microwave
JOURNAL

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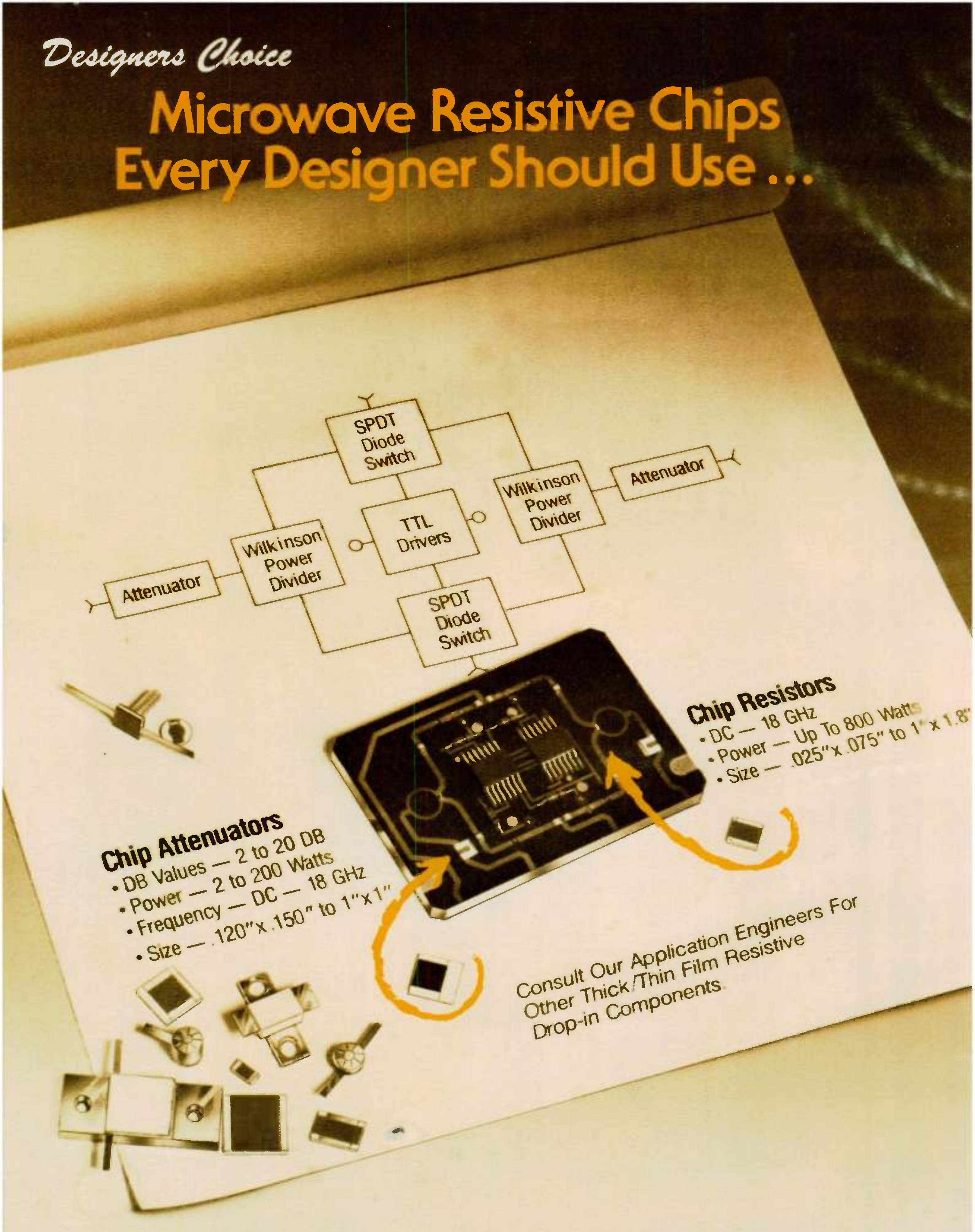

Designers Choice
Microwave Resistive Chips
Every Designer Should Use ...
SPDT
Diode
Switch
Wilkinson
Power
Divider
Attenuator
TTL
Drivers
Wilkinson
Power
Divider
Attenuator
SPDT
Diode
Switch
Chip Resistors
• DC — 18 GHz
• Power — Up To 800 Watts
• Size — .025" x .075" to 1" x 1.8"
Chip Attenuators
• DB Values — 2 to 20 DB
• Power — 2 to 200 Watts
• Frequency — DC — 18 GHz
• Size — .120" x .150" to 1" x 1"
Consult Our Application Engineers For
Other Thick/Thin Film Resistive
Drop-in Components.

YOUR BEST BET -

Model Number	Freq. Range (MHz)	Min. Gain (dB)	Gain Flat. (±dB)	Noise Figure Typ.	Noise Figure Max.	VSWR Max. In	VSWR Max. Out	Dynamic Range: 1 dB Gain Comp. (dBm Min.)	Dynamic Range: 3rd Order Inter. (Typ.)	DC Power Volt	DC Power mA
	To 500 MHz										
AU-2A-0110	1-100	30	.5	1.0	1.3	2:1	2:1	0	+10	15	30
AU-3A-0110	1-100	45	.5	1.0	1.3	2:1	2:1	+5	+15	15	50
AU-4A-0110	1-100	60	.5	1.0	1.3	2:1	2:1	+10	+20	15	75
AU-2A-0120	1-200	30	.5	1.2	1.4	2:1	2:1	+5	+15	15	55
AU-3A-0120	1-200	45	.5	1.2	1.4	2:1	2:1	+5	+15	15	50
AU-4A-0120	1-200	60	.5	1.2	1.4	2:1	2:1	+10	+20	15	75
AU-2A-1045	100-450	30	.5	1.3	1.6	1.5:1	1.5:1	+7	+17	15	50
AU-3A-1045	100-450	45	.5	1.3	1.6	1.5:1	1.5:1	+10	+20	15	50
AU-4A-1045	100-450	60	.5	1.3	1.6	1.5:1	1.5:1	+10	+20	15	80
AU-1A-0150	1-500	15	.5	3.5	4.0	2:1	2:1	+10	+20	15	40
AU-2A-0150	1-500	30	.5	1.5	1.8	2:1	2:1	+10	+20	15	55
AU-3A-0150	1-500	45	.5	1.5	1.8	2:1	2:1	+10	+20	15	70
AU-4A-0150	1-500	60	.5	1.5	1.8	2:1	2:1	+10	+20	15	100
	To 1000 MHz										
AM-1A-0510	500-1000	8	.5	2.0	2.5	2:1	2:1	-5	+5	15	20
AM-2A-0510	500-1000	20	.5	1.5	1.8	2:1	2:1	0	+5	15	45
AM-3A-0510	500-1000	30	.5	1.5	1.8	2:1	2:1	+5	+15	15	75
AM-4A-0510	500-1000	40	.5	1.5	1.8	2:1	2:1	+10	+20	15	105
AM-1A-000110	1-1000	10	.5	2.0	2.5	2:1	2:1	-10	0	15	20
AM-2A-000110	1-1000	25	.5	1.8	2.2	2:1	2:1	+5	+15	15	70
AM-3A-000110	1-1000	35	.5	1.8	2.2	2:1	2:1	+10	+20	15	60
AM-4A-000110	1-1000	50	.5	1.8	2.2	2:1	2:1	+10	+20	15	105

MITEQ AMPLIFIERS

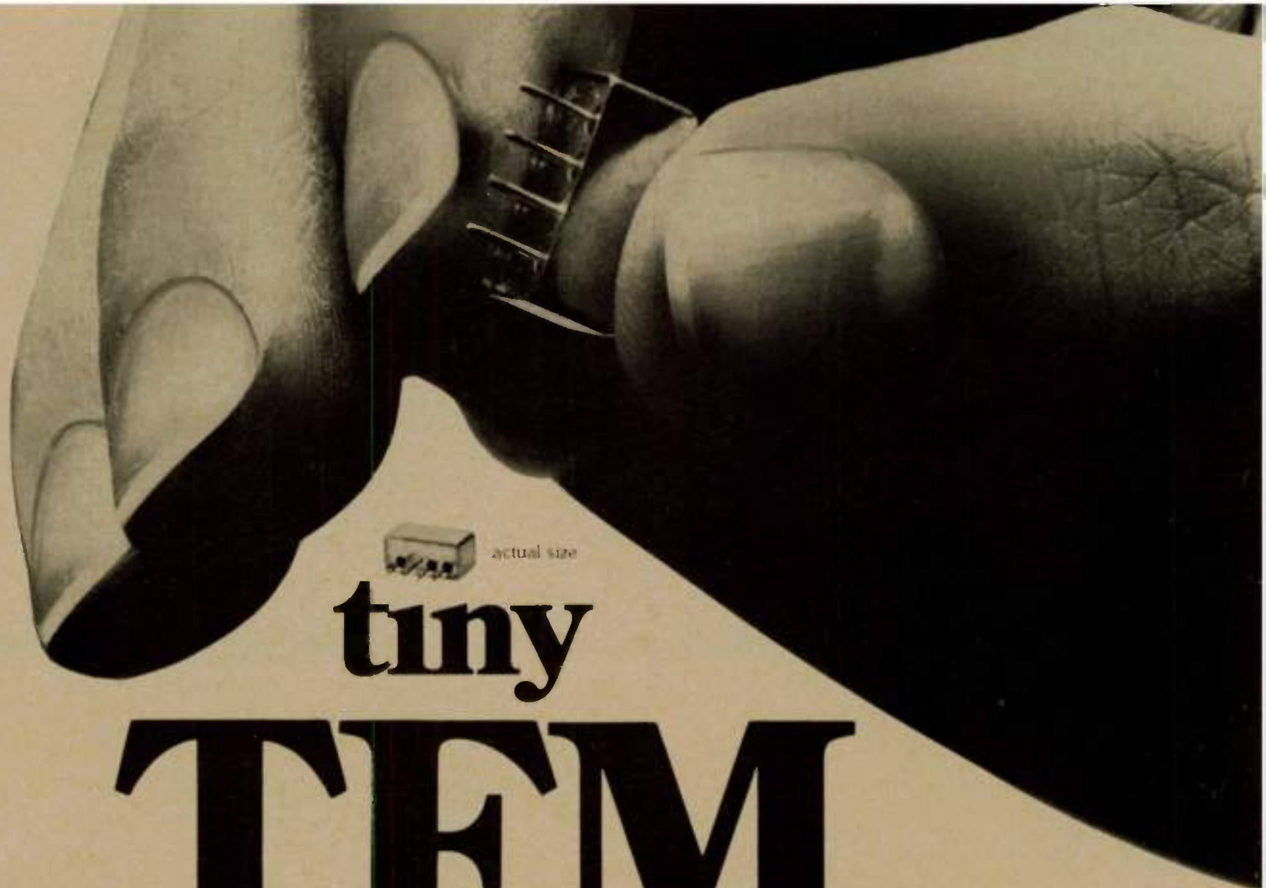
Model Number	Freq. Range (MHz)	Min. Gain (dB)	Gain Flat. (±dB)	Noise Figure Typ.	Noise Figure Max.	VSWR Max. In	VSWR Max. Out	Dynamic Range: 1 dB Gain Comp. (dBm Min.)	Dynamic Range: 3rd Order Inter. (Typ.)	DC Power Volt	DC Power mA
	To 1500 MHz										
AM-1A-0515	500-1500	10	.5	2.0	2.5	2:1	2:1	-5	+5	15	25
AM-2A-0515	500-1500	18	.5	2.0	2.5	2:1	2:1	-5	+5	15	50
AM-3A-0515	500-1500	30	.5	2.0	2.3	2:1	2:1	+7	+17	15	75
AM-4A-0515	500-1500	40	.5	2.0	2.3	2:1	2:1	+7	+17	15	100
AM-1A-000515	5-1500	10	.5	3.0	3.2	2:1	2:1	-10	0	15	20
AM-2A-000515	5-1500	25	.5	3.0	3.2	2:1	2:1	+10	+20	15	60
AM-3A-000515	5-1500	30	.5	3.0	3.2	2:1	2:1	+10	+20	15	100
AM-4A-000515	5-1500	40	.5	3.0	3.3	2:1	2:1	+10	+20	15	100
	To 2000 MHz										
AM-1A-1020	1000-2000	10	.5	2.0	2.5	2:1	2:1	-10	0	15	10
AM-2A-1020	1000-2000	18	.5	2.0	2.5	2:1	2:1	0	+10	15	50
AM-3A-1020	1000-2000	30	.5	2.0	2.5	2:1	2:1	+10	+20	15	75
AM-4A-1020	1000-2000	35	.75	2.0	2.5	2:1	2:1	+10	+20	15	105
AM-5A-1020	1000-2000	50	.75	2.0	2.5	2:1	2:1	+10	+20	15	120
AM-3A-0322	300-2200	30	.75	2.2	2.75	2:1	2:1	0	+10	15	75
AM-1A-0420	400-2000	7	.5	2.4	2.8	2:1	2:1	-10	0	15	20
AM-2A-0420	400-2000	18	.5	2.0	2.5	2:1	2:1	0	+10	15	50
AM-3A-0420	400-2000	30	.5	2.0	2.5	2:1	2:1	+5	+15	15	75
AM-4A-0420	400-2000	40	.5	2.0	2.5	2:1	2:1	+5	+15	15	100
AM-5A-0420	400-2000	50	.75	2.0	2.5	2:1	2:1	+10	+20	15	125
AM-1A-000520	5-2000	7	.5	3.0	3.5	2:1	2:1	-10	0	15	20
AM-2A-000520	5-2000	15	.5	3.5	4.0	2:1	2:1	-10	0	15	50
AM-3A-000520	5-2000	23	.5	3.5	4.0	2:1	2:1	+5	+15	15	75
AM-4A-000520	5-2000	30	.75	3.5	4.2	2:1	2:1	+5	+15	15	100

Contact factory for custom options - often at no extra charge.

MITEQ INC./100 RICEFIELD LANE/HAUPPAUGE NY 11787/(516)543-8873

CIRCLE 4 ON READER SERVICE CARD

Microwave Journal (USPS 396-250) is published monthly per request by Horizon House-Microwave Inc., 610 Washington Street, Dedham, MA 02026. Second Class postage paid at Dedham MA 02026 and additional offices. **POSTMASTER:** Send form 3579 to Microwave Journal, 610 Washington Street, Dedham MA 02026.



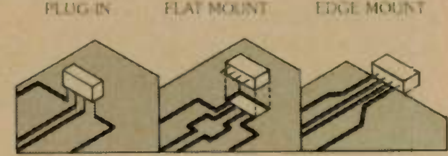
tiny TFM

the world's smallest hermetically sealed mixers
40 KHz to 3 GHz MIL-M-28837 performance
The TFM Series from Mini-Circuits from $11.95

Increase your packaging density, and lower your costs...specify Mini-Circuits miniature TFM Series. These tiny units 0.5" x 0.21" x 0.25" are the smallest, off-the-shelf Double Balanced Mixers available today.

Requiring less PC board area than a flat-pack or TO-5 case, the TFM Series offer greater than 45 dB isolation, and only 6 dB conversion loss.

Manufactured to meet all the requirements of MIL-M-28837, the tiny but rugged TFM units have become the preferred unit in new designs for military equipment.



E-Z Mounting for circuit layouts
Use the TFM series to solve your tight space problems. Take advantage of the mounting versatility—plug it upright on a PC board or mount it sideways as a flatpack.

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

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

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

For Mini Circuits sales and distributors listing see page 41

Mini-Circuits
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

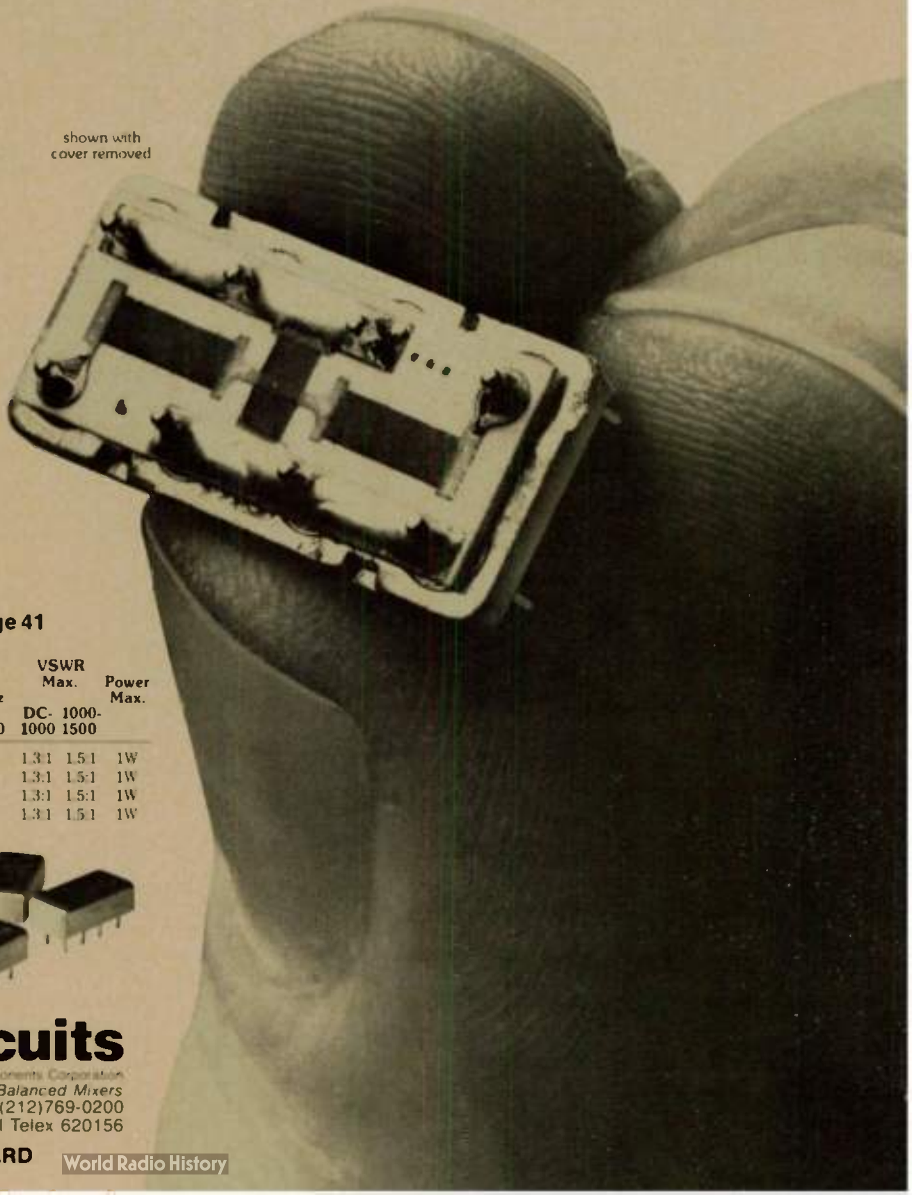
CIRCLE 5 ON READER SERVICE CARD

fixed attenuators

the world's lowest priced attenuators 3,6,10 or 20dB
from DC to 1500 MHz...hermetically sealed

The AT Series from Mini-Circuits

$1.95
1000 Quantity
$3.95 (10-49)



Check these features:

√ High stability; thick film construction in a hermetically sealed case
√ Rugged construction: Meets requirements of MIL STD 202
√ Miniature Size: 0.4" by 0.8" by 0.2" high
√ Flat frequency response: Typically ±0.3 dB
√ Excellent VSWR: typically less than 1.2:1
√ Low cost: $1.95 (1,000 quantity), $3.95 (10-49)
√ Delivery: From stock

DESIGNERS KIT AVAILABLE, KAT-1
4 attenuators of each type
AT-3, AT-6, AT-10, AT-20 only $39.95

For Mini Circuits sales and distributors listing see page 41

Model	Attenuation, dB Nominal Value	Attenuation Tolerance from Nominal	Frequency Range MHz	Attenuation Change From Nominal Over Frequency Range, MHz		VSWR Max.		Power Max.
				DC 1000	1000-1500	DC-1000	1000-1500	
AT-3	3	±0.2dB	DC-1500	0.5dB	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	0.5dB	0.8dB	1.3:1	1.5:1	1W

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World's largest manufacturer of Double Balanced Mixers
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67-2 REV. B

CIRCLE 6 ON READER SERVICE CARD

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 popular mixers in the world. Rugged construction and tough inspection standards insure MIL-M-28837/1A performance.*

Check these features...

SRA-1 the world standard, covers 500 KHz to 500 MHz, Hi-REL, 3 year guarantee, HTRB tested, MIL-M-28837/1A-03 S performance.* $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, MIL-M-28837/1A performance.* $11.95 (6-49).

SBL-1 world's lowest cost industrial mixers, only $3.95 (100), 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).

*Units are not QPL listed

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 41

finding new ways...
setting higher standards

Mini-Circuits

A Division of Scientific Components Corporation

World's largest manufacturer of Double Balanced Mixers

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CIRCLE 8 ON READER SERVICE CARD

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microwave JOURNAL

contents

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USPS 396-250
NOVEMBER 1981

BUSINESS/SPECIAL REPORTS

TECHNICAL/APPLICATIONS SECTION

DEPARTMENTS

ON THE COVER: Systrom Donner Microwave Division's new Model 5220 Transline Analyzer for locating transmission line faults is shown in use in the UHF transmitter room of the USS John F. Kennedy (CV-67). See the Technical Feature beginning on page 49.

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 Microfilm copies of Microwave Journal 300 N. Zeeb Rd., Ann Arbor, MI 48106. are available from University Microfilms.

Some things we *can* say about one of our Supercomponents™.

This MIC (we proudly call it a Narda Super Component) is a switched oscillator assembly used in existing EW systems. It's already at work in places we can't disclose. However, we can talk about our "vertical integration" that put it there in the first place --innovative engineering and design under direct in-house control. All the disciplines necessary to make it are at Narda. It probably couldn't be completed anywhere but here.

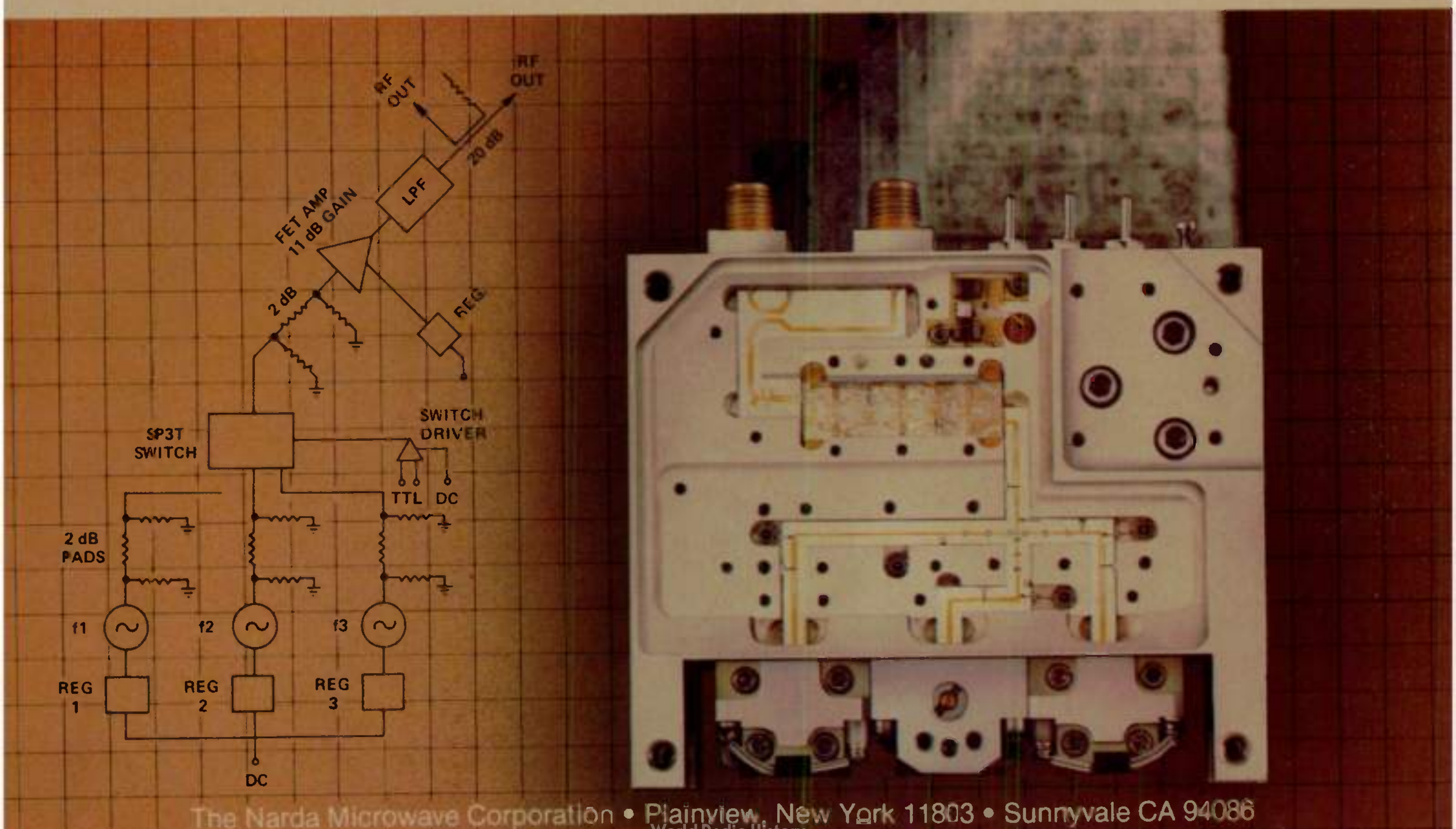
Take design, for example. To get stability and low noise in its C-Ku band operating range, we successfully combined three ultra-stable GUNN oscillators with a broadband, low noise, GaAs FET amplifier. Simultaneously, we developed a rather unique PIN switch that terminates (not reflects) energy to get instant broadband switching with high isolation (80 dB or 1000x higher than typical switches).

And the story continues. Low pass filters, hybridized regulators and drivers, directional couplers and all the electronics --- all designed and manufactured by Narda right down to the diodes and transistors. The payoff is that we are delivering Supercomponents with exacting requirements and on-time. We can do the same for you. Call us.

For further information, write or call Narda Mircowave Corporation, Plainview, L.I., New York 11803 • (516) 349-9600. TWX: 510-221-1867 • CABLE: NARDACORP PLAINVIEW NEW YORK.

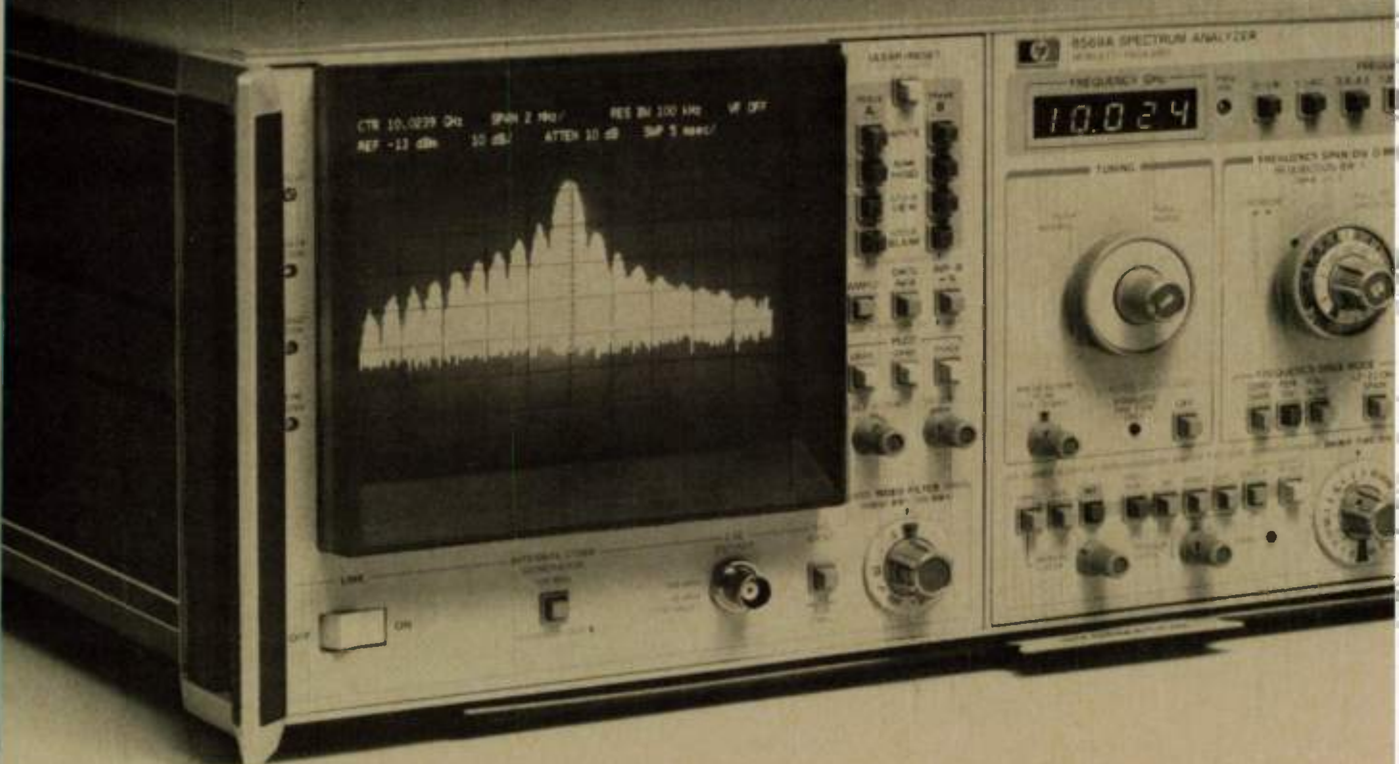
The biggest job was to put it into the smallest package.

narda



CIRCLE 9 ON READER SERVICE CARD

This new microwave displays its talents in



The new HP 8569A microwave analyzer makes it easier than ever to collect, view and analyze spectral information. The CRT system uses digital techniques to give you a variety of data presentation modes. Built-in HP-IB input/output capability lets you transfer displayed information directly to a digital plotter. And, should you want to automate some of the operation, you can connect the 8569A to an HP-IB-compatible controller that will process and format the data, as well as provide operating instructions and feedback on the 8569A's CRT.

The flexible CRT display.

Two independent signal traces allow you to store trace data and still monitor changes. Traces can be displayed individually or simultaneously. Built-in processing capability permits techniques such as: *digital averaging* to reveal low-level signals without sacrificing sweep speed; "INP-B→A" to show the difference between traces to highlight changes between scans; *maximum hold* to display time-related data such as long-term signal drift. All major control settings are displayed above the graticule, to avoid interference with the traces.

High performance levels.

Frequency coverage is from 10 MHz to 22 GHz, with built-in preselection from 1.7 to 22 GHz. Using external mixers, coverage can be extended to 170 GHz. Advanced mixer and LO designs produce high sensitivity (−95 dBm at 18 GHz for 1 kHz BW), and excellent stability and spectral purity for close-in measurements. Ten IF filters—from 100 Hz to 3 MHz—

83107

spectrum analyzer more ways than one.

let you select the optimum trade-off between resolution and sweep speed. With the optional 100 MHz internal comb generator, frequency accuracy can be increased to ±0.007%, even at 22 GHz.

Operator convenience.

The HP 8569A retains the convenient set-up and operating procedures traditional with HP spectrum analyzers. Coupled controls simplify the procedure: the operator merely tunes to the signal, decreases the frequency span, and then sets the signal to the amplitude reference level. The other operating parameters are automatically optimized. The easy-to-access display modes also include direct transfer of displayed information to any HP-IB-compatible plotter without requiring a controller.

HP-IB interface, an entree to automation.

Thanks to the 8569A's built-in HP-IB display interface, two-way communication is possible with external computing and peripheral devices. Besides having direct plotter output capability, the 8569A can transfer display data to a computer—such as the economical HP 85F—for data logging and processing.

Conversely, computer-generated trace data and alphanumeric prompting messages can be displayed on the 8569A's CRT. These semi-automatic capabilities are especially useful in environments such as production test, where operator interaction is required. They make the 8569A a very practical first step in automating spectrum analysis.

The 8569A is priced at $26,500. For the internal comb generator Option 001, add $1,425. To find out how the HP 8569A can display its talents for you, call your nearby HP sales office or write to Hewlett-Packard, 1820 Embarcadero Road, Palo Alto, CA 94303. U.S. domestic prices only

It keeps getting easier to make HP your automatic choice for spectrum analysis.

HEWLETT PACKARD

CIRCLE 10 ON READER SERVICE CARD

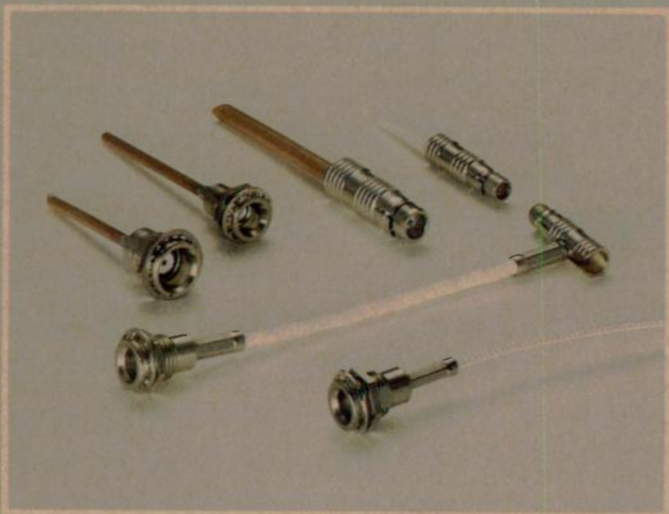
OSM/OSSC (SMA) Miniature Coaxial Connectors

- Laser-welded right-angle connectors for higher performance at lower cost
- 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
- Panel and Bulkhead mount receptacles
- 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)

Terminations/Attenuators

- Meet MIL-E-5400 and MIL-E-16400 environmental requirements
- OSM, Type N, BNC and 7mm connectors

We've made several important changes at OMNI SPECTRA to insure that you will get the best connectors and components available for less money and with less waiting.

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**

Omni Spectra, Inc., 21 Continental Boulevard, Merrimack, NH 03054

It's easier than ever to get the best for less.

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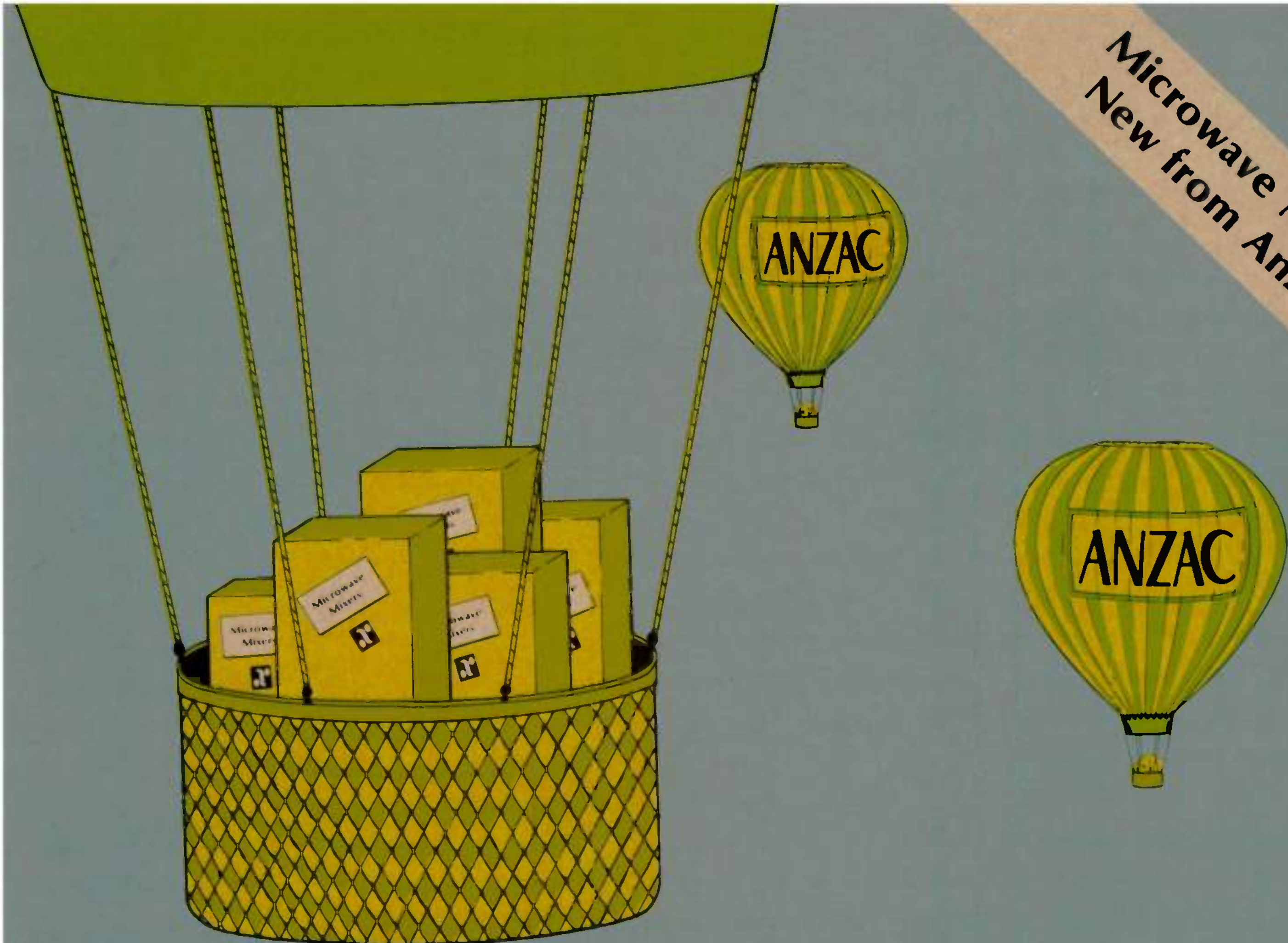

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FREQUENCY (GHz)	RF, LO	1 - 7	1 - 6	0.5 - 9	0.5 - 8	0.5 - 18	0.5 - 18	0.5 - 18	2 - 4	2 - 8	2.6 - 5.2	3.7 - 4.2	4 - 8	5.9 - 6.4	3 - 13
	IF	0.01 - 2	0.01 - 2	0.01 - 2	0.01 - 2	DC - 5	0.5 - 8	DC - 5	DC - 0.3	DC - 0.5	DC - 0.3	DC - 0.3	DC - 0.5	DC - 0.5	DC-400
CONVERSION LOSS** (dB)		6.0	7.0***	6.5	9.0***	8.0	8.0	8.0	4.5	5.0	4.5	4.5	4.5	4.5	6.5

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Sponsor: IEEE Los Angeles Bay Section. Place: Snowmass, Colorado. Topics: Systems concepts and management, aerospace electronics and measurement, energy and space, millimeter and microwave technology. Contact: R. A. Gaspari, Hughes Aircraft MS S12/V305, P.O. Box 92919, Los Angeles, CA 90009.

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.

MEDICAL APPLICATIONS OF ELECTRO—MAGNETIC ENERGY WORKSHOP JUNE 18, 1982
Call for papers Sponsors: IEEE MTT-S and IEEE committee on Man and Radiation (COMAR). Place: Hyatt Regency Hotel Dallas, TX. Topics: Hyperthermia, NMR and electromagnetic imaging, radiometry, blood and organ thawing, etc. Authors are asked to submit 5 copies of both a 35-word abstract and a 500 — 1,000-word summary (up to 6 illustrations), clearly explaining their contribution, its originality and relative importance by January 8, 1982 to Dr. Gideon Kantor, Bureau of Radiological Health, 12721 Twinbrook Parkway, Rockville, MD 28057 Tel: (301) 443-3840.

CONFERENCE ON PRECISION ELECTRO—MAGNETIC MEASUREMENTS JUNE 28-JULY 1, 1982
Call for papers Sponsor: National Bureau of Standards. Place: NBS, Boulder, Colorado. Topics: Design, performance, or application of electromagnetic measurements, techniques, instruments or systems. Submit both a 35-40 word abstract and a 500-1000 word summary in camera ready form by February 15, 1982. David W. Allen, CPEM, '82, National Bureau of Standards, 325 Broadway, Boulder, CO 80303. (303) 497-3981. ■

MONOLITHIC INTEGRATED CIRCUIT TRENDS

There is little disagreement with the thesis that microwave systems will ultimately incorporate a high proportion of monolithic microwave integrated circuits. As pointed out in the Special Report in this issue, a major goal of microwave research during the past several years has been to develop the technologies required for the fabrication of those circuits. The report reviews the substrates considered the most promising candidates for the circuits. It discusses the active and passive elements suitable for various circuit functions and the innovative approaches which have been explored for realizing practical designs. Finally, it reviews the particular advantages and disadvantages of monolithic microwave integrated circuits and suggests the applications for which they appear to be most appropriate.

Sum Up

THE NAVY'S NEW TRANSMISSION LINE FAULT FINDER

Transmission line faults are located quickly and accurately with the AN/PSM-40, the Navy's new analyzer which makes use of digital signal processing algorithms and microprocessor based hardware. The instrument utilizes frequency domain techniques involving digital processing of swept RF signals. Spectrum analysis of the detected waveform provides information about each fault within the transmission line system; multiple faults are identified in a single measurement. The design and operation of the interactive equipment is described in detail.

MAXIMIZING MICROWAVE FET POWER AMPLIFIER EFFICIENCY

While ohmic losses have an obvious effect on FET amplifier efficiency, a study of the effect of saturation of the devices under large signal conditions has identified two additional factors which directly affect operating efficiency. The paper describes a waveform measurement system which yields current waveforms at a number of points in the FET circuit and it discusses the use of that data to optimize amplifier efficiency.

QUASI-LUMPED ELEMENT IMPEDENCE MATCHING FOR WIDE-BAND GaAs FET AMPLIFIERS

A quasi-lumped element impedence matching technique for multi-octave bandwidth FET amplifiers from S through Ku band is described. The technique is applicable to short run designs for which a monolithic approach would be expensive on a unit basis. It also allows for circuit tunability to optimize gain, noise figure or power output. A 10 mil thick alumina substrate carries the matching circuit. Performance data on a number of low noise and power amplifiers for the 2-6 GHz, 4-12 GHz and 4-18 GHz bands are shown.

A 110 TO 170 GHz BWO

The expanding activity in mm waves exerts continuing pressure on mm wave source suppliers to expand the capabilities of sources for that effort. One of the latest entries, a backward wave oscillator for the 110 to 170 GHz band, is described. During the design effort for this tube, particular attention was paid to making it compatible with existing power supplies and to reducing its size and weight so that it would be competitive with the solid state oscillators which are available in the frequency range above 100 GHz.

Howard Ellowitz

Workshops & Courses

SATCOM SYSTEMS SHORT COURSE

Topic: The Earth Station: A Practical Approach to Implementation.
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Site: George Washington University
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Topic: Millimeter Wave Systems and Technology
Sponsor: The Vice President for Research and The Engineering Experiment Station
Site: Georgia Institute of Technology, Atlanta, GA
Date: January 26-28, 1982
Contact: Department of Continuing Education, Georgia Institute of Technology, Atlanta, GA 30332 Tel: (404) 894-2400

FIBER OPTICS SHORT COURSE

Topic: Fiber Optical Communications
Sponsor: Center for Professional Development
Site: Arizona State University
Date: March 15-17, 1982
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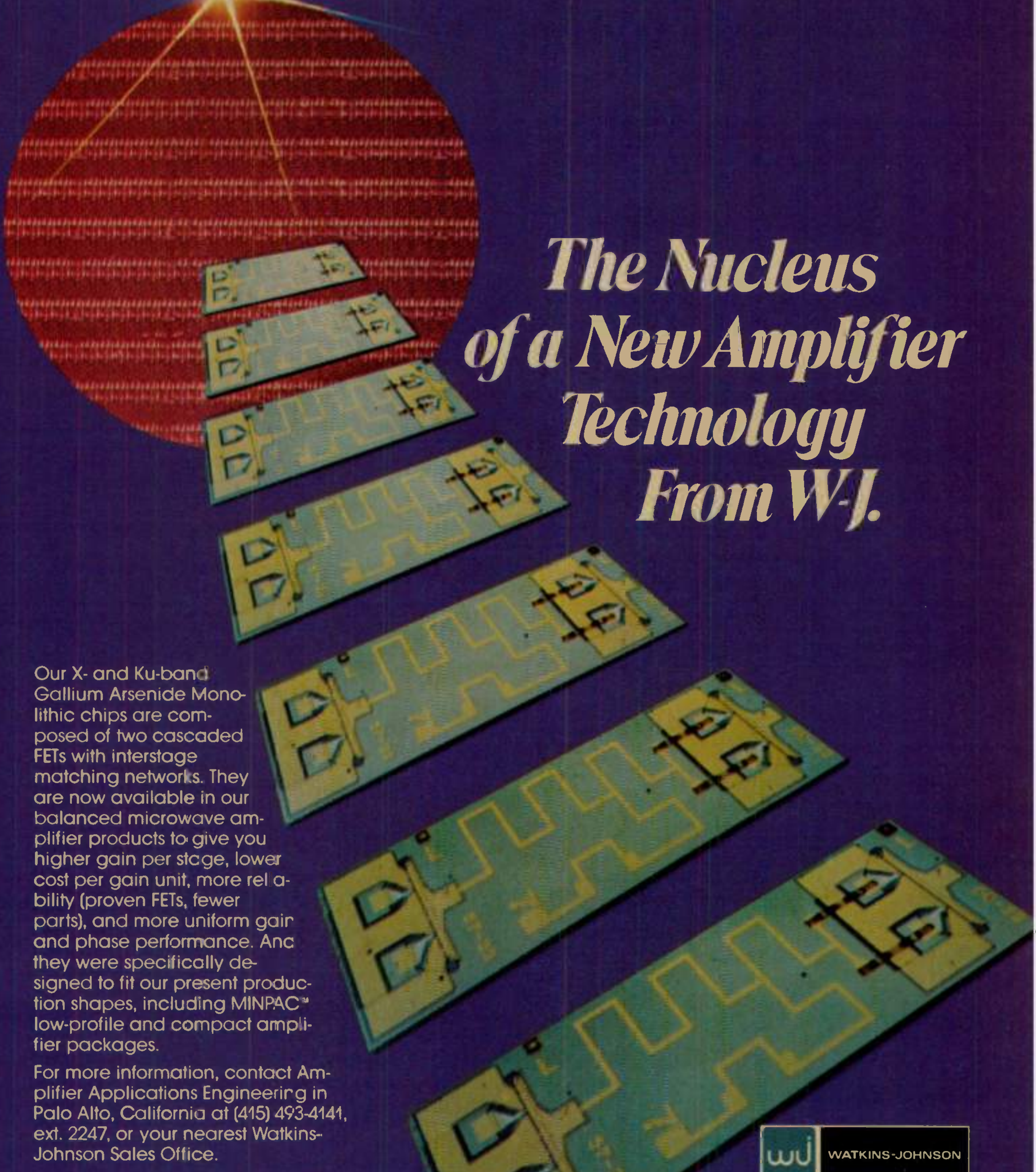
Monolithic Superfets



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Trends in Monolithic Microwave Integrated Circuits

FRED STERZER
RCA LABORATORIES
Microwave Technology Center
Princeton, NJ

INTRODUCTION

One of the major goals of microwave research for the past several years has been to develop the technologies needed to fabricate monolithic microwave integrated circuits (MMICs). This goal has now been attained. MMICs that can perform a variety of receive, transmit, or signal processing functions have been successfully fabricated in a number of laboratories.

As would be expected, different laboratories are using somewhat different technologies and design philosphies to build their MMICs. Still, among the diversity of approaches for building today's MMICs, general trends are beginning to emerge. It is the purpose of this paper to briefly review some of these trends.

SEMI-INSULATING GaAs

The overwhelming majority of MMICs fabricated to date use semi-insulating GaAs as the substrate, and it appears highly likely that GaAs (or some other semi-insulating III-V compound) will remain the favorite substrate for MMICs for years to come. The major competing substrate material is sapphire, which is used in an MMIC technology based on silicon-on-sapphire. This technology is not only not as advanced as GaAs MMIC technology, but more importantly, does not seem to offer as much long-term promise. This is primarily because much higher electron mobilities can be achieved in GaAs than in silicon-on-sapphire, electron mobility being of critical importance in determining the performance of most active microwave devices. A further important advantage of GaAs over silicon for MMICs is that the peak electron velocity in GaAs is higher than in silicon.

The technologies required for fabricating MMICs on semi-insulating GaAs have been painstakingly developed over the past two decades; they are described in hundreds of papers in the literature. The key technologies include growth of large semi-insulating GaAs crystals, expitaxial growth of thin n-type GaAs layers on semi-insulating GaAs, whole wafer or selective ion implantation into semi-insulating GaAs substrates or exitaxial layers, laser annealing of ion-implanted GaAs, ohmic and Schottky barrier contacts to GaAs, two-level metalization, deposited oxides, plated-through holes, etc.[1] The current large industry-wide effort on GaAs diodes and transistors ensures continuing improvement of these technologies.

Silicon Substrates for Monolithic Integrated Millimeter Wave IMPATT Circuits – A substrate material that is likely to offer competition to semi-insulating GaAs, particularly at millimeter wave frequencies, is high-resistivity intrinsic silicon. Use of silicon substrates allows one to monolithically integrate silicon IMPATT circuits, silicon IMPATT diodes being at the present time the most widely used solid-state power sources at millimeter wave frequencies. The resistivities that can be obtained in intrinsic silicon (2000-10,000 ohm-cm)* are not as high as those of semi-insulating GaAs (10^7-10^8 ohm-cm). Still, the small circuit losses associated with substrate resistivities of a few thousand ohm-cm can be tolerated in most millimeter wave circuits.

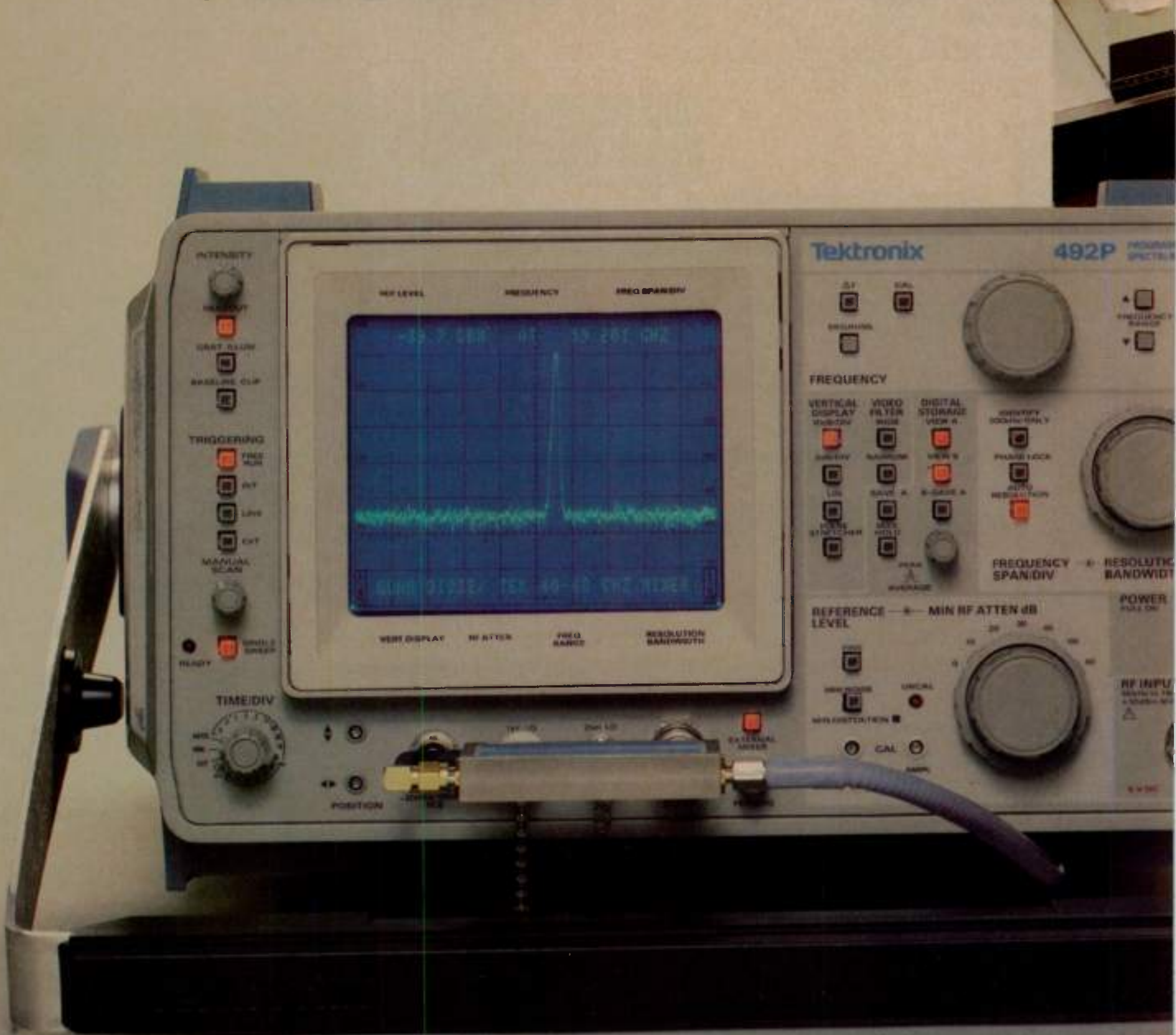
GaAs FIELD EFFECT TRANSISTORS

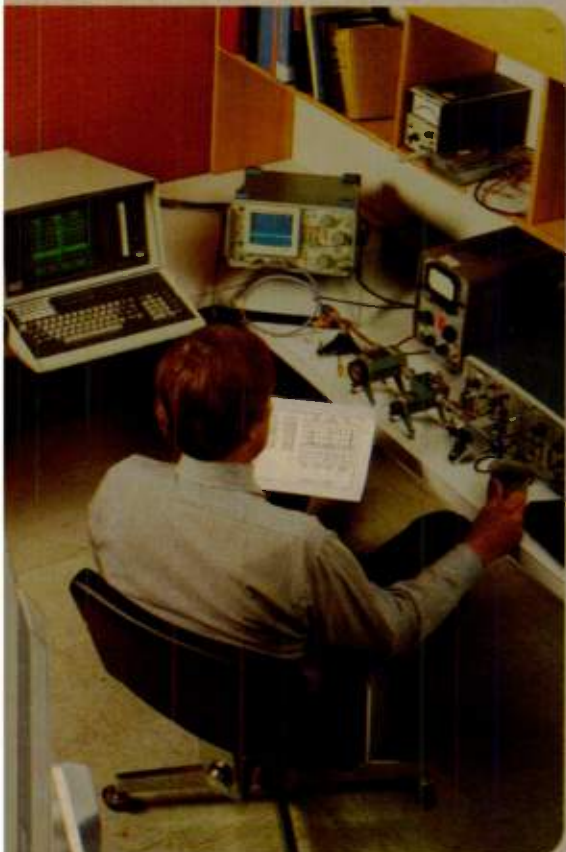
GaAs FETs are almost universally used as the active elements in today's MMICs[2]. They are versatile devices that can provide excellent performance in a variety of microwave and digital functions as shown in Table 1. In fact, the microwave portion of a number of systems can be built entirely around GaAs FET circuits. No creditable challenger to the GaAs FET has yet appeared, although in the future competition could possibly come from FETs fabricated from very high mobility

*These high resistivities can be maintained during IC fabrication provided low temperature (<800° C) processing techniques such as ion-implantation and laser annealing are employed.

[*Continued on page 24*]

Now Tektronix offers two fully programmable, easy-to-use, lab performance spectrum analyzers: 50kHz - 220GHz or 1.0kHz - 1.8GHz





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The 492P and 496P are friendly. Their high level language lets you concentrate on measurements instead of programming. Most commands are simply abbreviations of the front panel nomenclature. For example, to set the center frequency to 5.2GHz, just send "FREQ 5.2GHz" over the bus. To read the frequency, send "FREQ?".

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TABLE I

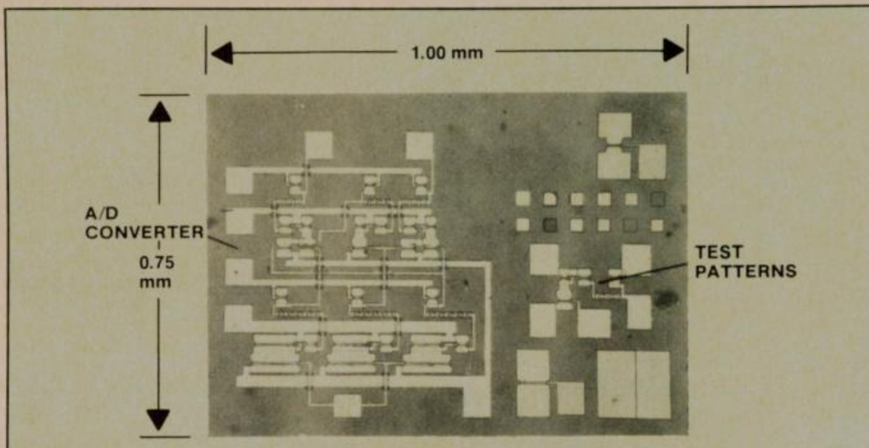
Range of Applications of GaAs FETs

Low-Noise Amplifiers	Mixers
Low-Power Amplifiers	Filters
Medium Power Amplifiers	Discriminators
Oscillators	Variable Attenuators
Switches	Power Splitters
Modulators	Power Combiners
Phase Shifters	Digital Circuitry
Multipliers	

materials such as $Ga_{0.47}In_{0.53}As/InP$. Also, for applications requiring high-peak powers, it might be advantageous in some cases to monolithically integrate GaAs IMPATT diodes.

LUMPED CIRCUITS

One of the major objectives of MMIC design is to minimize the substrate area occupied by a given microwave circuit. Lumped element circuits* being much smaller than distributed transmission line quencies appeared in the literature during the mid-1960s, but it was not until the late 1960s and early 1970s that Dr. Martin Caulton and his co-workers at RCA Laboratories developed a thin-film technology for batch-fabricating a variety of lumped element circuits on sapphire that provided good performance at microwave frequencies. The design rules, measurement techniques and thin film technology employed to fabricate hybrid



Courtesy of Dr L C Upadhyayula, RCA Laboratories, Princeton, NJ

Figure 1. 2-bit A/D converter chip.

circuits, are therefore fast becoming the favorite passive elements (in addition to Schottky diodes) of MMIC designs. Another factor contributing to the popularity of lumped elements is that it is easier to synthesize a wide range of impedance with them than with transmission line circuits.

Suggestions for using lumped element circuits at microwave frelumped element microwave circuits are described in References 3-5; they are for the most part directly applicable to monolithic lumped element microwave circuits.

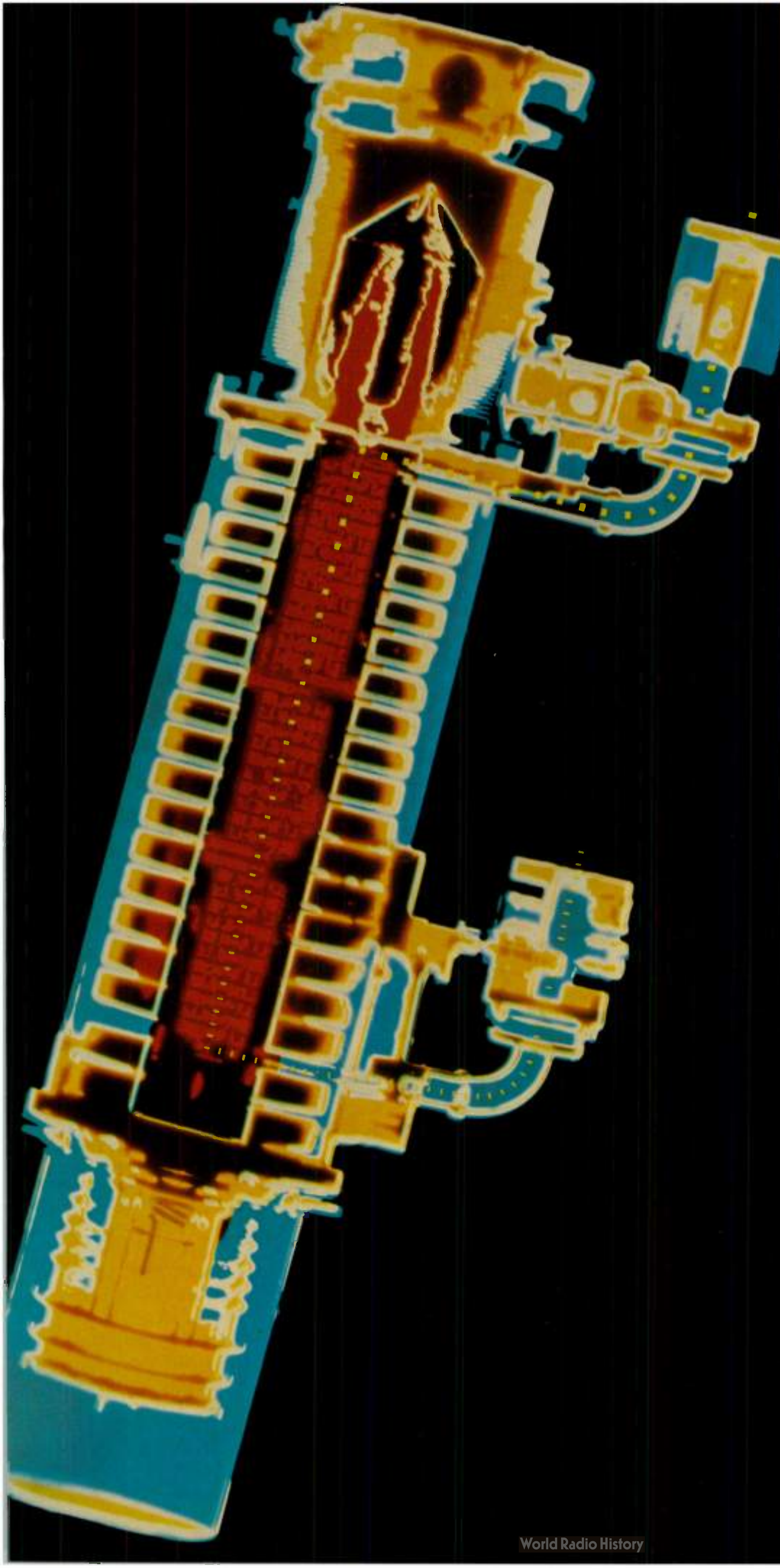
MMIC DESIGNS

An increasing number of MMIC designs are no longer patterned after conventional thin-film metal-ceramic hybrid integrated circuits, but rather are innovative designs that take advantage of unique features of MMIC technology. Here are a few representative examples of what has already been done in this area and what is likely to be possible in the future.

Actively Matched Amplifiers and Mixers-Small signal FETs take up

[*Continued on page 26*]

*A lumped element circuit is a circuit that to a good approximation can be represented from baseband to the highest frequency of interest by fixed capacitors, inductors, and resistors. In order for a circuit element to behave like a lumped element and not exhibit significant distributed properties, the rf phase across the element at the highest frequency of interest must be nearly constant, i.e., the element must be much smaller than the shortest rf wavelength of interest.



[From page 24]

little space on an MMIC chip and are therefore - - in principle at least - - inexpensive MMIC components. As a result, it is often advantageous in MMIC designs (but usually not in hybrid designs) replace relatively large passive circuits with small signal FETs. This principle has been successfully applied to active matching networks for amplifiers and mixers, where size reductions by a factor of two have been reported. Furthermore, it has been shown that at least in certain applications, active matching networks can be designed to be more tolerant to parameter variations than passive matching networks.[6,7]

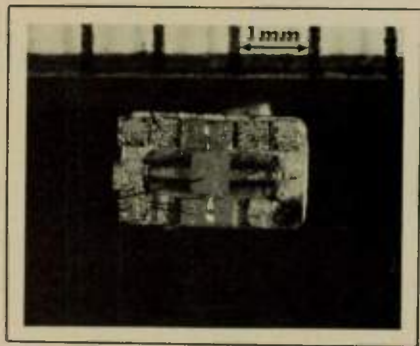
Circuits Containing Analog and Digital Functions- Among the most promising applications of MMICs circuits that combine analog and digital functions on a single chip. A good example of a circuit of this type is the 2-bit A/D converter shown in Figure 1. This circuit, which contains 26 FETs and 24 Schottky barrier diodes, has successfully operated at sampling rates of 1 GHz.

MMICs Containing Surface Acoustic Wave [SAW] Circuits- Since GaAs is a good piezoelectric material, it is possible to fabricate SAW circuits on semi-insulating GaAs substrates. Several such GaAs SAW circuits have already been successfully fabricated and tested.[8] This opens up the interesting possibility of fabricating compact, sophisticated signal processing circuits by combined SAWs and MMICs on the same semi-insulating substrate, the fabrication technologies for GaAs SAWs and MMICs being fully compatible.

*Increasing Q's with Negative Resistances-*The performance of many MMICs is limited by the inherent low Qs of their passive circuitry.* A promising approach to overcoming this limitation is to use the negative resistance that can be generated by FETs to reduce the positive resistances of the passive reactances, thus increasing their Qs. Stable Q values approaching infinity have been obtained with this technique in hybrid circuits.[9]

HYBRID INTEGRATED CIRCUITS

A new generation of hybrid integrated circuits with improved characteristics is being developed. These new circuits are certain to provide stiff competition for MMICs, particularly for MMICs of more conventional designs. Some of the new hybrid integrated circuits use lumped elements in order to achieve dramatic reductions in size.[10,11] This is illustrated in Figure 2 which shows the rf section of 1/2 W - 15 GHz lumped element GaAs FET amplifier. Note that this hybrid circuit is nearly as small as a comparable MMIC. The



Courtesy of Dr R L Camisa RCA Laboratories, Princeton NJ

Fig. 2 RF section of 1/2 watt, 15 GHz GaAs FET amplifier.

rf performance that can be achieved in many of the newer hybrid integrated circuits is usually considerably better than the best rf performance that has been obtained so far in MMICs. There are several reasons for this. Hybrid circuits have, in general, higher Qs than MMICs; in hybrid circuits, unlike in MMICs, one can individually select FETs with optimum characteristics; in hybrid circuits one can usually make small tuning adjustments to obtain optimum performance, while in MMICs this is usually not possible, etc. A good example of the type of performance achievable with modern hybrid circuits is a cascadable 4-8 GHz, 1 W GaAs FET amplifier module housed in a shielded 6x16x3.5mm package. The weight of the complete amplifier module is only about 1 gm.[12]

Much effort is going into lowering the cost of fabricating hybrid integrated circuits. One promising approach is to batch-fabricate

[Continued on page 28]

*The Qs of the passive circuitry of MMICs are inherently low because the circuitry must be very small in order to fit on an MMIC chip of practical size. Small sizes lead to high losses, and therefore low Qs.

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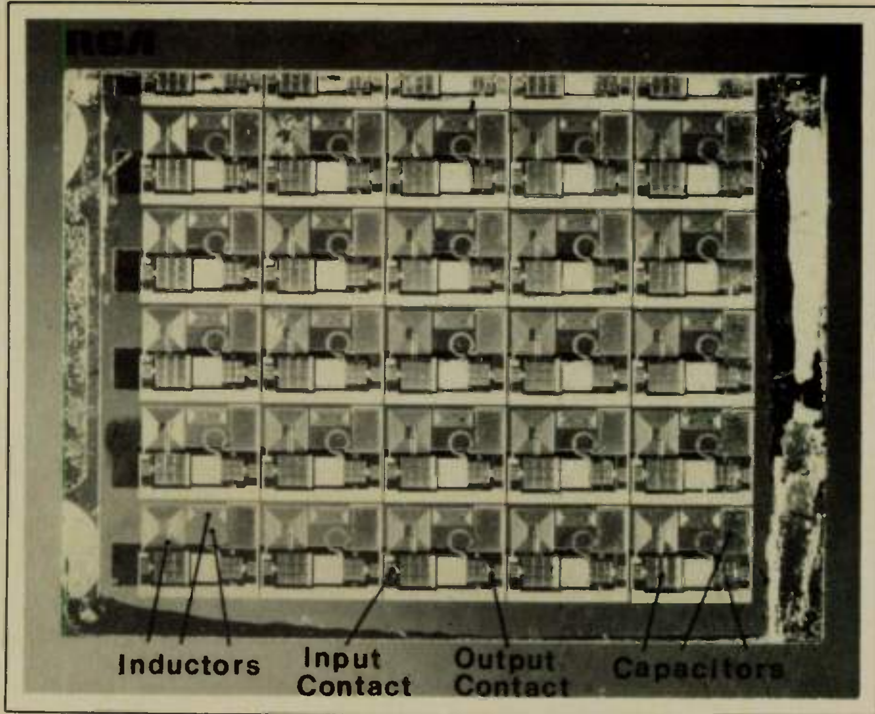

Figure 3. 25 Batch-fabricated passive circuits for S-band hybrid amplifiers. Substrate measures 3/4" x 1".

all passive circuitry and then bond discrete semiconductors into these batch-fabricated circuits. Figure 3 shows 25 hybrid circuits that were batch-fabricated on a sapphire substrate. One attractive feature of this approach is that high throughputs can be achieved with relatively modest capitol investments. In contrast, fabrication of high performance MMICs increasingly involves use of the most modern semiconductor equipments, including ion-implanters, ion-beam millers, electron-beam exposure systems, etc.

OUTLOOK

Like other types of microwave circuits, MMICs have advantages and disadvantages. Advantages include small size, batch fabrication, elimination of most wire bonds, and ability to bring matching elements very close to active devices. Disadvantages include the relatively low Qs of reactive passive elements, a difficult technology that requires many processing steps and very large capital investment, and poor utilization of GaAs substrates as far as active device area is concerned.

MMICs appear to have the best chance of succeeding in the following applications and circuits: applications that require large quantities of similar circuits and therefore can take full advantage of batch processing (phased array radars, fuzes, expendable ECM decoys, etc.); circuits that use large numbers of transistors (A/D converters, certain types of phase lock loops with digital countdown circuits, etc.), circuits with few transistors but where passive elements take up relatively little space (baseband amplifiers, FETs with internal matching, etc.); and novel circuits such as SAWs monolithically combined with FETs.

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[Continued on page 30]

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Fred Sterzer *received the BS in Physics from the City College of New York in 1951, and the MS and PhD degrees in Physics from New York University in 1952 and 1955, respectively. His PhD thesis was on microwave spectroscopy. He joined RCA Corporation in 1954 and has worked there on the development of travelling-wave tubes, optical components, high speed logic, and microwave solid-state devices and circuits. His most recent work involves the application of microwave heating to the treatment of human cancers.*

Dr. Sterzer is currently Director of the Microwave Technology Center at RCA Laboratories heading a group of approximately 85 scientists, engineers and technicians engaged in developing new microwave technologies.

Dr. Sterzer is a Fellow of the IEEE and a member of Phi Betta Kappa, Sigma Xi, and the American Physical Society. He is the author of over 75 papers and holds more than 30 patents.

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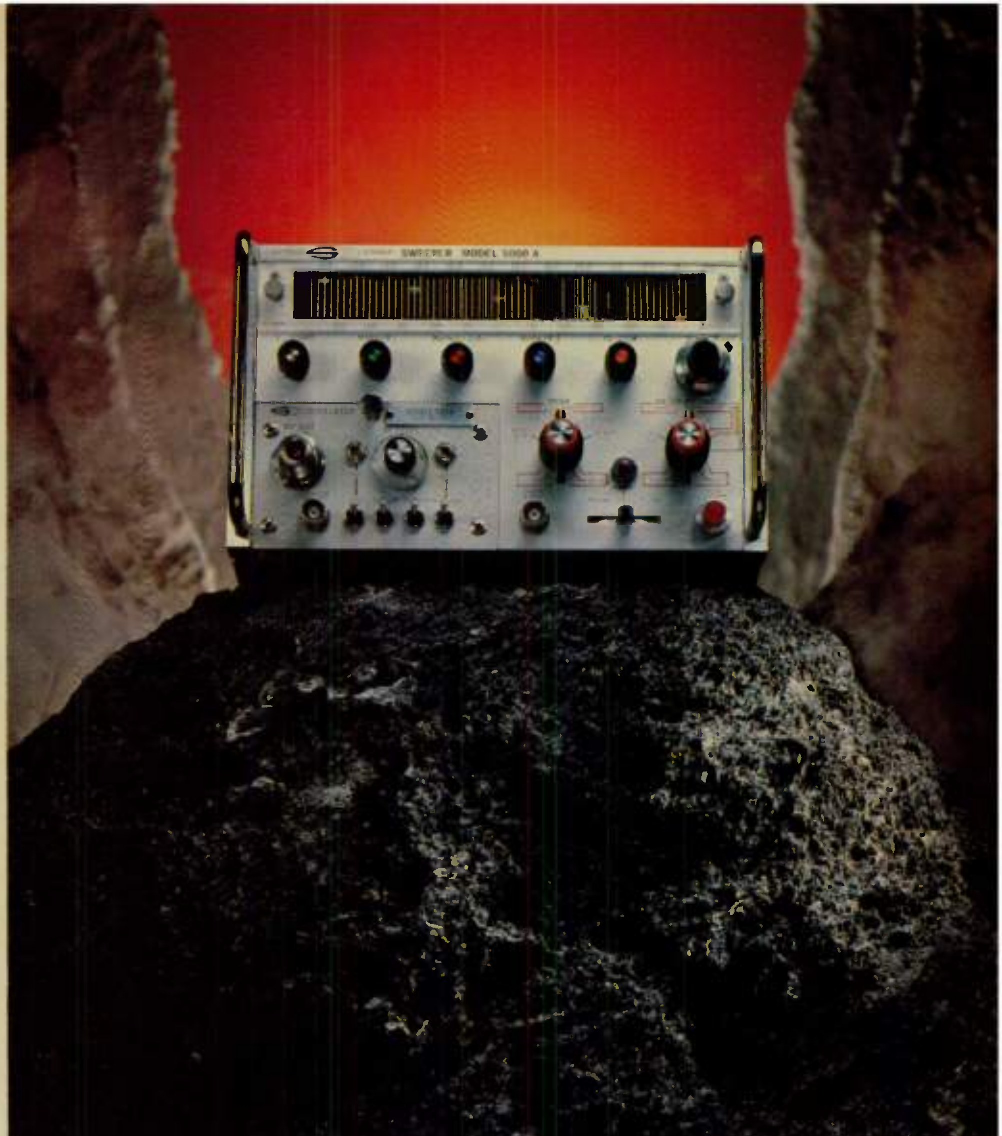

You can still get solid performance and solid value in a microwave sweeper. More than 5,000 Systron Donner 5000A mainframes are in the field. Averaging better than 100,000 service hours each. And only 43 have come back for even minor repairs.

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The 575 and 578 counters phase-lock virtually any swept signal source with an FM input to the same accuracy and long-term stability as the time-base oscillator in the counter.

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Model 578 provides coverage to 26.5 GHz standard, with options that extend frequency coverage to 40, 60, 90 or 110 GHz.

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GPIB (IEEE-STD-488-1978) is standard on the 575/578 counters, so it's easy to integrate your sweeper and counter into a controller-based automatic test system.

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A software routine, stored in the counter, automatically controls the interface to the sweeper, making interconnection a simple matter of plugging in three cables.

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And for production testing or other repetitive tests, the 575/578 can store and recall up to nine pre-programmed frequencies, so the operator need not set the desired frequency every time.

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For only a modest increase over the cost of a microwave counter, the 575/578 gives you the combined capabilities of a lockbox, a counter, a power meter, and a controller, plus spectrum analysis. And its ability to stabilize the frequency of swept sources to 110 GHz often eliminates the need for an expensive, synthesized signal generator as well.

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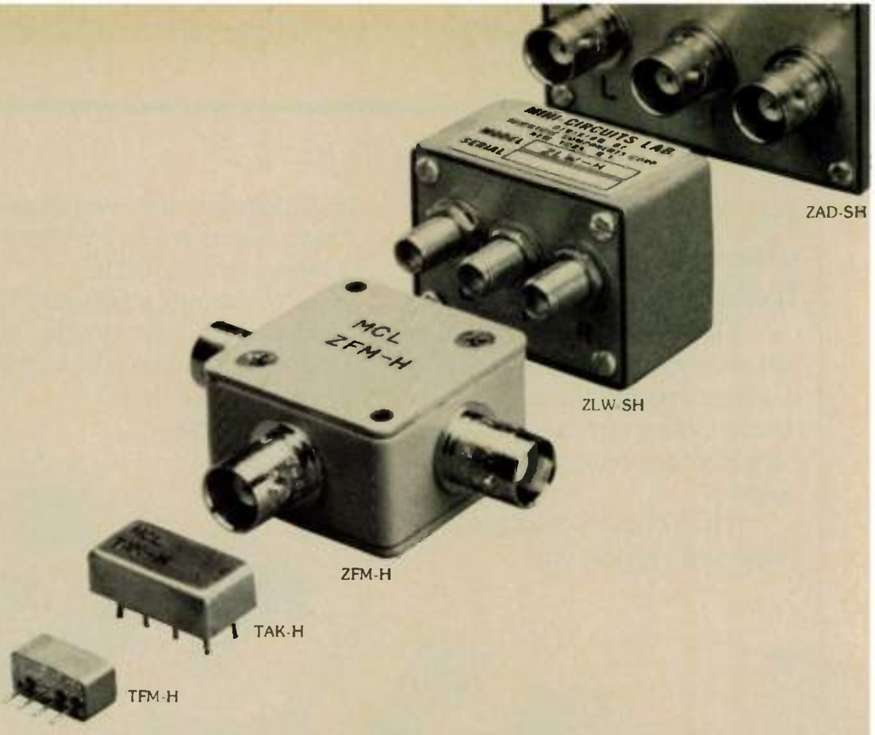

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For complete specifications and performance curves refer to the 1980-1981 Microwaves Product Data Directory, the Goldbook or EEM.

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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

58 REV. C

CIRCLE 26 ON READER SERVICE CARD

For Mini Circuits sales and distributors listing see page 41

News from ★ Washington

GERALD GREEN,
Washington Editor

ENGINEERS NOW ABLE TO CALCULATE ENERGY LEVEL OF TRANSIENT PULSES

High-energy power surges that pass through electrical equipment and power systems in the form of transient voltage pulses can cause equipment to malfunction. Furthermore, these transient pulses can damage equipment, causing burnout in transistors and silicon-controlled rectifiers, for example.

A technique, developed by the David Taylor Naval Ship Research and Development Center in conjunction with Virginia Polytechnic Institute and State University, enables electrical engineers to calculate the energy level of a transient pulse, which is an important parameter used to judge the extent of damage caused by a transient. This technique uses fast Fourier transforms to obtain accurate estimates of the energy content of these pulses.

A computer program was written to implement the technique and has been tested on several simple pulses, achieving an error between the results and the computed values on the order of 10 percent.

MANUFACTURING TECHNOLOGY DEMOS SET FOR NOVEMBER BY NAVELEX

Two manufacturing technology end-of-project demonstrations that should be of interest to electronics specialists are being planned by the Naval Electronics Systems Command.

Demonstrations of Automated Traveling Wave Tube Manufacturing (project E 024) and High Burnout Schottky Barrier Diodes (project E 124) should occur sometime in November 1981, however, specific dates have not yet been set.

For additional information, address written requests on company stationery to the Naval Electronics Systems Command, Attention Mr. R. Hill, Code 8134, Washington, D.C. 20360.

SECOND PRODUCTION CONTRACT AWARDED FOR JAM-RESISTANT VOICE COMMUNICATIONS

The Air Force Systems Command's Electronic Systems Division has announced award of a second production contract for additional HAVE QUICK jam-resistant voice communications systems for use by military pilots and ground-based controllers.

The $23.6M contract to Magnavox Government and Industrial Electronics Company of Fort Wayne, Indiana followed two production options to the original contract which were awarded to the company in 1980.

According to HAVE QUICK program manager Lt. Col. Galen Rose, systems purchased through the new production contract will be installed in a variety of aircraft — such as the F-15 and A-10 — at Tactical Air Force bases in the United States, Europe and the Pacific. The Tactical Air Forces will also use them in vans, jeeps and transportable shelters.

Military Airlift Command plans to operate the systems in a variety of helicopters and some of its aircraft, including the C-130 and C-141.

Deliveries under the new contract will begin in May 1982 and be completed by mid-1983.

News from ★ Washington

RADIO FREQUENCY TESTING STANDARD BEING DEVELOPED BY NBS

Scientists of the National Bureau of Standards (NBS) are developing a standard radiating device that government and industry can use to calibrate the accuracy of various radio frequency test facilities. The small, spherical dipole radiator is self-contained and is intended to be used as a transfer standard.

The new device has been tested successfully at 30 MHz and resonant frequencies up to 240 MHz. NBS hopes eventually to provide a frequency range from 10 MHz up to 1 GHz.

ENGINEERING RESEARCH EQUIPMENT GRANTS AVAILABLE FROM NSF

A little known fact that could be valuable to microwave specialists engaged in research is that the National Science Foundation (NSF) provides funds for research equipment as part of regular research grants, and also makes separate awards exclusively for specialized research equipment.

NSF considers the following in making research equipment awards: Quality and importance of the research for which the equipment is to be used, the appropriateness of the equipment and its expected contribution to the research, qualifications of the principal investigator and associated staff, and provisions for essential supporting facilities and maintenance of the equipment. Other considerations are the likelihood that the equipment will be useful for several different research projects and that the proposing institution or company considers the equipment sufficiently important to make a reasonable contribution of its own funds toward the projected purchase.

Additional information can be obtained by contacting the appropriate engineering division at NSF. For electrical, computer and systems, the telephone number is (202) 357-9618. For administrative inquiries, contact Ms. Janice Apruzese, Directorate for Engineering (202) 357-9834.

ELECTROMIGRATION STUDIED

The interaction of thermal and electrical currents with respect to electromigration in integrated circuits (ICs) with geometries down to 0.7 microns is being studied for the Department of Defense by Honeywell's Solid State Electronics Division (SSED) under a contract issued by the Rome Air Development Center.

Current flow in integrated circuit conductors can cause microscopic particles of metal to move from one point to another along the conductor. This phenomenon, called electromigration, can cause voids in some spots thereby breaking conductor continuity. Difficulties due to electromigration have been largely circumvented for IC geometries of five microns and larger. The problems, however, increase significantly as conductor widths shrink to submicron size. (A strand of hair is about 75-100 microns thick).

Results from the electromigration study should be applicable to Very Large Scale Integration (VLSI) conductive film development throughout industry.

Honeywell is also involved with DoD's VHSIC (Very High Speed Integrated Circuits) program and is currently working on the Phase 3 portion of that program. ■

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CIRCLE 32 ON READER SERVICE CARD

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1.535–1.66	AM-1660	2.5	2.0
3.7–4.2	AWC-4200	1.5	1.0
7.25–7.75	AW-7720	2.5	2.0
10.7–11.7	AW-11700	4.0	3.0
11.7–12.2	AW-12200	4.5	2.5

Intermediate Power

Frequency, GHz	Series	Po (1dBGCP) Std.	Best
1.0–2.0	APG-2000	1 W	2 W
5.9–6.4	AMP-6420	1 W	5 W
7.9–8.4	AMP-8420	1 W	2 W

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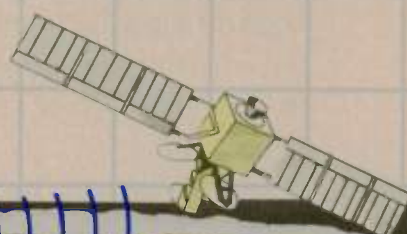

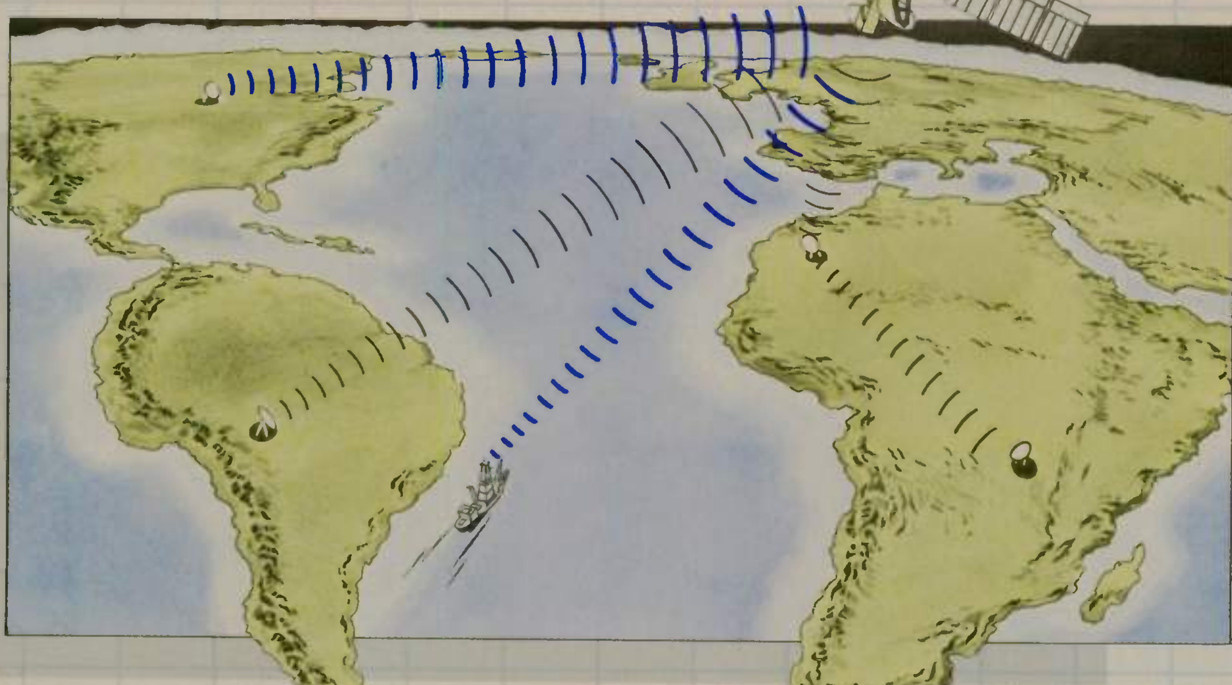

Avantek

International Report

GERALD GREEN, *Washington Editor*

CANADIANS TESTING NEW SATELLITE COMMUNICATION SYSTEM

Canadian government scientists and engineers are testing a new satellite communications system for Canadian Forces ships.

The flexible, low-cost UHF (ultra-high frequency) satellite terminal, which may be ready for widespread use by 1984, will be used by Canada's navy.

The experimental shipborne terminal is expected to provide reliable radio communications for Canadian Forces ships and, at the same time, be operated with various types of UHF radio equipment used by allied navies.

The system transmits a variety of data rates, making it compatible with transmission from Canada's proposed MSAT satellite or from the U.S. FLTSATCOM, GAPFILLER or LEASAT satellites.

The system is currently undergoing testing by the Canadian Department of National Defence (DND) which has funded the project. The department hopes to transfer the technology to Canadian industry through contracts for development of an industrial prototype and equip Canadian Forces ships with the terminal starting in 1984.

UNIT FORMED TO ASSIST SMALL COMPANIES INTERESTED IN INVESTING IN SCOTLAND

Electronics companies with sales of less than $100 M interested in markets in Europe can now find help from a group formed specifically to service their needs.

A New Ventures Unit, operating out of Glasgow, has been established by the Scottish Development Agency to assist small companies which are looking to invest in Scotland for the first time or are looking for joint venture opportunities with Scottish companies.

The unit was set up primarily to provide venture capital, through stock, loans and financial guarantees, to new and developing high-technology companies. According to its director, Ken Smith, the unit will help make it financially possible for small companies to come to Scotland, which, with over 200 companies, has one of the largest and most diverse electronics industries in Europe.

The unit also can provide information on British and European markets for company products and services and can help provide specialist help and advice on locating sites and resources.

Smith said, "We're not just advisors, we're prepared to put up money for electronics ventures and we're not afraid to take risks."

A Scottish Development Agency has been set up in San Francisco and New York to provide specific information to interested U.S. firms.

BRITISH AND DUTCH INDUSTRY COMPETING FOR RADAR BUY

Britain's Ministry of Defense (MOD) is taking a lot of heat these days concerning procurement of light-weight missile tracking radars for Royal Navy frigates.

Britain's Marconi Radar Systems Company is claiming that Britain's defense industry will be severely damaged if the Ministry of Defense buys a Dutch radar system rather than one built by Marconi.

The Dutch system, manufactured by the Hollandse Signaal Apparaten Company, a subsidiary of Phillips, is being evaluated by MOD for use aboard British frigates along with the competing Marconi system.

International Report

BRITISH TELECOM ORDERS MICROWAVE LINKS FOR LONDON

In an order valued at over 1 million pounds, British Telecom is to purchase microwave radio links for installation in London.

With 2 ft. parabolic antennas, the microwave radios will be installed on London roof tops in the first phase of B.T.'s 17 million pound program to provide the capital's business community with one of the most advanced telecommunications services in the world.

To be built by the Farinon Division of the Harris Corporation in San Carlos, California, for delivery by the end of 1981, the order will comprise eighty 19GHz digital microwave transmitter/receivers as well as the antennas, spares and test equipment.

Housed in compact weather-proof cabinets, the links have a range of up to 10 kilometers and will carry 30 or 120 voice channels, each voice channel pulse code modulated at 64K bits.

B. T. spokesmen cite two principal advantages in using microwave radio links in London. First, the high transmission speeds, 2 or 8 million bits per second, suitable for voice, data and possibly in the future, video transmission or video phone; and secondly, the speed and ease with which the links may be installed and brought into operation.

UK'S CIVIL AVIATION AUTHORITY CALLS FOR MONOPULSE INTERROGATOR

A monopulse radar interrogator has been specified by the United Kingdom's Civil Aviation Authority (CAA) for their new radar replacement program.

The interrogator, the SSR 950, built by Cossor Electronics Limited of England, was recently displayed at the Asian Exhibition and Conference in Singapore.

The monopulse radar overcomes many of the problems of signal interference encountered with current radars operating in England's high density air traffic. It has been tested as part of ADSEL, the selective address system that Cossor is developing with the Civil Aviation Authority for use in air traffic control later in the eighties.

DOD PLANS TO PROVIDE SAUDI NAVY WITH SHORE-BASED C^3 SYSTEM

While Congress and the Reagan Administration were embroiled in the headline dispute over the proposed sale of AWACS and enhanced F-15s to the Saudis, a steady stream of other military hardware sales, including a sophisticated Command, Control, and Communications (C^3) system for the Saudi Navy, was proposed by the Department of Defense.

The proposed C^3 sale took the form of an amendment to an existing Letter of Offer to the Government of Saudi Arabia for the sale of the services and materials required to complete the development, installation, and check-out of the Saudi Naval Forces C^3 system; installation of a microwave interconnect system between the Saudi communications stations at Jidda and Jubail; support services; and completion of communications stations at Jidda, Jubail, and Riyadh at an estimated cost of $180 million.

The prime contractors for the C^3 sale will be Science Applications, Inc. of McLean, Virginia, and Page Communications Engineers, Inc. of Vienna, Virginia.

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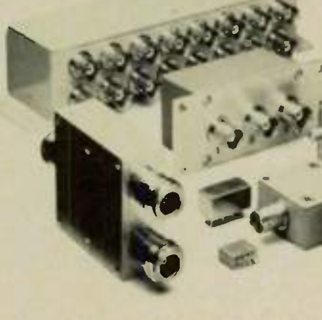

For complete specifications and performance curves refer to the 1980-1981 Microwaves Product Data Directory, the Goldbook or EEM.

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CIRCLE 30 ON READER SERVICE CARD

Around the Circuit

PERSONNEL

George Birutis has resigned as Pres. of Comtech Antenna Corp. to assume the same post at Adria Communications Group, Inc. . . The newly elected V. P. of ITT Avionics Division is **Ralph L. Asher . . . Dr. Robert Fossum** has been appointed V. P. of Avantek, Inc.'s Telecommunications Div., and **Mr. Robert Mullaley** who previously held that post is now V. P. -Business Development . . . **Jack Barbera** has been promoted to V. P. of the Times Wire & Cable Div. of Times Fiber Communications, Inc. . . At Systron Donner, **Jack Relchel** has been appointed Corporate V. P. -Sensor and Systems Group, and **Dr. Lawrence A. Wan,** General Manager-Electronic Systems Div. . . **Anthony A. Martinelli** has been appointed general sales manager of Scientific-Atlanta Inc.'s instrumentation group.

NEW MARKET ENTRY

Scientific Devices, Inc., a new corporation headed by **John A. Caruso,** formerly General Manager of Microwave Semiconductor Corporation's Diode Division, has purchased the business of that division. Manufacturing and engineering operations for the entire division product line will be continued at the North Billerica, MA plant and all existing warranties, orders and quotations for its products will be honored. Address, Telephone, Telex and TWX numbers formerly applicable to the MSC Diode Division will be assumed by Scientific Devices, Inc.

INDUSTRY NEWS

RHG Electronics has appointed the **M. Lader Co.** to handle its line of MIC microwave and IF/RF products in the Washington, D.C. and surrounding markets . . . **California Microwave, Inc.,** has an agreement-in-principle with **Dexcel, Inc.,** to exchange 887,000 shares of its common stock for all of the outstanding shares of **Dexcel . . . Chomerics, Inc.** has established its European headquarters, Chomerics Europe, in the London suburb of Hendon where it will house the Materials and Shielding Technology (formerly Metex Electronic Shielding Products) divisions. **Amplica, Inc.** has reorganized its manufacturing divisions with the Telecommunications Products Div. supporting the broadcast, commercial and home satellite T.V. markets, while the Defense Electronics Div. will serve the military electronics industry. **Adams-Russell, Co.,** has completed the acquisition of **Microwave Products, Inc.,** of Chatsworth, CA.

CONTRACTS

Ford Aerospace & Communications Corp. will produce an additional three INTELSAT VA international communications satellites. The contract is valued at more than $75M . . . **EPSCO, Inc.,** has received a contract for $620K to design and manufacture high power microwave instrumentation for a major medical equipment manufacturer . . . **The Sperry Division** of Sperry Corp. has received a $1.6M contract from the US Naval Air Development Center to integrate the NAVSTAR Global Positioning System (GPS) capability into the current navigation system aboard a US Navy aircraft carrier . . . **TRAK Microwave Corp.,** a subsidiary of Tech-Sym Corp., has received a sub-contract from Raytheon Co. valued at more than $1M for the second production release of crystal controlled oscillators for the AIM-7M Advanced Monopulse Sparrow Missile . . . **Loral Corp.** has been awarded a $9M contract to supply the US Army with an improved Radar Warning System for its fixed wing and rotary wing aircraft. Negotiations were concluded for additional systems and services valued at $4.5M which the company expects the Army to exercise before the end of 1981. . . **Itek's Applied Technology Division** received U.S. Air Force contracts totaling $16M for ALR-46/69 radar warning systems and ground support equipment, with deliveries slated to begin in mid-1982 . . . **Microdyne Corp.** has received a $6M contract from White Sands Missile Range to supply microprocessor controlled telemetry receivers and diversity combiners.

FINANCIAL

Adams-Russell reports third quarter earnings of 32¢ per share compared with 26¢ per share over the same period in 1980. Net income for the quarter was $1.2M on sales of $12.9M compared to last year's income of $840K on sales of $10.7M in the same quarter . . . **Microdyne Corporation** reports third quarter sales of $6.8M, up from $6.1M for the comparable period in 1980. Net income of $1.1M or 24¢ per share is up 39% from $822K or 20¢ per share in the third quarter of 1980. . . **Watkins-Johnson Company's** Board of Directors declared a quarterly dividend of 12¢ per share payable to stockholders of record October 1, 1981. . . **M/A COM, Inc.,** reports third quarter net income of $11.2M or 30¢ per share as compared with figures for the same period in 1980 of $7.9M or 23¢ per share. Sales for the third quarter were $132M up from $93M reported for the same period in 1980. . . **AEL Industries, Inc.** reports a net loss from continuing operations of $1.1M or 76¢ per share on sales of $12.5M for the secnd quarter ending August 28, 1981 compared with the similar period in 1980 when income from continuing operations was $449K or 26¢ per share on sales of $14.1M. ■

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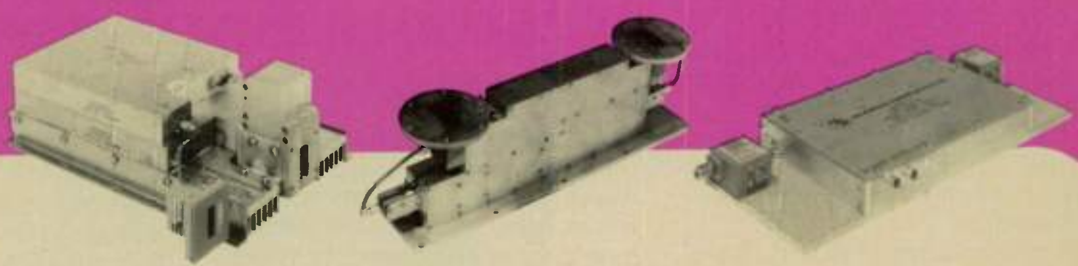

50 watts, 7.9-8.4 GHz.
Class A linear
GaAs FET power amplifiers.
Ideal TWT replacements.

During the past few years, lots of people have been doing lots of talking about the merits of GaAs FET amplifiers. But for quite some time, we've been way beyond the talking stage. For example, we are the only company—*repeat, the only company*—that has already delivered production units of GaAs FET Class A linear power amplifiers with power outputs of 10 watts in the 7.9-8.4 GHz frequency range! And we're now ready to deliver up to 50 watts! We also have models for other telecommunications and radar frequency bands including 3.7-4.2, 4.4-5.0, 5.9-6.4, 7.1-7.7 and 9.0-10.0 GHz.

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We have delivered large production quantities of these amplifiers for a wide spectrum of commercial and military applications such as: satellite ground station up links, point-to-point microwave drivers and final stage transmitters, drivers for airborne EW systems, ECM jammers, EW simulators and test sets.

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ERADCOM Microwave Device and Circuit Contracts

V.G. GELNOVATCH
US Army Electronics Technology and Devices Laboratory
Fort Monmouth, NJ

Texas Instruments under Contract DAAB07-78-C-2966 has delivered a 4.5 watt CW GaAs FET Ku-band transmitter package for airborne data link application. This 16 cubic inch transmitter demonstrated 2 GHz bandwidth, 18% overall efficiency, 26 dB gain and 1¼ lb weight. The output was achieved by 8 circuit paralleled 600 um gate width devices. The shorter gate width devices gave larger bandwidth performance even though more devices were required.

A program with Hughes, Contract DAAK20-80-C-0527, to develop a 3.5 watt Ku-band GaAs FET and a 6 watt amplifier with 10 dB gain and 30% efficiency has produced 2.15 watts at 15 GHz, 5 dB gain and 24% efficiency. This was obtained from two - 3 cell devices using a 6 way combiner/splitter. Each cell was 600 um by .7 um, optically defined gate device. This program is a back up to eventually obtaining a 10 W 30% efficiency data link transmitter for RPV or other air mobile platforms.

Under the US Army monolithic GaAs IC thrust a recently completed contract with Rockwell (DAAB07-78-C-2999), a Ku-band monolithic receiver on a ¼ x ¼ inch, GaAs chip has been delivered. The constituant circuits of this receiver were an RF preamplifier, a dual gate FET mixer and a wideband IF amplifier (1 - 10 GHz). An LO was not required for this early feasibility effort. For a complete report, the reader is directed to the March 1981 issue of the *Microwave Journal*. To overcome the limitations of the previous contract (i.e., lack of a local oscillator) a contract at TRW (DAAK20-80-C-0279) will develop, in monolithic format, the key frequency conversion elements for a Ku-band front end. Accordingly, the objective is the development of a voltage tuned local oscillator over 13 - 15 GHz with at least 20 mW output, a 14 - 16 GHz monolithic mixer with 6 dB SSB noise figure and 1 GHz IF. Progress to date has been the completion of the first mask set and several wafer lots using discrete components have been tested. Single gate device performance has been good but dual gate devices are not operating to design objectives. Sample L's and C's have been evaluated and Q's of 12 - 16 demonstrated. These Q's will be adequate. Second generation masks have been ordered. In the department of lessons learned, it can be stated that microwave circuit design (monolithics), just as discrete device design in the past, is now at the mercy of the mask makers with 3-6 months turn around time required between iterations.

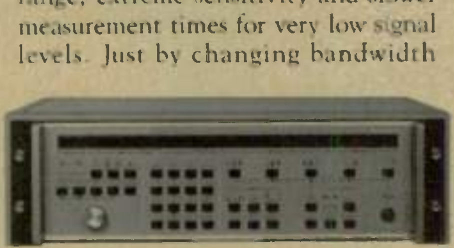
Avantek under Contract DAAB20-80-C-0284 is developing a K-band low noise GaAs FET with an objective of 2 dB noise figure at 22 GHz with 8 dB gain. Progress to date has been an optically defined .3 um gate device (75 um wide) using three feeds which has achieved 3.5 dB noise figure. The problems encountered in achieving the required gain and noise figure appear to be in obtaining the right channel doping profile and adequate buffering from impurities in the substrate. Currently gettering is being used to push chrome impurities to the back of the substrate. New material sources and profiles will be investigated.

A new MMT program to establish production processes and in line controls needed to produce multicell bipolar transistors has been awarded to Microwave Semiconductor Corp under Contract DAAK20-81-C-0386. Objectives are 30 watt power output, 6.5 dB gain and 40% collector efficient over the 3.1 to 3.5 GHz band.

Norden is developing a 17 GHz 40 W peak power solid state amplifier for the MARFS missile seeker, DAAK20-81-0377. To date a single diode coax cavity IMPATT amplifier has been developed which is stable in the band and produces 2.5 watts of output power. A coax-capacitively coupled cavity combiner has been demonstrated for combining enough diodes to achieve the desired output power over the full 1000 MHz bandwidth. Small hybrid modulator modules have been fabricated. Future plans include a breadboard which will combine in parallel to achieve 40 W output and in cascade to achieve the required overall gain.

In the millimeter wave arena, Alpha Industries has been recently awarded an MMT contract (DAAK20-81-C-0403) to establish low cost production techniques for millimeter wave mixers at 56, 94 and 140 GHz utilizing precise and accurate control of device processing and test procedures. This includes effort in the area of material processing and evaluation, Schottky contact and beam lead metallization, and device packaging and testing. The objective is low cost

[*Continued on page 46*]



[From page 44] **CONTRACT NEWS**

reproducible mixers that are small, light and tolerant to harsh environmental factors. This program represents an attempt by the Army to expand the millimeter wave industrial base. This is a new effort.

Hughes has been awarded a contract to continue work in developing high cutoff monolithic beam lead Schottky barrier mixers at 60, 94, 140, and 220 GHz using advanced semiconductor processor techniques such as E-beam lithography, proton bombardment and MBE under Contract DAAB20-81-C-0417. This is a new contract and there is no progress yet. Also in the new contract award status is an effort awarded to Microwave Associates (DAAK20-81-C-0395) to design and develop an all solid state duplexer which will be able to stand off 1 KW peak and 50 W average transmitter power levels at 94 GHz. Objectives are to limit spike leakage to .005 ergs (50 mW) and flat leakage to 5.0 mW.

As an additional effort to expand the industrial base, a contract has been awarded to Martin Marietta to develop an advanced concept for beam leading silicon IMPATT diodes at 94 GHz (DAAK20-80-C-0308). Progress to date has been the development of a 3 watt beam lead device, development of a quartz ring package and specialized tools required for assembly and the fabrication of a complete package. Currently RF measurements are being made. Future effort will concentrate careful evaluation of the uniformity properties of these devices in relation to non-beam lead diodes.

In 1980 Hughes (DAAK20-80-C-0316) was awarded an MMT program to establish production techniques for 56, 94, and 140 GHz silicon IMPATT devices. Current first engineering samples have just been delivered consisting of one oscillator and 10 diodes at each of the three band (V, W, D). CW power output ranged from 775 mW at V-band to 540 mW at W band to 80 mW at D band. Preformed beam lead ribbons will be added and sample uniformity will be evaluated in the future. A pilot run will be established at 94 GHz and a contractor demonstration will follow. ■

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Avantek creates more active area on our power FETs by using our unique airbridge interconnects. This provides better heat dissipation and greater reliability. It also means you get more power out with less power in! Avantek's rugged FETs have all-gold metallization and are space-qualifiable.

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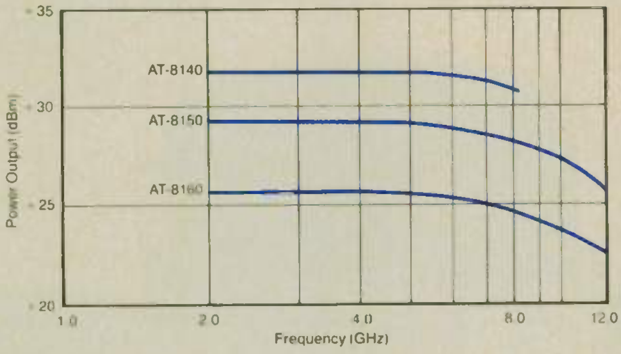

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As with all **Optimax** hybrid modules, these TO-8 amplifiers are a product of Alpha's unique high volume processing technique. Their high output power, wide bandwidth, and superior efficiency make them well suited for such demanding applications as manpack radios, missile telemetry, and radar receivers.

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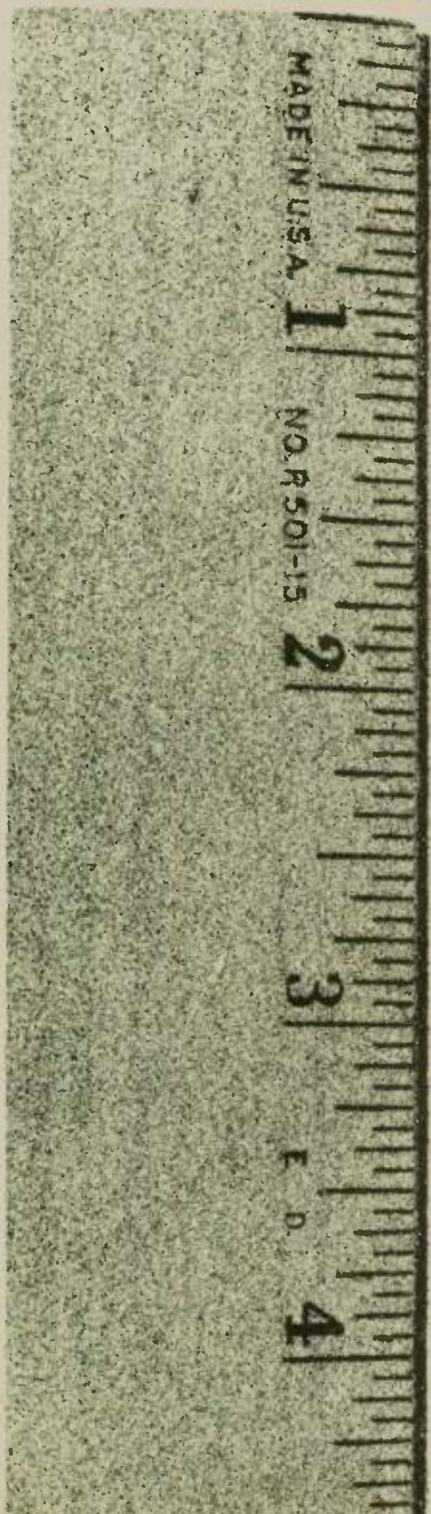
		Small Signal Gain dB		Gain Flatness ±dB	Power Output At 1dB Compression dBm		Intercept Point dBm	Noise Figure dB		VSWR In/Out	D.C.	
Model	Frequency Range MHz	Typ.	Min. -54/85°C	Max. -54/85°C	Typ.	Min. -54/85°C	Typ.	Typ.	Max. -54/85°C	Max. -54/85°C	Volts Nom.	mA Typ.
AP-1	10-100	13.0	11.5	1.0	26.5	24.5	36.0	8.0	10.0	2.0	24	95
AP-2	10-300	11.0	9.0	1.0	25.0	23.5	38.0	8.0	10.0	2.0	24	95
AP-3	5-500	14.5	13.0	1.0	24.5	22.5	34.0	7.5	9.5	2.0	15	130
AP-3-1	5-500	14.5	13.0	1.0	22.5	20.0	38.0	7.0	9.0	2.0	15	98
AP-10	10-1000	10.0	9.0	1.0	22.5	20.0	36.0	8.5	10.0	2.0	15	95
AP-10-1	10-1200	9.5	8.5	1.0	22.5	19.5	35.0	8.5	10.0	2.0	12	95
AP-12	10-1200	9.5	8.5	1.0	22.5	19.5	35.0	8.5	10.0	2.0	15	95
AP-12-1	10-1200	9.5	8.5	1.0	22.5	19.5	35.0	8.5	10.0	2.0	12	95
AH-29	10-1500	10.0	7.0	1.0	20.0	19.0	34.0	7.5	13.5	2.0	15	90
AH-39	10-2000	10.0	6.0	1.0	22.0	20.0	34.0	7.0	10.0	2.5	15	90

Alpha

The Alpha Advantage

Microprocessor Based Fault Finder Pinpoints Transmission Line Faults Within Inches . . .And Within Minutes



ASAD MADNI
Program Manager
and
DR. ZOLTAN TARCZY-HORNOCH
Research Manager
Systron Donner Corp.
Microwave Division
Van Nuys, CA

INTRODUCTION

Historical methods for evaluating transmission line systems are: measurements of the voltage standing wave ratio (VSWR) by a conventional reflectometer set-up, by a time domain reflectometer system (TDR), or by swept frequency techniques. Useful as they are, certain limitations in each of these approaches have precipitated the need for a sophisticated measurement system which can overcome these limitations.

The conventional reflectometer technique[1] works by feeding an RF signal into the transmission line under test by using a slotted line and observing the minimum and maximum amplitudes of the standing waves on the line. When the system includes several discontinuities, the VSWR measurement does not resolve them. Furthermore, in the case where one discontinuity generates a reflection whose phase and magnitude partially or even fully cancels the reflection of a second discontinuity, the reflectometer gives a false measurement of the actual quality of the system. To avoid this problem, one could use the very time-consuming process of repeating the measurement at widely different frequencies. The evaluation of this data would require considerable operator skill.

In the time domain reflectometer (TDR)[2], an extremely narrow pulse is fed into the test system and the total transit time of the reflections from the discontinuities is observed. Reflections occur each time the pulse encounters an impedance mismatch. The time required for the reflected pulse to arrive at the point of pulse insertion, determines the location of the fault. The magnitude of the reflection as compared to the incident pulse determines the degree of mismatch. The TDR method, however, is basically broadband testing. A typical 50-picosecond rise time TDR pulse for example, implies a 7.0 GHz bandwidth. Testing low frequency or narrowband communication systems with these pulses is very difficult or not possible. In case of multiple discontinuities where the already reflected pulses are re-reflected, the evaluation of the CRT display may become too difficult even for an experienced user.

The swept frequency measurement[3] is relatively easier than the conventional VSWR measurement since no slotted line is required. The detector is stationary and the frequency is swept generating standing waves which are typically displayed on a CRT. Multiple discontinuities again require great operator skill for interpretation and can be quite error prone.

The use of microprocessor based hardware and advancements in digital filtering and signal processing techniques have made it possible to not only overcome many of the existing limitations, but have led to the development of a very sophisticated "intelligent" instrument capable of characterizing transmission lines within minutes. The Systron Donner Model 5220 Transline Analyzer utilizes frequency domain techniques which involve digital processing of swept RF signals. A standing wave pattern is produced by interactions of a swept frequency incident and reflected RF signal as in the previous tech-

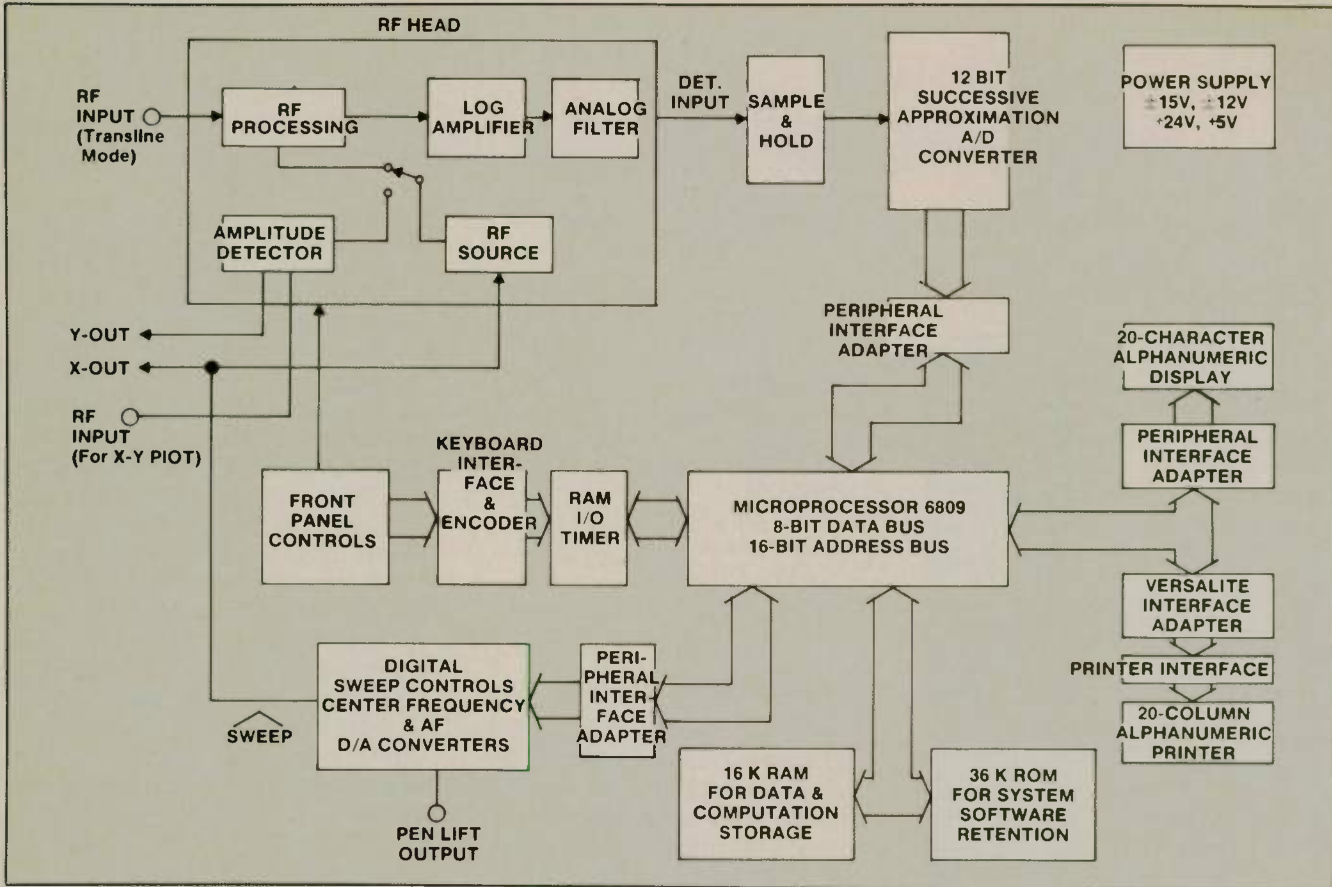


Fig. 1 Overall block diagram.

nique. The sweep is relatively narrow band (40 MHz). Test frequencies as low as 2 MHz can be processed to characterize the transmission line system.

The detected waveform, when spectrum analyzed, contains information on each fault within the transmission line system under test. Spectrum analysis is accomplished by digitizing the detected waveform, and processing the digital information by a Fast Fourier Transform (FFT)[4] algorithm. Each spectral peak represents the distance of a transmission line fault, the amplitude is a function of the VSWR caused by the fault. A significant advantage of the frequency domain processing arrangement is the instrument's ability to detect multiple faults in one pass and to reject interference signals by digital filtering and correlation techniques.

A special advanced version of Systron Donner's Model 5220 Transline Analyzer was developed under the funding of Naval Surface Weapons Center (NSWC), Dahlgren, Virginia, for the Combat Readiness Electromagnetic Analysis Measurement (CREAM) program and has been assigned the nomenclature AN/PSM-40 by the United States Navy.

The Transline Analyzer is comprised of the mainframe unit and several RF heads. These heads are provided in octave bandwidths to cover the frequency range from 1 to 26 GHz, one additional head covers the range from 2 to 1000 MHz.

The Analyzer contains an alphanumeric display and a hard copy printer which, in conjunction with specially designed human interface software, provides the user with step-by-step instructions and prompts, guiding him through the entire measurement. The total measurement time to characterize a given transmission line setup is typically 5 minutes. The system performs a complete self-check during the calibration routine. In addition, it computes the complete system VSWR, return loss, line attenuation, and the location and magnitude of each fault along the line. The measurement parameters, status, and results are provided on the display and a hard copy printout. Additionally, a second measurement mode is available and is referred to as the X-Y mode of operation. This consists of a reflectometer which plots return loss vs. frequency. The user has the ability to select the start and stop frequencies and control the sweep time. In this mode of operation, the system may be interfaced with an X-Y display or plotter. Each of the measurement routines will be discussed in detail in the following paragraphs.

SYSTEM BLOCK DIAGRAM

Figure 1 shows the simplified overall block diagram of the Transline Analyzer test set. A swept frequency source (RF head) is used to generate the test signals for the transmission line system under test. The RF head in addition to the swept source, contains all of the RF processing circuitry required to process the detected standing waveform pattern set up by the impedance mismatches along the transmission line. The frequency of the signal source is set up by a linear ramp function tuning signal generated by the sweep generator section under microprocessor control. The detected analog waveform is passed through a logarithmic amplifier to provide increased dynamic range. This, in addition to a digital AGC algorithm provides the necessary dynamic range (>90 dB) to make meaningful measurements under widely varying attenuation conditions. As an example, typical shipboard communication lines have 10 to 40 dB directional couplers installed at the test ports, making other measurement techniques almost not usable.

The output of the log amplifier, after signal conditioning, is sent to the sample-and-hold circuit, which under digital control and in synchronization with the sweep, collects 2^N (typically 1024) analog samples of the waveform over the specified swept bandwidth (ΔF). These analog samples are converted to digital format by a 12-bit successive approximation analog-to-digital converter (ADC). The output of the ADC is presented to one of the ports of a peripheral interface adapter (PIA) and under processor control is stored in the Random Access Memory (RAM). The entire system software is contained in Read-Only-Memory (ROM). The data values stored in RAM are digitally processed and the final results are decoded, displayed on the front panel, and printed out by the printer. The system is controlled by a 6809 microprocessor.[5]

RF SIGNAL PROCESSING

All of the RF signal processing circuitry is contained in the RF head. It consists of the signal source, the Transline processing section, and the X-Y (return loss vs. frequency) processing section. Figure 2 shows the simplified overall block diagram for the 2 to 1000 MHz RF head. The RF head is basically a homodyne receiver in which the transmitted signal is also used as the LO. The signal source generates the variable 2 to 1000 MHz test signal that is internally leveled and produces an RF output of +0 dBm. A mode select switch controls a relay in the RF head and connects the source to either the power divider for the Transline processing or to the impedance bridge for the X-Y mode of operation.

The output signal is derived from a down conversion process of mixing signals from a variable 2.002 to 3.000 GHz YIG oscillator (f_1) with a fixed 2.000 GHz signal (f_2) multiplied by a comb generator from a 100 MHz crystal oscillator. The advantages of this frequency are:

- A linear tuning function inherent in a YIG-tuned oscillator eliminates the need for complex linearizing hardware.
- A single oscillator is sufficient to cover the entire 2 to 1000 MHz frequency range.

A 1 GHz low-pass filter is used at the IF port of the mixer to attenuate all harmonics and subharmonics by 30 dBc and spurious by 40 dBc. The IF signal output is boosted by a broadband amplifier. The output of the signal source is leveled by sampling through a directional detector. The voltage at the detector output is compared to a reference voltage in an error amplifier. The error signal is used by a PIN diode attenuator to correct for level variations. The leveling loop, as shown in Figure 2, does not level either the tuned or 2 GHz fixed frequency source per say, but does level the port of interest, the 2 to 1000 MHz signal, and results in a somewhat simpler implementation.

When the Transline processing section is selected, the 2 to 1000 MHz test signal (through a power divider) serves as the LO input to a mixer/detector module which is

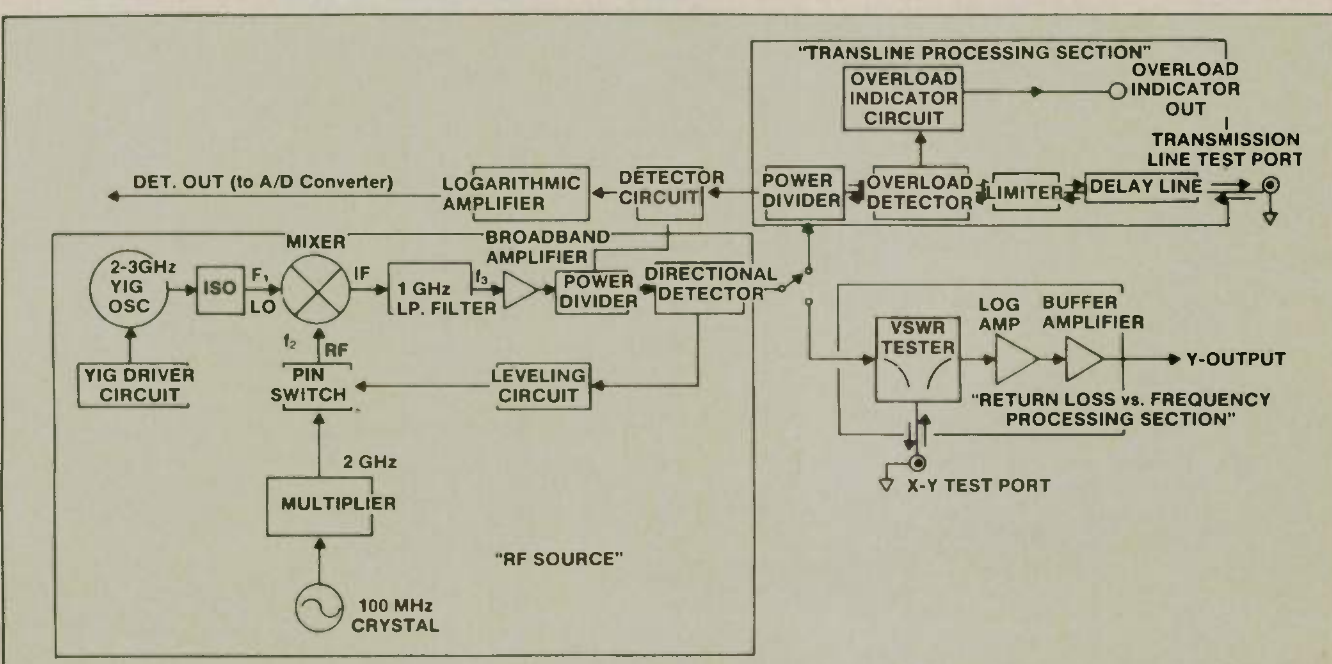


Fig. 2 RF head block diagram.

[*Continued on page 52*]

[*From page 51*] **FAULT FINDER**

part of the detection circuitry. Additionally the reflected power from the transmission line under test appears as the RF signal to the mixer/detector module. The IF output from this detection circuitry contains ripple waveform frequencies which are digitally processed. This IF signal is the standing wave pattern set up by the signal source and the reflected power from faults in the transmission line under test. As noted previously, sweeping the signal source produces a dynamic standing wave whose frequency is a function of the swept bandwidth and distance to fault, and whose amplitude is a function of the reflection coefficient of the fault. A bandpass filter at the output of the mixer/detector module filters out the low frequency components caused by connector mismatch between the RF head and the line under test and high frequency components which are above the range of interest. Included in the Transline processing section are a power limiter and a calibration cable. The limiter is used to protect against high power 'foreign signals' and turns on an overload indicator denoting that the front end is saturated. The built-in coaxial delay line is utilized for calibration. The propagation constant and attenuation characteristics of the delay line are stored in memory for data correction purposes.

When the return loss vs. frequency processing mode is selected, the measurements are made by using an impedance bridge with at least 40 dB directivity to process the amplitudes of the incident and reflected signals. The incident signal which propagates down the transmission line under test passes through the bridge to the output signal port. The returned signal reflected back through the test port is added to the incident wave by the impedance bridge. The result is detected by a built-in crystal detector, and log amplifier. This output signal is a linear function of power in dB and is presented directly to the Y-output jack on the rear panel of the analyzer where it can be utilized for external X-Y recording or display. The dynamic range in the X-Y measurement mode is restricted by the directivity of the bridge.

DATA ACQUISITION

In this system, data acquisition is the process of sampling the detected ripple wave, converting the samples to digital format, and storing them in RAM for digital signal processing. In general, the required system sampling rate is determined by the Nyquist criterion which is mathematically expressed as: $f_s \geq 2f_h$ where f_s is the sampling frequency and f_h is the highest frequency component in the signal waveform. From the Nyquist sampling theorem, a minimum of two samples per cycle of the data bandwith is required in an ideal sampled data system to reproduce the sampled data with no gross distortion of the information. Thus, an important consideration in determining the system sampling rate is "aliasing"[4] error, i.e., errors due to information being lost by not taking a sufficient number of samples per cycle of signal frequency. Figure 3 illustrates "aliasing" error caused by an insufficient number of samples per cycle of data bandwidth.

The sampling rate of the Transline Analyzer is determined by the fact that in order to generate a precise FFT spectrum, 1024 sam-

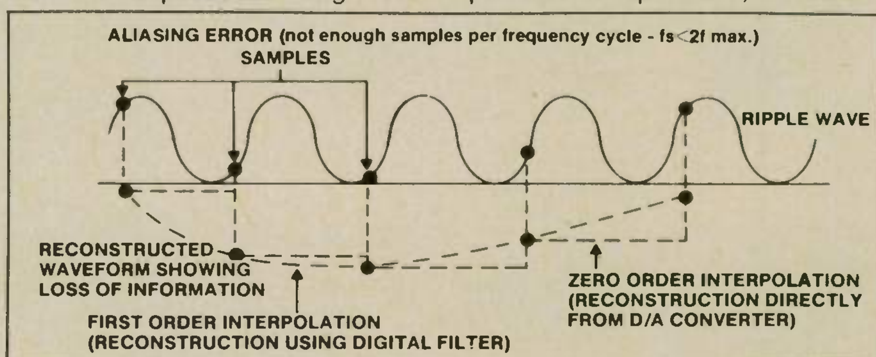


Fig. 3 Aliasing error vs. sampling rate.

ples of the analog signal are taken over the specified bandwidth (ΔF). The frequency of the ripple wave for a fixed ΔF is directly proportional to the distance of the mismatch from the source. Hence, the further away the mismatch, the larger the frequency of the ripple wave and greater the effect of aliasing due to sampling error for a fixed sampling rate.

The frequency increments (f_i) before samples are acquired are represented by: $f_i = \frac{f_2 - f_1}{N} = \frac{\Delta F}{1024}$

where f_2 = stop frequency, f_1 = start frequency, and N = number of samples. Likewise the time interval (t_i) corresponding to each frequency increment on the sweep is given by: $t_i = \frac{\Delta T}{1024}$

where ΔT = Time to sweep the source from f_1 to f_2. There are ADC conversion rate limitations which prevent acquiring a large number of samples over one sweep especially at fast sweep times. As an example, in order to acquire 1024 samples at a ΔT of 1 msec, it would require the ADC to complete the entire 12-bit conversion in less than 1 μsec. This implies the use of very expensive ADCs and even with those, less amplitude resolution. To overcome this limitation, the Transline Analyzer performs the data acquisition in multiple sweeps. The waveform generated in synchronism with the sweep is sampled 32 times during the first sweep pass. These 32 samples (S_0, S_{31} - - - -- S_{991}) are equally spaced in time by $\Delta T/32$.

During the next sweep pass, 32 more equally spaced samples (S_1, S_{32}, - - - - - S_{992}) are acquired such that they are offset with respect to the previous samples by a time difference of $\Delta T/1024$. In this manner, over a total of 32sweeps, 1024 samples of the detected ripple wave are acquired and digitized, using a relatively slow and inexpensive converter. Using this technique,increasing the number of sampling points in the instrument does not require the changing of the ADC, but may be accomplished by changing software variables: the number of samples per sweep and the total number of sweeps.

DIGITAL SIGNAL PROCESSING AND MEASUREMENT ROUTINES

The measurement routines use four fundamental digital signal processing algorithms: signal averaging, FFT, interpolation, and correlation. Only a short discussion of each will be given.

Signal Averaging

Signal averaging is the process of sampling the ripple waveform (1024 samples) repeatedly and taking the point-by-point average. If S_{k1} represents the first 12-bit sampled data value corresponding to a given sweep amplitude,

$S_k = \sum_{1}^{N} \frac{S_{kn}}{n}$ gives the average of that point. In a similar manner, all of the 1024 sample values are averaged to reduce the effect of noise on the signal.

Fast Fourier Transform (FFT)Algorithm

Analysis of the complex, composite, time varying ripple wave generated as a result of multiple mismatches along a transmission line could be performed either in the time or in the frequency domain. Since the physical location of discontinuities is represented by spectral lines as stated before, frequency domain analysis is the logical choice. The computation of the frequency spectrum from the time varying signal X(t) is done by the well known equation:

$$X(f) = \int_{\infty}^{\infty} X(t)e^{-j2\pi ft}dt$$

where X(f) is the Fourier Transform of X(t). In the sampled data system, a modified form called the Discrete Fourier Transform (DFT) is used and is represented by:

$$f_k = \sum_{n=0}^{n-1} X_n W^{nk}$$

where X_n is the nth sample of data, f(k) is an individual frequency element out of N elements, and W^{nk} equals $e^{-j(2\pi nk/N)}$. W is a complex rotating vector where the angle is divided into 360/N segments. When the k - 1 frequency is calculated, W_1 is rotated through one cycle (0 to 359°). When the k = 2 element is calculated, W_2 is rotated through 0 to 719° for N complex multiplications. In general the computation is a long procedure as it requires N^2 complex operations in order to calculate

[Continued on page 54]

the entire spectrum. The Fast Fourier Transform (FFT) algorithm provides the same result as the DFT, but with fewer computations, as it is optimized by eliminating all redundant calculations. The Transline Analyzer uses the so called in-place FFT algorithm to conserve memory. The FFT algorithm allows identification of the spectral component generated by each fault along the transmission line.

Interpolation

After the frequency spectrum is generated by the FFT algorithm, it consists of spectral peaks, each of which determines the location of a mismatch. In general, the resolution accuracy (ΔL) of locating distance to faults is inversely proportional to the swept bandwidth (ΔF) and is mathematically expressed by:

$\Delta L = v/2\Delta F$

where v is the velocity of propagation. As an example, for Teflon with a ΔF of 100 MHz, ΔL is equal to 3.4 feet. The larger the swept bandwidth, the greater the number of cycles present in the ripple waveform for a given mismatch. Each additional cycle may be looked upon as the quantization level of the distance resolution. Interpolation is the technique which looks at each spectral peak and its neighboring spectral lines which define a second order curve, and estimates the probable location of the spectral peak by relatively simple calculations. This results in a more precise distance figure for the fault. In the Transline Analyzer, interpolation in conjunction with the FFT algorithm results in a distance accuracy of ±18 inches for a ΔF of 40 MHz in the 2 to 1000 MHz range. Accuracies to better than 3 inches were achieved for higher frequencies and wider bandwidths.

Correlation

There are two environmental conditions affecting the accuracy of the Transline Analyzer in field use and, especially in shipboard use. One is the random or semi-random electromagnetic noise present from various sources, for example from wind and wave action on ships. In addition, there are non-random signal sources such as transmitters, radar, other navigation equipment, plus the numerous communication signals received by the ship's antennas. All these signals can distort the FFT spectrum. The narrow spectrum signals can actually appear as 'phantom' discontinuities on the tested transmission line.

To eliminate these 'foreign signal' interferences a correlation algorithm is employed. The Analyzer is programmed to make several location measurements with widely varying sweep parameters. For each true discontinuity, there is a predictable mapping of parameters from measurement to measurement. In contrast, foreign signals do not follow the expected transformation. Thus, if the correlation integral is computed for each spectral line, preprogrammed correlation threshold can easily distinguish between true and phantom discontinuity. In real life tests, this technique works so well that the Analyzer was able to make correct multiple discontinuity location measurements in the presence of foreign signals more than 20 dB above the test signal from the RF head.

Measurement Routines

The measurement routines are initiated in a manner which allows the user maximum flexibility of operation. At power turn on, the user is asked if he requires instructions. If instructions are requested by activating the "Yes" key, or if nothing is done for 15 to 20 seconds, the instrument assumes an inexperienced operator and starts printing out the instructions which give a clear understanding of the measurements and data entry to the user. The instructions may be bypassed by an experienced user by use of the "No" key.

The next step is the data entry sequence where the user is prompted to input the center frequency of operation, minimum VSWR value of interest, the cable number, etc. At this point the user is asked if he can short the far end of the

[Continued on page 56]

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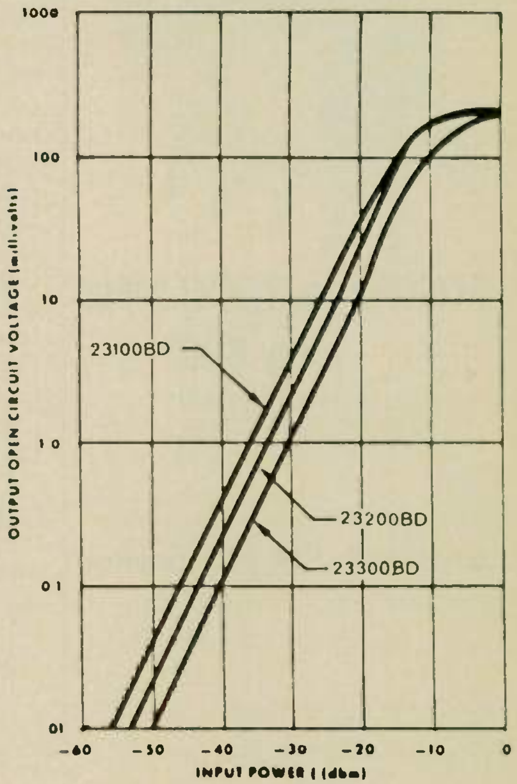
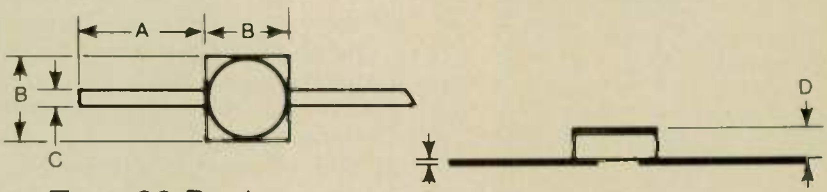


GERMANIUM BACK DIODES

Part Number	I_p µA Range	V_R * mV (Typ)	V_F ** mV (Typ)	R_S ohms (Typ)	C_T pF (Max)
100BD	50 150	540	110	8	0.7
150BD	50 150	540	100	7	1.0
200BD	150 250	550	110	7	0.7
250BD	150 250	550	100	6	1.0
300BD	250 350	550	110	7	0.7
350BD	250 350	550	100	6	1.0
400BD	350 500	560	100	6	0.7
450BD	350 500	560	80	5	1.0 *

* V measured at 500 µA I_R ** V measured at 3 mA I_F



Type 23 Package

DIM	A	B	C	D	E
MIN	13	064	018	026	003
MAX	17	076	022	034	006

ENVIRONMENTAL RATINGS

Storage Temp —65°C to +100°C
Operating Temp —65°C to +100°C
Temperature Cycle—MIL-STD-202, Method 107 TC A
Shock—MIL-STD-750, Method 2016
Vibration—MIL-STD-750, Method 2056
Acceleration—MIL STD-750, Method 2056
Fine Leak—MIL-STD-750 Method 1071 TC G
Gross Leak—MIL-STD-750, Method 1071, TC C

Typical Performance

Type	Rv (Ω)	Tss (−dbm)	M $\left(\frac{K}{\sqrt{Rv}}\right)$	K $\left(\frac{mv}{mw}\right)$
23100BD	400	−52	80	1600
23200BD	180	−51	90	1200
23300BD	80	−50	100	900

NOTES: 1. 2MHz Band Width **2.** Input Power −20 dbm **3.** Fo 10GHz

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[*From page 54*] **FAULT FINDER**

line. If he responds with a "Yes", the system will allow computation of the true line attenuation and length. If not, the user is asked to input the line length of interest. Additionally, if the cable number was programmed in the instrument's memory, it will use the theoretical attenuation in dB/100 feet and the velocity of propagation of the specific cable. If the cable number is not programmed in the system memory and if the user cannot short the far end (this is typical of shipboard runs to the antenna) the system will ask the user to input the theoretical attenuation in dB/100 feet and the velocity of propagation for the cable under test.

Upon the completion of the data entry parameters, the system requests the operator to use the reference line by putting a short at its end. In the 2 to 1000 Mhz head, the calibration delay line is self-contained and the user needs to place a short at the connector of the RF test port.

The system then looks at this total reflection coming back from the short. It acquires 1024 samples of this ripple wave, averages the signal 8 times and stores the values in RAM. It computes the maximum and minimum data values and computes the peak-to-peak calibration amplitude (CA) as a reference. It then performs an FFT algorithm to generate the frequency spectrum. The index of the spectral peak and its height (Y_{cal}) are computed and interpolation is used to verify the calibration length (CL) of the delay line reference by the following equation:

$CL = (IP) \bullet (NF) \bullet (\nu)$

where IP is the index of the spectral peak, i.e. number of cycles in the ripple wave, and NF is the normalizing factor equal to $1/2\Delta F$. ν again is the velocity of propagation. The preceding equation is derived from the relationship that the frequency of the ripple wave (f_D) is given by:

$f_D = \Delta F/\Delta T \bullet 2L/\nu$ where ΔF = swept bandwith, ΔT = sweep time, L = length of mismatch from source and ν = velocity of propagation.

If the length is computed by the instrument to be within 2% of what it was programmed for, the display and printer indicate a "CALIBRATED" message. Otherwise the system will show a "CALIBRATION FAILURE". Upon completion of the calibration routine and if the end of the line is shorted, the attenuation measurement is initiated. The ripple wave being generated from this short is digitized, averaged, and stored in RAM. The measured peak-to-peak amplitude (MA) is computed and after the FFT and interpolation algorithms are performed the location (cable length KBL) of the short is determined. The 2-way line attenuation is now computed by the following equation: 2-way attenuation $(ATT)_{total} = 20 \log[CA/MA] - 2\ CPATT$ where CPATT is the one-way coupler attenuation, if any. The cable length is computed as before: $KBL = (IP) \bullet (N.F) \bullet (\nu)$

The attenuation in dB/100 feet (ATT) is computed by:

$ATT = {}^{50}(ATT)_{total}/KBL$

The computed attenuation and cable length are subsequently used in the location measurement routines.

The next measurement prompt tells the user to terminate the end of the line with antenna or dummy load before initiating the return loss measurement. If the load is a perfect termination, no reflections should occur there. Any reflections that are detected are due to mismatches (imperfections) along the line under test. As before, peak-to-peak value (MA) is computed and the composite return loss (RL) as it appears at the connector of the RF head is computed and printed:

$RL_{dB} = 20 \log [CA/MA] - 2\ CPATT$

where CPATT = Coupler Attenuation, if any.

The return loss value is then converted to VSWR:

$VSWR = 1 + \rho/1 - \rho$

where ρ is equal to the reflection coefficient.

The last measurement routine involves the location measurement. The complex ripple wave is digitized, as before, and the FFT spectrum generated. After interpolation and correlation the indices of each spectral peak are computed and the distance (D_n)

of each mismatch from the source determined.

$D_n = (IP_n) \bullet (NF) \bullet (\nu)$

An estimation is made of the percentage contribution of each mismatch to the composite VSWR by using the individual spectral heights. The percentage contribution (%) is computed by:

$\% = \left(Y_m / \sum_{1}^{N} Y_n\right) \bullet 100$

where Y_m = individual spectral height and $\sum_{1}^{N} Y_n$ is equal to the sum of spectral heights.

An alternate printout gives the percentage of each mismatch as compared to the short in the calibration routine, after being corrected for line attenuation. If the attenuation corrected VSWR is greater than the set minimum VSWR threshold value, the discontinuity is printed and if not, it is suppressed. In the case where the corrected VSWR for all discontinuities is below the set threshold, a "NO REFLECTIONS" message is printed. At the end of the measurements, a full report is printed out giving the user a hard copy of all data conditions, measurement parameters, and results. A typical printout is shown in Figure 4.

```
********************
   FULL REPORT
********************
C.F.=22.ØØ MHz
RG=142.ØØ
THRESHOLD VSWR=1.ØØ
VELOCITY FACTOR=Ø.69
COUPLER ATT=Ø.ØØ DB
2-WAY ATT=4.32 DB
ATT=2.Ø4 DB/1ØØFT
LENGTH=1Ø5.73FT
RET LOSS=4.18 DB
VSWR=4.24
PERCENT REFLECTION
 AT CONNECTOR
2.76% AT 33.68 FT
2.65% AT 63.54 FT
94.59% AT 1Ø5.69 FT
PERCENT OF SHORT
WITH ATT CORRECTION
2.Ø8% AT 33.68 FT
2.3Ø% AT 63.54 FT
1ØØ.ØØ% AT 1Ø5.69 FT
END OF LOC LIST
```

Fig. 4 Typical printout

INSTRUMENT ACCURACIES

Location measurement accuracy at a 40 MHz bandwith is ±18 inches. At higher bandwiths, e.g. X-band with 500 MHz BW, this figure is greatly increased and results in an accuracy of 3 inches. The system uses different bandwidths for different RF heads. In general the bandwidth of the measurement is approximately 5% of the overall frequency coverage of the specific head.

The composite VSWR measurement accuracy (i.e. VSWR as measured at the connector) is ± 0. 15 typical at 2:1. The dynamic range for the Transline Analyzer is 80 dB. The dynamic range for the return loss vs frequency measurement (X-Y mode) is 40 dB, RF output is 0dBm leveled to ±0.2 dB flatness and the attenuation vs frequency measurement accuracy is better than 10% typical. "The percentage contribution at the connector" readout and the "percentage of short" readout are really distribution estimates and a spec cannot be laid on it.

ACKNOWLEDGEMENT

We wish to express our gratitude to Mr. Stanley M. Wolf and Dr. Lawrence Wan for giving us the opportunity to develop the Transline Analyzer; to the management staff for their support; to Messrs. Jim Brown, Joe Fala, Thang Vu, Jim Vuong, John Scordo, and Richard Mangalos for their unending dedication and contributions; and last but not least, Mr. Robert Windle of NSWC for his cooperation and efforts in realizing the full potential of the instrument.

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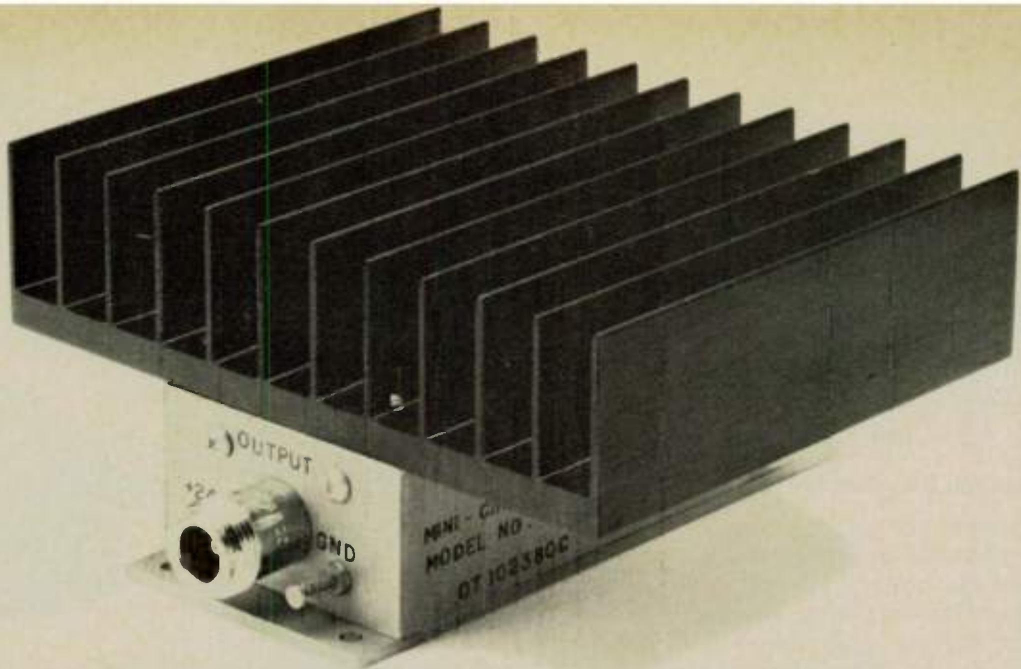

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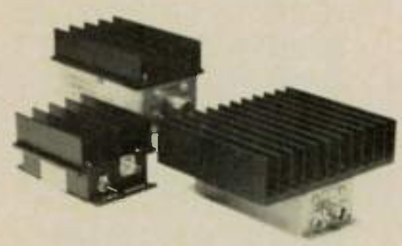

* Model No.	Freq. MHz	Gain dB	Gain Flatness dB	Max. Power Output dBm 1-dB Compression	Noise Figure dB	Intercept Point 3rd Order dBm	DC Power Voltage	DC Power Current	Price $ Ea.	Price Qty.
ZHL-32A	0.05-130	25 Min	±1.0 Max	+29 Min	10 Typ	+38 Typ	+24V	0.6A	199.00	(1-9)
ZHL-3A	0.4-150	24 Min	±1.0 Max	+29.5 Min	11 Typ	+38 Typ	+24V	0.6A	199.00	(1-9)
ZHL-1A	2-500	16 Min	±1.0 Max	+28 Min	11 Typ	+38 Typ	+24V	0.6A	199.00	(1-9)
ZHL-2	10-1000	15 Min	±1.0 Max	+29 Min	18 Typ	+38 Typ	+24V	0.6A	349.00	(1-9)
ZHL-2-8	10-1000	27 Min	±1.0 Max	+29 Min	20 Typ	+38 Typ	+24V	0.65A	449.00	(1-9)
ZHL-2-12	10-1200	24 Min	±1.0 Max	+29 Min*	10 Typ	+38 Typ	+24V	0.75A	524.00	(1-9)
ZHL-1-2W	5-500	29 Min	±1.0 Max	+33 Min	12 Typ	+44 Typ	+24V	0.9A	495.00	(1-9)

Total safe input power +20 dBm, operating temperature 0° C to +60° C, storage temperature −55° C to +100° C, 50 ohm impedance, input and output VSWR 2:1 max.
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SMA, TNC and Type N connectors are also available

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An experimental study was performed on a variety of GaAs FETs from different companies to identify device and circuit parameters that lead to high efficiency operation of microwave FET amplifiers. FETs were characterized by means of a unique system that measures the current and voltage waveforms of the device operating at microwave frequencies under large signal conditions. We found that some characteristics of the gate - namely gate resistance and breakdown voltage of the Schottky barrier - have a large effect on the maximum operating efficiency and output power. A device, whose circuit was optimized for highest efficiency, operated with a power-added efficiency of 72% and an output power of 1.2W at 2.4GHz.

INTRODUCTION

Many parameters affect the efficiency with which an FET converts DC into RF power. Internal losses are important and their minimization; obtained for instance by increasing the carrier mobility in the semiconductor and by improving the metal — semiconductor contacts, has resulted in significant efficiency improvements. However, we found that other factors are also very important, particularly those that affect the saturation of the device under large-signal condition. One of these, the breakdown of the gate Schottky barrier, prevents full pinch-off of the channel thereby limiting the RF power and consequently the maximum operating efficiency.

Also, forward-conduction of the Schottky barrier and the associated DC current prevents full opening of the channel and therefore limits the efficiency.

Indications of such phenomena were brought out by a unique system that measures the current and voltage waveforms of the device operating at microwave frequencies under large-signal conditions. A variety of GaAs FETs from different manufacturers were tested to identify device and circuit parameters that lead to high efficiency operation.

The driving force behind this effort is the realization that GaAs FETs have the potential of operating at very high efficiency. In effect, recent results were most encouraging: a power-added efficiency of 72% was obtained from a device delivering 1.2 W of output power of 2.45 GHz. To our knowledge, this is the highest efficiency reported for a solid state device operating at that frequency.

WAVEFORM MEASUREMENT

The waveform measurement system, specifically developed for studying nonlinear effects in GaAs FETs'[1], is shown with its main components in Figure 1. A source of microwave power at 2.45 GHz feeds the input of a text fixture in which the FET is mounted. The fixture includes input and output tuning circuits that match the FET to the input generators and to the output load and shape the waveforms for optimum efficiency and power performance. Four very small resistive probes, integrated with microstrip tuning circuits, sample the waveforms at different points in the circuit. The four signals are then routed through switches toward a dual-channel sampling oscilloscope. The time axis of the oscilloscope is stepped by the output of a computer-controlled D/A converter. The vertical outputs from the oscilloscope are digitized and the data are stored in the computer. The basic operation is as follows: The measured data are processed with a fast Fourier transform algorithm and resultant spectral components are corrected for the amplitude and phase frequency response of the probes, previously characterized over a frequency range covering typically five harmonics. This defines the corrected spectra of the voltage waveforms present in the circuit at the probing points. These spectra are then used, in conjunction with the ABCD parameters of various sections of the circuit. to compute the current

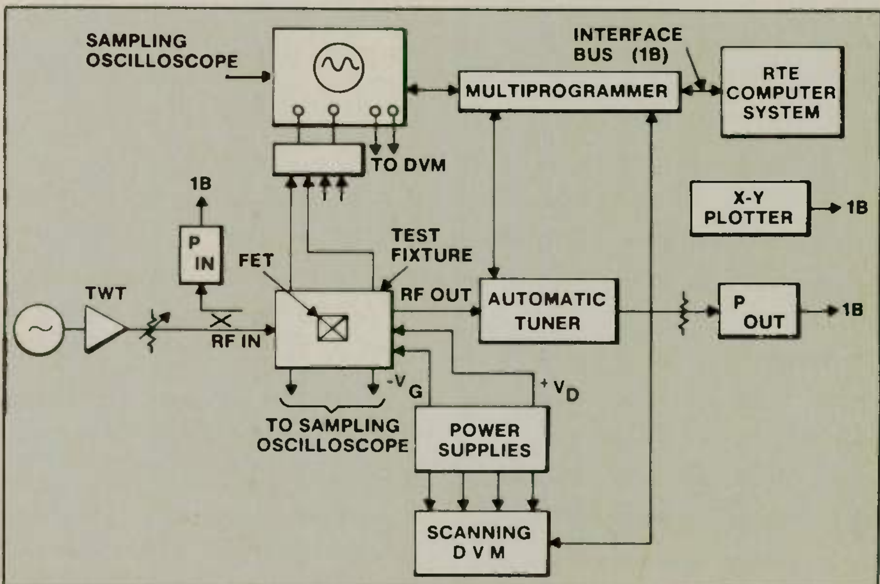

Fig. 1 Waveform measurement system.

*This work was supported by Rockwell International, Anaheim, CA. Contract No. A9EA-766939-910.

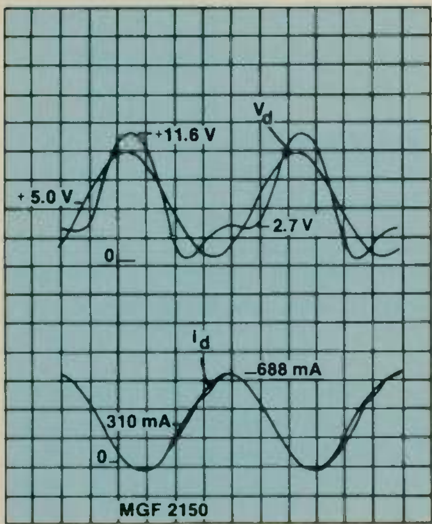

Fig. 2 Mitsubishi MGF 2150 drain waveforms.

spectra and finally, the current waveforms. This procedure, based on Fourier transform and reverse transform of signals, allows us to derive current waveforms more accurately than it could be done by direct current measurements.

Examples of measured waveforms are shown in Figures 2 through 7. The data of Figures 2 through 5 — measured waveforms with superimposed fundamental components — were obtained from a device type MGF 2150 made by Mitsubishi and specified for an output power of 2 W at 12 GHz. Both theory and experimental evidence indicated that a high cutoff frequency is a necessary, although not sufficient, requirement for achieving high operating efficiency. This device provided some of the best results, specifically a power-added efficiency of 71% with an output power of 0.97 W at 2.45 GHz.

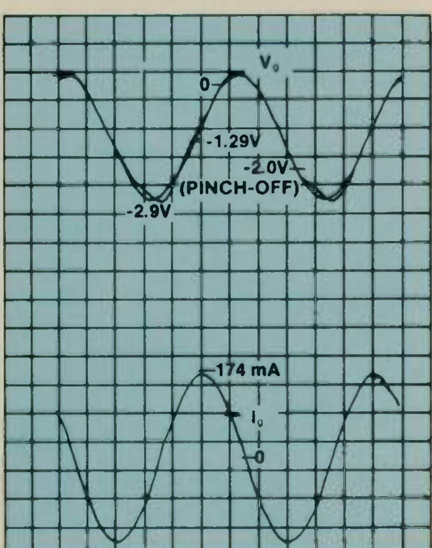

Fig. 3 Mitsubishi MGF 2150 gate waveforms.

The waveforms of Figures 2 and 3 measured at a condition which corresponds closely to the one that provided maximum efficiency. The drain voltage waveform, v_d, approaches a square wave. This is a mode of operation which is conducive to high efficiency. The current is mostly sinusoidal, in agreement with the theory for maximum operating efficiency[2]. Only a small amount of distortion can be seen near the waveform peak.

The corresponding gate waveforms, at a bias of -1.29V, are shown in Figure 3. Because of the high transconductance of 630 mS and low pinch-off voltage of -2 V of this device, the peak-to-peak voltage swing is low, only 3 V, and the distortion of the waveform is rather small. The gate current is mostly capacitive (70° advanced with respect to v_g) and reaches a high peak value because of the rather high input capacitance, approximately 5.5 pF.

Figure 2 shows the FET fully turned-on ($v_d \approx 0$) only for a small fraction of the cycle, approximately 60°. During the following 120° of the cycle, v_d rises to a value of approximately 2.7 V. This rather high voltage across the FET during the on-time is detrimental to the efficiency. If, for instance, the voltage were to be maintained at a constant value of 0.7 V — corresponding to a low-field resistance of 1.0 Ω and peak current of 666 mA — the efficiency could reach a value of approximately 80%.

A seemingly certain way for achieving a full turn-on and turn-off of the device and for better approximating a switching mode of operating is to increase the level of RF input drive. The expected result is an increase of the power-added efficiency, provided that the device gain is still sufficiently high to have a net increase of output power. However, our experiments have shown this not to be the case. When the RF input power was increased approximately 2 dB above the optimum efficiency condition, the drain voltage waveforms, as shown in Figure 4, revealed a significant increase of the "on" voltage. This effect clearly reduces the operating efficiency. The cause of this unexpected result is attributed to the presence of forward conduction currents flowing through the resistance of the gate circuit. Specifically, a sufficiently large RF signal overcomes the negative bias of the gate and causes the gate to draw current in the forward conduction direction during parts of the RF cycle developing an additional negative voltage on the gate, which is then stored by the gate capacitance. The net result is a constriction of the channel with

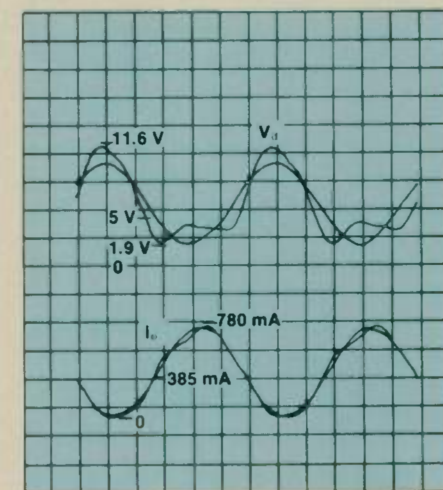

Fig 4 Mitsubishi MGF 2150 drain waveforms - 2.3 dB overdrive.

an increase of channel resistance at a time when the drain current is at its peak. The gate voltage waveform of Figure 5 clearly shows clipping above 0.7V, due to forward conduction of the Schottky barrier.

Another interesting device is an FLC 30 made by Fujitsu and designed for C-band operation, which provided the highest efficiency at 2.45 GHz (72% at 1.2W of output power). The waveforms measured at a high efficiency operating condition are shown in Figures 6 and 7. The bias level was 10.5V and 130 mA for the drain, and -4.0V for the gate. The drain voltage waveform shows a pronounced distortion, although they do not clearly indicate the switching mode of operation which would lead to an even higher efficiency.

The drain current is virtually sinusoidal, as it is expected in FET's operating with tuned output circuits[2]. In addition, during approximately 30% of the cycle, the drain current is negative. This is caused by an apparent increase of the amplitude of the current in the active device due to the device output capacitance. The amplitude ratio, K, between the current in the active device and the total current (including output capacitance), is given by:

$K = 1/(1 - \omega^2 LC + j\omega RC)$

where C is the output capacitance, and L is the output tuning inductance in series with the output load resistance R. At resonance, K assumes the values of 0.4 for R = 25 Ω and C = 1 pF, which is consistent with the values of the negative current swing that have been observed during these experiments.

The instantaneous gate voltage v_b varies between -0.5V and -8.3V around a dc bias of -4.0V. The flattening of v_g around -0.5V is caused by a large change of the gate-to-source capacitance due to the varactor-like behavior of the Schottky barrier. The gate current is mostly capacitive and sinusoidal because of the filtering effect of the high impedance-ratio input tuning circuit. Interestingly, when the device was heavily overdriven (a 7 dB increase in RF input power) the drain waveform remained practically unchanged. This indicates that the device saturation is caused by the gate, which has lost control of the current in the channel.

Finally, waveform measurements were carried out on a device, type Fujitsu FLS50, capable of delivering 5 W of output power at S-band. The device was tuned for highest efficiency (50%) at a drain voltage of 10V and a drain current of 25% of I_{dss}. The results show that the drain voltage and current are fully modulated. However, even at the higher RF drive, the device did not approach the switching mode of operation which would be conducive to higher efficiency. This behavior may be explained by the slow response of the gate circuit due to the charging of the gate capacitance through the resistance of the gate stripes. The low-pass characteristic of this circuit prevents rapid switching of the device. A supporting piece of evidence is that the measured gate waveforms were also practically sinusoidal even at this high drive level. Therefore, it is the rather slow response of the device that limits the maximum efficiency to approximately 50% — the limit value for an unsaturated amplifier.

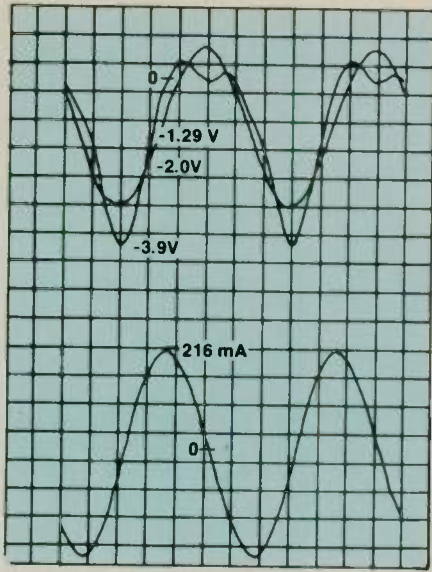

Fig. 5 Mitsubishi MGF 2150 gate waveforms - 2.3 dB overdrive.

HIGH EFFICIENCY AMPLIFIER PERFORMANCE

The performance of a higher efficiency amplifier, designed around the device whose waveforms are depicted in Figures 6 and 7, is shown in Figure 8 where the out-

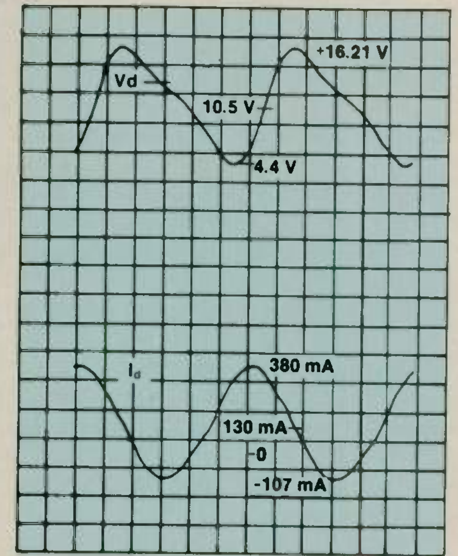

Fig. 6 Fujitsu FLC30 drain waveforms - maximum efficiency point.

[*Continued on page 62*]

put power P_{out} and the power-added efficiency η are plotted as functions of the input power P_{in}. The power-added efficiency is here defined as

$$\eta = \frac{P_{out} - P_{in}}{P_{dc}}$$

where P_{dc} is the DC input power. A power-added efficiency of 72% was obtained at an output power of 1.27W and a gain of 8 dB. This value of power-added efficiency is, to our knowledge, unsurpassed by any solid state device operating in the same frequency range. The mode of operation is Class AB with the DC drain current varying from 52 mA at low RF drive to 135 mA at full drive. The change of drain current, and the associated change in device transconductance, causes the gain to be relatively low (8 dB) at low RF drive and to gradually increase with drive up to peak value of 10 dB. A further increase of drive produces device saturation and a consequent decrease of gain.

Figure 9 is the photograph of an amplifier without the protective cover. The FET is matched by input and output tuning circuits printed on $A1_2O_3$ substrates. Clearly visible are the coiled RF chokes and the 10-μF electrolitic capacitors connected at the bias input terminals.

CONSIDERATIONS OF EFFICIENCY OPTIMIZATION AND CONCLUSIONS

Some of the parameters that limit the efficiency of GaAs FETs are well known such as, carrier mobility of GaAs, contact resistance, and resistance of the thin-film metallization. Some phenomena, however, have a more subtle effect in limiting the efficiency. One of these phenomena is the voltage breakdown of the Schottky barrier under large RF drive conditions[1]. The onset of the breakdown prevents the device from being fully turned-off or from being turned-off for a sufficiently long fraction of the RF cycle. An attempt to overcome this affect by reducing the drain bias voltage and thereby reducing the gate to drain peak voltage, is not effective beyond the point where the saturation voltage becomes a large fraction of the bias voltage.

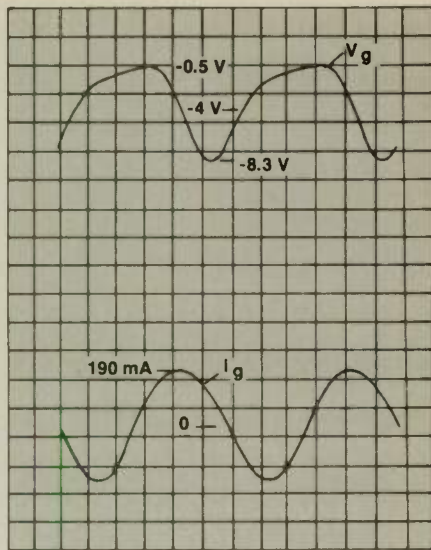

Fig. 7 Fujitsu FLC30 gate waveforms - maximum efficiency point.

It should be noted also that forward conduction of the gate is often unavoidable. High efficiency operation requires full saturation and rapid switching which is obtained by applying at the input large sinusoidal RF signals. It is conceivable that, in the future, high efficiency devices might include low resistance diodes clamping the gate voltage and limiting both the positive and the negative swing of the RF signal. These diodes, in order to be effective, will have to be fabricated on the same substrate as the active device. This would avoid the filtering effect of an intermediate circuit.

Another factor which affects the operating efficiency is the device cut-off frequency. For instance, higher power devices, which have typically lower cut-off frequencies, are sometimes less efficient simply because of their longer switching time. This was exemplified by the waveform measurement results of the FLS50.

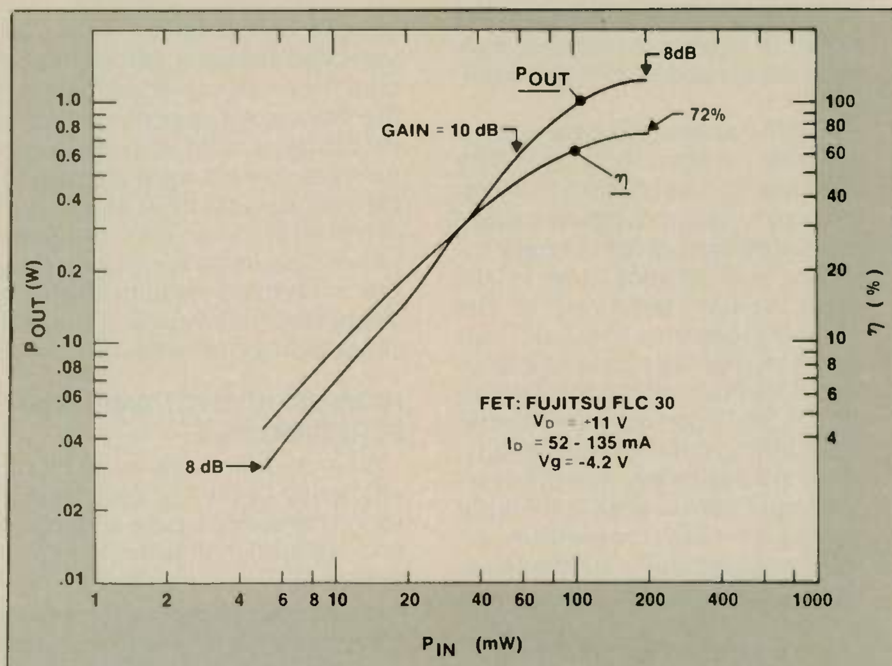

Fig. 8 Performance of FLC30 tuned for maximum power-added efficiency.

Waveform measurements of an FET, which was otherwise very efficient, showed an unexpected effect (i.e. an increase of saturation voltage when the RF drive was raised above the level required for optimum efficiency.) This additional saturation voltage was attributed to the presence of forward conduction current flowing through the resistance of the gate. This made apparent the importance of a low gate resistance for achieving full saturation of the FET.

Also, power and waveform measurements made apparent that neither the Class C nor the Class Emode of operation[3] were suited for high efficiency GaAs FET amplifiers. The basic limitation is the Schottky barrier of the gate that, when biased at pinch-off (or beyond for Class C) breaks down at

[Continued on page 66]

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Frequency (MHz) RF, LO	IF	Conversion Loss (dB)	Isolation (dB) L-R	L-I	3rd Order IP (dBm)*	Model No.
5-500	DC-500	5.0	45	40	+34	M37T
10-1000	DC-600	5.5	45	40	+32	M47T
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*3rd Order Intercept measured at +23dBm LO

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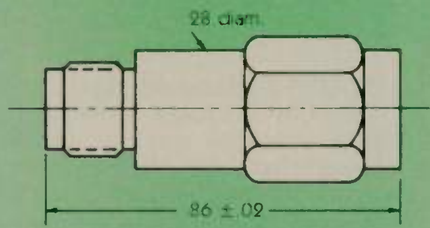
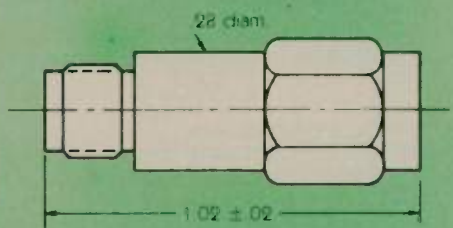
- DC to 18.0 GHz
- 1 thru 30dB
- -65°C to+125°C
- 2 watts at +25°C
- MIL-E-5400 environment
- MIL-A-3933 requirements
- MIL-E-16400 environment
- 0.86 in. long × 0.28 in. diam.



1 through 12dB Weight: 5 grams (0.175 oz.)



13 through 30dB Weight: 6 grams (0.212 oz.)

DC to 18.0 GHz
HIGH PERFORMANCE

- Model 290, M290, F290
- Maximum VSWR: 1.07 +0.015fGHz
- Input Power: 2 watts average at +25°C derated linearly to 0.5 watts at +125°C
- Operating Temp. Range: -65°C to +125°C
- Connectors: Stainless Steel SMA per MIL-C-39012

ATTENUATION VALUE	ACCURACY
1,2,3,4,5, and 6dB	±0.3dB
7,8,9,10 thru 20dB	±0.5dB
21 thru 30 dB	±1.0dB

DC to 12.4 GHz
HIGH PERFORMANCE

- Model 291, M291, F291
- Maximum VSWR: 1.07 +0.015fGHz
- Input Power: 2 watts average at +25°C derated linearly to 0.5 watts at +125°C
- Operating Temp. Range: -65°C to +125°C
- Connectors: Stainless Steel SMA per MIL-C-39012

ATTENUATION VALUE	ACCURACY
1,2,3,4,5 and 6dB	±0.3dB
7,8,9,10 thru 20dB	±0.5dB
21 thru 30dB	±1.0dB

DC to 8.0 GHz
HIGH PERFORMANCE

- Model 292, M292, F292
- Maximum VSWR: 1.07 +0.015fGHz
- Input Power: 2 watts average at +25°C derated linearly to 0.5 watts at +125°C
- Operating Temp. Range: -65°C to +125°C
- Connectors: Stainless Steel SMA per MIL-C-39012

ATTENUATION VALUE	ACCURACY
1,2,3,4,5,6,7,8,9,10dB	±0.3dB
11 thru 20dB	±0.5dB
21 thru 30dB	±1.0dB

DC to 2.0 GHz
HIGH PERFORMANCE

- Model 294, M294, F294
- Maximum VSWR: 1.15
- Input Power: 2 watts average at +25°C derated linearly to 0.5 watts at +125°C
- Operating Temp. Range: -65°C to +125°C
- Connectors: Stainless Steel SMA per MIL-C-39012

ATTENUATION VALUE	ACCURACY
1 thru 20dB	±0.3dB
21 thru 30dB	±0.5dB

DC to 18.0 GHz
INEXPENSIVE

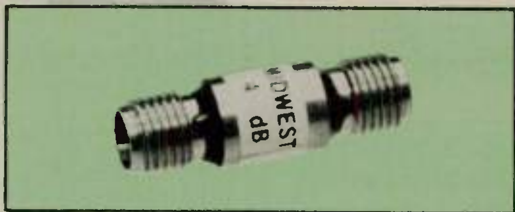
- Model 444, M444, F444
- Maximum VSWR: DC to 4.0 GHz 1.25 • 4.0 to 12.4 GHz 1.45 • 12.4 to 18.0 GHz 1.65
- Input Power: 2 watts average at +25°C derated linearly to 0.5 watts at +125°C
- Operating Temp. Range: -65°C to +125°C
- Connectors: Stainless Steel SMA per MIL-C-39012

ATTENUATION VALUE	ACCURACY DC to 12.4 GHz	ACCURACY 12.4 to 18.0 GHz
1,2,3,4dB	±0.75dB	±0.75dB
5,6,7,8dB	±0.75dB	±1.00dB
9,10,11,12dB	±1.00dB	±1.25dB
13 thru 20dB	±1.50dB	±1.50dB
21 thru 30dB	±2.0dB	±2.0dB

Double Male Connectors
(M PREFIX TO MODEL NO.)

dB VALUE	OVERALL LENGTH
1-12dB	.98"
13-30dB	1.12"

Double Female Connectors
(F PREFIX TO MODEL NO.)



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[*From page 62*] **HIGH EFFICIENCY**

low RF drive before the device is fully saturated. The result is rather low power and efficiency. In addition, since the transconductance is usually low near pinch-off, the gain is also low, which further reduces the power-added efficiency.

Saturated Class B and Class AB modes of operation were found to be better suited for high-efficiency FET amplifiers. Here, a better compromise could be maintained between the conflicting requirements of device bias, RF drive level, and circuit impedances. Indeed, experiments have consistently shown that best efficiency - or a best efficiency/power compromise could be obtained by operating the devices in Class AB or near Class B (bias current approximately 10% of I_{dss}).

During the course of this study it became apparent that maximum efficiency can be obtained only at approximately one-half of the

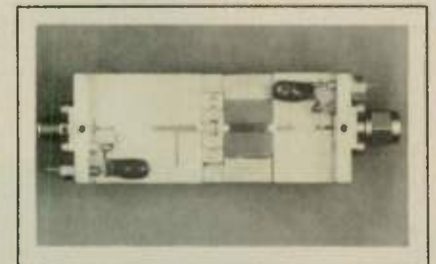

Fig. 9 FLC30 amplifier without protective cover.

maximum power available from the device. Therefore, a device designed for optimum efficiency should be oversized by approximately a factor of two. The inevitable higher input and output capacitance of the device can be overcome by internal matching techniques.

Finally, excellent efficiency performances were obtained from selected commercial devices type MGF 2150 (Mitsubishi) and FLC 30 (Fujitsu), the latter providing an efficiency of 72% with an output power of 1.2 W and a gain of 8 dB at 2.45 GHz. Even higher efficiency can be expected from devices designed specifically for high efficiency operation.

ACKNOWLEDGEMENTS

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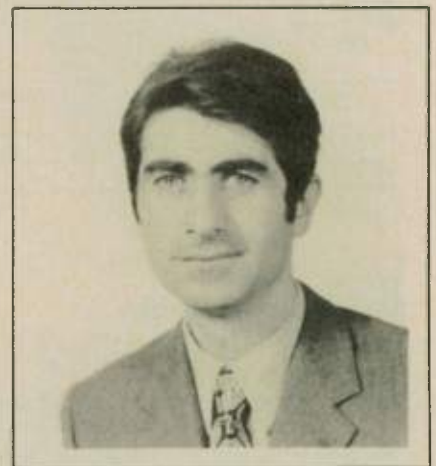

Franco N. Sechi, *received the degree of Doctor in Electrical Engineering in 1964 from the Polytechnic Institute of Milano, Italy. From 1965 to 1968, he was employed by ITT in Milano, where he was concerned with the design of solid-state microwave radio-link equipment.*

In 1968, he joined RCA, Electronic Components, in Harrison, New Jersey, as a design engineer in the Solid State Product Design Group. In this position he designed transferred-electron oscillators. In 1973, he transferred to the Microwave Technology Center, RCA Laboratories, Princeton, New Jersey. In his present position he is involved in the development of power transistor amplifiers.

For his work on linear microwave power amplifiers, he received an RCA Laboratories' Outstanding Achievement Award in 1976. He received a second Achievement Award in 1979 for his work on a solid-state radar system for aircrafts.

Dr. Sechi authored papers on transferred-electron oscillators, thermal and large signal characterization of microwave devices, and on high-power microwave transistor amplifiers. He currently holds six U.S. patents. Dr. Sechi is a member of the IEEE. ■

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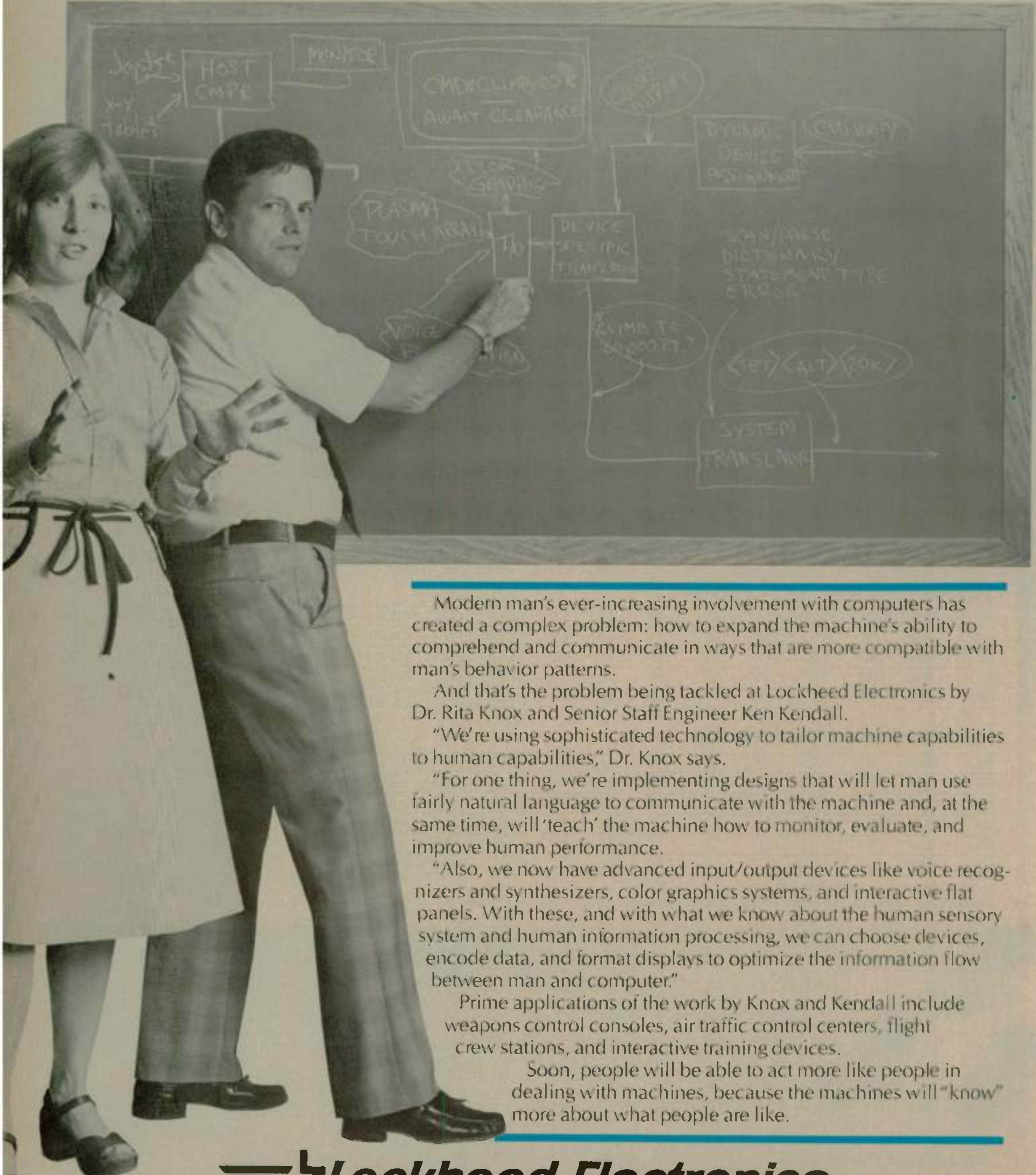

Modern man's ever-increasing involvement with computers has created a complex problem: how to expand the machine's ability to comprehend and communicate in ways that are more compatible with man's behavior patterns.

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Quasi-Lumped Element Impedance Matching Networks —

for wideband miniature GaAs FET amplifiers

SANJAY B. MOGHE
Raytheon Company,
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A quasi-lumped element impedance matching technique was developed for multi-octave bandwidth FET amplifiers. The lumped elements were realized by parallel plate capacitors and high impedance lines etched on two pieces of 10 mil thick alumina substrates. Various low noise and power amplifiers for S thru Ku-bands were constructed using this method. Power amplifiers with 4 dB gain and more than half a Watt power output were realized for 2-12 and 4-18 GHz bands.

INTRODUCTION

The performance of wideband amplifiers has improved steadily over the past few years. Improved bandwidth is achieved by placing the impedance matching network located physically close to the device. Some circuit designers employ the monolithic integrated circuit approach[1,2,3] to attain this "close-in" matching while some others use MOS or MOM capacitors to realize lumped elements in a so-called hybrid circuit approach.[4,5] The monolithic circuit approach requires a precise "device level" process control and provides no flexibility for any circuit adjustment. The cost of monolithic amplifiers cannot generally be justified without a very large volume. In the hybrid approach MOS or MOM capacitors are incorporated in the circuit either by mounting directly on the ground plane without a substrate or by mounting on top of the substrate with plated thru holes. Either method requires additional fabrication processes as part of the assembly. Furthermore, the MOS and MOM lumped capacitor values are hard to control within a tolerable window and are less flexible for circuit tuning.

This paper describes a quasi-lumped element impedance matching technique for multi-octave bandwidth FET amplifiers from S thru Ku-bands. The lumped capacitors were realized as parallel plate capacitors on 10 mil thick alumina substrates and the inductors were approximated by high impedance bond wires or by 1 mil wide lines etched on the substrate. The complete quasi-lumped element impedance matching network can be fabricated by photolithographic methods without using any discrete lumped elements except for the bond wires. The fabrication process is greatly simplified resulting in lower manufacturing cost. Unlike the monolithic approach, this approach allows for circuit tunability to obtain flat gain, minimum noise figure or maximum power output.

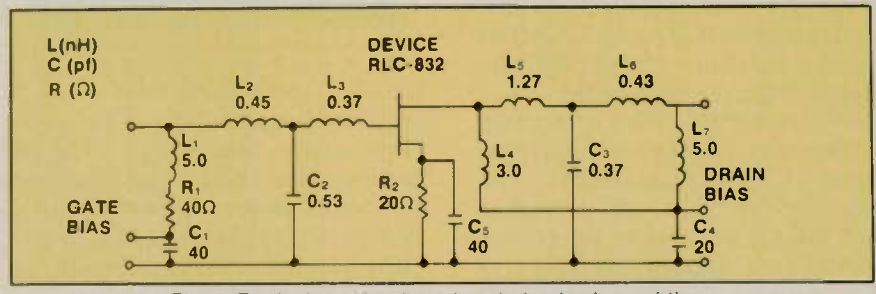

Fig. 1 Equivalent circuit and optimized values of the circuit elements of 4-12 low noise amplifier.

Due to the small size of the amplifier, it is easy to produce in large quantities and can be sealed in a hermetic package. This quasi-lumped approach is suited for broadband matching of both low noise and power amplifiers.

Performance data on a number of low noise and power amplifiers across 2-6, 4-12, and 4-18 GHz bands are presented. The overall size of the amplifiers is significantly smaller than those using the conventional distributed circuit approach. A dual FET 4-12 GHz amplifier with complete matching and bias circuits is realized on a 200 x 380 mil metal carrier. A power amplifier with similar circuit produces half watt saturated power output across 4-12 and 4-18 GHz bands.

AMPLIFIER DESIGN

Raytheon GaAs FETS, RLC832 and RPX 4328 and single cell type 872-50 devices were used in the amplifiers for low noise and high power applications, respectively. These devices have typical dimensions of 1 x 500, 1 x 200 and 0.5 x 400 μm respectively. The detailed performance of these devices is described elsewhere.[6,7] Devices RLC 832 abd RPX 4328 were used in amplifiers through x-band frequencies and type 872-50 and NEC 137 through Ku-band frequencies. The small signal S-parameters were measured on an automatic network analyzer and de-embedded to obtain the FET chip characteristics. The small signal S-parameters give first approximate matching circuit which has to be optimized for maximum power output or low noise as desired. This can be easily accomplished in the quasi-lumped approach since the circuit has considerable tunability. FET chips rather than packaged devices were

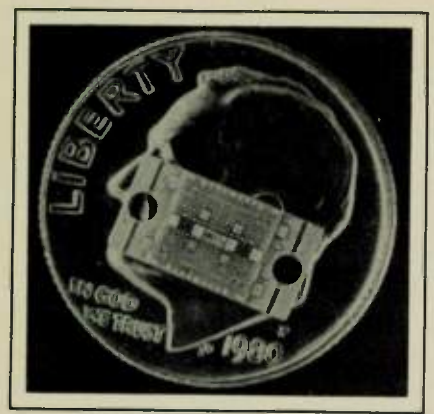

Fig. 2 A 4-12 GHz dual FET amplifier with complete bias and matching circuit.

used to obtain matching circuits close to the device.

Figure 1 shows the circuit topology used to match the impedance of a 4-12 GHz low noise FET. This topology was selected because it can be realized easily in a quasi-lumped form and provides a good match over a wide band. The circuit element values required to achieve flat gain across 4-12 GHz were obtained by CAD using the COMPACT® optimization program. Capacitors C_1, C_5 and C_4 are 20 pf bypass capacitors and are not part of the matching circuit. R_2 is the self-bias source resistor which is used to obtain gate to soure bias for minimum noise figure and eliminates the need for two bias power supplies. R_1 is used to reduce the gain and improve input VSWR at low frequencies and compensates 6dB/octave device gain roll off. Although the circuit was designed for 4-12 GHz band, it exhibited 5.5±0.5dB gain across the 2-12 GHz band.

The complete input matching circuit was realized on a single piece of alumina substrate. The inductances were approximated by 1 mil diameter gold bond wire and by 1 mil wide lines etched on the substrate. Lumped capacitors were realized as parallel plate capacitors formed between top and bottom conductors on a 10 mil thick alumina substrate. For inductors and capacitors to act as lumped elements, their physical size should be much less than quarter wavelength at the maximum operating frequency. For a 10 mil thick alumina substrate (ε = 9.8), the capacitance of a 10 mil square pad is 0.022 pf. Therefore, the area required for a 0.5 pf lumped capacitor is in the vicinity of 50 mil square which is less than quarter wavelength at 12 GHz ($\lambda/4 \alpha$ 12 GHz 100 mils). Similarly the length of a 1 nH inductor realized by 120 Ω impedance line (1 mil line) is about 50 mils. Thus the choice of 10 mil thick substrate is an important factor in realizing both capacitors and inductors in quasi-lumped form. The circuit was actually reoptimized to take into account slight degradation at the band edges due to distributed effect of circuit elements.

Figure 2 shows the realization of this circuit mounted on 0.200 x 0.380 inch gold plated copper carriers. Since the overall amplifier has a balanced configuration, two FETs with their matching circuits are realized on one carrier. Circuit elements L_1, L_2, C_2, L_3, L_4, L_5, C_3, L_6, and L_7 for two FETs are all realized on two pieces of 0.170x 0.085 inch alumina substrates. The circuit elements C_1, C_5, C_4, used for RF bypass and R_1, R_2 are non-critical and are realized as lumped chip capacitors and resistors. The

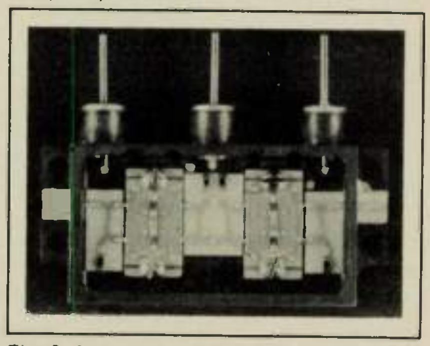

Fig. 3 A two-stage balanced 4-12 GHz amplifier module.

inductors realized by 1 mil wide lines have a few tenths of a dB greater loss than those realized by bond wires. However, this loss is not critical for most applications. In some applications where low circuit losses are essential, the smaller value inductors L_2, L_3 and L_6 could be realized by 2 mil wide lines and inductors L_4 and L_5 by bond wires.

The layout of Figure 2 also shows the tunability of the circuit. Tuning can be done by bonding in extra capacitor pads to increase C_2 and C_4 or by connecting a 1 mil bond wire from the 1 mil etched lines to the capacitor pads to decrease L_3, L_2, L_5 or L_6. Due to the inherently large bandwidth of the circuit, there is less variation in response from unit to unit. Also, unlike bond wires, whose lengths can vary, etched one mil wide lines result in a more reproducible inductor. In practice, most of the trimming can be done by adjusting only one or two circuit elements.

Figure 3 shows the photograph of a two-stage amplifier module.

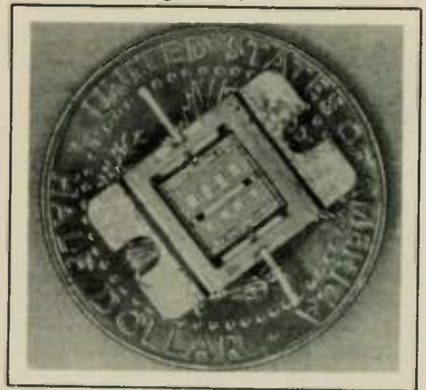

Fig. 4 A 2-6 GHz cascadable packaged amplifier.

Three dB Lange couplers were used to achieve low input and output VSWR. The couplers were realized on 25 mil thick alumina substrates and were soldered to the gold plated copper housing. Each balanced stage has low VSWR and can be cascaded without any readjustment. The overall size of the two stage amplifier module is 0.55 x 1.2 x 0.3 inches.

Similar matching circuits were designed for 2-6 and 4-18 GHz amplifiers using low noise and power FETs. For power FETs, circuit topology similar to that of Figure 1 can be used with increased values of capacitors C_2 and C_3 and decreased values of inductors L_2, L_3, L_5 and L_6. A 6-12 GHz power amplifier was designed using Raytheon's RPX 4328, 1200 μm gate periphery device. For larger periphery power FETs input and output matching capacitor C_2 and C_5 becomes too large ($>$ 1pf) to be realized as a lumped element on 10 mil thick alumina substrate. These large capacitors can still be realized on a substrate with high dielectric constant or on five mil thick alumina.

Figure 4 shows the realization of a 2-6 GHz amplifier on a 300 x 300 mil alumina substrate. The complete cascadable balanced amplifier is housed in a miniature hermetic package. Figure 5 shows a 4-18 GHz balanced amplifier using Raytheon type 872-50 power FETs. For both of these amplifiers the 3 dB couplers and the matching circuits were realized on 10 mil thick alumina. Figure 6 shows a 37 dB gain 6-18 GHz amplifier using eight low noise and power FET stages.

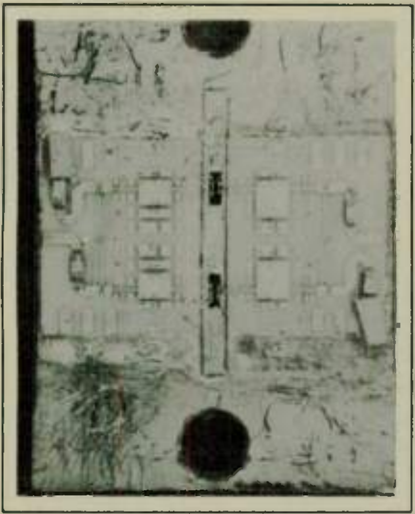

Fig 5 A single stage 4-18 GHz power amplifier.

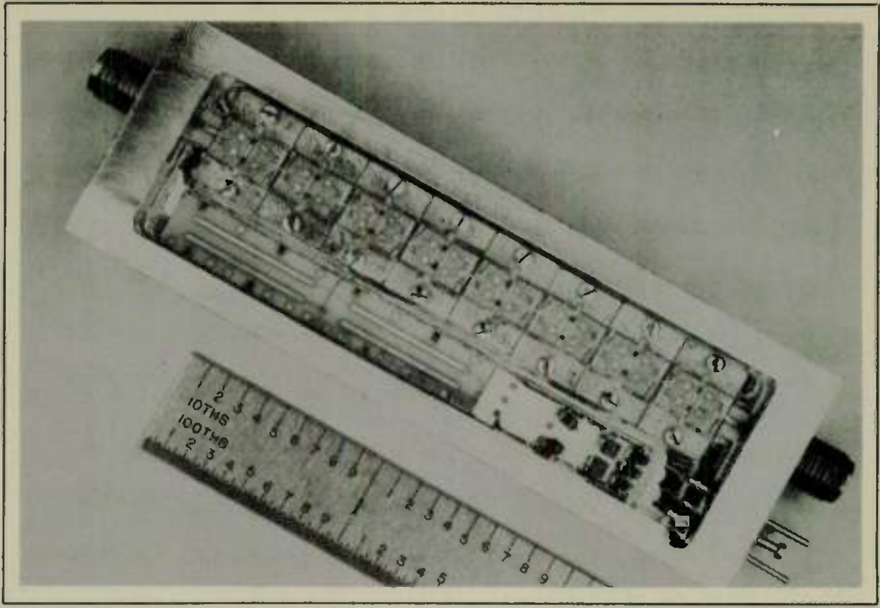

Fig. 6 Eight stage 6-18 GHz amplifier.

AMPLIFIER PERFORMANCE

Figure 7 shows the performance of a balanced single stage, low noise amplifier using the RLC 832 device. Small signal gain of 5.0 ± 1 dB is obtained across the desired band. A single balanced stage gives a minimum saturated power output of 27 dBm. The initial amplifier circuit was based on small signal scattering parameters of the device, however, the output circuit was later tuned to achieve maximum broadband power output. This shows the usefulness of the tunability of the quasi-lumped impedance matching approach.

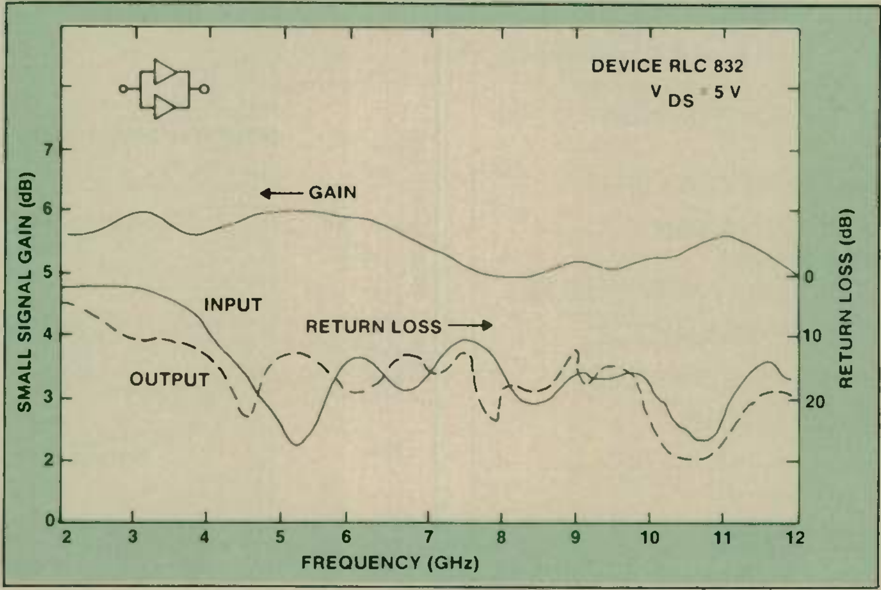

Fig. 7 Gain and return loss of a balanced single stage 4-12 GHz amplifier.

Figure 8 shows the gain, return loss and output power of a 4 - 12 GHz power amplifier using the RPX 4328 device. Small-signal gain of 5.0 ± 1 dB is obtained across the desired band. A single balanced stage gives a minimum saturated power output of +27 dBm. The initial amplifier circuit was based on small-signal scattering parameters of the device. However, the output circuit was later tuned to achieve maximum broadband power output.

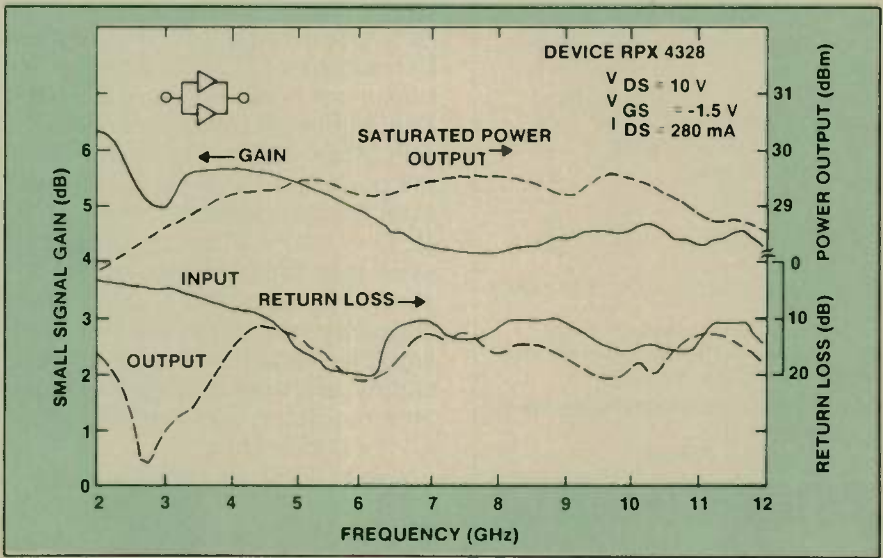

Fig. 8 Gain, return loss and output power of 4-12 GHz power amplifier

Similar performance was achieved with a 2-6 GHz ampli-

[Continued on page 74]

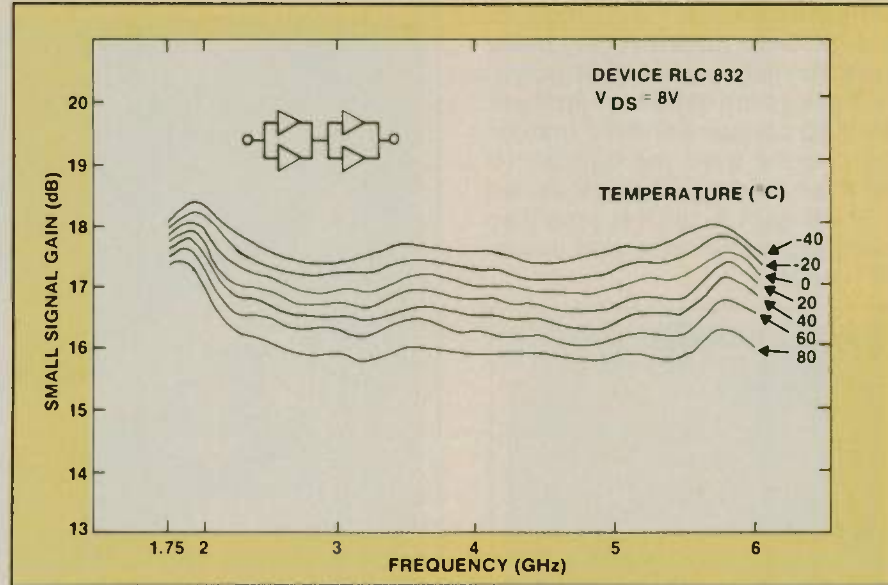

Fig. 9 Temperature dependence of gain of a 2-6 GHz amplifier.

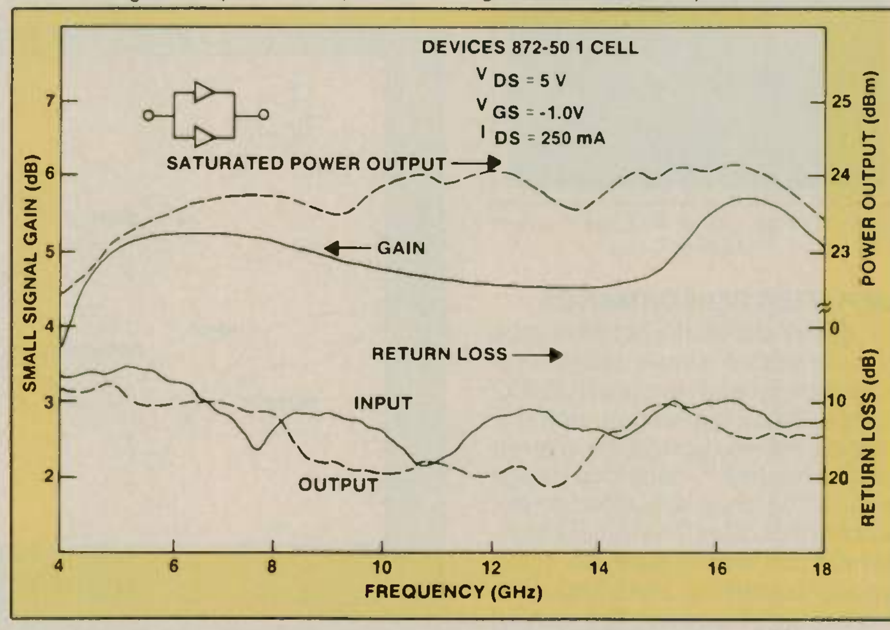

Fig. 10 Gain, return loss and output power of a 4-18 GHz power amplifier

fier. A two-stage 2-6 GHz amplifier achieved 17 ± 1 dB gain with a maximum noise figure of 3.5 dB. The minimum saturated power output is +15 dBm. Figure 9 shows the performance of this amplifier over the temperature range of -40°C to + 80°C. The gain change over this temperature range is about 1.5 dB, without using any temperature compensation. The amplifier requires a +10V power supply and has an internal voltage regulator, and measures 1.5 x 1.1 x 0.85 inches.

A 4-18 GHz amplifier was designed using a Raytheon type 872-50 one cell device. Figure 10 shows that gain of 5 ± 1.0 dB can be achieved with a two-stage amplifier. Input and output return loss is better than 10 dB across the band. The amplifier exhibited a minimum saturated power output of 200 mW across the 4-18 GHz band.

CONCLUSIONS

A quasi-lumped element impedance matching network for multioctave bandwidth amplifiers has been developed. The complete matching network can be realized of two pieces of alumina substrate without using any discrete lumped elements except for bond wires. This approach provides circuit tunability and can be used for amplifiers from S to Ku-band. This technique is suitable for the low cost production of both low noise and power amplifiers.

ACKNOWLEDGMENT

The author wishes to thank R. Gray and W. Tsai for many helpful discussions. The help of F. Morese, B. Hines, Z. Freitas and J. Lubenau in circuit fabrication, assembly and testing in gratefully acknowledged. Thanks are also due to K. Ringdahl for photographs and D. Babigian for manuscript preparation.

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Sanjay B. Moghe *received B. S. and M. S. degrees in 1972 and 1976 from Delhi University, India and the University of Louisville, KY, respectively. He joined Raytheon Company Special Microwave Device Operation in 1979 where he was engaged in development of low noise and power GaAs FET amplifiers. Presently he is at Rensselaer Polytechnic Institute on a leave of absence completing his Ph. D. degree in Electrical Engineering. Mr. Moghe is a member of IEEE and Sigma Xi.* ■

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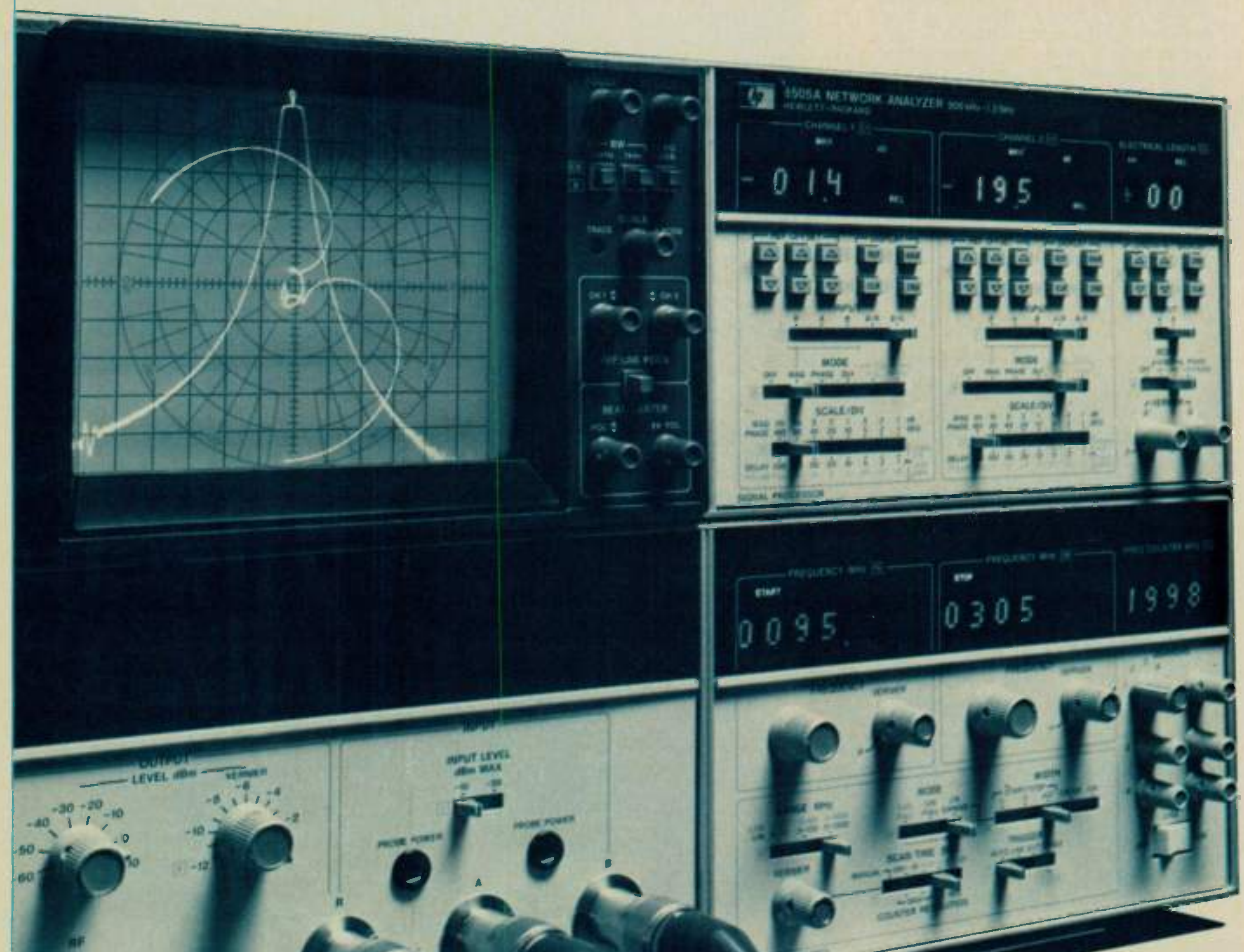

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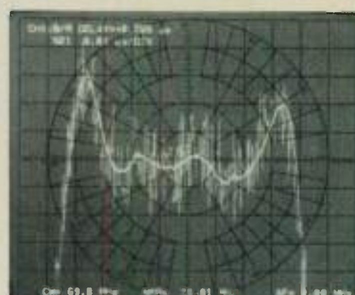

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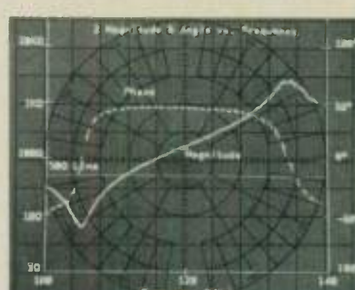

Reflection Coefficient data reformatted to impedance magnitude and angle

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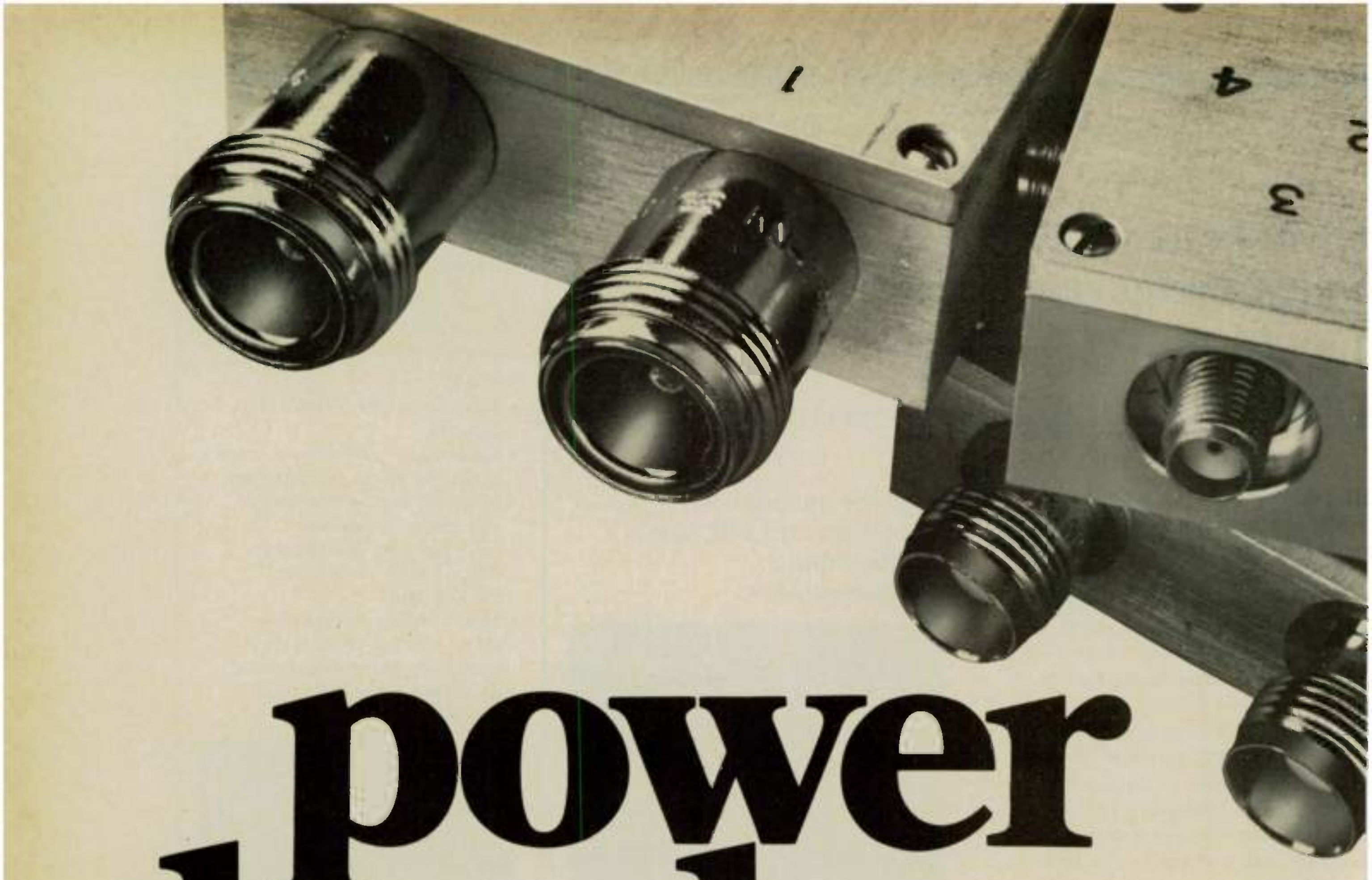

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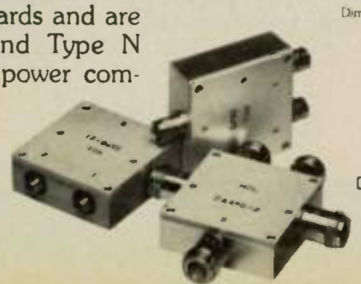
			Typ	Max	Typ	Min		Typ	Divider	Combiner	
ZAPD 1	2 WAY	0.5-1.0	0.3	0.6	25	19	0.3	1.2	10W	10mW	39.95
ZAPD 2	2 WAY	1.0-2.0	0.3	0.6	25	19	0.3	1.2	10W	10mW	39.95
ZAPD 21	2 WAY	0.5-2.0	0.4	0.7	25	18	0.4	1.2	10W	10mW	49.95
ZAPD-4	2 WAY	2.0-4.2	0.4	0.8	25	18	0.4	1.2	10W	10mW	39.95
ZA4PD 2	4 WAY	1.0-2.0	0.3	0.8	25	19	0.7	1.2	10W	10mW	79.95
ZA4PD 4	4 WAY	2.0-4.2	0.5	1.0	25	18	0.8	1.2	10W	10mW	79.95

Dimensions 2" x2" x0.75" Connectors Available BNC, TNC. $5.00 additional for SMA and Type N. BNC not available for −4 models.



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A Backward Wave Oscillator For The 110 To 170 GHz Band

ERICH GLASS
Siemens AG
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In recent years, the application of mm-waves, especially above 100 GHz, has increased more and more. The first kind of application had been pure research, the spectroscopic investigation of material properties in the mm-wave range, e.g. the ferromagnetic, antiferromagnetic and spinflop resonances in solids and gases[1]. Another example is the investigation of non-thermal effects in biological cell growth, the measurement of microwave reflectivity and transmissivity of semiconductors at very low temperatures, or the investigation of dielectric properties at high frequencies.

A second kind of measurement application is in plasma physics. The electron cyclotron emission from plasmas is a useful diagnostic means for the electron temperature. Swept frequency microwave reflectrometry is a valuable method for obtaining density profiles on plasma discharges. These measurements have to be made in the region from 75 to 300 GHz.

A third type of applying frequencies above 100 GHz is the study of mm-wave propagation. Between typical absorption ranges by O_2 and H_2O there are propagation "windows," e.g. at 35, 95, 140 and 220 GHz. The atmospheric loss at 4 km altitude for these frequencies is only 10% compared to sea level attenuation at clear weather. Therefore, the application for military radar and communication is possible.

For all these applications, waveguide components, (filters, mixers, circulators, PIN-attenuators, isolators, directional couplers, antennas, etc.) are required and must be developed. Most of

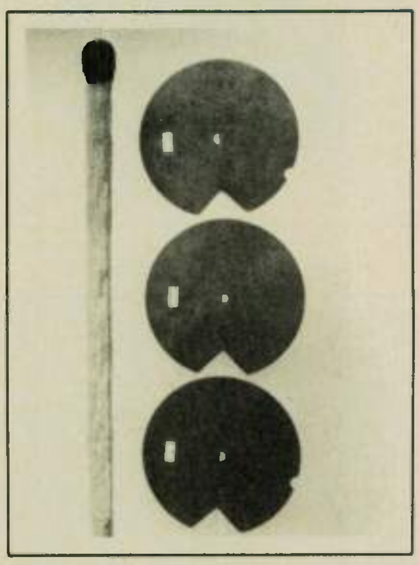

Fig. 1 Discs of the slow wave structure.

these components are made in microstripline technique or have been designed as microwave integrated circuits. Narrowband de-

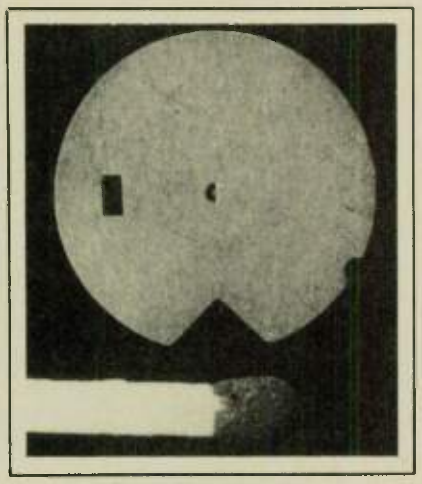

Fig. 2 Enlarged disc.

vices are sufficient in most cases, but for a special purpose broadband devices may be required.

The measurements, either in material research, plasma physics, wave propagation or in the development of components, need a frequency tunable signal source which can be swept over a wide band and leveled to a constant output power.

In the frequency range up to 40 GHz, the solid state oscillators are nearly equivalent to backward wave tubes in tuning range, but not entirely in output power. From 40 to 100 GHz, only narrow band solid state oscillators are commercially available with low power levels. In the frequency range above 100 GHz, there is no competition for a backward wave oscillator, especially in the wide tuning range 110 to 170 GHz. This oscillator is a further development of traditional BWO types, each covering a whole H_{10}-waveguide band. In four partly overlapping frequency ranges, they have worked till now in the total region from 22 to 110 Ghz.

Increasing need for shorter wavelengths was a reason for developing an oscillator in the next frequency band. The dimensions of the slow wave structure were designed by a straight forward scaling from a larger model which was the basis also for the other tubes. In order to be able to use the same power supply, the operating voltages, especially for frequency tuning, i.e. the delay line voltage, also had to be in the same range from 600 to 2200 V, which is relatively low compared with other BWO's in this high RF region.

The slow wave structure is a coupled cavity type with an inner diameter of 0.65 mm, consisting of 200 punched copper discs of three different kinds, each 30 μm thick. They are stacked and pressure welded in vacuum. A brazing, e.g. by thin gold foils, is precluded here because the cavity diameters become different from one to the other and thereby cause high mismatch.

Smallness of the slow wave structure is revealed in Figures 1-3. Figure 1 shows the three types of discs, one with the coupling slot below, one with it above, and between them the cavity disc. The rectangular hole is the waveguide with the dimensions of 1.65 mm x 0.825 mm. The enlargement of one disc is shown in Figure 2. Since the hole for the electron beam has a diameter of only 0.11 mm, the low mechanical tolerances caused severe electron beam focusing problems. The end of the slow wave structure is attenuated by two small wedges made from hard pencil lead. Figure 3 shows one of these graphite wedges.

Focusing problems required the formulation of a new design of electron gun and magnetic field. The permanent magnet of earlier model BWO's was not powerful enough for this new oscillator, and it was not possible to increase the field density with this magnetic material and design (maximum 2700 G).

A cross-section of the electron gun and the adjacent RF components of the tube is shown in Figure 4. A special cylindrical electrode called Grid 3 has the low potential of 200 to 300 V with respect to cathode and allows a sufficient beam transmission over the total range of delay line voltage from 600 to 2200 V.

As a first step, to investigate the required magnetic field density, a solenoid magnet was designed which allowed a variation of field density. Oscillation could be obtained with fields above 3700 G, the optimum was between 4600 and 5000 G. However, this solenoid magnet was very heavy (15 kg) and needed a current of 18 to 20 A at 20 V; furthermore, water-cooling was necessary.

A samarium-cobalt magnet ring with an outer diameter of only 80 mm and a height of 26 mm was sufficient for a maximum field density of 4700 G. Specially-shaped pole pieces were experimentally designed. It was possible to get nearly the same rise of the magnetic field from the cathode to field maximum as in the solenoid magnet. Figure 5 shows the shape of the field density vs Z-coordinate along the axis in both constructions. The position of the cathode and the slow wave structure is also marked in this graph.

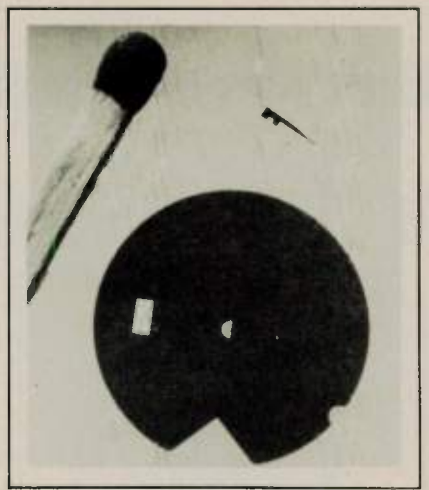

Fig. 3 Disc of the slow wave structure with graphite damping wedge.

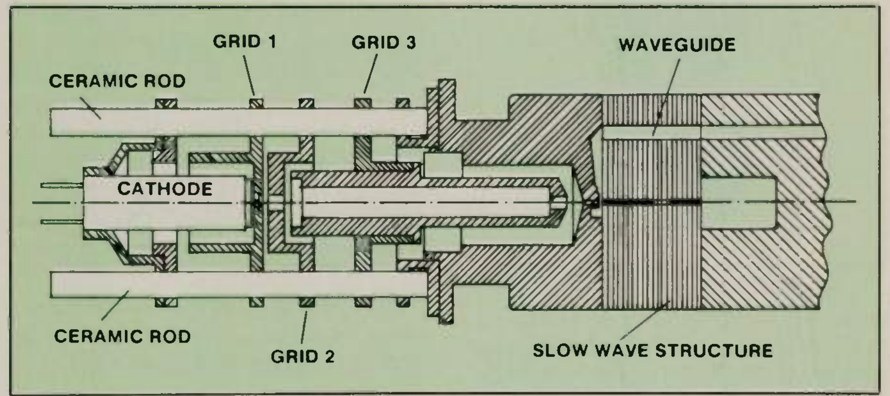

Fig. 4 Cross section of the electron gun and RF section.

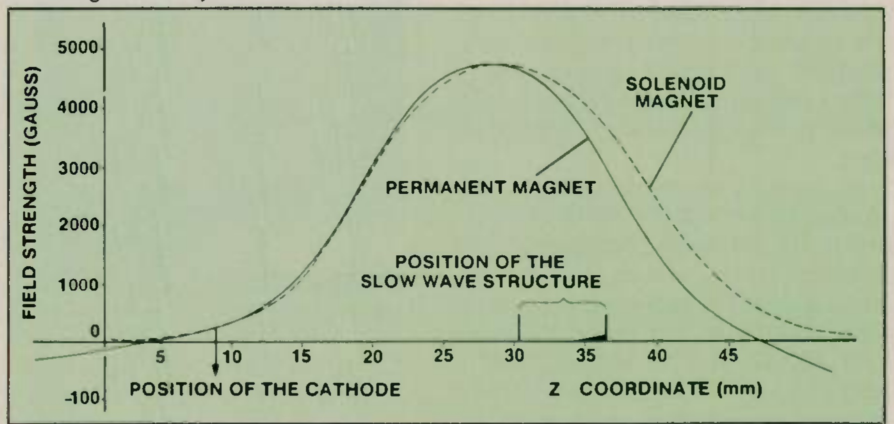

Fig. 5 Shape of the magnetic field.

In order to eliminate mechanical inaccuracies, the tube can be adjusted axially and radially in the permanent magnet. Figure 6 shows a cross-section of this magnet. After adjusting, the oscillator is sealed in the housing with a special compound. The criterion for adjusting is maximum output power.

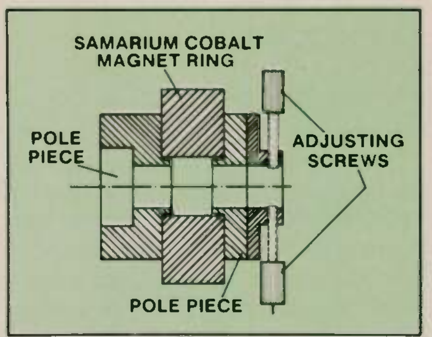

Fig. 6 Cross section of the permanent magnet.

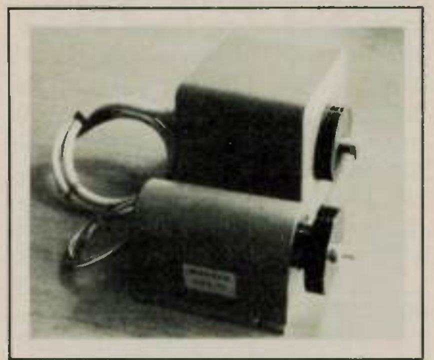

Fig. 7 Backward wave oscillator RWO 170 compared with previous focusing design BWO.

The weight of the whole oscillator is 2 kg. Figure 7 compares the size of the old permanent magnet (7 kg) and new magnet. In the near future all BWO types, including those in the lower frequency ranges, will be modified to utilize small samarium-cobalt ring magnets.

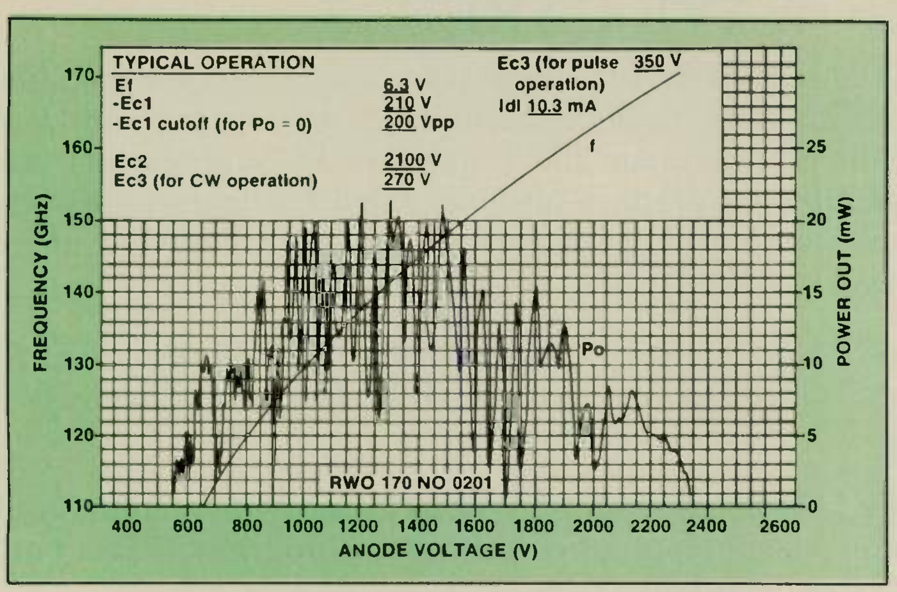

Fig. 8 Plot of power and frequency versus anode voltage.

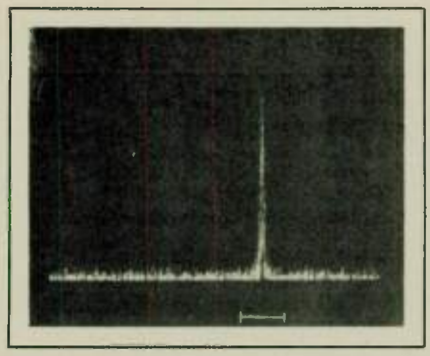

Fig. 9 Spectrum analyzer signal at 150 GHz, horizontal scale 1 MHz/div.

The cathode current of the tube is about 12 mA, and the oscillator is continuously tunable from 110 to 170GHz by the delay line voltage U_V=650 - 2300V. The recorded output power and the frequency vs U_V of the first sample is shown in Figure 8 with the operating voltages for the electron gun listed in tabular form. The spectrum analyzer signal at 150GHz as well as the spectrum width of about 300 kHz is shown in Figure 9.

Table I lists the frequency and power pushing factors for the tube and Table II, the nominal pulling figures. Among other characteristics:

- Spurious outputs are 30dB below the signal
- Power vs frequency is rather random depending on mismatch and beam transmission properties of each tube
- Load SWR's up to 3.3 will not damage tubes or lead to power or frequency holes.

A suitable power supply for operating this oscillator is available from Micro-Now Instrument Company, Chicago. All features required for quick and accurate high frequency measurements, e.g. frequency sweeping over selectable ranges and rates, internal and external modulation, tube protection, power leveling, etc. are provided with this power supply. The oscillator tube, in combination with it, is a broadband millimeter-wave source for various applications.

TABLE I
PUSHING FACTORS

Voltage	Δf (MHz/V)	ΔPo (dB/V)
E_{c1}	4	0.25
E_{c2}	0.4	0.02
E_{c3}	20	0.75

TABLE II
PULLING FIGURES

SWR	Δf (MHz)
1.4	15
2.0	30
3.3	60

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3. *International Journal of Infrared and Millimeter Waves,* Vol. 1, No. 2, June 1980, pp 255 and 295. ■

Determination of Varactor Parameters By A Modified DeLoach Method

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Burlington, MA

INTRODUCTION

In 1964, B. C. DeLoach of Bell Telephone Laboratories introduced an important new method by which varactor diode characteristics could be determined at microwave frequencies[1]. The usual difficulties of high frequency characterization were avoided by elimination of waveguide supporting posts and lossy tuning elements. Specifically, DeLoach's method involved the series resonance of a varactor in shunt with a reduced height rectangular waveguide.

DeLoach's measurements were restricted to either chip devices or packaged devices with negligible enclosure parasitics. The equivalent varactor circuit in these cases (shown in Figure 1) consists of a relatively fixed resistance (R_S) in series with a voltage variable junction capacitance [$C_j(V)$].

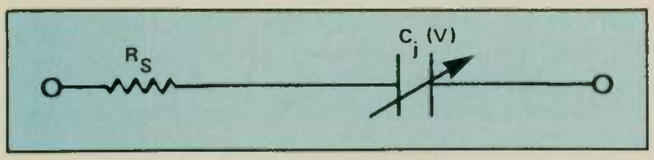

Fig. 1 Equivalent electrical circuit of a varactor chip.

In many higher frequency tuning, multiplier and parametric amplifier applications, parasitic reactances are not negligible. As an example, at high frequencies varactor junction capacitance must be kept quite low in order to maintain reasonable operational efficiencies. It is not uncommon to have a diode enclosure with a case capacitance (Cp) that is larger in magnitude than the junction capacitance. Also, in many applications, the reactance of the contacting lead inductance (L_S) is several times larger than the diode resistance (R_S). Thus, no circuit simplifications can be allowed when characterizing these varactors. Figure 2 shows the total equivalent circuit for the varactor diode.

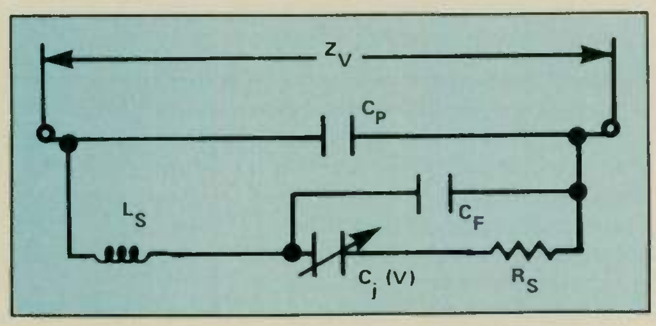

Fig. 2 Total equivalent circuit for a packaged varactor diode.

Z_V = Total Varactor Impedance

C_P = Case Capacitance

C_F = Fringing Capacitance existing primarily between the contacting strap and case pedestal.

L_S = Inductance contributed primarily by the contacting wire, ribbon or mesh.

C_J = Voltage-variable Junction Capacitance

R_S = Diode resistance due to (1) the contacts, (2) the spreading resistance in the chip, and (3) the undepleted area within the chip.

It can be shown (See Appendix I) that the total varactor impedance is given by:

$$Z_V = \frac{(-X_{C_P})\,[A + (X_{L_S})\,(B)] + j\,(X_{C_P})\,(R_S)\,(X_{C_F})^2}{R_S\,(X_{C_F})^2 + j\,[A + (X_{L_S} + X_{C_P})\,(B)]} \quad (1)$$

where:

$$A = (X_{C_F})\,[R_S{}^2 + X_{C_j}{}^2 + (X_{C_j})\,(X_{C_F})]$$

$$B = R_S{}^2 + (X_{C_j} + X_{C_F})^2$$

Multiplying the top and bottom of Equation (1) by the complex conjugate of the denominator, and separating the real and imaginary parts, we have:

$$\mathrm{Re}\,[Z_V] = \frac{(-R_S)\,(X_{C_F})^2\,(X_{C_P})\,[A + (X_{L_S})\,(B)] - [A + (X_{L_S} + X_{C_P})\,(B)]}{(R_S)^2\,(X_{C_F})^4 + [A + (X_{L_S} + X_{C_P})\,(B)]^2} \quad (2)$$

$$\mathrm{Im}\,[Z_V] = \frac{(R_S)^2\,(X_{C_F})^4\,(X_{C_P}) + X_{C_P}\,[A + (X_{L_S})\,(B)]\,[A + (X_{L_S} + X_{C_P})\,(B)]}{(R_S)^2\,(X_{C_F})^4 + [A + (X_{L_S} + X_{C_P})\,(B)]^2} \quad (3)$$

Analysis of these equations shows that there are, in addition to the lead inductance, three different capacitances that influence device performance (Cp, C_F and Cj). The package capacitance (Cp of the diode) is determined by measuring the capacitance of an empty diode case assembly (i.e., no die or connecting element present) on a 1 MHz capacitance bridge. Package capacitance, however, is not the only influential non-voltage variable capacitance present in an encased varactor. In low capacitance ($Cj0 \leq 0.4$ pF) varactors, especially devices designed to minimize lead inductance, the capacitance between the diode's connecting lead or leads and the case assembly's pedestal can greatly influence the determination of both varactor figure of merit (cutoff frequency-F_C) and device nonlinearity. This capacitance is known as fringing capacitance (C_F) and can be easily determined for any case assembly and lead configuration. Preformed leads or straps will assure repeatability of the fringing capacitance between devices of the same type.

Appendix II describes a method for determining fringing capacitance. Junction capacitance, Cj (V), can be measured as with Cp, on a 1 MHz capacitance bridge.

The figure of merit for a varactor diode is a function of voltage and is given by:

$$F_{C_V} = \frac{1}{2\pi R_S C_{jV}} \quad (4)$$

METHOD OF MEASUREMENT

The DeLoach[1] method of determining varactor cutoff frequency requires that the diode under test be resonated in a well-matched reduced height waveguide environment.

Figure 3 shows an axial pronged ceramic package imbedded in an appropriate X-band test holder. The selection of the test holder and the test band is determined by the physical dimensions of the diode under test as well as by the diode's junction capacitance. Over twenty different holders have been designed ranging in frequency from 4 GHz to 40 GHz to use for these measurements. Figure 4 shows a typical X-Band DeLoach holder. In each holder, the holder height is reduced by use of a well matched step transition to the ceramic height of the diode under test. Figure 5 shows a cross section of a typical DeLoach holder.

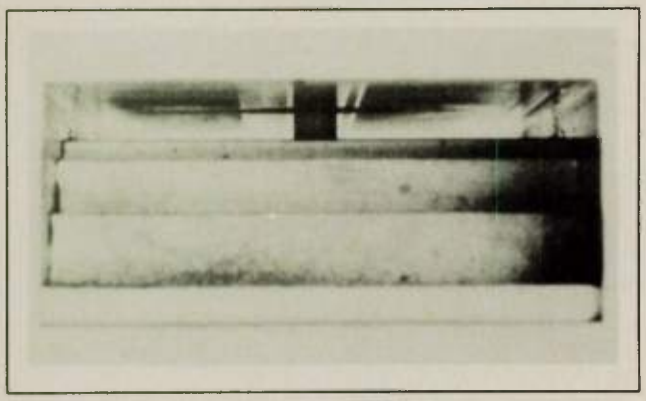

Fig. 3 Diode under test imbedded in DeLoach holder.

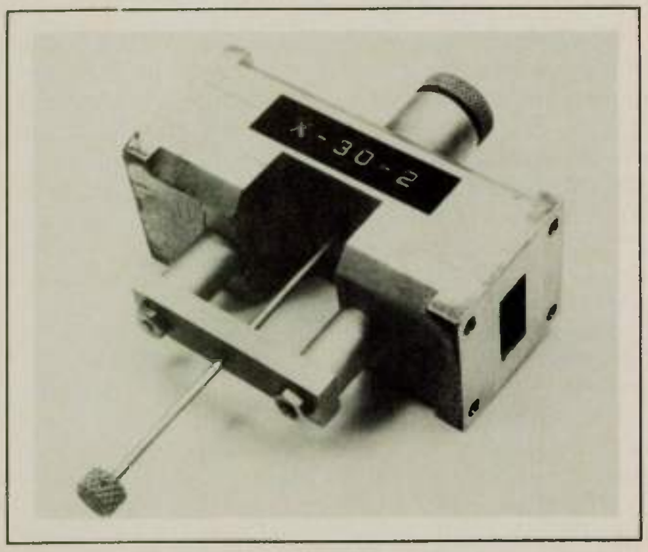

Fig. 4 X-band DeLoach holder.

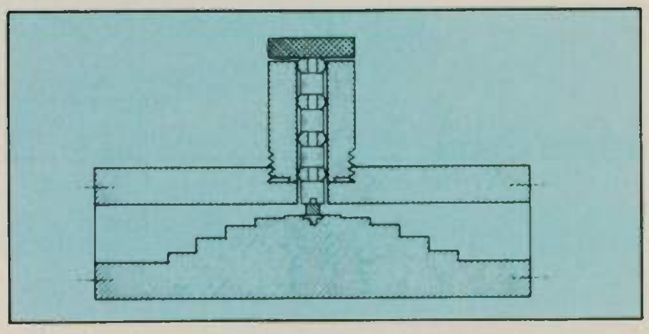

Fig. 5 Typical DeLoach holder.

The characteristic impedance of the reduced height environment can be calculated if we asssume the diode junction to be electrically short and not unsimilar to a small diameter wire connected transversely between the centers of the broad walls of the reduced height guide[2,3]. This being the case, then:

$$Z_0 = 754 \left(\frac{B}{A}\right)\left(\frac{\lambda_g}{\lambda_0}\right) \quad (5)$$

Where "A" and "B" are the broad waveguide dimension and reduced height dimension, respectively λ_g is the dominant mode guided wavelength in the diode environment at the frequency of measurement and λ_o is the free space wavelength.

At resonance, the varactor impedance is pure real (though not equal to R_S as shown in Equation 2) and can be represented in its waveguide environment as a lumped circuit.

Appendix III shows the derivation of the real part of the varactor with respect to the power transmission loss ratio. Equation 6 shows this relationship.

$$Re\,[Z_V] = \left(\frac{377}{T_P - 1}\right)\left(\frac{B}{A}\right)\left(\frac{\lambda_g}{\lambda_0}\right) \quad (6)$$

The test setup for the actual measurement is shown in the block diagram of Figure 6. It is very important to keep the incident power on the test holder below about 200 microwatts so as not to self bias the varactor. GaAs varactors, since they do not store charge, are good RF rectifiers and will readily self bias when pumped with an RF signal. The step by step measurement procedure is as follows:

1. Using the test set of Figure 6 (in the appropriate waveguide band) obtain a zero loss reference on the scope with the diode under test removed from the circuit and with the holder choke in place. The generator should be adjusted to sweep over the whole waveguide band.
2. Insert the varactor to be tested into the holder and apply the desired bias voltage.
3. Locate the characteristic varactor resonance typified by reduced power to the circuit detector.
4. Reduce sweep range to center resonance on scope face and locate and record resonant frequency using frequency meter, as shown in Figure 7.

5. Measure the transmission loss at the resonant frequency (using the precision 50 dB attenuator) by the substitution method.
6. Measure package capacitance (Cp) of a typical package at 1 MHz as well as junction capacitance at the bias level of the transmission loss measurement.
7. Refer to Appendix II to determine the fringing capacitance (C_F).
8. The real part of the varactor impedance at resonance can now be found using Equation 6.
9. Using a digital computer (in conjunction with equations 2, 3 and 4) R_S, L_S and $FC_{(V)}$ can now be determined by following discrete steps.

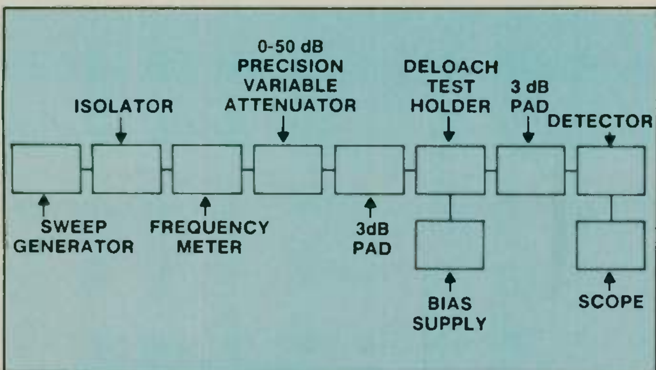

Fig. 6 Test set-up for DeLoach measurement.

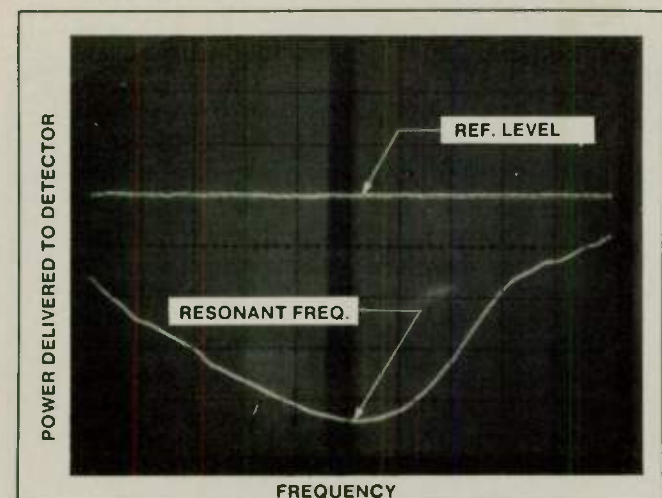

Fig. 7 Varactor resonance.

The first step is to assume that L_S is some small value, such as 0.1 nH and vary R_S in small increments until Equation (2) is satisfied. Using the final value of R_S determined here, enter Equation (3) by setting Im $[Z_V] = 0$, (a known condition at resonance), and stepping L_S. The resulting final value of L_S is brought back to Equation (2) and the whole procedure is reiterated until a value of R_S and L_S has been found that will both satisfy Equation (2) and set Equation (3) equal to zero (within reasonable limits). Cutoff frequency [$FC_{(V)}$] is then calculated using Equation (4).

PRACTICAL EXAMPLE

In order to help clarify the procedure, if we wish to characterize an MA-46600K-30 GaAs abrupt junc-

[Continued on page 86]

tion tuning varactor over an extended bias range we would first select an X-band DeLoach Holder with a reduced height of 0.062 inches (the ceramic height of the type 30 case style). The capacitance, transmission loss and resonant frequency data, as a function of bias voltage, would typically be as shown in Table 1.

TABLE 1

DeLoach Data

Reverse Bias Voltage (Volts)	C_{jV} (PF)	T_P (DB)	F_O (GHz)
0	0.466	28.4	8.20
0.24	0.420	28.6	8.35
0.48	0.381	28.8	8.91
0.81	0.345	28.8	9.28
1.26	0.308	29.1	9.68
1.76	0.280	29.3	10.07
2.37	0.253	29.7	10.45
3.15	0.229	30.4	11.44
4.08	0.207	30.8	11.23
5.27	0.188	31.3	11.61
6.58	0.172	32.8	11.97

Using procedures similar to those described in Appendix II, the fringing capacitance (CF) is determined to be 0.027 pF. The package capacitance of a gutless package is measured to be typically 0.18 pF. Using the computer program previously described, the lead inductance (L_S) is determined to be typically 0.813 nH (in this specific reduced height environment). The additional calculated results are shown in Table 2.

TABLE 2

Computed Results

Reverse Bias Voltage (Volts)	Re [Z_V] (Ohms)	R_S (Ohms)	F_C (GHz)	Q @ 50 MHz
0	1.71	1.91	178.8	3,576
0.24	1.62	1.82	229.5	4,590
0.48	1.45	1.65	253.2	5,063
0.81	1.38	1.60	288.3	5,767
1.26	1.28	1.51	342.2	6,844
1.76	1.22	1.46	389.3	7,787
2.37	1.13	1.37	459.2	9,184
3.15	0.99	1.23	565.0	11,301
4.08	0.95	1.20	640.7	12,814
5.27	0.88	1.14	742.6	14,852
6.58	0.73	0.97	953.9	19,079

Since this particular device is a tuning varactor, the classical tuning varactor figure of merit is Q as referenced to a frequency of 50 MHz. Q at 50 MHz is determined from the values of R_S and C_j by:

$$Q = \frac{1}{2\pi(50)(10)^6(C_j)(R_S)} \qquad (7)$$

A graphical representation of 50 MHz Q as a function of bias voltage for the MA-46600K-30 is shown in Figure 8.

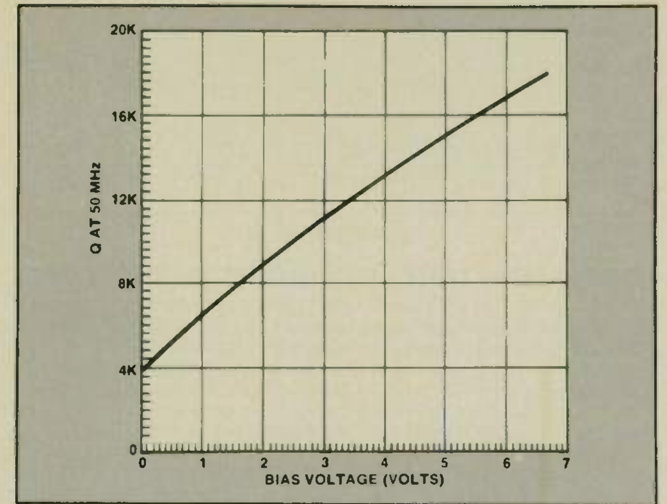

Fig. 8 MA-46600-K-30 varactor-typical Q vs bias voltage.

CONCLUSION

We have presented a method for accurately characterizing high quality packaged varactor diodes that takes into account the influence of package parasitics. As can be seen from the results of Table 2, a measurement that results in the real part of varactor impedance does not (without computer analysis) yield the true R_S of the device and thus will result in an optimistic varactor figure of merit. Additionally, by its very nature, this method yields the series resonant frequency of each varactor tested in a defined waveguide environment. This information can provide acute selectivity for matching diodes within a production lot.

The author wishes to thank Dr. Joseph White for his helpful suggestions and Dave McQueen for his help with the measurements.

APPENDIX I.

DERIVATION OF VARACTOR IMPEDANCE

The derivation of total varactor impedance can be determined as follows:

$$Z_V = \frac{\left|\dfrac{(R_S + jX_{C_j})(jX_{C_F})}{R_S + jX_{C_j} + jX_{C_F}} + jX_{L_S}\right| \left|jX_{C_P}\right|}{\dfrac{(R_S + jX_{C_j})(jX_{C_F})}{R_S + jX_{C_j} + jX_{C_F}} + jX_{L_S} + jX_{C_P}}$$

The total equivalent circuit of Figure 2 (see text) can be broken down into the following separate parts thus aiding derivation:

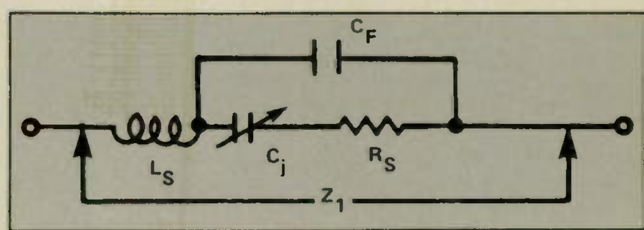

Fig. 9 Branch 1 of equivalent circuit.

[Continued on page 88]

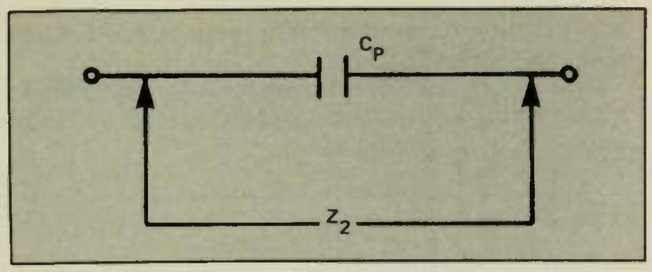

Fig. 10 Branch 2 of equivalent circuit.

Analyzing Branch # 1

$$Z_1 = \frac{(R_S + jX_{C_j})(jX_{C_F})}{R_S + jX_{C_j} + jX_{C_F}} + jX_{L_S}$$

Expanding:

$$Z_1 = \left| \frac{[\,j\,(R_S)(X_{C_F}) - (X_{C_j})(X_{C_F})]}{[R_S + j\,(X_{C_j} + X_{C_F})]} \right| \cdot \cdot \cdot \cdot \cdot \cdot$$

$$\left| \frac{[R_S - j\,(X_{C_j} + X_{C_F})]}{[R_S - j\,(X_{C_j} + X_{C_F})]} \right| + jX_{L_S}$$

and:

$$Z_1 = \frac{j\,(R_S)^2\,(X_{C_F}) - R_S\,(X_{C_j})(X_{C_F})}{R_S^2 + (X_{C_j} + X_{C_F})^2} \cdot \cdot \cdot \cdot \cdot \cdot$$

$$+ \frac{R_S\,(X_{C_F})(X_{C_j} + X_{C_F}) + j\,(X_{C_j})(X_{C_F})(X_{C_j} + X_{C_F})}{R_S^2 + (X_{C_j} + X_{C_F})^2} + jX_{L_S}$$

collecting:

$$Z_1 = \frac{R_S\,(X_{C_F})^2}{R_S^2 + (X_{C_j} + X_{C_F})^2} \cdot \cdot \cdot \cdot \cdot$$

$$+ \frac{j\,X_{C_F}\,[R_S^2 + X_{C_j}^{\,2} + X_{C_j}\,X_{C_F}]}{R_S^2 + (X_{C_j} + X_{C_F})^2} \cdot \cdot \cdot$$

$$+ \frac{X_{L_S}\,[R_S^2 + (X_{C_j} + X_{C_F})^2]}{R_S^2 + (X_{C_j} + X_{C_F})^2}$$

Referring to Branch # 2:

$$Z_2 = jX_{C_P}$$

Combining the two impedances to get Z_V:

$$Z_V = \frac{(Z_1)(Z_2)}{(Z_1 + Z_2)}$$

or expanding:

$$Z_V = \frac{(-X_{C_P})\,[A + (X_{L_S})\,(B)] + j\,(X_{C_P})\,(R_S)\,(X_{C_F})^2}{R_S\,(X_{C_F})^2 + j\,[A + (X_{L_S} + X_{C_P})\,(B)]}$$

where:

$$A = (X_{C_F})\,[R_S^{\,2} + X_{C_j}^{\,2} + (X_{C_j})\,(X_{C_F})]$$

$$B = R_S^{\,2} + (X_{C_j} + X_{C_F})^2$$

By multiplying the top and bottom of this equation by the complex conjugate of the denominator, and separating the real and imaginary parts, we have:

$$\mathrm{Re}\,[Z_V] = \frac{(-R_S)\,(X_{C_F})^2\,(X_{C_P})\,\left|\,[A + (X_{L_S})\,(B)] - [A + (X_{L_S} + X_{C_P})\,(B)]\,\right|}{(R_S)^2\,(X_{C_F})^4 + [A + (X_{L_S} + X_{C_P})\,(B)]^2}$$

$$\mathrm{Im}\,[Z_V] = \frac{(R_S)^2\,(X_{C_F})^4\,(X_{C_P}) + X_{C_P}\,[A + (X_{L_S})\,(B)]\,[A + (X_{L_S} + X_{C_P})\,(B)]}{(R_S)^2\,(X_{C_F})^4 + [A + (X_{L_S} + X_{C_P})\,(B)]^2}$$

APPENDIX II.

DETERMINATION OF FRINGING CAPACITANCE

To identify C_F one must first carefully measure the capacitance of several empty case assemblies and maintain the identity of each. Next, a group of relatively high capacitance [$\geq$ (30) (C_P)] varactor dice are selected. It is necessary that these dice all be from the same wafer and preferably from the same area of that wafer. These dice are then bonded or soldered into the premeasured case assemblies. A repeatable top contact, of the fashion desired, is made to each die. At this point, some of the devices are etched so that their total capacitance $\cong$ (2) (C_P). All of the devices are then capped and total capacitance is measured at both zero and minus six volts at 1 MHz. The rate of capacitance change between zero and six volts (irrespective of the magnitude of the measured total capacitances) of each die is given by:

$$\frac{C_{j0}}{C_{j\text{-}6}} = \left(1 + \frac{6}{\phi}\right)^{\gamma}$$

In the case of a GaAs diode:

$$\phi \cong 1.1$$

Thus:

$$\frac{C_{j0}}{C_{j\text{-}6}} = (6.455)^{\gamma}$$

Defining the total capacitance measurement:

$$C_{T_V} = C_{j_V} + C_P + C_F$$

By substitution:

$$\frac{C_{T_0} - C_P - C_F}{C_{T\text{-}6} - C_P - C_F} = (6.455)^{\gamma}$$

In the case of the high capacitance devices [$\geq$ (30) (CP)], C_F is low enough so that it has a minimal effect on the above equation. Thus, it can be eliminated and the capacitance change law, γ, for the wafer section can be calculated using the total capacitance measurements and the pre-recorded C_P measurements.

Rearranging the equation:

$$C_F = \frac{\left|(6.455)^{\gamma}\right|\left|C_{T\text{-}6} - C_P\right| - \left|C_{T_0} - C_P\right|}{\left|(6.455)^{\gamma} - 1\right|}$$

Using the capacitance change law derived from the measurements on the high capacitance diodes, and the data recorded on the low capacitance [$\cong$(2) (C_P)] diodes, this equation will yield the value of the fringing capacitance for the particular lead and package configuration involved. Fringing capacitance will vary from diode to diode, but as a rule of thumb, using preformed 0.005" x 0.00025" connecting straps, is equal to approximately 15% of the case capacitance.

APPENDIX III.

VARACTOR IMPEDANCE

The object is to relate the transmission loss of the appropriately mounted varactor at resonance, to the real part of the varactor impedance.

[*Continued on page 90*]

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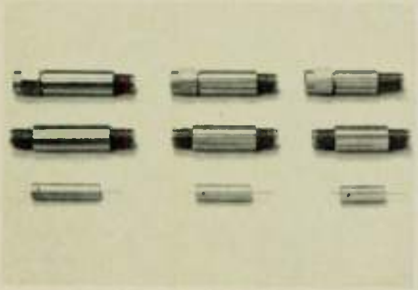

STANDARD DETECTORS

Omniyig Model #	Freq. Range (GHz)	k Factor	T_{ss} (dBm)
Tunnel			
DT0D1P	0.1- 0.5	800	−50
DT2C1P	1.0- 2.0	800	−50
DT9A1P	2.0-18.0	450	−48
Schottky*			
DS1C1N	0.5- 1.0	2000	−54
DS6A1N	8.0-16.0	1500	−52
DS9A1N	2.0-18.0	1000	−51

*Bias = 100 - 200 μamp

STANDARD LIMITERS

Omniyig Model #	Freq. Range (GHz)	Ins. Loss (dB)	Lkg. Pwr. (dBm)**
Pin			
PL0E1	0.1-0.5	0.5	+20
PL2D1	1.0-2.0	0.6	+20
PL4C1	4.0-8.0	1.2	+18
Schottky Turn-on			
SL0E1	0.1-0.5	0.5	+15
SL1E1	0.5-1.0	0.7	+15
SL3D1	2.0-4.0	1.2	+14

**Measured at P_{IN} = +30 dBm

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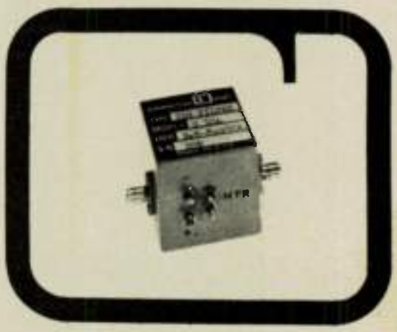

STANDARD UNITS

Model	Freq. (Ghz)	Insertion Loss (Max)	Min. Peak Rejection
P202R	.5-1	1.5db	20db
L202R	1-2	1.5db	20db
S202R	2-4	2db	30db
C202R	4-8	2db	30db
X202R	8-12	2db	35db
Ku202R	12-18	2db	45db
P204R	.5-1	1.5db	25db
L204R	1-2	2db	30db
S204R	2-4	2db	30db
C204R	4-8	2db	30db
X204R	8-12	2db	30db
Ku204R	12-18	2db	30db
M202R	.5-4	2db	20db
M203R	2-8	2db	30db
M204R	4-12	2db	35db
M205R	8-18	2db	30db

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YM1002	100Mhz	1-12	-33dBm
YM1003	200Mhz	1-12	-28dBm
YM1004	500Mhz	1-12	-10dBm
YM1026	1-2Ghz	2-18	4dBm
YM1027	100Mhz	1-18	-40dBm
YM1028	200Mhz	1-18	-34dBm
YM1029	500Mhz	1-18	-22dBm
YM1087	.1-.2Ghz	1-12	-30dBm

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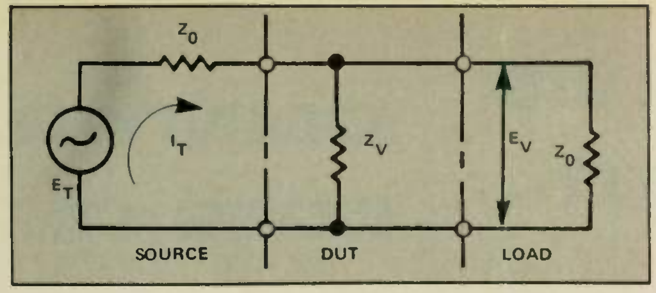

Fig. 11 Equivalent constant voltage representation of varactor diode at resonance.

E_T = Constant source voltage
Z_V = Varactor impedance
Z_0 = Characteristic waveguide impedance

Now:

$$I_T = \frac{E_T}{Z_0 + \left(\frac{(Z_V)(Z_0)}{Z_V + Z_0}\right)}$$

$$E_V = \left(\frac{E_T}{Z_0 + \frac{(Z_V Z_0)}{Z_V + Z_0}}\right)\left(\frac{(Z_V)(Z_0)}{Z_V + Z_0}\right)$$

Simplifying:

$$E_V = \frac{E_T}{2 + \left(\frac{Z_0}{Z_V}\right)}$$

The power delivered to the load (P_L) can now be calculated:

$$P_L = \frac{|E_V|^2}{Z_0}$$

Now by substitution:

$$P_L = \left(\frac{|E_T|^2}{Z_0}\right)\left(\frac{1}{2 + \left(\frac{Z_0}{Z_V}\right)}\right)\left(\frac{1}{2 + \left(\frac{Z_0}{Z_V^*}\right)}\right)$$

Since we are at resonance, and Z_V is not complex;

$$P_L = \frac{|E_T|^2}{Z_0\left(2 + \left(\frac{Z_0}{\mathrm{Re}\,[Z_V]}\right)\right)^2}$$

The maximum power available (P_A) to the load from the source can be determined if we envision Z_V at anti-resonance. (Z_V approaching infinity.)

Now:

$$E_V = \frac{E_T}{2}$$

and:

$$P_A = \frac{|E_V|^2}{Z_0} = \frac{\left|\frac{E_T}{2}\right|^2}{Z_0}$$

or:

$$P_A = \frac{|E_T|^2}{(4\,Z_0)}$$

By definition:

$$T_P = \frac{P_A}{P_L}$$

where:

T_P = Power transmission loss ratio

P_A = Power available from the source

P_L = Power delivered to the load

By substitution:

$$T_P = \frac{\left(\frac{|E_T|^2}{4\,Z_0}\right)}{\frac{|E_T|^2}{Z_0\left(2 + \left(\frac{Z_0}{\mathrm{Re}\,[Z_V]}\right)\right)^2}}$$

simplifying:

$$T_P = \frac{\left(2 + \frac{Z_0}{\mathrm{Re}\,[Z_V]}\right)^2}{4}$$

rearranging:

$$\mathrm{Re}\,[Z_V] = \left(\frac{Z_0}{2}\right)\left(\frac{1}{\sqrt{T_P} - 1}\right)$$

substituting into Equation 4 (see text):

$$\mathrm{Re}\,[Z_V] = \left(\frac{377}{\sqrt{T_P} - 1}\right)\left(\frac{B}{A}\right)\left(\frac{\lambda_g}{\lambda_0}\right)$$

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These units are internally terminated circulators (isolators) with SMA female connectors. Many are available from stock

Frequency (GHz)	Model No.	Isolation (dB min.)	Insertion Loss (dB max.)	VSWR (max.)	Height	Size Width	Thickness
0.5 - 1.0	T-0S63T	17	0.8	1.35:1	5.78	5.59	1.35
1.0 - 2.0	T-1S63T-18	18	0.5	1.30:1	2.75	2.75	0.88
2.0 - 4.0	T-2S63T-6	17	0.5	1.35:1	1.63	1.63	0.75
2.6 - 5.2	T-2S63T-44	17	0.5	1.35:1	1.25	1.25	0.70
4.0 - 8.0	T-4S63T-10	17	0.4	1.35:1	1.06	1.00	0.76
4.5 - 9.0	T-4S63T-13	17	0.5	1.35:1	1.13	0.95	0.76
5.2 - 10.4	T-5S63T	17	0.5	1.35:1	1.06	1.00	0.76
8.0 - 16.0	T-8S63T-18	17	0.5	1.35:1	0.75	0.63	0.40
10.0 - 20.0	T-10S63T-5	17	0.7	1.35:1	0.68	0.51	0.56

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Both circulators and isolators are available with either SMA-male or female connectors. Model Nos. shown are isolator versions with SMA-female connectors.

Frequency (GHz)	Model No.	Isolation (dB min.)	Insertion Loss (dB max.)	VSWR (max.)	Height	Size Width	Thickness
1.7 - 4.2	T-1S83T-2	16	0.7	1.50:1	1.70	1.63	0.76
2.0 - 4.5	T-2S73T-4	16	0.6	1.40:1	1.70	1.56	1.10
3.7 - 8.2	T-3S73T-2	16	0.7	1.40:1	1.06	1.00	0.76
4.4 - 10.0	T-4S73T-2	16	0.7	1.40:1	1.13	0.95	0.76
5.9 - 13.0	T-5S73T-1	17	0.6	1.35:1	0.81	0.63	0.80
7.6 - 18.0	T-7S83T-20	16	0.8	1.50:1	0.76	0.63	0.62

POPULAR NARROW BAND — STANDARD DESIGNS

Frequency (GHz)	Model No.	Isolation (dB min.)	Insertion Loss (dB max.)	VSWR (max.)	Height	Size Width	Thickness
.95 - 1.225	T-0S23T-2	20	0.5	1.25:1	1.20	1.20	0.75
1.2 - 1.6	T-1S23T-7	17	0.5	1.35:1	1.25	1.25	0.70
1.9 - 2.3	T-1S13T-2	20	0.4	1.30:1	1.25	1.25	0.75
2.2 - 2.3	T-2S03T-11	20	0.4	1.35:1	1.05	1.00	0.58
3.7 - 4.2	T-3S13T-9A	25	0.25	1.10:1	0.75	0.75	0.50
4.4 - 6.5	T-4S33T-1	17	0.5	1.35:1	0.75	0.75	0.50
5.9 - 6.4	T-5S03T-3A	26	0.3	1.10:1	0.75	0.75	0.69
7.0 - 11.0	T-7S43T-6	28	0.4	1.10:1	0.85	0.75	0.50
8.0 - 12.4	T-8S43T-1A	17	0.4	1.35:1	0.78	0.63	0.70
12.4 - 18.0	T-12S43T-8	18	0.5	1.30:1	0.68	0.51	0.56
18.0 - 26.5	T-18S33T-7	16	1.0	1.50:1	0.68	0.51	0.53

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Frequency Range of Large-Scale TEM Mode Rectangular Strip Lines

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Much confusion exists regarding the correct criteria for determining the cut-off frequency of the first higher order mode in rectangular strip lines. These criteria are reviewed, and it is shown that the actual frequency range of commercially-available TEM mode transmission lines is considerably less than that advertised by the manufacturers. Valid methods of extending the frequency range of these structures are discussed, and it is concluded that optimum performance is obtained by reverting to a parallel-plane strip line structure.

INTRODUCTION

Large-scale rectangular strip transmission lines which contain a propagating transverse electromagnetic (TEM) field are finding widespread application in such areas as EM susceptibility and emissions testing, calibration of radiation survey meters and electric field probes, as well as studies on the biological effects of RF exposure. These structures are characterized by an air dielectric and a thin center conductor surrounded by a rectangularly-shaped shield. This provides for an optimally-sized test space within the line in which equipment, field probes, or animals, etc. may be exposed to a well-defined and reasonably uniform field. Crawford[1] has discussed the properties of such lines, as well as their advantages, and has described a family of TEM "cells" constructed at the National Bureau of Standards. Other uses have been reviewed by Weil[2] in a paper which primarily discusses design criteria for rectangular strip transmission lines. These devices are now commercially available from two sources: Instruments for Industry, Inc., (IFI), Farmingdale, NY, and Narda Microwave Corporation, Plainview, NY. They have been termed "Crawford cells" or "TEM transmission cells" by their respective manufacturers.

The ability to create an accurately defined and reasonably uniform field represents the chief asset of these devices. There is undoubtedly universal agreement that this can only occur under conditions in which the line is operating in the basic TEM mode only. The presence of any higher order modes will seriously disturb the field uniformity within these lines. Thus, any experiments or testing procedures that are critically dependent on a known and uniform field structure will surely be compromised or invalidated if knowingly or unknowingly performed under conditions in which higher order modes are present. As Crawford[1] has stated: "Higher order modes would greatly complicate the interpretation of the measured results of the cell."

In order to avoid problems with higher order modes, an upper frequency limit, where the first higher order mode begins to propagate, has generally been established for these lines. Regrettably, there appears to be considerable confusion regarding the proper criteria by which this upper frequency limit is defined. The purpose of this article is to point out the discrepancies which exist in the criteria used by different designers for determining an upper frequency limit, to establish the correct criteria, and to provide researchers with corrected data on recommended upper limits for commercially-available devices. This last purpose appears particularly important and urgent in as much as many researchers may be using such equipment in critical testing and calibration experiments under conditions where higher order mode fields may be unknowingly present. Valid methods of extending the upper frequency limits will also be discussed.

DISCUSSION OF CUT-OFF CRITERIA

An examination of the technical specifications published by both manufacturers of commercially-available devices shows that the advertised upper frequency limit appears to have been derived using the criteria $\lambda_c/2=b$, where "b" is the internal height of the line's cross section as shown in Figure 1 and λ_c is the cut-off wavelength. In the case of the IFI Crawford cells, a 20% safety factor reduction appears to have been applied to the cut-off value predicted using the above criteria in the case of Model Nos. CC-101, 101.5, 102, and 103, but not for the smaller Model Nos. CC-105 and 110. No safety factor has been applied in

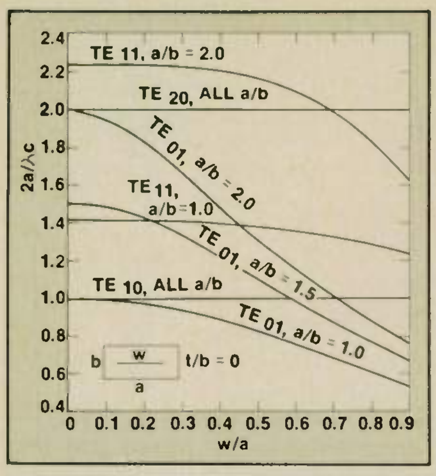

Fig. 1 Normalized cut-off frequency versus w/a for different H-modes in rectangular strip lines with a/b = 1.0, 1.5 and 2.0.

the case of the Narda Models 8801 and 8802 TEM transmission cells.

While the $\lambda_c/2 = b$ criteria is quite correct for the case of the first higher order mode in parallel plane strip-line and microstrip-type structures which have open side walls, it is definitely not correct for the case of the rectangular strip-lines with closed side walls. A closer analogy to this structure is that of rectangular waveguide and not the open strip-line. Consequently, a more accurate approximation might be to use the cut-off criteria for the dominant TE_{10} mode in waveguide, namely $\lambda_c/2 = a$, where "a" is the internal width. Crawford[1], proposed this criteria for predicting the cut-off of the first higher mode. This is indeed the correct cut-off criteria for the TE_{10} mode in rectangular strip-line structures, but, as will be seen shortly, the TE_{10} mode is generally not the first higher order mode to propagate in these structures.

From the material which follows it will be shown that the first higher order mode to propagate in rectangular strip-lines is frequently the TE_{01} mode. Consequently it must be concluded that both the $\lambda_c/2 = b$ and $\lambda_c/2 = a$ criteria for defining the upper frequency limit of these lines are incorrect.

That the TE_{01} mode rather than the TE_{10} mode is generally the first higher mode in rectangular strip lines can be physically explained as follows: if the presence of the center strip conductor is temporarily ignored, then it is well known that the cut-off wavelengths for the TE_{10} and TE_{01} modes for a rectangular waveguide are $\lambda_{c10} = 2a$ and $\lambda_{c01} = 2b$ respectively. Consequently, $\lambda_{c10} = a/b\ \lambda_{c01}$; if this is expressed in terms of cut-off frequencies, we obtain $f_{c01} = a/b\ f_{c10}$. Thus it is seen that as long as $a > b$, then $f_{c01} > f_{c10}$ and the TE_{10} mode is always the first higher order mode. The presence of the center strip will not substantially alter this analysis provided that the strip is narrow in width relative to the dimension 'a'. However, when the center strip is relatively wide, its presence can no longer be neglected owing to capacitive coupling between the edge of the center strip and the side walls of dimension 'b'. This coupling is sufficient to cause the structure to have electrical dimensions that appear greater in height 'b' compared to width 'a'. In such cases, the first higher mode will be the TE_{01} mode even when $a > b$; i.e. $f_{c01} > f_{c10}$.

These predictions have been recently confirmed by Gruner[3] who has analyzed the higher order mode properties of rectangular strip lines with thin center conductors. In his paper, Gruner shows that a close analogy does exist between rectangular waveguide and rectangular strip-line. However, the cut-off for TE_{mm} and TM_{mn} modes is altered relative to that in waveguide for all cases in which the integer "n" is odd while the cut-off for the "n" even modes remains unchanged for the strip line with an infinitely thin center conductor. Of the two sets of modes whose cut-off is altered, the H-modes (TE) are more significant, since cut-off is generally lowered relative to that of waveguide for these modes. In Figure 1, Gruner's data have been replotted in terms of the normalized cut-off frequencies versus the parameter w/a, for four different H-modes and three aspect ratio values a/b = 1.0, 1.5 and 2.0 (Data on the TE_{01} mode characteristics were obtained from Reference 3 while the TE_{11} mode data were derived from a subsequent paper dealing with crossed rectangular strip-line structures[4].) The cut-off for the TE_{10} mode remains unaltered by the presence of the strip conductor and continues to be governed by the criteria $\lambda_c = 2a$. Since the TE_{10} cut-off is invariant with respect to w/a, it appears in Figure 1 as a horizontal line at $2a/\lambda_c = 1.0$. When the curves labelled TE_{01} in Figure 1 pass below the horizontal TE_{10} line a condition exists whereby the TE_{01} mode rather than the TE_{10} mode is the first higher order mode to propagate. It can be seen that this condition always exists in the square structure (a/b = 1.0) while for larger values of a/b it occurs over a more limited range of the parameter w/a. Data on the TE_{11} mode have been included in Figure 1 because it was important

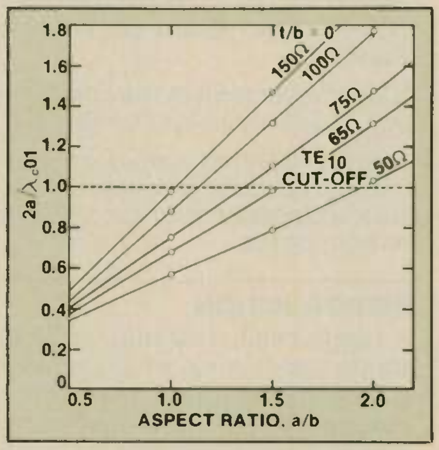

Fig. 2 Normalized cut-off frequency versus a/b for the TE_{01} mode in rectangular strip lines of different characteristic impedance.

to know whether or not this mode is also capable of propagating at frequencies below the cut-off for the TE_{10} mode. It can be seen that this is not possible since the curves labelled TE_{11} never pass below the TE_{10} line.

In order to eliminate the parameter w/a, it is instructive to replot the data of Figure 1 in terms of the normalized TE_{01} frequency cut-off versus aspect ratio a/b for several different values of characteristic impedance, Z_0. The characteristic impedance data for rectangular strip-lines were derived from Weil's review[2]. From Figure 2, it is apparent that this relationship is close to linear and can be extrapolated to the left, giving TE_{01} cut-off data for values of a/b < 1. The curve labelled '50Ω' in Figure 2 provides some useful and practical data on most existing strip-line facilities which generally possess a characteristic impedance of about 50 ohms. It can be seen that for aspect ratios a/b < 1.93, the first higher order mode in 50 ohm rectangular strip lines is always the TE_{01}; for a/b > 1.93 the first higher order mode is the TE_{10}.

Using Figure 2, it can be shown that the correct TE_{01} cut-off criteria for the IFI series of Crawford cells, which have an a/b ratio of 1.5 and a characteristic impedance of 55 ohms, is given by $2a/\lambda_c = 0.80$. Similarly, the correct criteria for the Narda TEM transmission cells, which have an a/b ratio of 1.667 and a 50 ohm characteristic impedance is given by $2a/\lambda_c = 0.85$. The corrected values for TE_{01} cut-off derived using these criteria, are listed in Table 1 for both series of commercially-available equipment. The upper frequency limit of operation recommended by the manufacturer is also shown for comparison purposes, as well as the intermediate cut-off for the TE_{10} mode. Similarly, Table 2 gives the corrected upper frequency limit for the TEM transmission cells designed and built by Crawford at NBS[1].

The data given in Tables 1 and 2 demonstrate that the corrected upper frequency limit for commercially-available lines is considerably less than that recommended by the manufacturer. For the IFI Crawford cells, there is a 34-47% reduction in advertised bandwith, depending on the particular model in use, while for the Narda TEM transmission cells, the corresponding figure is 49%. Similarly for the NBS transmission cells there is a 42% reduction in the bandwith claimed by Crawford for the cells of square cross section (a/b = 1), while for the rectangular cells, the reduction is only 14%.

From the data presented in Table 1, it should be readily apparent that if commercially-available rectangular lines are operated at or near the upper frequency limit recommended by the manufacturer, then at least two higher order modes (TE_{01} and TE_{10}) may propagate in the line. Note that the NBS transmission cells will exhibit only one higher order mode (TE_{01}) when operated near their upper limit.

TABLE 1
COMPARISON OF PREDICTED CUT-OFF FREQUENCIES IN MHz FOR COMMERCIALLY-AVAILABLE EQUIPMENT

Manufacturer and Model No.	Upper limit recommended by manufacturer	Predicted from $\lambda c/2 = b$ criteria	Predicted from $\lambda c/2 = a$ criteria	TE_{01} mode, corrected upper limit
IFI CC-101	100	125.4	82.9	66.3
IFI CC-101.5	150	187.8	125.1	100.2
IFI CC-102	200	250.0	166.8	133.3
IFI CC-103	300	382.5	249.2	204.4
IFI CC-105	500	513.3	333.5	272.4
IFI CC-110	1000	1010.6	663.5	530.5
Narda 8802	250	250.0	150.0	128.2
Narda 8801	500	500.0	300.0	256.4

Table 2
COMPARISON OF PREDICTED CUT-OFF FREQUENCIES IN MHz FOR NBS TEM TRANSMISSION CELLS

TE_{10} Mode predicted by Crawford using $\lambda c/2 = a$ criteria	TE_{01} Mode Corrected Upper Limit	
	Square Cell a/b = 1	Rectangular Cell a/b = 1.667
100	58	86
300	174	258
500	290	430

EXPERIMENTAL MEASUREMENTS

Experimental detection of higher order modes in TEM mode transmission lines is not as straightforward as might first appear. First, it is known that higher order modes will not necessarily propagate in well-matched co-axial or strip transmission lines unless deliberately excited by means of some impedance discontinuity or irregularity[5]. In the case of the large-scale rectangular strip-lines being considered in this article it is possible that discontinuities which may exist at the co-axial connector-to-strip-line transitions and at the tapered section-to-straight section junction are sufficient to actually excite higher order modes. They will almost certainly be excited by the presence of any form of load within the line.

Several workers[5,6] have shown that higher order modes can be detected using swept insertion loss techniques. The presence of higher order modes can excite spurious resonances at certain frequencies which appear on a swept transmission loss plot as a series of very sharp lines with attenuation values in the range of 3-8db. The exact mechanism by which these spurious resonances are created does not appear to have been studied as yet. It is in principle also possible to detect these resonances by means of swept VSWR measurements[5]. However the VSWR changes are small and not as easily detectable as the transmission loss changes. In attempting to make these measurements, care must be taken to ensure that what is being observed is not caused by the transmission line behaving like a low-Q cavity due to possible mismatches at the co-axial connector transitions. This can be checked by noting whether the line is any multiple of a half-wavelength long at the frequencies at which the resonant lines are observed.

The spurious resonances created by higher order modes do reveal the presence of such modes at certain discrete frequencies. However, it is unclear whether such modes continue to exist at adjacent frequencies and, more importantly, at what frequency they will cease to propagate. Even though not detectable by transmission loss measurements at these frequencies, theory tells us that these modes are still capable of propagating and therefore should presumably exist. Another method for experimentally detecting the presence of higher order modes, which has the potential for greater sensitivity, involves actual probing of the internal field structure of the line. Since the H modes (TE) are characterized by a magnetic field component which acts in the direction of wave prop-

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agation, it may be possible to detect their presence by means of a small loop probe, suitably aligned in the transverse plane of the line. Such a probe needs to be carefully designed and constructed so that it will respond only to magnetic fields and not to the electric fields that are also present; the use of a Moebius loop probe for measuring magnetic fields has recently been discussed by Iskander et. al.[7]. Provided that the loop probe has sufficient sensitivity, it should be possible to experimentally determine, using an RF spectrum analyzer, where the first higher order mode begins to propagate. We plan to attempt such measurements in the near future.

Some internal field measurements recently made on a 50 ohm rectangular strip-line, built at our EPA laboratory for purposes of irradiating animals, appears to definitely show the existence of the first higher order mode. For this facility, the TE_{01} mode cut-off, as predicted from the Gruner theory, is 275 MHz (a/b = 1.1), while that for the TE_{10} mode is about 450 MHz. The field mapping was performed at a frequency of 425 MHz, where, in addition to the fundamental TEM mode, a single higher order mode, the TE_{01} is predicted to occur. Actual field probing measurements were performed using the miniature nonperturbing field probe developed by the Bureau of Radiological Health[8] and the data were computer processed into a two-dimensional plot of power density distribution over the cross-sectional area of the line as illustrated in Figure 3. An examination of this plot reveals a serious asymmetry in power density values as the probe travelled across the line's cross section in a direction parallel to the center conductor (or septum). Reference to published mode charts for rectangular co-axial structures[9] shows that the TE_{01} mode is typically characterized by an electric flux that is greatest at the edge of the center conductor, progressively drops to zero at the midpoint, then reverses itself in mirror image across the rest of the center plate.

Our asymmetric field plots appear to represent the scalar summation of both the TEM fundamental mode and TE_{01} mode fields, thereby confirming the existence of the first higher order mode.

The highly non-uniform nature of the field plot shown in Figure 3

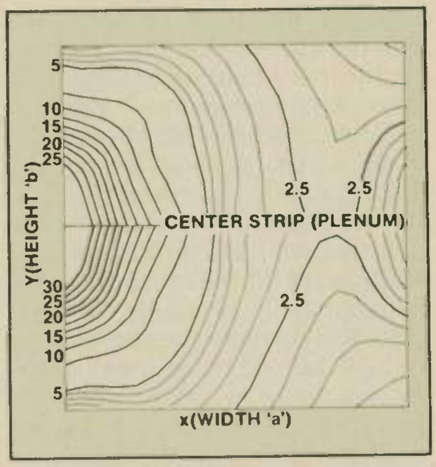

Fig. 3 Cross-sectional distribution of measured power density in EPA 50 ohm rectangular strip line; a = 33.5 cms, b = 30.5 cms, w = 28 cms.

serves to emphasize that it is generally inadvisable to operate such lines when this higher order mode is present, particularly where the objects under exposure occupy a significant part of the line's cross sectional area. However, it may be acceptable to extend the calibration of a small field probe in a well-matched line beyond the TE_{01} cut-off provided one is reasonably sure that higher order modes are not being generated.

METHODS OF EXTENDING FREQUENCY RANGE

Extending the frequency range of rectangular strip-lines basically involves selecting a design such that the TE_{01} mode cannot propagate at frequencies below the TE_{10} mode cut-off. In general, this can be accomplished by reducing the value of the parameter w/a; this in turn raises the characteristic impedance of the line. This process can be accomplished with the aid of Figure 2. As pointed out already, there exists a minimum aspect ratio a/b of 1.93 for the 50 ohm line if the above criteria is to be met. Lowering the aspect ratio requires a progressive increase in the characteristic impedance in order to realize full bandwith capability. For example, the impedance needs to be increased to about 65 ohms for a line with a/b = 1.5. However, it is evident from Figure 2 that a limit to this process is reached as a/b approaches 1 since the TE_{01} mode is always the first higher order mode in the square (a/b = 1.0) structure for any non-zero value of w/a.

Some further improvement in frequency range can be realized by incorporating corner ridges in the outer rectangular shield of this structure. In a recent paper, Gruner[4] has shown that, in general, frequency range can be improved some 10-15% by using symmetrical corner ridges with an optimal ridge aspect ratio of approximately 0.25. This technique appears to raise the cut-off of the TE_{10} mode; that of the TE_{01} mode is usually, but not always, raised.

As usual, there is a practical price which must be paid in exchange for the increased frequency range. Using higher impedance lines involves center conductors of narrower width. This, in turn, will create greater nonuniformity of the internal field structure in the line. The presence of corner ridges will tend to further compound this problem. These effects will be particularly pronounced for structures having aspect ratios that are close to square. Crawford[1] has pointed out that for his 50 ohm lines, the field uniformity is poorer in the square structure than that for the rectangular one (a/b = 1.667). This uniformity will be considerably worsened in the square structure if attempts are made to extend its frequency range by raising line impedance. Whether or not such techniques can be used to extend the frequency range of these devices depends on the nature of the application. Due to excessive nonuniformity of the internal fields, large objects which occupy a substantial proportion of the cross sectional area of the line cannot be exposed to satisfactorily defined fields. This, therefore, eliminates such applications as equipment susceptibility testing and animal exposures. However, if the object is small, such as a field probe undergoing calibration, the field nonuniformity is not a problem and for this

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[From page 97] **STRIP LINES**

application, such techniques are usable.

It should be apparent that the frequency limitations of the rectangular strip line are entirely due to the presence of the vertical side walls which cause higher order modes to propagate at frequencies lower than is possible if these walls are absent. This, therefore, suggests that maximizing the operating frequency range can best be achieved by removing the vertical side walls; i.e., to revert to parallel-plane strip line or large scale microstrip type structures having open side walls. As discussed earlier, the first higher order mode cut-off in these structures is governed by the simple criteria $\lambda_c/2 = b$ where "b" is the ground plane separation.

Large scale microstrip lines which consist of a single strip conductor mounted over a ground plane have been used by Roseberry and Schulz[10] for equipment susceptibility testing and by D'Andrea, Ghandi, and Kessner [11] and Sanders and Joines[12] for purposes of whole animal irradiation. The only problem associated with this form of transmission line is that the fields are not entirely enclosed and there will be some undesirable leakage of RF energy from the system. (The degree of leakage depends largely on the width of the ground plane employed.) If such a system is excited under high power conditions, the resulting leakage fields may pose a hazard to laboratory personnel as well as susceptible equipment nearby. If these problems do occur, such a system needs to be enclosed in a shielded room or anechoic chamber.

Most of the leakage problems associated with the large scale microstrip line are largely eliminated by placing a second ground plane over the strip conductor; i.e., the parallel plane strip line. In this structure, the RF fields are almost entirely confined to the region between ground planes. A very small fraction of this energy may leak out of the open sides of the structure, the amount again depending on the width of the ground planes employed. Hence, there appears to be a definite advantage in using this kind of configuration over the rectangular strip line configuration. Besides the increased frequency range, such lines also possess a more uniform internal field structure. These advantages must be weighed against the small leakage problem associated with parallel plate lines as well as the increased physical size created by the need for wide ground planes.

Editor's Note: In light of the data presented in Table 1, the supplier's of the equipment involved were given an opportunity to comment and their responses follow.

Relative to the possibility of modeing in the TEM cells; I have mapped the fields in Narda's 8801 TEM cell using the BRH model 15 probe. The probe is an isotropic, three axis probe with dipoles of 1.5mm length, and fine line high impedance transmission line. The same lateral pattern exists over the septum at 500 MHz as exists at 200 MHz. The field is uniform and constant. At and below 500 MHz the frequency was varied and a search made for other than the TEM, none were found.

The modes which you monitored in the cell constructed at your laboratory may have been caused by some assymetry in the cells design and construction.

The Narda cells will not support other modes except under the condition where the field is shorted at a position near the edge of the septum. If the shorting stub is placed at the center of the septum, the modeing is not present or less pronounced. If the perturbing body placed in the cell is limited to the maximum recommended dimensions, indicated in the data sheets, of 5 cm high, 10 cm wide and 15 cm long for the Model 8801 cell, and proportionately higher for the 8802 cell, no modeing will occur.

Edward Aslan
Principal Research Engineer
The Narda Microwave Corp.

Commentary on "Frequency Range of Large-Scale TEM Mode Rectangular Strip Lines," by Weil, Joines and Kinn.

1. One initial factor should be made clear prior to any technical notes; this is in regard to the IFI CC-105 and CC-110 TEM Cells referenced in the article. It is mentioned by the authors that these two cells do not have the same "margins" as the other models in the product line. This is quite true, if the measurements were referenced to early models of this series. It was determined in January, 1980, that these two models were erroneously rated for 500 and 1000 MHz first resonance, due to an error in dimensional scaling. The two models in question were reassigned as CC-104 and CC-108, which correctly establishes the region of first resonance at 400 and 800 MHz respectively. A new CC-100 was then configured, with proper scaling, and is now part of the standard product line.

2. We are in agreement that the TE10 mode can set up in the IFI line of Crawford Cells; in fact it is such that this will occur 16.7% below our stated first resonance point. However, there is considerable difference between the potential existance of a particular mode to exist and the proba-

[Continued on page 100]

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Comb Spacing (MHz)	500	500	950	100
Freq Stability (ppm)	30	30	30	30
Power Output (dBm)	-40 min	+6 max -22 min	-20 min	0 max -40 min
Load VSWR (max)	1 5 1	1 5 1	1 5 1	1 5 1
In-band Spurious (dBc max)	>24	>30	>40	>60
Power Supply	+15 Vdc at 200mA max	+15Vdc at 200mA max	+15Vdc at 150mA max	+15Vdc at 300mA max
Size (inches nominal)	1 9 x 1 33 x 5	1 33 x 1 33 x 15	1 9 x 1 33 x 0 6	1 9 x 1 33 x 0 6

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FERRITES

OCTAVE AND BROAD BANDWIDTH CIRCULATORS AND ISOLATORS

Freq. Range (GHz)	MODEL # Circulators	MODEL # Isolators	Isolation dB Min.	Insertion Loss dB Max.	VSWR Max.
1 0 - 2 0	50A1021	60A1021	16	0 6	1 35
2 0 - 4 0	50A3001	60A3001	18	0 5	1 30
2 6 - 5 2	50A3011	60A3011	18	0 5	1 30
4 0 - 8 0	50A6001	60A6001	18	0 5	1 30
5 0 - 10 0	50A6071	60A6071	18	0 5	1 30
8 0 - 12 4	10B9201	20B9201	20	0 4	1 30
8 0 - 16 0	50A2001	60A2001	17	0 5	1 35
12 0 - 18 0	10B2201	20B2201	18	0 5	1 30
8 0 - 18 0	50A2051	60A2051	16	0 6	1 [illegible]

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OSCILLATORS

DIELECTRICALLY STABILIZED OSCILLATORS (DSO)

Model Number	Frequency Range (GHz)	Output Power (mw Minimum)	Size, inches nominal
6600 - 1300	3 65 - 4 05	30	2 7 x 1 7 x 1
6600 - 1610	4 25 - 4 75	30	2 6 x 1 5 x 9
6600 - 1611	4 75 - 5 25	30	2 5 x 1 5 x 9
6600 - 1612	5 10 - 5 60	30	2 5 x 1 5 x 9
6600 - 1613	5 40 - 5 90	30	2 5 x 1 5 x 9
6600 - 1614	5 60 - 6 10	30	2 5 x 1 5 x 9
6600 - 1615	5 70 - 6 20	30	2 5 x 1 5 x 9
6600 - 1616	5 90 - 6 40	30	2 5 x 1 5 x 9
6600 - 1617	6 40 - 6 90	30	2 5 x 1 5 x 9
6600 - 1618	7 00 - 7 45	15	2 2 x 1 3 x 9
6600 - 1910	8 00 - 8 40	15	2 2 x 1 3 x 9
6600 - 1911	8 60 - 9 00	15	2 2 x 1 3 x 9
6600 - 1912	9 00 - 9 50	15	2 2 x 1 3 x 9

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

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[From page 98] **STRIP LINES**

bility of that mode sustaining and propagating within the cell. The TE01 mode is one that cannot be easily sustained, even with great effort to assist and encourage its existence. We have explored this possibility in detail, using up to 4 E-field sensors positioned within the center section of the cell.

3. Our test results do substantiate that the TE10 and the TE01 (if it can be set up) modes cannot propagate, due to the tapered end sections which cannot support higher order modes. In a similar fashion, the relatively small waveguide aperture at each end severely discourages any tendency to moding at the connector area.

4. Discounting any propagation of higher order modes, only the self-resonant effects remain as a potential source of difficulty; the extremely high-Q situation will create a tight region of instability at the resonant point. For example, in the CC-101.5, the resonance occurs at 125 and 150 MHz in an unloaded cell; object loading will tend to reduce these values.

5. We would emphasize the results obtained utilizing E-field sensors in evaluating the field conditions within the cell. Multiple sensors are valuable in determining the presence of another mode by detecting the alteration of an existing field distribution. In each test, cross-polarized components were not found to be present; testing with four sensors clearly would have shown these effects, had they been present.

6. We endorse the fact that, although cavity resonance does appear in all TEM cells, it is still reasonable and valid to use the cell at points between the resonant frequencies. Additionally, we encourage the continuous monitoring of the E-field within the cell, using a mounted and calibrated E-field sensor, to facilitate the detection of any deviations from the expected field values.

7. We feel there are distinct disadvantages in the use of a parallel plate line recommended by the authors. Although we agree that the plate system may have an advantage in its upper first-order moding, there are far more negative aspects that would lead one to choose the TEM cell as a more practical and efficient alternative. Just a few are outlined below:

- Of primary concern is the authors' apparent disregard for the consequences of discontinuities in the parallel plate lines, which are far more serious than those found in TEM cells.
- The plate line presents a more serious effect on both the user and the electromagnetic environment.
- Harmonics from the RF power source must be controlled and eliminated for proper use of the plate; E-field enhancement at resonance can be 20 dB or more, dependant on circuit Q.
- The parallel plate line will tend to radiate when stimulated by mismatch loading or resonance.

8. In summary, we feel that, at a given line impedance, the TEM cell has a larger test area for the required lab space; that the TEM cell does not require a shielded enclosure for safe operation and provides a valid and accurate vehicle for producing known E-field levels.

Instruments for Industry, Inc.

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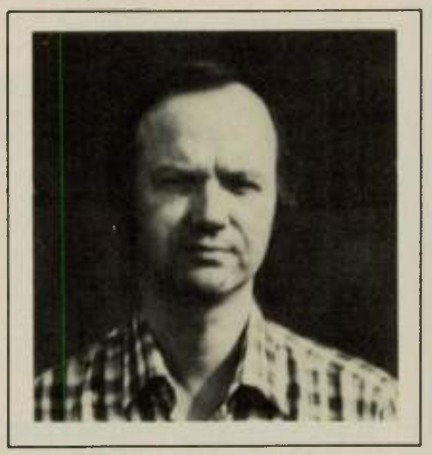
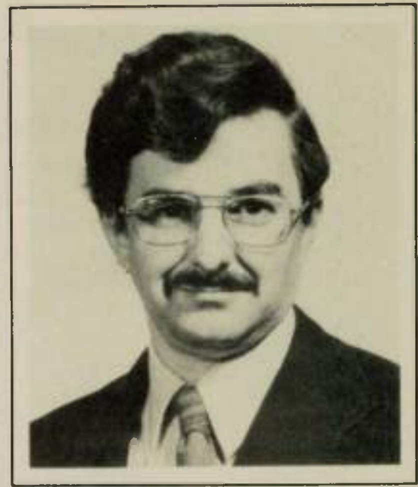
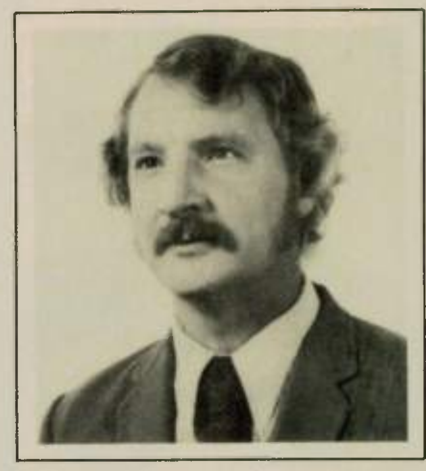


James B. Kinn, *received the B.S. degree in Physics from the University of Detroit in 1960. From 1961 to 1971 he was employed by the Federal Aviation Agency developing physiological monitoring instrumentation. In 1970 he joined the U. S. Environmental Protection Agency and has been conducting research in the dosimetry of non-ionizing radiation. Mr. Kinn is a member of the IEEE and the Bioelectromagnetics Society.*



Claude M. Weil, *received the B.S. degree in 1959 from the University of Birmingham, England, the M. S. E. degree in 1963 from George Washington University and Ph. D. in 1970 from the University of Pennsylvania, Philadelphia [all in Electrical Engineering].*

He has been employed in the past as a Navy systems and instrumentation engineer and has designed microwave components and antennas. He has also been an instructor in Electronics. He joined the Environmental Protection Agency, Office of Research and Development, in 1971 and is currently engaged in research activities associated with EPA's program of nonionizing radiation health effects



William T. Joines, *received the B.S.E.E. degree with high honors from North Carolina State University, Raleigh, in 1959, and the M.S. and Ph. D. degrees in electrical engineering from Duke University, Durham, N. C., in 1961 and 1964, respectively.*

From 1959 to 1966, he was a member of the Technical Staff at Bell Telephone Laboratories, Winston-Salem, N. C., where he was engaged in research and development of microwave components and systems for military applications. He joined the faculty of Duke University in 1966, and is currently a Professor of Electrical Engineering, as well as working for the U. S. Environmental Protection Agency. His research and teaching interests are in the area of electromagnetic wave interactions with materials. He has published numerous papers in this area. ■

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Primarily concerned with the problem of measurements in the frequency domain with spectrum analyzers, **Spectrum Analyzer Theory and Application** by Morris Engelson and Fred Telewski, follows a dual approach. Chapters one through five have a mathematical, process-oriented technique; the latter part of the volume applies the former theory to specific measurement problems. The authors present a unified, mathematical, and philosophical rationale for the use of spectrum analyzers, making this book an excellent text and reference for classroom use. Also included are sample problems and an extensive bibliography.

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The Dielectric Resonator Oscillator- A New Class of Microwave Signal Source

MERT PURNELL
Frequency Sources,
West Div.,
Santa Clara, CA

INTRODUCTION

The availability of highly stable, low loss dielectric material has led to the development of this new class of microwave signal source, the dielectric resonator oscillator. The generation of stable, narrow-band microwave signals has typically been performed by cavity-stabilized Gunn oscillators or crystal-multiplier chains. However, with the need for smaller, more efficient, more reliable sources, the DRO is finding increasing application in such areas as telecommunications systems, radar beacons, ECM receivers, missile transponders, BITE, and weather radar.

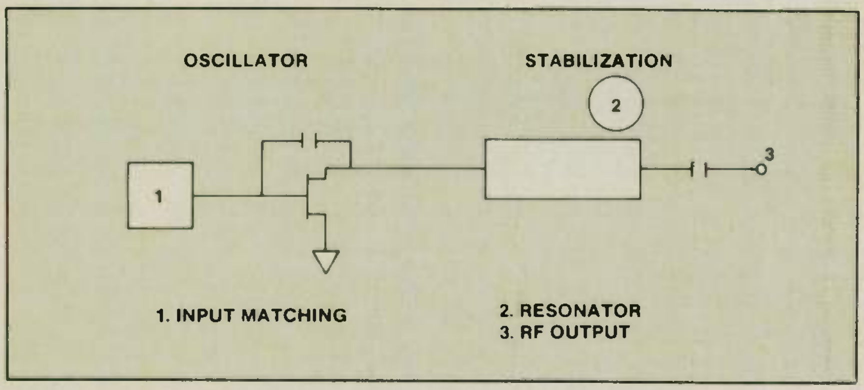

Fig. 1 Negative resistance oscillator.

The dielectric resonator can be integrated with a variety of active devices to cover the microwave frequency spectrum. GaAs FET's are the most commonly used device due to their low cost and high f_t. Presently, FET DRO's to 14 GHz are available in production quantities and prototype designs are being evaluated to 19 GHz. Gunn DRO's to 50 GHz and bipolar DRO's as low as 1 GHz are known. The current state-of-the-art performance for both bipolar and FET DRO's in the 1-19 GHz range are discussed.

DIELECTRIC MATERIAL AS THE RESONATOR

Although the idea of using dielectric material as the resonator in a microwave oscillator has existed for many years, it is only recently that suitable material has become commercially available. Early dielectric material exhibited poor resonant frequency and temperature stability. However, the high Q dielectric material now available has a dielectric constant of approximately 38 and a resonant frequency temperature coefficient of 0 ppm/°C ± 1 ppm/°C. The high frequency stability is achieved by compensating thermal expansion of the resonator with a corresponding change in the dielectric constant temperature coefficient, as given in Equation 1[1].

$$\tau_f = -\left(\frac{\tau_K}{2} + \alpha_L\right) \quad (1)$$

where τ_f is the resonant frequency temperature coefficient, τ_K is the dielectric constant temperature coefficient, and α_L is the coefficient of thermal expansion.

DRO CIRCUIT CONFIGURATIONS

The two most commonly used circuit configurations for the DRO are the negative resistance oscillator and the feedback oscillator. In the negative resistance oscillator, Figure 1; the DRO is comprised of the FET oscillator section, and the dielectric stabilization section. The oscillator is designed with drain-to-gate feedback and gate matching, though other oscillator configurations may be used. The signal, coupled out of the drain, is stabilized by the dielectric resonator mounted $\lambda/2$ away. Comparing performance of the stabilized and unstabilized oscillator, the stabilized oscillator shows typically a 20 dB reduction in FM noise and a factor of 20 improvement in frequency stability.

In the feedback oscillator, Figure 2, a FET amplifier is initially designed to operate at the desired oscillator frequency. By coupling back part of the output power to the amplifier input via the dielectric resonator, the amplifier oscillates at the resonator frequency. The performance of the feedback oscillator is comparable to the negative resistance oscillator.

[Continued on page 104]

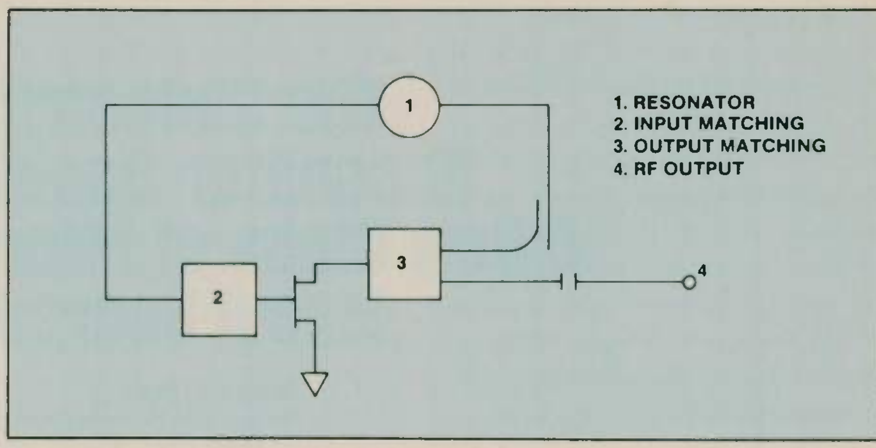

Fig. 2 Feedback oscillator.

Frequency/Power

Both the frequency and power limitations of the DRO are determined primarily by the active device and the circuit configuration, as the low loss tangent of the dielectric material allows high power and high frequency operation. Practical GaAs FET designs have been demonstrated up to 19 GHz, with operation at higher frequencies feasible. The lower frequency limit is determined by size constraints; at frequencies below 1 GHz the resonator becomes quite large, hence, other oscillator technologies become more attractive.

For LO applications low power FET's with 5-20 mW of output power are suitable and efficiencies are typically 15-20%. The DRO approach is also suitable for high power applications. Over ½W at 9 GHz with 20% efficiency has been achieved.

Bandwidth

The bandwidth of the DRO is determined by how far the resonator frequency can be tuned by raising or lowering the ground plane/tuner over the resonator. Bandwidths of up to 10% can be achieved. Figure 3 shows the output power vs. frequency of a 12 GHz DRO, with a tuning bandwidth of 6.5%.

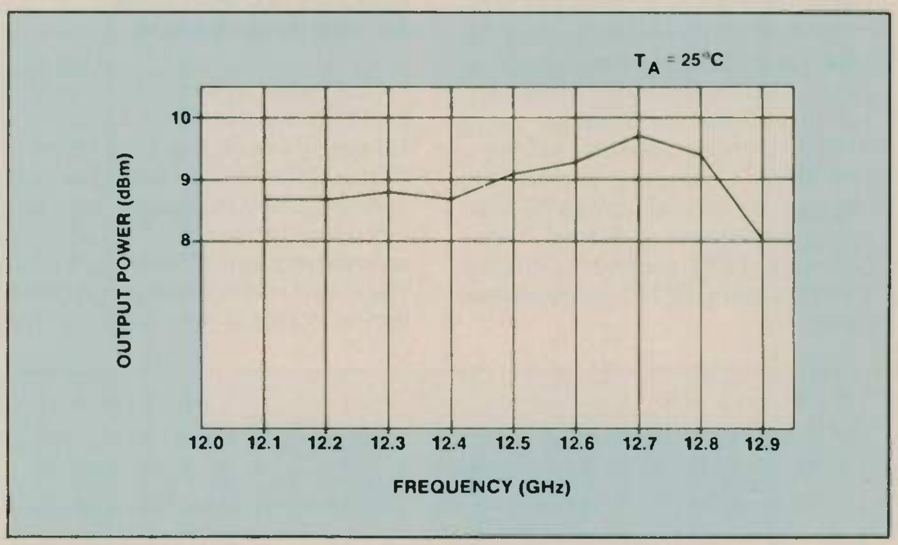

Fig. 3 Typical DRO output power as a function of frequency.

Frequency Stability

Measured stabilities of DRO's using temperature stable dielectric material are typically 5 ppm/° C or less. Figure 4 shows the frequency stability of a 5.8 - 6.3 GHz DRO operated from -50° C to +100° C. The worst case temperature stability averaged over the

[*Continued on page 106*]

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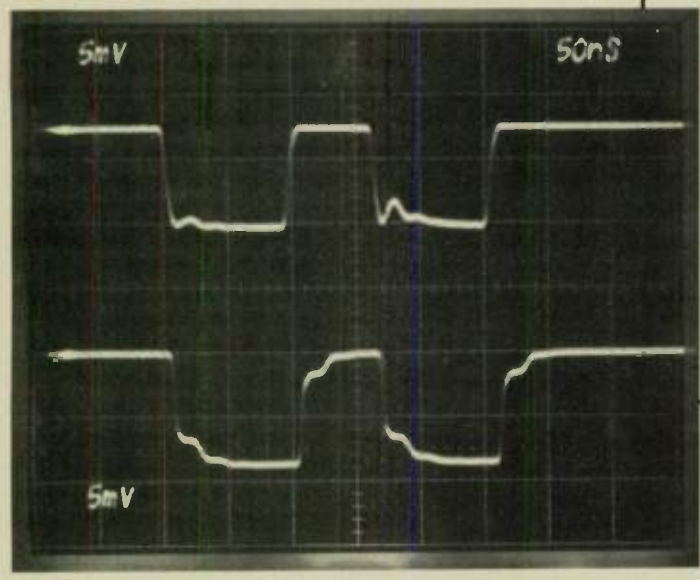

Data for Aertech 7.5 to 18 GHz Limiting Amplifier at 16 GHz...

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Input: -5 dBm

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100 nsec pulses:

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No need for a lot of fancy words. When it comes to performance, Aertech Limiting Amplifiers practically speak for themselves. Our amplifiers feature the exceptional pulse fidelity and excellent response in high pulse density environments which is essential to EW system requirements. And proprietary Aertech circuitry enhances the small signal suppression and saturated output power flatness. The difference between 1 dB compression and hard saturation is typically less than 3 dB.

Aertech Limiting Amplifiers are available in discrete frequency ranges from 2 to 18 GHz. Performance similar to that shown above is available in our 2 to 6 GHz band model, as well as in the standard operating band units.

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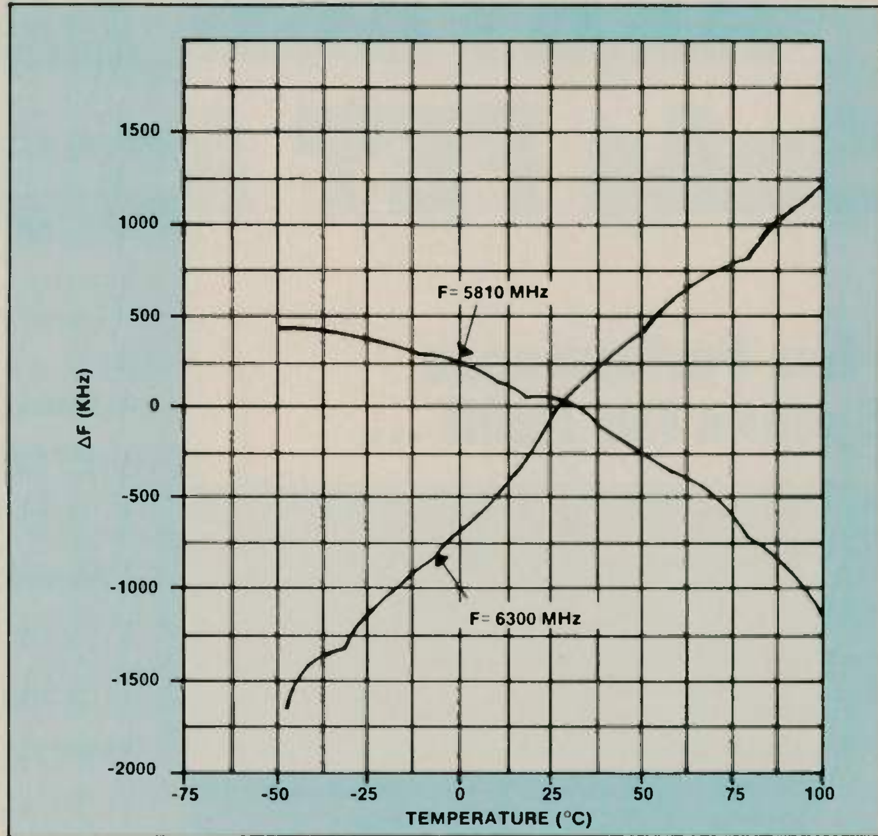


Fig. 4 Frequency stability vs. temperature for a 5.8 - 6.3 GHz DRO.

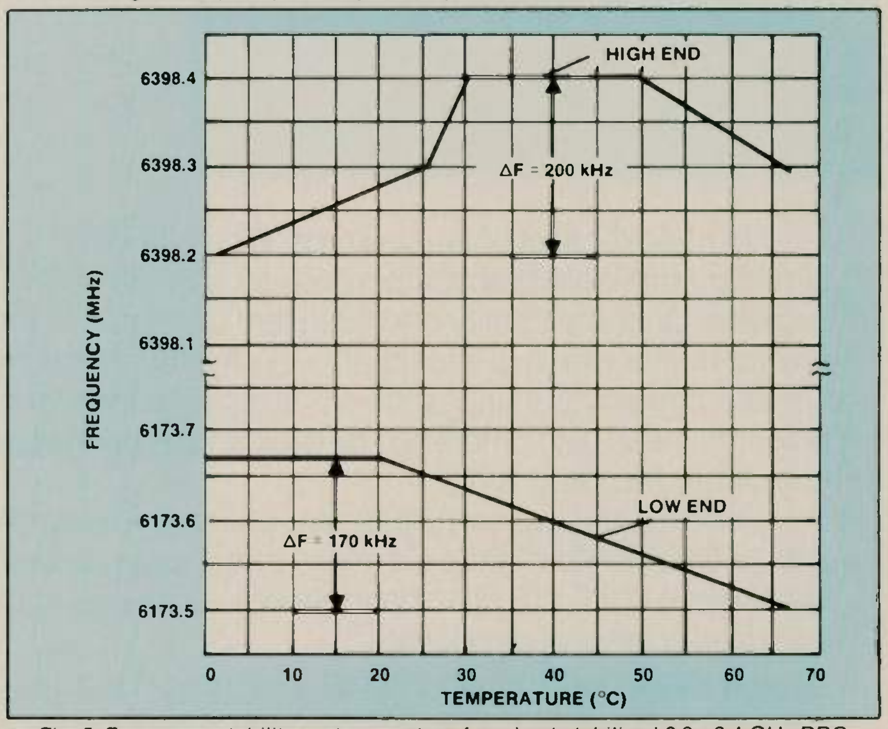


Fig. 5 Frequency stability vs. temperature for a heat stabilized 6.2 - 6.4 GHz DRO.

temperature range was 3 ppm/°C.

Where better frequency stabilities are required, for example in telecommunications systems, there are at least three approaches: heating of the oscillator, phase-locking, or injection locking. The use of a heater is feasible because the small size of the oscillator requires only several watts of heater power to maintain the DRO at an elevated temperature. Figure 5 shows the frequency stabil-

[Continued on page 108]

PLENICOM 82

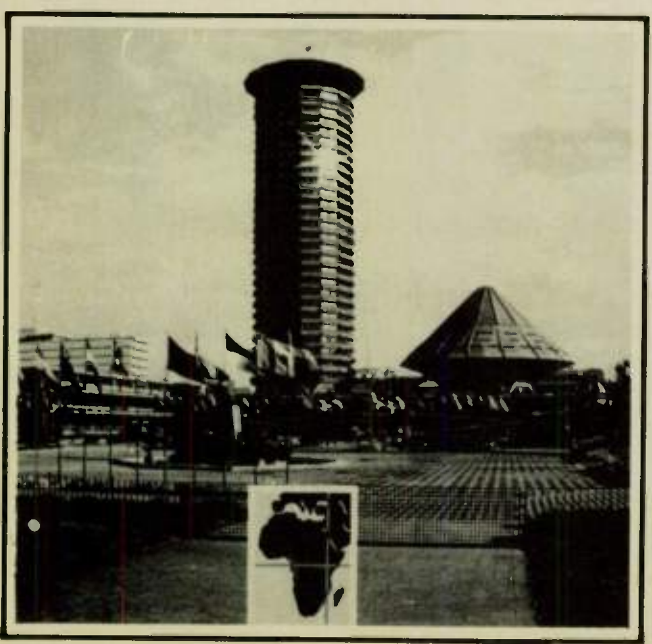
NAIROBI, KENYA — OCTOBER 1982
AN INTERNATIONAL TELECOMMUNICATIONS EXHIBITION

Representatives from 155 Countries

Horizon House Expositions, Inc. announces PLENICOM, an international telecommunications exhibition to be held concurrent with the ITU Plenipotentiary Conference in Nairobi, Kenya, October 1982. The merging of the developed and developing worlds for telecommunications becomes possible at the time of PLENICOM. The market for products is best identified with the needs and programs of all 155 member countries attending the Plenipotentiary Conference. There is no other occasion in the decade of the 80's that represents the prestigious assemblage meeting in Nairobi, Kenya, in October of 1982 for the Plenipotentiary Conference. It is inconceivable that any company providing a product or service can afford to miss this priceless opportunity. The Government of Kenya has recognized the great value of this meeting by establishing the agreement and authorizing the establishment of an exhibition for this occasion.

The ITU Plenipotentiary Conference is an assembly of senior officials - Ministers, Vice Ministers, Directors General, Chief Planners and Systems Engineers - those individuals directly responsible for purchases of equipment used in the national networks.

As the UN agency that coordinates the planning and operation of the world's national and international networks, the ITU convenes its Plenipotentiary Conferences at irregular intervals - the last being in 1973. Thus, it is almost a decade since there has been a comparable gathering of top telecommunications officials. Their stature and competence emphasize the importance of decisions made at a Plenipotentiary - major economic and political policy issues as well as a range of technical, operating and equipment considerations affecting radio, telephone, telex, satellite, digital switching and transmission, and data.



Front View of Kenyatta Conference Center and Plaza

At present, the plans are to open the Plenipotentiary Conference on September 28, 1982. The exhibition, while presently scheduled for October 11, reserves the right to move the schedule forward by one week to open on October 4 if the plans and programs of the Conference change. The exhibition is to start at the time nearest the conclusion of the opening ceremonies, at which time the key political figures involved with this exhibition are able to schedule themselves to participate in this program. Adequate notification will be provided to all exhibitors and potential attendees to the exhibition of any changes should they occur.

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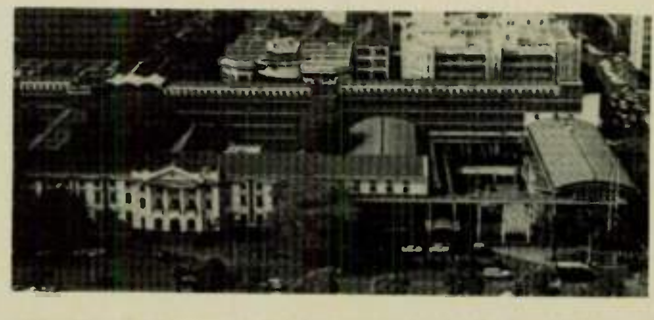

City Hall, directly across the street from Kenyatta Conference Center.

[From page 106] RESONATOR

TABLE I

DRO PERFORMANCE SUMMARY

Center Frequency	1-19 GHz
Bandwidth	Up to 10%
Power	1 mW - 500 mW
Freq. Stability w/temp	.2 - 5 ppm/°C
FM Noise	-95dBc/Hz @ 10 kHz from carrier for 6 GHz DRO
Power Supplies	± 12 VDC + or - 12 VDC ± 15 VDC + or - 15 VDC
Environment	-54°C to +100°C up to 50 G's, 50-2000 Hz
Size	< 2 in^3

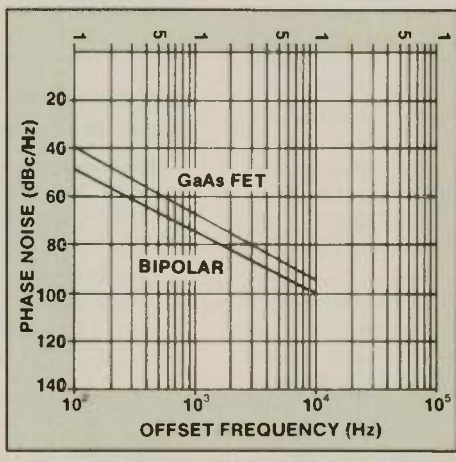

Fig. 6 SSB phase noise for 6 GHz bipolar and FET DRO's

ity of a heated 6 GHz DRO, where the stability over a 4% bandwidth was less than ± 20 ppm. Fixed frequency, heated DRO's have demonstrated ± 5 ppm stability over a 0°C to 65° C ambient.

Figure 6 shows the measured SSB phase noise of both a bipolar and GaAs FET DRO at 6 GHz. At 12 GHz, the SSB phase noise is typically -90 dBc/Hz at 10 KHz from the carrier.

Mechanical Stability

The dielectric resonator is not seriously affected by high levels of shock and vibration. Tests of DRO designed for missile transponder applications at vibration levels of up to 50 G's at 50-2000Hz yielded less than 5kHz peak-to-peak FM.

Size

The high dielectric constant of the resonator minimizes its space requirements. At 10 GHz, a resonator is 0.2" in diameter x 0.1" high. Oscillator overall volumes vary from 0.5" inch^3 to 2.0 inch^3, depending on frequency and configuration. Table 1 is a summary of the typical characteristics of bipolar and FET DRO's.

Fig. 7 Typical DRO.

CONCLUSION

The DRO is appropriate to applications that require high stability, small size and high efficiency. They are presently being supplied in production quantities to both commercial and military systems.

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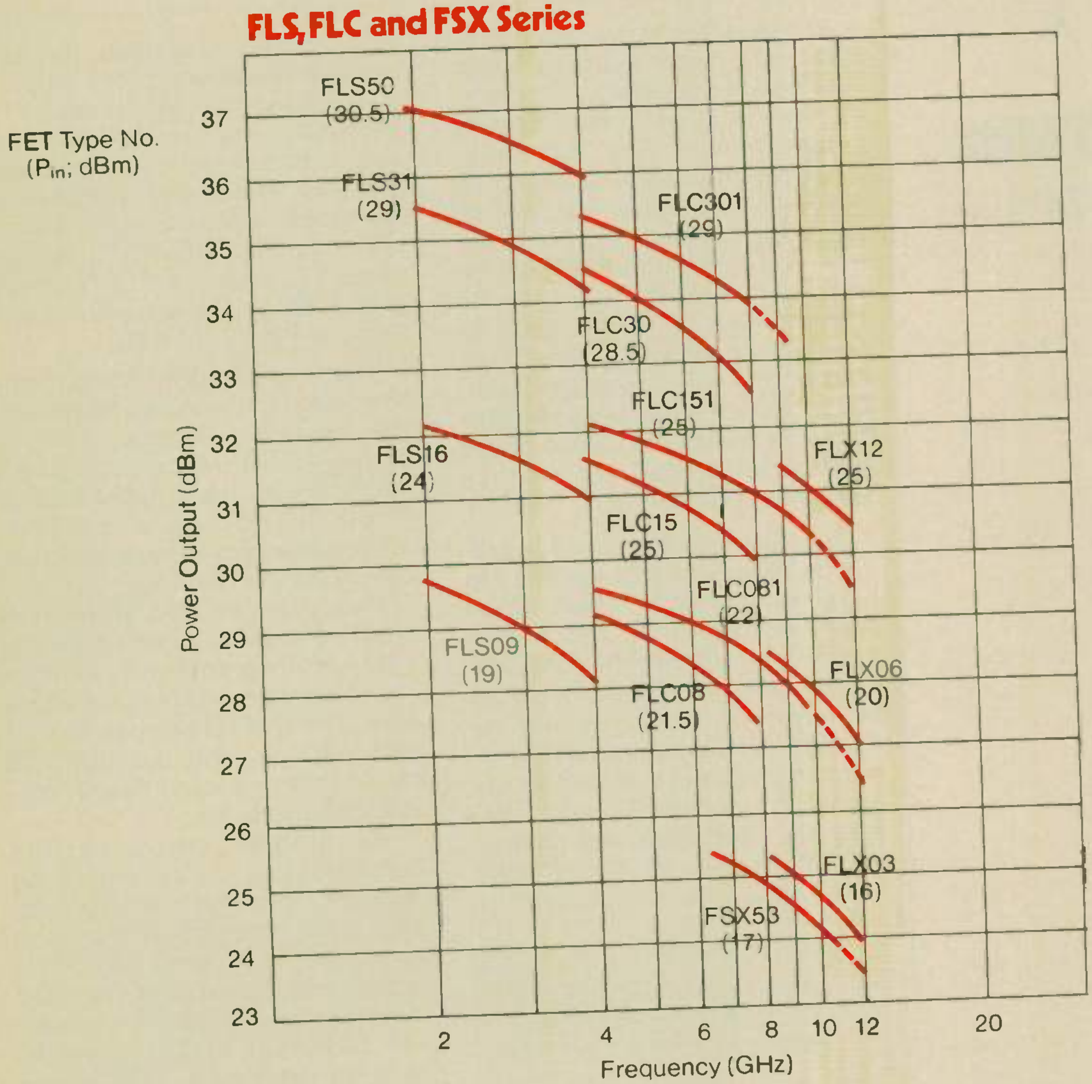

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Device Type	FLM3742-5	FLM4450-5	FLM5964-5	FLM6472-5	FLM7177-5	FLM7984-5
POUT (SAT) Min (+dBm)	36	36	36	36	36	35
Frequency Range GHz	3.7—4.2	4.4—5.0	5.9—6.4	6.4—7.2	7.1—7.7	7.9—8.4

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Program Separates Device Under Test From Fixture

MADE-IT ASSOCIATES
Burlington, MA

The Automatic Network Analyzer has many features which increase measurement accuracy. A new program called MAMA, (Measurement And Microwave Analysis) has been released by Made-It Associates and will be marketed by Compact Engineering. The program uses the conversion of reflection data into the time-domain to reveal each reflection as it appears in time. Eliminating the reflections produced by connectors, fixtures and adaptors and leaves the device under test de-embeded from these unwanted reflections.

To illustrate the program, a precision 10 cm line was modified to have two screws, .9" apart, which form two capacitive discontinuities. The test data for this device is shown in Figure 1, plotted as VSWR vs frequency. Converting the reflection coefficient into the time-domain produces the time display shown in Figure 2a. This time display has been evaluated using a special function of the program called RLC which interprets the time-domain and gives the approximate capacitance, inductance, resistance, and location in the time-domain. The two screws used in this fixture were evaluated and their values presented below the graph by the program. A simple command, REMOVE, can be used to "erase" the reflections produced by the precision connectors. (See Figure 2b).

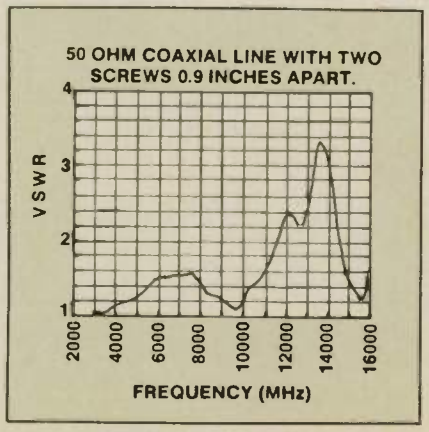


Fig. 1 Measured VSWR vs. frequency.

The resulting time-domain can then be deconvoluted back into the frequency-domain. The resulting VSWR is shown in Figure 3 overlaid with the original data of Figure 1. This illustrates the improved accuracy of measuring the device by this method.

This process also eliminates small reflection errors which occur near the first precision bead helping to further improve the accuracy. The two screws can further be separated using the REMOVE feature leaving each screw individually interpreted.

The program also allows the transmission coefficient to be changed into the time-domain. This gives a display of distortion vs time. A 14 in. long microstrip circuit was tested and changed into the time-domain to illustrate the power of the program to separate two main TEM waves from the interfering waves possible with microstrip. The transmission coefficient has increasing loss vs frequency. The time display for this transmission coefficient is shown in Figure 4. The energy transmitted by the TEM wave causes the larger positive peak to occur in the time-domain.

To separate this time display into its component parts a loss-

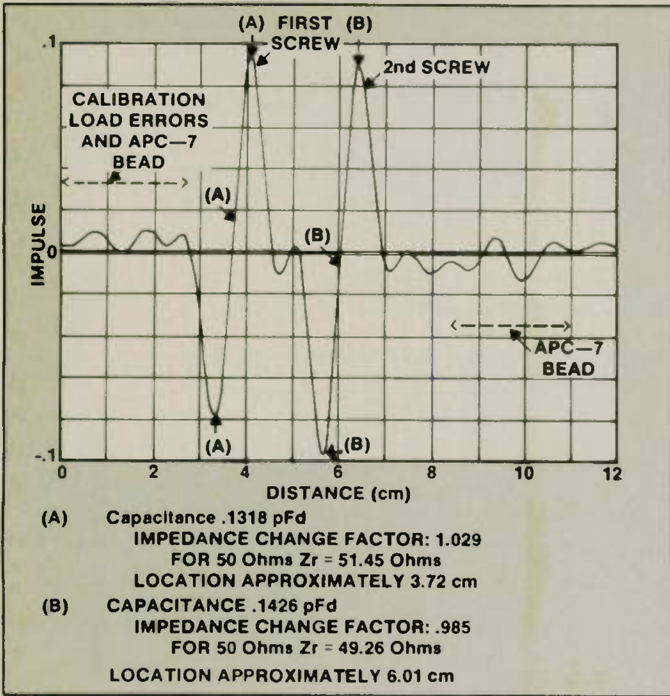


Fig. 2a Impulse time-domain display showing both screws 0.9" apart with APC-7 connectors.

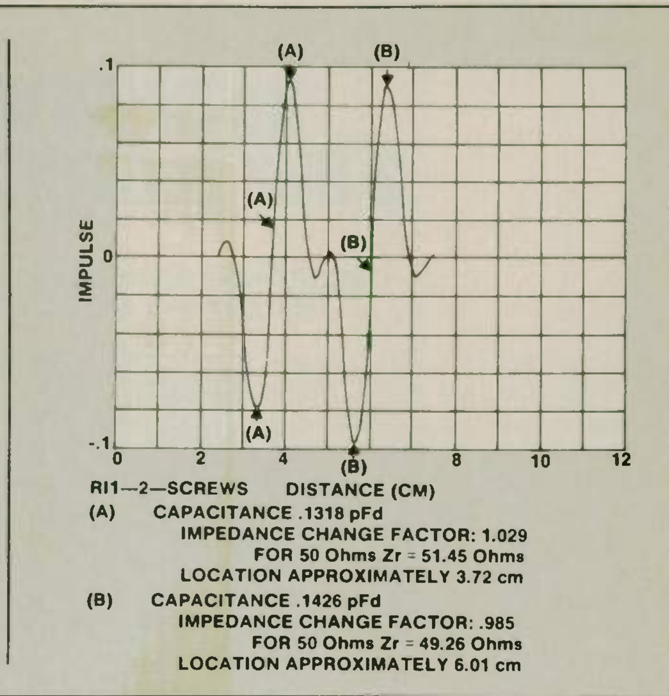


Fig. 2b Impulse time-display after using REMOVE command: REM 20, 2.5 and REM 27.5, 12.

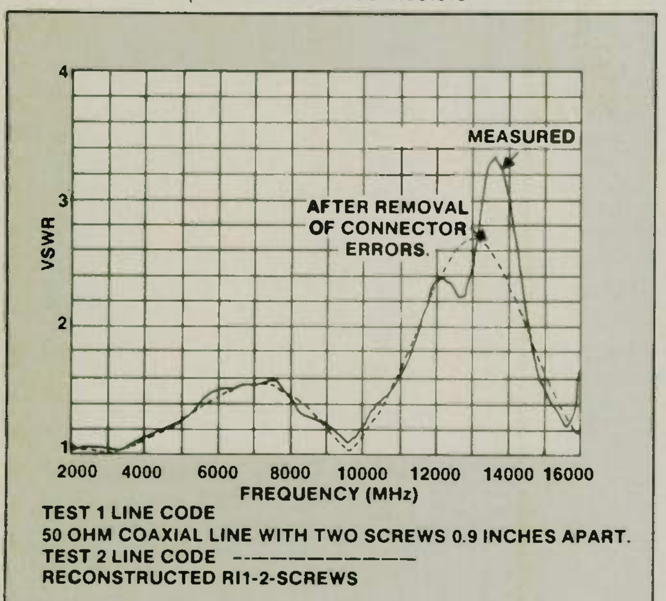


Fig. 3 Measured and reconstructed VSWR vs. frequency.

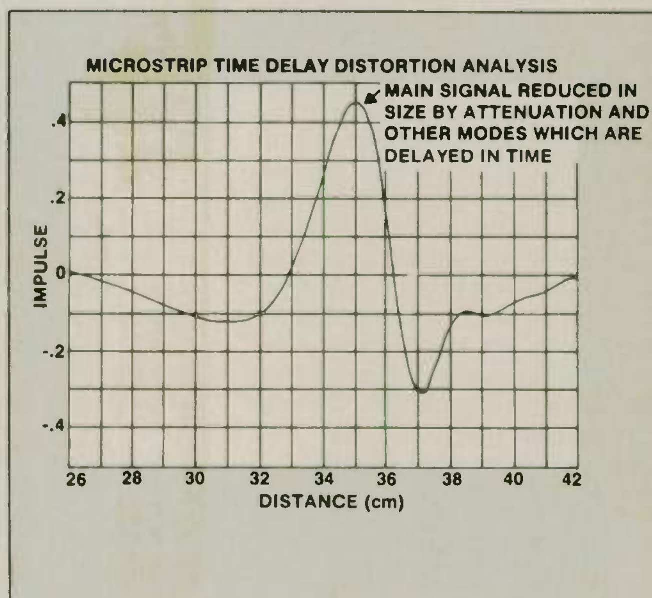


Fig. 4 Microstrip transmission time distortion analysis (TDD) using impulse function and MAMA program.

less transmission coefficient of the same length was converted to the time-domain and overlaid, resulting in the time display of Figure 5. Since the main signal had been attenuated by the microstrip, whereas the comparing signal was loss-less, the main signal at the center is now reduced indicating the loss due to the microstrip transmission. The energy appearing before the TEM signal is responsible for some of its attenuation and now appears in the time display. The time of arrival

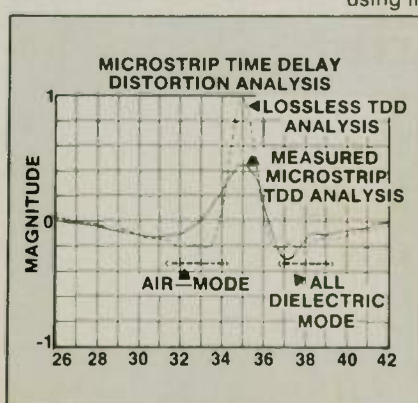


Fig. 5 Overlay of microstrip TDD analysis with lossless TEM delay.

corresponds to the velocity of free space for the circuit under test. A third signal appears later in time (beyond the TEM wave) which is characteristic of a transmission pattern which is magnetically coupled with a main path indicating that 1e mode of transfer had a common magnetic field with the TEM microstrip wave.

The program has other features of impedance, admittance, and relative phase or group delay. ■

Circle 132.

Ka BAND VCO WITH 2 GHz ELECTRONIC TUNING

MICROWAVE ASSOCIATES Ltd., *Dunstable, England*
MICROWAVE ASSOCIATES, *Burlington, MA*

The MA-87912 Series of electronically tunable Gunn diode oscillators provide electronic tuning bandwidths greater than 2 GHz across the 26.5 to 40 GHz band. Typical output power for these oscillators is 65 mW from -30°C to +70°C, (Figure 1) with power flatness over the tuning range better than ±1 dB. Electronic tuning linearity is better than 4.5:1.

A constant gamma GaAs hyperabrupt junction tuning varactor and a GaAs Gunn diode developed by Microwave Associates are combined in a waveguide cavity. This design has resulted in a millimeter wave VCO with a broad electronic tuning range, flat output power, excellent electronic tuning linearity and wide deviation (2 GHz) frequency modulated operation at rates up to 50 MHz.

These electronically tuned Gunn oscillators are ideally suited for use as FM transmitters and interferometers. The excellent linearity of better than 4.5:1 over the 2 GHz tuned bandwidth, coupled with low noise characteristics, make these sources also attractive for use as millimter local oscillators in AFC loops and phase lock systems. The linearity (Table 1) which is the ratio of maximum to minimum frequency sensitivity across the tuned bandwidth has been achieved without the need for internal linearizers.

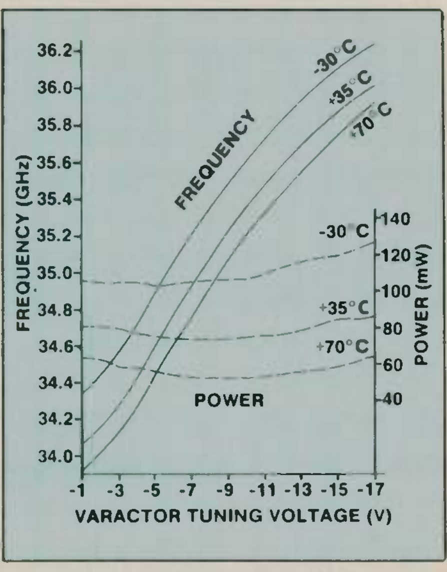


Fig. 1 Typical Characteristics

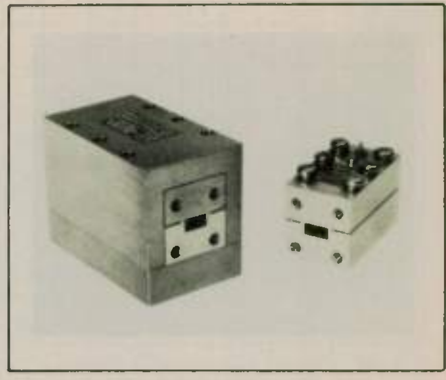


Fig. 3 Ka band VCOs

The standard unit is equipped with Gunn bias supply line filtering and transient protection. The oscillator requires +3.0 to +5.5 volts. With an optional internal regulator the required input voltage is +8.0 to +9.0 volts. The operating and threshold currents are 1.5 and 1.9 A typical. The output is WR-28 waveguide and mates with the UG-599/u flange.

TABLE 1

FREQUENCY TUNING LINEARITY AT FIXED TEMPERATURE

TUNING BANDWIDTH (GHz)	WORST CASE LINEARITY
<0.5	1.15:1
<0.75	1.30:1
<1.0	1.50:1
<1.25	1.80:1
<1.5	2.50:1
<2.0	4.50:1

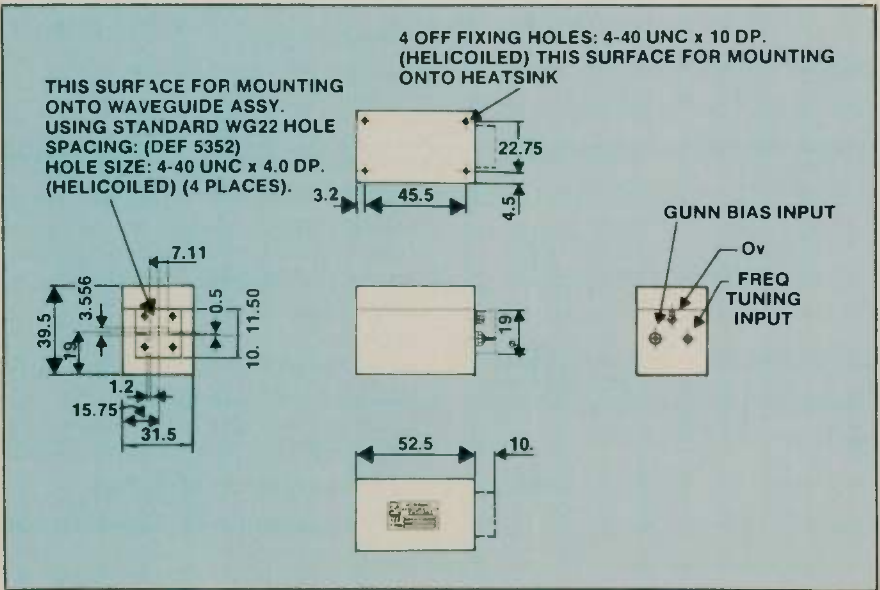


Fig. 2 Outline drawing (electronically tuned model)

Each unit is contained in a rugged housing, giving protection against shock and rough handling. An optional waveguide window can be provided for added environmental protection.■

Microwave Products

Instruments

RF CALORIMETER/LOAD

The Calorimetric MODULOAD line terminations have continuous power ratings of 10 kW or 25 kW and offer an SWR of less than 1.1 from 1 kHz to 900 MHz. A digital display indicates power in kW with an accuracy of ± 3% of indication and may be located several feet from the load. Models are available for 1 5/8" and 3 1/8" EIA lines, flanged or unflanged, and for operation from 115 V or 230 V, 60 or 50 Hz. **Bird Electronic Corp, Cleveland, OH. H. H. Heller (216) 248-1200.**

Circle 167.

ERROR RATE TEST SET

Model 5105 Error Rate Test Set is designed for systems operating at the DS1 transmission rate. It measures both bit errors and bipolar violations and can test automatic protection switches. Controls and connections are located on the front panel. The system can perform standard end-to-end tests on digital radio, fiberoptic and T-carrier systems and is fully compatible with all T1 bit-error-rate test sets. Price: $1,450. Delivery: 8 to 12 weeks ARO. **Tau-tron, Inc., Chelmsford, MA. Jim Hanley (617) 256-9013.**

Circle 166.

CONSTANT CURRENT SOURCE

The AT-SM37 Constant Current Source can deliver 5 ma into 50K ohms, 50 ma into 5K ohms or 500 ma into 500 ohms. A 3½ digit LED panel meter displays the output current which, once set, remains constant regardless of load impedence changes. Size: 11"W x 6½"H x 15"D. Weight: 24 lbs. Price: $975.00. Delivery: 30-60 days. **AD-TECH Microwave Inc., Scottsdale, AZ, G. A. Herlich (602) 998-1584.**

Circle 168.

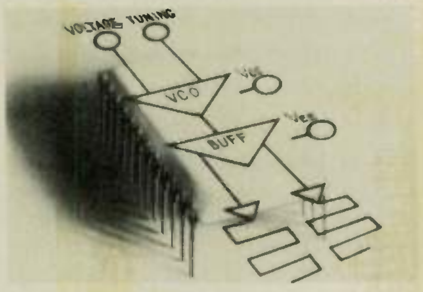
Device

LOW COST GaAs FET

The ALF 1020 series GaAs FET's feature an attractive price/performance ratio. ALF 1020 chips offer typical 1.6 dB noise figures and 11.5 dB associated gains at 4 GHz. Price: Under $10.00 for quantities of 200. **Alpha Industries Inc., Woburn, MA Nicholas A. Bishop (617) 935-5150.** **Circle 173**

Components

VOLTAGE CONTROLLED OSCILLATORS COVER THE 21 to 55 MHz RANGE



Series KJ1000 voltage controlled oscillators are intended for phase lock loop or clock applications and cover the frequency range 21-55 MHz. They feature dual buffered complementary outputs for 10K, 10 HK and 100 K levels and can be shifted for TTL. Floating tuning voltage inputs of 1-5 V produce a 20% frequency range, 1-20 V gives 50% bandwidth. Packaging is 24 pin DIP and operation is from a +5 or -5.2V single supply. Price: $14.35 (100-999). Delivery: Stock. **Frequency Sources, Semiconductor Div., Chelmsford, MA. (617) 256-8101.**

Circle 144.

DIODE ATTENUATER COVERS 960 to 1215 GHz BAND

Model AG-9423 digitally controlled PIN diode attenuator covers the 960 to 1215 GHz frequency range with a maximum insertion loss of 1.4 dB and a maximum attenuation of 80 dB. Accuracy at 25° C is ±0.7 dB to 40 dB and ±1.1 dB to 80 dB. Over the -55 to +85° C range, accuracies are ±1.0 dB to 40 dB and ±1.5 dB to 80 dB. DC voltage is ±15 V at ±50 mA, switching speed is 10μ sec. Maximum SWR is 1.5. Availability: 10 weeks ARO. **Triangle Microwave, Inc., East Hanover, NJ. Martin Rabinowitz (201) 884-1423.** **Circle 146.**

COAXIAL FEED THROUGH TERMINATIONS

Series FT coaxial feed through terminations are available in 50, 75 and 93 ohm impedance designs and with BNC, TNC, N and SMA male to female connectors. Model FT-50 operates from DC to 1 GHz, FT-75 from DC to 500 MHz and Model FT-90 from DC to 150 MHz. The units have a 1 W CW or 1kW peak power rating. Price: from $10.50. Delivery: stock to 30 days ARO. **Elcom Systems, Inc., Boca Raton, FL. Leonard Pollachek (305) 994-1774.**

Circle 145.

TUNABLE FILTER COVERS 4.4-5.0 GHz

Model DMT-4450-4T dual mode tunable filter covers the frequency range 4.4 to 5.0 GHz providing 4 pole bandpass response. The filter exhibits a maximum insertion loss of 0.7dB, with a 30 dB stopband of f_0±26MHz and an input SWR no greater than 1.5 and a 3dB bandwidth of 20±2 MHz. The unit is available with direct-reading tape, dial or uncalibrated knob. Price: $1775 (with tape read-out). Delivery: 45 days. **RS Microwave Company, Inc., Butler NJ. Richard Snyder. (201) 492-1207.**

Circle 136.

OSCILLATORS COVER 0.4 to 5.2 GHz

A series of phase locked oscillators covering the frequency range 0.4 to 5.2 GHz are fundamental frequency devices utilizing high Q stabilized oscillators. RF power output of standard units is +20 dbm between 0.4 and 2.0 GHz and +13 dbm between 2.0 and 5.2 GHz. Internal reference stability is ±0.003% over the operating temperature range of 0 to +50° C. Harmonics are specified at -20 dBc and spurious at -80 dBc. Power supply requirements are positive or negative 12 to 28 VDC at 150ma typical. Options include lock limit alarm, custom mounting, operation from -30 to +70° C and higher RF output power up to +26 dbm from 0.4 to 2.0 GHz and +16 dbm from 2.0 to 5.2 GHz. Overall size is 1.25" x 2.25" x 2.25" excluding projections. Delivery: 10 weeks. **RFD, Inc., Tampa, FL. (813)872-1502.**

Circle 134.

Microwave Products

1000 MHz CRAWFORD CELL

Model CC-110 Crawford Cell is suitable for lab and RF susceptibility testing in the dc to 1 GHz range. It provides a controlled environment for generating a homogeneous E-field of up to 1500 V per meter with an accuracy of ±1dB. To prevent hazardous radiation exposure a side door is provided for insertion of objects under test. Bulkhead transistion plates are located at each end of the unit for access of cables and connections to the test sample. The CC-110 measures 18" x 8" x 5" and can accommodate test objects of up to 15lbs. **Instruments for Industry, Farmingdale, NY. Ronald Richards (516) 694-1414.**

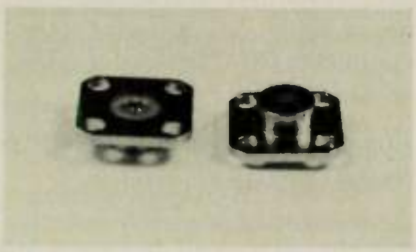
Circle 164.

5 W to 18 GHz TERMINATION

Model TO-18N broadband termination covers dc to 18 GHz with a maximum SWR of 1.25. The unit absorbs 5 W CW at 95°C. Connector is type N male. Model TO-18N meets enviromintal requirements of MIL-E-16400, Class 1 and MIL-E-5400, Class 3. MTBF as calculated per MIL-HDBK-217B for airborne (unattended) service is in excess of 400,000 hours. Price: $50.00. Delivery: stock to 2 weeks. **Micronetics, Inc., Norwood, NJ. Gary Simonyan (201)767-1320.**

Circle 133.

10 W FLANGE TERMINATION



Model 229-4000 is a 10 W SMA-style flange termination with a maximum SWR of 1.30 up to 18 GHz; an option of 1.2 SWR up to 18 GHz is also available. Operating terperature range is -50°C to 125°C. Flange terminations are also available in TNC and N type hole patterns. Contact for either slotted or tab configurations are available. Price: 1-9, $12,50 each. Delivery: 6 weeks ARO. **Solitron Microwave Connector Division, Port Salerno, FL (305) 287-5000.**

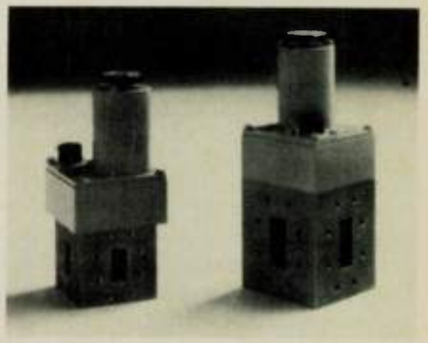
Circle 147.

10-100MHz POWER DIVIDERS/ COMBINERS

A line of in-phase power dividers/ combiners operating in the 10 to 100MHz frequency range are available with 2, 3, 4, 8, 12 or 16 input/output ports with equal division or combination of power. The units are in compact packages with BNC, TNC or SMA connectors. Construction is lumped element on low loss, stable dielectric material. Price: from $35.00 in unit quantity. Delivery: 4 to 6 weeks (small quantity). **RLC Electronics, Inc., Mt. Kisco, NY. Alan Borck (914)241-1334.**

Circle 135.

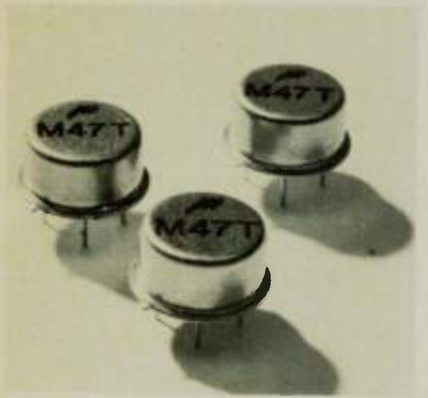
80 mS WAVEGUIDE SWITCHES



A line of latching transfer switches in WR62, WR75, WR90 and WR112 waveguide sizes feature 60 dB minimum isolation, 1.1 maximum SWR, 0.1 dB maximum insertion loss and 80 mS switching time. Power ratings are 3kW (CW) for the WR62 units and 5kW (CW) for the other waveguide sizes. Indicating switch contacts close before the waveguide rotors are activated and they can double as interlocks to inhibit HPAs and other system sources during switching. 28 VDC and 48 VDC drive voltages can be accomodated and all models can be pressurized to 5 PSIG. Price: from $550.00. Delivery: 60 days. **NEICO Microwave Co., Hopkinton, MA. Robert Ranslow (617) 435-6366.**

Circle 157.

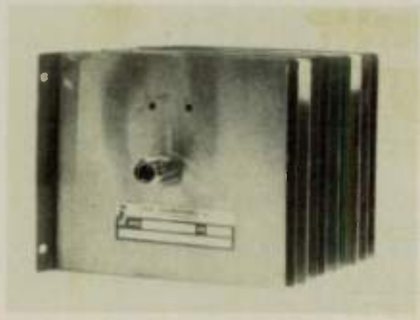
10-1000 MHz HIGH LEVEL MIXERS



Model M47T double balanced mixer covers the 10-1000 MHz range with an intercept point of +27 dBm, typical, with an LO drive level of +23 dBm. The mixer features IF bandwidth up to 600 MHz and typical conversion loss of 5.5 dB. The unit is hermetically sealed in a TO-8 package. Price: (100 quantity) $51.50. Delivery: two weeks ARO. **Magnum Microwave Corp., Sunnyvale, CA. David Fealkoff (408) 738-0600.**

Circle 158.

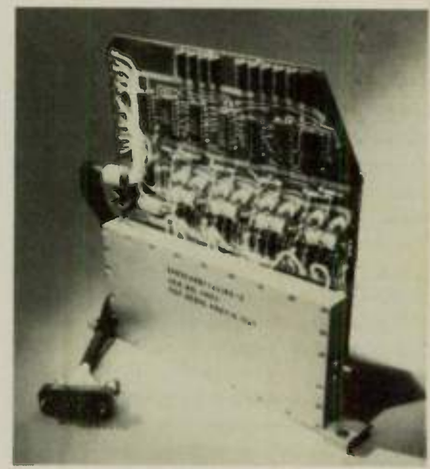
800-8000MHz POWER DRY LOAD



Model FT2974 is an air-cooled medium power dry load covering 800 to 8000 MHz and optimized over the TACAN band. The unit offers an 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 137.

3kW X-BAND PHASE SHIFTER



Model VSX-9738 is a 4-bit PIN-diode phase shifter that handles 3kW of peak RF power over the 8.5 to 9.6 GHz frequency range. The unit will operate within any 250-MHz bandwidth within its range. Switching time is less than 10 μsec and switching rate is 6000 Hz. The unit is a key component of the AP/TPN-22 precision approach and landing radar (PALR) system. It measures 7" x 7" x ¾"; all non-RF interconnections with the system are made through a single MS 24020-1 multi-pin connector. **Varian Associates, Beverly, MA. (617) 922-6000.**

Circle 159.

BROADBAND DOUBLE BALANCED MIXERS

Model DMJH4-18 double balanced mixer has a 1-8 GHz IF range and 4 to 18 GHz RF/LO coverage. Model DMJ2-18 provides dc to 500 MHz IF, 2 to 18 GHz RF/LO coverage and is also useful as a third harmonic mixer. Using fundamental mixing from 2 to 6 GHz and third harmonic mixing from 6 to 18 GHz allows a 2 to 6 GHz local oscillator to cover the entire 2 to 18 GHz range. Both mixers have a typical conversion loss of 8 dB; LO-RF isolation is typically 25 dB for the DMJ4-18 and 20 dB for the DMJ2-18. Connectors are SMA. Size: 1.25"x .375" x .750". Price: Model DMJ2-18: $475.00, Model DMJH4-18: $156.00. Delivery: 90 days ARO. **RHG Electronics Laboratory, Inc., Deer Park, NY. Sid Wolin (516) 242-1100.**

Circle 156.

SILICON POWER TRANSISTORS

Two ranges of transistors are offered. The linear (Class A) transistors have output powers up to 2.4 W and can be used to drive one of several class-B power transistors to provide up to 4 W over the 3.7 to 4.2 GHz band. These transistors are intended for use in communication links. Also available is a range of pulsed-power transistors for powers up to 700 W and frequencies up to 3.5 GHz. Typical applications for these transistors include transmitters for IFF, DME transponder and radar systems. **Amperex Electronic Corp., Hicksville, NY. John Salvey (516) 931-6200.**

Circle 148.

100-1000MHz SAW RESONATORS

Low cost SAW resonators for narrowband filtering and oscillator applications in the 100 to 1000 MHz range are offered. Typical performances at 700 MHz includes 10 dB insertion loss, spurious rejection greater than 15 dB and 50Ω loaded Q of 4500. Temperature stability is 40 ppm from 0° to 70°C and 125 ppm from –40°C to 125°C. Packaging is TO-5, TO-8 or HC-18. Delivery: stock to 8 weeks. **Sawtek. Inc. (305) 299-4441.**

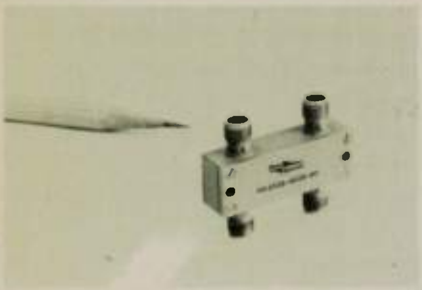
Circle 138.

WAVEGUIDE ROTARY JOINT OPERATES FROM 3.1 to 3.7 GHz

Model 21013 single channel waveguide rotary joint operates from 3.1 to 3.7 GHz with 170 kW peak power and 15 kW average power ratings. The unit is designed for use in the elevation axis of ground terminal surveillance radar, and measures 15.5" long x 4.5" diameter. **Kevlin Manufacturing, Woburn, MA. E. Lattanzi (617) 935-4800.**

Circle 141.

7-18 GHz QUADRATURE HYBRID



Model 2035-4035-00 is a miniature dielectric quadrature hybrid covering the 7-18 GHz band with a power handling rating of 100 W CW output when used as a combiner. The unit is a non-crossover type hybrid with 15 dB minimum isolation, mean coupling of 3.0 ±0.5 dB and amplitude balance of ±0.5 dB. **Omni-Spectra, Inc., Merrimack, NH. John C. Callahan (603) 424-4111.**

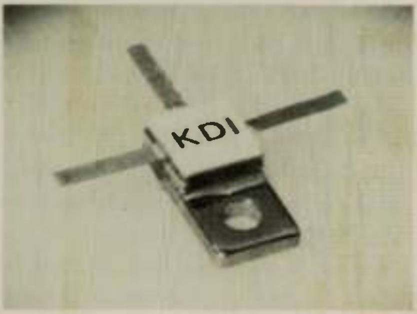
Circle 140.

MILLIMETER BAND BIAS - TUNED OSCILLATORS

A series of bias-tuned Gunn oscillators available in U-band (40-60GHz) through W-band (75-90 GHz) and in several output/tuning bandwidth combinations have bandwidths varying from 200 to 1000 MHz and power output levels varying from 5 mW to 100 mW. The W-band units are available with the standard 20 mW minimum power output and 200 MHz tuning bandwidth or as broadband units with 5 mW minimum power output and 1 GHz tuning bandwidth. Custom oscillators are available with voltage regulators and frequency stabilizing circuits. Price: from $1600. Delivery: 60 days ARO. **Hughes Electron Dynamics Division, Torrance, CA. (213) 517-6400.**

Circle 143.

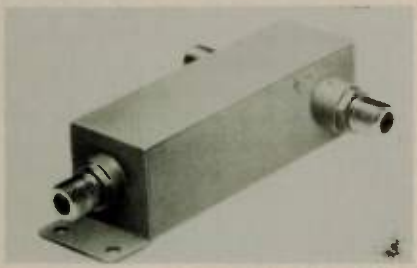
20 W POWER DIVIDER



Model PPD 20-2 20 W two-way power divider is designed for leveling loop applications. The unit features flat tracking between outputs and operates in the dc to 2.5 GHz range. The unit operates at power levels up to 20 W with an effective source SWR of 1.10 maximum. Price: 1-9 is $49.75. Delivery: from 4-6 weeks. **KDI Pyrofilm Corp., Whippany, NJ. Al Arfin (201) 887-8100.**

Circle 142.

75 Ohm POWER DIVIDER



Model 3809 TVRO 75 ohm power divider has type F connectors and operates over the 975-1425 MHz band. The unit has isolation of 20 dB minimum and return loss (to each leg) is 3.5 dB maximum. Price: $85 (1-9 quantity). Delivery: 2 weeks. **Microwave Filter Co., Inc., Syracuse, NY. Emily Bostick (800) 448-1666.**

Circle 139.

ISOLATED FEED-THROUGH ADAPTORS

BNC and TNC feed-through adaptors which eliminate ground loops and common mode current in coax systems are offered. The Amphenol 31 series have voltage ratings of 500V peak, center contact resistance of 1.5 ohms, dielectric withstanding voltage of 1,000V RMS and insulation resistance of 500 ohms minimum. Mating for both types is in accordance with MIL-C 55339. Price: $6.43 (1-24). Delivery: 2 weeks. **Amphenol North America Division, Oak Brook, IL. Art Morse (312) 986-2322.**

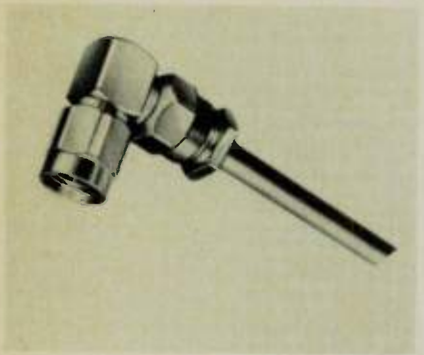
Circle 163.

LOW COST TRIMMER CAPACITORS

The DV-2 series of trimmer capacitors are available with 50 VDC and 100 VDC voltage ratings and cover the capacitance ranges from 10 to 120 pF at 50 VDC to 10-60 pF at 100 VDC. Mounted case height is less than .197". Temperature coefficients range from NPO to N1200. Price: less than 20¢ each in quantity. Delivery: from stock. **JFD Electronic Components, Douglasville, GA. (404) 949-6900.**

Circle 149.

ADJUSTABLE RIGHT ANGLE SMA CONNECTOR



Model 50-611-9702-31 right angle SMA connector for semi-rigid cable permits orientation adjustment in the field. The connector is mounted on the semi-rigid cable with a solder-clamp assembly. A clamping nut permits adjustment of the right angle body orientation. **RF Components Division, Sealectro Corp., Mamaroneck, NY. (914) 698-5600.**

Circle 151.

WAVEGUIDE DELAY LINES FOR 4 to 40 GHz

Long waveguide runs are offered for use as delay lines with nanosecond values, as power attenuators or as power loads. Waveguide sizes range from WR-137 to WR-28 including double ridged sizes WRD-350 to WRD-180 covering the 4 to 40 GHz range. SWR and insertion loss values are assured by bilaterial swept measurement techniques. Various flange types and combinations are available. **Microtech, Cheshire, CT. (203) 272-3234.**

Circle 152.

RIGHT ANGLE SMA CONNECTORS OPERATE TO 18 GHz

SMA right angle connectors and adaptors cover the dc to 18 GHz range with a SWR rated at 1.05 + .010f (GHz) maximum, or 1.23 maximum at 18 GHz. The connectors and adapters are mechanically interchangeable with low profile swept radius connectors. Available in passivated stainless steel and gold plating. Delivery: stock to 8 weeks. **Omni-Spectra, Inc., Merrimack, NH. John Callahan (603) 424-4111.**

Circle 155.

PULSED POWER METER COVERS 950 — 1250 MHz

Model PPM-101A pulsed power meter covers the frequency range 950 to 1250 MHz and is used to measure peak power levels within the TACAN and IFF frequency ranges. The unit is capable of measuring pulsed power ranges from 20 to 500 and 200 to 5000 W with an accuracy of ± 0.5 dB (traceable to NBS). The instrument requires no adjustments for calibration and it can operate in either automatic or manual modes, the latter for measurement of multi-peak complex waveforms. **Republic Electronic Industries, Corp., Melville, NY (516) 249-1414.**

Circle 165.

Microwave Products

OPTICAL FIBER LIGHT CHOPPER

The Model SWC in-line optical fiber light chopper provides sine wave and/or square wave modulation of a fiber contained light beam. It operates from a small remote driver and is available in any frequency from 200 to 5,000 Hz with an on-off duty cycle of either 50%-50% or 90%-10%. Other duty cycles are available on special order. Stability is better than 1% of operating frequency from -40° to +60° C, attenuation is 2 dB. Drive voltage is either 24 VDC or 115 VAC with input power approximately 60 mW at 24 VDC. Amphenol connectors are standard. **American Time Products Division, Frequency Control Products Inc., Woodside, NY. John Murray (212) 485-5811.**

Circle 161.

4-18 GHz FLATPAK DOUBLE-BALANCED MIXERS

The first pair of a planned complete line of double-balanced microwave mixers in the AVANPAK microwave flatpak designed for direct installation in 50 ohm stripline/microstrip systems and are also suitable for field attachment of RF connectors are offered. The DBX-184 covers the 4-18 GHz RF/LO range with a dc-4 GHz IF range; the DBX-185 covers a 5-18 GHz RF/LO range with an IF range of dc-6 GHz. Both are available in +7 dBm LO drive ("L"-suffix) and +10 dBm drive ("M"-suffix) versions. All four feature 6.0 dB typical conversion loss/noise figure, 30 dB typical LO-RF and 25 dB RF-IF isolation. Two-tone input intercept points are +15 dBm and +20 dBm for the "L" and "M" versions, respectively. Packages are hermetically sealed. Price: DBX-184: $425.00; DBX-185: $490.00 in small quantites. Delivery: 3 weeks. **Avantek Inc., Santa Clara, CA. Jim Cochrane (408) 727-0700 ext. 129.**

Circle 162.

Amplifiers

7.25-7.75 GHz 170°K GaAs FET LNAs

The AW-7720 low noise amplifier series covers the 7.25-7.75 GHz band and individual models are available with 18, 27, 36 or 44 dB of gain. All models offer 2 dB noise figure, 170°K noise temperature, across the full band, input and output SWR's are 1.25 maximum, output power is +10 dBm at 1 dB compression and the gain slope/group delay characteristics are suitable for wide band communications applications. Standard input power is 115 VAC; +15 to +28 VDC input power versions are available. Price: $3,000 (single quantities). Delivery: 120 days. **Avantek Inc., Santa Clara, CA. Jonnie Danielson (408) 496-6710 ext. 16.**

Circle 169.

Microwave Products

Material

CORROSION RESISTANT RFI CONDUCTIVE CAULKS AND SEALANTS

ECCOSHIELD® VY-NC and VY-NN non-noble metal materials are single component, electrically conductive formulations for containment or exclusion of RF energy which are particularly suitable for use under sea and salt conditions. Metal and plastic enclosures, structures, cabinets, conduits, modules and components can be sealed with the materials. **Emerson & Cuming Microwave Products, Canton, MA. Joe Flaherty (617) 828-3300.**

Circle 176.

ALUMINA AND QUARTZ SUBSTRATES

A line of alumina and quartz substrates for thin film, hybrid and microwave IC applications is supplied in either 99.6% alumina with a dielectric constant of 9.9 or fused silica (quartz) with a dielectric constant of 3.8. Both substrates are available in sizes from 1" x 1" through 2" x 2". Standard thickness is 0.025" ± 0.001" or ± 0.0005" polished on both sides to better than one microinch surface finish. Other dimensions are available on special order. Availability: standard sizes from stock. **Valley Design Corp., Littleton, MA (617) 486-8933.**

Circle 177.

CIRCUITS ETCH TO 1.5 MIL

Etching of micro-strip and strip-line microwave circuits to etched line and space widths as small as 1.5 mils with tolerances of less than ± 0.0005" on line widths is offered. Circuits can be produced from woven and nonwoven teflon fiberglass, Polyguide, rexolite, ceramic teflon and other stock materials with dielectric thicknesses from 0.001" to ¼". The circuits can be bonded, double-sided, plated and finished to specifications and are manufactured to MIL SPEC 45208. Price: varies with material and specifications. Availability: two to three weeks for engineering quantities. **Microwave Printed Circuitry, Inc., Lowell, MA. Michael P. Casper (617) 452-9199.**

Circle 178

Cable

FLEXIBLE COAXIAL CABLE

A 50 ohm flexible microwave cable for high frequency applications is offered. Part No. 50H155R parallels the performance of 141 semi-rigid cable through 18 GHz. The cable has a diameter of .155"; it is larger than its semi-rigid counterpart and 10% lighter. **Insulated Wire Inc., Ronkonkoma, NY (516) 981-7424.**

Circle 179.

Microwave Products

Antenna

SUSCEPTIBILITY-TESTING ANTENNA

The AT2000 CAVITENNA radiator is designed to create high intensity fields for rfi and emi susceptibility testing by acting as a cavity exciter within shielded test rooms. The system will handle up to 1250 W from 30 to 1000 MHz, maximum input from 30-250 MHz is 3500 W and from 250-500 MHz, 2000 W. At rated power it will generate fields that exceed 600 V/meter in the 200 MHz region. The system uses a wall or ceiling of the room as a ground plane providing wide bandwidth performance in a small package. The radiator measures 46" x 24" x 20" and weighs 30 lbs. It is available with optional magnetic clamp mounts for quick repositioning. **Amplifier Research, Souderton, PA (215) 723-8181.**

Circle 171.

MULTI-BAND HIGH PERFORMANCE ANTENNA

The SHX horn antenna is an all-metal conical design with a field-replaceable TEGLAR radome. With the appropriate combining network, the antenna is capable of multi-band operation at 4, 6 and 11 GHz and, in these bands, it meets or exceeds KS-21972 Coordinating Specifications and US FCC Standard "A" (Part 21) requirements. **Andrew Corp., Orland Park, IL.**

Circle 172.

Sub Systems

Ku TO C-BAND TRANSLATOR

Model DC12/4 is a Ku to C-band frequency translator providing block down conversion of the entire 11.7-12.2 GHz frequency band to 3.7 - 4.2 GHz. Alternatively, the unit can be used to downconvert a Ku-band video signal into an unused channel in a 24 channel C-band TV receiver at a cable TV head end. The unit interfaces directly with a 12 GHz LNA and 4 GHz receivers or converters and is suitable for video message or data carriers. Unit is self-contained, measures 1 3/4" in height and is designed for rack mounting. **LNR Communications Inc., Hauppauge, NY. Nancy Wagner (516) 273-7111.** **Circle 175.**

Tube

75 W TWT FOR EARTH STATION

Model TH 3641 is a 75 watt, 6 GHz travelingwave tube for use in earth station uplink transmiters. The tube is designed for small "domestic" satellite communications stations. The unit exhibits better than 25% efficiency at saturation. Other features include simple conduction cooling, compactness and ppm beam confinement. **Thomson CSF Electron Tube, Clifton, NJ (201) 779-1004.** **Circle 174**

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New Literature

SATCOM RECEIVER CATALOG

A 17-page fully illustrated short form catalog details a series of telemetry and meteorological satellite receivers and related equipment. Photographs, application information and specifications are included for the complete series. **Microdyne Corporation, Rockville, MD. K. B. Boothe (301) 762-8500.**

Circle 181.

ATE SWEEP GENERATOR BROCHURE

An 18-page brochure describes a series of sweep generators designed for the automatic test equipment market. The programmability of the Series 6600 is demonstrated by a listing of the mnemonics recognized by the sweep generator. A sample program, written for the Model 85 controller shows how the mnemonics are used. Graphics illustrate the ATE applications and benefits of programmability. Frequency accuracy, stability and spectral purity features of the 6600 suit it particularly for ATE applications. Also outlined in the brochure are features of the 560 Scalar Network Analyzer. **Wiltron Co., Mountain View, CA. Walt Baxter (415) 969-6500.**

Circle 182.

4.3 GHz TRANSMITTER OSCILLATOR APPLICATION NOTE

A 4-page application note (AN81401) "4.3 GHz Pulsed Transmitter Oscillator with 11 W Output Power" covers theory of operation, bias circuit operation, results and circuit adjustments. Also included are diagrams of the circuit, RF output spectrum and power, emitter current pulse, collector to emitter voltage, PC board layout and dimensions, and oscillator chassis. **California Eastern Laboratories, Inc., Santa Clara, CA. Jerry Arden (408) 988-3500.**

Circle 183.

"IN-A-CABLE" FILTER ASSEMBLIES BULLETIN

A 12-page bulletin details the patented "In-A-Cable" band-pass and low-pass filters and cable assemblies which are implanted within continuous sections of Micro-Coax[R] semi-rigid coaxial cables. Standard cable diameters of 0.0865", 0.141" and 0.250" with eight types of connector options are included. Charts and graphs illustrate performance characteristics, size and weight of the assemblies, and a special section describes compatible military environmental conditions. **Uniform Tubes, Inc., MicroDelay Div., Collegeville, PA. (215) 539-0700.** **Circle 185.**

GaAs FET LOW NOISE AMPLIFIER CATALOG

A two-color, short form catalog details low noise and ultra low noise GaAs FET amplifiers covering VHF through Ku-band. Dimensions and performance specifications are provided. **California Amplifier, Inc., Newbury Park, CA. Jacob Inbar (805) 498-2321.**

Circle 184.

2 TO 18 GHz AMPLIFIER GUIDE

A 20-page guide book provides technical data on a complete line of GaAs FET amplifiers covering the 2 to 18 GHz frequency range. The standard line specializes in broadband octave and multi-octave amplifiers for use in electronic warfare systems and narrow band amplifiers for use in radar, telecommunications and telemetry applications. Detailed specifications and outline drawings are included. **Narda Microwave Corp., Plainview, NY. (516) 349-9600.**

Circle 188.

LOG AMPLIFIER DATA SHEET

A 6-page data sheet describes a line of log amplifiers. Included in the description are characteristic curves, log linearity, typical characteristics and typical outline dimensions. A short form catalog of products and ordering instructions is part of the data. **Varian/Beverly, Beverly, MA.**

Circle 186.

MICROWAVE COUNTER COMPARISON GUIDE

A 6-page point-by-point comparison analyzes specifications, features, benefits and trade-offs of automatic frequency counters manufactured by EIP Microwave, Inc., and Hewlett-Packard. EIP Models 545A/548A and the HP 5342A/5343A are compared based on published specifications and in-house testing of frequency range, sensitivity, amplitude discrimination, acquisition time, power measurement capability and basic design architecture. **EIP Microwave, Inc., San Jose, CA. (408) 946-5700.**

Circle 187.

TVRO COMPONENTS CATALOG

A complete line of TV Receive-Only components is described in this short form catalog. Switches, filters, terminations, dividers, attenuators, and blocks available in the 3.7 to 4.2 GHz and 50 to 90 MHz ranges are detailed. **RLC Electronics, Inc., Mt. Kisco, NY. Alan Borck (914) 241-1334.**

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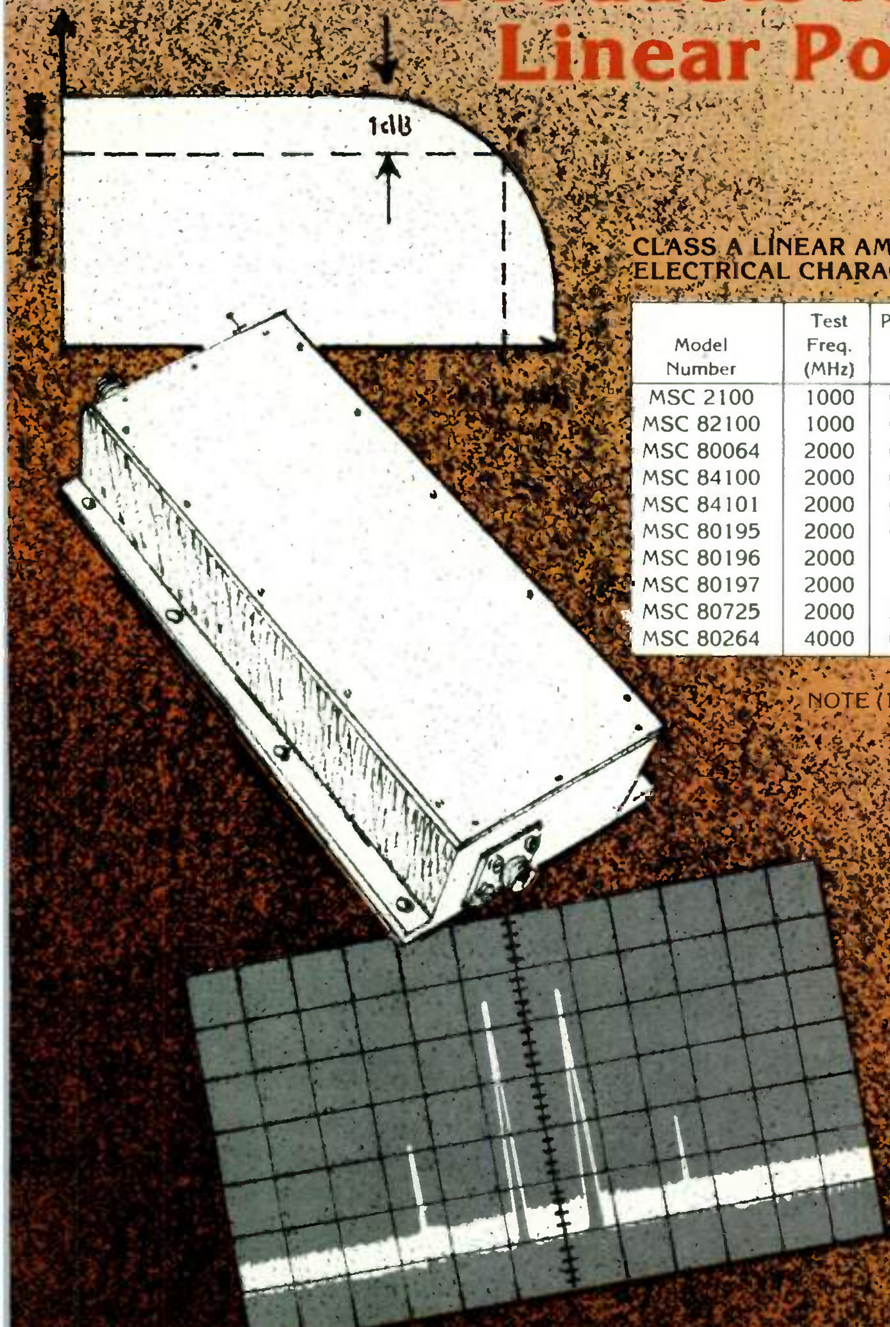

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ELECTRICAL CHARACTERISTICS (@ 25°C)

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

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

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