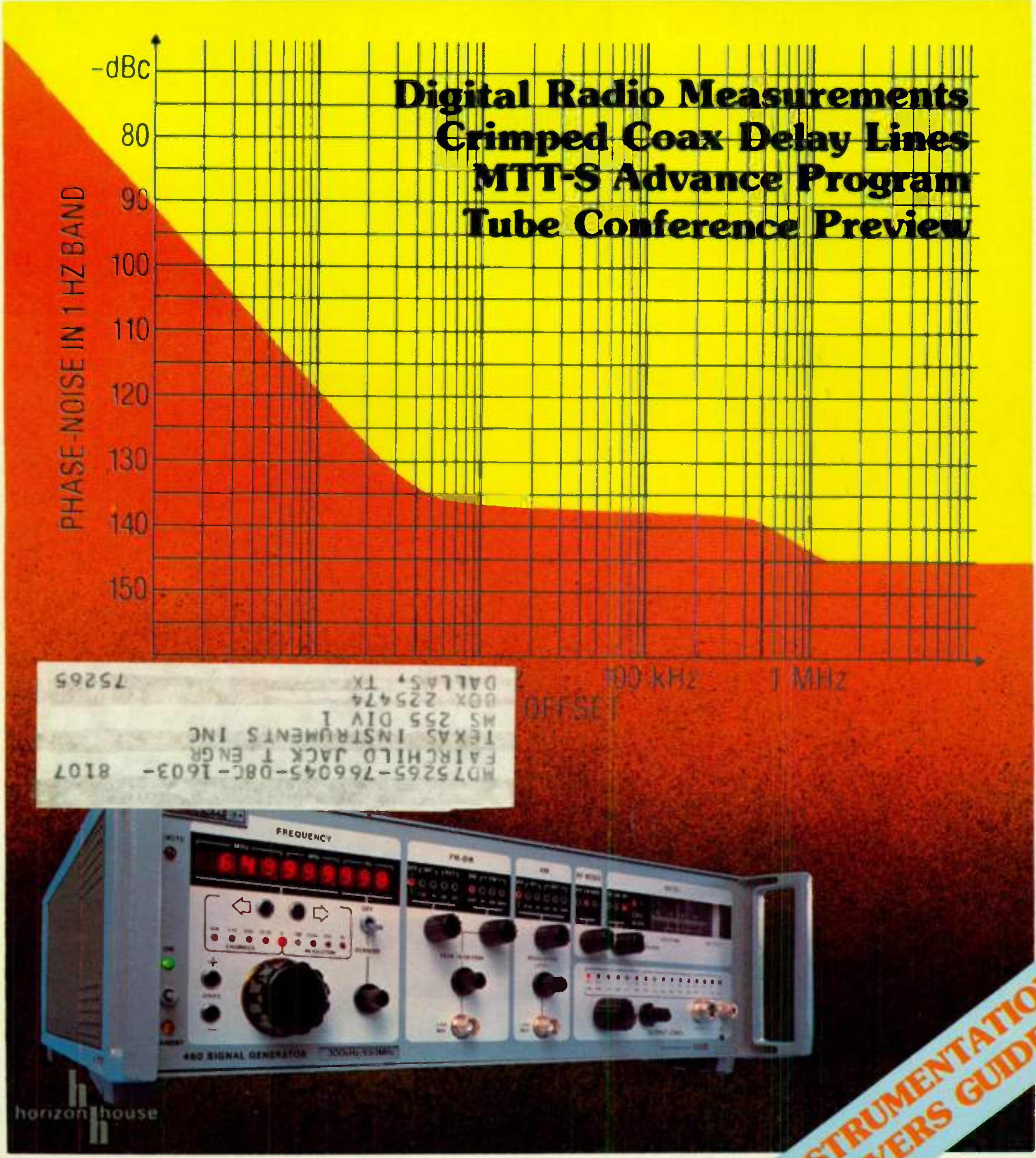
microwave JOURNAL

INTERNATIONAL EDITION □ VOL. 23, NO. 4 □ APRIL 1980

Digital Radio Measurements
Crimped Coax Delay Lines
MTT-S Advance Program
Tube Conference Preview



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INSTRUMENTATION BUYERS GUIDE

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Stable, low-noise klystron cavity oscillators with solid-state plug-removeable modules.

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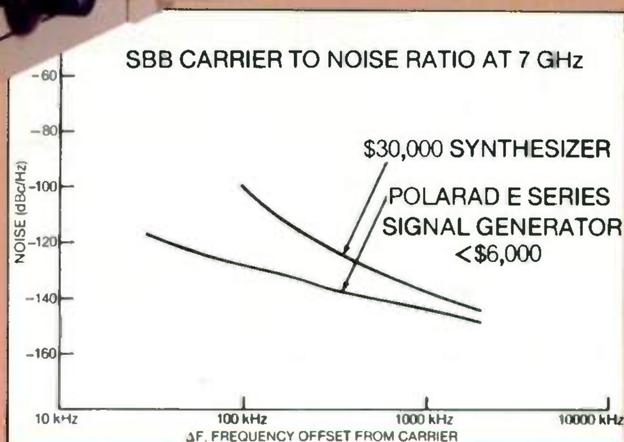

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VOLUME 23, NUMBER 4
USPS 396-250
APRIL 1980

ON THE COVER: AILTECH's Model 460 Programmable Signal Generator offers synthesizer features together with phase noise characteristics approaching those of cavity-based sources.

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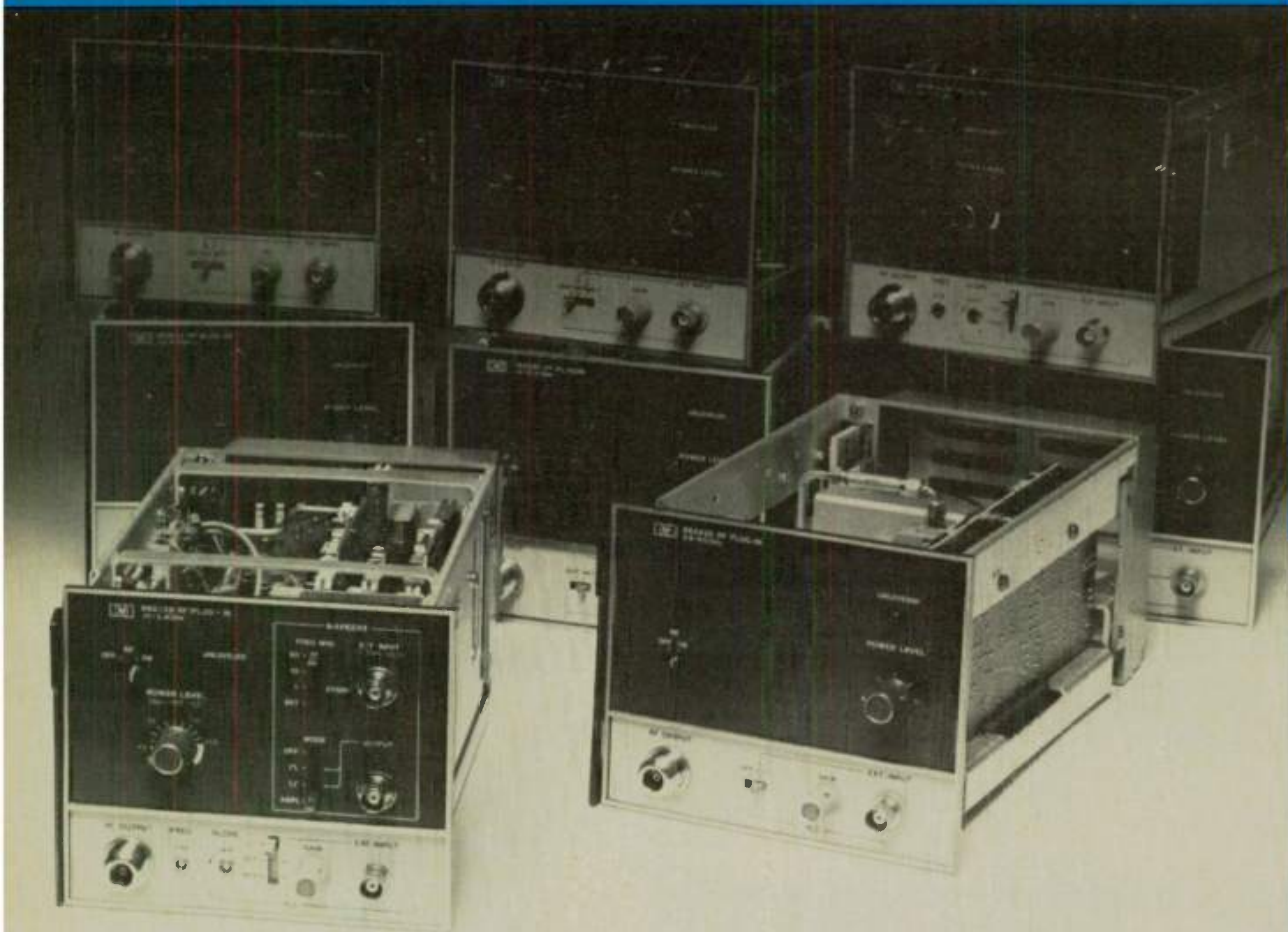
Microwave Journal is issued without charge upon written request to qualified persons working in that portion of the electronics industry including governmental and university installation that deal with VHF through light frequencies. Other subscriptions, domestic, $25 per year, two-year subscriptions $45, foreign $30 per year, two-year subscriptions $54, back issues (if available) and single copies $2.50.



Popular plug-ins include:

Model No.	Freq. Range	Output Power	Price
86222A/B	10 MHz-2.4 GHz	20 mW	$4100/4800
86235A	1.7-4.3 GHz	40 mW	$3500
86240A	2.0-8.4 GHz	40 mW	$4150
86240C	3.6-8.6 GHz	40 mW	$4950
86245A	5.9-12.4 GHz	50 mW	$4600
86260A	12.4-18 GHz	10 mW	$3950
86290A	2-18 GHz	5 mW	$14,250
86290B	2-18.6 GHz	10 mW	$16,250
86290B opt. H08	2-22 GHz	2 mW	$19,250

HP-IB programmability option adds value.

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HP-IB: Not just IEEE-488, but the hardware, documentation and support that delivers the shortest path to a measurement system.

When you add the HP-IB option ($950) to the 8620C Mainframe, you can program up to 10,000 frequency points per band on up to 4 bands with a variety of sweep modes. With the precision 86222 or 86290 RF units installed in the 8620C, excellent repeatability is achieved thanks to the exceptional frequency accuracy and linearity of these plug-ins.

A real advantage of HP-IB is you get not only the bus architecture you need, but the documentation support that can help you get your system up and running in weeks, instead of months.

A prime example is the scores of HP-IB automatic microwave network analyzer systems now in use. Major elements of these systems are the 8620C/86290B Sweeper, HP 8410 Network Analyzer, and HP 9825A Desktop Computer. Systems like this are fully described in HP Application Notes 221 and 187 Series. You can get copies of these plus information on the 8620C from your local HP sales office, or write Hewlett-Packard, 1507 Page Mill Rd., Palo Alto, CA 94304.

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Coming Events

1980 MTT-S INT'L MICROWAVE SYMPOSIUM/ EXHIBITION MAY 28-30, 1980 Sponsor: IEEE MTT-S (Microwave Theory and Techniques Society). Place: Shoreham Hotel, Washington, DC. Contact: B. Sheleg, Code 5733, Washington, DC 20375. Tel: (202) 767-2297.

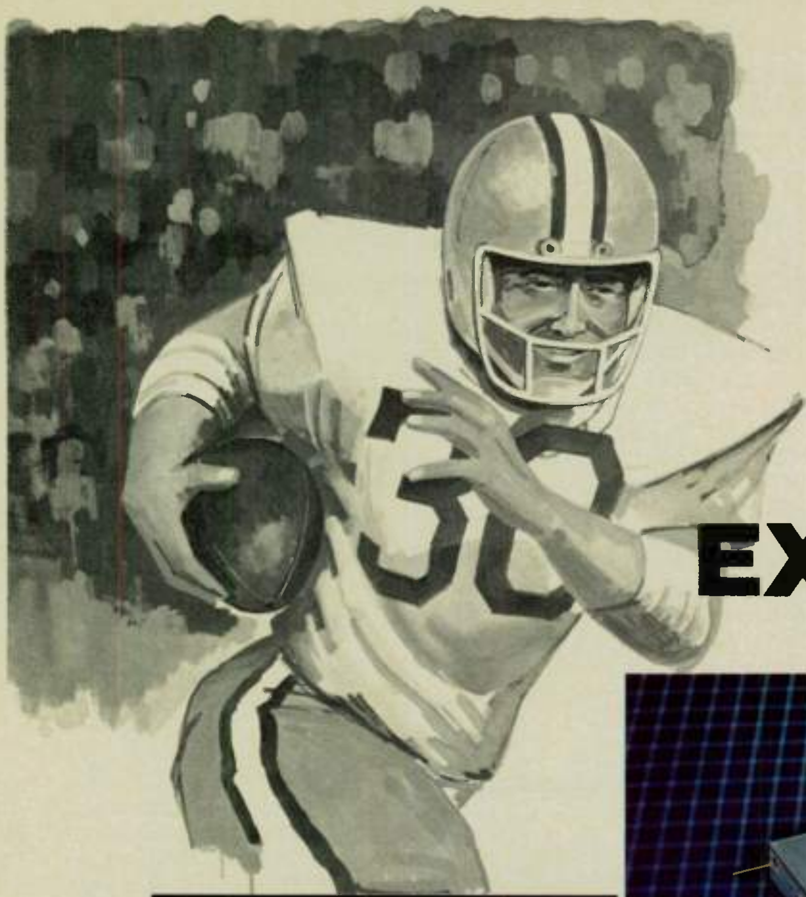
15TH SYMPOSIUM ON ELECTRO-MAGNETIC WINDOWS JUNE 18-20, 1980 Sponsor: Georgia Institute of Technology. Place: GIT, Atlanta, GA. Topics: radomes and radome techniques. Contact: D. J. Kozakoff, Engineering Experiment Station/EML/RSD, GIT, Atlanta, GA 30332 Tel: (404) 894-3505.

38TH DEVICE RESEARCH CONFERENCE JUNE 23-25, 1980 Sponsor: Rockwell Int'l. Place: Cornell University, Ithaca, NY. Contact: Fred A. Blum, Conference Chairman, Rockwell International, P.O. Box 4761, Anaheim, CA 92803. Tel: (714) 632-2584.

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. Chrm., Industrial Institute of Electronics, u1. Dluga 44/50, 00-241, Warszawa, Poland.

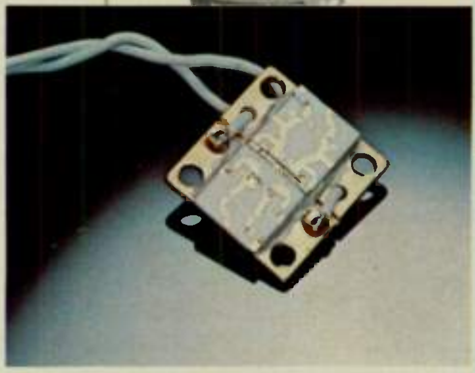
5TH INT'L CONFERENCE ON INFRARED AND MILLIMETER WAVES OCT. 6-10, 1980 **Call for Papers.** Sponsors: Int'l. Union of Radio Science, Physikalisches Institut. Place: Würzburg, Fed. Rep. Germany. Submit 35-word abstract by June 1, 1980; deadline for conference digest ms. is September 1, 1980, to: Kenneth J. Button, General Chairman, MIT National Magnet Laboratory, 170 Albany St., Cambridge, MA 02139. Tel: (617) 253-5561.

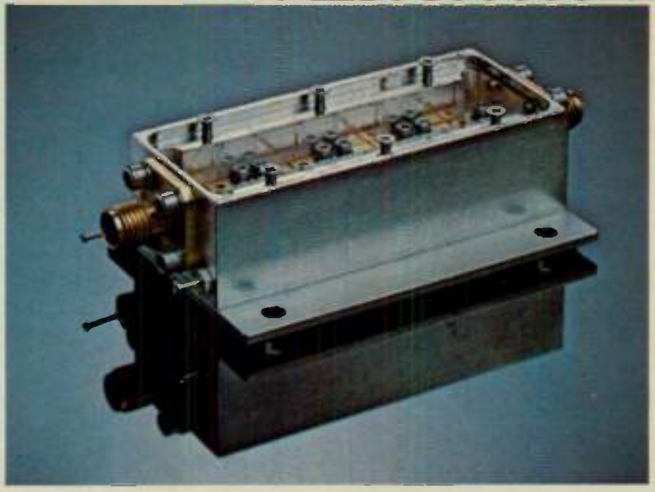
MILITARY MICROWAVES '80 CONFERENCE AND EXHIBITION OCT. 22-24, 1980 Sponsor: Microwave Exhibitions and Publishers Ltd. Place: Cunard Int'l. Hotel, London. Topics: Military applications of microwave engineering, including defense systems, communications, EW, radar, guided weapons and RPVs, mm-wave systems, and technology. Contact: R. C. Marriott, Managing Director MEPL, Kent TN13 1JG. Tel: (0732) 59533/4. Telex: 95604 YNLTD G.



OUR 20th YEAR

EXPERIENCE COUNTS





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Part Number	Frequency Range (GHz)	Noise Figure (db max.)	Small Signal Gain (db min.)	Power Output (dbm min.)	
8000-1300	2.0- 4.0	3.5	25	+10	
8010-1610	2.0- 6.0	4.0	25	+10	
8010-1600	2.0- 8.0	4.5	25	+10	
8000-1600	4.0- 8.0	4.5	25	+10	
8010-1210	8.0-18.0	8.0	25	+10	
8010-1220	7.0-18.0	8.0	25	+10	
8010-1900	6.0-16.0	7.0	25	+10	
8010-1232	6.0-17.0	10.0	18	+ 8	*stock item
8010-1230	6.0-18.0	8.0	25	+10	
8000-1200	12.0-18.0	8.0	25	+10	
8000-1209	12.0-18.0	10.0	18	+ [illegible]	*stock item

Standard octave bandwidth and narrow bandwidth amplifiers are also available.

*These units are available for delivery 1 week ARO (U.S.A.) subject to prior sale.

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(813) 884-1411 Telex: 52-827.

A PREVIEW OF MTT-S IN WASHINGTON

Dr. Larry Whicker, chairman of the Steering Committee for the 1980 MTT-S Symposium, previews the May 1980 meeting which promises to be the largest in the Society's history. Workshop sessions on May 26 and 27, the days preceding the formal Symposium, are identified. The extensive social program is described and a condensed version of program for the three-day Symposium is shown.

MEASURING DIGITAL RADIOS WITH A SPECTRUM ANALYZER

In a comprehensive paper, the authors review digital radio principles and the regulatory requirements for commercial installations and proceed to a discussion of the techniques for measuring FCC-specified as well as other digital radio characteristics. Precautions applicable to the use of spectrum analyzers for obtaining accurate measurement results are discussed. In this section of the paper, the importance of proper input level, resolution bandwidth selection, on-screen amplitude display and analyzer calibration are emphasized. Finally, a section titled "Practical Hints for FCC Measurements" discusses and illustrates those tasks in detail.

THE MICROWAVE POWER TUBE CONFERENCE

The fourth in the Microwave Power Tube Conference series is scheduled for May 1980 in Monterey, CA. Since 1976, these conferences have provided a forum for communication among tube companies, system manufacturers and the Department of Defense which has raised the level of mutual understanding among these groups considerably. Since there are no Proceedings published, free expression is encouraged and, frequently, incomplete work and unsolved problems are discussed at length. The preview provides a broad outline of tne program. In the general area, various sessions will address business climate, technology directions and industry interfaces with DoD and systems suppliers. In addition, reliability, fast-wave devices, applications and materials and processes will be covered in the technical portion of the program. A strong interest in the Conference is anticipated and a summary of the meeting will be published in the July issue of the Journal.

CRIMPED COAX PROVIDES BROADBAND PULSE COMPRESSION

Channelized and pulse-compression receivers for electronic warfare systems depend heavily upon acoustic wave devices and other similar devices to provide the filter technology which they need. The article describes a study which addresses the feasibility of fabricating broadband microwave dispersive delay lines using conventional semi-rigid coaxial cable in a reflective mode. Procedures for the introduction of discontinuities into the coaxial line are discussed and the analysis of these discontinuities is shown. The design, construction and test of an experimental dispersive line for a 5 to 7 GHz baseband and a second with a broad baseband spectrum of 4.2 to 7.8 GHz are described. Special design considerations for receivers which are to incorporate the coaxial dispersive lines are noted.

Sum Up

INSTRUMENTATION BUYERS GUIDE

The second of our issue theme-related buyers guides list suppliers of microwave instrumentation beginning on page 73. Inquiries against the Reader Service numbers shown in the table will be forwarded to a number of suppliers for reply.

Howard Ellowitz

Workshops & Courses

GWU SHORT COURSES

Sponsor: George Washington U Cont. Engrg. Ed. Program
Place: GWU Library, 2130 H St., N.W., Washington, DC.
Titles and Dates: Applied ECM — No. 651, May 5-9, 1980, Digital Trans. Systems Engrg.—No. 535, June 16-20, 1980
Description and Fee: No. 651 — review of electronic countermeasures field as applied to microwave, radar-controlled, weapon systems. Fee per person, $625. No. 535 — advantages of digital vs analog transmission. Fee, $650.
Contact: Director, Cont. Engrg. Ed., GWU, Washington, DC 20052. Tel: (202) 676-6106; Toll Free: (800) 424-9773.

SHORT COURSE ON RADAR REFLECTIVITY MEASUREMENT TECHNIQUES

Sponsor: Georgia Institute of Tech. Engrg. Experiment Station
Date: June 3-5, 1980
Place: GIT, Atlanta, GA
Subject: collection and interpretation of radar reflectivity data.
Contact: Director, Dept. of Cont. Ed., GIT, Atlanta, GA 30332
Tel: (404) 894-2400.

1980 ENGINEERING SUMMER CONFERENCE

Sponsor: U of Michigan, College of Engrg., Cont. Engrg. Ed.
Date: June 23-27, 1980
Topic: Microwave Sensing Technology
Fee: $475
Contact: Engineering Summer Conf. 800 Chrysler Cent, N. Campus The U. of Michigan Ann Arbor, MI 48109

SPREAD SPECTRUM COMMUNICATIONS SEMINAR

Sponsor: Hellman Associates
Dates: August 11-13, Los Angeles, CA Sept. 22-24, Palo Alto, CA
Lecturer: Dr. David Nicholson, Tech. Asst. of V.P., E-Systems, MELPAR Div.
Description: jamming and anti-jamming strategies.
Fee: $595 per person.
Contact: Hellman Associates, 299 S. California Ave., Palo Alto, CA 94306. Tel: (415) 328-4091.

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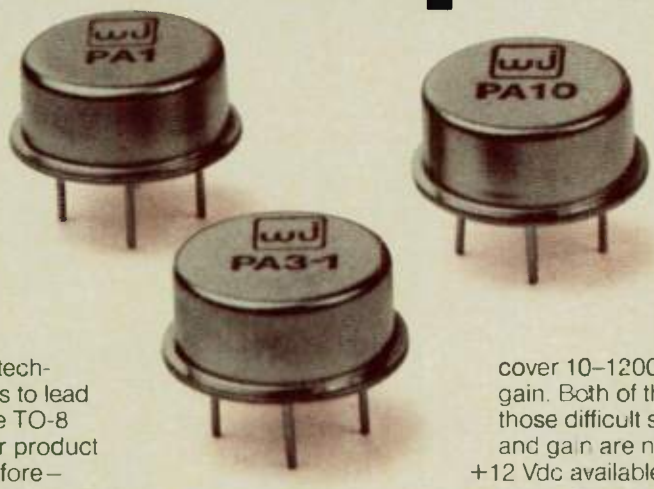

Using state-of-the-art RF amplifier technology, Watkins-Johnson continues to lead the industry with high-performance TO-8 amplifiers. The latest editions to our product line offer more power than ever before—up to almost one-half of a watt with the WJ-PA1!

The WJ-PA1 and WJ-PA2 also offer high-efficiency outputs for applications requiring a minimum current drain with maximum output power. Both models draw typically only 58 mA at +15 Vdc, while providing +20 dBm at the 1 dB compression point. For battery-operated radios or missile applications, more efficient amplifiers cannot be found.

The WJ-PA3 and WJ-PA3-1 cover the full 5–500 MHz band and provide 14 dB of gain with outputs in excess of +20 dBm. The WJ-PA10 and WJ-PA12 cover 10–1200 MHz and provide 9 dB of gain. Both of these are great amplifiers for those difficult situations where both power and gain are needed. If you only have +12 Vdc available, then take a look at the WJ-PA3, WJ-PA10-1 and WJ-PA12-1. You'll receive better than +20 dBm from all of them. And, if your application calls for a hi-rel amplifier, we've still got you covered. These new models are all available with our standard "S" series screening to MIL-STD-883B. We will also be happy to discuss any additional screening which your application might require.

To find out more about these powerful TO-8 amplifiers, contact the Watkins-Johnson Field Sales Office in your area or phone Amplifier Applications Engineering in Palo Alto, California at (415) 493-4141, ext. 2117.

Then you NEED Watkins-Johnson!

Model	Frequency Range (MHz)	Gain (dB) Typ	Gain (dB) −54/+85°C	Noise Figure (dB) Typ	Noise Figure (dB) −54/+85°C	Output Power (dBm[1]) Typ	Output Power (dBm[1]) −54/+85°C	Vdc Nom.	DC Current, mA Typ
WJ-PA1	10–100	13	11.5	8.0	10.0	+22.5 (+26.5)	+20.5 (+24.5)	+15 (+24)	58 (95)
WJ-PA2	10–300	11	9.0	8.0	10.0	+21.0 (+25.0)	+19.0 (+23.5)	+15 (+24)	58 (95)
WJ-PA3	5–500	14.5	13.0	7.5	9.5	+23.0 (+24.5)	+20.0 (+22.5)	+12 (+15)	98 (130)
WJ-PA3-1	5–500	14.5	13.0	7.0	9.0	+22.5 (+25.0)	+20.0 (+22.5)	+15 (+20)	91 (132)
WJ-PA10	10–1000	10.0	9.0	8.5	10.0	+22.5 (+24.5)	+20.0 (+23.0)	+15 (+20)	95 (130)
WJ-PA10-1	10–1000	10.0	9.0	8.5	10.0	+22.00	+20.00	+12	98
WJ-PA12	10–1200	9.5	8.5	8.5	10.0	+22.5 (+24.5)	+19.5 (+22.5)	+15 (+20)	95 (130)
WJ-PA12-1	10–1200	9.0	8.0	8.5	10.0	+22.0	+19.5	+12	97

Note 1. At 1 dB compression point. For WJ-PA3-1, WJ-PA10, and WJ-PA12 at 20 Vdc operation and WJ-PA3 at 15 Vdc operation, the maximum case temperature is +71°C.

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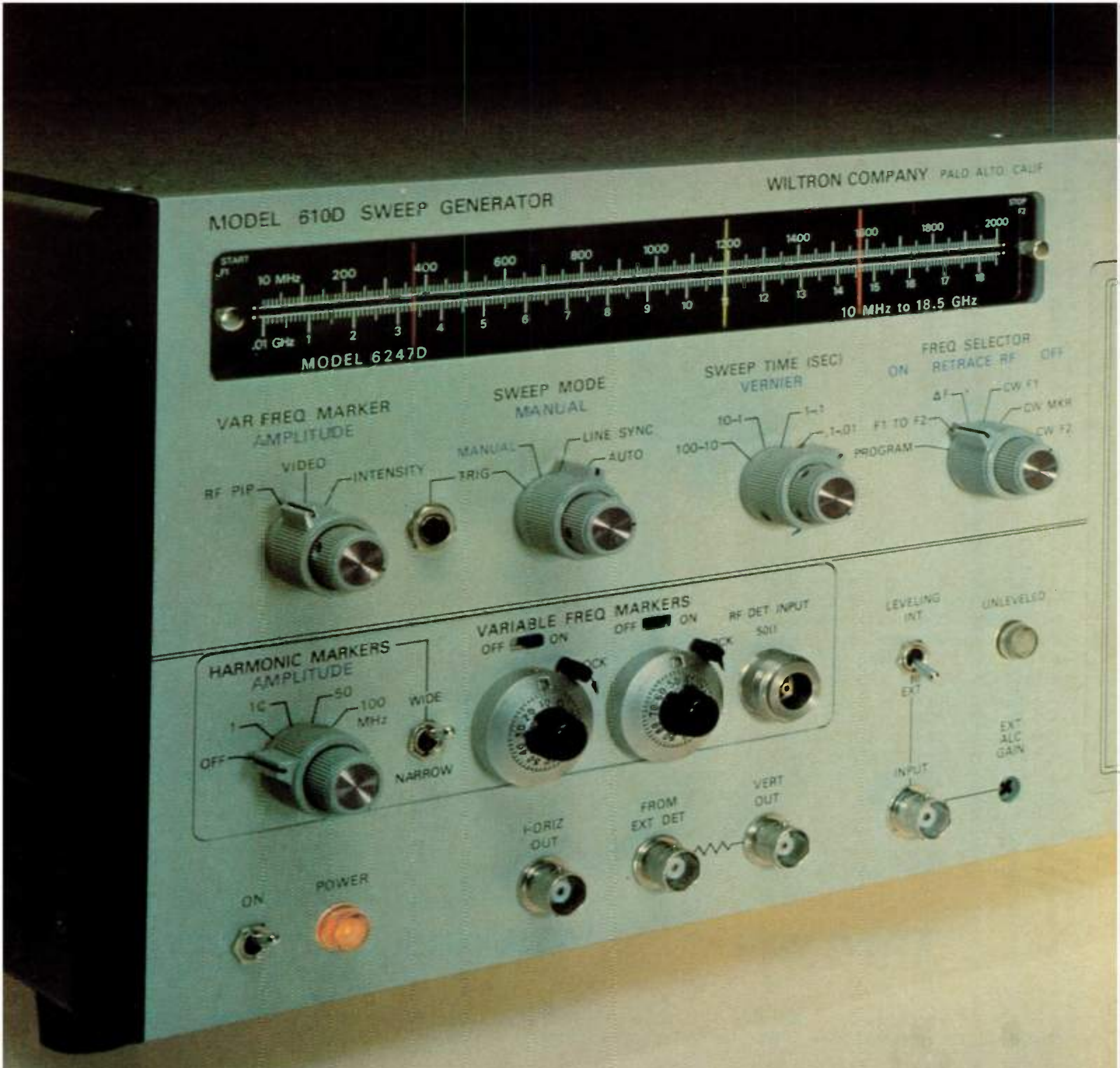

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You build the Wiltron 610D!

Simply, the best sweeper for automatic testing.

Cover the 10 MHz to 18.5 GHz range in one continuous sweep. No need to stop and change plug-ins or external components. One test set-up using Wiltron's new 10 MHz to 18 GHz SWR Autotester is all you need.

It's clearly the superb system for broadband and high directivity (40 dB) SWR measurements.

Plug-ins for cost effectiveness.

There are 26 RF solid state plug-ins covering single, dual and multiple bands—the only complete line of sequentially swept units. Pick the exact unit you need. All units feature outstanding spectral purity, flatness, and the latest YIG tuned oscillators. The 6247D plug-in sweeps from 10 MHz to 18.5 GHz in one continuous sweep. Others go up to 40 GHz.

Advanced features all the way.

Field proven reliability is so good Wiltron gives you a two-year warranty on the oscillator. The 610D features a selectable Intensity, Video or RF marker. Phase lock to an external standard for crystal controlled accuracy. All plug-ins feature slope control to compensate for cables and connectors. Main frame weighs 18 lbs., plug-ins, 10 lbs. And it's compatible with our new 560 Scalar Network Analyzer as well as the other fellow's. Options include crystal controlled frequency markers, preset frequencies, additional markers and GPIB (IEEE 488-1975).

Excellent service, free field seminars.

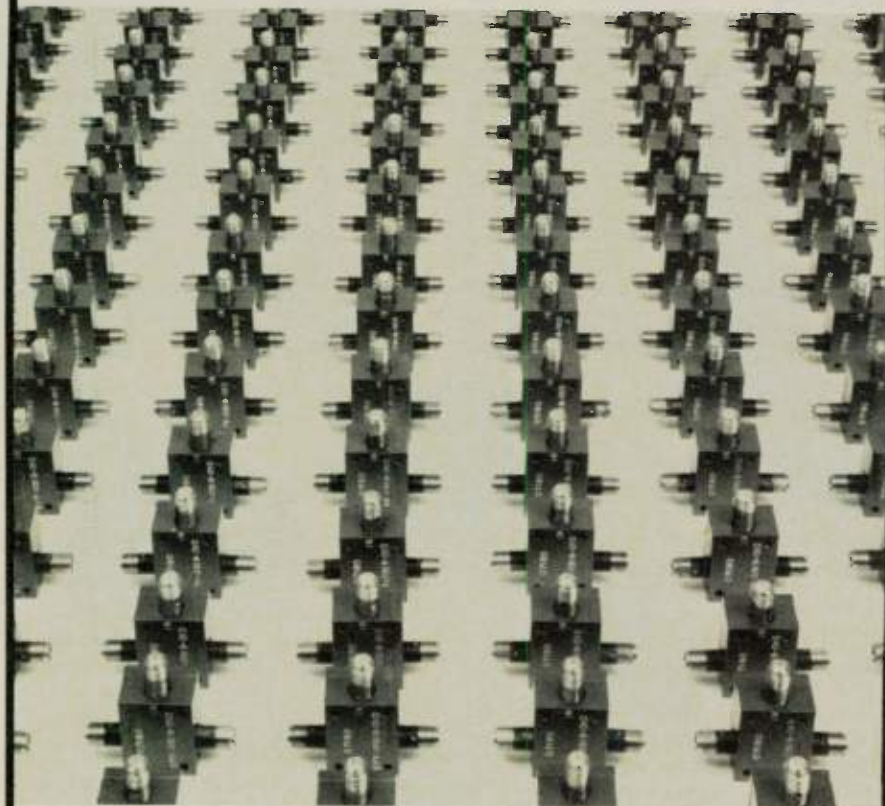
No one provides faster service than Wiltron. If you're in a spot, we'll help you with a loaner. So you can get the most from your Wiltron 610D, we offer user seminars and technical notes.

Isn't it time for the complete sweeper?

For an early demo or literature, call us today. Phone Walt Baxter at (415) 969-6500, or write Wiltron, 825 East Middlefield Road, Mountain View, CA 94043.

WILTRON

The first FP2562 was good. The 1,000th was even better.



But then again, that's no surprise to our customers who expect and receive continued value engineering throughout an entire project.

That's why Rockwell's International Collins Transmission System Division brought us their power divider project. They needed a new communication band power divider which promotes high isolation in a small package.

We gave them the FP2562, a two-way power divider which typically offers greater than 30 dB isolation between outputs from 5.94-to-6.41 GHz, a typical insertion loss of 0.1 dB and typical VSWR of 1.15/1 maximum — all in a unit smaller than 1 cubic inch.

We worked along with Rockwell's engineers on the project to produce what they needed, and they liked what we delivered. So they ordered 1,000 more — each with the same or better high performance and reliability as the first, and at a surprisingly low cost.

If you need microwave components, systems or subsystems — whether its one or thousands, call Sage. We'll work on your project with you. And you'll have the full benefit of our 25 years of experience.

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There are two ways to conserve wildlife.

Zoos are one—but they're mainly for the enjoyment and education of people. The other way: natural habitat that provides what all animals need to survive—food, water, cover and a place to raise their young.

But does it have to be either/or?

In too many places the chance for a fair balance between human needs and wildlife habitat is being recklessly destroyed by chainsaws and bulldozers.

As part of its drive to protect habitats without stifling necessary development, the National Wildlife Federation recently acquired a 2,765-acre tract in northern California's Shasta Valley. The new Lava Lakes Wildlife Area and Nature Center provides habitat for a wide range of species—mammals, birds, waterfowl, fish, reptiles, amphibians. *That's* the way to conserve wildlife.

To help, write Department 404, National Wildlife Federation, 1412 16th Street, N.W., Washington, D.C. 20036.

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cut Avenue with convenient bus/rail service to the center city area. The National Zoo is a short walk from the Conference site. Further details concerning the many attractions will be provided at the Conference.

The hotels require that reservations with one night's deposit be received by April 25, 1980. Since the number of rooms available at the special conference rate is finite, it is advisable to make reservations as soon as possible.

I gratefully appreciate the contributions of the Steering Committee members and Chairmen listed on these pages. Their continuing support will make the 1980 International Microwave Symposium a truly memorable one.

The 1980 IEEE/MTT-S Social Program

Annual Banquet, A Complete Evening's Entertainment

The annual banquet will be held Thursday, May 29, from 7:00-10:30 P.M. and it immediately follows the exhibitors cocktail party taking place in the Blue Room of the Shoreham Hotel. The banquet hall will be transformed into an elaborate setting with the use of flowers and decorations

A proper prelude to the dinner will be a concert by the "President's Own" U.S. Marine Band or the U.S. Navy Band. Following the concert will be the Presentation of the Colors by the Joint Armed Forces Color Guard.

Following dinner, entertainment will be provided by Mr. Mark Russell, Washington's renowned political satirist, night club entertainer and TV celebrity. Mark has promised to provide us with a special show for our group which

Mark Russell

will address issues of interest to the microwave community.

In the awards portion of the banquet the Microwave Theory and Techniques Society will honor the following members and present to them the listed prizes and awards:

- Microwave Career Awards:
  Dr. S. B. Cohn
  Dr. W. J. Klein
- Microwave Applications Award:
  Erwin F. Belohoubek
- Microwave Prize:
  E. R. Carlson,
  M. V. Schneider, and
  T. F. McMaster

An award will be presented for the paper:

"Subharmonically Pumped Millimeter-Wave Mixers,"
MTT Trans., Oct. 1978

Newly elected IEEE Fellows who are members of MTT will be recognized and additionally, IEEE President, Leo Young, will present the IEEE/USAB Award for Engineering Professionalism to Bruno Weinschel, Past USAB Chairman.

Champagne and Dessert Tour, "Sparkling" Introduction to the Nation's Capitol

This tour on Wednesday, May 28, from 8:00-10:30 P.M. will provide a magnificent overview of the Washington area. The tour should interest both conference attendees, their spouses and older children.

An experienced guide will narrate a leisurely tour of Washington monuments and government buildings, beautifully lighted for nighttime viewing. On our bus we will tour the city and see an overview of the Capitol while enjoying champagne and French pastries served by a waiter.

We will pass the Capitol, the White House, the Washington Monument, galleries and museums and the now famous Watergate complex.

Special stops will be made at the Lincoln Memorial, the Jefferson Memorial, the Iwo Jima Memorial and the John F. Kennedy Center for the Performing Arts—all more impressive and inspiring under illumination.

Spouses Program

The spouse's program will begin each day (Wednesday through Friday) at 8:30 A.M. Tour No. 1, on May 28 from 9:30 A.M. - 3:30 P.M. will explore Georgetown Houses. Tour No. 2, on May 29 from 9:30 A.M. - 3:00 P.M. will cover Capitol Hill and Tour No. 3, on May 30 from 9:00 A.M. to 12:00 noon will visit Mt. Vernon and Alexandria.

MTT Technical Program

Program will feature five days of state-of-the-art workshops and sessions, including:

(continued on page 26)

SYMPOSIUM SCHEDULE

MONDAY, MAY 26, 1980 (MEMORIAL DAY OBSERVED)

EXECUTIVE ROOM	FORUM ROOM	PALLADIUM ROOM
GIGABIT LOGIC FOR MICROWAVE SYSTEMS 9 00 5 00 P M	AUTOMATIC RADIO FREQUENCY TECHNIQUES 9 00 5 00 P M	MILLIMETER WAVE DEVICES USING GYROTROPIC MEDIA 9 00 5 00 P M

TUESDAY, MAY 27, 1980

AMBASSADOR ROOM	FORUM ROOM	EMPIRE ROOM
ELECTROMAGNETIC DOSIMETRIC IMAGERY 9 00 5 00 P M	AUTOMATIC RADIO FREQUENCY TECHNIQUES 9 00 5 00 P.M	MONOLITHIC MICROWAVE ANALOG IC's 9 00 5 00 P M

WEDNESDAY, MAY 28, 1980

AMBASSADOR ROOM	DIPLOMAT ROOM	FORUM ROOM	TUDOR ROOM	EXHIBIT HALL
A OPENING SESSION 8 30–10 00 A.M.				1980 MICROWAVE EXHIBITION 9 00–6 00 P.M.
B MICROWAVE POWER TRANSISTOR 10 30 noon	C MILLIMETER RECEIVERS AND COMPONENTS 10 30 noon	D MICROWAVE ACOUSTICS AND MAGNETOSTATIC WAVES 10 30–12 40 P.M	E MICROWAVE ENGINEERING FOR EXPORT 10 30 – noon	
F MILLIMETER WAVE POWER GENERATION 2 00–5 00 P.M.	G HIGH POWER DEVICES AND TECHNIQUES 2 00 5 10 P M.	H MICROWAVE SYSTEMS 1 1 30–5 00 P.M.	J MICROWAVE FILTERS AND MULTIPLEXERS 2:00–5 10 P.M.	
	PANEL SESSION ON MILLIMETER WAVE IC's 8:00–10:00 P.M.	PANEL SESSION ON THE SOLAR POWER SATELLITE (SPS) SYSTEM 8 00–10 00 P.M.		

CHAMPAGNE AND DESSERT TOUR OF THE NATION's CAPITOL – 8 00–10 30 P M

THURSDAY, MAY 29, 1980

AMBASSADOR ROOM	DIPLOMAT ROOM	FORUM ROOM	TUDOR ROOM	EXHIBIT HALL
K HIGH POWER SOLID-STATE CIRCUITS 8 30–11 40 A.M.	L SATELLITE COMMUNICATION SYSTEMS IN EUROPE 8 30–noon	M GUIDES AND COMPONENTS 8 30–11 40 A M	N FERRITE APPLICATIONS 8 30–11 30 A M.	1980 MICROWAVE EXHIBITION 9 00–5 00 P.M.
O MILLIMETER WAVE IC's 1 30–4 40 P.M	P SATELLITE BROADCASTING IN JAPAN 1:30–5:00 P M.	Q GUIDED WAVE OPTICS AND INTERACTIONS 1 30–4 30 P.M.	R BIOLOGICAL APPLICATIONS AND EFFECTS 1:30–5 00 P M.	

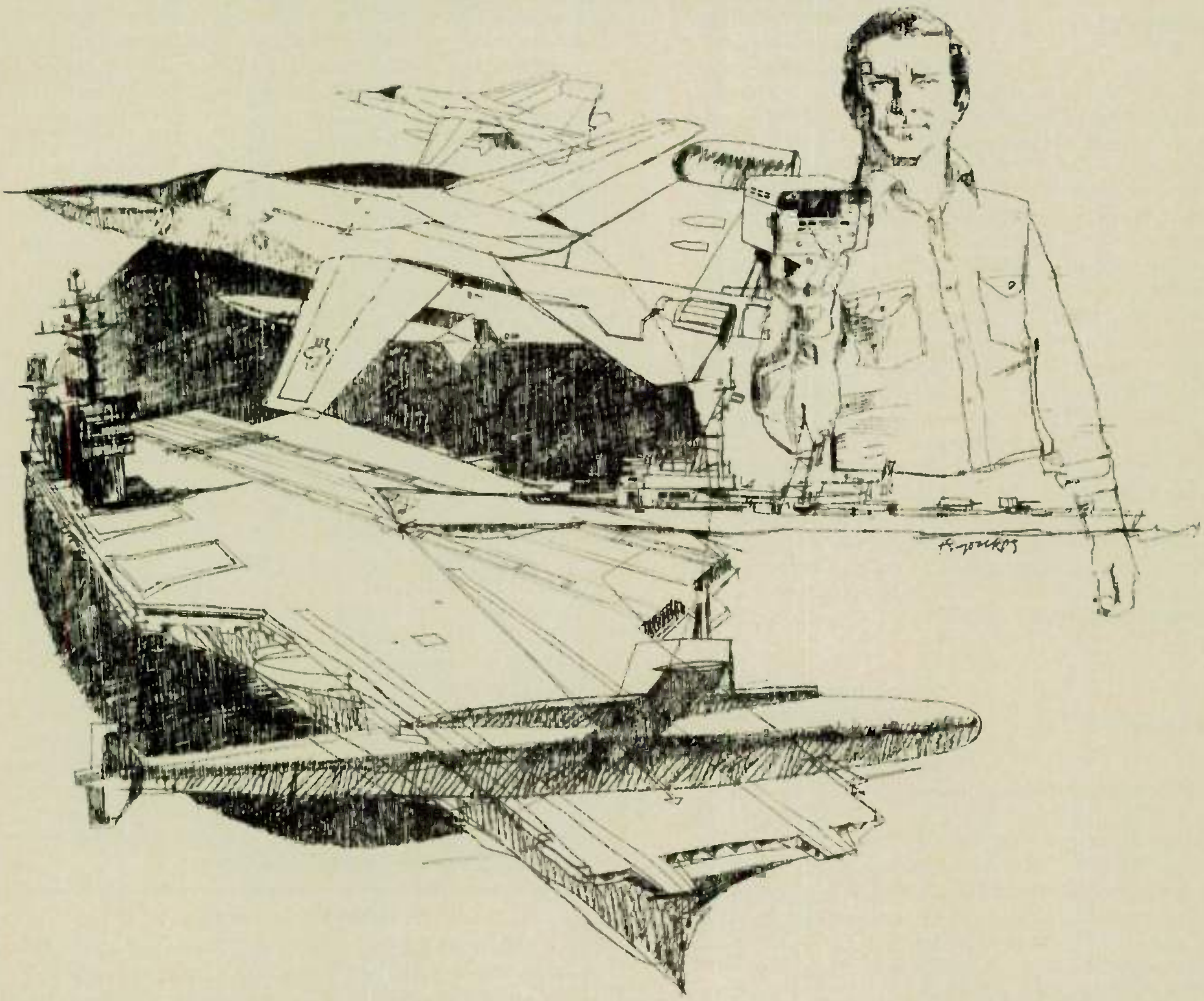
EXHIBITORS' RECEPTION, PALLADIUM ROOM – 5 15 – 6 45 P.M.

1980 IEEE/MTT S ANNUAL BANQUET, BLUE ROOM – 7:00 – 10 30 P.M.

FRIDAY, MAY 30, 1980

AMBASSADOR ROOM	DIPLOMAT ROOM	FORUM ROOM	TUDOR ROOM	EXHIBIT HALL
S GaAs FET CIRCUITS 8 30 noon	T COMPUTER-AIDED TECHNIQUES 8:30–11:50 A.M.	U MICROWAVE MEASUREMENTS 8 30–11 40 A.M.	V FIELD THEORY 8 30–11 40 A.M	1980 MICROWAVE EXHIBITION 9 00–3 00 P M
W MICROWAVE IC COMPONENTS 1 30–2 50 P.M.	X MICROWAVE SYSTEMS 2 1 30–2 50 P.M.			

SANDERS' NEWEST NOISE FIGURE TEST SET. LIGHTWEIGHT, PORTABLE. COVERS 1 MHz TO 40 GHz.



Our new Model 5440C Test Set is tough. Yet it's light (8 lbs.) and small (5" x 7" x 10.5") for easy storage and carrying.

It measures noise figures from 2 to 35 dB, with typical accuracy of ± 1 dB, and presents the output as a digital readout.

A continuously variable IF preselector allows measurements anywhere in the 10-200 MHz range.

This Test Set can also operate directly from system video or audio output. Result? A more accurate representation of complete system performance.

Four LED indicators alert the operator to improper system setting or procedures.

Another feature. The 1 MHz-18 GHz noise source — a solid state component of the Test Set — can be positioned up to 300 feet from the Set. This way, the entire system can be included in noise figure measurements. Defective components at the front end can be quickly isolated by connecting the noise source before or after the suspect unit or cable.

The Model 5440C Noise Figure Test Set from Sanders. For details and a demonstration, contact Sanders Associates, Inc., Microwave Division, Grenier Field, Manchester, NH 03103. Telephone: 603-669-4615, X-447. TWX: 710-220-1846.

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Around the Circuit

PERSONNEL

Cablewave Systems, Inc. appointed **Henry Pessah** as Manager of Engineering. . . **Gilbert F. Johnson** joins California Microwave, Inc. as Executive V.P. – Operations. . .Comtech Telecommunications Corp. announced the appointment of **George Birutis** as President of CTC's subsidiary, Comtech Antenna Corp. . .**Harry B. Sefton Jr.** was named Engineering Mgr. for the American Nucleonics Corp., Westlake Village, CA. . .**Michael Mulcay** joins TerraCom, a division of Loral Corporation, as V.P. of Marketing. . .Scientific-Atlanta, Inc. promoted **Marvin Shoemake** to General Manager for the antenna products division. . .Materials Research Corporation has added two key members to its management ranks. **Dr. Charles Ristagno** joins the company as Manager of Precious Metals, Advanced Materials Div. and **Vijay Borase** advances from Operations Mgr. to General Mgr., Ceramic Substrates Div. . .**David C. DeGree** has been appointed Dir. of Marketing and Product Development and **Richard Betz** was named V.P. and Gen. Mgr. for the Laminates Division of Keene Corporation. . .At Systron-Donner's Instrument Division, **Harry H. Hollington** has been appointed Director of Operations. . .Valtec Corporation elected Chairman of the Board of Directors and appointed **James R. Kanely** as Pres. and Chief Operating Officer.

CONTRACTS

Harris Corporation's Government Electronic Systems Division received a $7.3M contract for a classified electronic communication system from the US Army's Maryland Procurement Office. . .US Army Satellite Communications Agency in Fort Monmouth, NJ awarded a $4.37M fixed price contract to **Ford Aerospace & Communications Corp.'s Western Development Labs** for a satellite earth terminal with a 60-foot diameter antenna dish to be used by the Gov't of Australia. . . **Norden Systems, a United Technology** subsidiary, received a $30M contract award to begin production of the AN/APS-130 airborne navigational radar system. Initial funding of $10.7M was received from the Naval Air Systems Command (NASC). Another $60M contract was awarded by the **Sylvania Systems Group of GTE** to **Norden Systems** to supply the data processing equipment for the C^3 systems of the MX missile system. GTE was named as prime contractor for the M program in a $325M grant awarded by the Air Force Ballistic Missile Office, Norton AFB, CA. . .NASC awarded a $1.7M contract to **Xerox Electro-Optical Systems** for helicopter infrared countermeasure system. . .**Electromagnetic Sciences, Inc.** was granted a contract by **Motorola's Government Electronics Div.** to develop electronic equipment for use on the US Army's Stand-Off Target Acquisition System. . .**Alpha Ind.** received a contract over $1.4M from **Loral Electronic Systems** for ECM system devices.

INDUSTRY NEWS

Satellite Business Systems' FCC Order of February 8, 1977 was affirmed en banc by the US Court of Appeals in Washington, DC. . .**RCA** announced plans to develop a military Modular Airborne Search and Track Radar (MASTR). . .**General Telephone & Electronics** announced multimillion dollar expansion plans for its communication research facilities in Phoenix, AZ and St. Petersburg, FL. . .**Omni Spectra, Inc.** opened a 10,000 square-foot manufacturing facility in Lawrence, MA to expand its **Microwave Connector Div.** in Waltham, MA. . . .**Anderson Labs, Inc.** (Bloomfield, CT) has formed a new company, **Signal Technologies, Ltd.** in a joint venture with **Plessey Co. Ltd.** (UK). **Signal Technologies** is a SAW device manufacturer located in Swindon, England. . . .**Weinschel Engineering Co., Inc.** has begun construction of a 10,000-square-foot electronic calibration equipment manufacturing facility at the Airport Industrial Park near Frederick Municipal Airport, MD. . .**M/A-COM, Inc.** and **Omni Spectra, Inc.** announced they have entered into a definitive Plan and Agreement of Merger pursuant to which *Omni Spectra* will become a *M/A-COM* company.

FINANCIAL NEWS

Adams-Russell reported first quarter results for the period ended December 30, 1979 of sales of $7.98M and net income of $529K on earnings per share of 29¢. This compares with 1978 quarterly sales of $6.39M and net income of $366K on earnings per share of 21¢. . . **Electromagnetic Sciences, Inc.** announced yearly net earnings for 1979 of $262K or 31¢ per share and sales of $5.05M. This compares with 1978 earnings of $214K or 6 per share and sales of $3.77M. . .**Omni Spectra, Inc.** reported first quarter results for the period ended December 29, 1979 of net income of $205K or 8¢ per share and sales of $6.98M. For the same period of 1978, net income was $126K or 5¢ per share and sales totalled $6.21M. . .**Raymond Industries Inc.** announced sales of $34.6M and net earnings of $1.49M or $1.42 per share for the year ended December 31, 1979. During the end of fiscal year 1978, sales were $31.5M and net income was $1.17M of $1.12 per share. . .**California Microwave, Inc.** reported second quarter results for the period ended December 31, 1979 or net income of $104K, or 5¢ per share on sales of $9.0M. During the same quarter last year, net income was $584K, or 29¢ per share (adjusted for 50% stock dividend) on sales of $9.1M.

ROBERT L. RIDDLE

Robert L. Riddle, co-founder, President and Chairman of the Board of Directors of Locus, Inc., died February 19, 1980, a victim of leukemia.

Mr. Riddle was born October 28, 1922 in Smithland, IA. He received a B.S. in 1949 and an M.S. in 1951, both in E.E. from the State University of Iowa. From 1951-1956, he was associated with BTL, RCA, and the Univ. of Iowa on the application of solid state devices for spaceborne radiation experiments. From 1951 to 1963, Mr. Riddle was assistant professor of E.E. at Penn State. In 1963, he became V.P. of HRB-Singer, Inc. and since 1968 served as President of Locus, Inc., State College, PA.

He was an Air Force veteran of World War II and the Korean conflict. Mr. Riddle was a member of Tau Beta Pi, Eta Kappa Nu, Sigma Xi and IEEE, Amer. Soc. of Engineers, Assoc. of Old Crows, the Air Force Assoc. and the Assoc. of the US Army.

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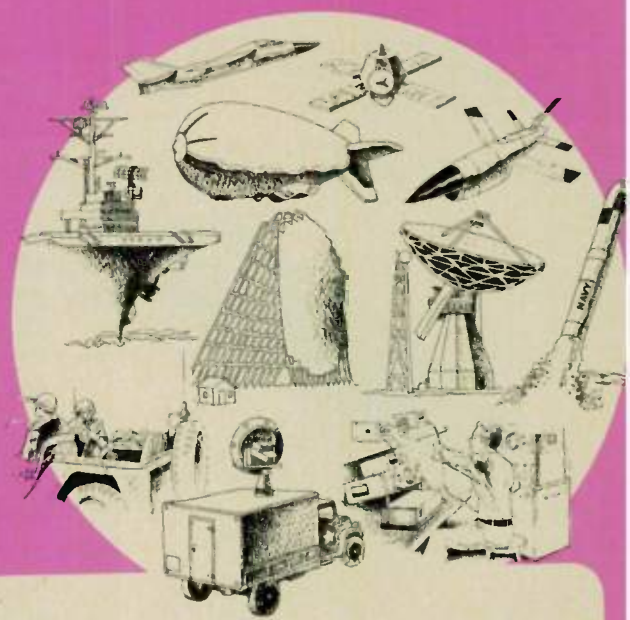

RF/Microwave products for telecommunications, defense electronics, laboratory instrumentation and test—frequencies from less than 1 MHz up to 8.4 GHz—power outputs from less than 1 watt up to several kilowatts—all solid state—that's the world of MPD. It's a world that's getting wider every day, with a constantly growing spectrum of commercial, industrial and military applications:

Space Satellite Amplifiers
- Class A linear or Class C
- 1500 to 2300 MHz
- "Space-Qualified" transmitters

Satellite Ground Stations
- GaAs FET power amplifiers, up to 8.4 GHz
- 1 and 5 MHz distribution amplifiers, up to 26 outputs
- FM carrier baseband video distribution amplifiers
- IF amplifiers, 70, 700 and 1100 MHz

Terrestrial Microwave Amplifiers
- Microwave LOS, 100 watts
- High power troposcatter, L-band, 1000 watts
- Microwave and UHF radio relay, 1000 watts

Broadcast
- UHF/VHF color TV transmitters, up to 1.5 KW peak synch
- Airborne TV visual/sound power amplifiers
- FCC type-accepted driver amplifiers
- UHF TV internal 3-tone amplifiers

Avionics
- FAA and MIL TACAN transmitter systems— power amplifiers, modulators, synthesizers, power supplies
- L-band digital transmitters (JTIDS)
- Data link transmitters
- Up/down converters
- Airborne pulse amplifiers

Missile Systems
- Command/destruct transmitters
- Guided weapon data link amplifiers
- Military drone transmitters

Radar Amplifiers
- L-band transmitters
- S-band pulse drivers for 3-D radar
- Shipboard drivers for AN/SPS-48 radars

Electronic Warfare
- Communication jammers
- Class A linear power amplifiers
- Linear AB wideband jammers
- Jamming simulators

Military Communications Amplifiers
- Long-pulse data links
- Communication command links
- UHF transceiver amplifiers/modulators
- MIL RF power boosters (ECCM)

Laboratory Instrumentation/Test
- Class A linear amplifiers
- RFI/EMI test amplifiers
- Power meter calibration systems
- Commercial and MIL power supplies— high efficiency, compact packaging, up to 6000 watts
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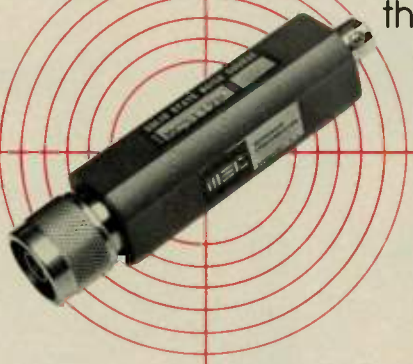

Solid-State Noise Sources that Generate Broadband Noise from 1 MHz to 40 GHz.

Features

- Long term stability
- Extreme temperature stability
- Fast switching capability
- Wide bandwidths.

Instrument Noise Sources

MODEL NUMBER	FREQUENCY RANGE (MHz)	ENR	FLATNESS	POWER REQ. MAX.
WIDEBAND COAXIAL				
MC 1000	10 - 1500	35.0 dB	±0.50 dB	+28V, 10mA
MC 1100	10 - 1500	15.5 dB	±0.50 dB	+28V, 10mA
MC 1040	10 - 4000	25.5 dB	±0.50 dB	+28V, 15mA
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				
MC 5012	1000 - 2000	30.0 dB	±0.50 dB	+28V, 15mA
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
MC 51218	12400 - 18000	28.0 dB	±0.50 dB	+28V, 15mA
WAVE GUIDE BAND				
MC 5046W	3950 - 5850	15.5 dB	±0.50 dB	+28V, 15mA
MC 5068W	5850 - 8200	15.5 dB	±0.50 dB	+28V, 15mA
MC 5812W	8200 - 12400	15.5 dB	±0.50 dB	+28V, 15mA
MC 51218W	12400 - 18000	15.0 dB	±0.50 dB	+28V, 15mA
MC 51826W	18000 - 26500	25.0 dB	±2.00 dB	+28V, 20mA
MC 52640W	26500 - 40000	23.0 dB	±3.00 dB	+28V, 20mA
ADAPTORS				
MC 601A	ADAPTOR FOR HP 340 and HP 342 Series Automatic Noise Figure Meters			
MC 602	ADAPTOR FOR AIL Model 74 Noise Figure Meter			

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A Preview of the Microwave Power Tube Conference

J. E. GRANT
Hughes Aircraft Company
Torrance, CA

The fourth Microwave Power Tube Conference will be held May 12-14, 1980, at the Naval Postgraduate School in Monterey, CA. The theme for this year's Conference is "The Decade of the '80s." The principal objective is to provide a forum for open communications among tube companies, systems manufacturers and DoD. Both technical and institutional issues will be discussed. To promote the free expression of views, no Conference proceedings are published and neither photographs nor tape recordings are permitted. Papers include incomplete work, reports on unsolved problems, future requirements, and conclusions to problems raised at past conferences. If the past is any indication, the atmosphere will be casual and the discussion, at times, heated. For the most part, however, the debates are constructive and most attendees come away with a sense of accomplishment.

Broadly, the first day of the Conference will be devoted to discussions of the Tube Industry/DoD interface, the business climate, DoD investment strategy, and a forecast of where the technology is headed. The second and third days will cover technical progress and problems and the Tube Industry/OEM interface.

Perhaps the single most important subject at the First Monterey Conference was reliability. This year's organizing committee hopes to rekindle interest in this subject. A synopsis of the Transmitter Reliability Workshop will be given This workshop was held at NASA's Lewis Center last September to discuss problems associated with spacecraft tube-type transmitters. Reliability papers will also cover radar and ECM applications. Since it is generally recognized that the bulletproof tube is an impractical concept, papers on both tube and transmitter reliability improvements will be presented.

Following the reliability session, there will be four invited papers addressing cost drivers. Two of these papers will cover expendables. Since it makes little sense to plug a low-cost tube into a high-priced expendable transmitter, the transmitter will be considered as well.

Since a large number of fast-wave abstracts were submitted, a separate session will be devoted to this topic. Subjects include gyrotrons, approaches to wide bandwidth at millimeter waves, electron gun problems, and the low frequency gyrotron.

A variety of subjects will be covered in the applications session. Two papers will address new radar techniques and their transmitter requirements. The need for dual mode (10 dB pulse up) radar tubes will be pointed out. Included in this session will be a paper concerning Soviet tube technology and their spectrum trends will be discussed. A comparison between the US and USSR's power versus frequency capabilities will be made.

Given that materials and processes are the major factor in limiting the performance of microwave tubes today, there should be considerable interest in this session at the Conference. Cathode papers oriented towards device applications will be a dominant subject. There is no doubt that the industry is in need of a better cathode. Over the past several years, controlled life tests and sketchy field data reveal that cathode emission degrades more rapidly with time than the original, close-spaced diode work suggested. With the advent of higher frequency operation and inherent electron gun design limitations, the availability of an improved cathode becomes even more important. Included in this session will be a synopsis of the 1980 Tri- Services Cathode Workshop. The synopsis will include cathode technology, problem areas, identification of specific requirements, and recommendations for the future. Papers will address such topics as improvements in the dispenser cathode, performance of gold-magnesium-oxide secondary emitting cathodes, and field emitter arrays. Other subjects to be included are brazed helix technology, thin film deposition of helix attenuators, permanent magnets, high voltage encapsulation, and shelf life considerations.

After several false starts, it now appears that millimeter waves are here to stay. In fact, millimeter waves may soon become a major growth segment of the tube market. Historically, frequency has been steadily pushed upward due either to spectrum overcrowding at lower frequencies or the quest for enhanced systems performance at the higher frequencies. Both DoD and industry planners are actively pursuing next generation systems employing millimeter waves. There is also a steadily increasing availability of solid state sources, passive components, and high power tubes. Because of the growing interest, a millimeter device session will be held covering such topics as novel slow-wave structures, gridded guns, and future requirements.

Finally, the Conference would not be complete without a broad spectrum of papers covering the latest and greatest in microwave devices and technology.

An equally important aspect of the Conference is a frank and honest discussion of the institutional and management problems confronting the industry. The entire first day of the Conference will be devoted to this subject. The morning will cover three broad topics:

- Technology forecast
- Business climate and expectations
- DoD investment strategy

The afternoon session will delve into more specific issues including such topics as the role of AGED, status of the AFTER Program, R&D funding problems, DoD program planning and management, manufacturing methods activities, and material problems.

Predicting the tenor of these management topics at the Conference is difficult. It is expected, however, that the discussions will center on the unpredictable business picture facing the industry and the precarious world economic condition. Our industry is characterized by significant capital investment requirements, relatively low profit return, small production quantities, and high technology and quality requirements. It is not an environment which encourages major investments. This industry characteristic is probably why the number of major tube companies has dwindled from 16 in 1970 to 7 in 1980.

The industry is being subjected to rapidly increasing prices and material shortages. This situation is not expected to ease in the foreseeable future. In the past, the cost of raw materials was an insignificant cost element, but that is no longer true. There is certainly no need to reiterate what gold and silver prices have done recently. There are other raw materials whose prices have escalated at a rapid pace and some that are likely to become extinct altogether. We face a significant challenge of finding either substitutes or technical alternatives. More and more purchased items are priced at time of delivery making it difficult to recover escalating costs in the fixed price environment.

The tube industry is an exciting, challenging field. Significant changes have occurred demanding a concerted effort to find new and different solutions to the problems facing the industry. An open discussion of these problems and, in particular, exploring possible solutions is what the Microwave Tube Conference is all about.

Jeff Grant received an A.B. (Physics) degree from the University of California at Berkeley in 1964 and a M.S. (E.E.) from the University of California at Los Angeles in 1968. As an engineer, he has worked on the design and development of a variety of TWT's at Hughes. Mr. Grant is now Manager of the Microwave Amplifier Product Line at the Electron Dynamics Division of Hughes Aircraft Company. He is also the General Co-Chairman of the 1980 Microwave Power Tube Conference.

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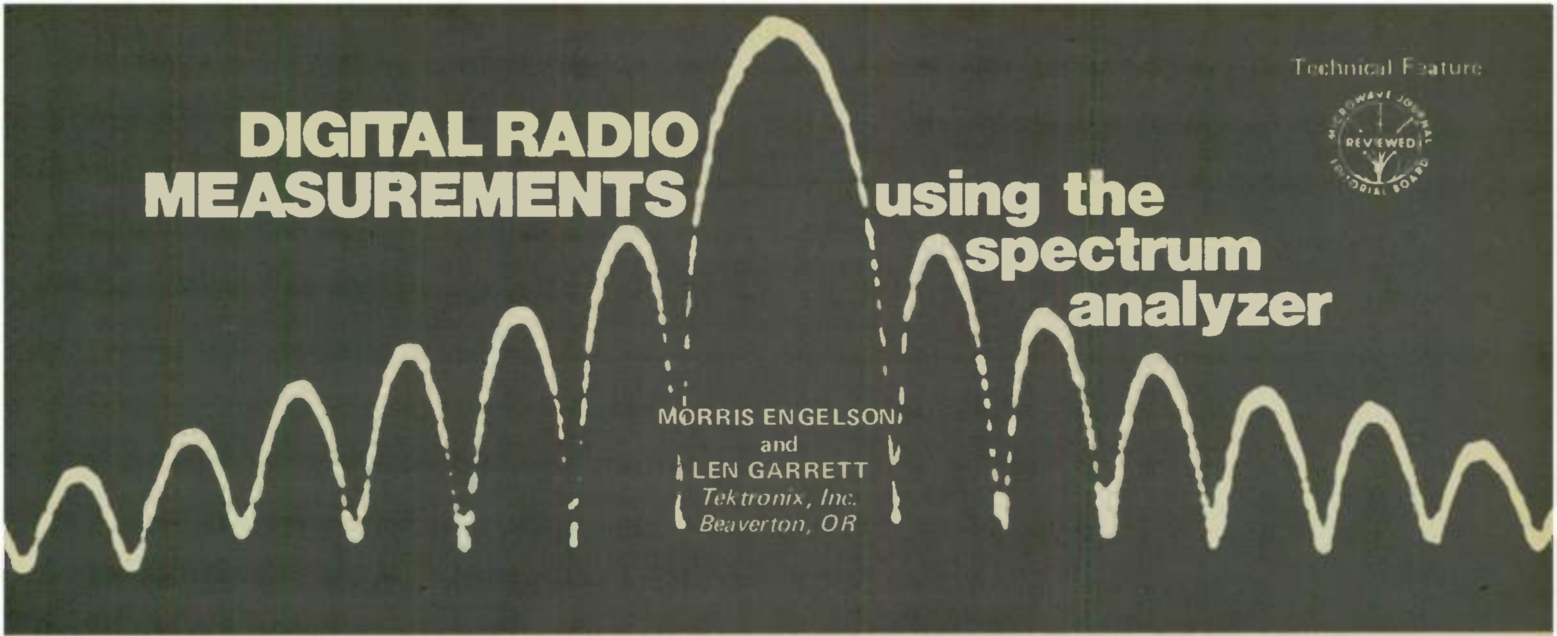

DIGITAL RADIO MEASUREMENTS using the spectrum analyzer

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and
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Beaverton, OR

Digital radio differs from analog radio not in the radio transmission part, but rather in the scheme of modulation. Digital modulation, known as pulse code modulation (PCM), was first described by A. H. Reeves in the 1930s. However, it was not used commercially until the 1960s, when semiconductor technology made the scheme practical. In digital modulation, a signal is sampled in frequency and quantized in amplitude to convert the original analog information to discrete binary symbols. Various bit streams of discrete symbols are interleaved in time, creating a time-division-multiplex (TDM) composite which is the time equivalent of the well known frequency-division multiplex (FDM) for analog modulation.

The desire to go digital derives from two advantages — performance and economics. Digital symbol regeneration is possible, at least in theory, without the usual degradation due to noise and other problems of analog signals. The economic advantages derive from the inherent simplicity of digital switching equipment and from the ability to combine diverse types of signals into one bit stream. The fact that existing transmission media, such as cable or point-to-point microwave, can accept digital transmission without major redesign makes the switch to digital easier.

FORMS OF DIGITAL MODULATION

The binary symbols representing the sampled-quantized original signal can be modulated onto a carrier in various ways, such as phase, frequency, or amplitude. Many variations and combinations of these three primary schemes are possible. The most popular scheme at this time is 8PSK (that is, 8-phase-shift keying), which results in eight phase vectors at 45-degree intervals. These vectors represent three possible binary states (since $2^3 = 8$). Earlier schemes provided fewer binary levels, while future developments are moving toward more levels. The push for more levels is a result of the desire for improved transmission efficiency in the form of bits per hertz of transmitted bandwidth. Sixteen-level systems using AM and PM are now becoming available. It would appear, therefore, that the popular 8PSK system is only a temporary step in digital radio development. Nevertheless, all examples in this note are based on the 8PSK system since the measurement principles, if not the members, remain the same regardless of modulation scheme.

As in FDM, so also in TDM, digital bits are combined in a hierarchical system. The first unit in North America is 24 voice channels consisting of 1.544 Mb/s; for Europe the first combined level consists of 30 voice channels at 2.04 Mb/s. These channels can be combined further so that the fourth level in the USA is a 274.176 Mb/s stream. These numbers result from the basic encoding. In the USA, a 0.3 to 3.4 kHz voice channel is sampled at an 8 kHz rate and encoded into an 8 b signal. Therefore, 24 channels require (24 x 8 =) 192 b. Adding one bit for frame synchronization results in 193 x 8 kHz (sample rate) = 1.544 Mb/s. A sample from each of the 24 channels is contained in a 193-bit frame lasting 124 µs (1/8 kHz).

It should be clear from the above discussion that the digital modulation signal is a complex combination of AM, FM, or PM pulse combinations. Such a signal creates a rather wideband spectrum that must be carefully shaped, adjusted, and monitored in order to meet transmission fidelity and interference criteria and regulatory requirements (the FCC in the USA). The primary measurement instrument for this spectrum is the Spectrum Analyzer.

MEASUREMENT NEEDS

FCC regulations § 21.106 specify that for digital modulation transmission below 15 GHz, "in any 4 kHz band, the center frequency of which is removed from the assigned frequency by more than 50% up to and including 250% of the authorized bandwidth: As specified by the following equation but in no event less than 50 dB.

$$\text{"}A = 35 + 0.8\,(P - 50) + 10 \log B$$

"(Attenuation greater than 80 dB is not required.)"

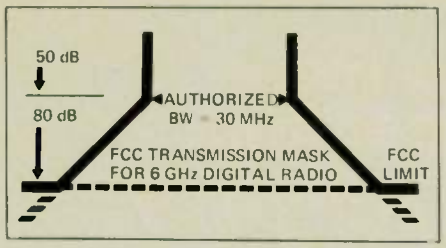

Fig. 1 Graphical representation of FCC transmitted spectrum specifications.

In this equation, A is the attenuation (dB) below the mean output power level, P is the percentage removal from the carrier, and B is the authorized bandwidth in MHz.

A different equation applies for transmission above 15 GHz. Likewise, the FCC specifies authorized bandwidths as 30 MHz at 6 GHz, 40 MHz at 11 GHz, etc.

Applicable specifications are usually given by the radio manufacturer. These specifications can also be found in FCC publications, as indicated above. A graph of the specifications for a 6 GHz radio is shown in **Figure 1**. Two points need to be adjusted and/or verified. The spectrum width 50 dB from the mean transmitted power should not exceed the authorized bandwidth, and the output level should be at least 80 dB down outside the frequencies within the FCC mask.

Besides performing the spectrum occupancy tests, the spectrum analyzer can be used in many other digital radio applications. These applications include checking or adjustment for maximum peak power output, spectrum shape and symmetry, comparison of pre- and post-output filter performance, spurious emissions far from the carrier, amplitude-level variations among several transmissions on the same antenna, interference pick-up antenna alignments, etc.

Before proceeding with a discussion on how to perform these measurements, it may be well to consider the spectrum-analyzer display of the digital radio signal.

SPECTRUM ANALYSIS OF DIGITAL RADIO SIGNALS

The digital radio signal is composed of a large number of individual components multiplexed together. In the aggregate, the output is of a random nature and has a noise-like character. For the measurement bandwidths involved (kHz to MHz), the spectrum analyzer responds as it would to random noise. The spectrum shape depends on the form of modulation used. For PSK, the spectrum shape is determined by the Fourier transform of the bit stream, which, if of ideal rectangular pulses, generates a sinx/x spectrum. The nulls of this spectrum occur at the signalling, or bit, rate. Such a spectrum is shown in **Figure 2**.

Because the signal is noise-like, it follows that noise measurement theory applies. Display power level is proportional to spectrum analyzer resolution bandwidth, changing by 10 log (bandwidth ratio). Smoothing, by use of either video filter or digital averaging, needs to be used since the peak signal level will fluctuate. Absolute power-level measurements call for the usual 2.5 dB log mode correction factor, as well as a knowledge of the random-noise bandwidth rather than just the spectrum-analyzer resolution bandwidth (see **Appendix A**).

As long as the measurement resolution bandwidth is relatively narrow (Less than one tenth) in relation to the spectrum shape to be measured, the displayed spectrum shape will not be distorted. Bandwidth changes will change only the displayed amplitude, not the shape. The displayed amplitude depends on the mean transmitted power, the measurement random-noise bandwidth (Bn), and the signaling bit rate (fs). For the PSK signal the relationship is: dB display relative to total power = 10 log (Bn/fs). (See **Appendix C** for derivation.) A greater signaling rate means more signal spreading, and thus less power output at the peak of the mainlobe. However, a greater signaling rate means more efficient spectrum utilization since the efficiency in terms of transmitted bits per hertz of output bandwidth is proportional to signaling rate (see Appendix B).

The signaling baud rate is also optimized with respect to transmitted power bandwidth, which has to be restricted to meet maximum output-bandwidth regulations. Consequently, the power in the sidelobes is filtered out, and sometimes some of the mainlobe, too, if the signaling baud rate is sufficiently high. The 10 log (Bn/fs) relative-amplitude relationship assumes that power loss due to filtering is negligible. Sometimes manufacturers correct for this assumption, as discussed later.

MEASURING TO FCC SPECIFICATIONS

Occupied Bandwidth: The bandwidth is measured at the 50 dB-down point, as shown by the FCC mask in **Figure 1**. The 50 dB bandwidth calls for a relative-level measurement in dB rather than an absolute power determination in dBm. What has to be measured is the occupied spectrum width at the point where the spectrum is 50 dB down from the "mean output-power level," as illustrated in **Figure 3**. If the power outside the output-filter bandwidth is ignored, then the mean output power is the same as the unmodulated level, and the relative-level formula 10 log (Bn/fs) holds.

The measurement consists of two steps. The first step is to determine the relative level between signal display peak and the mean power level. Manufacturers will usually specify this number. Alternately, the user can compute it from 10 log (Bn/fs). The remainder of the 50 dB is measured with respect to the mainlobe of the signal. It is important to note that the actual measurement bandwidth need not be 4 kHz because this measurement is a relative-level one.

Consider the following example. If the specified baud rate is 30.086 MHz, then 10 log (4 kHz/30.086 MHz) = -38.76 dB. With this high a baud rate, the output-filter truncation error is close to 0.8 dB, and the manufacturer specifies that the mainlobe is 38 dB down from mean output

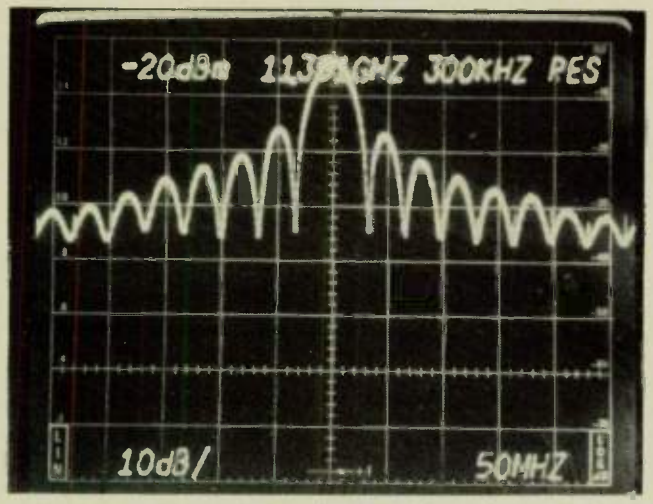

Fig. 2 Sinx/x spectrum generated by 8PSK digital radio transmission.

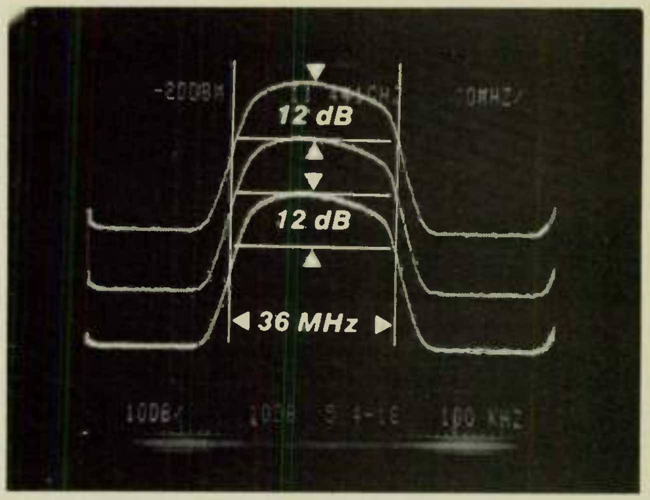

Fig. 4 Transmission-bandwidth illustration using different measurement resolution bandwidths.

power after the filter. Both numbers meet the intent of the FCC specifications, although referencing to the output of the filter makes the specification slightly tighter. The remainder of the measurement is illustrated in **Figure 4**, where the spectrum shape is observed using three different resolution bandwidths. In each case, the shape is the same and the bandwidth is 36 MHz at 12 dB down (note 38 + 12 = 50). The change in bandwidth, in each case by a factor of 10 times, moves the spectrum shape by 10 dB (10 log 10 = 10 dB), but this movement has no effect on the bandwidth measurement.

Filter Leakage: Output level in a 4 kHz noise bandwidth must be at least 80 dB down from the mean output level outside 250%

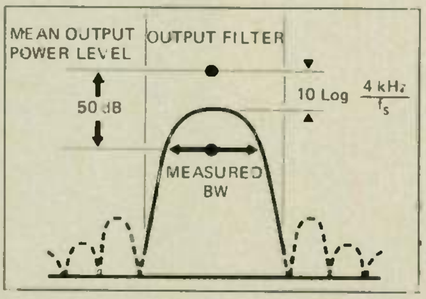

Fig. 3 Spectrum output level relative to mean output power.

of specified bandwidth offset. Without an output filter, sidelobes are only about 11 dB down, as shown in **Figure 5**. The function of the output filter is to reduce these sidelobes to the necessary −80 dB level.

Figure 6 shows such a measurement. The result can be interpreted in two ways. The simplest technique is to consider the peak of the display as 10 log (4 kHz/fs) down, and add the relative level of the leakage to this value. Thus, 38 + 53 = 91 dB. The other technique is to consider the effect of the bandwidth used in the measurement. A 300 kHz resolution bandwidth equals a (300 x 0.8 =) 240 kHz noise bandwidth (see **Appendix A**). Thus, the mainlobe is 10 log (240 kHz/30.086 MHz) = −21 dB from the mean power, plus 53 dB down for the leakage, or 74 dB total. The specification, however, is based on a 4 kHz bandwidth. For 240 kHz, the number would be 80 − 10 log (240/4) = 62 dB. This measurement indicates a level over 10 dB better than required.

An important point is that the spectrum analyzer internal noise

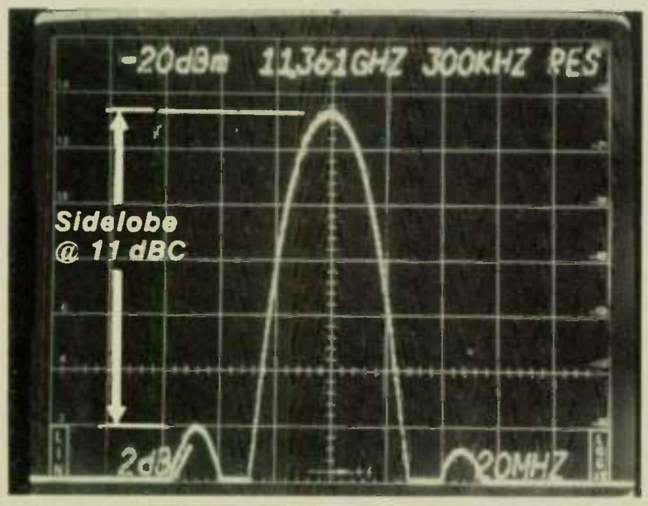

Fig. 5 Unfiltered modulation sidelobe is only 11 dB below mainlobe.

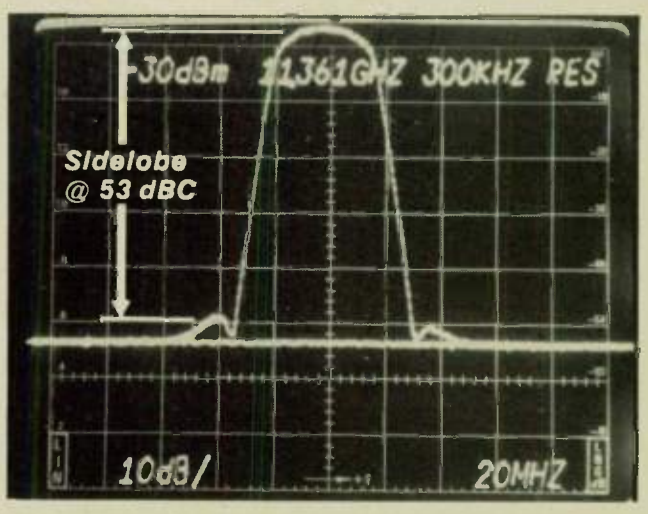

Fig. 6 After the filter, the sidelobe is 53 dB below mainlobe, as measured in a 300-kHz resolution bandwidth (240-kHz noise bandwidth).

(continued on page 39)

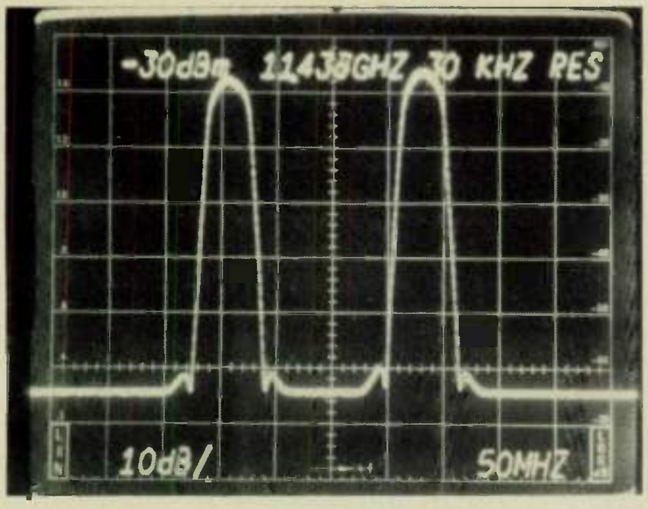

Fig. 7 Two digital radio signals on the same antenna.

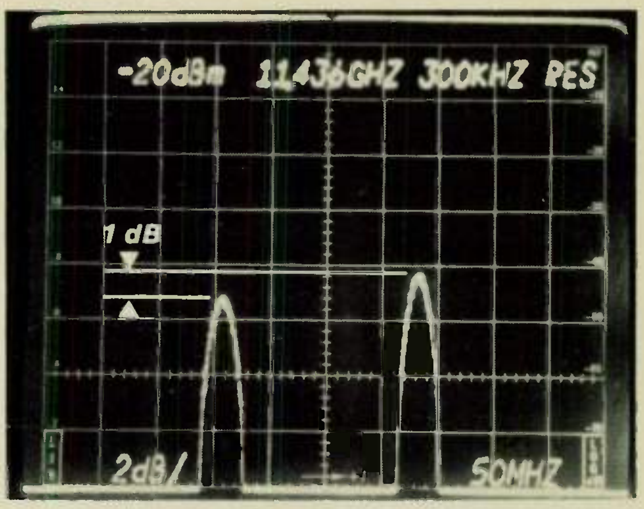

Fig. 8 Same as Fig. 7, showing an amplitude difference of 1 dB between the two signals.

This feature is discussed in a later section.

OTHER MEASUREMENTS

Numerous measurements besides those specified by the FCC are possible, as indicated previously. Some examples are given below.

Multiple-Carrier Level Balance: **Figure 7** shows two digital radio signals transmitted on the same antenna. Although these signals are supposed to be at the same amplitude level, clearly they are not. The use of a vertical display of 2 dB/div., as shown in **Figure 8**, illustrates an amplitude difference of 1 dB.

Spectrum Symmetry: **Figure 4** clearly shows an unsymmetrical mainlobe that indicates transmitter misadjustment.

Another form of asymmetry is shown in **Figures 9** and **10**, taken before the output filter. **Figure 9** shows that one sidelobe is more than 2 dB different in amplitude from the other one. By activating the amplitude-difference measurement mode (unique to the Tektronix 492 spectrum analyzer), the large sidelobe is adjusted to full screen and its amplitude relative to the mainlobe is determined at 13.25 dB (upper left read-out). Since a perfect sinx/x gives 13.26 dB, it is obvious that the right sidelobe is the one that is incorrect.

Interference: Stray interference can be captured by using digital storage display with maximum hold function. This technique holds random interference hits even when these hits occur for a short time. **Figure 11** shows an interference problem by the stray output within the signal nulls on the right side.

Spurious Outputs: The spectrum analyzer is an excellent tool in checking for spurious outputs. **Figure 12** shows a spurious signal offset by the 70 MHz intermediate frequency above the 6 GHz main signal.

MEASUREMENT PRACTICE

Amplitude Level: The input mixer of the spectrum analyzer will generate intermodulation distortion if driven by too high an input signal. For digital radio, the most serious intermodulation is third-order distortion whose components blend into the original signal and cause a wider appearance at the base. The lower the input level, the lower the distortion components. Spectrum analyzers are usually specified as having good input linearity for

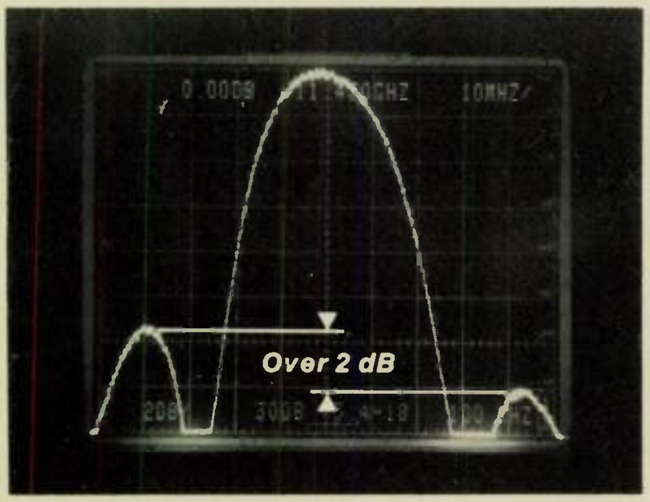

Fig. 9 Unequal sidelobes of digital radio transmission.

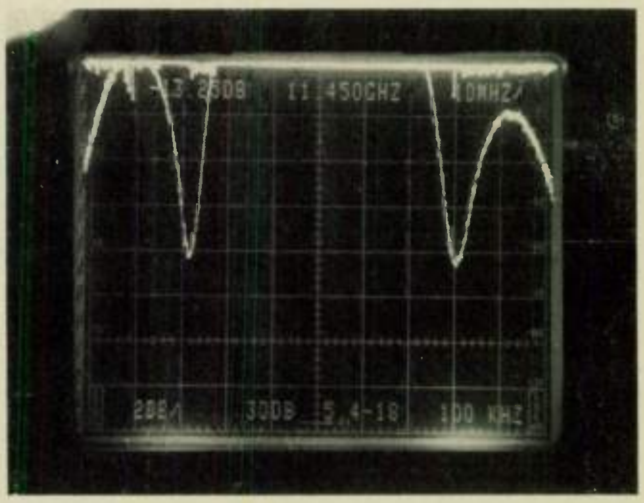

Fig. 10 Left sidelobe measures 13.25 dB down from mainlobe, as compared to 13.26 dB for ideal sinx/x shape.

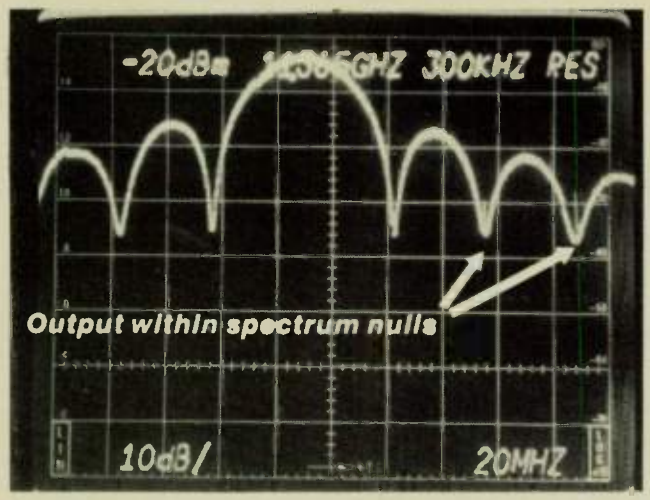

Fig. 11 Stray output within the spectrum nulls (measured before the output filter).

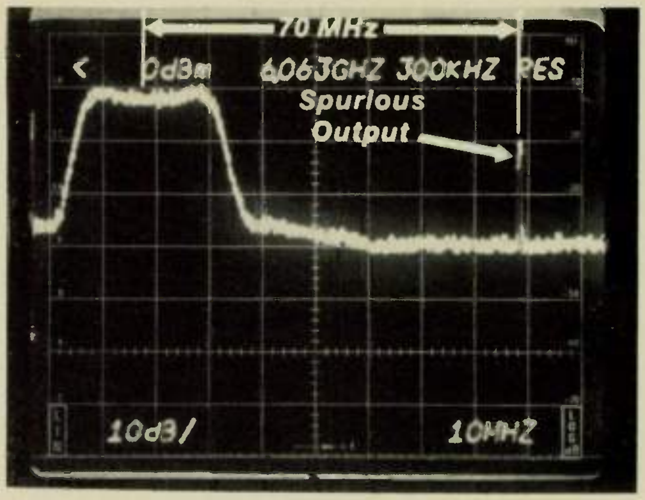

Fig. 12 Spurious output at 70-MHz IF offset.

input levels up to −30 dBm. At this point, the third-order products are between 70 and 80 dB down. The 50 dB down bandwidth measurement is consequently not affected; however, the 80 dB down leakage level can be in error. It should be noted that an intermodulating spectrum analyzer will show more low-level signal than is real. A radio that meets specifications under these conditions is actually better than measured.

The input levels necessary to make FCC-required measurements can be determined as follows: Peak signal display level with respect to mean transmitted power is 10 log (4 kHz/fs). For accurate measurement at 50 dB down, the internal noise level must be at least 60 dB down (assuring a measurement at least 10 dB above noise level). Therefore, the mean transmitted level should be 60 dB above internal noise, and the peak display level should be 60 − 10 log (fs/4 kHz) above internal noise. Based on a signaling rate of about 30 MHz, the peak signal display level needs to be at least 20 dB above the internal spectrum analyzer noise level. The immediate tendency might be to reduce the resolution bandwidth so as to cut the internal noise level; however, the signal is also noise-like, and the input-signal-to-internal-noise-level ratio will not change.

Since signal-to-noise ratio is essentially independent of resolution bandwidth setting, the choice of bandwidth must be established on the basis of other factors. Too wide a bandwidth prevents faithful reproduction of the input spectrum shape. A narrow bandwidth requires exceedingly narrow post-detection smoothing filtering and a very slow measurement sweep. These factors dictate a measurement bandwidth between 10 kHz and 1 MHz, with 100 kHz and 300 kHz the preferred resolution bandwidth positions.

A 100 kHz measurement bandwidth and 30 MHz signaling rate mean that the peak display is 10 log ($10^5/3 \times 10^7$) = −25 dB from mean transmitted power. A 20 dB signal-to-noise ratio represents a mean transmitted power level only 45 dB above internal noise. Thus, for a maximum peak transmitted level into the mixer of −30 dBm, the spectrum analyzer must have −75 dBm sensitivity. Most spectrum analyzers easily meet this requirement.

The 80 dBc leakage specification requires a minimum on-screen display of 80 − 10 log (fs/4 kHz) = 42 dB; a value of 50 dB includes some safety margin. Thus, a typical instrument such as the Tektronix 492, which as a sensitivity of −90 dBm at 100 kHz resolution at 6 GHz input frequency, has to be driven by a (−90 + 50 + 25 =) −15 dBm mean transmitted power level This level is within the linear display portion of the instrument, but some degree of low-level intermodulation is inevitable.

To reiterate, the best resolution-bandwidth settings are 100 kHz and 300 kHz. The input level to the mixer can be kept at a minimum by using spectrum analyzer front-end RF attenuation. Intermodulation-distortion errors should have no effect on the measurement accuracy of the 50 dB bandwidth. A few dB of intermodulation sidebands will usually appear around the 80 dBc leakage level. Any radio that meets specifications under these conditions is certainly within specifications might actually be good. Under these conditions, the input level to the mixer should be reduced as much as possible. One way of checking for the degree of error due to intermodulation is to check the relative out-of-band leakage level at two different input levels. If the relative leakage level does not change, then intermodulation is not a factor.

Signal Averaging: The noise-like behavior of the digital radio signal makes it necessary to average the display signal. Figures 13 and 14 show averaged and peak displays for 6 and 11 GHz signals. The difference in display level between peak and average is about 10 dB, the same as for ordinary random noise. Signal averaging is accomplished by actuating the video filter, use of digital averaging, or both.

Spectrum Analyzer Calibration: The horizontal and vertical calibration of the spectrum analyzer is quite important since a

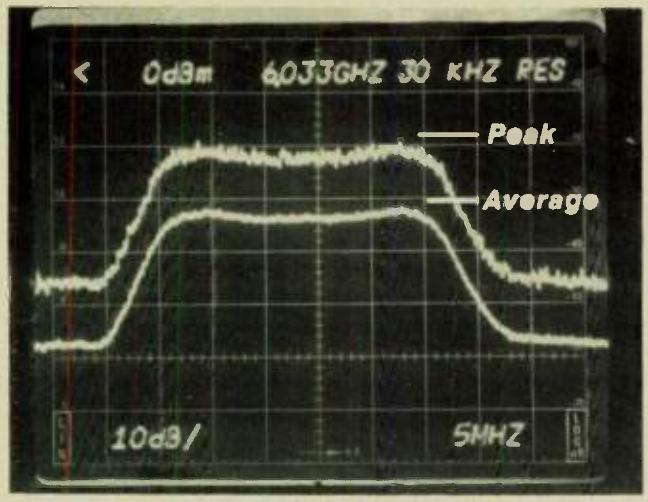

Fig. 13 6 GHz digital radio signal showing peak and average levels

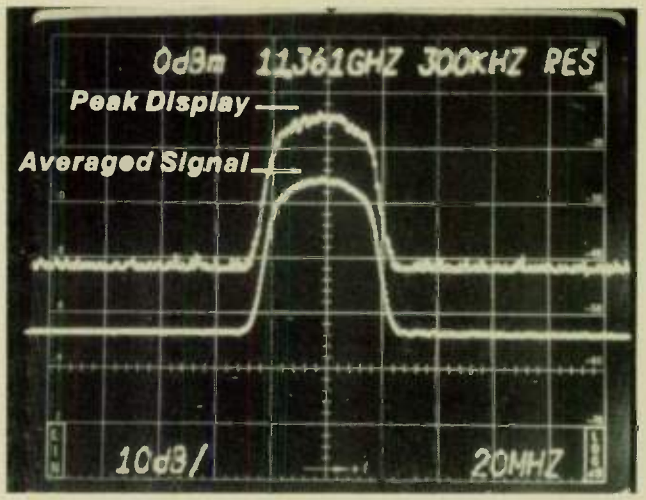

Fig. 14 11 GHz digital radio signal, showing peak and average levels.

small error may mean the difference between meeting or not meeting specifications for a marginal radio.

The vertical dB/div. logging accuracy is much more important than the absolute accuracy since most measurements are of a relative nature. Therefore, in tradeoffs between "Log Cal" and "Amplitude Cal," the choice should be to optimize logarithmic linearity.

Horizontal accuracy depends on the initial span/div. setting at calibrator frequency (usually 100 MHz) and the errors due to the span attenuator and frequency tuning. These errors are very low Nevertheless, in a marginal case, the user may wish to check the horizontal accuracy at the measurement frequency. This check can be made with any modulated signal source whose modulation frequency is accurately known. A comb-line generator such as the Tektronix #067-0885-00 will do the job.

Setting the Amplitude Level: Spectrum analyzer front-end attenuators operate in 10 dB steps. This attenuation may be too coarse when operating near the linearity limit of the spectrum analyzer. For example, at a typical test-point output of +3 dBm, the input level can be set to +3 dBm, -7 dBm, -17 dBm, etc. The display reference level, of course, can be set to other levels by means of IF controls. However, the input to the front-end circuits of the spectrum analyzer can only be set in 10 dB steps.

Thus, if 80 dBc leakage-measurement requirements call for -13 dBm, the input would have to be set at -7 dBm. The additional 6 dB of input will increase the intermodulation and, in some instances, cause input limiting which prevents full signal amplitude display. To check for input limiting, the user should add 10 dB of RF attenuation and check that the signal shape and level are correct. If the shape changes much, or if the signal level does not follow the reference-level change, then input limiting is a possibility. In this case, the spectrum analyzer cannot be operated correctly at the higher input level. Even if no limiting occurs, it may be wise to reduce spectrum analyzer drive level to limit intermodulation. However, intermodulation makes the leakage appear worse than actual. If specifications are met, the input drive level need not be changed.

To change the input drive level, it is necessary to insert an external attenuator between the radio and the spectrum analyzer. Three dB and 6 dB are good values. This attenuator need not be a precision type since absolute power levels are not in question and the exact value of the attenuator need not be known.

PRACTICAL HINTS FOR FCC MEASUREMENTS

Display with Carrier: The unmodulated carrier, when available, can be used as a reference of the mean transmitted power. Thus, **Figure 15** shows the unmodulated carrier and the modulated signal 21 dB down at a resolution setting of 300 kHz. Therefore, the display using a hypothetical 4 kHz bandwidth would be 21 + 10 log (240/4) = 38.8 dB down. The 50 dB bandwidth should be measured at the (50 - 38.8) = 11.2 dB points. The result shows that the 6 GHz signal just barely meets the 30 MHz specification.

Figure 16 shows a similar measurement at 11 GHz using the Tektronix 492. Modulated display level is 25 dB down from the carrier at an 80 kHz noise bandwidth. A 4 kHz measurement bandwidth would result in a display amplitude difference of 25 + 10 log (80/4) = 38 dB. The 50 dB bandwidth is measured at (50 - 38 =) 12 dB down. The result shows a 37 MHz bandwidth, well within the specified 40 MHz maximum.

Display Without a Carrier: The display level relative to the transmitted mean power is usually given by the radio manufacturer. The level difference can also be computed from 10 log (fs/4 kHz). For a 30.086 MHz signaling rate, the result is 10 log (30.086 MHz/4 kHz) = 38.8 dB for an 11.2 dB difference relative to 50 dB. **Figure 17** shows a 38 MHz bandwidth for an 11 GHz radio.

Checking for Leakage: For a 38 dB mean transmitted power

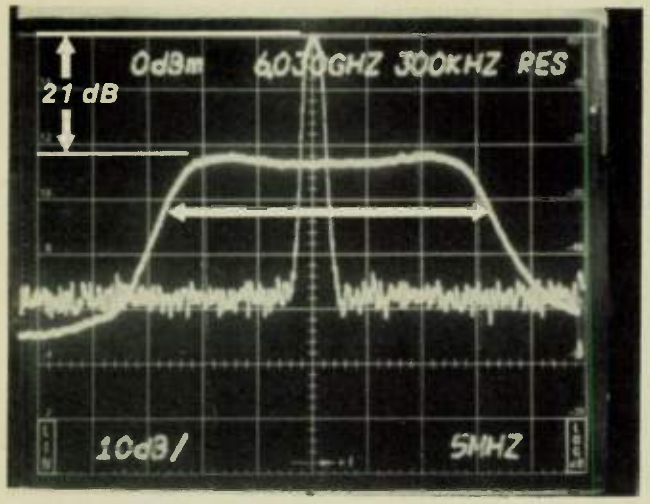

Fig. 15 Modulated spectrum level in relation to unmodulated carrier (6 GHz 8PSK radio).

Fig. 16 11 GHz radio, showing the modulated spectrum in relation to the unmodulated carrier.

to display level (4 kHz hypothetical bandwidth), the 80 dB down point is (80 - 38 =) 42 dB from the display peak.

The FCC formula for relative power level is A = 35 + 0.8 (P - 50) + 10 log B. For a specified bandwidth (B) of 30 MHz (at 6 GHz transmission), the percent off-set (P) from the carrier at which the attenuation must reach 80 dB is 87.5%, or 30 x 0.875 = 26.25 MHz. A = 35 + 0.8 (87.5 - 50) + 10 log 30 = 79.77 dB.

Figure 18 shows that at 25 MHz off-set, the level is more than 42 dB down, thus the specification is met. Because the precise center of the spectrum is difficult to establish, it might be more accurate to check for the vertical level that intersects the double-sided (52.5 MHz) spectrum width.

With a spectrum display level of -22 dBm and 10 dB RF attenuation, the mean transmitted power input for Figure 18 is -22 - 10 + 10 log (30 x 106/240 x 103) = -11 dBm. At this level, the 7L18 spectrum analyzer should be experiencing some degree of intermodulation. However, the specification is met even under these conditions. Removing the 10 dB of RF attenuation and letting the signal go 10 dB off screen makes it possible to check how badly the spectrum analyzer might be intermodulating. Even when driven 10 dB more, as shown in Figure 19, the 80 dBc relative to mean power specification is met.

Using the FCC Mask: A relatively painless way of checking to all the parameters of the FCC specification is to use a transparent CRT overlay (mask) that graphs the requirements of the specification equation A = 35 + 0.8 (P - 50) + 10 log B. Tektronix has such masks available for both the 492 and 7L18/7603 spectrum-analyzer displays at 6 GHz and 11 GHz. These masks are available under part number 020-0612-00. Figure 20 shows the 492 11 GHz mask. The mask is a graph similar to that shown in Figure 1. If the signal falls outside the mask, the specification is not met. If the signal falls within the mask, it is within specification.

The spectrum analyzer controls are set in accordance with the mask, that is, 10 dB/div. vertical and 10 MHz/div. horizontal. The digital radio signal is centered within the mask, and the reference level/gain is adjusted so that the computed 50 dB down floor permits this measurement.

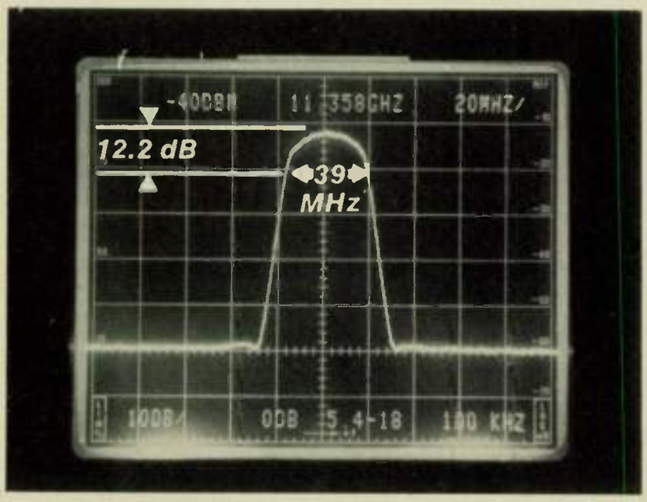

Fig. 17 Bandwidth measurement without a reference carrier.

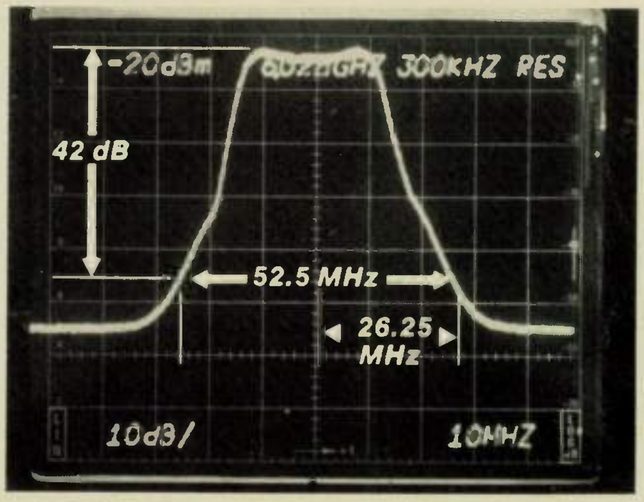

Fig. 18 Out-of-band transmission measurement.

(continued on page 46)

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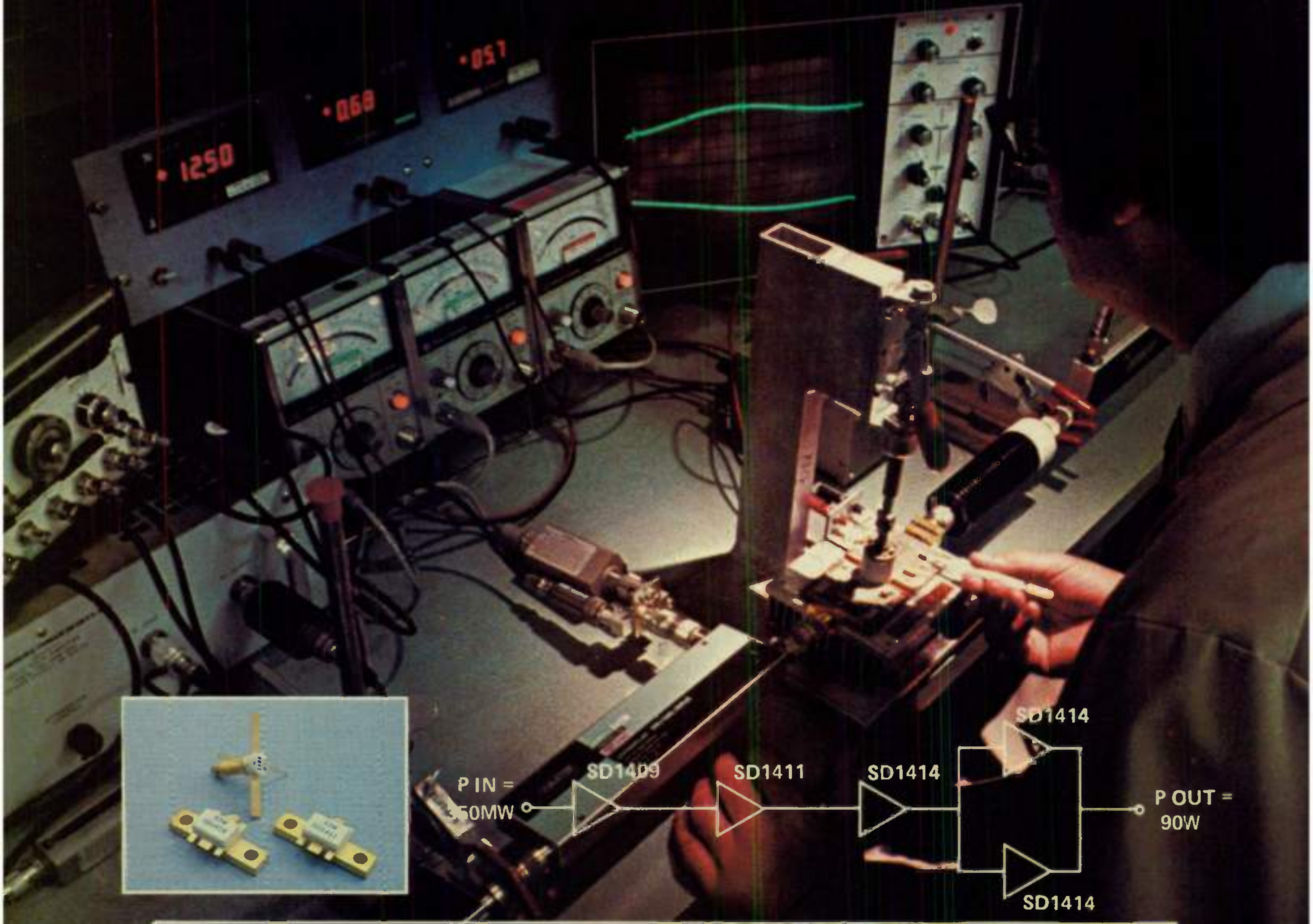

DEVICE TYPE	P OUT POWER OUTPUT WATTS	P IN INPUT POWER WATTS	PG POWER GAIN dB MIN	fo TEST FREQUENCY MHz	PACKAGE
SD1409	2.0	.35	7.6	870	XO72
SD1411	20	4.5	6.5	836	.230 6LFL
SD1414	50	15	5.2	836	.230 6LFL

V_{cc} (supply voltage: 12.5 volt)

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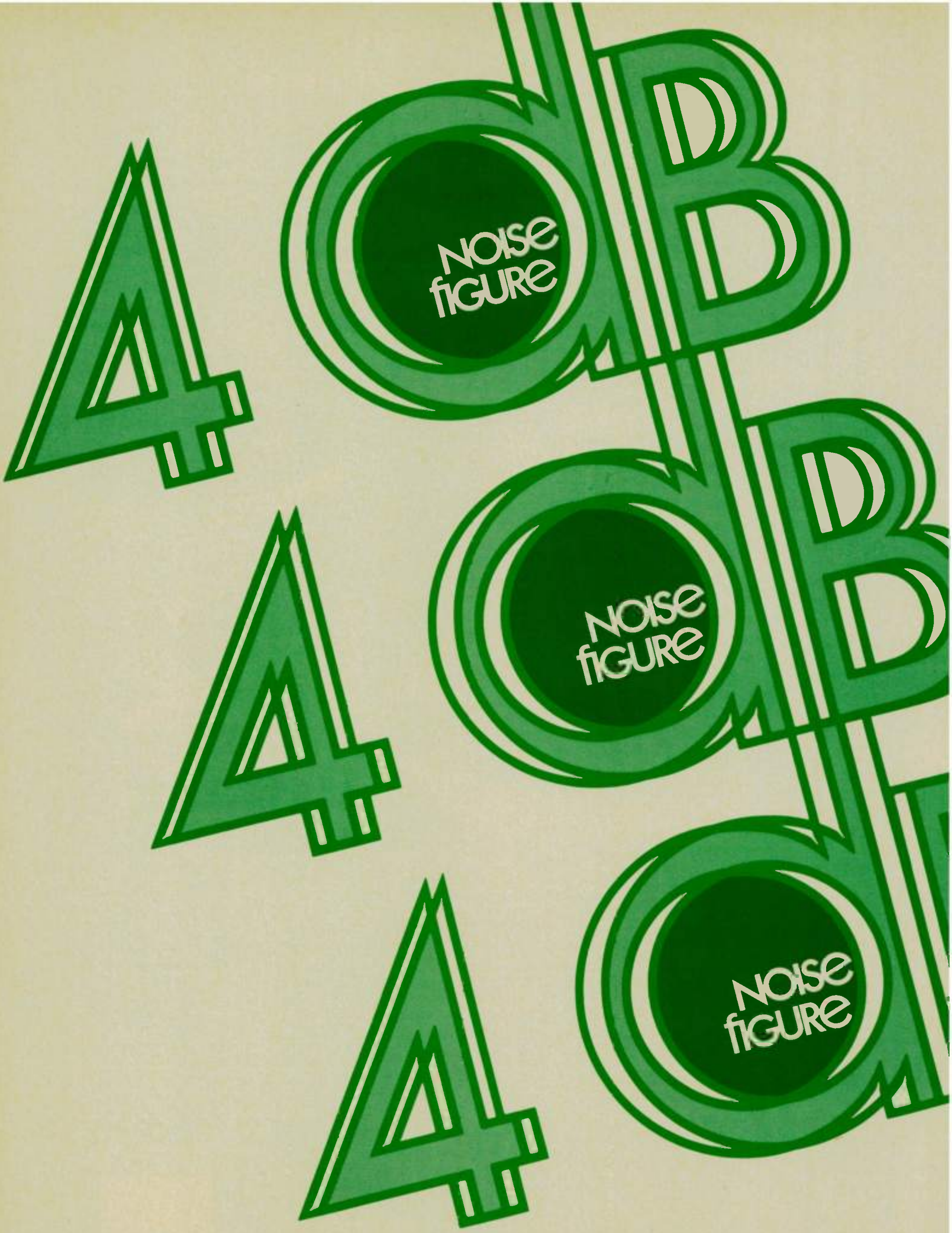
4 dB
NOISE figure
4 dB
NOISE figure
4 d
NOISE figure

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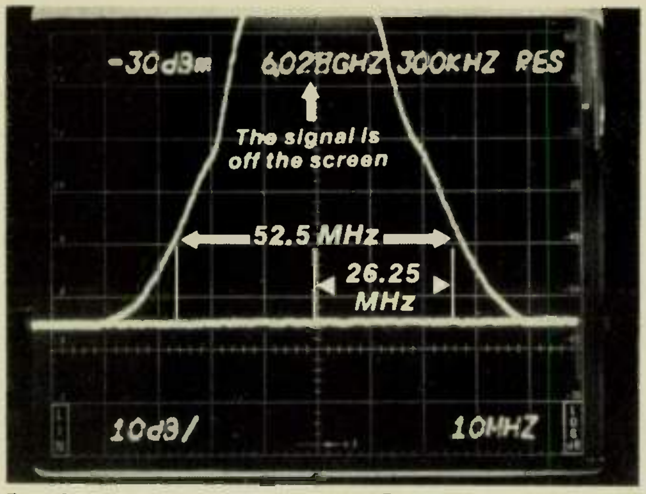

Fig. 19 Measurement with the signal of Figure 19 above full screen level.

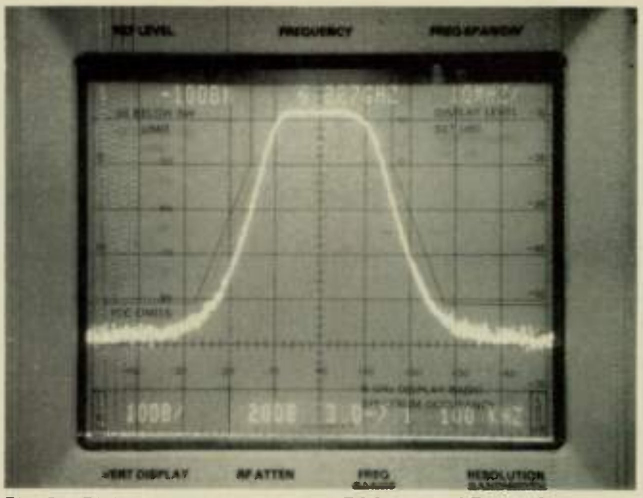

Fig. 21 Digital radio spectrum with FCC mask of Figure 21.

point coincides with the mask marking. For example, if the peak display is computed (from 10 log fs/4 kHz) or specified by the manufacturer to be 38 dB below mean transmitted power, then the peak display should be set (50 - 38 =) 12 dB above the 50-dB point of the mask.

The signal/mask combination is shown in Figure 21. Since the signal falls within the mask, the FCC specification is met.

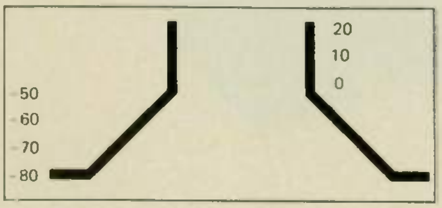

Fig. 20 The "FCC MASK" CRT overlay per FCC specifications equation.

ACKNOWLEDGMENT

The authors thank Dave Shores, whose research has been of help in preparing this applications discussion.

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APPENDIX A:
When Making Random Noise Measurements*

Random noise is displayed as a power spectral density and measured in watts/Hz. Display level is proportional to the random noise bandwidth of the spectrum analyzer. If a particular amplitude is displayed with one bandwidth (B1), the amplitude for a different bandwidth (B2) will change by 10 log B2/B1. The resolution bandwidth (Br) specified for a spectrum analyzer does not equal the random noise bandwidth (Bn). The relationship is approximately Bn = 0.8 Br when resolution is specified at the 6 dB points as at Tektronix, and Bn = 1.2 Br when resolution is specified at the 3 dB points.

The peak amplitude of random noise is unpredictable, changing from moment to moment. The desired rms value, though, does not change. However, spectrum analyzers do not measure the rms value directly. Rather, they respond to the peak value and are calibrated in rms. For random noise, the signal is averaged by smoothing by use of a narrow post-detection filter (video filter), digital averaging, or both. The display then corresponds to the average value of the signal. The rms value is greater than the average value by $(4/\pi)^{1/2} \rightarrow 1.05$ dB. When the noise signal undergoes logarithmic compression prior to averaging, then an additional 1.45-dB error is introduced. Thus, to get to true rms value it is necessary to multiply by $(4/\pi)^{1/2}$ in the linear voltage mode, or add 2.5 dB (1.05 + 1.45) in the logarithmic dBm mode. These correction factors apply only to absolute-level measurements. The correction factor drops out in relative-level measurements.

The total noise displayed consists of incoming noise plus the spectrum analyzer internal noise. Thus, when incoming noise equals internal noise, the total displayed will be twice either for a 10 log 2 = 3 dB error. Accurate low-level noise measurement requires correction for the spectrum analyzer internal noise. Table I can be used for such correction.

For accurate determination of a noise shape, the measuring bandwidth should be no greater than one-tenth of the shape to be measured.

* See Tektronix application notes AX-3260 — Noise Measurements Using the Spectrum Analyzer, and AX-3861 — Swept Selective Level Measurements.

APPENDIX B: Bits/Hz Derivation

The number of transmitted bits equals the product of the signaling rate and the number of binary states per signaling vector.

The number of hertz equals the output bandwidth. Thus, for an 8-vector system with 3 bits per vector at a bandwidth of 30 MHz, a signaling rate of 25.72 MHz corresponds to (3x25.72)/30=2.57 bits/Hz.

A signaling rate of 30.086 MHz and 40 MHz bandwidth corresponds to 2.26 bits/Hz.

APPENDIX C:
Derivation of Relative Level Formula

Phase modulation does not change the total power within the signal; only the frequency distribution is changed. Thus, the output power of the modulated signal equals the power of the unmodulated carrier.

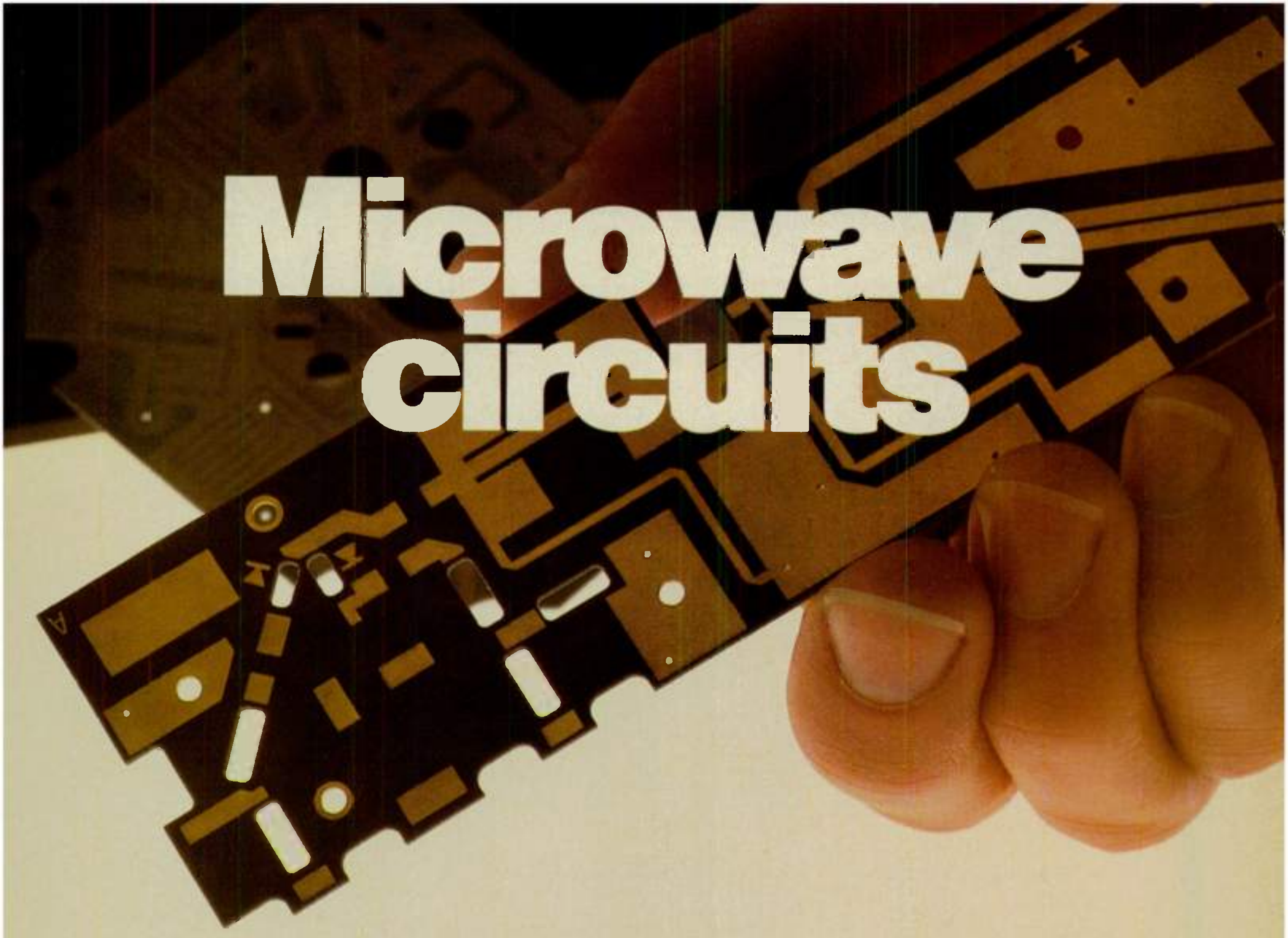
The modulated carrier is a sinx/x spectrum with a total power level of

$$\int_{-\infty}^{+\infty} \frac{\sin^2 x}{x^2}\, dx = \pi$$

TABLE I

Displayed dB above internal noise	1	1.5	2.0	3.01	4	5	6	8.0	10.0
Actual input dB above internal noise	−5.87	−3.85	−2.33	0	1.80	3.35	4.74	7.25	9.54
Actual input dB below displayed level	6.87	5.35	4.33	3.01	2.20	1.65	1.26	0.75	0.46

(continued on page 48)



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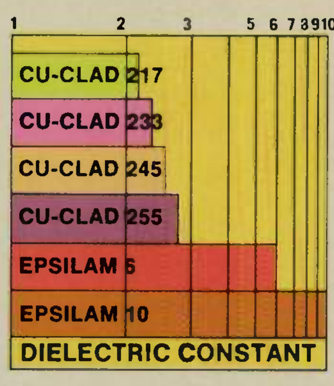

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A rectangle π radians wide and the peak of the mainlobe in amplitude has an area equal to that under the $\sin^2 x/x$ curve which equals the total transmitted power. Pi radians also correspond to a spectrum null which occurs at the signaling frequency (fs). When the peak of the mainlobe is measured using a relatively small noise bandwidth (Bn), the normalized output equals Bn. Thus, the ratio of display level to total transmitted level is Bn/fs.

These relationships are illustrated in **Figure 22.**

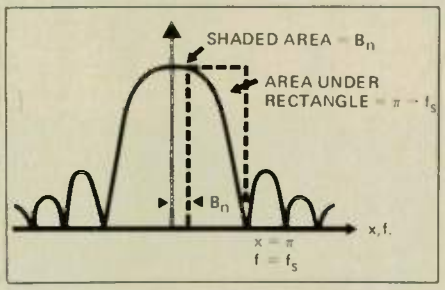

Fig. 22 Relative power levels.

APPENDIX D:
Choosing a Spectrum Analyzer

Frequency Range: Most modern microwave spectrum analyzers provide sufficient performance to permit checking to FCC specifications directly at carrier frequency. Under these conditions, the user might decide to trade special performance features, such as very high resolution and stability, and cost versus low-end frequency range. Such an instrument is the Tektronix 7L18, whose lowest operating frequency is 1.5 GHz.

For some measurements it might be desirable to go to a 70 MHz measurement at IF. Such an instance would be when looking for very low-level interference beats on multi-signal antenna systems. The sensitivity at microwave frequency of current spectrum analyzers is usually not sufficient. Use of a narrow resolution bandwidth does not improve sensitivity because the interfering signal is noise-like; therefore, the level goes down when the bandwidth is cut. Under these conditions, a full-frequency-range instrument such as the Tektronix 492 might be a better choice.

Preselection, Sensitivity and Dynamic Range: The ideal instrument for digital radio would have very high sensitivity and high dynamic range. Unfortunately, better sensitivity does not necessarily mean better dynamic range.

The dynamic range is the difference in dB between the sensitivity noise level and the largest permissible input signal. Typically, preselected spectrum analyzers have 5 dB better noise sensitivity than unpreselected ones. The preselector is a tracking filter that eliminates spurious responses. Furthermore, while the sensitivity is indeed better for such low-level applications as looking for spurious signal beats, the dynamic range is not really improved because the largest input signal is controlled by the level at the mixer. With the preselector in front of the mixer, one can simply drive the input that much harder. Therefore, while a non-preselected instrument option will save money, it will not improve measurement capability to FCC specifications, which require dynamic range for relative-signal-level checks rather than absolute-level sensitivity.

Likewise, going to a higher-stability, narrower-bandwidth instrument will not help, as discussed previously, because of the noise-like nature of the signal.

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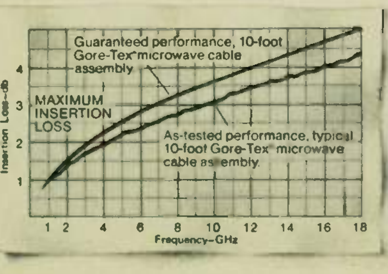

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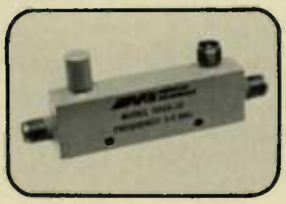

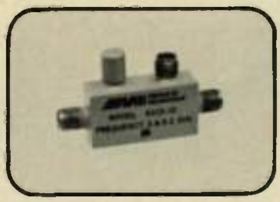
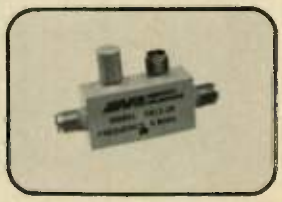

SPECIFICATIONS

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

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

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

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

DIMENSIONS
(See Engineer Drawing Above)

All dimensions are in inches

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

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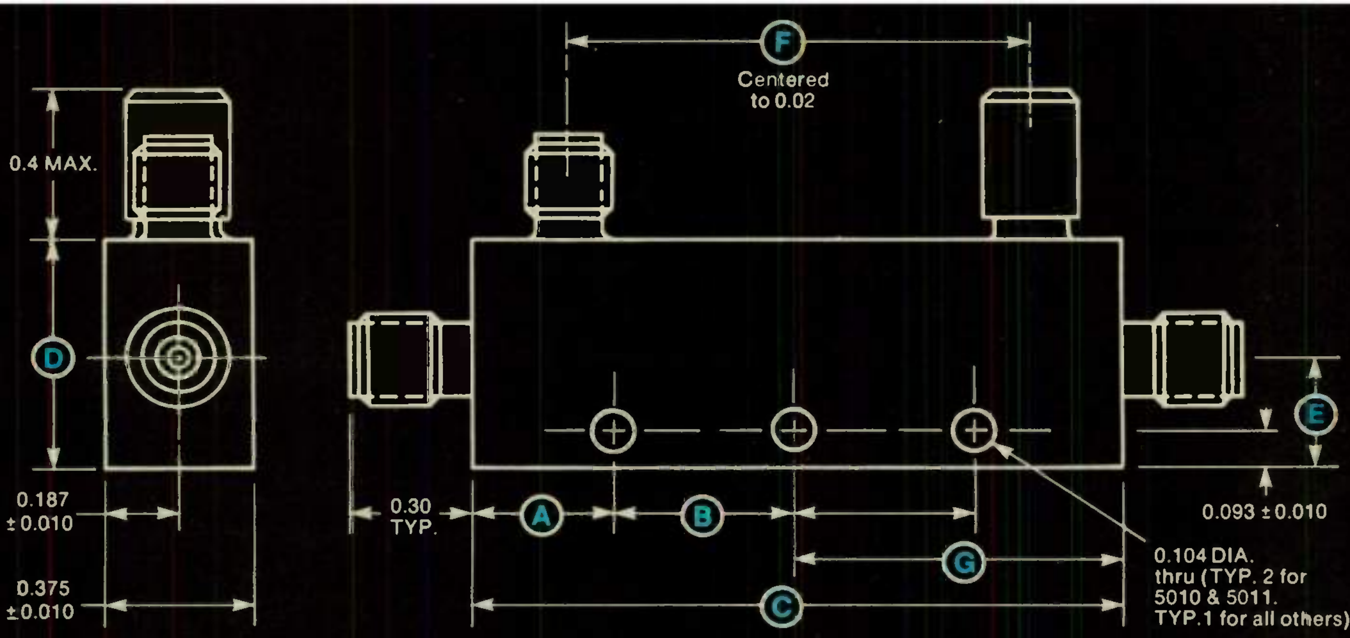

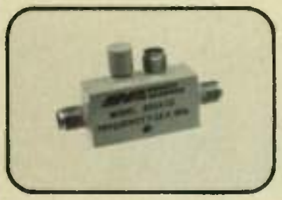

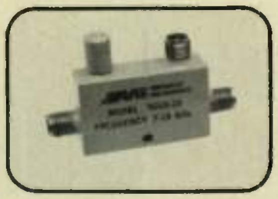

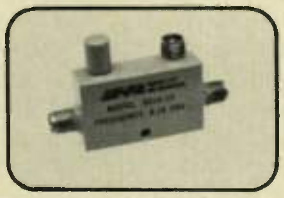

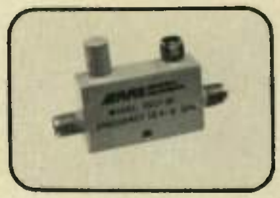

5014-6	5014-10	5014-20
ALL ARE 7-12.4		
6 ± 1.00	10 ± 1.25	20 ± 1.25
ALL ARE ± 0.50		
15	BOTH ARE 17	
ALL ARE 1.30		
ALL ARE 0.40		

5015-6	5015-10	5015-20
ALL ARE 7-18		
6 ± 1.00	10 ± 1.25	20 ± 1.25
± 0.60	BOTH ARE ± 0.75	
BOTH ARE 12		15
ALL ARE 1.40		
ALL ARE 0.50		

5016-6	5016-10	5016-20
ALL ARE 8-16		
6 ± 1.00	10 ± 1.50	20 ± 1.25
± 0.60	BOTH ARE ± 0.75	
BOTH ARE 12		15
ALL ARE 1.40		
ALL ARE 0.50		

5017-6	5017-10	5017-20
ALL ARE 12.4-18		
6 ± 1.00	10 ± 1.00	20 ± 1.00
ALL ARE ± 0.50		
ALL ARE 15		
ALL ARE 1.40		
ALL ARE 0.40		

N/A	
N/A	
ALL ARE 1.15	
BOTH ARE 0.60	0.65
ALL ARE 0.30	
ALL ARE 0.36	
ALL ARE 0.57	

N/A	
N/A	
ALL ARE 1.25	
BOTH ARE 0.75	0.80
ALL ARE 0.30	
ALL ARE 0.77	
ALL ARE 0.63	

N/A	
N/A	
ALL ARE 1.25	
BOTH ARE 0.75	0.80
ALL ARE 0.30	
ALL ARE 0.77	
ALL ARE 0.63	

N/A	
N/A	
ALL ARE 1.25	
BOTH ARE 0.75	0.80
ALL ARE 0.30	
ALL ARE 0.77	
ALL ARE 0.63	

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INTRODUCTION

The need for present-day electronic warfare systems to process in real time all signals in a dense signal environment has stimulated a substantial effort in recent years to develop specialized high-performance receivers. The performance requirements on these receivers include broad instantaneous bandwidth, separable response to simultaneous signals of varying types, large dynamic range, instantaneous frequency measurement and tolerance to military environments, among others, as well as low cost, weight and volume an ever-present and basic demand. Channelized and pulse-compression receivers have attracted particular developmental attention recently, notably drawing on acoustic wave devices[1,2] and the newer magnetostatic wave devices[3] to provide the high-performance filter technology needed in such receivers. For pulse compression receivers, advances in dispersive SAW filters,[4] and now magnetostatic wave dispersive delay lines,[5] hold out the promise of practical devices with operational microwave bandwidths of 1 GHz and above, but at the expense of substantial insertion loss, temperature sensitivity, and costly fabrication technology and materials.

An alternative technique, and the results of a feasibility study, are presented here for fabricating broadband microwave dispersive delay lines of moderate time-bandwidth product. This technique uses conventional semi-rigid microwave coaxial cable in a reflective mode that offers the following practical advantages: employs readily available inexpensive mass-produced precision coax cable; has pulse-compression bandwidth of several GHz with low insertion loss in a single line; requires no special fabrication tooling or techniques with easy

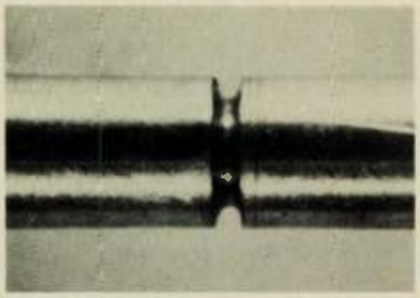

Fig. 1 Annular-discontinuity reflector in 141 mil copper coax made with tube cutter.

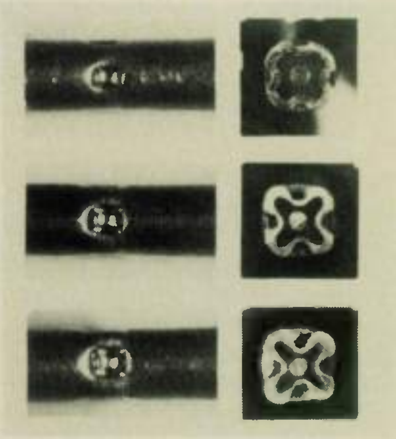

Fig. 2 Light, medium and heavy crimps in 85 mil copper made with modified crimping tool.

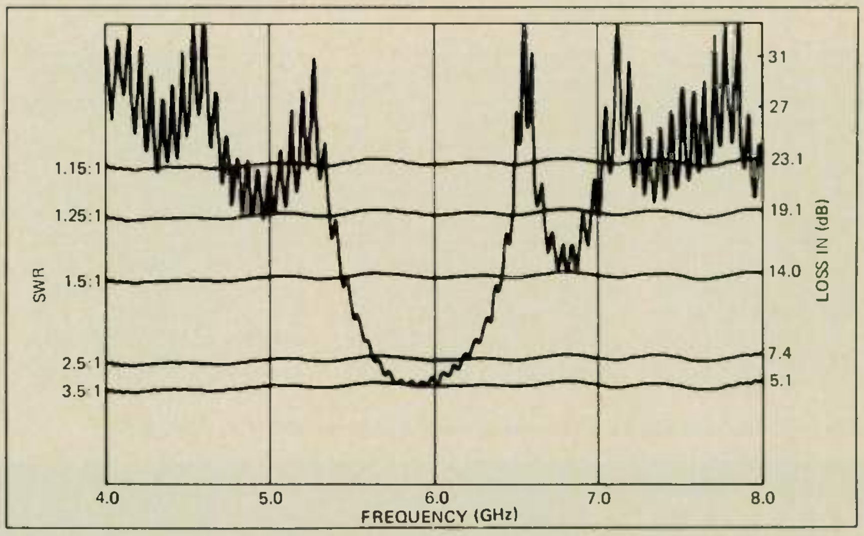

Fig. 3 Experimental reflection signal (return loss) of nine-crimp line, design-centered at 6.0 GHz; uniformly spaced heavy crimps (ref. Fig. 2).

transfer of technology; exhibits automatic 50 ohm characteristic impedance without matching networks; has low temperature coefficient of delay with multi-line packaging which eliminates the need for temperature control in a military environment; features independent amplitude-weighting and time-delay compensation for stringent performance requirements within the line; and has compact structures with no shielding or special packaging requirements (e.g., approximately 50 cubic centimeters and 200 grams for a 100 nsec line with a 1 GHz dispersive-delay bandwidth.) There is, furthermore, a substantial precedent for the use of coaxial delay line in electronic warfare equipment: simple, reliable wideband coaxial delay line has been used for many years in many thousands of repeater jammers as the microwave memory element, and no other device has proven as economically effective in this application.

CONSTRUCTION AND PROPERTIES OF REFLECTIVE CRIMPED-COAX-LINE FILTERS

The following discussion traces the development of the Coax Reflective Dispersive Line (CRDL). The reflector fabrication description includes annulus and crimp geometries, and the electrical nature of the crimp discontinuity, with support photographs and data. **Figure 1** shows an annular reflector geometry. This discontinuity geometry is attractive in being axially short, and radially symmetrical, thus most closely matching the purely capacitive diaphragm discontinuities treated theoretically in the literature (Reference 6, **Figure 9**). The annular depression in **Figure 1** was produced in 141 mil copper coax, and was made with a rolling wheel tube cutter having the wheel edge rounded to a 10 mil radius of curvature. In spite of this blunting, the side walls of the annular depression can become too thin and can rupture. This may occur before the capacity of the discontinuity is large enough to have a sufficient reflection coefficient, and is ac-

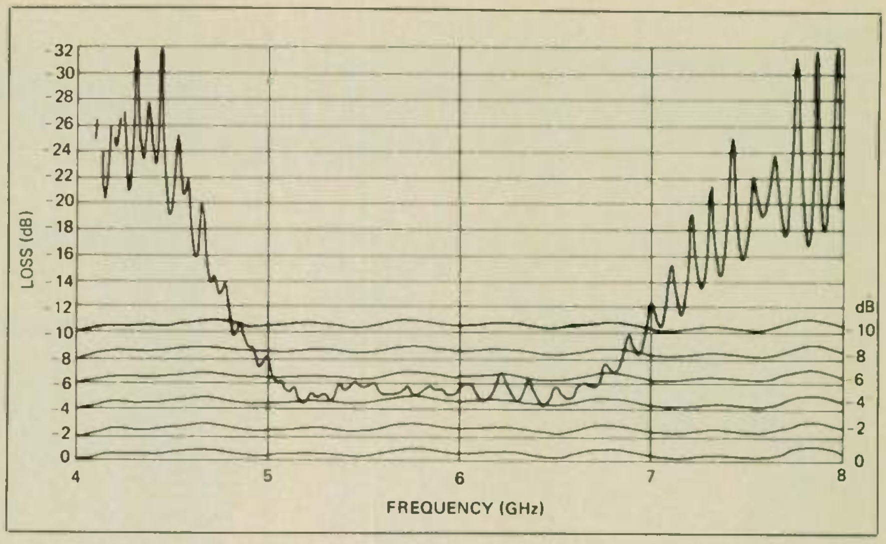

Fig. 4 CW reflective insertion loss for 2.0 GHz-bandwidth line driven from low-frequency end.

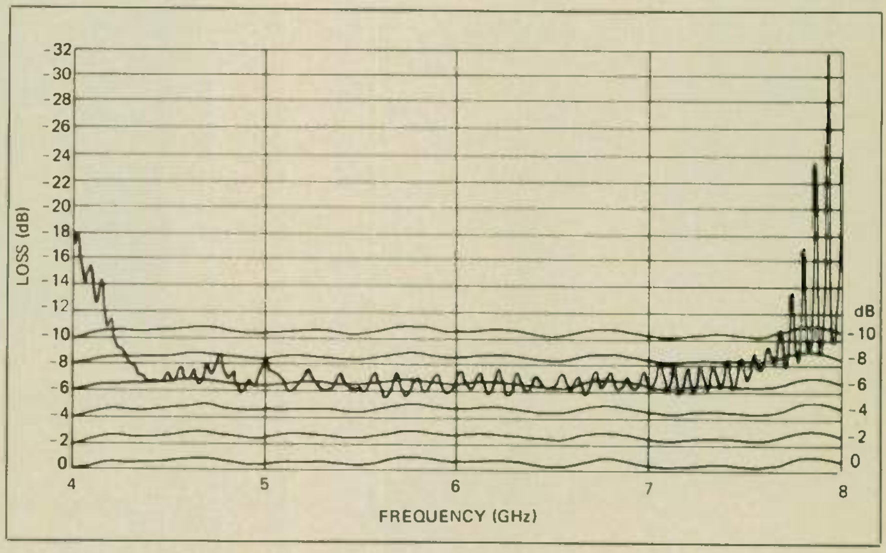

Fig. 5 CW reflective insertion loss for 3.6 GHz-bandwidth line driven from low-frequency end.

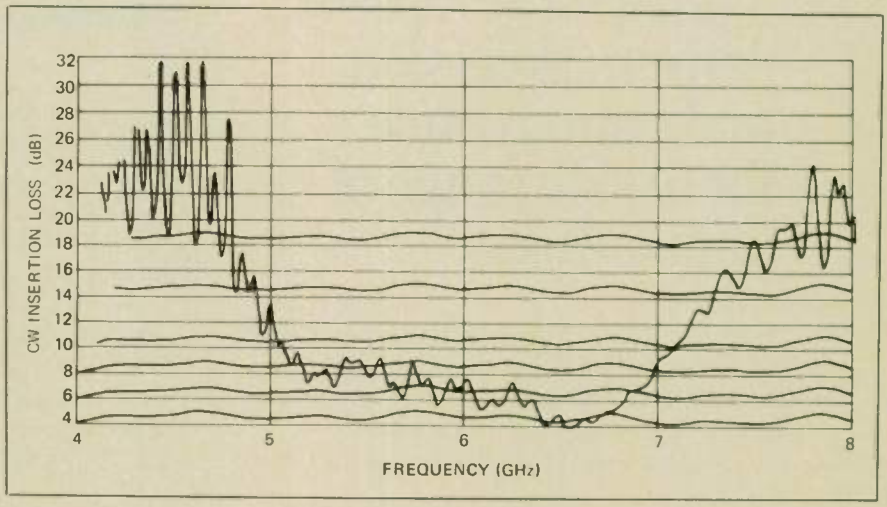

Fig. 6 Response slope resulting from driving the 2.0 GHz-bandwidth line from high-frequency end.

(continued on page 55)

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companied as well by an appreciable elongation of the cable.

A modified crimping tool was then used successfully to produce the discontinuities shown in **Figure 2**; these are in 85-mil copper coax, and are shown externally (one of four symmetrical depressions) and in cross section for light, medium and heavy crimps. The modified crimping tool has a keyed dial allowing eight different resettable crimp depths; the deepest crimp caused a slight rupture of one of the four copper indentations, resulting in the use of the next-deepest setting for the heavy crimp of **Figure 2**, and for the experimental reflection calibration plus dispersive lines described below.

The equivalent circuits for idealized diaphragm-shaped coax line discontinuities were found some time ago[6,7] by J. R. Whinnery, et al. On the assumption that the crimp discontinuities of **Figure 2** are axially sufficiently short, these discontinuities behave like a simple shunt capacity.

The capacity of the heavy shunt in **Figure 2** can be calculated by using the three radii within the 85-mil diameter line: the inner-conductor radius (10.6 mils), the inner radius of the outer conductor (31.9 mils), and the effective inner radius of the outer conductor in the center region of the crimp (16.8 mils). Using the results of Reference 6, taking $\epsilon = 2$ for the teflon insulation, and using a design center frequency of 6 GHz, this shunt capacity has a reactance of $|X_c(6\ \text{GHz})| \doteq 736\Omega$. The reflection coefficient of such a reflector in a 50Ω line is approximately $\rho \doteq 50/736 = 0.068$, with phase shift $= \tan^{-1} \rho = 87.7°$.

This reflection coefficient was experimentally calibrated and confirmed by fabricating a nine-crimp deep-crimp filter in 85-mil coax. The return-loss recording is shown in **Figure 3**, with calibration SWR/loss recordings included. Note that the design center frequency of 6.0 GHz became about 5.85 GHz by two effects: the heavy crimps cause a physical elongation of the line of about 7 mils per crimp, accounting for about a 0.7% physical elongation of the line for a frequency factor of 0.993; and an additional RC delay produced by the 50 Ω line impedance and 736 Ω capacitive reactance (each 1/2 wavelength) results in a phase delay of $\tan^{-1}$ (50/736) for each half wavelength, (0.021 fractional), or a frequency factor of 0.979. The theoretically-corrected center frequency is thus:

$$f_c = (6\ \text{GHz}) \times (0.993) \times (0.979) = 5.83\ \text{GHz},$$

in good agreement with the experiment.

The experimental peak-response insertion loss in **Figure 3** is about 5.2 dB, compared with a theoretical reflective return loss (using f = 5.85 GHz) of:

$$\text{Loss} = -20 \log(9 \times 0.066) = 4.5\ \text{dB}$$

within experimental error.

While the response of the nine-reflector filter should be approximately sin x/x in form, it is clear from **Figure 3** that the high frequency sidelobes exceed the low. This asymmetry results from two 6 dB per octave effects: the linear increase in (small) reflection coefficient (for a capacitive shunt) with frequency, and the number of reflectors per incremental frequency interval that increases linearly with frequency. Note in the experimental data of **Figure 3** that the first high fre-

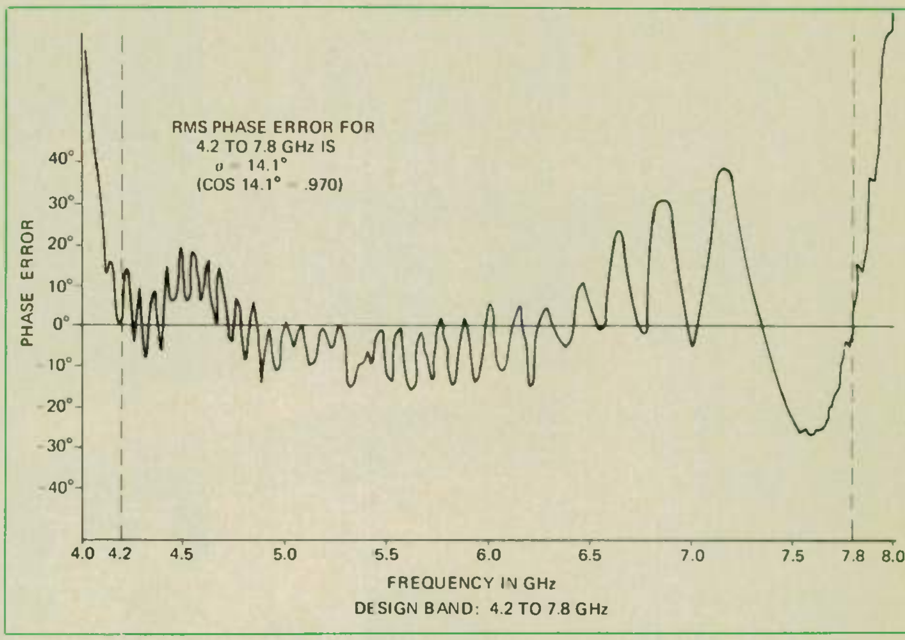

Fig. 7 Experimental phase measurement of 3.6 GHz-bandwidth line from HP network analyzer. Plot is the phase error between experimental 4.2 to 7.8 GHz linear-FM dispersive line, and best-fit quadratic phase function (rms error of 14.1°).

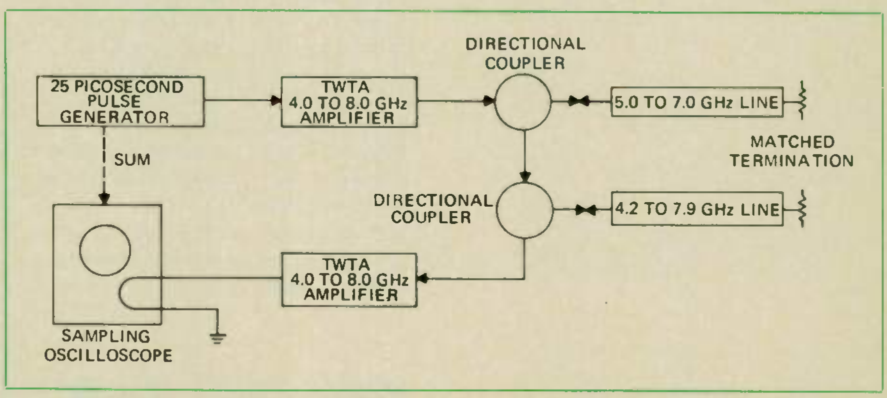

Fig. 8 Block diagram of basic test circuit for swept-LO and pulse-compression lines.

(continued on page 57)

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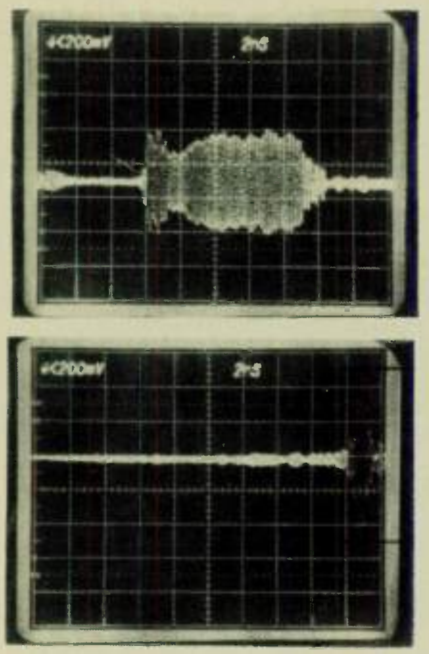

Fig. 9 Pulse expansion (upper-a) and recompression (lower-b) at 6.0 GHz center frequency and 2.0 GHz compression bandwidth, common time scale. (Auxiliary lines in lower waveform accent peaks of compressed pulse.)

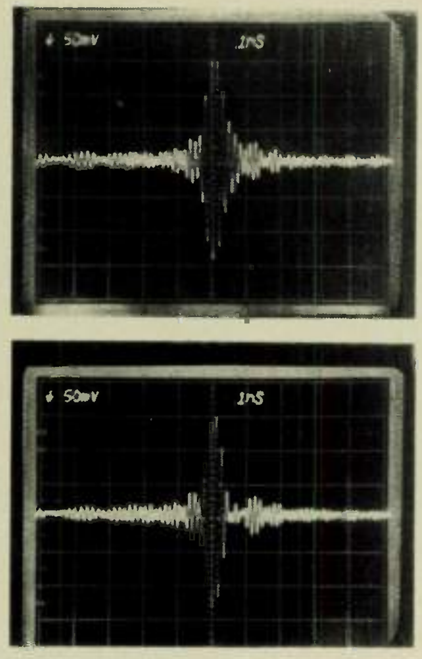

Fig. 10 Compressed pulse with straight swept-LO line (upper-a) and with coiled swept LO line (lower-b).

quency lobe exceeds the first low frequency lobe by about -15 + 20.6 = 5.6 dB, while by the preceeding 12 dB per octave prediction, the difference is expected to be (with Figure 3 frequency peaks of 6.82 GHz and 4.95 GHz):

$$20 \log \left(\frac{6.82}{4.95}\right)^2 = 5.7 \text{ dB} ,$$

a satisfactory self-consistent match.

From the above results, a theoretical and experimentally-confirmed reflection capacity of 0.036 pfd can be utilized for deep crimps. The same analysis yields capacities for the medium and light crimps of Figure 2 that are 0.0162 pfd and 0.0074 pfd, respectively. Thus it is straightforward to achieve a 10-to-1 range of crimp reflection coefficients, all that is required to permit building amplitude weighting directly into the line.

The extensive line elongation of a heavy crimp demands that this elongation be considered in a line-design crimp-position distribution. A second order amplitude and phase correction is available by very light crimps placed between initial crimps for fine-tuning amplitude and phase corrections of the measured line. The elongation effects of these light crimps can be ignored. Amplitude corrections are made midway between heavy crimps to produce a 180° phased reflection, thus affecting amplitude and not phase. Phase corrections that leave amplitude response undisturbed are made at 1/4 or 3/4 positions between crimps, depending on whether a positive or negative phase correction is required.

DESIGN, CONSTRUCTION AND TEST OF EXPERIMENTAL DISPERSIVE LINES

Two short unweighted lines were designed, fabricated and tested to show the ability to produce broadband crimped-coax dispersive filters: the swept-LO line was designed for a baseband of 5.0 to 7.0 GHz with an active length of 36.0 inches, and a pulse-compression line with a broad baseband spectrum of 4.2 to 7.8 GHz occupying 64.8 inches.

Starting with the low frequency end of the chirp line, the recurrence relationship used in calculating crimp positions is:

$$x_{i+1} = x_i + \frac{Kc/2f_L}{1 + x_i\,(f_H - f_L)/Lf_L}$$

This is derivable from the continuous model for a linear delay-vs-frequency characteristic,

$$f(x) = \frac{x}{L}\,(f_H - f_L) + f_L \,,$$

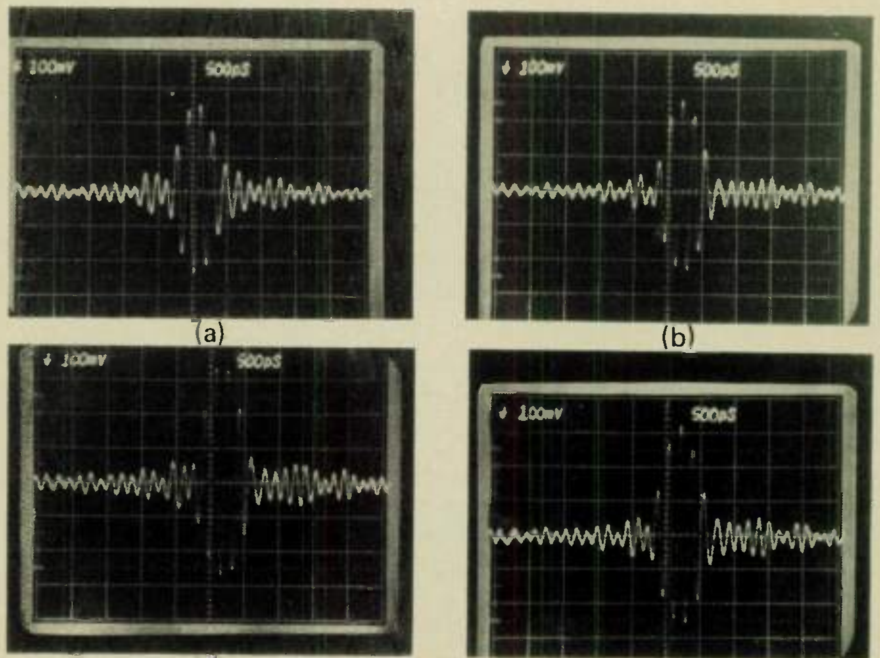

Fig. 11 Compressed pulse under temperature testing of coiled 5.0 to 7.0 GHz swept-LO crimped-coax line: (a) +25°C, (b) +125°C, (c) -65°C, and return to (d) +20°C (pulse-compression line at steady +20°C).

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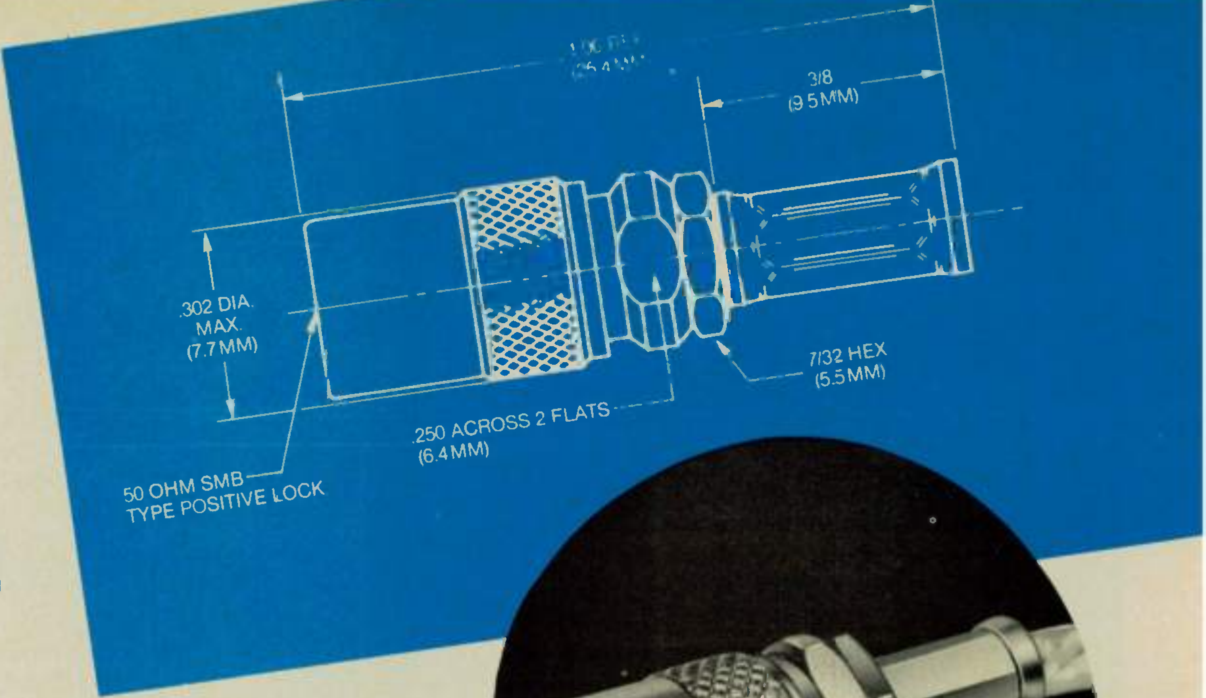

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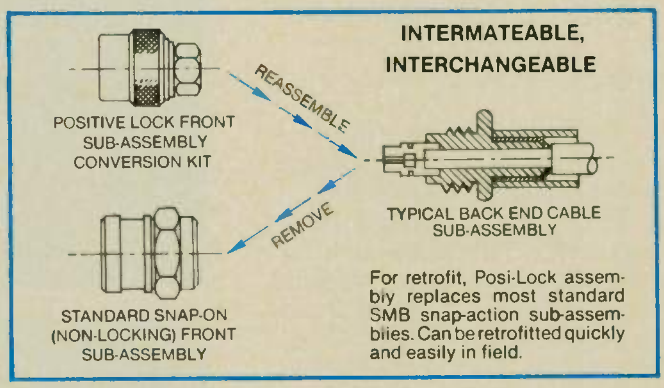

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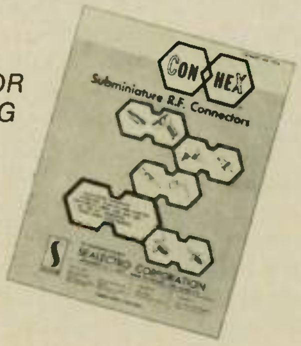

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where x is the length variable along a delay line of total (differential delay) length L, f_H and f_L are the high and low dispersive-filter band edges respectively, x_i are the reflector positions along the line, c is the velocity of light, and K is the fractional velocity, $\epsilon^{-1/2}$, caused by a coax dielectric medium of dielectric constant, ϵ.

The broadband line was designed using $K = 1/\sqrt{2}$, $c = 3 \times 10^{10}$ cm/sec, $x_o = 0$, $f_L = 4.2$ GHz, $f_H = 7.8$ GHz, and L = 64.8 inches. The lines were fabricated with a table of $x_i - x_{i-1}$ values from which 7 mils was subtracted to allow for line stretching due to heavy crimping. Note that the reflective pulse compression delay line has an unweighted compression ratio of TB = 15.55 nsec X 36 GHz = 55.8, while that for the chirp line is TB = 8.62 nsec X 2 GHz = 17.2.

Figures 4 and **5** show the CW insertion loss of the 2 GHz and 3.6 GHz lines respectively when driven from their low frequency ends. The flatness of both amplitude characteristics is noted, with inband insertion losses of 4.5 dB and 6.0 dB respectively; the two positive 6 dB per octave effects described earlier closely compensate for increasing propagation losses in the lines with frequency. **Figure 6** shows the slope that results from driving the 2 GHz swept-LO line at its high-frequency end, since this is how the 2 GHz line was employed to obtain the pulse compression performance of the 3.6 GHz line, as shown in the next section.

A Hewlett-Packard automatic network analyzer was used to measure the all-important phase characteristic of the 3.6 GHz dispersive line between 4.0 and 8.0 GHz at 20 MHz intervals. These experimental results were used to derive the coefficients of a best-rms-fit parabola, the ideal form of the phase-vs-frequency characteristic for a linear-FM dispersive filter. The phase error between the experimental measurements and the best-fit parabola was computed, and is plotted in **Figure 7**. The standard deviation of this error was found to be 14 degrees in the design band between 4.2 GHz and 7.8 GHz. This rms error reduces the ideal compressed-pulse amplitude by 3 percent. It is seen in **Figure 7** that a refinement in crimp positioning could probably reduce the phase error by a factor of 2 to 4, leaving only the fine-grain contributions.

DYNAMIC PERFORMANCE OF CRIMPED-COAX DISPERSIVE LINES

Figure 8 shows the block diagram for the experimental test of the dynamic performance of the chirp-LO and compressed-pulse dispersive lines described above A 25 psec pulse generator was used to impulse the 5 to 7 GHz chirp line. Directional couplers are used at the inputs of each line to isolate the reflected outputs from the line inputs. **Figure 9** (upper) shows the impulse-excited linear-FM output of the 7.0 to 5.0 GHz line used as a swept LO (about 9 nsec sweep time). **Figure 9** (lower) shows the compressed-pulse output of the 4.2 to 7.8 GHz compressive line with the time base corresponding to that of the swept LO wave form above. Note in **Figure 9** that the compressed-pulse peak-to-peak amplitude is almost 6 dB greater than the swept-pulse input above it. The swept-LO impulse input is seen to be less than −26 dB peak-to-peak at the compressed-pulse output (lower left of **Figure 9**). The effective bandwidth of the compressed pulse output is 2 GHz owing to the restricted 7 to 5 GHz sweep of the swept-LO signal.

Figure 10 (upper) is an expanded view of the compressed pulse showing a width of about 0.5 nsec, in expected correspondence with the effective compression ratio of 17. **Figure 10** (lower) compares the compressed-pulse shape after coiling the 5.0 to 7.0 GHz dispersive line. A small change occurs in the sidelobe structure of the compressed pulse, which could be further reduced by a more deliberate line-coiling procedure; the width and amplitude of the compressed pulse has not changed noticeably.

Figure 11 shows the effects on the compressed-pulse structural detail of temperature cycling from room temperature (upper left) to −65°C (lower left), to

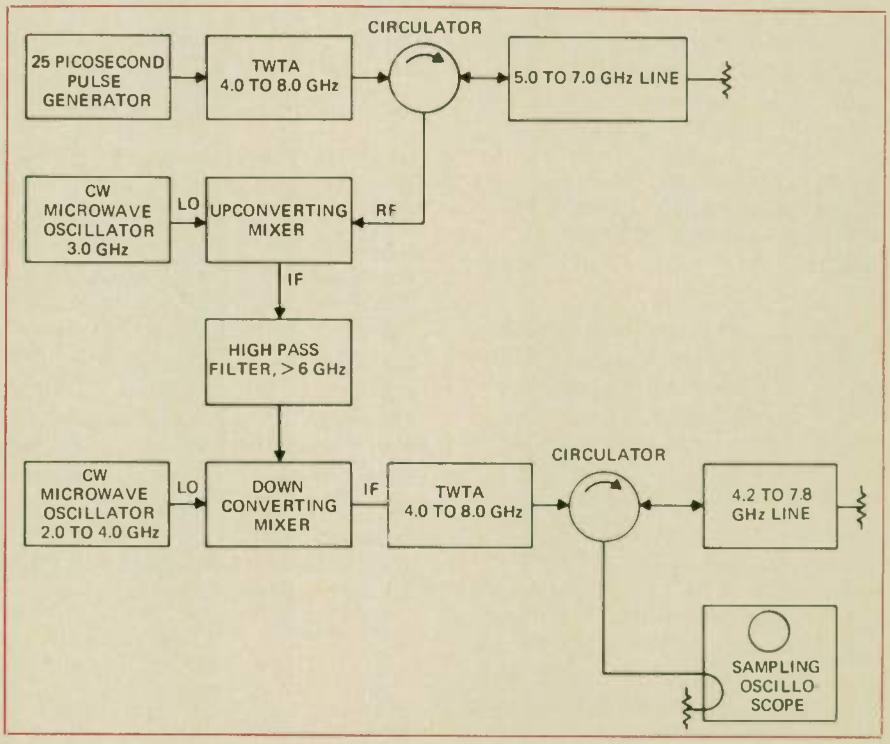
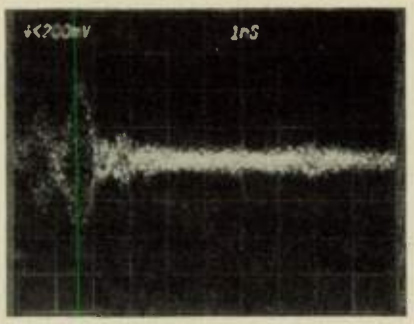


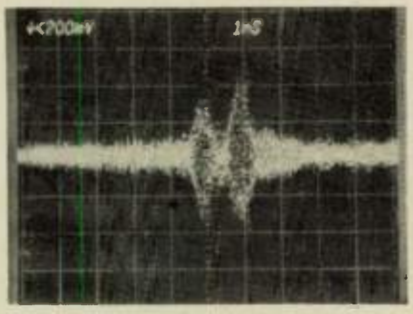
Fig. 12 Block diagram of test circuit to demonstrate the effect of CW signal tuning on compressed pulse characteristics, and of spectral resolution for simultaneous CW signals.

+125°C (upper right) back to room temperature (lower right). The peak-to-peak amplitude changes by perhaps 1 dB (gains were not monitored), but little change occurs in compressed-pulse width or sidelobe structure. Note that only one line of the pair was temperature cycled, the other remaining at room temperature. More complete temperature compensation for longer lines could be achieved by packaging both flat/coiled lines of the pair in close thermal contact within the same thermally insulated case. Temperature control is then unnecessary for maintaining designed compressed-pulse width and sidelobe levels.

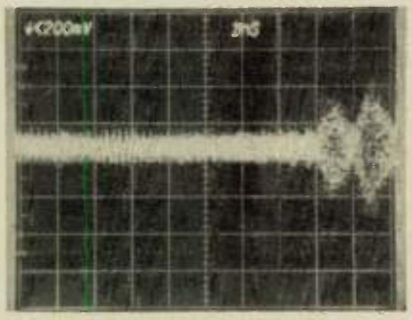
Figure 12 shows the block diagram for test of the swept-LO and pulse-compression lines for multiple variable-frequency signals tuned across the 2 GHz baseband width. The broadband MD-112-1 mixers are from Anzac and the 7904 main frame and sampling scope accessories are from Tektronix. Figure 13 shows the compressed-pulse output of two CW signals separated by 200 MHz, first at the lower end of the 2 GHz baseband, then at the middle, then at the high end of the 2-GHz baseband. The background noise level in these photographs is abnormally high because of lack of an optimal design of the feasibility demonstration circuit.



A. LOWER END OF 2.0 GHz BAND



B. MIDDLE OF 2.0 GHz BAND



C. HIGH END OF 2.0 GHz BAND

Fig. 13 Compressed-pulse outputs of two simultaneous CW signals separated by 200 MHz, and tuned across 2 GHz bandwidth (ref. Fig. 12).

RECEIVER DESIGN CONSIDERATIONS USING CRIMPED-COAX DISPERSIVE LINES

Semi-rigid coax transmission lines can be used to fabricate inexpensive broadband dispersive delay lines of small size and rugged construction for satisfying the component requirements of ECM/ESM pulse compression receivers. A first important receiver design consideration is that the frequency-dependent attenuation of propagation in the line must be compensated when determining the crimp depths along the line that provide the desired spectral weighting function required for sidelobe suppression, especially for lines of large time-bandwidth product. A large range of crimp reflection coefficients are possible for attaining this goal.

A second important difference in receiver design between using a conventional pulse-compression line and the present CRDL is that the latter is a single-port device and so has no natural isolation between the (simultaneous) input and output signals. The dynamic range of operation due to input-output leakage, R, (at IF, prior to detection) is therefore set in the CRDL by the relative amplitudes of these two signals, i.e.,

$$R = G_C + I_C - L_I - L_W,$$

$$L_I = L_T + L_R,$$

where G_C is the compression gain (~ 20 dB), I_C is the isolation furnished by the circulator inserted to separate input from output (~ 35 dB), L_I is the insertion loss of the line, and L_W is the weighting loss. For a single-line construction, L_I is made up of the transmission loss, L_T, at the center frequency of the line (i.e., that for transmission to the center of the line and back), and L_R is an additional "reflective" loss. L_R accounts for the impossibility of obtaining 100 percent reflection of the energy in each frequency interval without introducing second order effects on the dispersive characteristics of the line that could adversely affect the sidelobe structure of the compressed pulse.

As an example, consider a CRDL with 100 nsec dispersive delay and a bandwidth of 1 GHz; this line would comprise an approximately 35 foot length of 85 mil semi-rigid coax line. The line (50 nanoseconds single-way delay) after "crimping" will provide a nominal 100 nsec differential delay over a 1.2 GHz to 2.2 GHz bandwidth. Crimp reflection loss, L_R, will be approximately 2 dB without weighting, i.e., in the center of the filter band. Cable transmission loss, L_T, will be 14 dB, 9 dB and 0 dB for the 1.2 GHz, 1.7 GHz, and 2.2 GHz frequency points respectively with the low-frequency reflective crimps located at the far end of the cable. The dynamic range then consists of the compression gain (20 dB), plus the directional coupler isolation (35 dB), minus the midband transmission loss (9 dB), the crimp reflection loss (2 dB) and the weighting loss (6 dB), for an overall dynamic range of 38 dB. This sample line has a volume of 50 cc and a weight of 200 grams. An additional 4 dB of dynamic range is possible using 141 mil coax, giving a line of volume 140 cc and of weight, 500 grams.

To avoid altogether the input-output leakage limitation on dynamic range, one method used is to apply the broadband capability of the CRDL technique, using two pulse compression lines, each with 2 GHz bandwidth (2.2 GHz to 4.2 GHz), and with 100 nsec

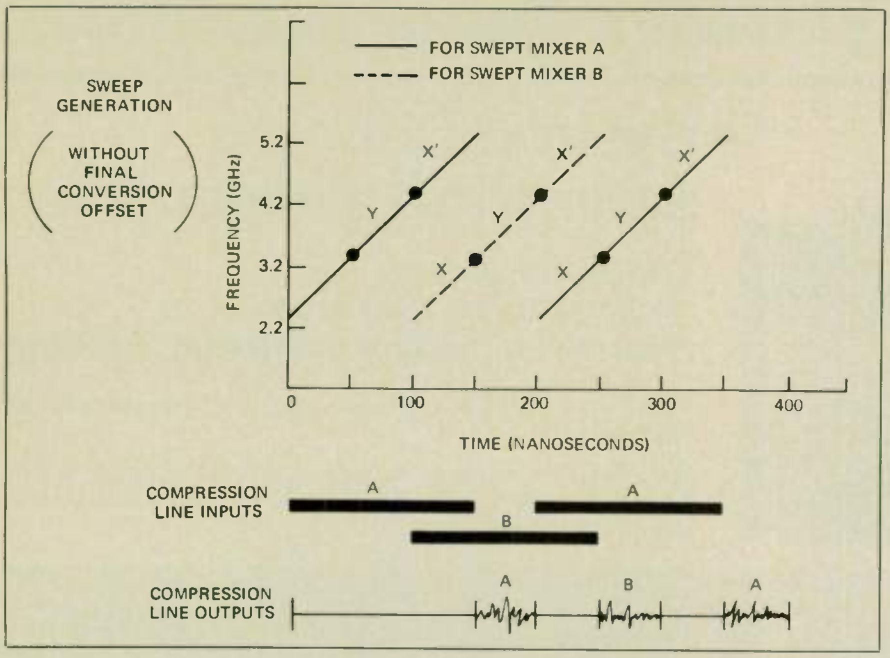


Fig. 14 Use of sectioned swept-LO generation lines and dual pulse-compression lines for reducing input-to-output leakage limitation on dynamic range of operation.

differential delays. In addition, there would be a 50 nsec total nondispersive delay segment (25 nsec single-way delay) at the input/output end of each line. Each line would consist of 52.5 feet of 85 mil coax. Total volume and weight of the two lines would be 150 cc and 600 grams. Insertion loss (including 2.0 dB reflection loss) would be 27.0 dB at the midpoint of the band, 3.2 GHz. The LO sweep would cover 3 GHz in 150 nanoseconds and be generated by two 1.0 GHz bandwidth, 50 nsec differential-delay lines. The sweeps are depicted in Figure 14, where X and Y represent the two 1.0 GHz line outputs, with X′ being derived from X, upconverted by 2.0 GHz. For this system configuration, the instantaneous signal bandwidth remains at 1.0 GHz, but the signals are dispersed over a 2.0 GHz bandwidth during the 100 nsec fill time. Frequency resolution remains the same (reciprocal of fill time, 100 nsec), but the compression line outputs, as shown in Figure 14, are compressed within 50 nsec segments for a 1 GHz pulse compression bandwidth. The additional 50 nsec nondispersive delay in the compression lines allows for non-simultaneous input/output operation, and removes input/output leakage as a limitation to dynamic range. The LO-sweep CRDL lines would each consist of 17.5 feet of 85 mil coax. Total volume and weight of the two lines would be 50 cc and 200 grams. Low frequency (1.2 GHz) line loss and time dispersion loss would be 9 dB and 19 dB respectively, which with other attendant losses leads to an impulse input requirement of +8 dBm to preserve the sensitivity potential (−80 dBm) of the receiver. The average power requirement per line would be −11 dBm.

Since the insertion loss of a single-line CRDL pulse compression receiver directly affects the leakage-limited dynamic range, and since the insertion loss is in large part determined by the propagation attenuation of the center frequency reflected from the midpoint of the line, study has been made of a "bifurcated" CRDL wherein the pulse compression line is divided at the midpoint, and the two halves are simultaneously filled from separate half-length upswept and down swept signals; the reflected-signal outputs are added to constitute the composite compressed pulse. The bifurcated design significantly reduces the insertion loss for increased dynamic range, reduces the fill time of the line, and makes possible wideband operation at a much reduced time for fully-processed 100% detection probability.

ACKNOWLEDGEMENTS

The author wishes to acknowledge the contributions and aid of several Sanders colleagues including C. Field, E. Hollis, S. Risteen, C. Stromswold, and H. D. Witthun.

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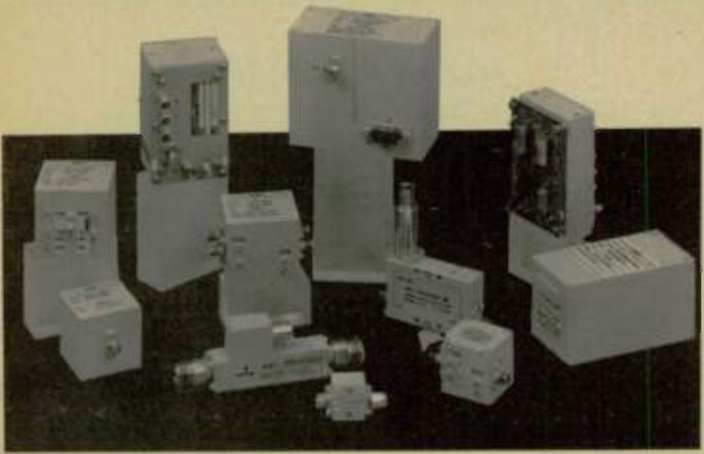
Harrison W. Fuller received his B.S. degree in Physics from Worcester Polytechnic Institute, and his Ph.D. in Applied Physics from Harvard. He joined Sanders Associates, Inc., in 1967 where he is currently a Scientific Fellow in Physics, and Manager of New Technology Applications.

Systron-Donner Advanced



Systron-Donner's setting the standards for:

- Yig-tuned devices
- Voltage-tuned oscillators
- Mechanical and voltage-tuned oscillators

Yig-tuned oscillators are available in transistor and Gunn-diode types.

Specifications are given for typical standard models. In most cases, standard units with higher (100 mW) or lower (10 mW) power are also available.

MODEL:	Frequency Range (GHz)	Power Output Min. (mW)	Power Variation vs. Frequency	Spurious Signals:		Residual FM 1 Hz—30 kHz	Frequency Stability:			
				Harmonic Min.	Non-Harmonic Min.		Vs. Temperature	Vs. Power Supply	Vs. Load Variation	Hysteresis
SDYX-3038	0.5—1.0	20	5 dB p-p	12 dBc	60 dBc	10 kHz p-p	0.03%/°C	1 MHz/V	500 kHz	2 MHz
SDYX-3034	1.0—2.0	20	5 dB p-p	15 dBc	60 dBc	10 kHz p-p	0.03%/°C	1 MHz/V	500 kHz	2 MHz
SDYX-3034-114	0.5—2	20	6 dB p-p	12 dBc	60 dBc	10 kHz p-p	0.03%/°C	1 MHz/V	1 MHz	4 MHz
SDYX-3036	2.0—4.0	20	5 db p-p	15 dBc	60 dBc	10 kHz p-p	0.03%/°C	1 MHz/V	500 kHz	4 MHz
SDYX-3036-125	1.0—4.0	20	7 dB p-p	15 dBc	60 dBc	10 kHz p-p	0.03%/°C	1 MHz/V	500 kHz	6 MHz
SDYX-3039-107	2.0—6.0	10	7 dB p-p	15 dBc	60 dBc	10 kHz p-p	0.03%/°C	1 MHz/V	6 MHz	7 MHz
SDYX-3000	8.0—12.4	25	6 dB p-p	30 dBc	60 dBc	10 kHz p-p	0.01%/°C	10 MHz/V	10 MHz	10 MHz
SDYX-3001	12.4—18.0	25	6 dB p-p	30 dBc	60 dBc	10 kHz p-p	0.01%/°C	10 MHz/V	10 MHz	15 MHz
SDYX-3001-111	8.0—18.0	10	8 dB p-p	30 dBc	60 dBc	10 kHz p-p	0.01%/°C	10 MHz/V	10 MHz	25 MHz
SDYX-3003	18.0—26.5	10	6 dB p-p	20 dBc	60 dBc	10 kHz p-p	0.01%/°C	10 MHz/V	10 MHz	35 MHz
SDYX-3004	26.5—40.0	5	8 dB p-p	20 dBc	60 dBc	10 kHz p-p	0.01%/°C	20 MHz/V	20 MHz	100 MHz

YIG DRIVERS

Any Systron-Donner YIG device may be ordered with a matched YIG driver to provide accurate voltage/frequency conversion and to facilitate installation of the YIG device in a system. Two types of drivers are available: a standard version and a high-stability version. Both types operate with input power of ±15 V (±20 or ±12 V available with some frequencies on special order), control voltage of 0 to 10 V, and minimum input impedance of 10kΩ. Units meeting either commercial or military environmental requirements may be provided. Options available with 12-bit digital tuning.

VOLTAGE-TUNED OSCILLATORS

MODEL:	Frequency Range (MHz):	Power, Min. (mW):				Spurious Signals:		Residual FM:	Amplitude Control:	
		Leveled	Power vs. Frequency	Unleveled	Power vs. Frequency	Harmonic Min.	Non-Harmonic Min.	in 1 Hz-30 kHz Band	Full Output	Down 40 dB
SDVX-2011	470-1030	20	<4 dB	30	<4 dB	20 dBc	60 dBc	2 kHz p-p	−5 V @ 30 mA	0 V @ 0 mA
SDVX-2012	940-2060	20	<4 dB	30	<4 dB	20 dBc	60 dBc	4 kHz p-p	−0.6 V @ 0 mA	+5 V @ 30 mA
SDVX-2013	1240-2060	20	<4 dB	25	<4 dB	20 dBc	60 dBc	4 kHz p-p	−0.6 V @ 0 mA	+5 V @ 30 mA
SDVX-2108	0.1-32	20	±0.3 dB[2]	25	<4 dB	20 dBc[1]	50 dBc	4 kHz p-p	−5 V @ 15 mA	0 V @ 0 mA
SDVX-2110	8-112	20	±0.1 dB[1]	30	<4 dB	30 dBc[1]	60 dBc	2.5 kHz p-p	−5 V @ 15 mA	0 V @ 0 mA
SDVX-2111	25-305	20	±0.2 dB[2]	30	<4 dB	30 dBc[1]	60 dBc	1.5 kHz p-p	−15 V @ 15 mA	0 V @ 0 mA
SDVX-2112	90-510	20	±0.2 dB[2]	30	<4 dB	30 dBc[1]	60 dBc	1.5 kHz p-p	−5 V @ 15 mA	0 V @ 0 mA
SDVX-2000	235-515			50	<4 dB	20 dBc	60 dBc	1 kHz p-p		
SDVX-2001	470-1030			50	<4 dB	20 dBc	60 dBc	2 kHz p-p		
SDVX-2002	940-2060			50	<4 dB	20 dBc	60 dBc	4 kHz p-p		
SDVX-2003	1340-2460			50	<4 dB	20 dBc	60 dBc	4 kHz p-p		

1) @ 20 mW leveled 2) Internal Detector

MECHANICAL AND VOLTAGE-TUNED OSCILLATORS

MODEL:	Mechanical Tuning Frequency Range (GHz):	Voltage Tuning Bandwidth (MHz)	Power Output (mW)	Power vs. Frequency (dB)	Spurious Signals:		Voltage Tuning Range (volts)	DC Power (VDC) See Note 1
					2nd Harmonic	Non-Harmonic		
SDYX-2015-105	5.925-6.425	120	50	1	30 dBc	70 dBc	2-25	+15 @ 500 mA
SDVX-2016-110	7.25-7.75	120	50	1	30 dBc	70 dBc	2-25	+15 @ 500 mA
SDVX-2016-114	7.9-8.4	120	50	1	30 dBc	70 dBc	2-25	+15 @ 500 mA
SDVX-2016-107	8.5-9.1	120	50	1	30 dBc	70 dBc	2-25	+15 @ 500 mA
SDVX-2016-108	9.0-9.6	120	50	1	30 dBc	70 dBc	2-25	+15 @ 500 mA
SDVX-2017-106	10.7-11.2	100	30	1	20 dBc	70 dBc	2-30	+15 @ 600 mA
SDVX-2017-107	11.2-11.7	100	30	1	20 dBc	70 dBc	2-30	+15 @ 600 mA
SDVX-2017-112	12.7-13.2	100	30	1	20 dBc	70 dBc	2-30	+15 @ 600 mA
SDVX-2017-120	14.0-14.5	100	30	1	20 dBc	70 dBc	2-30	+15 @ 600 mA

Note 1: Current is steady state. Surge current will be 70% higher.

Components: Update 1980

YIG-TUNED FILTERS

Systron-Donner SDYF-4000 Series reciprocal bandpass filters are available in one, two, three, four, and dual-two stage versions in single bands and multi-octave versions from 0.5 to 40 GHz. These filters are ideal for use in receiver systems, frequency synthesizers, or test sets for preselection, signal sorting, or any other application in which a tunable filters must pass a desired signal or band of signals with minimal attenuation and reject undesired out-of-band signals.

SDYF-4000 Series filters may be ordered individually, with a YIG driver, as part of a tracking filter/oscillator/driver assembly meeting specific system applications.

		MODEL:	Frequency Range *(GHz)*	Bandwidth *(MHz, Min.)*	Insertion Loss *(dB, Max.)*	O.R.I. *(dB, Min.)*	O.R.S. *(dB, Min.)*	PB Ripple & Spurious *(dB, Max.)*	Linearity *(MHz, Nom.)*	Hysteresis *(MHz, Nom.)*
BANDPASS	Two-Stage	SDYF-4021	0.5–1	12	6.0	40	25	2.5	±2	4
		SDYF-4022	1–2	20	3.0	40	25	2.5	±2	4
		SDYF-4023	2–4	20	3.0	50	25	2.5	±3	6
		SDYF-4024	4–8	25	3.0	50	25	2.5	±5	8
		SDYF-4025	8–12.4	30	3.0	50	25	2.5	±8	15
		SDYF-4026	12.4–18	30	3.0	40	30	2.5	±10	15
		SDYF-4027	18–26.5	35	4.0	40	30	2.5	±15	20
	Three-Stage	SDYF-4028	0.5–1	12	6.0	70	35	2.0	±2	4
		SDYF-4029	1–2	18	5.0	70	40	2.0	±2	4
		SDYF-4030	2–4	20	4.0	70	40	2.0	±3	6
		SDYF-4031	4–8	25	4.0	70	40	2.0	±6	8
		SDYF-4032	8–12.4	25	4.0	70	40	2.0	±10	15
		SDYF-4033	12.4–18	30	4.0	70	40	2.5	±10	15
		SDYF-4034	18–26.5	35	5.0	70	40	2.5	±15	20
	Four-Stage	SDYF-4035	0.5–1	10	8.0	70	40	2.8	±2	4
		SDYF-4036	1–2	15	6.0	70	50	2.8	±2	4
		SDYF-4037	2–4	15	5.0	70	50	2.8	±3	6
		SDYF-4038	4–8	20	5.0	70	50	2.8	±8	8
		SDYF-4039	8–12.4	20	5.0	70	50	2.8	±10	15
		SDYF-4040	12.4–18	25	5.0	70	50	2.8	±10	15
		SDYF-4041	18–26.5	30	5.5	70	50	2.8	±15	20
	Dual Two-Stage (Per Channel)	SDYF-4042	0.5–1	12	6.0	40	25	2.0	±2	4
		SDYF-4043	1–2	20	3.0	40	25	2.8	±2	4
		SDYF-4044	2–4	20	3.0	50	25	2.5	±3	6
		SDYF-4045	4–8	25	3.0	50	25	2.5	±5	8
		SDYF-4046	8–12.4	25	3.0	50	25	2.5	±8	15
		SDYF-4047	12.4–18	25	3.0	40	25	2.5	±10	15
		SDYF-4048	18–26.5	30	4.0	40	25	2.5	±15	20
	Multi-Octave	SDYF-4000	1.8–18	20	5.0	70	60	1.5	±10	15
		SDYF-4000-102	1.8–26.5	15	8.5	70	60	1.5	±20	20
		SDYF-4000-113	2–18	30	3.0	40	40	1.5	±10	15
		SDYF-4000-114	2–12	30	3.0	40	40	1.5	±10	15
WIDE-BAND		SDYF-4235	8–18	250	7.5	70	50	2.8	±15	20

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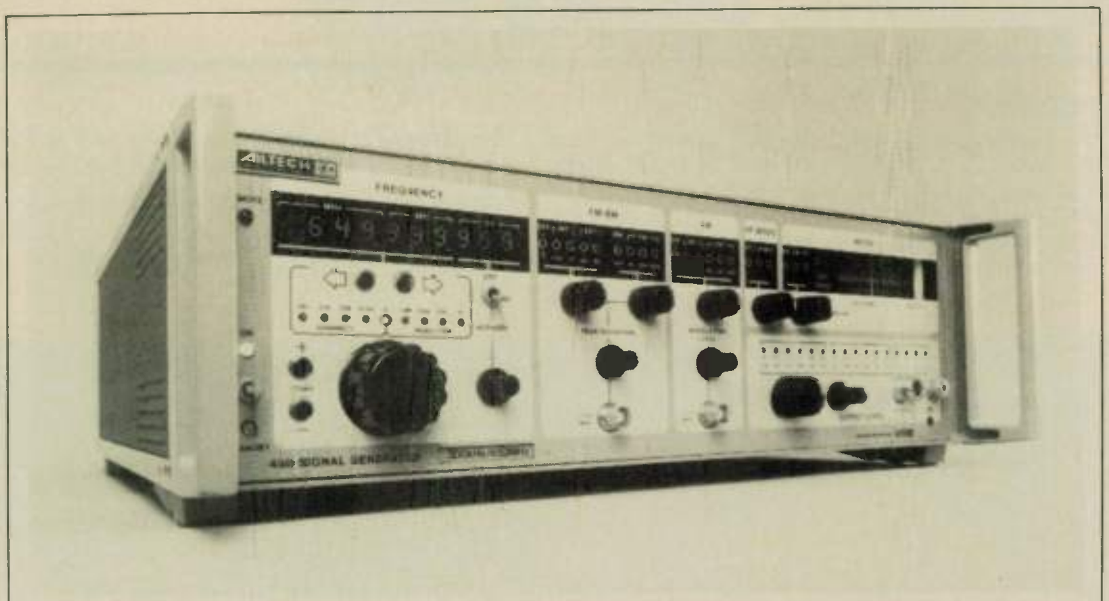
Circle 32 for Demonstration
Circle 57 for Literature



The Quiet Synthesizer

AILTECH
Ronkonkoma, NY

Under manufacturing rights acquired from ADRET Electronique of France, AILTECH is offering its Model 460 frequency synthesizer, a programmable generator combining the desirable characteristics of synthesizers and cavity-based signal generators.

The Model 460 delivers 100 mW over an operating frequency range of 300 kHz to 650 MHz. An internal doubler option extends the range to 1300 MHz with 10 mW available from the doubler. 1 Hz frequency resolution is available throughout the range including the doubled portion and output level is resolved to 0.1 dB. Simultaneous AM and FM or AM and phase modulation are available and all functions are programmable.

While offering the features of a synthesizer, the instrument achieves a phase noise of −132 dBc/Hz at 12.5 kHz from the carrier and this decreases to −145 dBc/Hz at 2 MHz from the carrier. The cavity-like performance of the model 460 does not depend on new component technology but on the method of synthesis which employs three phase lock loops in an indirect design controlled by a microprocessor.

DESIGN

The heart of the instrument design is an 80 MHz crystal controlled oscillator phase-locked to a 10 MHz oven controlled frequency standard. At 10 kHz from the carrier, the 80 MHz oscillator noise is −165 dBc.

The intermediate phase locked loop which generates frequencies from 300 to 670 MHz in 10 MHz increments has a loop bandwidth sufficiently wide so that the noise on the output is essentially that of the multiplied 80 MHz oscillator. The low multiplication factor avoids severe degradation of the 80 MHz oscillator noise characteristics. The combination of the wide loop bandwidth and high reference frequency succeed in imposing the excellent noise characteristics of the reference oscillator on the output VCO.

In the 10 MHz loop, the 80 MHz oscillator drives a harmonic generator, and a varactor tuned filter picks out harmonics from 320 to 640 MHz which drive the RF port of the mixer. The LO port is driven by a sample of the 300 to 670 MHz VCO. Since the harmonics are spaced at 80 MHz intervals, the VCO frequency is never separated by more than 40 MHz from a harmonic and the mixer output is limited between 0 to 40 MHz. A 40 MHz low pass filter follows the mixer output, and the output drives one port of a sampling phase detector. The 10 MHz sampling frequency is developed from an eight-to-one divider from the 80 MHz oscillator. The 10 MHz sampling rate allows the VCO to be locked in 8 intervals of 10 MHz between the 80 MHz harmonics providing 10 MHz steps over the 300 to 670 MHz range.

The output loop shown in **Figure 1** tunes the 320 to 650 MHz range and provides the higher resolution steps between the 10 MHz increments while still preserving the low noise performance of the 300 to 670 MHz stepped oscillator. The two oscillators are fed to both input ports of a mixer. The mixer output is passed through a bandpass filter covering 20 to 25 MHz,

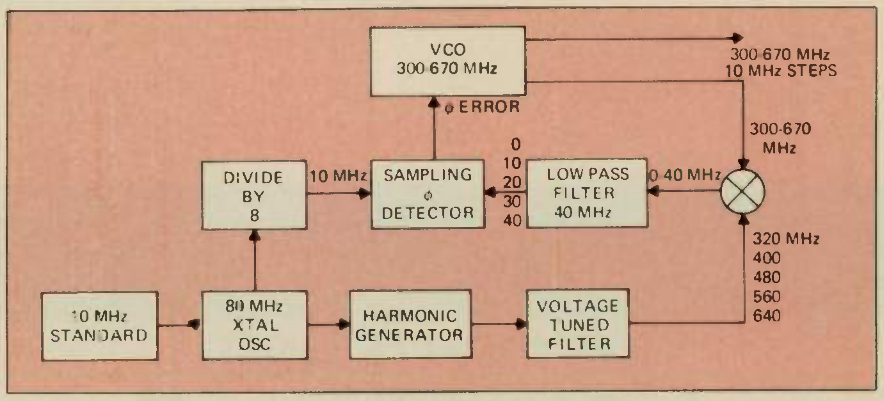


Fig. 1 10 MHz loop.

(continued on page 66)

FERRITE ISOLATORS AND CIRCULATORS

STANDARD DESIGNS

POPULAR OCTAVE BANDS — STANDARD DESIGNS

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

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

S-T-R-E-T-C-H OCTAVE BANDS — STANDARD DESIGNS

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

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

POPULAR NARROW BAND — STANDARD DESIGNS

Frequency (GHz)	Model No.	Isolation (dB min.)	Insertion Loss (dB max.)	VSWR (max.)	Size Height	Size Width	Size Thickness
.95 - 1.225	T-0S23T-2	20	0.5	1.25:1	1.20	1.20	0.75
1.2 - 1.6	T-1S23T-7	17	0.5	1.35:1	1.25	1.25	0.70
1.9 - 2.3	T-1S13T-2	20	0.4	1.30:1	1.25	1.25	0.75
2.2 - 2.3	T-2S03T-2	20	0.4	1.35:1	1.00	1.00	0.62
3.7 - 4.2	T-3S13T-9A	25	0.25	1.10:1	0.75	0.75	0.50
4.4 - 6.5	T-4S33T-1	17	0.5	1.35:1	0.75	0.75	0.50
5.9 - 6.4	T-5S03T-3A	26	0.3	1.10:1	0.75	0.75	0.69
7.0 - 11.0	T-7S43T-6	28	0.4	1.10:1	0.85	0.75	0.60
8.0 - 12.4	T-8S43T-1A	17	0.4	1.35:1	0.78	0.63	0.70
12.4 - 18.0	T-12S43T-8	18	0.5	1.30:1	0.68	0.51	0.56
18.0 - 26.5	T-18S33T-7	16	1.0	1.50:1	0.68	0.51	0.53

SPECIAL DESIGNS

- Multi-Junction 4, 5, 6 or more Ports
- Stripline Tab Drop-Ins
- Special RF Power Requirements
- N or TNC Connectors (lower frequency units only)

TELEDYNE MICROWAVE

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Circle 33 on Reader Service Card

which is the maximum offset range between the two oscillators. The filter output feeds one port of a phase comparator and a 20 to 25 MHz signal from the third loop feeds the other port. The loop covers the 20 to 25 MHz range in steps as small as 1.0 kHz. As will be explained, the noise of this loop is -140 dBc/Hz at 10 kHz from the carrier and has little effect on output noise level. **Figure 2** illustrates the frequency relationships between the three loops as the output frequency is changed.

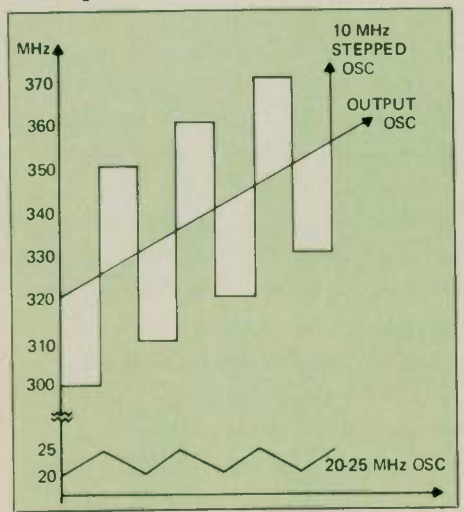

Fig. 2 Frequency plan.

Since the reference frequency for the output loop is relatively high, 20-25 MHz, a wideband loop is used in the output without the problem of filtering the reference sidebands. Thus the low noise of the 300 to 670 MHz stepped oscillator is imposed on the output VCO, and the noise floor is -145 dBc/Hz, approaching the performance of a cavity oscillator.

Since the 20 to 25 MHz loop output is added directly to the 320 to 650 MHz output loop, its noise performance is of primary consideration. This loop operates as a divide-by-n synthesizer with a 1 kHz resolution and, to avoid the shortcomings of large division ratios and 1 kHz sidebands, a low noise VCO, permitting use of a very narrow loop bandwidth to reduce divider noise and reference sidebands, is employed. The low noise VCO designed to cover the 80 to 100 MHz range achieves a noise of -135 dBc/Hz at a 10 kHz offset. The VCO drives a divide by 4 to produce the 20 to 25 MHz, and the division improves the noise an additional 12 dB to less than -145 dBc/Hz.

The loop design is slightly complicated by the fact that a directional frequency change is required in covering a 10 MHz range. As the 1 MHz digit is tuned from 0 to 4.999 MHz, the oscillator frequency increases from 20 to 24.999 MHz. As the 1 MHz digit is tuned from 5 to 9.999 MHz, the frequency decreases from 25 to 20.001 MHz. If this reversal were uncompensated, it would result in a phase reversal of the dc FM modulation which is coupled through the 2 MHz input of the loop, as the generator is tuned across every 4.999 to 5 MHz transition. Operation of the frequency vernier would also be reversed. Compensation is achieved by interleaving a second loop with a VCO operating at 32 to 42 MHz, and 48 to 58 MHz. The two ranges are switched at the 4 and 5 MHz transition and provide low and high frequency mixing with the 20 to 25 MHz signal. The low and high mixing provides a fixed 4 MHz mixer output with the two phases required for compensation.

The 32 to 58 MHz oscillator tunes in 1 kHz steps. Since it is followed by a divide-by 2, it forces the 20 to 25 MHz signal to tune in 500 Hz steps. This resolution is necessary for generators equipped with the doubler option to 1300 MHz, so as to allow 1 kHz steps over the entire frequency range.

A 2 MHz input signal to the 20-25 MHz loop is developed directly from the 10 MHz frequency standard when the vernier is not used. For vernier operation, a 2 MHz signal with a tuning range of 4 kHz is required to interpolate the 1 kHz steps (in the 80-160 MHz output range, the output VCO 320 to 640 MHz is divided by 4). The noise contribution by the 2 MHz oscillator into the 20 to 25 MHz loop is so small as to be undetectable when operating with the vernier either on or off.

Since the output VCO covers the 320 to 650 MHz range, the lower frequencies are generated by successive binary division. The ranges covered by division are 80-160 MHz and 160 to 320 MHz. One advantage of the division is that the noise decreases by 6 dB, after each division, thereby improving the performance in the lower frequency ranges. Frequencies below 80 are generated by a heterodyning process because of the increased difficulty of filtering harmonics using further binary division.

THE RESULTS

The objective of the design was to obtain cavity like performance from a synthesizer by imposing the low noise of an 80 MHz crystal oscillator on a VCO which tunes the 320 to 650 MHz range. The degree of success is shown in the comparison noise plot of Figure 3. Cavity signal generator performance below 2 kHz is not shown, however, the new 460 synthesizer has 10 dB lower noise out to an offset frequency of 20 kHz, where both are about the same at -132 dBc/Hz. The noise floor of the cavity, -145 dBc/Hz, is reached sooner than the synthesizer at 100 kHz offset. The synthesizer characteristic flattens out at 132 dBc/Hz from 10 kHz to 400 kHz, and then decreases to the -145 dBc/Hz floor at 2 MHz. The small advantage of the cavity signal generator in the 20 kHz to 2 MHz offset range has little significance for narrow channel receiver measurements (up to 20 kHz offsets) where the 460 synthesizer has the performance advantage. For image and spurious measurements, the wideband noise floor beyond several megahertz is important, and here both instruments are about equal.

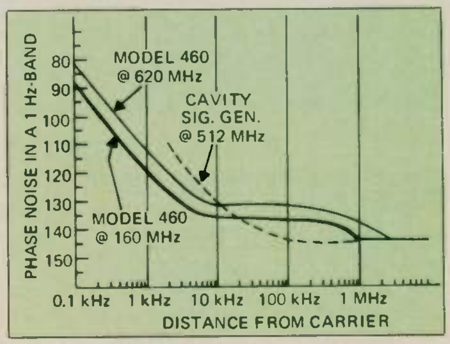

Fig. 3 Phase noise of 460 and cavity signal generator.

The instrument provides AM, FM, and Phase Modulation. AM at 100% modulation is available to output power levels up to +13 dBm; distortion is less than 3% for 80% and modulation at output levels to + 14 dBm. Separate modulators are used to allow simultaneous AM and FM or AM and Phase Modulation.

In CW mode, the maximum output power is 100 milliwatts (+20 dBm) and is adjustable to -140 dBm using a stepped attenuator and vernier. A meter provides calibration between the 10 dB steps and output leveling is less than ±0.5 dB.

Frequency is controlled using a spin-wheel with adjustable resolution from 1 kHz to 1 MHz. A vernier interplates the 1 kHz steps to a resolution of 1 Hz and the actual output frequency is indicated on an LED display.

All functions are controllable using the GPIB (IEEE-488) option. Frequency can be programmed to 1 Hz resolution, level to 0.1 dB resolution, AM level is 1% resolution, and FM as Phase Modulation to 1/300 of full scale resolution.

The instrument is packaged in a 5¼" high rack mountable cabinet and weighs 51 pounds.

Circle 119 on Reader Service Card

Finally a way to make noise figure measurements– 10 MHz to 18 GHz–that are accurate and repeatable.

HP's new 346B Noise Source can cut your uncertainty in half.

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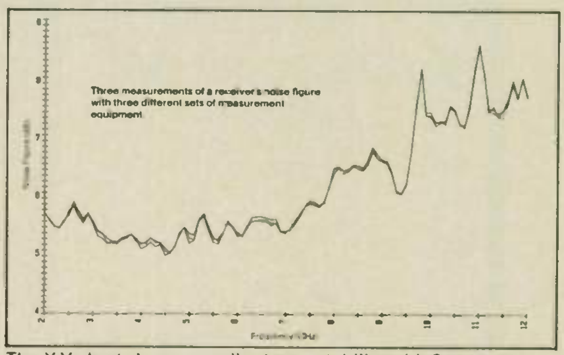

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Technology Notes

Signal Processing For Satellite Mapping

by Lloyd W. Martinson
Richard P. Perry

(This brief report summarizes the problems and findings of a recent technology study undertaken by engineers at RCA Moorestown for the Jet Propulsion Laboratory, sponsored by NASA)

Introduction

The formation of mapping images with a satellite-borne synthetic aperture radar (SAR) presents challenging signal processing problems. Current processors are located on the ground and process the raw, unprocessed data transmitted from the spacecraft in non-real time. That is, images can be formed from only a small portion of the total swath scanned by the radar. In addition, the wide bandwidth, unprocessed signal places great demands on the communication link. Typical SAR processors use multiple looks at an image field to reduce speckle effects and thus improve image quality. This multiple look processing reduces the required bandwidth of the processed SAR data by a factor equal to the number of looks integrated. If the multiple-look processing could be done on board the spacecraft, a similar reduction would be achieved in the communication link requirements.

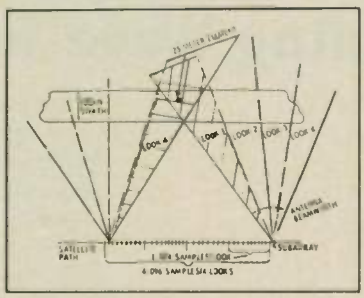
A synthetic aperture processing procedure using subarrays (which RCA calls the step transform algorithm) has three basic operations:

1. Focus data over short subarray time.
2. Store subarray data over large array length.
3. Combine subarray outputs to form large array.

The System Definition Problem

A typical SAR azimuth processing system requirement is represented by NASA's SEASAT and is shown diagrammatically in Figure 1. As the satellite maps an image field, four separate looks are generated for each image sample. Each look consists of 1024 azimuth samples which cover a 100 km range swath. Four looks require a total of 4096 azimuth samples. One fourth of the total of 1024 (or 256) azimuth elements complete the four-look integration during a look. Thus the basic word data rate from the input to the output of the processor is reduced by a factor of four.

The SAR azimuth processor must process 1024 spatial element (azimuth) samples of a synthetic array into an equal number of azimuth angle elements (beams) for each range element in the 100 km swath. The process is complicated by the fact that during the four-look interval, a particular image element will walk, or migrate, in range up to a maximum of 128 range elements.



The objective of the step transform subarray approach is to subdivide the SAR processing in a manner which permits range migration compensation to be applied on the coarse subarray resolution element rather than for each individual azimuth and range sample. Care must be taken to minimize image smearing during the integration of subarrays.

The Implementation

The power dissipation of a digital system is essentially determined by the power-delay product of the technology employed, assuming an efficient design. The high speed and low power of CMOS/SOS technology makes it an attractive choice for spaceborne LSI and VLSI systems. The primary hardware constituent of a SAR signal processor is memory. CMOS/SOS RAM's on the market today consume about 1 microwatt per bit. The total bulk memory in a 20 km swath SAR baseline processor is about 25 x 10^6 bits, which projects a system memory power dissipation of only 25 watts. The rapid advancements in commercial memory device capacity promises even lower power and greater density as the integrated circuit feature sizes are reduced from the 3 to 6 microns of today to less than 1 micron in the future.

A six-stage pipeline FFT structure is employed in the SAR processor to achieve real-time processing rates. Optimum performance capability in the FFTs is obtained by using a special floating point architecture developed by RCA.

The study verified that the step transform subarray technique is a viable approach in a satellite-borne SAR processor. Resolution achievable in SAR images with the baseline range walk compensation technique is within two percent of theoretical. Further study to be undertaken includes detail design and system simulation and development of actual programmable SAR functional units for future SAR system needs.

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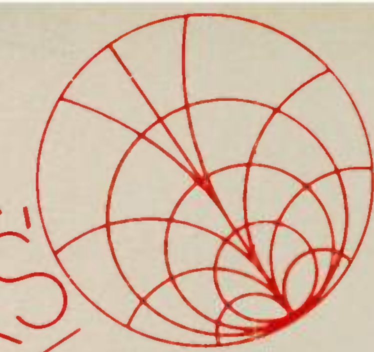

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Devices

TWT OFFERS 250 W IN K_u-BAND

TWT, Model 881H provides more than 250 W of CW power in the 14.0-14.5 GHz band. The tube is a metal-ceramic design with PPM focusing and forced-air cooling. A modulating anode is utilized for fast turn-on and turn-off of beam current during normal operating sequencing and under fault conditions. Projected cathode life of greater than 25,000 hours is offered. Tube body is conduction-cooled through its base-plate support, and an integral collector thermal switch is provided to allow tube turn-off under excessive temperature conditions. Bandwidth is 500 MHz minimum, gain at rated power output is typically 45 dB, and nominal efficiency is 25%. Size: 21.6" L x 5.2" W (54.9 cm x 13.2 cm). Weight: 11 lbs. (5 kg). **Hughes Aircraft Co., Electron Dynamics Div., Torrance, CA. (213) 534-2121. Circle 116.**

LOW DIVERGENCE CW LASER

Model LCW-30 is a low divergence CW laser combining multimode CW laser with an NSG SELFOC lens. It offers less than 7 milliradians divergence at 5 mW output. Typical characteristics include a threshold of 100 mA and an operating current of 135 mA at a peak wavelength of 820 nm. Price: $500 each in single quantities. Del: 4-6 wks. ARO. **Laser Diode Laboratories, Inc., New Brunswick, NJ. Richard Klein, (201) 249-7000. Circle 147.**

Materials

GaAs FET CHIP CARRIERS

Gel-Pak is a chip carrier designed for handling and shipping devices such as GaAs FET's. Product incorporates non-slip non-adhesive Si elastomer, which has been cast into a transparent hinged container. Chips loaded onto this flat silicone surface are held tightly in position, but are removed easily by vacuum, or tweezer. Chips are immobilized on surface of elastomer by use of separator fabric and compression foam. Gel-Pak carriers can be supplied with company logo or other identifying markings and a single carrier will accommodate a variety of chip sizes. **Gel-Pak Carrier Systems, Stanford, CA. Victor E. Althousc, (415) 941-9347. Circle 114.**

(continued on page 78)

SERIES OF BROADBAND ABSORBING MATERIALS

ECCOSORB® CV is a series of electrically tapered broadband absorbing materials intended for use in anechoic chambers. Power reflectivity of 0.01% (-40 dB) and less is assured at frequencies above the low-frequency limits (6-9 GHz) to frequencies above 100 GHz. At frequencies two octaves below the lower limit of the -40 dB region, reflectivity of these materials remains better than -29 dB. Models CV-3, CV-4, and CV-6 are made from lightweight artificial dielectric loaded flexible foam. Front surface is rippled for optimum impedance match. Size: 24" x 24" per sheet. Thickness: 3 to 6 inches, depending upon model. All materials are self-extinguishing. **Emerson & Cuming, Canton, MA. Jeanne B. O'Brien, (617) 828-3300. Circle 113.**

Instrumentation

THERMOCOUPLE POWER SENSOR OFFERS LOW SWR, CALIBRATION DATA

Model 8485A is a coax power sensor capable of making measurements from 50 MHz to 26.5 GHz in a power range of -30 dBm to + 20 dBm. SWR specification is 1.25 max. at 26.5 GHz. Each sensor is calibrated for CAL FACTOR, which is plotted at 34 frequencies on an attached label and is traceable to the US National Bureau of Standards. The measurement root sum of squares uncertainty at 26.5 GHz is ± 3.21%. Input connector is APC-3.5 and it mates directly with SMA. Price: $700 per power sensor. Del: 12 wks. **Hewlett-Packard Co., Palo Alto, CA. (415) 857-1501.**
Circle 117.

PROGRAMMABLE, BROADBAND NOISE GENERATOR SERIES

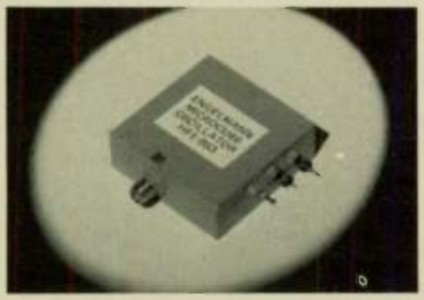
Model PNG 5100 through 5110 is a series of noise generators which feature full noise source capability from 10 Hz to 1 GHz in a compact instrument or rack mounted, 7-inch high, panel. Output is greater than +10 dBm (10 mW) across the given frequency range. IEEE-488 Interface Bus remote programming is featured, with on and off standby, external pulse and attenuation control, as well as local and remote capability. Series has the following standard frequency bands: 10 Hz - 20 kHz, 10 Hz - 100 kHz, 10 Hz - 500 kHz, 100 Hz - 3 MHz, 100 Hz - 10 MHz, 100 Hz - 25 MHz, 100 Hz - 100 MHz, 1 MHz - 200 MHz, 1 MHz - 500 MHz, and 10 MHz - 1 GHz. Optional frequency bands are available up to 40 GHz. Micronetics Inc., Norwood, NJ. (201) 767-1320. Circle 148.

Components

SEAMLESS FLEXIBLE WAVEGUIDE

Two styles of seamless flexible waveguide are offered: WR 90, with an 8.2-12.4 GHz frequency range and 1.100" x 0.530" nominal O.D.; and WR 62, with a 12.4-18.0 GHz frequency range and a 0.757" x 0.400" nominal O.D. Both feature corrugated walls that can be bent, flexed and twisted; and they are constructed of either brass or beryllium copper. Waveguides come in lengths up to 1 foot with or without flanges. **Microwave Development Laboratories, Inc., Natick, MA. Ernest Bannister, (617) 655-0060.** Circle 146.

SPURIOUS FREE TRANSISTOR OSCILLATOR FOR 2090 MHz BAND



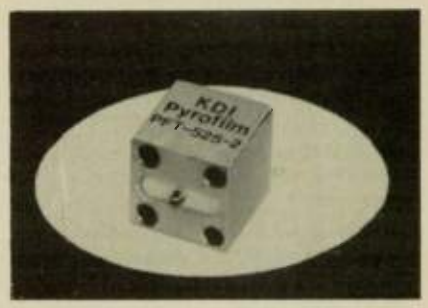
Model HFE-B53 is a fundamental transistor oscillator which provides 17-20 dBm output over a ± 50 MHz tuning range centered at 2090 MHz. Spurious responses of the oscillator are guaranteed to be less than 80 dBc in or out of band. Second harmonic content is 25 dBc min. with an optional low pass filter provided at additional cost. A variety of input voltages can be accepted over the ± 12 to 24 Vdc range. A fine resolution screwdriver shaft provides tuning; AFC tuning is optional. Free-running, long-term temperature stability is ± 0.15% over the 0° to 50°C temperature range. Price: $325 each, plus $35 each for the AFC option and $50 each for the low pass filter option. **Engelmann Microwave Co., Montville, NJ. Carl Schraufnagl, (201) 334-5700.** Circle 106.

(continued on page 80)

GaAs FET LNA SERIES FOR 3.7 - 4.2 GHz BAND

Two lines of GaAs FET LNAs cover the 3.7-4.2 GHz band. Both lines feature 100°K typical noise temperature. Series AW-4290 offers ac (115 or 230 Vac), dc (+15 to +28 Vdc) and ac/dc (- 18 to - 28 Vdc) powered versions. Series AWC-4200 is dc (+15 to +28 Vdc) powered through the RF output connector. Both series are offered with a choice of 49/53 dB or 58/63 dB (min/max) gain. Other characteristics include: ± 0.5 dB full-band gain flatness, .01 dB MHz max. gain slope, .01 ns/MHz max. linear group delay, 0.1/ dB max. AM/PM conversion (at - 5 dBm P_0) and 1.25:1 max. input and output SWR. Series amplifiers also provide + 10 dBm linear output capability (1 dB gain compression). Third-order intercept point is +20 dBm. Price: AW-4295, under $1600; AWC-4205, under $1500. Del: 120 days ARO. **Avantek, Inc., Santa Clara, CA. Jim Lindauer, (408) 249-0700.** Circle 104.

STRIPLINE FLANGE TERMINATION WITH 25 W DISSIPATION



The PFT-525-2, a conduction cooled power termination, dissipates 25 W average power at a heat sink temperature of 100°C. This stripline flange termination features a solderless clip contact. The standard temperature coefficient is less than 150 ppm /°C. SWR from dc to 2 GHz is 1.20 max.; from 2-4 GHz is 1.30 max.; and from 4-6 GHz is 1.50 max. Price: $17.95 each in qty. of 50-99 pieces. **KDI Pyrofilm Corporation, Whippany, NJ. Al Arfin, (201) 887-8100.** Circle 108.

SCHOTTKY DIODE DETECTOR SERIES FOR 0.1 - 18 GHz BAND

A series of miniature, Schottky diode detectors, #2086-6040-00, are multi-octave, matched zero-bias detectors. Series feature MIC construction, provide broadband (0.1-18 GHz) performance and a voltage sensitivity of 600 mV/mW. Response over the specified frequency range is typically ± 1.0 dB. Tangential sensitivity is - 45 dBm minimum. SWR of the detector (zero-bias) is 1.6:1 typical and 1.8:1 maximum. Avail: from stock, in both zero-bias and biasable versions as well as in matched and unmatched models. **Omni Spectra, Inc., Microwave Component Division, Merrimack, NH. (603) 424-4111.** Circle 139.

HIGH TEMP. TEFLON HELIAX® CABLE

FT4-50 is a 1/2", 50 ohm cable suitable for operation up to 200°C. Its Teflon foam dielectric features a velocity propagation of 85% and exhibits low phase change with temperature. An annularly corrugated outer conductor, in conjunction with the connector "O" ring seals, provides a longitudinal moisture block. Differential expansion is eliminated by mechanically locking the outer conductor and bonding the inner conductor to the closed-cell foam dielectric. Connectors feature a self-flaring design. Cable is normally supplied unjacketed; custom jacketed versions are available. **Andrew Corporation, Orland Park, IL. (312) 349-3300.** Circle 103.

SAW BANDPASS FILTERS

A line of surface acoustic wave bandpass filters is offered for communication systems and radar IF systems. Center frequencies range from 70 to 150 MHz. Group delay variations range from 100-500 nS, insertion losses from 19-22 dB and ripple from 0-1.0 dB. Size: 2.35" x 1.15" each, in production quantities, varying according to specific model. **Rockwell International, Filter Products, Newport Beach, CA. (714) 833-4324/4544.** Circle 130.

RF TERMINATIONS HANDLE UP TO 50 WATTS

Models 1426 and 1427 are broadband coaxial terminations for use from dc to 8 GHz. Units have a maximum average power input of 50 W and 25 W, respectively, and 5 kW peak for both models. Max. SWR for both is 1.2 to 4 GHz and 1.3 to 8 GHz. Connectors are semi-precision type N male or female which mate with same type connectors per MIL-C-39012. Price: $175, #1426, $100, #1427 (both prices are for male or female connectors), Avail: stock to 90 days ARO. **Weinschel Engineering, Gaithersburg, MD. (301) 948-3434.** Circle 133.

PRECISION TRIMMER CAPACITOR SERIES

SF/SP is a line of sealed glass dielectric precision trimmer capacitors with temperature coefficients of ± 50 or ± 150 ppm/°C. Seal withstands 40 psi. Typically, a 1-20 pF, SP20, PC style mounts horizontally and is .440" long. The high resolution trimmer has 72 turns per inch and tuning is linear without reversals. Standard maximum capacitance ratings are 4.5, 5.5, 8.5, 10, 11, 12, 14, 16, 17, 20, 22, 25, 28, 30, 3 and 40 pF in both horizontal and vertical PC mounting. Typical prices: $2.60 each for 4.5 pF rating and $6.15 each for 20 pF rating, 1000 qtys. Del: 6-8 wks for large quantities, small quantites from stock. **Voltronics Corporation, East Hanover, NJ. Richard J. Newman, (201) 887-1517.** Circle 129.

MINI QUADRATURE HYBRID COUPLERS

A series of miniature quadrature hybrid couplers, Model H, are 3 dB 90° couplers which span the 250 MHz - 18 GHz frequency range. These devices exhibit a SWR in the range of 1.1 to 1.5 and an isolation in the range of 30 dB to 15 dB, depending upon model. Units have an amplitude balance ranging from ± .25 dB to ± .5 dB. Stripline construction incorporates precision etching and tightly controlled material tolerances. Model's materials insures reproducibility of electrical parameters as well as excellent phase and amplitude tracking. Units are packaged in lightweight solid aluminum cases with convenient mounting holes. Prices: $115, in unit qty. Del: 4 wks. in small qty. **RLC Electronics, Inc., Mt. Kisco, NY. (914) 241-1334.** Circle 137.

(continued on page 82)

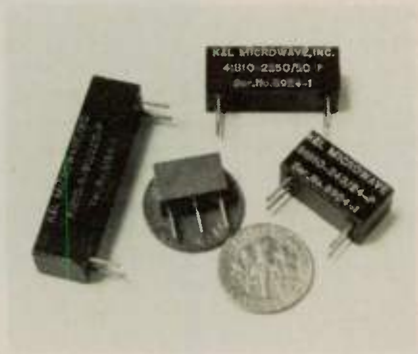
SPECTRUM/COMB GENERATOR SERIES

A series of spectrum/comb generators, SCG 1000, SCG 2000 and SCG 3000, cover frequency ranges from 1 kHz - 5 MHz, 5 MHz - 100 MHz, and 100 MHz to 1.0 GHz in octave bandwidths. Fundamental to the fifth harmonic is typically less than 2.0 dB variation, and up to the tenth harmonic is less than 5 dB variation. Input power is 0 to 10 dBm and output is 5 dBm, typically. Units have an operating temperature from −55°C to + 71°C and input power is + 5V typically. Price: starts at $75. Del: 90 days ARO. **Frequency Engineering Laboratories, Farmingdale, NJ. (201) 938-9000. Circle 107.**

LOW ESR CERAMIC CHIP CAPACITORS

Low equivalent series resistance (ESR) HF series ceramic chip capacitors are offered for requirements above 100 MHz. ESR values are in the 0.05 ohms range. Available in NPO and P100 formulations. Sizes: 0.080" x 0.050" and 0.125" x 0.095". Del: 12 wks. Prices: 50% higher than standard RF chips. **Centre Engineering, State College, PA. Richard N. Stover, (814) 237-0321. Circle 105.**

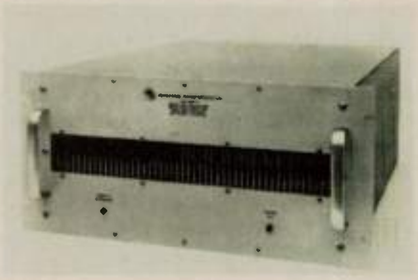
MINIATURE BANDPASS FILTER SERIES



Series IB10 are miniature bandpass filters with center frequencies from 160 MHz - 3.0 GHz. Bandwidths can be specified at 3% to 70% of the center frequency and the devices can be manufactured with 2 to 8 resonant sections. Filters are designed to meet full range of military environmental requirements, including temperatures from −62°C to + 125°C. Input and output terminals are glass feed-through pins, volumes can be SMA, SMB, or SMC connectors can be provided. Price: $250-$475, (1-2 pcs.) dependent upon design. Avail: 6-8 weeks.

K & L Microwave, Inc., Salisbury, MD. (301) 749-2424. Circle 112.

LINEAR POWER AMPLIFIER COVERS 500-1000 MHz



Model LWA 510-210 is a Class A linear power amplifier which operates in the 500-1000 MHz frequency range at output power levels of 200 W saturated and 120 W at 1 dB gain compression. Other characteristics include: 58 dB small signal gain; ± 1 dB small signal gain flatness; 10 dB noise figure; harmonics of −20 dB at 1 dB gain compression; input/output SWR of 2:1. Load SWR may be infinite, all phase angles, to saturation. Operation is from a 24 Vdc, 60 A power source. The amplifier's cooling fans require an additional 120 or 240 Vac power source. Type N, female connectors are supplied. **Microwave Power Devices, Inc., Plainview, NY. (516) 433-1400. Circle 142.**

PLUG-TO-JACK COAXIAL SMA ADAPTOR

Model 50-678-000 is an SMA jack made of gold-plated stainless steel which mates with an SMA plug to provide the transition in a right-angle configuration. Unit is designed for testing of components on subsystems. Size: .625" x .585". Sealectro Corporation, RF Components Division, Mamaroneck, NY. (914) 698-5600. Circle 136.

MINI ROTARY ATTENUATOR SPANS dc - 1 GHz

Miniature rotary attenuator, Model MA-211, offers 3 W power handling, dc-1 GHz frequency coverage, and minimal package density. Unit provides 0-60 dB attenuation in 10 dB steps. Impedance is 50 ohms; insertion loss, 0.2 dB; maximum SWR is 1.2 at 500 MHz and 1.3 at 1 GHz; accuracy is ± 1% or 0.3 dB at 30 MHz, ± 2% or 0.3 dB at 500 MHz and ± 3% at 1000 MHz. Peak power handling capability is 100 W, 3 μs pulse and operating temperature range is 0° to +55°C. SMA connectors are standard. Weight: 6½ oz (185 g). Price: $150, each. Del: 4-6 wks. Texscan Corporation, Indianapolis, IN. Raleigh B. Stelle, III, (317) 357-8781. Circle 135.

DOUBLE-BALANCED MIXER COVERS 6-18 GHz

ER-1011 is a 6-18 GHz double-balanced mixer. With −3 dBm LO power, SWR is 2.0 maximum and conversion loss is 6.5 dB max. LO-signal isolation is 20 dB and LO and signal isolation to IF is 40 dB. The IF band is 100-500 MHz. Size: 2" x 2" x 3/8". Unit uses a SMA connector. Price: (small qty) $475. Avail: from stock. Triangle Microwave, Inc., East Hanover, NJ. Martin Rabinowitz, (201) 884-1423. Circle 134.

SCR PULSE MODULATOR

VXX-3415 is a standardized SCR pulse modulator with all solid-state components capable of driving a magnetron at from 7-10 kW peak output. Modulator has three pulse widths: 0.05 μs at a PRF of 4 kHz, 0.25 μs at a PRF of 2 kHz and 1.00 μs at a PRF of 100 Hz. Operation is from a single-phase, 120 V power supply (either 50-60 or 400 Hz); output is 5 kV at 5 A. A 28 V dc power supply option is available. Size: 7" x 7" x 5". Avail: off the shelf. Varian Associates, Beverly, MA. John Denman, (617) 922-6000. Circle 132.

CARCINOTRON POWER SUPPLY

A carcinotron power supply works with the Thomson-CSF C010, C010.1, C020 and C08 tubes, and it can be modified to handle other BWO's. Ripple of the unit is in the 10 mV peak-to-peak range with regulations of 0.0005% for line and load changes. Crowbars are provided for the anode, collector and line supplies with response time of 5 μs. Automatic on/off sequencing, interlocks, functional indicating lamps, meters and provisions for external modulation of the anode and line supply are also provided. Megavolt Corporation, Hackensack, NJ. Harry Tekel, (201) 487-0100. Circle 131.

FAST ACQUISITION FREQUENCY AGILE SOURCE

PLA-FA-4853 is a fast acquisition frequency agile source. It features low phase noise, and covers the 4.8-5.32 GHz frequency band with +13 dBm power output. Acquisition time is 1 ms maximum, multiplication factor is x48, with locking to an input signal at 0 ± 3 dBm. Alternate frequency bands are offered. Price: $1390 in qty. of 1-4. Avail: 60-90 days ARO. Miteq Inc., Hauppauge, NY. (516) 543-8873. Circle 141.

(continued on page 84)

94 GHz RADAR RECEIVER AND TRANSMITTER

Model 4194B radar receiver and transmitter operate at 94 GHz. The transmitter may be frequency modulated by external source; AM or pulse modulated may be derived from an internal 1 KHz source or external TTL compatible voltage modulator. Receiver has a discriminator for FM modulation and a detector for AM/pulse modulation. Dual-mode receiver includes a low noise balanced mixer, Gunn local oscillator, pulsed or FM/CW reception modes, AFC lock and AGC in the IF amplifier. Transmitter consists of a 20 mW Gunn oscillator. Frequency deviation is ± 100 MHz at rates to about 5 MHz. AM is on/off at rates to 500 kHz (square wave). Price: $68,250. Del: 5 months ARO. **Epsilon Lambda Electronics Corp., Batavia, IL. (312) 879-6006.** Circle 111.

DRM RECEIVER SERIES

Model DRM6-D3 is a message receiver for the 6 GHz FM/FDM band. It features thumbwheel tuning from 5.925-6.425 GHz; synthesizer steps of 10 kHz; downconverter local oscillators phase-locked to internal or external 5 MHz reference (switchable), threshold extension for improved performance at low carrier-to-noise levels, plug-in IF filters and baseband modules for changes of channel capacity and four baseband outputs. **LNR Communications Inc., Hauppauge, NY. Howard Carlin, (516) 273-7111.** Circle 109.

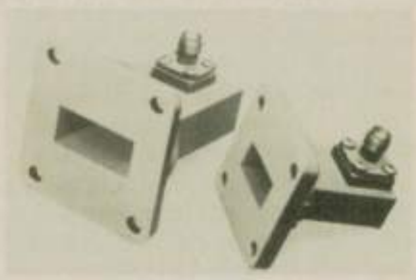
FAST ACQUISITION FREQUENCY AGILE SOURCE

PLA-FA-4853 is a fast acquisition frequency agile source. It features low phase noise, and covers the 4.8-5.32 GHz frequency band with +13 dBm power output. Acquisition time is 1 ms maximum, multiplication factor is x48, with locking to an input signal at 0 ± 3 dBm. Alternate frequency bands are offered. Price: $1390 in qty. of 1-4. Avail: 60-90 days ARO. **Miteq Inc., Hauppauge, NY. (516) 543-8873.** Circle 141.

HYPERABRUPT TUNING DIODES

A hyperabrupt tuning diode offers C4 = 20pf ± 10%, PIV of 15 V, tuning ratio C4/C8 of 1.8 min., 2.4 typical. Diodes provide octave tuning or straight line frequency tuning. Del: 2-4 wks., depending upon quantity. Price: $4.10 each, for 100 pieces, package style D07. **Eastron Corp., Haverhill, MA. (617) 373-3824.** Circle 115.

ADAPTORS FOR WAVEGUIDES AND SMA COAXIAL CONNECTORS



Model 2089-4 is a waveguide-SMA connector adaptor for the 8.2-12.4 GHz range, and Model # 2089-3 covers the 12.4-18 GHz range. Units provide maximum SWR of 1.25 and both have 50 ohm SMA female passivated stainless steel receptacles for the coaxial line. Adaptor # 2089-4 has a UG-39/U flange on RG-52U silver plated waveguide; # 2089-3 has a UG-419/U flange, RG-91/U waveguide, silver plated. Weight: 2089-4 - 120 g.; 2089-3 - 75 g. Dimensions: 2089-4 - 1.85" x 1.6" x 1.6"; 2089-3 - 1.23" x 1.3" x 1.3". **Kings Electronics Company, Inc., Tuckahoe, NY. (914) 793-5000.** Circle 110.

ERRATA

On page 79 of our Microwave Products section run in Feb., 1980, incorrect specifications and photo were included for the FS-1000 frequency synthesizer made by Micro-Tel Corp. The announcement should read:

FREQUENCY SYNTHESIZER CONVERTS YIG OSCILLATORS

Model FS-1000 is a frequency synthesizer which converts YIG oscillators to synthesized, digitally-controllable operation. It is also designed to control most microwave sweep generators. Frequency is controlled to an accuracy of 3×10^{-9} per day and residual FM is reduced to less that 100 Hz. Frequency can be controlled in 100 Hz steps remotely through IEEE-488 bus, parallel BCD, or manually from the front panel. Unit has a .01 - 18 GHz frequency range and resolution can be selected to 10 kHz (with 100 Hz option). Price: standard model - $17,000 Del: 45-60 days. **Micro-Tel Corporation, Baltimore, MD. (301) 823-6227.**

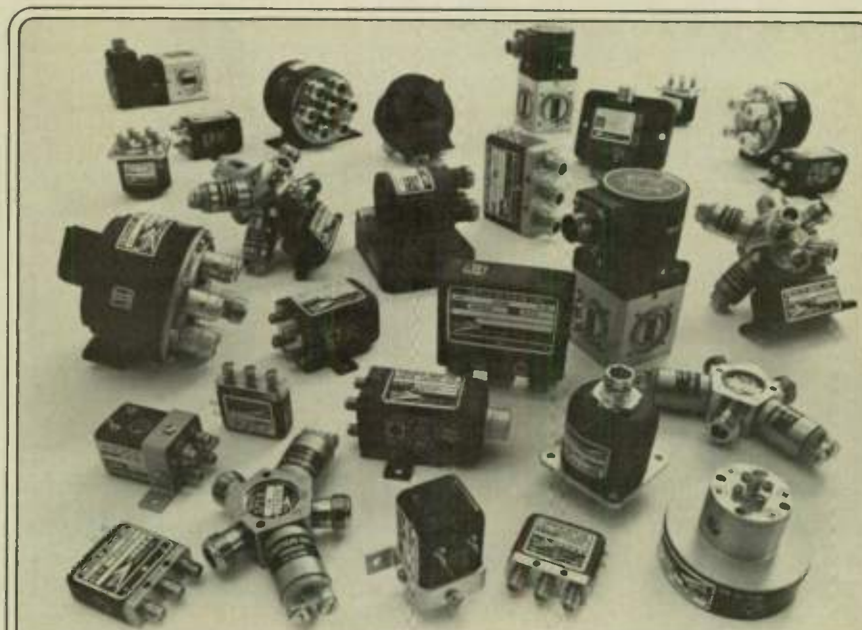
Circle 149.

Figure 4, p. 78 of the February 1980 *Microwave Journal* incorrectly identifies the frequency range of the data shown. The figure caption and callout should identify the frequency range as 0.6-0.8 GHz.

The following appeared in the New Literature section of our Feb. 1980 issue (p. 84). The correct version should read:

SMA CONNECTOR CATALOG

A 28-page catalog on SMA microwave connectors describes a line of standard and high performance and MIL-C-39012 qualified connectors for use with flexible cable, semi-rigid cable and for stripline applications. A line of in-series, and "tee" type adapters is also shown. **B & W Associates, Inc., Burlington, MA.** Robert W. Gray (617) 272-4420.



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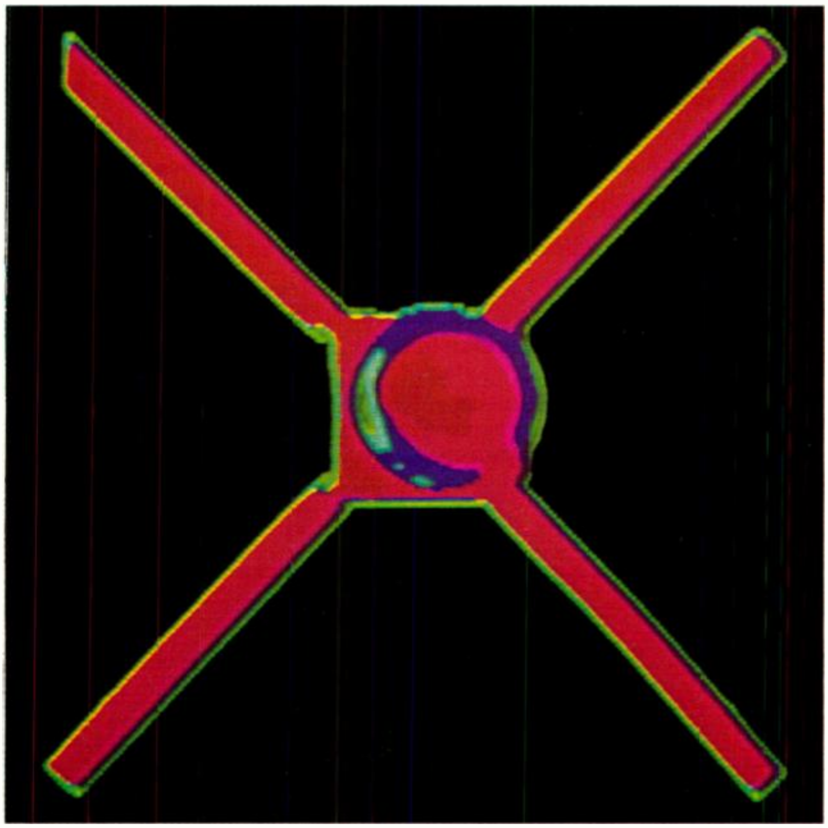


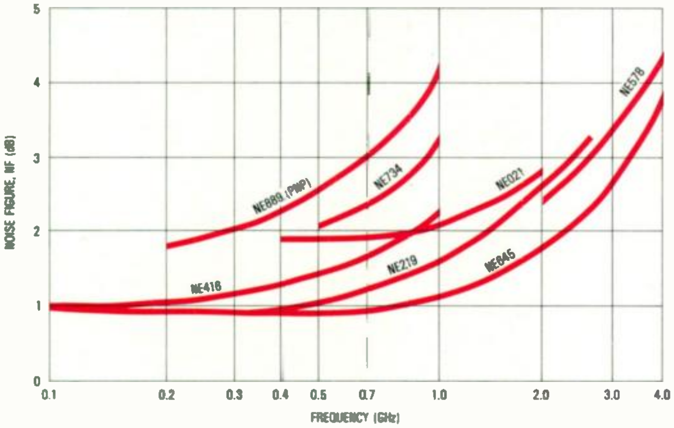
The Best Package Deal Available—In Performance, Reliability and Price

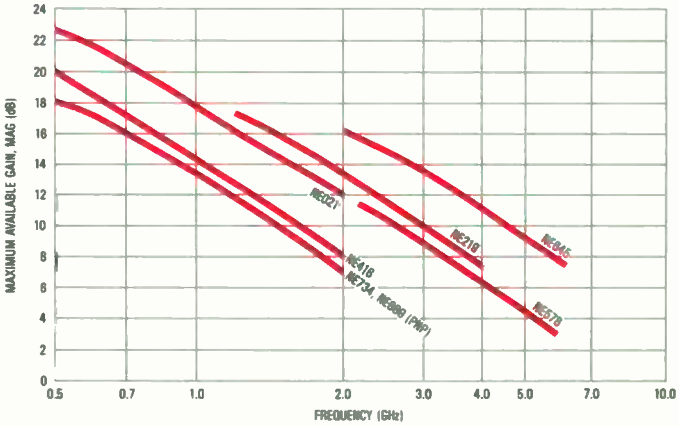
The key to NEC's newly-developed MICRO-X series of hermetically sealed transistors is in manufacturing these high-performance microwave components, with a proprietary ceramic package, using high-speed automated assembly and testing equipment.

The result is a series of state-of-the-art transistors economically priced for large volume OEM's where performance and reliability are primary design considerations.

Ideal for industrial and military applications such as low noise front ends, IF amplifiers and oscillators where plastic transistors cannot be used.

Many devices are priced below $5.00 at 1K. All components available from stock, and Grade C (TXV) screening is available for military applications. Contact your local Cal Eastern representative for a free application note, specification sheet and samples.





NEC microwave semiconductors

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