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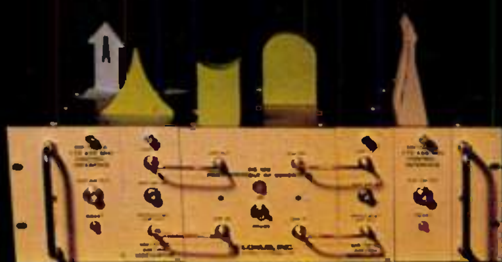
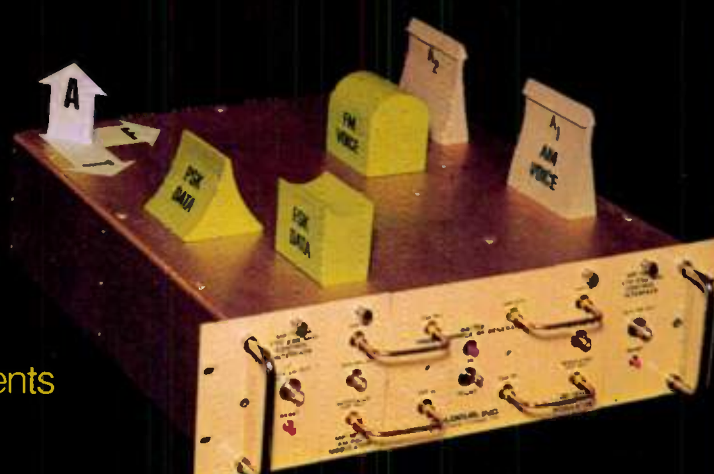


COMMUNICATIONS

- mm Satellite System
- Semiconductor Light Sources
- 18 GHz Repeaters



AND

Cooled L-Band Amplifier
Suspended Substrate Components

horizon house

NEW SPECTRUM ANALYZERS

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

100 kHz to 40 GHz

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

- 80 dB on-scale range for 10 dB/div. log scale.
- Improved accuracy for 2 dB/div. and linear scales.
- Greater sensitivity, flatness and noise-free dynamic range.
- All signal levels directly measured in dBm.
- Improved dynamic range for harmonics and IM.
- Improved display stability and resolution through reduced residual FM and noise sidebands.
- Greater accuracy for Frequency readout and Spans.
- Expanded frequency coverage; 3 MHz to 40 GHz (Models 630B & 640B).
- Cooler operating temperatures increase service life and reduce warmup time.

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

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

Model	Freq. Range	U.S.A. Prices (Sept. 80) Without Memory	Including Memory
632B-1	100 kHz to 2 GHz	N/A	$ 9,250
630B	3 MHz to 40 GHz	$11,325	$12,975
640B	3 MHz to 40 GHz with internal preselector	$15,325	$16,975

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

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

CIRCLE 1 ON READER SERVICE CARD

The most significant price breakthrough in DOUBLE-BALANCED MIXERS!

...from Mini-Circuits of course!

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For demanding industrial and commercial applications, where low-cost and high-performance are critical; model SBL-1 will fill your need.

Don't let the low price mislead you. As the world's number one manufacturer of double-balanced mixers, Mini Circuits' has accumulated extensive experience in high-volume production and testing, a key factor in achieving a successful low cost/high performance line of products.

The tough SBL-1 covers the broad frequency range of 1-500 MHz with 6 dB conversion loss and isolation greater than 40 dB. Only well-matched, hot-carrier diodes and ruggedly constructed transmission-line transformers are used Internally, every component is bonded to the header for excellent protection against shock, vibration and acceleration.

Here are some of the steps taken to ensure quality: Every SBL-1 is RF tested two times, every solder connection is 100 per cent inspected under a high power microscope, all transformer leads are double-wrapped, and all components are rated for more than + 85 °C operation.

Of course, our one-year guarantee applies to these units.

Model SBL-1

metal case, non hermetic-seal

Frequency Range, MHz

LO 1-500 RF 1-500 IF DC-500

Conversion Loss, dB		Typ	Max
One Octave from Band Edge		5.5	7.5
Total Range		6.5	8.5
Isolation, dB		**Typ**	**Min**
Lower Band Edge to	LO-RF	50	35
One Decade Higher	LO-IF	45	30
Mid Range	LO-RF	45	30
	LO-IF	40	25
Upper Band Edge to	LO-RF	35	25
One Octave Lower	LO-IF	30	20

Signal, 1dB Compression Level +1dBm
Impedance, All Ports 50 ohms
Electronic Attenuation Min (20mA) 3dB

World's largest manufacturer of Double Balanced Mixers

Mini-Circuits

MINI-CIRCUITS LABORATORY

A Division of Scientific Components Corp

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Domestic and International Telex 125460 International Telex 620156

21 Rev IBL

CIRCLE 5 ON READER SERVICE CARD

POWER DIVIDERS

Broadband, 0.5 — 4.2 GHz • Only 0.2 dB insertion loss
Isolation over 30 dB midband, 25 dB at bandedges • Octave bandwidths
Two way • up to 10 W (matched output)

- High performance microstrip construction
- Housed in rugged RFI-shielded aluminum case
- Available with BNC, TNC, SMA and Type N connectors
- Meets MIL-202E standards
- Also useful as power combiners at signal levels up to +10 dBm

$39.95 1-9

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

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

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

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

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

World's largest manufacturer of Double Balanced Mixers

Mini-Circuits
A Division of Scientific Components Corp.

2625 East 14th Street Brooklyn, New York 11235 (212) 769-0200
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CIRCLE 6 ON READER SERVICE CARD

Coming Events

29TH INT'L WIRE AND CABLE SYMPOSIUM NOV. 18-20, 1980 Sponsor: US Army Communications Research and Development Command. Place: Cherry Hill Hyatt House, Cherry Hill, NJ. Sessions: Outlook for Voluntary Standards in USA, Cable Design I and II, Materials, Connectors, Fiber Optics I, II and III, etc. Contact: Elmer F. Godwin, Dir., GEF Associates, Shrewsbury, N.J. Tel: (201) 741-8864.

1981 DoD-INDUSTRY FIBER OPTICS STANDARDS CONFERENCE APRIL, 1981 Sponsor: Electronic Industries Association (EIA). Place: Washington, DC. Theme: Fiber optic standardization. Contact: EIA, 2001 Eye St., N.W., Washington, DC 20006. Tel: (202) 457-4981.

3RD INT'L CONFERENCE ON I-O AND OPTICAL FIBER COMMUNICATION APRIL 27-29, 1981 Call for Papers. Sponsor: Optical Society of America. Place: Hyatt Regency San Francisco, San Francisco, CA. Topics: Research and development in integrated optics and optical fiber communication, including such areas as fibers and cables, connectors, couplers, equipment systems and transmission techniques and integrated optics and active devices. Submit 2 copies of 35-word abstract and 200-to-500-word summary by Nov. 28, 1980 to: Optical Society of America, 1816 Jefferson Place, N.W., Washington, DC 20036. Tel: (202) 223-8130.

16TH SYMPOSIUM OF THE INT'L MICROWAVE POWER INSTITUTE JUNE 9-12, 1981 Call for Papers. Sponsor: Int'l Microwave Power Institute. Place: Royal York Hotel, Ontario, Canada. Theme: Non-communication aspects of microwaves, such as bioeffects, microwave ovens, industrial microwave systems, RF systems and applications, microwave power generation, equipment safety, etc. Submit up to 40-word abstract and 500-word summary by Jan. 5, 1981 to: Dr. S. Kashyap (US and Asia), National Research Council of Canada, Div. of E.E., Ottawa, Ontario, KIA OR8 Canada. Tel: (613) 993-9214.

1981 IEEE/MTT-S INT'L MICROWAVE SYMPOSIUM JUNE 15-17, 1981 Call for Papers. Sponsor: IEEE MTT-S (held jointly with IEEE AP-S on June 17-19, 1981). Place: Bonaventure Hotel, Los Angeles, CA Theme: "Around the World with Microwaves," including such topics as CAD and measurement techniques, microwave, and mm-wave solid-state devices and IC's, etc. Submit 35-word abstract and 500 to 1000-word summary by Jan. 15, 1981 to: Dr. Don Parker, TPC 1981 MTT-S Symposium. Hughes Aircraft Co., Bldg. 268, M.S. A54, Canoga Park, CA 91304.

System Components FROM ENGELMANN

Integration of Stripline Components Yields TACAN Switch Circuit

High peak power diode switching is made possible through the integration of a SPDT switch with antenna couplers and filters in a stripline configuration.

The switch diodes are mounted in an air-evacuated and dielectrically loaded circuit, using a special process that increases high voltage breakdown ratings significantly.

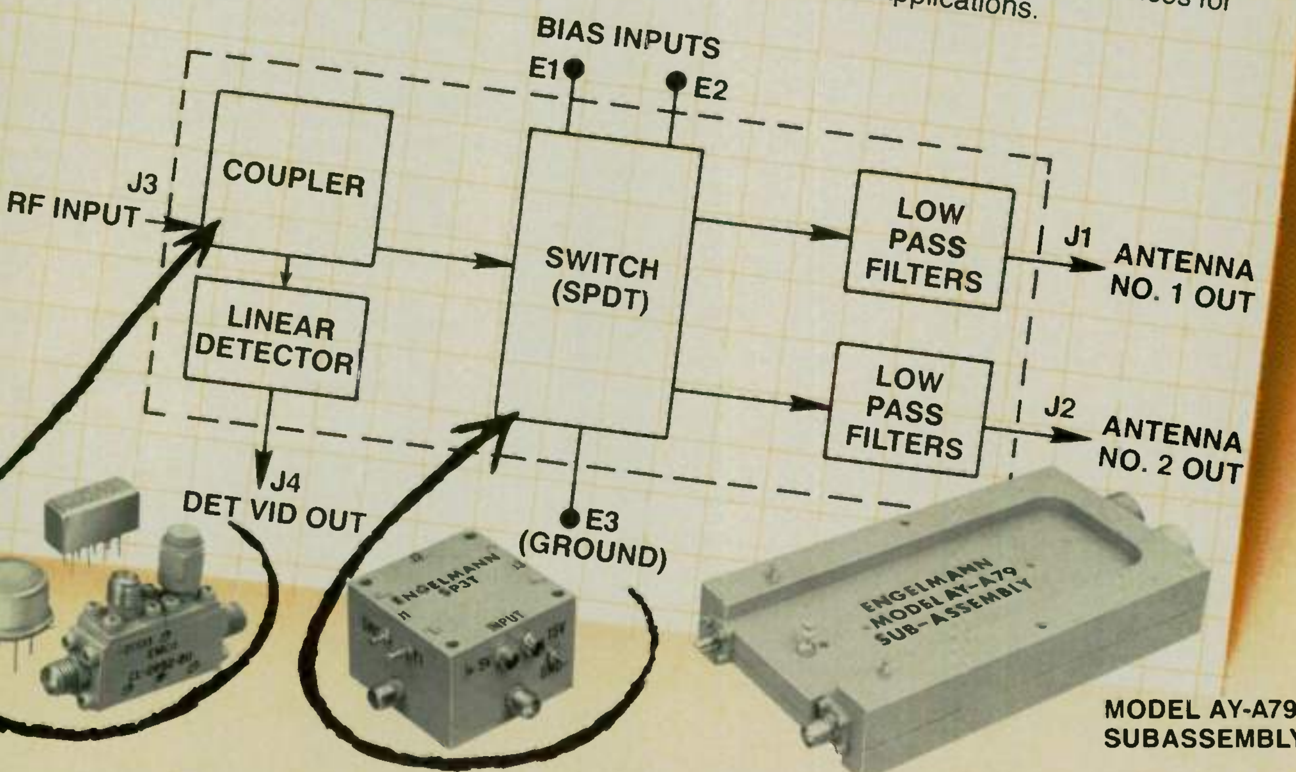
The process permits hot switching into open or short circuit loads of all phases with a peak power of 4.5KW in the 935-1260MHz frequency range, at 85°C and altitude of 80,000 feet.

Secondary circuitry in the integrated unit includes dual low-pass filters and a linear directional detector which provides an accurate voltage output proportional to RF input, as an input power monitor function.

The antenna filtering suppresses transmitter harmonic content to assure compliance to MIL-STD-461 EMI requirements out to 12.4GHz.

Additional specifications include a maximum insertion loss of the circuit of 1.35db in the "on" position, and 20db minimum in the "off" position.

The switch module is available separately, or may be integrated with other devices for similar special applications.



MODEL AY-A79 SUBASSEMBLY

We'll Supply the Components...or the Complete Integrated Subsystem!

DIRECTIONAL COUPLERS
A full line...from
10KHz to 18 GHz
- P.C. Board plug-in
- All connector types
- Stripline
- Octave and multioctave

DIODE SWITCHES
A full line...from
10MHz to 18GHz
- SPST up to SP16T
- P.C. Board plug-in
- All connector types

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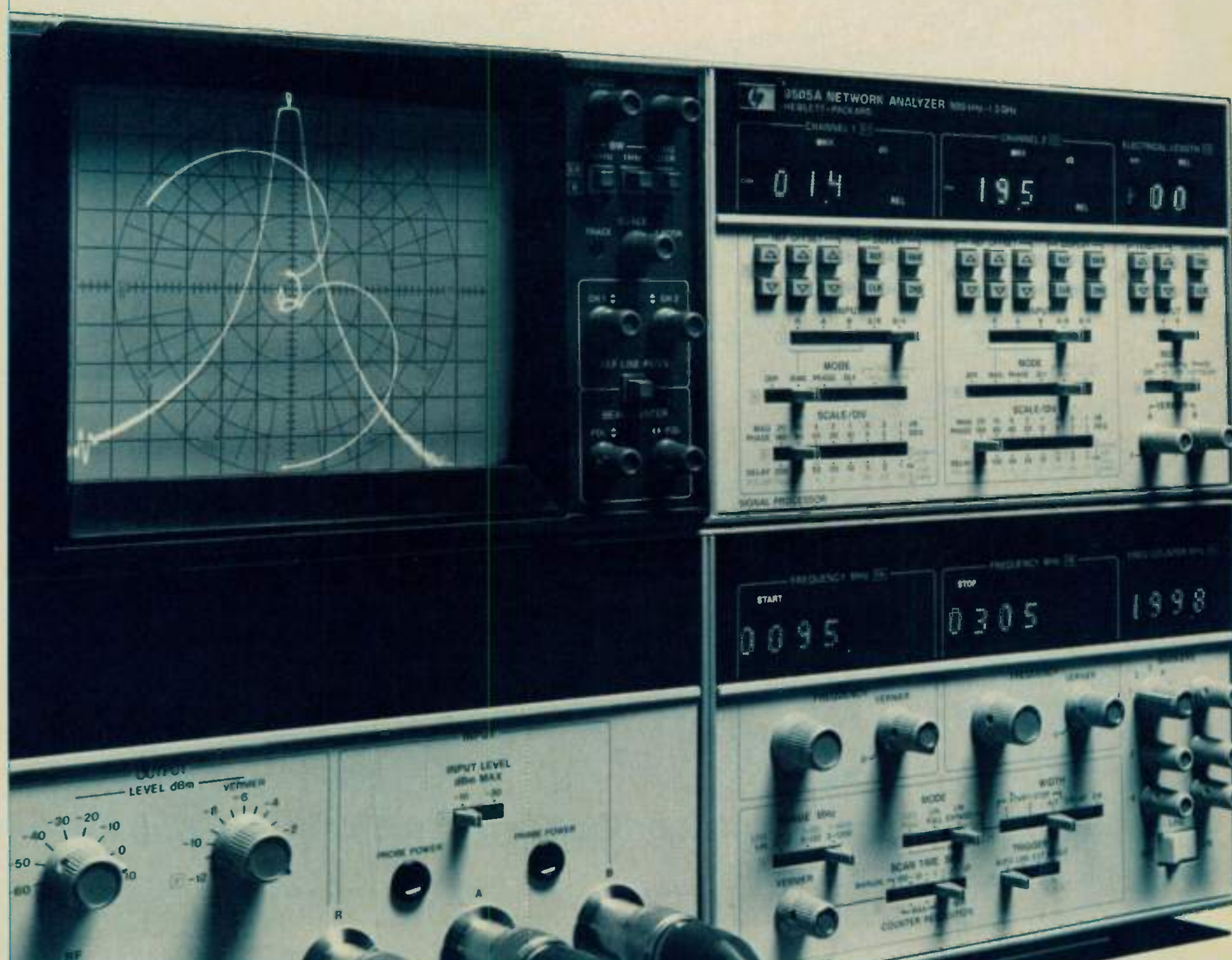
E80-3

The 1300 MHz Network Analyzer–

Until now performance like this was beyond reach:

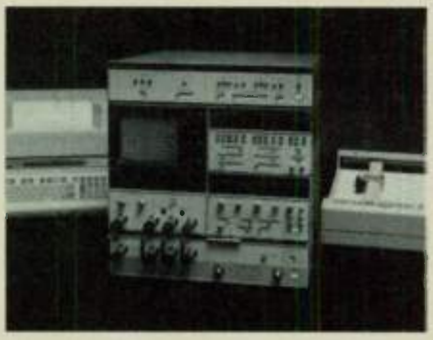


- **Over 3 Decades of Swept Frequency Coverage**
- **100 dB Dynamic Range**
- **Direct Measurement of Group Delay**

HP's 8505A Network Analyzer brings the precision, resolution and range you need for the measurement of phase and magnitude of transmission and reflection, group delay and deviation from linear phase. And any two parameters can be measured and displayed simultaneously.

- Test signals come from the 8505A's built-in high performance sweeper with exceptional spectral characteristics and a wide variety of sweep modes (including two independent start/stop sweeps) to accommodate virtually any test requirement.
- The 8505A's 500 kHz to 1.3 GHz frequency range gives you the broad coverage you need to characterize such networks as filters, transistors, antennas, cables, SAW devices and crystals.
- Your measurements are fast and accurate thanks to a swept display with a marker system that provides a high resolution digital readout of the parameter's value at the frequency of any of five variable markers. And group delay measurements are made directly; no calculations required. Or you can observe phase distortions directly in the form of *deviation* from linear phase using the 8505A's revolutionary electronic line stretcher.
- With optional phase-lock capability, the 8505A can be locked to such precision signal sources as the HP 8640 and 8660 Signal Generators. This provides the stability and resolution needed to characterize ultra narrowband devices such as crystal filters.

Get the speed, precision and efficiency of automatic measurements.



Because the analyzer is programmable, via the Hewlett-Packard Interface Bus (IEEE-488), you can combine the 8505A with a computing controller such as HP Model 9825A Desktop Computer to configure a powerful automatic measurement system. With remarkably simple programming you can make many measurements quickly and with enhanced accuracy, and easily format the data to the form you want. The result is high throughput for cost-effective operation in both production test and design lab applications.

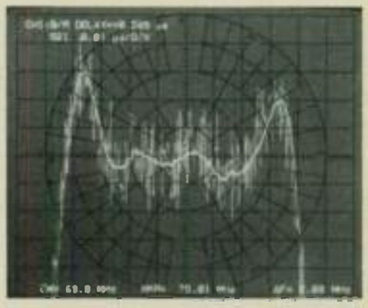
Find Out More.

We've only touched on the highlights of the 8505A's performance and capabilities here. For complete data, contact your nearby HP field sales office, or write 1507 Page Mill Road, Palo Alto, CA 94304.

Add capability with HP's Storage-Normalizer.

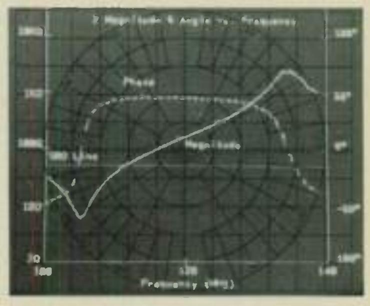
The companion HP 8501A Storage-Normalizer brings these additional features to the 8505A Analyzer:

- *Digital storage* for flicker-free displays.
- *Normalization* to remove errors and make direct comparisons.
- *Magnifier* for up to a tenfold increase in resolution.
- *CRT Labeling* that presents major 8505A settings and marker data.
- *Signal Averaging* that raises signal-to-noise ratio, thereby improving narrowband group delay and low signal level measurements.



Group Delay of 70 MHz bandpass filter with and without averaging (Vert. scale 5nsec/div)

When the HP-IB programmable 8501A is combined with the automatic 8505A/computing controller combination, the system offers versatile display capabilities for text and graphics plus high-speed digitizing for fast, yet precise and comprehensive measurements.



Reflection Coefficient data reformatted to impedance magnitude and angle

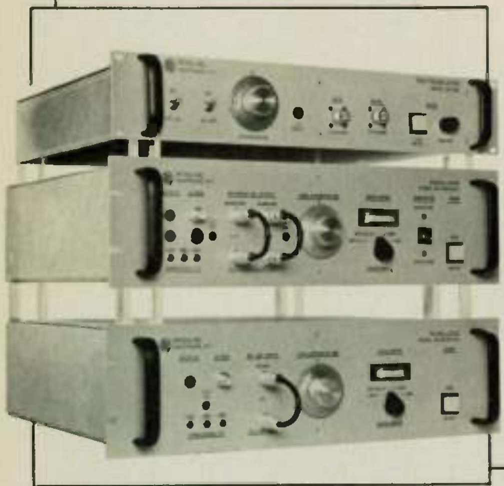
45906A

HEWLETT PACKARD

MITEQ

- * C and X Band Downconverters
- C and Ku Band Upconverters
- Frequency Translators to permit use of existing C Band receiving equipment for new X Band applications
- Frequency Translators for upconverting existing C Band transmitters for Ku Band applications
- Loop Test Translators for C, X and Ku Band applications

***NEW COST EFFECTIVE DESIGNS TIMELY DELIVERY.**



Functional specifications include:

Conversion Loss (maximum) . . . . . . 12 dB without options, 15 dB with options

Amplitude Response . . . . $\pm$.25 dB maximum over any 40 MHz band, $\pm$1 dB maximum over 500 MHz

Frequency Stability . . . . 1 X 10^{-6}/day, 5 X 10^{-6}/month, 1 X 10^{-7}/day optional, phase locked to station, 5 MHz reference optional

Level Control . . . . 30 dB continuous standard, 60 dB optional

Intermodulation Distortion . . . . With two inband signals each at −13 dBm, the third order intermodulation products are less than −50 dBc

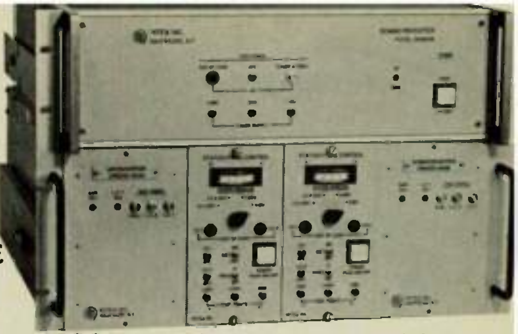
Input/Output Isolation. . . . . 60 dB maximum

Miteq Test Translators for satellite communications feature minimum amplitude and delay distortion, along with a low intermodulation distortion and a high frequency stability. Options include internal and external LO selection, internal and external reference selection, input filtering, an input PIN attenuator, input/output amplifiers, a waveguide input/output and a synthesized LO to 5 MHz reference.

Model	Input Frequency (GHz)	Output Frequency (GHz)	Return Loss (dB) In	Return Loss (dB) Out	LO Frequency (GHz)
DN-8011	5.925-6.425	3.7-4.2	23	23	2.225
UP-6-12	5.925-6.425	11.7-12.2	23	20	5.775
UP-6-14	5.925-6.425	14.0-14.5	23	20	8.075
UP-8011	3.7-4.2	5.925-6.425	23	23	2.225
UP-4-12	3.7-4.2	11.7-12.2	23	20	8.0
UP-4-14	3.7-4.2	14.0-14.5	23	20	10.3
DN-12-4	11.7-12.2	3.7-4.2	20	23	8.0
UP-12-14	11.7-12.2	14.0-14.5	20	20	2.3
DN-12-6	11.7-12.2	5.9-6.4	20	23	5.775
DN-14-4	14.0-14.5	3.7-4.2	20	23	10.3
DN-14-6	14.0-14.5	5.925-6.4	20	23	8.075
DN-14-12	14.0-14.5	11.7-12.2	20	20	2.3
DN-14-10	14.0-14.5	10.95-11.20	20	20	3.05
DN-14-11	14.0-14.5	11.45-11.70	20	20	2.55
DN-10-4	10.95-11.20	3.70-3.95	20	23	7.25
DN-11-4	11.45-11.70	3.95-4.20	20	23	7.50
DN-4245	(10.95-11.20) (11.45-11.70)	3.7-4.2	20	23	(7.25) (7.50)
DN-10-4WB	10.95-11.70	3.45-4.20	20	23	7.5
DN-10-4HO	10.95-11.70	4.20-3.45	20	23	15.15

MITEQ INC. 100 RICEFIELD LANE/ HAUPPAUGE, N.Y. 11787, (516) 543-8873, TWX: 510-226-3781

UPDATE



ELIMINATE NON-RECURRING ENGINEERING COSTS, ASSURE

Featuring solid state reliability, low phase noise, status monitors, summary alarm, ICSC compatibility, low intermodulation distortion, no spectral inversion and dual conversion with mechanically tunable or frequency agile phase lock oscillator.

	UPCONVERTERS		DOWNCONVERTERS	
	UP-8201	**UP-8205**	**DN-8001**	**DN-8012**
Type —	Dual conversion	Dual conversion	Dual conversion	Dual conversion
Tunability —	Second local oscillator only	Second local oscillator only	First local oscillator only	First local oscillator only
Frequency Sense —	No inversion	No inversion	No inversion	No inversion
Tuning Range —	500 MHz	500 MHz	500 MHz	250-1000 MHz
Input Characteristics:				
Frequency —	50-90 MHz 100-180 MHz optional	50-90 MHz 100-180 MHz optional	3.7-4.2 GHz	10.95-11.2 GHz 11.45-11.7 GHz 10.95-11.7 GHz 11.70-12.2 GHz
Input Level —	−20 dBm	−20 dBm	−20 dBm	−20 dBm
Input Impedance —	75 ohms	75 ohms	50 ohms	50 ohms
Return Loss —	26 dB minimum	26 dB minimum	23 dB minimum	20 dB minimum
Output Characteristics:				
Frequency —	5.925-6.425 GHz	14.0-14.5 GHz	50-90 MHz (100-180 MHz optional)	50-90 MHz (100-180 MHz optional)
Bandwidth —	40 MHz, 80 MHz optional	40 MHz, 80 MHz optional	40 MHz, 80 MHz optional	40 MHz, 80 MHz optional
Impedance —	50 ohms	50 ohms	75 ohms	75 ohms
Return Loss —	23 dB	20 dB	26 dB	26 dB
Level —	−5 dBm, up to +30 dBm with optional output amplifiers	−5 dBm, up to +20 dBm with optional output amplifiers	+10 dBm nominal, +20 dBm optional	+10 dBm nominal, +20 dBm optional
Transfer Characteristics:				
Noise Figure —	12 dB typical	12 dB typical	10 dB typical, 12 dB maximum, as low as 1.5 dB with optional amplifier	10 dB typical, 12 dB maximum, as low as 4 dB with optional amplifier
Gain —	15 dB nominal	15 dB nominal	30 dB	30 dB
Image Rejection —	80 dB minimum	80 dB minimum	80 dB minimum	80 dB minimum
Level Stability —	±.25 dB at constant temperature ±.5 dB 0-50°C ±1 dB with optional amplifier	±.25 dB at constant temperature ±.5 dB 0-50°C ±1 dB with optional amplifier	±.25 dB at constant temperature ±.5 dB 0-50°C ±1 dB with optional amplifier	±.25 dB at constant temperature ±.5 dB 0-50°C ±1 dB with optional amplifier
Frequency Response —	40 MHz at .5 dB 36 MHz at .4 dB 20 MHz at .2 dB	40 MHz at .5 dB 36 MHz at .4 dB 20 MHz at .2 dB	40 MHz at .5 dB 36 MHz at .4 dB 20 MHz at .2 dB	40 MHz at .5 dB 36 MHz at .4 dB 20 MHz at .2 dB
Group Delay — (±18 MHz)	Less than .03 ns/MHz linear Less than .01 ns/MHz² parabolic Less than 1 ns peak-to-peak ripple	Less than .03 ns/MHz linear Less than .01 ns/MHz² parabolic Less than 1 ns peak-to-peak ripple	Less than .03 ns/MHz linear Less than .01 ns/MHz² parabolic Less than 1 ns peak-to-peak ripple	Less than .03 ns/MHz linear Less than .01 ns/MHz² parabolic Less than 1 ns peak-to-peak ripple
Intermodulation Distortion — (third order)	At −20 dBm output 50 dBc	At −20 dBm output 50 dBc	With two −40 dBm input signals 60 dBc	With two −40 dBm input signals 60 dBc
AM/PM Conversion —	<.1°/dB to −5 dBm	<.1°/dB to −5 dBm	.1°/dB to +5 dBm output	.1°/dB to +5 dBm output
Gain Slope —	<.02 dB/MHz maximum	<.02 dB/MHz maximum	<.02 dB/MHz maximum	<.02 dB/MHz maximum
Spurious Outputs —	−90 dBm in band	−90 dBm in band	−65 dBc	−65 dBc
Gain Adjustment —	±3 dB nominal continuously variable	±3 dB nominal continuously variable	optional	optional

MILLIMETER-WAVE COMMUNICATIONS SATELLITES – PART II

The final portion of this article deals with the technologies critical to the implementation of millimeter-wave satellite systems and identifies the research and development required in some of those fields. The need for substantially more data describing the propagation characteristics of 40/50 GHz signals in a geosynchronous satellite application is emphasized. The alternatives for providing reliable spatial diversity are discussed. Since the useable data rates of the millimeter-wave system exceed the capabilities of interfaces to users, the need for economical new bulk data storage techniques is pointed out. Additional development of space switching equipment, receivers, transmitters and satellite antennas is recommended and other recent studies on the subject are mentioned.

SEMICONDUCTOR LIGHT SOURCES

Beginning with a general comparison between the light emitting diode (LED) and the injection laser light sources for fiber optic communications applications, the article proceeds to a detailed description of one of each type of device. The LED chosen is the etched-well design, applicable to systems up to several kilometers long operating at moderately high data rates. Fabrication of the device is described in detail; its characteristics and special features are discussed. The single mode injection laser diode is used to represent that family of devices and its construction and application are similarly discussed. Criteria governing the selection of one or the other device for a given system are also provided.

18 GHz PASSIVE REPEATERS

Considerations for the design and installation of repeater stations in an 18 GHz terrestrial system are discussed. Free space and atmospheric path losses are defined. Rainfall attenuation data is shown. The use of passive reflectors for obstructed paths is discussed and a table of passive reflector gain data is provided. Finally, calculations for a number of sample link legs are shown.

MILLIMETER-WAVE COUPLED LINE FILTERS

Printed circuit media hold the promise of reducing the cost of millimeter-wave system components and, further, offer an opportunity for size reduction through component integration. Suspended substrate and microstrip are candidates for such applications. The article describes the design of parallel coupled filters and diplexers in both media. A practical transition from waveguide to suspended substrate is illustrated. A step-by-step procedure for developing a four-section coupled line filter from a lowpass prototype demonstrates the approach taken. Experimental results for a number of K_a-band filters are shown. Characteristics of a diplexer made up of two of the filters are also shown.

L-BAND COOLED GaAs FET AMPLIFIER

The development of an L-Band GaAs FET amplifier designed for low-noise characteristics at cryogenic temperatures is described. The two-stage amplifier cooled to 18°K in a closed cycle helium refrigerator achieves a noise temperature of less than 20°K from 1.2 to 1.5 GHz. A number of transistors were evaluated for the design and some are recommended for cryogenic applications. Four of the amplifiers are used routinely in a radio astronomy application at the University of California Hat Creek Observatory.

Sum Up

Howard Ellowitz

Workshops & Courses

PRINCIPLES OF MODERN RADAR SHORT COURSE

Sponsor: Georgia Institute of Technology
Date: November 3-7, 1980
Site: GIT, Engineering Experiment Station
Subjects: Radar systems analysis, synthesis and evaluation; plus demonstrations.
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Low-Profile GaAs FET Amplifiers

CIRCLE 12 ON READER SERVICE CARD

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Dale Peterson *is Executive Vice President and Chief Operating Officer for Frequency Sources, Inc. He has been a part of the microwave component business since 1959 when he was building voltage tunable magnetrons and reflex klystrons at Eitel-McCullough. After the purchase of Eimac by Varian Associates, Mr. Peterson held various engineering and management positions in the microwave tube and solid-state source areas until he joined Frequency Sources in 1973. He holds patents on microwave tubes and RF circuits for microwave tubes.*

Microwave Components Make or Buy?

DALE L. PETERSON
Frequency Sources, Inc.
Santa Clara, CA

In the last twenty plus years I have heard almost all the reasons (OEM's) equipment manufacturers have given for wanting to make microwave components in house.

- We have to be cost effective!
- We can't be dependent on a small supplier for our critical needs!
- We must have vertical integration to be competitive!
- We need technical and schedule control over critical components!
- Our production is so large you can't meet our needs.

With the above "reasons" in the minds of the OEM's, how does the microwave component house respond? The proper response is simple. *Be well enough focused and managed to solve the customer's microwave component problems more efficiently and better than he can.*

The average system engineer does not really want to be in the component business, no matter what he says. He wants the component problem(s) solved so that his resources can be focused on system technology.

The component supplier is and must be a service to the equipment manufacturer. As long as it is a service function, done professionally and with dependability, it will be a service in demand. If we in the microwave component business fail our customer then we deserve to lose. When we are not dependable, we are not satisfying a need but creating a problem. Who in the systems business needs another problem?

System houses almost always have a large number of talented microwave engineers capable of doing an excellent job of microwave component selection, and, if necessary, design. The missing ingredient that precludes the system manufacturer from being the best microwave component source is shared experience and focus. The major factor in a make or buy decision should be the relative experience between the equipment manufacturer and the component supplier since that impacts both technical and cost performance. The microwave component supplier that makes his advantages visible can respond to the arguments for in-house production given by his system house customer and come out a winner. The following paragraphs discuss these arguments.

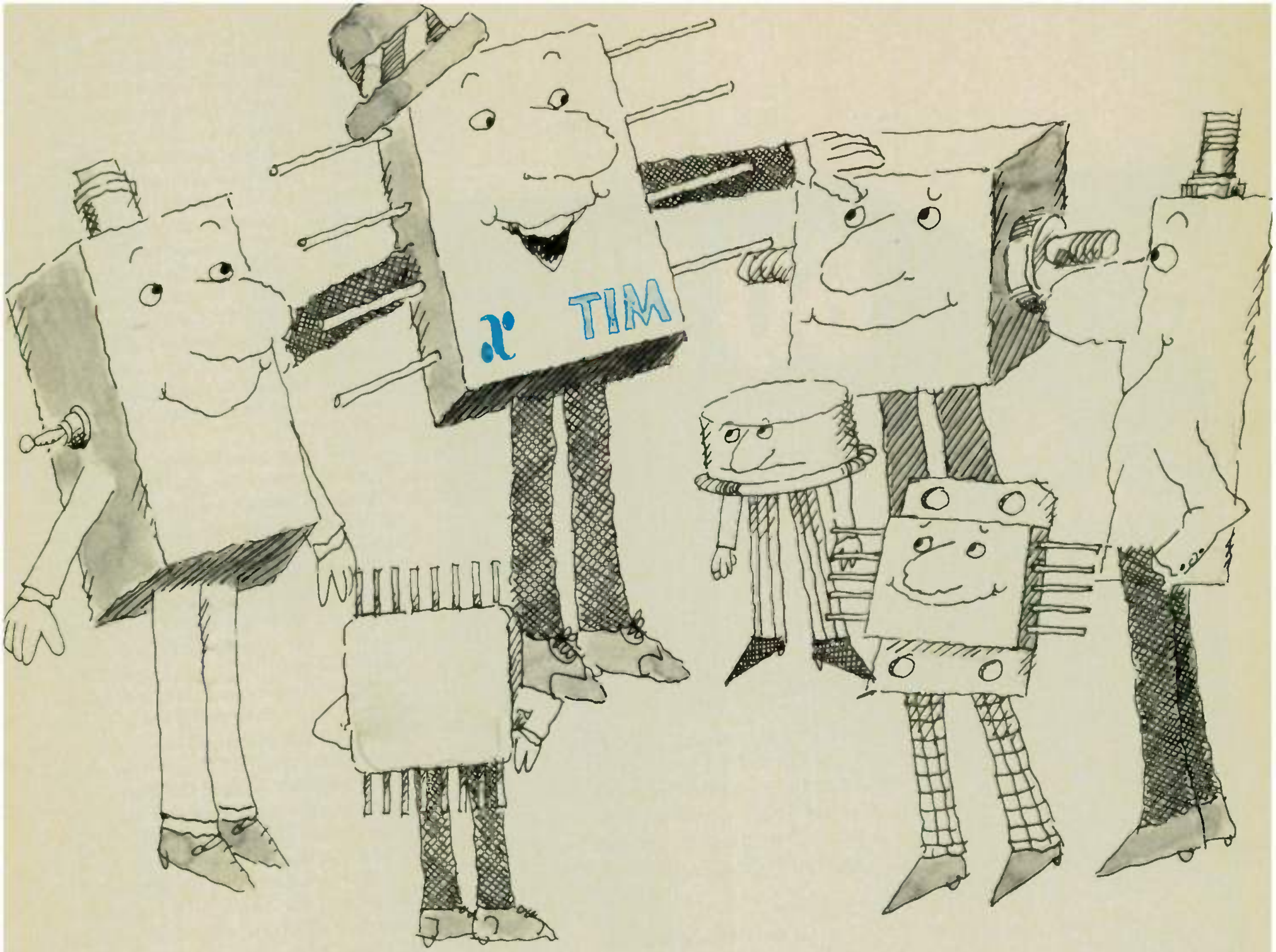
"We have to be cost effective." The equipment manufacturer that builds its own components will in all probability be limited by market factors to servicing only its own needs. Market share and product focus advantages are required to be cost effective. If the microwave component manufacturer cannot offer his larger product and manufacturing base, (i.e. all the equipment manufacturers who use his components) as sufficient competitive advantage then he may not deserve to win.

The sales leader in any business has a major competitive advantage. The cost advantages of volume are automatic but only if managed. The specific cost reduction drivers of volume are: learning curve effect, scale effect, and investment for cost reduction.

The component manufacturer will be a sales leader in respect to the equipment manufacturer by virtue of his larger selling base and large volume of focused production.

"We can't depend on such a small supplier for our production program." Is it better to have 20% of a component manufacturer's production and engineering tied up on your job, with

(continued on page 18)



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(from page 16) **COMPONENTS**

80% in reserve or have a 100% of a small group in the equipment manufacturer's shop working on this job only? I think the answer is obvious. How can an internal group start up and gain the cost and quality advantages of extensive learning curve experience without the benefit of continuing volume production and program continuity. The answer has to be that the well managed leader in production volume will have the most accumulated experience and the lowest cost. Product focus and competitive pressures produce survivors who force costs down as volume expands. Isn't this what the equipment supplier really wants?

"We must have vertical integration." The best answer for that question is as a question. Why? In the past few years an increasing number of component manufacturers have gone into the business of building equipment. Not to be outdone, several equipment manufacturers are getting into the components business. My answer is still a question. Why? Sometimes it makes good economical sense to vertically integrate, other times it may make sense to do what you know how to do and let others do what they know how to do. When business is good the decision is easy but when business gets tight we all get smart in areas far from our specific expertise.

"Our production is so large you can't meet our needs." The component manufacturer should be the best at manufacturing high quantities and at high production rates. That is what the "focused factory" concept is all about. The component manufacturer is to begin with a focused business. A "focused factory" can produce high volume products much more cheaply than a plant designed for flexibility. Why then can't the component manufacturer supply the "production needs" of the equipment manufacturer. Is it possible that the real statement being made by the system manufacturer is how can I avoid letting this large order out of my shop.

"We need technical and schedule control over critical components." Is it not true that systems engineers and subcontracts people can evaluate and control vendor status if they are serious about it and they select responsible component houses? This is normally a statement used to avoid good management practices regardless as to whether things are made or bought.

In summary, I believe microwave components should be built where the most value in terms of cost, quality, product innovation and delivery can be provided. This should be at well focused and managed component houses.

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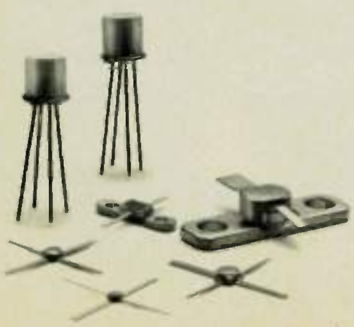

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AT-1845	2.5dB	14dB	1.0GHz
AT-1825	3.0dB	13dB	1.0GHz
GaAs FETS UP TO 12GHz			
AT-8110 Chip AT-8111	1.3dB(TYP)	12dB	4GHz
AT-8060 Chip AT-8061	2.8dB(TYP) 2.5dB(TYP)	8dB 9dB	12GHz 12GHz

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PART II

Solid-State Amplifiers as TWT Substitutes

FERDO IVANEK
Harris Corporation
Farinon Electric Operation
San Carlos, CA

Part I of this Special Report (pp. 65-68, *Microwave Journal* September, 1980) contained background information on the ICC'80 Panel Session* "Solid-State Amplifiers as TWT Substitutes," and individual presentations by three panel members: R. L. Metivier, P. G. Debois, and J. Gewartowski.

Part II contains the presentations by the other three panel members, C. C. Hsieh, I. Haga and K. Morita.

- *C. C. Hsieh* presents the advantages of silicon bipolar power transistors for Class C operation below 5 GHz, and Class A operation below 4 GHz.
- *I. Haga* summarizes the results of several GaAs FET power amplifier implementations in the various bands between 4 and 14 GHz, elaborates on the reliability aspects and assesses the possible advances over the next five years.
- *K. Morita* presents data on the usage of TWT and GaAs FET power amplifiers in Japan, defines the relative power requirements for a digital radio application, and points out the heat-sink size disadvantage of present-day GaAs FET amplifiers.

* Panel Session held at the 1980 International Conference on Communications, Seattle, Washington.

This Special Report concludes with summaries of the short presentations from the floor, and notes on the panel discussion with audience participation — both provided by the moderator.

BIPOLAR POWER TRANSISTOR AMPLIFIERS AS TWT SUBSTITUTES

CHI C. HSIEH
Farinon Electric Operation
Harris Corporation
San Carlos, CA

Power GaAs FET amplifiers have been the focus of all microwave development activities for the past five years; very few publications have dealt with the development of high power bipolar transistor amplifiers.[1,2] However, based on the commercially available data the bipolar power transistor amplifiers offer superior Class-C performance below 5 GHz, and below 4 GHz as linear power amplifiers. Included in the following two paragraphs are some up-to-date data of bipolar power transistors both in Class-C and Class-A operations. Many TWT amplifiers in the 2-3 GHz frequency range have been replaced by bipolar amplifiers over the past decade. (10-20 W output power), whereas only limited numbers of GaAs FET amplifiers

TABLE I

Some commercially available Class-C Bipolar Power Transistors (internally matched)

Freq. Range (GHz)	P_{sat} (W)	G (dB)	η (%)	Manufacturer
1.7 - 2.0	20	6	35	MSC
2.0 - 2.3	20	6	35	TRW, MSC, CTC,
2.4 - 2.7	20	6	30	MSC
2.7 - 3.2	40	8	42	TRW
3.0	15	5	38	TOSHIBA
3.7 - 4.2	10	7	35	TRW
4.4 - 5.0	5	7	30	TRW

TABLE II

Some Class-C TWT Replacement Bipolar Power Amplifiers for the OEM market

Freq. Range (GHz)	P_{sat} (W)	Gain (dB)	Manufacturer
2.3 - 2.7	20	40	MPD
3.1 - 3.5	20	40	MPD
3.7 - 4.2	10	40	MPD, MSC,
2.2 - 2.3	45	40	MSC
1.6 - 1.7	45	27	MSC
1.75 - 1.85	2.5 kW	66	MPD
1.7 - 2.0	500 W	60	MPD
2.0 - 2.3	500 W	60	MPD

are presently being used, while at lower output power.

CLASS-C BIPOLAR POWER TRANSISTOR AMPLIFIERS

At frequencies below 5 GHz, the bipolar power transistors offer high efficiency, high output power and high reliability. Also, as of today the bipolar power transistors still cost less dollars per watt, as compared to power GaAs FET's. Most of these devices are used in common-base configuration; devices marketed by TRW, MSC and CTC are mostly internally matched in the package to a specific bandwidth, which simplifies the amplifier design. Some of the higher performance devices are listed in Table I.[3,4,5]

Several power amplifier manufacturers are currently delivering bipolar power amplifiers as TWT replacements, although most radio system manufacturers have in-house design capability. Some of the bipolar power amplifiers which are currently produced for the OEM market are listed in Table II.[6]

CLASS-A BIPOLAR POWER TRANSISTOR AMPLIFIERS

For linear amplifications below 4 GHz, the bipolar power transistor amplifiers offer a distinct advantage in linearity performance as compared to the GaAs FET power amplifiers. A number of publications have reported the "ill-behavior" of the 3rd-order intermodulation distortion of GaAs FET power amplifiers.[8,9,10,11] Figure 1 shows a comparison of the 3rd-order IMD between a bipolar transistor amplifier and a power GaAs FET amplifier at 2 GHz. The curves clearly show that the 3rd-order intercept is meaningless when used to indicate the linearity of a

TABLE III

Linear Bipolar Power Transistors

Freq. Range (GHz)	P_o at 1 dB Comp. (W)	G (dB)	Manufacturer
2.3	5.0	7.0	FUJITSU
2.3	6.0	6.5	CTC
2.2	2.5	9.0	FUJITSU

TABLE IV

Linear Bipolar Power Amplifiers

Freq. Range (GHz)	P_o @ 1 dB Comp. (W)	Intercept point (dBm)	Gain (dB)	Manufacturer
1.0 - 2.0	9	+49.5	43	MPD
1.7 - 2.4	2	+43	37	MPD
2.1 - 2.2	10	+50	60	FARINON
2.2 - 2.3	10	+50	40	MPD

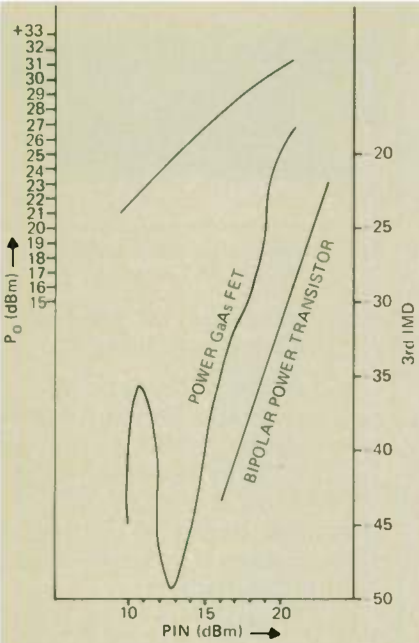

Fig. 1 Comparative examples of 3rd-order IMD in power amplifiers using bipolar silicon transistors and GaAs FET's, respectively.

GaAs FET power amplifier. It is also a difficult task to apply the IF pre-distortion technique to linearize this type of device, for it is not a simple 3rd-order device.[12] The bipolar power transistor amplifiers, when operated in Common-Emitter configuration and Class-A mode offer "well-behaved" 3rd-order IMD characteristics; the IMD level at any output power level can be estimated from the 3rd-order intercept point by using the well known 2:1 relationship.

Three of the highest performance linear bipolar power transistors are listed in Table III.[7]

The linear bipolar power amplifiers listed in Table IV are currently in production and are being used in place of traveling-wave tube amplifiers.

CONCLUSION

In summary, the bipolar power transistor amplifier represents a mature technology. The saturated mode amplifier can readily replace TWT amplifiers below 5 GHz. The bipolar power transistor amplifiers offer a "well-behaved" 3rd-order linearity char- ... radio systems shows superior performance as compared to GaAs FET Power Amplifiers at frequencies below 4 GHz.

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GaAs FET POWER AMPLIFIERS AS A SUBSTITUTE FOR THE TWT

ISAO HAGA
Microwave & Satellite Comm. Div.
Nippon Electric Company, Ltd.
Yokohama, Japan

NEC has been producing various kinds of solid-state power amplifiers using Si bipolar transistors, Gunn diodes or IMPATT diodes which have been used in the various communication equipments. Several thousand solid-state microwave power amplifiers employing Si bipolar transistors, Gunn diodes or IMPATT diodes have been produced during 15 years, and it is confirmed that these amplifiers have provided reliable service in the field. Today, the most interesting and excellent solid-state active semiconductor for microwave power amplifier is the GaAs FET.

RECENT ACHIEVEMENTS

The mass production of the power GaAs FET amplifier was begun at NEC in Oct. 1977, and about 1310 units, such 5 W/10 W in 6 GHz, 5 W in 8 Hz, 2.5 W in 11 GHz and 2 W in 14 GHz, have been manufactured during 2.5 years.

Figure 1 is the internal view of the 14 GHz band, 2 W GaAs FET amplifier. This amplifier consists of 8 stages. The final stage is a hybrid coupled balanced type amplifier. The 1st, 2nd and 3rd amplifier modules consist of 2 stage amplifiers respectively, while the 4th and 5th modules consist of single-stage amplifiers. Each unit amplifier module is provided with internal matching circuits.

GaAs FET power amplifiers provide the best dc-RF conversion efficiency, the lowest intermodulation distortion and the lowest AM-PM conversion characteristic compared with the other type solid-state amplifiers:

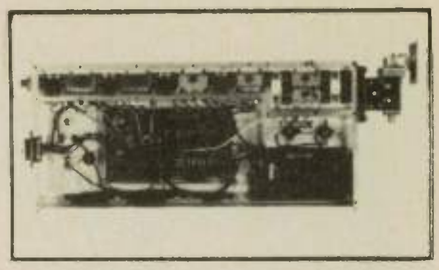

Fig. 1 Internal view of the 14 GHz band 2 W GaAs FET amplifier.

- The dc-RF conversion efficiency of the GaAs FET power amplifier with 38 dB in gain and 10 W in output power at 6 GHz band is more than 14%, and that of the 14 GHz band, 35 dB gain, 2 W amplifier is more than 10%.
- IM_3 of the 6 GHz band 38 dB gain, 10 W amplifier is less than 25 dB at 3 dB output back-off point, and that of the 14 GHz band, 35 dB gain, 2 W amplifier is less than 20 dB at 3 dB output back-off point. We have not observed "Dip Phenomenon" of the IM_3

characteristics on the mass production type GaAs FET amplifier employing the graded-recess type GaAs FET.

- The AM-PM conversion coefficient of the 6 GHz band 38 dB gain, 10 W amplifier is less than 1.5°/dB, and that of the 14 GHz band 35 dB, 2 W amplifier is less than 2°/dB.
- Small signal gain deviation due to ambient temperature variation of the high gain GaAs FET amplifier is about ±2 dB/25°C.

RELIABILITY

The high power GaAs is operated at a large voltage amplitude. Therefore, a high temperature storage test and a dc burn-in test alone are not sufficient to estimate the reliability of a high power GaAs FET.

Since reliability of a high power GaAs FET has to be estimated under the worst operating conditions, NEC conducted high temperature operation test with an RF signal drive using 60 NE 868 series GaAs FETs. Test conditions were 135°C in channel temperature, 8 V in drain-source voltage and saturation in output power. As a result, we can confirm that the MTBF is better than 450,000 hours under operation at a channel temperature of 130°C and saturation output power.

There are now more than 2,250 power GaAs FET's which have operated for more than 4600 hours in the field, total device-hours already exceed 25,000,000 hours. During this time, only four failures have occurred. Therefore, the MTBF of the GaAs FET in the field is 6,250,000 hours, it corresponds to 160 fit.

When the power GaAs FET amplifier is used as a TWT substitute, seven or eight GaAs FET's are used with a total gain of about 35 dB. Therefore, the MTBF of the power GaAs FET amplifier is 900,000 to 700,000 hours based on the field data, so we can assume that the MTBF of the power GaAs FET amplifier will be at least ten times better than the TWT.

To guarantee high reliability of the power GaAs FET amplifier, it is necessary to perform systematic screening tests of the channel temperature, the dc bias conditions, the gate leakage current, and the aging under saturated output power.

PROJECTIONS

Figure 2 shows the realized output power characteristics by NEC today, and the expected output power characteristics of the GaAs FET amplifier in the near future. We have already realized 12 W at C band, 3 W at 11 GHz band and 2 W at 14 GHz band.

The output power limitation of the GaAs FET power amplifier is caused by lower gate breakdown voltage and higher channel temperature. The lower gate breakdown voltage causes impedance imbalance between individ-

(continued on page 28)

CIRCLE 24 ON READER SERVICE CARD

ual cells or chips, and thereby causes gain expansion.

Increasing gate breakdown voltage is necessary for increasing the output power, for eliminating expansion, and for higher efficiency amplification in Class B or Class C operation.

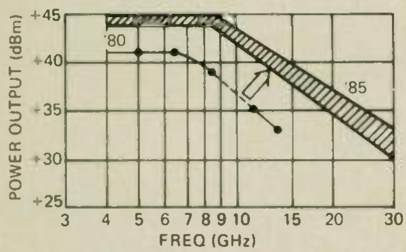

Fig. 2 Present and projected power output characteristics of GaAs FET power amplifiers. Means for obtaining the projected improvement are:
(a) development of the new internal matching technique
(b) development of the new heat radiation technique
(c) increase gate breakdown voltage.

To increase the output power of the C-band and X-band GaAs FET amplifier, it is necessary to develop new heat reduction techniques and to increase the gate breakdown voltage. Also, to increase the output power of the K_u-band and K-band GaAs FET amplifier, it is necessary to develop the new internal matching technique and to develop the new monolithic IC devices.

As a result of many development efforts, we can now anticipate that in another half decade it will be possible to produce GaAs FET amplifiers with output power of about 25 W at 6 GHz, 10 W at 11 GHz and 5 W at 20 GHz.

GaAs FET AMPLIFIERS IN MICROWAVE SYSTEMS

KOZO MORITA
Nippon T&T Public Corp.
Yokosuka, Japan

STATE-OF-THE-ART GaAs FET's IN JAPAN

Progress in the development of GaAs FET's has been so remarkable during the past five years that medium power output TWT's have already been replaced by GaAs FET amplifiers.

According to the latest report on FET's in the experimental stage, 25 W power output at 6 GHz with 24% power-added efficiency and 700 MHz bandwidth is attainable. At 10 GHz, a flip-chip FET with 28.8 mm total gate width has achieved 11.5 W power output with 800 MHz bandwidth and 14% power-added efficiency.

Recent reports on GaAs FET amplifiers describe a seven-stage amplifier with 12.5 W saturation power, 40 dB gain, 500 MHz bandwidth, 1.6°/dB maximum AM/PM conversion and 17% overall efficiency in the 6 GHz band. In the 12 GHz band, a 3.5 W amplifier with a 28 dB gain and 1 GHz bandwidth was obtained using recessed-gate FET's. A five-stage amplifier using flip-chip FET's has also been reported to have 2 W saturation power with 29 dB linear gain in the same band.

APPLICATIONS IN MICROWAVE COMMUNICATION SYSTEMS

FET amplifiers as TWT substitutes have been applied in microwave systems since 1977 as shown in **Figure 1**. More than a thousand FM transmitters with 0.5-5 W power output in the 6-8 GHz bands have been manufactured and a 10 W transmitter in the 6 GHz band became available recently.

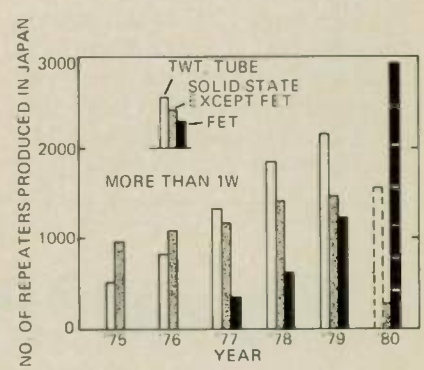

Fig. 1 GaAs FET amplifiers progress in Japan.

(continued on page 30)

THE CASE OF THE DISAPPEARING LINEARIZERS

The baffling disappearance of linearizers from varactor-tuned systems can at last be explained. Our new GaAs Hyperabrupt Tuning Varactors are so incredibly linear – as low as 0.1% – that they need no expensive linearizing circuitry whatsoever. In fact, a simple process of deduction (after a close examination of our revealing linearity curves) will prove that our GaAs Hyperabrupts are more inherently linear than Silicon Abrupts with their linearizers.

So, be baffled no more. The solution is elementary. GaAs Hyperabrupt Varactors by Microwave Associates. A complete dossier is available for your examination. Call or write the Semiconductor Sales Department, Microwave Associates, Inc., Burlington, MA 01803 (617) 272-3000.

CIRCLE 26 ON READER SERVICE CARD

(from page 30) **AMPLIFIERS**

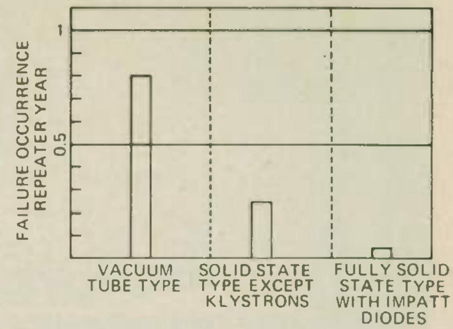

Fig. 3 Failure occurrence probability of 11 GHz radio repeaters.

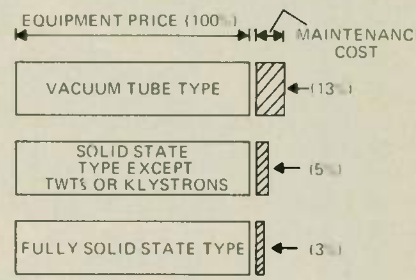

Fig. 4 Maintenance cost.

Akira Hashimoto, of the NTT Engineering Bureau, Tokyo, Japan, gave a reliability and maintenance cost comparison of radio-relay repeaters operating in their network, which convincingly illustrates the benefits of the transition from all tube to all solid-state equipment. His data are reproduced in Figures 2 through 4. Note that this survey does not include all solid-state equipment using GaAs FET's.

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*Unable to attend the conference, B. Maalsnes asked F. Ivanek to make the presentation using the prepared vugraphs.

**Presentation made by J. Gewartowski, substituting for F. Paik who was unable to attend.

NOTES ON THE DISCUSSION

The discussion was most productive, as could be expected from the presentations that provided so much up-to-date factual information, and blended so well into a coherent picture.

Since no deliberate attempt was made to reach consensus, everybody was left to reach his own conclusions. The following notes represent the moderator's personal views of what transpired from the presentations and the discussion.

The development of power GaAs FET's broadened and intensified the solid-state competition

(continued on page 98)

Around the Circuit

PERSONNEL

Recent promotions include **Warren Gould**, who moves from CATV Product Line Manager to Operations Manager for TRW RF Semiconductors, and at HRB-Singer, Inc., **George N. Doliana** was named V.P., Programs Development and **Richard L. Wales**, V.P., Programs Management. Mr. Doliana leaves his post as Director, Command and Information Programs and Mr. Wales the job of Director, Special Programs. . .In a senior management reorganization at Eaton Corp.'s Electronic Instrumentation Div. (EID), EID's new Operations Managers are **Francis X. Geissler**, former General Mgr. of the Cramer Div. of Conrac Corp. and **Howard Cooper**, an Eaton Gen. Mgr. at the City of Industry, CA facility. And at Eaton's AIL Division, Dr. Curtis Schleher has been appointed Dir. of Planning, a new post. . .**Ed V. Roos** was appointed Mgr. of Crystal Technology's Optical Device Dept. . . **Ed Mendel**, Pacific Measurements, Inc.'s Marketing Manager, assumes the newly created post of V.P. of Marketing. . .E-Systems, Inc. has promoted **Billy C. Hooker** to V.P. - Advanced Systems, **Frank E. Stapp**, V.P.-Electronic Warfare, and Dr. **Phil H. Rogers**, V.P.-Electronic Systems Software. . .**Robert Trouard** was named Mgr. of the Equipment Engineering Dept. in the Strategic Reconnaissance Organization at the Western Div. of GTE Sylvania Systems Group. . .**Lyman "Rusty" S. de Camp** was named Senior Systems Engineer for California Microwave's Satellite Communication Div. . .At the Pulsecom Div. of Harvey Hubbell, Inc., **Steve Petty** was named Regional Sales Manager for the Pacific Northwest, West Coast and South West. . .**Alan H. Rice** became Manager of the Thin Film Div. of Tek-wave, Inc., the Frequency Electronics, Inc. subsidiary.

CONTRACTS

American Electronics Labs (AEL) received a $7.3M series of orders from the Ft. Monmouth US Army operation to install EW suite components in aircraft. AEL also was awarded a $2.5M contract from the AF Systems Command, ASD for ECM equipment for the EF-111A. . . **Westinghouse Electric Corp.**'s Defense Group was granted a $25M USAF contract for Full Scale Development of modifications to the F-16's AN/APG-66 fire control radar. . .Naval Electronic Systems Command awarded **GE's Electronic Systems Div.** a $134M contract for AN/TPS-59 radar systems. . .**Scientific-Atlanta, Inc.** signed an agreement to furnish its 3-meter antenna to **California Microwave, Inc.** for the latter's satellite ground stations. Some 400 antennas will be delivered during 1980, and 900 by the end of 1981. . .NRL has developed a process to grow semi-insulating GaAs for direction ion implantation. These new crystal growth methods have been accepted by the Naval Air Systems Command and contracted to **M/A, Inc.** for manufacturing technology. . . **GTE** received six contracts valued at $5.2M for terrestrial and on-board satellite communication systems for various European domestic satellite systems.

INDUSTRY NEWS

Georgia Institute of Technology's Radar and Instrumentation Lab has been active in the development of jeeps, trucks, tanks, missiles, ships as well as aircraft which are radar "proof." Also, **GIT's Engineering Experiment Station** engineers are conducting a study to improve the AN/APS-94F radar system on the OV-ID Mohawk aircraft for the US Army. . .On Aug. 15, 1980, **Loral Corp.** completed the merger of **Frequency Sources, Inc.** into a wholly-owned subsidiary of Loral. In a transaction valued at $55M, Loral exchanged .75 shares of common stock for each of FSI's 2.3M shares. . .Another merger on Sept. 3, 1980, made **MBAssociates** (MBA) into a wholly owned subsidiary of **Tracor**. Tracor issued .3125 of one share of its common stock for each of MBA's approximately 1.2M shares outstanding. . .**Metex Corp.**'s Electronic Products Div., Edison, NJ., announced a 21% increase in prices of their entire line of EMI/RFI shielding products. . .Expansion plans of **Itek's Applied Technology Div.** call for a new 78,444 sq. ft. building at Sunnyvale, CA. By March 1981, **E-Systems, Inc.** will complete a 280,000-sq. ft. expansion of the Garland, TX facility. . .**M/A COM, Inc.** acquired **LINKABIT Corp.** LINKABIT shareholders will receive 1.0575M shares (after exercise of outstanding options) of M/A-COM common stock. M/A-COM, Inc. also acquired **Omni-Spectra, Inc.** following approval of the transaction by stockholders of both companies on Aug. 25, 1980. Under the agreement, stockholders of Omni-Spectra, including certain option holders, will receive a maximum of 838,244 shares of M/A-COM common stock and Omni-Spectra will operate as a wholly-owned M/A-COM subsidiary under its present management. Also, M/A-COM, Inc. announced that it will build its own nationwide digital sat. com. network connecting the Burlington, MA operations with its operating companies in MD, CA and NC. . .**Aydin Corp.** formed an operation to develop and market products in digital telecom., specifically in Time Div. Multiple Access systems for sat. com. Dr. Joseph Deal has joined Aydin to head new product development for this new operation, based initially at Aydin Microwave Div. in San Jose, CA.

FINANCIAL NEWS

Raymond Industries, Inc. reported half-year results for the period ended June 30, 1980 of sales of $18.6M, net earnings of $3.8M or $3.37 per share. This compares with 1979 net sales of $16.5M, net earnings of $595K or 73¢ per share. . .**Adams Russell** reported third quarter net sales of $9.2M, net income of $714K or 39¢ per share for the period ended June 29, 1980. During the same 1979 quarter, net sales were $7.38M, net income was $545K and earnings per share were 31¢. . . **Microwave Power Devices, Inc.** announced net sales for the fourth quarter of the year ended July 31, 1980 of $1.8M (unaudited). This compares with 1979 quarterly results of $1.41M. **Sealectro Corp.** reported on June 17, 1980, second quarter net income of $609K, or 41¢ a share on sales of $10.9M. For the comparable 1979 quarter, net income was $574K, or 39¢ per share and sales were $8.7M. . .**Hewlett-Packard Co.** reported third quarter sales of $810M, net earnings of $70M or $1.15 per share for the period ended July 31, 1980. This compares with 1979 quarterly results of $620M, net earnings of $52M or 89¢ per share. . .**Harris Corp.** reported year-end sales of $1.3B, net income of $79.7M or $2.63 per share for the period ended August 21, 1980. This compares with 1979 annual results of $1.075B sales, $68.8 net income or $2.32 per share.

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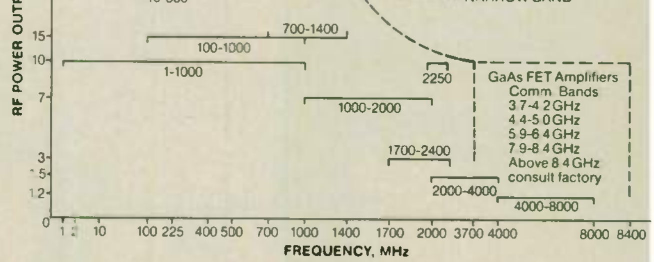
You can now choose from 225 standard MPD models, vailable in module or rack-mounted cabinet configura- ons for systems applications, as well as self-contained ıstruments for laboratory use. Ultra-broadband frequency anges from 1-1000MHz up to 7900-8400MHz, including ur newest high power model with 200 watts saturated power ating at 500-1000MHz.

These Class A amplifiers are particularly recommended ɔr applications requiring exceptional linearity and wide ynamic range. They combine low noise figure with high ower output to yield low distortion amplification of both ingle and multi-carrier signals with any form of modulation.)ther standard features of most models include: high fficiency; high gain; built-in protection against DC input eversal, thermal overload and infinite load VSWR; raceful degradation; drift-free output; field-replaceable ıodular construction.

In addition, MPD manufactures and markets the wideband linear transistor power amplifier product line acquired from the Hughes Electron Dynamics Division.

Complete technical details and specifications are yours for the asking—then, get ready to move to the head of the class in solid state linear power amplifiers!



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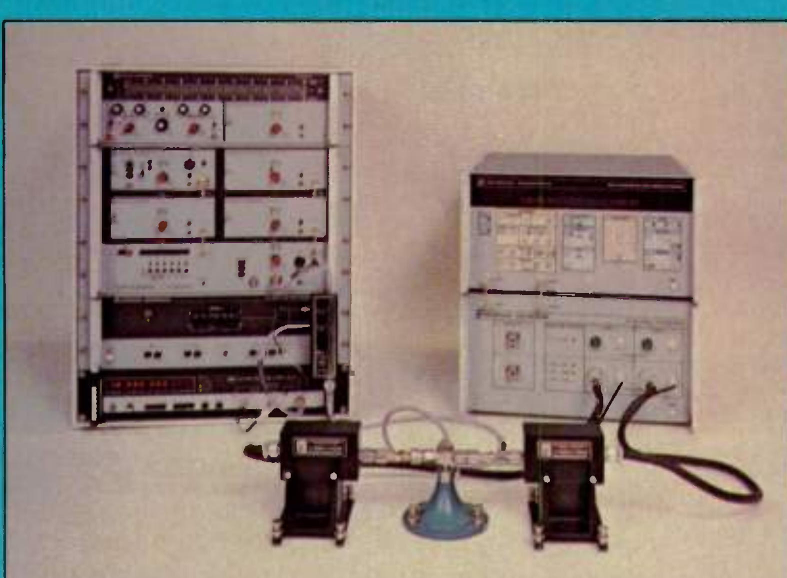
Programs, measurement results, and prompting messages are displayed on a 40-character flat gas-discharge display. Programs and measurement results can be printed externally via the RS-232 interface.



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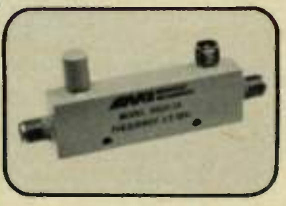
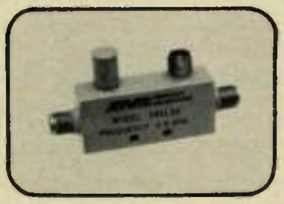
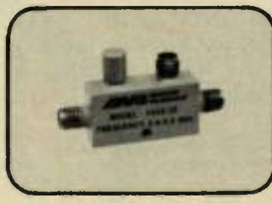
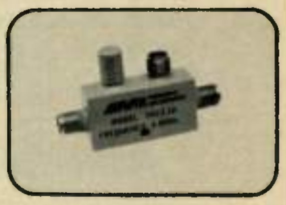

SPECIFICATIONS

MMI Model (1)	5010-6	5010-10	5010-20
Frequency Range (GHz)	ALL ARE 1-2		
Coupling in dB*	6 ± 1.00	10 ± 1.25	20 ± 1.25
Frequency Sensitivity in dB	± 0.60	BOTH ARE ± 0.75	
Directivity dB min.	BOTH ARE 25		27
VSWR max.	1.15	BOTH ARE 1.10	
Insertion Loss** dB max.	ALL ARE 0.20		

MMI Model (1)	5011-6	5011-10	5011-20
Frequency Range (GHz)	ALL ARE 2-4		
Coupling in dB*	6 ± 1.00	10 ± 1.25	20 ± 1.25
Frequency Sensitivity in dB	± 0.60	BOTH ARE ± 0.75	
Directivity dB min.	ALL ARE 22		
VSWR max.	ALL ARE 1.15		
Insertion Loss** dB max.	ALL ARE 0.20		

MMI Model (1)	5012-6	5012-10	5012-20
Frequency Range (GHz)	ALL ARE 2.6-5.2		
Coupling in dB*	6 ± 1.00	10 ± 1.25	20 ± 1.25
Frequency Sensitivity in dB	± 0.60	BOTH ARE ± 0.75	
Directivity dB min.	18	BOTH ARE 20	
VSWR max.	ALL ARE 1.25		
Insertion Loss** dB max.	ALL ARE 0.25		

MMI Model (1)	5013-6	5013-10	5013-20
Frequency Range (GHz)	ALL ARE 4-8		
Coupling in dB*	6 ± 1.00	10 ± 1.25	20 ± 1.25
Frequency Sensitivity in dB	± 0.60	BOTH ARE ± 0.75	
Directivity dB min.	18	BOTH ARE 20	
VSWR max.	ALL ARE 1.25		
Insertion Loss** dB max.	ALL ARE 0.25		

DIMENSIONS
(See Engineer Drawing Above)

All dimensions are in inches

	5010		5011		5012		5013	
A	ALL ARE 0.51		ALL ARE 0.51		N/A		N/A	
B	ALL ARE 0.94		ALL ARE 0.34		N/A		N/A	
C	ALL ARE 1.95		ALL ARE 1.35		ALL ARE 1.15		ALL ARE 1.15	
D	ALL ARE 0.65		BOTH ARE 0.60	0.65	BOTH ARE 0.60	0.65	BOTH ARE 0.60	0.65
E	ALL ARE 0.30		ALL ARE 0.30		ALL ARE 0.30		ALL ARE 0.30	
F	ALL ARE 1.35		ALL ARE 0.75		ALL ARE 0.56		ALL ARE 0.61	
G	N/A		N/A		ALL ARE 0.57		ALL ARE 0.57	

Average Incident Power - 50 Watts *Also includes frequency sensitivity. **Excluding coupled Power.

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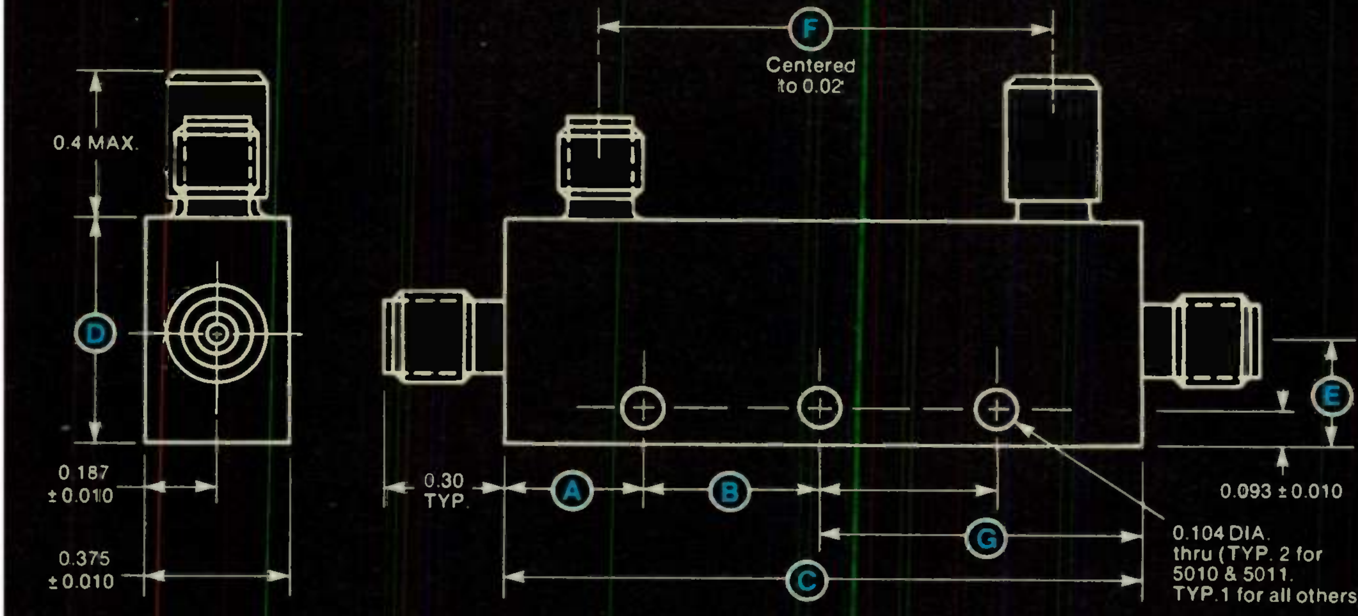

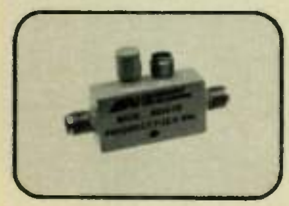

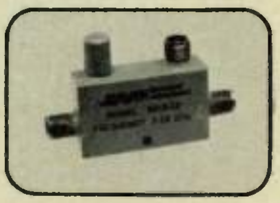

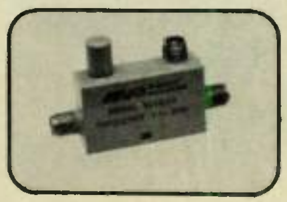

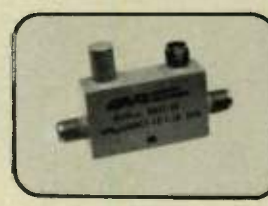

5014-6	5014-10	5014-20
ALL ARE 7-12.4		
6 ± 1.00	10 ± 1.25	20 ± 1.25
ALL ARE ± 0.50		
15	BOTH ARE 17	
ALL ARE 1.30		
ALL ARE 0.40		
N/A		
N/A		
ALL ARE 1.15		
BOTH ARE 0.60		0.65
ALL ARE 0.30		
ALL ARE 0.36		
ALL ARE 0.57		

5015-6	5015-10	5015-20
ALL ARE 7-18		
6 ± 1.00	10 ± 1.25	20 ± 1.25
± 0.60	BOTH ARE ± 0.75	
BOTH ARE 12		15
ALL ARE 1.40		
ALL ARE 0.50		
N/A		
N/A		
ALL ARE 1.25		
BOTH ARE 0.75		0.80
ALL ARE 0.30		
ALL ARE 0.77		
ALL ARE 0.63		

5016-6	5016-10	5016-20
ALL ARE 8-16		
6 ± 1.00	10 ± 1.50	20 ± 1.25
± 0.60	BOTH ARE ± 0.75	
BOTH ARE 12		15
ALL ARE 1.40		
ALL ARE 0.50		
N/A		
N/A		
ALL ARE 1.25		
BOTH ARE 0.75		0.80
ALL ARE 0.30		
ALL ARE 0.77		
ALL ARE 0.63		

5017-6	5017-10	5017-20
ALL ARE 12.4-18		
6 ± 1.00	10 ± 1.00	20 ± 1.00
ALL ARE ± 0.50		
ALL ARE 15		
ALL ARE 1.40		
ALL ARE 0.40		
N/A		
N/A		
ALL ARE 1.25		
BOTH ARE 0.75		0.80
ALL ARE 0.30		
ALL ARE 0.77		
ALL ARE 0.63		

(1) 5000 Series Couplers available with removeable termination. Add suffix "R" to model number.

MIDWEST MICROWAVE

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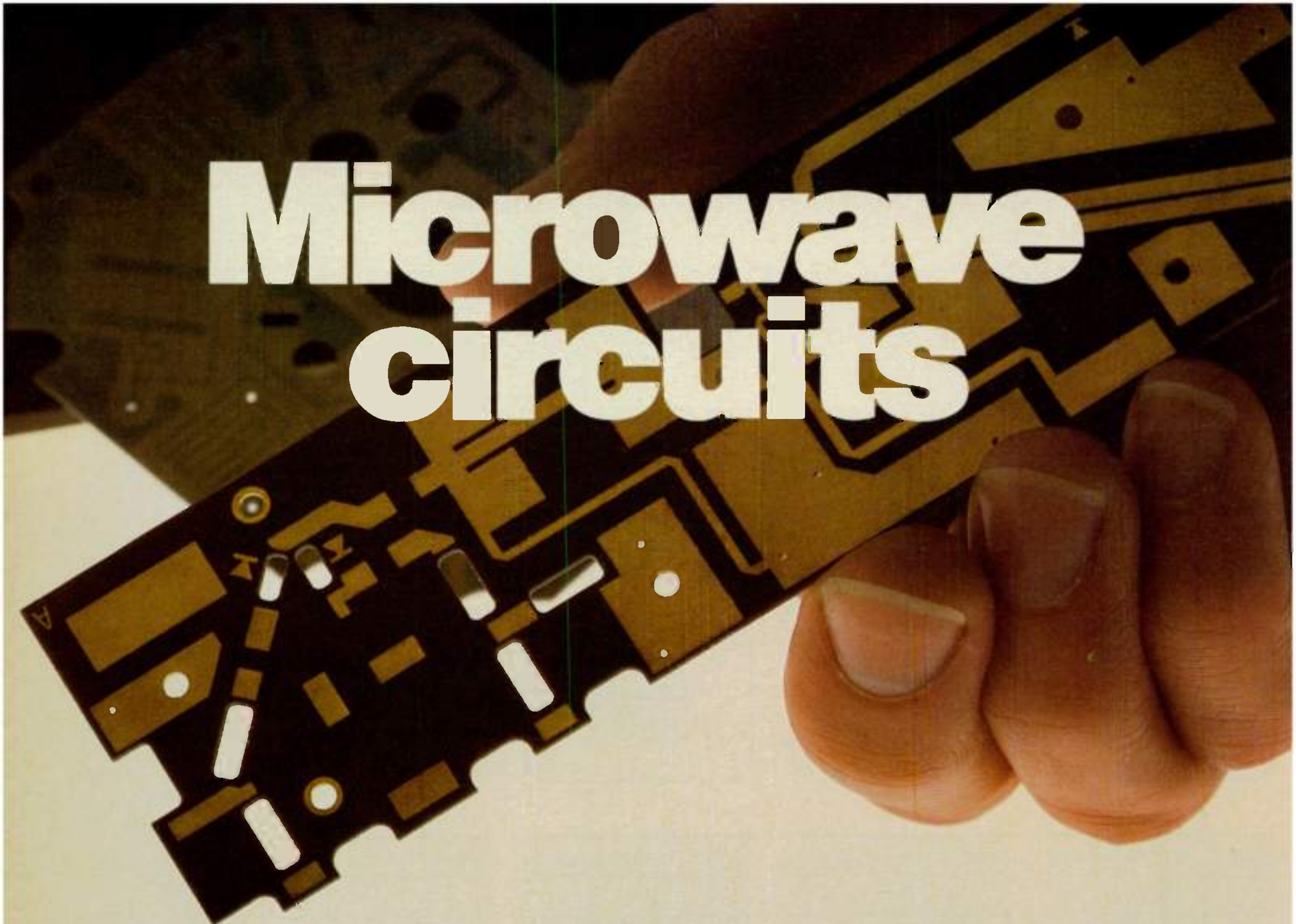

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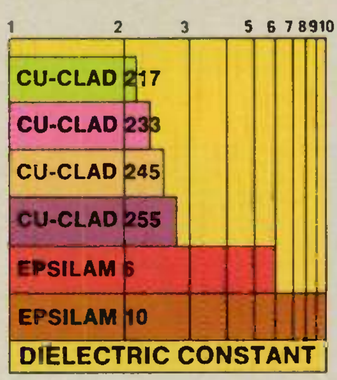

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System Analysis for Millimeter-Wave Communication Satellites

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INTRODUCTION

Part I of this article considered the various critical technologies necessary for the development of millimeter-wave communication satellites. It treated cost, weight and performance models for subsystems, methodology for design trade-off studies and conceptual designs for high frequency communication satellite systems.

TECHNOLOGY ASSESSMENT

One of the objectives of the study has been the identification of technologies critical to implementation of millimeter-wave space communication systems. Brief scenarios which describe research and development needs follow for several of the more critical technologies.

Propagation

By far, the one item of greatest impact on the results of this study was the assumed propagation fade statistics. Consequently a more refined engineering analysis of 40/50 GHz communications should await basic data from satellite experiments in the 40/50 GHz region. The scale of these data should be comparable with the work performed at lower frequencies. The propagation studies are more difficult at these wavelengths, not only because of the increased clear air attenuation over that existing at lower frequencies, but also because of the increased attenuation resulting from rain and cloud coverage. As a result of these factors, propagation of millimeter waves has exhibited severe fluctuation effects and has been difficult to analyze. The research required for millimeter wave propagation could be done in conjunction with other experimental work requiring geosynchronous satellites and allowing the additional payload of a group of millimeter-wave beacons.

High Data Rate Diversity Link

In choosing the means of transmitting between two spatial diversity sites, several techniques were considered. From the viewpoint of size and operation during inclement weather, the buried millimeter wave link and fiber optic system have the greatest potential. These two schemes also provide the greatest capacity for high data rate transmission. Substantial research and development efforts are already under way in both these areas and it is doubtful that additional effort would be called for. At this time it would appear that the buried waveguide and optical fiber technologies will be competitive. However, because of its large contribution to the overall cost of the satellite communication system (Application I), the diversity link costs must be substantially reduced and/or the link operated with high traffic loads.

Bulk Data Storage

The attractive capability of millimeter wave communications to provide near 1 Gbit data rates is severely limited by the interface of the communication to the users. It is always necessary to provide buffer storage which operates at these high data rates. Currently, solutions require high parallelism in digital equipment and correspondingly large costs. Several technologies have been suggested which may eventually accommodate these applications, but none is sufficiently developed to allow estimates of availability.

Since there is strong motive for the development of high data rate storage in the computer industry, it is likely that additional research sources will not speed the process. Rather, research should be limited to determining new advances in the area and judging their impact on the attractiveness of millimeter digital communications.

Space Switching Equipment

Switches for application in millimeter wave communications applications are currently available but are considered too bulky for the large capacity systems of interest. The development program for these components would be to provide reliable ferrite switches while taking advantage of the inherent small size of millimeter devices. Special attention should be given to the use of these switches in matrix arrangements with configurations adapted to satellite communication requirements.

This problem is primarily one of engineering design; most of the work is that of prototype construction and testing. Flight tests are required primarily for reliability and life-time analysis. After the switching capacity requirements are specified it is estimated that development can be completed in 2 years.

Receiver and Transmitter

Because of the high loss propagation characteristics of millimeter waves, improvements in system performance will depend heavily on the availability of high performance receivers and transmitters. In particular, the weight of the spacecraft transmitter is especially critical. With our assumed models, it appears these devices would account for a substantial portion of spacecraft weight. In some configurations the required satellite weight exceeded launch capabilities.

By our estimates, a 2 lb. reduction in spacecraft weight can be realized for every 1 lb. reduction of transmitter weight. The 2:1 leverage occurs because of the reduced requirements for structure, attitude control and station keeping. Our analyses indicate only a modest RF power requirement per device for Application I. However, the total RF power required is substantial, requiring a significant weight penalty in the thermal control system. In Application II, the required RF power and thermal control capacity per transmitter were substantial and severely restricted satellite payload.

Therefore, emphasis in the technology effort on spacecraft transmitters should be on lightweight devices, efficient operation, and modest to high power outputs. Both the spacecraft and ground terminal receivers should have a relatively low noise performance. It appears appropriate to consider cryogenically cooled types for the ground terminals while uncooled types may suffice for the spacecraft.

Satellite Antennas

Two areas of satellite antenna development are of interest in millimeter wave communication applications. One is to improve the tolerance of dish or lens fabrication to reduce error tolerances. At millimeter wavelengths this allows significantly improved antenna gain. The second is further development of multibeam antenna techniques, an important adjunct to the switch capacity of a communication satellite. Each of these areas requires further engineering studies to improve construction techniques and to decide among alternative designs. Work is currently underway for both of these design efforts. Consequently, it is expected that two years is sufficient for adequate development after system requirements are defined.

CONCLUSIONS

For the trunking application, typical annual cost to the user for a simplex voice channel via a high capacity 40/50 GHz satellite is approximately $950. However, the rain margin assumed is sufficiently only to support 99.9% availability with respect to rain attenuation. Cases having higher reliability, such as 99.99%, were not evaluated due to excessive system costs and/or excessive spacecraft weight. This is significantly lower than current simplex channel tariffs of $3,500 to $6,500 annually. The bulk of this difference is due to economy-of-scale effect arising from the use of high-capacity millimeter wave satellites. Other operational factors, such as differences in assumed versus actual utilization, would account for the remainder.

For the wideband direct-to-user application, the annual costs were about $200K per user in a 360 terminal 40/50 GHz satellite network. The spacecraft provided 120 half-duplex channels to this network so that on the average each terminal could access the spacecraft 1/3 of the time. However, no site diversity was assumed for this application.

Compensation for rain attenuation aboard the spacecraft led to excessive cost and/or spacecraft weight for configurations having rain reliability in excess of 96%. A similar commercial service using existing satellites for television and radio would cost approximately $500K/year/terminal; however, it would provide higher reliability.

One of the unknowns which will significantly influence the design and cost of a millimeter space communication system is the propagation statistics for the ground station locations. One of the primary results of the study relates the link reliability (percent of the time the link is operational) to assumed weather statistics and, in the case of the point-to-point service, an assumed ground station diversity.

Technology risks have been defined in the research program for those technologies deemed most critical to the cost of an overall millimeter communication system. The critical technologies include all receivers and transmitters, bulk data storage, diversity landline, satellite switching and satellite antennas.

Recommendations as a result of this study include additional experiments and analysis of atmospheric propagation characteristics and specific technology research scenarios intended to reduce the costs of subsystems. It is also recommended that the methodology and models developed here be extended to other applications such as navigation satellites where maximum advantage can be taken of existing methodology and models.

Further investigation of satellite broadcast applications at millimeter frequencies is required. Such investigations should be directed toward increasing the link reliability by the use of multiple satellites and massive satellites to provide sufficient RF power to assure communications through moderate rainstorms. The commercial marketability of applicable services should also be investigated.

ive to implementation of advanced communication satellites would include additional research in on-board signal processing, direct modulation for receive/transmit at 50/40 GHz, data regeneration for use with digital transmission, storage and methods for efficient use of space communication links with variable data rate users.

Since completion of these investigation, several technology advances and new concepts have been demonstrated for millimeter wave satellite communications. None of these, however, negate the results of this study, and all could be employed with the methodology which has been discussed here. A new model for estimating rainfall attenuation along an earth-satellite path has been introduced recently by Crane.[9] This model is perhaps the most accurate employed thus far, as Crane has proposed a set of eight regions into which the earth is separated for estimating rainfall rates.

Ricardi[10] has used the Crane model to calculate the cumulative distribution of total atmospheric attenuation along the satellite path at various frequencies in the band from 7-50 GHz. From Ricardi's work, it is noted that a link operating at 40 GHz will be disrupted on the order of 2-3% of the time. It is pointed out, however, that space diversity (as discussed above) can substantially reduce the outage due to excess rain attenuation.

In addition to propagation losses, Ricardi has presented other factors which should be integrated into the systems analysis discussed here. Among these factors which influence the choice of frequency of a millimeter SATCOM system are frequency variation of a terminal's EIRP, spatial discrimination, and bandwidth available for spread spectrum anti-jamming. In discussing these factors, Ricardi has indicated that operating at millimeter wavelengths improves the system's ability to discriminate spatially between desirable and un- spread spectrum techniques to improve resistance to undesirable signals, and to reduce the difference in EIRP of a potentially interfering terminal and the SATCOM terminal. The trade-off of all these factors with the effects of propagation loss should be further considered as part of the systems analysis of this paper.

A recent technology assessment by Frediani[11] provides an updating of many sub-systems investigated originally in this program.[5,6,7]. The work of Frediani should be reviewed to include his recommendations where appropriate.

In addition to the concepts and technology discussed in the above references, Eaves[12] has investigated the use of the upper SHF and EHF regions to provide secure satellite communications to large numbers of users. In particular, he has considered affordable and secure mobile communications. As alternatives to conventional transponders, he has presented system concepts and technologies which include uplink coverage through directive and nulled beams, on-board signal processing, and downlink beam hopping. Eaves has indicated that, to achieve the goal of interference-resistant EHF SATCOM for large numbers of small mobile users, a complex space segment is necessary. The systems analysis described here should be applicable to the following concepts which were shown by Eaves to provide the basis of the system he considered

- On-board signal processing for demodualtion-remodulation and decoupling the uplink and downlink;
- Multiple-beam reception for reliable uplink communications with small terminals;
- Power-efficient signal processing with SAW technology to demodulate large numbers of FDM users;
- Provision of satellite EIRP to serve large numbers of small terminals at modest power consumption through a TDM downlink with beam hopping.

coupling of uplink and downlink by the use of on-board signal processing is the fact that this permits an FCM uplink/TDM downlink structure to avoid many of the limitations of transponder satellites. Navy SATCOM requirements parallel many of the concepts discussed by Eaves[1] and Ricardi[10] in that the Navy community is one with a large number of highly mobile platforms each having a comparatively light total communications load.[13] The use of a processing satellite has been shown to be important for this application as is the use of the highest frequency which technology will support and which permits acceptable link performance.[13] The use of millimeter wavelengths for all applications of interest here must be compatible with the regulations set forth recently by the World Administrative Radio Conference '79 (WARC-79).[14]

In extending SATCOM links to higher frequencies, consideration has been given to employing frequencies as high as 300 GHz.[15] As technology and atmospheric models improve for these higher frequencies, analysis as described in this article can be applied to describe feasibility and affordability of systems operating at frequencies as high as 300 GHz. Technology, in particular that of receivers and transmitters, has advanced considerably in recent years for frequencies above 100 GHz[16,17] In addition to the papers discussed above, several sessions of EASCON '79[18] were devoted to advanced satellite communication technology, and should be integrated where appropriate into the work reported here.

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(continued on page 93

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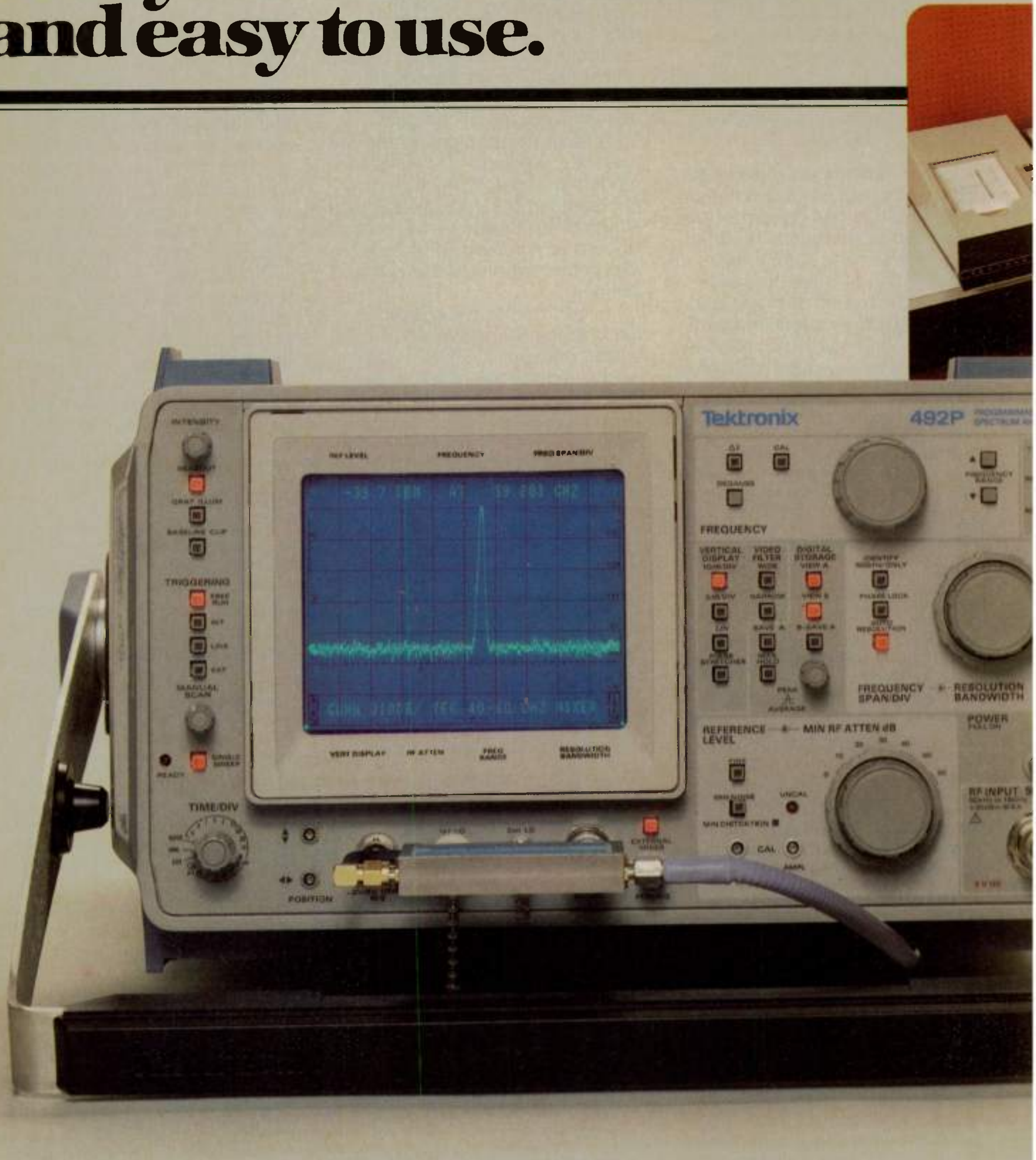

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Semiconductor Light Sources for Fiber Optic Applications

T. E. STOCKTON and R. B. GILL
Laser Diode Laboratories, Inc.
New Brunswick, NJ

INTRODUCTION

Common to any fiber optic data transmission link, regardless of format, are the following essential components: the photon source (LED, laser), the transmission medium (optical fiber), and the photon sensor (detector). Among these elements, the source is usually considered the active component. It is the objective, in what is to follow, to shed more understanding on the element which is the source of light in fiber optics. The fundamental function of any fiber optic light source is to efficiently convert electrical energy (current) into optical energy (light) in a manner which permits the light output to be effectively launched into the optical fiber. Also, the light signal generated by the source must accurately track the input electrical signal to minimize distortion and noise. From the overall systems point of view, enough signal power must be generated at the source to overcome attenuation of the fiber plus interface (coupling and connector) losses with enough optical power left over to drive the detector. Above all, reliability and cost factors must demonstrate substantial improvements over conventional data transmission techniques in order to justify the use of fiber optics.

System power budget, distortion, reliability and cost considerations require that fundamental system elements be "performance optimized", and it is in this context that the subject of semiconductor light sources will be discussed. Although a myriad of device types and configurations exist on today's market, we have chosen the two most widely used light sources for detailed discussion: the etched-well surface LED, and the single mode CW laser diode. Taken together, both devices adequately cover the majority of analog and digital data transmission applications via fiber optics.

A GENERAL COMPARISON BETWEEN LIGHT EMITTING DIODES AND INJECTION LASERS

Power Efficiency and Beam Characteristics

Although LEDs and lasers are fabricated from the same basic semiconductor compounds, and, in fact, have similar heterojunction structures, they differ substantially in their performance characteristics. LEDs are generally less efficient than injection laser diodes (ILDs) and their spatial intensity distribution in Lambertian (cosine). Lasers, on the other hand, exhibit a relatively high degree of waveguiding and, hence, for a given acceptance angle, it is usually possible to obtain at least an order of magnitude improvement in coupled power over that which can be achieved with LEDs. The increased output within a narrow beam gives the laser a distinct advantage when used with low loss and graded index optical fibers, since these fibers generally have small core diameters ($< 100\ \mu m$) and low numerical aperture. (The numerical aperture is the sine of the acceptance half-angle; usually N.A. <0.25 for low loss fiber). For comparison, at the same electrical input power (e.g. 200 mW) and fiber acceptance angle (N.A. $= 0.2$), the LED is capable of launching about 300 μW of optical power into the core of the fiber, and an injection laser about 3.0 mW. The total optical power produced by the LED in this case is about 7.0 mW, whereas the laser emits about 10.0 mW for the same electrical input. Thus, the coupling efficiency is about seven times better for the laser. It is

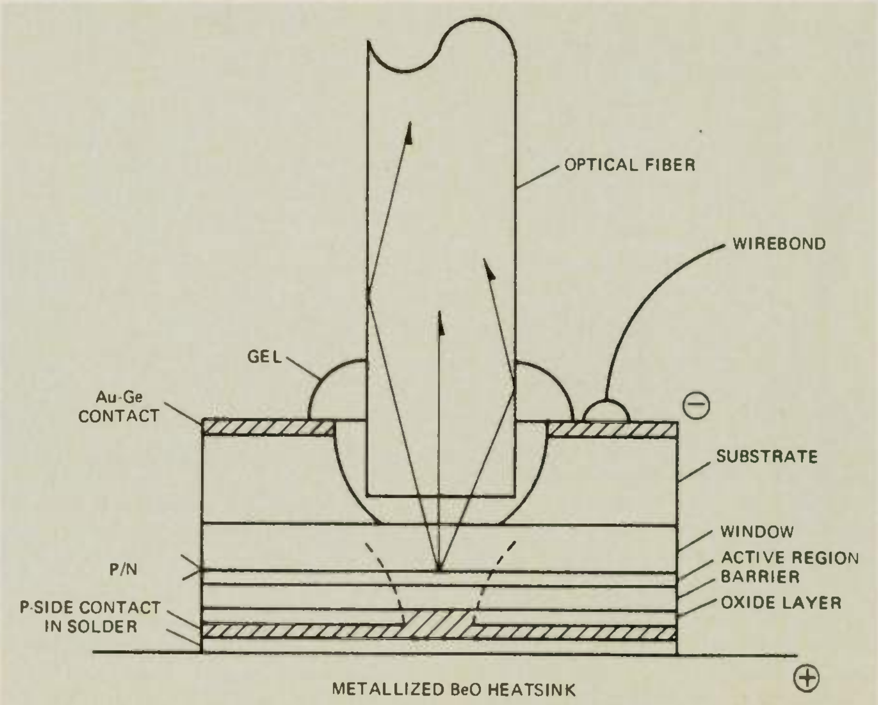

Fig. 1 Etch-well light emitting diode schematic.

efficiency considerably for both LEDs and lasers through the use of microlenses.

Since injection lasers and LEDs can be fabricated from the same semiconductor compounds, it is possible to optimize both devices to match the attenuation characteristics in optical fibers by varying the composition, and hence the bandgap and emission wavelength, of the active region of the light source. Although the loss curves for the many varieties of glass fiber vary in detail, generally they possess minima near 820 nm and beyond 1200 nm. Typical losses on the order of 5 dB/km or less can be achieved in the 800 nm to 850 nm region and, more recently attenuation of only 0.2 dB/km has been achieved at 1500 nm. GaAs-GaAlAs (Gallium-Arsenide, Gallium-Aluminum-Arsenide) semiconductor compounds span the wavelength region between 780 nm and 940 nm. For the longer wavelengths, the quaternary alloys based on InP-GaInAsP (Indium-Phosphide, Gallium-Indium-Arsenide-Phosphide) are used to cover the wavelength region from 1000 nm to 1600 nm.

Signal Distortion and Noise

Signal distortion can originate in both the light source and the transmission medium. For sources, both LEDs and lasers, pulse response (digital) and linearity (analog) are parameters of prime importance. LEDs exhibit slower pulsed response than lasers and have rise and fall times in the range of 3 to 15 nanoseconds. Lasers, however, are much faster devices, capable of responding within 200 picoseconds when biased at or slightly above threshold. Recently, modulation in excess of 4 GHz has been achieved using injection lasers. Because there is a trade-off between speed and efficiency in most LEDs, their useful signal bandwidth extends only up to about 200 MHz. Linearity, expressed as the percentage of power contained in the second harmonic of the demodulated light signal, is better than 40 dB below both LEDs and lasers. Lasers, because they require a finite amount of current to reach their "on" state and because this "threshold current" is strongly temperature dependent, require more complex drive circuitry than do LEDs. Feedback stabilization of the light signal, with respect to temperature and power output, is often required for the satisfactory use of lasers at higher frequencies. For high frequency analog modulation (> 200 MHz) of injection laser diodes, harmonic distortion must be below −52 dB. This value is attainable only with single mode lasers

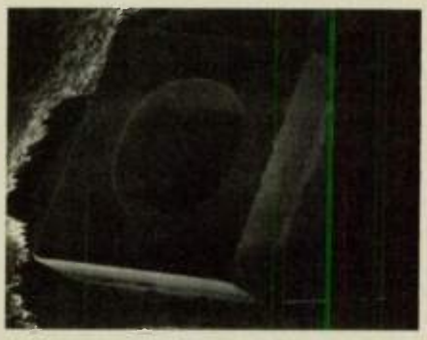

Fig. 2 SEM photograph of LED chip (200X).

such as the structure described in the section on single mode injection laser diodes.

Signal distortion can also arise from the transmission medium. Pulse spreading in optical fiber is due primarily to modal and material dispersion. In step index fiber, i.e. a circular waveguide composed of a high index transparent "core" surrounded by a lower index of refraction "cladding" material, modal dispersion is a simple consequence of the difference in path length for axial and skew rays. The problem of modal dispersion is greatly reduced through the use of "graded index" optical fiber, in which a near parabolic index of refraction profile is used to slow down axial rays in order to precisely compensate for differences in path length. Material dispersion, on the other hand, is due to the fact that the index of refraction (i.e. speed of light) in the core is wavelength dependent Thus, material dispersion is minimized when the spectral bandwidth of the source is kept narrow. The spectral width for the 40 nm, whereas for lasers the spectral width (full width at half intensity) is less than 3 nm.

Recently, injection lasers operating in a single longitudinal mode have been fabricated with an emission spectrum less than 0.1 nm. For these devices, material dispersion in the fiber is essentially eliminated as a significant source of signal distortion. Much of the recent activity in longer wavelength devices (InP-GaInAsP) stems from the fact that in many fibers, material dispersion passes through zero in the wavelength region between 1200 nm and 1300 nm. Together with the very low attenuation at longer wavelengths, zero material dispersion is the driving force behind the development of quaternary LEDs and lasers.

Reliability

In the past, LEDs have generally been considered to be much more reliable than their laser counterpart. This is not surprising since although the material structures are almost identical for both devices, additional degradation mechanisms exist in the case of ILDs which at first glance may make them appear to be somewhat less reliable than LEDs.

Gradual degradation, however, occurs by the same bulk mechanism in both LEDs and lasers. Dislocation networks in the vicinity of the active region can propagate under the influence of mechanical stress and heat. If they penetrate the active area, non-radiative recombination occurs and causes a reduction in internal quantum efficiency. Unfortunately, in the case of laser, the effect is much more noticeable since these increased optical losses are manifested as a shift in threshold. Because of the high external differential quantum efficiency of the laser, large changes in power output can result from relatively small changes in threshold current. For the LED, the same reduction in efficiency results in a much smaller decrease in light output. Bulk degradation in both LEDs and lasers is minimized by using substrate material having the lowest possible defect count and by minimizing those forces which tend to induce defect migration. Device thermal impedance and bonding stress are therefore of paramount importance in reducing degradation rates in fiber optic sources.

In addition to bulk degradation, injection lasers suffer the additional problem of facet erosion. Changes in the reflectivity of the end mirrors of the laser caused by the photochemical interaction of GaAlAs with elements in the external environment result in a gradual increase in threshold with time. This problem virtually has been eliminated through the use of facet passivation films composed of aluminum oxide.

More recently, the reliability of both lasers and LEDs has improved to the point where degradation rates of 1%/1000 hours or less have been routinely observed. Lifetime estimates made through the use of high temperature accelerated life tests indicate that useful lifetime in excess of 10^5 hours can be expected.

THE ETCHED-WELL LIGHT EMITTING DIODE

For applications requiring moderately high digital and analog data transmission up to several kilometers, the high radiance etched well emitter has proven to be a reliable and inexpensive alternative to the injection laser. Shown schematically in **Figure 1**, the etched well emitter consists of a double heterojunction waveguide which has been selectively contacted on the p-side (heat sink face) to form a circular emitting aperture nominally 50 μm in diameter. The substrate above the aperture is removed via chemical etching thus forming the "etched-well" for which the device is named. Although, as shown here, the substrate is mounted up, fabrication of the device structure is accomplished by depositing the active semiconductor layers on top of the substrate in a process commonly referred to as liquid phase epitaxy (LPE). In this process, layers of ternary $Ga_{1-x}Al_xAs$ are precipitated onto a single crystal GaAs substrate from molten Gallium-rich melts. Crystal growth is accomplished by controlled cooling of the saturated melts at approximately 850°C. By exposing the substrate to each metal for a predetermined period of time during the cooling cycle, the thickness of each layer is precisely controlled. In **Figure 1**, the GaAs substrate is n-type, and in the completed device has a thickness of approximately 100 μm. The first grown layer is a 10 μm thick $Ga_{.7}Al_3As$ n-type region referred to as the "window"; this layer forms a transparent window through which light generated in the active layer passes. The next grown layer, the active layer, is p-type Ge doped $Ga_{.95}As_{.05}As$ about 1 μm thick. It is followed by a 2 μm p-type $Ga_{.7}Al_{.3}As$ barrier layer which completes the three-layer waveguide structure. Normally, a fourth layer of GaAs (not shown) is grown as a passivating cap. Aluminum is employed to increase the bandgap and reduce the index of refraction of the semiconductor compounds so that, when the structure is forward biased, carriers injected across the P/N junction are confined to the active layer. Highly efficient radiative recombination takes place in this region creating photons at the bandgap energy of the active layer which are emitted in all directions. **Figure 2**, is a Scanning Electron Micrograph (SEM) of a single LED chip in which the etched-well is clearly visible. The chip is about .020" square by .004" thick; the well is approximately .009" in diameter at the bottom. The circular emitting aperture (not visible) is centered in the well bottom of the opposite side of the chip facing the heatsink surface and is approximately .002" in diameter.

It is important to point out that although the internal quantum efficiency of the double heterojunction LED is extremely high, only a relatively small percentage of the light escapes from the semiconductor. This is primarily because the high index of refraction of GaAs and its GaAlAs alloys, about 3.2, defines a small angle of total internal reflection within the chip. Thus, although the absorbing GaAs substrate has been removed by selective etching, only light incident within a 17° cone normal to the window layer surface can be extracted. The remaining optical energy is reabsorbed within the bulk of the chip. Also, within certain limits, the LED rise time, quantum efficiency, and peak emission wavelength can be adjusted by varying the acceptor concentration and aluminum content in the narrow active recombination region. Heavier Ge doping in this layer reduces the carrier lifetime and results in faster rise and fall times, but at the expense of quantum efficiency. Alternatively, a lightly doped active layer results in high efficiency but with much slower response. Optimum doping occurs when $N_A \sim 1 \times 10^{18}$ which yields a total efficiency of 5% with rise time of 15 nsec. for the structure described here. As mentioned previously, the peak emission wavelength, λp, can be adjusted by varying the solid composition of $Ga_{1-x}Al_xAs$ of the active layer. For $x = 0$, is GaAs, $\lambda_p \sim 900$ nm; for $x = 0.10$, $\lambda_p \sim 800$ nm with a linear dependence within these composition limits.

One convenient feature of the etched-well emitter, is the ability to use the etched well to position the fiber pigtail for optimum alignment. In general, the LED aperture diameter ($\sim 50\ \mu$m) is smaller than the core diameter of typical step and graded index fiber pigtails. This minimizes the alignment accuracy necessary to achieve optimum coupling efficiency between LED and fiber pigtail. For large core (125 μm)

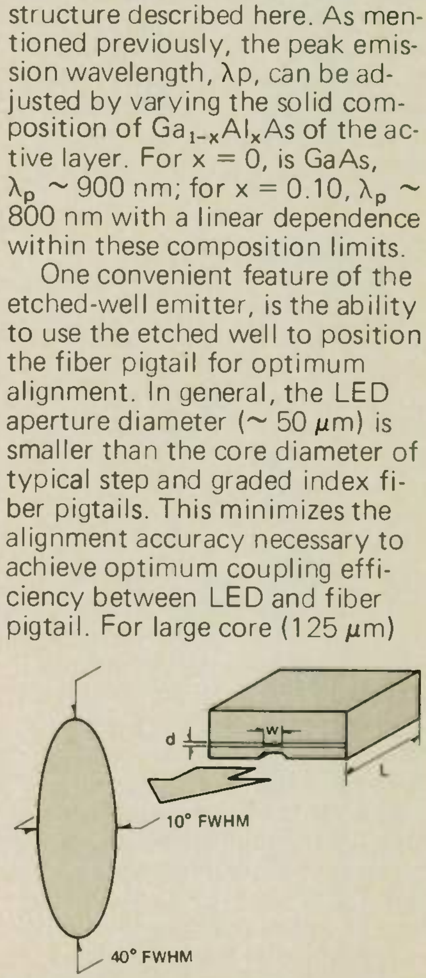

Fig. 3 Schematic of laser diode chip showing far field emission.

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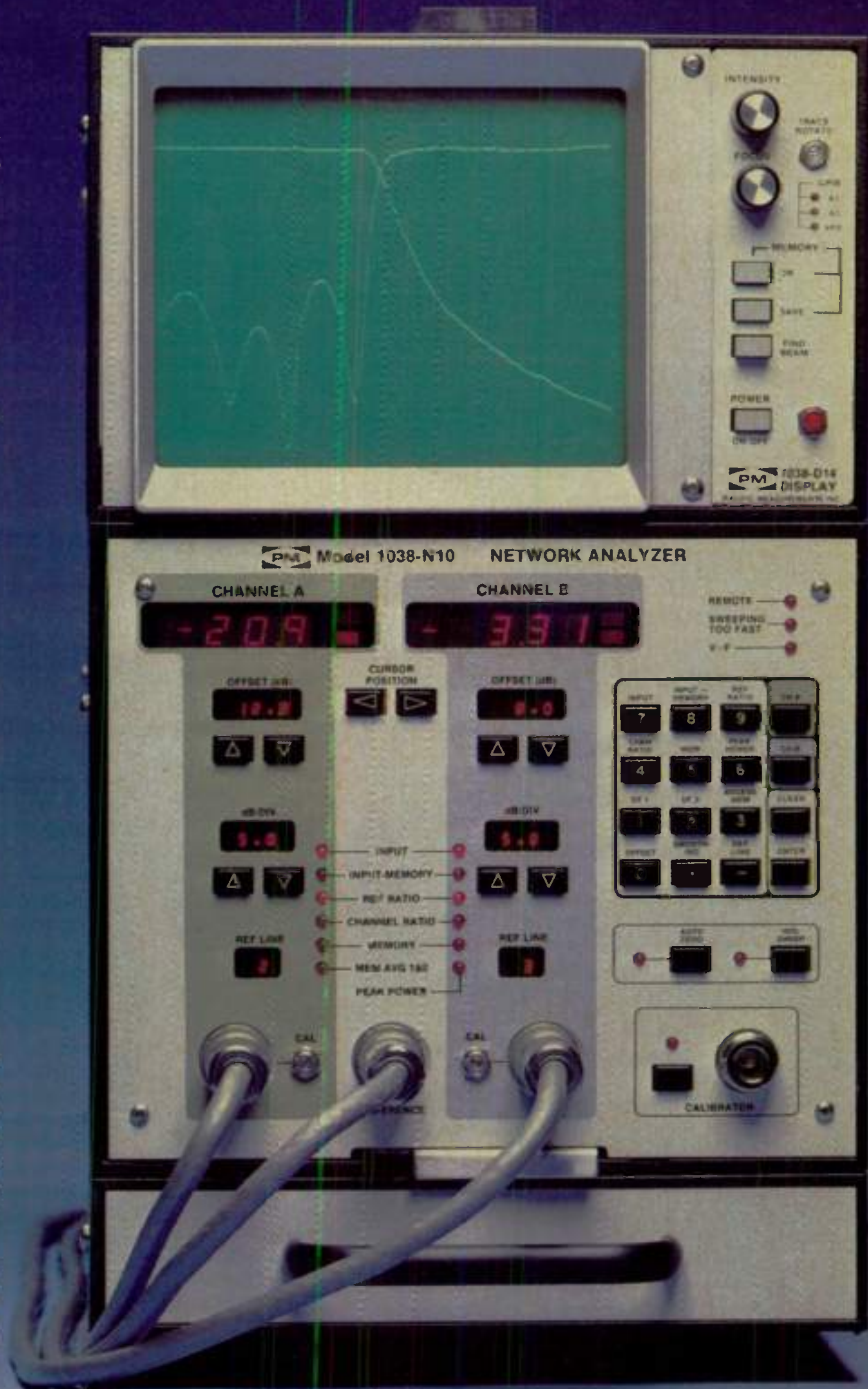

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high N.A. (0.5) step index fiber, coupling efficiency of about 25% has been achieved with simple butt alignment. Index matching gel can result in a substantial improvement in coupling efficiency even though the effective N.A. of the fiber is reduced in the presence of a higher index gap material. This is due to the fact that LED external efficiency is improved by increasing the angle of total internal reflection within the LED chip. One novel version of the LED involves the use of a spherical microlens in place of the fiber pigtail shown in **Figure 1**. Microbead lenses can be used in large core or fiber optic bundle applications in which the source is fabricated as a discrete component without the pigtail. A 150 μm diameter spherical microbend placed within the etched-wall of the LED chip collimates the output beam so that 1.5 mW can be coupled into a standard 0.5 N.A. .045" diameter fiber optic bundle.

THE SINGLE MODE INJECTION LASER DIODE

Since the commercial introduction of CW semiconductor lasers in 1975, many improvements in performance and reliability have been achieved through the use of increasingly more sophisticated laser structures. Recently, the development of the Double-Carrier-Confined (DCC) structure described at the end of this section has resulted in exceptionally stable performance suitable for both high frequency analog and digital applications. **Figure 3** shows a schematic diagram of a laser chip along with a cross-section of the emitted beam. In its most fundamental form, the injection laser consists of a simple P/N junction slab waveguide which contains end mirrors formed by cleaving the crystalline semiconductor along parallel plates. Photons generated by carrier recombination in the vicinity of the junction provide the pumping mechanism when the diode is forward biased. Because of the high index of refraction of GaAs and its related compounds, the semiconductor-air interface provides a built-in reflectivity of about 30%; this reflectivity provides optical feedback required to operate the laser. To achieve continuous wave (CW) operation, the conditions of effective optical waveguiding parallel and perpendicular to the P/N junction, extremely low operating current density, and low thermal impedance are necessary. As with the LED structure described previously, the double heterojunction (DH) waveguide provides an excellent means of controlling both carrier and optical confinement

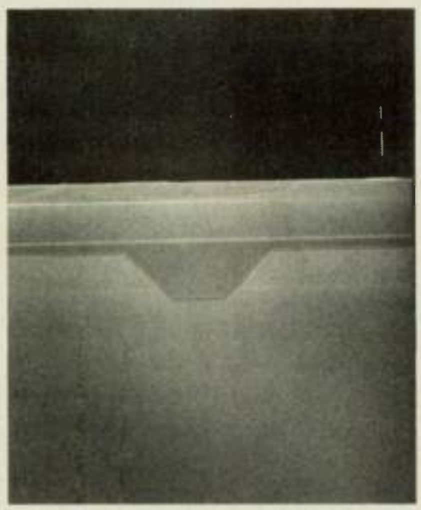


Fig. 4 SEM photograph of DCC laser structure (5000X).

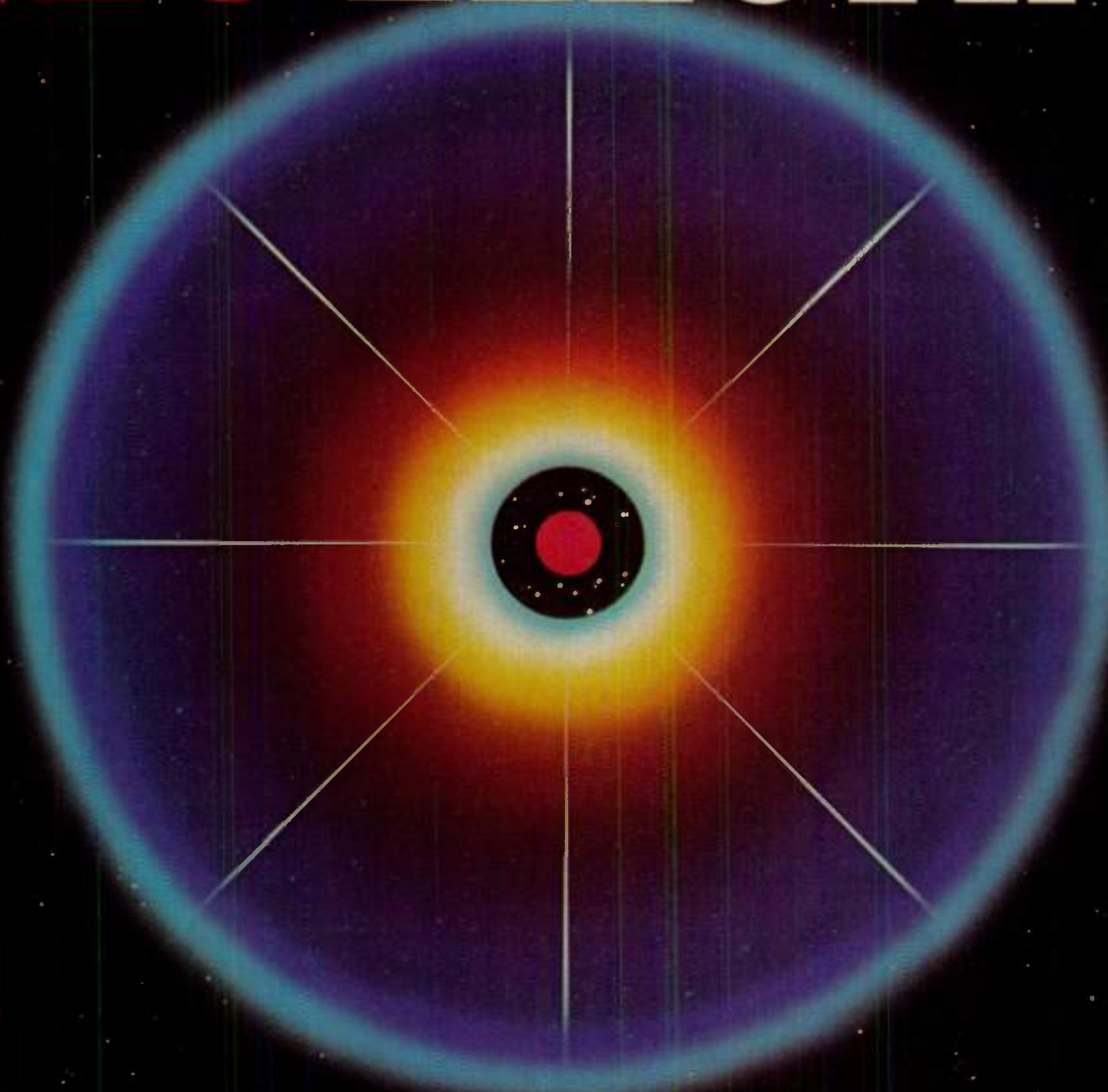
by using $Ga_{1-x}Al_xAs$ to modify the bandgap and index of refraction profile of the laser waveguide. Similarly, the peak emission wavelength of the laser is determined by the aluminum fraction in the active layer. It has been shown experimentally that the threshold current density required for CW operation of the DH laser must be below about 2.5 KA/cm^2; J_{th} greater than this value results in excessive self-heating due to the relatively low thermal conductivity of GaAs alloys. To achieve low current densities and minimize Joule heating, extremely thin active layers of less than 0.3 μm (d in **Figure 3**) are required. For a composition difference between guiding layers and active region of $\Delta x = 0.3$, threshold current densities less than 1.0 KA/cm^2 have been obtained for DH diodes with thin active layers. An unfortunate consequence of the need for low current density and thin active layers is the increased divergence of the emitted beam due to diffraction as light exits the laser facet; i.e. $\lambda p \sim d$. Beam divergence perpendicular to the plane of the junction between 30° and 50° FWHM typically results. Semiconductor injection lasers differ vastly from macroscopic gas and solid state laser systems in this one respect.

In order to minimize the absolute threshold milliamperes, "stripe geometry" must be employed. Many methods exist for fabricating stripe geometry DH diodes, but all of these techniques primarily are aimed at reducing the surface area of the waveguide in order to isolate a single parallel transverse mode. This is usually accomplished by selectively contacting the surface of the laser chip so that current flows in a narrow stripe, usually less than 10 μm wide (w in **Figure 3**) along the length of the optical waveguide. With conventional diodes of this type, lateral current spreading occurs as the drive current is increased so that the optical width in the near field increases. These variations cause the excitation of higher order parallel transverse modes and changes in the parallel beam divergence as a function of drive current. Nonlinearities in output vs. current usually develop as a result and it is this type of behavior which is particularly undesirable for fiber optic applications.

Figure 4 is a Scanning Electron Micrograph of the DCC single mode laser cross section, in which output stabilization has been achieved by funneling current through a conducting channel formed in the GaAs substrate prior to LPE deposition of the GaAlAs laser waveguide structure. In this photograph, the laser structure has been magnified 5000 times; the channel width is approximately 5 μm and the active layer, which appears as a continuous single white line extending across the top of the channel, is less than 0.2 μm thick. In this more recently developed structure, only a single parallel transverse mode is generated over

(continued on page 56)

NEC ELECTRO OPTICS



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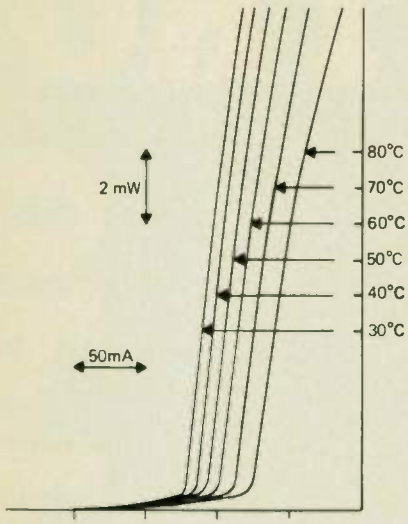


Fig. 5 Light output vs. drive current for single mode CW lasers.

the entire operating range of the diode. Also, because all of the funneled current contributes to the lasing stripe, i.e. no current is lost in spreading, threshold current has been reduced to well below 100 mA in the DCC structure. Typical power output versus input current curves as a function of ambient temperature are shown in Figure 5. High power capability in excess of 14 mW is achieved with room temperature threshold current of about 75 mA. The device exhibits highly linear behavior and its parallel beam divergence is maintained at 6° FWHM over the entire operating range; perpendicular beam divergence for this device is approximately 35°.

DCC lasers are available as discrete components or coupled to a fiber optic pigtail. Figure 6 shows a typical single mode laser assembly (SCW) which incorporates a graded index fiber optic pigtail and a monitor detector. The detector, by sensing scattered light, allows the power level at the source to be monitored as part of a feedback control network to compensate for threshold shifts as a function of ambient temperature. Coupling efficiency between the DCC laser chip and 62 μm core diameter, 0.2 N.A. fiber is 40 to 50% when a spherical lens is formed on the input end of the pigtail. Alignment of the fiber to the 7 μm x 0.2 μm source is accomplished using a special microprocessor controlled stepping motor manipulator having a step resolution of 0.1 μm. Using this configuration, the tracking error between laser and monitor output is less than 2% up to the maximum output from the pigtail (5 mW). Current laser development is directed toward the incorporation of additional elements within the laser package. For example, the entire laser transmitter including TE cooler, thermistor, feedback network, detector chip, and signal interface can be incorporated in a single dual in-line package (DIP) modified to accept either a pigtail or F/O connector.

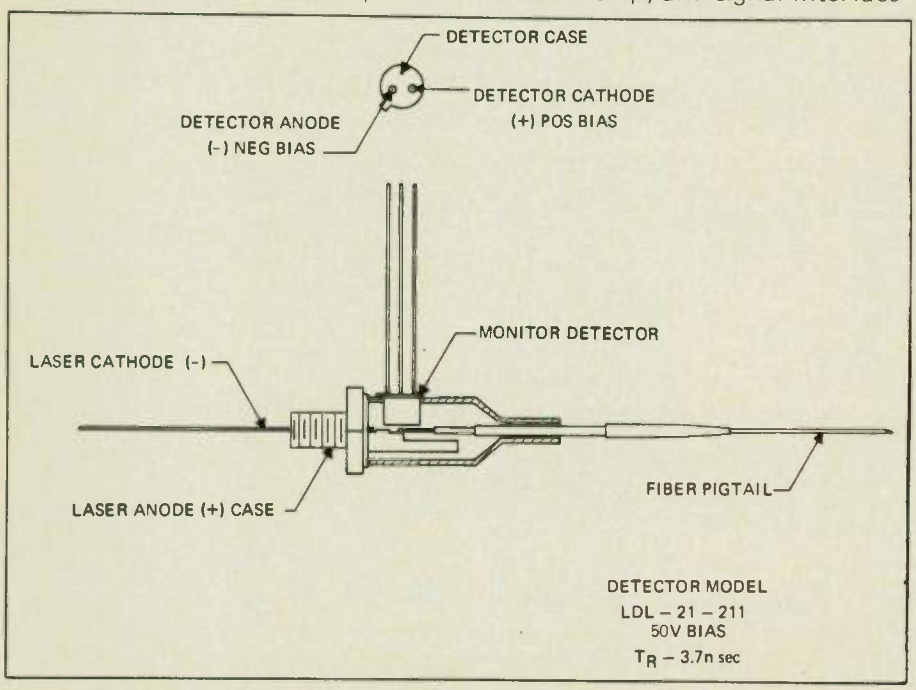


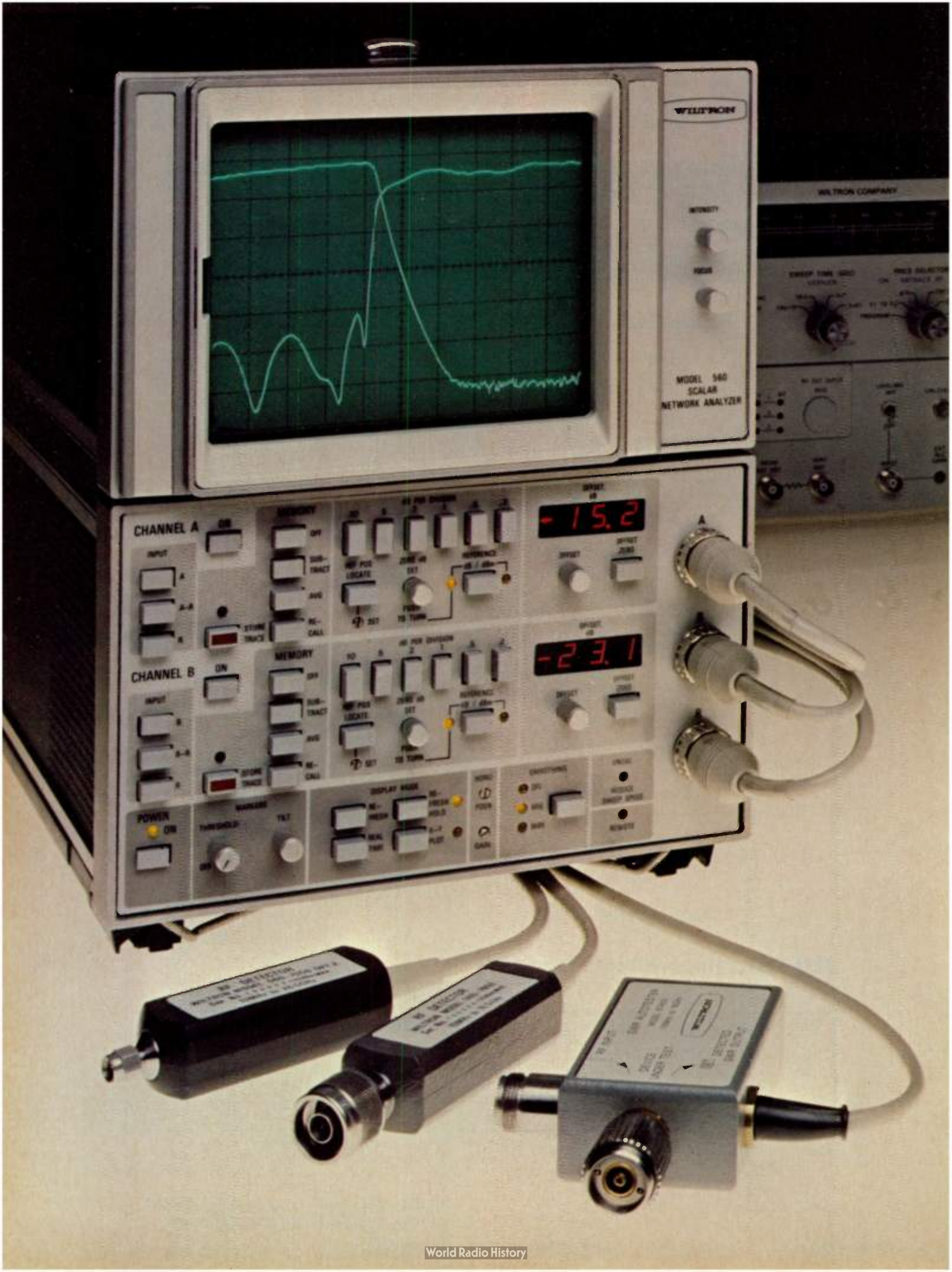
Fig. 6 SCW-laser assembly.

SUMMARY

Two important light sources for fiber optic applications, the high radiance etched well LED and the single mode injection laser, respectively, have been described in terms of their performance characteristics and structure. Requirements for very high power and data rate may dictate the use of lasers over LEDs while low bandwidth links LEDs are more appropriate. When comparing the two devices for the purpose of evaluating their relative suitability in any given F/O systems application, one must fully recognize the wide range of overlap in their performance. There is a general tendancy to utilize LEDs whenever possible, simply on the basis of reliability and cost. These factors will become less important in the future, since many of the laser reliability problems experienced in the past are rapidly being resolved. Also, because of the nature of semiconductor devices, future costs are expected to fall considerably as the device market matures.

Thomas Stockton graduated with high distinction in Physics from Rutgers University, where he received his Bachelor Degree in 1971. He was employed at RCA Laboratories in 1972 where he engaged in optoelectronic device research, design, and development. Mr. Stockton came to the staff at Laser Diode Laboratories, Inc., in the fall of 1974 as Operations Manager, Devices. And in 1977, Mr. Stockton joined the staff of Spectronics, Inc. where he was Director, Injection Laser and Fiber Optic Source Development. He rejoined LDL in 1978 as Director of Research. In his current position at LDL, he has direct responsibility for all R&D programs.

Robert Gill received the B.S. degree in physics from Fairleigh Dickinson University in 1963 and the M.S. degree in physics from Stevens Institute of Technology in 1967. He was employed at RCA Laboratories in 1963, where he worked on the development of semiconductor devices, including lead-salt infrared photoconductive and photovoltaic detectors and thin-film gallium arsenide solar cells. Mr. Gill transferred to the Optoelectronic Products Department of the RCA Solidstate Division in 1968, where he was assigned to Semiconductor Engineering. In 1971, Mr. Gill joined the staff of Laser Diode Laboratories, Inc., as Operations Manager, Devices, and became the President in 1973. In this position, he is in charge of R&D as well as production engineering on materials, designs and processes for the injection lasers and laser arrays.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18 GHz Passive Repeater Considerations

M. J. SHEPHERD
Farinon Electric Operation
San Carlos, CA

A previous article (see Reference 1) has described in detail the DM18 radio's purpose and its environmental and technical qualities. Path design aids will be described that explain how to overcome intervening obstacles and circumvent on-site construction/maintenance problems. Sample calculations are given for several paths.

The calculations contain the system gain of the radio, the gain and loss of the antenna-transmission line combination, and the variable quantities that make up the loss of the path. These path losses, free space attenuation, atmospheric absorption, and rain attenuation are considered first. Then the system and antenna gains are taken into account as part of the examples.

FREE SPACE

Attenuation in free space is a well-known quantity and will be shown here without the effects of atmospheric refraction (fading) being considered.

The loss due to free space may be calculated as shown:

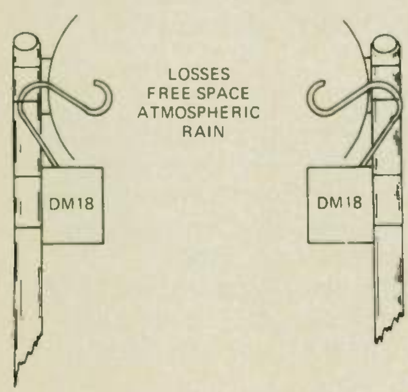


Fig. 1 Free Space consideration showing typical tower mounting (which may also be a short support on top of or within a building).

$$\text{Loss (dB)} = 96.6 + 20 \log f \text{ (GHz)} + \ldots + 20 \log d \text{ (miles)}$$

$$\text{Loss (dB)} = 92.4 + 20 \log f \text{ (GHz)} + \ldots + 20 \log d \text{ (kilometers)}$$

Figure 2 shows ease of calculation, and for comparative purposes, the common frequencies of 2 GHz and 6 GHz are also given.

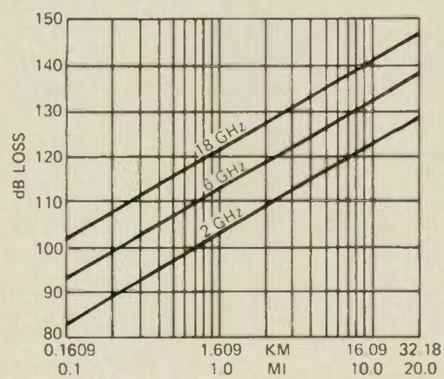


Fig. 2 Free space path attenuation.

ATMOSPHERIC ABSORPTION

Attenuation of a microwave signal by moisture in the air has little or no effect on a signal that is commonly used as a communication frequency below 6 GHz, but as can be seen in Figure 3, higher frequencies are affected.

As the wavelength becomes shorter, the quarter wavelength of the desired signal approaches that of the droplets in the air. Figure 3 shows the commonly known first absorption band due to water vapor peaking at 22 GHz. Note that at 18 GHz the attenuation is 0.1 dB per kilometer (0.1609 dB per mile). This quantity is still quite small, but it will add up for longer paths, i.e., 10 miles = 1.6 dB.

For 18 GHz a composite graph has been made (Figure 4) which gives the sum of free space and atmospheric absorption attenuation. This graph may be used as a practical tool in the design of a path whether it be direct line of sight or via a passive reflector.

RAIN

Rain attenuation is the third factor affecting the signal received. As with atmospheric absorption, the higher the frequency the greater the effect. But since rainfall quantities vary dramatically with geographic areas, this effect must be applied to each individual path. Figure 5 gives the effect of rainfall at various rates for path loss versus frequency. As before, the attenuation is greater at higher frequencies and this must be taken into account for 18 GHz:

Figure 6 is a map of the United States giving the rate of rainfall for different regions. Combining Figures 5 and 6, or other known rain rate, will produce the attenuation to be added to Figure 4. One last note about rain attenuation – rain cells (storms) usually do not occur over the whole path at one time, but have an average diameter of 4 miles (6.4 km); therefore, reducing the quantity

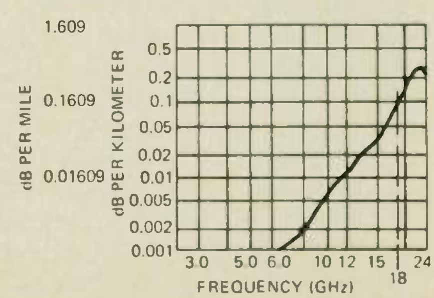


Fig. 3 Atmospheric absorption losses.

of the dB/distance times the rain cell diameter. (See Reference 2.)

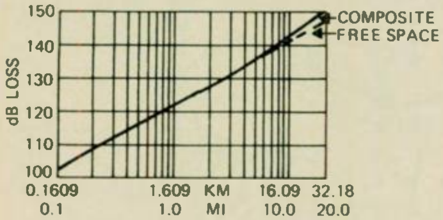


Fig. 4 Composite of free space and atmospheric absorption losses versus distance for 18 GHz.

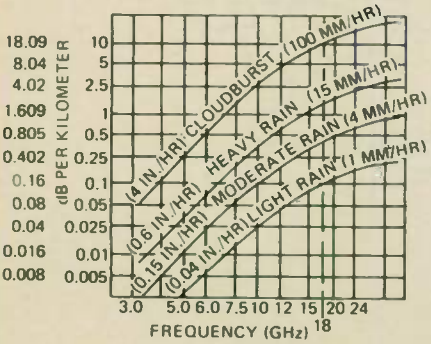


Fig. 5 Rainfall attenuation.

OBSTRUCTED PATHS
(Use of Passive Reflectors)

Path obstructions can be avoided by placing the antenna high enough to transmit over the obstruction or by using a passive reflector to transmit around the obstruction. The DM18 radio, with integral antenna, has been designed to mount at whatever height is necessary to overcome an obstruction, but some users may prefer to place the radio at a more convenient, lower location and use a passive reflector to transmit over the obstruction. The passive reflector can be used either on the obstacle or a high tower near one end (similar to a periscope).

In the United States, the Federal Communication Commission has decreed that periscopes, where the active antenna is pointing vertically, may no longer be installed. Present installations must be changed by 1985. This is due to the inherent possibility of interference with this configuration. One may ask, "When is a passive a periscope?"; the answer may be that the active antenna is not to be pointed vertically (90°) to the horizontal and that the installation does not cause interference with any other users.

The gain of a passive reflector varies with its size, operating frequency, and included angle i.e., angle between the two signal paths.

TABLE 1

PASSIVE REFLECTOR GAIN (dB)

Included Angle (degrees)	Size (feet) 4x6	6x8	8x10	10x16
120	94.0	99.3	103.5	110.5
100	96.1	101.2	105.5	112.5
90	97.0	102.0	106.2	113.5
80	97.7	103.0	107.0	114.0
60	98.7	104.0	108.0	115.2
40	99.4	104.3	109.0	115.8
20	99.8	105.0	109.2	116.2

This gain may be calculated by the use of the following formula, but for simplicity Table 1 is provided at 18 GHz for several passive reflector sizes vs. included angle.

Passive Repeater Gain =

$$\frac{20 \log 4\pi(\text{Area})(\cos\alpha)}{\lambda^2}$$

EXAMPLES

With the data given in the preceding pages concerning path loss and passive antenna gain, we will now present a few typical examples on the following pages.

Mid-Path Passive

a_1 = 2 miles = 128 dB (from Figure 4)

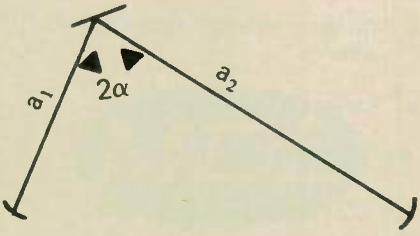
a_2 = 4 miles = 134 dB (from Figure 4)

$2\alpha = 100°$

Antennas at sites (1) and (2) are 4 foot = 44 dB gain each.

DM18 transmitter power = +18 dBm

Receiver threshold = −74 dBm for 4 T1 lines at 10^{-6} BER



The question here is to determine the size of the passive gain and realize a satisfactory receive level i.e., fade margin = 20 dB.

distance a_1 and a_2. No loss is being considered for transmission line because the proposed installation has the antenna no more than two feet from the tower mountable radio.

The gains are the two antennas and the passive which we are to determine.

	Losses	Gains
a_1	128 dB	Two antennas 88 dB
a_2	134 dB	Passive ?
	262	

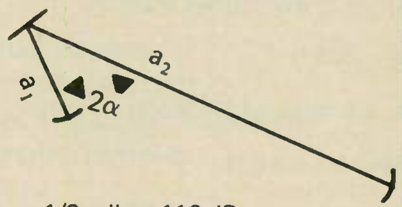
Radio equipment system gain
18 dB + 74 dBm = 92 dB

Required gain of passive = . . .
= losses − ant. gain − system gain
+ an acceptable fade margin

= 262 − 88 − 92 + 20 = 102 dB

Referring to Table 1, we find that for an angle of 100° and a required gain of 102 dB, a passive size of 6′ x 8′ would minimally meet the requirements, but better margin would be obtained using an 8′ x 10′ passive.

Passive Near To One End (not near field): Find fade margin



a_1 = 1/2 mile = 116 dB

a_2 = 8 miles = 141 dB

$2\alpha = 20°$

Using the first passive found in the previous example (6′ x 8′), we have a passive gain = 105 dB.
Radio equipment gain = 92 dB

	Losses	Gains	
a_1	116	2 antennas	88
a_2	141	Passive	105
	257 dB	Radio Equipment	92
			285 dB

Therefore, fade margin equals 285 − 257 = 28 dB.

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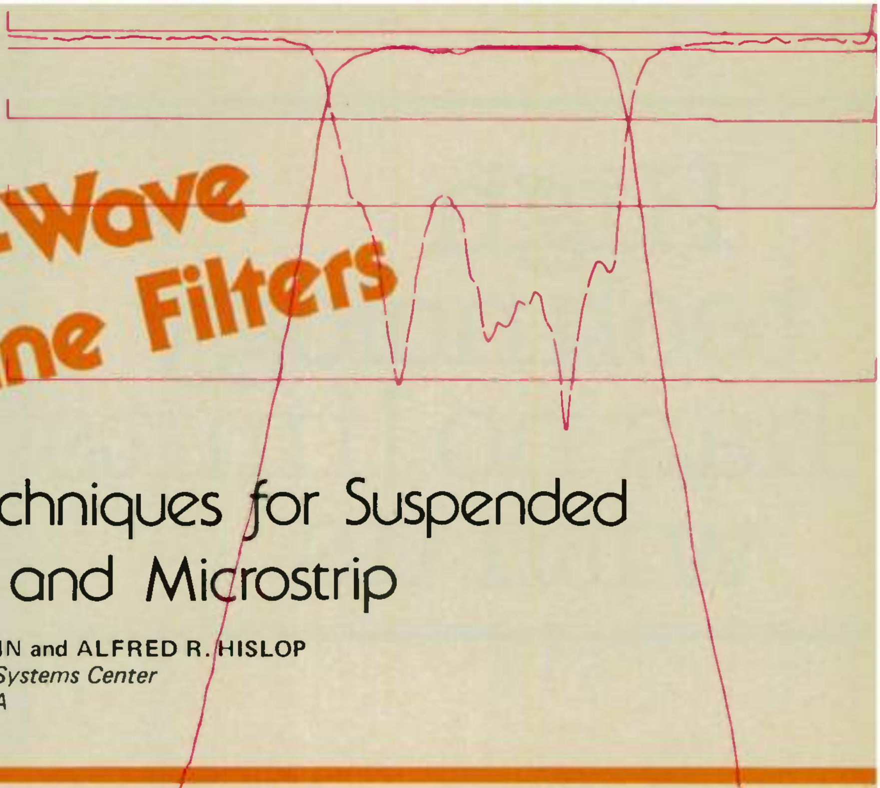

Millimeter-Wave Coupled Line Filters

Design Techniques for Suspended Substrate and Microstrip

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

INTRODUCTION

Low cost compact filters are an important aspect of millimeter-wave systems design. Almost all receiver systems require input filters or multiplexers for image rejection or channelization. A major obstacle to the production of millimeter-wave systems has been high cost. The use of printed circuits helps to overcome this obstacle and also allows for great reduction in size through component integration. Suspended substrate and microstrip are two such printed circuit media, with suspended substrate having the potential for lower loss.

This article should provide some useful techniques for the design of parallel coupled filters and diplexers in suspended substrate and microstrip. Printed circuit transitions have been fabricated to allow waveguide inputs to the circuits. The filters we have fabricated in suspended substrate utilize stripline equations and certain assumptions. The performance of these filters closely approximates the calculated performance. To avoid extremely narrow gaps in tightly coupled end sections of wideband filters, these sections can be modified to eliminate the gaps. Use has been made of a method for the compensation of the effects due to the discontinuities at the open ends of each filter section. Finally, diplexers can be designed by appropriately spacing two filters from a "T" junction.

TRANSITIONS

Figure 1 shows a half section of the suspended substrate channel used in our experimental work. A simple probe transition from waveguide to suspended substrate was found to work well from 26.5 to 40 GHz. The substrate material is .010 inch 5880

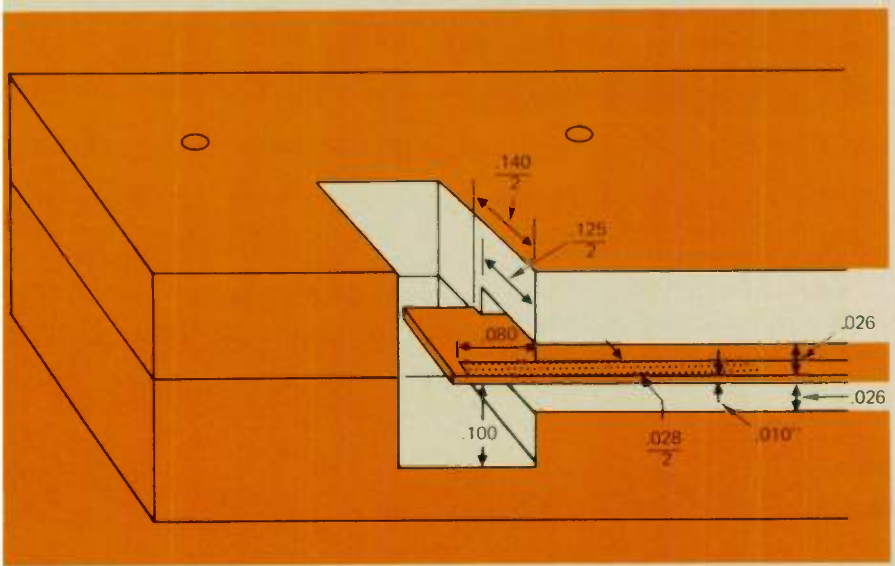

Fig. 1 Probe transition half section.

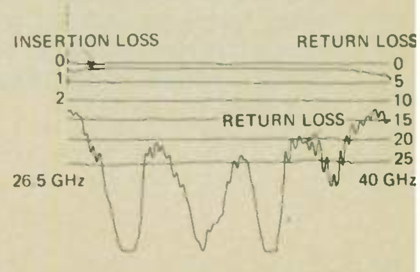

Fig. 2 Suspended substrate probe transition and line loss.

(continued on page 69)

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Duroid, copper thickness .0005" or less. The waveguide back short is non-adjustable. Insertion and return loss of two probe transitions and 1.4 inch of suspended substrate line are shown in Figure 2.

Printed circuit transitions from waveguide to microstrip have been previously demonstrated along with various coupled line filters and diplexers.[1] Both suspended substrate and microstrip transitions and filters were fabricated within channels which were cutoff to waveguide modes of the frequencies of interest. Microstrip has the advantage of having less dependence on spacing from the filter edges to the channel walls. Suspended substrate filters are more critical in that respect. Computer programs involving suspended substrate are available,[2,3] however, ones we know of calculate parameters for lines which are symmetrically located within and parallel to the walls of the channel. The coupled sections of a multisection filter are normally skewed within the channel, causing impedance and velocity changes in the coupled sections.

EXTENDING THE BANDWIDTH OF COUPLED LINE FILTERS

There are both electrical and physical limitations encountered when designing coupled line filters. One can design a wide bandwidth filter (25% or more) using appropriate lengths and even and odd mode impedances. Network analysis and computer graphics will show that the band edges do not have the ideal response (e.g., Tchebysheff or maximally flat). This is because the coupled-line sections are used to approximate ideal circuit elements. A review of Cohn's articles[4,5] will reveal the electrical approximations used.

Aside from electrical limitations, one will find that for wide bandwidth filters the end sections may require such physically close coupling that they may either be impossible to fabricate or be totally unrepeatable. As pointed out by Mosko[6] one may not need to have coupled end sections at all. Instead, they may be replaced by two quarter-wave sections of simple transmission line.

In order to understand why this is so, an example should prove illuminating. Figure 3 shows the development of a four-section coupled line filter starting from a three reactance element low pass prototype. A short description of each step follows:

A. The N=3 prototype: Note that prototypes are symmetric for odd N only. The procedure is not too different for even N. The prototype g values (Henrys, Ohms, Farads) are found from synthesis, the values given by tables or through formulas.

B. Passing from the low pass prototype to the band pass equivalent: The series and parallel elements are all resonant at the center frequency ω_0. Values of L and C are found by equating the impedances of the ladder circuits between 0 and 1 Hz with the impedances of the bandpass equivalent between ω_0 and

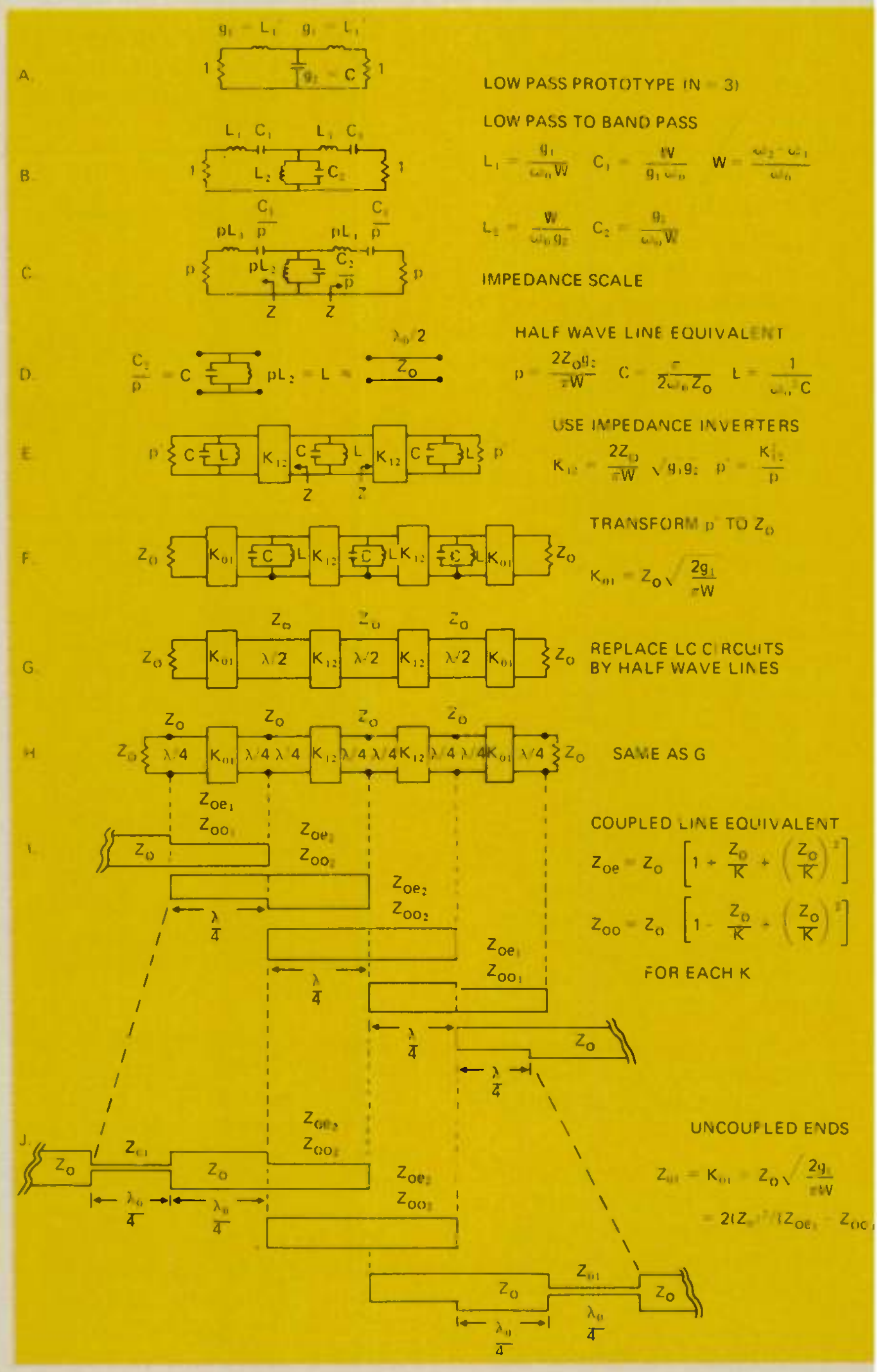


Fig. 3 Development of filter from prototype.

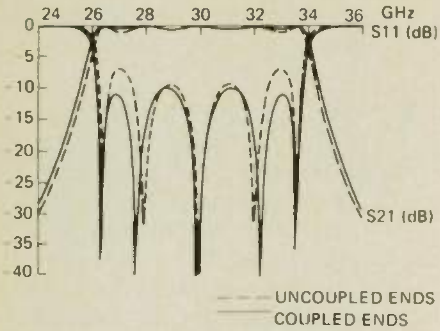


Fig. 4 Computed performance of 8 GHz BW filter with and without coupled end sections.

the upper band edge frequency ω_2. See e.g., Altman[7] for more details.

C. The circuit is impedance scaled: Resistances and reactances are all scaled up a factor p (to be determined).

D. A half-wave transmission line of impedance Z_o approximates a parallel tuned circuit if the latter is very lightly loaded. Conversely it approximates a series tuned circuit if heavily loaded. We take the case here of light loading. The parallel tuned circuit in step C can be approximated by a nalf wave distributed line of impedance Z_o if the p value is adjusted as given.

E. Using ideal impedance inverters all series tuned circuits can be converted to parallel tuned circuits. To have the same impedance characteristics the previous resistance p must be changed to a new value p′. Note that the inverter value K_{12} is picked so that the previous series tuned circuits are converted to parallel tuned circuits with the same L and C values as the single parallel tuned circuit given above.

F. A simple impedance inverter (transformer) K_{01} allows the use of resistance Z_o instead of p′.

G. All the parallel tuned circuits are loaded lightly by high impedance inverters and therefore can be approximated by half-wave lines of impedance Z_o.

H. This network is identical to G, however, the filter is broken up into four segments with each segment consisting of an impedance inverter between two quarter-wave lines of impedance Z_o. Cohn[4] has shown that the transfer characteristics of the two port formed by each segment closely approximate those of a coupled line section. This approximation requires that the even and odd mode impedances of the coupled line be dependent upon the impedance K of the inverter as shown in I.

I. This is the four-section coupled line equivalent of the three tuned circuit bandpass filter of step C.

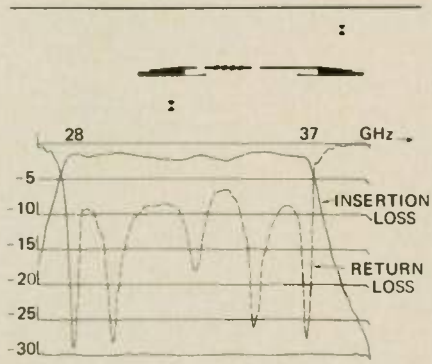


Fig. 5 Wideband microstrip filter with uncoupled end sections.

J. For wide bandwidth filters, the gap widths for the two end sections in step I may be too small to be easily fabricated. Instead of using coupled lines, the end sections may be replaced by quarter wavelength lines of impedance K_{01} followed by another quarter wave line of impedance Z_o. Note that the input and output quarter wave Z_o lines of step H are no longer required.

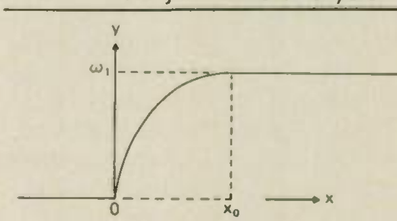
Figure 4 compares two six-section (N=5) Tchebysheff coupled line filters with design ripple = .5 dB. One of the two filters utilizes uncoupled end sections and has slightly less than ideal behavior. The filters are over 25% bandwidth. Similar filters with less than 15% bandwidth would be virtually indistinguishable from each other. In practice, SWR, inaccuracies in estimating even and odd mode velocities (in the case of suspended substrate), and losses, obliterate the design ripple. Figure 5 shows the measured results of a wideband microstrip filter with uncoupled end sections.

In the Appendix, a small program in BASIC gives the required even and odd mode impedances for double terminated Tchebysheff filters for any desired number of sections, frequency range, and ripple.

PRACTICAL DESIGN OF MICROSTRIP AND SUSPENDED SUBSTRATE FILTERS

Unless provisions are made to eliminate or reduce the effects of open end step discontinuities, coupled line filters will appear lower in frequency and exhibit less than ideal behavior. For microstrip coupled lines, we have used tables generated by the Bryant and Weiss MSTRIP[8] program to determine the gaps and widths of each coupled section. Knowing the even and odd mode velocities, the coupled lengths were determined by the Dell-Imagine formula,[9] which differs slightly from that obtained from the average of the two velocities. Given the widths of the coupled sections, the open end fringe capacitances can be determined from Silvester and Benedek.[10] It is necessary to incorporate these end capacities into the coupled line four port[11] to find the equivalent two port used in most network analysis.

Using the above design methods, microstrip coupled line filters were subjected to analysis



GIVEN: WIDTH ω_1; THICKNESS h; REL. DIEL. CONST ϵ_r; FRINGE CAP/LENGTH C_f

LET $\omega_0 = \dfrac{h\,C_f}{\epsilon_r}$

AND $K = 1 + \dfrac{\omega_1}{\omega_0} + \sqrt{\dfrac{\omega_1}{\omega_0}\left(2 + \dfrac{\omega_1}{\omega_0}\right)}$

THEN $y = \omega_1 - \dfrac{\omega_0}{2Ke^{\frac{-x}{\omega_0}}}\left(Ke^{\frac{-x}{\omega_0}} - 1\right)^2$

AND $x_0 = \omega_0 \ln K$

Fig. 6 Elimination of step discontinuities.

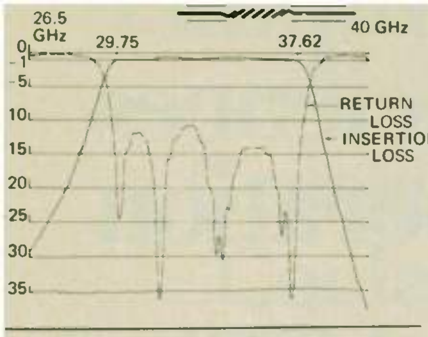


Fig. 7 Eight GHz bandwidth suspended substrate filter.

and graphics and the shift to lower frequencies determined. All filter lengths were then decreased by the proportional amount derived from this frequency shift.

A simpler and more effective way to eliminate the effect of step discontinuities is to eliminate the steps themselves. This was shown recently by Malherbe and Steyn[12] for the case of stripline. Their method alters the step shape which is used when one impedance line joins another. The total capacitance associated with each incremental strip along the propagating direction consists of parallel plate capacitances between strip and grounds, and edge capacitance. By modifying the shape of the edges so that all incremental strips have the same capacitance, the impedance stays the same up to the edge discontinuity. We have used this method for both microstrip and suspended substrate by compensating for different edge capacitance/unit length. Figure 6 gives the shape of the compensating curve using equations that reduce to those of Malherbe. The capacitance per unit length, C_f, has been derived by Cohn[13] for the case of stripline and is $C_f = .043\ \epsilon_r$ pFd/inch for zero thickness lines where ϵ_r is the relative dielectric constant of the substrate. Experimentally, for .010" Duroid, we have had the closest predicted frequency response with filters using C_f = .17 pFd/inch for microstrip, and .13 pFd/inch for the suspended substrate configuration shown in Figure 1.

The method we have used to experimentally determine the even and odd mode impedances and velocities for suspended substrate is not precise, but very useful.

- Design a simple coupled line filter (using the program given in the Appendix) as if the dielectric did not exist. Design for a frequency near the higher end of the band since the dielectric will lower it considerably. Use the coupled strip impedances of Cohn[14] with $\epsilon_r = 1.0$.
- Use the step compensation given by Malherbe with C_f that of Cohn.
- Test the filter for frequency response. The effective dielectric constant will be $\epsilon_r' = (f_o/f_o')^2$ where f_o = design center, f_o' = measured filter center.

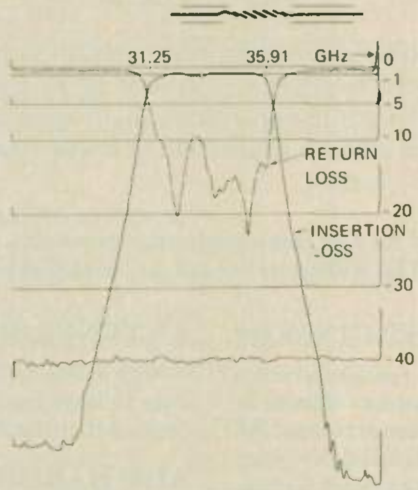


Fig. 8 Measured performance of 4 GHz bandwidth suspended substrate filter.

- Scale all impedances so that $Z' = \sqrt{\epsilon_r'}$ where Z' is the determined suspended substrate impedance and Z is the impedance with $\epsilon_r = 1.0$.
- Design desired filter using the above determined dielectric constant and impedances. It may be necessary to change the value used for C_f to obtain the desired response.

Figures 7 and 8 show two six-section suspended substrate filters designed for .5 dB ripple and bandpass ranges of 30-38 and 32-36 GHz, respectively. The 8 GHz bandwidth filter did not require uncoupled end sections since the lower effective dielectric constant of suspended substrate generally leads to wider line widths and gaps than those of equivalent microstrip filters. For higher frequencies, where dimensions are smaller, uncoupled end lines may be required. As will be seen, these type filters are almost ideal for some T junction diplexers.

WIDE-BAND T JUNCTION DIPLEXERS

Previously, we have described coupled line microstrip diplexers[15] which were used to separate contiguous or non-contiguous frequency ranges of about 4 GHz in K_a band. Computed lengths of Z_o line connected each bandpass filter to the common T junction so that the extremely large out-of-band susceptance of each filter appeared, at the junction, to be almost zero at the band center of the other.

When decoupled end sections are used, the added 90° line section nearest the junction reduces the out-of-band susceptance drastically. Depending on the actual filters and frequency bands used, it may not be necessary to add any line at all to minimize the out-of-band susceptance. Figure 9 shows to contiguous 8 GHz bandwidth suspended substrate filters connected together at their common junction without added diplexer lines. The reader should be forewarned, however, to use computer analysis on any diplexing structure before fabrication. In the case of contiguous diplexers, the susceptances at the cross-

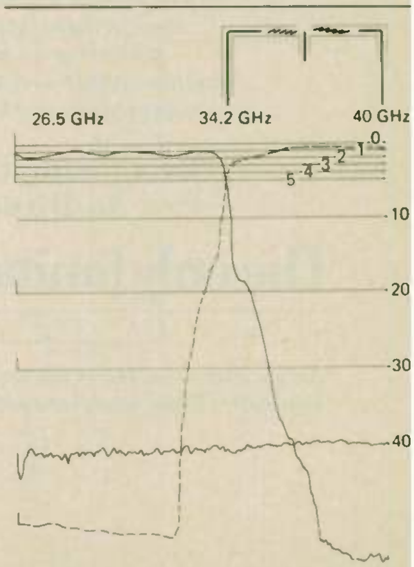


Fig. 9 Measured response of suspended substrate diplexer.

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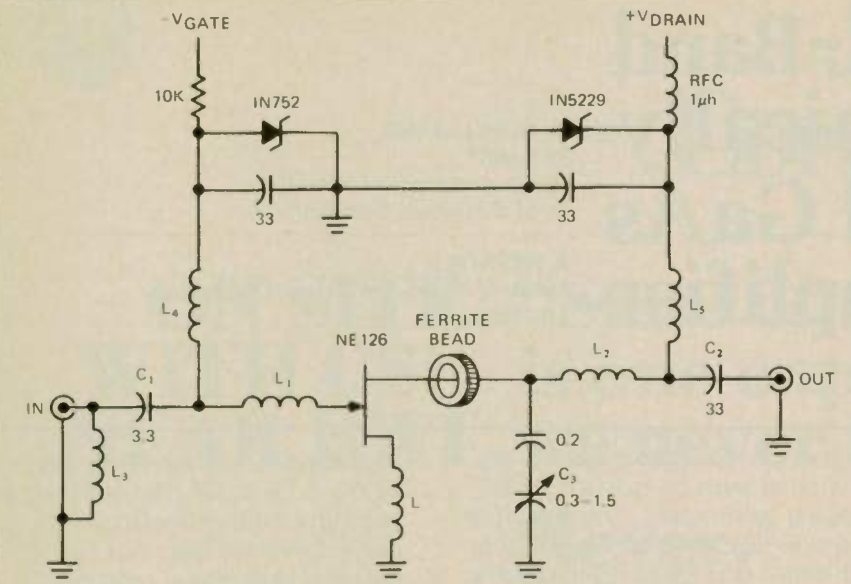

Fig. 2 Schematic diagram of one of two identical stages in the cooled amplifier. All inductors are made with .008" diameter phosphor-bronze wire wound to make a .080" inner diameter coil. L_1 and L_2 are 3 turns with .100" length and ∿10nh inductance. L_3 and L_4 are 4 turns with .125" length and ∿16nh inductance. L_5 is 5 turns and is non-critical. Note that L_3, L_4, and C1 comprise a 500 MHz high-pass filter. All capacitor values are in pf. C_3 is a Johannson Giga-Trim type 7264 variable capacitor. The ferrite bead is Micrometals Type T10-6. L is the added source inductance ∿1nH (see text).

$$g_n = \frac{K_g\ (\omega C)^2}{g_m}$$

$$r_n = R_g + R_s + K_r/g_m$$

where T_{min} is the minimum noise temperature, T_O = 290°K, $R_{opt} + j\ X_{opt}$ is the optimum source impedance, g_m is the transconductance, C is the gate to source capacitance, $f_T = g_m/(2\pi C)$ is the gain-bandwidth product, K_g, K_r, and K_c are dimensionless factors defined by Pucel, et al.,[18] g_n is the current noise conductance, r_n is the voltage noise resistance, and $R_g + R_s$ is the sum of gate and source resistances external to the channel. (Resistance in the channel produces noise through g_n.) For a typical Nippon Electric Company NE 126 at 290°K, g_m = .027, C = .45 pF, f_T = 9.5 GHz, r_n = 13, K_g = .14, K_c =1.8, K_r = .06 and T_{min} = 129° x f/f_T.

At a frequency of 5 GHz the above theory gives a noise temperature of 68° (0.9 dB noise figure) which is somewhat lower than measured[10] for the NE 126. However, the noise temperature predicted at 1.3 GHz is 18° which is a factor of 2.7 lower than we have measured. This discrepancy cannot be explained by circuit losses (< 10°K) and most likely is due to another noise mechanism not contained in the present theory.

At a temperature of 20°K the thermal noise generators in a FET are reduced in power by a factor of 15 and probably have a negligible contribution to the total noise. The transconductance increases by 40% due to an increase in saturation velocity[10] and this decreases the effect of non-thermal noise in the drain circuit. The temperature dependence of the non-thermal noise (i.e., hot-electron noise and high-field diffusion noise) is not known at present but is not likely to be very large. Some further discussion of the noise temperature theory at cryogenic temperatures is given in Reference 10.

EFFECT OF SOURCE INDUCTANCE

The purpose of the added inductance in the source lead of the FET is to apply negative feedback which increases the real part of the input impedance to obtain a near-matched condition.[19] Considering the simple equivalent circuit of Figure 1, an input current, i, causes an e_g of $i/j\omega C$, a drain current of $g_m \cdot i/j\omega C$, and a voltage drop across L of $j\omega L \cdot g_m \cdot i/j\omega C$. Thus the effect is the same as adding a frequency-independent resistance, $R_{FB} = g_m\ L/C$, to the input. This result is valid for $\omega C << g_m$ and for an output load impedance much less than the FET output impedance. If the latter is not true R_{FB} is multiplied by the complex fraction of the drain current generator flowing through the output load.

The addition of L has negligible effect upon noise temperature. As shown in a very general way by Haus and Adler[20] the addition of lossless elements to a network does not change the minimum noise measure, $M_{min} = T_{min}/[T_o\,(1 - G^{-1})]$, of the network. Thus, in the case of high G, adding an inductance may lower G and slightly lower T_{min} if the source impedance is reoptimized. In the particular case of the FET, Vendelin[21] has shown that for small L the effect upon optimum source impedance is to reduce

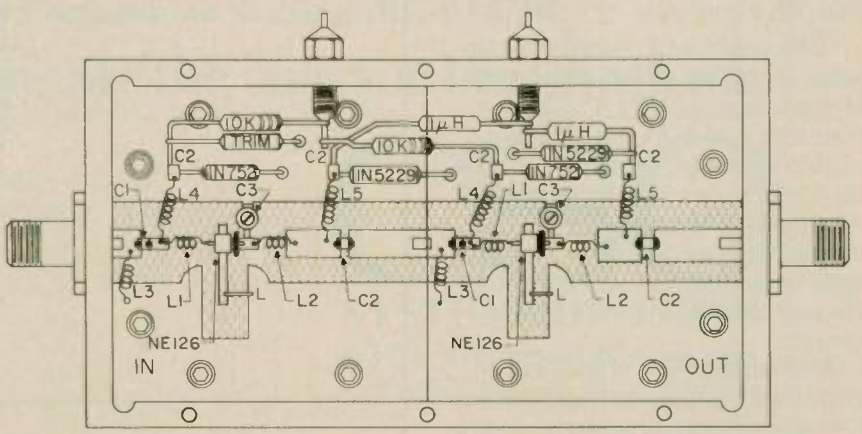

Fig. 3 Physical layout of the two-stage amplifier. Circuit board is .062" teflon-glass board Rogers Duroid type 5880. Each stage is tested separately for best noise performance and input and output return loss before assembly into the two-stage amplifier.

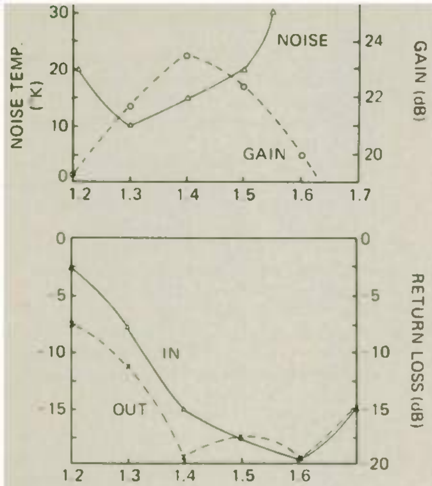

Fig. 4 Measured gain, noise temperature, input return loss, and output return loss for two-stage amplifier in the 1.2 to 1.7 GHz frequency range.

the reactive part by ωL; as far as noise temperature is concerned the effect is to add L (∿lnh) in series with the gate.

AMPLIFIER CONSTRUCTION

A two-stage amplifier has been built embodying the above principles, using an NE 12683 packaged FET, and cooled to 18°K in a closed-cycle helium refrigerator.[1] Each amplifier stage utilizes the source inductor matching technique described above to achieve a 50 ohm input and output impedance, prior to cascading the two stages. The advantage of this technique is that it eliminates the need for an input cryogenic isolator with its large thermal mass and its associated losses.

The circuit in **Figure 2** shows the series tuned input configuration used, in which the input inductor L1 is found to determine the frequency at which the noise minimum occurs. In a separate measurement, a variable quarter-wave transformer was used ahead of the input to determine the value of the optimum source resistance, R_{opt}, which was found to be 70 ohms. However, as a practical consideration a source resistance of 50 ohms is close enough to R_{opt}, avoids the loss of an input transformer, and improves the match. **Figure 3** shows the physical layout of the lumped microwave circuit and biasing arrangement on a teflon circuit board. A teflon circuit board rather than a ceramic substrate is utilized to avoid substrate and solder joint fracture which may result from temperature cycling of too rigid a structure. It will be noted that the source inductance L is achieved by the use of the source tab of the packaged FET together with a small grounding wire. The small variable capacitor on the output of the FET is adjusted together with L2 for best output match. Note also the use of the ferrite bead on the drain tab to help quench parasitic oscillations at high gigahertz frequencies (additional lossy material is sometimes needed in the lid of the amplifier box; 1/8" sheet Radite 75 is used). All solder connections for the cryogenic amplifier were made using a low-temperature Indium solder (Indalloy #3 10% Ag, 90% In). Zener diodes are employed on the gate and drain of the FET to protect from overvoltage static. Selection was made among various manufacturers of FETs and also cular type. Twelve individual single stage amplifiers were built and tested and the ones with best cryogenic performance were selected for the two-stage NE 126 amplifier described here. Some correlation was found between the best manufacturers noise figure test data and best cooled performance.

The two-stage amplifier was attached directly to the 18°K station of the refrigerator as shown in **Figure 5**. A special low-loss 4-inch long input air-line was fabricated to bring the signal through the dewar wall to the 18°K input of the amplifier. The 1/4 inch diameter air-line was constructed from thin wall (.006") stainless steel; the inner wall of the outer and the outer wall of the inner conductors were copper plated and gold flashed to reduce microwave losses. The estimated noise contribution from this line is 1°-2°K in the cold condition, and its contribution is included in the overall measured amplifier temperature.

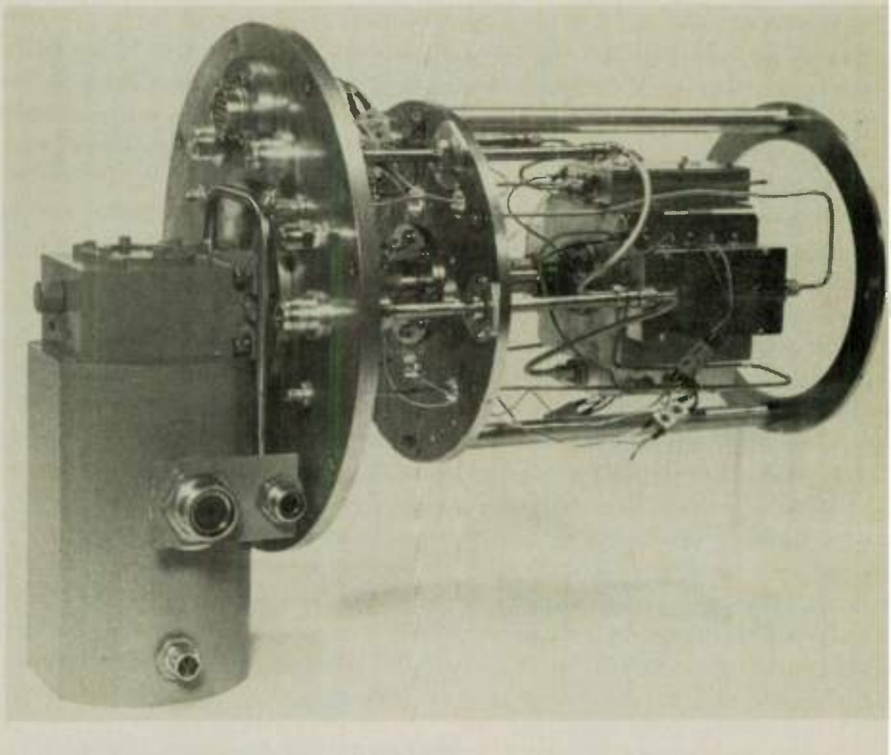

Fig. 5 Photograph showing four two-stage amplifiers attached to the 15K station of the CTI Model 21 refrigerator, at left. The outer dewar and the heat shield (attached to the 77°K station) have been removed to show assembly. The low-loss input lines (4) are gold-plated thin-wall stainless steel. They pass through the dewar wall on Type-N connectors to SMA at the amplifier input and are heat sunk to the 77°K station at mid-point. The output lines are .085 stainless steel semi-rigid coax.

TEST RESULTS

At optimum frequency, room temperature amplifiers exhibit noise temperatures of 48° ± 3°K without any corrections for input losses or second stage contribution; the input and output return losses are > 20 dB. Over the 1.3 to 1.6 GHz band the noise temperature is below 60°K and the return losses > 10 dB. A similar performance is also obtained with other FETs, notably some selected NE 388, NE 218 and Mitsubishi MGF 1412 FETs, and Avantek AT8110.

At cryogenic temperatures we measure a minimum amplifier noise temperature of 13° ± 1°K at 1.3 GHz, and under 20°K from 1.2 to 1.5 GHz (see Figure 4) on one of a pair of amplifiers built. Over the frequency range 1.3 to 1.6 GHz the gain is greater than 20 dB. The input return loss is > 20 dB at 1.6 GHz and > 10 dB over the 1.35 to 1.7 GHz range. Other manufacturers of FETs which are found to give cooled noise performance in the 20°K range are the Plessey GAT5, NEC NE 218 and Mitsubishi MGF 1412. The latter two FETs are recommended for new cooled designs.

The amplifiers described here are used in a radio astronomy application on the 25 m radio telescope of the University of California at Hat Creek Observatory. A total of four amplifiers provide polarization diversity for spectral line observation in the H and OH radio-astronomy bands at 1.42 and 1.66 GHz respectively.

CONCLUSION

An amplifier with very low noise in L-band has been developed. It is easily constructed and has proved reliable in the 20 units so far constructed. Some of the units have been temperature cycled to cryogenic temperatures 20-30 times with no reported electrical or mechanical failures.

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STANDARD DESIGNS

RF COAXIAL SWITCHES

DC TO 18 GHz

- SPDT - CS-33 SERIES

Type	Model No.
Failsafe	CS-33S10
Failsafe w/indicators	CS-33S1C
Latching	CS-33S6D
Latching w/indicators	CS-33S6C

- TRANSFER - CS-37 SERIES

Type	Model No.
Failsafe	CS-37S10
Failsafe w/indicators	CS-37S1C
Latching	CS-37S6D
Latching w/indicators	CS-37S6C

- MULTI THROW - CS-38 SERIES

Type	Model No.
SP3T Basic Unit	CS-38S13
SP3T w/indicators	CS-38S13C
SP4T Basic Unit	CS-38S14
SP4T w/indicators	CS-38S14C
SP5T Basic Unit	CS-38S15
SP5T w/indicators	CS-38S15C
SP6T Basic Unit	CS-38S16
SP6T w/indicators	CS-38S16C
SP7T Basic Unit	CS-18S17*
SP7T w/indicators	CS-18S17C*
SP8T Basic Unit	CS-18S18*
SP8T w/indicators	CS-18S18C*

*Operating frequency limited to 12 GHz.

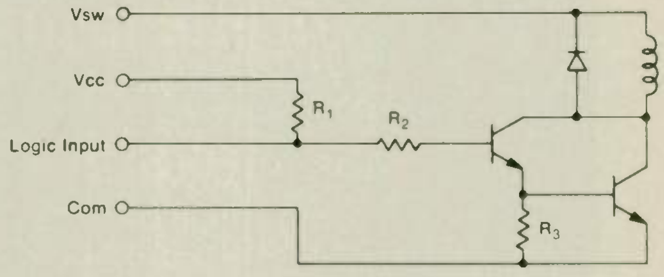
Larger size units with N or TNC Connectors, operating DC-12 GHz, are available.

TTL SWITCH DRIVERS

As a special option, on both failsafe and latching type switches, drivers can be provided which are compatible with industry standard low power Schottky TTL circuits.

Two options are provided as follows:

1. All units are provided with a 5 volt (Vcc) connection and an internal pull-up resistor (R1). When the 5 volt connection is made, the Logic Input current drain closely resembles two low power Schottky TTL loads (40 μA).
2. If a high level Logic Input current drive (450 μA @ 2.4 Vcc) is available, the 5 volt (Vcc) connection need not be made.



R_1 = 6.2K ohms
R_2 = 2.7 - 6.3K ohms

Switch requires one of the above drivers per position (except Failsafe). VSW, Vcc, and Com terminals are common to all positions.

SPECIAL FEATURES AVAILABLE

- Special Actuator Voltages, i.e., 12 VDC, 15 VDC
- MS Connectors Can Be Installed On Most Models
- Arc Suppression Diodes

RF PERFORMANCE

	CS-33, 38 SERIES			CS-37 SERIES		
Frequency	0-6 GHz	6-12 GHz	12-18 GHz	0-4 GHz	4-12 GHz	12-18 GHz
VSWR (max.)	1.25:1	1.40:1	1.50:1	1.25:1	1.40:1	1.50:1
Insertion Loss (max.)	0.2 dB	0.4 dB	0.5 dB	0.2 dB	0.4 dB	0.5 dB
Isolation (min.)	70 dB	60 dB	60 dB	70 dB	60 dB	60 dB

TELEDYNE MICROWAVE

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OTHER PRODUCTS: • Circulators and Isolators • Multiplexers and Integrated Components • VCOs over frequency can often be made to subtract from each other, generally decreasing the SWR and allowing the lowest loss at the crossover.

CONCLUSIONS

Both suspended substrate and microstrip bandpass filters and diplexers can be readily constructed in below cutoff channel housings. In general, suspended substrate is preferable because of lower loss. However microstrip, with the availability of good design data and ease of parts integration, may be preferable where 1 or 2 dB of additional loss can be tolerated.

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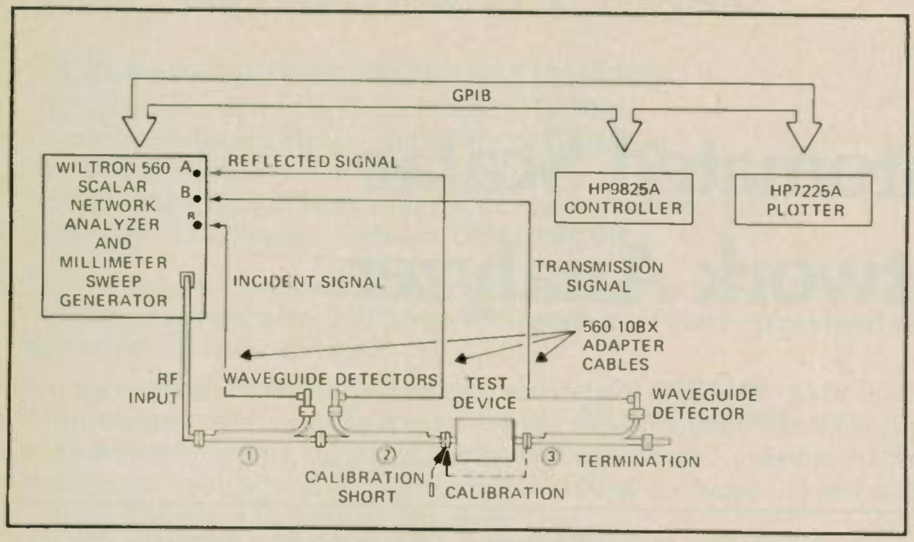

Fig. 3 Millimeter waveguide reflectometer setup for measurements above 34 GHz. Available components are listed in Table 2.

TABLE 1

REFLECTOMETER COMPONENTS FOR USE BELOW 40 GHz

Freq. Range (GHz) (1)	Maury 10 dB Directional Coupler Model No.	Maury (2) Waveguide/Coax Adapters Model No.	Wiltron RF Detectors Model No.	Maury Termination Model No.
8.2 - 12.4	X410C	X210D	560-7S50	X301A
10 - 15.5	M410C	M210D	560-7S50	M301A
12.4 - 18.0	P410C	P210D	560-7S50	P301A
15 - 22	N410C	N210D	560-7S50 Opt. 2	N301
18 - 26.5	K410C	K210C	560-7S50 Opt. 2	K301
26.5 - 40	U410C	U210C N/A	560-7S50 Opt. 3 to 34 GHz Hughes 44820H-500 waveguide mount above 34 GHz	U301

(1) Waveguide couplers are also available below 8.2 GHz.

(2) Adapters have SMA or MPC-3 (SMA compatible) coaxial output connectors. Other adapters are available in frequency ranges to match the available directional couplers.

TABLE 2

MILLIMETER REFLECTOMETER COMPONENTS FOR USE ABOVE 40 GHz

	10 dB Directional Couplers		Detectors & Mounts		
Freq. Range (GHz)	Baytron Model No.	Hughes Model No.	Baytron Model No.	Hughes Model No.	Hughes Fixed Termination Model No.
40 - 60	3Q-40/10	45333H-2010	1Q-6F 1Q-6/X	47323H-2200	45613H-2000
60 - 90	3E-40/10	45335H-2010	1E-5W 1E-5/X	47325H-2200	45615H-2000
75 - 110	3R-40/10	45336H-2010	1R-5W 1R-5/X	47326H-2200	45616H-2000

by subtracting R from A (A-R mode) and storing the resultant in the network analyzer memory. Later, the resultant is subtracted from the actual measurement data to enhance accuracy.

For 0 dB transmission calibration, couplers (2) and (3) are connected directly together, and the resultant is stored in B-R memory. In this setup, coupler (3) is recommended for the most accurate transmission measurements. Because the waveguide termination provides a mainline match of greater than 40 dB return loss, error-producing reflections back through the DUT are very small and the measurement can be made accurately. The effect of mismatch introduced by the adapters and detectors on couplers (1), (2), and (3) is small. Typical test results are shown in **Figure 2**.

For measurements above 34 GHz, the recommended setup is shown in **Figure 3**. Notice that the only difference between **Figure 3** and **Figure 1** is that above 34 GHz waveguide detector mounts and Wiltron 560-10BX Adapter Cable are used in place of waveguide-to-coax adapters and the WSMA detectors. Because the SWR of millimeter waveguide detector mounts is very high, it is essential that coupler (3) be used as shown. The adapter cable interfaces the BNC output of the waveguide detector mounts to the network analyzer inputs. Built into the adapter cable are resistors which simulate the thermistor and offset impedances of the coaxial detectors and which are required to maintain the balance of bridge circuits in the network analyzer. Under these conditions, the network analyzer closely maintains with the waveguide detectors the logarithmic conformity and measurement accuracy that is achieved with the coaxial detectors in **Figure 1**.

A few of the manufacturers that supply waveguide directional couplers with 40 dB directivity are listed in **Tables 1** and **2**. One source for the more difficult to find detector mounts is given. The Wiltron Network Analyzer is compatible with all presently available sweep generators. The 5610 system shown in **Figure 1** is available with sweep generator plug-ins operating up to 40 GHz. For applications above 40 GHz, Hughes offers a plug-in sweeper operating up to 110 GHz.

The availability of components and an automated network analyzer with required interface cables allows waveguide measurements up to 110 GHz with ease and accuracy.

INTELCOM '80
Components for Communications

10-13 November
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The role of semiconductor development in shaping the future of telecommunications will be the focus of eight special sessions during the INTELCOM '80 Conference which will be held at the Convention Center in Los Angeles, California from November 10-13. One Executive Forum, an Application Panel and six Components Symposia sessions will be held.

Technology in the predominantly analog telecommunications industry has been lagging considerably behind technical advancements in the computer industry, which has benefited from the immediate application of innovations in semiconductor devices.

Recent growth in digital communications, where analog voice is converted to digital numbers for transmission and switching, resulting in operational and economic advantages coupled with computer controlled switching designed to provide feature flexibility, has created a need for more semiconductor technology applications. In particular, microprocessor and memory elements, which provide considerable processing power and large storage capacity at minimum cost — along with high packaging density — are finding rapid acceptance in the telecommunications industry.

Along with these standard computer devices, the digitization of voice has created a large demand for new components, such as CODEC and filters. Due to this digitization of voice, semiconductor development will play a significant role in the future of the telecommunications industry by providing highly reliable subsystem elements in LSI packages.

EXECUTIVE FORUM

Tuesday, 11 November
9:00 am—11:30 am, Room 217A

The Infrastructure Needed to Develop a Telecommunications Industry
C. Lester Hogan
Former President
Fairchild Camera and Instrument Corporation
Speakers include:
Richard Lee, President
Siliconix

APPLICATION PANEL

Tuesday, 11 November
3:00 pm—5:30 pm, Room 216A

The Role of Semiconductor Technology in Shaping the Future of Telecommunications
S. G. Pitroda
Vice President & Managing Director
WESCOM Switching

The objective of this Panel is to explore the role of semiconductor development in the changing telecommunications environment from predominantly analog to integrated digital transmission and switching networks.

Robert John Paluck
Vice President of Data and Communications Products
Mostek Corp.

Tom Longo
Vice President
Semiconductor Operations
Fairchild Industries

Earl Rogers
President
PMI

Richard Pieranunzi
Manager-Strategic Marketing
Communications Segment
Motorola Semiconductor Group

COMPONENTS SYMPOSIA

Monday, 10 November
3:00 pm—5:30 pm, Room 216C

Custom LSI/VLSI Design for the Systems Engineer
Douglas Fairbairn
Publisher
LAMBDA Magazine

This session examines custom integrated circuit alternatives for electronic systems.

Alternatives for Semi-Custom LSI
Robert Hartmann
President
IC Cost Consultants

Hierarchical Design for VLSI
James A. Rowson
Research Fellow
Computer Science Dept.
California Institute of Technology

A Description of a Single-Chip Implementation of the RSA Cipher
Ron Rivest
MIT

The Role of VLSI Implementation Systems in Interfacing the Designer and Fabricator of VLSI Circuits
Alan Bell
Xerox Palo Alto Research Center

Fast Turnaround Fabrication for Custom VLSI
Gunnar Wetlesen
Vice President-Engineering
VTI

Tuesday, 11 November
3:00 pm—5:30 pm, Room 216C

Satellite Earth Station Components
Lou Cuccia
Manager, Technical Development-Transportable/Mobile Ground Stations
Ford Aerospace and Communications Corp.

The modern commercial earth terminal has benefited from the development of new subsystems and components using the field effect transistor (FET), new TWT developments, and the use of the microprocessor and mini-computer. These new components and subsystems now provide sensitivity, radiated power, and terminal and network control for optimal information handling with a cost-effectiveness not possible only five years ago. This session will provide an update on these key components and subsystems which are causing a virtual revolution in user acceptance of satellite communications.

The FET 110 Satellite Earth Terminals
Jerry Arden
President
Neil Corpron
Manager of Engineering
California Eastern Laboratories

K-Band Paramp LNA's For Satcom Earth Terminal Applications
E. Ng
J. Degruyl-Vice President,
Research & Development
Leonard J. Steffek-
Dept. Head of Microwave Technology
H. C. Okean
LNR Communications, Inc.

Modern 10-20 W TWTA's for High-Capacity Microwave Links and Satellite-Communications Earth Stations
J. Boulange
J. P. Desne
Division Tubes Electroniques
Thomson-CSF

The Microprocessor and the Mini-computer for Earth Terminal and Network Control
Carl Hellman
Manager of Earth Terminal Engineering
Western Development Labs.
Ford Aerospace and Communications

Wednesday, 12 November
9:00 am—11:30 am, Room 216C

Tubes for Communications Systems
Chet Lob
Technical Director
Varian

This session looks at the latest developments in microwave tubes for terrestrial and satellite communications.

Amplifier Klystrons for WARC 79 Up-Link Frequencies
Earl McCune
Darrell Green
Varian

(continued on page 87)

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Millimeter Waves Amplifiers: A Survey of Device Choices
Jack Ruetz
Varian

A 250 Watt K_u-Band TWT for Satellite Communication
Richard Ohtomo
Hughes Electron Dynamics Division

A 750 Watt K_u-Band TWT for Satellite Communication Systems
Ronald Leborgne
Hughes Electron Dynamics Division

Wednesday, 12 November
3:00 pm–5:30 pm, Room 216B

Digital Modems
Estil Hoversten
Vice President-Development
LINKABIT

Speakers include:
Loic Guidoux
Head of Modem Research and Development Dept.
TRT

James Bright
Director of Marketing
Novation Inc.

Wednesday, 12 November
3:00 pm–5:30 pm, Room 216C

Fiber Optics Components in Communications Systems
Hank Maas
National Accounts Manager
Lightwave Communications Systems
Harris Corp.

Optical sources, detectors, couplers, and multiplexing are some areas that will be addressed. Specifically injection laser diodes, avalanche photodiodes, light emitting diodes and wavedivision multiplexing.

Speakers include:
Peter Bark,
Siecor

Robert Chow,
Nippon Electric Company, Ltd.

Alex Ceruzzi,
Laser Diode Corporation

Thursday, 13 November
3:00 pm–5:30 pm, Room 216C

Solid State Microwave Sources: The Next Decade
Carl Sirles
Manager of RF Component Development
Collins Transmission Systems Division
Rockwell International

The 60s saw the development of the basic technology used in today's solid state microwave sources. The 70s brought many refinements to this technology. As the 80s begin, new technologies are emerging which promise higher reliability and lower cost for this important class of components. This panel session will present an industry perspective on what system designers might expect from solid state microwave sources in the 80s.

J. W. Sedin
Vice President & General Manager
Frequency Sources, West Division

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R. M. Knowles
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48005B

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HEWLETT PACKARD

Designers Tip

A New Type of Dual Mode Circular Cavity Filter

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Frequency Engineering Laboratories
Farmingdale, NJ

Filters with circular cavities, TE111 mode excited, (**Figure 1**) are commonly used in microwave applications with cavities of the single TE111 mode type, as well as TE111 dual mode type (1). The structure of a conventional 6-pole, 3-cavity dual mode filter is shown in **Figure 1(b)**. Typically:

- Consecutive irises are positioned 90° to each other.
- Every dimension of the filter greatly exceeds the diameter of a single cavity.
- The input and output are rectangular waveguides.

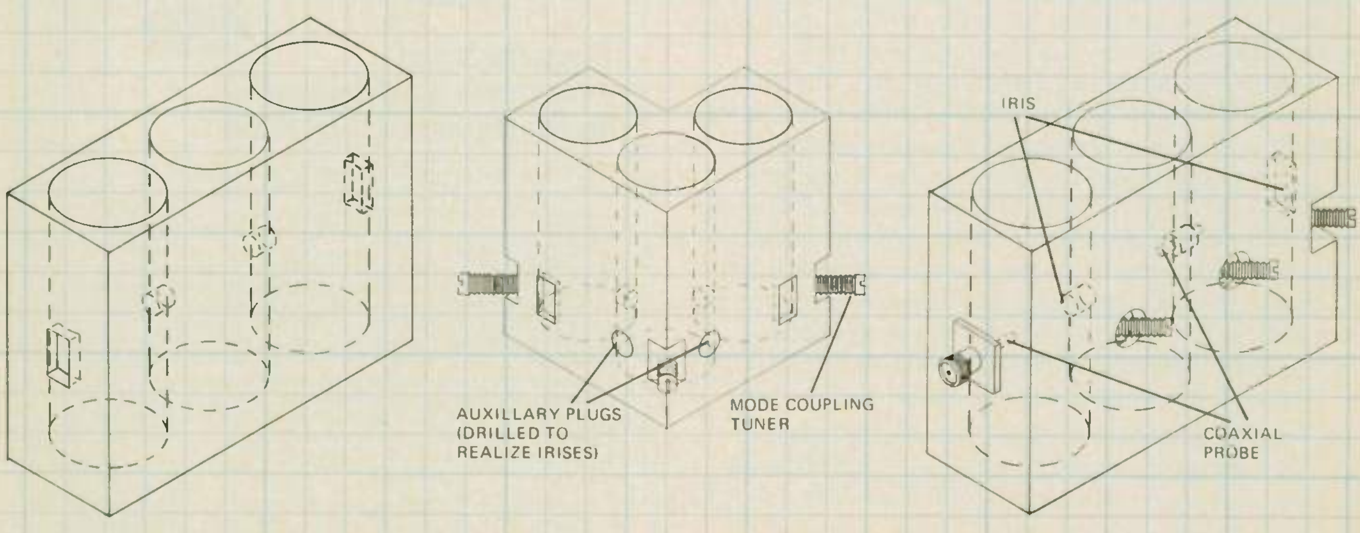
The first characteristic is the greatest burden in the construction of this type of filter. Drilling of the irises is performed in different directions in several steps. During this phase of construction, necessary auxiliary holes are being drilled, to be plugged eventually as shown in **Figure 1(b)**.

The single TE111 mode cavity filter shown in **Figure 1(a)** is simple to construct but requires twice as many cavities to match the dual mode filter (**Figure 1(b)** out of band rejection properties. Combining: (1) Volume efficiency of dual mode filters. (2) Manufacturing simplicity of single TE111 mode filters. The result is an attractive new filter.

The E-Field pattern in a circular cavity supporting a TE111 mode and the E-Field pattern excited by a coaxial probe in the same cavity, are shown in **Figure 2(a)** and **2(b)**. The orthogonal nature of the two E-Field patterns explains the difficulties and the inefficiency encountered in attempting to drive microwave signals through the cavity in **Figure 2**. Indeed, to a first approximation, only the deviation from perfect symmetry causes energy transfer at all. A mode coupling tuner positioned as shown in **Figure 2(c)** or as in **Figure 2(d)**, will couple the E-Field excited by the coaxial probe, to a TE111 mode supported in the same cavity, resulting in a cavity operating in the dual mode.

The mode coupling tuner in **Figure 2(d)** is located a distance away from the coaxial probe, in an E-Field with more resemblance to the E-Field of the conventional dual mode (**Figure 1(b)**. Therefore, one may expect the configuration in **Figure 2(d)** to demonstrate superior dual mode quality, as compared to the configuration in **Figure 2(c)**.

A trial filter was constructed, consisting of two cavities (1.13 inch diameter), designed to yield, 35 MHz 3 dB bandwidth, at 8.0 GHz. The mode coupling tuners, inserted into the cavities, effectively couple an E-Field supported by the coaxial probe in the cavities to a TE111 mode. As expected,



(a) 3-Pole filter single TE111 cavities

(b) Conventional 6-pole dual mode TE111

(c) New 6-pole filter dual mode TE111 cavities

Fig. 1 Circular cavity filters side wall coupled.

the two modes, coupled and "folded,' in each cavity yield a dual mode operation, demonstrated simultaneously by the numbers of minimas in the return loss and the out of band rejection skirt of the filter. The configurations in **Figure 2(c)** and 2**(d)** were tested, both yield comparable results. It may well be that the difference between these two configurations will become more pronounced for cavities with smaller diameters.

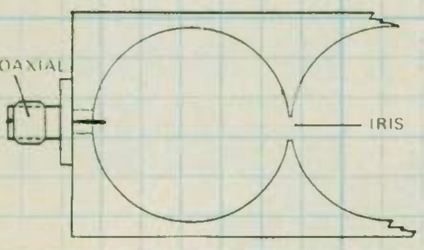


Fig. 2 Initial coupling structure, in the input (and output) cavity.

An equivalence was demonstrated between an iris in a side wall of a circular cavity to a coaxial probe (instead of the iris) with a mode coupling tuner properly positioned in the same cavity. This equivalence suggests a method for building a new type of dual mode filters.

The new dual-mode, filter shown in **Figure 1(c)** contains none of the structural disadvantages, typical of the "old" conventional dual mode filters, is simpler and eventually cheaper to construct, lending itself to more versatile applications. Together with the "old" dual-mode filter and a possible hybrid combination between the two, they make an effective, flexible tool to address microwave filtering applications.

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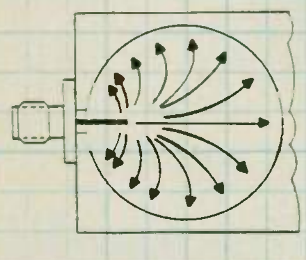
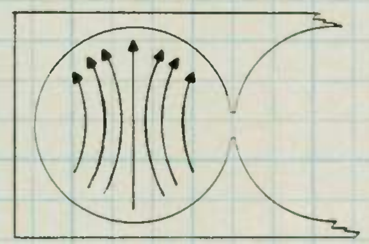
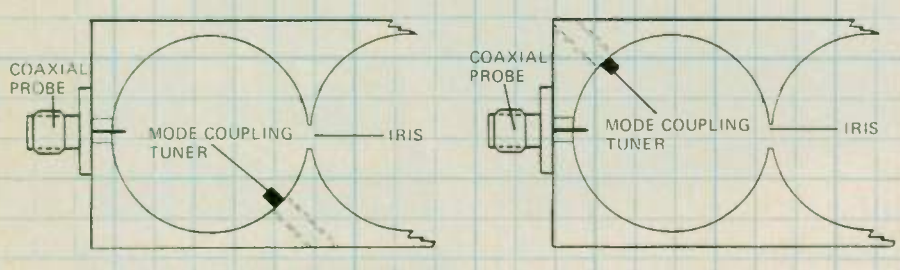


(a) Estimated E-field pattern supported in a cavity excited by a coaxial probe



(b) The E-field pattern supported in a cavity (TE111 mode) excited through an iris



Coupling Arrangement - (c) and (d) Coaxial probe E-field to TE111 mode coupling

For the last two years, **Israel Galin** has worked at Frequency Engineering Laboratories as a Senior Microwave Engineer. He recently joined Aerojet Electro Systems. At FEL, he designed filters, multiplexers, gain equalizers, as well as mixers, Gunn oscillators and varactor multipliers in the MW/MMW frequencies. Mr. Galin got his M.S.E.E. from University of Massachusetts, Amherst, Massachusetts, in 1978 after completing the development of "A 94 GHz Balanced Mixer Using Beam Lead Schottky Diodes." In 1976 he spent six months in University of Eindhoven, Holland, working with Baritt Diodes. Mr. Galin got his B.S.E.E. in Israel in 1971 where he worked until 1976 for the Ministry of Defense as a R & D Microwave Engineer.

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Larry D. Holland is a Senior Research Engineer and is Chief of the Electronic Support Measures Division at the Engineering Experiment Station. He has performed and directed research in the analysis and design of rocket and satellite systems, communication satellites, radar, and electronic warfare. He received B.E.E. and M.S.E.E. degrees from the Georgia Institute of Technology in 1961 and 1963, respectively. Mr. Holland is a member of Sigma Xi and Eta Kappa Nu, a Senior Member of IEEE, and a registered Professional Engineer in the State of Georgia

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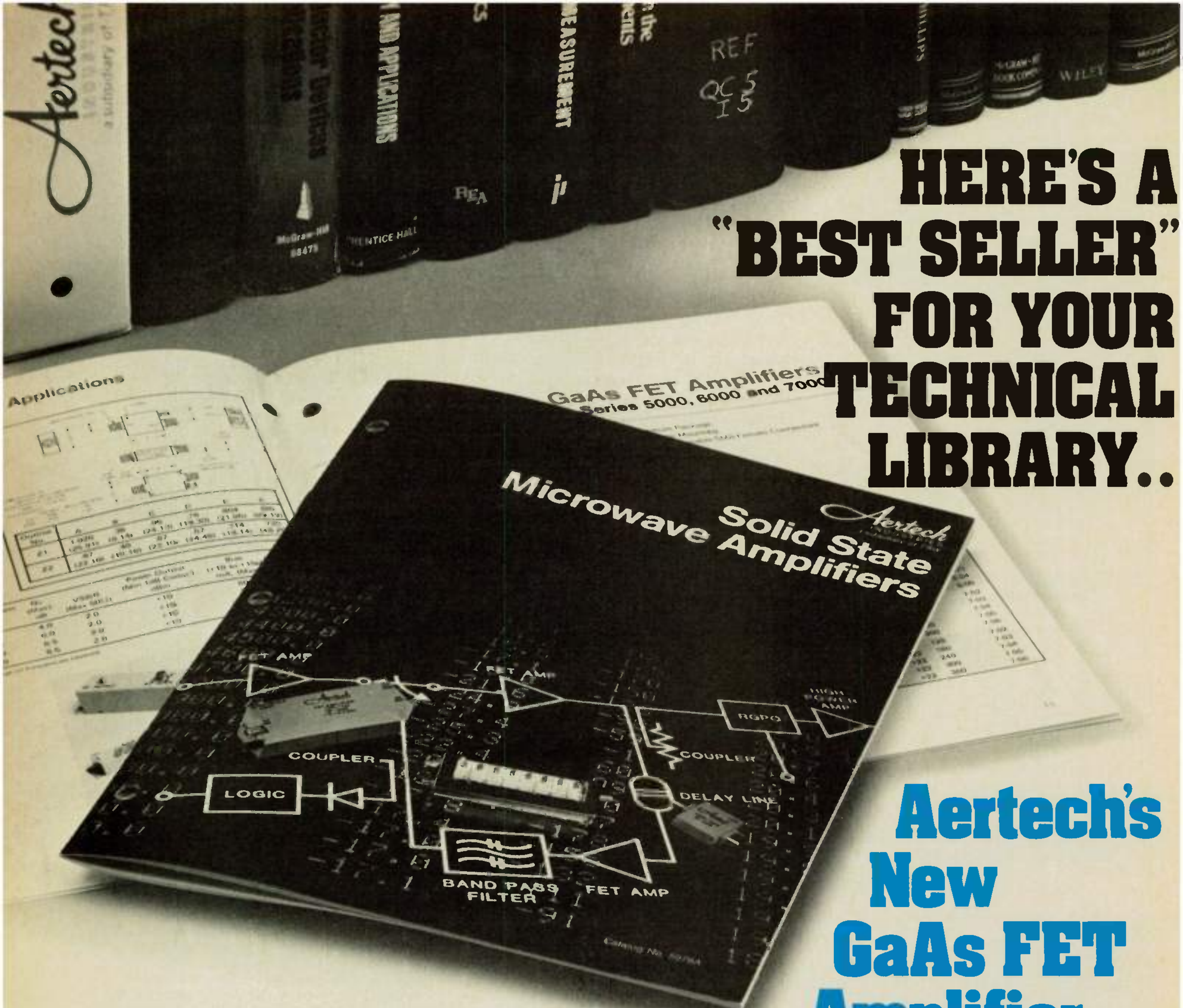
James J. Gallagher is a Principal Research Scientist in the Electromagnetics Laboratory of the Engineering Experiment Station at Georgia Tech. He has been involved in millimeter wave technology for over 28 years. Mr. Gallagher's millimeter wave activities include frequency control, spectroscopy, atmospheric propagation, laser sources, and systems analysis.

Grady Stevens is currently a communications engineer with NASA-Lewis in Cleveland, Ohio. His primary function is definition and overall analysis of future NASA experimental satellite communication systems. In particular, Mr. Stevens is intimately involved in definition of NASA's current 30/20 GHz satellite effort. He is a graduate of North Carolina State University and holds an MS/EE from Case Western Reserve University as well as an MS/ES from Toledo University.



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To meet tomorrow's simulation ieeds today, the Locus, Inc. SG-122 Communications Simulator/Programmable RF Generator provides modeling evaluation, test and training services for C³I Systems. As a stand-alone unit or integrated into a computer-controlled SIGINT simulator, the SG-122 can create a replica of the most desired RF signal environments.

An array of SG-122 units operating under computer control provides effective RF emitter scenario replication of high density signal environments. The sources can be remotely controlled by means of a computer interface which regulates the parameter of each simulated signal including modulation content and type.

The SG-122 is a solid-state unit which provides two simultaneous, independent, high-quality RF output signals in the 1 to 256 MHz range in a single 5¼" x 19" rack-mountable mainframe. Modular plug-ins providing coverage to 512 MHz and 1280 MHz are available. The unit offers microprocessor control of carrier frequency, modulation type and index, and signal level. The carrier frequency resolution of the SG-122 is 10 kHz over the 1 to 256 MHz range.

The SG-122, with a standard MP-1211C AM/PSK Modulator plug-in, provides the following modulation capability as shown in **Table I.**

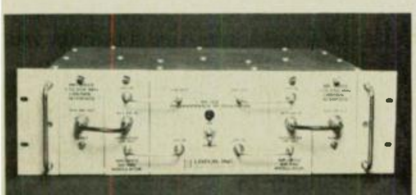


Fig. 1 LOCUS SG-122 – Front View.

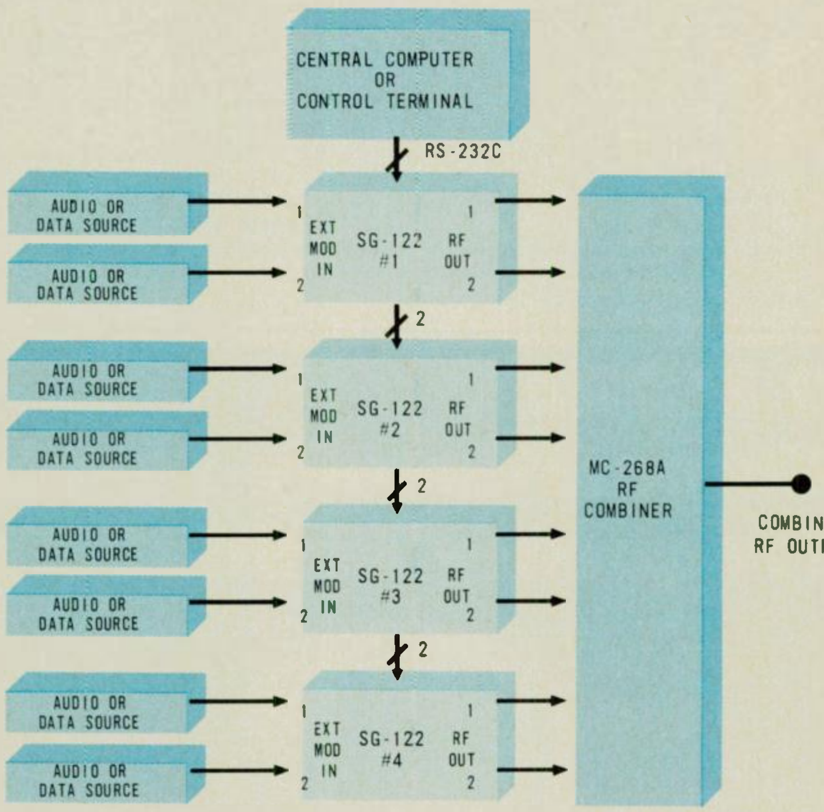


Fig. 2 A typical test configuration using four SG-122's.

TABLE I

• AM	- Modulation Bandwidth	20 Hz to 100 kHz
	Percent Modulation	0 to 100% (1/ steps)
• FM	- Bandpass	dc to 50 kHz
	Deviation	0 to 100 kHz peak (1 kHz steps)
• FSK	- Keying Rate	50 kHz, maximum
	- Deviation	0 to ± 100 kHz
• PSK	- Keying Rate	50 kHz, maximum
• OOK	- On/Off Ratio	50 dB, nominal
	Rise/Fall Time	2 μs, typical

TABLE II

CONDENSED SG-122 SPECIFICATIONS WITH MP-1202A CONTROL/INTERFACE MODULE PLUG-IN AND MP-1211C AM/PSK MODULATOR PLUG-IN

Frequency Characteristics (2 Independent Signals Per Mainframe)	
Frequency Range*	1-256 MHz (10 kHz steps)
Frequency Switching Time	< 20 ms
Output Level Range	-20 to -120 dBm (2 dB steps)
Spurious Outputs (harmonic and non-harmonic)	-50 dBc
Modulation Capability (AM, FM, FSK, PSK, OOK standard)	
AM Modulation Depth	0 to 100% (1% steps)
AM Bandpass	20 Hz to 100 kHz
FM Modulation Deviation	0 to 100 kHz peak (1 kHz steps)
FM Bandpass	dc to 50 kHz
FSK Deviation	0 to ± 100 kHz max.
FSK & PSK Maximum Rate	50 kHz max.
OOK (On-Off Keying)	50 dB ON/OFF ratio, nom.
Rise Time/Fall Time	2 μs typical
SSB (MP-1212 Modulator	USB or LSB
Modulation Bandwidth	4 kHz voice
Internal Modulation Source	+2 V pk squarewave
Programmable Range	7 Hz to 230 kHz
Interface Characteristics	
Impedance at RF Outputs	50 ohms, nominal
EXT MOD IN	5K, nominal to 100 kHz
Control Interface**	RS-232C with "daisy chain" output
Data Rate	110 or 9600 baud
Data Format	1 start, 2 stop, no parity bits
Control Characters	Standard ASCII, 8 bits
Total Parameter Change Time	< 75 ms at 9600 baud
Mechanical/Electrical	
Size	5-¼" H x 19" rack mountable
Power	115 Vac ± 10%, 80 watts

* Extended range mainframes and control/interface modules to 12 MHz and 1280 MHz are available as options

** GP-IB interface available as an option.

Other modulator plug-ins are available to provide single sideband (SSB) and quadrature phase shift-keyed (QPSK) modulations.

The modulated RF output level is programmable in 2 dB steps for -20 to -120 dBm. Both harmonic and non-harmonic spurious signals are < -50 dBc, providing greater than 50 dB of spurious free dynamic range over the 1 to 256 MHz band (see **Table II**).

Each SG-122 unit communicates with the host computer or control terminal via an RS-232C link utilizing standard ASCII characters. An isolated output interface is provided for "daisy chain" control to other SG-122s. Each modulated RF signal is controlled by an internal microprocessor which responds when its unique address is transmitted in the preamble of the control character string. Once addressed, the unit echoes to the terminal and enables control of that unit. An optional GP-IB interface can be used instead of, or in tandem with, the RS-232C interface. The GP-IB interface permits the SG-122 to be used as a control or monitor element for other GP-IB equipment such as tape systems, audio sources, frequency counters, or receivers.

The versatility of the LOCUS SG-122 permits evaluation of communications receiving, recognition and distribution systems, as well as communication jamming equipment under repeatable conditions that closely resemble operational signal environments. A front view of the SG-122 is shown in **Figure 1**, and a typical interconnection of four SG-122s capable of providing eight simultaneous or multiple non-simultaneous signals is shown in **Figure 2**.

A full line of RF hardware accessories to complement the SG-122, as well as software for scenario generation, is available.

Circle 101 on Reader Service Card

ERRATUM

In the July, 1980 Microwave Tube Buyers' Guide, (*Microwave Journal*, p. 67) a listing for the ITT Electron Tube Division was omitted. The division address is: P.O. Box 100, Easton, PA. Tel: (215) 252-7331; Contact: Travis Gerould. It manufactures TWT's, klystrons and special devices.

Microwave Products

Antennas

LOW COST, 3 M ANTENNA SYSTEM

A 3-meter antenna system is designed for economical installation and fast satellite acquisition time. Constructed of fiberglass in three sections, the standard antenna is delivered with spars to mount the single polarization corrugated feed horn radiator, 4 GHz transformer and mast mount. Basic reflector contains a splice plate design to assure surface tolerance for high performance in C-band and allows 12/14 GHz update. Prime focus feed design provides real gain figures in excess of 40 dB, and option to convert to a dual polarization receive mode or a transmit/receive mode via modular addition of an orthomode transducer. Price: $2500, for standard 3-meter antenna. System weight: 450 lbs., with heaviest section 150 lbs. Shipping container has a volume of 150 cu. ft. **COMTECH Antenna Corporation, St. Cloud, FL. (305) 892-6111.** Circle 115.

Materials

HONEYCOMB CORE ABSORBERS WITH GRADIENT LOADING

A honeycomb core microwave absorber line, AAP Type HC, are of hexagonal, open-cell fiberglass construction. Material is treated with a lossy layer to achieve broadband electrical characteristics. Gradient loading maximizes adsorption and maintains desired insertion loss. Material can handle in excess of 10X the RF power of typical foam absorbers. Resistant to shock, vibration and humidity, and operating up to 500°F, these absorbers are available in various configurations — flat with gradient or uniform loading, pyramidal, wedge, and custom machined shapes. **Advanced Absorber Products, Inc., Amesbury, MA. Maurice A. Pennisi. (617) 388-1963.** Circle 110

Au-PLATED Cu CONDUCTIVE ELASTOMER GASKETS

Xecon® SPC, a homogenous conductive elastomer, consists of high grade silicone in which are suspended microscopic silver-plated copper particles. In gasket form the material is suited for wideband EMI/RFI shielding (better than 110 dB at 18 MHz, 72 dB at 10 GHz) and environmental sealing. Control of dispersion of conductive particles gives predictable physical and electrical properties. The material's 60 durometer strength makes it particularly suitable for high closure force applications. **Metex Corporation, Electronic Products Div., Edison, NJ. (201) 287-0800.** Circle 113.

STABLE SUBSTRATE

Cu-Clad 217 is a dimensionally stable microwave substrate which combines woven glass mechanical stability and economy with non-woven products low-loss characteristics. Material has high copper peel strength for fine-line striplining. Size: Standard sheet — 17" x 36", also up to 36" x 36". Thicknesses come in 10-mil increments from 10-120 mils. Conforms to military specification MIL-P-13949F. **3M, Electronic Products Div., Dept. EP80-18, St. Paul, MN. (612) 733-9214.** Circle 112.

LOSSY FLEXIBLE FOAM SHEET

ECCOSORB® LS is a series of low-density, high-loss flexible foam sheet materials designed for cavity loading, wrapping of radiating elements or use as waveguide loads. It is available in a wide variety of forms with range of dielectric properties. Weight: 5 lbs./cu. ft. (0.08 g/cc). Size: Standard sheet — 24" x 24" with 1/8", 1/4", 3/8" or 3/4" thickness. Sheets can be cemented to each other absorber to form a graded absorber. Spray coating is available to seal material for outdoor use. **Emerson & Cuming, Canton, MA. Jeanne B. O'Brien, (617) 828-3300.** Circle 111.

Devices

STANDARD NPN Si RF TRANSISTOR

A standard NPN silicon transistor, NE021, comes in 36 and 37 Micro-Disk packages. The device features a 1.5 dB noise figure at 500 MHz, and can operate at currents up to 50 mA for low distortion applications. Price: $12, 100 qty. **California Eastern Laboratories, Inc., Santa Clara, CA. (408) 988-3500.** Circle 118.

PIN DIODES FOR 500-900 MHz ANTENNA SWITCHES

A series of PIN diodes, UM9601-UM9608, are designed for use with 2-way radio antenna switches and for microwave switching elements. Operates over a range of 100 MHz to 4 GHz. When used in antenna switching circuits of 800-900 MHz 2-way radio, model UM9601 provides > 30 dB of receiver isolation during transmit at < 0.4 dB receive insertion loss. Intermodulation distortion of received signals is < −100 dB. **Unitrode Corporation, Lexington, MA. (617) 861-6540.** Circle 142.

300 W CW SPACE TWT FOR 2.0-2.3 GHz

Model 8281H TWT offers 300 W CW in the 2.0-2.3 GHz frequency band. Conversion efficiency is greater than 50% and integral heat pipe cooling permits low temperature operation. Average baseplate flux density is 2.3 W per sq. in. Cathode loading is 184 mA per sq. cm. Weight: 18.5 lbs. (8.4 kg). Size: 22" (55.9 cm) L. **Hughes Aircraft Co., Electron Dynamics Div., Torrance, CA. (213) 534-2121.** Circle 143.

MAGNETRON

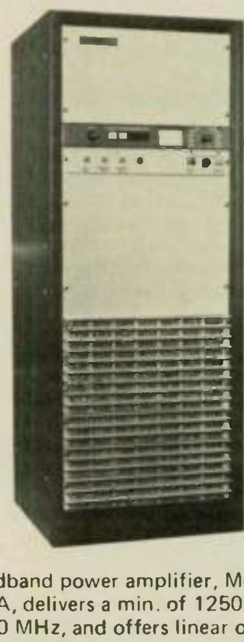
Miniature K_u-band magnetron tube, Model VMU-1255, delivers a min. peak power output of 350 W (at a duty cycle of 0.002) and has an operating life expectancy of over 500 hrs. This device is mechanically tunable over the 16.2-16.3 GHz frequency range. Designed for radar beacon transponders, it has a max. frequency deviation of 5 MHz over the −54° to 71°C temperature range. Size: 1.32" x 1.26" x 2.4". Weight: 8.5 oz. **Varian Associates, Electron Device Group, Palo Alto, CA. (415) 493-4000.** Circle 144.

Systems

EARTH STATION TRAINING SYSTEMS

A low noise amplifier training system was developed for use in an earth station training facility. Consisting of an ultra-low noise amplifier, hot/cold load, low loss transmit/reject filters, manual waveguide switch, and precision test receiver, the system trains technicians and engineers in a classroom situation. **LNR Communications, Inc., Hauppauge, NY. Jeannie Piotrowski, (516) 273-7111.** Circle 114

1250 W, 220-400 MHz RANGE AMPLIFIER



A broadband power amplifier, Model 1000HA, delivers a min. of 1250 W from 220-400 MHz, and offers linear operation over a 90 dB dynamic range — from noise level to 1 dB gain compression at 700 W. Unit employs vacuum tubes in the output stage. 60+ dB gain figure guarantees full output with an input level of 1 mW or less; a continuous gain adjustment of 14 dB is available on the front panel. Model operates into any magnitude and phase of load impedance without damage. An optional IEEE bus interface is available. Size: 22" x 60" x 23". Price: $37,000 FOB. Del: 120 days ARO. **Amplifier Research, Souderton, PA. (215) 723-8181.** Circle 145

Instrumentation

SPECTRUM ANALYZER GPIB INTERFACE

The 757-57 General Purpose Interface Bus (GPIB) is designed for use with a spectrum analyzer and permits remote calculator control for the processing of commands related to the sweep, trace data and alphanumeric annotations which are displayed on the CRT. GPIB Interface uses a microprocessor in conjunction with the interface adapter; it will interpret and execute commands issues by a standard controller. **Electronic Instrumentation Div., Eaton Corporation, Ronkonkoma, NY. (516) 588-3600.** Circle 116.

JOSEPHSON JUNCTION ANALOG

The JA-100 is a Josephson junction analog which simulates the action of cryogenic, superconducting Josephson junctions, at room temperature. It translates the 10-15 pS switching time of a junction into 100 μs and the 2-3 mV gap voltage into 1 V so that they may be observed with low cost test equipment. Unit can be used a circuit component, in the development of Josephson circuit prototypes, and as a means of reviewing Josephson characteristics. It has a self-contained power supply. Input power is 107-135 Vac 60 Hz 10 W. Unit is electronically isolated and protected against accidental overload. Price: $245, 1-4 qty; $215, 5-24 qty. Del: 30 days, FOB. **Philipp Gillette and Associates, Beaverton, OR. Harold Philipp, (503) 645-6339.** Circle 117.

Components

S-BAND FREQUENCY SYNTHESIZER

Model 6765 is a frequency synthesizer which covers the 2080 to 2108 MHz band in 500 kHz steps with +10 dBm output. The 10 MHz reference oscillator provides ±10 ppm frequency stability over the 10 to 55°C range. Unit can be monitored externally and adjusted to compensate for aging. Spurious outputs are −50 dBc, maximum, FM noise is −40 dBc in 3 kHz band at 20 kHz offset. Commands are BCD, TTL; settling time to within 50 ppm of final frequency is 100 ms. dc inputs are 28 and 5 V. Size: 6″ x 4.5″ x 2.7″. **Zeta Laboratories, Inc., Santa Clara, CA. (408) 727-6001.** Circle 141.

HIGH-LEVEL NOISE MODULE

A high-level noise module NOD45, covers the 500-1000 MHz frequency band. Component operates from a +15 V supply and uses less than 300 mA. Minimum output is −82 dBm/$\sqrt{Hz}$ across 50 ohms, and a peak factor greater than +12 dB is provided. Temperature coefficient is < .01 dB/C° over a range of +10° to 50°C. Long-term drift after warm-up is < .05 dB per 24 hrs. Comes with SMA output for noise, solder lugs for bias. Size: 9.9″ x 4.4″ x 1.5″. Del: 8 wks., ARO. **Micronetics, Inc., Norwood, NJ. (201) 767-1320.** Circle 135.

HYBRID TEST FIXTURE

Model T-1160 is a hybrid text fixture which was designed for making RF measurements on 4 pin TO-8 RF modular amplifiers. Maximum frequency recommended for making accurate "S" parameter measurements is 1.3 GHz. Fixture comes with 2 modules to give short, open, through, and 50-ohm impedance for calibration purposes. Pin contacts and header contact plate are gold plated for repeatable and reliable interfaces. **Q-bit Corporation, Palm Bay, FL. (305) 727-1838.** Circle 136.

GUIDED ENTRY SLEEVE FOR SUBMINI RF CONNECTORS

A guided entry sleeve for both slide-on and snap-on type plugs for SMB and SMC RF connectors is offered to aid multiple connector mating. A long chamfered entry on the connector sleeve compensates for accumulated tolerance build-ups in rack and panel, or modular interfaces on center-to-center dimensions. Guided entry aligns the mating components and prevents damage of the connector due to misalignment. **RF Components Div., Sealectro Corp., Mamaroneck, NY. (914) 698-5600.** Circle 138.

CUSTOM WIDEBAND HIGH PASS FILTER

Models in Series F-90 and F-100 are custom wideband, highpass, filters for the 20 MHz to 18 GHz frequency range. Units provide low insertion loss, good SWR and rejection through the use of both distributed and lumped component techniques. Miniaturized construction is featured. Price: from $150, unit qty. Del: 4 wks. **RLC Electronics, Inc., Mt. Kisco, NY. (914) 241-1334.** Circle 137.

SMALL PROGRAMMABLE ATTENUATORS

Series 3200 are programmable step attenuators which operate over the dc to 2 GHz frequency range. They are available in 8, 5, and 1 cell configurations and 5 standard attenuation ranges (127 dB/10 dB steps, 63.75 dB/0.25 dB steps, 0.5 dB/0.5 dB step. Incremental accuracy is: ±0.2 dB or 0.5% to 500 MHz; ±0.2 dB or 1% to 1 GHz; and ±0.3 dB or 2% to 2 GHz. Switch life is 10^7 operations; switching speed is 6 ms, maximum at nominal 12 V/14 mA per cell. Monotonic performance is guaranteed to 1.5 GHz. Price: starts at $135 (US). Avail: Stock to 90 days ARO. **Weinschel Engineering, Gaithersburg, MD. (301) 948-3434.** Circle 140.

(continued on page 100)

(from page 32) AMPLIFIERS

to TWT amplifiers. Gunn and IMPATT diodes seem to be overtaken, but silicon bipolar transistors continue to compete for applications in saturated amplifiers below 5 GHz and in linear power amplifiers below 3 GHz.

The modern high-efficiency TWT's offer unmatched overall performance, reinforcing the systems users confidence that has been building up during the past three decades. This tends to postpone the transition to solid-state TWT substitutes wherever the advantages thereof are not yet quite convincing.

Leaders in the development and systems applications of GaAs FET power amplifiers are manufacturers with superior in-house device capability. They manage to absorb the start-up costs and tend to avoid the use of TWT's wherever technically justifiable.

The principal performance disadvantage of present-day GaAs FET power amplifiers is their low efficiency in comparison with the modern TWT's. Overcoming this disadvantage may take another few years of systematic development efforts. However, there is a way of circumventing it in a significant number of systems applications, where the solid-state amplifiers are not required to match the standard 10 W or higher TWT power output. Such is the case wherever a sizable tradeoff (3-6 dB) between transmitter output power and receiver noise figure is technically and economically feasible, and does not result in excessive system interference.

Equipment manufacturers depending on commercially available active microwave devices are presently at a distinct disadvantage in terms of power GaAs FET availability, performance and cost. GaAs FET's will have reached maturity when this distinction becomes negligible, as it has for TWTs and silicon bipolar power transistors.

In summary, while solid-state microwave power amplifiers made most encouraging progress during the past five years, the TWTs kept advancing, as well and we may well plan another panel session on the same subject in another five years.

Now measure frequency, power, AM, FM, ΦM from 150 kHz to 1300 MHz.

10 Hz resolution to 1000 MHz

Single keystroke autotunes frequency; autoranges modulation and power

Keys enter numeric data and special codes

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Low distortion recovered modulation

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Average and + or — peak detectors with peak hold for transients

Ratio keys for relative dB or %

"Special" key for manual "upshift" functions and service routines

HP's New 8901A Modulation Analyzer.

For the first time all the capability you need to measure RF frequency, power and modulation is in one instrument.

The new HP 8901A Modulation Analyzer—essentially a precision calibrated heterodyne receiver—quickly characterizes transmitters and accurately calibrates signal generators.

The 8901A design minimizes measurement error, e.g. residual FM on the internal local oscillator is <1Hz at 100 MHz (50Hz-3kHz BW).

AM & FM accuracy is ± 1% (ΦM ± 3%). Well isolated detectors let you separate small values of incidental AM or FM from large values of primary modulation.

Easy to use too, the HP 8901A automatically selects the strongest signal, makes the measurement and displays the result. All in all, it's not only a new concept in RF measurements, but an extremely powerful tool for the analysis of modulated signals.

Take a look at the features above, then if you want more information call your nearby HP sales office or write Hewlett Packard Co., 1507 Page Mill Road, Palo Alto, CA 94304.

Price $7800*, Calibrator, $500*

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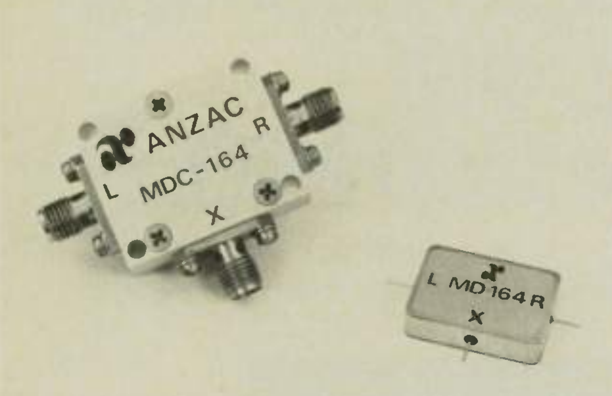
04902B

HEWLETT PACKARD

TERMINATION INSENSITIVE MIXER COVERS .5-.9 GHz BAND



MD-164 is a termination insensitive mixer (TIM) which covers the .5-.9 GHz range. Unit offers 8.5 dB typical conversion loss with IF port response to 2 GHz; +8 dBm compression point and +17 dBm 3rd order intercept over the entire band. Price: $395, flatpack version; $470, connectorized version. Del: from stock. **Anzac Div., Adams-Russell Company, Inc., Burlington, MA. (617) 273-3333.** Circle 120.

SMA QUICK-DISCONNECT RACK AND PANEL CONNECTOR

The SMA quick-disconnect rack and panel plug connector, 705535-003, has a spring-loaded float mount and is intended for blind mating rack and panel applications. SWR is 1.25 max. to 28 GHz. The plug mates with a standard SMA jack and meets all applicable portions of MIL-C-39012. Price: $6.32, 1000 pieces. **Cablewave Systems, Inc., North Haven, CT. Steven Raucci, Jr., (203) 239-3311.** Circle 121.

INTEGRATED UHF DUAL ISO-FILTER

Model 103600027 is a UHF dual iso-filter with sharp cutoff comb-line filter and two high performance ferrite isolators integrated on a common ground plane structure. Unit covers the 494-554 MHz frequency range with an insertion loss of 1.3 dB maximum. Rejection at 614 MHz and 438 MHz is greater than 40 dB; input and output SWR remain below 1.3 for all load conditions. Both single and dual iso-filters covering bandwidths as wide as 70% from 500 MHz to 18 GHz are available. Phase linearity and loss variation specifications are optional. Model 103600027 price: $1,395. Del: 8-12 wks., ARO. **Addington Microwave Components, Eaton Corp., Sunnyvale, CA. Jim Wilson, (408) 738-4940.** Circle 122.

LOW LOSS COAXIAL CABLE ASSEMBLIES FOR 26.5 GHz BAND

The Gore-Tex® .190" O.D. line of coaxial cable assemblies offer low insertion loss at all frequencies up to 26.5 GHz. Impedance is 50 ± 1 ohm; time delay is 1.2 nsec/ft (85% speed of light); capacitance is 26 pf/ft; RF leakage is more than 100 dB down at 1 GHz, corona extinction voltage is 2500 V, rms and dielectric withstanding voltage is 1000 V, rms. Operating temperature range is -55°C to 200°C (cable), -55°C to 150°C (assembly). Minimum bend radius is 1 in., standard assembly. Weight: 30 gm/ft. of cable,14 gm per connector pair. **W. L. Gore & Associates, Inc., Electronic Assembly Div., Newark, DE. (301) 368-3700.** Circle 124.

BROADBAND OVERLOAD PROTECTOR FOR RF INSTRUMENTS

A limiter which protects RF instruments from overload damage, Model 11867A, provides limiting action at signal levels around 1 mW. With applied levels of 10 W CW (or 100 W peak), the output from the limiter stays below 100 mW. effect on lower level measurements is minimal, typically introducing frequency response variations of less than 0.25 dB across the dc to 1800 MHz range. Price: $225 (USA). Del: 2 wks. **Hewlett-Packard Co., Palo Alto, CA. (415) 857-1501.** Circle 125.

TVRO MIXER SPANS .5-3 GHz RANGE

A TVRO mixer, MLK-124, covers the 1-4.2 GHz range and has an IF output range of 0.5-3 GHz. LO power required is +7 to +13 dBm, minimum. Isolation for all ports is specified at 20 dB, with typical figures of approximately 30 dB. Unit has a conversion loss specified at 8.5 dB maximum for the full frequency range, with typical figures in 6.5 dB area. Price: $75, with SMA connectors, a flatpack version is also available. **Engelmann Microwave Company, Montville, NJ. Carl Schraufnagl, (201) 334-5700.** Circle 126.

LOW CONVERSION LOSS MIXERS FOR 1-10 GHz

A line of double balanced mixers (FC-325), feature conversion loss of 4.5 dB max. for the lower frequency models, and no higher than 5.5 dB max. for the higher frequency units. Line covers the 1-10 GHz frequency band. Series has flat conversion loss over the entire frequency range, 1 dB compression points range from 0 to +3 dBm and 3rd order intercept points from +9 to +14 dBm. Models are available both as down-converters and up-converters and are equipped with SMA connectors. **Lorch Electronics Corp., Englewood, NJ. (201) 569-8282.** Circle 130.

DOUBLE BALANCED TO-8 MIXERS

Miniaturized high and low level double-balanced TO-8 mixers cover the 1-1000 MHz frequency range. Series M40T mixers also come in DIP package (relay header) from 10-1000 MHz. Units operate at +7 dBm LO (low-level model M43t) or +13 dBm LO (high-level model M46t); typical conversion loss is 6.5 dB; typical isolation is 45 dB from 1-100 MHz; 35 dB from 100-500 MHz; 30 dB from 500-1000 MHz. **Magnum Microwave Corporation, Sunnyvale, CA. David Fealkoff, (408) 738-0600.** Circle 131.

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New Literature

PUBLICATION ON MIXER MODULATION MEASUREMENTS

Publication SP 400-16 describes modulation measurements for microwave mixers. This 86-page NBS document discusses the measurement of mixer conversion loss using periodic or incremental modulation of the local oscillator, plus evaluation and minimizing of associated systematic and random uncertainties encountered with an X-band mixer measurement system. It covers improvements in periodic and incremental modulation techniques and novel circuits for measuring intermediate-frequency output conductance and local-oscillator return loss. Price: $3.75. **Supt. of Documents, Govt. Printing Office, Washington, DC.** Circle 106.

SMB/SMC COAXIAL CONNECTOR CATALOG

Catalog CX-9 details a line of SMB/SMC coaxial connectors. It includes information on an expanded line of ConheX connectors, including the Posi-Loc SMB, both 50 and 75 Ω types, and direct solder series. A complete cross-reference to MIL-C-39012, Series SMB/SMC is provided, along with detailed specifications and drawings. **RF Components Div., Sealectro Corporation, Mamaroneck, NY.** (914) 698-5600. Circle 107.

DATA SHEET ON VHF POWER TRANSISTOR

A two-color data sheet, No. SD1480, describes a 125 W, 28 V VHF power transistor. Includes application description, features, absolute maximum ratings, electrical characteristics, typical curves and a circuit diagram for the test circuit. **Thomson-CSF, Solid State Microwave Division, Montgomeryville, PA.** (215) 362-8500. Circle 108.

COAXIAL MAGNETRON BOOKLET

A 38-page booklet describes coaxial magnetrons developed for use in instrumentation, surface and airborne radars and ECCM applications. A history of coaxial magnetrons is presented, as well as theory of operation, typical parameters and specifications plus photos for full line of these devices, excluding classified types. Modification kits for upgrading older radars are also offered. **Varian, Electron Device Group, Palo Alto, CA.** (415) 493-4000. Circle 109.

APPLICATION NOTE ON AUTOMATED SYSTEMS FOR NOISE MEASUREMENTS

An application note, 64-3, gives details on how to assemble an automated system to make accurate noise figure measurements. This 32-page booklet describes a system using a microwave source and mixer as a tunable down-converter plus power meter, noise source and off-the-shelf components. General noise figure measurement principals are introduced in addition to instructions for assembly and operation. Demonstration software routines are provided, as well as typical curves and block diagrams. **Hewlett-Packard Co., Palo Alto, CA.** (415) 857-1501. Circle 103.

RENTAL TEST AND MEASUREMENT EQUIPMENT CATALOG

A catalog lists over 2,600 electronic test and measurement items offered for rent to government agencies. The General Services Administration Master Price List describes all testing and measurement devices available for short-term rental (one year or less) for a variety of applications. **Leasametric, Foster City, CA.** (415) 574-4441. Circle 104.

FERRODISC MICROSTRIP CIRCULATORS AND ISOLATOR CAPABILITY BROCHURES

Bulletins 1147 and 1141B are capability booklets for lines of microwave ferrodisc circulators and isolators. They provide circuit designers with comprehensive specifications, including key mechanical and thermal parameters, technical applications information and procedures for installing the devices into microstrip circuits. **Microwave Associates, Burlington, MA.** (617) 272-3000. Circle 105.

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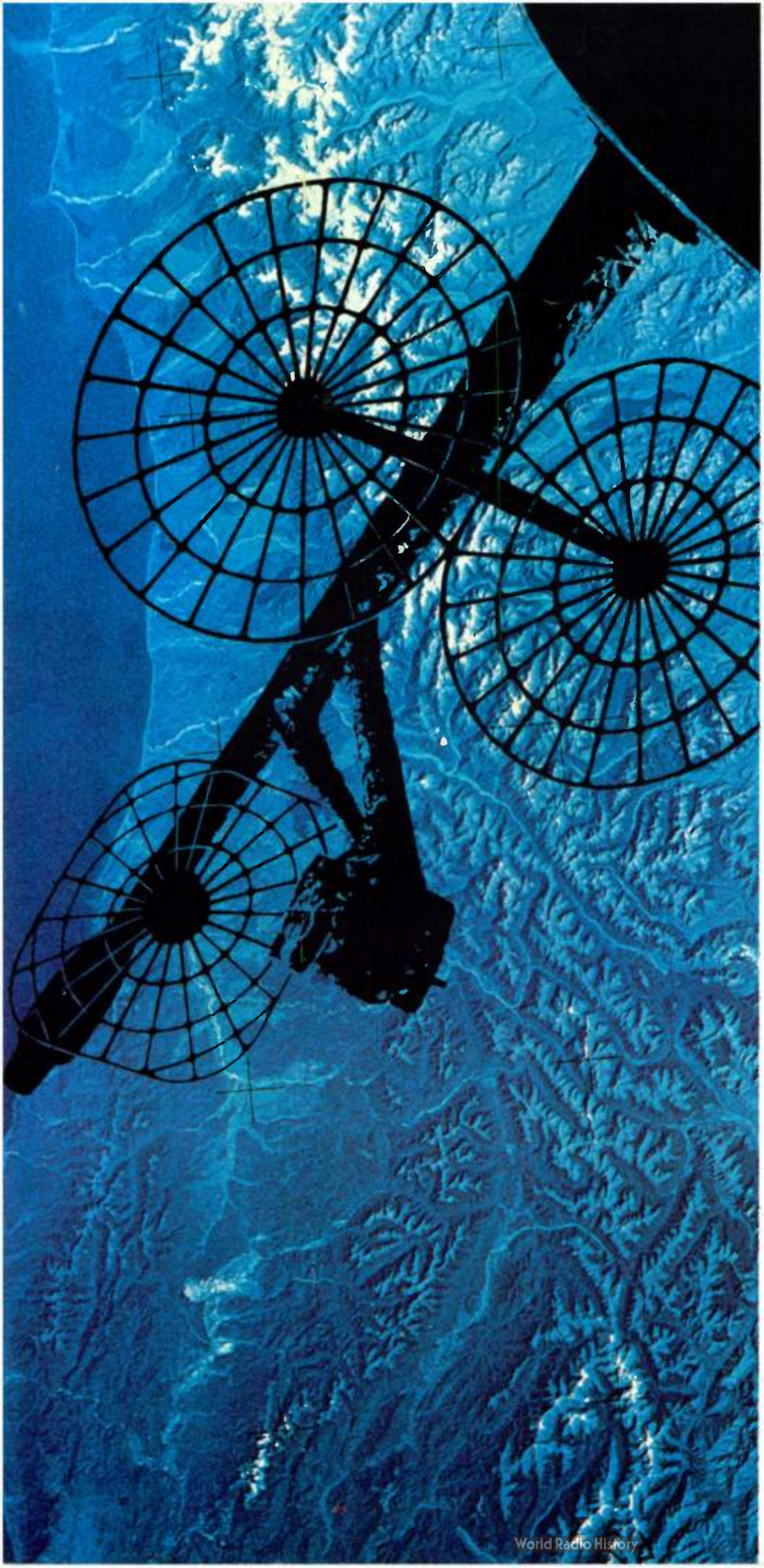

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