microwave JOURNAL

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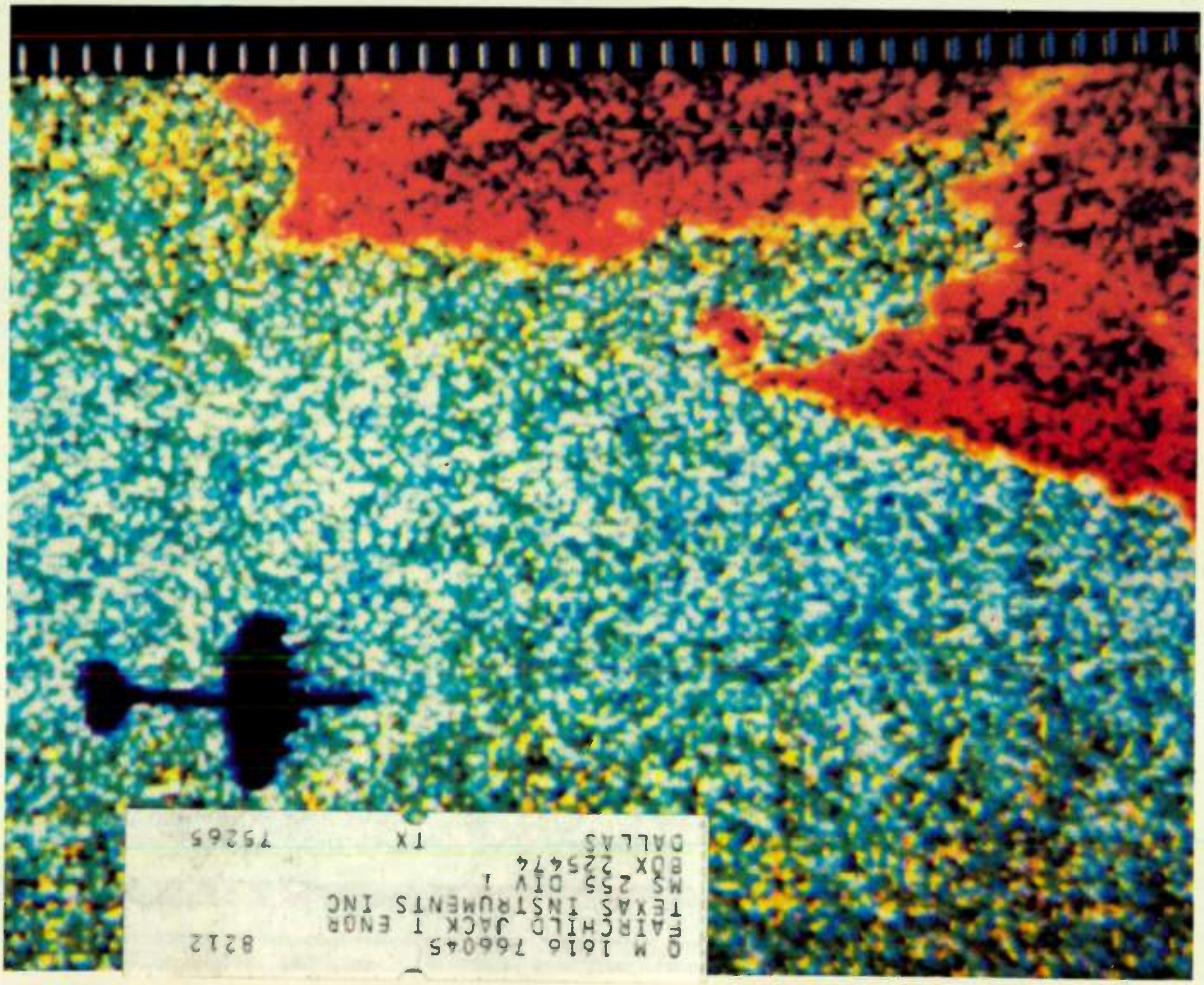

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632B-1	100 kHz to 2 GHz	N/A	$10,175
630B	3 MHz to 40 GHz	$12,460	$14,285
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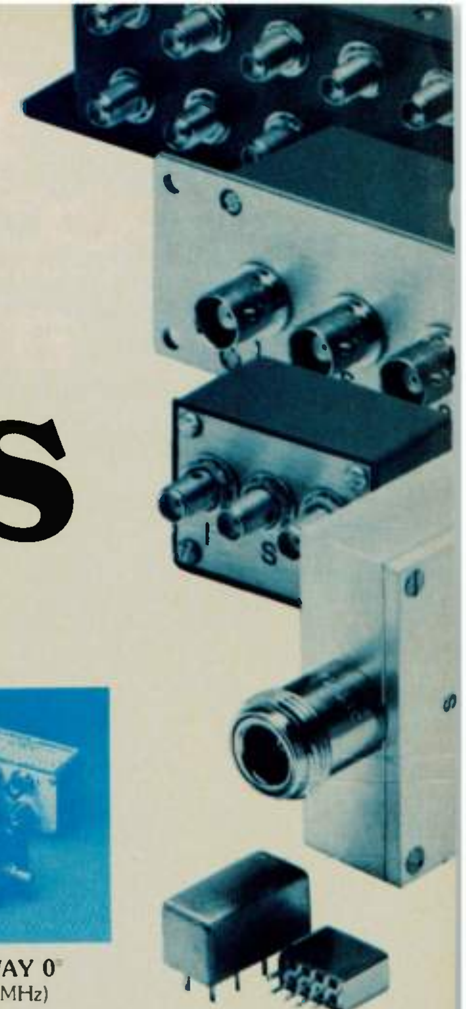
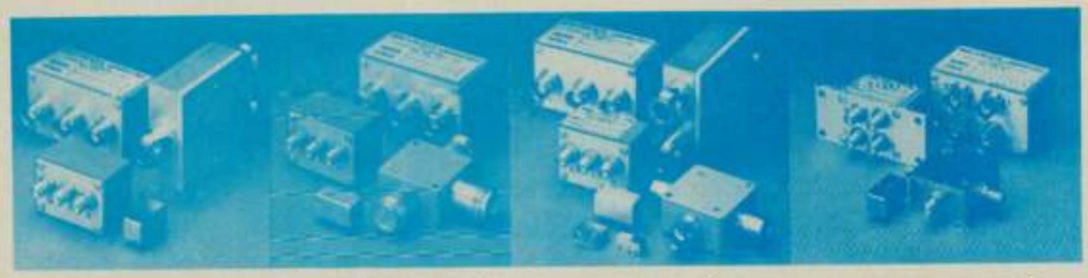


power splitters



TWO WAY 90° (1.4-4200 MHz) | TWO WAY 180° (10 KHz-500 MHz) | TWO WAY 0° (2 KHz-4200 MHz) | THREE WAY 0° (0.01-750 MHz)

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For complete specifications and performance curves refer to the Microwaves Product Data Directory, EEM, or the Gold Book.

Model	Freq. range (MHz)	Min. isol-dB (Mid-band)	Max. insert. loss-dB (Mid-band)	See notes below	Price (Qty.)
2-WAY 90°					
PSCQ2-1.5	1.4-1.7	25	0.7	2	$12.95 (5-49)
PSCQ2-3.4	3.0-3.8	25	0.7	2	$16.95 (5-49)
PSCQ2-6.4	5.8-7.0	25	0.7	2	$12.95 (5-49)
PSCQ2-7.5	7.0-8.0	25	0.7	2	$12.95 (5-49)
PSCQ2-10.5	9.0-11.0	20	0.7	2	$12.95 (5-49)
PSCQ2-13	12-14	25	0.7	2	$12.95 (5-49)
PSCQ2-14	12-16	25	0.7	2	$16.95 (5-49)
PSCQ2-21.4	20-23	25	0.7	2	$12.95 (5-49)
PSCQ2-50	25-50	20	0.7	2	$19.95 (5-49)
PSCQ-2-70	40-70	20	0.7	2	$19.95 (5-49)
PSCQ-2-90	55-90	20	0.7	2	$19.95 (5-49)
PSCQ-2-120	80-120	18	0.7	2	$19.95 (5-49)
PSCQ-2-180	120-180	15	0.7	2	$19.95 (5-49)
PSCQ-2-250	150-250	18	0.8	2	$19.95 (5-49)
PSCQ-2-400	250-400	16	0.9	2	$19.95 (5-49)
PSCQ-2-450	350-450	16	0.9	2	$19.95 (5-49)
ZSCQ-2-50	25-50	20	0.7	2.3	$39.95 (4-24)
ZSCQ-2-90	55-90	20	0.7	2.3	$39.95 (4-24)
ZSCQ-2-180	120-180	15	0.7	2.3	$39.95 (4-24)
ZMSCQ-2-50	25-50	20	0.7	2.4	$49.95 (4-24)
ZMSCQ-2-90	55-90	20	0.7	2.4	$49.95 (4-24)
ZMSCQ-2-180	120-180	15	0.7	2.4	$49.95 (4-24)
ZAPDQ-1	500-1000	20	0.9	2.13	$59.95 (1-9)
ZAPDQ-2	1000-2000	18	0.9	2.13	$59.95 (1-9)
ZAPDQ-4	2000-4200	20	0.9	2.13	$59.95 (1-9)
2-WAY 180°					
PSCJ-2-1	1-200	25	0.8		$19.95 (5-49)
PSCJ-2-2	0.01-20	25	0.5		$29.95 (5-49)
ZSCJ-2-1	1-200	25	0.8	3	$37.95 (4-24)
ZSCJ-2-2	0.01-20	25	0.5	3	$47.95 (4-24)
ZMSCJ-2-1	1-200	25	0.8	4	$47.95 (4-24)
ZMSCJ-2-2	0.01-20	25	0.5	4	$57.95 (4-24)
ZFSCJ-2-1	1-500	25	1.5	5	$49.95 (4-24)
ZFSCJ-2-3	5-300	25	1.5	5	$39.95 (4-24)

1. 75 ohms impedance
2. Average of coupled outputs less 3 dB
3. BNC connectors standard TNC available
4. SMA connectors only
5. BNC connectors standard. TNC available. SMA & Type N available at $5 additional cost
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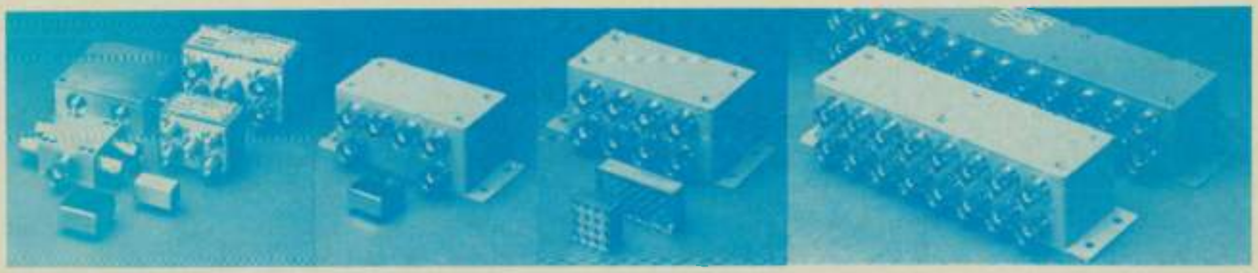
and combiners



FOUR WAY 0° (2 KHz-4200 MHz) | SIX WAY 0° (1-175 MHz) | EIGHT WAY 0° (0.01-750 MHz) | SIXTEEN WAY 0° (0.5-125 MHz) | TWENTY FOUR WAY 0° (0.2-100 MHz)

Model	Freq. range (MHz)	Min. isol-dB (Mid-band)	Max. insert. loss-dB (Mid-band)	See notes below	Price (Qty.)
2-WAY 0°					
PSC 2-1	0.1-400	20	0.75		$9.95 (6-49)
PSC 2-1W	1-650	20	0.9		$14.95 (6-49)
PSC 2-2	0.002-60	20	0.6		$19.95 (6-49)
PSC 2-1-75	0.25-300	20	0.75	1	$11.95 (6-49)
PSC 2375	55-85	25	0.5	1	$19.95 (6-24)
PSC 2-4	10-1000	20	1.2		$19.95 (6-49)
MSC 2-1	0.1-450	20	0.75		$16.95 (5-24)
MSC 2-1W	2-650	25	0.8		$17.95 (5-24)
ZSC 2-1	0.1-400	20	0.75	3	$27.95 (4-24)
ZSC 2-1-75	0.25-300	20	0.75	1.3	$29.95 (4-24)
ZSC 2-1W	1-650	20	0.8	3	$32.95 (4-24)
ZSC 2-2	0.002-60	20	0.6	3	$37.95 (4-24)
ZSC 2375	55-85	25	0.5	1.3	$37.95 (4-24)
ZMSC 2-1	0.1-400	20	0.75	4	$37.95 (4-24)
ZMSC 2-1W	1-650	20	0.8	4	$42.95 (4-24)
ZMSC 2-2	0.002-60	20	0.6	4	$47.95 (4-24)
ZFSC 2-1	5-500	20	0.6	5	$31.95 (4-24)
ZFSC 2-1-75	0.25-300	20	0.75	5	$32.95 (4-24)
ZFSC 2-1W	1-750	20	0.8	5	$35.95 (4-24)
ZFSC 2-2	10-1000	20	1.0	5	$39.95 (4-24)
ZFSC 2-4	0.2-1000	20	1.0	5	$44.95 (4-24)
ZFSC 2-5	10-1500	20	1.0	5	$49.95 (4-24)
ZFSC 2-6	0.002-60	20	0.6	5	$36.95 (4-24)
ZFSC 2-6-75	0.004-60	20	0.8	1.5	$38.95 (4-24)
ZAPD-1	500-1000	19	0.6	6	$39.95 (1-9)
ZAPD-2	1000-2000	19	0.6	6	$39.95 (1-9)
ZAPD-21	500-2000	18	0.7	6	$49.95 (1-9)
ZAPD-4	2000-4200	19	0.8	6	$39.95 (1-9)
3-WAY 0°					
PSC 3-1	1-200	25	0.7		$19.95 (5-49)
PSC 3-1W	5-500	15	1.4		$29.95 (5-49)
PSC 3-1-75	1-200	25	0.7	1	$20.95 (5-49)
PSC 3-2	0.01-30	25	0.45		$29.95 (5-49)
PSC 3-13	1-200	35	0.6		$24.95 (5-49)
ZSC 3-1	1-200	25	0.7	3	$37.95 (4-24)
ZSC 3-1-75	1-200	25	0.7	1.3	$38.95 (4-24)
ZSC 3-2	0.01-30	25	0.45	3	$47.95 (4-24)
ZSC 3-2-75	0.02-20	25	0.6	1.3	$48.95 (4-24)
ZMSC 3-1	1-200	25	0.7	4	$47.95 (4-24)
ZMSC 3-2	0.01-30	25	0.45	4	$57.95 (4-24)
ZFSC 3-1	1-500	20	0.9	5	$39.95 (4-24)
ZFSC 3-1W	2-750	20	1.0	5	$41.95 (4-24)
ZFSC 3-13	1-200	35	0.6	5	$39.95 (4-24)

Model	Freq. range (MHz)	Min. isol-dB (Mid-band)	Max. insert. loss-dB (Mid-band)	See notes below	Price (Qty.)
4-WAY 0°					
PSC 4-1	0.1-200	20	0.75		$28.95 (6-49)
PSC 4-1-75	1-200	20	0.9	1	$24.95 (6-49)
PSC 4-3	0.25-250	20	0.75		$23.95 (6-49)
PSC 4A-4	10-1000	15	1.1		$49.95 (6-49)
PSC 4-6	0.01-40	25	0.5		$29.95 (6-49)
ZSC 4-1	0.1-200	20	0.75	3	$46.95 (4-24)
ZSC 4-1-75	1-200	20	0.8	1.3	$46.95 (4-24)
ZSC 4-2	0.002-20	25	0.5	3	$69.95 (4-24)
ZSC 4-3	0.25-250	20	0.75	3	$43.95 (4-24)
ZMSC 4-1	0.1-200	20	0.75	4	$56.95 (4-24)
ZMSC 4-2	0.002-20	25	0.5	4	$79.95 (4-24)
ZMSC 4-3	0.25-250	20	0.75	4	$53.95 (4-24)
ZFSC 4-1	1-1000	18	1.5	8	$89.95 (1-4)
ZFSC 4-1W	10-500	20	1.5	8	$74.95 (1-4)
ZFSC 4375	50-90	30	1.2	1.8	$89.95 (1-4)
ZA4PD-2	1000-2000	18	1.0	14	$79.95 (1-9)
ZA4PD-4	2000-4200	18	1.0	14	$79.95 (1-9)
6-WAY 0°					
PSC 6-1	1-175	18	1.0		$68.95 (1-5)
ZFSC 6-1	1-175	20	1.2	9	$89.95 (1-4)
8-WAY 0°					
PSC 8-1	0.5-175	20	1.1		$68.95 (1-5)
PSC 8-1-75	0.5-175	20	0.8	1	$69.95 (1-5)
PSC 8A-4	5-500	18	1.8		$89.95 (1-5)
PSC 8-6	0.01-10	23	1.1		$79.95 (1-5)
ZFSC 8-1	0.5-175	20	1.1	10	$89.95 (1-4)
ZFSC 8-1-75	0.5-175	20	1.0	1.10	$90.95 (1-4)
ZFSC 8375	50-90	25	1.3	1.10	$119.95 (1-4)
ZFSC 8-4	0.5-700	20	1.5	10	$129.95 (1-4)
ZFSC 8-6	0.01-10	23	1.1	10	$109.95 (1-4)
16-WAY 0°					
ZFSC-16-1	0.5-125	18	1.6	11	$174.95 (1-4)
24-WAY 0°					
ZFSC-24-1	0.2-100	20	2.0	12	$264.95 (1-4)

7 TNC, SMA & Type N at $5 additional cost. Please specify connectors.
8 SMA connectors standard. BNC on request.
9 BNC connectors standard. TNC available. SMA available at $15 additional cost.
10 BNC connectors standard. TNC available at $10 additional cost. SMA at $25 additional cost.
11 BNC connectors standard. TNC available at $20 additional cost. SMA available at $45 additional cost.
12 BNC connectors standard. TNC available at $35 additional cost. SMA available at $65 additional cost.
13 BNC connectors (not available for ZAPDQ-4). TNC available. (SMA [3MM] and Type "N" on request. Add $5 per unit.) Please specify connectors.
14 TNC, SMA, Type N please specify connectors.

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Site: Hyatt Regency at Reunion
Dates: June 15-17, 1982

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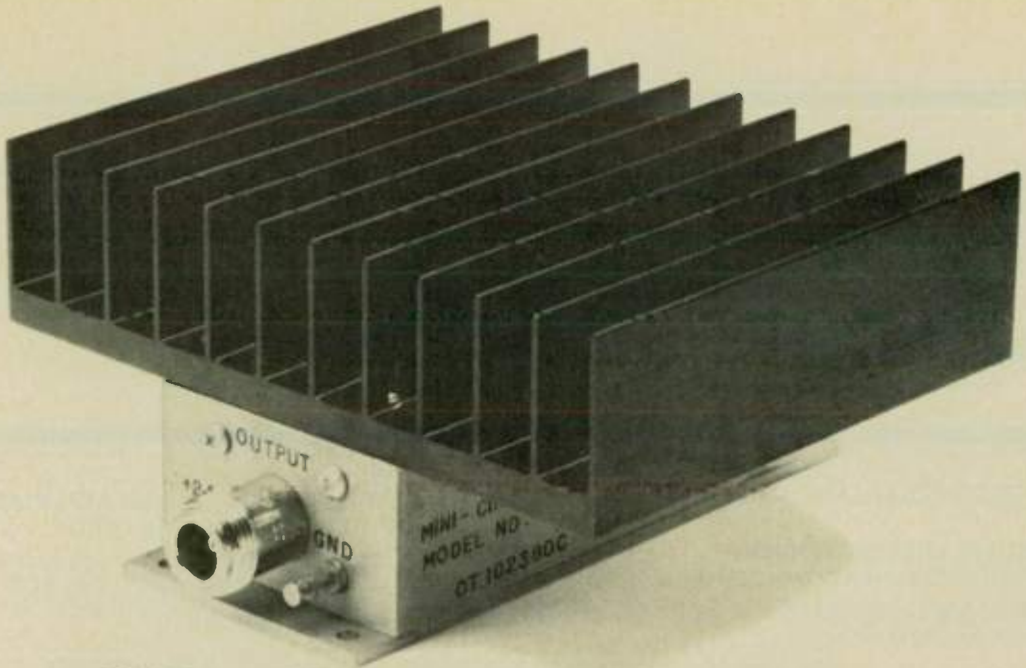

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1 Watt and now. . . 2 Watts linear output **from 50KHz to 1200 MHz** from $199

If your application requires up to 2 watts for intermodulation testing of components... broadband isolation... flat gain over a wide bandwidth... or much higher output from your frequency synthesizer or signal/sweep generator... Mini-Circuits' ZHL power amplifiers will meet your needs, at surprisingly low prices. Seven models are available, offering a selection of bandwidth and gain.

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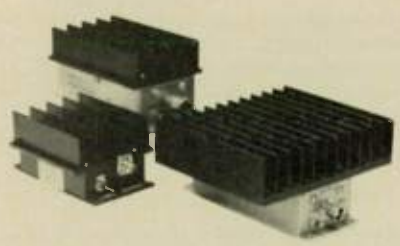

Of course, our one-year guarantee applies to each amplifier.

So from the table below, select the ZHL model for your particular application ...we'll ship within one week!

* Model No.	Freq. MHz	Gain dB	Gain Flatness dB	Max. Power Output dBm 1-dB Compression	Noise Figure dB	Intercept Point 3rd Order dBm	DC Power Voltage	DC Power Current	Price $ Ea.	Price Qty.
ZHL-32A	0.05-130	25 Min.	±1.0 Max.	+29 Min.	10 Typ.	+38 Typ.	+24V	0.6A	199.00	(1-9)
ZHL-3A	0.4-150	24 Min.	±1.0 Max.	+29.5 Min.	11 Typ.	+38 Typ.	+24V	0.6A	199.00	(1-9)
ZHL-1A	2-500	16 Min.	±1.0 Max.	+28 Min.	11 Typ.	+38 Typ.	+24V	0.6A	199.00	(1-9)
ZHL-2	10-1000	15 Min.	±1.0 Max.	+29 Min.	18 Typ.	+38 Typ.	+24V	0.6A	349.00	(1-9)
ZHL-2-8	10-1000	27 Min.	±1.0 Max.	+29 Min.	10 Typ.	+38 Typ.	+24V	0.65A	449.00	(1-9)
ZHL-2-12	10-1200	24 Min.	±1.0 Max.	+29 Min.*	10 Typ.	+38 Typ.	+24V	0.75A	524.00	(1-9)
ZHL-1-2W	5-500	29 Min.	±1.0 Max.	+33 Min.	12 Typ.	+44 Typ.	+24V	0.9A	495.00	(1-9)

Total safe input power +20 dBm, operating temperature 0° C to +60° C, storage temperature −55° C to +100° C, 50 ohm impedance, input and output VSWR 2:1 max.

+28.5 dBm from 1000-1200 MHZ

For detailed specs and curves, refer to 1980/81 MicroWaves Product Data Directory, Gold Book, or EEM

* BNC connectors are supplied, however, SMA, TNC and Type N connectors are also available

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JUNE 1981

BUSINESS/SPECIAL REPORTS

TECHNICAL/APPLICATIONS SECTION

DEPARTMENTS

ON THE COVER: Radiometric image of an eastern coastline obtained by a Georgia Tech 140 GHz imaging radiometer during airborne measurements by Naval Research Laboratories. The aircraft appearing in the image was flying beneath the aircraft in which the radiometer was located. Photo courtesy of NRL, Dr. Jim Hollinger.

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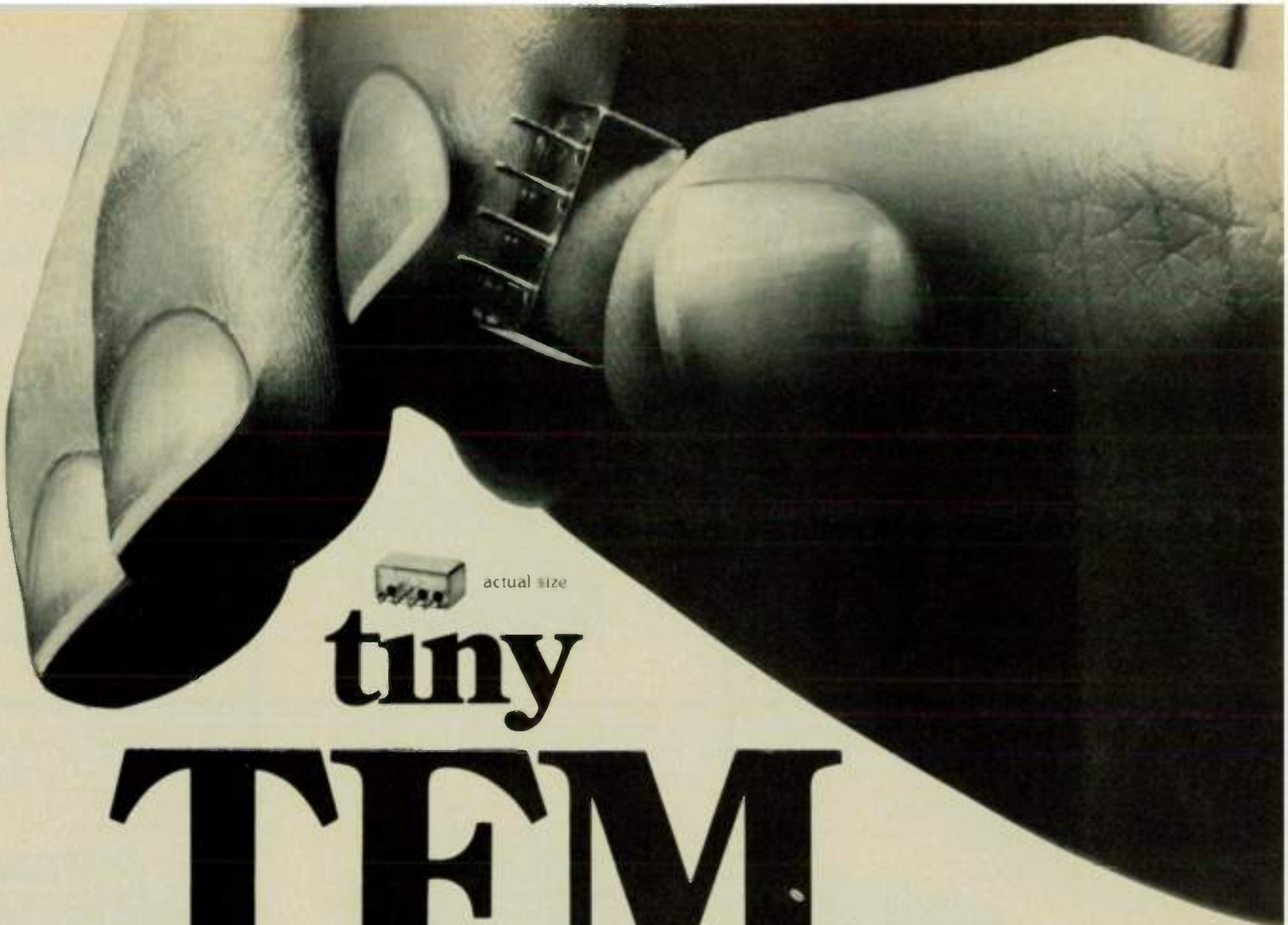

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tiny TFM

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The TFM Series from Mini-Circuits from \$11.95

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Model No.	Frequency Range MHz		Conversion Loss dB, Typical		Isolation dB, Typical						Price	
					Lower Band Edge to One Decade Higher		Mid Range		Upper Band Edge to One Octave Lower			
	LO RF	IF	One Octave from Band Edge	Total Range	LO-RF	LO-IF	LO-RF	LO-IF	LO-RF	LO-IF	\$ EA.	QTY.
TFM 2	1-1000	DC-1000	6.0	7.0	50	45	40	35	30	25	11.95	(1-49)
TFM 3	.04-400	DC-400	5.3	6.0	60	55	50	45	35	35	19.95	(5-49)
TFM 4	5-1250	DC-1250	6.0	7.5	50	45	40	35	30	25	21.95	(5-49)
•TFM 11	1-2000	5-600	7.0	7.5	50	45	35	27	25	25	39.95	(1-24)
•TFM 12	800-1250	50-90	—	6.0	35	30	35	30	35	30	39.95	(1-24)
••TFM 15	10-3000	10-800	6.3	6.5	35	30	35	30	35	30	49.95	(1-9)
••TFM 150	10-2000	DC-1000	6.0	6.5	32	33	35	30	35	30	39.95	(1-9)

•IF Port is not DC coupled
•• +10 dBm LO, +5 dBm RF at 1dB compression

For complete specifications and performance curves refer to the 1980-1981 Microwaves Product Data Directory, the Goldbook or EEM

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Coming Events

1981 PANAMANIAN AND CENT. AMERICAN CONF., IEEE JULY 9-12, 1981 Sponsor: IEEE Panama Chapter. Place: Panama. Topics: electrical and electronic issues of primary concern to Central America. Subjects such as electrical energy, telecommunications systems, data systems, marine communications, will be featured. Approximately 60 exhibiting companies are expected to participate. Contact: Ing. Antonio Raven, Apartado Postal No. 6-1997, El Dorado, Panama, Panama. Tel: 52-7763.

3RD ANNUAL SATELLITE COMMUNICATIONS USERS CONF. AUG. 19-21, 1981 Sponsor: *Satellite Communications* magazine. Place: Regency Hotel, Denver, CO. Subject: Satellite applications and technology, with presentations by experts and exhibition of hardware and services. Will hold live video-teleconference via satellite with feeds to three remote locations. Contact: Irl Marshall, SCUC '81 Conference Director, *Satellite Communications*, 3900 S. Wadsworth Blvd., Denver, CO 80235. Tel: (303) 988-4670.

EASCON '81 NOV. 16-19, 1981 Sponsor: IEEE — Wash. Sect. and Aerospace & Electronics Society, Place: Washington Hilton Hotel, Washington, DC. Subject: Government-Industry Interchange, including increased federal military budget, new aerospace system developments, and reduced regulations of communication services. The 1981 Electronics and Aerospace Systems Conference will also feature exhibition as well as technical and classified programs. Contact: Dr. Delbert D. Smith, Chrmn. EASCON '81, COMSAT General Corp., 950 L'Enfant Plaza S.W., Washington, D.C. 20024. Tel: (202) 863-6822.

VI IR AND MM WAVE CONFERENCE DEC. 7-12, 1981 Call for Papers. Sponsor: IEEE, MTT-S, Place: Carillon Hotel, Miami Beach, FL Sessions: Millimeter Sources, Devices or Systems, Mm and Sub-mm Propagation, Spectroscopy, Lasers, Imaging, etc. Submit 35-40 word abstract by June 30, 1981 to: K. J. Button, M.I.T., National Magnet Laboratory, Cambridge, MA 02139. Tel: (617) 253-5561.

The New Microprocessor-Based Scalar Network Analyzer System.

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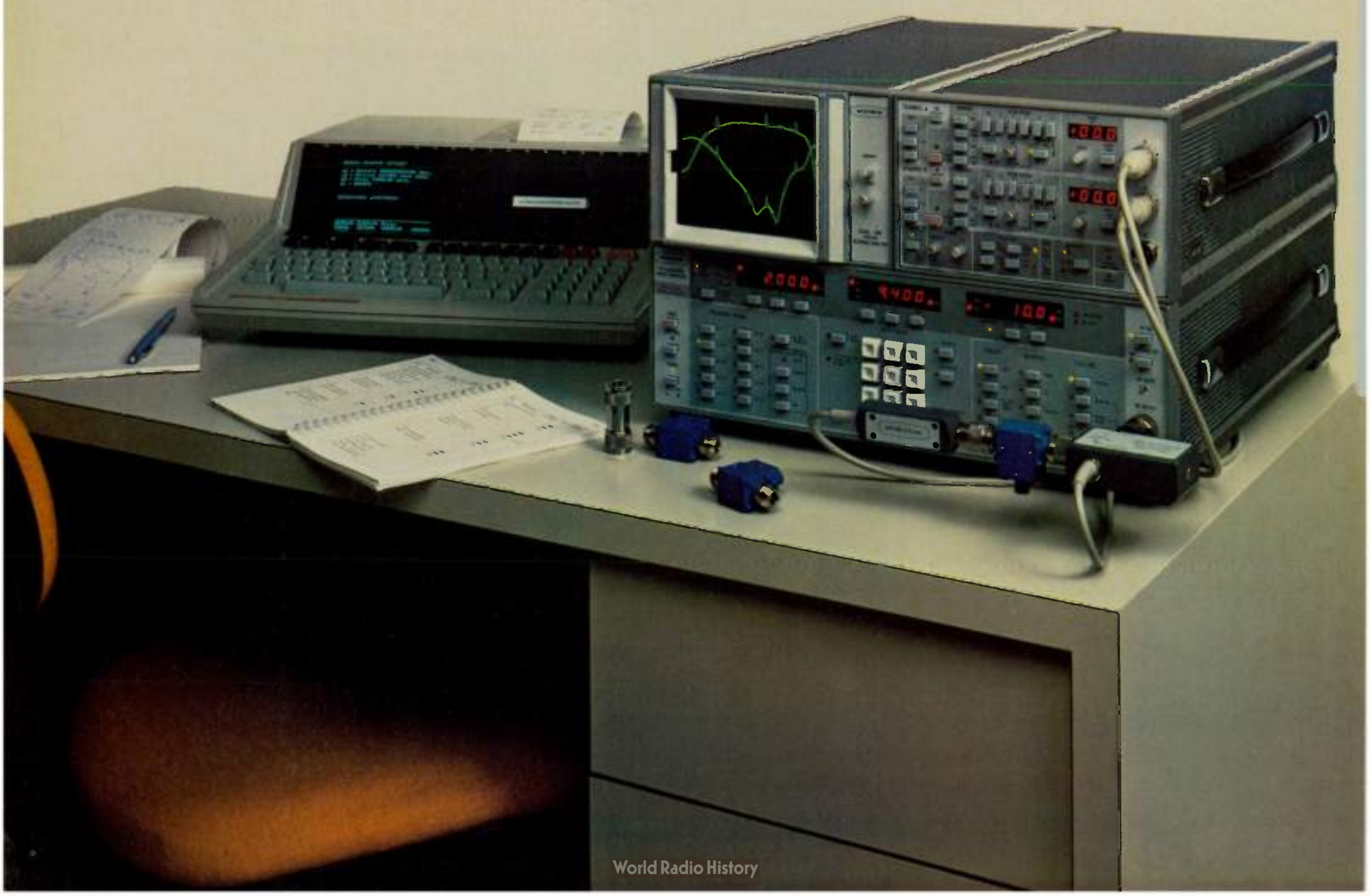

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You're going to make accurate automated microwave measurements far easier with Wiltron's new Series 5600 Automated Scalar Network Analyzer System over the 10 MHz to 40 GHz range. And you're going to make them for less.

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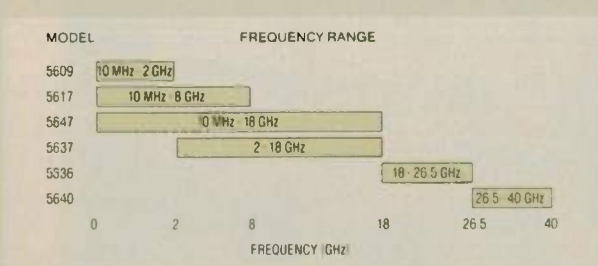
Only Three Elements to the System

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 preprogrammed 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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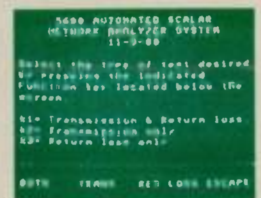
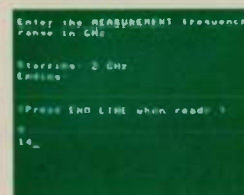
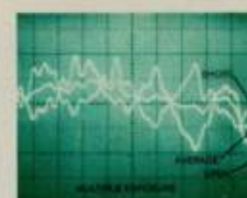


A. System Setup
Enter 1) Date
2) Type of measurement to be made



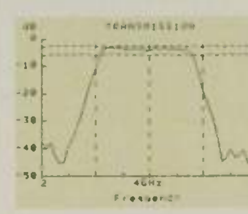
B. Frequency Selection
Enter 1) Frequency range limits
2) Frequency step size or number of test points



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
Select 1) Plotted curves
2) Tabular data

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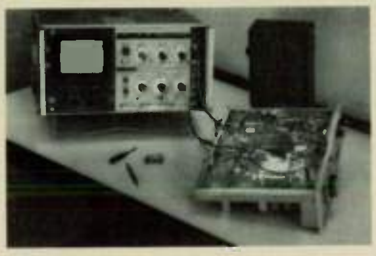
140T	Normal Persistence Display	$2250
141T	Variable Persistence/Storage Display	$3050
8552A	Economy IF Section	$3950
8552B	High Resolution IF Section	$5300
8556A	20 Hz-300 kHz RF Section	$3375
8553B	1 kHz-110 MHz RF Section	$3650
8443A	Companion Tracking Generator/Counter	$5650
8554B	100 kHz-1250 MHz RF Section	$5650
8444A	Companion Tracking Generator	$3950
8555A	10 MHz-40 GHz RF Section	$9600
8445B	10 MHz-18 GHz Automatic Preselector	$4900

20 Hz to 300 kHz



The 8556A tuner covers 20 Hz to 300 kHz and comes with a built-in tracking generator. It's calibrated for measurements in both 50 and 600 ohm systems, with accuracies better than ±1 dB. Highest resolution is 10 Hz.

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100 kHz to 1250 MHz

Use the 8554B tuning section to cover the 100 kHz to 1250 MHz range. Maximum resolution is 100 Hz. Measure with ±1¼ dB accuracy. The HP 8444A Tracking Generator (500 kHz to 1300 MHz) also works with the 8555A tuning section.

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1 kHz to 110 MHz

The 8553B takes you from 1 kHz to 110 MHz with −140 dBm sensitivity and resolution as high as 10 Hz. Signals can be measured with ±1¼ dB accuracy. Choose the companion HP 8443A Tracking Generator/Counter for wide dynamic range swept frequency measurements and precise frequency counting.

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10 MHz to 40 GHz

For 10 MHz to 40GHz, choose the 8555A. Its internal mixer covers to 18 GHz, accessory mixer for 18-40 GHz. Maximum resolution is 100 Hz. Measure with ±1¾ dB accuracy to 6 GHz, ±2¾ dB to 18 GHz. For wide scans free from unwanted response between 10 MHz and 18 GHz, add the HP 8445B Automatic Preselector.

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HEWLETT PACKARD

Publisher's Note: This month's Sum Up is provided by one of our Associate Editors, Dr. J. C. Wiltse of the Engineering Experiment Station at Georgia Tech. Dr. Wiltse was responsible for assembling the contributed articles on mm-wave subjects which are featured in the issue.

THE COMING OF mm-WAVE FORWARD LOOKING IMAGING RADIOMETERS

Passive radiometer sensors have been used from aircraft to produce images of surprisingly good quality. These systems have much better potential for penetrating cloudy, foggy, hazy weather and/or smoke than infrared sensors. Advances in hardware components have pushed the state-of-the-art to the point where minimum detectable temperatures as low as 10 millidegrees are obtainable (with a 1 second integration time) at 94 GHz. Sensitivities are not that good at 140 GHz, but as this article shows, excellent imagery is possible. Trends for future applications are also discussed.

Sum Up

SOLID STATE mm WAVE POWER SOURCES AND COMBINERS

This article is a tutorial survey of the present state-of-the-art for IMPATT and Gunn sources, power combiners, phase-locked oscillators, and FET devices. CW IMPATT's have produced ≈ 1 W in the 60 to 90 GHz range and 50 mW at 245 GHz. For pulsed IMPATT's, peak powers of 18 watts at 94 GHz and over .5 watt at 240 GHz have been obtained from individual diodes. By employing 8 diodes in a combiner, a peak power output of 63 W has been obtained at 94 GHz. This combiner has also been used as an injection-locked oscillator with 13 dB gain. Injection-locked oscillators have great importance in several applications, including wide-bandwidth amplification or obtaining high coherent output power (in conjunction with a phase-locked oscillator). At frequencies up to 94 GHz the system designer has almost a full range of possible waveform choices.

While GaAs Gunn devices have been useful up to about 94 GHz, InP Gunn oscillators which give considerably higher powers at 94 GHz and offer usable signal levels up to 140 GHz are discussed.

Another very recent and somewhat surprising development is the extension of FET devices to millimeter-wave frequencies. Oscillators have operated as high as 69 GHz, and amplifiers to as high as 38 GHz.

IC's FOR 94 GHz RADAR APPLICATIONS IN DIELECTRIC IMAGE GUIDE

One of the approaches to integrated circuits at millimeter wavelengths is the use of dielectric image guide, and this paper reports the development of 94 GHz balanced mixers and Gunn oscillators and their incorporation into several 94 GHz radar test units. A conversion loss of 11.1 dB was obtained for the balanced mixer and a power output of +5 dBm for the Gunn oscillator. Details are given for the use of these components in a short-pulsed non-coherent radar and a frequency-shift keyed radar. The image guide circuits offer potential for low cost and rugged design in quantity production.

140 GHz SILICON IMPATT POWER COMBINER DEVELOPMENT

This article complements the tutorial summary by Dr. Kuno, and gives specific results for pulsed IMPATT combiners at 140 GHz. Diode fabrication and packaging considerations and the resonant waveguide combiner are described. The equivalent circuit and theoretical considerations are explained, and the structural configuration is given for a four-diode combiner which produced 9.2 W peak power at 139 GHz. A two-diode combiner was also constructed, and it provided 5.2 W peak at 142 GHz. A discussion of modulator design is also included.

SUSPENDED SUBSTRATE K_a-BAND MULTIPLEXER

An increasing need today is the development of very wideband receivers at various frequency regions between 30 and 100 GHz. The authors' analysis covers theoretical and measured phase velocities and impedances for suspended substrate lines, and applications to 3 dB branch couplers and a four-channel multiplexer. These developments are expected to lead to a compact, wideband, channelized, down-converter front end, which could find particular use in surveillance receivers.

JAMES C. WILTSE,
Associate Editor

Workshops & Courses

ADVANCED RADAR TECHNOLOGY

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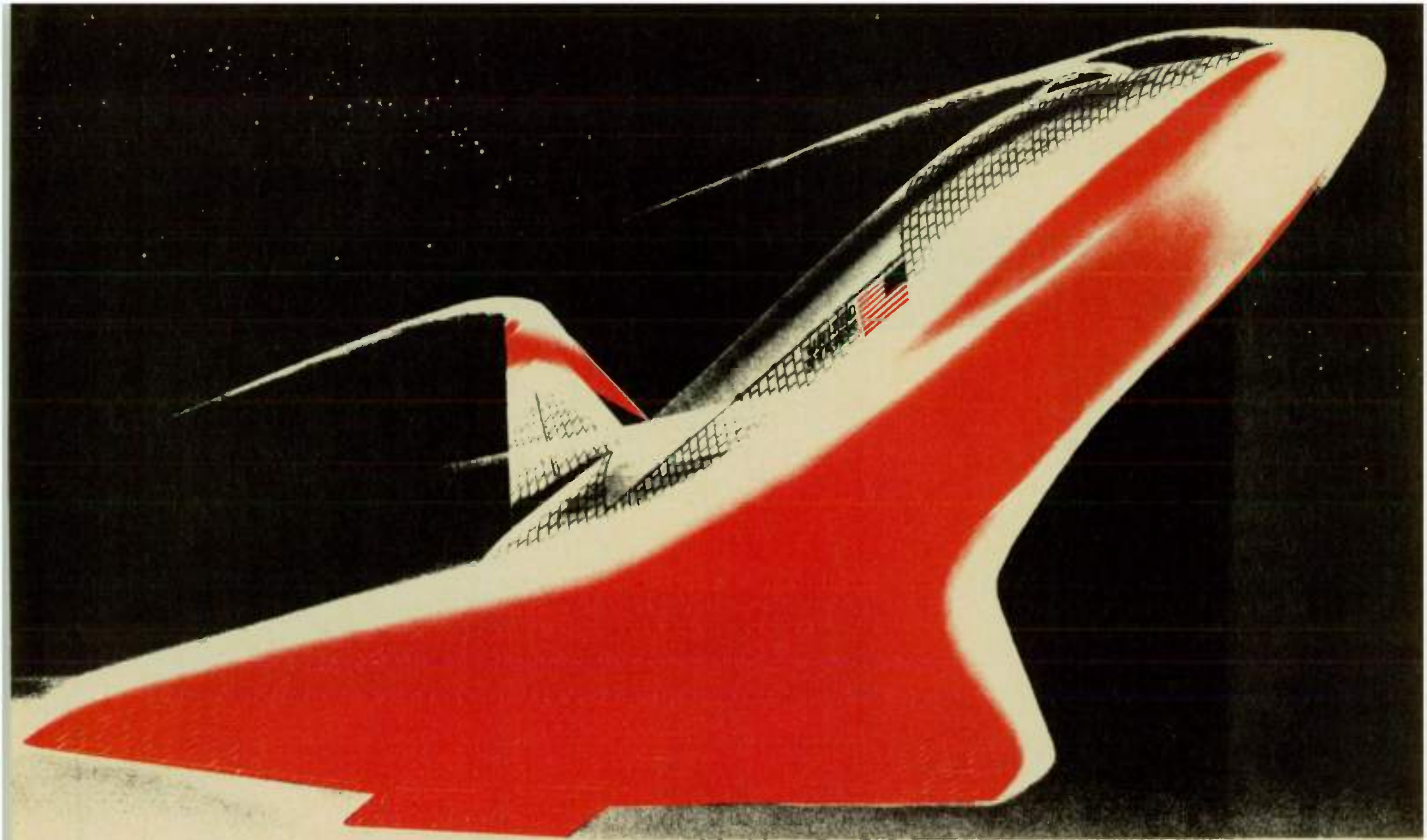

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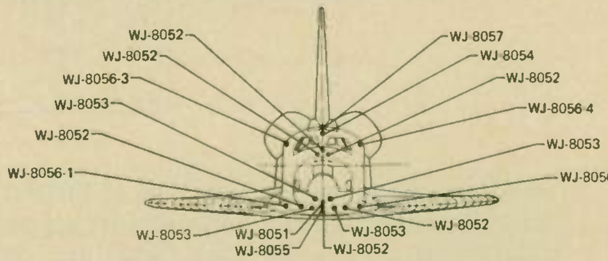

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Model Number	Frequency (MHz)	Min. Gain (dB)	Flatness (dB)	Noise Figure (dB) typ.	Noise Figure (dB) max.	Pwr. Out @ 1 dB Compression Pt. (dBm)	Case/ Connectors*
W50ETD	0.01-50	50	± .5	1.3	1.5	0	C/SMA
W50ETC	0.01-50	20	± .5	4.0	4.5	+ 23	C/SMA
W250G	5-250	43	± .5	1.3	1.5	+ 25	B/SMA
W500E	5-500	30	± .5	1.3	1.4	0	C/SMA
L60E-2	50-70	60	± .5	0.9	1.0	+ 10	C/SMA
L450E	400-500	27	± .5	1.2	1.4	+ 5	C/SMA
WIG2H	5-1000	30	± .5	1.3	1.5	+ 5	C/SMA
W2GHH2	1-2 GHz	30	± .5	2.3	2.5	+ 5	AB/SMA

Special Purpose Amplifiers

Model Number	Frequency (GHz)	Gain (dB)	Noise Figure (dB)	Pwr. Out @ 1 dB Compression Pt. (dBm)	Case/ Connectors
L13GE	1.25-1.35	25	2.2	+ 5	C/SMA
W89DGA	0.47-0.89	25	2.0	+ 5	C/SMA
L215GA	2.15-2.165	11	3.2	−3	C/N
L215GC	2.15-2.165	29	2.9	+ 7	C/N
W2GH	0.5-2.0	25	3.0	+ 10	B/SMA
P150P	0.08-150 MHz	60	1.5	+ 30	H/BNC
W15GB1	0.05-1.5	20	1.8	−3	C/SMA
W23GA	0.1-2.3	8	9.0	+ 20	C/SMA

Wide Dynamic Range Amplifiers

Model Number	Frequency (GHz)	Min. Gain (dB)	Pwr. Out @ 1 dB Compression Pt. (dBm) typ.	Pwr. Out @ 1 dB Compression Pt. (dBm) min.	Noise Figure (dB)	Case/ Connectors	Typical Intercept Pt. (dBm)
P60F	30-90 MHz	30	+ 32	+ 31	5.5	H/BNC	+ 43
P150H2	0.1-150 MHz	27	+ 31.5	+ 30	6.5	H/BNC	+ 44
P175M	150-200 MHz	23	+ 34	+ 33	8.0	H/BNC	+ 45
P400C	10-400 MHz	20	+ 31	+ 30	7.0	H/BNC	+ 42
P500N	2-500 MHz	17	+ 31	+ 30	8.0	H/BNC	+ 42
P10GL	0.5-1.0	30	+ 31	+ 30	5.0	H/SMA	+ 42
P2GS-7	0.5-2.0 GHz	30	+ 30	+ 29	10.0	FS/SMA	+ 42
P24GB	1.4-2.4	16	+ 20	+ 19	8.0	A/SMA	+ 32

CASE DIMENSIONS:
(Others Available)

	L (in.)	W (in.)	H (in.)
C	1.875	1.875	0.465
A	3.375	1.875	0.465
H	3.75	2.60	1.95
AB	3.00	1.875	0.465
B	2.625	1.875	0.465
FS	4.5	2.8	1.1

*Standard this model; others may be specified.
VSWR all models:
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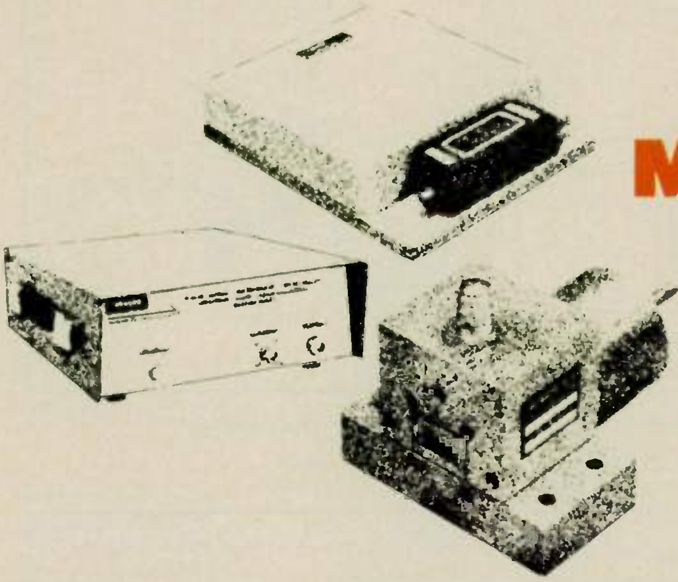

Solid State Millimeter-Wave Power Sources and Combiners

H. J. KUNO
Hughes Aircraft Company
Electron Dynamics Division
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INTRODUCTION

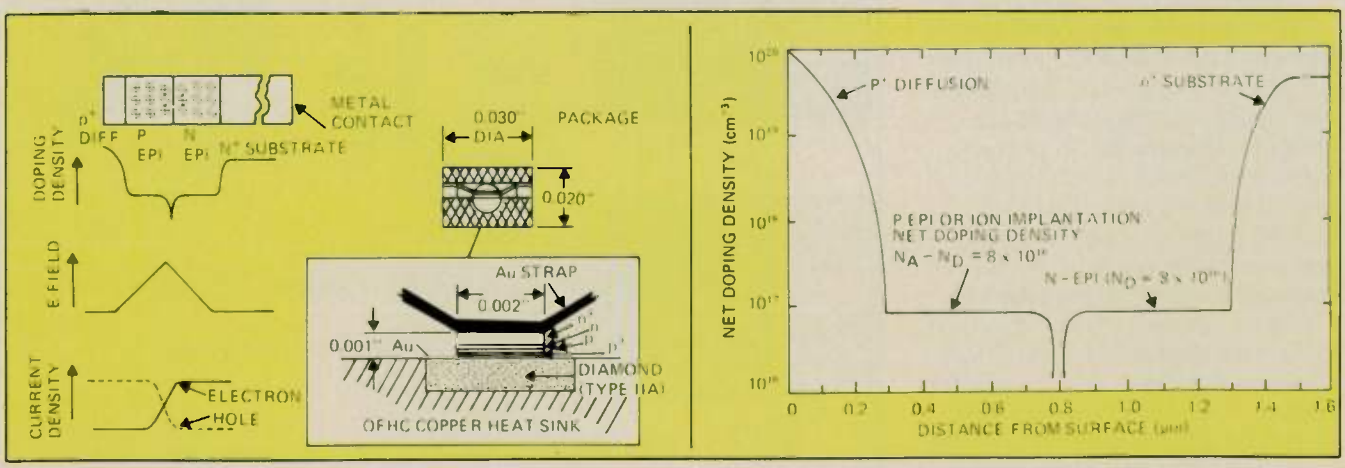
The development of millimeter-wave systems, such as fire control radars, missile seekers, projectile sensors, EW receivers, and line-of-sight communications, has become increasingly important in recent years to meet the operational requirements through smoke, dust, fog, haze, and clouds with small aperture antennas. Realization of millimeter-wave system developments can be attributed in large part to the program achieved in solid-state millimeter-wave device technology of the past several years. The system requirements in turn have stimulated further development and refinement of device technology. At the present time, solid state devices are ready for system applications at least through 100 GHz. The frequency coverage is rapidly being extended into the 140 GHz range and beyond.

Recent progress of solid state millimeter-wave power sources is reviewed. IMPATT and Gunn sources, power combiners, phase-locked oscillators, and FET devices are covered. Applications of these devices in millimeter-wave systems are discussed.

This paper presents an overview of recent developments and trends of solid state millimeter-wave power sources, that play a key role in system developments. Power generation and amplification with IMPATT, Gunn and FET devices, power combining, and noise characterization and reduction at millimeter-wave frequencies up to the 250 GHz range are covered. Applications of the devices in millimeter-wave systems are also discussed.

IMPATT OSCILLATORS

At millimeter-wave frequencies, silicon double-drift IMPATT diodes have been the most effective solid state power sources for both CW and pulsed operations. At the low end of the millimeter-wave spectrum (i.e., K_a-band),



(a) Diode structure

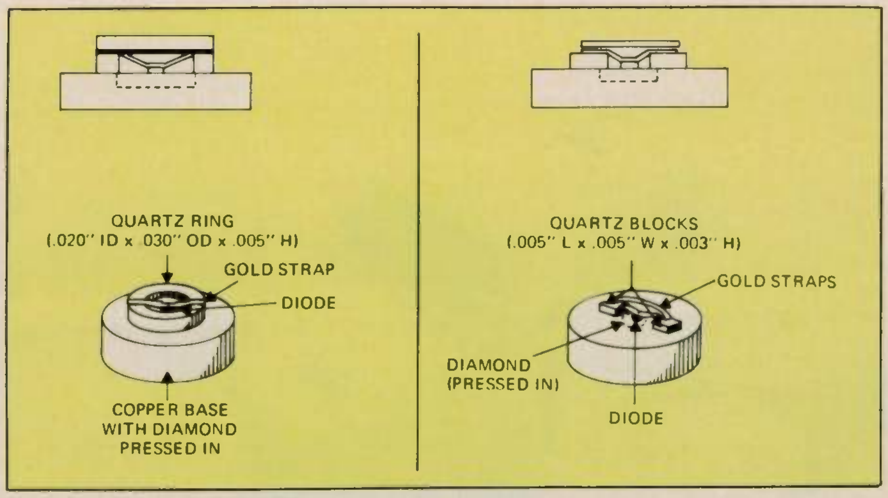
(b) Doping profile design for 60 GHz operation

Fig. 1 Millimeter-wave double drift IMPATT diode.

GaAs IMPATT diodes have shown potential for higher dc-to-RF conversion efficiencies (20%). However, the frequency coverage of the GaAs IMPATT operation has not been extended much beyond 35 GHz.

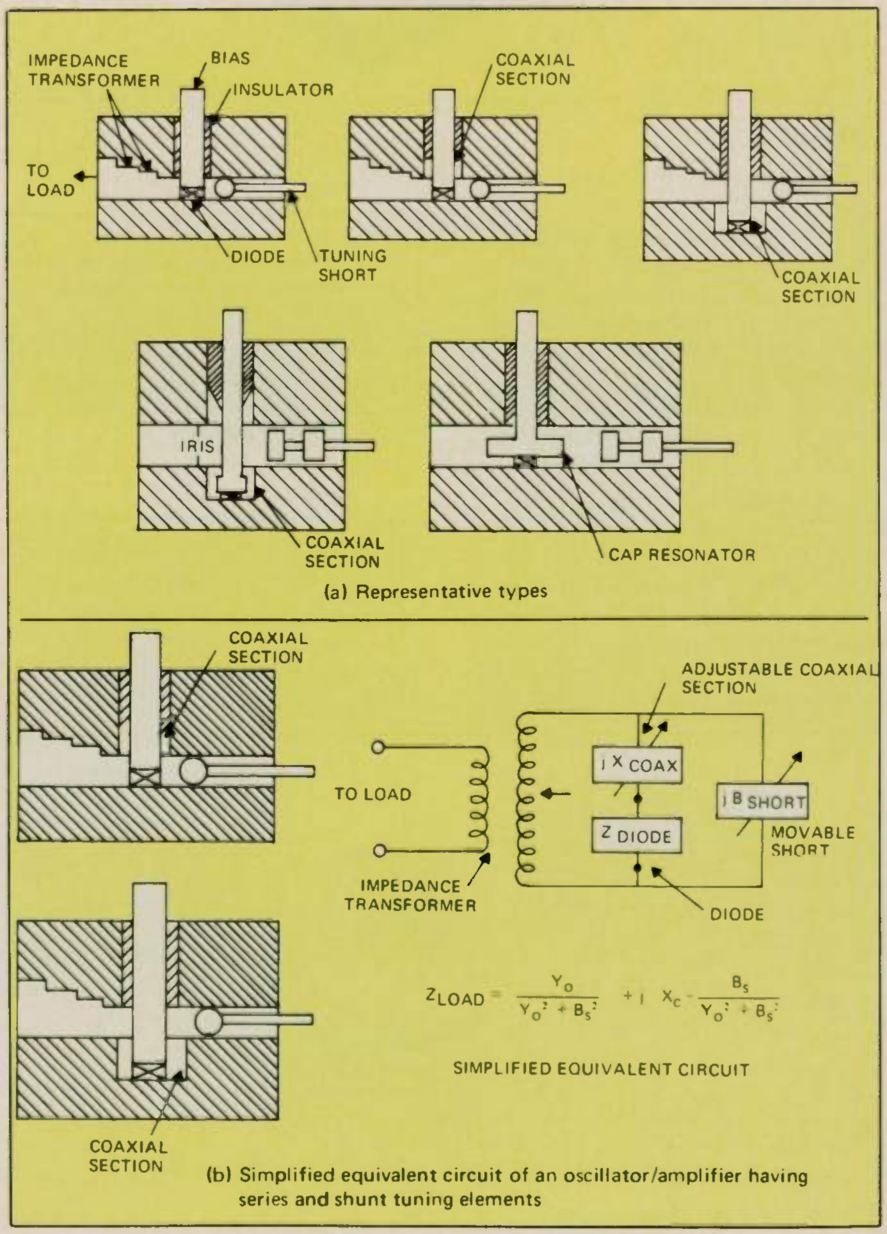
A double-drift IMPATT diode consists of a p^+-p-n-n^+ structure as depicted in **Figure 1**. In fabricating IMPATT diodes, the most critical parameter that determines RF performance is the doping density profile. As an example, a 60 GHz IMPATT diode doping density profile design is shown in the figure. Since the total active region thickness is less than 1 μm, it is evident that the diode fabrication process requires extremely fine controls. For system applications, it is desirable to mount devices in sealed packages. Millimeter-wave operation requires small parasitics. In addition, the package must provide low thermal resistance so that heat generated in the extremely small active region of the diode can be removed efficiently. Shown in **Figure 2a**, is a small package developed for millimeter-wave devices. The type IIA diamond used in the package provides an improvement in thermal resistance by a factor of 2 over OFHC copper. This package design is used for IMPATT devices up to 110 GHz. Above 110 GHz, an open type package with further reduced parasitics as shown in **Figure 2b** is used. With this type of open package, sealing must be implemented at a component or subsystem level.

Various types of oscillator circuits have been developed for millimeter-wave IMPATT oscillators. Structural features of representative design configurations are shown in **Figure 3a**. In order to achieve optimum performance at a desirable frequency from an IMPATT device, both the real and imaginary parts of the circuit impedance must be matched to those of the device. For this purpose, two degrees of tuning freedom are desirable in an oscillator circuit. The circuits with a tunable coaxial section and a tuning short provide both series and shunt tuning elements as illustrated in **Figure 3b**.



(a) Sealed package (b) Open package

Fig. 2 Package configurations for millimeter-wave devices.



(a) Representative types

(b) Simplified equivalent circuit of an oscillator/amplifier having series and shunt tuning elements

Fig. 3 Millimeter-wave IMPATT oscillator circuits.

(continued on page 24)

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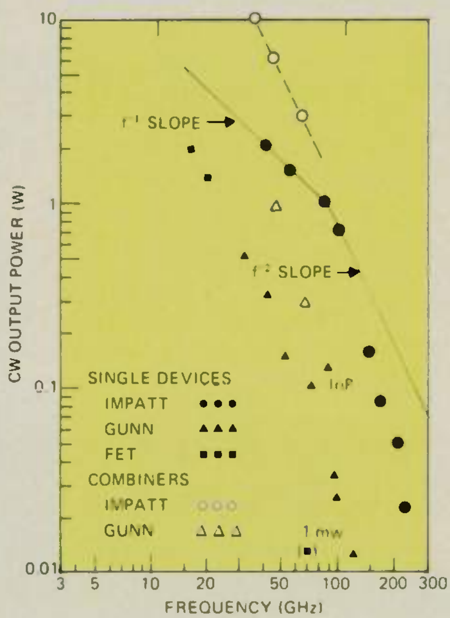
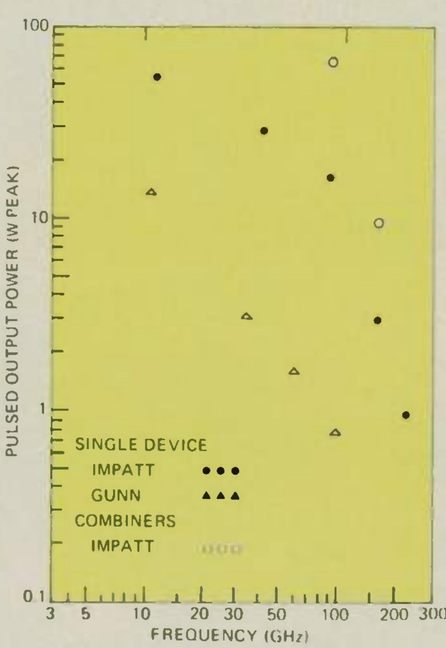
(from page 22) **SOLID STATE**

Silicon IMPATT device development efforts currently cover the frequency region from 20/30 GHz to 250 GHz. The development efforts for CW devices are centered around 20/30 GHz, 40 GHz, 60 GHz, and 220 GHz. They are aimed mainly at communication system applications. Pulsed device development efforts are aimed primarily at radar/seeker applications centered around 35 GHz, 94 GHz, 140 GHz, and 220 GHz. The present state-of-the-art of IMPATT oscillators is summarized in **Figure 4**. With silicon CW IMPATT oscillators, output power levels achieved at various frequencies are 2.3 W at 40 GHz, 1.4 W at 60 GHz, 900 mW at 94 GHz, 150 mW at 120 GHz, 130 mW at 140 GHz, 50 mW at 245 GHz, and 12 mW at 255 GHz.[1] Conversion efficiencies of these devices are less than 10%. At 35 GHz, 3 W CW output power was achieved with 20% efficiency from a GaAs IMPATT diode. However, GaAs IMPATT diodes have not yet shown much promise beyond 35 GHz.

Millimeter-wave IMPATT diodes are very well suited for high peak power and short pulse applications. Peak output powers achieved with pulsed Si IMPATT oscillators are 28 W at 35 GHz, 18 W at 94 GHz, 6.5 W at 130 GHz, 5.6 W at 140 GHz, 1 W at 220 GHz, and 620 mW at 240 GHz.[1] These results were for short pulse (50-100 ns), low duty factor (1% or less) operation. For increased pulse width or duty factor, the peak power output would decrease. Note that the data shown in **Figure 4** represent the best laboratory experimental results. For system applications, the expected power levels should be derated from those shown.



(a) CW operation



(b) Pulsed operation

Fig. 4 State-of-the-art of millimeter-wave oscillators.

GUNN OSCILLATORS

A typical millimeter-wave Gunn diode structure is illustrated in **Figure 5**. The Gunn device is made with GaAs or InP material that exhibits negative differential mobility as shown in **Figure 6**. GaAs Gunn devices have been effective in the millimeter-wave region up to 90 GHz. Beyond 90 GHz, the output power falls very sharply with increasing frequency as shown in **Figure 4**. This may be attributed to the upper cut-off frequency limitation of the transferred electron effect of GaAs. The cut-off frequency for InP has been predicted to be much higher. In addition, since the peak-to-valley ratio of InP is higher than that of GaAs as shown in **Figure 6**, InP Gunn devices should yield higher power

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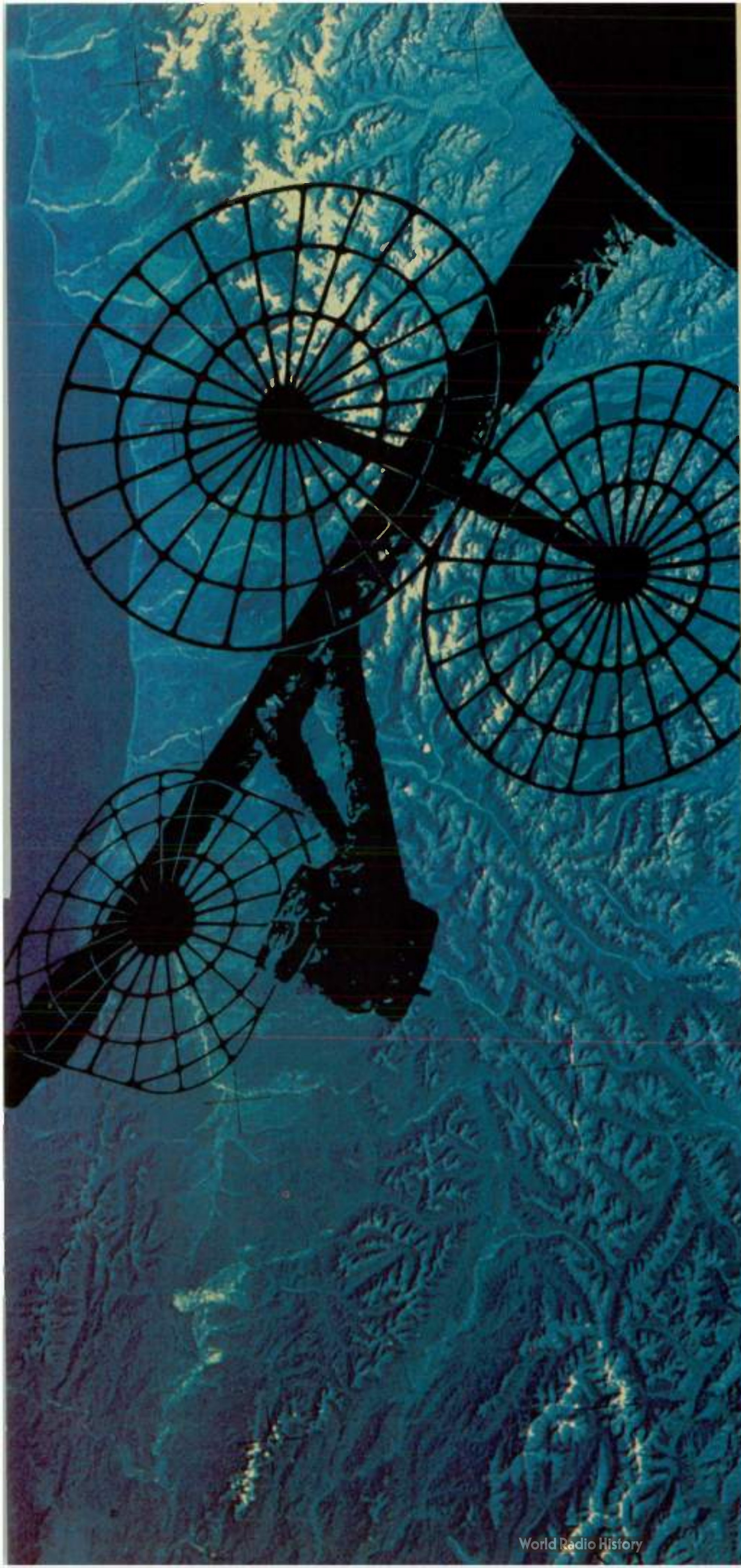

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output than GaAs Gunn devices. Recently, 126 mW CW output power at 90 GHz and peak output power of 236 mW at 90 GHz (short pulse, low duty factor operation) have been obtained from InP Gunn oscillators.[2] The state-of-the-art of Gunn oscillators is summarized in **Figure 4**. It can be seen that power levels achievable with Gunn devices are an order of magnitude lower than those with IMPATT devices. However, Gunn devices have less AM noise than IMPATT devices, particularly at modulation frequencies away from the carrier, as shown in **Figure 7**. For this reason Gunn oscillators are better suited for receiver mixer LO applications than the IMPATT oscillators, which are better for transmitter applications.

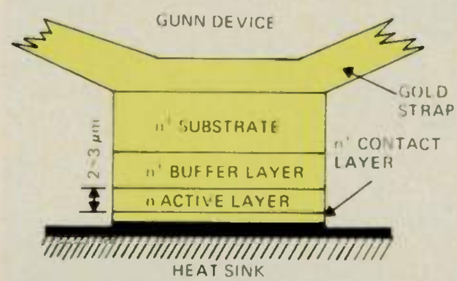

Fig. 5 Millimeter-wave Gunn device structure.

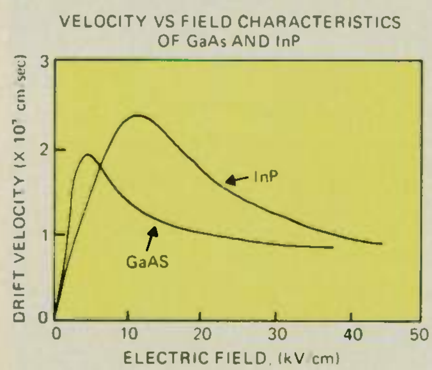

Fig. 6 Carrier velocity vs. electric field properties of GaAs and InP.

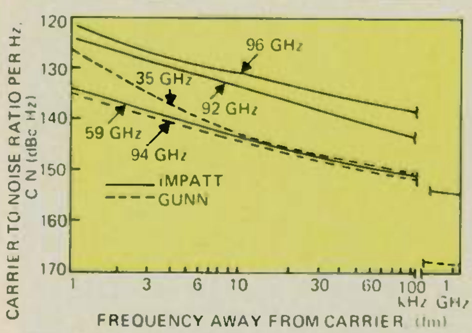

Fig. 7 AM noise characteristics of millimeter-wave IMPATT and Gunn oscillators.

POWER AMPLIFIERS

The requirements for solid state power amplifiers for millimeter-wave coherent radar and communication system applications have increased rapidly in recent years. Both IMPATT and Gunn devices can be used as millimeter-wave amplifiers. However, IMPATT diodes are better suited as power amplifiers. Pulsed or CW power amplification can be achieved either with a stabilized amplifier or an injection-locked oscillator (ILO). Generally speaking, a stabilized amplifier provides a wider bandwidth (greater than 1 GHz), a lower gain, lower output power, and better linearity; while an injection locked oscillator is better suited for a narrow bandwidth, (less than 1 GHz), higher gain, higher output power application.[3] However, locking bandwidths greater than 2 GHz at 40 GHz with 10-16 dB gain have been achieved with IMPATT ILO's. In addition, injection-locked oscillators are easier to implement than stabilized amplifiers at millimeter waves.

For those reasons, a great deal of effort has recently been directed toward the development of injection-locked oscillators for both CW and pulsed coherent radar and communication systems in the 35 GHz, 40 GHz, 60 GHz, 94 GHz, and 140 GHz regions. Shown in **Figure 8** are locking gain bandwidth characteristics of CW and pulsed IMPATT ILO's. It can be seen that they follow the classical relationship, $\sqrt{G} \times BW \cong$ Constant. The figure of merit of an ILO is often expressed by the gain-bandwidth product. With millimeter-wave IMPATT ILO's, a gain-bandwidth product ($\sqrt{G} \times BW$) of 3-6 GHz is typical. Although stabilized single-drift IMPATT amplifiers are demonstrated the feasibility of achieving broadband amplification (as much as 8 GHz with 10 dB gain in the 60 GHz range), stabilized, high power, double-drift IMPATT amplifiers are yet to be developed.

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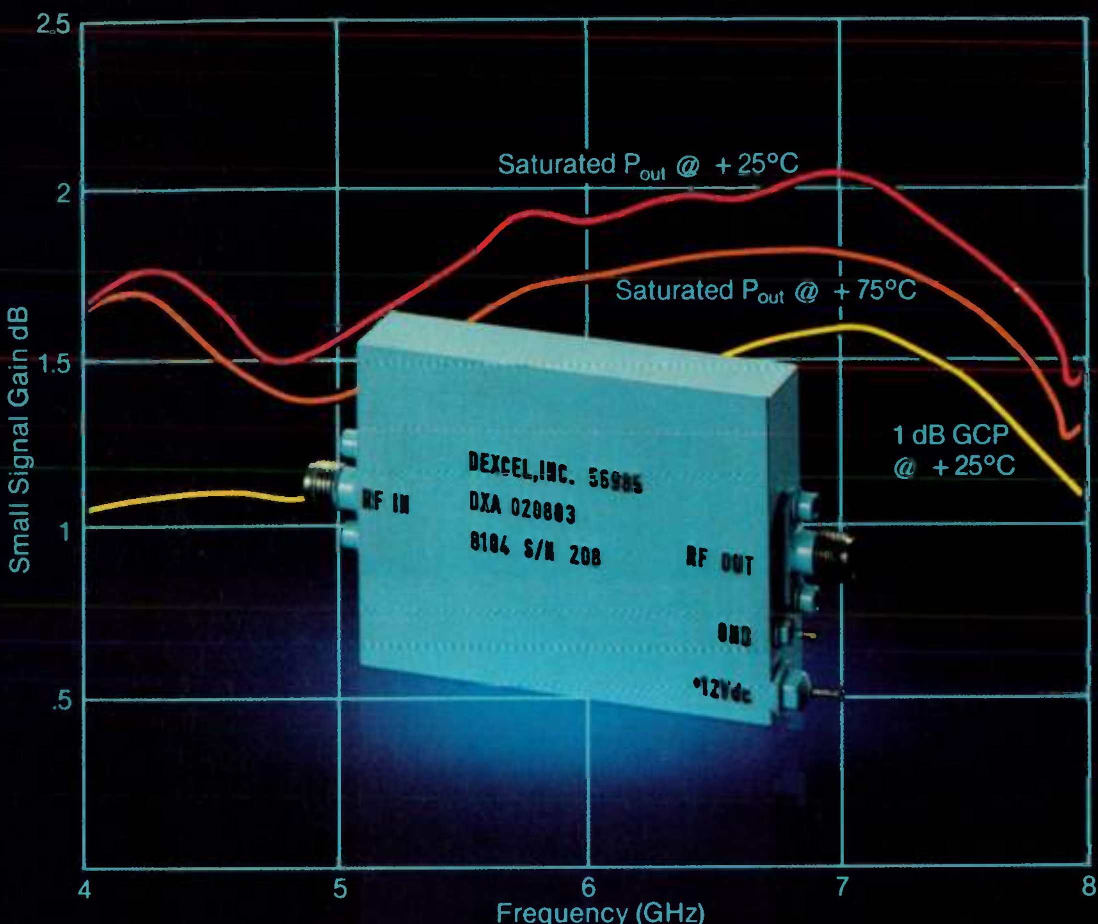

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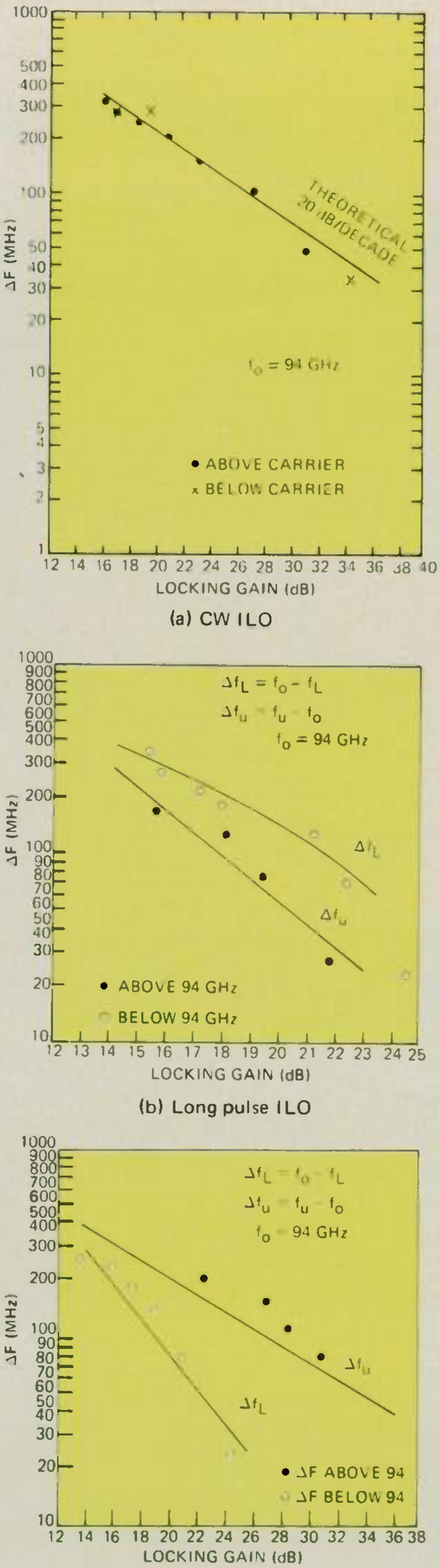

(a) CW ILO

(b) Long pulse ILO

(c) Short pulse ILO

Fig. 8 Locking gain-bandwidth characteristics of millimeter-wave IMPATT ILO's.

POWER COMBINERS

By combining a number of devices operating coherently, achievable power output can be increased greatly. There are a variety of ways of combining millimeter-wave power. Chip level combining techniques where a number of devices are connected in series and/or parallel have been developed at X-band.[4] Similar techniques and structures (as illustrated in Figure 9) are being developed to combine millimeter-wave diode chips. Status and progress of these investigations at 35 GHz will be reported at this year's International Microwave Symposium.

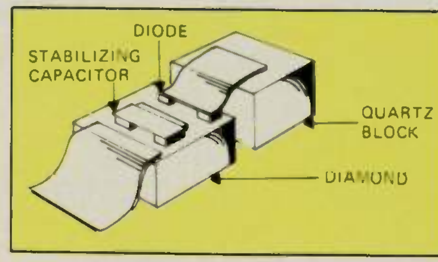

Fig. 9 Chip level combining of millimeter-wave IMPATT diodes.

Much effort has recently been expended on the development of circuit level power combining schemes, examples of which are shown in Figure 10. Multi-device resonator circuits, hybrid coupled circuits, and a combination of both techniques have successfully been used to combine as much as 8 IMPATT diodes. With 8 diodes mounted in a circular TM_{010} cavity, CW power of 10 W has been obtained at 30 GHz.[5] This circular cavity approach has not been extended beyond the K_a-band.

Modified Kurokawa waveguide[6] combiner schemes have been used at millimeter waves.* It is difficult, however, to combine more than four diodes in a single waveguide cavity. With a combination of hybrid couplers and the waveguide combiners, eight diodes have successfully been combined. With this technique, peak output power of 63 W has been achieved at 94 GHz[7] with 100 ns pulse width, 0.5% duty factor, and 76% combining efficiency. The combiner was operated as an injection-

* Specific structures and accomplishments at 140 GHz are described in an associated paper in this issue by Chang, Hayashibara, and Thrower, p. 65.

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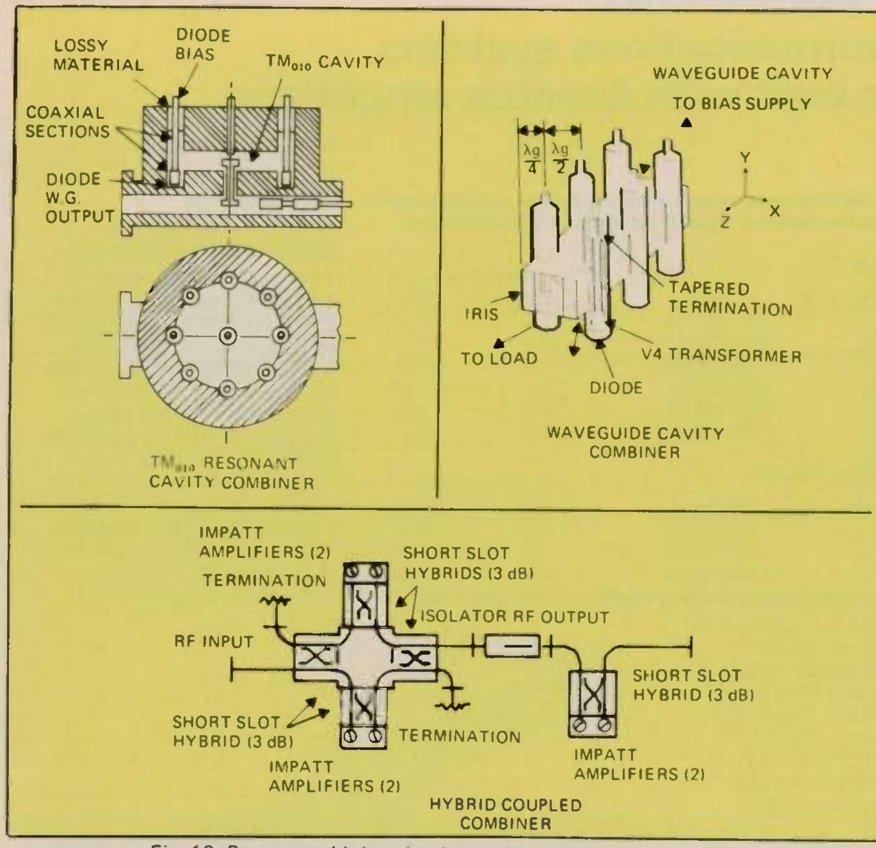


Fig. 10 Power combining circuits used at millimeter waves.

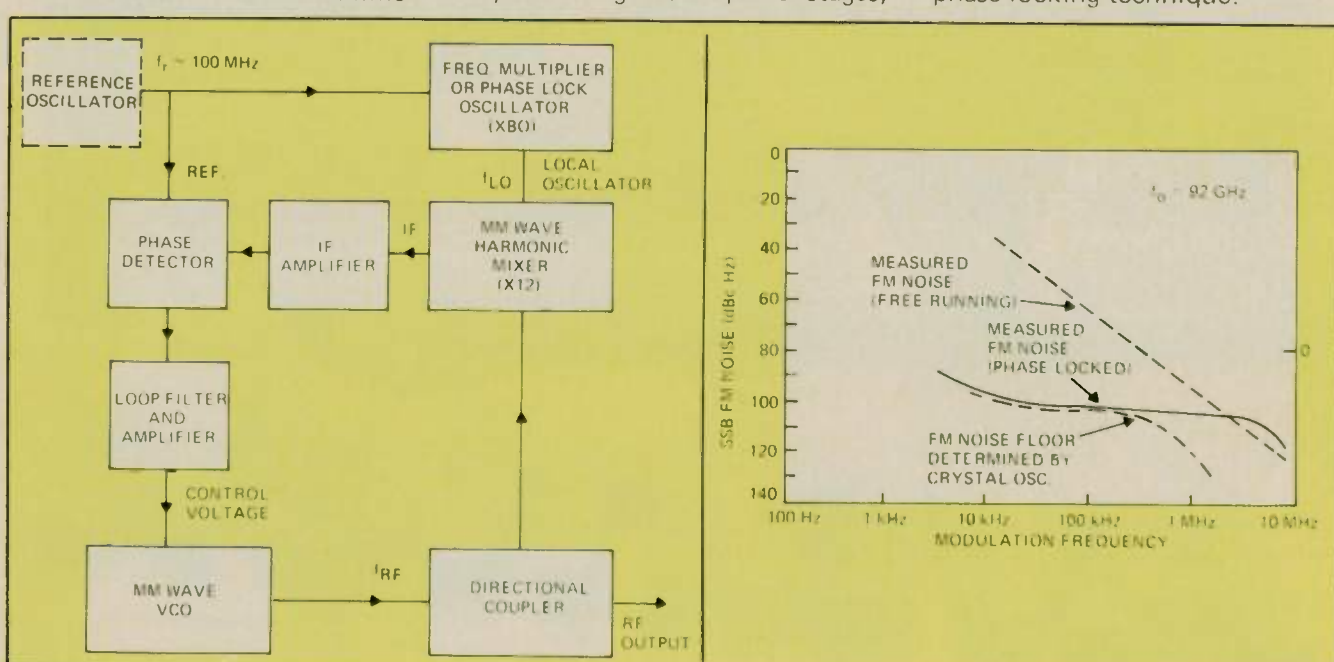
locked oscillator with a locking gain of 13 dB and locking bandwidth of 900 MHz.

FET DEVICES

Frequency coverage of GaAs FET devices has been extending into the millimeter-wave region. This newcomer into the millimeter-wave region has already shown potential as a viable millimeter-wave source. In K_a-band, amplifiers have been developed. A bandwidth of 3 GHz with 6 dB gain has been achieved in the 30 GHz region. Single sideband noise figure of 3.6 dB was measured.[8] By cascading two amplifier stages, the gain was increased to 10 dB. At 38 GHz, 8 dB gain over 1 GHz bandwidth has been achieved. These amplifiers can be used as low noise preamplifiers for millimeter-wave receivers or perhaps they may even be used as IF amplifiers for high frequency millimeter-wave broadband mixers for receiver applications. FET devices can also be operated as oscillators. Output powers of 2 W at 15 GHz and 1.25 W at 18 GHz have been reported.[9] Recently, a FET device mounted in a waveguide oscillator cavity has produced 1 mW at 69 GHz.[10] Thus FET devices have entered into the millimeter-wave region.

PHASE-LOCKED OSCILLATORS

In addition to output power level, spectral purity of signals is often important in system applications. Phase-locking techniques have recently been developed for millimeter-wave sources. With the phase-locking system as shown in **Figure 11a**, IMPATT and Gunn oscillators can be phase-locked to a crystal controlled reference oscillator. Shown in **Figure 11b** is the measured noise spectrum of a phase-locked 94 GHz Gunn oscillator. It can be seen that the spectral quality of the crystal reference oscillator can nearly be reproduced at 92 GHz by the phase-locking technique.



(a) Functional block diagram

(b) Measured spectrum

Fig. 11 92 GHz phase-locked Gunn oscillator.

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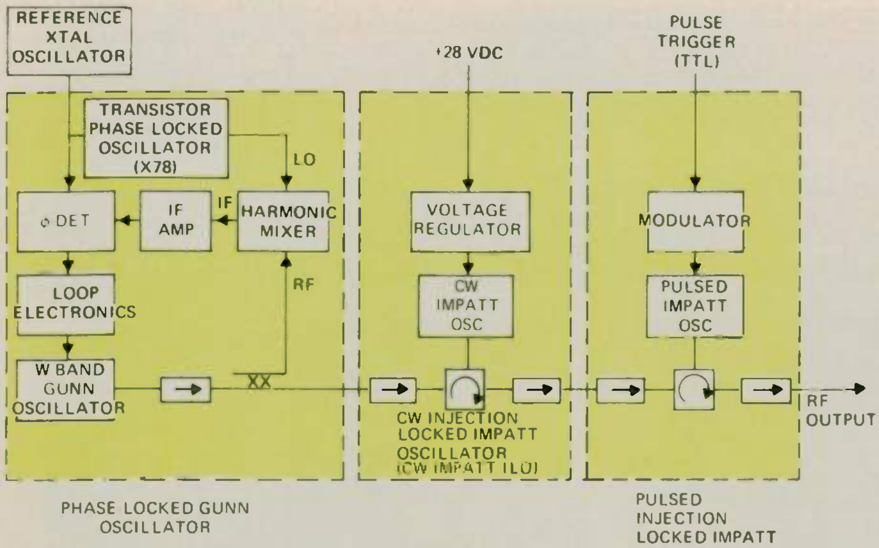
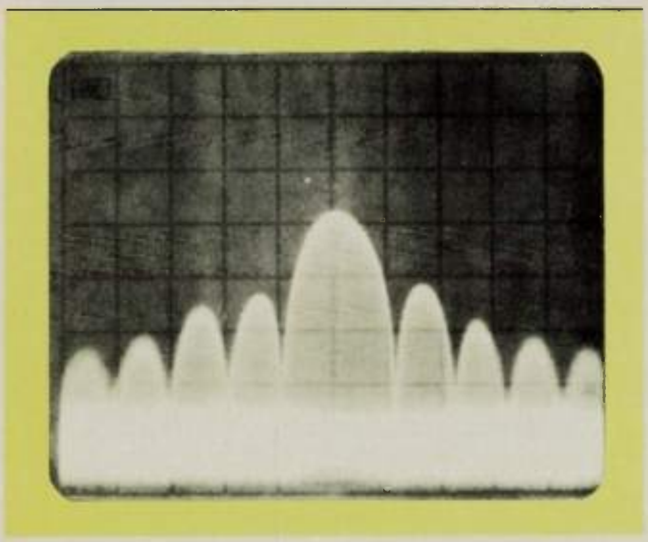


Fig. 12a 94 GHz pulsed coherent transmitter — functional block diagram.

A phase-locked oscillator can be combined with an injection-locked oscillator (either CW or pulsed). This technique (shown in Figure 12a) is useful for obtaining coherent high output power in Figure 12b shows the measured spectrum of a 94 GHz pulsed coherent transmitter output generated by this method.

CONCLUSIONS

Millimeter-wave solid state sources have certainly become ready for system applications. Depicted in Figure 13 is the relationship between the progress of sensor technology and solid state device technology. The sensor technology progress closely relates to the device technology



HORZ:	20 MHz/div.
BW:	300 kHz
VERT:	10 dB/div.
TOP:	-15 dBm
PRF:	100 kHz
PULSEWIDTH:	60 ns

Fig. 12b 94 GHz pulsed coherent transmitter — output spectrum.

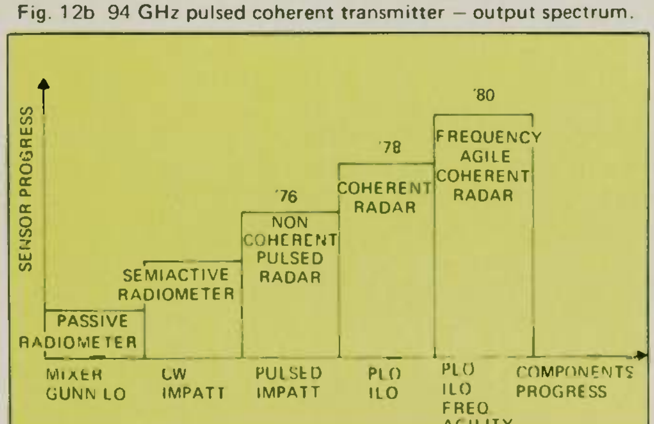


Fig. 13 Millimeter-wave sensor technology development reflects solid state components progress.

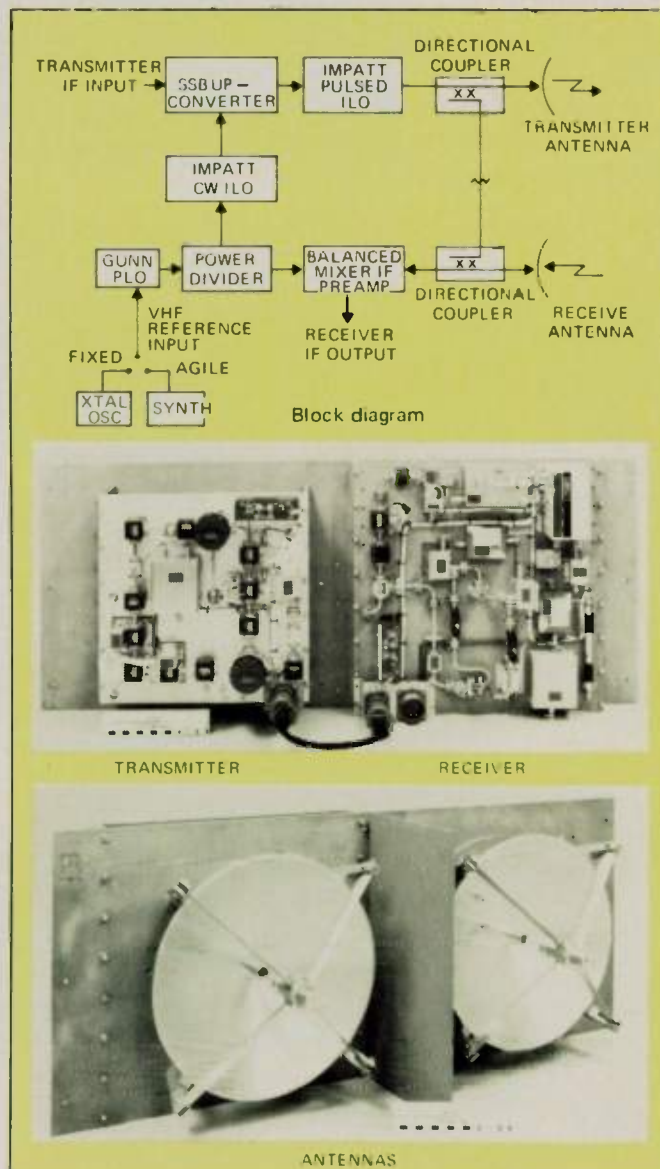


Fig. 14 94 GHz solid state coherent pulsed radar front end.

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advancements. Millimeter-wave device technology has reached the level where frequency agile coherent radars can be developed, at least through the 100 GHz range. This demarcation is moving up rapidly into the 140 GHz range. An example of a 94 GHz solid state coherent pulsed radar front end with frequency agility[12] is shown in **Figure 14**. This illustrates the progress in system development provided by solid state device technology. Immediate millimeter-wave system applications include radars/missile seekers, fuses, altimeters, projectile guidance radars, line-of-sight communication equipment, space communication equipment, EW receivers, and transponders. Solid state power sources are now ready for system applications through the 100 GHz range and the technology is rapidly being extended to 140 GHz. However, this does not imply that the technology is mature; there remains much room for further development. Power combiner development is a good example. The feasibility of achieving high power output by combining a number of devices has been demonstrated. However, the millimeter-wave power combiner development effort has barely started. Chip level or space level power combining techniques should be pursued to complement circuit level combiners. The power output levels of solid state millimeter-wave sources have been increasing at a rate of 3 dB per year for the past several years, as illustrated in **Figure 15**. This trend may be expected to continue, with power

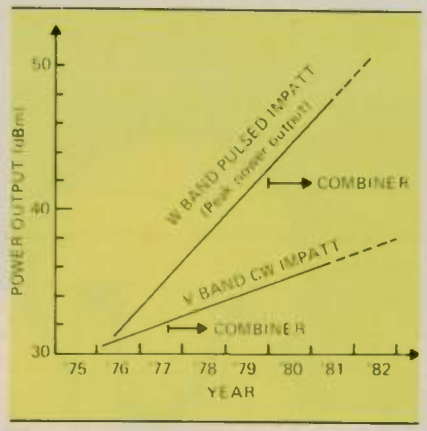

Fig. 15 Solid state millimeter-wave sources progress.

combining techniques, at least into the near future.

In addition to IMPATT and Gunn devices, the development of other power generating devices is currently under way and includes FET and TUNNETT devices, frequency multipliers, and single sideband upconverters. These devices may prove to be useful in many millimeter-wave systems. InP material may also be viable material for millimeter-wave devices. Whether or not these devices mature to be practical millimeter-wave sources, it is certain that solid state devices will play key roles in millimeter-wave systems in the 80s and beyond.

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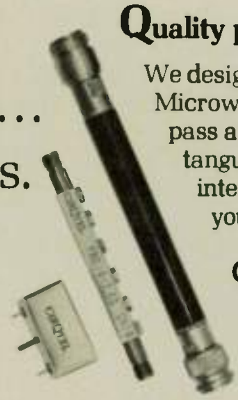
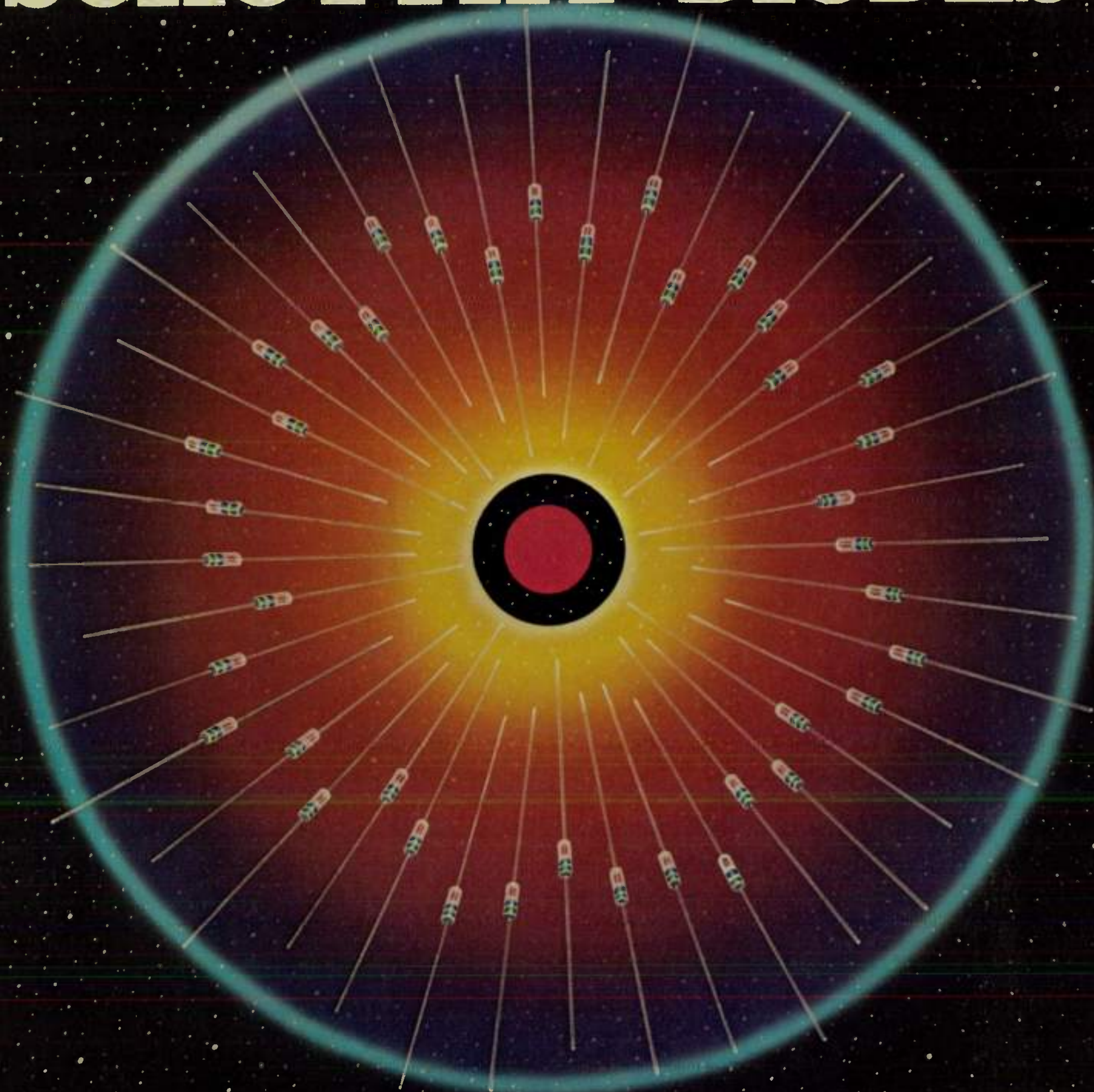
(continued on page 48)



SCHOTTKY DIODES



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ND4972-7E	ND4971-7E	5082-2811
ND4981-7E	ND4982-7E	5082-2835

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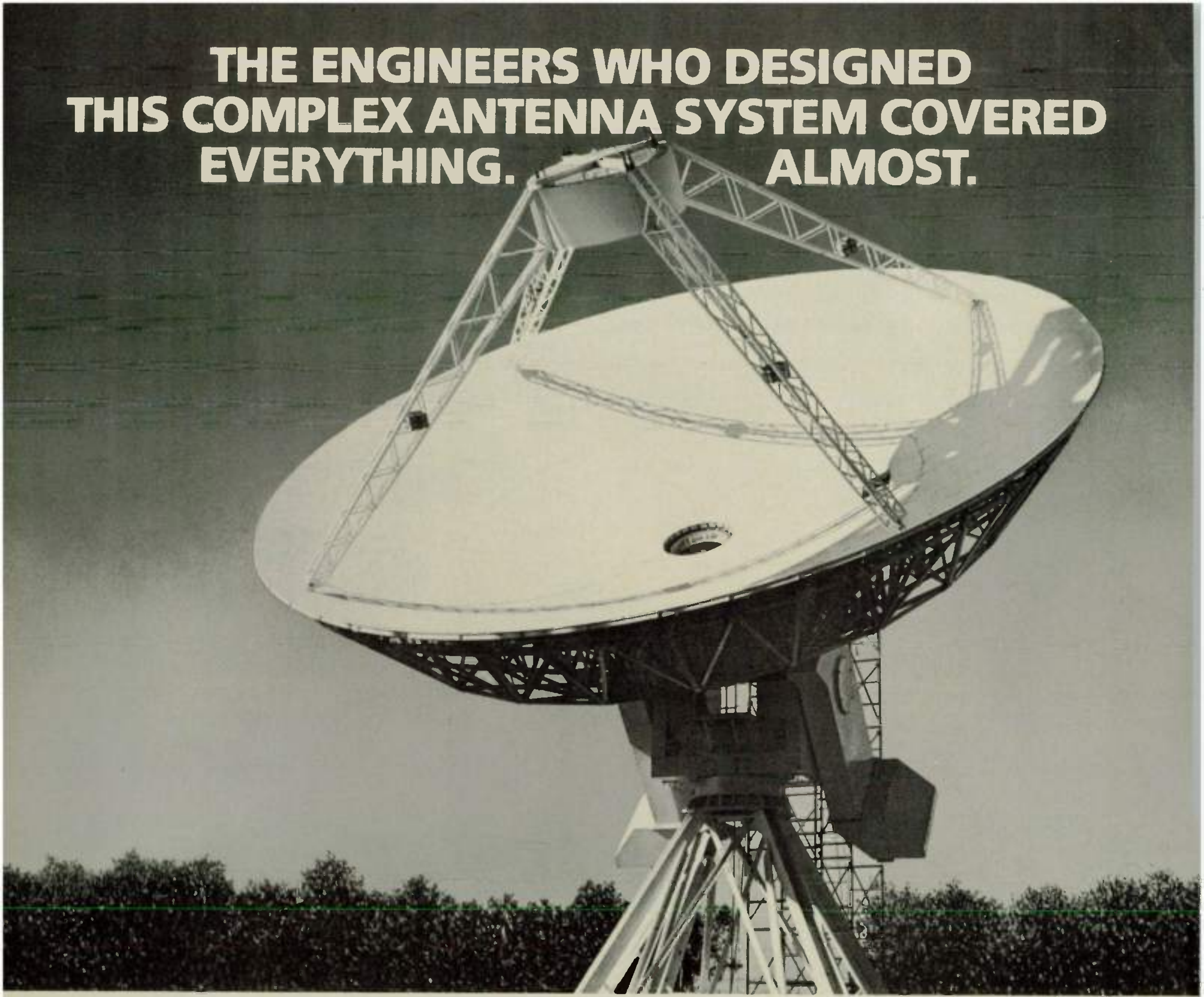

The one thing they didn't cover is the antenna system itself. Field experience has proven time and again that antennas protected by radomes will perform better over significantly longer periods of time than antennas exposed to the environment.

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CIRCLE 26 ON READER SERVICE CARD

News from ★ Washington

GERALD GREEN, Washington Editor

DoD's Research & Engineering Reorganization Outlined

A new organization of the Office of the Under Secretary of Defense for Research and Engineering has been proposed by the Pentagon's "vicar" of technology, Dr. Richard D. DeLauer, the Under Secretary for Research and Engineering ((USD (R&E)).

The reorganization, which was proposed in a recent memorandum to the Secretary of Defense, reflects a shift in priorities towards acquisition management, including a downstream change in the title of the office from USD(R&E) to Under Secretary of Defense for Research, Engineering and Acquisition (RE&A).

Also, additional emphasis will be directed towards oversight of development and support of deployed systems and equipment.

Dr. DeLauer delineates an organization headed by a three man top-management oversight team, led by himself. Specifically, the triad team would consist of the Under Secretary of Defense for Research and Engineering, a new Assistant Secretary of Defense for Development and Support, who would also be the Principal Deputy (USD), and a new Assistant Secretary of Defense for Research and Technology, who would be double-hatted as the Director of the Defense Advanced Research Projects Agency (DARPA).

Dr. James P. Wade, Jr., Acting Principal Deputy ((PDUSD (RUE)), and Assistant to SecDef for Atomic Energy, is expected to fill the dual ASD for Development and Support Principle Deputy position.

The nominee for the position of ASD for Research and Technology Director of DARPA has not yet been identified as we go to press.

Radar Could Help Protect Miners From Black Lung

Radar technology developed for government use may help to protect American miners from black lung disease.

Researchers at Georgia Tech and the National Aeronautics and Space Administration (NASA) have built a prototype radar sensor designed to make remote control of coal mining operations possible.

"The aim of the research is to remove miners from areas where coal dust is heavy and to do it without impairing the efficiency of mining," said Dennis J. Kozakoff, project director at Georgia Tech's Engineering Experiment Station (EES).

If used commercially, these radar sensing devices would be installed on mechanical shearers which remove sections of coal in mine shafts and transport them out of mines on conveyor belts. At present, human operators must walk along side these shearers to position the digger for proper cuts.

The Tech-NASA radar unit would electronically measure the distance between the cutter and the ceiling of a shaft, making it possible for an operator to control this equipment at a remote location hundreds of yards from the excavation site.

Tech has built four prototypes, using radar developed by NASA at its Marshall Space Center in Huntsville, AL. Engineers in Tech's Electromagnetics Laboratory developed the unit's antenna and a special lens which focuses the radar beam on targets several feet from the apparatus. Georgia Tech packaged the radar, antenna and lens into an explosion-proof box.

No decision has been made yet to release the radar sensor design for commercial development. However, tests undertaken recently at the Marshall Space Center were promising enough that Georgia Tech has received a contract to build five new prototypes with modification for improved performance.

NASA is the agency in charge of the coal mining radar program, and is using funds made available by the Department of Energy. Georgia Tech is working as a subcontractor in the program through Foster-Miller Associates, a Massachusetts engineering firm which is handling mining test of the radar for NASA.

Australia Selects Contractor to upgrade Defense Communications

The Australian Defence Ministry and the Department of Administrative Services have announced that Collins International Service Company (CISCO) has been selected, subject to successful completion of contract negotiations, to upgrade the Australian major defense communications links with a new network known as DISCON (Defence Integrated Secure Communications Network).

CISCO, a wholly owned subsidiary of Rockwell International Corporation, was selected to implement DISCON following competition with Ford Aerospace, Litton and Plessey. Collins Communications Systems, a division of Rockwell's Defense Electronics Operations, will be responsible for the program.

News from ★ Washington

The development of the Australia-wide integrated communications network is expected to cost between $150 million and $200 million and would extend over 10 years. It will use modern digital electronic techniques to enable the rapid and secure handling of many types of information, including voice, facsimile and data communications as well as telegraph, to interlink major national defense establishments throughout Australia and to provide a national defense communications requirement to the year 2000.

JTIDS Terminals Pass Navy Flight Tests

The Joint Tactical Information Distribution System's (JTIDS) distributed time division multiple access (DTDMA) functions implemented in command and tactical terminals have been successfully flight tested by the US Navy.

The tests involved integrating JTIDS command and tactical terminals into a hot mockup of Navy and Marine Corps command, control and communication (C^3) suites for CV aircraft carrier and E-2C and F-14 aircraft installations.

The JTIDS equipment, developed by ITT Avionics Division under contract awarded by the Naval Air Development Center (NADC), has accumulated more than 10,000 hours of ground and flight test time at the Naval Ocean Systems Center (NOSC), San Diego, CA.

Equipment tested included two AN/USQ-72 JTIDS command terminals and two AN/USQ-75 JTIDS tactical terminals, one of which was installed aboard a Navy P-3 aircraft operated by NADC, and a JTIDS RF simulator. The tests concluded at NOSC with a flight demonstration of all DTDMA functions.

Further tests with multiple platforms (both airborne and ground-based) will be conducted in the near future at the NOSC Pt. Mugu facilities.

European Military Semiconductor Market to Grow 13.6% Annually Through 1986

Average annual growth of 13.6% is forecast for the European military semiconductor market through 1986, says Frost & Sullivan, international market analysts.

Sales are projected to increase from $529.9 million in 1980 to $1.14 billion by 1986 (in constant dollars), with a total of $5.64 billion spent over the entire period, the firm reports in a new study, "Military Semiconductor Market in Europe."

Annual expenditures for discrete semiconductors are expected to build from $196 million in 1980 to $326 million by 1986, up 66% for the period. At the same time, the market for MOS digial IC's is forecast to expand by 187%, from $124 million to over $357 million. Purchases of bipolar digital IC's are seen growing 131% from $70 million to $161 million, while the market for linear IC's advances 86% from $73 million to nearly $136 million.

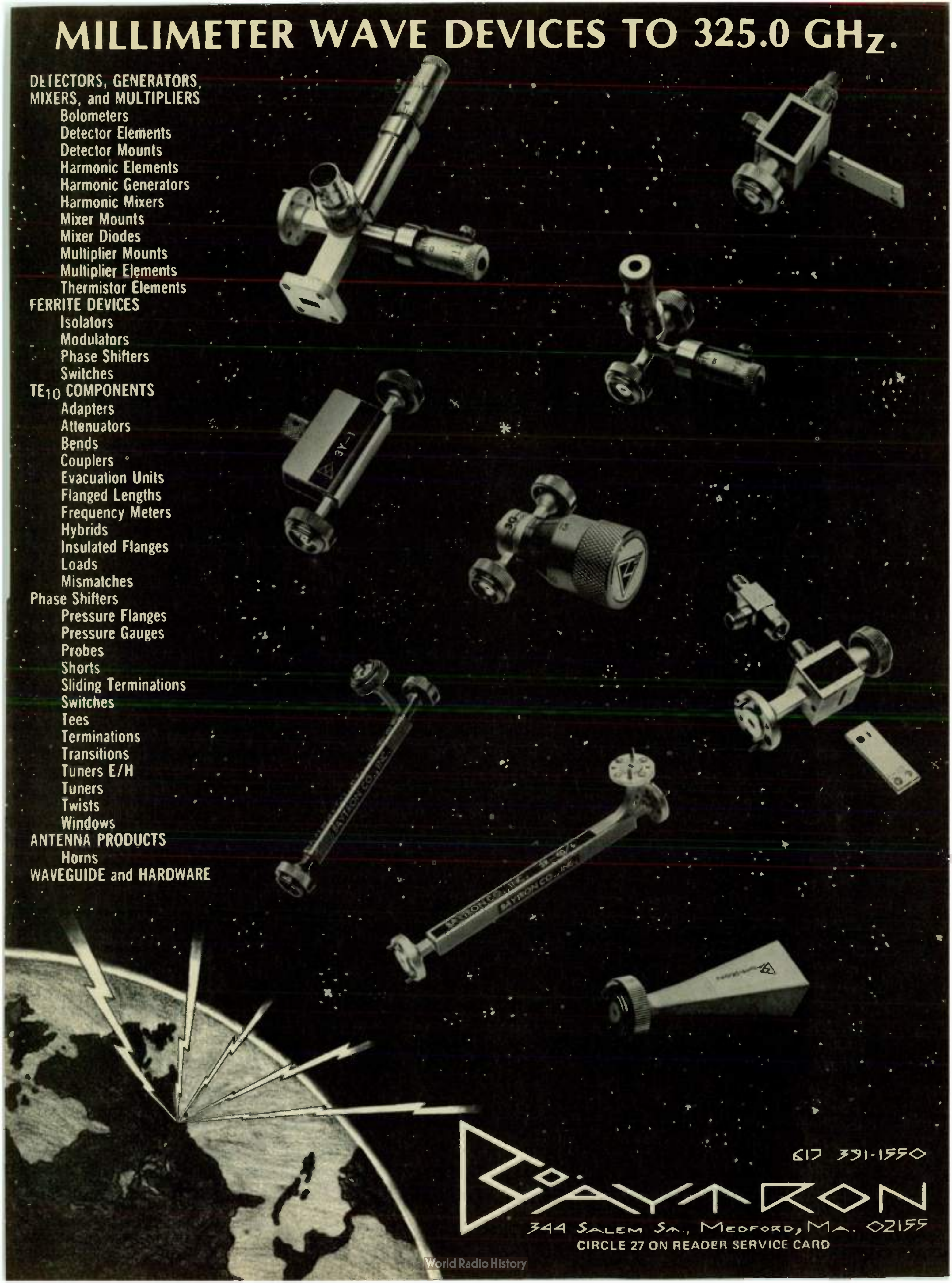
"The most important trend affecting the use of semiconductors in the European military market is the tendency for an increased proportion of the cost of a piece of equipment to be spent on electronics, and hence, on semiconductors," the market research firm observes.

F & S submits that while European defense spending is forecast to rise at an average annual rate of 3.4% during the 1980-1986 period ($99.7 billion to $120 billion), expenditures on defense electronics are seen increasing 4.3% per year ($10.9 billion to $14.2 billion). With that, the proportion of overall spending dedicated to electronics will climb from 10% to 11.8%. At the same time, semiconductors will build their share of the defense electronics market from 4.9% to 8%.

Within the semiconductor market itself, the most significant trend will be the strengthening of the relative position of integrated circuits and MOS digital IC's in particular, the study indicates. The share of all IC's in the total market is forecast to escalate from 57.6% in 1980 to 64.5% in 1986, while MOS integrated circuits share moves up from 23.5% to 31.4%.

Analyzed by nation, Europe's four major markets for military semiconductors are West Germany, France, the United Kingdom and Italy. Purchases in West Germany, France, the United Kingdom and Italy. Purchases in West Germany are forecast to increase 110.7% from $136.7 million in 1980 to $287.9 million in 1986, with France growing 122.4% ($123.4 million to $274.3 million), the United Kingdom climbing 100.6% ($102.8 million to $206.2 million) and Italy rising 127.7% ($38.8 million to $88.4 million).

Rapid growth, however, is expected to occur as well in the smaller Southern European nations of Spain and Greece. The Spanish market is seen increasing 150.2% to $51.2 million in 1986, while Greek purchases escalate 134.2% to $22.3 million.



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CIRCLE 27 ON READER SERVICE CARD

Around the Circuit

PERSONNEL

Andrew Walczak joined M/A Electronics Canada Ltd as Field Marketing Manager for Eastern Canada. . .Hughes Aircraft Co.'s Industrial Electronics Group announced these promotions: **Dr. John T. Mendel** to Group V.P., **Carl L. Flores** to Group V.P. and Manager, Connecting Devices Div., and **Dr. Arne Lavik**, Group V.P. and Manager, Electron Dynamics Div. . .**Dale L. Peterson** was promoted to Pres., Frequency Sources West Div. and **Eugene J. Veracka** was promoted to Pres., Frequency Sources East Div. . .Watkins-Johnson Co. named **Robert F. Welte** as Manager, CEI Division. . .**Ken Carr** of Microwave Associates, Inc. was appointed a Research Associate at Eastern Virginia Medical School for his work on the use of microwaves for cancer detection. . .Comstron Corp. named **Leonard J. Borow** as its President and Chief Executive Officer. . .**Nancy B. Knowlton** was promoted to position of Marketing Services Manager at Alpha Industries. . .**Dr. O. Thomas Purl** was appointed V.P., Shareowner Relations and Planning Coordination and **Dr. Williams E. Kunz** replaces Dr. Purl as V.P., Systems Group at Watkins-Johnson. . .Midwest Microwave, Inc. appointed **Walter Stockman** as Sales Manager.

CONTRACTS

American Electronic Laboratories, Inc., the **AEL Industries, Inc.** subsidiary, received a $4.78M contract from US Army's Rock Island, IL facility for the production of Vulcan Radar Systems, a gun control system used against low-flying aircraft. . .**Electromagnetic Sciences** was granted a $500K-plus contract from AT&T for the design, manufacture, and installation of the High Seas Control System, a subsystem of AT&T's High Seas Radio-Telephone Communications System. . .**Harris Corp.** received a $11.6M contract award from the US Army CORADCOM for Light Terminal Antennas and Dual Capability Servo Control Units with options for $3.8M. . . . **Itek Corp.'s Applied Technology Division** has received a $19.8M letter contract from the US Air Force's Warner Robbin's Air Logistics Center for updated AN/ALR-67/69 radar warning systems. . .**Scientific-Atlanta, Inc.** received orders valued at $1.2M for Model 3055M MARISAT satellite communications terminals from **Texaco, Inc.** and **Phillips Petroleum Co.**

NEW MARKET ENTRY

Component General Inc. of Pinellas Park, FL, announced its entry into the microwave component manufacturing field. The first of several product lines to be announced, consists of a series of roads and discs in both conventional and high power resistors designed for use in loads, coaxial attenuators, terminations, and probes from dc to 18 GHz. The standard microwave line utilizes carbon film as the resistive element; metal film is available upon request. Contact: James A. Cook, at Component General, Inc., 10460 68th St. North, Pinellas Park, FL 33565. Tel: (813) 541-7516.

FINANCIAL NEWS

Scientific-Atlanta, Inc. reported results for the third quarter ended March 31, 1981 of sales of $73.3M, net earnings of $5.2M or 25¢ per share. This compares with 1980 quarterly sales of $48.9M, net earnings of $3.5M or 17¢ per share. . .For the first quarter ended April 3, 1981, **Watkins-Johnson Co.** reported sales of $34M, net income of $1.78M and earnings per share of 56¢. In the comparable 1980 quarter, sales totaled $31.8M, net income was $1.75M or 55¢ per share. . . **EPSCO, Inc.** announced first quarter results for the period ended April 21, 1981 of net sales of $4.1M, net income of $250K or 20¢ per share. In the first quarter of 1980, net sales were $3.65M, net income totaled $227K or 24¢ per share. . .**Adams-Russell Co., Inc.** announced a cash dividend payment of 5¢ per share on May 20, 1981 to shareholders of record as of May 6, 1981. The company also reported second quarter results for the period ended April 4, 1981 of net sales of $11.5M, net income of $900K or 27¢ per share. In the comparable period of 1980, net sales totaled $8.42M, net income was $571K or 21¢ per share. . .**Electromagnetic Sciences, Inc.** announced a 10% stock dividend payable on May 26, 1981, to stockholders of record of May 14, 1981. . .**Radiation Systems, Inc.** reported nine-month earnings of $689K or 80¢ per share on sales of $6.7M for the period ended March 31, 1981. This compares with 1980 nine-months results of earnings of $574K, or 72¢ per share on sales of $4.6M. . .**Aydin Corporation** reported first quarter sales of $25M, net income of $1.7M and earnings per share of 49¢ for the period ended March 28, 1981. During the comparable 1980 quarter, sales totaled $22.2M, net income was $1.5M and earnings per share were 43¢. . . **California Microwave, Inc.** declared a 100% stock dividend to shareholders of record on May 15, 1981, with payment set for June 15, 1981. The company announced third quarter results for the period ended March 31, 1981 of sales of $15.5M, net income of $713K, 29¢ per share. During the comparable 1980 quarter, sales totaled $8.3M resulting in a loss of $958K or 47¢ loss per share. . . .**AEL Industries, Inc.** reported year-end results for the period ended February 27, 1981 of sales of $54.5M and a loss of $4.0M, $2.07 per share. During the four quarter of 1980, net sales totaled $50.9M, net loss was $177.1K or 9¢ per share. . .During the second quarter ended March 28, 1981, **M/A-COM, Inc.** generated net sales of $118.7M, net income of $8.9M or 24¢ per share. During the comparable 1980 quarter, net sales were $77.6M net income was $5.5M or 17¢ per share. . .**Sealectro Corporation** announced first quarter sales of $12.1M, net income of $1.6M or $1.07 per share for the period ended March 27, 1981. During last year's first quarter, sales totaled $10.5M, net income was $535K or 36¢ per share. . .**Sage Laboratories, Inc.** reported its nine-month results for the period ended March 28, 1981 as sales of $2.1M, net income of $216K or 49¢ per share. This compares with 1980 nine-month sales of $1.65M, net income of $204K or 48¢ per share. . .**Alpha Industries** reported 1981 fiscal results for the year ended March 31, 1981 of $29.6M, net income of $2.78M, or $1.16 per share. This compares with 1980 year-end sales of $21.8M, net income of $1.87M or 95¢ per share. . . **Narda Microwave Corporation** declared a regular quarterly cash dividend of 5¢ per share to stockholders of record of May 8, 1981 with payment slated for May 26, 1981. . .**Microdyne Corporation** declared a regular semiannual cash dividend of 3¢ per share to stockholders of record of May 15, 1981, with payment set for May 29, 1981.

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Unmatched quality. Proven reliability. Competitive price.

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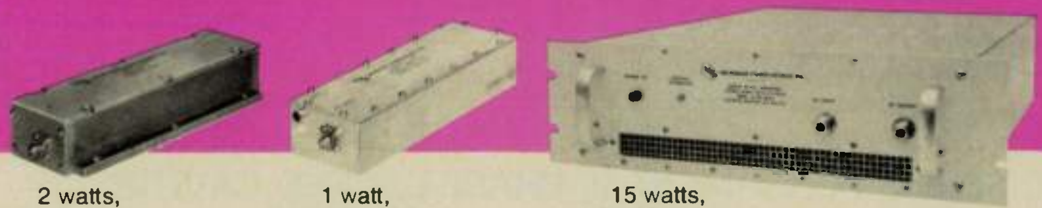

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Other power ratings up to 200 watts, frequency ranges from 1-1000 MHz to 4-8 GHz instantaneous bandwidth.

As the original pioneer and world leader in Solid State RF/Microwave Power Amplifiers, we have noted with interest the recent price and performance claims being made by some competitive companies in the field. When all is said and done, however, there emerges one inescapable fact: there is no other company—*repeat, no other company*—that can match MPD's technical design expertise, manufacturing know-how and proven track record of product reliability under actual field operating conditions!

One of our strongest product areas is Class A linear power amplifiers. Here are just a few design and performance facts that have made our Series LWA the world-wide standard of quality:

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  Exceptional linearity characteristics combined with wide dynamic range performance.
- MINIMUM DISTORTION
  Low noise figure plus high power output.
- UNCONDITIONALLY STABLE
  For any source/load VSWR.
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- GRACEFUL DEGRADATION
  Isolated circuit design.
- THERMAL DESIGN
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Our Series LWA offers more than 70 standard models, in module packages and rack-mount cabinets, including ultra-broadband frequency ranges from 1-1000 MHz up to 7900-8400 MHz, and saturated power ratings up to 200 watts.

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Model Number	Output Frequency Range (MHz)	Power Output (dBm)	Mult. Factor	Reference Frequency Range (MHz)
PLO SERIES FUNDAMENTAL PHASE-LOCKED L-C OSCILLATOR				
PLO—	270-360	+13	X3	90 -120
*Center	360-460	+13	X4	90 -115
Frequency	460-575	+13	X5	92 -115
in MHz	575-690	+13	X6	95.8-115.0
	690-805	+13	X7	98.6-115.0
* ±2% tuning	805-920	+13	X8	100.6-115.0
minimum	920-1000	+13	X9	102.2-111.1
PLC SERIES FUNDAMENTAL PHASE-LOCKED CAVITIES				
PLC-0910	0.98- 1.05	+20	X9	108.9-116.7
PLC-1012	1.05- 1.20	+20	X10	105.0-120.0
PLC-1214	1.20- 1.40	+20	X12	100.0-116.7
PLC-1315	1.30- 1.50	+20	X13	100.0-119.2
PLC-1517	1.50- 1.75	+20	X15	100.0-116.7
PLC-1720	1.70- 2.00	+20	X17	100.0-117.6
PLE SERIES FUNDAMENTAL PHASE-LOCKED CAVITIES				
PLE-2022	2.0 - 2.2	+13	X20	100.0-110.0
PLE-2225	2.2 - 2.4	+13	X22	100.0-113.6
PLE-2428	2.4 - 2.8	+13	X24	100.0-116.7
PLE-2832	2.8 - 3.2	+13	X28	100.0-114.3
PLE-3236	3.2 - 3.6	+13	X32	100.0-112.5
PLM SERIES PHASE-LOCKED CAVITY/MULTIPLIERS				
PLM-3642	3.6 - 4.2	+13	X36	100.0-116.7
PLM-4347	4.3 - 4.7	+13	X42	102.3-111.9
PLM-4853	4.8 - 5.35	+13	X48	100.0-111.5
PLM-5459	5.4 - 5.9	+13	X52	103.8-113.5
PLM-5965	5.9 - 6.5	+13	X56	105.3-116.1
PLM-6570	6.5 - 7.0	+10	X64	101.5-109.4
PLM-7075	7.0 - 7.55	+10	X70	100.0-107.9
PLM-7580	7.5 - 8.0	+10	X75	100.0-106.7
PLM-8085	8.0 - 8.5	+10	X80	100.0-106.3
PLM-8590	8.5 - 9.0	+10	X84	101.1-107.2
PLM-9095	9.0 - 9.5	+10	X90	100.0-105.6
PLM-9510	9.5 -10.0	+10	X90	105.5-111.1
PLM-100105	10-0 -10.5	+10	X100	100.0-105.0
PLM-105110	10.5 -11.0	+10	X105	100.0-104.8
PLM-110115	11.0 -11.5	+10	X105	104.7-109.5
PLM-115120	11.5 -12.0	+10	X114	100.8-105.3
PLM-120125	12.0 -12.5	+10	X120	100-0-104.2
PLM-125130	12.5 -13.0	+10	X120	104.1-108.4
PLM-130135	13.0 -13.5	+10	X126	103.1-107.2
PLM-135140	13.5 -14.0	+7	X133	101.5-105.3
PLM-140145	14.0 -14.5	+7	X140	100.0-103.6
PLM-145150	14.5 -15.0	+7	X140	103.5-107.2

FOR MECHANICALLY AGILE, SYNTHESIZED SOURCES

FREQUENCY AGILE

Model Number	Output Frequency (GHz)	Power Output Minimum (dBm)	Reference Frequency Range (MHz)	Multiplication Factor
PLA-AA SERIES				
PLA-AA-3742	3.7 -4.2	+13	123.3-140.0	X30
PLA-AA-4449	4.4 -4.9	+13	91.6-102.1	X48
PLA-AA-4853	4.8 -5.32	+13	100.0-110.8	X48
PLA-AA-6570	6.55-7.05	+10	109.1-117.5	X60
PLA-AA-7075	7.0 -7.55	+10	116.6-125.8	X60
PLA-AA-7277	7.2 -7.7	+10	120.0-128.3	X60
PLA-FA SERIES, FAST SWITCHING (1 ms)				
PLA-FA-3742	3.7 -4.2	+13	102.7-116.7	X36
PLA-FA-4449	4.4 -4.9	+13	91.6-102.1	X48
PLA-FA-4853	4.8 -5.32	+13	100.0-110.8	X48
PLA-FA-6570	6.55-7.05	+10	109.1-117.5	X60
PLA-FA-7075	7.0 -7.55	+10	97.2-104.9	X72
PLA-FA-7277	7.2 -7.7	+10	100.0-106.9	X72

FEATURING a thumbwheel frequency selection, low phase noise, 10 MHz frequency steps, 5 MHz input reference and options for 10 MHz and 100 MHz input reference.

SYNTHESIZED

Model Number	Output Frequency (GHz)	Power Output Minimum (dBm)
10 MHz FREQUENCY STEPS		
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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Integrated Beam Lead MIC Mixer/Preamplifiers

Available from 18 to 230 GHz in 9 Bands

Series 9600 Fundamental balanced mixer/preamplifiers are available in many popular frequency and IF amplifier combinations. State-of-the-art beam lead suspended stripline technology is used to achieve the best commercially available noise figures in receivers which have demonstrated outstanding reliability in critical missile guidance and radiometric applications. Both bipolar and GaAs FET IF amplifiers are integrated into low weight, compact assemblies directly useable in production programs. A sample of the diverse selection of available mixer/preamplifiers includes:

Model	RF Center Frequency	DSB Noise Figure	LO Drive Level	Instantaneous IF Output BW (MHz)	$ Unit Price	Delivery ARO
A 9600-9	35 GHz	3.2 dB DSB	1 mW	100-1000	3,650	Stock to 30 days
B 9600-9	44 GHz	3.5 dB DSB	1 mW	100-1000	3,950	Stock to 30 days
U 9600-9	60 GHz	4.0 dB DSB	1 mW	100-1000	4,250	Stock to 30 days
W 9600-9	94 GHz	5.0 dB DSB	1.5 mW	100-1000	5,150	Stock to 30 days
F 9600-9	140 GHz	6.0 dB DSB	3.0 mW	100-1000	7,050	Stock to 60 days

Noise figures quoted include 1.5 dB IF Noise Figure Contribution

Beam Lead MIC Full Band Switches

Available from 18 to 110 GHz in 5 Bands

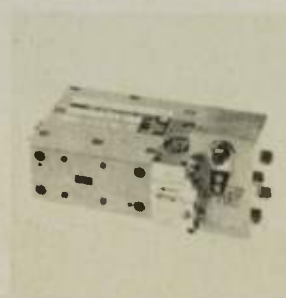
Series 979 SPST PIN switches have full waveguide instantaneous bandwidth and can be used as either switches or continuously adjustable attenuators in a variety of applications at millimeter wavelengths. These switches have lower insertion loss and higher isolation than other commercially available narrowband switches and are supplied with an integral hybrid thick film driver module for direct control from TTL logic levels.

Model	Instantaneous Frequency Range	Insertion Loss (Max.)	Isolation (Min.)	Switching Speed	Unit Price $	Delivery ARO
K 979	18-26 GHz	1.5 dB	30 dB	50 nsec	2,700	Stock to 30 days
A 979	26-40 GHz	1.5 dB	30 DB	50 nsec	3,150	Stock to 30 days
B 979	33-50 GHz	2.0 dB	25 dB	50 nsec	3,650	Stock to 45 days
V 979	50-75 GHz	2.5 dB	20 dB	50 nsec	4,850	Stock to 90 days
W 979	75-110 GHz	2.5 dB	20 dB	50 nsec	Available Early 1981	

†Includes integral thick-film driver

Beam Lead MIC Mixers With Full Waveguide Instantaneous Bandwidth

Available from 18 to 110 GHz in 5 Bands



Series 9700 full waveguide band mixers utilize extensive MIC technology to achieve flatness and state-of-the-art conversion efficiency in broadband applications at millimeter wavelength. These down converters can be supplied with integral local oscillator sources as well as a variety of GaAs FET IF amplifiers to suit system requirements. Presently available in this series are.

Model	Instantaneous Bandwidth RF	Instantaneous Bandwidth IF	LO Drive Required	SSB Conversion Loss	Flatness	$ Unit Price	Delivery ARO
K 9700	18-26 GHz	0.1-4 GHz	1.0 mW at 22 GHz	6 dB	±1 dB	3,100	Stock to 30 days
A 9700	26-40 GHz	0.1-8 GHZ	1.0 mW at 33 GHz	7 dB	±1 dB	4,200	Stock to 30 days
B 9700	33-50 GHz	0.1-8 GHz	1.5 mW at 42 GHz	7.5 dB	±1 dB	4,900	60 days

50-75 and 75-110 GHz version are currently under development

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The Coming of mm-Wave Forward Looking Imaging Radiometers

J. M. SCHUCHARDT, J. M. NEWTON, T. P. MORTON, J. A. GAGLIANO
Georgia Institute of Technology, Engineering Experiment Station, Atlanta, GA

Advances in hardware components and improved system understanding, combined with a need for operation in a variety of weather and climate situations, have come together to provide an impetus for the development of new and advanced millimeter-wave radiometer imaging systems.[1-16] Today's sensors are operating near the earth's surface, in high altitude aircraft and in both low and synchronous orbit satellites. Applications include surface sensing of geographic features, weather, status and object location. Future applications will require high resolution and rapid image generation for military, scientific, and consumer use. These latter systems will utilize multibeam sensors capable of fast beam scanning, incorporating sensitive integrated multichannel receivers.

These devices, referred to as MFLIR's, may be utilized in many applications that currently utilize infrared imaging sensors. Several of these applications are shown in Table 1. The millimeter-wave versions possess an outstanding advantage over the infrared sensor — namely, an ability to provide high quality images in cloudy, hazy, and foggy weather. Such operation is vital in many applications, as, for example, battlefield smoke and dust, adverse weather aircraft landing, reconnaissance and surveillance, vehicle navigation, and remote sensing of critical signatures.

Recent projects at several organizations have led the effort necessary to advance both the technical knowledge and techniques available for the coming generation of MFLIR's. Hardware methods have accelerated the art, permitting advanced radiometer techniques such as total power imaging for good sensitivity accompanied by periodic calibration for stability, RF noise injection for long term radiometer stability, and quasi-optical chopping for simultaneous multiband radiometric imaging using a common aperture.

Measurements conducted with surface based sensors have provided atmospheric test data including sky temperature and atmospheric losses in fog, rain, and clouds. Long and short term fluctuation data have also been gathered. Airborne sensors have provided information regarding surface features. Both water and land features have been investigated as well as manmade fixed and mobile structures.

ENVIRONMENT AND SIGNATURES

Imaging radiometer system design depends on the purpose of the system and on the radiometric properties of the environment in which the system is to be used. Radiometers find applications

TABLE 1
APPLICATIONS OF MFLIR's

PLATFORM	IMAGE/SUBJECT	REMARKS
Spacecraft	Planets, comets	Multifrequency
Satellite	Surface targets (land or sea)	Low orbit case
Aircraft		
Fixed Wing	Tactical targets, landing aid	Weapons delivery, navigation
Missile Drone	FEBA* engagements	Guidance, reconnaissance
Helicopter	Battlefield surveillance	Real time data, adverse weather and obscurant penetration
Tower	Area security, range safety	Real time data, scene/reference changes
Ship	Horizon scanning for aircraft, ships and cruise missile	High spatial resolution
Land mobile	Battlefield surveillance	Covertness, real time data
Man portable	Perimeter defense	Real time data

* Forward Edge of the Battle Area

ranging from radio astronomy to satellite-borne earth surface meteorology; to air-surface and air-air missile guidance; to battlefield target detection through smoke, fog, and other obscurants; to thermographic medical diagnostics. In each of these applications the radiometric background may differ considerably from the others. It is, therefore, necessary to radiometrically characterize each background and operating environment in order to optimally design the radiometer system for use therewith.

The apparent radiometric temperature of a surface consists of both a thermally emitted power T_ϵ and a reflected power T_ρ. The emissivity ϵ of a surface is defined as the ratio of that surface's brightness temperature to the brightness temperature of a blackbody at the same physical (kinetic) temperature ($0 \leqslant \epsilon \leqslant 1$). For an object that allows no incident RF to pass through (i.e., its transmissivity $t = 0$), $\epsilon + \rho = 1$. It is known from physical experiments that the reflectivity ρ is a function of incidence angle as well as of material type and wavelength. The emissivity ϵ is, therefore, also a function of aspect angle, and the apparent radiometric temperature of a surface also depends on aspect angle (grazing angle or depression angle).

In order to make predictions of apparent radiometric temperature for design purposes, it is helpful to use standard tabulated data gathered during measurement programs to estimate atmospheric path loss, radiometric sky temperature, radiometric temperatures of various environmental surfaces, etc. Most portions of the earth's surface (soil, grass, foliage) are quite emissive, rather than reflective; they tend to have radiometric temperatures that depend mainly on their kinetic temperatures and are relatively independent of sky temperature. Table 2 lists some typical earth surface radiometric temperatures.[17-20]

Table 3 presents average atmospheric attenuation (expressed in dB/km) as a function of frequency across the MMW (millimeter-wave) spectrum. The absorption peaks centered around 60 and 118 GHz are caused by oxygen O_2 resonant absorption. Absorption peaks around 183 and 320 GHz are due to water vapor. References 21 through 27 discuss the quantum mechanical explanations of these absorption lines.

Within the MMW region of the spectrum, there are four low-loss bands commonly referred to as transmission windows, found around the frequencies 35, 94, 140, and 220 GHz. These are the

TABLE 2

SOME TYPICAL 95 GHz RADIOMETRIC TEMPERATURES (T_{BKG}) OF COMMON BACKGROUND SURFACES UNDER CLEAR CONDITIONS AND MODERATE RAIN

SURFACE	CLEAR CONDITIONS (°K)	MODERATE RAIN (°K)
Tall Grass	250	250
Short Grass	240	240
Bare Dirt	220	230
Ice	210	220
Concrete	200	210
Gravel	200	210
Water Surface	190	200

TABLE 3

TYPICAL ATMOSPHERIC LOSS/TEMPERATURE PARAMETERS

Frequency (GHz)	Atmospheric Status	Clear Weather Attenuation (dB/km) Horizontal Path Sea Level	Horizontal Path 4 km alt.	Total Zenith	Zenith Sky Temperature (°K) Clear	Partly Cloudy	Medium Rain
35	Window	0.15	0.02	0.2	25	50	130
60	Absorption (O_2)	15.0	5.0	100+	290	290	290
95	Window	0.4	0.10	1.0	60	150	240
118	Absorption (O_2)	2.0	0.90	100+	290	290	290
140	Window	1.5	0.15	1.5	130	200	250
183	Absorption (H_2O)	30.0	3.0	100+	290	290	290
220	Window	5.5	0.20	2.6	170	260	290
320	Absorption (H_2O)	35.0	6.0	100+	290	290	290
350	Window	9.0	0.80	6.0	220	280	290

Note: Under clear air conditions standard pressure and temperature are assumed, with about 7.5 gm/m^3 of H_2O.

frequencies most commonly used by MMW radars, radiometers, and communications systems. Some short range (1-2 km) tactical communication systems use a 60 GHz RF in order to assure minimal interceptable transmission beyond the range of intended communication. However, for most purposes minimum attenuation is usually preferred.

The radiometric contrast temperature for unresolved target detection may be determined using:

$$T_c = \frac{A_T \eta \sigma (T_{BKG} - T_{TGT})}{\pi \tan^2 (\frac{1}{2}\beta) R^2 10^{0.1 \ell R}} \quad (1)$$

where

- T_c = radiometric scene/background contrast temperature (°K)
- A_T = effective target radiometric area (m^2)
- η = beam efficiency (-3 dB beam assumed, $\eta \doteq 0.6$)
- σ = target position factor ($\sigma \approx 1$)
- T_{BKG} = radiometric background temperature (°K)
- T_{TGT} = radiometric target temperature (°K)
- β = beamwidth (-3 dB beam width assumed) (degrees)
- R = range to target (R in km)
- ℓ = atmospheric attenuation (dB/km).

Another factor involved in the tradeoff for frequency selection is the beamfill factor Γ:

$$\Gamma = \frac{A_T \eta \sigma}{\pi \tan^2 (\frac{\beta}{2}) R^2}$$

in Equation (1). The larger Γ is $(0 \leqslant \Gamma \leqslant 1)$, the larger the contrast temperature T_c which may be achieved. For a given aperture and given target size, Γ increases with increasing RF frequency ($\beta \approx 70\ \lambda/D$ degrees; D = aperture diameter).

The target/background contrast ($T_{BKG} - T_{TGT}$) is also a function of frequency. When the target is a metallic object it may reflect a cold sky temperature toward the radiometer, thereby, taking on an apparent radiometric temperature T_{TGT} approximately equal to the sky temperature, and the sky temperature depends on frequency as is shown by Table 4.

The magnitude of contrast temperature measured by a 94 GHz radiometric target detector can now be determined from Equation (1). Assuming a 10 percent fill-factor ($\Gamma = 0.1$), a target background temperature ranging from 280 to 300°K, and typical clear weather values of ℓ and T_{SKY} at 94 GHz, namely 0.4 dB/km and T_{SKY} = 20 to 80°K. With these values, a contrast temperature (ΔT) ranging from 10 to 14°K could be observed from 7 km. Since typical millimeter radiometers can resolve temperature differences on the order of 1°K or less (depending on integration time), a metal target would provide a distinct signal to a passive target detection system.

TABLE 4

NOMINAL LOSS FACTORS AND SKY TEMPERATURES FOR 35, 95, 140, 220 GHz FOR VARYING ATMOSPHERIC CONDITIONS

	35 GHz			140 GHz	
	T_{SKY} (K°)	α (dB/km)		T_{SKY} (°K)	α (dB/km)
Clear	20	.02	Clear	120 (average 130)	1.4
Over-cast	50	.05	Overcast	150	1.45
Fog	80	.1	Fog	190	1.5
Light Rain	110	.25	Light Rain	220	1.5
Moderate Rain[2]	130	2	Moderate Rain	260	3.5
	95 GHz			220 GHz	
Clear	50	.4	Clear	150	4
Overcast	150	.45	Overcast	180	4.1
Fog	180	.5 - .8	Fog	200	4.5
Light Rain	210	1	Light Rain	230	4.6
Moderate Rain	240	3	Moderate Rain	270	6

Note 1: Assume visibility near 100 m and temperature of 18 to 20°C. Note 2: Light rain (< 4 mm/hr) and moderate rain (> 4 mm/hr).

Associated with the apparent target contrast temperature is the radiometer sensitivity needed to detect the available contrast temperature change. Two radiometer configurations (Dicke or total power) are commonly used. The sensitivity equation is:[28,29]

$$\Delta T_{min} = \left[\left(\frac{C\,T_{sys}}{\sqrt{B\tau}}\right)^2 + (T_y)^2 \left(\frac{\Delta G}{G}\right)^2\right]^{1/2} \qquad (2)$$

where,

ΔT_{min} = minimum detectable temperature

C = radiometer constant (2 to 2.2 for a Dicke), (1.0 for a total power)

(continued on page 50)

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(from page 48)

T_{sys} = double sideband system temperature

$= \left(\frac{T_A + T_C}{2} + T_{RCVR}\right)$ (Dicke)

$= (T_A + T_{RCVR})$ (Total Power)

T_{RCVR} = receiver noise temperature (DSB)

τ = post detection integration time

B = IF bandwidth

$\frac{\Delta G}{G}$ = amplifier stability

$T_y = (T_A - T_C)$ (Dicke)

$= T_{sys}$ (Total Power)

T_A = antenna temperature

T_C = Dicke load temperature

Typical system and component parameters permit, with careful design, the achieving of ΔT_{min} values of 0.1 to 1.0 degrees in typical imaging applications. The required ΔT_{min} for a system is based on the expected temperature change (contrast) in the scene to be imaged. Typically ΔT_{min} is 1/3 to 1/10 the value of the smallest contrast temperature within the scene.

PRACTICAL ANTENNA CONSIDERATIONS

Antenna performance is extremely critical in determining how well a radiometer can perform. Some brief considerations of antenna parameters, important for airborne applications, can be made.

The directional response of antennas varies as a function of wavelength and antenna size. For pencil beam antennas, the half-power beamwidth is approximately, $\beta = 70\lambda/D$ (degrees) where D is the antenna diameter and λ is the wavelength.

When utilized in a conventional radiometer, the antenna can resolve a target equal to the beam's projected half-power diameter at the distance of observation. This is given as:

$$\text{Resolution Spot} = \frac{\beta h \pi}{180} \text{ (meters)}$$

where h = distance in meters. Tradeoffs in receiver parameters and resolution spot size for conventional airborne mapping radiometers have been treated.[30]

A second important property of the antenna response depends on its sidelobe level, that is, the degree to which it is sensitive to radiation incident upon the antenna from directions outside the main beam. The fraction of the energy accepted from within the main beam relative to that outside of the main beam is called the beam efficiency. Low sidelobes and high beam efficiencies are obtained with tapered aperture distributions.

Sidelobe effects in a radiometer antenna can be shown by considering the basic antenna transfer function. The total signal temperature seen by the receiver is:

$$T_{RS}(\theta_o, \phi_o) = \eta_1 \eta_m \bar{T}_{ap}(\theta_o, \phi_o) + \eta_1 (1-\eta_m) \bar{T}_{s1}(\theta_o, \phi_o) + (1-\eta_1) T_o \quad (3)$$

where

θ_o, ϕ_o = direction angles of observation for the radiometer

T_{ap} = apparent temperature (average) of resolution cell in the main beam

η_1 = radiation efficiency

η_m = main beam efficiency ($\eta_m \sim 0.9$ to 0.98)

$\bar{T}_{sl}$ = average sidelobe temperature

T_o = antenna's loss mechanism physical temperature.

The last term in Equation (6) shows that a lossy antenna not only attenuates the incoming signal but adds additional noise power.

Some applications require scanned beam antennas. These antennas can be mechanically or electronically scanned. The advantages of mechanical scanning include a superior multifrequency capability, lower antenna losses and electronic simplicity. Electronically scanned antennas are more compact, need have no moving parts and can scan more rapidly.[31]

MFLIR APPLICATIONS

Airborne Mapping

The airborne mapping radiometer uses an antenna beam that is scanned through or forward of nadir across a swath beneath the aircraft (**Figure 1**) and the output signal of the radiometer is displayed or stored to present an image of the scene or terrain below. Either multiple or single beams can be used. In this application, the radiometer integration time, τ, is often equated to the dwell time (t_d) of the beam in each resolution spot. The dwell time is in turn dependent on the beamwidth, angular swath width, number of beams, the aircraft velocity, and height.

The scanning of the scene can be accomplished in different ways. The "push-broom" arrangement uses a number of beams arranged in a row normal to the flight path so that the total instantaneous field of view is fan-shaped. It is also possible to scan one or more radiometers along a line normal to the flight path. For the case of an airborne line scan mapper with multiple beams, the dwell time is:

$$t_d = \frac{n\beta}{\mathring{\Omega}} = \frac{n}{\psi}\,\frac{\beta^2}{(v/h)}$$

where

β = beam width

ψ = angular swath wide

n = number of beams employed simultaneously

v,h = velocity and altitude, respectively, of the aircraft

$\mathring{\Omega}$ = total solid angle mapped per unit time

It is assumed that there is no dead time (100% scanning efficiency) and no overlap.

Equations have been given[32] for targets located directly beneath the aircraft. For example, the difference between contrast temperature with the target in

(continued on page 52)

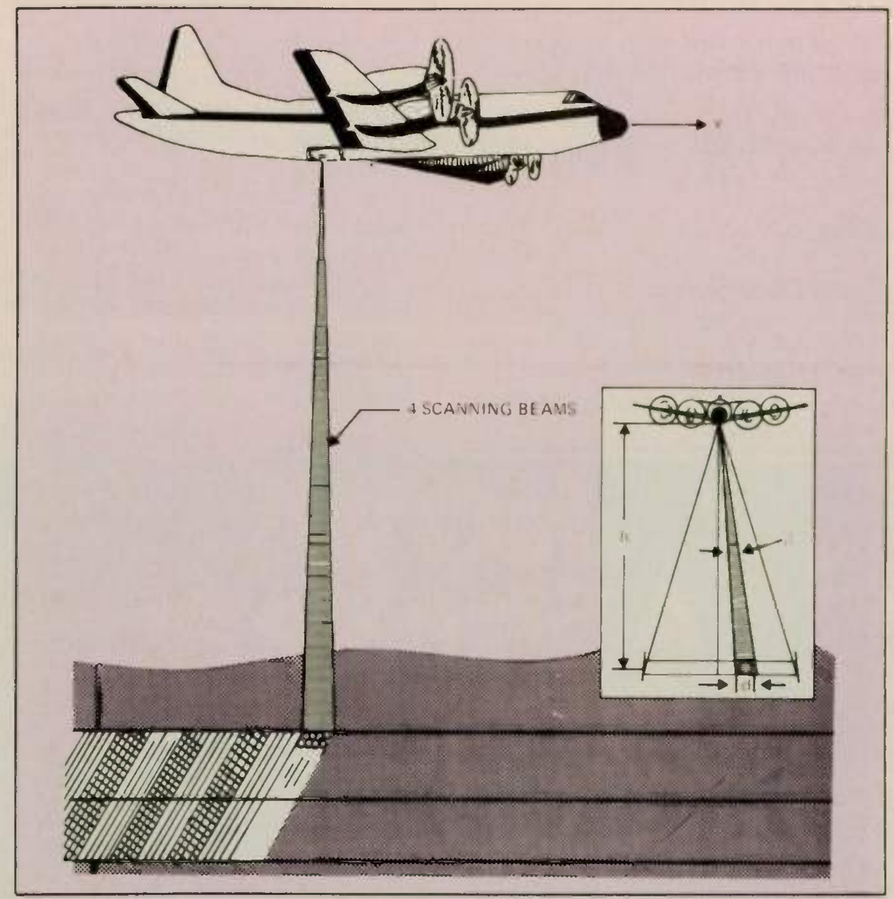

Fig. 1 Airborne imaging with an MFLIR.

beam and the target not in beam divided by ΔT_{min} is (S/N) and is given as:

$$\left(\frac{S}{N}\right) = \left[A_t(\epsilon_t - \epsilon_b)\right] \times$$

$$\times \left[\frac{1}{L}(T_b - T_{sky})\sqrt{\frac{1}{h^3 v}}\right]$$

$$\left[\frac{\sqrt{n}}{\sqrt{\psi}\,(\beta)\,\Delta T_{min_i}}\right]$$

Here ΔT_{min} is the minimum detectable temperature (with $\tau = 1$), A_t is the target area, ϵ_t and ϵ_b are the emissivities of the target and background respectively, T_b and T_{sky} are the brightness temperatures of the background and sky, and L is the atmospheric loss.

For contiguous coverage of the scene on the ground by one beam, a radiometer usually scans a portion of the scene previously scanned. The contiguous scan requirement implies that the beam be scanned in a time equal to the time required for the aircraft to

(continued on page 56)

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Comb Spacing (MHz)	500	500	950	100
Freq Stability (ppm)	30	30	30	30
Power Output (dBm)	-40 min	+6 max -22 min	-20 min	0 max -40 min
Load VSWR (max.)	1.5:1	1.5:1	1.5:1	1.5:1
In-band Spurious (dBc max.)	>24	>30	>40	>60
Power Supply	+15 Vdc at 200mA max	+15Vdc at 200mA max	+15Vdc at 150mA max	+15Vdc at 300mA max
Size (inches nominal)	1.9 x 1.33 x .5	1.33 x 1.33 x .15	1.9 x 1.33 x 0.6	1.9 x 1.33 x 0.6

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Freq. Range (GHz)	MODEL # Circulators	MODEL # Isolators	Isolation dB Min.	Insertion Loss dB Max.	VSWR Max.
1.0 - 2.0	50A1021	60A1021	16	0.6	1.35
2.0 - 4.0	50A3001	60A3001	18	0.5	1.30
2.6 - 5.2	50A3011	60A3011	18	0.5	1.30
4.0 - 8.0	50A6001	60A6001	18	0.5	1.30
5.0 - 10.0	50A6071	60A6071	18	0.5	1.30
8.0 - 12.4	10B9201	20B9201	20	0.4	1.30
8.0 - 16.0	50A2001	60A2001	17	0.5	1.35
12.0 - 18.0	10B2201	20B2201	18	0.5	1.30
8.0 - 18.0	50A2051	60A2051	16	0.6	1.50

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Model Number	Frequency Range (GHz)	Output Power (mw Minimum)	Size, inches nominal
6600 - 1300	3.65 - 4.05	30	2.7 x 1.7 x 1
6600 - 1610	4.25 - 4.75	30	2.6 x 1.5 x .9
6600 - 1611	4.75 - 5.25	30	2.5 x 1.5 x .9
6600 - 1612	5.10 - 5.60	30	2.5 x 1.5 x .9
6600 - 1613	5.40 - 5.90	30	2.5 x 1.5 x .9
6600 - 1614	5.60 - 6.10	30	2.5 x 1.5 x .9
6600 - 1615	5.70 - 6.20	30	2.5 x 1.5 x .9
6600 - 1616	5.90 - 6.40	30	2.5 x 1.5 x .9
6600 - 1617	6.40 - 6.90	30	2.5 x 1.5 x .9
6600 - 1618	7.00 - 7.45	15	2.2 x 1.3 x .9
6600 - 1910	8.00 - 8.40	15	2.2 x 1.3 x .9
6600 - 1911	8.60 - 9.00	15	2.2 x 1.3 x .9
6600 - 1912	9.00 - 9.50	15	2.2 x 1.3 x .9

FREQUENCY STABILITY: 10 ppm/°C max
5 ppm/°C. typical

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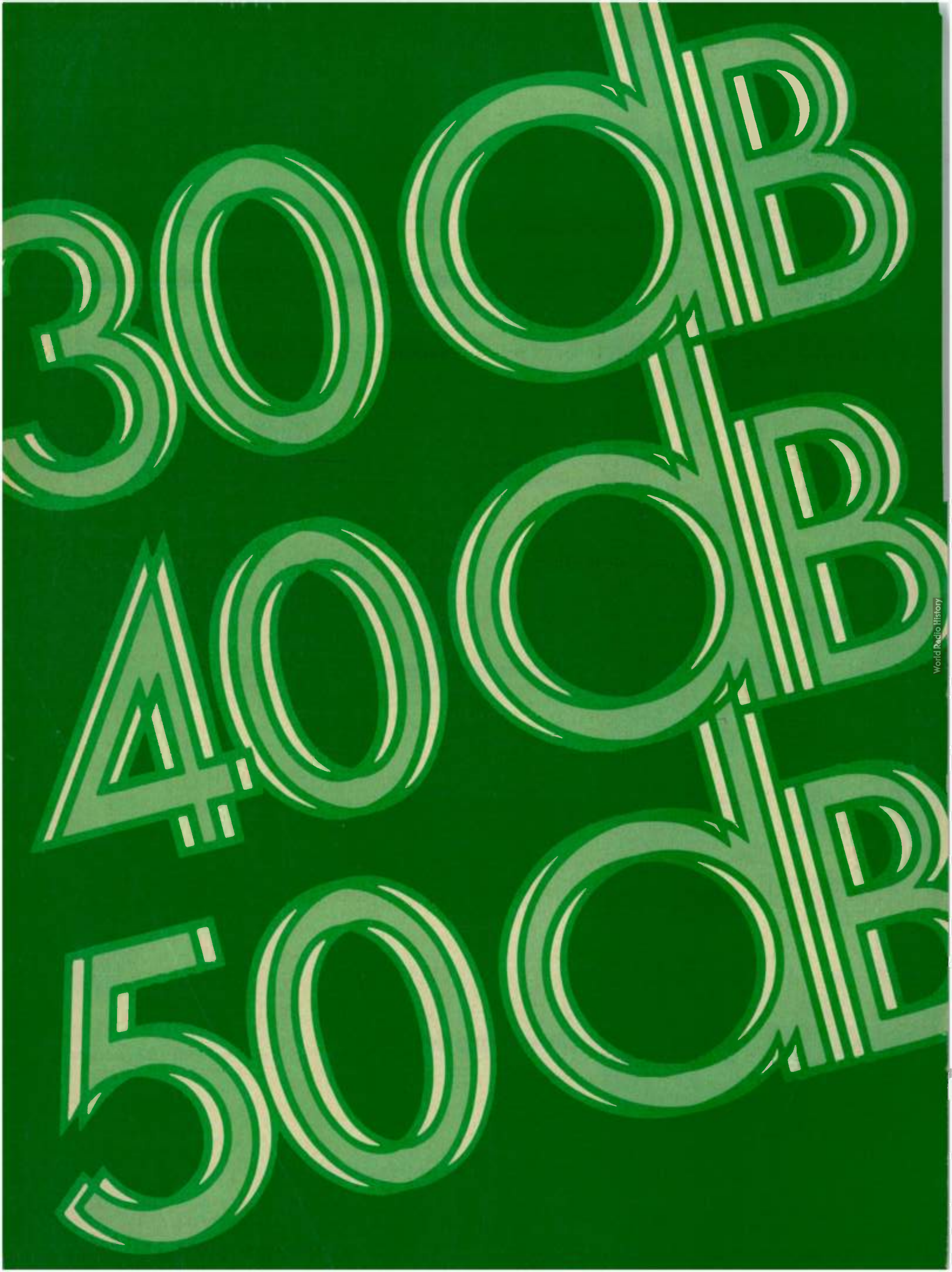
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move forward a distance equal to the projected beam spot diameter, or

$$vt_c = h\beta\left(\frac{\pi}{180}\right)$$

where v is aircraft velocity, t_c is contiguous scan time, and h is aircraft altitude. The main consideration is that the scanner observes each beam width of the scene for a time not less than τ. This places constraints on the sensor integration time, the beamwidth and the velocity/altitude ratio. The integration time τ must be compatible with the minimum detectable temperature that may be observed by the MMW radiometer.

Tracking

Tracking radiometers are used for target designation, weapon delivery and terminal guidance (Figure 2). Usually there is an angle measuring system (conical scan, sequential lobing or monopulse) designed to generate an error signal having magnitudes and phase or polarity that indicate the angle error from the tracker to the target. The radiometer integration time in these cases is very situation dependent and consideration of response time, target extent, and allowable tracking accuracy is necessary.

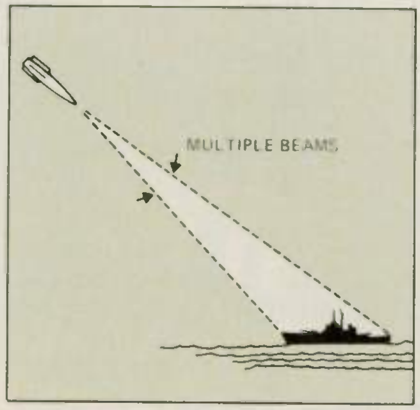

Fig. 2 Target tracking during terminal guidance.

For the tracking radiometer, the S/N for resolved and unresolved targets is given by Reference 32:

$$\left(\frac{S}{N}\right) = \frac{T_C A_t \sqrt{\tau}}{\Delta T_{min_i} R^2 (\beta)^2}$$

where T_c is the contrast temperature between target and background, R is the range to the target and β is the antenna beamwidth.

Target Detection

Target detection criteria is best handled on a probabilistic sense wherein a calculation is made of a detection probability based on the particular situation. Here several factors must be evaluated. The nature of the background (uniform or cluttered), the background dynamics (such as sea state) and the physical size of the target (which can range from tens of feet for an armored vehicle to thousands of feet for fixed sites) must be compared to the antenna resolution spot. Further, the emissivity difference between the target and background must be considered. This difference may be high for an armored vehicle in a grassy field ($\epsilon_T - \epsilon_{GF} = 0.9$) and quite low for the air field compared to a grassy field ($\epsilon_{AF} - \epsilon_{GF} = 0.15$). The effect of the inter-

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CIRCLE 40 ON READER SERVICE CARD

(from page 56) MM-WAVE

vening atmosphere at the frequency of interest and weather conditions must also be considered.

The probability of detecting unresolved targets is shown[32,33] to be 0.9 for S/N = 4.1. For targets too large to be contained within the beam, S/N is increased by $\sqrt{n}$ where n is the number of separate beam positions to cover the target completely. Thus,

$$(S/N)_D = \left(\frac{S}{N}\right)\sqrt{n}$$

The value of $(S/N)_D$ decreases with increasing aircraft altitude in all cases, because the increased dwell time available at higher altitude is more than offset by the decreased beamfill ratio (or, at low altitudes, the decreased number of beam positions affected), and the increased atmospheric attenuation. An abrupt change occurs when the beamfill ratio passes through unity and at the cloud boundaries. In clear weather, excellent performance is achieved at all frequencies. Calculations[32] indicate motor vehicles are detectable (with a P_D of 0.9) up to altitudes of 10, 20, and 35 thousand feet at frequencies of 140, 94, and 35 GHz, respectively. With a moderate overcast, these altitudes decrease to 5, 8.5, and 26 thousand feet. Large manmade targets and geographic formations benefit enormously from spatial integration, and are detectable from very high altitudes in all cases.

Atmospheric data collection from high altitude aircraft[39] is being carried out. In the future, a satellite weather sensor for the 183 GHz water vapor line (humidity) and the 118 GHz oxygen line (temperature) could aid in weather data collection. Future satellite based radiometers, when placed in geosynchronous orbit with a 4-meter dish, would yield resolution of 15 to 25 km on earth. In this orbit, atmospheric temperature and humidity sounders would start at a given spot on earth and continuously monitor rapidly changing phenomena, such as severe local storms. Such a sensor would provide excellent mezzoscale meteorology data.

Another recent application involves launch-pad operations in the detection of ice formation on rocket cryogenic fuel tanks using

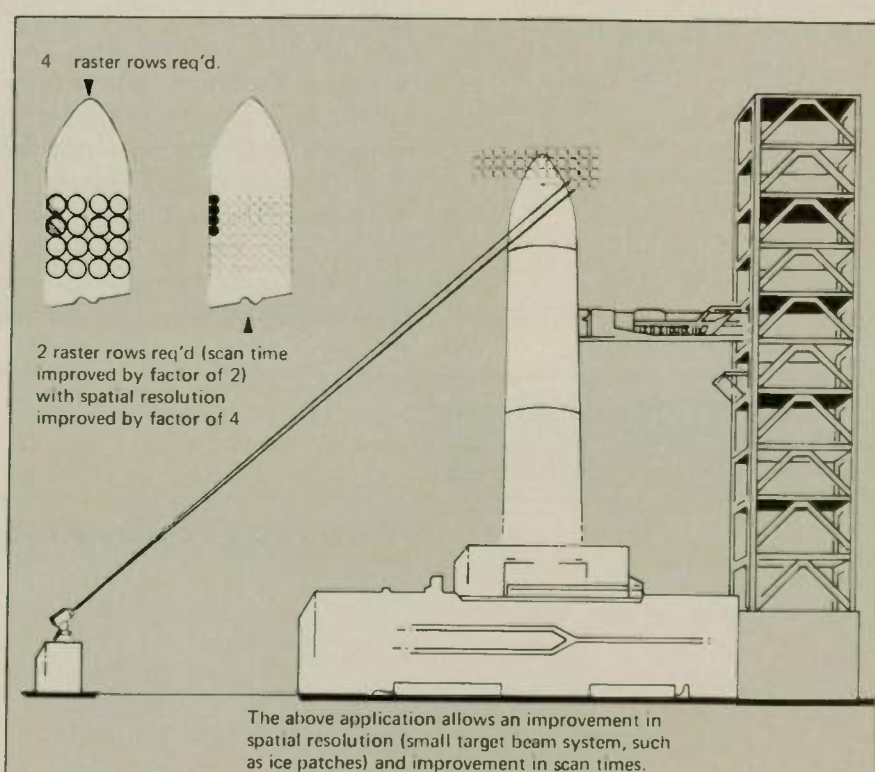

Fig. 3 Scan raster geometrics.

a 35/95 GHz multichannel scanning radiometer. Figure 3 depicts the single beam raster scan approach currently implemented by Georgia Tech on a fuel tank. Future versions of such a scanner might use four simultaneous beams for improved spatial resolution and shorter scan time. One of the critical parameters for this application is the total time required to scan the target, which must be minimized to allow rapid detection of ice or frost formation in isolated regions of the rocket fuel tank. Since a rocket can be 50 m in length and 3 m in diameter, the short scan time per raster requires a fast antenna beam scanner. The spatial resolution of the millimeter-wave receiver must be sufficient to allow the detection of ice patches with dimensions as small as 30 cm by 30 cm. The use of a multiple beam antenna system would reduce the total scan time and simultaneously improve the receiver's spatial resolution as shown in Figure 3.

RECEIVER HARDWARE

Compact and rugged receiver circuits will be needed to satisfy the multiple beam antenna system requirements. Each antenna beam will likely require the use of a sensitive integrated multichannel receiver. The reliability and environmental aims of military type systems imply that active and passive circuits must be combined in compact structures. The millimeter receiver design philosophy includes the development of multifunction millimeter integrated circuits containing integral mixers, IF amplifiers, and local oscillator power-on circuitry.[35–38] Low-noise, solid-state devices become an integral part of the integrated millimeter-wave receiver.

Due to the desire to use an all solid state local oscillator at the higher millimeter-wave frequencies, subharmonic mixing is being employed in millimeter receivers.[39–44] Recently developed subharmonically pumped mixers which use antiparallel mounted diodes are seen as a valuable approach for MFLIR fabrication. The electrical properties of the

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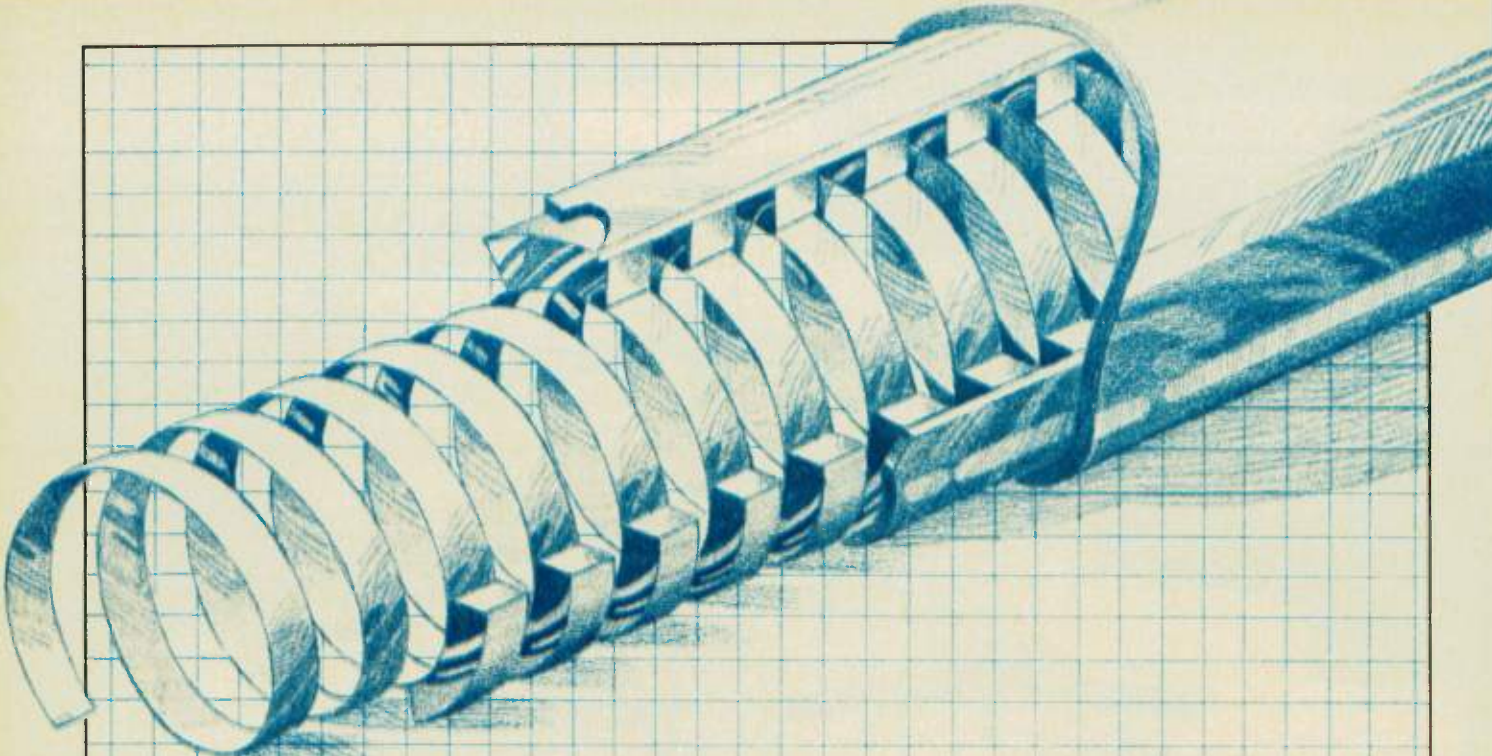

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140 GHz Silicon IMPATT Power Combiner Development

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Hughes Aircraft Company
Electron Dynamics Division
Torrance, CA

Waveguide cavity power combiners using silicon IMPATT diodes have been developed to generate high pulsed power near 140 GHz. Peak output power of 5.2 W for a two-diode combiner and 9.2 W for a four-diode combiner have been achieved with 100 ns pulsewidth and 25 kHz pulse repetition rate.

INTRODUCTION

The rapidly increasing demand for millimeter-wave radar systems has created an urgent need for high power solid-state transmitters. Millimeter-wave systems require smaller antennas and provide better resolution than the microwave systems. Compared with optical systems, the millimeter-wave systems offer better penetration capabilities through fog, clouds and dust. At the low loss atmospheric window of 140 GHz, recent progress in IMPATT diode development has resulted in the generation of 3 W peak output power.[1] Although the output power from a single diode can be further increased by better diode design and processing, it will eventually reach a limit due to the thermal and impedance problems. To meet many systems requirements, it is necessary to combine several diodes in a circuit to achieve higher power levels. Many power combining approaches have been proposed and tried in the past decade. They have fallen mainly into four categories: resonant cavity combiners, nonresonant combiners, chip-level combiners and spatial combiners. Chip-level combiners connect multiple mesa chips electrically in series and thermally in parallel.[2] Spatial combiners accumulate a large number of energy sources by using a dense array of radiating elements.[3] Nonresonant combiners include hybrid coupled combiner[4] and Wilkinson-type combiner.[5-7] An X-band resonant cavity combiner was first proposed and demonstrated by Kurokawa and Magalhaes[8] and later modified and improved by Harp and Stover.[9] The basic module for the Kurokawa's type combiner is a cross-coupled coaxial-waveguide diode mounting structure. Recently a four-diode power combiner operating at a center frequency of 94 GHz with 40 W peak output power has been reported.[10] The combiner uses a circuit similar to that proposed by Kurokawa. This paper summarizes the extension of this technology to 140 GHz with a peak output power of 9.2 W.

DIODE DEVELOPMENT

At 140 GHz, diode fabrication requires a considerable amount of care because of the extremely small dimensions involved. For example, the typical epitaxial thickness is only on the order of a few tenths of a micron, which is taxing the capability of most epitaxial reactors. The performance of the IMPATT diode therefore depends critically on how well the diode parameters are controlled. The single-drift diode, shown in Figure 1 was chosen over the double-drift diode because of its relatively simple structure which is easier to realize. The diode design was based on a small-signal computer analysis.[11] For a specified current density, junction temperature, and junction doping profile, the small-signal computer program calculates and plots the dc electric field as a function of distance. With this dc solution, the device small-signal RF conductance and susceptance per unit area and the device Q can be calculated as a functions of frequency for a specified frequency range. By means of an iterative process for different values of input parameters such as doping density, a condition is reached for which the device Q displays its maximum value near the desired frequency of operation. The parameters of the device which produce this condition are then taken as the design values. For 140 GHz operation, the doping concentration of the N epitaxial layer was selected to be 1.9×10^{17} atoms/cm^3. The breakdown voltage was computed to be 9.7 volts and the operating voltage was 14 volts.

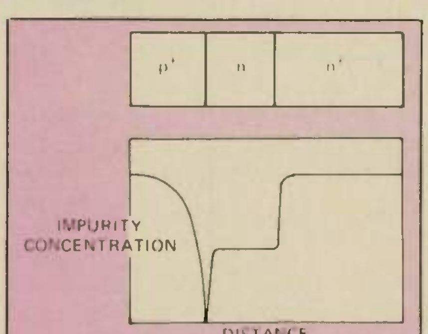

Fig. 1 Single-drift IMPATT diode structure.

The single-drift IMPATT profile was first formed by epitaxially growing an n-layer on an n^+ arsenic doped substrate wafer, followed by a p^+ boron diffusion forming a p-n junction. After profile formation, an array of inverted mesa diodes was fabricated from the wafer and individually

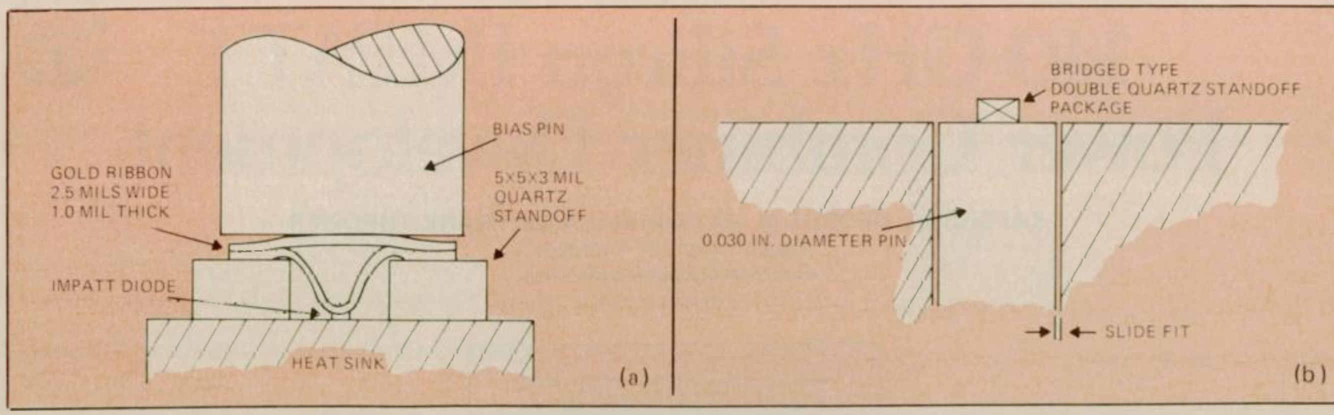

Fig. 2 Diode package (a) Bridged type double-quartz-standoff package (b) Heatsink pin configuration

bonded on copper heatsinks.

At 140 GHz, the diode package design is extremely important. The package transforms the low diode chip impedance to a higher level waveguide impedance. To accomplish this, a bridged double-quartz-standoff package configuration as shown in Figure 2(a) was developed.

To achieve the highest possible power output in a combiner circuit, it is desirable to preselect diodes which have similar RF characteristics. This calls for first evaluating each diode individually in a reduced height waveguide cavity test circuit and if the RF performance is satisfactory, removing the diode from the single diode test circuit and mounting it in the multi-diode circuit. Conventionally, such a process involves two soldering steps and one unsoldering step. Because of the small size and fragility of the double-quartz-standoff package, these steps often result in irrepairable damage to the diode.

In order to solve this problem, a heatsink pin approach was devised to entirely eliminate any such soldering and unsoldering of the diode package in and out of the heatsink slab. In this method, a 0.031 in. diameter hole is drilled through the heatsink slab at the place where the diode is mounted. A cylindrical pin with exactly the same diameter as the hole is cut to the desired length and polished at one of the end faces to a high degree of flatness and smoothness. The diode chip and the two quartz standoffs are bonded onto the polished pin surface. To test the diode, the "heatsink pin package" is simply inserted into the hole in the heatsink slab and then secured in place by a set screw. Figure 2(b) shows the heatsink pin configurations. Although the small air gap increases the thermal resistance, it is expected to pose no problems for pulsed operation with low duty factor.

COMBINER CONFIGURATION

Our combiner design was based on the resonant waveguide combiner developed by Kurokawa at X-band frequencies.[8] A schematic diagram of the cavity design is shown in Figure 3. There are, however, certain noteworthy departures from the conventional low frequency circuit design, which are:

- An oversized waveguide is used. The cavity width, which is 0.100 in., is wider than the standard WR-7 waveguide width of 0.065 in. for the 110-170 GHz frequency range.
- The diode spacing in the longitudinal direction is one guide wavelength compared to the conventional one-half-guide wavelength spacing.
- The distance from the cavity opening to the first diode row is $5\lambda_g/4$ and not $\lambda_g/4$.
- The cavity is formed by the sliding short and the discontinuity due to the transition from the change of waveguide width. This transition introduces an inductance element at the interface plane.

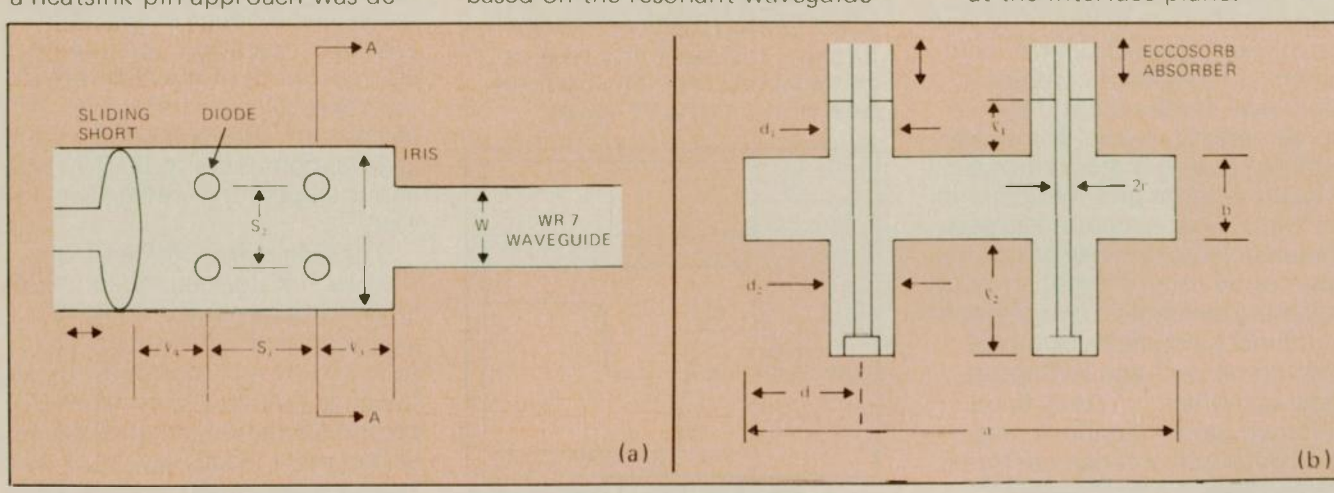

Fig. 3 Four-diode waveguide cavity. (a) Top view. (b) Cross-sectional view at AA.

- Eccosorb® terminations with flat ends (mismatched type) are used instead of those with tapered ends (matched type).

- The diodes are not mounted at the edge of the waveguide; instead, they are located close to the center line of the waveguide.

The reason for the first three modifications is primarily due to the mechanical limitations imposed by the dimensions of the diode modules. The reason for the last three modifications will be explained next.

EQUIVALENT CIRCUIT AND THEORETICAL DISCUSSION

The basic components in the Kurokawa-type resonant waveguide combiner consist of coaxial modules and a waveguide cavity. The power generated from each IMPATT diode couples to the waveguide cavity through the coaxial module.

Resonant Cavity

The resonant cavity is formed by the sliding short and the waveguide discontinuity. For a rectangular resonator as shown in **Figure 4**, the resonance frequency is given by:

$$f_{nmp} = \frac{1}{2\sqrt{\mu\epsilon}}\sqrt{\left(\frac{n}{a}\right)^2 + \left(\frac{m}{b}\right)^2 + \left(\frac{p}{c}\right)^2} \quad (1)$$

where a is the waveguide width (x-axis), b is waveguide height (y-axis) and c is the resonator length (z-axis), with the corresponding eigen numbers (mode numbers) n, m and p. The resonance frequencies can be readily calculated for given resonator dimensions. Since the IMPATT diode has negative resistance from 100 GHz to 170 GHz, all modes in this frequency range must be considered.

For a = 0.100 in. and b = 0.0325 in., all higher order modes with $m \geqslant 1$ have resonance frequencies of no less than 180 GHz. Since this is most likely beyond the negative resistance frequency range of the IMPATT diode, we may set m = 0. At 140 GHz, the cavity length is given by:

$$c = 2.5\lambda_g = 0.2325 \text{ in.} \quad (2)$$

Equation (1) can be rewritten as:

$$f_{nop} = 59.05 \sqrt{n^2 + 0.185\,p^2} \text{ (GHz)} \quad (3)$$

All the resonance frequencies below 170 GHz are listed in **Table 1**. Later it will be seen that the spurious mode near 128 GHz might be excited in tuning the four-diode combiner.

Equivalent Circuit

The basic component of the combiner is a coaxial-waveguide diode mounting module. This module consists of a cross-coupled coaxial-waveguide diode mounting structure with a diode mounted at the bottom of the post. A general structure is shown in **Figure 5(a)**. The coaxial line can be of different diameters in the upper and lower sections; and Z1, Z2, Z3, and Z4 are the load impedances at each port, respectively. The equivalent circuit shown in **Figure 5(b)** can be found from Reference 10.

Z_o is the characteristics impedance of the waveguide and Z_{01} and Z_{02} are those of the coaxial lines. Z_{op} is an inductive component due to the post in waveguide excited by TE_{no} modes. Y' and Y_{1p}, Y_{2p} account for the effects of waveguide-coaxial junctions.

Computer programs were developed, based on the circuit model. The accuracy of the model was checked by comparing the computer results with experimental results at lower frequencies where the impedance measurement could be easily conducted. The general agreement between the theoretical results and measurement is good, as described in Reference 10. An exact quantitative design should include interactions among posts and is thus very difficult, if not impossible. However, our purpose was to determine which parameters were most critical and then treat these highly sensitive parameters with extreme care or make them adjustable elements. With the circuit analysis, it was found[10] that the location of the mismatched load (i.e., flat type Eccosorb® termination) is critical in achieving high power output and it also gives one more degree of freedom of tuning the circuit. This provides capability of individual tuning for each diode to accommodate the differences among the diodes. Also, the tuning is essential for the impedance matching from diode to circuit. With proper adjustment of Eccosorb® positions, the power dissipated in the absorber can be reduced to a minimum. The results of the computer program at lower frequencies also indicated that the sliding short position, cavity output loading and diode separation distance are important parameters to the achievement of efficient power accumulation. Em-

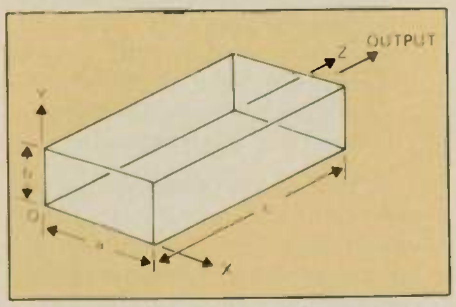

Fig. 4 Waveguide resonator cavity coordinate system.

TABLE 1

RESONANCE FREQUENCIES f_{nop} (GHz) BELOW 170 GHz

n \ p	1	2	3	4	5	6
1	64.28	77.89	96.40	117.50	140.05	163.43
2	120.80	128.56	140.54	155.78		

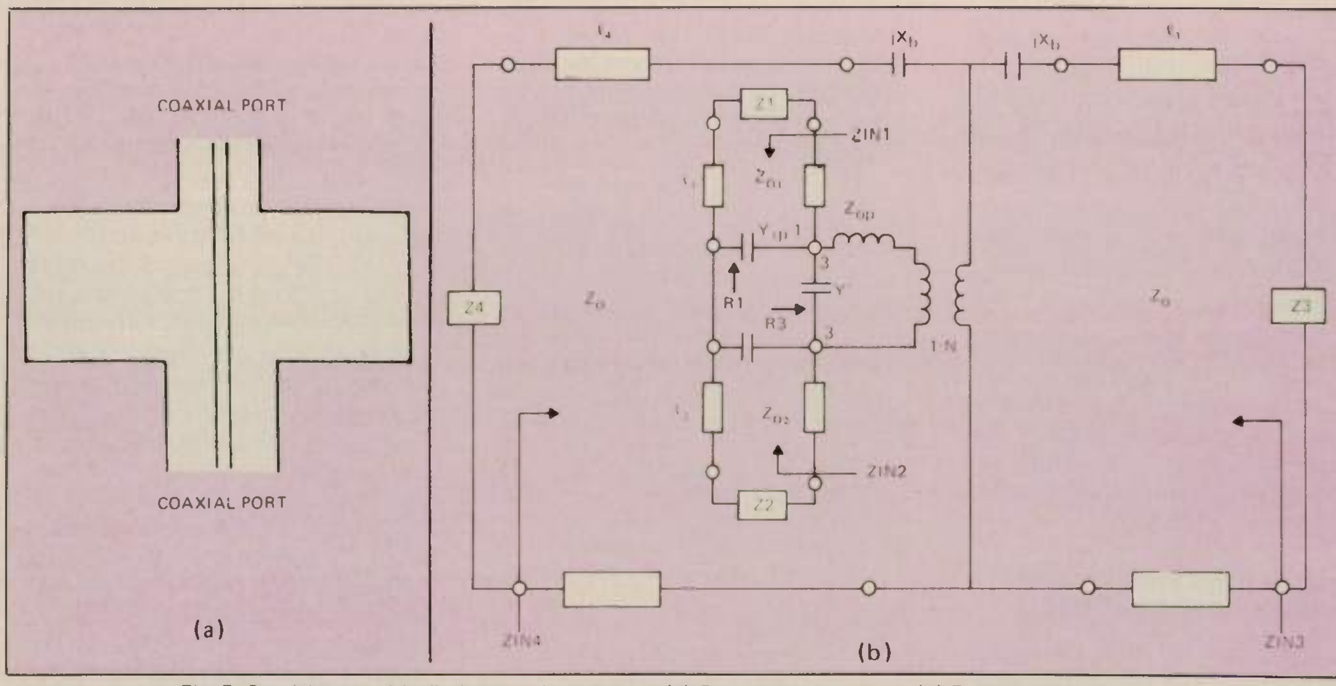

Fig. 5 Coaxial-waveguide diode mounting structure (a) General configuration (b) Equivalent circuit.

pirical effort was thus spent on optimizing these parameters in the combiner development.

COMBINER DEVELOPMENT

Based on the previous discussion, a four-diode combiner and a two-diode combiner were constructed and tested.

Four-diode Combiner

Photographs of the four-diode combiner are shown in **Figures 6(a)** and **(b)**. To improve the quality of bias current waveform, careful design on the bias circuit is essential. Stray capacitances should be reduced and the cable length should be kept as short as possible. As shown in **Figure 6(b)**, a microstrip bias board was designed and fabricated, so that four bias pulses generated from modulators could be connected to the four corners of the board and transmitted to the center through 50 Ω microstrip lines. Four bellows were soldered to the microstrip lines to insure good contact between the bias pins and diodes. The spacers, sandwiched between the microstrip board and top cavity piece, were used to adjust the length of the top cavity piece. The waveguide is formed by the bias slab and the heatsink slab. With an open window on each side of the top cavity piece, the positions of the four Eccosorb® terminations can be adjusted. The diodes were mounted in heatsink pin packages as described in **Figure 2**.

The four-diode combiner has a number of parameters which can be varied discretely or continuously in order to optimize the circuit:

- Several different pin diameters can be used to optimize the output power and frequency. They are 0.020 in., 0.022 in. and 0.027 in. It was found that 0.022 in. diameter provided the best performance.
- The separation between two diodes in the same transverse plane can take on any one of the four values, namely, 0.070 in., 0.080 in., 0.090 in. and 0.100 in. This is accomplished through the use of four separate bias slabs and four separate heatsink slabs. Through experiments the best separation was determined to be 0.070 in.
- Because each heatsink pin can slide up or down in the common heatsink slab when the set screw is loosened, the top face of the pin can be made to be recessed, flush or protruding with respect to the upper surface of the heatsink slab. Different spacers can be inserted between the microstrip board and the top cavity piece to compensate for these length differences. This degree of freedom can be very effective in optimizing the circuit and has been used very extensively in low frequency single diode oscillators and combiners. Its implementation in practice, however, is quite cumbersome and was not extensively utilized for our combiner development.
- Each Eccosorb® termination can be adjusted up or down the bias pin continuously for optimum tuning.
- The sliding short position can be adjusted continuously for optimum tuning.
- Each bias current level can be adjusted continuously through the pulse modulator. The phase differences among the bias current pulses can also be varied continuously. The waveform of each bias current can be shaped discretely by changing the pulse-forming network in the modulator.

A four-diode combiner has been built and tested with diodes which were first individually tested and selected in a single-diode oscillator circuit.

The structural dimensions of

this combiner (refer to **Figure 3**) are as follows:

a = 0.100 in.
b = 0.0325 in.
w = 0.065 in.
r = 0.011 in.
d_1 = 0.052 in.
ℓ_1 = adjustable
ℓ_2 = 0 in.
ℓ_3 = 0.1375 in.
ℓ_4 = adjustable
S_1 = 0.110 in.
S_2 = 0.070 in.

The above dimensional values listed and used for ℓ_3 and S_1 were computed by mistake using the guide wavelength calculated from the standard WR-7 waveguide with width of 0.065 in. For an oversized waveguide with a width of 0.100 in., the guide wavelength is about 15 percent smaller. Fortunately the effects of these structural errors could be compensated by the sliding short adjustment and Eccosorb® tuning.

A peak output power of 9.2 W at a center frequency of 139.3 GHz has been achieved. The RF-output video pulse from the combiner is shown in **Figure 7**. The diodes used for the combiner were selected to have a low breakdown voltage and similar output power (3-4 W) and frequency. The diodes with higher breakdown voltage and thus lower operating frequency had a tendency to excite the lower cavity mode (around 134 GHz, i.e., mode f_{202} in **Table 1**) in the combiner circuit. The diodes were operated with 100 ns pulsewidth at 25 kHz pulse repetition rate. Frequency chirp of less than 2 GHz

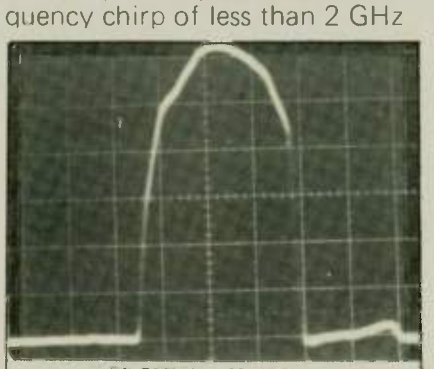

Fig. 7 The RF-output video pulse from a four-diode combiner.

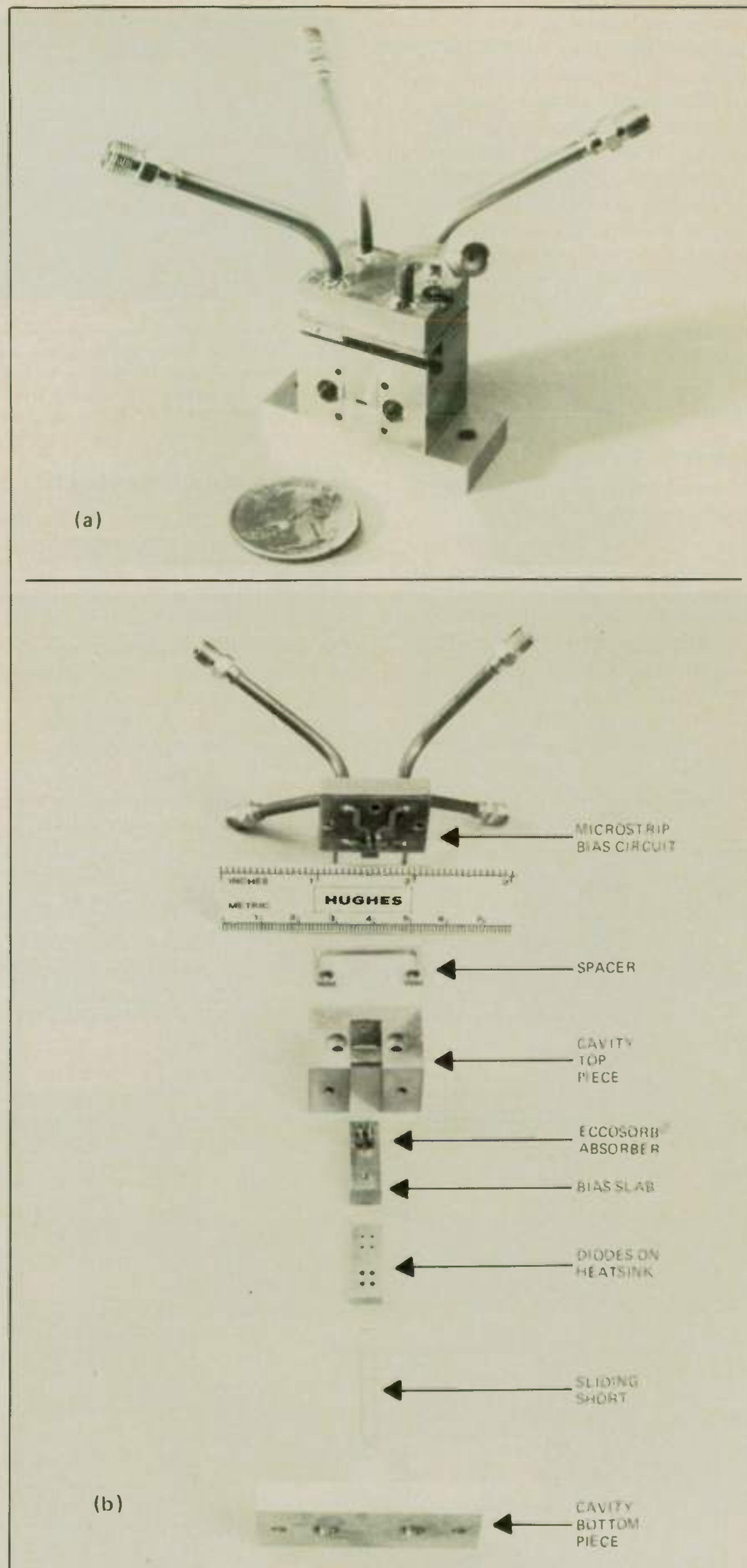

Fig. 6 Four-diode power combiner with 9.2 W peak output power. (a) Assembled circuit. (b) Disassembled circuit.

and power variation across the pulse of less than 1 dB were achieved. Combining efficiency of 60 percent was generally achieved. This low combining ciency of 60 percent was generally achieved. This low combining efficiency is partially attributed to the lower bias current rating applied to each individual diode in the combiner compared to that used in the single diode testing in an effort to improve the combiner reliability. The bias current applied to the combiner is about 10 percent lower.

Two-diode Combiner

In order to establish the two-diode combiner capability, the four-diode combiner was modified to become a two-diode combiner fixture simply by inserting two plug-in pins into the two unused holes in the bias slab. Peak output power of 5.2 W at a center frequency of 142.2 GHz was obtained. A photograph of the RF-output video pulse and the two bias current pulses are shown in **Figure 8**. Combining efficiency of over 80 percent was achieved.

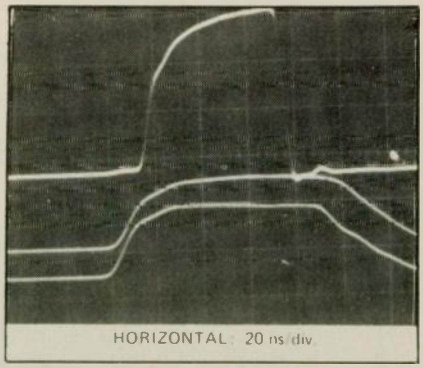

Fig. 8 The RF-output video pulse and the two bias current pulses for the two-diode combiner. Top trace is RF-output video pulse and the bottom traces are current pulses.

MODULATOR DEVELOPMENT

An important item in the development of a high power pulsed IMPATT combiner is the multichannel pulse modulator. The pulse modulator not only must provide high current up to 10 amperes with very small jitter but must also have fast rise and fall times, because the pulsewidth of interest is only on the order of 50 to 200 nanoseconds. Moreover, because of the frequency chirping of the pulsed IMPATT diodes, capability for current waveform shaping must be provided to minimize the chirp bandwidth such that all the diodes in the combiner can be properly combined. The features of the four-channel modulator are its small package size and easy access to all adjustments.

The block diagram of a four-channel modulator is shown in **Figure 9**. A multiple output trigger serves as a current amplifier and distributes the trigger signal to all four channels. The modulator utilizes four pulse-forming networks (PFN) to achieve the desired pulse current waveforms. Each pulse forming network incorporates a seven-section adjustable L-C network to control the magnitude and slope of the current at different positions of the bias pulse. By adjusting the lumped inductors and capacitors of PFN, it is possible to vary the shape of the drive pulse to the IMPATT diode. This feature provides a versatile method to control frequency characteristics of the pulsed IMPATT oscillation. The line is discharged into the

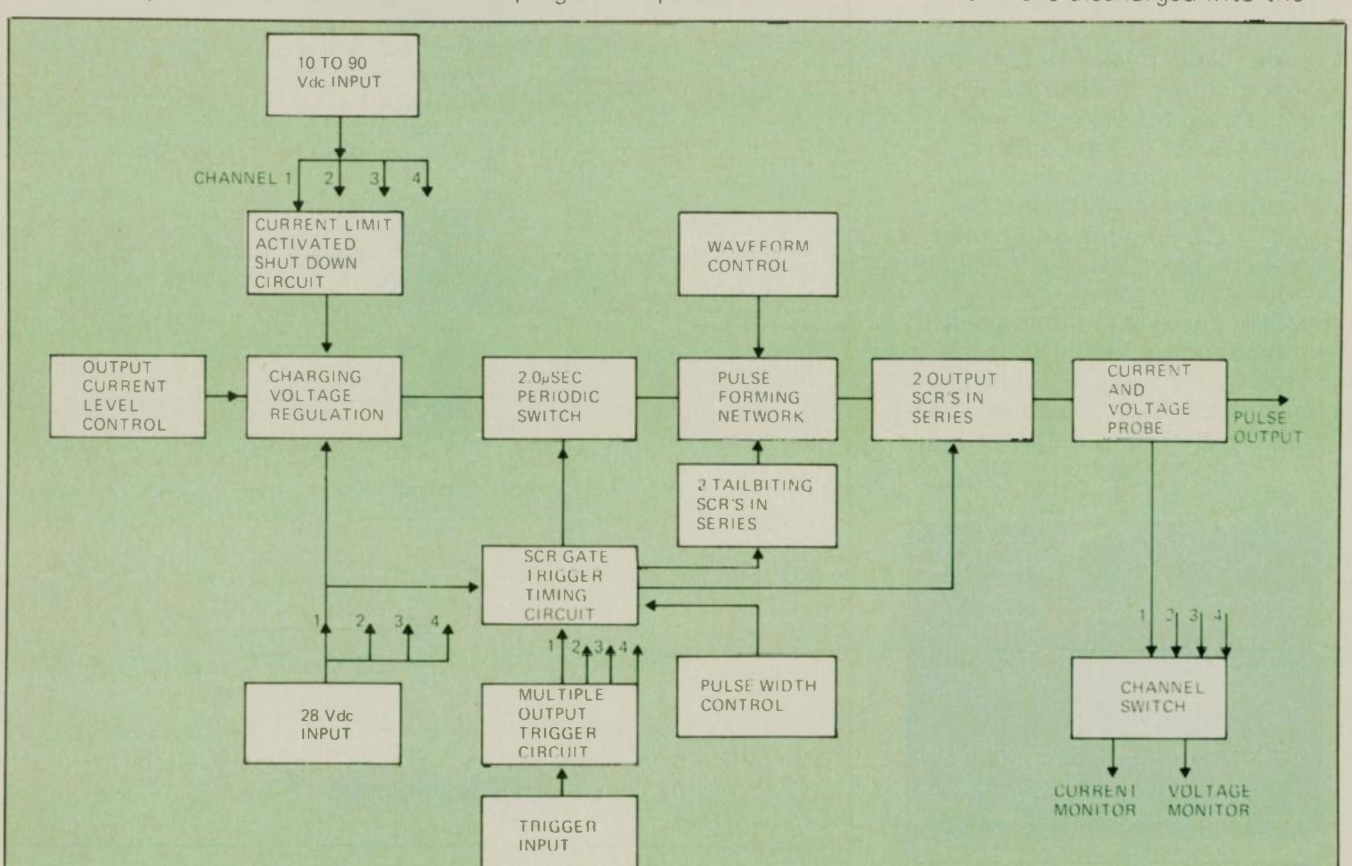

Fig. 9 Four-channel modulator block diagram.

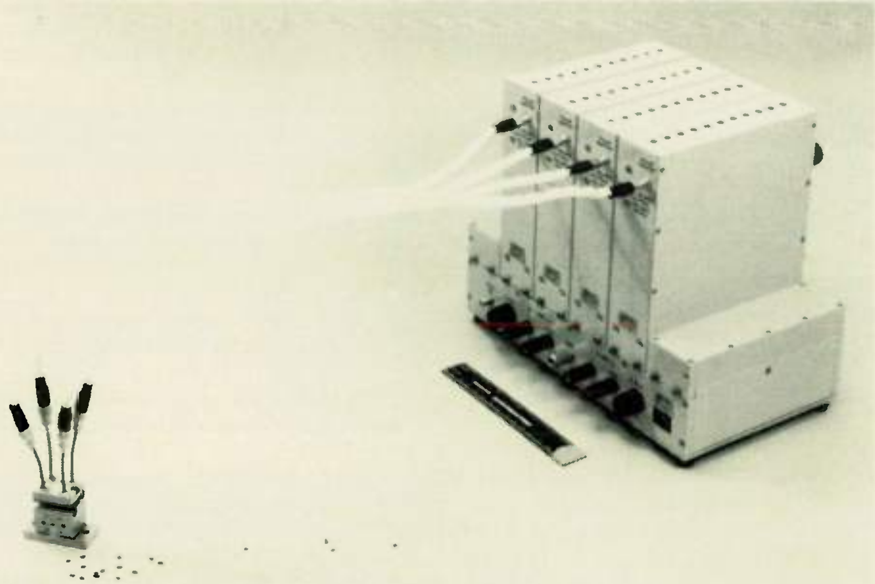

Fig. 10 A four-diode combiner connected to a four-channel modulator.

load using a SCR switching circuit. Tailbiting SCR's short the PFN at the end of the desired pulse time so that a short fall time can be obtained from the relatively slow PFN. The output current and voltage of each channel can be easily monitored one at a time by a switch control.

The modulator was designed to provide up to 15 amperes peak current with a regulation of about 1%. Controls are provided to adjust the operating current and pulsewidth. A TTL triggering circuit is provided in the modulator to control the repetition rate. The minimum supply voltage required at the modulator depends on the actual operating voltage of the IMPATT diodes. A sufficient margin above this voltage must be provided to properly bias the transistors in order to ensure adequate regulation of the output current. For this reason, a maximum +70 Vdc is required as the modulator supply voltage. In addition, +28 Vdc is required for biasing the integrated circuits used for voltage regulation. This modulator is capable of 0 nsec to 200 nsec pulsewidths up to 1% duty.

The modulator also provides auxiliary circuits for short circuit protection. If an external short causes SCR lockup, the power supply goes to a preset current limit. The sensor then detects this condition and shuts off the charging voltage regulator until the load is removed. A photograph of this four-channel modulator with the four-diode combiner is shown in Figure 10.

CONCLUSIONS

A 140 GHz four-diode power combiner using a rectangular waveguide cavity circuit has been described. Peak output power of 9.2 W was achieved from the four-diode combiner. A heatsink pin diode package and a four-channel modulator have also been described, to facilitate the combiner development. The design presented here firmly established the feasibility of power combining at 140 GHz.

ACKNOWLEDGEMENTS

The authors wish to thank E. N. Nakaji, R. S. Ying and H. J. Kuno for many helpful suggestions and discussions. They also wish to thank L. E. Anderson and D. L. English for their technical assistance in modulator development. The support by the Army Electronics Research and Development Command (Contract No. DAAK20-79-C-0259) under project monitors J. Armata and L. Wandinger is gratefully acknowledged.

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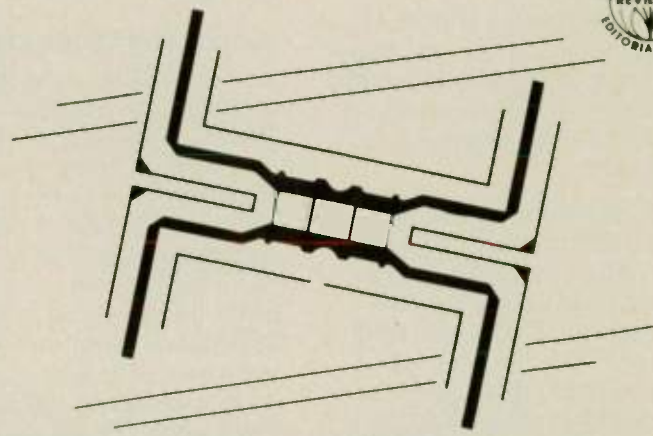
Suspended Substrate K_a-Band Multiplexer



ALFRED HISLOP and DAVID RUBIN
Naval Ocean Systems Center
San Diego, CA

INTRODUCTION

Channelized downconverter techniques are finding increased use in surveillance receivers, especially at millimeter-wave frequencies where extremely wide bandwidths must be covered. Wideband mixers with switched or swept local oscillators have been used, but these systems are limited to an instantaneous bandwidth equal to the downconverted IF bandwidth, and also suffer from lack of image rejection.

The key front-end component of the channelized downconverter is the RF multiplexer. Millimeter-wave multiplexers have been demonstrated previously in microstrip[1] and finline[2] structures. In the microstrip case, three mixers were later integrated on the same substrate with the multiplexer to produce a channelized downconverter.[3]

As part of a program to develop small, inexpensive, low loss millimeter-wave circuits, we have been investigating the properties of low dielectric constant suspended substrate structures. In October 1980, we published an article in *Microwave Journal*[4] which discussed several techniques found useful in the design of transitions and filters in this medium. This article will compare theoretical and measured phase velocities and impedances of suspended substrate lines, show some very useful 3 dB branch couplers, and describe a four channel suspended substrate multiplexer.

SUSPENDED SUBSTRATE TRANSMISSION LINE PARAMETERS

Swept frequency transmission measurements were made on suspended substrate lines of various

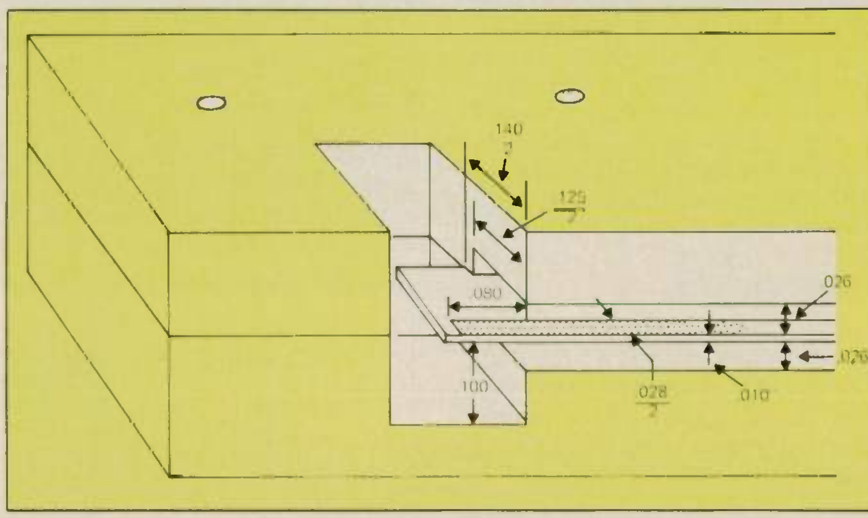


Fig. 1 Probe transition half-section[4].

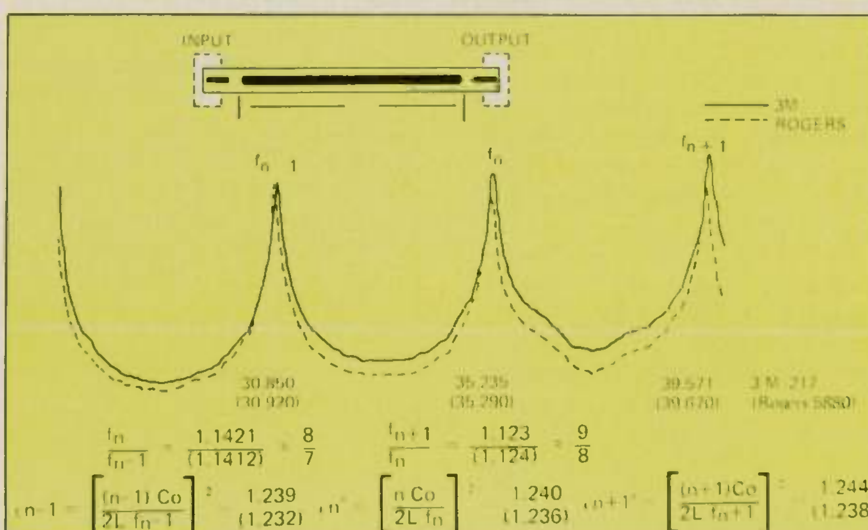


Fig. 2 Measurement of effective dielectric constant. Example shows .050" width line on .010" thickness 3M Cu Clad and Rogers Duroid.

widths to determine their effective dielectric constants. The channel dimensions and probe transition details are shown in **Figure 1**. Measurements of transmission loss using various lengths of .028 inch width transmission line resulted in an average loss of .19 dB/inch.

To determine effective dielectric constant, lines of various widths were loosely coupled to input and output probes, and swept transmission measurements were taken. The frequencies of the transmission peaks were accurately determined using a frequency counter. As shown in **Figure 2**, the effective dielectric constant can be found at the transmission peak frequencies, with each peak corresponding to an integral number of half-wavelengths in the line length, L. The length should be made several wavelengths long in order to avoid errors associated with coupling and end capacitance, particularly for the wider lines. The ends of the lines were shaped to eliminate the effects of step discontinuities[4] although these effects are not significant for thin or very long lines.

The effective dielectric constants for the various line widths showed little dispersion, varying less than 3% over K_a band. Very little difference was found between Rogers 5880 Duroid (rolled copper) and 3M 217 Cu Clad material, both in these measurements and in the filter and multiplexer circuits built later. The average measured effective dielectric constants are compared in **Figure 3** with the theoretical values calculated using the programs of Mirshekar-Syahkal and Davies.[5,6] **Figure 3** also contains a plot of the characteristic impedances calculated using the above programs. Using the measured dielectric constant values and assuming pure TEM propagation, characteristic impedances were also calculated using Cohn's formulas for triplate in a homogeneous medium.[7] The results clearly agree, indicating the utility of this classical paper after 27 years! It must be emphasized that all measured and computer results were for the particular channel configuration used.

3 dB QUADRATURE COUPLERS

A synchronous four branch coupler with impedances modified from those given by Levy[8] was used for the primary design. Minimum line width considerations (about .001") limited the outside branch lines to a maximum of 266 ohms. Using computer graphics, the effects of line impedance variations were readily determined. It was found (**Figure 4**) that changing the outside branch lines of a 4 dB ($\lambda go/\lambda g2 = 1.3$) Tchebysheff design from 362 ohms to 266 ohms gave the required performance.

To allow for the large width of the coupler, the suspended substrate channel must be widened. The usual moding problems associated with channels large enough to support waveguide modes were soon evident. To eliminate the moding problems, holes were drilled through the substrate between the branch lines, and .020 inch diameter metal rods were inserted from the top of the fixture

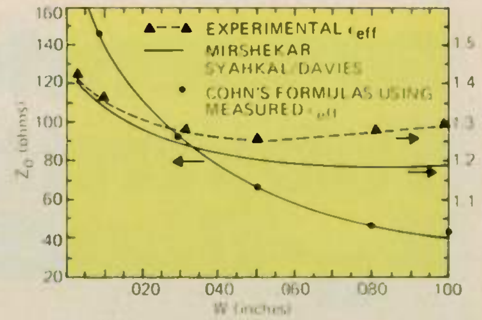


Fig. 3 Measured values of effective dielectric constants (and associated TEM characteristic impedances) compared to derived values of Mirshekar-Syahkal & Davies.

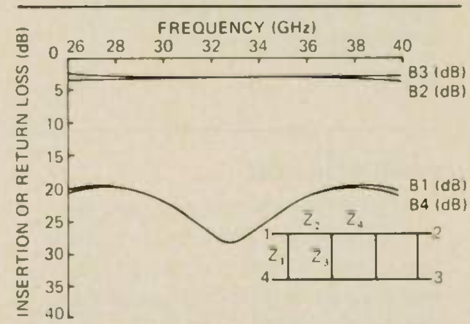


Fig. 4 Modified four branch synchronous coupler design. $\bar{Z}_1 = 2.96$, $\bar{Z}_2 = .8121$, $\bar{Z}_3 = 1.557$, $\bar{Z}_4 = .6855$.

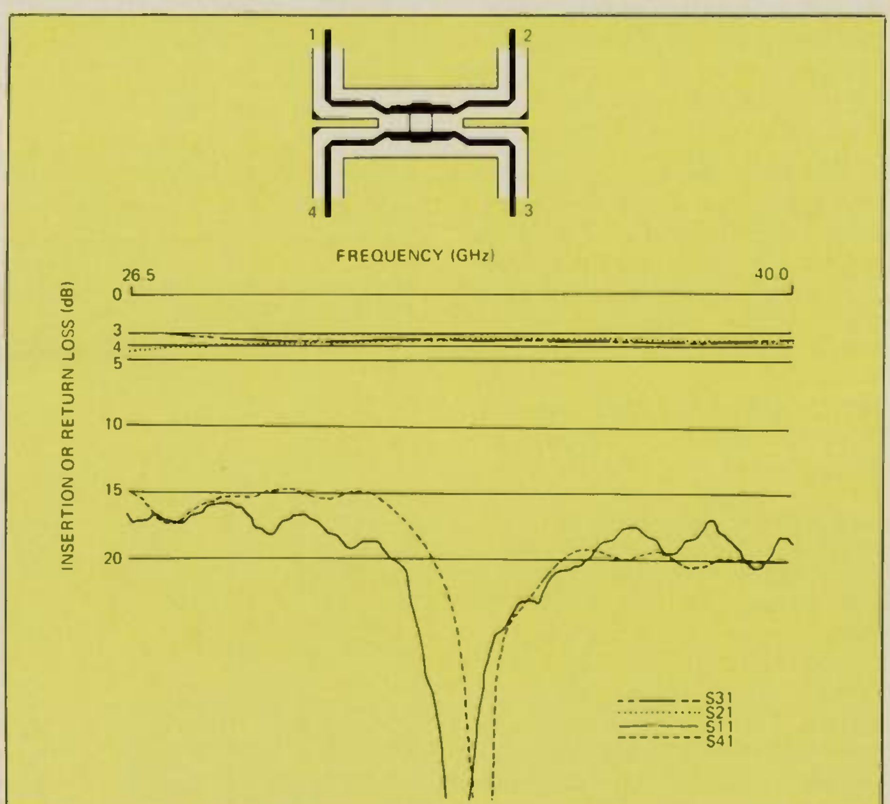


Fig. 5 Coupler based on parameters of Figure 4. $\bar{Z}_1$ changed to 1.77 for good performance, possibly countering the capacitive effects of the moding screws.

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SIGNAL ACQUISITION TO 40 GHz WITH ONE ANTENNA.

CIRCLE 50 ON READER SERVICE CARD

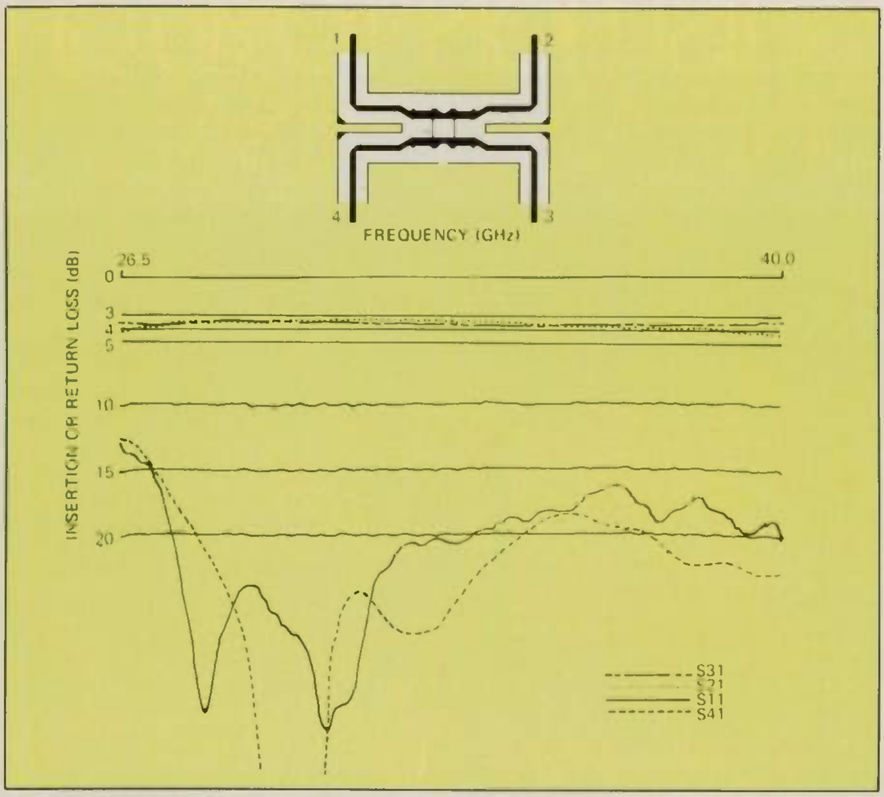

Fig. 6 Final coupler design. Experimentally determined T junction compensation resulted in better directivity and input SWR.

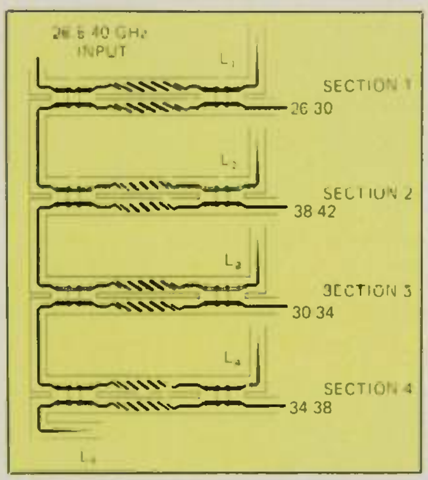

Fig. 7 Multiplexer pattern.

through the substrate to the bottom of the channel. The moding problem was eliminated, but wideband coupling was not achieved.

To obtain wideband coupling close to 3 dB, the width of the inner branch lines was decreased. This may have been necessary due to the close proximity of the mode suppression rods. We had previously found that good performance of microstrip branch line couplers could be achieved at these frequencies if the distance between the inside edges of the through lines is λ/4, and the distance between the nearest edges of the branch lines is λ/4. That is, neglecting differences in phase velocity for the different width lines, the areas enclosed by the branch and through lines are squares. This same technique was applied to the suspended substrate couplers, and the result is shown in **Figure 5**. Further compensation of the branch line T junctions was done empirically, and the results are shown in **Figure 6**. This coupler was used in the multiplexer due to its increased directivity.

K_a-BAND MULTIPLEXER

A channel dropping K_a-band multiplexer was made using the hybrid of **Figure 6** and the filters described in Reference 4. The circuit pattern is shown in **Figure 7** and consists of four hybrid-filter-hybrid sections and ferrite wedge loads (L1-L5) above and below the tapered lines. The filters of section 1 pass signals from 26 GHz to a 3 dB upper band edge frequency of 30 GHz. Power rejected by this section goes to the next section. Section 2 has a lower 3 dB band edge frequency of 38 GHz, and is designed to pass signals to 42 GHz. (Moding problems increase the insertion loss of this section above 41 GHz). The filters of section 3 pass 29.8 GHz to 34 GHz. The lower band edge of the output of section 3 is actually determined by the upper band edge frequency of section 1. The filters of section 4 pass 33.8 - 38.2 GHz, and the lower and upper band edge frequencies of the output of this section are de-

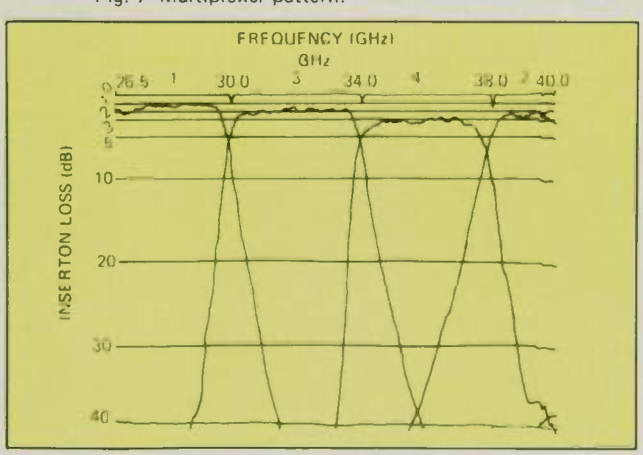

Fig. 8 Measured performance of complete multiplexer, all losses included.

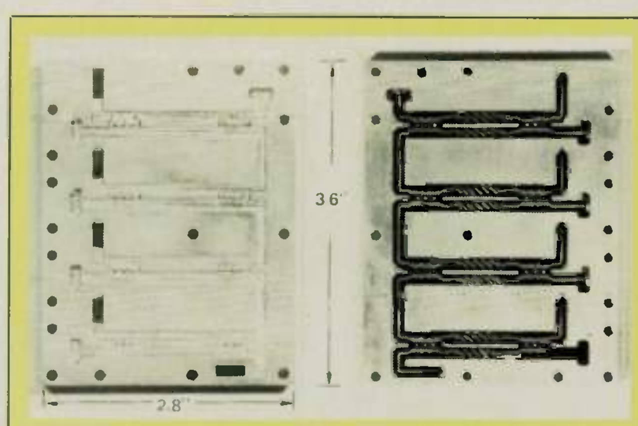

Fig. 9 Opened multiplexer assembly.

termined by the upper and lower band edge frequencies of section 3 and 2 respectively.

Power rejected by the filters of section 4 goes into load L5. Reflected power from any section is routed back either to the input or the loaded port of the previous section, not to its output port. This greatly reduces interactions due to the SWR's of filters as might occur in other types of multiplexers. Measured performance of the multiplexer is shown in **Figure 8. Figure 9** shows the opened multiplexer assembly. In the next phase of this development, mixers will be integrated on the substrate within the spaces between the channel dropping sections. This will result in a channelized downconverter front-end for K_a band with no increase in total size of the substrate. If performance above 41 GHz is required, sections 1 and 2 could be interchanged and a smaller width channel used for the 38-42 GHz section.

CONCLUSIONS

The methods used for these suspended substrate designs should prove useful for channel configurations and substrates other than those presented.

A small, low cost, low loss multiplexer has been demonstrated for the 26-40 GHz band. Continued development will result in a compact channelized down converter front-end.

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(from page 71) IMPATT

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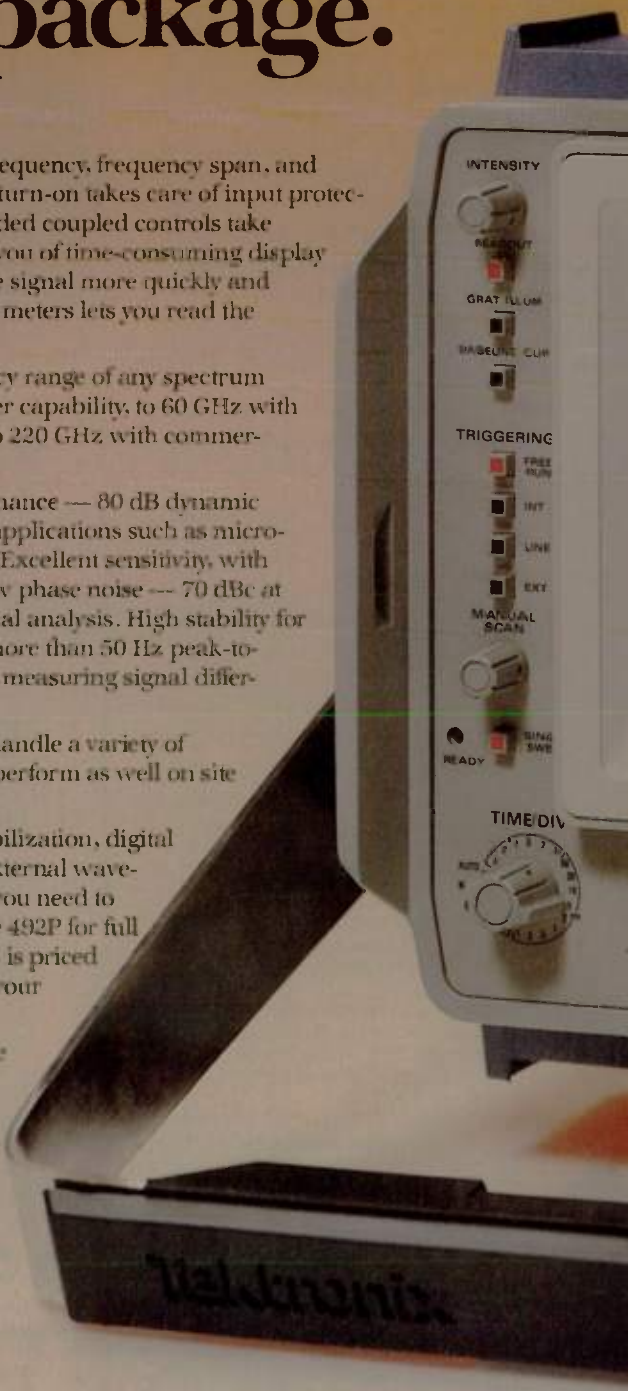

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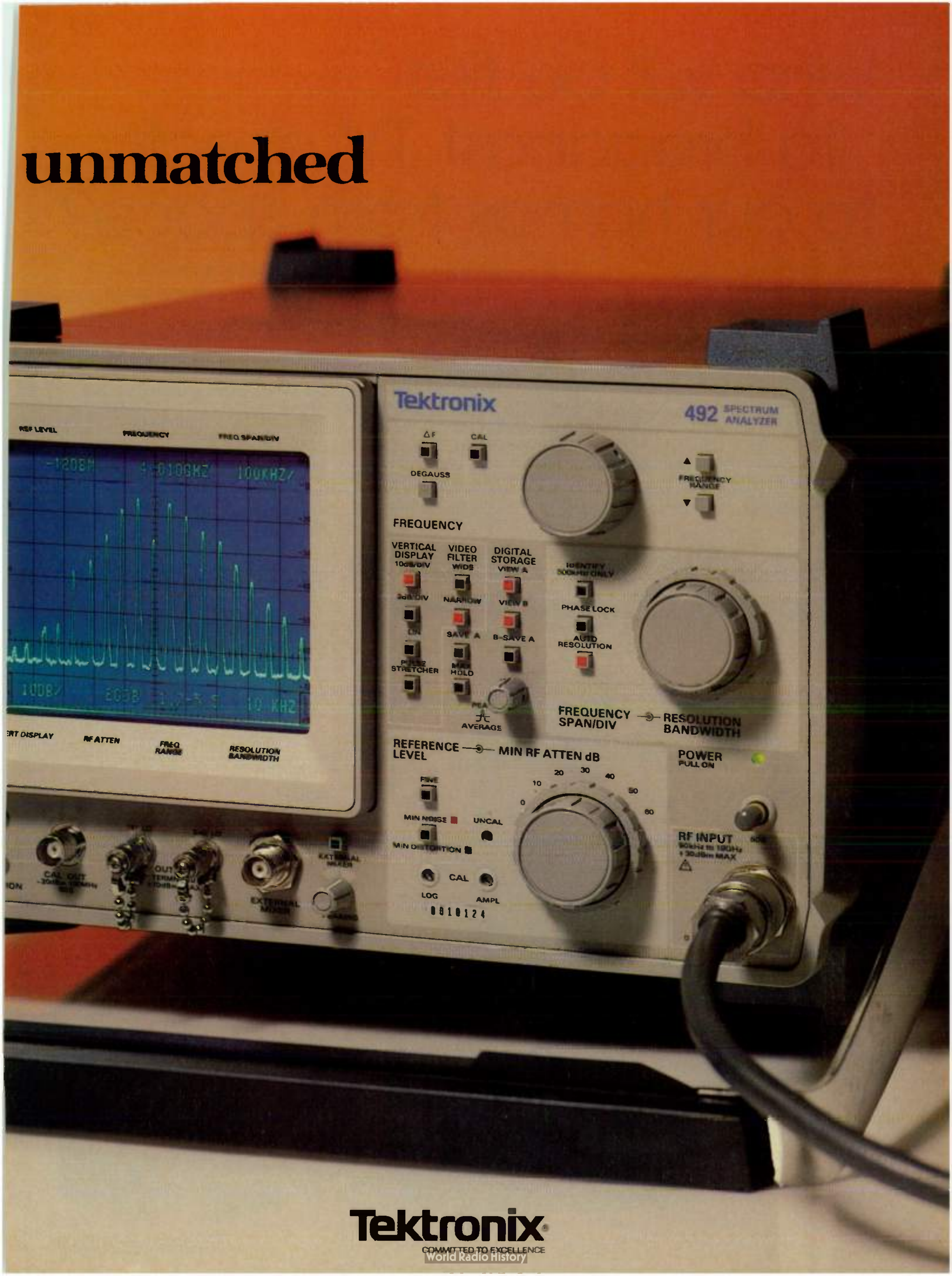
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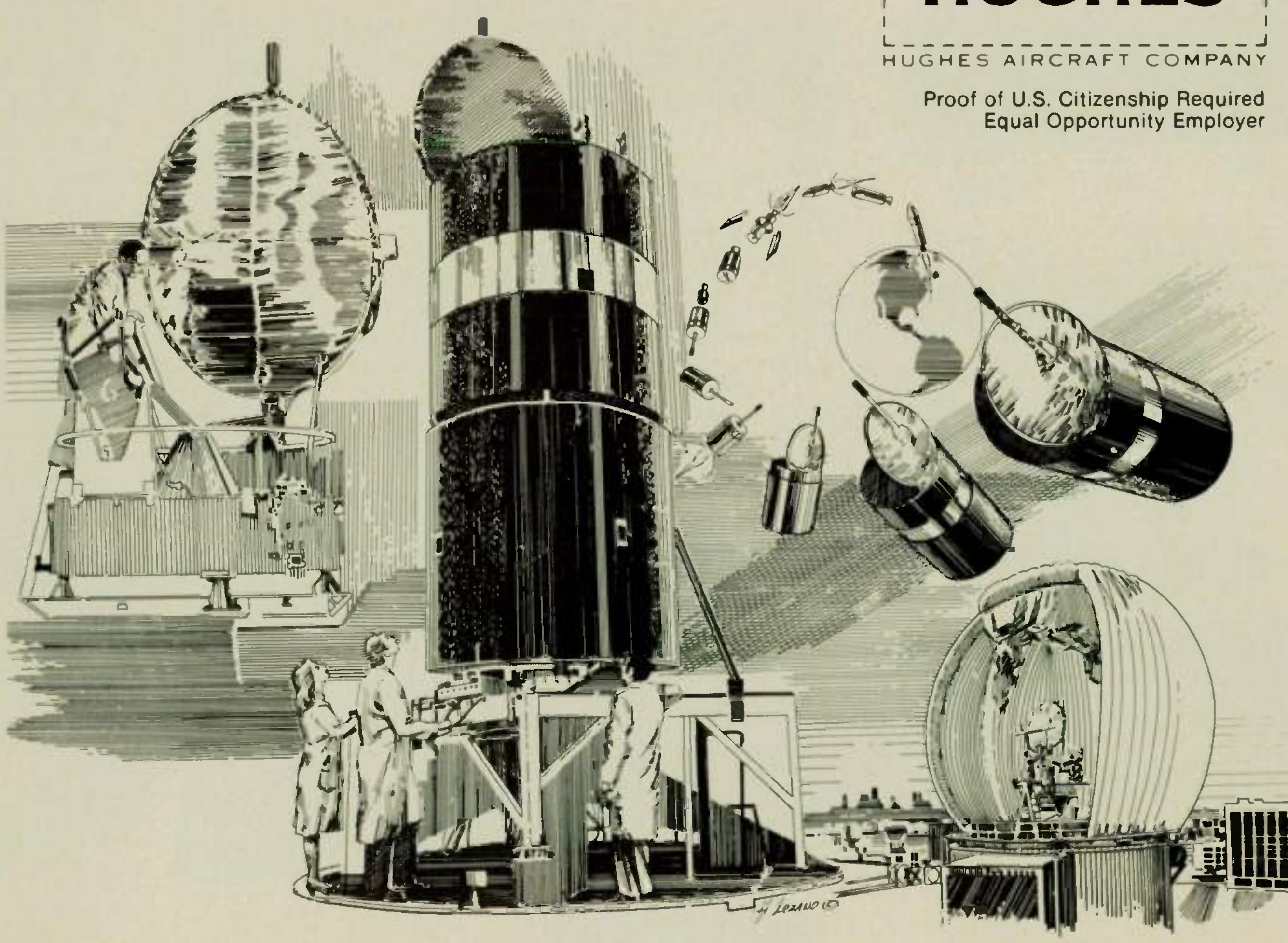

I.C.'s for 94 GHz Radar Applications

In Dielectric Image Guide

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Electron Dynamics Div.
Torrance, CA

Dielectric image guide integrated circuit balanced mixers and Gunn oscillators have been developed at 94 GHz frequencies, obtaining a conversion loss of 11.1 dB for the balanced mixer and an output power of +5 dBm for the Gunn oscillator. These components have been incorporated into several 94 GHz radar test units, including a noncoherent pulsed radar and a frequency-shift keyed CW radar for digital signal processing. This paper describes the development and tests of the integrated circuits and the radar test units. *

INTRODUCTION

Tactical weapon systems demand small radars with high resolution and tracking accuracy, adverse weather penetration, anti-jam capability and low probability of intercept. 94 GHz radar and sensors have been given great attention recently because of their short wavelength (~3mm), narrow beamwidth associated with small antenna, and large bandwidth for frequency spreading. However, it has always been a challenge to develop extremely low cost, rugged and compact 94 GHz components for the tactical radars and sensors. Conventional 94 GHz waveguide circuits cannot meet the above requirements because of the small waveguide size (0.050 x 0.100 inch) and the associated tight dimensional tolerance. Millimeter-wave integrated circuits (MMIC's) are emerging as potential solutions. For many years, US Army ERADCOM (previously known as ECOM) has been supporting the development of MMIC's, and has been the driving force behind the development of the dielectric guide MMIC's, which promise low propagation loss in the dielectric guiding structures at millimeter-wave frequencies.

The dielectric guide MMIC's have demonstrated their circuit's performance in both the United States and abroad. Published results include passive components such as couplers, filters, phase shifters, detectors, and balanced mixers; active components such as IMPATT and Gunn oscillators; and antennas such as electronic scanned array. System applications of dielectric guide MMIC's have also been demonstrated including the 60/70 GHz communication modules, the binocular radios, and radar front ends.[1-8]

This paper presents our recent results of dielectric image guide balanced mixer and Gunn oscillator development and the 94 GHz radars using the dielectric image guide front ends. The 94 GHz radars have been fabricated to determine the dielectric image guide MMIC front end performance by demonstrating various radar waveforms, including a short pulsed radar, a frequency-modulated CW radar, and a frequency-shifted keyed (FSK) radar for digital signal processing. Only the short pulsed radar and the FSK radar will be reported here.

BASIC RADAR FRONT END MMIC COMPONENTS

The basic radar front end components include the MMIC Gunn oscillator and the MMIC balanced mixer. Figure 1 is a photograph of an image guide Gunn oscillator. A packaged GaAs Gunn diode is placed in a hole formed in a dielectric image guide. The image guide material is hot pressed boron nitride (BN), which has low loss characteristics (loss tangent better than 0.001) at 94 GHz, and is thermally stable. Pyrolitic boron nitride has also been used for the MMIC Gunn oscillator. Image guide surfaces surrounding the Gunn diode mount are metallized by sputtering Cr-Au. A thin metal cap with a pin contacts the diode from the top. Since the image guide's top surface around the diode mount is metallized and soldered to the ground (the metal image plane supports the image guide), the top of the diode is grounded. Bias voltage to the Gunn diode was applied through the metal substrate. An anodized aluminum sleeve provides insulation between the diode soldered to a metal stud and the ground. The tip of the image guide is inserted into a waveguide as a simple low loss image-guide-to-metal guide transition. The transition loss is typically no greater than 0.2 dB. Because of the metallized surfaces surrounding the Gunn diode mount, we can not detect any radiated power leakage from the Gunn diode. Performance of the MMIC Gunn oscillators is plotted in Figure 2. Output power of +5 dBm at 92 GHz and +8 dBm at 85.6 GHz was measured.

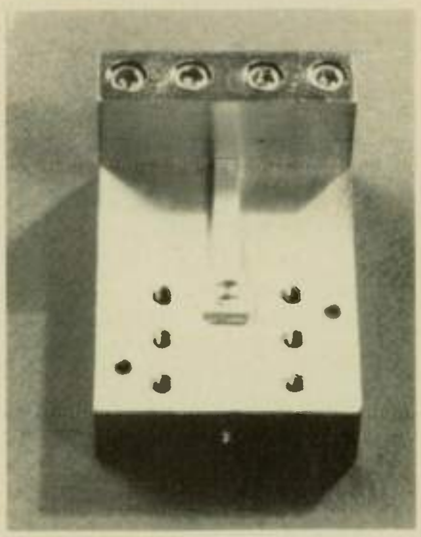

Fig. 1 Dielectric image guide Gunn oscillator.

* Work Performed Under US Army ERADCOM Contract "Millimeter Wave Integrated Circuits," Contract No. DAAB07-78-C-3003, Sept. 28, 1978 - Sept. 28, 1980.

The oscillator circuit is apparently very compact and rugged as shown in Figure 1.

The MMIC Gunn oscillator requires no external metal shielding; therefore, it should be relatively easy for adaptation into various multifunctional circuits. This is especially so for Gunn oscillators. The oscillator circuit quality factor Q is relatively high so that frequency pulling by load mismatch will not seriously affect the oscillator performance. Hence, there will be no need for ferrite isolator at the Gunn oscillator output.

The MMIC Gunn oscillator has also been tested as a self-oscillating mixer. The RF-to-IF conversion characteristics of the MMIC Gunn oscillator was found to follow the characteristics of the waveguide Gunn oscillators.

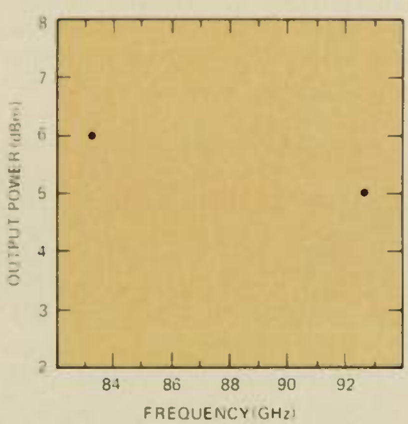

Fig. 2 Dielectric image guide MMIC Gunn oscillator test results.

Conversion loss of about 6 to 12 dB has been measured for MMIC Gunn oscillator. Improvement of the conversion loss, however, is expected by modifying our Gunn bias design so that IF signals from the self-oscillating mixer will not be blocked by the bias circuit. The MMIC Gunn oscillators tested have narrower frequency tuning range (≤ 50 MHz) compared with a metal waveguide Gunn oscillator (~100 MHz). The oscillator frequency jitter is the same as the waveguide oscillator, and is caused mainly in the voltage bias circuit of the Gunn devices.

Figure 3 shows the picture of an image guide MMIC balanced mixer at 94 GHz frequencies. The balanced mixer is formed by coupling the local oscillator power and the signal through a 3 dB hybrid coupler into two beam-lead mixer diodes (Model No. AEL 1309). The coupler is formed by placing two image guides close to each other with a small spacing over a proper coupling length.

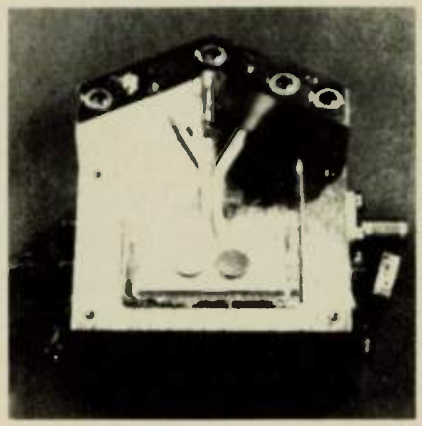
Fig. 3 94 GHz image guide balanced mixer.

Again, BN was used to form the image guide circuit. The beam-lead mixer diodes are mounted at the end of the image guide, leading into an air cavity. BN surfaces surrounding this cavity are metallized, and a thin cap soldered over each cavity to prevent radiated power from the mixer diode mount. Intermediate frequency (IF) output of the mixer diodes are summed into an IF amplifier placed under the substrate surface. The IF amplifier is in TO-8 case package with a bandwidth from 5 to 500 MHz. Since the beam-lead mixer is mounted in an air cavity surrounded by metallized BN, the cavity dimensions affect the balance mixer's performance. Another factor that also affects the balanced mixer's performance is the transition of BN image guide

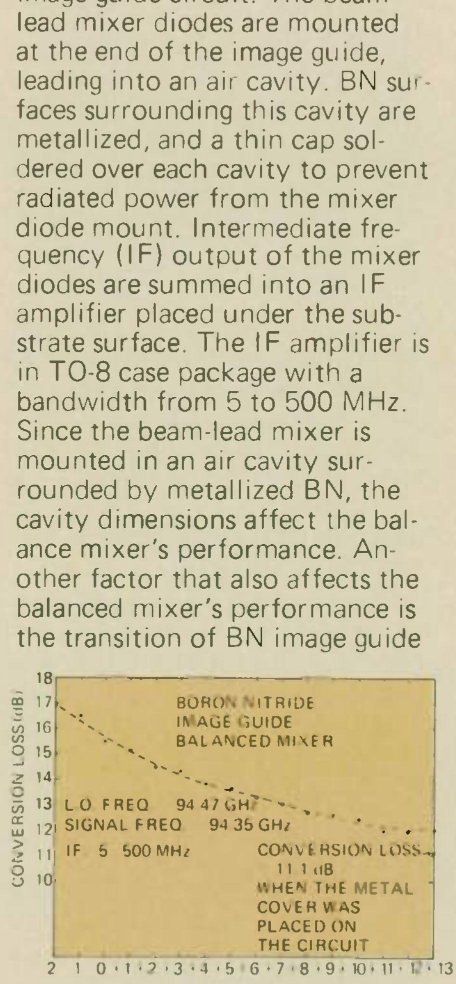

Fig. 4 94 GHz dielectric image guide balanced mixer conversion loss data.

into the metallized mixer cavity. These factors have not yet been thoroughly investigated. The third factor is the beam-lead mixer diode itself. Cutoff frequency of the diode must be high, at least 10 times the operating frequency of the balanced mixer. For 94 GHz applications, parasitic circuit parameters of the beam lead diodes must be minimized. The diode type AEI 1309 seems to be better than the diode type 1308, since the latter always resulted in MMIC balanced mixers of larger conversion loss (13 to 14 dB).

Figure 4 is the measured single-sideband conversion loss of the balanced mixer, pumped with a local oscillator at 94.47 GHz. AEI 1309 mixer diodes were used in the balance mixer. The data was taken as the local oscillator power varied from 12 to 13 dBm. Conversion loss was 12 dB at pump power of 11 dBm. The large pump power required has been typical of the AEI beam lead mixer diodes we have tested. The mixer conversion loss was improved to 11.1 dB when a metal cover was placed over the mixer for sealing. This indicates that radiation loss might have existed at the 3 dB coupler and was suppressed by the metal cover. It was found that diodes were also not matched. After examination of their I-V characteristics, we conclude that better conversion loss results can be achieved with a pair of matched diodes. Instantaneous bandwidth of the MMIC balanced mixer measured was typically in the order of 3 to 4 GHz. However, at higher conversion loss (13 to 14 dB), bandwidth as large as 10 GHz was observed.

MMIC 94 GHz RADARS

Under the ERADCOM contract, several small 94 GHz radars have been developed to demonstrate the MMIC front end performance and for laboratory and short-range field tests. Only a short-pulsed noncoherent radar and a frequency-shift keying (FSK) radar will be described here. All the radars are battery-operated for quick field tests.

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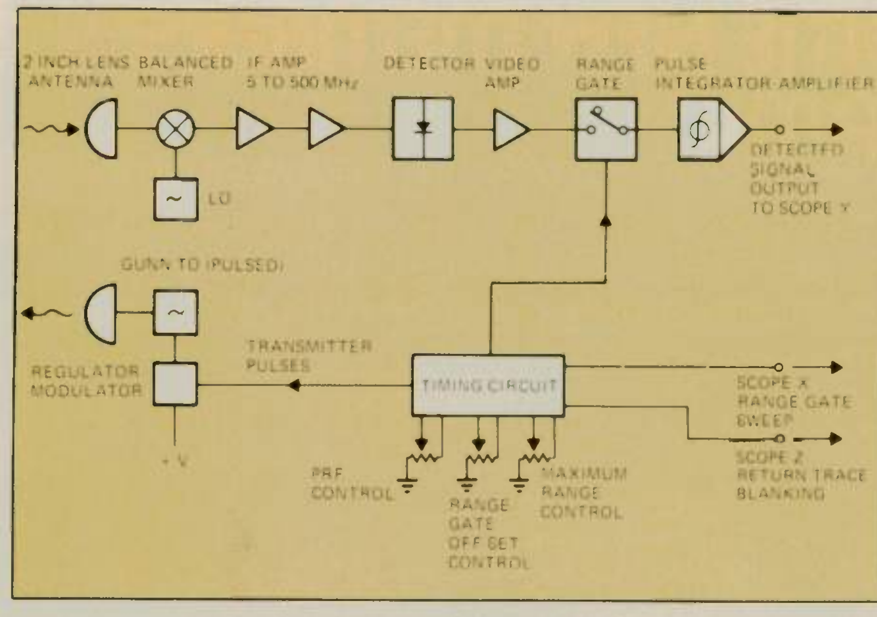

Fig. 5 Test radar unit system block diagram.

Figure 5 is the system block diagram of the 94 GHz pulsed radar. The radar has two separate lens antennas for the transmitter and the receiver. The two 2-inch diameter antennas are separated by about 6 inches so that their main beams overlap at a distance of 6 feet from the radar. The radar transmitter is a CW Gunn source. Pulsed power output is obtained by on-off keying the

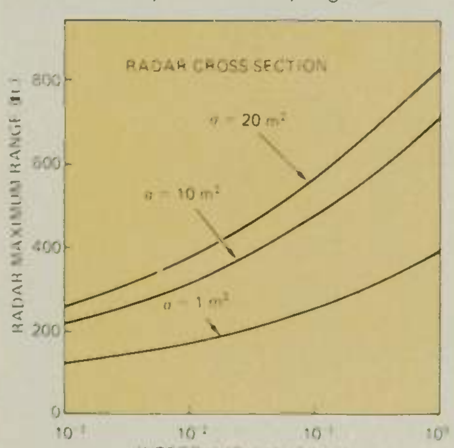

Fig. 6 Estimated radar range based on parameters listed in Table 2-1.

oscillator. Pulses as short as 30 nsec and as long as a few microseconds can be generated. The pulsewidth and the pulse repetition frequencies are controlled by the timing circuit. The radar receiver consists of a balanced mixer pumped by a local oscillator. The target returned signals in the 5 to 500 MHz IF are detected by a video detector. An integrator/amplifier integrates these pulses in a duration of one millisecond to one second, depending on the integrator time constant. A range gate, placed after the detector and the video amplifier, is controlled by the timing circuit. The timing circuit sweeps the radar range gate from a target distance of 30 to about 300 feet with a minimum gate width of 30 nsec for target resolution of 15 feet.

Multiple target display is accomplished with a standard oscilloscope. The integrated signal output connects to scope input Y, and the swept range gate to scope input X. The timing circuit generates a blanking waveform to the scope input Z for blocking the scope retrace. Table 1 lists

TABLE 1

RADAR PARAMETERS

Frequency	94 GHz
Antennas	31 dB gain
(2 inch-diameter antenna)	
Transmitter	
Power	10 mW
Pulse width	50 nsec
PRF	1 MHz
Receiver	
Noise	15 dB
Bandwidth	500 MHz
Integration Loss	3 dB
Probability of Detection	0.999
Probability of False Alarm	10^{-5}

the radar parameters. With a chosen probability of detection, 0.999, and a probability of false alarm, 10^{-5}, the radar with 50 nsec short pulse, 1 MHz pulse repetition frequency and 10 mW transmitter power is estimated to have the ranges as shown in Figure 6. The range calculation is based on post-detection (noncoherent) pulse integration with a wide front end bandwidth of 500 MHz. Figure 7 shows the 94 GHz pulsed radar unit with the battery pack. The radar is housed in a standard instrumentation box of 3.5 (height) x 8.4 (width) x 12.7 (length) inches, and is mounted on a tripod for field tests. Beam patterns of the antennas as shown in Figure 7 have been measured. They have a beamwidth of 4.2 degrees and a maximum sidelobe level of −20 dB from the peak of the mainlobe.

Fig. 7 94 GHz test radar unit with battery pack.

FSK radar has also been under development. The FSK radar consists of frequency-shifted pulses with pulsewidths from 35 to 100 nsec to provide a target resolution of 17.5 to 50 feet. Using an 8-bit shift register, a radar code consisting of $2^8 - 1$ (or 255 bits) can be generated. For a bit length of 50 nsec, a code length will be 255 x 50 nsec, corresponding to an unambiguous radar range detection of 6,375 feet (=25 feet × 255). The code length also determines the dynamic range of the radar. For a code length of 255, a compression ratio of 24 dB is expected in the radar receiver. The FSK radar system block diagram is shown in Figure 8. An FSK code generator drives a modulator circuit, which in turn modulates the Gunn transmitter oscillator bias voltage to generate FSK radar waveforms. A

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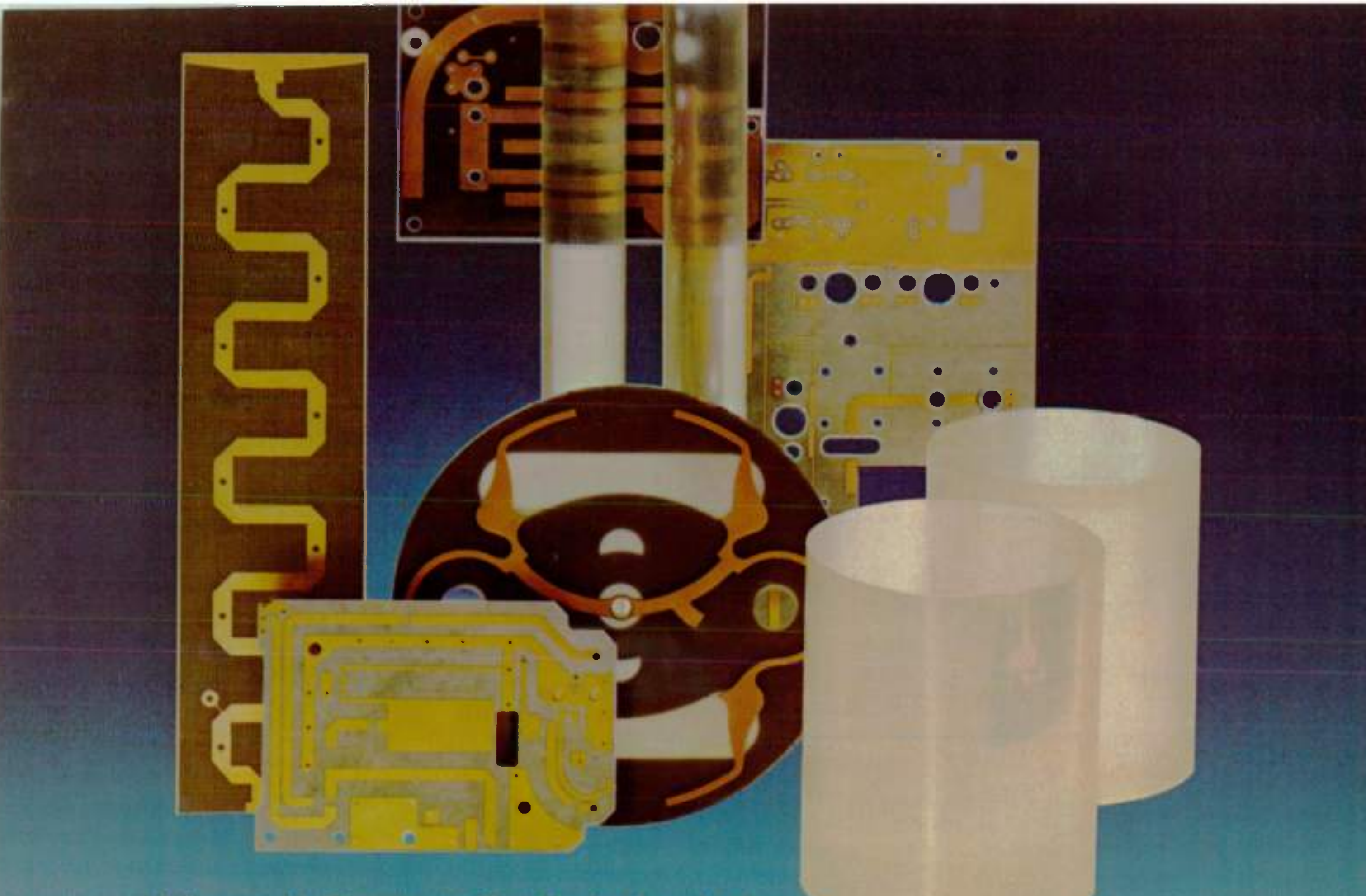

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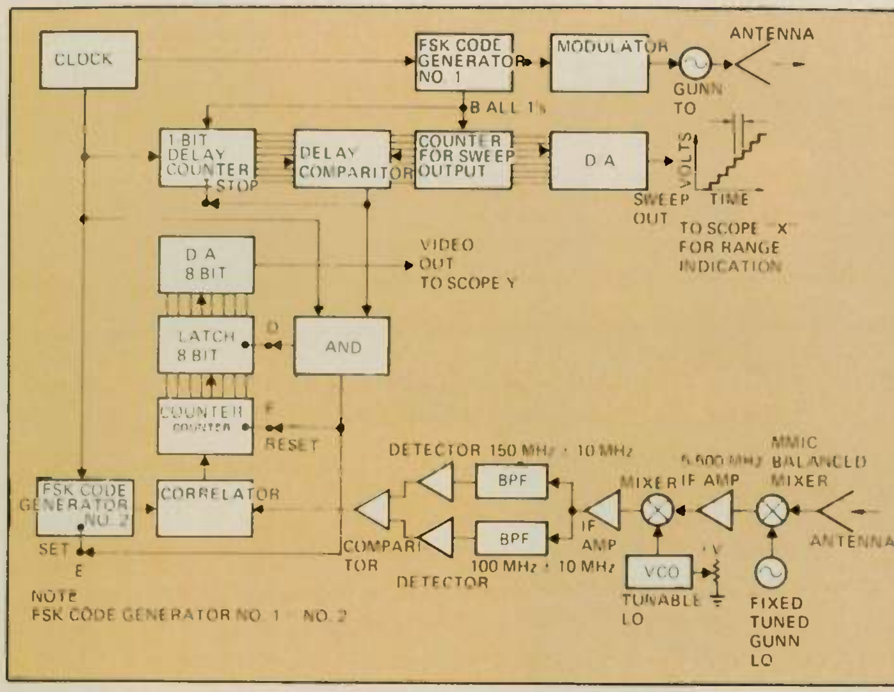


Fig. 8 FSK radar system block diagram.

4-inch diameter lens antenna is used for the transmitter and for the receiver. The target returned signals are detected by the MMIC balanced mixer pumped by a MMIC LO with an IF bandwidth of 5 to 500 MHz.

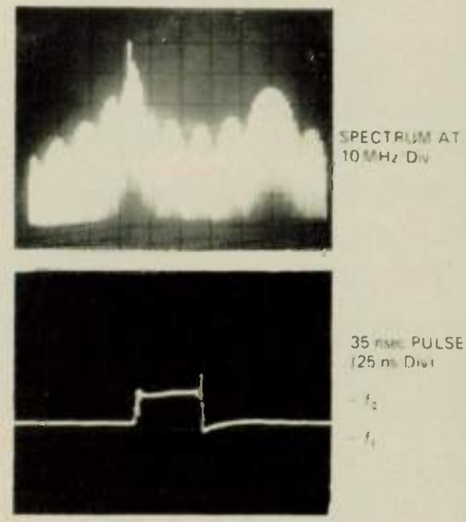
A second downconverter, a double balanced mixer pumped by a turnable voltage-controlled oscillator (VCO), converts the signals to an IF covering from 70 to 180 MHz. This IF is further diplexed into two frequency channels with a bandwidth of 20 MHz centering at 100 and 150 MHz, respectively. The diplexer is required to detect the two FSK frequency levels. After detection with video detectors, the two signals are then summed in a comparator to recover FSK digital signals.

A reference FSK code generator, progressively delayed with respect to the clock pulses, compares the coded signal with the reference code in the correlator. The correlator then drives the counter to provide the detected signal which is converted into analog pulsed signals after D/A conversion. Display of this signal is provided in the scope Y-terminal, and a digital sweep signal in the scope X-terminal provides the range gate sweep to indicate the target range.

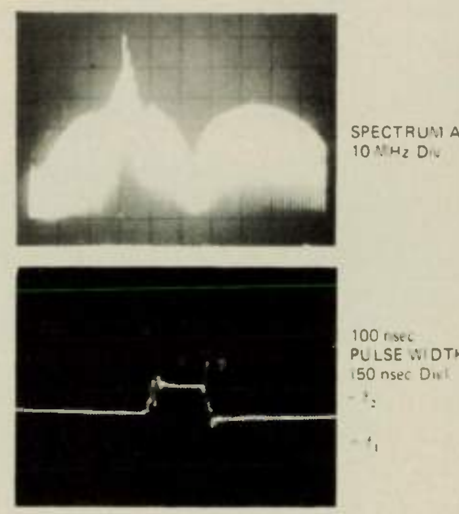
Figure 9 shows the FSK radar mounted on a tripod. The two lens antennas are clearly shown. The antenna beam patterns have been measured; the experimental beamwidth is about 2 degrees. Lobes of the antennas are down by at least 19 dB from the main-lobe peak gain.

The FSK modulation-pulsed waveforms are recorded in **Figure 10**. **Figure 10 (a)** shows a 100 nsec pulsewidth modulation, indicating a frequency separation of 50 MHz and a pulse repetition frequency of 1.8 MHz. The modulated Gunn oscillator spectrum is also shown, indicating the 50 MHz separation. **Figure 10 (b)** shows the modulation pulse waveform and spectrum of a 35 nsec pulse at a repetition rate of 1.73 MHz.

Fig. 9 94 GHz FSK radar mounted on tripod for field test.



(a) 100 nsec pulse width modulation.



(b) 35 nsec pulse width modulation.

Fig. 10 FSK radar modulation waveforms.

CONCLUSIONS

94 GHz dielectric image guide balanced mixers and Gunn oscillators have been developed, and their performance demonstrated in several small 94 GHz solid state radars of various radar waveforms. The performance of the balanced mixers is limited by the availability of high cutoff frequency beam mixer diodes. Gunn oscillator performance is circuit-limited presently, but higher output is expected with dielectric guide circuit improvement.

The demonstration of these image guide components in sev-

(continued on page 88)

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eral small 94 GHz radars with various radar waveforms indicates the feasibility of utilizing image guide integrated circuit for tactical weapons and sensor applications. These circuits are attractive because they offer a low cost and rugged design for mass production. At present, the circuits are ready for multifunctional integration to form compact and rugged front ends.

ACKNOWLEDGEMENT

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8. Chang, Y., and L. T. Yuan, "Millimeter Wave Binocular Radio," *Microwave Journal*, Vol. 23, March 1980, pp. 31-36.

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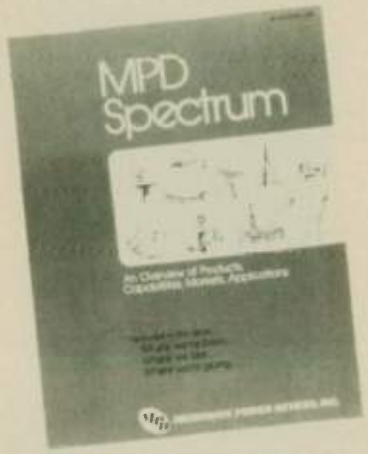

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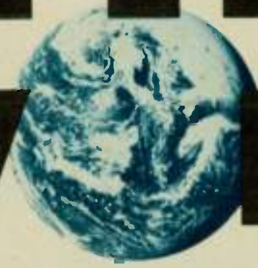
The wrld

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WJ-6855 Series	4 to 8 GHz	WJ-6859 Series	12 to 20 GHz

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Watkins-Johnson—U.S.A.: • **California.** San Jose (408) 262-1411; El Segundo (213) 640-1980 • **Florida.** Fort Walton Beach (904) 863-4191 • **Georgia.** Atlanta (404) 45[illegible] • **Illinois.** Palatine (312) 991-0291 • **District of Columbia.** Gaithersburg, MD (301) 948-7550 • **Massachusetts.** Lexington (617) 861-1580 • **Missouri.** Bridgeton (314) 291-6532 • **Ohio.** Fairborn (513) 426-8303 • **Texas.** Dallas (214) 234-5396 • **United Kingdom:** Dedworth Rd., Oakley Green, Windsor, Berkshire SL4 4LH • **Tel: Windsor 69241** • Cable: WJUKW-WINDSOR • Telex: 847578 • **Germany.** Federal Republic of: Manzingerweg 7, 8000 Muenchen 60 • Tel: (089) 836011 • Cable: WJDBM-MUENCHEN • Telex: 529401 • **Italy:** Piazza G. Marconi, 25 00144 Roma-EUR • Tel: 59 45 54 • Cable: WJROM-I • Telex: 612278

(from page 92) UPDATE

AUTOMATED SCALAR NETWORK ANALYZER SYSTEMS 10 MHz-40 GHz

A new family of Automated Scalar Network Analyzer Systems covering the 10 MHz to 40 GHz range is described in this 16-page brochure. Included are specifications and application notes for measuring return loss, transmission loss/gain and power in coax or waveguide. Block diagrams show how the system consisting of a new programmable sweep generator, network analyzer, desktop controller and precision components with 40 dB directivity. **WILTRON Company, 825 E. Middlefield Rd., Mountain View, CA 94043. Tel: (415) 969-6500. Circle 142.**

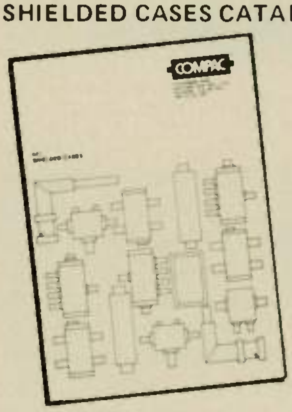
FERRITE DEVICES AND FILTERS BROCHURE

A brochure detailing a complete product line of ferrite devices and filters is offered – including complete table specifications and updated information on applications for isolators, circulators, filters and special designs. This 10-page catalog includes product photographs, performance curves and schematic diagrams. **UTE Microwave Inc., 3500 Sunset Ave., Asbury Park, NJ 07712. Tel: (201) 922-1009; TLX: 132-461 UTE APK. Circle 143.**

TUNNEL DIODE LINE BROCHURE

A complete line of tunnel diodes for amplifiers, detectors, mixers and switches is described. Each product section includes specifications, performance curves and applications, plus product features and schematic drawing. Package outlines for type 23 package provided. **Custom Components, Inc., Box 334, Lebanon, NJ 08833. Tel: (201) 236-2128; TLX: 132-445. Circle 144.**

RFI SHIELDED CASES CATALOG



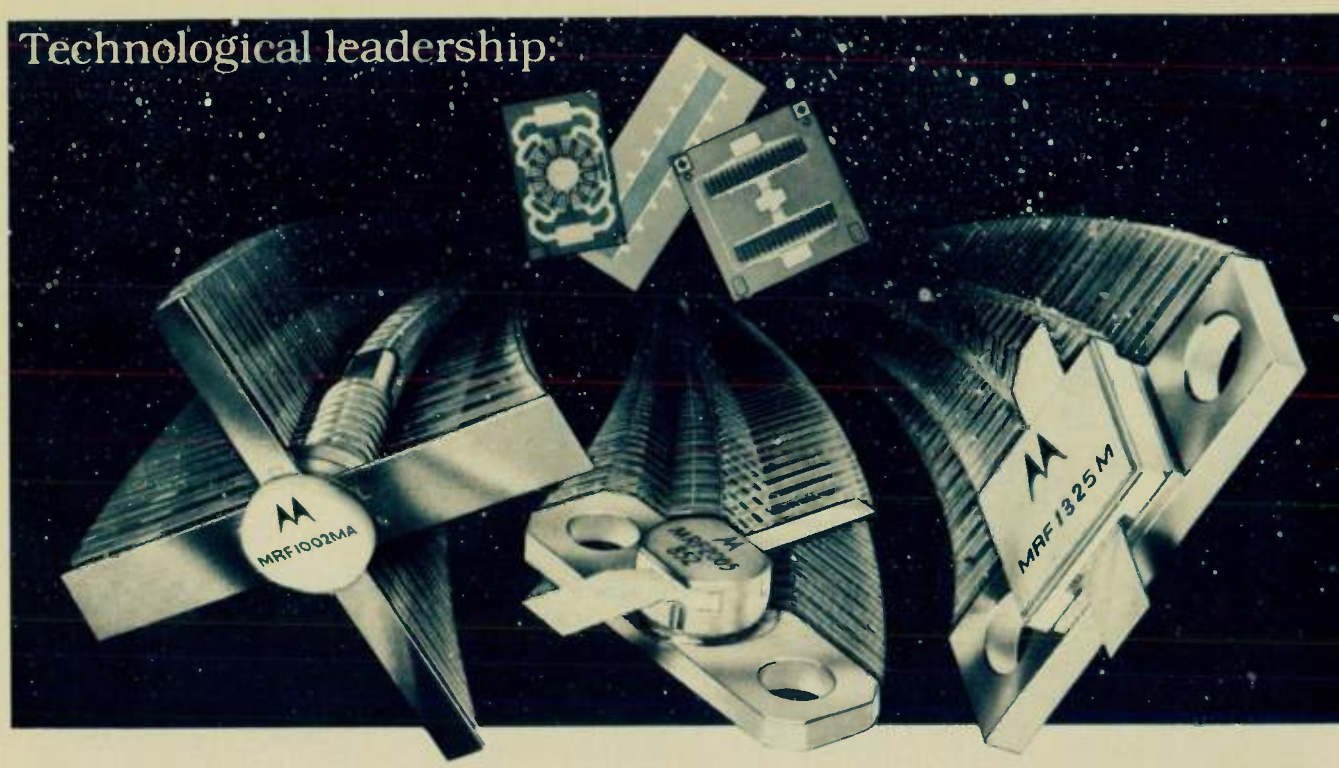
An 8-page catalog describes low cost RFI shielded cases and accessories. Catalog contains photos and drawings describing a variety of blank cases, standard size cases and a custom series plus accessories. It also features the RFT series which offers greater shielding effectiveness. Cases are effective from 60 to > 100 dB at 100 MHz and are available with an optional nickel plate finish. Various configurations are noted in the numerous outline drawings. A series of die cast boxes and a comprehensive group of gaskets are also shown. **COMPAC, 279 Skidmore Road, Deer Park, NY 11729. Tel: (516) 667-3933. Circle 145.**

MICROWAVE COMPONENTS AND NOISE CATALOGS

Updated catalogs featuring microwave components and solid state noise products has been issued. The Microwave Components Catalog #MC/180 features an extended line of coaxial switches and includes waveguide switches, dummy loads, crystal detectors, bolometers, and RF micropotentiometers. The Noise Catalog features solid state noise diodes and sundries. Ask for Catalog #SSN5/681. Both catalogs are available upon request. For further information, write or call: **MICRONETICS, INC., 36 Oak Street, Norwood, NJ 07648, Tel: (201) 767-1320; TWX: 710-991-9603. Circle 146.**

PRODUCT LINE CATALOG

The 1980 catalog offers over 200 products in the areas of amplifiers, mixers, power dividers, hybrids, attenuators and directional couplers. This 272-page booklet includes detailed specifications which are guaranteed over temperature, typical performance curves on each device including (S parameters and Smith Charts) and small quantity pricing. The catalog features products including termination insensitive mixers, thin film amplifiers and passive devices. **Adams-Russell Co., Inc., Anzac Division, 80 Cambridge Street, Burlington, MA 01803, Tel: (617) 273-3330, TWX: 710-332-0258. Circle 153.** *(continued on page 96)*



Come to one reliable source for all microwave power from 0.2 W to 325 W. Motorola.

Motorola is the well-established, quality-in-quantity producer of RF small signal and power devices and modules (15 million units shipped this year). We have the backing of a $4.2 billion corporation. We're a price leader. We have an excellent quality record. And we're worldwide.

It's only natural for Motorola to be the leading contender in the exploding, high-technology microwave power arena. Our product lineup shows capability. Our die geometries show reality. And our technology shows up.

All units provide guaranteed performance — they're 100% tested including load mismatch at all phase angles with 10:1 VSWR and offer gold-metallization and emitter ballasting for long life and resistance to metal migration.

Later, we'll offer higher-power, 400 W, short pulse units and long-pulse microwave devices in volume. And, we provide plastic and ceramic drivers for 1.5 to 5 GHz application.

Contact Motorola Semiconductor Products, Inc., P.O. Box 20912, Phoenix, AZ 85036 or call (602) 244-6394 for data on microwave power for all your

Device Type	P_{out} Output Power Watts	G_{PB} Power Gain dB Min	V_{CC} Supply Voltage Volts
.96-1.215 GHZ			
MRF1000MA,B	0.20	10	18
MRF1002MA,B	2.0	10	35
MRF1004MA,B	4.0	10	35
MRF1008MA,B	8.0	10	50
MRF1015MA,B	15	10	50
MRF1035MA,B	35	10	50
MRF1090MA,B	90	10	50
MRF1150MA,B	150	7.8	50
MRF1250M	250	6.0	50
MRF1325M	325	6.0	50
1.7-2.3 GHZ			
MRF2001M	1.0	8.5	24
MRF2003M	3.0	8.0	24
MRF2005M	5.0	7.5	24
MRF2010M	10	7.0	24
MRF2016M	16	6.5	24
2.0 GHZ			
MRF2001,B	1.0	9.0	28
MRF2003,B	3.0	7.8	28
MRF2005,B	5.0	8.0	28
MRF2010,B	10	6.0	28

Innovative systems through silicon.

MOTOROLA INC.

TO: Motorola Semiconductor Products Inc., P.O. Box 20912, Phoenix, AZ 85036.

Please send me data on microwave power devices.

96 MJ 6/81

Name ______________________

Title ______________ Tel.: (   ) ________

Company ______________________

Address ______________ Mail Drop ________

City ______________________

State ______________ ZIP ________

CIRCLE 62 ON READER SERVICE CARD

(from page 94) UPDATE

COMPONENTS AND SUBSYSTEMS FOR SIGNAL PROCESSING CATALOG

A 48-page, 2-color brochure covers line of broadband electronic components and subsystems and is subdivided into 17 basic product lines with extensive tabular data on each product. Typical categories include: power dividers/combiners; directional couplers; pulse, biphase and quadriphase modulators; double balanced mixers; power sensors; quadripole networks and comparators; subsystem capabilities, etc. **Olektron Corporation, 61 Sutton Rd., Webster, MA. Tel: (617) 943-7440. Circle 149.**

ATTENUATORS AND OTHER COMPONENTS

A new 40-page attenuator catalog, No. 82-13, has just been released by Alan Industries. This catalog features high performance fixed attenuators and terminations operating from dc to 18 GHz and step attenuators that operate by dial, toggle, push-button or rocker switches, from dc to 2 GHz. Other components are programmable attenuators, broadband detectors for video through UHF, return loss bridges, RF fuses and reactive power dividers/combiners. **Alan Industries, Inc., 745 Greenway Drive, P.O. Box 1203, Columbus, IN 47201. Bill Kennedy, Tel: 812-372-8869. Circle 152.**

COAXIAL CABLE ASSEMBLIES CATALOG

A 24-page illustrated catalog describes the line of precision coaxial cable assemblies for ECM, avionic, and ground-based applications. Complete mechanical, environmental, and electronic specifications are given for each cable type along with large, easy-to-interpolate curves of insertion loss and power handling. Manufacturing, testing, and special services are described as well as details of cable-construction.

Adams Russell
ANTENNA & MICROWAVE DIVISION

Haverhill Road, Amesbury, MA. 01913. Tel: (617) 388-5210, (617) 665-2750, TWX: 710-347-6360. Circle 151.

VARACTOR FREQUENCY MULTIPLIERS

This catalogue describes a line of varactor frequency multipliers from 10 - 12 GHz which includes multiplications of up to 16 and bandwidths to 35%. The catalogue gives examples of custom designs with multiplication factors of as much as 240 and output frequencies to 15 GHz. Passive and active multiplier chains are described at low, medium and high power. **A. I. Grayzel Inc., 3 Common Street, Waltham, MA 02154. Tel: 617-893 4210. Circle 154.**

SOLID STATE DEVICES AND COMPONENTS CATALOG

Product Guide for 1981 describes standard microwave components and semiconductors, including GaAs Power FET's, bipolar power transistors and amplifiers and GaAs FET amplifiers, noise sources. GaAs FET device line has capabilities through K_u band and the silicon bipolar power transistor product line offers performance through 6.0 GHz for both CW and pulse applications. The power amplifier product lines offer performance from L through X band and noise sources are designed for instrument applications and system applications. **Microwave Semiconductor Corporation — A Siemens Company, Somerset, NJ. Tel: (201) 469-3311. Circle 158.**

NEW LITERATURE

Short Form Catalog for 1981 now offered which describe mixers, mixer/preamplifiers, wide band, low noise amplifiers, C, X and K_u-Band communication converters and frequency translators for satellite communications applications. **MITEQ Inc., 100 Ricefield Lane, Hauppauge, NY 11787. (516) 543-8873. Circle 159.**

Product Feature

Broadband ECM Antenna

ADAMS RUSSELL ANTENNA & MICROWAVE DIVISION
Amesbury, MA

Broad frequency coverage with excellent control over the axial ratio is offered by the AN-364 series of circularly polarized horn antennas for ECM applications. The new series of antennas utilizes a quadridged horn design with lenses on the aperture to provide broad azimuth coverage and a "compensated" elevation aperture to control axial ratio at wide azimuth and elevation angles.

These 3:1 bandwidth antennas are available with double-ridged waveguide inputs or high-power coaxial to waveguide transitions. Figure 1 shows return loss vs. frequency of an antenna configured with a coax to waveguide transition.

The AN-364 antennas are unique in that radiation pattern shaping is achieved by the utilization of a finned elevation aperture combined with a low loss dielectric lens for azimuth pattern control. Figure 2 shows a typical principle plane azimuth pattern. Nominal azimuth half power beamwidth is 120° with an elevation half power beamwidth of 65°. Radiation pattern coverage at the 6 dB level is nominally 160° for azimuth and 90° for elevation.

The basic antenna design permits utilization in any 3:1 bandwidth portion of a broad range of frequencies. The AN-364 is the first of a new series of antennas that satisfies system level requirements of wide frequency bandwidths with specially shaped radiation patterns (sample radiation patterns are available upon request). Contact: David W. Ryan, (617) 388-5210.

Circle 143 on Reader Service Card

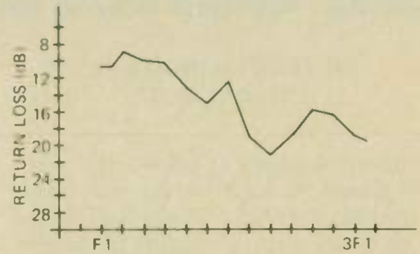


Fig. 1 SWR vs frequency.

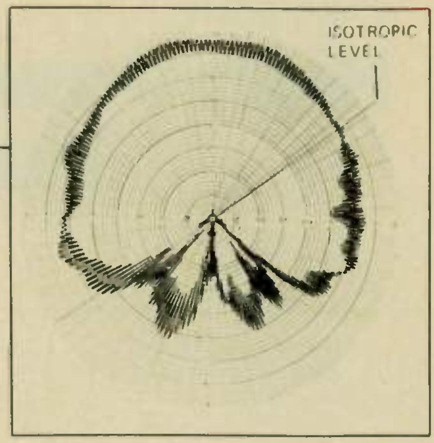
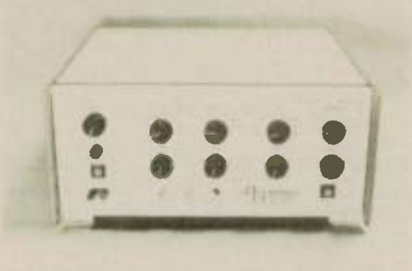


Fig. 2 Typical azimuth pattern.

AN-364 Antenna.

High Power Constant Current Pulse Generator

AD-TECH MICROWAVE INC.
Scottsdale, AZ



AT-SM33 Pulse Generator.

The Model AT-SM33 high power, constant current pulse generator is specifically designed to power IMPATT diodes in pulsed IMPATT diode oscillator and amplifier applications. This new generator incorporates features which make it versatile enough to serve in research and development laboratories as well as in production tests.

The wide ranges of pulse repetition frequencies, pulse widths, duty cycle options and output currents make the AT-SM33 a universal pulse modulator for a variety of pulsed IMPATT diode oscillator and amplifier designs.

Built-in protection circuitry automatically shuts down the output pulse and guards against "burnout" of IMPATT diodes during development and test work. Convenient front panel controls allow the operator to preset pulse repetition frequencies and pulse widths not to be exceeded during the operation of the generator. The output current pulse can be turned off without changing the output current setting and internal circuitry prevents undesireable stretching of the output pulse.

The AT-SM33 can pulse IMPATT's requiring bias voltage anywhere between 0 and 180 V. Short circuits at the output terminals can be handled indefinitely.

An abbreviated list of performance specifications for the AT-SM33 is shown in Table I.

Circle 156 on Reader Service Card

TABLE I

Current: 0-3 amps. peak

Voltage: 0 to +180 V

Output power: 90 W average, 540 W peak

Risetime: 150 ns maximum

Falltime: 200 ns maximum

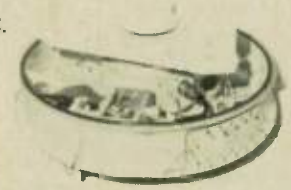
Pulse repetition rates: 8 ranges from 20-100 Hz to 1-3 MHz

Duty cycle: up to 90%

Pulse width: 6 ranges from 0.2-1 μs to 100 - 400 μs

Multi-Octave DF Antenna System

SANDERS ASSOCIATES INC.
Manchester, NH



The Model AS140/TR DF antenna system covers the entire 1-40 GHz band and avoids the problems posed by a rotary joint design for that multi-octave range by employing a reflector which rotates around a fixed spiral feed.

Circular, RH or LH polarization is available. System SWR is typically below 3 throughout the frequency range and axial ratio is 2 dB, typical 5 dB, maximum. Typical antenna gain rises from 5 dB at 1 GHz to 20 dB near 8 GHz and remains above 20 dB through 40 GHz. Typical azimuth and elevation beamwidths are shown. (see **Figure 1**).

The pedestal characteristics for the Model 1 option are shown in **Table 1**.

Circle 157 on Reader Service Card

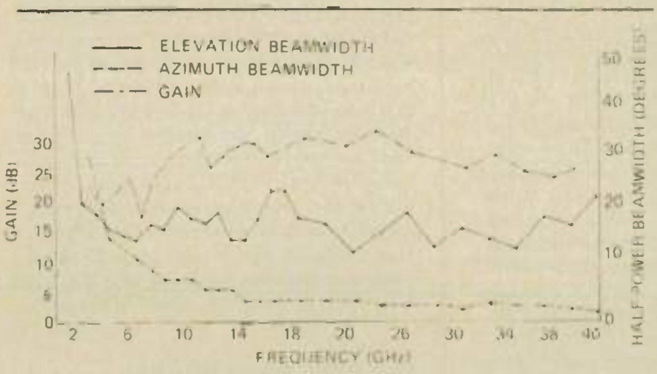


Fig. 1 Gain and beamwidth characteristics.

TABLE I

AS140/TR MODEL 1 PEDESTAL OPTION

Azimuth Travel	Continuous/Sector Scan/Variable
Continuous Scan Mode	0-60 rpm
Sector Scan Mode	For any sector 1 sec. max.
Sector Scan Width	From 30° to 180°
Stop Point Mode Accuracy (0-360°)	±0.5°
Data Output	Digital Format 10 Bits
Power Requirements	27 Vdc @ 5 amps
Controller	TTL Compatible
Weight	24.3 lbs.
Temperature	−54° to +55° C

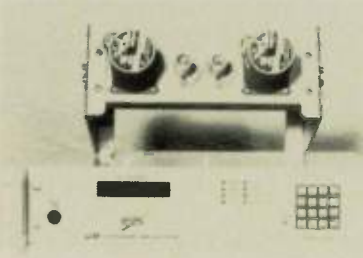
Microwave Products

Components

TUNABLE BANDPASS FILTER

A six-section waveguide tunable filter, Part No. 501850, covers the 13.25-14.5 GHz band. Unit is designed in the TE_{111} cylindrical mode to achieve an insertion loss of only 2.0 dB maximum. Component has a 3 dB BW of 50 MHz minimum and SWR of 1.7 maximum. Filter has rejection of 35 dB minimum at f_o + 50 MHz Model offered with type N female connectors and direct frequency readout. Price: $1990 each. Avail: 10-12 wks ARO. **Coleman Microwave Co., Edinburg, VA 22824. Circle 171.**

C-BAND MICRO-PROCESSOR-CONTROLLED KLYSTRON CHANNEL SELECTOR



A C-Band klystron channel selector, M/N 13049, features micro-processor-controlled channel changes programmble up to 7 days with up to 99 operational entries. Display shows time (day of week, current hour, minutes and seconds) and also channel and frequency selected. Mechanized selector fits all Thomson CSF and Varian C Band klystron 6- or 12-channel communication tubes. RS 422 bus operation provides for computer control of channel selection, programming and for transmitting status information. Rapid tuning achieved by selector's automatic determination of the shortest route to the new channel and immediate turn to the new position. Options include programmable control chassis, mechanized selector and remote control. Del 90 days ARO. **MCL, Inc., LaGrange, IL. Frank Morgan, (312) 354-4350. Circle 173.**

LOW COST SMA ATTENUATOR SETS

Model AT-50-SET/SMA and AT-51-SET/SMA are 50 Ω coaxial SMA attenuator sets. Each set contains a 3, 6, 10, and 20 dB attenuator and they come in both calibrated and uncalibrated models. Attenuation accuracy is 0.5 dB from dc to 1000 MHz and 1 dB from 1000 to 1500 MHz. Sets have an SWR of less than 1.35:1 at 1500 MHz, averaging 1.2:1 over the band. Units can dissipate 0.5 W CW or 1 kW peak power. Design for models uses MIL type connectors and MIL resistors in a MIL plated housing. Price: AT-50-SET/SMA, $69; AT-51-SET/SMA, $64, FOB. Del: stock to 30 days ARO. **Elcom Systems, Inc., Boca Raton, FL. Leonard Pollachek, (305) 994-1774. Circle 165.**

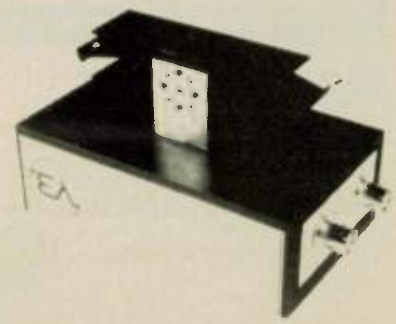
GaAs FET THIN-FILM AMPLIFIERS

Models AWT-6054 and AWT-18057 are aluminum-packaged, thin-film GaAs FET amplifiers which combine 100 mW output power (at 1 dB gain compression) with 35 dB gain. This pair of MIC amplifiers offers 2.0:1 maximum input and output SWR over the 2-6 GHz (Model AWT-6054) and 8-18 GHz (Model AWT-18057) bands. These moderate-power amplifiers provide maximum respective noise figures of 6.0 and 8.0 dB and 1.5 and 2.0 dB gain flatness. Either version operates from 12 Vdc, requiring only 350 mA for operation. Units meet MIL-Q-9858A requirement, MIL-E-5400, MIL-E-16400, and MIL-E-4158 environmental specifications as well as MIL-STD-461 EMI conditions. Size: 1.6 and 2.2 cu. in. Weight 68 and 90 grams respectively. Del: 120 days ARO. **Avantek, Inc., Santa Clara, CA. Peter Campbell, (408) 727-0700. Circle 162.**

WAVEGUIDE AND COAXIAL LNAS

Model AXM123201 (waveguide) and Model AXM 123202 (coaxial) low noise amplifiers provide a 3.0 dB noise figure (290°K) over the 11.7-12.2 GHz satellite communications receive bandwidth. These units achieve the designated noise figure performance by using low loss isolator/single ended GaAs FET input circuitry. The coaxial module features a design, which includes ultra-low noise and balanced input. Input stages of both models exhibit low SWR. Output stages are balanced to insure wide dynamic range and minimum cascading interaction problems. LNA's meet MIL-E-5400, MIL-E-16400 and MIL-E-4158 environments. **Amplica, Inc., Newbury Park, CA. Nick Pena, (805) 498-9671. Circle 160.**

EXPANDED GUNN OSCILLATOR LINE

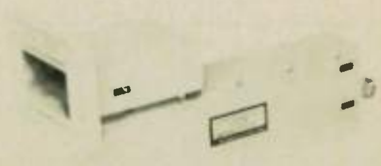


A line of Gunn oscillators is offered for use in the V, E and W bands. Model 4560A for WR 15 covers 50-75 GHz frequency band and has 175 MW power, maximum. Model 4575A for WR 12, spans 60-90 GHz frequency range and has 75 MW power, maximum. Model 4575A for WR 10 covers 80-110 GHz band and offers 20 MW power, maximum. A modulator/regulator option provides both FM and AM (pulse, square wave) modulation. Del: 30-45 days, typ. and 60 days with isolators. **Epsilon Lambda Electronics Corporation, Geneva, IL. Robert M. Knox, (312) 232-9611. Circle 166.**

1 GHz FILTER WITH 2-12.4 GHz STOPBAND

Model F183CS is a tubular low pass filter with a passband of 1020-1100 MHz, a maximum insertion loss of 0.3 dB and a maximum SWR of 1.30 in the passband. Unit features a stop-band of 40 dB minimum between 2150-12.4 GHz (60 dB minimum from 2150-500 MHz). Filter is designed for airborne application up to 100,000 ft. with operational power levels of 4 kW peak and 10 watts average. Component operates from −54° to +95°C and meets environmental requirements of MIL-E-5400, Class 3. Size: 4½" x 4½" with SMA male/female type connectors; SMC connectors also offered. Price: $25.00, 500 pieces. Del 60 days. **Engelmann Microwave Company, Montville, NY. Carl Schraufnagl, (201) 334-5700. Circle 172.**

FET AMPLIFIER COVERS 3.7-4.2 GHz BAND



Model 4055 is a FET amplifier which features a maximum of 55°K noise temperature over the 3.7-4.2 GHz band. This series of amplifiers is thermo-electrically cooled and provides stable performance over the −10° to +50°C temperature range. Size: 10" L x 8½" H x 7" W. Weight 22 lbs. **Mitsubishi Electronics America, Compton, CA. Ric Fochtman, (213) 979-6055. Circle 180.**

COAXIAL ROTARY JOINTS

A series of coaxial rotary joints is offered in short, compact versions. Model 120RK36 has type N connectors, covers the dc to 18 GHz band, and is 1.95" long. The Model 180RS36 has SMA connectors, spans dc to 18 GHz range, and is .95" long. The component No. 120RK36 provides a 1.15 maximum SWR up to 6 GHz (1.25 up to 12.4 GHz) and a 0.2 dB maximum insertion loss up to 10 GHz (0.3 dB up to 12.4 GHz). Model 180RS36 offers a 1.25 maximum SWR up to 10 GHz (1.5 up to 18 GHz) and a 0.2 dB maximum insertion loss up to 10 GHz (0.5 up to 18 GHz). Price: $270 – Model 120RK36; $245 – Model 180RS36. **Microwave Development Laboratories, Inc., Natick, MA. (617) 655-0060. Circle 177.**

DIGITALLY CONTROLLED TUNABLE BANDPASS FILTER

A digitally controlled tunable bandpass filter, No. 4DBT-225/400-3N-F/R, covers the military communications band from 225-400 MHz. Filter has its own built in microprocessor which operates on 28 V for aircraft but has options for other voltages. In addition to its binary control system, the filter can be constructed using other digitally controlled techniques. Unit has an automatic shut down for low power consumption when tuned to a specified frequency. Available in standard ATR rack housings or rack and panel. Price: from $2000. Avail: 6-8 wks. **K & L Microwave Inc., Salisbury, MD. Charles J. Schaub, (301) 749-2424. Circle 168.**

(continued on page 102)

SMA SWEPT RIGHT ANGLE ASSEMBLIES

Series 705970 SMA swept right angle assemblies include male/male and male/female units. Assemblies use CT 141-50 semi-rigid cable for the right angle bend. The SMA plugs use the cable center conductor as the center contact and are gold-plated. Typical SWR for the series is 1.2:1 to 18 GHz. Pt. No. 705976-001 is SMA male/male and Pt. No. 705979-001 is SMA male/female. **Cablewave Systems, Inc., North Haven, CT. Steven Raucci, Jr., (203) 239-3311. Circle 163.**

HI-BAND (400 MHz) CHANNEL BANDPASS

A 3820-series bandpass filter is offered for each TV channel in the 300-400 MHz band. The model is a seven-resonator microwave cavity filter. Unit provides 25 dB minimum rejection to nearest adjacent carriers. Impedance is 75 ohms with F connectors. Mounting is on a 19" rack panel (7" high) and other connectors and impedances available. Price: $335. Del: 10 days. **Microwave Filter Co., Inc., East Syracuse, NY. Emily Bostick, (315) 437-3953. Circle 175.**

COAXIAL CRYSTAL DETECTORS

A line of coaxial crystal detectors covers the 0.01-12.4 GHz frequency spectrum and offers flat frequency response within ± 0.5 dB absolute. Relative matching, excluding bias sensitivity, is within ± 0.2 dB. Unit has output impedance of less than 15 Ω shunted by 10 pf; 100 mW peak or average maximum power. RF input is type N, male and output polarity is normally negative, positive optional. **Micronetics, Inc., Norwood, NJ. Gary Simonyan, (201) 767-1320. Circle 174.**

OSCILLATOR WITH WAVEGUIDE OUTPUT

Model 3004 W is a YIG-tuned oscillator which covers the 26.5-40.0 GHz frequency range with waveguide output, minimum power output of 5 mW. Model has spurious signals of 20 dBc (harmonic), 60 dBc (non-harmonic), residual FM, 10 kHz p-p, frequency stability of 0.01%/°C and power variation vs frequency of 8 dB p-p. **Systron-Donner, Advanced Components Division, Sunnyvale, CA. (408) 735-9660. Circle 183.**

DOUBLE BALANCED MIXER COVERS 5.9-6.4 GHz BAND

Model MD-181 is a double balanced mixer designed to cover the 5.9-6.4 GHz communications band. The model features extremely flat 4.5 dB typical, 6 dB maximum conversion loss, 27 dB L-R isolation and is capable of 0 dBm starved LO operation at zero bias. Price: Units are available in 1-5 qty for $275 for module and $300 for SMA connectorized versions. **Anzac Division, Adams-Russell Co., Inc., Burlington, MA. (617) 273-3333. Circle 161.**

DROP-IN MIXERS

FM Series of drop-in mixers operate in stripline and microstrip designs and provide a double balanced design and thin film beam-lead construction. Operating temperature is −54° to +100°C and storage temperature is −65° to +100°C. The RF power is 100 mW maximum and RF/LO frequency range is from 4-8 GHz to 14.4 - 17.7 GHz, depending on model. Conversion loss ranges from 6.5 to 9.0 dB maximum and LO/RF isolation is from 16-20 dB minimum. The IF bandwidth spans dc to 2.0 GHz and dc to 6.0 GHz with 1 dB compression point ranging from +1 to +2 dBm and 3rd order intercept point of +12 to +13 dBm, depending on model. Mixers meet MIL-STD 883 specifications to Class B. **Western Microwave, Inc., Applications Engineering Group, Sunnyvale, CA. (408) 734-1631. Circle 187.**

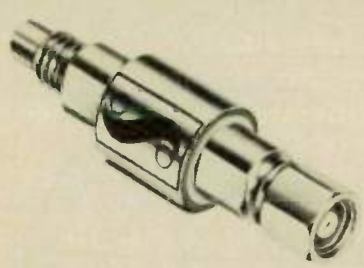
LOW CAPACITANCE BEAM LEAD PIN DIODE

Model HPND-4005 is a beam lead PIN diode designed for use in stripline and microstrip circuits. Series resistance is 4.7 ohms, capacitance is 0.017 pF, breakdown voltage is 120 V and reverse recovery time is 20 ns. Applications include switching, attenuating, phase shifting, limiting and modulating at microwave frequencies to 18 GHz. Diode's rugged construction includes a polymide surface layer for scratch protection, and leads with 6 gram typical pull strengths. Price: $13.60, 50-99 qty. Avail: from stock. **Hewlett-Packard Company, Palo Alto, CA. (415) 857-1501.** **Circle 167.**

SERIES OF LOW FREQUENCY FIXED ATTENUATORS

Series 3401 are low frequency fixed attenuators available in attenuation values from 1-30 dB with 3, 6, 10, 20 and 30 dB models. Frequency range spans the dc to 2.0 GHz range, SWR is 1.3 maximum, RF power rating is 1 watt average and connectors of the SMA male/female type are provided. Price: $25, small qty. Avail: from stock. **ARRA, Inc., Bay Shore, NY. Mike Geraci, (516) 231-8400.** **Circle 170.**

50 OHM COAXIAL ATTENUATORS



Broadband 50 ohm coaxial attenuators are offered for SMC systems. Standard unit provides attenuation values of 3 dB ± 0.3 dB; 6 dB ± 0.3 dB; 10 dB ± 0.5 dB; 20 dB ± 1.0 dB. SWR is 1.15 maximum, dc to 4 GHz range; 1.20 maximum, 4-8 GHz range; and 1.25 maximum, 8-12 GHz range. Power rating is 2.0 W average, 500 W peak. **RF Components Division, Sealectro Corporation, Mamaroneck, NY. (914) 698-5600.** **Circle 182.**

SERIES OF TUBULAR FILTERS

SF 103 Series tubular filters provide low insertion loss and high power capability in same length and volume. Filter Series SF 103 covers 200 MHz to 5.5 GHz (1" x 1" cross-section); Series SF 102 covers 1-10 GHz frequency spectrum (.8" x .6" cross-section) and the SF 101 Series spans 9-18 GHz frequency band (.56" x .38" cross-section). Bandwidths range from 1-70%. Available with pins or connectors and built-in mounting provisions. Price: $125. Avail: 2 wks. **RS Microwave Company, Inc., Butler, NJ. (201) 492-1207.** **Circle 179.**

SUBMINIATURE PROGRAMMABLE ATTENUATOR

PA-5010 is a subminiature programmable attenuator which provides 0-127 dB of attenuation in 1 dB steps. The frequency range of the unit is dc to 1300 MHz and per cell accuracy is 0.2 dB or 1% at 1000 MHz. Component's SWR is 1.5, maximum at 1000 MHz. Size: 3 cu. in. Available with control voltages of 26.5 Vdc, 12.0 Vdc or 5.0 Vdc and SMA type connectors. Price: $390, 1-9 qty. Del: 8 wks ARO. **Texscan Corporation, Indianapolis, IN. Raleigh B. Stelle, (317) 357-8781.** **Circle 184.**

IMAGE REJECT (IMAGELESS) MIXER

An image reject (imageless) mixer converts RF frequency of 3.0 - 3.5 GHz to an IF frequency of 25 - 35 MHz. Unit offers conversion loss of 7.5 dB maximum, SWR of 1.6 maximum (RF and LO) and image rejection of 25.0 dB minimum. The LO to signal isolation is 15 dB minimum and LO and signal to IF isolation is 40 dB minimum. Available with SMA female connectors and 3½" x 3" x ¾" dimensions. **Triangle Microwave, Inc., East Hanover, NJ. (201) 884-1423.** **Circle 185.**

(continued on page 104)

(from page 103) PRODUCTS

Materials

Al CLADDING ON TEFLON PC SUBSTRATES

Aluminum cladding for Teflon®-based microwave printed circuit substrates is introduced to enhance dimensional stability and to serve as a heat sink to protect heat-sensitive components on the PC board. The aluminum is clad to such low loss substrates as Di-Clad grades 522 and 527, Teflon® -fiberglass fabric and Di-Clad grades 870 and 880, Teflon® -woven fiberglass and grade 810, ceramic-loaded Teflon®. A special manufacturing process permits .25" of Al to be bonded to the Teflon®. **Keene Corporation, Chase-Foster Division, Bear, DE. Frank Yoerg, (302) 834-2100.** **Circle 169.**

Antennas

WIDEBAND, CIRCULARLY POLARIZED ANTENNA

WJ-8338 is a wideband circularly polarized omnidirectional antenna with a frequency range of 1.0-12.0 GHz. The antenna pattern's coverage is 360° in azimuth at elevation angles of 30° to 90° from the antenna axis. Right or left-hand circular polarization may be specified and unit can be used in conjunction with a broadband direction-finding system as an acquisition antenna. Antenna functions, with reduced performance, from 0.75 - 1.0 GHz and from 12.0 - 18.0 GHz. **Watkins-Johnson Company, Antenna Applications Engineering, San Jose, CA. (408) 262-1411.** **Circle 186.**

Devices

SILICON TUNING VARACTORS

Series MA-45200 are silicon abrupt junction microwave tuning varactors with minimum Q at −4 V from 800 to 5500 and total capacitance C_T at −4 V ranges from .50 to 8.20 pF and minimum capacitance ratios (0-25 V) from 2.7 to 7.4, depending upon model. Each device in the series has a high density silicon dioxide passivation, which results in low leakage currents and low post tuning drift. Contacts are sputtered for consistency and reliability and diodes meet MIL-S-19500, MIL-STD-202, and MIL-STD-750 specifications. **Microwave Associates, Inc., a M/A-COM Company, Burlington, MA. (617) 272-3000.** **Circle 176.**

SILICON UHF MIXER AND DETECTOR DIODES

Series IN82 point contact barrier mixer diodes and Series IN830 silicon UHF detector diodes are designed for FM, UHF TV and airborne communication circuit applications. UHF mixer diode series has T = 25°C, V_R (reverse voltage) = 5 V minimum and C_{JO}, junction capacitance of .3 pf typical and an overall noise figure at f = 700 MHz of (Model IN82A) 14 dB maximum and (Model IN82AG) 12 dB maximum. UHF detector diode series has CW burnout of 375 mW at T = 25°C or 15 W peak for 1 μs. Model IN830 has detector rectification efficiency of 65% minimum at f = 100 MHz. Price: IN82 series, 35¢, 10,000 qty; IN830 series, 38¢, 10,000 qty. **Microwave Diode Corp., W. Stewartstown, NH. D. E. Shea, (603) 246-3363.** **Circle 178.**

Instrumentation

IEEE-488 GPIB OPTION FOR SWEEP GENERATOR

Option IEEE-488 GPIB, for a broadband, continuous sweep generator, No. 6600B/9514D/9616D, provides automatic operation under the control of an external controller calculator. Continuous sweep from 10 MHz to 18 GHz is executed as a smooth analog function instead of digital steps. This instrument features band selection of full .01-18 GHz or four dual band frequency ranges. Selection of 10,000 steps per band for any frequency function; F1-F2, FϕΔF, FϕCW, Fϕ external FM, and 10,000 step call up of three markers. All other sweeper controls are programmable – including RF on-off, four leveling modes, modulation modes and special sweep-pause mode. **AIL-TECH Electronic Instruments, Eaton Corporation, Los Angeles, CA. Roy E. Wendell, (516) 588-3600.** **Circle 164.**

PROGRAMMABLE RECEIVER

Series 1780 programmable microwave receiver operates at fixed frequencies to over 100 GHz and is programmable from 1-40 GHz. An optional low frequency converter extends coverage down to 100 MHz. The receiver is IEEE-488 Bus compatible. Receiver automatically calibrates the IF system with a stable internal crystal oscillator. **Scientific-Atlanta, Inc., Atlanta, GA. Bruce K. Hudson, (404) 441-4000.** **Circle 181.**

Advertising Index

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New Literature

SIGNAL GENERATOR BROCHURE

A 6-page brochure describes the Model SMS Signal Generator, including all specification data for the instrument. In addition to the basic feature and application description, the product literature explains all options and the unit functions are highlighted on a one-page callout illustration. **Polarad Electronics, Inc., Lake Success, NY. (516) 328-1100.** **Circle 188.**

TRANSISTOR DESIGNERS CATALOG

A catalog for transistor designers includes complete data sheets for silicon bipolar and GaAs FET transistors along with a selection guide and information on the company's high reliability screening program. The 160-page handbook for 1981 provides the user of VHF, UHF and microwave transistors with comprehensive information on the complete transistor line. In addition to the data sheets, the catalog features abbreviated specifications and typical performance curves and applications of each product type. **Avantek, Inc., Santa Clara, CA. Charles Cochran, (408) 496-6710.** **Circle 195.**

PRODUCT LINE CATALOG

A product line catalog features 20 pages of information and data on cesium beam frequency and time standards, digital clock, quartz crystal oscillators, quartz frequency standards and satellite timing receiver. This two-color booklet introduces the company, its capabilities and methods. All product types are illustrated with performance curves and specifications as well as applications and features. **Frequency and Time Systems, Inc., Beverly, MA. (617) 927-8220.** **Circle 190.**

FREQUENCY SYNTHESIZER CAPABILITY BOOKLET

Microwave frequency synthesizers from UHF to X bands are described in a capability booklet which includes color photographs, functional diagrams and performance curves. The booklet outlines the design considerations for high-stability tunable microwave sources in both radar and communications applications. It treats a basic single-loop indirect synthesizer arrangement, the relationships between FM noise sideband levels, loop bandwidth and switching speeds. These are followed by descriptions of methods used to achieve high degrees of frequency agility in radar drive sources, including frequency-swept transmitter pulse chirp generation in pulse compression radar systems. **Microwave Associates Ltd, A M/A-COM Company, Woodside Estate, Bedfordshire, England. Ian Williamson, (0582) 605012/3/4/5, TLX 82295.** **Circle 192.**

CATALOG ON OSCILLATOR PLUG-INS

A recently released catalog features specifications and prices for 18 oscillator plug-in units that cover five microwave frequency ranges and power levels to 250 mW. Booklet includes a general description of the modular construction and interchangeability of the YIG oscillators used. Specifications and product illustrations are provided. **Electronics Surveillance Components, Inc., Palo Alto, CA. Joe Balaty, (415) 494-7803.** **Circle 196.**

SCALAR NETWORK ANALYZER APPLICATION NOTE

The automation of a scalar network analyzer is described in Application Note 155-3. This 15-page, two-color note discusses the analyzer system components, the measurement sequence and how each system is connected. Operation is described with details of each operator task. Also provides Appendix I, containing a complete program listing, Appendix II which covers calibration, Appendix III which discusses the use of sweep oscillator and Appendix IV which lists variables. **Hewlett-Packard Co., Palo Alto, CA. (415) 857-1501.** **Circle 189.**

DATA SHEET ON HIGH PERFORMANCE RECEIVER TRANSISTOR

A two-color data sheet describes a high performance receiver transistor designed for applications up to 2 GHz. This one-page sheet, No. 590, summarizes Model LT-4700's electrical characteristics with schematic drawings and features a noise/frequency graph. **TRW RF Semiconductors, Div. of TRW, Inc., Lawndale, CA. (213) 679-4561.** **Circle 191.**

DATA SHEET ON VARIABLE ATTENUATOR

Data sheet describes the Model 910 Continuously Variable Attenuator. Full specifications and features of the unit are detailed as well as the model's typical incremental shift curve and typical incremental insertion loss curves. **Weinschel Engineering, Gaithersburg, MD. (301) 948-3434.** **Circle 194.**

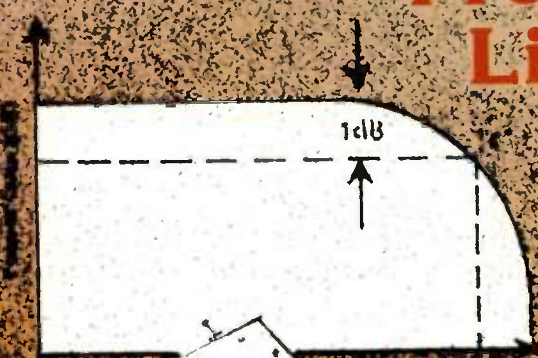
TEST EQUIPMENT LINE CATALOG

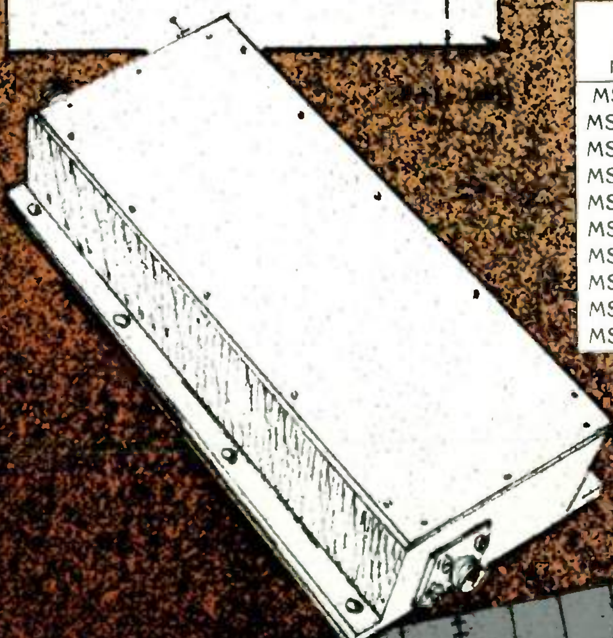
A complete line of microwave test equipment is featured in this 1981 catalog. Over 1400 waveguide and coaxial components are listed with electrical specifications. Literature also describes company's capability to design and manufacture to stringent specifications, plus special component types for modular system applications. **Waveline, Inc., West Caldwell, NJ. R. H. Koenig, (201) 226-9100.** **Circle 193.**

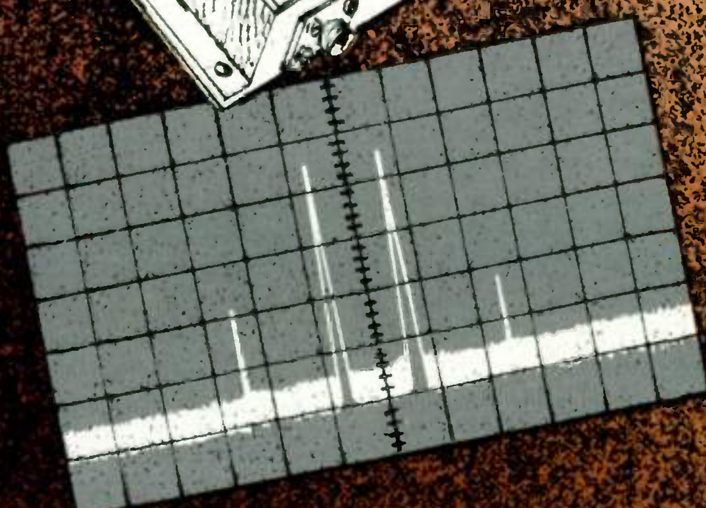
MSC

Linear Power Transistors

Products for Solid State Linear Power Amplifier Applications







CLASS A LINEAR AMPLIFIERS
ELECTRICAL CHARACTERISTICS (@ 25 °C)

Model Number	Test Freq. (MHz)	POUT (1) Min. (W)	PIN (W)	Bias VCE (V)	IC (mA)	θJC Max. (°C/W)	VCEO Min. (V)	IC (mA)
MSC 2100	1000	0.316	0.028	18	50	30.0	20	5
MSC 82100	1000	0.316	0.028	18	50	20.0	20	5
MSC 80064	2000	0.112	0.014	18	50	45.0	20	5
MSC 84100	2000	0.250	0.025	20	60	45.0	21	5
MSC 84101	2000	0.500	0.080	20	120	25.0	21	5
MSC 80195	2000	0.630	0.110	18	140	35.0	20	5
MSC 80196	2000	1.000	0.200	18	220	17.0	20	5
MSC 80197	2000	1.500	0.370	18	360	8.5	20	5
MSC 80725	2000	2.500	0.630	18	450	8.5	20	5
MSC 80264	4000	0.100	0.025	12	60	45.0	15	5

NOTE (1) Gain Compression is ≤ 1.0dB at Pout

YOUR TOTAL MICROWAVE RESOURCE

The MSC Linear Power Transistor Series provides low intermodulation distortion performance for amplifier applications requiring capability up to 4 GHz with power levels ranging from +20 dBm to +34 dBm. Please call or write for the complete MSC Linear Product Data Packet.

MSC MICROWAVE SEMICONDUCTOR CORP.
A Siemens Company
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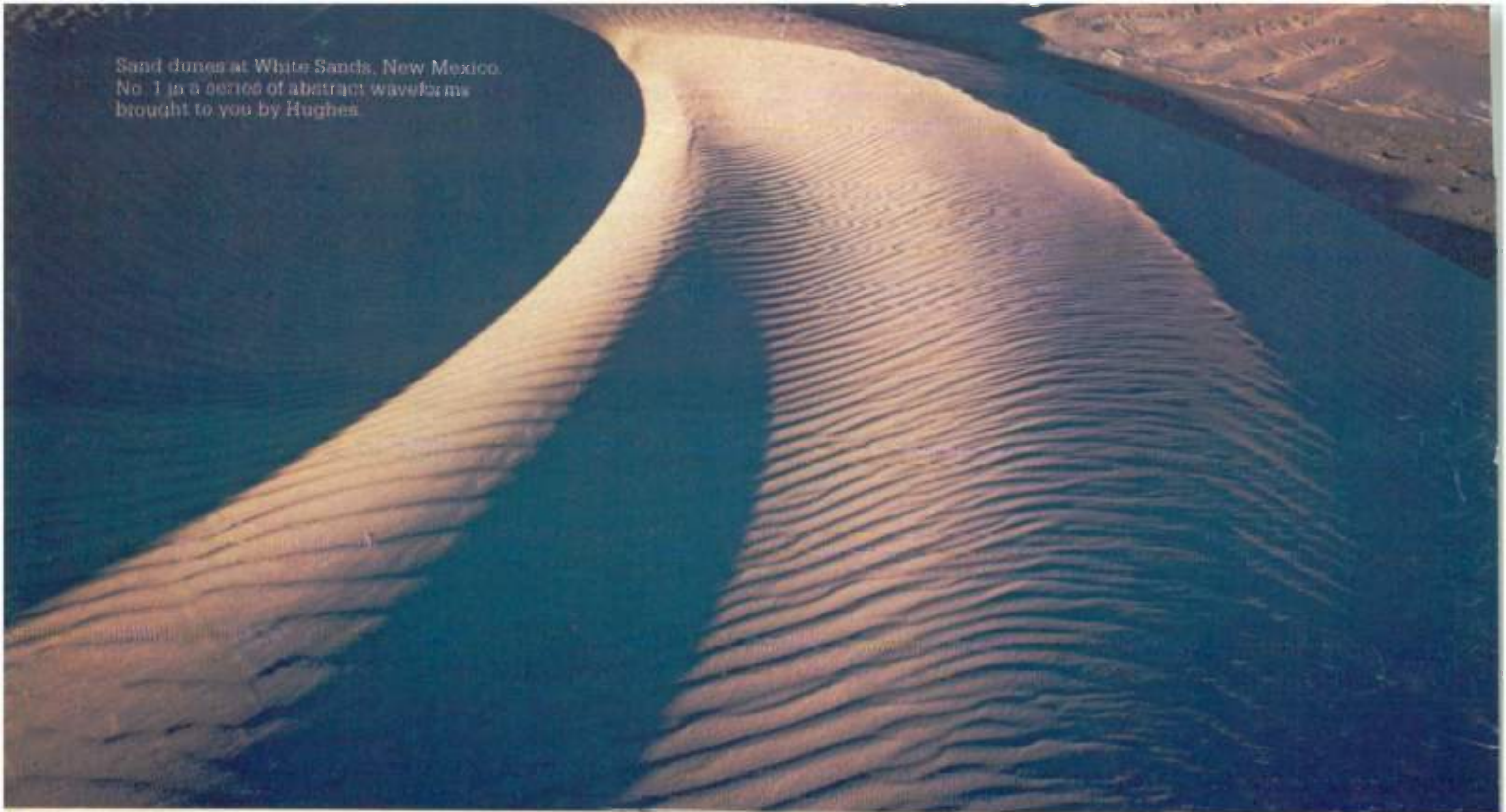
Somerset, New Jersey 08873 U.S.A.
(201) 469-3311 TWX (710) 480-4730 TELEX 833473

CIRCLE 2 ON READER SERVICE CARD



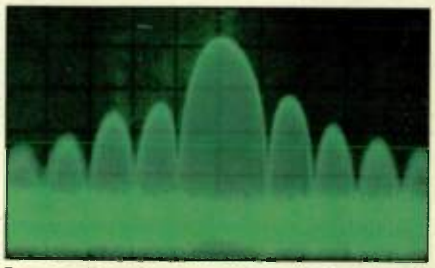
Hughes is making new waves with a shortcut to mm-wave radar design.

Now designers don't have to reinvent the radio when they take on millimeter-wave radar development. That's because Hughes short cuts the millimeter-wave learning curve by offering preassembled and tested subsystems right off the shelf.

For example, our new 42266H subsystem is a complete, all-purpose, coherent instrumentation radar front end. It's made specifically for use in gathering a broad base of fundamental data immediately. . . like you'd need in developing weapon sensors. Or you can use it to incorporate the technology into your own system. We'll even help customize it if you'll tell us your needs.

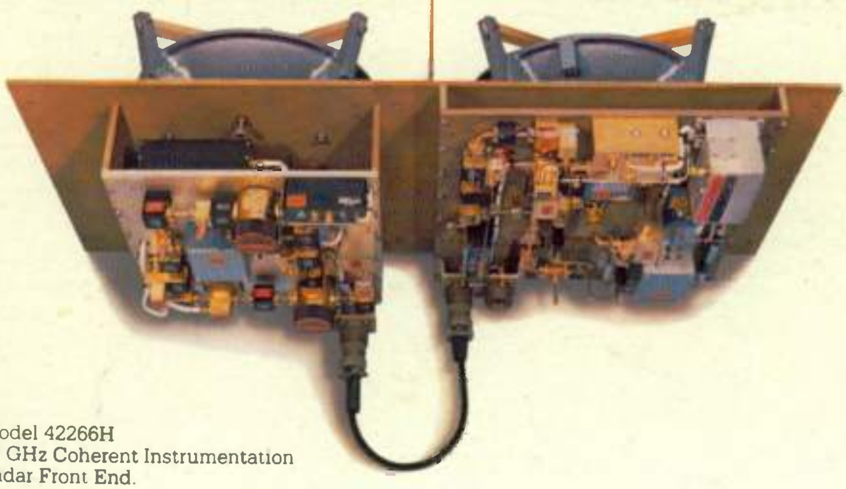
Operating at 94 GHz this fully calibrated breadboard combines transmitter, receiver and antenna functions all in one subsystem. It performs all mixing, detecting, switching and modulating functions needed to provide an S-band IF output signal.

There's lots more to tell, so write or call Hughes Electron Dynamics Division, Solid State Marketing, 3100 West Lomita Blvd., Torrance, CA 90509; (213) 517-6400.



Outstanding signal-to-noise ratio is demonstrated in the spectrum of pulsed IF output signal (Sin x/x). Total peak-to-valley ratio is in excess of 50 dBc/KHz.

The 42266H can be used as a multifunction instrumentation radar capable of either CW or pulsed coherent operation. In the CW mode it can perform as a single or multi-frequency CW coherent front-end. In pulsed mode, the system performs as a single-frequency coherent pulsed radar front-end or as a coherent system with frequency diversity in the form of pulse-to-pulse frequency agility, coherent chirp, or combinations of both. It can also be used to implement pulse compression techniques. The receiver consists of an antenna with a single-plane monopulse comparator and two balanced mixer/IF preamplifiers.



Model 42266H
94 GHz Coherent Instrumentation Radar Front End.

HUGHES

HUGHES AIRCRAFT COMPANY
ELECTRON DYNAMICS DIVISION

Making waves in millimeter-wave technology.

CIRCLE 3 ON READER SERVICE CARD