microwave JOURNAL

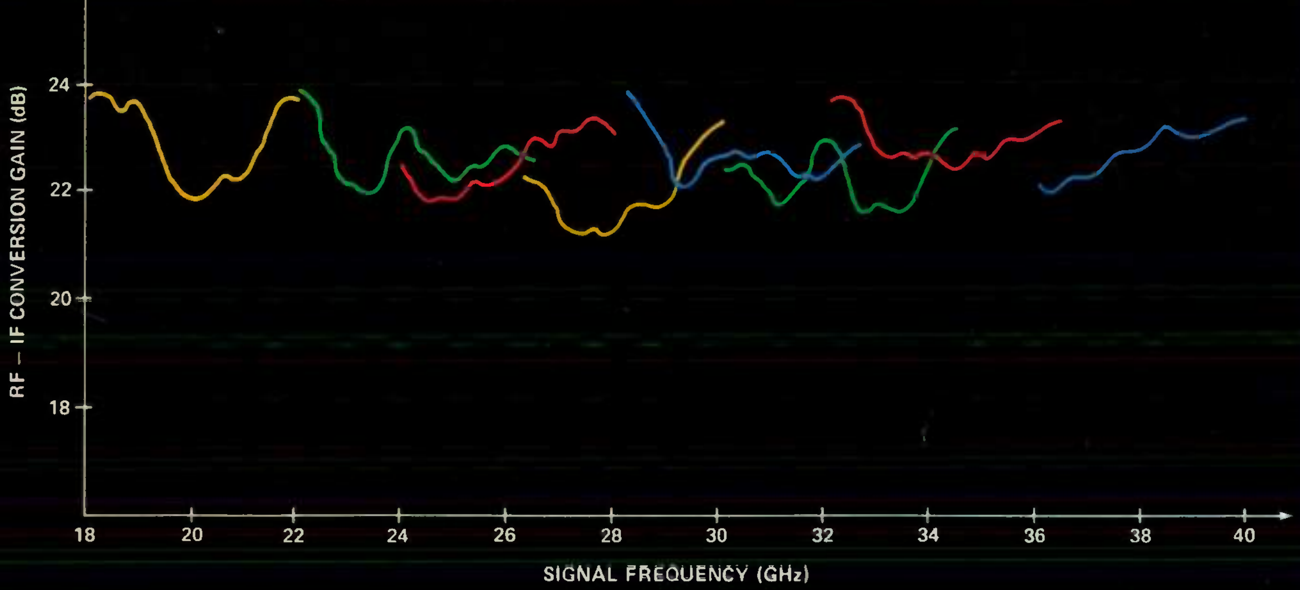
INTERNATIONAL EDITION □ VOL. 24, NO. 2 □ FEBRUARY 1981

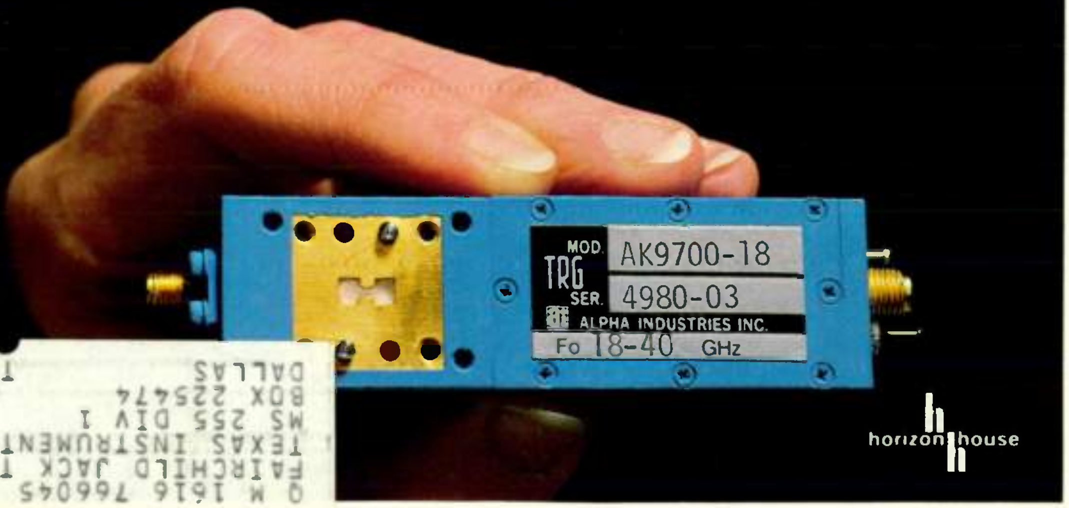
DEFENSE ELECTRONICS

- Phased Array Technology
- Radant - a new Scanning Principle
- An Integrated 18-40 GHz Receiver
- Wide Band ESM Receivers - Part II

and

Waveguide Applicators for Tumor Treatment





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NEW SPECTRUM ANALYZERS

The performance you need... the economy you want.

100 kHz to 40 GHz

Here's what Polarad's new refined 3rd generation, 600 "B" Series Spectrum Analyzers offer you...

- 80 dB on-scale range for 10 dB/div. log scale.
- Improved accuracy for 2 dB/div. and linear scales.
- Greater sensitivity, flatness and noise-free dynamic range.
- Improved dynamic range for harmonics and IM.
- Reduced residual FM and noise sidebands.
- Greater accuracy for Frequency readout and Spans.
- Expanded frequency coverage; 3 MHz to 40 GHz (Models 630B & 640B).
- Cooler operating temperatures increase service life.

NEW ACCESSORIES for the Internal Digital Memory and Programmable Data Processing Interface:

- Model 6488 Adapter for the GPIB, IEEE-488 Bus.
- Model 6700 Digital Cassette Recorder stores and recalls up to 120 displays per cassette.

U.S.A. Prices (Jan. 81)

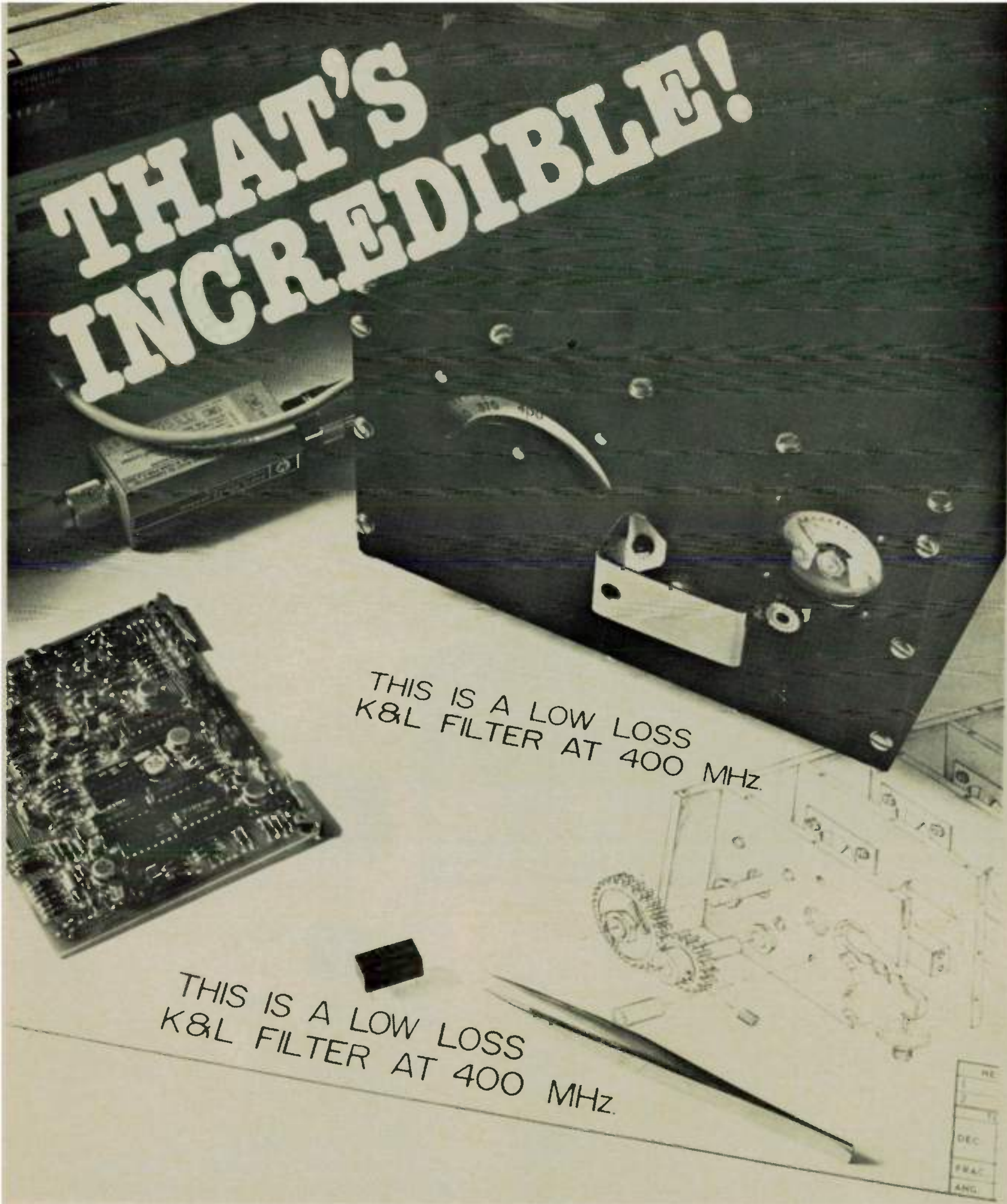
Model	Freq. Range	Without Memory	Including Memory
632B-1	100 kHz to 2 GHz	N/A	$10,175
630B	3 MHz to 40 GHz	$12,460	$14,285
640B	3 MHz to 40 GHz with internal preselector	$16,170	$17,995

Call or write for complete specifications or to request a demonstration.

d polarad polarad **polarad** p

Polarad Electronics. Inc. 5 Delaware Dr. Lake Sucess N.Y. 11042 Tel: 516-328-1100 TWX: 510-223-0414

CIRCLE 1 ON READER SERVICE CARD



It is incredible...both filters have approximately the same low loss (true, there is a difference in bandwidth.) However, K&L's IB10 Series utilizes high Q circuits to achieve a surprisingly low insertion loss for the size. (A size that will fit perfectly on a circuit board.)

Frequencies from 30 MHz to 12 GHz...Contact us with your requirements.

K&L
MICROWAVE INCORPORATED

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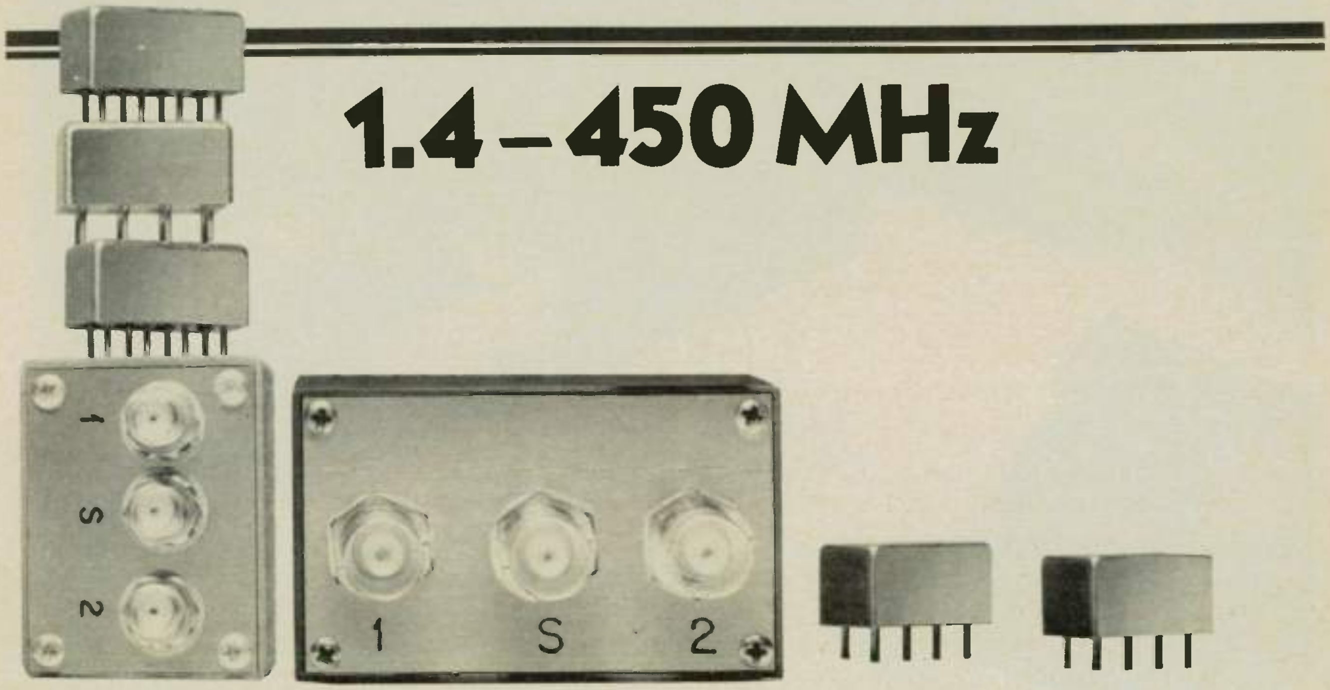
CIRCLE 4 ON READER SERVICE CARD

90° POWER SPLITTERS



1.4 – 450 MHz

Model No.	Freq. Range MHz	Isolation dB		Insertion Loss dB*		Phase Unbalance Degrees	Amplitude Unbalance dB	Price	
		Typ.	Min.	Typ.	Max.	Max.	Max.	$ Each	Qty.
PSCQ-2-1.5	1.4-1.7	29	25	0.4	0.7	3.0	1.2	12.95	(5-49)
PSCQ-2-3.4	3.0-3.8	30	25	0.4	0.7	3.0	1.2	16.95	(5-49)
PSCQ-2-6.4	5.8-7.0	30	25	0.4	0.7	3.0	1.2	12.95	(5-49)
PSCQ-2-7.5	7.0-8.0	35	25	0.4	0.7	3.0	1.2	12.95	(5-49)
PSCQ-2-10.5	9.0-11.0	25	20	0.4	0.7	3.0	1.2	12.95	(5-49)
PSCQ-2-13	12-14	29	25	0.4	0.7	3.0	1.2	12.95	(5-49)
PSCQ-2-14	12-16	30	25	0.3	0.6	3.0	1.8	16.95	(5-49)
PSCQ-2-21.4	20-23	30	25	0.4	0.7	3.0	1.2	12.95	(5-49)
PSCQ-2-50	25-50	30	20	0.3	0.7	3.0	1.5	19.95	(5-49)
PSCQ-2-70	40-70	25	20	0.3	0.7	3.0	1.2	19.95	(5-49)
PSCQ-2-90	55-90	30	20	0.3	0.7	3.0	1.2	19.95	(5-49)
PSCQ-2-120	80-120	25	18	0.3	0.7	3.0	1.5	19.95	(5-49)
PSCQ-2-180	120-180	23	15	0.3	0.7	4.0	1.2	19.95	(5-49)
PSCQ-2-250	150-250	23	18	0.4	0.8	4.0	1.5	19.95	(5-49)
PSCQ-2-400	250-400	22	16	0.4	0.9	4.0	1.5	19.95	(5-49)
PSCQ-2-450	350-450	22	16	0.4	0.9	4.0	1.5	19.95	(5-49)
ZSCQ-2-50	25-50	30	20	0.3	0.7	3.0	1.5	39.95	(4-24)
ZSCQ-2-90	55-90	30	20	0.3	0.7	3.0	1.2	39.95	(4-24)
ZSCQ-2-180	120-180	23	15	0.3	0.7	4.0	1.2	39.95	(4-24)
ZMSCQ-2-50	25-50	30	20	0.3	0.7	3.0	1.5	49.95	(4-24)
ZMSCQ-2-90	55-90	30	20	0.3	0.7	3.0	1.2	49.95	(4-24)
ZMSCQ-2-180	120-180	23	15	0.3	0.7	4.0	1.2	49.95	(4-24)

*Average of coupled outputs less 3 dB   Impedance 50 ohms all models

FROM **$12.95** 5-49 qty.

- **Over 20 models available**
- **Compact PC. . .**
  0.4" × 0.8" × 0.4" H
- **High isolation. . .**
  better than 0.3 dB
- **Low insertion loss. . .**
  less than 0.3 dB

2-way, 90°, pin or connector models. . . Mini-Circuits offers a wide variety of Power Spitters/Combiners to chose from- with immediate delivery. But there are always "special" needs for "special applications". . . We can supply them at your request. . . with rapid turnaround time. . . and at standard catalog prices!

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World's largest manufacturer of Double-Balanced Mixers

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49 Rev Orig.

POWER DIVIDERS

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Isolation over 30 dB midband, 25 dB at bandedges • Octave bandwidths
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- High performance microstrip construction
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$39.95 (1-9)

Now you can specify and purchase state-of-the-art power dividers at 1/3 to 1/2 the price of competitive units, with immediate off-the-shelf delivery. . . from Mini-Circuits, of course.

This breakthrough in price/performance is a natural extension of our extensive experience in high-volume manufacturing, exacting quality control and thorough testing. This expertise assures you highly reliable power dividers with guaranteed repeatability of performance at lowest cost.

So, if you are among the thousands of companies now using Mini-Circuits' signal-processing units in your systems designs, add power dividers to the list of price/performance industry standards available from Mini-Circuits.

Model	Frequency Range, GHz	Insertion Loss, dB Typ.	Insertion Loss, dB Max	Isolation, dB Typ.	Isolation, dB Min	Amplitude Unbalance, dB	VSWR (All Ports) Typ.	Power Rating-W Divider	Power Rating-W Combiner	Price	Qty.
ZAPD-1	0.5-1.0	0.2	0.4	25	19	±0.1	1.20	10 W	10 mW	$39.95	1-9
ZAPD-2	1.0-2.0	0.2	0.4	25	19	±0.1	1.20	10 W	10 mW	$39.95	1-9
ZAPD-4	2.0-4.2	0.2	0.5	25	19	±0.2	1.20	10 W	10 mW	$39.95	1-9

Dimensions 2" × 2" × 0.75"   **Connectors Available:** BNC, TNC, available at no additional charge
$5.00 additional for SMA and Type N

World's largest manufacturer of Double-Balanced Mixers

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34 REV C

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CIRCLE 6 ON READER SERVICE CARD

microwave JOURNAL

contents

VOLUME 24, NUMBER 2
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FEBRUARY 1981

BUSINESS/SPECIAL REPORTS

TECHNICAL/APPLICATIONS SECTION

DEPARTMENTS

ON THE COVER: Alpha Industries' 18-40 GHz mixer-preamplifier with its integral 2-6 GHz GaAs FET IF preamp is a significant step in the quickening search for an affordable, rugged millimeter system for tactical applications. See the Technical Feature beginning on p. 37.

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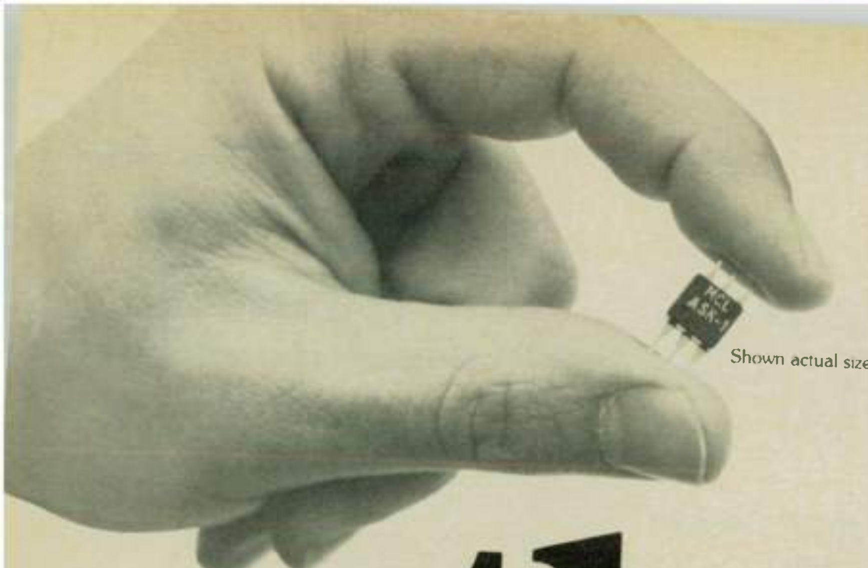
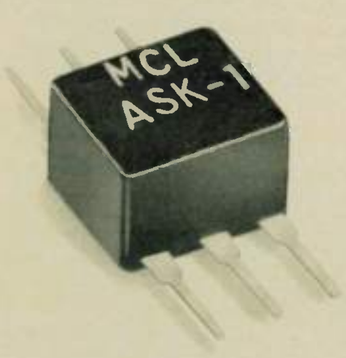


Shown actual size

the shrinker

the worlds smallest and lowest priced flatpack mixer **shrinks size and cost.**

The ASK-1 from Mini-Circuits $5.95 (10-49)



Mini-circuits
Model ASK-1 Plastic Case

Until now, the smallest mixer flatpack available was 0.510 by 0.385 inches or 0.196 sq. inches.

Now, Mini-Circuits introduces the ultra-compact ASK series, measuring only 0.300 by 0.270 inches or 0.081 sq. inches, more than doubling packaging density on a PC board layout.

Utilizing high production techniques developed by Mini-Circuits, the world's largest manufacturer of double-balanced mixers, the ASK-1 is offered at the surprisingly low price of only $5.95 (in 10 quantity).

Production quantities are available now for immediate delivery. And, of course, each unit is manufactured under the high quality standards of Mini-Circuits and is covered by a one-year guarantee.

ASK-1 SPECIFICATIONS

FREQUENCY RANGE
RF, LO 1-600 MHz,
IF DC-600 MHz

CONVERSION LOSS
One Octave from Bandedge 8.5 dB Max
Mid Range 7.0 dB Max

ISOLATION
L-R 45 dB Typ., L-I 30 dB Typ

ABSOLUTE MAXIMUM RATINGS
Total Input Power 50 mW
Total Input Current, peak 20 mA
Operating Temperature −55°C, +100°C
Storage Temperature −55° to +100°C
Pin Temperature (10 sec) +260°C

WEIGHT .35 grams (.01 ounces) *CASE*: Plastic

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CIRCLE 7 ON READER SERVICE CARD

Coming Events

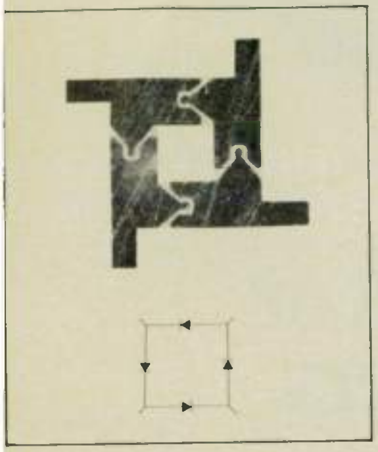
1981 IEEE/MTT-S INT'L MICROWAVE SYMPOSIUM JUNE 15-17, 1981
Sponsor: IEEE MTT-S (held jointly with IEEE AP-S on June 17-19, 1981). Place: Bonaventure Hotel, Los Angeles, CA. Theme: "Around the World with Microwaves," includes such topics as CAD and measurement techniques, microwave and mm-wave solid-state devices, IC's, etc. Contact: Al Clavin, Hughes Aircraft Company, Bldg. 268/A-55, Canoga Park, CA 91304. Tel: (213) 702-1778.

27TH ANNUAL TRI-SERVICE RADAR SYMPOSIUM JUNE 23-25, 1981
Sponsors: US Navy and DoD agencies. Place: The Naval Postgraduate School Monterey, CA. Theme: Development and operation of military radar systems. DoD security clearance through SECRET and a need-to-know endorsement will be required for attendance at the Symposium. Contact: Radar Symposium Coordinator, Environmental Research Institute of Michigan, P.O. Box 8616, Ann Arbor, MI 48107. Tel: (313) 994-1200, ext. 324.

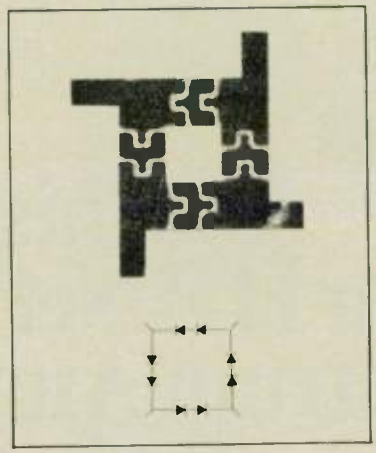
20TH GENERAL ASSEMBLY OF THE INT'L UNION OF RADIO SCIENCE AUG. 10-19, 1981
Sponsors: Int'l Union of Radio Science (URSI) US Nat'l Committee (USNC). Place: Hyatt Regency Hotel, Washington, DC. Symposia Sessions: Millimeter and Submillimeter Waves, Remote Sensing, Mathematical Models of Radio Propagation, Interaction of Electromagnetic Waves with Biological Systems, plus meetings of General Assembly Commissions. Contact: R. Y. Dow, Exec. Sec. of the 1981 URSI General Assembly Organizing Comm., National Academy of Sciences, 2101 Constitution Ave., N.W., Washington, DC 20418. Tel: (202) 389-6478.

6TH INT'L CONF. ON INFRARED & MM WAVES DEC. 7-12, 1981
Call for Papers. Sponsors: IEEE MTT-S and IEEE Quantum Electronics and Applications Society. Place: Carillon Hotel, Miami Beach, FL. Topics: Millimeter sources, devices or systems, mm and sub-mm propagation, atmospheric physics and propagation, plasma interactions and diagnostics, guided propagation and devices, etc. Submit 35- or 40-word abstract by June 30, 1981 to: Mr. K. J. Button, Program Chairman, MIT Francis Bitter Nat'l Magnet Lab, 170 Albany St., Cambridge, MA 02139. Tel: (617) 253-5561.

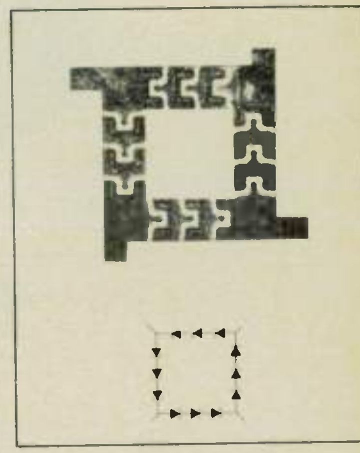
Monolithic Multiple Schottkys Technical Hot Line: (617) 935-5150



4 Diode Ring

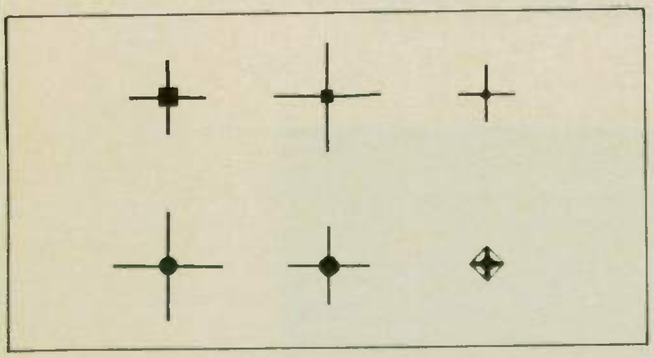


8 Diode Ring



12 Diode Ring

Package Styles



Partial Listing of Available Devices

Type Number	Frequency Range (Typical) GHz	Drive Level (Typical) dBm
DDC 4383	8-12	0
DMF 5829	8-12	+7
DMF 4574	12-18	+7
DME 6549	2-4	+10
DME 4541	12-18	+10
DMJ 4007	2-4	+17
DMJ 4708	2-4	+23
DMJ 4766	2-4	+25

Get the Alpha Advantage: Free samples and quick reaction time to your design application inquiries when you call the technical hot line.

Use of multiple junctions with proper selection of barrier metal enables us to offer you the largest selection of ring diodes for use over a wide range of drive levels.

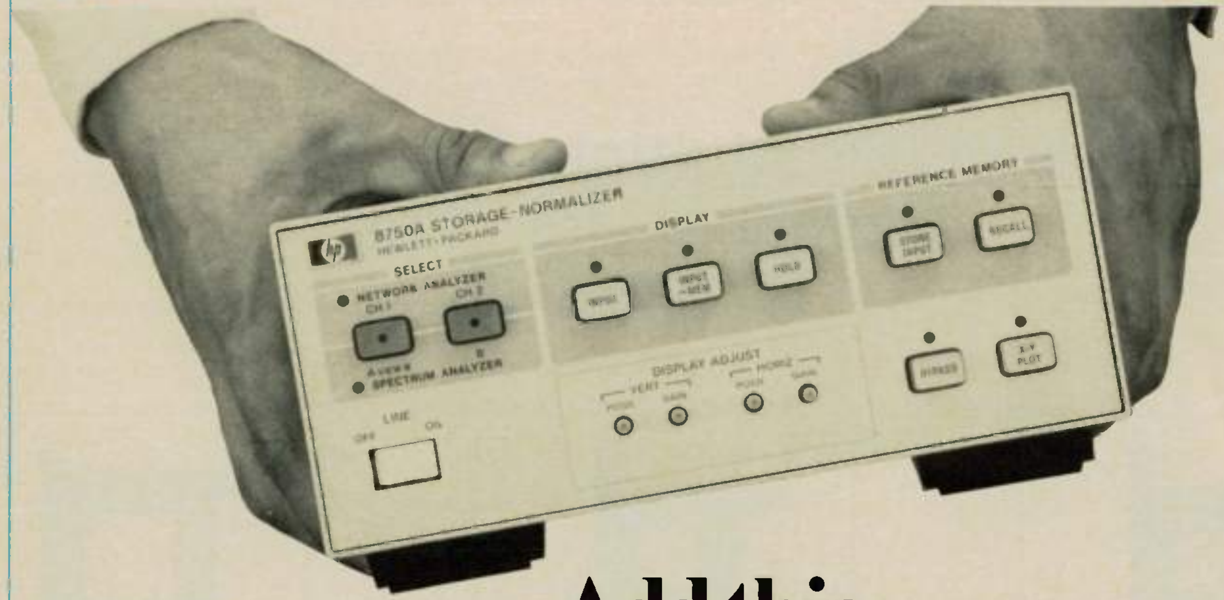
These devices are available in a variety of resin encapsulated or hermetic packages for operating frequencies to 18 GHz.

Call: Alpha Industries, Inc., Semiconductor Division, 20 Sylvan Road, Woburn, Massachusetts 01801 (617)935-5150, TWX: 710-393-1236, Telex: 949436.

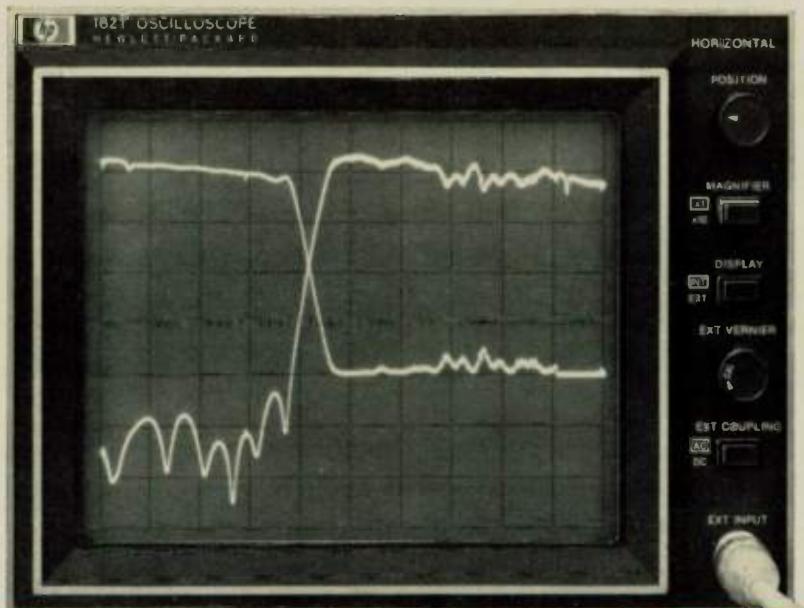
ai Alpha

The Alpha Advantage

CIRCLE 9 ON READER SERVICE CARD



Add this to your HP Network or Spectrum Analyzers and immediately enhance their capability.

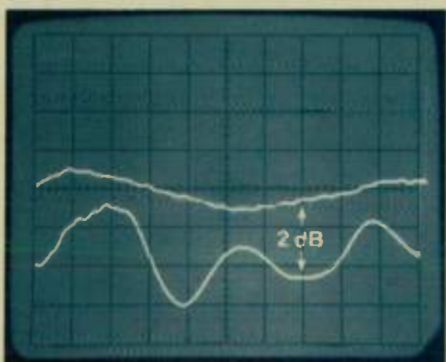


The HP8750A Storage-Normalizer: It brings additional accuracy and simplicity to swept frequency measurements.

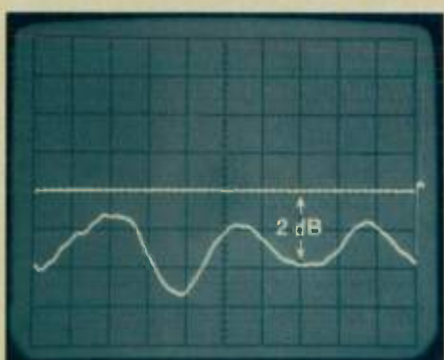
Here's an extremely useful and versatile accessory for most HP Network and Spectrum Analyzers. The 8750A Storage-Normalizer employs memory techniques to "normalize" – that is, remove system response from measured data. And its digital storage, constantly updated, provides a continuous flicker-free display regardless of sweep speed.

Here are some examples of the improvements it can bring to your swept frequency measurements:

High Accuracy Measurements.



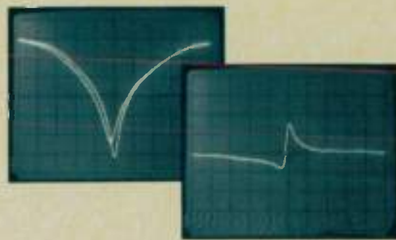
Before Normalization



Normalized

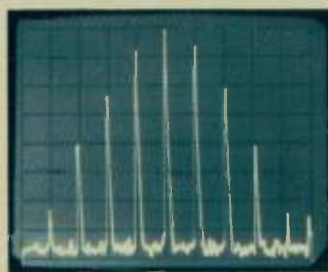
Frequency response or tracking errors in transmission or reflection measurements are eliminated with normalization. You can calibrate the test system's response and store it, then subtract it from the measured data. The resultant difference represents the corrected measurement that's displayed directly in dB.

Comparison Measurements.



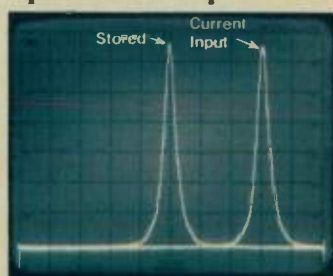
No longer is it necessary to visually scale deviations between two traces. With the HP 8750A, you can now display the *difference* between the two. Deviation between test devices is displayed directly in dB with a single trace.

Slow Sweep Measurements.



Use it for high resolution measurements when slow scan times are needed and get a bright, flicker-free display. Measurement data are displayed from memory with continuous refresh, independent of scan time and scope adjustments.

Spectral Comparisons.



Using the 8750A in spectrum analysis applications, a signal spectrum can be frozen on the CRT and then compared directly with the current input signal.

Because the HP 8750A can "freeze" the display, photography is simplified and hard copies such as X-Y recordings can automatically be plotted, even while new measurements are being made.

Domestic U.S. price of the Storage-Normalizer is $1750.

Call your HP field engineer for more information on how the 8750A enhances measurements made with HP 8407, 8410, 8505, 8754 and 8755 Network Analyzers, HP 8557, 8558, 8559 and 8565 Spectrum Analyzers, plus other instruments. Or write 1507 Page Mill Road, Palo Alto, CA 94304.

45905C

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CIRCLE 10 ON READER SERVICE CARD

PHASED ARRAY TECHNOLOGY WORKSHOP

Joe White, our Consulting Editor, reports on the Phased Array Workshop hosted by Naval Research Laboratory in September 1980. The meeting attracted representatives from all the Services and from industry and it reviewed the achievements to date and examined the directions in which future work should be concentrated. Every major phased array system built since 1950 was discussed. Active programs at each Service laboratory and facility were described by representatives from those organizations. This Special Report provides as complete a summary of US phased array efforts as is available on an unclassified basis.

Sum Up

MULTIMODE TWTA DESIGN CONSIDERATIONS

Following a review of basic traveling-wave tubes operation, the article examines the advantages of multi-stage depressed collector designs. The practical limitations of that approach and the common problems encountered in its application are discussed. The benefits available from computer design simulations are also covered. The effectiveness of the multi-stage depressed collector design for multimode (electronic countermeasures) applications as well as for high efficiency satellite communications tubes is shown.

AN INTEGRATED RECEIVER FOR 18 - 40 GHz

An integrated balanced mixer and GaAs FET IF amplifier with an input frequency range of 18-40 GHz and an output range of 2-6 GHz is described. A fused quartz suspended stripline medium is used in the balanced mixer to provide high tolerance to the extreme environmental conditions which can be anticipated in the tactical applications for which the design was created. The design approach is discussed and complete performance data is provided. In addition, data for mixer performance, by itself, are also included.

A NEW ELECTRONIC SCANNING METHOD

The RADANT (a contraction of the words "radome" and "antenna") principle of electronic scanning involves the modification of the refractive index of a lens made of an artificial dielectric. The article describes such an artificial dielectric consisting of grids of wires containing numbers of diodes connected together. Changing the bias states of the diodes changes the index of refraction of the dielectric and the beam position. The theory of operation is covered in detail and the advantages of the new approach over the conventional ones are discussed.

WIDEBAND ESM RECEIVING SYSTEMS

In the second part of this two-part article, the channelized receiver which combines the high intercept probability of the wide open receiver with the sensitivity and resolution of the superheterodyne receiver is discussed. In addition, the importance of the relationships between the receiving subsystem, the processing sub-system and the technique used for determining direction of arrival are covered. Finally, a tabular comparison of the advantages and disadvantages of the various receiving systems discussed in both parts of the article is presented.

WAVE GUIDE APPLICATORS FOR TUMOR TREATMENT

Radiation at 27 MHz can be used for hyperthermia treatment of deep-seated tumors which are not accessible with either 915 or 2450 MHz radiation. The paper describes the design of ridged waveguide applicators for use at 27 MHz which can handle the power levels involved, minimize the level of power to healthy tissue, are consistent with the physical comfort of a patient undergoing treatment and minimize radiation into free space. Calculations of temperature profiles generated with plane waves at 27 MHz in living tissue models are included and the use of the applicators in animal experiments and clinical trials is described.

Howard Ellowitz

Workshops & Courses

MM-WAVE TECHNOLOGY SEMINAR

Sponsor:	Palisades Institute
Site:	Coply Plaza Hotel, Boston, MA
Date:	March 25-27, 1981
Fee:	$550
Description:	Sessions: introduction, solid state and high-power sources, periodic beam interaction tubes, etc.
Contact:	Mr. Leonard Klein, Dir. Conf. Act., Palisades Inst. 201 Varick St. New York, NY 10014 Tel: (212) 620-3377

ONE-DAY WORKSHOP

Sponsor:	Dallas IEEE Section, MTT-S Chapter and UT, EE Dept., Arlington
Site:	UT, Arlington Student Center
Date:	Sat., March 28, 1981,
Fee:	Students – $10, IEEE members – $20, non-members – $30 (+$5 for onsite registration)
Chairman:	John Wassel, TI
Topics:	I – Automated RF Testing, II – RF Filter Tech., III – Sig. Processing with SWD's, IV – MW Monolithic GaAs Tech.
Contact:	Fern Arteburn, UTA Tel: (817) 273-2671

SHORT COURSE ON RADAR REFLECTIVITY OF LAND AND SEA

Sponsor:	Georgia Institute of Tech., Engrg. Exp. Station
Site:	GIT, Baker Bldg. Aud.
Date:	April 6-10, 1981
Fee:	$400
Description:	Radar clutter discussion.
Contact:	Department of Cont. Ed., GIT, Atlanta, GA 30332 Tel: (404) 894-2400

RADAR TECHNOLOGY SEMINAR

Sponsor:	IEEE/AESS
Site:	Hilton Hotel, Huntsville, AL
Date:	April 20, 1981
Fee:	$125, IEEE member; $140, nonmember
Lecturer:	Dr. Eli Brookner, Consulting Scientist
Subjects:	Radar systems, signal processing, solid-state tubes, etc.
Contact:	Dr. Eli Brookner, Raytheon Co., Wayland, MA 01778 Tel: 358-2721, ext. 2366

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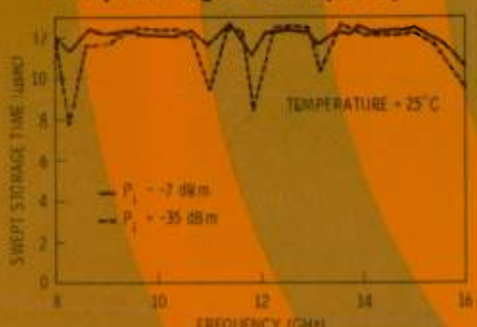

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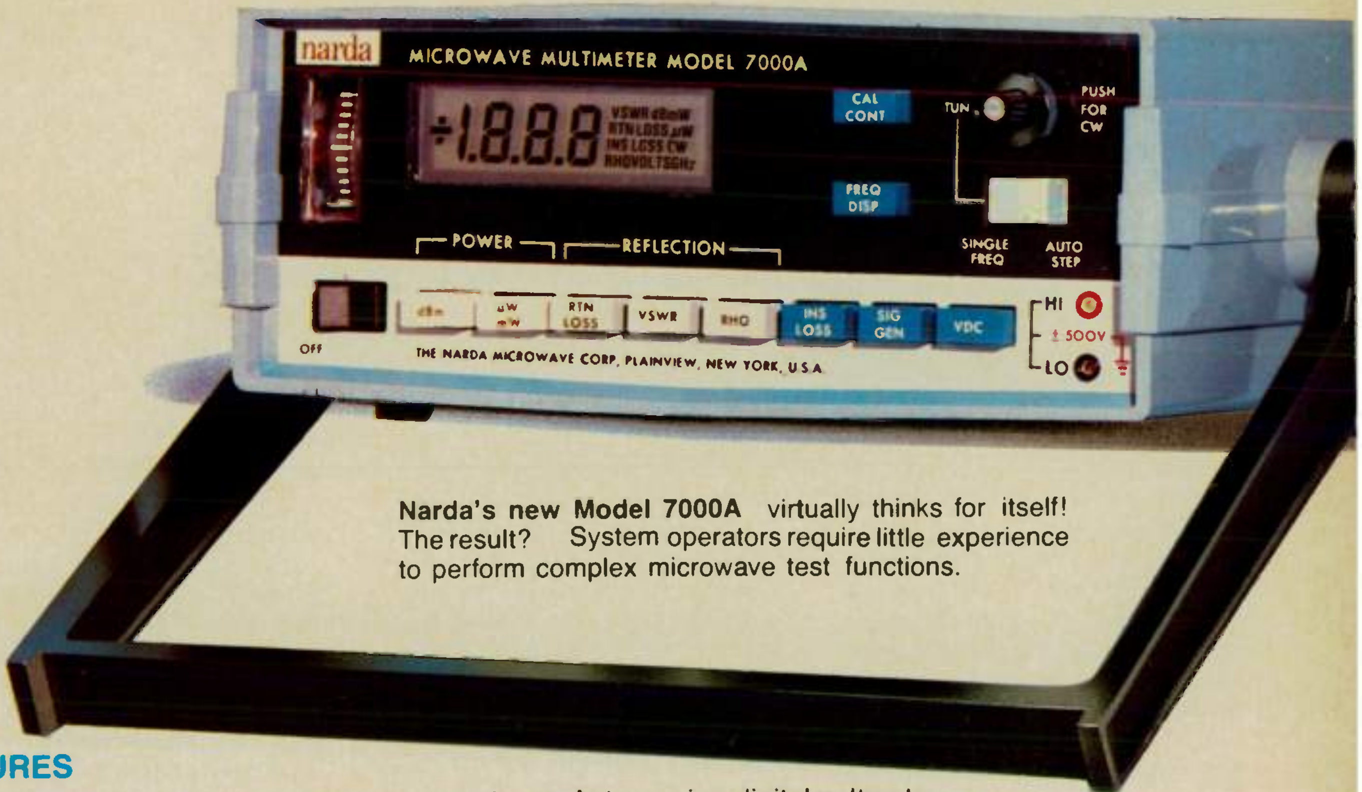

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JOSEPH F. WHITE, Consulting Editor

Phased Array Technology Workshop

September 9-10, 1981 NRL, Washington, DC

This phased array workshop was hosted by Naval Research Laboratory (NRL) September 9 and 10, 1980 to establish the current technology of phased arrays as well as to examine the trends since 1950, about 30 years from the earliest planning of actively steered arrays. Chairman, J. Paul Shelton stated that this was intended as a Government-wide shared workshop and that, over the two days of papers and talks, representatives from all of the Services and from industry would participate in the definition of not only the technological achievements and realities of this field but, as each speaker perceived it, the future advantages and utilizations of radar systems employing electronically steered phased array antennas. Paul Shelton opened the meeting with the questions, "What should we be doing that isn't being done?" and "Is there something we're doing that we shouldn't be doing?"

Merrill Skolnik followed with a review of the history of phased arrays beginning in 1950 with the early frequency scan, Huggins phase shifters, followed in the mid-1960s with the diode phase shifter utilizing space feed, and finally with the most powerful phased array antenna thus far built, the COBRA DANE in 1975. Merrill pointed out some experiences gained during the funding of early phased array systems. For example, the Bendix FPS-85 was solicited in a government RFP that defined the mission of the radar system but left the choice of operating frequency band to the contractor. Merrill pointed out that this was then and is today an effective move. At the contractor's discretion, UHF was selected and since the system was intended primarily for surveillance work this was doubly appropriate. First of all, with surveillance the lower the frequency the better, consistent with antenna size limitations and so forth; and second, the lower the frequency the more easily can the necessary array steering modules be built. "Surveillance is best done at low frequency, tracking at high frequency. If you mix them, a compromise must be accepted," noted Merrill. Presently the Government has been "looking at, not funding arrays," and he suggested that perhaps the Government should consider a concerted effort at this time to identify, based on present and anticipated technology, the applications which phased arrays would best serve. This might, he suggested, require funding five contractors for five years at a million dollars per year just for unstructured research in phased arrays.

Bob Hill (NAVSEA) presented a paper titled "Phased Array Radars – the Commitment, the Experience, the Challenge," in behalf of RAdm, Wayne E. Meyer, who could not attend the conference. This paper pointed out that new Soviet threats will require the best front ends, antennas included. Missile targets will be smaller and faster. AEGIS was started by RCA in 1969. In 1969 the Navy evaluated "piece parts," i.e. phase shifters and feed networks. Today that myopia would be inappropriate. AEGIS is a pioneering effort in phased array manufacturing from the specs, in process tests, computer control support and manufacturing technique. In AEGIS was developed ACE (AEGIS Cost Effectiveness) and though useful, it is limited in that it is only a set of cost-containing methods; totally new techniques will be needed to further reduce array costs to practical levels.

Even if AEGIS ferrite material were free, each antenna face would cost \$2.5 million, but we need \$1 million antennas (4500 elements/array face). AEGIS uses 4 faces/ship, 18,000 elements per ship.

Bob Hill went on to add his own overview observing that

(continued on page 19)

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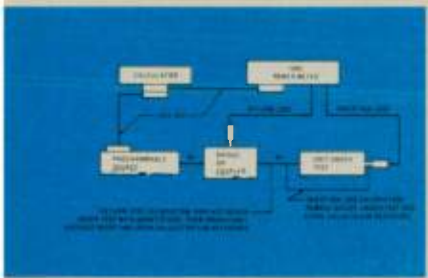

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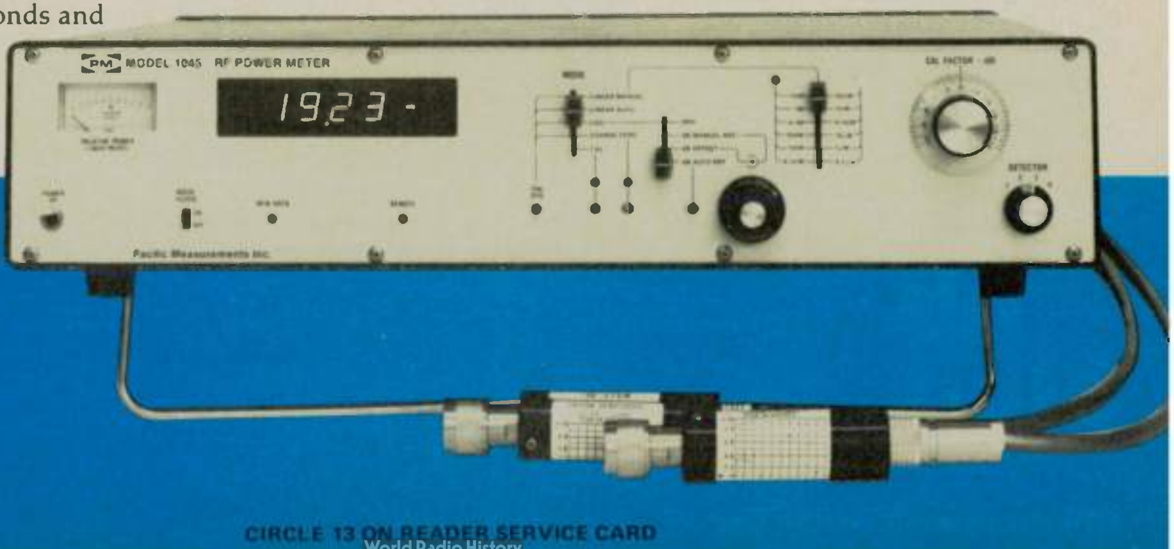

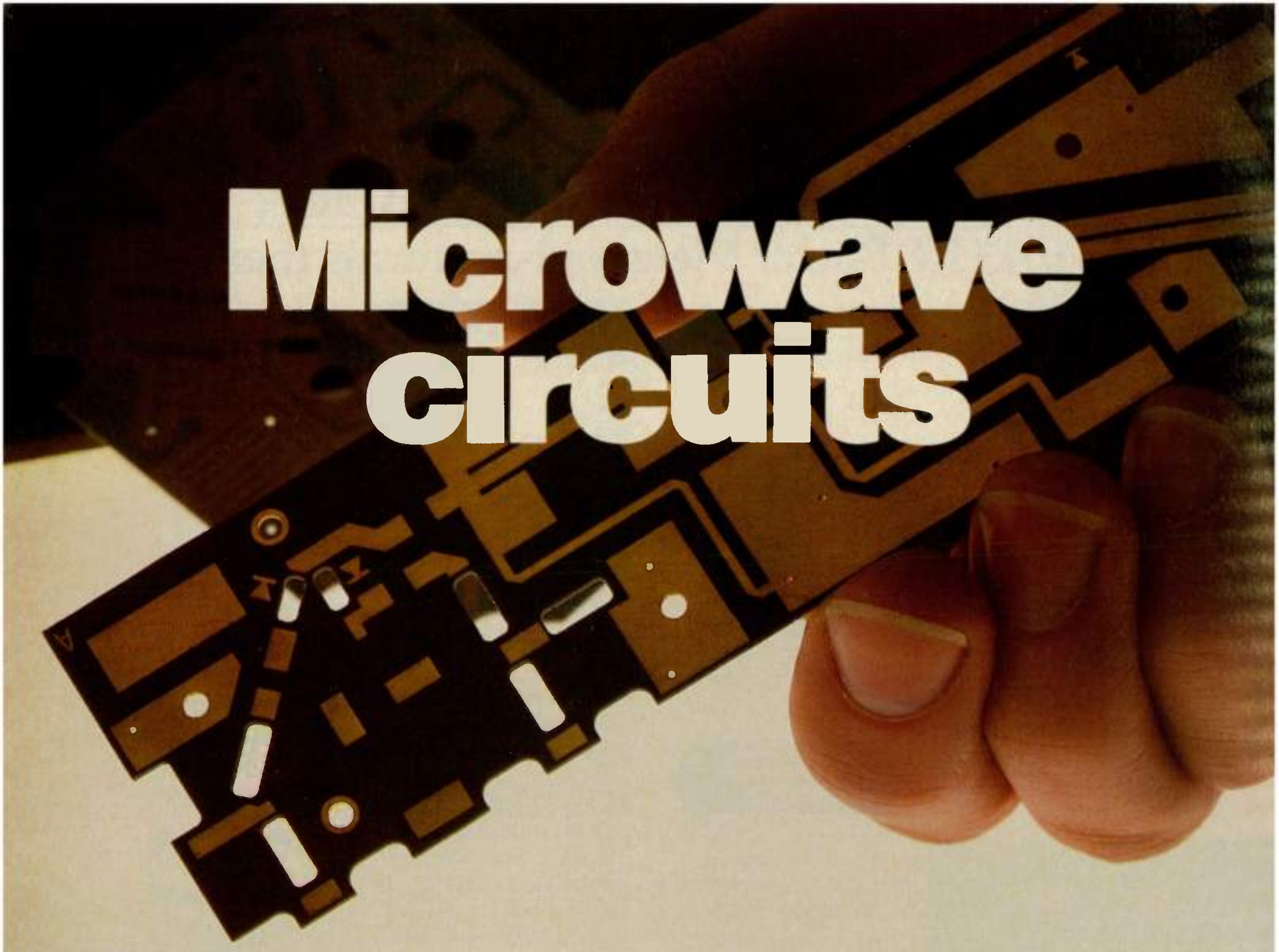

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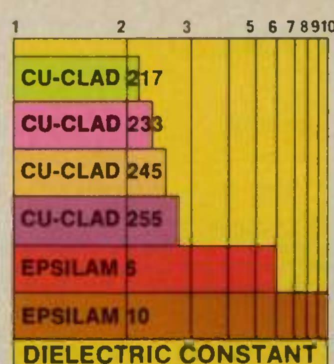

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vhat we need is a fresh approach. n general, Bob feels the antenna hould be viewed as an "amor- hous controllable medium," vhose transmissive phase can be lectrically distorted to steer and o shape one or more beams as hey emerge from or converge on vhat has thus far been referred to s a microwave antenna. Taking a road view is needed, he feels, if ovel approaches, departing from lassical techniques to realize a ontrollable microwave radiating tructure are to emerge. Some ovel configurations include the DOME antenna concept (more bout this later) and the RAD- NT technique (which consists f a lens containing strings of di- des that act to steer microwave eams via a distributed phase hift technique).

Leon Poirier (RADC) in "The hased Array Technology Pro- ram at RADC" described work roceeding at RADC from the iewpoint of radar jamming vul- erability versus cost. To mini- nize the susceptibility of the ra- lar to jamming, broadband oper- tion and steering of sidelobe ulls (adaptive nulling) are em- loyed. Given the performance equirements then, cost is ad- ressed through use of printed ircuits, monolithic techniques, pherical lens arrays such as the DOME lens system and alternat- ng beam control, i.e. use of more han one beam simultaneously.

George Jones (ABMDATC) de- cribed the problem of ballistic nissile defense as it applied to adar requirements. The fact that n the terminal phase (as an CBM would reenter the atmos- here prior to impact) an ABM efensive missile terminal has ess than one minute for reaction, nd during this time may be re- uired to cope with multiple mis- ile targets moving at very high peed, mandates an electronically gile, phased array antenna. The echnical details of George Jones' resentation are classified SEC- RET but certain general observa- ions can be made. For long-range shortly after launch) tracking nd/or surveillance a high power, ow frequency (L-band for exam- le) radar like the COBRA DANE with its one degree beam is appropriate. However the use of interceptor missiles required to defend MX sites mandates higher frequency bands both for increased tracking accuracy and because the antennas must be small enough to fit on relatively small cross section missiles. Even for a mission as critical as ballistic missile defense (BMD), cost is still an important consideration. If the cost for BMD is too high, regardless of whether that cost is engendered in phased array antennas or expensive rockets, then the competing alternative is simply the building of more offensive missiles.

Charles Jedrey (NAVSEA) is a Program Manager of Surveillance Radar Development whose talk was titled "The Future of Shipboard Phased Array Technology." He amplified the observation that high performance at low cost relative to what is currently available is needed for wide spread shipboard use of phased arrays. More was said on this topic throughout the conference. He pointed out that at NAVSEA, low sidelobe arrays with adaptive nulling are an important and inherent advantage of the phased array systems.

Bill Spaulding (MICON, Army Missile Command, Redstone, Alabama) made several observations in "Phased Array Research at MICOM," regarding how phased arrays relate to the ECM threat.

- The ECM threat requires the use of pencil beams and low sidelobes.
- Broad bandwidth is needed for high resolution.
- Much information is lost by phased array combinatorial techniques. We should instead be designing systems which convert the received RF phase and amplitude information at each array element directly into a multibit digital word. Then a computer could process simultaneously all array information, factoring in sidelobe data, creating multiple beams numerically, and so forth. (This was a point acknowledged by many of the later speakers in the Worksho as an important and unique opportunity available throug use of phased arrays.) This technique may require use of the multistatic principle (see below) to allow enough integration time at each module to achieve a satisfactory signal-to-noise ratio.
- The multistatic radar technique (whereby one or more receive only antennas are situated for security purposes some distance from one or more transmitting sites) can provide a way of protecting an expensive phased array, which need only be used for reception and therefore woul not be vulnerable to tracking through its emissions. Moreover, the receive array is less expensive since it need be neither reciprocal nor able to handle high power.
- Using the above techniques, Bill envisions a system operat ing with 200 simultaneous receive beams (and suitable digi tal processing). The multiple beams could stay on their tar gets long enough to provide sufficient integration time th only a single 10 kilowatt CW transmitter would be needed for illumination.
- This analog-to-digital process array might be made possible through VHSIC (Very High Speed Integrated Circuits) presently enjoying large Government sponsorship. Throug this program, Bill postulates a butterfly chip using 40 ns log and having an area of 1 cm^2 Twelve tiers might be used which, operating at a 2.08 MHz sample rate, would provide 75-meter radar resolutio

James Fraser (DARPA) in "DARPA Solid State Module Development Program," described research on a concept by Grumman Aircraft, Inc., for a 70-meter diameter phased array to be located in space. It would contain 100,000 to 1 million modules, depending on the frequenc of operation, L- through X-banc being considered. Several contractors including Texas Instruments, Raytheon Equipment and Research Divisions, and General Electric were awarded contracts for critical portions of the research. In addition, Ball Brothers Research, Inc., Aerospace Division, is conducting system studies. Low cost monolithic microwave techniques, including silicon on sapphire and gallium arsenide, are being examined. This work began in 1979 and the research portion is expected to be completed by 1982. RADC is the prime contractor with tri-service funding. The goal is to achieve all monolithic state-of-the-art performance (with affordability $10 to $100 per transceiver module) in the large module quantities.

Eli Brookner (Raytheon) presented a talk entitled "New Achievements in Array Radars" which illustrated new techniques as employed in such recent phased arrays as the COBRA DANE and the PAVE PAWS radars. The COBRA DANE is special in that its system characteristics are unclassified. Operating in the band 1175 to 1375 MHz with a peak radiated power of 15 megawatts, average power of 1 megawatt, and pulse length up to 2,000 msec, it has a range of 8500 kilometers with a signal-to-noise ratio of 13 dB on a one square meter target. This is achieved using 15,360 radiating elements arranged in groups (subarrays) of 160 elements each. A traveling-wave tube of 160 kW peak power feeds each subarray for a total of 96 tubes. The 95-foot diameter array is thinned, having approximately 45,000 elements, about two thirds of which are terminated in dummy loads. While fairly far advanced for its period (completed in 1975) Eli points to still newer phased arrays which include PAVE PAWS (built by Raytheon), EAR (built by Westinghouse), and TPQ37 and TPQ36 (built by Hughes), LUXOUR (built by Thomson CSF, France) and the Sperry DOME array antenna.

New features include (for the PAVE PAWS radar) the fact that dummy elements need not be matched load terminated. The array is thinned without the need for expensive absorptive array elements. PAVE PAWS operating at UHF is entirely solid state. Eli showed photographs of the radar's interior which, although designed along the same lines as the COBRA DANE, has very little RF plumbing behind the array face due to the all solid-state-module design.

The Sperry DOME antenna (more about this later in a separate talk by Jerry Hanley) built at S-band achieves 360° coverage with a single controllable lens antenna. This is accomplished through a hemispheric passive lens systems and could logically be a candidate for the AEGIS shipboard system, Eli noted. Two factors mitigating the consideration are the fact that it was not developed at the time AEGIS was begun (in 1969) and, although offering the attractive feature of 360° coverage with a single lens, more volume is required to realize the space feed aspect of the array.

Eli also described the new gallium arsenide substrate monolithic work at Raytheon which achieves 2.1 W of power at X-band and 1.4 W over a 20% bandwidth at X-band. More about this later.

Earl Maine (NRL) presented a classified talk entitled "Fixed Array Surveillance Radar (FASR)" in which he described a UHF shipboard surveillance array designed for low sidelobes.

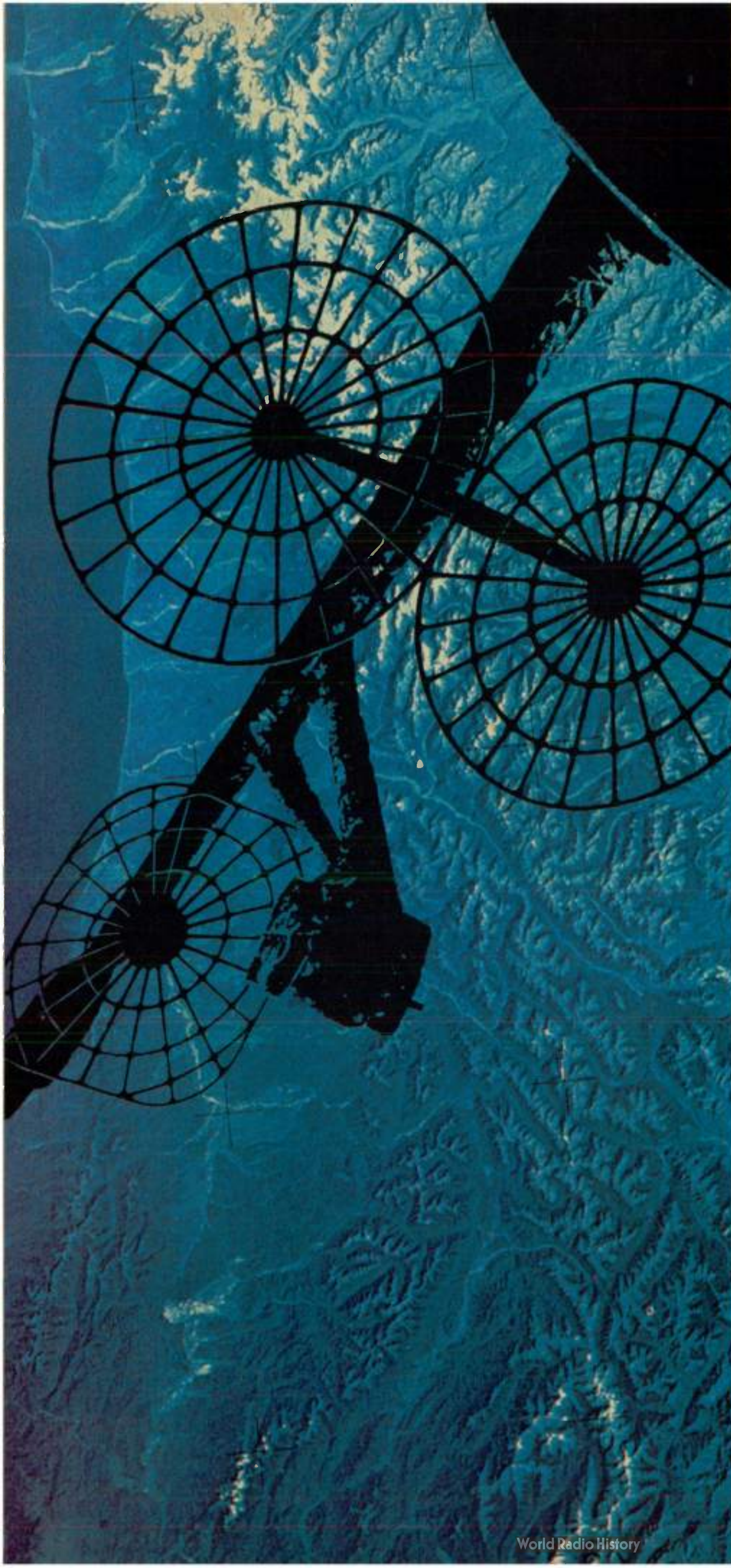
Gerald Hanley (Sperry Gyroscope) in "Hemispheric Coverage Antenna" presented the fruition of a very innovative scheme under development since about 1970, which uses a two dimensional phased array located in a horizontal plane. Above it is a near-hemispheric honeycomb (DOME) structure which contains (in the model reported) about 25,000 cells each providing a fixed phase shift to refract the vertically oriented beam leaving the steerable array beneath the DOME into a more horizontally directed beam. Control of the active phase shifters influences which sector of the DOME is illuminated. The net result is a pencil beam of 1.8° which can be steered through 75° in elevation and 360° in azimuth. The actual operating specifications of the antenna are classified but most interesting is the fact that Jerry feels the array could be manufactured for about $200,000.

Thomas Tice (NOSC) presented "Planar Phased Array for Shipborne Radar," a paper describing an antenna imaging technique which provides the gain associated with multiple radiating elements through use of multiple reflections from a conducting surface located behind a single radiating element. While a theoretical paper, the concept has important practical implications because it offers the gain of many elements while incurring the construction cost of only one element. It is also expected to reduce the effect of mutual coupling which would have occurred had many elements rather than an imaging of one element been employed.

Jean Claude Sureau (Lincoln Lab) described a novel array antenna structure in a talk entitled "Advanced Ground Surveillance Radar." This design offers 360° azimuth scanning and weighs only 180 lbs, easily carried by two men. Practical measurements made on this C-band system, which resembles a pill box in shape having a diameter of about 4 feet and a height of about 2 feet, show that it provides 21 dB of gain, 25 dB sidelobes and a 5° pencil beam in azimuth and elevation. Phase control is accomplished by multithrow switches, each of which has 1.5 dB loss and 35 dB isolation. Typical application is for an air terminal surveillance system.

Willard Patton (RCA) talked about "Low Sidelobe Phased Array Antennas for Tactical Radar." This paper, by one of the principal designers of AEGIS, described some of the modern theoretical work directed toward reduction of antenna cost through array thinning and other practical array architecture expedients. Bill referenced these comments to

(continued on page 22)



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with an active module measuring 5.1 inches by 1.5 inches by 0.325 inches and having a weight of 1.5 ounces. At the present time, it exhibits 7° RMS phase error (over the band and all phase states) and develops 2 watts of output power with 15% overall efficiency at 0.3 to 0.5 duty cycle. Anticipated efficiency is expected to increase to 30% for the module by 1982. This performance is realized using FET's having a maximum channel temperature of 180° C which, applying MIL 217, yields an estimated failure of 29 failures per million hours. This accommodates a 3,000-hour array life.

The present Air Force Electronically Agile Radar (EAR) array uses a tube and ferrite phase shifters and weighs 400 pounds. Using the solid state modules, it is envisioned that a 350-pound array weight is possible. Especially interesting is the projected economic posture for the solid state modules when compared with the present EAR figures. The EAR antenna costs $465,000 in small quantities and the whole radar $600,000, which increases to $717 K over a 10-year period when cost of maintenance is included. This works out to $256 per element. Assuming a 3,000-hour life with no maintenance, solid-state modules would be considered equivalent at a $350 per module Harold indicated, a figure that Gene Gregory, Program Manager at Hughes, feels might be realized. Presently small quantity costs of these active modules are in the vicinity of $2,000 per element.

Peter McVeigh (Eaton, AIL) described a technique for using very few phase shifters in his paper entitled "High Accuracy, Cost-Effective, Electronic Steered Array." This system was designed for the Microwave Landing System and used only six precision 0 - 360° and ten 0 - 90° phase shifters to provide single axis scan. Very accurate pointing (1/100 of a beamwidth) is described for this system, which accommodates the 20° per millisecond scan array needed for MLS. Part of the technique involves the use of multiple subarrays (11 in this experiment) with the phase shifters located in series fashion rather than the customary parallel to drive the subarrays. Ferrite phase shifters were used and the resulting array antenna has a 2°, 3 dB beamwidth, 18 dB of gain, 28 dB average sidelobes and 20 dB peak sidelobes. Using ferrite toroidal phase shifters, a 100 kilowatt peak capability is estimated for the system (although the MLS operates at only a few watts of radiated power). The antenna is nonreciprocal but could be used in a bistatic radar or made reciprocal using reciprocal phase shifters. Peter suggested that a 15,000 element antenna with 1° square beamwidth could be replaced using 640 regular phase shifters and 128 precision line source phase shifters utilizing this technique.

Jim McDade (GE, Utica, NY) described the results of a cost effectiveness study in "Phased Arrays and LSI," through which he concluded that using present technology one way antenna loss would be 1.5 dB for the flat plate mechanically steered antenna, 2.5 dB using ferrite phase shifters, 3.5 dB using diode phase shifters

(continued on page 28)

and only 0.5 dB using active modules. Presently GE is developing a 4 bit, 1.7 dB loss, suspended substrate, four-element diode phase shift module with IC custom made drivers. In support of this system would be an X-band tandem series feed structure.

Roger Sudbury (Lincoln Laboratory) described "Millimeter Wave Phase Arrays for BMD." The missile descriptions of this paper are classified. Lincoln Lab is evaluating a K_a-band GaAs monolithic active module for this application. On the assumption that a 1 micron geometry is dictated for practical production using existing technology, the plan is to limit the RF source frequency to the 10-20 GHz range and then to use a diode multiplier to reach K_a-band.

Present progress at Lincoln has included a GaAs mixer at 30 GHz with 7 dB noise figure and a low noise amplifier with 3.5 dB noise figure, 11 dB gain and 2.5 GHz bandwidth. They have cascaded 3 chips (0.3 x 0.1 inch each) to yield 30 dB overall gain and 3.5 dB noise figure. The mixer amplifier combination gives net gain of 4 dB.

PHASED ARRAY PANEL DISCUSSION

Following the prepared papers, was an open discussion entitled "Future Directions In Phased Array Research," which was moderated by Merrill Skolnik. Panel members included Dave Barton and Eli Brookner (Raytheon) Peter Kahrilas and Ray Tang (Hughes) Leon Schwartzman (Sperry) and Harold Weber (AFAL).

Peter Kahrilas reemphasized the point that radar antennas should become more "mosaic-like" by converting the radar data (amplitude and phase) directly to fully processed information right within the antenna structure through the use of analog to digital techniques. Having made this conversion, radar processing might be performed directly in the antenna, including such things as rejecting unlikely targets, eliminating clutter, and so forth.

Ray Tang gave some remarks emphasizing that low cost array techniques are, in fact, presently available and that it's up to the radar community to use these techniques to promote large scale production of array antenna systems.

Eli Brookner reviewed several of the radar concepts described in the workshop and pointed to future trends which he feels will include adaptive array processing (for jammer rejection), the increasing need for still lower sidelobe levels, cost reduction by the elimination of bonding wires implicit in monolithic circuitry and low weight. Eli pointed out that two monolithic chips used in an output active module themselves only weigh 0.00002 pounds, and the advent of space radars will be made possible by modules which produce 0.5 watts of RF power, weigh 0.02 pounds and cost in the vicinity of $50 each.

Leon Schwartzman emphasized that with the DOME antenna, while 30% more elements are needed in the feed array, since the antenna scans 360° in azimuth it replaces a four-faced array and thereby does so at a projected cost of only 40% of the 4 faced array antenna cost. The DOME array, Leon suggests, could be built today for $200,000 (presumably in a configuration similar to that of the S-band embodiment described earlier by Jerry Hanley). Designers have done a lot to drive down the cost, Leon asserts, through the use of monolithic circuitry, array thinning, the DOME array concept and so forth. It's now time for the system designer to stop listing the "whole store" for radar performance, thereby driving the cost back up to unacceptable levels for array systems.

Harold Weber feels that waveform processing and the use of analog to digital conversion is increasing rapidly in other areas, therefore radars can be expected to reflect this trend. He emphasized that we (in Government) must solidify the state of the art through MM&T (Manufacturing Methods and Techniques) programs. He points out that programs the Air Force funded early on in diode and ferrite phase shifters have paid the dividends that were seen in this workshop with respect to the modern array performance results.

Dave Barton took the devil's advocate role in pointing out that, while phased arrays are wonderful, too often modern system performance specifications are written in such a fashion that only the phased array technology can address them and that, having once done so, the development of the radar is often held back too long for want of the funds to realize the system in phased array format. He pointed out some of the paradoxical situations that can occur in modern radar planning. For example, lots of active jammers in the radar scenario favor the phased array, but not rain or chaff. Doppler processing which can separate the target from rain and chaff clutter requires 10 to 30 hits per scan for 60 dB of rejection, however phased arrays typically only spend 1 hit per scan on the target, mainly, no doubt, because they are able to move quickly on to the next target. He noted facetiously that this will be quite allright as long as we can convince the enemy not to use chaff or make attacks during rain. Also, the solid-state radar usually requires long duty cycle (typically 40%) due to the fact that transistors can't trade off high peak power with low duty cycles to achieve their average power output. This he says is tough if you are trying to fly at 1,000 feet or if the radar is ground based and needs to see targets only 1,000 feet away (hopefully, however, targets this close wouldn't require the full average power and therefore active modules could operate with shorter pulses for such close in targets). An active question and answer period followed the panelists' prepared talks.

There were over 90 registered participants in the Workshop including members from the triservices and industry. It was surprising to see as many elements of novelty in the phased array field, frequently referred to as a mature technology during the late 1970s.

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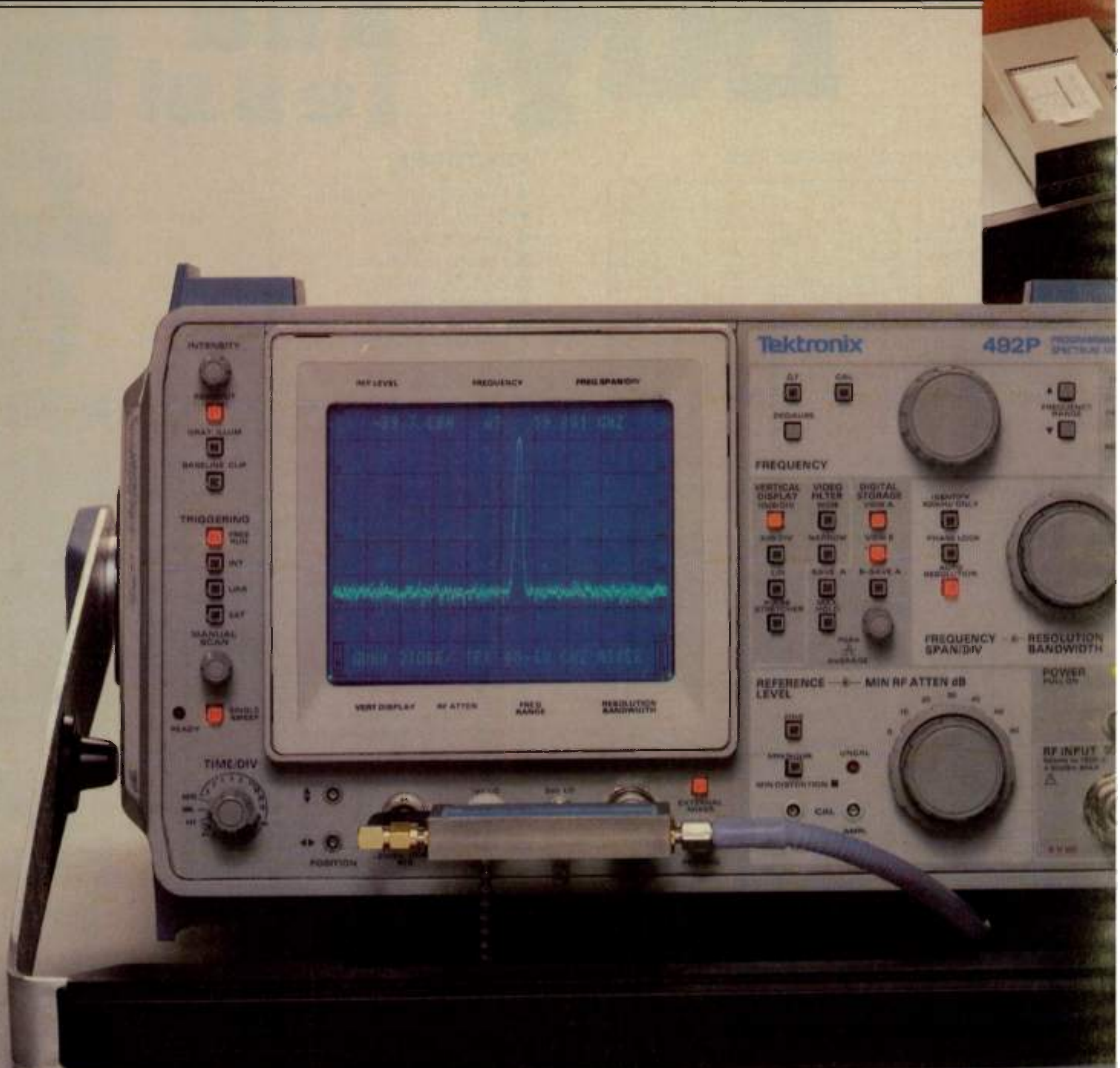

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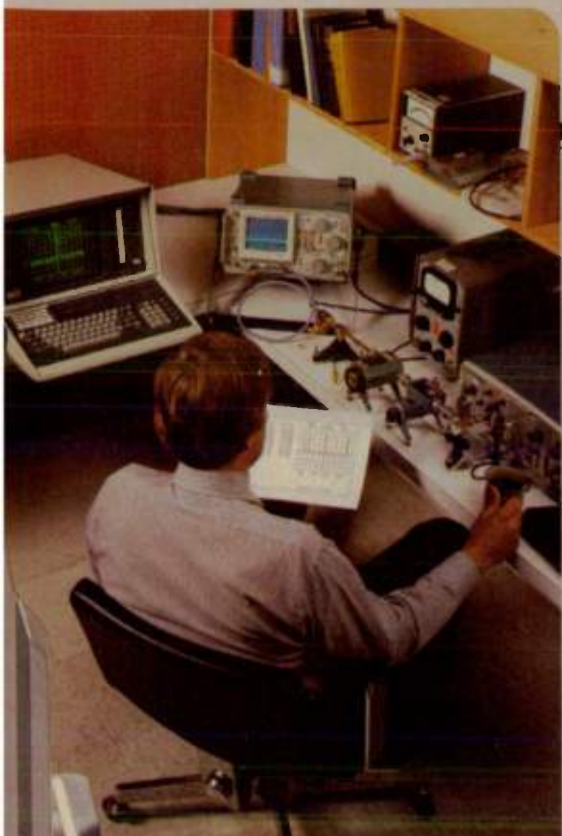

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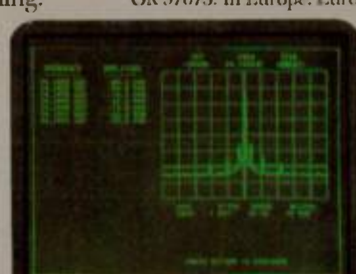

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World Radio History

Around the Circuit

PERSONNEL

Watkin-Johnson's Chairman of the Board, Dr. **Dean A. Watkins**, was elected President of the California Chamber of Commerce for 1981. . .**Len Johnson** has been promoted to V.P., Marketing, of Integra Inc.'s Microwave Div. after serving as that division's Marketing Manager this past year. . .**Dean Lattman** was appointed Program Manager for Military Electronics Systems of Datron Systems, Inc. . .The Board of Directors of AEL Industries, Inc. elected **Jesse H. Riebman** a Vice President. . .Alpha Industries, Inc. appointed **William F. Brady** as Manager of Human Resources, **Jim Lautermilch** as Northwest Regional Sales Manager of the Semiconductor Division and **Carl Roland** as Customer Service Manager. Alpha also announced the following promotions: **Mark E. Madden** to the position of New England Area Sales Manager for its Solid State Div. and **Steve Scannell** to the post of Sales Engineer, Int'l Sales Dept. . . **Dr. James T. Hoffman** was promoted to V.P., Operations for Interstate Electronics Corp. . .At Eaton Corp.'s AIL Div., **Michael J. Philbin** has been promoted from Division Director to V.P., Electronic Warfare Systems. . . California Microwave, Inc. appointed **Fred P. Storke** V.P. – Engineering and Chief Technical Officer. . .Midwest Microwave elected **William D. Painter** V.P./General Mgr. . . .Adams-Russell Co. elected **Michael Alfieris** a Vice President; Mr. Alfieris continues as President of the firm's Digital Processing Div. . .**Michael J. Bruno** was appointed Sales Manager for Microlab/FXR's Microwave Components Div. . .**John L. Blazejewski** was promoted from Regional to National Sales Manager for Shielding Products at Tecknit, Inc.

CONTRACTS

Datron Systems, Inc. received a $2.6M contract from the US Army Test & Evaluation Command to produce a 3-tone, continuous-wave radar system. . .**California Microwave, Inc.** was awarded a $15.5M order from **AT&T Long Lines Dept.** for approximately 6000 of its Model CA42 transmitting amplifier systems. . . **Adams-Russell, Antenna & Microwave Div.** received contracts valued at over $500K to develop user antennas for the US Navy's Navstar GPS Program. . .**E-Systems, Inc.** was granted orders valued at $10.5M for the production of AN/WSC-3 "Whiskey-3" UHF communications terminals and associated equipment for the US Navy. . .**Aydin Corp.** recently received a $3.1M add-on to its present $73.5M subcontract with **Litton Saudi Arabia Ltd.** a Troposcatter Communications Network for an Air Defense System. . .Naval Ocean Systems Center awarded a $320K contract to **Eaton Corp.'s AIL Division** to develop a high frequency receiver to extend RF coverage of radar surveillance and warning receivers into the mm bands. . .**Digital Communications Corp.**, a M/A-COM company, announced the receipt of a contract from **RCA American Communications, Inc.**to supply two redundant TDMA terminals, with options to purchase an additional 5 terminals. . .**Microtel** has received an order valued at $4.8M from the US Air Force for a quantity of their Model 1295 precision measuring receivers.

NEW MARKET ENTRY

RS Microwave Co., Inc. has been formed by Richard V. Snyder. The company will supply high performance microwave filters, couplers and assemblies to the military and communications market. Located at 22 Park Place, Butler, NJ 07712, the company will concentrate on devices requiring computer-aided design.

INDUSTRY NEWS

Teledyne MEC is planning May, 1981 occupancy of an additional 110,000 sq. ft. facility which will bring its total floor space to 210,000 square feet. . .**Southern Pacific Communications Co.** announced it was granted permission by the FCC to construct three satellites (two for launching and one as a ground spare) for use in its US communications network. . .**Alpha Industries, Inc.** has occupied a 10,000-sq. ft. addition to the **Optimax Microelectronics Div.** in Colmar, PA, which will include a new thick film processing and assembly area. . .**M/A-COM, Inc.** reached an agreement in principal with **U.T.G., Inc.** for the purchase of all the outstanding capital stock of **Alanthus Data Communications Corp.**, whereby Alanthus would become a M/A-COM company. . .**COMSAT General Corp.** has acquired by merger **COMPACT Engineering, Inc.** for 35,714 shares of common stock of COMSAT. . .The controlling interest in **International Microwave Corp.** (IMC) stock has been purchased by **NR Technology, Inc.** Under the terms of the sale, IMC will continue to operate as an independent company. . . The Aeronautical Systems Division (ASD) at Wright-Patterson AFB announced that two of its cargo aircraft will conduct "proof of concept" flight tests of a new radar system with a radar transmitter aboard a C-141 and the radar receiver on the C-130 aircraft.

FINANCIAL NEWS

Radiation Systems, Inc. filed a registration statement with the SEC for a proposed public offering of 225K shares. The Company will use the proceeds principally to increase working capital. . .**California Microwave, Inc.** filed a registration statement with the SEC in connection with an offering of 400K shares of its common stock. Proceeds of the offering will be used to repay about $6M of bank debts and for general corporate purposes. . .During the first quarter ended September 30, 1980, **Scientific-Atlanta** reported sales of $55.6M and net earnings of $3.6M or 35¢ per share. This compares with 1979 quarterly net earnings of $2.2M or 24¢ per share on net sales of $39.5M. . .**Microdyne Corp.** reported year-end results for the period ended November 2, 1980 of sales of $23.7M, and net income of $3.2M or $1.16 per share. This compares with 1979 sales of $17.5M, and net income of $2.95M or $1.18 per share. . . **Cubic Corp.**'s year-end results for the period ended September 30, 1980 showed sales of $196.4M, and net earnings of $8.1M or $2.75 per share. This compares with 1979 sales of $171.7M, and net after-tax earnings of $15.1M or $2.71 per share. . .**General Instrument Corp.** reported third quarter results for the period ended November 30, 1980 of revenue of $210.9M, and net income of $17.7M or $1.97 per share. In the third quarter of 1979, revenue was $196.2M, net income was $14.2M or $1.61 per share.

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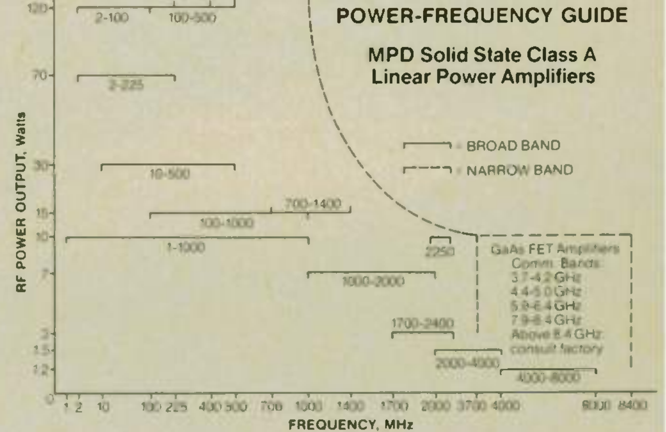

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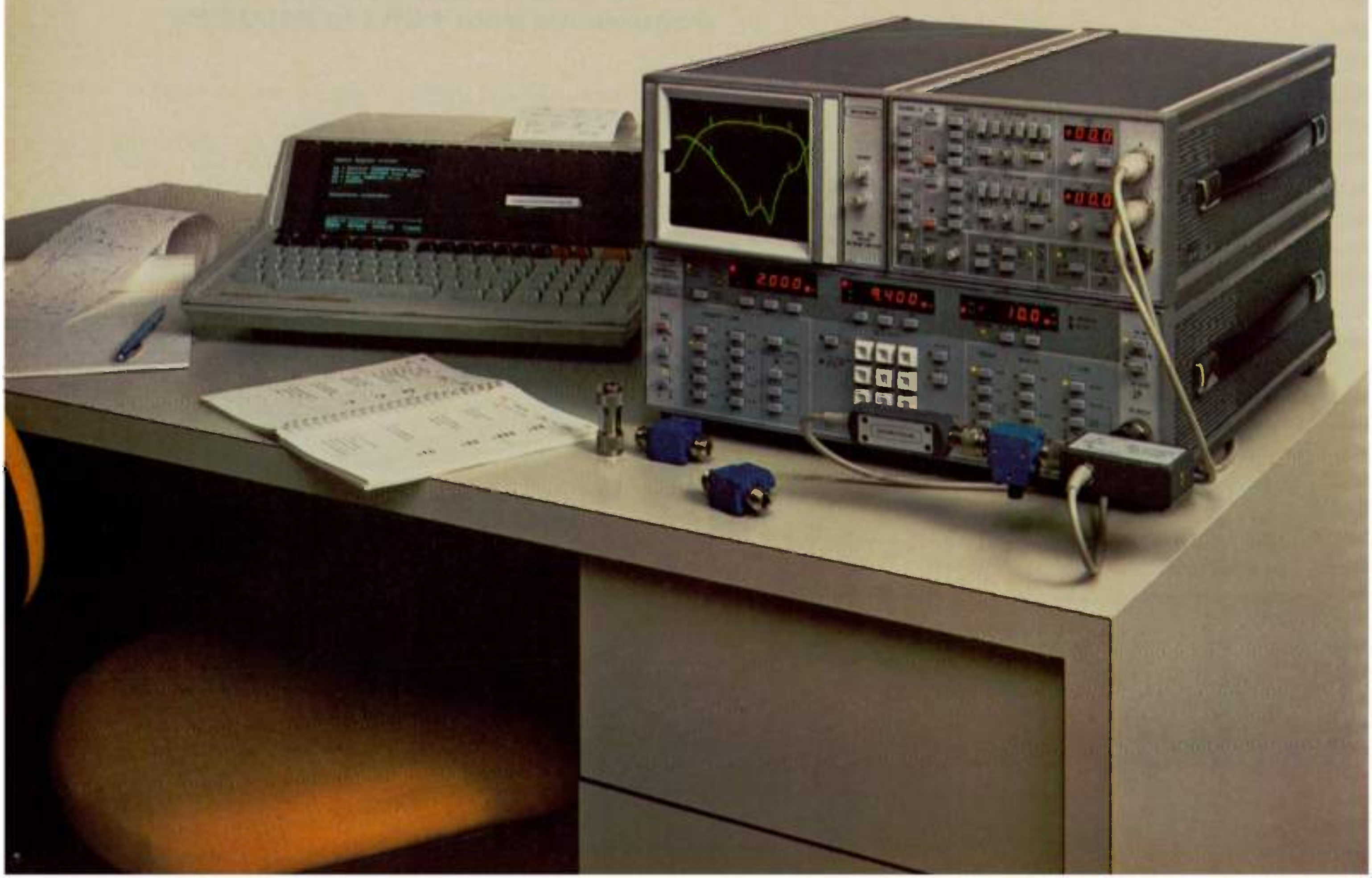

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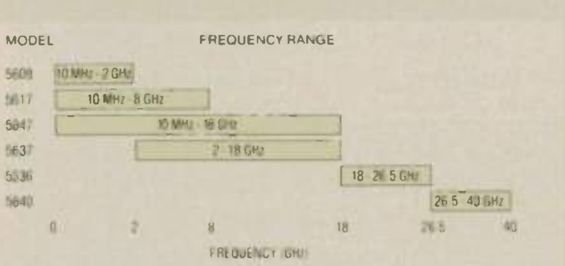
Each system consists of the Model 560 Scalar Network Analyzer, a 6600 Series Programmable Sweep Generator and the Model 85 Controller. Connect the SWR Autotester and Detector supplied, plug-in the factory pre-programmed cartridge and the system is ready to go.

40 dB Directivity

Wiltron's 5600 Series offers 40 dB directivity over a 10 MHz to 18 GHz continuous sweep range. Dynamic range is 66 dB with −50 dBM sensitivity. The system offers 82 dB programmable attenuation in 0.1 dB steps. ROM-corrected frequencies are accurate to ±10 MHz from 10 MHz to 18 GHz. Six models span the 10 MHz to 40 GHz range.



A key part of the system is the new Series 6600 Programmable Sweep Generator. This sweeper uses fundamental oscillators to avoid substantial errors generated by the harmonic products of multiplier type oscillators. The result, broadband coverage with the lowest harmonic content (−40 dBc, 2-18.6 GHz), low residual FM and greater stability.

A pre-processor chip separately scans front panel controls and interfaces directly with the main processor. Response speed is never a problem. All interfaces with the bus are internal, eliminating the need for an external interface box. For user convenience, up to 99 test set-ups can be stored in the control cartridge for future use.

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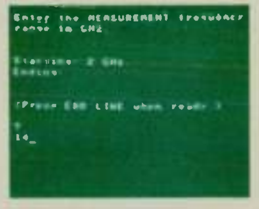
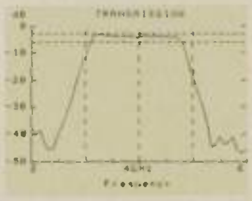


B. Frequency Selection
Enter 1) Frequency range limits
2) Frequency step size or number of test points

C. Calibration
Enter 1) DUT identification.
Select 1) Averaging of open/short residuals
2) Storing of normalized residuals

D. CRT display of DUT characteristics
1) Select marker frequencies and amplitude limits.
2) If necessary, adjust DUT
3) If not, continue

E. Measurement
1) Press key to start automatic measurement sequence



F. Hard-copy output
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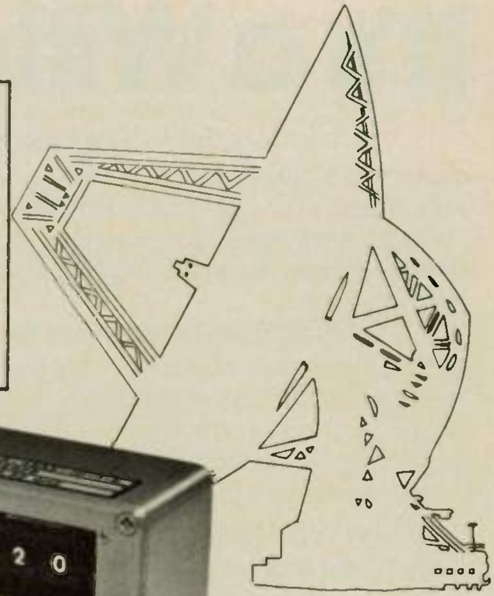

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Reference Input - Frequency:	5 MHz standard, 10 MHz and 100 MHz options
Power:	0±3 dBm
Frequency Step Size:	10 MHz
Frequency Stability:	Same as reference input
Spurious Outputs - Inband:	-70 dBc
Outband:	-55 dBc
Power Output Variation - Frequency:	±1.5 dB
Temperature:	±1.5 dB
DC Power:	+20V at 700 mA
Operating Temperature:	0 to 50°C
Weight:	23 oz.
Connectors - RF:	SMA female
DC:	Solder filter

Model Number	Output Frequency (GHz)	Power Output (Min.) (dBm)
PLS-3742-10	3.7-4.2	+13
PLS-4449-10	4.4-4.9	+13
PLS-4853-10	4.8-5.32	+13
PLS-6570-10	6.55-7.05	+10
PLS-7075-10	7.0-7.55	+10
PLS-7277-10	7.2-7.7	+10

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INTRODUCTION

A new series of full waveguide band millimeter-wave receiver components developed for the EW and ELINT programs of the 80s utilizes suspended microstrip to achieve levels of performance required for practical systems. In recent articles, several authors have addressed the requirements for broadband low cost receiver systems at millimeter wavelengths.[1,2] Much of the discussion has centered around the use of "fin-line" and soft-duroid microstrip techniques to achieve low-cost receivers at millimeter wavelengths. This is a logical development in the industry, since "soft-substrate" technologies are readily available in many laboratories that are geared to the production of lower frequency microwave circuits. On the other hand, those organizations with high technology resources and know-how that allow for the successful implementation of stripline or microstrip technologies on fused quartz at millimeter wavelengths, have produced rugged components which are found to be cost competitive and extremely reliable in actual system environments.

The success of fused quartz suspended stripline components in achieving state-of-the-art noise figures in relatively narrow band receiver applications has been well documented in the literature,[3,4,5,6] but the use of fused silica suspended stripline techniques to achieve full waveguide instantaneous bandwidths is a new and desirable benefit of a circuit medium which can be accurately and reproducibly fabricated for low cost rugged components useful in practical systems.

The use of full waveguide band receiver components is especially desirable for radar warning and ELINT applications that must survey large bandwidths searching for possible emitters in the millimeter spectrum. Meeting mission requirements with the minimum number of receiver channels required to do the job is the principle reason for seeking full waveguide-band capability in millimeter-wave receiver components now being developed.

By extending design principles successfully applied to the design of relatively narrow band fixed frequency receiver components such as radar receivers and missile-seeker radiometers, a new series of millimeter-wave mixer preamplifiers with full waveguide instantaneous bandwidths has been developed. The series includes 18-26 GHz, 26-40 GHz and 33-50 GHz balanced mixers with various LO and IF configurations to meet varied systems requirements. Additionally, a fully tunable 18-40 GHz mixer preamplifier with integral 2-6 GHz GaAs FET IF preamp (as pictured on this month's cover) is now being supplied for several prototype EW and ELINT systems. Common to all of these mixer preamplifiers are the benefits of small size and weight, very low cost in large quantities and tolerance to adverse environmental stress.

CONSTRUCTION DETAILS

The AK9700-18, 18-40 GHz mixer preamplifier with integral 2-6 GHz GaAs FET IF preamplifier utilizes a planar cross-bar mixer design in suspended quartz microstrip to achieve the desired broadband performance. The direct connection of the mixer substrate to the IF amplifier input stage without use of connectors allows for a rugged and reproducible integration of the multioctave 2-6 GHz IF preamplifier with the balanced millimeter downconverter. An OSSM connector is utilized to couple the local oscillator signal into the mixer and facilitates the use of 0.085" semi-rigid coaxial cable to route LO signals in practical system realizations.

In **Figure 1**, the various circuit elements of the receiver are illustrated. The LO signal, which is coupled to the mixer substrate from the OSSM connector using a short length of gold-ribbon, propagates in a suspended stripline circuit fabricated from 0.021" thick fused silica. In excess of 2.5 microns of gold is metallized and then photoetched on the millimeter substrate to in

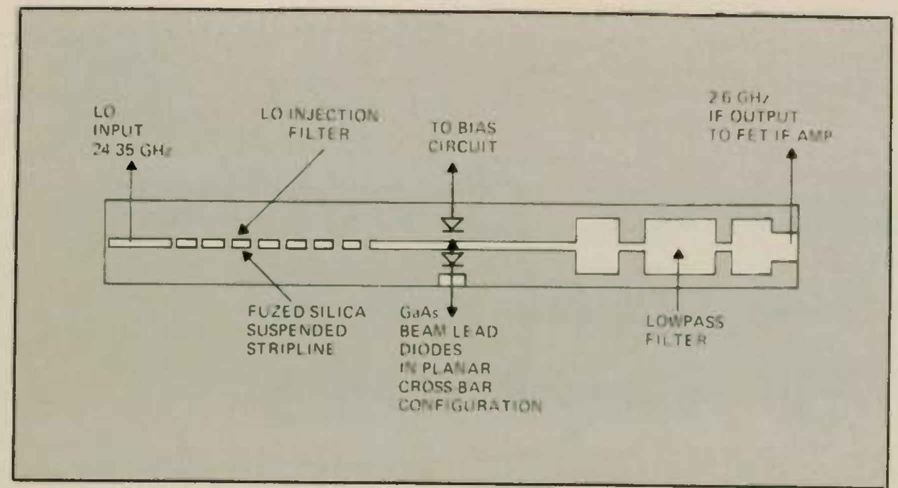

Fig. 1 Mixer substrate circuit elements.

sure low circuit loss and proper operation of the various filter structures. A seven-section injection filter is used to couple the LO signal to the matched pair of mixed diodes. The diodes are excited "out of phase" by the local oscillator signal and "in-phase" by the RF input signal. The RF signal is radiatively coupled to the diodes which appear in an equivalent series configuration to the fields propagating in the double-ridged waveguide mode. This provides for AM noise rejection in excess of 35 dB typically, and sufficient balance is achieved in this circuit to maintain at least 25 dB RF-to-LO isolation over the band.

The LO injection filter appears as an open circuit at the microwave IF frequency range used in this design. This allows the IF signal extracted from the diode pair to couple to a standard five section L-C low pass stripline filter and series matching network located between the diode pair and the FET amplifier. Miniature bypass capacitors enable dc bias lines to be connected to the series pair of mixer diodes, allowing the best overall dynamic operating point for the receiver to be realized. Bias is derived from a protected integrated circuit regulator, and is a low impedance voltage source when viewed "looking-back" from the diodes.

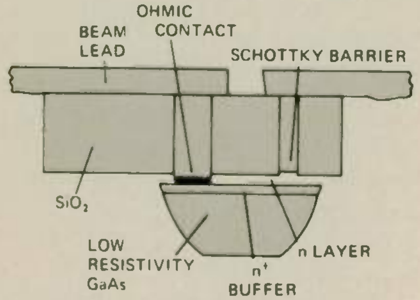

Fig. 2 Mm wave beam lead diode cross section.

Optimization of the location of the various mixer filter elements was accomplished using a X10 scale model operational from 1.8 to 4.0 GHz. In this model, conversion loss, noise figure and SWR were measured and later found to adequately represent the results obtained from the actual prototype 18-40 GHz receivers.

Specially developed millimeter beam lead GaAs semiconductors are used in the mixer (cross section shown in **Figure 2**). Typical devices exhibit 0.025 pfd total capacitance with an associated dc series resistance of 4 ohms, and are usable up through 230 GHz in properly designed stripline circuits.

The noise performance of the receiver is enhanced by the use of an integrated Alpha 2-6 GHz FET IF amplifier with a typical noise figure of 3.5 dB. This noise figure in a production design is derived from low loss, miniaturized MIC's together with low noise recessed-gate MESFET's with 1 μm gate lengths and an overall chip size of 0.12" x .016". Use of microprocessor controlled vapor phase epitaxy insures reproducible characteristics in production quantities. Typical devices (AFL-1000), illustrated in **Figure 3**, exhibit 1.2 dB noise figures with associated gains of 12 dB in room temperature applications near 4 GHz. The amplifier consists of three balanced stages with integral regulators similar to the commercially available Alpha AMA-3264-03 2-6 GHz GaAs FET amplifier. Performance data for a typical amplifier is shown in **Figure 4**.

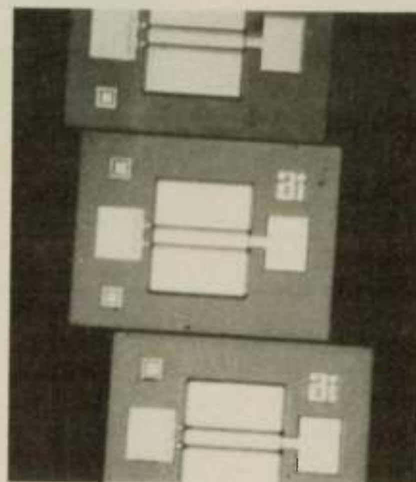

Fig. 3 ALF-1000 GaAs FET chips.

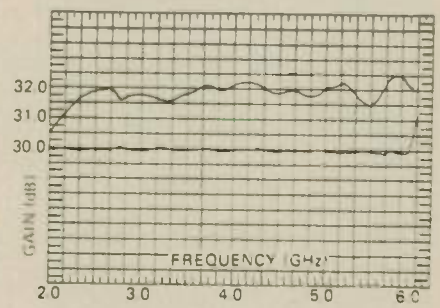

Fig. 4 AMA-3264-03 GaAs FET amplifier performance.

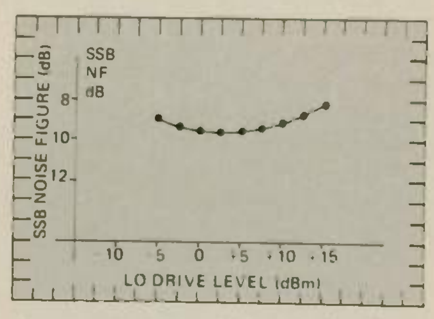

Fig. 5 18-40 GHz receiver SSB noise figure as a function of LO drive.

Bias circuitry, utilizing integrated regulators and filtering circuits, are also contained in the receiver enclosure which weighs less than 75 grams and occupies less than 1.80 cubic inches. The

(continued on page 40)

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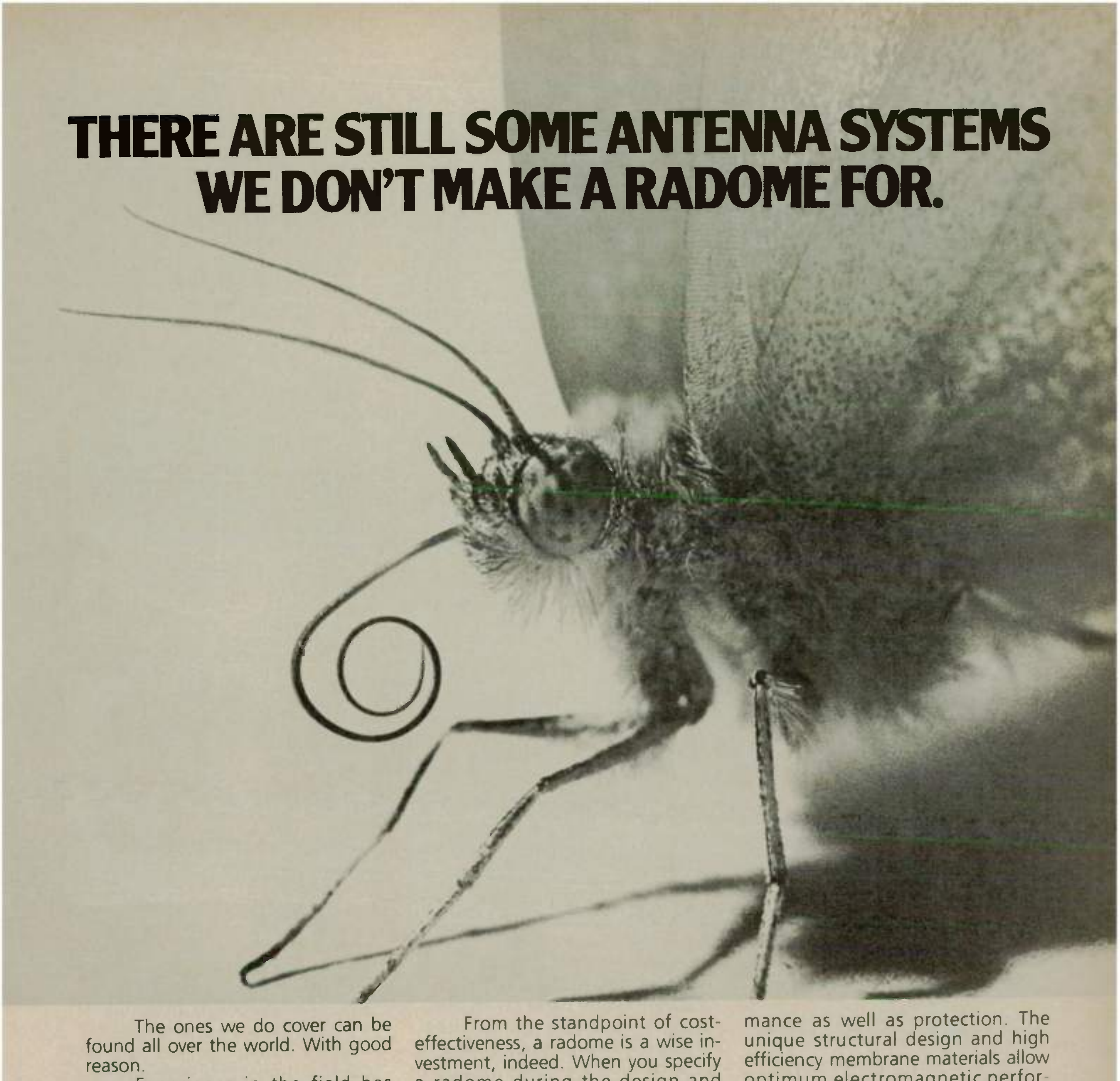

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TABLE I

TECHNICAL SPECIFICATION FOR 18-40 GHz INTEGRATED RECEIVER

RF INPUT:	18-40 GHz double-ridged waveguide to mate with UG-1586/U cover flange
LO INPUT:	23-34 GHz OSSM connector
IF OUTPUT:	2-6 GHz OSM connector
NOISE FIGURE:	10-11 dB SSB (including FET amplifier contribution)
RF TO IF GAIN:	23 dB nominal
LO TO RF ISOLATION:	25 dB minimum
LO TO IF OUTPUT ISOLATION:	60 dB minimum
RF TO IF OUTPUT ISOLATION:	60 dB minimum
2x2 SPURIOUS RESPONSE:	−29 dBc minimum for −15 dBm RF signal input
OTHER SPURIOUS SIGNALS:	−65 dBm minimum
GAIN TRACKING:	±2 dB between any two units
LO POWER LEVEL:	0 dBm to +13 dBm, +6 dBm optimum
−1 dB COMPRESSION POINT:	+10 dBm referred to IF output port (2-6 GHz)
MAX CW INPUT LEVEL:	+20 dBm (18-40 GHz)
RF, LO SWR:	2.5:1
IF SWR:	2.0:1 (50 ohm reference)
dc POWER:	+15 Vdc 130 mA typical

Mixer diodes are internally biased from integral regulator.

overall specifications of the receiver are outlined in Table I.

LOCAL OSCILLATOR REQUIREMENTS AND NOISE FIGURE

For practical millimeter systems, a receiver must be capable of operation over one full decade of LO drive levels (usually 0 to +10 dBm) and must be continuously tunable over a wide bandwidth with only minor changes in system performance. This is especially important where the millimeter mixer must be located at considerable distance from the local oscillator source. Use of coaxial cable for LO runs at millimeter wavelengths will be commonplace in ELINT and EW systems to reduce package size and complexity and to allow for ease of installation as retrofit/upgrade to previously built non-millimeter systems.

Fig. 6 18-40 GHz receiver Y-Factor over 4.5-5.5 GHz IF Band.

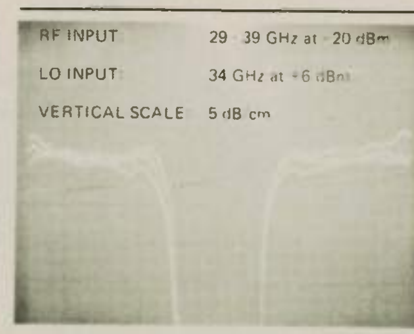

Fig. 8 Tracking characteristics for three 18-40 GHz receivers.

The 18-40 GHz receiver operates from −5 dBm to +15 dBm LO drive levels. Figure 5 illustrates single sideband noise figures for the unit, including the contribution of the integrated IF amplifier, over this drive range. The optimum LO drive level occurs in the range from 0 to +6 dBm where, typically, the lowest noise figure is achieved. The mixer can be pumped from 24 to 34 GHz enabling operation over the full 18-40 GHz bandwidth in either the upper or lower sideband.

In Figure 6, the broadband Y-Factor of one 18-40 GHz receiver is displayed at the 5 GHz IF center frequency at the IF output port. During this measurement, the receiver was pumped at 33 GHz at +6 dBm and a calibrated noise lamp with a WR-28

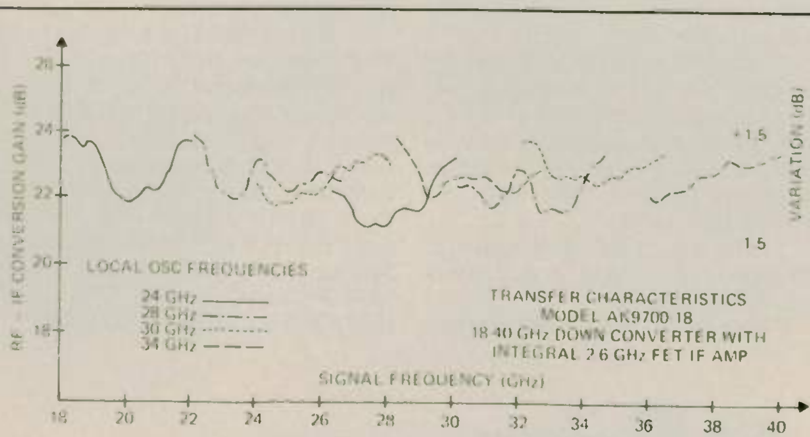

Fig. 7 18-40 GHz receiver transfer characteristics.

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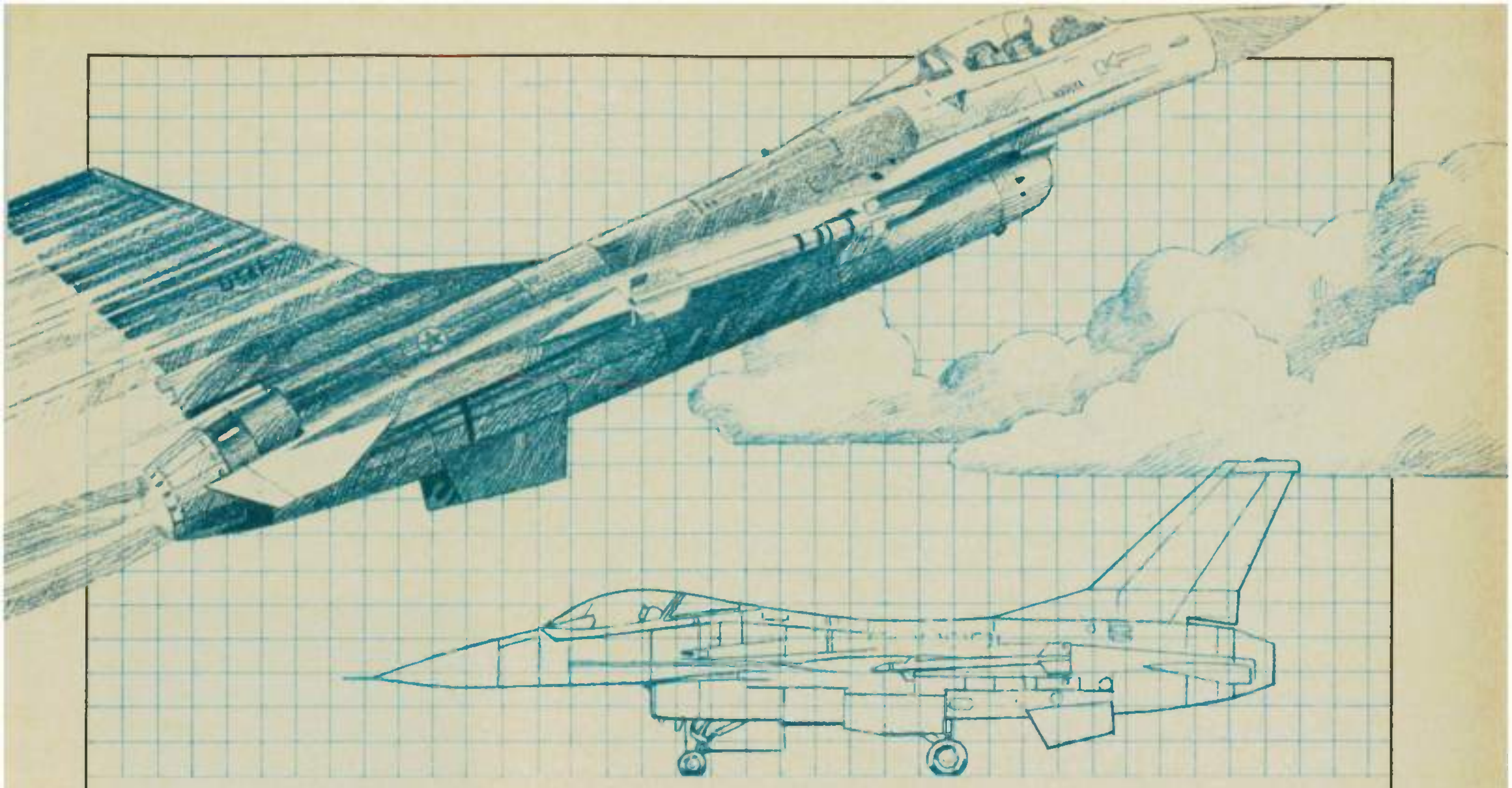

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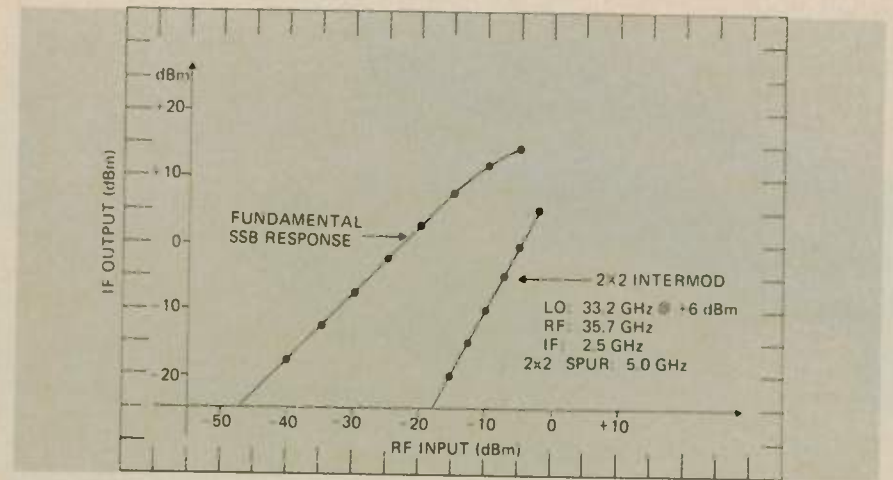

Fig. 9 RF-to-IF conversion gain and 2x2 intermodulation products for 18-40 GHz receiver.

to 18-40 double-ridged waveguide adaptor was coupled to the RF input port. The effective ENR of this source was 16.1 dB, referred to the mixer input flange. Double sideband Y-Factors in excess of 9 dB are obtained over the entire passband of the receiver. This corresponds to single sideband noise figures of 10 to 11 dB including the 2-6 GHz FET IF amplifier contribution.

FLATNESS AND TRACKING

In many applications, matched pairs of receivers will be used for DF or other applications where absolute flatness and tracking from unit to unit must be achieved. **Figure 7** shows the effects of tuning the local oscillator to various points in the band and demonstrates a maximum excursion of ± 1.5 dB peak to peak around the nominal conversion gain. **Figure 8** compares three complete 18-40 GHz receivers at one particular LO frequency (34 GHz) and indicates that the variation from receiver to receiver is less than ± 2 dB. This general tracking performance is also observed for other levels of LO drive (0 to +10 dBm), RF frequency and IF frequency that can be applied to the receivers. The RF to IF conversion gain and 2x2 intermodulation products measured for typical receivers is illustrated in **Figure 9**. The nominal +10 dBm (-1 dB) compression point of the transfer characteristic is primarily determined by the output stage of the GaAs FET IF amplifier, while the 2x2 intermodulation product is determined by the LO drive level applied to the singly balanced mixer. In this data, the RF signal was applied at 35.7 GHz.

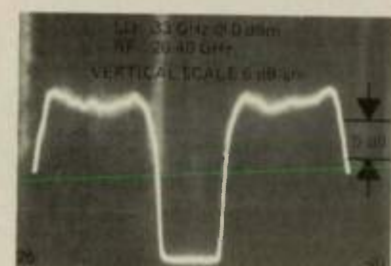

Fig. 10 18-40 GHz receiver swept RF transfer characteristics – upper and lower sidebands.

The flat nature of the RF transfer characteristic is also evident in the scope photograph of **Figure 10**, where the input of the receiver is swept over 26-40 GHz and the two responses (upper and lower sideband) are viewed centered about the LO set at 33 GHz. One can see the symmetric IF amplifier response 2 to 6 GHz below and above the "zero-beat" at 33 GHz. A closer look at the typical swept response of one sideband centered on 34 GHz appears in **Figure 11**, where flatness of the order of ± 0.5 dB is achieved.

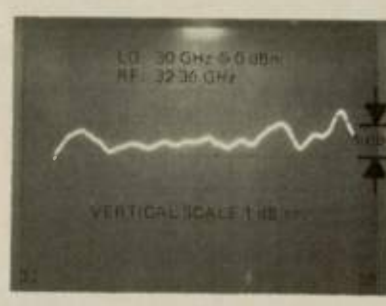

Fig. 11 Upper sideband swept RF transfer characteristic of 18-40 receiver.

FUNDAMENTAL IF RESPONSE BROADER THAN 2-6 GHz

The mixer can be assembled without the integral FET IF amplifier for use with systems having non-standard IF requirements or where existing FET IF amplifiers are available in retrofit applications.

The balanced mixer exhibits flat response from 20 MHz out to 8 GHz at the IF output port. Typical flatness is better than ± 0.5 dB out through 1 GHz and better than ± 1.0 dB through 8 GHz (see **Figure 12**). Two mixers with slightly different IF coupling capacitors were used in these measurements, one opti-

Fig. 12 18-40 GHz balanced mixer IF response.

(continued on page 44)

100 dB Attenuation Measurements to 18 GHz

The 1290 Receiver Offers...

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(from page 42) MILLIMETER RECEIVERS

mized for VHF/UHF operation and the other for microwave IF response. The mixers exhibit a 50 ohm IF output with residual SWR of less than 1.7 over the entire 0.02 to 8.0 GHz IF range. Illustrated in Figure 13, the mixer occupies less than 0.95 cu. in. and weighs less than 40 grams.

COST-EFFECTIVE FUSED SILICA

The use of fused silica for construction of millimeter integrated circuits is both cost effective and mass producible. Soft-substrate approaches may prove worthwhile for many less demanding applications, but where considerable environmental stress and large temperature extremes must be tolerated, the use of fused silica substrates is a reliable and proven approach. This is especially true when one considers that physically small millimeter-wave semiconductors must be bonded to a reliable temperature-stable media.

Substantial shock and vibrational testing done to assess performance of fused silica millimeter circuits in several missile seeker programs have shown tolerance of this medium to shock levels greater than 10,000 g's. Continuous and random vibration of magnitudes encountered in tactical weapon systems do not damage fused silica substrates. Microcircuit bonds made between millimeter semiconductors and gold conductor patterns on fused silica substrates are highly reliable and have a long history of use in the micro-circuit community. Bonds to gold stripline conductors can easily be accomplished in production line environments using several approved bonding approaches including ultrasonics, thermal-compression and micro-welding techniques. On the other hand, attempts to fabricate soft substrate circuits with smaller and more delicate beam lead devices used at millimeter-wave lengths requires specialized soldering or welding techniques. These are not as well suited to avoiding damage to the semiconductor over large and rapid fluctuations in ambient temperature. Soft substrate finline circuits can deform to some extent and place mechanical stress upon mounted semiconductors as a response to rapid (greater than 5°C/min.) temperature variations in systems applications.

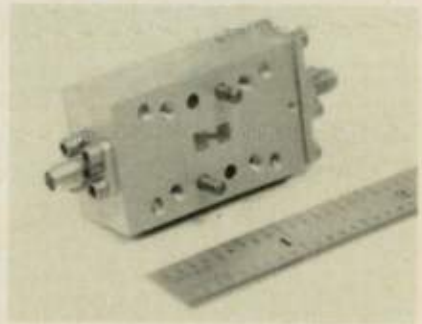

Fig. 13 Model AK9700 18-40 GHz balanced mixer.

The actual cost of the fused silica substrate is inconsequential (typically less than $5.00 per mixer substrate in 10,000 piece quantities for a substrate with metallized millimeter circuit ready for the bonding of millimeter-semiconductors) in the cost-performance economics of millimeter componentry, but used of fused silica as a circuit medium does require a substantial technology base not found in every manufacturing facility. For these reasons, the use of system components made from fused silica for full waveguide band operation at millimeter wavelengths is desirable, and will allow state-of-the-art performance in actual systems applications until late in this decade when monolithic technologies may develop new receiver components of even higher performance at a reasonable cost.

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RADANT:
New Method of Electronic Scanning

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INTRODUCTION

In the case of a conventional phased array antenna,[1] the phase gradient across the array required to change the beam direction is obtained by changing the phase of discrete components. The process of electronic scanning described in this article uses the principle of modifying the refractive index of a lens made of an artificial dielectric. By this technique, discrete phase shifters are eliminated and the phase control becomes truly distributed within a radiating lens aperture made up of a diode controlled artificial dielectric. The artificial dielectric consists primarily of grids of wires containing many diodes connected together. By changing between the forward biased and reverse biased states the desired change is made in the index of refraction.

This method of electronic scanning is called the RADANT® principle and is the result of the contraction of the words "radome" and "antenna." This principle has led to studies[3,4] and antenna developments for ground and airborne radar applications.

Editor's Note
This article first appeared in French in *L'Onde Electronique*, Dec. 1979, Vol. 59, No. 12. Initially this work was performed completely in France at the Société d'Etude du Radant, Orsay, France. Currently, research, development and production capability is available in the United States as well in the recently opened US affiliate, Radant Systems Inc., Stow, MA, a modern and complete antenna facility. Inquiries may be directed to either the US or French facilities.

EFFECT OF A GRID OF METALLIC WIRES ON A PLANE ELECTROMAGNETIC WAVE

To understand the way in which the artificial dielectric controls phase it is best to consider the transmission phase network representation of an admittance Y which shunts a transmission system of Z_0 impedance. Transmission line phase shifters of this kind have been described in the literature.[2] The analysis will be described here as it applies to the RADANT embodiment. The reflection and transmission of a plane electromagnetic wave incident to a plane can be described by referring to Figure 1.

In general the equivalent normalized admittance, Y, of the plane is of the form

$$Y = G + jB$$

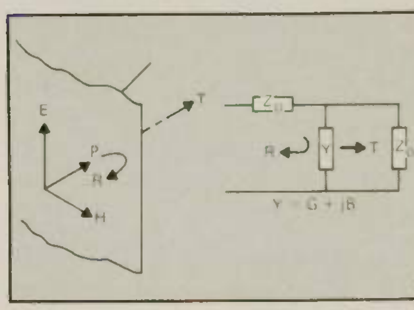

Fig. 1 Diffraction of a plane wave through a plane—equivalent circuit.

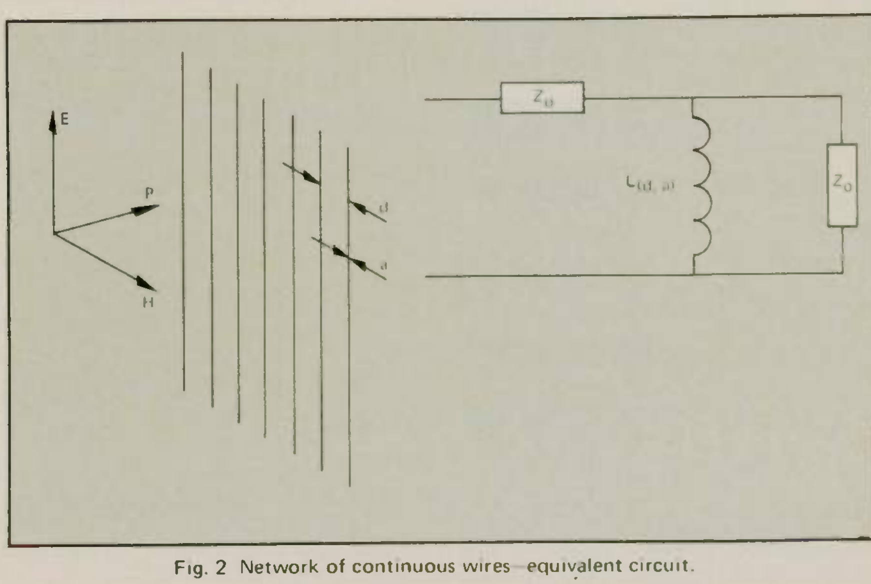

Fig. 2 Network of continuous wires—equivalent circuit.

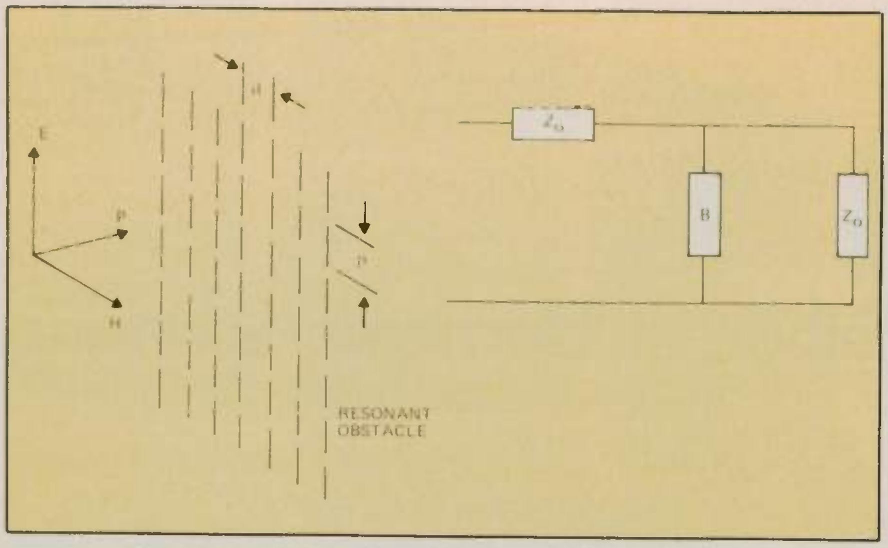

Fig. 3 Network of discontinuous conductors–equivalent circuit.

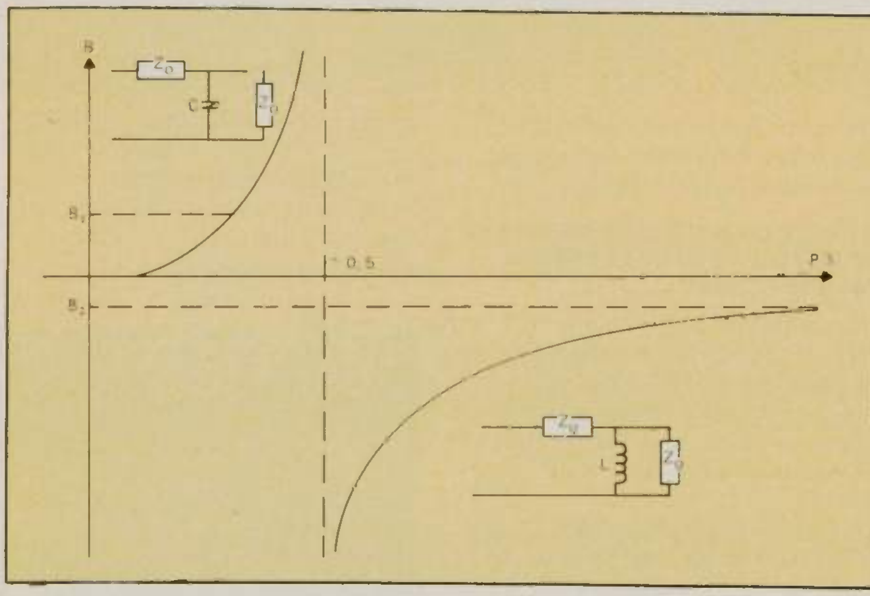

Fig. 4 Susceptance variation of a discontinuous wire network as a function of the length of the cut wires.

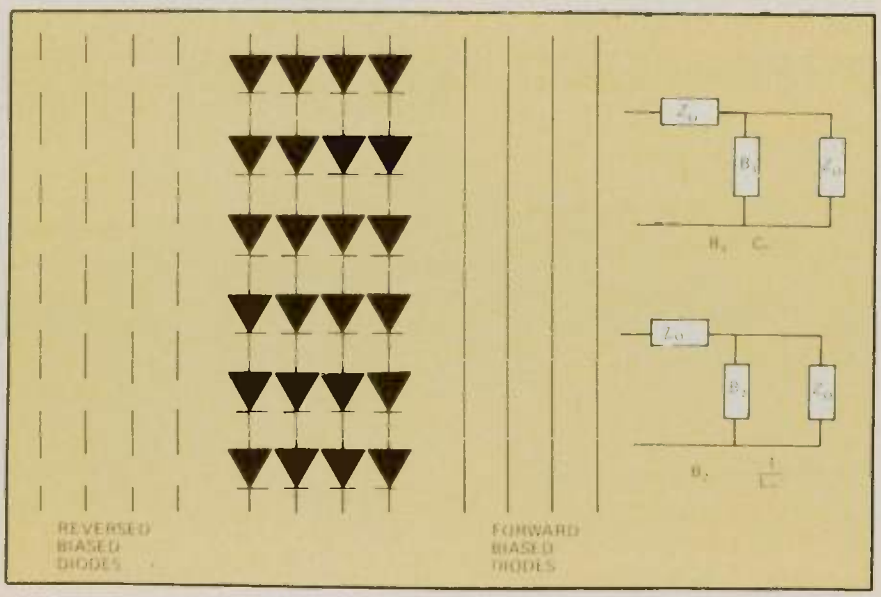

Fig. 5 Wire–diode grid and equivalent circuits for both bias conditions.

If we assume the admittance is lossless, then Y = jB and there are simple expressions for the susceptance, B, and the coefficients of reflection, R, and of transmission, T.

$$R = \frac{|B|}{\sqrt{|B|^2 + 4}}$$

Where tan (ϕ) = − B/2 and ϕ is the phase shift during transmission.

Also: $|R|^2 + |T|^2 = 1$

It is convenient to consider the admittance Y to represent a "plane," that is, surface within the dielectric lens in which controlling elements are located. The presence of the control elements in an otherwise uniform dielectric result in an overall lens having a nonhomogeneous or, as more commonly termed, "artificial" dielectric. If the plane is a grid of continuous conducting wires parallel to the electric field, E, of the incident wave (Figure 2), the equivalent reactance, x = 1/B, is inductive and its value depends on the wavelength, the spacing, d, between the wires, on the diameter, a, of the wires, and on the incident angle θ of the plane wave.

$$X = \frac{-B}{G^2 + B^2} = \frac{d \cos\theta}{\lambda} \cdots \left[L_n \frac{d}{\pi a} + \frac{1}{2} \sum_{\substack{m=-\infty \\ m \neq 0}}^{+\infty} \left(\frac{1}{\sqrt{m^2 + \frac{2md}{\lambda} \sin\theta - \frac{d^2 \cos^2\theta}{\lambda^2}}} - \frac{1}{|m|} \right) \right]$$

Use of a continuous wire grid is well known[5] and is often used in polarizers or to tune dielectric panels to make them matched at specific frequencies for microwave "windomes" and "radomes."

The method to be described is derived from a lesser utilized configuration using interrupted wires, which produce resonant structures[6,7], as shown in Figure

. In the RADANT steerable lens pproach, however, the wire grids re interrupted using PIN diodes n order that their susceptance an be electronically varied. This epresents a new and hitherto un-escribed method for realizing lectronic antenna scanning.

A grid consisting of equal ength cut wires (no diodes) has n equivalent susceptance, B, hat primarily depends on the ength, p, of the cut wire lengths Figure 4).

The other parameters that deter-nine B are as follows in decreas-ng order of importance:

. Distance "d" between parallel wires,
2. Diameter of the wires, a.
3. Distance "e" between collin-ear dipoles.

For a length, p, less than a half-vavelength, and for a fixed dis-ance, d, and diameter, the wire rid is capacitive. With p greater han a half-wavelength the grid ppears inductive. For a larger p he susceptance approaches that of a continuous wire. Resonance s obtained for a value of p close o that of a half-wavelength.

DESCRIPTION OF A PHASE SHIFTING

If, instead of the grid of cut vires, we connect wire segments ogether using PIN diodes (Figure 5), a two state admittance (sus-eptance, neglecting loss) results. When the voltage applied to the diode strings causes the diodes to onduct, the grid is equivalent to network of wires which has a usceptance, B_2. If we apply a oltage in the reverse direction, he grid is equivalent to a net-vork of cut wires of susceptance, B_1, and this can be capacitive vith the proper choice of the alues of p, d, wire diameter, and diode parameters.

The diode characteristics also affect the value of susceptance, B_1. Between the two states, for-vard and reverse bias, of the di-odes, the equivalent susceptance s different. On the curve shown n Figure 4, note the two values of susceptance, B_1 and B_2. This change of susceptance state causes a change in the transmis-

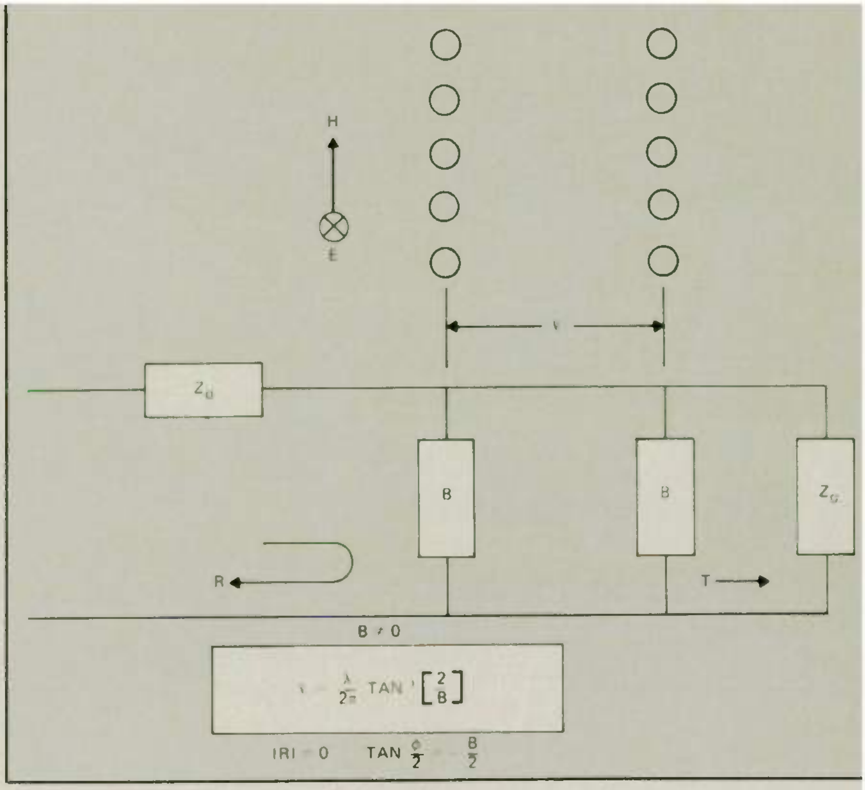

Fig. 6 Matching principle using two identical wire networks.

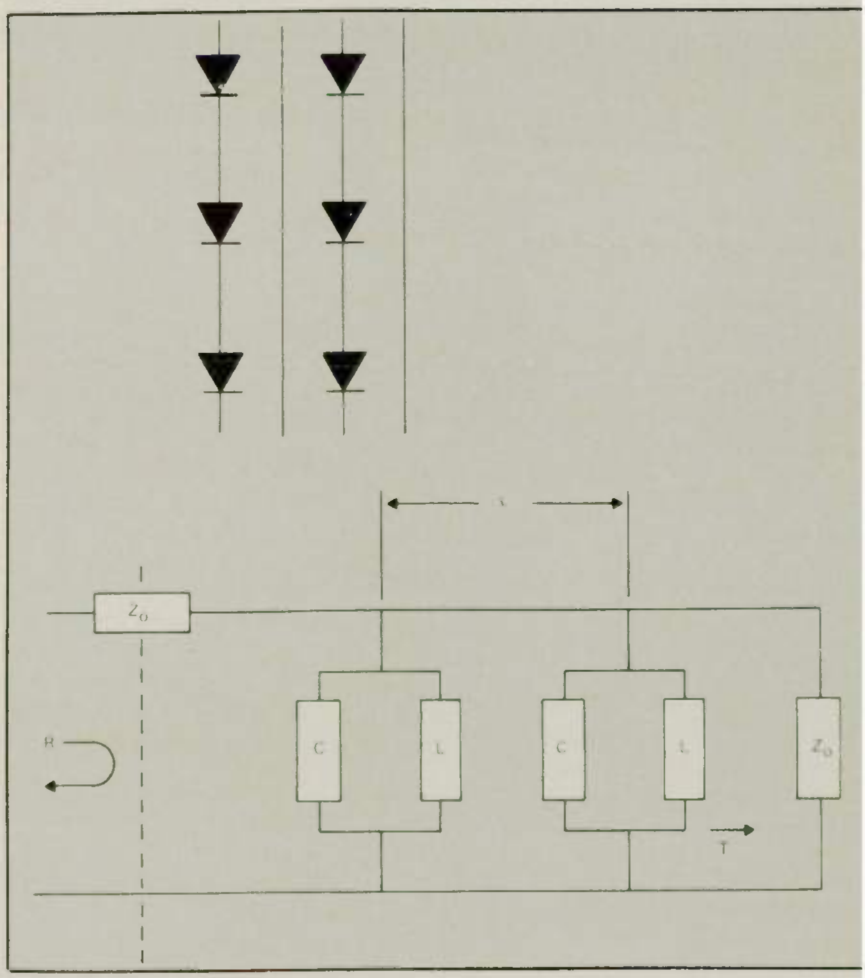

Fig. 7 Phase shifting panel—equivalent circuit—diodes reverse biased.

(continued on page 5

The price is righ

sion phase of the incident wave as it passes through the lens plane containing the diode string. Along with the desired phase change, however, there is an attendant change in the coefficient of reflection.

Since we desire to provide a match of the incident wave to free space, that is to say to prevent reflections (and have a SWR equal to unity), a "panel" consisting of two grid planes identically switched, and parallel to one another are used. As will be shown, the use of a number of such matched panels provides the means to steer a beam electrically. In **Figure 6**, two planes are shown separated by a distance, ℓ, from each other. If the susceptance B of the grids is nonzero, the distance, ℓ, for which the two susceptances B match one another is given by:

$$\ell = \frac{\lambda}{2\pi} \tan^{-1} \left(\frac{2}{B}\right)$$

for which case the net reflection coefficient of the pair is $R = 0$. The change in transmission phase $\Delta\phi_1$ for the wave after it has encountered the panels is:

$$\tan \frac{\Delta\phi_1}{2} = - \frac{B_1}{2}$$

where B_1 is the grid susceptance in one of the diode states. Because the diode switched grids can have two states, two separate nonzero susceptances (B_1 or B_2) can be perfectly matched for only one of these values, because the distance, ℓ, is fixed. Therefore, we must find another independent design parameter to adjust in order to provide a match for both values of B.

This independent parameter can be obtained practically by interleaving a continuous wire grid with the diode network as shown in **Figure 7**.

The distance between the wires and the spacing of the diodes are chosen in such a way that, when reverse biased, the capacitive effect of the diodes is parallel resonated by the inductive effect of the continuous wires.

At a given frequency ($f = \omega/2\pi$) the parallel resonance is obtained when $LC\omega^2 = 1$. The net equivalent susceptance in this case, as well as the associated transmission phase shift $\Delta\phi_2$ are zero.

When we then apply a voltage so that the diodes conduct (see **Figure 8**), there are two inductive networks for which a new value of ℓ can be selected for a transmission match, satisfying

$$\ell = \frac{\lambda}{2\pi} \tan^{-1} \left[\frac{2}{B}\right]$$

$$= \frac{\lambda}{2\pi} \tan^{-1} \left[- L\omega\right]$$

Where the net susceptance B is now $-2/L\omega$ and the reflection coefficient, $R = 0$.

Between the two states of the diodes, the differential change in phase is:

$$\Delta\phi = \phi_1 - \phi_2 = \phi_1$$

and the net reflection coefficient for the pair of grid planes making up a panel remains zero in both states.

This basic concept of producing matched transmission phase shift can be used to construct a lens which consists of many pairs of panels to achieve "multibit" phase control.

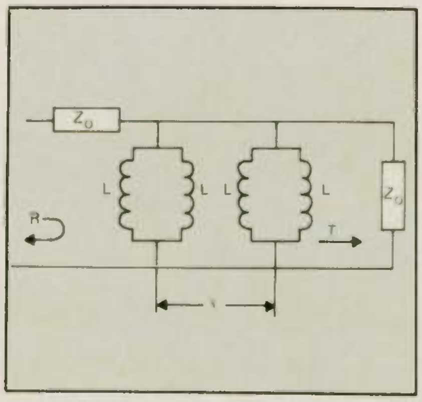

Fig. 8 Phase shifting panel—equivalent circuit—diodes forward biased.

RADANT LENS

Using the preceding panel as an example, suppose that the diodes in one half of the panel are conducting, while those in the other half are non-conducting (see **Figure 9**). The wavefront, after traversing the panel evidences a discontinuity in transmission phase between the two

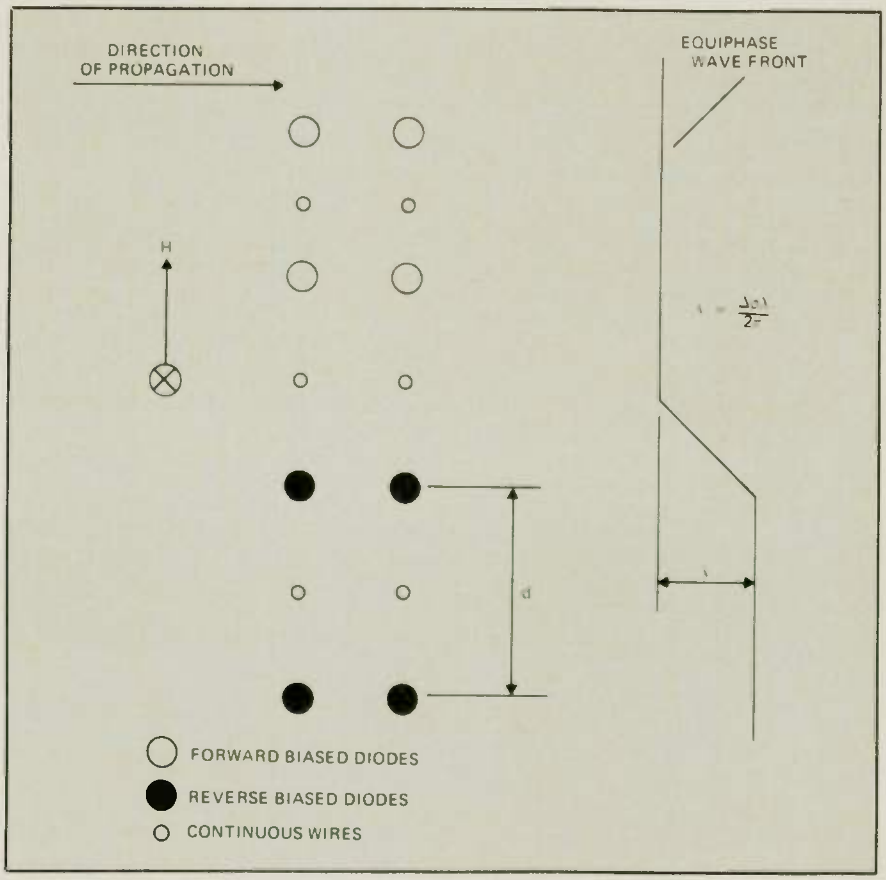

Fig. 9 Phase shift as a function of diode biasing.

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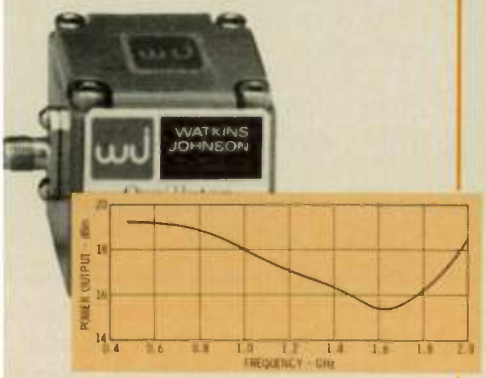
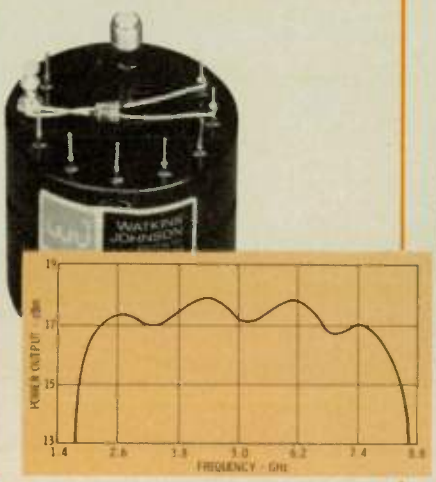
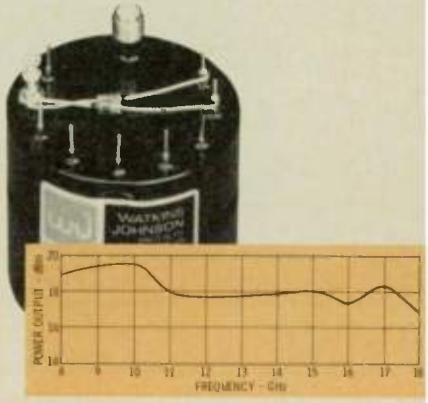
Watkins-Johnson Yig-tuned Oscillators—0.5-18 GHz in Three Easy Steps.



1. 0.5–2 GHz



2. 2–8 GHz



3. 8–18 GHz

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$$\tau = \frac{\Delta\phi\lambda}{2\pi}$$

A Radant lens (Figure 10) is made by using N panels identical to the one shown in Figure 9 and for which:

$$\Delta\phi = \frac{2\pi}{N+1}$$

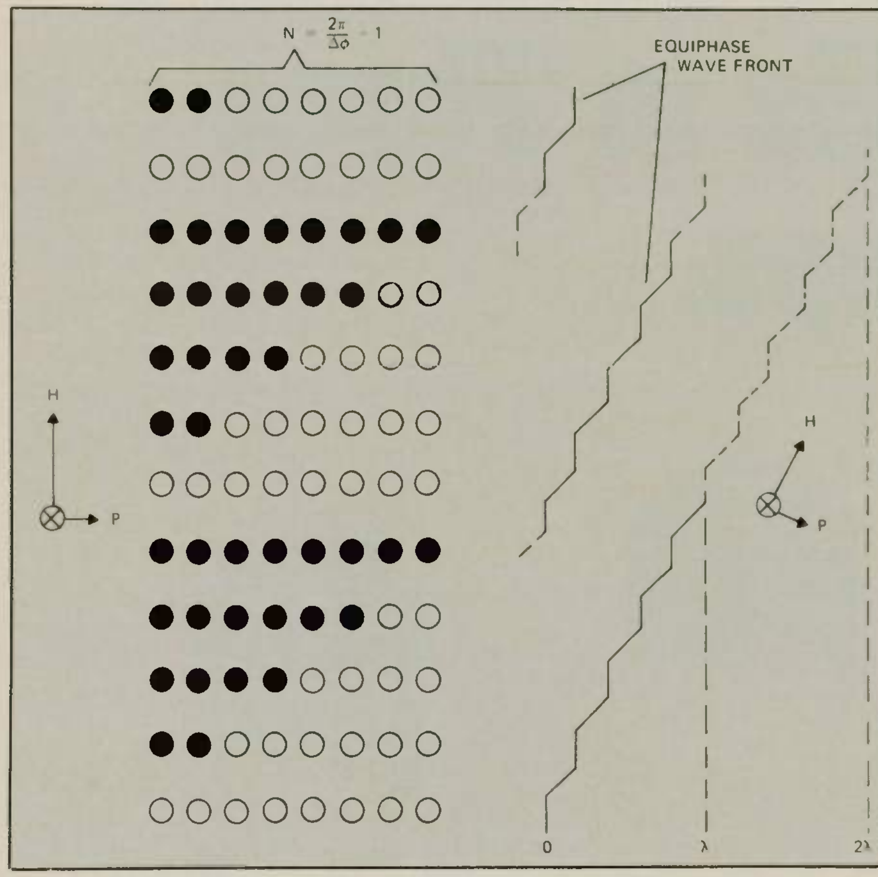


Fig. 10 Change of scanning angle through a radant lens.

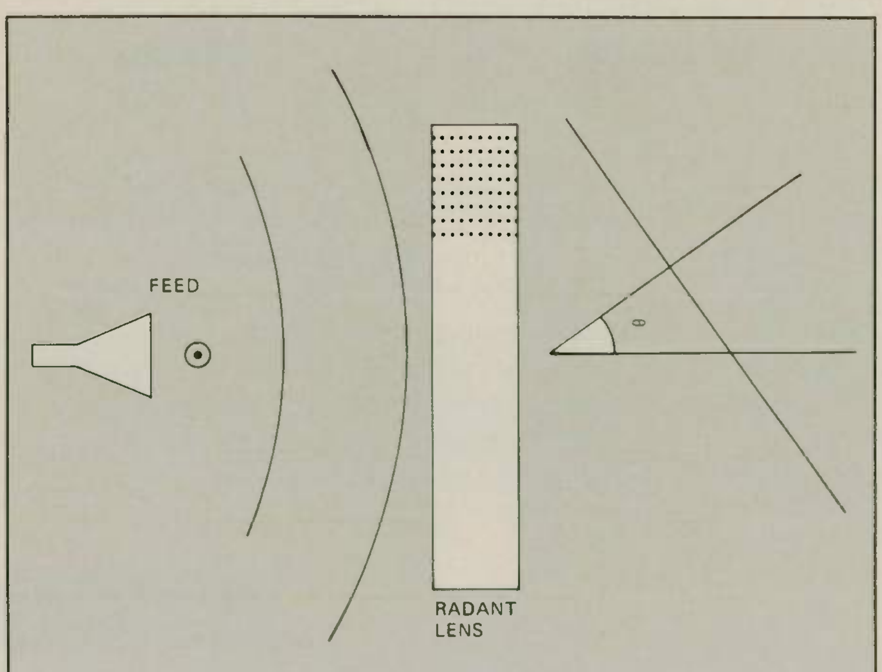


Fig. 11 Radant antenna single plane scanning.

If all of the panels are to be made identically, then $\Delta\phi$ must be designed to be a sub-multiple of 2π. With a lens constructed in this manner, we can increment the phase by $\Delta\phi$ from 0 to 2π.

If the diodes are made to be either conducting or non-conducting within the panels in such a way that from one section of diodes to the other the change in phase varies by $\Delta\phi$, the wave front will no longer be parallel to the plane of the incident wave. Instead, the resulting beam will have been steered in direction by an angle θ, based on the relationship:

$$\frac{2\pi d \sin\theta}{\lambda} = \Delta\phi$$

Rapid change in beam direction can be accomplished by high speed electronic control of the diode bias states. Second order effects, to be discussed later, occur when there is coupling from one section of the diode array to the other, in which case the algorithm required to steer the beam is not as simple an application as the preceding formula.

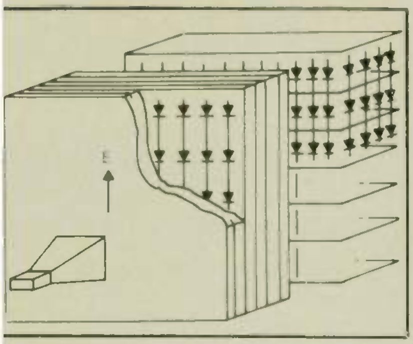
RADANT ANTENNA

To make a complete antenna employing the Radant lens scanning principle, a variety of feed systems to illuminate the lens can be used (see Figure 11), including:

- A feed horn that can provide a spherical wave with a single polarization.
- A lens whose wires are parallel to the electric field of the incident wave.

The lens, besides steering the beam, can focus the spherical incident wave so as to generate a plane wave perpendicular to the axes of the wires (or diodes). Focusing in the orthogonal plane can be accomplished at the feed or by a passive lens placed between the feed and lens.

For three dimensional scanning, it is necessary to place two Radant lenses (see Figure 12) in front of the feed. The first Radant lens focuses and scans the beam in the horizontal plane. The second lens focuses and scans in the vertical plane.



ig 12 Radant antenna dual plane scanning.

CONCLUSION

The Radant lens array has certain advantages over the conventional, phase shifter arrays.

- A fundamentally simpler design that leads to a cost saving potentially one half or greater than a conventional array.
- With a flat feed, a thin antenna array can be built.
- The need for fewer control circuits results in reduced cost (n + m instead of n x m circuits).
- A reduction in scan losses (i.e. cos θ instead of $\cos^{3/2}\theta$ or $\cos^2\theta$).
- Distributed phase shifting which allows a high power handling capability, and decreases the parasitic side lobe level due to tolerances.

The greatest advantage of the Radant lens array is to accomplish electronic scanning at a lower cost than presently possible using conventional techniques. In a second part (to be published shortly) practical methods and results of the RADANT technique will be described.

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Chang, W. H., Infinite Phased Bipole Arrays, *Proc. IEEE*, 1966, Vol. 56 pp. 1892-1900.

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† Patent Pending

Model Number	Attenuation Range/Steps (dB)	Cell Increments (dB)
3200-1	127/1	1, 2, 4, 8, 16, 32, and 64
3200-2	63.75/0.25	1, 2, 4, 8, 16, 32, 0.5 and 0.25
3200-3	16.5/0.1	0.1, 0.2, 0.4, 0.8, 1, 2, 4 and 8
3200-4	150/10	10, 20, 20, 20, 20, 20, 20, and 20
3201-1	31/1	1, 2, 4, 8, and 16
3201-2	120/10	10, 20, 30, and 60
3201-3	12/1	1, 2, 3, and 6
3201-4	1.2/0.1	0.1, 0.2, 0.3, and 0.6
3202-1	0.5/0.5	0.5

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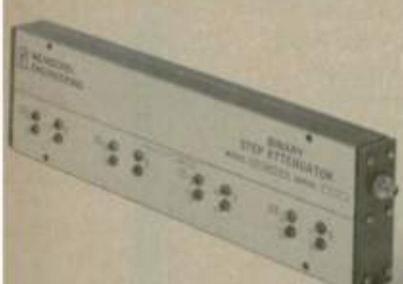

- dc — 18 GHz, covered in four models
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- VSWR — 1.80 Maximum
- Magnetic-latching switching mechanism

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  - — 0 to 9 dB/1 dB steps
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- VSWR — 1.75 Maximum
- WPM Connectors — mate with SMA per MIL-C-39012

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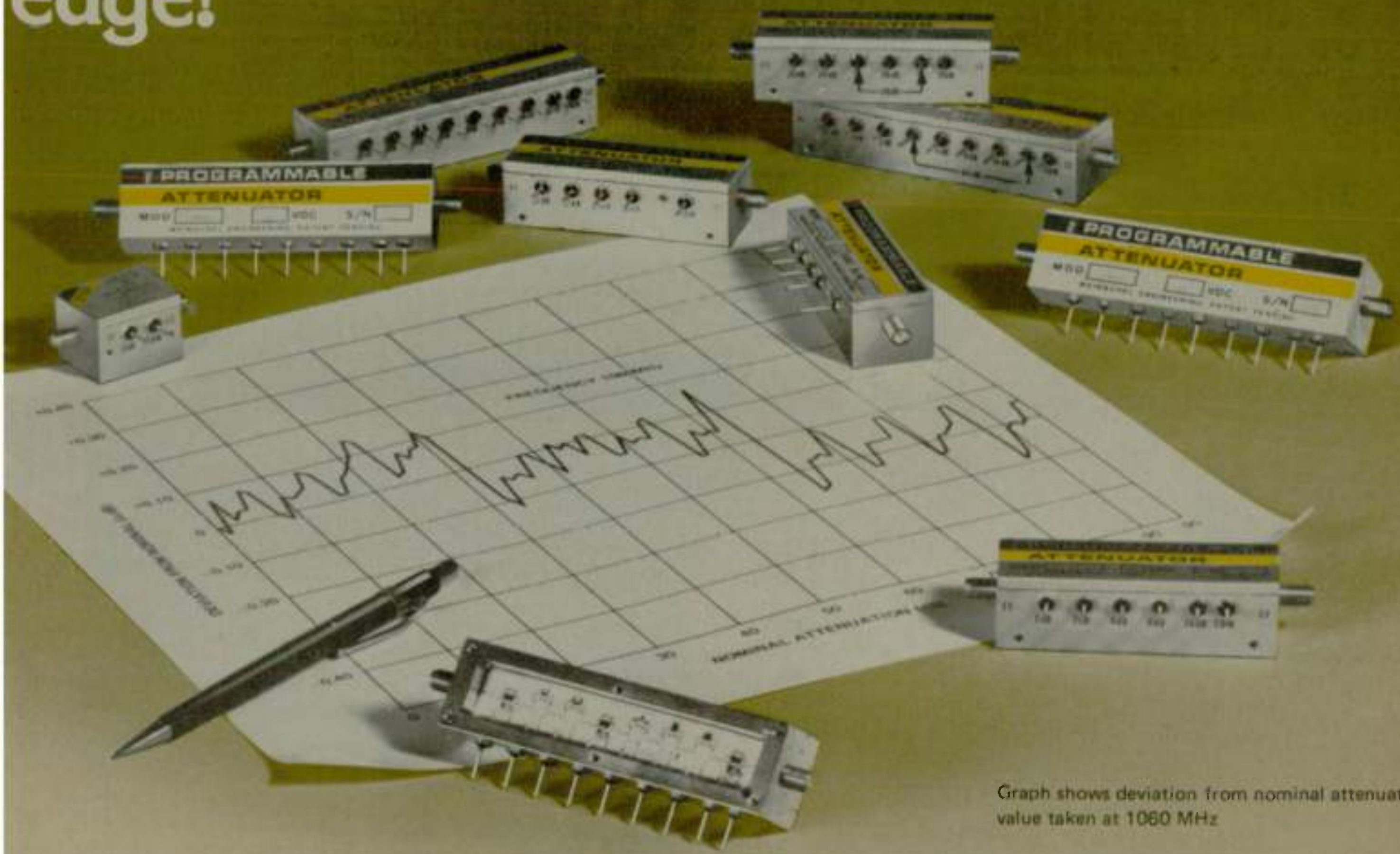
edge!



Graph shows deviation from nominal attenuation value taken at 1060 MHz

C30

Miniature Step Attenuator Model 115A

- dc — 18 GHz, covered in four models
- Five Standard Attenuation Ranges/Steps — up to 0-99 dB/1 dB steps
- VSWR — 1.70 Maximum
- Choice of Connectors — WPM (m or f), Type N (m or f), TNC (m or f), or 7 mm precision.

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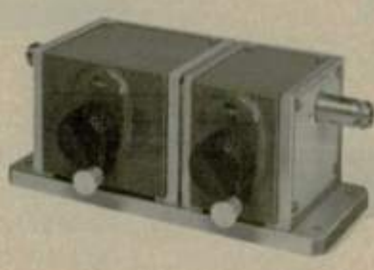
Precision Manual Step Attenuators Models 93, 94 and 95



- dc — 18 GHz
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- VSWR — 1.50 Maximum, Model 93
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PART II

Wideband ESM Receiving Systems

C. B. HOFMANN
A. R. BARON
Amecom Division
Litton Systems, Inc.
College Park, MD

Part I of this article, published in September 1980, discussed the basic need for wideband receiving systems for ESM missions and described suitable crystal video, acousto-optic, compressive and IFM receiver designs. Part II continues with a discussion of the channelized receiver and a tabular comparison of the advantages and disadvantages of the various receiving systems.

CHANNELIZED RECEIVER

The channelized receiver offers the advantages of the high intercept probability of the crystal video or wide-open IFM receiver with the sensitivity and frequency resolution of the superheterodyne receiver. In addition, the channelized receiver is finding increasing use in applications where accurate frequency measurement under conditions of overlapping pulses resulting from either high pulse density and/or deployment of multibeam radar systems is of concern. The resolution of such signal conditions has become a necessary requirement in many instances if accurate signal identification is to be performed. The wide-open IFM receiver technology although capable of recognizing when a simultaneous pulse condition exists can only measure the strongest signal and cannot provide information that will allow two or more signals to be separated.

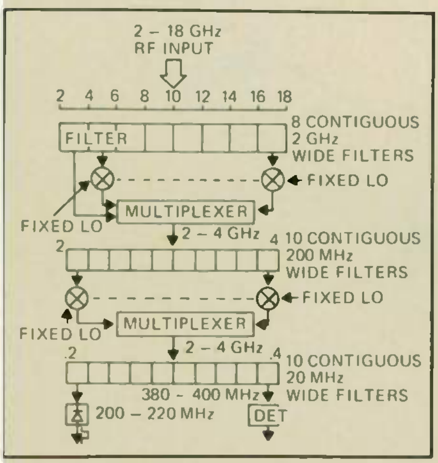

Fig. 11 Conceptual channelized receiver block diagram.

The channelized receiver consists of banks of contiguous filters that cover the frequency band of interest followed by individual signal detectors that determine when a signal is present within the filter bandwidth. In practice, multiplexing and down conversion to a common baseband is usually employed to reduce the complexity and overall complement of hardware. For example, as shown in **Figure 11**, a channelized receiver covering 2-18 GHz might be configured as follows: (1) a set of contiguous filters covering 2 GHz each from 2-18 GHz, followed by down converters that transform the 4 to 18 GHz range to 2-4 GHz; (2) the independent 2-4 GHz outputs are multiplexed into a contiguous filter bank of ten 200 MHz wide filters covering 2-4 GHz; and (3) these outputs could then be further down converted and multiplexed to a common 200 MHz wide IF output. A third bank of 10 filters with associated detectors might then be used to further resolve the pulse frequency to ±10 MHz.

It is important that the initial selectivity be good if dynamic range on the order of 60 dB is to be obtained. Wideband balanced mixers driven by at least 20 dBm of LO power are also usually required to keep spurious responses down.

Figure 11 shows two multiplexers; the second multiplexer is sequenced through each of its steps for each step of the first multiplexer. This is essentially equivalent to a superheterodyne receiver stepping through the R band with a 200 MHz step. However, alternate schemes to improve intercept probability may be accomplished by adding detectors at the outputs of either the first or second converters so that "smart" or adaptable switcing may take place. Another modification to reduce intercep time is to use a 2 GHz wide IFM receiver at the output of the fir multiplexer, thus providing a 2000 MHz instantaneous bandwidth. The IFM receiver, althou having somewhat less sensitivity can be used to detect the stron non-overlapping signals as well a adaptively direct the second mu tiplexing. The narrowband filte banks will look for pulse overlapped signals or closely spaced CW signals.

The advances in GaAs FET technology and the application of microwave integrated circuit packaging techniques are also ideally suited for use in the fror end and down converter section of the channelized receiver.

Of significant importance in the third or IF filter bank of th channelizer is to prevent/eliminate adjacent filter slots from b ing activated by the occurrence of a high power, narrow pulse. This is of particular concern when a pulse overlap condition exists and the two pulses are rel tively close in frequency but greatly different in power.

This problem of selecting the appropriate filter output is one of the most significant problem in designing a channelized recei er. For example, for a 0.1 μsec pulse the sideband energy 60 dI down from the peak extends ±3000 MHz out from the carrie for a zero rise time signal. For a triangular shaped pulse this reduces to about ±150 MHz. For pulse whose rise and fall time is approximately a cos^2 curve which is more typical, the side band energy still extends out to ±70 MHz, 60 dB down from the peak of the signal. Thus, two 0.1 μsec overlapping signals that are 60 dB apart in amplitude cannot be resolved unless they are separated in frequency by more than 70 MHz. This is independent of the selectivity of the filter.

The basic frequency accuracy and resolution desired is often incompatible with the final filter bandwidth necessary to permit measurements on the narrowest expected pulse. This problem can be resolved if the frequency response of the filter skirts are carefully controlled. An instantaneous fine-frequency capability can be achieved by using adjacent filters to form a frequency discriminator. When the outputs of the filter cells are strobed, the signal amplitude of all three cells can be measured and used to determine the position of the signal within the center cell.

Channelized receivers have seldom been employed in operational systems since they require a large amount of hardware to achieve broad bandwidth coverage. The introduction of surface acoustic wave (SAW) devices, has greatly improved the ease of packaging and decrease the cost of such receivers and there is now renewed interest in channelized receivers. The use of SAW devices potentially allow a number of contiguous bandpass filters to be implemented on a single piezoelectric crystal – combining a small size, easily reproduced and low cost filter bank.

Figure 12 is a channelized IF/video processor module developed by Texas Instruments for the Air Force Avionics Laboratory. This unit uses SAW filters to provide bandwidth channelization to a slot width of 20 MHz while providing processing for 100 nsec pulses.

SYSTEM IMPLICATIONS

The interrelation between the receiving subsystem, the processing subsystem and the technique used for direction of arrival meas-

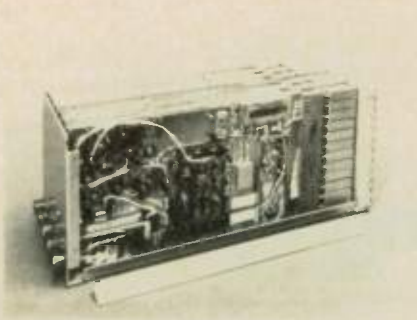

Fig. 12 Channelized IF video processor module having 20 MHz slot width SAW filters. (courtesy of Texas Instrument Inc.)

urement must be addressed in the design of the ESM system. Figure 13 illustrates these interrelationships. To minimize the time required to intercept a signal, a system wide open over the entire frequency range and having instantaneous 360° azimuthal field of view is desired. The system configured in this wide-wide open configuration may produce such a vast quantity of data due to the signal density that without other aids the processing system could be overwhelmed. The two dimensional (frequency/DOA) acousto-optic system presently in the R&D stage could ultimately provide the desired performance. Signal separation into discrete frequency/DOA cells permits isolation of even time coincident signals and would yield precision measurement of frequency and direction of arrival so that a processing system could be organized that can reliably handle a multiple mega-pulse per second signal environment.

The data rate is reduced by constraining the instantaneous field of view. Many existing narrowband systems provide instantaneous 360° azimuthal coverage with precise DOA measurement via amplitude or phase comparison techniques. Total frequency coverage is achieved by sequentially tuning the narrowband receiver across the frequency range.

Alternatively, a very broadband directional search reduces data rate, also providing an overall increase in system sensitivity due to antenna gain, increasing the ESM system detection range. The directional search can be achieved by mechanically or electronically scanned antennas. The organization of a processor for this type of system is shown in Figure 14. The objective is to detect significant changes in previously seen emitters and then to immediately identify new emitters from the dense redundant, mostly uninteresting signal environment.

As an example, consider that the directional antenna rotates continuously. The signal environment that has previously been

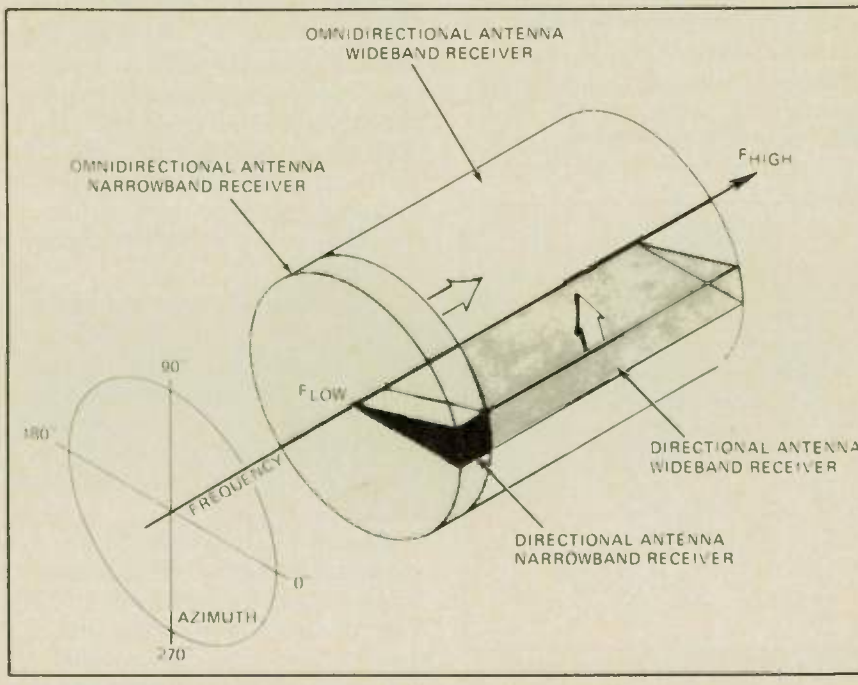

Fig. 13 Frequency/azimuth coverage trade-offs.

(continued on page 60)

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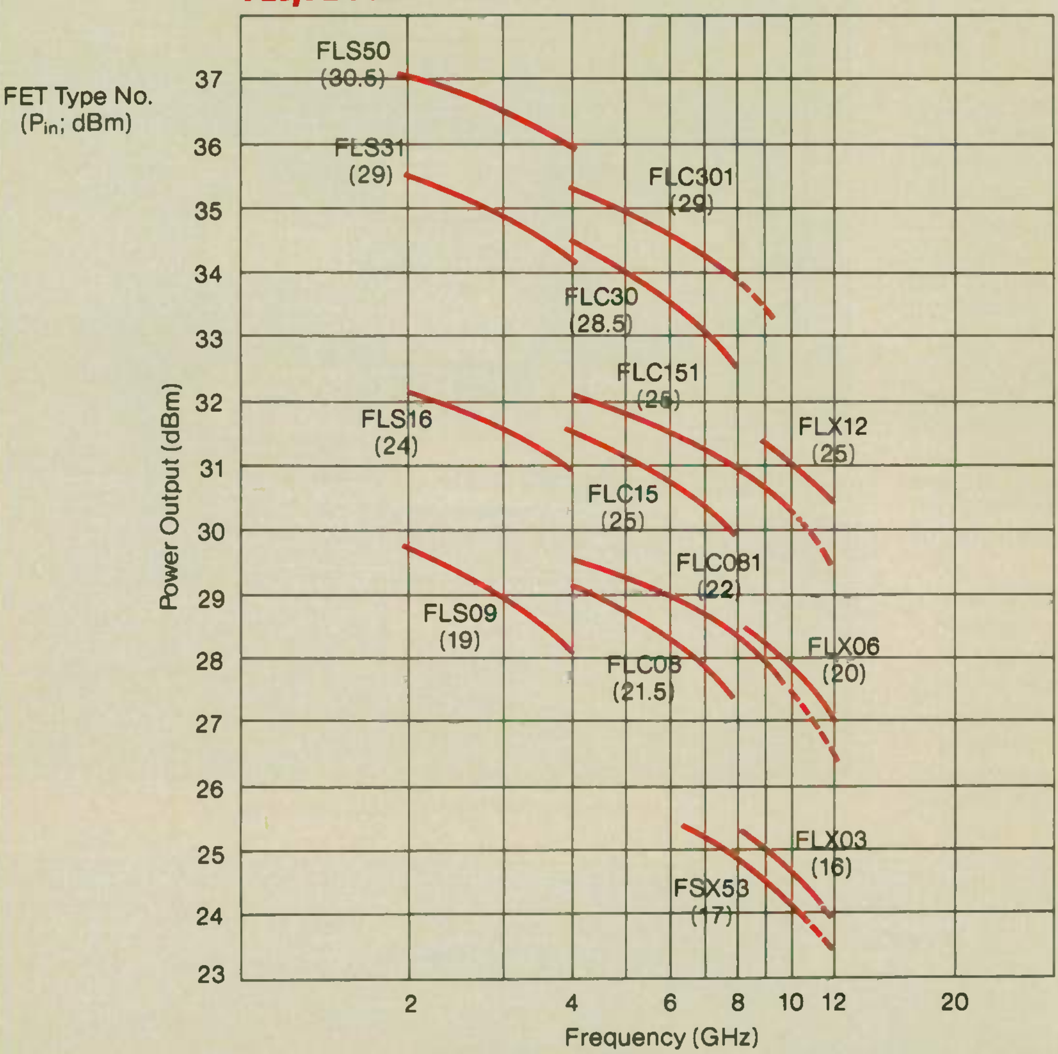

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Frequency Range GHz	3.7—4.2	4.4—5.0	5.9—6.4	6.4—7.2	7.1—7.7	7.9—8.4

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seen at each discrete azimuth position of the antenna is contained within the pulse sorter/content addressable memory (CAM) and signal monitor unit (SMU). As the antenna rotates, the CAM and SMU are updated for each discrete position of the antenna. The CAM could have multiple levels; the primary sort is by frequency. If a signal had previously been detected at the designated frequency, a secondary CAM could check for pulse width agreement. If the intercept's frequency and pulse width match the contents of the CAM, the SMU analyzes the pulse train pattern. If this pattern matches a previously stored pattern, the signal has been seen before, and no further analysis other than updating its direction of arrival is required. If there is mismatch at any stage, the new signal is routed to the Pulse Data Analyzer microprocessor. This approach only requires sufficient CAM and SMU storage capacity to handle the signal environment in the most dense directional sector and not that seen over the entire 360° field of view. This approach can also function with nondirectional DF systems e.g., interferometer, that provide a DOA tag for each received pulse.

The omnidirectional antenna of **Figure 14** provides the function of sidelobe discrimination, i.e., prevents the system from detecting signals on the side and backlobes of the directional antenna. This is a vital system function since it: (1) ensures correct measurement of the emitter di-

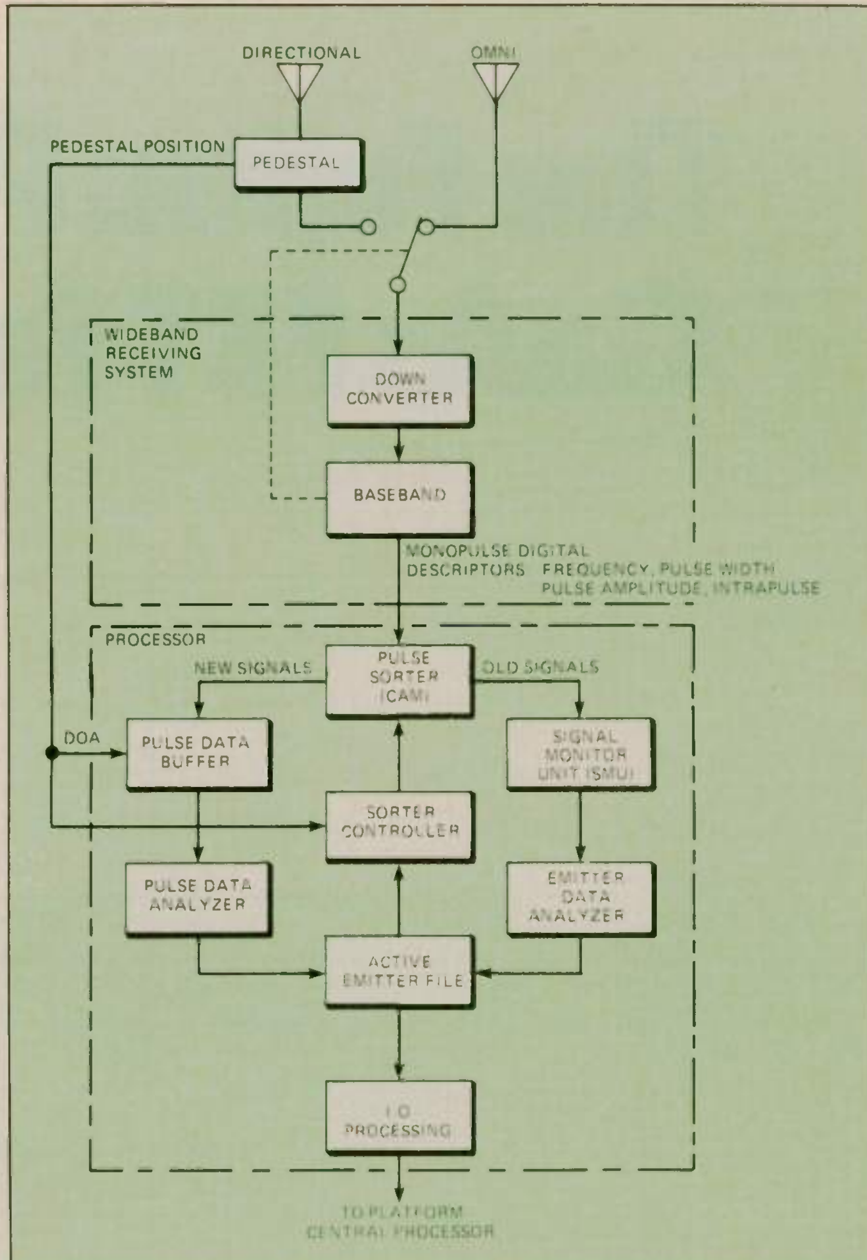

Fig. 14 Wideband ESM system/processor block diagram.

TABLE 3

COMPARISON OF WIDEBAND RECEIVERS

	Sensitivity	Dynamic Range	Frequency Accuracy	Simultaneous Signal Capability	TOA Availability	Instantaneous Bandwidth	Processing Complexity	Hardware Complexity	Cost	Stage of Development
CVR	Low	Low	Poor	Poor	Excellent	Excellent	Low	Low	Low	Mature
A/O	High	Low	Good	Excellent	Poor	Fair	Med	Med	Med	Early
Channelized	High	Med	Good	Good	Excellent	Fair	High	High	High	Emerging
IFM	Med	High	Excellent	Poor	Excellent	Excellent	Med	Med	Med	Mature
μscan	High	Med	Good	Good	Poor	Good	High	High	High	Emerging

ection of arrival and (2) reduces the data rate into the processing system. Sidelobe suppression is accomplished by requiring the signal from the directional antenna to exceed the signal from the omnidirectional antenna by a fixed amount; otherwise, the received signal is rejected.

While the system makes use of the direction of arrival to perform the functions mentioned above, it could also operate in an omnidirectional mode. In this mode, the system sensitivity is sacrificed but an instantaneous 360° coverage is provided. This instantaneous coverage coupled with the wide open frequency coverage ensures detection of extremely short transmissions. This configuration requires no modification in hardware, only software. In this mode the system loses some capability in handling very dense environments, in that the CAM and SMU capacity must accommodate the entire azimuthal environment. For many scenarios the capacity resident within these memories should be adequate to enable the system to operate in this wide, wide, open mode.

SUMMARY

In conclusion, Table 3 presents a comparison of the advantages and disadvantages of the various wideband receiving systems discussed. The comparisons have been made for comparable complexity of use and for the present state of receiver development. Note that on a subjective basis there is a strong correlation between hardware complexity, the difficulty associated in processing the receiver output, and relative cost.

Allan R. Baron has been engaged in the development of electronic warfare systems for 22 years. He is presently Director of Electronic Warfare Systems Engineering at the Amecom Division of Litton Systems, Inc., where he has been employed for the past 12 years. While at Amecom he was responsible for programs that demonstrated the feasibility of the Binary Beam Precision Direction of Arrival and Instantaneous Frequency Measurement Interferometer Techniques. He played a key role in the acquisition and development of a number of major electronic warfare systems including the Passive Detection System (AN/ALR-59) for the Navy's E-2C Airborne Early Warning Aircraft, and the AN/ALQ-125 Tactical Electronic Reconnaissance Sensor (TEREC) for the Air Force's RF-4C fighter aircraft. He has also directed Amecom's Independent Research & Development (IRAD) program and helped transition its products into DoD-sponsored exploratory development projects.

Prior to his employment at Amecom, Mr. Baron worked at AIL for 10 years where he was engaged in the development of EW systems and techniques. Mr. Baron received a B.E.E. from the City College of New York in 1956 and an M.S.E.E. from the Polytechnic Institute of Brooklyn in 1962. His graduate studies emphasized communication and control systems.

Charles B. Hofmann has been engaged in the development of electronic warfare systems for 17 years. He is presently Vice President of Engineering at the Amecom Division of Litton Systems, Inc., where he has been employed for the past 11 years. He is responsible for all engineering technical and administrative activities for Amecom's entire product line that includes electronic warfare, communications, telecommunications and radio navigation systems. Previously, at Amecom, Mr. Hofmann was Program Manager for the development of the Passive Detection System (AN/ALR-59) for the Navy's E-2C Aircraft. He was Engineering Program Director for the development of the Terminal Control Voice Switching System presently operational at the Dallas/Fort Worth Airport. Mr. Hofmann was also Director of the Engineering Product Development Directorate, responsible for the design of all electronic subsystems.

Prior to his employment at Amecom, Mr. Hofmann worked at the Maxson Electronics Company developing EW trainers, and at AIL where he developed amplitude and phase tracked logarithmic IF amplifiers used in electronic warfare systems. Mr. Hofmann received a B.S.E.E. from Lehigh University in 1961 and an M.B.A. from the Columbia University Graduate School of Business in 1963.

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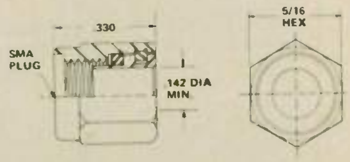
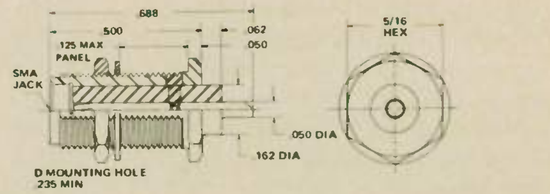


PART NO. 300-1
Straight Cable Plug - 141 Semi Rigid Cable



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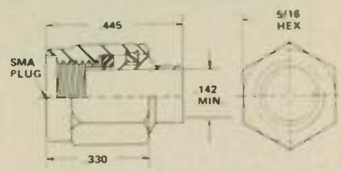
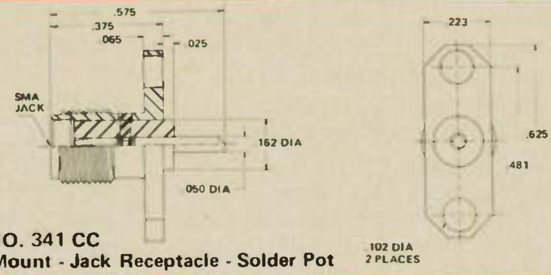


PART NO. 300
Straight Cable Plug - 141 Semi Rigid Cable



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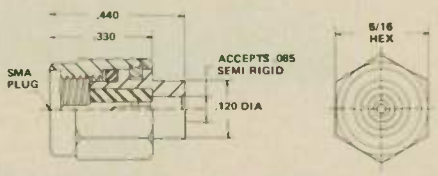

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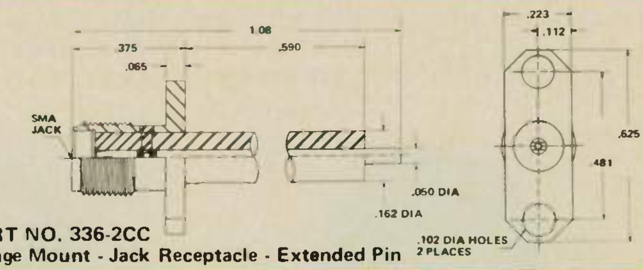

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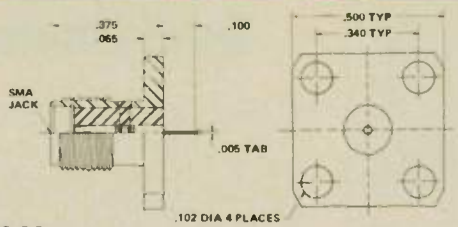
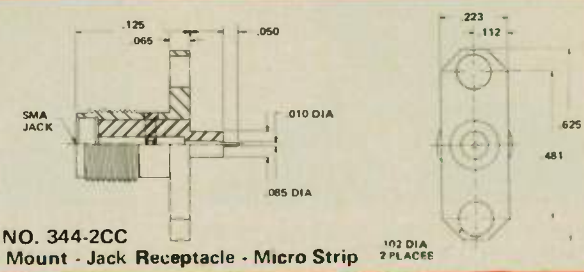
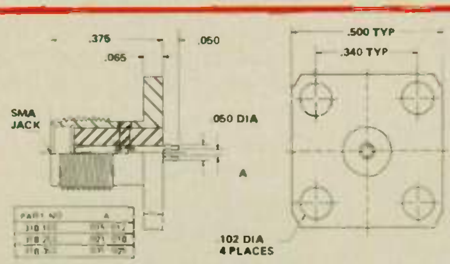


PART NO. 340 CC
Flange Mount - Jack Receptacle - Tab Contact



PART NO. 344-2CC
Flange Mount - Jack Receptacle - Micro Strip



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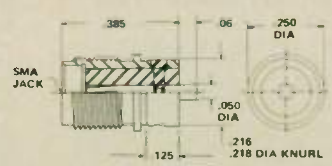

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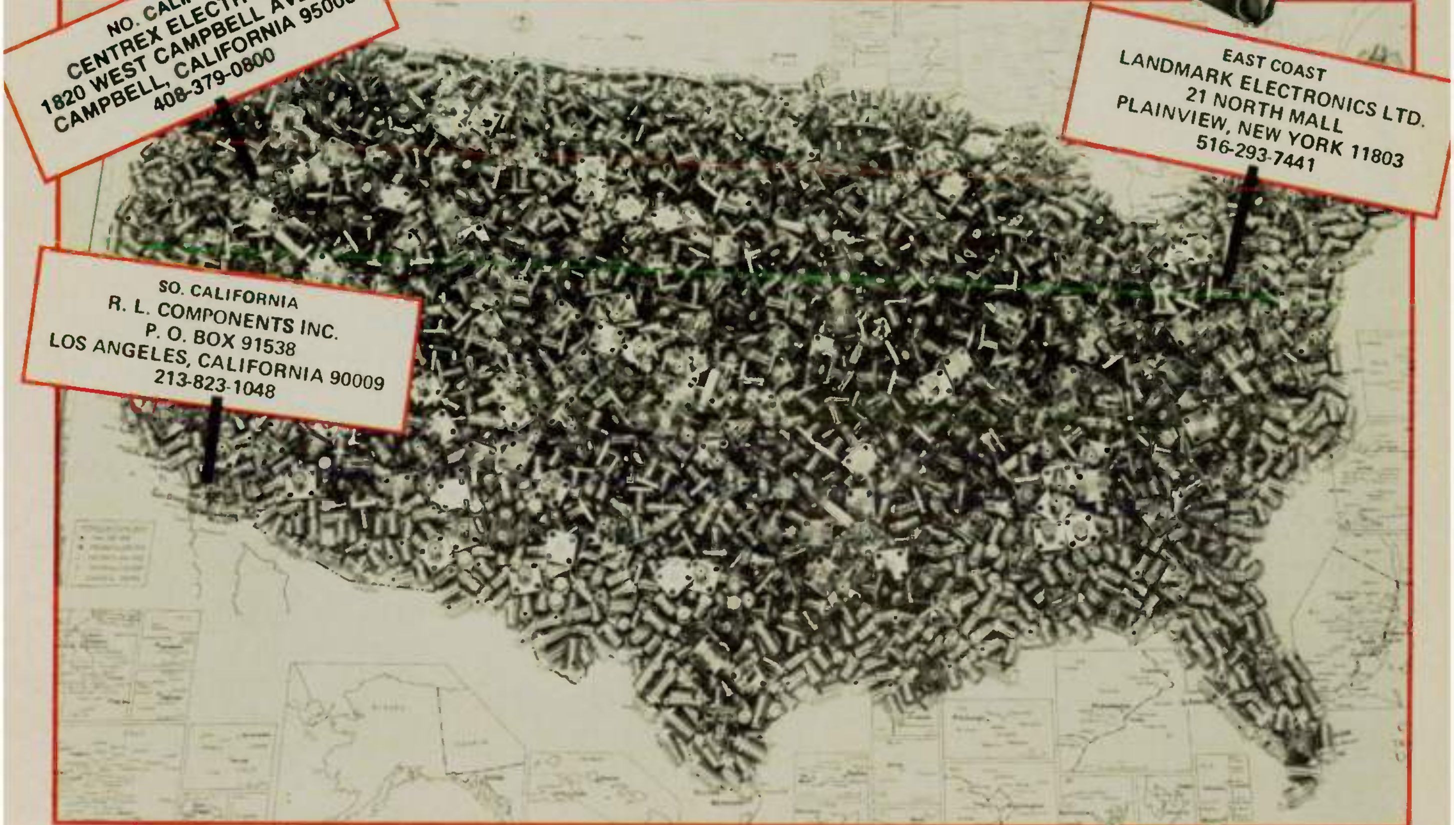

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FINISH	3.3.1.	Contacts .0001 min. gold Per MIL-G-45204
INSULATION RESISTANCE	3.11	5000 MEGOHMS min.
DWV	3.18	1000 VRMS
CONTACT RESISTANCE Milivolt Drop	3.17	Initial after Environment 3.0 4.0
CONNECTOR DURABILITY	3.16	Insertion withdrawal 500 cycles min. at 12 cyc/min. max.
VIBRATION	3.19	MIL STD. 202 Method 204 test cond. D
SHOCK	3.20	MIL. STD. 202 Method 213 test cond. 1
TEMP CYCLING	3.21	MIL. STD. 202 Method 102 test cond C
CORROSION	3.14	MIL. STD. 202 Method 101 test cond B
MOISTURE RESISTANCE	3.23	MIL-STD. 202 Method 106 IR shall be 200 MEGOHMS

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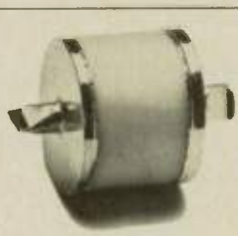
MAG19	35·0	11·0	11·0	0·5	0·001
MAG23	1·50	2·25	3·0	0·25	0·001

2.



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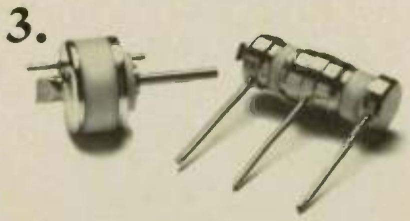

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Multimode Travelling-Wave Amplifier Tubes

ERIK BUCK, Major, USAF
Aeronautical Systems Division/XRE
Wright-Patterson AFB, OH

If one wants a broadband power amplifier at frequencies of a gigahertz or higher, one will probably choose a traveling wave tube (TWT), TWT's find applications in communications, especially in satellites, and in various military systems, particularly broadband jammers. Bandwidths of an octave or more are common, and efficiency can, but usually doesn't, approach 55%. Even at K band, power outputs in the hundreds of watts are achieved. Even the most optimistic proponents of solid state devices admit that when it comes to brute power. TWT's are the way to go. This article is concerned with recent developments which make the TWT even more attractive as a microwave amplifier.

THE BASIC WORKINGS

The traveling-wave tube, depicted very schematically in Figure 1, is a nearly axisymmetric tubular device with an electron beam running down its length. At one end is an electron gun, a cathode and focusing optics, and at the other end is a collector, one or more electrodes which collect the electrons. The beam is focused magnetically, most commonly with periodic permanent magnets (PPM) which form a series of magnetic lenses and keep the electrons, which tend to spread apart because of space charge, from "intercepting" the RF circuit.

The RF circuit is a transmission line, from the input coupler to the output coupler, which is characterized by a phase velocity, relative to the axis of the tube, which is a fraction of the speed of light. In broadband tubes, the common RF circuit is a helix. Some narrowband tubes use coupled cavities. The electron beam, under the influence of the RF input, tends to form bunches of charge. When the velocity of the beam is very slightly higher than the phase velocity along the circuit, the bunches of charge induce a wave which travels along the circuit and is amplified at th expense of the kinetic energy of the electron beam. Hence the name, traveling-wave amplifier tube, which is descriptive but does not lend itself to an acronym. The classic text on the sub ject is Pierce, *Traveling-Wave Tubes*.[1]

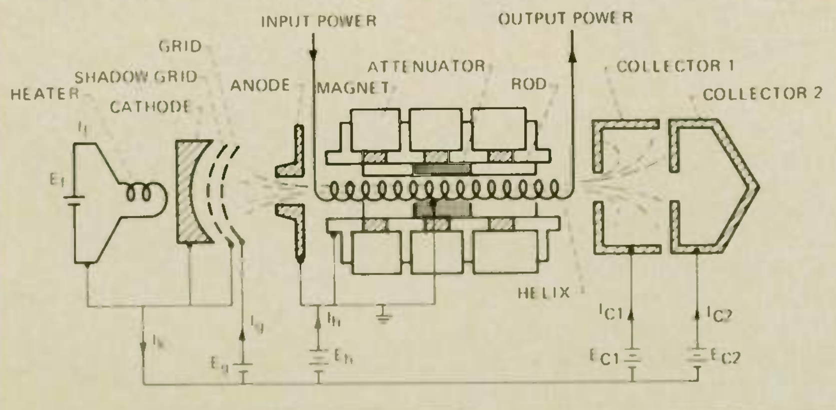

Fig. 1 Schematic of TWT and power supplies.

EFFICIENCY

The overall efficiency of a TWT, as shown in Table I, is the ratio of the useful power output into a load to the dc input powe from the power supply, excluding, for convenience, the small amount of cathode heater power Another useful term is electroni efficiency, the ratio of power in duced in the circuit to the powe of the electron beam, I_o, V_o. Ty ically, electronic efficiency may be in the range of 10 to 20%; if you take more power out of the beam, it will not stay in phase with the RF wave. Circuit efficiency has to do with the efficiency of the circuit as a transmission line, typically 80 to 90% because of resistive losses and d electric losses in the insulators which support, and also cool, th helix. At this point you will appreciate that if electronic efficiency is 20% and circuit efficie cy is 90%, only 18% of the bear

TABLE 1. – DUAL MODE TWT/MDC

Performance at and below saturation for a fixed set of MDC operating conditions (9.2 GHz)

(a) FIVE-STAGE COLLECTOR

Mode	Pulse-up ratio P_{RF} (high mode) / P_{RF} (low mode)	Overall efficiency-no MDC, %	Overall efficiency with MDC, %	Collector efficiency, %
High	1.0 (saturation)	17.2	42.6	76.6
Low	2.0	11.1	39.8	83.6
	3.0	7.4	35.4	87.1
	4.0	5.5	32.4	89.2
	5.0	4.4	29.2	90.1
	6.0	3.7	27.2	90.9
	7.0	3.1	24.9	91.4
	8.0	2.8	23.1	91.6
	9.0	2.5	21.7	91.9
	10.0	2.2	20.6	92.3
	dc beam	——	——	95.2

(b) THREE-STAGE COLLECTOR

Mode	Pulse-up ratio P_{RF} (high mode) / P_{RF} (low mode)	Overall efficiency-no MDC, %	Overall efficiency with MDC, %	Collector efficiency, %
High	1.0 (saturation)	17.0	40.9	74.8
Low	2.0	11.1	38.0	82.0
	3.0	7.3	31.7	84.6
	4.0	5.5	27.4	85.9
	5.0	4.4	23.8	86.4
	6.0	3.7	21.2	86.8
	7.0	3.2	19.0	87.2
	8.0	2.8	17.4	87.6
	9.0	2.5	15.9	87.6
	10.0	2.2	14.4	87.5
	dc beam	——	——	92.5

power will get out the output window. How can overall efficiency reach 50%?

Enter the concept of collector efficiency. If, in actuality, the electrodes of the collector were at ground potential, all of the kinetic energy left in the beam would be simply dissipated as heat when the beam impacted the collector. Then overall efficiency would be limited by the electronic efficiency. However, the collector is typically depressed, operated at a potential closer to cathode potential than the potential of the tube body (ground). Hence, as the electrons enter the collector, they give up their kinetic energy traversing the potential gradient and strike the collector electrodes with less wasted power. Typical two-stage depressed collectors, like the one depicted in **Figure 1**, can recover about half of the kinetic energy of the spent beam. The rest generates heat. One can conceive that if a collector and circuit were 100% efficient, then the tube could be 100% efficient. All of the dc power actually consumed would result in useful RF power (**Figure 2**). Poor circuit efficiency, however, lowers overall efficiency, as shown in **Figure 3**[2].

CHOOSING AN OPERATING POINT

The performance of a given TWT is highly dependent upon the qualities of the beam and the manner of its focusing. A useful concept is perveance.

$$\text{perveance} = \frac{I_o}{V_o^{3/2}} \text{ pervs.}$$

In a high power tube with a beam current, I_o, of 1 ampere and a beam voltage, V_o, of 10,000 volts, the perveance is 10^{-6}, or 1 microperv. The difficulty of designing the gun and the magnetic focusing increases with perveance, and values much above 2 microperv require fairly heroic measures; PPM focusing gives way to heavy, power-consuming solenoids.

In a multimode tube, designed to operate at different power levels, the traditional approach has been to vary the perveance of the beam by varying the beam current, typically with a grid (or several) in the gun. With fixed focusing fields, from permanent magnets, the lower perveance beam, which has less space charge tending to expand the beam, will shrink in diameter. The interaction with the RF circuit is less strong, so the beam power, the electronic efficiency (the portion of the beam power converted to RF), and the gain of the tube all decrease. For the tube which was the basis of **Table 1**, a beam current (at 9.93 kV) of 480 ma (high mode) yielded 830 watts of RF output at saturation. When the beam current was reduced to 380 ma, the saturated output was reduced to 500 W. The tube would not operate usefully with much less beam current.

The gain of a TWT is constant over a range of small signal values. As the RF input increases, the output becomes less linear until the tube saturates and further increases in RF input result in decreased output. The common tubes available today are usually

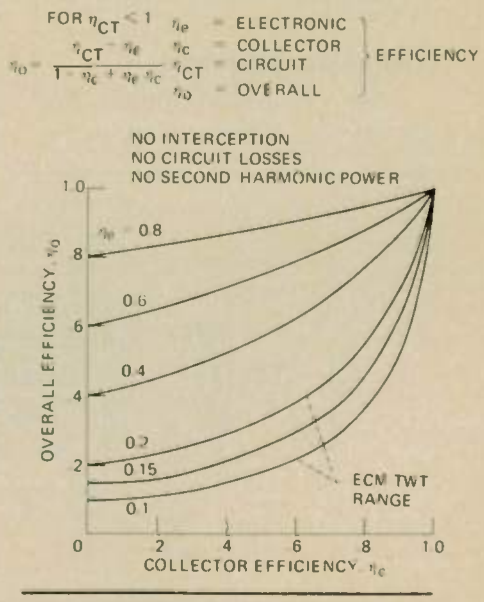

Fig. 2 Overall efficiency vs collector efficiency.

operated close to saturation, because that is where the electronic efficiency is highest, and overall efficiency is dependent upon electronic efficiency.

THE LINEAR TUBE

Suppose, however, that the overall efficiency were relatively independent of electronic efficiency. It would then be practical to operate tubes in the linear region (low electronic efficiency) controlling the RF output by varying the RF input, rather than by changing the gain characteristics of the tube. (For some reason, system designers often resist this concept, preferring to saturate the RF driver, so that RF input is constant. I assert that there should be less discomfort controlling milliwatts of RF input than there is controlling kilowatts of dc with grids). If a large range (10 dB) of output is desired, both methods can be used. That is what was done in our experimental tube (**Table 1**). At the lowest P_{RF} (low mode), only 83 watts, the beam current was still 380 ma. The low electronic efficiency was compensated for by an efficient collector. The linear, "constant efficiency," tube, to be attractive, *must* have two features: (1) a collector efficiency of 90% or better, and (2) low losses, with very little interception of the beam or backstreaming electrons.[3]

THE MULTISTAGE DEPRESSED COLLECTOR

The difficulty with conventional depressed collectors (as in **Figure 1**) is that a space charge builds up which limits the depression, and thus the efficiency.[4] Several years ago, Dr. Henry Kosmahl, at NASA Lewis Research Center, took a fresh look at the depressed collector and the electron optics, which resulted in patents for improved multistage depressed collectors (MDC) and for a refocusing section, inserted between the end of the RF circuit and the beginning of the collector, in which the electron beam is expanded and collimated which, in conjunction with a

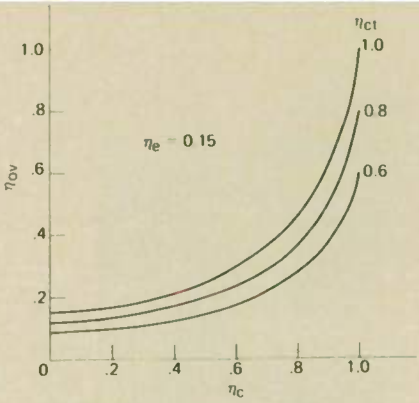

Fig. 3 Effect of circuit losses on the overall tube efficiency for electronic efficiency $\eta_e = 0.15$.

good collector, makes possible near-total recovery of the energy in the spent beam. These devices were spectacularly successful in their first application, the Communications Technology Satellite (CTS). The CTS tube, a 12 GHz, coupled-cavity, 200 watt TWT was operated in space for nearly 4 years (satellite was turned off in October 1979). It was both the most powerful and the most efficient (~50%) communications TWT ever flown.

The design of an efficient multistage depressed collector requires some sophistication and is best done by digital simulations.[5] **Figures 4** and **5** depict calculated electron trajectories in two different collectors. The collectors are symmetric about the axis; only half of a section is shown. The lines which are horizontal at the axis are equipotential lines, which illustrate the electrostatic lenses formed in the apertures. The arrowheads denote probable secondary electrons. The keys to a successful MDC are these: (1) collect the spent beam electrons at as low a potential as possible for maximum energy recovery, and (2) avoid backstreaming, which may be either primary electrons which are reflected back into the tube by the potential gradient or secondary electrons which are accelerated into the tube. Thus, in the successful collector there are divergent lens effects to disperse the beam toward the collector walls, and the electrons are collected on the side of the plates away from the tube, in suppressing fields, so as to avoid secondaries finding the way into the tube.

The ill effects of secondaries can be further reduced by using collector surfaces with a low yield. Oxidized copper will usually yield more than one secondary electron for each primary electron collected. For laboratory use only, we at NASA coat our collectors with ordinary soot, which improves collector efficiency by 2 - 4%, typically. That can be very significant when one is striving for 90% efficiency. Some tube makers use a sputtered coating of titanium carbide which seems to help and is obviously more durable than soot. At Lewis, we are now fabricating collector plates from pyrolytic graphite, which is light, strong, and highly conductive in the plane of the plate. It should be compatible with bake-out in vacuum, and we have demonstrated by measurement of roughened samples, that the secondary yield is low, about 0.3. Practical assembly techniques (eg. brazing) have not yet been mastered at Lewis, but Siemens has done it.

It is commonly said that too much collector depression causes increased interception. It would be more correct to say that increased depression in a poorly engineered collector will cause excessive backstreaming. Electrons which are going the wrong way down the tube will, of course, have different and deleterious effects as compared with the interception of the edges of the beam by the circuit. However, they both show up as "body current" and cannot be separately measured in a typical production tube. The point is this: if the body current increases significantly when the collector is depressed, go back to the drawing board.

Another misconception is that the MDC is necessarily big. The CTS collector is intentionally large, because it is radiation cooled. The collector in **Figure** however, is very close to the size of the collector it was designed to replace; it is roughly an inch in inside diameter. Note also th

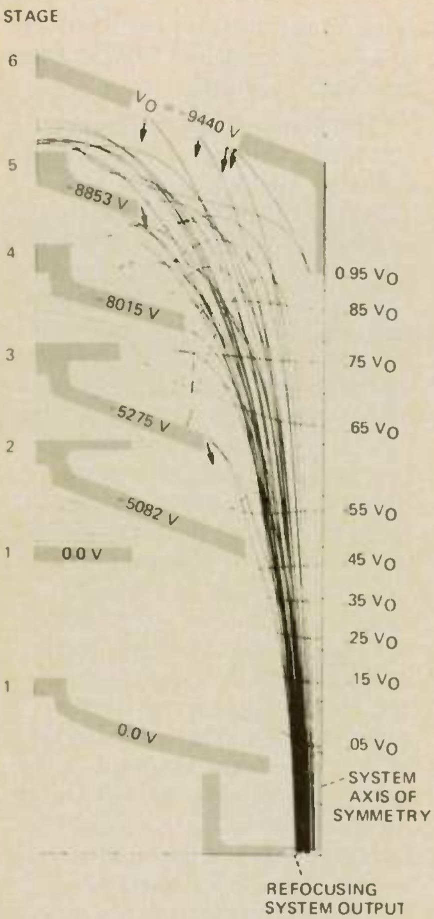

Fig. 4 Electron trajectories in experimental collector with five depressed stages; TWT operating at 3 decibels below saturation. Secondary electron emission yield, δ, 1/2

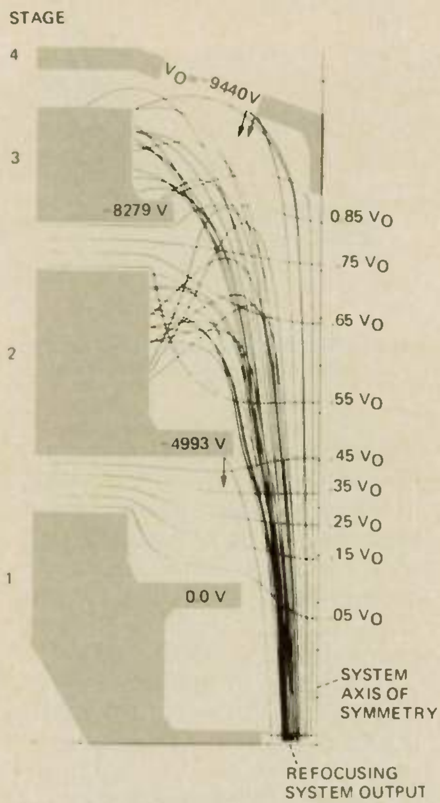

Fig. 5 Electron trajectories in scaled-down collector with three depressed stages; TWT operating at saturation. Secondary electron emission yield, δ, 1/2.

most of the impacts are on the walls, which facilitates cooling.

How many stages of depression does one need? The answer depends upon the perveance of the beam and the electronic efficiency. When either or both are reduced, the variance in the velocities of the electrons is reduced (**Figure 6**).[5] Typically, part of the collector is at ground, where it attaches to the body, and the last stage is at cathode potential. One plate is approximately at the potential which corresponds to the energy of the slowest electrons, and one or more plates are at intermediate potentials, for a total of three to five depressed stages. If the range of velocities is great (the high perveance or high electronic efficiency case), then it is more difficult to sort out the electrons by velocity and to recover their energy. At low RF drive levels (low electronic efficiency), the beam is more homogeneous, so the electrons are more efficiently collected. The fact that the MDC automatically becomes more efficient as the electronic efficiency is reduced leads to the concept of the "constant efficiency" tube.

PROOF OF CONCEPT

In January and February of 1979, Ramins and Fox (at NASA Lewis, as part of a joint program with the Air Force) demonstrated the "constant efficiency" concept with a modified production tube, the Teledyne MEC MTZ 7000 fitted with a NASA-fabricated[6] collector. The data in **Table 1** is but a sample, not the best that can be done, but an example of a tube which operates over a 10 dB output range varying only the focus electrode voltage (dual mode gun) and the RF input power. Optimized for saturated performance, the overall efficiency exceeded 50%. Optimized for collection of the dc beam (no RF input), the collector efficiency exceeded 97%.

To the system designer, the development of the high-efficiency MDC, which makes possible the multimode "constant efficiency" tube, will have a direct impact on what he can do. If one is in the electronic countermeasures business, one can double the efficiency of the current generation of tubes, reducing power consumption and cooling requirements while increasing output power. Perhaps one can replace separate pulse and CW tubes with one multimode tube.

For satellite communications capability, the high efficiency technology will be incorporated in a new 20 GHz tube for NASA Lewis Research Center.[7] The new tube will be extremely linear, so as to be able to handle three simultaneous QPSK modulated carriers without undue intermodulation distortion. The saturated output will be approximately 75 W. However, it is intended to operate in the linear region, not at saturation. Normally it will operate 14 dB down from saturation, at an overall efficiency of about 20%. In the event of attenuation by rain, the output signal can be boosted 10 dB, for the duration of the thunderstorm only, to a point 4 dB down from saturation but greater than 40% efficiency, thus conserving prime power. With 98+% recovery of beam power in the no-drive condition, it will be feasible to leave the tube on all the time, thus improving reliability.

Obviously, a tube with a good MDC will cost more, both to de-

Fig. 6 Computed energy distributions for tubes of same efficiency as function of perveance.

sign and to build, than a tube with lower parts count. However, the increment in cost is small compared with the advantages of higher efficiency. There will be savings in the weight of power supplies and collector cooling, though the power supply may need more taps. (CTS had eight). However, two or three-stage collectors are becoming common; four or five-stage collectors are not a big step. We are looking at future power supplies using capacitor-diode voltage multipliers (CDVM), modern versions of the 1932 Cockcroft-Walton device. With the CDVM, multiple taps are "free." High voltage stand-off is no problem; as the number of plates increases, the potential between them decreases. Regulation is not critical; a voltage variation of a few percent has little effect on over-all collector efficiency. The power supply should, however, have enough energy storage to handle the rapid changes in collector plate current which can occur when the tube is pulsed. Here, again, the CDVM looks like an economical solution.

It may be noteworthy that multi-stage collector efficiency is not sensitive to the RF frequency of the tube, as such. MDC efficiency varies with electronic efficiency, which tends to vary with frequency.

CONCLUSIONS

Significant advances have been made in traveling-wave tube engineering, and digital models allow new concepts to be converted into hardware more quickly, with fewer cut-and-try iterations than used to be the case. Thus, tubes which were once marketed when they were "good enough" can be much improved. The application of efficient refocusing and multistage collectors, made possible by computer simulations, will greatly increase tube efficiency and, thereby, make practical new multimode tubes which operate, at least partly, in a linear fashion. These new tubes will lead to major changes in system architecture and will generate needs for advances in the control of RF input.

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Major Erik Buck heads the Electromagnetic Combat Group (XRE) of the Deputate for Development Planning at Aeronautical Systems Division, Wright-Patterson Air Force Base. Before this assignment, he was a "Laboratory Associate" and Project Engineer in the Communications Technology Branch at NASA Lewis Research Center, Cleveland, Ohio. For the past three years he has been working on joint USAF-NASA projects to improve the efficiency of traveling-wave tubes. His Bachelor's degree is in Physics, from the State University of New York at Albany. He joined the Air Force to become a weatherman, spending a year at Penn State and five years as a Forecaster and Staff Meteorologist. The Air Force sent him for an M.S. in Engineering Physics (laser optics) at the Air Force Institute of Technology (AFIT), Dayton, Ohio. That led to work in the development of reconnaissance/strike systems and "optical warfare." Optical countermeasures led to electronic countermeasures, which led to his being sent to work with NASA to learn about TWT's.

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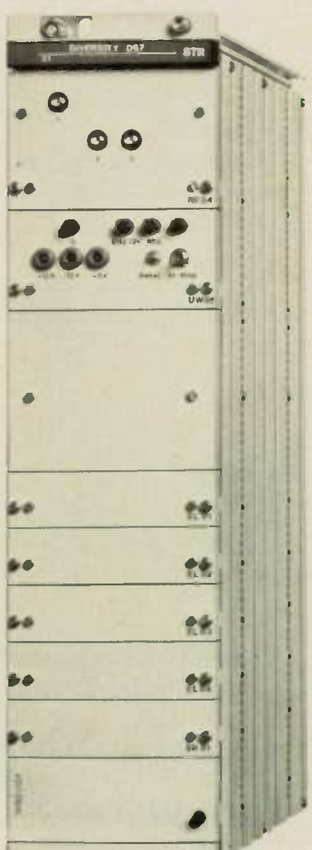

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27 MHz Ridged Waveguide Applicators for Localized Hyperthermia Treatment of Deep-Seated Malignant Tumors

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Microwave Technology Center
RCA Laboratories, Princeton, NJ

J. MENDECKI, E. FRIEDENTHAL and C. BOTSTEIN
Department of Radiotherapy
Montefiore Hospital and Medical Center, Bronx, NY

Ridged waveguide applicators suitable for localized hyperthermia treatment of deep-seated malignant tumors are described. When driven with several hundred watts of 27 MHz RF power, these applicators can raise the temperature of deep-seated tumors to the hyperthermic range (42.5-43.5°C), i.e., the temperature range which appears to be optimum for the treatment of cancer. In initial clinical trials encouraging results were obtained with no discernible side effects.

INTRODUCTION

Localized hyperthermia has been shown to be effective in the treatment of a variety of malignant tumors, either as a stand-alone therapy, or more often in conjunction with radiation therapy.[1,2] A typical treatment with localized hyperthermia consists of raising the temperature of the tumor mass to about 42.5-43.5°C, taking care to minimize the heating of the surrounding healthy tissues. The tumor is maintained at the high temperature for approximately one-half to one hour at a time. Multiple treatments are usually given. Ionizing radiation when added is usually administered in reduced dosages either during the hyperthermia treatment, or immediately before or after the treatment.

One of the most useful methods of producing localized hyperthermia in tumors is dielectric heating with radio-frequency (RF) radiation.[1] Here, power from an RF generator is transmitted into the tissue volume to be heated by an antenna or applicator. The RF travels through the tissues of the body in the form of an exponentially decaying wave, giving up energy to the tissues (dielectric heating) as it traverses them.

The depth to which RF waves can penetrate into tissues and produce heating is primarily a function of the dielectric properties of the tissues and of the RF frequency.[3,4] In general, the lower the water content of the tissue the deeper a wave at a given frequency can penetrate into it. Thus, for example, RF waves can penetrate much deeper into fat (low water content) than into muscle (high water content). Also, at the RF frequencies of interest, the lower the RF frequency the deeper the depth of penetration into a tissue with a given water content. This is illustrated in **Figure 1**, which shows the results of calculating the heat produced by plane waves at the five lowest ISM frequencies* in a simple tissue model (infinite layer of fat 2 cm thick followed by an infinite layer of muscle infinitely thick).[3] The values of the complex dielectric constant use

* ISM frequencies are frequencies set aside by the Federal Communications Commission for *I*ndustrial, *S*cientific and *Medic*al applications.

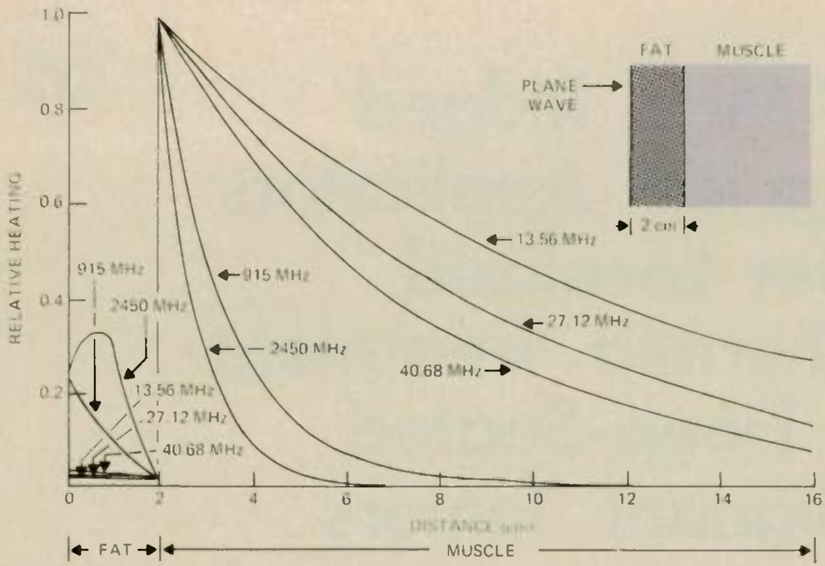

Fig. 1 Calculated relative heating in fat and muscle as a function of distance for five different ISM frequencies.

in the calculations were taken from Reference 9. Note the relatively small amount of RF power that is dissipated in the fat layer. In particular, note that virtually all of the energy at 13.56 MHz, 27.12 MHz and 40.68 MHz is transmitted through the fat into the muscle where it is dissipated.

In the January, 1980 issue of the *Microwave Journal* we described the theory and construction of hyperthermia applicators designed for operation at 915 and 2450 MHz.[5] Applicators at these frequencies are useful for noninvasive hyperthermia treatments of cutaneous and subcutaneous tumors, tumors located within or in the vicinity of natural body cavities, and tumors located in the breasts. On the other hand, tumors that are shielded from an accessible body surface by more than about 2 cm of tissue with high water content are usually difficult to treat with 915 or 2450 MHz radiation because of the high absorption of 915 and 2450 MHz radiation in tissues of this type (see Figure 1).

In the present paper, we describe applicators designed for operation with 27 MHz RF radiation. Unlike 915 or 2450 MHz radiation, 27 MHz radiation can penetrate deeply into tissues with high water content (see Figure 1), and can therefore be used to noninvasively heat deep-seated tumors that are not accessible with either 915 or 2450 MHz radiation. The major limitations of heating with 27 MHz are comparatively poor focusing (wavelengths in tissues are greater than 1 meter), and the large size of the present applicator designs. Also, accurate temperature measurements with thermocouples on thermistors are difficult in the presence of large 27 MHz fields.

The paper is divided into three parts: The first part describes the construction and principles of operation of the 27 MHz applicators. Despite their low frequency of operation, the design of these applicators is based on the microwave concept of dielectrically-loaded waveguide radiators. In the second part of the paper, the results of calculations of temperature profiles generated with plane waves at 27 MHz in simple living tissue models are presented. The third part covers the use of the applicators in animal experiments and as therapeutic tools in clinical trials involving various types of malignant tumors.

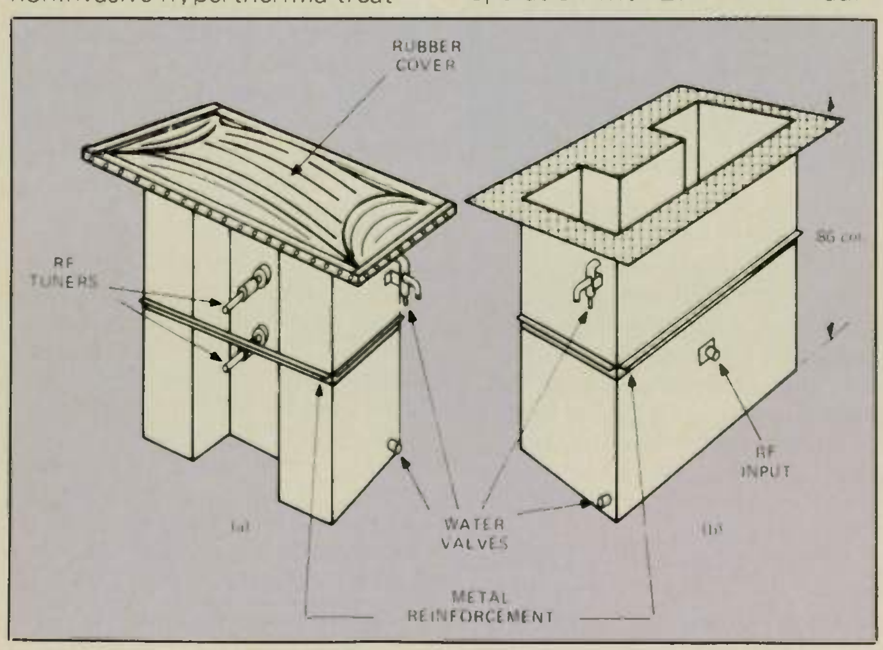

Fig. 2 Sketch of a 27 MHz ridged-waveguide applicator with (a) rubber cover in place and (b) with rubber cover removed.

DESCRIPTION OF 27 MHz APPLICATORS

Several considerations must be kept in mind when designing applicators for producing localized hyperthermia with RF radiation.

- The applicator must be able to handle the RF power required to raise the temperature of the tumor or tumors to be treated to the hyperthermic range. Power levels as high as several hundred watts CW are often required.
- The design of the applicator must minimize the amount of RF power being delivered to healthy tissues.
- The applicator design must be consistent with the physical comfort of a patient who has to undergo one or more treat-

(continued on page 74)

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ments at a time, each treatment lasting up to one hour, or in some instances even longer.

- Radiation into free space from the applicator must be kept to a minimum to protect the patient and the technicians administering the treatment from unnecessary exposure to RF.
- The applicator must be rugged, its cost must be moderate, and its size must be consistent with the space available in a typical hospital treatment room.

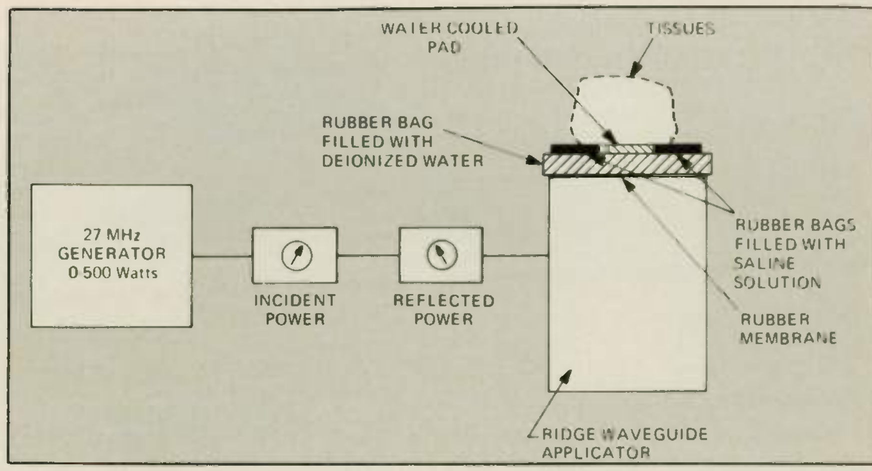

Fig. 3 Setup for hyperthermia treatment using 27 MHz ridged-waveguide applicator.

Based on the above design considerations, we have developed 27 MHz ridged waveguide applicators that have proven to be satisfactory for both animal studies and clinical work. **Figure 2** is a sketch of a typical applicator. The applicator is built from sheet metal in the shape of a shorted section of ridged waveguide. The open end of the guide is covered with a rubber membrane and the guide is filled with deionized water (ordinary water is too lossy). The 27 MHz power is introduced into the applicator via a coax-to-waveguide transition and is radiated from the applicator through the rubber sheet. Reflections of 27 MHz power back into the coaxial RF input port are minimized by means of three capacitive tuners.

The dimensions of the ridged waveguide are chosen to allow propagation of the TE_{10} mode at 27 MHz, typical cutoff frequencies for the TE_{10} mode being on the order of 20 MHz. Such low cutoff frequencies can be achieved with reasonable guide dimensions because: (1) the high dielectric constant of deionized water ($\epsilon \sim 81$) reduces the linear dimensions of the guide for a given cutoff frequency by a factor of approximately 9 over an air-filled guide and (2) ridged waveguides have lower cutoff frequencies than rectangular guides of the same outer dimensions. Suitable design equations for calculating the cutoff frequencies of ridged waveguides are given in References 6 and 7.

Figure 3 is a diagram showing a complete setup for inducing hyperthermia with a 27 MHz ridged-waveguide applicator. The applicator is driven by an RF generator whose power output can be varied between 0 and 500 watts. Incident and reflected powers are measured with conventional RF power meters. A rubber bag large enough to cover the entire rubber membrane is filled to a thickness of about 5 cm with deionized water and is placed on top of the applicator. The purpose of this bag is to spread the high electric fields present at the edge of the applicator; this prevents any excessive local heating of the patient. On top of the 5 cm thick water bag is a thin water-cooled pad for cooling the skin of the patient, and one or more small bags filled with saline solution—the saline solution is a good absorber of 27 MHz radiation and, therefore, it protects the tissues that one does not want to heat.

The heating patterns produced by the applicators were studied with the setup of **Figure 3**. Blocks of ground meat were placed on top of the 5 cm thick water bag and heated. As expected from standard ridged-waveguide theory, most of the heating took place in the volume of meat placed above and between the ridge and the wall opposite the ridges, i.e., above the shaded area of **Figure 4**. The hottest point was usually in the center of the shaded area (point A of **Figure 4**). This fall off in heating with distance from the applicator along a line perpendicular to the applicator and going through point A followed approximately the calculated curve for 27 MHz radiation shown in **Figure 1**.[†]

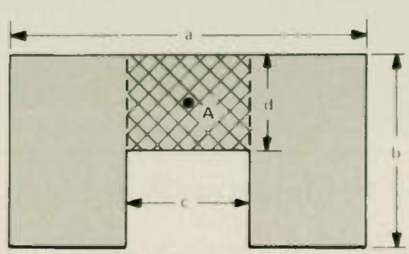

Fig. 4 Cross-section of ridged-waveguide applicator. Most of the RF energy in the guide is concentrated in the cross-hatched area.

While heating with one applicator is adequate in a number of clinical situations, it is often desirable to use two applicators in a cross-fire arrangement, since two applicators in such an arrangement can often produce deeper and more uniform noninvasive heating than a single applicator. This is illustrated in **Figure 5** which shows the results of calculating the heating produced by two RF waves that impinge on a simple tissue model from opposite directions. (The tissue model consists of an infinite layer of fat 2 cm thick followed by an infinite layer of muscle whose thickness is chosen so that the relative heating due to the two RF waves in the center of the muscle is approximately 80% of

† Figure 1 can be used to determine the maximum depth of penetration of waves emitted from symmetrical apertures (round, square, rectangular, etc.) into the tissue model, since the center ray emitted from any symmetrical aperture parallel to the surface of the model must, from symmetry considerations, always propagate in the same direction as the plane wave of Figure 1.

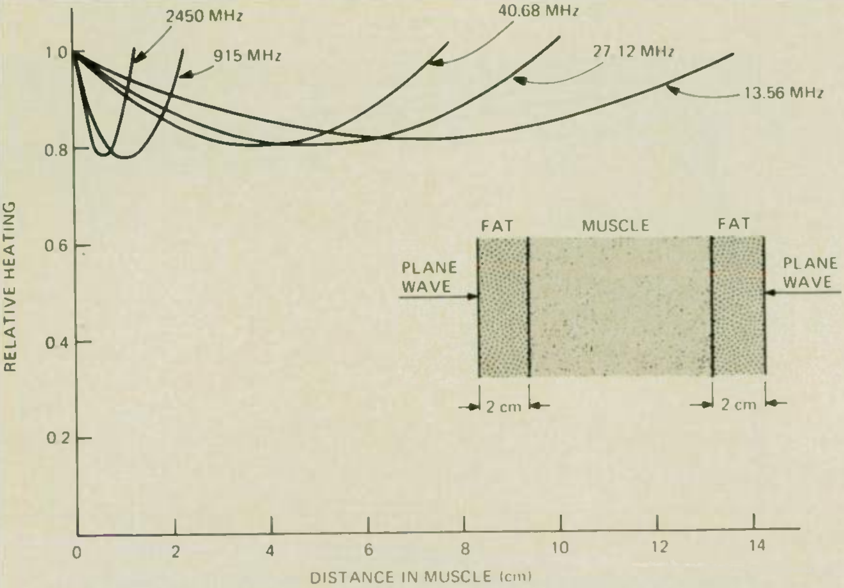

Fig. 5 Calculated relative heating in muscle as a function of distance for five different ISM frequencies due to two noncoherent plane waves of approximately the same frequency traveling in opposite directions.

the heating at the fat-muscle interface, followed again by an infinite layer of fat 2 cm thick.) The two plane waves incident on the tissue layers are assumed to be of approximately the same frequency but uncorrelated, and any reflections from the second tissue interface (muscle-fat) are neglected. **Figure 5** indicates that 10 cm of muscle can be heated nearly uniformly with two 27 MHz waves, a significant improvement over heating with a single 27 MHz wave (**Figure 1**).

We have built several ridged-waveguide applicators that can be used in "cross-fire" arrangements. The inside of these applicators is compartmentalized with solid plastic sheets, and each compartment is individually filled with deionized water. Compartmentalized applicators can be placed in a horizontal position without danger of rupturing the rubber membrane, since the plastic sheets protect the membrane from most of the water pressure. Two cross fire arrangements are possible with these applicators: two applicators in horizontal positions facing each other, or one applicator in a horizontal position and the other in a vertical position.

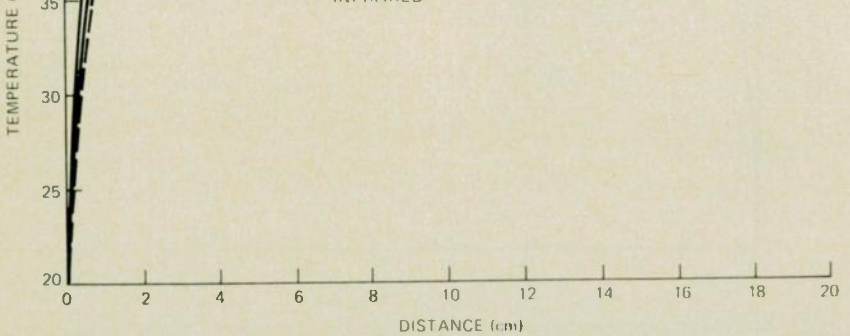

Fig. 6 Calculated temperature distributions in a semi-infinite slab of muscle when the muscle is heated with infrared radiation or with radiation at any one of five ISM frequencies.

CALCULATIONS OF TEMPERATURE DISTRIBUTIONS PRODUCED IN LIVING TISSUES

The temperature distributions produced by RF waves propagating through living tissues can be calculated by determining the complex propagation constants of the waves and then solving the heat transport equation. Such calculations are too difficult for the actual tissue geometries encountered when treating malignancies with RF hyperthermia; however, much information of practical value can be obtained by calculating temperature distributions in simplified tissue models.

Figures 6-10 are plots of calculated temperature distributions in semi-infinite slabs of muscle based on steady-state solutions to the one-dimensional heat transport equation in living tissues that are heated by RF waves. These solutions, which were given by Foster, Kritikos and Schwan,[8] assume a uniform plane wave incident on a homogeneous semi-infinite layer of tissue, and take into account blood flow (assumed to be temperature independent), tissue heat conductivity, and temperature difference between the surface of the tissue and the environment. In calculating these graphs, it was assumed that the arterial blood entering the tissues is at a temperature of 37°C. The values of the complex dielectric constants needed in the calculations were taken from Reference 9.

Figure 6 is a plot of the calculated temperature distributions in a semi-infinite slab of muscle when the muscle is heated with infrared radiation (assumed depth of penetration ∿0.001 cm), or with waves at one of five ISM frequencies. The following values of power densities transmitted into the muscle were used in the calculations: 0.05 W/cm² (infrared), 0.27 W/cm² (2450 MHz), 0.3 W/cm² (915 MHz), 0.56 W/cm² (40.68 MHz), 9.7 W/cm² (27.12 MHz), and 0.87 W/cm² (13.56 MHz). These values just raise the temperature in part of the muscle to the hyperthermic range (42.5-43.5°C). A value of $\lambda = 1.3\ cm^{-2}$ was assumed in the calculations for this figure and for Figures 7 and 8. (λ is the product of flow and heat capacity of blood divided by the coefficient of tissue heat conductance.)

Figure 6 shows that waves at the five ISM frequencies can produce temperatures in the hyperthermic range at distances of several centimeters into muscle without causing excessive temperature increases near the surface. On the other hand, most of the temperature increase produced by infrared radiation is confined to a few millimeters near the surface. As expected from the curves of Figure 1, the three lower ISM frequencies can produce hyperthermic temperatures at a significantly greater depth than 915 or 2450 MHz.

Figure 7 illustrates the dependence of temperature rise on the power density of 27 MHz radiation transmitted into muscle. Note that with an RF power density of 0.7 W/cm^2, temperature rises to the upper limit of the hyperthermic range are obtained; while with RF power densities about one-third smaller (0.5 W/cm^2) the maximum temperature is well below the hyperthermic range.

Figure 8 compares active cooling of the surface to 20°C with passive cooling in room temperature surroundings (22°C). Significantly deeper heating is obtained with active cooling, since more power per unit area can be transmitted into the muscle without causing excessive heating near the surface.

Figures 9 and 10 illustrate the dependence of temperature profiles on blood flow. Figure 9 shows heating to the hyperthermic temperature range with 27 MHz radiation for various values of blood flow. It can be seen from the figure that the lower the blood flow, the smaller the power density required to raise the temperature of the muscle to the hyperthermic range, and the greater the depth at which one obtains hyperthermic temperatures. Figure 10 illustrates the temperature distribution due to heating with 27 MHz radiation in a model consisting of a 4 cm thick tumor buried at a depth of 6 cm in muscle. The tumor is assumed to have properties similar to normal muscle except that the blood flow in the tumor is assumed to be 1/4 that of the flow in the muscle. (Blood flow in malignant tumors is usually much lower than the blood flow in healthy tissues.)[10] Note that part of the tumor can be raised to the hyperthermic range without significantly increasing the temperature of the normal muscle. This phenomenon is of the greatest importance in hyperthermia treatment with RF because, generally, solid malignant tumors have poor blood circulation and therefore it is possible to selectively raise the temperature of such tumors with RF heating.

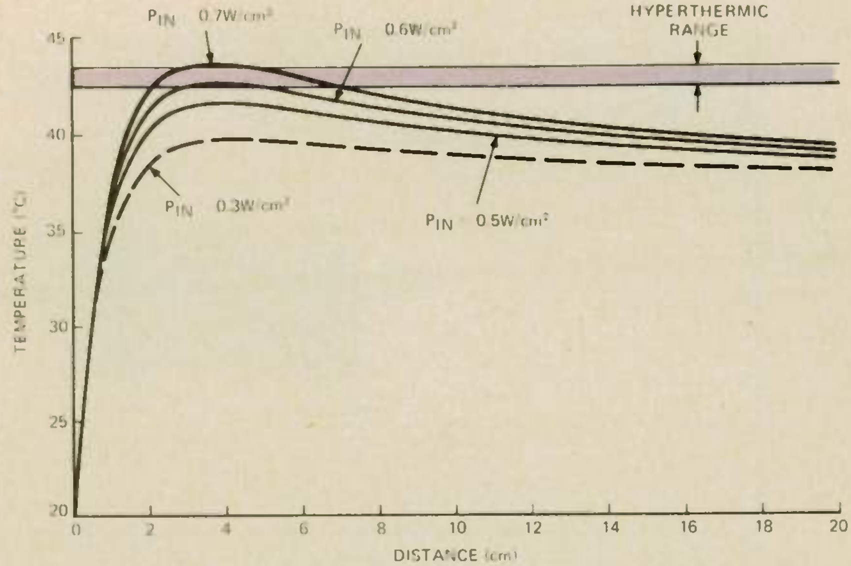

Fig. 7 Calculated temperature distributions in a semi-infinite slab of muscle for four different power densities when the muscle is heated with RF radiation at a frequency of 27 MHz and the surface of the muscle is maintained at 20°C by active cooling. (P_{in} = RF power density transmitted into muscle.)

ANIMAL EXPERIMENTS AND CLINICAL TRIALS

We have used the experimental arrangement shown in Figure 3 in both animal experiments

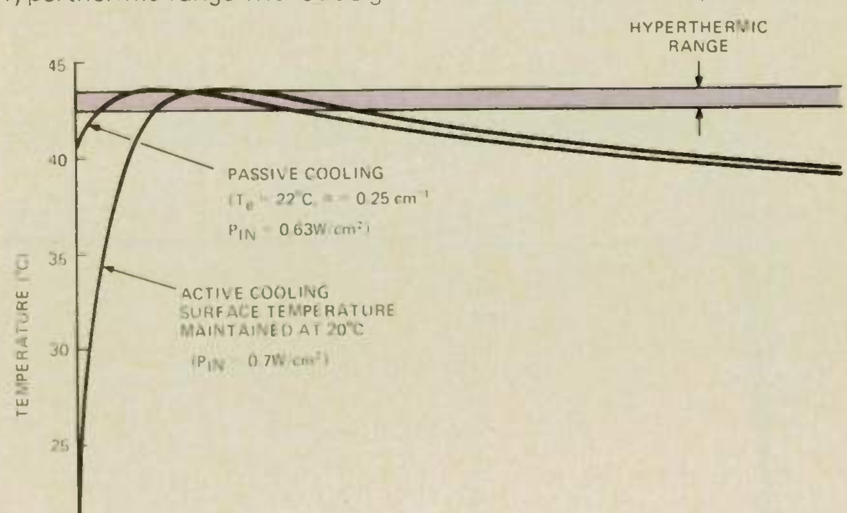

Fig. 8 Calculated temperature distributions in a semi-infinite slab of muscle for passive and active cooling when the muscle is heated with RF radiation at a frequency of 27 MHz. (T_e = temperature outside of tissue plane, α = heat loss coefficient from surface of muscle, P_{in} = RF power density transmitted into the muscle.)

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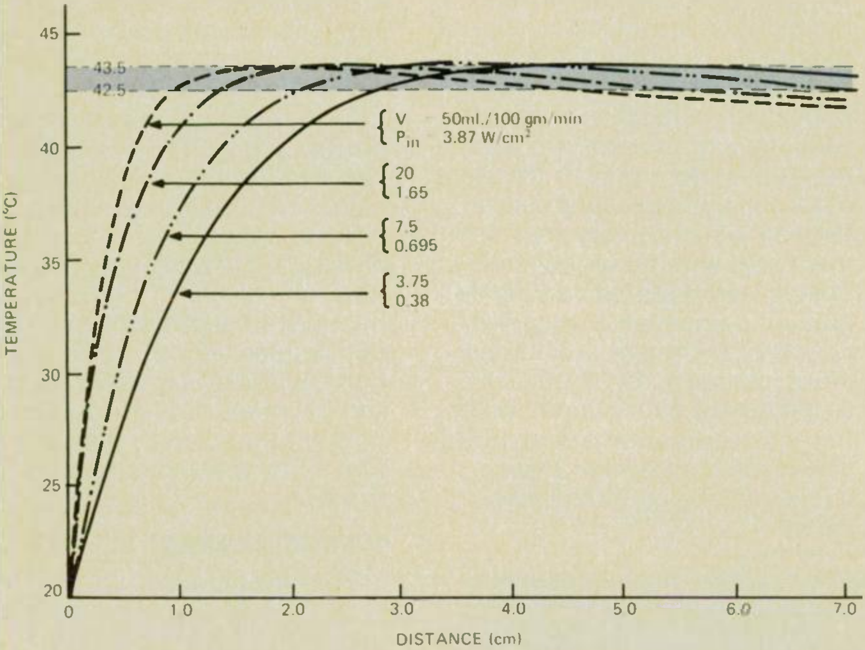

Fig. 9 Calculated temperature distributions in a semi-infinite slab of muscle for four different values of blood flow when the muscle is heated to the hyperthermic temperature range with RF radiation at a frequency of 27 MHz, and the surface of the muscle is maintained at 20°C by active cooling. P_{in} = RF power density transmitted into muscle.

and clinical trials. Tissue temperatures in animals or patients were measured with 3-mil diameter thermocouples immediately before and after heating. The thermocouples were removed during RF heating because 27 MHz RF fields induce currents in the thermocouple wires. These currents cause heating of the thermocouple junction and, therefore, produce misleading temperature readings. Various arrangements were used to support the animals or patients during RF heating. For example, in many of our animal experiments the animals were supported on a wooden table with a cutout for the ridged-waveguide applicator. Patients with tumors in the rectal area were treated while sitting on a specially constructed chair with a cutout for the applicator.

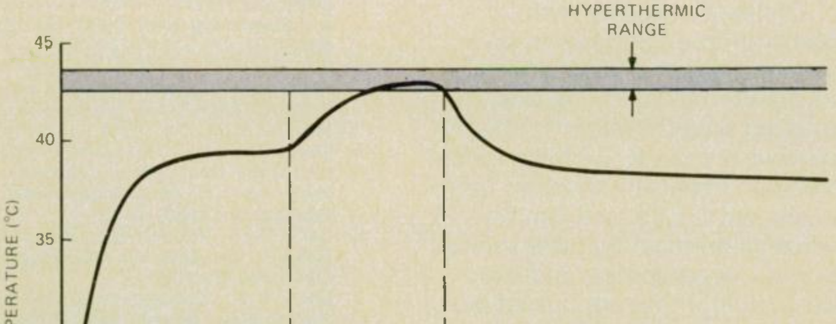

Fig. 10 Calculated temperature distribution in a layered one-dimensional structure of muscle-tumor-muscle when heated with RF power at a power density of 0.25 W/cm² at 17 MHz and the surface of the muscle is maintained at 20°C by active cooling. [λ (muscle) = 1. λ (muscle) = 1. λ (tumor) = 0.325 cm⁻².]

Patients who needed treatments in chest areas or in the extremities, etc., were treated lying dov on stretchers built around appli cators. Applicators of various di mensions were used. The dimen sions of the smallest applicator were (see **Figure 4**): a = 50.8 cm, b = 22.9 cm, c = 25.4 cm, d = 7.6 cm, height = 38.1 cm; the dimensions of the largest ap plicator were a = 58.4 cm, b = 26.3 cm, c = 29.2 cm, d = 13.7 cm, height = 86.4 cm. When properly matched, all applicato could raise well-vascularized mu cle to the hyperthermic range within several minutes. The required RF input power was typ cally 300-400 watts. Assuming that about 80% of the RF inpu power was concentrated across the ridge of the guide (i.e., the shaded area of **Figure 4**), this corresponds to power densities of roughly 0.8 W/cm², which is in good agreement with the calculated values of the previous section.

The ability of the waveguide applicators to produce uniform temperature distributions in healthy tissues was checked by heating the gluteal region of a 35 kg pig. The pig was anesthetized with IV nembutol and pos tioned on one of its sides with the gluteal region resting on a cooling pad placed on top of an applicator. Four plastic angiocaths were introduced into the muscle mass at distances of 2, 5 7 and 9 cm from the skin in cor tact with the cooling pad. The pig was then heated with 250 watts of 27 MHz power. Muscle temperature was measured by inserting thermocouples at five-minute intervals into the angiocaths. During these temperature measurements, the RF power was interrupted for 20-30 seconds. **Figure 11** is a plot of the measured temperatures as a fun tion of time. Note that after 20 minutes of heating the temperature distribution in the muscle i reasonably uniform, the peak temperature occurring 5 cm fro the skin.

In order to test the concept c differential heating of poorly va cularized deep-seated tumors with 27 MHz radiation (see **Figure 10**), the following experiment was performed: A 60 g piece of meat was formed into an egg-shaped mass and placed in a rubber balloon. This simulated tumor was then inserted into the rectum of a 40 kg anesthetized male dog. The dog was positioned on its side with the *gluteus maximus* resting on a cooling pad that is placed on top of the rubber membrane of an applicator. Plastic angiocaths were placed in (1) the center of the "tumor," (2) 3 cm above the "tumor," and (3) 3 cm below the "tumor." The "tumor" itself was approximately 6 cm above the applicator.

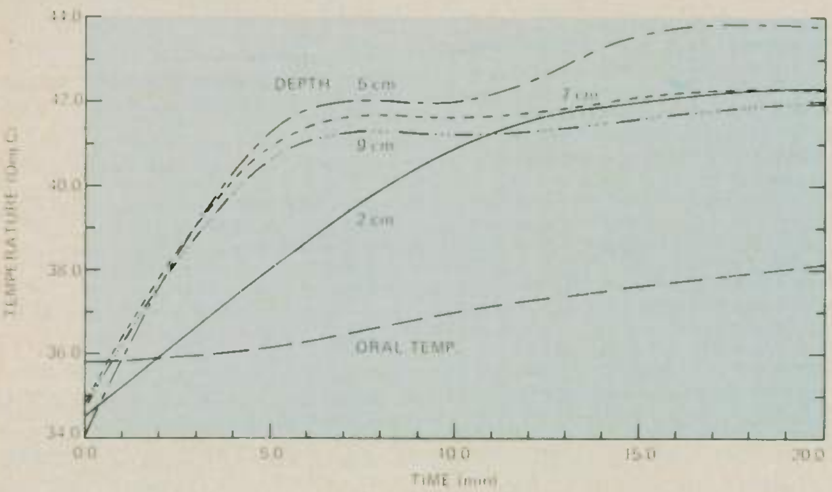

Fig. 11 Measured temperatures as a function of time in a gluteal muscle of a pig. Temperatures at 2, 4, 5, 7 and 9 cm from the skin are plotted.

Temperatures were measured at four-minute intervals by inserting thermocouples into the embedded angiocaths.

With 300 watts of 27 MHz power the non-vascularized "tumor" could be heated to 45-48°C within eight minutes. At the same time the normal tissues above and below the "tumor" remained below the hyperthermic range. The systemic temperature of the dog as measured by a fourth thermocouple inserted into a remote muscle rose to 39°C, indicating active dissipation of heat from the directly heated area. Also, the respiratory and cardiac rates of the dog increased markedly during the localized hyperthermia.

A number of patients with advanced malignancies were offered treatments of 27 MHz hyperthermia. Tumors treated included primary carcinoma of the lung, hip metastasis from carcinoma of the uterus, recurrent liposarcoma of the leg, recurrent breast carcinoma with metastasis to the ribs, carcinoma of the prostate, and metastatic carcinoma of the scapula from a primary lung cancer. Patients generally tolerated treatment well with no serious side effects, and reported varying degrees of pain relief. During the period of treatment oral temperature increased, usually to 39°C, accompanied by moderate sweating and acceleration of heart rate. The medical aspects of these studies will be published elsewhere.

CONCLUSIONS

Dielectric heating with 27 MHz radiation appears to be a safe and efficient means for noninvasively raising the temperature of deep-seated tumors to the hyperthermic range. 27 MHz radiation can penetrate through fat layers with little loss, can traverse anatomical features such as bones‡ or air spaces, and can heat tumors that are buried several centimeters deep inside muscle.

The water-filled ridged-waveguide applicators that are described in this paper are of reasonable size and are relatively inexpensive to fabricate. The rest of the equipment needed for hyperthermia treatment with 27 MHz radiation (RF power generators, power meters and temperature-measuring instruments) is commercially available.

Calculations of temperature profiles produced by RF radiation in simple models of living tissues are useful as guides to the effects of variations in RF power density, surface cooling, blood flow, etc. Much further work in this area is needed, particularly calculations taking into account finite aperture and tissue sizes, and the nonlinearity of blood flow as a function of temperature in normal and malignant tissues.

‡ Heating of bones can be minimized by orienting the electric fields in the RF wave perpendicular to the bones in the tissue being heated.

ACKNOWLEDGMENT

The authors wish to thank Elvira E. Petersack, Markus Nowogrodzki, Steven Weber and Francis J. Wozniak for their invaluable help during the course of this work.

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Micro– Coax® Cables

Qualify Under New MIL–C–17E

UNIFORM TUBES, INC.
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Twelve Micro-Coax® semi-rigid coaxial cables have been qualified under MIL-C-17E Qualified Products List M17. The newly qualified cables include all the most popular sizes of semi-rigid coax, and are available in both plated and non-plated versions.

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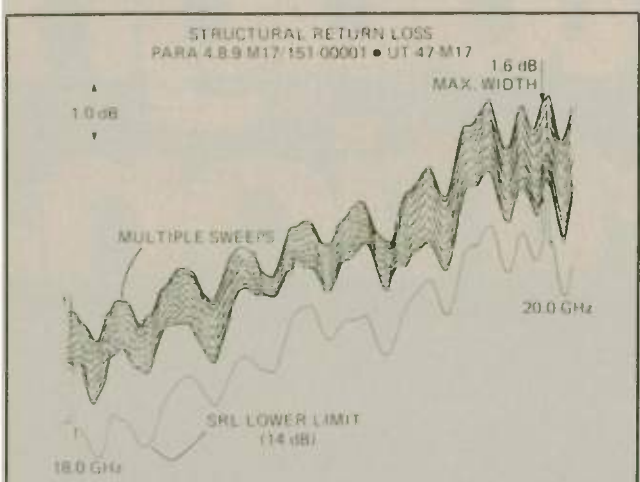
Four of the M17 qualified cables are in .047 inch and .034 inch OD sizes. These sizes are finding increased usage in densely packaged and minimum weight systems such as advanced airborne requirements.

The MIL-C-17 specification was developed to establish broad-band performance requirements for attenuation, and add specifications for SWR (structural return loss), coaxial concentricity, conductor adherence, dimensional stability, and weight. MIL-C-17E specifies quality assurance requirements and inspection methods.

All M17 Cables are carried in factory and distributor stocks in completely tested condition and the Uniform Tubes M17/129, 130 and 133 Cables are marked continously to provide quick identification of cable type and inner conductor type. Users requiring outer conductor finishes other than no plating or tin plating can specify the desired plating be applied on unplated M17 Cable.

PRODUCTS QUALIFIED UNDER MIL C 17E

Government Designation	MicroDelay (UT) Designation	Supersedes
M17/129/RG401	UT 250A M17	RG401/U
M17/129 00001	UT 250A TP M17	–
M17/130 RG402	UT 1415A M17	RG402/U
M17/130 00001	UT 1415A TP M17	–
M17/133 RG405	UT 85 M17	RG405/U
M17/133 00001	UT 85 TP M17	–
M17/133 00002	UT 85C M17	–
M17/133 00003	UT 85C TP M17	–
M17/151 00001	UT 47 M17	–
M17/151 00002	UT 47 TP M17	–
M17/154 00001	UT 34 M17	–
M17/154 00002	UT 34 TP M17	–



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CIRCLE 43 ON READER SERVICE CARD

LINE OF COAX ATTENUATORS AND TERMINATIONS

A line of low cost, coaxial attenuators and terminations come with SMA, BNC, or TNC connections and are offered in seven attenuation values 1, 2, 3, 6, 10, 20 and 30 dB with each connector. Both attenuators and terminations are effective over the dc to 2.5 GHz frequency band. Price: $19 for 1-24 units. Del: Stock to 2 wks. ARO. **COMPAC, Deer Park, NY. (516) 667-3933. Circle 186.**

BROADBAND SCHOTTKY DETECTORS

A Schottky detector module is designed for stripline and/or microstrip applications and operates from 0.5 GHz to 18 GHz with flat frequency response. T_{ss} is > -50 dBm and sensitivity at -20 dBm is > 1250 mV/mW. Circuit is hermetically sealed and two 2-56 through holes are compatible for microstrip mounting. Price: $< \$50$, 1-99 qty. **Litchfield Microwave Laboratory, Campbell, CA. Michael L. Litchfield, (408) 378-0377. Circle 187.**

PILLSHAPE ATTENUATOR WITH DUAL dc BLOCK

Pillshape attenuators with built-in dual dc blocks, Series PSA-DCB* models employ 10 Pf capacitors integral with the input and output terminals. Attenuation values range from 2 dB to 20 dB. Attenuators meet or exceed the requirements of MIL-E-5400 and MIL-E-10509. Price: $17.55 each, 100 pieces. **KDI Pyrofilm Corp., Whippany, NJ. (201) 887-8100. Circle 143.**

SINGLE-STAGE TUNABLE FILTER

Model TMF-1800C is a tunable single-stage microwave filter. It features an insertion loss of 3 dB maximum, frequency range of 2-18.5 GHz and 3dB bandwidth of 15-70 MHz. Selectivity is 6 dB/octave, nominal; passband ripple and spurious is 0.5 dB, maximum. Off-resonance isolation and spurious are both 30 dB, minimum. Filter has +10 dBm limiting level (minimum), typical SWR of 2, 50 impedance, hysteresis maximum of 20 MHz and input power of +30 dBm, maximum. Connectors are SMA type. Size: 3" W x 3" H x 4" D. Price: $3450. **Integra Microwave, Inc., Santa Clara, CA. Len Johnson, (408) 727-9601. Circle 140.**

MINIATURE DETECTOR LOG VIDEO AMPLIFIERS

Series LDS 1500 are miniature detector log video amplifiers with a package size of 2.5" x 3.35" x 0.59" – a reduction to approximately one-third the size of previous units. Units cover five frequency bands from 0.75-20 GHz. Each model within the series provides 15 MHz video bandwidth, -44 dBm tangential signal sensitivity, 40 dB instantaneous dynamic range, and 23 ns rise time. Amplifiers are qualified for shipboard environment over the -30°C to +85°C temperature range. dc offset, gain matched sets, linear output and other options are available. **Aertech Industries, Sunnyvale, CA. W. S. Patton, (408) 732-0880. Circle 145.**

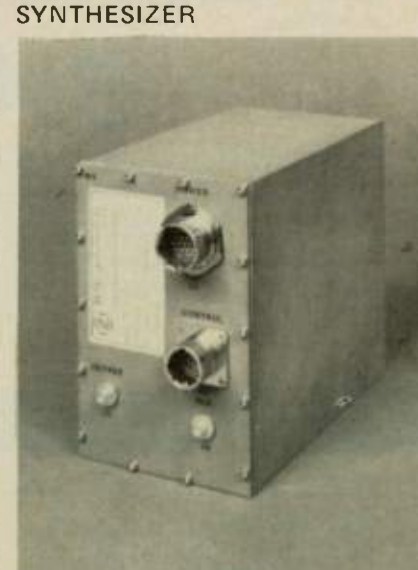
COAXIAL CABLE FEED THROUGHS

Series of coaxial cable feed throughs provide method for terminating coaxial cable. They are designed for use in installations that do not require the capability of connect/disconnect functions. Feed throughs are offered for feeding coaxial cable through bulkheads, terminating at a bulkhead, or for terminating on a printed circuit board, and can be supplied with either clamp-on or crimp-on cable connection. They provide proper strain relief while assuring reliable electrical termination of both center and outer cable conductors. **RF Components Div., Sealectro Corporation, Mamaroneck, NY. (914) 698-5600. Circle 148.**

THICK FILM MICROWAVE VARIABLE ATTENUATOR

Model ATN-4002 is an integrated electronically controlled attenuator covering the 5-1000 MHz band with an attenuation range of at least 15 dB. Insertion loss, maximum, is 2.5 dB @ 5-500 MHz; 3.0 dB @ 5-1000 MHz. Gain flatness is ± 0.25 dB, maximum at 5-550 MHz and ± 0.5 dB, maximum at 5-1000 MHz. Attenuator has a SWR, in/out maximum of 2.0. Supply current is 15 mA, supply voltage 15 Vdc, and control current is 6 mA. Housing is a hermetic TO-8 package. Thick-film processing techniques render it capable of withstanding MIL-Q-883B screening. **Optimax, Div. of Alpha Industries, Inc., Colmar, PA. (215) 822-1311. Circle 147.**

S-BAND FREQUENCY SYNTHESIZER



Model 6748-01 is an S-band frequency synthesizer which covers the 2.5-3.5 GHz range in 1 MHz steps and settles to within ± 1 PPM of the final frequency in 500 microseconds. Output power is +10 dBm and spurious rejection exceeds 65 dB. Single sideband phase noise is less than -85 dBc/Hz at 10 kHz. Inputs are 10 MHz, +5 dBm reference, +28 V and +5 V and TTL compatible commands. Size: 8" x 6.25" x 4". Weight: 7 lbs. **Zeta Labs, Inc., Santa Clara, CA. (408) 727-6001. Circle 183.**

THIN-FILM CASCADABLE AMPLIFIER MODULES

A series of TO-8 and TO-3 packaged thin film cascadable amplifier modules, particularly for wide dynamic range IF amplifiers, incorporate GaAs and VMOS FETs. Model UTO-161 (TO-3) employs a VMOS FET to produce a minimum output power of +32 dBm (+20 V input) and +15 V input) over the -54° to +71°C temperature range at frequencies from 10-100 MHz. Minimum gain is 8 dB, noise figure is 6 dB maximum, third order intermodulation intercept point is +43 dBm. Model UTO-2012 and UTO-2013 (TO-8) use GaAs FETs and offer up to +18 dBm output power, 8.5 dB gain, noise figures as low as 5 dB and +32 dBm third order intercept point over the 500-2000 MHz range. Price: UTO-161, $160; UTO-2012, $195 and UTO-2013, $295, 1-9 qty. Del: 10-12 wks. ARO. **Avantek, Inc., Santa Clara, CA. John Hyde, (408) 727-0700.** **Circle 141.**

DIGITAL CONTROLLED MICROWAVE PRESELECTOR

Model DR-218A is a digital controlled microwave preselector which covers the 2-18 GHz frequency band and is suited for use in processor-controlled microwave receiving systems. Tuning is accomplished via an 8-bit serial digital word in ASCII format in less than 10 ms. Overall tuning accuracy is 0.5%, insertion loss is 4 dB, maximum and 3 dB bandwidth is 35 minimum, 55 MHz maximum. Power requirements are +24 V @ 2.0 A, -15 V @ 200 mA. Connectors are SMA for RF and DM-25 for control/power. Size: 4" x 6" x 8". Price: $4300, 1-9 qty. Avail: 180 days. **American Microwave Corp., Damascus, MD. Raymond L. Sicotte, (301) 253-6782.** **Circle 175.**

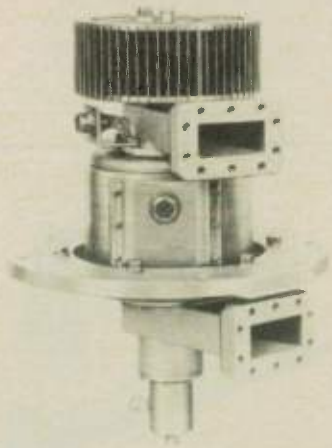
(continued on page 88)

(from page 87) NEW PRODUCTS

MODULAR RF SWITCHES

Modular RF coaxial switches are available in three through ten position models with latching and normally open configurations for dc-3 GHz; 3-8 GHz, and 8-12.4 GHz applications. The modular concept reduces replacement time by eliminating the need for soldering operations. SWR, maximum, is 1.2-1.4, depending upon frequency band; insertion loss (maximum) ranges from 0.2-0.4 dB and isolation (minimum) from 60 dB-80 dB. Impedance is 50 ohms, switching time is 20 ms, maximum and rated RF power is 50 W, average, Actuating current for unit is 115 mA maximum at 28 Vdc. Models meet MIL-E-5400, Class 2 environmental standards; finish is electroless nickel per MIL-C-260741 with gold-plated contacts per MIL-A-8625. Price: varies according to specifications. Del: 6-8 wks. **ARO. U.Z., Inc., Culver City, CA. Jerry Hoffman, (213) 839-7503.** **Circle 149.**

40 kW DUAL CHANNEL ROTARY JOINT



A dual channel, 40 kW average power rotary joint designed for use in ship surveillance radar provides a waveguide channel with a 40 kW average power, 3.5 MW peak power rating over the 2875-3125 MHz band. Channel 2 is coaxial and is rated at 150 W average, 15 kW peak power over 1015-3125 MHz. Units have a 12-circuit slip ring. Size: 7" D x 18.312" L. **Kevlin Manufacturing Company, Woburn, MA. (617) 935-4800.** **Circle 144.**

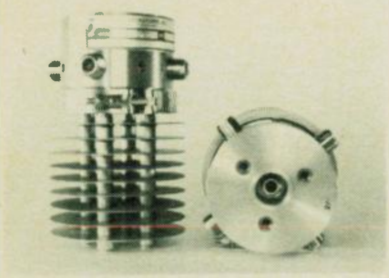
LOG IF AMPLIFIERS FOR 2 GHz BAND

An IF-to-log video amplifier series, ICL-5, cover the 600-2000 MHz frequency range. Log amplifier accuracy is typically less than ± 1 dB deviation from ideal log plot at temperatures to 85°C. Options such as hermetically sealing and power supply protection and regulation are offered. **Varian Associates, Beverly, MA. (617) 922-6000.** **Circle 176.**

MULTI-OCTAVE HIGH POWER COAXIAL QUADRATURE HYBRIDS

A series of multi-octave coaxial quadrature hybrids for 2-8 GHz (EG504A/EG504AP) feature compact, high power double octave design (18 dB typical and 15 dB minimum). Frequency sensitivity is ± 0.3 dB typical and ± 0.5 dB maximum, isolation is 15 dB minimum. SWR is 1.4 maximum, phase difference is 90° ± 5 between output ports and power rating is 400 W CW, 6 kW peak. Temperature range for series is -54° to +100°C or that of the connector used (type N, TNC and SC Female offered). Typical power rating of 800 W average, 6 kW peak can be supplied with SC connectors. **Microwave Research Corporation, North Andover, MA. Robert A. Schmidt, (617) 685-2776.** **Circle 179.**

S-BAND POWER DIVIDER WITH 6 kW PEAK, 360 W, AVERAGE



A high power S-band power divider, Model FP2757, operates from 3.1-3.5 GHz, with 6 kW peak and 360 W average input power. Built-in terminations will dissipate power reflected from a short or open circuit load without damage. Specification limits include input SWR of 1.5, output SWR of 1.2, amplitude unbalance of 0.4 dB and phase unbalance of 4°. Isolation between outputs is 20 dB. **Sage Laboratories, Inc., Natick, MA. Ken Paradiso, (617) 653-0844.** **Circle 139.**

TUNABLE BANDPASS FILTER

Model 201575, a tunable bandpass filter, is constructed of silver-plated brass and Invar. Screwdriver range is 4.9-5.52 GHz, 3 dB bandwidth is 16 ± 1 MHz, 25 dB bandwidth is ± 60 MHz maximum. Insertion loss is 1.5 dB maximum at fo ± 2 MHz min. and SWR is 1.4 maximum. Connector is SMA Female. Price: $525, small qty. Del: 4-6 wks ARO. **Coleman Microwave Company, Edinburg, VA. (703) 984-8848.** **Circle 177.**

PUSHBUTTON ATTENUATORS FOR dc-750 MHz

A series of pushbutton attenuators cover the dc to 750 MHz (50 Ω) or dc to 500 MHz (75 Ω) bands. Attenuation range is 0-45.5 dB in .5 dB steps or 0-65 dB in 1 dB steps. Overall attenuation accuracy from dc-250 is ± .5 dB and from 500-750 is ± 1.3 dB. Attenuators have SWR's of 1.3 (75 Ω) or 1.4 (50 Ω), maximum. Maximum power rating is 1 W average and BNC, TNC or "F" connectors are available. Price: $80, from 1-9 qty. **JFW Industries, Inc., Beech Grove, IN. J. L. Walker, (317) 783-9875.** **Circle 178.**

(from page 89) NEW PRODUCTS

COMPACT LUMPED ELEMENT IF FILTER FOR 25 MHz - 2 GHz BAND

A line of compact lumped element IF filters, Series OPL, covers the 25 MHz to 2 GHz band. A typical unit (OPL-071), with a center frequency of 70 MHz and a bandwidth of 15 MHz, exhibits insertion loss of 1 dB, SWR of 1.5, or less, 60 dB rejection, minimum and a 60 dB bandwidth of 44 MHz. Parabolic group-delay in the pass band in 60 ns. This filter measures 3.5" x 1.0" x .563", not including the standard SMA connectors, and it is built to meet MIL-E-5400, Class 2. **Omniyig, Inc., Santa Clara, CA. William Capogeannis, (408) 988-0843.** Circle 152.

DOUBLE BALANCED MIXER COVERS 0.5-18 GHz BAND

Model MD-166 is a double balanced mixer which spans the 0.5-18 GHz frequency band. Unit exhibits a 8 dB typ., 10 dB maximum conversion loss, an IF bandwidth to 5 GHz, 1 dB conversion loss flatness and isolation better than 20 dB on all ports. Other features include a 3-PIN hermetic module for stripline and microstrip integration. SMA connectorized version also can serve as a module test fixture. Mixer can be screened to MIL STD 883. Price: $550 for module in 1-5 qty; $625 for SMA model. Del: from stock. **Anzac Division, Adams-Russell Co., Inc., Burlington, MA. (617) 273-3333.** Circle 138.

COAXIAL FEED THROUGH TERMINATIONS

A series of coaxial feed through terminations are available in 50, 75, and 93 ohms models with BNC, TNC, N or SMA male to female connectors. Model FT-50 operates from dc to 1 GHz; Model FT-75 from dc to 500 MHz, and Model FT-90 from dc to 150 MHz. Units have a dc accuracy of 5% and a 1 W CW, 1 kW peak power rating. Price: from $11.10 each, small qty. Del: 30 days ARO. **Elcom Systems, Inc., Boca Raton, FL. Leonard Pollachek, (305) 994-1774.** Circle 146.

360° DIGITAL PHASE SHIFTER

Model QQ74 is a digitally controlled analog phase shifter which operates over the 8.0-18.0 GHz frequency band. Phase shift is 360° min. and phase flatness is ± 15° to 180°, ± 35° to 360° over the entire band and ± 3° to 180°, ± 5° to 360° over any 200 MHz segment. Insertion loss is 17 dB maximum, SWR is 2.5 maximum; amplitude ripple is ± 1 dB maximum and switch speed is 150 ns maximum. Phase shifter has 10 bits capability, and response is monotonic. Requires: ± 15 V at 50 mA. **Triangle Microwave Inc., East Hanover, NJ. James Beard, (201) 884-1423.** Circle 151.

ONE-PIECE SOLDER TYPE SMB SNAP-ON PLUGS

A one-piece, semi-rigid SMB cable plug (#701908-002) is designed for .086" cable. Units meet all applicable portions of MIL-C-39012B; metal parts are gold plated per MIL-C-45204, Type 1, Class 2, Grade C over copper plate per MIL-C-14550, Class 4. Other plating is offered. **Cablewave Systems, Inc., North Haven, CT. Steven Raucci, Jr., (203) 239-3311.** Circle 142.

Shielded Chamber

ODULAR RF-SHIELDED HAMBER FOR EMC EASUREMENT

ystem 86/3 is a pre-engineered modular F-shielded anechoic chamber for indoor round reflection EMC measurement. Chamber can be used for testing under FCC ocket, 80-284, CISPR, VDE and MIL-TD-285, 461, 462 requirements. Interior all and ceiling RF absorbers keep ambient mplitude levels below ± 2 dB at 30 MHz GHz and above. RF-shielded panel contruction provides up to 120 dB attenuaion from 14 kHz to 100 GHz. Floor deign provides a reflective ground plane. ize: 24' L x 12' W x 11' H, with variable eiling height. **Keene Corporation, Ray roof Division, Norwalk, CT. Bob Barbour, 203) 838-4555.** **Circle 131.**

Systems

-BAND RADAR PACK

ew Japan Radio Co. announces availabilty in the USA of its X-band radar pack, a icrowave system containing a magnetron, unn oscillator, balanced mixer, primerless R-limiter and series tee. It is available at .375 GHz and 9.4 GHz with transmitted ower up to 50 kW, depending upon the hoice of magnetron. The overall mixer oise figure is under 7 dB and TR spike akage is under 0.1 erg measured at 40 kW eak, .001 duty. Price for typical 5 kW radar pack complete with magnetron: under 500, FOB USA. **Calvert Electronics, Inc., ast Rutherford, NJ. Bernard Fudim, 201) 460-8800.** **Circle 180.**

Design Aid

AD FOR MICROWAVE CIRCUITS

ICRO COMPACT™ is a computer-aided esign program which analyzes and optiizes passive and active microwave circuits. is specifically designed to run on the ewlett Packard 9845 desktop computer ith B and T options, and utilizes several xtended capabilities of the 9845T/B, inuding graphics and data cartridge storage. nalysis and optimization in the frequency omain for two-port linear circuits are proided. Complex interconnections and casade connections can be handled and an inernal text editor is provided. **Compact ngineering, Inc., Palo Alto, CA. Jim Linauer, (415) 858-1200.** **Circle 132.**

(continued on page 92)

Instrumentation

NOISE SOURCE DRIVER PROVIDES 28 VOLTS

Model DR-1 is a noise source driver designed to provide an accurate 28 volts from 0-40 MA to drive solid state noise sources. Internal 400 Hz square pulse modulation or external TTL level pulse modulation may be used. Power requirements are 115 V at 50/60 Hz. The driver is designed to operate over the 0-70°C temperature range. Size: 3 1/2" x 7 7/8" x 5 11/16". Price: $225, each. Del: Stock. **Micronetics, Inc., Norwood, NJ. Gary Simonyan, (201) 767-1320.** Circle 135.

SWEEP OSCILLATOR MAINFRAME WITH IEEE BUS

Model 430C is a sweep oscillator mainframe for the 0.01-40 GHz frequency range which features an integrated IEEE-488 Interface Bus Option -09 which may be factory or field installed. Functions controlled via Option -09 and the bus include RF power (with up to 256-point resolution over a 15 dB range) frequency (with 10,000-point resolution between any start and stop frequency) and band select, sweep on/off and RF on/off modes. Size: 5.63" H x 16.75" W x 15.75" D. **Weinschel Engineering, Gaithersburg, MD. (301) 948-3434.** Circle 137.

PRECISION AUTOMATIC NOISE FIGURE INDICATOR

Model 7514 is an improved precision automatic noise figure indicator. Accuracy and versatility improvements over earlier model are offered and an option provides six preselected, front-panel switched input frequencies – 21.4, 36, 45, 60, 70, 160 MHz – in addition to 30 MHz. The same option, when used with an external local oscillator, permits noise figure measurements on devices with output frequencies from 10-1000 MHz. Full scale ranges are 0, 3, 6, 12 and 18 dB, resolution is better than 0.05 dB and direct-reading results for a range of noise generator outputs is provided. Size: 14 7/8" L x 17" W x 5 1/4" H. Weight: 21 lbs. **AILTECH Electronic Instruments, Eaton Corp. Div., Ronkonkoma, NY. (213) 965-4911.** Circle 133.

ISOTROPIC RADIATION MONITOR SENSES LEVELS DOWN TO 1 μW/cm^2

The RAHAM Model 4A is a high-sensitivity, ultra-broadband radiation meter for measuring nonionizing radiation from levels as low as 1 μW/cm^2. Instrument operates over a frequency range from 200 kHz to 26 GHz with a single omnidirectional probe. **General Microwave Corporation, Farmingdale, NY. Moe Wind, (516) 694-3600.** Circle 173.

BASEBAND ANALYZER FOR MICROWAVE RADIO MEASUREMENTS

Model 3724A/25A/26A Baseband Analyzer combines all of the instruments used on the baseband of a radio system in one integrated test set and links them through a common CRT and keyboard and permits measurement modes to be changed at the press of a key. The instrument replaces: wideband power meter, selective voltmeter, synthesized signal generator, counter, spectrum analyzer and white-noise test set. Test set is HP-IB (IEEE-488) compatible and all front-panel controls may be accessed via the bus. The analyzer complies with CCIR, CCITT, Intelsat and Bell recommendations for choice of bandwidths and frequencies. Price: #3724A/25A/26A, $44,000. Del: From March, 1981. **Hewlett-Packard Co., Palo Alto, CA. (415) 857-1501.** Circle 134.

DIGITAL STORAGE RF SPECTRUM ANALYZER

Model 7L14 is an RF spectrum analyzer for the 10 kHz to 1800 MHz frequency band with digital storage capabilities. It features 70 dB on-screen dynamic range, spurious free and −130 dBm sensitivity with 30 Hz resolution. CRT readout of control settings, 4:1 shape factor resolution filters and a display mainframe compatible with more than 25 7000 Series plug-ins are provided. Signal levels up to 1 W can be connected to the input for any setting of the RF attenuator without damage to the limiter-protected first mixer whichis also protected from large (up to 50 V) line frequency (50/60 Hz) signals which may be present along the wanted signal. Size: 5" x 8.15" x 14.75". Weight: 16 lbs. Price: $15,000. Avail: 4 wks. **Tektronix, Inc., Beaverton, OR. (503) 644-0161.** Circle 136.

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New Literature

SWEPT ANALYZER APPLICATIONS MANUAL

A 16 page application manual, AN 19, covers the Model 1038 D 14 swept analyzer. The manual offers step by step swept analysis procedures, including on how to program the analyzer, using the General Purpose Interface Bus (GPIB). GPIB makes the analyzer compatible with both computerized analysis systems and automated test equipment. **Pacific Measurements, Inc., Sunnyvale, CA. Jim Forbes, (415) 968-7570.** **Circle 158.**

BROCHURE ON MINIPAC™ AMPLIFIER LINE

A two-color brochure describes the Minipac™ line of low profile solid state amplifiers. This 15-page booklet includes an applications section, details on microwave connectorless package design benefits plus product line specifications charts as well as amplifier and outline drawings. **Watkins-Johnson, Palo Alto, CA. (415) 493-4141.** **Circle 164.**

PULSED POWER METER BOOKLET

Literature describes the pulsed power meter (PPM-101A) designed for measurements of instantaneous power within the L band frequency range with an accuracy of ± 0.5 dB. Specifications and modes of operation are detailed. A manual mode permits power to be measured at any point on the waveform. **Republic Electronic Industries Corp., Melville, NY. John Michaels, (516) 249-1414.** **Circle 161.**

SHORT-FORM RF TRANSISTOR CATALOG

Catalog 503A is a 12 page, short form catalog which lists basic specifications for some 156 RF and microwave transistor types. Ten categories of products are shown, along with photographs and engineering drawings of the various package types available. An alphanumeric index and cross referenced table is included. **TRW RF Semiconductors, Lawndale, CA. Dan Faigenblat, (213) 679-4561.** **Circle 162.**

BNC COAXIAL CONNECTOR CATALOG

Sixteen-page catalog details 250 standard BNC coaxial connectors. Two-color literature contains illustrations and cross-sectional views of four lines of clamp field-serviceable connectors. The standard wedge-lock, wedge-eze and improved "V" groove plus three crimp lines (X, econo and improved econo) are described. **Automatic Connector, Inc., Commack, NY. Edward Selig, (516) 543-5000.** **Circle 193.**

MARINE RADAR REPLACEMENT TUBE SUMMARY

A summary of New Japan Radio Company's microwave tubes for commercial marine radars is offered. This 16 page booklet contains cross references to over 400 NJR magnetrons, TR tubes and solid state local oscillators suitable for replacement use in 325 commercial marine radars made by 23 manufacturers in Europe, Canada, US and Japan. **Calvert Electronics, Inc., East Rutherford, NJ. Bernard Fudim, (201) 460-8800.** **Circle 155.**

WORLD STANDARD FOR TNC CONNECTORS

The new international standard for TNC type connectors is available in IEC Publication 169-17. This 35-page booklet provides the type designation, standard ratings for grade 2 connectors, preferred climatic categories, dimensions for grade 2 general purpose connectors, requirements and test sequences for gauges and standard test connectors and a schedule for type tests for first through sixth lot testing. Price: S.francs 42. **International Electrotechnical Commission, Geneva, Switzerland. 34 01 50; Telex: 28872 CEIEC-CH.** **Circle 156.**

COAXIAL AND WAVEGUIDE SWITCH CATALOG

Catalog No. 5 describes a broad line of coaxial switches for application up to 18 GHz. Average power ratings range up to 5 kW. Electrical and mechanical specifications are shown for SPDT through SP10T and transfer switch models with a variety of connector options, housing styles, switch circuit options and such special options as T² logic drivers and terminations. In addition, a waveguide SPDT model for X band and a trimming machine for semi rigid coaxial cable are described. **UZ, Inc., Culver City, CA. Robert Hamilton, (213) 839-7503.** **Circle 163.**

TUNABLE FILTER LINE BROCHURE

A 4 page brochure for a tunable filter line includes electrical, dimensional and environmental data on more than 30 standard filters. Passband frequency, 3 dB and 30 dB bandwidths, insertion loss, connectors and tuning range are shown in tabular form. Diagrams and drawings indicate standard filter designs and availability of higher and lower frequencies and bandwidths. **Premier Microwave Corp., Port Chester, NY. Jules Simmonds, (914) 939-8900.** **Circle 160.**

MODPACK CATALOG AND PRICE LIST

A 17 page catalog and 4-page price list covering the Modpak product line of RF shielded circuit enclosures are offered. The enclosures are designed to minimize assembly time and they provide direct access to both sides of the PC board. N, TNC, SMA, or BNC connectors are available. Mechanical drawings and photographs for each product type are provided. **Adams Russell Co., Inc., Modpak Division, Burlington, MA. (617) 273-3330.** **Circle 154.**

SIGNAL GENERATOR SERIES BROCHURE

The C/X Signal Generator Brochure provides specification data and highlights features of the Polarad "E" Series generators which operate from 3.7 11.0 GHz. Options and accessories also are described. **Polarad Electronics, Inc., Lake Success, NY. (516) 328-1100.** **Circle 159.**

MIXER PRODUCTS CATALOG

A 16 page catalog features microwave mixer products and provides detailed specifications and operating characteristics on seven different models covering 0.5-18.0 GHz with octave, multi-octave and special purpose models. All units are designed for military environments and are available in hermetic modules or SMA connectorized versions from stock. **Anzac Division, Adams-Russell Company, Inc., Burlington, MA. (617) 273-3333.** **Circle 153.**

MICROWAVE COMPONENT CAPABILITY BROCHURE

A six-page brochure outlines the product capability and facilities of this microwave component supplier. Products for high power applications in the 0.1 to 40 GHz range include absorptive and reflective filters, PIN switches, couplers, adaptors, transitions and dummy loads. A multiplexer capability is also discussed. **Wincom Corp., Lawrence, MA. Robert Antonucci, (617) 685-3930.** **Circle 165.**

ENGINEERING DATA ON VULCANIZED STRIP GASKETING

Bulletin ESG 850 features photographs, tables, charts, and graphs that provide data on Combo Seal, a vulcanized strip gasketing that combines silicone rubber with knitted metal mesh to provide an effective EMI/RFI shield as well as a seal against severely hostile environments. Brochure contains specifications on shielding effectiveness and sealing and shielding material. Graphs depicting compressed height vs force per inch of gaskets and required dimensions for typical applications are included. **Metex Corporation, Electronic Products Div., Edison, NJ. John Severinsen, (201) 287-0880.** **Circle 157.**

RF CAPACITOR/INDUCTOR CATALOG

Line of fixed variable capacitors and inductors for RF applications from HF to microwave ranges is described in a 36-page catalog. Describes features and technical specifications pf multi-layer porcelain, glass-encapsulated, and single-layer chips, chips for IC's, high power and precision air variable capacitors, plus ceramic trimmers, metallized inductors and LC networks. **JFD Electronic Components, Div. of Murata Corp. of America, Douglasville, GA. (404) 949-6900.** **Circle 192.**

MSC

Power GaAs FETs

Products for Terrestrial Radio and Satellite Communications Systems

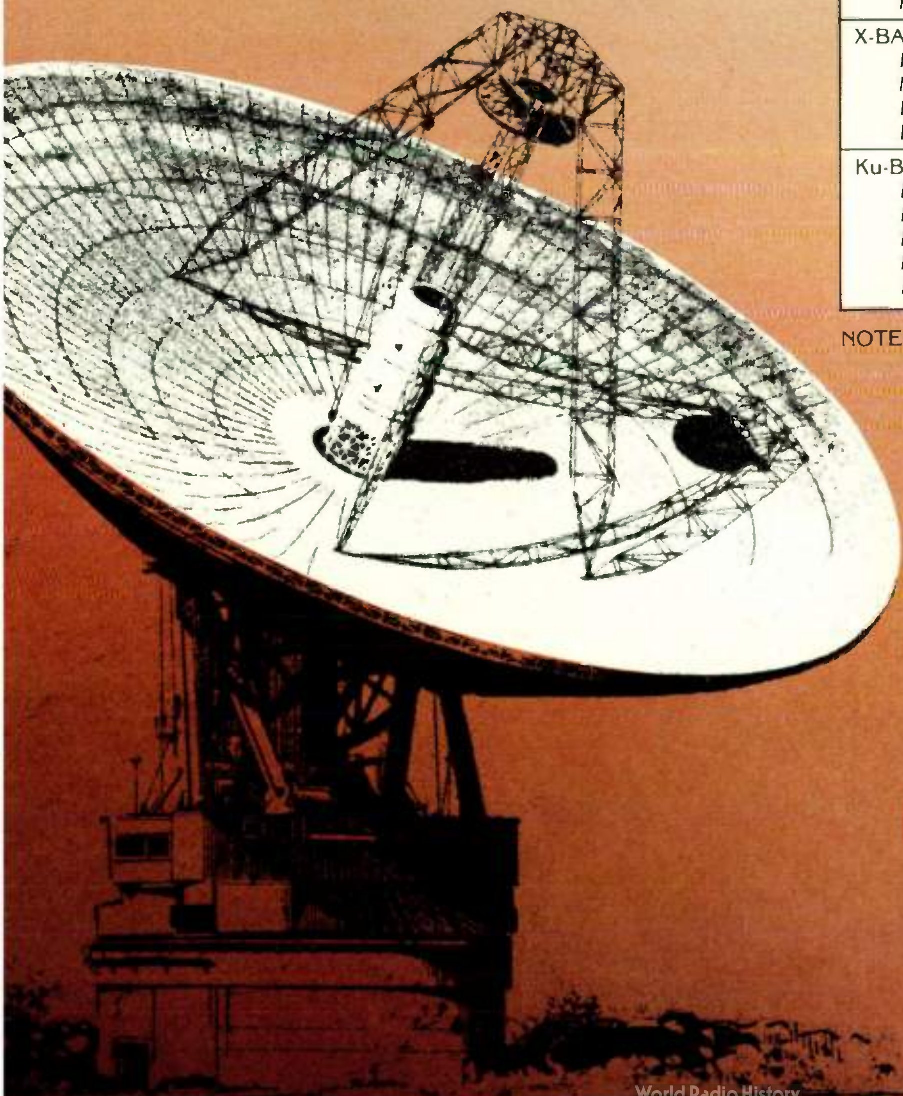
MSC 88000 SERIES

Electrical Characteristics (@ 25°C)

MODEL NUMBER	TEST FREQ (MHz)	P_{OUT}(1) MIN (W)	P_{IN} (mW)	θ_{CC}(2) TYP (°C/W)
C-BAND SERIES				
MSC 88000	6000	0.050	8	45
MSC 88001	6000	0.175	40	35
MSC 88002	6000	0.350	90	25
MSC 88004	6000	0.800	200	20
MSC 88012	6000	3.500	800	7
X-BAND SERIES				
MSC 88100	12000	0.050	16	45
MSC 88101	12000	0.175	56	35
MSC 88102	12000	0.350	125	25
MSC 88104	12000	0.800	280	20
Ku-BAND SERIES				
MSC 88199	15000	0.025	6	40
MSC 88200	15000	0.100	25	35
MSC 88201	15000	0.200	70	29
MSC 88202	15000	0.400	140	23
MSC 88204	15000	0.800	316	15

NOTES (1) Power Output at the 1dB Gain Compression point is defined as the point where further increases in input power cause the output power to decrease 1dB from the linear portion of the curve.

(2) Thermal Resistance determined by Infra-Red Scanning of Hot-Spot Channel Temperature at rated RF operating conditions. Reference MSC Application Note TE-212.



YOUR TOTAL MICROWAVE RESOURC

The MSC series of GaAs FETs are designe for linear power amplifiers and for oscillat applications from 2-15GHz. Higher frequenc state-of-the-art devices are also available.

MSC **MICROWAVE SEMICONDUCTOR CORP.**
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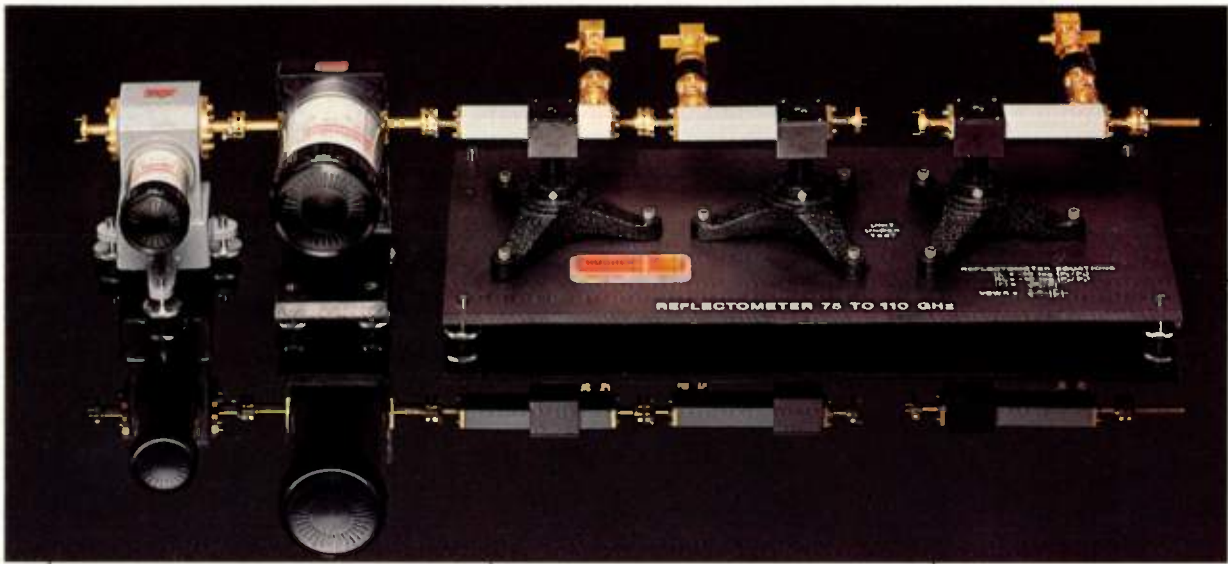
CIRCLE 2 ON READER SERVICE CARD

Impedance Measurement. Hughes makes it fast, accurate and simple.



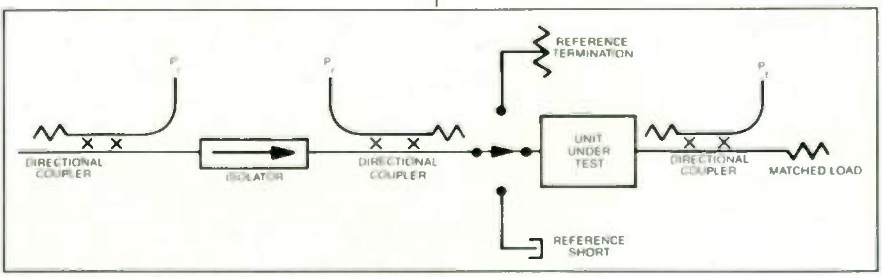
Our IMPATT-sourced millimeter-wave reflectometer is far more simple to use than slotted lines and hybrid impedance bridges.

Its unique design lets you sweep the full 75-to-110 GHz bandwidth automatically, for fast, accurate data—without time consuming point-to-point measurements. And it's packed with advanced technology.

Like Hughes-built broadband isolators and detectors to reduce the effects of unwanted reflections and mismatch error. Or the 1-KHz square wave modulator that delivers increased sensitivity and dynamic range. Or our rotary vane attenuator for constant attenuation with frequency.

A small computer or desk top calculator can be added to give you calibration factors instantly when you need them. This lets you factor out the frequency dependence of test line insertion loss, coupling coefficients and detector response.

Hughes reflectometer systems cover frequency ranges from 40 to 110 GHz and can be ordered as a complete system or millimeter-wave components only.



Simplified reflectometer block diagram. The Hughes technique extends reflectometer measurements into the millimeter-wave region.

So if you need a fast, accurate and simple system for measuring millimeter-wave impedance, come to Hughes. Our involvement in millimeter-waves covers nearly two decades and includes a broad spectrum of components and instrumentation.

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ELECTRON DYNAMICS DIVISION

Making waves in millimeter-wave technology.