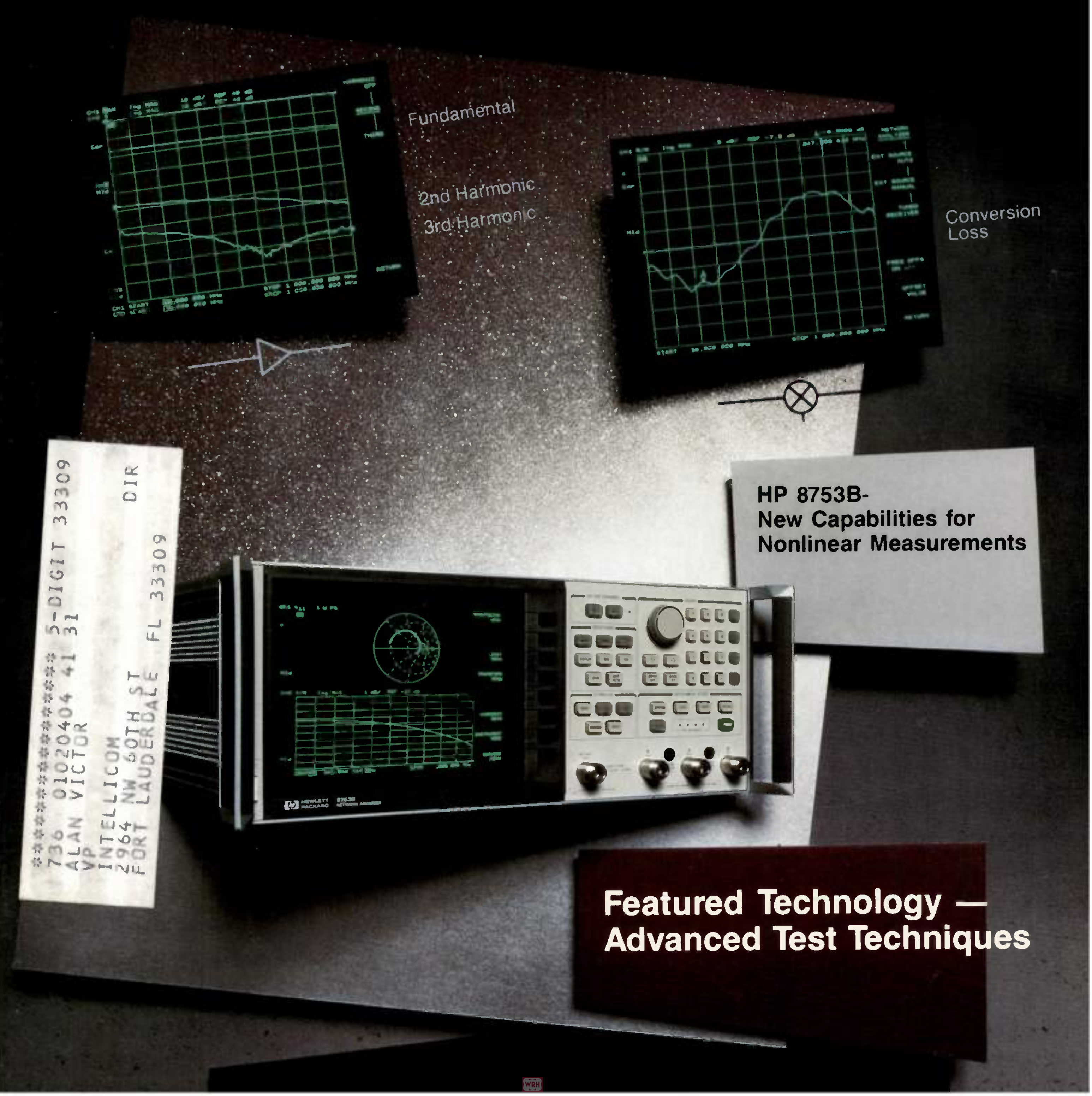
rf design

engineering principles and practices

1978–1988 10th ANNIVERSARY

April 1988



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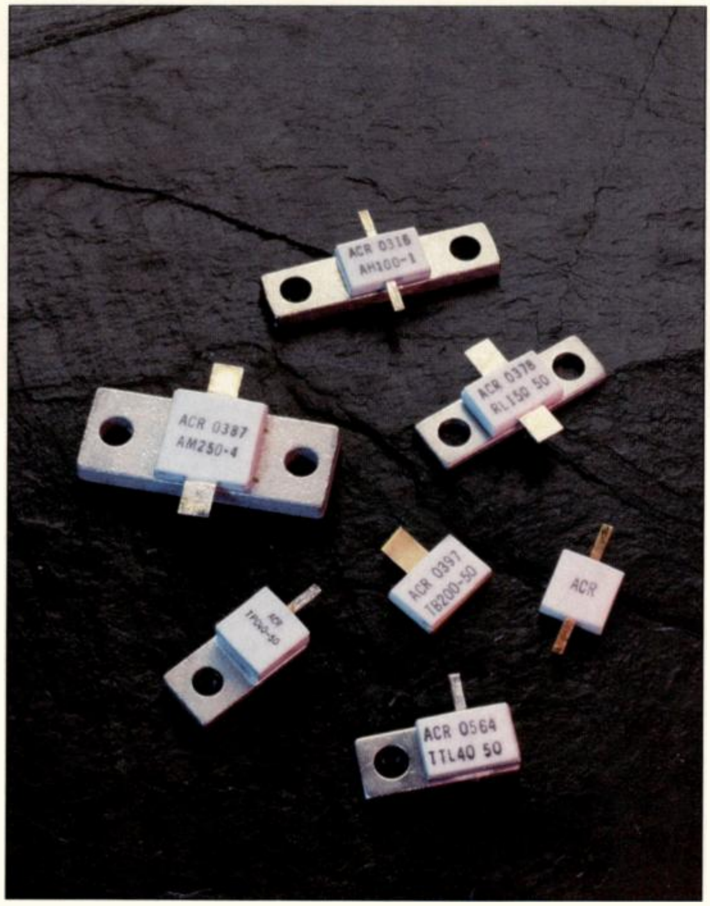

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Western (408) 988-3500 **Eastern** (301) 667-1310
Canada (613) 726-0626
Europe NEC Electronics GmbH 0211/650301

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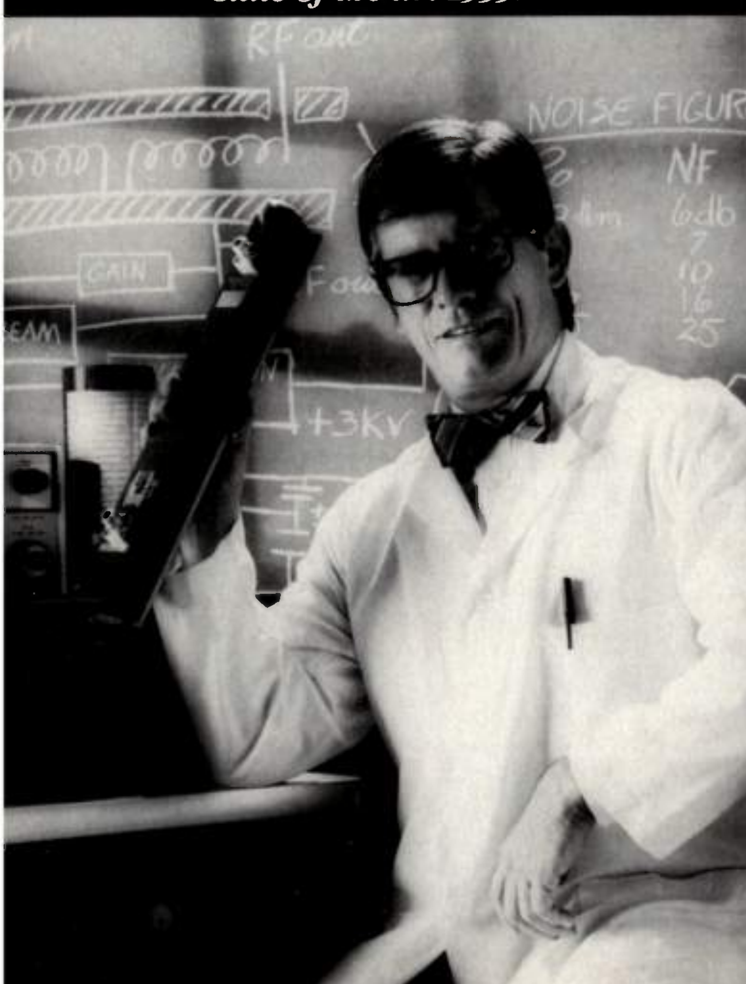
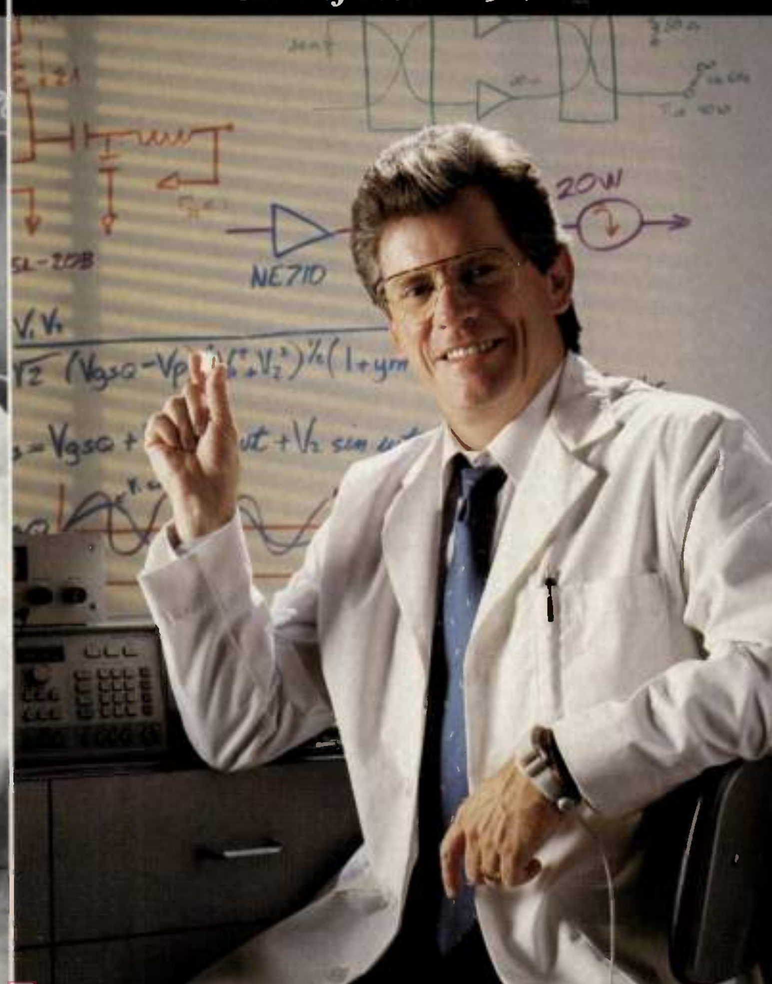


State of the art 1987.



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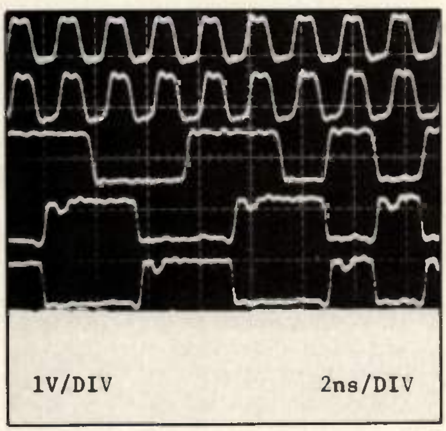
INFO/CARD 3

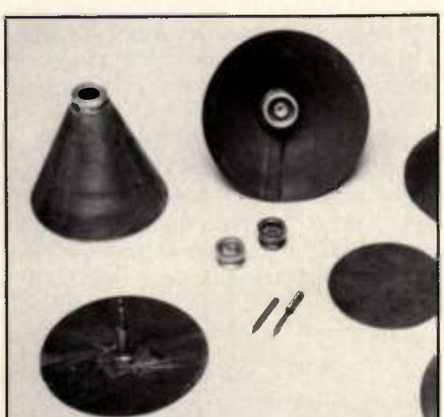
rf design

April 1988


Page 21 — New Network Analyzer



Page 26 — Fast Pulses for GaAs ICs


Page 37 — Discone Test Antennas

Cover Story

Featured Technology Section

Departments

R.F. DESIGN (ISSN: 0163-321X USPS: 453-490) is published monthly plus one extra issue in September. April 1988, Vol. 11, No. 4. Copyright 1988 by Cardiff Publishing Company, a subsidiary of Argus Press Holdings, Inc., 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111 (303) 220-0600. Contents may not be reproduced in any form without written permission. Second-Class Postage paid at Englewood, CO and at additional mailing offices. Subscription office: 1 East First Street, Duluth, MN 55802, (218-723-9355). Domestic subscriptions are sent free to qualified individuals responsible for the design and development of communications equipment. Other subscriptions are: $33 per year in the United States; $41 per year in Canada and Mexico; $45 (surface mail) per year for foreign countries. Additional cost for first class mailing. Payment must be made in U.S. funds and accompany request. If available, single copies and back issues are $5.50 each (in the U.S.). This publication is available on microfilm/fiche from University Microfilms International, 300 N. Zeeb Road, Ann Arbor, MI 48106 USA (313) 761-4700.
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***rf* editorial**

RF Education: Still a Problem

By Gary A. Breed
Editor

"You can't hire an RF engineer. You have to grow your own."

This is a quote from a department head at a major aerospace company. His dilemma is the same one faced by managers everywhere in the RF industry: There are very few experienced RF engineers available in the job market. Promising new graduates have to be found and educated with the slow combination of on-the-job training, seminars and conferences, additional college course work, and self-study. Hopefully, this significant investment results in a decent RF engineer.

Those new graduates still aren't well enough prepared. Two years ago, we reported hopeful signs in the engineering departments of colleges and universities. They seemed to understand the need for teaching analog and electromagnetic principles. Unfortunately, the word isn't getting to the incoming freshmen EE students that RF is a good college track to follow. Mass media news coverage of technology isn't directed to prospective EEs, either. They should be told that superconductors, supercomputers, Star Wars and stealth aircraft all require leading-edge RF techniques.

You can help: A letter to your alma mater might be the right "first step;" a visit would be even better. Let the young EE students know that RF is not a dinosaur left over from the days of Marconi. If their interest is digital, point out that digital engineers often have to find an "RF guy" to solve a problem in their digital circuitry. Remind them that computers with multi-megaflops need engineers who can keep track of stray coupling and crosstalk. Even the mundane world of power supplies (remember DC?) now has megahertz "resonant power converters." That's RF in anybody's book.

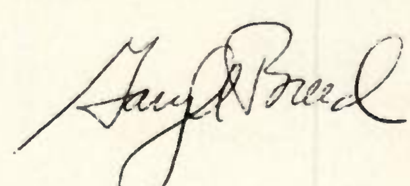
Our Own Role

In response to the need for education, our RF Expos will have more classroom instruction, beginning this fall in Philadelphia. They day prior to the opening of the exhibition (October 24), we will offer two classes. Les Besser's "Fundamentals of RF Circuit Design" is one, familiar to those who have attended previous Expos. For engineers with a little more experience, Randy Rhea will be teaching "Computer-Aided Filter Design," a course based on his top-rated session at last years' RF Expo East.

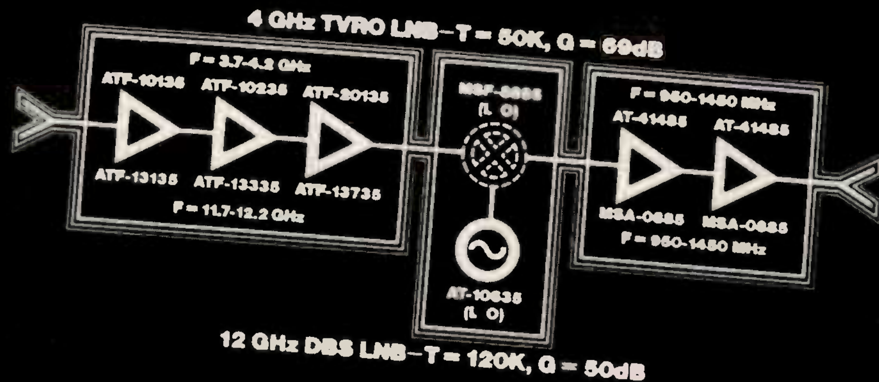
The following day, Les will be teaching a new course, "RF Circuit Design II," intended for engineers who have attended the "Fundamentals" at this or a previous Expo. With these courses, plus Carl Erikson's guidance of the technical sessions, RF Expo East will target the educational needs of the RF engineer. And I wish we could do even more!

In the meantime, we will do what we can. We are always looking for solid, basic tutorial articles; the kind of stuff many of you are teaching junior engineers. We try to mix in plenty of applications information to let everyone know what is happening in areas of RF outside of their own specialty. We want to be a good informational resource.

You and I know that RF is a strong part of engineering. To borrow a phrase from Joe Johnson of MMD, there's *sizzle* in RF, and it's not just bacon cooking in the microwave. Until the rest of the world gets that message loud and clear, we have our work cut out for us.



Just What The Doctor Ordered



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rf letters

Letters should be addressed to: Editor, *RF Design*, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111.

Index of Articles

Editor:

It may be of interest to your readers that I have assembled a five-year cumulative index to *RF Design*, available on 3.5 inch diskettes in MacWrite format for Apple Macintosh computers. The index is organized by subject and title, and includes article corrections and updates published in subsequent magazine issues.

I will provide the diskette for $10.00 to interested readers.

Peter Vizmuller
207 Harding Blvd.
Richmond Hill, Ontario L4C 4X8
Canada

PLL Confusion

Editor:

As an engineer needing to design a phase lock circuit to allow external TV to be displayed in sync with a computer graphics source, I eagerly anticipated Mr. Przedpelski's "PLL Primer" in the November 1987 issue.

However, I got thrown by the first paragraph and Figure 1. Surely the integrators shown in 1a and 1b both have first order transfer functions, the difference being that 1b is a "lag-lead" filter. Therefore, since the VCO acts as an integrator, the PLL behaves as a second order system in both cases, always.

I would appreciate any suggestions from your readers on digital phase comparators that could be used to detect coincidence of two TV horizontal sync pulse trains at 32 kHz, and be sensitive enough to allow 25 ns resolution.

Gordon Coverly
Computing Devices Co.
MS 351
Ottawa, Ontario K2H 6T6
Canada

The Author's Response

Editor:

Reader Coverly is correct, of course. I should not have used Fig. 1a, since first order loops do not have integrators and are not used in frequency synthesizers. Type and order are better described in Part 1. I was mainly trying to show the different integrator configurations and their time responses, which, to a large extent, determine the "slew rate" of the loop. Of particular interest, in some applications, is the different integrator slew rate (third order PLL) for small and large frequency changes, as shown in Fig. 6.

Andrzej Przedpelski
A.R.F. Products

A Correction

The article "Distortion in Nonlinear Circuits" appearing in the February 1988 issue had two errors we would like to correct. First, in Figure 1, the labels of the two axes are reversed. The vertical axis is "output" and the horizontal axis "input." Also, equation (2) has a sign reversed, and should read:

$$y = 10.246X - 0.2308X^3 - 2.1237X^5$$

The signs are correct in the program (Table 1) which uses the equation.

Simple Spectrum Analyzer Fans

Editor:

I liked the article on p. 35 of Jan. 1988, "A Simple Spectrum Analyzer," an inexpensive approach to an expensive piece of equipment. Glad to see some engineers out there who have in mind technical approaches to money matters.

Tom Crawford
Electronic Communications and Service
Elma, WA

Editor:

Keep up the good work! Articles such as "A Simple Spectrum Analyzer" are great.

Mike Ichino
Northrop, Corp.
Pico Rivera, CA

A Request for an Article

Editor:

An article describing printed wiring layour techniques which minimize or eliminate the effects of parasitic circuit elements at high switching frequencies (i.e., printed wiring or component lead inductance, stray capacitance, etc.), as well as the proper utilization of ground planes, would be quite helpful.

Thomas L. MacDonald
Aydin Vector Division
Newtown, PA

Help celebrate the 10th Anniversary of *RF Design* during 1988 — send us your observations, recollections, or philosophical musings about the past ten years of RF technology. We will print as many letters as we can.

rf calendar

April 19-22, 1988
IEEE Instrumentation/Measurement Technology Conference
San Diego Princess Hotel, San Diego
Information: Bob Myers, IMtc, 1700 Westwood Blvd., Los Angeles, CA 90024. Tel: (213) 457-4571

April 20-27, 1988
Hannover Fair Industry 88
Hannover, West Germany
Information: Hannover Fairs USA, Inc., P.O. Box 7066, 103 Carnegie Center, Princeton, NJ 08540. Tel: (609) 987-1202

April 24-28, 1988
Association of Old Crows Technical Symposium of the Mountain and Western Regions
Red Lion Inn, Colorado Springs, CO
Information: Dan Odum, SIERRA Technical Group Inc., 43 Inverness Drive East, Englewood, CO 80112. Tel: (303) 790-1700

May 9-11, 1988
38th Electronic Components Conference
Biltmore Hotel, Los Angeles, CA
Information: EIA, 2001 Eye St. N.W., Washington, DC 20006

May 10-12, 1988
Electro '88
Bayside Exposition Center,
Boston World Trade Center, Boston, MA
Information: Electronic Conventions Management, 8110 Airport Boulevard, Los Angeles, CA. Tel: (213) 772-2965

May 10-12, 1988
EMC Expo 88
Washington Hilton, Washington, DC
Information: Karen Smith, EMC Expo 88, P.O. Box D, Gainesville, VA 22065. Tel: (703) 347-0030

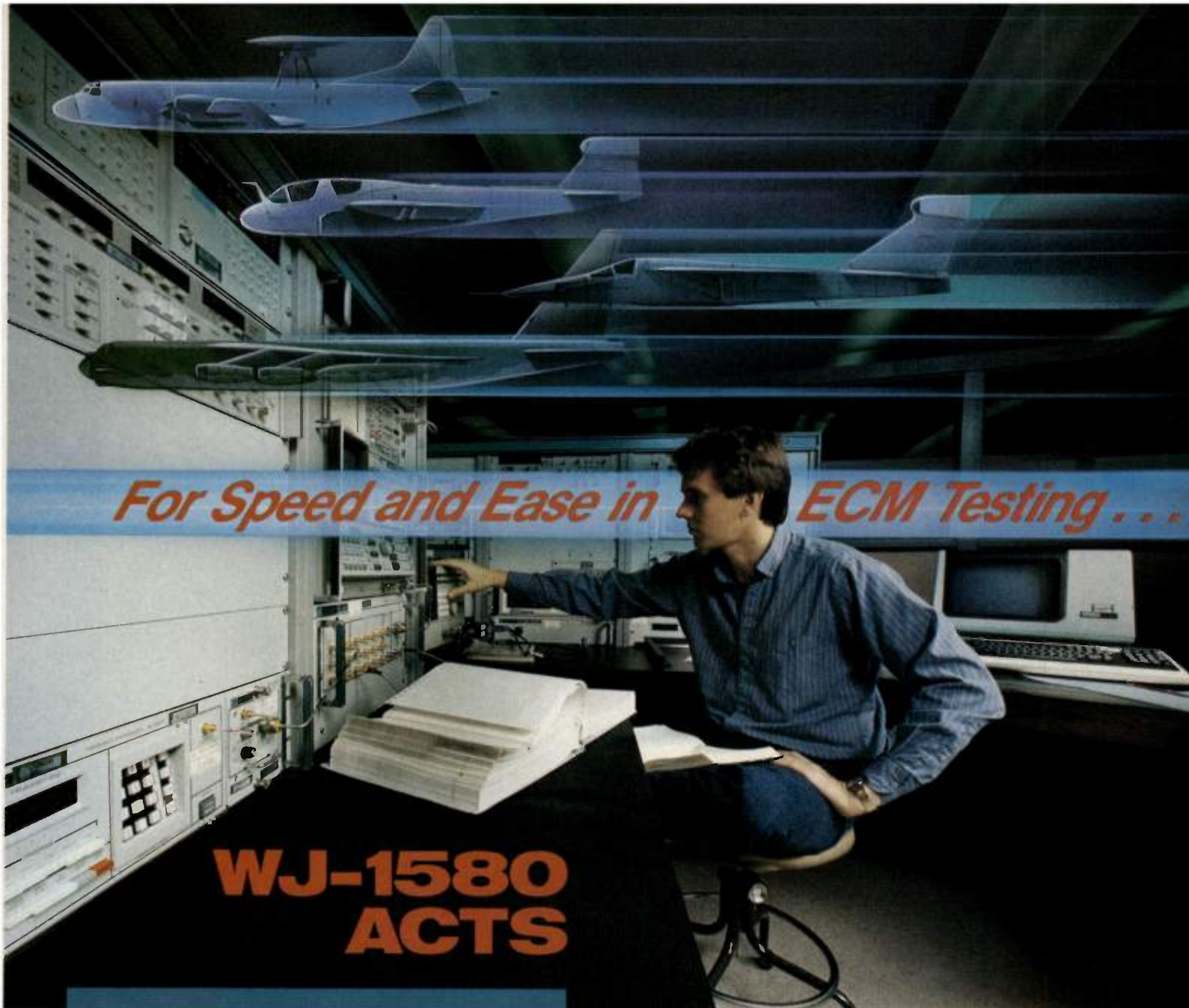
May 24, 1988
ARFTG 31st Conference
Marriott Hotel, New York City, NY
Information: Raymond W. Tucker, Jr., Rome Air Development Center, RADC/RBCM-M, Griffiss AFB, NY 13441. Tel: (315) 330-2841

May 25-27, 1988
1988 IEEE MTT-S International Microwave Symposium
Javits Auditorium, New York City, NY
Information: Charles Buntschuh, Narda Microwave Corp., 435 Moreland Road, Hauppauge, NY 11788. Tel: (516) 231-1700

June 13-15, 1988
Isratech
Jacob Javits Center, New York, NY
Information: Beth Belkin, GreyCom International, 777 Third Avenue, NY 10017. Tel: (212) 546-2200

June 1-3, 1988
42nd Annual Frequency Control Symposium
Stouffer Harborplace Hotel,
Baltimore, MD
Information: Raymond L. Filler, Frequency Control and Timing Branch, Department of the Army, Electronics Technology and Devices Laboratory, Fort Monmouth, NJ 07703-5000

July 5-7, 1988
Military Microwave '88
Wembley Conference Centre,
London, England
Information: P.G. Pinches, Microwave Exhibitions and Publishers Ltd., 90 Calverley Road, Tunbridge Wells, Kent TN1 2UN, England. Tel: (0892) 44027



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- RF Processing Assemblies
- Digital Control Assemblies

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rf courses

The George Washington University

Electronic Warfare Systems: Technical and Operational Aspects
May 2-6, 1988, Washington, DC
August 1-5, 1988, Washington, DC

Modern Spectrum Estimation, Array Processing, and Digital Filtering
May 16-20, 1988, Washington, DC

Video Transmission and Broadcasting via Satellite
June 13-17, 1988, Washington, DC

Spectrum Management
June 20-24, 1988, Washington, DC

Fiber Optics Technology for Communications
June 28-30, 1988, Washington, DC

Global Positioning System (GPS): Principles and Practice
July 6-8, 1988, Washington, DC

Antennas and Arrays: Analysis, Synthesis and Applications
July 18-22, 1988, Washington, DC

Electromagnetic Interference and Control
August 1-5, 1988, Washington, DC

Information: Shirley Forlenzo, Continuing Education Program, George Washington University, Washington, DC 20052; Tel: (800) 424-9773, (202) 994-8530

Besser Associates

Microwave Circuit Design I: Linear Circuits
June 20-24, 1988, Los Angeles, CA
August 15-19, 1988, Baltimore, MD

Microwave Circuit Design II: Non-linear Circuits
August 22-26, 1988, Baltimore, MD

Information: Les Besser, Besser Associates, Inc., 3975 East Bayshore Road, Palo Alto, CA 94303; Tel: (415) 969-3400

UCLA Extension

Introduction to Automatic Testing and ATE Engineering
May 9-12, 1988, Los Angeles, CA

Advanced Topics in Automatic Test Equipment
May 16-20, 1988, Los Angeles, CA

Microwave Circuit Design I
June 20-24, 1988, Los Angeles, CA

Information: UCLA Extension, P.O. Box 24901, Los Angeles, CA 90024; Tel: (213) 825-1901; (213) 825-1047; (213) 825-3344

Test Systems, Inc.

MIL-STD-1553
May 10-11, 1988, Phoenix, AZ

Information: Leroy Earhart, Test Systems, Inc., 217 W. Palmaire, Phoenix, AZ 85021; Tel: (602) 861-1010

Bloom Associates
Modern Power Conversion Design and Analysis Methods
May 16-20, 1988, Boston, MA
June 20-24, 1988, San Diego, CA
July 11-15, 1988, Chicago, IL
September 19-23, 1988, San Francisco, CA

Information: Bloom Associates, Inc., 115 Duran Drive, San Rafael, CA 94903-2317. Tel: (415) 492-8443

Integrated Computer Systems
Digital Signal Processing: Techniques and Applications
May 10-13, 1988, Washington, DC
June 14-17, 1988, Boston, MA
June 28-July 1, 1988, Washington, DC

C Programming: Hands-On Workshop
May 17-20, 1988, Palo Alto, CA
May 24-27, 1988, Washington, DC

Advanced C Programming: Hands-On Workshop
May 10-13, 1988, Toronto, Canada
May 24-27, 1988, Palo Alto, CA
June 7-10, 1988, Washington, DC
July 5-8, 1988, San Diego, CA

Image Processing and Machine Vision
June 14-17, 1988, Toronto, Canada
June 21-24, 1988, Washington, DC
June 28-July 1, 1988, San Diego, CA
July 26-29, 1988, Los Angeles, CA

Information: Barbara Fischer, Integrated Computer Systems, 5800 Hannum Avenue, P.O. Box 3614, Culver City, CA 90321-3614; Tel:(800) 421-8166, (213) 417-8888

Compliance Engineering
Compliance Seminars: EMI, Safety, ESD, Telecom
June 7-10, 1988, Boston, MA

Information: Compliance Engineering, 593 Massachusetts Avenue, Boxborough, MA 01719. Tel: (617) 264-4208

EMC Services, Inc.
EMI Control in Switched Mode Power Supplies
June 27-30, 1988, Boston, MA

Filter Design for Switching Supplies
July 1-2, 1988, Boston, MA

Information: Mark Nave, EMC Services, 11833 93rd Avenue North, Seminole, FL 33542. Tel: (813) 397-5854

Georgia Institute of Technology
Principles and Applications of Millimeter-Wave Radar
June 27-July 1, 1988, Atlanta, GA

Information: Deidre Mercer, Education Extension Services, Georgia Institute of Technology, Atlanta, GA 30332-0385. Tel: (404) 894- 2547.

Southeastern Center for Electrical Engineering Education
Antennas: Principles, Design, and Measurements
August 2-5, 1988, San Diego, CA

Information: Ann Beekman, SCEEE, 1101 Massachusetts Ave., St. Cloud, FL 32796. Tel: (305) 892-6146

R & B Enterprises
Understanding & Applying MIL-STD-461C
June 7-8, 1988, Washington, DC

Electromagnetic Pulse (EMP) Design & Test
June 6-7, 1988, Philadelphia, PA

Identification & Control of Microwave/RF Hazards
May 11-13, 1988, Philadelphia, PA

TEMPEST-A Detailed Design Course
May 16-20, 1988, Philadelphia, PA

Grounding, Bonding & Shielding
June 16-17, 1988, Philadelphia, PA

Worst Case Circuit Analysis
June 20-22, 1988, Philadelphia, PA

EMI Suppression Methods
June 28-30, 1988, Philadelphia, PA

The R & B EMI Training Institute
August 8-19, 1988, Philadelphia, PA

Information: Greg Gore, R & B Enterprises, 20 Clipper Road, West Conshohocken, PA 19428. Tel: (215) 825-1684

Design & Evaluation, Inc.
The Worst Case Circuit Analysis Training Seminar
May 9-11, 1988, Washington, DC
July 11-13, 1988, Honolulu, HI
September 12-14, 1988, Boston, MA
October 17-19, 1988, San Francisco, CA

Information: Design & Evaluation, Inc., 1000 White Horse Road — Suite 304, Voorhees, NJ 08043. Tel: (609) 770-0800

University Consortium for Continuing Education
Modern Microwave Techniques
September 26-29, 1988, Washington, DC

Information: University Consortium for Continuing Education, 16161 Ventura Boulevard, M/S C-752, Encino, CA 91436. Tel: (818) 995-6335

Directory Lists Resources of Federal Laboratories

Hundreds of Federal laboratories, engineering and information centers now make their expertise available to assist U.S. businesses and researchers. The participating laboratories are listed in a directory of Federal laboratory and technology resources. Many of these labs offer capabilities and expertise not available from other sources.

The publications describes the individual capabilities of each organization, along with the name and telephone number of a contact person at that facility. A new feature in this edition is a list of all U.S. Government laboratories and their technology transfer offices.

Subject areas listed which are of interest to RF researchers include: Atmospheric and Astronomical Sciences, Computer Technology, Electrotechnology, Energy, Engineering, Manufacturing, Materials, Nuclear Technology and Physical Sciences. This directory is published by the Center for the Utilization of Federal Technology (CUFT) to encourage and increase technology transfer to U.S. private industry.

To obtain the Directory, order PB88-100011/KCS, *Directory of Federal Laboratory and Technology Resources — A Guide to Services, Facilities and Expertise,* 1988-1989, (28 pages, softcover, ISBN 934213) $36, plus $3 handling fee, from National Technical Information Service (NTIS), Springfield, VA 22161, (703) 487-4650.

Three New Companies to Pursue GaAs ICs, Capacitors and Connectors

The new startups have announced their entry into the RF marketplace. The first is Gazelle Microcircuits Inc., Santa Clara, Calif., which will be producing large-scale-integration GaAs IC devices for large volume applications. To support their CAD/CAE computer models, the firm has modified existing silicon-design software packages. Formed in 1986, Gazelle has just received a first-round venture capital investment of $5.5 million, and expects to announce products later this year.

Another new company is Sierra Aerospace Technology Inc. of Carson City, Nev. Sierra has just begun operations in the manufacture of multilayer ceramic capacitors and EMI filters. The main direction of this company is in high reliability, producing devices for military and demanding commercial specifications. A ceramic manufacturing process with a protective environment to protect the material from contamination has been developed to assure quality.

National Tel-Tronics, Inc. has announced the formation of a new company, NTT/Waveconn, a new venture to design, manufacture and sell microwave connectors. Waveconn's offices and factory will be located in Meadville, Pa., initially producing over 125 styles of SMA, SMB and SMC coaxial connectors. Although military approved products are not among the first offerings, the products will all be designed to MIL C39012.

Amplifonix in Leveraged Buyout

The management team at Amplifonix, Inc. has made an amicable leveraged buyout of the company from SAMCO Investors, Inc. The new ownership, including president and chairman Dr. Arthur Riben, vice president of sales Richard DuBois, and vice president of operations Richard Leodore, feels the company is in a stronger financial position through arrangements with local banks, and is in a position to increase sales and earnings in 1988. Established in 1980, Amplifonix now employs 62 people in the manufacture of RF amplifiers used in military and communications systems.

Pirelli Establishes Superconductivity Chair at M.I.T.

Pirelli Cable Corporation and the worldwide Pirelli Group has established a Career Development Professorship at the Massachusetts Institute of Technology dedicated to teaching and research in superconductivity. The chair will be awarded every three years to a young M.I.T. faculty member whose work is focused on this specialized area.

The first holder of the chair is Dr. David A. Rudman, as assistant professor in the Department of Materials Science and Engineering. His research centers on the investigation of fundamental properties of superconducting materials, as well as the fabrication and characterization of those materials for use in microelectronics.

'RF Circuit Design Fundamentals' on Videotape

Besser Associates announces the release of a video tape course entitled "RF Circuit Design Fundamentals," a revised version of the one-day tutorial offered at the RF Expo symposia. The course includes six sessions: RF Concepts and Lumped Component Models, Filters and Resonant Circuits, Transmission Lines and Scattering Parameters, Impedance Matching Techniques, Smith Chart and its Applications, and Small Signal Amplifier Design. Each session is concluded with a review example to test comprehension. The tapes for each session can be purchased individually for $199 each, or as an entire package for $995, including the text and course notes. The text book is *RF Circuit Design* by Chris Bowick. For additional information on this videotape course, circle INFO/CARD #176.

Gould Sells ASW Unit to Westinghouse

Gould, Inc. has announced the sale of its antisubmarine warfare (ASW) operation in Cleveland, Ohio, to Westinghouse Electric Corp. for approximately $100 million. Proceeds from the sale will be used to reduce debt. The Ocean Systems Division, which employs about 1,200 persons, produces underseas weapons systems for the U.S. Navy.

New Contracts for W-J

Watkins-Johnson Company announces that it has been awarded contracts with a potential value over $18 million. One package is from Sanders Associates to produce solid-state amplifiers and digitally-tuned oscillators for the ALQ-126B defensive countermeasures system used in Navy and Marine aircraft. Another contract has been awarded by Hughes Aircraft Company's Missile Systems Group to deliver RF processors for the AIM-120 (AMRAAM).

Hughes to Buy Westar Satellite System

Hughes Aircraft Company has reached an agreement to buy Western Union's Westar Satellite system, including three satellites now in orbit, one yet to be launched, plus telemetry, tracking and command equipment. Final sale is contingent on a definitive agreement and approval by the Boards of Directors, plus approval by the Federal Communications Commission.

Hughes Communications Inc., a subsidiary which currently operates three Galaxy C-band satellites, will also operate the Westar satellites. Hughes also intends to implement the planned launch of Westar VI-S as a replacement for Westar III, which will soon be phased out of ser-

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A Decisive Advantage In Winning The Wide-Band Dynamic Range Game. 135 dB Spurious-Free Dynamic Range.

At frequencies from HF to microwave, Paramixer™ Technology is a design architecture which realizes the highest dynamic range frequency converters available today.

Jamming Immunity (out-of-band dynamic range) of radar and communication receivers can be significantly enhanced even in the harshest environments with a Paramixer front end. The Paramixer frequency converter is capable of withstanding out-of-band interfering signals of up to one watt without suffering in-band desensitization. The unequalled second-order intercept performance of the Paramixer frequency converter can permit frequency hopping tuning speed to be maximized by eliminating the need for preselection filtering. Spur levels due to out-of-band jammers are reduced to levels presently attained by preselector designs.

Optimal in-band dynamic range can be attained by incorporating Paramixer technology throughout the RF chain. Triple-conversion architectures can be realized with more than 90-db in-band spurious-free dynamic range in a 1 MHz bandwidth.

Wide Frequency Coverage Applications—HF To Microwave. Paramixer technology can be used to realize low spurs and high dynamic range in designs for wide-band—

- Frequency Converters
- Translators
- RF-to-Digital Converters
- Exciters
- Synthesizers

Reduce synthesizer spurs by incorporating Paramixer technology into critical frequency translator stages. Paramixer frequency converters have been designed having M × N products (M>3, N>2) at more than − 100 dBc with 0 dBm input.

Specifications For A Typical Paramixer Frequency Converter.

Input intercepts:	
Third-order	>+ 45 dBm
Second-order	>+ 82 dBm
Noise figure	<6.5 dB
LO power required	+ 17 to + 20 dBm
DC power required	<7 Watts

High-level Paramixer frequency converters have been designed to handle continuous signals at power levels *exceeding ten watts.* Gain a winning advantage in the wide-band dynamic range game—call us today for details.

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It is a Steinbrecher goal to provide our customers with a decisive advantage, gained through applied technology, superior engineering and high quality manufacturing.

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vice. Both systems currently serve radio, television, cable, data, print media and private telecommunications customers.

Avantek to Utilize New GaAs Facility

Avantek Inc. plans to move a substantial portion of its GaAs semiconductor production to its Newark, Calif., facility during 1988. Occupancy of the building will provide increased capacity for the company's GaAs product line, including varactor, PIN and limiter diodes, GaAs FETs, and MMICs. The manufacturing facility includes 34,000 sq. ft. of wafer processing area, with 17,000 sq. ft. of Class-10 cleanrooms. All Avantek silicon transistors and MMICs have been produced at this plant since March, 1987. By moving the GaAs line to Newark, Avantek will also facilitate the pilot production of circuits under the DOD's MIMIC program, which Avantek participates in as part of the Westinghouse (prime contractor) team.

Electrospace Receives New Contracts

Electrospace Systems Inc. announces contracts totalling over $9 million. GTE Government Systems has contracted the firm to manufacture and test active training aids and signal entry panel mock-ups for Mobile Subscriber Equipments. Another contract from the Navy is for a Communications Digital Switching System (CDSS) for the Range Coordination Network (RCN) at the Atlantic Fleet Weapons Training Facility in Puerto Rico and at outlying ranges. The RCN includes RF, microwave and landline communications equipment. Also announced is a $3.95 million contract from TRW Defense Systems for a Red Digital Circuit Switch to be installed at Fort Belvoir, Va.

Jammer Contract to AEL

American Electronics Laboratories has been awarded a $5.5 million contract to supply the Navy with 25 Low Band Jammer (LBJ) amplifiers for the EA-6B Prowler aircraft. This award continues AEL's involvement in the LBJ, which constitutes the two lowest bands of the AN/ALQ-99, a tactical jamming system for the disruption of radar. The system is also used on the Air Force's EF-111A aircraft.

PAMTECH Makes a Move

Passive Microwave Technology Inc. (PAMTECH) has announced its relocation to 1151 Avenida Acaso, Camarillo, CA 93010, from the former Canoga Park address. PAMTECH manufactures components from DC through mm-wave frequencies, with a standard line of ferrite components, and a specialty in custom and High-Rel waveguide and coaxial components.

Richardson's Canadian Facility Moves to Larger Plant

Richardson Electronics Canada Ltd. has moved to a larger facility, located at 4 Baker Road — Unit 2, Brampton, Ontario L6T 4E3. The new facility will increase the capability of the office in its distribution of power semiconductors, tubes and other components.

Flam & Russell to Upgrade USAF Radars

A $1.7 million contract has been awarded to Flam & Russell Inc. by the Eastern Space and Missile Center at Patrick AFB to upgrade the range antenna systems. The improvements will enhance performance of the high-powered C-Band track-

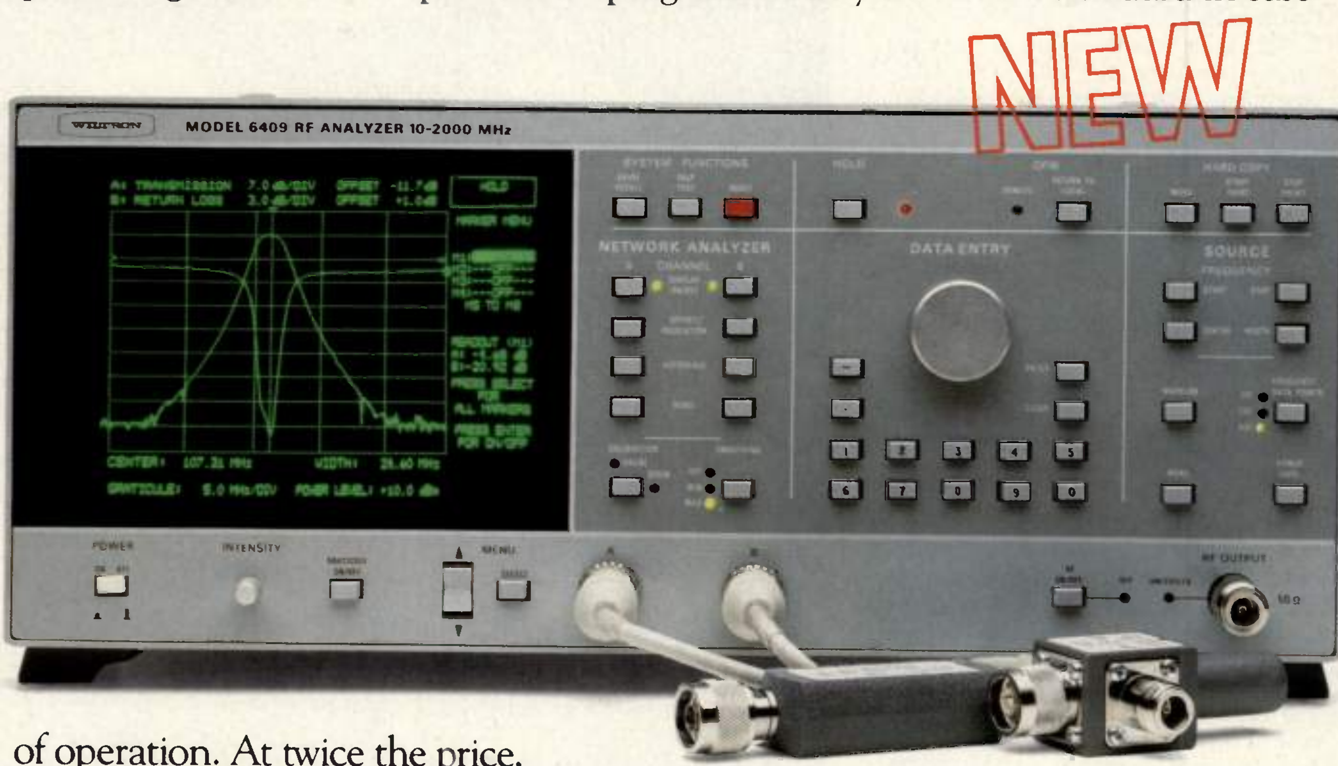
You don't g

With the NEW Wiltron RF Test System.

The new 1 to 2000 MHz Wiltron 6400 RF Analyzer. In a single, compact instrument, Wiltron brings you an affordable crystal-derived swept signal source, a scalar network analyzer, precision measurement components, and a display. Measure transmission, return loss, and absolute power.

It's a remarkable breakthrough. For far less than you would expect to pay, Wiltron gives you synthesizer-like performance over the entire 1 to 2000 MHz.

The 6400 has resolution and accuracy 10 times better than anything in its price range. It offers complete GPIB programmability and a new standard in ease



of operation. At twice the price, the 6400 would still be a bargain.

Our engineers gained exceptional stability and accuracy by "locking" the frequency to a crystal marker at the beginning of every sweep. Because frequency is always on the mark, you won't waste time looking for traces that have drifted off the screen. You'll have excellent repeatability of test data taken on very narrow bandwidth devices. Even if tests are made days apart.

et the drift.

Dynamic range is 71 dB (+16 to −55 dBm). Use the Wiltron 6400 for the most demanding applications including TV tuner, cellular radios, filters, amplifiers, diplexers and DBS. For a permanent record, test data are plotted in graphical or tabular format on an optional ink-jet printer.

The 6400 is a joy to use on the production line, in the laboratory, or out in the field.

Autoscaling can be used to automatically select the optimum display for your device. The display is fully annotated to ensure accurate, confusion-free interpretation of test data.

Specifications
Signal Source
Frequency Range:
  Model 6407: 1 to 1000 MHz
  Model 6409: 10 to 2000 MHz
Frequency Accuracy: ± 100 kHz
Frequency Resolution: 10 kHz
Leveled Output Power Range: +12 dBm to +0.1 dBm
  Optional attenuator: +10 dBm to −70 dBm
Harmonics: < −30 dBc
Nonharmonic spurious: < −40 dBc
Network Analyzer
Dynamic Range: +16 dBm to −55 dBm
Vertical Display Resolution: 0.003 dB maximum
Horizontal Display Resolution: 101, 201, or 401 points.
Normalization: 800 points, automatically interpolated for ranges less than full range.
Markers: up to 8.
SWR Autotesters
Directivity: 40 dB
Impedance: 50 or 75 ohms
Test Port Connector: Type N or BNC
RF Detectors
Impedance: 50 or 75 ohms
Test Port Connector: Type N or BNC

Test Data on Friday, 10:15 AM:
M1 Marker Frequency: 104.72 MHz
Passband Amplitude at M1 Marker: −4.82 dB

Test Data on Wednesday, 3:30 PM:
M1 Marker Frequency: 104.72 MHz
Passband Amplitude at M1 Marker: −4.86 dB

You get fast production test times since frequencies can be changed without recalibration. Go/no-go limit lines and up to eight markers, which can be set to test points of interest, make readings easier. Costly setup time is eliminated by storing up to nine front panel setups in memory.

In addition to the 6400's 17.8 cm (7 in.) display, there's a composite video output which will drive a larger screen. Production people love it. And with a weight of only 16 kg (35 lb), the 6400 can be carried to the most inaccessible repeater station.

Now in a single instrument, you get everything you need to make fast, accurate RF measurements. And you don't get the drift.

For more information, contact Wiltron, 490 Jarvis Drive, Morgan Hill, CA 95037-2809. Tel: (408) 778-2000.

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ing radars located at sites on Merritt Island and Ascension Island.

K & L Microwave to Sell Through Distributors

K & L Microwave has added stocking distributors to its existing sales and representative force, for its coaxial switches. Customers may now obtain standard products in this product line in five regions of the U.S. plus Japan and West Germany. In addition to the coaxial switch line noted here, K & L makes microwave filters, crystal filters, integrated subassemblies and RF switching matrices.

Penstock Moves to Larger Facility

Penstock Inc., distributor of RF/IF and microwave components has moved to a new facility at 520 Mercury Drive, Sunnyvale, CA 94086-4018. The additional space will allow the firm to expand its stocked inventory and add complementary product lines.

Report Shows Japanese Commitment to GaAs Semiconductors

Tokyo-based EGIS K.K., a market research and consulting firm, has released a report showing how Japanese industry is working to develop ICs and discrete semiconductors using GaAs and other advanced compounds. Much of the growth has been in optical electronics, but a substantial portion is in devices for digital, RF and microwave applications.

A key factor in the development of these products is that Japanese semiconductors are produced in large vertically-integrated, mass production oriented electronic companies. This assures an in-house customer, justifying large sums for capital equipment and R & D activities. An additional advantage for the Japanese is reduced reliance on U.S. manufactured production equipment. The report states that 80 percent of semiconductor fabrication equipment purchased in Japan is of Japanese manufacture.

The complete report, *GaAs/Compound Semiconductors — Japanese Development*, is available for a price of $4,800 from International Planning Information, Inc., 465 Convention Way #1, Redwood City, CA 94063, (415) 364-9040.

Agreement Promotes Blind Mate Connectors

M/A-COM Omni Spectra and Radiall announce an agreement to promote standardization of the OSP and OSSP Blind Mate RF Coaxial Connector technology. Under the agreement, Radiall has obtained a license for use of the OSP patent and will manufacture compatible connectors, which are designated BMA series by the IEC and DESC.

Gould Semiconductor Division to be Centralized in Idaho

Gould, Inc. will transfer all marketing and remaining administrative functions to its semiconductor facility in Pocatello, Idaho, from former headquarters in Santa Clara, Calif. This action is intended to strengthen the division's position in ASICs, an intensely customer-driven portion of technology. By centralizing all engineering, manufacturing and marketing functions, closer interaction is possible.

rf cover story

Simplify Your Amplifier and Mixer Testing

New Network Analyzer Has Nonlinear Measurement Capability.

By Jim Curran
Hewlett-Packard Co.

The HP 8753B offers new features for nonlinear measurements and direct control of external instruments.

When measuring a device's linear and nonlinear characteristics, multiple test configurations are typically required. Network analyzers have traditionally been used to measure the linear reflection and transmission characteristics of RF components, such as filters and cables, by applying a known swept signal and then measuring the magnitude and phase of the transmitted and reflected signals. Active devices such as amplifiers and mixers, however, require both linear and nonlinear characterization. Each device must be moved between one test set-up to measure impedance, amplifier gain and mixer conversion loss and another set-up to measure harmonics and intermodulation distortion. The latest RF network analyzer from Hewlett-Packard Co., the HP 8753B, dramatically simplifies and speeds these tests. The ability to measure amplifier harmonics and mixer conversion loss is added to the contributions of its predecessor, the HP 8753A. In addition, the HP 8753B offers other new features designed to aid manufacturing test applications and provides frequency coverage from 300 kHz to 6 GHz.

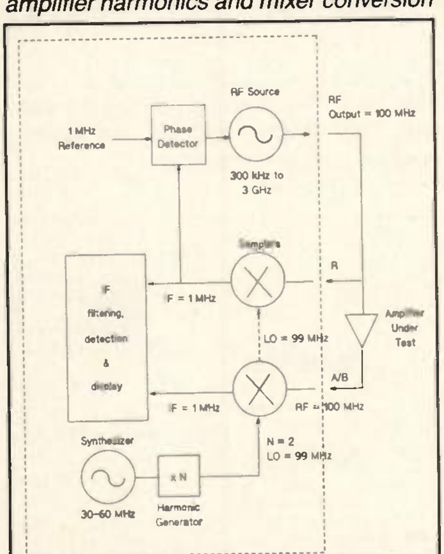

Figure 1a. Functional diagram for linear measurements.

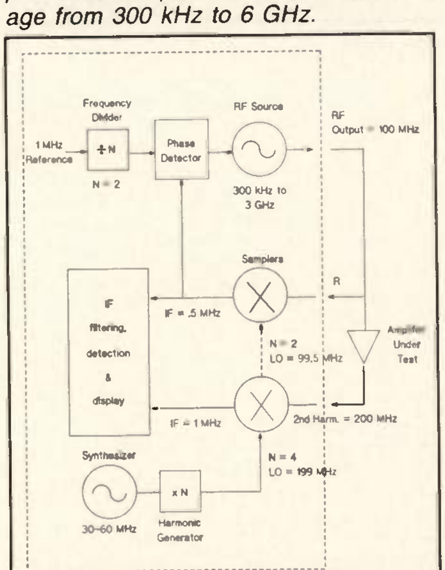

Figure 1b. Functional diagram for 2nd harmonic measurement.

A device's behavior is linear when a sine wave input produces a sine wave output at the same frequency with only an amplitude and phase change. Examples of linear devices are filters and cables. The output of a nonlinear device is dependent on the power level of the input signal and is usually composed of multiple signal components at harmonically-related frequencies. Examples of such devices are saturated amplifiers and mixers.

Figure 1a contains a block diagram which illustrates how the HP 8753B is normally configured to make linear measurements. The RF source stimulates the amplifier under test (100 MHz in this example). The incident and test signals are received and mixed down to a low frequency IF (1 MHz). The 1 MHz IF signal is filtered to remove unwanted signals and then measured. The 1 MHz IF signal in the Reference (R) channel also goes to one input of a phase detector. The other input is fed by a stable 1 MHz frequency reference. The phase detector produces an output voltage which is proportional to the phase difference between the two input signals. This voltage is used to fine

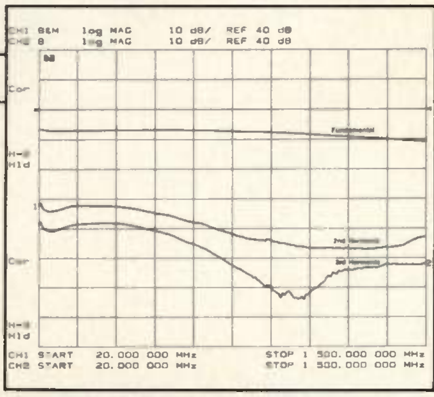

Figure 2a. Swept frequency amplifier measurement of absolute fundamental, 2nd and 3rd harmonic output level (dBc).

tune the source to a precisely specified frequency. This particular block diagram is capable of measuring the signals only at the RF source frequency.

To measure nonlinear performance, such as an amplifier's second harmonic output level, a different configuration (shown in Figure 1b) is required. The incident stimulus frequency is used as the reference signal (again, 100 MHz) while the test channels is at two times the fundamental's frequency to measure the harmonic. By multiplying the LO frequency (99.5 MHz) in the output test channel by 2 (2 × LO = 199 MHz) the second harmonic (200 MHz) is mixed down to the receiver's desired 1 MHz IF. In the reference channel, the fundamental (100 MHz) is mixed down (LO = 99.5 MHz) to 0.5 MHz. When measuring the second harmonic, the phase detector reference signal is also 0.5 MHz. Measurement of either the fundamental, second or third harmonic signals, individually or as ratios (dBc), can be displayed.

Similar modification to the network

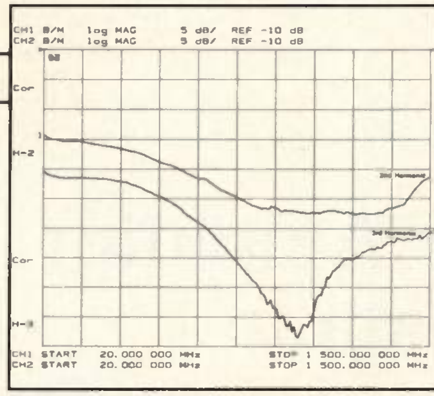

Figure 2b. Swept frequency amplifier measurement of 2nd and 3rd harmonic distortion (dBc).

analyzer's block diagram is necessary to measure a frequency translating device such as a mixer. An external frequency source is required to generate the LO signal. The HP 8753B's built-in source can be used as the mixer's RF stimulus with all of the receiver input channels tuned to the mixer's output IF (RF minus LO product only). In this case the analyzer's source is offset from the receiver by the specified LO frequency.

Amplifier testing

The majority of amplifier measurements are linear but many applications also require nonlinear information. Today, the typical configuration uses an external RF source to stimulate the amplifier at a CW frequency with a spectrum analyzer connected at the device output. The output spectrum of the amplifier is then displayed. This technique provides high dynamic range and the ability to measure total harmonic distortion (including all harmonics, spurious and intermodulation products).

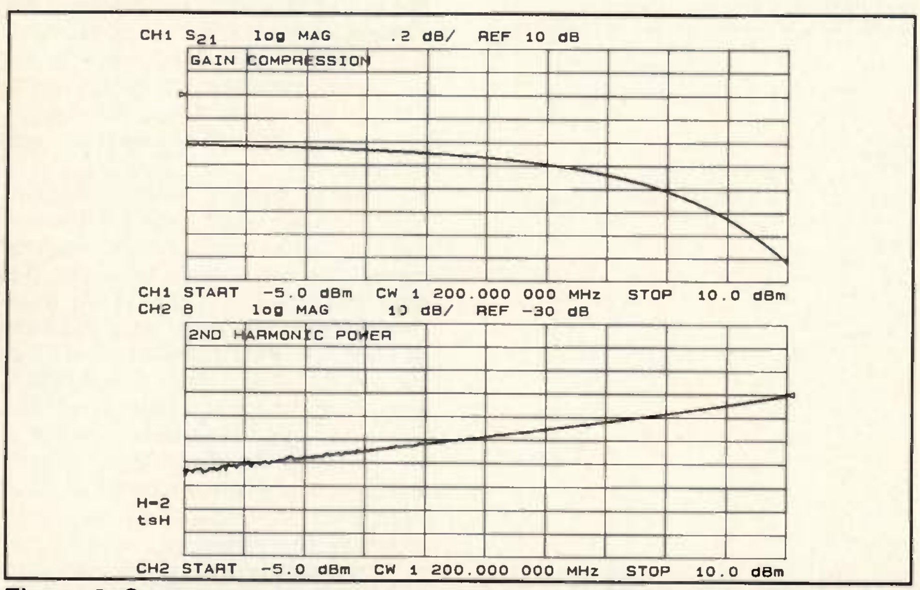

Figure 3. Swept-power measurement of an amplifier's fundamental gain compression and 2nd harmonic output level.

The HP 8753B can now make a swept-frequency measurement of an amplifier's second or third harmonic as shown in Figure 2a. The second/third harmonic response can be displayed directly in dBc, or dB below the fundamental or carrier (see Figure 2b). The ability to display harmonic level vs. frequency or RF power allows "real-time" tuning of harmonic distortion. Further, this swept harmonic measurement, as well as well as all of the traditional linear amplifier measurements can be made without reconnecting the device to a different test configuration.

Vector network analyzers are commonly used to characterize amplifier gain compression vs. frequency and power level. This is essentially linear characterization since only relative level of the fundamental input to the fundamental output is measured. The narrowband receiver is tuned to a precise frequency and, as a result, is immune from harmonic distortion. Sometimes it is desired to quantify the harmonic distortion itself. Figure 3 illustrates a simultaneous measurement of fundamental gain compression and second harmonic power as a function of input power.

In a compression measurement it is necessary to know the RF input or out-

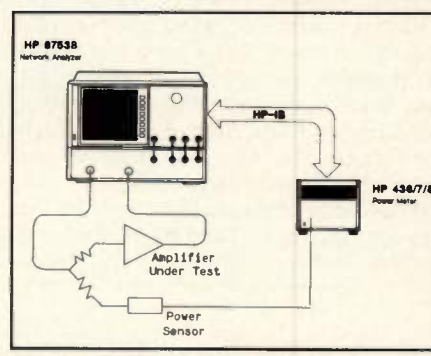

Figure 4. Test configuration for setting RF input using automatic power meter calibration.

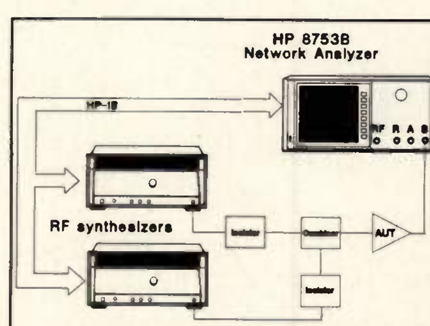

Figure 5. Test configuration for two-tone third-order intermodulation measurement.

put power at a certain level of gain compression. Therefore, both gain and absolute power level need to be accurately characterized. Uncertainty in a gain compression measurement is typically less than 0.05 dB. Also, each input channel of the HP 8753B is calibrated to display absolute power (typically within ±0.5 dBm up to 3 GHz, and ±1 dB up to 6 GHz).

However, when measuring a device that is very sensitive to absolute power level, it is important to be able to accurately set the power level at either the device input or output. Power meters are typically used to accurately monitor absolute power level. This presents an additional level of complexity, because the RF source must be re-adjusted to provide the correct signal level at the device's input or output. In automated production test areas, an external computer is often used to control both the network analyzer and the power meter. The HP 8753B is now capable of using an external HP-IB power meter and controlling source power directly. Figure 4 shows a typical test configuration for setting a precise leveled input power at the test device input.

Intermodulation distortion can also be characterized using the HP 8753B and two external RF sources. Figure 5 shows a typical configuration. The receiver is adjusted to measure the signals of interest, the two fundamental tones as well as the third order products. Because the receiver is not phase-locked to any of the incoming signals, the two external sources must be synthesized. Hewlett-Packard low cost RF signal generators, the HP 8653B or HP 8657A are recommended because of their frequency stability in this application. An example measurement third order intermodulation with a two tone stimulus is shown in Figure 6.

Mixer Testing

Mixers or frequency converters, by definition, exhibit the characteristic of having different input and output frequencies. For a single-sideband mixer measurement, the RF source can be offset in frequency from the input receiver frequency, allowing for a swept RF stimulus over one frequency range and measurement of the IF response over another (in this case the output IF). Figure 7 shows a suggested test configuration for the HP 8753B in the Frequency Offset mode. External signal separation schemes rather than two-port S-parameter test sets are generally required for testing multiple-port devices.

Attenuation at all mixer ports is used to reduce errors associated with port mismatches. IF filtering at the mixer's IF port is recommended to prevent unwanted mixing products from entering the receiver. The swept RF-to-IF conversion loss of a mixer with a fixed LO of 1500 MHz, is shown in Figure 8. In this same test configuration, conversion loss compression (loss vs. power level) can be measured on a single frequency basis by sweeping input power level. Stepped RF and LO frequency, fixed IF mixer testing can be achieved by stepping two external sources while measuring the output IF.

Amplitude and phase matching of mixers can be performed using the test configuration shown in Figure 9. The R channel mixer is used as part of the phase and amplitude reference. A second mixer is added to either the A or B input channel of the HP 8753B receiver. Using power splitters to divide the RF and LO signals between the reference and test mixers, comparative measurements are made by displaying the ratio of the test and reference channels.

Built-In Automation

As instrumentation becomes more complex and test requirements more exhaustive, it is even more important to maintain a balance between a simple operator interface and minimum test time. Complex test systems require more operator training and the possibility of human error increases as the number of steps in the test process increase. When production volumes are high enough, many test systems are completely automated to simplify and speed device testing. However, there are many low and medium volume test applications where automation would reduce test time but do not justify the initial cost of purchasing a computer and developing custom software.

The new Test Sequence function lets you automate a test without a computer and does not require any additional programming expertise. Operation is similar to programming a hand-held calculator. The HP 8753B simply "learns" the keystrokes normally used to make a measurement, which can later be executed with the push of a key. Each sequence can hold approximately 200 instructions. All of the analyzer's test features can be automated, augmented by some basic decision-making capability (i.e., IF LIMIT TEST FAILS, DO TUNE SEQUENCE). A simple example is shown in Figure 10. This sequence recalls a set-up, measures second harmonic in dBc, checks a test limit, and plots data if test limits are passed.

Another application of test sequencing is illustrated in one of the previous examples, The mixer test set-up in Figure 8 includes an external RF source to provide the LO signal. The test sequence function allows the HP 8753B to set the frequency and power of the RF source through the HP-IB interface. A similar application would be measurement of an RF amplifier at different bias levels automatically by controlling a programmable

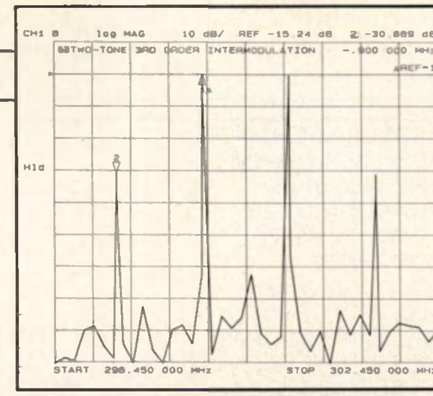

Figure 6. Measurement of third order intermodulation (TOI) products of an amplifier with a two-tone RF stimulus.

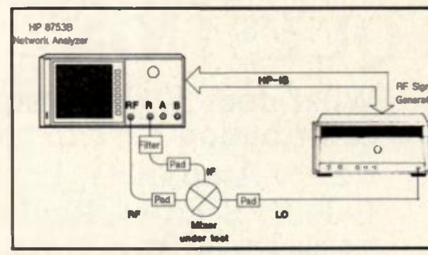

Figure 7. Mixer conversion loss test set-up.

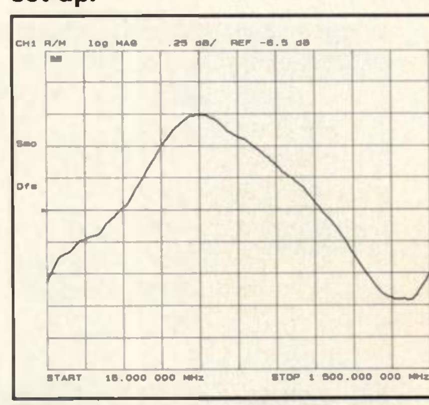

Figure 8. Swept RF-to-IF conversion loss measurement.

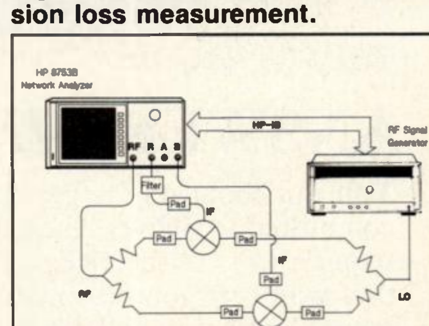

Figure 9. Test configuration for swept-frequency amplitude and phase tracking of mixers.

```
SEQUENCE AMPTEST
Start of Sequence
RECALL 3
HARMONIC OFF
DATA MEMORY
DATA/MEM
SECOND
LIMIT TEST ON
IF LIMIT TEST PASS THEN DO PLOT
IF LIMIT TEST FAIL THEN DO
  SEQUENCE 3
```

Figure 10. Example test sequence.

DC bias source.

The HP 8753B has a variety of new measurement capability and productivity enhancement features. Among these features are a built-in plotter/printer buffer and interpolative calibration (which allows the operating frequencies to be changed without recalibrating). However, just as important is the ability to configure a system that meets specific test requirements. A family of Transmission/Reflection and S-parameter test sets for 50 ohm (3 GHz) and 75 ohm (2 GHz) systems are available.

Alternatively, a complete 6 GHz system can be configured including the new HP 85047A 6 GHz S-parameter test set. This test set includes a frequency doubler that can be switched in to measure 3 MHz to 6 GHz in a single sweep or switched out to measure 300 kHz to 3 GHz. This test set exhibits less than 6 dB of insertion loss between the RF source and the test port for as high as 14 dBm output power at the test port in the 3 GHz band.

Optional time domain analysis, harmonic measurement capability and 6 GHz receiver operation can be added to any existing HP 8753A or 8753B network analyzer. An upgrade kit provides the hardware, software and documentation necessary to retrofit any existing HP 8753A.

Performance summary:

Frequency range (standard	300 kHz to 3 GHz
with Option 006 and HP 85047A	300 kHz to 6 GHz
Frequency resolution	1 Hz
Measurement range	
300 kHz to 3 GHz	0 dBm to −100 dBm
3 GHz to 6 GHz	0 dBm to −95 dBm
Dynamic accuracy	
(over a 40 dB range)	±.05 dB, ±0.3 deg
Harmonic measurement	
(Option 002)	
Frequency range	16 MHz to 3 or 6 GHz
Dynamic range	40 dBc

The HP 8753B family also includes other new products including a new HP 85024A 300 kHz to 3 GHz high impedance probe. Low input capacitance and high shunt resistance minimizes the loading to the circuit under test. The HP 8347A RF amplifier is capable of providing leveled power of +5 to +20 dBm with 25 dB gain over the broad 100 kHz to 3 GHz frequency range. The amplifier's ALC can be used to extend the dynamic range of the HP 8753B by 20 dB.

The HP 8753B is $25,500 U.S. list price and available May 1 with eight weeks delivery ARO. Harmonic measurement capability (add $3,000), 6 GHz receiver operation (add $3,000) and time domain analysis (add $4,800) are optional. The HP 85044A Transmission/Reflection test set is $3,500 and HP the 85046A S-parameter test set is $7,800. Both test sets cover the 300 kHz to 3 GHz frequency range and are currently available. The new HP 85047A 6 GHz S-parameter test set is $9,800 and is available May 1 with 16 weeks delivery ARO. The HP 85024A probe is $1,900 and is currently available. The HP 8347A RF amplifier is $3,750 and is available starting May 1.

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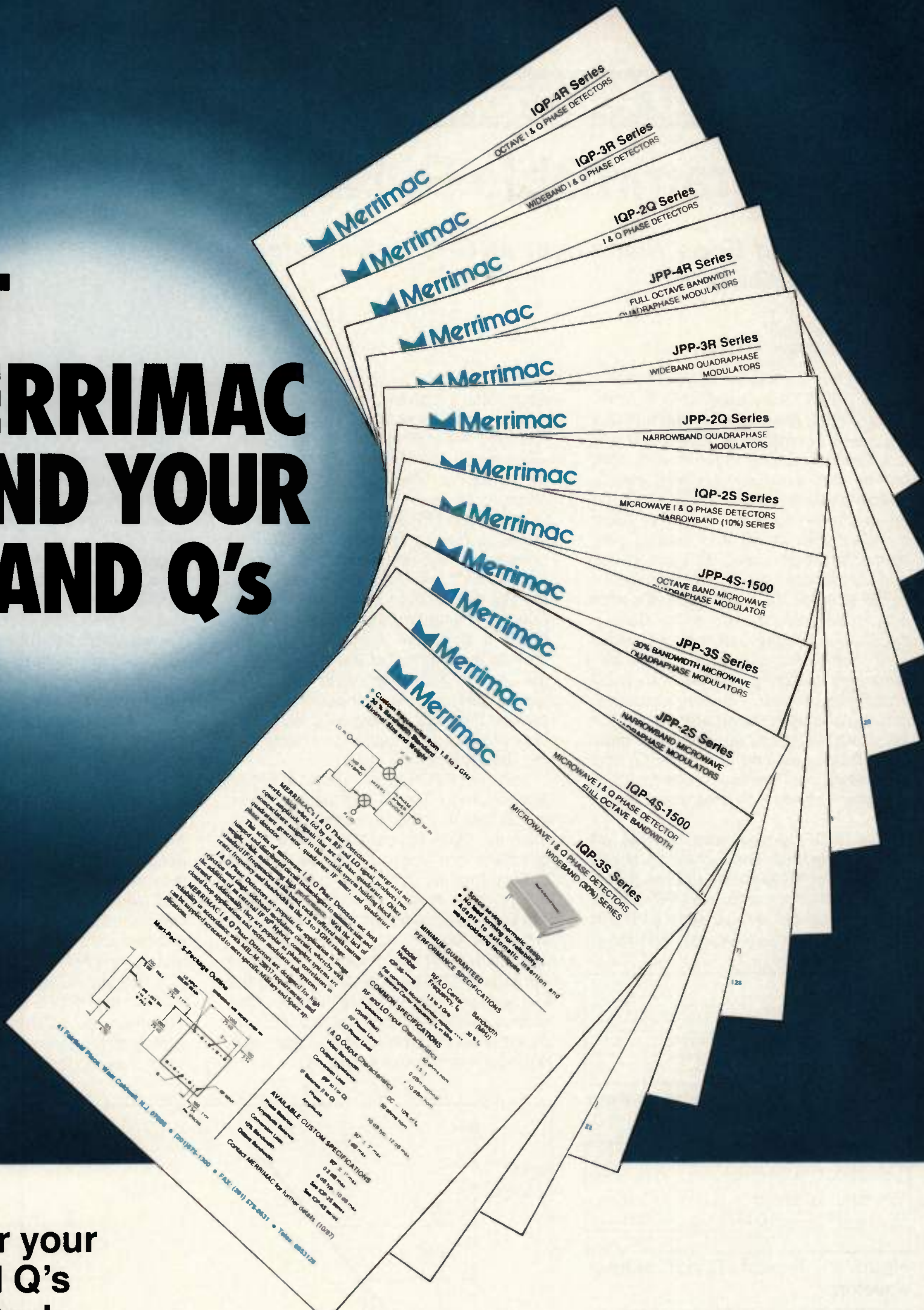

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INFO/CARD 5

A Programmable Pulse Driver Formatter

Picosecond Edge Placement at GHz Pulse Rates with New GaAs IC

By Karl C. Zabel
Harris Microwave Semiconductor

The Programmable Pulse Driver Formatter (PPDF) is a digital GaAs IC developed to meet the measurement and testing needs of high speed GaAs and ECL ICs. Until now those requirements could not be met because available circuits operated with less accuracy and functioned at slower speeds. The PPDF uses standard ECL inputs and creates complementary, high-speed, ECL compatible outputs that are user-programmable. The PPDF's GaAs/ECL compatible I/Os allow it to directly interface with either GaAs or Si ECL ICs. The medium scale integration level PPDF formats data pulses and generates complex, gigahertz rate pulse trains with variable, independent control of rising and falling edge placement. This stand-alone GaAs IC accomplishes many of the tasks of existing, complex ECL word generators, but it performs their functions at higher speeds with greater accuracy.

The PPDF formats data in three different modes. The first mode, shown in Figure 1, is Non-return to Zero (NRZ). During each clock cycle the NRZ data remains in the same logic state as it's input. If the input is a logical *High* (one), then the NRZ data also stays in a logical *High* state. When the input changes to a logical *Low* (zero), the NRZ data also changes to a logical *Low* during the next clock cycle. The NRZ data then remains in the logical *Low* state until the input changes state.

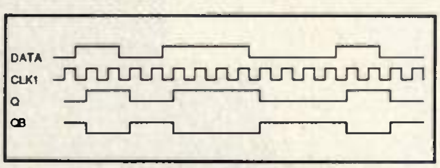


Figure 1. Typical NRZ data output patterns.

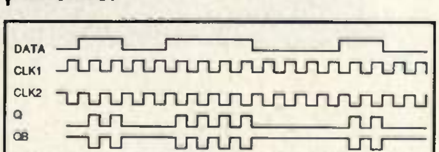


Figure 21. Typical RZ data output waveform.

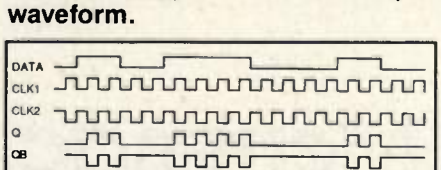


Figure 3. Typical R1 data output waveform.

The second PPDF mode, shown in Figure 2, is Return to Zero (RZ). RZ data does not always remain the same logical state as it's input. During each clock cycle the RZ data always returns to a logical *Low* after outputing the input data. If the input is a logical *High*, the RZ data spends the first half of the clock cycle in the logical *High* state and returns to logical *Low* for the remainder of the clock cycle. If the input is a logical *Low*, the RZ data remains in a logical *Low* state during the entire clock cycle.

The third PPDF mode, shown in Figure 3, is Return to One (R1). R1 data is similar to RZ data in that it does not always remain the same logical state as it's input, but there is a major difference between the two modes. During each clock cycle the R1 data always returns to a logical *High* after outputing the input data. If the input is a logical *Low*, then the R1 data spends the first half of the clock cycle in the logical *Low* state and returns to the logical *High* state for the remainder of the clock cycle. If the input is a logical *High*, then the R1 data remains a logical *High* during the entire clock cycle. Due to the PPDF's high degree of accuracy, the user-programmable output pulse train pattern can be controlled in picosecond increments. This allows the user to independently modify the placement of the leading and trailing edges of the pulses in the output waveform. The PPDF's differential outputs also operate up to 1 GHz with typically less than 50 ps skew between them. In addition to the pulse placement and skew accuracies, the PPDF also has 150 ps typical output rise and fall times. These sharp output edges are important because they increase the accuracy of the output waveforms. The purpose of the PPDF is to create accurate waveforms that can be used as a known reference input to the DUT so that timing information about the DUT can be determined. The DUT has a threshold voltage (Vth), which is known only to within a finite range of values (ΔVth). If the input waveform to the DUT has a finite slope, as shown in Figure 4, there is additional uncertainty in when the DUT started to respond to the input stimulus. As the rise and fall times of the input waveforms is decreased by the PPDF, the uncertainty is decreased and the accuracy of the measurements is increased. Typical Si ECL rise and fall times are as large as 600 ps while the PPDF's short, 150 ps rise and fall times are able to increase the accuracy of the reference waveform.

All these characteristics can also be

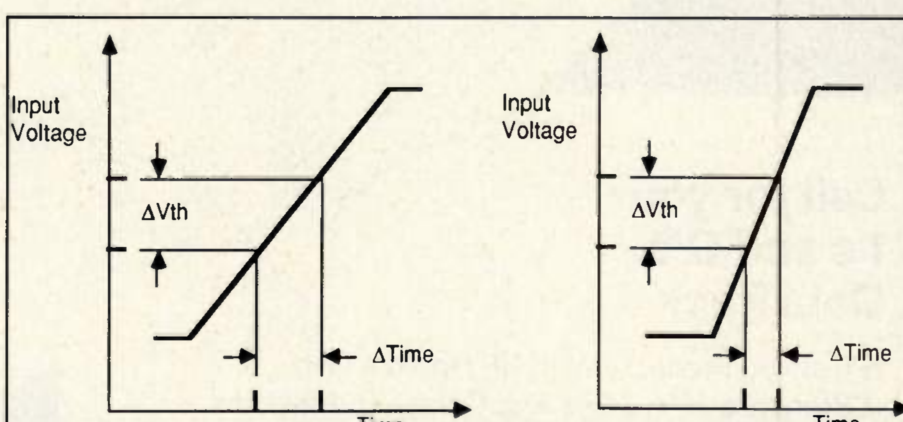


Figure 4. The slope of the input of the input reference waveform affects the accuracy of the timing measurements.

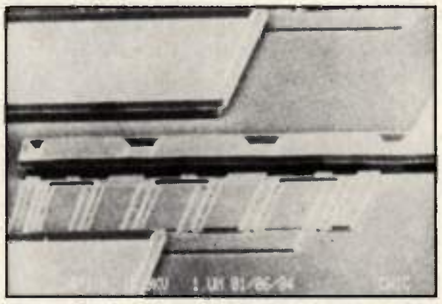


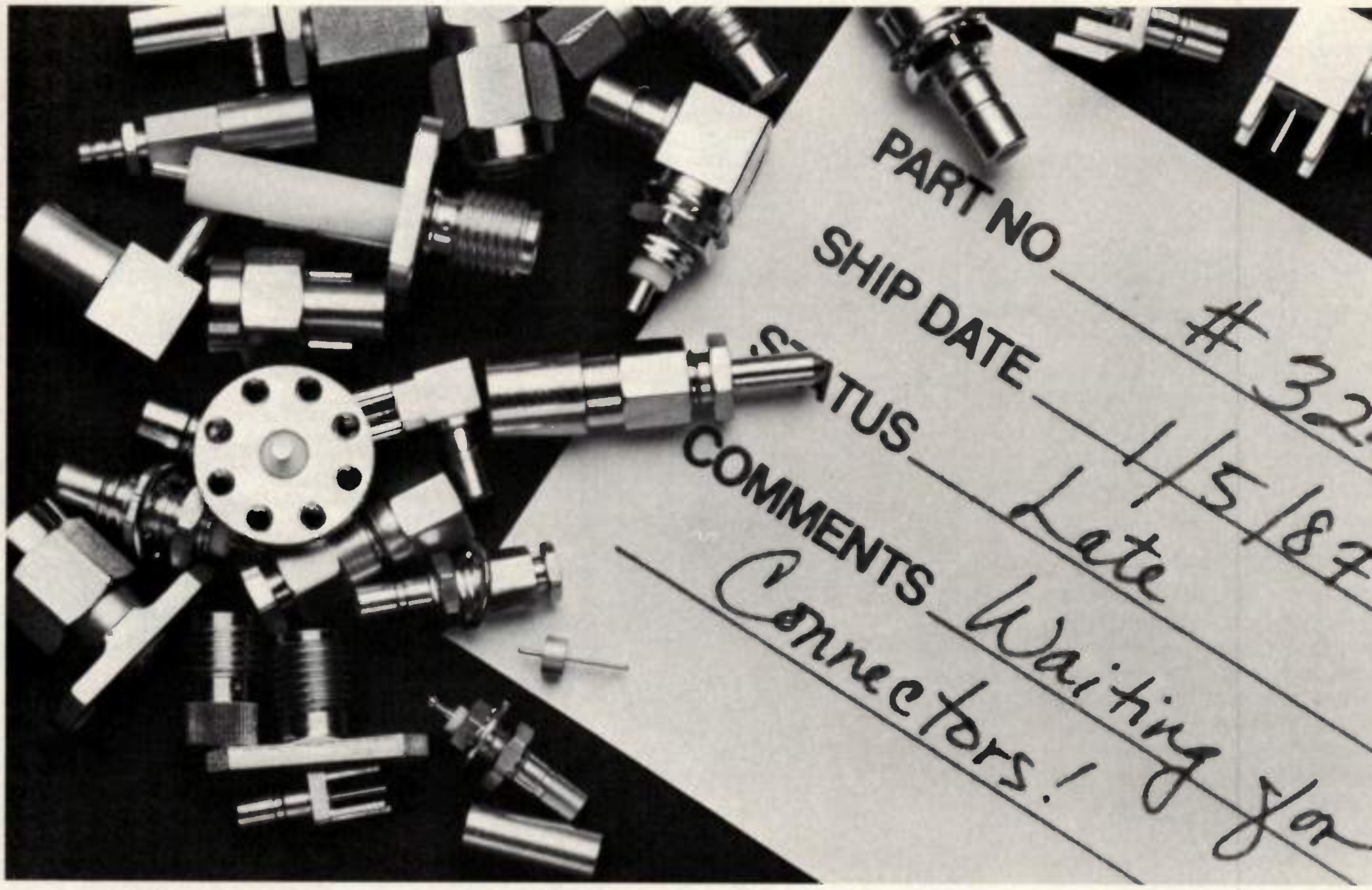
Figure 5. This photo shows the dual metal airbridge interconnect system.

achieved using slow edge rate, sinusoidal inputs, so the high speed rate output waveforms are not dependent on providing high speed input signals. In fact, clock signals of a few MHz work fine if only low duty cycle output signals are required. These properties make the PPDF unique and ideal for pulse/word generation, system timing formatter, single shot trigger generator, and memory testing applications.

Process Information

The PPDF is fabricated using Harris' standard DIGI-1 GaAs process. The DIGI-1 process is a direct, multiple ion-implant, two level interconnect process utilizing an undoped, liquid encapsulated Czochralski (LEC) grown semi-insulating GaAs substrate. The process is highly planar (flat), leading to high density and high yield. The PPDF's MESFET and diode structures have minimum geometry 1.0 um gate lengths, and the MESFETs are — 2.0 V pinch-off, depletion mode structures. The PPDF is capable of operating at gigahertz frequencies because it's able to reduce interconnection parasitics by using dual level metal interconnects. The first level of metal interconnect sits directly on the GaAs substrate while the second metal is an airbridge layer which is suspended above the GaAs substrate. Both of the interconnect levels are processed using layers of Ti/Pt/Au to reduce parasitic resistance, but the airbridge level is thicker than the first metal level which adds to it's strength and also reduces it's resistance. First metal resistivity is 0.04 ohms/sq and airbridge resistivity is 0.02 ohms/sq. The parasitic capacitance of the airbridge level is also about one half the capacitance of the first level metal because the airbridge level takes advantage of the air between it and the ground plane to provide a low ϵ_r dielectric layer.

The PPDF was designed with buffered FET logic (BFL). BFL was chosen because it has a robust logic swing, that



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INFO/CARD 21

TRUTH TABLE - NRZ MODE

MODE1	MODE2	DATA	CLK1	CLK2	Q	$\bar{Q}$
L	X	L	↑	X	L	H
L	X	H	↑	X	H	L
L	X	X	↓	X	Q(t-1)	$\bar{Q}$(t-1)

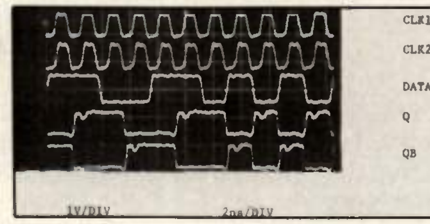

Figure 6. In the NRZ mode th PPDF operates like a Master/Slave D flipflop that is positive edge triggered.

allows large internal fanouts with minimum speed degradation. It also maintains noise margin and resistance to environmental factors without complex circuitry for temperature compensation.

Functional Description

The PPDF has three different modes of operation; Non-return to Zero (NRZ), Return to Zero (RZ), and Return to One (R1).

In the NRZ mode the PPDF operates as a positive edge triggered master/slave D Flip-flop and the outputs track the Data input. The output transitions are dependent upon the rising edge of the Clk1 input and independent of the Clk2 input. In the NRZ mode the V_{trim} input adjusts the output pulse width.

In the RZ mode Clk1 controls the rising edges of the Q output pulse and Clk2 controls the falling edges. Clk1 also controls the falling edges of QB while Clk2 controls the rising edges. The phase relationship between Clk1 and Clk2 controls the output duty cycle. The output duty cycle can be either increased or decreased by changing the phase relationship of the two input clock signals. Changing the phase relationship of these two inputs causes the independent movement of either the rising or falling edge of each pulse in the output signal. When Clk1 is advanced in time with respect to Clk2 the Q output pulse duty cycle decreases. The rising edges of Q shift to the right and the falling edges remain unchanged. When Clk2 is advanced in time with respect to Clk1 the Q output pulse duty cycle increases. The falling edges of Q shift to the right and the rising edges remain unchanged.

In the R1 mode Clk1 controls the falling edges of the Q output pulses and Clk2 controls the rising edges. Clk1 also controls the rising edges of QB while Clk2 controls the falling edges. The phase relationship between Clk1 and Clk2 controls the output duty cycle which can be either increased or decreased by changing the phase relationship of the two input clock signals. Changing the phase relationship of these two inputs causes the independent movement of either. The rising or falling edge of each pulse is the output signal and when Clk1 is advanced with respect to Clk2 the Q output pulse duty cycle increases. The falling edges of Q shift to the right and the rising edges remain unchanged. When Clk2 is advanced in time with respect to Clk1 the Q output pulse duty cycle decreases. The rising edges of Q shift to the right and the falling edges remain unchanged.

Functional Block Diagram

The functional block diagram in Figure 9 explains the complex functionality of the PPDF. The two interconnected flipflops are the logic blocks that give the PPDF it's edge placement control. The flipflops are configured so that they each have the ability to form one edge of each output pulse. In the NRZ mode one flipflop controls the placement of both output edges,

TRUTH TABLE - RZ MODE

MODE1	MODE2	DATA	CLK1	CLK2	Q	$\bar{Q}$
H	L	L	X	X	L	H
H	L	H	↑	X	↑	↓
H	L	H	X	↑	↓	↑
H	L	X	↓	X	Q (t-1)	$\bar{Q}$ (t-1)
H	L	X	X	↓	Q (t-1)	$\bar{Q}$ (t-1)

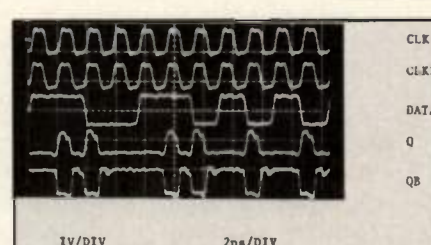

Figure 7. In the RZ mode the PPDF operates similar to the NRZ mode except that the output data is Return to Zero, and the PPDF is again positive edge triggered.

TRUTH TABLE - R1 MODE

MODE1	MODE2	DATA	CLK1	CLK2	Q	$\bar{Q}$
H	H	H	X	X	H	L
H	H	L	↑	X	↓	↑
H	H	L	X	↑	↑	↓
H	H	X	↓	X	Q (t-1)	$\bar{Q}$ (t-1)
H	H	X	X	↓	Q (t-1)	$\bar{Q}$ (t-1)

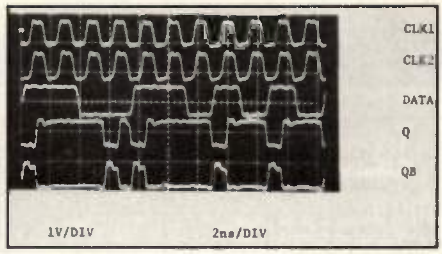

Figure 8. In the R1 mode the PPDF operates similar to the NRZ mode except that the output data is Return to One, and the PPDF is again positive edge triggered.

but in the RZ and R1 modes each flipflop determines the placement of one edge of each pulse. The mode inputs are simply used to configure the flipflops so that they execute the correct output pulse patterns which are determined by the Clk and Data inputs.

NRZ Mode

In the NRZ mode the Mode1 input is used to drive flipflop #2 so that it disables the Set and Clr capabilities of flipflop #1. This allows flipflop #1 to function as a standard D-type master/slave flipflop. In this mode flipflop #2 is only used to disable the Set and Clr functions of flipflop #1, so the Clk2 input, which drives flipflop #2, and the Mode2 input, which drives the Set and Clr inputs of flipflop #1, are in "don't care" states.

RZ and R1 Modes

In the RZ and R1 modes the Set and Clr inputs are periodically enabled by flipflop #2, which allows flip-flop #1 to form the appropriate output waveform. In this manner each of the two flipflops control either the rising or falling output edge placement. Each of the two Clk inputs is responsible for generating either the rising or falling edges of the output pulse train. Clk1 controls the rising edges of the RZ pattern and the falling edges of the R1. Likewise, Clk2 controls the falling edges of the RZ pattern and the rising edges of R1. The phase relationship of the two input clocks controls the edge placement of the RZ and R1 output waveforms.

The differential output buffer stage is an important part of the PPDF because it increases the accuracy of the output skew between Q and QB. Figure 10 shows a typical output eye diagram which has less then 50 ps skew.

Operating Information

In all three modes the differential output buffer stage allows the PPDF to adjust the output skew between the Q and QB outputs. The V_{trim} input serves as an adjustment to the differential amplifier's output current source. By adjusting V_{trim}, the output voltage can be adjusted in either the positive or negative direction. When the V_{trim} input is adjusted more positive, the PPDF decreases the output duty cycle by making the Q output pulses narrower and the QB output pulses wider. Similarly, when the V_{trim} input is adjusted more negative, the PPDF increases the output duty cycle by making the Q output pulses wider and the QB output pulses narrower. V_{trim} is typically set at −3.5 V and it's range is ±0.5 V. If low out-

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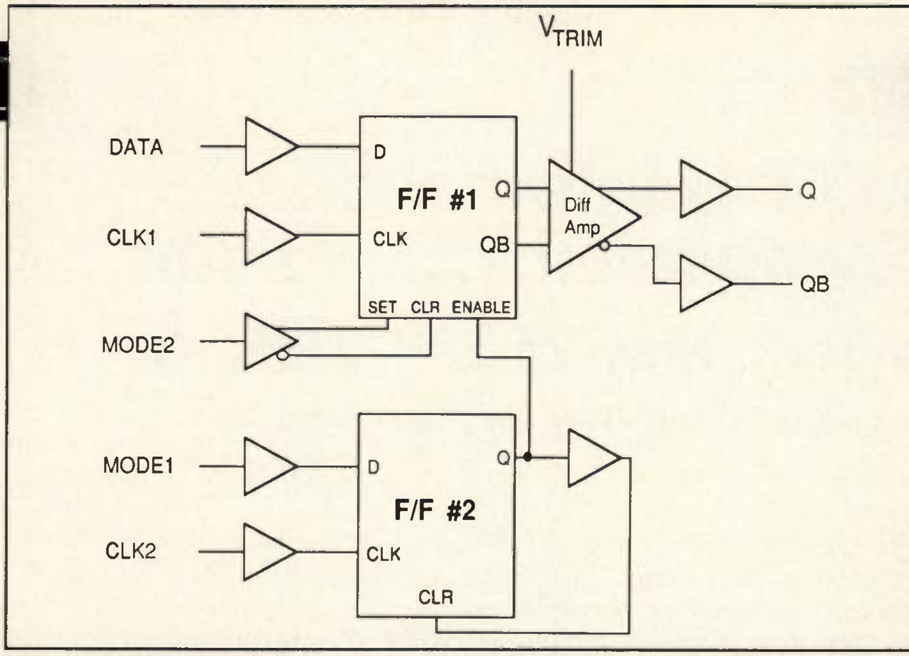

Figure 9. The functional block diagram shows how the PPDF is able to format complex variable output waveforms.

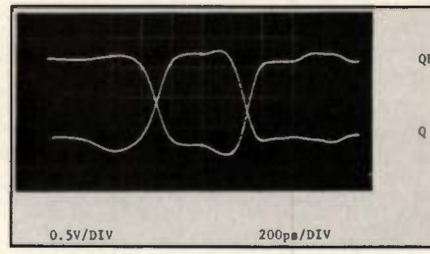

Figure 10. PPDF output skew is typically less than 50 ps.

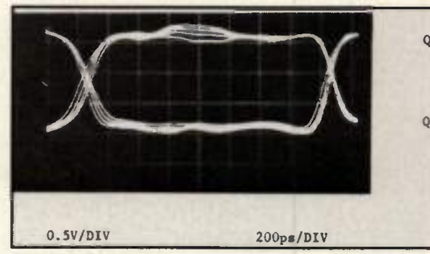

Figure 11. The photo shows a 10 percent change in duty cycle.

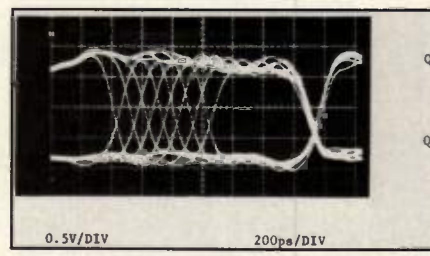

Figure 12. In the RZ mode the PPDF can modify the placement of the rising edge of the output waveform.

put skew is important for your application, then the V_{trim} adjustment can be used. For most applications the V_{trim} can be connected to the −3.5 V supply and output skew will still be very close to the 50 mV specification. Figure 11 shows the effect of V_{trim} vs. change in duty cycle.

Figures 12 and 13 show the PPDF's ability to modify the output pulse train by independently moving either the rising or falling edges of the waveform without moving both at the same time. In both cases the phase difference between the Clk1 and Clk2 inputs determines the amount by which the edges move. In Figure 12 the rising edge of Q is moved in approximately 100 ps increments while the falling edge of Q remains fixed in time. Depending on the frequency and phase difference of the two input clocks either the rising or falling edges can be independently incremented in steps as large as several nanoseconds, or as small as a few picoseconds.

In Figure 13 the falling edge of Q is moved in 10 ps increments while the rising edge remains fixed in time. Both the rising and falling edges can be independently moved in both the RZ and R1 modes.

Application

The PPDF's three different functional modes, it's GaAs/ECL compatible I/Os, high speed, and picosecond accuracy make the PPDF a perfect candidate for high speed measurement and test applications. Among these are pulse/word generation and memory testing applications. Independent control of complex pulse train edge placement make the PPDF an evolutionary extension of existing ECL pulse/word generators. Note that the high speed PPDF has additional accuracy that isn't available in current ECL products. For memory testing applications the output pulse trains can be formatted to known references so that critical access and propagation delay timing information can be accurately measured. The differential pulse trains patterns are also variable and user chosen so that DUTs can be tested with specific address and data input waveforms.

Other applications include a signal shot trigger generator and a system timing formatter. Figure 14 shows a single 300 ps pulse that could be used as a single shot trigger generator. The pulse can be generated from several different types of inputs that vary in both their frequency and edge rates. The frequency of the input clocks is used to determine the output pulse repetition rate and the phase difference between them determines the pulse width. Notice that this 300 ps, 1 Vpp pulse has approximately 150 ps edges. The pulse has almost been reduced to only a rising and falling edge. By increasing the phase difference between Clk1 and Clk2 the pulse width can be increased.

The PPDF is also able to operate as a system timing formatter because it

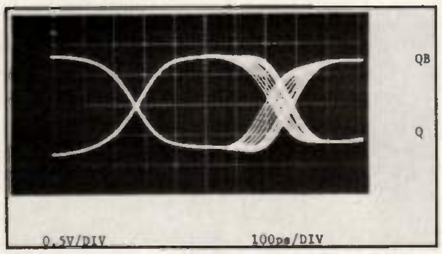

Figure 13. In the R1 mode the PPDF can modify the placement of the falling edge.

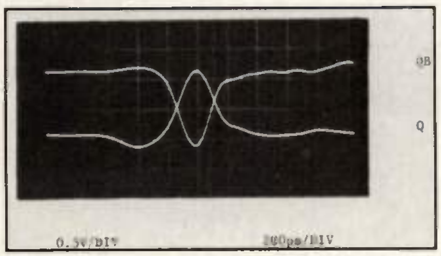

Figure 14. The PPDF can produce a 300 ps pulse with 150 ps rise and fall times.

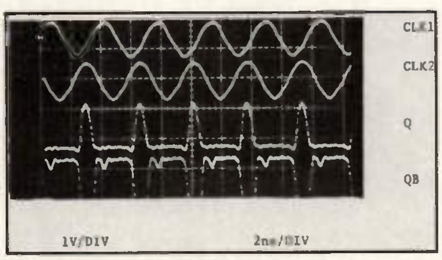

Figure 15. In this photo, low edge rate sinusoidal inputs are used to generate an output pulse drain with 150 ps edges.

sharpens-up slow, sinusoidal inputs and reduces the amount of skew between them. Figure 15 shows what happens to slow edge rate, sinusoidal inputs when the PPDF processes the differential outputs. These high speed edge rate waveforms could then be distributed around a system to be used as clocks and timing pulses.

The author wishes to thank all the Harris Microwave Semiconductor people that helped with the article including: Jerry Schappacher, Dick Davis, and Nancy Pressel with special thanks to Steve Nordblom for the technical expertise and patience necessary to build fixtures, and to test and photograph the devices. rf

About the Author

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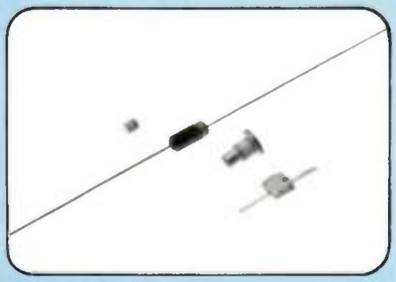

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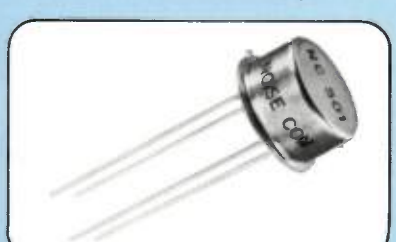

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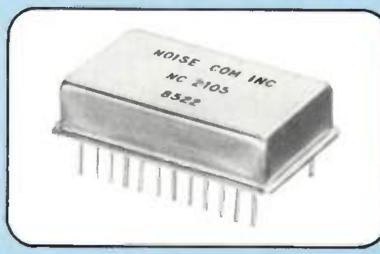

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Wide-Band Test Antennas

Simple-to-Build Discones Provide an Excellent Match at L Band

By Theodore S. Rappaport
Virginia Polytechnic Institute and State University

As part of an indoor multipath measurement system, discone antennas featuring simple N-connector feed systems have been designed for the 1.0 to 2.0 GHz band. Extensive experimentation reveals that excellent performance (VSWR < 1.5:1 across the band) can be obtained with a simple "snap-on" feed/mount method, and that VSWR is most sensitive to the diameter of the disc feed conductor. Performance data from over 70 discone antennas having a variety of flare angles, disc-to-cone spacings and feed conductor diameters are summarized here. Data shows that for N-connector mounts, best flare angles range between 45° and 75°, and that the disc feed conductor should be 0.33 times the diameter of the cone top. The empirical data reveals that the antennas may be tuned for even better match by using a simple clamp nut tuning scheme. Design equations, which differ from Nail's (2) because of the large minimum cone diameter, are presented.

The discone antenna is well documented and has been used extensively (1-6). The discone's main virtue is that it provides low VSWR over a bandwidth of several octaves (1, 2). The antenna may be modeled as a high pass filter which has a high pass cutoff frequency equal to the reciprocal of four times the cone slant height (2). As part of an experimental wideband indoor multipath measurement system (7), several discone antennas for the 1.0 to 2.0 GHz band have been developed. Each antenna uses a standard male N-connector as both an RF feed and mechanical support. This technique affords quick and inexpensive antenna construction and deployment. The data shows that antenna performance is not comprised despite the large diameter ($\lambda/12$ at high pass cutoff) of the feed connector.

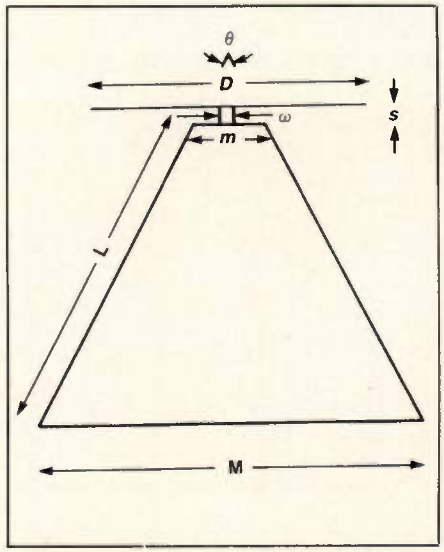


Figure 1. The discone antenna.

Antenna Construction

As shown in Figure 1, the discone may be characterized by the dimensions D, L, M, θ, m, s and ω, where m is the minimum cone diameter, ω is the diameter of the disc feed conductor and s is the disc-to-cone spacing. In the literature (refs. 2 and 5) it is usually assumed that $s \ll D$, and ω is not considered. In fact, the author has not seen previous data discussing the effects that ω and s have upon antenna loading when m is large (on the order of $\lambda/10$) with respect to the high pass cutoff wavelength of the antenna (as is the case here). Nail found that discone design formulas (for $m \cong \lambda/75$ at high pass cutoff) are:

$$s = 0.3m; \; D = 0.7M \tag{1}$$

regardless of θ, where L is slightly larger than $\lambda/4$ at cutoff (2). For a discone designed to operate at L band using a direct N-connector feed, m is equal to $\lambda/12$ at high pass cutoff, and Nail's design formulas were found to be helpful but incomplete.

Four cones made of pliable copper sheet were built with flare angles (θ) of 45°, 60°, 75° and 90°. The cones were formed by cutting and rolling the copper sheet around a wooden conical block of the desired flare angle, and then by soldering the sheet onto itself so that the cone would keep its shape. The four cone dimensions are given in Table 1. Slant lengths of all four cones were cut for $\lambda/4$ at 1000 MHz.

Each cone was soldered to the body of a UG-21D/U N-connector, giving each antenna a minimum cone diameter(m) of 19.0 mm (3/4"). With the rear end of the connector made flush with the (small) top of the cone, solder was carefully applied to the connector/cone junction to form a mechanical and electrical connection.

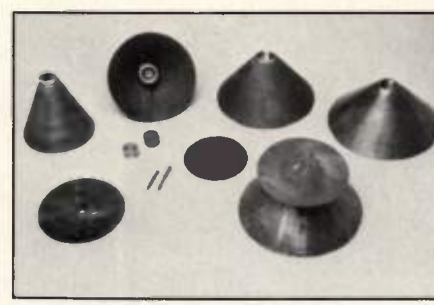


Figure 2. Discone antennas constructed on N-connectors.

Care was taken so that solder did not flow into the clamp nut threads on the rear of the connector (the clamp nut is useful for tuning the antenna). A Teflon nut inserted into the rear of the connector while soldering is an excellent way to prevent solder from blocking the clamp nut threading.

Discs of varying diameters were centered and soldered on copper rods that were beveled and soldered to the removable center pins of the N-connectors. A discone antenna was formed by plugging a particular disc into a cone-mounted con-

Parameters	Ant. 1	Ant. 2	Ant. 3	Ant. 4
θ	45°	60°	75°	90°
M	74.2 mm (3")	95.2 mm (3-3/4")	108 mm (4-1/4")	120.6 mm (4-3/4")
m	19.1 mm (3/4")	19.1 mm (3/4")	19.1 mm (3/4")	19.1 mm (3/4")
L	73 mm (2-7/8")	73 mm (2-7/8")	73 mm (2-7/8")	73 mm (2-7/8")

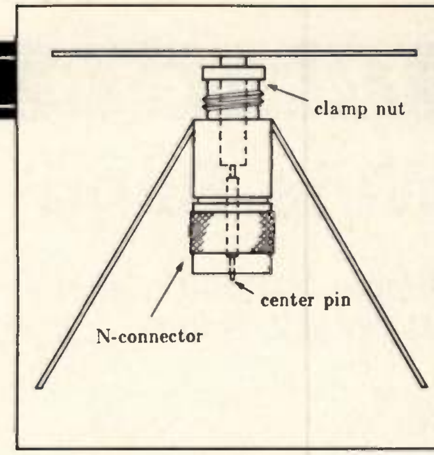
Table 1. Antenna Cone Dimensions.

nector. Antenna feed was accomplished by simply mating the discone with 50Ω coaxial cable having a female N-connector termination. Figure 2 shows several discone antennas in various phases of construction. Figure 3 details the novel disc/cone mounting method.

Experiments

Set-up

Reflection coefficient measurements were made by placing the antenna under test in a large clear area surrounded by electromagnetic absorption material. The measurement set-up consists of a UHF signal generator, a dual directional coupler (HP 776 D) and a dual channel power meter (HP 438 A) that was programmed to read return loss (= $-20 \log |\Gamma|$, where Γ is the reflection coefficient) referenced to a 50 ohm load. The system was calibrated across the band to remove the loss of the antenna feed cable as well as the frequency dependent variations of the equipment. Antenna return loss measurements were made at 50 MHz intervals across the 1000-2000 MHz band. Care was taken to insure power levels were sufficient to provide reliable measurements. Antenna performances were evaluated by the average value of the reflection coefficient across the 1.20 to 2.0 GHz band.



Results

Equation 1 was first used to develop four discones described in Table 1. Of particular importance, the antennas had parameters $s = 0.33\ m$ and $\omega = 0.16\ m$. Clamp nuts were not installed on the connectors, and a value of $D = 0.7M$ was used for each antenna. The antenna loading characteristics of all four discones were disappointing, and are shown in Figure 4.

The disc separation, s, was then varied in increments of $0.082\ m$ (m is not "meters," but refers to the minimum cone width) over the range *[0m, .66m]* for each of the four antennas to obtain lowest average VSWR over the band. For antenna 1, optimum separation was found to be $0.41\ m$ (8.0 mm, 5/16"). Antennas 2, 3 and 4 were optimized when $s = 0.5\ m$ (9.6 mm, 3/8"). These values are 50 percent larger than the values suggested by Nail in Equation 1. Even with optimum disc-to-cone spacing, the loading characteristics of all four antennas were still poor. For each antenna, the VSWR averaged about 2.0:1 throughout the band. Thus, the data indicates that for large m ($m \cong \lambda/10$), the common discone design equations (e.g. found in refs. 2 or 5) are lacking.

With s optimized, the disc parameter D was varied in 0.05 M increments over the range *[0.55M, 0.75M]*. The loading characteristics of each antenna was slightly affected, most noticeably at the low frequency end and at the first harmonic of the cutoff frequency. A value of $D = 0.70\ M$ provided the best broadband response for antennas 2 and 3, whereas antennas 1 and 2 performed slightly better with $D = 0.75M$. All subsequent measurements were made with disc values of $D = 0.75M$. This is in close agreement with Nail (2).

Changing the diameter of the disc feed connector dramatically affected the loading performances of the antennas. Values used for ω range from 0.082 m to 0.41 m in 0.082 m increments. For each value of ω, the disc-to-cone spacing (s) on each antenna was readjusted for lowest VSWR. It was found that best matching for nearly all antennas occurred for values $\omega = 0.33\ m$, $s = 0.5\ m$. Figure 5 illustrates the effect of variations in ω upon the loading characteristics of antenna 3 with $s = 0.5\ m$ and all other antenna parameters fixed. The other three antennas demonstrated very similar behavior, although the 90° antenna provided a consistently poor match over the lower part of the band. Figure 6 displays the loading characteristics of the four optimized discone antennas, each having parameters $\omega = 0.33\ m$, $s = 0.5\ m$, $D = 0.75\ M$, and cone dimensions given in Table 1.

From the data it appears that a large disc feed conductor diameter mitigates a substantial impedance mismatch within the connector. Due to the structure of the UG-21D/U connector, the disc feed conductor travels a non-negligible distance ($\cong\lambda/10$) within the throat of the connector before reaching the connector end. Thus it seems that a simple coaxial transmission line model of the feed conductor within the connector is applicable. The impedance of an air-dielectric coaxial transmission line is well known to decrease with increasing center wire diameter (8) and is solely a functon of the ratio (m/ω). Thus the results shown in Figure 5 are not surprising. For $\omega = 0.33\ m$, the characteristic impedance offered by the feed conductor/connector transmission line segment is 66.5Ω, very close to a 50Ω match.

The experimental results justify the modification of Nail's original design equations to include relationships for ω and s when m is large, on the order of $\lambda/10$ at high pass cutoff (lowest frequency of operation). Our data suggests the following design equations for the case when the cone top has a diameter of $\lambda/10$ at high pass cutoff:

$$s = 0.5m;\ \omega = 0.33m;\ D = 0.75M;\ L = 1.15\lambda/4 \qquad (2)$$

In Equation 2, the value for λ corresponds to the lowest desired frequency of operation for the antenna. Because the impedance relationship within the feed housing is a function of both m and ω, Equation 2 should hold for *any* discone coaxial feed system with a large m dimension.

A second independent set of four discone antennas was fabricated using Equation 2 to test the reliability of the design formula. Measurements performed on the second set of antennas yielded results within ±2.0 dB of those shown in Figure 6 for most frequency points.

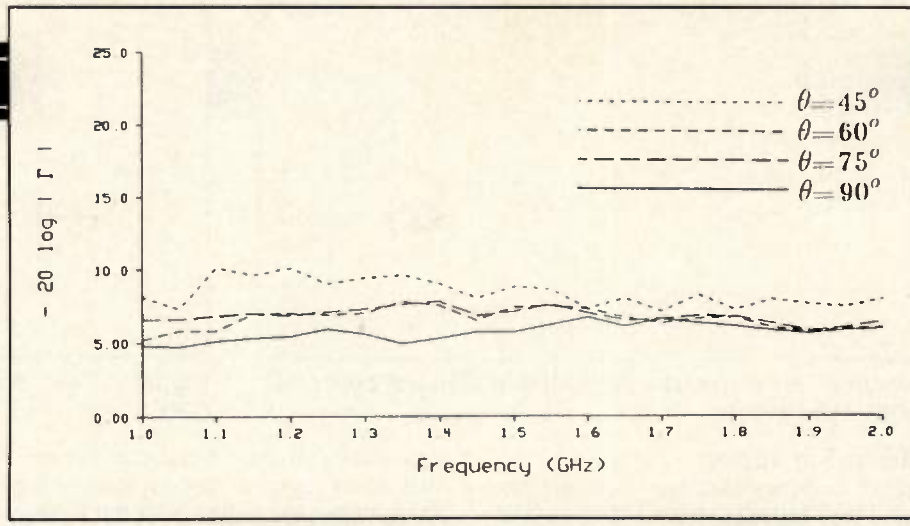


Figure 4. Loading characteristics of discone antennas (designed from Equation 1).

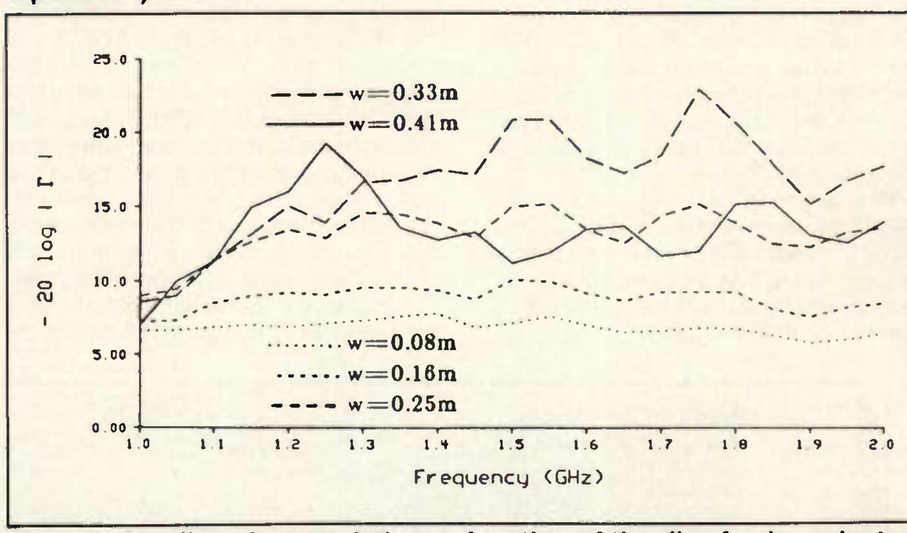


Figure 5. Loading characteristics as function of the disc feed conductor diameter (θ = 75°).

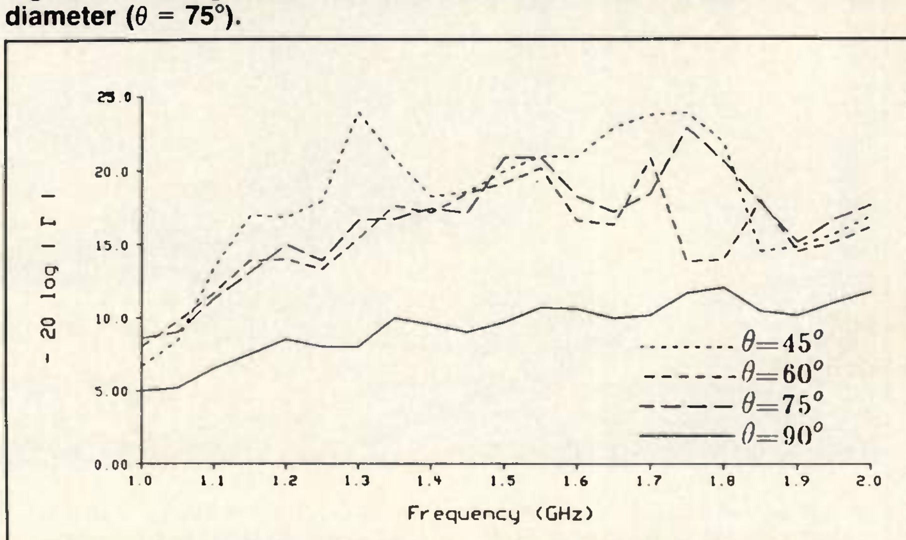


Figure 6. Loading characteristics of discone antennas (designed from Equation 2).

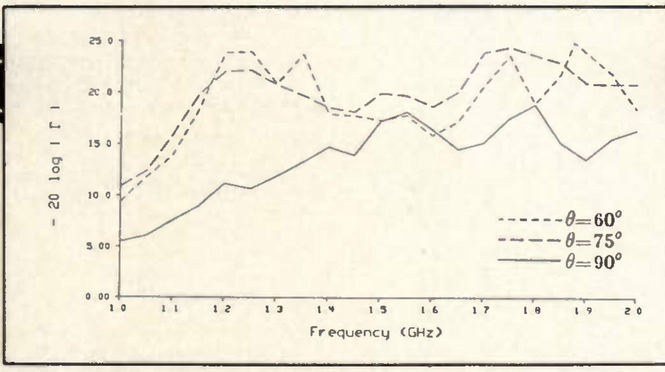


Figure 7. Loading characteristics of discone antennas (with clamp nut).

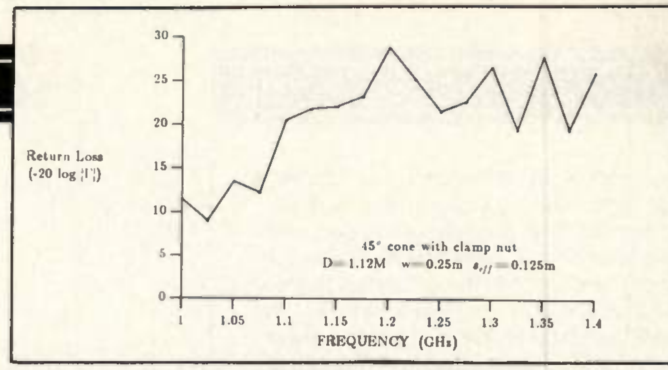


Figure 8. 45° discone antenna optimized for 1.1-1.4 GHz.

Clamp Nut Tuning

Further experiments were conducted to see how inserting the clamp nut into the connector would affect antenna loading characteristics. An immediate benefit of the clamp nut is dimension *s* can easily be adjusted. In addition, the clamp nut serves to further reduce the impedance mismatch created within the connector, and may be thought of as a low impedance transmission line segment in a short tapered line.

By fastening the clamp nut into the connector, improvement in VSWR performance for three of the four antennas was accomplished. The 45 ° discone performance deteriorated with the clamp nut installed (when D = 0.75M).

An optimum ω value of 0.25 *m* (4.8 mm, 3/16") was found to hold for each of the three discones using clamp nut tuning. Disc-to-cone spacing was kept at 0.5 *m* (9.6 mm, 3/8"); however, with the clamp nut fully inserted, the distance from the top of the clamp nut to the disc (s_{eff}) was decreased to 0.33 *m* (6.4 mm, 1/4"). All other parameters remained the same. Figure 7 illustrates the loading behaviors of the three discone antennas with clamp nuts securely fastened on the top of the cones. For other disc feed conductor diameters, clamp nut tuning improved the reflection coefficient (see Figure 5) by 5 dB on the average throughout the band.

For the 45° discone, additional experiments were conducted to determine the effects of disc diameter upon loading when clamp nut tuning is used. In particular, we strived to lower the high pass cutoff frequency without changing the cone dimensions. Figure 8 shows the matching characteristics of an optimized 45° discone measured at 25 MHz increments across the 1.0-1.4 GHz band. The data illustrates that with clamp nut tuning, better low frequency response may be obtained by simply increasing the disc diameter. The results also suggest that for large minimum cone diameters (on the order of λ/10 at high pass cutoff), antenna loading anomalies due to disc feed conductor diameter, ω, and disc-to-cone spacings, *s*, can be neutralized by a cable clamp nut (a standard part supplied with all coaxial connectors).

Conclusion

Discone antennas are quickly and easily constructed as appendages to coaxial connectors. Such antennas are easily deployable and are suitable for use in bench testing in UHF/microwave wideband indoor communication systems. In this article, extensive experimental data were analyzed to determine design equations for discone antennas mounted on N-connectors. Such a mounting forces the minimum cone diameter m to be on the order of $\lambda/10$ at high pass cutoff. For this case, the data illustrates the significance of antenna parameters s and ω upon antenna loading. For discone antenna design using coaxial connector feeds, Nail's equations (2) must be modified to include the effect of the diameter of the disc feed conductor. By selecting the disc feed conductor diameter to be 0.33 m, the experiments reveal that a VSWR below 1.3:1 is easily achievable across an octave and further suggests that cone flare angles between 45° and 75° yield best results. Equation 2 is a suitable modification of the well-accepted discone design equation and should hold for any discone coaxial feed system when m is large. By using a simple clamp nut adjustment and slightly decreasing the diameter of the disc feed conductor (ω), it is possible to optimize VSWR characteristics of a discone antenna mounted on a coaxial connector. The construction, design and tuning techniques described here should be valid for discones designed to operate at much higher frequencies, although this must be borne out by experimentation.

The antenna described is being patented and any party wishing to license, buy or market the product can either contact the author at the number listed or Dr. Bill Baitinger of Purdue University at (317) 494-5785. rf

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Quality Assurance Issues For EMC

Part I: Continued Compliance Requirements

By Mike Howard
Norand EMC Test Lab

Is the task of EMI/EMC testing completed with the certification test demonstrating compliance of the engineering model or the initial production unit? The answer is "No". Continued compliance or audit testing is required to successfully ensure that production quantities maintain compliance with the respective EMC test standard or requirement. This article discusses a structured approach in auditing for continued compliance using a 10 dB guardband for design, auditing and sampling guidelines. It is focused around FCC Part 15 Subpart J for computing devices. However, this structured audit approach for continued compliance may be used for other EMI/EMC standards as well.

What constitutes the necessity for continued compliance? Mandatory requirements, manufacturing tolerances, process variables, quality defects, product design changes, component selection and cost reduction changes are just a few examples of items that may be considered insignificant at the time of implementation. However, it can significantly alter the EMI/EMC compliance of the product. The fines caused by non-compliance for FCC Part 15-J can be expensive and serve as an additional incentive to maintain compliance. For example, in 1982, the FCC issued $11,350 worth of fines, while in 1986 this figure rose to $878,750. During this period the number of citations rose from 33 to 959 (1).

A case history demonstrating the necessity for continued compliance involves a manufacturer of a control device utilizing a remote control receiver powered by 115 VAC. From the initial certification test of this product, it was empirically determined that a .033 uF decoupling capacitor from Vendor A was required across the hot and neutral lines of the AC power to decouple the differential mode interference noise present to comply with the conducted emission requirements of FCC Rule Part 15.63(b) for receivers in the frequency range of 450 to 600 kHz. The .033 uF capacitor provided a 4 dB margin of compliance. Six months later the same manufacturer requested a retest of this product to qualify a new vendor (Vendor B) of the .033 uF capacitor. It was found that the use of the new part from Vendor B resulted in conducted emission levels 2 dB over the FCC limit as specified in 15.63(b) from 450 to 600 kHz. Vendor A and Vendor B capacitors both were found to be within their specified tolerances for a .033 uF capacitor. Upon further investigation, Vendor B's capacitor that contributed to the product's non-compliance was found to have a much higher ESL (Equivalent Series Inductance) and ESR (Equivalent Series Resistance) as compared to Vendor A. This is an excellent example to demonstrate that all components are not created equal when considering Electromagnetic Compatibility of the product. If the manufacturer of this product had not considered an audit test to verify compliance of his product with the new vendor part, he would have run the risk of producing large quantities of product that were non-compliant. This would have resulted in possible fines, rework costs, lost sales, or worse, the seizure of the non-compliant product.

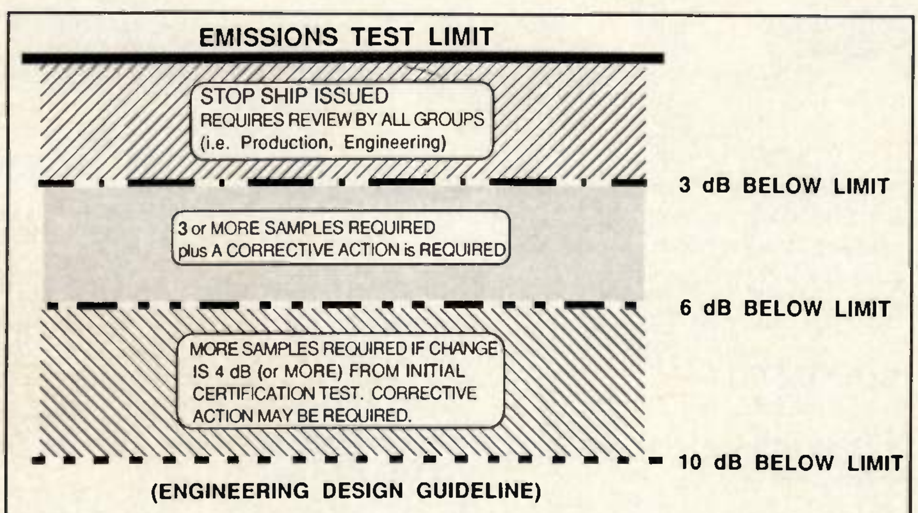

Figure 1. Audit/sample requirements based on a 10 dB guardband.

Mandatory Requirements For Continued Compliance

This section takes a look at some of the legal requirements as specified under the U.S's FCC regulations and West Germany's VDE-0871 requirements for continued compliance. Table 1 shows the rule parts within FCC regulations that deal with continued compliance and audit requirements (2).

To date, two public notices from the FCC, issued on April 7, 1982 and September 18, 1986, have cautioned manufacturers against changes in their product and the effect these changes may have on the compliance of the product with commission rules. The rule parts are also specific in the manufacturer's responsibility pertaining to changes made to their product. As to specific guidelines for determining and maintaining continued compliance, the mentioned rules leave this to the manufacturer's interpretation. Guidelines for the manufacturer in this area are presented later in this article as well as in Part II of this article.

Section 4.1.3.4 of VDE-0871 states that for the verification of production run product, it is sufficient to select items at random for test purposes. If a limit is exceeded by 2 dB or less, the production run is still acceptable if it can be proven that a minimum of 80 percent of the equipment does not exceed the limit with an 80 percent confidence level (3). This method is commonly referred to as the 80/80 rule. A method of computing this 80/80 rule can be seen in Figure 2.

Simply stated, compliance with the requirements of FCC Part 15 and VDE-0871

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SAS-200/511	1000 - 12000 MHz	Log Periodic
SAS-200/512	200 - 1800 MHz	Log Periodic
SAS-200/518	1000 - 18000 MHz	Log Periodic
SAS-200/530	150 - 550 MHz	Broadband Dipole
SAS-200/540	20 - 300 MHz	Biconical
SAS-200/541	20 - 300 MHz	Bicon'l Collapsible

MODEL #	FREQ. RESP.	DESCRIPTION
SAS-200/542	20 - 300 MHz	Biconical, Folding
SAS-200/550	.001 - 60 MHz	Active Monopole
SAS-200/560	per MIL-STD-461	Loop - Emission
SAS-200/561	per MIL-STD-461	Loop - Radiating
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B. Sampling of the production line at regular intervals

Variability

The importance of ensuring compliance from the initial design to the production line cannot be overstressed. The opportunities to have the product become non-compliant from the initial certification test to the production line are in abundance. The seemingly insignificant variables are what must be closely controlled, from the initial certification test to implementation in production. Where does this variability come from? The following are just a few examples of variables that must be addressed and can be detected from continued audit and compliance testing for EMC (4).

In the manufacturing process variability can arise from insufficient documentation, assembly errors, unqualified components, contact contamination, improper plating, vendor process changes, human errors, product workmanship, and tool wear and utilization. Variability in product design and functionality is also a factor that has to be considered. This can arise from the number of configurations and operations involved, end usage and application of the product, implementation of EMI/EMC fixes on a production basis, and cost reduction programs without consideration of EMC. In test facilities, the number of samples evaluated, errors in instrumentation and test site, test methodology and calibration methods used, EUT setup, configuration and operation, and inadequate quality assurance program for test site and personnel are also variables that should not be neglected.

By addressing the variables that exist in each of these three areas, one can then attempt to maintain compliance of the product throughout it's expected life cycle. Specific guidelines need to be in place to adequately ensure this continued compliance while addressing variability within the process of EMC compliance.

First, let's look at the relationship of product RF emissions in comparison to an emission limit. We know there will be a certain degree of error in the system used to make the emission measurements, as well as from variability in product to product measured. Structured procedures must be applied both to the measurement task and the product being evaluated and measured. Repeatability and accuracy must be maintained separately in both cases. If we start from the beginning with a prototype or engineering sample, we would like to strive for an engineering design goal. In this case, as seen in Figure 1, a 10 dB margin is desired.

Where is the 10 dB figure derived from? Table 2 shows where some of the sources of error can originate (5). These represent typical examples of potential errors. Cases exist where I/O cable placement alone resulted in errors of 10 dB or more. The 10 dB guardband is not a guarantee of maximum errors that can exist in the measurement process of product emissions. It serves only as a starting point for a measurable reference level in achieving overall compliance of a product.

The purpose of this 10 dB margin is to give greater flexibility for errors that may arise from the measurement task as well as process variables that could occur

$$\bar{X}_n + k s_n \leq L$$

Where $\bar{X}_n$ = Arithmetic mean value of the measured interference values of n test samples:

S_n^2 = Sample variance $s_n^2 = \frac{1}{n-1} \sum (X - \bar{X}_n)^2$

X = The measured RFI value of each test sample

L = The limit at each measured frequency

k = A factor derived from the tables of the non-central t-distribution for a 80% confidence that at least 80% of the production will be below the limit:

n = The number of test samples.

n	3	4	5	6	7	8	9	10	11	12
k	2.04	1.69	1.52	1.42	1.35	1.30	1.27	1.24	1.21	1.20

Figure 2. Calculations for 80/80 rule as derived from VDE-0871/6.78, section 4.1.4.

within the product from the time of the engineering sample to final production versions. If the initial engineering sample does comply with the design goal of all product emanations being 10 dB or greater below the required test limit, then, once the pilot or production units have been constructed incorporating the EMI fixes from the engineering model, these units will have a greater margin of acceptability. The 10 dB guardband has been selected on the basis of limited engineering samples (i.e three or less). If more engineering samples could be secured (i.e. four or greater) then the guardband could possibly be reduced to perhaps 6 dB. From the author's personal experience, though, if the engineering samples can be developed to meet the 10 dB guardband criteria, then the percentage of pilot and production units found un-acceptable for FCC/VDE requirements is significantly reduced. The 10 dB guardband also has the advantage of a built in design margin for the addition of future options or peripherals to the product. It in effect minimizes additional development time for EMI fixes for the overall system by having a higher margin of acceptability for each item or component of a system for FCC/VDE requirements.

This approach can be of particular benefit in today's personal computer market. A whole host of manufacturers are producing a broad range of peripherals and features for today's personal computers. If both the personal computer and peripheral manufacturer would strive to have their product meet this 10 dB (or even 6 dB) design guideline, then the ability to comply with FCC/VDE limits when integrating these individual products/ components together as a system would be more easily attainable.

A current frustration for today's test lab and manufacturer is the mix and match procedure required to find peripherals and personal computers that, when combined as a system, will comply with FCC or VDE test requirements. Many times this mix and match procedure has to be completed before the manufacturer can even consider testing his/her computer or peripheral within the system for qualification to FCC/VDE requirements. This even occurs with outright purchases of computers and peripherals off the computer store shelves and, when assembled as a system, the system is found to be unacceptable to the FCC/VDE regulations. However, individual items tested separately will more than likely be in compliance. The 10 dB guardband design goal is not the total solution to resolve this current mix and match problem, but can greatly reduce the frustration of integrating your product with another manufacturer's product and finding that the final assembled system is non-compliant with FCC/VDE emission limits. Additional investigation of the mix and match problem should be required by the FCC, VDE, ANSI, IEEE, and other EMI/EMC committees with findings of their investigations reflected in the rules and measurement procedures of the regulating agencies such as the FCC and VDE.

What happens when the 10 dB engineering design goal is met and the manufacturer has moved on to pilot or production run audits of his product? Again, referring to Figure 1, if the audit of the product has shown product emissions located in the 6 to 10 dB region below the limit, no additional samples would be required for test to demonstrate compliance. However, if the emanations have risen more than 4 dB from the initial engineering qualification test emissions recorded, then it would be advisable for engineering to review and investigate this delta increase in emissions. The question that must be asked is why the change and how did this change come about. Some form of corrective action may result from this investigation to prevent the delta emissions recorded from varying more than 4 dB from the initial engineering test units emission levels recorded.

If, during the audit, the emission levels are found to be 3 to 6 dB below the limit, then a minimum of three additional test samples would be required to ensure that compliance is maintained statistically for the production run at the time of sample. Mandatory review by engineering is required to investigate the upward shift in emission levels from the initial qualification test of engineering units. If it can be proven that the process is not changing in a continuing upward trend and the emission levels are maintained within this 3 to 6 dB range with statistical data from sample lots of 3 to 12 units, then these production run units would still be acceptable for continued compliance. The rate of sampling may need to be increased here at this time to ensure that the process is acceptably stable. If emanations are found to be within 3 dB of the limit, or higher, then *Stop Ship* requirements are in order at this point. This is taking into consideration the initial design goal of product emissions at least 10 dB below the emission limit. Emissions within 3 dB of the limit can easily be enhanced by many factors that can place these emis-

FCC Rule Part	Description
2.931	Responsibility of the grantee
2.932	Modification of equipment
2.936(c)	FCC Inspection
2.937	Equipment defect and/or design change
2.938(a)(2)	Retention of Records
2.945	Sampling tests of equipment compliance
2.953	Responsibility of manufacturer or importer
2.955(a)(2)	Retention of records (for verification)

Table 1. FCC rule parts with descriptions.

Source	Probable Error
Test setup and operation (i.e., cable placement)	4 dB
Test instruments and sensors (i.e., receivers, RF cables and antennas)	4 dB
Misc. errors	2 dB
Total	10 dB

Table 2. Derivation of the 10 dB margin of error.

sions at or above the emission test limit. Many of these factors include the installation of the product, product operation, or I/O connections as just a few examples. The reason for issuing a *Stop Ship* at this point, is that the production emissions are of an unacceptable margin as referenced to the limit and the possibility that these emanations could easily be at or over the emission test limit with the right stimulus or condition. All parties involved from engineering to production must give this the highest priority to ensure that this does not become reality. The *Stop Ship* gives the needed priority required to accomplish this task.

The requirements for production run compliance for VDE-0871 is quite different than what has been described previously as a guideline. VDE states that if, during an audit of production units, a unit is found with emissions exceeding the limit, the production run is still acceptable if a minimum of 80 percent of the units in the production run do not exceed the limit with an 80 percent confidence level. Five to twelve samples are required to demonstrate this 80/80 rule. However, if five or more samples are not available then three or four test units would be acceptable. The statistical equation for the 80/80 rule can be found in Figure 2.

The process of performing and acquiring emissions data for FCC/VDE test requirements requires a quality assurance program to be in place. Here again, the repeatability and accuracy of the measurement process must be maintained from day to day and from test to test.

Some key items that must be maintained by the measurement facility to achieve repeatability and accuracy are

- Calibration standards traceable to NBS
- Maintain test artifacts/calibrators for daily site verification and instrumentation checks
- Knowledgeable and qualified test personnel
- Proven and written measurement procedures implemented
- Quality assurance & calibration procedures developed and implemented
- Continuing EMC education programs in place for all personnel

When the above items are closely controlled, then more attention can be given to the product being evaluated for compliance without having to investigate measurement abnormalities resulting from these items. The issues surrounding the evaluation of the product for EMC are complex enough without having to differentiate errors resulting from the EMI/EMC measurement task or process.

Both from the test facility and the product evaluation, detailed and accurate records must be maintained. Having thorough and detailed documentation summarizing all testing and design implementation will aid all future audits of the product for compliance. Without this detailed documentation, it becomes very difficult to try and duplicate the original certification test. If a facility has the storage space and funds, it would be beneficial to archive all product initially certified, along with the documentation, for comparison to units audited in the future. If this is not possible, then the requirement for detailed documentation of the actual EMI test and product status becomes paramount. This documentation must be maintained at, and during, the time of the initial certification test.

Summary

The responsibility for continued compliance is shared by several groups including engineering, the EMC test lab, quality assurance, and production. Audit testing and continued compliance cannot be overlooked! Both the FCC & VDE have regulations specifically dealing with continued compliance of the product. The processes of design, testing, and production contain variables that need to be recognized and controlled efficiently to maintain a level of acceptable compliance. The 10 dB guardband has been reviewed herein to give a starting point for considering engineering design goals as well as audit requirements. It is not the total solution to continued compliance but one of many tools that may be used to achieve and maintain compliance for EMC.

The design engineer, EMC test lab personnel, QA and production personnel need to have a better understanding of process changes and variables that can affect the compliance of a product to EMI standards such as FCC Part 15-J and VDE-0871. Part II of this article, to be published next month, discusses a tool called Statistical Process Control (SPC) using X-BAR and R charts to provide a histogram of the process, both at the EMC Test Lab and Production. It will give the necessary details to implement an SPC program for continued compliance for EMC.

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High-Frequency Design With Analog Master Chips

These Versatile Devices Have Some Distinct Advantages Over Discrete Components.

By John Shier
VTC Inc.

An analog master chip (AMC) is a monolithic integrated circuit that contains a large number of uncommitted circuit elements. A unique pattern of element connections (wires) is generated for each application of the AMC, which creates amplifiers, flipflops, etc. Because only the last 1-3 masks needed for the interconnections are unique to a given customization of the AMC, the cost and fabrication time for an AMC is much less than alternative IC methods (cell libraries and full custom designs). This article sets forth basic principles for high-frequency design using AMCs, and gives an example which illustrates their application in a particular case.

AMCs have been available since the early 1970s, but until recently the processes used yielded high-voltage low-frequency transistors typical of conventional analog ICs. The unity-gain frequency f_t was 300 MHz for the NPNs and only about 5 MHz for the PNPs, inadequate for most RF designs.

There are many opportunities for the equipment designer in using AMCs, but to fully realize them the designer must re-examine current practices with discrete-element designs, since the optimum circuit solution is usually significantly different. The fundamental AMC advantage is that it is smaller, consumes less power, has less power dissipation and costs less. However, for the RF designer there are some other advantages which are less obvious.

Any RF circuit element radiates electromagnetic waves, that can cause undesired crosstalk and interference in other parts of the circuit. The use of an AMC greatly reduces the radiated energy for two reasons: a) The signal currents can be much smaller, so the generated fields are of much lower amplitude and, b) The on-chip interconnections are extremely short and thus make very poor antennas for either receiving or radiating RF energy.

One of the main things dividing RF design from low-frequency design is the need to consider distributed elements for side effects such as transmission line effects. This makes RF design substantially more complex then conventional lumped circuit design. The very short interconnections used on an IC mean that the lumped approximation remains valid to frequencies an order of magnitude higher than for discrete-element designs. Line lengths rarely exceed 1-2 mm, and the on-chip propagation delays are usually less than 50 ps. On-chip inductances are generally negligible.

Design Objectives for Analog Master Chips

An AMC is a different medium for circuit design, and it is not surprising that methods and objectives change. The optimal integrated design requires a different design philosophy, and a rethinking of conventional RF design practices. Below are some basic factors that should be considered.

Cost — An AMC can have hundreds of transistors. Their cost in volume may be only .1 cent each, and there are often lots of them left over after the design is finished. Much of conventional RF design practice makes the unstated assumption that active elements are expensive and discretes are cheap. When using AMCs, if you can eliminate a capacitor or inductor with 10 transistors it is almost always cheaper to do so. A consequence of this is that active circuit solutions are preferred over passive ones.

Signal Path — The designer should keep as much of the signal path on-chip as possible if the benefits of low-power lumped design are to be realized. Whenever the signal goes off-chip there are sizable parasitic elements to deal with. An on- chip node rarely has more than 1 pF of capacitance, while off chip nodes seldom have less than 10 pF. Thus extra gain stages are needed to drive the parasitic load elements with a corresponding increase in power dissipation and possibilities for increased noise, distortion, RF interference, etc.

Off-Chip Discretes — If the same R, L, and C elements off-chip are used, merely shrinking the active circuitry with an AMC is hardly worth it. Off-chip discretes have other disadvantages. They serve as antennas to pick up RF interference. Also, the pins themselves are expensive. Each added pin will add 2-5 cents to the package cost. An off-chip discrete usually re-

Input capacitance (to adjacent pins and ground)	3-10 pF (DIP) ~1 PF (28-pin LCC)
Pin self inductance	8-25 nH (DIP) ~4nH (28-pin LCC)
Chip bonding pad capacitance	.3-.5 pF
Chip wire bond inductance	1-2nH

Table 1. Typical parasitic elements for IC packages.

Transistor Type	Name	R_B ohms
Minimum-size NPN	T6NW	450
Low-Noise NPN (large)	TLN1	9
Minimum-size PNP	TVPA8	560
Large PNP	TVPA35	70

Table 2. Room temperature base resistance for transistors used in the VJ900 series.

quires several mA of drive current and an additional driver gain stage, while on-chip 0.1-1 mA is more typical.

This change in design emphasis and objectives means that RF design with AMCs involves more than just translating a design transistor-by-transistor. Truly optimal design will often require that the whole circuit design of a given block be redone.

The overwhelming majority of analog ICs are DC coupled. This is done because it is impossible to get large-value capacitors on a silicon chip. The use of interstage coupling capacitors and transformers should be abandoned or minimized.

Differential Signals

Differential signals are quite popular in analog ICs, and they help in solving some of the problems mentioned above. A traditional objection to differential signals is that they need twice the number of components. However, this doesn't apply to ICs. Differential signals help greatly in solving the coupling problem, which accounts for much of their popularity. It is all right for a given stage to have a large and temperature dependent DC offset between output and input, since the other paired differential signal will have the same offset. The good device matching makes this attractive on an IC. One can readily achieve differential stages with input offset voltages of 1-2 mV. Often several stages can be cascaded before it is necessary to go off-chip with coupling capacitors to get rid of the accumulated offset.

Differential signals also help with RF interference problems, both between chips and on-chip. As mentioned earlier, on-chip elements are relatively immune to radiated RF fields. They do, however, have enough capacitive coupling to other on-chip elements that one can sometimes get undesired crosstalk. Note that this can be minimized with proper chip layout.

Integrated Resistors

Integrated resistors are readily available on-chip. Compared to discrete resistors, however, they do have some limitations. Perhaps the most important limitation is on resistor value. An entire master chip may have only 1 to 2 M ohms of resistors, i.e., if all were connected in series 1-2 M ohms would be realized. In typical IC designs most resistors are between 200 ohms and 10 k ohms.

The on-chip resistors also have a significant nonlinear temperature dependence, amounting to around 1000-2000 ppm/°C. Since they are doped silicon regions embedded in a substrate and isolated by a reverse-biased PN junction they also have a small voltage coefficient, which amounts to about .1-1 percent per volt.

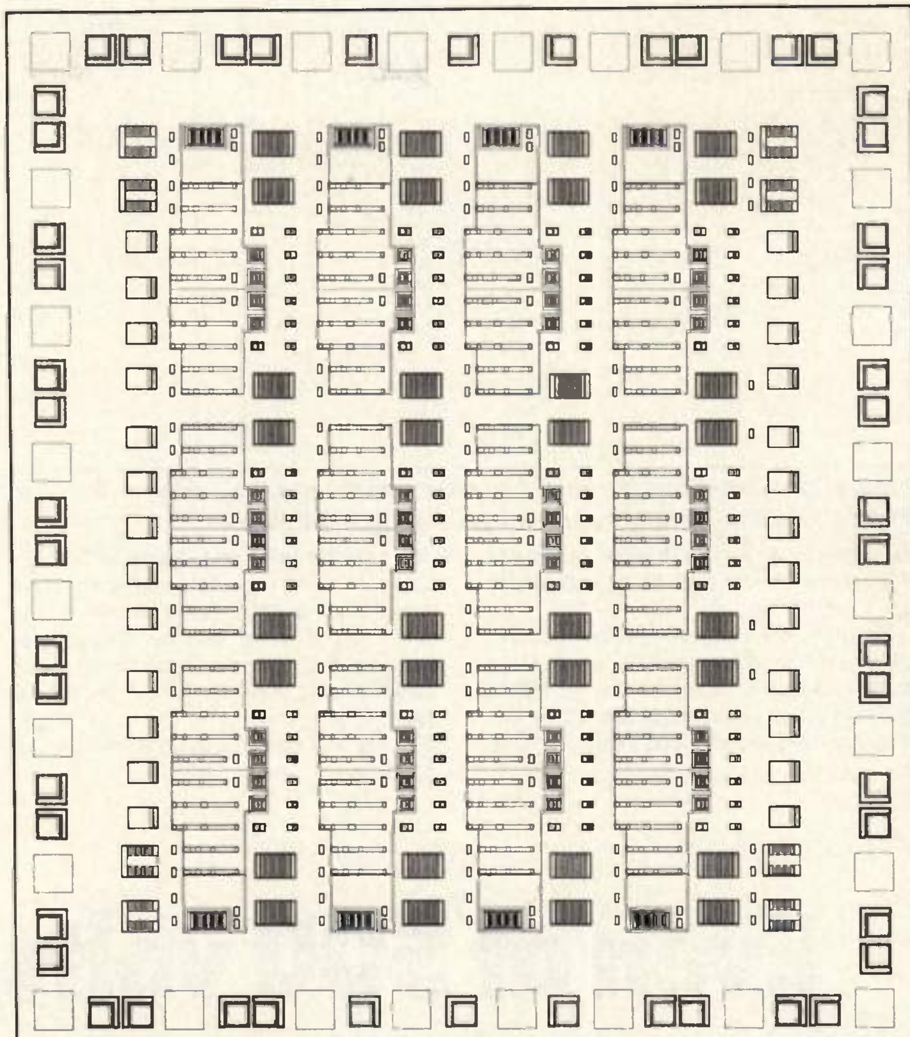


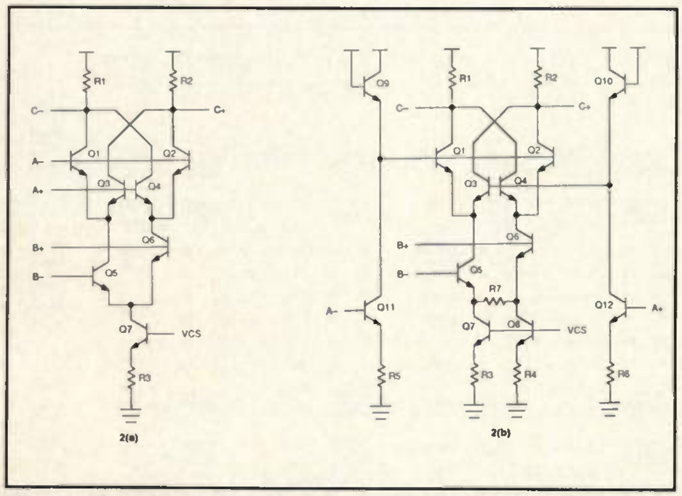
Figure 1. An example of an AMC — the VTC VJ960.

The control of absolute resistor value is relatively poor — about ±20 percent. Resistor ratio matching, however, is good with a ratio standard deviation around .1-3 percent. Silicon is an excellent thermal conductor, and all of the circuit elements on a chip will be at almost the same temperature. Temperature differences within a single chip rarely exceed .1-1°C.

Integrated Design Using Capacitors and Inductors

One form of capacitor often used by IC designers is the junction capacitor. This is basically a reverse-biased PN junction. Such capacitors are often implemented by connecting the emitter and collector of an NPN as the + terminal and the base as the – terminal. A disadvantage of this capacitor is that it is really a diode and shows forward conduction when forward biased. It can only be used where the capacitor bias voltage is always of the same polarity. An NPN connected as described also has a reverse (Zener) breakdown voltage around 5 V, and thus the magnitude of the capacitor bias must be limited. An additional problem is that such a capacitor is nonlinear. It does not obey Q = CV but a different, nonlinear charge-voltage equation. It is, in fact, a kind of varactor. The most common use of junction capacitors is for stabilizing (compensating) DC and other low-frequency feedback circuits.

Some AMCs provide a number of linear (oxide) capacitors, and by paralleling several of them, values of several pF can be obtained. These values can be quite useful on an IC at high frequencies and with small signal currents. Note that these are not MOS capacitors. Although they use a thin thermally-grown SiO_2 film as the dielectric, the underlying silicon is



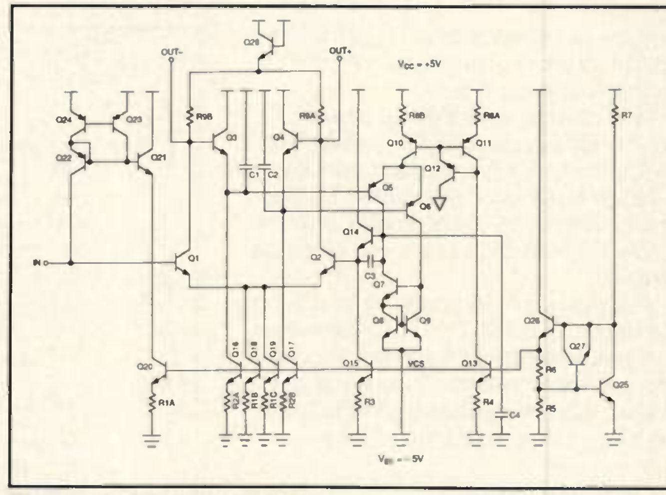


Figure 2(a) Basic circuit for a balanced modulator.
Figure 2(b) The configuration for an analog multiplier.

Figure 3. RF input stage.

very heavily doped so that the nonlinear C-V relation seen with MOS capacitors does not occur.

A significant restriction on these linear capacitors is their parasitic substrate capacitance. An IC is a planar structure and one of the capacitor plates will be adjacent to the substrate. This plate normally has a substantial capacitance to the substrate. In certain devices the lower plate is a P type silicon region and this substrate capacitance will be a nonlinear junction capacitance. Thus when using these on-chip capacitor circuit configurations, where one of the capacitor plates is at the VCC or VEE supply voltage, the parasitic capacitance is unimportant. This can be frustrating at times because many of the most important active filter configurations involve floating capacitors with signal voltages at both ends.

Apart from these parasitic effects, the on-chip capacitors are nearly ideal. They have extremely low (negligible) temperature and voltage coefficients, and are entirely free of dielectric absorption. The series resistance arises primarily from the lower (P type) plate.

Inductors must still be discrete elements. A few efforts have been made to create on-chip inductors using spiral coil geometries, but the relatively high substrate conductivity and the resistance of the aluminum itself result in Q values around 1 with the usual silicon IC parameters. Circuits using inductors should be redesigned to eliminate them if possible. Interstage coupling transformers, for example, can often be eliminated by using DC coupling and a differential signal arrangement.

Circuit Packaging

The most common conventional package for integrated circuits is the DIP. Unfortunately, this package has serious limitations for RF integrated circuits. The most favorable packages commonly available today are the SOIC (Small Outline IC) and LCC (Leadless Chip Carrier). The latter are commonly called surface-mount packages since their leads do not extend through holes in the PC card. This also means that their PC card traces are short and their parasitic lead capacitances can be small. Several types of flatpack are also available, and these also have small parasitics.

It is desirable to make a package choice as early in an RF design project as possible, so that the parasitic elements can be modeled as part of the SPICE simulation of the circuit. Table 1 gives typical element values for the package parasitics. If more accurate values are needed, the parasitic element values should be measured directly on sample packages.

The values quoted for DIPs vary over a wide range. This reflects the fact that DIPs come in many sizes, and the longer pins (those farther from the die cavity) have several times more capacitance and inductance than the shorter ones. It can be seen from this data that large DIPs can have LC resonances at a few hundred MHz. The SOIC and LCC packages have about the same parasitic element values, and the package resonances are typically above 1 GHz.

An even more favorable packaging scheme is the use of chip-and-wire hybrids. This kind of packaging often reduces parasitic elements to negligible values. Another type of hybrid scheme uses TAB (Tape Automated Bonding) assembly methods.

Noise

Random noise is an important parameter in RF design. The primary sources are Nyquist noise and shot noise, and these are the same on an AMC as in a conventional discrete design. The base resistance of the input transistor is often the primary factor in noise level. Table 2 gives the room-temperature nominal base resistance of several transistors used in the VTC900 series AMCs. Figure 1 shows the VTC VJ960, a typical example of a UHF AMC.

The minimum-sized transistors have higher base resistances and more noise. The TLN1 geometry was especially designed for low noise, and is rated for up to 10 mA collector currents.

The achievable noise is illustrated by the following example. A common-emitter amplifier was simulated using the TLP1 transistor biased with I_c = 4.5 mA and a 260 ohm load resistor. The source resistance was 50 ohms, and it was coupled

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in through 1000 pF. This resulted in a voltage gain of 37 V/V (31 dB) and a 3 dB bandwidth of 75 MHz. The SPICE-simulated input noise was 1.18 nV/$\sqrt{}$Hz at 1 MHz and 1.03 nV/$\sqrt{}$Hz at 10 MHz (dominated by noise arising from the 50 ohm input resistance). With a 5 ohm source resistance the input noise was .77 nV/$\sqrt{}$Hz at 1 MHz and .52 nV/$\sqrt{}$Hz at 10 MHz. Wider bandwidths can be achieved by reducing the gain, using cascoding, etc., at some cost in noise.

Usually 1/f noise is not a problem with the VJ900 series. The noise corner is typically only a few 10s of a Hz.

Mixers

Mixers suitable for RF use can be readily integrated. The basic circuit is the 6-transistor balanced modulator (1) (Figure 2a), or a more elaborate version of it called an analog multiplier (2). This is shown in Figure 2b. It has predistortion circuitry to increase the range of signal amplitudes for which it remains linear. These are differential DC-coupled circuits, and are effective mixers from DC to 100 MHz and more when implemented on a UHF AMC. Signal amplitudes need not be large, hence power can be saved.

Filters

Thus far there are no widely-used methods for implementing video-frequency filters on-chip. However, several papers have shown how active filters can be integrated for television receiver filters in a full-custom design (3, 4). For most low-volume applications, off-chip filters are still the most cost-effective approach.

Phase-Locked Loops

Phase-locked loops are a very important class of frequency-selective circuits, and they lend themselves readily to monolithic integration (5, 6). Voltage-controlled oscillators of the RC type for frequencies above 10 MHz can often be integrated entirely on chip using either junction or oxide capacitors (7). When a varactor-tuned LC VCO or a crystal oscillator is needed, the tuning elements must be placed off-chip.

The phase detector is commonly an exclusive-OR logic circuit or a sequential logic circuit using flipflops (8), both of which are readily integrated as CML or ECL circuits. Operation to frequencies beyond 100 MHz is possible with careful design.

The loop filter can also be integrated on-chip. The most popular filter is the charge-pump type (9, 10). Control of the center frequency may present problems for some narrow-band PLLs.

Circuit Design Using Simulation Tools

RF circuits have traditionally been developed using a breadboard which contains the same circuit elements to be used in the production version. This works well for discrete-element designs, but breadboards are of limited use with AMCs because the breadboard cannot imitate the <1 pF parasitic capacitances typical of AMCs. Integrated design is done using an analog circuit simulator. A large variety of these are available today (11), nearly all of them elaborations of the SPICE simulator.

An Example

The circuit shown in Figure 3 illustrates many of the ideas set forth above. Here the input is a continuous-wave RF signal with significant amounts of low-frequency interference superimposed on it. In keeping with IC design principles, this signal is converted to a differential signal on-chip. A common solution would be to put the signal into a transformer with a center-tapped output to achieve single-ended to differential conversion. Figure 3 shows how the transformer can be eliminated by suitable circuit design and the use of feedback.

The basic amplifier is the emitter-coupled pair Q1, Q2. The inverting and noninverting outputs are connected to the peak rectifiers Q3 and Q4 with current source leads Q16, Q17 and filter capacitors C1, C2. The rectified and filtered peak values of the output voltages are then fed to the PNP amplifier Q5, Q6, and the proper polarity output is fed back through the emitter follower Q14 to the base of Q2 so that the feedback is negative. This feedback loop forces the base voltage of Q2 to a value which causes the positive peaks of both the OUT + and OUT − waveforms to be at the same voltage and thus accomplishes single-ended to differential conversion.

The stability of this feedback loop must be carefully evaluated using SPICE. Capacitors C1, C2, and C4 will usually be off-chip, while C3 is only 1-2 pF and can be an on-chip capacitor. C3 is usually the largest capacitor, and provides the dominant pole of the response. Note that none of these capacitors lie in the signal path. They are all involved in the feedback loop which has a much lower bandwidth than the amplifier signal path. Since C1 and C2 feed a differential stage on chip, pick-up in their leads will be a common-mode signal for Q5 and Q6 if they are laid out symmetrically on the PC card.

Because the off-chip components are not in the main signal path and do not

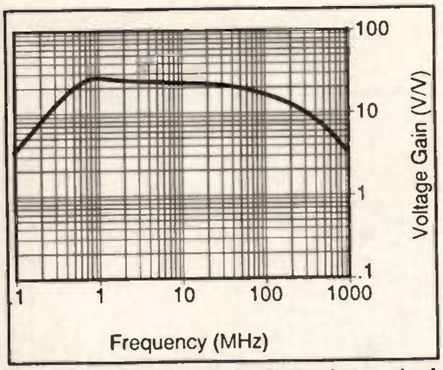

Figure 4. Bode plot of single-ended to diff converter.

have large bandwidths, their power dissipation can be kept low. The circuit shown dissipates only 17 mW.

Transistors Q20-24 provide input bias current and are an example of base current cancellation. Transistors Q18, Q19, and Q20 are identical as are resistors R1A, B, C. Thus the emitter current in Q21 is approximately equal to that of Q1 in the balanced condition of Q1, Q2, so that the base current of Q21 will also equal the average base current of Q1. This base current is then mirrored through the Wilson mirror Q22, Q23, Q24 and fed back to the input. Conventional discrete design would use high-value resistors to provide base bias for Q1, whereas only one low-value resistor (RIA) and five transistors were used.

This input bias circuit relies heavily on the tight matching of integrated components on a single chip, and would not be feasible with discrete components. An advantage of this method is that it provides just the right amount of base bias current for Q1 over any condition of beta, temperature, and supply voltage. The output impedance of Q22 is several mega ohms, a value which is entirely out of reach using on-chip resistors.

The input arrangement allows the circuit to be used quite flexibly. It can be DC coupled to inputs over a quite broad range of common-mode voltages (about ±3 V). It can also be capacitor-coupled at the input, or it can be operated from a high-impedance source such as a photodiode.

At the right of the drawing is a DC bias generator (Q25-Q27, R5-R7). This circuit uses negative feedback to provide a simple tracking output voltage, V_{be}. The output is clamped at a value (R5 + R6)/R5)V_{be}(Q25), which tracks V_{be} with temperature. Q25 and R7 makeup an inverting common-emitter amplifier which is the error amplifier of the feedback loop. If the output voltage VCS, emitter of Q26, should fall below its DC value, Q25 would begin to turn off, resulting in an increase in the base voltage of Q26, i.e., negative feedback. At room temperature the Q25, R7 combination has a voltage gain of about 315 V/V with ±5 V supplies, and this high gain results in a very low output impedance for VCS. Far more than the seven current sources shown here could be driven without serious loading effects. The high gain also makes VCS-VEE largely independent of supply voltage (good power supply rejection). At high frequencies the feedback phase shift can reach 360 degrees, which will result in oscillations, if many current sources are fed from a single bias generator. To prevent this the junction capacitor Q27 is placed where it is a Miller capacitor for Q25. This rolls off the voltage gain at high frequencies and effectively stabilizes the DC output voltage.

The V_{be} tracking bias generator shown here is simple and effective with excellent regulation and PSRR. It does have a temperature-dependent output voltage and this will result in a voltage gain which falls with increasing temperature. By using more sophisticated bandgap techniques (12) the voltage gain of Q1, Q2 can be made nearly independent of temperature.

The SPICE-simulated frequency response of the amplifier is shown in Figure 4. In this example C1 = C2 = 200 pF (on chip), and C4 = 5000 pF. There is peaking at about 1 MHz which arises from the offset-cancelling feedback loop. Below this frequency the peak rectifiers have excessive ripple and the offset cancellation loop no longer operates properly. By suitable choice of C1, C2, and C4 this frequency can be made arbitrarily low, but at the cost of slow response to changes in the DC level of the input. Input frequencies substantially below the peak at 1 MHz (DC, noise, interference) are strongly rejected, as intended.

Above 4 MHz the frequency response is quite flat, with a 3 dB point at 130 MHz. The single-ended to differential gain is 20 (40 dB) in the flat region. By designing a more sophisticated gain stage (cascading transistors Q1 and Q2 for example) even better bandwidth could be achieved. The input impedance is approximately 6 k ohms.

Conclusion

Analog master chips offer large improvements in power, physical size, cost, and RF interference if suitable design approaches are used. The optimum techniques are often quite different than those for conventional discrete-element design, and require the designer to review the circuit design at the system level. rf

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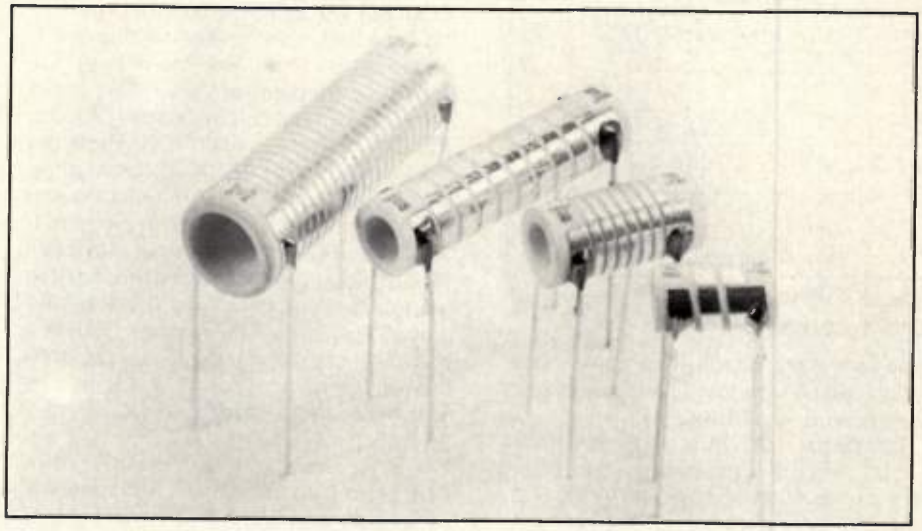
About the Author

John Shier is strategic marketing manager for VTC, Inc., 2401 E. 86th Street, Bloomington, MN 55420. He received a B.S. degree from the California Institute of Technology in 1960, and M.S. and Ph.D. degrees from the University of Illinois in 1962 and 1966, all in physics. John has held IC processing and design positions at Signetics, AMD, Intersil, Sperry Univac, and CDC. His phone number is (612) 851-5228.

High Stability Inductors From Oxley

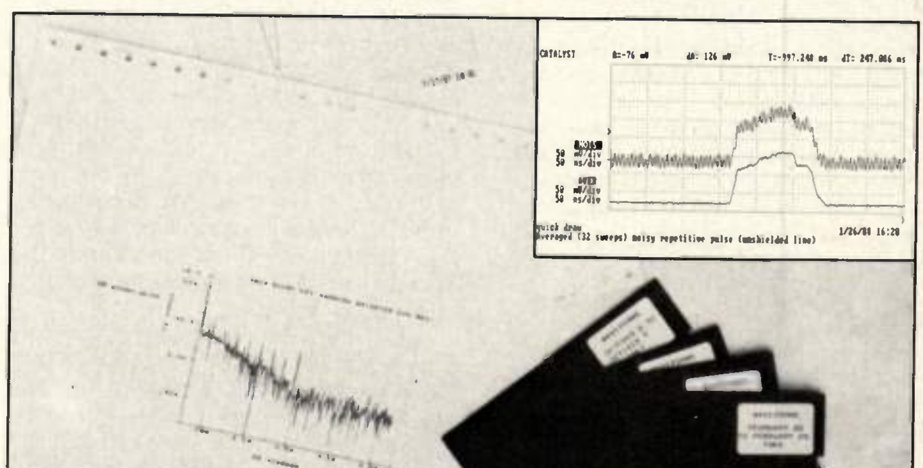


Type QLC 1603 high stability metal-on-ceramic inductors feature construction consisting of high Q silver spiral fused to a low loss alumina former, providing an extremely stable temperature coefficient of ±10 ppm over the −55°C to 125°C temperature range. Heavy deposits of electroplated copper and silver provide a minimum Q of 120 at 130 MHz for a nominal inductance of 0.07 uH. High melting point solder joints (305°C) bond thinned copper wire lead to the inductor, preventing damage during normal soldering operations.

The inductance values range from 0.045 uH to 1.830 uH with maximum direct current of 0.5 A. The QLC 0802 has an inductance of 0.045 uH at 1 MHz with a minimum Q of 100 at 75 MHz. Typical self resonant frequency is greater than 1000 MHz. The inductance value for the QLC 826 at 1 MHz is 0.245 uH with a minimum Q of 150 at 75 MHz and typical self resonant frequency at 400 MHz. An inductance of 1.830 uH at 1 MHz can be obtained with the QLC 1643. The Q for this device is 150 at 25 MHz and typical self resonant frequency is 220.

All inductances mentioned are specified to ±5 percent and the inductances conform to MIL-C-15305. The applications for this family of inductors include VCOs, PLLs, and various other applications where stability is a major concern. **Oxley Inc., Branford, CT. INFO/CARD #222.**

LeCroy Introduces Automated Waveform Digitizers



The Century™ Series of automated single channel waveform digitizers provides numerous benefits when compared to all analog and some digital oscilloscopes. The advantages include higher accuracy measurements, extended memory-length waveform recording, both pre- and post-trigger waveform recording, high resolution transient capture, digital signal processing and analysis, and total programmability.

The Century 2005 has a sample rate of 5 Ms/sec, 2.5 MHz bandwidth, 12-bits of resolution, and 512 K of memory that can be expanded to 8 M. The price is $8950. The specifications on the 2101 include a sample rate of 100 Ms/sec, bandwidth of 100 MHz, 8-bits of resolution, 32 K memory expandable to 2 M and price of $8450. The 2102 has the same bandwidth and resolution with sample rate of 200 Ms/sec, memory of 64 K that can be increased to 2 M, and price at $12,950. The Century 2100 has a sample rate of 1.3 Gs/sec, bandwidth of 400 MHz and resolution of 8 bits, while memory is 10 K expandable by another 10 K. The price on this digitizer is $16,950.

The modular Century Series design allows for channel expansion. Additional digitizing can either operate independently, with different time base and trigger setups, or be synchronized for simultaneous multichannel capture. Digitizer modules of higher sample rate, bandwidth or resolution may be added to any single-channel Century Series system.

The modular design also provides for waveform memory expansion on each channel. Thus, a combination of more memory and an additional higher speed channel can provide both extended waveform recording time and a more detailed second channel record of a waveform section.

Features include differential inputs, segmentable memory with trigger point time stamps, battery backed memory, variable clock input, window triggering and hysteresis triggering. All digitizer operating functions are fully programmable. These functions include gain, sample rate, channel enable, trigger level, trigger slope, trigger delay offset, pre- and post-trigger percentages, and active memory length. **LeCroy Corp., Chestnut Ridge, NY. INFO/CARD #219.**

RF Step Attenuators

Daico Industries has announced a family of seven bit step attenuators with 63.5 dB of range in 0.5 dB steps. These devices operate from 30 MHz to 500 MHz with up to +13 dBm of RF input power over the temperature range from −55°C to +125°C. The DA0617 boasts 7 ns RF transition time, 25 ns total switching

speed including the propagation delay of the integral TTL driver. This PC mountable (DIP) version is available for $630 each (10-24 pieces). The DA0285 and DA0295 are low power alternatives offering CMOS and TTL controls respectively. Each consumes only 30 mA and 50 mA respectively at +5 V. Both costs $511 each when quantities of 10-24 are purchased. **Daico Industries, Inc., Compton, CA. Please circle INFO/CARD #218.**

Linear Phase SAW Filter

FB199.5-6, a bandpass filter based on SAW technology, provides linear phase and flat amplitude response over a 6 MHz bandwidth centered at 199.5 MHz. With 60 dB ultimate rejection, it is used in channelized applications to reduce unwanted

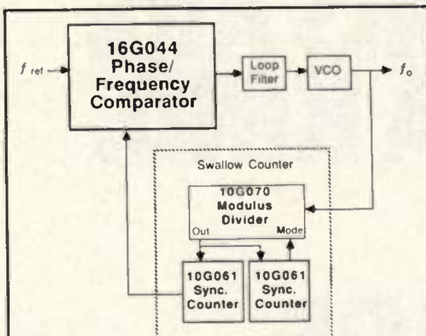
image, and sideband and LO signals. The FB199.5-6 achieves ±0.25 dB amplitude ripple and ±2 degrees phase error. Maximum input power is +25 dBm. These hermetically sealed filters are packaged in a 14 pin DIP measuring 0.8"× 0.5"× 0.2". **Phonon Corp., Simsbury, CT. Please circle INFO/CARD #217.**

DIP VCXO

Connor-Winfield introduces a VCXO packaged in a standard 14- pin DIP. Nominal frequencies between 450 kHz and 20 MHz are available. The maximum average deviation is 80 ppm/V with a maximum of 10 control volts. **Connor-Winfield Corp., West Chicago, IL. INFO/CARD #216.**

Phase/Frequency Comparator

The 16G044 1 GHz phase/frequency comparator handles input signals up to 1 GHz. It improves noise performance for single frequency PLLs and in low divide ratio applications. Other applications include high speed PSK and FSK demodulators, high resolution time delay measurements, and frequency discriminators.



FREQUENCY SYNTHESIZER

The 16G044 is a rising edge triggered device with high gain differential amplifier inputs. When the inputs are unequal in frequency and/or phase, the differential outputs act as pulse streams which when subtracted and integrated, provide an error voltage for control of a VCO. Inputs are AC-coupled and ECL/Picologic compatible. The 1 GHz device costs $90.30 in quantities of 100. **GigaBit Logic, Newbury Park, CA. INFO/CARD #215.**

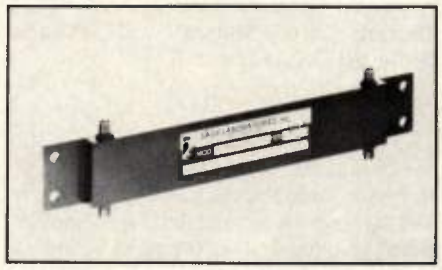
Enhancements to Spectrum Analyzers

Tektronix announces that the 492B, 492BP, 2755A and 2577AP spectrum analyzers now offer expanded capabilities. The enhancements include 10^{-6} frequency accuracy, external reference lock to provide the flexibility of tying in with a system clock, macro capability for automated measurements without a controller, maximum resolution bandwidth of 3 MHz, and multiband sweep of 1.7 GHz to 21

GHz. **Tektronix, Inc., Beaverton, OR. INFO/CARD #214.**

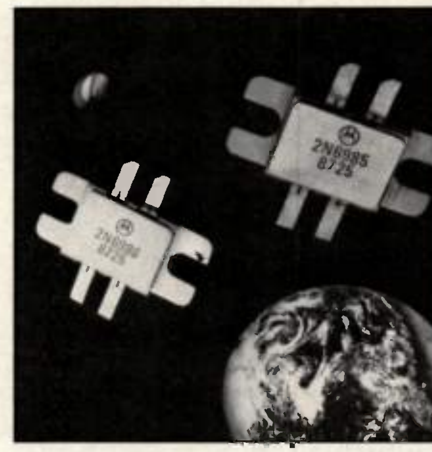
Quadrature Hybrid

The Model FH4206 hybrid operates from 475 to 4200 MHz and features less than 1.2 dB unbalance, greater than 21 dB return loss, greater than 23 dB isolation, and net insertion loss from 0.2 dB at 475 MHz and 0.8 dB at 4200 MHz. **Sage Laboratories, Inc., Natick, MA. Please circle INFO/CARD #213.**



UHF Power RF Transistors

The 2N6985 and 2N6986 28-volt UHF RF power transistors are intended for operation in the 225 to 400 MHz (2N6985) or 225 to 500 MHz (2N6986) frequency band and offers 125 or 100 watts of output power in a push-pull hermetic package. In single quantities, the 2N6985 is $156.90 and the 2N6986 costs $162.80 **Motorola, Inc., Phoenix, AZ. Please circle INFO/CARD #212.**



Frequency Counter/Wattmeter

The Model 7510 frequency counter/wattmeter provides frequency measurements as well as forward and reflected power readings. The instrument is supplied with a remote dual socket line section, and adjustable X-tractor RF sampling element and two cable assemblies to connect the line section to the instrument. **Coaxial Dynamics, Inc., Cleveland, OH. INFO/CARD #210.**

SMT Digital Attenuator

KDI introduces a four-bit surface-mount digital attenuator with step size of 1 dB. It has a VSWR of 1.35:1, insertion loss of 2.5 dB max., and 0.5 dB cumulative accuracy. The switching speed is 100 ns measured from 50 percent of logic input

to 90 percent of RF output. **KDI Electronics, Whippany, NJ. Please circle INFO/CARD #211.**

SMT Micro Inductors

Designed for surface mounting, the inductors are capable of withstanding shock and vibration. The frequency range is 500

kHz to 100 MHz with inductance values from 0.1 mH to 100 mH. **Interconnection Products Inc., Santa Ana, CA. Please circle INFO/CARD #209.**

Dual Crystal Filter

K&L Microwave unveils the Model 001-54400 dual crystal filter operating at 39.875 MHz. Used in a receiver system, this combined filter eliminates the need for two separate components. Channel A is designed to customer's specifications (bandpass) while Channel B replaces a required SAW filter. **K & L Microwave, Inc., Salisbury, MD. INFO/CARD #208.**

MMIC Down Converters

CEL introduces a family of UHF/VHF MMIC down converters for CATV, TV tuners, cellular radios and UHF/VHF communications equipment made by NEC. The devices feature wide-band operation, and require a 5 V power supply. **California Eastern Laboratories, Inc., Santa Clara, CA. INFO/CARD #207.**

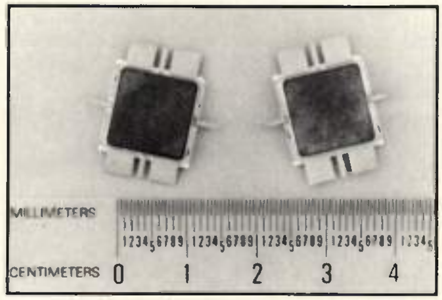
HF Notch Filter

Model 6002 narrow band notch filter can be obtained for any frequency from

2 to 40 MHz. Typical performance of a 30 MHz notch filter is 60 dB notch attenuation with a 3 dB bandwidth of 2.4 MHz and a 50 dB bandwidth of 150 kHz. Impedance is 50 ohms with SMA connectors. Price of single units is $950. **Microwave Filter Company, East Syracuse, NY. INFO/CARD #206.**

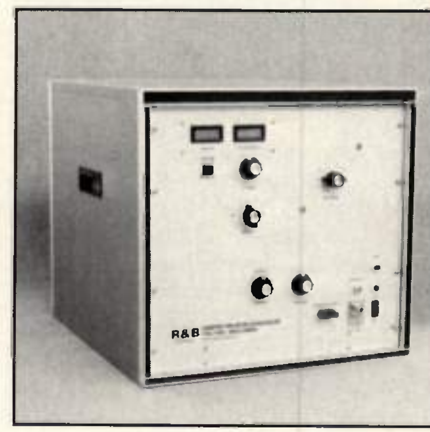
MMIC Amplifier Module

The TCWL-0103 contains two cascaded TCWL-0100 series 4 GHz bandwidth MMIC amplifiers. The module provides 28 dB typical gain over the 50 MHz to 4 GHz range and noise figure is typically 5 dB at 3 GHz. Input and output VSWRs are less than 2:1 from 100 MHz to 4 GHz. In single units the TCWL-0103 is $165. **Tachonics Corp., Plainboro, NJ. Please circle INFO/CARD #205.**



EMP Signal Generator

R & B Enterprises introduces the PGV-7003 EMP signal generator designed to produce waveforms for EMP testing as specified by MIL-STD-461C. It incorporates testing circuitry to enable tuning of signal frequencies between 3 MHz and 20 MHz. It also provides signal for direct lead and pin injection testing, and also capacitive and inductive coupling for bulk cable testing. Selection of trigger mode enables manual triggering of single events and oscillator triggering for various repetition rates. The PGV-7003 is priced at $16,000. **R & B Enterprises, West Conshohocken, PA. INFO/CARD #204.**



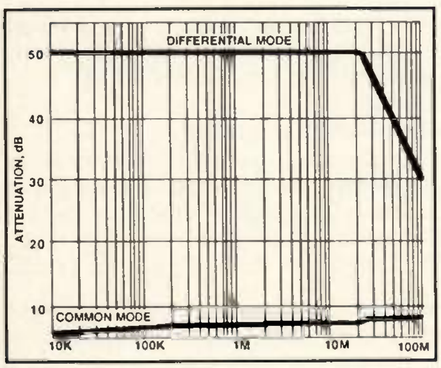
Laboratory Amplifiers

The P/N W50ATC-PS has a frequency range of 10 kHz to 50 MHz with gain of 50 dB and flatness of ±0.5 dB. The noise figure is typically 1.3 dB and power out is +5 dBm. This amplifier is suited for laboratory uses including increasing the

dynamic range of spectrum analyzers, boosting output signal of wideband sweepers to +30 dBm, and for producing low noise, wide frequency amplification for input into high level mixers. **Trontech, Inc., Neptune, NJ. INFO/CARD #203.**

Differential Mode Rejection Network

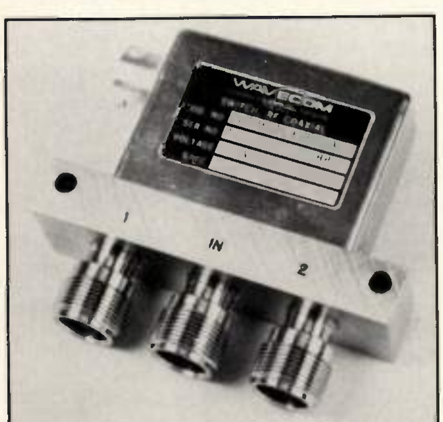
EMC Services announces the availability of its LISN MATE differential mode rejection network. It enables the user to separate differential mode and common mode noise while performing LISN-based



conducted emissions measurement. It attenuates the differential mode by 50 dB minimum and the common mode by less than 4 dB. It is priced at $590. **EMC Services, Inc., Seminole, FL. Please circle INFO/CARD #202.**

High Power SP2T Switch

Wavecom introduces a SP2T coaxial switch with insertion loss of 0.05 dB, and operating range of DC to 3 GHz. It is guaranteed for 500,000 cycles per switch



position and handles power levels up to 4.4 kW CW. VSWR is 1.15:1 (typ.) and isolation is 90 dB typical. The Series 027 is available with either latching or failsafe actuation. **Wavecom Division of Loral Corp., Northridge, CA. Please circle INFO/CARD #200.**

Single Chip Spectrum Analyzer

Analysis functions performed by the MISA09 include windowing, FFT calculations, magnitude conversion and logarithmic compression. The component processes 128 complex data points. A 16-bit bi-directional bus provides simple I/O operations. The real time performance for the device is sufficient for all audio frequency applications and the 128 point complex transform is executed within 2.5 ms. Output dynamic range is 14 bits while input data may be processed by either a rectangular or Hanning window. **Monolithic Instruments, Inc., Florissant, MO. INFO/CARD #201.**

Quartz Crystals

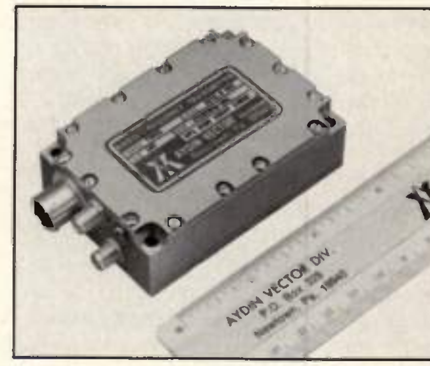
The SMX line of ceramic surface mount AT quartz crystals is being introduced by Standard Crystal. The frequencies available range from 5 MHz to 160 MHz with

INFO/CARD 43

typical drive level of 150 microwatts and price ranging from $4.80 to $9.00. The crystal measures 0.070"× 0.400"× 0.250". **Standard Crystal Corp., El Monte, CA. INFO/CARD #199.**

UHF Telemetry Transmitters

The Aydin Vector T-800S/L Series UHF telemetry transmitters are 2 W and 5 W units that operate between 1453 and 2500 MHz with frequency agility stability of ±0.002 percent. To maximize RF isolation, the internal modules (modulator, series regulator, power amplifier, multipliers and



filters) are housed in separate machined enclosures. **Aydin Vector Division, Newtown, PA. INFO/CARD #198.**

Iron Powder Toroids

Based on Core-Tronics P7 iron powder, these cores can be used in applications where other magnetic materials require an air gap to prevent saturation. Initial permeability of the P7 material is 75 and

positive temperature coefficient is 830 ppm. All cores are epoxy coated for insulation and identification and come in standard sizes ranging from 0.255" to 1.840". Guaranteed A_L values range from 235 to 1640. **Core-Tronics, Orange, NJ. INFO/CARD #197.**

Multiple Crystal Oscillators

The Series PXSM contains up to 16 independently selectable crystal oscillators at frequencies to 200 MHz. Each oscillator

is externally selectable by a BCD code. The selected crystal oscillator is automatically phase locked to an external reference standard at frequencies to 20 MHz. Each crystal oscillator can be at any frequency to 200 MHz and be phase locked to any external reference to 20 MHz. Spurious output is better than 90 dBc. **Communication Techniques, Inc., Whippany, NJ. INFO/CARD #196.**

Low Noise GaAs MMIC Amplifier

The HMM-10620 GaAs MMIC circuit is designed for use in gain stage applications where low DC power consumption is a major requirement. It has an operating frequency range of 2 GHz to 6 GHz, small signal gain of 11.5 dB (typ.), gain flatness of ±0.5 dB and 1 dB gain compression output power of +13 dBm. VSWR is 1.75:1 and noise figure is 5.5 dB. **Harris Microwave Semiconductor, Milpitas, CA. INFO/CARD #194.**

Silicon Chip Capacitors

The MA4M series of silicon chip capacitors have values from 1 pF to 600 pF and are offered in .018" × .050" × .050" outlines. The capacitors have a temperature stability that is at least five times better than that of ceramic chips. **M/A-COM Semiconductor Products, Burlington, MA. INFO/CARD #192.**

Programmable Signal Simulator

S.T. Research unveils the RA-100 AM signal simulator that uses a 16-bit processor and operates at 100 MHz. Simulation of jitter ranges from 0.01 to 99.99 percent in 0.01 percent increments. It has a 20 ns minimum pulse width and 16 position pulse stagger/group per channel. Non-volatile memory is standard. **S.T. Research Corp., Newington, VA. Please circle INFO/CARD #175.**

Signal Generator Combines Generator Combines Frequency Agility and Spectral Purity

The HP 8645A is a 2060 MHz agile signal generator that combines 15 us switching speed with −128 dBc phase noise and high rate, high deviation modulation.

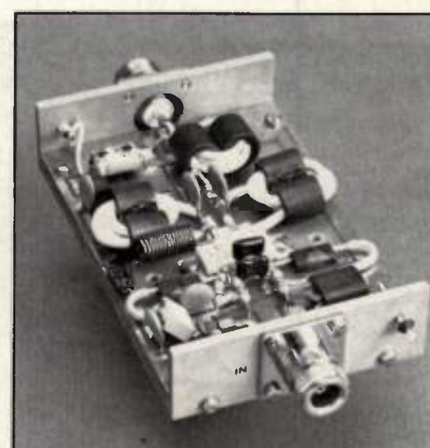
For in-channel measurements, the instrument offers simultaneous FM, AM and pulse modulation along with less than 2 Hz of residual FM. It can be internally modulated with DC to 400 kHz rates or externally modulated with 20 MHz of deviation and up to 10 MHz rates. In fast hop, maximum deviation is 4 MHz with 10 MHz rate. AM is available with up to 100 kHz rates and 99 percent depth. Pulse modulation allows a 35 dB On/Off ratio with 200 ns rise/fall times. The base model of the HP 8645A is $32,000. **Hewlett-Packard Company, Palo Alto, CA. Please circle INFO/CARD #174.**

Analog Transmitter/ Receiver Modules

General Optronics announces the availability of laser-based wide bandwidth analog transmitter/receiver modules for fiber optic communication links. The modules feature an analog single mode 1300 nm laser transmitter and PIN diode optical receiver which accommodates bandwidths up to 1.5 GHz. The transmitter incorporates circuitry which monitors and controls average optical output power and the operating temperature of the company's 1300 nm single mode laser. The receiver is constructed with an InGaAs PIN diode for optimum stability and low dark current. Link bandwidths up to 1.5 GHz are available with loss budgets of up to 20 dB. The optical dynamic range it typically better than 20 dB. Price for the single mode, 1300 nm link is $8,900. **General Optronics Corp., Edison, NJ. INFO/CARD #173.**

MOSFET Power Modules

Kalmus introduces two linear power modules designated the 162F, 174, and 460F. Model 162F covers from 10 to 200 MHz at 100 W, Model 174 is rated at 200



W from 15 to 150 MHz, and Model 460F goes from 200 to 400 MHz at 75 W. Gain ranges from 10 to 20 dB and price ranges from $795 to $995 depending on model. **Kalmus Engineering International Ltd., Woodinville, WA. INFO/CARD #172.**

Single Chip Radio Receiver

Plessey introduces a radio receiver on a single chip which consumes less than 4 mW. Designated the SL6638, it utilizes direct conversion techniques to receive digital information at rates to 1200 bits per second in the form of shift keyed transmissions. It is designed for operation up to 200 MHz and provides a typical input sensitivity of 200 nV. In operation, the chip splits the incoming signal into two parts which are frequency converted to base band. If the waveform at the limiter in one signal path leads the waveform on the other path by 90 degrees, decoder output will be low; while if it lags by 90 degrees, decoder output will be high, thereby reproducing the digital data stream. The price is $10.31 each in 1000 piece quantities for j-lead and $17.70 for LCC. **Plessey Semiconductors, Irvine, CA. INFO/CARD #171.**

SMT MMIC VCOs

Pacific Monolithics unveils the PM-OVO6XX-B varactor tuned MMIC oscillators that cover from 1 to 6 GHz with a minimum 20 percent tuning bandwidth. They feature low frequency load pulling, output power of +13 dBm (min.), and fast tuning and settling times. The oscillator is available in a 10-lead surface mount package that measures 0.270" square and 0.065" thick. **Pacific Monolithics, Sunnyvale, CA. INFO/CARD #170.**

RF Amplifiers

The PF849 family of RF amplifiers feature quadrature coupling of the GaAs FET devices to provide unconditional stability and high reliability. These amplifiers are available in 30 MHz bandwidths from 800 to 1000 MHz with gain of 16 dB and noise figure less than 1.0 dB. Other features include a third order intercept point greater than +20 dBm. **Janel Laboratories, Inc., Corvallis, OR. INFO/CARD #169.**

Broadband RF Amplifier

The Model 1000L has a frequency range of 10 kHz to 220 MHz with linear output at less than 1 dB gain compression of 750 W (min). A low range output (100 W CW nominal) is provided as a convenience for low power experiments without the need for moving to a smaller unit. **Amplifier Research, Souderton, PA. Please circle INFO/CARD #168.**

Low Noise Preamplifier

WI-COMM introduces a preamplifier covering the 2 MHz to 1000 MHz range with gain of 13 dB, noise figure of 3 dB, and 1 dB gain compression level of 8 dBm. Third order intercept point is 20 dBm (min) and VSWR is 2:1. **WI-COMM Electronics, Inc. Massena, NY. Please circle INFO/CARD #167.**

S-Parameter/Noise Parameter Library

This software library provides access to S and noise parameters for 35 selected NEC small signal bipolar transistors, low noise GaAs FETs and power GaAs FETs. This data can be used with Touchstone or Super-Compact. **California Eastern Laboratories, Inc., Santa Clara, CA. INFO/CARD #179.**

Analysis, Optimization, and Synthesis Software

The TAME (Top Algorithms for Microwave Engineering) software package analyzes, optimizes, and synthesizes microwave and RF networks in the frequency domain. The circuit library for analysis and optimization includes RLC elements, transmission lines, controlled sources, gyrators, directional couplers, and one-, two-, and three-ports described by tables of network parameters. Graphics output includes a choice of magnitude and phase plots of any S-parameter, as well as Smith Chart plots. In the synthesis mode, one can design lumped, distributed, or even mixed lumped-distributed matching networks between complex sources and loads. The terminations may be described by tables of S, Y, or Z parameters or modeled by circuit elements. TAME runs on IBM PCs and compatibles. The is priced at $49.95 plus $5.00 shipping. **Spefco Software, Stony Brook, New York, NY. INFO/CARD #178.**

1-518

Filter Synthesis and Design Program

S/FILSYN, the general purpose filter synthesis and design program has been updated. For passive/microwave designs, the program can now design bandpass filters with sloping passbands for the design of interstage and impedance matching purposes, and a shifted-bandpass for arithmetic symmetry and/or linear phase. Also, active RC and switched capacitor realizations are now available even for filters with complex transmission zeros, providing optimal designs in these forms for both loss and delay requirements. For switched capacitor filters, one can specify the type of section to be used out of the four standard sections available. The IIR digital filter segment has added capabilities, including the availability of an ASCII file containing filter design data. **DGS Associates, Inc., Santa Clara, CA. INFO/CARD #177.**

Active Filter Design Package

The Active Filter Design program designs Butterworth, elliptic, Chebyshev, and Bessel low pass, high pass, bandpass, and bandstop active filters. It allows the direct entry of pole and zero locations or transfer functions and can convert low pass prototype poles and zeros to the desired filter configuration. A menu driven format is featured and the program supports manual or automatic pole/zero pairing as well as uneven gain distributions. Active implementation of type MFB, VCVS, biquad, state variable and Reticon Switched capacitor filters are available using components of a specified tolerance. Transfer functions and poles and zeros can be regenerated based on the calculated or modified component values. The Active Filter program is $525, while the Spice file conversion utility is $125. The programs can be purchased together for $625. **RLM Research, Boulder, CO. INFO/CARD #221.**

RF Connector Poster

A spectrum chart depicts typical applications for connectors ranging from Twinax for DC to various RF and microwave connectors to 50 GHz. It shows the appropriate AMP products for each application. Each product family shown above the spectrum chart at its maximum operating frequency is further defined in a table. The information shown includes product line, nominal impedance, maximum frequency, temperature rating, cable retention, coupling mechanism, connector family, product specification, durability, and connector body finish. The photographs in the poster show typical connectors including a wide selection of printed circuit board connectors. **AMP, Inc., Harrisburg, PA. INFO/CARD #223.**

Application Note on Building PGAs

This application note, "CMOS DACs and Op Amps Combine to Build Programmable Gain Amplifiers, Part II," investigates the performance of dual CMOS DACs as gain determining elements in a PGA (programmable gain amplifier) system. The note focusses on how greater accuracy over a wider dynamic range can be achieved with a dual DAC circuit as opposed to a single DAC solution. Some of the topics discussed include "The Basic Equations for a Dual-DAC PGA," "Comparing the Errors," "Dynamic Problems/ Stability and Compensation," "Small Signal Bandwidth," "Dynamic Gain Errors," "Noise and Distortion," and "Test Results." **Analog Devices, Norwood, MA. INFO/CARD #190.**

SAW Delay Lines Brochure

This brochure from Sawtek describes non-dispersive delay lines, tapped delay lines, SAW dispersive delay lines, linear FM, non-linear FM, modified non-linear FM, and dispersive delay line implementation. The final section focuses on definitions and parameter clarifications. **Sawtek, Inc., Orlando, FL. Please circle INFO/CARD #189.**

VHF and UHF Semiconductor Catalog

Matcom introduces a catalog on Toshiba's VHF and UHF semiconductors. The products listed includes bipolar silicon transistors, various silicon diodes, MOS transistors, and GaAs MESFETs. Some of the products listed are available in surface mount configurations. **Matcom, Inc., Palo Alto, CA. INFO/CARD #188.**

Note Describes Chip Capacitors

Thin Film Chip Capacitor With Very High-Q For UHF/VHF describes the development of thin-film capacitors for use in the VHF/UHF frequency range. The topics discussed include Market Need; Available Solutions; Thin-Film Technology; Accu-F Principle; Characteristics and Performance of the Accu-F; and Future Developments. **AVX Ltd., Hampshire, England. INFO/CARD #187.**

1988 Publication Catalog

This catalog lists the books, conference publications and journals published by the institution of Electrical Engineers and Peter Peregrinus Ltd. Titles are grouped by subject area and may be located by reference to either the author index or title index. The subjects listed include telecommunication; electromagnetic waves; radar, sonar, navigation and avionics; computing; materials and devices; electrical measurement; energy; control; power; management of technology; history of technology; and bibliographies. **IEE/PPL, Piscataway, NJ. Please circle INFO/CARD #184.**

Custom Hybrids Brochure

The RF & Microwave Subsystems division of M/A-COM introduces a brochure on custom hybrids. It outlines their capabilities with respect to design, manufacturing and quality assurance of products such as high power voltage regulators, IF attenuators, and video detectors. **M/A-COM RF & Microwave Subsystems, Burlington, MA. Please circle INFO/CARD #186.**

Application Notes on Digital Oscilloscopes

A series of application notes for the LeCroy Model 9400 digital oscilloscope is available. The notes describe how to measure RMS using the 9400; linking the 9400 with an IBM-PC/AT via an RS-232-C asynchronous interface; and the 9400 in ultrasonics. The instrument is a dual channel digital oscilloscope with 8-bit ADCs capable of sampling transient waveforms at 100 megasamples/sec and repetitive signals at 5 Gs/sec. **LeCroy Corp., Chestnut Ridge, NY. Please circle INFO/CARD #185.**

RF Amplifiers and Power Dividers Catalog

This short form catalog lists specifications for power dividers/combiners and RF amplifiers. The dividers/combiners listed are 2-, 3-, 4-, 5-, 6-, 8-, 10-, 12-, 16-, and 18-way. The amplifiers listed cover the spectrum from 1 MHz to 2 GHz. A price sheet is also included. **Janel Laboratories, Inc., Corvallis, OR. Please circle INFO/CARD #183.**

1988 Membership Directory

The *1988 IEEE Membership Directory* provides name, current location and title of over 240,000 IEEE members and affiliates in over 137 countries (student members are not included). Listings include telephone numbers when authorized. In addition to the roster, the directory contains a listing of over 3900 IEEE fellows, including their citations and awards; winners of IEEE awards for outstanding achievements in science and technology; and IEEE past presidents and directors. Also included is a section on the purpose, organization and history of the IEEE and requirements for the attainment of various membership grades. It is priced at $126 for non-members and $50 for members. **IEEE, New York, NY. Please circle INFO/CARD #181.**

Capabilities Brochure

This brochure focuses on digital RF memories, threat simulators, and EW integration. The digital RF memories section emphasizes new capabilities for RF countermeasures while the threat simulators section gives an overview of the capabilities of the products Rockwell DEL manufactures. The final section is about the company's experience to solve integration problems. **Rockwell DEL, Inc., Huntington Beach, CA. Please circle INFO/CARD #180.**

Test, Repair and Assembling Products Catalog

Featured in this catalog are products for testing, repairing and assembling electronic equipment. The products are described with specifications, photos and pricing. **Contact East, No. Andover, MA. INFO/CARD #182.**

Advertiser Index

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How to make high-speed measurements more accurately.

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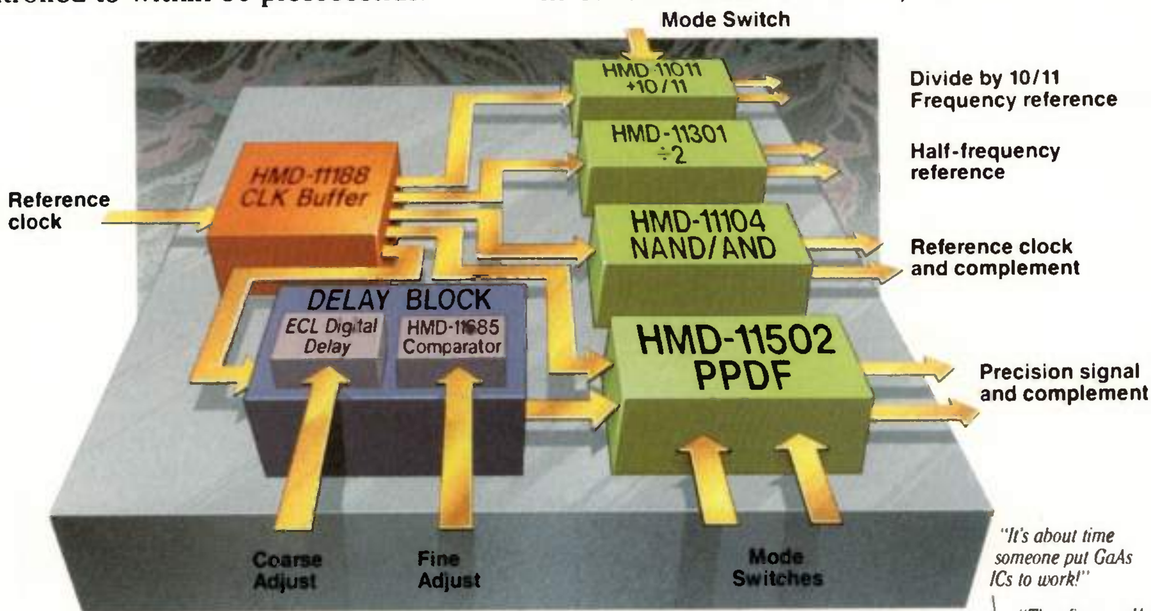
For more information, write or call us: Harris Microwave Semiconductor, 1530 McCarthy Blvd., Milpitas, CA 95035. In U.S., phone 1-800-4-HARRIS, Ext. 1510, or (408) 433-2222, Ext. 202 (TWX: 910-338-2247). In Canada: 1-800-344-2444, Ext. 1510.



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