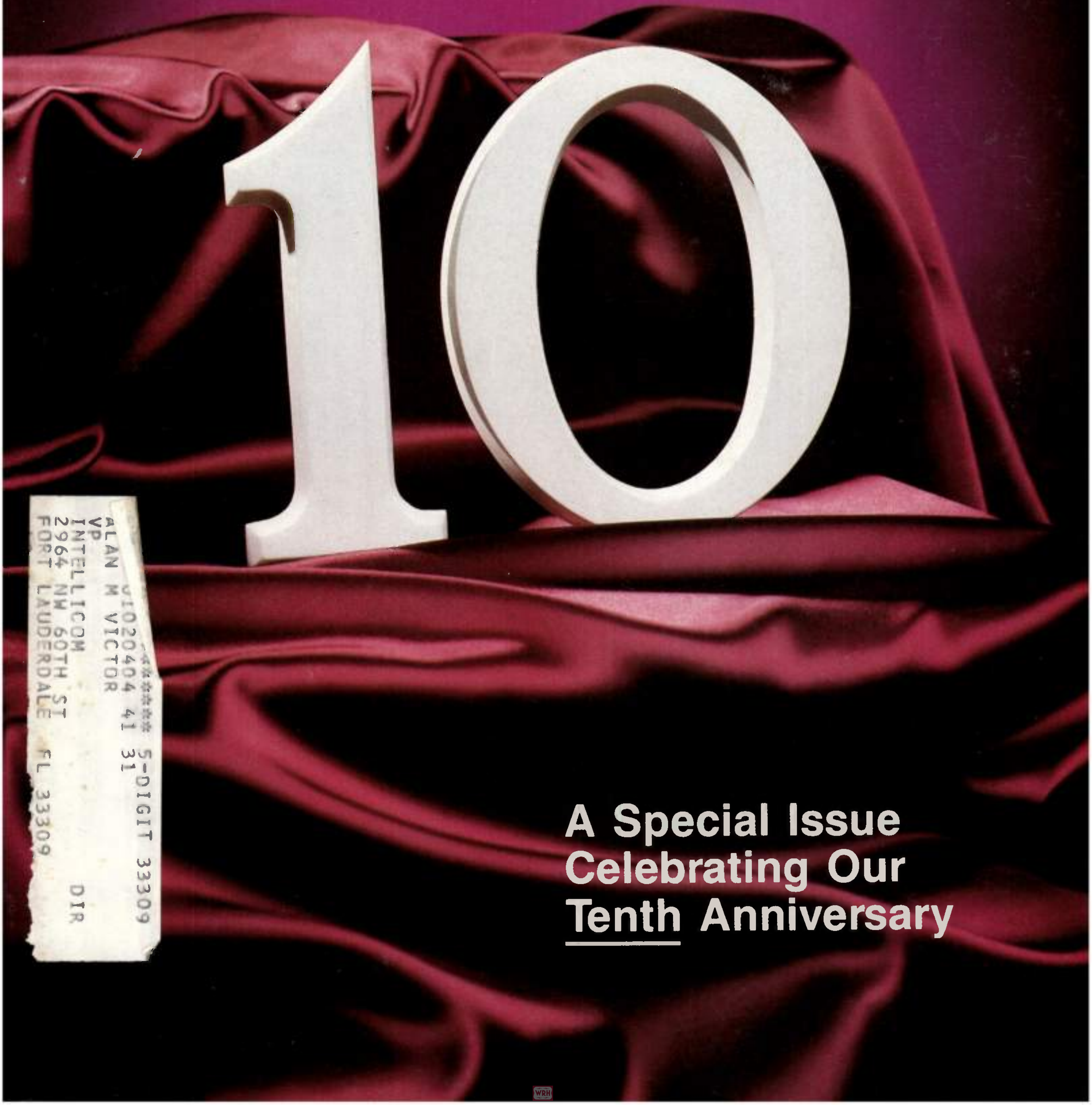
rf design

engineering principles and practices

October 1988

1978–1988 10th ANNIVERSARY



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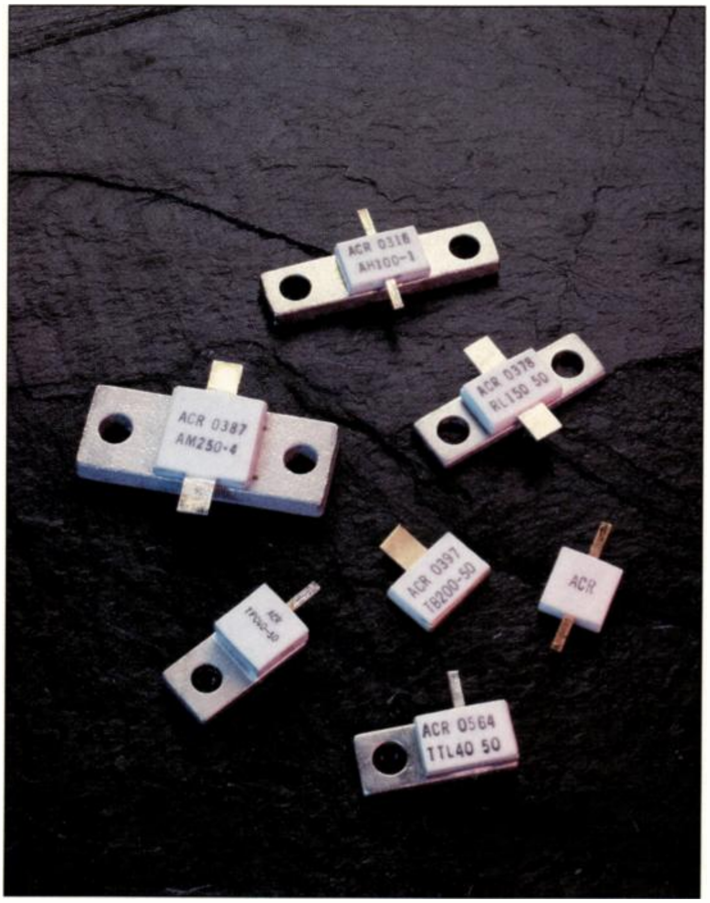

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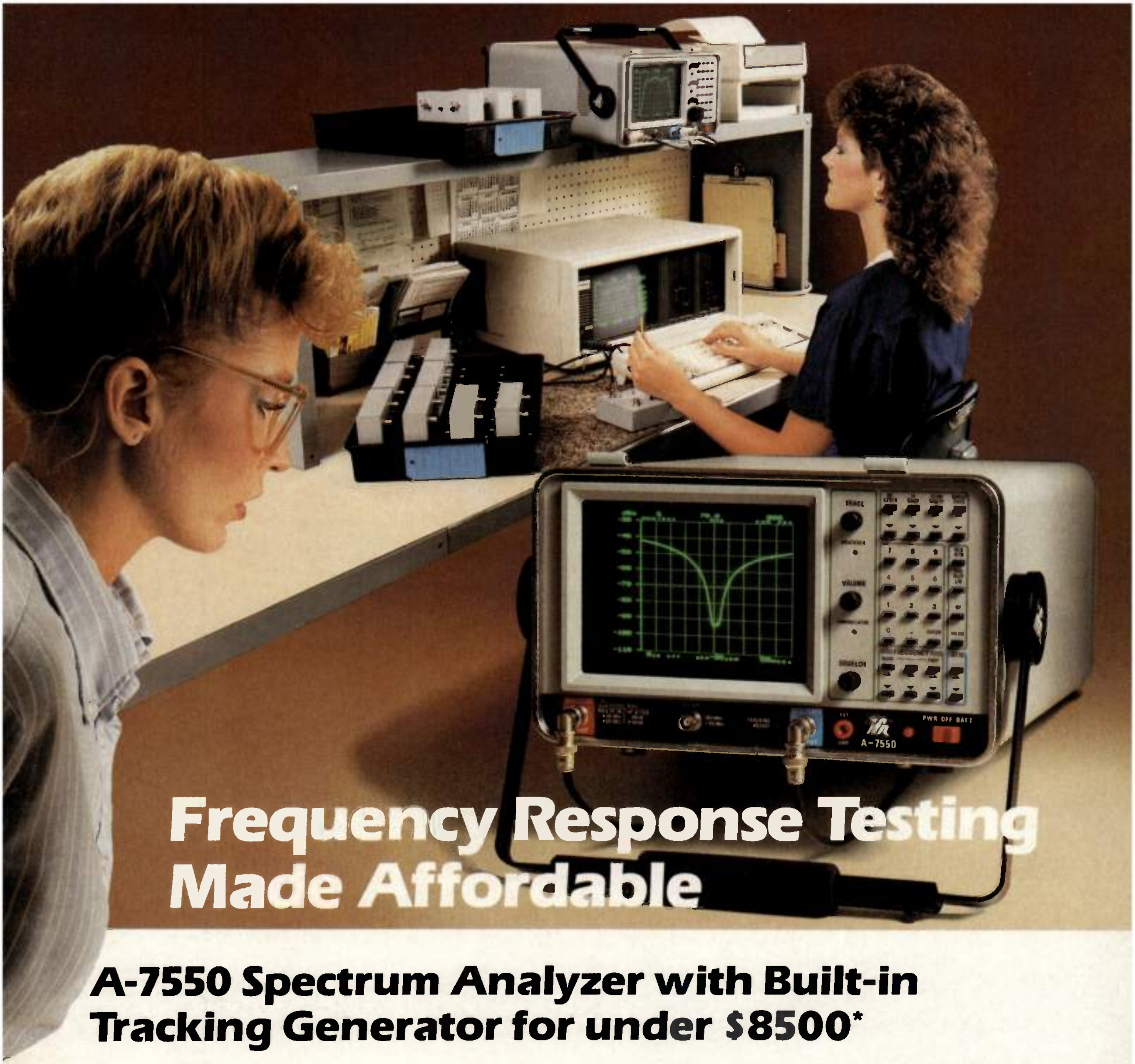

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rf design

October 1988

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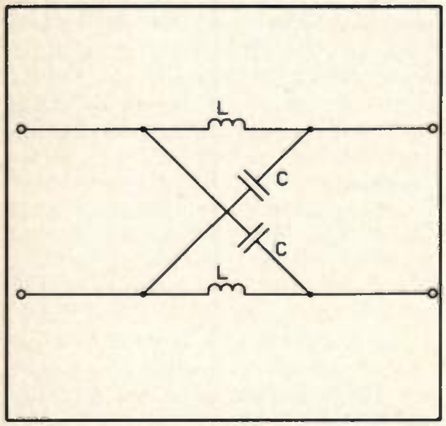
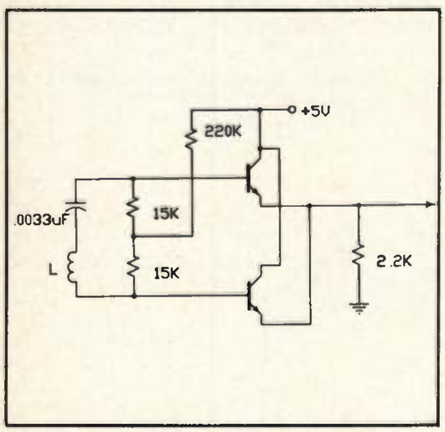


Page 38 — Phase Shifter



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Cover Story

Features

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rf editorial

RF Design — Ten Years is Just the Beginning

By Gary A. Breed
Editor

Where were you 10 years ago?

Jimmy Carter was president, and the U.S. was struggling through a post-Vietnam war recession. We still hadn't recovered the confidence lost a few years earlier during Watergate and Richard Nixon's resignation. The country was in a transition between the political activism of the '60s and the complacency of the '80s.

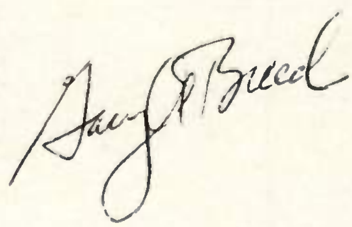
Electronics technology was in a transition, too — or so some folks thought. The hottest word in our vocabulary was *microprocessor*, and the big electronics magazines had abandoned RF engineering to focus on the "digital revolution." To fill the void, *RF Design* was created. For 10 years, this magazine has survived the ups and downs of business (both our business and yours) to become a successful, established engineering journal. The success we have achieved is entirely a reflection of the success of the RF industry.

If there is any one thing that exemplifies the RF industry in the last 10 years, it is the demand for greater performance. No segment of technology can single-handedly meet that demand. The vision of an entirely digital world has faded, and analog techniques have taken their place in the spotlight alongside their digital counterparts. (They even share space on the same monolithic ICs.)

This issue celebrates our 10th anniversary with a look back at the developments of the past decade. But we are also celebrating by doing what has brought us this far — providing useful information for engineers. This month, the exchange of ideas among engineers covers topics that are at the heart of RF: a voltage-controlled phase shifter and a review of Lange microstrip line couplers (both contest prize winners), a description of an unusual oscillator circuit that grew out of an unrelated application, and a look at filters that dissipate stopband power instead of reflecting it back to the load. There also is a complete listing of the technical program at the upcoming RF Expo East.

Ten years is both a long time and not very long at all. It depends on your perspective. It is pleasant to look back and see how far we have come since that first issue of *RF Design* was produced. But, it is far more exciting to view those ten years as just a small segment in the life of an idea that has only begun to address the needs of RF engineers.

Here's to the past 10 years, they're just the beginning!



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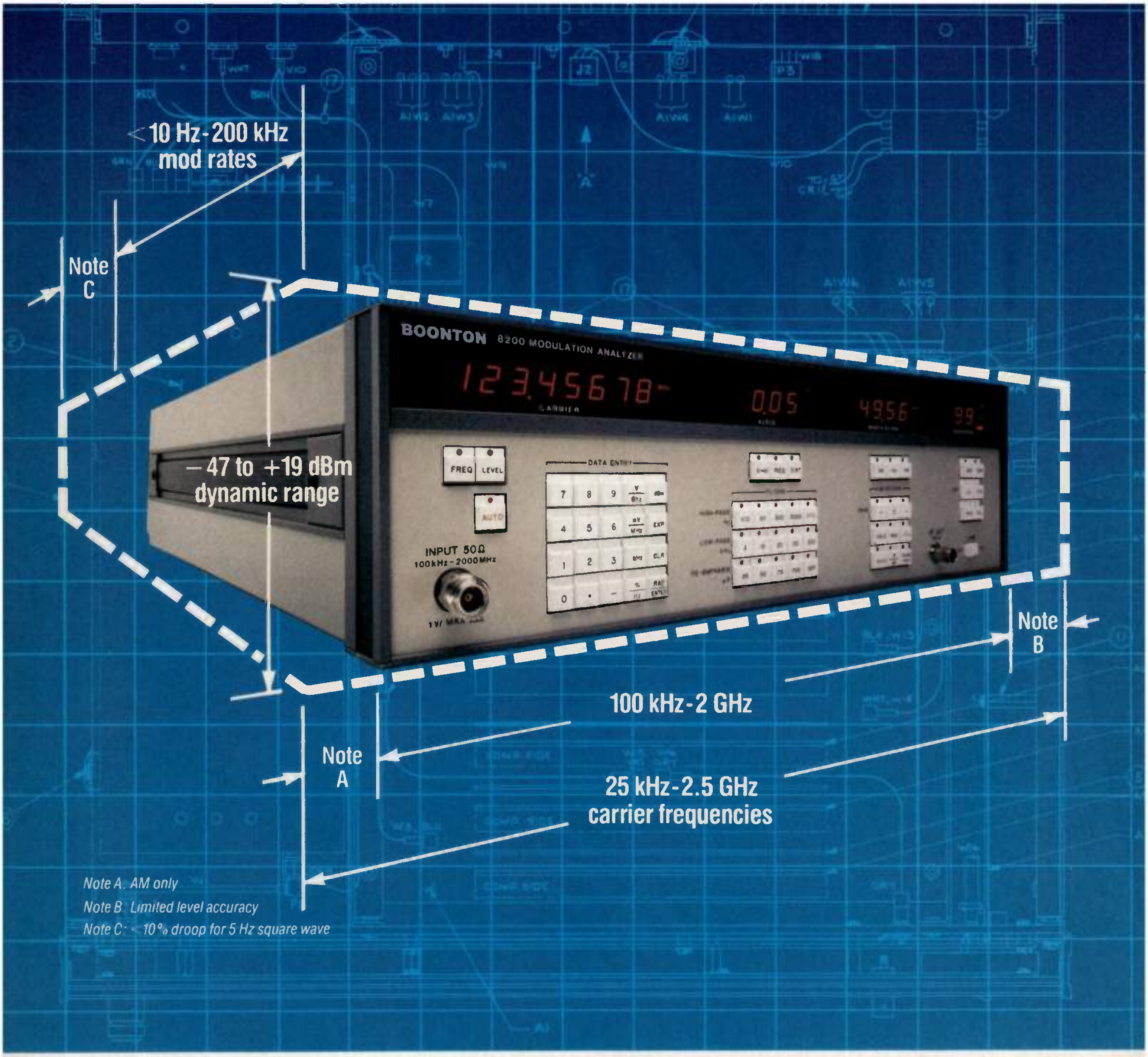

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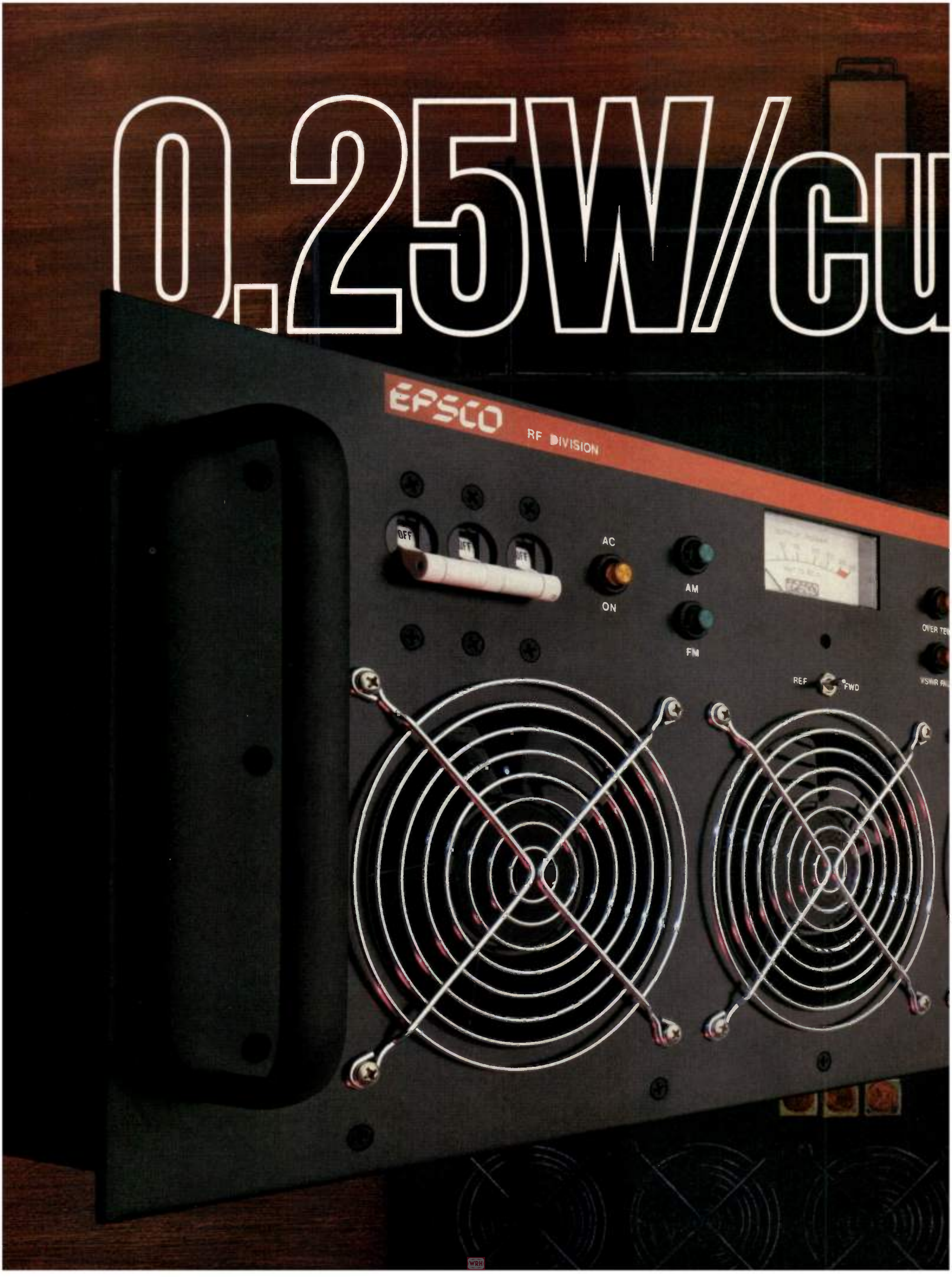

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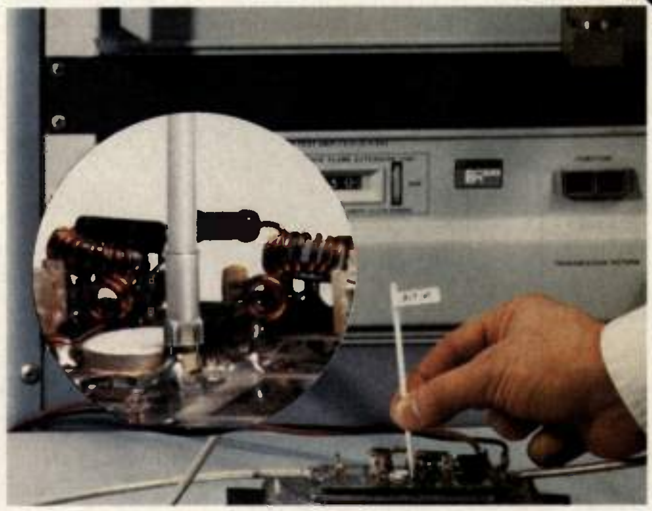

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rf letters

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

A Note on Terminology

Editor:

Please exhort your correspondents to join the majority of the world's engineering community in using metric dimensions for variables. I thought Britain was steeped in the traditions of the Industrial Revolution — we occasionally use thousandths of one inch — but at least we colloquially refer to these as "thou" (well, at least when designing steam engines for mine pumps) and not "mils." To me, it's like the Eagle landed using thrust measured in "slugs."

Roger Marmion
DMW Associates (Electronics) Ltd.
Braunston, Daventry, Northants, England

(RF Design *is completely in favor of standardization of measurements in metric terms. However, many products have been conceived, designed and built in the English system, and it is only logical to maintain the original terminology.— Editor)*

Microstrip Advice

Editor:

Regarding the problems experienced by Bob Hyde of the University of Texas in his letter entitled Microstrip Comments (August 1988, *RF Design*), he is right in concluding that a microstrip circuit having ground plane on both sides is neither truly microstrip nor truly planar. The Line Calc program which is part of Touchstone by EEsof does treat this type of line, simply calling it coplanar waveguide with ground plane. It also treats true coplanar waveguide in the same program.

To address his specific problem, I believe that if he makes the upper ground plane truly ground, his problem will go away. Regardless of the gap width of the line, there will be edge coupling as in a coplanar line and the upper ground must truly be the same ground as the lower ground. Grounding at the board edges will not do it. Physically tying the upper and lower ground planes together with rivets, Z-wires or plated through holes in as many places as practical is absolutely necessary. A practical number would be the number required to suppress the unwanted board resonances in the particular design. Theory tells us that the grounds must be a minimum of one half wavelength apart at the frequency of highest board resonance.

Bill Beauregard
Olektron Corporation
Webster, MA

French Correction

Editor:

Thank you for publishing my paper entitled "Lumped Element Phase Shifting and Matching Networks" in the July 1988 issue of *RF Design*. As some typographical errors have occurred, would you please print the following corrections:

- Figure 3 — the capacitor C2 on the right hand side should read C3.
- Figure 9 — the formula CG = .2979 pF should be replaced by CG = 2.979 pF.

I apologize for any inconvenience that this may have caused.

Andre Boulouard
CNET
Lannion-B, France

Article Postscript

Editor:

Mark Gomez' article "EMC Organizations and Societies" (August 1988, *RF Design*) is good except for a few spots. These are mentioned below.

The SAE AE4 does not represent all of SAE's EMC elements. AE4 is primarily concerned with aerospace. Land and water vehicles and devices are covered by two technical committees, Electromagnetic Radiation (EMR) and Electromagnetic Interference Standards and Test Methods (EMI). The EMR Technical Committee is concerned with electromagnetic emissions from the vehicle or device that may interfere with roadside or adjacent vehicle receptors, and is the U.S. National Committee participating in CISPR. The EMI Technical Committee is concerned with the electromagnetic susceptibility of vehicle electronics to roadside or adjacent vehicle emitters and intra-vehicle system compatibility, and is the U.S. National Committee for ISO.

ANSI-accredited standards committee C63 (ASC C63) has much more activity than implied in the article. In addition to its Ad Hoc committee on consumer electronics EMI (PL 97-259), it coordinates all other ANSI EMC-related standards and develops such documents as ANSI C63.2, C63.4, C63.5, C63.6, C63.7, C63.8, C63.12 and others.

An organization unmentioned in the article is the Radio Technical Commission for Aeronautics (RTCA). This body deals with EMC, and other topics, in commercial and private aircraft.

Edwin L. Bronaugh
Vice-Chairman, ASC C63
Amsterdam, NY

(For more information about RTCA contact: James Lohr, RTCA, 1428 K St., N.W., Suite 500, Washington, D.C. 20005.)

rf calendar

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October 17,18, 1988, Ottawa, Canada
Information: Contact local H-P sales offices listed in the white pages of the telephone book.

October 23-26, 1988
MILCOM '88
San Diego Marriott, San Diego, CA
Information: Timothy L. Dolan, MILCOM '88, TRW Space Defense Sector, P.O. Box 2568, Redondo Beach, CA 90278. Tel: (213) 812-4722

October 25-27, 1988
RF Expo East 88
Philadelphia Civic Center, Philadelphia, PA
Information: Linda Fortunato, Cardiff Publishing, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80110. Tel: (303) 220-0600; (800) 525-9154

October 26-28, 1988
Undersea Defense Technology
The Novatel, London, England
Information: P.G. Pinches, Microwave Exhibitions and Publishers Ltd., 90 Calverley Road, Tunbridge Wells, Kent TN1 1BR, England. Tel: (0892) 44027

October 27, 1988
3rd Annual EMC Event
Minneapolis Hilton Inn, Minneapolis, MN
Information: Diane Swenson, Tel: (612) 462-7001

November 8-12, 1988
Electronica '88
Munich Trade Fair Centre, Munich, W. Germany
Information: Gerald Kallman, Kallman Associates, 5 Maple Ct., Ridgewood, NJ 07450-4431. Tel: (201) 652-7070

January 25-26, 1989
14th Annual San Diego Electronics Show
Del Mar Fairgrounds, Del Mar, CA
Information: Epic Enterprises, Show Management, 3838 Camino Del Rio North, Suite 164, San Diego, CA 92108. Tel: (619) 284-9268

January 30-31, 1989
Electronic Warfare
The Catamaran Hotel, San Diego, CA
Information: Susan Call, Frost & Sullivan, Inc., 106 Fulton St., New York, NY 10038-2786. Tel: (212) 233-1080

February 5-10, 1989
1989 Aerospace Applications Conference
Breckenridge, CO
Information: Leo Mallette, Hughes Aircraft, MS: Bldg R-10, A9026, P.O. Box 92919, Los Angeles, CA 90009. Tel: (213) 334-2909

February 14-16, 1989
RF Technology Expo 89
Santa Clara Convention Center, Santa Clara, CA
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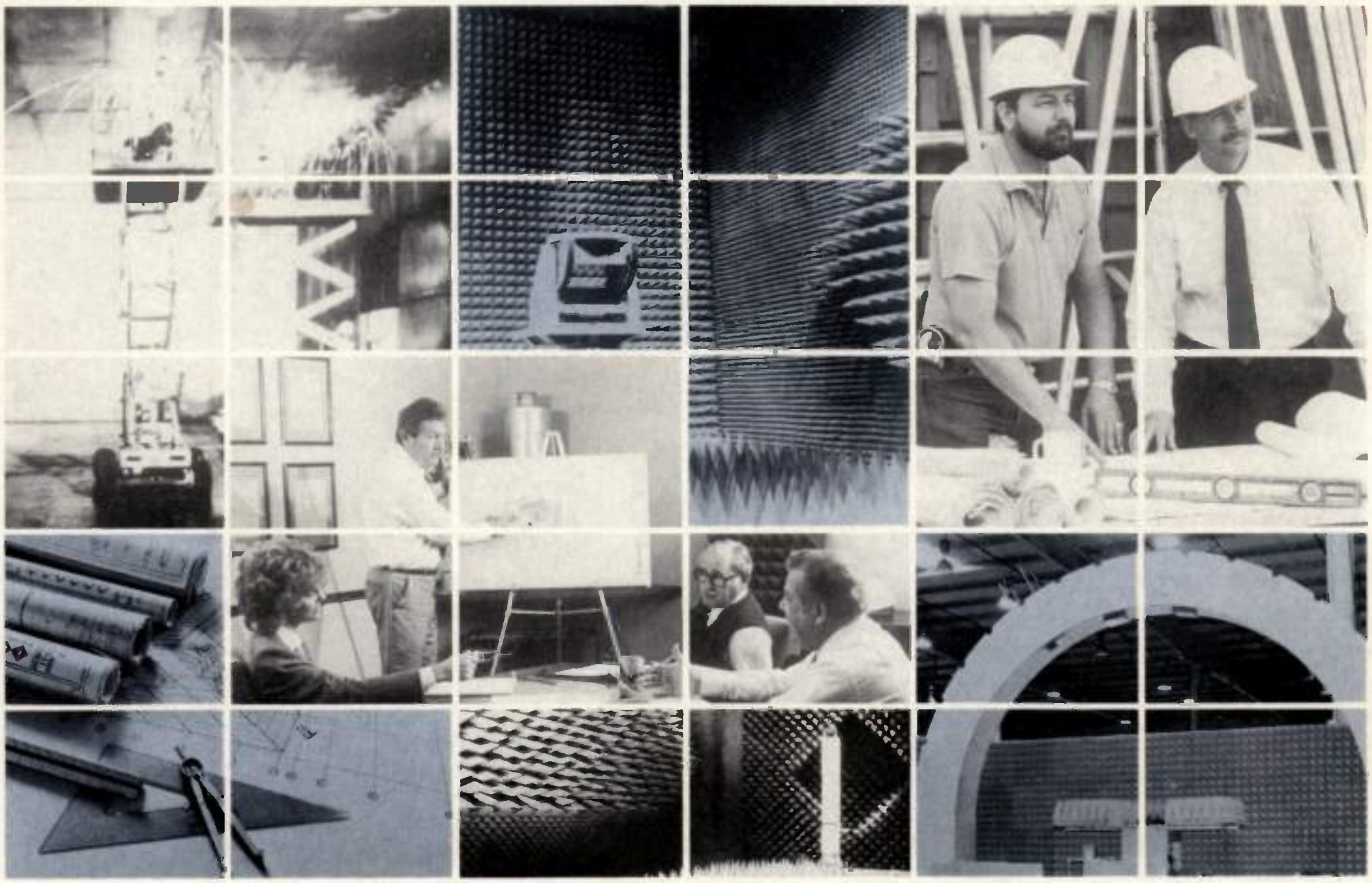

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November 1-3, 1988, Los Angeles, CA

Kalman Filtering II
November 28-December 2, Los Angeles, CA

Microwave Circuit Design I
November 28-December 2, 1988, Los Angeles, CA

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

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New HF Communications Technology: Advanced Techniques
October 17-21, 1988, San Diego, CA
December 12-16, 1988, Washington, DC

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

R and B Enterprises

Grounding, Bonding & Shielding
October 31-November 1, 1988, Washington, DC
November 21-22, 1988, Orlando, FL

TEMPEST: Documentation and Reports
October 26-28, 1988, Philadelphia, PA

Worst Case Circuit Analysis
October 31-November 2, 1988, Philadelphia, PA

TEMPEST: A Detailed Design Course
November 14-18, 1988, Philadelphia, PA

Real Life Solutions to EMI Problems
November 29-December 1, 1988, Philadelphia, PA

MIL-STD 461C/462 Test Workshop
December 8-9, 1988, Philadelphia, PA

Information: Grant R. Brown, R & B Enterprises, West Conshohocken, PA 19428. Tel: (215) 825-1960

EEsof, Inc.

Introduction to Microwave Computer Aided Engineering
November 7-8, 1988, Westlake Village, CA

Introduction to Microwave Computer Aided Layout and Design
November 9, 1988, Westlake Village, CA

Advanced CAE
November 14-15, 1988, Westlake Village, CA

Advanced Microwave Computer Aided Layout and Design
November 16, 1988, Westlake Village, CA

Nonlinear Circuit Design
November 21-22, 1988, Westlake Village, CA

Information: Sandra Scoredos, EEsof, Inc., 5795 Lindero Canyon Road, Westlake Village, CA 91362. Tel: (818) 991-7530

Interference Control Technologies, Inc

Grounding and Shielding
October 24-28, 1988, Atlanta, GA
October 31-November 4, 1988, Las Vegas, NV
December 5-9, 1988, Los Angeles, CA

Practical EMI Fixes
October 31-November 4, 1988, Atlanta, GA
December 5-9, 1988, Las Vegas, NV

EMC Design and Measurement
November 7-11, 1988, Dallas, FL
December 12-16, 1988, Washington, DC

Intro to EMI/RFI/EMC
November 15-17, 1988, San Diego, CA

TEMPEST Design and Measurement
November 15-18, 1988, Palo Alto, CA

Information: Penny Caran, Registrar, Interference Control Technologies, Inc., State Route 625, P.O. Box D, Gainsville, VA 22056. Tel: (703) 347-0030

Integrated Computer Systems

Digital Signal Processing: Techniques and Applications
October 18-21, 1988, Los Angeles, CA
November 1-4, 1988, Washington, DC
November 15-18, 1988, San Francisco, CA
November 29-December 2, 1988, San Diego, CA
December 13-16, 1988, Washington, DC

Image Processing and Machine Vision
November 1-4, 1988, Los Angeles, CA
November 29-December 2, 1988, San Diego, CA
December 6-9, 1988, Washington, DC

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

Georgia Institute of Technology

Principles and Applications of Millimeter-Wave Radar
November 14-18, 1988, Atlanta, GA

Phased-Array Antennas: Theory, Design and Technology
November 15-18, 1988, Atlanta, GA

Information: Education Extension, Georgia Tech, Atlanta, GA 30332-0385. Tel: (404) 894-2547

Compliance Engineering

EMI
October 18, 1988, Boston, MA
December 6, 1988, Orlando, FL

Safety
October 19, 1988, Boston, MA
December 7, 1988, Orlando, FL

ESD
October 20, 1988, Boston, MA
December 8, 1988, Orlando, FL

Telecom
October 21, 1988, Boston, MA
December 9, 1988, Orlando, FL

Information: Compliance Engineering, 271 Great Road, Acton, MA 01720. Tel: (508) 264-4208

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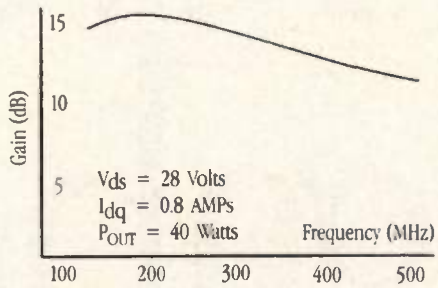

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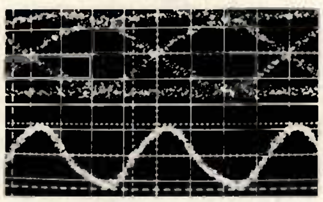

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UPG702B	3-Input NOR/OR	400 ps	3.0GHz
UPG703B	Four Stage Ripple Counter	500 ps	2.5 GHz
UPG704B-15	4:1 Multiplexer	800 ps	1.5 GHz
UPG704B-20	4:1 Multiplexer	800 ps	2.0 GHz
UPG704B-25	4:1 Multiplexer	800 ps	2.5 GHz
UPG705B-15	1:4 DeMultiplexer	800 ps	1.5 GHz
UPG705B-20	1:4 DeMultiplexer	800 ps	2.0 GHz
UPG705B-25	1:4 DeMultiplexer	800 ps	2.5 GHz
UPG706B-1	High Speed Flip Flop	400 ps	4.0 Gbps
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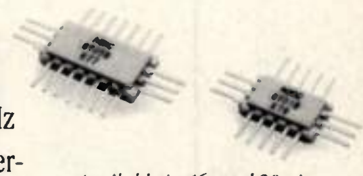

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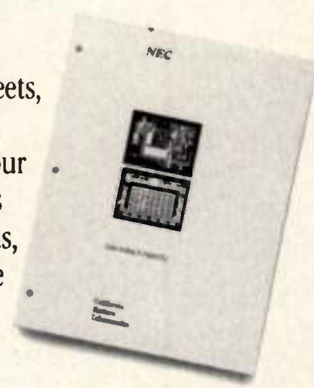

California Eastern Laboratories

CEL, 3260 Jay Street, Santa Clara, CA 95054; (408) 988-3500 ☐ Los Angeles, CA (213) 645-0985 ☐ Bellevue, WA (206) 455-1101 ☐ Scottsdale, AZ (602) 945-1381 or 941-3927 Richardson, TX (214) 437-5487 ☐ Burr Ridge, IL (312) 655-0089 ☐ Cockeysville, MD (301) 667-1310 ☐ Peabody, MA (617) 535-2885 ☐ Hackensack, NJ (201) 487-1155 or 487-1160 Palm Bay, FL (305) 727-8045 ☐ Norcross, GA (404) 446-7300 ☐ Nepean, Ontario, Canada (613) 726-0626 ☐ Europe, NEC Electronics GmbH 0211/650301

INFO/CARD 17

New Name, New Agenda for NBS

As a result of the Omnibus Trade and Competitiveness Act, signed into law on August 23, the National Bureau of Standards (NBS) officially became the National Institute of Standards and Technology (NIST). The new institute retains all of the traditional functions and services of NBS and takes on several new assignments designed to boost American industry in the world marketplace. The new responsibilities build on the technical expertise of NBS, the only federal laboratory with a mission specifically to support U.S. industry. NIST will work with several new constituencies, including state and local economic development organizations. Four major new programs are called for by the legislation: 1) the development of regional centers for transfer of manufacturing technology, 2) the creation of a focal point within the federal government to work with and support state and local industrial extension services, 3) the creation of an advanced technology program to support and encourage the rapid commercialization of promising new inventions and technologies and 4) the creation of a national clearinghouse of information on state and local technology development initiatives.

EMC Products and Services Report Predicts Growth in U.S. Demand

U.S. end-user demand for electromagnetic compatibility (EMC) products and services is expected to grow from the 1987 level of $1.1 billion to $2 billion in 1992, according to a market research report by Frost and Sullivan, Inc. of New York. This represents a substantial, real, compound annual growth rate of 11.7 percent. The overall EMC marketplace is being influenced by the higher standards of protection being specified by the military and government security sectors of the market. Demand for electromagnetic interference (EMI) shielding is being driven by the need to protect ever more sensitive military equipment used in closer proximity and by the concern for security from a more sophisticated surveillance of eavesdroppers in telecommunications and computing. In addition, the 1983 FCC specifications are not only being more stringently enforced, but are expected to be applied to a broader spectrum of equipment.

The complete report (#A1938) is available at a cost of $2,150 from: Frost and Sullivan, Inc., 106 Fulton Street, New York, NY 10038. Tel.: (212) 233-1080.

Foreign Technology Newsletter

Foreign research and development advances are the subject of the new *Foreign Technology Abstract Newsletter,* a joint effort of the National Science Foundation, the U.S. Department of State and the National Technical Information Service of the U.S. Department of Commerce. The newsletter provides coverage of foreign R&D in such fields as advanced microelectronics, structural ceramics and superconductivity. Contributing sources include Science and Technology Counselors at U.S. Embassies, the Office of Naval Research, the Foreign Broadcast Information Service and American scientists and scholars abroad. Coverage includes reviews of major research programs, reports on activities of specific organizations and laboratories, conference summaries, and descriptions of specific new technologies, processes and discoveries. One to two hundred new foreign technical reports are also summarized in each issue. Annual subscription rate is $125. To order: Request PB88-903900/KKA, NTIS, Springfield, VA 22161. Tel. (703) 487-4630.

Gould Sold to Nippon Mining Company

Gould, Inc. and Nippon Mining Company, Ltd. of Tokyo announced the execution of an agreement for the acquisition of Gould by Nippon Mining in a transaction valued at approximately $1.1 billion. Gould and Nippon Mining have worked together for close to seven years. In 1981, the two companies entered into a joint venture, Gould Nikko Foil Company, Inc., to produce copper foil in Japan. Plans to build a manufacturing plant in Hong Kong were recently announced, with completion scheduled for 1990. The acquisition of Gould will enable Nippon Mining, one of Japan's 50 largest industrial companies, to enhance its position in several key markets, including electro-deposited copper foil, fuses, fiber optic components and optoelectronics. Nippon Mining has indicated its intention to proceed with Gould's ongoing plans to divest itself of its remaining defense systems businesses.

U.S. and West German Companies Team Up For Patriot System Work

Raytheon Company has entered into teaming agreements with American and

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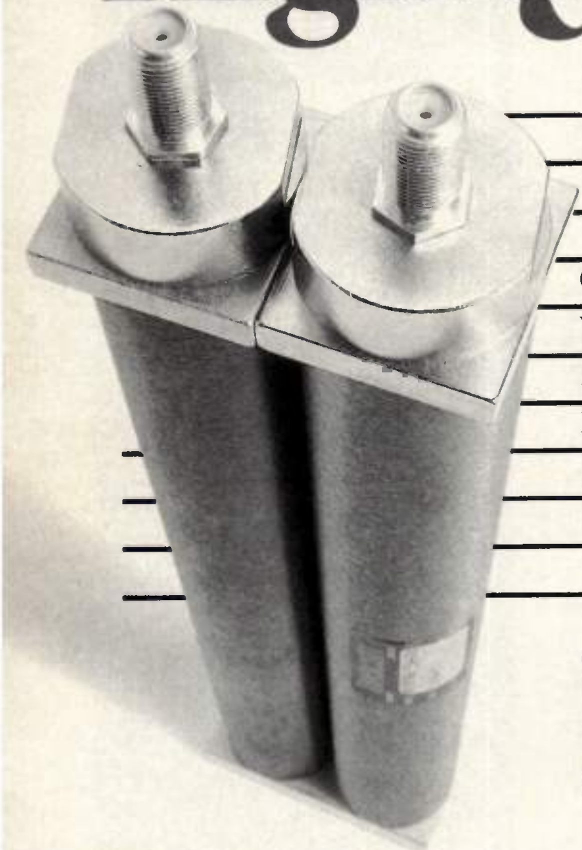

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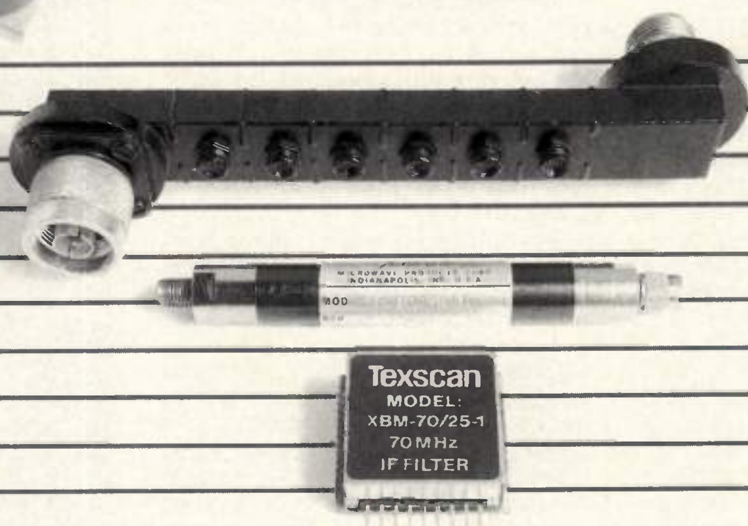

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rf news Continued

West German companies to cooperatively extend the capabilities of the Patriot missile system. Raytheon, the system's prime contractor, has joined Martin Marietta, AEG, Messerschmitt-Boelkow-Blohm (MBB) and Siemens to perform research and development efforts aimed at improving the Patriot system's Anti-Tactical Missile (ATM) capability as well as the system's performance against advanced Air-Breathing Threats (ABT). Initial activities include studies and engineering development models to investigate advanced missile and radar concepts. Funding for these bilateral cooperative efforts comes from both the Federal Republic of Germany and the United States in a multi-phased program.

M/A-COM Receives MIMIC Subcontract

The Advanced Semiconductor Division (ASD) of M/A-COM, Inc. has received a letter subcontract in excess of $5 million for development work, GaAs monolithic IC wafer fabrication and other activities in support of the Hughes/GE team on the Department of Defense-sponsored Microwave/Millimeter Wave Monolithic Integrated Circuits (MIMIC) program. The award was made by the Radar Systems Group of Hughes Aircraft Company of El Segundo, Calif. The MIMIC program is aimed at enhancing the availability and affordability of analog monolithic IC chips for a wide variety of microwave defense and space systems. In addition to development and support activities, M/A-COM ASD will fabricate GaAs MIMICs for line certification and system brassboard insertions for airborne phased array radar, the Advanced Tactical Surveillance Radar (ATSR) and Global Positioning System (GPS).

High Accuracy Satellite Time Transfer

Scientists at the National Institute of Standards and Technology (NIST) report that they can now utilize several commercial communication satellites to disseminate a time signal whose eventual coverage will be worldwide. It will be an extremely accurate signal — to within one nanosecond (one billionth of a second). The experimental service was initiated recently using domestic U.S. satellites and will be extended to an Intelsat satellite; users require both a receiver and a transmitter.

Information Handling Services Acquires D.A.T.A. Business Publishing

Information Handling Services (IHS) of Englewood, Colo., has purchased D.A.T.A. Business Publishing (DBP) from Mitchell International, Inc., a division of International Thomson Organization, Ltd. As part of its vendor catalog product line, IHS currently produces and distributes several electronic components-parts databases, distributed on microfilm and microfiche as well as through an online service. DBP is the developer of several significant databases describing the comparative electrical and physical characteristics of integrated circuits, discrete semiconductors and other special electronic devices. IHS foresees converting the DBP databases so that they can be distributed on CD-ROM or other electronic format. These databases may also be integrated into current IHS information services.

EEsof Selected as Phase 1 MIMIC Team Member

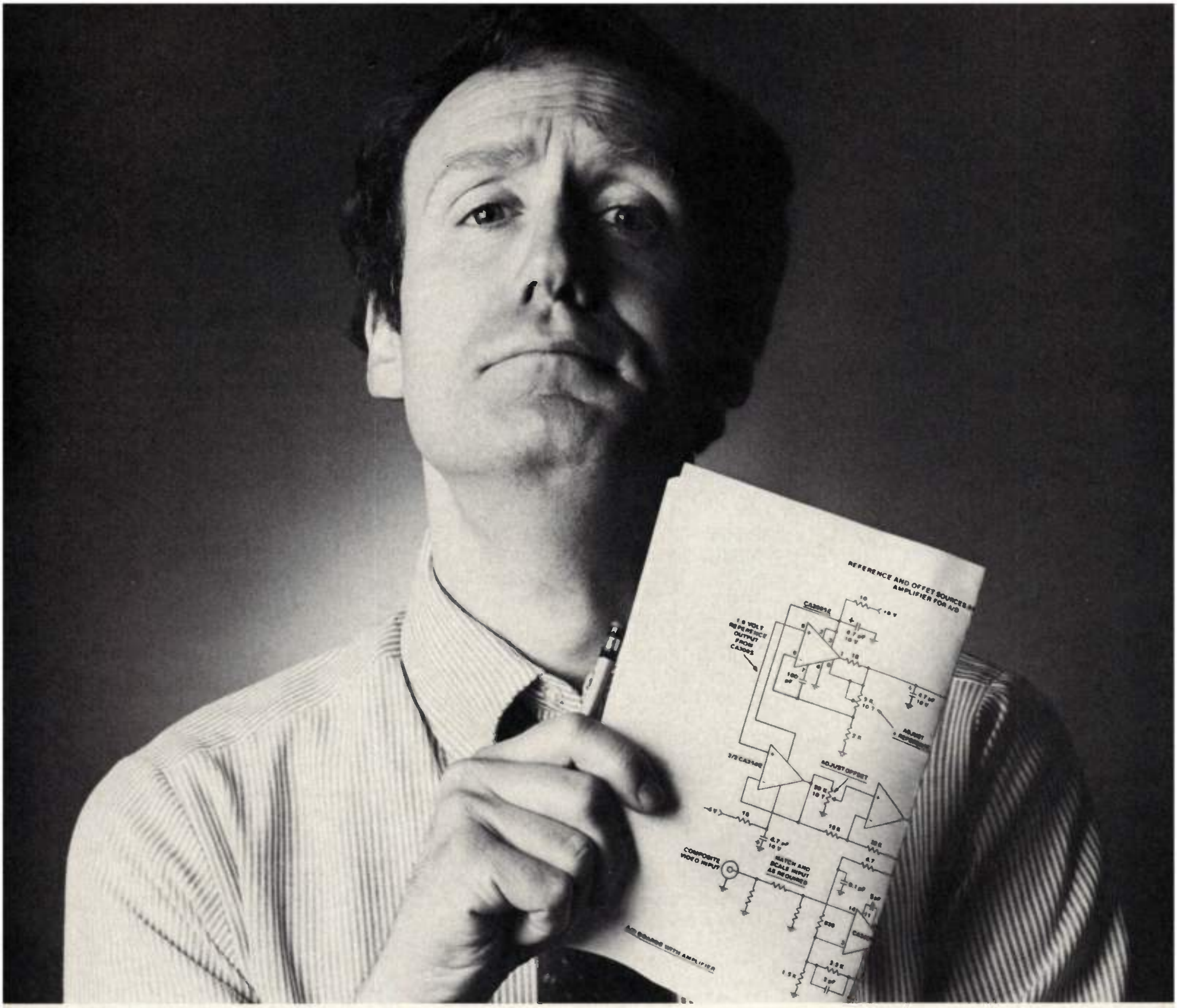
EEsof, Inc. of Westlake Village, Calif., has been named a member of the Hughes/GE team to participate in Phase 1 of the MIMIC program on a $50 million contract, under which EEsof holds a subcontract for $2 million. EEsof's primary responsibility will be the development of a MIMIC design workstation, a fully-integrated turnkey system of computer-aided engineering tools for the simulation and design of GaAs MIMICs. The workstation will be a full-solution, user-friendly CAD system with a design library containing verified electrical models and physical designs for multiple GaAs MIMIC foundries. EEsof's role will be to expand MIMIC GaAs technology in the area of software development and to provide microwave/RF engineers with accurate and flexible software engineering tools. EEsof has been encouraged by the MIMIC Program Office and Hughes/GE to commercialize the software technology developed for the program, and this technology will be made available as upgrades to existing EEsof products or as new products.

ARX Subsidiary Awarded Raytheon Subcontract

Aeroflex Laboratories, Inc., a subsidiary of ARX, Inc., has received a $2.5 million subcontract from Raytheon Company's Equipment Division. The subcontract calls for two AN/FSS-17 Interference Analysis Systems to be installed as part of the upgrade of two PAVE PAWS radar sites, a program sponsored by the U.S. Air Force Electronic Systems Division, Hanscom AFB.

Ericsson to Install Trial Digital Mobile Telephone System

Ericsson and the Swedish Telecommunications Administration have agreed to install a trial system to test the new pan-European plan for digital mobile tele-



phony within the Stockholm area. The trial system will be installed and operating by the beginning of 1989 to test important functions of the new digital transmission technique from vehicles in Stockholm traffic. The trial will determine if the system can be designed in accordance with the joint European recommendations and function technically in actual operating conditions. Sixteen European nations, including Sweden, have agreed that a new digital mobile telephone system will be installed and operating on a pan-European basis by 1991.

Raytheon Awards Soladyne Standard Missile II Subcontract

The Soladyne Division of Rogers Corporation has received a one-year $1 million subcontract from the Missile Systems Division of Raytheon Company, Bristol, Tenn. Soladyne, based in San Diego, will supply Raytheon with complex microwave frequency stripline circuit boards for the Standard Missile II program.

Tracor Aerospace Receives $15.6 Million Air Force Contract

Tracor Aerospace, Inc., a subsidiary of Tracor, Inc. of Austin, Texas, has received a $15.6 million contract for full-scale development of an AN/ALE-47 advanced countermeasures dispenser system for use on board U.S. Air Force and U.S. Navy tactical aircraft. The four-year development contract was awarded by the Department of the Air Force, Headquarters, Aeronautical Systems Division, Wright Patterson Air Force Base, Ohio, and includes five production options with probable requirements totaling $77 million.

Teledyne Microwave to Supply French Satellite Program

Teledyne Microwave of Mountain View, Calif., has been selected as the coaxial switch supplier to Alcatel Espace, coprime for France's Telecom-2 satellite. The Telecom-2 geosynchronous satellites will provide communication paths for commercial, military and television signals within France and between the French nation and its territories. Teledyne has supplied Hi-Rel switches to the Apollo-Soyuz and Space Shuttle missions, as well as to various maritime and classified military space programs.

Cornix Technologies Purchases U.K. Amstar Electronics Unit

Cornix Technologies Plc, a newly-founded U.K. defense electronics company, has purchased the U.K. subsidiary of Amstar Corporation of the United States. The subsidiary, Amstar Electronics Products Ltd., is a manufacturer of communications intercept, location and analysis receivers as well as communications distribution and switching systems. The acquired company, which will be known as Cornix Systems Ltd., is headquartered in Coventry, England.

PMI Acquires SSMT

Precision Monolithics, Inc. (PMI) of Santa Clara, Calif., has announced the acquisition of SSM Audio Products (SSM), formerly known as Solid State Micro Technology for Music, Inc. (SSMT), also of Santa Clara. The acquisition should afford greater financial and manufacturing resources to SSM while easing PMI's entry into the high-end consumer/audio IC market. The markets served by the SSM product line include applications such as synthesizers, organs, mixing consoles, studio equipment and broadcast systems. Current SSM products use semicustom linear bipolar arrays for low-noise, low-distortion amplifiers and specialized audio circuits. PMI, a Bourns company, is a manufacturer of precision analog and converter integrated circuits in both CMOS and bipolar technologies. The acquisition of SSM marks PMI's entry into the consumer marketplace.

AMP and Matrix Science Finalize Merger

AMP Inc. of Harrisburg, Pa., has announced the merger of Matrix Science Corporation following approval by Matrix shareholders of a deal valued at $110 million. The merger with Matrix Science enables AMP to offer a broad range of interconnects available from a single source. Matrix provides AMP with a strong U.S. customer base, while AMP enhances the international opportunities.

AEL Awarded $4.3 Million Navy Contract

AEL Defense Corporation of Lansdale, Pa., has been awarded a $4.3 million production contract by the Naval Avionics Center in Indianapolis for aircraft amplifier systems. AEL will produce eight High Power Microwave Amplifier Systems (HPMAS) and test equipment for the Multi-Service Electronic Warfare Support Group (MEWSG). HPMAS is a training device that provides high continuous wave (CW) power over a broad range of microwave frequencies. In all, AEL has been awarded contracts on the HPMAS program totaling $21.5 million. AEL designed and developed the prototype system and has delivered eight production systems to date. The new award will increase AEL's HPMAS production to 46 systems.

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All Instrument Specialties shielding strips can be attached quickly and easily to the surface desired. Some carry a self-adhesive strip for this purpose. Others clip-on and are friction-held; many can be supplied with barbs that bite and grab, preventing slippage. Still others are supplied with holes for riveting.

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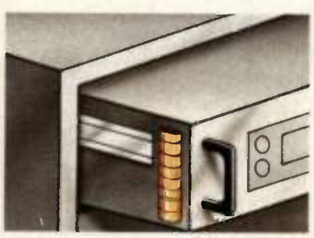

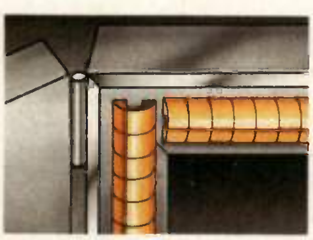

The Symmetrical Slotted Shielding (S³) series is perfect for sliding drawers and door enclosures, where frequent opening and closing occurs.

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INFO/CARD 24

rf cover story

RF Design 1978-1988: A 10th Anniversary Celebration

With this issue, RF Design *completes 10 years of publication, and the RF industry has seen plenty of change since that November/December 1978 inaugural issue. Let us look back over those 10 years with a little history from the perspective of* RF Design, *plus the observations of two engineers who were there when the magazine began.*

It was 1978, and the digital era was growing at an exponential rate. RF and other analog design techniques had been well covered by several major electronics publications up until then, but they had recently chosen to focus their efforts on microprocessors, data communications, and the rest of digital electronics. At the same time, there were successful microwave magazines, covering the unique aspects of the laws of physics that dominate electronics at GHz frequencies. RF technology was rather suddenly left alone in the trade publishing industry.

It was only natural, then, that Cardiff Publishing Company would consider a new publication: *RF Design.* Although nearly all of the people who inaugurated this magazine are no longer associated with it, the idea has survived, and that is the best sign of success.

Developments in RF Technology

Ten years isn't a terribly long time in history, but there are a few areas that have seen remarkable progress from the viewpoint of *RF Design:*

Solid State RF Power. The transition from tubes to transistors was well under way in 1978, but has skyrocketed since then. A couple kilowatts at AM broadcast frequencies, a couple hundred watts at HF and VHF, and less than 100 watts at higher frequencies was the state-of-the-art. The "big" bipolar devices put out around 80 watts, and RF power MOSFETs had just been developed.

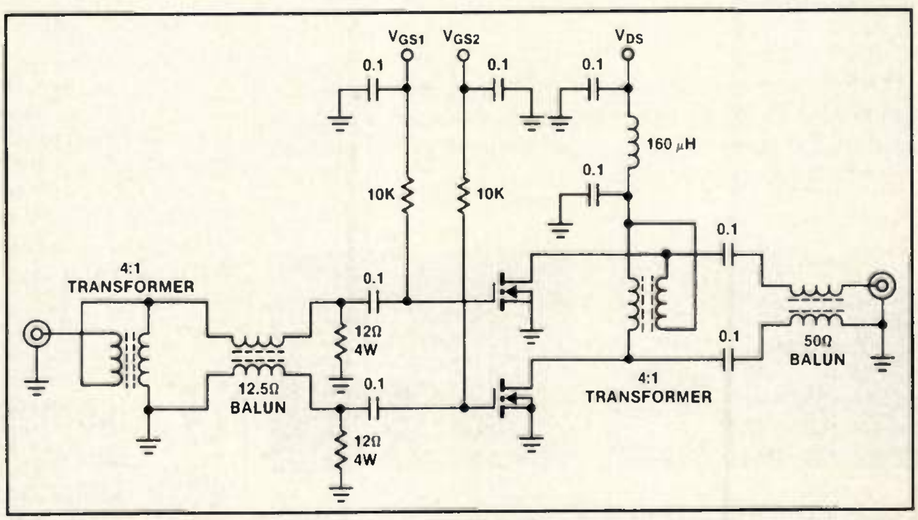
Now there is no upper limit to solid state RF power. Devices are available to produce up to 600 watts each. Amplifier design and combiner technology have been honed to a fine art. Solid state devices are now the first choice in virtually every RF power amplifier design because they are reliable, available and offer the necessary performance.

The first issue of RF Design *reached the desks and benches of RF engineers in November 1978.*

Computer-Aided Design. CAD was an infant 10 years ago, in the RF business. Engineers just didn't have computing power readily available to them. Paralleling the growth of the personal computer has been software for RF engineers. Offerings range from small, specialized programs for daily work, to incredible number crunching for system performance analysis, S-parameter based circuit design with libraries of components, or complete method-of-moments electromagnetic radiation analysis. The benefit to engineers is unmeasurable.

Digital Techniques. Although digital devices and techniques are the biggest chunk of the electronics industry, they haven't taken over every analog function as some soothsayers predicted. What has happened is a recently developed analog/digital cooperation to achieve high performance in RF systems. Digital controls are now essential to manage the functions of complex systems, but this is just an external application. Digital circuits themselves are now fast enough to be considered RF-like, another area where cooperation has begun.

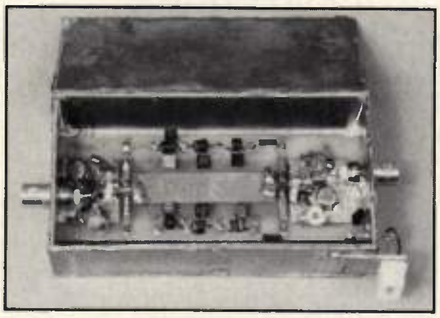
The biggest impact on RF, however, is the direct use of digital circuitry for RF applications. After achieving an analog conversion to IF, digital signal processing is now the principle analysis method in radar, EW and ECM systems. A few systems convert from analog transmission to digital processing right after the antenna. Frequency synthesizers are the most common application, with both phase locked loops and direct digital synthe-



With this amplifier, Gary Appel and Jim Gong introduced many readers to the potential of RF power MOSFETs.

sizers representing this RF/digital hybrid area. A couple more subtle areas are square wave (digital) drive to mixers for IMD reduction and switch-mode power amplifiers using digital drive circuitry.

Integrated Circuits. This is one more area where significant progress has been made, affecting the size and cost of RF equipment. Log amplifiers, commercial and consumer receivers, universal amplifier building blocks, audio functions, and A/D or D/A converters are some prime examples. The result is twofold. First, the monolithic devices are cheaper and may even be higher performance than discrete designs. The other benefit is reduced design time, since interfacing a functional "building block" is much easier than designing a whole subsystem from scratch.



An example of RF Design's *tutorial contribution is this linear phase filter, from a May 1982 article by Robert Sellers.*

There are many other areas that deserve attention, such as the growth of modular products and the rapid rise of GaAs technology. The new applications of RF that have developed in the past 10 years could fill an encyclopedia set. The few significant enhancements to RF technology which have been highlighted here are truly the "tip of the iceberg." RF is a strong, diverse, and active technology.

Some Significant Things That Have Happened To RF In The Last 10 Years

John Minck
Hewlett-Packard

I have always considered myself present at the birth of *RF Design*. To me it was an obvious idea. In HP's business, RF instruments were neck and neck with microwave instruments. Whether you put the dividing line at 1000 MHz or 2000 MHz, they were somewhere near equal. It's not hard to rationalize that, with the enormous amount of activity in those HF, RF, VHF, UHF bands, whatever acronyms you use.

This was the first magazine in my memory that HP supported with advertising from almost the first issue. We felt that strongly about having it succeed. And succeed it did, but not without some years of heartburn. I remember especially the early sophomoric graphics, and some of the articles were pretty lightweight.

You see, there was an arrogance in the early '70s that was more than convinced that all circuitry would either be on a silicon chip, or be in microwave hybrid microcircuitry on sapphire or ceramic. Certainly those technologies were rushing ahead fast, but a lot of practitioners knew that some things were just always going to be best (cheaper) on old-fashioned printed circuit boards. When you see an elegant trick circuit made by wrapping the wires just right around a ferrite core, or using some particularly-shaped p.c. traces, you just have to stop and admire the innovation. That's what such a magazine is for.

I like to think of RF as the backbone technology, not just in its own territory, but also the underpinning of microwave, millimeter-wave and even lightwave. Look at the inside of a microwave generator or a microwave communications system. About 35 percent is bus interface, memory and smarts, front panel interface, and general digital things. About 35 to 40 percent is baseband, IF, RF and VHF frequency. These are the voltage-controlled oscillators, the phase-lock loops with fancy digital divide-by-n components, the IF amplifiers, filters, converters, etc.

Somewhere less than 25 percent is true microwave, the up-converters, the power amps, the filtering, the attenuators and signal control, the switching, etc. The same thing happens when you divide the lab engineers by function, except that there are far more firmware and software folks than you would imagine.

The point is that if you don't have the backbone, you don't have a system. Microwave CAE and component technology has gotten a whole lot of press for 10 years. But those things that have happened to a variety of RF technologies in the past 10 years are also pretty dramatic. Yet the glamor and glitz doesn't show it. Here are some of those changes in RF technology:

EMI

Those EMI voices crying out in the wilderness for 20 years have finally gotten our attention. They were right all along, and now a lot of it is writ in regulations and law. We simply can't live in a modern society with electronic smog all over the place.

The trouble was that 10 years ago, EMI was still a black art. You could easily measure emissions coming from a product. But try to duplicate it tomorrow or on the other side of the world, and you'd get 10 to 20 dB differences. An HP case in point was a product that passed the EMI specs in design and qualification, but on entering a small European country, the testing showed non-compliance, and the product was confiscated along with the threat for all future ones.

What got better during that time was that the National Bureau of Standards in Boulder got going on the TEM-cell measurement method which finally provided better understanding of the vagaries of EMI measurements, even when using recognized equipment and antennas and methods. Following that, the inrush of spectrum analyzer technology into the area really put the solution-set at a point in the design cycle where some dramatic improvement could happen, right on the design bench. No longer were the Band-Aids applied to the outer cabinet at the end of the design cycle, but instead they were applied far upstream where costs of fixup were smaller.

RF designers can take a bow for their recognition of the problem and efforts to do something about it. *RF Design* can take a bow on their continued harping (some would call it nagging) on the issue. Instrument suppliers can take a bow on equipment and software configurations with personality sets that turn bench-top spectrum analyzers into EMI sniffers.

Metrology

Pity the poor U.S. National Bureau of Standards (just re-christened as the National Institute of Standards and Technology — NIST). Attached to the Department of Commerce, probably the least technical of all the cabinet organizations, yet mandated by the founding "Organic Act" of 1907 to standardize all parameters of a modern high-tech society. The NBS struggles under a woefully meager budget to research measurement technology and supply measurement services to support the measurements revolution of the '80s and '90s. Truly an impossible task.

In the first place, there are so many more RF parameters than just a decade before. Phase noise came out of nowhere. Noise figure got 10 times more accurate. Fiber optics barely existed in 1978, but now will preside over the wiring of the world. Materials science runs rampant with new semiconductor materials, surface effects of thin film and deposition technologies of all types along with "designer" layered composites. All require an understanding of the physical effects of materials as well as some kind of handle on measurement methods.

The "easy" measurements of 1907 and 1970 whereby everything was stabilized or "soaked" before taking the readings are gone forever. Now, it must be dynamic. And I don't just mean automatic, which is also true. Agile signal generators now

This radar chirp signal was generated by digital means to be highly accurate for 100 MHz chirp bandwidth.

hop, hop, hop and chirp, chirp, chirp and jitter, jitter, jitter (to quote a recent ad). But in doing so, how does one measure the effective carrier phase noise if the carrier itself is hopping to another frequency in 8 nanoseconds, like our HP 8770A Arbitrary Waveform Synthesizer?

The traditional AM-FM/Phase Mod-pulse generator is still with us and will be for a while. But vector modulation, the I/Q type with baseband bandwidths of DC to 250 MHz could give a metrologist fits. So far, most of them have chosen to ignore it.

National productivity and competitiveness are terms that plague every session of Congress. With the disastrous budget deficits, getting a few million more for NBS initiatives is terribly difficult. But isn't something wrong with a system that sets the entire NBS budget at perhaps $130 million (maybe $10 million for measurement services), while a B-1 bomber costs $500 million per plane and a commercial 747 which depends on accurate navigation and time standards or GPS costs $150 million each?

Yes, measurement services are a long way from glamor. But they underpin so much of our technology. And the constituency is so diverse that no one seems to care. Does anyone care?

Reliability

Something dramatic has been going on for about 10 years. In 1980, our HP President John Young set a goal for designers to shoot for that many thought much too aggressive. He asked for a 10 to 1 improvement in MTBF reliability, to be achieved in the decade of the 1980s. With about 1.5 years to go, HP has come very close to achieving that goal. It is a noteworthy fact, and not limited to HP alone.

At recent meetings of the Equipment Management Forum (EMF), an ad-hoc committee of the National Conference of Standards Labs (NCSL), the operations data coming in from dozens of users shows that the repair portion of calibration/metrology costs are coming down dramatically. For example, the latest HP 437B Power Meter now has a stated MTBF of 120,000 hours. The confidence is high since an earlier HP 438A Meter stated 25,000 hours and actual warranty experience bore out an actual number of 45,000 hours.

Even complex products show dramatic effects. A microwave sweeper with over 6000 active components has a stated MTBF of 8000 hours. Why not? Consumer stereos and TVs regularly come out of the box and turn right on and run for years and years without a second thought of calibration or repair. Certainly instrumentation is far more complex and accurate in general, but once expectations are raised, whether with good cars or good cameras or good TVs, the other products get similar treatment.

Digitalization of Everything

Inroads of digital technology on the obvious front-panel functions and the computer control of measurements and measuring instruments are obvious, but the replacement of analog functions in the VHF domain has been nothing short of amazing.

Certainly we were ready for digital phase lock loops and divide-by-n circuitry because, after all, that was like familiar digital counter circuitry. We even looked on benignly as the A-to-D converters which started in digital voltmeters got up to the point where you could use their 20 MHz response for fast-Fourier spectrum analysis. That's where the waveform analysis measurement function came from, and it makes a whale of a close-in-to-the-carrier analyzer.

But next came terms like synthetic video. It was the idea of digitizing the IF of a radar and sending the video on a

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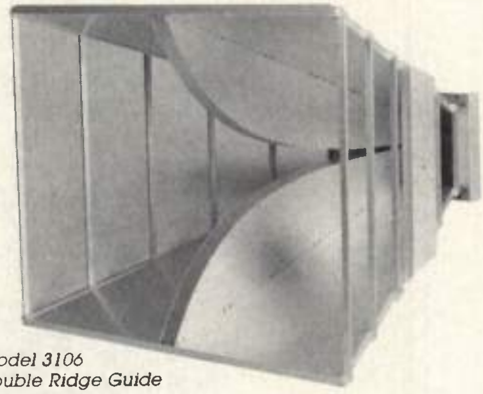
1–18 GHz
Model 3116 Double Ridge Guide
18–40 GHz



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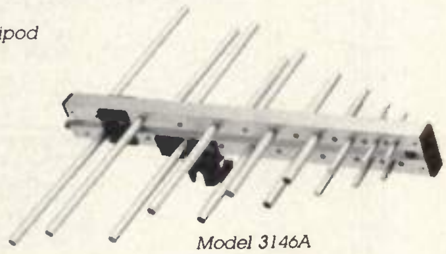

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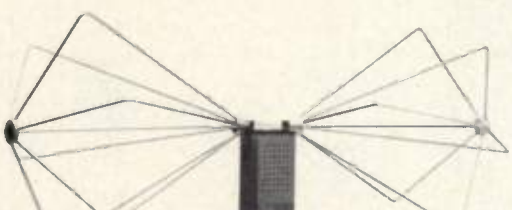

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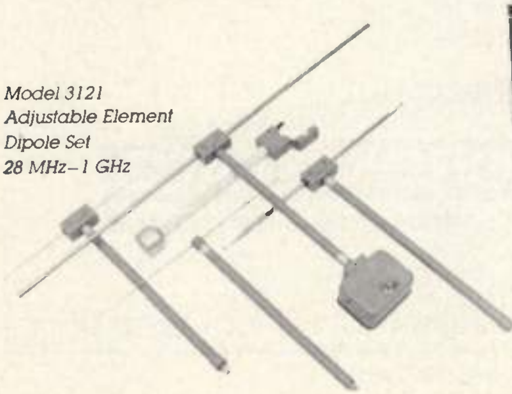

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digital bus back to the airplane's computer or wherever. And next, digital oscilloscopes which hit 500 MHz, then 1000 then 20 GHz. HP even dropped the remnants of their analog scope line a couple years back.

And the communications systems went to digital PSK modulations formats with terms like 64QAM and 49QRS. The "spook" systems went to agile and phase-mod and who-knows-what? The radars went to phase-tagged pulses and I & Q signal processing to dig more detail out of the echoes by use of the essential vector data in the signal. (Think of it as a vector network analyzer where you have control of the input test signal and have a long path in the test loop.) These days you have a lifetime job if you're into signal processing. It's mostly digital, but at RF speeds.

Then on the signal generation side of things, it's the fast DAC. DC to 50 MHz baseband digital signal generators can lead to a whole series of complex simulation possibilities. You can feed two of them into an I/Q signal generator, one for I and one for Q and get an exceptionally complex modulated signal. You can take that 50 MHz agility (8 nanoseconds) and move that window anywhere up to 3000 MHz in another 200 nanoseconds. And you can up-convert that into microwave. The most interesting thing to us old analog generator folks is that these digital simulators are absolutely repeatable. You can ask for a 1 percent AM modulation from two units at each end of the country and they are dead on. With 12- bit DACs, you're talking about repeatability of 1 part in 4092.

And what of the future? Is there any reason whatever to assume that these trends won't continue? None! And interestingly, the RF engineer is in the middle of it. Because most of these things are happening in dramatically-expanding basebands and ever-rising IFs, signal processors that go to 350 MHz, and lightwave bandwidths of 2 GHz. That's all RF-Land to many of us. And that is *RF Design*.

And away we go for another 10 years. rf

Reflections — The Past Decade Of Our RF Technology

Doug DeMaw
Oak Hills Research

One of the notable changes in the RF communications technology is the advance in receiver dynamic range. Ten years ago and before, we found commercial receivers deficient in minimum discernible signal (MDS) levels, notorious for phase noise from synthesizers and sorely lacking in IMD performance. IMD dynamic range numbers as low as 68 dB were common, with the better receivers in the low 80 dB range. Past receiver designs resulted in countless spurious responses through the receiver tuning range.

Then homemade receivers designed by engineers such as Wes Hayward (W7ZOI) and myself (W1FB) were demonstrated, exhibiting superior dynamic range over top-of-the-line commercial products.

Dynamic range figures in excess of 100 dB were proven inexpensive and practical when attention was paid to gain distribution and the correct choice of active devices in the critical RF amplifier, mixer and post-mixer stages. The use of diplexer circuits after the mixer also enhanced IMD performance. These published designs helped to stimulate receiver manufacturers toward better receiver performance. Phase noise (which became an obvious problem when the dynamic range was improved) has now been minimized and spurious responses are seldom a problem.

Another forward step in our technology is in power MOSFET design and application. Higher efficiency, compared to comparable bipolar transistors in RF amplifier service, is a keynote of power FETs. Typical bipolar power devices yield Class C and Class AB efficiencies on the order of 50 to 60 percent in broadband service. Power FETs can provide an efficiency as great as 90 percent, according to my colleague, Ed Oxner (formerly of Silconix, Inc.). I have achieved power-FET efficiencies of 70 to 80 percent with amplifiers of my design.

Furthermore, a properly designed FET RF amplifier is less prone to damage from high VSWR, compared to a bipolar amplifier. Thermal runaway is practically unheard of with power FETs. Finally, amplifier IMD is markedly lower when using power FETs. If there is a negative factor to be considered, it's the fragility of the FET input circuit when excessive gate current flows, or when the gate-source voltage is excessive. The need for operation at 24 volts or more presents a deterrent to the designer of 12 volt land-mobile equipment.

Digital techniques have displaced analog ones with respect to tuning and frequency readout in receivers and transmitters, and this technology has allowed computer interface, memory channels and video display of CW and RTTY signals. Readout resolution has moved from 5 to 10 kHz increments to Hz in the past decade. Synthesized LOs have ensured a form of short and long-term stability that we never experienced in the analog days!

Gigantic strides have been made in the RF IC technology, particularly with regard to MMICs. Gain blocks for instrumentation and other applications are now possible at moderate cost and with minimum component count because of MMICs. High-frequency op amps are now available, and this has taken us well above the previous 1 to 3 MHz op amp range. Doubly balanced mixer ICs, such as the Motorola MC1496, helped lead the way to improved receiver performance.

VVC (voltage-variable capacitance) diodes have reduced the cost and bulk of VFOs and VCOs by eliminating the need for large mechanical tuning capacitors. There has also been a significant improvement in high-speed switching diodes and varactor power diodes.

The list of technological advances is certainly much longer than this, but I am pleased to share with *RF Design* readers my thoughts on those areas that have affected my own design and marketing interests. rf

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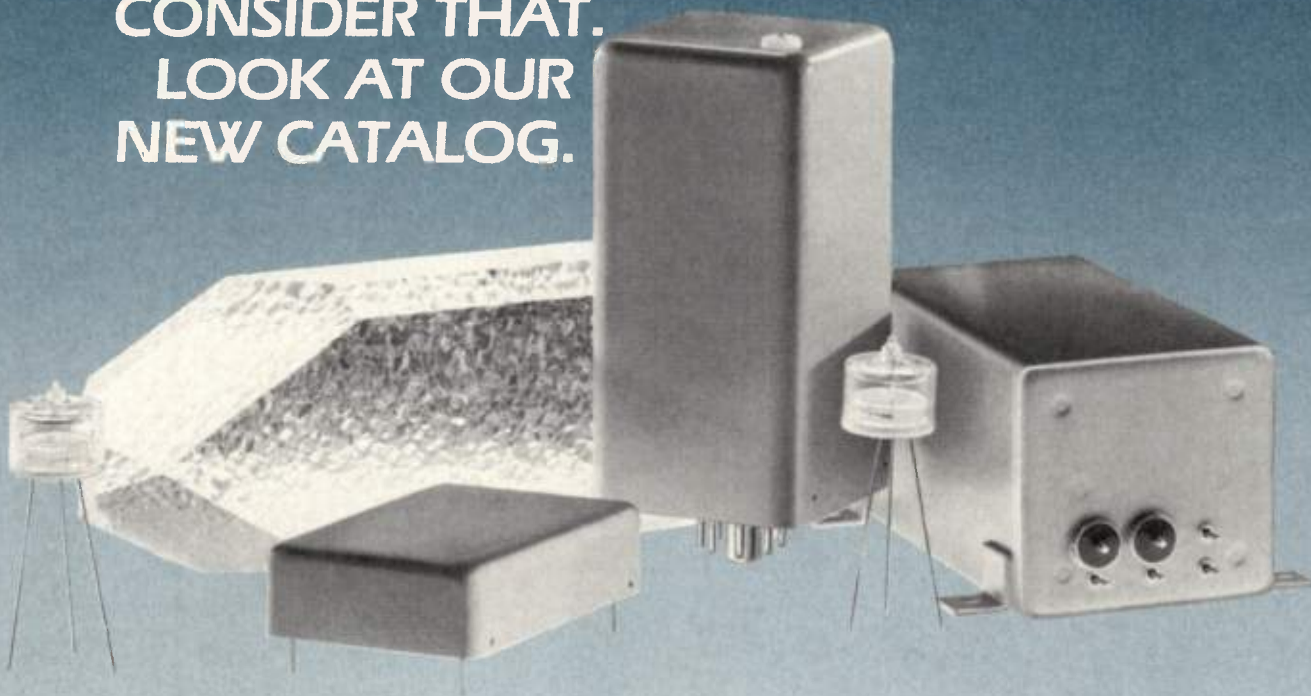

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A Voltage-Controlled Phase Shifter

Contest Winner Provides Design Flexibility for Phase-Sensitive Designs

By Thomas R. Mathews
Wavetek RF Products, Inc.

Many of today's complex systems require the ability to control the phase relationship of two signals. Typical applications include delay line descriminators, phased antenna networks and oscillators. Quite often these systems are realizable with fixed delay networks, but for many applications there is a need to provide a voltage-controlled phase shift. The design presented here won the author a runner-up prize in the Third Annual RF Design Awards Contest.

Traditionally, the problem of adjusting the phase of an RF circuit has been approached with variable length lines often resembling a trombone slide. Engineers have also been known to vary phase by cutting many lengths of coaxial cable and substituting until the desired phase is obtained. This approach is tedious to say the least and does not allow the phase to be controlled dynamically. A new approach to this problem uses varactor diodes to provide continuously variable phase control.

The ideal voltage-controlled phase shift

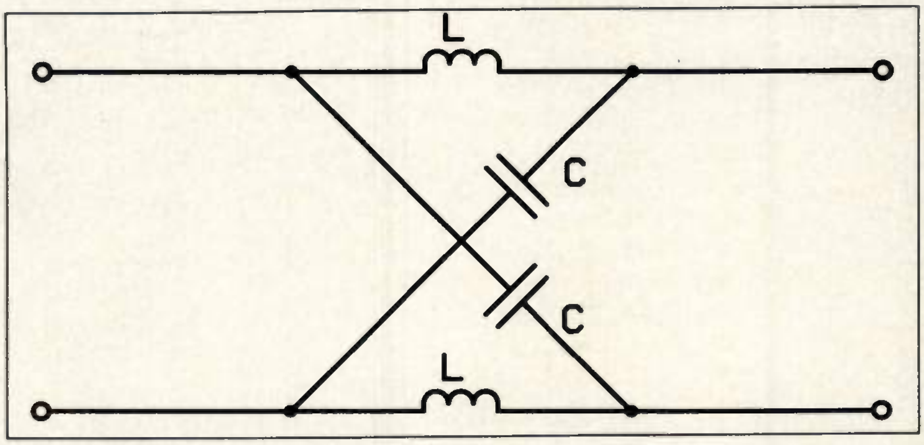

Figure 1. Phase shift network.

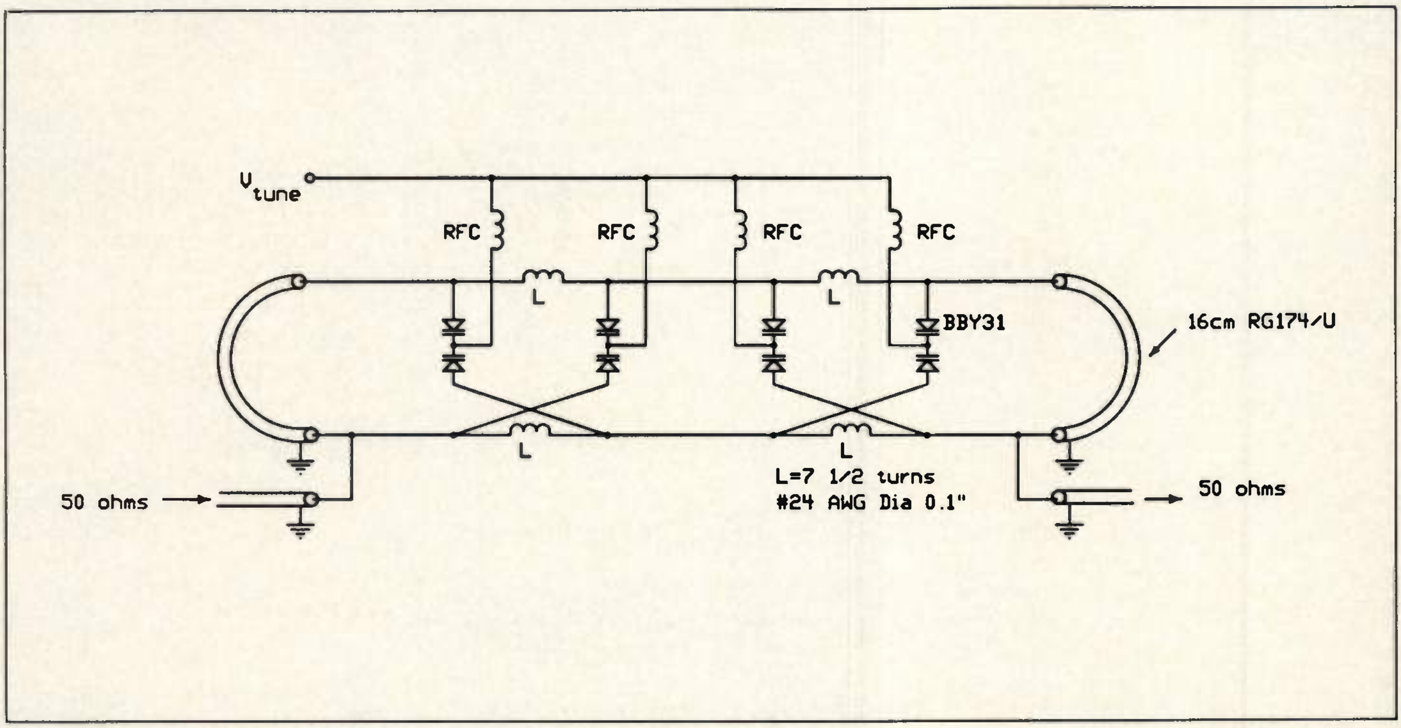

Figure 2. Final circuit with input and output baluns.

(VCPS) would have several desirable characteristics. Obviously, these would include a wide range of phase control, minimal variation of insertion loss with control voltage, and a flat magnitude response. The range of phase control is, in part, determined by the selection of the tuning diodes. Varactor selection is not critical. However, for a wide range of phase control, the varactor's capacitance should be in the range given by equation 1.

$$C \approx \frac{1}{(R\omega)} \quad (1)$$

Substituting R = 200 ohms and $\omega = 2\pi$ (500 MHz) into equation 1 indicates that C should be in the range of 1.6 pF. In this design, the BBY31 will be used. These diodes will typically provide 8 pF at 6 volts and since two are used in series the capacitance will be 4 pF. In addition, these diodes have about a 4:1 capacitance ratio over the 2 to 12 volt range. This should result in a wide range of phase control.

To insure a flat magnitude transfer characteristic, the phase shift network shown in Figure 1 was chosen. This network is also known as an all-pass filter or delay equalizer. Since this is a balanced circuit, baluns were included on the input and output of the final design (Figure 2). Transmission line baluns were chosen because they are easily realizable at UHF frequencies. For lower frequency designs, a transformer balun could be substituted for wider bandwidth and reduced size while the transmission line baluns should be one half wavelength at the VCPS center frequency. For RG174/U the velocity of propagation is about 200×10^6 m/s. The balun length is found with equation 2.

$$l = \frac{(200 \times 10^6)}{(2f)} \quad (2)$$

At 500 MHz, l is 20 cm. The actual circuit operated over the desired range of 450 to 550 MHz when this balun was trimmed to 16 cm. The 4 coils shown in Figure 2 are all the same value and were calculated with equations 3 and 4.

$$\alpha_o = \frac{1}{RC} \quad (3)$$

$$L = \frac{R}{\alpha_o} \quad (4)$$

The value for C is assumed to be fixed and is the geometric mean of the range

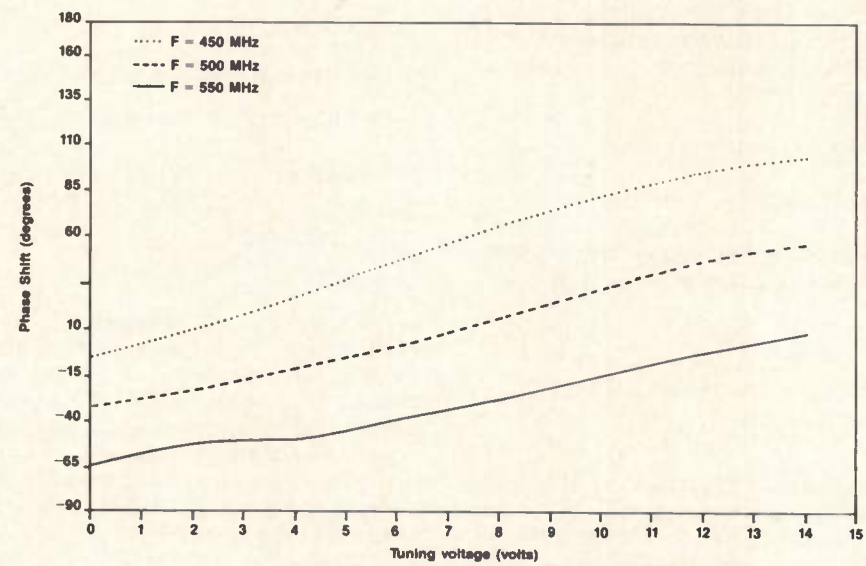

Figure 3. Voltage-controlled delay.

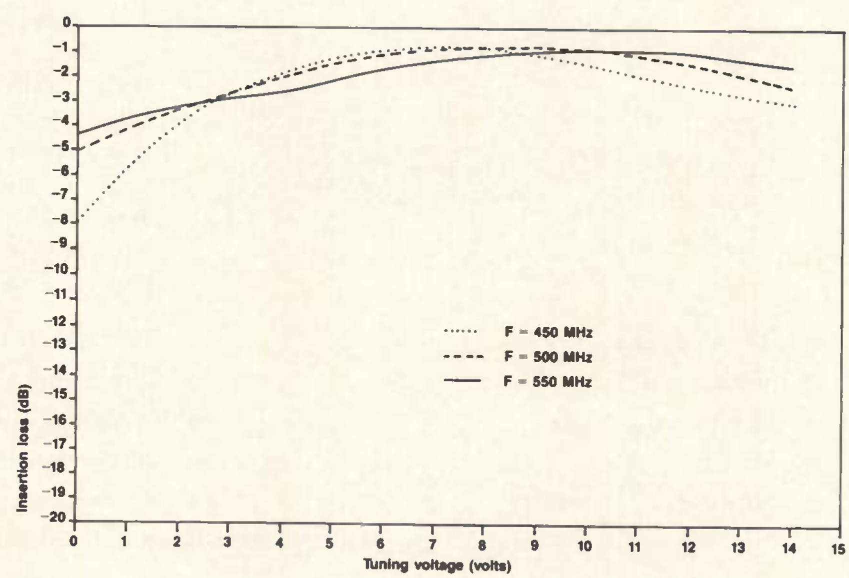

Figure 4. VCD insertion loss.

of capacitance provided by the varactors. For two BBY31s in series this is about 4 pF (The BBY31 provides about a 4-16 pF range from 2 to 12 volts.). Since the transmission line baluns provide a 4 to 1 impedance transformation, the value for R is four times the input impedance or 200 ohms. Evaluating equations 1 and 2 results in the values for α and L shown here.

$$\alpha_o = 1.250 \times 10^9 \text{ rad/sec}$$

$$L = 160.0 \text{ nH}$$

The phase shift through a single stage can be calculated with equation 5.

$$\phi = -2 \arctan\left(\frac{\omega}{\alpha_o}\right)$$

where $\omega = 2\pi f$

The amount of phase control through this stage can be estimated by calculating the network phase shift at the extremes of the varactor capacitance and taking the difference. This difference is found by finding α_o at varactor capacitances of 2 and 8 pF. Since the network impedance deviates from 200 ohms at these extremes, one should use equations 6 and 7 for this estimate.

$$\alpha_o = \frac{1}{LC} \qquad (6)$$

$$\alpha_{01} = 1.768 \times 10^9 \text{ for } C = 2 \text{ pF}$$

$$\alpha_{02} = 883.9 \times 10^9 \text{ for } C = 8 \text{ pF}$$

$$\Delta\phi = -2 \arctan\left(\frac{\omega}{\alpha_{01}}\right) + 2 \arctan\left(\frac{\omega}{\alpha_{02}}\right) \qquad (7)$$

For $f = 500$ MHz

$$\Delta\phi = 0.477$$

Since the all-pass networks are constant impedance, the amount of phase control can be easily increased by cascading several stages. Since this design uses two stages, the total estimate for phase control at 500 MHz is 54.6 degrees. The actual circuit provided 70 degrees of control at 500 MHz over a 2 to 12 volt tuning range (Figure 3). The performance of this circuit is limited to the 450 to 550 MHz range. This bandwidth restriction is due primarily to the limited bandwidth of the transmission line baluns and could no doubt be improved with lumped baluns.

Insertion loss is shown in Figure 4. As long as low varactor voltages are avoided, this loss is less than 3 dB. More important than the absolute insertion loss is the variation of loss versus tuning voltage. Figure 4 shows that, for the 2 to 12 volt range, this variation is less then ±1.0 dB. rf

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About the Author

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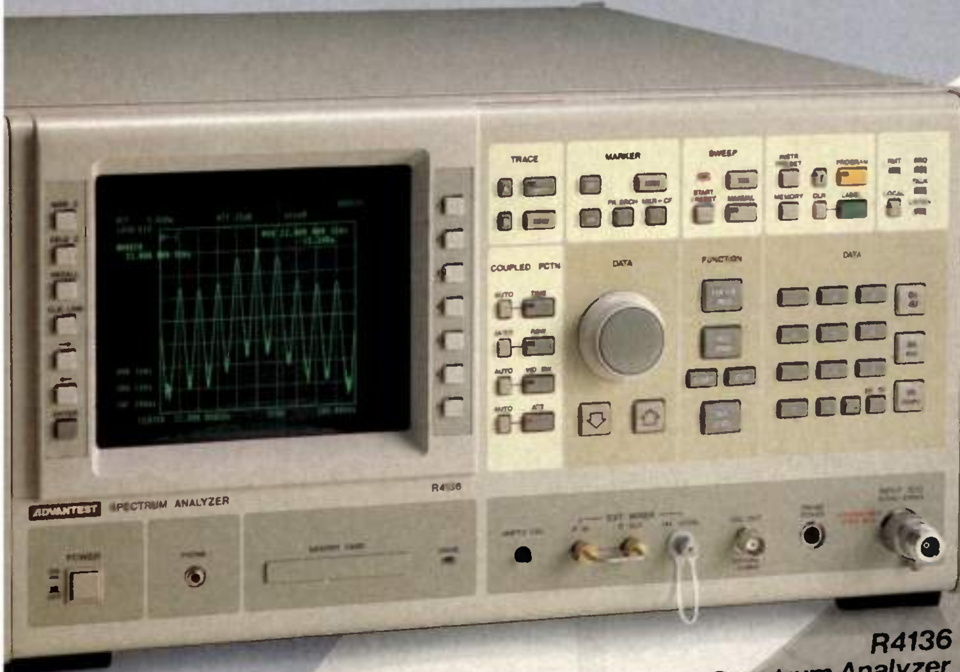

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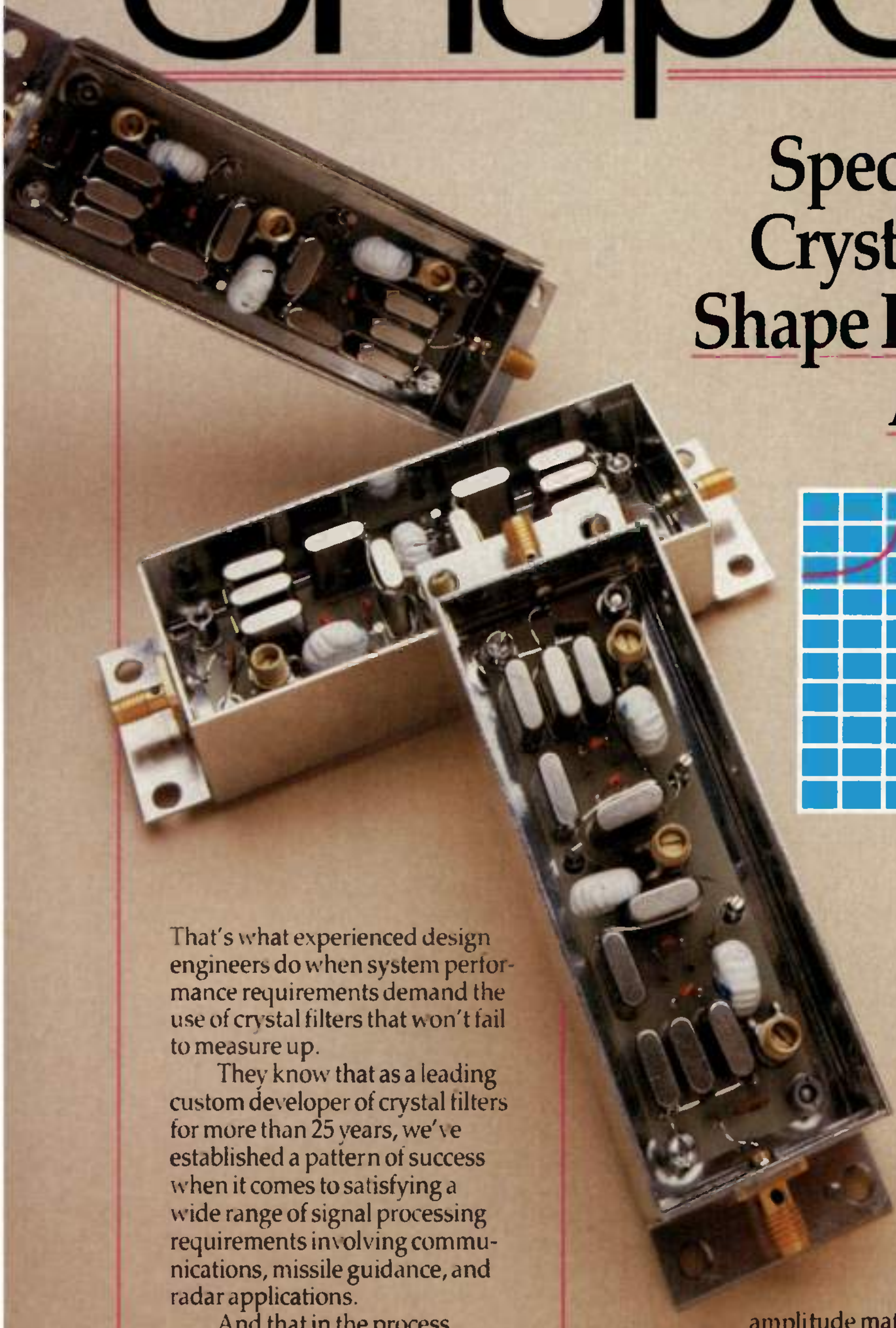

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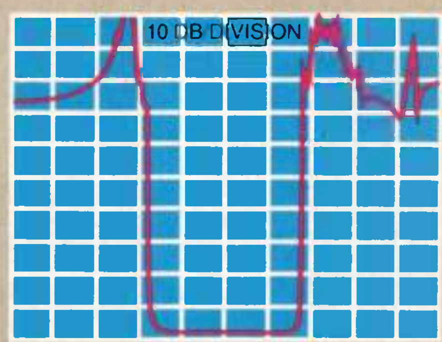

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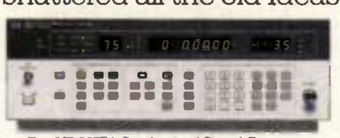

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Designing with the Double Lange Coupler

Contest Prize Winner Analyzes Interdigitated Design

By Derek Fitzgerald
Raytheon M.S.D

The interdigitated broadband 3 dB Lange coupler overcomes the problem of manufacturing repeatability associated with minute gap parallel edge coupled microstrip line structures. However, the coupler is of the quadrature type which implies that a signal applied to the input port will emerge from the coupled output port with a 90 degree phase shift relative to the input port. A novel hybrid-coupler design adaptation realizes a variety of wideband components. These include a 180 degree Lange type hybrid, an in-phase hybrid, an arbitrary phase hybrid and an adjustable phase hybrid.

One design adaptation employs the simple means of connecting two Lange couplers together in the manner shown in Figure 1. The resulting 180 degree Lange type hybrid has desirable properties, such as:

a. broadband frequency coverage
b. compactness of size
c. planar geometry
d. compatibility with existing hybrid and monolithic microwave integrated circuit (MMIC) technology
e. adaptability to balanced hybrid application in push-pull amplifiers
f. simplification in receiver and mixer applications that use quadrature hybrids in conjunction with phase shifting networks
g. simple planar realization of balanced mixer, balanced duplexer and discriminator circuits

A simple theoretical explanation of the coupler's operation follows. First, consider a perfect 3 dB quadrature coupler shown in Figure 2a. An input signal a_1 transmits half its power to port 2 and couples half its power to port 3. No signal flows to port 4. Also shown, is a flowgraph for this coupler with the input assumed matched. The voltage coupling relationship to port 2 and port 3 are:

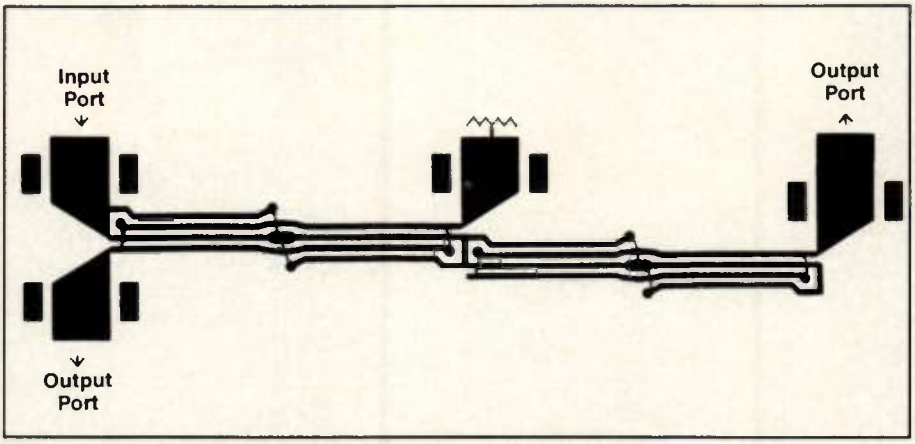

Figure 1. Physical layout of 180 degree hybrid.

$$\frac{b_2}{a_1} = \frac{1}{\sqrt{2}} \quad \text{and} \quad \frac{b_3}{a_1} = j\frac{1}{\sqrt{2}}$$

Now consider the flowgraph in Figure 2b where port 2 and port 3 end in open circuits. The input signal reflects at ports 2 and 3 and recombines constructively at port 4. The voltage coupling coefficient at port 4 is:

$$\frac{b_4}{a_1} =$$

$$\frac{1}{\sqrt{2}} \cdot (+1) \cdot j\frac{1}{\sqrt{2}} + j\frac{1}{\sqrt{2}} \cdot (+1) \cdot \frac{1}{\sqrt{2}} = j$$

Using this analysis, consider the configuration in Figure 2c consisting of two cascaded hybrid couplers. As illustrated, two of the ports connect to open circuits. The coupling arm of the input hybrid connects to the input port of the second hybrid. The direct path and coupled outputs of the second hybrid end in open circuits. The input/output voltage coupling relationships are given by:

$$\text{and} \quad \frac{b_2}{a_1} = \frac{1}{\sqrt{2}}$$

$$\text{and} \quad \frac{b_3}{a_1} = \frac{j}{\sqrt{2}} \cdot j = -\frac{1}{\sqrt{2}}$$

Thus, b_2 and b_3 are of equal amplitude and are out of phase by 180 degrees. Note the result of connecting two 3 dB quadrature hybrids together in general, and not restricted to just Lange-type hybrids. The concept is applicable to in-phase hybrids, arbitrary phase hybrids and adjustable phase hybrid coupling structures.

Simple analysis shows that if short circuit elements replace the open circuit terminations, in-phase coupling will result. Such in-phase Lange type hybrids require

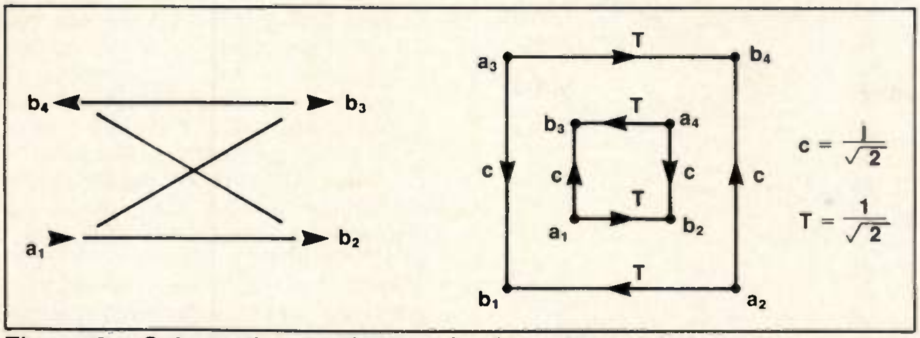

Figure 2a. Schematic and flowgraph of quadrature hybrid.

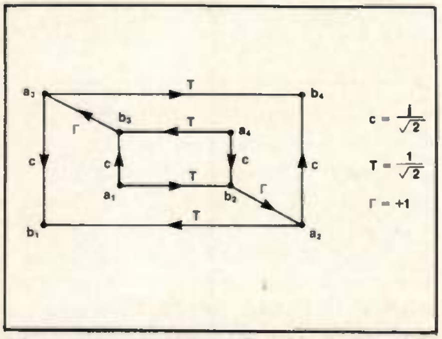

Figure 2b. Flowgraph of hybrid with two ports terminated with open circuit elements.

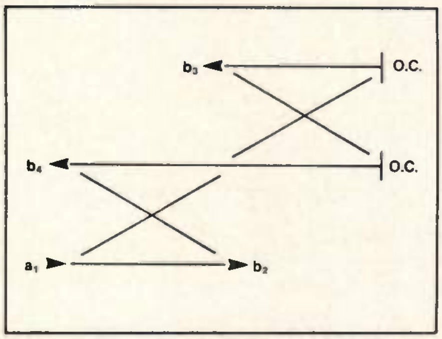

Figure 2c. Two hybrids cascaded.

metallized via holes or wraparounds. A simpler realization uses the same open circuit termination structure as the 180 degree type hybrid, except the input signal latches at port 4 of the first coupler. Thus, the direct arm of the input hybrid connects to the input of the second hybrid. Note the output signals are in phase or out of phase depending on the input signal launching port. So, the structure has magic-tee junction properties.

Extending the concept, reactive or adjustable reactive elements such as diodes or ferrites can replace the open circuit elements. It is easy to show that the phase difference between the two outputs is equal to the phase of the reactive elements. Therefore, arbitrary phase and adjustable phase hybrid designs are realizable with this structure.

Figure 3 diagrams the interdigitated topology of the standard Lange coupler and the reverse Lange coupler (where the coupling and isolation ports are on one side). Jumpers, which are normally thermo-compression bonded, connect the finger structure. Practical in-phase and 180 degrees phase hybrids use both types of couplers as shown in Figure 4. The computer simulation results, shown

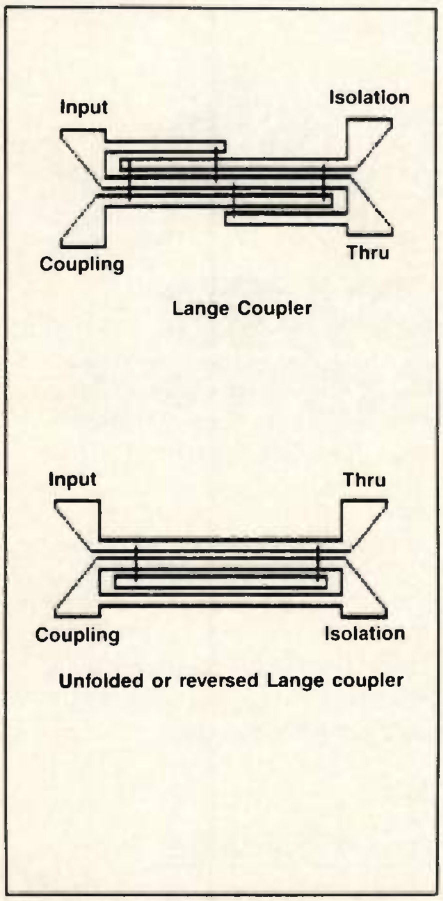

Figure 3. Lange coupler types.

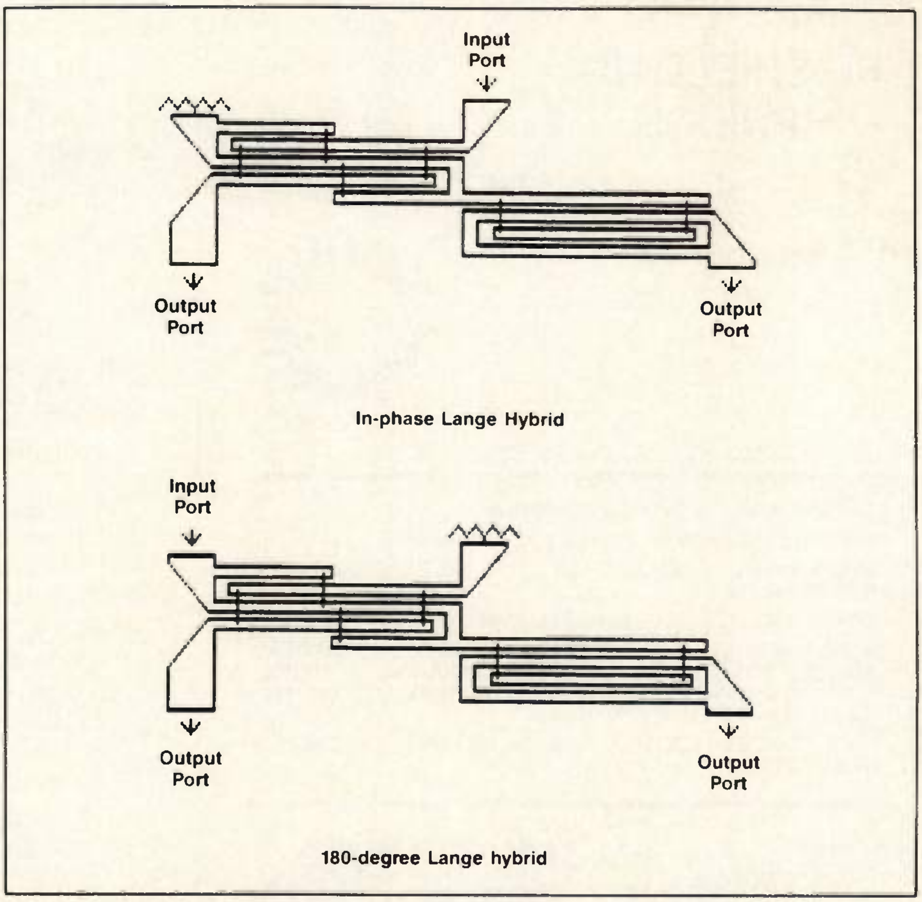

Figure 4. In-phase and 180 degree hybrids.

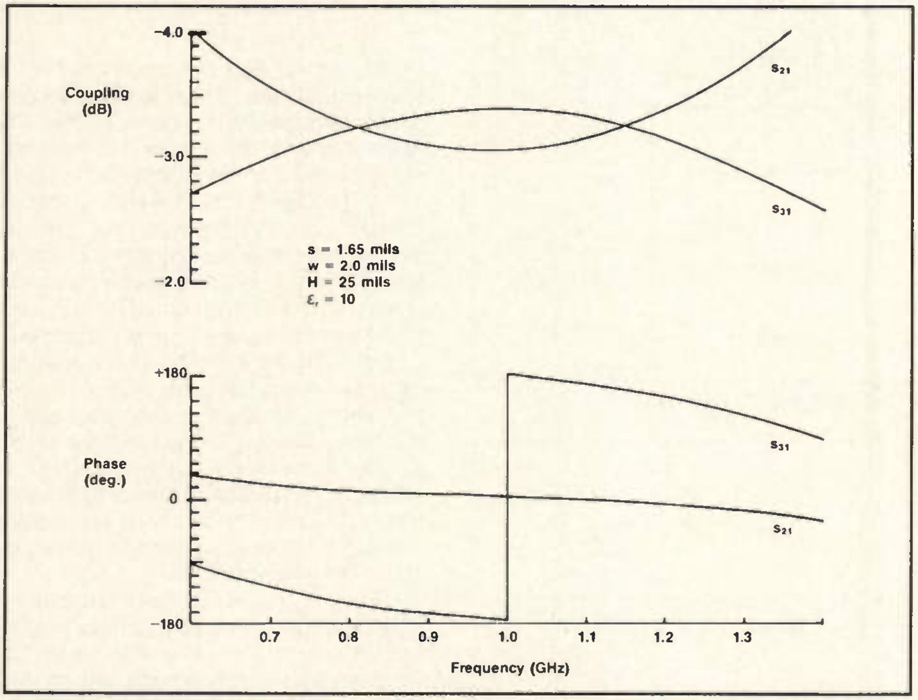

Figure 5. 180 degree Lange coupler characteristics using Supercompact simulation.

in Figure 5, illustrate the performance characteristics of the 180 degree Lange coupler. While the circuit is not optimized, the data clearly demonstrates the potential capabilities of this component.

In conclusion, simple theory and computer-aided simulation show an extension of the Lange quadrature hybrid capabilities. The evolved component has magic-tee-like properties and is attractive for implementing a general class of broadband hybrid and MMIC applications. These include push-pull amplifiers, balanced mixers, balanced modulators and other circuits requiring baluns structures.

The author gratefully acknowledges the constructive editorial comment of his engineering colleagues Al Hieber and Zvi Galani. rf

About the Author

Derek Fitzgerald is principal engineer at Raytheon Missile Systems Division, 50 Apple Hill Drive, Tewksbury, MA 01876-0901. He can be reached at (508) 858-4242. This design received an honorable mention in the 1988 RF Design Awards Contest.

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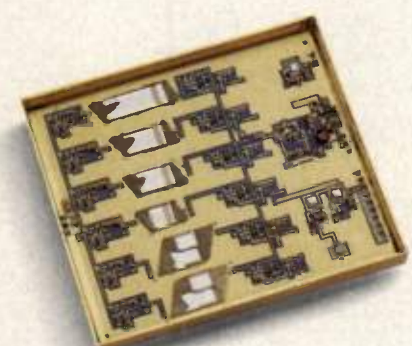

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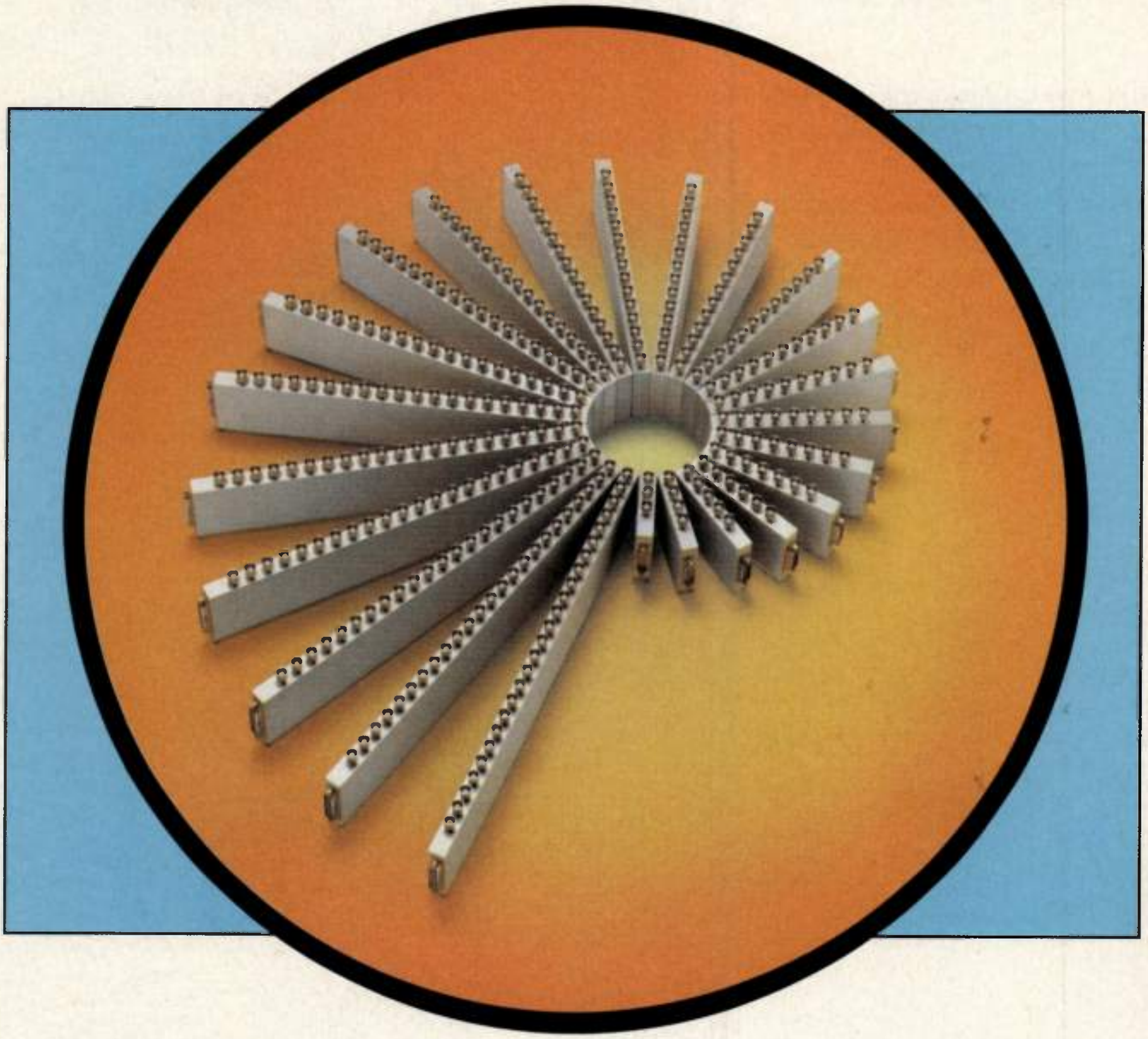

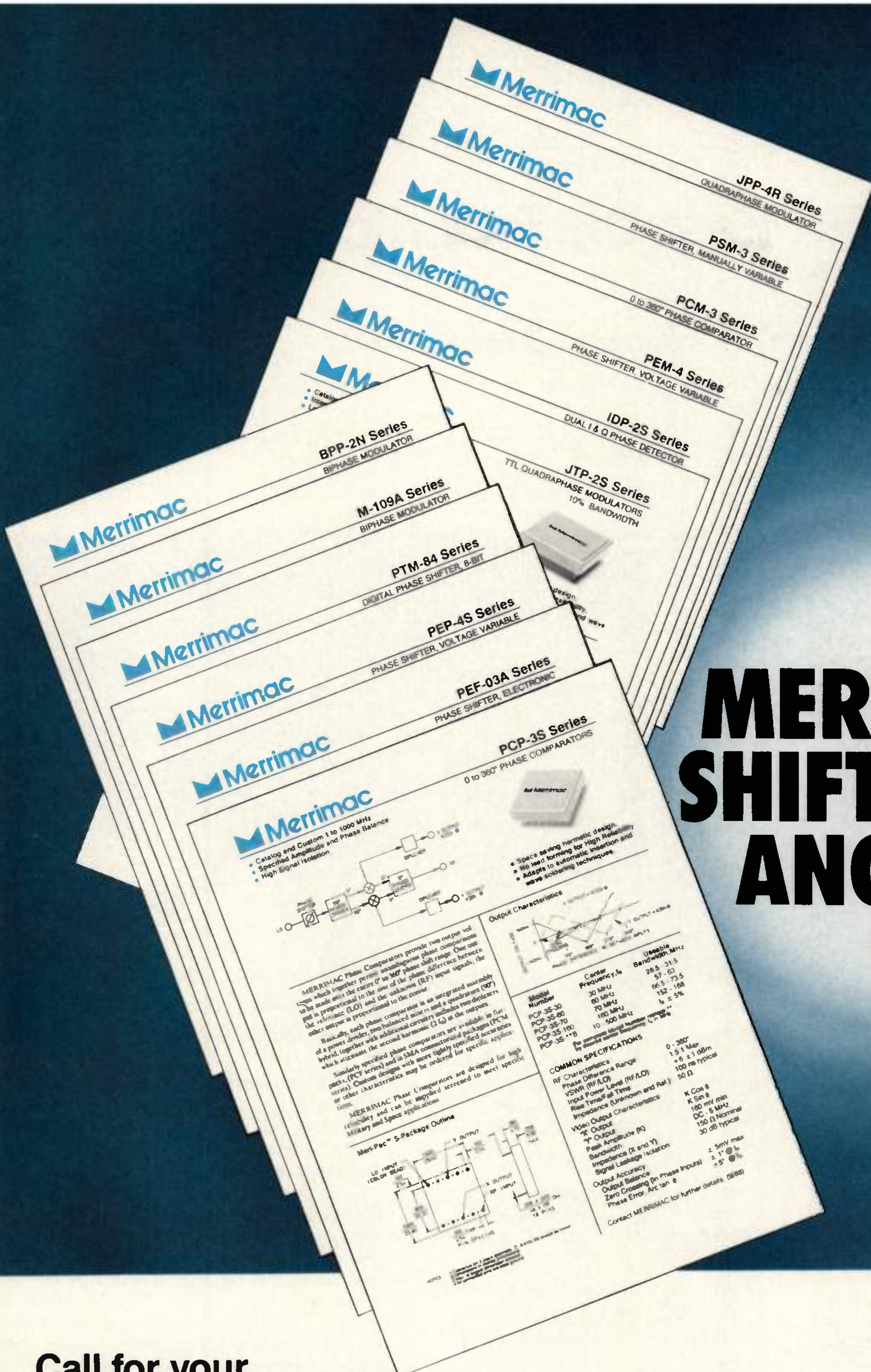

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The RF Expo East 88 Technical Program

A Wide Variety of Topics Offers Something For Everyone

RF Expo East 88, October 25-27 at Philadelphia's Civic Center, offers RF engineering professionals an opportunity to exchange ideas, solve problems and brush up on basics. Tutorials on radiowave propagation, mixer design and antenna theory head the list of more than 35 technical sessions directed specifically at RF engineering concerns. Under the direction of Program Chairman Carl Erikson, the technical program for RF Expo East represents a relevant and diverse collection of papers presented by top experts from across the industry.

Tuesday — 8:30 to 11:30 a.m.

Session A-1 — SAW Devices
Chair: Emmanuel Sang, Tektronix, Inc.

Real-Time SAW Spectrum Analyzers Appropriate for Tactical Use
Maura Fox, Phonon Corporation

System application and design information is presented for the Sliding Transform Analyzer (STA), a Surface Acoustic Wave (SAW) spectrum analyzer configured to perform a partial Fourier Transform. The STA is particularly well suited to use in a tactical system due to excellent wideband frequency resolution.

SAW Products: Custom Subsystems and Fully Integrated Hybrid Modules
Ron Hays, T. O'Shea, J. Andersen and M. Lewis, SAWTEK

SAW components play critical roles in modern systems due to their reliability, size and performance advantages. This paper addresses the challenge to SAW manufacturers of achieving the highest performance in systems that employ SAW components. Topics discussed include complete SAW signal processing subsystems, the need for miniaturization and integration, and innovations in packaging such as fully integrated hybrid assemblies.

Selectable Delay Line Using an Acoustic Charge Transport Device
Dan Fleisch, Electronic Decisions, Inc.

Session A-2 — Receiver Design
Chair: Malcolm Levy, Racal-Dana Instruments, Inc.

Receiver Design — Simplified
Jerry Iseli, RF Monolithics, Inc.

This paper discusses the use of SAW filter technology to simplify receiver systems, resulting in lower cost at reduced size and power consumption. Specifically, a two-channel Guard Receiver developed by RF Monolithics using this technology is presented.

GPS Receiver RF Subsystem Design Overview
B.D. Hammell, G.K. Nelson and J.G. Klopfenstein, Rockwell International

An overview of the Global Positioning System (GPS) is presented, with a description of the Rockwell GPS receiver RF subsystem architecture, including an antenna, antenna electronics and receiver signal processing unit.

Optical/Acoustic Component Technologies for Frequency Sorting Receivers
Richard V. Johnson, Crystal Technology, Inc.

In electronic warfare, channelized microwave receivers must function successfully in a spectrally dense environment. They can do so by dividing the RF spectrum into a sufficiently large number of narrow channels, each of which is independently processed and detected. Advantages and limitations of several alternative analog component technologies, including SAW filters, acoustooptic Bragg cell spectrum analyzers, integrated optic (I/O) versions of Bragg cell spectrum analyzers, bulk acoustic wave (BAW) filters and an electrooptic spatial light modulator are reviewed.

Session A-3 — Design Techniques
Chair: Ted Dudziak, Wavetek RF Products, Inc.

Optimum Impedance Matching Using Minimum Number of Components
Donald J. Lanzinger, Loral Defense Systems Division

Impedance matching can significantly impact systems parameters. This paper investigates impedance matching circuitry design with respect to bandwidth (and cut-off characteristics), power transfer, group delay, types of components (lumped and/or distributed) and minimum number of components needed.

Complex Signals: Their Processing and Transmission
Noel Boutin, University of Sherbrooke

Complex signals are omnipresent in modern analog and digital communication systems. This tutorial paper discusses what a complex signal is, how to filter a complex baseband signal and how to shift frequency in a complex signal. Quadrature Amplitude Modulation (QAM), Single-Sideband (SSB), Vestigial-Sideband (VSB) and Phase/Frequency Modulation (PM/FM) systems will be reviewed highlighting the presence of complex signals in each case.

Unequal Splitter Design
Douglas K. Linkhart, Micon Incorporated

RF power distribution systems and phased-array antennas require power splitters which provide a specific fraction of the input power at each output. A procedure for designing low-loss splitters, based on quarter-wavelength impedance transformers, is presented, along with supporting theory and a computer program for design calculations.

Tuesday — 1:30 to 4:30 p.m.

Session B-1 — Synthesizer Topics
Chair: Dave Badger, Wavetek RF Products, Inc.

Optically Coupled VHF Voltage Controlled Oscillator
Albert Helfrick, Dowty RFL Industries, Inc.

Voltage controlled oscillators (VCOs) for frequency synthesizers must be well isolated to prevent the VCO from being frequency modulated by external fields or from the effects of varying loads. This paper describes an interesting way of eliminating load pulling while also eliminating the last unfiltered connection to a shielded oscillator — the output cable.

Low Spurious Techniques and Measurements for DDS Systems
Robert J. Zavrel, Jr. and Earl W. McCune, Jr., Digital RF Solutions

A major design objective in direct digital synthesizers (DDS) is the minimization of spurious signal responses between the DC and the Nyquist frequency. This paper reports on research aimed at this goal with particular consideration given to quantized spurious limitations of the waveform map and to DAC linearity limitations.

A Hybrid FSK Modulator for Use in RF Signal Generators
John Vendely,
Wavetek RF Products, Inc.

This paper describes a hybrid carrier frequency-shift keying (FSK) modulator designed primarily for use in RF signal generators used in testing land-mobile digital signalling systems. A comparison of the traditional methods, a description of the modulator and performance test results are given.

Session B-2: Reliability and Component Performance

Failure Analysis of RF Devices
Andrew Blackwood, Structure Probe

Determination of the M.T.T.F. for Pulsed Power L-Band Microwave Transistors Used in Radars and Navigation Apparatus
Georges Bobin, RTC and
John C. Salvey,
Amperex Electronic Corporation

RF Expo East 88 Offers Special Courses

RF Circuit Design Part I: Fundamentals
Les Besser,
Besser Associates, Inc.
Monday, October 24,
8:00 a.m. to 5:00 p.m.

This popular course in RF design fundamentals begins with an introduction to high frequencies. Lumped and distributed components, parasitic effects, and use of the Smith chart are discussed, as well as theory and various physical realizations of transmission lines, with emphasis on practical design aspects. Small signal amplifier design is also reviewed, with the afternoon session covering unilateral and bilateral design using constant gain circles; computer-aided synthesis and optimization; stability consideration and more.

Computer-Aided Filter Design
Randy Rhea, Circuit Busters
Monday, October 24,
8:00 a.m. to 5:00 p.m.

This course covers L-C filter design and analysis, with emphasis on principles and solving practical problems using the personal computer. The historical development of filter design introduces the course, followed by discussions of the modern lowpass prototype, its transformation to highpass, bandpass and bandstop configurations, and the transfer function amplitude and delay characteristics of each. Additional topics include elliptical transfer function filters, determination of required filter order, effects of component Q and parasitics, filters as matching networks and distributed (transmission line) filters.

RF Circuit Design Part II: Advanced Topics
Les Besser,
Besser Associates, Inc.
Tuesday, October 25,
8:00 a.m. to 5:00 p.m.

A continuation of RF filter design, using structures beyond those introduced in the first course. Modern filter design, including computer-aided synthesis, is also covered. Impedance transformations are extended to broadbands, using both lumped and distributed elements and complex terminations. The small signal amplifier design is also extended to broadbands, and impedance matching and feedback methods are discussed. Large signal amplifiers of Class-B and -C types are evaluated through the latest CAD tools. Other nonlinear circuits, such as oscillators, mixers and frequency multipliers are introduced; their behaviors are examined both in the frequency and time domain.

An additional fee of $185 is charged for each course, and advance registration is strictly required. Registration for a second course can be obtained for $135.

Are the limits placed on junction temperature by various government agencies reasonable for a new generation of radar equipment? Classically, junction temperature has been related to M.T.T.F. of RF and microwave transistors. This paper details evidence that many factors influence the junction temperature, and that all these factors have direct consequences in predicting the M.T.T.F. of pulsed L-band microwave power transistors.

RF Components at Low Temperatures
Otward Mueller,
General Electric Company

It is a well-known fact that the properties of many electronic components improve if they are cooled. This paper presents a review of these properties and includes a discussion of electronic, thermal and magnetic conductors. An experiment with a 2-kilowatt RF linear pulse power amplifier is described.

Session B-3: New RF Components Chair: Gene Niemec, Merrimac Industries, Inc.

TV Linear Power Transistors: 150 Watts, Class AB, Band IV-V
Serge Juhel,
SGS-Thomson Microelectronics

This paper covers recent advances in TV Linear Power Transistors and the die characterization data required to optimize the device parameters for best application performance.

A Silicon RF Amplifier Using a Dielectrically Isolated Monolithic Microwave Integrated Circuit (DIMMIC) Process
P. Bachert, Michael McCombs, Paul Sanders,
Motorola Semiconductor Product Sector

A three-stage silicon monolithic RF power amplifier is described. Performance characterization, including a comparison to a thin-film hybrid discrete prototype, is discussed. Also described is the dielectrically isolated high frequency silicon bipolar process used to fabricate the amplifier.

Applications and Measurement of Small Value Inductors in SMT and Hybrid Circuits
Richard Dunlap, Micoil

The small values of inductance and physical size of chip inductors present many application problems in today's RF circuit design, fabrication and assembly. This paper details some of the chip inductor fabrication, measurement and assembly problems and presents discussion and analysis of some solutions. Specific topics regarding measurement, fixtures, physical mounting and reliability will be evaluated.

Wednesday — 8:30 to 11:30 a.m.

Session C-1: Radiowave Propagation Tutorial
Daniel R. Dorsey, Jr.,
DocSoft Enterprises

This session presents an introduction to basic principles and characterizations involved in radio wave propagation. Discussion includes a study of VLF propagation, especially seawater propagation and submarine communications. This is followed by a look at HF propagation, with an examination of the structure and characteristics of the earth's atmosphere and ionosphere and an introduction to several "exotic" modes of propagation. The final section of this session covers propagation at VHF and above, with a consideration of atmospheric effects and satellite communications.

Session C-2: Quartz Crystal Applications Chair: Colin Lanzl, Alpha Industries

Fundamentals & Stability of Crystal Resonators
John Vig,
U.S. Army Electronics Technology & Devices Laboratory (LABCOM)

A Review of Crystal Filters
William Pond,
McCoy Electronics Company

The three major parts of filter design require the definition of a specification, the development of a suitable approximation and the realization of a circuit. This paper presents a discussion of these three items and how they apply to crystal filters, as well as what a crystal filter consists of, why crystal filters are used and what their limitations are.

Practical Considerations in Specification of High Stability Crystal Oscillators
Glenn R. Kurzenknabe,
Piezo Crystal Company

Problems for the user and manufacturer of high-stability crystal oscillators can be minimized by increased understanding of parameters affecting design and cost. This paper addresses system design considerations and critical performance parameters for oscillators utilizing "SC" cut crystals in the 5-10 MHz and 90-110 MHz regions.

Session C-3: Computer-Aided Design
Chair: Jack Koscinski, General Optronics Corporation

CAD Package for RF Applications
Alec Clerihew, Hazeltine Corporation

A Surface Acoustic Wave Filter Computer-Aided Design Workstation
Sam M. Richie, Benjamin P. Abbott, Carlton D. Bishop and Donald C. Malocha, University of Central Florida

This paper presents a description of "SAWCAD," a surface acoustic wave computer-aided design workstation system implemented on a personal computer class hardware system. A description of device modes, analysis techniques and CAD design structure will be discussed.

High-Speed Printed Circuit Board Analysis and Simulation in a Workstation Environment
Joseph Hall, Sasan Ardalan, M.S. Basel, Donna O. Riddle and Michael B. Steer, North Carolina State University; Real Pomerleau, Bell Northern Research

Performance of digital systems on printed circuit boards (PCBs) is currently limited by transmission line effects rather than the technology of digital devices. This paper presents a review of transmission line theory, a technique for recursively solving transmission line networks and a review of CAPNET, the CAD tool for complex transmission line networks.

Wednesday — 1:30 to 4:30 p.m.

Session D-1: Elements of Antenna Theory
Chair: Gary Breed, *RF Design*

Benjamin Rulf, Lockheed Electronics Company, Inc.

The purpose of this tutorial is to explain in an elementary way some basic concepts and factors of antenna theory. Following a review of the basics of wave theory, there are sections on wire antennas, small aperture antennas, antenna arrays and reflector antennas. The session is intended for RF engineers who have had some exposure to electromagnetics and have a working knowledge of transmission lines and of basic engineering mathematics.

Session D-2: Power Amplifiers
Chair: Samuel J. Klein, Microwave Modules and Devices

Effects of VSWR Upon Class-E Power Amplifiers
Frederick H. Raab, Green Mountain Radio Research Company

The class-E power amplifier achieves high efficiency at high frequencies by using the output-filter network to discharge the drain-shunt capacitance immediately prior to switching on the transistor. This paper investigates the effects of VSWR upon the power, efficiency and stress

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characteristics of an ideal class-E RF power amplifier through numerical evaluation.

RF Hybrid Linear Power Amplifier With Diamond Heat Sink
Nafiz Karabudak, Aydin Vector Division

This paper addresses the design of a high power output RF linear small signal amplifier using diamond heat sink. The benefits of using diamond heat sinks will be reviewed and future trends and possible applications discussed.

Development of a Linear L-Band High Power Amplifier for Satellite Application
Gilles Brassard, Spar Aerospace Limited

Providing an economically viable mobile communications service via satellite necessitates onboard linear power amplifiers with good efficiency. This paper describes the requirements for and design of such an amplifier, capable of producing 50 watts of RF power with a multitone signal at an efficiency of 25 percent and with intermodulation products better than 17 dB below carrier.

Session D-3: Non-Linear Circuit Simulation Chair: Michael K. Ferrand, Microlab/FXR

Nonlinear Circuit Simulation for Microwave and High-Speed Circuits
P.K.U. Wang, K.T. Lin, M. Sango and C. McGuire, EEsof, Inc.

Circuit simulation programs such as SPICE have been available for nonlinear circuit design and simulation since the late 1960s and early 1970s. This paper presents approaches to making a generic nonlinear circuit program, such as SPICE, into a program for high-speed frequency applications. Techniques for improvements in the areas of equation formulation, numerical integration, device modeling and simulation data processing are discussed.

Simulation of Nonlinear RF and Microwave Circuits
C.R. Chang, P.L. Heron, Michael B. Steer, G.W. Rhyne, Donna O. Riddle and R.S. Gyurcsik, North Carolina State University

Nonlinear analysis methods for RF and microwave circuits can be classified as time domain, frequency domain or as hybrid (time and frequency domain) depending on how the linear and nonlinear elements are analyzed. Two promising techniques, harmonic balance methods and series methods, are reviewed and BJT and MESFET amplifiers are analyzed using harmonic balance (a hybrid method) and generalized power series analysis (a frequency domain method).

Thursday — 8:30 to 11:30 a.m.

Session E-1: Mixer Tutorial
Donald Steinbrecher, Steinbrecher Corporation

Fundamentals of mixer design using various configurations are presented in this tutorial. Emphasis is placed on the effects that design parameters and external influences have on mixer performance.

Session E-2: Manufacturing-Related Topics Chair: Lisa Boyd, McCoy Electronics Company

Special Requirements of RF Packaging
Rudy Sachs, ASPE, Inc.

A discussion of RF package design is presented with design solutions for meeting electrical performance criteria.

An Inexpensive Silicon Monolithic IC Process for Microwave Low and Medium Power Circuits
Paul Sanders, Motorola Semiconductor Products Sector

Transistors fabricated with dielectric isolation technology have long been recognized for their advantages over conventional junction isolated ICs. However, fundamental disadvantages such as low density limitations and device tub (collector region) thickness tolerance still exist. This paper describes and evaluates patented modifications to the fundamental isolation process aimed at minimizing these factors.

Simulation of Complex Coupled PCB Layouts with Non-Linear Digital Device Termination
Daniel Winkelstein, Real Pomerleau and Michael B. Steer, Bell Northern Research

Performance of digital printed circuit board technology is currently limited by transmission line effects of PCB tracks. This paper presents a numerical method for determining the transient response of arbitrarily complex transmission line systems terminated with digital devices. Experimental and analytic results are compared and guidelines for design and simulation presented.

Session E-3: Transmission Line Topics Chair: Robert C. Kane, Motorola, Inc.

Through Symmetric Fixture: A Two-Port S Parameter Calibration Technique
Jeffery S. Kasten and Michael B. Steer, North Carolina State University; and Real Pomerleau, Bell Northern Research

A straightforward procedure for the calibration of a measurement system with symmetric test figuring is described and experimentally verified.

Use of Pads and Baluns for TAHQ Alignment
Nathaniel Ersoz, Thomson Consumer Electronics

The topic of this paper is an analysis of a measurement technique used to evaluate performance of RF devices with a balanced input. The technique, involving a "balanced pad" as a restrictive device, is examined and found to be very poor.

The Design of a High Reliability All-Electronic Step Attenuator
Marcus Da Silva and David Whipple, Hewlett-Packard

This paper describes a step attenuator used in the HP 8657A Signal Generator, implemented using a resistive ladder network optimized for broadband operation, 50 ohm characteristic impedance and PIN diode switches.

rf design feature

An Emitter Follower Oscillator

By Harvey L. Morgan
TDC, Inc.

An emitter follower is generally not considered capable of oscillation since the input and output are in phase and output voltage is less than input voltage. However, since an emitter follower (EF) has power gain, emitter follower oscillation is possible. The EF oscillator described in this article is believed to be new. Its mode of operation is strange, to say the least, but it works quite well, and is presented for consideration by the readers.

A circuit being developed for broadband full-wave rectifier use was observed to oscillate at a 10 MHz rate at the zero-crossings of the output waveform. Since the output active elements were a pair of transistors with a common emitter resistor, that oscillation was an interesting phenomenon deserving further investigation. The full-wave rectifier consisted of a phase splitter driving the emitter followers with opposite polarities of the signal (Figure 2) so that when one EF was following a positive-going voltage, the other was following a negative-going voltage. With their emitters at the same potential at all times, conduction would be alternate. At cross-over, their collector currents and base bias would be almost identical, but not quite, due to difference in beta. Oscillation started slightly before cross-over and continued about an equal interval after cross-over. Oscillation could be suppressed by a capacitor shunting the common emitter resistor.

A normal EF has a steady-state current in the bias resistor modulated by the signal. When the transistor Q1 with the negative-going signal has its emitter pulled up by the other transistor (Q2), the base current of Q1 will decrease at the same time, reducing the bias resistor drop which would tend to pull the Q1 base up and renew conduction. The coupling capacitor-base resistance time constant appears to be the controlling factor.

The two EFs were disconnected from the phase splitter and a series capacitor and inductance was coupled between bases (Figure 3). An oscillation occurred in a restricted supply voltage range and with small amplitude. It was quickly

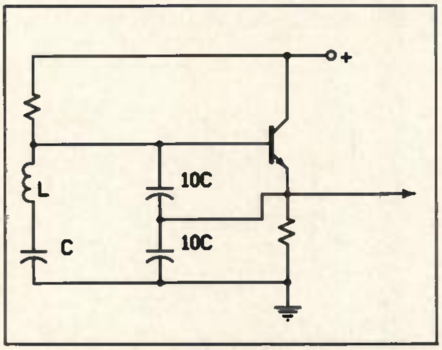

Figure 1. Transistorized Clapp oscillator (for reference).

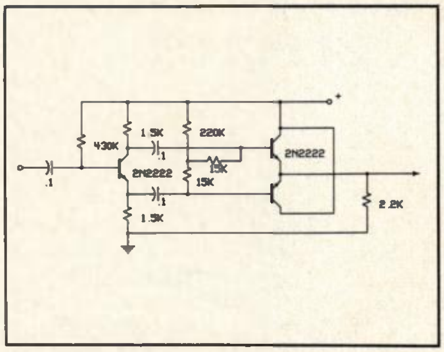

Figure 2. Broadband full-wave rectifier.

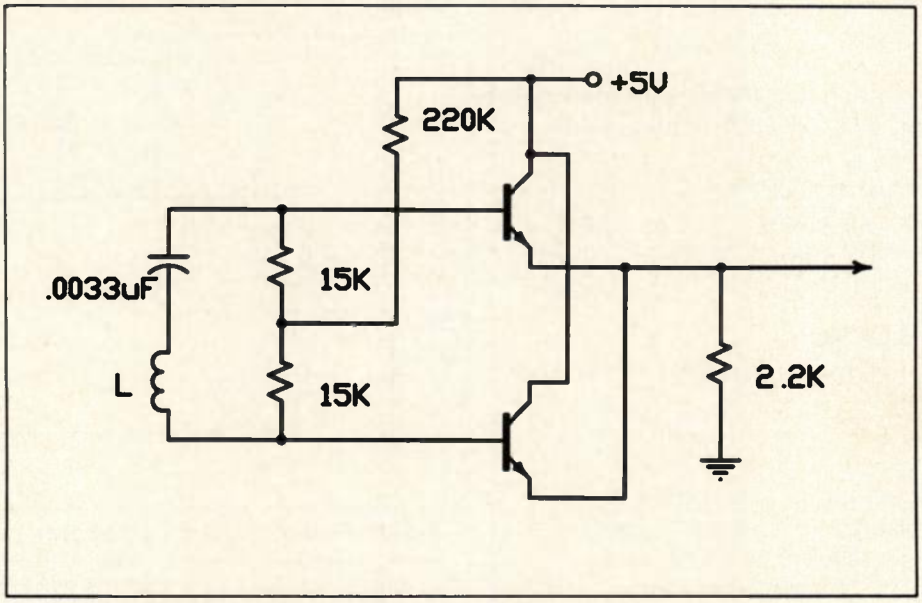

Figure 3. Initial form of emitter follower oscillator.

established that the oscillation was not a series resonance but rather a parallel resonance of the inductor with its distributed capacity and the input capacities of the EFs. Oscillation occurred at low voltages (on the order of 5 volts). A variable capacitor across the inductor allowed some frequency tuning, but too much capacity stopped oscillation.

RF grounding of one EF base (Figure 4) increased the amplitude of oscillation to on the order of the supply voltage (peak-to-peak) and the output waveform was a fair sine wave. The sensitivity to capacitive loading of the inductor remained. It was noted that the voltage developed across the RF-grounded base was a small asymmetrical sawtooth.

The output waveform was noted to swing above Vcc. The EF swinging positive at the time would have to be drawing enough base current to provide emitter output current under those conditions. The negative swing did not reach ground potential. Scope probe capacity was too great to observe the waveform on the base of the non-RF-grounded EF. Obviously, drive to the resonant circuit was provided by base current pulses alternately from the EFs.

To clarify what was happening, the inductor was split and the center-tap grounded (Figure 5). This worked from 16 MHz to about 2 MHz as the inductor value was varied. The waveform remained sinusoidal and equal to or greater than the supply voltage. The supply voltage sensitivity was greatly reduced (5 to at least 15 V). The extreme sensitivity to capacitive loading across the inductor remained.

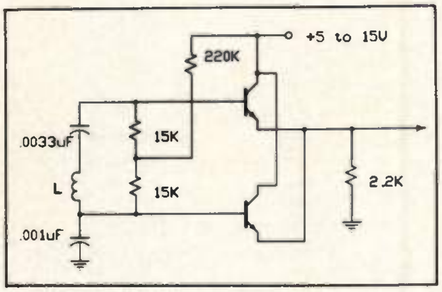

Figure 4. Grounded base EF oscillator.

However, it was found that a variable capacitor could be used to vary the frequency when shunted across the emitter resistor (Figure 6). The capacitive impedance reflected onto the resonant circuit by the transistor bases provided the tuning. The ratio of actual emitter capacity to effective capacity across the inductor was not the same as the beta of the transistors, as would be the case with resistive loading due to an EF. The transistors (2N2222) used probably had an average operating point beta on the order of 100. The ratio of actual to effective capacity in one case was about 1:4. The circuit has fairly good output of a sinusoidal waveform with low output impedance. rf

About the Author

Harvey L. Morgan is senior engineer at TDC, Inc., 621 Six Flags Drive, Arlington, TX 76011. Tel: (817) 649-4563.

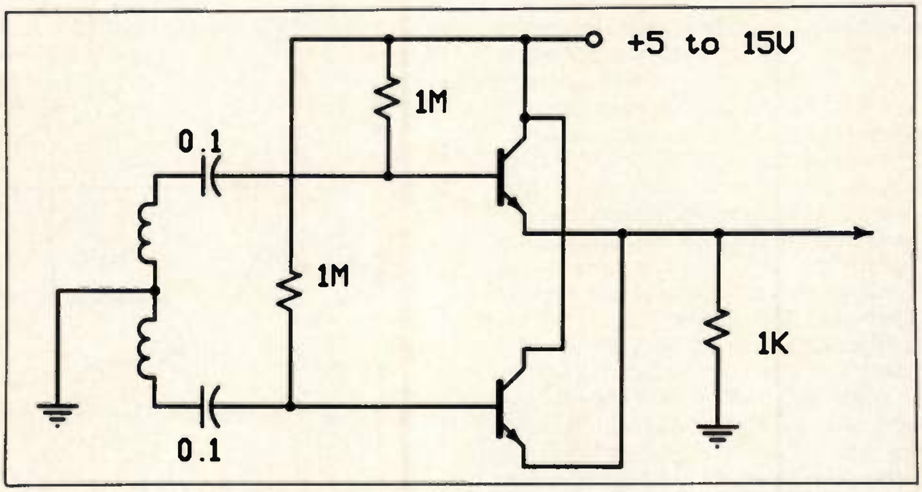

Figure 5. Improved EF oscillator.

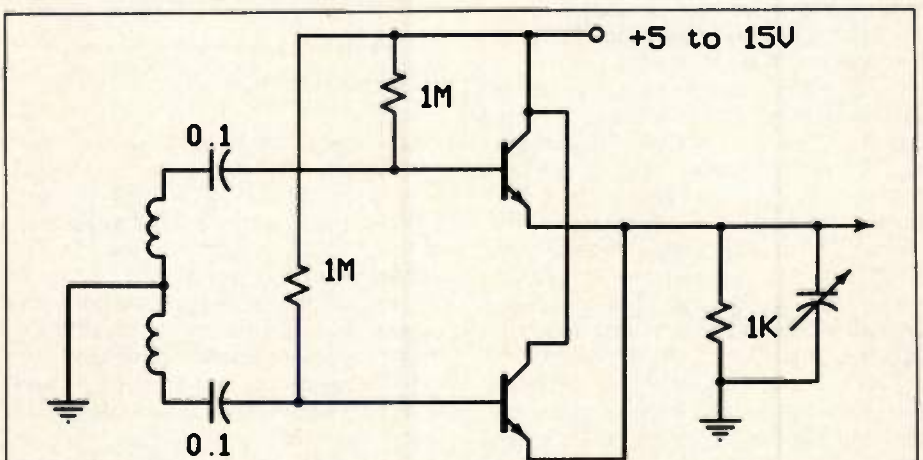

Figure 6. Variable frequency EF oscillator.

Reflections on EMC Measurements

By Tom Minnis
David Systems

The many ways by which EMC measurement can be done are dependent upon available equipment. The following is a discussion of various methods used, starting with the very old through the most modern equipment and concluding with some improvements the author feels can and should be implemented to improve efficiency. It is possible to conceive of all sorts of remote controlled antenna positioners, EUT turntables and antenna switches that would speed up the process. For the sake of discussion, it is assumed that this part of the system is the same in all cases in order to focus on the data gathering instrumentation and methodology.

Making calibrated field strength measurements on the spectrum emitted by electronic equipment is tedious and it can be difficult to maintain calibration. The fact that the FCC has required the measurements be made on an open site makes the testing even more interesting because most of what is seen on a spectrum analyzer are ambient signals. Sometimes the signals are continually present like broadcast stations, other times they come and go as in the case of two-way radio communications such as cellular radio telephone, police, fire and air traffic control. Typically, the equipment under test (EUT) will have a characteristic spectrum "signature" once the product is stable. In the development phase, the EUT signature may change, sometimes dramatically as various modifications are tried in an attempt to make the EUT meet whatever emission standard is required. It is possible to reduce one frequency component to a passing level while increasing the level of the other components that may have gone unrecorded because they were so insignificant before the modification. In the development phase of a product, the ability to quickly see the effects of a change is very important in order to maintain continuity of thought in addition to saving time.

Method I

This old widely used method makes use of a wideband communications receiver with a calibrated signal strength meter. The NF105 with numerous plug-in front ends to cover the spectrum was very popular. This method requires the equipment operator to manually record each frequency of interest, the level, line loss and antenna loss. Making the unit conversions, adding the numbers up and comparing the result to the standard for each frequency tells the user if the EUT is in or out of spec.

As the receiver is manually tuned across the band, the ambient signals are instantly recognized and ignored by simply listening to them. The EUT signals usually have a characteristic modulation or sound to them. Sometimes the EUT will put out a relatively pure unmodulated carrier that has no sound to it. Momentarily turning off the EUT and observing what happens to the signal can easily resolve the ambiguity. After a few runs, the quiet ambients and EUTs can be identified thus saving the time it takes to cycle the power on the EUT. Except for changing plug-ins, the tuning process is relatively fast compared to the time it takes to stop and record the data. Manually processing the data to make the determination as to whether the EUT passes or fails the emission standard consumes lots of time and is prone to clerical errors.

Method II

Method II takes advantage of somewhat newer test equipment with a spectrum analyzer hooked to an XY plotter with preprinted calibrated chart paper. The primary components of a typical system include the HP141T spectrum analyzer system using the 8552B IF and the 8554L RF plug-in along with the HP 7014B XY recorder and the HP 8447F wideband amplifier. This system eliminates all the clerical work in Method I because of the use of preprinted chart paper. The plot directly shows the operator if the EUT is in or out of spec.

First, a slow scan is directly plotted. If a speaker is hooked up to the video, the operator can easily tell which parts of the spectrum contain the most EUT signals by listening and watching the plotter as the analyzer slowly sweeps the band. When the plot is done, the operator lifts the pen and manually tunes in the peaks one by one and marks the ambient peaks that exceed the limit. When an EUT component is tuned in, the peak level can be visually integrated and maximized with the ability to disregard noise hits because the operator is listening to the peak detected audio.

Direct graphical output is helpful in viewing the signature and quickly determining if the EUT passes or fails. While eliminating clerical errors associated with manually manipulating large amounts of numbers, the system is dependent on a constant site attenuation since the charts are preprinted with that assumption. Hopefully, by maintaining a clean unobstructed repeatable test setup, variations in ground conductivity will introduce only minor errors. Positioning the chart paper on the plotter is a concern as well as the linearity. An experienced operator who is aware of all the sources of error in the system can minimize these errors and produce respectable charts very quickly.

The procedures for discriminating between the ambients and the EUT are the same as in Method I. In the checking phase, the spectrum analyzer is manually tuned just like a radio except the entire band can be seen at once, yet one listens only to where the analyzer is tuned. In this way the user can safely skip over large chunks of spectrum that do not have signals close to the limits. The beauty of this system is that everything is recorded, even the EUT signals far below the limits. As the design evolves, the components that were ignored in the past come up and vice versa. This is by far the fastest method for determining the relative strengths and weakness of different designs. The tabular methods may lead to more precise numbers, but the precision of the numbers may lull the operator into a false sense of accuracy that in the end may be no more accurate than can be judged by interpolating a scale on the chart.

Method III

This method takes advantage of computer technology using a computer hooked to a spectrum analyzer and various other devices via the general purpose interface bus (GPIB). A typical setup consists of the HP 8566B or the 8568B spectrum analyzer, the HP 85650A Quasi-peak adapter, a GPIB plotter and printer and one of the HP 200 series computers running the HP 85864A EMI measurement software or its equivalent.

When an EUT component is tuned in, the operator simply pushes a button on

the computer console and the computer reads the frequency and level to which the spectrum analyzer cursor is set. Since all the site attenuation data and specification limits are already known, the normalized plot can be directly displayed and plotted. While it is amazingly efficient at collecting data, it is surprisingly inefficient and difficult to use in situations where it is necessary to discriminate between ambient and EUT signals.

If the operator wants to check a signal to determine if it is coming from the EUT by listening to it, the user first sets the cursor on the signal and centers it on the screen. Then the sweep span is changed to 0 Hz and the volume is turned up on the quasi-peak detector speaker. At this point, the display has changed from a frequency domain to a time domain display of the peak detected carrier with the associated bandwidth of the IF and video filters. The operator has lost touch with the surrounding spectrum and cannot easily judge what amplitude to attribute to ambient or EUT if there is even an EUT present. After going through all this trouble to check a signal, the majority of the time the signal will be an ambient. Getting back to the frequency domain requires the operator to open up the span to a few MHz and turn down the volume on the speaker, since by this time the audio coming from it is useless.

With the spectrum now visible again, the operator tunes the cursor to the next peak not yet knowing if this peak is ambient or the EUT. After all this button pushing, all that is obtained is a tabular list of frequencies and levels that the operator thought were significant enough to record. Since it is so much work to actually listen to a small chunk of spectrum, there is a temptation to make the judgment based on how the spectrum looks rather than on how it looks and sounds. Going back and forth between the time domain and frequency domain, the operator runs the risk of both missing EUT signals and mistaking ambients for EUTs. An inexperienced operator who realizes that he can be fooled by simply looking at the spectrum will diligently tune each one in and listen (this can take many hours). One possible method of avoiding the time it takes to constantly switch between the frequency domain and the time domain would be simply to stay in the time domain with 0 Hz span and tine the center frequency. This tuning method is equivalent to Method I with the advantage of automated logging.

Method IV

An improvement over Method III would eliminate the disadvantages of Method III and include the advantages of Method II. The author believes that this would only require a firmware change to the spectrum analyzer. The equipment setup would be the same as in Method III but with an improved spectrum analyzer and software.

The start and stop frequencies are set by the antenna in use such as a frequency span from 30 to 200 MHz. Next, a slow scan is saved in memory and displayed including the limits. To verify the peaks, the operator manually tunes the cursor over the stored display, listening while tuning. The analyzer is now in a zero span mode. The stored display remains and the cursor follows the stored display and indicates the level of the detector without permanently overwriting the stored display. When an EUT signal is tuned in with a peak or quasi-peak detector, the level can be determined. This level can be marked in memory and on the display and read into a tabular file on the computer. When the verify phase is completed, the operator plots the screen which clearly shows the EUT signals against the limits and the ambient. The tabular listing of each component of the EUT spectrum signature can be printed while the screen is being plotted. During the printing and plotting phase, antennas and/or EUTs can be changed. Method IV was envisioned as a minimal design change solution to existing hardware which makes the system acceptable in this application but not optimal. Since the author is not intimately familiar with the inner workings, perhaps someone who is could comment on the feasibility of this modification.

Method V

This method would use a modified HP 8568B or 8566B spectrum analyzer with a new version of the HP 85650A quasi-peak detector. The GPIB plotter, printer and 200 series computer would obviously not need to change but the software could be improved.

In this scenario, the operator would start out by defining the limits of the scan among other things as is the case now. Next, a slow scan is started with the video stored in the spectrum analyzer memory

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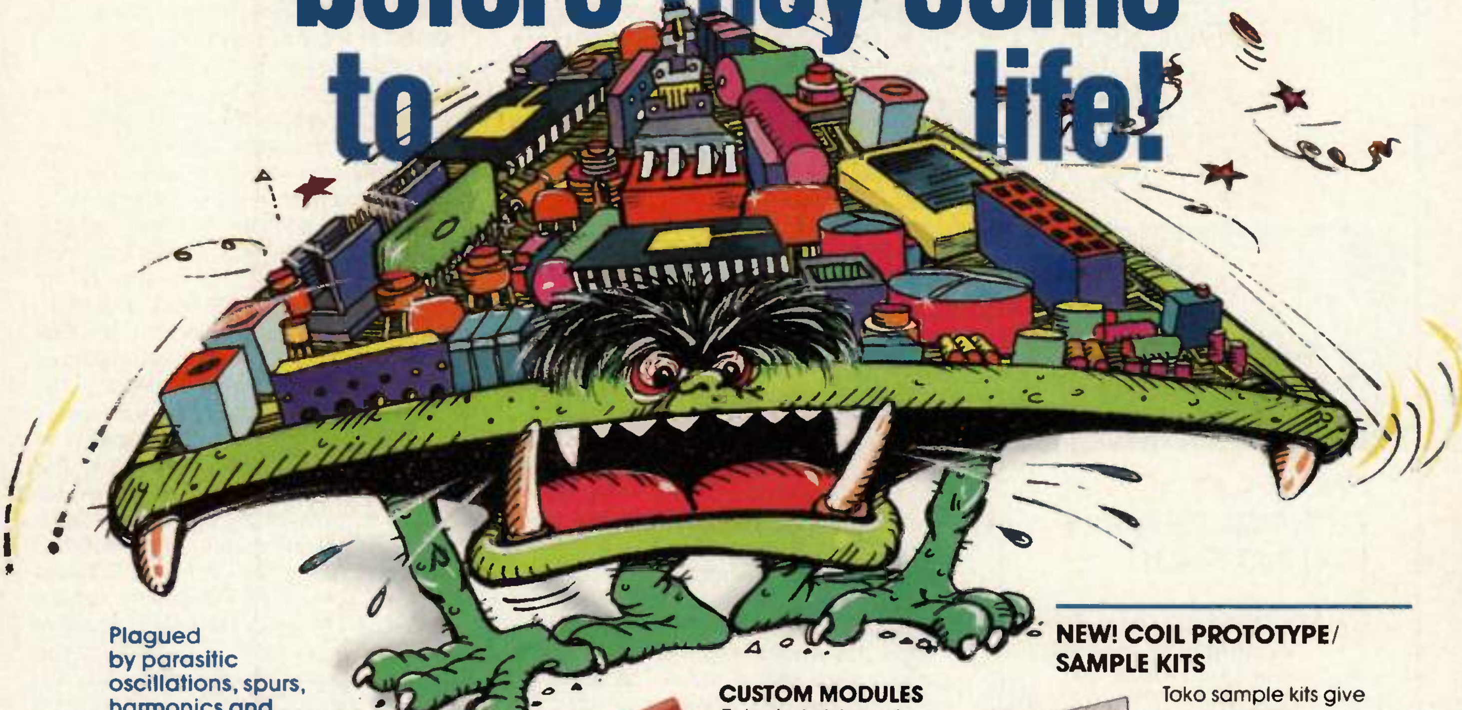

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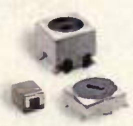

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and displayed on the tube with the limits. As the sweep slowly moves up the band, the detected video is audible, just as in Method II. In the verification phase, the operator manually tunes in the peaks and listens to where the cursor is positioned over the stored plot. Typically, the frequency scale of a complete scan is too big to see the detailed spectrum. So, in this mode the stored spectrum, while it is saved, is displayed in low intensity and a detailed spectrum is displayed in normal intensity. The detailed spectrum would be centered about where the cursor is tuned on the stored display and the frequency to which the analyzer was tuned would always be centered on the real time detailed spectrum. The span, bandwidth, video and filtering would all work normally but in addition, the audio would be representative of what the user would hear if the analyzer was not sweeping and was tuned to the frequency at the center of the display. This would require an additional mixer and IF detector chain that would track the bandwidth settings of the spectrum analyzer. When the EUT signal is tuned in and maximized, the operator could mark the level and frequency with a vertical line that touches the frequency scale and whose length is precisely the level of the detected signal. The spectrum plot and the EUT lines could have different colors and each line would be described in detail on the tabular listing. As the data is taken for each line, the operator could type in comments without having to exit the program. This is especially important when the signal is maximized because during the maximization process, EUT setup conditions can change and a means of noting these changes on the fly is important during the development phase of a product.

The ability to see in the frequency domain on either side of where a receiver is tuned is not a new concept. The 1950's vintage Hammerlund SP600 was available with a panoramic adapter. Heath has offered at least two pan adapter models over the years for their communications receivers (both now discontinued). More recently, Kenwood offered the SM220 to work in conjunction with their newer equipment, and ICOM's recently introduced IC-781 communications transceiver includes a pan adapter. All these designs are receivers first and simple spectrum analyzers second. The author is not aware of a modern GPIB controlled spectrum analyzer that can function as a simple receiver *and still display the spectrum.*

This article illustrates the need for the test equipment designer to adequately familiarize himself with the application for which he is designing the equipment. The author (and readers) would appreciate additional information on new model capabilities from equipment manufacturers.

Special thanks goes to Noel Duckett at Ford Aerospace for his help with describing the EMI measurement techniques of the past. rf

About the Author

Tom Minnis graduated from Cal Poly San Luis Obispo in 1971 with a BSEL and holds amateur extra class license WB6HYD. He most recently worked at David Systems as a senior design engineer. He can be reached at 1132 Reinclaud Court, Sunnyvale, CA 94087. Tel: (408) 720-1650.

EMC Directory to Debut in January

Liberty Labs will be publishing the 1989 Directory of EMC Test Laboratories and Consultants in January 1989. The publication will contain detailed information on EMI/EMC test labs and consultants, telecommunication labs and consultants, product safety labs and consultants, and calibration services for test labs. The specific entries will include national and international services, services by region, test types and capabilities.

The directory will contain seven main sections. The first section will feature guest articles and publications relating to EMC test labs and consultants. This will be followed by a resource summary section with history, capabilities and services, facilities, major instrumentation, personnel, hours of operation, lead times, and contacts. This is followed by a capability index (FCC Part 15-J, FCC Part 90, MIL-STD 461, etc.), state index, national and foreign index, accredited labs, certified EMC personnel and professional engineers, indexes, plus an advertising section.

The cost of the directory is $65. Interested consultants and test labs that wish to be listed in the directory (for free) can contact Liberty Labs, Inc., P.O. Box 8268, Cedar Rapids, IA 52408. Tel: (319) 390-3646; Fax: (319) 390-3802. To receive ordering information, circle INFO/CARD #220.

Dissipative Filters

These Filters Work by Absorption Rather Than Reflection.

By Matthew Fivash
Honeywell, Inc.

Common filter types, such as Butterworth, depend on reflection for their performance. Networks can be designed to achieve filtering action by absorption rather than reflection. These networks may be called dissipative filters (DFs) as the stopband energy is dissipated rather than reflected.

All filters whose constituents are lossless components (inductors and capacitors) operate by selective reflection of input energy. This type of filter may be modeled with the Feldkeller energy equation.

$$|T(j\omega)| = \frac{1}{1 - |\Gamma|^2}$$

where Γ is the input reflection coefficient.

This equation, a statement of the principle of energy conservation, shows that all energy which enters a lossless network must appear at the output. Stopband energy, then, is reflected back to the source by impedance mismatch. As can be seen, the source and load impedances play an important role in filter performance.

When termination impedances are poorly controlled, reflection filters can produce disappointing results. These limitations might be mitigated by series or shunt resistances or resistive attenuators at the filter's ports. While this will place limits on the port impedances, a passband loss penalty is extracted. When reflecting stopband energy back to the source is undesirable, diplexers can be designed to pass stopband energy to an alternate port. Diplexer design, however, requires three port networks and the transition band can be a source of trouble. Both of these cases can be addressed using dissipative filtering.

Dissipative Filter Topology

Ideal passband and stopband two-port models of a dissipative filter (DF) are shown in Figure 1a and b, respectively. A little investigation suggests the networks in Figure 2 as possible approximations of the ideal two-port model. The network in Figure 2b is easily recognized as an amplitude equalization network (1). Cuthbert (2) presents a brief analysis of both networks in Figure 2. He shows that both networks have an ultimate attenuation slope of 6 dB per octave. The input and output impedances of both networks converge to the value of the resistors.

Another filtering network with dissipative properties is the RC ladder filter. This network, analyzed in (3), introduces a constant loss dependent on source and load terminations. The DF circuits shown in Figure 2 do not have this loss restriction.

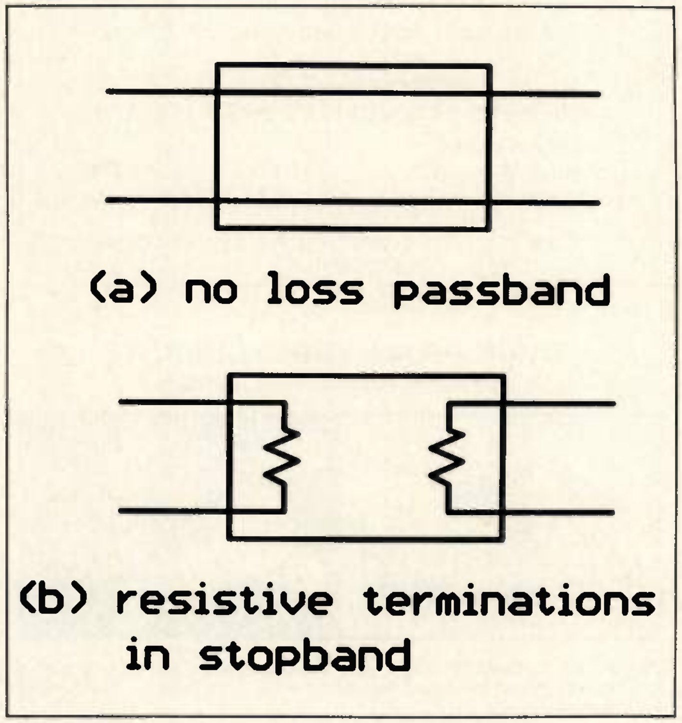


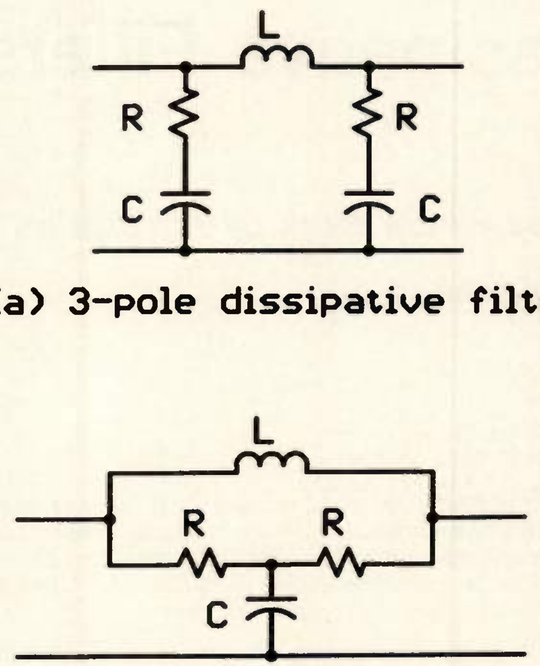
Figure 1. Ideal non-reflective filters.

DF Performance

Two-port theory is a useful means of analyzing filters. To develop insight into DF operation it is instructive to compare the response of a three-pole Butterworth filter with that of a three-pole DF (Figure 2a). Worst case filtering performance of both filters will occur when the input and output are simultaneously conjugately matched at all frequencies. While this is a contrived case, it provides an unexpected view of filtering action. Maximum available gain (MAG) is given by

$$\mathrm{MAG} = \frac{|Y_{21}|}{|Y_{12}|}\left(K + \sqrt{K^2 + 1}\right) \qquad K = 2\left[\frac{g_{11}g_{22} - g_{21}^2}{|Y_{21}|^2}\right] + 1$$

Figure 3 shows the frequency response for a three-pole Butterworth and DF filter. Best case (constant real impedances on



(a) 3-pole dissipative filter

(b) 2-pole dissipative filter

Figure 2. Approximations to ideal model.

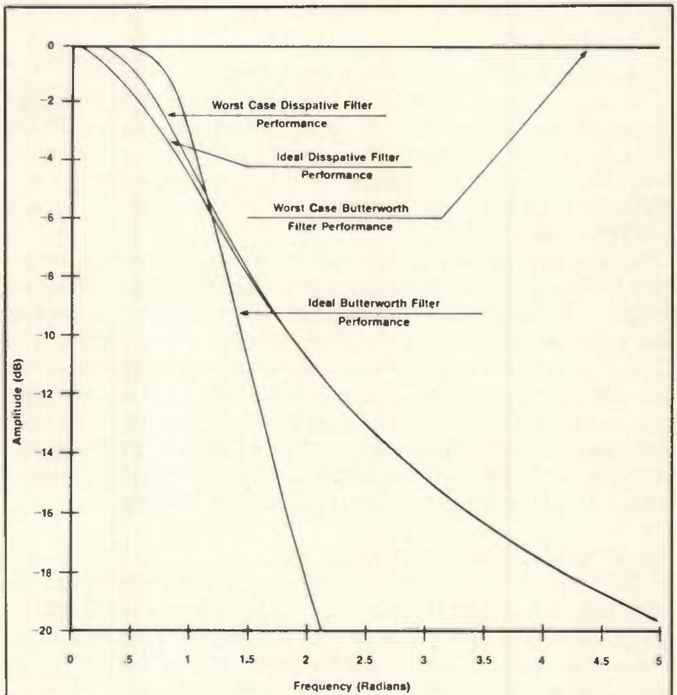


Figure 3. Frequency response.

both ports) and worst case performance are shown. As can be seen, the lossless Butterworth design provides no attenuation when the ports are conjugately matched. The performance of the DF is not strongly affected by source and load impedances.

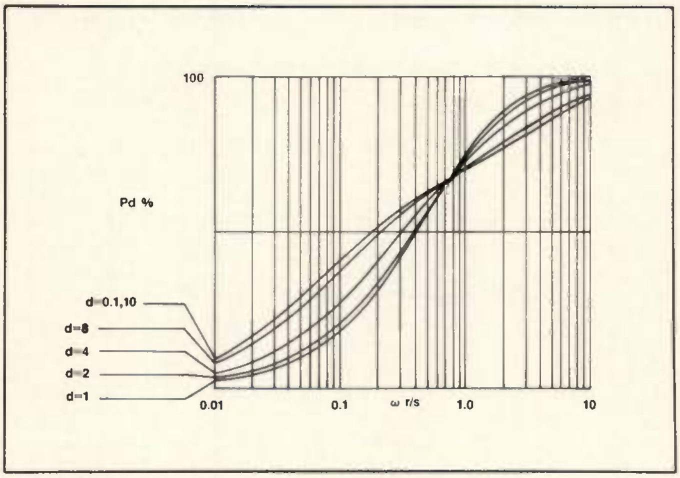


Figure 4. Input power vs. frequency.

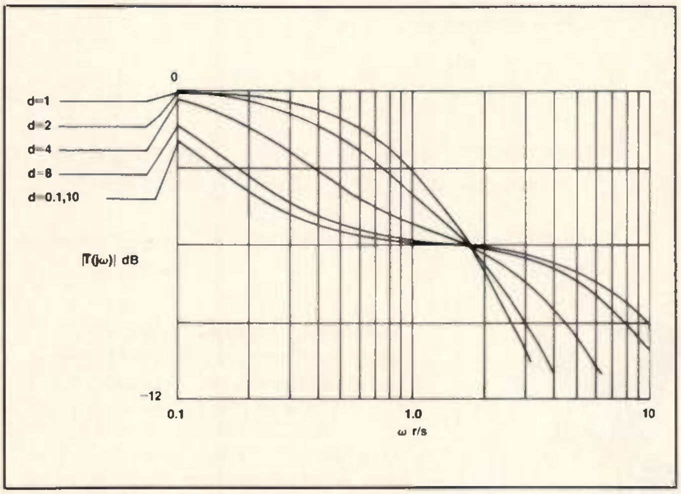


Figure 5. Magnitude vs. frequency for a two-pole DF.

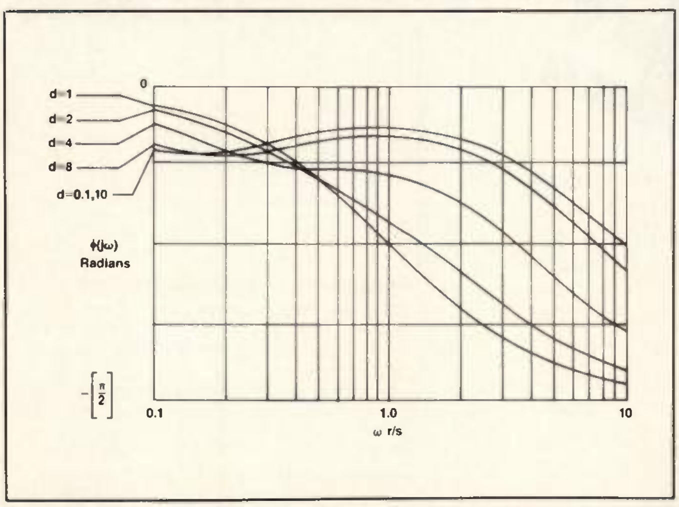


Figure 6. Phase vs. frequency for a two-pole DF.

Most of the analysis presented here is based on ABCD or chain parameters. Impedances and transfer functions of an embedded two-port can be cast in terms of these parameters. Reference 4, for instance, tabulates these relations in terms of ABCD parameters. The ABCD matrixes for the two-pole and three-pole

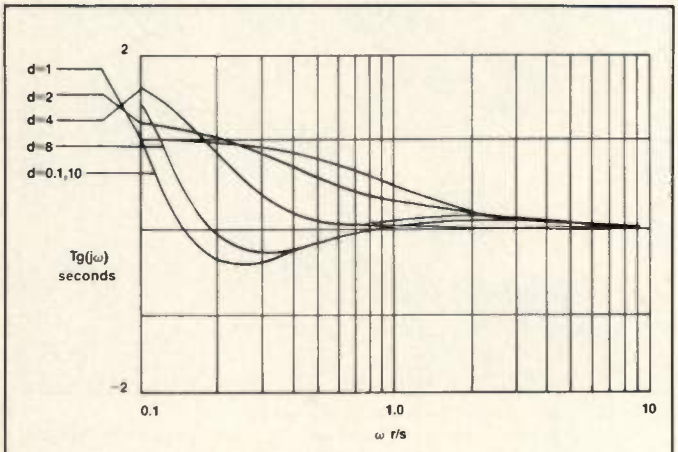


Figure 7. Group delay vs. frequency (two-pole DF).

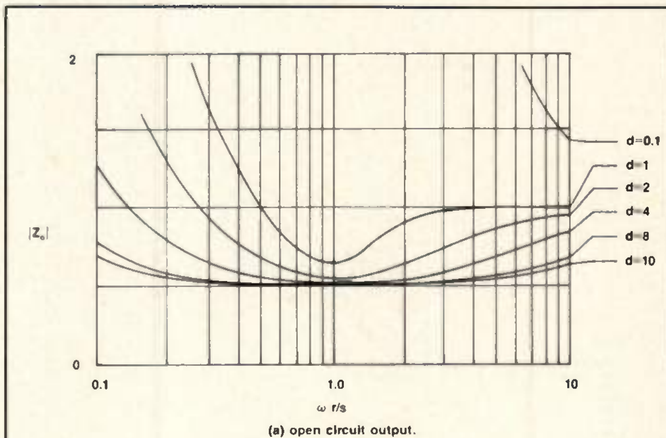


Figure 8a. Input impedance vs. frequency (two-pole DF — open circuit).

(Figure 2b and 2a, respectively) DF networks are:

$$2b: \begin{pmatrix} \dfrac{S^2RLC+S(R^2C+L)+2R}{S(R^2C+L)+2R} & \dfrac{S^2R^2LC+2SRL}{S(R^2C+L)+2R} \\[2ex] \dfrac{S^2LC+2SRC}{S(R^2C+L)+2R} & \dfrac{S^2RLC+S(R^2C+L)+2R}{S(R^2C+L)+2R} \end{pmatrix}$$

$$2a: \begin{pmatrix} \dfrac{S^2LC+SRC+1}{SRC+1} & SL \\[2ex] \dfrac{S^3C^2L+2S^2C^2R+2SC}{(SRC+1)^2} & \dfrac{S^2LC+SRC+1}{SRC+1} \end{pmatrix}$$

Power dissipated in the filter is easily analyzed in terms of

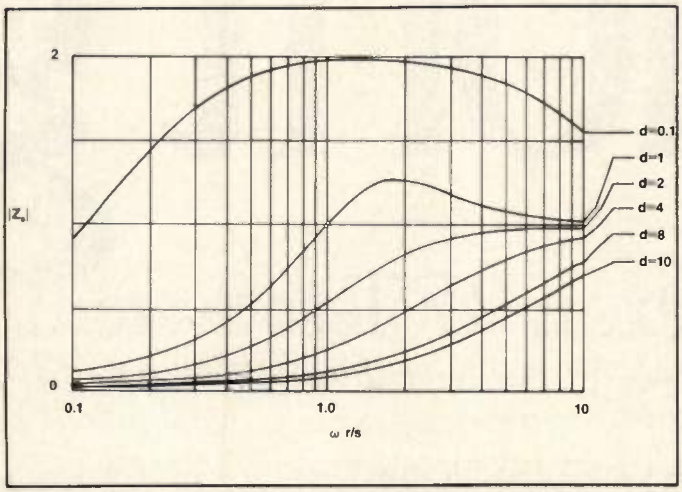


Figure 8b. Input impedance vs. frequency (two-pole DF — short circuit).

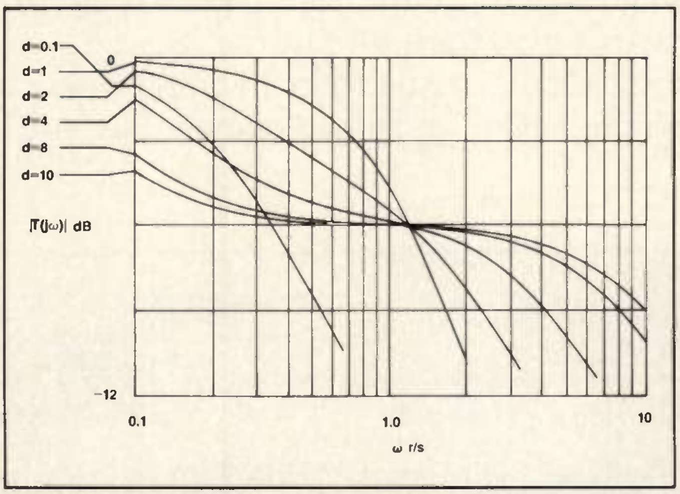


Figure 9. Magnitude vs. frequency (three-pole DF).

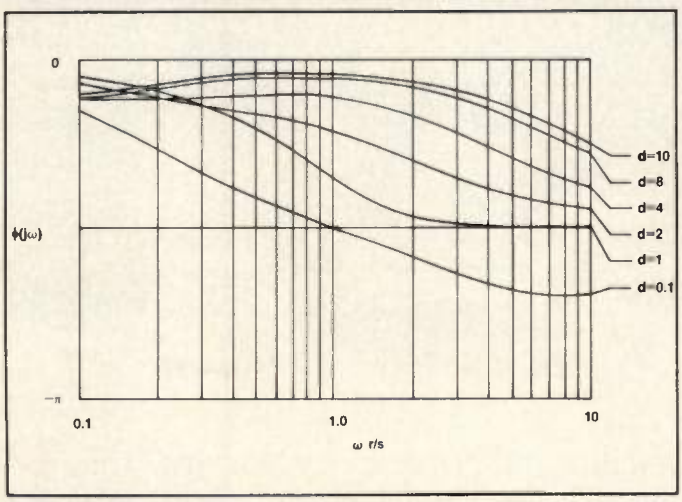


Figure 10. Phase vs. frequency (three-pole DF).

the difference between input and output power as a percent of available input power. Given the DF input impedance and power available from the source (P_{as}), the power entering the DF is given by:

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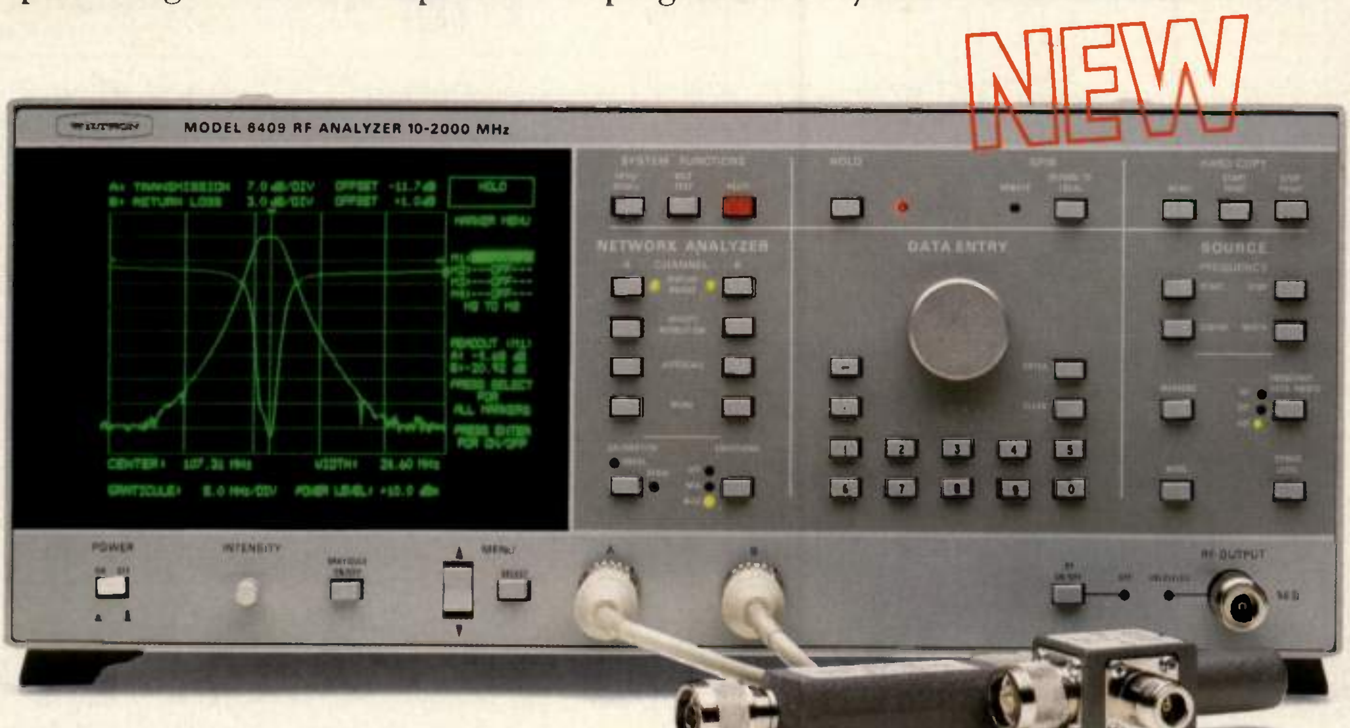

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Harmonics: < −30 dBc
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Dynamic Range: +16 dBm to −55 dBm
Vertical Display Resolution: 0.003 dB maximum
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Normalization: 800 points, automatically interpolated for ranges less than full range.
Markers: up to 8.
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Directivity: 40 dB
Impedance: 50 or 75 ohms
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Test Port Connector: Type N or BNC

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Passband Amplitude at M1 Marker: −4.82 dB

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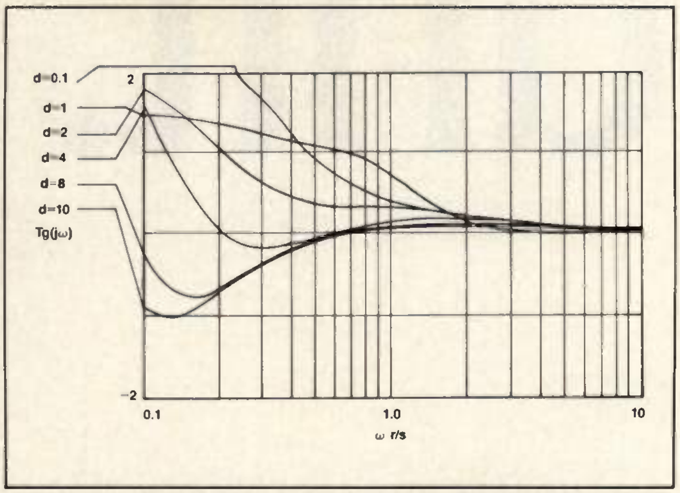

Figure 11. Group delay vs. frequency (three-pole DF).

$$P_{in} = \left[1 - \left| \frac{Z_{in} - R}{Z_{in} + R} \right|^2 \right] P_{as}$$

The percent input power absorbed by the DF is then:

$$P_d\% = 100 \frac{P_{in} - P_{out}}{P_{as}} \quad \text{where } P_{out} = \frac{[R_e V_{out}]^2}{R}$$

As can be seen in Figure 4 as the input frequency enters the stopband, most of the input power is absorbed by the filter. It is therefore important that the resistive elements in the DF design be capable of dissipating stopband energy.

The ABCD parameters are used to plot frequency response, input impedance (with open and short loads), and group delay characteristics of DF filters. A design parameter, d, defined by Cuthbert (2) is varied to show the range of DF performance.

$$d = \omega_o RC \quad \text{where } \omega_o = \frac{1}{\sqrt{LC}}$$

All the DF response plots are designed with constant characteristic impedance (the value of the resistive elements) and constant cutoff frequency. The definition of cutoff frequency is:

$$\omega_o = \frac{1}{\sqrt{LC}} \text{ r/s}$$

Note that the 3 dB point on the magnitude response corresponds to the cutoff frequency only when d = 1.

The performance of the two-pole DF is symmetric with respect to d and 1/d. Thus, a design with d = 10 has the same response as a design with d = 0.1. The three-pole DF does not possess this symmetry.

Designing with the Dissipative Filter

Dissipative filters can improve system or circuit performance

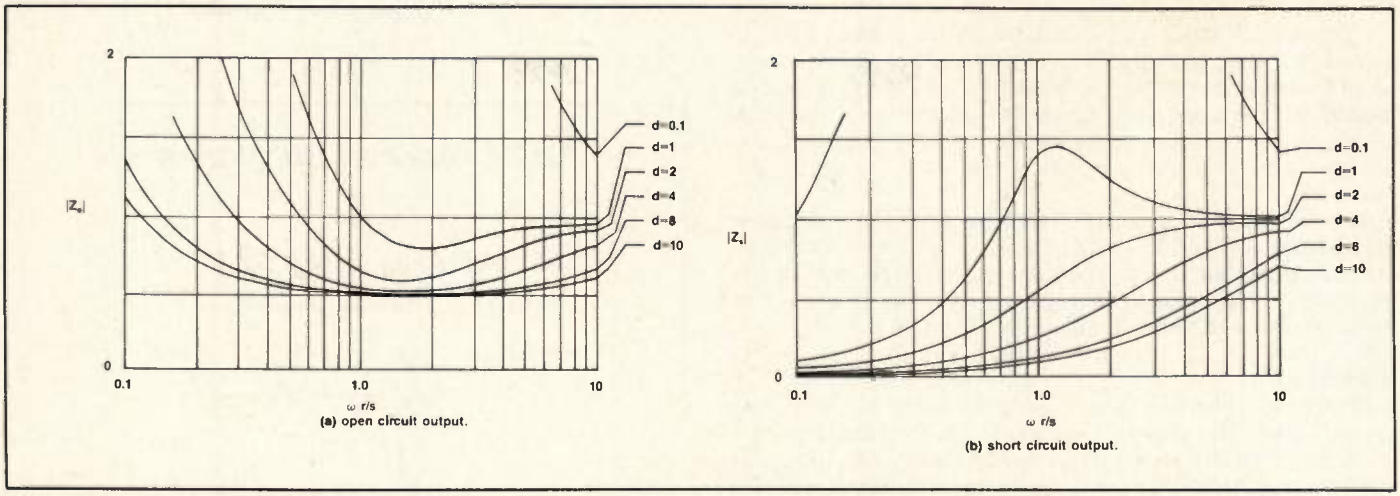

Figure 12. Input impedance.

in a variety of ways. Dissipative filters can be used as terminating sections for lossless filters greatly improving transfer function predictability in the face of unknown or changing impedance environments. DFs can be used on the output of circuits that have a low tolerance for reflected energy in the stopband. The DF can be thought of as a frequency selective attenuator.

The minimum attenuation through a DF is given by:

$$L_{min} = 8.6858 \ \mathrm{Sinh}^{-1} \sqrt{R} \ \mathrm{dB} \qquad \text{where } R = \frac{g_{11}g_{22} - g_{21}^2}{|Y_{12}|^2}$$

DF sections can be cascaded and the attenuation through the cascade will be at least the sum of the minimum attenuations through each DF section. In the case of the two-pole DF with $d = 1$, the attenuation will be exactly the sum of the minimum attenuations.

Highpass, bandstop, and bandpass DF sections can be designed. Standard transformations apply to this network as well as to lossless filter networks. Williams (5) suggests a scaling method for both cutoff frequency and characteristics impedance. DF lowpass prototypes can be transformed to highpass, bandpass, or bandstop filters by using the standard component transforms.

As an example, suppose a 100 MHz bandpass DF is required to terminate a mixer. The characteristic impedance will be 50 ohms, and choose d = 1. The lowpass prototype is shown in Figure 13a and the transformed filter is Figure 13b. The projected performance of the filter is shown in Figure 14.

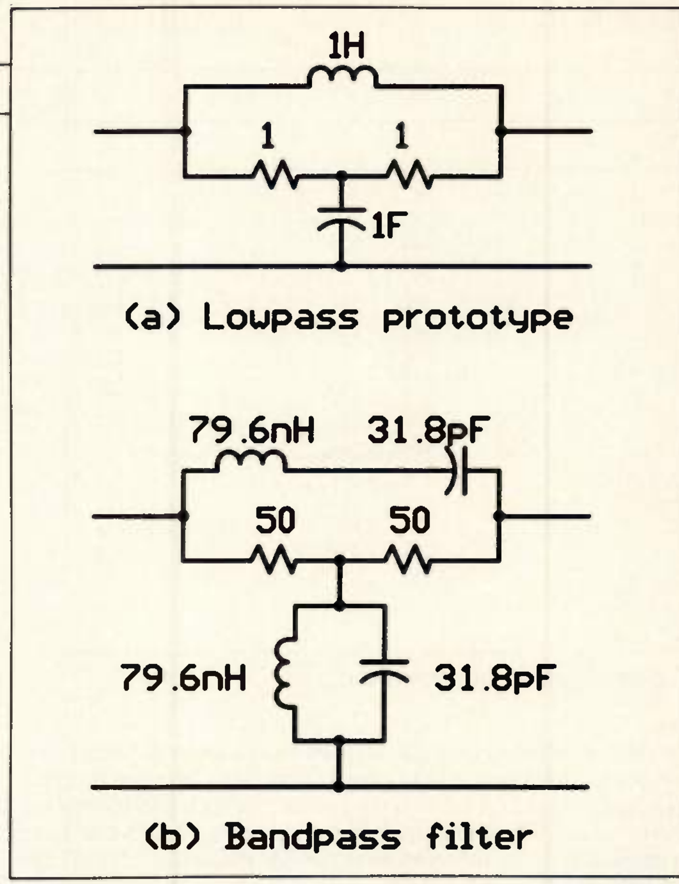

Figure 13. Lowpass prototype and final revised filter.

Summary

Dissipative filters can serve useful functions in circuit and system design. By offering a method to control stopband impedances in low frequency networks, DF circuits can reduce or eliminate the effects of uncontrolled terminating impedances. DF networks do not offer the attenuation slope of an equivalent lossless network, but they can be combined with lossless filters to improve performance of the overall system. rf

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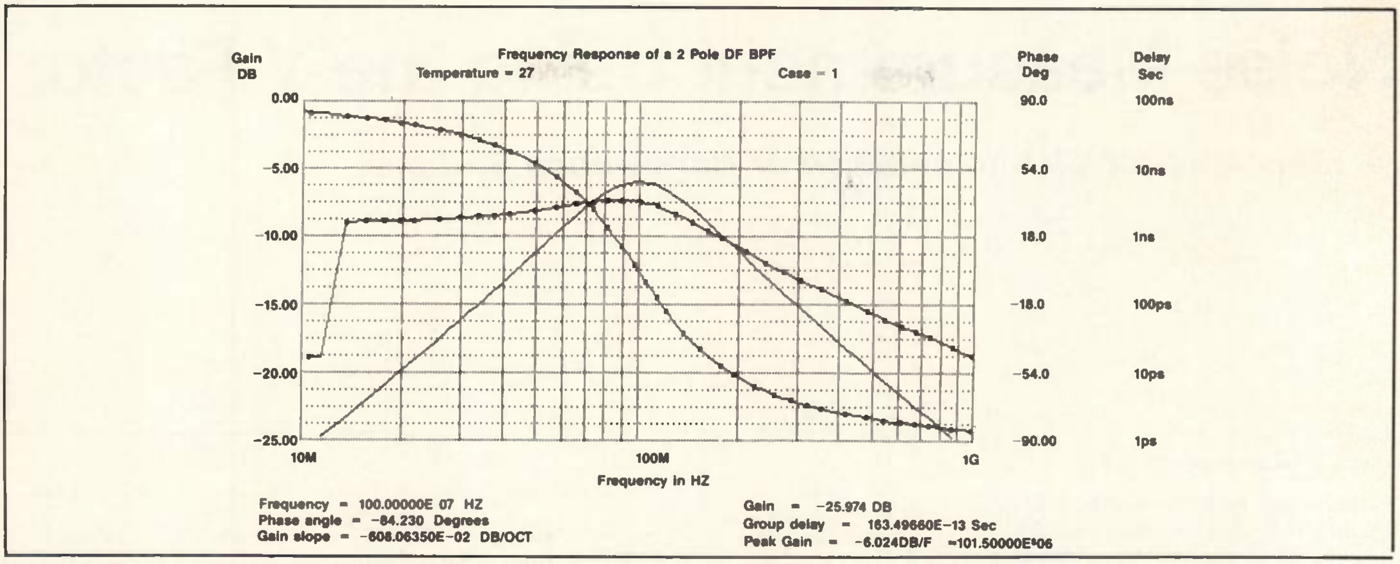

Figure 14. Frequency response of a two-pole DF.

March 31, 1987.

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Noise Measurement Using the Y Factor

A Technique for Characterization of Narrowband Sources.

By Paul Drexler
Naval Air Development Center

Receiving system sensitivity is often degraded by the presence of undesired broadband energy, or "noise." Noise is of particular significance when signals are close to the receiver noise threshold, as in GPS (Global Positioning Satellite) downlink channels, LPI (Low Probability of Intercept) systems, and weak signal communication modes such as meteor scatter. Noise may be attributed to sources that are both external and internal to the receiving system. The design and manufacture of high performance receive systems requires careful attention in the characterization and reduction of undesired noise. This article presents a laboratory measurement technique that allows simple measurement and characterization of receiver noise performance.

Receiver sensitivity may be measured and evaluated using several laboratory techniques. Signal-to-noise-ratio, quieting, and noise figure are three popular methods of characterizing a receiver's ability to detect weak signals. Noise figure is often used since it may apply to entire receive systems as well as individual components and sub-systems. Noise figure meters, although adequate for many measurements, are restricted to measuring wideband frequency devices and require external frequency downconversion when evaluating high frequency systems. The Y factor method presented here overcomes these drawbacks, requiring only a laboratory spectrum analyzer and calibrated diode noise source. This technique has become particularly beneficial in characterizing the narrowband devices finding increased application in modern receiver systems, such as amplifier stages using narrowband SAW filters.

Measurement Theory

Any network when measured at a temperature other than zero Kelvin will have

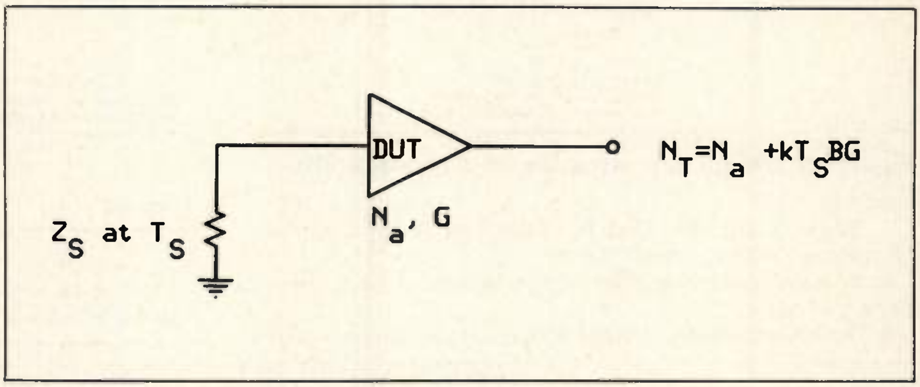

Figure 1. Nonideal amplifier noise output.

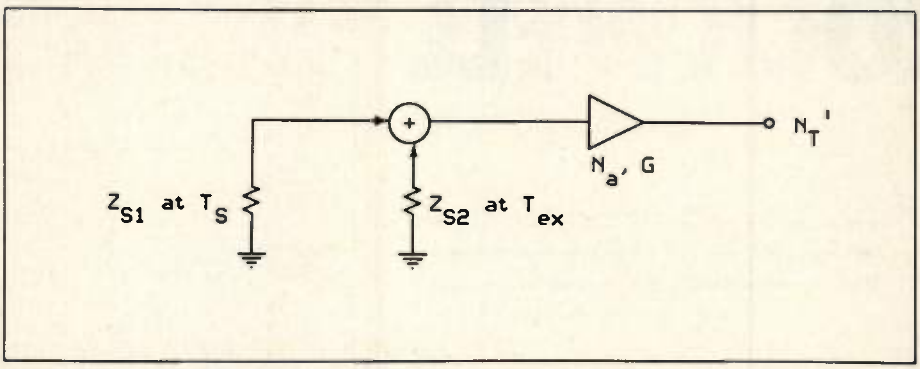

Figure 2. Amplifier noise output with external noise source input.

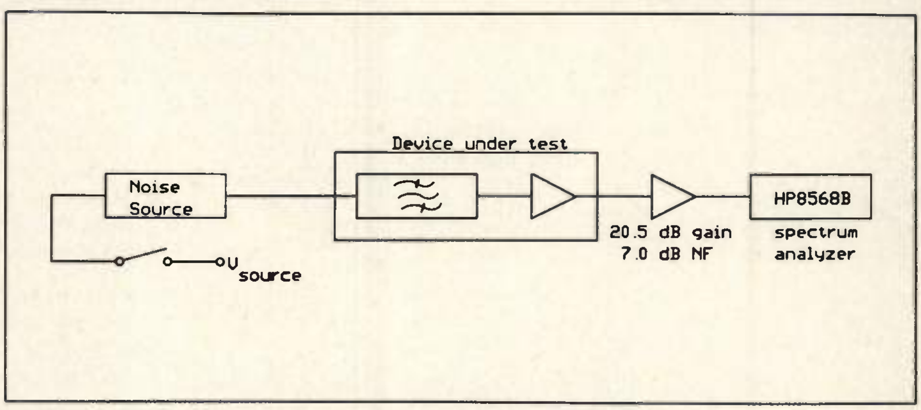

Figure 3a. Y factor noise measurement.

an output component that is unrelated to its input signal, and this unrelated output signal is in the form of a random noise signal. The simplest device that's often considered for such a measurement is a pure resistance at some finite temperature T. A resistor will have a noise output power that is primarily due to thermally agitated electrons within the resistor itself. This power is termed the available thermal noise power, and is expressed by

$$P_a = kTB \qquad (1)$$

where P_a is the available noise power measured in watts, k is Boltzmann's constant, equal to 1.38×10^{-23} watts/K, B is the bandwidth in Hz in which the measurement is made, and T is the resistor temperature measured in absolute Kelvin. It should be noted that the noise power is totally independent of the resistor ohmic value, but is proportional to the physical temperature of the resistor, which is responsible for the thermal agitation of free electrons. When the units of equation 1 are evaluated at room temperature, 290 K, P_a is equal to 4×10^{-21} watts/Hz, or −174 dBm/Hz. Thus it is apparent that at a given temperature the available noise power is proportional to the bandwidth of the measurement device.

The relationship of equation 1 may be extended to characterize a typical amplifier, where results offer greater significance. Consider an amplifier as the device under test (Figure 1) having a power gain G and adding an additional noise component N_a to the desired signal output. Note from equation 2 that the total noise output power, N_T, is equal to the sum of the amplified input noise due to a source impedance Z_s at a finite temperature, as well as the additional noise component generated within the DUT.

$$N_T = N_a + kT_sBG \qquad (2)$$

Several terms contribute to N_T, the total output noise power of the DUT. Note however that the N_a term of equation 2 is independent of G, T_s and B. Since it is the N_a term that characterizes the noise performance of system, or DUT, having a specific absolute temperature, frequency bandwidth, and gain, it is this term that should be minimized for superior detection of small signals.

Since the total noise output, N_T, from such a device may be difficult to measure directly, another measurement technique is desirable. One such method is to apply an input signal to the DUT and then measure the resulting output signal. Since any input signal will consist of the sum of a desired signal plus an undesired noise component, there exists a ratio of signal power to noise power (SNR) at the device input terminals.

Furthermore, since the device is non-ideal the SNR observed at its output terminals will be degraded from the input SNR. The ratio of the input SNR to the output SNR is termed the noise factor (F) of a system, and is described by equation 3.

$$F = \frac{S_i/N_i}{S_o/N_o} = \frac{S_i\,N_o}{S_o\,N_i} \qquad (3)$$

where S_i and S_o are the input and output signal powers, and N_i and N_o represent the respective noise powers. When equation 3 is expressed as a log ratio, the noise

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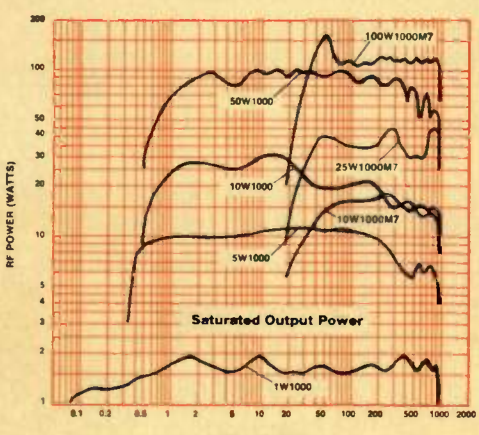

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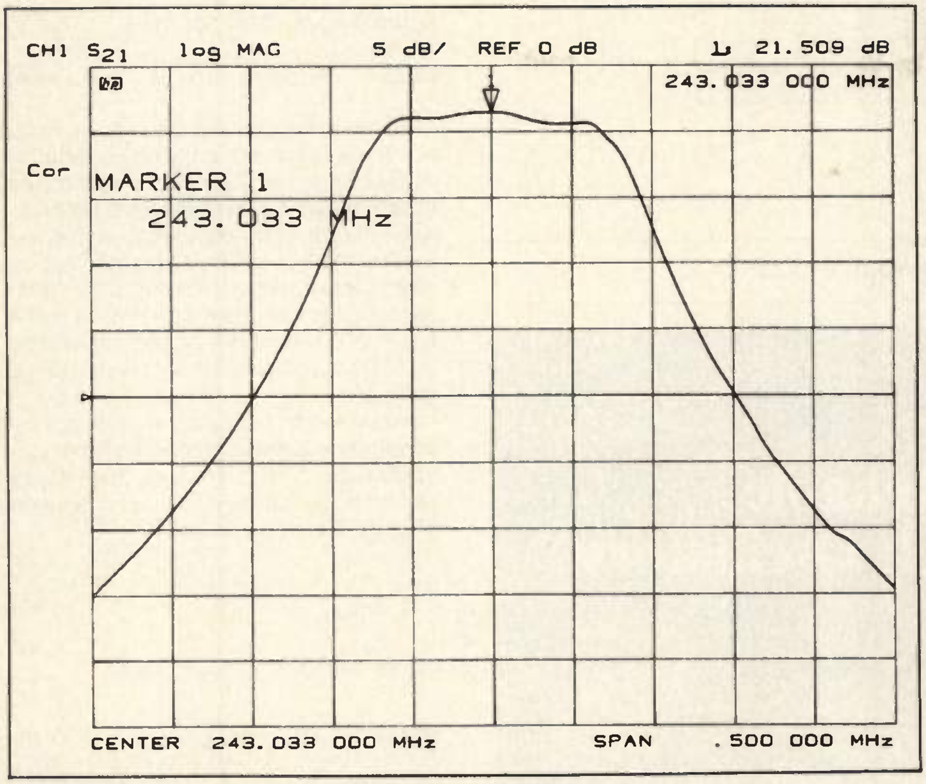

Figure 3b. 243 MHz narrowband device under test.

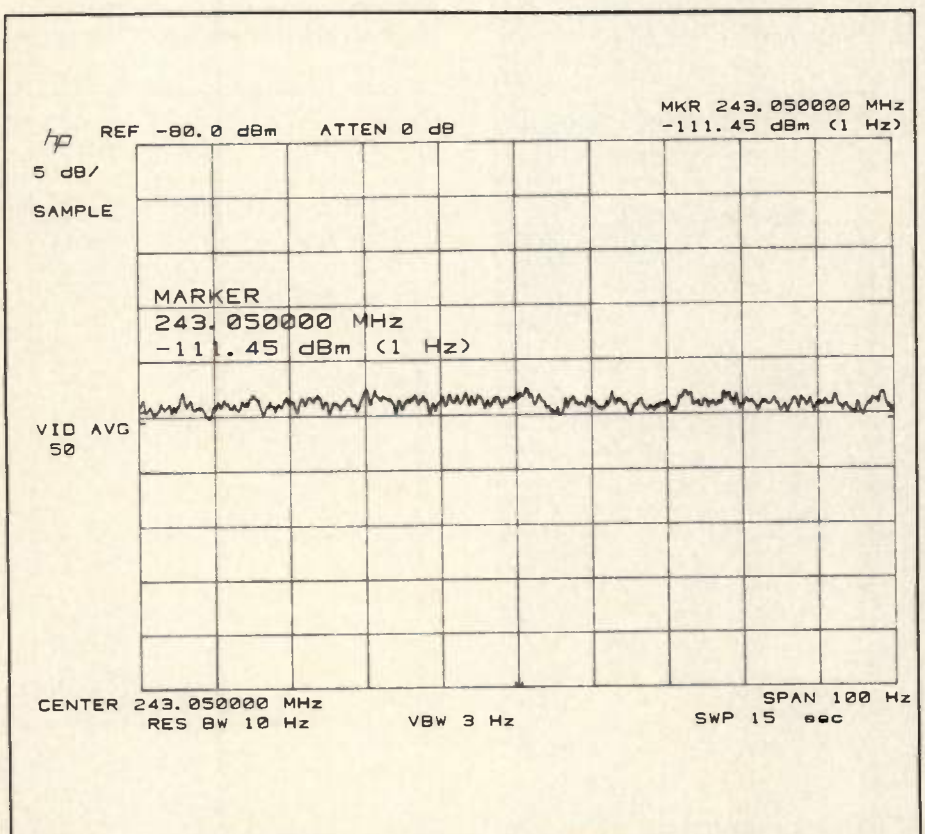

Figure 4a. N_T' total noise power with noise source "On."

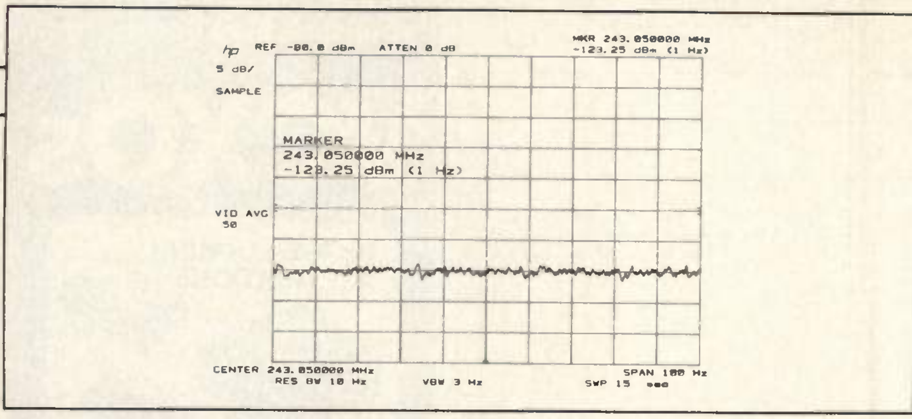

Figure 4b. N_T noise power with noise source "Off."

figure (NF) is obtained; thus,

$$NF(dB) = 10 \log F \tag{4}$$

For an ideal amplifier the ratio of equation 3 will result in a unity noise factor or 0 dB noise figure. Care must be exercised in distinguishing the difference between noise figure and noise factor since the two are easily confused and may not be directly interchanged. Although this represents the fundamental definition of noise figure, SNR is difficult to measure and so an alternative measurement technique is desirable.

Noise factor may also be defined by considering the ratio of total available output noise power (N_T) to the noise generated by the input source impedance (kT_sBG). Referring to equation 2,

$$F = \frac{N_T}{kT_sBG} \tag{5}$$

or expressed as a noise figure,

$$NF(dB) = 10 \log \frac{N_T}{kT_sBG} \tag{6}$$

Direct NF measurements, as implied by equations 5 and 6 are difficult to perform and hence rarely used since the denominator is not easily evaluated. This uncer-

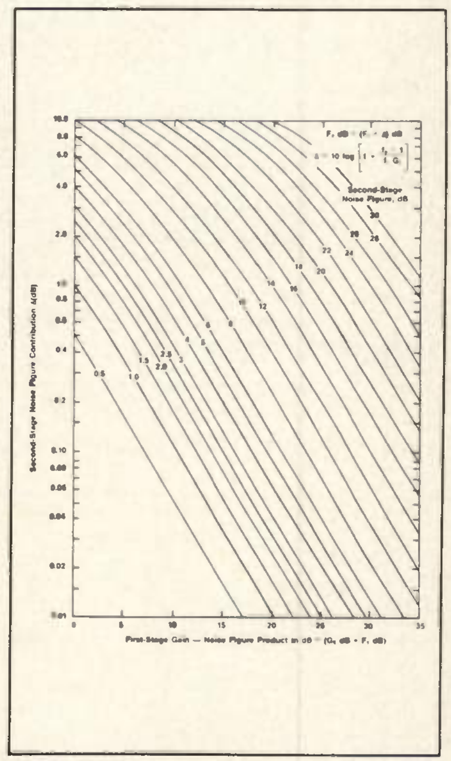

Figure 5. Cascaded noise figure.

tainty however, may be all but eliminated by considering not only the output noise N_T of equation 2, but also including an extra noise input due to an added noise source, illustrated in Figure 2.

If the extra noise source of Figure 2 has a temperature T_{ex}, then the new total noise power N_T' may be given by

$$N_T' = N_a + kT_sBG + kT_{ex}BG$$

or,

$$N_T' = N_a + (T_s + T_{ex})\, kBG \qquad (7)$$

Upon algebraic substitution of equations 2 and 7 into equation 5, a useful result may be obtained:

$$NF(dB) = 10 \log \left(\frac{T_{ex}}{T_s}\right) - 10 \log \left[\frac{N_T'}{N_T} - 1\right] \qquad (8a)$$

or,

$$NF(dB) = ENR(dB) - 10 \log \left[\frac{N_T'}{N_T} - 1\right] \qquad (8b)$$

The result of equation 8b is the fundamental basis for all noise figure measurements. T_{ex} is the external noise source temperature and T_s is the temperature of the DUT. N_T' represents the total output noise power including the external noise source (Figure 2), while N_T is the noise power output without the added noise source (Figure 1). The ratio N_T'/N_T is determined by measuring the ratio of available power with the noise source "On" to that with the noise source turned "Off." The ratio N_T'/N_T is termed the "Y" factor.

The first term in equation 8a is due to the noise source and is very nearly equal to the excess noise ratio (ENR) of the external source used. More accurately, ENR is defined as:

$$ENR = 10 \log \left(\frac{T_{ex}}{T_s} - 1\right) \qquad (9)$$

Notice however, that since the ratio T_{ex}/T_s is usually much larger than one, the approximation used in the first term of equation 8a may be considered acceptable and the result in equation 8b is obtained. Commercial solid state diode noise sources are available with ENR's ranging from 5 to 15 dB, and ENR data versus frequency is normally supplied as part of the manufacturer's data. Traditional automatic noise figure meters utilize the described Y factor relationship by switching a noise source and obtaining an out-

N_T	N_T'	NF (dB)	Corrected NF (dB)
−123.25 dBm	−111.45 dBm	3.69	3.64

Table 1. Measurement results.

put display calibrated in noise figure.

Measurement Example

A reasonably accurate noise figure measurement may be obtained using a calibrated noise source and a spectrum analyzer. Figure 3 illustrates the test set-up used to measure a narrowband front-end at 243 MHz, used in an emergency guard channel receiver. The noise source used is a Hewlett-Packard HP346B noise source having a 15.2 dB ENR at a frequency of 243 MHz. The HP8568B spectrum analyzer's video averaging function was used to negate signal amplitude variations. Fifty averages were made. Since signals are close to the noise floor of the analyzer, no RF attenuation was used. A broadband amplifier follows the DUT in order to raise the signal level available to the spectrum analyzer. Without voltage applied to the noise source, a power measurement is obtained on the spectrum analyzer using a 1 Hz bandwidth and a convenient frequency span; the measurement bandwidth of 1 Hz is narrower than the DUT, as required for accurate NF measurement. The procedure is then repeated with the noise source "On" and the ratio N_T'/N_T is obtained. Both N_T' and N_T are first converted from dBm using the relationship:

$$\mathrm{mW} = 10^{\mathrm{dBm}/10} \tag{10}$$

Several trials of N_T'/N_T are completed in order to assure a reasonable accuracy. The spectrum analyzer display and corresponding measurement results are shown in Figure 4. It should be recognized that the noise generated by the amplifier stages following the DUT should be considered in the overall measurement results. The noise figure contribution of the second stage may be computed by applying the relationship:

$$F_{total} = F_1 + \frac{F_2 - 1}{G_1} + \ldots + \frac{F_n - 1}{G_1 G_2 \ldots G_{n-1}} \tag{11}$$

or, in dB,

$$NF_{total} = 10 \log (F_{total}) \tag{12}$$

where F_n is the noise factor of the nth stage and G_n is the corresponding gain of the added stage (here n = 2 since only one additional second stage was used). Alternately, the chart in Figure 4 may also be used to determine the noise contribution of a second stage. The corrected noise figure in Table 1 reflects the contribution made by the second stage.

Note that in this example the corrected noise figure is nearly equal to that obtained by simply neglecting the second stage contribution. As a rule of thumb, if the gain of the first stage is 15 dB or greater, the second stage NF contribution may usually be considered negligible.

Conclusion

The fundamentals of noise figure measurement theory have been reviewed and the Y factor noise figure measurement technique has been presented. The availability of a calibrated diode noise source and a modern spectrum analyzer enable the RF engineer to perform basic noise figure characterization when a noise figure meter is unavailable. Additionally, the Y Factor technique is most useful in characterizing the noise figure of narrowband devices. *Several HP41 calculator programs are available from the author to aid in NF calculations — please include an SASE.*

The author wishes to thank Gordon Heal of NADC, whose technical suggestions and inspiration made this article possible.

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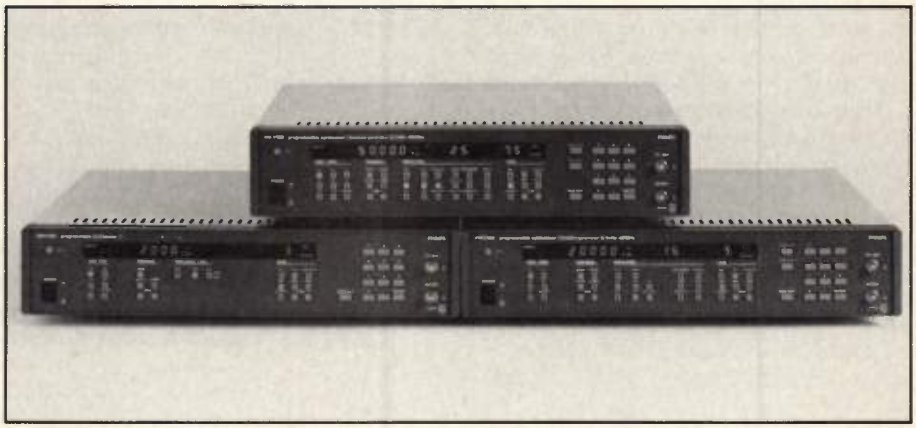
Paul E. Drexler is an RF/microelectronics engineer at the Naval Air Development Center. He is involved in the development and miniaturization of various HF, VHF and UHF systems and holds a BSEE from Drexel University in Philadelphia, Pa. He can be reached at NADC code 4041, Warminster, PA 18974. Tel: (215) 441-1402.

Synthesizers/Function Generators from Fluke/Philips



A series of Philips synthesizers/function generators combines GPIB programmability with high resolution. The series, available from Fluke, is comprised of the Philips PM 5191 SM, PM 5192 SM, PM 5193 SM and PM 5193 VM. The equipment is designed for use in applications where it is necessary to synchronize two or more instruments such as generators and counters. These function generators can be synchronized at 10 MHz or any sub-harmonic of this frequency. This allows a complete set-up of different instruments to be synchronized with a single central clock, such as an in-house standard. In addition to all standard bus functions, the units support bus learn mode and can transmit status and setting information to the controller at any time, making the instruments suitable for flexible, bus-based automatic test systems.

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Price for the PM 5191 SM is $3,695, PM 5192 SM is $4,295, PM 5193 SM is $4,895 and PM 5193 VM is $5,895. **John Fluke Manufacturing Co., Inc., Everett, WA. INFO/CARD #219.**

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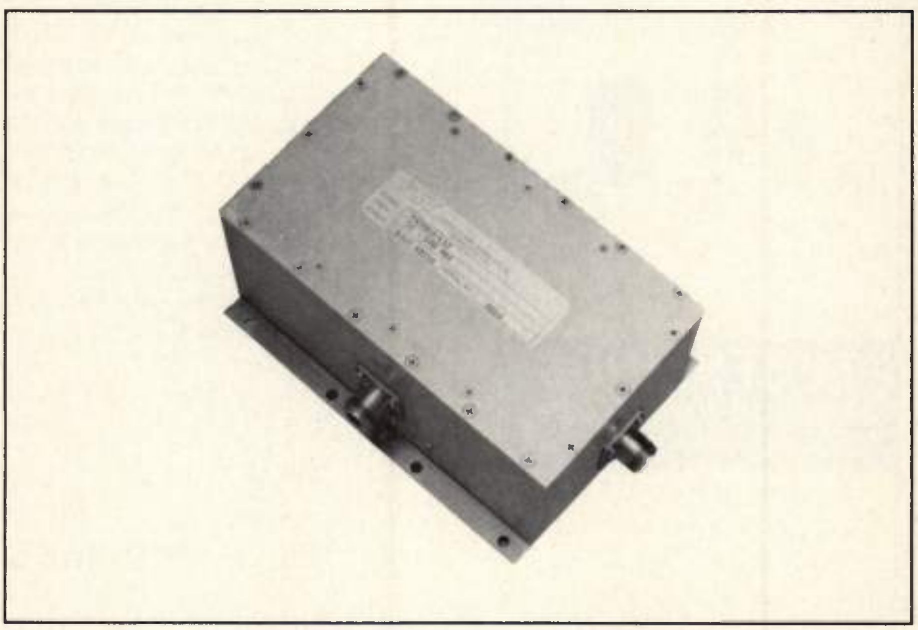

Model TR 2010-152 is a high power PIN diode transmit-receive switch that operates from 20 to 100 MHz with a built-in high speed driver. The switch can handle 1500 watts CW, can switch in less than 1 us and is used to switch an antenna between a sensitive receiver and a high power transmitter.

Maximum antenna VSWR is 2:1 and insertion loss is 0.1 dB (typ) from transmitter to antenna while isolation from transmitter to receiver is 85 dB min. Insertion loss increases to 0.2 dB (typ) from antenna to receiver and isolation from antenna to transmitter is 55 dB min. Maximum VSWR from transmitter to antenna is 1.2:1 and from antenna to receiver it is typically 1.2:1. Command input is two-line balanced TTL and switching speed is typically 500 ns, while maximum switching rate is 20 kHz. The switch measures 5" × 7" × 3.5". **Power Systems Technology, Inc., Hauppauge, NY. INFO/CARD #218.**

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Chip Inductors

Dale's IMC-1812 series of inductors provides an inductance range from 0.01 to 1000 uH. Tolerance is 20 percent from 0.01 uH to 0.39 uH and 10 percent from 0.47 uH to 1000 uH. Tolerances of 5 and 3 percent are available upon request.

Also from Dale is the IMS-2 shielded inductor with a range of 0.1 uH to 560 uH. It is designed to meet the requirements of MIL-C-15305 and has a standard tolerance of 10 percent. A typical IMS-2 (10 uH at 10 percent tolerance) is priced at 80 cents each when purchased in quantities of 500. **Dale Electronics, Inc., Columbus, NE. INFO/CARD #215.**

Yagi Satcom Antenna

This unit features a crossed-yagi design with frequency range of 240 to 320 MHz. Polarization is right or left-hand circular with 1.5:1 VSWR. Power is 200 watts CW and RF impedance is 50 ohms. The FCY-2432B/1 has a gain of 9.5 at 240 MHz and gain of 14.5 at 320 MHz. Other available gains include 8.5 and 6 dB at 240 MHz and 13.0 and 10.0 dB at 320 MHz. Length ranges from 96 inches for the FCY-2432B/1 to 60 and 36 inches for other gain levels. **Austron Corporation, Herndon, VA. INFO/CARD #214.**

ECL Gate Array

Sony introduces a high speed 200-gate ECL gate array. Applications for the E3G200 include digital oscilloscopes, high speed testing, high speed computers and lightwave communications. Typical internal gate delay is 150 ps/gate at IEF of 400 uA and maximum toggle frequency is 2.5 GHz. Available packaging is 24-lead and 32-lead flatpacks. **Sony Corp. of America, Cypress, CA. Please circle INFO/CARD #213.**

Pulse Compression Subsystem

The multimode pulse compression subsystem processes three switchable linear FM waveforms with greater than 14 dB processing gain. Dynamic range is 65 dB min and mode select is TTL driven. Power consumption is 11 watts. **Phonon Corp., Simsbury, CT. INFO/CARD #212.**

Hybrid A/D Converter

Comlinear Corp. and TRW LSI Products have introduced a hybrid analog-to-digital converter which provides a resolution of 12-bits and a +70 dB signal to noise ratio at a sampling rate of 10 million samples per second. The CLC925 (Comlinear)/THC1202 (TRW) is packaged in a 40-pin DIP, has total harmonic distortion of −64 dBc and signal to noise ratio of +70 dBc. DC differential linearity error is 1/2LSB and the integral linearity error is 1/2LSB. Features include a "gain adjust" pin which allows the voltage gain of the circuit to be adjusted ±10 percent over the full scale range. Typical power consumption for this device is 5.3 watts and the device requires three supplies (+5 V, +15 V and −5.2 V). Harmonic distortion at 5 MHz is −68 dBc and −3 dB analog bandwidth is 70 MHz (min). Typical applications include radar, medical electronics, optical electronics and instrumentation. In single quantities, the CLC925A1 is priced at $1,120. **Comlinear Corp., Fort Collins, CO; TRW LSI Products Inc., La Jolla, CA. INFO/CARD #211.**

Digital Sampling Oscilloscope

Tektronix introduces a line of 20 GHz digital sampling oscilloscopes that feature multi-channel differential time domain reflectometry (TDR) measurement capability. The 11800 Series oscilloscopes include the 11802 four-channel mainframe, the 11801 eight-channel mainframe, the SD-24 dual channel/TDR sampling head, the SD-26 dual channel sampling head plug-in unit and the SM11 multi-channel unit. Also available is the P6150 9 GHz passive probe. The 11801 and 11802 combine 17.5 ps or less risetime with 200 KS/s sampling rate and eight-bit vertical resolution at all vertical sensitivities. They can acquire record lengths of 512 to 5120 points. **Tektronix, Inc., Beaverton, OR. INFO/CARD #210.**

GaAs DDS

The ADS-2 GaAs direct-digital synthesizer (DDS) covers 180 MHz of bandwidth with better than 6 Hz resolution. An internal SAW oscillator provides the system clock that can be locked to an internal or external 10 MHz reference. Spurious signals are less than −45 dBc, while harmonics are −25 dBc and an option reduces the harmonics to the same level as the spurs. Phase noise is less than −110 dBc/Hz at 100 Hz, −115 dBc/Hz at 1 kHz, −130 dBc/Hz at 10 kHz and less than −140 dBc/Hz at 100 kHz (measured at 150 MHz). The device offers a switching speed of 20 ns between any two output frequencies. Price ranges from $2,000 to $3,500 depending on quantity and configuration. **Sciteq Electronics, Inc., San Diego, CA. INFO/CARD #209.**

Ceramic-Metal Tetrode

The TH 582 tetrode is a ceramic-metal tetrode of coaxial structures designed for use in amplifiers delivering up to 22 kW of output power at peak-of-sync. The amplifiers are for use in UHF TV transmitters. **Thomson Electron Tubes and Devices Corp., Dover, NJ. INFO/CARD 208.**

Analog Array

The DataLinear Division of Sipex introduces the SP1104 analog array. The features include dielectric isolation; complementary, bipolar vertical NPN and PNP transistors; thin-film nichrome resistors; low temperature coefficient; and dual-layer aluminum metallization. The NPN transistor characteristics feature a beta of 200 and FT of approximately 1 GHz while the PNP features a beta of 100 and FT of 600 MHz. Specifications include a bandwidth of 30 MHz, gains to 120 dB and offset voltages of less than 100 uV. The array is available in either 20 volt or 35 volt transistor Vceo's. Non-recurring prices for the SP1104 start at under $20,000. **Sipex Corp., Milpitas, CA. INFO/CARD #207.**

Flash Converter Evaluation Board

Designed to ease observing and analyzing the performance of Analog Devices' AD770 flash A/D converter, the ADEB770 evaluation board provides the support circuitry required when using this eight-bit, 200 megasample device. Analog signal inputs are applied directly to the converter or via an AD9611 high-speed buffer amplifier. Digitized outputs are available through high-speed 100 K ECL latches and drivers. Output data is provided at 200 MSPS. Jumper selectable decimation provides every second, fourth, eighth or 16th digital output sample for users that cannot accept data at that rate. The ADEB770 evaluation board requires +5 V at 600 mA and −5.2 V at 1.8 A. Price ranges from $635 to $725 depending on grade. **Analog Devices, Norwood, MA. INFO/CARD #206.**

500 MHz Waveform Digitizer

The Model 6880B/6010 waveform digitizing system from LeCroy combines 11-bit single shot resolution with 1.35 gigasample/second digitizing rate, 500 MHz bandwidth (typical) and 10,000 point memory. It offers greater than 14-bit resolution for repetitive waveforms. The minimum system consists of one 6880B single shot digitizing front-end module and one 6010 controller/IEEE-488 interface module installed in an IEEE-583 (CAMAC) standard instrument mainframe. Either a host computer or video monitor completes the system. **LeCroy Corp., Chestnut Ridge, NY. INFO/CARD #205.**

Airborne Diplexer

Model 6127 diplexer is used at a microwave antenna to combine two transmitters to one antenna output. Lowband is 1426 to 1463 MHz and highband is 1510 to 1546 MHz. Passband loss is 3 dB maximum and mutual isolation is 50 dB minimum. Impedance is 50 ohms and connectors are BNC. Power handling is 15 watts average and 75 watts peak power per channel. Price is $1,575. **Microwave Filter Company, Inc., East Syracuse, NY. INFO/CARD #204.**

Coaxial Adapter

Model PE9273 is a 50 ohm coaxial adapter with type SMA male to SMA male radius right angle coaxial adapter. Frequency range is DC to 18 GHz and VSWR is 1.15:1. The adapter has a passivated stainless steel body with TFE fluorocarbon insulation and a gold plated contact.

Price is $27.16 each in 100 piece quantity. **Pasternack Enterprises, Irvine, CA. INFO/CARD #194.**

CMOS Flash A/D

The ADC-208 is Datel's eight-bit CMOS video flash converter. Specifications include 20 MHz sampling rate and 10 MHz full power bandwidth. The design architecture delivers bit rates to 30 MHz in the burst mode, one shot mode conversion times of 35 ns and power modes of 150 mW. **Datel, Inc., Mansfield, MA. Please circle INFO/CARD #196.**

Radiated Susceptibility Test System

The EMPS-2000 radiated susceptibility test system is designed to meet the 50 kV/meter fast rise time specification (RS05) of MIL-STD-461C/462 Notice 5. The unit comes with a one meter antenna and a 50-100 kV variable pulser with remote control triggering. A two meter antenna is available as an option. The system is priced at $39,900. **R & B Enterprises, West Conshohocken, PA. Please circle INFO/CARD #192.**

Miniature Crystal Oscillator

Piezo Crystal Company introduces the Model 2880065 SC-cut crystal oscillator. The unit features an aging rate of 5×10^{-10} per day and a noise floor of −160 dBc/Hz. Final frequency is achieved within 1×10^{-7} after 5 minutes at 25°C. The time domain stability is better than 1×10^{-11} per second and the vibration sensitivity is 2×10^{-9}/g. The oscillator is available in standard frequencies of 10 MHz to 10.23 MHz. **Piezo Crystal Company, Carlisle, PA. INFO/CARD #195.**

High Power Combiner

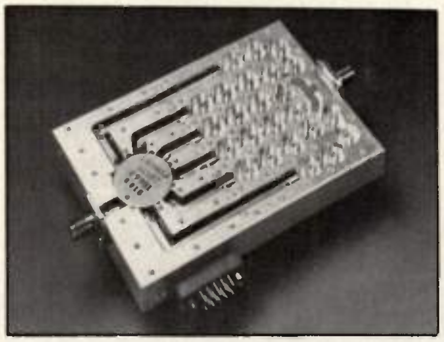
The high power VHF switchless combiner can combine and/or switch two transmitter inputs to one or both of the combiner outputs. The switching operation is done under full power without interruption of programming. Model 55005 has a frequency range of 174 to 216 MHz and combined output power of 100 kW. **Micro Communications, Inc., Manchester, NH. INFO/CARD #203.**

Voltage Controlled Crystal Oscillator

Piezo Technology offers a hybrid voltage-controlled crystal oscillator housed in a four-pin package and capable of operating at a frequency of 44.736 MHz. Model CXH-1004 has a 10K ECL output signal with linearity better than ±10 percent. Maximum rise and fall time is 2 ns and frequency stability is ±25 ppm from −40°C to 70°C. **Piezo Technology, Inc., Orlando, FL. INFO/CARD #202.**

Dual Directional Coupler Detector

Sage introduces a dual directional coupler-detector which provides a 3 ±0.15 volt output with 40 watts input over the band. Model FCD4147 incorporates potentiometers which permit the adjustment of output voltages to calibrate out system meter errors. Directivity is greater than 20 dB for both forward and reverse signal outputs. Insertion loss is less than 0.2 dB, VSWR is less than 1.2:1 and maximum input power is 100 watts CW. **Sage Laboratories, Inc., Natick, MA. Please circle INFO/CARD #201.**



Filter/Multiplexer to 20 Channels

Custom switch/mulitplexer combinations are available in a variety of bands. The I band, six-channel switch/mulitplexer has 1 percent bandwidth and 70 dB isolation. VSWR is 1.8:1 and the SP6T switches in 100 ns (max). **Teledyne Microwave, Mountain View, CA. INFO/CARD #199.**

Analog Design and Analysis Program

Lincad-PC V 2.0 is an analog electronic circuit design and analysis program for PC/DOS, MS/DOS or compatible computers. The features of this version include poles, zeros, characteristic equation, open and short circuit time constants, pulse response graphics, Fourier spectra graphics and standard one percent circuit element substitution. It will plot up to four variables on any of the frequency, time or spectra graphics. The program costs $159. **Sofcad Electronics, Inc., Columbus, OH. INFO/CARD #189.**

GPIB Software

A software package from Philips turns a PC into a GPIB controller by reducing programming and debugging time. The Philips PM 2240/001 TestTeam software provides programmers of GPIB instrumentation systems with a versatile environment to develop GPIB application programs. It features standard functions, libraries and an easy-to-use editor. All facilities are available via pull-down menus and pop-up windows. In addition to providing an environment for writing test and measurement programs, the software allows program testing with facilities for syntax checking, interactive debugging and error handling. The program is priced at $925. **John Fluke Mfg. Co., Inc., Everett, WA. INFO/CARD #188.**

AutoCAD and HP Interface to MiCAD

Two translator programs from EEsof allow Autodesk's AutoCAD™ and Hewlett-Packard's EGS™ to directly interface with MiCAD™. ASM 1000 translates MiCAD's mask files to AutoCAD DXF file format so that designers can utilize MiCAD's CAD functions while integrating circuit layout files directly into AutoCAD. The ASM 1200 likewise translates MiCAD mask files into the HP EGS archive file format. The 1000 operates on PCs and compatibles while the 1200 requires the HP 9000 series 300 workstations. **EEsof, Inc., Westlake Village, CA. Please circle INFO/CARD #187.**

Mathematical Software Packages

The Minpack1-Lib focuses on solving systems of nonlinear equations and nonlinear least squares problems. The latter capability allows the user to perform a least squares fit of virtually any nonlinear curve to data. It can solve algebra, calculus and trigonometry problems. The package costs $465.

FITLIB, another package from PC Scientific, is a curve and surface fitting package that is based on tension splines which provide the flexibility for tightness or looseness of fit. Source code is included to allow the subroutines to be customized for specific purposes and uses. The FITLIB software package sells for $695. **PC Scientific Software, Inc., St. Paul, MN. INFO/CARD #186.**

Network Analysis Software

Solution #5 allows design data to be analyzed with an operator interface and calculation feature set similar to those found on network analyzers. S-parameters can be obtained either from CAE files or HP network analyzers. Data can be displayed in the frequency or time domain in any of 13 formats. The software operates on the HP Series 200 and 300, and IBM-AT compatible machines. **Innovative Measurement Solutions, Atlanta, GA. INFO/CARD #167.**

EMI Filter Catalog

This EMI filter catalog contains information on a broad range of EMI/RFI filters and accessories. Included are general descriptions, insertion loss values, schematics and ordering information on general application, switching transient, three-phase and connector series filters as well as IEC power receptacles. Specifications listed include voltage rating, line frequency, leakage current, test voltage, current overload tests, insulation resistance and performance. Mechanical details include case styles and dimensions, panel cutout sizes, terminal types and sizes and attachment data. Components also described include fuses, switches, power connectors, voltage selectors and indicators. A glossary of terms is included. **Stanford Applied Engineering, Santa Clara, CA. INFO/CARD #185.**

Simulation Tool Description

A description of OmniSys, a system simulation tool, is provided in this booklet. This technical summary from EEsof discusses such details as characterization of nonlinear components, external and internal component models and component model equations. Included are three technical examples to illustrate the capabilities of the software. **EEsof, Inc., Westlake Village, CA. INFO/CARD #184.**

Microwave Products and Instruments Catalog

Products covered in this catalog include attenuators, couplers, power dividers, mechanical switches, control products, sources, MICs, radiation and power monitors, and microwave measurement systems. In addition to alphabetical and numerical listings, quick reference guides introduce each product group followed by technical notes that provide basic function and application data. **Narda Microwave Corp., Hauppauge, NY. Please circle INFO/CARD #183.**

Technical Description of Microwave Cable Assemblies

This description is centered around SUCOFLEX cables. These flexible microwave cable assemblies are designed to substitute semi-rigid cables. The information presented includes a product digest, technical data on the cables, armours for SUCOFLEX, interchangeable connectors, application specialties, mechanical dimensions and ordering information. **Huber + Suhner AG, Herisau, Switzerland. INFO/CARD #182.**

Antenna and Antenna Systems Catalog

This revised catalog presents in-depth application information and product specifications on the company's antenna line. The document also includes an antenna designer's guide, containing a wide range of technical data to assist the systems engineer in determining antenna performance parameters for specific applications. **Watkins-Johnson Company, San Jose, CA. INFO/CARD #181.**

Catalog Describes Cascadable Amplifiers and Signal Processing Components

RF and microwave cascadable amplifiers and signal processing components from 0.3 to 3000 MHz are highlighted in this catalog. Shown are graphs for gain, power, single and double stage noise figure, and third order two-tone intercept point. Parameters listed include small signal gain, gain flatness, noise figure, power output, intercept point, and VSWR. Other information shown includes cascading application information, single and cascaded TO-8 amplifiers with SMA connectors and outline drawings. **Cougar Components, Sunnyvale, CA. Please circle INFO/CARD #180.**

Crystal Oscillators and Quartz Crystals Catalog

Bliley Electric Company has released its catalog detailing high rel crystal oscillators and quartz crystals. Included among the expanded crystal oscillator section are a variety of oven oscillators (OCXO), digitally compensated crystal oscillators (DCXO), as well as voltage controlled (VCXO), temperature compensated voltage controlled (TCVCXO) and clock oscillators. The quartz crystal section includes surface mount designs and SC cut units among general purpose, MIL-Spec, vacuum sealed and ultra high stability components. **Bliley Electric Company, Erie, PA. INFO/CARD #179.**

Application Note Features Using JFETs for Speed and Accuracy

Application note AN-108, "JFET-Input Amps are Unrivaled for Speed and Accuracy," presents high speed application circuits such as fast sample-and-hold amplifiers, logarithmic amplifiers, peak detectors, etc. that are possible with the OP-42 and OP-44 fast settling op amps from PMI. Accompanying each application is a detailed description of how the circuit works along with pertinent performance characteristics. The note includes practical design hints to show users how to take full advantage of the particular products. In addition, a section is devoted to helping designers minimize circuit noise and error. **Precision Monolithics, Inc., Santa Clara, CA. INFO/CARD #178.**

Short Form Catalog Describes Pulse Generators

Avtech Electrosystems introduces its 1988-1989 short form catalog of nanosecond waveform generators and accessories. It describes over 150 models of pulse generators, impulse generators, monocycle generators and accessories such as transformers, power splitters, delay generators, sample and hold amplifiers and scope probes covering the PRF range of 0 to 250 MHz, rise times from 40 psec to 10 nsec, pulse widths from 130 psec to 100 usec and amplitudes from 5 to 500 volts. **Avtech Electrosystems LTD., Ottawa, Ontario, Canada. INFO/CARD #176.**

Filters Brochure

Lark Engineering's brochure gives information on microminiature lowpass, bandpass, and high pass filters covering from 1 MHz to 6000 MHz. Packages are either TO-8 cans or flat packs with axial or radial pins. **Lark Engineering Company, San Juan Capistrano, CA. Please circle INFO/CARD #175.**

Quartz Crystals and Oscillators Catalog

This catalog provides technical specifications on quartz crystals and crystal clock oscillators. Designed for both surface mounting and thru-hole applications, the resonators cover the 10 kHz to 35 MHz range. Package dimensions are shown for hi-temp, leadless surface mount version, as well as for thru-hole units with leads. **Micro Crystal Division/SMH, New York, NY. INFO/CARD #174.**

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8	18	28	38	48	58	68	78	88	98	108	118	128	138	148	158	168	178	188	198	208	218	228
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