rf design

engineering principles and practices

November 1989

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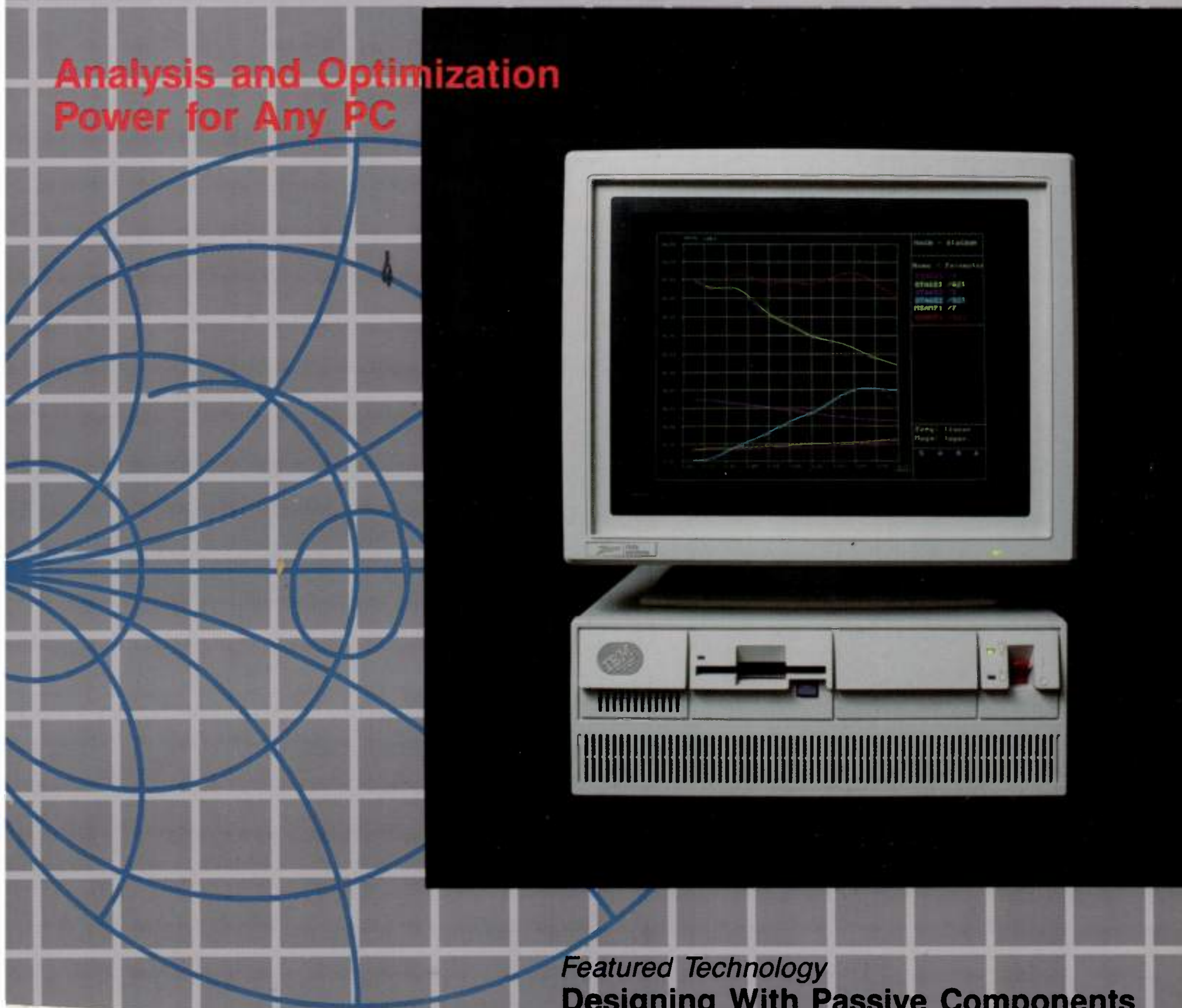

Featured Technology
Designing With Passive Components

Official Show Issue
RF Expo East

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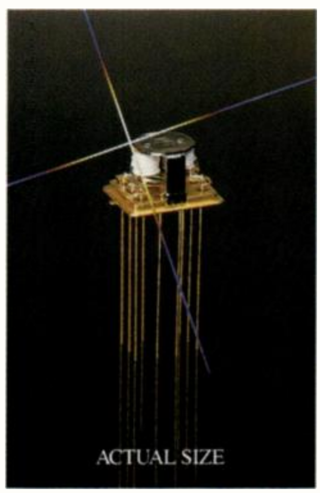
- Military and commercial versions



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INFO/CARD 1

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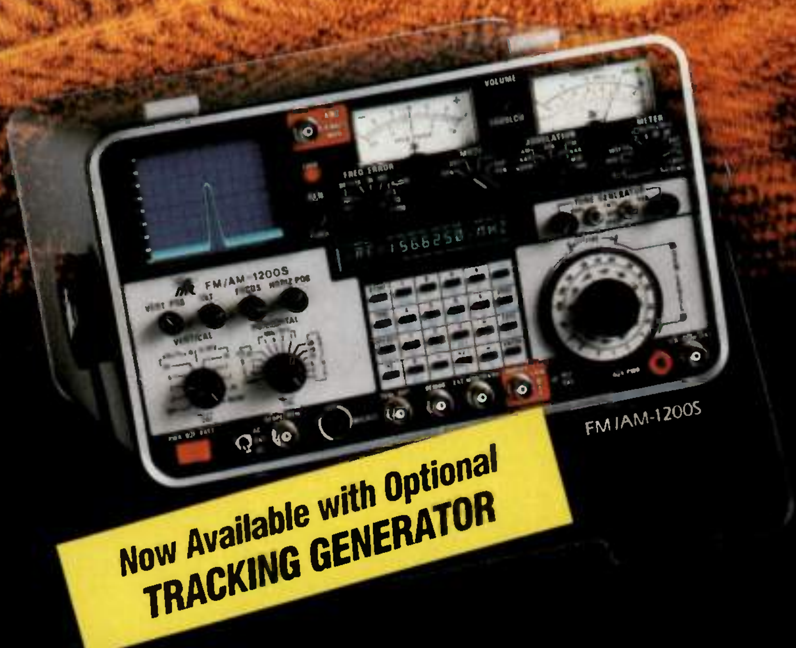

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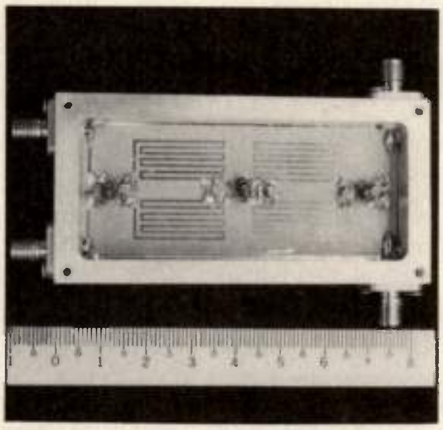
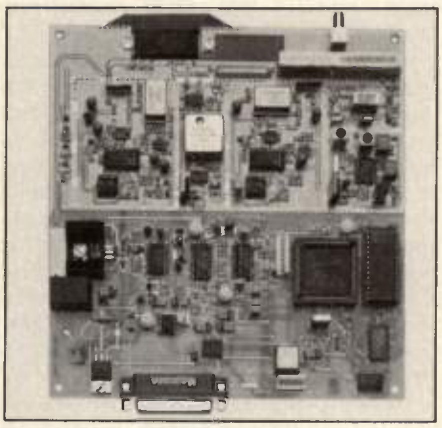
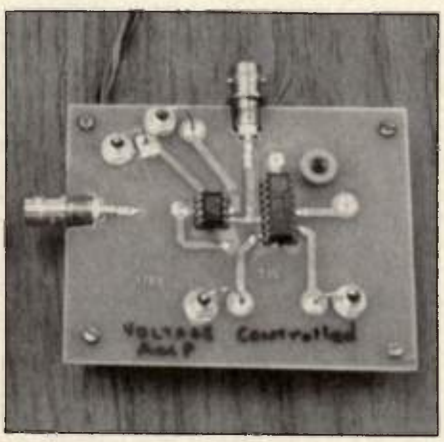
November 1989



Page 58 — Quadrature Couplers



Page 70 — RF Expo East



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rf editorial

Glasnost, Perestroika, People and Technology

By Gary A. Breed
Editor

The rapid and radical changes happening in the Soviet Union and Eastern Europe have been making big headlines. Socialist nations are now admitting that they cannot survive using ideas that were developed about the same time radio became practical. As a result, these formerly closed societies are opening their borders, as well as their internal barriers.

Unlike the situation in China, I believe these changes are irreversible. China's economic goals have not been accompanied by a social reform policy like glasnost, which is directed toward the people. This is a lot more than just propaganda.

One major reason I think these changes are real is from first-hand experience — amateur radio. Ham radio has been popular in the USSR for many years, but with restrictions on the type and content of communications with others. For example, the "QSL" cards that hams exchange after contacting one another had to be sent through a central bureau. Only short conversations were permitted, restricted to basic information of name, location, and a few pleasantries. Like all types of contact with "the outside world," amateur radio was controlled along with the rest of Soviet society.

All that has changed — in a hurry. One of the first events to demonstrate glasnost as a serious policy was the Armenian earthquake. Foreign assistance was welcomed, and for the first time in history, amateur radio operators were allowed to relay personal messages in and out of the Soviet Union. It was once unthinkable that American hams ever would be allowed to operate from the USSR, but the first joint US-USSR operation took place earlier this year, and more are planned. Citizens under Soviet influence have been allowed freer travel, and hams have made trips to operate from some very rare (amateur radiowise) places like Mongolia and Vietnam.

The amount of information exchanged between Soviet hams and the rest of the world is now nearly unlimited. Direct mailing of cards and letters is commonplace, as is the donation of radio and computer equipment to radio clubs in the USSR. Early model personal computers have been sent, along with some very late model radio equipment, including packet radio terminals for digital communications. Soviet hams are anxious to get modern technology, something which they have been denied.

Since a large percentage of hams are engineers and technicians, the taste of current technology they are getting via amateur radio will certainly influence their professional lives. More than ever, they will want components, instruments, design techniques, and engineering tools to meet their product development goals. Many nations will be waiting to take advantage of this new market for technology, especially Western Europe and Japan.

The United States should be looking closely at the political and economic situation in these changing Socialist countries. Reform takes time, but each passing day increases the chance that these changes are permanent. We need caution when dealing with a long-time enemy, but we should also have an eye on marketplace realities. We should be ready for battle — not for World War III, but for sales wars as the world clamors to serve a new customer.

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DTO		
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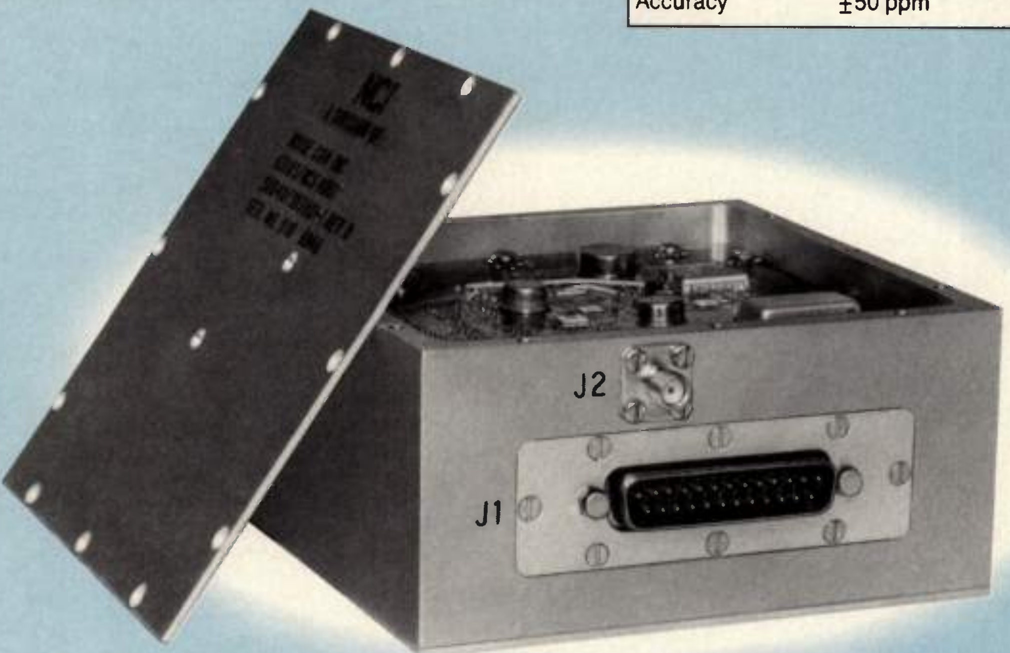
SYNTHESIZER 1 MHz resolution, 10 Hz to 1 MHz modulation		
Stability	±0.03%	50 ppm
Accuracy	±50 ppm	±10 ppm



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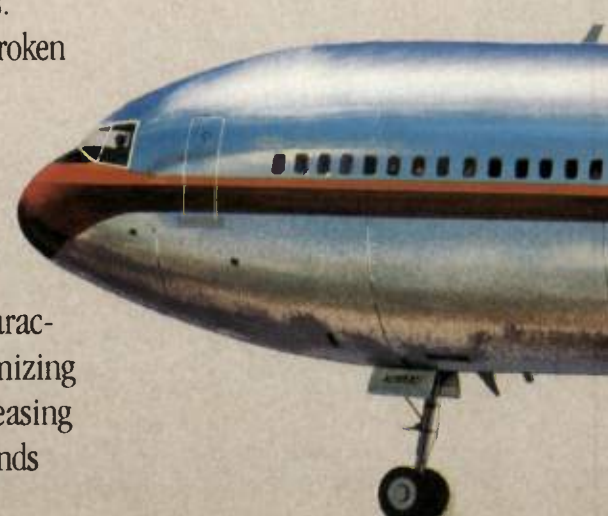

TION SCAN, JAM, TRACK, CALL, ACRIAN RF TRANSISTORS.

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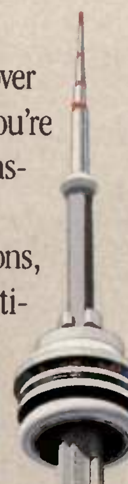

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rf letters

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

Thoughts on De-Specialization

Editor:

Many of us can remember when RF engineers needed to be concerned with little other than their principal specialties. This situation began to erode with the integration of digital control circuits into RF equipment, a trend greatly accelerated by the advent of the microprocessor. Today, it seems a legitimate debate could be conducted as to whether a modern radio or test instrument is primarily an RF device or a computer. The problem is that most RF engineers lack the skills to design and develop such micro-controllers. So, a digital designer must be relied upon to handle this aspect of product development. In many instances, a development effort would proceed more smoothly if RF engineers could handle this phase of the effort on their own.

There are precedents for such de-specialization of effort. At one time, for example, few RF engineers were involved with programming; however, most are now able to handle routine programming needs on their own. And the emergence of PC-based CAD is encouraging more RF engineers to layout their own PC boards. In both cases, the efficiency lost to de-specialization is more than offset by better continuity and reduction in human interface problems.

Recently, new micro-controller development tools have emerged that offer the user a more simplified development environment that can be controlled by a PC. Specifically, BASIC language development environments have appeared for the Intel 8051 series of micro-controllers, and the better of these would likely work very well in many RF applications. Since most RF engineers understand and are comfortable in BASIC, is the time right for articles explaining the mysteries of micro-controller development to RF engineers?

Alex J. Burwasser
R.F. Products
San Diego, California

rf calendar

November 14-16, 1989
Wescon/89
Moscone Convention Center, San Francisco, CA
Information: Wescon/89, 8110 Airport Boulevard, Los Angeles, CA 90045-3194. Tel: (213) 772-2965

November 16-17, 1989
U.S. Cellular Standards: The Digital/Analog Evolution
Inner Harbor, Baltimore, MD
Information: Suzanne Mullendore, TIA Mobile Communications Division, 1722 Eye Street N.W., Suite 440, Washington, DC 20006. Tel: (202) 457-4937

December 3-6, 1989
1989 International Electron Devices Meeting
Washington Hilton, Washington, DC
Information: Courtesy Associates, 655 15th Street, N.W., Suite 300, Washington, DC 20005. Tel: (202) 347-5900

December 13-14, 1989
Technology '89
Mesa Convention Center, Mesa, AZ
Information: C/S Communications, Inc., P.O. Box 23899, Tempe, AZ 85285. Tel: (602) 967-7444

January 15-18, 1990
SMART VI Conference and Exhibition
Buena Vista Palace Hotel, Orlando-Lake Buena Vista, FL
Information: EIA Components Group, 1722 Eye Street, N.W., Suite 300, Washington, DC 20006. Tel: (202) 457-4930

February 26-March 1, 1990
Electronic Imaging West 90
Pasadena Convention Center, Pasadena, CA
Information: MG Expositions Group, 1050 Commonwealth Avenue, Boston, MA 02215. Tel: (800) 223-7126; (617) 232-3976

March 26-29, 1990
Space Commerce 90
Montreux, Switzerland
Information: Space Commerce 90, P.O. Box 97, CH/1820 Montreux, Switzerland. Fax: +41 21 963 78 95

March 27-29, 1990
RF Technology Expo '90
Anaheim Convention Center, Anaheim, CA
Information: Kristin Hohn, Cardiff Publishing Company, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111. Tel: (303) 220-2600; (800) 525-9154

April 2-5, 1990
35th International SAMPE Symposium and Exhibition
Anaheim Convention Center, Anaheim, CA
Information: SAMPE, P.O. Box 2459, Covina, CA 91722. Tel: (818) 331-0616

rf courses

The George Washington University

Radar Operation and Design: The Fundamentals
November 27-30, 1989, Washington, DC
Spread-Spectrum Communications Systems
December 4-8, 1989, Washington, DC
Fiber-Optic Technology for Communications
December 5-7, 1989, Washington, DC
New HF Communications Technology: Advanced Techniques
December 11-15, 1989, Washington, DC

Information: Misael Rodriguez, Continuing Engineering Education, George Washington University, Washington, DC 20052. Tel: (800) 424-9773; (202) 994-6106

R & B Enterprises

Understanding and Applying MIL-STD-461C
November 15-17, 1989, Philadelphia, PA
Introduction to EMI for Non-EMI Personnel
December 4-5, 1989, Orlando, FL
Real-Life Solutions to EMI Problems
December 6-8, 1989, Orlando, FL

Information: Registrar, R & B Enterprises, 20 Clipper Road, West Conshohocken, PA 19428. Tel: (215) 825-1966

Compliance Engineering

EMI
December 5, 1989, New Brunswick, NJ
Safety
December 6, 1989, New Brunswick, NJ
ESD
December 7, 1989, New Brunswick, NJ
Telecom
December 8, 1989, New Brunswick, NJ

Information: Compliance Engineering, 629 Massachusetts Avenue, Boxboro, MA 01719. Tel: (508) 264-4208

EEsof, Inc.

Computer-Aided Engineering for Linear Microwave Circuits (Touchstone)
November 13-15, 1989, Westlake Village, CA
Computer-Aided Drafting for Microwave Circuits (MiCAD)
November 16-17, 1989, Westlake Village, CA
Computer-Aided Engineering/Drafting for Microwave Circuits (Academy)
December 4-6, 1989, Westlake Village, CA

Information: Sande Scoredos, Training Coordinator, EEsof, Inc., 5795 Lindero Canyon Road, Westlake Village, CA 91362. Tel: (818) 991-7530, ext. 197

Hewlett-Packard Co.

Designing for Electromagnetic Compatibility (EMC)
November 14-15, 1989, Las Colinas, TX
December 4-5, 1989, Santa Clara, CA
December 11-12, 1989, Mountain View, CA
December 14-15, 1989, Mountain View, CA

Information: Hewlett-Packard Co., 3000 Hanover Street, Palo Alto, CA 94304. Tel: (800) 2HP-EDUC

Learning Tree International

Introduction to Datacomm and Networks
November 28-December 1, 1989, San Diego, CA
December 12-15, 1989, Washington, DC
C Programming Hands-On Workshop
November 28-December 1, 1989, Boston, MA
December 12-15, 1989, Seattle, WA

Information: John Valenti, Learning Tree International, 6053 W. Century Boulevard, P.O. Box 45974, Los Angeles, CA 90045-0974. Tel: (800) 421-8166; (213) 417-8888

Interference Control Technologies, Inc.

Fundamentals of EMI and EMC
November 28-30, 1989, Las Vegas, NV
Grounding and Shielding
December 5-8, 1989, Washington, DC
EMC Design and Measurement
December 11-15, 1989, Orlando, FL
Practical EMI Fixes
January 22-26, 1990, Orlando, FL

Information: Registrar, Interference Control Technologies, Inc., State Route 625, P.O. Box D, Gainesville, VA 22065. Tel: (703) 347-0030

Compact Software, Inc.

A Systematic Approach to Microwave Circuit Design
November 14, 1989, Dallas, TX

Information: Helen Shapiro, Compact Software, Inc., 483 McLean Boulevard, Paterson, NJ 07504. Tel: (201) 881-1200

Design and Evaluation, Inc.

Worst Case Circuit Analysis
December 4-6, 1989, Orlando, FL

Information: Design and Evaluation, Inc., 1451B Chews Landing Road, Laurel Springs, NJ 08021. Tel: (609) 228-3800

National Semiconductor

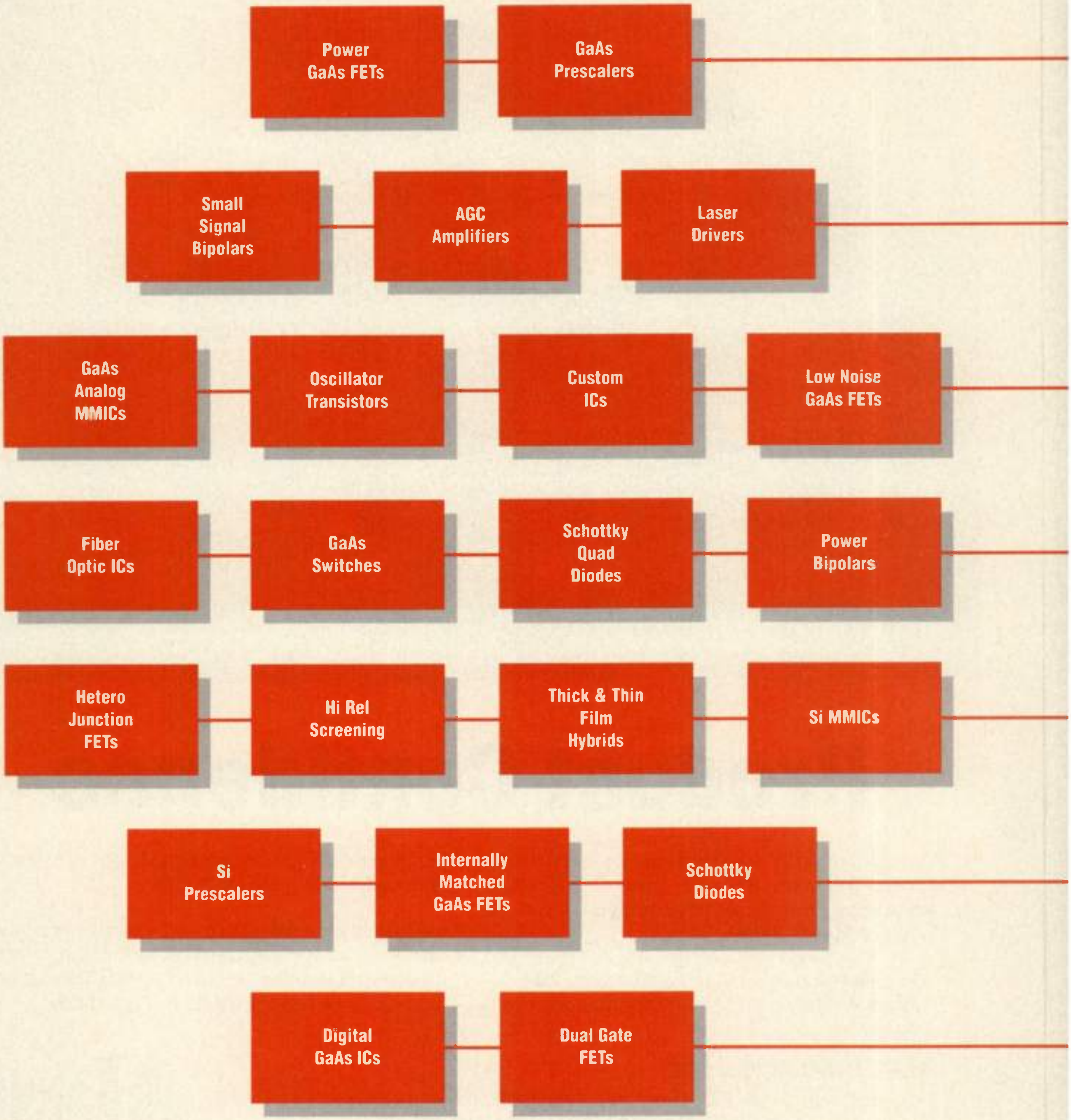
Linear Applications Seminar
November 15, 1989, Phoenix, AZ
November 16, 1989, Albuquerque, NM
November 17, 1989, Denver, CO
November 28, 1989, Anaheim, CA
November 29, 1989, Los Angeles, CA
November 30, 1989, Woodland Hills, CA

Information: National Semiconductor, 2900 Semiconductor Drive, P.O. Box 58090, Santa Clara, CA 95052-8090. Tel: (800) 548-4529; (408) 721-5000

Praxis International, Inc.

Facility Grounding, Shielding and Lightning Protection
November 15-17, 1989, Washington, DC
EMI Design and Analysis
December 5-7, 1989, Denver, CO

Information: Praxis International, Inc., Exton Professional Building, Suite 103, 319 N. Pottstown Pike, Exton, PA 19341.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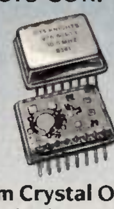

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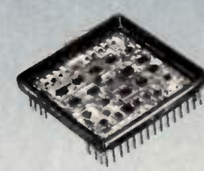

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Banquet Honors New Jersey RF/ Microwave Supporters

The New Jersey RF and Microwave Industries Association honored contributors to the Center for Microwave and Lightwave Technology at New Jersey Institute of Technology (NJIT) at the association's second annual awards banquet October 6, 1989. Over one million dollars has been raised to establish the new state-of-the-art laboratory at NJIT.

The contributors include individuals, manufacturers, the New Jersey Council of the American Electronics Association (AEA), the New Jersey Commission on Science and Technology, and the National Science Foundation. All have donated time, capital or equipment to the microwave and lightwave laboratory, which is chartered to provide a hands-on learning environment for graduate and undergraduate students at NJIT. The laboratory will enable students to learn skills in microwave theory and techniques, as well as lightwave principles. The university has committed a floor in its new engineering center to the Microwave and Lightwave Technology Laboratory, upon completion of the facility in 1990.

Speakers at the awards banquet included Dr. Saul Fenster, president of the New Jersey Institute of Technology, who presented awards to nearly two dozen contributors. The New Jersey RF and Microwave Industries Association is a non-profit organization that was formed in 1987 to achieve three goals: to build interest in microwave technology within the undergraduate and graduate academic environment; to bring together New Jersey's RF and microwave companies in a quarterly forum; and to present distinguished speakers from throughout the academic, industrial and government communities.

Dr. Saul Fenster, president, New Jersey Institute of Technology; Robert Asdal, executive director, New Jersey Council of the American Electronics Association; and John J. McCrea, co-founder, New Jersey RF and Microwave Industries Association (left to right).

RF Technology Expo 90 Call for Papers—Authors are invited to present a paper at the technical program at RF Technology Expo 90, to be held March 27-29, 1990 in Anaheim, Calif. Proposals for individual papers or complete three-hour sessions are welcome. Topics of current interest to RF engineers include: RF integrated circuit applications; digital signal processing; high-performance circuits; electromagnetics modeling; frequency synthesizers; power amplifiers; circuit analysis and design; and complex modulation.

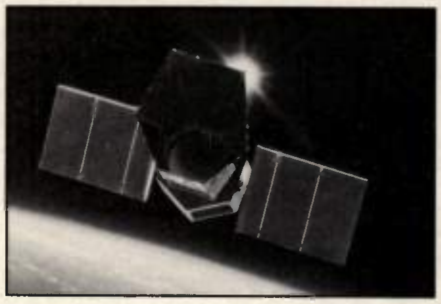
In addition to these topics, quality presentations on almost any RF subject have been well-received at previous RF Expos. Of special interest are fundamental tutorial papers on classic RF topics: oscillators, amplifiers, filters, modulators, mixers, antennas, etc.

Interested authors are urged to submit a proposal as soon as possible; the final program will be determined in mid-December. Send an outline or abstract of the proposed paper to: Gary A. Breed, Editor, RF Design Magazine, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111. Fax: (303) 773-9716.

Call for Frequency Control Papers—A call for papers has been issued for the 44th Annual Frequency Control Symposium, to be held May 23-25, 1990 in Baltimore. Authors are invited to submit papers dealing with recent progress in research, development and applications in areas represented by the following topics: fundamental properties of piezoelectric crystals; theory and design of piezoelectric resonators; resonator processing techniques; filters; surface acoustic wave devices; quartz crystal oscillators; microwave and millimeter-wave oscillators; synthesizers and other frequency control circuitry; atomic and molecular frequency standards; noise phenomena and aging; frequency and time coordination and distribution; sensors and transducers; applications of frequency control; and measurements and specifications.

Two copies of a paper summary (at least 500 words), together with the author's name, address and telephone number, should be sent to: Dr. Thomas E. Parker, Raytheon Research Division, 131 Spring Street, Lexington, MA 02173. Deadline for submission is January 10, 1990. Further information on the symposium can be obtained by contacting Dr. R.L. Filler, U.S. Army Electronics Technology and Devices Laboratory, Attn.: SLCET-EQ, Fort Monmouth, NJ 07703-5000. Tel: (201) 544-2467

Ball Receives Contract for Radio Astronomy Payload—Ball Corp. has been awarded a $14.5 million contract from the Smithsonian Astrophysical Observatory for the instrument design and fabrication for the Submillimeter Wave Astronomy Satellite (SWAS). As prime contractor, Ball's Electro-Optics/Cryogenics Division, Broomfield, Colo., will build, integrate and test the radio astronomy payload. Several of Ball's other



aerospace operations will also be involved in the instrument system: the Communication Systems Division will do radio receiver design analysis; the Efratom Division will make the oscillator used for precise timing control for the receiver; and the Space Systems Division will develop spacecraft interface specifications. Ball will also provide the star tracker for spacecraft navigation.

Two subcontractors, Millitech Corp. and the University of Cologne, will supply the submillimeter radio receiver and acousto-optical spectrometer for on-board signal analysis, respectively. The satellite's mission is to measure the chemistry of dense, interstellar clouds using submillimeter lines of water, molecular oxygen, carbon atoms and carbon monoxide. This information will aid scientists in advancing the theory of star formation. The instrument is scheduled for delivery in October 1992, with the tentative satellite launch date set for 1993.

Rohde and Schwarz to Supply Munich Airport DF System—Rohde and Schwarz has been selected by the Federal Administration of Air Navigation Services in West Germany to provide direction-finding (DF) and radio equipment for air traffic control installations at the new airport Munich II. The DF system will consist of direction finders from 118 to 400 MHz. There will be six parallel channels for VHF and UHF each, as well as one VHF and one UHF distress frequency channel. Two VHF/UHF channels can be switched to every in-service channel to replace it in the event of failure. By means of two interface multiplexers, the bearings obtained in the 12 in-service channels are converted and adapted to the radar data format and levels, and are then transferred to each radar operator position via the communication switching system. The bearings are used for additional echo identification on the radar screens. The system will be installed at Munich II in the fall of 1990.

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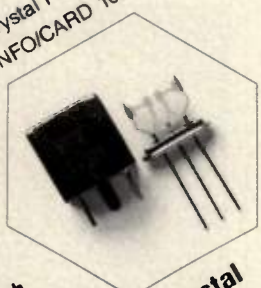

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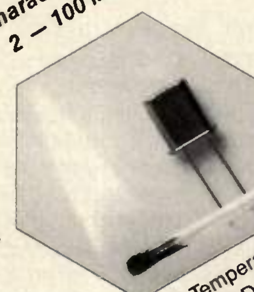

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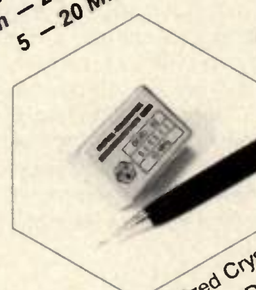

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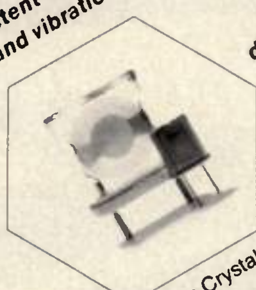

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Hughes Team Selected for Radar Warning System Upgrade—The Radar Systems Group of Hughes Aircraft Co. has been awarded a multi-million-dollar U.S. Navy contract to upgrade the Navy's AN/ALR-67 radar warning system. The Radar Systems Group is teamed with AEL Defense Corp. of Lansdale, Pa., and Phase Two Industries (PTI) of Santa Clara, Calif. The Navy contract calls for production of 14 full-scale engineering development systems of the ALR-67 Advanced Special Receiver (ASR), an electronic warfare system utilized by a number of Navy aircraft. The ASR's mission is to counter surface-to-air threats by expanding frequency coverage, providing more accurate threat detection and increasing range. The new receivers will be installed in the Navy's existing fleet of F-14, F/A-18, AV-8B and A-6 aircraft.

Hughes will provide a form-fit replacement for portions of the existing ALR-67, including the new advanced special receiver and a new processor. AEL will produce half of the 14 systems called for in the contract, and will be qualified as a production second source. PTI will assist in software design.

System Developed to Measure Superconductivity—NKK Corp. of Japan and Yokogawa Hewlett-Packard (YHP) have jointly developed a measuring and analysis system to test materials for degree of superconductivity. The new system measures the factors involved in inducing a superconductive state 10 times faster than conventional systems. Previously, analysis of a single sample has taken up to a full day to perform. The shortened analysis time offered by the NKK/YHP system should facilitate scientific research into superconductivity. The system consists of YHP's HP 9000 16-bit microcomputer connected to a 14 in. diameter cryosat. A sample, placed in a holder in the cryosat's center, is charged with a magnetic field while the screening current generated within the sample is measured. The microcomputer then analyzes the sample's superconductivity based on relationships among critical current density, critical magnetic field and critical temperature. Measuring hardware for the system was developed by NKK, and the automated data analyzing system by YHP.

Power Meter Measurement Techniques Described—A publication from the National Institute of Standards and Technology describes techniques for evaluating high-frequency power meters. Measurements at RF, microwave and millimeter frequencies are affected by many factors. Impedance mismatch, interference, leakage, nonlinear effects and other sources of error must be assessed and minimized. *Performance Evaluation of Radiofrequency, Microwave, and Millimeter Wave Power Meters* describes measurement techniques for evaluating the electrical performance of certain commercially available power meters that use bolometric sensors and operate typically from 10 MHz to 26.5 GHz. The publication is available from the Superintendent of Documents, U.S. Government Printing Office, Washington, DC 20402. Order by stock no. 003-003-02931-9 for $8.50 prepaid.

PCs Spell Growth for RF Software

By Katie McCormick
Assistant Editor

There's no doubt about it — developers and vendors of software for computer-aided design (CAD) are confident and excited about what lies ahead for the field of RF design software. They describe a field with tremendous room for growth in both capabilities and influence. The advent of more powerful personal computers, coupled with more affordable prices, promises to provide the design engineer with greatly expanded capabilities.

According to Randy Rhea, president of Circuit Busters Inc., "New PCs are approaching the performance of workstations at the same time that the prices are coming down. The availability of desktop computers with considerable power costing under $2000 is introducing more people to computers." Rhea is one of a number of software developers who see a lot of potential in the market for affordable, competitive, PC-based software for design applications. As Charles Hymowitz, chief applications engineer at Intusoft, states his company's goal, "We try to specialize in giving engineers everything they can get from their workstation at an affordable price and on a PC." Steve Lafferty, president of Tesoft, expresses a similar sentiment: "One of our key goals was to make sure that our product is available to be on every engineer's desktop." For Tesoft, this meant introducing a software package that was easy to use and low in cost. The goal in pricing, Lafferty explains, was to keep the product in the "why not?" category.

Many software developers anticipate that the introduction of Intel 80486-based PCs will result in greatly improved performance capabilities. "With the '486s coming along," Lafferty observes, "it's not clear that the workstations are going to stay dominant." Alan Kafton, marketing communications manager at Hewlett-Packard's Network Measurements Division, characterizes the 386 and 486 machines as computing platforms that are rivaling the performance of low-end (for the 386) and even mid-level (for the 486) workstations. He forecasts "a blurring of the workstation line," and growth in the market for high-end PCs. Ray Pengelly, vice-president of marketing and sales at Compact Software Inc., agrees: "These days it's becoming more and more difficult to know where the so-called PC ends and the workstation starts."

The market for workstation-based software will remain strong, however, according to Tom Reeder, vice-president for marketing at EEsof Inc. He predicts, "The workstation market will continue to grow strongly, certainly 30 to 40 percent per year, maybe more." He explains, "The workstation market is extremely important for the bench engineer." Specifically, because of computing power and time requirements for some applications, there will continue to be a demand for high-performance software for the workstation platform.

The Macintosh

The Macintosh computer is coming of age for RF applications, and designers should expect to see a greater variety of software packages available to them. Bradley Nedrud is owner of Nedrud Data Systems and developer of a Macintosh CAD program. "I think the potential for the Macintosh is super," he says. "The Macintosh is something that has to be considered now when you're buying for a scientific application." Zelko Jagaric, manager of technical support and development at Ingsoft Ltd., stresses the Mac's capabilities for doing the variety of tasks that confront today's engineer. "Our main objective," he explains, "is to introduce one machine per engineer." According to Jagaric, this is where the Mac has a niche to fill.

Both users and developers of Macintosh software point out how easy it is to learn and to use. In fact, says Nedrud, "One of the big selling points for Macintosh software is that you can quickly recoup the cost of the computer and the software by cutting down on the training time." The standardized user interface offered by Macintosh software makes it much easier for a user to get acquainted with new programs. And, says Hymowitz, "Its networking is as good as a workstation, and its graphics are as good."

The Future

For the future, design engineers can expect increased accuracy and better models from RF software packages. Software will keep pace with the advances being made in hardware. "The software will evolve and take advantage of the increased power of these machines," Randy Rhea asserts. "In the future," he continues, "computers, via optimization and numerical techniques, will allow the engineer to attack problems more effectively."

Dick Webb, president of Webb Laboratories, stresses that the new PCs will give workstation performance from DOS. "I think you'll see more accurate software and software with greater utility available for use on the PC and the Mac," he predicts. "With existing and planned language extensions," he continues, "the incentives to move to UNIX and OS/2 will be minimized."

EEsof's Tom Reeder says, "I expect to see useful GaAs bipolar models within the next year, and silicon bipolar models very soon after that." Alan Kafton adds, "In the future, there will be software and high-performance platforms available that will make complete optimization of nonlinear circuits practical, with speeds that rival today's linear optimization." According to Compact's Pengelly, features that will soon be available to the designer include temperature and bias dependency for bipolars and FETs in a nonlinear simulator, and the ability for users to define their own models.

Of the industry in general, Pengelly cautions, "While the totally integrated solution is nice for some people, it is by no means what everybody needs." He stresses the importance of offering the user a package that is priced right, and does what they want and no more than they want. "It is the responsibility of the CAD houses to offer a portfolio that serves the entire industry," he says, and asserts that the RF designer will definitely see more software targeted for his particular applications. rf

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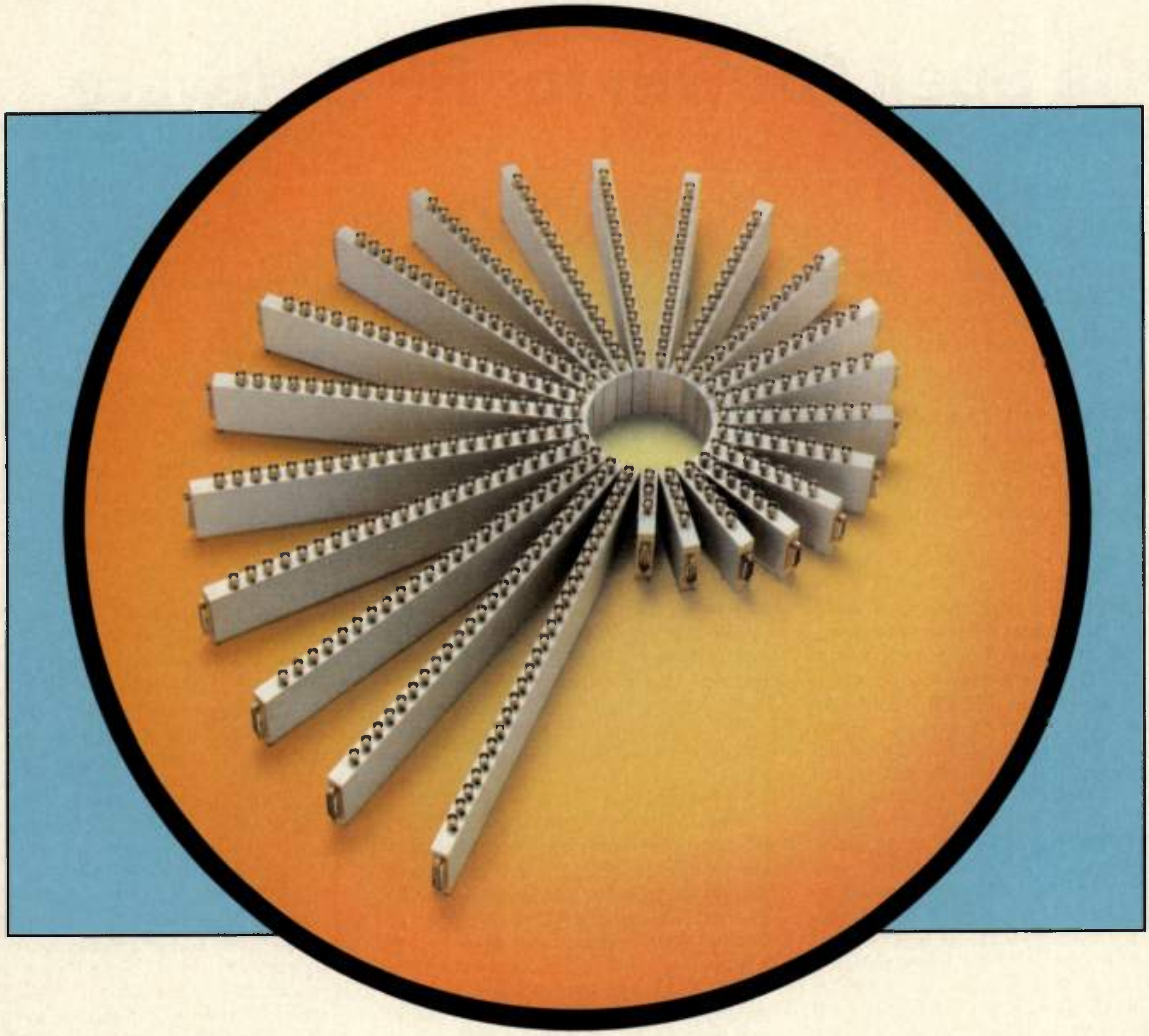

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rf cover story

SANA: Nodal Network Signal and Noise Analysis and Optimization

By Richard C. Webb
Webb Laboratories

Today's engineer is not only tasked with the responsibility for the design of modern, state-of-the-art high frequency components and subsystems, but is also expected to carry out these developments in record time. Furthermore, emphasis on cost, reliability and producibility has never been greater. In today's RF and microwave engineering world, there is little or no time to experiment with designs in either the proposal, prototype or production phases of a program.

A key factor in the success of engineering efforts is the ability to predict the behavior of the hardware of interest, including the numerous anticipated, and unexpected, circuit effects and phenomena. While there is no substitute for sheer engineering skill and experience in the conceptual stages of a design, the ability to analyze and effectively optimize complex lumped and distributed networks has become a necessity. Unfortunately, available software tools often lack the needed analysis capability or are so costly to obtain and support that only a minority of working engineers have access to the needed capability.

In order to deal accurately and thoroughly with today's sophisticated networks, a simulation tool must offer a number of features. First, and consistent with the characteristics of large and complex circuit arrangements, it is imperative that network definition be performed in nodal fashion. With the execution speed achievable with the application of sparse-matrix techniques to admittance parameter reduction, there is no longer a good argument for alternative network interconnection schemes. Second, the simulator must allow for the inclusion of noisy devices in general. That is, uncorrelated noise contributors, such as transistors, must be taken into account such that the noise characteristics of the network are at the engineer's immediate disposal.

Third, a modern CAE package must possess not only a very comprehensive library of distributed and lumped net-

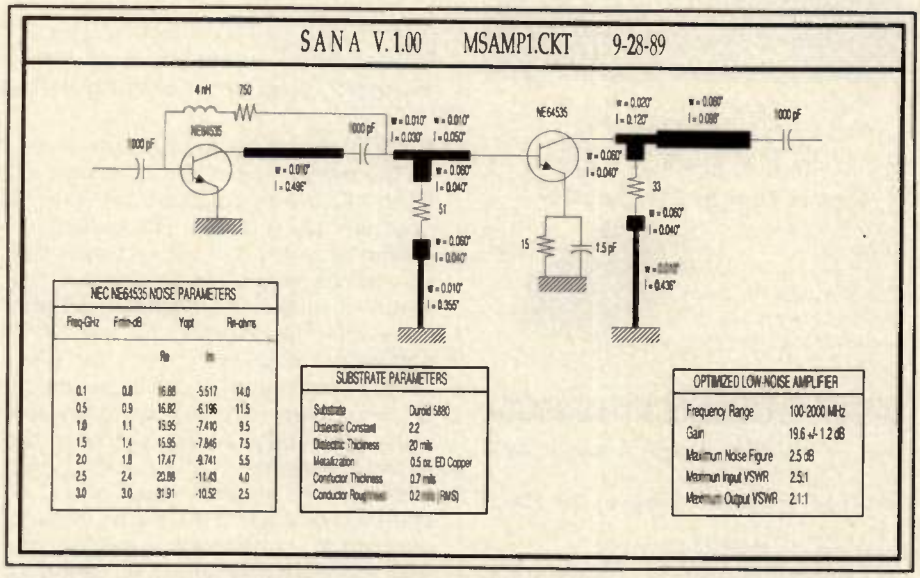

Figure 1. The broadband low noise amplifier is represented as a 32 node network including all microstrip discontinuities.

work elements, but must include the effects of related discontinuities as well. Finally, circuit optimization must be performed in a useful fashion. Optimization must be thorough and must reach the best solution. The network optimizer also must be smart enough to adjust parasitic and discontinuity elements consistent with other circuit perturbations. The practicing engineer can no longer be satisfied with anything less.

SANA

As a result of many years of effort, the SANA simulator is now available. SANA (*S*ignal *A*nd *N*oise *A*nalysis) is the product of a group headed by Dr. Peter Russer of RTI and the Technical University of Munich in West Germany. Provided and supported by Webb Laboratories, SANA allows very large and complex topologies to be entered and modeled with ease.

The software employs a familiar input file format to describe topologies comprised of both lumped and distributed elements. Parasitic elements, such as microstrip bends, tees and steps, are entered such that they accurately track interconnecting lines under the influence of the optimizer. That is, the discontinuity model modifies itself as the adjoining element is being adjusted by the optimization process. Finally, in addition to the agility of passive network definition, SANA allows device noise and linear signal parameters to be recalled from a library format.

Design Example

It was decided to apply SANA to the task of optimizing the design of a single-ended two-stage low-noise amplifier covering the 100 to 2000 MHz band. The network of Figure 1, using a pair of NEC NE64535 silicon bipolar transistors, was first designed to realize maximum matched gain at 2 GHz (over 55 dB, but very narrow band). The SANA input file was then modified, adding feedback around the first stage and defining eleven parameters to be subject to adjustment by the SANA optimizer. Basically, the lumped resistances and several of the microstrip line widths and lengths were allowed to vary. All variables were constrained to remain within practical regions. Line widths

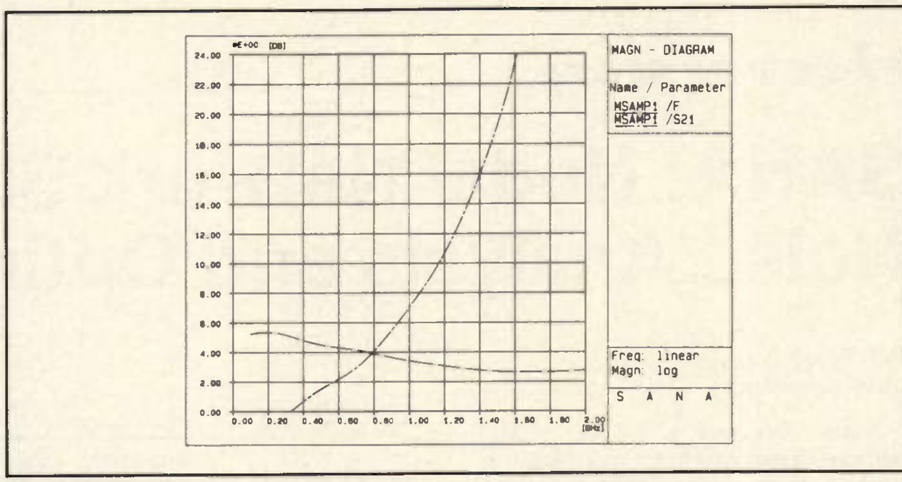

Figure 2. Gain and noise figure before optimization.

stayed between 10 and 100 mils, etc.

The plotter output shown in Figure 2 illustrates the amplifier noise figure (F) and gain (S21) in dB. The preliminary design is clearly narrowband and offers little in the way of noise performance. Figure 3 shows the amplifier performance after modification by the SANA compound deterministic optimizer. While computer optimization of a large number of parameters is not always consistent with good engineering practice, the example illustrates that it is possible with SANA. The program was asked to achieve noise performance as close as possible to the theoretical single-tuned limit while simultaneously adjusting for flat gain of 20 dB with good input and output match. The final network realization, shown in Figure 1 and originally intended to serve as one leg of a balanced amplifier, displays sufficiently good input and output VSWR (S11 and S22 are shown in Figure 4) that it will satisfy many broadband applications in the single-ended configuration.

Optimization

The above results were achieved by application of the SANA Simplex optimization algorithm, which uses a strategy dramatically superior to the gradient approach. Simplex is relatively fast and very reliable. Of note is the fact that SANA offers a choice of ten optimization algorithms. Four deterministic choices and five stochastic procedures may be specified in the circuit input file. While it might be argued that the availability of choices complicates the user interface, one need only consider the results of the above example to see that the Simplex approach alone performs at a very high level. The remaining SANA algorithm performs linked optimization using both stochastic and deterministic procedures. This approach offers the best possible likelihood of achieving the desired circuit performance.

Execution Speed

SANA employs a very sophisticated analysis engine. The nodally-connected network is first reduced to an equivalent admittance parameter matrix. Subsequently, proprietary sparse matrix techniques are applied to the Y-matrix, yielding an extremely simplified network

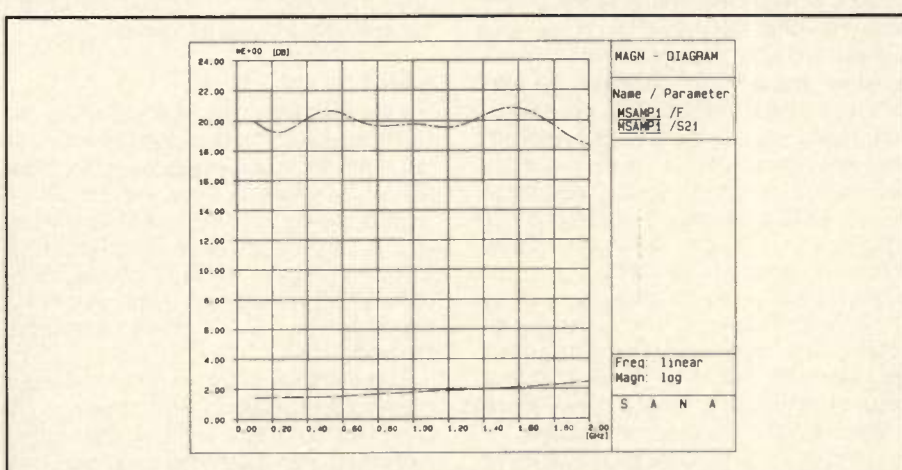

Figure 3. Gain and noise figure after optimization.

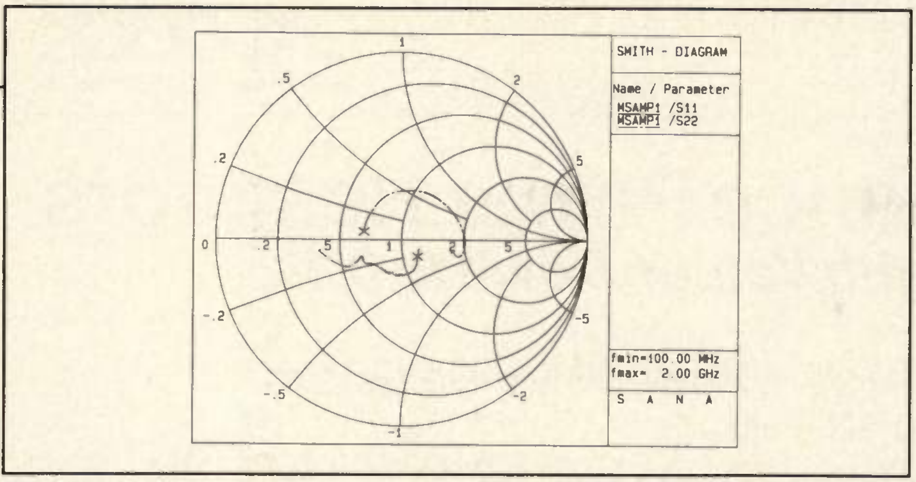

Figure 4. Input (S11) and output (S22) characteristics presented in Smith chart format.

representation which lends itself to fast computation. On an IBM PS/2 Model 50 (10 MHz 80286/80287), SANA performed the analysis of the example of Figure 1 (32 network nodes including discontinuity elements) in less than 1.5 seconds per frequency. This time included all file preparation and data (S-parameter and noise) interpolation.

It must be emphasized that SANA, unlike some other CAE simulation packages, performs all computations necessary for complete graphics output after the fact. That is, a unique graphics post-processor allows rectangular, polar and Smith Chart representation of desired outputs to be delivered without returning to the analysis mode! The graphics post-processor allows S-, Y- and Z-parameters, noise figure, minimum achievable noise figure, group delay, and numerous other parameters of interest to be presented, printed and plotted on a choice of displays, all after the analysis and without re-analyzing the network of interest. SANA also saves all necessary graphics information so that the file last analyzed may be graphed immediately upon reloading the program.

System Requirements and the SANA Memory Manager

The intelligent platform management capability of SANA allows very large analysis and optimization jobs to be performed on conventional and readily available machines. The creation of up to 40 circuit blocks, with up to 40 nodes per block is allowed. Analysis will accommodate up to 512 frequencies, while optimization is capable of an unlimited number of variable parameters and up to 16 different optimization goals.

SANA surprisingly operates on IBM-PC/XT/AT and PS/2 machines with 640K RAM (565K available), DOS 3.0, and the math coprocessor, and has been verified on major 80486/80386/80286/8086/8088 compatibles. SANA supports Super VGA (800x600), VGA, EGA, Hercules and CGA graphics, with outputs delivered to HPGL plotters as well as conventional printers.

Perhaps the most unique feature of SANA is its ability to adapt to the platform at hand. While other analysis packages might require alternate operating systems, or might not operate on older 8088/8086 machines, this package possesses the ability to operate on all PC-DOS computing platforms. Although the SANA executable file is well in excess of 1 MByte in size, compatibility with older machines is provided through data caching via the hard disk or, preferably, added machine memory. On machines with extended and/or expanded memory, SANA utilizes the added RAM without special installation. In short, SANA allows large networks to be simulated in comprehensive fashion without forsaking the large population of DOS users and machines.

Summary

The SANA package brings a previously unavailable combination of computational power and affordability to the IBM-PC world in general, permitting signal and noise analysis to be performed on extremely large and complex network topologies without the requirement for a Unix workstation or the OS/2 environment.

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Richard Webb is President of Webb Laboratories, 139 E. Capitol Drive, Suite 4, Hartland, WI 53029. He can be reached by telephone at (414) 367-6823.

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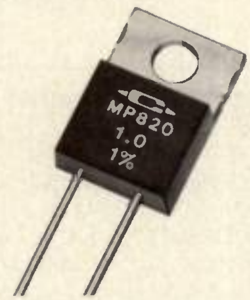

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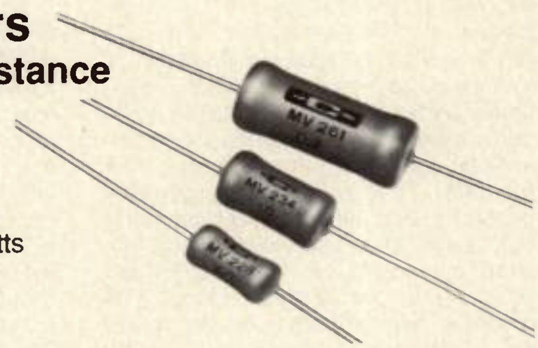

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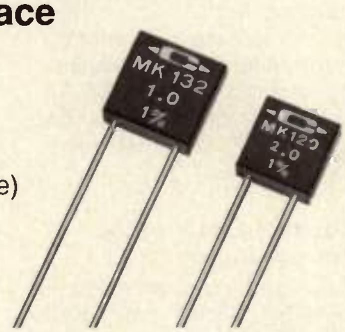

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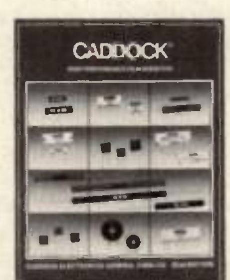

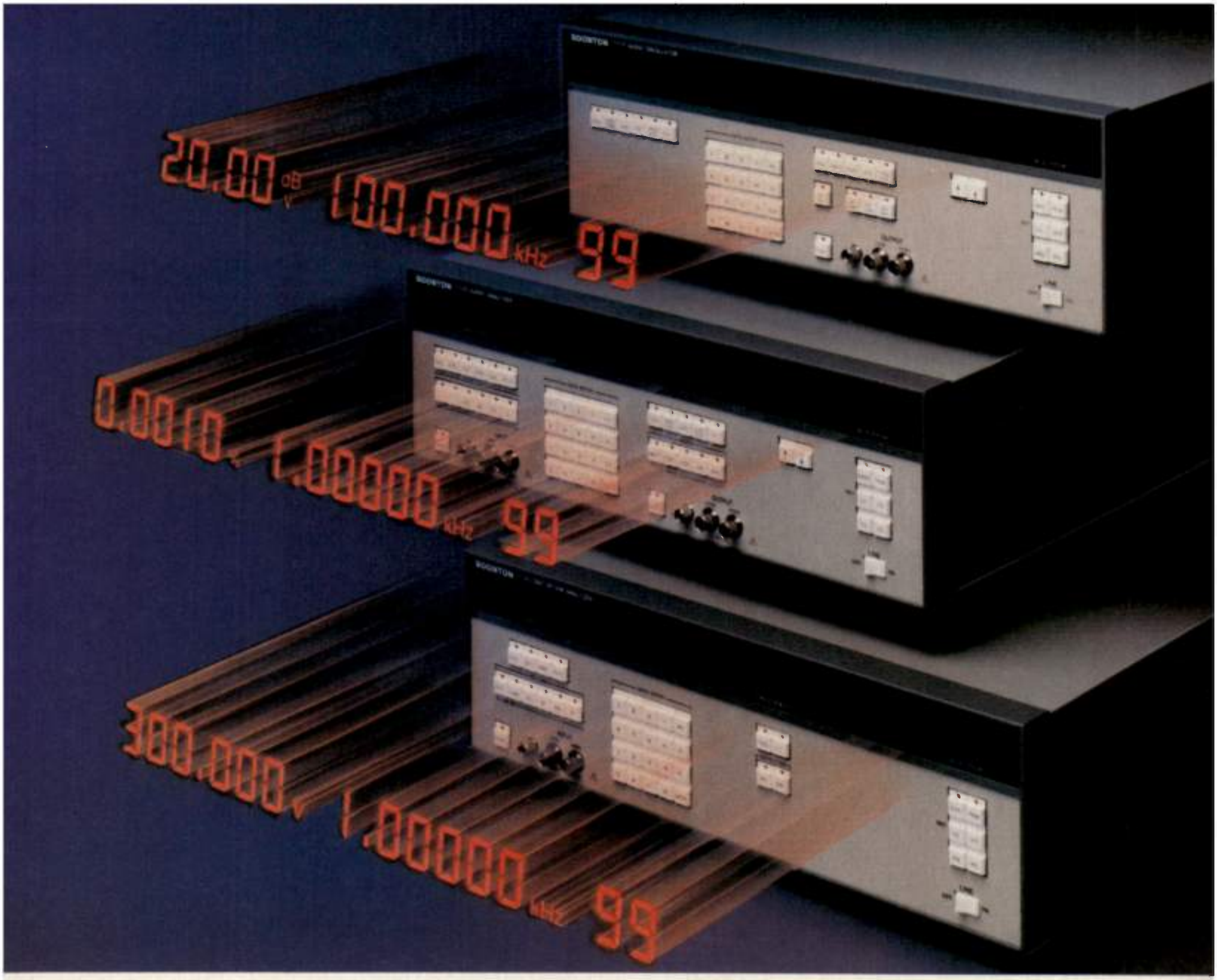
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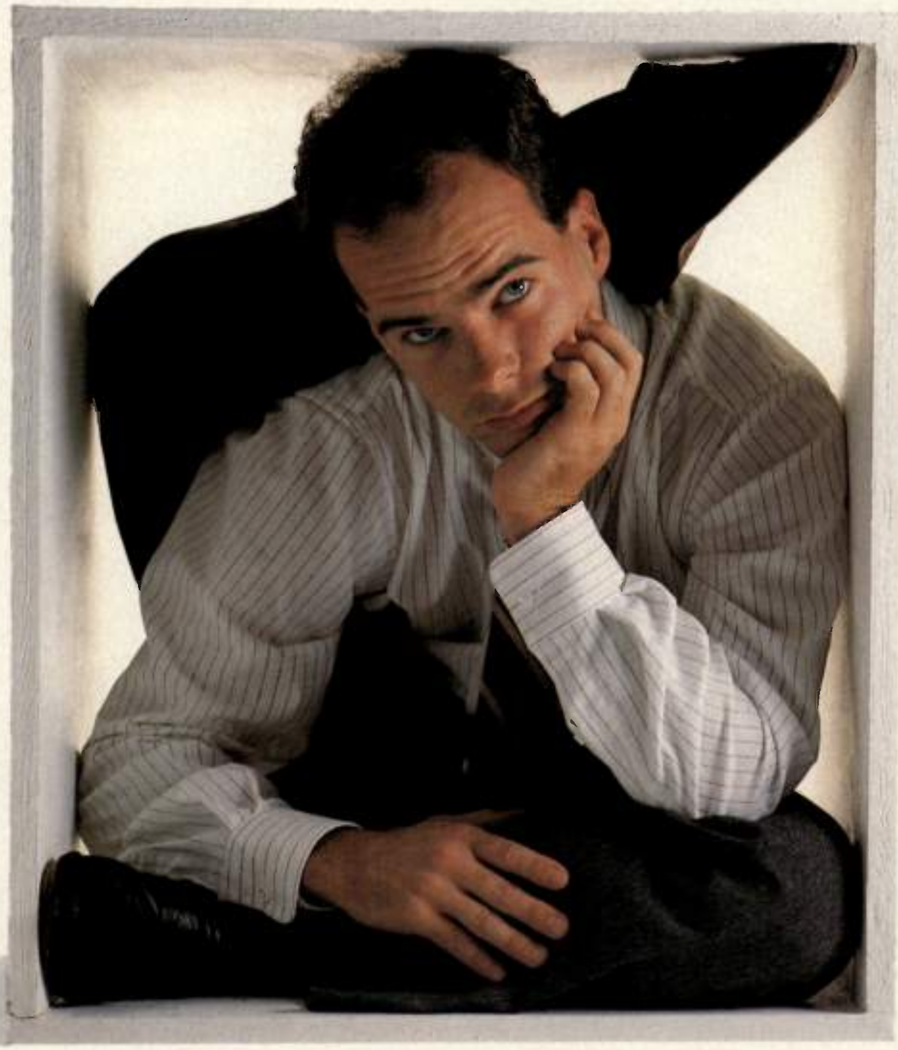

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An Unconventional Varactor-Tuned Bandpass Filter

By Gary Thomas
General Electric Mobile Communications

This prize-winning entry in the 1989 RF Design Awards Contest describes an electrically tunable bandpass filter which consists of a varactor-tuned, narrow bandstop filter (notch filter) and a directive bridge network. When connected as described below, the resultant circuit has a narrow bandpass characteristic which can easily be tuned using a single control voltage. This design was originally developed to improve front end selectivity in a VHF (150-174 MHz) mobile radio receiver (by placing the filter between the RF input amplifier and first mixer). However, it is adaptable to a large number of applications over a wide range of frequencies.

Electrically tuned bandpass filters are usually designed using multiple section coupled resonators. Each section is made tunable by placing a voltage variable capacitor (varactor diode) across the resonator. A control voltage is applied to the varactor diodes to tune the filter to the desired center frequency. An example of this type of filter is shown in Figure 1. The passband response of this filter is illustrated in Figure 2.

This filter could be made narrower by adding more sections and by increasing the Q of the resonators. But in both cases, this adds to the cost, complexity and size. In addition, the use of multiple sections requires multiple varactor diodes which must track as the tuning voltage varies. Finally, since the diodes are placed directly across the resonators, they may be subjected to high RF levels at all frequencies and have the potential to generate intermodulation products within the filter itself.

An Improved Filter

The circuit shown in Figure 3 has been developed to overcome many of the problems associated with other electrically tuned bandpass filters. A complete parts list is included in Table 1. It should be noted from Figure 3 that the tuning voltage is used to control a single point in the circuit. The two varactors are placed back to back to provide more linear tuning of center frequency vs. voltage and to provide higher intermodulation immunity. Thus the need for tracking among multiple diodes is avoided.

Figures 4 through 6 characterize the performance of the improved electrically tuned bandpass filter shown in Figure 3. Comparison of Figure 2 and Figure 5 shows that the improved filter achieves a much narrower bandwidth using only one tuned stage as opposed to a conventional filter which uses two. In addition, the increased selectivity has been achieved without resorting to the use of a large, expensive resonator.

Theory of Operation

The improved electrically tuned bandpass filter uses the signal isolating properties of a balanced resistive bridge to transform a series tuned reflective notch filter into a selective bandpass filter. In this respect, any directive device (directional coupler, ferrite circulator, Wilkinson splitter, etc.) which could be used to measure return loss can be used to provide this transformation. The resistive bridge was chosen because it can at the same time be used to provide an apparent "Q multiplication" in the tuned circuit, yielding a sharper selectivity.

To understand the operation of the filter, refer first to Figure 7 which shows a conventional VSWR bridge. In the network in Figure 7(a), if

$$(R_1/R_2) = (R_3/R_4) \tag{1}$$

then both sides of the load resistor will always be at the same potential and no current will flow through it regardless of the source voltage V_1. Thus, the source and load are isolated from one another. As long as the relationship in equation 1 holds, the impedance seen by the source will be the parallel combination of R_1+R_2 and R_3+R_4:

$$R_{in} = \frac{(R_1+R_2)(R_3+R_4)}{R_1+R_2+R_3+R_4} \tag{2}$$

Similarly:

$$R_{out} = \frac{(R_1+R_3)(R_2+R_4)}{R_1+R_2+R_3+R_4} \tag{3}$$

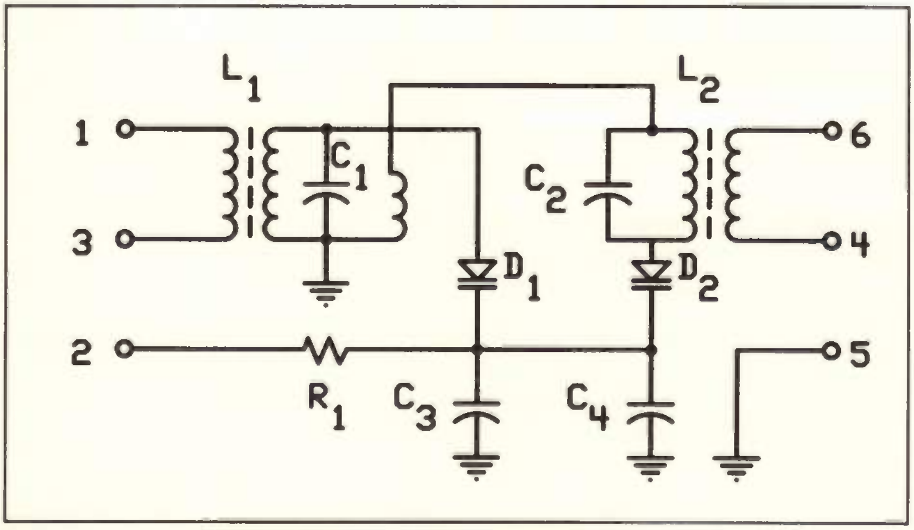

Figure 1. Typical conventional varactor tuned filter.

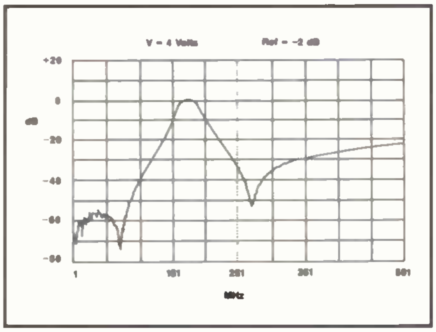

Figure 2. Response of the conventional filter in Figure1.

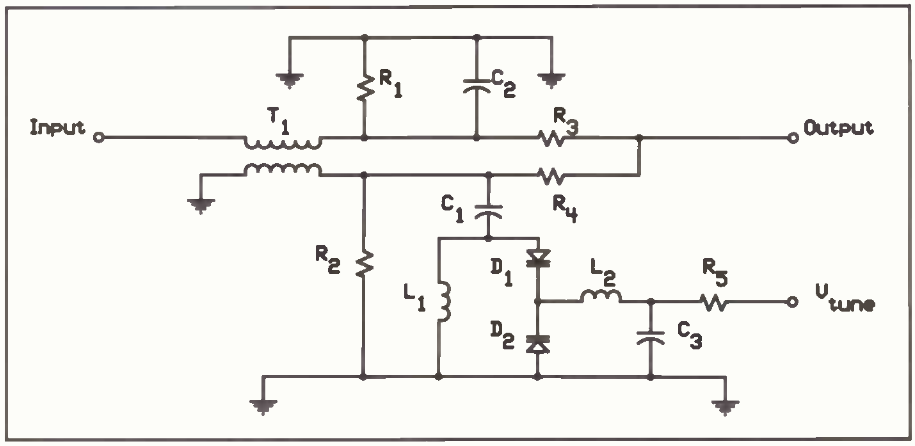

Figure 3. The improved electrically tuned bandpass filter.

Under normal conditions R_1, R_3 and R_4 are each set to 50 ohms. If a 50 ohm resistor is also used as R_2, equation 1 shows that the bridge is balanced and there is no output voltage across R_L. In this case R_{in} = R_{out} = 50 ohms so that the network is matched to the source and load impedance. In normal operation R_S and R_L are usually replaced with external ports for connection to a signal generator and measuring device, while R_2 is replaced with an external port for connection to a device under test. A broadband transmission line balun is used as shown in Figure 7(b) to accommodate an unbalanced source without grounding node d and shorting out the bridge.

The VSWR bridge is normally used to compare an impedance connected to the external port with the internal 50 ohm reference. The signal present at the output will be a minimum when the impedance at the test port is exactly 50 ohms, and will increase to a maximum value with either a short or open circuit at the test port.

It should be noted that when there is a short or open at the test port, the input and output impedances according to equations 2 and 3 are no longer 50 ohms. The equivalent circuit for the shorted condition is shown in Figure 8. This network has a 12 dB return loss and a 12 dB insertion loss in a 50 ohm system.

If the impedance presented to the test port is as shown in Figure 9, the tuned circuit consisting of C and L will present a short circuit to the test port at resonance, and there will be a maximum transfer of signal from input to output. At all other frequencies, the series combination of C and L presents essentially an open circuit across the 50 ohm resistor, so that the test port is matched and no transmission occurs. Clearly, the inclusion of the series tuned circuit has produced a passband characteristic from input to output. If a varactor diode is used for C, the center frequency of the passband can easily be changed by changing the tuning voltage on the varactor. This is the basic principal of operation of the improved electrically tuned bandpass filter.

Note that with this arrangement the varactor diode is protected from high signal levels which are not within the narrow passband of the filter. For this reason the filter is inherently less susceptible to the generation of intermodulation products than is the conventional design.

The circuit of Figure 3 contains several refinements compared to the simplified model presented above. The simple series tuned trap is replaced with the circuit shown in Figure 10. In this implementation of the notch filter C_1 is a small fixed capacitor. D_1 and D_2 form a voltage variable capacitor which is parallel resonant with L at a frequency above the desired tuning frequency of the filter. Thus, at the desired frequency, the parallel combination of L and D_1/D_2 appears to be a high value of inductance which can be varied by varying V_{tune}. This inductance in turn resonates with C_1 to yield a low impedance at the desired center frequency.

The most important refinement to the simple model is obtained by changing

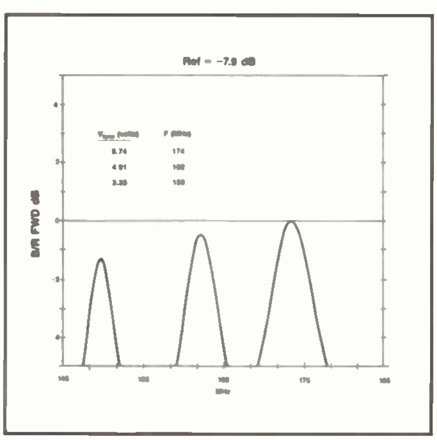

Figure 4. Transmission response (narrow sweep).

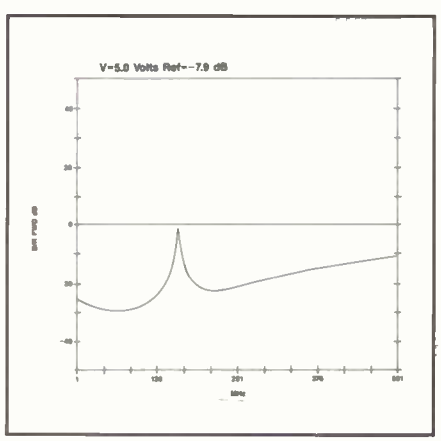

Figure 5. Transmission response from 1 MHz to 501 MHz.

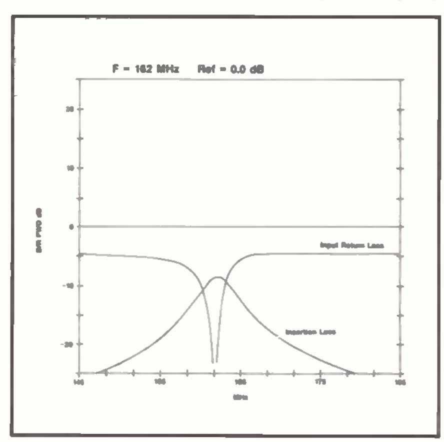

Figure 6. Passband return loss and insertion loss.

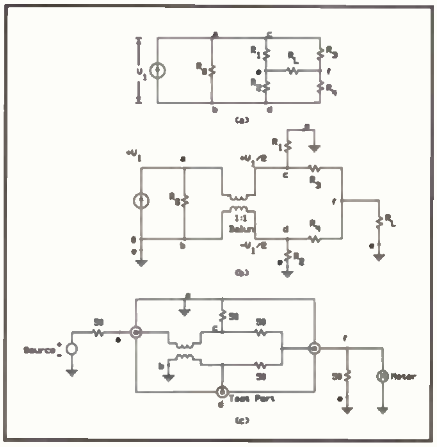

Figure 7. Conventional VSWR bridge.

C_1	Capacitor, Ceramic Disc, 1.5 pF ±0.25 pF, N150 50DCWV
C_2	Capacitor, Ceramic Disc, 15 pF ±5%, N150 50DCWV
C_3	Capacitor, Ceramic Disc, 33 pF ±5%, N150 50DCWV
D_1	Diode, Varactor, Motorola MMBV105G
D_2	Diode, Varactor, Motorola MMBV105G
L_1	Coil, RF, 181 nH ±5%, Paul Smith #SK954-1
L_2	Coil, RF Choke, 1 uH
R_1	Resistor, Chip, 220 ohm ±5%, .125 watt
R_2	Resistor, Chip, 1800 ohm ±5%, .125 watt
R_3	Resistor, Chip, 27 ohm ±5%, .125 watt
R_4	Resistor, Chip, 220 ohm ±5%, .125 watt
R_5	Resistor, Carbon Film, 1800 ohm ±5%, .250 watt
T_1	Balun, wideband transmission line transformer 2 strands, 38 Gauge polyurethane-coated copper wire, 2 in. long, 4.5 twists per inch, bifilar wound on Fair-Rite products balun core #28430002402

Table 1. Parts list for the improved filter.

the resistors in the bridge to the values shown in Figure 3. Since only the input and output need be matched to 50 ohms, R_1, R_2, R_3 and R_4 can be chosen to optimize other parameters subject to the constraint imposed by equation 1.

In order to maintain a 50 ohm input and output impedance in the passband of the filter (i.e., when the series tuned notch presents a short circuit to node d), R_1, R_3 and R_4 are chosen to simulate a matched 50 ohm Pi attenuator (Figure 8). Once the value of the attenuator is chosen, standard equations are used to calculate R_1, R_3 and R_4. Next, the value of R_2 is chosen to satisfy equation 1. This ensures that at frequencies away from resonance the bridge will be balanced and transmission from input to output will be minimum. Since nominal passband insertion loss is equal to the value of attenuation chosen, it is desirable to design for the lowest attenuation possible. The values of R_1, R_3 and R_4 chosen for the prototype circuit of Figure 3 correspond to a matched 4.4 dB attenuator. Thus, the nominal insertion loss of the filter is 4.4 dB. (The actual attenuation exceeds this due to additional stray losses in the circuit.) The resulting impedance to ground from node d is 86 ohms compared to 25 ohms for the standard 50 ohm bridge. This results in a potential 3.4-fold increase in loaded Q!

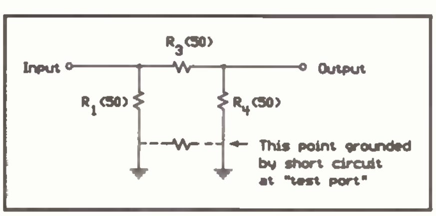

Figure 8. Equivalent circuit in shorted condition.

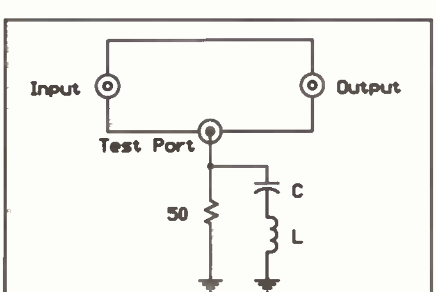

Figure 9. Test port load which produces bandpass transmission characteristics.

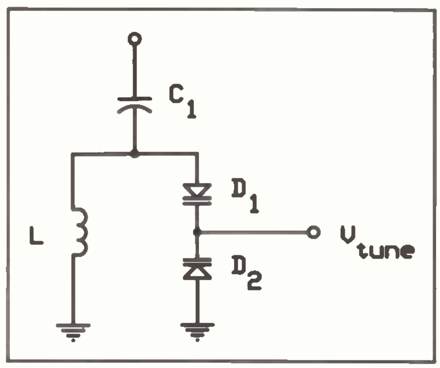

Figure 10. The improved tuning circuit.

It should be noted that the 4.4 dB attenuator was chosen for the prototype circuit because it and the resulting termination at R_2 could be implemented using standard chip resistors. By using non-standard resistors and accurately trimming their values, lower loss and higher apparent Q may be achieved.

Summary

An improved electrically tuned bandpass filter has been presented. The circuit offers several advantages over currently available filters: It is simple and cheap to build, can achieve higher selectivity in a smaller size using less complex circuitry, and is inherently less susceptible to the generation of intermodulation products.

A patent application for this invention has been filed with the U.S. Patent Office on behalf of General Electric Company.

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About the Author

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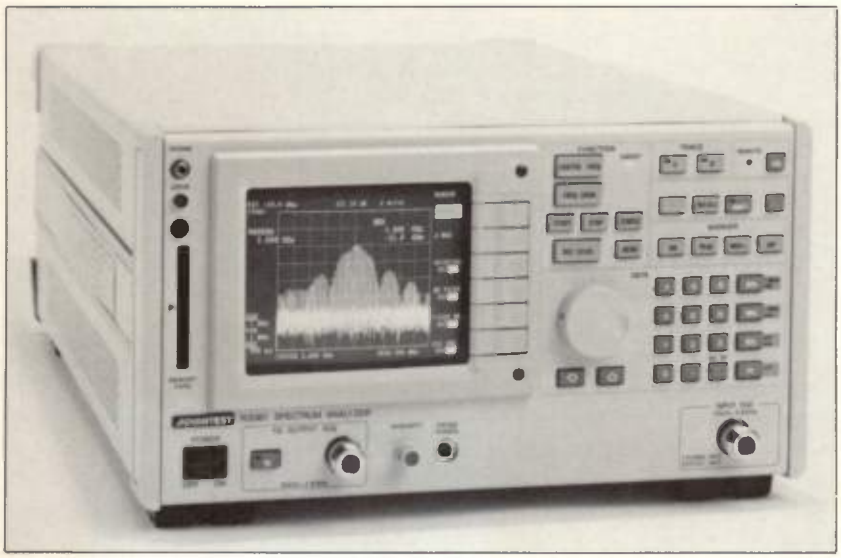

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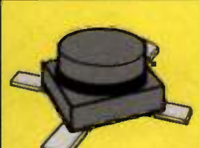 CGY 40	Micro-X	17.5	10	2.9	1.4	32.5	20	100 MHz–1.8(4) GHz
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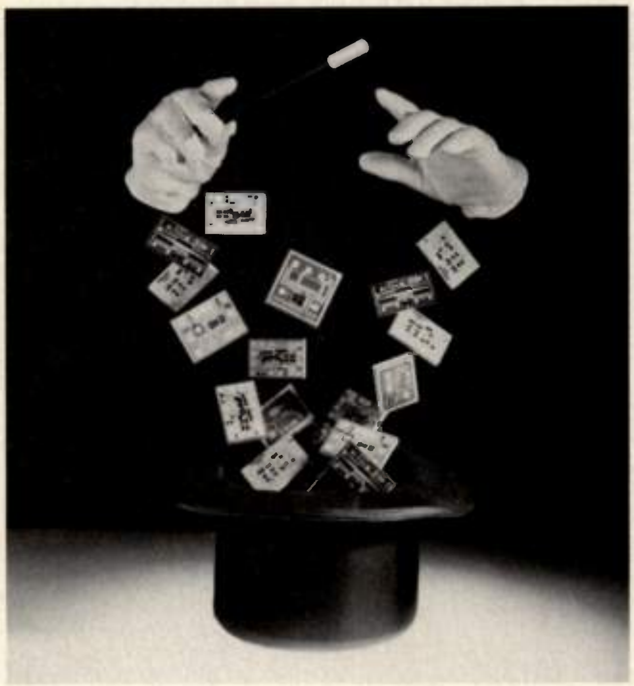

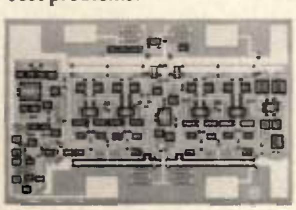

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Absorptive Directional Filters

By Michael Fithian
General Electric Company

Absorptive directional filters (ADF) are a class of bandpass filters that provide not only an acceptable in-band impedance match but also an excellent out-of-band match. This differentiates them from filters which achieve out-of-band rejection by providing large impedance mismatches. This article examines the design of ADF circuits and presents measurements of a microstrip version of the filter at low L-band frequencies. Several possible applications of the circuit are also discussed.

The filter consists of an identical pair of parallel coupled lines interconnected by quarter-wavelength sections of transmission line (Figure 1(a)). The analysis of the ADF circuit begins with an examination of the parallel coupled line structure.

Parallel Coupled Line Circuit

From reference 1, for a perfect, lossless coupler (Figure 1(b)), 1 volt applied at the input at port 1 will result in:

$$V_2 = \frac{\sqrt{1-k^2}}{(\sqrt{1-k^2}\cos\theta) + j\sin\theta} \qquad (1)$$

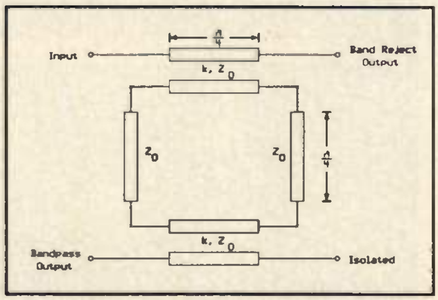

Figure 1(a). ADF schematic diagram.

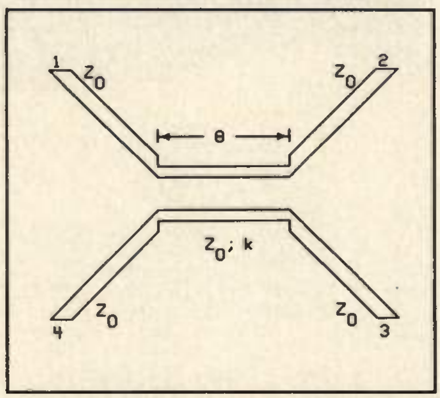

Figure 1(b). Parallel coupled line circuit.

$$V_3 = 0 \qquad (2)$$

$$V_4 = \frac{jk\sin\theta}{(\sqrt{1-k^2}\cos\theta) + j\sin\theta} \qquad (3)$$

where $k = V_4/V_4 = V_4/(1\text{ volt})$

Let $\theta = 90$ degrees, then:

$V_2 = -j\sqrt{(1-k^2)}$
$V_3 = 0$
$V_4 = k$

Using the above results to evaluate the ADF circuit (Figure 1(c)) and including a loss of α in nepers/wavelength yields:

$$\begin{aligned} V_{out} &= k_1k_2e^{-\alpha\lambda_o/4} \measuredangle{-90°} \\ &+ k_1k_2\sqrt{(1-k_1^2)(1-k_2^2)}e^{-\alpha\lambda_o(5/4)} \measuredangle{-90°} \\ &+ k_1k_2(1-k_1^2)(1-k_2^2)e^{-\alpha\lambda_o(9/4)} \measuredangle{-90°} \\ &+ \ldots \end{aligned}$$

$$\begin{aligned} V_{out} = k_1k_2e^{-\alpha\lambda_o/4}[1 + \\ \sqrt{(1-k_1^2)(1-k_2^2)}e^{-\alpha\lambda_o} + \\ (1-k_1^2)(1-k_2^2)e^{-2\alpha\lambda_o} + \ldots] \measuredangle{-90°} \end{aligned}$$

From the series $1/(1-X) = 1 + X + X^2 + \ldots$,

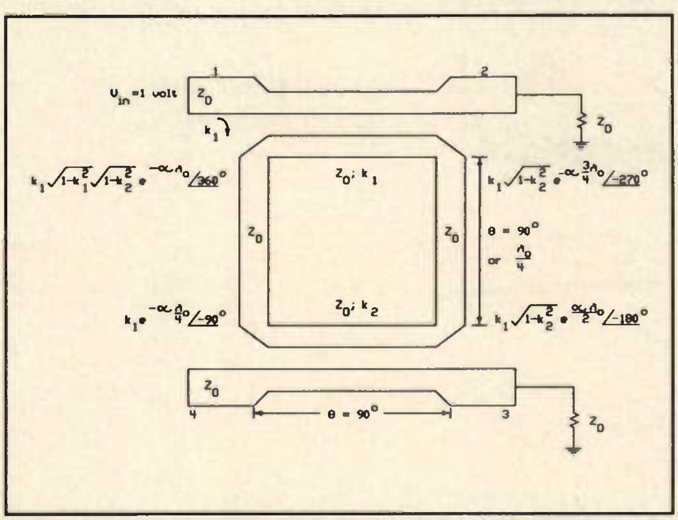

Figure 1(c). The ADF circuit with α considered .

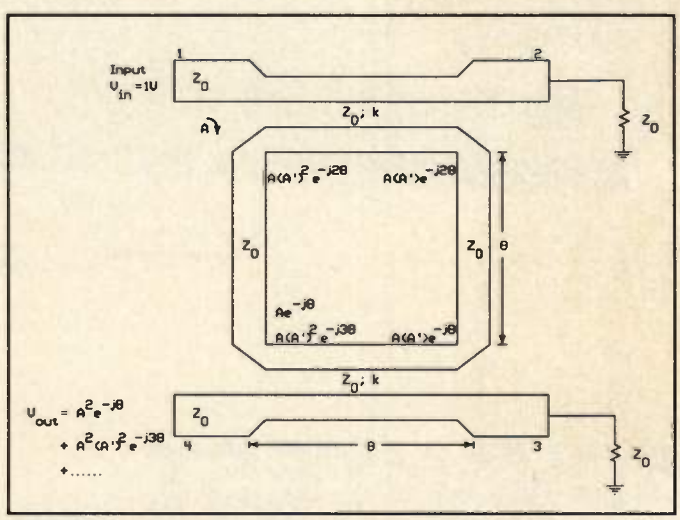

Figure 1(d). Frequency response model.

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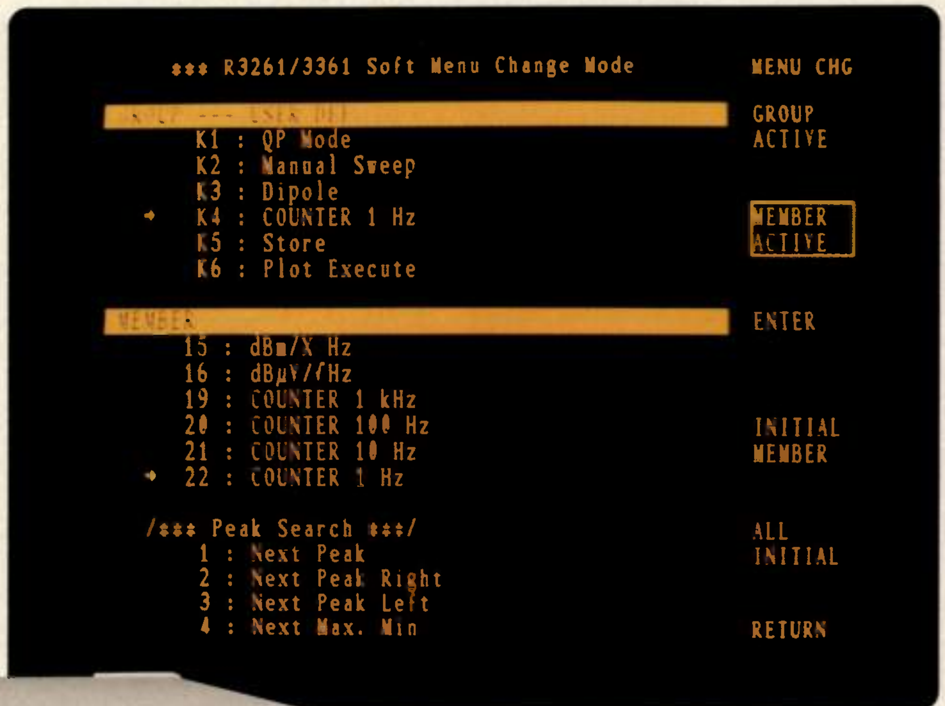

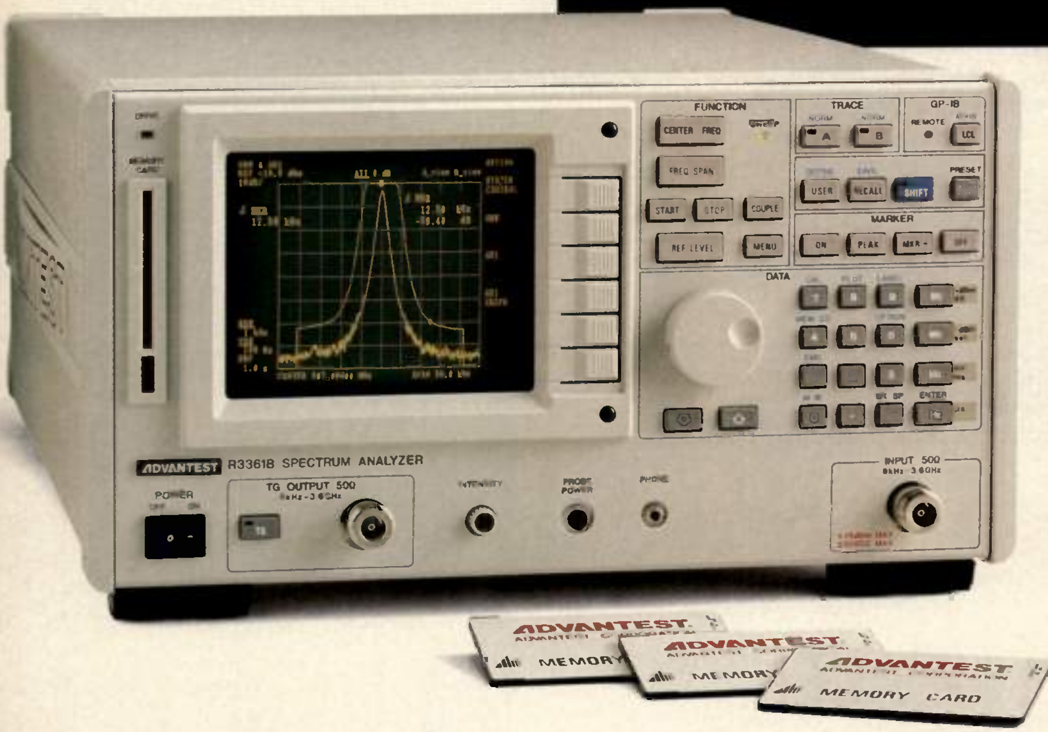

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$$|V_{out}| = \frac{k_1 k_2 e^{-\alpha\lambda_0/4}}{1 - \sqrt{(1-k_1^2)(1-k_2^2)}e^{-\alpha\lambda_0}} \quad (4)$$

Now if $k_1 = k_2 = k$ and $\alpha = 0$ (for lossless lines), then:

$$|V_{out}| = \frac{k^2}{1-(1-k^2)} = 1 \text{ at } \theta = 90°$$

Note that the result is independent of k at the frequency at which $\theta = 90$ degrees.

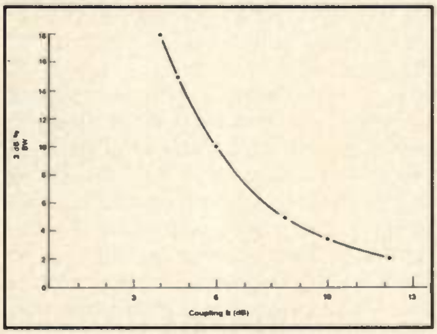

Figure 2. 3 dB percent bandwidth vs. coupling factor.

```
Absorptive directional filter
3 dB bandwidth of 10 percent
Designed for microstrip soft substrate
Er=10.5, H=0.05 in.

DIM
FREQ MHZ
RES  OH
LNG  MIL

VAR

LEN1=873, LEN2=950, WC=27.9, GAP=5.0, W1=40.3
(Filter dimensions)

CKT
Substrate Data

MSUB ER=10.5 H=50 T=1.4 RHO=1.1 RGH=0.4E-3
TAND TAND-0.0015

MCLIN  1 2 3 4   W^WC  S^GAP  L^LEN1
MLIN   2 5       W^W1  L^LEN2
MLIN   3 6       W^W1  L^LEN2
MCLIN  8 5 6 7   W^WC  S^GAP  L^LEN1
RES    7 0       R=50
DEF3P  1 8 4     ADF

FREQ

SWEEP 1000 1500 10

OUT

ADF DB[S11]
ADF DB[S22]
ADF DB[S21]
ADF DB[S31]
```

Figure 3. CAD filter model.

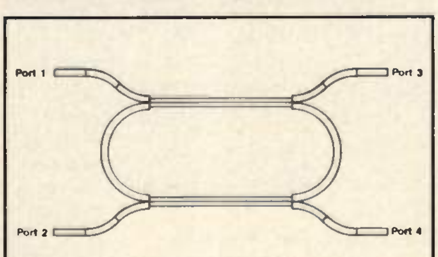

Figure 4. ADF circuit layout.

In order to calculate the frequency response for $k_1 = k_2 = k$ and lossless lines, a similar analysis is required. Using equations 1 and 3, let:

$$A = \frac{jk \sin\theta}{(\sqrt{1-k^2}\cos\theta) + j\sin\theta} \quad (5)$$

$$\text{and } A' = \frac{\sqrt{1-k^2}}{(\sqrt{1-k^2}\cos\theta) + j\sin\theta} \quad (6)$$

Again, for a 1 volt signal at the input (Figure 1(d)):

$$V_{out} = A^2 e^{-j\theta} + A^2(A')^2 e^{-j3\theta} + A^2(A')^4 e^{-j5\theta} + \ldots$$

$$V_{out} = A^2 e^{-j\theta}[1 + (A')^2 e^{-j2\theta} + (A')^4 e^{-j4\theta} + \ldots]$$

From the infinite series $1/(1-X^2) = 1 + X^2 + X^4 + \ldots$,

$$V_{out} = \frac{A^2 e^{-j\theta}}{1-(A')^2 e^{-j2\theta}}$$

(Note when $\theta = 90$ degrees, $V_{out} = 1$)

It is convenient to characterize the frequency response in terms of 3 dB bandwidth:

$$|V_{out}|^2 = \frac{|A^2|^2}{|1-(A')^2 e^{-j2\theta}|^2} = 0.5 \quad (7)$$

Using equations 5 and 6 in equation 7 and simplifying yields:

$$\theta = \tan^{-1}\left\{\frac{2[(1-k^2)+\sqrt{1-k^2}]}{k^2}\right\} \quad (8)$$

$$\text{and 3 dB \%BW} = \frac{2(90° - \theta)}{90°} \times 100 \quad (9)$$

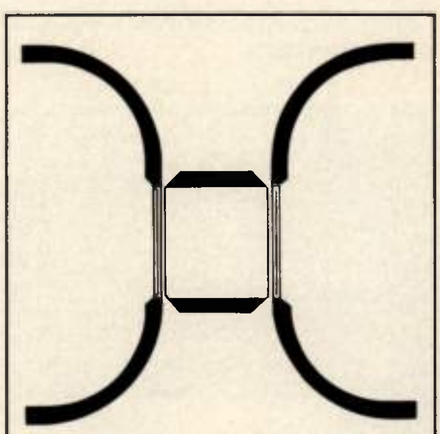

Figure 5. ADF circuit using Lange couplers.

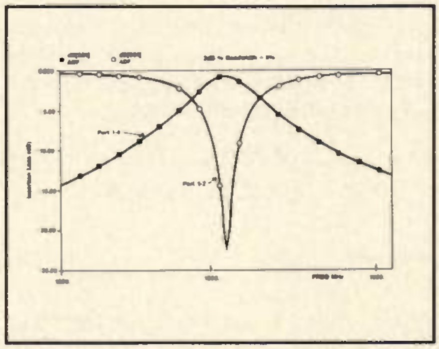

Figure 6(a). ADF bandpass/bandstop characteristics.

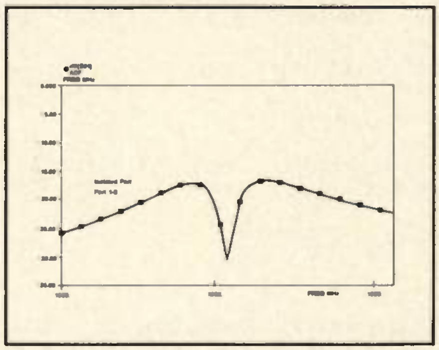

Figure 6(b). ADF isolated port characteristics.

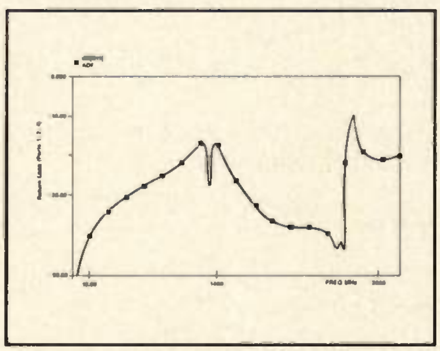

Figure 6(c). ADF impedance match characteristics.

A plot of 3 dB percent bandwidth vs. coupling factor k(dB) is shown in Figure 2. Note that the percent BW is inversely proportional to the coupling factor.

Experimental Results

A filter design that would yield a 10 percent 3 dB bandwidth centered at 1250 MHz was performed. Figure 2 shows that a 10 percent bandwidth requires a coupling value of 6 dB. These requirements were factored into the CAD model given in Figure 3. The design was optimized for a microstrip, soft substrate application (ε_r=10.5, H=50 mils, 1 oz. copper). The line widths, lengths and spacings were used to generate the circuit layout (Figure 4). Here, a coupling value of 6 dB can be achieved using a pair of parallel coupled lines. A wider bandwidth filter requires tighter coupling values which are not readily achieved with this approach. If this is the case, Lange interdigital couplers can be used (Figure 5).

The CAD-generated results are presented in Figures 6(a), 6(b) and 6(c). Note the bandpass characteristic at port 4 and the corresponding bandstop characteristic at port 2. These characteristics are achieved while maintaining an excellent broadband impedance match, DC-3000 MHz, at all ports (Figure 6(c)). Rather than reflect out-of-band frequencies, the filter absorbs them into its 50 ohm terminations.

Measured results for the filter are given in Figure 7. The 3 dB passband is 11 percent with an in-band loss of 1.2 dB. The actual results are highly dependent on etching tolerances for coupler line widths and spacings. Etching tolerances vary with conductor thickness, and tighter tolerances can be achieved with gold-plated ceramic substrates rather than with copper-clad soft substrate materials.

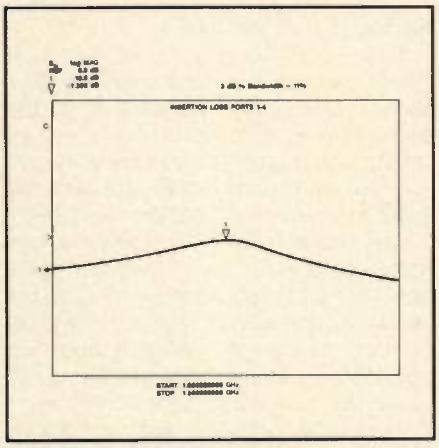

Figure 7(a). Measured bandpass insertion loss.

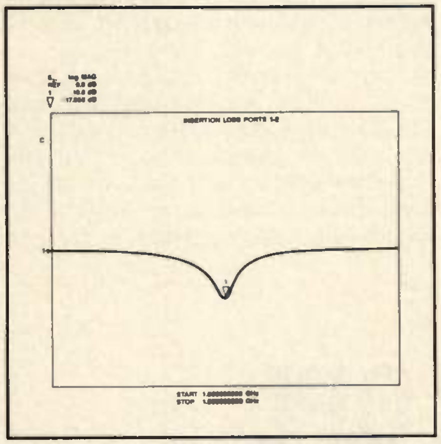

Figure 7(b). Measured bandstop insertion loss.

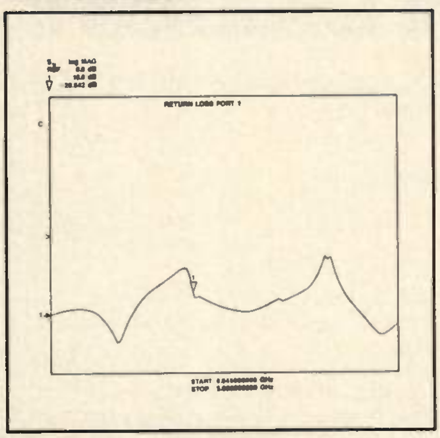

Figure 7(c). Measured return loss.

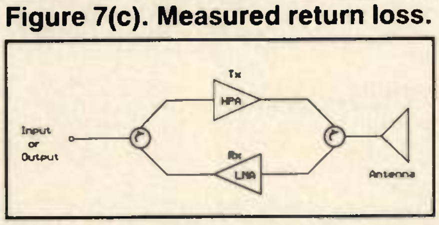

Figure 8. Radar transmit/receive module.

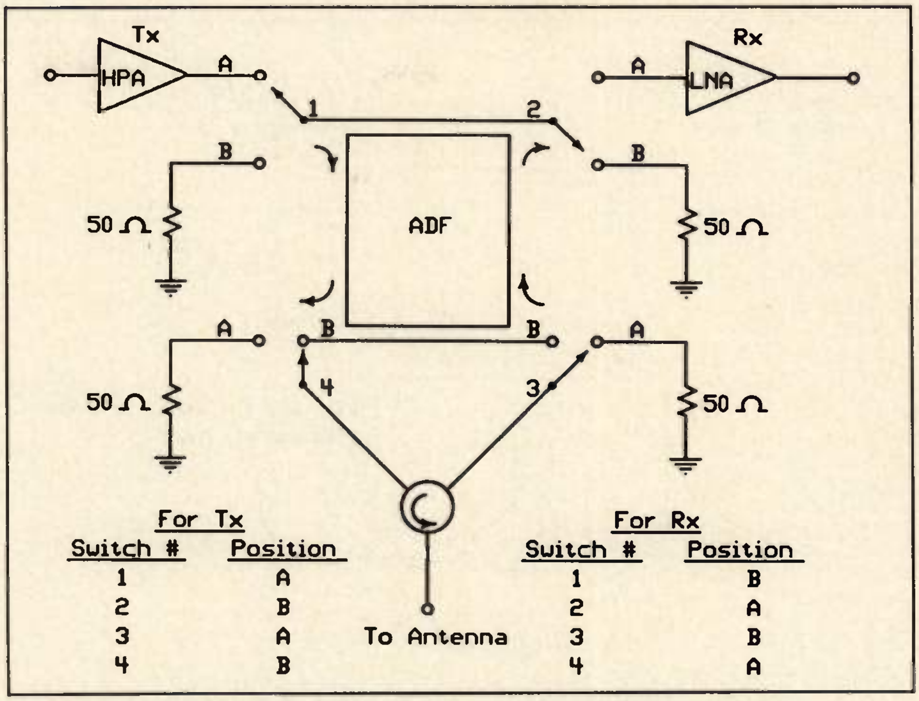

Figure 9. T/R module stabilizing scheme for radar applications.

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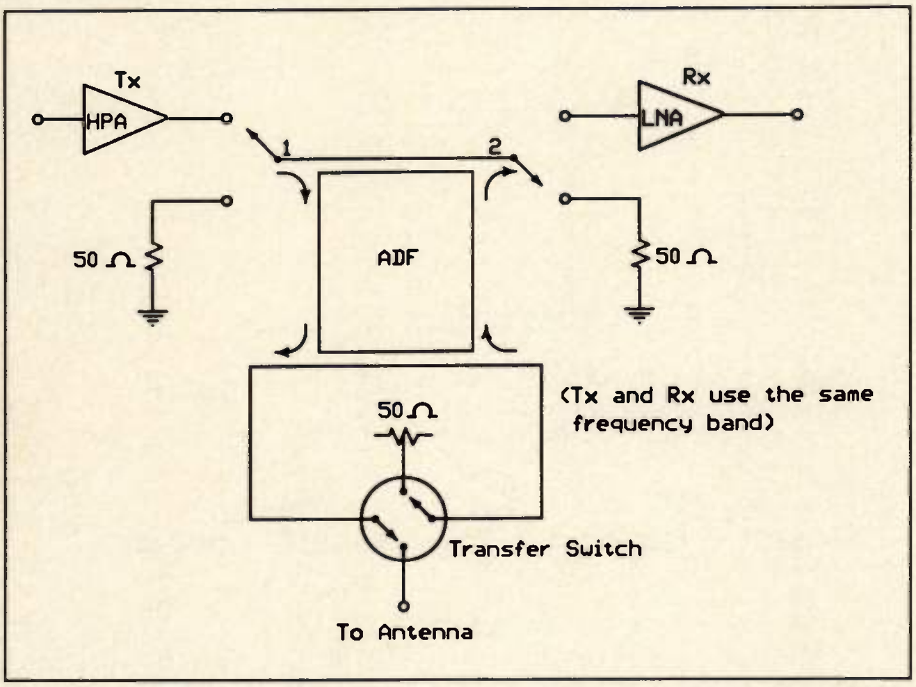

Figure 10. Alternate approach for the T/R module switching scheme for radar applications.

Applications for ADFs

The unique properties of the ADF circuit can be beneficial in many situations. One such application is the transmit/receive module shown in Figure 8. A multitude of these modules are used to populate an active phased array radar antenna. Both the transmit high-power amplifier (HPA) and receive low-noise amplifier (LNA) require well-behaved load and source impedances for stable operation. Severe in-band as well as out-of-band mismatches can frequently cause high performance GaAs FET amplifiers to oscillate. In a phased array antenna, such conditions are present due to the antenna mismatches associated with electronically scanning the beam off its main axis. One method for insuring a wideband, absorptive match for each amplifier is presented in Figure 9. This scheme uses an ADF circuit in conjunction with four SPST absorptive switches and a circulator as a buffer between the transmit and receive amplifiers and the antenna. In either the transmit or receive mode, the appropriate switch position is set and

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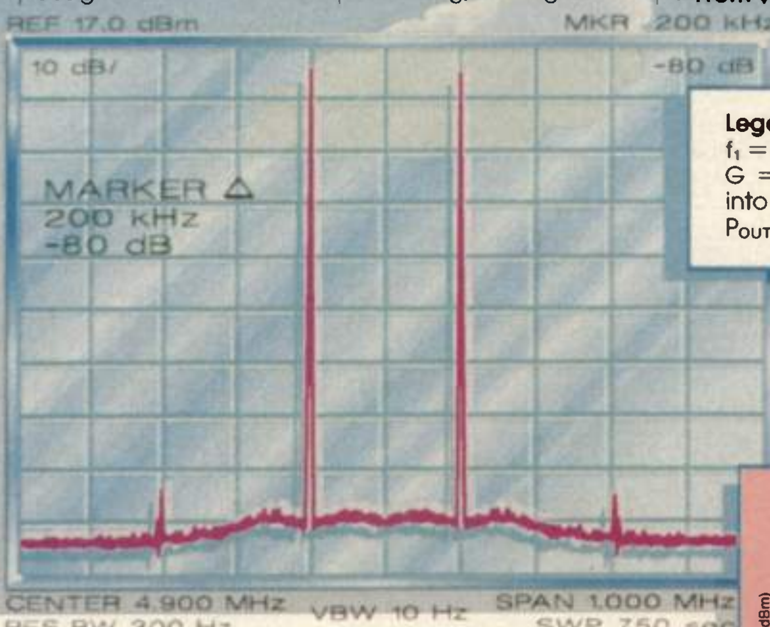

Legend:
f_1 = 4.8MHz, f_2 = 5.0MHz
G = 2V/V, V_{OUT} = $5V_{p-p}$ into 50Ω.
P_{OUT} = 12dBm/tone

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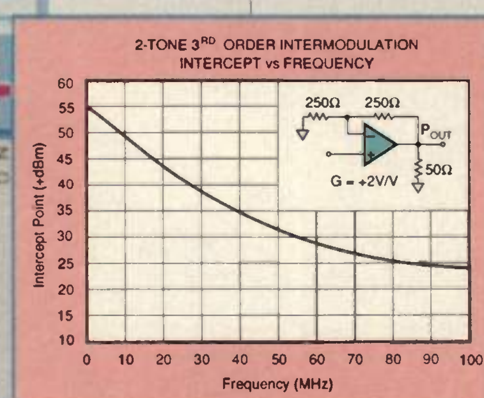

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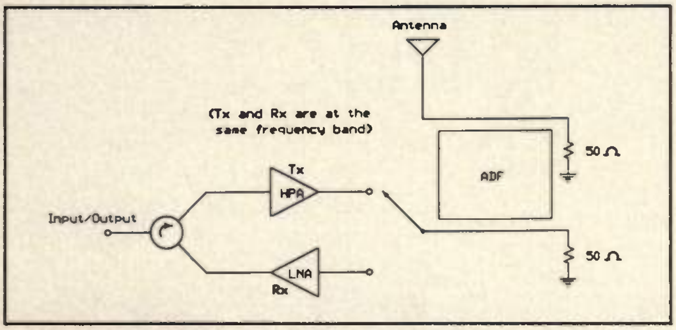

Figure 11. This configuration uses an SPDT switch.

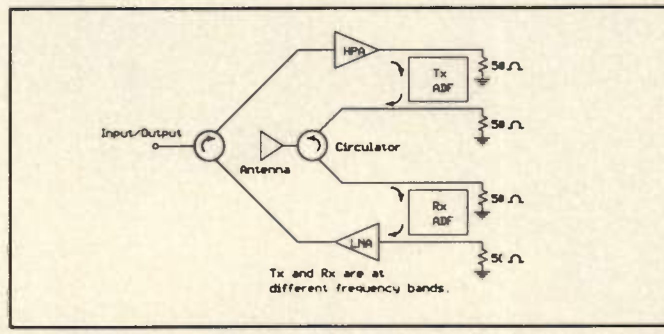

Figure 12. Two ADF circuits used to form a diplexer.

the ADF circuit and the 50 ohm loads insure a broadband match for the amplifiers. This match is independent of the scanned antenna impedance.

Two alternate configurations are shown in Figures 10 and 11. The approach of Figure 10 uses a transfer switch to replace the circulator, while the approach of Figure 11 uses an SPDT switch. Unlike the method shown in Figure 9, both these approaches are susceptible to in-band mismatches due to antenna impedance variation. If these mismatches are not too severe, however, the HPA and LNA will be presented with acceptable terminations.

Another application for the ADF circuit is as a diplexer in a communications system where the transmitter and receiver operate simultaneously (Figure 12). The two ADF circuits are tuned to different frequency bands and supply a 50 ohm termination to frequencies outside their passband, thus insuring amplifier stability.

In conclusion, the ADF circuit is a relatively simple design which can be used in applications where its filtering and its out-of-band absorptive properties are beneficial.

Reference

1. E.M.T. Jones and J.T. Bolljahn, *IRE Transactions on Microwave Theory and Techniques*, April, 1956, pp. 75-81.

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Perspectives on Ceramic Chip Capacitors

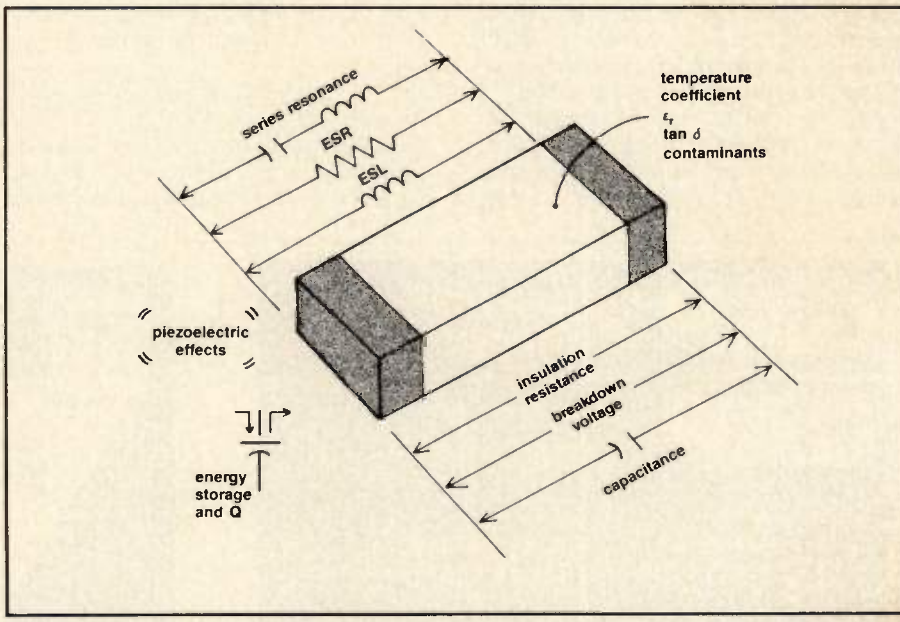
Part I: Device Properties

By Mark W. Ingalls
Dielectric Laboratories, Inc.

It is important to note that there simply aren't any "laws" which describe the behavior of natural phenomena in their totality. What have historically been know as laws (Ampere's, Faraday's, Ohm's, etc.) are only sketches from one point of view. Real objects in real systems, such as capacitors in electronic circuits, sometimes give responses that are absurd from the perspective of a single theory. To gain a clearer picture of how ceramic chip capacitors work, their behavior will be discussed using acoustics, optics, and electromagnetic theory, as well as circuit theory. This article (the first of two) will establish the definition of terms, and describe the properties of chip capacitors.

Circuit theory models the behavior of electrical devices in a single dimension — the order of devices along the path of the circuit. The actual length of the circuit is not considered and the devices themselves are taken as functions. The variables that these ordered functions (devices) operate on are voltage and current, which may in turn be time dependent. Just as any two functions may be combined to form a new function, separate device models can be combined to form new models, or broken down into several models for different characteristics of the device. One shortcoming of the theory is that some very simple physical devices, such as transmission lines, can have very complex models. Another is the tendency to forget that the device models are not real.

Electromagnetic theory models reactions of objects to forces resulting from stationary and moving electrical charges. These forces are summed or integrated over the appropriate physical dimensions and taken as a function called a field. Electromagnetics can be used to derive circuit theory, as well as to deal with problems that can't be solved with circuit theory alone. The main shortcoming of electromagnetics as a practical tool is that it takes years of practice to be able to do all but the simplest of calculations. A related problem is that calculations tend to be taken as the only way to comprehend the theory.



Many different characteristics are required to define a "real" ceramic chip capacitor.

Below are definitions of some commonly used terms, from the viewpoints of circuit theory and electromagnetics. These will be the tools used to develop a perspective on capacitors. (Careful reading can lead to new perspectives on any high frequency device!) Naturally, the two definitions of each term are intended to be inconsistent, but each emphasizes different qualities of the objects they describe.

For brevity's sake, the circuit perspective is labeled 1), and the field perspective is labeled 2). Appropriate example expressions follow some of the definitions.

Phasor. 1,2) An ordered pair of numbers signifying a magnitude and a time dependence (phase): Re[A exp(jωt)].

Vector. 1) Not defined. 2) An object

that consists of a magnitude and a direction: $A\hat{n}$.

Phasor-Vector. 1) Not defined. 2) An object that consists of both a magnitude and a direction, with a time dependence: $\mathrm{Re}[\sqrt{2}\,A\hat{n}\,\exp(j\omega t)]$.

Charge. The fundamental aspect of electricity which affects, or is affected by: 1) the circuit, or 2) the electromagnetic field.

Voltage. 1) Energy per unit of charge: joules per coulomb. 2) The force necessary to move a charge from point to point in the presence of an electric field: newton-meters per coulomb.

Current. Flow of charge per unit time: 1) through a conductor, or 2) normal to a proscribed area.

Conductor. A substance: 1) which allows the free transfer of charges, or 2) which cannot support an electric field.

Dielectric. A substance: 1) which prevents the free transfer of charges, or 2) which can support an electric field.

Flux. 1) The (nonlinear) relation between motive and induced current. 2) A measure of field strength across a surface. For electric fields, the surface is taken to be closed; for magnetic fields, the surface is open.

Capacitance. 1) The ability of a device to store an electric charge. 2) The proportionality constant between charge and voltage.

Inductance. 1) The ability to "store" energy in the form of a moving charge (not correct). 2) Proportionality constant between magnetic flux and volume current.

Resistance. 1) The ability of a device to dissipate electric energy. 2) see Impedance.

Impedance. 1) The constant of proportionality between voltage and current phasors. 2) The matrix relationship between electric and magnetic field phasor-vectors.

Power. 1) Voltage times current: P = V x I. 2) The rate that energy leaves a given volume through its surface:

$$P = -\oint_s \bar{E} \times \bar{H}\, d\bar{s}$$

The right-hand side of this equation includes energy flowing in an electromagnetic field as well as in a moving charge.

tan δ. 1) Ratio of equivalent series resistance to reactance (in a capacitor). 2) Ratio of imaginary to real dielectric constant (in a material).

Properties of Ceramic Chip Capacitors

Having established a set of definitions, the behavior of these devices can be discussed. In addition to the usual electrical parameters, selected special characteristics of interest to RF engineers are included.

validInsulation Resistance. The most basic measurement of ceramic chip capacitor quality is its DC resistance. High insulation resistance indicates that the capacitor is well-designed, the body is properly manufactured, and the surface of the ceramic between contacts is free of contaminants such as dirt or flux. Class "A" dielectrics exhibit room temperature insulation resistance in the millions of megohms at up to three times the working voltage. Low insulation resistance usually indicates a problem in one of the above areas. (Troubleshoot-

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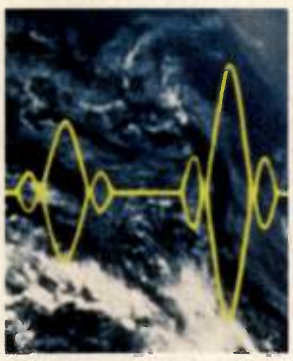

A typical "Spin Echo" pulse sequence

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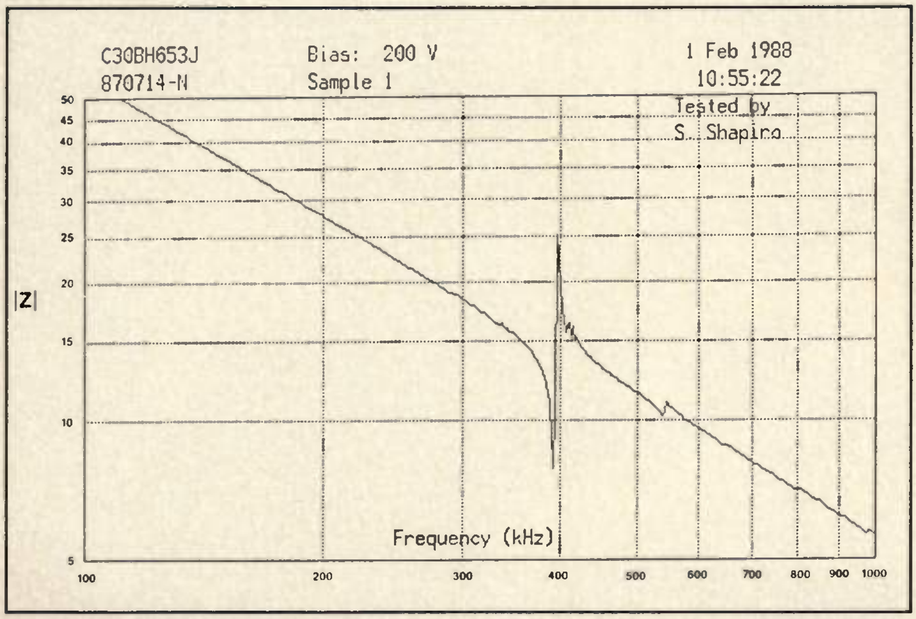

Figure 1. Impedance variation of a chip capacitor due to piezoelectric effects.

ing will be discussed in Part II.)

Q, or Dissipation Factor. A popular capacitor quality measurement is its Q factor, or the equivalent reciprocal relation known as dissipation factor. Though commonly requested by users and buyers, it seems not well understood. (Seven textbooks gave three different definitions: one based on capacitance or inductance and resistance, one based on stored and released energy, and one with no discernible meaning.) The perspective is very important.

The most general definition of Q would be: a measure of the ability of a system to store energy. The original intent was to measure the energy stored and released in a periodic system such as a harmonic oscillator, where the quantity has a clear meaning:

$$Q = \frac{2\pi\,(\text{avg. energy stored})}{(\text{energy dissipated})} \text{ per cycle}$$

Using ideal circuit elements, its not hard to take Q apart:

$$Q = \omega L/R = 1/\omega CR = \omega LG = G/\omega C$$

But if the element is distributed, like a transmission line, capacitance and inductance, resistance and conductance can not be separated. Here is where things get difficult, since real capacitors can behave as distributed elements. In practice, the validity of a Q measurement is related to how the application fits the energy storage definition. In a lumped element filter it is valid to ask about Q factors, but in coupling or decoupling application it may not be a valid question.

Equivalent Series Resistance (ESR) and Inductance (ESL). Consider two common problems. First, a monolithic chip capacitor (MLC) has these characteristics:

C(1 MHz) = 500 pF
ESL = 0.5 nH
ESR(f) = $.025\sqrt{f}$ ohms

Since Ceq will go to infinity at f = 318.3 MHz, will the Q factor of the device, at 318.3 MHz, be:

(a) $2\pi f(.5\text{ nH})/(.025\sqrt{f}\text{ ohms})$
(b) $1/[2\pi f(500\text{ pF})(.025\sqrt{f}\text{ ohms})]$
(c) zero

What is the Q factor at 1, 10, or 100 MHz?

Second, a manufacturer gives these specifications for a single layer capacitor:

$\varepsilon_r = 144$
$Q = 10{,}000(f/10\text{ MHz})^{-3/2}$
l = w = 1.25 cm
t = .025 cm
C(1 MHz) = 800 pF

The application bandwidth is 50 to 1250 MHz and requires fairly low loss, so the design engineer decides to test a sample. He finds the loss unacceptably high at the upper band edge. Should the device be rejected for nonconformance to the claimed Q factor?

The answer in both cases is "the customer is always right." This is little comfort to the engineer (or the vendor) when the product doesn't work. In the first example, the power dissipated will be at a broad local minimum because the capacitor is series resonant. Here, a practical alternative to the ambiguous Q factor is for the user to ask for ESR and ESL data. ESR can be found at radio frequencies by the use of a resonant line, such as the Boonton 34A. ESL can be estimated by the following relationship:

$$ESL = (1/C_1 - 1/C_2)(\omega_2^2 - \omega_1^2),\ \omega_2 > \omega_1$$

The second example is trickier, although the alert reader should be tipped off by the inclusion of dielectric constant and physical dimensions. The manufacturer has not lied about the Q factor in this instance. In fact, this effect becomes stronger with increasing Q factor because the capacitor is showing a parallel resonance effect, making it look lossy. Once again, specification of ESR (not Q) over the frequency band leads to the correct conclusion that the part will not fit the application.

Because electrical resonance is very important when fitting a capacitor into a particular application, a detailed discussion will be included in Part II of this article, which will emphasize applications and troubleshooting.

Piezoelectric Effects. One common ingredient for the dielectric in class "B" ceramic capacitors is barium titanate ($BaTiO_3$). In addition to having a dielectric constant of 1000 or greater, $BaTiO_3$ exhibits piezoelectricity — the property of generating voltage in response to physical pressure, or undergoing mechanical stress in response to an electric field. Ceramists have added special additives to reduce or eliminate this effect when $BaTiO_3$ is used in capacitors, but piezoelectric effects may still be present at high DC bias voltage.

One observable effect of piezoelectricity in ceramic capacitors is an abrupt change in impedance at a given frequency. This effect is a function of dimensional change resulting from the application of a voltage and a vibrational mode related to the size and shape of the capacitor and the sound velocity within the device (Figure 1). It is not known whether these resonances have a deleterious effect on capacitor performance. While reactance changes abruptly, no change in ESR at acoustical resonant frequencies has been detected by the measurements group at Dielectric Laboratories.

Another effect which is probably due to piezoelectricity is that a $BaTiO_3$ based device tends to retain some of its stored energy in its crystal lattice when first discharged. The crystal structure then changes back to its relaxed configuration, releasing the remaining "bound" charges to the electrodes. This effect can be hazardous as well as inconvenient.

Temperature Coefficient. The temperature coefficient of capacitance can be tailored to many applications by adjusting material formulations. This is widely known and does not need to be discussed in detail. Temperature dependence of loss is less well known, at least in capacitors. It is possible to design ceramic dielectrics with positive or negative temperature coefficient of loss.

Conclusion

Hopefully, this review of electrical and electromagnetic operating mechanisms sheds some light on the performance of ceramic chip capacitors at radio frequencies. This article is intended to instruct engineers in the basics of capacitor behavior at radio frequencies. Part II will build on this groundwork with an in-depth look at how capacitor characteristics affect performance in various RF applications.

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Computers as Victims of RF Interference

By Daryl Gerke, P.E.
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Most RF engineers are well aware of the potential for interference to computers and microprocessors due to nearby radio transmitters. Often lacking, however, are quantitative tools to actually predict such problems. This article provides some specific guidelines and methods to predict and assess this particular EMI (electromagnetic interference) threat.

For computer and microprocessor systems, interference problems generally fall into one of several categories: power disturbances, electrostatic discharge, lightning, FCC/VDE/MIL-STD regulations; and electromagnetic fields due to nearby radio transmitters. Several of these threats have been covered in previous articles (1,2). This article addresses the effects of high-level electromagnetic fields on modern computers and microprocessor-based equipment.

Due to these multiple threats, one of the first troubleshooting challenges is to prioritize them. Much can be learned by simply observing and asking questions. Has anyone felt static discharges? Do the lights flicker? Any thunderstorms recently? Any nearby radio, television, or radar transmitters? Anyone using handheld radios? Watch out for this last case — even low power, when it's nearby, is a big threat.

If radio transmitters are present, additional questions should be asked. First, are they powerful enough to cause a problem? If not, they can be ruled out. Second, what are typical symptoms? Third, what are typical failure modes?

When analyzing and trying to understand any EMI situation, it helps to divide the problem into categories. At the highest level, the three categories are the source, the victim and the coupling path. At the second level, typical categories address the coupling path. These include electromagnetic radiation, power and signal lines, crosstalk, and grounding. An example is when the source is a nearby radio transmitter, the victim is a computer/microprocessor system, and the coupling path is electromagnetic radiation.

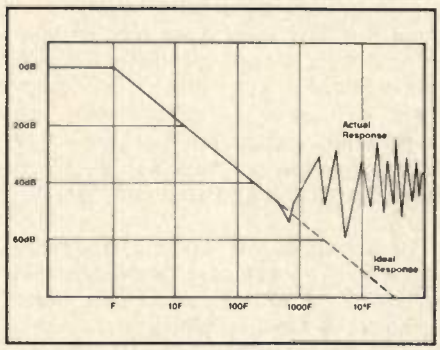

Figure 1. Analog amplifier frequency response.

Computers as Interference Victims

The first point to be made is that digital circuits behave as broadband receivers, with a bandwidth of 1/(π x risetime). For example, logic with a 3 nsec rise/fall time responds to energy from DC to about 100 MHz with no rejection; for 1 nsec logic, that frequency rises to over 300 MHz.

The second point is that the connecting cables act as receiving antennas, and can become very effective in the HF and VHF ranges. At about 1/20 wavelength, cables start to become quite effective as antennas. Thus, at 150 MHz (a wavelength of 2 meters), even a few inches of wire can be effective in picking up radiated energy. Even in the 27 MHz Citizens Band (11 meters), two feet of cable can cause problems.

It is this combination of high frequency response and efficient antennas that make modern computer equipment so susceptible to HF and VHF fields. The problems will get worse, no doubt, as rise/fall times become shorter and thus frequency responses go even higher.

Failure Modes

A primary failure mode for digital circuits is the creation of false signals due to exceeding circuit noise margins. Since most digital circuits have worst case noise margins of under 1/2 volt, a relatively low induced voltage at a logic circuit can cause a gate to switch states. The effect can be immediate, such as changing a control line, or latent, such as changing a memory location. In either case, the computer system has been corrupted.

A primary failure mode for analog circuits is rectification at the signal input. Here, the RF energy can be well outside the intended frequency range and still cause problems. If the induced voltage forward-biases a diode junction, the energy is demodulated and moved down in frequency. A demodulated CW carrier results in a DC bias, which can drive a low-level analog circuit into cutoff or saturation.

Rectification explains how a VHF radio at 150 MHz can jam a 100 Hz sensor, or a how a CB radio at 27 MHz

Military	1-200 volts/meter
Automotive	20-200 volts/meter
Industrial	1-10 volts/meter
Local Area Nets	2-5 volts/meter
Medical	400 volts/meter

Table 1. Typical electric field susceptibility specifications.

	1W	10W	100W	1kW	10kW	100kW
1 volt/meter	5.5 m	17 m	55 m	170 m	550 m	1.7 km
10 volts/meter	55 cm	1.7 m	5.5 m	17 m	5.5 m	170 m

Three Guidelines:

1 W @ 1 meter = 5.5 volts/meter

30 W @ 10 meters = 3 volts/meter

100 kW @ 1 km = 1.7 volts/meter

Table 2. Electric field levels vs. distance and power.

Frequency	λ	$\lambda/(2\pi)$	$\lambda/20$
1 MHz	300 m	48 m	15 m
10 MHz	30 m	4.8 m	1.5 m
30 MHz	10 m	1.6 m	50 cm
100 MHz	3 m	50 cm	15 cm
150 MHz	2 m	30 cm	10 cm
450 MHz	67 cm	10 cm	3.3 cm
1 GHz	30 cm	5 cm	1.5 cm

λ = wavelength
$\lambda/(2\pi)$ = near-field to far-field transition
$\lambda/20$ = cable becomes effective antenna

Table 3. Frequency, wavelength and critical lengths.

can interfere with a 20 kHz audio amplifier. In fact, it is quite usual for rectification to occur when the frequency of the source is 100 to 1000 times the frequency of the victim, due to parasitic effects. This is shown in Figure 1, which compares the theoretical and the real-world rejection of a hypothetical analog amplifier. The high-frequency degradation is due to parasitic capacitances and inductances (3).

Another failure mechanism that should not be overlooked is undesired coupling into a circuit's power or ground. These two sneak paths can allow undesired high-frequency energy into a circuit and cause upsets to both digital and analog devices.

Typical Failure Levels

It is difficult to predict the exact failure levels and mechanisms, and to predict which type of failure will occur first. In the author's experience, analog circuits such as sensors are usually more susceptible, but each system is unique. Note that upsets are a function of cable length (antenna), circuit margins (sensitivity), and frequency (bandwidth). Add in nonlinear effects such as rectification, and it's a tossup.

Guidelines do exist, however, and are

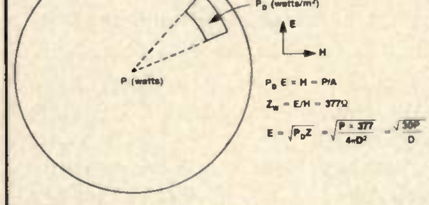

Figure 2. Electric field vs. power and distance.

based on practical experience. For today's microprocessor-based systems, a good rule of thumb is an electric field level of 1 to 10 volts per meter. In field levels below 1 volt/meter, most equipment will work; at greater than 10 volts/meter, equipment will likely fail unless special precautions, such as shielding and high-frequency filtering, are taken.

How do these levels compare with the real world? It's not unusual to have field levels of 10 to 100 volts/meter near commercial radio or television transmitters. Levels of several hundred volts per meter have been reported on cars and trucks due to on-board land mobile transmitters. Near airports or topside of naval vessels, the fields can easily reach hundreds of volts per meter due to the plethora of communications and radar systems.

Recognizing this, a number of specifications limits have evolved, which are summarized in Table 1. For equipment to operate in any of these severe environments, the designer will likely need to perform susceptibility tests.

Predicting and Assessing Threats

To make a quick assessment of a potential threat, the following example questions should be considered. Is that 100 kW radio station two miles away a problem? What about the security guard's 1 watt hand-held radio? Is the microwave link on the next skyscraper a potential problem?

Fortunately, it is relatively easy to make that initial prediction if some simplifying assumptions are made. First, assume a point source; second, assume far-field conditions; and third, assume free space. Then, the electric field at any distance, knowing the radiated power of the source, can be predicted by the following formula:

$$E = \sqrt{30P}/d$$

where E is the electric field in volts/meter, P is the effective radiated power in watts, and d is the distance from the source to the victim in meters. The simplified model is shown in Figure 2. The formula was derived using the surface area of a sphere and using 377 ohms as the wave impedance of free space.

To illustrate the levels, Table 2 shows distances vs. transmitter power levels at the 1 and 10 volts per meter guidelines. To answer the questions posed above, the 100 kW transmitter at 2 miles is not a threat, but the 1 watt hand-held radio is. For the microwave link, the engineer only needs to know the power and the direct distance to make a quick assessment.

Are the assumptions valid? For a first approximation of radio transmitters, they are. The free space assumption should give a worse case that neglects reductions for shielding. The far-field assumption is valid if the distance between the source and victim is greater than $\lambda/2\pi$, or about 1/6 wavelength. The point source assumption is valid if the effective radiated power (transmitter output power multiplied by the antenna gain in the direction of the victim) is used rather than transmitter power.

Incidentally, precision is not required here, just a quick go/no-go decision. If more precision is needed, tests can be run with a spectrum analyzer or receiver and antennas.

For reference, Table 3 shows frequency, wavelength, the near-field/far-field distance of $\lambda/2\pi$, and $\lambda/20$, a criteria for when cables become effective antennas. It lists frequencies from 1 MHz to 1 GHz, resulting in wavelengths from 300 m to 30 cm.

Case Histories

This material has been used numerous times to predict and assess potential problems with nearby radio transmitters. The following situations are examples.

1. Power Disturbances

The computers controlling a production line were failing at random, resulting in production halts. In fact, occasionally the computer power supplies were being destroyed.

The computer vendor's field engineer was insisting that a radio station two miles away was the source of the problem, and was recommending extensive shielding. (Have any of you hams out there been blamed like this because of your big antennas?) A quick assessment predicted a field level well below the 1 volt/meter criteria.

The real problem was due to power disturbances, and was resolved with isolation transformers and some changes in the system design (4).

2. Mobile Radio Interference

In this case, a new fire truck worked fine, until the on-board mobile radio was used. At that point, the water pressure in the pumper would abruptly increase when the transmitter was keyed. In fact, even a nearby hand-held radio would cause the system to malfunction.

A quick assessment showed field levels easily in excess of 1 volt/meter. Not known, however, was the actual failure mode. Was it the analog pressure transducer, the microprocessor controller, or the engine control electronics?

Since the pressure transducer was connected to the microprocessor via a long cable (also acting as an undesired antenna), it was immediately suspected. Further investigation with a hand-held radio confirmed that rectification in the transducer was probably the problem.

This case had a second problem. After clearing up the transducer rectification, it was discovered that the DC voltage to the transducer was also being affected by the transmitter. Not an RF problem, this was simply a common impedance power distribution problem. Nevertheless, it's believed this second problem contributed to the first, since the op amp transducer was on the very edge of its operational range, and thus more susceptible to RF.

3. Microwave Links and Software

The manager of a law firm had problems with a newly networked computer system. The obvious possibilities included software, hardware, power disturbances and electrostatic discharge.

But, what about that microwave link on the next building, at eye level with the law offices? After finding the power level, a quick assessment showed that the field levels could be in the 1 volt/meter range. Ferrites were installed on several cables as a precaution against this potential threat.

In this situation, however, it was finally determined that software was the problem. Fortunately, not every computer malfunction is an EMI problem.

Summary

Computers can indeed be victims of electromagnetic fields from nearby radio transmitters. A quick assessment of potential threats can be made with a simple formula that predicts a free space field intensity. If this prediction is greater that 1 volt/meter, the threat is real.

Both digital and analog circuits can be affected. Digital circuits are vulnerable because of their broad bandwidth, while analog circuits are vulnerable due to rectification and circuit parasitics. Cables act as pick-up antennas for both types of circuits, and any cable longer than 1/20 wavelength is a possible suspect.

Finally, expensive test equipment or complicated calculations are not necessary to make a quick assessment of the problem. Checking some vital signs can provide a lot of information and insight into the problems with very little effort. ■

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Design of Wideband Quadrature Couplers for UHF/VHF: Part I

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This article presents design information for a class of wideband quadrature couplers consisting of three narrowband quadrature couplers separated by two pairs of transmission lines of adequate length and characteristic impedance Z_o. Due to the use of lumped elements, the coupler is particularly useful in the UHF/VHF band. Experimental results agree well with the theoretical computations.

The configuration of the narrowband quadrature coupler is shown in Figure 1. On a toroid, two strands of insulated wire are tightly twisted together to form a bifilar pair. The self inductance and mutual inductance of the transformer toroid are each equal to L. (This assumes unity magnetic coupling between the inductors.) Two lumped capacitors of value C/2 are shunted across the primary and secondary windings of the transformer. The design equations for this coupler can be found in References 1 and 2 and are summarized below. Let Z_o be the input and output impedance of the coupler. L and C are chosen such that:

$$Z_o^2 = L/C \quad (1)$$

The coupler is then perfectly matched for all frequencies, and port 1 (or 2) and port 3 (or 4) are perfectly isolated for all frequencies. The voltages at ports 2 and 4 with 1 volt applied at port 1 are:

$$V_{21} = \frac{j\omega L}{Z_o + j\omega L} \quad (2a)$$

and

$$V_{41} = \frac{Z_o}{Z_o + j\omega L} \quad (2b)$$

The phase of V_{21} leads that of V_{41} by 90 degrees for all frequencies. The amplitude responses of $V_{21}{}^2$ and $V_{41}{}^2$ are plotted in Figure 2, showing the 0.6 dB amplitude ripple bandwidth to be about 10 percent.

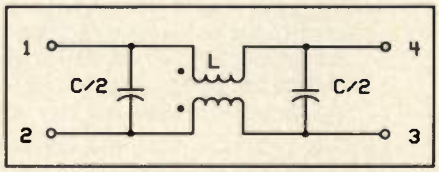

Figure 1. A narrowband quadrature coupler.

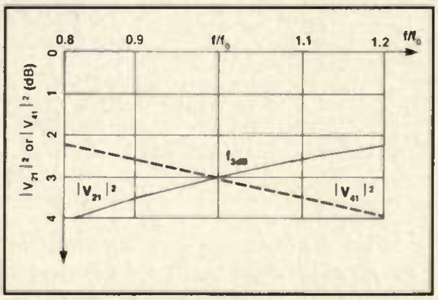

Figure 2. Amplitude response of the narrowband quadrature coupler.

The frequency at which $V_{21} = V_{41}$ is designated as f_{3dB}. At f_{3dB}:

$$Z_o = \omega L = 2\pi f(3\ dB)L \quad (3a)$$

or,

$$L = \frac{Z_o}{2\pi f(3\ dB)} \quad (3b)$$

The capacitor which is related to L by equation 1 can be expressed as:

$$C = \frac{1}{2\pi f(3\ dB)Z_o} \quad (4a)$$

or

$$\frac{C}{2} = \frac{1}{4\pi f(3\ dB)Z_o} \quad (4b)$$

If the f_{3dB} of the narrowband quadrature hybrid is known, the capacitor C and inductor L can be computed by equations 3 and 4. The amplitude response

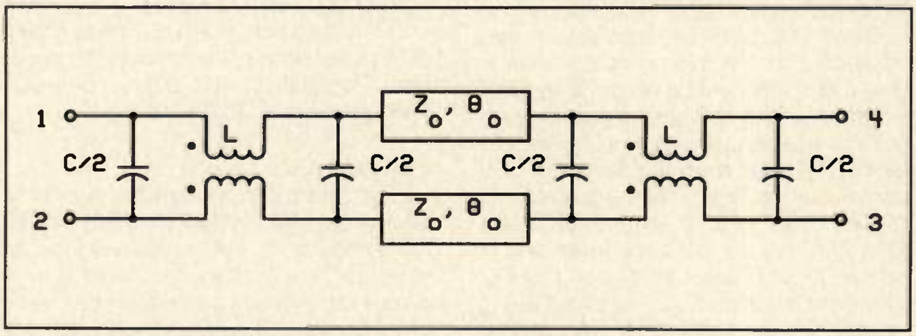

Figure 3. Two identical narrowband quadrature couplers.

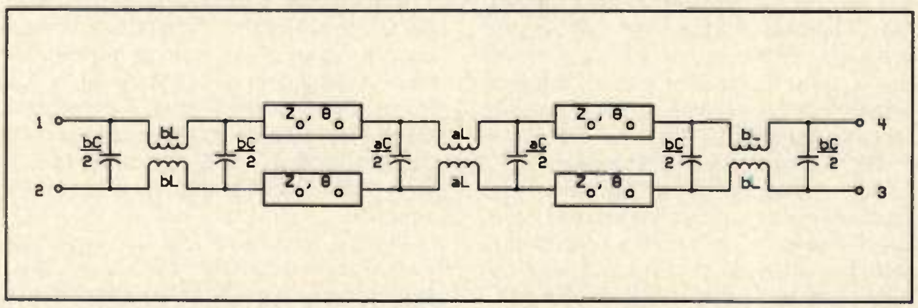

Figure 4. Configuration of the broadband quadrature coupler.

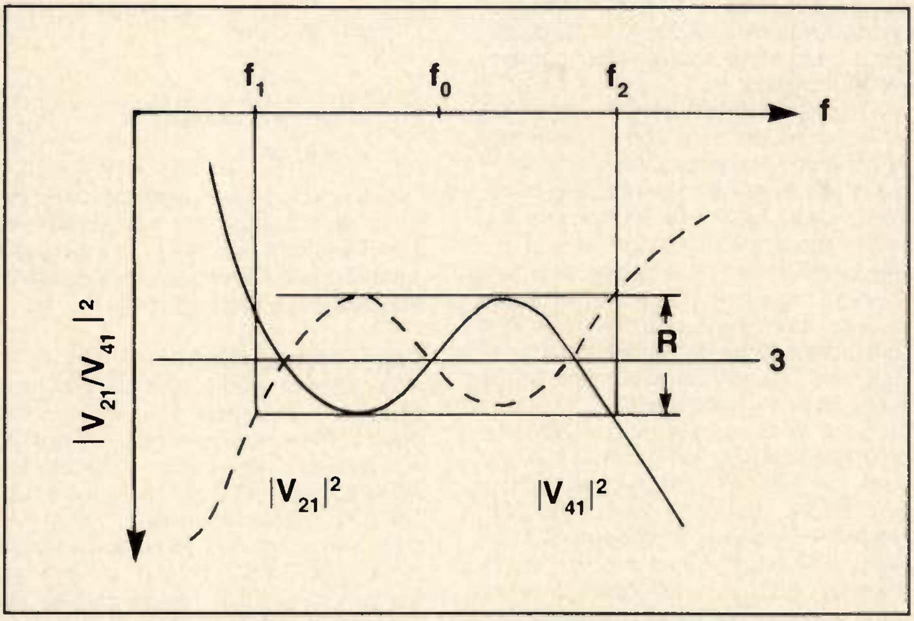

Figure 5. A typical frequency response of the broadband quadrature coupler.

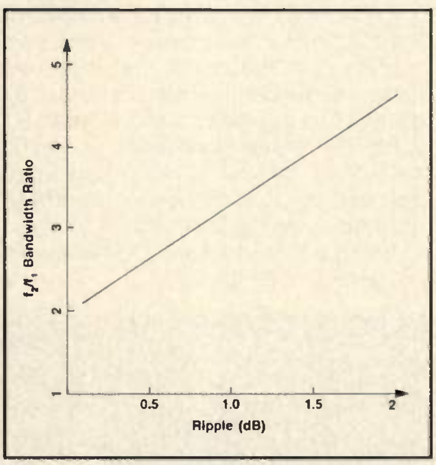

Figure 6. f_2/f_1 vs. Ripple R(dB).

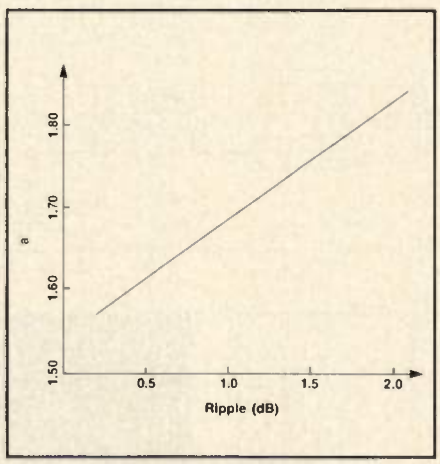

Figure 7. a vs. Ripple (dB).

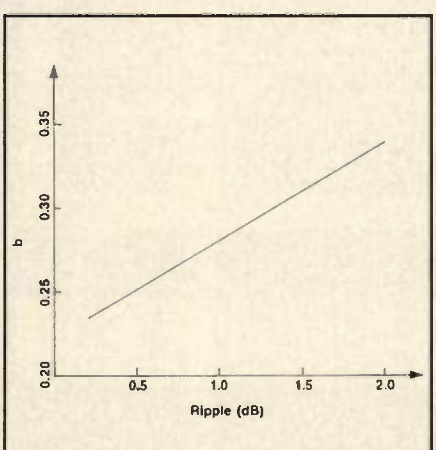

Figure 8. b vs. Ripple (dB).

of the coupler can be seen in Figure 1. Therefore, f_{3dB} completely represents the characteristics and the design of this narrowband quadrature hybrid.

The operating frequency bandwidth of the coupler can be broadened by using two identical narrow couplers of this type interconnected by a pair of transmission lines of characteristic impedance Z_o and of length θ, as shown in Figure 3.

The design parameters of this coupler can be found in References 1 and 3. Several design curves are available to facilitate the design. It has been reported that the 0.6 dB amplitude ripple bandwidth of this coupler is about 67 percent.

Design of Broad Bandwidth Quadrature Couplers

The broadband quadrature coupler under consideration has the configuration shown in Figure 4. It consists of two identical narrowband couplers (described in the previous section) at both ends and another narrowband coupler of different f_{3dB} in between.

The couplers are separated by two pairs of transmission lines of characteristic impedance Z_o and electrical length θ_o at f_o, the center frequency of interest. Note that the narrowband couplers at both ends have f_{3dB} equal to f_o/b, and the middle coupler has its f_{3dB} at f_o/a. Design parameters for this broadband coupler can then be simplified to the three parameters a, b and θ_o. The inductor L and capacitor C are related to the center frequency f_o and source/load impedance Z_o by:

$$L = \frac{Z_o}{2\pi f_o} \tag{5a}$$

$$C = \frac{1}{2\pi f_o Z_o} \tag{5b}$$

Analysis of the broadband coupler of Figure 4 can be carried out either by an even-mode and odd-mode approach of analysis, or by cascading techniques which use a transmission matrix and scattering parameter matrix (4). Both approaches arrive at identical results. Let R be the output power ratio between port 2 and port 4 for input power at port 1. Then:

$$R = \frac{|V_{21}|^2}{|V_{41}|^2} \tag{6}$$

$$= \left[\frac{a\omega L}{Z_o} + \frac{2b\omega L\cos(2\theta)}{Z_o} - \frac{(2a+b)b\omega^2L^2\sin(2\theta)}{Z_o^2} + \frac{4ab^2\omega^3L^3\sin^2(\theta)}{Z_o^3}\right]^2$$

$$= \left[a\left(\frac{f}{f_o}\right) + 2b\left(\frac{f}{f_o}\right)\cos\left(\frac{2\theta f}{f_o}\right) - (2a+b)b\left(\frac{f}{f_o}\right)^2\sin\left(\frac{2\theta_o f}{f_o}\right) + 4ab^2\left(\frac{f}{f_o}\right)^3\sin^2\left(\frac{\theta_o f}{f_o}\right)\right]$$

To find the design parameters a, b and θ_0 for a given R, a computer program has been written to compute the response of the broadband quadrature coupler. Given the desired amplitude ripple R in a nominal 3 dB design and center frequency f_0, the program (after some optimization) will provide these design parameters. Figure 5 shows a typical frequency response of the broadband quadrature coupler after parameter optimization.

The optimization criteria are established to adjust a, b and θ_0 such that equal ripple response of the amount R dB is achieved with the resulting bandwidth ratio, f_2/f_1. The relationship between the amplitude ripple R and the bandwidth ratio f_2/f_1 is shown in Figure 6 where higher passband ripple results in wider bandwidth ratio. Figures 7, 8 and 9 present the design information for a, b and θ_0, also as a function of the passband amplitude ripple.

As a design example, f_0 = 80 MHz, Z_0 = 50 ohms and R = 0.6 dB; f_2/f_1 = 3.0 with f_2 = 120 MHz and f_1 = 40 MHz; a and b are equal to 1.63 and 0.26, respectively; and θ_0 = 35 degrees at 80 MHz. The narrowband coupler at the ends has an f_{3dB} of 307.7 MHz and the middle coupler has an f_{3dB} of 49.08 MHz.

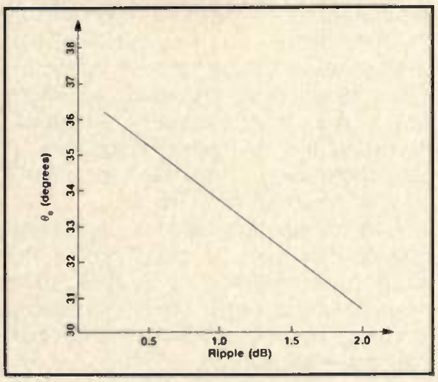

Figure 9. θ_0 vs. Ripple (dB).

$$L = \frac{50}{(2\pi)(80 \times 10^6)} = 99.47 \text{ nH} \qquad (7a)$$

$$\frac{C}{2} = \frac{1}{2(2\pi)(80 \times 10^6)(50)} \qquad (7b)$$

$$= 19.89 \text{ pF}$$

The element values of the broadband quadrature coupler for this design are summarized in Figure 10. The computer simulation of the frequency response of the design is shown in Figure 11.

Experimental Results

A broadband quadrature coupler was fabricated and tested based on the design information provided in the previous section. The circuit was fabricated using 25 mil thick Epsilam-10 substrate. The circuit was constructed using ATC chip capacitors and Johanson variable capacitors. Eight turns of 34 bifilar wires are used on a T20-10 toroid core for the realization of 162.1 nH and five turns of 34 bifilar are used on a T20-0 toroid core to obtain 25 nH.

A photo of this coupler is shown in Figure 12 and the measured frequency responses in Figure 13. The measured passband amplitude ripple is about 0.8 dB and the bandwidth ratio is 143/45 = 3.18. The center frequency is slightly shifted to 84 MHz. The agreement between the theoretical and the measured responses is excellent.

Part II will be published in an upcomming issue of *RF Design*.

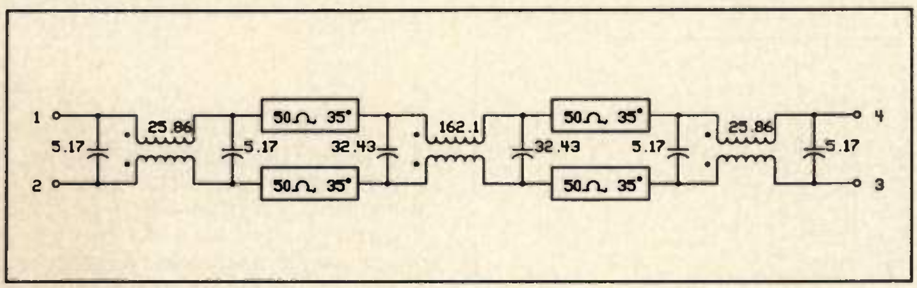

Figure 10. Experimental circuit.

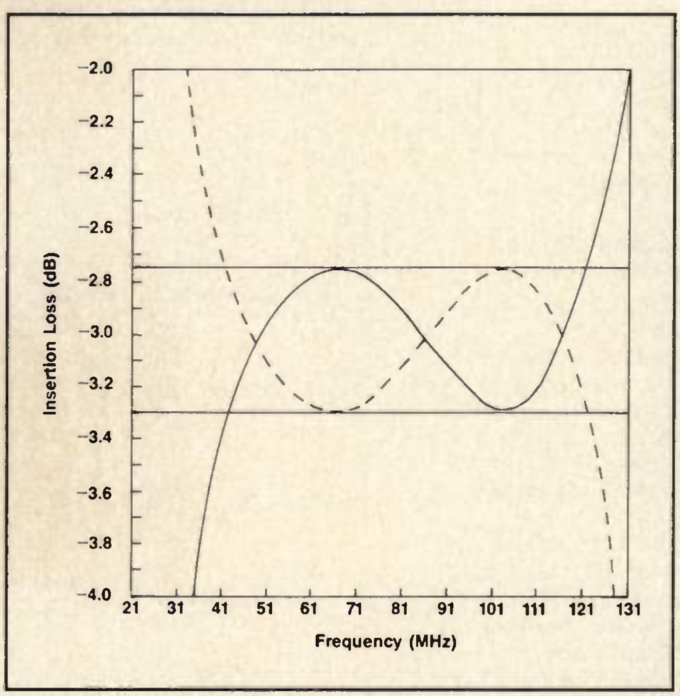

Figure 11. Computer simulation frequency response.

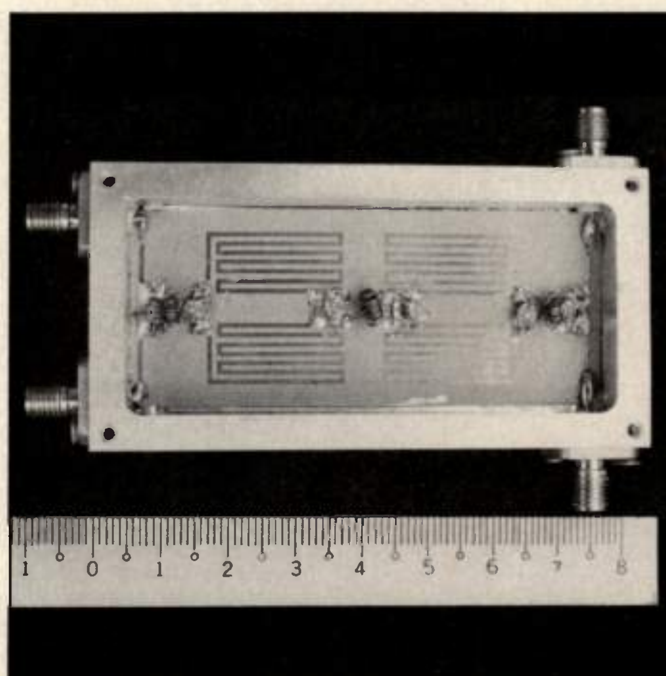

Figure 12. Photo of the coupler.

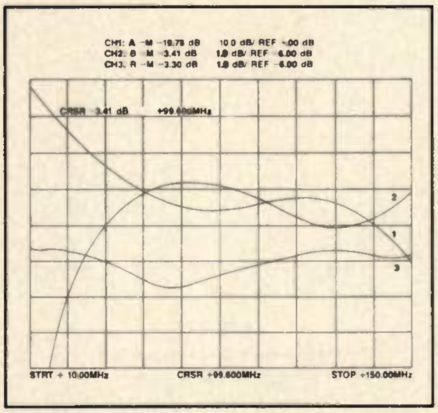

Figure 13a. Measured frequency response.

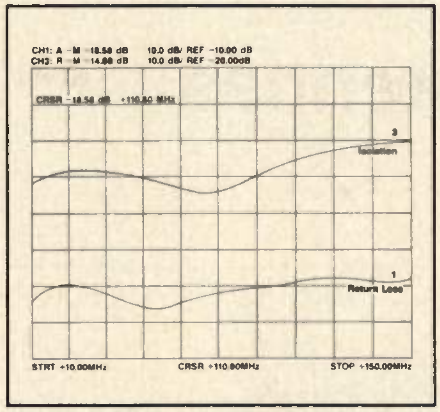

Figure 13b. Return loss and isolation.

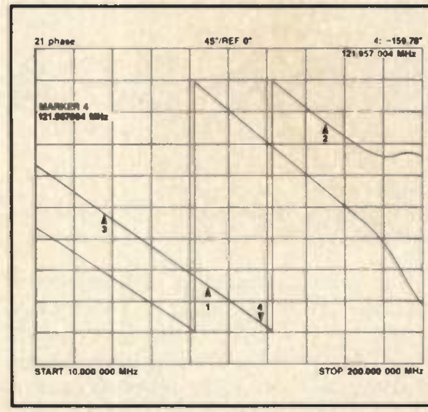

Figure 13c. Measured phase relation between T2I and T4I.

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Active DF Antennas

By Henry W. Anderson and R. Stephen Smith
Watkins-Johnson Company

An active antenna is defined as a passive element, usually electrically small, terminated at its input by an active device (amplifier). The principle advantage in using an active antenna over a passive one is its smaller physical size. In an attempt to utilize this advantage, a design was undertaken to reduce by over 50 percent the size of a 20-140 MHz direction finding (DF) array. The typical design parameters for an active antenna are bandwidth and sensitivity; however, additional constraints are imposed by the DF application. These constraints consist of amplitude and phase matching the array elements. The successful design of an HF active monopole DF array provided additional motivation for attempting to use a similar technique at VHF frequencies.

Active antenna technology has other potential benefits in addition to the substantial reduction of element length. One advantage is that the output of an active element provides a more constant termination for connecting coaxial cables. In this application, the reverse isolation of the amplifier also significantly reduces electromagnetic interference (EMI) from the DF array.

Unfortunately, using active antennas does not solve all problems. Design tradeoffs still exist, especially in the area of amplifier topology and circuit design. These compromises are accentuated by the additional amplitude and phase matching requirements of DF array design. Specifically, an optimum impedance match over the entire frequency range may not be possible. Decreased sensitivity at low or high frequency band edges may then occur. This may be acceptable at the lower frequencies if the antenna is externally noise limited, since the signal-to-noise ratio will then not degrade. At the higher frequencies, a low-noise amplifier is required, to assure that antenna performance is limited only by external noise and not by internal amplifier noise.

This article begins with a brief review of antenna terminology and fundamental theory. The concept of effective area is introduced and used to derive an antenna conversion factor. This antenna conversion factor provides a convenient link for calculating antenna output signal levels from electric field strength. External noise sources are listed and their effects on system design requirements tabulated. Design tradeoffs are then explored as they relate to the amplifiers, the elements and their interaction. Experimental results based on the implementation described are then given.

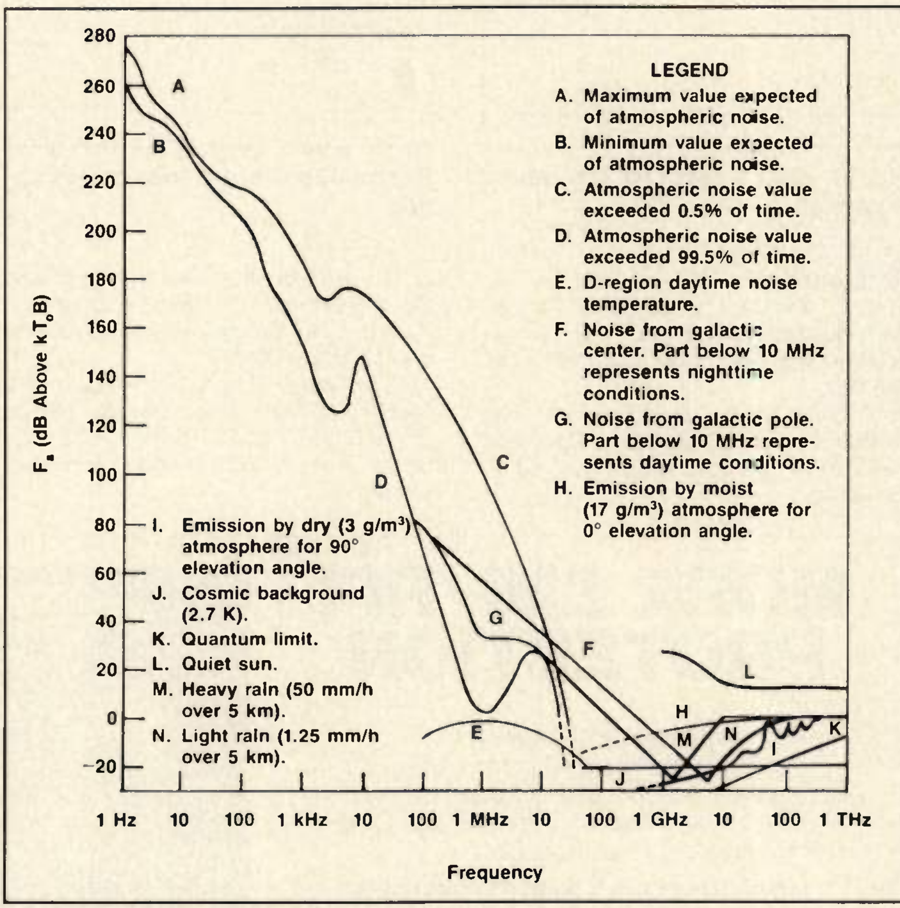

Figure 2. Sky temperature.

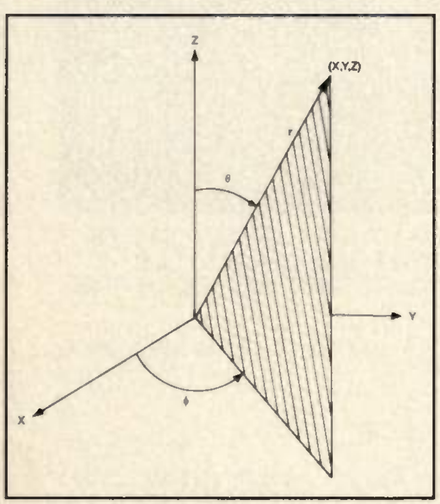

Figure 1. Spherical coordinate system (x, y, z).

Antenna Terminology

Antenna elements are physical structures capable of supporting an electric current distribution **J(r)** which, in turn,

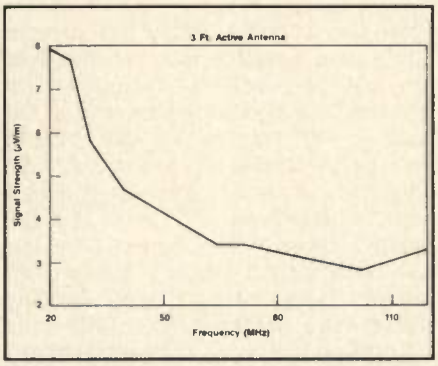

Figure 3. Minimum useable signal.

produces a radiated field. (It should be noted that the boldface letters indicate vector quantities.) In antenna analysis, the current distribution is usually either known, assumed or approximated numerically (e.g., method of moments). Once the current distribution is known, the radiated fields are derived from Maxwell's equations (1). In the far field of an antenna (radial distance, r, much greater than a wavelength) located at the origin of the coordinate system shown in Figure 1, this solution reduces to:

$$\mathbf{E} = j\omega\mu\mathbf{A} \quad (1)$$

$$\mathbf{H} = (\hat{r}\mathbf{x}\mathbf{E})/(120\pi) \quad (2)$$

where **A** is the magnetic vector potential, given by:

$$\mathbf{A} = \frac{e^{-j\beta r}}{4\pi r}\iiint_{v'} J(r')e^{j\beta\hat{r}\cdot r}dv' \quad (3)$$

In the above, **r** indicates the field coordinate, **r**′ the source points and v′ the source volume.

The radiation pattern of an antenna can now be defined as the variation of the electric field at a fixed radius. Equation 1 thus becomes:

$$\mathbf{E}(r) = \mathbf{E}(\theta,\phi) \quad (4)$$

Normalizing equation 4 yields:

$$F(\theta,\phi) = E(\theta,\phi)/E_{max} \quad (5)$$

This equation leads to the definition of directivity, an important antenna parameter. Directivity is simply a measure of how well the antenna concentrates energy in a certain direction. In mathematical terms:

$$D(\theta,\phi) = \frac{|F(\theta,\phi)|^2}{\frac{1}{4\pi}\int_\Omega |F(\theta,\phi)|^2\, d\Omega} \quad (6)$$

This equation is basically an expression for the power radiated in the (θ,ϕ) direction divided by the average power radiated. If the radiation efficiency of the antenna, e, is included, the gain is obtained as:

$$G(\theta,\phi) = eD(\theta,\phi) \quad (7)$$

It is common practice to describe the maximum value of equation 7 as the gain of an antenna, usually given in decibels. For example, an ideal short dipole has a gain of:

$$G = 10 \log 1.5 = 1.76 \text{ dB} \quad (8)$$

Another important property of an antenna is the input impedance presented by the antenna at its terminals. The input impedance of an antenna is expressed as:

$$Z_a = R_a + jX_a \quad (9)$$

where R_a is the antenna resistance, representing the combination of radiated and ohmic power dissipation; and X_a is the antenna reactance, representing the power stored in the near field of the antenna. The antenna resistance, R_a, is determined by:

$$R_a = \frac{2(P_r + P_{ohmic})}{|I_{in}|^2} \quad (10)$$

where I_{in} is the current at the antenna terminals; P_r is the radiated power, found by integration of the far field complex Poynting vector over a closed surface surrounding the antenna; and P_{ohmic} is simply the ohmic loss obtained from knowledge of the resistivity of the antenna and the current distribution. These definitions conveniently provide an expression for the antenna efficiency introduced in equation 7:

$$e = \frac{P_r}{P_r + P_{ohmic}} \quad (11)$$

The receiving properties of an antenna can now be investigated. A common and intuitive way of expressing the voltage induced across the terminals of an antenna by an electric field is in terms of effective length, or:

$$V_a = l_e|\mathbf{E}| \quad (12)$$

where l_e is the effective length and is given by:

$$l_e = \frac{1}{I_o}\int_{-l/2}^{l/2} I(z)\, dz \quad (13)$$

The concept of effective length is useful, but is not general enough for all antennas. A more rigorous concept is that of effective aperture, given by:

$$A_e(\theta,\phi) = \frac{G(\theta,\phi)\,\lambda^2}{4\pi} \quad (14)$$

with the gain as defined in equation 7. The power received by the antenna and transferred to a matched load is simply the effective aperture (area) multiplied by the power per unit area, or:

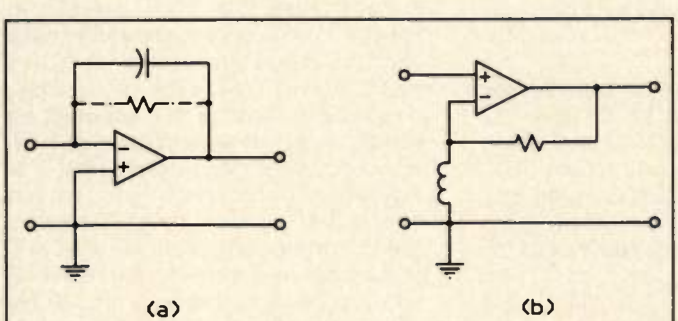

Figure 4. Conventional (a) and Sallen-key (b) integrator topologies.

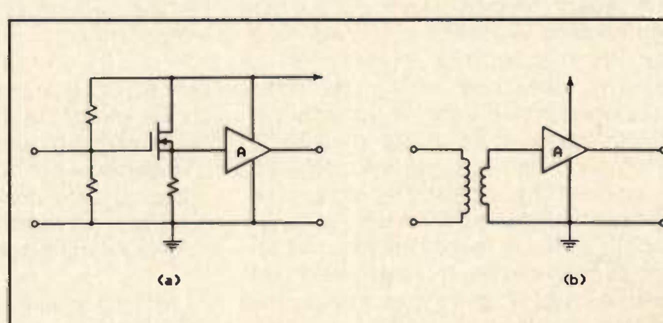

Figure 5. High input impedance amplifiers: (a) FET input amplifier and (b) transformer input amplifier.

$$P = A_e S \tag{15}$$

with S being the magnitude of the Poynting vector of the incoming field. Equation 15 can be modified to include the effects of mismatch and polarization as follows:

$$P = qpA_e S \tag{16}$$

where q is the loss due to mismatch and p is the loss due to polarization.

Equation 16 provides the desired relationship between received power and field strength. This relationship can be expanded upon by noting that the magnitude of the Poynting vector is:

$$S = \frac{|\mathbf{E}|^2}{120\pi} \tag{17}$$

and that power P is:

$$P = V^2/50 \tag{18}$$

for a 50 ohm system in free space. Thus, equation 16 can be rearranged to obtain:

$$|\mathbf{E}|^2 = \frac{2.4\pi V^2}{qpA_e} \tag{19}$$

The above equation can be expressed in decibel form as follows:

$$20 \log|\mathbf{E}| = 8.8 - 10 \log (qpA_e) + 20 \log V \tag{20}$$

$$20 \log|\mathbf{E}| = ACF + 20 \log V \tag{21}$$

$$ACF = 8.8 - 10 \log (qpA_e) \tag{22}$$

where ACF is the antenna conversion factor or transform between voltage output by the antenna into a 50 ohm load and electric field strength. This is a commonly used parameter in the characterization of an antenna.

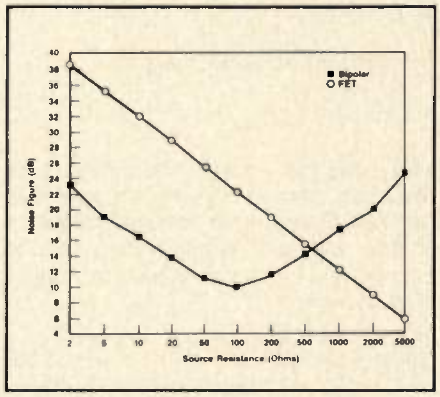

Figure 6. Noise figure vs. source resistance.

Noise

A major limiting factor in system performance is noise. It is preferable that the system be limited only by external noise and not by the noise generated in the antenna amplifier or accompanying components. In general, external noise is composed of both ground and sky noise and is represented by noise temperatures, such as T_{ground} and T_{sky}. For purposes of this discussion, a ground noise temperature of 290 K is assumed. This is a good approximation for nonreflecting surfaces.

Sky noise consists primarily of atmospheric, galactic and manmade components. At low frequencies, the largest of these is atmospheric noise, which results primarily from lightning noise propagating via the ionosphere. For this reason, atmospheric noise is dependent on factors such as time, weather, season and location.

Galactic noise originates from sources outside the earth's atmosphere such as the sun, moon and stars. The contribution of galactic noise is primarily at frequencies above 15 MHz, but it usually is not a major noise source below 100 MHz. Variations occur due to phenomena such as sunspots and solar flares.

Manmade noise is produced by electric motors, power lines, etc. In most situations, manmade noise is insignificant. In urban areas, however, it can be dominant at low frequencies.

When considering antenna performance, it is convenient to represent the external noise by a constant called the antenna noise temperature. An expression for antenna temperature is developed by Krauss (2) by integrating the noise power multiplied by the effective aperture of the antenna, over the beam solid angle covered by the antenna. The result is:

$$T_a = \frac{1}{4\pi} \iint_{\Omega} T(\theta,\phi)\, G(\theta,\phi)\, d\Omega \tag{23}$$

where $T(\theta,\phi)$ is the temperature of the noise source. In this case, equation 23 can be used to determine the noise temperature of a short dipole. For simplicity, the assumption is made that the sky temperature is uniform (i.e., independent of position). This results in:

$$T_a(\text{short dipole}) = (1/2)(T_{sky} + 290)\ \text{K} \tag{24}$$

The antenna temperature concept allows external noise to be modeled as a system parameter. By knowing the characteristics of the other system components, it is possible to relate the field strength and external noise at the antenna to a signal-to-noise ratio at the radio input. This relationship (and a knowledge of the signal-to-noise ratio necessary at the DF processor) allows one to determine the field strength needed, given a specific sky temperature. A computer program was written to perform this transformation. The sky temperature was taken from Reference 3 (see Figure 2) using the relationship:

$$kT_a B = F_a + kT_o B \tag{25}$$

The internal noise of the amplifier was approximated from the manufacturers's noise figure data in a 50 ohm system. All other noise components were assumed to be limited by thermal noise. The computer program then calculated the minimum useable signal. This signal level is based on a 10 dB signal-to-noise ratio in a 10 kHz bandwidth. The results are shown in Figure 3. Analysis of the external noise power at the output of the amplifier verified that the antenna is indeed externally noise limited up to approximately 70 MHz.

Amplifiers

There are many considerations and tradeoffs in any amplifier design; however, there are additional factors to consider when that amplifier is to be used in an active antenna. Noise performance, as previously discussed, is a key characteristic, as it is with most VHF radio frequency design. The reliability (or robustness) of the circuit is of particular importance due to the environmental extremes to which the antenna could be exposed. Cost, as in all but a few cases, is also a concern.

In addition to noise performance, reliability, and cost, an active DF array application requires that other factors be considered. Since an antenna is exposed to the entire electromagnetic spectrum, including nearby transmitters and occasional nearby lightning strikes, some type of input protection must be included. A dipole antenna element is a balanced output device; therefore, if a single-ended amplifier is used, ground currents will be induced in the cable shields. These ground currents will substantially alter the predicted antenna performance. It is therefore necessary that some type of balanced input topology be used.

A final characteristic required by the amplifier relates to direction finding. The

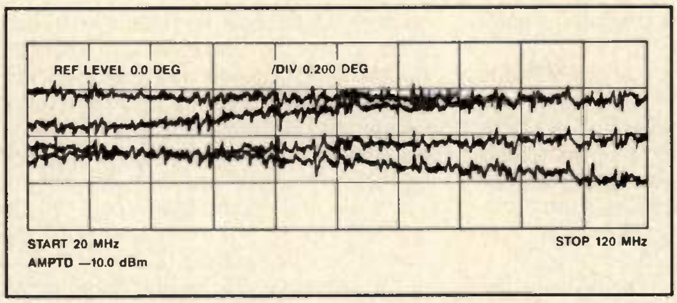

Figure 7. Measured phase matching between transformer-coupled amplifiers.

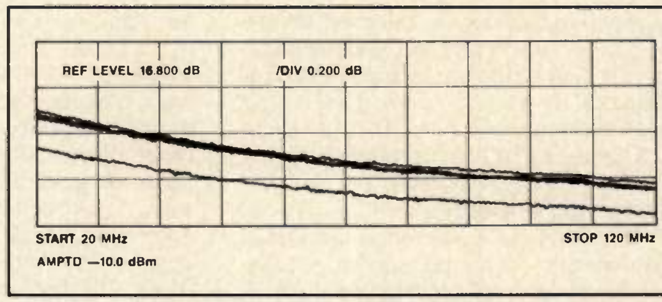

Figure 8. Measured amplitude matching between transformer-coupled amplifiers.

particular DF system for which this antenna is designed derives angle-of-arrival information from the phase of the incoming plane wave. Therefore, signal paths from each of the sensor elements must be of the same electrical length, i.e., phase matched. Calculated angles-of-arrival will be distorted and accuracy reduced by small variations in phase. Phase matching between the active antenna amplifiers is, therefore, a key specification. All amplifiers (four, in this case) must be matched to within 0.5 degrees at the lower frequency band edge.

Various design approaches are available which may meet the required characteristics. The tradeoffs involved in each approach must be weighed to select the best overall circuit. Much has been written about active antennas and their application in the HF frequency spectrum (8,9,10). The articles referenced discuss only two approaches: the more typical FET input (high impedance) design and a virtual ground or integrator (low impedance) design. A third logical choice is to design a conjugate-matched amplifier, thereby providing maximum power transfer from the antenna element (dipole, in this case) to the amplifier. Significant differences exist between the HF frequency spectrum (2-20 MHz) and the VHF spectrum (20-140 MHz) where the proposed antenna must operate. The major difference is that external atmospheric noise is much higher at HF and decreases as frequency increases (see Figure 2). This phenomenon requires that the integral antenna amplifiers at VHF have better noise performance than their counterparts at lower frequencies. Tradeoffs exist for each of the above techniques and will now be discussed.

A circuit design attempting to conjugate match the antenna element to the amplifier quickly becomes complex due to the change in element impedance over the frequency range. Fano's limit gives the number of circuit components required to provide the impedance match over this frequency range. Since many components are required, phase matching can only be accomplished by tuning or by requiring very tight tolerances on these components. Either of these techniques is expensive and, due to complexity, not adequately reproducible.

Low input impedance amplifier designs typically incorporate feedback not only to set the input impedance and circuit transfer function, but also to reduce parameter sensitivities. High-speed operational amplifiers are commercially available; however, their use is not without problems. Care must be taken in the physical implementation of the feedback path. Instabilities can easily surface as spurious oscillations or notches in the gain response. Also, some circuit topologies are unrealizable at higher frequencies due to stray capacitance. In particular, the conventional integrator circuit topology must be modified due to the instabilities caused by capacitive feedback. An alternative Sallen-key topology is shown in Figure 4. Careful circuit layout is still required to prevent oscillations.

A high input impedance circuit design as shown in Figure 5 converts voltage at its input to power at its output. Its high input impedance (ideally an open circuit) does not load the antenna element output terminals. The degree to which the antenna voltage is loaded down is a function of the amplifier impedance and antenna element impedance. This interface can be viewed as a frequency-dependent voltage divider; therefore, a reasonable compromise for the amplifier input impedance (load) is between three and five times that of the antenna element (source). This represents a compromise between extremely high impedance levels and excessive loss through the equivalent voltage divider. An amplifier output impedance of 50 ohms is also desired. The FET source follower (A = 0.9) configuration is the simplest way to accomplish this conversion. It is then typically followed by a gain stage to restore the correct signal level. Its useable high-end frequency or cutoff frequency is limited by source resistance and C_{gd}, gate-to-drain capacitance. A bipolar equivalent approach is to use a broadband transformer and 50 ohm input impedance amplifier to provide a high input impedance. This approach also has the added capability of a balanced input which can be directly connected to the dipole element.

High and low impedance amplifiers were evaluated further, while the conjugate-matched approach was abandoned. Conjugate matching was dropped due to the number of close tolerance components required to perform an adequate match over the 20-140 MHz frequency range. Several amplifier configurations were tested. The low impedance amplifier was found to have excessive noise and be susceptible to spurious oscillations. Two high input impedance designs were also tested (see Figure 5). The first circuit had a FET input device to provide the desired input characteristic. The second used a transformer in front of a bipolar gain block. Tests showed that the transformer input amplifier not only had better noise characteristics, but also had more repeatability in its phase and amplitude responses. A plot of noise figure vs. drive impedance reveals why these results are expected (see Figure 6). Notice that low source resistances, such as presented by short dipole antennas (see equation 26), favor the bipolar design by a wide margin. The FET amplifier's greater variations in frequency response are due to device transconductance (g_m) changes. In the bipolar gain blocks, large amounts of negative feedback substantially reduce these parameter variations. Response variations in the transformer are small if a wideband device is selected. Phase and amplitude matching measurements on the final transformer input amplifier (Figures 7 and 8) show phase and amplitude responses matched to within 0.5 degrees and 0.3 dB, respectively.

Antenna Elements

The choice of the antenna element is governed by both sensitivity requirements and the DF application. For DF purposes, an antenna element must be chosen which is characterized by an omni-directional radiation pattern in the horizontal plane. In addition, a vertical pattern is desired with a main lobe narrow enough to reduce extraneous signals, yet wide enough to receive signals 30 degrees from horizontal. From the sensitivity standpoint, the impedance of the antenna element is critical. The reactance of the antenna, which is very large in electrically small antennas, must be reduced in order to allow better power transfer between the element and amplifier. In addition, the radiation resistance of the antenna should be as large as possible, since a low input resistance is a limiting factor in the system noise performance (4).

An obvious antenna which begins to satisfy the above constraints is a short dipole (or some variation). Clearly, a short dipole will have the necessary radiation pattern; however, the impedance characteristics are less than desirable. For example, at 20 MHz a 3 ft. wire dipole with a radius of 0.2 in. has Z_a = 0.74 – j2164 ohms. Increasing the antenna resistance while maintaining a short length proves to be a very difficult task. It has been shown that an antenna has a minimum Q which is related to its electrical size (5). Thus, there is a physical limit to the maximum resistance for an antenna of a given length. Maximum resistance as well as maximum broadside gain is obtained with a uniform current distribution. A common method for approximating a uniform current distribution is to provide charge storage at the ends of the dipole. This can be accomplished by metal plates or radial wires. This arrangement is referred to as a capacitor-plate or top-hat antenna. If the antenna is small, the current distribution is close to uniform and a fourfold increase in radiation resistance is possible over that achieved with a short dipole (which has a triangular current distribution). In practice, a truly uniform current distribution cannot be achieved, and the advantages offered by this technique seem to be outweighed by the increased size and mechanical complexity.

An alternative technique for manipulation of the current distribution of the antenna element is the use of resistive materials. This technique has been used extensively in previous Watkins-Johnson DF arrays and offers the additional advantage of reducing mutual coupling between array elements while broadbanding the antenna. The basic theory of a resistively loaded antenna is that a resistance along the element is established which approximates the impedance needed to support a pure outward traveling wave (6). The obvious drawback of the resistively loaded dipole is the loss in sensitivity which occurs by adding even small resistances.

Careful investigation of the above

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elements (along with others such as a normal mode helix, folded dipole, etc.) showed that their advantages over a short dipole were minimum for the tradeoffs involved. Thus, a short dipole was chosen as the element for the active antenna. Some improvements were possible, however. Specifically, the diameter of the dipole was increased to reduce the reactive component of the input impedance. This can be illustrated by considering the following approximation:

$$X_a = j240(k_l)^{-1}\,[\ln(l/2a) - 1] \text{ ohms} \quad (26)$$

where $2l$ is the length fo the antenna and a is the radius (7). From equation 26, it can be seen that X_a will decrease with increasing radius. For the final antenna, a radius of 0.625 in. and a length of 36 in. were selected as a compromise between input impedance and physical size. At 20 MHz, this gives:

$$Z_a = 0.74 - j1459 \text{ ohms} \quad (27)$$

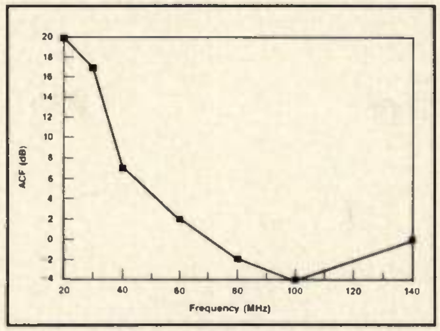

Figure 9. Antenna conversion factor vs. frequency.

Conclusion

Based on the rationale given above, the combination of a fat dipole element and a balun-coupled amplifier was selected. The balun provides the impedance transformation of the transformer in addition to converting from a balanced to an unbalanced system. The increased diameter-to-length ratio of a fat dipole reduces the element reactance (increases the capacitance). The ratio chosen represented a reasonable compromise between increased diameter, constrained by wind loading and manufacturability, and increased capacitance.

Active antenna elements were built and tested. They were then configured as a 20-140 MHz DF array and integrated into the WJ-9881 20-500 MHz DF antenna. Antenna conversion factor was measured using a calibrated half-wave dipole reference. A plot of the resulting antenna conversion factor vs. frequency is given in Figure 9. This figure shows a degradation at the low frequencies due to impedance limitations of the transformer input amplifier. Better sensitivity at these frequencies might have been achieved by using a higher impedance FET amplifier. However, the improved system accuracy and lower noise figure obtained with the transformer input amplifier were judged desirable characteristics at the cost of a slight decrease in sensitivity. The noise floor of the active antenna was not measured directly, due to the fluctuations in external noise that can be expected in any field test situation. DF system sensitivity was, however, measured using this active antenna array. The electric field strength required to reduce line-of-bearing errors to 2 degrees RMS accuracy was experimentally determined by field measurements. This field strength is plotted in Figure 10. These results can be compared to those in Figure 3. While actual performance is slightly worse than that predicted, the basic response is very similar. A peak can be seen in Figure 10 around 100 MHz. This is a result of the very crowded FM band at the test site where the measurements were made.

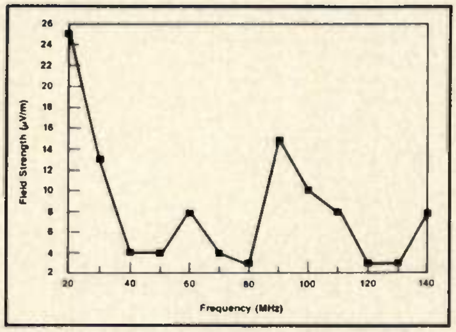

Figure 10. Field strength required for 2 degree RMS error.

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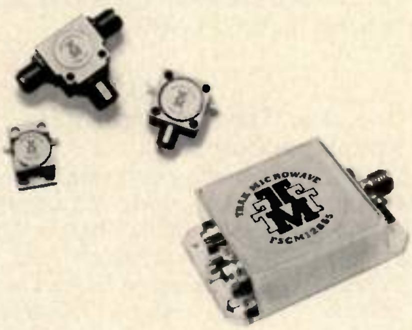

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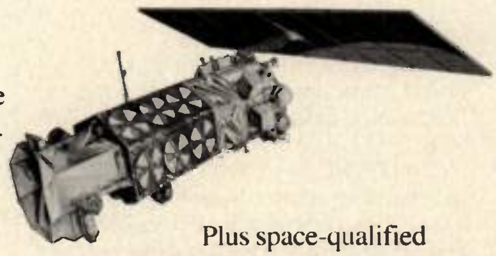

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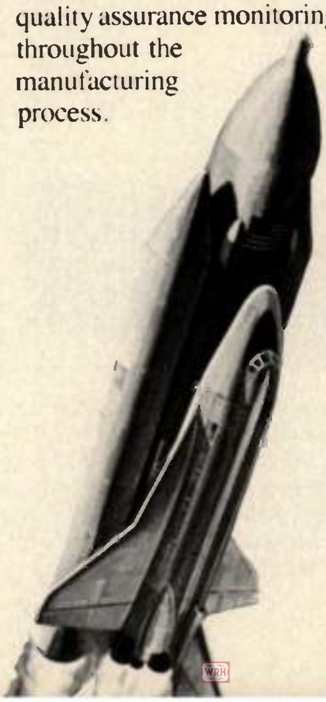

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INFO/CARD 39

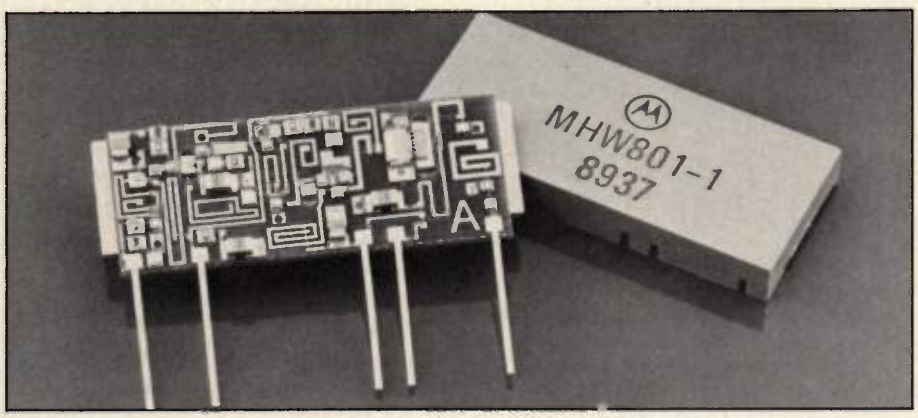
New Products Featured at RF Expo East

Motorola Introduces Modular Amplifiers for Cellular Radio

The MHW 801-1 Series of modules for cellular radio has an output of 1.6 W for a 1 mW RF input from 820 to 850 MHz. The MHW 801-2 has an 870 to 905 MHz frequency range. The modules operate from a 6 VDC power supply.

Motorola also introduces a 20 watt UHF power module (MHW 721A2) that is designed for 450 to 460 MHz cellular radio applications. It operates from 12.5 volts and has minimum gain of 21 dB. Also available is the MRF 650, a 50 watt power transistor for UHF mobile radios. Frequency range is 440 MHz to 512 MHz.

The PAA0200 and PAA1000 Series wideband medium power amplifier family features bandwidths of 1 to 200 MHz and 10 to 1000 MHz with power outputs from 0.4 W to 4 W. **Motorola, Inc., Semiconductor Products Sector, Phoenix, AZ. INFO/CARD #230.**



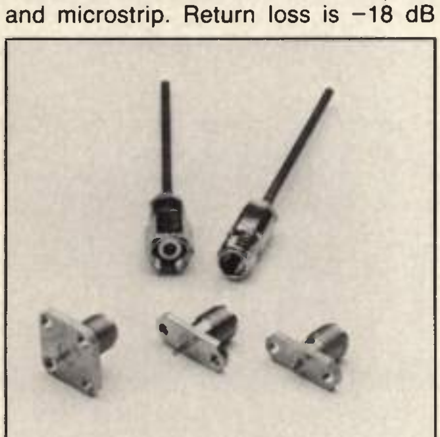
Programmable Transversal Filter
Electronic Decisions

This programmable transversal filter is a software-programmable wideband signal processor based on acoustic

charge transport (ACT) technology which can be used to generate, correlate and filter signals with bandwidths up to 100 MHz. The functions of delay, amplitude weighting, and summation are performed on a single chip, allowing programmable finite-impulse-response transfer functions. This GaAs device features low power. **Electronic Decisions, Inc., Urbana, IL. INFO/CARD #229.**

50 GHz Coaxial Connectors
Amphenol

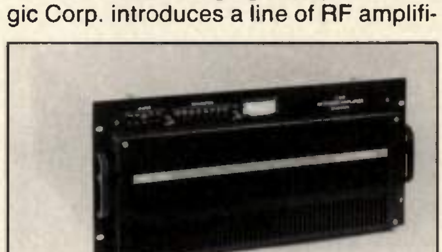
The A-50™ Series is a production-grade 50 GHz connector for use in millimeter wave and related microwave system applications. The connectors include plugs and jacks for 0.086 in., 0.096 in. and 0.141 in. semi-rigid coaxial cable as well as launchers for stripline and microstrip. Return loss is −18 dB



min from DC to 50 GHz. In OEM quantities, the price per mated pair is $84. **Amphenol Corp., Danbury, CT. Please circle INFO/CARD #228.**

RF Amplifiers
Analogic Corp.

The Medical Imaging Group of Analogic Corp. introduces a line of RF amplifi-



ers for pulsed RF from 3 to 90 MHz with peak envelope power up to 25 kW. The hybrid version, AN8036 delivers up to 25 kW at a specified frequency. Models AN8061 and AN8063 deliver 1 kW and 2 kW, respectively. **Analogic Corp., Peabody, MA. INFO/CARD #227.**

Precision Trimmer Capacitors
Voltronics Corp.

This line of precision trimmer capacitors features glass, quartz, air, sapphire and PTFE dielectrics. A high-voltage version with an air dielectric will be introduced. Also being introduced is an expanded line of DRO and microwave cavity tuners, and a new set of tuning tools. **Voltronics Corp., East Hanover, NJ. INFO/CARD #226.**

Surface-Mount Filters
Trilithic

The Z-Series of surface-mount filters covers the 30 to 1500 MHz frequency range. Configurations available include Chebyshev, Butterworth, linear phase

and Bessel. 3 dB percent bandwidths from 2 to 150 percent can be specified with greater than 60 dB stopbands. The package features built-in lugs for enhanced grounding and isolation. **Trilithic, Inc., Indianapolis, IN. Please circle INFO/CARD #225.**

3-Way Power Divider
TRM, Inc.

Model DL 335 is a three-way power divider that operates from 5 to 2000 MHz. Maximum insertion loss is 1.6 dB, minimum isolation is 20 dB, amplitude balance is 0.5 dB max, maximum phase balance is 6 degrees, and maximum output VSWR is 1.6:1. **TRM, Inc., Manchester, NH. INFO/CARD #224.**

Microwave Analysis and Optimization Software
Nedrud Data Systems

DragonWave is a microwave analysis and optimization software tool written for the Macintosh computer. Features include integrated schematic capture and an S-parameter library management system. Reflection coefficient and impedance readings at any point on the Smith Chart are provided by a cursor readout. The element library includes microstrip and stripline discontinuities, coupled lines, ideal transformers and models for ribbon cables and wires. The software is priced at $1380. **Nedrud Data Systems, Las Vegas, NV. Please circle INFO/CARD #223.**

1 GHz TCXO
Vectron Laboratories

The Series CO-256H TCXOs provide stability of $\pm 1 \times 10^{-6}$ from 0 degrees C to +70 degrees C and $\pm 2 \times 10^{-6}$ from −40 degrees C to +75 degrees C. The aging rate is 1×10^{-6}/year. It is available for frequencies from 630 MHz to 1 GHz. A companion series, CO-250, is available from 1 MHz to 630 MHz. **Vectron Laboratories, Inc., Norwalk, CT. Please circle INFO/CARD #222.**

Linear RF and Microwave Simulation Package
Webb Laboratories

Webb Labs unveils SANA, a linear RF and microwave simulation package. It performs signal and noise analysis of nodally defined lumped and distributed networks. The user may define networks in a modular fashion and may optimize circuit performance based on the behavior of defined blocks. The program allows the inclusion of multiple noise-contributing devices. Graphical outputs include S-, Y- and Z-parameters, group delay, noise figure and minimum achievable noise figure. It permits the inclusion of up to 40 network blocks and analyzes and optimizes at up to 512 frequencies at one time. The software is priced at $2950. **Webb Laboratories, Inc., Hartland, WI. INFO/CARD #221.**

SAW Bandpass Filter
Phonon Corp.

The FB70 SAW bandpass filter achieves a bandwidth of 100 kHz and phase linearity of ±3 degrees. Shape

factor is better than 3:1 and maximum insertion loss is 20 dB.

Also from Phonon is a narrowband SAW filter centered at 400 MHz with 1.5 MHz bandwidth. Amplitude tracking is ±1.5 dB and phase tracking is ±5 degrees. The filter is housed in a 6-pin DIP. **Phonon Corp., Simsbury, CT. INFO/CARD #220.**

Spectrum Analyzer Preselector
Advantest America

Advantest unveils the R3551 preselector that assists a spectrum analyzer in making EMI measurements. The unit can measure 100 Hz, 0.044 uVsec and 100 V input signals for conforming to CISPR standards. **Advantest America, Inc., Lincolnshire, IL. Please circle INFO/CARD #219.**

Switch/Limiter/Attenuator
AEL Defense Corp.

The SLA switch/limiter/attenuator features a TTL driver for applications from

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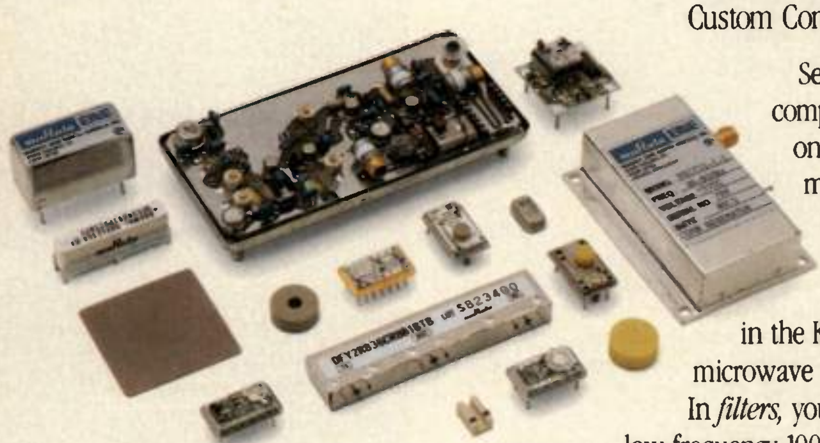

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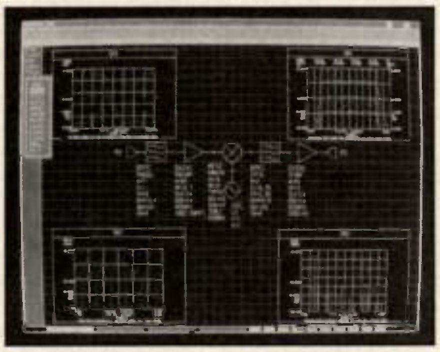
4 to 18 GHz. Insertion loss is 1.5 dB and minimum isolation is 60 dB. Switching speed is less than 15 ns with VSWR of 2.0:1. The assembly handles CW power levels of 1 watt and peak power levels of 10 watts. **AEL Defense Corp., Lansdale, PA. INFO/CARD #218.**

Workstation Simulation Software
EEsof, Inc.

EEsof introduces OmniSys 1.1, a workstation-based microwave system simulation program. It combines the



linear and nonlinear characteristics of interconnected components to form representations of systems and subsystems. **EEsof, Inc., Westlake Village, CA. INFO/CARD #217.**

Multilayer Capacitor
American Technical Ceramics

ATC's STA-TUNE™ is a multilayer capacitor that features the advantages of both trimmers and multilayer capacitors in a 1/8 in. package. Also being introduced is a new tuning stick kit that uses ATC A-case 100 Series Superchips® which measure 0.055 in. square. **American Technical Ceramics Corp., Huntington Station, NY. Please circle INFO/CARD #216.**

Linear Pulsed Amplifier
Erbtec Engineering

This 20 kW, 64 MHz, linear pulsed amplifier is designed for MRI applications. Features include internal diagnostics and digital control. **Erbtec Engineering, Inc., Boulder, CO. Please circle INFO/CARD #215.**

Frequency Sources
RF Monolithics

Model HO1133 is a frequency source that operates at 1000 MHz, while Model HO1135 operates at 500 MHz. Both products use multi-pole quartz SAW resonators for frequency control elements. Voltage tuning is a feature that allows them to be phase-locked to a low frequency master reference. SSB phase noise is −95 dBc/Hz for the HO1133 and −97 dBc/Hz for the HO1135 at 1 kHz offset. The noise floor for both sources is typically −160 dBc/Hz. Price is $550 each. **RF Monolithics, Inc., Dallas, TX. INFO/CARD #214.**

Cellular Radio Design Kit
Signetics

This design kit for cellular radio product developers includes software and layout examples and a working cellular telephone that utilizes the Philips/Signetics cellular chip set. **Signetics, Sunnyvale, CA. INFO/CARD #213.**

SPDT Switch
California Eastern Labs

UPG 107B is an S-band SPDT GaAs FET switch that operates from DC to 3.5 GHz. Typical switching speed is 5 ns.

The NEO2130 is a mini-mold bipolar package that features an f_T of 5.0 GHz.

V_{CE} is 10 volts and I_C is 20 mA. Also available is a 2.5 GHz wideband AGC amplifier. Frequency range is 100 kHz to 2.5 GHz and gain is 20 dB typ. **California Eastern Laboratories, Inc., Santa Clara, CA. INFO/CARD #212.**

Chip Capacitors
Compex Corp.

Compex introduces a line of single-layer parallel-plate microwave chip capacitors in either gold or tin finish. Capacitance values range from 0.1 pF to 10,000 pF. **Compex Corp., Medford, NJ. INFO/CARD #211.**

IF Log Amplifiers
Optimax, Inc.

These high dynamic range logarithmic amplifiers have an operating frequency to 2.5 GHz. **Optimax, Inc., Hatfield, PA. INFO/CARD #210.**

rf expo products *Continued*

Cable Prep Tool
Andrew Corp.

The Easiax cable prep tool is a cutting tool to make improved connector attachments for Heliax[R] cables. **Andrew Corp., Orland Park, IL. Please circle INFO/CARD #209.**

TTL Logic Oscillator
CTS Corp./Knights Div.

MXO-SSL is a TTL logic oscillator that covers from 2.4576 MHz to 70.0 MHz. It features an enable/disable and complementary output option and is packaged in a 14-pin compatible DIP. **CTS Corp., Knights Div., Sandwich, IL, INFO/CARD #208.**

Microwave Training Kit
Lectronic Research Labs

LRL 550B is a microwave student training kit that teaches the theory and application of microwaves. It includes the components and power supply needed to perform the 14 laboratory experiments that are detailed in the manual. The kit costs $2450. **Lectronic Research Labs, Camden, NJ. Please circle INFO/CARD #207.**

Digital Frequency Synthesizer
Infrared Fiber Systems

This digital frequency synthesizer has a 17 ns hop time, −45 dB noise and 16-bit resolution. It supports a pure cosine wave output from 0.005 to 100 MHz with 5 dBm/50 ohm output. It has an optional built-in oscillator and has the capability to accept external oscillators. **Infrared Fiber Systems, Inc., Silver Spring, MD. INFO/CARD #206.**

Distance-to-Fault Analyzer
Marconi Instruments

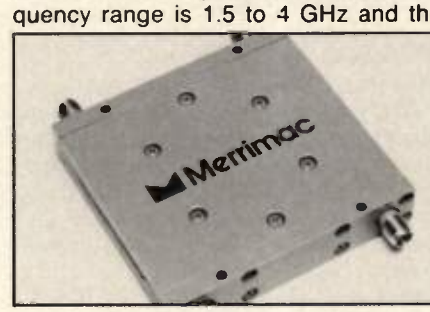
Model 6580 is a distance-to-fault analyzer transmission line measurement system. It consists of two instruments. **Marconi Instruments, Inc., Allendale, NJ. INFO/CARD #205.**

GaAs DDS
Sciteq Electronics

This hybrid GaAs direct-digital synthesizer has a DC to 300 MHz frequency range. It is housed in a package measuring 1 in. X 2 in. Also on display will be a DC to 8 MHz synthesizer/function generator/arbitrary waveform generator designed to work inside a PC. **Sciteq Electronics, San Diego, CA. Please circle INFO/CARD #204.**

I and Q Demodulator Merrimac

Merrimac introduces the IQF-25L Series of I and Q phase detectors. Frequency range is 1.5 to 4 GHz and the device provides 20 degrees of phase trimming. **Merrimac Industries, Inc., West Caldwell, NJ. INFO/CARD #203.**



Microwave Cable Assemblies
UTI/Rosenberger Micro-Coax

A line of coaxial RF connectors together with UTiFLEX flexible microwave cable assemblies will be unveiled. **UTI/ Rosenberger Micro-Coax, Inc., Collegeville, PA. INFO/CARD #202.**

Ultra-Low-Noise Frequency Source
Techtrol Cyclonetics

The NB Alpha 200 is a frequency source with low noise figure. It offers outputs at 10, 100 and 640 MHz. Phase noise at 640 MHz output is −86 dBc at

1 Hz, −143 dBc at 1 kHz and −163 dBc at 10 kHz per Hz.

Techtrol Model R1500 is a tunable receiver that is available for L and S band. The S band offers 0.1 MHz resolution at 200 MHz tuning. **Techtrol Cyclonetics, Inc., New Cumberland, PA. INFO/CARD #201.**

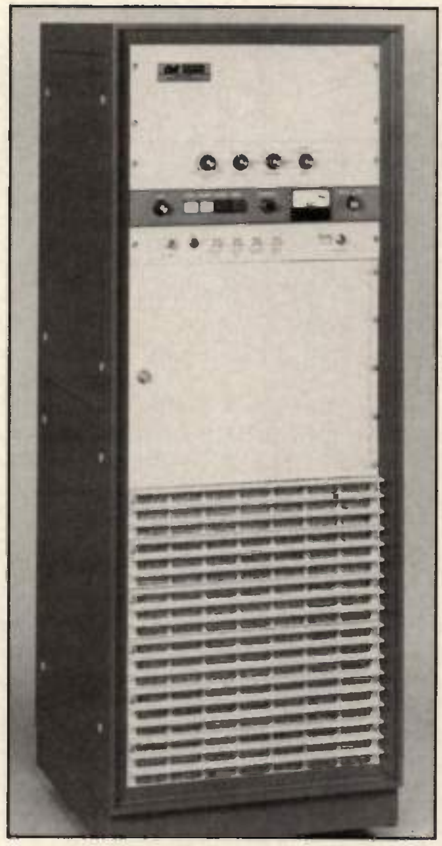
90 Degree Hybrid
Synergy Microwave

Synergy introduces a class of quadrature hybrids which offer 3:1 bandwidths up to 700 MHz. The Series 300 is available in standard flatpack and plug-in packages.

Also being introduced is a line of signal processing modules which include switchable filter networks. In these subassemblies, one filter can be selected at a time from a number of filter types (lowpass, highpass, bandpass). The modules are rated from DC to 1000 MHz and are available in connectorized packages. **Synergy Microwave Corp., Paterson, NJ. INFO/CARD #200.**

Broadband Power Amplifier
Amplifier Research

The Model 300HB delivers a minimum of 300 watts at 55 dB minimum gain from 400 to 1,000 MHz. It uses push-pull circuitry to achieve Class A



operation. A front-panel power meter monitors both forward and reflected power, and an internal RF detector provides an output for use in the self-test or operational mode. **Amplifier Research, Souderton, PA. Please circle INFO/CARD 199.**

Modular Amplifier
Avantek

The UTO/UTC-547 is a thin-film amplifier that is available in a TO-8 package for PC board installation or in an aluminum case with RF connectors. Frequency range is 10 to 500 MHz and two-tone third-order intercept point is +31 dBm. Noise figure is 2.8 dB. **Avantek, Inc., Milpitas, CA. Please circle INFO/CARD #198.**

Frequency Counters
EIP Microwave

EIP introduces VXIbus pulse/CW microwave frequency counters. The EIP 1230A and 1231A are preselected counters that can automatically detect and measure frequencies. The 1230A features a 100 Hz to 26.5 GHz range with optional frequency extension to 170 GHz. The 1231A provides coverage to 20 GHz. **EIP Microwave, Inc., San Jose, CA. INFO/CARD #197.**

High Frequency Op Amp Makes Precise Detector and AGC

By Brian Mathews and Dave Riemer
Harris Semiconductor

The combination of an advanced complementary bipolar process with design techniques that emphasize device matching and minimization of parasitics has resulted in a precision, wideband operational amplifier. This op amp is designed with a combination of speed, precision and low power dissipation, providing system designers added flexibility for achieving their intended design goals. Typical applications for this device are described in this article.

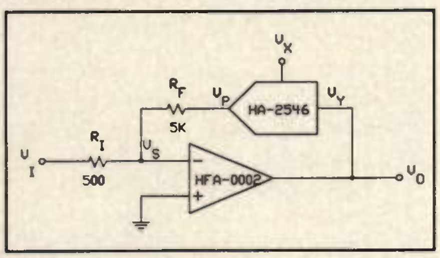


Figure 1. Wideband voltage-controlled amplifier (VCA).

Precision is combined with high speed and gain-of-10 stability in the Harris HFA-0002 operational amplifier. The circuit features a gain bandwidth product of 1 GHz with DC precision previously available only in devices with one-tenth the bandwidth. Typical performance data is described in the appendix.

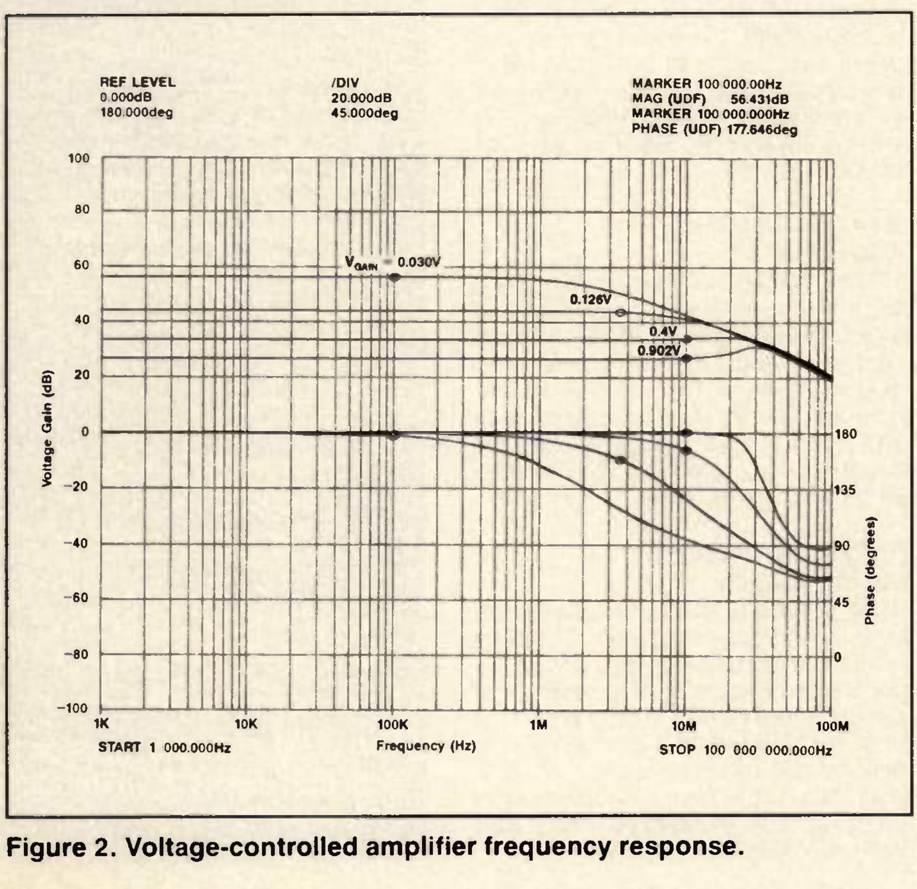


Figure 2. Voltage-controlled amplifier frequency response.

Voltage-Controlled Gain Amplifier (AGC Amplifier)

When an analog multiplier is placed in the feedback loop of an op amp, the gain can be continuously varied using an analog controlling voltage input. This is shown in Figure 1 where the HFA-0002 op amp and a wideband analog multiplier (HA-2546) are used as examples.

In this circuit, V_P is the output of the multiplier and has the following transfer function:

$$V_P = \frac{V_X \cdot V_Y}{2} \qquad (1)$$

Given that $V_Y = V_O$ and $V_S = 0$ (virtual ground), the following derivation provides the overall transfer function:

$$\frac{V_I}{R_I} + \frac{V_P}{R_F} = 0 \quad \text{(Node Equation)} \qquad (2)$$

Substituting:

$$V_P = \frac{V_X \cdot V_Y}{2} \qquad (3)$$

Equation 2 is now:

$$\frac{V_I}{R_I} + \frac{V_X \cdot V_Y}{2R_F} = 0 \qquad (4)$$

Substituting $V_Y = V_O$ yields:

$$\frac{V_I}{R_I} + \frac{V_X \cdot V_O}{2R_F} = 0 \qquad (5)$$

Rearranging terms:

$$\frac{V_O}{V_I} = \frac{-2R_F}{R_I \cdot V_X} \qquad (6)$$

If $R_F/R_I = 10$, then:

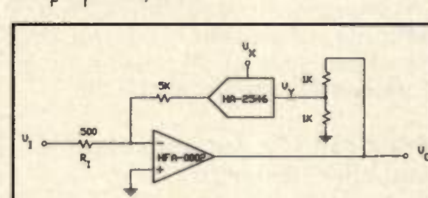


Figure 3. Voltage-controlled amplifier for higher gains.

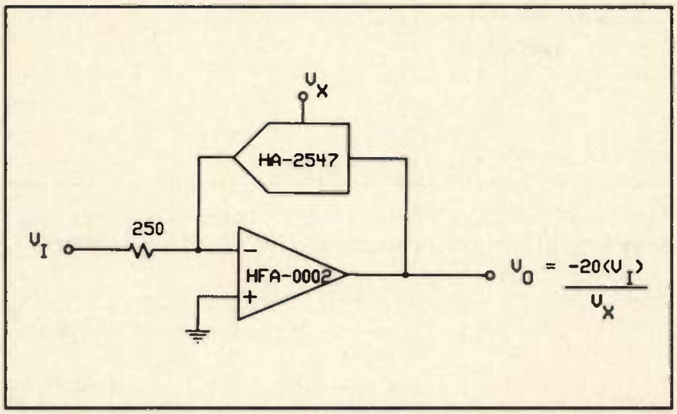


Figure 4. VCA for lower gain and wider bandwidth.

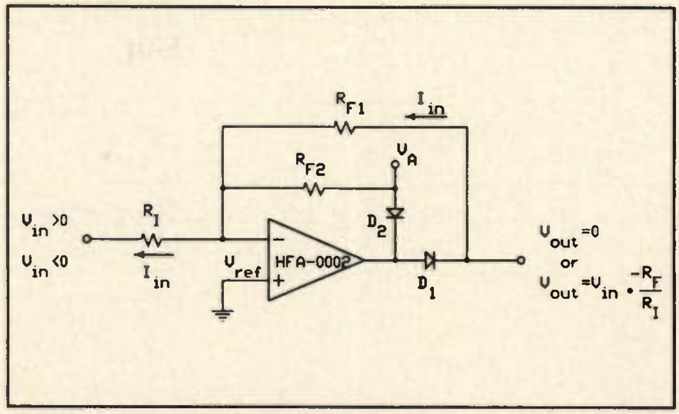


Figure 5. Rectifier schematic.

$$\frac{V_O}{V_I} = \frac{-20}{V_X} \qquad (7)$$

In this configuration, a practical range of V_X has been observed to be between 0.02 V and 1.4 V, yielding a voltage range of 1000 V/V to 14.3 V/V or 60 dB to 23 dB.

As the frequency response plot shows in Figure 2, when V_X is small, the gain is high. The top curve indicates a gain of 56 dB with a bandwidth of greater than 1 MHz. The lines converging on 40 dB (100 V/V) at 10 MHz show that the HFA-0002 has a gain-bandwidth product of (100 V/V)(1E7 Hz) = 1000 MHz or 1 GHz.

While this configuration works well as shown, it suggests several variations. The HFA-0002 has over 100 kV/V low frequency open-loop gain and should work well at gains greater than 1000 V/V. To achieve this, some pre-attenuation should be used in the loop prior to the multiplier. A possible configuration is shown in Figure 3. Nulling of the DC offset may be accomplished using a 10K potentiometer between pins 1 and 8 with the wiper connected to V_{cc}.

Another variation may help with achieving lower gains and more bandwidth. The HFA-0002 is stable at gains of 10 or greater so this AGC circuit should operate at gains down to 20 dB, but it begins to oscillate between 28 dB and 26 dB. At these gains the op amp has about 50 MHz of bandwidth, which is the same as the multiplier's bandwidth. When the op amp has the same or more bandwidth than the multiplier, there is too much phase shift in the feedback path and the circuit is unstable. As long as the gain is higher than 26 dB, the op amp has less bandwidth than the multiplier and there is no instability. To get lower gain and higher bandwidth, a wider bandwidth multiplier such as the HA-2547 could be used. This multiplier has more than 100 MHz bandwidth when driving low impedances. One possible HFA-0002/HA-2547 configuration is shown in Figure 4.

High-Speed Precision Rectifier (Detector)

The precision rectifier configuration shown in Figure 5 is well-known and is frequently used in audio, instrumentation and other low-frequency systems. But, with modern high-speed op amps

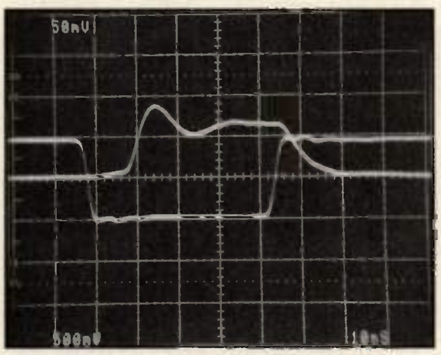


Figure 6. Rectifier transient response.

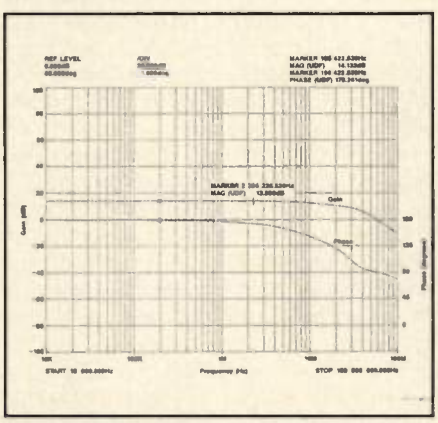


Figure 7. Rectifier frequency response.

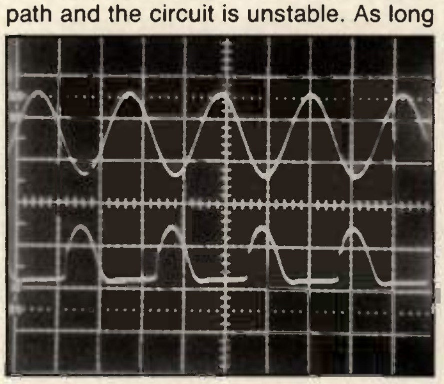


Figure 8. Rectifier input (top; 50 mV/div) and output (bottom; 500 mV/div) for a 2.3 MHz waveform.

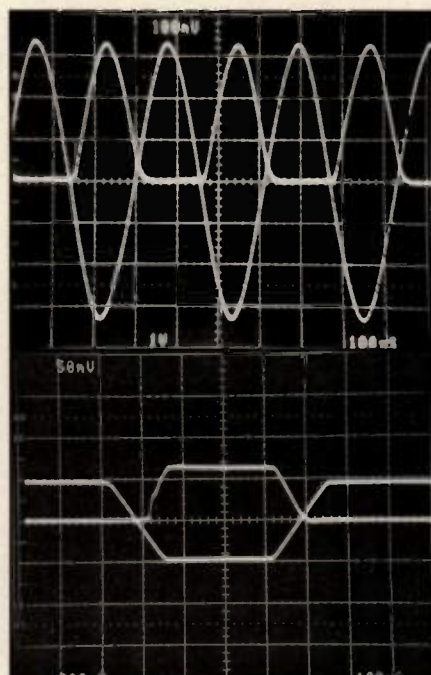


Figure 9 Other rectifier responses.

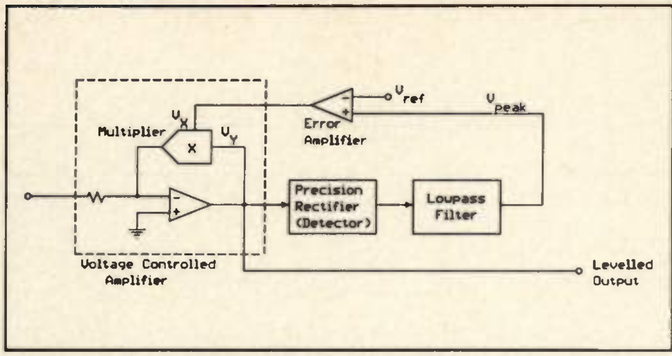


Figure 10. Automatic gain control (AGC) feedback loop.

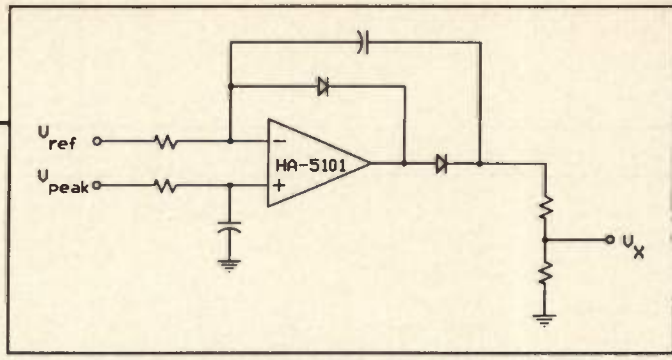


Figure 11. AGC error-amp with lowpass filter, clipping and attenuation to satisfy V_x input limitations.

and proper construction techniques, this circuit can be applied to video, IF and RF signal rectification needs.

The op amp in Figure 5 will respond to maintain the same voltage at both of its inputs (virtual ground). If V_{in} is a lower voltage than V_{ref}, then I_{in} will flow as indicated. Since the input current of the op amp is negligible (1 uA), this current must flow through R_{F1}. This results in $V_O = (I_{in})(R_F)$ and, since $I_{in} = V_{in}/R_{in}$, $V_o = V_{in}(R_F/R_{in})$, the standard op amp transfer function. So, as long as V_{in} is a lower voltage than V_{ref}, the amplifier functions normally and tracks the input signal exactly.

When V_{in} is a higher voltage than V_{ref}, current cannot flow through diode D_1 and so comes through diode D_2. With no current flowing through diode D_1 or R_{F1}, the voltage across R_{F1} must be zero. Thus the output voltage of the circuit is the same as that on the left side of R_{F1}, namely V_{ref}, which would normally be ground. Note than when V_{in} is a higher voltage than V_{ref}, V_A is tracking V_{in} and could be used as an additional output node to capture the opposite polarity portion of the signal.

As the oscilloscope photos in Figure 6 show, this circuit is capable of rectifying very fast waveforms. The 10 nanosecond per division time scale shows that when a +50 mV to −50 mV transition occurs, less than 40 nanoseconds later the output is a stable level near 500 mV. The apparent gain error is due to resistor value error. When the input makes a transition from −50 mV to +50 mV, the output falls to zero within 20 nanoseconds.

An HP 3577 network analyzer sweep, shown in Figure 7, indicates that the 0.1 dB bandwidth of the rectifier is around 2.3 MHz, and the oscilloscope photo in Figure 8 shows the appearance of the waveform at that frequency. Figure 9 shows other transient and CW responses.

This circuit is useful in automatic gain-control loops for level sensing, squelch circuits and sync-stripping in video systems. The diodes chosen should be fast Schottky types in order

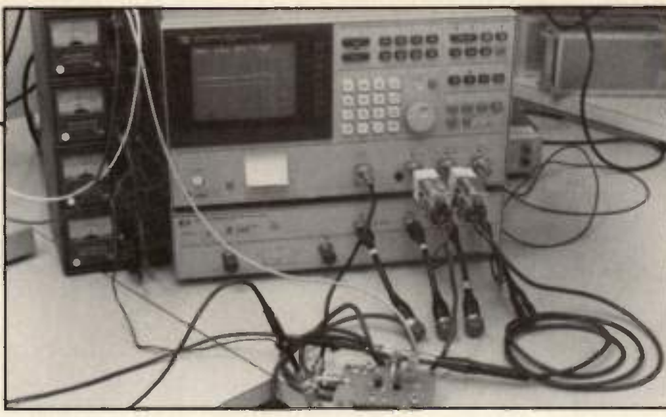


Figure 12. The lab test set-up.

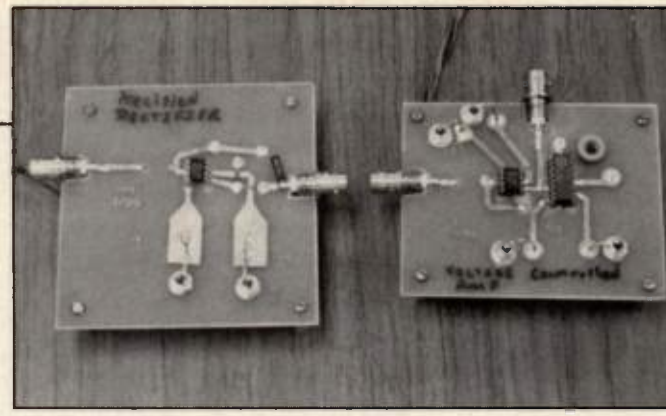


Figure 13. Test circuit PC boards.

to achieve maximum performance.

The voltage-controlled amplifier could be combined with the rectifier and an error amp to form a precision AGC loop. A block diagram is shown in Figure 10. Figure 11 shows a possible configuration for the error amplifier with clip diodes and output attenuator to ensure proper startup conditions and to keep the control signal within the desired range.

Test Set-Up

The lab set-up photo in Figure 12 indicates the type of high-frequency test equipment used to ensure the performance of these circuits. Also, any oscilloscope used to check these devices should have at least 100 MHz bandwidth since the HFA-0002 has about 100 MHz bandwidth in a gain-of-10 configuration. It is recommended that an oscilloscope be used to monitor the circuit when a network analyzer is used since the analyzer will continue to deliver apparently valid results even when the amplifier is oscillating or clipping. The board photos in Figure 13 show that standard high-frequency construction practices have been used. RF feedthroughs are used to get DC power to the board. A pair of chip capacitors (0.01 uF and 0.1 uF) are used for decoupling each IC in the circuit at (or very near) the power supply pins. Gold-plated, low-profile, individual pin sockets are used to minimize socket parasitics. A continuous ground plane covers the bottom of the board. Carbon resistors are recommended for use at high frequencies, as are good quality coaxial connectors and cables. An appendix is on page 81.

About the Authors

Brian Mathews is a lead engineer, market development, for the linear product line at Harris Corporation's Semiconductor Products Division, P.O. Box 883, Melbourne, FL 32901. He can be reached at (407) 729-4644. David Riemer is a lead design engineer at the same division. His number is (407) 729-5332.

KAMAN SiO_2 CABLE

The proven lifeline to EW success

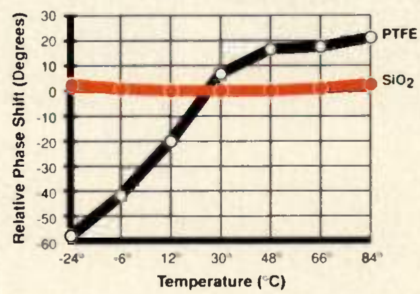
ELECTRICAL STABILITY THAT SETS THE STANDARD

Kaman's SiO_2 is the cable of choice for all EW applications, from the most routine to the most demanding. It is unequaled for superior phase stability over temperature. In fact, at frequencies as high as 18GHz, SiO_2 changes as little as 70ppm over a 75°C shift. Cables with connectors can fully meet specs from −70°C to +371°C; cables alone can operate from −269°C to +1000°C. And unlike PTFE and other dielectrics, SiO_2 is inert, rugged and stable, and is virtually hysteresis-free.

CABLES AS TOUGH AS THE APPLICATION

Our semi-flexible cables feature hermetically-sealed all-welded stainless steel construction. Connectors — SMA, TNC, or Quick Disconnect — are laser welded to withstand five times the pull force of other connectors. In critical airframe applications, SiO_2 cable saves weight and space, and can be easily formed and reformed without compromising performance. Kaman's 0.270" diameter SiO_2 cable has the same insertion loss as a 0.350" PTFE flexible cable. But SiO_2 weighs only two-thirds as much, and bends with a radius of 0.810" (0.405" factory-bent), versus 1.05" for PTFE. And, SiO_2 cable can be bent directly at connector junctions, eliminating angled connectors.

PHASE SHIFT vs. TEMPERATURE
0.141" Dia. Cable @ 5.5 GHz



LOW LIFE-CYCLE COSTS

With an MTBF of 1,000,000 hours — far exceeding any other cable type — Kaman's rugged SiO_2 is truly a cable you can forget about after installation. Furthermore, time and cost of installation are reduced, since SiO_2 resists kinking and overbending; bundles are tighter; rigidity eases complex routing; and pre-formed assemblies reduce routing errors.

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Kaman's SiO_2 cable has proven successful in major programs: Space Shuttle Orbiter, Gallileo, F-14, F-16, Trident I and II, and B-1B. Kaman routinely meets rigorous qualification requirements, including MIL-T-81490 and MIL-C-39012.

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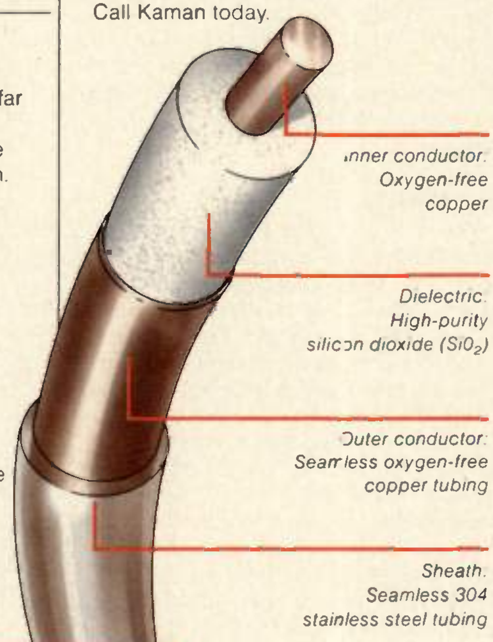
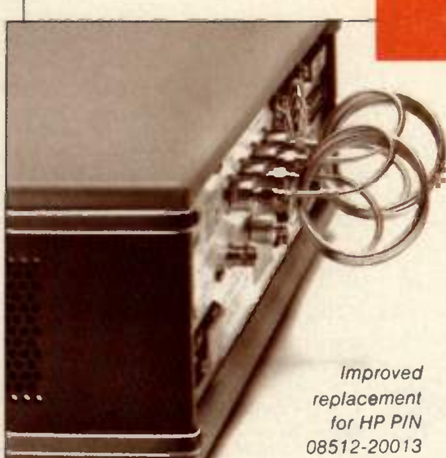
Supplying the best EW cable means more than having the best specs. It means total commitment to your needs — from concept through design and implementation. With timely RFP/RFQ turnarounds, the best technical solutions, an experienced team of engineers and program managers, and on-time delivery of SiO_2 cables custom-configured to your specifications, Kaman responds. Get in touch with your lifeline to EW success. Call Kaman today.



OFF-THE-SHELF SiO_2



Improved replacement for HP P/N 08512-20013

VANA reference lines improve HP 8510 performance nine times

Kaman's SiO_2 Vector Automatic Network Analyzer (VANA) Reference Lines are an improved replacement for cables supplied with the Hewlett-Packard HP 8510. Used in pairs on the back-panel ports, the VANA cables match the phase length to the 8510's front-end test cables. VANA is nine times more stable over temperature than the original PTFE lines, reducing random errors and phase drift, and eliminating frequent calibration. These SiO_2 lines allow accurate measurement over a broader temperature range (+/− 12°F versus +/− 1.8°F), and may eliminate expensive temperature-controlled environments. Two-year warranty standard; lifetime guarantee and quantity discounts available.

Kaman Instrumentation Corporation
Microwave Products Group
P.O. Box 7463, 1500 Garden of the Gods Road
Colorado Springs, CO 80933
719-599-1821 FAX 719-599-1823
Telex: 452412 KAMAN CSP

KAMAN

INFO/CARD 61

The HFA-0002 Operational Amplifier

The Harris HFA-0002 features a gain bandwidth product of 1 GHz. DC performance includes a 1 mV maximum input offset voltage (V_{IO}) with a typical magnitude of 300 uV. Such a low V_{IO} greatly reduces, and often eliminates, the need for additional offset voltage nulling circuitry. Stability and wide bandwidth, combined with a minimum slew rate of 220 V/usec, helps the HFA-0002 achieve a 0.1 percent settling time of 100 nanoseconds. Maximum input offset voltage temperature coefficient (V_{IOTC}) is 0.5 uV/degree C.

The HFA-0002 has an A_{VOL} (open-loop voltage gain) of 100 kV/V and rejection ratios of greater than 100 dB. High gain and rejection ratios were achieved, in part, by the use of two gain stages and high output impedance current sources. The input referred voltage noise is 2.4 nV/Hz.

Two voltage gain stages were used in the implementation of the HFA-0002 to meet stringent DC requirements (Figure 14). Frequency compensation for the HFA-0002 is provided internally. The HFA-0002 is compensated for closed loop gains of 10 or greater. With external compensation, the HFA-0002 can be made stable for gains less than 10. Compensation of each gain stage along with feedforward and Miller compensation maximizes bandwidth while providing a very stable amplifier.

Double-level interconnect allows devices to be placed close together, minimizing the distance signals must travel from device to device. Gold is used for its low inductance and resistance per unit length and its ability to carry high currents without electromigration problems, providing high reliability.

Another feature of the HFA-0002 is its monolithic construction. This allows the small 8-pin standard op amp packages to be used: TO-99 can, ceramic mini-DIP and plastic mini-DIP with SOIC packaging in development. Other benefits of monolithic construction are low power dissipation (less than 200 mW), high reliability and low cost. This op amp offers a combination of features that provide designers with added flexibility in their design.

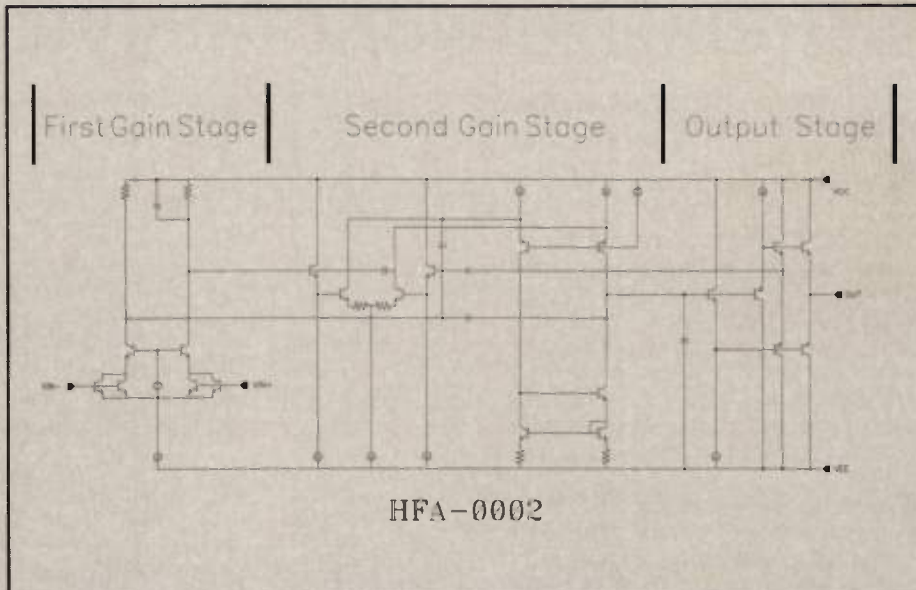


Figure 14. The HFA-0002 schematic.

A Nodal Network Analysis Program

This month's computer program has been contributed by Gary Appel. Named ACANAL (AC ANALysis), this program performs a nodal analysis of an electronic network over a specified frequency range. The output format has been selected primarily for analyzing radio frequency circuits, but lower frequency circuits can be analyzed as well.

This program will accept up to 50 nodes, with the convention that node 0 (zero) is always ground. There is no absolute limit on the number of circuit elements, since memory is allocated dynamically during program execution. Data entry is done in tabular form, defining the element type and value, and the connection nodes. Frequency limits with log or linear sweep, input and output response formats are also part of the entry data.

Circuit Elements

ACANAL contains ten circuit elements in its model library:

Resistor (ideal). The value is entered.

Capacitor (ideal or lossy). The entry includes the nominal capacitance and a value of Q which can be fixed or frequency-dependent.

Inductor (ideal or lossy). Enter the inductance value plus fixed or frequency-dependent Q.

Transconductance (G_m). Specify value and nodes for connection of current sources and sinks, plus positive and negative voltage sense.

Transmission Line (unbalanced). Data entry includes impedance, frequency, length in wavelengths at that frequency, and connection nodes of inner conductor.

Open Stub. Connection nodes, impedance, frequency, and length in wavelengths at that frequency are required input data.

Shorted Stub. Uses same parameters as the open stub.

Transformer. This can represent either a transformer or coupled inductors. The primary and secondary inductances are specified, plus the coupling coefficient and the sense of the particular windings.

Crystal or Ceramic Resonator. Motional capacitance and inductance, mounting capacitance and optional series resistance are defining parameters.

Two-Pole Monolithic Crystal Resonator. Modeled basically as back-to-back crystals with coupling capacitance to ground at the midpoint, this element requires motional capacitance and inductance (assuming identical poles), coupling capacitance, and optional series resistance.

Transistor. Base, collector, and emitter nodes are identified, plus parameters R_{pi}, G_m, C_{pi}, C_{mu}, and optional R_x as shown in the model in Figure 1. Parameters can be chosen to represent NPN, PNP or FET devices.

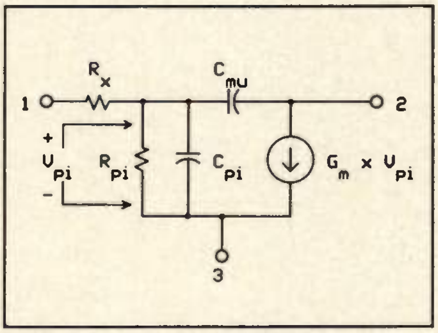


Figure 1. ACANAL transistor model.

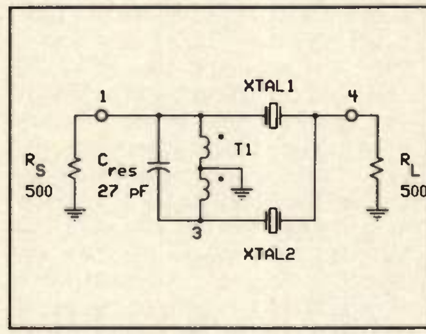


Figure 2. Example filter circuit.

Inputs and Outputs

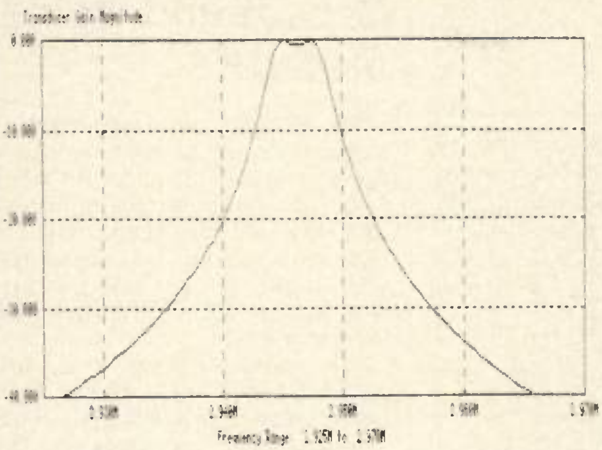
Input and output response format options include: impedance, VSWR, return loss, reflection coefficient (S_{11}/S_{22}), and mismatch loss. The gain response format can be: transducer gain (S_{21}), voltage gain, current gain, transconductance (Y_{21}), or transimpedance (Z_{21}). Source and load impedances, as well as the response formats, may be selected independent of the network description.

The format options determine the output of the analysis, which is performed over the frequency range and number of steps specified. Responses are printed to the screen, with the option of echoing them to the printer. The response may also be plotted on the screen and dumped to a graphics printer. Although the program plotting routine is CGA compatible (an EGA version is also included), inexpensive third-party software will allow the use of Hercules or other display systems. The plot shown in Figure 2 was created in this manner.

Example

A sample calculation, among six that are included with ACANAL, models a two-crystal half-lattice filter. The diagram of the network is shown in Figure 2, with the circuit file and response plot in Figure 3. Other example files include a common emitter amplifier, transmission-line transformer, lowpass filter, ladder crystal filter, and monolithic crystal filter.

This program has been contributed for distribution by the RF Design Software Service by Gary A. Appel, 1318 Old Abbey Place, San Jose, CA 95132. A complete manual is included on the disk, with additional details on program operation. The manual may be printed on any printer by simply using the MS-DOS Print command. See the announcement on page 102 for ordering information.




```
RS =  500.00  1
RL =  500.00  4

1  CAP     Cres  1  3          27.0000p
2  XFRMR    T1   1  0  0  3    27.0000u  27.0000u  0.9990
3  XTAL   XTAL1  1  4          54.0000f  54.0000m  29.0000p  0.0
4  XTAL   XTAL2  3  4          54.1200f  54.0000m  29.0000p  0.0

1  LIN   2.9250M   2.9700M   250.0000
```

Figure 3. Circuit element list and response plot of example.

CALL FOR PAPERS

You are invited to present a paper at the technical program at RF Technology Expo 90, to be held March 27-29, 1990 in Anaheim, Calif. Proposals for individual papers or complete three-hour sessions are welcome. Topics of current interest to RF engineers include:

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Send an outline or abstract of the proposed paper to:

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RF Design Magazine
6300 S. Syracuse Way - Suite 650
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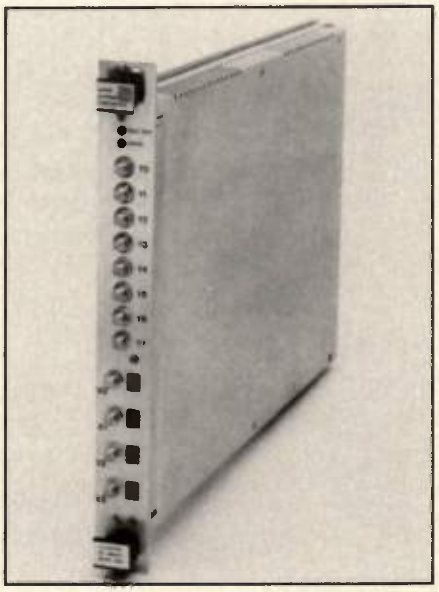
INFO/CARD 63

Matrix Introduces a VXIbus Switching Matrix

Model 10081 is a VXIbus (Revision 1.3) compatible coaxial 4 X 8 shielded message-based, C-sized switching module. The frequency range is DC to 625 MHz. It is designed for both military and commercial automatic test equipment (ATE) systems where instrumentation such as frequency counters, signal generators, oscilloscopes, pulse generators, RF analyzers and other devices must be connected using the VXIbus specification.

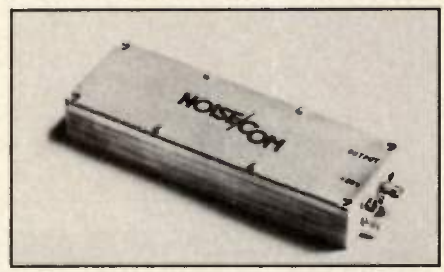
It is available in 50 and 75 ohm configurations and features 12 front panel SMA signal connectors to provide access to the bi-directional DC to 625 MHz blocking matrix at a switching speed up to 1 ms per crosspoint. The module has a seven-segment display next to each of the four X-axis signal connectors to provide front panel indication of any crosspoint connections to a Y-axis signal connector. Interlocking on the Y-axis signal prevents the user from connecting any two Y-axis signal connectors together. Remote crosspoint verification is also included. When purchased in single quantity, the switching module is priced at $3,000. **Matrix Systems Corp., Calabasas, CA. Please circle INFO/CARD #196.**



Broadband Noise Module

The NC1112A is an amplified noise module that produces white noise in a Gaussian distribution from 20 MHz to 2 GHz, with power output of −10 dBm ±2.5 dB (−106 dBm/Hz). It has a minimum crest factor of 5:1, operates from −35 degrees C to +100 degrees C, requires 28 VDC and utilizes female SMA connectors.

The module is based on a hermetically packaged noise diode that has been burned-in for 168 hours. A lowpass filter attenuates noise above the output frequency band. Options include operation from 15 VDC or 24 VDC, and other types of RF connectors. **Noise Com, Inc., Paramus, NJ. INFO/CARD #194.**



Vector Network Analyzer Upgrade From LMT

This product from Linear Measurement Technology Corp., is designed to provide a digital interface for the HP 8410. With this interface, extended processing capabilities can be achieved to approach that of the newer HP 8510.

The interface controller and software (Model IC-5500) mates a standard HP 8410 vector network analyzer system to an Apple Macintosh Plus, SE or II computer for computer control and full term error correction. The system swept source is controlled via an IEEE-488 bus.

The program provides simultaneous, corrected or uncorrected Smith, polar, rectangular and tabular displays of S-parameter data. All data can be stored and exported to word processing programs as charts or as file data.

Model IC-5500 functions over the 0.11 to 18 GHz frequency range of the 8410 and will accommodate waveguide, coaxial, microstrip and stripline transmission line measurements. **Linear Measurement Technology Corp., Lafayette, CO. INFO/CARD #195.**

Phase Noise Detectors

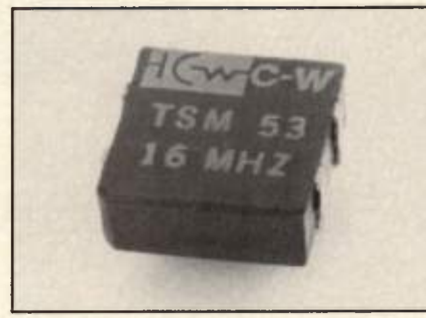
Model FSS600 and FSS1000 phase noise detectors are designed to measure the phase noise in oscillators, amplifiers, passive components, frequency multipliers, and dividers for Fourier frequencies from approximately DC to 400 kHz from the carrier. Features include a built-in monitor, adjustable phase delay, and low noise floors. Input frequency ranges are 1 MHz to 600 MHz for the FSS600 and 5 MHz to 1 GHz for the FSS1000. At 400 kHz, the noise floor is −177 dBc. **Femtosecond Systems, Boulder, CO. INFO/CARD #193.**

EMI Suppressant Tubing Kit

A kit containing 15 one-foot lengths of electromagnetic interference suppressant tubing, with inside diameters ranging from 0.0500 to 0.500 in., in increments of 0.020 to 0.025 inch, is available from Capcon. This selection of flexible and lossy absorptive tubing for RFI/EMI suppression is priced at $35 per kit. **Capcon, Inc., New York, NY. Please circle INFO/CARD #192.**

Crystal Clock Oscillator

Designated the TSM53, these crystal clock oscillators cover from 250 kHz to 24 MHz. The output waveform is a square wave at 5 volts rail-to-rail and is TTL/HCMOS compatible. Standard frequency tolerance is ±0.01 percent. The price ranges from $50.00 each in 5-



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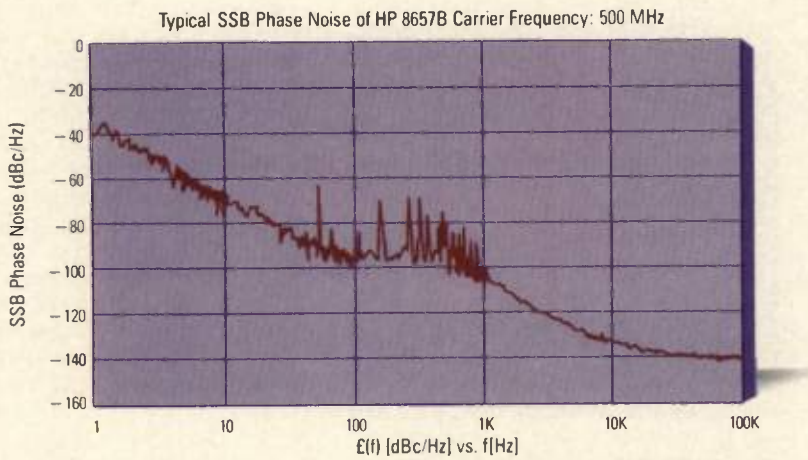

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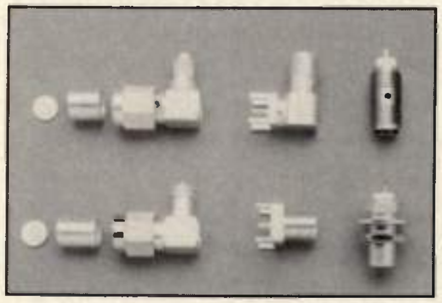
piece quantity to $3.95 each in 1000-piece quantity. **Connor-Winfield Corp., Aurora, IL. INFO/CARD #191.**

SMA Connectors

Cambridge Products introduces an array of SMA connectors. Included in the product line are plugs and receptacles in straight and right-angle crimp-type connectors for flexible cable (in-



cluding double shielded), bulkhead, and flange-mount plug and jack connectors. Frequency range is DC to 18 GHz and characteristic impedance is 50 ohms. **Cambridge Products Corp., Bloomfield, CT. INFO/CARD #190.**

Proportionally Controlled Heater

The DN-501 is a miniature proportionally controlled heater whose temperature can be programmed with an external resistor. It is suited for regulating the temperature of sensitive electronic components such as microwave filters and oscillators. It is packaged hermetically and can supply up to 28 watts of power from an unregulated supply. **Dawn Electronics, Inc., Carson City, NV. Please circle INFO/CARD #189.**

RF Pulse Amplifier

The 3200 Series amplifiers cover from 6 to 220 MHz and are available in 300 and 1000 watt models. Standard features include pulse droop of under 5 percent, blanking time of under 2 us and linearity of ±1 dB. **American Microwave Technology, Inc., Fullerton, CA. INFO/CARD #188.**

PTFE Coaxial Cables

Times Microwave Systems introduces small core PTFE coaxial cables which can be used over a broad frequency range. The M17/60-RG142 is designed to be used in applications where M17/158-00001 or RG 142B/U can be used. **Times Microwave Systems, Wallingford, CT. INFO/CARD #187.**

Unity Gain Filter Module

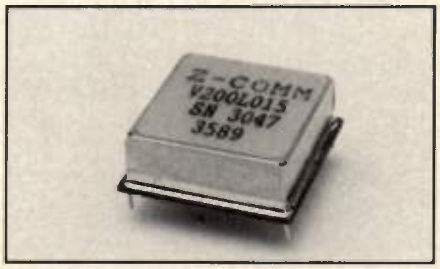
This product is designed as a testing tool for the commercial/military customer. The device, which accommodates Sawtek's 70 MHz filters, allows the customer to evaluate the filters in his system. The products feature 3 dB bandwidths from 0.25 to 40 MHz. **Sawtek, Inc., Orlando, FL. Please circle INFO/CARD #186.**

Low-Frequency Coupler

These dual directional couplers have a frequency range from 20 MHz through 1450 MHz. Continuous wave power ratings are from 500 watts to 2000 watts. The couplers use SC or N type connectors on the through line and SMA type connectors on the sampled ports. **Amarin Company, Inc., Manchester, NH. INFO/CARD #185.**

Voltage-Controlled Oscillators

Model L-200 features a tuning range of 200-300 MHz with close-in phase noise characteristics of −95 dBc at 1 kHz and −105 dBc at 10 kHz. The L-200 has an output of 10 dBm ±2 dBm into 50 ohms. Frequency pulling with a 1.3:1 VSWR is less than 5 MHz and pushing is less than 1 MHz/volt. In quantities of 1 to 9, the price is $35 each. Other frequencies are available. **Z-Communications, Inc., Ft. Lauderdale, FL. Please circle INFO/CARD #184.**



Surface Mount Amplifiers

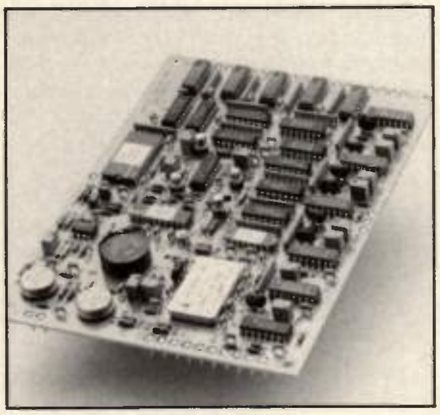
Model PA397M is an amplifier with a 100 to 3000 MHz range. It is housed in a surface mount package. Gain is 34 dB, noise figure is 4 dB and output power is +10 dBm. **Phoenix Microwave Corp., Telford, PA. INFO/CARD #183.**

RF Power Sensors

Bird unveils the Thrulne[R] Model 4024 directional power sensor that is designed to measure up to 10 kW power from 1.5 to 32 MHz at ±3 percent of reading accuracy. It carries its own calibration profile and corrects readings automatically. **Bird Electronic Corp., Cleveland, OH. INFO/CARD #182.**

10-Bit 40 MHz A/D Converter

The ASA-1040 is an A/D converter with an input bandwidth of 100 MHz. S/N and noise power ratios are 58 and 48 dB, respectively. The 10-bit parallel output is provided by using digitally corrected subranging. The 40 million samples-per-second encode rate is achieved through the use of a T/H amplifier, a subranging architecture and digital correction circuitry. **Addacon, Inc., Greensboro, NC. INFO/CARD #181.**



High Dynamic Range Mixer

The UNCL-R1H mixer has a 10 to 500 MHz range with noise figure of 4.5 dB and gain of 6 dB. A built-in 50 ohm impedance matching amplifier is included. LO-RF isolation is 55 dB, LO-IF isolation is 35 dB and the 1 dB compression point is at +1 dBm. In quantities of 1 to 9, the price is $27.95. **Mini-Circuits Laboratory, Brooklyn, NY. Please circle INFO/CARD #180.**

Surface Mount Inductor

RL2525 is a surface mount chip inductor with an inductance range of 0.22 uH to 220 uH. These products are available in tape and reel for automated assembly. Price is $0.50 in quantities of 2500. **Renco Electronics, Inc., Deer Park, NY. INFO/CARD #179.**

RF Coaxial Switch

Loral Microwave-Wavecom unveils the Model 8027, a latching SPDT RF coaxial switch that handles up to 1 kW CW and 10 kW peak input power from DC to 1 GHz. VSWR is 1.2:1, insertion loss is 0.2 dB and minimum isolation is

rf products Continued

80 dB. Switching time is less than 50 ms. **Loral Microwave-Wavecom, Northridge, CA. INFO/CARD #178.**

Surface Mount Crystal

The SX2050P surface mount crystal is an AT-cut quartz unit packaged in a high temperature epoxy package. Seventeen frequencies from 4.0 to 24.0 MHz are available. It is available on EIA RS481 24 mm tape and reel. **M-tron Industries, Inc., Yankton, SD. Please circle INFO/CARD #177.**

RF Amplifier

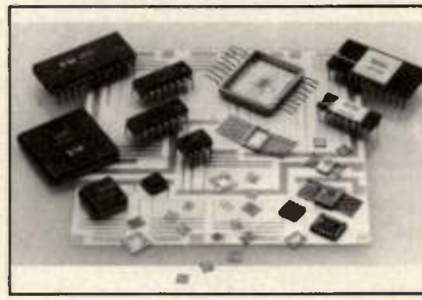
Model RC113-2 provides a minimum of 2 watts of linear RF power from 15 Hz to 100 MHz. Power gain is 40 dB and harmonic rejection is −23 dBc. Typical applications include medical and ultrasonics. **Wessex Electronics Ltd., Downend, Bristol, England. Please circle INFO/CARD #176.**

Hi-Rel Inductors

Delevan introduces a line of axial leaded inductors. Series 1537 are unshielded coils with an inductance range of 0.15 to 33 uH. Shielded components (Model 1641) are available from 0.10 to 12 uH. **Delevan Division, American Precision Industries, East Aurora, NY. INFO/CARD #175.**

Hybrid/Microwave Chip Resistors

Solitron Devices unveils a line of precision thin-film microwave chip resistors having a resistance range from 5 to 500 ohms. Depending on the applica-



tion, the resistors can operate up to 18 GHz with power ratings up to 0.15 W. **Solitron Devices, Inc., Precision Resistor Group, Riviera Beach, FL. Please circle INFO/CARD #174.**

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INFO/CARD 79

INFO/CARD 80

Quick-Disconnect Connectors

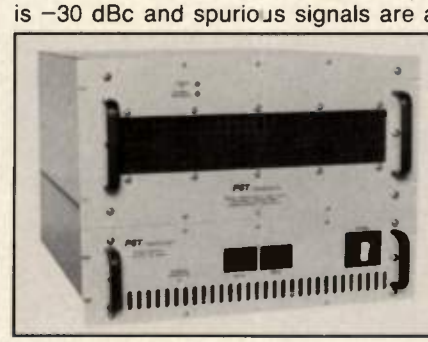
Kaman introduces the availability of its SiO_2 cable assemblies with quick-disconnect (QD) connectors. It features a spring-loaded, positive locking connector system. The connector is rated up to 18 GHz and is hermetically sealed. **Kaman Instrumentation Corp., Colorado Springs, CO. INFO/CARD #173.**

RF Switching Unit

Model ACS20500-8 is a 20 to 500 MHz, eight input, three output RF matrix. Any input may be switched to one or more outputs. Digital switching commands are by RS232 or IEEE interfaces. The third-order intercept point, referenced to the input, is +24 dBm. **AML, Inc., Camarillo, CA. INFO/CARD #172.**

1 kW Pulse Amplifier

PST introduces the Model BHPC 787967-1000/1225 pulse amplifier with a 780 to 960 MHz frequency range. Power output is 1000 watts min and pulse rise/fall time is under 50 nsec. Input VSWR is 1.5:1, harmonic output is −30 dBc and spurious signals are at



−60 dBc. The unit features type N female connectors. **Power Systems Technology, Inc., Hauppague, NY. Please circle INFO/CARD #171.**

TCAS Switch

The Model D2-729B001 is a switch designed for use in traffic alert and collision avoidance systems (TCAS). Frequency range is DC to 2 GHz, VSWR is 1.3:1 max, and insertion loss is 0.25 dB max. Minimum isolation is 60 dB. **Dynatech Microwave Technology, Inc., Calabasas, CA. Please circle INFO/CARD #170.**

Portable Radio Modem

Monicor introduces the IC-15ME portable radio modem which interfaces to a computer via an RS-232. Up to 27 of these devices within a mile radius can be polled by the IC-2·0A base network modem on a single simplex 12.5 kHz,

450 to 470 MHz radio channel. CRC validated variable length packets may be transferred over the radio at 2400 or 4800 bits/sec. Rates from 50 to 9600 baud and other parameters may be set through the serial port. The portable unit is priced at $1195 and the base unit is priced at $1495. **Monicor Electronic Corp., Fort Lauderdale, FL. Please circle INFO/CARD #169.**

Crystal Oscillator

This oscillator is designed for applications from 8 to 12 MHz. Typical SSB phase noise characteristics are −113 dBC/Hz at 10 kHz, −133 dBc/Hz at 100 kHz, and −143 dBc/Hz at 10 kHz. The aging rate is 7 x 10^{-10}/day. Typical vibration sensitivity is measured at 3 X 10^{-9}/g. Frequency stability is ±5 X 10^{-9} over a 0 to +60 degrees C temperature range. Price ranges from $300 to $400 in small quantities for the 2870060. **Piezo Crystal Company, Carlisle, PA. Please circle INFO/CARD #168.**

4 GHz Signal Generator

The Rohde & Schwarz Model SMHU is a 100 kHz to 4.32 GHz signal generator. Phase noise at 100 MHz is −144 dBc/Hz for a 20 kHz offset and residual FM at 500 MHz is less than 0.5 Hz. Output power is +19 dBm. AM is

available with less than 2 percent distortion, FM up to 3.2 MHz deviation with 0.2 percent distortion and PM to 320 rad with 0.5 percent distortion. The instrument costs $36,700. **Rohde & Schwarz, Lanham, MD. INFO/CARD #167.**

rf software

Waveform Software

Analogic Corp. introduces DPCOM™, a set of software utilities which gives IBM PC and compatible users the ability to control, store and manipulate waveform data from Analogic/Data Precision's line of waveform synthesizers and generators. With the software and IEEE-488 interface, users can send commands, load and save polynomials, and upload and download waveform data points from the Analogic 2020-25, 2020-100, 2040 and 2045 waveform synthesizers. The software costs $150 when purchased separately and $100 when purchased with an Analogic waveform synthesizer/generator. **Analogic Corp., Peabody, MA. INFO/CARD #163.**

Program Runs HP BASIC on IBM PCs

HTBasic is a software product that emulates the HP 9000 Series 200/300 workstation BASIC. It runs on MS-DOS based computers, HP VECTRA, IBM AT or IBM PS/2. The software provides a method of developing and running HP BASIC programs on a PC. Utility programs that transfer data files and ASCII programs between HP LIF format diskettes and MS-DOS disks are included. The package is priced at $500. **TransEra Corp., Provo, UT. Please circle INFO/CARD #162.**

RF/Microwave Circuit Design Utilities

SData + is a scattering parameter and noise data enhancement utility. It reads S-parameter and noise data DOS library files from bipolar, FET and IC device manufacturers. The program also performs stability and gain calculations and will generate tubular device impedance data, noise match data or circuit files for use by other Microwave Software programs. The program costs $195. **Microwave Software, Laguna Hills, CA. INFO/CARD #161.**

Continuous-Measurement Software

HP introduces a software product for use with the HP 5371A frequency and time-interval analyzer and the HP 9000 Series 200/300 or Vectra computer. The software quadruples the number of continuous measurements available with the instrument. In addition, multiple measurements can be indefinitely taken and be precisely related to one another in time. The software also allows access to the unprocessed measurement data from the 5371A. Price is $1200. **Hewlett-Packard Company, Palo Alto, CA. INFO/CARD #160.**

Amplifier Simulation Program

A.S.P. Version 1.0 provides the user with menu-driven options for designing solid state amplifiers. Included are automatic or user-interactive design routines, matching circuits, a help menu, and a utility to optimize noise figure. A sample library, documentation and theory files are also included. The software runs on IBM compatible computers and requires at least 360K of memory. CGA and math co-processor is recommended but not required. The amplifier simulation program is priced at $54.95 plus $2.50 shipping and handling. **SW.I.F.T. Enterprises, Hoffman Estates, IL. INFO/CARD #159.**

rf literature

Component Selection Diskette

Analog Devices introduces a component selection guide in the form of a diskette. It contains information on analog and digital signal processing components and data converters. The user basically enters the minimum and maximum desired specifications for key parameters and the software searches its data base and displays up to 17 components meeting the criteria together with actual specifications. The products are listed in order of ascending price. The user can also enter any part number to see the key specifications. The data base contains over 1600 parts including analog-to-digital converters, op amps, voltage references, instrumentation amplifiers, isolation amplifiers, and multipliers/dividers. **Analog Devices, Norwood, MA. INFO/CARD #158.**

RF Mixer Catalog

This catalog describes mixers that operate from 10 kHz to 14 GHz. A special section defines mixer terms and discusses performance aspects. A selection chart is also provided. Most of the mixers listed are available in high-reliability versions for military applications. **Lorch Electronics Corp., Englewood, NJ. INFO/CARD #157.**

RF Switch Product Guide

Dynatech introduces an eight-page product guide which contains information on microwave coaxial switching products. Specifications listed include operating frequency, VSWR and switching speed. Graphs, illustrations and photographs are used in the descriptions. Design considerations together with general technical data are provided. **Dynatech Microwave Technology, Inc., Calabasas, CA. INFO/CARD #156.**

Coaxial Components Catalog

HP's coaxial components catalog provides specifications as well as operating and installation information on switches, detectors and fixed and step attenuators. Products highlighted include the HP 33314 SPDT coaxial switch and families of step attenuators and coaxial detectors. **Hewlett-Packard Company, Palo Alto, CA. INFO/CARD #155.**

Literature Selection Guide

Abstract of Application Notes and Selection Guides describes 17 articles and selection guides available from the M/A-COM Semiconductor Products Division. The publications are available upon request. The titles listed include PIN Diode Selection Guide, Multiplier Selection Guide, Applications of Multiplier Diodes, Designing with PIN Diodes, Principles, Applications, and Selection of Receiving Diodes, Distortion in PIN Diode Switches, and Comparison of Gallium Arsenide and Silicon PIN Diodes for High Speed Microwave Switch Circuits. **M/A-COM Semiconductor Products Div., Burlington, MA. INFO/CARD #154.**

DTO and Synthesizer Catalog

NCI, a division of Noise Com, Inc. has released a catalog that features a line of DTO and synthesizer products. Specifications listed include those for the ULDTO Series closed-loop oscillators for use as modulators or sources in phase-locked loops (PLLs), the OLDTO Series of open-loop oscillators for use in PLLs or as coarse-tuning LOs, the FSDTO Series of fast set-on oscillators for applications requiring rapid acquisition of unstable signals, and the FSS Series of custom frequency synthesizer subsystems. **NCI, Paramus, NJ. Please circle INFO/CARD #153.**

RF Connector Catalog

This RF connector catalog contains a selection guide, a military cross-reference index, a part number index, a cable-to-connector reference table and segments on cable retention methods and plating. Connectors listed include BNC, TNC, C, SC, N, HN, UHF, TWIN, SHV, MHV, Triax, TRB, TRT, SMA, K-LOC, terminations, adapters, and crimp tools and kits. **King Electronics Co., Inc., Tuckahoe, NY. INFO/CARD #152.**

Switched Multiplexer and Channelizer Data Sheet

Information on switched multiplexers which are used as signal thinners at IF stages are discussed in this data sheet. These devices operate from 3 to 5 GHz and allow the sorting of signals before they enter a discriminator frequency measurement subsystem. Also covered are one-input, N-output channelizers that divide the input frequency band into a relatively large number of channels (from 6 to 64). **K & L Microwave, Inc., Salisbury, MD. INFO/CARD #151.**

Telemetry Components Catalog

This catalog from AML presents a range of telemetry components. Products described include polarizers, vectorial comparators, scan converters, filters and low-noise amplifiers. Specifications together with outline drawings and schematics are provided. **AML, Inc., Camarillo, CA. INFO/CARD #150.**

Rental Catalog

A 1989-1990 rental catalog featuring products such as analyzers, meters, generators, oscilloscopes, desktop computers, and telecommunications is available from Genstar Rental Electronics. The manufacturers represented include Hewlett-Packard, Tektronix, Intel and Fluke. **Genstar Rental Electronics, Inc., Woodland Hills, CA. INFO/CARD #149.**

VXIbus Literature

A booklet of application notes for the VXIbus is available from Racal-Dana Instruments. It discusses the scope, purpose, benefits and an overview of the standard. Information presented includes integrating an electronic test system with VXIbus instruments, ATE improvements with VXI, enhanced pulse measurement with generation using VXIbus-based instruments, and the benefits of VXIbus architecture for RF applications. **Racal-Dana Instruments, Inc., Irvine, CA. INFO/CARD #148.**

Filter Catalog

This catalog features microwave tubular lowpass, bandpass and highpass filters from Micro-Coax. Product information and general technical information on filters and impedance transformers is included. A connector option guide is also provided. Graphs provide response, frequency, insertion loss and rejection characteristics. **Micro-Coax, Inc., Collegeville, PA. INFO/CARD #147.**

RF and Microwave Products Catalog

Information about Triax's thick-film digital and analog components designed and manufactured by Optimax is featured. Products described include cascadable amplifiers, RF and microwave cascaded assemblies, GaAs FET amplifiers, switches and step attenuators, integrated subassemblies, custom microcircuits, log amplifiers and constant phase limiting amplifiers. A section on packaging capabilities is included. **Optimax, Inc., Hatfield, PA. INFO/CARD #146.**

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This month's disk (RFD-1189)
ACANAL, AC nodal-analysis program by Gary Appel, described in the article "Nodal Network Analysis Program." Information provided on the disk and the figures contained in the article provide a complete "instruction manual" for the program.

***Coming soon:* Disk RFD-DR89: Annual Directory Issue**
A set of excellent filter design programs by David Greene of NAP Consumer Electronics. The programs assist in filter order selection, then compute component values from filter specifications.
1. "Chebyshev Lowpass Filter Design"
2. "Elliptic Lowpass Filter Design"
3. "A Design Program for Butterworth Lowpass Filters," from March 1989 issue.

Disk RFD-1089: October 1989
"A Spurious Response Program for Wideband Mixer Circuits," by William Sabin of Rockwell-Collins. Computes and plots intermodulation products for wide bandwidth frequency conversions.

(All disks to date are MS-DOS compatible)
Programs have been available on disk since February 1989.
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Disks are $9.00 each (5¼ in.) or $10.00 (3½ in.). Outside U.S. and Canada, add $8.00 per order. Foreign checks must be in U.S. funds, drawn on banks with U.S. offices or agents. Prices include postage and handling.

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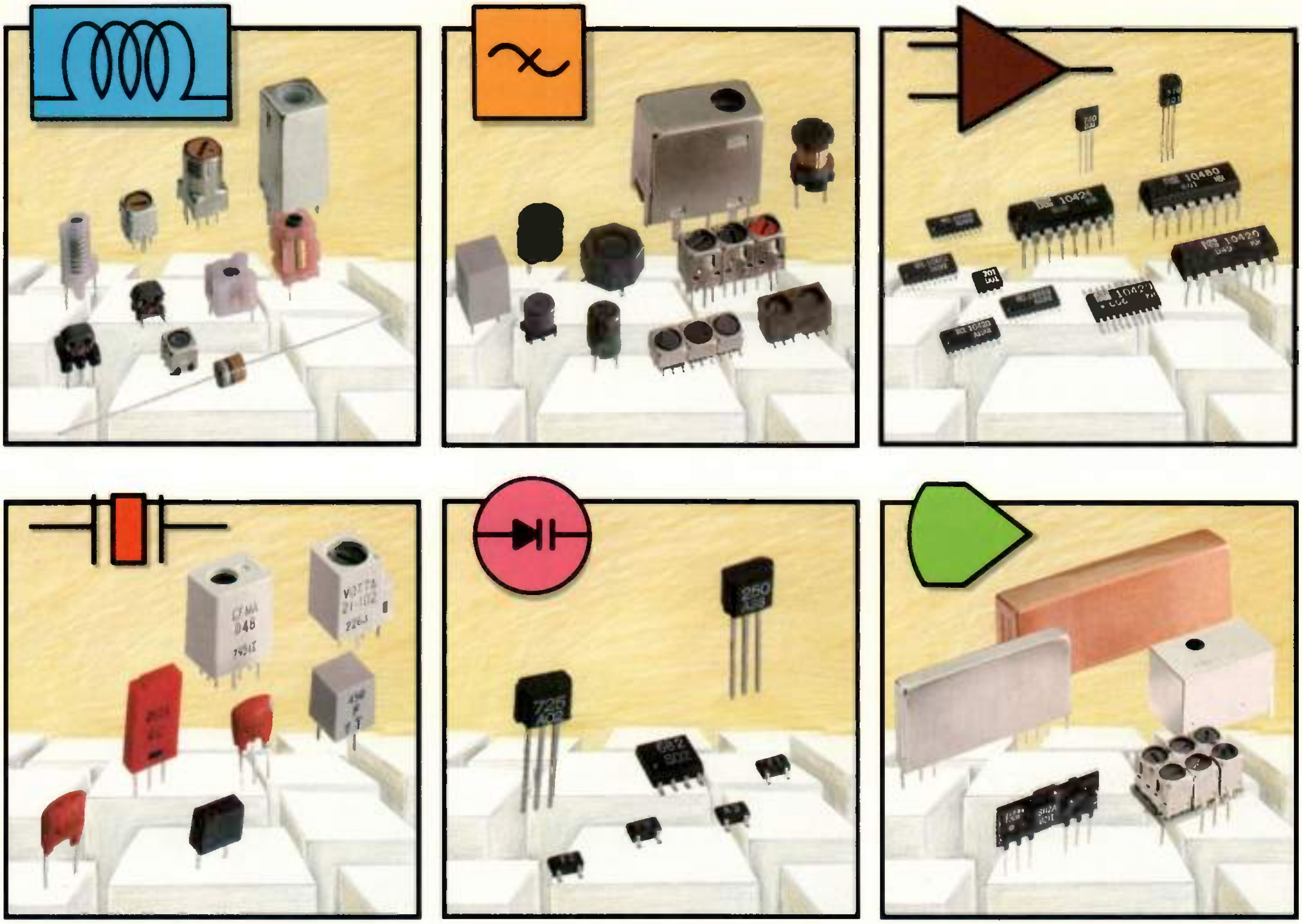
Questions and comments should be directed to *RF Design* magazine.

Advertising Index



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