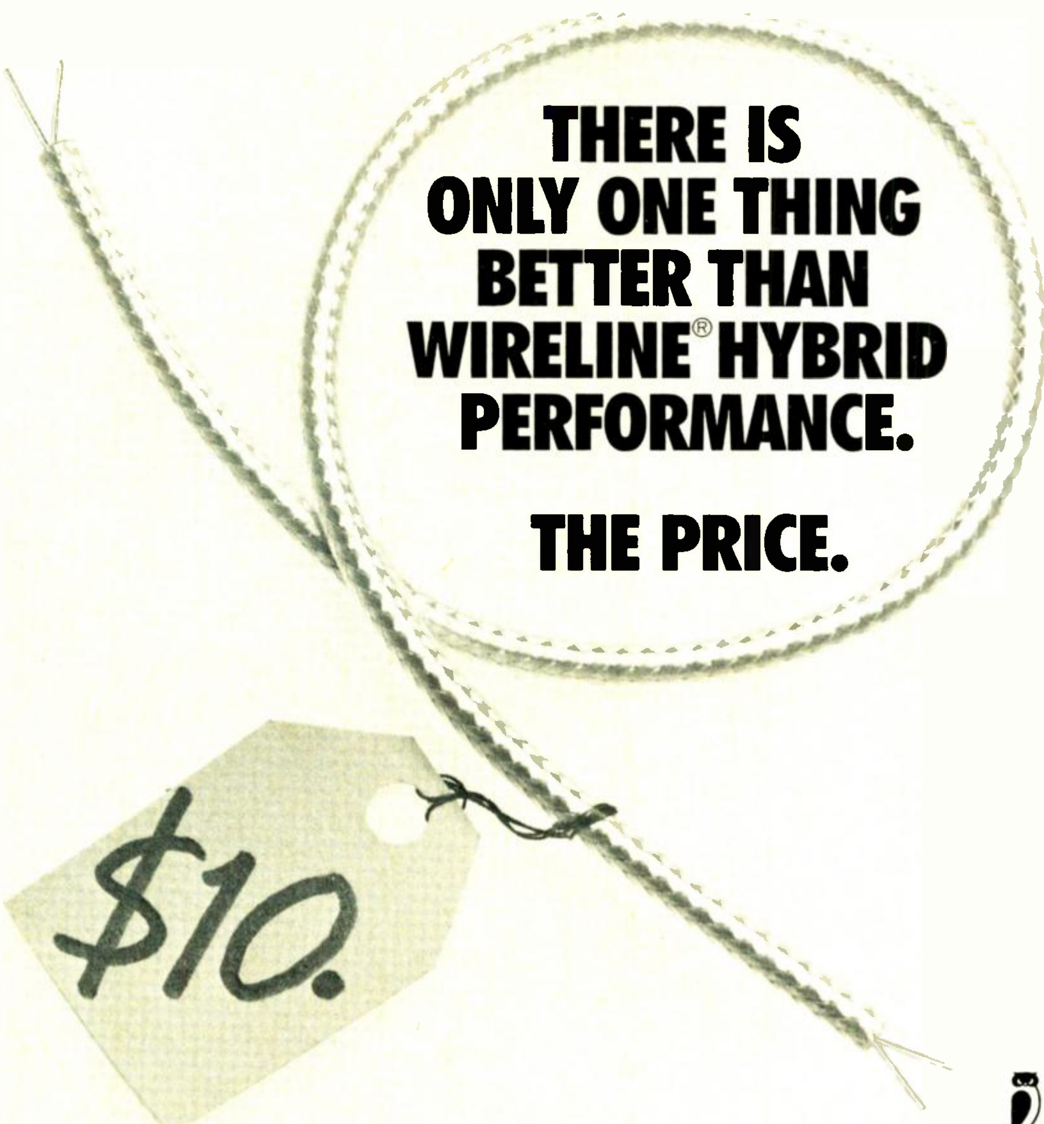

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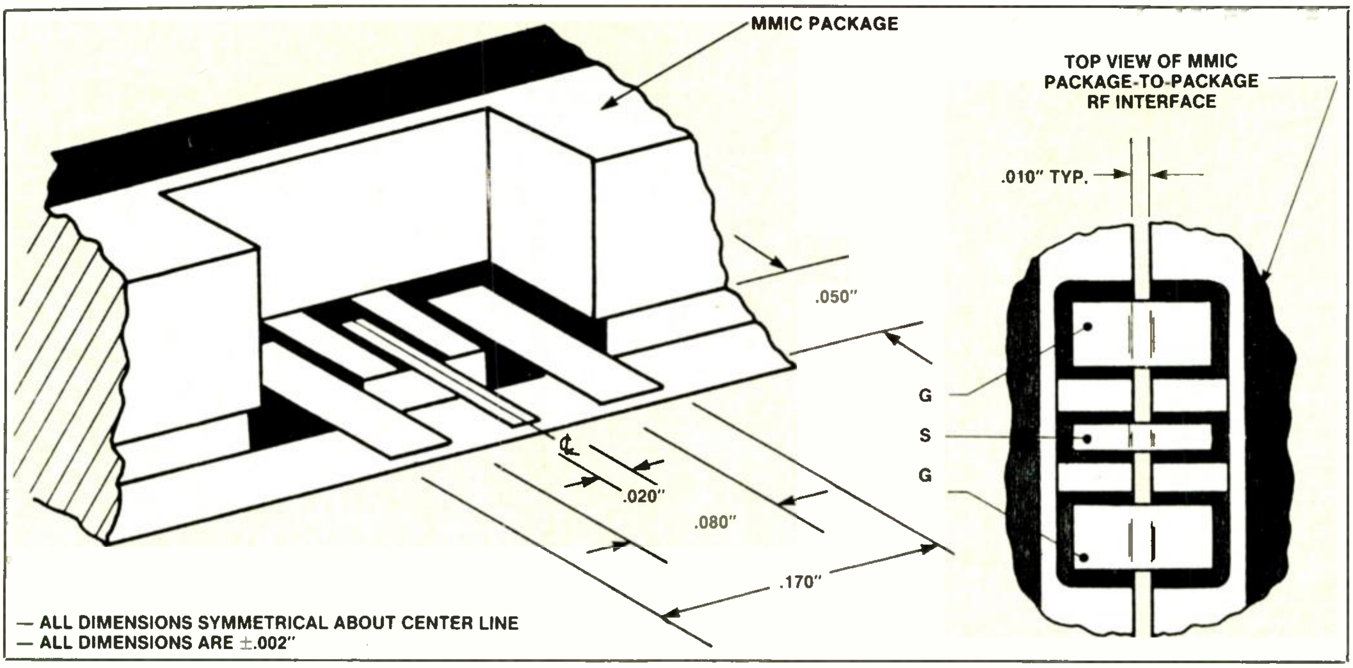

Fig. 4 Planar G-S-G RF interface footprint.

package base is a kovar structure featuring a braze/solder moat, thermal via (when needed) and a key slot for accurate positioning and cascading in the subsequent RF subassembly.

The thermal via is a brazed copper plug in the kovar base. The lead frame for the bias and control signal access is fabricated from 0.005" thick kovar by conventional photo-etch processing. The MMIC die is mounted on its molybdenum or laminar carrier and then tested for DC and thermal characteristics prior to its assembly into the package.

The MMIC chip/carrier assembly is illustrated in Figure 6. The die attach to the carrier is effected with a eutectic gold-tin alloy. The attach-

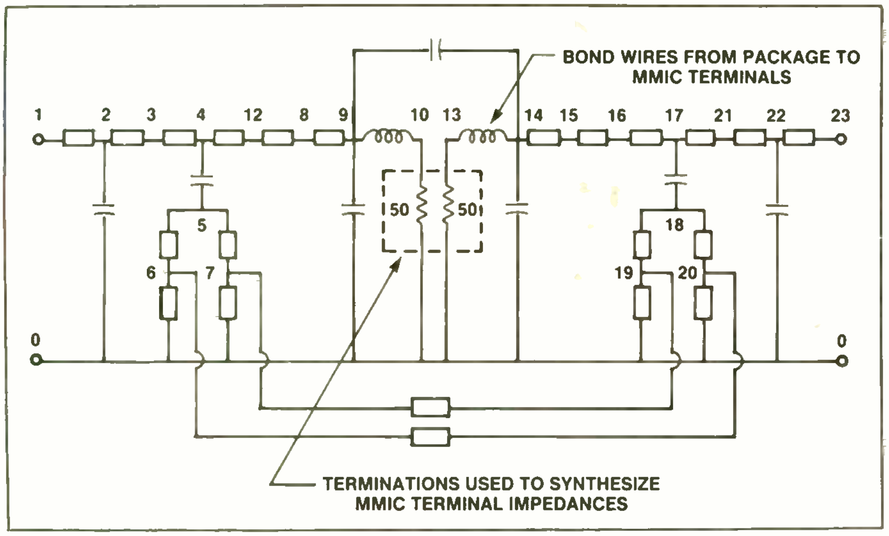

Fig. 5 MMIC package electrical model.

ment of the chip/carrier into the package is performed with a tin-silver alloy. Thermal sonic wire and ribbon bonding are used to connect the MMIC to the package. The bias bypass and RF tuning capacitors are attached with the eutectic gold-tin alloy prior to installation of the MMIC chip/carrier assembly. Hermetic lid sealing is effected with a tin-silver alloy.

Integration into RF Subsystem

The design of a high performance and cost-effective package must account for its subsequent installation in an RF subsystem assembly. Repeatable RF transitions in cascaded MMIC applications or to the accompanying mother board is critical to realizing the full potential of the packaged MMIC performance. It was to this end that the standardized (G-S-G) RF interconnection was adopted. The planar RF connection approach removes the dependency of RF interconnection performance from the variations inherent in the ground connection through the base mounting to the RF subsystem. The package mounting base is designed to provide the low thermal impedance from the MMIC chip to the RF subsystem housing with a reliable metal-ceramic package structure.

The standardized planar (G-S-G) RF interconnection footprint is illustrated in Figures 2, 3 and 4. The transmission medium is essentially microstrip from the package exterior to the MMIC chip terminals. Edge metalization on the package substrate connects the ground plane to the upper metalization surfaces, effecting a G-S-G transition. Some infringing of the grounded substrate metalization into the microstrip transmission fields is tolerated in order to optimize the G-S-G transition geometry at the MMIC chip and at the package edge, as well as reduce the undesired coupling to the seal area metalization.

The DC and control connections to the package are made with a standard photo-etched lead-frame brazed to the ceramic substrate. The lead-frame connections allow either a planar connection to the accompanying motherboard or connection to circuitry above the MMIC package in denser RF subsystem packaging.

The package is designed to be used with 0.025" thick motherboard material, although other thicknesses can be used at some compromise in the planar (G-S-G) transition performance. The package illustrated in Figure 2 is designed to be soldered into the RF subsystem. Keyways in the base locate the package in the subassembly. The solderable base permits a high subsystem packaging density to be achieved with a minimum thermal impedance mounting. Other base designs can be used if flange or stud mounting is desired.

Package RF Performance

The RF evaluation of the MMIC package was conducted by first in-

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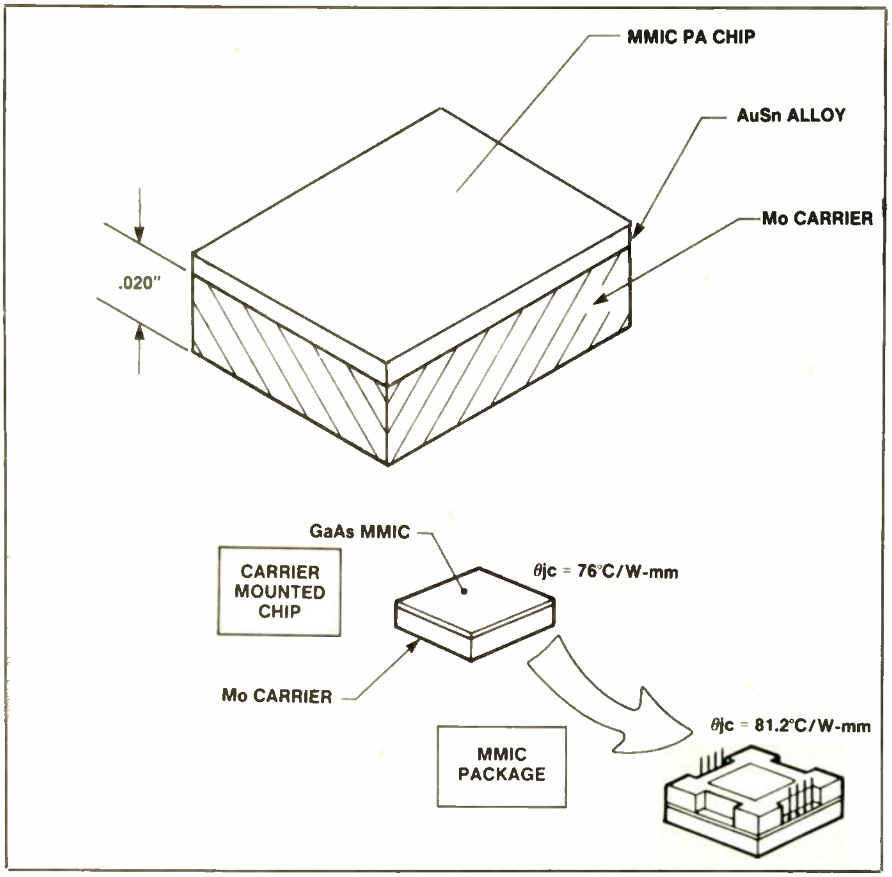

Fig. 6 MMIC chip/carrier.

stalling the package in a fixture that provided the transition from an SMA coaxial environment to the G-S-G microstrip interface. Shielding of the input and output microstrip transition regions of the fixture was used to reduce the fixture feedthrough and thus enhance the dynamic range of the measurements.

A 28 pS time domain reflectometer (TDR) measurement technique was used to assess the transition and through-wall microstrip performance. Both a short circuit and 50 ohm line terminations were used on the package interior to obtain the desired measurements. Analysis of the TDR measurements indicates that the line impedance of the microstrip line in the package is about 51 ohms with the area under the package wall being about 53 ohms. The G-S-G transition was measured to be approximately 53 to 55 ohms with a gap of 0.010" used between the package and fixture.

The insertion loss of the input and output (I/O) sections of the package were measured using both a through-line substitution and the short circuit reflection method. Both methods produced nearly identical test results. The measured performance is plotted in Figure 7 and indicates less than 0.25 dB loss per I/O port through X band. The isolation of the input terminal from the output terminal, measured with both a short circuit and a 50 ohm termination within the package, also is shown in Figure 7.

Isolation of greater than 50 dB was achieved through X band. A comparison of the measured performance to the electrical model used is shown in Figure 8. Reasonable correlation of measured and modeled performance was realized.

Conclusion

A hermetic package is being produced for the enclosure of MMIC products. A standardized (G-S-G) RF interconnection technique is used to facilitate installation into RF subassemblies. A reliable and low thermal impedance mounting of the MMIC chip is assured by the chip/carrier subassembly and the package thermal vias. Insertion loss of each input and output port is less than 0.25 dB through X band with the input-output isolation greater than 50 dB. The use of contempor-

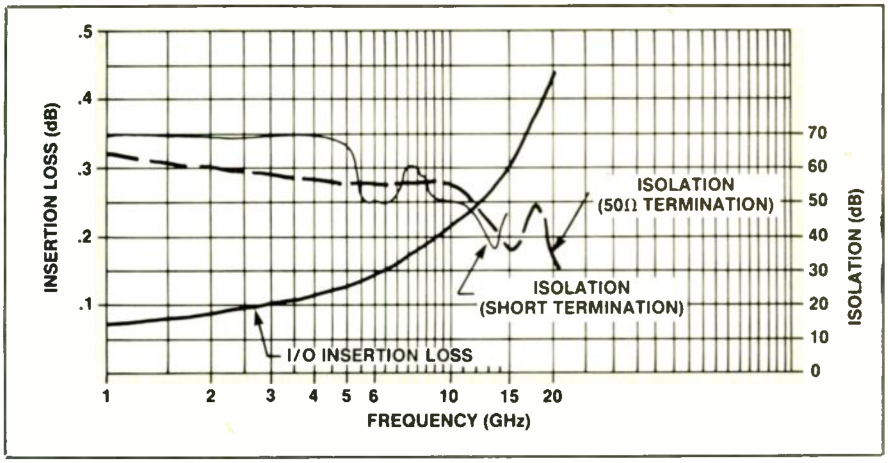

Fig. 7 Insertion loss and isolation performance of MMIC PA package.

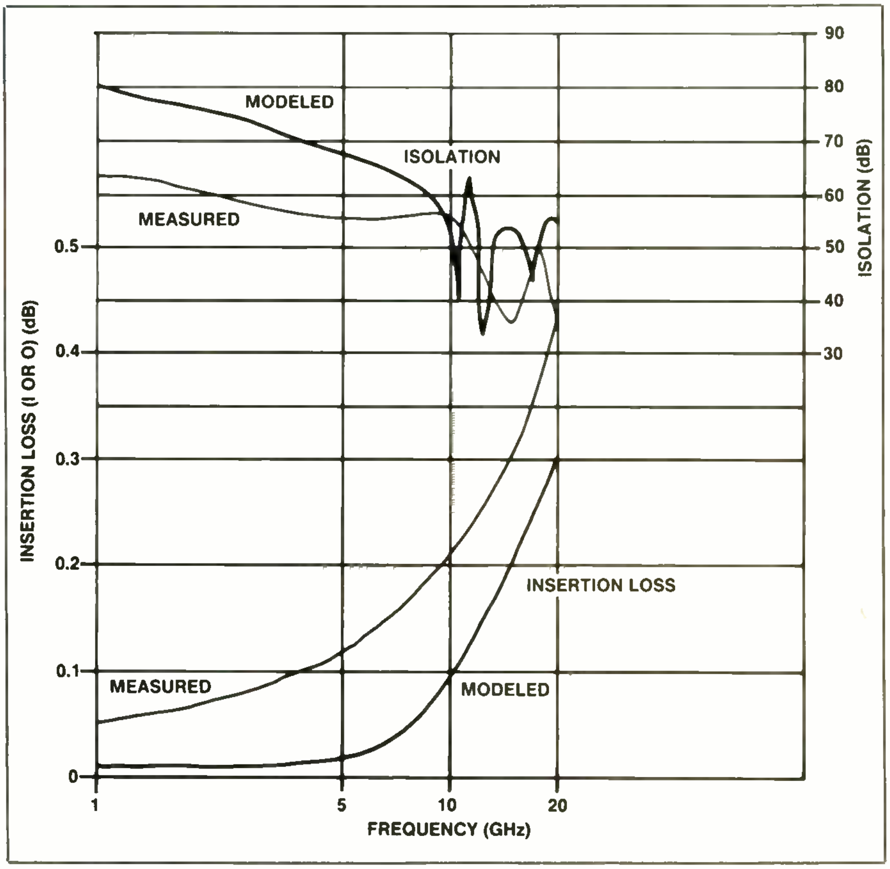

Fig. 8 Comparison of measured and computed loss and isolation of ODS-2001 packages.

ary electronic ceramic technology for the high performance MMIC package ensures that volume applications can benefit from the economics of the established manufacturing methods.

Acknowledgments

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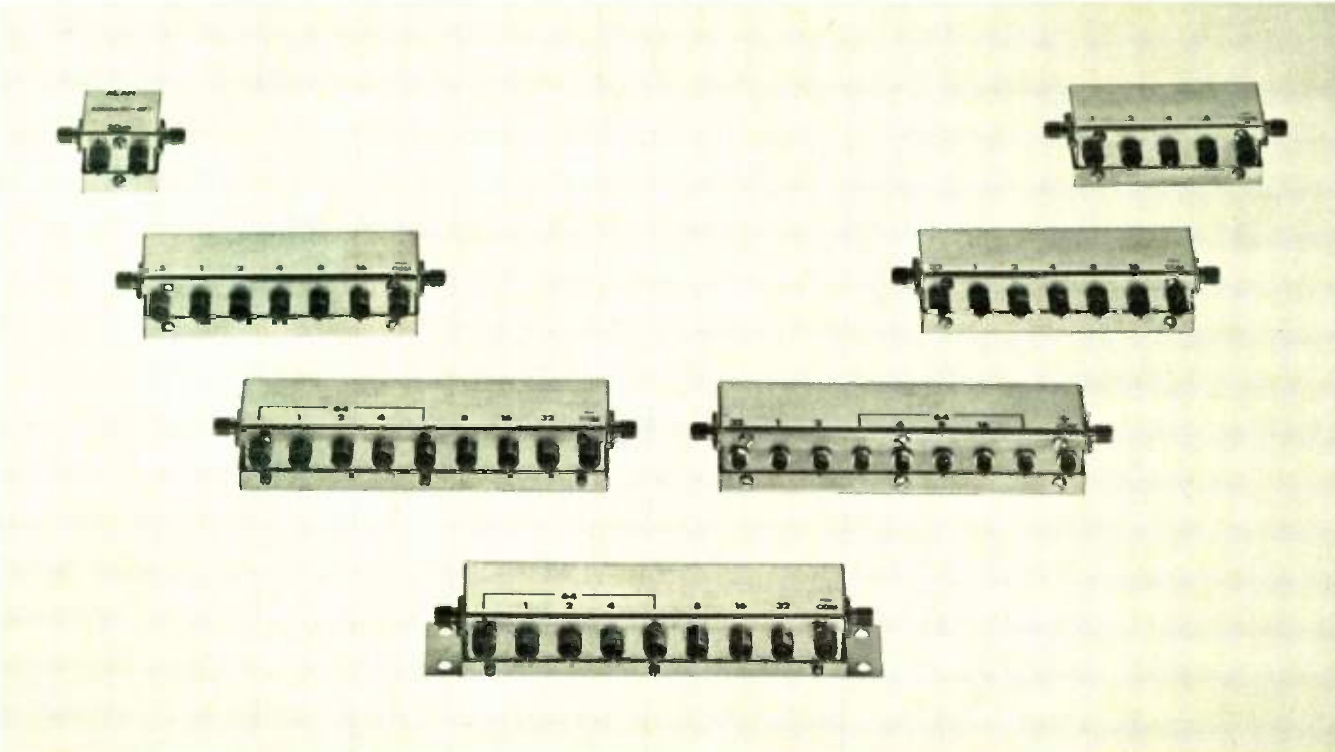
Bernhard Ziegner *is responsible for the development of internally matched FET assemblies, automated testing of FET products, packaging of MMIC devices and the reliability aspects of MMIC/FET products at M/A-Com Advanced Semiconductor Operations, Lowell, MA.*

He acquired an extensive background in MIC packaging, testing, processing and design while employed at Motorola GEG and SCG in Arizona.

From 1966 to 1981 he acquired hands-on experience at Motorola GEG in the development of a capability for processing MIC substrates as well as MIC and hybrid assemblies.

Ziegner is a member of the International Society of Hybrid Microelectronics.



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Practical Determination of Dielectric Resonator Coupling Coefficients*

Jerry C. Brand and **James F. Ronnau**
Motorola GEG
Chandler, AZ

Introduction

The coupling coefficient between dielectric resonators must be known to construct filters with a desired passband structure and bandwidth. A brief description of a basic experimental technique for determining the coupling between dielectric resonators is reviewed. Additional considerations important to the use of this method are discussed. These considerations include coupling when tuning probes are present, higher-order mode couplings and coupling when the resonators are not symmetrically placed in their enclosure.

Experimental Technique

Microwave and RF bandpass filters constructed using dielectric resonators (DRs) usually are designed in a direct-coupled configuration.[1] This coupling is achieved through the external magnetic field of the DRs and is adjusted by varying the distance between the resonators. The construction of a direct-coupled filter requires that the coupling coefficient between the transmission line and the end resonators and between individual resonators is known. This is illustrated in Figure 1. A common method of determining the coupling coefficient between resonators is the insertion loss method.[2] However, using this test with DRs can involve a great deal of machining and testing and possibly result in poor coupling data.

An experimental technique[3] that has been described recently offers an alternative to the method referenced above. This technique involves the model of a DR, image theory and the equivalent circuits of coupled circuit elements. The DR equivalent circuit can be represented near resonance as a coupled parallel RLC network, as shown in Figure 2. The resonant frequency of the DR is then given by the basic equation

$$f_0 = \frac{1}{2\pi\sqrt{L\,C}}. \tag{1}$$

The impedance of the RLC network is given by[4]

$$Z = \frac{2\,B}{1 + 2\,Q_u\,x} \tag{2}$$

where Q_u is the unloaded quality factor of the DR, B is the coupling of the DR to a test transmission line and $x = (f-f_0)/f$. The equivalent circuit values of L and C near resonance are calculated using the measured data for B and Q_u along with Equation 1.

The coupling between two DRs may be modeled as two RLC networks coupled by a mutual inductance L_m. The coupling coefficient k may be determined by using equivalent circuit representations for the coupled circuits in conjunction with a measurement technique.[3] The measurement configuration is shown in Figure 3 along with the equivalent image representation that can be applied in this case.[5] The image theory problem allows the determination of the fields that would exist if two DRs were physically in place. The transfer of energy does not take place since the model represents a mathematical equivalent. The energy is reflected as if the image resonator were short circuited. This causes a shift in the resonant frequency of the DR. The shift in the resonant frequency is measured and the mutual inductance that would cause this

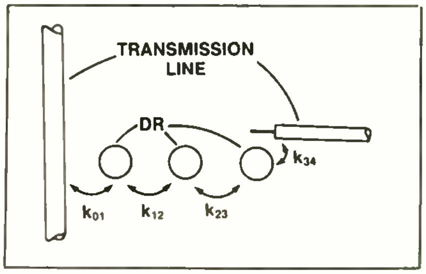

Fig. 1 Filter coupling coefficients.

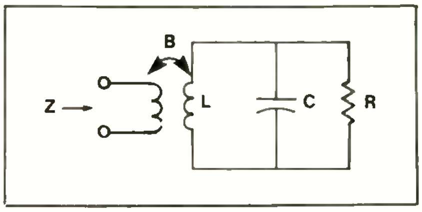

Fig. 2 Equivalent circuit of the dielectric resonator.

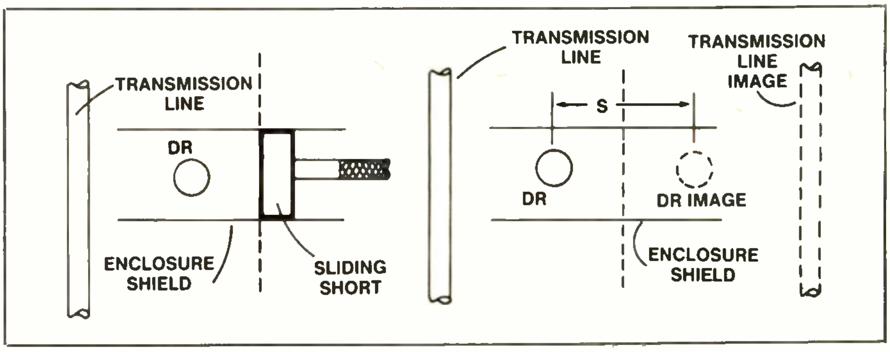

Fig. 3 Measurement configuration and the equivalent image representation.

*Invited paper.

shift is found by

$$L_m = \frac{L' L}{L - L'} \tag{3}$$

where L′ is the inductance required to resonate C at f_0. The coupling coefficient is then given by

$$k = \frac{L^2}{L^2 + LL_m}. \tag{4}$$

Thus, by measuring the resonant frequency shift for different short positions, a set of coupling coefficients vs DR separation may be found.

Measured Results

The experiment described above may be used with the DR in several ways. The position of the DR may be varied within the enclosure — for instance, in a non-symmetrical configuration. Tuning probes also may be used so that their effect may be included. Higher order modes are handled in a similar fashion.

A comparison of the experimentally determined coupling coefficients with calculated data[1] for a symmetrically placed DR is given in Figure 4. This data was used to design, construct and test a three-pole, 15 MHz bandwidth filter. The measured vs predicted response is illustrated in Figure 5.

Tuning probes are placed above the DR to vary the resonant frequency by disturbing the magnetic fields of the DR. This obviously can change the coupling between resonators. The method was used to calculate the coupling for the case of Figure 4 with tuning probes added. The resonant frequency was moved up 100 and 200 MHz, a reasonable amount for filters of this type. The results are shown in Figure 6. The RF magnetic fields of the DR also are changed when placed asymmetrically in the enclosure. The configuration and coupling coefficients for this case are shown in Figure 7. Note that the coupling changes only slightly with tuning at the more loosely coupled positions. This shows that tuning over most ranges will have little effect on the filter's response.

The higher-order modes of the DR previously have been used for certain filter applications.[6] One application that may be of interest is to use the higher-order modes for wider bandwidth filters. The use of high Q resonators of this type typ-

[Continued on page 144]

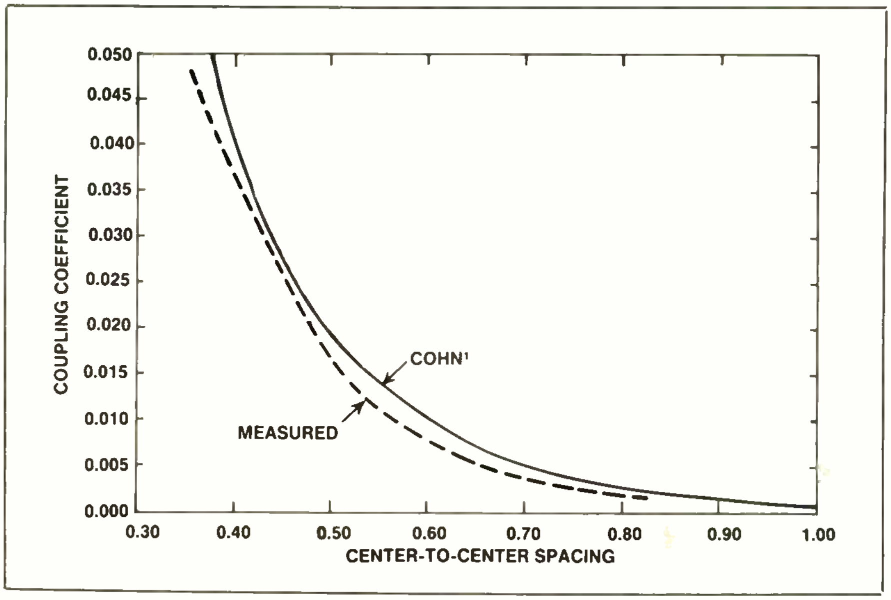

Fig. 4 Comparison of the calculated[1] and experimental coupling coefficients.

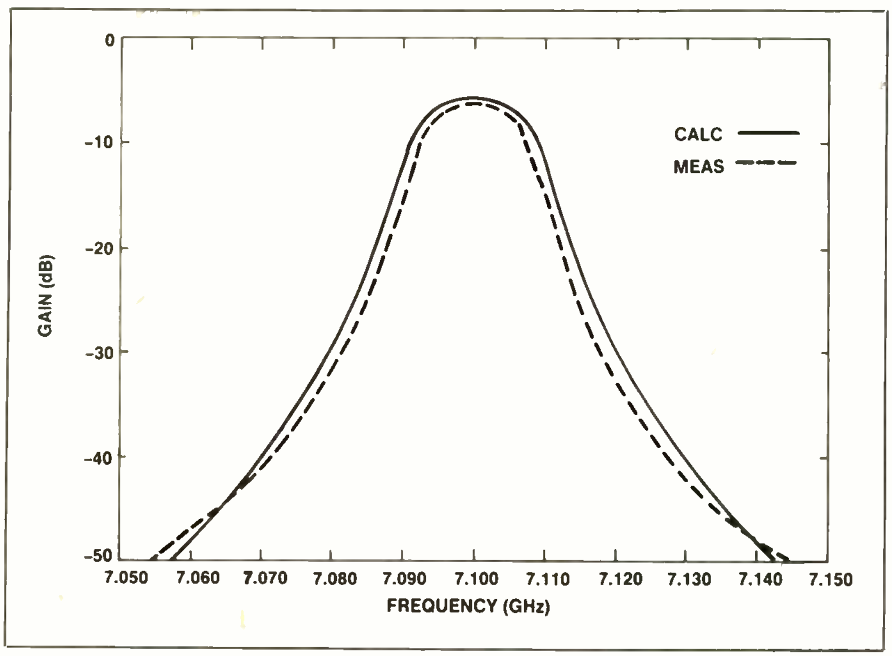

Fig. 5 Calculated and measured response of a three-pole, 0.5 dB ripple Chebyshev filter

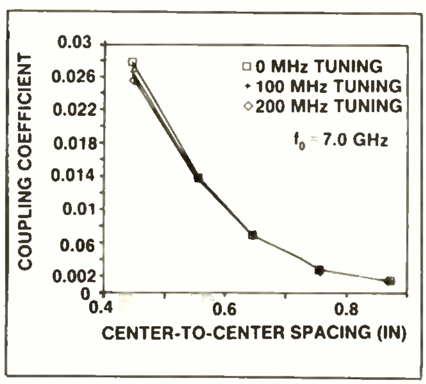

Fig. 6 Comparison of the coupling coefficients for various probe depths.

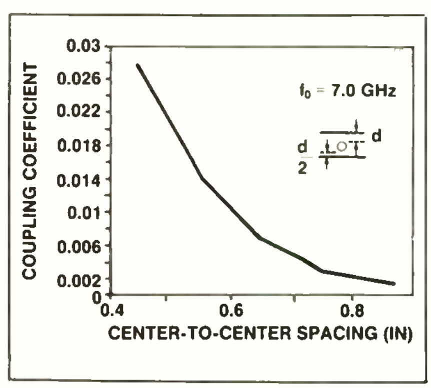

Fig. 7 Coupling coefficients for an asymmetrically placed DR.

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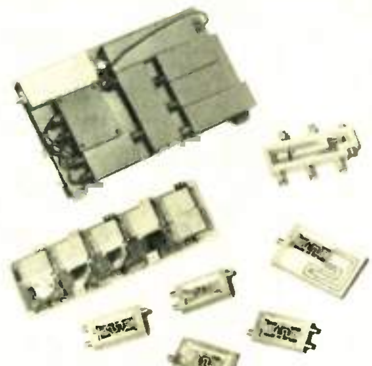

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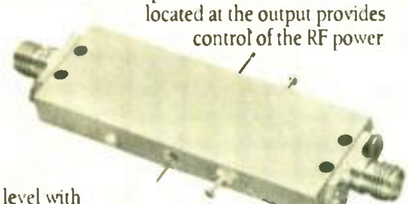

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DFD SPECIFICATIONS			
MODEL NUMBER	DFD 0206-10	DFD 0612-12	DFD 1218-12
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Unambiguous Frequency Range (GHz)	1.6-6.72	5.47-13.15	11.35-19.03
Output Data Bits	10	12	12
Mean Frequency Resolution (MHz)	5	1.875	1.875
Absolute Frequency Accuracy (MHz)	25	12	12
Signal Duration (nsec)	50 to CW	45 to CW	45 to CW
Throughput Time (nsec) Max.	175	150	150
Dynamic Range/Sensitivity (dBm) Min.	-60 to 0	13 to 19*	13 to 19*
VSWR	2.0:1	2.0:1	2.0:1
Logic	TTL	TTL	TTL
Size (Inches) Max.	5 x 10 x 1	9.3 x 6.9 x 1.9	9.3 x 6.9 x 1.9
Weight (Pounds) Max.	4	13.2	13.2

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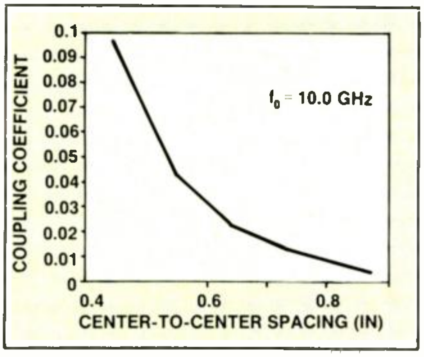

Fig. 8 Coupling coefficients for the first higher-order mode of the DR.

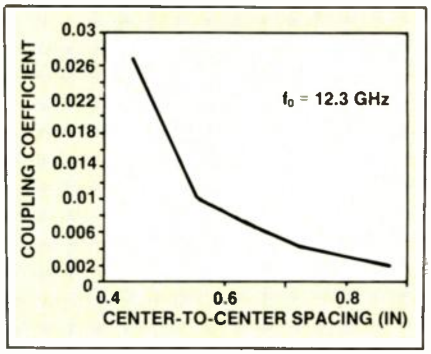

Fig. 9 Coupling coefficients for the second higher-order mode of the DR.

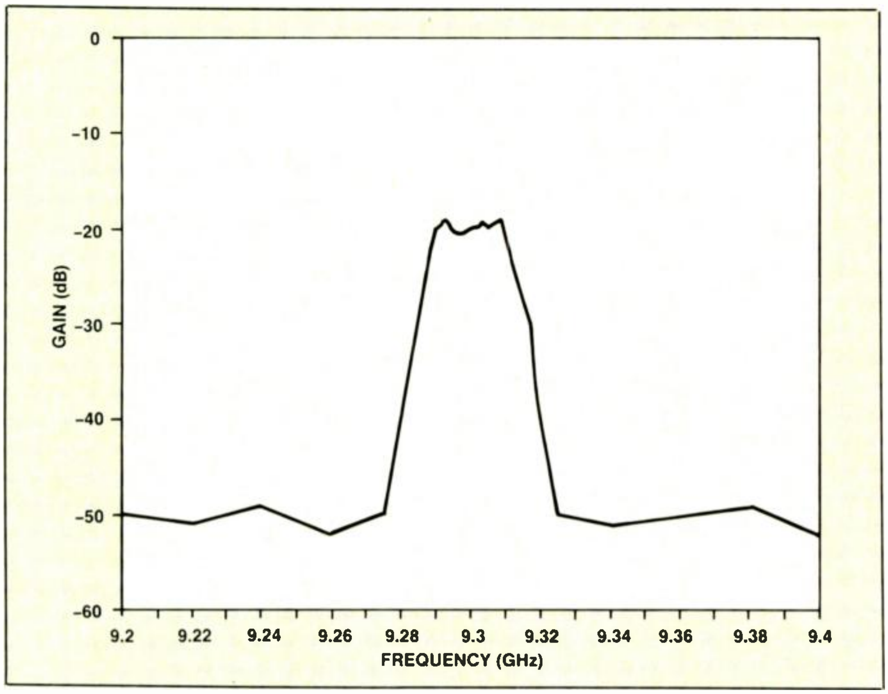

Fig. 10 Measured response of a 26 MHz bandwidth, five-pole filter constructed with the structure of a dominant mode filter but utilizing the second mode of the DR.

ically is limited to a bandwidth range of 0.1 to 2 percent. Some of the higher-order modes may well be used to construct filters with bandwidths of 0.4 to 8 percent due to the higher coupling obtainable. The coupling coefficients of two higher order modes are given in Figures 8 and 9.

The next higher-order mode shown in Figure 8 is seen to have larger coupling coefficients than the dominant mode. The coupling coefficients of the second higher-mode shown in Figure 9 are about the same magnitude as those of the dominant mode, but they have a flatter slope. The second higher mode was used to construct a filter with a 26 MHz bandwidth, with a structure in place that was used with the dominant mode for a 15 MHz design. The measured data is shown in Figure 10. This data indicates one of the problems that exists when trying to use these higher order modes. The loss has doubled over that of the dominant mode due to a decrease in the Q of that resonator mode. This problem can be addressed when incorporating the filter in a particular system.

Conclusion

An experimental method that allows the determination of coupling coefficients between two dielectric resonators has been demonstrated. Measured data have been given and are shown to compare well with calculated data. The coupling coefficients have been analyzed for various configurations and modes and the data has been used to construct filters. The filters are shown to have response curves close to those predicted. Overall, the practical determination of coupling coefficients is seen to be a useful tool for filter design and construction. ■

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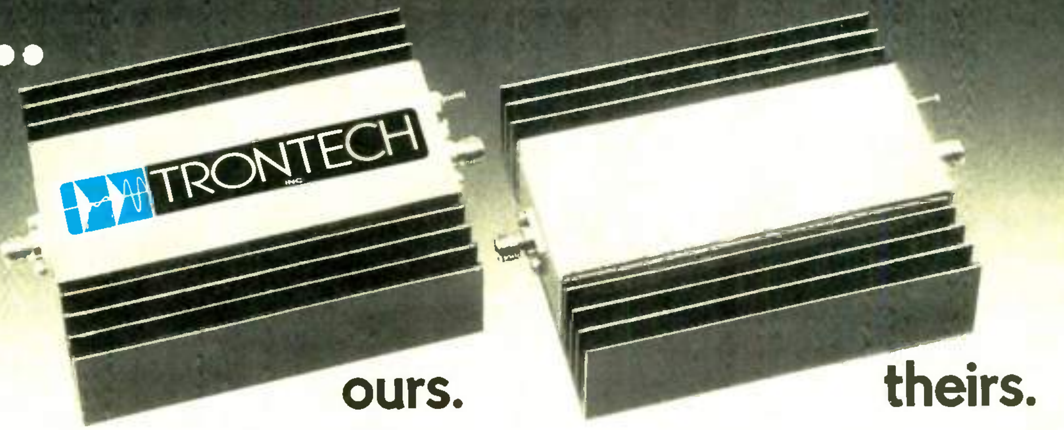

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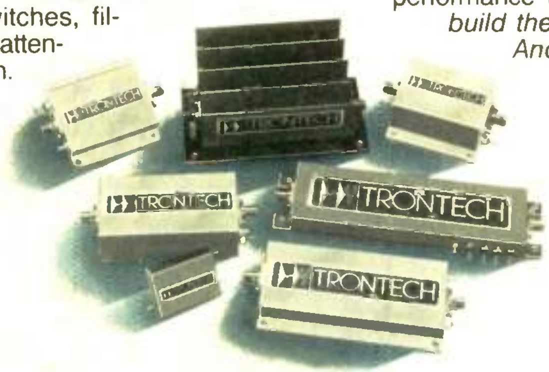

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An N-Way Broadband Planar Power Combiner/Divider*

W. Yau
and
J.M. Schellenberg
Hughes Aircraft Co.
Microwave Products Division
Torrance, CA

An N-way broadband planar power combiner/divider for ultra-broadband MMIC applications was developed using a Dolph-Chebyshev transmission line taper. The combiner/divider demonstrated a 5 to 18 GHz bandwidth with an insertion loss of less than 0.2 dB and an input SWR of no more than 1.35.

Introduction

Recently, ultra broadband (6 to 18 GHz, 2 to 20 GHz, 2 to 40 GHz)[1-3] MMIC amplifiers have been developed. However, to meet future power requirements of commercial and military systems, an efficient method of combining RF power over broad bandwidths must be developed. Conventional power combining techniques have limitations in either bandwidth, insertion loss (efficiency), or physical configuration. For example, the N-way Wilkinson divider has the advantage of low loss, moderate bandwidth and good amplitude and phase balance. However, its major disadvantage for power applications is the "floating starpoint" isolation resistors. These resistors require a non-planar crossover configuration, which limits the power handling capability of the combiner. Furthermore, it results in a non-planar structure, which makes an MIC circuit realization impractical.

To overcome this difficulty, a new broadband N-way planar power combiner/divider circuit that utilizes the Dolph-Chebyshev tapered transmission line[4,5] has been developed. The distinguishing feature of this approach is that the isolation resistors connect between adjacent coupled transmission lines, as opposed to a common floating node as in the case of an N-way Wilkinson combiner.[6] In addition, this system of tapered transmission lines simultaneously provides impedance transformation of, in general, N 50 ohm distributed ports to one 50 ohm central port.

Design Approach

Power combining can be considered, in part, an impedance transformation problem. The total N-way load (50/N) must be transformed to 50 ohms over the desired bandwidth. The key to obtaining ultra-broad bandwidth performance is in employing non-uniform transmission lines. The impedance transforming properties of non-uniform transmission lines are well documented in the literature. One major use is in the matching of unequal resistances over a broad range of frequencies. As such, the functional variation of the tapered section greatly affects its frequency response. While the exponential and hyperbolic transmission lines are examples of commonly employed transmission line tapers, they are not the optimum design. It has been shown that the Dolph-Chebyshev taper is a better choice, since it has minimum reflection coefficient magnitude in the passband for the specified length of taper. Or, for a specified maximum magnitude reflection coefficient in the passband, the Dolph-Chebyshev taper has minimum length.[1]

The design of the N-way power combiner begins with the Dolph-Chebyshev tapered transmission line. The contour and the length of the taper determine the in-band reflection coefficient and the lower cutoff frequency, respectively. The frequency response is essentially highpass, with equi-ripple in the passband.

The tapered transmission line is then segmented into N nominally identical strips. The strips are connected at one end, forming the input port of the power divider, and disconnected at the other end, forming the output ports. An array of isolation resistors connects between the tapered strips, as shown in Figure 1a. To achieve equi-amplitude and equi-phase power division at each port over the frequency band, the following conditions and approximations[7] must be fulfilled:

- The N conductors must have the same per unit length (PUL) capacitance to ground
- The PUL capacitances between adjacent conductors must be identical and

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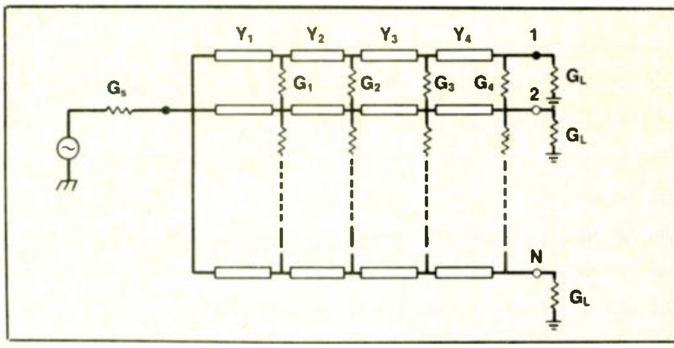

Fig. 1a Equivalent circuit of N-way power combiner.

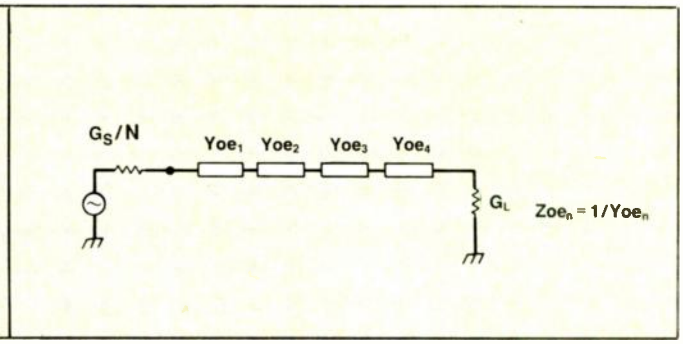

Fig. 1b Even mode representation of N-way power combiner.

- The PUL capacitance between nonadjacent conductors must be negligible.

The even mode representation of the power combiner is shown in Figure 1b. Each segment represents a non-uniform coupled transmission line. The power combiner approach is similar in concept to the N-way planar power divider proposed by Nagai[8] and Schellenberg,[9] except that non-uniform coupled transmission lines are employed. The impedance Z_{0e_n} represents the even mode impedance of the transmission line segment.

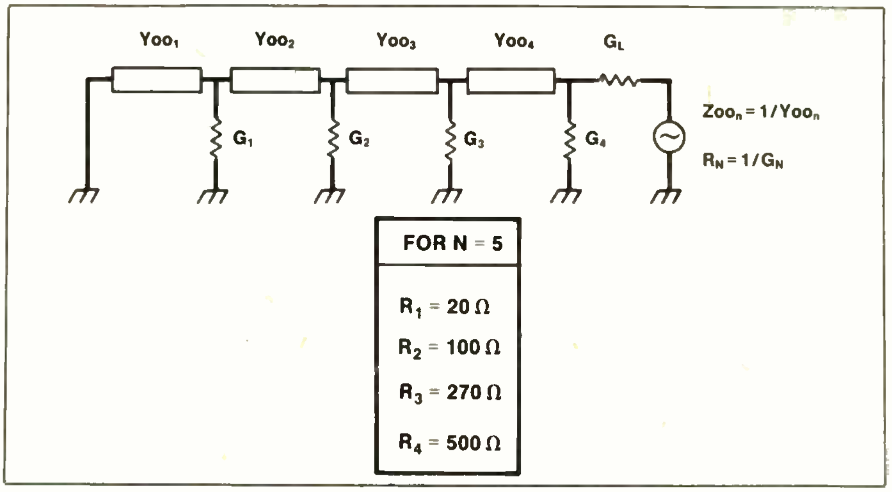

Fig. 1c Odd-mode representation of N-way power combiner.

The odd mode representation of the power combiner, shown in Figure 1c, is required to analyze the isolation resistor network. Z_{00_n} represents the equivalent odd mode impedance, which was calculated using the variational method[10] and spectral domain method.[11] The analysis of the isolation resistor network is well documented.[7,8] The isolation resistor values can be calculated by using the expressions in the appendix at the end of this article.[8]

Note that fulfilling the above conditions requires conductors with different widths and spacings. The two end conductors, in particular, have to be narrower than the intermediate conductors to account for the fringing capacitances at both ends. As will be shown in our design example, all gap spacings are kept identical, violating the second bulleted condition above. The result is a slight degradation in the isolation. However, this degradation can be compensated for by adjustment of the isolation resistors.

The above concepts have been applied to realize a five-way power combiner/divider on a quartz (ϵ_r=3.78) substrate, 25 mils thick (Figure 2). Gap spacings between adjacent conductors were kept relatively small (1.5 mil) to ensure that the coupled structure conforms to the Dolph-Chebyshev tapered line condition. Chip resistors were employed in the isolation resistor network. Their calculated values are shown in Figure 1c. Each power divider/combiner circuit has dimensions of 0.710 x 0.710 inch.

Measured Results

The S-parameters of a set of Dolph-Chebyshev tapered transmission lines, connected back-to-back, were measured (Figure 3) from 2 to 18 GHz. A power combiner/divider set was measured in the same fashion (Figure 4). The two measured results agree closely, implying that the coupled structure conforms to the tapered transmission line case. The frequency response also demonstrates the high-pass characteristics of this type of circuit with a Chebyshev equal ripple response in the passband. Due to multiple reflections, the measured 11 dB return loss corresponds to 17

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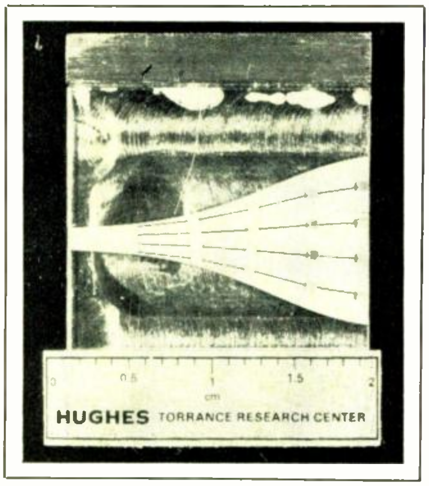

Fig. 2 A five-way planar power combiner/divider using Dolph-Chebyshev tapered transmission line.

dB (SWR=1.35) for each combiner/divider circuit. The total insertion loss of 0.35±0.5 dB implies a loss of approximately 0.2 dB for each divider.

The power division loss was measured to be −7.0±0.5 dB across the whole described frequency band (Figure 5). Between each of the five output ports, the variation in power division is less than ±0.5 dB. The equi-power split between ports, along with the low insertion loss, is a good indication of equi-phase match at each port.

Port-to-port isolation measurements also were performed. A minimum of 15 dB isolation between output ports was obtained across the 5 to 18 GHz bandwidth, as shown in Figure 6.

Conclusion

A broadband five-way power combiner/divider, using Dolph-Chebyshev taper transmission lines, has been demonstrated. A bandwidth of at least 100 percent has been achieved. It is believed to be the widest bandwidth achieved for an MIC power combiner of this complexity. The important aspect of this approach is that it allows power combining and impedance transformation in the same structure. Furthermore, the floating common node, required with the N-way Wilkinson divider, is eliminated with this design, thereby yielding a planar structure.

The development of this power combiner/divider is an important step toward meeting the increasing demand for wideband and high power hybrid/monolithic amplifiers in future commercial and military applications.[12]

Acknowledgments

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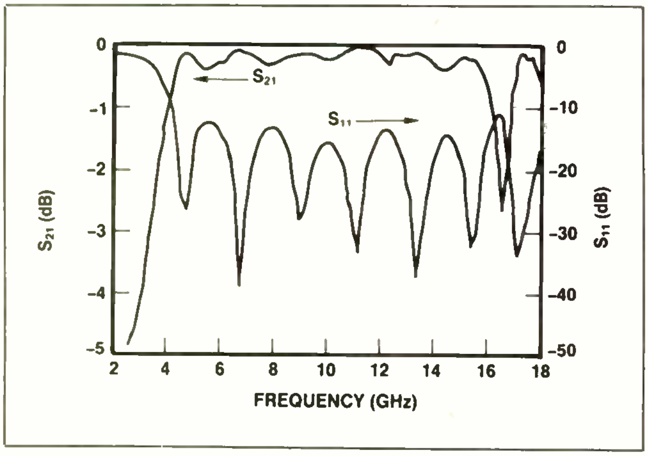

Fig. 3 S_{11} and S_{21} measurement of Dolph-Chebyshev tapered transmission line. (The resonance frequency at approximately 16 GHz is caused by the transverse resonance mode supported by the cross-section of the tapered transmission line.)

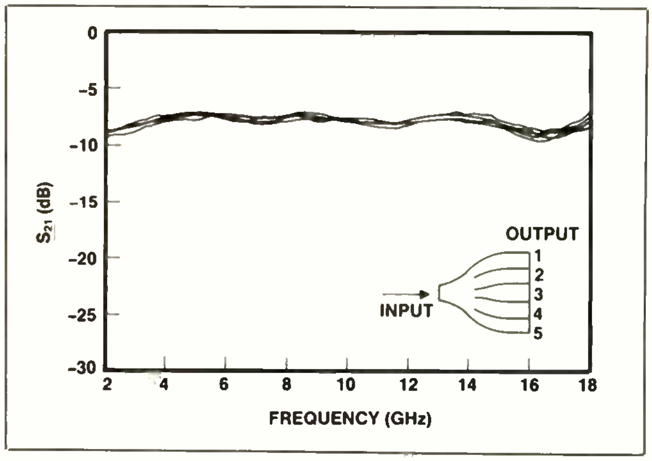

Fig. 5 Magnitude of power splits of five-way power divider.

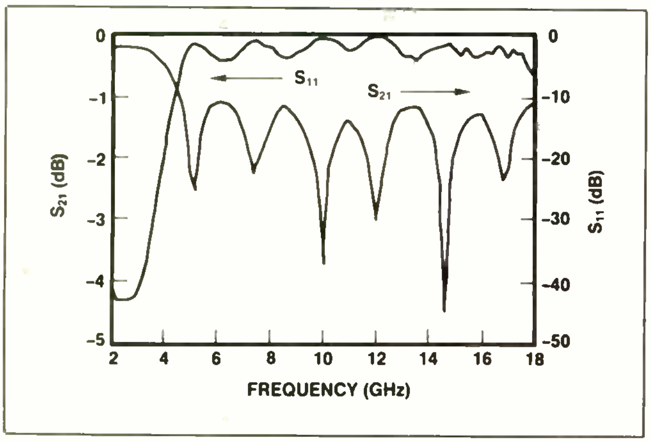

Fig. 4 S_{11} and S_{21} measurement of a set of five-way power combiner/divider using Dolph-Chebyshev tapered transmission line.

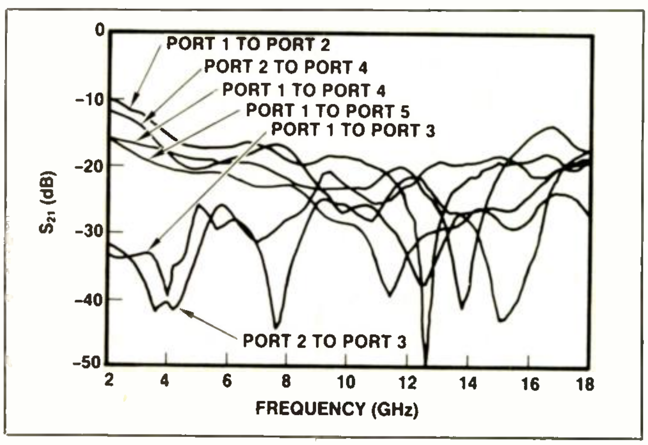

Fig. 6 Port-to-port isolation of five-way power combiner/divider.

Appendix

$$G_L = h_i G_M + \cfrac{Y_M^2}{h_i G_{M-1} + \cfrac{Y_{M-1}^2}{\vdots \atop h_i G_3 + \cfrac{Y_3^2}{h_i G_3 + \cfrac{Y_2^2}{h_i G_1}}}}$$

$$(i = 2, \ldots, N)$$

where $h_i = 2 - 2\cos\pi\,(i-1)/N$

$Y_M = Y_{00_n}$ and

$G_M = G_N = 1/R_N$; RN = isolation resistance

$G_L = 1/R_L$.

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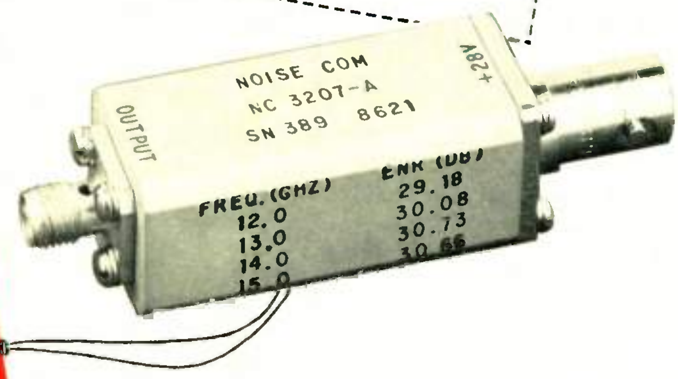

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NC 3202	.001 – .6	30 – 35	± 1DB		.01, .1, .6
NC 3203	1 – 2	30 – 35	± 1DB	A	1, 1.5, 2.0
NC 3204	2 – 4	30 – 35	± 1DB	to	
NC 3205	4 – 8	30 – 35	± 1DB	H	1GHz STEPS
NC 3206	8 – 12	28 – 33	± 1DB		
NC 3207	12 – 18	26 – 32	± 1DB		
NC 3208	1 – 18	26 – 32	± 1DB		

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NC 5228	26.5 – 40	23.0	± 2.0DB TYP ± 3.0DB max	UG599/u	2GHz STEPS	WR-28
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Matched, Dual Mode Square Waveguide Corner*

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A design method for matching a square waveguide right angle corner for both E- and H-plane (TE_{10} and TE_{01} mode) operation is presented. Results show SWR less than 1.05 for a 1.0 GHz bandwidth at X band with mode-to-mode isolation greater than 30 dB.

Introduction

Square waveguide often is used for a dual-polarization application because it can support two orthogonal modes (TE_{01} and TE_{10}) with identical phase velocity. Such an application requires waveguide bends to construct a square waveguide system. A right angle bend or corner is the most severe and represents the most difficult case to design. See Figure 1 for definition of terms. The conventional mitered corner (fixed L, see Figure 1) offers a good match for one mode; however, that mitered corner would give poor performance for the other. This paper explains a technique that results in good performance for both modes.

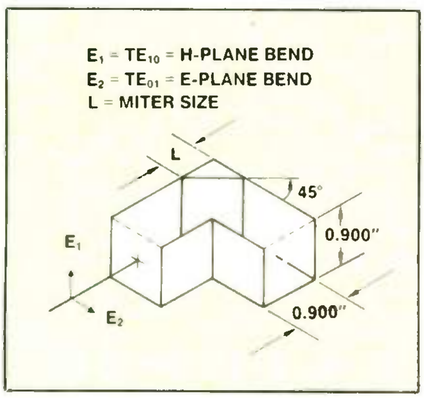

Fig. 1 Description of the square corner for low X-band operation.

Design Approach

The design philosophy was based upon the conventional mitered corner. That is, for a given frequency, there is a mitered size (L) that is well matched for a given mode (TE_{10} or TE_{01}).

Figure 2 shows representative match data vs frequency for both E-plane and H-Plane operation with the appropriate values of L to match at 7.95 GHz. Also shown are the accompanying results for the non-matched mode in each case.

Measurements taken with a range of L values allow the establishment of two empirical design curves (shown in Figure 3), which represent the optimum L dimension vs frequency for each mode. Note that the L values for the E-plane (TE_{01}) bend always are larger than those for the H-plane (TE_{10}). Consequently, a given value of L cannot provide a match for both modes simultaneously.

If, however, a reflecting surface that appeared mode-dependent in position were used in the miter, then both modes could be matched simultaneously. At this point, the concept of the polarized mitered corner was introduced. The orthogonality between the two modes allows us to use a polarizer to provide this mode-dependence.

By using a flatplate reflector that has added many raised ridges oriented parallel to the E-field of the TE_{01} mode, as shown in Figure 4, the effective shorting plane for the TE_{01} mode will be coincident with the ridge tops. The TE_{10} mode, which will have the E-field perpen-

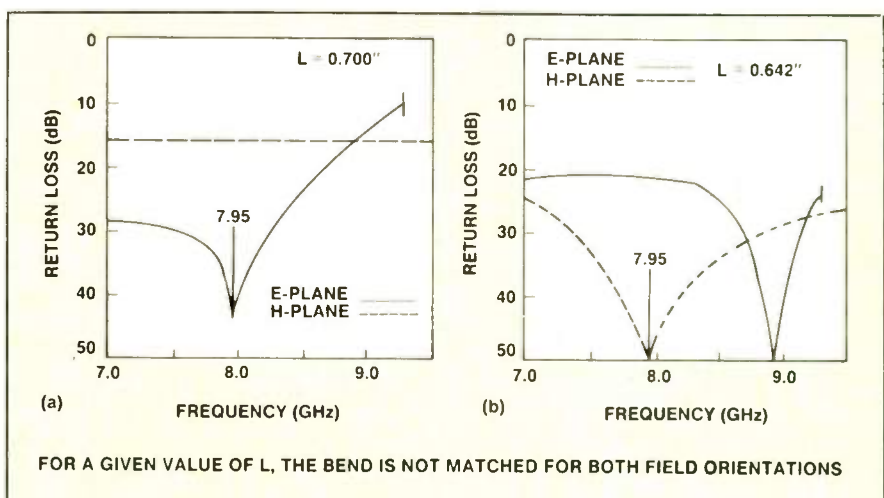

Fig. 2 Typical match characteristics for a mitered corner square waveguide for (a) E-plane bend and (b) H-plane bend.

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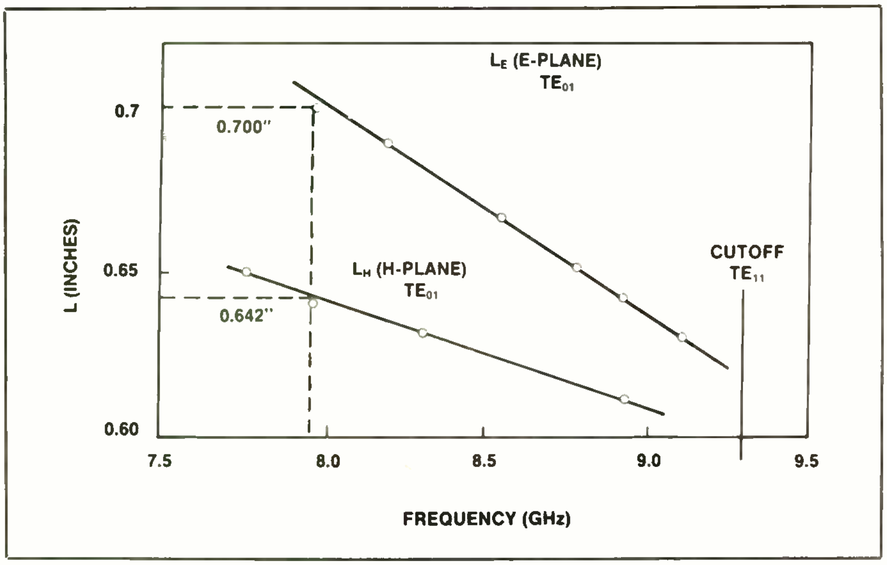

Fig. 3 Experimental design curves. Optimum miter size for matched square corner.

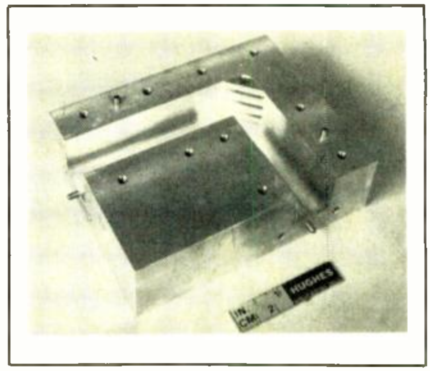

Fig. 5 Open test fixture showing a polarized mitered corner with a three-ridge reflector.

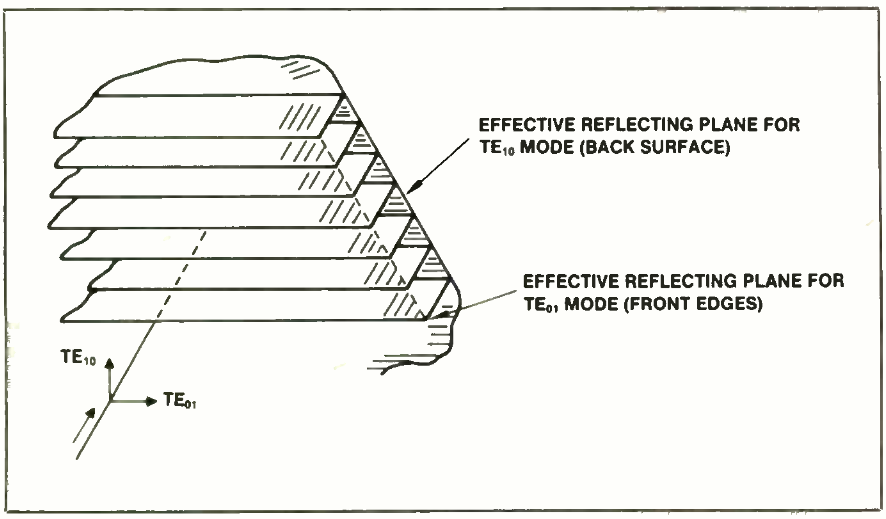

Fig. 4 Polarized reflecting surface.

dicular to the ridges, will be little influenced by the ridges, and the effective shorting plane will be approximately the original flatplate surface.

This produces an effect in which L_E for the TE_{01} mode is larger than L_H for the TE_{10} mode. The experimentally established values for L in Figure 3 can then be used to design the reflecting surface.

A photograph of the corner with the top removed is shown in Figure 5. Three- and seven-ridge reflecting corners, as shown in Figure 6, were used to optimize input impedance match for both polarizations. Figures 7 and 8 show the impedance match for three-ridge and seven-ridge polarized reflecting corners, respectively. The fact that these L values do not match those of Figure 3 at 7.95 GHz implies that there is a slight amount of mode interaction, requiring minimal design iteration.

Interaction between the inner and outer surfaces for the seven-ridge corner was less than that for the three-ridge corner, as expected. (See Table 1.)

Experimental Results

A corner designed for 7.95 GHz showed an SWR less than 1.05 for both E-plane and H-plane operation over a band of approximately 1.0 GHz. Cross-polarization isolation was typically 30 dB across the 7 to 9.6 GHz band.

Conclusion

A technique that provides a design for dual-mode operation of a square waveguide through a 90° corner has been demonstrated. This approach is applicable to lesser bends as well, allowing the use of square waveguide for dual-mode operation in complex waveguide runs. ■

[Continued on page 156]

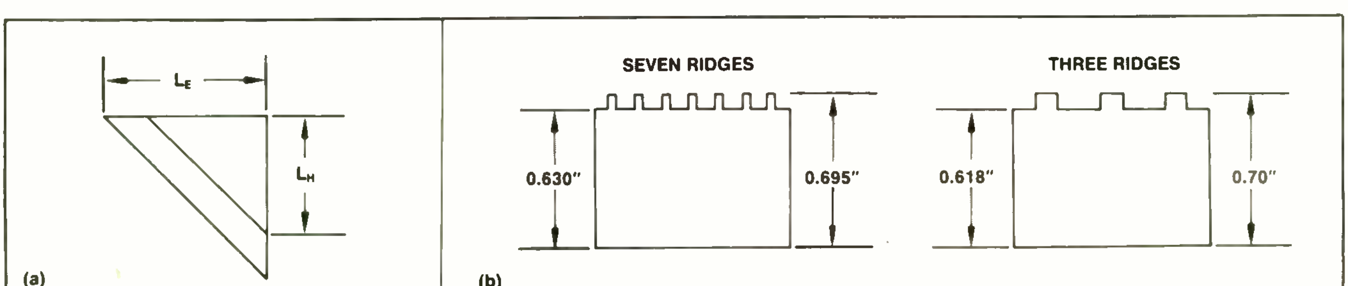

Fig. 6 Ridge dimensions: (a) top view; (b) side view.

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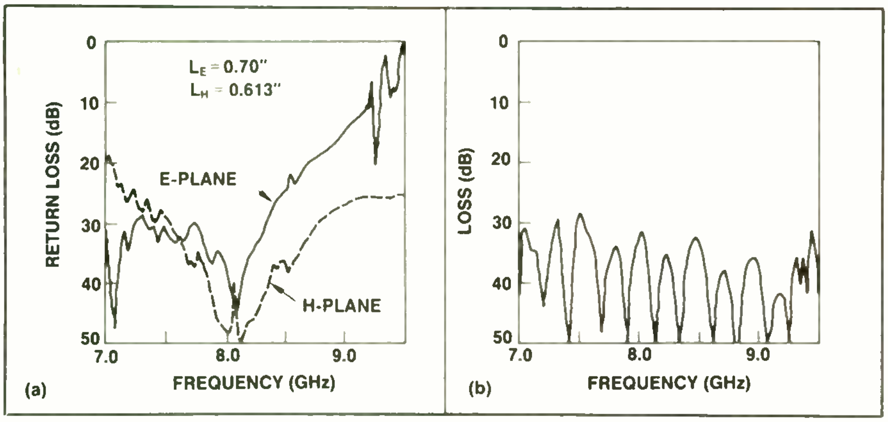

Fig. 7 Performance of the matched, dual mode square waveguide corner using three-ridge polarized miter: (a) input match; (b) cross-polarization isolation.

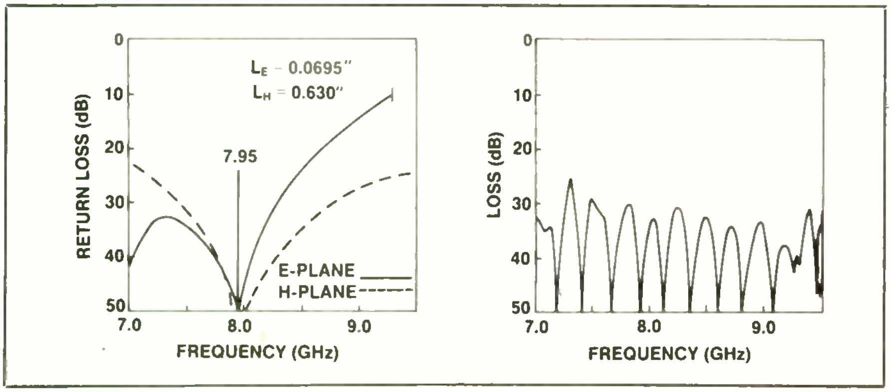

Fig. 8 Performance of the matched, dual mode square waveguide corner using seven-ridge polarized miter: (a) input match; (b) cross-polarization isolation.

INNER AND OUTER SURFACE INTERACTION

	L_E	L_H	Remark
Design value	0.7"	.0642"	from design curves
7-Ridge	0.695"	0.63"	less interaction
3-Ridge	0.7"	0.613"	more interaction

Pyong K. Park *received the BSEE degree from In-Ha University, Korea, the MSEE degree from Yon Sei University, Korea and the PhDEE degree from UCLA.*

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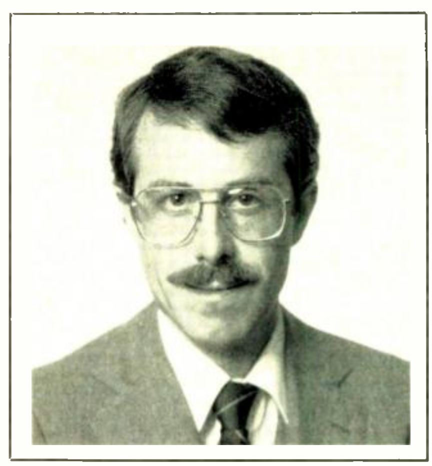
He spent two years in the US Army Security Agency and, upon discharge from the service in 1962, remained with the Agency in a civilian capacity to become a staff engineer with its European Headquarters in Frankfurt, West Germany. During this period, he received a letter of commendation from the assistant secretary of defense, Eugene G. Fubini, for outstanding contributions to the Army Security Agency.

He returned to school in 1965 and, until 1970, was engaged in the theoretical and experimental analysis and design of microwave circuitry. This led into waveguide equivalent circuit modeling, which was the basis for his thesis work. Upon completion of the doctoral program at the University of Michigan, he went to work for Hughes Aircraft Co., Culver City, CA in the Radar Systems Group. In 1977 he transferred to the Hughes Missile Systems Group in Canoga Park. He is still at Canoga Park as a chief scientist, RF Sensors and Subsystems Laboratory, continuing his work in R&D of microwave circuits.

He taught a microwave theory course at Loyola University of Los Angeles, CA and has been a guest lecturer at the University of California, Los Angeles and at the University of Michigan, Ann Arbor. He directed the research on five Masters theses at UCLA and one at California State University at Northridge. He also has published 18 papers in the field of microwave circuits covering a wide variety of subjects. He has three patents and four pending patents.

Eisenhart is a member of IEEE Microwave Theory and Techniques Society and Antennas and Propagation Society and the honoraries Sigma Xi, Tau Beta Pi, and Eta Kappa Nu. He is listed in Who's Who in the West *and* Who's Who in Technology Today.



Steven E. Bradshaw *received the BSEE degree from The Pennsylvania State University in 1982.*

He is presently employed at Hughes Aircraft Co.'s Missile System Group, where his primary responsibilities are in antenna design and analysis. He is a member of the Antennas and Propagation Society of the IEEE.

Design and Performance of a Wideband, Multilayer Feed Network*

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MIT Lincoln Laboratory
Lexington, MA

This paper describes the design and construction of a novel, high performance, broadband, stripline, 8 x 8-port feed network. Hybrid ring couplers with improved performance are used as building blocks. The division of the circuit into two stackable even-mode and odd-mode subcircuits has greatly simplified the realization and testing of the feed network. Experimental results for various phase modes are presented.

Introduction

Several microwave applications, such as electronically scanned antenna arrays, phased array receivers,[1] communication systems to and from mobile platforms,[2] and directionfinding systems[3] require the use of a microwave or mm-wave feed network. A typical feed network (or beamforming network) usually is composed of 2N ports: N input ports and an equal number of output ports or radiating elements. A signal introduced at one input port produces a specific set of excitations at the output ports, while a signal introduced at another input port results in a different set of excitations. Therefore, each input port of the feeding network is capable of exciting a separate orthogonal spatial or phase mode.

This work was sponsored by the Department of the Air Force. The views expressed are those of the author and do not reflect the official policy or position of the US Government.

*Invited paper.

In general, beamforming networks can be put to use for receiving purposes and/or transmitting purposes. For example, when a receiving antenna array is excited by an incident plane wave, the beamforming network can be used to analyze this excitation and hence expresses it in terms of its basic Fourier components or phase modes.[3] Indeed, by observing the relative phases of the excited phase modes, one can determine the angle of incidence of the excited plane wave and, thus, directionfinding can be carried out. In addition, one may use the beamforming network to feed a number of antenna elements and thereby synthesize a specific farfield pattern. This normally is done by exciting the network input ports with the appropriate values of amplitude and phase.

In this paper, the design and construction methodology of a wideband beamforming network will be reported. Many of the cogent issues surrounding the building block selection and design, the hardware realization and the testing of this wideband compact RF circuit are addressed.

Design and Description of the Feed Network

The first step toward the design of any beamforming network is the selection of the basic element or the basic building block. The choice of a particular circuit architecture normally is governed by one's desire to arrive at the simplest design possible (with the minimum number of components) and yet attain a wide bandwidth peformance. Frequently, beamforming networks are found to consist of phase shifting elements in combination with 3 dB directional couplers. However, in certain cases it is possible to design wideband beamforming networks without using separate wideband constant phase shifting elements. In such cases, the bandwidth limitation of the beamforming network is entirely due to the bandwidth limitation of the individual hybrid couplers being used.

The microwave feed network under consideration has 16 ports — eight input ports and eight output ports — and is matched at all ports. The desired performance is outlined in Table 1. This table describes

[Continued on page 160]

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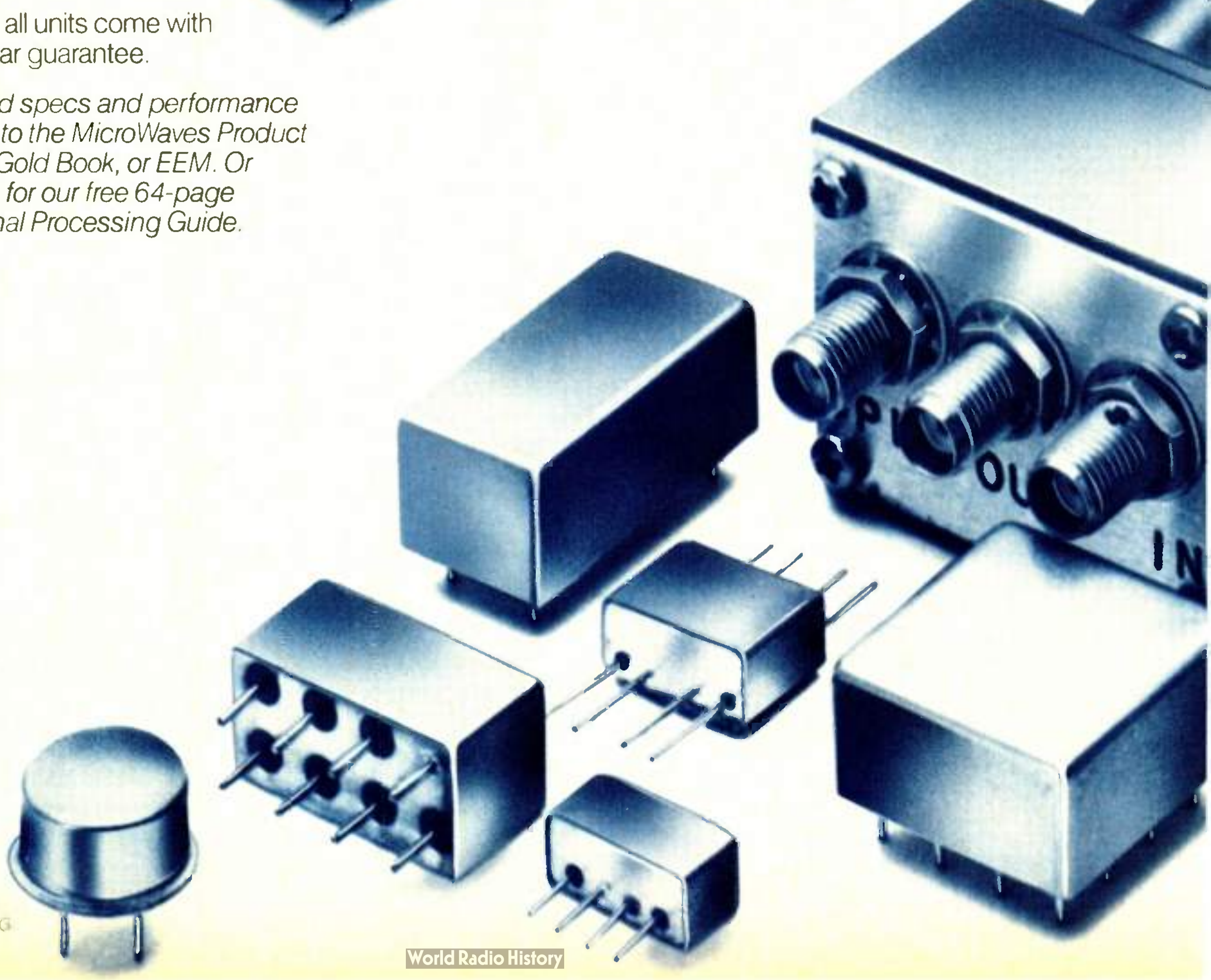

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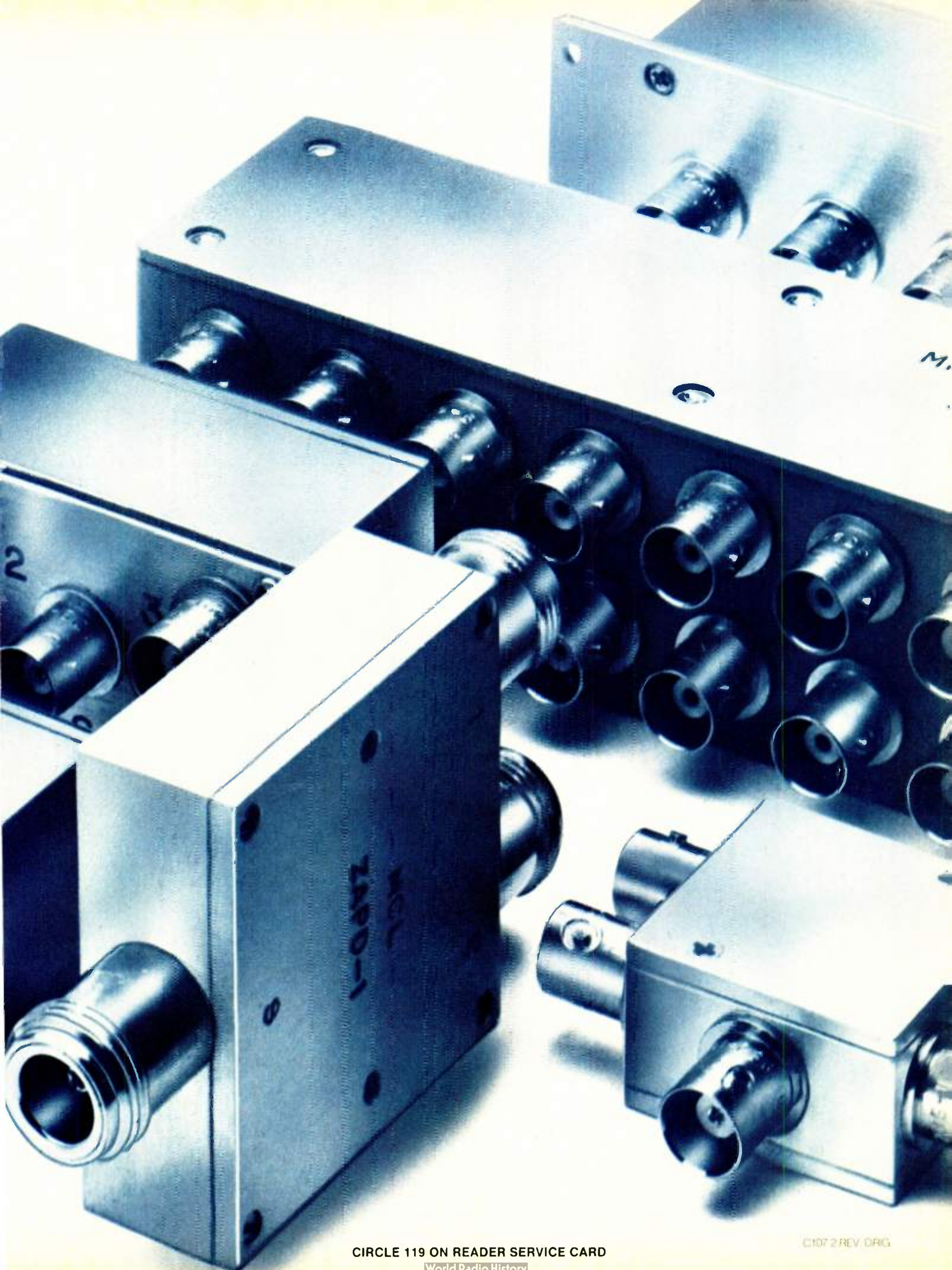
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TABLE I
THEORETICAL PERFORMANCE MATRIX

Mode Designation	Output Port Excitations 1	2	3	4	5	6	7	8
0	+1	+1	+1	+1	+1	+1	+1	+1
+1	+1	$+\frac{1}{\sqrt{2}}$	0	$-\frac{1}{\sqrt{2}}$	-1	$-\frac{1}{\sqrt{2}}$	0	$+\frac{1}{\sqrt{2}}$
-1	0	$+\frac{1}{\sqrt{2}}$	+1	$+\frac{1}{\sqrt{2}}$	0	$-\frac{1}{\sqrt{2}}$	-1	$-\frac{1}{\sqrt{2}}$
+2	+1	0	-1	0	+1	0	-1	0
-2	0	+1	0	-1	0	+1	0	-1
+3	+1	$-\frac{1}{\sqrt{2}}$	0	$+\frac{1}{\sqrt{2}}$	-1	$+\frac{1}{\sqrt{2}}$	0	$-\frac{1}{\sqrt{2}}$
-3	0	$-\frac{1}{\sqrt{2}}$	+1	$-\frac{1}{\sqrt{2}}$	0	$+\frac{1}{\sqrt{2}}$	-1	$+\frac{1}{\sqrt{2}}$
+4	$+\frac{1}{\sqrt{2}}$	$-\frac{1}{\sqrt{2}}$	$+\frac{1}{\sqrt{2}}$	$-\frac{1}{\sqrt{2}}$	$+\frac{1}{\sqrt{2}}$	$-\frac{1}{\sqrt{2}}$	$+\frac{1}{\sqrt{2}}$	$-\frac{1}{\sqrt{2}}$

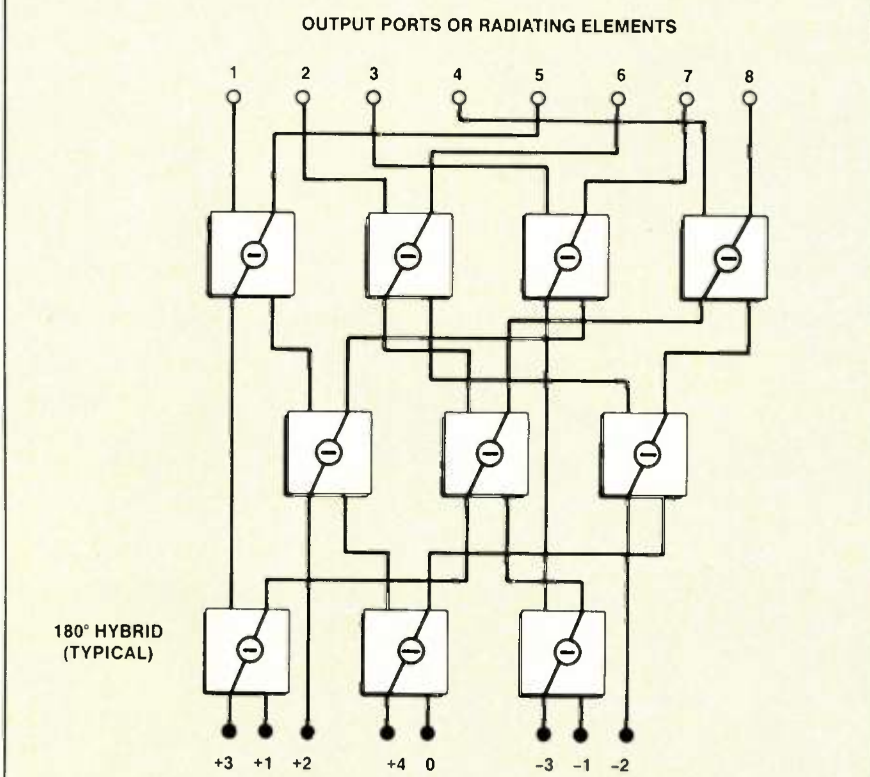

Fig. 1 Schematic diagram of the Butler-type 8 x 8 feed network.

the resulting excitation at each output port for all of the desired eight phase modes.

Figure 1 shows a schematic of the 8 x 8 port matrix, with each block representing a 180° hybrid coupler, which is yet to be selected. Conventional hybrid ring couplers have a narrow bandwidth and hence could not be chosen. However, other 180°, 3 dB directional coupler configurations were found and, therefore, were considered.[4-6]

The classical rat-race coupler and three of its variants are shown in Figure 2. When compared to the

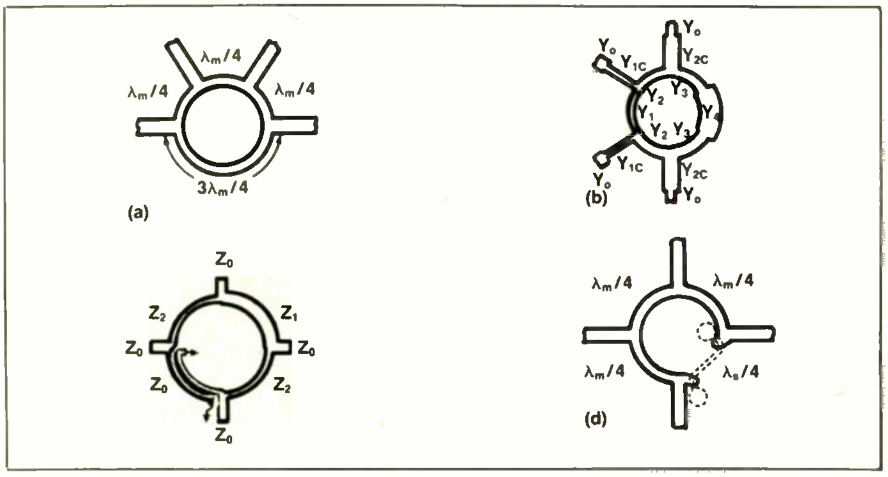

Fig. 2 Schematics of the conventional rat-race coupler and three of its broadband variants: (a) conventional; (b) unequal admittance; (c) reverse-phase; (d) slot line.

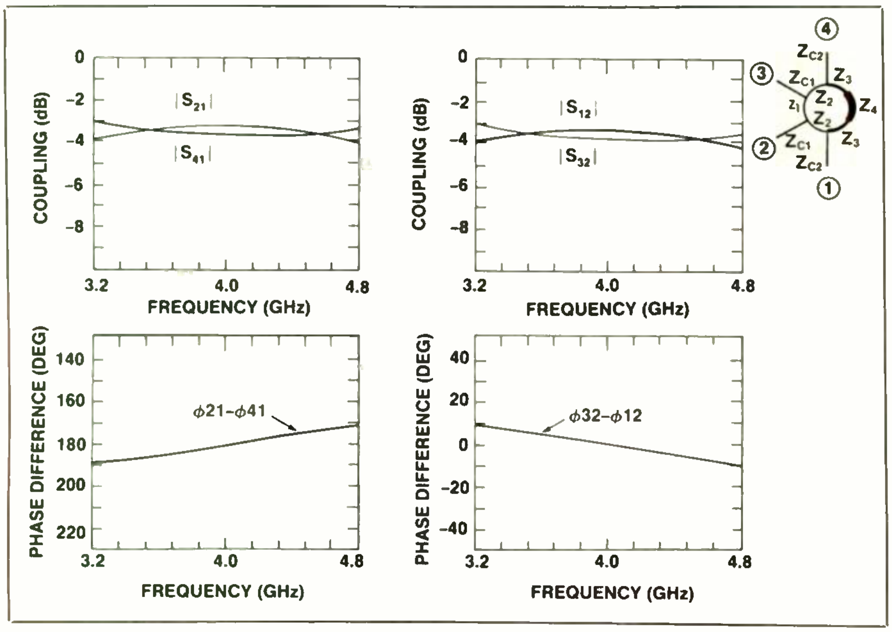

Fig. 3 Measured characteristics of the unequal admittance hybrid.

conventional hybrid ring coupler, these three alternative designs are less frequency-dependent and hence offer wider bandwidth. A detailed examination of their respective properties led to the selection of the first configuration. This is because the unequal admittance configuration has the advantage of a unique and simple geometry that lends itself to stackable topology, in addition to being easy to design and fabricate.

This particular version of the hybrid ring coupler, whose characteristics can be analyzed using any of the commercially available CAD programs, has been studied in detail by Kim et al[5] and, more recently, by Agrawal et al.[6] Indeed, reference 5 has shown that a broader bandwidth is achievable by dividing the three quarter-wave equal admittance section of the conventional hybrid ring geometry into three unequal admittance quarter-wave sections. This is done in addition to adding a quarter-wave transformer to each of the feeding ports.

An experimental stripline model of this improved hybrid ring geometry has been designed and constructed on RT/duroid 6006 substrate. The values of the characteristic impedances Z_1, Z_2, Z_3, Z_4, Z_{c1} and Z_{c2}, were chosen (by means of an optimization process) so that coupling, matching and isolation are within certain tolerance limits over the 3.2 to 4.8 GHz frequency range. The values of the imped-

[Continued on page 162]

ances were 72, 62, 43, 33, 53 and 43 ohms, respectively. the measured performance of this optimized design is illustrated in Figure 3. This measurement demonstrates that a useful bandwidth (in terms of ±0.3 dB amplitude imbalance and ±8° phase variation) of about 35 percent has been achieved. This improvement in performance has been achieved by: overcoupling S_{21} and S_{41}, and flattening the frequency response of S_{41} (relative to S_{41} of the conventional configuration).

Due to the stringent size requirement of our particular application, a planar topology could not be adopted. Instead, a substrate with a dielectric constant of $\epsilon_r=6$ (RT/duroid 6006,) and stackable topology were selected. The packaging constraints on the feed network dictated that the final circuit should be shielded and fit entirely within a 3.5 inch diameter cylindrical structure. The total volume was required to be 5.0 cu-inch or less.

A full-scale model was realized using a stripline design and adopting a multilayer topology. The feed circuit was divided into two subcircuits: even-mode and odd-mode. The even-mode subcircuit is composed of seven hybrid ring couplers

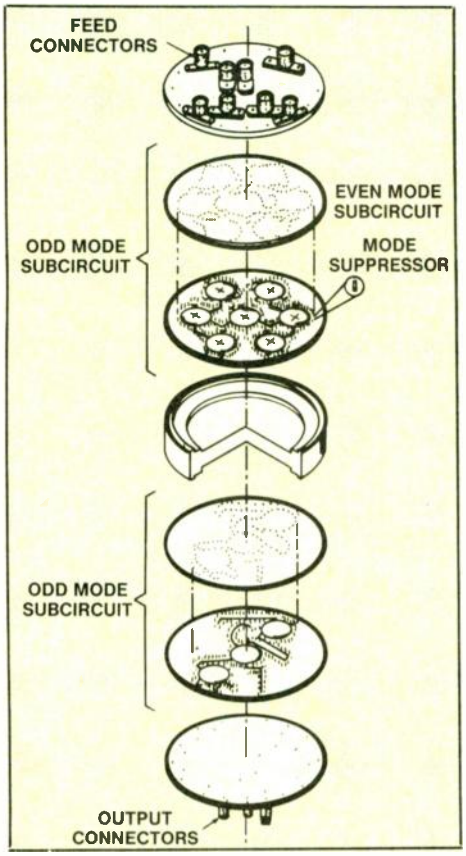

Fig. 4 Exploded view of the network configuration and assembly.

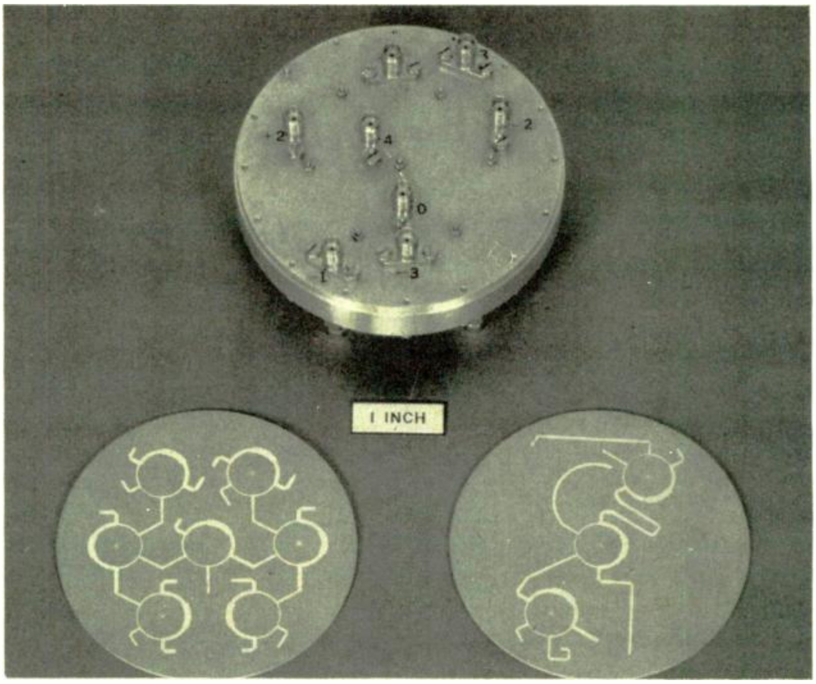

Fig. 5 The feed network.

capable of independently producing the four even-phase-modes m=0, ±2,+4. The second layer, or subcircuit, encompasses three hybrid ring couplers that, together with the first layer, produce the odd phase modes m=±1, ±3. Both layers are etched on separate soft copper-clad substrates, which are then housed in a cylindrical structure to yield the desired multilayer configuration.

An exploded view of the network assembly is outlined in Figure 4. The two layers are interconnected together by means of four short sections of 50 ohm coaxial lines running vertically through the conductive wall separating them. The lengths of the four coaxial interconnecting lines are chosen carefully so that the overall phase balance of the feed network is retained. Due to the presence of these vertical interconnections and, as a consequence of using broadwall launching, two problems arose and required attention.

The first problem was the spurious excitation of the planar mode. The excitation was caused by the abrupt change in the direction of propagation (from horizontal to vertical or vice versa). This problem was resolved by introducing mode suppressors (resilient low impedance, multipath wire mesh contact elements) into the circuit, particularly into areas with sudden transitions. Secondly, the deployment of the broadwall launching and the vertical interconnections has resulted in a slight degradation of the hybrid ring isolation. Proper mounting of the coaxial interconnections and the two-hole flange mount cable jack connectors has resulted in obtaining a minimum isolation of 17 dB per hybrid over the frequency band of interest. Figure 5 shows a photograph of the layout of both layers, as well as the cylindrical housing structure.

This approach of dividing the overall feed circuit into two subcircuits greatly simplified the development task. It permitted separate construction and testing of each subcircuit. It was then a simple matter to subsequently connect the two subassemblies together to characterize the whole feed network.

Experimental Results

The characterization of the constructed beamforming network was performed through measurement of its scattering parameters. An HP 8510 automatic network analyzer was used to conduct the measurements. The variation of the scattering parameters (amplitude and phase) was measured over the frequency range from 3.2 to 4.8 GHz. Curves describing the amplitude variation with frequency for the m=0, +1, +2 and +3 modes are given in Figure 6 through 9.

Figure 6 depicts the network performance when the m=0 mode is excited. For this mode of operation, all of the eight outputs are excited and the network has an insertion loss of about 0.75 dB. The amplitude and phase imbalances are, respectively, within ±0.5 dB and $\pm20°$ over 30 percent of the band. The m=2

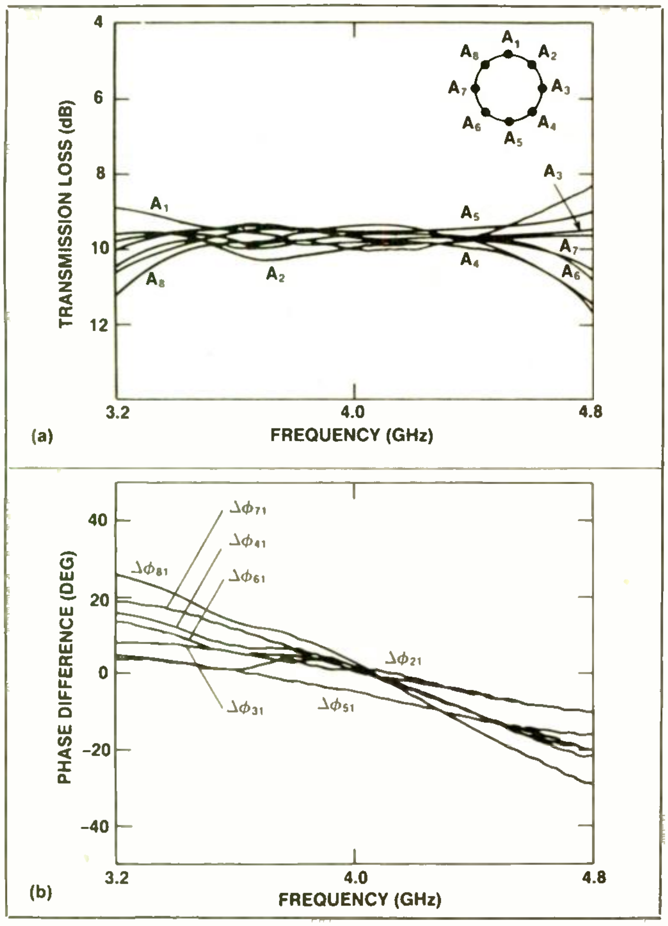

Fig. 6 Measured amplitude (a) and phase (b) characteristics of the m = 0 mode. Phase is relative to port A_1.

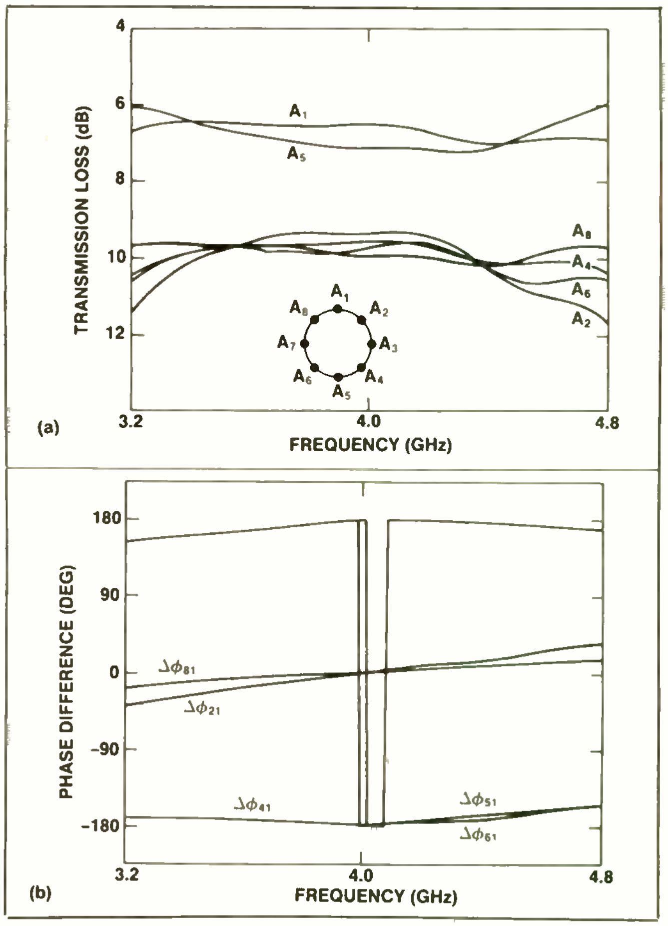

Fig. 7 Measured amplitude (a) and phase (b) characteristics of the m = 1 mode. Phase is relative to port A_1.

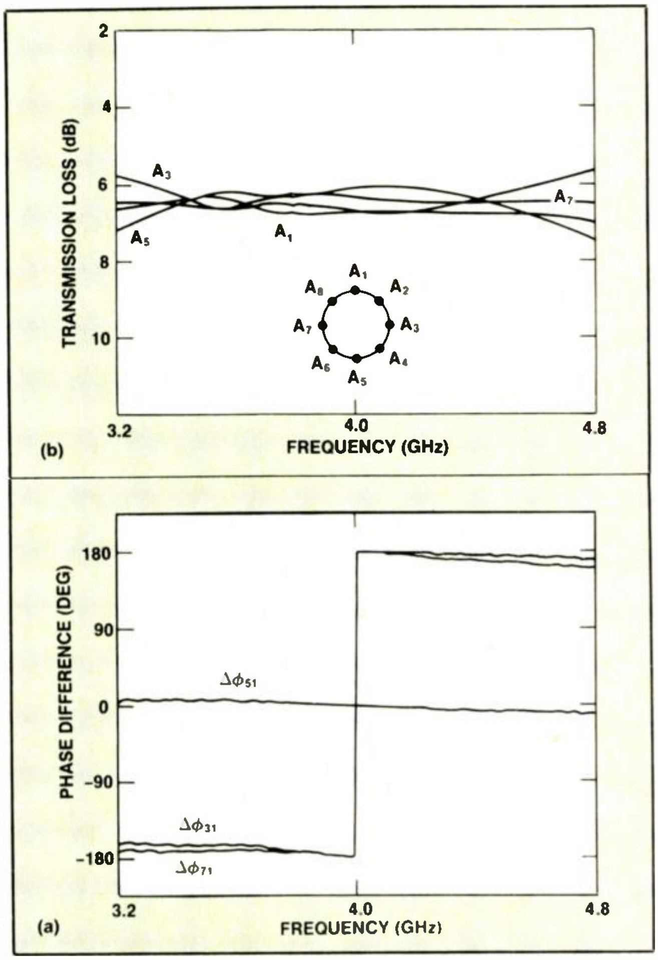

Fig. 8 Measured amplitude (a) and phase (b) characteristics of the m = +2 mode. Phase is relative to port A_1.

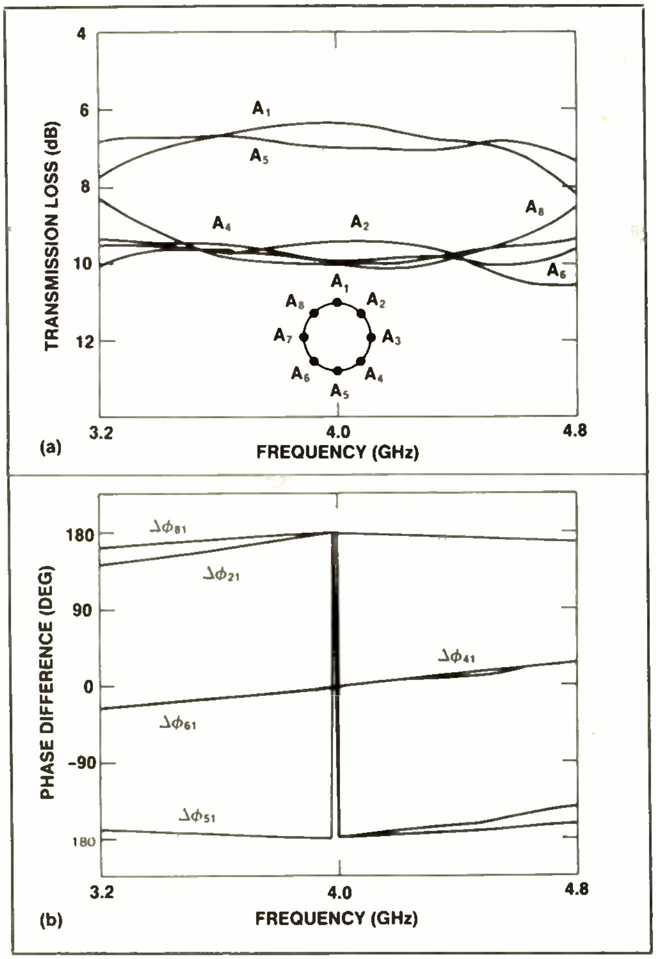

Fig. 9 Measured amplitude (a) and phase (b) characteristics of the m = +3 mode. Phase is relative to port A_1.

mode excites four radiating elements and is depicted in Figure 8. Its insertion loss is about 0.5 dB while the amplitude and phase imbalances are less than ±0.5 dB and ±20°, respectively, over 35 percent of the band.

Figures 7 and 9 depict the performance of the m=1 and 3 modes, which excite only six of the eight radiating elements. The amplitude and phase imbalances are within ±0.5 dB and ±20°, respectively over 32 percent of the band. Return loss of better than 15 dB has been realized and mode-to-mode isolation of better than 17 dB was obtained over the full 40 percent bandwidth.

Conclusion

A broadband, multilayer feed network has been developed. The wideband performance of the basic component has led to an improvement in the performance of the overall circuit. The division of the circuit into two subcircuits, even-mode and odd-mode, simplified the construction and testing tasks. ■

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From 1979 to 1984 he worked as a research and teaching assistant at the University of Colorado, Boulder. Since 1984 he has been a member of the Countermeasures Technology Group at the Massachusetts Institute of Technology, Lincoln Laboratory. He has worked on microstrip line discontinuities, wideband dielectric directional couplers and dielectric image lines. He is currently interested in computer-aided design of microwave and mm-wave wideband planar circuits.

Stopband Filter Resonance Frequency Shift Due to an Uncoupled Length of Resonator

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Broadband, high cutoff frequency stopband filters require narrow cavities that restrict resonator placement. Normally, resonators are placed as shown in Figure 1a for low cutoff frequency filters. But for high cutoff frequency stopband filters (narrow cavity), the resonators are arranged as shown in Figure 1b. This arrangement requires an additional uncoupled length at the electrically shorted (grounded) end of the coupled resonator. This additional length lowers the resonanace frequency of the resonator to a larger extent at high frequencies and a smaller extent at low frequencies. (In the case of cavity filters, this length is equivalent to the gap between the coupled resonator and ground, which ultimately is determined by even- and odd-mode impedances of the resonator. See Figure 1c.

The coupled length of the resonator is different than a quarter wavelength at the resonance frequency. Since the additional resonator length lowers the resonance frequency, the coupled resonator should be a quarter wavelength long at the higher frequency. In this article, an attempt has been made to calculate this quarter wavelength frequency and, hence, the length of coupled resonators, in order to obtain resonance at the required frequency for a given uncoupled length of a resonator.

A typical parallel-coupled configuration used for the design of a stopband filter is shown in Figure 2. Ports 1A and 1B are input and output ports, respectively. Port 2A is loaded with additional uncoupled shorted transmission line of characteristic impedance Z_0 and length L, and port 2B is an open circuit. The resonators are a quarter wavelength long at frequency f_1 (which is to be determined) and the required resonance frequency f_0 (which is known). The length L can be determined by knowing the even- and odd-mode impedance as well as the cavity dimensions of the filter/resonator, i.e. L is known. The load presented by the shorted transmission line is given by Z_L:

$$Z_L = j\tan(\beta L) \qquad (1)$$

where $\beta = 2\pi/\lambda$
$\lambda = c/(\sqrt{\epsilon_r}\, f)$
c = velocity of light
ϵ_r = relative dielectric constant of medium and
f = frequency of operation.

The [Z] matrix of a transmission line of length L and characteristic impedance Z_0 is given by:[1]

$$[Z] = \begin{bmatrix} Z_0/S & Z_0\sqrt{1-S^2}/S \\ Z_0\sqrt{1-S^2}/S & Z_0/S \end{bmatrix} \qquad (2)$$

where frequency variable S is defined as[2]

$S = j\tan(\beta L)$ for lossless line
$= \tanh(rL)$ for lossy line
and $r = \alpha + j(\beta L)$

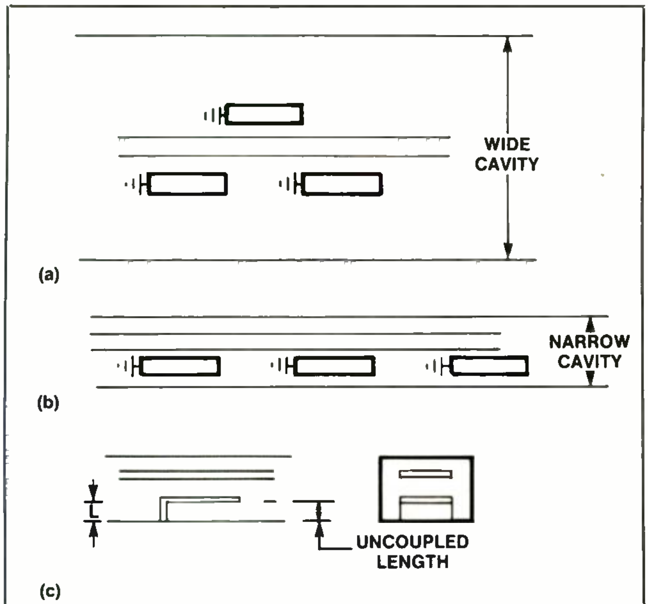

Fig. 1 Stopband filter configuration: (a) wide cavity; (b) narrow cavity; (c) practical example of a narrow cavity stopband filter.

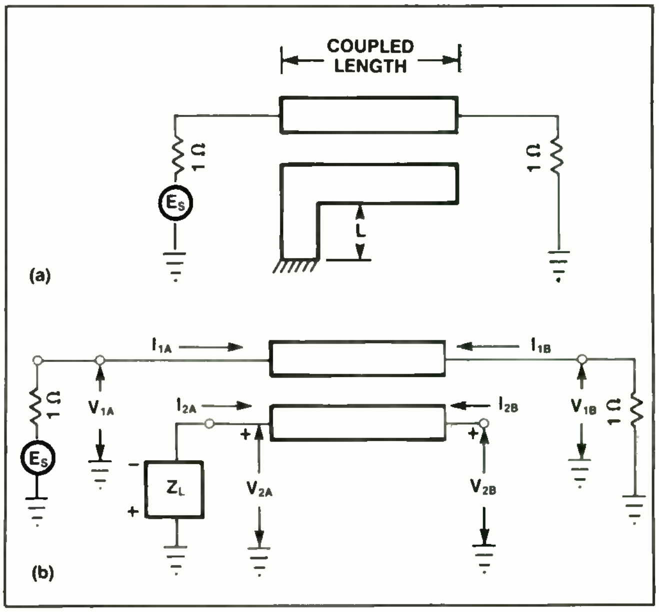

Fig. 2 (a) Typical terminated resonator and (b) equivalent circuit, Z_L = tan (L).

where α = line loss in nepers/unit length
= 0 for lossless line.

For simplicity, we will consider the case of symmetrical coupled lines. If Z_{0e} and Z_{0o} are even- and odd-mode impedances of each line of Figure 2, then the [Z] matrix for symmetrical coupled lines is given by:

$$[Z] = \begin{bmatrix} A/S & B/S & A\sqrt{1-S^2}/ & B\sqrt{1-S^2}/S \\ B/S & A/S & B\sqrt{1-S^2} & A\sqrt{1-S^2}/S \\ A\sqrt{1-S^2}/S & B\sqrt{1-S^2}/S & A/S & B/S \\ B\sqrt{1-S^2}/S & A\sqrt{1-S^2}/S & B/S & A/S \end{bmatrix} \quad (3)$$

where $A = (Z_{0e} + Z_{0o})/2$ and $B = (Z_{0e} - Z_{0o})/2$.

From Equation 3,

$$V_{1A} = (A/S)\, I_{1A} + (B/S)\, I_{2A} + (A\sqrt{1-S^2/S})\, I_{1B} + (B\sqrt{1-S^2}/S)\, I_{2B} \quad (4)$$

$$V_{2A} = (B/S)\, I_{1A} + (A/S)\, I_{2A} + (B\sqrt{1-S^2}/S)\, I_{1B} + (A\sqrt{1-S^2}/S)\, I_{2B} \quad (5)$$

$$V_{1B} = (A\sqrt{1-S^2}/S)\, I_{1A} + (B\sqrt{1-S^2}/S)\, I_{2A} + (A/S)\, I_{1B} + (B/S)\, I_{2B} \quad (6)$$

$$V_{2B} = (B\sqrt{1-S^2}/S)\, I_{1A} + (A\sqrt{1-S^2}/S)\, I_{2A} + (B/S)\, I_{1B} + (A/S)\, I_{2B}. \quad (7)$$

Boundary conditions for the coupled lines shown in Figure 3 are as follows:

$$V_{2A} = -I_{2A} Z_L \quad (8)$$

$$I_{2B} = 0. \quad (9)$$

Using the boundary conditions of Equations 8 and 9 and Equations 4, 5, 6 and 7, the [Z] matrix for a two-port network (input port V_{1A} and output port V_{1B}) is given by (see appendix A):

$$\begin{bmatrix} V_{IA} \\ V_{IB} \end{bmatrix} = \begin{bmatrix} \dfrac{A}{S} - \dfrac{B^2}{Z_L S+A} & \dfrac{\sqrt{1-S^2}}{S}\left(A - \dfrac{B^2}{Z_L S+A}\right) \\ \dfrac{\sqrt{1-S^2}}{S}\left(A - \dfrac{B^2}{Z_L S+A}\right) & \dfrac{A}{S} - \dfrac{B^2(1-S^2)}{Z_L S+A} \end{bmatrix} X \begin{bmatrix} I_{1A} \\ I_{1B} \end{bmatrix} \quad (10a)$$

Equation 10a can be rewritten in [Z] matrix form as follows:

$$\begin{bmatrix} V_{1A} \\ V_{1B} \end{bmatrix} = \begin{bmatrix} Z_{11} & Z_{12} \\ Z_{21} & Z_{22} \end{bmatrix} X \begin{bmatrix} I_{1A} \\ I_{1B} \end{bmatrix} \quad (10b)$$

The transmission coefficient S_{21} for [Z] matrix is given by:

$$S_{21} = \frac{2Z_{21}}{1 + Z_{11} + Z_{22} + Z_{11} Z_{22} - Z_{21} Z_{12}} \quad (11)$$

At resonance frequency f_0, $S_{21} = 0$, which gives $Z_{21} = 0$

or, from Equation 10,

$$\frac{\sqrt{1-S_2}}{S}\left(A - \frac{B_2}{Z_L S+A}\right) = 0$$

or

$$\sqrt{1-S^2}/S = 0 \quad (12)$$

or

$$A - [B^2/(Z_L S + A)] = 0. \quad (13)$$

Equation 12 gives sec (β1) = 0, which is not true.

Equation 13 gives

$$A = B^2/(Z_L S + A)$$

or

$$AZ_L S + A^2 = B^2$$

or,

using Equation 1 with the definitions following Equation 2,

$$[jZ_c \tan(\beta L)]\,[j\tan(\beta L)] + A^2 = B^2$$

or

$$(A^2 - B^2/[Z_L A \tan(\beta L)] = \tan(\beta L). \quad (14)$$

Since resonators are a quarter wavelength long at frequency f_1 and resonance frequency is f_0, tan (βL) is given by:

$$\tan(\beta L) = \tan\left(\frac{2\pi}{\lambda_0}\,\frac{\lambda L}{4}\right) = \tan(\pi f_0/(2 f_1)) \quad \partial f = f_0. \quad (15)$$

Also,

$$\tan(\beta L) = \tan(2\pi L/\lambda_0) = \tan(2\pi\sqrt{\epsilon_r}\, f_0 L/c \quad \partial f = f_0. \quad (16)$$

Substituting the "where" Equation that follows Equation 3, Equation 15, and Equation 16 into Equation 14,

$$f_1 = f_0\,\pi/(2Y). \quad (17)$$

where

$Y = \tan^{-1}(X)$ and

$X = 2 Z_{0e} Z_{0o} / [Z_L (Z_{0e} + Z_{0o}) \tan(2\pi\sqrt{\epsilon_r}\, f_0 L/c)]$.

Knowing the resonance frequency f_0, the load impedance Z_L and the load length L, f_1 can be computed from Equation 17 and coupled sections are a quarter wavelength long at f_1.

For multisection stopband filters, the gap L is different for each resonator and, hence, each resonator length has to be computed using Equation 17 in order to obtain the correct resonance frequency of a complete filter.

In the case of asymmetrical coupled lines, let the even- and odd-mode impedances of two lines (A) and (B) be Z_{0eA}, Z_{0oA} and Z_{0eB}, Z_{0oB}, respectively. A,B,C and D are defined;

$$A = (Z_{0eA} + Z_{0oA})/2 \qquad B = (Z_{0eA} - Z_{0oA})/2$$
$$C = (Z_{0eB} + Z_{0oB})/2 \qquad D = (Z_{0eB} - Z_{0oB})/2.$$

Equation 17 is true for an asymmetric case also, but X, defined just after Equation 17, is replaced as follows:

$$X = [(AC) - (BD)]/[AZ_L \tan(2\pi\sqrt{\epsilon_r}\, f_0 L/c)].$$

The analysis of a five-section filter was performed by considering coupled lengths a quarter wavelength long at f_0 and f_1 (using Equation 17). The results are plotted in Figure 3. Comparing these plots, not only is the resonance frequency lowered, but stop- as well as pass-bandwidths are wider when resonators are

[Continued on page 168]

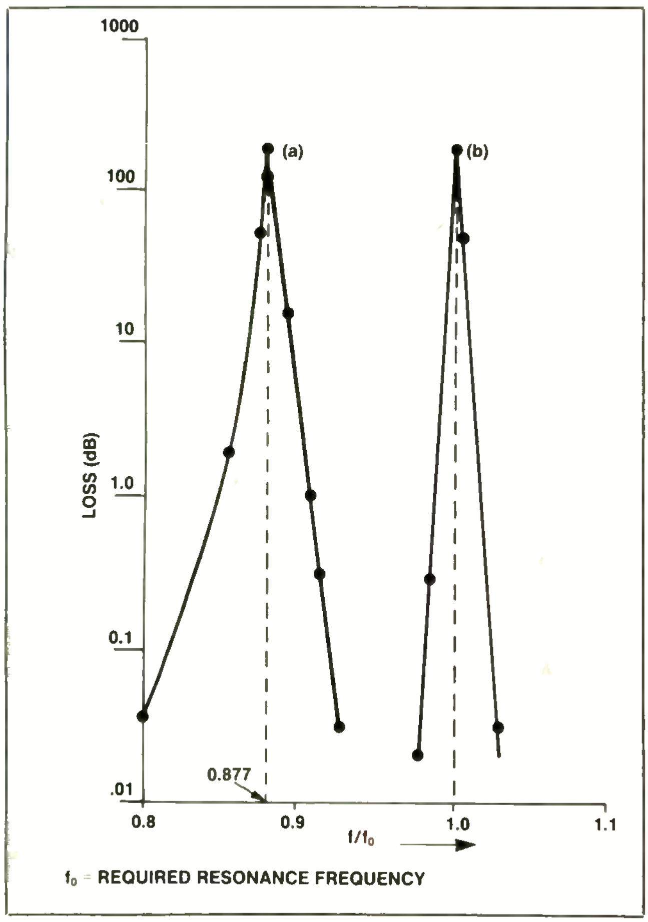

Fig. 3 Five-section filter response. Coupled sections are quarter wavelength long at: (a) resonance frequency; (b) frequency computed from Equation 17.

[From page 167]

90° long at f_0. While using Equation 17, the resonance occurs at the required frequency and, at the same time, filter parameters are preserved as well. The frequency response of the filter is not quite symmetrical because of small errors in computing the gap-length 'L', and also because of the frequency dependence of the resonator length.

Close agreements between practical measurements and theoretical calculation for a single resonator can be seen in Figure 4. For this section, Z_{0e} and Z_{0o} are 62.76 and 40.79 ohms respectively. The physical structure used is shown in Figure 1c with cavity dimension .250 in. x .250 in. Broadside coupled configuration was used. The complete nomograms relating physical dimensions and Z_{0e}, Z_{0o} were developed. For this particular case, the dimension L (gap) was .039 in. The required resonance frequency was 9.8 GHz. This requires that, using Equation 17 for an air dielectric, coupled length be one quarter wavelength long at 11.277 GHz (0.262) in.).

Figure 4a shows the response when resonance

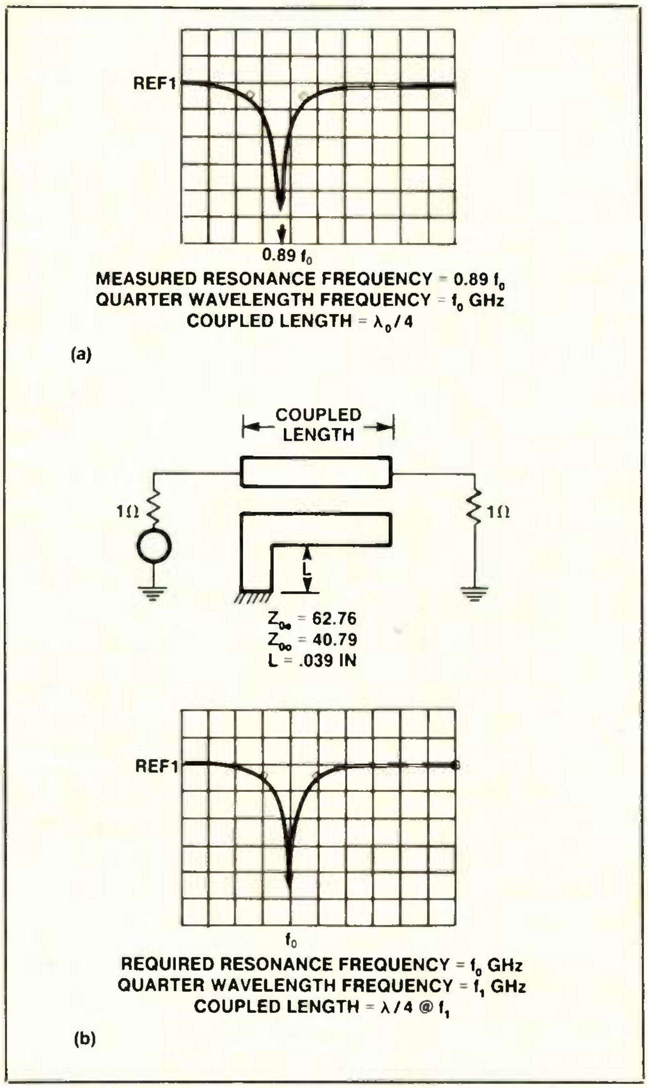

Fig. 4 Performance: (a) measured; (b) required.

frequency equals quarter wavelength frequency of 9.8 GHz [i.e. coupled length is one quarter wavelength long at 9.8 GHz (.301 in.)] The resonance occurs at

8.705 GHz i.e. resonance occurs at a lower frequency.

Figure 4b shows response when the resonance frequency is 9.8 GHz and coupled sections are one quarter wavelength long at 11.277 GHz (.262 in.) and resonance occurs at a required resonance frequency of 9.8 GHz. It should be noted that calculated coupled length and practical physical length were exactly equal.

Conclusion

An uncoupled shorted length of a coupled resonator (Figure 2) lowers the resonance frequency. This can be corrected by using Equation 17, which provides the exact quarter wavelength frequency and length of the coupled section. The agreement between measured results and theoretical predictions is shown in Figure 4.

Appendix

Derivation of Equation 10a

Substituting Equations 8 and 9 into Equation 5 and solving for I_{2A}.

$$-Z_L I_{2A} = (B/S) I_{1A} + (A/S) I_{2A} + (B\sqrt{1-S^2}/S) I_{1B}$$

or

$$I_{2A} = -[(B/S) I_{1A} + (B\sqrt{1-S^2}/S) I_{1B}]/((A/S) + Z_L). \qquad (18)$$

Substituting Equation 18 into Equation 4 and solving for V_{1A}.

$$V_{1A} = [(A/S) - (B^2/(S(A+Z_L S)))] I_{1A} + [(A\sqrt{1-S^2}/S) - (B^2\sqrt{1-S^2}/(S(A+Z_L S)))] I_{1B}. \qquad (19)$$

Substituting Equation 19 into Equation 6 and solving for V_{1B}.

$$V_{1B} = [(A\sqrt{1-S^2}/S) - (B^2\sqrt{1-S^2}/(S(A+Z_L S)))] I_{1A} + [(A/S) - (B^2(1-S^2)/(S(A+Z_L S)))] I_{1B}. \qquad (20)$$

Equations 19 and 20 can be represented in matrix form as shown in Equation 10a. ■

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Synthesizer Switching Speed Measurements

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and
Michael R. Hagins
Wavetek San Diego Inc.
San Diego, CA

Fast switching frequency synthesizers are used in a number of applications requiring frequency agility. These include spread spectrum communications, frequency agile radars, frequency shift keying (FSK) modulation and ECM systems. They also are useful for increasing the efficiency of ATE systems requiring a large number of accurate stepped frequencies and for providing the reference signal in commercial magnetic resonance imaging (MRI) systems.

Measuring synthesizer switching speeds can be difficult. This article discusses a method that can be used for the measurement of fast switching synthesizers. It covers several definitions, measurement limitations, a typical test setup and a detailed test procedure.

Definitions

For the purposes of this discussion, switching speed is defined as the period of time from the initial change-frequency command to the time when the output frequency has settled to the new frequency within the specifications of the instrument. This time interval can be separated into two parts:

- Programming delay. This is the time interval from the change command to the start of the frequency change
- Switching transient. This is the transition time from old to new frequency. (The signal spectrum is adversely affected by this portion of the total switching time interval.)

With slower synthesizers, the programming delay time is small compared to the switching transient time and, therefore, has been traditionally ignored. For synthesizers capable of submicrosecond switching transients, programming delay cannot be ignored.

Delay and switching transient durations are measurable quantities and can be specified to within some phase or frequency error. The method discussed here uses a double-balanced mixer as a phase detector. In this case, the switching transient duration is the time during which the phase offset from the new settled condition exceeds the specified phase error. For many synthesizers, this phase error is specified to be 0.1 radian. Frequency error can be calculated from the slope of the phase curve since frequency is the mathematical derivative of phase:

$$\Delta f = \frac{\Delta\phi}{2\pi\Delta t}.$$

Measurement Setup

The test setup in Figures 1a and 1b and the described procedure used to measure frequency switching speed may be used as a general guide for testing synthesizers with fast switching capabilities.

In Figure 1, the reference oscillator provides an accurate, stable signal to drive both the synthesizer unit under test (UUT) and the reference synthesizer. The reference synthesizer does not require a fast switching capability and needs only to cover the frequency ranges involved in the test. It must, however, be phase locked to the same reference as the UUT. It is very helpful if it has a very fine resolution, such as 0.1 Hz or smaller, as discussed later.

Outputs of the UUT and reference synthesizer are attached to the LO and RF inputs, respectively, of the double-balanced mixer. The mixer acts as a phase detector in this setup to determine the phase difference between the outputs of the two synthesizers. The output of the mixer (the IF port) is routed through the lowpass filter that attenuates all but the difference frequency. The resulting signal is then observed on the oscilloscope as a DC voltage level when the UUT and reference synthesizer are set to the same frequency. When the UUT and reference synthesizer are set to different frequencies, the oscilloscope presents the sine waveform of the difference frequency.

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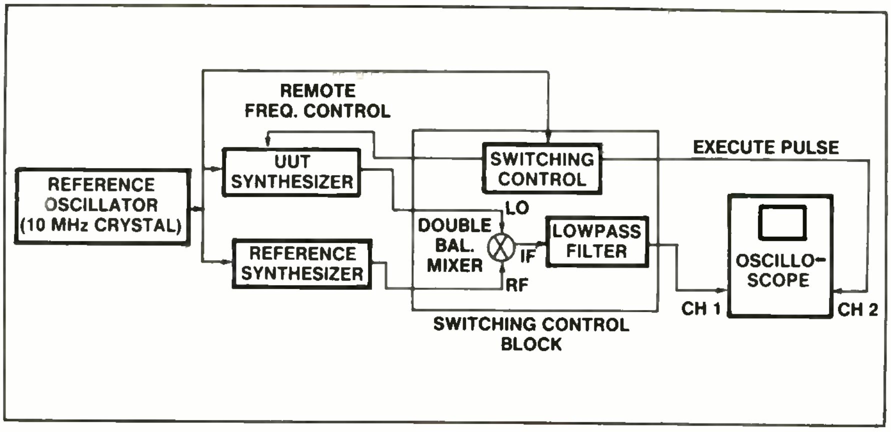

Fig. 1a Switching speed measurement setup block diagram.

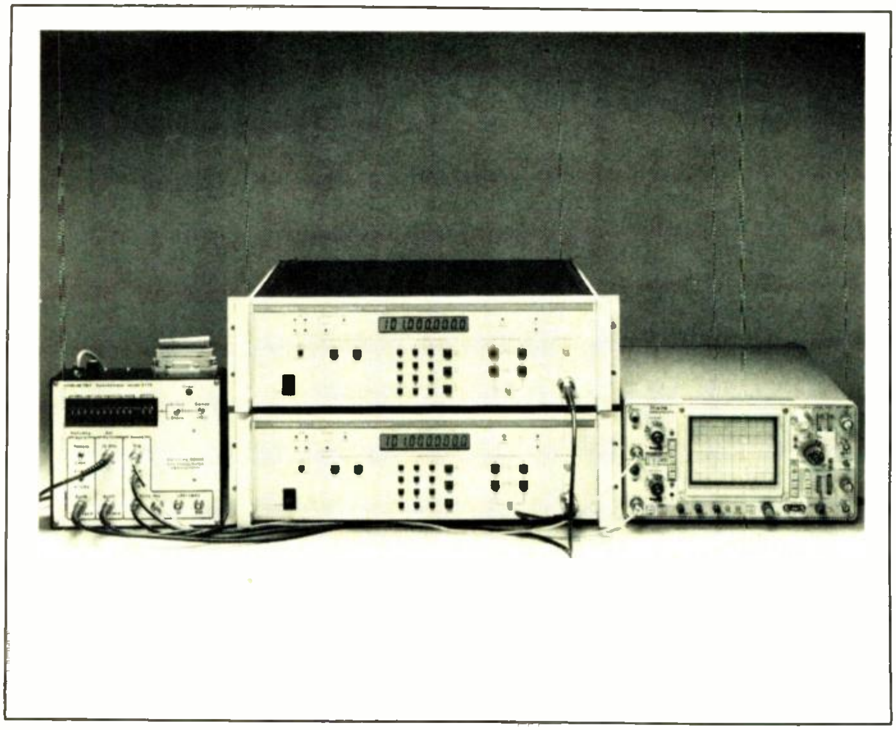

Fig. 1b Switching speed measurement setup.

The switching control block in Figure 1 provides the frequency commands to the UUT for hopping between two operator-selected frequencies. It also provides to the oscilloscope a sample of the execute pulse that initiates each frequency change.

The waveform at the top portion of Figure 2 represents the phase detector output as seen on the oscilloscope when the UUT is switched from a frequency equal to the reference synthesizer frequency (101 MHz), down to a different frequency (100 MHz) and back up to the original common frequency (101 MHz). The lower waveform is the execute strobe that initiates the UUT frequency changes. The leading edge 50 percent point of each strobe pulse identifies the starting time (T_0) of a frequency change command. By identifying the point at which the phase error passes through 0.1 radian (as described later), the delay time of a downward hop (T_1) and the total switching time of an upward hop ($T_{2'}$) can be read from the oscilloscope display.

Figure 3 shows the same frequency hops as those shown in Figure 2; however, the reference synthesizer frequency has been changed to the lower of the two frequencies that the UUT hops be-

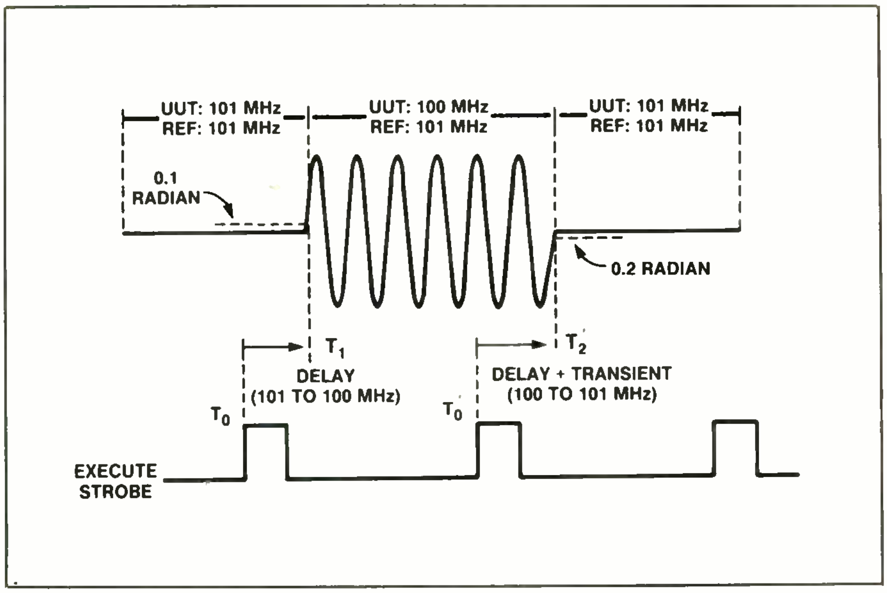

Fig. 2 Phase detector output for UUT hops between 101 and 100 MHz with reference synthesizer at 101 MHz.

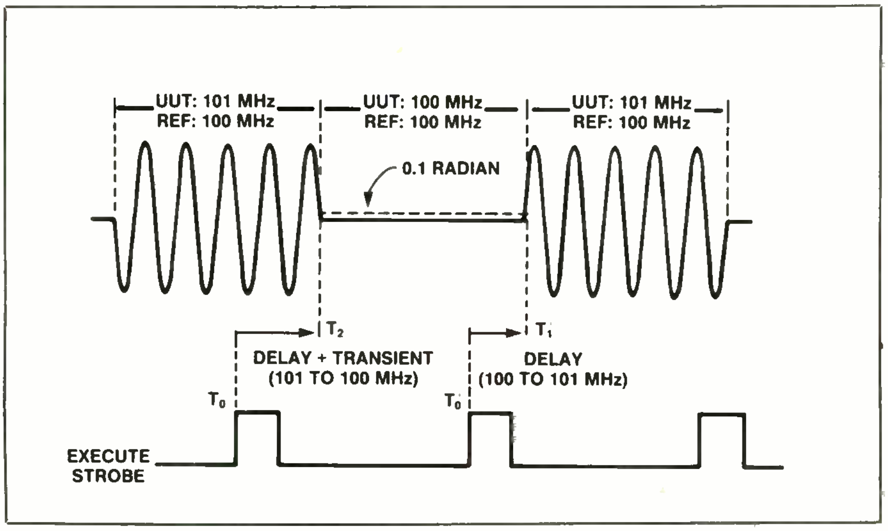

Fig. 3 Phase detector output for UUT hops between 101 and 100 MHz with reference synthesizer at 100 MHz.

tween (i.e. 100 MHz). In this case the total switching time of the downward hop (T_2) and delay of an upward hop ($T_{1'}$) can be read from the oscilloscope.

Direct measurement of switching transient time is very difficult, but it can be calculated by subtracting the delay time (T_1) from the total switching time (T_2) for the downward hop, or the respective times ($T_{1'}$ from $T_{2'}$) for the upward hop.

Before the procedure to identify the 0.1 radian point and the measurement of switching times for a specific frequency hop are discussed, several measurement limitations and pitfalls should be mentioned.

Measurement Pitfalls and Limitations

Almost any arrangement for measuring a difficult parameter will have some pitfalls and limitations. The same applies to testing switching speed over a wide range of frequencies, with various equipment and with differing requirements.

One potential pitfall is a phase detector without sufficient bandwidth to cover the frequency switched. The output (IF) port also must operate down to DC. For very large frequency hops, such as 1 GHz or more, the test setup must be carefully evaluated for adequate

[Continued on page 174]

frequency response.

The lowpass filter used to eliminate the sum frequency from the phase detector must have a transient response that does not mask the transients produced by the synthesizer under test. This imposes a low limit on the filter bandwidth. For instance, to measure a switching speed of 1 μs, a filter bandwidth of at least 10 MHz should be used so that the filter transient will be at least one tenth the transient generated by the synthesizer. Quite often the bandwidth limit setting on the oscilloscope input is sufficient to make the test. The oscilloscope also should have sufficient bandwidth to adequately display the switching transient.

The quadrature relationship between the UUT and the reference tone may not always be maintained. The quadrature relationship can change slowly or quickly from one frequency hop to another. This incoherence can be caused by the method the synthesizer uses to switch frequencies and, within the same synthesizer, the switching techniques may differ from one digit of resolution to another.

For synthesizers that maintain phase continuity when switched from one frequency to another, the DC portion of the display (Figures 2 and 3) will be at different levels unless an integral number of cycles in the difference frequency are displayed. In these cases, two frequencies resulting in a stable DC display should be selected.

Low bandwidth frequency hopping can create difficulties in measurement, as can a low slope when crossing the 0.1 radian point. A number of similar limitations are beyond the scope of this article and may require the test engineer to contact the synthesizer manufacturer.

Detailed Test Procedure

To illustrate the switching speed test procedure, the UUT is first programmed to 101 MHz and will be switched between 100 and 101 MHz at a 10 kHz rate. The reference synthesizer has a 0.1 Hz or greater resolution. The BCD frequency commands to the UUT and the execute pulse signal to the oscilloscope are provided by the control box.

The reference synthesizer is set initially to the higher of the two switched frequencies, 101 MHz. The two-channel oscilloscope is set to trigger on Channel 2 input with AC coupling, with time division and other parameters set to display both traces of Figure 2 conveniently. Power out from the two synthesizers should be set for appropriate levels into the mixer.

The initial output of the mixer will be a DC voltage level indicating that the two synthesizers are at identical frequencies. The lowpass filter should have a cutoff frequency that will ensure that the sum frequencies are attenuated and that only the difference frequency is passed.

Establishing the 0.1 Radian Point

Many manufacturers specify phase error at 0.1 radian (5.73°). The problem is to conveniently locate this point on the difference frequency sine wave.

[Continued on page 176]

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Since the sine of a small angle is approximately equal to the angle expressed in radians, at 0.1 radian, the amplitude (x) can be derived from:

$$\frac{x}{0.1} = \frac{V_{peak}}{\sin 90°}$$

and $x = 0.1V_{peak}$

To establish the 0.1 radian point, one synthesizer should be offset from the other by approximately 0.1 Hz. The scope vertical gain is then adjusted so that the minimum and maximum points (i.e. V_{peak}) of the oscillation fall on the second graticle lines above and below the center line. Th- vertical sensitivity is then increased by a factor of five and the first graticles above and below center line become 0.1 V_{peak}, or 0.1 radian levels.

Establishing Quadrature

After establishing the 0.1 radian graticle lines, the oscillating frequency offset must be removed just as the voltage passes through the center line zero value. (It may take a few tries to achieve the zero value.) With the zero voltage level achieved, the two instruments are in quadrature, the state that exists when the inputs to the mixer (the LO and the RF) are 90° apart in phase.

Measuring Programming Delay (T_0 to T_1)

The UUT then is placed in its remotely controlled hopping mode so that the oscilloscope presentation contains the elements of Figure 2. The DC portion of the waveform corresponds to that period of time when both synthesizers are operating at the identical 101 MHz frequency. The portion of the oscilloscope display showing a sine wave indicates that the UUT is generating the difference frequency of 1 MHz in the case. The time between the 50 percent point of the trigger pulse leading edge (T_0) and the point where the waveform first *deviates* from the DC level by more than 0.1 radian (T_1) is measured. This time period is the programming delay for the UUT hop from 101 to 100 MHz.

Measuring Total Switching Time (T_0 to T_2)

To measure the total switching time of a 101 to 100 MHz hop, the UUT and reference synthesizer are set to 100 MHz, offset slightly to establish the 0.1 radian level and then set in quadrature using the methods described above. The UUT is set to the hopping mode and the scope presentation should look like Figure 3. The time from the 50 percent point of the trigger pulse leading edge (T_0) to the point where the waveform first remains within 0.1 radian of the DC level (T_2) is measured, and this time period is the total switching time of the UUT synthesizer.

Calculating The Switching Transient (T_2 minus T_1)

The switching transient for a hop from 101 to 100 MHz may be calculated by subtracting the programming delay time (T_1) from the total switching time (T_2).

To calculate the transient from 100 to 101 MHz, the delay ($T_{1'}$) of Figure 3 and the total switching time ($T_{2'}$) of Figure 2 are measured and $T_{1'}$ is subtracted from $T_{2'}$. ■

Applications of GaAs Heterojunction Bipolar Transistors in Microwave Dielectric Resonator Oscillators*

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This article describes the application of GaAs heterojunction bipolar transistors (HBTs) in microwave oscillators. Microwave characteristics of HBTs are measured to 12 GHz. Using an NPN grounded emitter HBT with 1.2 to 1.5 m emitter width, 4 GHz microwave oscillator power in excess of 10 dBm with 30 percent efficiency is achieved. A dielectric resonator with a relative dielectric constant $\epsilon_r = 38$ is used in the feedback circuit and as a frequency stabilizing element. Oscillator frequency stability of 3 ppm/°C over –30 to +70°C and FM noise of –73 dBc/Hz at 1 kHz off-carrier were measured. This phase noise performance of a GaAs HBT is comparable to that of a silicon bipolar transistor and is superior to performance of a GaAs FET. Further improvements in microwave performance is predicted with device optimization.

Dielectric and metal plate tuning are used to mechanically adjust the frequency of the oscillators. For a 1 dB change in the output power, 3 percent tuning is obtained for the dielectric resonator tuning and a 9 percent for metal plate tuning. The dielectric resonator tuning provides a controlled tuning slope but was found to be susceptible to mode jumping.

Introduction

In communication receivers, strict frequency stability standards are mandated by the FCC. These standards necessitate the use of a stable, low phase noise and temperature-stable local oscillator. Such requirements have been met with microwave oscillators phase locked to 25 to 125 MHz quartz crystal oscillators using several amplifiers and frequency multiplier ICs. Phase-locked oscillators are complex, have a large number of parts, reducing system reliability, and are high in cost. Dielectric resonator oscillators offer the alternative of being highly frequency-stable, simple, reliable and easy to fabricate; they have good phase noise at a reasonable cost.

Typically, Si bipolar transistors to 6 GHz and GaAs FETs at 4 GHz and higher frequencies have been used as active devices in these oscillators. In recent years, substantial progress has been made in growing materials and device fabrication using molecular beam epitaxy (MBE) and metal organic chemical vapor deposition (MOCVD) processes. This has led to the realization of bipolar devices on semi-insulating GaAs substrates. Such heterostructures were initially used to fabricate high speed digital ICs. Similar techniques have been used recently to fabricate microwave heterojunction bipolar transistors (HBTs). For a small, 20 μm emitter periphery, a gain of 11 dB at 12 GHz has been reported; for a larger, 320 μm device, CW output power of 32 mW with 7 dB associated gain at 3 GHz has been reported.[1,2] Higher gain, lower phase noise and even higher power levels exceeding the performances of GaAs FETs are expected with further device optimization and improved material growth technologies.

Microwave amplifiers and oscillators to 20 GHz are realizable with the present devices that have maximum oscillation frequencies, f_{max}, of 25 GHz or more. This work at 4 GHz presents a comparative study of spectral purity, phase noise and other microwave performance parameters of a GaAs HBT oscillator with Si bipolar transistor and GaAs FET oscillators. For all cases, similar design methods are used and all

*Invited paper.

three types use a dielectric resonator (ϵ_r = 38) as a feedback and frequency stabilizing element. Conventional mechanical tuning of the oscillators using a metal tuner allows an accurate frequency setting for many transmitter/receiver type applications. Results of a dielectric tuner used to achieve linear tuning with control on the tuning rate are presented.

HBT Microwave Characteristics

Microwave GaAs HBTs used are general-purpose NPN grounded emitter devices, fabricated by MBE and mounted in a 70 mil commercial stripline package. Ion implanted base and optical contact lithography were used to define emitter widths of about 1.2 μm. The DC breakdown voltage, V_{CE}, is 4 to 5 V, and the DC current gain, β, is on the order of 30. The microwave measurements, at a bias of 3 V at 15 mA, resulted in the S-parameters as listed in Table 1. Over the typical operating bias range, the small-signal S-parameter showed insignificant bias dependence (V_{CE} of 2 to 3 V and I_c of 10 to 20 mA). The microwave gain compression characteristics measured at 12 GHz (Figure 1) showed a 6 dB gain, 35 percent power-added efficiency and P_{sat} of 17 dBm (2.5 W/mm). Higher narrowband gain of up to 11 dB at 12 GHz has been observed in the amplifier mode for the HBT devices.

Oscillator Design

A parallel feedback-type oscillator circuit configuration was selected and the HBT was operated in the common emitter mode. The circuit was constructed on a 0.025-inch-thick duroid substrate with ϵ_r = 10.2. Oscillator design was carried out using the S-parameter method and a dielectric resonator coupled to a pair of 50 ohm microstrip lines. The material characteristics of the dielectric resonator are listed in Table 2. S-parameters in the transmission mode were measured for the dielectric resonator. The measured data were used to derive the coupling coefficients needed in the oscillator design. Figure 2a shows the circuit and layout of the parallel feedback DRO. A photo of the oscillator with dielectric tuners is shown in Figure 2b. A similar technique was used to design FET and Si bipolar dielectric resonator-stabilized oscillators for comparative studies. A 50 ohm resistor on the base or gate was used to suppress undesired modes and for achieving good out-of-band frequency stability.

The location of the resonator was selected to give a light coupling (high Q), thereby improving phase noise characteristics and reducing harmonic signals (resulting from the overdriven gate or base of the device). Additionally, the dielectric resonator is placed on a thin quartz spacer above the duroid substrate. A high strength, low loss epoxy was used for assembly.

MICROWAVE SCATTERING PARAMETERS

Frequency (GHz)	S_{11}		S_{21}		S_{12}		S_{22}		Max AVL Gain (dB)	Stab Fac (K)
	Mag	Ang (deg)	Mag	Ang (deg)	Mag	Ang (deg)	Mag	Ang (deg)		
4.0	0.41	131.0	3.12	-117.0	0.06	-135.6	0.64	126.8	12.50	1.63
8.0	0.39	-89.8	1.87	0	0.08	9.0	0.63	-103.8	7.94	2.01
12.0	0.22	-5.0	1.46	92.0	0.07	126.0	0.55	-33.0	5.06	3.33

f_{max}: 21 GHz, bias: 3 V/15 mA, f_t: 30 GHz.

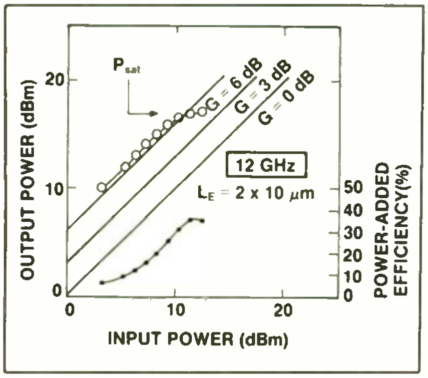

Fig. 1 Saturation and efficiency characteristics of GaAs HBT in amplifier mode. P_{sat} = 17 dBm (50 W) = 2.5 W/mm.

MATERIAL CHARACTERISTICS OF THE DIELECTRIC RESONATOR

Material	(Zn Sn) TiO_9
Rel. dielectric constant (ϵ_r)	37 ± 1.0
Temperature coefficient of resonant frequency	-2 ± 2 ppm/°C
Quality factor (Q)	~ 10,000
Shape	cylindrical, doughnut
Resonator support	quartz

Phase Noise and Other Performance Characteristics

Figure 3 shows the frequency response of the HBT-DRO biased at 2.5 V/15 mA. This response is quite clean, free of spurious signals and similar to the responses of the GaAs FET and Si bipolar transistor. Output power, harmonic content (second and third) and efficiency of the oscillators were measured under varying bias conditions and are listed in Table 3. Output power of 10 dBm with 30 percent efficiency in oscillator operation was achieved. Second harmonic content was at least 15 dB below the fundamental.

Mechanical frequency tuning and power variations resulting from such tuning were examined (Figure

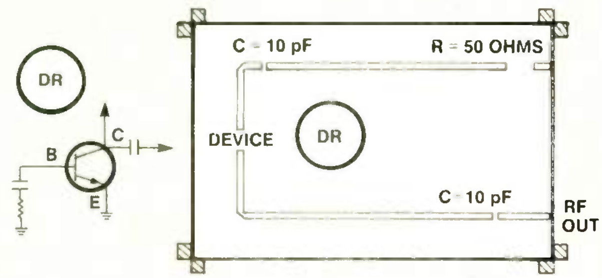

Fig. 2a Layout of shunt feedback DRO.

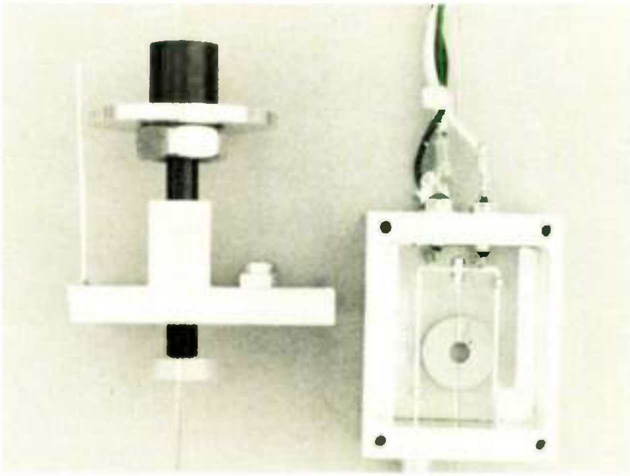

Fig. 2b DRO showing layout and dielectric tuner.

4). A tuning range of about 400 MHz was achieved with output power variations of less than 2 dB. This tuning range is comparable to other types of DROs, but with somewhat less output power variation (2 dB vs 3 dB for a FET).

DROs were tested for frequency stability with temperature over −30 to +70°C (Figure 5). The HBT-DRO has a 3 ppm/°C frequency stability.

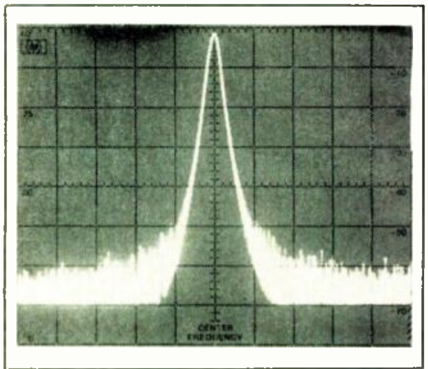

Fig. 3 Frequency response of a 4 GHz HBT-DRO, biased at 2.5 V/15 mA.

Comparable figures for a FET of 2 ppm/°C and Si bipolar transistor of 5 ppm/°C were measured.

Close-in phase noise performance of −73 dBc/Hz at 1 kHz away from the carrier was measured for an HBT-DRO at 4 GHz (Figure 6). This is about 12 dB better than the GaAs FET and almost the same as a Si bipolar transistor oscillator. As one moves away from the carrier, the phase noise of the HBT-DRO degrades and is almost equal to that of a FET-DRO (see Figure 6 at 10 kHz and beyond). The cause of this crossover phenomenon is under further investigation.

The measured results are summarized in Table 4 for comparison of the achieved performance of the HBT, FET and bipolar oscillators.

Frequency Tuning of DROs

Mechanical tuning with metal and high dielectric material tuners was measured (Figure 7). Since microwave output power of the oscillators changes with tuning, 1 dB or less change in the output power was selected as the criteria to compare tuning range and capability. The metal tuner had about 9 percent tuning capability (curve 1). For the dielectric tuner ($\epsilon_r = 38$), a dielectric material with $\epsilon_r = 38$ was used. The thickness of the resonator was divided in various ratios, resulting in

4 GHz HBT DRO OSCILLATOR OUTPUT FREQUENCY AND POWER AS A FUNCTION OF V_{CE} AND I_E

V_{CE} (v)	I_E (mA)	Output Frequency (MHz)	Output Power (dBm) Harmonic Signals 1st	2nd	3rd	Efficiency (η%)
2	10	4007.905	7.6	−6.0	−12.0	28.7
2	13	4007.209	8.6	−15.0	−5.0	27.8
2	15	4006.737	8.7	−17.0	−7.0	24.71
2	20	4005.781	9.7	0.0	−9.0	23.33
2.5	10	4012.773*	8.1	−1.0	−20.0	25.82
2.5	13	4007.955	9.8	−2.0	−16.0	29.4
2.5	15	4007.630	10.4	−6.0	−14.0	29.2
2.5	20	4006.604	11.0	−7.0	−7.0	20.14
3	10	4017.043*	9.0	−1.0	−22.0	26.47
3	13	4008.269	10.4	−1.0	−20.0	28.11
3	15	4008.189	11.3	0.0	−13.0	29.97
3	20	4007.361	12.4	−11.0	−4.0	28.96
3.5	10	4008.151*	9.3	−1.0	−21.0	24.31
3.5	13	4007.252*	10.8	−2.0	−18.0	26.42
3.5	15	4008.250*	11.5	−2.0	−19.0	26.90

*Output was distorted

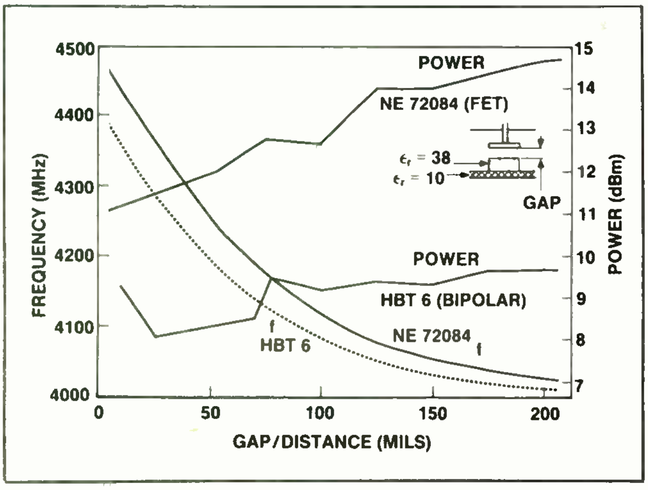

Fig. 4 Measured tuning and output power of DROs.

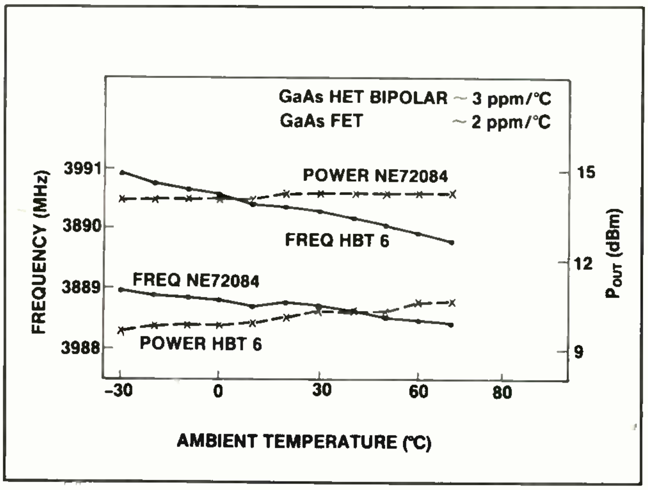

Fig. 5 Frequency stability and power variation with temperature for GaAs oscillators.

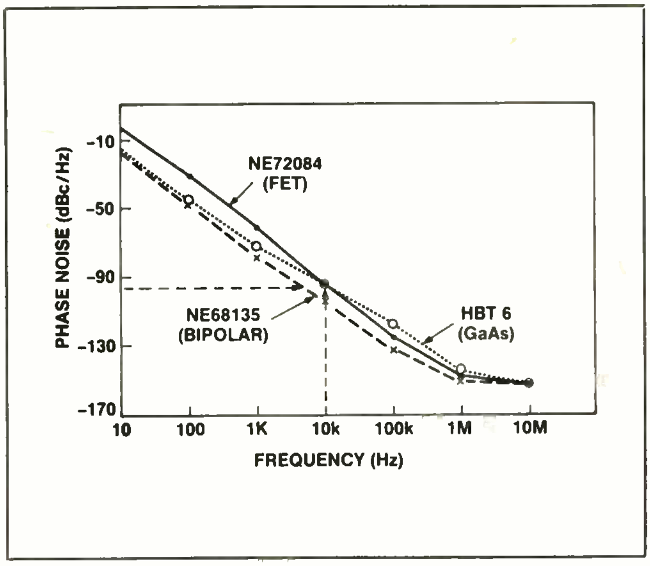

Fig. 6 FM noise behavior of GaAs HBT, GaAs FET and silicon bipolar transistor.

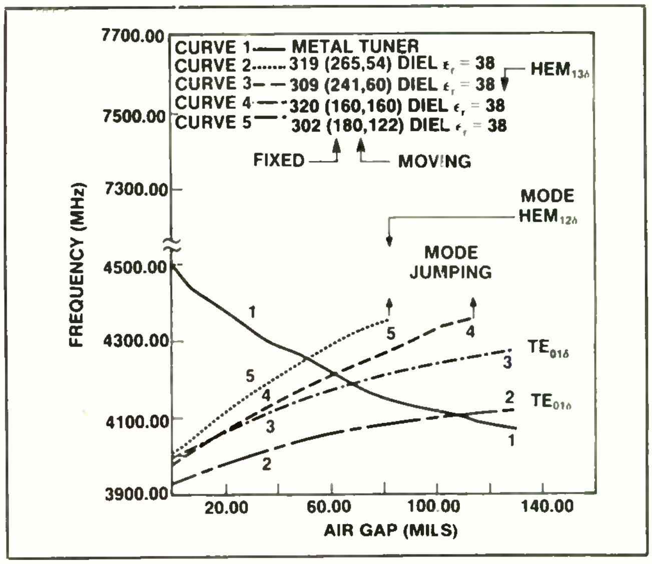

Fig. 7 Measured tuning characteristics of DROs (1-metal tuner, 2-5 dielectric tuners).

a variable tuning slope (MHz/turn or degree). A maximum tuning range of 3 to 4 percent was achieved with good frequency stability. For higher tuning slopes and for wider tuning range (up to 8 percent), one has to be careful because of the tendency of the oscillators to jump to higher-order resonance modes. This problem is minimized by fabricating the main resonator to nearly its correct size and using a very thin dielectric tuning plate or a seperate thick dielectric tuner made of high dielectric material for frequency adjustment.

Conclusion

It has been demonstrated that GaAs heterojunction bipolar transistors are suitable as microwave oscillators. They offer good output power, high efficiency and low phase noise in oscillator applica-

[Continued on page 182]

SUMMARY OF DRO PERFORMANCE CHARACTERISTICS

Device	HBT Rockwell	FET NE72084	Bipolar NE68135
Oscillation frequency (MHz)	3990.4	4006.8	3968.0
Output power (dBm)	10.2	16.0	14.9
Power of second harmonic frequency (dBm)	-6.0	-1.0	-8.0
Efficiency (%)	29.4	44.2	12.9
Frequency stability (ppm/°C) -30 to +70°C	2.802	1.361	5.29
Tuning range for 1 dBm power var (mech) (%)	2.8	2.1	3.0
FM noise (dBc/Hz) at 1 kHz off-carrier frequency	-73.0	-61.0	-78.0
At 10 kHz off-carrier frequency	-95.0	-95.0	-108.0

tions. Excellent frequency stability with temperature can be realized using GaAs heterojunction bipolar transistors and high dielectric constant resonators. Phase noise comparable to that of Si bipolar transistors is possible with GaAs HBTs. It is expected that advances in material technology, device design and circuitry would lead to further improved performance with heterojunction bipolar transistors on GaAs at frequencies to 20 GHz or more.

Acknowledgment

I would like to thank Peter Asbeck for providing the HBT devices. In addition, the technical assistance of Ching Ho in the fabrication and testing of oscillators used during this investigation is gratefully acknowledged. ■

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In 1979 he joined the senior staff at TRW Space & Defense. He managed breadboard development of the satellite transponder under contract from Bell Labs and was part of the team responsible for solid-state power amplifier design for Ku-band satellites.

Since 1982, he has been with Advanced Technology, Rockwell International in Dallas, where he is manager of optics and RF components. He has been doing applied research in GaAs and high speed devices and components for applications in fiber optics and microwave product lines of Rockwell.

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He received the President of India award for merit and is the recipient of a Philips Research Fellowship, Ford Foundation Fellowship and several merit-based scholarships and fellowships. He is a member of Eta Kappa Nu, Tau Beta Pi, and Pi Mu Epsilon. He has published several papers in journals on microwave subsystems and components and continues to do research with emphasis on circuit applications of advanced materials and devices.

DC to 40 GHz Switches and 33 GHz Attenuators

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Introduction

A new technology has been developed to fill the need to go from DC to above 26 GHz in coaxial transmission lines for mechanically controlled switches and attenuators. Refining the current state of the art to go above 26 GHz required innovative solutions for dealing with the demands of increased precision on mechanical parts and connectors without sacrificing reliability and longevity. The basic building block of the switch is the coplanar concept using two planar technologies sandwiched together. The development of the attenuator builds on the concept of using coplanar technology with the added feature of attenuation pads.

DC to 40 GHz Microwave Switch

The development of the 40 GHz SPDT and transfer switches was undertaken to fill an internal need to go above 26 GHz in coax. The current technology, moving reeds in a switch cavity, served the industry well for many years and is well understood to 18 GHz. Where special care is taken to assemble a precision unit, acceptable return loss to 26 GHz can be achieved. A combination of factors, such as connectors, cavity and reed dimensions, renders the performance of reed switches unacceptable above 26 GHz.

A new approach was needed to successfully assemble 40 GHz switches in a manufacturing environment where the mechanical design parameters of the switch must be able to accommodate reasonable manufacturing tolerances. To achieve the precision necessary to go to 40 GHz, the design of the switch took advantage of existing technologies applicable to thin-film hybrids, and the issues of reliability and durability were addressed early in the program.

Switching takes place in a compensated 50 ohm microstrip environment. The K connector by Wiltron is used to make the launch onto a 10 mil thin-film quartz substrate with microstrip runs. A circular flexible polyimide disk with a photoetched contact is sandwiched against the quartz hybrid and pressure is applied by foam backing. Switching takes place by rotating the polyimide disk until the appropriate lines contact and the RF circuit is completed. The word "coplanar" denotes that switching is done with two superimposed surfaces in the same plane, that of the quartz substrate and that of the polyimide disk.

Rotational motion was chosen over linear motion to do the switching because it is much easier to control machining tolerances in a rotating bearing than in a linear one. Cavity resonances are eliminated by the addition of blocks of microwave absorbing material designed to fill the vacant cavity spaces. Port-to-port isolation is greater than 60 dB for the SPDT switch and greater than 50 dB for the transfer switch at 40 GHz. The 40 GHz performance is made possible by the fact that thin-film microstrip line widths and locations can be controlled to 0.0002 inch and the quartz substrate thickness can be held to 0.0005 inch. These tolerances permit accurate positioning of the rotating contacts and precise maintenance of line impedance.

In addition, the rotating polyimide disk covers and protects the contacts from debris, and the contacts are self-cleaning due to the wiping action. With this construction, the switch can handle high levels of shock and vibration without momentary contact opening. No failure of this kind was noted in shock testing to the 800 g level in all axes and in vibration from 30 to 2000 Hz at 50 g's.

Finally, the K connector launcher is soldered in place, ensuring a stable frequency response over time and temperature. Using microstrip on quartz instead of stripline in air does compromise insertion loss. Typical insertion loss and return loss are shown in Figure 1.

In searching for compatible switch materials, it was discovered that gold placed on polished quartz had excellent dry bearing qualities. There were, however, two unexpected problems that affected the life of the switch. Life testing showed that the liquid applied polyimide used to embed the contact lines could not hold up to the wear on quartz. Early failures were due to an increasing population of polyimide granules that had been ground off and which eventually contaminated the contacts to the point of opening. The factory cured polyimides have been found to be tougher materials, but the disadvantage is that the contacts no longer could be embedded.

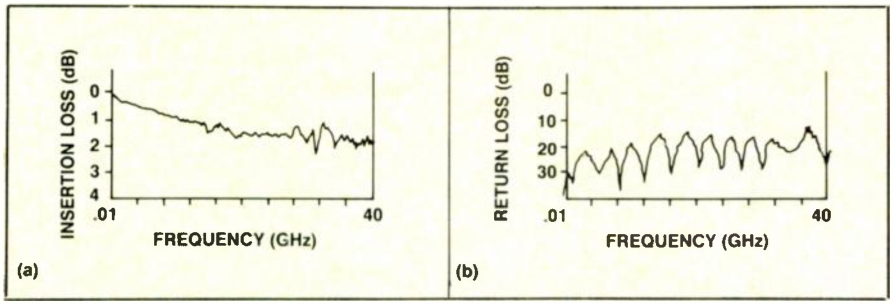

Fig. 1 Typical switch (a) insertion loss; (b) return loss.

The material that now is in use is copper clad polyimide. A mask is used to gold plate the contacts. Then the copper is etched back, leaving the gold-plated contact run. Since the polyimide disk is flexible and compliant, the raised contacts deflect into the foam backing, allowing the contacts to slide over each other.

The second problem that affected the life of the switch was a metallurgical one. Thin-film quartz substrates are usually plated with soft gold. Since the coefficient of friction of gold on gold is very high and the switch works by wiping contacts over one another, galling of the contacts occurred. The contacts would wear completely away. The use of harder material plated on the thin-film quartz substrate stressed the metal film until adhesion was lost. The solution involved plating the contact areas on the thin-film substrate and polyimide disk with compatible materials.

Typically, the drive mechanism wears out before the switch does. After a million cycles, changes in insertion loss are barely detectable. To date, 6 W at 10 GHz have been switched on a continuous basis. Measurements made on the motorized transfer switch at the National Bureau of Standards in Boulder, CO indicate that good repeatability between positions is achievable within specifications. The Hewlett-Packard 8510 was used to obtain the data, which was then cross-checked with the NBS six-port measurement system.

The switching system eliminates the need for special cavity shapes and straight reeds. As shown in Figure 2, the contact runs on the polyimide disk are curved to reduce reflections. The general design approach permits the designer freedom to use any number of contact shapes and compensations on both the quartz substrate and the polyimide disk. In addition, as in the attenuator, there is space on the quartz substrate to integrate any passive or active device.

The actuator developed for the switch is a spring-loaded overcenter cam driven by a DC gearmotor. It has 90 degree rotation and sufficient torque to overcome the considerable frictional losses of the switch, and it precisely lines up the

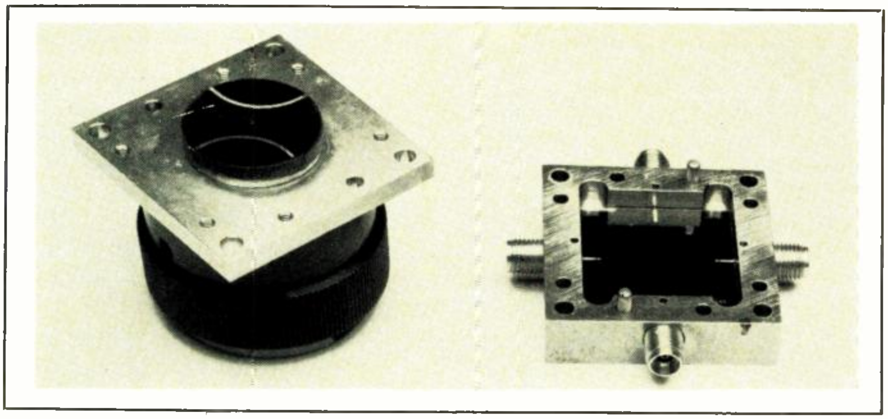

Fig. 2 Major subassemblies of the 40 GHz switch.

contacts. A limit switch disconnects the motor when switching is completed. Reversing polarity on the leads reverses the gearmotor, changing the switch to its original state. Switching time is about 1 second. Motor power requirement is 6 V at 60 mA.

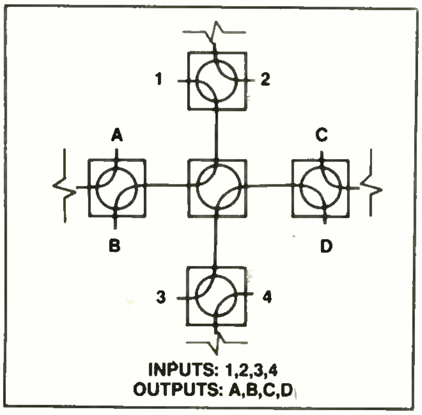

Fig. 3 Five switches used in a 4 x 4 matrix configuration. Note the equal path lengths between the inputs and outputs.

The K connectors fan out radially from the switch, providing ready wrench access to cable connectors. This layout lends itself well to matrix configurations because switches can be connected with short, straight cables, minimizing insertion loss. Figure 3 shows an example of a 4 x 4 matrix that uses five transfer switches. Note that all port combinations have equal line length for constant phase. Unused ports can be terminated with 50 ohm SMA terminations.

DC to 33 GHz
10 dB Step Attenuator

The basic building block for an attenuator is a switch that can serially switch in and out of different levels of attenuation. Typically, this has been done with slab line, employing flexible reeds driven by actuators to conduct the RF energy to a through path or to an attenuation hybrid. This method has been used successfully to 26 GHz. However, these assemblies have many small parts requiring the highest degree of precision in manufacturing and assembly to produce a reliable attenuator with good flatness.

Historically, attenuators have been one of the least reliable components in instruments due to their delicate nature. Going to higher frequencies means smaller parts and tighter tolerances and further degradation of reliability.

With coplanar technology, all the critical line width and spacing tolerances can be held very accurately by photolithography and thin-film processing on quartz techniques. In addition, thin-film resistors for attenuation pads easily can be included in the manufacturing process.

The complete attenuator is built on one thin-film quartz hybrid. Figure 4 shows four 33 GHz, 10 dB pads mounted on one side of the hybrid attenuator housing. The pads are switched in and out by a pair of counter-rotating contacts on polyimide disks. Since there are four pads, eight contacts are necessary to achieve the appropriate switching. Standing wave ratio at 33 GHz is 1.9 (nominal impedance 50 ohms). Insertion loss and return loss for the 10 dB pad are shown in Figure 5.

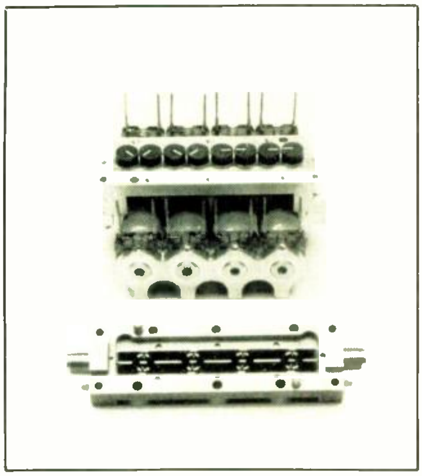

Fig. 4 Four 33 GHz, 10 dB pads mounted on one side of the hybrid attenuator housing.

This attenuator topology puts some unusual constraints on the pad design. Ground is accessible on one side only, and a minimum power handling capability of 1 W was desired. Additionally, the design had to allow for trimming in order to achieve the desired precision of DC attenuation values.

Extensive computer modeling was used to arrive at this unique (patent pending) lumped asymmetrical pad design. The input and output resistances of 50 ohms, as well as the correct attenuation, were major considerations in that model. A 3D field program was used iteratively to converge the pad geometry (Figure 6) to the targeted DC values. The shape distributes the incoming current across the input side, enhancing power dissipation while keeping parasitic capacitance to a minimum.

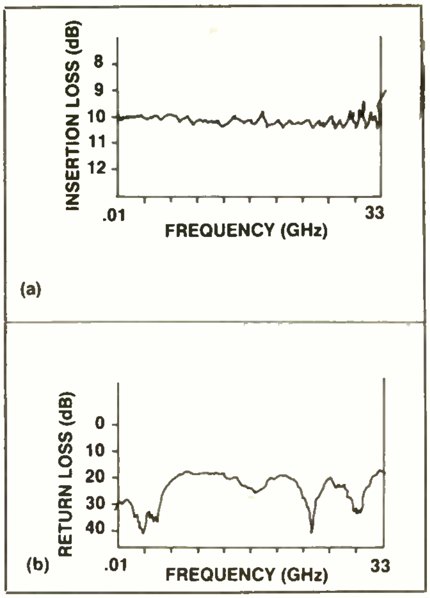

Fig. 5 (a) Insertion loss and (b) return loss of the 10 dB pad referenced to the through position.

An analog computer design program was then used to obtain the AC response. Optimization was achieved by the use of a compensating stub and a peaking inductor. The final pad configuration was selected from a set of experimental pads built according to the theoretical result. The pad is grounded by wrapping gold ribbon from the pad to the back side metalization and welding it in place. Since the pads have this unusual asymmetrical design, one connector on the attenuator is designated as the input port.

This attenuator topology has an interesting manufacturing benefit. Since the pads are asymmetrical and occupy only one half of the hybrid, a second set of pads can be placed on the opposite side with little added cost. If a defective pad is identified during assembly or in the field, the attenuator can be repaired by removing the housing, which is held by six screws, reversing it end-to-end and reassembling it to the actuator.

Push-pull solenoids, which convert linear motion to rotational motion through pairs of cams, are used to move the four pairs of contact wipers in the precise counter-rotating manner required to control the attenuator. These actuators have successfully operated the component through more than 1 million cycles per stage.

Solenoid drives are particularly suited to the application because they can provide the dynamic torque required by the wiping contacts as well as sufficient pulling force during travel.

A new approach to the design of mechanically operated switches and attenuators, for use above 26 GHz in coaxial line systems, has been demonstrated, and these components are practical to manufacture. In conjunction with newly available coaxial connectors suitable for that frequency range, a new class of coaxial components is available to the system designer working beyond 26 GHz.

Acknowledgments

The author wishes to acknowledge the work of Bob Beckman, who solved the polyimide wear issues, and John Reagan, who developed the metallurgical process. ■

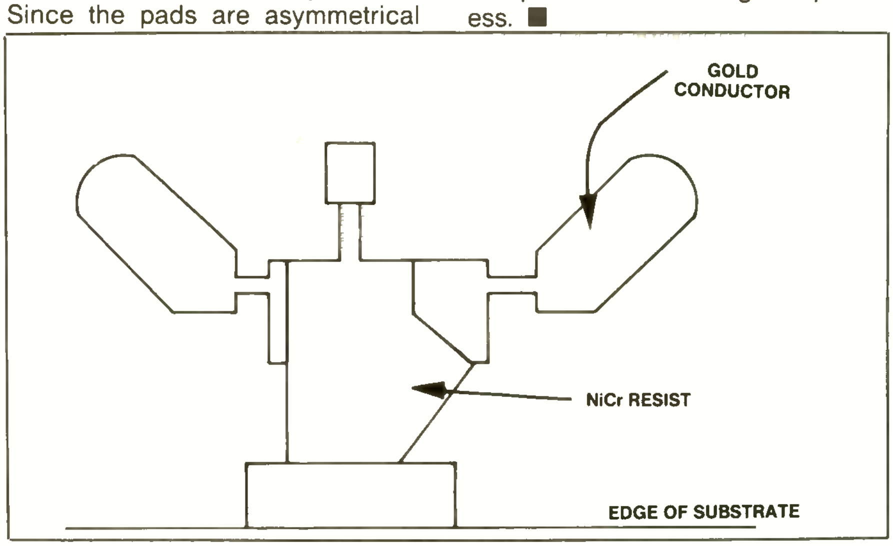

Fig. 6 10 dB pad.

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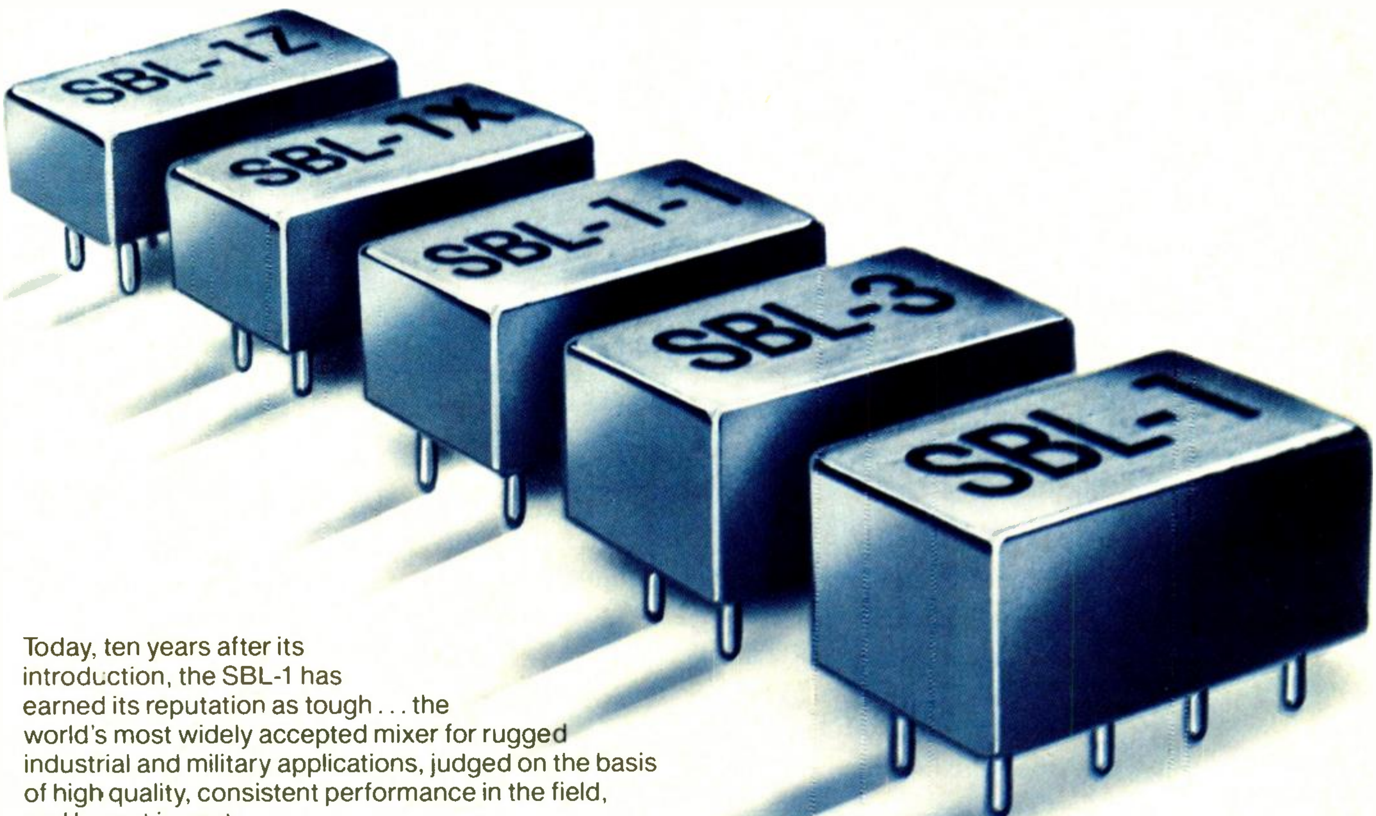

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SBL SPECIFICATIONS (typ.)

Model	Freq. (MHz)	Conv. Loss	Isolation,dB L-R	Isolation,dB L-I	**Price (10-49)**
SBL-1	1-500	5.5	45	40	**$4.50**
* SBL-1X	10-1000	6.0	40	40	**$5.95**
SBL-1Z	10-1000	6.5	35	25	**$6.95**
SBL-1-1	0.1-400	5.5	35	40	**$6.50**
SBL-3	0.25-200	5.5	45	40	**$7.50**

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C104 REV. ORIG.

CIRCLE 139 ON READER SERVICE CARD

Sensitivity Analysis of 3 dB Branchline Couplers

A.F. Celliers
National Institute for Aeronautics System Technology
Council for Scientific and Industrial Research
Pretoria, South Africa

A sensivity analysis was done on a –3 dB branchline coupler to determine the effect of small changes in line lengths and impedance levels on the amplitude of the coupled and through ports, respectively.

Introduction

The performance of 3 dB branchline couplers can be severely degraded by junction effects due to finite line widths. From reference 1, the optimized line lengths and impedance levels can be obtained. The dimensions of the final etched coupler can, however, differ from the optimized design. The difference can be due to etching errors, layout errors and calculating errors. To determine the effect of small deviations on the amplitude of the through and coupled ports, respectively, sensitivity calculations were performed. The equivalent circuit of a T-junction from reference 1 was used in a computer program to calculate the sensitivity of small changes in line lengths and widths. The results were compared with measured and theoretical results at 6 GHz.

Sensitivity Calculations

The sensitivity of a function b(a) with variable "a" is defined as the normalized change in magnitude "b" due to a normalized change in magnitude "a" and is defined as

$$S\Big|_{a}^{b} = \frac{\dfrac{\partial b}{b}}{\dfrac{\partial a}{a}} \qquad (1)$$

If the sensitivity factor, S, is known, the incremental change in magnitude "b" due to an incremental change in magnitude "a" can be calculated.

This paper is based on a dissertation submitted to the Department of Electrical and Electronic Engineering, University of Stellenbosch, South Africa, in partial fullfillment of the requirements of the M Eng degree by A.F. Celliers.

Figure 1 shows the general configuration of a branchline coupler. The sensitivity of the output power at port 3 with small changes in impedance Z_1 is given by

$$S\Big|_{Z_1}^{B_3} = \frac{Z_1}{B_3}\,\frac{\partial B_3}{\partial Z_1} \qquad (2)$$

where B_3 is the output power at port 3.

The output power at ports 3 and 4 both are obtained from references 2 and 3 by computing the [ABCD] transmission parameters of the equivalent circuit of the branchline coupler, as shown in Figure 1, and transferring that to the S-parameter matrix.

The derivative of the output power is obtained numerically.

From reference 4,

$$f'(x_0) \approx \frac{1}{\Delta x}\left\{f(x_0 + \Delta x) - f(x_0)\right\}. \qquad (3)$$

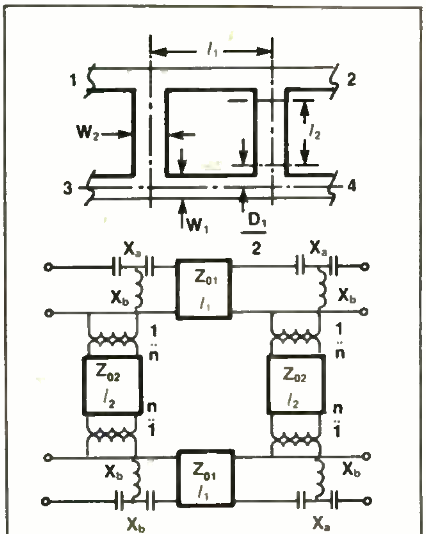


Fig. 1 Equivalent circuit of branchline coupler including junction discontinuities.

Substituting Equation 3 in Equation 2, the sensitivity is obtained as

$$S\Big|_{Z_1}^{B_3} = \frac{Z_1}{B_3} \times \frac{B_3(Z_1 + \Delta Z_1) - B_3(Z_1)}{\Delta Z_1} \qquad (4)$$

Equation 4 is calculated for Z_1, Z_2, l_1 and l_2 with respect to the output power at ports 3 and 4, respectively.

Computer Calculations

The sensitivities

$$S\Big|_{Z_1,\,Z_2,\,l_1,\,l_2}^{B_3}$$

and

$$S\Big|_{Z_1,\,Z_2,\,l_1,\,l_2}^{B_4}$$

were calculated at 6 GHz for a stripline branchline coupler with the following: conductor thickness = 1 oz copper, ground plane separation = 3.175 mm and ϵ_r = 2.32. These are shown in Figures 2 and 3, respectively. A coupler was manufactured for a 6 GHz center frequency. The

[Continued on page 190]

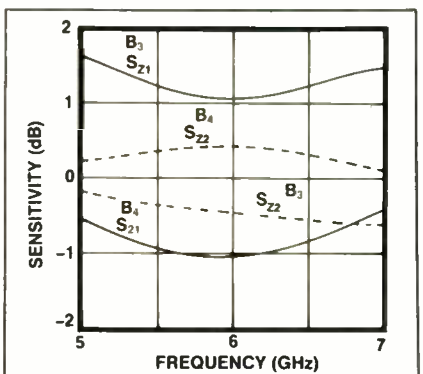
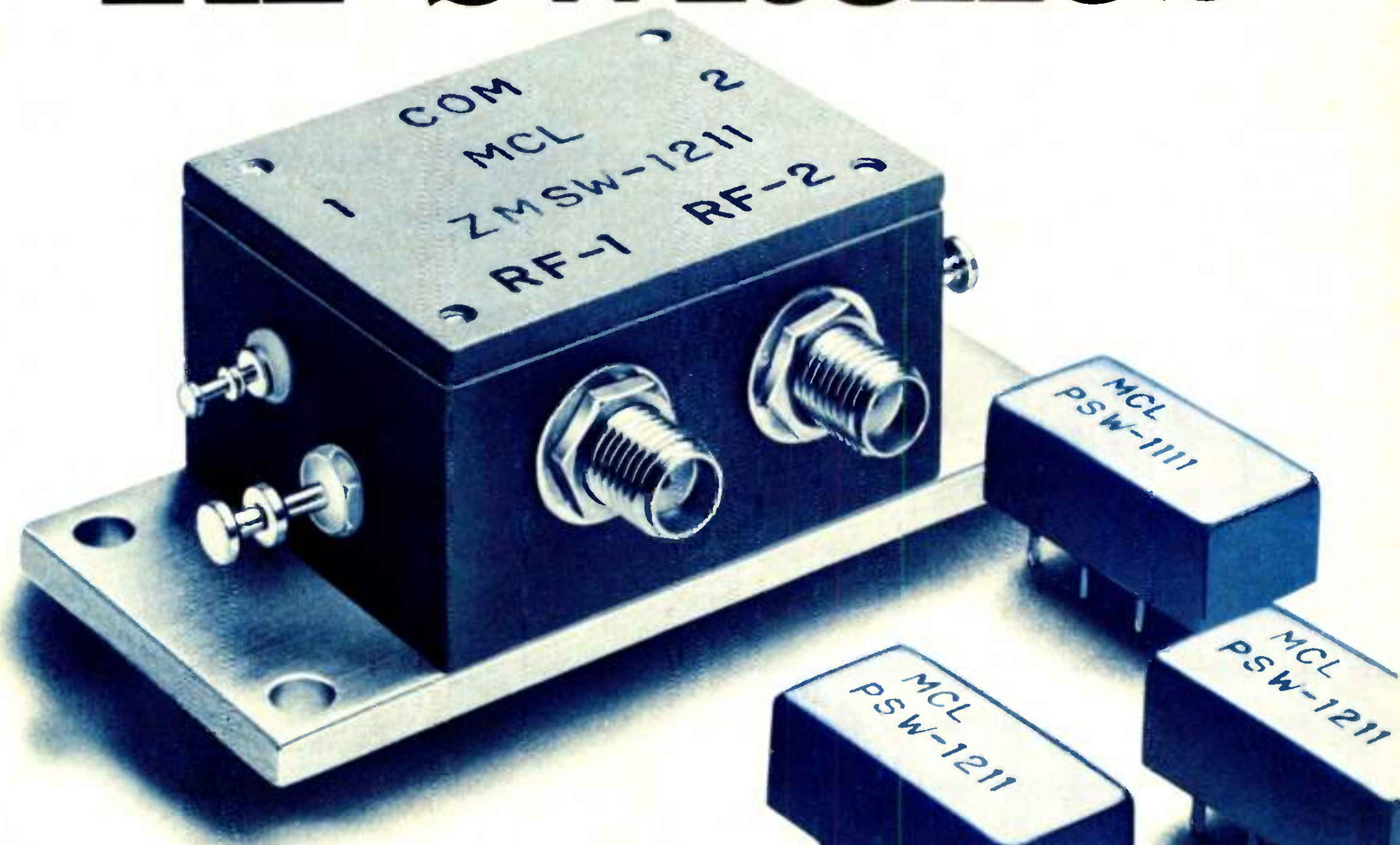


Fig. 2 Sensitivity of coupler with respect to Z_1 and Z_2.

RF switches



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ZMSW 1111 and ZMSW 1211

FREQUENCY RANGE	10-2500 MHz
INSERTION LOSS	
10-2000 MHz	1.7 dB max.
2000-2500 MHz	2.7 dB max.
ISOLATION	
10-500 MHz	40 dB min.
500-1000 MHz	30 dB min.
1000-2000 MHz	25 dB min.
2000-2500 MHz	20 dB min.
SWR	1.5 max. ("on" state)
SWITCHING SPEED	1 μsec. (max.)
MAXIMUM RF INPUT	+20 dBm
CONTROL	+5 V (5 mA max.)
OPERATING TEMPERATURE	−54°C to +100°C
STORAGE TEMPERATURE	−54°C to +100°C

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C80 REV. B

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measured transfer characteristics are shown in Figure 4. The dimensions of the coupler were measured and are shown in Table 1.

By using the sensitivity curves of Figures 2 and 3, one can calculate the variation of output power at ports 2 and 3, respectively to be

COMPARISON OF DIMENSIONS OF OPTIMIZED AND PHYSICAL COUPLER

	Measured Dimensions	Optimized Dimensions	Deviation (%)
l_1	9.025 mm	9.2 mm	-1.9
$l_2 = l_3$	7.379 mm	7.496 mm	-1.56
Z_1	51.0 Ω	50.0 Ω	2.0
$Z_2 = Z_3$	34.1 Ω	32.49 Ω	4.96

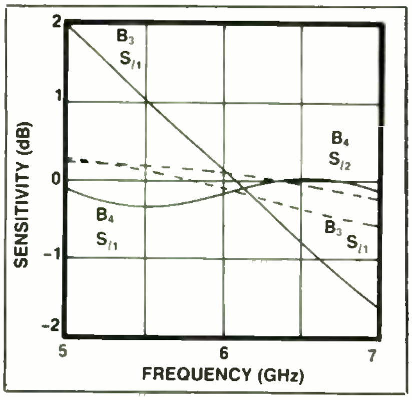

Fig. 3 Sensitivity of coupler with respect to l_1 and l_2.

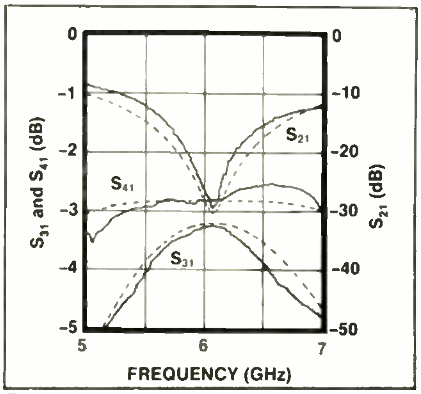

Fig. 4 Measured (solid line) and predicted (dotted line) response of coupler.

$$\Delta B_3 = \Delta B_{3Z_1} + \Delta B_{3Z_2} + \Delta B_{3l_1} + \Delta B_{3l_2} = -0.2 \text{ dB.}$$

$$\Delta B_4 = \Delta B_{4Z_1} + \Delta B_{4Z_2} + \Delta B_{4l_1} + \Delta B_{4l_2} = 0.16 \text{ dB.}$$

This is exactly the same as was measured in Figure 4. From Figures 2 and 3 it is seen that the output power at the various ports is more sensitive to variations in impedance Z_1 and Z_2 and less sensitive for variations in line lengths l_1 and l_2. Figure 4 also shows the computer-simulated characteristics of the coupler.

Conclusion

From the results obtained, it is found that branchline couplers are sensitive to line width variations. The results shown in Figure 4 prove that the equivalent circuit of reference 1 is accurate.

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2. J.L. Altman, *Microwave Circuits*, D. von Nostrand Co. Inc., pp. 151, 315, 400, 402 and 405, 1964.
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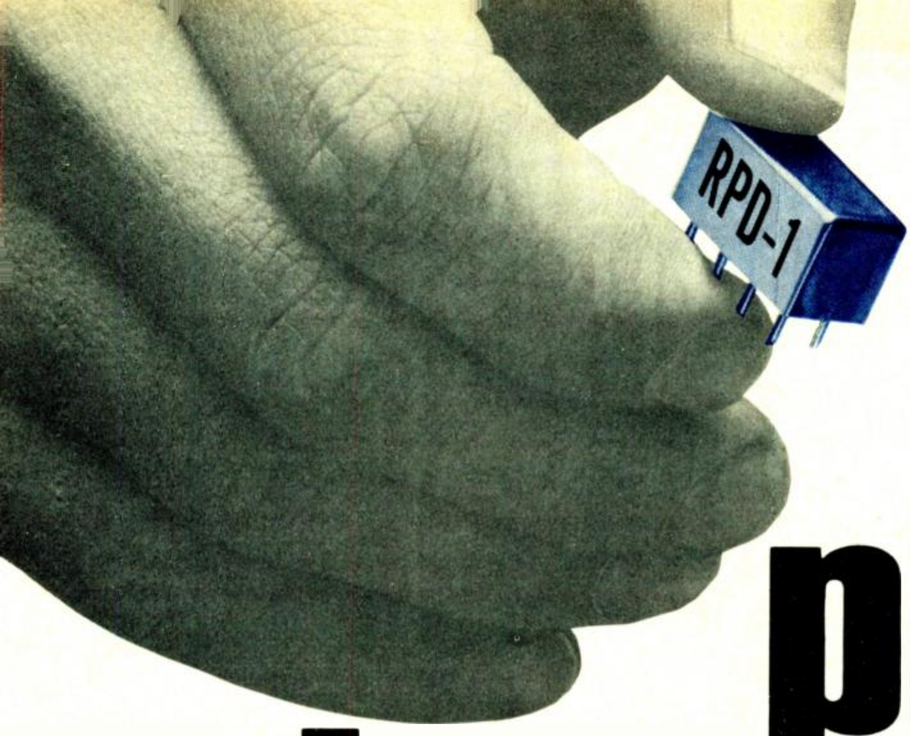
4. C.R. Wylie, *Advanced Engineering Mathematics*, McGraw-Hill Book Co., pp. 116-133, 1960.

Abraham F. Celliers *was born in July 1954. He received the degrees BSc, Hons BSc and MSc in electronics engineering from the University of Stellenbosch in 1976, 1977 and 1980, respectively.*

He joined the National Institute for Aeronautics and Systems Technology, Council for Scientific and Industrial Research, Pretoria, South Africa. His main interests are active and passive microwave and mm-wave components. Currently, he is working on phase stabilization techniques for semiconductor sources. ■



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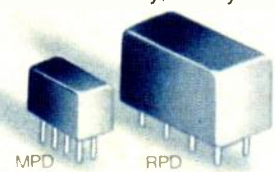

SPECIFICATIONS	MPD-1 RPD-1	RPD-2	MPD-2
FREQUENCY RANGE (MHz)			
L and R ports	1-100	5-150	10-200
Output port	DC-50	DC-50	DC-50
SCALE FACTOR (mV/degree)	8	8	8
IMPEDANCE (ohms)			
L and R ports	50	50	50
I port	500	500	500
L and R SIGNAL LEVELS (dBm)	+7	+7	+7
ISOLATION, L-R, min. (dB)	40	40	40
MAX. DC OUTPUT (mV) typ.	1000	1000	1000
min.	750	750	750
DC OUTPUT POLARITY (L and R in-phase)	neg	neg	neg
DC OUTPUT OFFSET (mV) typ.	0.2	0.3	0.3
max.	1	1	1
FIGURE-OF-MERIT, M (typ)	143	143	143

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C63 REV. C

Determination of Power Dissipation Characteristics of Broadband, In-Phase Power Combiners

Marc E. Goldfarb
M/A-Com Microwave Subsystems Group
Burlington, MA

Editor's Note: The following application note in our computer-aided design series is based on the use of Touchstone software.

In the design of in-phase Wilkinson power combiners, it is often necessary to be able to combine two coherent signals of arbitrary phase. The designer must, therefore, determine the power dissipation in the isolating resistors for a worst-case condition — namely, odd-mode excitation of the power combiner. This situation occurs when the two input signals are of equal amplitude and 180° out of phase with each other.[1] In a single-section, narrowband design, the analysis is a trivial matter since all of the incident power will be dissipated in the single isolation resistor. Additionally, the highest power dissipation will occur at the design center frequency, as this is the point at which the combiner exhibits the smallest reflection coefficient.

In a multiple-section divider, however, the power is not evenly distributed through the resistors. Additionally, the frequency at which the power dissipation is greatest is not necessarily the design center frequency. This application note will present a simple method of determining the power dissipation in the isolating resistors of a broadband power combiner. (See Figure 1.)

Modeling the power dissipation in these resistors can be performed in one of two ways. Either the odd-mode equivalent circuit of the combiner may be analyzed or a suitable source of odd-mode power may be used to excite the complete network. The latter method was chosen for this analysis because the complete model is used, thus eliminating the likelihood of making an error when generating the odd-mode model. A second reason is that it illustrates a noteworthy method of creating an odd-mode signal generator.[2,3]

To excite the combiner properly in the odd mode, it is necessary to have a perfectly matched device that will generate two equal amplitude signals with exactly 180° of phase difference at all frequencies. In the software used, this is achieved using an ideal transformer to form a 180° power dividing balun. As shown in Figure 2, the transformer is configured with a turns ratio that is the square root of 2. This ratio provides the proper impedance relationship at the output terminals of the transformer. When properly matched, this device exhibits the correct power division and phase relationships at its output ports; however, it has no way to terminate in-phase signals reflected back into the balun. For this reason, ideal isolation elements that solve the problem of reflected energy are added to the network.

To determine the power dissipated in the isolation resistors, a voltage "sniffer" is used to probe the resistor voltages. A complete description of the voltage sniffer was presented in an earlier application note, available on request from the software manufacturer.[4]

The software file and data file for this examination are given in Printout 1. The power combiner was designed using the method of Cohn[5] and the performance was optimized using the tune mode in the software. The balun is defined in the first definition block, and the combiner and sniffers are defined in the second. The third block combines the two devices into a single, four-port device: the input port and the three sniffer voltages. The output port of the combiner is terminated.

The time-consuming part of this type of analysis always has been determination of the actual dissipation in the isolation resistors once the magnitudes of the voltages across each resistor are known. This is solved using the processor (PROC) block[6] and an external file which contains the isolation resistor values normalized to 50 ohms. The first line of the processor block squares the S-parameters in the 'test' network. The second line creates the 'tstpwr' network by dividing the square of the S-parameters in 'test' by the normalized

[Continued on page 194]

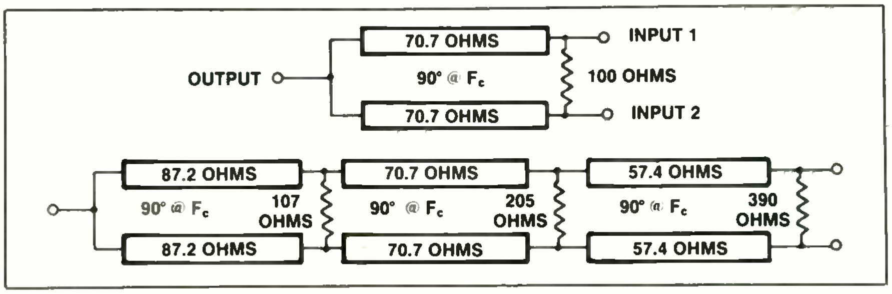
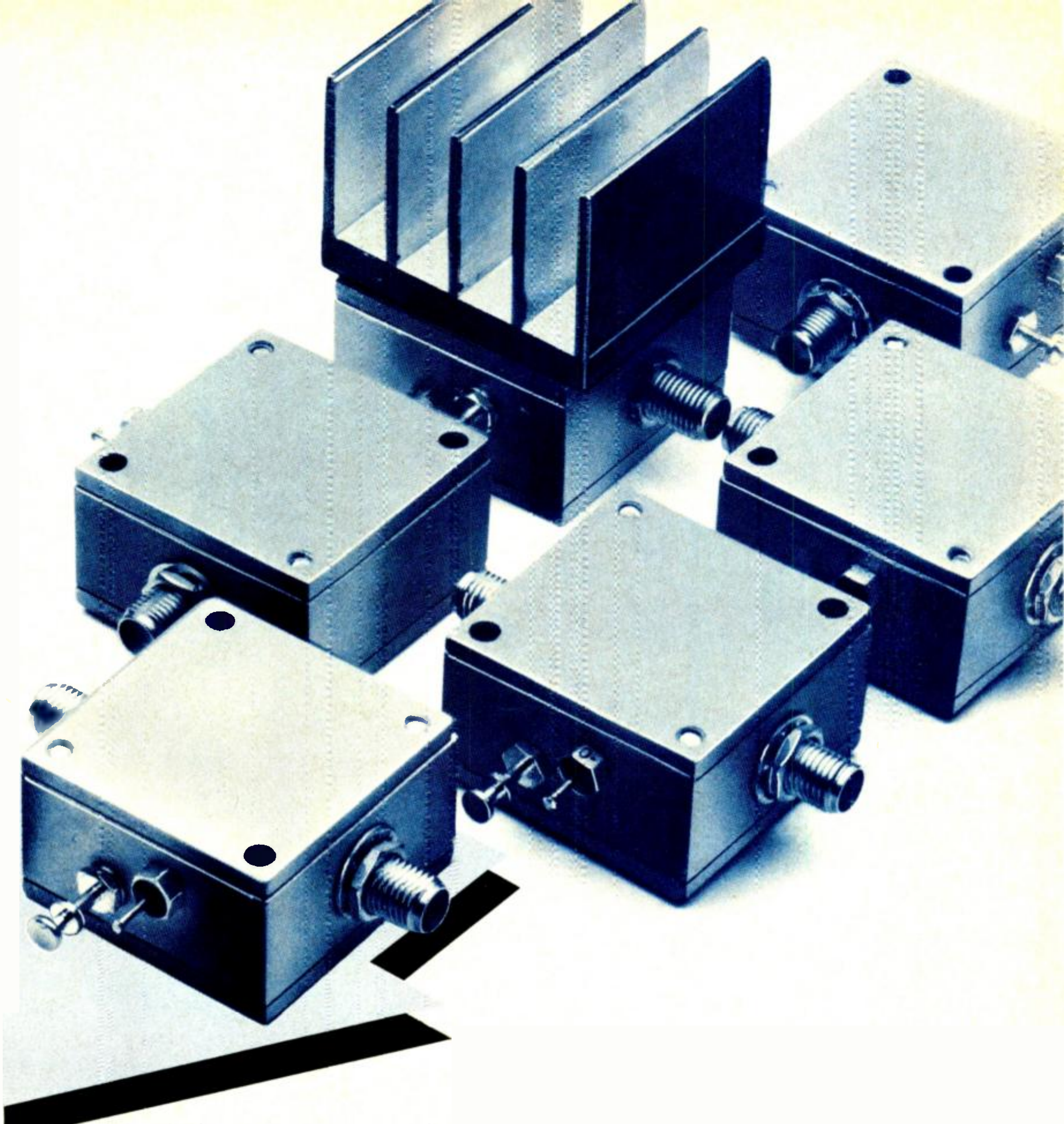


Fig. 1 Single- and triple-section in-phase combiners.



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Freq (MHz)	0.05-500	10-1000	10-2000	10-1000
Gain (dB), Min	20	17	20	28
Gain Flatness (dB) Max	±1.0	±1.5	±1.5	±1.0
Max. Power (dBm) (1dB compression)	+10	+3	+17*	+20
NF (dB) typ.	5.3	12.0	7.0	5.0
3rd order Intercept pt (dBm)	+18	+13	+25	+33
Current at 15V dc	80mA	90mA	100mA	150mA
Price $	69.95	199	179	219
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C101 REV. A

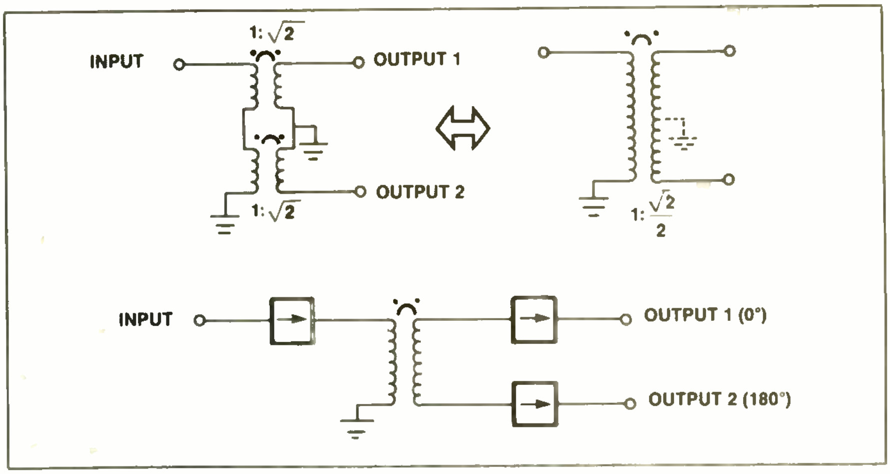


Fig. 2 Schematic of the 180° power divider.

PRINTOUT I

TOUCHSTONE PROGRAM FILE FOR THE EXPERIMENT

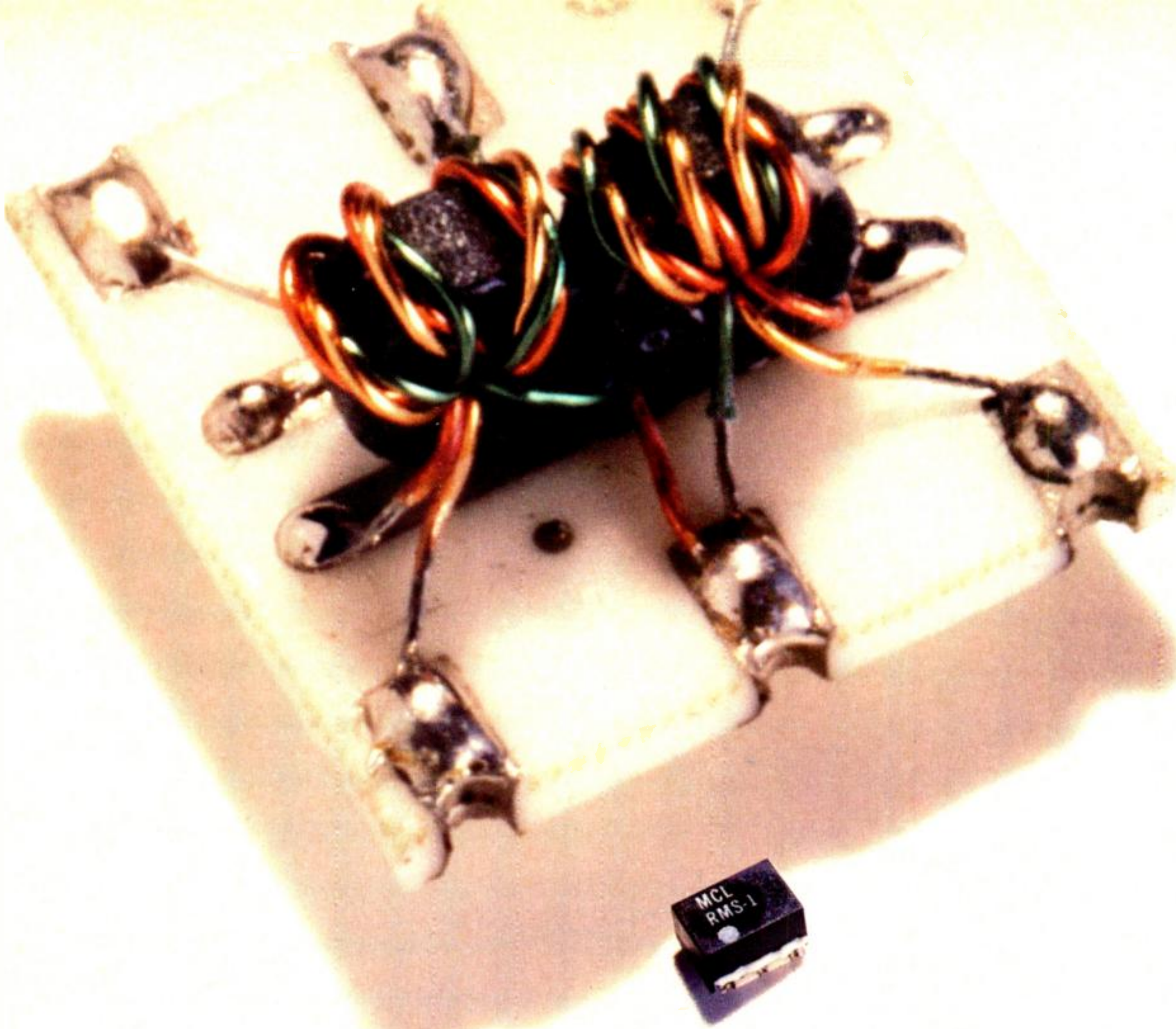
```
' ODDPWR3 - Measurement of power dissapation in an inline, zero phase in a
'           3 section power combiner using a balun for odd mode excitation,
'           and a VCVS as a voltage sniffer.
ckt
' Define odd mode excitation balun
 isolator   1 2
 xfer 2 3 0 5 n=.7071
 isolator   3 6
 isolator   5 7
def3p 1 6 7 balun

' Define power combiner and VCVS'
 tlin 1 2 z=87.1  e=90 f=1
 tlin 1 3 z=87.1  e=90 f=1
 tlin 2 4 z=70.7  e=90 f=1
 tlin 3 5 z=70.7  e=90 f=1
 tlin 4 6 z=57.4  e=90 f=1
 tlin 5 7 z=57.4  e=90 f=1
 res  2 3 r=107
 res  4 5 r=205
 res  6 7 r=390
 vcvs  2 10 3 0 m=-1 a=0 r1=1e8 r2=0  f=0 t=0
 vcvs  4 11 5 0 m=-1 a=0 r1=1e8 r2=0  f=0 t=0
 vcvs  6 12 7 0 m=-1 a=0 r1=1e8 r2=0  f=0 t=0
 match 1
def5p  6 7 10 11 12 combnr

' Define test configuration
 balun 1 2 3
 combnr 2 3 4 5 6
def4p 1 4 5 6 test

' Define resistor network for proc. block
 s4pa 1 2 3 4 a:isores.s4p
def4p 1 2 3 4 isores

proc
test1  =  test * test              ' Square S21 term in  test
tstpwr = test1   isores            ' Divide |S21| 2 by resistance
out
 tstpwr  mag[s21] gr2
 tstpwr  mag[s31] gr2
 tstpwr  mag[s41] gr2

freq
sweep .5 1.5 .1

grid
range .5 1.5 .1
gr1 -500 -100 50
gr2  0 .5 .1
```

values of the isolation resistors. This method is valid since it is desired to look at the value of the forward transmission parameter (S_{21}) only. The remaining parameters in 'tstpwr' are dummy variables and have no physical significance.

The output from this file is given in Printout 2. As expected, the power dissipated in the 107 ohm resistor is maximum in the center of the band. The power dissipated in the center resistor is minimum at the band center and maximum at the edges. The plot in Figure 3 shows separately the power dissipated in each individual resistor relative to that incident on the entire network.

The error associated with this analysis can be determined by summing the three separate power dissipations obtained at any frequency. In the worst case, the sum is .998.

[Continued on page 196]


MCL
RMS-1

PRINTOUT II

TABULAR RESULTS (a) AND DATA FILE (b)

(a)

FREQ-GHZ	MAG[S21] TSTPWR	MAG[S31] TSTPWR	MAG[S41] TSTPWR
0.500000	0.316	0.414	0.270
0.600000	0.343	0.407	0.250
0.700000	0.364	0.392	0.242
0.800000	0.379	0.374	0.246
0.900000	0.386	0.360	0.253
1.00000	0.389	0.354	0.257
1.10000	0.386	0.360	0.253
1.20000	0.379	0.374	0.246
1.30000	0.364	0.392	0.242
1.40000	0.343	0.40	0.250
1.50000	0.316	0.414	0.270

(b)

```
' Data file for analysis of 3 section power combiner
' File contains the value of the isolation resistors
' normalized to 50 ohms
' freq      |s11| <s11   |s12|/s12   |s13|/s13  |s14|/s14
  .1         1.0  0.0    2.14 0.0    4.1  0.0   7.8  0.0
             2.14 0.0    1.0  0.0    1.0  0.0   1.0  0.0
             4.1  0.0    1.0  0.0    1.0  0.0   1.0  0.0
             7.8  0.0    1.0  0.0    1.0  0.0   1.0  0.0

  2          1.0  0.0    2.14 0.0    4.1  0.0   7.8  0.0
             2.14 0.0    1.0  0.0    1.0  0.0   1.0  0.0
             4.1  0.0    1.0  0.0    1.0  0.0   1.0  0.0
             7.8  0.0    1.0  0.0    1.0  0.0   1.0  0.0
```

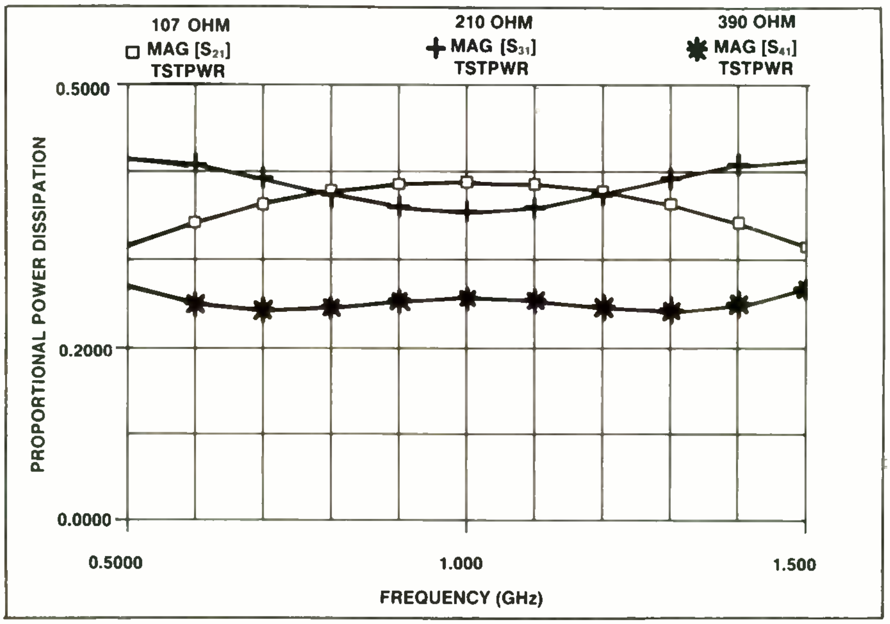


Fig. 3 Isolation resistor power dissipation vs frequency for a three-section in-phase combiner.

This corresponds to a .2 percent error. Farther from the band center, the sum of the three powers will decrease as a significant component of the input power is reflected and dissipated in the isolators. This example demonstrates how a graphical representation of the power dissipated in each resistor within a broadband, in-phase power combiner can be generated. This example could be easily modified to analyze combiners with greater or lesser bandwidths. ■

References

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4. Touchstone Application Note #2, EEsof Inc., 31194 La Baya Dr., Westlake Village, CA 91362, 1986.
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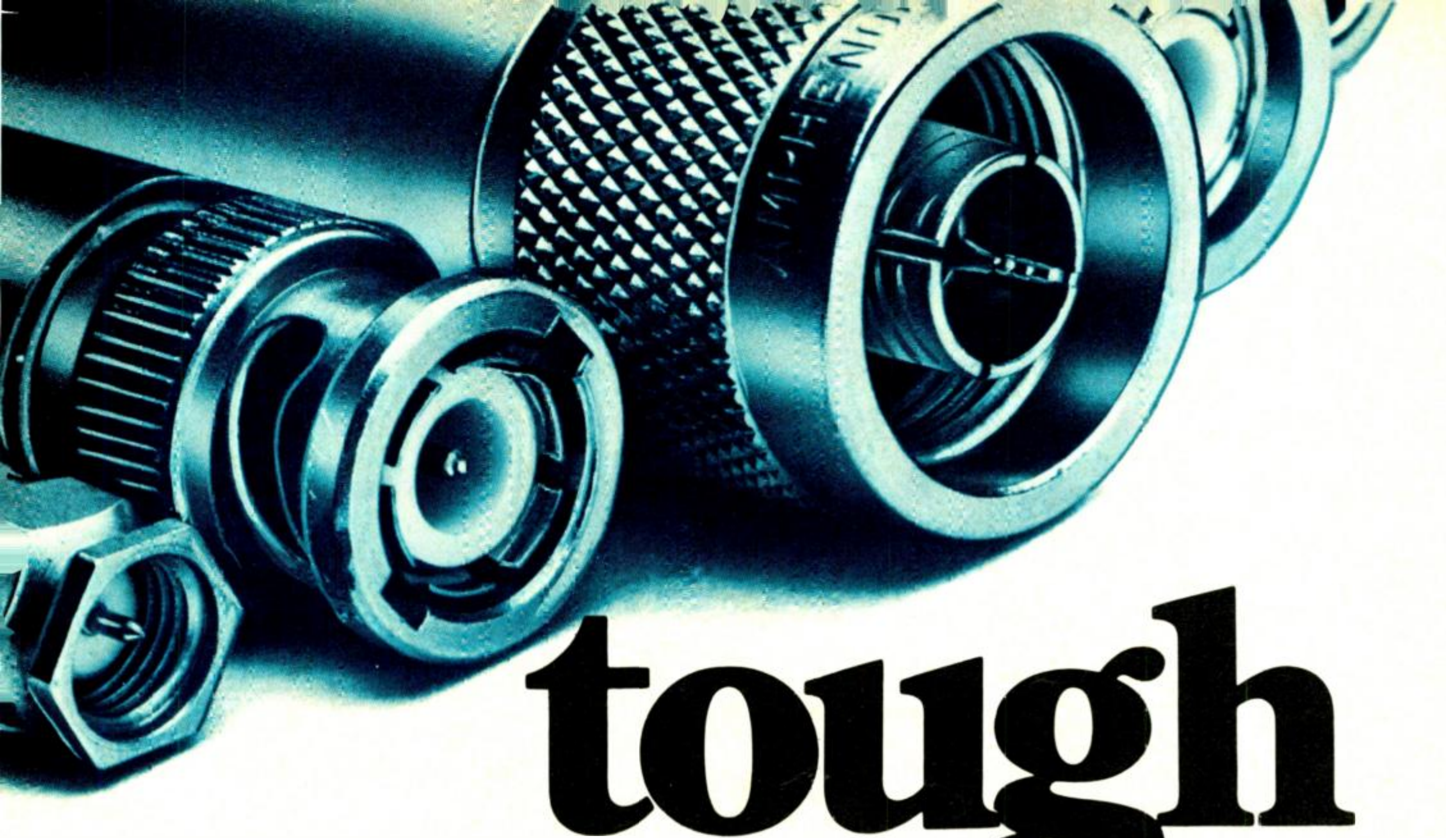

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*Freq (MHz)	Atten. Tol. (Typ.)	Atten. Change, (Typ.) over Freq. Range		VSWR (Max.)	
		DC 1000	1000-1500	DC-1000 MHz	1000-1500 MHz
DC-1500 MHz	±0.3	0.6	0.8	1.3	1.5

*DC 1000 MHz (all 75 ohm or 30 dB models) DC-500 MHz (all 40 dB models)

MODEL AVAILABILITY
Model no. = a series suffix and dash number of attenuation
Example: CAT-3 is CAT series, 3 dB attenuation.
■ denotes 75 ohms; add −75 to model no.
● denotes 50 ohms

ATTEN	SAT (SMA)	CAT (BNC)	NAT (N)	TAT (TNC)
1	●	●	●	●
2	●	●	●	●
3	●	● ■	●	●
4	●	●	●	●
5	●	●	●	●
6	●	● ■	●	●
7	●	●	●	●
8	●	●	●	●
9	●	●	●	●
10	●	● ■	●	●
12	●	●	●	●
15	●	● ■	●	●
20	●	● ■	●	●
30	●	●	●	●
40	●	●	●	●

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Computer-Aided Synthesis Tools for Microwave Amplifier Circuit Design

Radom Inc.
Meridian, Idaho

The SYNMAT CAD program performs network synthesis for the design of matching networks for wideband microwave amplifiers.[1] Significant design improvements have been made in the computer-aided synthesis process which provides a powerful, efficient and friendly tool for the microwave amplifier designer.

Passive network synthesis has long been used for the design of lumped-element filters. These circuits typically operate between real and equal terminations and have a flat frequency response in the passband.

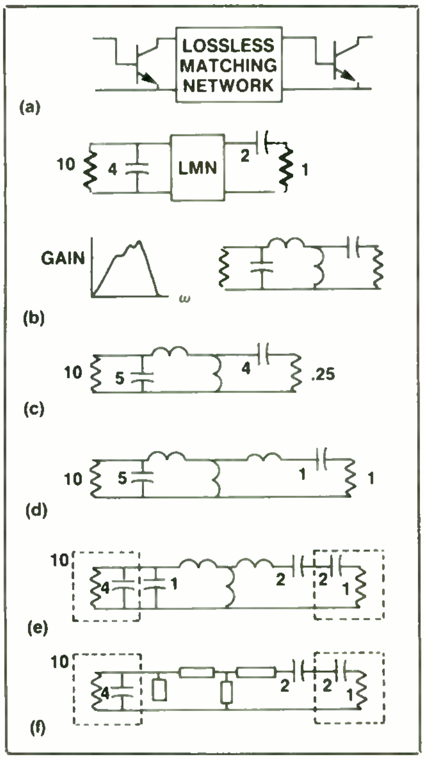


Fig. 1 Design procedure for synthesizing matching networks for microwave amplifiers: (a) model device impedances; (b) constrain frequency response and select topology consistent with parasitic elements; (c) synthesize network; (d) transform impedance; (e) separate out device impedances; (f) transform design to transmission line equivalent.

In SYNMAT, the synthesis method of filter design has been generalized to the design of interstage, input and output matching networks for microwave amplifiers, where the networks must operate between unequal impedance levels and include parasitic elements associated with the active devices. Also, the networks can provide gain compensation to offset the inherent gain rolloff of the transistors with increasing frequency.

SYNMAT can be used to provide for transformation of the lumped design to approximate transmission line equivalents for realization at microwave frequencies.

The steps required in network synthesis are outlined below. It can be seen that it is a simple-to-follow procedure, where the designer can check each circuit element as he proceeds in his circuit synthesis.

Steps in the Matching Network Synthesis Process

The steps in matching network synthesis are outlined in Figure 1 and listed here:

1. Model the input and output impedance of the active devices to be used in the microwave amplifier.
2. Select a topology consistent with device parasitics.
3. Adjust the gain-bandwidth to ensure inclusion of parasitics.
4. Select the reflection coefficient zeros consistent with inclusion of parasitics.
5. Transform impedances to desired levels.
6. Transform the lumped design to a transmission line realization.
7. Analyze the resultant design by itself and/or as part of the complete amplifier design.
8. Optimize the amplifier design (if needed).

Design Example

A single-stage amplifier was designed to operate over 6 to 12 GHz with the requirements of minimum noise figure and maximum gain consistent with minimum noise figure. The transistor used is an NEC GaAs FET biased for low noise op-

[Continued on page 200]

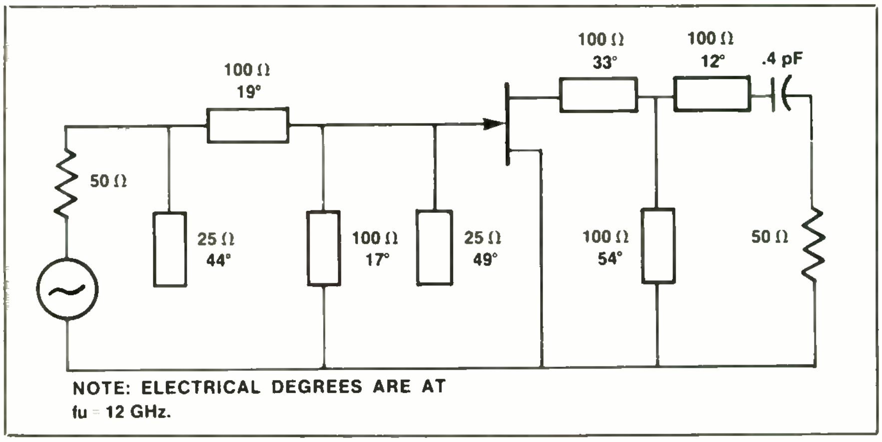
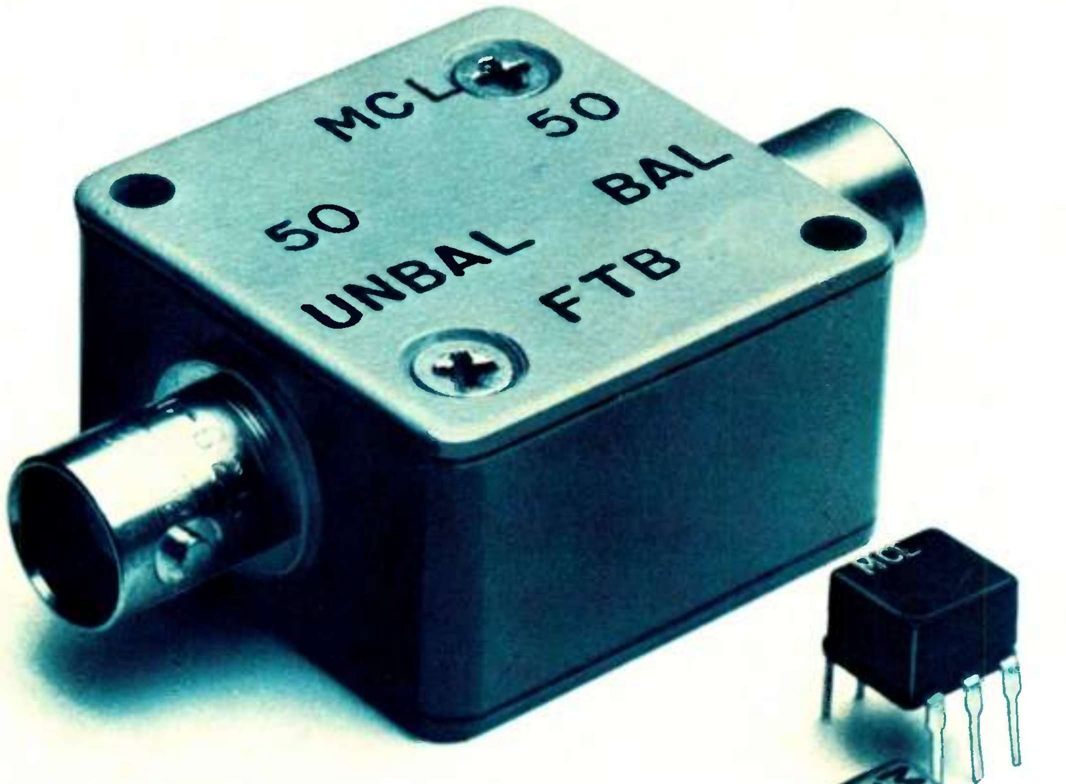
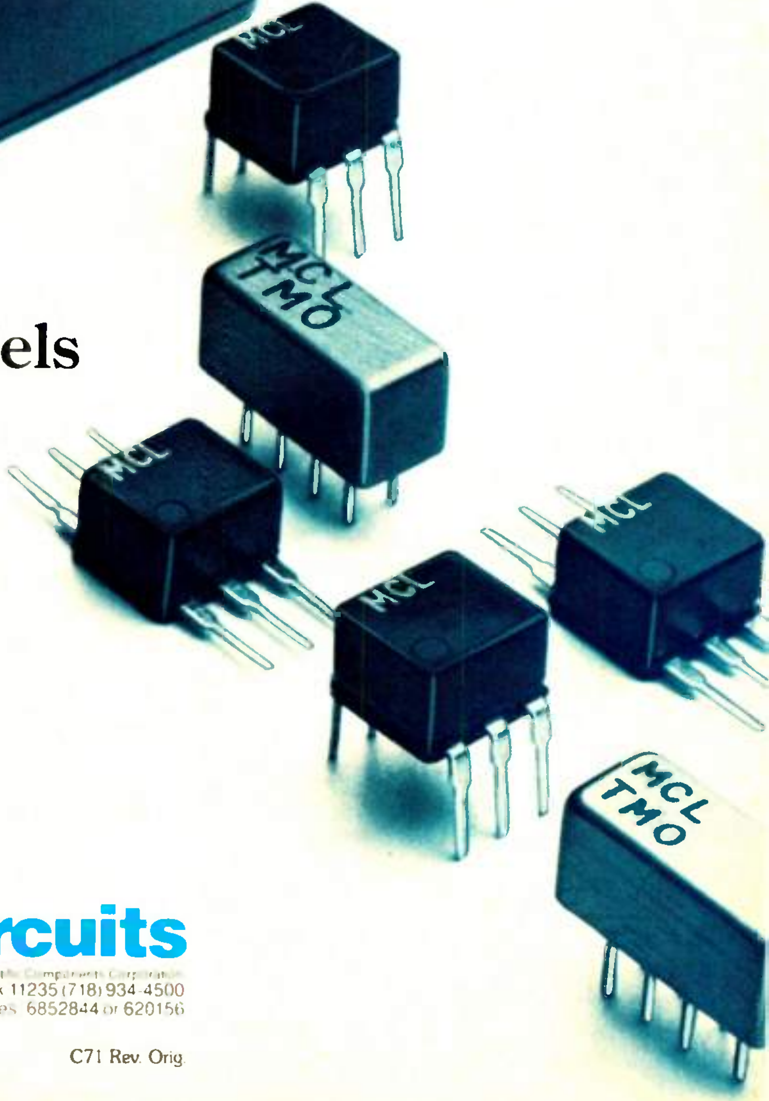


Fig. 2 Amplifier design.

RF transformers





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The requirements were easily translated into specific design specifications for the input and output matching network:

Input network:
Minimum noise figure.
Output network: Best match.

The abbreviated input/output of SYNMAT for the input and output network designs are shown in Printouts 1 and 2.

The resultant amplifier design is shown in Figure 2.

Figure 3 presents the analysis of the uncompensated circuit obtained directly from SYNMAT design.

The following is a summary of SYNMAT program capabilities.

- Device Modeling

— Automated modeling of device input impedance, output impedance and gain rolloff from measured S-parameters

— Automated modeling of noise equivalent input impedance from measured noise parameters

— Automatic frequency response compensation for parasitic element effects

- Synthesis

— Synthesis of lossless coupling networks of prescribed ripple, bandwidth and gain-vs-frequency slope

— Synthesis of a variety of network topologies having the desired frequency response

- Topology

— Automatic topology checking to ensure that a synthesis of an improper topology is not attempted

— Listing of all available topologies for a given response specification

— Provision of a default topology which meets the parasitic requirements and provides a good range of impedance transformation

- Impedance Transformation

— Automated network transformations to provide a desired load impedance. Includes the ability to au-

[Continued on page 202]

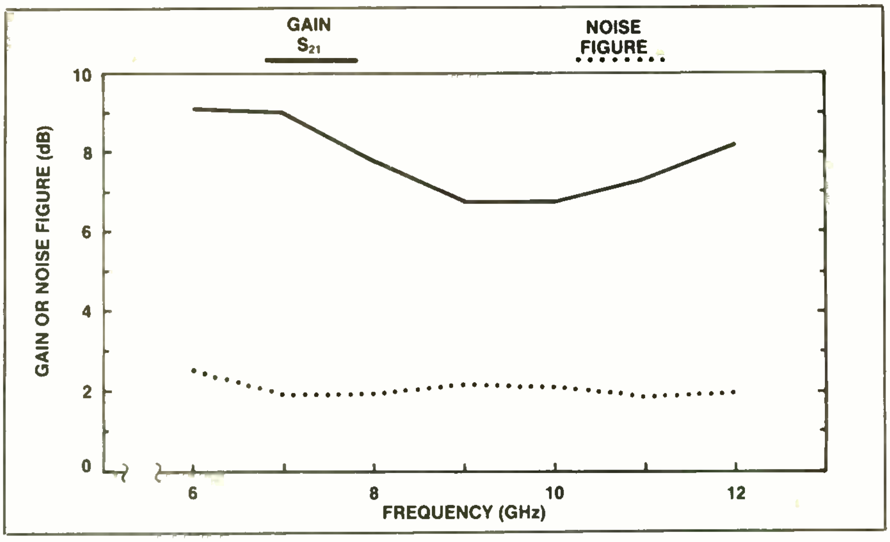

Fig. 3 Characteristic performance of GaAs FET amplifier.

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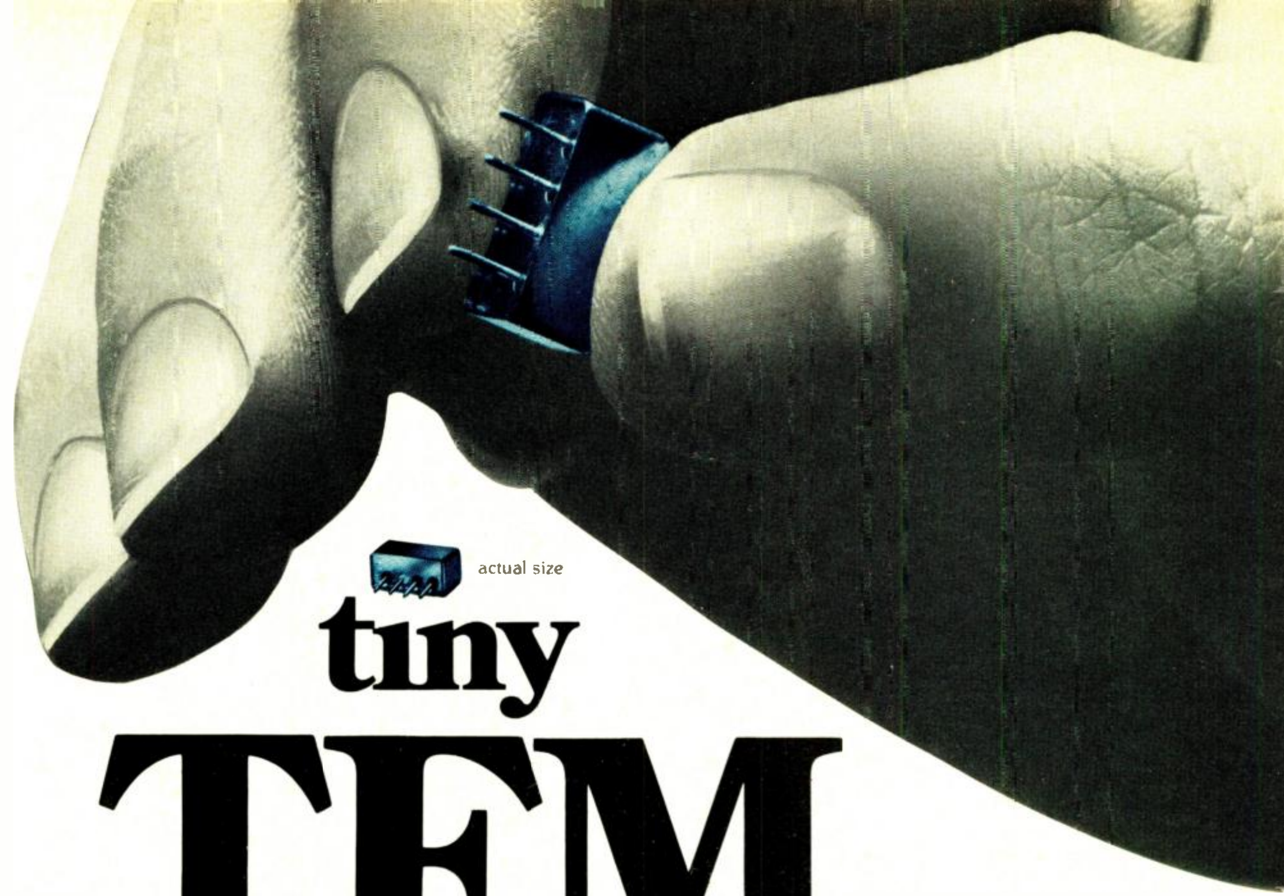

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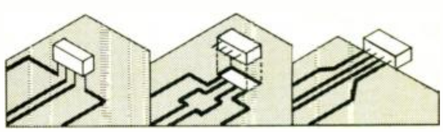

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					Lower Band Edge		Mid Range		Upper Band Edge			
	LO/RF	IF	1 Octave from Band Edge	Total Range	LO-RF	LO-IF	LO-RF	LO-IF	LO-RF	LO-IF	$ EA.	QTY.
TFM-2	1-1000	DC-1000	6.0	7.0	50	45	40	35	30	25	11.95	(1-49)
TFM-3	.04-400	DC-400	5.3	6.0	60	55	50	45	35	35	19.95	(5-49)
TFM-4	5-1250	DC-1250	6.0	7.5	50	45	40	35	30	25	21.95	(5-49)
•••TFM-11	1-2000	5-600	7.0	7.5	50	45	35	27	25	25	39.95	(1-24)
•••TFM-12	800-1250	50-90	—	6.0	35	30	35	30	35	30	39. 95	(1-24)
▬TFM-15	10-3000	10-800	6.3	6.5	35	30	35	30	35	30	49.95	(1-9)
▬TFM-150	10-2000	DC-1000	6.0	6.5	32	33	35	30	35	30	39.95	(1-9)

•••If Port is not DC coupled
▬+10 dBm LO, +5 dBm RF at 1dB compression
*Units are not QPL listed

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[From page 200] **RADOM**

PRINTOUT 1

Device Modeling

```
COMMAND ? FIL-NF
DEVICE FILE NAME ? N70000A
BEST 2-ELEMENT MODEL FOR S 1 1
FOR NOISE MATCHING
CP     R
0.4635 24.2496
```

Synthesis

```
RSOURCE ? 50
DESIRED LOAD ? 24.25
FLOWER, FUPPER (GHZ) ? 6,12
SLOPE..DB/OCTAVE, RIPPLE..DB.MINIMUM LOSS..DB ? 0,.15,0
MAXIMALLY FLAT OR EQUIRIPPLE...MF OR ER ? ER
SELECT A SOFTKEY COMMAND ? TOPOL
NUMBER OF ELEMENTS IN NETWORK ? 4
EMEMENT #1 ? CP
EMEMENT #2 ? LS
EMEMENT #3 ? LP
EMEMENT #4 IS CP
TRANS LINE NETWORK:SOLUTION 2/8
*************************************************************
SOURCE 50.0000 OHM
TLOC 44.504 DEG 25 OHMS
TLIN 19.002 DEG 100 OHMS
TLSC 17.202 DEG 100 OHMS
TLIN 0.432 DEG 100 OHMS
TLOC 49.072 DEG 25 OHMS
CP 0.463500 PF
LOAD 24.249997 OHM
*************************************************************
```

PRINTOUT 2

Device Modeling

```
COMMAND ? FIL-SP
DEVICE FILE NAME ? N70000A
BEST 2-ELEMENT MODEL FOR S 2 2
CP     R
0.1329    130.4236
```

Synthesis

```
RSOURCE ? 130.42
DESIRED LOAD ? 50
FLOWER, FUPPER (GHZ) ? 6,12
SLOPE..DB/OCTAVE, RIPPLE..DB, MINIMUM LOSS..DB ? 0,04,0
MAXIMALLY FLAT OR EQUIRIPPLE...MF OR ER ? ER
SELECT A SOFTKEY COMMAND ? TOPOL
NUMBER OF ELEMENTS IN NETWORK ? 4
1ST ELEMENT IS CP
EMEMENT #2 ? LP
EMEMENT #3 ? LS
EMEMENT #4 ? CS
TRANS LINE NETWORK:SOLUTION 2/8
*************************************************************
SOURCE 130.4200 OHM
CP 0.130000 PF
TLOC 0.254 DEG 25 OHMS
TLIN 32.543 DEG 100 OHMS
TLSC 53.793 DEG 100 OHMS
TLIN 12.186 DEG 100 OHMS
CS 0.407548 PF
LOAD 50.000001 OHM
*************************************************************
```

tomatically cascade multiple impedance transformations when necessary

- Transmission Line Realization

— Automated transformation of (lumped) synthesized networks to approximate equivalents in distributed elements

- Output Files

— Creation of standard format output circuit files of synthesized results.

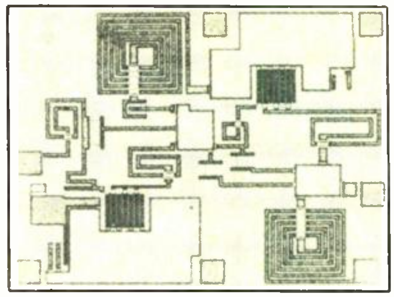
Radom Inc., Meridian, ID (208) 323-0318. **Circle No. 282**

Reference

1. Douglas J., Mellor, "Improved Computer-Aided Synthesis Tools for the Design of Matching Networks for Wideband Microwave Amplifiers," 1986 IEEE-MTT-S Digest.

Product Feature

2 to 6 GHz GaAs MMIC Amplifier Chip



Celeritek
San Jose, CA

The model CMM-2 monolithic GaAs amplifier covers 2 to 6 GHz with 9.0 dB minimum gain and consumes only 125 mW of DC power.

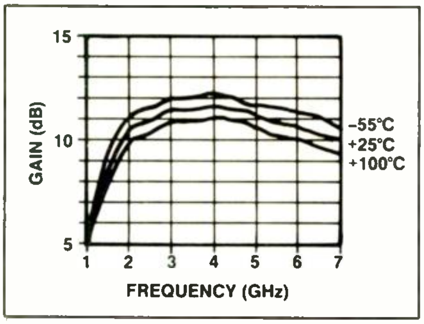


Fig. 1 Typical gain over temperature (V_{dd} = 5 V).

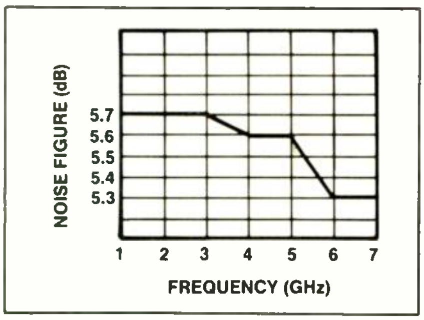


Fig. 2 Typical noise figure (V_{dd} = 5 V).

Typical gain flatness is ±0.5 dB over full bandwidth. Gain is stable over temperature with a typical variance of 1.5 dB from −54 to +100°C (see Figure 1).

The amplifier is designed for military applications calling for stable and reliable performance over extreme temperature ranges with minimum current drain. Applications include missiles, expendables and manpack equipment.

Operated from a +5 VDC, 25 mA supply, the CMM-2 typically provices 10.5 dB gain. Associated noise figure is less than 7 dB, typically 5.5 dB (see Figure 2). Power output at the 1 dB compression point is greater than +8.0 dBm. Typical input and output SWR is 1.8.

Size of the completely self-contained chip amplifier is 1 x 0.75 mm (see photograph above). The design includes two GaAs FET feedback stages, matching networks and DC bias circuitry. Of special interest is the DC bias circuit. Transistors are biased in series and provide amplification in cascade so that the circuit operates with ultra low power consumption.

The CMM-2 amplifier is fabricated in-house using a standard 0.5 micron GaAs process. The process uses ion-implanted active layers, Ti/Pt/Au metalization and silicon nitride passivation. On-chip circuit elements include mesa resistors, air bridge inductors and silicon nitride dielectric capacitors. Two inch GaAs wafers are ion-implanted and processed in a fully equipped Class 100 fabrication facility.

Finished wafers are 100 percent DC tested and then sample tested for RF performance. Amplifiers are available in chips and chips mounted in ceramic packages. Because of low SWR, the amplifiers are easily cascaded to provide higher gain.

Temperature performance of GaAs FETs and GaAs MMICs is wafer-dependent. Customers have the option of specifying chips from a single wafer; as a result, users of these chips can expect close matching of gain characteristics over temperature. In addition, phase matching is very consistent. Amplifier gain remains relatively level over temperature changes. (see Figure 1).

Model CMM-2 chips are available in 30 days in quantities of 100 or less. Larger quantities are available in 30 to 60 days. Price is under $450 each, depending on quantity.

Celeritek, San Jose, CA (408) 433-0335.

Circle No. 281

1 to 8 GHz Microwave Circuit Building Blocks

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The MICRO-S series of GaAs integrated circuits is designed to simplify the construction of microwave systems operating in the 1 to 8 GHz frequency range. The line consists of amplifiers, switches, power splitters and attenuators.

Triquint also has developed (with cooperation from its parent company, Tektronix) a component mount, shown in Figure 1. The MICRO-S series surface-mount package is designed for frequencies to 18 GHz and will accept all of the MICRO-S microwave component chips.

In effect, the new series provides the basis for a 'microwave glue' inter-stage component set, which will be used much like the 'digital glue' components that tie together diverse TTL and CMOS digital circuits today. By reducing the number of chips and interconnections required to build a given microwave circuit, the new series will reduce manufacturing costs and improve microwave circuit reliability. The two components shown on either side of the chip in Figure 1 are capacitors, part of the DC power input circuitry.

The Model TQ9111 1 to 8 GHz Amplifier

This MIC amplifier has a flat gain of 7.5±0.25 dB (typical) input and output SWR of 1.5 (typical), 2 (max.) and a typical noise figure of 4.5 dB at 2 GHz. Typical reverse isolation is 23 dB. Gain and loss characteristics are shown in Figures 2 and 3.

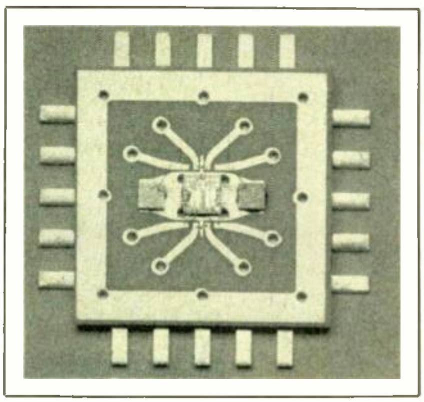


Fig. 1 18 GHz surface-mount package.

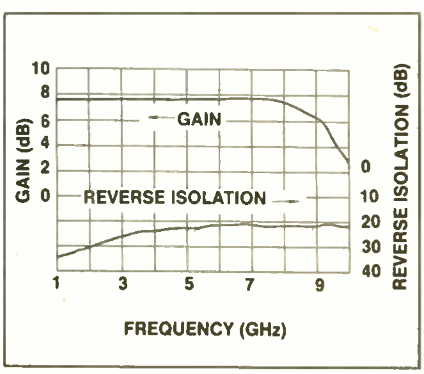


Fig. 2 TQ9111 gain and reverse isolation characteristics.

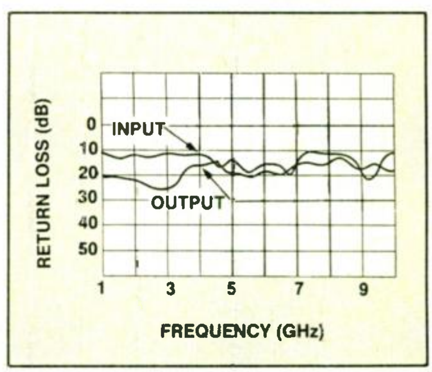


Fig. 3 TQ9111 return loss characteristics.

The Model TQ9161 1 to 10 GHz Attenuator

This MIC variable attenuator operates over the 1 to 10 GHz frequency range. It has less than 2 dB insertion loss, greater than 10 dB attenuation range and an input/output SWR of less than 2. Switching speed is less than 500 ns.

The TQ9161 is a voltage-controlled variable absorptive attenuator designed for gain compensation/control and leveling loop applications. Internal circuitry controls I/O return loss as attenuation is varied. The attenuator is available mounted in the surface-mount

[Continued on page 206]

package described above or in die form. Die size is 43 x 54 mils. Typical TQ9161 packaged performance is shown in Figures 4 and 5.

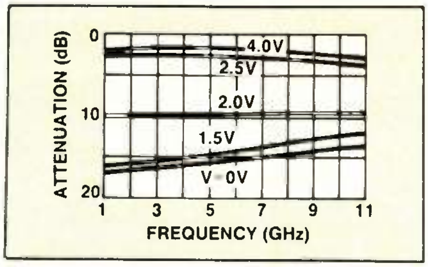


Fig. 4 TQ9161 attenuation vs frequency.

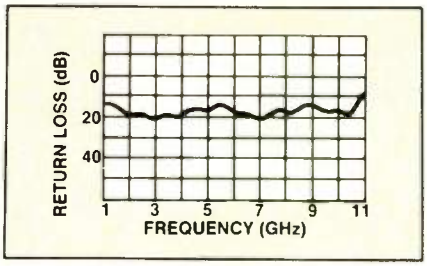


Fig. 5 TQ9161 input/output return loss.

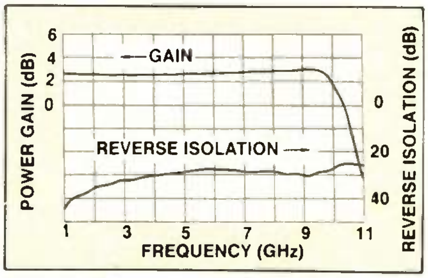


Fig. 6 TQ9141 power gain and reverse isolation vs frequency.

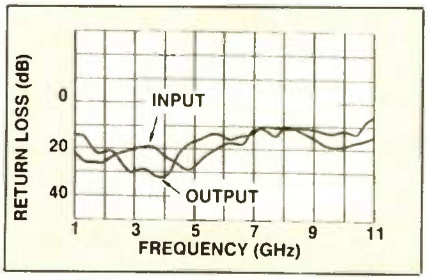


Fig. 7 TQ9141 input and output return loss vs frequency.

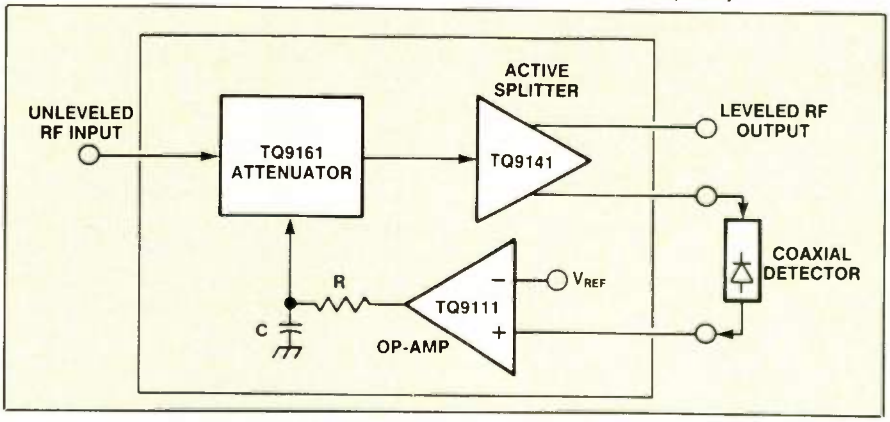


Fig. 8 Block diagram of leveling loop.

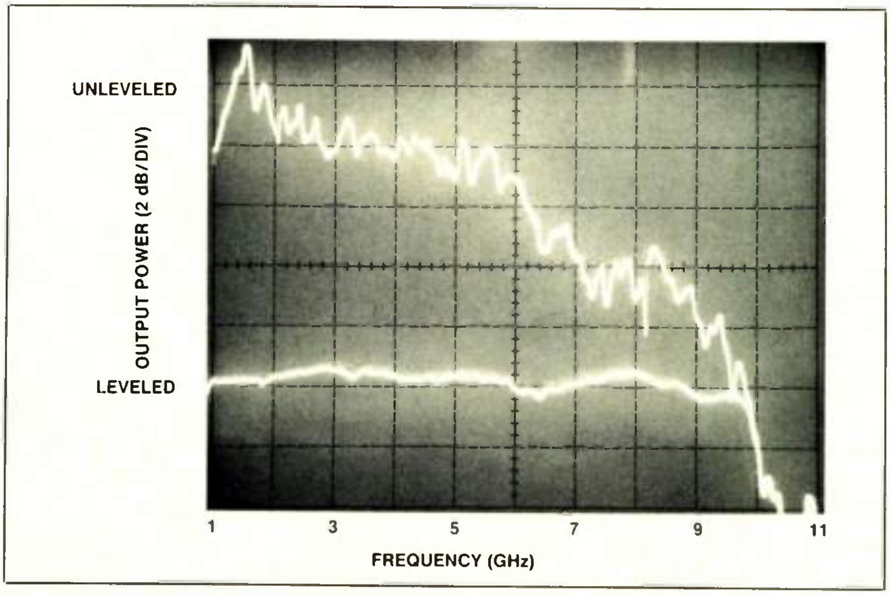


Fig. 9 Leveled and unleveled output circuit performance.

The TQ9141
1 to 10 GHz Active Power Divider

The TQ9141 is a general-purpose cascadable power divider providing both in-phase power division and gain. As an active device, it provides higher reverse isolation than is possible with passive power splitters. Positive gain slope is designed into the part to compensate for external losses incurred with hybrid substrates and packages. The power divider is available mounted in the surface-mount package or in die form. Die size is 54 x 54 mils.

The power divider has a bandwidth from 1 to 10 GHz, 2 dB gain and more than 10 dB output power at the 1 dB compression points. Input and output SWR is less than 2, amplitude balance is better than ±0.5 dB and phase balance is less than ±2°. Typical TQ9141 packaged performance is shown in Figures 6 and 7.

Microwave Leveling Loop

The block diagram in Figure 8 demonstrates how the MICRO-S series building blocks, described above, can be used to build a compact, low cost leveling loop. The leveling loop's performance characteristics are illustrated in Figure 9, which depicts the input circuit's unleveled output and the wideband performance of the leveling loop. The chip components described above produced constant power output within ±0.5 dB over the 1 to 10 GHz frequency range.

In die form for a quantity of 100: TQ9111 amplifier price is $60; TQ9141 power divider, $70; TQ9151 SPDT switch, not described in this article, $51; and the TQ9161 variable attenuator, $70. Corresponding prices for devices packaged in the 18 GHz surface-mount package are: TQ9111 ($144), TQ9141 ($155), TQ9151 ($143) and TQ9161 ($155). Delivery is 30 to 45 days ARO.

TriQuint Semiconductor Inc., Beaverton, OR, Dennis Powers (503) 629-4227.

Circle No. 280

CIRCLE 158 ON READER SERVICE CARD

Microwave Products

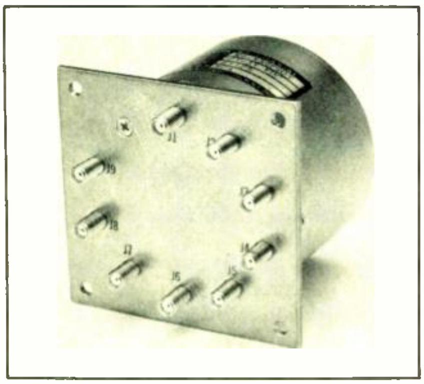
Components

0.1 - 20 GHz Chip Attenuators

This series of microwave chip attenuators covers the 0.1 to 20 GHz frequency range. Attenuation values may be customer-specified. A typical 13 dB attenuator has ±0.1 dB amplitude flatness from 0.1 to 12 GHz, ±0.25 dB flatness from 12 to 18 GHz and ±0.4 dB flatness from 18 to 20 GHz. Return loss is greater than 18 dB from 0.1 to 12 GHz and greater than 12 dB from 12 to 20 GHz. Size: 0.92" x 0.45" x 0.1". **Amitron, North Andover, MA (617) 686-1882.**

Circle No. 201

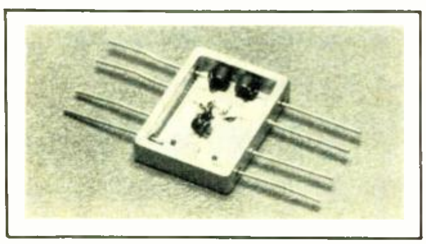
DC - 3 GHz Programmable Coaxial Switch



The model N9-439F903 latching type programmable switch operates from DC to 3 GHz over the -54 to +100°C temperature range. The switch has indicator and surge suppression diodes and is is of nitrogen-filled design. Delivery: 90 days. **Dynatech, Venice, CA, Evert Kjellberg (213) 392-9821.**

Circle No. 205

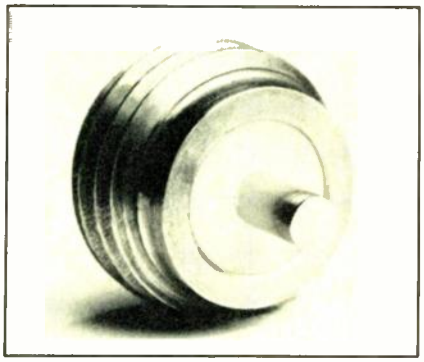
500 MHz - 2 GHz Electronic Phase Shifters



The PSEF-3A series of electronic phase shifters is available with center frequencies between 500 MHz and 2 GHz with minimum bandwidths of 10% and maximum insertion loss of 3 dB. Phase adjustment is continuous between 0 and 180° using a 0 to 30 VDC control voltage. The hermetically sealed flat-packs measure 3/8" x 1/2". Price: from $150 plus setup charges (small quantities). Delivery: within 8 weeks. **Merrimac Industries Inc., West Caldwell, NJ (201) 575-1300.**

Circle No. 212

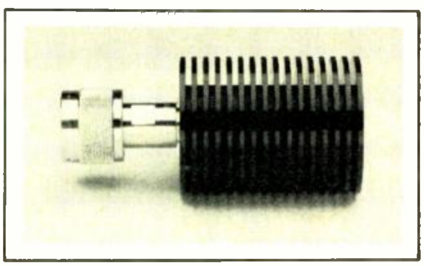
DC - 18 GHz Screw-In, Flush-Mount Termination



The model 4910 screw-in flush-mount termination operates from DC to 18 GHz and can dissipate 1 W of power at 60°C. SWR is 1.1 (max.) at 4 GHz and 1.3 (max.) at 18 GHz. Characteristic impedance is 50 ohms. The terminations are either gold-plated or passivated stainless steel with a gold-plated beryllium copper pin. Price: $9 (100). Delivery: stock to 8 weeks. **EMC Technology Inc., Cherry Hill, NJ (609) 429-7800.**

Circle No. 207

DC - 18 GHz Terminations



This line of 50 and 75 ohm terminations covers the DC to 18 GHz frequency range. Power handling capabilities range from 1 to 30 W and higher. Standard connectors include SMA, TNC, BNC and type N. Price: from $10.50 to $65, depending on frequency range and power handling requirements. Delivery: stock to 4 weeks ARO. No engineering fee for special designs. **JFW Industries Inc., Indianapolis, IN (317) 887-1340.**

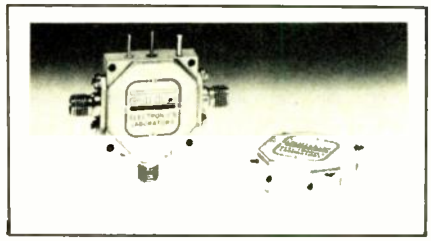
Circle No. 210

8 - 12.4 GHz 2-Way Power Divider

The model P8206-2 two-way power divider operates over the 8 to 12.4 GHz frequency band. SWR is 1.35 (max.) at all ports, insertion loss is 0.5 dB (max.) and isolation is 20 dB (min.). Amplitude imbalance is 0.2 dB (max.) and phase imbalance is 4° (max.). Size: 1" x 1" x 0.5". Connectors are SMA female. Delivery: stock to 90 days. **MAC Technology, Klamath Falls, OR (503) 883-3352.**

Circle No. 211

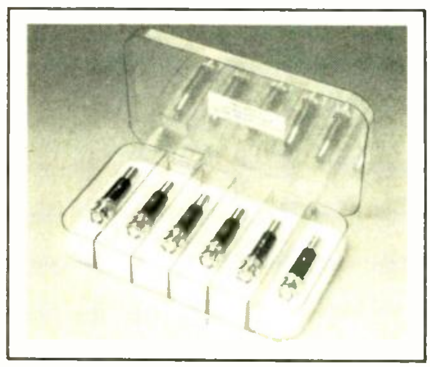
1 - 26 GHz Biasable Mixers



The DBL and DBLX series of hermetically sealed, ultra-miniature biasable mixers covers the 1 to 26 GHz RF frequency range. The DBL series covers the RF range to 26 GHz and has an IF response from 1 to 500 MHz. IFs from 0.01 to 8 GHz are available. Typical DBL1-12 models have RF/LO frequency coverage from 1 to 12 GHz and IF ranges from 1 to 500 MHz. Conversion loss is 9 dB at 0 dBm injection LO, 13 dB at -10 dBm LO injection. The DBLX2-18 model has an RF coverage of 2 to 18 GHz, 0.01 to 2 GHz IF coverage and 10 dB conversion loss at 0 dBm injection LO, 16 dB at -10 dBm LO injection. Price: DBL1-12, $895; DBLX2-18, $1,050. **RHG Electronics Laboratory Inc., Deer Park, NY (516) 242-1100.**

Circle No. 220

DC - 1.5 GHz Attenuator Kit



The RFA-4057-1 six-piece attenuator kit operates over the DC to 1.5 GHz frequency range. It includes six 1 W BNC in-line attenuators with values of 1, 2, 3, 6, 10 and 20 dB. Accuracy is ±0.2 to 0.6 dB. BNC fittings are silver-plated with gold pins. Housing is an unbreakable storage case. A 40 dB attenuator is available separately. Introductory price: $59.95 before Dec. 31, 1986. Regular price: $69.95. **RF Industries, Hialeah, FL (800) 233-1728.**

Circle No. 219

[Continued on page 210]

Mesa Beam Lead PIN Diodes

Beamed up and ready to go!

- *Rugged Construction*
- *Low capacitance*
- *Fast switching*
- *Oxide and polyimide passivation*

Metelics Mesa PIN Diodes are tough... tough enough to meet the challenge. Our new production process is carefully monitored to provide outstanding QUALITY and HIGH PERFORMANCE from all Metelics diodes. Our new ruggedly constructed Mesa Beam Lead PINs are ready to take on your application!

Preliminary Specifications @ 25°C

Model-Outline	V_{Br} @ I_R = 10μA (Volts) Typ.	R_s* (Ohms) Typ.	C_{TO}** (pF) Typ.	τ (nsec) Typ.
MPB 1030-B11	100	4.0	.020	80
MPB 1034-B11	100	3.0	.030	80

*Series resistance is calculated from insertion loss at 3 GHz, 50 mA
**Total capacitance is calculated from isolation swept over 2-18 GHz, zero bias

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TWX: 910-339-9537
FSCM & CAGE 59365

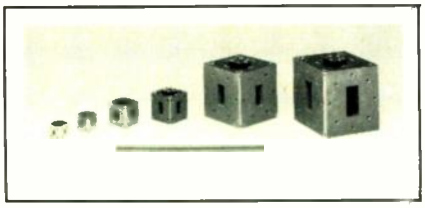
Microwave Products

0.5 - 18 GHz Switch Matrices

The RM100 series of modular switch matrices operates over the 0.5 to 18 GHz frequency range. Matrices are available in 3 x 8 to 24 x 64 configurations. They are built in a modular design and can be expanded in the field to any size up to 24 x 64 without degradation. The matrices include amplifiers so that no attenuation is introduced. Remote or computer control by means of an external RS-422 interface is standard; other interfaces are available. **Elsin Corp., Santa Clara, CA (408) 748-9900.**

Circle No. 206

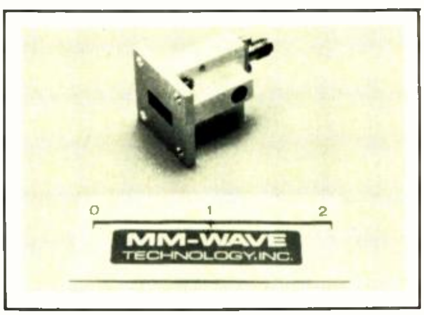
Miniature Waveguide Switches



This series of fast-latching miniature switches covers the waveguide bands from WR284 to WR28. A typical example is the WR62 switch, operating over the 12.4 to 18 GHz frequency band. Isolation is 80 dB, insertion loss is 0.1 dB and switching time is 20 ms. Life time is 100 million operations. Size: 38.1 x 38.1 x 41.5 mm. Weight: 140 g. **Forem Microwave Components, Caponago (Milano), Italy, tel. 02/9586397.**

Circle No. 276

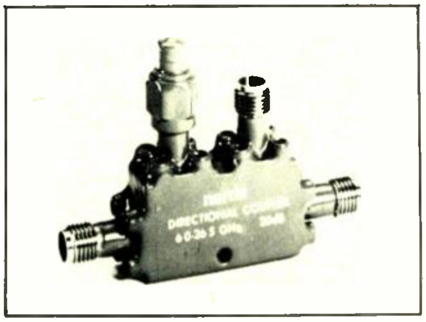
2 - 50 GHz Harmonic-Reject Filters



This series of harmonic frequency-reject filters can have cutoff frequencies from 2 to 50 GHz. The filters have a Chebyshev lowpass design in suspended substrate stripline. Input and output can be SMA connectors, K connectors or waveguide, depending on frequency band. Delivery: 4 weeks. **MM-Wave Technology, Torrance, CA (213) 212-7200.**

Circle No. 215

6 - 26.5 GHz Miniature Coupler



The model 4247-20 miniature coupler operates over the 6 to 26.5 GHz frequency range with flat frequency response. Coupling is 20 dB (±1 dB). Insertion loss is 0.7 dB, directivity is 13 dB (min.), average reflected power is 20 W and peak power handling capability is 0.2 kW. Weight: 1 oz. **Narda Microwave Corp., Hauppauge, NY (516) 231-1700.**

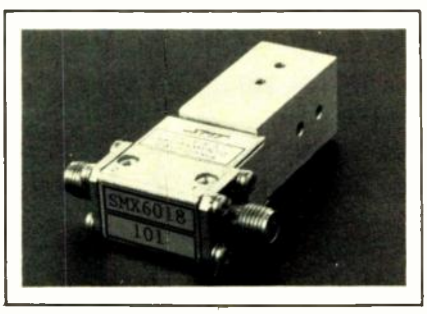
Circle No. 217

6 - 18 GHz Mixer Preamplifier

The model FP-1353 mixer covers the 6 to 18 GHz frequency range and has an IF range from 30 to 1,500 MHz. Noise figure is 9 dB from 6 to 12 GHz, 11 dB from 12 to 18 GHz. Conversion gain is 23 dB (min.) from 30 to 500 MHz IF and 10 dB (min.) from 500 to 1,500 MHz IF. Isolation is 22 dB (min.) LO to RF, 15 dB (min.) LO to IF and 20 dB (min.) RF to IF. Size: drop-in package, 0.56" D x 0.19"; connectorized package, 0.75" x 0.75" x 0.32". **Triangle Microwave, East Hanover, NJ, Gary Plowman (201) 884-1423.**

Circle No. 228

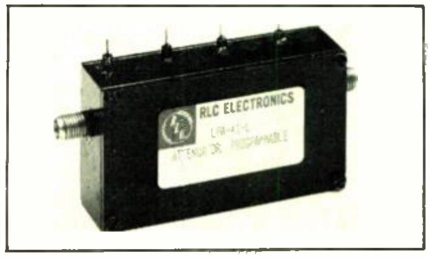
6 - 18 GHz High Power Isolator



The model SMX 6018 isolator operates in the 6 to 18 GHz frequency range. It handles reverse power to 50 W at 100°C. Isolation is 14 dB (min.) and insertion loss is 1 dB (max.). SWR is 1.5 and operating temperature range is -40 to +100°C. Price: $295 (1-4). Delivery: 6 weeks ARO. **Sierra Microwave Technology, Rancho Cordova, CA (916) 638-2002.**

Circle No. 223

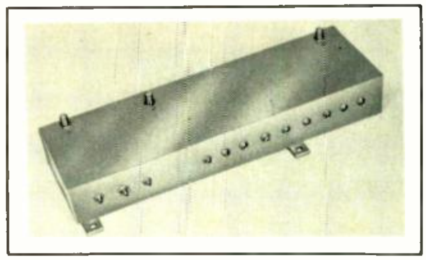
DC - 1,000 MHz Step Attenuators



The LPA miniature programmable step attenuators are available in four standard models and operate from DC to 1,000 MHz. Up to 63 dB of attenuation is available in 1 dB steps or 70 dB in 10 dB steps. Connectors may be SMA, BNC, TNC or type N. Coil voltages are 1.5 to 48 V with either latching or failsafe operation. Price: from $180. Delivery: stock to 10 weeks. **RLC Electronics Inc., Mount Kisco, NY (914) 241-1334.**

Circle No. 221

TACAN Frequency Band Diplexer



The model FF3400 diplexer meets FAA requirements for medium power TACAN band operations. The transmitter channel at 1,030 MHz has a midband loss of 0.3 dB (typical), 3 dB bandwidth of 15 MHz (min.) and 10 dB bandwidth of 45 MHz (max.). Rejection is 25 dB (min.) at 1,090 MHz. The 1,090 MHz channel has 0.3 dB typical midband loss, ±0.25 dB (max.) amplitude and ±3.5° phase ripple over a 6 MHz bandwidth, and ±1 dB (max.) amplitude ripple over a 28 MHz bandwidth. Insertion phase tracks within 0.5° over temperature from unit to unit. Rejection is 40 dB (min.) from 225 to 1,035 MHz and from 1,150 to 1,300 MHz. Both channels are rated for 100 W peak, 10 W average power. Size: 10.4" x 4.1" x 1.8" plus SMA connectors. Weight: 1.2 lb. **Sage Laboratories Inc., Natick, MA (617) 653-0844.**

Circle No. 222

Contiguous Quadraplexers

This line of contiguous band quadraplexers divides a specified frequency range into four bands. Typical insertion loss is 1 dB (max.) in the passbands, 4.5 dB at crossover frequencies. Selectivity is 55 dB (min.), 60 dB (special request). The quadraplexers are used to combine and separate microwave signals. **Time Microwave, San Jose, CA, Carl Martens (408) 434-6699.**

Circle No. 227

[Continued on page 212]

Microwave Products

2 - 18 GHz Coaxial Harmonic Absorption Lowpass Filters

The C70-A series of coaxial harmonic absorption lowpass filters is available with center frequencies in the 2 to 18 GHz frequency range and with stopband from 0.8 GHz center frequency to three times the center frequency. Passband insertion loss is from 0.3 to 1 dB and SWR is less than 1.5. Power handling is several hundred hundred watts. These filters do not reflect stopband signals back to the transmitter but absorb them internally, maintaining an SWR of 2 to 2.5 over the entire reject band. Size: 5" x 7" x 1". **Microwave Engineering Corp., Andover, MA (617) 685-2776.**

Circle No. 213

8 and 10 ns Switch Drivers

The DS-8002 series of switch drivers can operated at 10 MHz repetition rates from +5, -15 V power supplies at 8 to 10 ns (inverting/non-inverting) switching speeds. 30 mA output currents with 200 mA switching spikes are standard. Metal packages are seam-welded and units are screened to MIL-883B specifications. Price: $55 (volume), $75 (2). **Barry Industries Inc., TRX Div., Attleboro, MA (617) 226-3350.**

Circle No. 203

8 - 20 GHz Isolator

The model T-8S83U-10 isolator operates from 8 to 20 GHz. Isolation is 14 dB, insertion loss is 1.1 dB and SWR is 1.5 (max.). Operation is over full military temperature range. SMA female connectors are supplied but any combination of SMA female or male connectors is available. Magnetic and EMI shielding and 100% humidity requirements are available. **Teledyne Microwave, Mountain View, CA (415) 968-2211.**

Circle No. 225

DC - 18 GHz SMA Terminations

This series of coaxial terminations operates from DC to 18 GHz. The model 3900-9 miniature SMA plug termination has a maximum SWR of 1.15 to 18 GHz and is rated at 1 W average power. The female termination is the model 3900F-9. Construction is passivated stainless steel per QQ-S-764 and the beryllium copper center conductor is gold-plated per MIL-G-45204. Connector meets MIL-C-39012 requirements. Delivery: from stock. **Coaxial Components Corp., Huntington, NY, Janice Zaitz (516) 864-4747.**

Circle No. 204

DC - 18 GHz TNC and N Adapters and Connectors

This series of precision TNC and type N adapters and connectors, in bulkhead and 90° sweep designs, operates from DC to 18 GHz. The connectors are manufactured from stainless steel and silicon bronze to MIL-C-39012 and MIL-A-55339 specifications. **Astrolab, Warren, NJ (201) 560-3800.**

Circle No. 202

Coaxial Adapter Cables

The TPI-5000 kit includes 20 cables, each a different combination of BNC, TNC, N, UHF and mini-UHF connectors. Special RG-58 A/U cables are soft and easy to handle. Cables are bright yellow. Connectors are machined brass with silver-plated contacts. Furnished with two wall racks. Price: $125. **Test Probes Inc., La Jolla, CA (800) 368-5719 or, in CA (800) 643-8382.**

Circle No. 226

[Continued on page 214]

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Example:

6-18 GHz Isolator

Model 60A2601 Isolator

Frequency:	6-18 GHz
Temperature:	−40°C to +80°C
Isolation:	12 dB minimum
Insertion Loss:	1.0 dB maximum
VSWR:	1.67 maximum
Input Power:	100 watts peak, 1 watt average
Size:	.5 x .5 x .375 inches nominal, excluding projections
Weight:	0.6 ounces, nominal
Connectors:	SMA female

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Amplifiers

8.5 - 9.6 GHz GaAs FET Amplifier

The model CMA44450 GaAs FET amplifier operates in the 8.5 to 9.6 GHz frequency range. Peak pulse output power is 25 W over a 25 μs pulse width. Gain is 50 dB and rise time is less than 500 ns. **Alpha Industries, Central Microwave Co., Maryland Heights, MO, Bill Ruff (314) 291-5270.**

Circle No. 232

1 - 10 GHz Decade Bandwidth Amplifiers

This line of decade bandwidth amplifiers covers the 1 to 10 GHz frequency range in two power levels, +12 and +18 dBm. Gain options are 10 to 35 dB, input/output SWR is 2 (max.), noise figure is 6 dB (max.) and typical gain ripple is ±0.35 dB per 10 dB of gain. Temperature compensation is from −54 to +95°C (optional). Package size: 0.22" H x 0.99" W with removable SMA connectors; length varies from 1.32" to 2.32". Delivery: 45 days ARO. **Veritech Microwave Inc., South Plainfield, NJ, R. Stegens (201) 769-0300.**

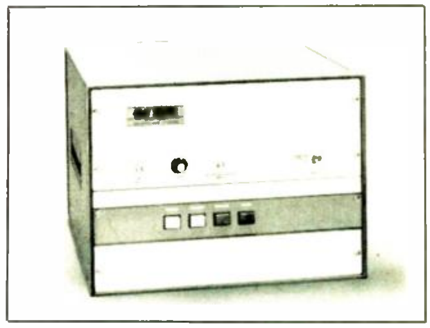
Circle No. 241

Ku-Band Power Amplifiers

The 11000 series of Ku-band power amplifiers covers the 11.7 to 14.5 GHz frequency range with nine models. Output power at the 1 dB compression point is +20 to +30 dBm and gain is 15 to 55 dB. Gain flatness is ±1 dB (max.) at gains from 15 to 20 dB, ±1.5 dB (max.) from 35 to 40 dB gain and ±2 dB (max.) at 55 dB gain. SWR is 2 (max.) and noise figure is less than 10 dB. Connectors are SMA female (standard). WR75 cover flange on output is optional. TTL gain control, output signal sampling and AC power are available. **Ramar Communications Inc., Chatsworth, CA (818) 341-7702.**

Circle No. 239

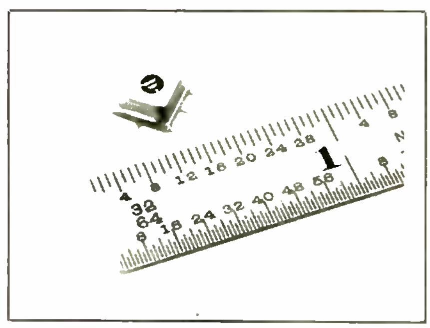
2 - 150 MHz, 1000 W RF Pulse Amplifier



The model 1000LP pulse amplifier delivers 1000 W of linear pulse power on up to 10% cuty cycle, 8 ms (max.) pulse in the 2 to 150 MHz frequency range. Fast blanking permits sampling of NMR signals. In addition to pulse operation, CW operation may be selected at power levels up to 200 W. Pulse operation requires a TTL gate pulse, synchronized with the RF input signal. Size: 50.3 x 38.1 x 55.1 cm. Weight: 82 kg. Price: $13,500. Delivery: 12 weeks ARO. **Amplifier Research, Souderton, PA (215) 723-8181.**

Circle No. 233

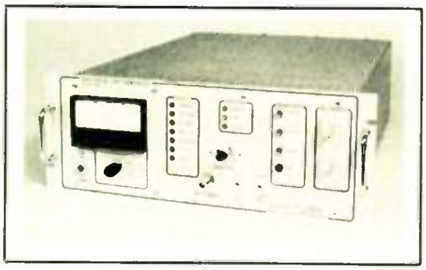
5 - 200 MHz Surface-Mount Amplifier



The model PPA-253 surface-mount amplifier operates over the 5 to 200 MHz frequency range. Gain is 29 dB (min.), 32 dB (typical) and noise figure is 4 dB (max.), 3 dB (typical). Output power at the 1 dB compression point is 0 dBm (min.), +3 dBm (typical). SWR is 2 (max.), 1.2/1.5 input/output (typical). Operating temperature range is 0 to +50°C; slightly reduced performance from −55 to +85°C. Size: 1/4" square. Weight: 0.25 g. **Avantek Inc., Santa Clara, CA (408) 970-2583.**

Circle No. 234

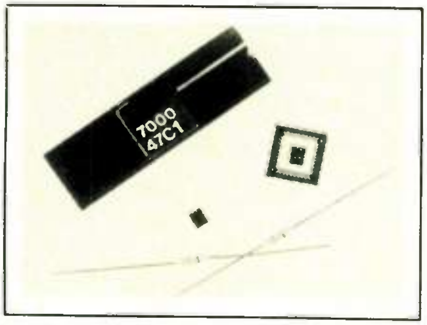
High Power Ku-Band Amplifiers



The R60 series of amplifiers is available in power levels from 160 to 300 W at Ku band. The amplifiers are packaged to fit in a 7" H rack chassis. Weight: 70 lb. **Keltec Florida, Shalimar, FL, Mark Yount (904) 651-9749.**

Circle No. 236

105 dB Dynamic Range Log Amplifier IC



The model M-7000 log amplifier IC contains seven independent logging stages. Each has a user-definable dynamic range of 6 to 17 dB. The stages may be cascaded together and summed through either a unipolarity amplifier or a differential amplifier for a system dynamic range of 105 dB. The amplifier also has a dual tracking voltage regulator, a bandgap reference, a precision die temperature sensor and system definable trim points. Price: $27 (up to 100 commercial, 40 pin plastic dip), $126 (up to 100 for MIL-STD-883C). **Megadyne Corp., Fairfax, VA (703) 280-5232.**

Circle No. 237

2 - 6 GHz FET Amplifiers

This series of miniature IF amplifiers operates from 2 to 6 GHz. Output power is +27 dBm (min.) at the 1 dB gain compression point. Temperature compensation is from −54 to +95°C. Gain flatness over frequency and temperature is ±1.5 dB (max.); noise figure with nominal gain of 40 dB is 3.5 dB (max.) at 25°C and 4.5 dB (max.) at +95°C. Package size: 0.32" x 0.66" x 2.5". **TRW Microwave Inc., Sunnyvale, CA (408) 732-0880.**

Circle No. 240

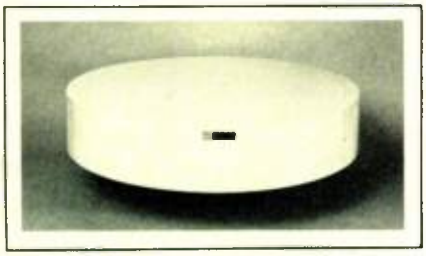
Ku-Band LNAs

The KLA series of low noise amplifiers covers the 10.95 to 12.75 GHz frequency range with three units. The KLA 1130 operates from 10.95 to 11.7 GHz, the KLA 1200 operates from 11.7 to 12.2 GHz and the KLA 1250 operates from 12.25 to 12.75 GHz. Nominal gain is 50 dB (40 dB optional), input SWR is 1.25 (max.) and output SWR is 1.5 (max.). Gain ripple is ±1 dB (max.); gain compression is 1 dB at +7 dBm output. Gain slope is ±0.25 (max.)/40 MHz. Third-order intercept is +15 dBm (min.). Noise temperatures from 155 K to 225 K are available. Operating temperature range is −40 to +60°C. Size: 6.9" x 1.9" x 1.5" (excluding connector). **Microwave Systems Engineering Inc., Phoenix, AZ, Steve Maziarz (602) 437-9040.**

Circle No. 238

Antenna

0.5 - 18 GHz Antenna Array

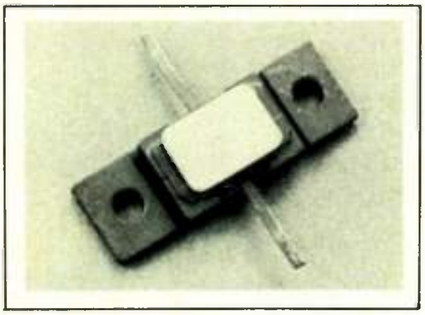


The LPR100A antenna assembly consists of five antennas covering the 0.5 to 18 GHz frequency range. The lowband antenna array covers the 0.5 to 1.75 GHz frequency range and is mounted back-to-back to the octave bandwidth, higher frequency antennas. The highband antennas occupy a minimum of volume, have narrow azimuth and have broad elevation beamwidths. The entire subsystem, including pedestal, is contained within a 32" D x 9" H cylinder. With an additional 1" for a radome and antenna clearance, the result is a maximum dimension below a surface of 10". The subsystem can be controlled by EC60 or EC70 series spinning DF controllers. **Watkins-Johnson Co., SSE Division Applications Engineering, San Jose, CA (408) 435-1400.**

Circle No. 231

Devices

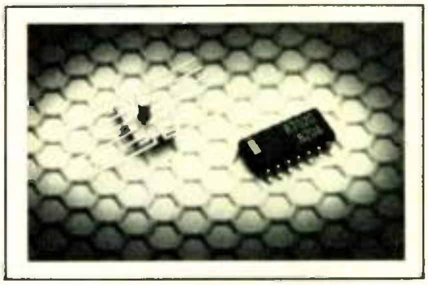
S-Band GaAs Power FETs



The MSC 0200 series of GaAs power FETs provides more than 14 W of power at the 1 dB compression point at a gain of 8 to 9 dB and a frequency of 2 GHz. Power-added efficiency is 30%. Price (100): from $200 to $250. Delivery: 4-8 weeks, sample quantities. **Microwave Semiconductor Corp., Somerset, NJ (201) 563-6300.**

Circle No. 245

9 GHz Transistor Arrays



These microwave logic devices (UPA101, 102, 103, 104) contain silicon transistor arrays that can be configured to meet various design needs. Connecting the pins in different combinations provides a variety of functions ranging from double-balanced mixer to high speed logic gates. These high frequency arrays can be used in equipment such as VLSI testers, new generation computers, precision timing systems and intelligent radar. The UPA101 contains six 9 GHz transistors configured as an active o ... -balanced mixer or doubler; the UPA1C ... tains two pairs of 9 GHz transistors co ... d differentially with a common bias ... istor for each pair; the UPA103 contain ... pair of 9 GHz transistors in differentia ... three individual 9 GHz transistors; th ... PA104 contains one 9 GHz transistor in ... llel to a differential pair and two individu ... GHz transistors. Two package types are a ... ble: Package B is a 14-pin ceramic packa ... with good thermal dissipation; Package G is ... 4-pin (8-pin for UPA101) mini-flatpack ... th 35% size reduction. **California Eastern L ... oratories, Santa Clara, CA (408) 988-350 ...**

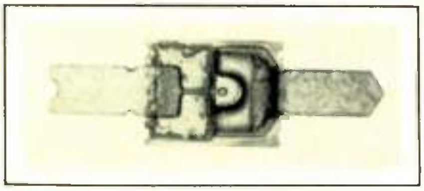
Circle No. 2...

500 MHz - 5 GHz GaAs Amplifier Chips

The HMR-10502 and HMR-10503 GaAs cascadable amplifier chips provide 10 dB gain with ±0.75 dB gain flatness over the 500 MHz to 5 GHz frequency range. The HMR-10503 is a directly cascadable broadband amplifier chip that does not require external DC blocking on the RF input or output ports. The chip includes two GaAs FET gain stages using negative feedback, with active and passive bias circuitry within 1.15 x 1.45 mm chip dimensions. The HMR-10502 is similar to the HMR-10503, but without internal source bypass circuitry. Input and output SWR is better than 2 for both amplifier types. Price (1000): HMR-10502, $29; HMR-10503, $39. Delivery: 3 weeks ARO. **Harris Microwave Semiconductor, Milpitas, CA, Bruce Hoffman (408) 262-2222.**

Circle No. 243

Beam-Lead Diodes



This series of mesa beam-lead diodes has 6 g typical beam strength. Maximum ratings include total power dissipation of 250 mW at 25°, operating temperature range of –65 to +175°C and storage temperature range of –65 to +200°C. V_{BR} at I_R = 10 μA is 100 V (min.) at 25°C. Series resistance, calculated from insertion loss at 3 GHz, 50 mA is 3.0 to 4.0 ohms (max.). Capacitance, calculated from isolation at 18 GHz, is 0.2 to 0.25 pF (typical) at V_R = 0 V. **Metelics Corp., Sunnyvale, CA (408) 737-8181.**

Circle No. 244

1 - 50 GHz AlGaAs/GaAs, Low Noise Microwave Transistor Chip

The 2SK676H5 AlGaAs/GaAs high electron mobility transistor (HEMT) chip operates in the frequency range from 1 to over 50 GHz. The chip is fabricated by metal organic chemical vapor deposition. Chip thickness is 200 μm and pad and backside metalization is 4500 angstroms gold. **Sony Component Product Division, Sony Corp. of America, Torrance, CA, Kou Togashi (213) 373-9425.**

Circle No. 247

Sources

75 - 96 GHz Source-Locked Gunn Oscillator

This source-locked Gunn oscillator covers the 75 to 96 GHz output frequency range and has up to 2 GHz of locking bandwidth. Output power is up to 20 mW. The unit is GPIB controllable. A varactor voltage control box is used to interface the VCO to the locking counter. **Honeywell, Santa Barbara Microwave Center, Santa Barbara, CA (805) 965-1013.**

Circle No. 250

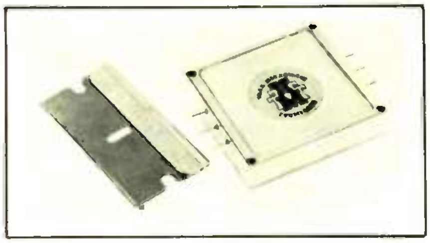
Microwave Products

500 MHz SAW Resonator Oscillators

The 812-0005-01/02/03 SAW resonator oscillators operate at a nominal frequency of 500 MHz with frequency tolerances of ±25 kHz (01), ±100 kHz (02) and ±250 kHz (03). Second harmonic signal level is -16 dBc (typical); third harmonic signal level is -20 dBc (typical). Minimum output power is +7 dBm, output SWR is 2 (max.) at 500 MHz and temperature coefficient is -2.5 ppm/°C (at greater than 25°C ambient temperature). Phase noise is -100 dBc at 0.2 kHz from oscillator frequency. Operating temperature range is -15 to +85°C. Long-term aging is 50 ppm in 10 years; short-term stability is 10^{-7} ($\Delta f/f_0$). Package is an 18 pin ceramic double wide DIP, hermetically sealed. Weight: 3.5 g. Meets MIL-STD-883, Class C. **Tektronix Inc., Beaverton, OR (503) 627-1299.**

Circle No. 252

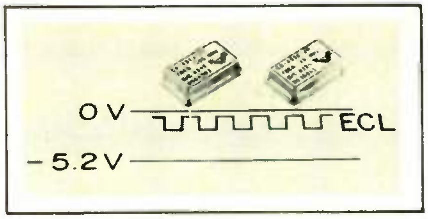
60 - 1,200 MHz Crystal Controlled Oscillator



This series of drop-in surface-mount/microstrip crystal oscillators covers the 60 to 1,200 MHz frequency range. Output power is 10 mW or 100 mW (min.); 1 W version is available over a limited temperature range. Frequency accuracy at room temperature is ±0.001%, frequency stability is ±0.003% (any temperature within -55 to +85°C) and aging rate is ±0.0005% per year. Power variation with temperature is ±2 dB. Spurious nonharmonic signals are more than 60 dBc in 1 MHz bandwidth. Maximum load SWR is 1.5. **TRAK Microwave Corp., Tampa, FL (813) 884-1411.**

Circle No. 254

5 - 200 MHz ECL Clock Oscillators



The CO-431 series of low profile DIP ECL clock oscillators is available at any frequency in the 5 to 200 MHz frequency range. Standard stability is ±25 ppm over 0 to +75°C. Options include ±50 ppm over the -55 to +125°C frequency range and ±5 ppm over the 0 to +50°C frequency range. Meets MIL-0-55310 requirements and meets MIL-STD-883 fine leak of 10^{-8} atm cc/s. Internal hybrid conforms to MIL-M-38510. Random vibration of 20 g to 2 kHz is met through the use of a rugged three-point crystal mount. Size: 0.2" (5.1 mm) H. Price: depends upon frequency and stability option. Delivery: stock to 10 weeks ARO, depending on frequency. **Vectron Laboratories Inc., Norwalk, CT, Larry Jawitz (203) 853-4433.**

Circle No. 255

Software

Microwave Circuit Analysis and Optimization Program

ANALOP 4.90 is a revised version of the ANALOP computer-aided microwave circuit analysis and optimization software, which operates on the IBM-PC/AT. In addition to the existing circuit optimization with dependently tuned discontinuities and graphical tracing, group delay analysis and optimization have been added to equalize user-supplied black-box data. The program features new circuit elements, such as first- and second-order all-pass bridge-T networks that tune with center frequency and shape factor. Also included are new elements such as tapped resonators. Price: $499. **ANALOP Engineering, Milpitas, CA (408) 942-8630.**

Circle No. 264

Ultra-Fast Version of CAD Program

Revision 1.9 is a vastly enhanced version of Super-Compact. It uses a dedicated sparse matrix technique. Standard practice has been to use the Gauss-Jordan matrix inversion in which the execution speed is essentially proportional to one over the square of the number of nodes. Each time the number of nodes is doubled, the execution speed of the program goes down to 1/4. Circuit execution may be performed in two-port or nodal analysis, the latter being necessary for three- and four-port circuits such as microwave couplers. By using a nodal sparse matrix technique, the execution time is essentially linearly proportional to the number of components — as fast as the two-port analysis. Users of the sparse matrix will, in some cases, show up to hundred-fold time savings. Revision 1.9 allows unlimited nodal analysis of arbitrary circuits. The software is being used in the Super-Compact PC and mainframe versions running on the HP 9000, series 500/300 computers, Apollo computers, Ridge computers, VAX computers and other large mainframe computers. **Compact Software Inc., Paterson, NJ (201) 881-1200.**

Circle No. 265

Microwave Layout and Drawing Program for MICs, Microstrip and Stripline Circuits

MICAD 1.05 is a microwave layout and drawing program for the preparation of camera-ready artwork, drawings and documentation associated with MICs, microstrip and stripline circuits. When used in conjuncion with the CAD program Touchstone, MICAD converts a Touchstone-generated circuit into a physical layout. Features not in previous MICAD versions include incorporation of all Touchstone 1.4 microstrip and stripline elements, additional hardware support, a geometric text editor and full disk operating system (DOS) pathname support. The autoprocessor supports Touchstone equations and SPICE syntax. Price: $8,400. Volume discounts available. **EEsof Inc., Westlake Village, CA (818) 991-7530.**

Circle No. 266

[Continued on page 220]

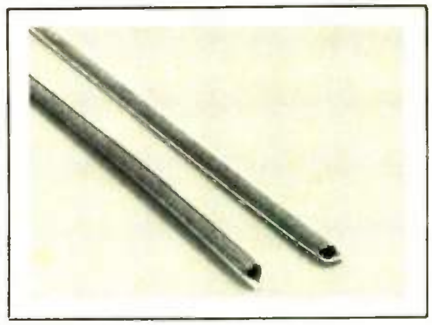
Hardware

Surface Probe with Replaceable Tip

This surface probe (MEL) has been developed for use with the model 750 fiber-optic temperature measurement system. The probe is designed for making surface temperature measurements. An extremely small mass of the phosphor coated over the elastomer responds quickly to temperature changes in its surroundings. Response time of a room temperature probe touching an aluminum block heated to 200°C is 25 ms. Temperature equilibrium is reached within 100 ms. The probe has a continuous working temperature range from –50 to +200°C. It can be used for short periods from –100 to +250°C without permanent damage. **Luxtron Corp., Mountain View, CA (415) 962-8110.**

Circle No. 270

Clip-On EMI Mesh Strip Gasket

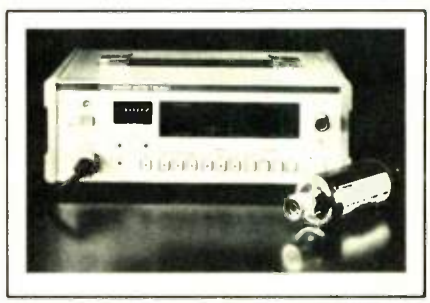


The MESHKLIP clip-on EMI mesh strip gasket makes possible quick, clip-on mounting of a wire mesh EMI gasket onto sheet metal, plastic or metalized cabinetry. MESHKLIP gasketing uses knitted wire mesh to provide repeatable shielding performance, with an elastomer core providing good compressibility and resilience. The clip passes UL94V-0 flammability tests. The standard clip profile can be modified to accommodate unusual flange designs. Size: standard 8′ lengths or custom-cut. **Chomerics Inc., Woburn, MA, Joan G. Green (617) 935-4850.**

Circle No. 269

Test Equipment

10 MHz - 140 GHz Digital Power Meter



The ML83A digital power meter operates from 10 MHz to 140 GHz and can be connected to any of its standard sensors without adjustment. Sensors include the MA72B (10 to 18 GHz), MA73A (10 to 18 GHz), MA81B1 (75 to 110 GHz), MP82B1 (90 to 140 GHz), MP712A (18 to 26.5 GHz); MP713A (26.5 to 40 GHz); MP714A1 (33 to 50 GHz) and MP715A (40 to 60 GHz). Price: from $2,140. Delivery: 3 weeks. **Anritsu America Inc., Oakland, NJ, Melissa Heck (800) 255-7234 or, in NJ, (201) 337-1111.**

Circle No. 258

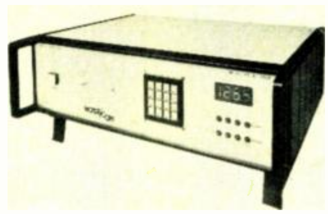
Microwave Products

30 MHz - 1 GHz Close Field Probe for EMI Measurements

The HP 11940A close field probe is a calibrated magnetic-field sensor that locates EM emission sources and makes repeatable relative measurements from 30 MHz to 1 GHz for EMI testing. The probe uses a dual-loop configuration and balun to provide common-mode rejection of electric field components. Calibration accuracy is ±2 dB in a 377 ohm field impedance. Since the probe provides frequency and amplitude information, its optimum use is with a spectrum analyzer for EMI characterization. No power supply is required; however, for local susceptibility measurements, a voltage can be fed into the probe to create a small magnetic field at the loops. Price: $500. Delivery: 12 weeks ARO. **Hewlett-Packard Co., Palo Alto, CA.**

Circle No. 261

10 Hz - 2 GHz Programmable Noise Generators



The NC 7107 series of noise generators is available with IEEE-488 (GPIB) MATE-CIIL protocol and covers the 10 Hz to 2 GHz frequency range with 11 models. White Gaussian noise signals of +10 dB or higher power into 50 ohm impedance is provided. Minimum crest factor is 5:1. Options are up to seven remotely switched filters for sub-bands, input signal modulation terminal and attenuation with 0.1 dB increments. Impedances other than 50 ohms may be specified. Price: $3,995. Delivery: 2 weeks ARO. **Noise Com Inc., Hackensack, NJ (201) 488-4144.**

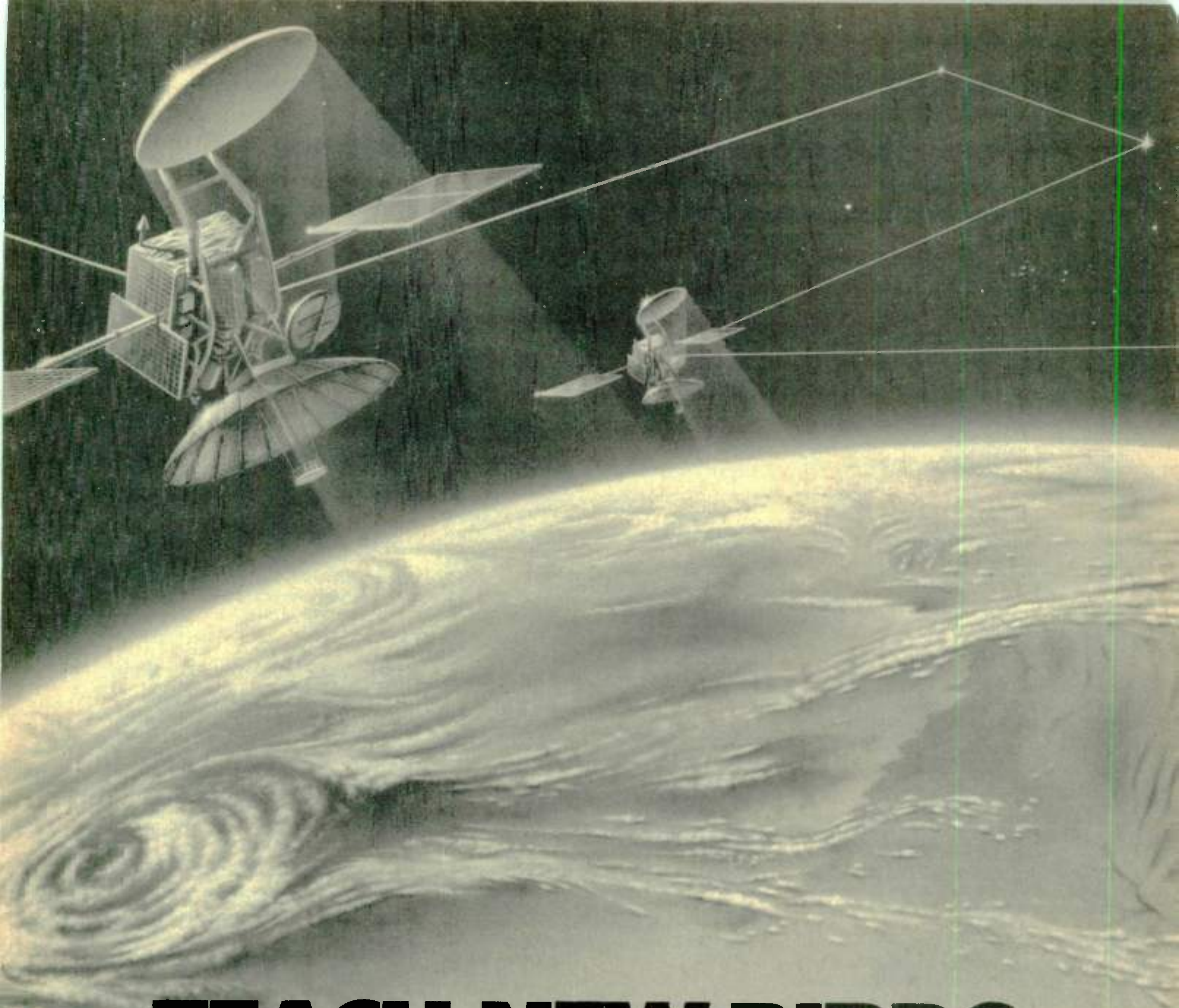
Circle No. 263

Errata:

The following list was omitted from the article "Design of Microstrip Receiver Supercomponents at mm-Wave Frequencies," published in the September 1986 *Microwave Journal*, beginning on p. 161.

References

1. C. Gupta, "Planar Waveguide Approach Optimizes mm-Wave Supercomponent Designs," *Microwave Systems News*, Dec. 1984.
2. P.J. Meir, "New Developments with Integrated Finline and Related Printed Millimeter Circuits," Digest of IEEE MTT-S Int. Microwave Symposium, pp. 143-145, May 1975.
3. J.A. Paul and Y.W. Change, "Millimeter-Wave Image Guide Integrated Passive Devices," IEEE Trans. Microwave Theory Tech., Vol. 26, pp. 751-754, Oct. 1978.
4. K.C. Gupta, R. Garg and I.J. Bahl, *Microstripline and Slotline*, Artech, 1979, pp. 1-193.
5. P. Bauhahn, et al, "94 GHz Planar GaAs Monolithic Balanced Mixer," Digest of IEEE MTT-S Int. Microwave Symposium, pp. 47-50, May 1984.
6. W.J. Gentsinger, "Microstrip Dispersion Model," IEEE Trans., Microwave Theory Tech., Vol. 21, pp. 34-49, 1973.
7. C. Gupta and A. Gopinath, "Capacitance Parameters of Discontinuities in Microstrip," IEEE Trans. Microwave Theory Tech., Vol. 26, pp. 831-836, 1978.
8. C. Gupta, "Design of Parallel Coupled Line Filter with Discontinuity Compensation in Microstrip," *Microwave Journal*, Dec. 1979.



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New Literature

SMA MIC Launcher Application Notes

These application notes discuss theoretical and practical design and application of SMA field-replaceable connectors used with hermetically sealed MIC packages. Connector design, connector/glass seal interface, coaxial line/microstrip interface, PIN-to-line attachment methods and test methods are discussed. Diagrams are provided. **Applied Engineering Products, New Haven, CT (203) 387-5282.**

Circle No. 317

Surface-Mounted Device Data Sheet

This data sheet contains information on SOT-23 packages with PINs, abrupt and hyperabrupt tuning varactors, multiplier varactors, capacitors and Schottky barrier diodes. Technical data is provided. **Alpha Industries Inc., Semiconductor Div., Woburn, MA (617) 935-5150.**

Circle No. 318

Microwave Semiconductor Capabilities Brochure

This 8-page brochure describes capabilities to design and produce microwave semiconductors, including bipolar transistors, GaAs FETs, diodes, hybrid ICs and monolithic microwave ICs. Relationship with affiliated Japanese company NEC Corp. is described. Customers, engineering, products, sales, chip processing and warehousing and service are discussed. Photographs are provided. **California Eastern Laboratories, Santa Clara, CA (408) 988-3500.**

Circle No. 320

Aluminum Dip Brazing/Precision Machining Capabilities Brochure

This 8-page brochure provides information on capabilities for aluminum dip brazing and contract machining. Technical discussions of aluminum dip brazing, its advantages, types of alloys that are dip brazeable, and heat treatment are included. A pictorial description of a precision machined assembly undergoing the dip brazing cycle is shown. Examples of NC capability, milling and turning centers, and the tolerances achieved on these machines are outlined. **Coleman Microwave Co., Edinburg, VA (703) 984-8848.**

Circle No. 321

Precision Etched Microcircuit Capabilities Brochure

This 8-page brochure describes capabilities for production of precision etched microcircuits for microwave applications. Brief discussions of metalized substrates, application of metallic films to ceramic substrates, film thickness vs power and frequency, film thickness vs work temperature, quality control, processing schedule, test data, cleanliness and proprietary control of specifications are provided. More information is given on deposition; dicing ceramic substrates; substrate preparation; chip, thin-film resistors; and chip capacitors. **Diablo Industries Inc., San Jose, CA (408) 436-0708.**

Circle No. 322

Noise Figure Application Note

This 4-page application note discusses the noise measurement of high frequency devices. Details are provided on the test configuration used in the 2075-2A noise-gain analyzer to measure the noise performance of receivers or mixers whose IF frequencies are greater than 1,850 MHz. A technique used for measuring the noise performance or any device under test at arbitrarily high input frequencies also is described. **Eaton Corp., Electronic Instrumentation Div., Los Angeles, CA (213) 822-3061.**

Circle No. 323

Microwave Frequency Generation and Control Product Catalog

This 42-page catalog provides information on products for microwave frequency generation and control, including semiconductors, microwave oscillators, multipliers, control devices and subsystems. Specifications, descriptions, photographs, block diagrams and performance graphs are provided. **Frequency Sources, Chelmsford, MA (617) 256-4113.**

Circle No. 324

Coaxial Connector and Microwave Component Distribution Data Sheet

This data sheet provides information on a service for distribution of Omni-Spectra coaxial connectors and microwave components. Service, inventory, technical expertise, pricing and stability are noted. A photograph is provided. **L-K-H Sales Inc., Haverhill, MA (617) 373-1313.**

Circle No. 325

Hi-Rel Capabilities Reference

This 20-page applications book contains information on hi-rel products, programs and capabilities. Signal processing components and supercomponents being used in current hi-rel and mil-screen programs are among the products described. **Merrimac Industries Inc., West Caldwell, NJ (201) 575-1300.**

Circle No. 327

Noise Figure Meter Brochure

This 4-page brochure provides information on the model 5420A noise figure meter. Communications system diagnostics are discussed, features and applications are listed and a general description is provided. Specifications, a photograph and a block diagram also are included. **Sanders Microwave Div., Manchester, NH (603) 645-6000.**

Circle No. 330

Radar Performance Calculation Software Brochure

This 28-page brochure contains information on the Radar Workstation, which provides a set of tools to conduct detailed radar performance calculations using a microcomputer. Applications, features, a product overview, detailed information on worksheets and other technical data are included. **Technology Service Corp., Silver Spring, MD (301) 565-2970.**

Circle No. 331

Microwave Absorber Data Sheet

This data sheet describes type HC honeycomb microwave absorbers. Four standard types are described: flat, flat with taper, pyramidal and wedge. Brief descriptions and a photograph are provided. **Advanced Absorber Products Inc., Amesbury, MA (617) 388-1963.**

Circle No. 288

Multiplexer, Demultiplexer, E/O and O/E Converter Brochure

This 8-page brochure describes units to be used with the ME522A error rate measuring equipment. The MH676A multiplexer and the MH677A demultiplexer for widening the measuring frequency range to 700 MHz to 1.4 GHz are described; also covered are the MH945A E/O converter and the MH946A O/E converter for optical PCM signal measurement. Features are listed, illustrated and discussed and specifications are provided. **Anritsu America, Oakland, NJ, Melissa Heck (201) 337-1111.**

Circle No. 289

Connector Catalog

This 36-page catalog covers a line of JCM subminiature coaxial connectors, commercial grade equivalents of military SMA, SMB and SMC connectors. Included are screw mating JCM-A and JCM-C connectors, snap-fit mating JCM-B connectors and JCM cable assemblies. Mechanical, electrical, material and performance specifications are given and outline drawings are provided for each product. Connector assembly instructions, mounting hole dimensional layouts, cable specifications and design formulas for determining connector specifications are included. **E.F. Johnson Co., Components Div., Waseca, MN (800) 247-8256 or, in MN, (507) 835-6307.**

Circle No. 293

III-V Epitaxial Service Capabilities Brochure

This 8-page brochure describes capabilities to supply custom and semi-custom GaAs and GaAlAs epitaxial crystal growth services. Production processes, product characterization, quality assurance, customer support and documentation programs are discussed. **Epitronics Corp., Phoenix, AZ (602) 581-3663.**

Circle No. 295

Semi-Rigid Cable Brochure

This 14-page brochure provides information on semi-rigid cable. Electrical and mechanical data are given for EasyBend .141"D, 086"D, .250"D, .047"D and .034"D cable, designed to meet MIL-C-17 requirements. Conductor and conductor finish codes, SWR data , mil-specs, manufacturing and attenuation also are covered. **Haverhill Cable and Manufacturing Corp., Haverhill, MA (617) 372-6386.**

Circle No. 298

Power Meter Technical Note

This 10-page technical note, 64-4, outlines major considerations used in choosing an RF/microwave power meter and sensors. Understanding the signal, sources of measuring uncertainty, power sensor alternatives and application considerations are discussed. Photographs and diagrams are provided, along with a selection table for HP power sensors. **Hewlett-Packard Co., Palo Alto, CA.**

Circle No. 299

RF Product Brochure

This 16-page brochure describes a line of RF circuits, formerly called the company's frequency control device line. Included are quartz crystals, oscillators, crystal filters and RF modules. Performance characteristics, typical specifications, photographs and tips on which designs are best for specific applications are included. **Hughes Aircraft Co., Microelectronic Circuits Div., Newport Beach, CA (714) 759-2952.**

Circle No. 302

DTO Capabilities Brochure

This 4-page brochure describes capabilities for mass production of custom digitally tuned oscillators to military specifications. Production is discussed, and typical DTO specifications and block diagrams of typical systems are provided. **M/A-Com Omni Spectra West, Tempe, AZ (602) 345-8188.**

Circle No. 305

Analog Power Meter Brochure

This brochure describes the model 6950 analog RF power meter, which can operate from 30 kHz to 26.5 GHz with a series of power sensors. Features and specifications are listed and a full description is provided. A series of power sensors also is described. **Marconi Instruments, Allendale, NJ, Paula Mecke (201) 934-9050.**

Circle No. 306

GaAs Capabilities Brochure

This 16-page brochure describes design, engineering and production capabilities for a range of GaAs devices, integrated circuits and subsystems. Discussed are design; processing and production; testing and ATE; packaging and assembly; and reliability. Photographs are included. **Microwave Semiconductor Corp., Somerset, NJ (201) 469-3311.**

Circle No. 307

Microwave Component Capabilities Brochure

This 4-page brochure provides information on capabilities to supply high performance standard and custom microwave components that use thin-film circuitry or discrete technology. Services from design to test are outlined. A table lists specifications for available low noise, wideband and special purpose amplifiers. Outline drawings are provided. **Microwave Solutions Inc., National City, CA (619) 474-8392.**

Circle No. 308

Bellows and Electroform Design Manual

This 12-page manual describes bellows and electroform design. Rigorous formulas and a table of convolution properties are included for bellows, while design considerations and typical parts are covered for electroforms. A listing of stocked bellows and their properties is included. Design parameters for electrodeposited nickel bellows with ODs from 0.63" to 1.250" are given, along with application suggestions. Chemical and physical properties for nickel are included. **Servometer Corp., Cedar Grove, NJ (201) 785-4630.**

Circle No. 313

Crystal Oscillator Brochure

This brochure describes crystal oscillators covering the 1 Hz to 1 GHz frequency range with stabilities from ±.01% to 1×10^{-10}. Clock oscillators (TTL, CMOS, HCMOS and ECL), low phase noise oven control crystal oscillators, TCXOs, VCXOs, VCOs and frequency standards are included. **Vectron Laboratories Inc., Norwalk, CT (203) 853-4433.**

Circle No. 316

Microwave Switch Catalog

This 20-page catalog provides information on 15 series of coaxial switches. Specifications, schematics and outline drawings are provided and options are listed. **DB Products Inc., Pasadena, CA (818) 449-3790 or (213) 684-2635.**

Circle No. 335

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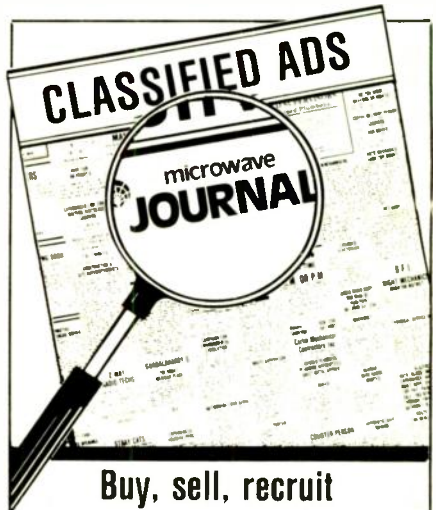

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