microwave JOURNAL

INTERNATIONAL EDITION □ VOL. 24, NO. 10 □ OCTOBER 1981

Hughes 12 GHz Monolithic Amplifier

RADAR
- Array Radar Review
- 95 GHz Instrumentation Radar
- Short Backfire Doppler Antenna
- Lookdown Capability Charts

horizon house

OUR SPECTRUM ANALYZERS NEVER FORGET A SIGNAL.

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Selection Guide

Model	Frequency
632B-1	100 kHz-2 GHz
630B-1	3 MHz-40 GHz
640B-1	3 MHz-40 GHz

Circle 1 on Reader Service Card

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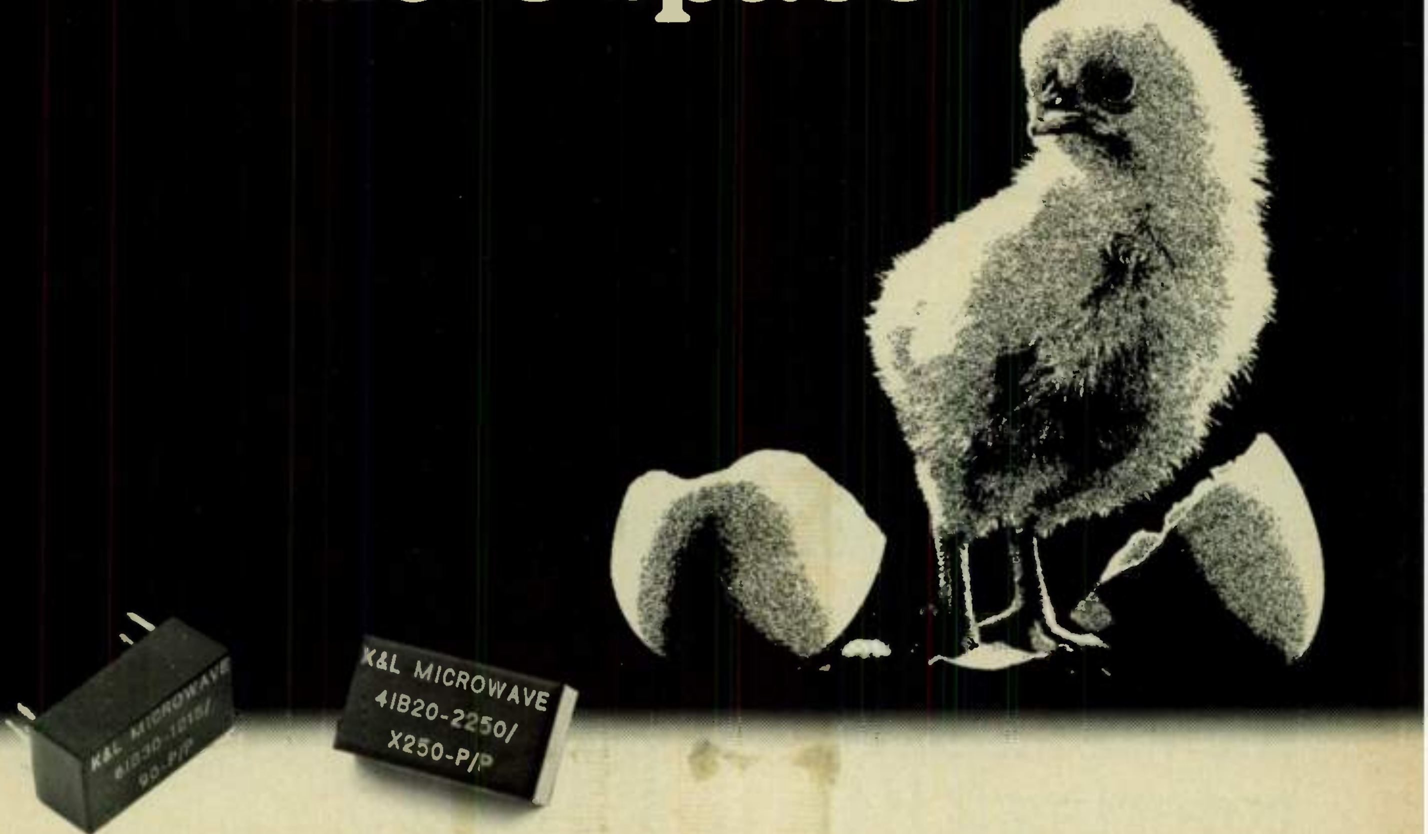
Polarad Electronics, Inc. 5 Delaware Dr. Lake Success, N.Y. 11042 Tel. 516-328-1100 TWX 510-223-0414

Are your customers pecking away for more space?



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CIRCLE 4 ON READER SERVICE CARD

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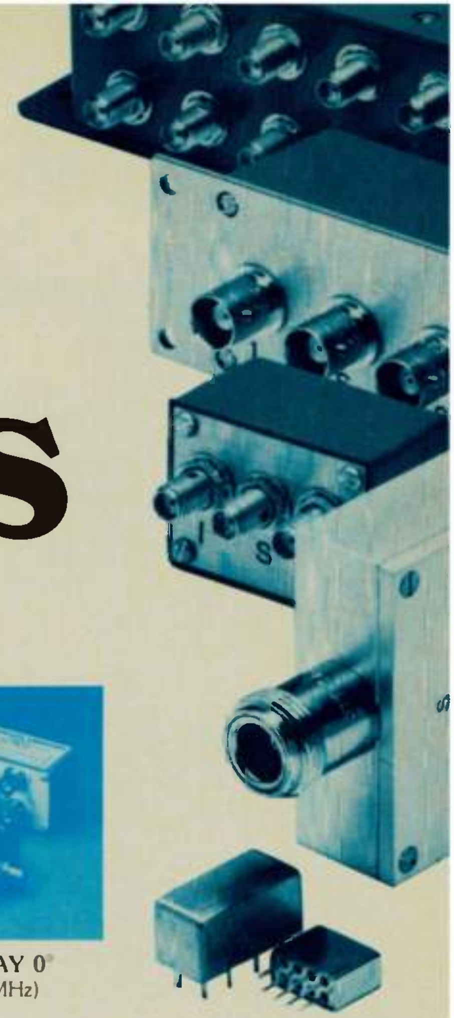
power splitters



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

the world's largest selection... covering 2 KHz to 4.2 GHz

from Mini-Circuits, from $9.95

Over 105 standard models 2-way to 24-way, 0°, 90°, 180°, pin or connector models... Mini-Circuits offers a wide variety of Power Splitters/Combiners to choose from, with immediate delivery. But there are always "special" needs for "special applications"... higher isolation, SMA and Type N connectors Intermixed, male connectors or wide bandwidths. Contact us. We can supply them at your request... with rapid turnaround time. Naturally, our one year guarantee applies to these units.

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
6. BNC and TNC connectors (SMA and Type "N" at $5 additional cost) (BNC not available on ZAPD-4)

Please specify connectors.

Mini-Circuits
A Division of Scientific Components Corp
World's largest manufacturer of Double Balanced Mixers
2625 East 14th Street, Brooklyn, New York 11235 (212)769-0200
Domestic and International Telex 125460 International Telex 620156

** Model PSC 3-1 Manufactured under license.
Protected by Patent 3,428,920

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CIRCLE 5 ON READER SERVICE CARD

and combiners

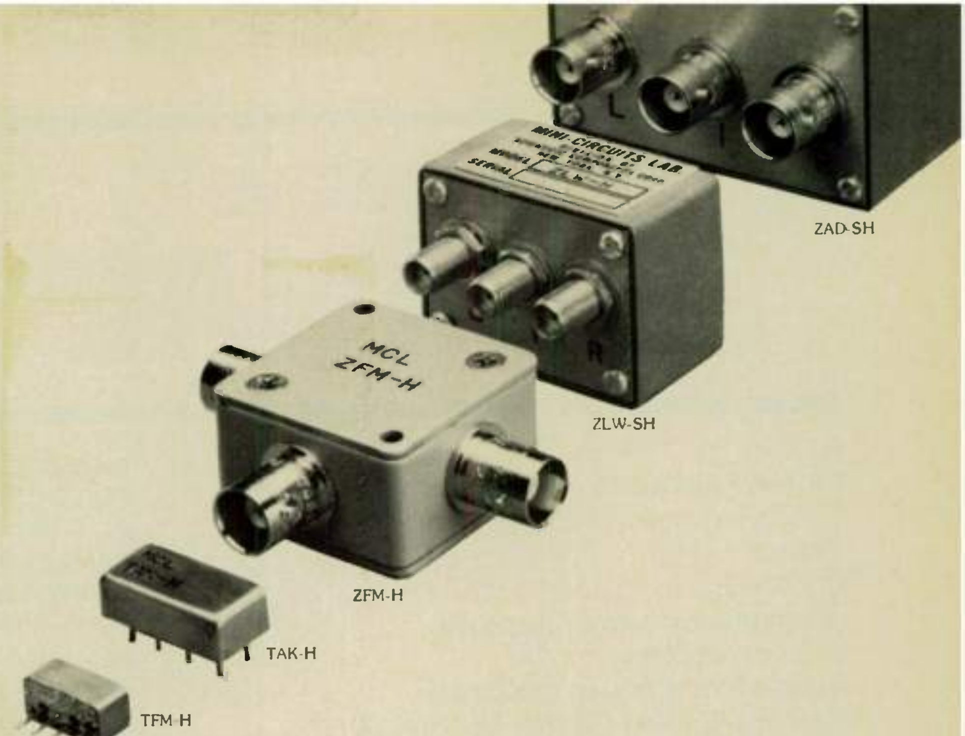
FOUR WAY 0°	SIX WAY 0°	EIGHT WAY 0°	SIXTEEN WAY 0°	TWENTY FOUR WAY 0°
(2 KHz-4200 MHz)	(1-175 MHz)	(0.01-750 MHz)	(0.5-125 MHz)	(0.2-100 MHz)

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	$39.95 (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	$36.95 (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)

"For Mini Circuits sales and distributors listing see page 35."

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.



the dynamic rangers

world's first guaranteed low-distortion mixers... -55 dB 2-tone 3rd order IM (0 dBm RF)

at unbeatable prices from Mini-Circuits from $19.95

Now...improve your systems intermod spec by as much as 10 dB with Mini-Circuits' state-of-the-art ultra-low distortion double-balanced mixers. These units are guaranteed −55 dB 2-tone 3rd order IM specs (below IF output); test conditions RF_1 = 200 MHz, RF_2 = 202 MHz at 0 dBm, LO = 180 MHz at +17 dBm. The models span 50 KHz to 1000 MHz, with only 6 dB insertion loss, isolation greater than 45 dB and 1 dB compression point typically at +15 dBm. Available for immediate delivery in two printed-circuit versions and three connector versions. Of course, each unit carries Mini-Circuits' one-year guarantee.

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

finding new ways...
setting higher standards

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A Division of Scientific Components Corporation
World's largest manufacturer of Double Balanced Mixers
2625 East 14th Street, Brooklyn, New York 11235 (212)769-0200
Domestic and International Telex 125460 International Telex 620156

SPECIFICATIONS LO + 17 dBm

Model No.	Freq (MHz)	Conv loss (dB max.)	Signal 1 dB compr level (dBm min.)	Connections	Size (in.) (w x l x ht.)	Price (Qty)
TFM 1H	2 - 500	8.5	+14	4 pins	0.21x0.5x0.25	$23.95 (5-24)
TFM 2H	5 - 1000	10	+14	4 pins	0.21x0.5x0.25	$31.95 (5-24)
TFM 3H	0.1 - 250	8.5	+13	4 pins	0.21x0.5x0.25	$23.95 (5-24)
TAK 1H	2 - 500	8.5	+14	8 pins	0.4x0.8x0.25	$19.95 (5-24)
TAK 1WH	5 - 750	9.0	+14	8 pins	0.4x0.8x0.25	$23.95 (5-24)
TAK 3H	0.05 - 300	8.5	+13	8 pins	0.4x0.8x0.25	$21.95 (5-24)
ZAD 1SH	2 - 500	8.5	+14	BNC, TNC	1.15x2.25x1.40	$40.95 (4-24)
ZAD 1WSH	5 - 750	9.0	+14	BNC, TNC	1.15x2.25x1.40	$44.95 (4-24)
ZAD 3SH	0.05 - 300	8.5	+13	BNC, TNC	1.15x2.25x1.40	$42.95 (4-24)
ZLW1SH	2 - 500	8.5	+14	SMA	0.88x1.50x1.15	$50.95 (4-24)
ZLW 1WSH	5 - 750	9.0	+14	SMA	0.88x1.50x1.15	$54.95 (4-24)
ZLW 3SH	0.05 - 300	8.5	+13	SMA	0.88x1.50x1.15	$52.95 (4-24)
ZFM 1H	2 - 500	8.5	+14	BNC, TNC SMA, N	1.25x1.25x0.75	$53.95 (1-24)
ZFM 2H	5 - 1000	10	+14	BNC, TNC SMA, N	1.25x1.25x0.75	$61.95 (1-24)
ZFM 3H	0.05 - 300	8.5	+13	BNC, TNC SMA, N	1.25x1.25x0.75	$54.95 (1-24)

Impedance 50 ohms, Isolation 30dB min.,
BNC standard, TNC on request, Type N and SMA $5.00 additional

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"For Mini Circuits sales and distributors listing see page 35."

microwave JOURNAL

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ON THE COVER: GaAs wafers containing several hundred 12 GHz single-stage, low noise amplifiers suitable for direct broadcast receivers and radar X-band applications. Artwork courtesy of Hughes Aircraft Co., Torrance Research Center.

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BPA

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pick a mixer

standard level (+7dBm LO)

from 500 KHz to 1GHz... hi-rel and industrial miniature, flatpack, and low profile from $3.95

Choose from the most demanded mixers in the world. Rugged construction and tough inspection standards insure compliance to MIL-M-28837/1A.

Check these features...

SRA-1 the world standard, covers 500 KHz to 500 MHz, Hi-REL, 3 year guarantee, HTRB tested, meets MIL-M-28837/1A-03S. $11.95 (1-49).

TFM-2 world's tiniest Hi-REL units, 1 to 1000 MHz, only 4 pins for plug-in or flatpack mounting, meets MIL-M-28837/1A, $11.95 (6-49).

SBL-1 world's lowest cost industrial mixers, only $3.95 (10-49), 1 to 500 MHz, all metal enclosure.

SBL-1X industrial grade, low cost, $4.95 (10-49) 10 to 1000 MHz, rugged all metal enclosure.

ASK-1 world's smallest double-balanced mixers, 1-600 MHz, flat-pack mounting, plastic case, $5.95 (10-49).

MODEL	SRA-1	TFM-2	SBL-1	SBL-1X	ASK-1
FREQUENCY, MHz					
LO, RF	5-500	1-1000	1-500	10-1000	1-600
IF	DC-500	DC-1000	DC-500	.5-500	DC-600
CONVERSION LOSS, dB					
one octave bandedge	6.5	6.0	7.5	7.5	7.0
total range	8.5	7.0	8.5	9.0	8.5
ISOLATION, dB, L TO R					
lower bandedge	50	50	45	45	50
mid range	40	40	35	30	35
upper bandedge	30	30	25	20	20

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

"For Mini Circuits sales and distributors listing see page 35."

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CIRCLE 9 ON READER SERVICE CARD 74 REV. ORIG

OSM/OSCS (SMA) Miniature Coaxial Connectors

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- Ceramic-sealed hermetic connectors for superior performance at high frequencies
- Flange mount, panel and bulkhead feedthroughs, microstrip launch and cable types

the best for less

OSSM Subminiature Coaxial Connectors

- Standard to 38 GHz, others to 40 GHz
- Stripline and Microstrip transmission line types
- Compatible with flexible and semi-rigid cable for direct solder or crimp attachment

OMQ/OSQ Miniature Quick-Connect/ Disconnect Coaxial Connectors

- Up to 40 connectors on a 4" square panel
- Positive interlock, impossible to disengage by pulling on cable
- Semi-rigid and flexible cable types
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- In-Series and Between-Series adapters

OSN (Type N) / OST (TNC) Coaxial Connectors

- Highest quality Type N and TNC connectors for high performance systems
- Standard through 18 GHz

Omni Spectra

A M/A-COM COMPANY

Precision 7mm Coaxial Connectors/Components
- DC to 18 GHz
- Low VSWR (to 1:06:1)
- Meets MIL-C-39012 requirements
- Adapters from 7mm to OSM, OSSM, OSN or OST (plug and jack configurations)

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- Meet MIL-E-5400 and MIL-E-16400 environmental requirements
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We've expanded our production capacity 50% and added new people and equipment . . . increased inventories at plants and distributors for immediate delivery of most standard connectors and components . . . doubled our field sales/service network by combining forces with M/A-COM. And, we've implemented an aggressive new price schedule that makes our commitment to give you the best for less hard to resist. The best quality, performance and delivery at the very best price. Put our commitment to the test. Today. **(603) 424-4111**

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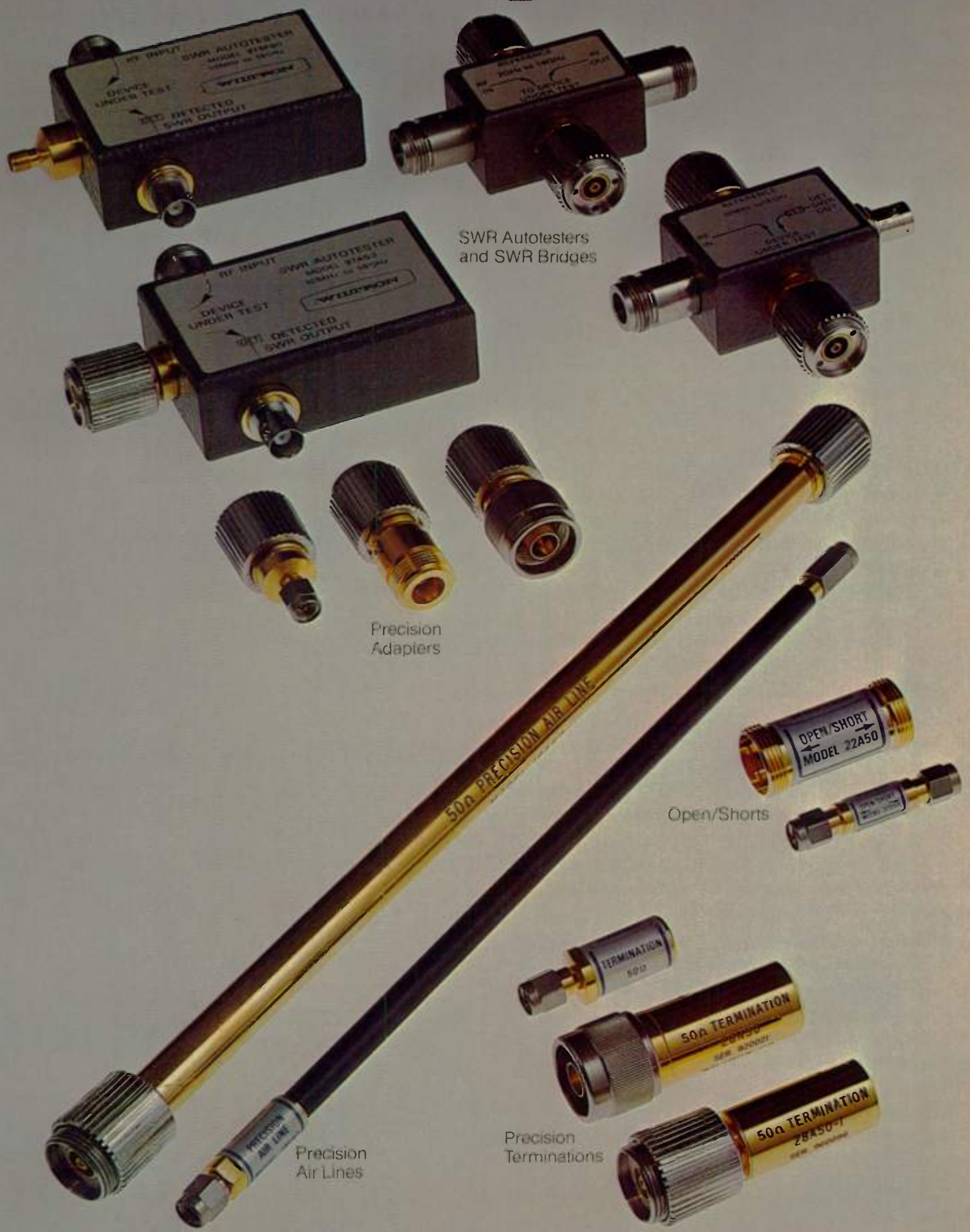

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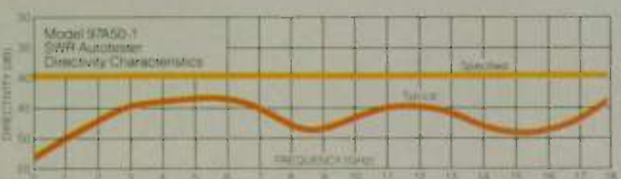

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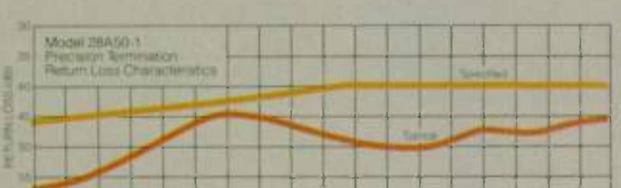

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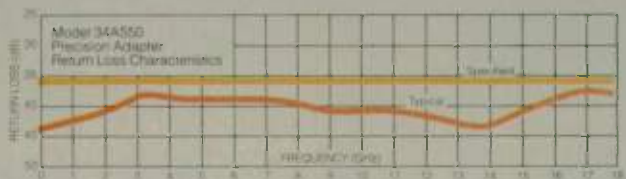

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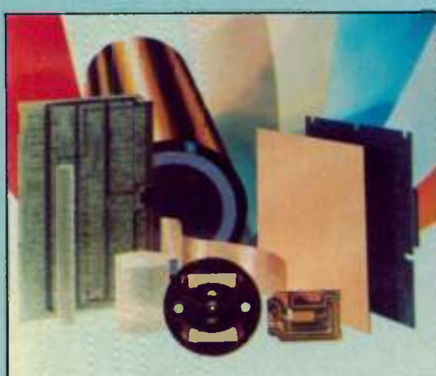

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Coming Events

ARFTG FALL 1981 CONFERENCE NOV. 5-6, 1981 Sponsors: Automatic RF Techniques Group. Place: Nassau Hotel, Princeton, NJ. Topics: Large ATE systems and recent hardware and software developments in computer-aided RF design and testing. Technical exchange will be accomplished by informal twenty minute talks using either 35mm slide or viewgraph illustrations. Manufacturers will demonstrate new equipment that has been specifically designed for use in computer-aided RF design and test. Contact: Edward J. Stevens, RCA Government Systems Division, Borton Landing Rd., Bldg. 101-124, Moorestown, NJ 08057. Tel: (609) 778-3905.

14th ANNUAL CONNECTOR SYMPOSIUM NOV. 11-12, 1981 Sponsor: Electronic Connector Study Group, Inc. Place: Franklin Plaza, Philadelphia, PA. Sessions: Connectors I & II, Materials I & II & III, Interconnection Techniques, Platings and Finishes I & II, Cabling Techniques and Connectors - Special Applications, plus workshops. Contact: Jim Pletcher, Ex. Dir. ECSG, Inc., P.O. Box 167, Ft. Washington PA 19034. Tel: (215) 279-7084.

EASCON '81 Nov. 16-19, 1981 Sponsors: IEEE-Washington 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. Also features exhibition as well as technical classified programs. Contact: Dr. Delbert D. Smith, Chrmn., EASCON '81, COMSAT General Corp., 950 L'Enfant Plaza S.W., Washington, DC 20024. Tel: (202) 863-6822.

1982 IEEE MTT-S INTERNATIONAL MICROWAVE SYMPOSIUM JUNE 15-17, 1982 **Call for papers** Sponsors: IEEE Microwave Theory and Techniques Society. Place: Hyatt Regency Hotel, Dallas, Texas. Topics: Original works in microwaves particularly computer-aided design and measurement techniques, radiometry and remote sensing, GaAs monolithic circuits, phased array and active array techniques, microwave field and network theory and other areas. Submit 5 copies of a 35 word abstract and a 500-1000 word summary (up to 6 illustrations) by Jan 8, 1982 to: Steven L. March, TPC 1982 MTT-S Symposium, COMPACT Engineering Div., CGIS, P.O. Box 401144, Garland, TX 75040.■

Guest Editorial

Radar Passages — The 40th Year

THOMAS F. CURRY
Associate Deputy Assistant Secretary of the Navy

Dr. Thomas F. Curry, *Dr. Curry is currently Associate Deputy Assistant Secretary of the Navy. He was formerly Staff Specialist for Tactical Reconnaissance and Passive EW in the Office of the Under Secretary of Defense for Research and Engineering, and the Assistant Director, SIGINT, in the Office of the Assistant Secretary of Defense for Intelligence, ASD[I]. Prior to joining the government in January 1976, he was Vice President of Microwave Systems, Inc., and from 1965-1974 he held various management and engineering positions with E-Systems. He was a founder and first director of the Electronics Research Laboratory at the Syracuse University Research Corporation and earlier was with the Bell Telephone Laboratories at Murray Hill, New Jersey.*

Dr. Curry has worked in the SIGINT, radar, reconnaissance, and Electronic Warfare field since 1952. He served five years in the Signal Corps during the World War II and Korean conflicts and holds Bachelors, Masters, and Ph.D degrees in electrical engineering, receiving his Doctorate from Carnegie Tech. He is an IEEE Fellow, Registered Professional Engineer, Member of Tau Beta Pi, Eta Kappa Nu, and Sigma Xi, and is a member of the National Board of Directors of the Association of Old Crows.

Just over 40 years ago radar experienced its full operational debut in the Battle of Britain. By today's standards, the Chain-Home System was crude hardware, but the British coaxed indications and warning out of it to alert and guide the limited number of Spitfires and Hurricanes. Almost 18 months later an SCR-270 recently installed on the north shore of Oahu detected the incoming Japanese aircraft on Sunday morning, 7 December 1941.

At this juncture, it is useful to briefly look back both to garner pride in progress made and to review lessons learned — but only a brief look — for the road ahead requires undivided attention.

In hindsight, the accomplishments during the period 1940 to 1945 are truly astounding. Design rapidly moved from VHF thru Ka band, inventing the necessary components as needed from a zero-based startup position and getting hardware into production and into the field in months. Skilling's delightful little book, "Fundamentals of Electric Waves", (Wiley, 1943) appeared everywhere to re-educate a rotating-machine trained generation of electrical engineers, and to introduce the new generation of mostly military training students to electromagnetic wave theory and microwaves. What had been a theoretical course in the physics department suddenly came center stage in the design and technique arena of the electrical engineering department.

We caught our breath in the 1950s, and Middleton, Peterson, Marcum and many others explained and extended radar detection theory from the classical formulation in Lawson and Uhlenbeck's **"Threshold Signals".**

The 1960s ushered in a new growth era, with frequency diversity (FD) coherent pulsed doppler, and phased array concepts being brought to fruition on all fronts and with a brand new target class, the earth satellite, driving requirements to the limit of available components. At this halfway mark (20 years) one might illustrate progress in the tracker category by placing an SCR-584 beside the MILLSTONE HILL radar. But most radars were still manually tended with single threshold detectors, and with a few exceptions, there were only few-of-a-kind of the advanced technique radars actually fabricated. In many cases, the acquisition of the radars rode an up-and-down airframe/platform procurement decision horse —for example the ASG-18 and APG-55 pulsed-doppler developments.

With one or two exceptions, the Vietnam war did not fuel a quantum jump in radar progress, the exceptions being related to special Nam-related problems, such as foliage penetration, personnel detection and shell-tracking capabilities. This assertion can be tested by careful examination of the Proceedings of the 1975 International Radar Conference, which contains no discernible evidence that the United States has just emerged from a relatively long Southeast Asian conflict. The initial operational use in Nam of a Synthetic Aperture Radar, the APQ-102A, was disappointing because of 50 foot resolution and slow optical processing.

The 1970s ushered in the full impact of digital processing on the radar art. In the tactical reconnaissance, surveillance and target acquisition (RSTA) arena, the data-linked radar return digitally-processed on the ground in near-real-time began to meet user requirements in both SAR and AMTI modes. The ALARM system progressed to SOTAS, and the Lincoln Lab MRS^3 effort progressed to PAVE MOVER. Beginnings were again being made in the millimeter-

[Continued on page 18]

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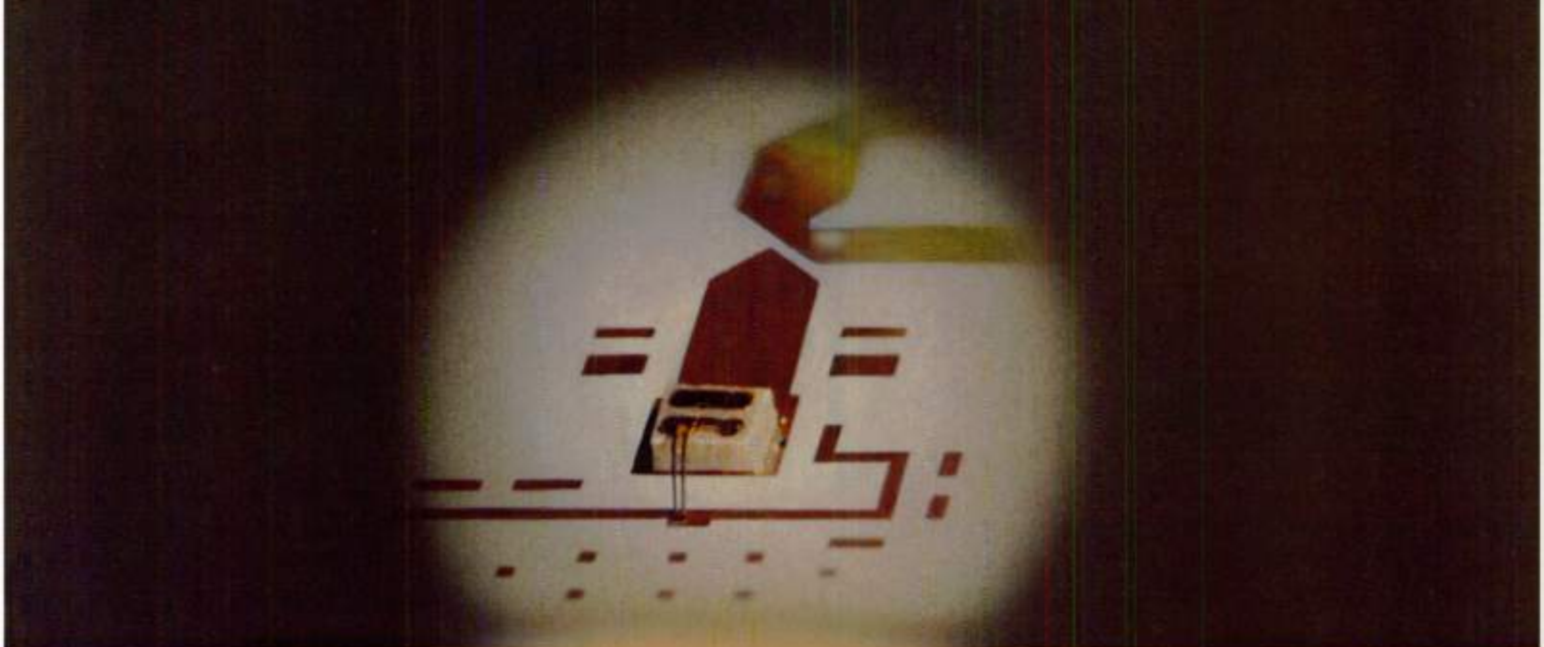

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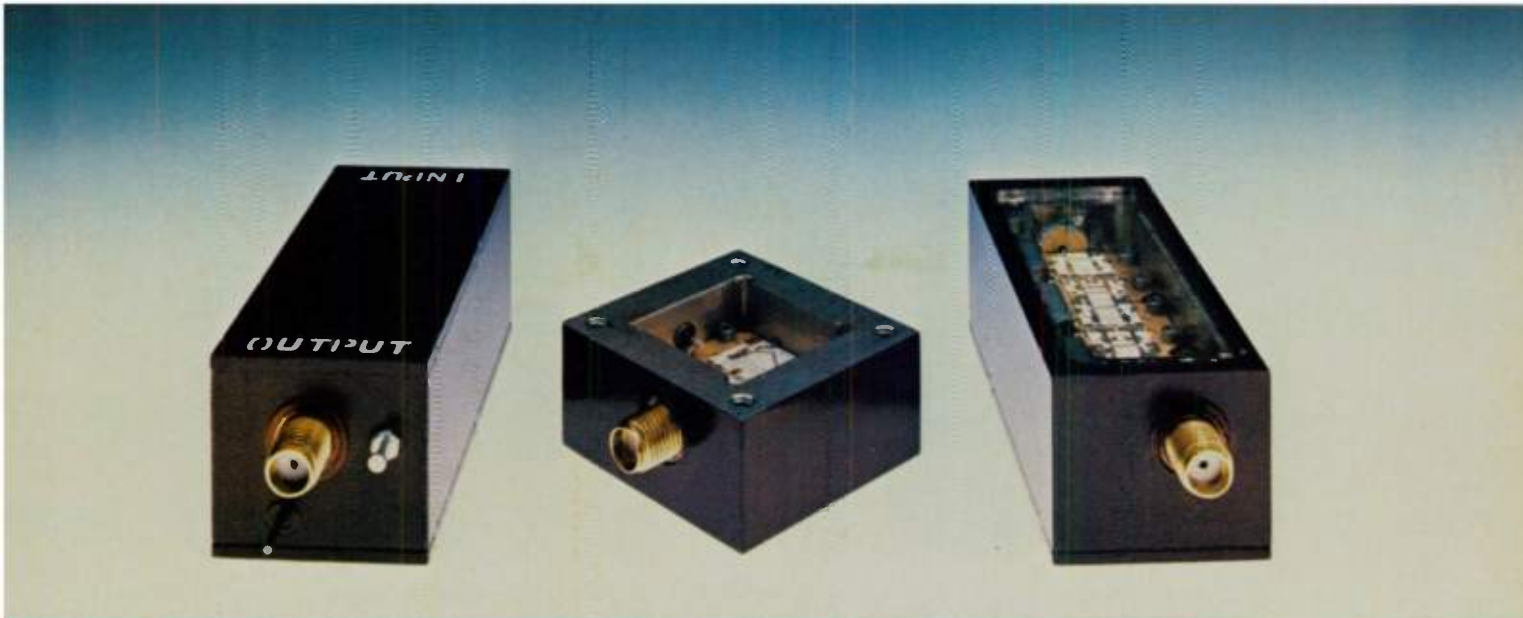
wave band with STARTLE and the Georgia Tech MM instrumentation efforts. But the large quantity production in the 1970s was associated with the culmination of the earlier coherent pulse doppler AI prototype developments in the AWG-9 and AWG-10 — truly lookdown — shootdown, multitarget, frequency jumping systems *in production*.

Now, imagine that Sir Robert Watson-Watt or Dr. Robert Page appeared at your office door today and asked, "whither goest our brainchild in 1981?". You could certainly shower acronyms on them, WAAS, SENRAD, FLEXAR, CFAR, SIAM, LPI, IADT, LORO, DCR, SLR, and SCV, to mention a few. But it would be better to describe where we are going in more fundamental ways:

- Brute Force Improvements — Power — Aperture products exceeding 10^7 and time bandwidth products exceeding 10^6.
- More coherent (cleaner) illumination exceeding $1/10^9$/millisecond and $1/10^{12}$/second allowing return fine-structure processing.
- Cleaner and more agile antenna directivity patterns with sidelobe levels approaching - 50 db and millisecond beam switching times.
- Adaptive detector/directivity strategies adjusting to the target/interference under microprocessor control.
- Radar as a wide area system element — netting and fusion of individual tracks with target identification.
- Near-stochastic spread spectrum waveforms processed by tailored SAW filters and microprocessor algorithms (for low probability of intercept and antijam qualities).
- False alarm management — complex moving threshold detectors working in coordination with the adaptive strategies.

These form a synopsis of the future directions in radar development. Expanding on each subject could lead to hours of interesting dialogue with your distinguished visitors. ■



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	SDA 9398-01	9.3 – 9.8	30	3.0	+13	+23
	SDA 117122-01	11.7 – 12.2	30	3.5	+13	+23
Broad band	SDA 2080-13	2 – 8	34	6.0	+18	+28
	SDA 80180-05	8 – 18	24	7.5	+10	+20
Medium power	SDA 2040-13	2 – 4	38	6.5	+21	+30
	SDA 4080-17	4 – 8	38	6.0	−21	+30
	SDA 80124-17	8 – 12.4	32	7.5	+21	+30

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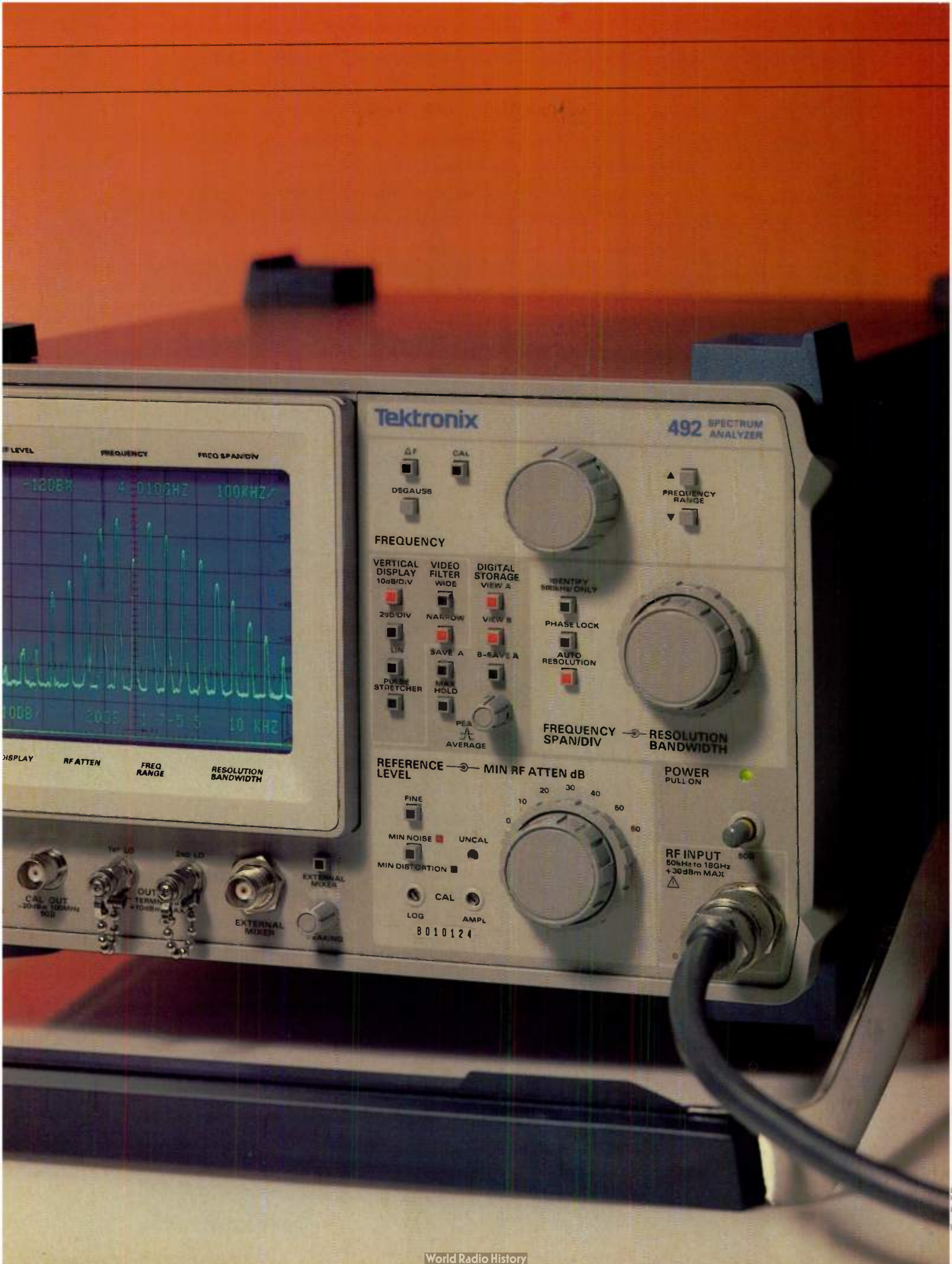
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Model 1295

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A Review of Array Radars

ELI BROOKNER
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INTRODUCTION

The achievements in array radars are exhibited by the operational deployment of the large high power, high range resolution (2.5 feet) COBRA DANE; the operational deployment of two all solid-state high power, large UHF PAVE PAWS radars; the development of the SAM multifunction Patriot radar which is nearing production; the development and production of the limited scan Precision Approach Radars (AN/TPN-25 and AN/GPN-22); the incorporation of polarization agility into an array radar; the development of hemispherical and omnidirectional coverage antennas; the development of arrays having low antenna side-lobes (-40 to -50 dB); the development of arrays requiring only N+1 controls instead of N^2 for phase-phase steering; the development of adaptive array processing techniques giving large (>50 dB) rejection of interfering signals; the development of X-band solid state power amplifiers; and the development of new low cost technology, such as monolithic power amplifiers. The large number of new array radars now under development, developed, in production, and in operation is also a measure of the advances made in the last decade in array radar technology.

The topics covered in this review include:

- Array radars steered in azimuth and elevation by phase shifting; (phase-phase steered arrays)
- Arrays steered ±60°, limited scan arrays, hemispherical coverage and omnidirectional coverage arrays.

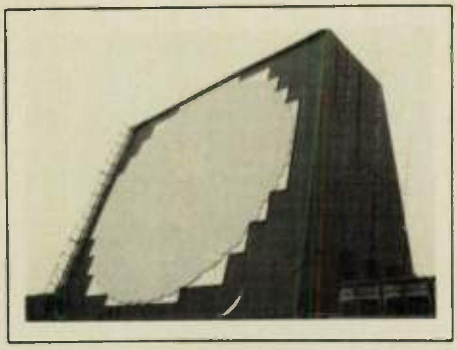

Fig. 1 COBRA DANE (AN/FPS-108).

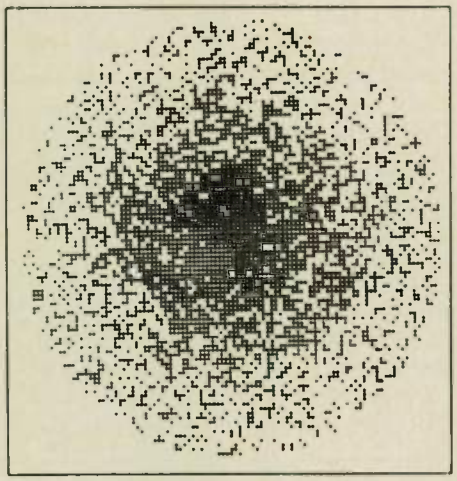

Fig. 2 Illustration of thinned circular array (not COBRA DANE).

- Array radars steering electronically in only one dimension, either by frequency or by phase steering.
- Array radar antennas which use no electronic scanning but instead use array antennas for achieving low antenna sidelobes.

PHASE-PHASE STEERED ARRAY SYSTEMS

COBRA DANE[1-5]

The COBRA DANE radar of Figure 1 represents an excellent example of the significant advances that have been made in phase-phase steered array radars in the past decade. This is a 95 ft diameter, 15.4 MW peak power and 0.92 MW average power L-band radar. It is intended for intelligence gathering on Soviet missile systems undergoing test firing, space track support and ICBM early warning.

A significant feature of this radar is its ability to provide high range resolution observations of targets at long ranges, on the order of 1,000 nmi. These wideband measurements are used for obtaining target size and shape data. A 200 MHz waveform which can provide a 2.5 ft range resolution is used to provide the high range resolution measurements. Normally it is not possible to obtain such high range resolution measurement from a 95 ft aperture antenna because of the time dispersion across the antenna. For example, when pointing to a target which is 22° off boresight in elevation, the signal from the lower part of the aperture arrives at the target 95' x sine 22° = 35.6 ft after the signal from the top of the array arrives at the target. As a result the 2.5 ft pulse is spread out over 35.6 ft, making it impossible to achieve the desired high range resolution.

To get around this problem the 95 ft diameter array is broken up into smaller arrays called subarrays. These subarrays are made small enough so that the time dispersion across them is small compared to the desired range resolution. Furthermore, the signal to each subarray is delayed by the amount necessary to make the signals from all subarrays arrive at the target simultaneously. Thus for the example given above the top subarray unit receives a time delay of about 35 ns relative to the bottom subarray unit. This process is called time delay steering. Over each subarray phase steering is used. The 95 ft aperture is divided into 96 subarrays with each subarray having 160 active radiating elements that are phase steered. There are a total of 15, 360 (160 x 96) active elements in the array face. Each subarray is fed by a Raytheon ring bar QKW 1723 160 KW peak power TWT.

Another interesting feature of the radar is its use of array thinning[49]. Figure 2 illustrates the concept of array thinning. The black dots in the figure indicate the positions over the array at which the active elements are placed. Where there are no black dots a dummy nonradiating element is used. Notice that most of the elements near the center of the aperture are active with the density of active elements decreasing as one goes out toward the edge of the aperture. This density variation effectively provides amplitude weighting of the aperture illumination on transmit. As a result low near sidelobes are achieved. The density tapering was designed to provide a 35 dB Taylor weighting on transmit. Array thinning also permits the achievement of a narrower beamwidth than would have been possible if a full array having the same number of active radiating elements had been used. The total number of dummy elements is 19,049 for a total of 34,769 elements in the array. If desired, it is

Fig. 3 Multifunction UHF phased array radar AN/FPS-115 (PAVE PAWS).

possible at a future date to replace the dummy elements by active elements in order to increase the system's sensitivity by 9 dB.

Detailed parameters for this radar, as well as many of the radars to follow, are given in Table 1 of Chapter 1 in Reference 1 and in Reference 2.

PAVE PAWS Radar[1,7,8]

The impressive advancement made in large UHF phase-phase array radars over the past decade is exemplified by the PAVE PAWS system of Figure 3. The first large phased array radar was the UHF AN/FPS-85[1]. This radar used separate antennas for transmit and for receive because of the expense of using a duplexer. The power generation was supplied by tetrode tubes driving each transmit radiating element. In contrast, the PAVE PAWS system transmits and receives on the same antenna. In addition, the transmitter is all solid state, using bipolar silicon UHF transistor power amplifiers with each active element of the array driven by a transceiver solid state module. These modules receive the exciter signal which is first passed through a phase shifter on transmit and then through a solid state UHF power amplifier. On receive the signal comes from a radiating element and passes through a low noise receiver in the transceiver module after which it is switched through the same phase shifter before going to a beam former. PAVE PAWS consists of two array antenna faces instead of one. Each site, in effect, consists of two radars (housed in one building) instead of one. The two faces provide continuous coverage for two 120° azimuth regions for a total coverage of 240° in azimuth.

Figure 4 shows a simplified block diagram of the radar. As in the case of the COBRA DANE, the array is divided into subarrays. Each subarray consists of 32 active modules feeding 32 radiating elements. There are 56 such subarrays. These subarrays are in turn driven by modules identical to the array element modules except that they do not use the unneeded receiver portion. A predriver drives the 1 to 56 divider which supplies the subarray signals. This predriver also uses a module identical to the array element module except that again it does not use the receiver portion. Thus to build the PAVE PAWS array a large number of the UHF modules are employed.

Array thinning is used as was done with the COBRA DANE. There are 885 dummy elements and 1792 active elements over a 72.5 ft diameter. Provision has

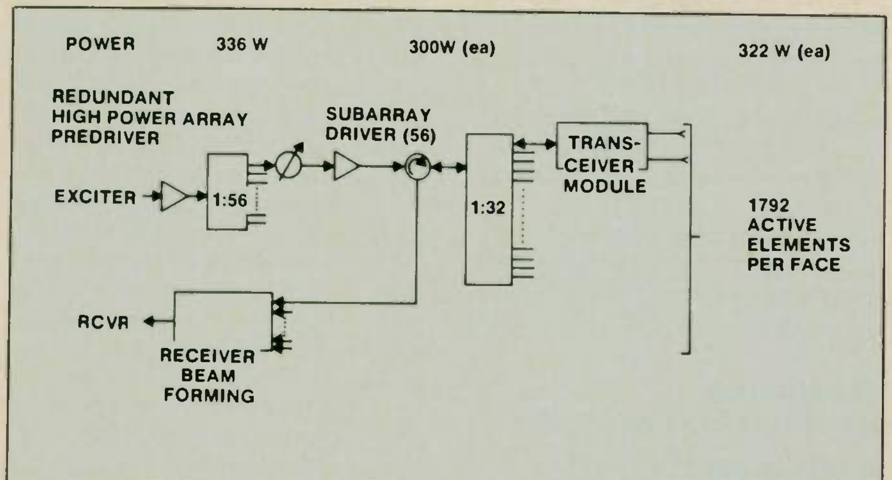

Fig. 4 Simplified block diagram of PAVE PAWS system.

Fig. 5 Patriot (formerly SAM-D) system.

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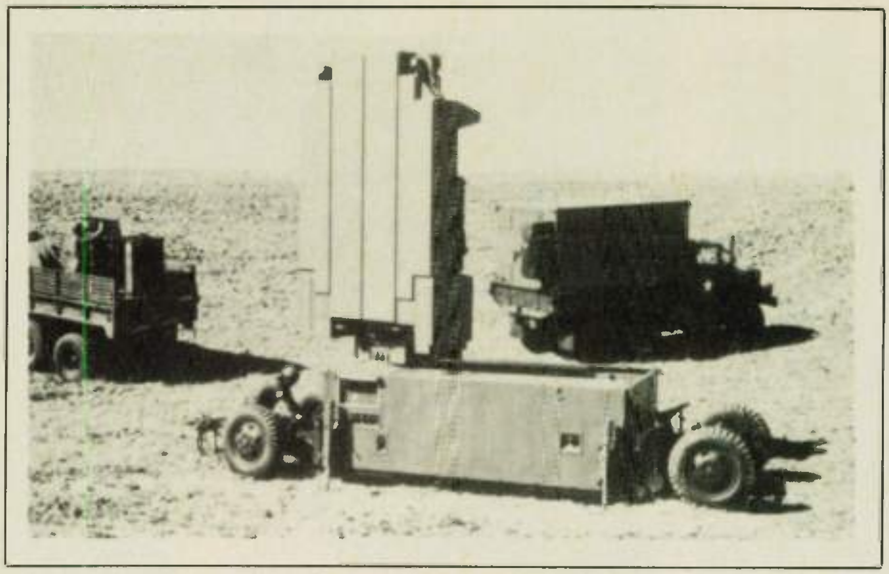

Fig. 6 Fire finder AN/TPQ-37 artillery locator radar.

been made to allow the active area of the antenna to grow to 102 ft if desired at a future time. The system is intended for early warning and attack characterization of submarine launched missiles. A secondary mission is the support of the Air Force Space Track program. This 420 to 450 MHz radar has a per face peak power of 600 kW and an average of 150 kW. Its range is 3,000 nmi for a 10 m^2 target.

PATRIOT Radar [1]

The PATRIOT system illustrates the ability of a single multifunction phased array radar to replace a system using many dish radars. The multifunction phased array PATRIOT tactical air defense system of Figure 5 replaces the HAWK type system which typically has five radars and even then is limited in the number of simultaneous engagements it can perform by the number of tracking/illuminator radars dedicated to each battery control center during an entire missile flight. The PATRIOT radar is space fed and uses about 5,000 ferrite phase shifters.

Fire Finder AN/TPQ-37 [10]

The AN/TPQ-37 of Figure 6 represents a phase-phase array radar now in production. An interesting feature of this radar is its use of the moving target detector (MTD) technique for suppressing clutter[2,9,10]. This S-band artillery locator radar scans ±45° in azimuth and a few degrees in elevation[2,9].

Multiple-Target Instrumentation Radar (MIR) [11]

Instrumentation radars are now taking advantage of the high accuracy, multiple target tracking capability afforded by phased array radars. One such radar is the MIR developed by Raytheon for the Naval Air Systems Command (NASC); see Figure 7. This C-band 12 ft space fed 8,973 element array system can track up to 16 targets simultaneously. Not only does the system reduce the number of tracking radars required, it also eliminates the radar-to-radar tracking errors that exist when multiple tracking radars are used to track many targets simultaneously. The MIR system has a relative tracking error accuracy of 0.05 mr, absolute tracking accuracy of 0.1 mr, and range accuracy of 2.0 yd rms.

Thomson-CSF Prototype Multimode Phased Array System [12]

Figure 8 shows the Thomson-CSF multimode phased array system being developed for the French Navy. This antenna system is intended for shipboard surveillance, tracking and weapons guidance involving multiple targets. In order to obtain 360° of azimuth coverage the phase-phase scan array is mechanically rotated about the vertical. The electronic scanning allows the beam to be scanned faster or slower than the mechanical azimuth motion. The array uses diode phase shifters and spiral radiators.

[*Continued on page 30*]

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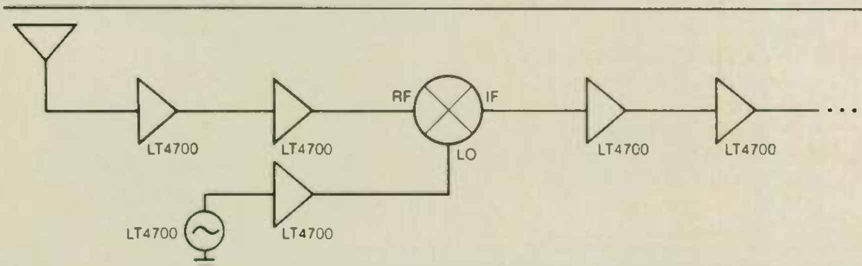

*Noise measure M_{min} is the system noise figure of an infinite cascaded chain of identical amplifier stages:

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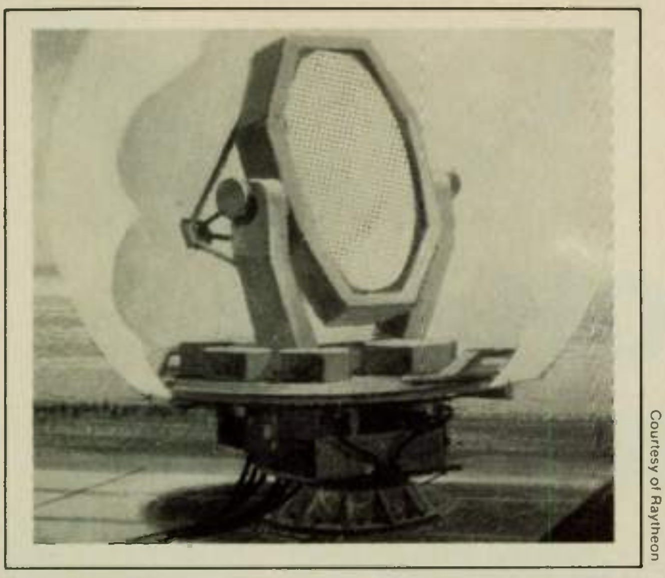

Fig. 7 Raytheon multiple-target instrumentation radar (MIR).

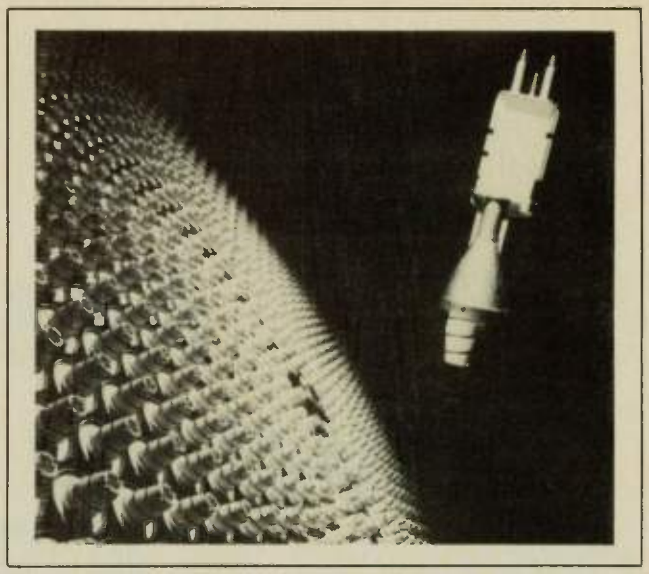

Fig. 8 Thomson-CSF multimode phased array.

Other Phase-Phase Arrays Systems Under Development or in Production

Examples of other phase-phase array systems include the airborne multifunction Electronically Agile Radar (EAR) originally intended for use aboard the B-1 and now possibly slated for use on the B-56 and F-111[1,13]; the Siemens 243 dipole element S-band VM 256 surface-to-air missile (SAM) multifunction adaptive array[14], see Figure 9; the 2200-element OREST under development by Siemens; the AEGIS AN/SPY-1A S-band shipboard radar which uses four array faces to achieve 360° azimuth coverage[15]; the Forschungsinstitut fur Funk und Mathematik experimental phased array system (ELRA) which uses sequential detection and adaptive processing[16,17,42]; the Toshiba and Mitsubishi phased array radars under development[18] (whether these are phase-phase array radars has not not been explicitly indicated in the open literature); the Thomson-CSF experimental Louxor 2,500 dipole phase array C-band system intended for short range air defense[19], see Figure 10; and the space based radar system for which technology is being developed, see Figure 11[20].

Fig. 9 Siemens VM 256 multifunctional radar system.

[Continued on page 32]

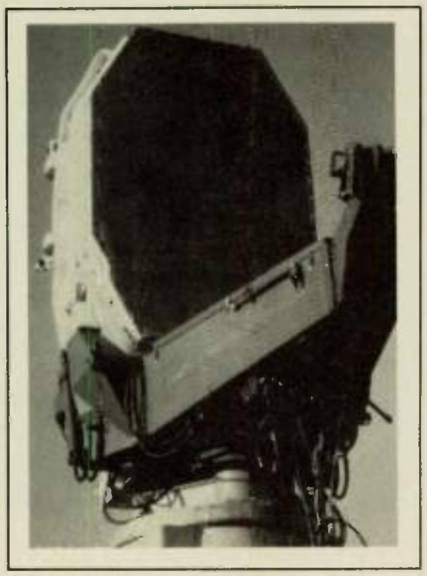

Fig. 10 Thomson-CSF Luoxor experimental electronically-scanned radar for short range air defense application.

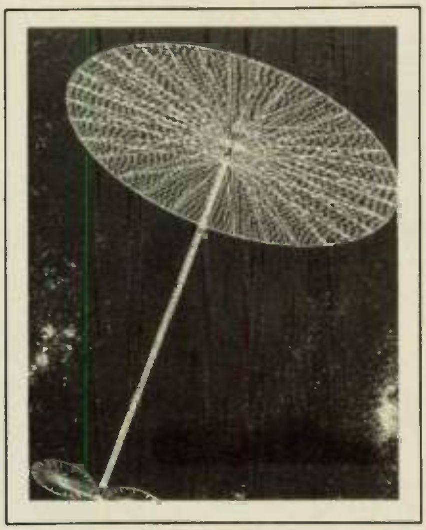

Fig. 11 Space based radar system.

LIMITED SCAN PHASED ARRAY SYSTEMS

In order to reduce the cost of phase-phase array radars the technique of limited scanning has been developed. This technique is useful where coverage over small angles such as ±10° is required instead of ±60°. By the use of this technique the number of phase shifters required for the system is significantly reduced, in some cases by an order of magnitude.

The X-band AN/TPN-25 precision approach radar (PAR) of the AN-TPN-19 system built by Raytheon is an example of such a system; see Figure 12. This system guides the aircraft during final approach and only requires a limited coverage sector, 15° in elevation by 20° in azimuth[1]. To achieve this limited angle coverage a phased array is used to illuminate a relfector having a hyperbolic surface. This feed phased array is space fed by a horn as shown in Figure 13. By changing the plane setting of the phase shifters in the array, the array images the horn feed to appear in a different position. Moving the horn feed image effectively moves the position of the beam formed by the main reflector. The number of phase shifters needed in the array feed is 824.

A fixed low cost version of the AN/TPN-25 radar has also been

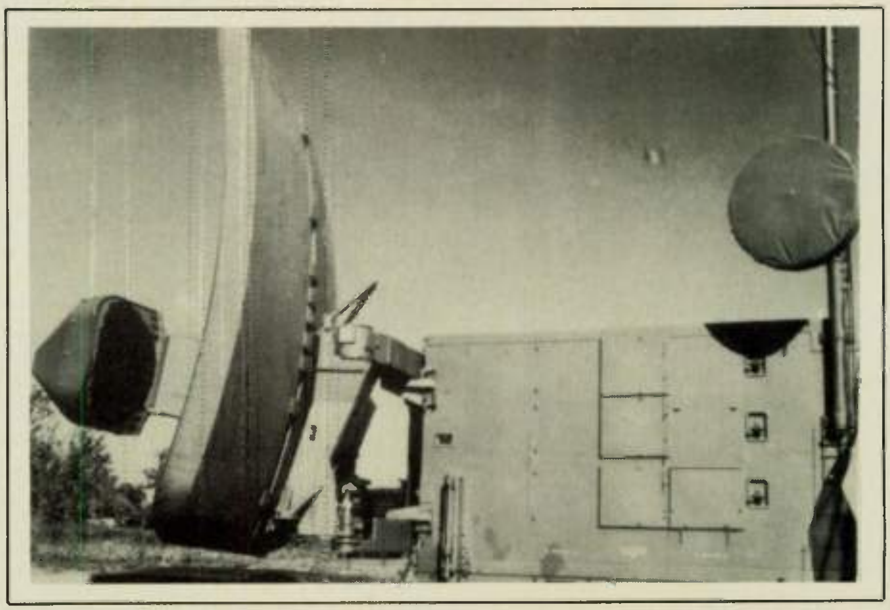

Fig. 12 AN/TPN-25 precision approach radar (PAR) of AN/TPN-19 system.

[Continued on page 34]

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W250G	5-250 MHz	43	1.3	1.5	+ 25		Wide Dynamic Range
P150P	.08-150 MHz	60	1.3	1.5	+ 30		Pulse Amplifier
P2GS	.05-2 GHz	20	5.5	6.0	+ 23		Increase Sweeper Power
L215GB	2.15-2.165 GHz	21	3.0	3.5	+ 7		MDS Preamp.
L13GE-1	1.2-1.4 GHz	25	1.7	2.0	+ 8		Low Noise/Radar
WV110A	1-110 MHz	60 dB, 70 dB Video	1.0	1.2	+ 15		Video Amp, IF & Video Output
2. ULTRA LOW NOISE AMPLIFIERS							
L60E	60 ± 5 MHz	60	1.0	1.1	+ 10		Low Noise IF
W500E	5-500 MHz	30	1.2	1.4	0		Low Noise Wide Band
L1GE	.9-1.1 GHz	25	1.6	1.8	+ 6		Low Noise
W1GE	.01-1 GHz	20	1.6	1.8	− 3		Wide Band Low Noise
W2G10B	.01-2 GHz	32	3.0	3.5	0		Wide Band Low Noise
W15GB3	.05-1.5 GHz	30	1.7	2.0	+ 5		Wide Band Low Noise
W2G2H	1-2 GHz	30	2.2	2.5	+ 5		Wide Band Low Noise
3. MEDIUM POWER LINEAR AMPLIFIERS					Typ.	Min.	
P100L	.08-100 MHz	27		5.0	+31	+30	Wide Dynamic Range
P150H2	.1-150 MHz	27		6.0	+31	+30	Low Distortion
P500A	2-500 MHz	37		3.0	+31	+30	Wide Dynamic Range
P1000E	.02-1 GHz	20		6.0	+23	+21	Wide Band Linear
P1GB	.01-1 GHz	30		6.0	+30.5	+30	Distribution, Driver Amp.
P10GD	.5-1 GHz	30		9.0	+34	+33	Wideband Linear
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produced. This is the AN/GPN-22. Its coverage in elevation is 8° instead of 15°[1]. Only 443 shifters are needed in its array feed.

These two PAR systems can track up to 6 aircraft simultaneously and maintain performance in heavy rain. Specifically the system is capable of tracking a 1 m^2 target such as a small jet aircraft at a range of 10 nmi through a 5 nmi cell of rain falling at a rate of 2 inches/hr[22]. To achieve this performance the system uses a high power cross field amplifier at the output of the TWT tube during heavy rain. This tube provides a 15 dB increase in the system output power. The system also uses circular polarization which provides 20 dB of rain attenuation. Noncoherent MTI is also used for rain attenuation. Finally the system range resolution is improved to about 5 ft during track to help reduce the clutter return relative to the target return.

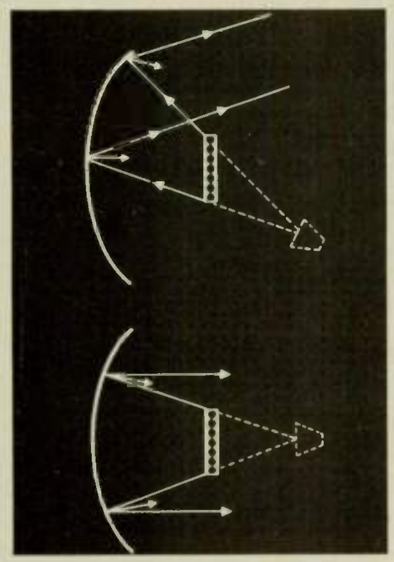

Fig. 13 Principle of limited scan antenna system.

Eleven of the AN/TPN-25 radars have been produced for the US Air Force and 50 AN/GPN-22's have been produced for the US Air Force and overseas. The AN/TPN-25 represents the first phased array radar to be put into production.

Another limited-scan phased array is the Thomson-CSF Artois shown in Figure 14. This system electronically scans in a cone having a vertex angle of at least 10°[19]. The antenna can be mechanically trained to any direction.

Still another limited-scan phased array is the Thomson-CSF 400

Fig. 14 Thomson-CSF Artois limited scan multitarget tracking radar.

[Continued on page 36]

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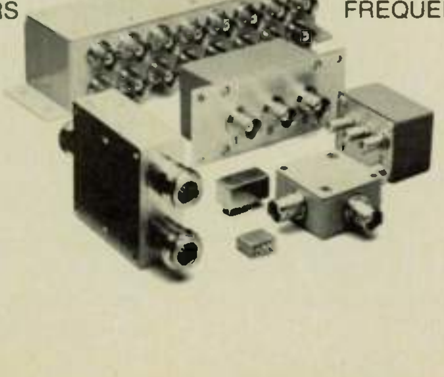

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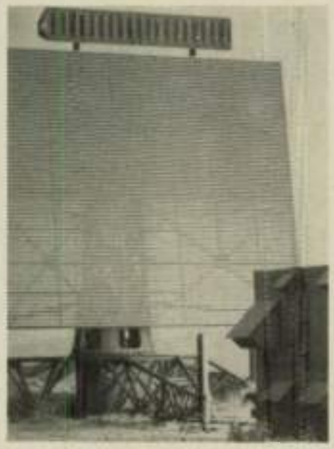

Fig. 15 Frequency scanned 3-D Series 320 radar.

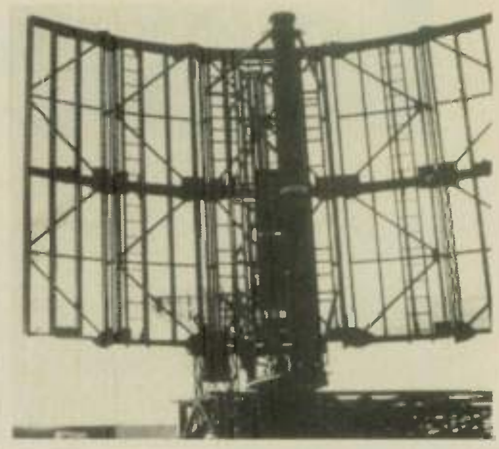

Fig. 16 Plessey S-band AR-3D frequency scanned radar.

MHz STRADIVARIUS satellite tracking radar using a 13m by 20m main cylindrical reflector[44,45]. Its sweep is ±20° in azimuth and 5.5° in elevation (from 12° to 17.5°).

ARRAY RADARS USING ONE-DIMENSIONAL ELECTRONIC SCANNING

Low cost 1-dimensional (1-D) electronic scanning systems have seen greater popularity. They are increasingly using such features as pulse compression, coherent processing, solid state transmitters, digital processing, CFAR, frequency agility, low antenna sidelobes, and improved built-in test equipment (BITE). 1-D electronic scanning is used extensively in 3-dimensional (3-D) radars where a pencil beam is electronically scanned in elevation while the antenna is rotated 360° in azimuth in order to provide full coverage in elevation and azimuth. The elevation scan can be obtained by either using frequency scanning or phase scanning.

Examples of recently developed successful 3-D radars using frequency scanning in elevation are the ITT Gilfillan Series 320 S-band radar of Figure 15 (which has low antenna sidelobes)[23] and the Plessey S-band AR-3D radar of Figure 16 which scans all elevation angles within one pulse (instead of covering one elevation angle with one pulse, the pulse frequency being changed from pulse to pulse to cover different elevation angles as is done with the Series 320)[24,25].

Examples of recently developed successful 3-D radars using phase scanning in elevation are the GE 592 solid state L-band radar of Figure 17 intended for modernization of the Alaskan Air Defense

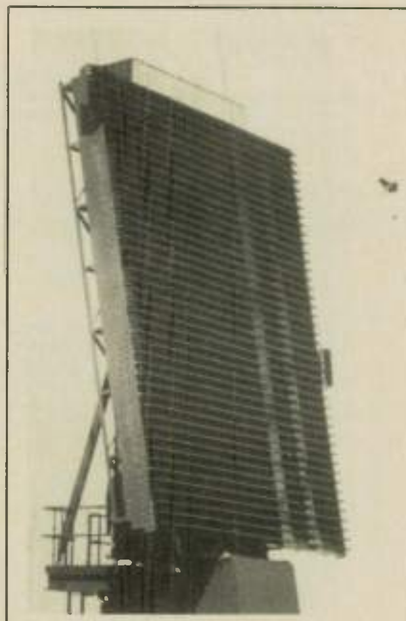

Fig. 17 GE 592 solid-state L-band 3-D radar.

radar network[26,27]; the coherent Thomson-CSF S-band defense TRS 2215 (Figure 18) and TRS 2230 mobile and fixed radars, respectively; the AWACS airborne surveillance system[37]; and the Selenia RAT-31S S-band surveillance radar of Figure 19 which has 3 beams in elevation that are independently phase scanned and which use 3 completely independent, randomly selected carrier frequencies[28,29]. The achievement of

[Continued on page 38]

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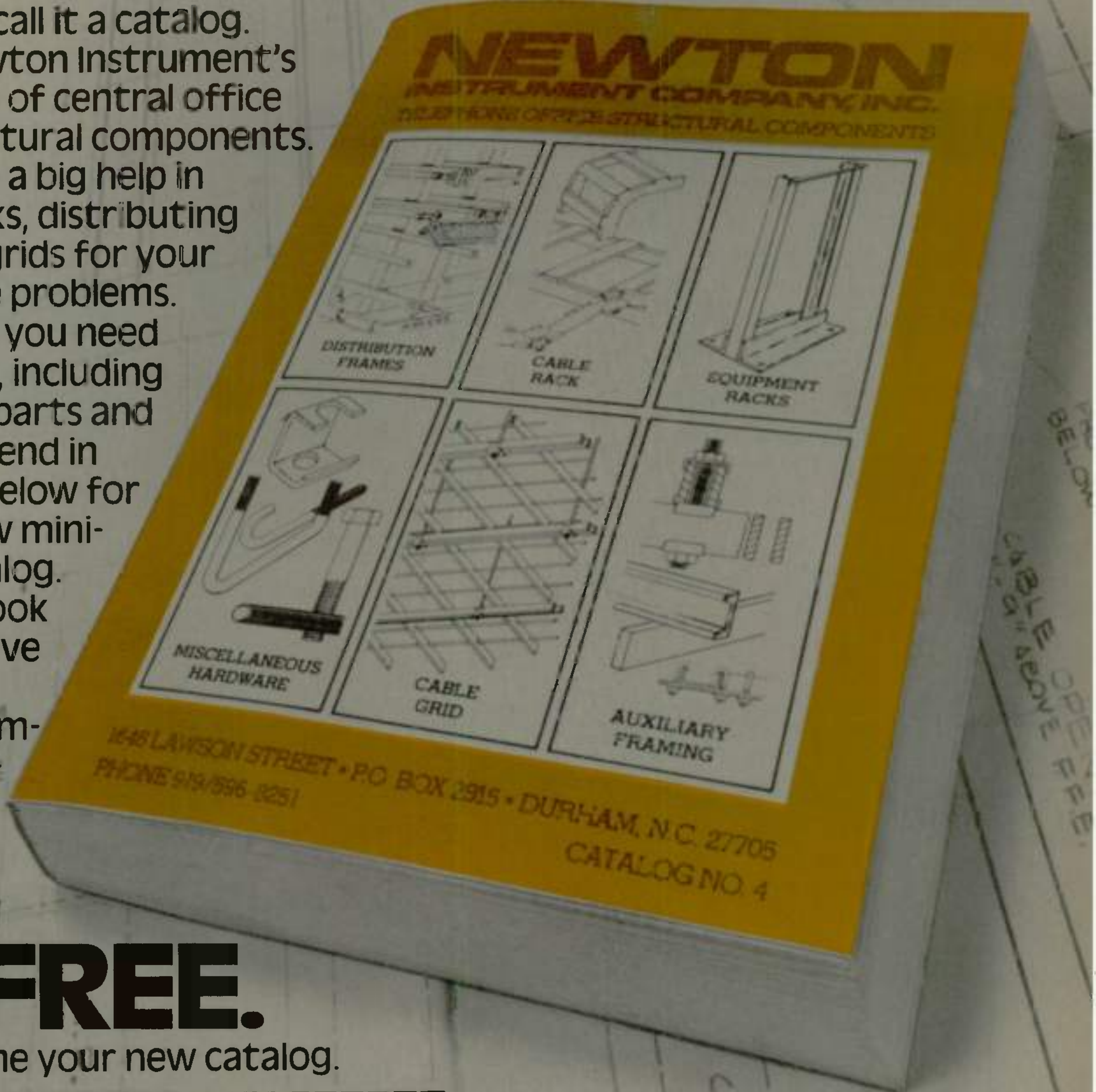

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low antenna sidelobes and the availability of multiple frequency operation provide the RAT-31S system with good ECCM characteristics. More than 20 RAT-31S radars have been ordered[30].

The prototype 3-D Telefunken Radar Mobile Search (TRMS)[19,31] is another example of a system which uses 1-D elevation scanning and mechanical scanning in azimuth. This mobile search C-band array radar, shown in Figure 20, has the useful featur s of polarization agility (horizontal, vertical, and circular), low sidelobes, the ability to continuously vary the antenna beamwidth in elevation if desired (by a factor of up to 8), and an ability to be raised higher which gives the antenna good low altitude target coverage capability. When using circular polarization the integrated can-

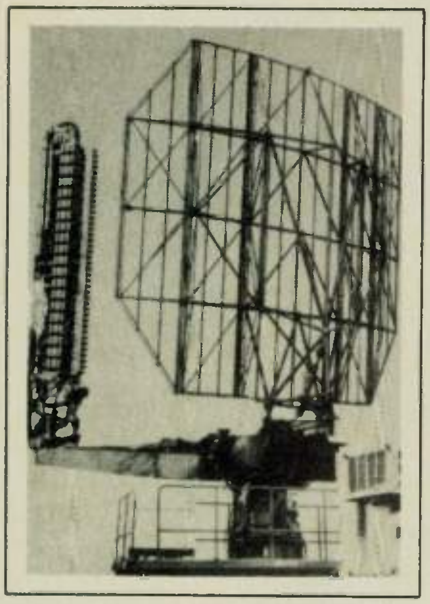

Fig. 18 Thomson-CSF 3-D S-band air defense TRS 2215 radar.

Fig. 19 Selenia S-band 3-D RAT-31S system.

Fig. 20 Telefunken radar mobile search (TRMS).

[Continued on page 40]

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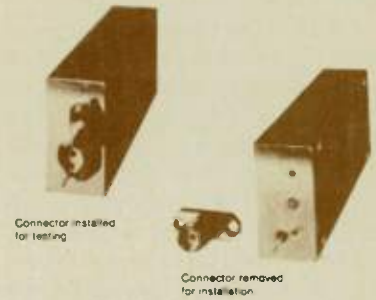
8081-1602	2.0-8.0	14.0	5.0	+7	2.0:1	80
8081-1603	2.0-8.0	22.0	5.0	+7	2.0:1	180
8081-1604	2.0-8.0	28.0	5.0	+7	2.0:1	220
8081-1605	2.0-8.0	34.0	5.0	+7	2.0:1	260
8081-1611	2.0-6.0	25.0	4.5	+14	2.0:1	140
8081-1612	2.0-6.0	34.0	4.5	+14	2.0:1	180
8081-1201	6.0-18.0	14.0	8.0	+9	2.0:1	180
8081-1202	6.0-18.0	18.0	8.0	+9	2.0:1	220
8081-1203	6.0-18.0	23.0	8.0	+9	2.0:1	260



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8 - 18 GHz

Percent BW	Isolation dB MIN	Insertion Loss dB MAX	VSWR MAX
10	20	0.4	1.25
11	18	0.4	1.30
12	17	0.5	1.35
13	16	0.5	1.40
14	15	0.6	1.45
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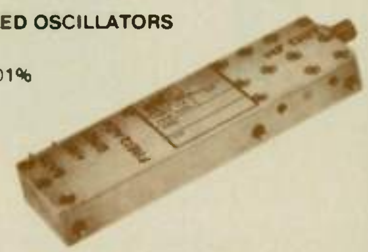

MODEL NUMBER	FREQUENCY (GHz)	POWER OUTPUT AT +25°C (dBm min.)	OUTPUT HARMONICS AND SPURIOUS (dBc)	CURRENT (ma nom.)
5044-1311	2.5 - 4.0	+10	>40	100
5044-1611	4.0 - 8.0	>10	>40	110
5045-1911	8.0 - 10.0	+10	>40	120

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[From page 38]

cellation ratio for rain clutter rejection is about 20 dB (approximately 0.8 dB axial ratio).

Polarization agility is achieved by using two sets of azimuth corporate feed structures per row of the array, with each corporate feed structure having its own ferrite phase shifters. The outputs of one pair of these azimuth corporate feed structures feed a row of dual polarized radiating elements; see Figure 21. Adjusting the relative phase between a pair of azimuth corporate feed structures provides the desired polarization control.

HEMISPHERICAL AND OMNIDIRECTIONAL COVERAGE

For many applications a hemispherical coverage is needed. Often a 3 or 4 faced phased array system is used to achieve such a coverage with a phase-phase array. An alternative approach is to use a single phase-phase array which is mechanically rotated 360° as was done for the Thomson-CSF multifunction array radar of Figure 8. Still another approach is to use the Dome antenna concept [1,2,32,33]. In this approach a phase-phase planar array is laid flat on its back with its boresight pointing straight up. Then over this array is placed in effect a "dielectric" lens. When the planar array scans the beam to 60° away from vertical the "dielectric" lens bends the rays down to 90° away from the vertical, that is, to point horizontally. When the planar array points the beam vertically, the lens does not bend the ray at all so that the beam is still pointed vertically. For all scan angles of the planar array between 0° and 60° the beam is scanned as it goes through the "dielectric" lens from 0° to 90°, thus achieving hemispherical coverage. If desired, the lens can be made stronger so that the beam is bent 20° or 30° below the horizon when the planar array scans the beam to 60° away from boresight. Such increased coverage is needed for shipboard applications to compensate for ship roll and pitch. The "dielectric" lens is actually a dome in which are embedded phase shifters which electrically cause the dome

[Continued on page 42]

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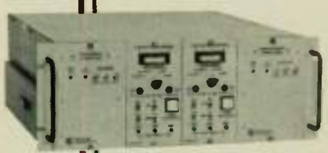

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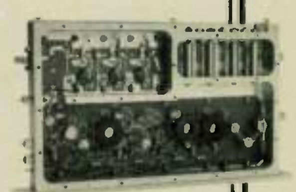

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[*From page 40*] **ARRAY RADARS**

to act like a "dielectric" lens as described above. Recently the Dome Antenna has been combined with the Rotman Lens Fed Multi-beam Array to realize octave band-widths, multiple simultaneous beam capability and better side-lobes[34]; see Figure 22. A 2-D demonstration system has been built which is capable of operating from 8 to 15 GHz. Twenty-nine beamports in the Rotman lens provide 180° of coverage[35]. No operational dome antenna has been built to date.

LOW SIDELOBE ARRAYS

Sidelobe levels of 40 to 50 dB down have been reported as achievable through the use of arrays made of edge-slotted wave-guides[36]. An operational slotted waveguide array which achieves very low sidelobes is that of the AWACS antenna[37]. A new ultra-low sidelobe (ULSA) slotted wave-guide antenna has been developed for the 3-D stacked beam AN/TPN-43[40].

The Marconi stacked-beam, 3-D, L-band Martello radar uses an array made up of 60 rows of dipoles accurately fed by stripline feed networks to provide low sidelobes (-45 dB 15° away from boresight)[41, 46,47]

The problem of attaining low sidelobes in a tactical phase-phase steerable array has been studied by Patton[38]. In this study the arrays were assumed to be subarrayed with column beamforming being done first followed by azimuth beamforming. He indicated that better than 50 dB rms sidelobes can be achieved for this antenna over all space if the array errors were: element-to-element, 4.8% in amplitude and 11.0° in phase; subarray-to-subarray, 4% and 7.8°, respectively; column-to-column, 2.8% and 3.4°, respectively. A 5-bit phase shifter was assumed.

COMPONENT DEVELOPMENT

Extensive work is going on in the development of monolithic microwave technology[43,48]. Such technology will eventually lead to low cost phased array phase shifters, low noise amplifiers and power amplifiers, and in turn low cost phased array radar systems.

[*Continued on page 44*]

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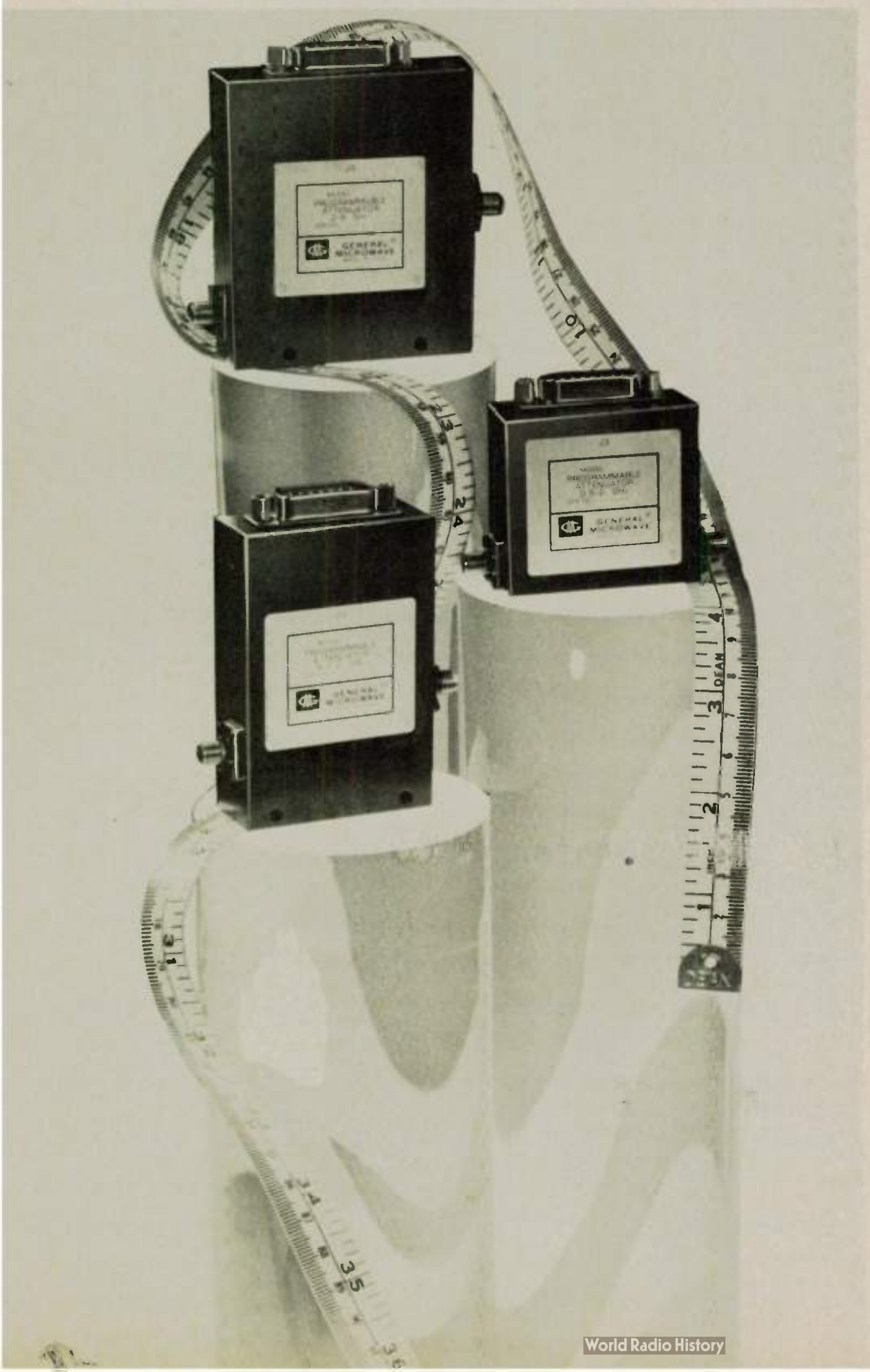

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[*From page 42*]

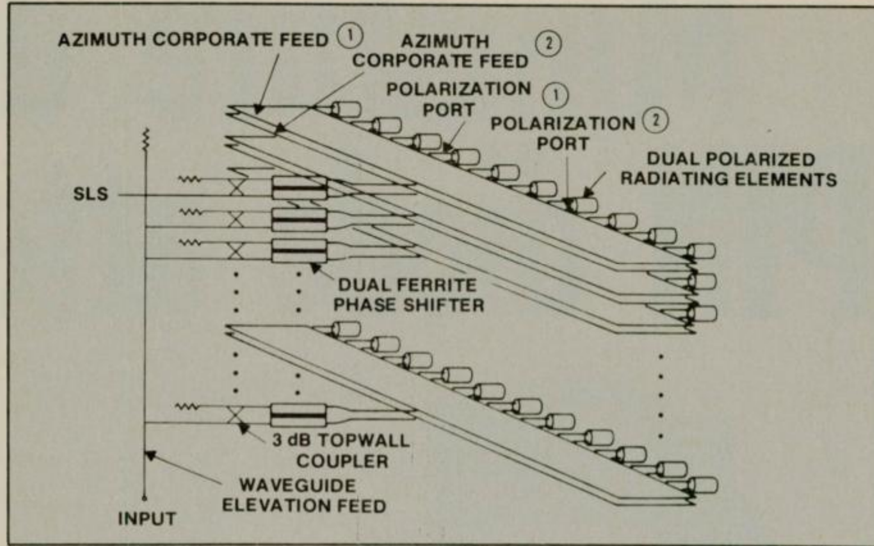

Fig. 21 Schematic array of TRMS antenna.

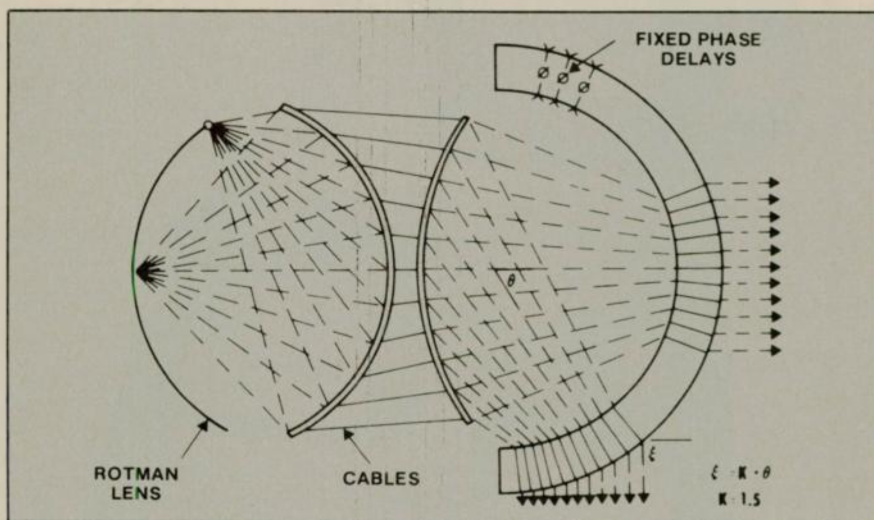

Fig. 22 Wide angle array fed lens (WAAFL)

GaAs FET's will lend themselves better to monolithic integration because they are fabricated on semi-insulating substrates[39].

CONCLUSIONS

From the large number of array systems that are in production, that have been built and that are under development and from the extensive technology efforts going on in the array components and techniques, it is apparent that the future will see significant advances in array systems and their increased use. On the other hand, we should bear in mind that simple dish systems will still be used where they satisfactorily do the job because of their low cost.

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[*Continued on page 61*]

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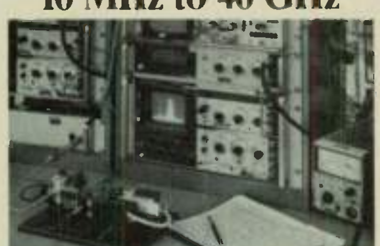

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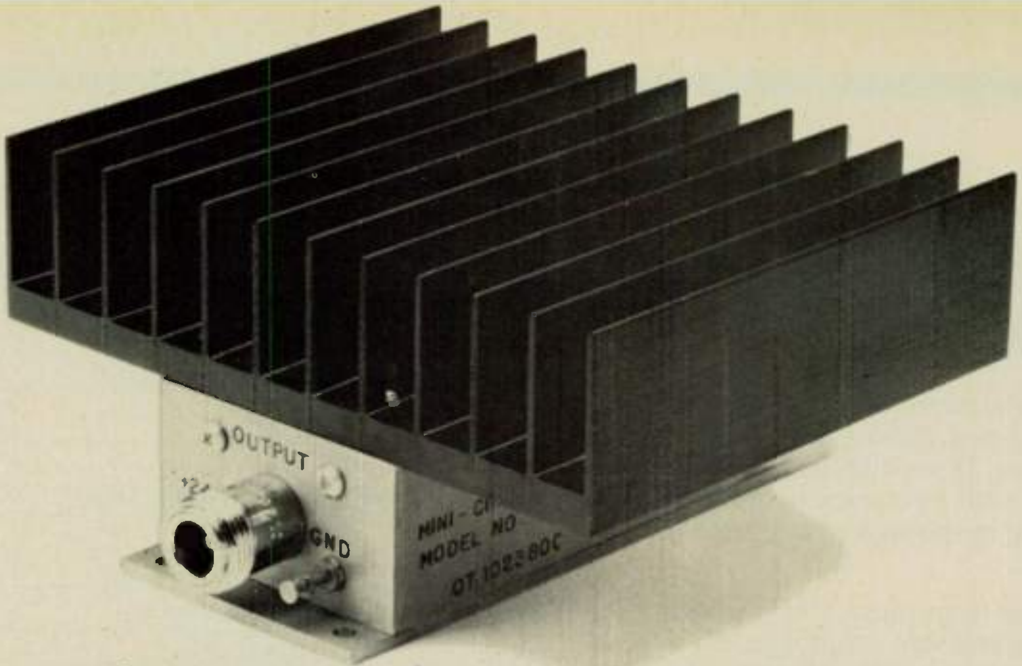

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ZHL-3A	0.4-150	24 Min	±1.0 Max	+29.5 Min	11 Typ	+38 Typ	+24V	0.6A	199.00	(1-9)
ZHL-1A	2-500	16 Min	±1.0 Max	+28 Min	11 Typ	+38 Typ	+24V	0.6A	199.00	(1-9)
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ZHL-2-8	10-1000	27 Min	±1.0 Max	+29 Min	10 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)
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REV ORIG

News from Washington

GERALD GREEN, *Washington Editor*

DOD, THE NEW SHERIFF

A little-known provision of the fiscal year 1982 Department of Defense authorization could create new microwave markets.

Posse usually refers to a band of men assisting the sheriff in pursuit outlaws. Posse Comitatus, however, refers to the Congressional Act of 1878 which banned the military from civilian law enforcement. A provision of the FY 1982 DoD authorization now being enacted by Congress would change the Act to allow the military to participate in civilian law enforcement.

The new provision would allow the use of military equipment, military personnel, and facilities in enforcing the law provided military preparedness is not adversely affected.

If enacted, the change to Posse Comitatus is expected to attract the interest of non-DoD agencies like Treasury's Customs Service, the Drug Enforcement Agency, and the Coast Guard in the utilization of surveillance systems aboard military aircraft like the AWACS, P-3, and the E-2C, to detect and track aircraft, ships, and even surface vehicles carrying contraband.

To microwave specialists previously concerned only with microwave requirements for military use, the change to Posse Comitatus could open up new markets with non-DoD agencies, as well as introducing new requirements for military aircraft, ships, and facilities.

JCS AWAITING RESPONSE TO $7 BILLION REQUEST FOR C^3

In the midst of proposed cutbacks in the Pentagon's budget the Joint Chiefs of Staff (JCS) are anxiously awaiting the administration's response to their recent request for a $7 billion set-aside over the next five years for command-control-communications (C^3) systems.

The importance of secure C^3 has been accentuated in recent weeks in Washington, D.C. partly because of growing awareness of the effects of electro-magnetic pulse (EMP). Public discussions have revealed that the detonation of a few nuclear weapons high above the U.S. could render useless most C^3 systems. Some of the requested funds would undoubtably be used to "harden" communications and radar systems within the U.S. and throughout the world.

Although $7 billion has been requested now some knowledgeable DoD officials believe that much more will be required (up to $25 billion) in order to provide the U.S. with a secure and hardened C^3 network.

U.S.-CANADIAN JSS NOW 75 PERCENT ON-LINE

A modernized air surveillance system is moving ahead with 62 radars now "on-line." The sensors, elements of the "Joint Surveillance System" (JSS), are being acquired by the Air Force Systems Command's Electronic Systems Division.

The JSS, an up-to-date peacetime surveillance and control system for the United States and Canada, will replace the 25-year-old SAGE (Semi-Automatic Ground Environment) System, BUIC (Back-Up Interceptor Control), and Manual Control Systems.

Tied into a continent-spanning network of mostly Air Force and Federal Aviation Administration (FAA) radars, the skies will be searched from these dual-use sites. Other military radars will also feed data into the system.

The latest sensors joining the network are at Whitehouse, FL, and Makah, WA. Seventy-five percent of the system's 86 radars (46 continental U.S., 24 Canada, 14 Alaska, and 2 Hawaii) now are operating; the remaining sensors will be tied into the network on a staggered basis between now and mid-1982.

The JSS streamlining consolidates the North American Aerospace Defense Command's (NORAD) military air surveillance radars with those of the FAA.

The contractor on the $200M program is Hughes Aircraft Company of Fullerton, CA.

ARMY PLANS COMMUNICATION MERGER

The Army Communications Command at Fort Huachuca, AZ, and the Army Computer Systems Command at Ft. Belvoir, VA, will be merged into a new Automation and Communications Command according to Army sources.

Specifics on the impact of the organizational change have not yet been announced but it is understood that the Army Communications Command at Ft. Huachuca has been given the lead in planning the new Command.

The date that the new Command structure is to take effect has not yet been announced.

NBS RESEARCH FORGES STRONG LINK WITH INDUSTRY

The National Bureau of Standards (NBS), the federal government's science and engineering measurement laboratory, has the answer to those in industry who would like to participate in government sponsored engineering and scientific projects and even join in the preparation of specifications and procedures — Become a partner in the Research Associate Program at NBS!

NBS is now hosting about 100 scientists and engineers from private companies and from trade and professional associations and other organizations. With their salaries paid by their sponsors, the associates work on a wide range of projects that includes electromagnetic interference, robotics, and other projects associated with electronics.

"The interaction is truly a two-way street," says Peter de Bruyn, NBS industrial liaison officer and coordinator of the program.

"Research associates can benefit from the use of our facilities and from the opportunity to consult with our diverse professional staff," he explains. "They then can take newly developed technology back to their organizations for prompt application."

NBS has invited companies and trade and professional organizations interested in the Research Associate Program to write to Peter de Bruyn, Industrial Liaison Officer, National Bureau of Standards, Washington, D.C. 20234 or call him at 301/921-3591. ■

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Here are some examples of their capabilities.

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Application	Limiter Amplifier	Gain Tracking	Power Amplifier
Frequency	I/J	12-18GHz	4 6 8 11 & 13GHz bands
Power Output	Level to ± 1.5dB	Track to ± 1.0dB	10-22W
Primary Source	115V AC 400Hz	—	21-28V DC 42-56V DC
Gain	—	32dB	40dB
N.F.	—	11dB	26dB
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International Report

GERALD GREEN, *Washington Editor*

PUERTO RICO ATTRACTING HIGH TECHNOLOGY FIRMS

The growth of electrical, electronic, precision instrument, and computer manufacturing in Puerto Rico is becoming more pronounced and today ranks among the fastest of any major industrial location in the world.

Employment in these segments increased by 84 percent to over 35,000 workers in just four years. Over 250 plants, representing more than $1 billion in invested capital, now generate 23 percent of total Island manufacturing employment compared with 13 percent four years ago in the high technology categories.

Meanwhile, computer product shipments from the Island to mainland customers grew from $26.8 M to $63.3 M in fiscal 1980, a one-year increase of 136 percent.

"We're attracting the high-technology companies for two main reason," says Mr. Jose R. Madera, Chief Executive of the Economic Development Administration of Puerto Rico. "Our workers' productivity and our incentives package, which is unrivaled under the U.S. flag, insure that output per dollar of production wages is 50 percent higher in Puerto Rico than it is for mainland U.S. manufacturing on average."

Currently, the Economic Development Administration has a backlog of some 90 manufacturing projects that are pending establishment on the Island.

COMMERCE PROBES DUMPING OF HIGH-POWER AMPS BY JAPAN

The Commerce Department is investigating whether imports of high-power, microwave amplifiers and components from Japan are being sold in the United States at less than fair value.

The investigation follows a petition from Aydin Corporation, Ft. Washington, PA., and MCL, Inc., LaGrange, IL., which alleges that the sales are being made at less than fair value and are materially injuring the U.S. industry.

The products being investigated are radio-frequency power amplifier assemblies and their components specifically designed for transmission from fixed earth stations to communications satellites. The assemblies are used primarily for final amplification of signals transmitted to communications satellites.

The particular transactions mentioned in the petition are valued at $3.3 M in two contracts scheduled to be completed in 1982.

The Commerce Department's preliminary determination which is due by December 31, 1981, follows an affirmative preliminary determination of the injury charge made by the U.S. International Trade Commission in September 1981.

RAN DEVELOPS RADAR EVALUATION SYSTEM

Reports from Royal Australian Navy (RAN) radar experts indicate that their defense scientists at Salisbury, South Australia, have scored a major success in developing a new system for checking the performance of naval search radar.

Known as the Vertical Coverage Diagram (VCD) measuring system, the equipment gives the RAN an economic and rapid method of evaluating the performance of a ship's electronic eyes.

International Report

Radar experts have acknowledged for some time that the actual operating efficiency of most shipborne radars is much less than the advertised "book" value. A ship's radar is often distorted by reflections from the water and the ship's superstructure. Environmental effects, (e.g. moisture in the transmission lines) also contribute to degradation of shipborne radar systems.

The VCD equipment developed by the Defense Research Center at Salisbury in cooperation with the RAN trials and Assessing Unit, Sidnew, operates on the standard technique of transmitting radar beams to follow the path of a metal-coated plastic sphere dropped from an aircraft at altitudes between 3,000 and 4,500 meters.

Since the size of the rigid plastic sphere is constant throughout its descent, and its range and height are known, the VCD monitoring equipment onboard the warship under test can record the strength of the signal from different elevation angles to construct a "vertical coverage diagram."

From this data the exact coverage of the radar transmission can be determined, including the range and height within which the radar can detect targets effectively.

MODERN MICROWAVE NETWORK FOR AFRICA SCHEDULED IN 1982

In 1982, the Malagagasy Democratic Republic (Madagascar) will be equipped with the most modern microwave telecommunications system in Africa, according to the selected telecommunications contractor.

The network, which will be partly supplied with solar energy, is to provide television and telephone coverage of the most densely populated coastal area of the island.

The 100 million franc contract was awarded to Thomson-CSF and includes the turnkey installation of 45 microwave stations together with the supply and installation of 19 television transmitters.

RAF ORDERS NEW CONCEPT DEVICES FOR HF FREQUENCY MANAGEMENT

Britain's Royal Air Force has ordered a real time HF frequency management system which introduces a new concept in frequency management.

The system, which is being supplied by MCL (Marlborough Communications Ltd.), provides, for the first time, ionospheric propagation data in real time identifying exact usable HF communications frequencies from 2 megahertz to 30 megahertz by oblique ionospheric chirpsounding techniques.

MCL spokesmen indicate that the basic equipment modules, which were designed and developed by BR Communications Inc., can be configured into tailor-made HF frequency management systems for operation on land for tactical HF command and control, at sea for management of ship/shore HF circuits, and in the air for HF data circuits in support of AWACS. ■

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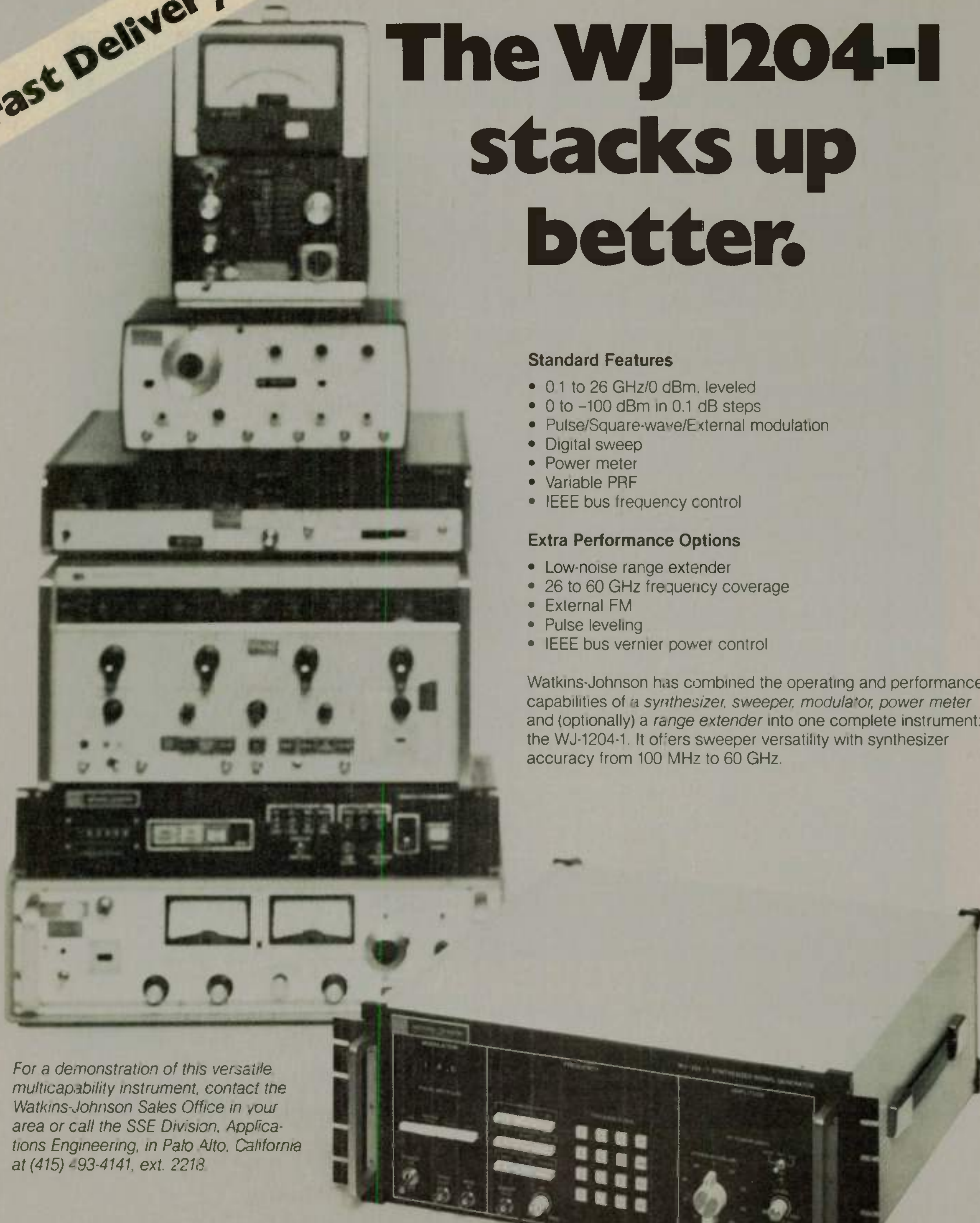

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Watkins-Johnson has combined the operating and performance capabilities of a *synthesizer, sweeper, modulator, power meter* and (optionally) a *range extender* into one complete instrument: the WJ-1204-1. It offers sweeper versatility with synthesizer accuracy from 100 MHz to 60 GHz.

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CIRCLE 36 ON READER SERVICE CARD

Around the Circuit

PERSONNEL

Irv Maltzer has been appointed General Manager of the Palo Alto Microwave Tube Div. (PAMTD) and a V. P. of Varian's Electron Device Group, of which PAMTD is a part . . . Systron Donner has appointed **Donald R. Rogers** V.P of the Washington, DC office . . . **Lionel Kirton** is the newly appointed V.P., Microwave Semiconductor Div., at Avantek, Inc. . . **M. V. Kreismanis** former assistant division manager of Hughes Aircraft electron dynamics division has been named division manager succeeding **Dr. Arne Lavik** who has been moved up to the Industrial Electronics Group office . . . M/A-COM announced that **Dr. Harry Van Trees** is the Senior V.P. of the new Eastern operations for LINKABIT . . . **Sidney K. Whiting** has joined Adams-Russell as Group V.P., Telecommunications and **Thomas W. Christenson** has joined Adams-Russell Antenna & Microwave Div. as eastern regional sales manager . . . **Kenneth Carr,** Senior V.P. of Microwave Associates, Inc., has been named a Research Associate in the Dept. of Radiation Oncology and Biophysics at Eastern Virginia Medical School . . . **Ernest W. Lattanzi** has been elected Pres. and Chief Operating Officer of Kelvin Manufacturing Co., replacing **James S. Galbraith** who was elected Chairman of the Board and CEO . . . **Bob Freischlag** has been promoted to Senior V.P. of Fujitsu Microelectronics, Inc. . . The new assistant chief engineer at Microwave Filter Co. is **William P. Johnson** who will head the company's new R&D department . . . Daico Industries, Inc has named **Kermit Heid** as Operations Manager and **Harold Ursenbach** is the new manufacturing manager of Daico's microwave integrated circuit department . . . **Bill Nicklin** has been named international marketing manager at Pacific Measurements, Inc.

CONTRACTS

Scientific-Atlanta Inc. has received a letter of intent providing for its manufacture and marketing of digital satellite earth stations for American Broadcasting Companies, Inc. (ABC) network radio affiliates. Typical earth station cost will be in the $10,000 range . . . A similar letter of intent from the NBC Radio unit of National Broadcasting Company, Inc. provides for Scientific-Atlanta, Inc. to supply digital satellite earth stations for use in the NBC Radio Network and the Source Network. . . **Aydin Corp.** has received orders for airborne and ground-based military communications equipment totalling $3.2M to support US government satellite receive station facilities . . . **Itek's Applied Technology Div.** was awarded a $16M contract for the US Air Force's ALR-46/69 radar warning systems and ground support equipment with deliveries scheduled for mid-1982. **Harris Corp.'s Farinon Div.** has received a $1.8M order for microwave radio equipment from British Telecom.

NEW MARKET ENTRY

New England Microwave Corporation of Hudson, NH announced its entry into the microwave solid-state components market. The corporation has launched a broad line of PIN diodes, multi-throw diode drivers and microwave switches. In addition to its standard switches, attenuators and detectors, custom-designed microwave circuits are offered. Contact: **Ronald J. Doherty,** New England Microwave Corp., 26 Hampshire Drive, Hudson, NH 03051 Tel: (603) 883-2900.

INDUSTRY

Leasametric Inc. has announced the formation of a new division, Metric Resources Sales Co., to sell used test equipment . . . **British Aerospace** and **McDonnell Douglas Corp.** have agreed to a joint AV-8B manufacturing program for the U.S. Marine Corps and U.K. Royal Air Force sales under which that work will be shared 40% — 60%. . . **The Sperry Division** of Sperry Incorporated has opened its new 55,000 square foot engineering and manufacturing facility in Rockland, Ontario. The $3.2M plant will initially employ 150 . . . HBR-Singer will begin relocation of key divisions from its current site in State College, PA to a 52,000 square foot facility under construction in Prince George's County . . . **Omni Spectra's** Microwave Component Division has completed relocation of the company's Cable Assembly Division from Tempe, AZ to Merrimack, NH. This product line will be integrated into the Microwave Component Division's Passive Component Group. **Microdyne Corporation** has commenced construction of a 40,000 square foot production facility adjacent to the corporation's existing satellite television products plant in Ocala, FL.

FINANCIAL

Varian Associates reports sales for the third quarter ending July 3, 1981 of $173M compared to $150M in 1980, earnings were $1.9M in 1981 compared to $5M for the third quarter last year. Earnings per share translate into 25¢ in 1981 and 62¢ for the 1980 period . . . **Sage Laboratories, Inc.** declared a year end dividend of 10¢ per share payable to stockholders of record September 25, 1981. Net income was $405K or 92¢ per share on sales of $3.1M compared to a net income of $395K or 91¢ per share on sale sof $2.4M in the 1980 period . . . **California Microwave, Inc.** results for the fiscal year ending June 1981 were net income of $2.5M, or 53¢ per share on sales of $57M compared with fiscal 1980 net income of $158K or 4¢ per share on sales of $38.1M. . . **Electromagnetic Sciences, Inc.** reports first half year net earnings of $423K or 31¢ per share on sales of $4.2M compared to first half 1980 net earnings of $106K, 11¢ per share, on sales of $2.5M ■

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High power DC output is delivered without the penalties of excess size, weight and heat generation of most conventional series-regulated, ferroresonant or SCR type power supplies. A principal advantage is the high efficiency design which allows the use of smaller components and heat sinks—heavy and bulky transformers are eliminated. Many other important design and performance features are standard, such as: self-cooling (including liquid-cooled models); protection against overload and excessive temperature; overvoltage and undervoltage load protection; modular design; BITE and remote operation.

Standard models are available with 24-30 or 35-50 VDC outputs up to 280 amps, 8400 watts. AC inputs of 110 thru 415 VAC, 50 thru 400 Hz. DC-DC converters up to 6000 watts with inputs from 12 thru 240 VDC, positive or negative. Typical 6000 watt converter has MIL-STD-704C DC input with regulated 28 VDC output. Standard 19-inch rack mounting for use as centralized power sources in aircraft, vans, aging racks, etc. Typical industrial and military applications include: transistor amplifiers, laboratory power, transmitters, phased arrays, lasers, computers, battery charging.

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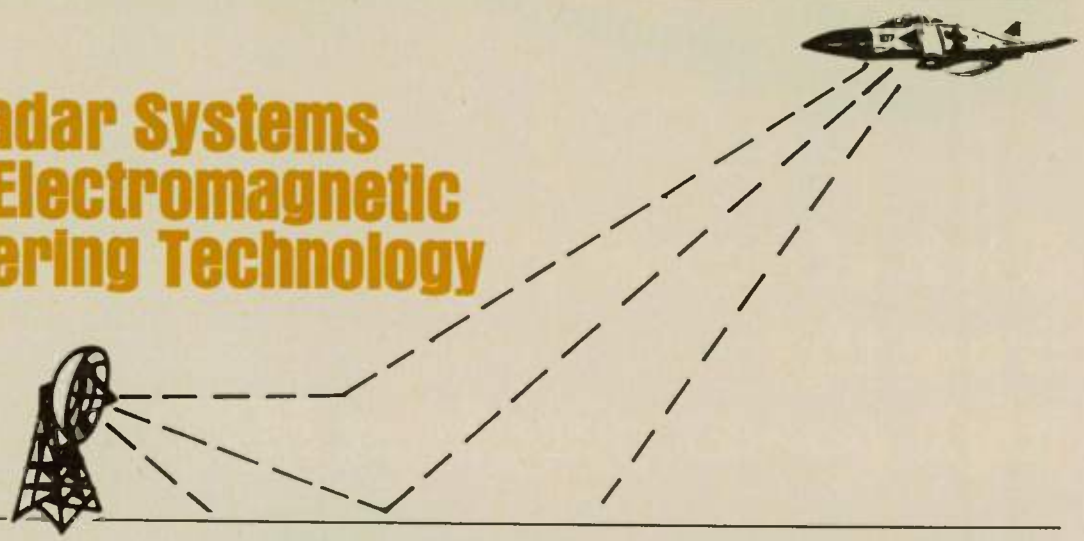
CIRCLE 41 ON READER SERVICE CARD

Radar Systems And Electromagnetic Scattering Technology



JOHN MCILVENNA
Electromagnetic Sciences Division
Rome Air Development Center
Hanscom AFB, MA

The aim of Contract F19628-78-C-0198 with Ohio State University was to increase the range of electromagnetic scattering problems that can be solved with relatively simple and economical computation techniques, the solutions to be used in the area of radar target signature identification. The approach was to combine separate techniques in hybrid schemes and to use these in the solution of certain diffraction problems for which the analytical diffraction coefficients are unknown. The ultimate goal was the determination of scattering over a large range of frequencies from targets of arbitrary shape. The major limitation of one such hybrid technique, often used in solving three-dimensional scattering problems (Geometrical Theory of Diffraction — Method of Moments), is the requirement of *a priori* knowledge of the current in the GTD region. A hybrid technique, developed under the contract, combines a moment method current and an asymptotic current. This technique does not require knowledge of the current away from the moment method region and seems well suited for handling arbitrarily shaped bodies. A third hybrid technique calculates the scattering from a body vertex by employing an exact eigenfunction solution only in a small region close to the vertex tip. Outside this region, the Uniform Theory of Diffraction is applied. The technique is computationally efficient and a significant improvement over the presently used techniques for vertex scattering.

The purpose of Contract F19628-78-C-0211 with Applied Science Associates, Inc., was to develop better theoretical models for terrain scattering by extending existing rough surface scattering models to include large scale surface tilt, finite conductivity, shadowing, and the effects of low grazing angles. Both backscatter (clutter) and forward scatter (multipath) were investigated to provide more accurate determination of the position of low flying aircraft and also to improve tactical radar system performance in a cost effective fashion. The contract has developed expressions for the normalized radar cross section of a lossy dielectric surface with two scales of roughness. The model handles all incident and scattering angles, including shadowing effects, and uses non-Guassian surface height statistics. This model will be incorporated into an in-house developed computer program which will use digitized terrain maps, weather data and visual observation of terrain class as input parameters. The goal is the real time prediction of complete clutter and multipath characteristics for any ground based tactical radar.

Contract F19628-79-C-0166 with Mark Resources Inc., seeks to optimize the apportionment of a tactical radar's operating parameters, such as scan time, transmitter power and antenna beam shape. This requires determining the effect of changes in environmental parameters and the uncertainty in the measurement of these environmental parameters upon the (S/N) ratio and the (S/C) ratio at the threshold detector in each resolution cell. The potential payoff from environmentally adapting radar parameters is great. The radar can be designed to meet average rather than worst case conditions and large reductions in search dwell time and/or operating wavelength can result. The baseline system that is modeled is typical of modern ground based mobile tactical radars operating in a European type environment. It performs wide area surveillance for aircraft targets as well as multiple target tracking in a heavy clutter environment of terrain (typical of dense forests), rain (typical of heavy storms) and chaff in an amount that could be carried by a large transport aircraft. Based on a recognition that short range includes ground clutter and requires rapid track updates, while long range does not, a schedule of wave forms that handles all ranges can be developed that is much more efficient than the use of a single waveform. Using as baseline mobile systems, a mechanically rotating antenna with electronic elevation scanning, and a phased array with electronic scan in both azimuth and elevation, two new design concepts have been developed. One involves the simultaneous use of multiple carrier frequencies while the other involves waveforms where blind range as well as blind speed zones must be resolved. The design of a real time adaptive radar including automatic algorithms for determining the appropriate PRF schedule have been developed and calculations of performance improvements associated with the use of such a system are available.

Contract F19628-78-C-0228 with Sperry Gyroscope set out to design a base-line surveillance radar for optimal, automatic, unattended operation in an adverse arctic environment. The resulting radar design featured the capability to adaptively manage the space-time-energy-hardware resources via a set of system of algorithms. These provided for (1) adaptable search/track burst-to-burst frequency, prf, pulsewidth and waveform selection to maximize detectability and accuracy of parameter estimates, (2) resolution of ambiguities, (3) suppresion of clutter and (4) adaptable track logic associated with window size, initiation, maintenance, and termination. The radar environment included effects such as severe surface, weather and bird clutter, terrain masking, non-optimal siting, low altitude penetration corridors, high reflectivity icebergs and multi-lobing. The final system design, optimized for an adverse arctic environment, included an azimuth monopulse channel in the antenna, receiver and signal processor, and the replacement of a vector hard limiter in the baseline design with a linear limiter and a sliding window CFAR threshold. It used an L-band, cylindrical array antenna 24 feet in diameter and 12 feet high, mounted on a high tower and capped by a protective dome. The system was capable of detecting a 0.5 square meter target at 30 nmi. The waveform and initiate track criterion provides a 0.95 probability of track initiation within sixteen seconds of penetrating the detection contour. Track-while-scan is provided on a minimum of 100 targets. ■

[From page 44] **ARRAY RADARS**

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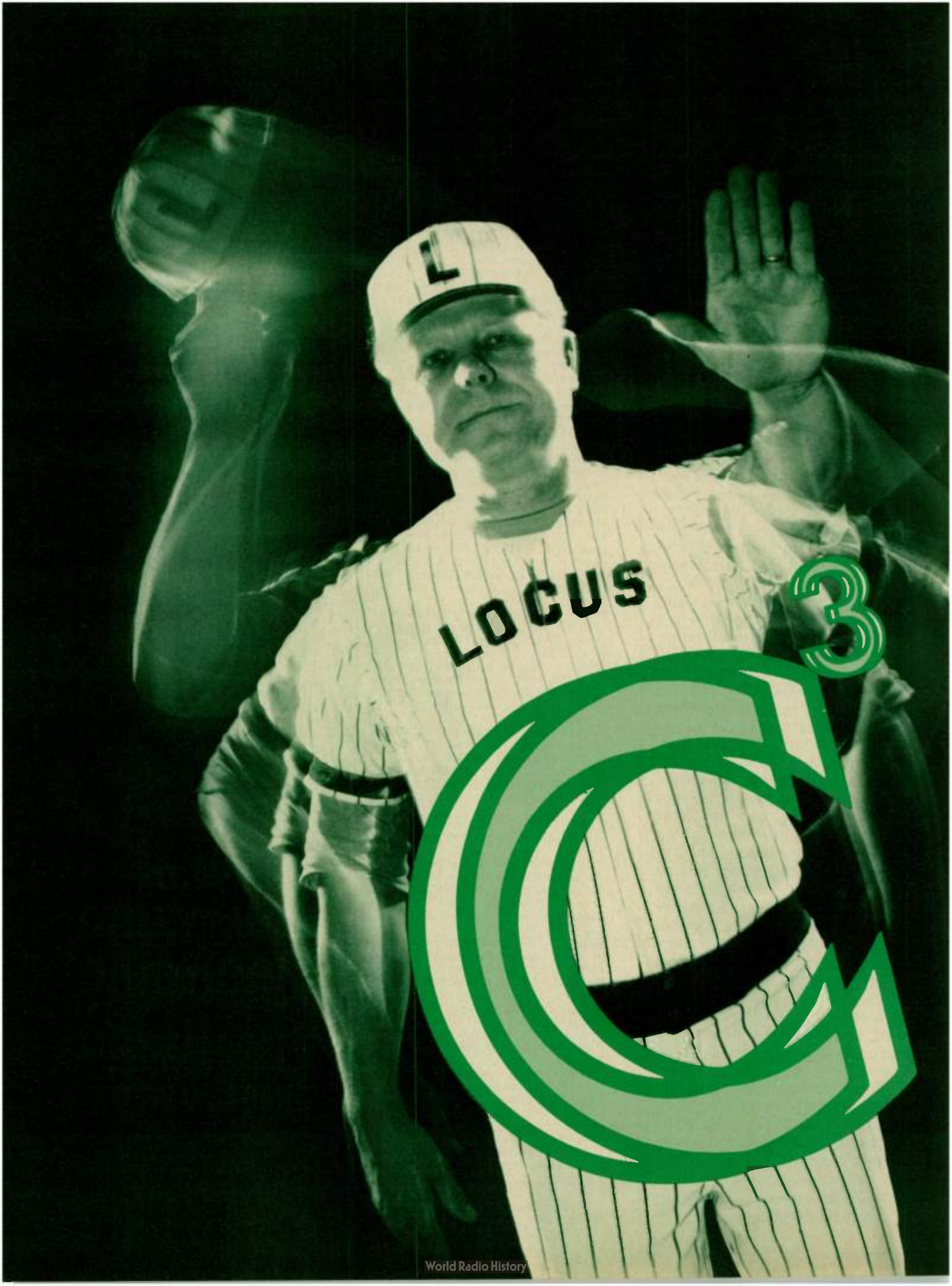
[Continued on page 114]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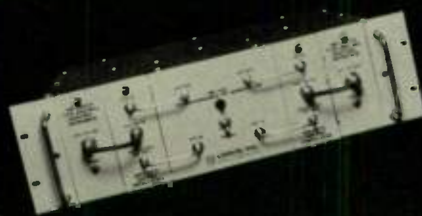

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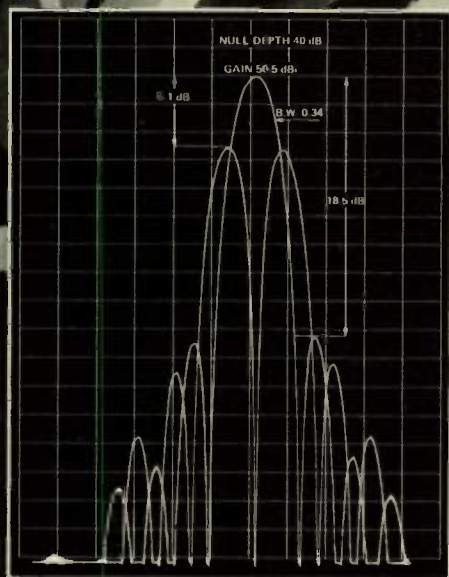

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	−3dB Beamwidth (Nominal)	Side Lobe Level (Nominal)	Net Gain (Minimum)	Null Depth (Minimum)	Insertion Loss (Minimum)
140GHz 24″ Diameter	0.25°	−18dB	56.5dBi	−28dB	2.5dB
94GHz 18″ Diameter	0.6°	−18dB	46.0dBi	−30dB	2.5dB
70GHz 18″ Diameter	0.65°	−18dB	46.0dBi	−30dB	1.5dB
53GHz 36″ Diameter	0.45°	−18dB	50.0dBi	−35dB	1.5dB
35GHz 96″ Diameter	0.25°	−18dB	54.0dBi	−35dB	1.0dB
9GHz 120″ Diameter	0.8°	−18dB	46.0dBi	−35dB	0.5dB

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Coherent 95 GHz Radar Achieves 40 dB Sub-Clutter Visibility

DAVID N. MCQUIDDY, JR.
Texas Instruments Incorporated
Dallas, TX

INTRODUCTION

Several promising military applications for radar systems operating at millimeter-wave lengths involve short-range targeting operations (detection, tracking, classification, etc.) in moderately adverse weather and under battlefield conditions (smoke, dust, and other obscurants). To be effective, the radar must often separate desired targets and target features that are in motion from much larger stationary objects. Coherent operation involving doppler processing is used to distinguish motions and vibrations of the radar's intended targets from the relatively stationary clutter returns.

This paper describes an approach for achieving the signal purity (phase or frequency noise) required for a 95 GHz radar to distinguish moving target returns that are 40 dB below the stationary clutter return. An instrumentation radar system was built, and tests have demonstrated the system will provide the desired sub-clutter visibility at doppler frequencies as low as 10 Hz. As a reference, a target moving 0.035 mph will generate a 10 Hz doppler frequency.

SYSTEM OVERVIEW

The application of millimeter-wave technology to electronic systems may provide effective solutions to a variety of military problems involving short-range targeting operations in clear to moderately adverse weather conditions. Millimeter-wave solutions can be implemented in the form of physically small equipment having high-resolution performance in range and angle with adequate doppler sensitivity.

Examples of military applications in which millimeter-wave technology provides attractive system alternatives include low-level interdiction strike missions in fixed or rotary-wing aircraft against ground armored vehicles. In the east European scenario, such missions are typically characterized by short time lines that support a requirement for automatic target acquisition and handoff for weapon delivery. Missile terminal-homing applications are another example. In this case, millimeter-wave technology may be used effectively in resolving high-value fixed targets or in identifying the unique signatures of selected tactical targets to enhance antiradiation homing by defeating the shutdown tactic. Finally, but of no less significance, is the area of ground-to-ground antiarmor operations that may include both fire control and battlefield IFF. In this area, millimeter-wave technology has demonstrated the potential of being an effective compliment to electro-optical (EO) systems in the adverse conditions of weather and battlefield-induced contaminants.

Within the battlefield scenario that is germane to millimeter-wave

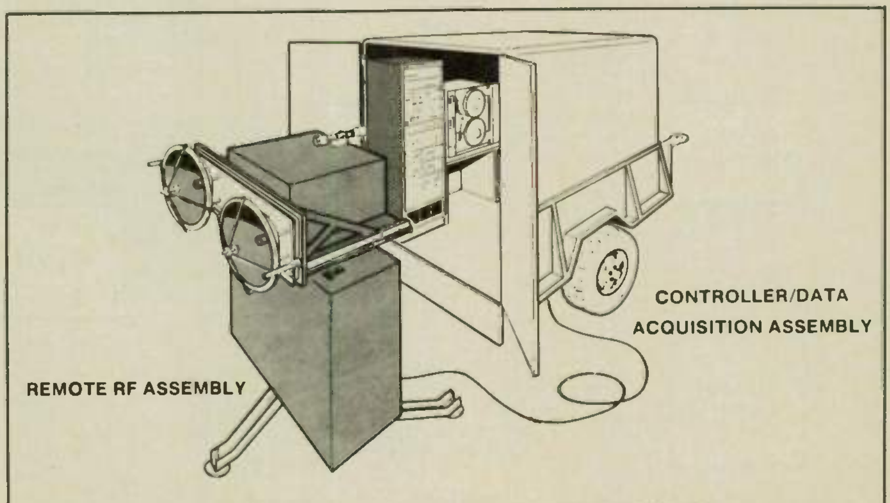

Fig. 1 Transportable 95 GHz instrumentation radar system.

* Paper first presented at "35th Annual Frequency Control Symposium," Philadelphia PA, sponsored by U.S. Army ERADCOM 27-29 May 1981.

applications, the high resolution and doppler sensitivity achievable at millimeter wavelengths may permit the integration of target acquisition and first-order classification techniques (e.g., target type through doppler signatures resulting from engine vibration and tread motion) in systems that are less complex, more readily integrable into a platform and, hence, potentially less costly than longer wavelength system alternatives.

Table I. KEY SYSTEM PARAMETERS.

PARAMETER	PERFORMANCE
Frequency	94.8 ± 0.15 GHz
Peak Transmit Power	0.5 W Minimum
Antenna Gain	9 inch: 42 dB Minimum 18 inch: 48 dB Minimum
Receiver Noise Figure	11.5 dB Maximum
Pulsewidths	25, 50, 100 nS 5 ns (Compressed from 100 nS)
System Loss	7.0 dB
Pulse Repetition Frequency (PRF)	
Transmit	40 kHz
Receive	40 kHz/(N+1) N = 0 TO 31

SYSTEM DESCRIPTION

The 95 GHz instrumentation system consists of a solid-state transmitter and receiver front-end, a digital data acquisition system, and a mini-computer data evaluator. The program objective was to develop a millimeter-wave measurement system that could be used in a variety of field conditions; hence, flexibility and portability were key system implementation considerations. Figure 1 is an artist's conception of the system as it would be deployed in field operations. The RF assembly was constructed as a remote sensing unit coupled to the system controller/data acquisition system via an IF link. Two men can transport the remote assembly to a location providing the optimum field of view. The unit can be trained in both elevation and azimuthal planes. A boresight telescope allows correct target alignment. An elevation or depression angle reading is provided. The combination is a rugged, self-contained, field-portable unit that can be readily deployed to acquire key millimeter-wave data under a variety of conditions that are not well suited to laboratory measurements. In field operations, the instrumentation system provides radar coherent processing interval (CPI) data records for subsequent off-line processing. System operation is computer monitored, thus permitting operator selection of a variety of radar parameters and system operating conditions that meet the needs of the planned experiment.

Figure 2 shows the instrumentation system in simplified block diagram form. The system exciter, transmitter, receiver, antennas and polarization control units are located in the remote RF assembly. The remaining units are located in the van. In general, the system gathers radar data reflected from the environment within its fixed beamwidth, provides syn-

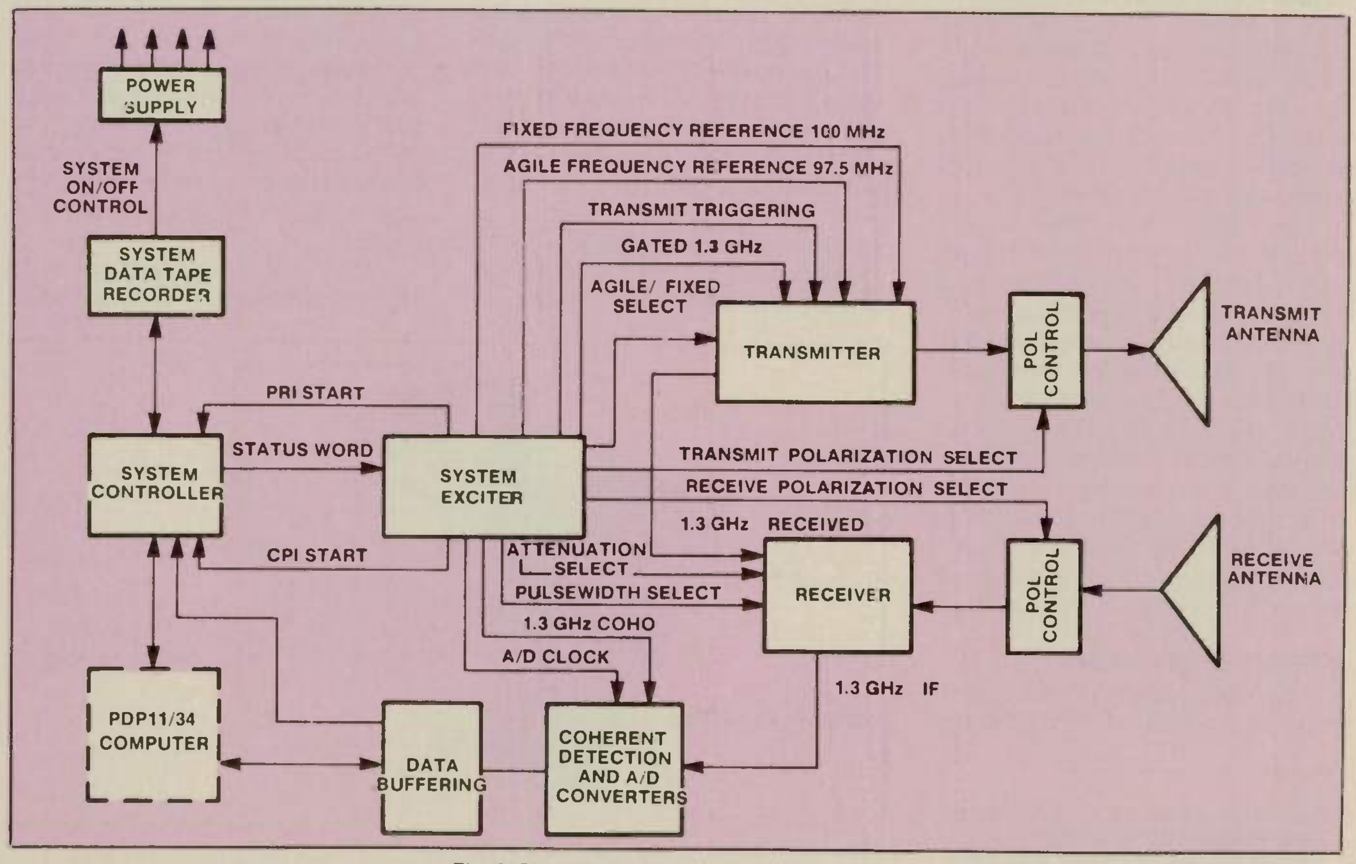

Fig. 2 Simplified system block diagram.

Table II. MODES OF OPERATION.

- **Pulse-to-Pulse Polarization Diversity**
  - **— RHC, LHC, Transmit RHC-Receive LHC and vice versa**
  - **— H, V, Transmit H-Receive V and vice versa**
- **Pulse-to-Pulse frequency agility**
  - **— 100 ns pulsewidth only**
  - **— 16 Discrete frequencies in 144 MHz bandwidth**
- **Coherent I & Q channels**
  - **— 4-bit A/D converters**
  - **— 200 MHz clock rate**
- **Coherent processing interval/memory store**
  - **— 64 contiguous range cells**
  - **— 1024 pulse repetition intervals (receive PRF)**

chronous detection (I & Q), formats the incoming time series data according to pulse repetition interval (PRI), and records the data on a CPI basis. Limited data analysis can be performed near real time by the mini-computer prior to passing the data on to the digital tape recorder.

A list of key system parameters is provided in Table 1. The system operates at 94.8 GHz and provides an instantaneous bandwidth of 300 MHz. The transmitter provides a minimum power of 0.5 W to the antenna input. Two antenna sizes can be used and are readily changed in the field. The 18 inch diameter antenna provides a minimum of 48 dB gain with a 0.49 degree beamwidth while the 9 inch diameter antenna provides 42dB gain with a 0.98 degree beamwidth. The cassegrain antennas (two of each size) are provided with boresight aligned telescopes. Four pulsewidth values can be selected. The shortest, 5 ns, is generated by compressing a 100 ns pulse that is nonlinear frequency chirped over a 200 MHz bandwidth. The 5 ns compressed pulse provides the system with a minimum resolvable range less than 1.0 meter. A constant transmit pulse repetition frequency (PRF) is used, and selectable receive PRF is achieved via programmed receiver blanking. Thirty-two receive PRF values are provided between 1.25 and 40 kHz.

The system operates bistatically (i.e. uses separate antennas for transmit and receive) and provides both coherent and noncoherent modes of operation. The system modes of operation are provided in Table II. The polarization diversity feature is implemented by electrically switching a waveguide ferrite polarizer on a pulse-to-pulse basis. The polarizer operates in a cylindrical waveguide mode and is followed by a fixed quarter-wave plate so that linear (H and V) and circular (RHC and LHC) polarizations can be attained. A switchable polarizer is contained in both the transmitter and receiver channels, allowing the transmission of one polarization state and the reception of the orthogonal state (linear or circular). Either fixed-frequency or pulse-to-pulse frequency-agile operation over a 144 MHz bandwidth may be selected with coherent operation provided in all fixed-frequency modes.

The function of the data acquisition system (DAS) is to preprocess the data collected by the remote RF assembly and record it in a format suitable for off-line processing. The initial DAS operation is synchronous detection of the receiver IF signal using the 1.3 GHz COHO reference to produce in-phase and quadrature video channels. Four high-speed 100 MHz A/D converters quantize the video data to a 4-bit level. Two A/D converters per channel are time phased to provide an I and Q channel instantaneous throughput rate at 200 MHz each. Data buffering and storage are provided by high-speed ECL memory that collects the incoming data on a PRI basis. Data reformation is accomplished by MOS bulk storage that accumulates range data from 1024 PRI's to form a coherent processing interval.

The 4-bit quantization level currently used in the DAS was a component availability limitation at the time of implementation and not a design constraint. All digital hardware in the DAS, exclusive of the A/D converters, has been designed to accept 8-bit data when 8-bit A/D converters with the high sampling rates used here become available.

A photograph of the instrumentation system hardware is shown in Figure 3. From left to right, the system components are the digital tape recorder (HP7970E), the data acquisition system and control panel, and the remote RF assembly equipped with 18 inch diameter antennas.

Recent tests performed with the system using a 1 m^2 trihedral reflector have verified that a single pulse signal-to-noise ratio of 9dB can be achieved at 1 km with the 18 inch diameter antennas. The measurements were made in

Fig. 3 System hardware.

clear weather. The transmitter power was measured and found to be 1 W. Assuming an atmospheric path loss of 0.72 dB/km, the actual system loss calculates to be 3dB with the receiver noise figure assumed to be 11.5 dB. Applying the full processing gain of a 1024-point CPI in a clutter free environment, the 1 m^2 reflector should result in a 11 dB signal-to-noise ratio when located at 5 km range.

SPECTRAL PURITY REQUIREMENTS

The system operating parameters have been selected to meet a wide variety of data collection needs. From an operational point of view, the PRF is selected to contain the doppler spectrum of the target of interest for the particular experiment being conducted. The data will be collected and the frequency spectrum from -PRF/2 to PRF/2 (or zero to PRF) will be filtered with digital transforms to characterize the target. Since the system must provide the capability to detect small moving targets in the presence (same range cell) of large stationary reflectors (discrete or point clutter), the combined transmitter and receiver local oscillator signal phase noise spectral density is constrained. The assumption is made that phase noise and not amplitude noise will be the primary mechanism degrading moving target detection. System tests have proven this assumption valid. The term sub-clutter visibility can be defined as the ratio of the large stationary reflector's backscatter cross-section to the backscatter cross-section of the smallest moving target that can be detected. The system employs a range-gated sampling process to provide range resolution. The sampling process folds the phase noise of the transmitted signal into the same spectral interval occupied by the target return. The folding process degrades sub-clutter visibility and provides an additional constraint on the system signal phase noise.

Since millimeter-wave radars provide adequate detection performance at short ranges only due to atmospheric loss, the short radar transit time can be used to advantage. The advantage results if a single master oscillator is used to generate both the transmit and the receiver local oscillator signals. Correlation of the transmitted signal, delayed by the propagation time to the target and back, with the local oscillator reference signal will suppress the phase noise near the carrier. Phase noise far removed from the carrier is degraded by 3 dB. The corner frequency occurs when:

$f_{corner} = 1/(4 T_d)$

where T_d is the propagation time to the target and back. Below the corner frequency phase noise is suppressed by 20 dB for each decade decrease in frequency. At the corner frequency the phase noise is 3 dB greater than the phase noise of the stable source. At two times the corner frequency, the phase noise has increased an additional 3 dB; however, the average degradation is only 3 dB when considering the spectrum out to the receiver IF bandwidth.

The system carrier-to-single sideband phase noise ratio (C/SSB) can be defined in terms of the factors discussed above. For frequencies out to the PRF/2, the ratio is set by the value of the sub-clutter visibility that is desired. Since it is most common to express the phase noise in terms of a per unit bandwidth, the doppler filter bandwidth (doppler resolution cell) must be multiplied times the sub-clutter visibility value to obtain the normalized C/SSB. If it is desired that the target signal appearing in a particular doppler resolution cell have a specified ratio to the noise in the cell, then this ratio (SNR) must be multiplied times the normalized C/SSB to obtain the improved C/SSB. Two other factors must also be considered. First, since range-gated sampling will raise the noise level in the frequency interval under consideration, it must be taken into account. This can be accomplished by requiring that the sum of the folded noise (normalized) contribute at the same level as the noise in the frequency interval under consideration. Multiplying the improved C/SSB by two then takes the folding into account. The second factor is the correlation effect associated with using a single stable source. The correlation factor can be expressed as:

$CF = 2(1 - \cos(2\pi f_{offset} T_d))$ for $f_{offset} < f_{corner}$

where f offset is the frequency offset from the carrier, and

$CF = 2$ for $f_{offset} < f_{corner}$

For the purpose of illustration, a particular set of system parameters have been selected. The system phase noise specification will be derived for this example and system data collected under the selected parameters will be presented later in the paper. A large stationary target is present at a range of 2.46 km. Assume that a small moving target 40 dB smaller in cross-section is present in the same range cell. Target motion is such that a 5 Hz doppler resolution cell is appropriate. The system will be operated at a PRF of 2.5 kHz and a 512-point discrete transform will be performed to produce the filtering function. The 100 ns pulsewidth sets the receiver IF bandwidth at 12 MHz. Figure 4 shows one phase noise spectral density function that will meet the requirements for moving target detection at a SNR of 1.

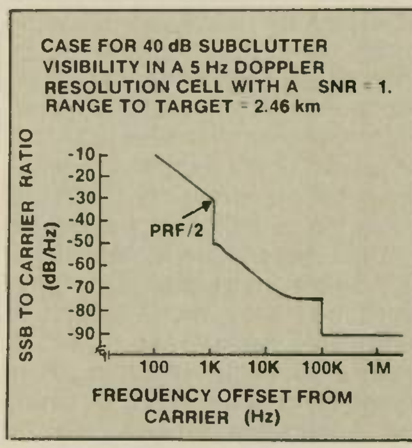

Fig. 4 System phase noise specification.

Figure 4 represents the system phase noise specification. A family of these curves can be generated for various PRF's and a worst case specification derived from them; however, the end result will not differ significantly from Figure 4.

SYSTEM SIGNAL GENERATION

The signal generation approach

that was selected to meet the system spectral purity requirements is shown in the block diagram in Figure 5. The primary signal path is highlighted with bold interconnecting lines. All signals including the system timing are derived from a single 100 MHz stable source. The 100 MHz signal is used as a reference to phaselock a transistor oscillator operating at 7.8 GHz. This oscillator was designed to achieve low noise performance for communications applications. The phaselock bandwidth is approximately 100 kHz. At frequencies greater than the loop bandwidth, the system noise is determined by the free running characteristics of the transistor oscillator. The multiplied phase noise of the 100 MHz source sets the system performance for frequencies within the loop bandwidth. The 7.8 GHz signal drives a harmonic mixer to mix its twelfth harmonic with the output of a Gunn oscillator operating at 93.5 GHz.

The 100 MHz difference signal from the mixer is compared in a phase detector with the signal from the 100 MHz stable source. The output of the phase detector is used to phaselock the Gunn oscillator. The loop bandwidth of this phaselock oscillator is slightly greater than 10 MHz. The output power of the phase locked Gunn oscillator is divided equally between two paths. One path provides the local oscillator signal for the receiver balanced mixer, the second path drives an injection locked IMPATT oscillator. This CW oscillator is used to amplify the small +5dBm signal from the Gunn oscillator to a power level of +17dBm. The injection locked bandwidth of the CW IMPATT oscillator is greater than 700 MHz. The output of this oscillator drives a single-sideband upconverter. The second input to the upconverter is a time gated 1.3 GHz signal that is obtained by multiplying a sample of the 100 MHz stable source by thirteen and then gating it with a diode RF switch. The gate time (pulsewidth) is determined by the selected system pulsewidth. The pulsed 94.8 GHz upconverter output signal is applied to a two-stage pulsed IMPATT injection locked oscillator assembly. The two oscillators are gated on for 140 ns. The pulsed injection signal is present during the latter 100, 50, or 25 ns of the 140 ns pulse depending on the system pulsewidth selection. The free running pulsed oscillator spectrum falls outside the matched receiver bandwidth and doesn't result in detectable target time sidelobes. The two-stage pulsed IMPATT oscillators amplify the upconverter output signal +5dBm up to the +30 dBm level at the transmitter antenna input port.

The 5 ns pulsewidth is achieved via nonlinear FM pulse compression techniques using surface acoustic wave (SAW) devices. This approach was selected over conventional linear FM pulse compression because it permits better control of time sidelobe structure and mismatch loss at the small time-bandwidth products required here (TB=20). The expanded transmit pulse is obtained by impulsing the SAW device with a 5 ns sample of the 1.3 GHz signal. The impulsed SAW device generates a 100 ns nonlinearly down-chirped signal that is phase referenced to the system stable source and covers a 200 MHz bandwidth. The

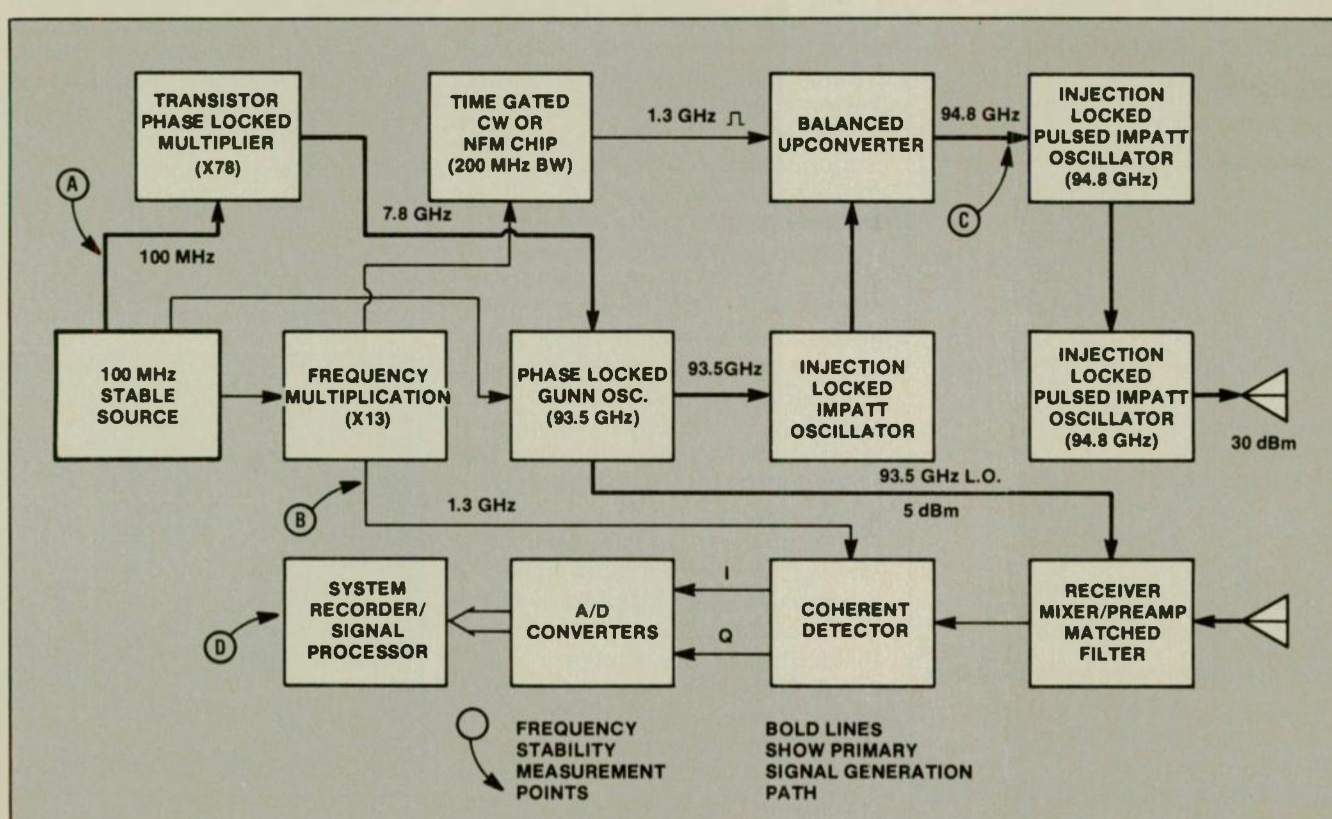

Fig. 5 System signal generation technique.

[Continued on page 70]

down-chirp approach was selected to provide a more natural match to the thermally induced free running chirp to be expected from the pulsed IMPATT oscillators. A matching SAW compressor is located in the receiver and compresses the 100 ns transmitted signal to a value near 5 ns having time sidelobes down more than 23 dB relative to the peak of the main lobe.

The signal flow for the frequency agile mode is not shown. Briefly, a stepping synthesizer, which covers the range of 97.34 to 97.66 MHz serves as a reference (in place of the 100 MHz signal) to the Gunn phaselock assembly. A times eighty multiplied version of the synthesizer output replaces the 7.8 GHz signal to the phase locked Gunn oscillator. This produces an output signal that steps in 9 MHz steps over a 144 MHz band at 93.5 GHz.

SYSTEM STABILITY MEASUREMENTS

Stable Source Performance

The phase noise performance of the 100 MHz stable source was measured by constructing two "nearly" identical sources. One source was designed with an electronic tuning capability. The tuning mechanism was lightly coupled to prevent the introduction of extraneous noise. The two oscillators were locked together with a narrow bandwidth phaselock loop. The loop bandwidth was approximately 10 Hz. The phase detector output was amplified by a low noise FET video amplifier and applied to a low frequency spectrum analyzer (HP3585A). The spectrum of the phase demodulated signal was recorded from 20 Hz (twice the loop bandwidth) out to 100 kHz.

The phase noise spectral density was then calculated assuming equal noise contributions from both sources. The results of this measurement are shown in Figure 6. Also shown in the figure is the system specification translated to 100 MHz. The translation results from the fact that frequency multiplication (100 MHz to 94.8 GHz) will increase the phase noise by the multiplication factor squared.

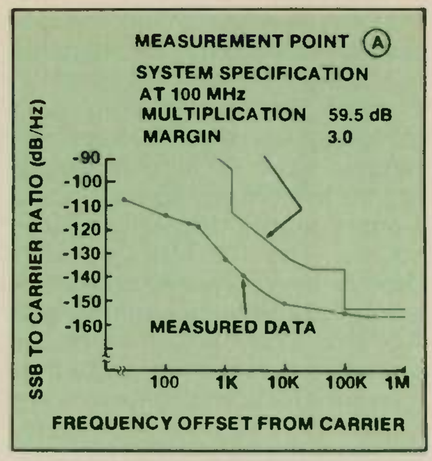

Fig. 6 Measured 100 MHz stable source phase noise.

Also, 3 dB has been added to account for any non-ideal multiplication effects. The translation amounts to 62.5dB. It is seen that adequate performance has been achieved.

Phase Locked Gunn Oscillator Performance

The phase noise was measured at 93.5 GHz by comparison with a second source that was derived completely by multiplication from a "clean" 97 MHz oscillator. The multiplied source was mixed with a sample of the system phase locked Gunn oscillator injection locked IMPATT oscillator combination to produce a difference frequency near 300 MHz. The difference signal was applied to a spectrum analyzer with a 1 KHz bandwidth. The results of this measurement adjusted to a 1 Hz bandwidth is shown in Figure 7. No secondary tests were performed to verify which of the two

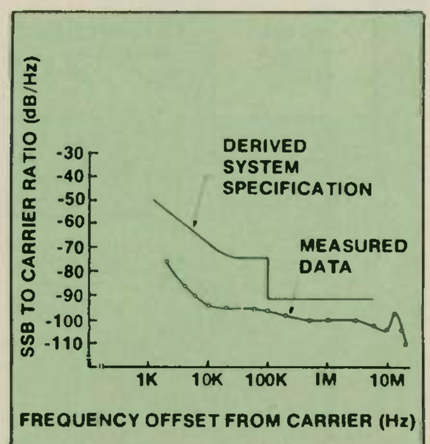

Fig. 7 Gunn PLO phase noise at 93.5 GHz.

sources dominate; however, the two noise plateaus in the vicinity of 100 kHz and 10 MHz seem to indicate the phase locked oscillator dominates. The system specification is repeated in this figure for reference.

System Sub-Clutter Visibility

During initial system operating tests, the instrumentation hardware was set up in an elevated test station overlooking the TI antenna range. The system was used to record radar returns from both fixed and moving ground targets. A number of runs were conducted using a variety of radar parameters. One experiment in particular was devised to evaluate sub-clutter visibility performance. A large stationary target was located at 2.46 km and illuminated with a 100 ns pulse train at a 2.5 kHz repetition rate.

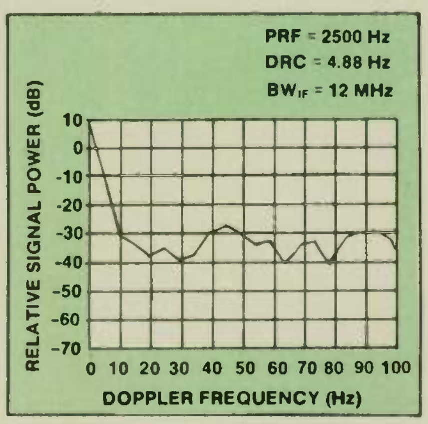

Fig. 8 System performance data for a large stationary target at 2.46 km.

A full CPI was recorded and processed off-line. The off-line processing consisted of performing a complex discrete Fourier-transform using 512 data points from the CPI. The result of the off-line processing is shown graphically in Figure 8. An expanded frequency scale was selected that shows the doppler performance near the carrier. The stationary target is seen in the zero doppler cell at a relative amplitude of +10dB. The power in the second doppler cell at 9.76 Hz has a relative amplitude of -31dB, and demonstrates a subclutter visibility greater than 40dB. Although not shown, the doppler spectrum was examined out to 2.5 kHz and most cells displayed greater than 40dB visibility. It should be commented here that the visibility limit could be a result of system thermal noise and not phase noise. The thermal noise floor is set by the 4-bit A/D drive requirements and the processing gain. In other words, the system phase noise performance supports the 40dB subclutter requirement at the minimum and could be significantly better.

CONCLUSION

A millimeter-wave instrumentation radar system has been described. The principal components of the system are a 95 GHz transceiver, a data acquisition system, and a digital recorder. The combined elements of the system provide a high degree of flexibility in data collection and signal processing. Radar returns may be processed coherently and mapped into a variety of range, amplitude, and doppler combinations.

Specifically, phase locked Gunn oscillators have demonstrated the degree of spectral purity required to support the coherent signal processing of radar returns at millimeter-wave frequencies. A cascade chain of pulsed, injection locked, IMPATT oscillators have provided transmitter signal amplification at 95 GHz while preserving signal coherence. A phase controlled waveform using a SAW pulse compression expander has been amplified by pulsed, injection locked, IMPATT oscillators. The target return has been compressed in a matched filter providing a range resolution better than 1m. The compressed waveform has been coherently processed.

In general, it can be concluded that a solid-state millimeter-wave radar can be realized in a form supportive of high-resolution range and doppler performance.

David N. McQuiddy, Jr., *is manager of the Microwave Laboratory in the Equipment Group at Texas Instruments Incorporated. He is responsible for the design, development, and product engineering support for a broad range of microwave and millimeter-wave components and subsystems.*

Dr. McQuiddy received the Bachelor of Electrical Engineering degree from Vanderbilt University and the MS and Ph. D. degrees in Electrical Engineering from the University of Alabama.

He joined the Equipment Group at Texas Instruments in 1968. He is Chairman of the 1982 International Microwave Symposium Steering Committee. ■

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3200-3	16.5/0.1	0.1, 0.2, 0.4, 0.8, 1, 2, 4 and 8
3200-4	150/10	10, 20, 20, 20, 20, 20, 20, and 20
3201-1	31/1	1, 2, 4, 8, and 16
3201-2	120/10	10, 20, 30, and 60
3201-3	12/1	1, 2, 3, and 6
3201-4	1.2/0.1	0.1, 0.2, 0.3, and 0.6
3202-1	0.5/0.5	0.5

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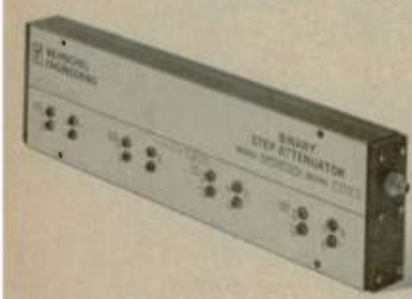

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  - — 0 to 90 dB/10 dB steps
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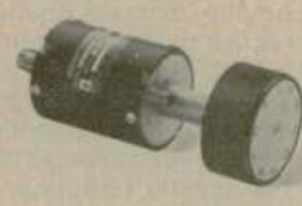

- dc — 18 GHz, covered in four models
- Five Standard Attenuation Ranges/Steps — up to 0-99 dB/1 dB steps
- VSWR — 1.75 Maximum
- WPM Connectors — mate with SMA per MIL-C-39012

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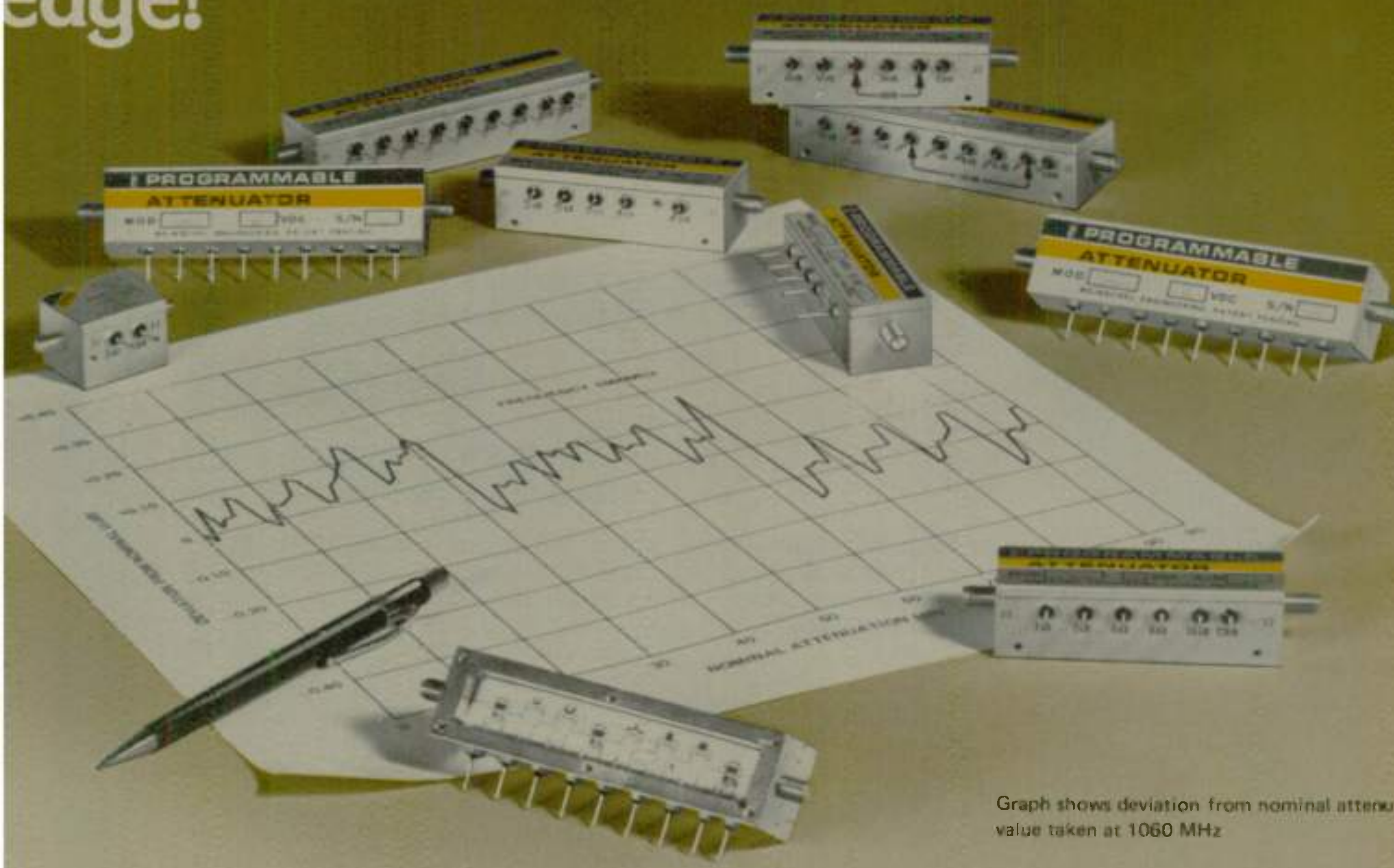
edge!



Graph shows deviation from nominal attenuation value taken at 1060 MHz

C30

Miniature Step Attenuator Model 115A

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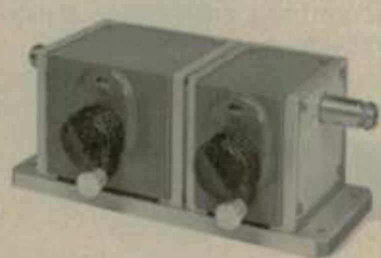

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Axial Ratio Measurements of Dual Circularly Polarized Antennas

WARREN L. STUTZMAN and WILLIAM P. OVERSTREET
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INTRODUCTION

Ideally circularly polarized antennas have unity (0 dB) axial ratios. In reality, good quality circularly polarized antennas have small axial ratios. For applications of dual, circularly polarized antennas such as on frequency reuse earth-space communications links, it is important to know the axial ratio value very accurately, for it will ultimately determine the isolation between channels. Axial ratio measurement is, however, difficult. For example, if a dual-polarized receiving antenna is aligned with a transmitting antenna the received signal in the cross-polarized channel arises from contributions from both the transmitting and receiving antenna; it is not obvious what the axial ratio values for the individual antennas involved are. In this paper two techniques are introduced for making this determination.

If either one or two dual circularly polarized receiving antennas and a circularly polarized transmitting antenna are available, the axial ratios of all can be determined. It is assumed that the receive antenna(s) can be rotated to locate the maximum and minimum isolation values. If two receiving antennas are available those maximum and minimum isolation values are all that is necessary to determine the axial ratios of the receive antennas and the transmitting antenna. If only one receive antenna is available but additionally the relative phase between the co- and cross-polarized ports of the receive and transmit antennas can be calculated. The theory behind these two methods is discussed first and then an example of some actual measurements is presented.

THE PROCEDURE WITH TWO RECEIVING ANTENNAS

With two receiving antennas the axial ratios can be determined from only isolation measurements. The polarization of a constant incoming wave is that of a distant transmitting antenna for a clear air, unobstructed propagation path. The polarization state of the incoming wave is characterized by ε_W and τ_W

$$\varepsilon_w = \cot^{-1}(AR_W) \qquad (1)$$

where AR_W is the axial ratio, the magnitude of which is the ratio of major axis to the minor axis of the polarization ellipse and sign of AR_W is plus for the left-hand sense of rotation and minus for right-hand sense. The tilt angle of the major axis of the incoming wave polarization ellipse relative to local horizontal is τ_W.

The two receive antennas (labeled 1 and 2) have cross-polarized channels with polarization states $(\varepsilon_{x1}, \tau_{x1})$ and $(\varepsilon_{x2}, \tau_{x2})$ where

$$\varepsilon_{x1} = \cot^{-1}(AR_{x1}) \qquad (2)$$
$$\varepsilon_{x2} = \cot^{-1}(AR_{x2})$$

and the tilt angles τ_{x1} and τ_{x2} are variable through rotation of the receive antennas. We define relative tilt angles as

$$\Delta\tau_{x1} = \tau_{x1} - \tau_W \qquad (3)$$
$$\Delta\tau_{x2} = \tau_{x2} - \tau_W$$

Dual polarized receiving antennas, such as a parabolic reflector with a dual circularly polarized feed, are usually such that the axial ratio magnitudes of the two channels are very close in value. Even if this is not true, deviations in the copolarized axial ratio magnitude $|AR_c|$ from that of the cross-polarized channel $|AR_x|$ for good quality antennas have very little impact on the subsequent development. Also, it is assumed that the tilt angle of the co- and cross-polarized receive antenna states are the same. These assumptions are

$$AR_x = -AR_c \text{ and } \Delta\tau_c = \Delta\tau_x = \Delta\tau \qquad (4)$$

for both antennas 1 and 2. The isolation between receive channels under these assumptions is [1]

$$I = \frac{A + B}{A - B} \qquad (5)$$

where

$$A = (AR^2_w + 1)(AR^2_c + 1) + (AR^2_w - 1)(AR^2_c + 1)\cos 2\Delta\tau$$

$$B = 4AR_wAR_c$$

With two dual-polarized receiving systems four measurements can be made, two for each antenna: when the tilt angle of the wave and cross-polarized receive antenna are aligned ($\Delta\tau = 0$ degrees, giving minimum isolation) and when orthogonal ($\Delta\tau = 90$ degrees, giving maximum isolation). The four isolation expressions are, from (5)

$$I_{1min} = I(AR_1, \Delta\tau_1 = 0) = \frac{C + D}{C - D} \quad (6)$$

where

$$C = (AR^2_w + 1)(AR^2_1 + 1) + (AR^2_w - 1)(AR^2_1 - 1)$$

$$D = 4AR_wAR_1$$

$$I_{2,min} = I(AR_2, \Delta\tau_2 = 0) = \frac{E + F}{E - F} \quad (7)$$

where

$$E = (AR^2_w + 1)(AR^2_2 + 1) + (AR^2_w - 1)(AR^2_2 - 1)$$

$$F = 4AR_wAR_2$$

$$I_{1max} = I(AR_1, \Delta\tau_1 = 90°) = \frac{G + H}{G - H} \quad (8)$$

where

$$G = (AR^2_w + 1)(AR^2_1 + 1) - (AR^2_w - 1)(AR^2_1 - 1)$$

$$H = 4AR_wAR_1$$

$$I_{2max} = I(AR_2, \Delta\tau_2 = 90°) = \frac{J + K}{J - K} \quad (9)$$

where

$$J = (AR^2_w + 1)(AR^2_2 + 1) - (AR^2_w - 1)(AR^2_2 - 1)$$

$$K = 4AR_wAR_2$$

The unknowns are AR_w, AR_1, and AR_2. By rotating the receive antennas to create the aligned ($\Delta\tau_1 = 0$ degrees, $\Delta\tau_2 = 0$ degrees) and orthogonal ($\Delta\tau_1 = 90$ degrees, $\Delta\tau_2 = 90$ degrees) conditions and measuring the channel isolations at each receiving antenna, the measured (known) quantities are $I_{1,min}$, $I_{1,max}$, $I_{2,min}$, and $I_{2,max}$. The unknowns AR_w and AR_1 can be found by solving (6) and (7) simultaneously. Unknowns AR_w and AR_2 are found by solving (8) and (9). The process for solving each pair is the same, so we outline the solution of the first pair. Because the sense of polarization is the same, the signs of AR_w, AR_1 and AR_2 are all the same. For simplicity we will henceforth assume this sign to be positive. We are really only solving for the magnitudes of the axial ratios anyway. Expanding (6) yields:

$$\frac{AR_wAR_1 + 1}{AR_wAR_1 - 1} = \sqrt{I_{1,min}} \quad (10)$$

Expanding (7) gives

$$\frac{AR_w + AR_1}{|AR_w - AR_1|} = \sqrt{I_{1,max}} \quad (11)$$

There are two cases of interest.

<u>CASE 1 $AR_w \geq AR_1$</u>

First we note

$$|AR_w - AR_1| = AR_w - AR_1 \quad (12)$$

Using this in (11) and solving for AR_1:

$$AR_1 = AR_w \frac{\sqrt{I_{1,max}} - 1}{\sqrt{I_{1,max}} + 1} \quad (13)$$

Substituting this into (10) and solving for AR_w: (14)

$$AR_w = \sqrt{\frac{(\sqrt{I_{1,min}} + 1)(\sqrt{I_{1,max}} + 1)}{(\sqrt{I_{1,min}} - 1)(\sqrt{I_{1,max}} - 1)}}$$

<u>CASE 2 $AR_w \leq AR_1$</u>

First note that

$$|AR_w - AR_1| = AR_1 - AR_w \quad (15)$$

Using this in (11) and rearranging,

$$AR_w = AR_1 \frac{\sqrt{I_{1,max}} - 1}{\sqrt{I_{1,min}} + 1} \quad (16)$$

Substituting this in (4.3-13) and solving for AR_1, (17)

$$AR_1 = \sqrt{\frac{(\sqrt{I_{1,min}} + 1)(\sqrt{I_{1,max}} + 1)}{(\sqrt{I_{1,min}} - 1)(\sqrt{I_{1,max}} - 1)}}$$

Note that (13) and (14) are the same as (16) and (17) with AR_w and AR_1 interchanged. This means unless we know *a priori* whether we had Case 1 ($AR_W \leq AR_1$) or Case 2 ($AR_w \leq AR_1$), we could only calculate the values of AR_w and AR_1 for one receiving station and would not know whether the values were interchanged. Of course, one usually would not know which case applied because AR_w and AR_1 are the unknown quantities. This problem can be resolved with a second set of measurements with antenna 2. The results are exactly the same as above for antenna 1 (e.g., Cases 1 and 2) but AR_1 is replaced by AR_2. Instead of reproducing this discussion for antenna 2 we present a summary of how the measurements/calculational scheme would proceed.

Procedure

a) Measure $I_{1,min}$ and $I_{1,max}$
b) Calculate

$$x = \sqrt{\frac{(\sqrt{I_{1,min}} + 1)(\sqrt{I_{1,max}} + 1)}{(\sqrt{I_{1,min}} - 1)(\sqrt{I_{1,max}} - 1)}} \quad (18)$$

$$y = x \frac{\sqrt{I_{1,max}} - 1}{\sqrt{I_{1,max}} + 1} \quad (19)$$

Then *either*

$$AR_w = x \quad \text{and} \quad AR_1 = y \quad (20)$$

or

$$AR_w = y \quad \text{and} \quad AR_1 = x \quad (21)$$

c) Measure $I_{2,min}$ and $I_{2,max}$
d) Calculate

$$u = \sqrt{\frac{(\sqrt{I_{2,min}} + 1)(\sqrt{I_{2,max}} + 1)}{(\sqrt{I_{2,min}} - 1)(\sqrt{I_{2,max}} - 1)}} \quad (22)$$

$$v = u \frac{\sqrt{I_{2,max}} - 1}{\sqrt{I_{2,max}} + 1} \quad (23)$$

Then *either*

$$AR_w = u \quad \text{and} \quad AR_2 = v \quad (24)$$

or

$$AR_w = v \quad \text{and} \quad AR_2 = u \quad (25)$$

e) Compare the results in b) and d) and choose solutions which yield identical AR_w values.

THE PROCEDURE WITH A SINGLE RECEIVING ANTENNA

The discussion in the preceding section indicates that the measurement of minimum isolation and maximum isolation with a single earth station antenna can be used to calculate the antenna and wave axial ratios, however, the results are ambiguous as to which axial ratio is that of the wave (or transmitter antenna) and which is that of the receiver

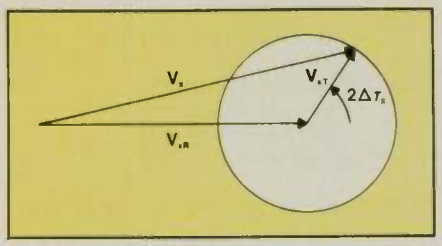

Fig. 1a Illustration of phasor addition as a function of rotation angle $\Delta\tau_X$.

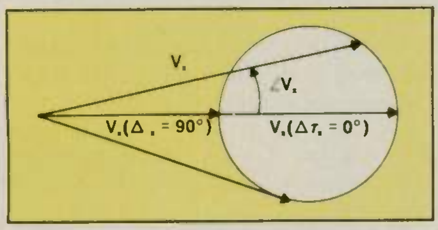

Fig. 1b Output voltage V_x for several rotation angles. Note that when $\Delta\tau_X$ is 0° and 90° the voltage is of the same phase.

antenna. If the phase of the receiver crosspolarized signal voltage output relative to the phase of the copolarized voltage is available, then the ambiguity of the axial ratios can be resolved. This can be explained with the aid of Figures 1 and 2 in which V_x is the receiver crosspolarized voltage. V_{xR} and V_{xT} are the contributions to V_x from the receiver and transmitter imperfections. In other words, the transmitter generates a wave polarization which can be decomposed into purely right and left hand sensed circular components; the small undesired component gives rise to the transmitter imperfection voltage V_{xT}. Similarly for the receiver, V_{xR} is a measure of how the receive antenna responds to an undesired polarization component. When $|V_x|$ is largest the isolation is minimum (worst) and the receiver imperfection voltage adds in phase to the transmitter imperfection voltage. The sum voltage is assigned a reference phase of 0°. As the antenna is rotated about its axis at an angle $\Delta\tau_x$, the phase difference between V_{xR} and V_{xT} increases from zero to $2\Delta\tau_x$. When $\Delta\tau_x = 90°$, V_{xR} and V_{xT} are 180° out of phase and combine to give maximum (best) isolation. In the case shown in Figure 1 the net phasor V_x has the same phase (0°) at maximum and minimum isolations. This is because the receiver imperfection (axial ratio) is greater than that of the transmit wave, $|V_{xR}| > |V_{xT}|$. If the reverse were the case, $|V_{xR}| < |V_{xT}|$, then the phase of V_x changes by 180° for a rotation between maximum and minimum isolations as seen in Figure 2.

By noting the phase change (or lack of phase change) in the cross-polarized signal voltage as the antenna is rotated from minimum to maximum isolation, one immediately knows which of the two calculated axial ratios is that of the receiver and which is that of the transmitter.

TABLE I

RESULTS OF ANTENNA FEED ROTATION TEST

(a) SIRIO Main Terminal (12-foot diameter reflector) — Antenna 1

Relative Feed Angle (degr.)		Isolation I(dB)	Phase (degr.)
5	min	17.9	0
15		17.7	10
20		18.2	10
40		20.2	17
60		24.0	40
80		33.8	60
90		35.0	20
95	max	35.0	0
100		34.5	-10
120		26.3	-50
140		21.5	-40
160		18.7	-25
180		17.8	0

(b) SIRIO Diversity Terminal (10-foot diameter reflector) — Antenna 2

Relative Feed Angle (degr.)		Isolation I(dB)
0		22.5
20		30.0
30	max	35.5
40		27.5
60		22.0
85		18.5
90		19.0
110		17.5
120	min	17.5
130		17.5
150		18.5
175		22.0

Procedure

a) Using one receiver antenna, measure

1) I_{max}, Φ_{max}

2) I_{min}, Φ_{min}

where Φ_{max} and Φ_{min} are the phases of the receiver voltage for maximum and minimum isolation conditions, respectively.

b) Calculate

$$x = \sqrt{\frac{(\sqrt{I_{min}} + 1)(\sqrt{I_{max}} + 1)}{(\sqrt{I_{min}} - 1)(\sqrt{I_{max}} - 1)}} \quad (26)$$

$$y = x \; \frac{\sqrt{I_{max}} - 1}{\sqrt{I_{max}} + 1} \quad (27)$$

which follows from (18) and (19).

c) If $\Phi_{min} = \Phi_{max}$, then:

$AR_R = x$

$AR_w = y$

or, if $\Phi_{min} = \Phi_{max} \pm 180$ degrees, then:

$AR_r = y$

$AR_w = x$

AN EXAMPLE

The 12-foot diameter reflector antenna at the Virginia Tech Satellite Tracking Station received the 11.7 GHz right-hand circularly polarized beacon from the CTS satellite from 1976 through 1979. In 1980 the dual circularly polarized antenna was converted to receive the 11.7 GHz beacon of the Italian satellite SIRIO. It was necessary then to perform in-service tests. In addition, a second portable terminal was acquired and located 7 km away to perform site diversity experiments.

The results of the feed rotation tests on both antennas are presented in Table 1[2]. The rotation angles shown are only relative. No phase data are available from the diversity terminal (antenna 2).

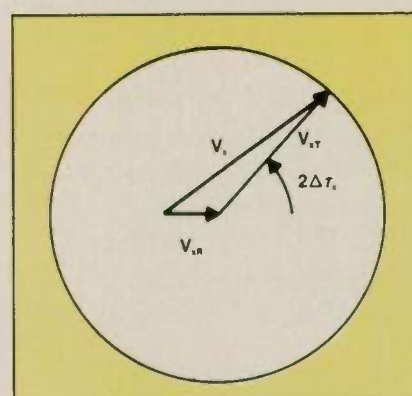

Fig. 2a Illustration of phasor addition as a function of rotation angle $\Delta\tau_X$.

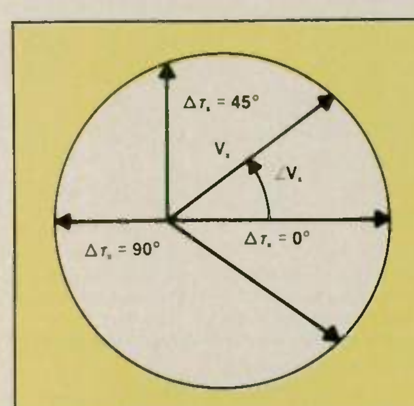

Fig. 2b Output voltage V_x for several rotation angles. Note that the voltage for $\Delta\tau_x = 90°$ is 180° out-of-phase with the voltage when $\Delta\tau_x = 0$.

[*Continued on page 78*]

[From page 77] **RATIO MEASUREMENTS**

Note that at both sites the minimum and maximum isolations are 90 degrees apart. This is the expected result and adds confidence to the measurement accuracy.

Notice in Table 1a that the phase is the same at the minimum and maximum isolation points. This absence of phase reversal indicates that $AR_R > AR_w$ and, for the main terminal,

$$I_{min} = 17.9 \text{ dB} = 61.66$$
$$I_{max} = 35.0 \text{ dB} = 3162.3$$

Then, from (26),

$$x = \sqrt{\frac{(\sqrt{61.66}+1)(\sqrt{3162.3}+1)}{(\sqrt{61.66}-1)(\sqrt{3162.3}-1)}} = 1.157$$

and, from (27)

$$y = x\,\frac{\sqrt{I_{max}}-1}{\sqrt{I_{max}}+1}$$

$$= (1.1157)(0.965) = 1.1165$$

Since $AR_R > AR_W$

$$AR_R = x = 1.157 = 1.27 \text{ dB}$$
$$AR_W = y = 1.1165 = 0.96 \text{ dB}$$

In this case the axial ratio of the receiver antenna is poorer than that of the spacecraft antenna. Therefore the receiver imperfection voltage $|V_{xR}|$ is greater than the transmitter imperfection voltage $|V_{xT}|$. These two voltages are plotted in Figure 1b for several different feed angles. Note that V_{xT} rotates as the antenna is rotated. The resultant of V_{xT} and V_{xR} gives the crosspolarized receiver voltage (magnitude and phase). Note that Figure 1 gives a result consistent with the measured data. Starting at the point of minimum isolation ($V_{xR} + V_{xT}$ maximum) the phase angle of the resultant is zero degrees. Then as the feed is rotated through its first 90 degrees, V_{xT} rotates 180 degrees and the resultant voltage phase increases then returns to zero. During the second 90 degrees of feed rotation the resultant goes to a negative angle then back to zero.

The diversity antenna results of Table 1b are very similar to those of the main site antenna, indicating that the axial ratios of the two antennas are very similar. Following the same procedure:

$$I_{min} = 17.5 \text{ dB} = 56.23$$
$$I_{max} = 35.5 \text{ dB} = 3548.13$$

and $x = 1.163 = 1.31$ dB
$y = 1.1246 = 1.02$ dB

So, for the diversity site

$$AR_R = 1.31 \text{ dB}$$
$$AR_W = 1.02 \text{ dB}$$

Note that $|AR_w| = 0.96$ dB determined by the main site agrees fairly well with that determined only by the diversity site, $|AR_W| = 1.02$ dB.

The two antenna method could be used also. Following that procedure we obtain

$x = 1.157$ $y = 1.1165$ (main, antenna 1)
$u = 1.163$ $v = 1.1246$ (diversity, antenna 2)

The values of y and v are nearly identical. Of course, they should be exactly equal, but measurement error is unavoidable. Averaging these two values gives

$$|AR_W| = \frac{1.1165 + 1.1246}{2} = 1.12 = 0.99 \text{dB}$$

then $|AR_1| = 1.157 = 1.27$ dB
$|AR_2| = 1.163 = 1.31$ dB

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2. Stuzman, W. L., W. P. Overstreet, C. W. Bostian, A. Tsolakis, and E. A. Manus, "Ice Depolarization on Satellite Radio Paths," *Final Report of INTELSAT Contract INTEL-123*, April 1981.

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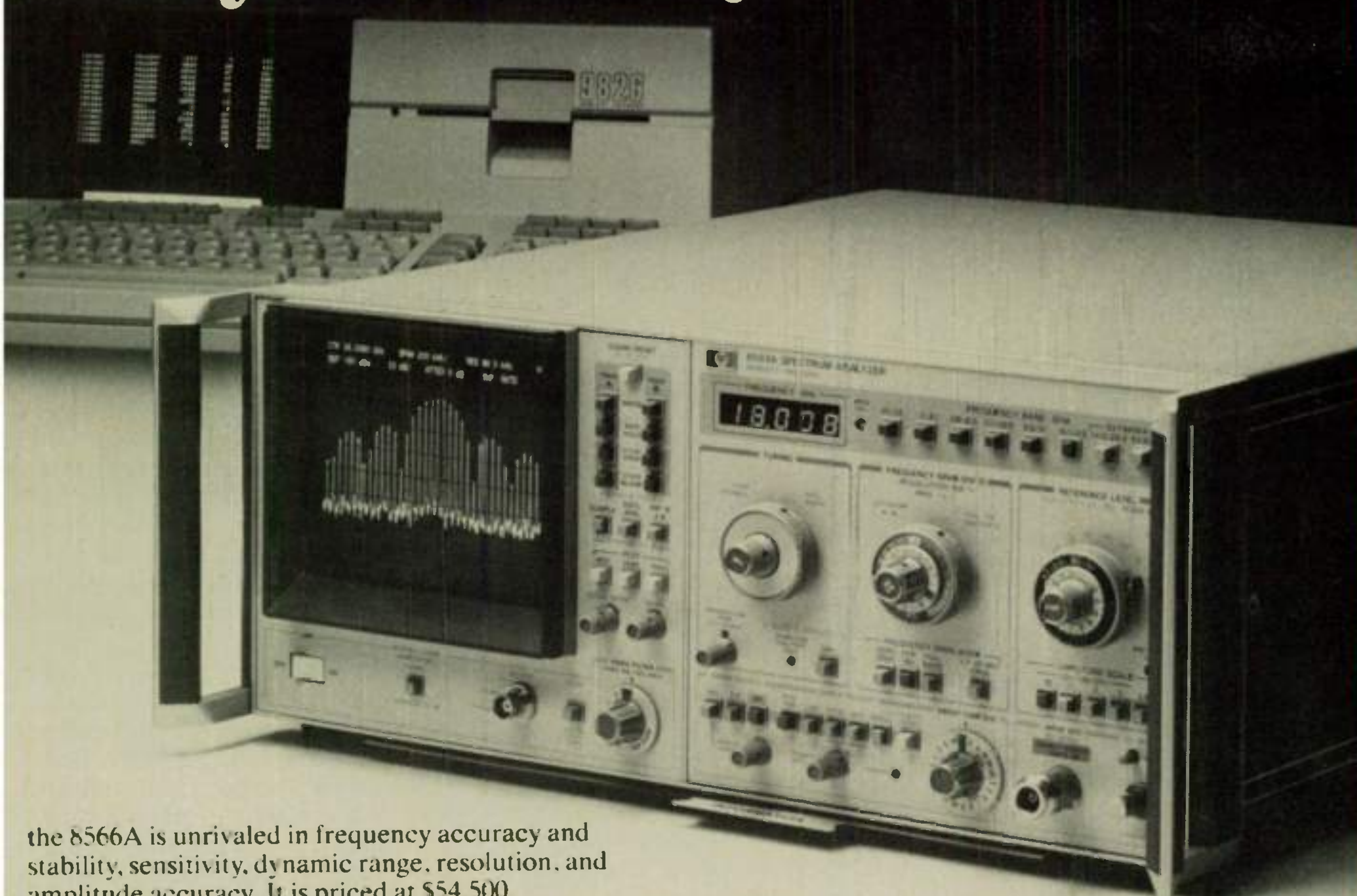

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INTRODUCTION

Antennas using large paraboloid reflectors are expensive to build, maintain and operate. If the reflector can be used simultaneously for more than one frequency band, significant savings can be made in acquisition, running costs and real estate. A popular method of feeding such a reflector is by using a frequency independent device such as a logarithmic periodic antenna. Unfortunately this type of feed degrades the ultimate performance of the antenna system because it cannot provide a truly focused system, it can cause beam squint and it is necessary to demultiplex the different frequency bands prior to amplification. If a feed can be designed with coplanar phase centers at each frequency and a feed port for each band, mutual interaction may be minimized so that optimum performance might be achieved at each frequency band enabling the benefit of low-noise receiving systems to be realized. An example of a coplanar feed is described in Reference 1. An alternative approach devised and used by the author is described in this paper and a model covering three separate frequency bands has been used successfully to illuminate a 45-ft diameter paraboloid reflector with a focal length to diameter (F/D) ratio equal to 0.43. The feed antenna described makes use of a combination of helices and a horn, all of which responded to right hand circularly polarized signals.

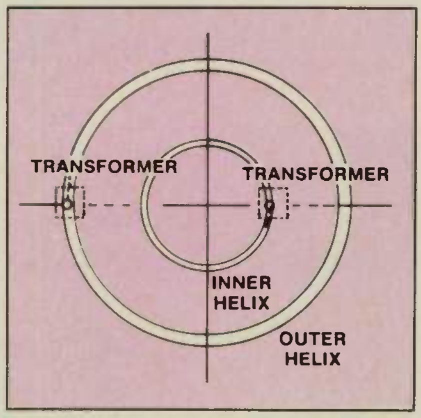

Fig. 1 Arrangement of impedance transforms

DESIGN FEATURES

The 0.9 - 1.0 GHz, 3.6 - 4.2 GHz and 6 - 8 GHz frequency bands were considered for the design study, however, the basic concept can be extended to cover other bands. When a feed is placed at the focus of a paraboloid reflector, the resulting main beam points along the axis of the paraboloid. If the phase center of the feed moves away perpendicular to the boresight axis, the main beam will squint away from the boresight direction by an angle proportional to the feed displacement perpendicular to the axis of the paraboloid. Any movement of the phase center along the reflector axis away from the focus will defocus the system and degrade the antenna performance. It is essential then that any multiple feed system should be designed to have the phase centers of each feed coaxial and coplanar in order to maintain focus and avoid squint. Helical antennas may satisfy this requirement since they can be mounted coaxially on a common ground plane. However, as frequency increases the dimensions of a helix become smaller and the helix is more difficult to fabricate. This was the case for the highest frequency band of the design study and a pyramidal horn was chosen as an alternative. The horn was mounted behind the ground plane with its aperture protruding through a hole and flush with the front face of the ground plane. When helices are placed close to each other mutual coupling between them effects the impedance and radiation patterns of each helix. Coupling can be reduced to a minimum by mounting the helices on a common axis but with their feed points arranged alternately 180° apart as shown in Figure 1.

The beamwidth of a helix is proportional to $1/\sqrt{n}$ where n is the number of turns[2]. Hence by choosing n appropriately a helical feed can be designed to give optimum illumination of a given reflector. The illumination is optimum in the sense that the antenna gain is usually maximum with a 10 dB edge illumination taper. For a helix antenna, the 10 dB beamwidth is approximately equal to

$$\frac{92}{C\lambda\sqrt{nS\lambda}}$$

where $C\lambda$ = circumference of helix in wavelengths
$S\lambda$ = pitch of helix in wavelengths
and n = number of turns.

For various reasons $C\lambda$ is usually made equal to 1 and $S\lambda$ = 0.25 so that

$$10 \text{ dB beamwidth} \simeq \frac{184}{\sqrt{n}}$$

The input impedance of a helix is approximately 140 $C\lambda$ ohms, hence a transformer is required to match this to a 50 ohm coaxial line. A tapered transmission line is a convenient way in which the impedance can be appropriately transformed giving a SWR less than 2 over a 15% band. The impedance of a helix and its radiation pattern depend upon the use of a ground plane at least $\lambda/3$ in diameter. The horn, which was used to cover the highest band, effectively removed the effects of a ground plane, at least as far as the smaller helix was concerned. This in turn changed the radiation pattern and SWR of the 3.6-4.2 GHz helix. This degradation in performance was overcome by introducing a metal ring into the aperture of the horn. If the dimensions of the ring are made equal to the diameter of the 3.6-4.2 GHz helix, this ring becomes a reflector for the helix and effectively restores the ground plane but is effectively invisible at the horn frequencies. This is indicated by the input impedance given in Figure 2. Figure 3 shows that the introduction of a ring of this type did not materially effect the horn but radically improved the polar diagram of the helix.

IMPEDANCE TESTS

The impedance characteristics of each port of the multiple feed antenna were measured. The feeds were tested as separate units and then assembled coaxially with variable terminations at their ports. Figures 3-7 are typical examples which show that the radiation pattern and impedance of an individual feed is essentially uneffected by the presence of the other feeds. The annular ring placed inside

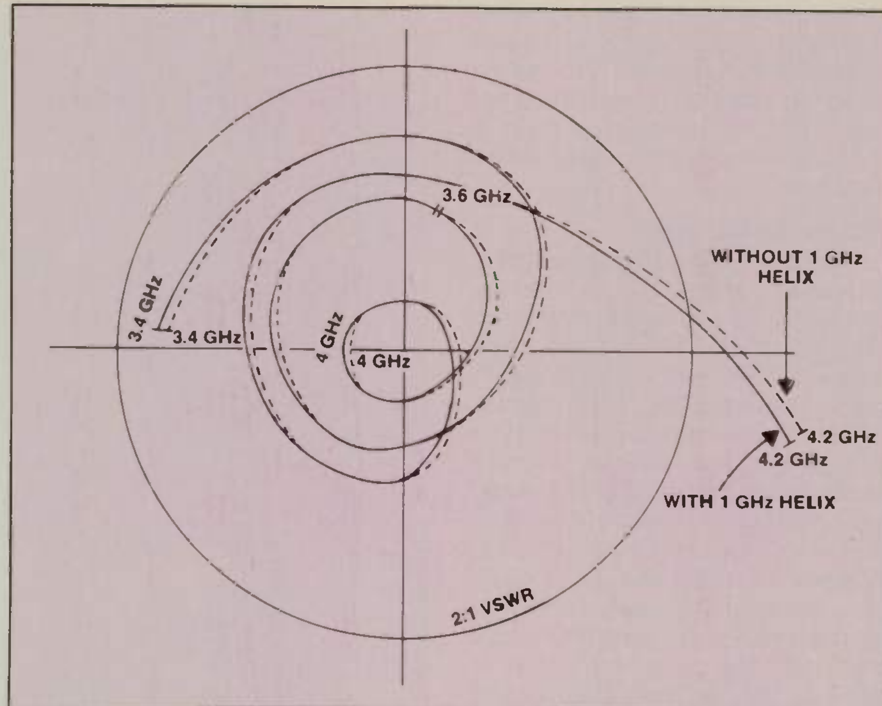

Fig. 2 Impedance charts for 3.4 — 4.2 GHz helix with and without 1 GHz helix.

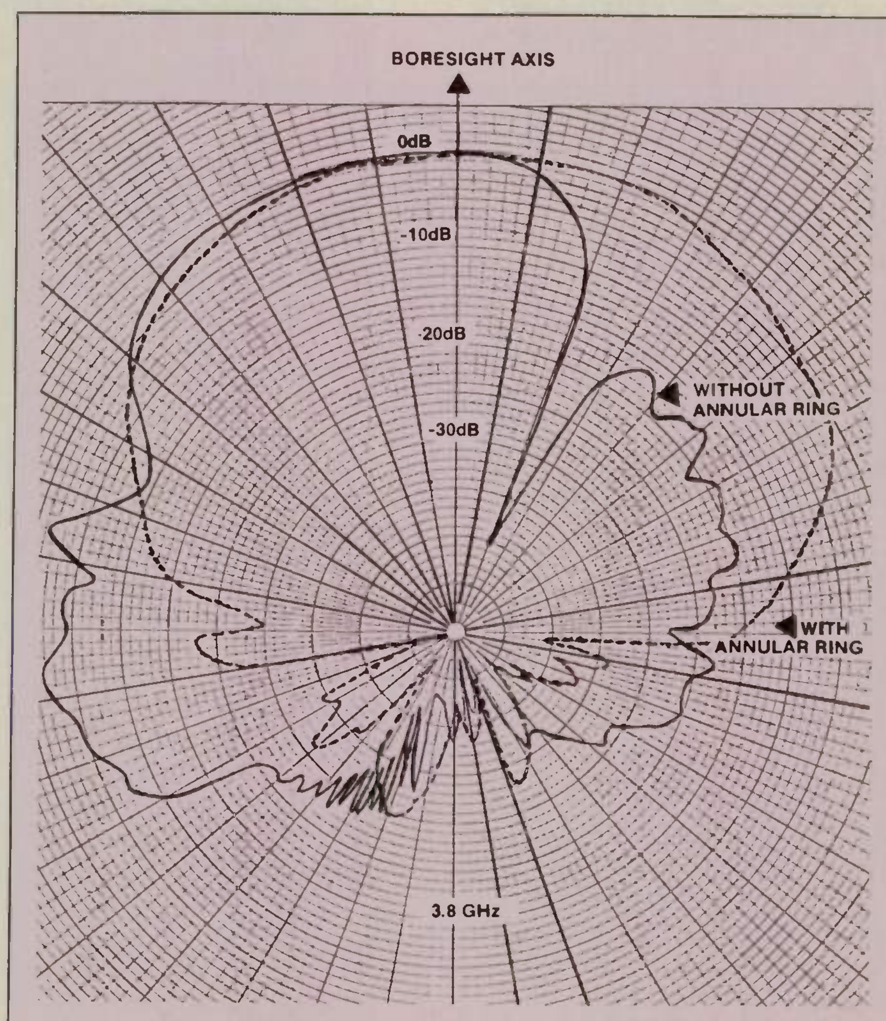

Fig. 3 Polar diagram of 3.4 — 4.2 GHz helix with and without annular ring.

the mouth of the horn degraded the SWR of the horn slightly but the helices themselves had essentially no effect. The corresponding SWR remained less than 2 over the operating frequency range of the horn.

POLAR DIAGRAMS

Radiation patterns were measured at various frequencies across the three bands; with each feed individually and with the other feeds added and assembled in place with variable terminations. All the tests were carried out on an open range with a linearly polarized transmitting antenna approximately 100 wavelengths distance from the receive antenna. Figures 3,5 and 6 show that each feed is only marginally effected by the presence of the other feeds and that the radiation patterns have a 120 — 10dB beamwidth and thus suitable for illumination of a paraboloid.

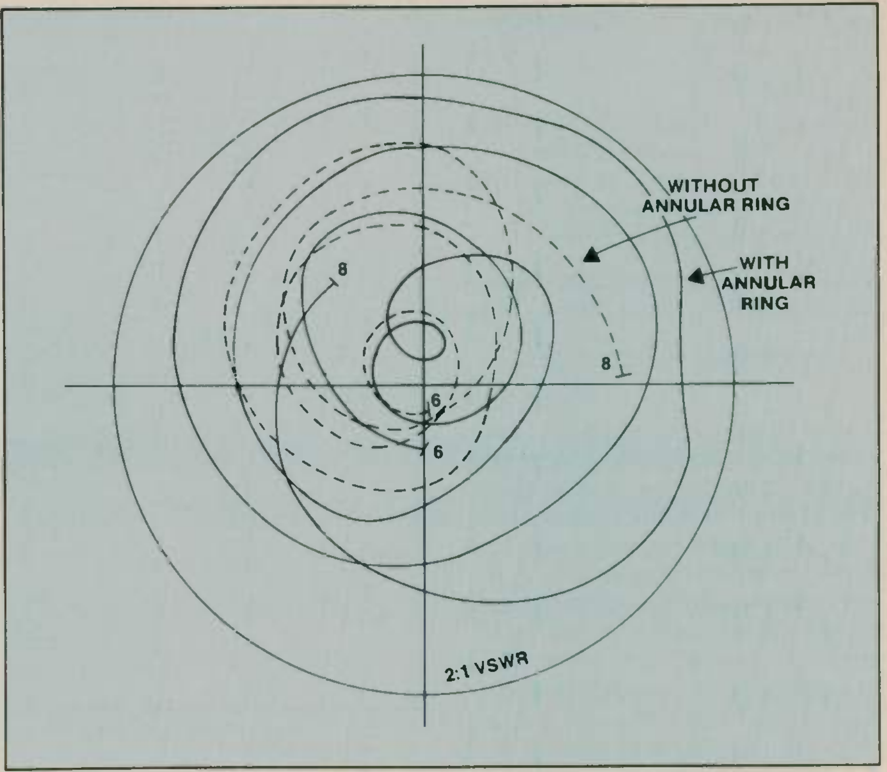

Fig. 4 Impedance chart for 6-8 GHz horn with and without annular ring.

In order to realize this helix with 2l/4 turns, $C\lambda = 1$ and $S\lambda = 0.25$ is required. This small number of turns can cause the polarization to become undesirably elliptical. This can be corrected by reducing the pitch ($S\lambda$) and increasing the number of turns (n) since

$$\text{Axial ratio (ellipticity)} = \frac{2n+1}{2n}$$

$$\text{and beamwidth} \quad \alpha \frac{1}{\sqrt{nS\lambda}}$$

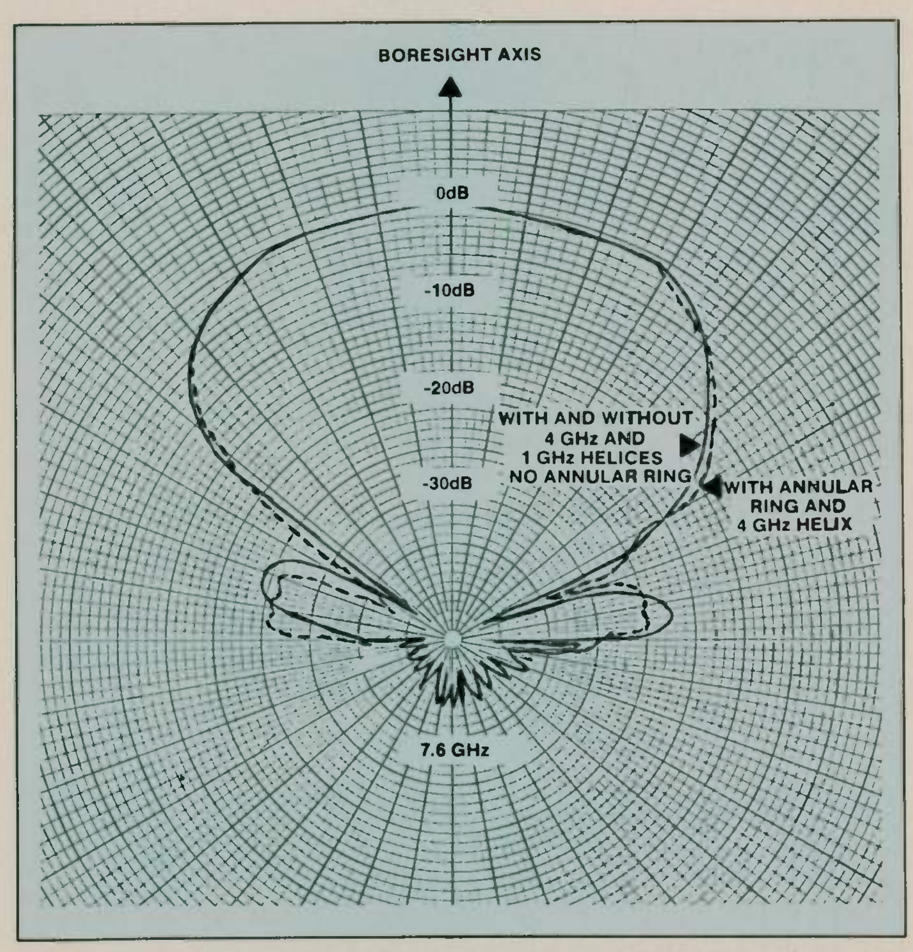

Fig. 5 Polar diagram of 6-8 GHz horn with and without 1 GHz and 4 GHz helices and annular ring.

CONCLUSION

Multiple feed antennas with 3 independent ports covering different frequency bands can be obtained for illuminating a paraboloid reflector. There are problems which need to be overcome such as mutual coupling between elements, squinting of each beam and impedance matching. To reduce beam squint and improve the antenna efficiency each feed must have its phase center at the focus of the reflector. Helices mounted coaxially on a common ground plane will satisfy these conditions. If they are arranged with their feed points spaced alternately at 180° intervals, the mutual coupling between them can be reduced to a tolerable level. Unfortunately the dimensions of

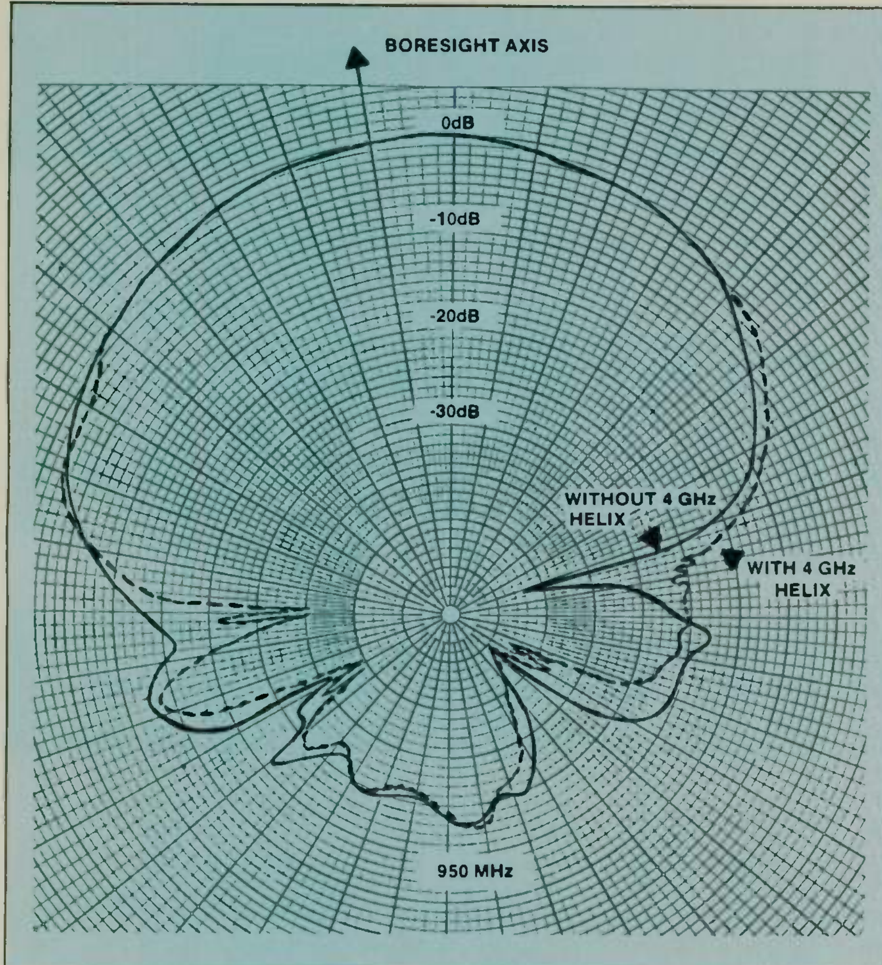

Fig. 6 Polar diagram of 900-1100 helix.

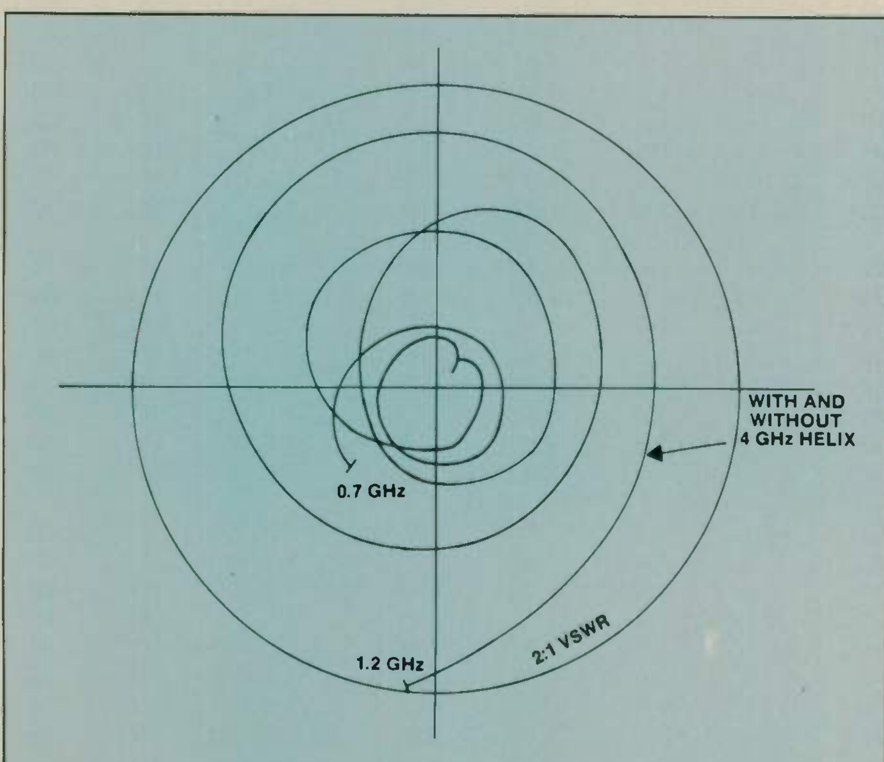

Fig. 7 Impedance chart for 1 GHz helix with and without 4 GHz helix.

a helix are small at frequencies greater than 5 GHz and hence the helix is difficult to fabricate. A horn can be used as an alternative if it is mounted behind the ground plane with its aperture flush with the ground plane.

This will satisfy the coaxial and coplanar requirements but the aperture of the horn will effectively destroy the effects of a ground plane for the smaller helices and drastically alter their impedance and radiation patterns. Annular rings having dimensions equal to the diameter of the helices and placed in the aperture of the horn do not affect the horn performance but effectively restore helix performance obtained with a ground plane. The nominal impedance of a helix is much higher than the 50 ohms transmission line employed; however, a fairly simple tapered transmission line can give a reasonable match. Better impedance matching should be possible over wider bandwidths by more careful design and construction of the tapered transmission line.

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2. J. D. Kraus, *Antennas*, (New York: McGraw Hill Book Co., Inc.) 1961.

J. Holland *served three years with the Royal Air Force as a radar mechanic and received a London University External Degree in Engineering. He then joined the Royal Naval Scientific Service. J. Holland is currently leading a team of scientists developing satellite communication systems.* ■

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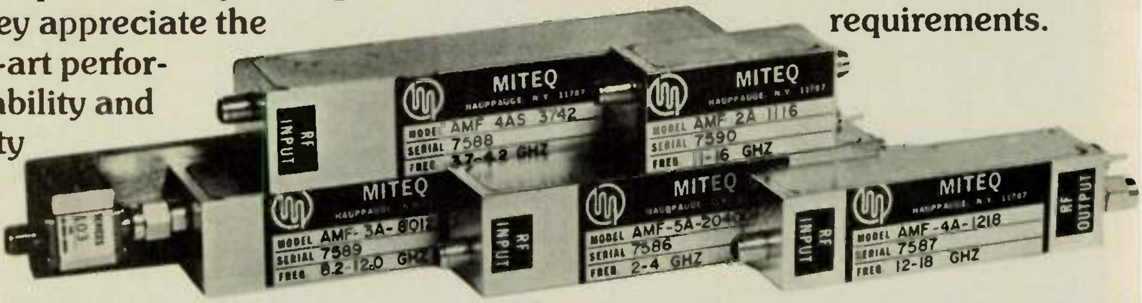

NOISE FIGURE OPTIMIZED FET/BIPOLAR DESIGNS

Model No.	Frequency (GHz)	Gain (dB) Min.	Gain Variation (±dB) Max.	Noise Figure (dB) Typ.	Noise Figure (dB) Max.	VSWR Max. Input	VSWR Max. Output	Dynamic Range 1 dB Gain Comp., Output (dBm) Min.	Dynamic Range 3rd Order Inter. Pt. (dBm) Typ.
AMF-2A-1213	1.2-1.3	25	0.5	1.3	1.5	1.25:1	1.5:1	5	15
AMF-3A-1213	1.2-1.3	35	0.5	1.3	1.5	1.25:1	1.5:1	10	20
AMF-2A-1314	1.3-1.4	25	0.5	1.3	1.5	1.25:1	1.5:1	5	15
AMF-3A-1314	1.3-1.4	35	0.5	1.3	1.5	1.25:1	1.5:1	10	20
AMF-2A-1415	1.4-1.5	25	0.5	1.3	1.5	1.25:1	1.5:1	5	15
AMF-3A-1415	1.4-1.5	35	0.5	1.3	1.5	1.25:1	1.5:1	10	20
AMF-2A-1516	1.5-1.6	25	0.5	1.3	1.5	1.25:1	1.5:1	5	15
AMF-3A-1516	1.5-1.6	35	0.5	1.3	1.5	1.25:1	1.5:1	10	20
AMF-2A-1617	1.6-1.7	25	0.5	1.3	1.5	1.25:1	1.5:1	5	15
AMF-3A-1617	1.6-1.7	35	0.5	1.3	1.5	1.25:1	1.5:1	10	20
AMF-2A-1718	1.7-1.8	25	0.5	1.3	1.5	1.25:1	1.5:1	5	15
AMF-3A-1718	1.7-1.8	35	0.5	1.3	1.5	1.25:1	1.5:1	10	20
AMF-3A-1520	1.5-2.0	25	0.75	1.6	1.9	1.5:1	1.5:1	5	15
AMF-4A-1520	1.5-2.0	35	0.75	1.6	1.9	1.5:1	1.5:1	10	20
AMF-2A-1720	1.7-2.0	22	0.75	1.6	1.9	1.35:1	1.5:1	5	15
AMF-3A-1720	1.7-2.0	30	0.75	1.6	1.9	1.35:1	1.5:1	10	20
AMF-2A-2124	2.1-2.4	22	0.5	1.6	1.8	1.25:1	1.5:1	5	15
AMF-3A-2124	2.1-2.4	30	0.5	1.6	1.8	1.25:1	1.5:1	10	20
AMF-2A-2223	2.2-2.3	22	0.5	1.3	1.5	1.25:1	1.5:1	5	15
AMF-3A-2223	2.2-2.3	35	0.5	1.3	1.5	1.25:1	1.5:1	10	20
AMF-2A-2425	2.4-2.5	22	0.5	1.6	1.8	1.25:1	1.5:1	5	15
AMF-3A-2425	2.4-2.5	30	0.5	1.6	1.8	1.25:1	1.5:1	10	20
AMF-2A-2530	2.5-3.0	18	0.5	1.6	2.0	1.5:1	1.5:1	5	15
AMF-3A-2530	2.5-3.0	27	0.5	1.6	2.0	1.5:1	1.5:1	10	20
AMF-2A-2729	2.7-2.9	20	0.5	1.6	1.8	1.25:1	1.5:1	5	15
AMF-3A-2729	2.7-2.9	30	0.5	1.6	1.8	1.25:1	1.5:1	10	20
AMF-3A-2729-1	2.7-2.9	30	0.5	1.5	1.6	1.25:1	1.5:1	10	20
AMF-2A-2931	2.9-3.1	20	0.5	1.6	1.8	1.25:1	1.5:1	5	15
AMF-3A-2931	2.9-3.1	30	.05	1.6	1.8	1.25:1	1.5:1	10	20
AMF-2A-3035-0	3.0-3.5	18	0.75	1.8	2.2	1.25:1	1.5:1	5	15
AMF-3A-3035-0	3.0-3.5	25	0.75	1.8	2.2	1.25:1	1.5:1	10	20
AMF-2A-3742	3.7-4.2	20	0.5	2.0	2.4	1.25:1	2:1	5	15
AMF-3A-3742	3.7-4.2	25	0.5	2.0	2.4	1.25:1	1.5:1	10	20
AMF-4A-3742	3.7-4.2	30	0.5	2.0	2.4	1.25:1	1.5:1	10	20

AMPLIFIER UPDATE

NOISE FIGURE OPTIMIZED THIN-FILM FET DESIGNS

Model No.	Frequency (GHz)	Gain (dB) Min.	Gain Variation (±dB) Max.	Noise Figure (dB) Typ.	Noise Figure (dB) Max.	VSWR Max. Input	VSWR Max. Output	Dynamic Range: 1 dB Gain Comp., Output (dBm) Min.	Dynamic Range: 3rd Order Inter. Pt. (dBm) Typ.
AMF-2S-3742-4	3.7-4.2	20	0.5	1.3	1.5	1.25:1	1.5:1	5	15
AMF-3S-3742-4	3.7-4.2	30	0.5	1.3	1.5	1.25:1	1.5:1	10	20
AMF-4S-3742-4	3.7-4.2	40	0.5	1.3	1.5	1.25:1	1.5:1	10	20
AMF-2S-4550-3	4.5-5.0	20	0.5	1.7	2.0	1.25:1	1.5:1	5	15
AMF-3S-4550-3	4.5-5.0	30	0.5	1.7	2.0	1.25:1	1.5:1	10	20
AMF-4S-4550-3	4.5-5.0	40	0.5	1.7	2.0	1.25:1	1.5:1	10	20
AMF-2S-5964-3	5.9-6.4	17	0.5	2.5	2.8	1.25:1	1.5:1	5	15
AMF-3S-5964-3	5.9-6.4	26	0.5	2.5	2.8	1.25:1	1.5:1	10	20
AMF-4S-5964-3	5.9-6.4	35	0.5	2.5	2.8	1.25:1	1.5:1	10	20
AMF-2S-7278-3	7.2-7.8	17	0.5	2.5	3.0	1.25:1	1.5:1	5	15
AMF-3S-7278-3	7.2-7.8	26	0.5	2.5	3.0	1.25:1	1.5:1	10	20
AMF-4S-7278-3	7.2-7.8	35	0.5	2.5	3.0	1.25:1	1.5:1	10	20
AMF-2S-8596-3	8.5-9.6	16	0.5	3.0	3.5	1.25:1	1.5:1	5	15
AMF-3S-8596-3	8.5-9.6	24	0.5	3.0	3.5	1.25:1	1.5:1	10	20
AMF-4S-8596-3	8.5-9.6	30	0.5	3.0	3.5	1.25:1	1.5:1	10	20
AMF-2S-109-117-3	10.9-11.7	14	0.5	3.0	4.0	1.25:1	1.5:1	5	15
AMF-3S-109-117-3	10.9-11.7	21	0.5	3.0	4.0	1.25:1	1.5:1	10	20
AMF-4S-109-117-3	10.9-11.7	28	0.5	3.0	4.0	1.25:1	1.5:1	10	20
AMF-2S-117-122-3	11.7-12.2	14	0.5	3.0	4.0	1.25:1	1.5:1	5	15
AMF-3S-117-122-3	11.7-12.2	21	0.5	3.0	4.0	1.25:1	1.5:1	10	20
AMF-4S-117-122-3	11.7-12.2	28	0.5	3.0	4.0	1.25:1	1.5:1	10	20
AMF-2S-140-145-3	14.0-14.5	12	0.5	4.0	5.0	1.25:1	1.5:1	5	15
AMF-3S-140-145-3	14.0-14.5	18	0.5	4.0	5.0	1.25:1	1.5:1	10	20
AMF-4S-140-145-3	14.0-14.5	24	0.5	4.0	5.0	1.25:1	1.5:1	10	20

WIDEBAND DESIGNS

Model No.	Frequency (GHz)	Gain (dB) Min.	Gain Variation (±dB) Max.	Noise Figure (dB) Typ.	Noise Figure (dB) Max.	VSWR Max. Input	VSWR Max. Output	Dynamic Range: 1 dB Gain Comp., Output (dBm) Min.	Dynamic Range: 3rd Order Inter. Pt. (dBm) Typ.
AMF-1B-2040-4	2.0-4.0	10	0.5	2.5	3.5	2:1	2:1	5	15
AMF-2B-2040-4	2.0-4.0	20	0.75	2.5	3.5	2:1	2:1	10	20
AMF-3B-2040-4	2.0-4.0	30	1.0	2.5	3.5	2:1	2:1	10	20
AMF-4B-2040-4	2.0-4.0	40	1.5	2.5	3.5	2:1	2:1	13	23
AMF-1B-4080-4	4.0-8.0	7	0.5	4.0	4.5	2:1	2:1	5	15
AMF-2B-4080-4	4.0-8.0	14	0.75	4.0	4.5	2:1	2:1	10	20
AMF-3B-4080-4	4.0-8.0	22	1.0	4.0	4.5	2:1	2:1	10	20
AMF-4B-4080-4	4.0-8.0	30	1.5	4.0	4.5	2:1	2:1	10	20
AMF-5B-4080-4	4.0-8.0	38	1.5	4.0	4.5	2:1	2:1	13	23
AMF-1B-8012-4	8.0-12.0	6	0.5	4.0	5.0	2:1	2:1	5	15
AMF-2B-8012-4	8.0-12.0	12	0.75	4.0	5.0	2:1	2:1	5	15
AMF-3B-8012-4	8.0-12.0	18	1.0	4.0	5.0	2:1	2:1	10	20
AMF-4B-8012-4	8.0-12.0	24	1.0	4.0	5.0	2:1	2:1	10	20
AMF-5B-8012-4	8.0-12.0	30	1.5	4.0	5.0	2:1	2:1	10	20
AMF-2B-1218-4	12.0-18.0	8	0.75	5.0	6.0	2:1	2:1	5	15
AMF-3B-1218-4	12.0-18.0	12	1.0	5.0	6.0	2:1	2:1	5	15
AMF-4B-1218-4	12.0-18.0	16	1.0	5.0	6.0	2:1	2:1	10	20
AMF-5B-1218-4	12.0-18.0	20	1.5	5.0	6.0	2:1	2:1	10	20
AMF-6B-1218-4	12.0-18.0	25	2.0	5.0	6.0	2:1	2:1	10	20
AMF-7B-1218-4	12.0-18.0	30	2.0	5.0	6.0	2:1	2:1	10	20

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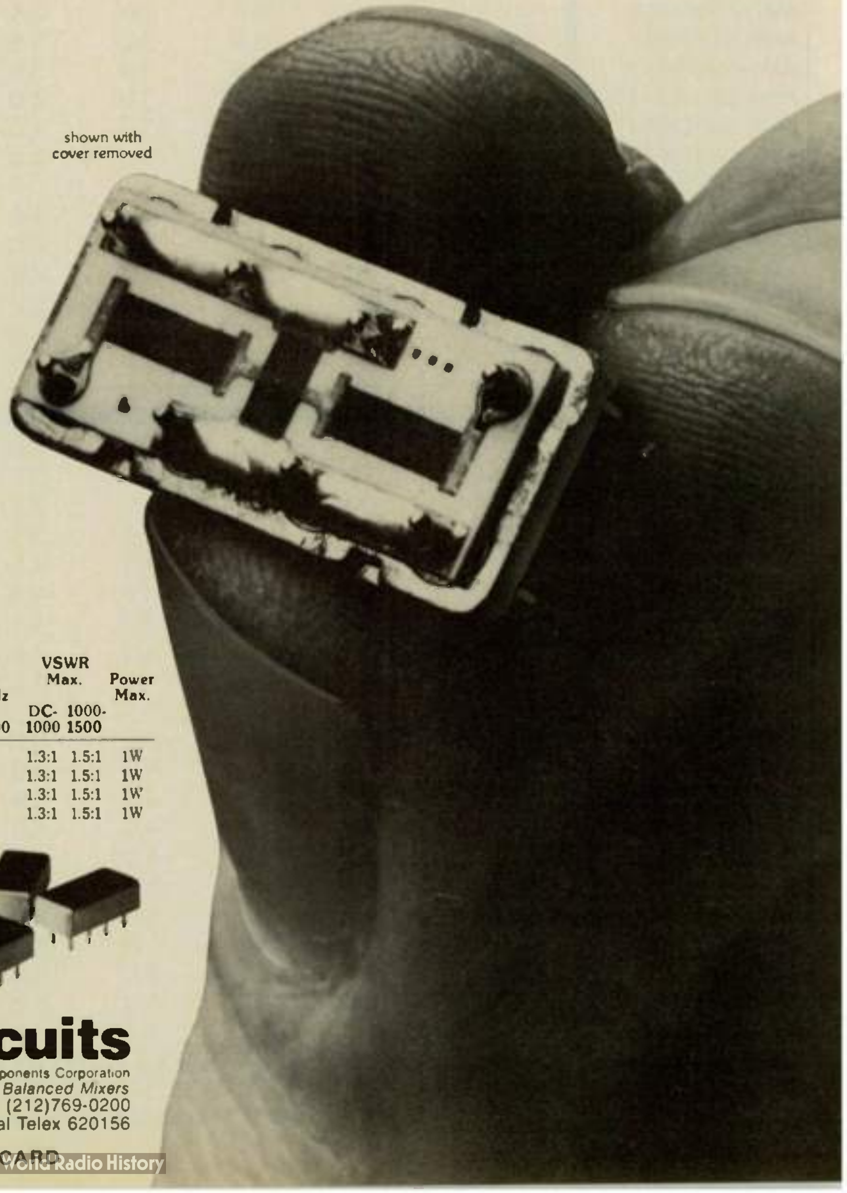
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AT-3	3	±0.2dB	DC-1500	0.6dB	1.0dB	1.3:1	1.5:1	1W
AT-6	6	±0.3dB	DC-1500	0.6dB	0.8dB	1.3:1	1.5:1	1W
AT-10	10	±0.3dB	DC-1500	0.6dB	0.8dB	1.3:1	1.5:1	1W
AT-20	20	±0.3dB	DC-1500	0.6dB	0.8dB	1.3:1	1.5:1	1W

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Range-Height-Angle Charts With Lookdown Capability

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Range-Height-Angle charts serve a variety of useful purposes in radio and radar engineering. Applications include the siting of microwave communication links as well as the theoretical determination of radar coverage. These charts have the important property of allowing electromagnetic ray paths to be depicted as straight lines; while in fact, they curve steadily downward when a ray is within the confines of the atmosphere. These refractive effects may be modeled in several different ways. This article will mention only two of the most popular methods: the 4/3 earth radius method and the exponential reference atmosphere method.

The 4/3 earth radius method is popular because of its analytical simplicity. Literature in the field of radio meteorology[2,3,4] points to a more accurate atmospheric model known as an exponential reference atmosphere. In particular, the exponential reference atmosphere model advanced by the Central Radio Propagation Laboratory[3] has enjoyed wide spread acceptance.

In many applications, particularly at low heights and short ranges, charts constructed on the 4/3 earth radius concept have proven adequate. However, with the advent of AWACS (Airborne Warning and Control System) and ballon borne sensors such as SEEK SKYHOOK, the gap between these atmospheric refraction models became more evident. A need was perceived for a Range-Height-Angle chart with a look down capability and based on an appropriate exponential reference atmospheric model.

The construction of Range-Height-Angle charts entails the following basic steps:

- Choose an atmospheric model.
- Choose the maximum and minimum values of range, height, and angle.
- Working within an angle/range coordinate system; determine a point by starting at the lowest angle and solve for the range to the lowest height (this procedure is dependent on the atmospheric model chosen). This point represents the first point in the constant height line for the lowest height desired. To complete this constant height line; the angle is incremented and range is recomputed at each iteration. After this procedure has been completed through the entire angular range, a smooth curve is drawn through all the computed points. To complete the chart the height is incremented through its entire range with the forementioned procedure completed at all heights.

The only difference between 4/3 earth radius charts and exponential reference atmosphere charts is the procedure in which the range, at a specific departure angle, to a height of interest is calculated.

The 4/3 earth radius model assumes that an electromagnetic ray path can be described by an arc of constant curvature. This arc then will be some multiple of the earth's radius. Four-thirds times the earth's radius is the usual multiple. To compute the range to a height in question, the Olmstead equation is employed[1].

The Olmstead equation is

$$h = A - E = \frac{2(r+E)R \sin\alpha + R^2}{2r + E + A}$$

where

A = Altitude above sea level of a point in space
E = Elevation above sea level of the radar antenna
R = Range (often called Slant Range)
α = Departure angle
r = 4/3 earth radius
h = Height difference between A and E.

If it is assumed that $2r >> E+A$ then the above equation can be rewritten as

$$R^2 + 2Ka\,R \sin\alpha - 2Ka(A - E) = 0$$

where Ka has been substituted for r, and K is the effective multiple (not necessarily 4/3) of the earth's radius, a. The roots of this equation are given by

$$R = Ka \sin\alpha\left[\pm\sqrt{1 + \frac{2(A-E)}{Ka \sin^2\alpha}} - 1\right] \quad (3)$$

Finding ranges, however, for an exponential reference atmosphere is a much more cumbersome task. Where the range in a 4/3 earth radius atmosphere is the root of a quadratic equation, the range in an exponential reference atmosphere requires a numerical integration. The numerical integration technique used in an exponential reference atmosphere is actually a summation of incremental ray tracings.

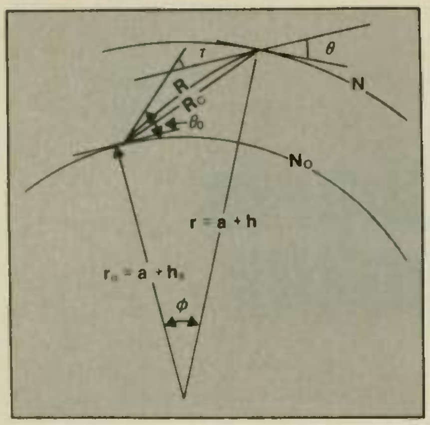

Fig. 1 Ray tracing geometry.

One ray tracing technique assumes an earth above which the atmosphere is spherically stratified into layers of constant air density. In an exponential reference atmosphere, this means the refractive index of each layer decreases in an exponential fashion from layer to layer with height. The refractive index within a given layer remains constant. This model allows computation of a ray path distance from layer to layer because all refractive effects will occur at the boundaries of the various layers. The problem geometry is presented in Figure 1. If the geometrical considerations of Figure 1 are coupled with Snell's Law of Refraction for polar coordinates, a ray tracing methodology can be systematically devised. Considering the first layer above the surface, θ_o represents the departure angle from the antenna, θ represents the departure angle from the boundary of the first atmospheric layer (after refractive effects), and R is the curved path a ray would follow if the index of refraction varied continuously over the atmospheric layer. R_o is the path the ray follows when there is no change in refractive index over the layer's width. The method of ray tracing to be discussed here divides the atmosphere into a sufficient number of small layers such that summation of the R_o distances across each layer approximates R. Φ is the interior angle traversed by the ray. The distance r_o is the radius of the earth (a) plus the height h_s at the bottom of the layer. r is the radius of the earth (a) plus the height h of the top of the layer. N_o is the refractivity at the bottom of the layer, and N is the refractivity at the top. τ is the angular measure called bending; it represents the angle a ray is bent from its original direction due to refraction in the layer.

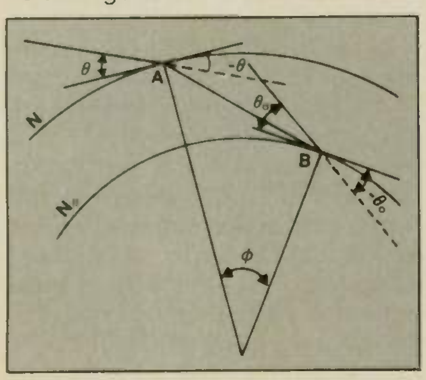

Fig. 2 Negative departure angle geometry.

The ray tracing methodology consists of three basic steps[3].

(a) Computation of the departure angle θ from the boundary of two atmospheric layers. It is given that the ray entered the layer at θ_o. This computation is accomplished via Snell's Law for a spherically stratified atmosphere. With geometry as given, that law is stated as
$nr \cos \theta = n_o r_o \cos \theta_o$
where n and n_o are the indices of refraction in the two layers and the refractivity $N = (n-1) \times 10^6$.

This expression can be manipulated by the use of certain trigonometric expressions to yield an exact functional relationship between θ_o and θ which is usable in single precision arithmetic on such computers as the IBM 360/370. The relationship $\theta = f(\theta_o)$ is (5)

$$\theta = 2 \arcsin \left[\frac{r_o}{2r} \left[2 \sin^2 (\theta_o/2) + \frac{r-r_o}{r_o} - \frac{N_o - N}{n} \times 10^{-6} \cos \theta_o \right] \right]^{1/2}$$

(b) Compution of the bending, τ, between the two layers. τ can be closely approximated by Schulkin's expression[2]

$$\frac{2(N_o - N) \times 10^{-3}}{\theta_o + \theta} = \tau \text{ (milliradians)} \qquad (6)$$

(c) Computation of R_o, the straight line path the ray follows inside of a given layer. When the layers are made sufficiently small R_o will approximate R, the curved path a ray would follow if it traveled through an atmosphere wherein the index of refractivity varied in a continuous fashion. This computation is made solely on the basis of geometric considerations.

Those considerations are

$$\Phi = \theta - \theta_o + \tau \qquad (7)$$

and from the law of cosines

$$R_o = \sqrt{2r_o(r_o + h)(1 - \cos \phi) + h^2} \qquad (8)$$

This methodology breaks down when a ray tracing is attempted from an initial departure angle less than zero degrees.

One method of tracing at negative angles involves exploitation of the geometry of Figure 1 which is redrawn in Figure 2. Notice Figure 2 is the mirror image of Figure 1. Figure 2 is designed to graphically illustrate the fact that a ray trace performed from a negative departure angle $-\theta$ at point A with height h_s to a point B with height h is equivalent to tracing a ray from B to A. In other words, a ray leaving height h_s at $-\theta$ is refracted to an angle of $-\theta_o$ at height h and inspection of Figure 2 shows this is the same as starting at an angle of θ_o at height h and tracing upwards to a height h_s where the ray is refracted to an angle of θ.

Hence to trace rays at negative departure angles, the same methodology as outlined above can be applied over each layer. All that has been changed, from layer to layer, is the exchange of the roles of θ_o and θ. However, caution is to be exercised. Under normal atmospheric conditions, when a ray leaves a given height at some positive departure angle it never returns to that height. When a ray leaves at a negative departure angle it travels downward until one of two things happen: the ray impacts the earth, or the ray passes tangent to the earth and begins to climb upward towards the height from which it originated. In Figure 3 a ray departs downward from H, passes tangent to a stratified layer at T, and climbs upward again to H. It should be noticed that the path from H to T is identical to the path from T back up to H again. This fact will be exploited later.

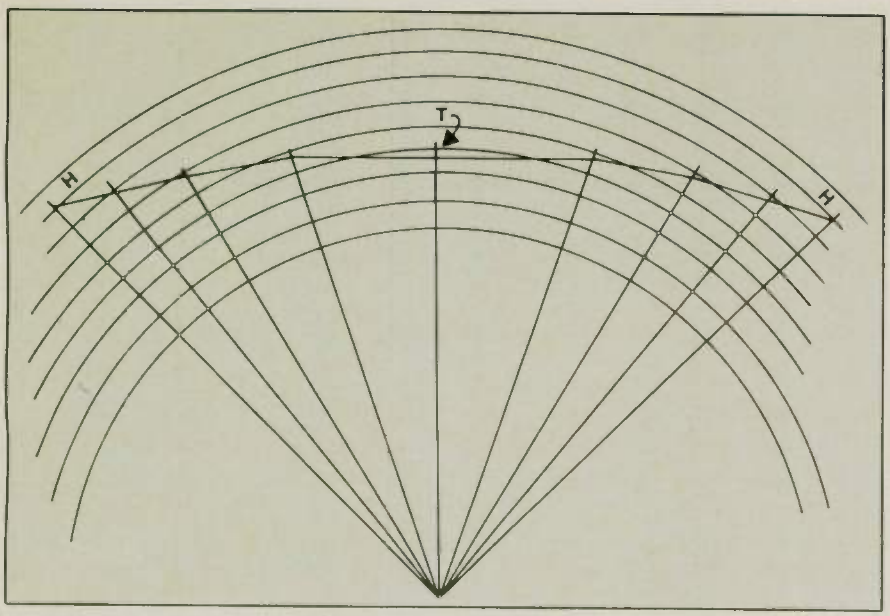

Fig. 3 Lookdown geometry.

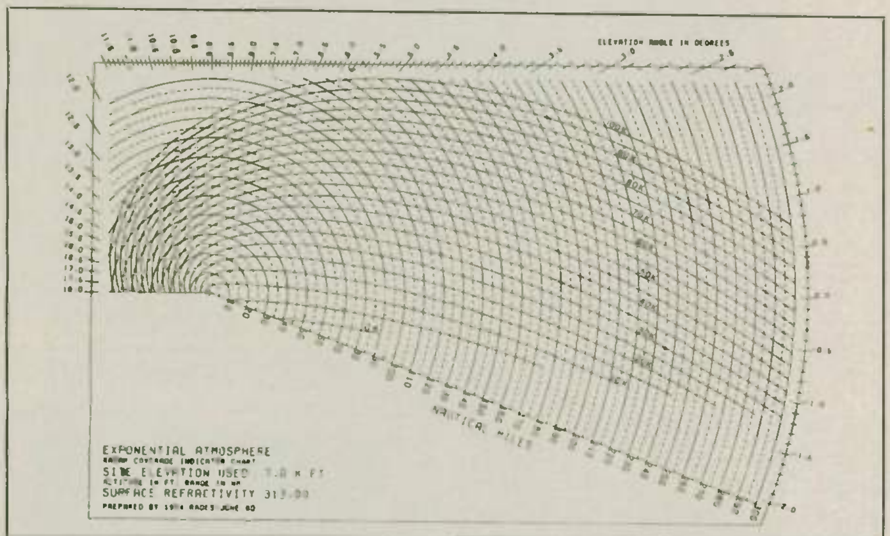

Fig. 4 Range-Height-Angle chart using the CRPL exponential reference atmosphere. (site elevation: 7.0 k FT)

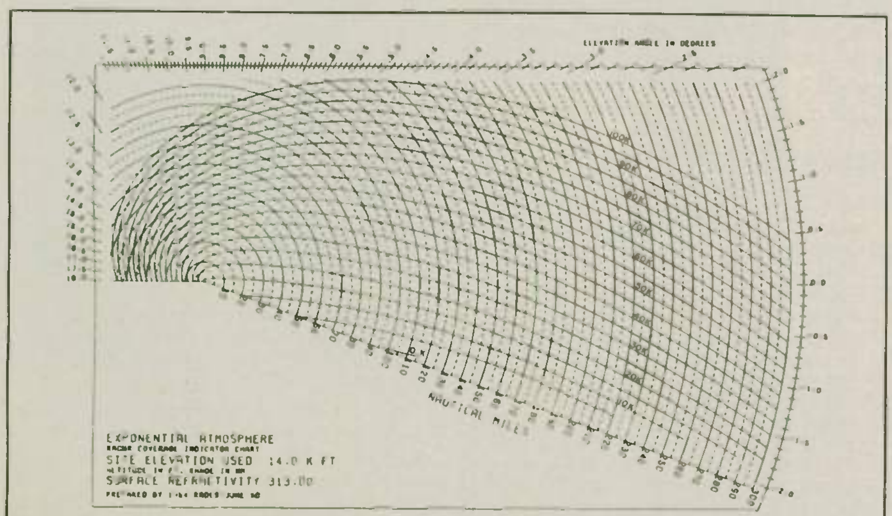

Fig. 5 Range-Height-Angle chart using CRPL exponential reference atmosphere. (site elevation: 14.0 K FT)

The ray tracing methodology, which was descibed above, can now be applied to ray tracings at negative departure angles. Now however, the departure angle from the upper boundary is known and the departure angle from the lower boundary is unknown. Hence θ is known and θ_o is unknown. Now from above

$\theta = f(\theta_o)$

but, what is needed is

$\theta_o = g(\theta)$

where g is the inverse function of f. The function g (θ) is found in similar fashion to $f(\theta_o)$. The function $\theta_o = g(\theta)$ is given by

$$\theta_o = 2 \arcsin \left[\frac{r}{2r_0} \left[2 \sin^2 (\theta/2) + \frac{r_{o-r}}{r} - \frac{(N - N_o)}{n_o} \times 10^{-6} \cos \theta \right] \right]^{1/2} \quad (9)$$

Notice the similarity between the respective equations. The geometry is as in Figure 2. The algorithm, then, for ray tracing at negative departure angles consists of the following steps. First the atmosphere is divided into small layers, .01 km, below the initial height. Proceeding layer to layer, θ_o is calculated by $g(\theta)$. τ is calculated as before. Φ is calculated as before and the range over the layer in question follows. This procedure is followed until g (θ) fails (the argument of the square root goes negative). Physically this occurs when an attempt is made to trace below the tangency height. To overcome this, θ_o is set to zero and an iterative procedure is used to calculate the height of the tangency point. Then the final range increment is calculated and the trace from a height to its tangency point has been accomplished. If the trace had been desired to reach its original height, all that is required is doubling of the computed range to the tangency point as shown in Figure 3.

Figures 4, 5, 6, and 7 are examples of Range-Height-Angle charts constructed using the CRPL exponential reference atmosphere. In all examples the refractivity at mean sea level (surface refractivity on the charts) has been chosen at 313. Elevation angles are given in degrees and run along the outer perimeter of the chart. All exam-

[Continued on page 92]

[From page 91] **CHARTS**

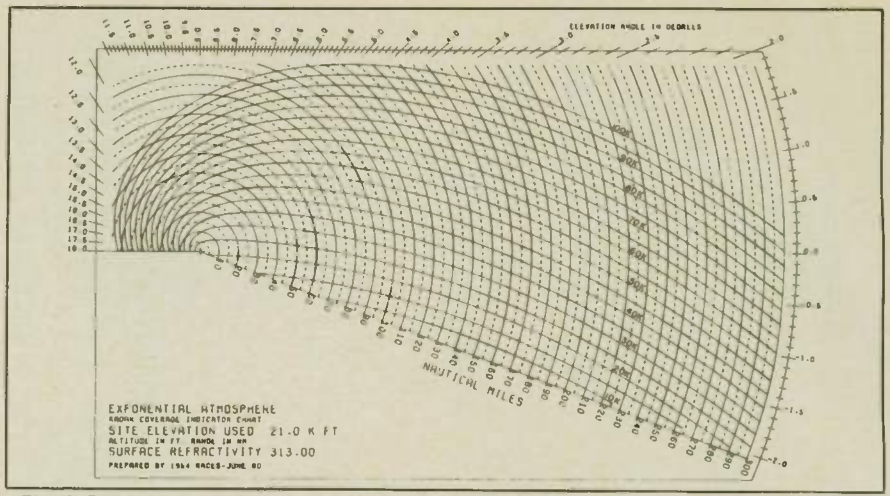

Fig. 6 Range-Height-Angle chart using the CRPL exponential reference atmosphere. (site elevation: 21.0 K FT)

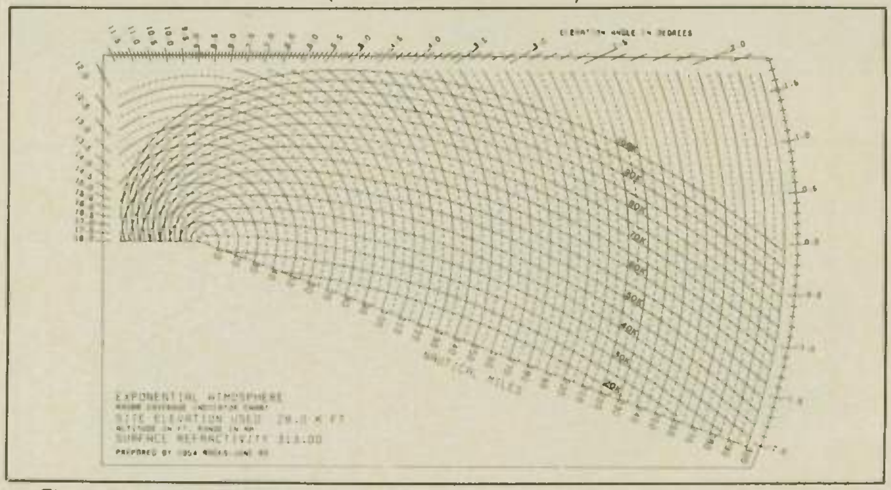

Fig. 7 Range-Height-Angle chart using CRPL exponential reference atmosphere. (site elevation: 28.0 K FT)

ples are constrained to an angular range from -2 degrees to 18 degrees. Charts are given for site heights (initial propagation heights) of 7, 14, 21, and 28 thousand feet. The ray tracing procedure is as presented in this paper and incorporates the CRPL exponential reference atmosphere equations[3].

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Short Backfire Antenna For Doppler Sensing

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This paper describes the construction of a circular short backfire antenna fed from a standard X-band waveguide (WR 90/WG 16) for use at 10.5 GHz. Frequency scaling is difficult with this antenna and an aperture transforming principle is discussed.

SHORT BACKFIRE ANTENNAS

The backfire principle is close to coming of age since its first description by H. W. Ehrenspeck[1] in 1960. During its earlier life many variants have been topics of research. Both the original and short backfire antenna reported by the same author[2] in 1965 concentrate, though not exclusively, on a design with a dipole feed between two reflectors spaced half a wavelength apart, as shown in Figure 1a. The principle is well known[3,4]. The parallel reflectors create multiple images of the feed dipoles giving the effect of a considerably longer end-fire array. By making the front reflector much smaller than the other, it becomes an imperfect or leaky mirror so that radiation takes place in the direction of the reflector, as in a leaky open resonator employed as a laser cavity.

Inter-reflector spacing can be any multiple of half a wavelength, in principle, to show resonant enhancement of gain. At the minimum separation this class of antenna is termed the short backfire and the best spacing is half a wavelength or slightly longer depending on the feed. More gain can be achieved by increasing the spacing to multiple half wavelengths but this "long backfire" antenna requires a surface wave structure as well as a dipole feed. Some designers have successfully used complicated surface wave and resonator combinations[5]. This surface wave structure can be a dielectric rod[6] fed from a rectangular or circular waveguide. The short backfire antenna can similarly be waveguide fed and a surface wave structure may or may not be used.

SHORT BACKFIRE WITH WAVEGUIDE FEED

A review of waveguide fed short backfires was given by A. C. Large in the *Microwave Journal*[7]. His best design is reproduced in Figure 1b. Particular features of the design are 1) increased gain obtained by the penetration of the waveguide into the center of the cavity, 2) optimum inter-reflection spacing of 0.6λ 3) employment of a rectangular front reflector 0.4λ x 0.9λ to give a good input match and nearly equal half-power beamwidths in the E and H planes. The test antenna design for a 9.0 GHz center frequency was reported to produce the results in summarized Table 1.

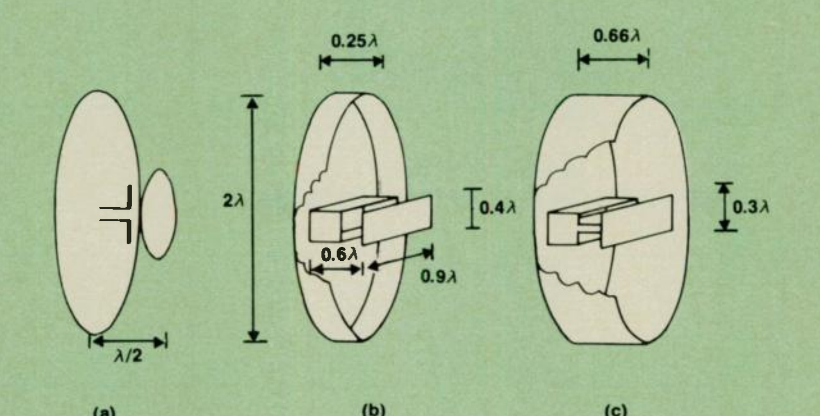

Fig 1. Short backfire antennas.

TABLE I
Test results for the antenna shown in Figure 1b.

GAIN	13.5 dB
HALF BEAMWIDTHS	
E plane	29°
H plane	31°
VSWR (9 GHz)	~1.1
BANDWIDTH	500 MHz

This design suggests that the useful antennas, shorter than a conical horn of similar gain, and having a waveguide input are possible. Note that gain is not sensitive to frequency and bandwidth is defined by the tolerable input mismatch say VSWR <1.5. A waveguide input permits increased power handling, increased compatibility, and easier fabrication for frequencies ~10 GHz and higher. For example, oscillator designs using Gunn diodes are usually easier to realize with waveguide cavities and hence are more compatible with a waveguide fed antenna.

HIGHER FREQUENCY DESIGNS

It is convenient for laboratory models to support the front reflector on expanded dielectric foam which permits adjustment of the inter-reflector spacing. This is unsuitable for repeatable performance or production since the tolerance, for good input impedance match, on the relfector spacing is less than 0.5mm and the reflectors must be maintained parallel. Thus the design shown in Figure 1c was evolved where the small front reflector was supported by a central web in the H-plane of the waveguide aperture. Since the web is normal to the TE_{10} mode fields in the aperture it has little effect on the radiation properties or VSWR, but does facilitate precise manufacture. Secondly, the design shows a deeper rim compared to the customary 0.25λ used by many others. The indicated rim depth 0.66λ results in a 0.25 dB decrease in gain, but also reduces sidelobes to ~-35 dB and the bandwidth ~ 500 MHz (5%). The main advantage of the deeper rim is that it provides a convenient means of mounting the antenna if it is to fit flush into a surface and, most importantly, gives a support for a flat dielectric dust cover over the aperture. The measured gain is 13.2 dB.

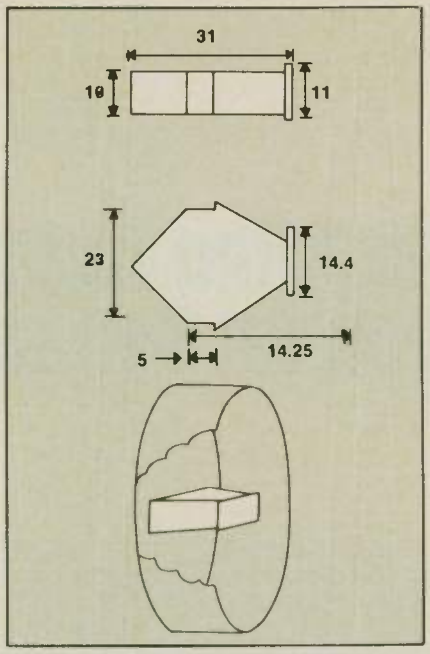

Fig. 2 10.5 GHz antenna showing dielectric insert and small reflector (dimensions in mm.)

Frequency scaling of this antenna by simple proportional dimensional changes is generally difficult because the waveguide aperture is fixed and waveguide transitions are inconvenient for minimum length. Reducing the dimensions beyond the point at which the front reflector is similar in size to the inserted waveguide cross section is unprofitable since excessive shadowing occurs.

One solution is to have the feed aperture flush with the back reflector, though this by itself is less effective. A flush fed cavity can be improved as shown in Figure 2 by the addition of a tapered dielectric insert. The insert is given a sharp taper to encourage leakage in the forward direction and improve bandwidth. Further, since the dielectric supports a surface type wave it tends to produce multiple reflections. Because of the surface guiding action the small reflector size is fixed by the size of the dielectric cross section and hence can be made smaller when the sharp taper is employed. The design shown in Figure 2 has a small reflector 0.32λ x 0.5λ substantially reducing the aperture blockage.

PERFORMANCE OF DIELECTRIC LOADED SHORT BACKFIRE

The performance data for the short backfire antenna with added dielectric surface wave structure is shown in Figures 3 and 4. Its radiation patterns are shown in Figure 3. Note that the H-plane pattern has no well defined sidelobes, and the E-plane pattern has a shoulder on the main beam, at -10 dB below boresight. These principal plane patterns are quite symmetrical with respective half-power widths of 31° and 28° in the H and E planes.

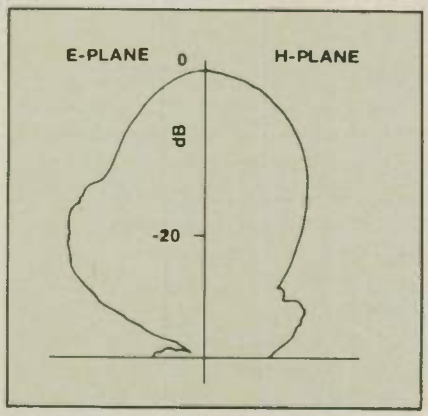

Fig. 3 Polar radiation pattern in principal planes of 10.5 GHz antenna.

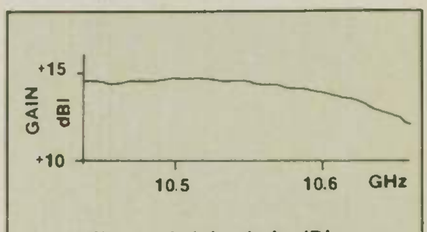

Fig. 4a Axial gain in dBI.

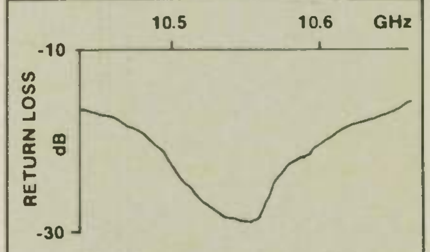

Fig. 4b Reflected return loss.

Fig. 4 Frequency response of 10.5 GHz antenna.

Figures 4a and 4b show the frequency variation of axial gain and return loss. Figure 4a has a peak of 14.6 dB at 10.5 GHz. Gain is not particularly sensitive to inter-reflector spacing, or, alternatively, frequency, and can be seen to remain fairly constant over a 100 MHz band. [Continued on page 96]

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[From page 95] **BACKFIRE**

Input reflection, measured in terms of return loss, in contrast with axial gain, is extremely sensitive to reflector spacing. Even if this spacing is correct, a dielectric insert adds a further reflection from its input face. Should this face be flat, good input match is only possible over a very narrow band placing severe limits on manufacturing tolerances. The design precision is eased, however, by a shallow taper on the input face of the dielectric. Figure 4a shows the input VSWR is less than 1.2 over a 100 MHz band and is a minimum of 1.06.

CONCLUDING COMMENTS

Waveguide-fed short backfire antennas have the advantage of high gain per unit length. The designs discussed here have 2λ diameter circular apertures and a best gain figure of 14.6 dB shown with an overall length of the dielectric insert of 1.1λ.

A comparable conical horn would require a length of 2λ. The disadvantage is that they are resonant antennas and precise control of the resonant frequency is required if low VSWR is to be maintained. At lower frequencies, say below the center frequency of the waveguide, the inserted feed design offers the shortest antenna of about 0.7λ. Frequency control from a manufacturing viewpoint is easily achieved by supporting the front reflector on a central web in the H-plane of the waveguide aperture.

For frequencies where the inserted waveguide shadows the front reflector it may be supported by a dielectric structure and the waveguide withdrawn flush with the back reflector. In both cases, the parts can be simply manufactured to good tolerances so that repeatable results are easily obtained. Further, both designs can have an extended rim for support of a dust cover or pressure seal.

The short backfire antenna is difficult to design because of a lack of theoretical base and must require empirical optimization for a particular case. Nonetheless, they have uses for fixed frequency operation such as Doppler sensing equipment where the length of a conventional horn may be intolerable. Though these results are presented for X-band, providing similar relationships between the parts of the antenna are maintained they should be applicable to any other frequency band.

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Dr. A. G. Martin graduated with B. Eng from the University of Sheffield, U. K., in 1963 and continued research there on transverse wave tubes, gaining a Ph.D. In 1966 he joined the staff of the University of Hull, U. K., at the inception of its Electronic Engineering Department. Alan's Research has been primarily concerned with microwave devices and particularly the modeling of radiators. He has published several papers in this connection and co-authored one book, "Linear Microelectronic Systems," by Macmillan Press. Alan intends to join British Aerospace, Space and Communication Division, to take responsibility for the satellite payload group in the near future. ■

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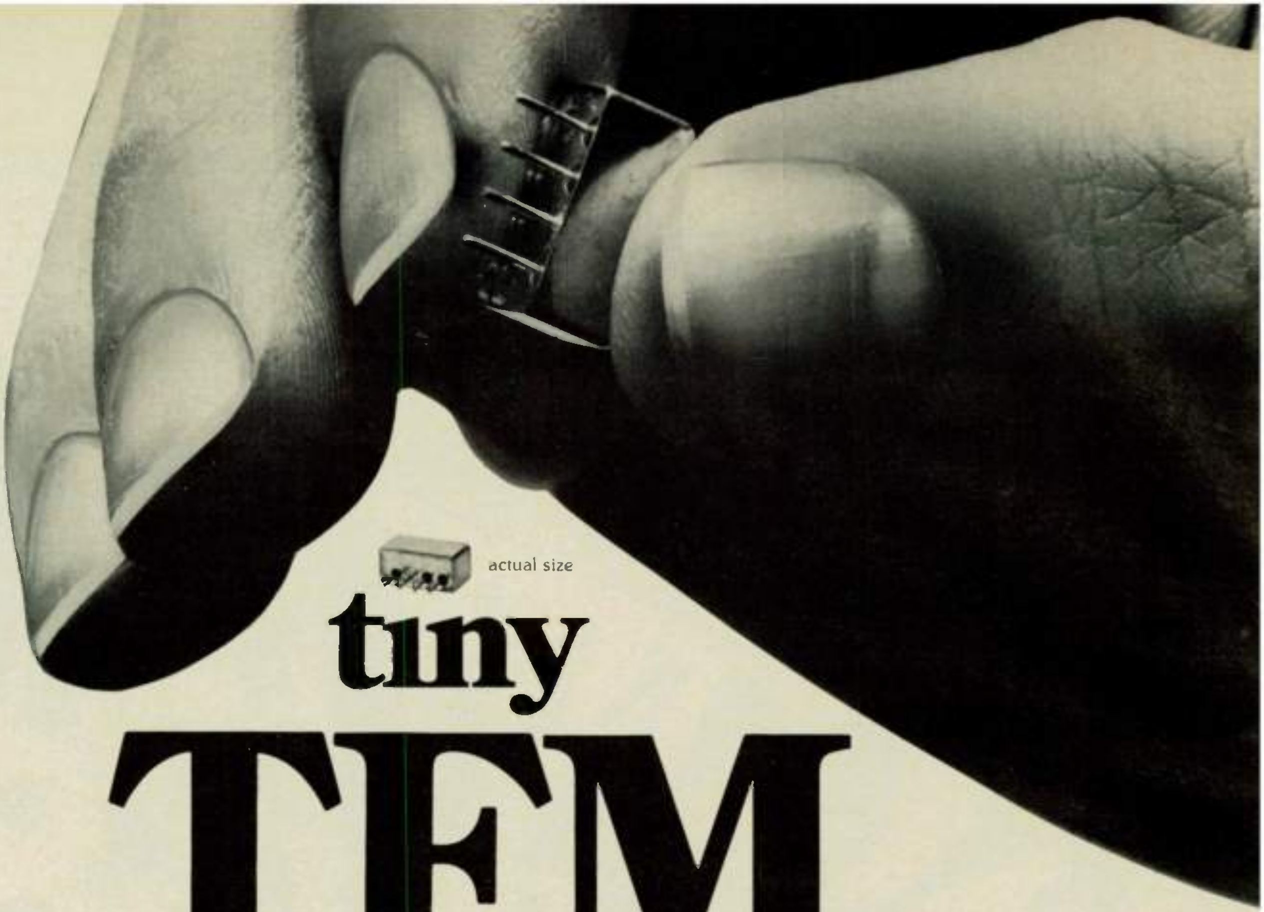

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TFM-4	5-1250	DC-1250	6.0	7.5	50	45	40	35	30	25	21.95	(5-49)
•TFM-11	1-2000	5-600	7.0	7.5	50	45	35	27	25	25	39.95	(1-24)
•TFM-12	800-1250	50-90	—	6.0	35	30	35	30	35	30	39.95	(1-24)
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Tangential Sensitivity of EW Receivers

JAMES TSUI
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INTRODUCTION

Sensitivity of an EW receiver is a very common and very important performance parameter. Although there are many papers discussing this subject, there are still many questions concerning sensitivity of receivers, such as:

- What is a practical way to find the effective noise bandwidth to be used in the sensitivity calculations of an EW receiver?
- How does the radio frequency (RF) gain in the receiver affect the sensitivity of the receiver?
- How do the RF bandwidth (B_R) and video bandwidth (B_V) affect the sensitivity?

The answers to all these questions can be found either explicitly or implicitly in References 1 through 6. This paper tries to systematically approach this problem and compares the calculated results with experimental data. The tangential sensitivity (TSS) is used in this paper rather than the probability of detection, since the tangential sensitivity is easy to observe experimentally. The tangential sensitivity is obtained through visual display on an oscilloscope at the output of a diode detector or the output of the video amplifier following the detector. On the display, when the bottom of the noise edge of a pulse is leveled with the top of the noise edge without the input pulse as shown in Figure 1, the receiver is at its tangential senstivity. It is generally agreed that at tangential sensitivity the signal is 8 dB above the noise level at the output of the detector[1,2]. However, a ±2 dB variation about the mean has been observed among different observers (Figures 1,2).

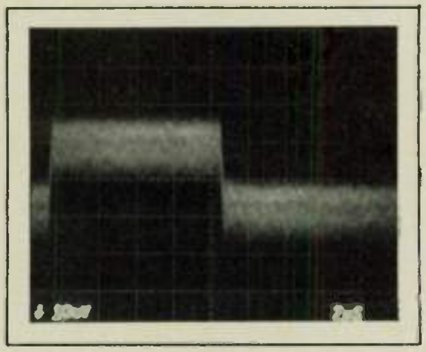

Fig. 1 Output of crystal detector when input signal is at TSS.

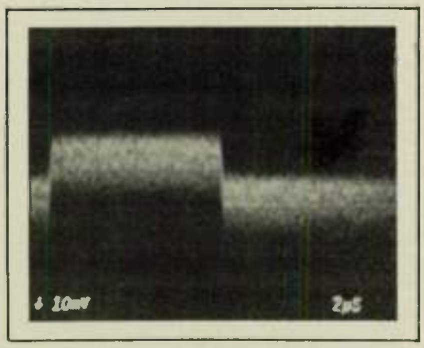

Fig. 2 Output of crystal detector when input signal is 2 dB below TSS.

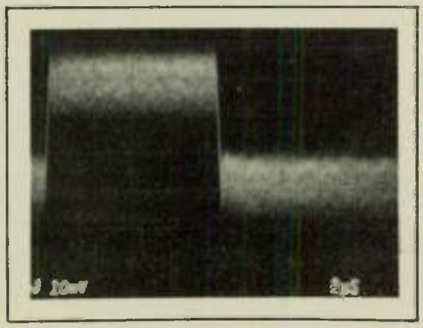

Fig. 3 Output of crystal detector when input signal is 2 dB above TSS.

BASIC EQUATIONS

In all practical microwave receiver designs, the minumum RF bandwidth (B_R) is at least as wide as the video bandwidth (B_V) in order to receive the information properly. Thus, the condition where $B_R < B_V$ is not discussed in this paper. Square law detectors are considered, since at the tangential sensitivity level most detectors work in the square law region.

The relations derived in References 1 and 2 are quite similar; however, different constants are used in the final expressions, and Lucas[2] has an additional term which is related to the property of the detector and the amplifier following the detector. However, the equations derived in Reference 2 are not expressed explicatively, therefore they are not convenient to use. This paper will extend some of the equations in Reference 2, discuss them in detail and try to make them self-explanatory. The tangential sensitivity from Lucas' paper can be written in the following forms:

(1a) $TSS = kTf_t\,[3.15\,B_R + 2.5\sqrt{2B_RB_V - B_V^2 + AB_V/(G_TF_T)^2}] \times 10^6$ watts

(1b) $TSS(dB) = -114 + 10\log F_T + 10\log[3.15B_R + 2.5\sqrt{2B_RB_V - B_V^2 + AB_V/(G_TF_T)^2}]$ dBm

for $B_V \leq B_R \leq 2B_V$

(2a) $TSS = kTF_T\,[6.31\,B_V + 2.5\sqrt{2B_RB_V - B_V^2 + AB_V/(G_TF_T)^2}] \times 10^6$ watts

(2b)

$$TSS(dB) = -114 + 10 \log F_T + 10 \log [6.31 B_V + 2.5 \sqrt{2B_R B_V - B_V^2 + AB_V/(G_T F_T)^2}\,] \text{ dBm}$$
$$\text{for } B_R \geq 2B_V$$

TSS(dB) represents the input signal power level in dBm that will produce an 8 dB signal to noise ratio (S/N = 8 dB) at the output of the detector/video amplifier, while TSS is the tangential sensitivity expressed in watts.

Where:

- F_T is the total noise figure from the input of the receiver to the detector and will be discussed later.
- B_V is the video bandwidth in MHz (3 dB bandwidth).
- B_R is the RF bandwidth in MHz (3 dB bandwidth).
- G_T is the total gain from the input of the receiver to the input of the diode.
- -114 dBm is thermal noise floor for a 1 MHz bandwidth. The thermal noise floor can be derived as 10 log (kTB)
- Where k is the Boltzman constant (= 1.38×10^{-23} W/°K);
- T is the room temperature in degrees Kelvin = 290° K.
- $kTB = 1.38 \times 10^{-23} \times 290 \times 10^6 = 4 \times 10^{-15}$ W/MHz $= 4 \times 10^{-12}$ MW/MHz
- 10 log (kTB) = -114 dBm

The noise figure F_T can be obtained from the usual equation (referred to as the Friis formula).

$$F_T = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \ldots + \frac{F_n - 1}{\prod_{i=1}^{n-1} G_i} \quad (3)$$

Where $F_1, F_2 \ldots F_n$ are the noise figure (expressed in power ratio) of the first, second and nth elements in the RF path of the receiver. $G_1, G_2 \ldots G_n$ are the gain of the first, second and nth elements. If the first element is a 3 dB attenuator then $F_1 = 2$ (corresponding to 3 dB) and $G_1 = 1/2$. The gain G_T can be calculated as $G_T = G_1 G_2 \ldots G_n$

The A's in Equations 1 and 2 is a constant related to the diode parameter and the noise figure of the video amplifier following. It can be expressed as[2]

$$A = \frac{4F}{kTM^2} \times 10^{-6} \quad (5)$$

where F is the noise figure of the video amplifier expresed as power ratio, and M is the figure of merit of the diode, which can be expressed as[7]

$$M = \frac{C}{\sqrt{R}} \quad (6)$$

where C is the detector sensitivity in MV/MW and R is the dynamic impedance of the diode in ohms.

VIDEO DETECTOR CONSTANTS

The value of A in Equations 1, 2 can be determined in the following manner. When a video detector is considered as the only element in a microwave receiver, there is no gain or loss in front of the detector then $G_T = F_T = 1$ in Equations 1 and 2. Under such conditions, the only dominate term in these equations is that containing A. Equations (1a) and (2a) can be reduced to

$$TSS = 2.5\, kT \sqrt{AB_V} \times 10^6 \text{ watts} \quad (7a)$$

$$TSS\ (dB) = -110 + 10 \log \sqrt{AB_V} \text{ dBm} \quad (7b)$$

substituting equations (5) and (6) into (7a) one can obtain

$$TSS = \frac{2.5\sqrt{4kTB_V F}}{M} \times 10^3 \text{ watts} \quad (8a)$$

$$= \frac{3.16 \times 10^{-4} \sqrt{B_V F}}{M} \text{ milliwatts} \quad (8b)$$

$$TSS(dB) = -35 + 10 \log \sqrt{B_V F} - 10 \log M \text{ dBm} \quad (8c)$$

The constant A can be determined in the following ways:

- The figure of merit M of the detector and the noise figure F are given by equation (5)
- The sensitivity C and dynamic impedance R of the detector are given by equations (5) and (6)
- The tangential sensitivity of the detector is given by equation (7).

RF GAIN LIMITED RECEIVER (TANGENTIAL SENSITIVITY IS SIGNAL STRENGTH-LIMITED)

An RF gain limited receiver is defined to be a receiver whose tangential sensitivity is dominated by the sensitivity of the detector. For example, a crystal video receiver having no RF amplification is an RF gain limited receiver. Even though a receiver may have RF amplification, the receiver sensitivity can still be dominated by the sensitivity of the detector. When the term $AB_V/(G_T F_T)^2$ in Equations 1 and 2 is larger than the term $(2B_R B_V - B_V^2)$, the receiver is considered RF gain limited. The term $AB_V/(G_T F_T)^2$ is frequently omitted in receiver sensitivity calculations and this omission can result in unacceptable errors as shown in the following examples.

Example 1

A crystal video receiver with an RF filter of B_R = 2000 MHz (from 2000 to 4000 MHz), followed by a detector and a 1 MHz bandwidth video amplifier as shown in Figure 4.

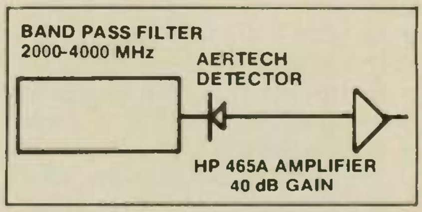

Fig. 4 A simple crystal video receiver.

The filter has a 2.1 dB loss at 3 GHz. The noise figure of this system is 1.62 (2.1 dB) and the gain is 0.62 (-2.1 dB). The measured TSS is -38 dBm. Since there is no data available on this detector, this value is used to determine A.

From Equation 2

$$TSS\ (dB) = -38 = -114 + 2.1 + 10 \log (6.31 + 2.5\sqrt{4000 - 1 + A})$$

one can obtain $A = 9.64 \times 10^{13}$

or from Equation 7

$$-(38 + 2.1) = -110 + 10 \log A$$

one obtains $A = 9.55 \times 10^{13}$

The difference in A reflects only .04 dB in tangential sensitivity which is caused by neglecting B_R, $B_R B_V$ and B_V^2 in Equation 2.

Example 2

An RF amplifier with a gain of 35.3 dB and noise figure of 6.5 dB (Avantek AM-4060M) is added before the detector. The total noise figure is calculated from Equation 4.

$$F_T = F_1 + \frac{F_2 - 1}{G_1} = 1.62 + \frac{3.47}{0.62} = 7.22 \text{ (8.6 dB)}$$

$$G_T = G_1 G_2 = 2089 \text{ (33.2 dB)}$$

$$TSS(dB) = -114 + 8.5 + 10 \log \left(6.31 + 2.5\sqrt{4000 - 1 + \frac{9.64 \times 10^{13}}{(7.22 \times 2089)^2}} \right) = -73.3 \text{ dBm}$$

If the A term is neglected, the tangential sensitivity will be calculated as -83.2 dBm. The error introduced will be 10 dB which is unacceptable.

Example 3

The RF amplifier in Example 2 is changed to an Avantek AM-4000M with a gain of 30.6 dB and noise figure of 6.5 dB. The total gain is 28.5 dB and noise figure 8.5 dB. The TSS calculated from Equation 2 is -68.6 dBm.

In all three examples $\frac{AB_V}{(G_F F_T)^2}$ is much greater than $2B_R B_V - B_V^2$ and $6.31 B_V$. It is the dominant factor in the term 10 log

$$\left(6.31 B_V + 2.5\sqrt{2B_R B_V - B_V^2 + \frac{AB_V}{(G_T F_T)^2}} \right)$$

of Equation 2. In other words, the video characteristics of the diode and video amplifier dominate the tangential sensitivity of the receiver.

NOISE LIMITED SENSITIVITY

In modern EW receiver designs, adequate RF gain is provided, that the sensitivity of the receiver is not limited by the input signal but rather by the noise of the receiver. If

$$\frac{AB_V}{(G_T F_T)^2} < 0.2\,(2B_R B_V - B_V^2) \text{ or}$$

$$G_T > \frac{2.24}{F_T}\sqrt{A/2(B_R - B_V)} \qquad (9)$$

then the term $\frac{AB_V}{(G_T F_T)^2}$ can be neglected in Equations 1 and 2, and the error induced will be less than 0.4 dB. Equation 9 should be satisfied if the receiver is properly designed for high sensitivity. It should be noted that the characteristics of the detector and video amplifier no longer affect the tangential sensitivity of the receiver. Two examples are given below.

Example 4

An instantaneous frequency measurement (IFM) receiver is used in this calculation with the following parameters:

B_R = 2000 MHz (2000 - 4000 MHz)
B_V = 1 MHz
G_T = 51 dB (1.259×10^5)
F_T = 9.7 dB (9.33)

The condition in Equation 9 is fulfilled in this design, based on A = 10^{13} which is a very conservative estimation of -40 dBm tangential sensitivity for the diode. The tangential sensitivity of the receiver is (by Equation 2b)

$$TSS\,(dB) = -114 + 10 \log F_T + 10 \log (6.31 + 2.5\sqrt{4000 - 1}) = -82.1 \text{ dBm}$$

Example 5

A channelized receiver with many parallel channels, each channel has the following parameters:

B_R = 10 MHz
B_V = 10 MHz
G_T = 86.7 dB (4.677×10^8)
F_T = 15.9 dB (38.9)

Again, Equation 9 is fulfilled and the tangential sensitivity of the receiver is given by Equation 1b

$$TSS(dB) = -114 + 10 \log F_T + 10 \log (31.5 + 2.5\sqrt{200 - 100}) = -78.7 \text{ dBm}$$

EFFECTIVE BANDWIDTH

In a receiver, it is highly desirable to use one bandwidth to calculate the sensitivity. This bandwidth is generally referred to as the effective bandwidth. However, from Equation 1b and 2b, it is obvious that an effective bandwidth cannot be easily defined. It is only when the receiver is noise limited and the RF bandwidth B_R is much greater than the video bandwidth B_V, then Equation 2b can be written as

$$TSS(dB) = -114 + 10 \log F_T + 10 \log (2.5\sqrt{2B_R B_V}) = -114 + 10 \log F_T + 4 + 10 \log \sqrt{2B_R B_V}$$

$$\text{with } 2B_R B_V - B_V^2 >> \frac{AB_V}{(G_T F_T)^2} \text{ and } B_R >> B_V \qquad (10)$$

The 4 dB in Equation 10 can be regarded as the input signal to noise ratio required to produce

[Continued on page 102]

[From page 101]

TABLE 1

EXAMPLE	B_R(MHz)	B_V(MHz)	G_T(dB)	F_T(dB)	TSS Cal (dBm)	TSS Meas (dBm)
*1	2000	1	-2.1	2.1		-38
2	2000	1	33.2	8.6	-73.3	-73
3	2000	1	28.5	8.6	-68.6	-68
4	2000	1	51	9.7	-82.1	-81
5	10	10	86.7	15.9	-78.7	-79

*Used to determine the value of A

the tangential sensitivity of the receiver.[7,8] The $\sqrt{2B_RB_v}$ can then regarded as the effective bandwidth which agrees with Klipper's results.[1]

EXPERIMENTAL RESULTS

The results obtained from the five examples are listed in Table 1. Experimental results with similar conditions were obtained in the laboratory. The calculated and experimental results agreed very well. This confirms that equations used in this paper to calculate TSS are valid.

CONCLUSION

In order to find the tangential sensitivity of a receiver, the following steps are followed:
1) The value of A corresponding to the diode detector must be determined either from the data sheet or through actual measurement; 2) Find the gain G_T and noise figure F_T of the receiver to determine whether the tangential sensitivity of the receiver is signal strength limited. Most modern receivers are noise limited; 3) Depending on the relation between B_R and B_V, the respective equation can be used to calculate the tangential sensitivity; 4) The effective bandwidth is a valid definition only the $B_R >> B_V$ and the receiver sensitivity is noise limited; and 5) It must be noted that increasing the input signal by a certain amount does not improve the output signal to noise ratio by the same amount. The output noise will increase when the input signal increases, because the output noise contains a signal-noise cross product term.[5,6] For example, if an output S/N = 14 db is required for a certain probability of detection, increasing the input signal by 6 dB above the tangential sensitivity does not necessarily increase the output signal to noise to 14 dB (6 + 8). The increase should be slightly higher than 6 dB.

ACKNOWLEDGEMENTS

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Monolithic Low Noise Amplifiers

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A GaAs monolithic low noise X-band amplifier has been developed which demonstrates a noise figure of 2.8 dB with an associated gain of 8.5 dB at 12 GHz. The circuits were formed using direct ion implantation into semi-insulating GaAs, which is the technology of choice for low cost, high volume applications.

In the last five years we have witnessed the rapid growth of GaAs monolithic integrated circuit technology.[1-3] The bulk of this effort in the U.S. has been devoted to the development of X-band transmit-receive (T-R) modules for phased array radar applications and power amplifiers, phase shifters and switches have been demonstrated. In Europe, however, the major emphasis has been on the development of monolithic direct broadcast television ground stations.

Current plans in many countries call for the direct broadcast of television signals to homes via geostationary satellites at 12 GHz. Japan and several European countries are planning to implement such systems in two to four years and an experimental system could be in operation in the U.S. as early as 1982.

Central to this planned system is the need for large numbers, certainly many millions, of high quality, inexpensive earth terminals available for purchase or lease to the public. Marketing research has shown that an installed cost of 500-1000 dollars for a system consisting of a one meter antenna, an outdoor receiver and an indoor receiver/demodulator, is necessary in order to attract a sufficient viewing audience to make the system economically feasible.

Current receiver technologies include hybrid FET amplifier front ends[4] and low noise, image enhanced mixer front ends[5,6]. Estimates of large scale production costs for these systems appear to place them out of contention for the mass home market, except during the early stages of system development when smaller quantities and higher prices could be tolerated.

A monolithic receiver appears to be ideally suited for this application, combining the cost benefits of batch fabrication with the RF performance obtainable with GaAs FETs. The monolithic low noise amplifier is a key component in both systems and is in many ways a more straightforward, lower risk component than many other monolithic elements. It does not have a significant thermal problem and the low noise FET has a higher impedance and is, therefore, easier to match than a power device. It also has a relatively low gate periphery per chip (600μm for a two stage amplifier) and its yield in production should be high.

The noise figure of the monolithic amplifiers already developed is sufficiently low to satisfy the requirements of both of the above systems and the fabrication techniques used are compatible with high volume, low cost production. It appears that limited production of these and similar devices is possible in 1982.

MONOLITHIC INTEGRATED CIRCUITS

Monolithic circuits consists of FETs, Schottky barrier diodes, lumped and distributed RF passive components, and bias circuitry integrated onto a GaAs substrate. Figure 1 gives a cross sectional view of a monolithic circuit. The passive elements, consisting of interdigital and overlay capacitors, resistors and lengths of transmission lines, are formed directly on the semi-insulating GaAs. Active device areas are formed by ion implanting the entire surface and then etching to form mesas. A micro-strip layout has been chosen due to field confinement and topological considerations. Inductors are formed from short lengths of high impedance microstrip line. Grounding is accomplished using ribbon bonding over

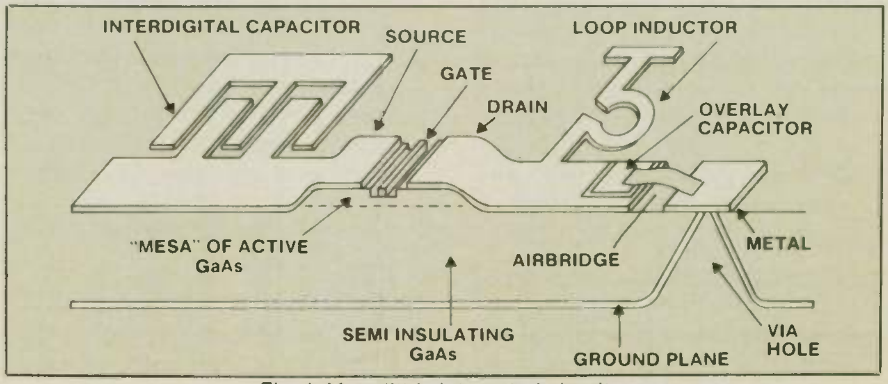

Fig. 1 Monolithic integrated circuit.

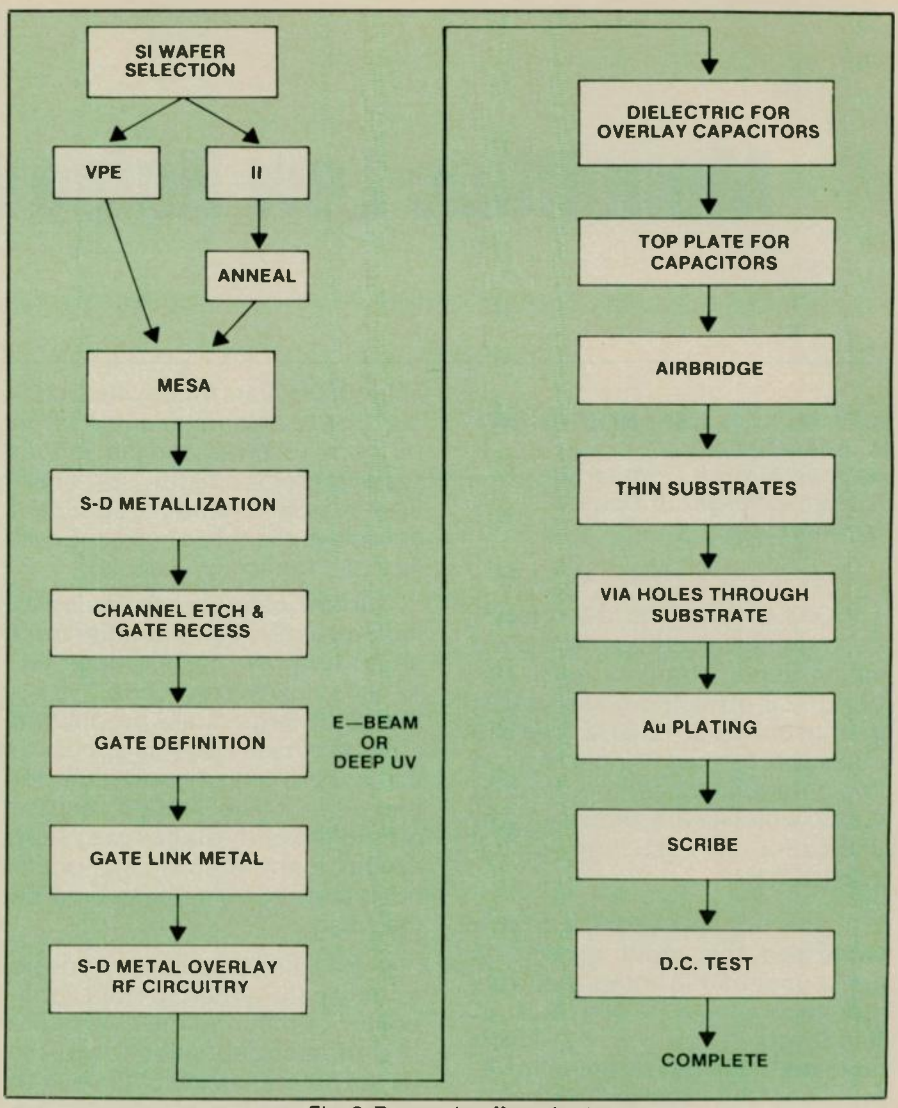

Fig. 2 Processing flow chart.

the chip edge or by via holes through the substrate. Solder and epoxy connections to components are eliminated and wire bonds are needed only at chip boundaries, thereby increasing reliability considerably and avoiding the labor intensive construction techniques used for hybrid circuits.

The details and benefits of monolithic circuitry have been discussed at great length in recent publications and will not be pursued here, but the difficulties inherent to the technology are, occasionally, lost in the enthusiasm. Included among these are:

- Before monolithic IC's can be produced large quantities, the material problems associated with pulling and achieving repeatable wafers on a boule after boule, month after month basis must be thoroughly solved. Preliminary results from LEC (Liquid Encapsulated Czochralski) high pressure crystal pullers, being studied at various laboratories, are encouraging, but it could be some time before the problems of background impurities and defects and their redistribution during annealing are finally understood.
- The state of the art in the processing of GaAs is such that the bulk of the work is done by hand using highly trained technicians and careful inspections between steps. To achieve the volume that some system designers say will be required in the not too distant future, automated batch processing equipment similar to that now in use on silicon wafers will need to be developed to handle the fragile GaAs wafer.
- Because they are small and batch processed, there will be little or no provision for tuning the circuits. Thus, the active devices and passive tuning elements must be well characterized and repeatable over any given wafer and from wafer to wafer.
- Although they are smaller than their hybrid counterparts, monolithic circuits are much larger than their active device area and the number of circuits per wafer will be correspondingly lower. In order to reduce the cost of the individual circuit, it will be necessary to reduce its size, increase the overall processing yield and increase the size of the GaAs wafer.
- The microwave Q of passive monolithic lumped components averages about 50 at 10 GHz. This figure compares to approximately 300 for a microstrip based hybrid circuit and several thousand for typical waveguide construction. Although adequate for the majority of tuning and bypass operations, this lower Q must be carefully accounted for in the design process.

MONOLITHIC FABRICATION

The processes necessary for the fabrication of the low noise amplifiers are outlined in Figure 2. The first step, wafer selection, is probably the most important and one of the most difficult in the sequence. There exist a large variety of tests one can make on GaAs, including mobility, light sensitivity, resistivity, activation of implants, etc., but it is difficult to correlate these data with the noise figure of a finished FET fabricated from that material. There are a number of GaAs, from a variety of sources, which from all D.C. tests appear to be good, yet which produce remarkably mediocre discrete devices. There are also a smaller number of boules which, when processed in an identical manner, produce near state of the art microwave results. For this reason, the wafer selection process includes the fabrication and RF testing of 300xl μm low noise FETs from every boule under consideration.

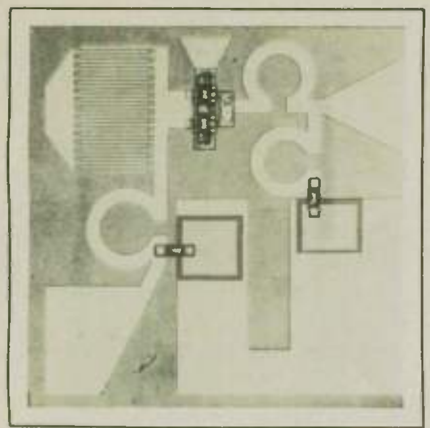

Fig. 3 Semi-lumped version of monolithic LNA.

It has been mentioned that the amplifiers use ion implanted active layers. In order to achieve uniformity, high throughput and low cost, an ion implanted technology is essential. VPE may be useful in the near term, especially for power amplifiers, when production levels are low but it does

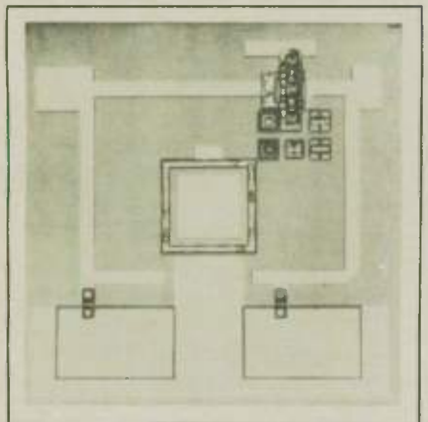

Fig. 4 Microstrip version of LNA.

not appear practical for the boule per week production levels which are projected for many system applications.

Once the active layer is formed, mesas are etched on which to form the FETs, and source-drain metal is deposited and alloyed to form ohmic contacts. The gate is aluminum and is deeply recessed to minimize parasitic resistance. The gates are defined either optically, giving 0.6 to 0.7 μm long gates, or using direct write E-beam fabrication, giving 0.5 μm or shorter gates. A multi-layer gate link is used to connect the Al gates to the gold circuitry. This link is omitted from test wafers to speed processing, but is essential for long term, reliable operation.

The source-drain overlay metal, deposited next through a photoresist aluminum mask[7] to a thickness of 1.5 μm, forms the RF circuitry and the bottom plates of the overlay capacitors. Silicon dioxide is RF sputtered, 2000°A thick, to form the dielectric layer and this is followed by a top metal layer of Cr-Au. Airbridges are then formed to interconnect the sources and to contact the top plates of capacitors which are formed as isolated pads. This greatly increases the breakdown voltage and yield of the devices. The wafers are then thinned to 200 μm and back metallized. This thickness was chosen instead of the standard 100 μm thickness to take advantage of the higher Q available with the thicker substrate. When this is done it is however, not possible use via holes for grounding since their yield drops rapidly when the substrate becomes greater than 100-125 μm thick, and grounding must be obtained over the edge of the substrate using wire bonds or ribbons.

RF RESULTS

Two versions of a single stage low noise amplifier[8] were fabricated and are shown in Figures 3 and 4. Each amplifier measures 1.27x1.27x0.20 mm and over 500 of them are contained on a single, small GaAs wafer. The amplifiers were fabricated by ion implantation directly into a semi-insulating substrate. Optical lithography was used throughout except for the gates which have been formed both using direct write E-beam and optically. The semi-lumped amplifier, shown in Figure 2, consists of a 300 μm FET, interdigital capacitors and microstrip transmission lines for RF tuning and overlay capacitors for RF bypass. The distributed amplifier, shown in Figure 4, consists of a 300 μm FET, microstrip transmission lines and overlay capacitors. The square in the center is a capacitor for C-V profiling.

Several wafers have been processed on both ion implanted and VPE material, with the best results being obtained using direct implantation of singly ionized ^{28}Si (100 KeV, 6E12) into semi-insulating Sumitomo Cr-doped GaAs. The wafers were annealed at 860°C with an SiO_2 cap and arsine overpressure for 20 minutes. The frequency response of the microstrip matched amplifier is shown in Figure 5. The amplifier, with no external tuning, exhibits 7.3± 0.5 dB gain from 10.6 to 13.7 GHz and 7.6 ± 0.1 dB across the 11.7 to 12.2 GHz band with a noise figure

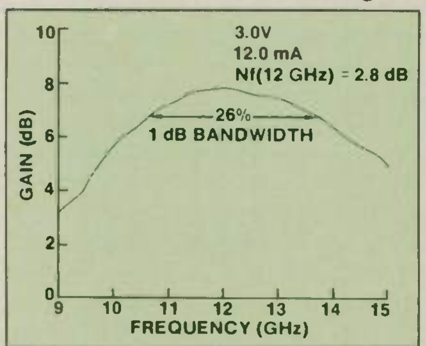

Fig. 5 Frequency response of single stage amplifier.

of 2.88 dB. Figure 6 shows an expanded curve of gain and noise figure over the 11.7-12.2 GHz frequency range. The input VSWR is 3:1 by design in order to achieve a noise match. The output VSWR, however, was also 3:1 over the 11.7-12.2 GHz band. With a slight adjustment of the drain tuning, 9.0 dB gain and 2.8 dB noise figure were achieved at 12 GHz. The drain tuning was necessary because the amplifier was designed using the S-parameters of a discrete FET fabricated on VPE, but the above amplifier was built on ion implanted material. Prelimi-

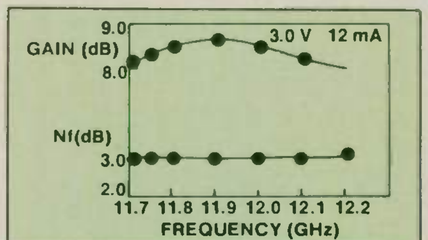

Fig. 6 Gain and noise figure vs. frequency.

nary results on amplifiers fabricated on VPE material showed a well matched output, 1.1:1 VSWR at 12 GHz, but lower gain and higher noise than the ion implanted circuits. More VPE wafers are being processed.

A two-stage amplifier has been designed and masks are being fabricated. The second stage will run at a higher current (20 mA) since, as is shown in Figure 7, an

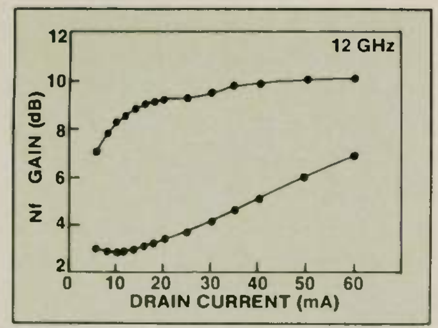

Fig. 7 Gain and noise figure vs. drain current.

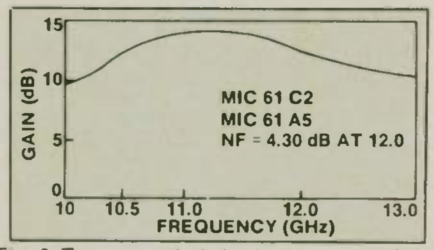

Fig. 8 Two cascaded single stage amplifiers.

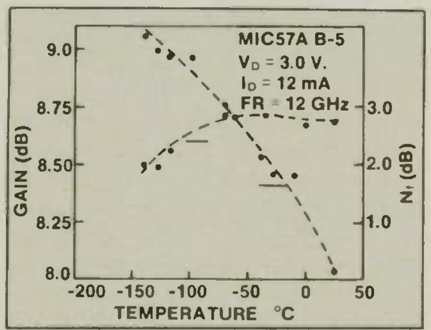

Fig. 9 Noise figure and gain vs. temperature.

extra dB of gain can be obtained for a very nominal increase in overall noise figure. As a test, two of the single stage amplifiers were directly cascaded, a distributed first stage followed by a semi-lumped amplifier. The resultant performance is shown in Figure 8.

Fourteen dB of gain with a 1-dB bandwidth of 1 GHz was achieved, centered at 11.2 GHz. The input reflection coefficient was unchanged from the single stage results, but the output reflection coefficient was near unity due to the mismatched first stage drain circuit. Overall noise figure was 4.3 dB which was within 0.1 dB of the value predicted from the individual measurements on the amplifiers. The predicted performance for the two stage amplifier being developed is 17 dB gain and 3.1 dB noise figure.

A subject of increasing interest is the operation of FETs and amplifiers at reduced temperature. Figure 9 shows the gain and noise figure of a monolithic amplifier cooled to —150°C. The lowest noise figure achieved was 1.95 dB with an associated gain of 9.0 dB. A redesign of the test fixture is necessary to cool the amplifier further.

MONOLITHIC RECEIVER

A direct broadcast receiver covering the 11.7—12.2 GHz band is being designed[9] which will use a combination of monolithic and standard hybrid components. Figure 10 shows a block diagram of the receiver, the current design philosophy of which is to use monolithic fabrication techniques wherever practical, and to use standard components where they are needed. The current configuration consists of a monolithic two stage preamplifier, bandpass filter and mixer, and a hybrid FET local oscillator with a dielectric resonator, and a silicon IC IF amplifier. A two pole bandpass filter, utilizing lumped element construction has been fabricated for this application. As illustrated in Figure 11, the filter consists of interdigital capacitors[10] and microstrip elements and is 0.048 x 0.019 inches in size. This is an order of magnitude smaller than an equivalent distributed circuit. The filter has 2.2 dB of loss over the 11.7-12.2 GHz band and 25-30 dB of rejection over the 8.9-9.5 GHz image band.

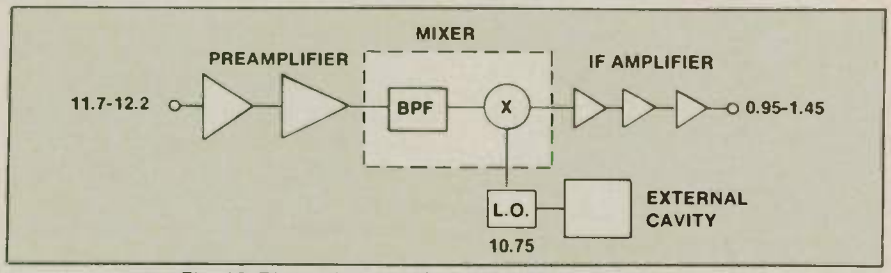

Fig. 10 Block diagram of monolithic/hybrid receiver

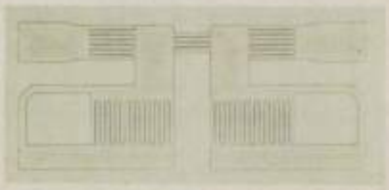
Fig. 11 Two pole monolithic filter.

CONCLUSION

Monolithic amplifiers with a noise figure low enough to satisfy a broad range of applications have been fabricated using direct ion implanted techniques. Several varieties of amplifiers are being fabricated or designed, including single ended and balanced units covering different frequency ranges within X-band.

ACKNOWLEDGEMENTS

The authors wish to acknowledge the work of the members of Hughes Torrance Research Center who have contributed to this effort, particularly L.H. Hackett for E-beam fabrication, L. Cochran for processing support, H. Yamasaki for device design, J.M. Schellenberg for device characterization, S.P. Kwok and W.F. Marx for processing development, S. McMillen for typing and T.A. Midford for continuing support and encouragement.

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A Two Stage 3.7 to 4.2 GHz GaAs FET LNA Using Two NE21889's or Two NE72089's

SHIGERU SANDO
CALIFORNIA EASTERN
LABORATORIES, INC.

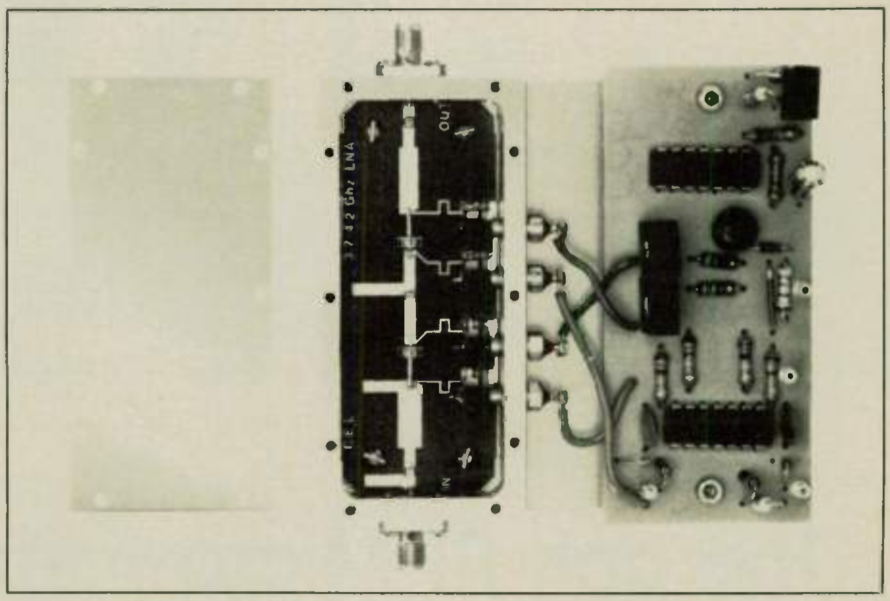


Fig. 1 Low Noise Amplifier and Power Supply Showing details of component placement and transistor mounting.

INTRODUCTION

The amplifier shown in Figure 1 is constructed using a 0.8 mm glass-teflon circuit board, three 10 pF capacitors, four 1000 pF capacitors, four 1500 pF feed-through capacitors, four ferrite beads and a housing. The typical noise figure is 1.2 to 1.7 dB in the 3.7 to 4.2 GHz frequency band as shown in Figure 2. The gain and input return loss is shown in Figures 3 and 4. The power supply described in the CEL Application Note, "Two Stage GaAs FET LNA Bias Supply" (AN80901) is recommended to bias this amplifier.

CIRCUIT DESIGN

The NE21889 and NE72089 noise parameters at 4.2 GHz with V_{DS} = 3 V and I_D = 10 mA are:

$$NF_{opt} = 0.9 \text{ dB}$$
$$G_a = 13 \text{ dB}$$
$$\Gamma_{in} = 0.68$$
$$R_n/50 = 0.39$$

The first stage input matching circuit uses a quarter wavelength transformer and a series inductor to match the input impedance to the optimum source impedance. The quarter wavelength transformer is 33 ohms in order to transform the input 50 ohms to 21 ohms, and the package gate lead is used to series resonate the FET input impedance. The interstage matching filter is less complicated be-

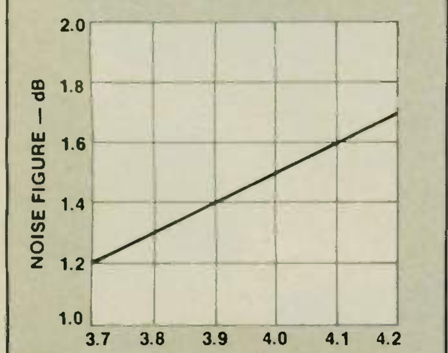


Fig. 2 Noise Figure performance of amplifier.

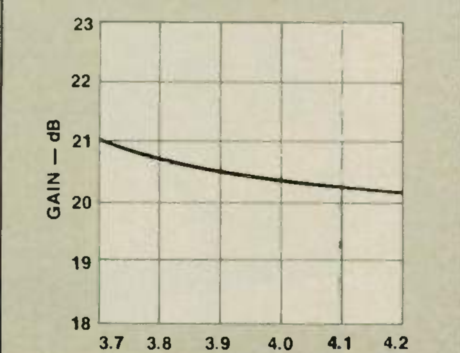


Fig. 3 Gain performance of amplifier.

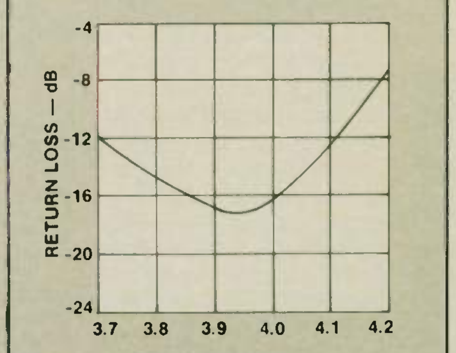
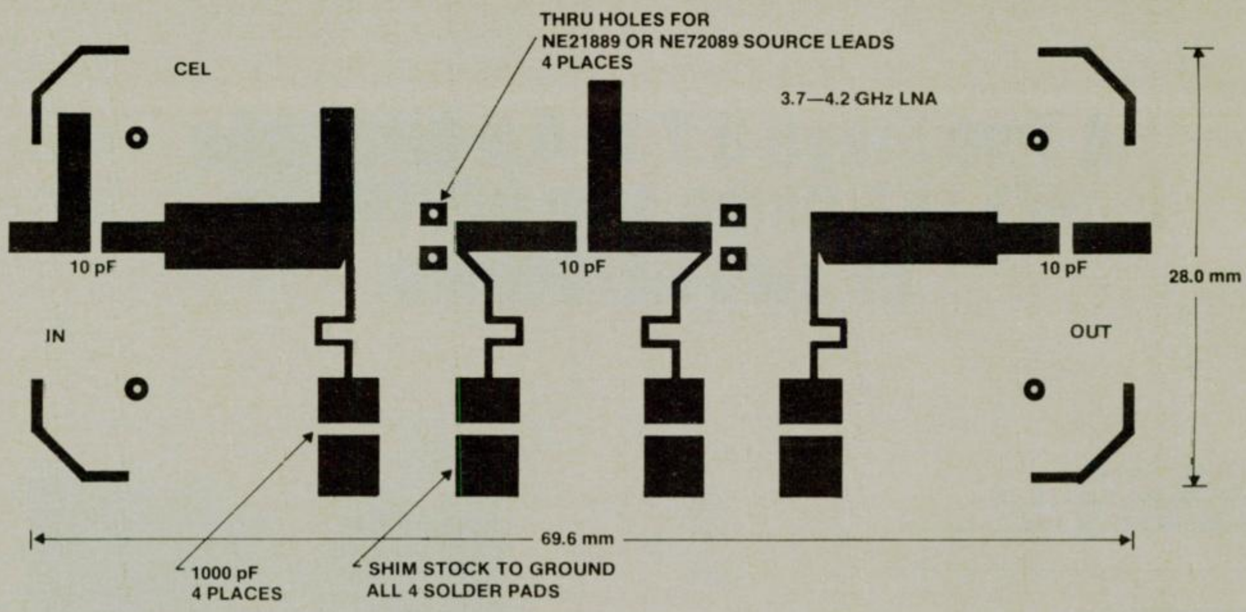


Fig. 4 Return Loss performance of amplifier.



Parts List: NE21889 or NE72089 FET (2 ea.) California Eastern Laboratories
10 pF Chip Capacitors (3 ea.) ATC #100A-100-J-P-X-50 or equivalent
1000 pF Chip Capacitors (4 ea.) Johanson #50R11W102KP or equivalent
1500 pF Feed-through Capacitors (4 ea.) Erie#2425-003-W5U0152AA or equivalent
Ferrite Beads (4 ea.) Fairrite #2643001301 or equivalent

Fig. 5 P.C. Board Layout.

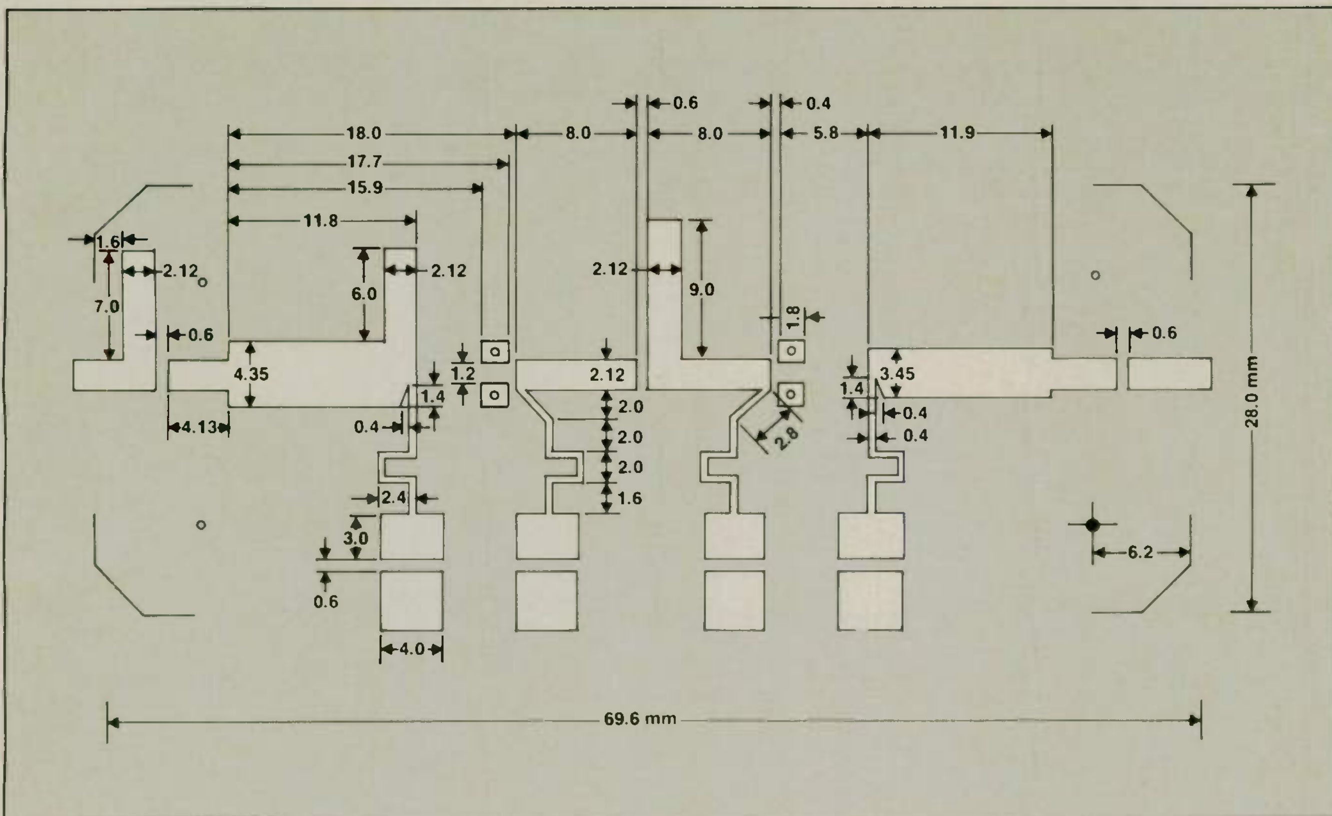


Fig. 6 P.C. Board Dimensions. (Units in mm)

[*Continued on page 110*]

[From page 109] NE21889'S

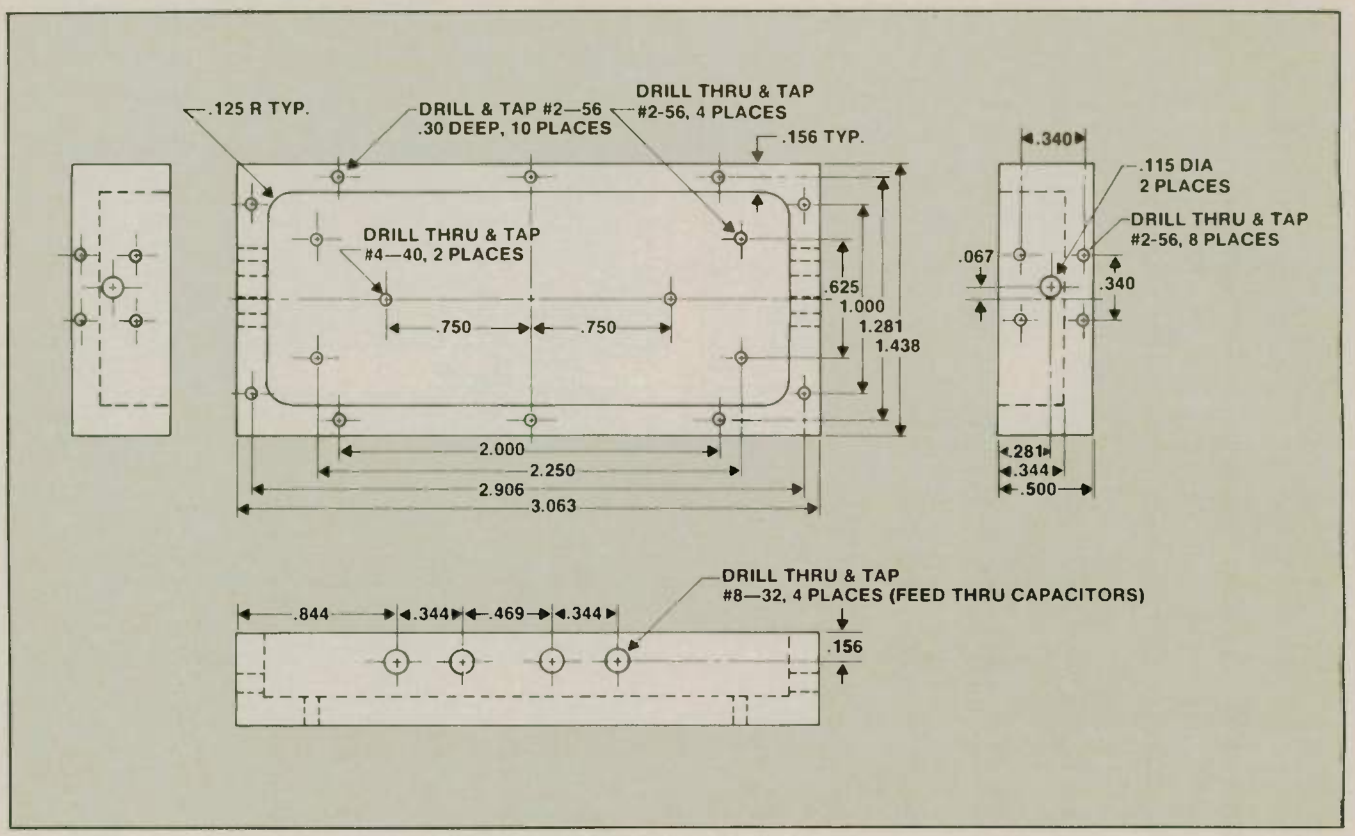


Fig. 7 Amplifier Chassis. (Units in inches)

cause the optimum source impedance and the load reflection coefficients are quite similar. Thus, a quarter wavelength 50 ohm transmission line can be used to match the first stage output impedance to the second stage input impedance.

The second stage output impedance is series resonated using the package led for an inductor. This is followed by a quarter wavelength 40 ohm transmission line which transforms the output stage series resonated load impedance of 30 ohms to 50 ohms. Open circuited 50 ohm transmission line stubs are empirically added to the circuit; two at the input and one at the interstage to compensate for the FET package source lead inductance and the blocking capacitor parasitics.

CONSTRUCTION

Set up a work area with a grounded work surface, i.e., a sheet of metal, copper clad P.C. board material, or conductive plastic. Use a grounded soldering iron of 18 to 27 watts. Never use a soldering gun.

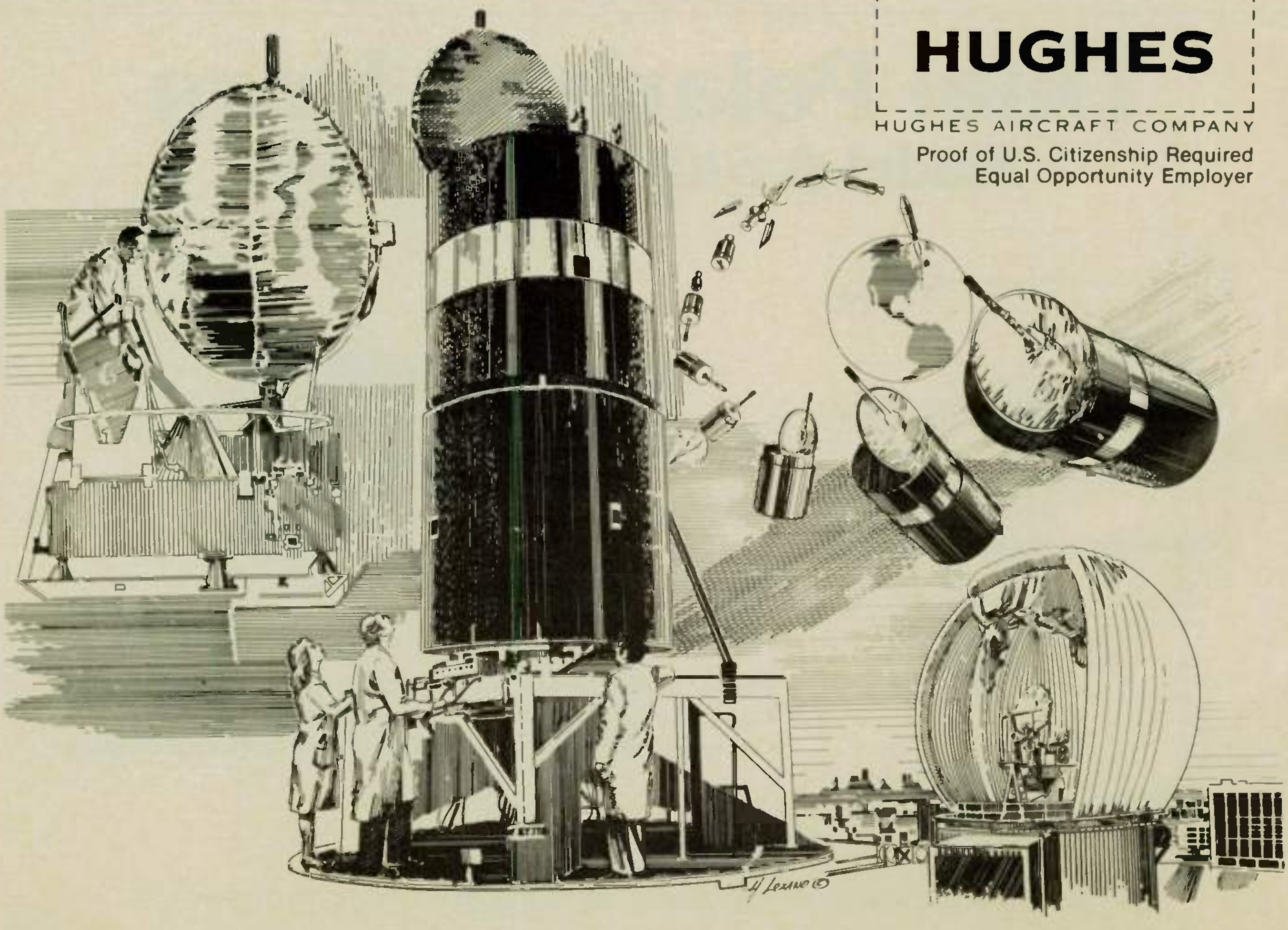
Wrap .001/.002 thick brass shim stock around the lower edge of the P.C. board, as shown in Figure 5, to create a low inductance path to ground for the bias circuit chip bypass capacitors. Install the three 10 pF DC blocking chip capacitors and the four 1000 pF RF bypass chip capacitors using a minimum amount of solder. Solder 4 wires (24/26 gage) to the bias line pads.

Provide a good ground by using a small chain or non-insulated flexible wire wrapped around the assembler's bare wrist. Take the first stage FET out of its shipping package and place on the grounded work surface. Form the source leads down so that they will fit through the two holes in the circuit board. Insert the FET into the circuit board with the gate lead toward the input end of the board (slashed lead). With the device seated firmly against the PC board, form the source leads flat against the back side of the PC board and solder them down. Trim the drain lead, fitting it to the very end of the input strip line and then solder both the drain and the gate lead.

Take the second FET and insert it into the PC board with the gate toward the input, using the same procedure as with the first FET. Solder the source leads. Next, trim the gate lead, fitting it to the very end of the output strip line, and then solder both the gate and drain lead.

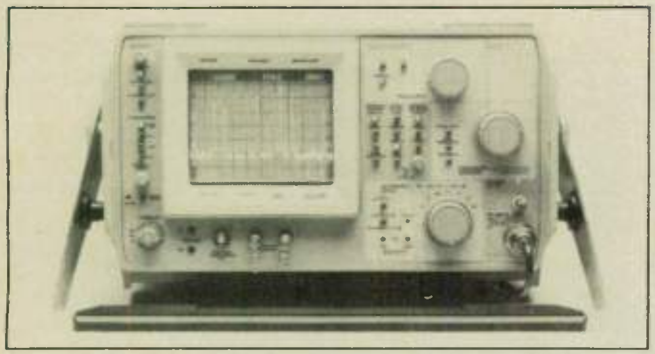
Install the board into the chassis, attach the connectors and solder them to the amplifier input and output transmission lines. See Figure 7. Solder the bias wires to the feed through caps on the chassis side after putting ferrite beads on the feed-through leads. Use a meter to set the V_{DS} to 3 volts and the I_{DS} to 10 mA. Never touch the gate connection with the meter leads as the transistors are prone to shorting out due to static discharge. ■

 CIRCLE 73 ON READER SERVICE CARD



New Spectrum Analyzers for 1 KHz to 1.8 GHz Range

TEKTRONIX, INC.,
Beaverton, OR



The 496 and programmable GPIB compatible 496P Spectrum Analyzers now available from Tektronix, Inc., are similar to the popular 492/492P Spectrum Analyzers in appearance, operation, and environmental specifications. While the 492/492P Spectrum Analyzers cover the RF spectrum from 50 KHz to 220 GHz; the new 496/ 496P Spectrum Analyzers cover frequencies from 1 KHz to 1.8 GHz.

With the 496/496P, resolution bandwidth can be varied from 1 MHz to 30 Hz over the entire frequency range. Automatic phase lock stabilization reduces incidental FM to 10 Hz p-p; phase noise sidebands are at least -75 dBc at 30 times the resolution offset. Frequency drift is no greater than 330 Hz in 10 minutes; and there is 30 Hz resolution bandwidth with -127 dBm sensitivity.

Digital storage eliminates time-consuming display adjustments. Constant tuning rate lets users position the signal quickly and accurately at any frequency span.

Microprocessor-aided functions automatically couple most frequently used controls, permitting true three knob operation.

The 496P is the fully programmable and GPIB compatible version of the 496 Spectrum Analyzer. Full programmability allows the user to operate the 496P under program control, change the front panel settings, read data from the CRT display, and send spectral waveforms from the internal digital source memory to other GPIB devices.

Prices: 496 — $22,950; 496P — $26, 950. Availability: 2 weeks.

Marketing Communications Department TEKTRONIX, INC., Beaverton, OR (800) 547-6711, Oregon, 644-9051. ■

CIRCLE 75 ON READER SERVICE CARD

[From page 61] ARRAY RADARS

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Eli Brookner *received the BEE degree from the City College of New York and the MS and DSc degrees in electrical engineering from Columbia University. Eli is presently a consulting scientist with the Raytheon Company Equipment Division. He conceived and helped design the Wake Measurement Radar; the first TWT radar put into space; has been involved in a number of space-based radar studies; and has been a technical consultant to COBRA DANE, Seasparrow, SIRCS, Milirad, Hardsite, COBRA JUDY and PAVE PAWS radars.* ■

Microwave Products

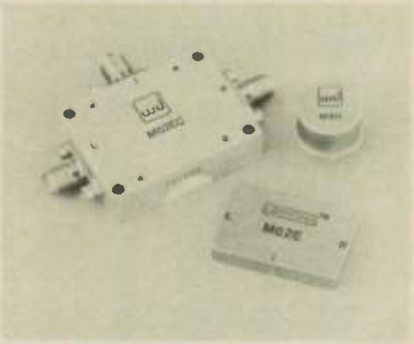
Components

STANDARD AND LOW SWR ADAPTORS

Type N and SMA series of standard and low VSWR waveguide-to-coaxial adaptors provide typical maximum SWR's of 1.25 for standard type N from 1.12 to 15 GHz and SMA from 2.6 to 18 GHz; 1.065 for low SWR type N from 2.6 to 15 GHz and SMA from 2.6 to 18 GHz. A double ridged adaptor for 7.50 to 18 GHz offers a maximum SWR of 1.3. The adaptors are available in brass or aluminum with stainless steel connectors. **Microwave Development Laboratories, Inc., Natick, MA. Ernest Bannister (617) 655-0060.**

Circle 159.

LOW COST TVRO MIXERS



Three 3.7 to 4.2GHz double balanced mixers feature IF coverage to 1125MHz. Model WJ-M8H in a TO-8B can is designed for direct mounting on a printed circuit board. Model WJ-M62E is for MIC integration and has a typical conversion loss of less than 5 dB and Model WJ-M62EC offers the same performance with SMA connectors. Price @100+: WJ-M8H and WJ-M62E: $59 each, WJ-M62EC: $106. Delivery: stock. **Watkins-Johnson Co, Palo Alto, CA. S. B. Witmer (415) 493-4141.**

Circle 147.

Microwave Products

PIN SWITCH DRIVER

Model 207cc driver features output current levels adjustable to 70 mA and switching speeds of 5 nanoseconds typical with extra current spiking for quick recovery of PIN switching diodes. Operating from +5 and -12V supplies, the unit provides an inverting output from TTL input for series, shunt and series-shunt PIN diode switches. The driver is contained in a .48" x .48" leadless chip carrier. Delivery: stock to 2 weeks. **New England Microwave Corp., Hudson, NH. Lucielle Robbins (603) 883-2900.**

Circle 155.

50 WATT AMPLIFIER COVERS 10 KHz to 250 MHz

Model M5360 solid state amplifier offers maximum gain of 50 dB and a variation of less than ± 1.5 dB over the frequency range of 10 KHz to 250 MHz. Typical harmonic distortion is -35dB at full output. Standard features of the M5360 are electronic gain control (manual or remote control) of 40 db and full protection against overvoltage in power supply and RF stages. The amplifier measures 21" x 22¼" x 15 3/4". **Instruments For Industry, Farmingdale, NY. Ronald Richards (516) 694-1414.**

Circle 156.

WIDEBAND GaAs FET AMPLIFIERS

Models A55G-6206 and A55I-7247 GaAs FET amplifiers cover 2 to 6 GHz and 6 to 12 GHz respectively with an output power of 100mW at 1 dB compression. The amplifiers have a gain level greater than 42 dB, with better than ±2.0dB flatness. Third order intercept on both units is above ±30dBm. Internal regulation allows for operation from ±15V or ±18V. Construction employs thin film hybrid technology. Delivery: 90 days ARO. **Aertech Industries, Sunnydale, CA. W. S. Patton (408) 732-0880.**

Circle 157.

POWER DIVIDER COVERS 1-18GHz RANGE

Model D301M isolated stripline power divider covers the 1-18GHz frequency range. The unit has a maximum phase variation between ports of 4° over the full band with a typical variation of less than 1°. Typical isolation figures over the full frequency range are 15 dB and typical insertion loss and SWR figures are 0.5 dB and 1.4. The unit has an internal resistor with a power rating of 500 mW CW. Model D301M measures 1.6 25" x 0.5" x 0.38" plus SMA female connectors. Price: $95. Delivery: stock. **Engelmann Microwave Company, Booton, NJ. Carl Schraufnagl (201) 334-5700.** **Circle 142.**

18 GHz TYPE N TO SMA ADAPTORS

Type N to SMA 18 GHz between-series adaptors feature stainless steel bodies and exhibit a SWR of less than 1.20 to 18 GHz. Price: 100 quantity less than $10.00. **United Microwave Products, Inc., Torrance, CA. Ed Jacobs (213) 320-1244.** **Circle 154.**

DIRECTIONAL COUPLER COVERS .5 — 18 GHz

Model 1851 directional coupler covers the frequency range from .5 to 18 GHz. Coupling (with respect to output) is 10 ± 1 dB nominal with less than ±1 dB ripple. Directivity is greater than 14 dB from .5-12.4 GHz and greater than 12 dB from 12.4-18 GHz. Maximum VSWR is 1.25 from .5-8GHz and 1.35 from 8-18 GHz. Insertion loss is less than 1.5 dB. Power rating (input) is 20W average and 3kW peak. Standard connectors are female N or female type SMA. Price: $575. Delivery: 4 weeks. **Krytar, Sunnyvale, CA. (408) 734-5999.**

Circle 167

HERMETIC SEAL SMA CONNECTORS

OSCS Series of SMA connectors have a VSWR rated at 1.04 + .009f (GHz) to 18 GHz. Laser welded contacts and brazed metal-ceramic seals allow exposure to 300° C for short periods. Maximum leak rate is 1 x 10^{-8} cc/sec. Price: $5.97. Delivery: stock to 8 weeks. **Omni Spectra, Merrimack, NH. John Callahan, (603) 424-4111.**

Circle 152.

PHASE SHIFTER COVERS 6.0-18 GHz

Model PQ-1122 is a 360° analog phase shifter that covers the 6 to 18 GHz frequency band with a maximum insertion loss of 15 dB. Maximum amplitude ripple is ±3dB, response time of the phase shifter is 50 nanoseconds and the maximum SWR is 2.5. Availability: 14 weeks. **Triangle Microwave, Inc., East Hanover, NJ. Martin Rabinowitz (201) 884-1423.**

Circle 160.

WRD 750 DOUBLE RIDGE CROSSGUIDE COUPLERS

A line of crossguide couplers covers 90% of the 2.4:1 bandwidth of WRD750 double ridge waveguide. The couplers have an input return loss greater than 19 dB with directivity greater than 15 dB. Coupling values from 30 dB and up are presently available with less than 5dB variation with frequency. Dual directional models have arms matching within ± 1 dB for integrating input to reflected power levels. Couplers come with SMA or flange output arms. **Microtech, Cheshire, CT. James McGregor (203) 272-3234.**

Circle 143.

500MHz — 18 GHz CAVITY OSCILLATORS

Series C fundamental and multiplier cavity oscillators cover the frequency range from 500 MHz to 18 GHz with power outputs ranging from 50 mW to 500 mW. Frequency stability is ± 0.05 % in an operating temperature range of - 30°C to +65°C. The series, designed for digital and analog telecommunications, telemetry and radar applications offers options that include: frequency modulation, AFC input, non-translating tuning shafts, rugged band edge stops and phase locking. Delivery: under 60 days. **Communication Techniques, Inc., East Hanover, NJ. G. Marshall (201) 884-2580.**

Circle 144.

Microwave Products

SINGLE SIDEBAND MODULATORS FOR 2 - 18 GHz

A series of miniature single sideband modulators for narrow and full active operation are available with IF frequencies of dc - 500 MHz. The units accomodate L.O. powers from +6 dBm to +12 dBm without bias, conversion loss ranges from 9 to 13 dB maximum and sideband suppression is 18 dB min. in the 6-18 GHz model. Separate models cover 2-6 GHz, 2-8 GHz and 6-18 GHz. Octave IF bandwidths from 20-40 MHz to 100-200 MHz may be selected. Operating temperature range is -5°C to + 95°C. **Norsal Industries, Inc., Central Islip, NY. Norman Spector (516) 234-1200.**

Circle 153.

800 — 8000 MHz POWER DRY LOAD

Model FT2974 is an air-cooled medium power dry load covering 800 to 8000 MHz optimized over the TACAN band. The unit offers and SWR of less than 1.15 from 960 to 1215 MHz and an SWR of less than 1.35 from 800 to 8000 MHz. It has a power rating of 5800 W peak and 210 W average and measures 6.95 x 5.18 x 6.00 inches. **Sage Laboratories, Inc., Natick, MA (617) 653-0844.**

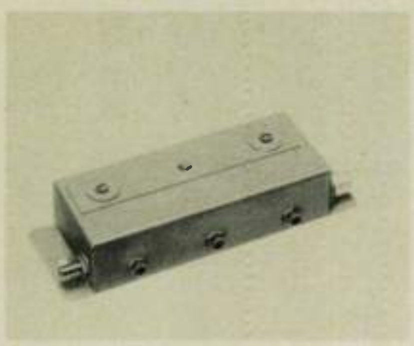
Circle 163.

THIN FILM OSCILLATORS COVER 2.7 TO 4.5 GHz

Models V72T-1 and V72T-2 are thin film voltage-tuned oscillators designed for satcom applications. The V72T series covers the frequency ranges 2.7-3.2 GHz and 2.8-3.3 GHz; the V82T series 3.0-3.5GHz, 3.7-4.2GHz and 4.0-4.5GHz. All offer a minimum of +10 dBm output power and cover their frequency range with maximum tuning voltage of +15 volts. Price: both series - $75.00 (100 quantity). Delivery V72T: 2 to 8 weeks; V82T: stock to 6 weeks. **Magnum Microwave Corp., Sunnyvale, CA. David Fealkoff (408) 738-0600.**

Circle 164.

VHF BANDPASS FILTERS



Series 3131 bandpass filters have a 2 MHz bandwidth and are available for any center frequency from 50 to 300 MHz. Insertion loss is approximately 3 dB with 30 dB rejection at ± 4 MHz. SMA 50 ohm connectors are standard; the filters are also available with 75 ohm type F connectors. Price: $125. Delivery: 1-9 quantity, 1 week. **Microwave Filter Company, East Syracuse, NY. Emily Bostick (800) 448-1666.**

Circle 145.

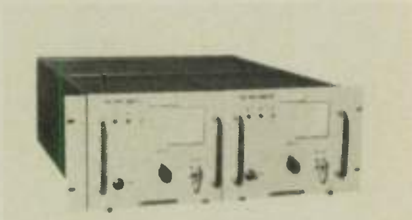
Microwave Products

L AND S BAND AMPLIFIERS COVER UP TO 2300 MHz

Low noise L and S band hybrid amplifiers combine GaAs FETs and silicon bipolar transistors and offer noise figures of 1.1 dB maximum. The four units have 35 dB minimum gain and 1.5 maximum input and output SWR. Model C14114 covers the 1430 MHz to 1540MHz range with a 1 dB noise figure; Model C17114 covers 1650 MHz to 1750 MHz with 1 dB noise figure and Model C22113 covers the 2200MHz to 2300MHz range with a maximum noise figure of 1.1 dB. A broadband amplifier covering the full 1400MHz to 2300MHz band is available with a maximum noise figure of 1.5 dB. **California Amplifier, Inc., Newbury Park, CA. Jacob Inbar (805) 498-2321.**

Circle 158.

CONVERTERS AND FILTERS FOR SATCOM



Series U-4536 converters operate in the 4 GHz down-link frequency band with residual phase noise less than 14 dB. Six options are available including: frequency synthesized HLO Type 1 (in 125 KHz steps, fully automated and applicable for FM-FDM carriers), and Type 2 (in 1 MHz steps with mechanical cavity tuning and noise characteristics suitable for SCPC traffic); external synthesizer input for the converter, group delay equalizer for compensation of group delay distortion produced in the first IF BPF, and up to four isolated and identical outputs. Inband ripple for gain frequency respose and group delay characteristics meet all INTELSAT IV and V requirements. **Mitsubishi Electronics America, Compton, CA. Ric Fochtman (213) 979-6055.**

Circle 146.

SMA TERMINATIONS

SMA terminations in plug and jack configurations cover the frequency range of DC to 18 GHz with a maximum VWSR of 1.05 + .005F(GHz). Average power ratings are 2.0 W at +25° C and 1.0 W at +125° C with a peak power level of 50 W. Plug (Part no. 705467-001) length is .547" and jack (Part no. 705482-001) is .578". Price: 1,000 quan., plug: $8.55. Availability: from stock. **Cablewave Systems Inc., North Haven, CT. Steven Raucci (203) 239-3311.**

Circle 148.

GUNN OSCILLATORS COVER 50-110GHz

A series of Gunn oscillators for millimeter wavelengths covers the 50 to 110 GHz range. Model 4560A covers the 50 to 75 GHz range with a 175 mW maximum power output; Model 4575A, the 60 to 90 GHz range with a maximum output power level of 75 mW and Model 4954C covers 80-110 GHz with 20 mW max. Modulator/Regulator options permit FM (± 75 MHz min) and AM (pulse, square wave) modulation and have built-in 1000 ± 25 Hz square wave modulators. Price: from $2350. Delivery: 30-45 days. **Epsilon Lambda Electronics Corp., Geneva, IL. Robert Knox (312) 232-9611.**

Circle 149.

BROADBAND MIXERS OPERATE OVER 2 - 18 GHz

Model RFX-4 broadband RF mixer operates over the 2 - 18 GHz frequency range and features a 1-7GHz IF frequency range. Typical conversion loss is 7dB with a maximum loss of 8.5 dB. SWR is 2.5 maximum at RF and LO and 2.0 IF maximum. Isolation is 22 dB typical, 18 dB minimum between all ports. LO power range is +14 dBm to +7 dBm and +10 dBm nominal with no burn-out to +22 dBm. Connectors are SMA female and operating temperature range is -54° C to +85° C. **Precision Tuning, Inc., Sunnyvale, CA. James Lautermilch (408) 734-4440.**

Circle 151.

Microwave Products

COAXIAL DETECTORS COVER .01 GHz TO 12.4 GHz

Coaxial crystal detectors cover the frequency range of .01 tc 12.4GHz and have a frequency response within ± 0.5 dB absolute. Relative matching excluding bias sensitivity is within ± 0.2 dB. The detectors have RF type N male connectors and a maximum power rating of 100 mW. **Micronetics, Ins., Norwood, NJ. Gary Simonyan (201) 767-1320.**

Circle 161.

1-2 GHz GaAs FET AMPLIFIERS

The APG-2000 series of GaAs FET amplifiers offer ±30 dBm minimum output power over the 1 to 2 GHz frequency range. The amplifiers feature a choice of 10, 20 or 30 dB gain with as low as ± 0.5 dB fullband gain flatness, 4.5 to 5.0 noise figures and +40 dBm dBm third order intercept point for intermodulation products. All versions operate from a single ± 15 VDC supply, requiring 875 to 975 mA. The amplifiers can be qualified to Mil-E-16400, MIL-E-5400 and MIL-E-4158. Price: $1,000 to $2,000. Delivery: 90 days ARO. **Avantek, Inc., Santa Clara, CA. Northe Osbrink (408) 496-6710.**

Circle 162.

4-8GHz MIXER

Model MD-180 double balanced mixer covers the 4 — 8GHz range with a 4.5 dB typical conversion loss and a 6.5 dB maximum loss. L-R isolation is 27 dB and the mixer is capable of 0 dBm starved L.O. operation without bias. Price: 1-5 quantity $240 for modules and $315 for SMA connectorized models. Availability: from stock. **Anzac Division, Adams-Russell Co., Inc., Burlington, MA. Mark Rosenzweig (617) 273-3333.**

Circle 165.

TO-8 AMPLIFIER PROVIDES +20 dBm UP TO 700 MHz

Model MHT-653 hybrid amplifier operates over the 5 to 700 MHz range with an output power of +20 dBm. Typical gain is 10 dB with a 6.5 dB noise figure and a +34 dBm third order intercept. Maximum SWR (input and output) is 2.0 at 50 ohms, and operating voltage is +15 V at 100 ma. Operating temperatures range from -54° C to 100° C. Delivery: stock to 6 weeks, including screening for MIL-STD-883B. **Aydin Vector Division, Newton PA. Vince Cipriano (215) 968-4271.**

Circle 166.

OCTAVE RANGE COAXIAL BANDPASS FILTERS TO 12 GHz

The series of bandpass filters covers the 1-2, 2-4, 4-8 and 8-12 GHz ranges with 2, 3 or 4 pole selectivity. Model OTF-4080-3I tunes 4 to 8 GHz with three individually tuned resonators. Bandwidth is 30 ± 3 MHz with a maximum insertion loss of 1.3dB and a passband VSWR of less than 1.5 dB. The units are available with dial, knob, or directr reading tape. Prices: from $1700. Delivery: 60 days. **RS Microwave Co., Inc., Bulter, NJ. (201) 492-1207.**

Circle 167.

COAXIAL SWITCHES PERFORM THROUGH 6 GHz

Series 303 coaxial switches offer performance through 6 GHz and are available in both remote failsafe and manual latching versions. The single-pole double throw subminiature switches offer a choice of three terminations: SMA, SubMinax 27 Series or printed circuit contacts. Insertion losses range from .1dB maximum at 3.0 GHz to .3 dB maximum at 6.0 GHz; rated power is 400 W at 0.1 GHz, 45 W at 6 GHz; switching time is 15 milliseconds maximum. Maximum VSWR for the SMA and 27 Series interfaces range from 1.1 at 2 GHz to 1.3 at 6.0 GHz. Minimum isolation ranges from 80 dB at 1.0GHz to 60 dB at 6.0 GHz. Coil voltages of 6, 12, 26 or 48 V are available. Price: $110. in quantities of 10 each. Delivery: 14-16 weeks. **Amphenol North America, Oak Brook, IL. Art Morse (312) 986-2322.**

Circle 168.

Microwave Products

ADAPTOR CONNECTS MICRO-MINIATURE TO SMA

Adaptor (part number 51-475-6201) is a bulkhead mounted transistion between SMA and nanohex RF connectors. The unit has an SMA jack at one end and a nanohex snap-on jack at the other end. There is a threaded section in the body with lock-nut and washer for mounting in a D-flat panel hole. **RF Components Division, Sealectro Corp., Mamaroneck, NY.**

Circle 169.

Instruments

100 MHz PROGRAMMABLE PULSE GENERATOR

The E-H Model 1516 programmable pulse generator provides frequencies from 1 Hz to 100 MHz with fixed rise and fall times of less than 500 picoseconds and full programmable control of other pulse parameters. The unit offers an output amplitude of ± 2.5 volts into 50 ohms on each channel with baseline offset variable in the range of ± 1 volt. Both positive and negative polarities and normal and complement mode are selectable under program control. The standard Model 1516 is supplied with rack mount flanges. A number of program bus options are available. **E-H International Inc., Oakland, CA. (415) 638-5656.**

Circle 170.

-125 dBm IN 1 KHz BANDWIDTH SPECTRUM ANALYZER

AILTECH Model 757A spectrum analyzer offers two features to the Model 757 without a change in price. It has a sensitivity of -125 dBm in 1 KHz bandwidth in the 0.001 to 2.0 GHz band and a 100 Hz IF resolution bandwidth. The unit features frequency coverage to 22 GHz, internal calibration, digital storage/ memory and a 100 dB display range. Domestic price: $23,000 or lease option. **Eaton Corp., Ronkonkoma, NY. Roy Wendell (516) 293-8905.**

Circle 171.

Devices

KLYSTRON AMPLIFIER FOR S-BAND RADAR

Model TH-2069 delivers from 1 MW to 2 MW of peak output power (up to 3 kW average power) between 2.7 and 2.9 GHz. Minimum gain is 52 dB, typical efficiency is 50% and the -1 dB bandwidth is over 40 MHz. The unit features digital tuning, forced-air cooling and an integral ion pump. **Thomson-CSF Electron Tube, Clifton, NJ. (201) 779-1004.**

Circle 172.

LOW NOISE GaAs FETs COVER 4—18 GHz

GaAs FET Models VSF-9320 and VSF-9330 are designed for operation in the 4 to 18 GHz frequency range. Typical noise figure for the 9320 series ranges from 1.4 dB at 4 GHz, to 3.0 dB at 18 GHz; at 10 GHz the noise figure is 1.9 dB. Typical gain at 10 GHz is 10 dB; maximum available gain is 12 dB. The VSF-9330 is a general purpose FET for X and Ku-band applications with slightly higher noise figure and lower gain. The devices are available in chips for use in thin film amplifier applications and in a hermetically sealed 70-mil square stripline package (VSF 9321, VSF 9331) for softboard use. Price: 1-24, VSF 9320 - $110; VSF 9330 -$65. Availability: stock to four weeks. **Varian Solid State Microwave Div., Santa Clara, CA. (408) 988-1331.**

Circle 173.

UNIT AMPLIFIER TRANSISTOR COVERS dc-500MHZ

Model MPA-201 is a unit-amplifier transistor which gives both broadband and narrowband performance from dc to 500 MHz. Device has a typical output power of 500 mW at 500 MHz, a typical 3rd-order intercept point of +45 dBm at 70 MHz and a typical low noise figure of 5.5 dB at 70 MHz. Small signal gain is 13 dB (typ.) from 2-500 MHz. Bias requirements are 12.5 V and 250 mA, typical. Package is hermetically sealed. Internal 50Ω matching makes cascading feasible. Units are also suitable for oscillator or modulator functions. **Communications Transistor Corp. San Carlos, CA. (415) 592-9390.**

Circle 174.

Material

CLOSE-TOLERANCE MICROWAVE LAMINATES

CuClad 217 and 233 teflon-glass microwave laminates have ± .02 dielectric-constant tolerances allowing for better design control in microwave integrated circuits. They are available in thickness increments of 5 mils from 5 to 125 mils. The dielectric constant is tested and certified for all thicknesses at X-band. Standard sheet size is 17" x 36" with 36" x 36" sheets also available. **3 M, Dept. EP-81-15, Box 33600, St. Paul. MN.**

Circle 175.

System

INTRA-CITY MICROWAVE SYSTEM

The Intra-City Microwave (ICM) system delivers 0.1 W of output power at the antenna and is designed for operation in the 10,550 to 10,680 MHz mobile band and the 12,200 to 12,700 MHz business radio band. The basic Model ICM01310M is a video only unit with optional add on data and audio subcarriers available. **International Microwave Corporation, Cos Cob CT. (203) 661-6277.**

Circle 176.

Antenna

MILLIMETER-WAVE ANTENNAS

Series 4581xH millimeter-wave parabolic antennas are available in eight different waveguide bands between 26.5 and 170 Ghz and in six different sizes. The smallest, a 4" dia. dish has a prime focus feed. The 10", 12", 18", 24" and 36" models use a Cassegrain feed. Cassegrain feed models have sidelobes typically 18 dB down. Construction is solid epoxy glass laminate with aluminum flame spray metallization. The 12" model weighs 7.5 lbs. Cassegrain types are equipped with integral boresighted telescope. Price: 12" - $2240. Delivery: 120 days ARO. **Hughes Electron Dynamics Division, Torrance, CA. (213) 517-6400.**

Circle 177.

Advertising Index

Sales Representatives

Bernard Bossard Vice President

Howard Ellowitz Publisher

Sandra Pasqualucci Mgr. Sales/Marketing Administration

NEW ENGLAND, NEW YORK STATE, MID-ATLANTIC, NORTHEAST, SOUTHEAST, AND MIDWEST
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Tel: (415) 969-3886

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Yamaguchi Bldg.
2-12-9 Kanda Jimbocho
Chiyoda-ku, Tokyo 101 Japan
Tel: 230-4117, 4118
Cable: REPRETIV Tokyo
TLX: J26860

New Literature

CALIBRATION SERVICES CATALOG

The National Bureau of Standards has issued a new edition of the agency's catalog of calibration services reflecting changes as of the second quarter of 1980. Over 300 services in the areas of mass and dimensional metrology, mechanics and acoustics, electrical and eletromagnetic qualities, time and frequency, thermodynamic qualities, optical properties and ionizing radiation are offered. Price lists, info contacts and service changes are updated every 6 months in a special supplement. Price: $4.50. Specify catalog number 003-003-02299-3. **Office of Management Services, National Bureau of Standards, Washington, DC 20234.**

SIGNAL PROCESSING COMPONENT CATALOG

A 16-page catalog features a complete guide to specifications, schematic representations and outline dimensions for signal processing components. Included in the listings are double balanced mixers, RF switches, broadband transformers, directional couplers, diode assemblies, frequency doublers, power dividers and quad hybrids. **Summit Engineering, Bozeman, MT (406) 587-4511.**

Circle 133.

PRODUCT CAPABILITY PROFILE

A four color 12-page booklet details capabilities for microwave components and subsystems spanning the DC to 40 GHz frequency range. Categories include GaAs FET amplifiers, high power microwave components, ferrites, filters, transmission line components and custom microwave assemblies. **M/A Electronics Canada, Ltd., Ontario. Sam Singer (416) 625-4605.**

Circle 134.

TRAVELING WAVE TUBE CATALOG

A short form catalog details Toshiba traveling wave tubes covering communications and satellite bands from 470 MHz to 31 GHz with saturated power ratings of 7W to 8kW. **MATCOM, Inc., Palo Alto, CA (415) 493-6127.**

Circle 136.

SIGNAL SOURCE BROCHURE

A brochure detailing the 520 line of solid-state microwave signal sources is available. Five units in the series cover the 0.85 to 18GHz frequency range. **General Microwave Corp., Farmingdale, NY. Moe Wind (516) 694-3600.**

Circle 139.

MICROWAVE ABSORBER CATALOG

A 32-page microwave absorber catalog also serves as a designer's manual for making rigid and flexible terminations, attenuators, chokes and gaskets. The publication delineates FERROSORB™ physical properties, material selection, design and molding and machine practice. Absorber applications are illustrated. **Microwave Filter Company, East Syracuse, NY. Emily Bostick (800) 448-1666.**

Circle 137.

PULSE GENERATOR BROCHURE

A 4-page four-color brochure describes the Model AT-SM33 constant current pulse generator specifically designed for use with IMPATT diodes, IMPATT diode oscillators and amplifiers. Applications, performance specifications, protection features and operator control functions are detailed. **Ad-Tech Microwave, Inc., Scottsdale, AZ. G.A. Herlich (602) 998-1584.**

Circle 138.

SOLID-STATE AMPLIFIER CATALOG

A 152-page catalog covering a full line of solid-state amplifiers includes more than 60 new low-profile amplifiers, compact amplifiers, limiting amplifiers and TWTA replacements. The catalog is a complete reference with selection charts, an amplifier glossary and applications information. **Watkins-Johnson Co., Palo Alto, CA. S. B. Witmer (415) 493-4141.**

Circle 140.

COAXIAL CONNECTOR CATALOG

Catalog QR-6 is a condensed quick-reference edition incorporating subminiature and microminiature connectors for applications from DC to 26 GHz. SMA, SMB, SMC and Nanohex coaxial connections are included. Complete electrical and mechanical specifications are provided for each series. **RF Components Division, Sealectro Corp., Mamaroneck, NY. (914) 698-5600.**

Circle 141.

ELECTRONIC INSTRUMENTATION CATALOG

A 28-page condensed catalog of electronic measuring instruments for telecommunications equipment and systems contains specification data, product features and photographs. Units included are standard signal generators, spectrum analyzers, microwave system analyzers, error rate measuring equipment and others. **Anritsu America, Inc., Oakland, NJ. Joseph Oliveri (201) 884-2550.**

Circle 135.

MSC

Pulsed Power Transistors

Products for L-Band and S-Band Radar System

CLASS C AMPLIFIER (Pulse Conditions Per Table)
ELECTRICAL CHARACTERISTICS (@ 25°C)

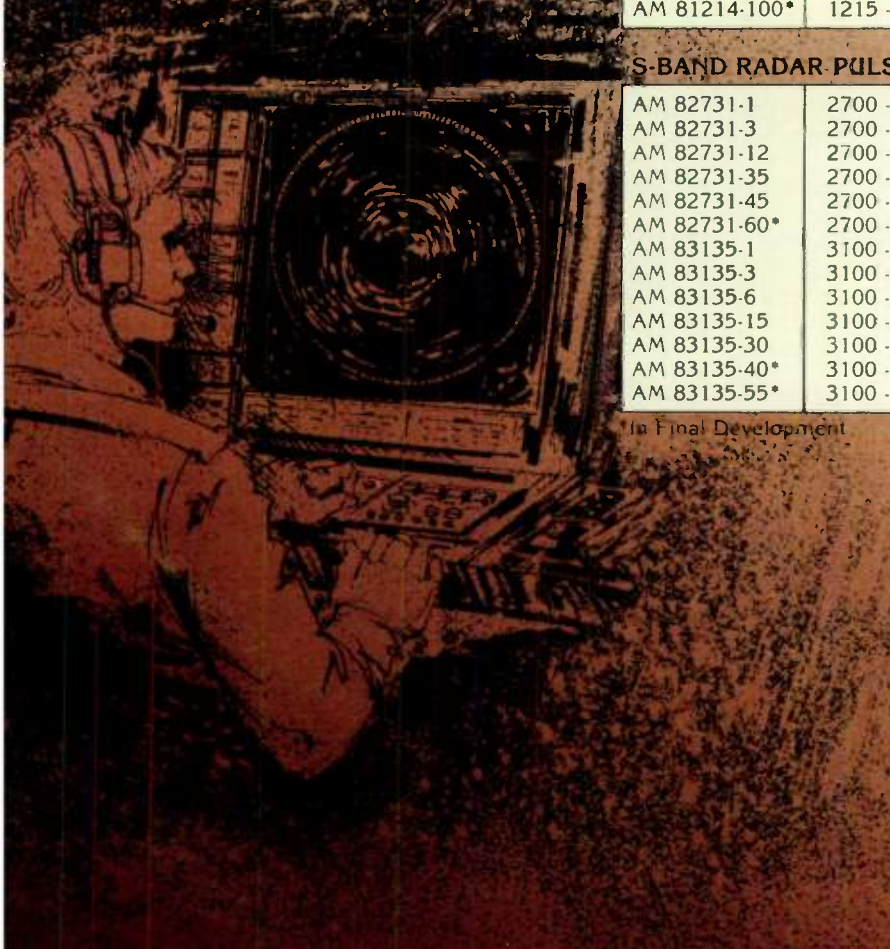
L-BAND RADAR PULSED POWER TRANSISTORS

AMPAC Model Number	Test Frequency (MHz)	P_{OUT} Min. (W)	P_{IN} (W)	Eff. Min. (%)	V_{CC} (V)	Pulse Width (usec)	Duty Cycle (%)
AM 1214-65P	1215 - 1400	65.0	16.00	40	42	50	10
AM 1214-75P	1215 - 1400	75.0	19.70	40	42	50	10
AM 1214-125P	1215 - 1400	110.0	29.00	40	42	50	10
AM 81214-250*	1215 - 1400	250.0	50.00	40	50	50	10
AM 81214-30	1215 - 1400	26.0	5.00	50	28	1000	10
AM 81214-60	1215 - 1400	55.0	12.00	50	28	1000	10
AM 81214-100*	1215 - 1400	100.0	22.00	50	28	1000	10

S-BAND RADAR PULSED POWER TRANSISTORS

AM 82731-1	2700 - 3100	1.0	0.20	30	28	100	10
AM 82731-3	2700 - 3100	3.5	0.80	30	28	100	10
AM 82731-12	2700 - 3100	12.5	3.20	30	40	100	10
AM 82731-35	2700 - 3100	35.0	11.25	30	40	100	10
AM 82731-45	2700 - 3100	45.0	15.00	30	40	100	10
AM 82731-60*	2700 - 3100	60.0	21.00	30	40	100	10
AM 83135-1	3100 - 3500	1.0	0.30	25	28	100	10
AM 83135-3	3100 - 3500	3.0	0.90	25	28	100	10
AM 83135-6	3100 - 3500	6.5	2.60	25	28	100	10
AM 83135-15	3100 - 3500	15.0	6.00	25	42	100	10
AM 83135-30	3100 - 3500	30.0	12.00	25	42	100	10
AM 83135-40*	3100 - 3500	40.0	16.00	25	42	100	10
AM 83135-55*	3100 - 3500	55.0	22.00	25	42	100	10

*In Final Development



YOUR TOTAL MICROWAVE RESOURCE

MSC provides a broad range of L and S-Band Radar Pulse Power Transistors with output pulse power levels up to 250 watts at 1.4GHz and 55 watts at 3.5GHz for the radar system designer. Please call or write for a complete MSC Radar Data Packet.

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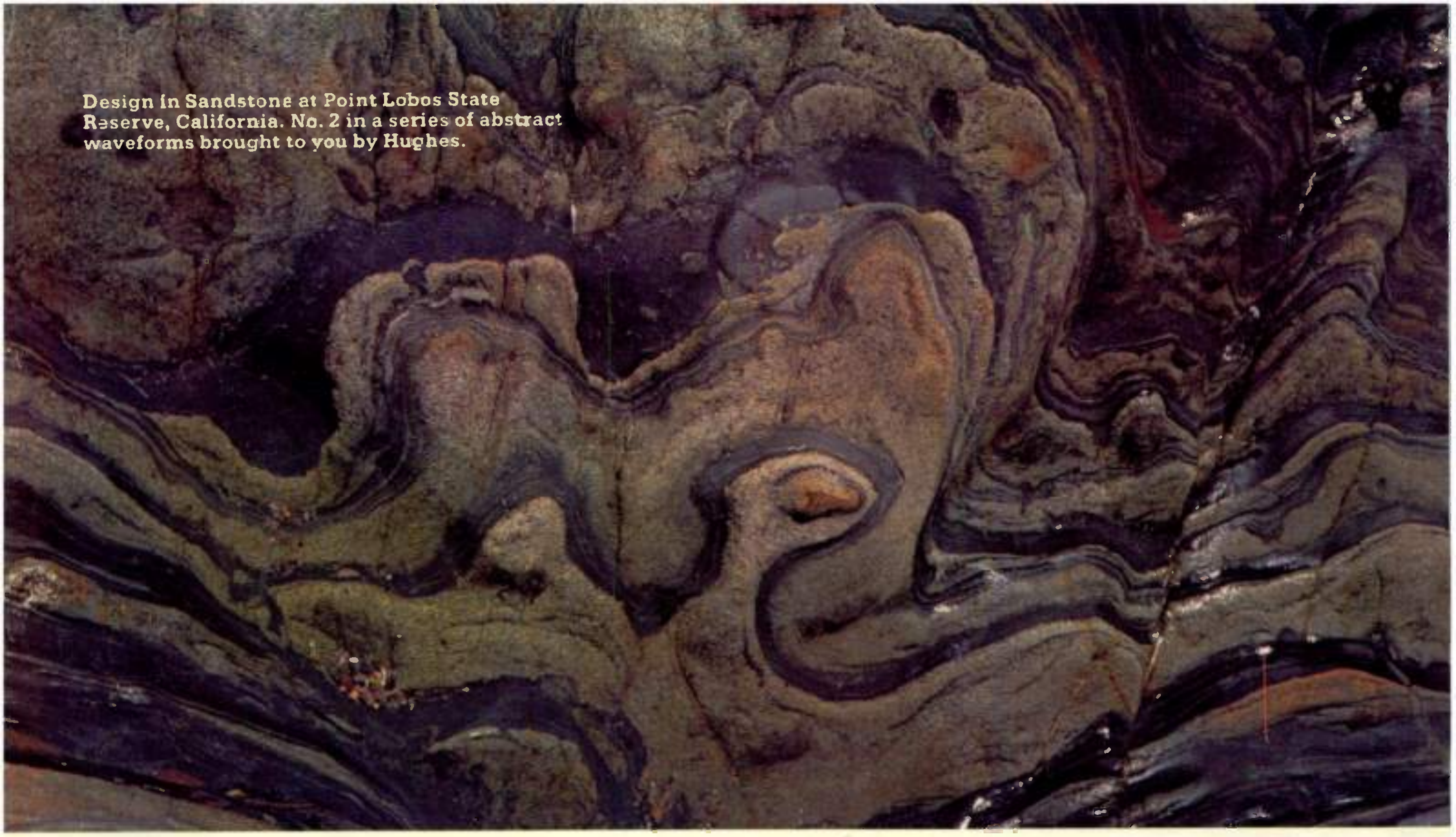
Circle 2 on Reader Service Card



Design in Sandstone at Point Lobos State Reserve, California. No. 2 in a series of abstract waveforms brought to you by Hughes.

New mm-wave computerized testing. For people who want to make waves.

Scalar transmission and reflection measurements up to 110 GHz.

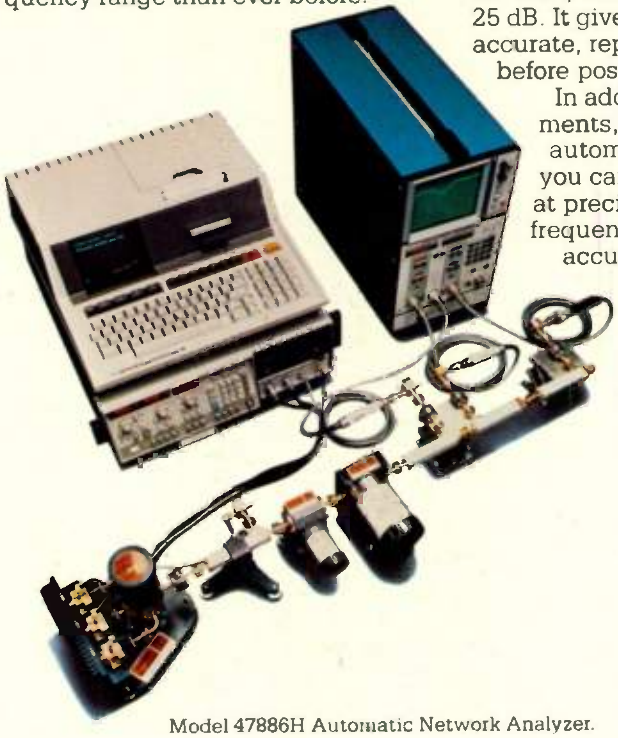
You're an innovator in millimeter waves.

You need a computer-controlled testing system to cover more frequency range than ever before.

You've got it.

Hughes' new 4788xH series of automatic analyzers lets you make automated scalar transmission and reflection measurements over full waveguide bandwidths up to 110 GHz, with a dynamic range of 25 dB. It gives you faster, more accurate, repeatable data than ever before possible.

In addition to swept measurements, with the optional automatic frequency control, you can make measurements at precisely controlled CW frequencies with repeatable accuracies of ± .01%. So it's easy to know exactly where your baseline is.



Model 47886H Automatic Network Analyzer.

The heart of the system is Hughes 4772xH mm-wave full-band sweep generator. We add an HP-85 computer; a full-band reflectometer with 40 dB directivity couplers and calibration short; a calibrated standard and an analyzer with built-in display... and you're ready to start making measurements in any of six different waveguide bands. It's quick, simple and easy, because the display shows all the instructions needed to run the system.

With options like frequency control and features like interfacing and systems analysis software, you'll be armed with the best means possible to prove and refine your designs—automatically.

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HUGHES

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Making waves in millimeter-wave technology.

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