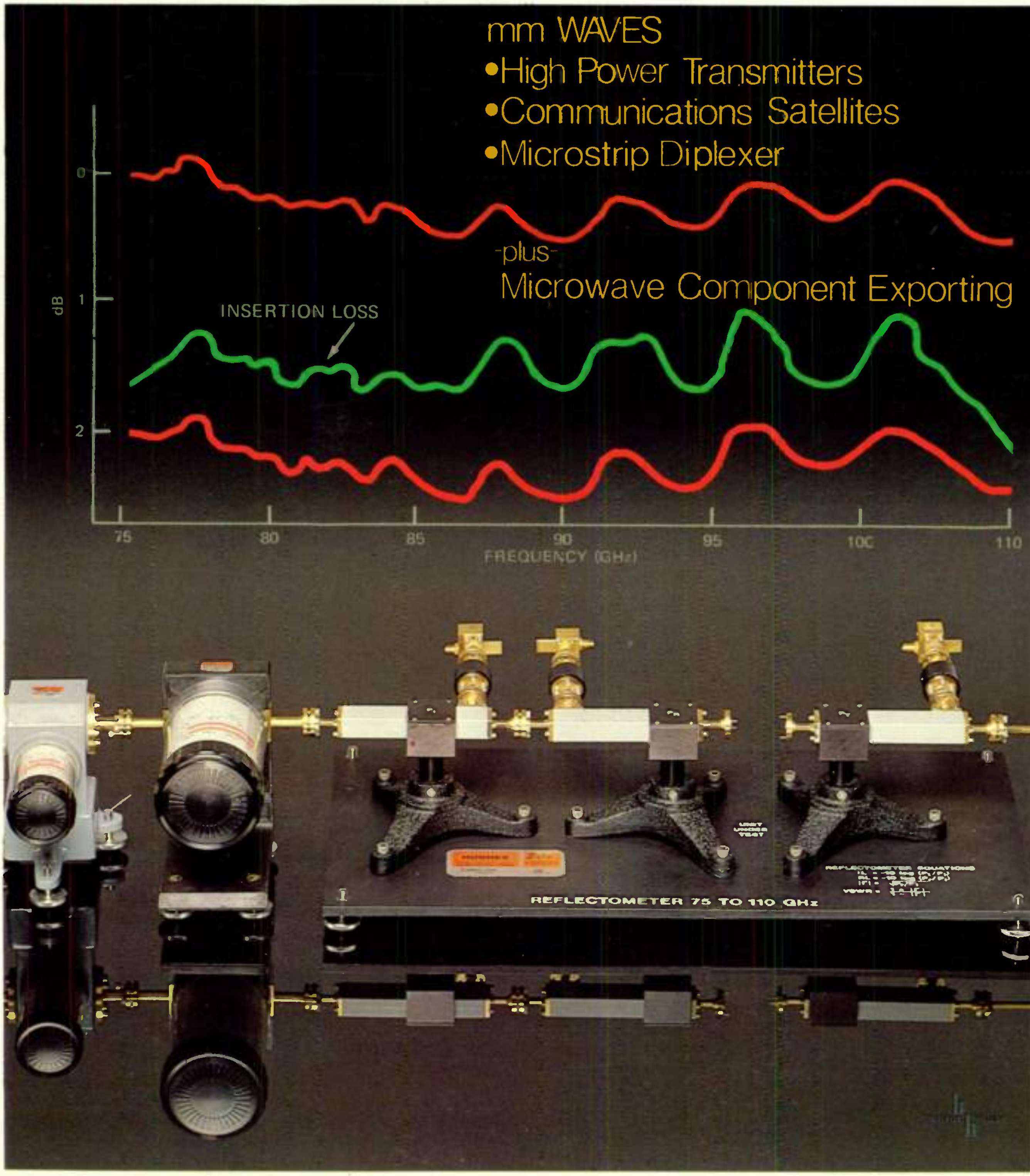
microwave
JOURNAL

INTERNATIONAL EDITION □ VOL. 23, NO. 6 □ JUNE 1980

mm WAVES
- High Power Transmitters
- Communications Satellites
- Microstrip Diplexer

-plus-

Microwave Component Exporting



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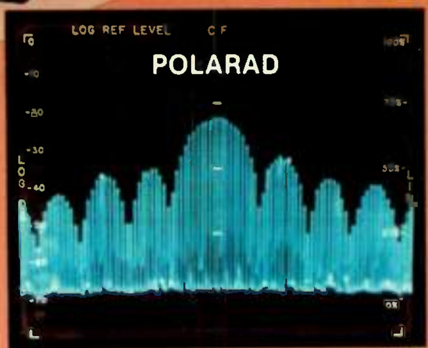

Pulsed output spectrum of Polarad "E" Series Generator.
Note: no carrier leakage.

Carrier leakage is clearly seen.

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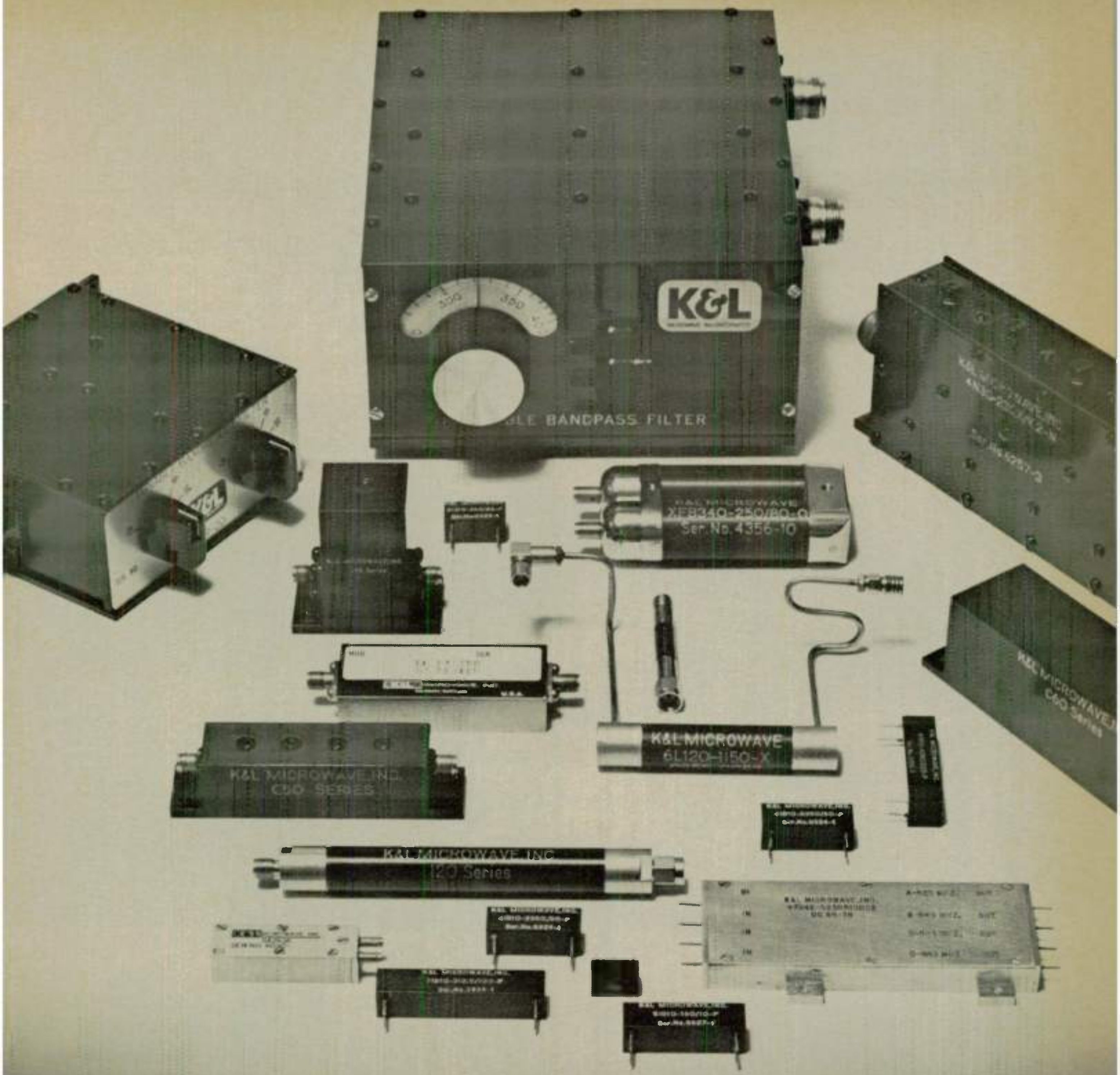
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ON THE COVER: Reliable impedance and transmission measurements from 26-110 GHz are available from a mm-wave reflectometer. A Cover Story starts on p. 59.

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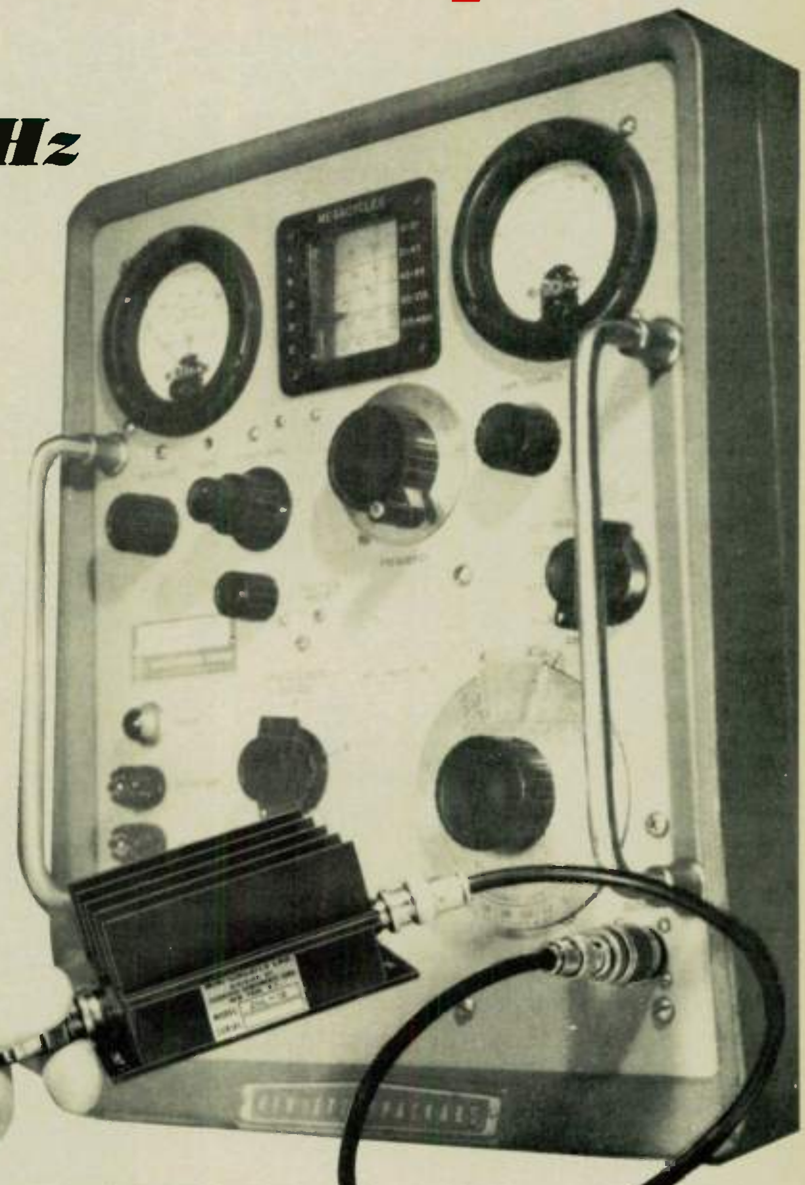

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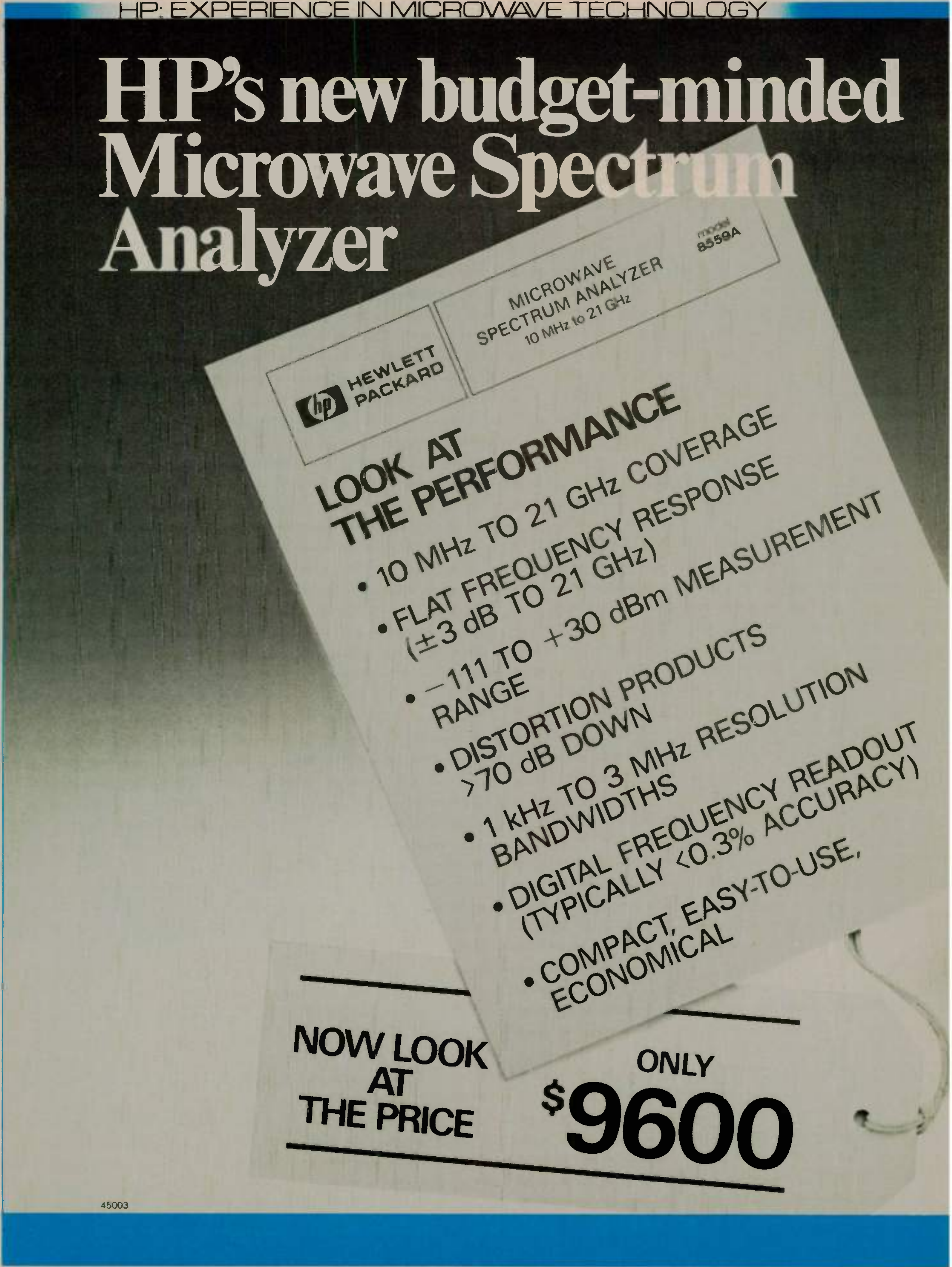
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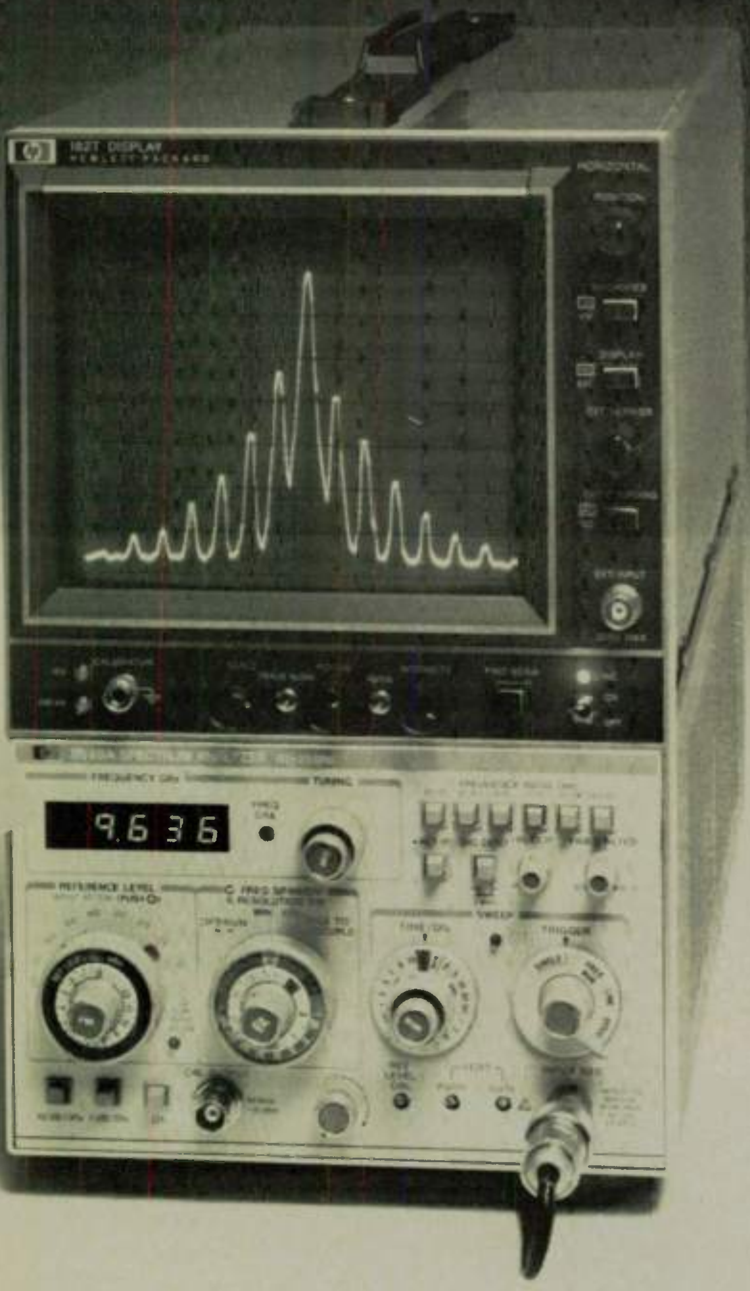

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Coming Events

26TH ANNUAL TRI-SERVICE RADAR SYMPOSIUM JULY 15-17, 1980 Sponsors: US Army Combat Surveillance and Acquisition Lab and Radar and Optics Div., Environmental Research Institute of Michigan (ERIM). Place: US Military Academy, Mahan Hall, West Point, NY. Description: advances in radar techniques, devices and applications. Contact: Henry A. Amble, Jr., ERIM, P.O. Box 8618, Ann Arbor, MI 48107. Tel: (313) 994-1200, ext. 324.

10TH EUROPEAN MICROWAVE CONFERENCE SEPT. 8-12, 1980 Sponsors: Association of Polish Electrical Engineers, EUREL, IMPI, URSI and IEEE Region 8 — in association with Microwave Exhibitions and Publishers, Ltd. Place: Warsaw, Poland. Contact: Prof. Andrzej Sowiński, EuMC Conf. Chrmn., Industrial Institute of Electronics, ul Długa 44/50, 00-241, Warszawa, Poland.

EASCON '80 SEPT. 29 - OCT. 1, 1980 Sponsors: IEEE - Washington Sect. and Aerospace and Electronics Systems Society (AESS). Place: Sheraton National Hotel, Arlington, VI. Theme: "The 1980s — Electronics Systems Decade." Contact: EASCOM '80, 608 H Street, S.W., Washington, D.C. 20024. Tel: (202) 347-7088.

1980 IEEE INT'L SYMPOSIUM ON ELECTROMAGNETIC COMPATIBILITY OCT. 7-9, 1980 Sponsor: IEEE Place: Baltimore Hilton Hotel, Baltimore, MD. Theme: "A Constellation of Ideas." Contact: Thomas J. Bode, Publicity, EMC '80, P.O. Box 1711, Annapolis, MD 21404. Tel: (301) 267-2898.

MILITARY MICROWAVES '80 CONFERENCE AND EXHIBITION OCT. 22-24, 1980 Sponsor: Microwave Exhibitions and Publishers Ltd. Place: Cunard International Hotel, London. Topics: Military applications of microwave engineering. Contact: R. C. Marriott, Managing Dir., MEPL, Kent TN13 1JG. Tel: (0732) 59533/4. Telex: 95604 YNLTD G.

GaAs IC SYMPOSIUM NOV. 4-6, 1980 Sponsor: IEEE Electron Devices Society. Place: Imperial Palace Hotel, Las Vegas, NV. Topics: linear monolithic signal and power and digital integrated circuit, development and applications, device physics, modeling and simulation, etc. Howard Phillips, Chrmn, Lockheed Microelectronics Cent., Space Systems Div., Dept. 62-46, Bldg. 151, Lockheed Missiles & Space Co., P.O. Box 504, Sunnyvale, CA 94086.

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

This issue concentrates on millimeter wave developments in several areas, ranging from component design to satellite systems analysis, and further to electropolitical questions relating to government support for millimeter wave programs. By observing current technical and trade journals, as well as symposium agendas, one can see that millimeter-wave research and development activity is at a high level. This was further confirmed recently by an announcement by Dr. William Perry (Department of Defense research director) that excellent progress is being made in the military application of millimeter waves to detection and guidance systems.

Sum Up

THE MICROSTRIP DIPLEXER

If, several years ago, someone had asked whether microstrip (or other forms of TEM lines) would become a useful and popular type of transmission line at frequencies as high as 100 GHz, probably most engineers would have been doubtful, particularly since several alternate waveguiding schemes were (and still are) under investigation. However, much very good developmental work has been accomplished by various workers, and the article by D. Rubin and D. L. Saul illustrates this point. They have produced a microstrip diplexer which provides excellent electrical properties in the 28 to 40 GHz region, and is also the basis for the design of a triplexer, an integrated downconverter, and a frequency stabilized oscillator. Moreover, Rubin and Saul give specific design and configuration layout information.

NEAR MILLIMETER-WAVE IC's

The paper by S. E. Schwarz and D. B. Rutledge addresses the possibility of developing integrated circuit techniques and components at much higher frequencies, using microfabrication technology. The frequencies discussed include both the submillimeter and millimeter wave region. The authors describe dielectric waveguides, antennas, a V-coupler, Schottky diode mixers, integrated receivers, and detector arrays. The results are preliminary, but the innovative approaches may help point the way for others.

MILLIMETER-WAVE COMMUNICATION SATELLITES

An area with a surprising amount of technical activity in recent years has been the consideration of millimeter-wave satellite communications. NASA, DoD, and the Services have supported numerous analyses, and study reports have been published by the RAND Corporation, MIT Lincoln Laboratory, and other university and industrial organizations. The IEEE EASCON '79 Symposium included several such reports, for example. In this issue L. Holland, N. Hilsen, J. Gallagher and G. Stevens describe an investigation into the considerations of cost, weight, performance, and design trade-offs for frequencies between 18 and 80 GHz, with primary emphasis on 40 and 50 GHz. In Part I of this article, conceptual designs are presented for point-to-point and broadcast communication satellites. Channel costs are compared with current tariffs. Since technical feasibility is already established, such cost information is very important for planning of future satellite systems.

US ARMY MILLIMETER-WAVE PROGRAMS

Additional subjects discussed by our other authors include relevant contract activity by the US Army Electronics Research and Development Command. This is Part II of the article dealing with the US Army Advanced Planning Briefing held in November 1979, the programs in the millimeter-wave region to be funded from 1980-1985 are shown. In general, funding is directed toward the design of low cost mm-wave components for high resolution radar, terminal homing device, wideband receiver and secure communications applications.

MILLIMETER-WAVE REFLECTOMETER

A swept 26.5 to 110 GHz reflectometer system is described. Above 40GHz, considerable inaccuracy is associated with impedance measurements employing slotted lines of hybrid impedance bridges. System components are described and charts relating system accuracy to coupler directivity and test mismatch are provided. A computer controlled version of the system is also supplied.

JAMES C. WILTSE, Associate Editor

Workshops & Courses

MICROWAVE CIRCUIT DESIGN

Sponsors: U. of Maryland, University College, UCLA Extension
Dates & Sites: July 7-11, 1980 – Cent. for Adult Ed., College Park, MD. October 27-31, Rm. 6266, Boelter Hall, UCLA
Fee: $575
Content: Microwave circuit design techniques, plus lab sessions.
Contact: (Coordinator and Lecturer) Technical Info. – Les Besser, Pres., Compact Engineering, Inc., Palo Alto, CA. Tel: (415) 858-1200
Program Info. – Conf. and Institutes Program, U. of Maryland.
Tel: (301) 454-5237
UCLA Extension,
Tel: (213) 825-1295

FIBER OPTIC COMMUNICATION SYSTEMS

Sponsor: Universityof California Extension, Santa Barbara
Date: July 7-11, 1980
Site: Rm. 1104, Engineering Bldg., UC, Santa Barbara
Fee: No. X457 - $550 (enrollment deadline – June 23)
Description: Review of fiber optic technology, with emphasis on practical system design and system components.
Contact: The Director of University Extension, 1833 Ellison Hall, UC, Santa Barbara, CA 93106.
Tel: (805) 961-2944.

RADAR TECHNOLOGY

Sponsor: Boston IEEE AESS (Aerospace and Electronic Systems Society)
Date: July 14, 1980
Site: Hotel Thayer, West Point, NY
Lecturer: Eli Brookner, Consulting Scientist, Raytheon Co.
Fee: (includes text) $105, member $120, non-member – by July 5
$120, member $135, non-member – after July 5
Description: Sessions on fundamentals of radar, trends in signal processing, and radar components.
Contact: Duane Matthiesen, V.P. IEEE IEEE AESS
Technical Info – Mitre Corp., Bedford, MA. Tel: (617) 271-2000, ext. 2309
Program Info. – Boston IEEE AESS, 282 Marrett Rd., Lexington, MA 02173

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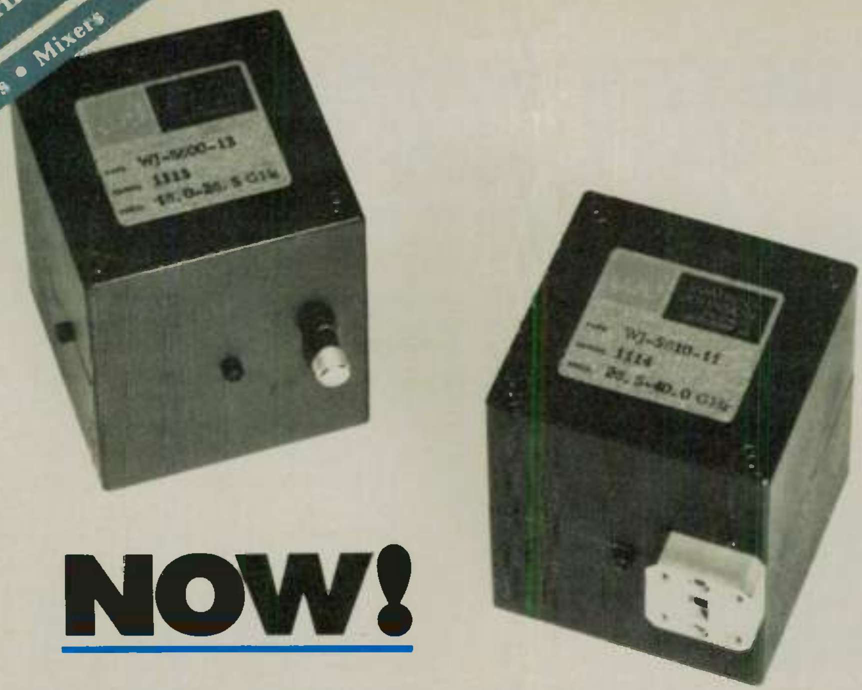

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WJ-5610	26.5–40.0 GHz	15 mW	±50	120 MHz	10-2V @ 650 mA

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MILLIMETER WAVES...
Who's Responding to the Challenge

Andrew S. Kariotis, *president and chief executive officer of Alpha Industries, joined Alpha just after its inception in 1962. He became a vice president of Alpha in 1964, executive vice president in 1972, and president in 1975. He has been a member of the board of directors since 1967.*

Prior to joining Alpha, he held various marketing positions with several high technology firms in the Boston area for eight years. These assignments followed his graduation with honors from the M.I.T. Sloan School of Management in 1954.

As was pointed out in these pages a few months back, the microwave industry has come of age in the past thirty years and established its place both in society and the business community. The acceptance has indeed become so complete that it has fairly well consumed the frequency spectrum once perceived as frontier territory. And although new technologies are making increasingly effective utilization of the spectral region, there has always been the knowledge that this resource was limited and that exploitation of the mm range was inevitable.

It has been difficult for people in the microwave field to be neutral in their perception of the frequency decade from 30 to 300 GHz. There have always been some strong advocates for its exploration and utilization, and a larger group who felt (or feel) that there was little, or no, potential here for serious consideration. The dichotomy — due largely to atmospheric vagaries — persists.

There were basic problems with these frequencies: they had no satisfactory solid state approaches to such essentials as oscillators and transmitters; and the component manufacture bore a strong resemblance to the jewelry business in both craftsmanship and price. In the world of financial reality this was an anathema. The promise of the laser and infrared applications had the glamour, the good promise, the financial support and some fantastic results. The mm region remained a void between mw and IR.

Lack of financial support from the government and visibility for market potential ties the hands of public corporations from investing disproportionate IR&D funds. So the evolution at this range has been very gradual. However, we believe that even in this rather laissez-faire environment, the diode and circuit developments of the past decade have been significant. They have brought the key elements of mm systems — i.e., the transmitter and receiver — to a level where a wide variety of mw systems can now have their mm counterparts, and function quite well.

We are recognizing, now, that for all its benefits, infrared is not a panacea for smart weapons in a battlefield situation. Mm waves can provide an adjunct or an alternative in the precision-guided munitions field.

We have long felt that the mm spectrum needed but one large systems push to open up the frequency range for a variety of potential applications. The millimeter seeker could provide this essential drive in this decade.

Once certain key components have been produced in quantity, and the proper production methods and techniques established, the long-standing barrier of price disparity in millimeter systems will quickly vanish. We should not, for example, sit quietly by and concede the high frequency communications market to foreign development.

The industry now has diodes and components for transmitters, pulsed or spread spectrum, which can provide adequate energy on target for active systems. Also, the industry has developed receivers at 94 GHz, for example, which are comparable in performance to X-band equipment.

For the short-range applications, tactical weapons and communications, we should stop postulating impediments and start fostering more rapid development. To date, many of the advances have been made with a significant amount of private investment as the millimeter frequency range languished through the infrared period.

If the disparity of armor in Europe is a serious tactical problem, and if the millimeter seeker presents a potential solution for the problem; then we would expect a clarion call to provide such systems with some associated urgency and funding. The systems houses venturing into the mm field cannot afford unilaterally to tool up for production, and the component suppliers are generally even less well endowed. If these weapons systems are to be deployed in the '80s, the signal must be very clear from the DoD that they are serious and committed. In the short run, it may be very cost effective to solicit a $1M system with a $0.5M budget and the prospect of production. There will be takers. But after a few such ventures with no production forthcoming, there will be a growing reluctance in the industry. Component houses can neither grow nor flourish if the systems houses are required to ask them to cost share every development. The same rationale can be applied to totally vertically integrated houses with internal cost or profit centers.

Industry has time and again proved itself capable of responding to a challenge when the commitment of the government is evident. There is an exciting potential here.

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Input VSWR:	1.30:1 (1.25:1 typ.) room temperature 1.40:1 (1.35:1 typ.) 0 to +50° C
Weight:	250 grams
Waveguide Size:	WR 10 (UG 385/U flanges)
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Switching Rate:	1 KHz
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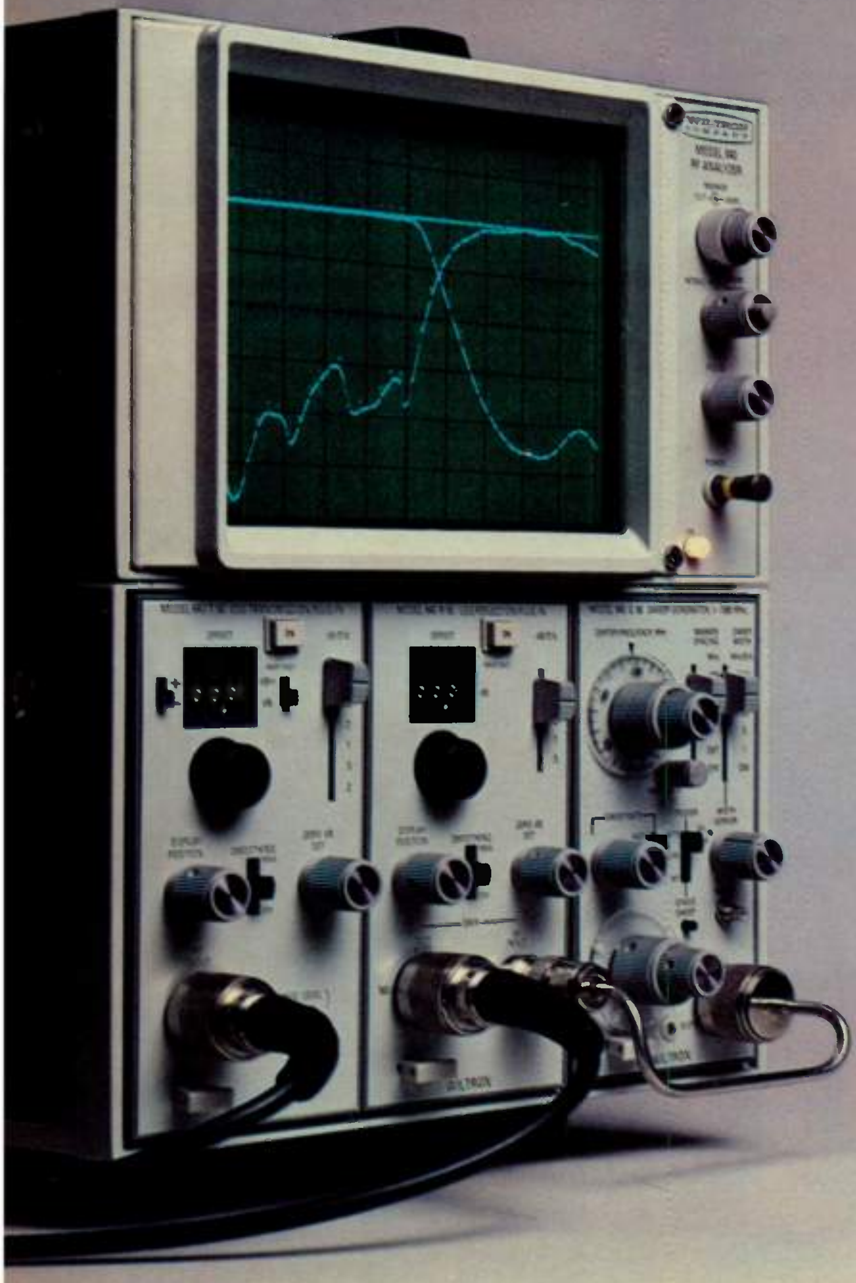

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Critique of Microwave Export Performance

TED SAAD
Sage Laboratories, Inc.
Natick, MA

INTRODUCTION

The history of the microwave industry has been one of feast or famine. Not too many years ago, most of us relied primarily on the Department of Defense for our well-being. In more recent years, however, with the growth in non-DoD microwave communications and radar and commercial and industrial applications, our reliance on the DoD has diminished. We have also seen a growth in non-U.S. markets. The objective of this paper is to discuss and critique the Export Performance of U.S. Microwave Companies.

To limit the study, I chose to concentrate my efforts on those companies exhibiting at the 1980 MTT Symposium in Washington, DC, and in particular those that could be classified as national rather than multinational organizations. From a list of exhibitors, I was able to select 68 companies that appeared representative of our industry. A questionnaire was prepared and mailed to each, along with a stamped addressed envelope and a completed questionnaire prepared by my own company to encourage participation.

In tabulating the results and analyzing the data, there appeared to be no surprises. It seems that because we all talk with one another, most of us conduct our export business in similar fashion. Of 68 questionnaires mailed, 28 completed questionnaires were returned. An additional 7 were returned, but with financial details omitted.

EXPORT SALES

To help understand the data, I have arbitrarily divided the companies into four major groups (A, B, C and D) based on sales volume as shown in Table I. The largest company, reporting sales in excess of $125 million, was omitted from this tabulation. Also, one can argue the validity of group D, with such a small sample. However, the first piece of data indicates that our percentage of export sales increase with sales volume. Not a surprising statistic.

TABLE I

	GROUP			
	A	B	C	D
Number of Companies	9	9	6	3
Sales Range (in millions)	.3-2.5	3.3-5.5	7.0-9.0	12.0-26.0
Total Sales (in millions)	12.74	39.61	49.38	53.20
Export Sales (in millions)	1.15	5.53	10.32	11.39
Export Sales %	9.0	14.7	20.9	21.0

EXPORT SALES BY TERRITORY

The next result to examine is to whom do we export. Here again, there are no surprises and once more the tabulation in Table II is based on the groupings indicated in Table I.

TABLE II

	GROUP			
Territory	A	B	C	D
Western Europe %	6.7	9.0	13.6	12.9
Eastern Europe %	.01	.8	--	.6
Japan	.5	2.5	1.5	2.6
Canada	.2	1.2	1.6	3.4
Other	1.5	1.2	4.2	5.8

If you check, you will note that the data for Group D in Table I does not correlate with the data for Group D in Table II. That is because only two of the three companies reporting chose to break out their export sales by country. Although export sales to Western Europe show an otherwise uniform progression with company size, sales to the other areas appear to follow a more random pattern. In many instances, it is the statistics from one or two companies that make the major differences. For exam-

(continued on page 21)

Now you can add famous Rust-Oleum® protection to aviation obstruction markings.

We're up to any job. Rust-Oleum offers a full line of tower coatings for marking aviation obstructions, such as microwave transmission towers, antennas and smokestacks. Our Aviation Surface White and Orange finish coats meet FAA specifications, including standards for luminosity and chromaticity. These durable coatings are available in high-gloss and flat finishes.

High quality, lasting protection. Backed by the right Rust-Oleum primer, our tough vinyl and water-base acrylic tower coatings give years of low-cost protection against weathering and corrosion without dulling or fading. No primer is usually needed on galvanized steel surfaces.

Easy on, easy clean-up. Rust-Oleum tower coatings can be applied with brush, roller, spray or paint mitten. And because they're water reducible, clean-up is fast and simple.

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ple, only one company in Group A and only two companies in Group B reported shipments to Eastern Europe. If this exercise were to be repeated, other countries and areas should be added to the list, for example, Israel, the Middle East, South America, Africa, etc. In addition, the countries in Western Europe should be listed separately.

The discussion to this point gives the results of the first two questions. What follows is a tabulation and analysis of the results question by question. Group E consists of those companies who did not submit financial data.

HOW DO YOU SELL OUTSIDE THE U.S.?

TABLE III

	GROUP				
	A	B	C	D	E
1. Sales Representative	6	7	6	3	4
2. Company Sales Personnel	1	1	1	2	
3. Export Import Agency	3	3			1
4. Catalogue					2
5. Other			1	1	1

The numbers will be greater than the number of companies, since several companies listed more than one sales technique. That will be true for most of the remaining questions.

The results clearly indicate that for the size companies we are discussing and for the business they are in, sales representatives are the most popular way to sell outside the U.S. Very often a small company or one with no experience outside the U.S. will use an export-import agency. This may be desirable with certain types of products but with conventional microwave products the company must be more involved in the sales effort. And as any good sales representative whether inside or outside the U.S. will state, the principal must respond, communicate and visit to achieve optimum results.

DO YOU HAVE A SEPARATE CATALOGUE?

Only two companies reported having separate catalogue for sales outside the U.S. The fact is that technical catalogues, in whatever language, are easily understood by most engineers. In addition, most non-U.S. engineers are familiar with spoken and written English. Only one additional cautionary note. Is it advisable not to include prices in the main text of the catalogue. One useful technique is to prepare a separate price sheet (even for U.S. use) since non-U.S. prices are affected by duty and taxes, over which the U.S. manufacturer has little control.

DO YOU PURCHASE SPACE ADVERTISING FOR SALES OUTSIDE U.S.?

TABLE IV

	GROUP				
	A	B	C	D	E
Yes	1	6	2	2	4
No	8	3	4	1	3

Here the results are not clear. One confusing element may be the excellent international circulation of the professional and trade publications produced in the U.S. An ad placed in the international edition of most good U.S. professional or trade publications will reach the market outside the U.S.

DOES YOUR COMPANY EXHIBIT IN SHOWS OUTSIDE THE U.S.?

TABLE V

	GROUP				
	A	B	C	D	E
Yes	1	9	6	3	5
No	2	0	0	0	2

Of those answering yes, there were the further questions as to whether they exhibited in their own stand (booth) or their Reps' stand and how many shows per year.

TABLE VI

	GROUP				
	A	B	C	D	E
Reps Stand	5	8	3		2
Own Stand			2	1	1
Both	2	1	1	2	2
Average No. of Shows Per Year	2.3	2.9	3.0	4.0	2.7

The exhibits are an important sales tool for sales representatives outside the U.S. The exhibits tend to be more sociable but no less business oriented. Here, if there is a language difference, it can be difficult for the U.S. principal unless he speaks the language. But the exhibits are worth attending periodically, but not every exhibit every year. The changes are too gradual to justify the time and expense.

HOW MANY WORKING MAN-DAYS DO PEOPLE FROM U.S. FACILITY SPEND ON SALES OUTSIDE U.S.?

TABLE VII

	GROUP				
	A	B	C	D	E
Average No. of Days	13	26	40	150	33

These results indicate the obvious, that the larger the company side the U.S. In our own experience this is perhaps the most important factor (other than product) in sales, both inside and outside the U.S.

HOW SOPHISTICATED ARE THE PRODUCTS YOU SELL OUTSIDE U.S. COMPARED TO PRODUCTS SOLD IN U.S.?

TABLE VIII

	GROUP				
	A	B	C	D	E
More sophisticated			1		
Less sophisticated	1	2		1	1
Same	8	7	6	2	6

This is not a surprising result. However, based on past experience (10 years ago) there would have been more votes for less sophisticated products. Products being sold outside the U.S. are essentially similar to those sold inside. Years ago, sales were probably made through the catalogue or advertising alone, but with the increase in travel and sales representatives' experience, the shift has been in the direction indicated.

HOW DO YOU PRICE PRODUCTS SOLD OUTSIDE THE U.S.?

TABLE IX

		GROUP				
		A	B	C	D	E
1.	Sell to rep at discount he sets final price	4	8	5	2	4
2.	Sell to rep at U.S. price he sets final price	4		1		1
3.	Same as U.S.	2	1	2		
4.	Special price list lower than U.S.					1
5.	Special price list higher than U.S.		2	1	2	2
6.	U.S. prices, plus shipping costs, plus import duty		1	1		1
7.	All prices negotiated			1		2
8.	Other	1	2			

One of the reasons for some companies indicating more than one pricing procedure stems from the fact that they use different pricing techniques for different countries. In examining the results as compared to % of sales outside U.S., there appeared to be no significant correlation. One of the advantages of selling to the sales representative at a discount (usually the catalogue price, less the U.S. rep commission rate) is that it reduces the import duty and hence makes the product a bit more competitive. The disadvantage is that unless you have an agreement with the rep as to the price to the customer, he is free to set his own price.

IS COMMISSION BASED ON A SLIDING SCALE?

TABLE X

	GROUP				
	A	B	C	D	E
Yes		3	4	1	1
No	9	6	2	2	6
At level of commission $10,000		1	1		
$25,000		2	1		
$50,000			2	1	
$100,000					1

Here too there was no correlation between these results and the % of sales outside the U.S.

TERMS OF SALE?

TABLE XI

		GROUP				
		A	B	C	D	E
1.	Open account	7	8	5	2	4
2.	Letter of credit	1	7	4	3	6
3.	Sight draft			2	2	2
4.	Time draft					
5.	Authority to purchase					
6.	Consignment					1

Two of the companies listing multiple terms of sale were specific in citing special terms for certain countries. This was perhaps true of the other companies listing more than one term of sale.

COLLECTION SPEED?

TABLE XII

	GROUP				
	A	B	C	D	E
30 days	1	2	1		
60 days	6	4	1	1	4
90 days	2	3	4	2	2
COD					1

Although the collection speed seems to be a bit slower than for U.S. sales it is a decided improvement over collection speed of a few years ago. But just as in the U.S., one must be persistent in collecting from customers outside the U.S.

RFQ VS. CATALOGUE ITEMS?

The results of this question were compiled and compared to % of total export sales as shown in Table XIII.

TABLE XIII

% OF TOTAL EXPORT SALES

	GROUP			
	A	B	C	D
Competitive Quotes	60	33	52	31
Catalogue Sales	38	66	31	64
Other	2	1	17	5

Here the results appear to be random. This may be due to the fact that the companies reporting vary from one extreme wherein the company has an extensive range of catalogue products and chooses not to deviate, to the other extreme of the engineering-systems type company that chooses to bid every program.

HAVE YOU NOTICED AN IMPROVEMENT BECAUSE OF DOLLAR DEVALUATION?

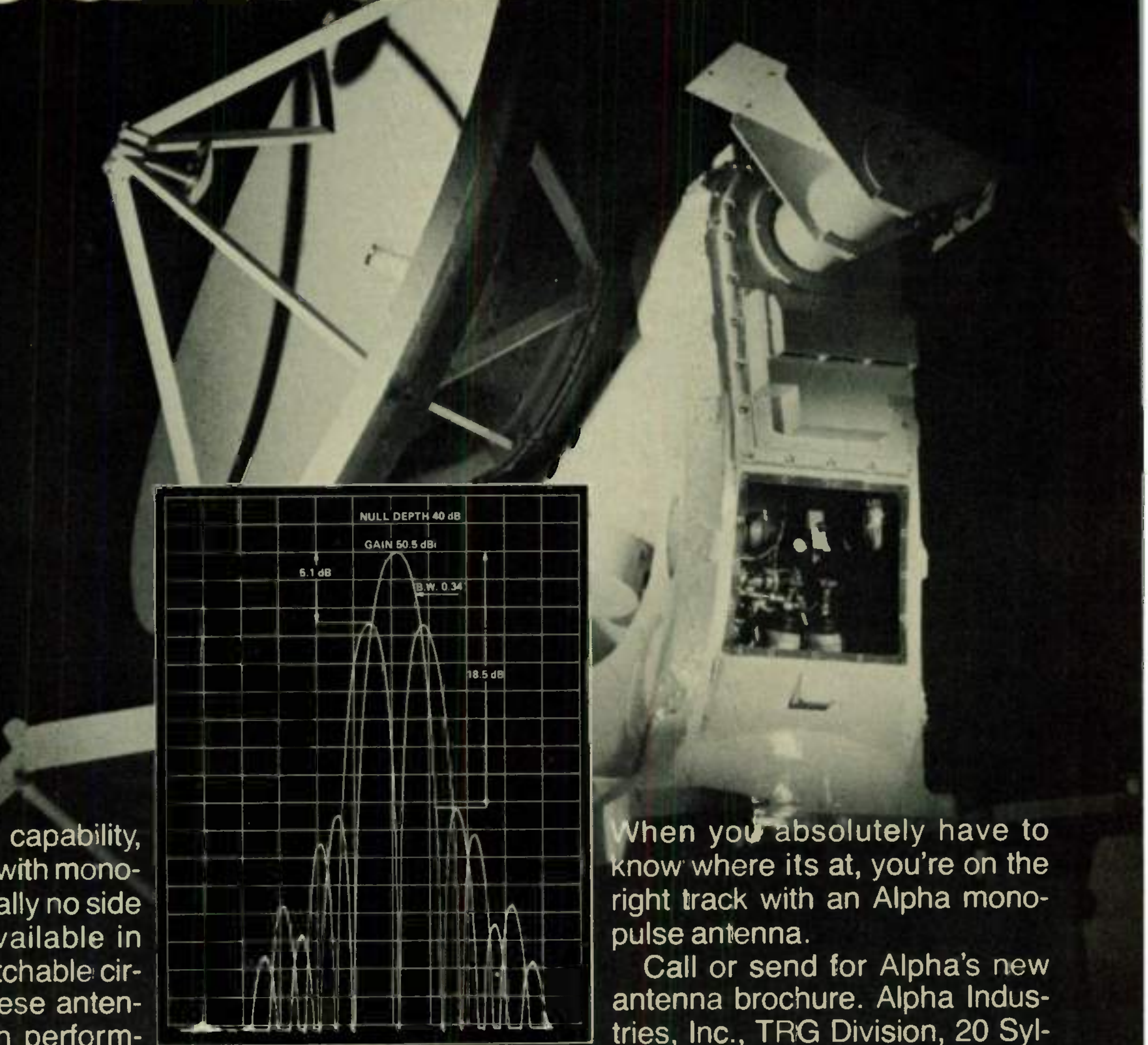
TABLE XIV

	GROUP				
	A	B	C	D	E
No	3	3	1		5
Some	6	4	5	2	2
Significant		2		1	

SUMMARY

In conclusion, it would appear that there are differences in our various export techniques. However, with the exception of travel outside the U.S. there does not appear to be one overriding technique which above all others gives rise to export sales. Rather, it goes back to the basics: good product, good sales effort, good communication, good service and occasionally being in the right place at the right time. And if you accept the definition that luck is when preparation meets opportunity, then one must add luck.

For some people tracking accuracy means survival.



Superior in tracking capability, Cassegrain antennas with monopulse feeds have virtually no side lobe degradation. Available in dual linear and/or switchable circular polarizations, these antennas combine the high performance of a Cassegrain system with sophisticated monopulse versatility . . . the Alpha advantage.

When you absolutely have to know where its at, you're on the right track with an Alpha monopulse antenna.

Call or send for Alpha's new antenna brochure. Alpha Industries, Inc., TRG Division, 20 Sylvan Road, Woburn, MA 01801. (617) 935-5150 TWX: 710-393-1236 TELEX: 94936

	−3dB Beamwidth (Nominal)	Side Lobe Level (Nominal)	Net Gain (Minimum)	Null Depth (Minimum)	Insertion Loss (Minimum)
94GHz 18" Diameter	0.6°	−18dB	46.0dBi	−30dB	2.50B
70GHz 18" Diameter	0.65°	−18dB	46.0dBi	−30dB	1.5dB
53GHz 36" Diameter	0.45°	−18dB	50dBi	−35dB	1.5dB
35GHz 96" Diameter	0.25°	−18dB	54dBi	−35dB	1.0dB
9GHz 120" Diameter	0.8°	−18dB	46dBi	−35dB	0.5dB

ai Alpha

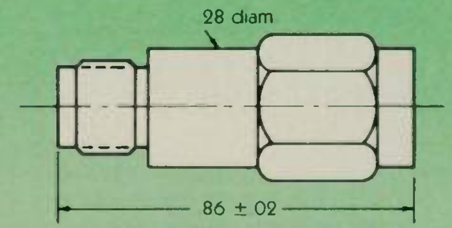
The Alpha Advantage

THE MINIPAD ATTENUATOR

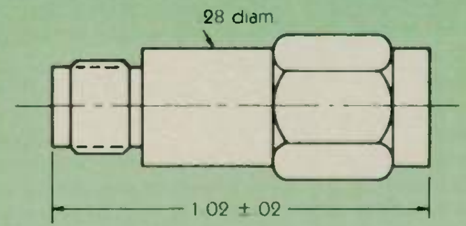
OUR SMALLEST, LIGHTEST, HIGH PERFORMANCE FIXED ATTENUATOR.

This miniature, high performance fixed attenuator has been designed to meet the needs of today's sophisticated microwave systems. The Minipad will operate within the stated specifications in most any hostile environment be it airborne, shipboard or land application. VSWR is typically less than 1.15 at 12.4 GHz and less than 1.25 at 18.0 GHz while the frequency sensitivity is normally less than 0.05 dB per GHz. The Minipad uses the Unique tube attenuator developed at Midwest Microwave. The tube attenuator is inherently a very reliable element because of its basic construction simplicity. The connectors for the Minipad attenuator are fabricated from stainless steel. The close tolerance machining plus the precise captivation techniques ensures an excellent match over the complete frequency range. All Minipads are production tested using the latest state-of-the-art swept frequency techniques. This complete testing assures that every attenuator will be within the published specifications.

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



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



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

DC to 18.0 GHz
HIGH PERFORMANCE

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

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

DC to 12.4 GHz
HIGH PERFORMANCE

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

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

DC to 8.0 GHz
HIGH PERFORMANCE

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

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

DC to 2.0 GHz
HIGH PERFORMANCE

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

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

DC to 18.0 GHz
INEXPENSIVE

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

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

Double Male Connectors
(M PREFIX TO MODEL NO.)

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

Double Female Connectors
(F PREFIX TO MODEL NO.)

dB VALUE	OVERALL LENGTH
1-12dB	.90"
13-30dB	1.03"

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Around the Circuit

PERSONNEL

Edward J. Sandor was appointed Sales Engineer for Rogers Corp.'s Minneapolis, MN territory. . .**John Mulroe** has joined Thomson-CSF Electron Tube Div. as Product Sales Mgr. of Power Grid and Image Tube products. . .Scientific-Atlanta, Inc. appointed **Samuel D. Davis** as General Mgr. of its Atlanta Instrument Div. . . **Eugene Kushner** was appointed to the newly created position of Dir. Corporate Development at Polarad Electronics, Inc. . .**Larry Peterson** joins Plessey Optoelectronics & Microwave as Nat'l Marketing Mgr. of Optoelectronics. . .**Glenn DeBella** becomes General Mgr. of Narda Microwave Corp.'s Pacific Coast Operation. . . Leasametric, Inc. appointed **David J. Sobo** as Nat'l Sales Mgr. . .**Harry Marschausen** was named Nat'l Sales Mgr. at JFD Electronic Components. . .**Michael D. Minkiewicz** becomes Dir. of Advanced Programs for Eaton Corp.'s AIL Div. . .Alpha Industries, Inc. appointed **Joseph J. Diesso** as Regional Sales Mgr., Optimax Div. . .**Frank E. McDonnel** was promoted to V.P., American Electronic Labs, Inc. . .Adams Russell's Anzac Division named **Mark R. Rosenzweig** as V.P. and Dir. of Marketing.

CONTRACTS

American Electronic Laboratories, Inc. received a $3M award from the Northrop Corp., Rolling Meadows, IL for mw equipment for an AF electronic device and a $792.8K contract from **Hughes Aircraft Co.** for a receiver and special antenna to be used in the US Roland program. . .**Anaren Microwave, Inc.** received a subcontract exceeding $3.5M from **Emerson Electric Co.'s Electronics and Space Div.** for production of advanced receivers for the US Army AN/MSQ-103 Teampack system. . .**California Microwave, Inc.** received a contract valued at $600K from the **Mutual Broadcasting System** to provide the uplink for the MBS satellite broadcast network. . .Royal Australian Navy awarded **E-Systems, Inc.** a $15M order for UHF shipboard radio equipment, and the **ECI Div.** of E-Systems was granted a five-year $46.2M contract to produce UHF shipboard radio terminals for the US Navy. . .**Harris Corp.** received a $1.5M order from **Martin Marietta Corp.** for a satellite communications network to handle high-volume integrated voice and data communications between Martin Marietta facilities in Denver, Baltimore and Orlando, FL. . . .**M/A-COM, Inc.** announced that one of its operating companies, **Microwave Associates, Inc.**, received a $1.4M contract from **Hughes Aircraft Co.** for mw components for the US Roland missile program. . .**Central Telephone Company (CENTEL)** of Texas has awarded **Valtec Corporation** a contract to supply a 8.57 km 14-strand fiber optic cable for the country's first 90MBPS fiber optic telephone system. . .**Watkins-Johnson Company** received a $3.5M contract from the USAF to extend the capabilities of the QRC-259 receiving system.

INDUSTRY NEWS

Soladyne, Inc. was acquired by **Rogers Corp.** in a stock transaction involving 103,500 shares of Rogers common stock with a market value of about $2M. . . **VARI-L Co.** moved its corporate headquarters into a 22,000 sq. ft. building in Denver, CO. . .**Antennas For Communications (AFC)** was acquired by **Microdyne Corp.** in an exchange of .875 shares of Microdyne for each share of AFC on April 16, 1980. AFC will continue its operations as a wholly owned subsidiary of Microdyne. . .**Loral Corp.** and **Frequency Sources, Inc. (FSI)** announced an agreement in principle for the merger of **FSI** into **Loral**. The transaction will involve an exchange of stock valued at $46.4M with an exchange ratio of .75 shares of **Loral** stock for each of **FSI**'s 2.2M outstanding shares, subject to adjustment prior to merger. . .**Hughes Aircraft Co.** and **ITT Corp.** announced the formation of TADCOM, a joint venture which is competing for full-scale development of Class 2 tactical terminals for JTIDS. . .**Satellite Transmission Systems, Inc.** will be acquired by **California Microwave, Inc.** for an undisclosed consideration following completion of Aug. 1980 audits of both companies. . .**Southern Pacific Communications Co. (SPCC)** announced that it has applied for authority to construct and operate a $200M domestic communications satellite system. . .An RFP has been issued by the AF for the design, full-scale development and testing of a low cost expendable mini-drone (LOCUST) harassment system. Proposals are due at Aeronautical Systems Div., Wright-Patterson AFB on July 1, 1980. . .**Erik A. Lindgren & Associates**, a Chicago-based manufacturer of screen rooms, has been acquired by former General Instrument exec. Williams E. Curran and is being restyled as **Lingren RF Enclosures, Inc.**

FINANCIAL NEWS

Adams-Russell reported second quarter results for the period ended March 30, 1980 of net sales of $8.4M, net income of $571K, or 31¢ per share (adjusted). This compares with 1979 quarterly net sales of $7.0M, net income of $395K, or 22¢ per share (adjusted). . .**Scientific-Atlanta, Inc.** reported nine-month net earnings of $8.7M, sales of $134.9M and earnings per share of $1.79 for the period ended March 31, 1980. For the nine months of 1979, net earnings were $5.3M, sales were $90.2M and earnings per share were $1.27. . . For fiscal year 1979, **Rogers Corp.** reported net sales of $92.9M, net income of $4.4M or earnings per share of $1.71. This compares with 1978 year-end net sales of $75.2M, net income of $3.2M or earnings per share of $1.29. . .**Omni Spectra, Inc.** reported for the six months ended March 29, 1980 sales of $15M, net income of $631K or 24¢ per share. This compares with 1978 half-year sales of $12.8M, net income of $149K or 6¢ per share. . .For the nine months ended March 31, 1980, **Narda Microwave Corp.** had sales of $14M, net income of $729K or earnings of 96¢ per share. This compares with 1979 nine-month results of sales of $12.3M, net income of $398K or earnings per share of 56¢. . .For the first quarter ended March 29, 1980, **Frequency Sources, Inc.** reported net sales of $8.1M, net income of $481K or earnings per share of 20¢. In the same quarter of 1979, net sales totaled $6M, net income was $320K and earnings per share were 17¢.

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linear power amplifiers...

We're in a class by ourselves!

Series LWA and LAB; up to 200 watts power; frequencies from 1MHz to 8400MHz

Higher powers, higher frequencies, more design and performance features, more models to fit your applications—that's the continuing story of MPD Class A solid state linear amplifiers. The result is a product line that nobody else in the industry—*repeat, nobody*—can even come close to matching!

You can now choose from 225 standard MPD models, available in module or rack-mounted cabinet configurations for systems applications, as well as self-contained instruments for laboratory use. Ultra-broadband frequency ranges from 1-1000MHz up to 7900-8400MHz, including our newest high power model with 200 watts saturated power rating at 500-1000MHz.

These Class A amplifiers are particularly recommended for applications requiring exceptional linearity and wide dynamic range. They combine low noise figure with high power output to yield low distortion amplification of both single and multi-carrier signals with any form of modulation. Other standard features of most models include: high efficiency; high gain; built-in protection against DC input reversal, thermal overload and infinite load VSWR; graceful degradation; drift-free output; field-replaceable modular 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!

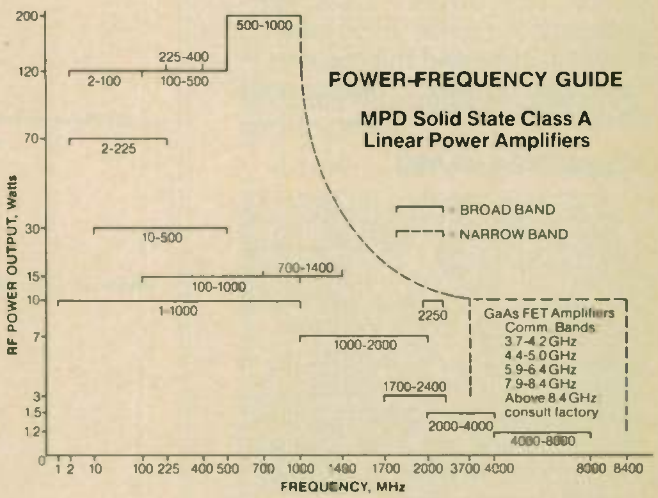

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TABLE III

MILLIMETER WAVE TUBES

FISCAL YEAR 80	81	82	83	84	85	PROGRAM TITLE AND KEY INFORMATION	APPLICATION
*	*	*				*MM/Near MM wave tubes* Complete developments of TWT and EIA and integrate with modulator from nanosecond pulser program into 20 pound, one cubic foot package	Tank and airborne (RPV) radar
	*	*				*High Current Density Cathodes* Tungstate cathodes to provide 10-fold increase in life to several thousand hours at 10 A/cm^2	1 - 100 kW peak power tubes
**	**	*				*Interaction Circuits for MM Wave Tubes* New circuits to overcome dimensional and thermal problems in metal circuits	MM & Near-MM tubes
**	**	*				*T/R Devices for Near-MM Region* Passive, high power techniques	Passive receiver protection devices for single antenna radar
**	**					*Low Magnetic Field Magnetron Oscillators* 95 GHz, 1 kW peak power, using samarium cobalt magnets and electron discharge machining for anode fabrication	Lightweight (.5kg) oscillator for radar
**	**					*RF Generation using Non-Linear Mixing in Crystals* Drive lithium niobate with 35 GHz magnetron	Radar Transmitters
**	**					*Backward Wave Oscillator Development* Establish US source for tunable near-MM power	O-type carcinotrons
*	*	*				*3.2 MM Wave Transmitter Tube (Advanced Development)* 1 kW peak, 50 dB gain, under 15 pounds using PPM and depressed collectors	Lightweight radar transmitters

* FUNDED CORE PROGRAM
** UNFUNDED INCREMENT 1

TABLE IV

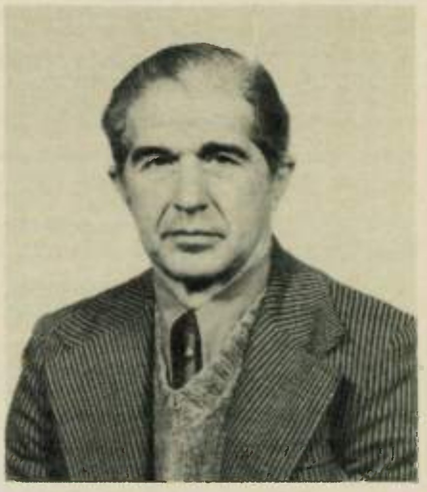
FISCAL YEAR 80	81	82	83	84	85	PROGRAM TITLE AND KEY INFORMATION	APPLICATION
*	*	*				*Nanosecond Pulsers* Fast plasma cathode switches, magnetic nanosecond pulsers, miniature high frequency inverter power supplies. Address laser application first, then MM wave.	MM radar and CO_2 lasers

* FUNDED CORE PROGRAM

and frequency of the TWT. Lightweight, periodic permanent magnet focusing and depressed collector techniques are applied in a current development to achieve 25% efficiency in a 50 dB gain TWT. Timing, application, and key information on the individual programs are tabulated in Table III.

NANOSECOND PULSERS

Target classification requires a minimum pulse width of 2 nanoseconds to recognize tanks, trucks, and artillery. A pulse repetition rate of at least 20 kHz is needed to maintain the required average power on the target. A package of 400 cm^3 or less is necessary for a mini-RPV. Laser risetime requirements are not as stringent, nor is the repetition rate requirement, but some missile beacon requirements need a 200 cm^3 package. Current technology does not meet the requirements and the planned programs (see Table IV) benefit both the mm wave and the laser systems.



George C. Uchrin received his B.S. in E.E. from Rutgers University in 1949 and joined the US Army Signal R & D Laboratories the same year. In the 1950s, he pioneered the development of transistorized power converters and engaged in the Army's early major drone surveillance programs, AN/USD-4 and 5, and guided the Cornell Aeronautical Lab in its development of mathematical modeling of complete drone surveillance systems. In 1960, Mr. Uchrin joined the Electronics Technology and Devices Laboratory as a member of the Army's management group which guided high power klystron tube developments for the Nike Zeus discrimination and target track radars. During the next decade, he served as ET&DL planning coordinator under QMDO (Qualitative Materiel Development Objective). He continues as the Army's chairman and principal contributor in the electronic devices technology area for the DoD.

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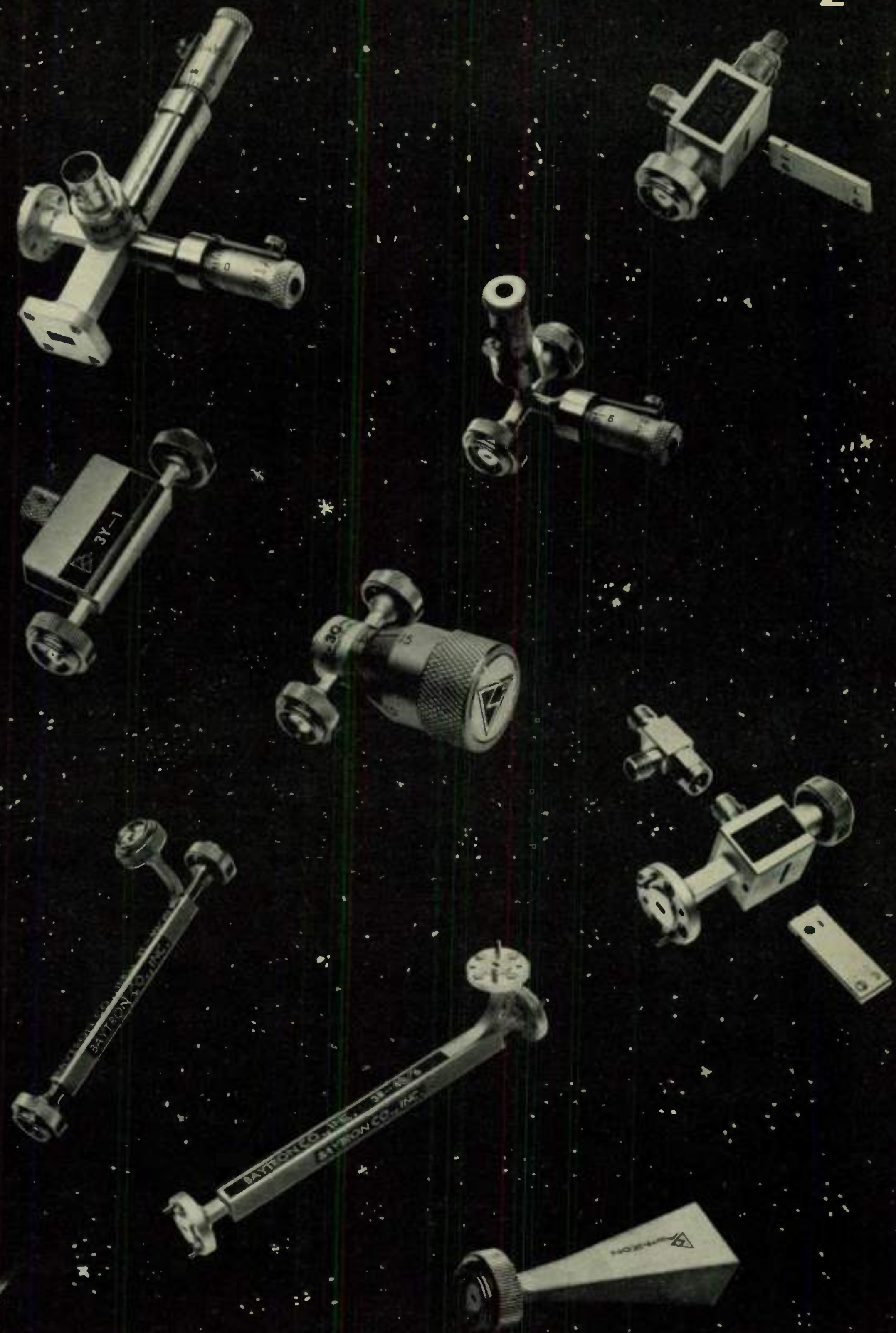

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System Analysis for Millimeter-Wave Communication Satellites

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INTRODUCTION

NASA studies[1,2] have indicated there will be a significant increase in both the applications and volume of satellite communications in the 1980-2000 time frame. Associated with an increase in demand is the potential problem of spectral crowding. Obviously some method of achieving higher capacity is necessary. One means of obtaining spectrum relief is to expand the communications services upward to the millimeter-wave region of the spectrum. The larger bandwidths available at these frequencies will provide capabilities for higher data rates, and the possibility of extremely narrow beams can lead to very high reuse of the frequency assignments.

Traditionally, United States industry has enjoyed a unique capability which has led to marketing of US satellite technology abroad. Introduction of proven US millimeter technology could have a part in maintaining this industrial position. Hence, there exists a need to investigate the technology associated with use of the millimeter-wave region of the spectrum for satellite communication applications.

Editor's Note: Part II of this article will appear in a subsequent issue. It will feature a technological assessment, including such aspects as propagation, bulk data storage, space switching equipment, receiver and transmitter and satellite antennas plus general conclusions.

With the potential millimeter services partially identified by previous studies, the overall objective of this program has been to identify the technologies necessary to satisfy those services and to assess the relative risks of these technologies. This paper is based upon specific objectives of the program. These are: (a) to develop a methodology based on the technical requirements of potential services that might be assigned to millimeter-wave bands which would identify viable and appropriate technologies for future NASA millimeter-wave research and development programs, and (b) to test this methodology with selected user applications and services.

While the current application of communication satellites is primarily for point-to-point communications among a small number of relatively sophisticated ground stations tied into terrestrial communications systems, future applications might also include a broadcast mode where many small, inexpensive ground stations would be able to communicate via a larger more powerful communication satellite. Applications of such a system might include direct wideband data or video lines (for teleconferencing) between corporation locations using rooftop antennas. The wide-bandwidth and narrow beam potential of the millimeter-wave frequency band offers advantages for such broadcast applications, but anticipated difficulties associated with high attenuation of the signal by atmospheric weather conditions must be overcome.

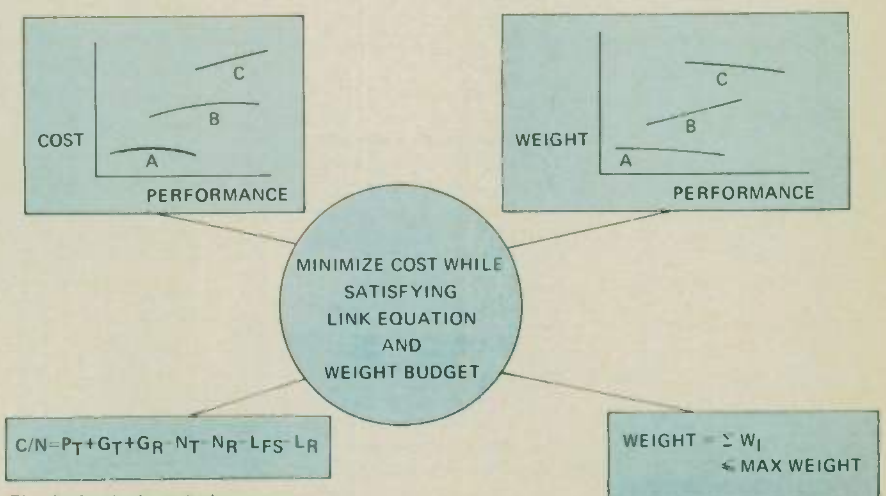
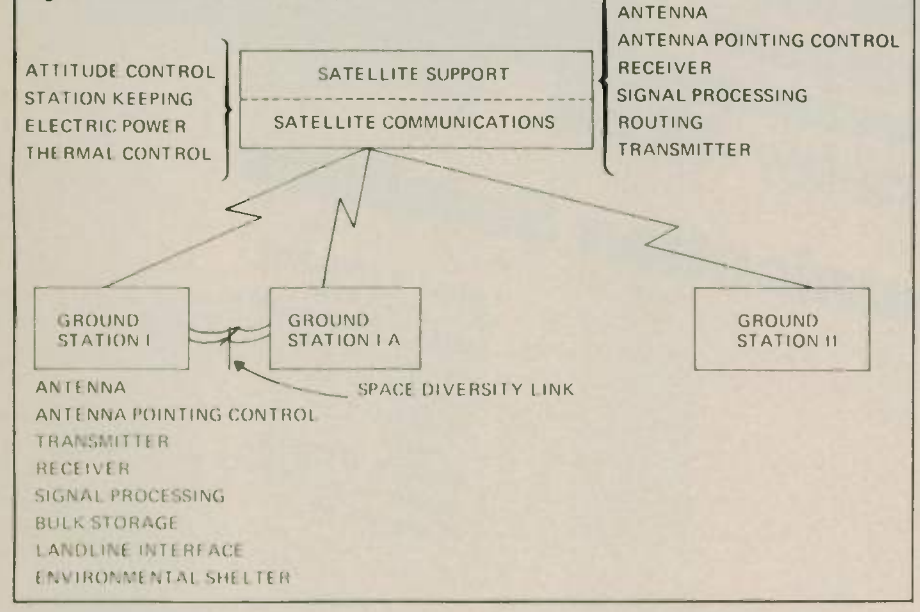


Fig. 1 Analysis technique.

Fig. 2 Ground and satellite subsystem categories.



METHODOLOGY

The program objectives have been met by a methodology which uses an appropriate level of detail in the subsystem models employed and in the numerical optimization procedure used for trade-off analyses. After a review of the pertinent literature, the applicable subsystem models available from SAMSO[3] and Hughes Aircraft Corporation[4] were selected as the basis for the subsystem model library. Models for the remaining subsystems were established from published specifications and from contact with personnel in the space communications industry. The overall communications link equation (received) carrier-to-noise ratio) was written in terms of the independent performance parameters in the subsystem models. The total satellite system weight was expressed in terms of the same independent variables. Lower and upper bounds on the performance variables of all subsystem models were established, and a computerized random-search optimization procedure was adopted for selection of the minimal cost (annual cost per channel to the user) system. The optimization procedure was then utilized to establish baseline design of the point-to-point application and of the broadcast application.

A satellite communication system requires both ground and satellite subsystems; the satellite subsystems can be further divided into the communications link and housekeeping subsystems by tracing the complete routing of a communication message from its initial arrival at the transmitting ground station to its final departure from the receiving ground station. The interrelationships between the cost models, weight models, link equation, and weight

TABLE 1

SUBSYSTEM COST AND WEIGHT MODEL DRIVING PARAMETERS

Subsystem Cost Models

Subsystem	Driving Parameters
Ground Antenna	Dish Diameter Transmitter Frequency
Radome	Radome Diameter
Ground Pointing and Control	Pointing Error Dish Diameter
Ground Transmitter	Transmitter Power Transmitter Frequency
Ground Receiver	Receiver Noise Figure Receiver Frequency
Ground Signal Processing	Baseband Channel Bandwidth
Bulk Data Storage	Data Rate Storage Volume
Landline Interface	Data Rate Number of Television Headins Number of Voice Multiplexers
Diversity Link	Diversity Range
Satellite Antenna	Antenna Diameter Operating Frequency Number of Feeds
Satellite Transmitter	Transmitter Power Operating Frequency
Satellite Receiver	Noise Figure Operating Frequency
Satellite Signal Processing	Number of Channels Number of Subchannels per Channel
Attitude Control System	Attitude Control System Weight
Station Keeping System	Station Keeping System Weight
Structure and Thermal Control	Structure and Thermal Control Weight
Satellite Power Supply	Prime Power Required

Subsystem Weight Models

Subsystem	Driving Parameters
Satellite Antenna	Antenna Diameter Operating Frequency Number of Feeds
Satellite Transmitter	Transmitter Power Operating Frequency
Satellite Signal Processing	Number of Channels Number of Subchannels per Channel
Attitude Control System	Attitude Control Error Satellite Weight
Station Keeping System	Station Keeping Accuracy Satellite Weight
Structure and Thermal Control	Satellite Weight
Satellite Power Supply	Prime Power Required

budget during system optimization are demonstrated in Figure 1. A detailed description of these elements of the analysis methodology follows.

The ground and space subsystems and their categorizations are indicated in Figure 2. This shows the specific subsystems that were modeled to represent the overall communication link. Parametric cost and weight models were formulated for each of the subsystems included in the satellite/ground configurations. In most cases there is one major driving parameter affecting the cost while several minor parameters are used to specify features of the configuration. The weight models normally have the same independent variables as the corresponding cost models. In cases where total satellite weight is the independent variable for a subsystem weight model, an iterative technique is required for computations. A summary of the cost and weight model driving parameters is given in Table 1. The cost models for the subsystems (and the weight models for the spacecraft subsystems) are in terms of the subsystem performance parameters which appear in the communication link equation. The individual subsystem models are applicable over a specified range of the performance parameters, and the models are continuous over the allowable range of the performance parameters.

The methodology for optimization of the communication link selects all subsystem performance parameters in such a way that the overall link carrier-to-noise ratio requirement and the satellite weight constraint are satisfied and the total system cost is minimized. A random search algorithm which uses a computerized random number generator to select trial points over the parameter intervals has been developed and used for most of the optimizations performed during the program. The algorithm reduces the parameter interval in successive optimizations until the density of random points selected is quite high in the final optimization step. This methodology has proven to be effective and efficient. However, for applications in which the optimal solution lies on the weight boundary, the random search algorithm requires a significant increase in computer time. As a result, an interactive man-in-the-loop gradient search algorithm was also developed as an option to the random search procedure. Use of this option (from a remote computer terminal) has significantly decreased the computer time for establishing the cost-optimal conceptual design of the satellite broadcast analysis.

The block diagram of the Satellite Cost Optimization Routine (SCOR) is shown in Figure 3. SCOR employs cost and performance models for satellite communication subsystems and numerical optimization routines to determine the satellite link design which will provide the specified carrier-to-noise ratio for a minimum total capital cost. Annual cost and channel capacity models have also been incorporated into SCOR and the computer program calculates and displays both the capital costs and the "annual cost per channel to the ultimate user" for the optimized communication system.

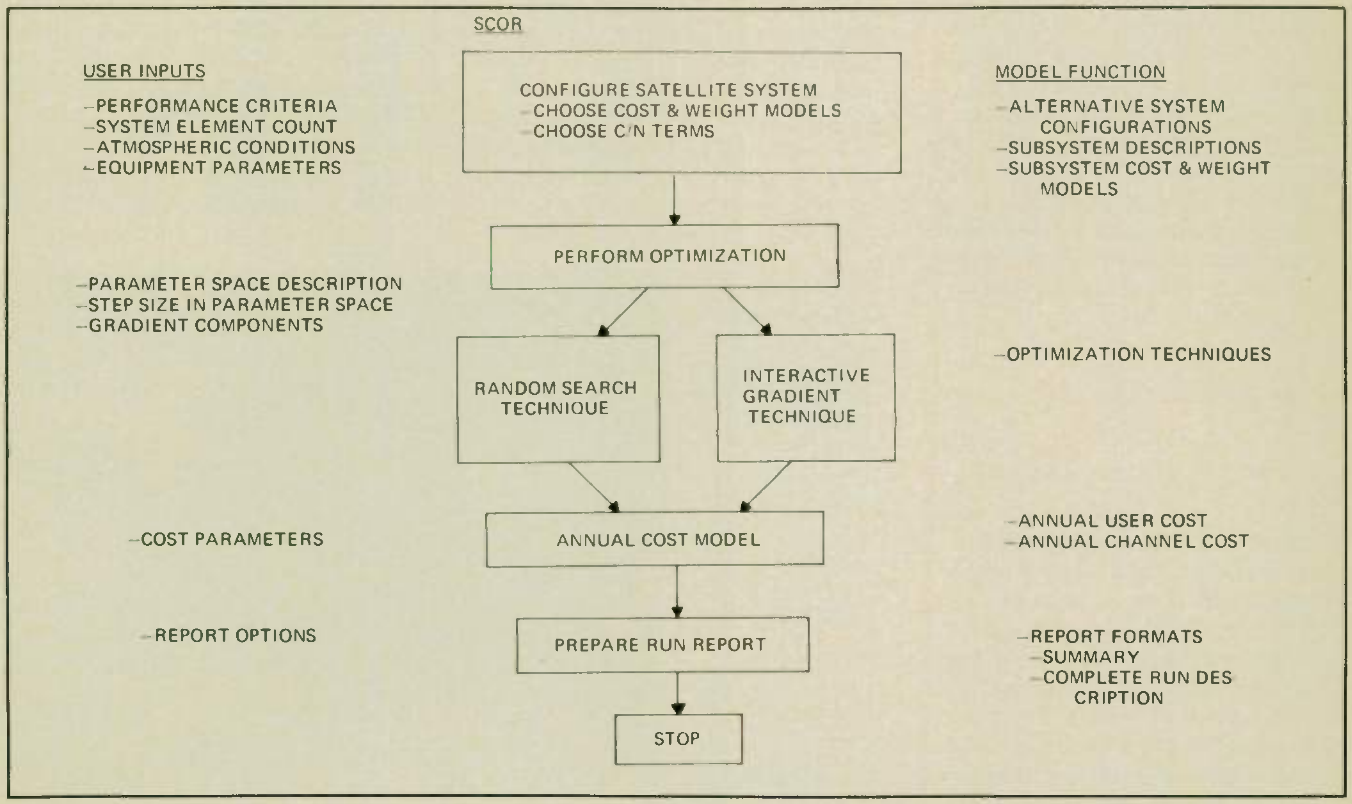


Fig. 3 Block diagram of the Satellite Cost Optimization Routine (SCOR).

Annual Cost and Channel Capacity Models

The following discussion indicates how these models are used within SCOR to provide additional insight into the economic viability of the Millimeter-Wave Satellite Concepts.

The annual cost model takes into account the capital investment for the satellite and ground systems, anticipated lifetimes of the satellite and the ground systems, and such financial parameters as the length of the financial planning horizon, the allowable return on investment in the regulated industry, the income tax rate applicable to the corporate venture, and an annual rate of escalation for operation and maintenance costs. Property taxes, fire insurance premiums, and ground system operation and maintenance costs are also included. The initial capital investment includes not only the satellites and the communication ground stations, but also the tracking telemetry, and control ground stations. The expression of annual cost of the system is a function of these parameters and takes into account the times at which the costs and revenues occur utilizing the concept of net present value. This uses a discount rate consistent with the rate of return allowed by the regulatory agency. The annual charge for the complete set of communication channels is related to the present value of the total allowable revenues by the following:

Annual Cost =

$$\frac{k}{1 - 1/(1+k)^H} \cdot PV(Rev_t)$$

where k is allowable rate of return and H is the length of the planning horizon. It is worth noting that the discounted annual cost is significantly greater than just the total revenue divided by the duration, H. For a typical case of 10% rate of return and an 8-year operation period, the annual cost is 50% greater than revenue divided by 8 years.

The above expression represents the annual charge for the entire communication system. The equivalent annual cost per channel to the user is determined by dividing that total annual cost by the effective number of channels; i.e., by the product of the number of simplex channels available and the utilization factor. Since channel capacity is a somewhat complex function of modulation, multiple-access technique, power levels, bandwidth, etc., the selected approach has been to start with results computed by COMSAT Corporation for INTELSAT IV and to denormalize those results to predict channel capacity for the millimeter concepts. The resulting channel capacity for the six transponders in the Application I Concept (point-to-point communications) is 66,300 simplex voice channels or 33,150 full duplex voice channels for frequency division multiplex (FDM). Similarly, the channel capacity for time division multiplex (TDM) in Application I is 123,672 simplex channels or 61,836 full duplex channels.

The annual cost calculation models are implemented within SCOR at a point which follows the equivalent capital cost optimization technique to minimize unnecessary computer time requirements. For those scenarios in which capital investments are made at different times throughout the financial planning horizon rather than just on initial investment, it would be necessary to locate the annual cost model inside the optimization loop in order to properly account for discounting of funds.

The parameters for variation in the annual cost to the user for each simplex channel with respect to utilization rate after the annual cost has been calculated for 100% utilization. The actual cost per channel is given by the fully-utilized rate per channel divided by the ratio of the leased channels to the total available channels. In this analysis no consideration has been made for primary and secondary channels with different charges for guaranteed channel availability.

APPLICATIONS

There are many potential applications of millimeter-wave communications satellites in both the public and private sector. This study used two basic systems which could be adopted for a variety of specific end users. For convenience, the two basic systems have been designated point-to-point and broadcast. The point-to-point system is considered to provide broadband (1 GHz) communications among a relatively small number of earth terminals, whereas the broadcast system provides narrowband (50 MHz) communications among a relatively large number of earth terminals. Both of the applications were based on a number of common assumptions. In order

Fig. 4 Application I — coverage area of point-to-point communications between cities (circled).

(continued on page 42)

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.

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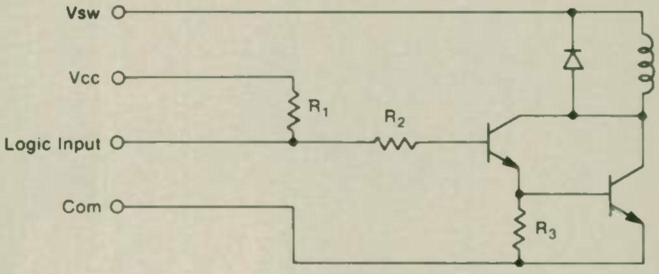
As a special option, on both failsafe and latching type switches, drivers can be provided which are compatible with industry standard low power Schottky TTL circuits.

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
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RF PERFORMANCE

Frequency	CS-33, 38 SERIES 0-6 GHz	CS-33, 38 SERIES 6-12 GHz	CS-33, 38 SERIES 12-18 GHz	CS-37 SERIES 0-4 GHz	CS-37 SERIES 4-12 GHz	CS-37 SERIES 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

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TABLE 2

POINT-TO-POINT APPLICATION BASELINE PARAMETERS

PARAMETERS			
Carrier/Noise Constraint Limit (dB)	15.00	Diversity Link Range (mi)	9.940
Weight Constraint Limit (lbs)	5000	Ground Station Building Cost (K$)	100.0
Downlink Frequency (GHz)	40.50	Diversity Station Building Cost (K$)	50.00
Uplink Frequency (GHz)	50.50	Marginal Income Tax Rate	0.48
Satellite Channel Bandwidth (MHz)	1000.	Rate of Return on Investment	0.13
Number of Channels (beams)	6	Financial Planning Horizon (years)	8
Number of Positions Per Beam	1	Life of Satellite (years)	8
Reliability (percent)	99.90	Life of Ground System (years)	14
Rain Rate (mm/hr)	50.00	Tax Constant	0.015
Number of TV Headins	12	Insurance Constant	0.012
Number of Voice Multiplexes	12	Cost of Debt	0.085
Digital Data Rate (Mbs)	3.000	Ratio of Debt to Total Capitalization	0.45
Bulk Data Rate (Mbs)	200.0	Fraction of Channel Sellable	0.50
Bulk Data Volume (Mb)	1000.	Average Growth of Operating Costs	0.065
Number of Ground Stations	6	Satellite Operating Cost Constant	0.01
Ground Transmitters Per Link	6	Ground System Operating Cost Constant	0.04
Ground Receivers Per Link	2	Launch Cost (K$/lb)	5.0
Channel Capacity	66,300	Launch Insurance Rate	0.1
Number of Subchannels Per Channel	5	Number of Satellites Purchased	3
Ground Station Bandwidth (MHz)	1000.	Number of Launches	2
Diversity Link Receive Cost (K$/mi)	100.7	Uplink Misc. Losses (dB)	7.000
Diversity Link Transmit Cost (K$/mi)	40.30	Downlink Misc. Losses (dB)	8.000
		Atmosphere Temperature (°K)	300.0

to avoid the necessity of using ground station tracking antennas, the satellites were assumed to be in a geostationary orbit (about 35,000 km) positioned over the middle of the continental United States. An available RF bandwidth of one GHz was assumed for both applications on the uplink and the downlink. The uplink frequency was considered to be in the 50 GHz band while the downlink was considered to be 40 GHz.

Application I: Point-to-Point

A baseline conceptual system was developed for the point-to-point application which uses six ground stations, each with single station diversity for both receive and transmit. Figure 4 shows the geographical coverage area. No radomes are used for the ground station antennas. For baseline analysis channelization is assumed to be by frequency-division multiplex. As for all analyses performed to calculate system cost, the cost for the baseline system was minimized under carrier-to-noise and weight constraints by the computer program SCOR. A complete set of the parameters required for input to this cost optimization is given in Table 2. Included are system constraints, system configuration parameters, and various assumed constants.

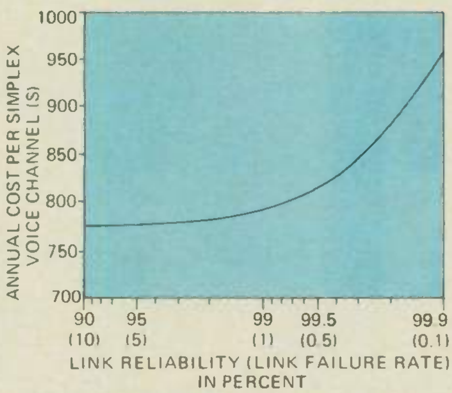

Fig. 5 Annual cost per channel versus link reliability for point-point service at 40/50 GHz (50% utilization).

The results for a three-satellite (two in orbit, one spare on ground) system present annual cost data in addition to capital costs. The total capital cost for the optimized system is $112.7M. This translates to an annual system cost of $31.8M and a per simplex voice channel annual cost of $959 (for 50% utilization). Figure 5 shows that as link reliability increases from 90% to 99.9%, the annual cost per simplex voice channel increases from $775 to $959 (for 50% utilization).

The number of ground stations was varied from 2 to 10 to examine the effect of this change on per terminal cost. This was done for both FDM and TDM signal processing to determine changes in the relative attractiveness of these two techniques. The results are plotted in Figure 6. The increasing cost for FDM as a function of the number of terminals and the generally lower cost for TDM than FDM are both due to the fact that FDM channel capacity decreases much more rapidly than TDM channel capacity as more terminals are added to the system.

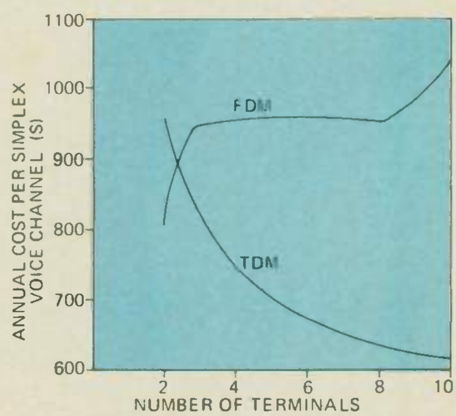

Fig. 6 Annual cost per channel versus number of terminals — FDM and TDM fixed point-to-point system at 40/50 GHz for 99% reliability with 50% utilization.

Application II: Broadcast

This application considers the interconnection of a large number of earth stations throughout the United States. Total ground coverage is required although not simultaneously. In concept, however, an earth station located anywhere within the US should be able to communicate with an earth station at any other point in the US through this satellite. Each earth station must be capable of transmitting full bandwidth television or 1.544 Mbps data as a minimum.

The objective of the broadcast application concept was to provide total US coverage using adjacent spot beams with 99.5% reliability (rain considerations only) for wideband uses such as video distribution. Preliminary power calculations indicated that very large (heavy) satellites would be required for this concept, and a compromise baseline design with limited simultaneous beam utilization and with on-board switching was developed. This design provides up to 96.5% link reliability with the assumed subsystem constraints (e.g., satellite weight). Other system configurations such as multiple satellites or a very large satellite could possibly achieve the desired 99.5% reliability; this is a subject for future investigation.

The weight of the on-board switches is the limiting criteria in performance of the baseline system. The resulting "broadcast" link is estimated to be able to maintain its design value carrier-to-noise ratio (12 dB) 95% of the time for the assumed rain attenuation statistics. Such a communication satellite system would not be commercially marketable in the sense of current communication satellites (e.g., video entertainment); however, there may well exist suitable applications such as high volume data transfer where the time of day for the data transfer is not critical. For example, the system being planned by Satellite Business Systems (SBS) is anticipated to accomplish data transfer using a satellite link with a bit error rate of 10^{-6} with 95% reliability.[8]

In order to achieve coverage of the entire continental United States, provisions were made for each of 6 channels to select from among 10 separate ground spot beams. To achieve the proper beam size, the satellite antenna

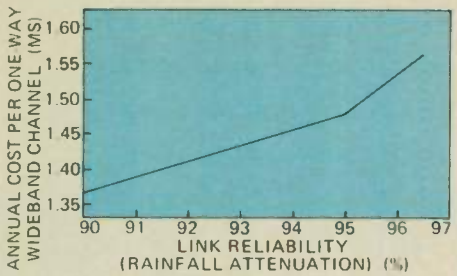

Fig. 7 Annual cost per wideband channel versus link reliability.

diameter was fixed at 0.6 meter rather than used as an optimization variable. For the required coverage, 60 spots with diameter 450 km are required. Once six receive beams and 6 transmit beams are selected, each beam carries 20 subchannels which are switched on the satellite. Any subchannel of a received beam may be transmitted on the corresponding subchannel of any transmitted beam.

Figure 7 gives a plot of the sensitivity of annual cost per wideband channel to changes in required system reliability. Reliabilities higher than 96.5% were not possible under the system constraints without the use of diversity stations. Note that there is approximately a 15% increase in cost per channel as the reliability increases from 90% to 96.5%. Also, the plot is shown as piece-wise-linear due to data points being generated at only 90%, 95%, and 96.5% link reliability.

In order to examine the cost per terminal for various numbers of ground terminals and for various communication capabilities, channel availability was defined as the ratio of the total number of channels to the number of ground terminals. Figure 8 gives annual cost per wideband terminal versus availability for 120, 360, and 1080 ground stations due to absolute launch weight limits.

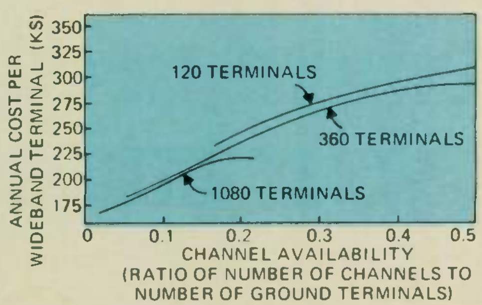

Fig. 8 Annual cost per wideband terminals versus channel availability for 95% reliability.

The increase in cost per terminal is approximately linear with increases in utilization for all numbers of ground stations. The increase is due to the cost of additional switching components and the effects of increased satellite weight on satellite operational systems and launch weight.

For a constant utilization, the cost may be studied for various numbers of terminals. For the increase to 360 from 120 ground stations, the drop in per terminal cost is a result of the further division of satellite cost. For the increase to 1080 ground stations, the decrease is less than would be expected due to substantially increased launch cost for the heavier satellite.

ACKNOWLEDGEMENTS

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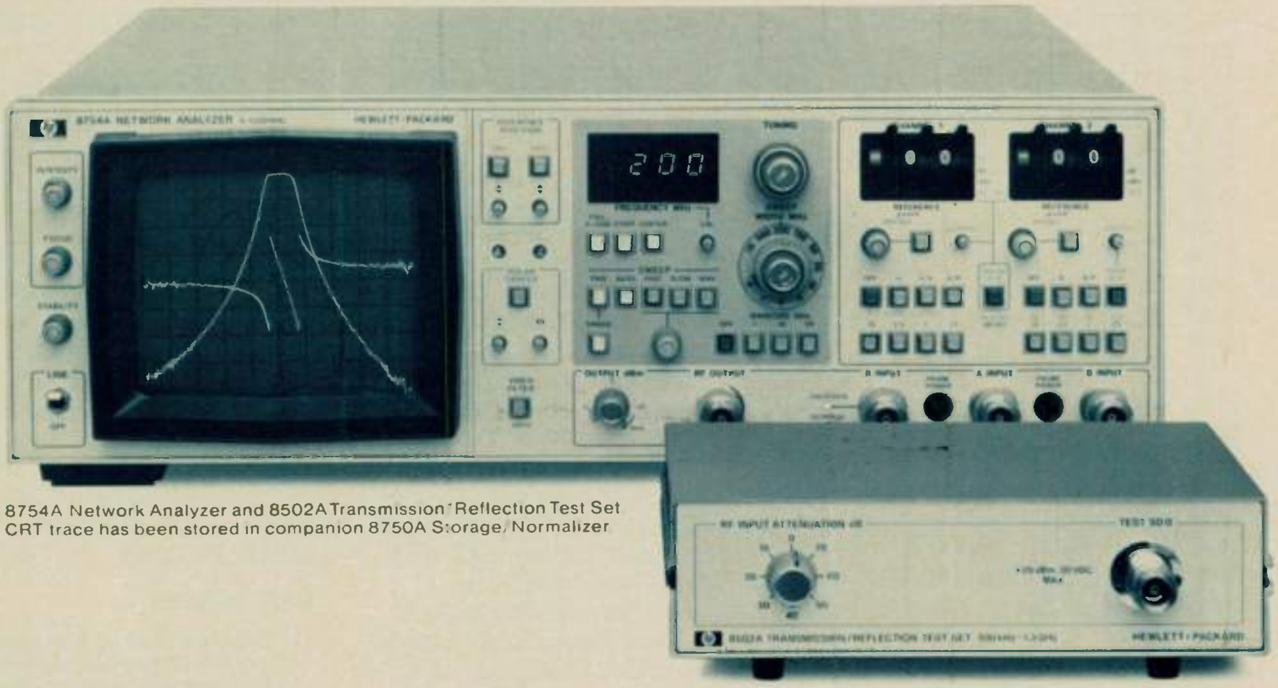

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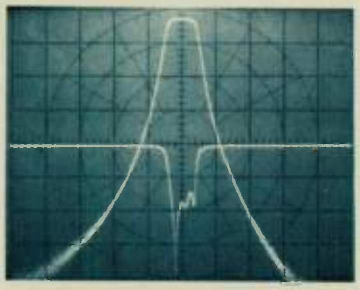

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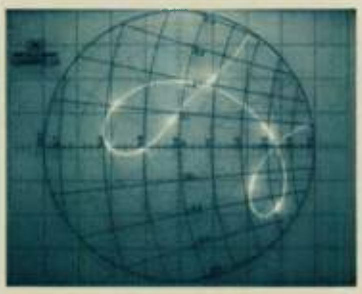

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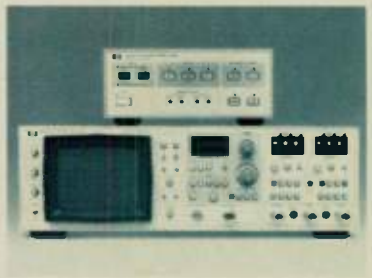

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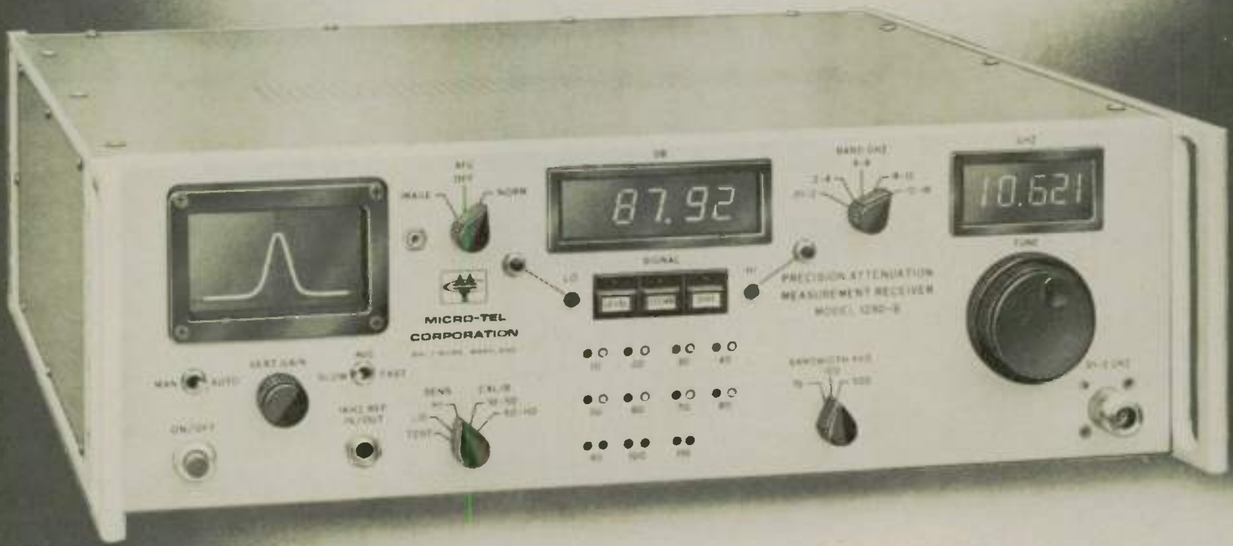

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Moving Toward NMM Wave Integrated Circuits

Interest in the near-millimeter (NMM) portion of the spectrum is presently increasing. More than 175 papers were contributed to last December's Fourth International Conference on Infrared and Millimeter Waves, and a new quarterly, the *International Journal of Infrared and Millimeter Waves*, has been announced.[1] Several emerging applications are encouraging development of technology for frequencies above 100 GHz; moreover, microfabrication technology is making it easier to develop this region of the spectrum. However, it appears that NMM apparatus will not consist of familiar components scaled down to the new size; interesting new engineering problems appear. This is, after all, the region of the spectrum where microwaves and optics meet, and the appropriate technology should be a judicious combination of techniques borrowed from both these fields. Finally, the NMM region represents something of a last frontier in the crowded radio spectrum. New services must now look to millimeter waves for new chunks of bandwidth. When the NMM spectrum has all been claimed, newcomers will have to make the long jump to optical frequencies, as the intervening far-IR is closed by heavy atmospheric absorption.

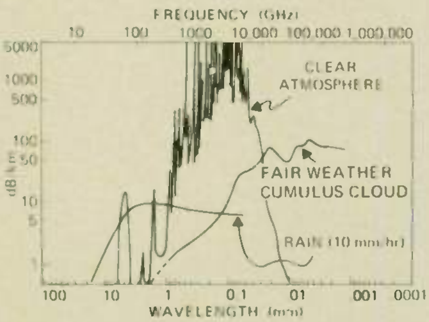

Fig. 1 Attenuation of the clear atmosphere. Also shown are additional attenuations due to fair weather cumulus cloud and to rain (10 mm per hour).

Undoubtedly, one reason for the past neglect of NMM waves has to do with the fact that atmospheric absorption has its onset here, as shown in **Figure 1**.[2] Maximum atmospheric penetration is obtained in windows located between the molecular absorption lines, especially near 136 and 230 GHz. In these windows useful propagation through a clear atmosphere can be obtained over several kilometers. Penetration of fog is better than with optical radiation, because fog particles tend to be much smaller than the NMM wavelength; however raindrops, which are larger, cause considerable loss. Another factor which has slowed NMM development is that dielectrics are relatively lossy. At $\lambda = 1$ mm some of the more transparent materials are[3] polyethylene (absorption coefficient 0.2 dB/cm), polypropylene (0.2 dB/cm), TPX (.26 dB/cm), optosil fused quartz (0.4 dB/cm), polystyrene (1.3 dB/cm), and silicon (2.2 dB/cm, provided that the resistivity is more than 10 Ω-cm) Fortunately several of the materials on this list, such as silicon and fused quartz, are quite convenient for fabrication of the small structures that are necessary, and the fact that semiconductors are reasonably transparent makes it possible to integrate dielectric waveguides with semiconductor devices.

Notwithstanding the atmospheric absorption, there is increasing interest in NMM waves for short-haul communications. Unoccupied NMM bands are available, and at these wavelengths high-gain antennas are conveniently small. There is also a great deal of new astronomical work to be done; here balloon-borne and satellite observing platforms as well as ground-based telescopes are used. There is a need for NMM-wave instrumentation for plasma diagnostics: detectors are needed for observation of synchrotron radiation, and plasma interferometers need to operate at frequencies near plasma frequencies of the dense fusion plasmas now being studied. Spectroscopic applications exist for pollution monitoring, atmospheric studies, and also general research.

All these applications require circuits analogous to those of a lower-frequency microwave receiver. Metallic waveguides, however, are quite lossy at NMM wavelengths, and the small dimensions of such guides make them expensive and difficult to use. It thus appears that low-loss dielectric waveguides will be the best choice in this waveguide region. Since free-space propagation is a part of most applications, antennas compatible with the dielectric waveguides must be designed. At the other end of the waveguide a transition structure, necessarily metallic, must be provided to concentrate the radiation across the depletion region of a suitable mixer diode. Local oscillator power must be supplied, combined with the signal, and applied to the mixer. The required components are likely

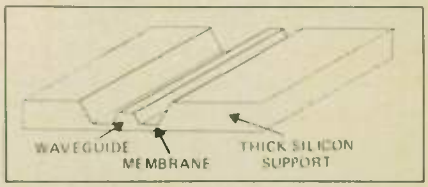

Fig. 2 Typical dielectric-rod waveguide supported on a dielectric membrane.

to be very small; for example, at 1 mm wavelength a typical single-mode silicon waveguide will have cross sectional dimensions around 200 micrometers. Such small circuits are most easily made using microfabrication techniques similar to those in IC technology. With this kind of fabrication it is natural to construct entire circuits as a single unit; they will then have the usual IC advantages of ruggedness and low cost. In some cases it may be desirable to construct arrays of identical structures, for example for imaging; with microfabrication this can readily be done. Eventually it may also be possible to integrate low-frequency components such as IF amplifiers as well. Aside from the problem of the local oscillator, which still presents difficulties, it does not seem far-fetched to envision small, inexpensive, integrated NMM receivers for, say, office-to-office communications use.

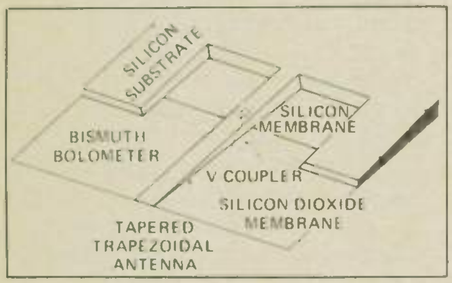

Fig. 3 Diagram of tapered-dielectric waveguide antenna.

Although production of integrated circuits can be rapid and inexpensive once the masks are made, the design of NMM devices can be tricky. Because of the rather odd shapes that arise in practice, mathematical analyses of new devices can only be approximate. On the other hand, one may well be reluctant to hazard the considerable effort of mask-making for a new IC when the design is unproven. Moreover, it is difficult to measure such things as impedance at NMM frequencies. For these reasons, microwave simulation of new designs is a very useful technique. We have tested various device designs by means of scale models enlarged for use at 4 - 8 GHz, with the tests being carried out in an anechoic chamber. One must, of course, use dielectric materials with refractive index equal to that expected in the actual device at the NMM wavelength. Usually dielectric losses and departures of metals from ideal conduction do not have much effect and can be neglected. Devices for wavelengths as short as .01 mm have been simulated, with good agreement always being obtained.

DIELECTRIC WAVEGUIDES

A promising approach to the basic NMM waveguide is shown in Figure 2. The dielectric waveguide is similar in principle to the glass light-guiding fibers now being introduced for telephone communications, but in an IC the guide must be supported in a different way. A basic principle is that no waveguide can be placed on a substrate that supports surface or bulk waves slower than the wave being guided.[4] If this condition is not fulfilled the energy will quickly escape from the guide, dispersing into shock waves in the substrate. For this reason the waveguide cannot simply be placed on a thick slab

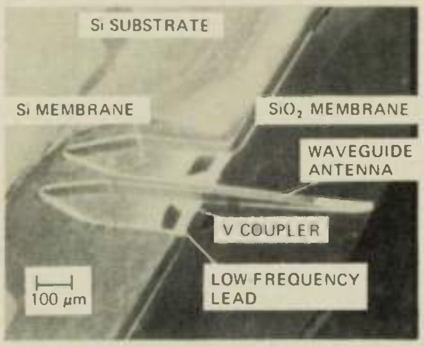

Fig. 4 Photograph of a 119-micron version of the antenna sketched in Figure 3.

of the same material, because the guided wave travels faster than the bulk wave in solid dielectric. However, the waveguide can be placed on a thin membrane, as the surface waves of the membrane are comparatively fast. Figure 2 illustrates a typical possibility: a silicon waveguide etched out of a silicon wafer and supported by a thin silicon membrane. The membrane is in turn supported by thicker silicon, sufficiently distant so as to not interfere with propagation in the guide.

After fabrication the structure can be placed on a slab of quartz for additional physical support. Because quartz has a lower refractive index than silicon, the support layer will not affect propagation much.

NMM ANTENNAS

In order to terminate the dielectric waveguide in an antenna, it is only necessary to taper it. An example of such an antenna[5] is shown in Figure 3. In this case the dielectric waveguide is simply constructed in a wedge-shaped piece of silicon to provide the desired taper. Figure 4 is a photograph of such an antenna for use at a wavelength of 0.119 mm. This antenna has a well-confined, single-lobed pattern, as shown in Figure 5, making it suitable for use as a feed horn for a primary mirror. The 3 dB beamwidth is about 35 degrees and the gain is about 13 dB.

Other kinds of NMM antennas have also been built. Metal antennas cannot be deposited on the surface of a thick dielectric, because the guided wave will be faster than bulk and surface waves of the dielectric; thus energy will quickly be lost into shock waves. However, it may be possible to build metal antennas on sufficiently thin dielectric membranes; conventional microstrip antennas are of this type. Another approach is to encase a metal antenna entirely in dielectric, in which case it operates just as it would in air, except for change of scale. An example of such a device, intended for use at 0.119 mm, is shown in Figure 6. In this case the metal V antenna is evaporated on a quartz substrate, and an identical quartz

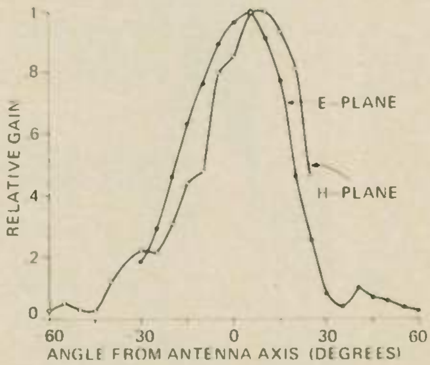

Fig. 5 E- and H-plane patterns of the antenna of Figure 4.

(continued on page 50)

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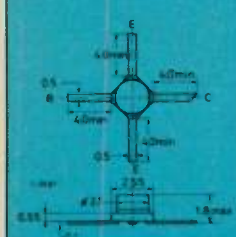

TYPE	Polarity	f_T (GHz)	GAIN S_{21e} (dB)	GAIN Gmax (dB)	GAIN V_{CE} (V)	GAIN I_C (mA)	GAIN f (GHz)	NOISE NF (dB)	NOISE Ga (dB)	NOISE V_{CE} (V)	NOISE I_C (mA)	NOISE f (GHz)
BFQ86	NPN	4	13	15	6	10	0.5	1	—	6	3	0.14
BFQ88	NPN	5	12	15	10	15	1	2.5	12.5	10	5	1
BFQ88A	NPN	5	11	14	5	30	1	2.5	10	5	5	1
BFQ88B	NPN	4.5	9.5	11	8	40	1	4.5	9	8	40	1
BFQ89	NPN	6	9.5	14.5	10	10	2	2	11	10	6	2
BFQ89A	NPN	6	9	14	10	15	2	5	11	10	15	2
BFQ98	PNP	5	12	15	10	15	1	2.5	12.5	10	3	1
BFQ98B	PNP	4.5	9.5	11	8	40	1	4.5	9	8	40	1

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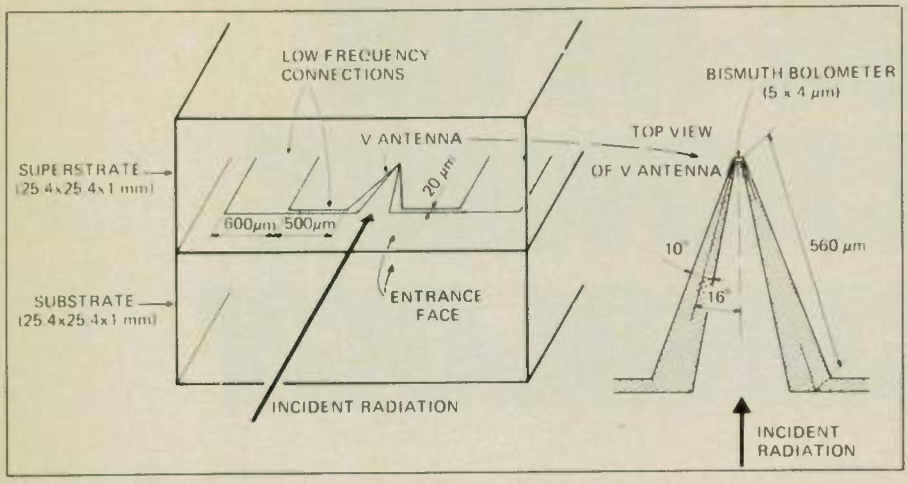

Fig. 6 V antenna and microbolometer detector for 119 microns, encased in crystal quartz "sandwich" structure.

superstrate is placed over it to give a "sandwich" structure.[6] This kind of antenna is more suited for use with two-conductor metallic transmission lines than with dielectric waveguides, but it has found an application in connection with far-infrared detection. A very small bismuth film bolometer can be deposited directly at the antenna terminals, as shown in the figure. Because the bolometer is quite small (1/25 - 1/200 wavelength in size), very little captured energy is required to heat it and a video NEP of around 10^{-9} watt/$Hz^{1/2}$ with 1 MHz bandwidth can be obtained at room temperature. These antenna-coupled bolometers may replace pyroelectric detectors for some far-infrared video detection applications (such as plasma diagnostics) because of their greater speed.

V COUPLER

When dielectric waveguides are being used, there must be a way to extract the energy from the guide and apply it to a mixer diode. In conventional microwave practice one might make a transition to hollow metal waveguide, and then mount the diode inside the metal waveguide on a post. With the microfabricated dielectric guide, however, it is more convenient to deposit a metallic antenna, which we call a V coupler, directly on the dielectric structure. The V coupler (which can be seen in **Figures 3** and **4**) resembles the V antenna used in the sandwich structure. In this case no superstrate is used, and advantage is taken of the tendency of the metal structure to couple to modes of the dielectric. Because in this case the dielectric is a single-mode waveguide, good coupling is obtained between the guide and a load connected at the terminals of the V. Coupling losses as low as 0.7 dB have been measured.

SCHOTTKY DIODE MIXERS

With a waveguide made of semiconductor it should be feasible to construct a diode mixer directly in the waveguide material, at the V coupler's terminals. Schottky diodes have often been constructed for use at 300 GHz and beyond, but in most cases they have been three-dimensional structures. Typically the diode will consist of a very small (about 1 micrometer) metallic dot on the semiconductor surface, to which contact is made by a kind of cat-whisker, which also serves as an antenna to couple energy into the diode. Diodes of this kind have given very good performance, although at 300 GHz and above, mixer noise temperatures still tend to be orders of magnitude above the "quantum limit" hv/k. It should be beneficial to eliminate the cat-whisker, because it is difficult to contact and reduces the ruggedness of the structure, but planar diode structures present their own difficulties. Small dimensions and carefully chosen doping profiles are required to assure sufficiently small values of their RC time constants. Further constraints are raised by the need for integration with a planar coupler or antenna. For example, one might wish to deposit a metallic V-coupler on epitaxial material, as the latter can help to reduce RC. But this cannot be done, inasmuch as the heavily-doped layer of the semiconductor will act as a metallic ground plane, short-circuiting the coupler. **Figure 7** shows one form of an integrated Schottky diode/antenna structure for $\lambda = 1$ mm. In this figure the rectangular areas at the terminals of the V coupler are photolithographic "windows" in which different levels of doping are produced by ion implantation, resulting in one ohmic and one rectifying contact. Very careful mask alignment is required to keep the diode's dimensions down to tolerable levels. Performance data are presently being obtained for this structure.

Fig. 7 Integrated Schottky diode/antenna structure for $\lambda = 1$ mm. Photo courtesy of G.-G. lau.

INTEGRATED RECEIVERS

It should be possible to combine antennas, waveguides, and diodes into integrated circuits, like the simple receiver sketched in **Figure 8**. Here the signal and LO channels are combined by a 3 dB hybrid coupler made by bringing two dielectric waveguides near each other so their fringing fields couple. A balanced mixer is used to avoid loss of power and to cancel local oscillator noise. Various other integra-

(continued on page 52)

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MC 5112	1000 - 12400	25.0 dB	±0.50 dB	+28V, 15mA
MC 5118	1000 - 18000	25.0 dB	±0.50 dB	+28V, 15mA
MC 50018	5 - 18000	25.5 dB	±0.75 dB	+28V, 15mA
STANDARD BAND COAXIAL				
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MC 5024	2000 - 4000	30.0 dB	±0.50 dB	+28V, 15mA
MC 5048	4000 - 8000	30.0 dB	±0.50 dB	+28V, 15mA
MC 5812	8000 - 12400	30.0 dB	±0.50 dB	+28V, 15mA
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MC 5812W	8200 - 12400	15.5 dB	±0.50 dB	+28V, 15mA
MC 51218W	12400 - 18000	15.0 dB	±0.50 dB	+28V, 15mA
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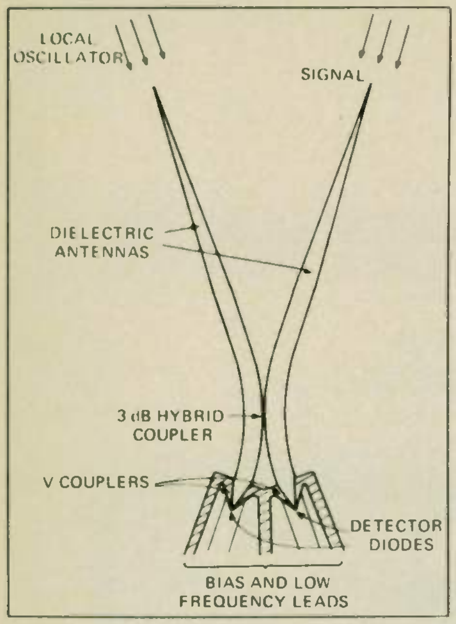

Fig. 8 Proposed integrated NMM receiver including two antennas, 3 dB hybrid coupler, and balanced mixer. An integrable LO has not yet been developed.

ble components based on the dielectric waveguide can readily be imagined. For instance it should be possible to use a closed circle of waveguide as a ring resonator, coupling to it by means of the fringing fields as in the case of the hybrid coupler. Development of such a resonator may eventually help to solve the vexatious problem of local oscillators for this frequency range. The convenience and cheapness advantages of the IC are largely dissipated if a bulky, expensive, external local oscillator must be used. One promising approach here is the use of harmonic mixers, which have recently provided noise temperatures as low as 2500 - 3000 K for reception at 200 - 230 GHz.[7] In addition to the noise-suppression advantages of the harmonic mixer, the reduced LO frequency should make it easier to produce an integrated LO. Some semiconductor negative-resistance devices, such as the IMPATT diode, are usable at frequencies as high as 300 GHz. One problem with semiconductor NMM oscillators is likely to be low power, because of the small dimensions such devices will probably be required to have. With IC technology it may be possible to get around the problem by replicating a large number of structures, provided they can be made to oscillate in phase. For example, a number of negative-resistance devices might be fabricated around the circumference of a single waveguide ring resonator. This however is only speculation; the LO problem is still a long way from solution.

DETECTOR ARRAYS

Microfabrication technology allows us to construct arrays containing numerous identical elements, such as the detector array shown in **Figure 9**. This detector consists of a planar metallic circuit, interspersed with small bismuth-film bolometer elements, constructed on the surface of a quartz cover slip. Radiation to be detected impinges normally on the detector. If we assume the metal circuit to be lossless and if no radiation is reflected, then all the incident energy will be used to heat the bismuth, as we desire. In order to eliminate reflection the detector must effectively terminate free space in its characteristic impedance. As far as the resistive component is concerned, this can be done by shaping the metal circuit into a transformer of appropriate dimensions, but there will still remain an unmatched reactive component arising from capacitance and inductance in the metal circuit and dielectric reflection from the thin substrate. Interestingly, this reactive mismatch and its associated reflection can be largely eliminated through the use of an adjustable metal shorting plane placed behind the detector plane and parallel to it. This refinement is similar to the shorting plunger used in a conventional waveguide mixer mount. In the present case, however, the adjustability is an interesting feature, since there are not many convenient ways to tune a NMM integrated circuit. A detector array of this kind, with 400 identical detector elements, has been tested at a wavelength of 1.3 mm, with the adjustable backshort working to reduce reflections as expected. With optimal adjustment, more than 50% of the incident radiation was coupled into the bolometers.

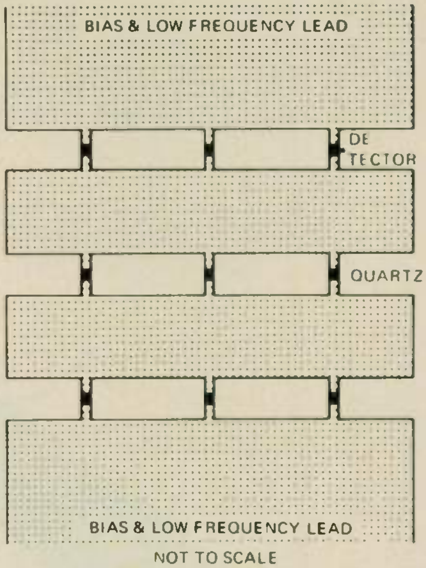

Fig. 9 Diagram of planar detector array for $\lambda = 1.3$ mm. (Not to scale.)

CONCLUSION

On the basis of what has already been done, it seems safe to say that integrated receivers for wavelengths near one millimeter can be made—with the exclusion of the local oscillator. The rate of progress in this field will depend mainly on the rate at which the motivating applications expand. Development of solid-state sources for this region, particularly integrable ones, will present interesting scientific problems. Spinoffs of the results into such related areas as far-infrared detection and imaging can be expected.

ACKNOWLEDGEMENTS

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The Microstrip Diplexer
A New Tool for Millimeter Waves

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INTRODUCTION

In 1976 we published an article in this journal[1] describing our efforts to date with millimeter-wave MIC's. We had begun to explore the many possibilities of using low dielectric substrates at the higher frequencies. Among the circuits which showed considerable promise was the edge coupled bandpass filter. Two of these filters, designed for different passband frequencies, were incorporated into microstrip diplexer form. The 1976 results, although encouraging at the time, fell far short of performance levels subsequently attained.[2] Examples, before and after, of contiguous diplexers are shown in **Figures 1** and 2. For these and all other measurement data, the figures include transition losses.

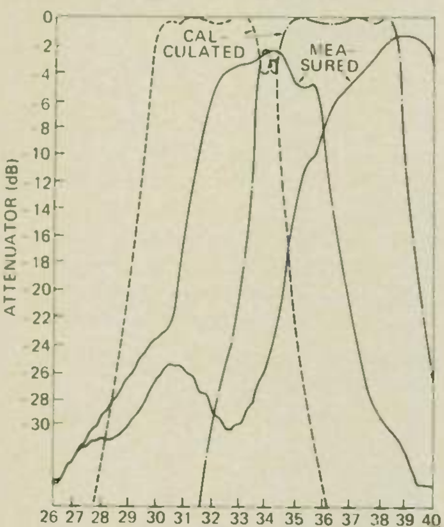

Fig. 1 Original attempt at diplexer (1976).

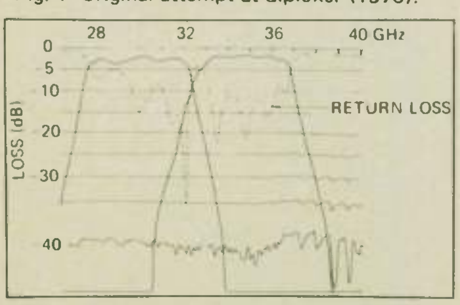

Fig. 2 Diplexer in below-cutoff waveguide using printed circuit transitions (1978).

EXPLAINING THE DIPLEXER

Using Z_0 as the output and input characteristic impedances, two different bandpass filters are connected to a common junction, preceded by different length lines, also of impedance Z_0. At the input port of a lossless filter, the out-of-band admittance is purely susceptive. The added line lengths between the junction and each input port are calculated to make the admittance of each bandpass filter appear as an open circuit at the center of the opposing filter's bandwidth. This transformation is easiest to understand with the aid of a Smith chart (Figure 3): Here Y1, the input admittance of Filter 1 (32-36 GHz bandpass), is shown over 28-32 GHz, bandpass range of Filter 2. As can be seen, Y1 is highly susceptive. If no transmission line were used between Filter 1 and the junction, the 28-32 GHz filter would be severely loaded. Adding a Z_0 impedance transmission line to the input of Filter 1 rotates the latter's susceptance around the Smith Chart so that it appears as Y2. Here the transmission line was of sufficient length to make Filter 1 appear as an open circuit at the center frequency (30 GHz) of Filter 2.

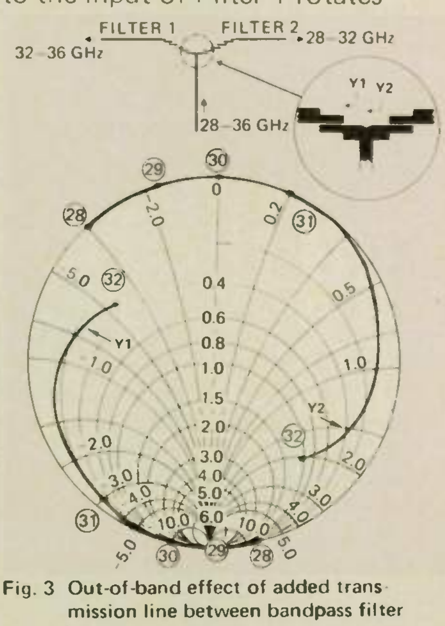

Fig. 3 Out-of-band effect of added transmission line between bandpass filter and diplexer junction.

Similarly, a transmission line is placed at the input of Filter 2 of sufficient length to make it appear as an open circuit at the center frequency (34 GHz) of Filter 1. Slightly different length lines could have been used to minimize the out-of-band susceptances somewhat more than shown. Except near the crossover point of contiguous filters, input RF within the diplexer range is directed to the proper bandpass filter almost as if the other filter were not present.

FABRICATION

Open-ended microstrip is well known for its radiative properties. This affects open resonators[3] by causing considerable rounding off of the attenuation at the low frequency end of the bandpass response. Radiation effects can be greatly reduced by enclosing the filter within a below-cutoff waveguide housing.

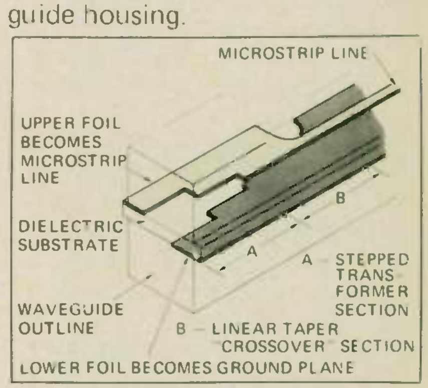

Fig. 4 Printed circuit to waveguide transition.

Since the addition of dielectric will increase the electrical dimensions of a waveguide, the frequency at which higher order modes can exist will be lowered. For the 0.01" thick Duroid used

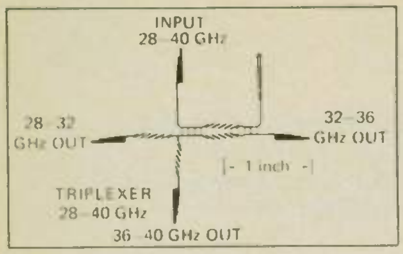

Fig. 5 28-40 GHz Triplexer.

Fig. 6 Total triplexer losses.

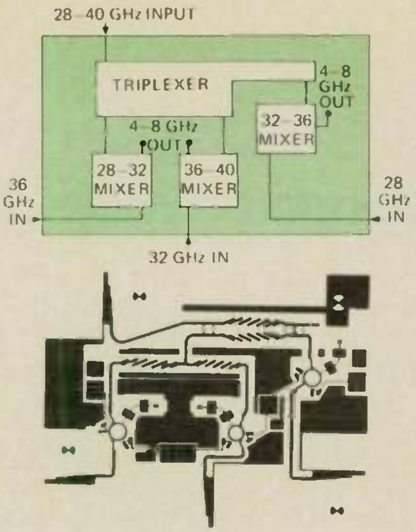

Fig. 7 Three-channel integrated downconverter – transparency used for MIC.

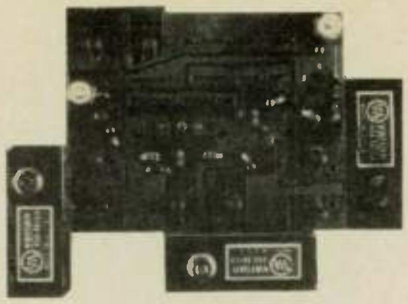

Fig. 8a Three-channel integrated downconverter, covers off.

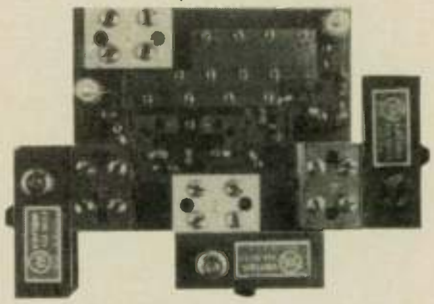

Fig. 8b Three-channel integrated downconverter, covers on.

up to 40 GHz and the 0.005" thickness above 40 GHz, a 10% reduction in the waveguide dimension parallel to the microstrip has been found to be sufficient. Figure 4 shows a diagram of the printed circuit to waveguide transition used for almost all of our final circuits.

The design of each bandpass filter proceeds as follows:

(a) Calculate the number of sections and even and odd mode impedances (Z_{oe} and Z_{oo}) required for the desired ripple and bandwidth.

(b) Use Bryant and Weiss MSTRIP[4] or other program to find the line widths and gaps needed to obtain these impedances. Also note the effective even and odd mode dielectric constants ϵ_{oe} and ϵ_{oo}. Use Dell-Imagine formula[5]

(c) $$L = \frac{\lambda_0}{4} \cdot \frac{Z_{oe} + Z_{oo}}{Z_{oe}\sqrt{\epsilon_{oe}} + Z_{oo}\sqrt{\epsilon_{oo}}}$$
to find the lengths needed to compensate for the differences in phase velocities of the two modes.

(d) Use Silvester and Benedek[6] or other end capacity compilation to determine the end capacitance at the ends of the half-wave resonators.

(e) Using a good graphic analysis routine, plot attenuation vs. frequency.

(f) Noting the percent shift in the center frequency (mainly caused by the end capacitance), decrease each line length by the same percent.

(g) Compute and graph again. One iteration is usually sufficient for results.

A FEW USES FOR THE DIPLEXER

Triplexer

Figure 5 shows the diplexer used as part of a microstrip triplexer. A 12 GHz input bandwidth is divided into three 4 GHz output channels. The output ports of the first hybrid coupler contain identical midband bandpass filters. Frequencies within their bandpass are transmitted to the next hybrid and recombined at its output. All other frequencies are reflected from the bandpass filters and combined at the isolated port of the first hybrid. A noncontiguous diplexer is then used to separate the high and low frequency bands. Figure 6 shows the result. The triplexer was integrated with three modified rat race mixers[7] to form an integrated three-channel downconverter with 4-8 GHz IF outputs. Figure 7 shows the transparency layout used for the integration, Figure 8 the unit itself. For this circuit, only the transition and triplexer were covered by the reduced height housing. Total input to output conversion loss for the three channels is shown in Figure 9. The triplexer losses were obtained by removing the mixers from the rubylith layout and extending the triplexer output lines to the LO input transitions. This allowed direct measurement, except for the addition of a second transition and some lossy microstrip transmission line. Conversion losses taken near the crossovers are probably in error due to inaccuracies caused in setting LO frequencies.

Diplexer-Mixer

Often, balanced mixers are used with high IF frequencies simply because it is inefficient to couple LO power through a 10 dB (or so) coupler to a mixer diode. The coupler provides isolation between the RF and LO ports. Figure 10 shows how full LO power can be applied to the diode through one arm of a diplexer with excellent isolation and without loading the RF port. In this case, the diode must be resonated and transformed to appear, at the junction, similar to Z_0 over the LO and RF ranges.

At the higher millimeter wave frequencies, high power local oscillators and mixer diodes can both be costly. A well designed diplexer-mixer would require half the LO power and half the diodes of a balanced mixer. The latter, necessary for reducing LO noise close to the carrier, is really not necessary for high IF frequencies. For 4 GHz bandwidths, and a LO to upper RF frequency spread of 8 GHz, our lowest conversion loss has been 11 dB. Not much time was spent with this mixer;

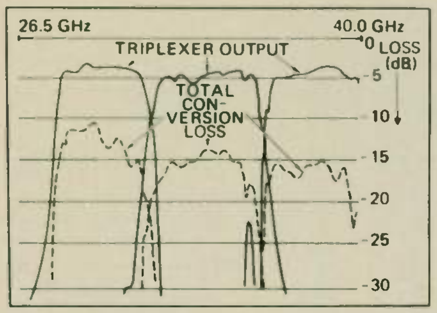

Fig. 9 Total conversion losses for integrated downconverter.

careful matching to the diode should work to lower these losses considerably.

Frequency Stabilizer

A noncontiguous diplexer can be used as a frequency discriminator by sensing its crossover frequency.[8] The crossover, which is designed to occur about 8-10 dB below the bandpass attenuation, is very sharp, i.e., many dB/GHz, presenting a much higher effective Q at this frequency than other planar resonators.[9] The detect-

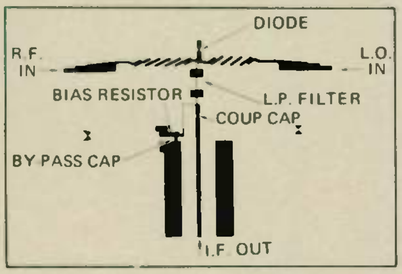

Fig. 10 Possible configuration for mixer-diplexer.

ed output voltages (Figure 11) must be differentially amplified with considerable gain and fed back to the VCO. If the output voltage difference is very slight for a small change in oscillator frequency, the dc feedback voltage will be noisy, resulting in oscillator phase noise. Our MIC Gunn oscillator used simple microstrip transmission lines (and the varactor) as a resonator. Without feedback the resonator, and the diode parameters, determined the free running frequency. When the feedback loop was closed, the output frequency was precisely that of the crossover. Phase noise was observed by mixing with a crystal comb generator, converting to 40 MHz, and applying to a low frequency discriminator. The phase noise was found to be lower by one or two orders of magnitude when compared to free running.

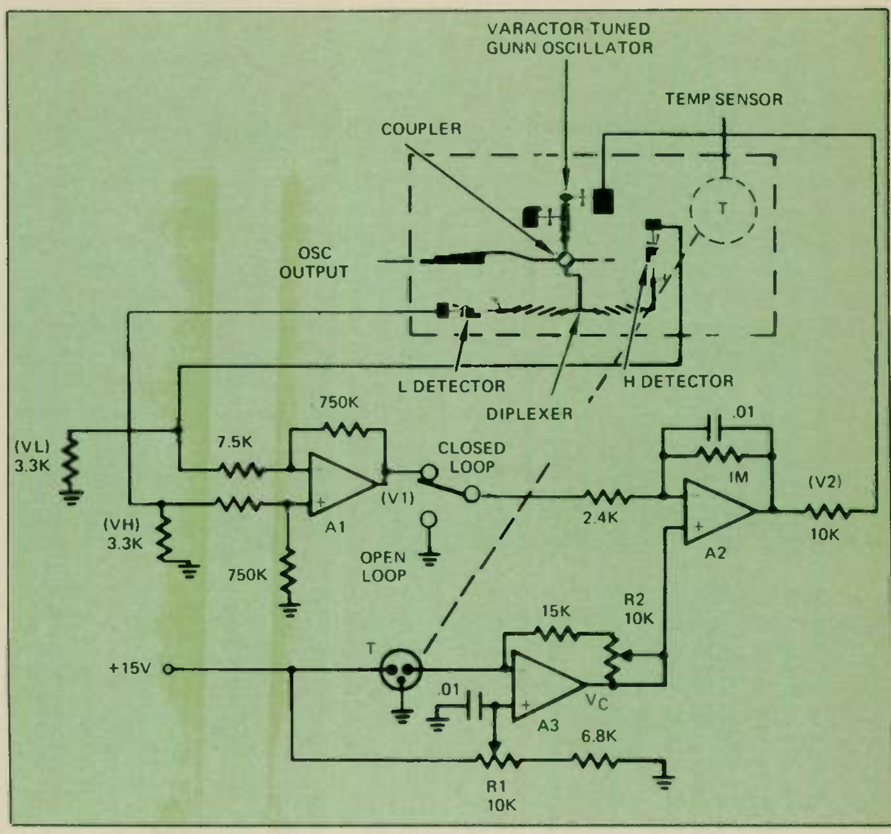

Fig. 11 Frequency stabilized microstrip oscillator.

Unfortunately the effective dielectric constant decreases with temperature, causing the crossover frequency to increase. Temperature compensation must therefore be used. This could be done simply by locating the diplexer away from the oscillator in a constant temperature enclosure. We chose, however, to use an IC temperature sensor* mounted in proximity to the diplexer. The sensor produces a current output proportional to the absolute temperature. This is used to form a voltage ramp (as a function of temperature) which is fed back as a correction bias to the VCO. The discriminator voltage modulates this bias. The integrated VCO has shown a temperature sensitivity of 0.35 MHz/°C, about three times better than many commercial waveguide VCOs.

WHAT'S NEXT?

Certainly three diplexers can be used to form a quadriplexer. One diplexer can be used to separate two frequency bands. Each output would go to another smaller bandwidth diplexer to divide its frequency band.

* Analog Devices Model AD590.

There are undoubtedly many other uses for these devices. Any ideas?

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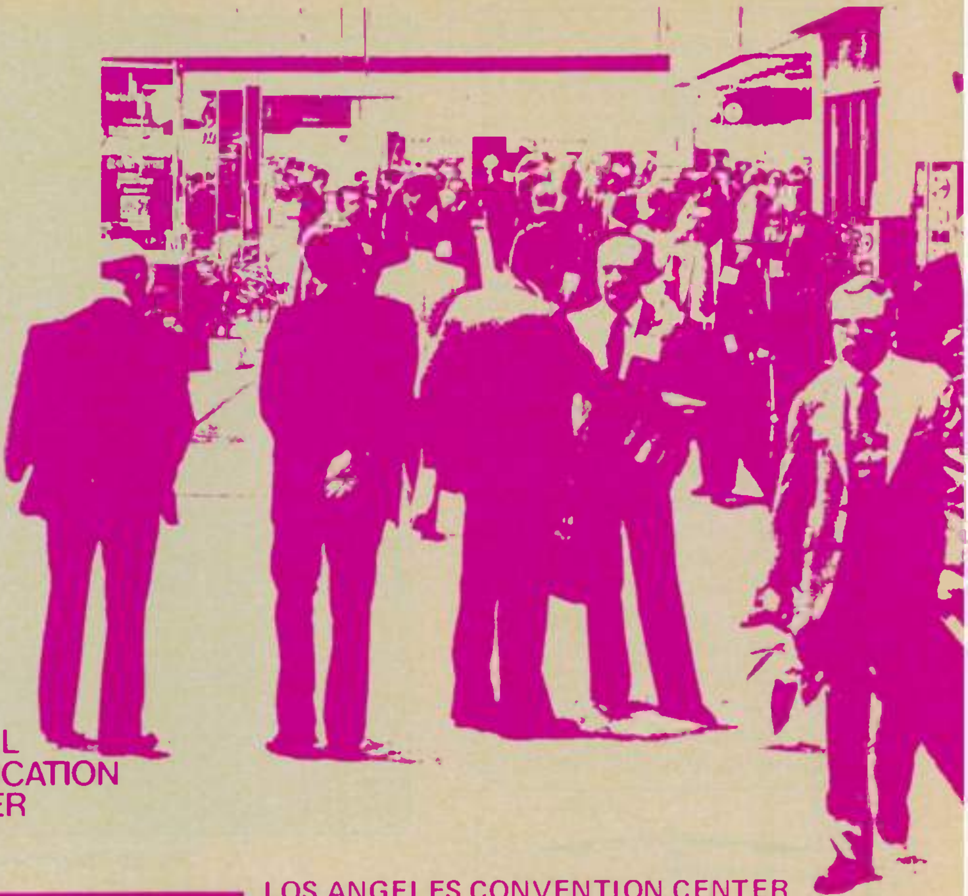

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REFLECTOMETERS
for Millimeter-Wave Measurements

MARK CRANDELL
PAUL A. CRANDELL
Hughes Aircraft Company
Electron Dynamics Division
Torrance, CA

There are various techniques available to measure impedance at millimeter-wave frequencies. Slotted lines and hybrid impedance bridges are available, but inaccuracies associated with residual SWR, probe interference, and coupling factors limit the usefulness of these devices. In order to overcome these limitations, the development of two basic types of network analyzer systems have been attempted for measuring impedance. The first method uses a parallel pair of reflectometers with downconverters, which gives a real-time plot of impedance vs. frequency when used with a low frequency network analyzer. This method works well at frequencies up to 40 GHz, but requires a synthesized, phase-locked sweeper at millimeter-wave ranges. The second and more promising method is the six-port network analyzer. However, this technique is still under development at the present time.

In the absence of a reliable impedance measuring instrument at millimeter-wave frequencies, a simple but reasonably accurate reflectometer system has been shown to offer an attractive alternative. In this article we present a broadband, practical, millimeter-wave reflectometer and discuss its performance characteristics, measurement range, and practical accuracy limits.

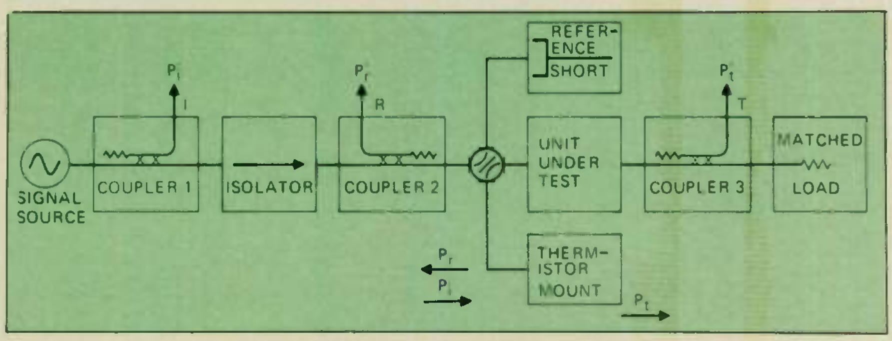

Fig. 1 Ideal reflectometer system.

A simplified block diagram of an ideal reflectometer system is shown in Figure 1. Power from the signal source is applied to the unit under test and incident, reflected, and transmitted power levels are measured at ports I, R, and T, respectively. From this information the insertion loss (IL), the return loss (RL), the voltage reflection coefficient (Γ), and the SWR of the unit can be determined from the following relationships:

$$IL = -10 \log (P_t/P_i) \text{ in dB} \tag{1}$$

$$RL = -10 \log (P_r/P_i) \text{ in dB} \tag{2}$$

$$|\Gamma| = (P_r/P_i)^{1/2} \tag{3}$$

$$SWR = \frac{1 + |\Gamma|}{1 - |\Gamma|} \tag{4}$$

where P_i, P_r, and P_t are the incident, reflected, and transmitted powers. It should be noted that P_i, P_r, and P_t are measured indirectly by measuring P_i', P_r', and P_t'. It is thus necessary to determine the system calibration factors.

To calibrate the system, the power calibration factors $K_i = P_i/P_i'$, $K_r = P_r/P_r'$ and $K_t = P_t/P_t'$ are first to be determined. This procedure can be done in a relatively straightforward manner.

To make a measurement, the unit under test is placed in the system and P_i', P_r', and P_t' are measured. By using Equations (1) through (4) with the following substitutions:

$$(P_t/P_i) = (K_t/K_i)(P_t'/P_i') \tag{5}$$

and

$$(P_r/P_i) = (K_r/K_i)(P_r'/P_i') \tag{6}$$

the unit parameters can be found.

In a practical reflectometer several potential sources of error must be considered in calibration

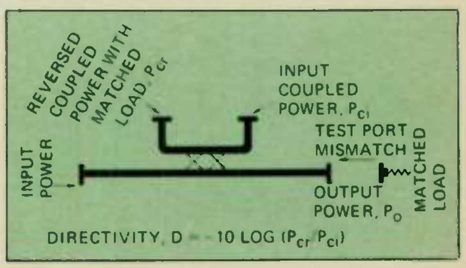

Fig. 2 Directional coupler definitions.

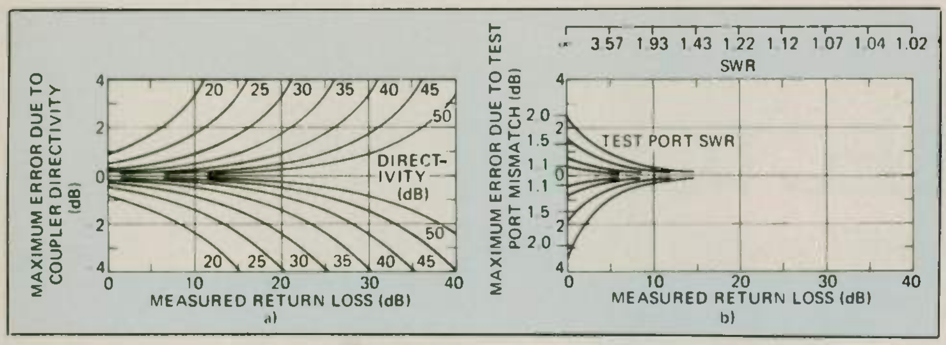

Fig. 3 Reflectometer error limits.

and measurement. One of the major sources of error in measurements of high return loss is caused by the directional coupler directivity. Referring to Figure 2, the directivity, D, of a terminated coupler, is defined as $D = -10 \log (P_{cr}/P_{ci})$.

When a unit under test is placed at the test port, the voltage reflection coefficient associated with it combines with the voltage reflection due to the directivity of coupler 2 ($\Gamma_D = 10^{-D/20}$). Depending on the relative phases of the two reflection coefficients, the resultant measured return loss will be:

$$-20 \log (|\Gamma| - |\Gamma_D|) \leq RL \leq -20 \log (|\Gamma| + |\Gamma_D|) \quad (7)$$

From this relationship, the error limits can be estimated as shown in Figure 3a where the error limits are plotted as a function of coupler directivity and measured

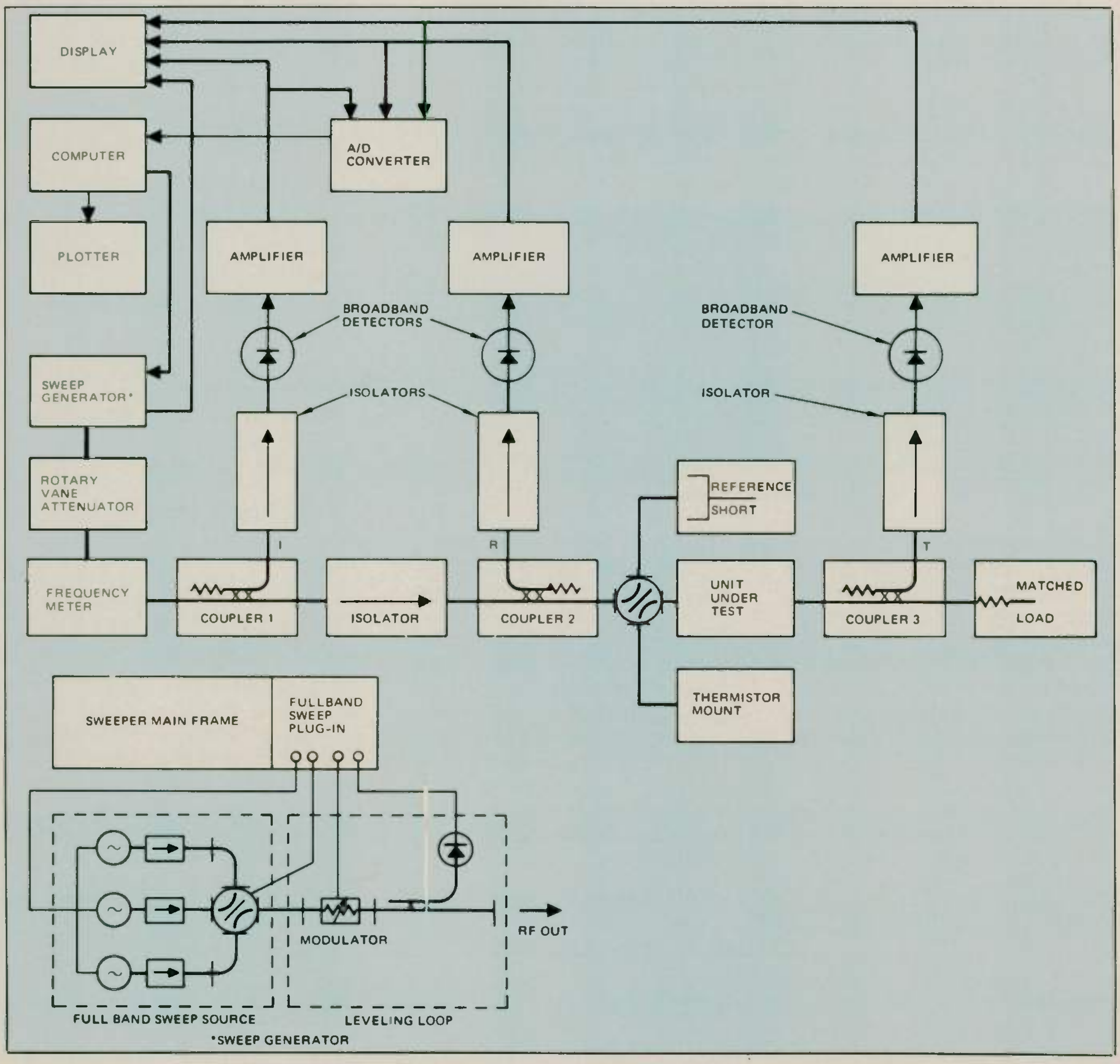

Fig. 4 Automated millimeter-wave reflectometer system.

(continued on page 63)

return loss. For example, a return loss of 22 dB as measured with a coupler having 40 dB directivity, which is typical for millimeter-wave directional couplers, gives an error limit of +1.2 dB and -1.0 dB. From these values we find that the input SWR of the unit lies between 1.15 and 1.20, i.e., $0.069 \leqslant |\Gamma| \leqslant 0.089$.

Another source of error, as shown in **Figure 2**, is the test port mismatch. Some of the power reflected from the unit under test is reflected from the test port due to its mismatch and then redirected to the unit under test. This multireflection arrives at the measurement port down from the primary reflection and introduces a measurement error depending on the relative phases of the two signals. **Figure 3b** is a plot of measurement error versus measured return loss as a function of test port mismatch. For example, if we measure a return loss of 5 dB for a particular unit and our test port has a SWR of 1.5:1, the error limits from **Figure 3a** are +0.55 dB and -0.60 dB.

When measuring low values of return loss (large SWR), there is a potential error due to any mismatch looking back into the signal source. This error is minimized by placing an isolator between couplers 1 and 2.

Both thermistors and detectors can be used for power measurements in reflectometers. An advantage of a thermistor is that it measures power directly; however it does not have sufficient dynamic range in this system. An advantage of a detector is that it is fast enough to work in swept measurements and has a large dynamic range; but its disadvantage is that it needs to be calibrated.

The reflectometer described above can be automated with a computer or desk-top calculator. The automated system is a fast, accurate method of measuring component parameters with a minimum of test error. By using the computer to determine the calibration factors K_i, K_t, and K_r the frequency dependence of test line insertion loss, coupling coefficients, and detector response can be factored out.

Figure 4 shows a block diagram of a computer-controlled millimeter-wave reflectometer system. An IMPATT sweeper source, controlled by the sweeper main frame, plug-in, and computer, is swept through the desired frequency range. The output power is directed through the test line as shown into the termination. The modulator is used to square-wave modulate the output signal at 1 kHz. This allows amplification of the detector output at 1 kHz with a narrow bandwidth, thus increasing sensitivity and dynamic range. The rotary vane attenuator is used to set the reference power level of the system and to ensure that the attenuation is constant with frequency. The frequency meter is used to measure specific frequencies and as an additional marker on the display. The isolator is inserted between couplers 1 and 2 to reduce the effects of unwanted reflections, particularly in measuring high SWR values of the unit under test. The three couplers with their isolators and detectors form the incident, reflection, and transmission channels. The broadband isolators are used with the broadband detectors to reduce the mismatch error. The matched load and short terminations are used to calibrate the system as discussed above. The waveguide switch is employed for measurement convenience. In using the waveguide switches, due care should be exercised when measuring low SWR values.

The outputs from the three amplifiers are fed into an A/D converter and then into the computer. The display can also be used for real-time monitoring and adjustment. The plotter is used for a permanent record of the test data.

The system shown in Figure 4 covers a full waveguide band. In this system detectors are used to measure power because of their advantages over thermistors. Reflectometer systems of the type described in this article have been built in the frequency ranges from 40-110 GHz. All the millimeter-wave components are built at Hughes Aircraft Company, Electron Dynamics Division. The reflectometer system can be supplied with just millimeter-wave hardware or as a complete system, depending on specific needs. Table I lists the components needed for a typical reflectometer.

TABLE I

MILLIMETER-WAVE COMPONENTS FOR REFLECTOMETER SYSTEM

Item	Qty.	Model No.	Description
1	1	4772XH-	Full Band Sweep Generator
2	1	4572XH-1000	Direct Reading Attenuator
3	1	4571XH-1000	Direct Reading Frequency Meter
4	3	4532XH-1010	Directional Coupler
5	1	4561XH-1000	Termination
6	3	4511XH-1000	Ferrite Isolator
7	3	4732XH-1100	Flat Detector
8	1	4567XH-1100	Adjustable Short

X = 1 – K_a-Band (26.5-40 GHz)
2 – Q-Band (33-50 GHz)
3 – U-Band (40-60 GHz)
4 – V-Band (50-75 GHz)
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6 – W-Band (75-110 GHz)

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Design Microstrip Oscillators with Coupled Line Matching

CARLOS CORREIA da SILVA BARTOLO
FRED E. GARDIOL
Ecole Polytechnique Fédérale de Lausanne
Chemin de Bellerive, Lausanne, Switzerland

SHAMSUR RAHMAN MAZUMDER
SPAR Technology Ltd.
Montreal, Canada

INTRODUCTION

Very small dimension oscillators can be realized using microstrip design where the solid state diode, TED or IMPATT, is placed directly on the substrate. All matching, tuning and biasing functions are carried out with microstrip circuits, allowing one to obtain in this fashion a very compact (flat) package. The reduction in size, in comparison with comparable waveguide realization, is, however, accompanied by a loss in flexibility: tuning and matching are more easily made in waveguide designs, where tuning screws can be used for this purpose. No similar adjustable devices exist in microstrip. Of course, it is possible to mechanically reduce the length of a matching stub by scraping part of the upper conductor (trimming). However, this is a destructive one-way process. Lengthening a stub is also feasible technically by using adhesive-backed metal strips. The operation, however, is tricky and must be carried out with great care.

Even so, the microstrip approach presents smaller capabilities as far as tuning and matching are concerned, in contrast with waveguide where such operations are done with successive adjustments of tuning screws, a quite reversible and non-destructive process. In order to effectively use the microstrip approach, it is desirable to separate the two processes of tuning (adjustment of signal frequency) and matching (adjustment of the impedance presented by the circuit to the microwave diode). In a practical design, one should be able to carry out the two operations quite independently.

STATEMENT OF THE PROBLEM

A microstrip microwave generator is schematically represented in Figure 1.

A tuning network, generally consisting of one or two open-ended stubs, provides a reactive impedance component across the diode terminals. Adjustment of this reactive component, by trimming off the stub lengths, serves to adjust the frequency of the signal: the oscillation occurs at the frequency at which the susceptances of the diode and of the outside circuit just cancel each other out.

The remaining real part of the generator impedance is in general different from the line impedance, most often chosen at the conventional value of 50 Ω, so that some kind of transformer is necessary to connect the generator to the load. The design is completed by a biasing network and also suitable filters and decoupling.

The conventional approach would be to use a quarter-wave transformer (either simple or multiple section). Unfortunately, the impedance ratio of such a transformer is fixed once and for all by the design. There is no simple way to modify the impedance ratio of a transformer realized on microstrip: one would have to reduce the width of the sections as well as their length.

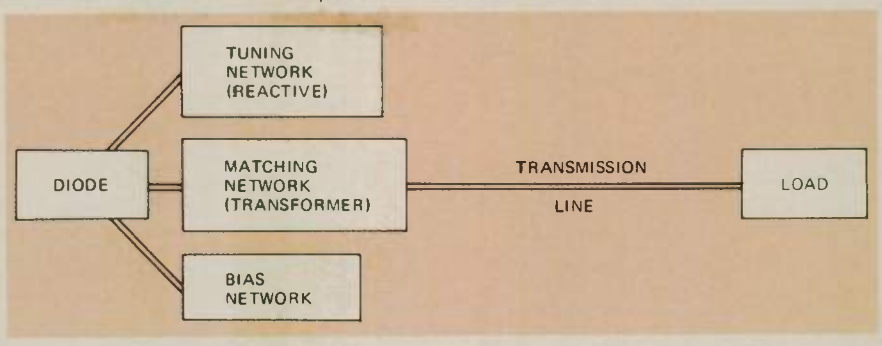

Fig. 1 Schematic representation of a solid state oscillator on microstrip.

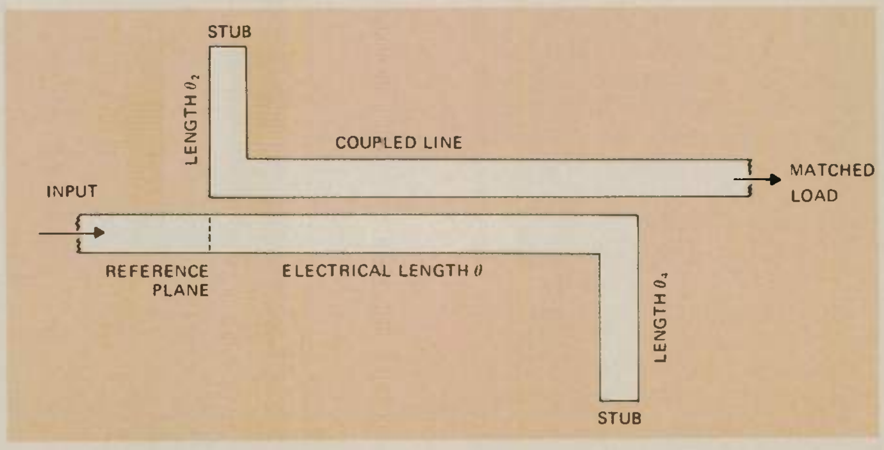

Fig. 2 Matching circuit consisting of a coupled line with two open-ended stubs.

The transformer cannot be adjusted to fit the particular diode used, hence a loss in performance and a need for different circuits corresponding to different types of diodes. Furthermore, the measurement of the diode impedance is a delicate and tedious process. In practice, it would be highly desirable to have an adjustable design, in which some limited amount of trimming is still possible. In the case of IMPATT oscillators, a large impedance is required for maximum power transfer to the matching network: this means that very narrow lines are required. They are difficult to realize accurately in practice, and may tend to heat up as the dc biasing current has to flow through them as well.

In essence, what is required is a variable, or at least adjustable transformer, in which the impedance ratio can be modified by some simple means, for instance by trimming a stub, as is done for tuning. As the process cannot be repeated easily in the case of error (do it right the first time!) the two operations should be non-interactive (orthogonal).

USE A COUPLER FOR MATCHING

An interesting way to solve this problem was found by considering the input impedance presented by a directional coupler, when two of its ports are terminated by reactances (open-ended stubs) and the fourth one is matched (**Figure 2**). The theoretical study of the structure, using impedance matrix characterization, is presented elsewhere.[1] Loci of constant conductance G and constant susceptance B lines are shown as a function of stub lengths in Figure 3 for one particular design. There is a singularity, near which small changes of phase produce large changes of admittance (high sensitivity). Far from the singularity, however, smooth variations can be obtained. Of particular interest is the possibility to adjust the real part G of the admittance without modifying the imaginary part B. As an example, setting $\theta_4 = 45°$, the value of G can be varied between 1 and 2.5 mS while the imaginary part remains approximately constant. This approximately covers the range required to match an IMPATT oscillator diode. The reactively terminated coupled line thus behaves as an adjustable transformer. An additional practical advantage is that it provides directly the required dc decoupling for the bias circuit.

PRACTICAL REALIZATION

The method was used to design an X-band IMPATT oscillator on an Epsilam 10 substrate (**Figure 4**). The IMPATT diode (type HP-5082-0435) is located at the T-junction, fitting in a hole drilled through the substrate, between the upper conductor and the ground plane. The length of the two open-ended stubs S_1 and

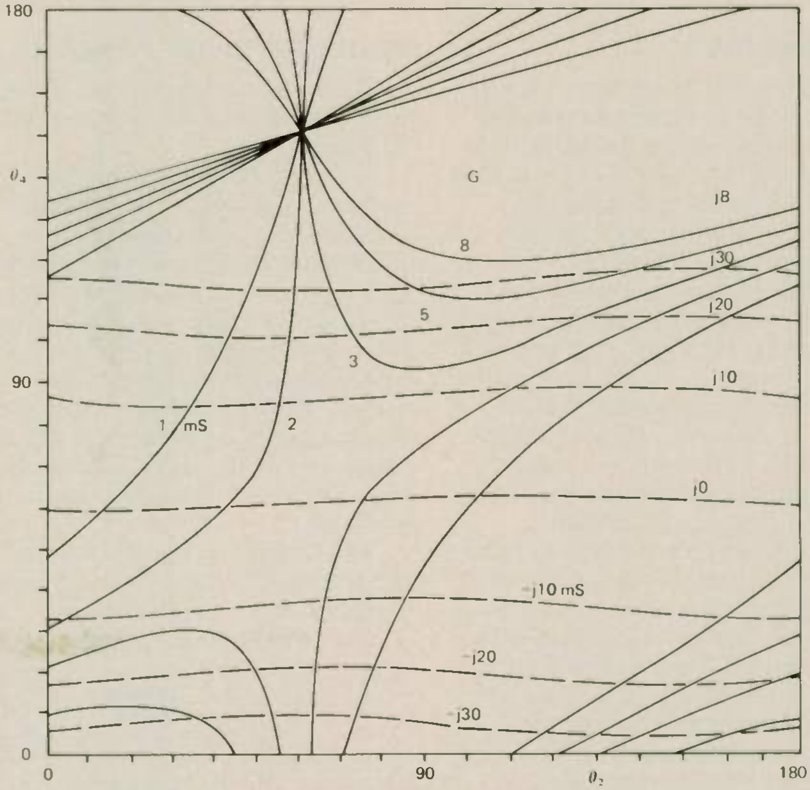

Fig. 3 Input admittance of coupled-line matching circuit for $\theta = 120°$, $Z_{oe} = 67\Omega$, $Z_{oo} = 33\Omega$. Solid lines represent the real part G (in millisiemens),* dashed lines the imaginary part B. Electrical lengths θ_2 and θ_4 are in degrees.

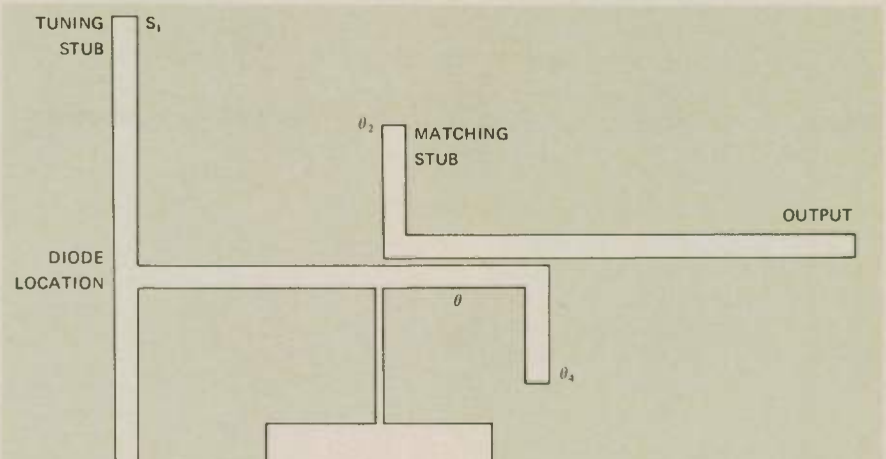

Fig. 4 Microstrip oscillator circuit layout showing tuning stubs S_1 and S_2 (frequency) and matching stubs θ_2 and θ_4 (power).

*(S) Siemen is a unit of conductivity in the MKS rationalized system.

S_2 was adjusted for oscillation at the design frequency of 10.4 GHz. A half-wavelength section of line between the diode and the coupler input serves to avoid interferences between the stubs. The length of stub θ_2 is adjusted for maximum output power (Figure 5): the adjustment must be made carefully, as the curve presents a rather sharp maximum. The signal produced was observed on a spectrum analyzer – the line is quite sharp and there are no spurious oscillations.

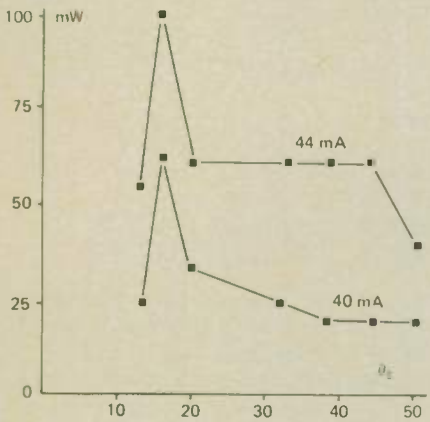

Fig. 5 Output power as a function of electrical length of stub θ_2, at 10.4 GHz, for $\theta_4 = 15°$, $\theta = 120°$, $Z_{0e} = 67\Omega$, $Z_{00} = 33\Omega$.

CONCLUDING REMARKS

A new approach to create solid state oscillators on microstrip was developed. The two operations of frequency tuning and impedance level matching were carried out independently. A signal power level of 100 mW was obtained at the design frequency of 10.4 GHz.

Further refinements considered are the addition of varactor diodes for fine tuning and matching. Variable capacitors with adequate biasing circuits placed at the ends of stubs S_1 and S_2 can be used to adjust the frequency of operation or to FM modulate the signal. At the end of stub θ_2, a variable capacitor can be used to adjust the power level (ALC) and for amplitude modulation of the signal.

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D. B. Rutledge was born in Savannah, Georgia, on January 12, 1952. He received his B.A. in mathematics from Williams College in 1973, the M.A. in electrical engineering from Cambridge University in 1975, and expects to receive the Ph.D. in electrical engineering in 1980 from the University of California, Berkeley. In 1975-76 he was associated with the Communications Group of General Dynamics Corporation, where he worked with airborne microwave video data systems. Mr. Rutledge is a member of Phi Beta Kappa and the American Physical Society.

S. E. Schwarz was born in Los Angeles, Calif. on Jan. 29, 1939. He received the B.S. in physics from Cal Tech in 1959, the A.M. in physics from Harvard University in 1961, and the Ph.D. in electrical engineering from Cal. Tech. in 1964. He has held positions with Hughes Research Laboratories, Bell Laboratories, and IBM Research Laboratories. Since 1964 he has been a faculty member of the Department of Electrical Engineering and Computer Sciences at the University of California, Berkeley, where he is engaged in work on millimeter-wave and infrared devices and quantum electronics.

He held a Guggenheim Fellowship in 1971-72 and is co-author (with W. G. Oldham) of the book "An Introduction to Electronics."

Constant Gamma Hyperabrupt Tuning Varactors

MICROWAVE ASSOCIATES, INC.
Burlington, MA

Microwave Associates is now offering a standard product line of GaAs Hyperabrupt Tuning Varactors with the added feature of "constant gamma." In this discussion, constant gamma is defined, the advantages to the circuit designer are described, and an approach to proper diode selection is discussed.

For PN junctions, the dependence of junction capacitance, C_j, on applied voltage, V, is given by:

$$C_j(V) = \frac{C_o}{(1 + \frac{V}{\varphi})^{\gamma}} \qquad (1)$$

where:

φ = the built-in potential ($\varphi = 1.3$ volts for GaAs)

C_o = a constant (mathematically equal to junction capacitance when V = 0)

γ = the capacitance - voltage slope exponent (gamma)

For simple abrupt junction varactors, gamma is constant and nominally equal to 0.5. The junction is referred to as hyperabrupt when $\gamma > 0.5$, and for most commercially available varactors, the value of γ varies widely with applied voltage.

From Equation (1), we observe that gamma is determined as the slope of the plot of junction capacitance C_j, versus total voltage (applied voltage plus built-in potential) on log-log graph paper. A typical plot of the new constant γ hyperabrupt tuning varactors is illustrated in **Figure 1** where the slope of curve ① is a constant $\gamma = 1.25$ over the applied voltage range of 2-20 volts. Notice that constant gamma is not maintained at low applied voltages, so $C_o = 6.6$ pF is a mathematical value determined by extending the constant slope to V = 0 (or $V + \varphi = 1.3$). The capacitance versus applied voltage curve ② is also shown in **Figure 1** for the chip, and curve ③ illustrates the C-V curve when the chip is mounted in an ODS-30 package having package capacitance of 0.17 pF.

The primary purpose of the constant gamma hyperabrupts is to permit the designer to achieve linear frequency tuning without the use of a linearizer. For a simple resonant circuit comprised of an inductance, L, and the varactor junction capacitance $C_j(V)$, the frequency-voltage relationship is given by:

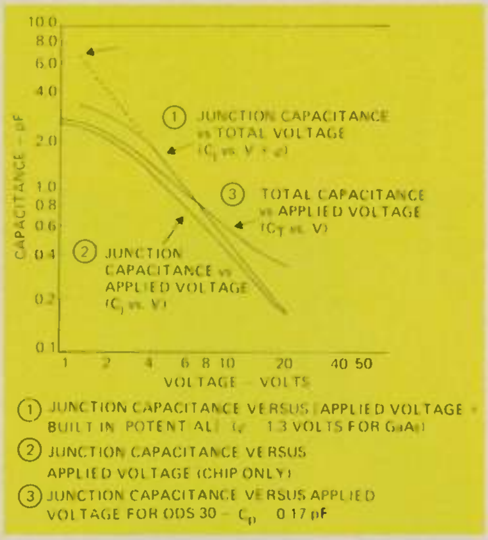
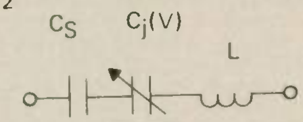


Fig. 1 GaAs Hyperabrupt C-V characteristics (Typical Hyperabrupt).

Fig. 2



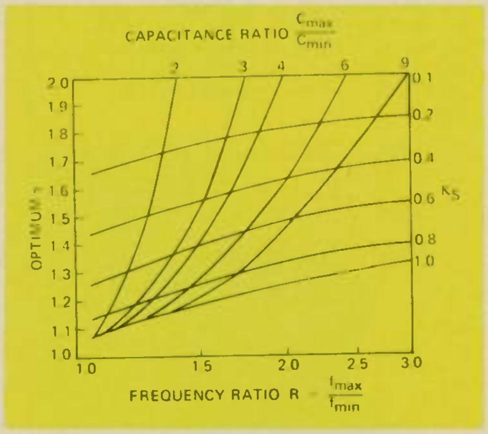


Fig. 3 Optimum γ selection for linear tuning.

$$f_r(V) = \frac{1}{2\pi\sqrt{L\,C_j(V)}} = \frac{1}{2\pi\sqrt{L\,C_o}} \cdot \left(1 + \frac{V}{\varphi}\right)^{\frac{\varphi}{2}} \qquad (2)$$

and the desired γ for linear tuning is 2.0. However, in nearly all microwave circuits, the varactor is not the only capacitance in the resonator. Instead, the capacitance of the active element, or parasitic or stray capacitance, or purposefully designed decoupling capacitance are a portion of the resonant structure. An analysis has been performed of the simple series circuit illustrated in **Figure 2** where the fixed capacitance, C_s, is in series with the varactor. The results of this analysis provides guidance to the selection of a suitable γ for the circuit designer. Now the total capacitance, C_T, of the resonant circuit can be expressed in terms of a coupling factor, K_s, as:

$$\frac{1}{C_T(V)} = \frac{1}{C_{TO}} \cdots \left[1 - K_s + K_s\left(1 + \frac{V}{\varphi}\right)^{\gamma}\right] \qquad (3)$$

where:

$$K_s = \frac{C_{TO}}{C_o} = \frac{\text{Total C of Resonator at V = 0}}{\text{Varactor Junction C at V = 0}}$$

When $K_s = 1$, the varactor is fully coupled and optimum $\gamma = 2$. When $K_s \to 0$, the varactor becomes heavily decoupled and only narrow-band frequency tuning is possible. For intermediate values of coupling in the range $0 < K_s < 1$, an optimum value of constant γ for linear frequency tuning is predictable. The result of the analysis is illustrated in **Figure 3**.

In this illustration, the optimum γ value is plotted versus the frequency ratio, f_{max}/f_{min}, with the coupling factor, K_s, as a parameter. Notice that linear tuning can be achieved for constant gamma within the limits $1.0 < \gamma \leqslant 2.0$ depending on the coupling factor.

This simplified analysis may be used for selection of constant gamma hyperabrupts under various coupling conditions. A decoupling limit occurs when the C_{max}/C_{min} is not available in the varactor. Most broadband tuning requirements are optimized in user circuits with $1.2 < \gamma < 1.4$, while circuits with narrow tuning bandwidth needs utilize $\gamma \to 1.0$.

Four new standard series of GaAs hyperabrupt tuning varactors are available with constant γ of 1.0 or 1.25 over the ranges 2 - 12 volts or 2 - 20 volts. Capacitance values are available from 0.4 pF to 10.0 pF at −4 volts in each series. Q performance is roughly equivalent to silicon abrupt junction varactors.

Other values of γ are available by special request. Data sheets with further details are available.

Circle 102 on Reader Service Card

Product Feature

Multi-Octave Single Band Sweep Generator

TEXSCAN CORP.
Indianapolis, IN

The new XR-1500 is a single band phase-locked sweep generator covering the 10 1/2 octaves from 1 - 1500 MHz. The instrument incorporates features which make it appropriate for a wide range of applications.

A sweep across the entire 1 - 1500 MHz range without band breaks and flat within ± 0.25 dB is available. In this mode, residual FM is approximately 20 kHz and drift is 500 kHz/5 min.

In sweep modes covering up to 2 MHz, however, a phase-lock loop is incorporated which improves stability to 1 kHz p-p residual FM and long-term drift of 50 kHz/5 min. These characteristics make possible the alignment and adjustment of circuits with bandwidths as narrow as 5 to 10 kHz.

The XR-1500 employs a clocked-marker system which avoids the "beat" problems which can be encountered with harmonically generated markers. A 100 MHz crystal source is divided by 10 and by 100 to provide 10 and 1 MHz markers whose amplitudes are held constant by a level and clip circuit. Marker width and amplitude may be adjusted for convenient display and markers may be tilted for greater visibility on steep-sided response curves.

Output level is controlled by two step attenuators and an infinite resolution vernier. The total range is over 70 dB.

Key specifications for the XR-1500 include:

Frequency Range	1-1500 MHz
Maximum Power Output	+ 7 dBm
Frequency Accuracy	± 20 MHz
Attenuation Accuracy	± 0.2 dB (1-9 dB)
	± 3% (10-60 dB)
Power Consumption	50 watts
Size	4"H.x12"W.x15"D.
Weight	18.5 lbs.

Price is $2,850 with delivery in twelve weeks.

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Microwave Products

Devices

HIGH POWER BALANCED RF TRANSISTORS FOR 800 MHz BAND

Models DBL45-12, DBL60-12 and DBL80-12 are balanced RF power transistors for the new 800 MHz communications band and offer rated RF power output across the entire 806-866 MHz band. Devices are designed for high power, base/mobile dispatch service. Model DBL45-12 has power output of 45 W, (min.) at power input of 15 W (min.). Model DBL60-12 offers power output of 60 W, (min.) at power input of 20 W. Unit DBL80-12 has a 80 W power output at 27 W power input. All operate from a 12.5 V supply voltage. **Communications Transistor Corp., San Carlos, CA.**

(415) 592-9390. Circle 105.

Components

HERMETIC SMA BULKHEAD FEEDTHROUGH ADAPTOR

Unit #705627-101 is a redesigned hermetic SMA bulkhead feedthrough adaptor. Model utilizes a 50 Ω solder-in Kovar glass seal with welded contacts. Component has SWR of 1.25 maximum from 2.0-18.0 GHz. Mating face is a SMA jack (female) on both ends in accordance with MIL-C-39012B. Avail: stock. Price: $7.82 in 1,000 piece qty. **Cablewave Systems Inc., North Haven, CT. Steven Raucci, Jr., (203) 239-3311. Circle 117.**

TUBULAR LC FILTERS COVER 1-1000 MHz RANGE

Series FA-3800 are miniature high performance LC filters for use in the 1-1000 MHz frequency band. These units replace larger tubular filters and offer excellent control of passband and phase response. Attributes of the models are achieved through synthesis procedures and component topology unavailable for tubular design. Available with SMA or BNC connectors for direct tubular replacement, or pins for PC board mounting. **Comstron Corp., Freeport, NY. Leonard J. Borow, (516) 546-9700. Circle 119.**

WR28 ROTARY JOINT

A single channel L-style WR28 waveguide rotary joint, Model RC12900, operates from 26.5-40.0 GHz. Unit has a maximum insertion loss of .5 dB, SWR of 1.50, maximum, and 10 kW peak power handling capability. Model can be pressurized up to 30 psig and rotates at 60 rpm with a torque of < 20 oz.-in. Size: 3" H. x 2" D. housing x 1.88" arm L. **MAST Microwave, Burlington, MA. Chris Theophile, (617) 273-4640. Circle 125.**

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Broadband waveguide bandpass filters, series 1300B, cover full waveguide bands in the 12-60 GHz range. Units have attenuation of 0.8 dB, typical, 1.5 maximum and SWR 1.4, typical. Rejection is 60 dB minimum at frequencies 15% above and below the edges of the passband. Dimensions: 1.0-1.32" square x 3.5-7.0" L. Price: $1450 for stan. full-band models. Del: 60 days. **Epsilon Lambda Electronics Corp., Batavia, IL. Robert Knox, (312) 879-6006.** Circle 123.

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A line of standard waveguide switches of modular construction, with interchangeable drive head assembly and full band specifications is offered. Both magnetic latching and failsafe types are available. Drive voltages, manual override and manually operated styles provide 42 interchangeable drive head combinations. Ten combinations of port and waveguide flange styles make 168 to 420 standard waveguide combinations per waveguide size. **Logus Manufacturing Corp., Deer Park, NY. (516) 242-5970.** Circle 124.

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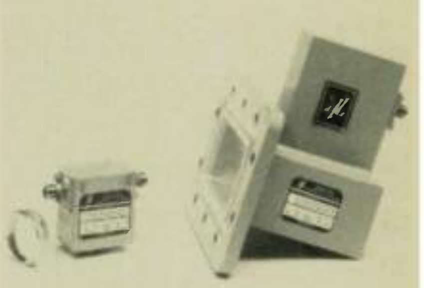
The PD-1000-4 is a 4-way power divider which covers the 10-1000 MHz frequency range. Isolation between outputs is 25 dB, min. from 10-20 MHz and 30 dB, min. from 20-1000 MHz. Insertion loss is typ. 0.6 dB from 10-400 MHz and 1.25 dB from 400-1000 MHz over split loss. SWR is 1.5 from 10-20 MHz; 1.2 from 20-1000 MHz. Impedance is 50 Ω. Offered with BNC, TNC and SMA connectors. Del: from stock. **American Microwave Corp., Damascus, MD. (301) 253-6782.** Circle 116.

MM WAVE VARIABLE ATTENUATOR SERIES

Model 75-110 A is a variable attenuator designed for the 75-110 GHz band. It has a power capability of 0.3 W; insertion loss is 0.5 dB and max. SWR is 1.2; max. attenuation is 25 dB. Model 75-110 AE is a direct-reading variable for the same band attenuator with a 0-50 dB range. Type 75-110 AV is a micrometer-controlled 0-30 dB for high resolution measurements in the 75-110 GHz band. **Thomson-CSF Components Corp., Electron Tube Div., Clifton, NJ. Anthony Laconti, (201) 779-1004.** Circle 128.

LOW LOSS ISOLATORS FOR 3.7-4.2 GHz BAND



Model 101104141, an isodaptor, and Model 101104142, an isolator, perform optimally over the 3.7 to 4.2 GHz range. Both the isodaptor, with waveguide input and SMA coax output, and the isolator, with SMA coax in and out, provide insertion loss of 0.15 dB max. while maintaining a 1.1-1 SWR and 26 dB minimum isolation. Either unit is available in a weather-proofed version for installation close to an antenna. Price: 101104141, $375; 101104142, $230, in small qty. Del: 6-8 wks. ARO. **Eaton Corporation, Communications Products Div., Addington Microwave Components, Sunnyvale, CA. Jim Wilson, (408) 738-4940.** Circle 120.

90° QUADTRATURE HYBRIDS SPAN 200-18,000 MHz

"H" series 90° quadrature stripline hybrids cover the 200-18,000 MHz range. Typical specifications for Model 7012 include: 7-12 GHz frequency range; 20 dB min. isolation; 0.5 dB max. insertion loss, SWR max. of 1.35; phase balance of ± 3 and ± 0.3 dB amplitude balance. Units are provided with sealed SMA steel connectors per MIL-C-39012, with captivated center conductors. Price: $95, except Model H1218 (12-18 GHz) – $125. Del: from stock. **Engelmann Microwave Company, Montville, NJ. Carl Schraufnagl,** (201) 334-5700. Circle 122.

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Line of square flange mounting microminiature RF connectors save 45% of the panel space of conventional SMA connectors. Nanohex units have a square mounting flange measuring .281" per side, feature a telescoping interface and provide low SWR up to 12.4 GHz. Units come in screw-on, snap-on, or slide-on matings and offer extended dielectrics. **Sealectro Corporation, R.F. Components Div., Mamaroneck, NY.** (914) 698-5600. Circle 114.

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A 2.5 W FLATSO resistor is power rated based on 70°C operation with maximum no load temperature at 150°C. Radial lead thick film resistor, Model FL2-½ has a voltage rating of 10,000 V, maximum, with temperature coefficients available at ± 100, ± 250 and ± 500 ppm/°C. Size: 1.125" L. x .690" H. x .175" T., with radial leads that are copper solder coated. For printed circuit mounting, lead bending is not required. Price: For standard TC, $1.34 to $2.34 in lots of 1,000 pieces depending on tolerance. Del: 6 wks. ARO **KDI Pyrofilm Corp., Whippany, NJ. Bill Dodge,** (201) 887-8100. Circle 121.

GaAs PULSED IMPATT SOURCE

Model AT-SP44 is a GaAs pulsed IMPATT source in the 8-10 GHz band. Power output is 12 W ± 1 dB over the 0-60°C heat sink operating temperature range. Source requires a constant current modulator to provide pulsewidths of 50 ns to 1 μs, pulse repetition frequencies of 0-3 MHz, and duty factors of 0-15%. Rise and fall times to 10 ns can be obtained. The dc to RF efficiency is nominally 18%. Frequency stability over temperature is typically ± 12 MHz. Size: 3¼" x 1" D. Weight: 3 oz. Price: $1900, 1-9 qty. Avail: 30-60 days. **Ad-Tech Microwave, Inc., Scottsdale, AZ.** R. C. Havens, (602) 941-1290. Circle 115.

VANE TYPE POWER DIVIDER SWITCH

Model P/N 39D06100 is a vane type power divider waveguide switch for failsafe redundant system integration. Switch covers the 7.05-10 GHz frequency band and operates from 28 Vdc ± 10% at .5 A max. current, with pull-in and hold at 20 Vdc and 28 Vdc. Switching time is 100 ms maximum. SWR is 1.15, max. for positions 1 and 3; 1.2, max. for position 2. Power handling is 1.5 kW CW. **Transco Products, Inc., Venice, CA.** Ray E. Williams, (213) 822-0800. Circle 129.

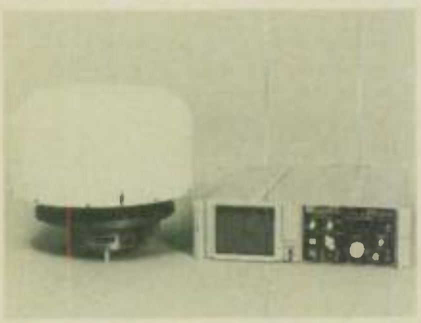
Hardware

MODPAK RFI SHIELDED PACKAGES

Modpak system of shielded packaging has RF connectors designed to provide typical SWR of 1.2 or less up to 1.5 GHz when a 50 ohm microstrip PC board is installed in the case. Mounting clip and connector locking system permit PC boards to be positioned and retained quickly. Other features include direct access to circuits and choice of connectors. Avail: from stock – 1 day, for 26 stand. packages; 3 wks., for custom models; 4 wks., for models fabricated from customer drawings. **Adams-Russell, Modpak Division, Burlington, MA. Susan MacDonald,** (617) 272-3330. Circle 103.

Antennas

COMPACT, SPINNING DF ANTENNA SYSTEM



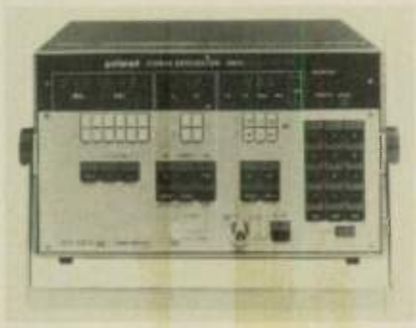
A spinning DF antenna system consists of L6-24 spinning DF antenna, EP-30 pedestal and EC-60 control/display unit. Antenna provides DF coverage over 1-18 GHz frequency range, with wide selection of band breaks and polarizations, as well as switched or simultaneous RF outputs. Antenna consists of a rotating dish and foam-filled feed, enclosed in dust cover to protect elements and provide a smooth surface to reduce rotation resistance. Permanent magnet dc torque motor rotates the antenna clockwise or counterclockwise at speeds variable to 200 rpm; maximum torque is 1.5 foot pounds. A synchro control transmitter/tachometer provides three-wire antenna position and velocity information. Control/display unit has a four-decade digital display for bearing readout, and has three front panel selectable modes of antenna operation: SCAN, SLEW and SECTOR SCAN. Size: L6-24, 8" x 15"; EC-60, 5.25" H. in standard 19" rack. **Watkins-Johnson Company, Palo Alto, CA.** (415) 493-4141. Circle 108.

CIRCULAR POLARIZED mm HORN ANTENNAS

Model CA-16 and CA-17 are circular, polarized horn antennas which feature half-power beamwidths of 60° nominal, input SWR of 2.0 maximum, and minimum gain of 6 dBi. Size: CA-16 (18.0-26.0 GHz) – 2.9" L. with an input flange mating to UG 595/U; Model CA-17, (26.5-40 GHz) – 2" L. with an input flange mating to UG-599/U. Weight: Model CA-16, 40 oz.; Model CA-17, 20 oz. **Sanders Associates, Inc., Microwave Division, Manchester, NH.** (603) 669-4615. Circle 107.

Instrumentation

SYNTHESIZED SIGNAL GENERATOR



Synthesized signal generator for programmable (IEEE-488) AM/FM measurements covers 0.4 to 520 MHz (standard), 0.4 to 1040 MHz. Model SMS supplies output signals calibrated from +13 to −137 dBm with calibrated modulation capabilities for AM, FM and PM. Instrument has microprocessor control and offers frequency resolution of 100 Hz on an 8-digit display and amplitude resolution of 0.1 dB on a 3½-digit display in µV, mV, dB µV or dBm. **Polarad Electronics, Inc., Lake Success, NY. Gene Kushner,** (516) 328-1110. Circle 109.

Systems

L BAND RF POWER AMPLIFIER OFFERS 1600 W CW

A solid state Class C RF power amplifier, PA 1385-122/4800 provides 63 dB gain and 1600 W CW output power over the 1370-1400 MHz frequency range. Output flatness is ± 0.5 dB. Harmonics are ≥ 85 dB, spurious ≥ 85 dB and input/output impedance is 50 ohms. Load SWR protection to infinity at all phase angles is provided and BITE status indicators monitor overall amplifier performance down to replaceable module level. Prime power is 208 Vac, three phase, 47 to 420 Hz. **Microwave Power Devices, Inc., Plainview, NY.** (516) 433-1400. Circle 106.

(continued on page 74)

(from page 73)

Materials

EPITAXIAL GaAs WAFERS FOR HYPERABRUPT VARACTORS

Material doubles the maximum tuning ratio of a varactor by tailoring the doping profile in the active region of the epitaxial layer to suit the capacitance change needed. Carrier concentration is given a negative ramp from the substrate interface towards the surface of the epitaxial layer, or the carrier concentration near the surface can be increased by growing a doping spike. Tuning varactor wafers for all frequencies, X to Q band, are available as is material for Read IMPATT devices. **EM Chemicals, 500 Executive Blvd., Elmsford, NY. (914) 592-4660.**

Circle 111.

LOSSY ROD, BAR AND SHEET STOCK FOR TRANSMISSION LINES

ECCOSORB® MF is a series of eight magnetically loaded epoxide rod, bar and sheet materials designed for use as absorbers, attenuators and terminations in waveguides, coaxial lines and striplines. Eight members of the series permit the choice of eight different loadings, for all of which the approximate relative impedance magnitude is in the 0.4-0.5 range. Attenuation at 8.6 GHz ranges from values as low as 1.5 dB/cm for the lightest loading to 65 for the highest loading. Materials have a service temperature of 180°C, density of 1.6-4.9 g/cc, durometer of 85 Shore D and tensile strength of > 8000 psi. Thermal expansion is $30 \times 10^{-6}/^{\circ}C$ and thermal conductivity is 10(BTU) (in)/(hr) (ft^2) (F) and water absorption in 24 hrs. is < 0.3%. **Emerson & Cuming, Canton, MA. Jeanne B. O'Brien, (617) 828-3300.**

Circle 110.

ERRATA

A number of errors and omissions were made in the publication of the article "Digital Radio Measurement Using the Spectrum Analyzer," Engelson & Garrett, in the April 1980 *Microwave Journal*. Interested readers may obtain an accurate reprint copy of the article by writing to Mr. Len Garrett, Tektronix Inc., DS 58-741, PO Box 500, Beaverton, Oregon 97077.

The Advertising Index of the May 1980 *Microwave Journal* improperly listed MAST Microwave/Megavolt Corp. as Divisions of Unaworld Corporation.

RESEARCH STUDY

DIGITAL MICROWAVE RADIO MARKETS

- Reviews key technological and market trends in conventional DMR market and also in the Digital Termination System market for systems such as the Xerox proposed XTEN satellite/terrestrial network.
- Analysis of tradeoff between analog and digital microwave radio for thin, medium and dense routes, with discussion of future probable trends.
- End-User market is analyzed by segment and also by frequency band used, with discussion of likely impact of fiber-optics and other new technologies.
- Supplier industry structure, with discussion of supplier strategies and analysis of market share by frequency band.
- Ten-year market projection of growth in DMR requirements by Bell, Independents, Specialized Carriers, etc.
- 158 pages; 33 illustrations; published March 1980; price $895.00.

For free descriptive literature and detailed table of contents, contact:

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New Literature

FILTER SHORT FORM CATALOG

A short form catalog describes over 1,000 RFI/EMC filters and feedthru capacitors. The 44-page booklet provides complete mechanical and electrical specifications for filters rated from 0.5 to 400 A. It describes parts manufactured for both military and commercial applications and provides a questionnaire form to assist in specifying non-standard parts. **RFI Corporation, Bay Shore, NY. (516) 231-6400.** Circle 155.

1980 COAXIAL FILTER CATALOG

Catalog C/80 describes filters and traps for use in the 5-1000 MHz frequency range (CATV, MATV, and CCTV) for 50 and 75 ohm systems. Information on over 700 filters, including bandpass (narrowband, wideband and diplexed), lowpass, highpass and band reject types is featured. Booklet is divided into six sections: Bandpass Filters; Low and High-Pass Filters; Pay TV Traps & Filters; Special Purpose Networks; Traps, General Purpose; and World Channel Allocations. **Microwave Filter Company, Inc., East Syracuse, NY. Emily Bostick, (315) 437-3953.** Circle 140.

DATA SHEET ON BeO SUBSTRATES

Data Sheet 7707-5 provides technical information about beryllium oxide substrates. This four-page sheet discusses thermal conductivity, surface finish and dimensional tolerances and lists available substrate specifications. **Valley Design Corp., Littleton, MA. Frank A. Reed, (617) 486-8933.** Circle 158.

ACOUSTIC SIGNAL PROCESSING HANDBOOK

Vol. III of a Handbook of Acoustic Signal Processing is on Pulse Expansion/Compression IF Subsystems for Radar. Booklet covers specifying resolution of bandwidth, side lobe suppression, system bandwidth, dynamic range S/N ratio, theme noise limitations, effects of Doppler frequency shift and other influential parameters. **Anderson Laboratories, Inc., Bloomfield, CT. F. Richard Cosma, (203) 242-0761.** Circle 132.

CATALOG ON RF DIODES AND TRANSISTORS

An illustrated, 24-page booklet on RF diodes and transistors describes function, application and essential parameters of the product line. Separate sections cover variable capacitance tuner diodes, hyperabrupt variable capacitance tuner diodes, Schottky barrier diodes and RF transistors. In addition to industry standard RF diodes and transistors, alternatives which offer same performance at lower cost are listed. A product index and package artlines for over 125 types are included. An O.E.M. price list is provided in an accompanying leaflet. **Ferranti Electric, Inc., Commack, NY. (516) 543-0200.** Circle 136.

COAXIAL CONNECTOR ASSEMBLY PROCEDURE MANUAL

A fully illustrated manual provides information on trimming instructions for flexible and semi-rigid coaxial cable. The manual includes a detailed display of tools, methods, and technical data needed to complete cable assemblies that meet requirements of MIL-C-39012 connector specification. **Omni Spectra, Inc., Microwave Connector Div., Waltham, MA. Ernest J. Devita, (617) 890-4750.** Circle 154.

RFI SHIELDED CASE CATALOG

Catalog No. 0380 is an eight-page, two-color catalog which describes a series of low cost RFI shielded cases, RF transfer switches and accessories. Illustrated with photos and drawings, the catalog has information on a variety of blank cases, standard size cases and a custom series. The RFT series is described and accessories such as circuit boards, feedthroughs, RF connectors, test cables, adapters, die cast boxes, gaskets, coaxial terminations, attenuators and an RF transfer switch line are shown. **COMPAC, Deer Park, NY. (516) 667-3933. Circle 135.**

CAGE JACK CATALOG

A condensed, 20-page two-color catalog describes cage jacks and includes information and data about plugs and patching products. Typical insertion/withdrawal force distributing curves, current ratings and current carrying capacities for the line are shown. Sixteen pages are devoted to listings/descriptions of PC jacks and plugs, both insulated and non-insulated types. Each product is identified by a part number. **Cambion, (Cambridge Thermionic Corporation), Cambridge, MA. William G. Nowlin, (617) 491-5400. Circle 134.**

BOOKLET ON EPOXY HAZARDS

A 24-page booklet on the hazards of working with epoxy products describes and identifies health and safety hazards, and features 11 "Do and Don't" rules. "Epoxy-Wise is Health-Wise" also presents a collection of case history accounts of how careless workers got into difficulty by not observing these precautions. A related section details certain procedures practiced by some workers to protect themselves and proceeds to show how and why the procedures can be inadequate. **TRA-CON, Inc., Medford, MA. M. M. Vitale, (617) 391-5550. Circle 157.**

SMA CONNECTOR CATALOG

Catalog SMA-9 covers a complete SMA connector line. It contains complete technical specifications and drawings as well as cross references to MIL-C-39012, Series SMA, listings. Reference plane dimensions are provided. **Sealectro Corporation, RF Components Division, Mamaroneck, NY. (914) 698-5600. Circle 141.**

CATALOG ON HERMETICALLY SEALED SMA IC LAUNCHERS

An eight-page catalog (No. 203A) describes a complete line of hermetically sealed SMA microwave integrated circuit launchers. Brochure features a description of products; electrical, environmental and mechanical specifications; installation instructions; and outline drawings. **Cablewave Systems, Inc., North Haven, CT. Steve Raucci, (203) 239-3311. Circle 133.**

CATALOG OF MILLIMETER WAVE PRODUCTS

Data for its complete solid state millimeter-wave product line is featured in a 100-page catalog. Products are listed in an alphabetical capability index which shows the frequency bands in which products or capabilities lie. **Hughes Aircraft Company, Electron Dynamics Division, Torrance, CA. (213) 534-2121. Circle 137.**

BROCHURE ON SCHOTTKY BARRIER DIODE QUADS

B-4216B is a brochure which describes a series of Schottky barrier diode quads developed for use as double balanced mixers and for other applications such as phase detectors, AM modulators and pulse modulators. Complete specifications for a low barrier, a medium barrier and a high barrier series of the beam lead quads are available in the brochure. **Microwave Associates, Inc., Burlington, MA. (617) 272-3000. Circle 139.**

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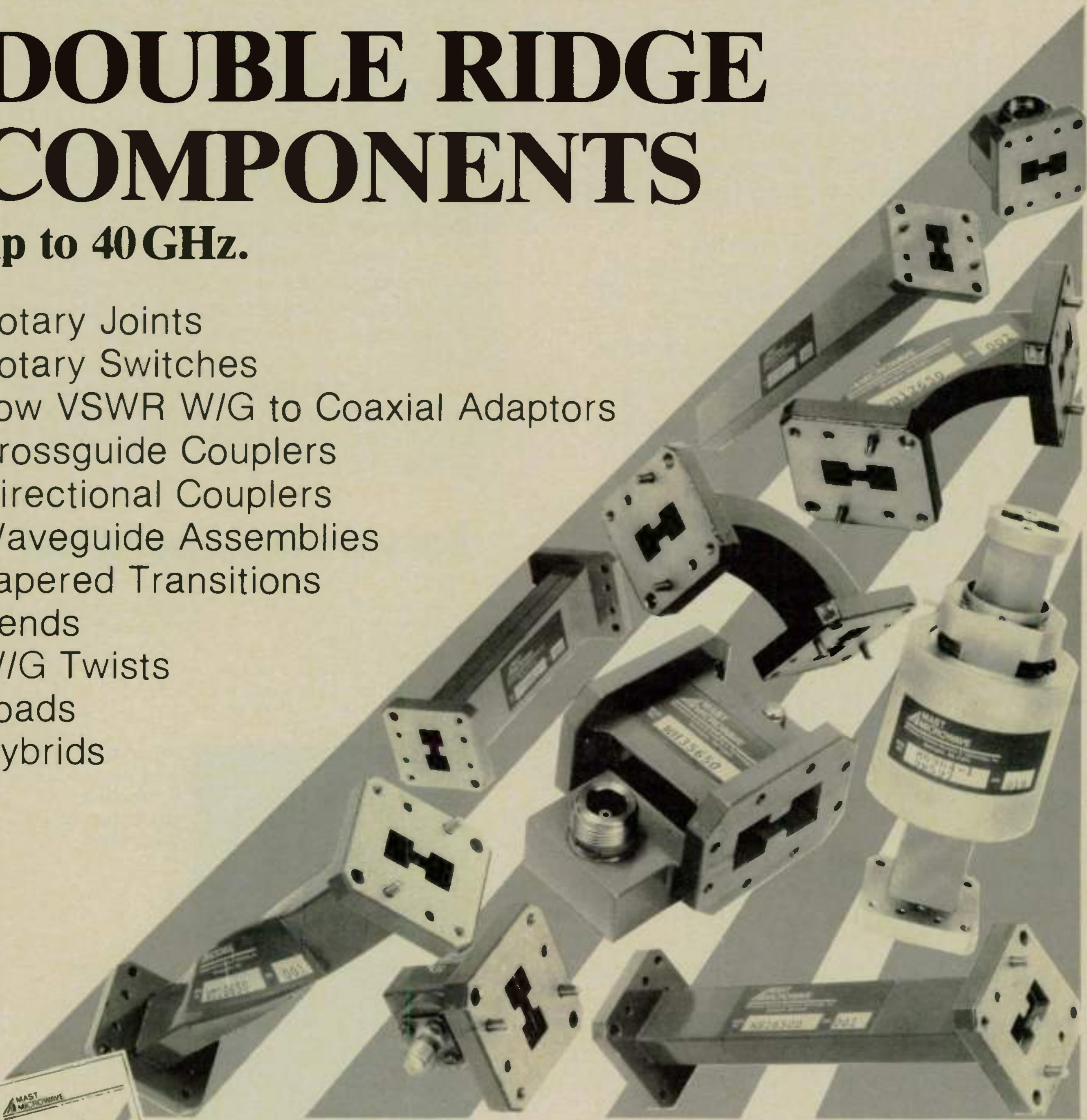

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