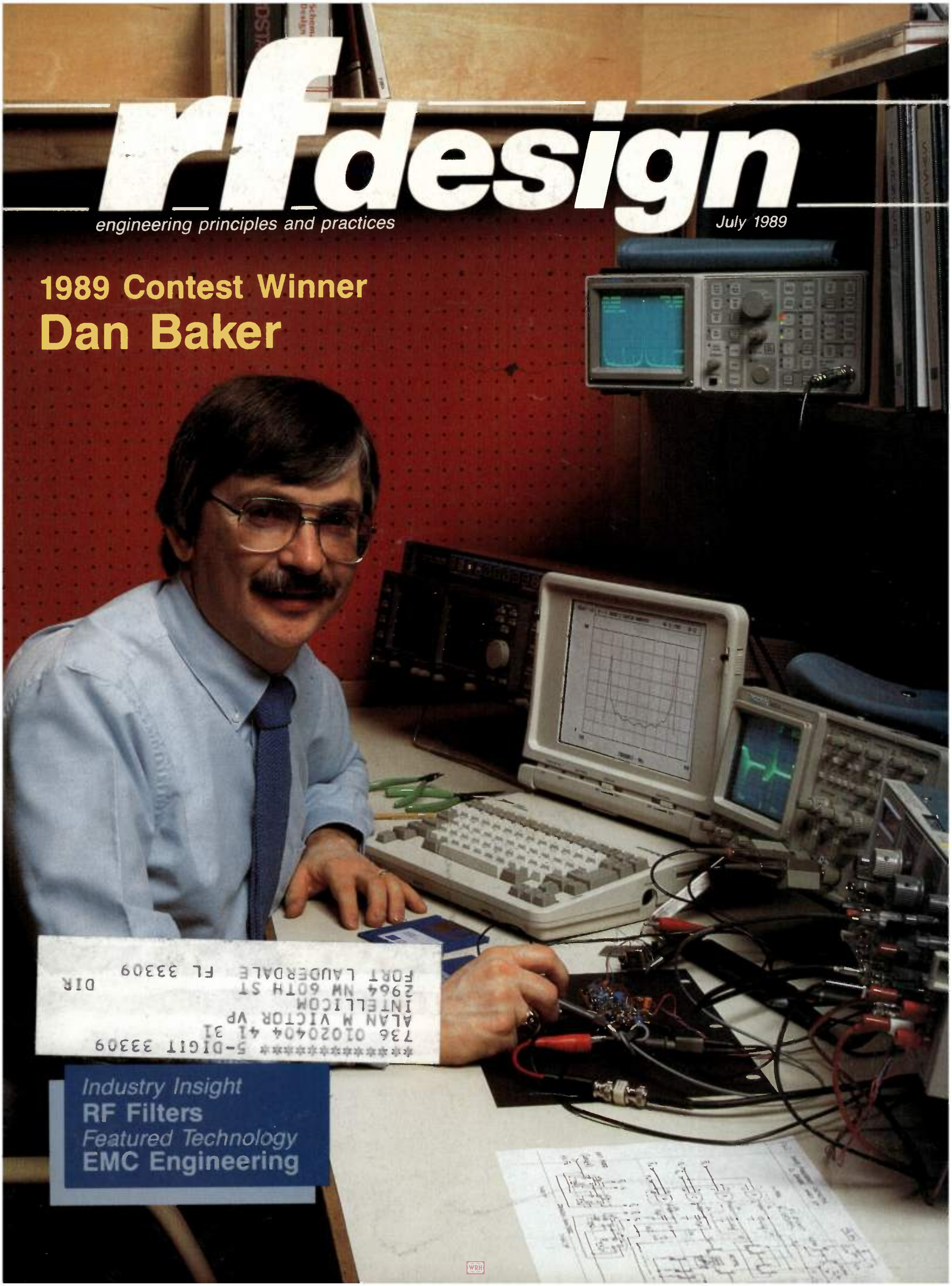
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engineering principles and practices
July 1989
1989 Contest Winner
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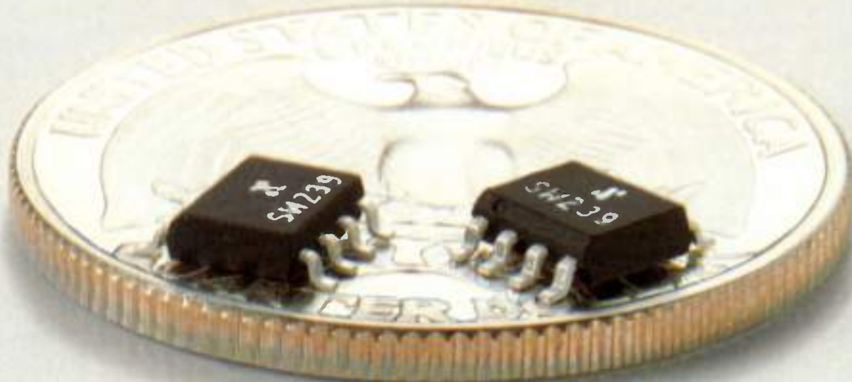
Two SPDT Bits.



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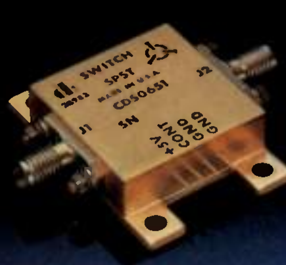

- Operating frequency: 10MHz-2GHz
- Transition time: 5ns
- Switching speed: 16ns
- Isolation: 57dB 10 - 1000MHz
  52dB 1000 - 2000MHz
- Power Requirements: 2mA at +5V
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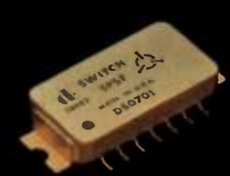

- Operating frequency: 10MHz-1GHz
- Transition time: 14ns
- Switching speed: 28ns
- Isolation: 70dB 10 - 200MHz
  62dB 200 - 500MHz
  52dB 500 - 1000MHz
- Power Requirements: 0.45mA at +15V
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- Operating frequency: 5MHz-4GHz
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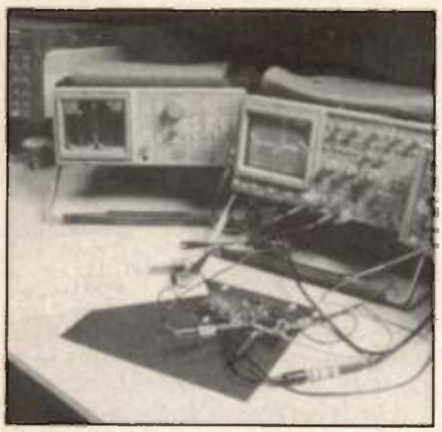
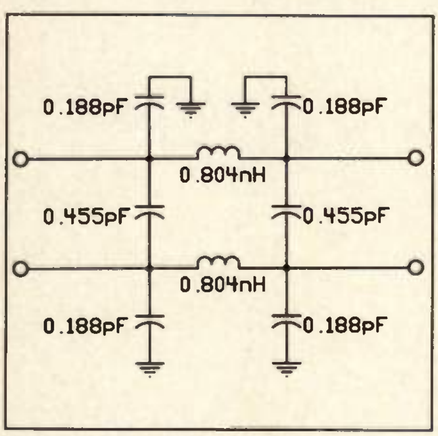
INFO/CARD 3

rf design

June 1989

Page 28 — Design Contest Winners


Page 35 — Phase/Frequency Detector



Page 56 — Quadrature Coupler

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rf editorial

The New Part 15: Proceed With Caution

By Gary A. Breed
Editor

The newly revised Part 15 regulations from the FCC represent a great opportunity for new product development, especially for wireless consumer convenience devices. Home security, VCR and stereo distribution, and computer communications devices are all possibilities. Industrial communications and control applications are another outstanding possibility.

I'd like to get up on my EMC soapbox, however, and issue a word of caution to those considering the development of such equipment: Put susceptibility to interference high on your list of design concerns.

My concern comes from several directions. First, I am a consumer of electronic equipment, and I live about one mile from a 5 kW AM broadcast station. The stray RF makes my frequency counter read 1430 kHz when no other input is applied, and I have to be careful how I run my stereo speaker wires. My background in the broadcast industry is a reminder that high power AM, FM and TV facilities are located near lots of other residential areas. Also, as a ham radio operator, I like to keep good relations with my neighbors.

Manufacturers are required to label Part 15 devices. Devices other than radio or TV receivers and cable TV selector switches require the following statement: "This device complies with Part 15 of the FCC Rules. Operation is subject to the following two conditions: (1) this device may not cause harmful interference, and (2) this device must accept any interference received, including interference that may cause undesired operation." This consumer warning will be even more important as new devices are marketed.

The FCC has stated that it is likely to consider additional labelling which would identify the responsible party (the manufacturer or importer). I wholeheartedly endorse this idea. Consumers need to know where to go when things don't work. As professionals, we understand how Part 15 works, but the average citizen only knows whether his electronic gadget works or not. These labels should go a long way toward making consumers aware of the limitations of the devices they are using.

In exchange for a more open marketplace for new ideas, we get more potential victims of interference. Let's help protect them through good design and correct information.

Gary A. Breed

Don't miss the contest results on page 28. This year's entries have been the best yet, representing some unique "fun" ideas and some very significant technical developments. To show off all of these great ideas, we will publish at least one contest entry every month from now on.

rf design

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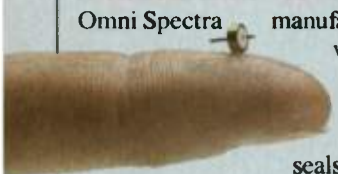

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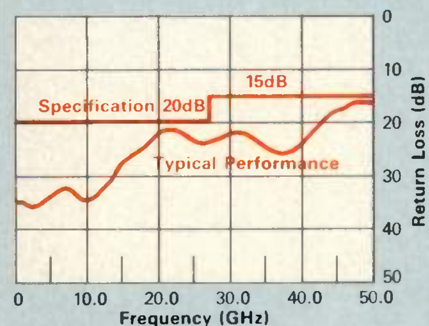

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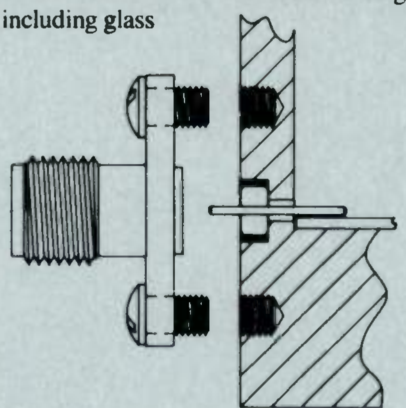

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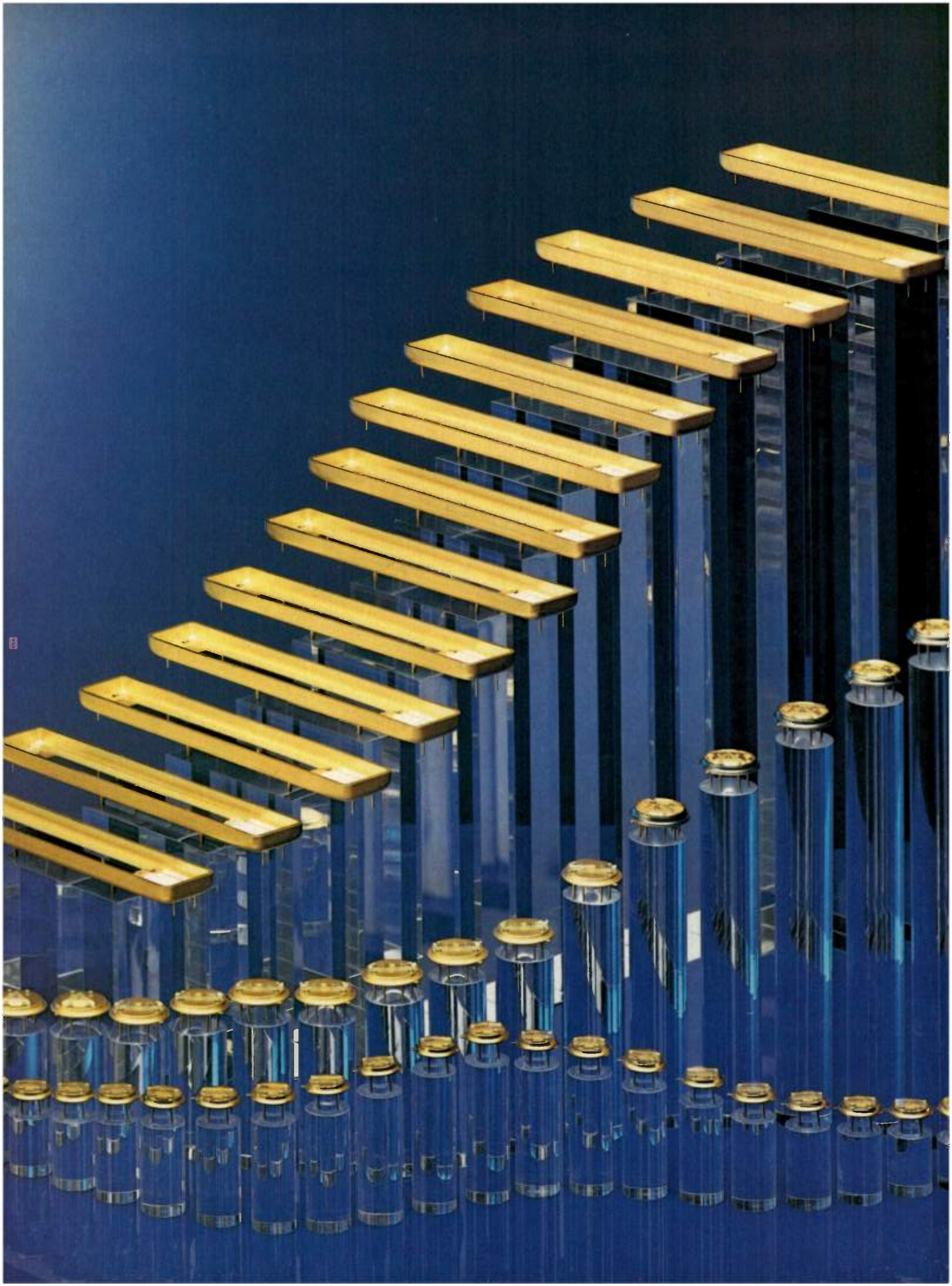

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INFO/CARD 10

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

A Reader's Compliment

Editor:

I want to thank you for seeing to it that all abbreviations and acronyms are defined early in *RF Design* articles. Reading ease is surely enhanced because of this attention to completeness. Just recently, I dropped my subscription to a magazine which catered mainly to individuals who apparently didn't need further enlightenment about such things. A newcomer like myself was too lost to read even the ads without a struggle.

No one should be offended because a well-known acronym is initially spelled out in an article, but when it is never understood by someone seeking knowledge in a relatively new field, there is good reason to be upset.

Franklin Swan
TV 38/WCFC
Chicago, Illinois

Cellular Amplifier Note

Editor:

As I read "New Transistors for the Next Generation of Cellular Base Stations" (Apr. 1989, *RF Design*), a few questions arose about the design.

The characteristic impedance of the transmission line comprising the balun transformer in Figure 3 was not specified. That value is apparently 25 ohms so that the balanced impedance will be 2 X 12.5 ohms. In that case, an additional impedance transformation from 50 to 25 ohms is required and is performed via line sections L2, L3.

Figure 6 depicts a balanced impedance transformation from base to base (about 6 ohms) to the balanced output impedance of the balun. Using a Smith chart with balanced lines is probably possible, but a simpler procedure would be to use the lines in the unbalanced mode and match the base-to-ground impedance to 12.5 ohms unbalanced.

Arie Shor
Acrian, Inc.
San Jose, California

Corrections

The following corrections to "A Mixer Spurious Plotting Program" (May 1989, *RF Design*) should be noted:

- In Figure 3, p. 34, the third line in column 1 should read:

$$LO > F_{out} > F_{in}$$

- The last sentence on p. 36 should read, "At an input frequency of 240 MHz ..."

- In Appendix 1, p. 41, under "Fixed input frequency option":

$$Delf_{x1} = [F_{min}(M-M1) - F_{in}(N1M-M1N)]/M1$$

$$Delf_{x2} = [F_{max}(M-M1) - Fin(N1M-M1N)]/M1$$

- In Appendix 1, p. 43, under "Fixed input frequency":

$$Deltaf_{x} = [F_{calc}(M-M1) - F_{rf}(N1M-M1N)]/M1$$

and the first equation in Part B should read:

$$F_2 = F_c + [BW1 + (BW1 \wedge 2 + 4F_c) \wedge .5]/2$$

Sunspots Herald Unusual Times Ahead

Communications around the world will be experiencing disruption, distortion, and just plain strange effects as sunspot cycle number 22 approaches its peak. The unusual phenomena, which will be occurring more frequently during the next few months, are the result of increased activity on the surface of the sun.

Sunspots occur in cycles averaging about 11 years in length, with a rise and then fall in the number of sunspots observed. Other kinds of solar activity, including solar flares and prominences, vary with the same period. The result is higher-than-normal levels of electrical activity in the ionosphere as the sunspot cycle moves toward its maximum.

Cycle 22, expected to peak sometime in early 1990, is shaping up to be one of the strongest on record. Forecasts about where a cycle will peak are based largely on the rate at which the rise to peak is occurring. In other words, the faster the rise in sunspot number, the higher the maximum is likely to be. The rapid and early rise in the number of sunspots and other solar phenomena associated with the present cycle has experts guessing that Cycle 22 may match or even surpass the most active sunspot cycle on record.

Sunspots are dark, relatively cool areas appearing in groups on the sun's surface. Intense magnetic fields, as much as several thousand times stronger than the average solar magnetic field, are associated with a sunspot region. It is thought that sunspots occur when strong magnetic fields interior to the sun periodically make their presence felt on the surface, perhaps as an effect of the unusual rotational movement of the sun.

The resulting dramatic increase in solar radiation can have a variety of consequences on Earth. The solar activity associated with sunspots, solar flares and other phenomena increases the ionization of the ionosphere. This can have some unusual effects on radio transmissions. Particles ejected from erupting solar flares can travel to Earth and cause power blackouts, disruptions in radio communication, and surges in power lines. A sunspot cycle with an exceptionally high maximum could result in television interference or reception of multiple stations on the same channel. Due to the increasing density and miniaturization in integrated circuits, avionics systems are becoming more susceptible to errors and failures caused by solar activity. Extremely strong solar flares can even present a radiation hazard to astronauts in space.

Electronic systems in satellites are vulnerable to heightened levels of solar radiation, and control from the ground can be interrupted or made impossible. The increase in solar activity causes a heating and expansion of Earth's atmosphere, a situation which threatens the useful lifetime of an orbiting satellite. The increased drag a satellite experiences as the solar cycle approaches its maximum can alter a satellite's orbit or even cause it to fall to Earth prematurely. This was the reason for the early re-entry and breakup of Skylab in 1979. The Solar Maximum Mission satellite, or Solar Max, which was launched in 1980 to study the sun, is expected to become a victim of the current sunspot cycle. Scientists estimate that Solar Max will fall from space and burn up in Earth's atmosphere sometime in 1990, or even sooner.

Not all the consequences of the sunspot cycle will be unpleasant ones. Beautiful auroras should occur more frequently, and will likely be observable over a greater area on Earth. One group of people likely to enjoy this period of high solar activity are ham radio operators. The increased activity on the sun's surface translates into greatly improved radio conditions, unusual propagation effects, and good signal quality. Of course, short wave radio blackouts will be more common, too, as the number of solar flares rises with the sunspot cycle.

The most active period of the cycle will probably last several years. So, at the very least, some interesting and exciting times are ahead as the sun continues to exert its considerable influence on our world.

EMC Expo 89 Convenes in D.C.— Designers and engineers will be gathering in Washington, D.C., August 1-3, 1989, for EMC Expo 89. The conference, to be held at the Sheraton Washington Hotel, will be offering attendees an agenda focused on current electromagnetic compatibility (EMC) issues, problems and solutions. There will be a variety of technical sessions from which to choose, along with an exhibit floor featuring the products and services of more than 100 companies in the EMC field. The technical program will be addressing EMI/EMC topics in three "tracks": Introduction to EMC, Advanced EMC, and Applications for EMC. Also scheduled is a one-day hands-on course on electrostatic discharge (ESD) control materials and techniques, conducted by 3M's Electrostatic Products Group. A Navy round table discussion concerning certification of EMC laboratories and engineers promises to be one of the highlights of this year's conference. For further details and registration information for EMC Expo 89, contact: *EMC Technology*, P.O. Box D, State Route 625, Gainesville, VA 22065. Tel: (703) 347-0030

Cause of Telescope Collapse Determined—A panel of engineers reporting to the National Science Foundation has determined the probable cause of the collapse last November of a 300-foot-diameter radio telescope, in operation at the National Radio Astronomy Observatory (NRAO) in Green Bank, W.Va. The April 29, 1989 issue of *Science News* reports the panel's conclusion that the telescope's collapse was probably caused by the fracture of a single steel plate in the supporting structure. The plate that failed was one which experienced great stress anytime the telescope was moving. Due to the location of the plate, however, routine inspection was difficult. Metallurgical analysis revealed small cracks in the plate, which investigators suspect grew and eventually led to the collapse of the telescope.

The instrument, a partially steerable parabolic radio antenna able to operate at wavelengths as short as 6 cm, had been in operation for 26 years. It was used in the study of cosmic radio sources, and had detected signals coming from nearly 10 billion light-years away (*Science News*, Nov. 26, 1988).

TIA Offers FCC Part 15 Seminar— The Telecommunications Industry Association (TIA) will be offering an in-depth informational seminar on the recent revisions to the Federal Communications Commission's (FCC) Part 15. The revisions, effective June 23, affect emissions requirements for most digital electronics and unlicensed RF equipment. Featured speakers at the TIA seminar will include the FCC authors of the changes, as well as industry experts. The seminar is scheduled for July 24-25, 1989, at the Mayflower Hotel in Washington, D.C. The registration fee is $395, with a discount available for members of TIA or its associated organization, the Electronic Industries Association (EIA). Registration and hotel information is available by contacting Suzanne Mullendore, TIA, 1722 Eye Street N.W., Suite 440, Washington, DC 20006. Tel: (202) 457-4937.

Report Details Japanese Research Efforts—A new report from the U.S. Department of Commerce provides information about Japanese superconductor, ceramic and semiconductor research activities. *Basic Research in Superconductor, Ceramic and Semiconductor Sciences at Selected Japanese Laboratories* is based on extensive visits to Japanese R&D institutions by Dr. Robert Gottschall of the Department of Energy. The report offers detailed notes on work being conducted at university, industrial and government laboratories in Japan. The report was sponsored jointly by the Technology Administration's Japanese Technical Literature Program, the Department of Energy, the Office of Naval Research, and the Congressional Office of Technology Assessment. The publication, PB89-172464, is available for $28.95 from the National Technical Information Service, 5285 Port Royal Road, Springfield, VA 22161. Tel: (703) 487-4650

Researchers Demonstrate Thin-Film Superconducting Microwave Filter—Operation of a thin-film, superconducting microwave component cooled by liquid nitrogen has been developed and demonstrated at the David Sarnoff Research Center, a subsidiary of SRI International, in a team effort with Bellcore (Bell Communications Research Inc.). The Sarnoff group has succeeded in fabricating a microwave comb filter that is 100 times more efficient than similar conventional filters made of copper. Tests conducted have shown that the superconductive filter has a Q of 3000 compared with a Q of 30 for conventional filters using a copper transmission line at 2 GHz. At 10 GHz, the filter is still 10 times more efficient than the comparable copper filter.

GAO Urges A Pause in Military HF Radio Spending—The May 8, 1989 issue of *Electronic News* reports that the

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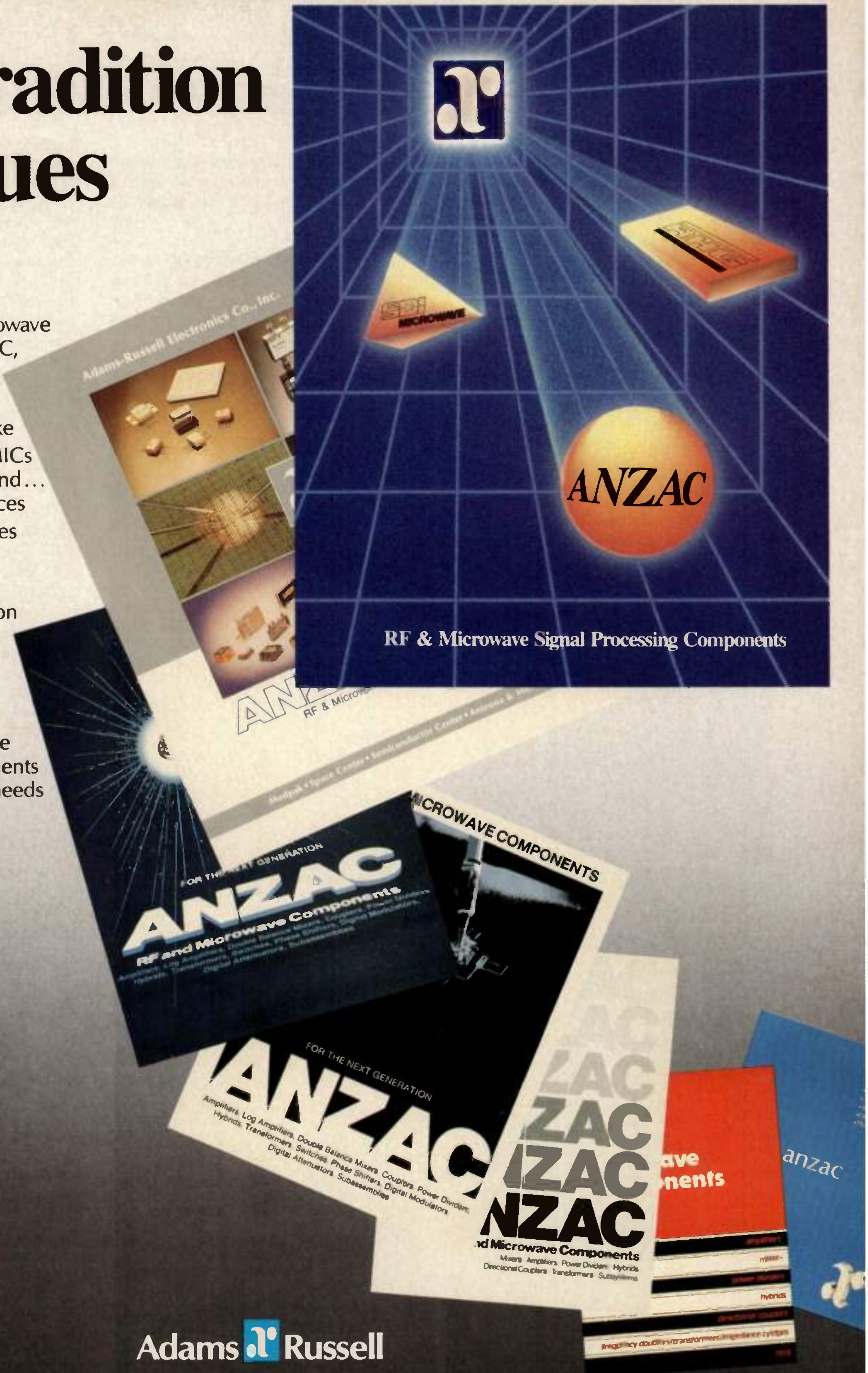

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INFO/CARD 16

General Accounting Office (GAO) has recommended that nearly $400 million in spending for new Army and Air Force antijam HF radio systems be put on hold until interoperability can be assured. According to the GAO, both services are going ahead with spending plans to enhance their HF radios' capabilities, despite the fact that the radios can't talk to each other. A Defense Department HF automatic link standard adopted last fall requires interoperability between the systems.

McCoy Wins $3.4 Million Contract From Hughes Aircraft—McCoy Electronics Co., of Mount Holly Springs, Pa., has received a $3.4 million contract from Hughes Aircraft Co. for high-reliability IF-strips to be used in U.S. Navy UHF communication satellites. The IF-strips will employ crystal filters, amplifiers and ground-controllable circuitry which will provide reliable, long-term, stable channel selectivity. These devices will be assembled by McCoy and supplied to Hughes over the next three years. The satellites are part of the UHF Follow-On program, which is the primary defense communication system in the United States. Through a network of satellites, this system supports the Navy's global communications network, serving ships at sea and a variety of other U.S. military fixed and mobile terminals.

Navy to Purchase Exciters From AIL Systems—The U.S. Navy has exercised a $33 million option under an existing contract with AIL Systems Inc. to purchase 70 additional advanced electronic assemblies for its EA-6B Prowler electronic warfare aircraft. The assemblies, known as universal exciters, are part of the AN/ALQ-99 tactical jamming system on board the Grumman Aerospace-built Prowler.

Varian to Acquire WJ Product Line—Varian Associates Inc. and Watkins-Johnson Co. (WJ) have reached an agreement in principle for Varian to acquire WJ's space communications product line for an undisclosed amount of cash. The product line includes high-reliability traveling-wave-tube amplifiers (TWTAs) and power supplies for use in satellite-based space communication systems. The line will be assigned to Varian's Microwave Equipment Division in Santa Clara, Calif., which makes power supplies, amplifiers and transmitters used in ground-based satellite communication, radar, electronic countermeasures, and varied scientific and instrumentation applications.

Phonon in Management Buyout—Phonon Corp., of Simsbury, Conn., has been purchased in a buyout by the company's management. Thomson-CSF of France, previously Phonon's majority stockholder, has sold all its stock to a U.S. management group led by Phonon's president and chairman of the board, Dr. Tom Martin. The buyout will allow Phonon to pursue classified military business and will permit the benefits of Federal Small Business status, both of which were impossible under the company's previous ownership.

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Published by ATC July 1989

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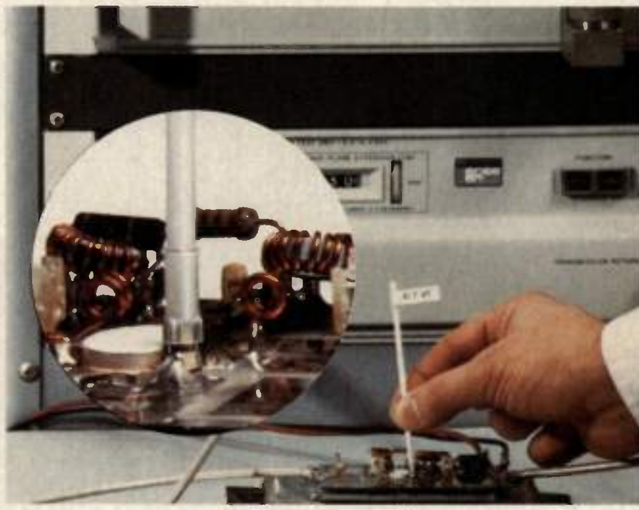

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See us at the MTT-S Show

Manufactured Filters See Growing Use

By Mark Gomez
Technical Editor

As the RF spectrum gets even more crowded, filters are finding more and more homes in RF systems. Also, with the advent of digital technology in RF, the importance of filtering is increasing. Control of spurious effects and tighter emissions regulations are just two of the contributing factors in the growing use of manufactured filters as RF "components."

This activity in filtering translates into market growth. "The communication spectrum is getting more and more crowded every day," says Chris Chrissie, new business manager at Telonic Berkeley. "One of the disadvantages of that crowding is that you must reduce spurious emission levels and reject them to a greater degree," he adds. According to Jim Lyons, vice-president of engineering at Allen Avionics, there is going to be growth in the filter market, especially in the area of anti-aliasing filters for use in digital circuits. "There are many applications in communication, digital signal processing and analog signal processing where filters play an important role," he comments. Scott Evans, product marketing manager at Comlinear, sees growing interest in anti-aliasing filters for analog-to-digital converters. "We see activity in the satellite band, the telephone industry, and the mobile radio and paging markets," observes Glyn Bostick, president of Microwave Filter Company. "Our particular niche is EMI filters," notes Dr. Jeff Chambers, technical director of Oxley Developments Company, "and we are supplying into a range of growing markets like industrial, avionics and the military."

Filters are seeing growing use due to various new applications. "I see more and more people using filters to solve emission problems inside systems where they used to use high-class amplifiers to prevent those emissions," observes Chrissie. "Now, instead of spending an extra $14,000 or $15,000 to buy an amplifier with reduced harmonics or a low third-order intermodulation figure, users tend to buy a much less expensive amplifier and a $100 or $150 filter."

The filter industry, like other RF segments, is experiencing miniaturization pressure. There is a growing need for manufacturers to supply the same performance in a smaller package. "We get many inquiries for things like surface-mount packages," comments Kent Truscott, applications engineer at EG&G Reticon. "A limitation in miniaturization is that the filter has to be larger than the smallest available component," notes Steve Sodaro, vice-president of marketing and sales at TTE. The laws of physics no doubt impose some limitations on packaging as well. "Packaging is a growing factor, and sometimes the requirements are unrealistic and defy the laws of physics," states Bostick. Chambers notes that manufacturers are under a great deal of pressure to squeeze filters into a much smaller volume. "We are having to build our filters which are ceramic dielectric-based devices inside a multi-way connector," he says. "About 99.9 percent of our customers are looking to reduce board space and increase performance," comments Lyons.

Together with denser packaging, improved performance is prevailing in filters. An example of improved performance can be seen in the switched capacitor filter area. "Switched capacitor filters are becoming more and more popular. It has become a much less noisy product than it has been in the past," notes Sodaro. "Since switched capacitor filters are low power, they seem to be the way to go, especially in portable equipment," says Truscott. He also points out that EG&G Reticon continues to get requests for quieter filters with less clock feed-through and better dynamic range.

Active filtering, a growing solution in RF, implies some sort of amplification device which, rather than just rejecting a certain band of undesired frequencies, combines amplification of the desired signal with some rejection of the undesired signal. "Active filters are nice because you do not have to match impedances to make the filter operate and up to 20 dB or so of gain is achievable," notes Sodaro. "However, most people prefer high performance passive filters because they have less harmonic distortion and less effects on the rest of the circuit," he adds. "Although the passive solution is attractive, many designers require active filtering, especially for integrator loops," says Scott Evans.

Filters by themselves are in a very competitive industry. In order to be more cost-effective and to obtain a competitive edge, a few companies are moving into higher levels of integration where they offer combinations of other building blocks together with filters. Jim Price, vice-president of sales at K&L Microwave, notes, "There is more and more demand for us to integrate our filters with switches, amplifiers, mixers, etc."

Since there are many filter companies in the market, competitive pricing is a key issue. "Pricing will be flat at best for standard units," says Price. Telonic Berkeley's Chrissie states that pricing has been headed down continuously for the past number of years while being offset by inflation levels. The contrary is observed by Dennis Hook, president of Daden Associates. "Pricing is going to be headed upwards," he says. Bostick shares this view and attributes the rise to greater levels of sophistication and rising inflation.

As the RF industry experiences continued change, manufactured filters seem to be reinforcing their place in RF systems. As a result of improved engineering and experience, filters are slowly moving from being an afterthought for cleaning up undesired byproducts to being considered in initial system design stages. Even though the filter industry is still considered to be rather custom, in special cases these products can be deemed components where off-the-shelf solutions are starting to become more common. rf

rf calendar

July 19, 1989
Electronic Materials Application Show
St. Louis, MO
Information: Ed Partyka, Complete Reading, 300 N. Mannheim Road, Hillside, IL 60162. Tel: (312) 547-5400.

July 24-27, 1989
1989 SBMO International Microwave Symposium/Brazil
Maksoud Plaza, Sao Paulo, Brazil
Information: Dr. Octavio M. Andrade, IMT—Escola de Engenharia Maua, Estrada das Lagrimas 2035, 09580 S. Caetano do Sul, Sao Paulo, Brazil. Tel: (011) 442-6944; Telex: 1145234 AUAT BR

August 1-3, 1989
EMC Expo 89
Sheraton Washington Hotel, Washington, DC
Information: *EMC Technology,* P.O. Box D, State Route 625, Gainesville, VA 22065. Tel: (703) 347-0030

August 14-17, 1989
Triennial URSI International Symposium on Electromagnetic Theory
Royal Institute of Technology, Stockholm, Sweden
Information: S. Strom, Dept. of Electromagnetic Theory, Royal Institute of Technology, S-100 44, Stockholm, Sweden.

August 22-25, 1989
1989 International Symposium on Antennas and Propagation
Nippon Toshi Center, Tokyo, Japan
Information: Dr. Takashi Katagi, Mitsubishi Electric Corp., 325 Kamimachiya, Kamakura, 247 Japan. Tel: (0467) 44-8862; Fax: (0467) 47-2005

August 29-31, 1989
11th Quartz Devices Conference and Exhibition
Kansas City Westin Crown Center, Kansas City, MO
Information: EIA, Components Group, 1722 Eye Street N.W., Washington, DC 20006. Tel: (202) 457-4930

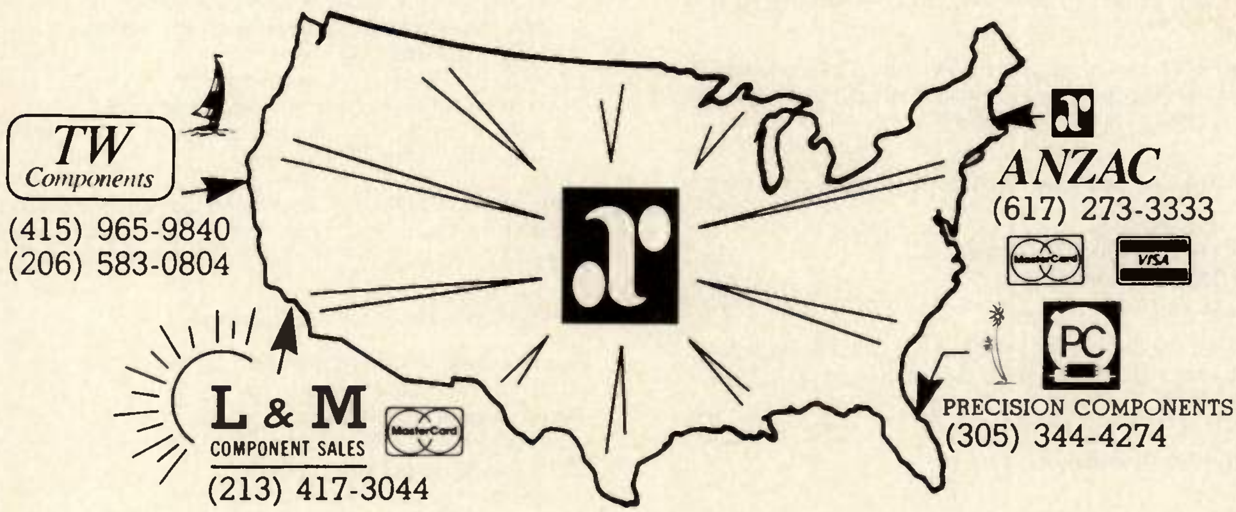
September 4-8, 1989
19th European Microwave Conference and Exhibition
Wembley Conference Centre, London, England
Information: Microwave Exhibitions and Publishers Ltd., 90 Calverly Road, Tunbridge Wells, Kent TN1 2UN, England.
Tel: (0892) 44027; Fax: (0892) 41023

September 4-8, 1989
2nd International Symposium on Recent Advances in Microwave Technology '89
Beijing, China
Information: Prof. Banmali Rawat, Dept. of Electrical Engineering/Computer Science, University of Nevada - Reno, Reno, NV 89557-0030. Tel: (702) 784-6927; Fax: (702) 784-1300

September 8-10, 1989
EMC '89/Nagoya
Nagoya Trade and Industry Center, Nagoya City, Aichi, Japan
Information: EMC '89/Nagoya, Prof. Yasumitsu Miyazaki, Toyohashi University of Technology, 1-1, Aza-Hibarigaoka, Tempaku-cho, Toyohashi-City, Aichi, 440 Japan. Tel: (81) 0532-47-0111, ext. 528; Fax: 0532-45-0480

RF & MICROWAVE NEWS

ANZAC Announces Coast to Coast Distributors



Burlington, Ma. The *ANZAC* division of Adams-Russell recently announced the opening of three distributors in the United States. The introduction of one distributorship in Florida, two on the West Coast, and *ANZAC*'s own standard product distribution center in Massachusetts now makes local procurement of *ANZAC* catalog components easier than ever.

A spokesperson for *ANZAC* expounded on the advantages design engineers and procurement agents are receiving by dealing with their local *ANZAC* distributor. "By opening regional distributors, we now offer 2 distinct advantages to our customers. The first is local delivery. Each distributor is fully stocked with *ANZAC* components and can provide off-the-shelf delivery in 24 hours or less. The second advantage is service. *ANZAC* distributors have years of technical experience in the RF & Microwave industry and are already familiar with the *ANZAC* product line. They can offer technical assistance to design problems and provide the devices to solve those problems right away."

Future plans for *ANZAC* distributors include the sale of components from other Adams-Russell Components Group companies such as RHG Electronics and SDI Microwave. *ANZAC* distributors are presently fully stocked. For more information, interested parties in these areas should call their local distributor direct.

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(206) 583-0804 - IN SEATTLE

So. California:
L&M Component Sales
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Suite 104
Los Angeles, CA 90045
(213) 417-3044

Southeast United States:
Precision Components
12020 N.W. 40th Street # 203
Coral Springs, FL 33065
(305) 344-4274

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The George Washington University

New HF Communications Technology
July 17-21, 1989, Orlando, FL
Radio-Wave Propagation for Communications System Engineering
July 24-28, 1989, Washington, DC
Grounding, Bonding, Shielding and Transient Protection
August 14-17, 1989, Hawaii

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

University Consortium for Continuing Education

Electronic Warfare
August 9-11, 1989, Washington, DC
Sonar Signal Processing
September 18-22, 1989, Washington, DC

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

Compliance Engineering

EMI
August 22, 1989, San Jose, CA
Safety
August 23, 1989, San Jose, CA
ESD
August 24, 1989, San Jose, CA
Telecom
August 25, 1989, San Jose, CA

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

EEsof, Inc.

Nonlinear FET Model Parameter Extraction (Xtract)
July 17-19, 1989, Westlake Village, CA
MMIC Design Workstation (MMIC)
July 24-28, 1989, Westlake Village, CA
Computer-Aided Engineering for Linear Microwave Circuits (Touchstone)
August 7-9, 1989, Westlake Village, CA
Computer-Aided Drafting for Microwave Circuits (MiCAD)
August 10-11, 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

EMC Services

Filter Design for Switching Supplies
July 24-25, 1989, San Francisco, CA
August 21-22, 1989, Chicago, IL
EMI Control for Switching Supplies
July 26-28, 1989, San Francisco, CA
August 23-25, 1989, Chicago, IL

Information: Sonya Nave, EMC Services, 11833 93rd Avenue North, Seminole, FL 34642. Tel: (813) 397-5854

Integrated Computer Systems

C Programming Hands-On Workshop
July 25-28, 1989, Los Angeles, CA
August 8-11, 1989, Washington, DC
Digital Signal Processing: Techniques and Applications
July 25-28, 1989, San Francisco, CA
July 25-28, 1989, Ottawa, Ontario, Canada
Introduction to Fiber Optic Communications
July 25-28, 1989, San Diego, CA
August 22-25, 1989, Los Angeles, CA
Troubleshooting Datacomm and Networks
July 25-28, 1989, Washington, DC
August 8-11, 1989, Boston, MA

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

Interference Control Technologies, Inc.

System Design and Integration for EMC
July 18-21, 1989, Palo Alto, CA
Grounding and Shielding
July 25-28, 1989, San Diego, CA
EMC Design and Measurement
August 7-11, 1989, Orlando, FL
Intro to EMI/RFI/EMC
August 8-10, 1989, Chicago, IL
Practical EMI Fixes
August 14-18, 1989, San Diego, CA

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

Praxis International, Inc.

EMP Hardening Design and Verification Techniques
August 7-9, 1989, Oklahoma City, OK
Facility Grounding, Shielding and Lightning Protection
August 21-23, 1989, San Diego, CA

Information: Praxis International, Inc., Exton Professional Building, Suite 103, 319 N. Pottstown Pike, Exton, PA 19341. Tel: (215) 524-0304

R & B Enterprises

Understanding and Applying MIL-STD-461C
October 11-13, 1989, Chicago, IL
Introduction to EMI for Non-EMI Personnel
October 16-17, 1989, Washington, DC
EMI/EMC in the Automotive System
October 16-18, 1989, Dearborn, MI
Printed Circuit Board and Wiring Design for EMI and ESD Control
October 18-19, 1989, Philadelphia, PA
Architectural Shielding
October 18-20, 1989, Washington, DC
Prominent E^3 Standards
October 23, 1989, Philadelphia, PA

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

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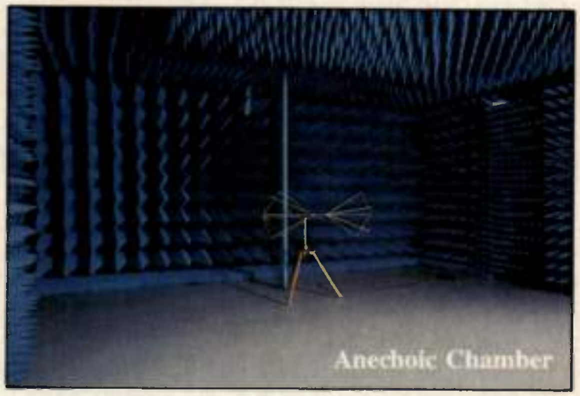
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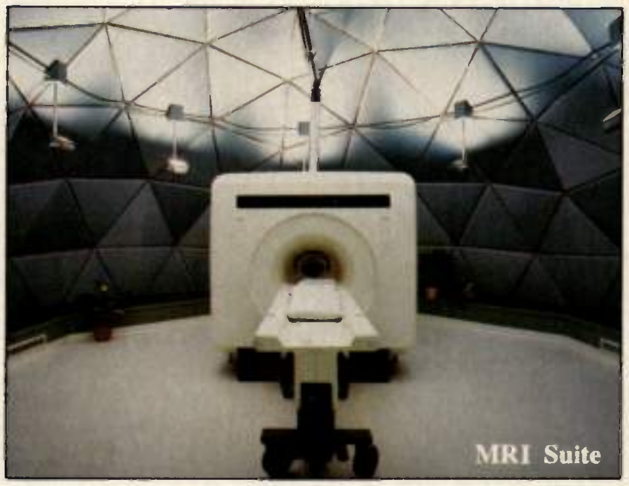
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1989 Contest Winners

Prize Winners in the Fourth Annual RF Design Awards Contest

By Gary A. Breed
Editor

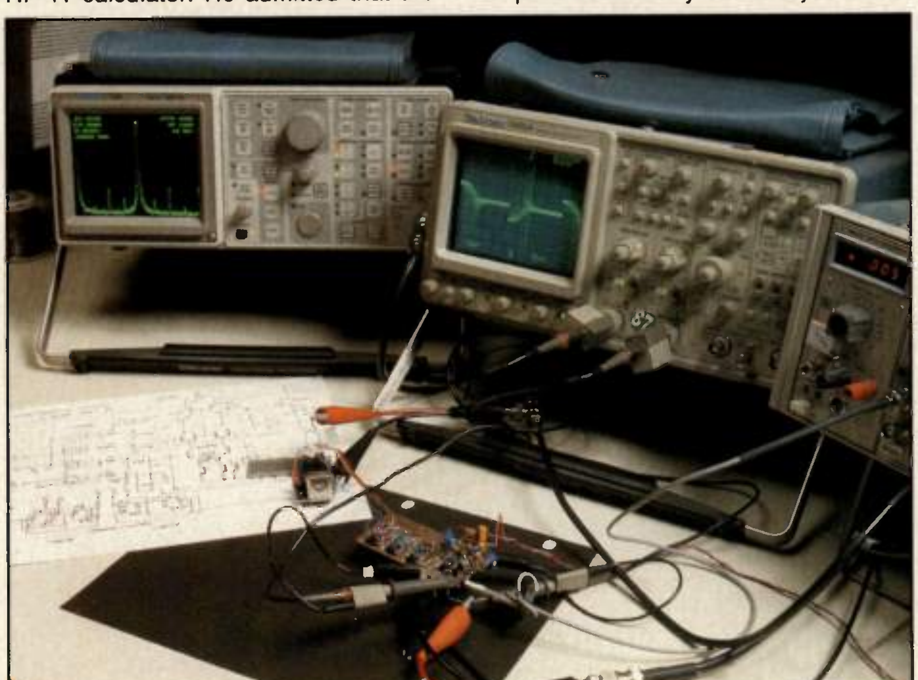
Six engineers have been awarded prizes in the Fourth Annual RF Design Awards Contest, representing just a few of many excellent ideas submitted to the panel of judges. For the fourth year in a row, the entries have gotten better, and the judging more difficult.

This year, the Grand Prize was donated by Webb Laboratories, a Zenith SupersPort laptop computer with "all the bells and whistles," including hard disk drive, math coprocessor, and carrying case. Of course, all of Webb Labs' software is included, too, including Filsolv, Syscad, Transcad, and the new Receiver Advantage.

This impressive package goes to Dan Baker of Tektronix' Television Products group in Beaverton, Oregon. Dan is a designer of video test equipment, with recent projects including a dual-display waveform monitor/vectorscope and a satellite-link monitor. Readers may remember him from the very first contest, when he won another Grand Prize, an HP-41 calculator. He admitted that the modest size of that prize, compared to subsequent contests, was a considerable part of his motivation to develop another championship entry. Well, after three years, he has won again, with a design that is sure to get PLL designers thinking.

The winning design is "A Reference-Cancelling Phase/Frequency Detector," which was developed for two purposes. First, it demonstrates that the "zero-order hold" model for sampling (digital) phase detectors is invalid. It also demonstrates that PLL lock takes only two sample periods in such a system. In order to demonstrate these things, a phase detector had to be designed which had a bandwidth of the same order as the reference frequency. The design uses a complex balanced mixing scheme to cancel the reference component at the detector output, allowing the bandwidth to equal, or even exceed, the reference frequency.

The complete winning entry starts on page 35. The article has been printed on perforated heavy stock for you to tear out, punch, and place in a three-ring binder. Every month from now on, at least one RF Design Awards Contest entry will be published, to add to your collection. We know that all articles in *RF Design* are worth keeping, but the contest entries represent their authors' very best ideas. They are a special collection to have!



Dan Baker's winning phase detector circuit cancels the reference frequency, reducing the reference sidebands at the output.

The Runners-Up

Five RF engineers received Fluke 87 digital multimeters from John Fluke Manufacturing Company. The model 87 features true RMS AC readings, an analog indicator, capacitance measurement, plus the usual voltage, current and resistance capabilities.

Alan Carr is the first among runner-up prize winners. Alan is a design consultant from Boulder, Colorado, whose primary work has been with Erbtec Engineering. He has developed a high power directional coupler which uses a divided line structure to solve the problems of coupling ratio, directivity, precision, and power handling. At low VHF (the coupler was designed for a 64 MHz system), neither ferrite toroidal couplers nor stripline parallel couplers are optimum choices.

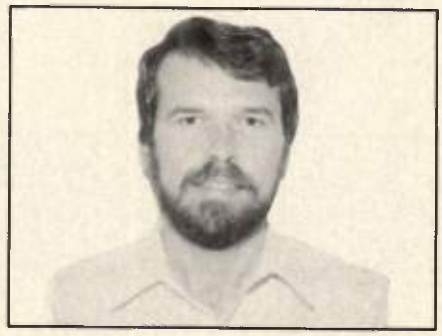
The coupler, for which a patent disclosure has been filed, is used in Erbtec's new 20 kW MRI (magnetic resonance imaging) power amplifier. The special requirements of MRI which required a better coupler are high pulse power, low duty cycle, and precision measurement to protect magnet coils. This coupler combines the properties of ferrite and stripline designs to achieve the required performance.

Gary Thomas is another prize winner. A senior staff engineer at General Electric Mobile Communications in Lynchburg, Virginia, Gary's entry is an improved electrically tunable bandpass filter. The filter uses a resistive bridge configuration to isolate a single-tuned notch, resulting in a bandpass filter which has much better Q than a single-tuned circuit.

The balanced bridge configuration of the filter also isolates the out-of-band signals from the tuned circuit, reducing

Alan Carr checks out an MRI amplifier with his new Fluke 87 multimeter.

Gary Thomas is a prize winner, for his voltage-tuned filter design.



A balanced oscillator design made Brani Petrovic a winner.

the RF level across the varactor diodes. What this filter design accomplishes is simpler tuning and a narrow passband, but with fewer components and lower cost than conventional multiple varactor-tuned resonators. A patent application has also been filed for this design.

A Varactor Frequency Divider

The list of winners continues with William Hoffert, who submitted "Frequency Division With Varactor Diodes" for the judges' consideration. Mr. Hoffert is a consultant to Los Alamos National Laboratory and lives in Albuquerque, New Mexico. His design is currently in use in an accelerator test stand, a divide-by-six unit providing a 68.3 MHz output from a 410 MHz input.

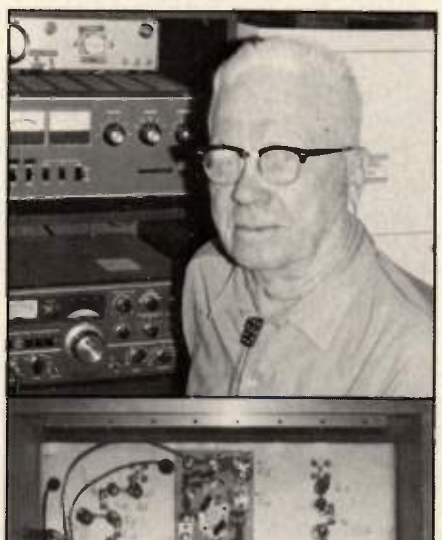
The principle of varactor divider is similar to multiplication, where distortion products (harmonics) of the fundamental are selectively recovered at the output. In the divider, the harmonics interact to generate lower frequency products. Idler circuits are used to enhance the desired frequencies involved in the division process. Unlike digital dividers, the varactor circuit can handle significant power levels. The prototype uses 1 watt input at 410 MHz.

George Vella-Coleiro, AT&T Laboratories, Murray Hill, New Jersey is the next winner to be introduced. His quartz watch time base monitor circuit solves a problem involving Heisenberg's Uncertainty Principle: The extremely low power circuitry in a quartz watch circuit will be affected by even a high impedance probe, so an attempt to directly measure the frequency of the 32,768 Hz clock crystal will be unsuccessful.

The solution he presents is to acoustically couple a transducer to the crystal, amplify the signal and measure the frequency. The low levels, however, guarantee that noise and interference will be present in such a system. To overcome interference, a differential transducer amplifier is used, with high common-mode rejection. The noise problem is solved by using a phase-locked loop with a very long time constant to recover the clock frequency.

The final prize winner is Branislav Petrovic. Brani is an engineer at General Instrument Corporation's Video Cipher Division in San Diego, California. His contest entry is a balanced RF oscillator circuit, to use when a balun transformer is not the best way to get 180-degree signal components for balanced mixers and modulators, phase detectors, frequency multipliers, or dividers.

The oscillator he describes is a VCO, tuning 1.3 to 1.9 GHz. Two identical oscillators share a common tank circuit,



William Hoffert's winning frequency divider circuit is used in an accelerator at Los Alamos.

which has its point of symmetry, or "center tap," floating. Common push-pull or push-push circuits would normally have a grounded center tap. This oscillator has low harmonic content, since its topology cancels even-order harmonics, with good amplitude and phase balance, although these are dependent on layout and component variations.

These entries will all be published in the next few issues of *RF Design*, in addition to other contest entries. Some other great ideas that you will see are: a phase shifter, a low-cost modem, a push-pull MMIC VCO, and an ultra-simple transmitter. These are just a few of the interesting circuits that didn't win prizes, but are certainly worthy of recognition and publication.

It wasn't planned, but the prize winning RF engineers represent a wide variety of geographical areas and industry types. Our six prize winners were from Oregon, California, New Mexico, New Jersey, Colorado and Virginia. Other excellent entries came from New England, the Southeast and Midwest, plus Canada, Greece and Switzerland. The prize recipients represent the test equipment industry, medical electronics, government scientific research, private research, the satellite and CATV industry, and land mobile communications. Now you know why the attitude at *RF Design* is that *diversity* is the key word in RF!

With this conclusion to the 1989 RF Design Awards Contest, we immediately begin the search for next year's winner. Many of the prizes have been selected, and the announcement of rules and deadlines will appear in the August issue. Our thanks to the winners for making this the best contest yet, and our thanks to all our readers for your enthusiastic response to the contest articles. Get ready for the Fifth Annual RF Design Awards in 1990! rf

Avoiding Ground Problems in High-Speed Circuits

By Jeff Barrow
Analog Devices, Inc.
Computer Labs Division

A designer typically considers ground as one of four nodes: chassis ground, analog common, digital common, or device zero. The first ground is used for questions of safety, the second for a high-quality return path, the third for a low-quality return path, and the last is used rarely as a zero-current reference point. While most designers fully understand these simple grounding principles, they often degrade their circuit's performance by applying these DC grounding practices to high-frequency circuits.

Consider the issue of ground planes. High-frequency circuits often use the ground plane to return power and signal currents and as a reference node for D/A converters, A/D converters, track-and-hold amplifiers and other related analog circuitry. Unfortunately, ground planes, even lots of them, do not guarantee a quality ground reference node for AC circuits. In many cases it is not how many ground planes the designer uses that matters, but where they are used.

It's not difficult to understand the problem with ground planes after carefully examining what occurs in a high-speed circuit. In a simple design constructed with a two-layer circuit board (Figure 1), the top layer is an AC and DC current source connected to a single U-shaped copper trace going to via 2. This via starts on the bottom of the circuit board and connects to a solid-sheet of ground plane. The same current source is connected to via 1 as well, which also goes through the circuit board and connects to the ground plane. If the designer can understand where the current flows in the ground plane from via 1 to via 2 for both DC and AC signals, then he is 95 percent assured of avoiding ground noise in a high-frequency layout. (High-frequency may be as low as 100 kHz when the design involves a 12-bit A/D converter.)

The DC current flows directly from via 2 to via 1, taking the path of least resistance (Figure 2). Though some spreading of the current occurs, it does not flow back in a uniform sheet. AC currents are another matter. The AC current does not take the path of least resistance; it takes the path of least impedance. To understand that path, it is useful to have a quick review of inductance.

From classical analysis, inductance is proportional to the area of the loop made by the current flow. By using the right-hand rule and examining the magnetic fields created by loops, the designer can determine the inductance in the ground paths (Figure 3). Inside the loop, current along all parts of the loop produces magnetic field lines that add constructively. Outside the loop, field lines from the top to bottom trace add destructively. As a consequence, the larger the loop, the larger the stored magnetic energy and the greater the inductance.

Because $X = j\omega L$, a larger loop has greater impedance than a small loop. The AC current path in the ground plane can be determined by finding the loop with the smallest area with respect to the top conductor. In the absence of any resistance, AC current will search out the smallest loop due to its lower impedance.

In most AC circuits, the ground path of lowest impedance is directly under the top conductor and not straight back from via 2 to via 1 (Figure 4). Due to finite resistance in the ground plane, the current may flow somewhere between straight back and the top trace. But at frequencies above 2 MHz, the return path is nearly under the top trace.

Once this phenomenon is understood, corrective action can be taken. First, any high-frequency and low-impedance paths should be indicated to the draftsmen. The draftsmen should be told to eliminate any vias, keep them away from digital and make the line as

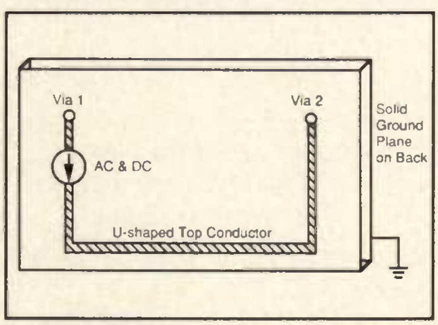


Figure 1. 2-sided PC board with U-shaped conductor on top and solid ground plane on bottom.

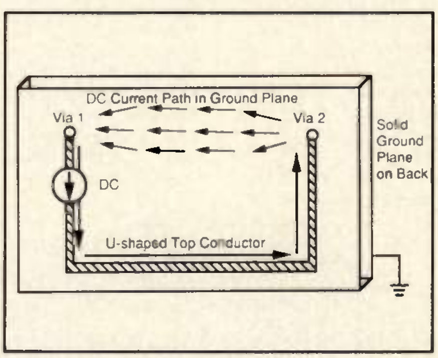


Figure 2. Typical current path for DC signals.

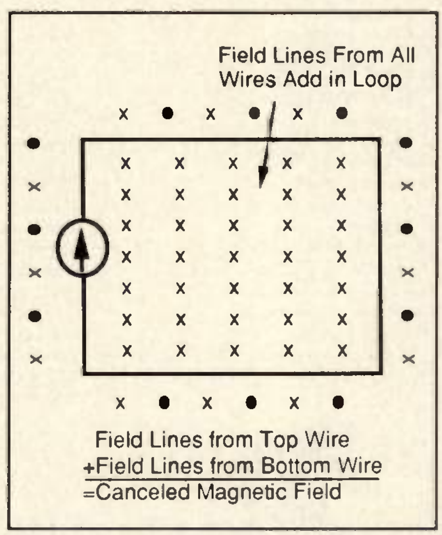


Figure 3. Magnetic field lines in inductor with right-hand rule.

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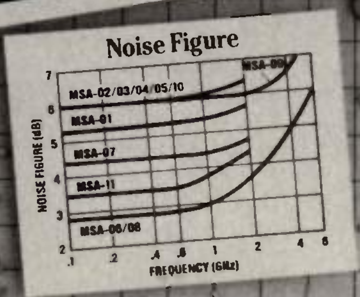

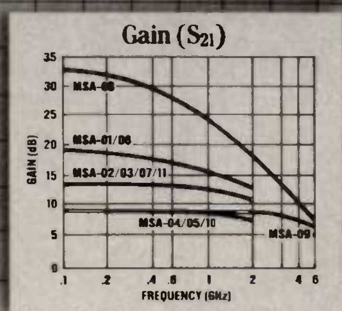

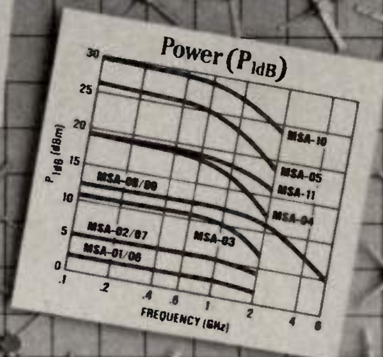

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short as possible. Designers should also make sure that the layout does not cut the ground plane under a critical path.

If the ground plane is cut so that path B can cross path A (Figure 5), inductance is introduced into the ground return of both signals because each return loop is made larger. Since path A conducts a high-frequency signal, an induced voltage drop will appear across the opening of the ground plane. The total voltage will equal L(di/dt). For the same reason, an AC current in path B will also cause ground plane noise.

For typical TTL and ECL signals, the resulting drop can be hundreds of millivolts. This can add several LSB errors to a high-speed 10 MHz, 12-bit A/D converter or even create errors in a lower resolution 20 MHz, 8-bit device. However, a simple fix to this problem is to install a wire directly across the cut in the ground plane, as shown in Figure 5.

Another problem with cutting the ground plane under a conductor is that it creates a time-varying magnetic field. These magnetic fields can pick up noise from transformers or other magnetic loops such as coils used in analog filters or relays.

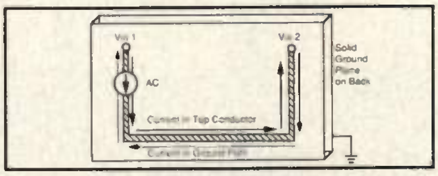

Figure 4. To follow the path of least impedance, AC current in ground plane traces the same path as the current in the top conductor.

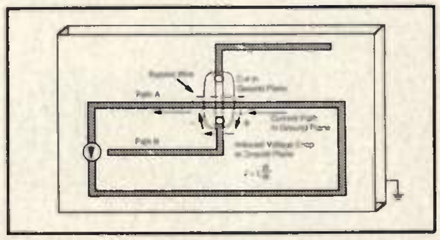

Figure 5. A cut in the ground plane can introduce voltage drops in the ground plane.

Power distribution can also suffer from poor ground management. Most designers think of power supply lines as low-impedance signal paths. In fact, a supply line without a ground plane beneath it can exhibit high impedance at frequencies as low as 1 MHz. Most designers choose to solve this problem by liberal use of 0.1 μF supply bypass capacitors. Although this is a good practice, if the problem is high impedance due to high inductance, then it could cause more problems than solutions. For instance, with a 0.1 μF capacitance and 30 nH inductance, the supply will sing at 3 MHz after every transient ($f=1/2\pi LC$). In general, to reduce noise levels, it is best to keep the inductance low with ground planes under supply lines and use bypass paths for transients.

About the Author

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A Reference-Cancelling Phase/Frequency Detector

By Dan Baker
Tektronix, Inc.

The prize-winning design presented here had its origin at RF Technology Expo 89 in Santa Clara, Calif. A paper presented during the conference mentioned the controversy over how to model the "sampling delay" of different phase detectors. This article describes a phase/frequency detector which uses the principle of SSB or complex mixing to cancel all but the desired signal component.

When modeling the sampling delay of different phase detectors, the Zero-Order-Hold (Z.O.H.) model shows more phase shift in the computer model than the real circuit exhibits. This "divider delay" or hold effect is only there if a true sample-and-hold phase detector is used. Since most phase-lock loops (PLLs) use digital (ECL, TTL, etc.) signals to drive the phase detector as opposed to sine waves driving linear, four-quadrant multipliers, phase information can only be conveyed on the transitions and, with some phase detectors, the PLL becomes essentially an impulse-sampled system.

Sampling by itself, however, does not cause any delay or lagging phase shift. The Z.O.H. model should thus not be included for most PLL designs, including ones using the 4044, 4046 type sequential phase/frequency detector. Since most PLLs have at least one, and often two, ideal integrators, some portion of the phase sample is held until the next sample and this does introduce excess phase shift. However, the phase shift would not be as much as the Z.O.H. model predicts since these integrators are not being set to the sample value in a single sample, as would be the case with a true sample-and-hold or Z.O.H.

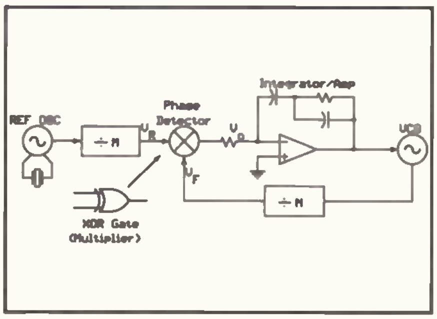


Figure 1(a). Basic PLL synthesizer as Type II, third-order loop.

By properly analyzing the effects of sampling, it should be possible to select the loop filter values and gain to design a phase-locked loop approaching the theoretical minimum settling time. In order to build a practical loop, it was necessary to get around the infamous "Catch 22" problem. Since the loop filter is needed to filter the reference frequency feedthrough as well as to control the loop dynamics, a phase/frequency detector was needed that had little, if any, output at the reference frequency or at any frequency except the difference between the feedback signal frequency and the reference frequency. When the loop is locked, only a DC component from the detector is desired.

Consideration of this characteristic spawned an idea for a phase/frequency detector to accomplish these goals. To verify that the idea would work, a prototype of the detector and a complete PLL was constructed. The object was two-fold: 1) To show that a Type II, third-order PLL could be optimized using sampling theory to essentially equal the minimum settling time theoretical limit (two samples for a Type II loop), and 2) By using this phase detector idea, to generate minimal reference frequency sidebands in the VCO output. The principles and concepts that led to the development of this new phase/frequency detector are discussed in the following sections.

Complex Mixing Principle

Figure 1(a) shows a common application of a PLL as a frequency synthesizer. The error amp and the VCO in this example act as essentially ideal integrators. This is an example of a Type II, third-order loop as discussed in Reference 1. The signals at the outputs of the dividers are usually binary logic levels such as ECL or TTL, and the multiplier can be implemented as an Exclusive-OR (XOR) gate. However, to understand how this multiplier acts as a phase detector, it is often easier to approximate these voltages as sine waves.

$$v_R(t) = \sin \omega t; \quad v_F(t) = \sin(\omega t + \theta)$$

Therefore,

$$v_o(t) = v_R(t)v_F(t) = (\tfrac{1}{2})[\cos\theta - \cos(2\omega t + \theta)]$$

This equation derives, for the sine approximation, the familiar result that two terms appear in the output of the multiplier. The first term is a DC component proportional to the cosine of the phase of the two signals. The second is the 2X reference frequency term that must be filtered by the integrator/amp in order to minimize VCO modulation and the resulting unwanted sidebands. For a practical TTL XOR gate, this component appears as a square wave at twice the reference frequency with a fundamental component of nearly 5 V peak-to-peak. The duty cycle of this square wave is linearly proportional to the phase of the inputs, and it is this component that pulls the loop into lock. The negative of the integral appears at the input of the VCO and is the negative feedback that maintains phase lock.

In order to make a fast, frequency-agile synthesizer, the loop bandwidth must be as wide as possible. As discussed before, however, the filter must attenuate the large square wave to minimize VCO sidebands. Typically, these opposing requirements are resolved by using a sequential phase/frequency detector or state-machine with a tri-state output. The output is in

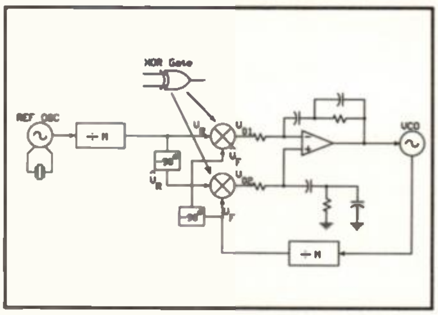


Figure 1(b). Complex mixing cancels 2X reference frequency.

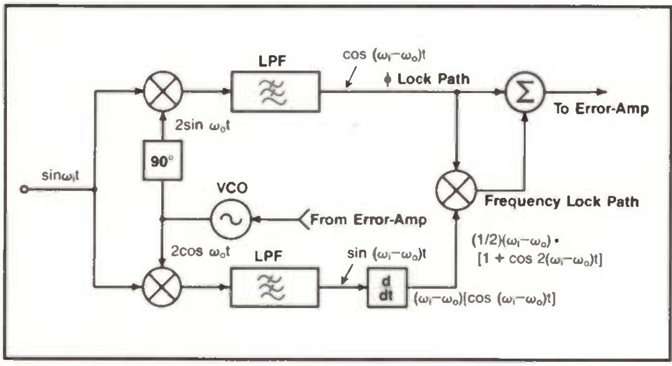


Figure 2(a). Quadricorrelator phase/frequency detector.

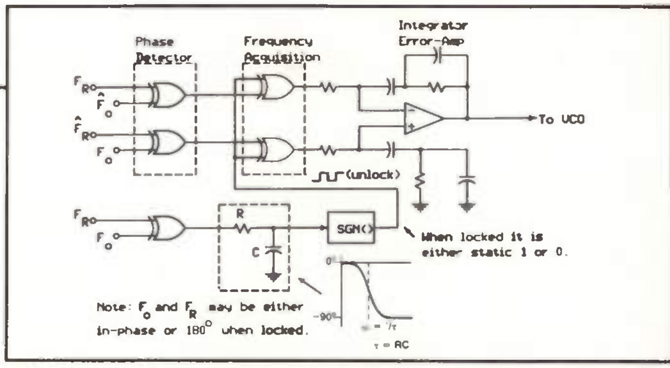


Figure 2(b). Complex phase det. with frequency det.

an open or high-Z state most of the time and, for small phase errors, the output is an impulse with a very small fundamental component at the reference frequency. The integrator/amp can then have a relatively wide bandwidth with minimal reference frequency attenuation. One problem with this sequential detector is that it has state memory. In some applications this is unacceptable, since if there is a missing or extra edge in one of the inputs, the VCO is pushed through 360° to regain phase lock.

Another problem is that the detector is an edge-triggered device and acts as a sampled phase detector with the sample rate equal to the reference frequency. In some applications this is a desired characteristic. In others, the feedback signal V_F already exists as a square wave and only half of the available phase samples are used, since both positive and negative zero crossings are not detected. The XOR gate discussed previously uses both zero crossings, and the effective sample rate is twice that of the sequential phase/frequency detector.

Figure 1(b) shows how, using the principle of complex mixing, the XOR gate function can be used as a phase detector and the 2X reference frequency cancelled out. Using the sinusoidal approximation again illustrates that the outputs of the two multipliers have the same components as before.

$v_R(t) = \cos \omega t \qquad v_F(t) = \cos (\omega t + \theta)$

$\hat{v}_R(t) = \sin \omega t \qquad \hat{v}_F(t) = \sin (\omega t + \theta)$

Therefore,

$$v_{o1}(t) = v_R(t)\hat{v}_F(t) = (1/2)[\sin \theta + \sin (2\omega t + \theta)]$$

$$v_{o2}(t) = \hat{v}_R(t)v_F(t) = (1/2)[-\sin \theta + \sin (2\omega t + \theta)]$$

Since the $(2\omega t + \theta)$ components cancel,

$$v_{o2}(t) - v_{o1}(t) = -\sin \theta$$

By using in-phase and quadrature inputs, however, the desired DC terms can be made to be of opposite polarity. These two outputs are effectively subtracted in the differential integrator/amp shown and the 2X reference frequency components cancel while the phase terms add. The negative polarity assures negative feedback to maintain phase lock.

An additional benefit is that any components common to both multiplier outputs, such as power-supply noise or correlated phase jitter, are cancelled as well. Yet another benefit may exist. To the extent that reference phase noise and internal phase noise are separated into uncorrelated in-phase and quadrature components, there may be a 3 dB improvement in the signal-to-noise ratio (S/N) of the detector. This is possible since the phase noise terms would add root-sum-square and the phase terms add arithmetically.

Figure 2(a) illustrates a famous method for phase and frequency detection invented back in the 1950s. As discussed by Gardner (2), this circuit takes advantage of quadrature mixing to drive a third multiplier in order to allow frequency lock outside the pull-in bandwidth of the phase-locked loop. This is especially advantageous in narrowband loop applications.

The circuit is best described by first considering operation after the loop is phase-locked. The upper multiplier acts as a phase detector. The difference frequency term becomes the cosine θ term discussed earlier, which is passed to the error amp to maintain phase lock. The lower multiplier output is differentiated and applied to the third multiplier. During phase lock the differentiator output is zero; therefore, the third multiplier output is zero.

When the loop is not phase-locked, the outputs are as labeled in Figure 2(a). The differentiator phase-shifts the lower frequency difference component to be in-phase with the upper frequency difference component. The third multiplier then acts as a squaring circuit, generating a positive DC component to push the VCO in a direction to reduce the frequency difference. If the frequency difference is negative ($\omega_i < \omega_o$), then the differentiator output is 180° out of phase with the upper multiplier frequency difference component. The third multiplier then creates a negative DC component to again push the VCO to minimize the frequency difference. When the frequency difference falls within the pull-in bandwidth, the loop phase-locks.

Figure 2(b) illustrates how this concept can be applied to the XOR gate reference frequency cancelling phase detector discussed earlier. This would allow the complex mixing idea to be used in narrowband applications without additional frequency acquisition circuitry.

The outputs of the two XOR gates that constitute the new phase detector are each applied to another XOR gate as shown in Figure 2(b). These gates act as switchable inverters to allow the phase detector output polarities to be reversed before they are subtracted in the difference amplifier.

The XOR gate at the bottom of Figure 2(b) serves the function of the lower multiplier in Figure 2(a). When the loop is phase-locked ($F_R = F_o$), the inputs of this gate are either in phase or 180° out of phase. The XOR gate output is then in either a steady low state or a steady high state. This signal is lowpass-filtered by the RC phase-shift network and applied to the sgn( ) box. The Quadricorrelator uses a differentiator which produces a +90° shift at all frequencies, while the lowpass filter shifts the phase −90° only at frequencies above the corner frequency. This requires that the polarity of the sgn( ) output be inverted from that required with the differentiator. In addition, the frequency lock action is only fully effective at frequencies above the lowpass filter cutoff.

The sgn( ) function acts like a comparator and outputs only a 1 or a 0 logic level, depending on the polarity of the signal at its input. In the phase-locked case, its output is a steady 1 or 0 independent of the noise at the output of the XOR gate. When the loop is unlocked, its output converts the frequency difference component at the output of the RC phase-shift network to a square wave.

The RC phase-shift network provides −90° of phase shift for frequencies

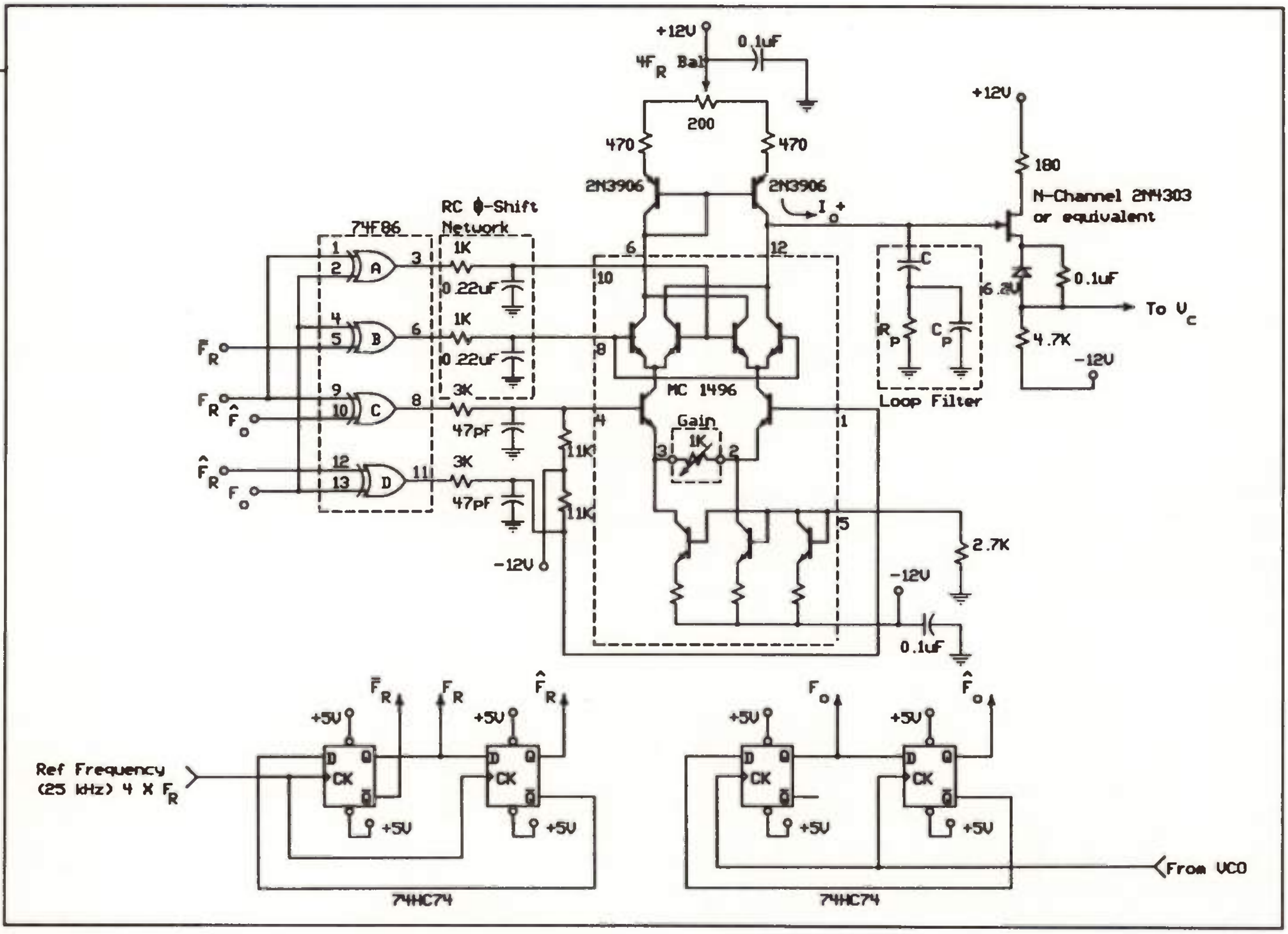


Figure 3. Complex mixing phase/frequency detector.

much above its corner frequency. Its time constant should then be selected to place the corner frequency below the pull-in range of the PLL. (This turns out to be fairly non-critical for most applications.)

When the loop is unlocked and the frequency difference is outside the PLL pull-in range, the frequency difference square wave at the output of the sgn( ) circuit is then either in phase with the SSB frequency difference term from the phase detector, or it is 180° out of phase, depending on the polarity of the frequency difference. The corresponding DC component at the input of the difference amp then forces the VCO inside the pull-in range in the Quadricorrelator of Figure 2(a). Since the RC phase-shift network is DC-coupled, the loop can lock the VCO either in phase or 180 ° out of phase with the reference.

A New Phase/Frequency Detector

Figure 3 shows the final schematic of the new detector concept. XOR gates A and B form the function of the lower gate in Figure 2(b). Using two gates produces complementary outputs to drive the upper stage of the MC1496 multiplier. The upper stage provides the sgn( ) function and is operated as a differentially driven switch without the required voltage reference for single-ended drive. In this way the proper switching threshold for small-signal detection is maintained independent of the actual logic levels. Although not shown in this schematic, some of the capacitance of the RC network could be connected differentially across pins 8 and 10 to improve frequency lock with low-tolerance capacitors.

XOR gates C and D act as the complex mixing phase detector discussed earlier. A balance adjustment was added to allow the 2X reference frequency component to be dipped. As discussed in the results section, this allowed over 60 dB of common-mode rejection to be obtained.

The lower stage of the MC1496 acts as the difference amp, and a 1K resistor between the emitters sets the detector gain. The resistor on pin 5 sets the bias current for the MC1496.

An active current mirror converts the differential output current of the MC1496 to single-ended and this is applied to the loop filter. Since this current output approaches an ideal current source, the large DC gain of the integrating error amplifier is accomplished here. In effect, this combines the function of the frequency detection multiplier and integrating differential error amplifier into one IC. To maintain the high low-frequency gain, the signal is then buffered by a FET follower and level shifter to drive the VCO of the test circuit. In applications where the VCO input is a DC-isolated varactor diode, the FET follower may not be needed.

The quadrature signals needed to drive the new detector were obtained by using two Johnson counters. The counters act as divide-by-four stages preserving the VCO phase variations without any additional delay. This is an important point, and will be discussed in the next section. For example, in the test circuit the VCO runs at 25 kHz and the counter outputs are 6.25 kHz square waves, making the reference frequency 6.25 kHz. The effective sample rate is 25 kHz, however, since four phase transitions occur in the detector for each

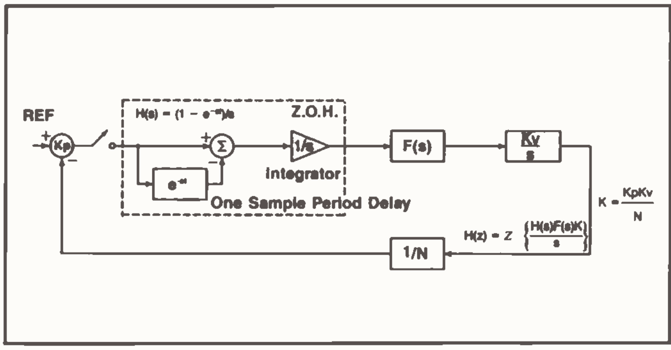


Figure 4. Model of a PLL incorporating a Zero Order Hold.

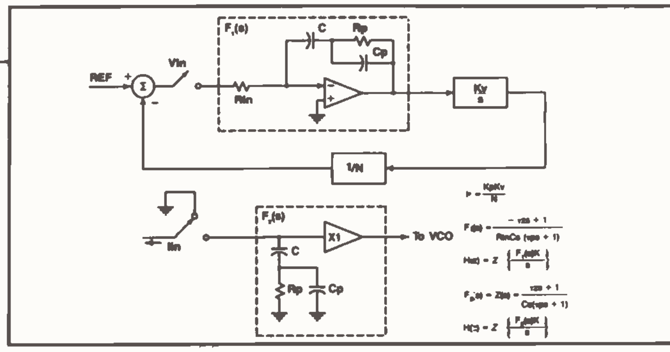


Figure 5. Voltage (top) and current impulse (bottom) output.

cycle of the reference frequency. This will allow the test PLL bandwidth to equal or even to exceed the reference frequency. It should be noted that if other phase-shift methods are employed, the reduction in sample rate or additional delay will reduce the maximum attainable bandwidth for a given reference frequency.

PLL Design As a Sampled System

As discussed earlier, the sample frequency for this new phase/frequency detector is 25 kHz. The phase difference between the reference and feedback signal is presented to the loop filter as current impulses for small phase errors. Since the loop is of Type II, it contains two ideal integrators. It may seem then that this would constitute a Zero Order Hold. There have been recent articles which advocate the inclusion of the Z.O.H. transfer function into the loop. Figure 4 presents a model of a PLL incorporating a Z.O.H.

To act as a true Z.O.H., the integrator must be ideal and fully charged on a single impulse sample. The output should remain at the previous value since successive impulses are cancelled by the one-sample delay path. This does not occur on most phase detectors/loops in common use, since a true S/H or DAC is not used. As shown in Appendix A, cascading the Z.O.H. model with the loop filter predicts too much group delay or excess phase. In fact, since the phase shift increases linearly from 0 to 25 kHz, the group delay of the Z.O.H. is a constant 20 μsec over this range.

An accurate model of most loops is shown in Figure 5. This diagram illustrates two types of phase detectors. The first example is a voltage output much like the popular 4044 sequential phase detector. For small phase errors, the output can be considered as impulses with areas proportional to the phase error of the inputs. The second example behaves much the same, except that the output is current impulses into a passive loop filter. Only an X1 buffer may then be needed to drive the VCO input. This is the preferred model for the new, complex-mixing phase detector.

The open-loop transfer functions for both examples are given in Figure 5. To analyze these loops as a sampled system, the Z transform should be taken as shown.

Appendix B details the mathematical analysis of the test PLL incorporating the new phase detector. The analysis was done using MathCad. This program is a generic math package and not a PLL modeling program. All the equations and constants for making the plots shown had to be derived for the analysis.

In Part A of Appendix B, the continuous-time transfer function of s and the sampled transfer function of z are defined for the Type II, third-order PLL. Both were analyzed in order to illustrate the inaccuracies of the continuous-time analysis near the sample frequency.

Part B compares the magnitude response and Part C compares the phase response of the two analysis methods. Note that the magnitude response increases and the phase accumulates more rapidly in the sampled system. For example, the phase margin at 5 kHz is 30 ° compared with 40 ° for the continuous-time equation. Still, this is much better than the −6° that the Z.O.H. model would show since it would add an additional 36 ° at this frequency.

Part D shows the root locus for various loop gains in both the S-plane and the Z-plane. Note that the optimum gain for the Z-plane is about half that which is predicted for the continuous-time analysis.

Part E compares the impulse and step response. Note that the continuous-time response is underdamped since the gain is 6 dB lower than the predicted optimum. If the gain were increased however, the real, sampled system would go unstable with only a 3 dB increase, as shown in the Z-plane root locus. Note also that by using the gain predicted from the Z-plane root locus, the real, sampled system completely settles in two sample periods, attaining the theoretical minimum.

Experimental Results

Measured/Calculated gains:
Optimum pole/zero
Fz = 3400 Hz
Fp = 25000 Hz
Kp = (2•3)/(π•530) amps/radian
Kv = 69K radians/V-sec
N = 4

Calculated values:
C = Kp•Kv/N•K amp-sec/V (Farads)
K = 62•10^7
C = 0.1 μF
Cp = 0.015 μF
Rp = 420 ohms

The values shown above were inserted into the loop filter and the gain and balance controls tweaked for best step response. The gain was only tweaked about 15 percent from the predicted value to optimize transient response. Figure 6 shows the photographs of the measured data from the test PLL.

Figure 6(a) shows the response to a frequency step. As shown in this waveform photo, the loop has settled in essentially 80 μsec or two samples. This validates the computer predictions and verifies that a practical Type II loop can settle in two sample periods. Figure 6(b) shows the VCO input and verifies that less than 4 mV of reference frequency component is present. The loop filter affords little attenuation to this component and this low value is due to the good balance of the MC1496. Figure 6(c) is an SA photo of the VCO sidebands at the 2X reference frequency or 12.5 kHz. A 3 mV component at this frequency with a VCO gain of 69,000 rad/V-sec should produce sidebands down 68 dB from the carrier. The measured sidebands are near this value as seen in the photo.

Figure 6(d) shows the SA display when white noise is added to the VCO input. The upper trace is with the VCO open loop and the lower trace is with the loop locked. Note the noise bandwidth of the loop is about 3 kHz. Figures 6(e) and 6(f) also compare the open loop VCO noise with closed loop for 500 Hz/div and 5 kHz/div, respectively. Fig-

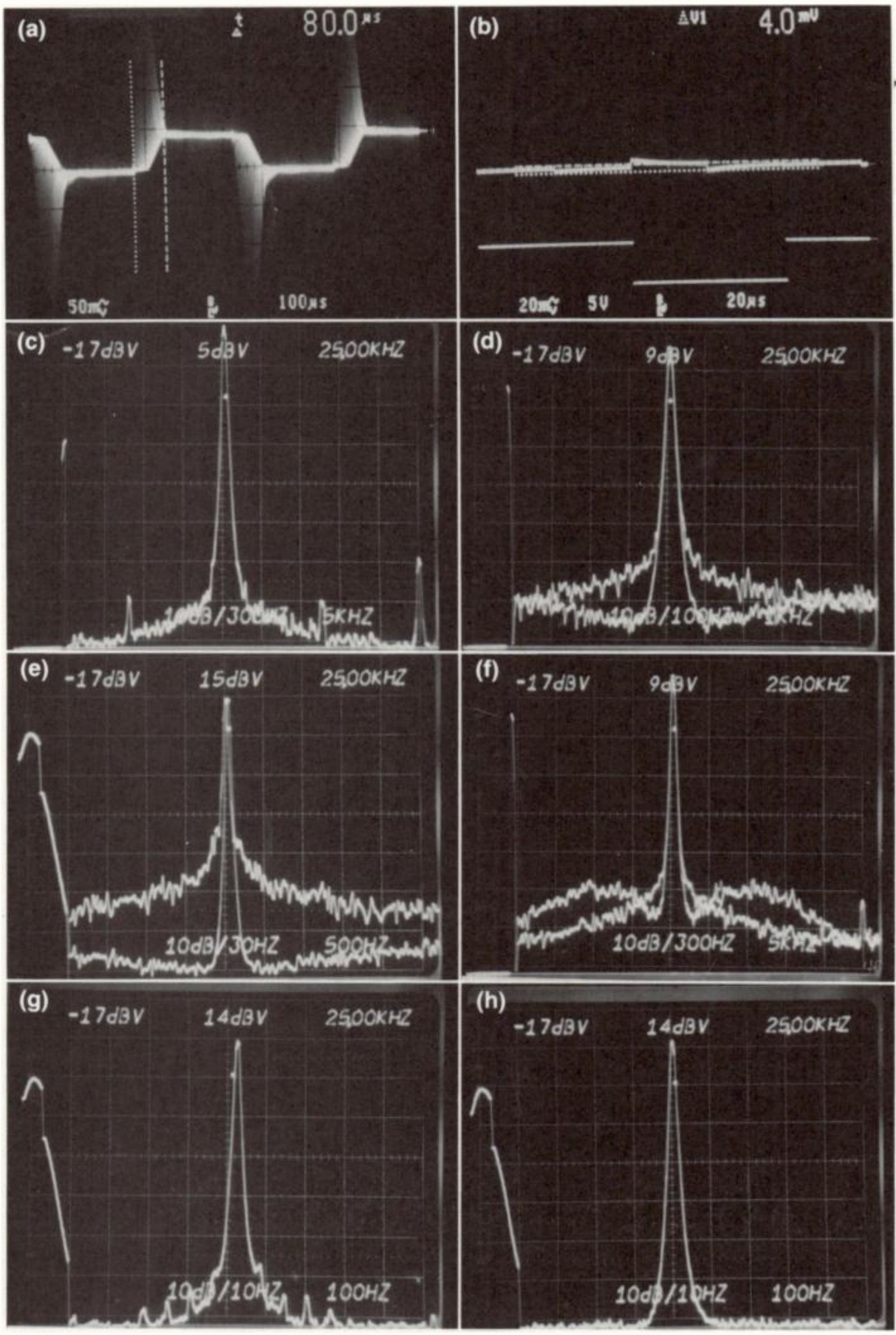


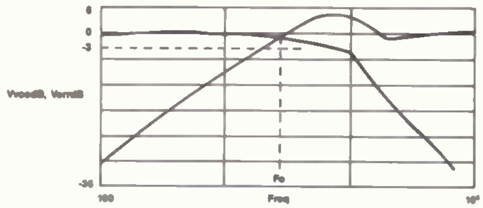
Figure 6. Measured data for the test PLL.

Freq	Vvco	Verr
100	3	100
200	7	100
600	22	100
1000	36	100
2000	73	96
6000	165	77
10000	160	60
20000	90	17
40000	97	6

i = 1...9

$$VvcodB = 20 \log \left[\frac{Vvco}{100}\right]$$

$$VerrdB = 20 \log \left[\frac{Verr}{100}\right]$$



Fo = Crossover Frequency = 2.6 kHz
Fn = Noise Bandwidth = 1.9 kHz
Fc = Control Bandwidth = 6.2 kHz

Figure 7. Bode plot of small-signal closed-loop characteristics.

ures 6(g) and 6(h) compare the VCO spectrum close in to the carrier with the VCO open loop in Figure 6(g) and closed loop in Figure 6(h). Note the 60 Hz sidebands and phase noise are removed by the PLL.

Figure 7 shows the measured Bode plots of both the control response and the response to perturbations within the loop. Note the control bandwidth is essentially equal to the reference frequency of 6.25 kHz, yielding a bandwidth to reference frequency ratio of 1. rf

References

1. A. Przedpelski, "PLL Primer," *RF Design*, July/August 1983.
2. F. Gardner, *Phaselock Techniques*, p. 87.
3. A portion of this design includes circuitry covered by U.S. Patent #4,072,909, February 7, 1978 (Zenith Radio Corporation).

About the Author

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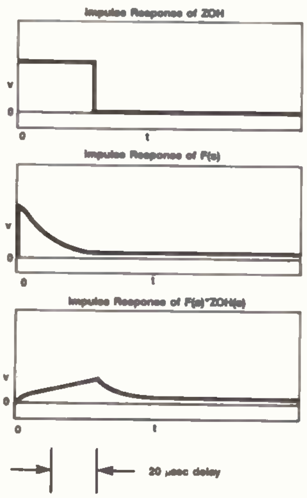
$Fr = 25 \times 10^3$
$T = 1/Fr$
$Wz = 3400$ Hz
$Wp = 25000$ Hz
$\tau z = 1/Wz$
$\tau p = 1/Wp$
$C = 0.1 \times 10^{-6}$
$R = 1 \times 10^3$

Transfer function for Z.O.H.:

$$Hzoh(s) = [(1-e^{-sT})/s](1/T)$$

Transfer function for filter:

$$F(s) = \frac{\tau zs + 1}{RCs\,(\tau ps + 1)}$$



Appendix A. Z.O.H. effects.

A. Open-Loop Transfer Function

1) Continuous System

$\tau z = 1/Wz$
$\tau p = 1/Wp$

For the open-loop function, $Gol\,(s) = \dfrac{K(s\tau z + 1)}{s^2(s\tau p + 1)}$

2) Sampled System

$z = \exp[j\omega T]$
$A = \exp[-T/\tau p]$

For the open-loop function,

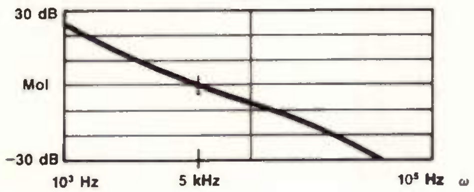
$$Gsol\,(z) = \frac{(T)(K)(z)}{(z-1)^2} \bullet \frac{(z-1)(1-A)(\tau z-\tau p)+T(z-A)}{(z-A)}$$

gain constant $K = 6.2 \times 10^8$
zero frequency $Wz/(2\pi) = 3.4 \times 10^3$
pole frequency $Wp/(2\pi) = 2.5 \times 10^4$

B. Magnitude Response

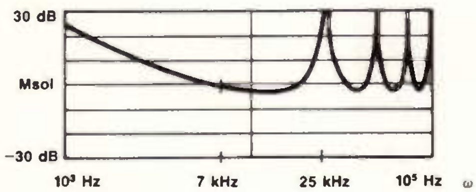
1) Continuous System

$M = |Gol[j\omega]|$
$Mol = 20\log[M]$ (in dB)



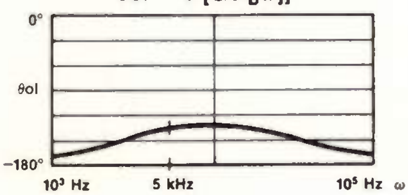
2) Sampled System

$Ms = |Gol[z]|$
$Msol = 20\log[Ms]$ (in dB)

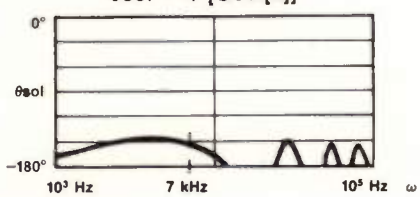


C. Phase Response

$\theta ol = \Phi\,[Gol[j\omega]]$



$\theta sol = \Phi\,[Gsol[z]]$

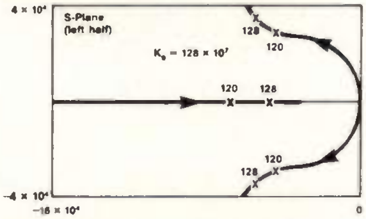


$\phi(z) = \tan^{-1}(Im(z)/Re(z))$

D. Root Locus Analysis

1) For a continuous system, define the characteristic equation as:

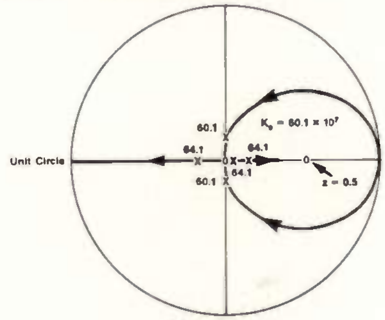
$CE(s) = (s\tau z + 1)(K) + s^2(s\tau p + 1)$



2) For a sampled system, define the characteristic equation as:

$CE(z) = [(T)(z)(K)] \bullet$
$[(z-1)(1-A)(\tau z-\tau p) +$
$\theta ol = \phi\ (z-1)^2(z-A)]$

zeros at: $Z1 = 0.5$, $Z2 = 0$

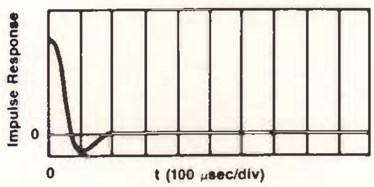


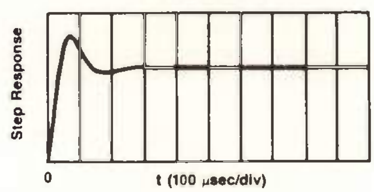
Z-Plane Root Locus

E. Transient Analysis

1) Continuous System

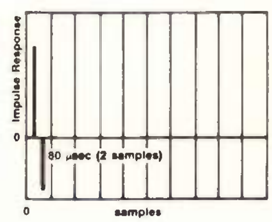
defined gain: $K = 62 \times 10^7$
closed-loop equation: $CL = \dfrac{Gol[j\omega]}{1 + Gol[j\omega]}$

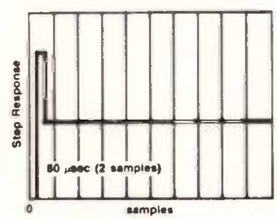




2) Sampled System

$z = \exp[j\omega T]$
closed-loop equation: $CLs = \dfrac{Gsol[z]}{1 + Gsol[z]}$





Appendix B. Third-order, Type II phase-locked loop analysis.

Lowpass Connector Filters: An Overview

By Andrew Dawson
AMP Incorporated

In the past, filtering options in input/ output (I/O) connectors were limited to tubular-Pi or feed-through constructions and were available only in limited capacitance ranges. Now, with recent advances in ceramic and ferrite processes, various filtering performance levels can be obtained in the most popular connector configurations. Single structure devices, combining ceramic and ferrite properties, allow a wide selection of performance and cost levels.

Predicting susceptibility or emission problems in a system and determining the level of filtering required can be simplified by increasing or decreasing the rate of attenuation (dB/decade) or by shifting the break or cutoff frequency on the system's I/O connections. Higher voltage levels can also be obtained for power or electromagnetic pulse (EMP) requirements for specific pins or for the entire connector.

Filter Selection Criteria

The most important element in the selection of any lowpass filter connector is maximum capacitance. This provides the capability to meet emission or susceptibility requirements. Calculations should be made to determine proper pulse fidelity and other electrical parameters. The more capacitance a system can handle, the more the filtering attenuation can increase for a particular frequency. After the upper capacitance limit is defined, the amount of attenuation required for a frequency point or over a frequency range is determined. A minimum attenuation level is specified for problem or susceptible frequencies. Filters are tested and specified to MIL-STD-220 at +25 degrees C (50 ohm source and load impedance). Filter attenuation curves provided by the manufacturer may need to be modified, depending on the source and load impedance of the system.

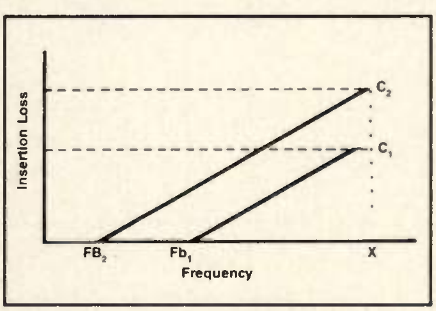


Figure 1. Plot of frequency vs. insertion loss shows Fb_2>Fb_1 when C_2>C_1.

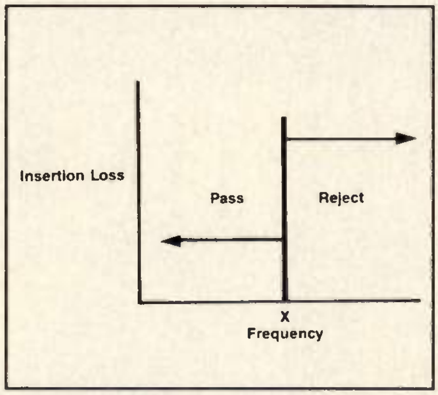


Figure 2. Ideal lowpass filter.

Break Frequency

By determining the 3 dB point or "break frequency" of a lowpass filter, many calculations on insertion loss can be simplified. The break frequency of the filter is defined as:

$$F_B = \frac{R_S + R_L}{R_S R_L (2\pi C)}$$

where R_S is the source impedance, R_L is the load impedance, and C is the filter capacitance. With R_S and R_L defined to be 50 ohms per MIL-STD-220, the above formula can be simplified to:

$$F_B = \frac{0.00637}{C}$$

Since capacitance is inversely proportional to break frequency for a given filter, the higher the capacitance, the lower the break frequency and vice versa. Figure 1 shows that for any lowpass filter element, the attenuation for frequency X can be increased simply by increasing the capacitance of the filter. In general, doubling the capacitance will increase the insertion loss at a given frequency by 6 dB. Providing more attenuation by adding capacitance is an easy solution to many emission or susceptibility problems.

Lowpass Filter Connector

With capacitance and attenuation parameters defined, the filter selection process is narrowed to examining the tradeoffs with each filter type and then choosing a filter that most closely meets the needs of the application. The primary purpose of a filter is to maximize attenuation of unwanted signals.

Figure 2 shows an ideal filter attenuation curve, passing all frequencies below point X and rejecting (providing infinite insertion loss at) all frequencies above point X. The filter has a vertical insertion loss curve. The steeper the insertion loss curve of a filter connector, the better the filter will perform. Obviously, tradeoffs are required when trying to maximize the insertion loss. Thus, many filter technologies are available.

Capacitive Filtering

The simplest ceramic feed-through tubular capacitor configuration is made with a hollow ceramic tube that is plated

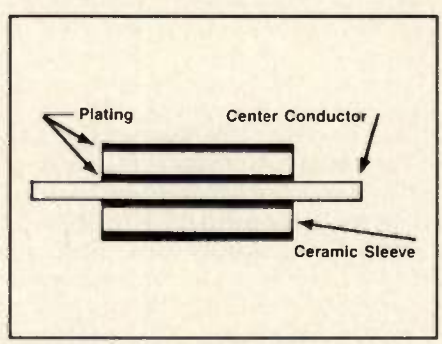


Figure 3. Ceramic feed-through capacitor.

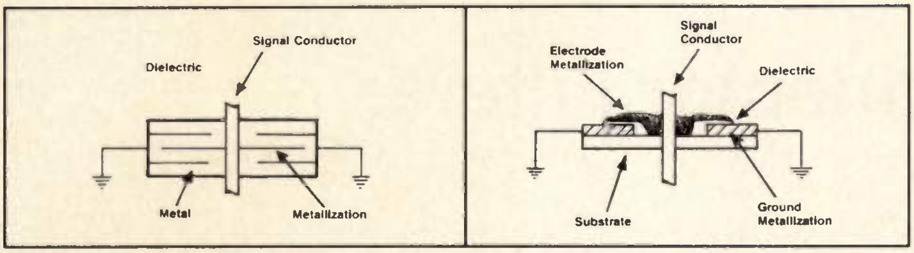


Figure 4. Multilayer (left) and thick film (right) planar capacitors.

on the inside and outside diameters (Figure 3). In most cases, the distance between the inner and outer plates is dependent on the mechanical construction of the connector (pin diameter and centerline spacing).

Capacitance is directly proportional to the area of the plates and the dielectric constant of the ceramic material. Thus, by changing sleeve length and material, capacitances ranging from 100 pF to 10,000 pF are available. Feed-through tubular sleeves are capable of withstanding voltages up to 1500 VDC due to the relatively large spacing between plates. A capacitively filtered connector will typically yield an attenuation (insertion loss slope) of 20 dB/decade and will reach and maintain a minimum value of 50 dB.

A relatively new type of capacitor, the planar type, is available in either single capacitors or arrays. They are either multilayer or thick film structures, as shown in Figures 4(a) and 4(b). In a multilayer type, several layers of dielectric, each covered with a conductor to form the capacitor plate, are stacked together into a single structure. The conductive layers are alternately attached to either an outside or inside metallization. Inside metallization is soldered to a conductor running through the center of the capacitor; outer metallization is soldered to ground.

In a thick film array, several layers of material are deposited on a ceramic substrate to form a monolithic capacitor. The first metallization layer serves as the ground plane. A dielectric is deposited over the metallization. The final layer is metallization to form the other electrode

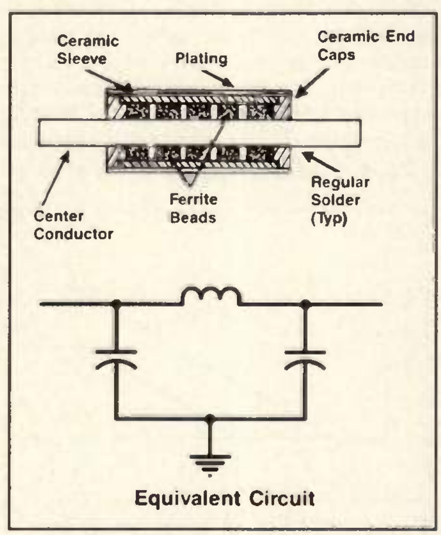


Figure 5. Typical lumped-element filter.

of the capacitor. An overcoating may also be added to protect against oxidizing of the metallization. The upper level of metal is soldered to the conductor pin running through the capacitor; the lower level is soldered to ground.

The relatively small planar arrays and their range of available capacitance values (up to 5 F) offer distinct advantages over feed-through capacitors. To apply, connector pins are soldered to the plated-through holes and the ground plane of the array is soldered to the metal shell of the conductor, which is then grounded.

Because of the high capacitances available in multilayer planar arrays, these arrays can provide extremely high attenuation at relatively low frequencies. As a result, planar arrays are becoming increasingly popular for applications that can handle a relatively fair amount of capacitance.

Lumped-Element Filters

Lumped-element reflective filters and distributed-element absorptive filters are higher performing filters. Lumped-element filters, available in T, L and Pi configurations, consist of ferrite beads inside a ceramic sleeve or alongside a planar array. Choice of the filter type depends on the circuit and insertion loss needs.

The L filter, for unbalanced circuits, provides about 40 dB loss per decade. To achieve proper performance, the inductor end of the filter must face the low impedance. The T filter, for low-impedance circuits, has an inductor facing the source and load. It provides an insertion loss of about 60 dB per decade. The Pi filter, for high-impedance circuits, also provides a loss of 60 dB per decade. Figure 5 shows the construction of a typical feed-through Pi filter.

Reflective filters have three main drawbacks. First, since they operate reflectively, they do not eliminate the electromagnetic interference (EMI) from the system. They simply shunt it to another part of the system, where it may or may not be acceptable. While the ferrite inductor is lossy enough to provide some dissipation, the action is still mainly reflective.

Also, the specified insertion loss (tested to MIL-STD-220) may not be realized in application. In a Pi filter, for example, as the circuit source or load impedance drops below 50 ohms, the insertion loss also drops. Conversely, as the circuit impedance rises above 50

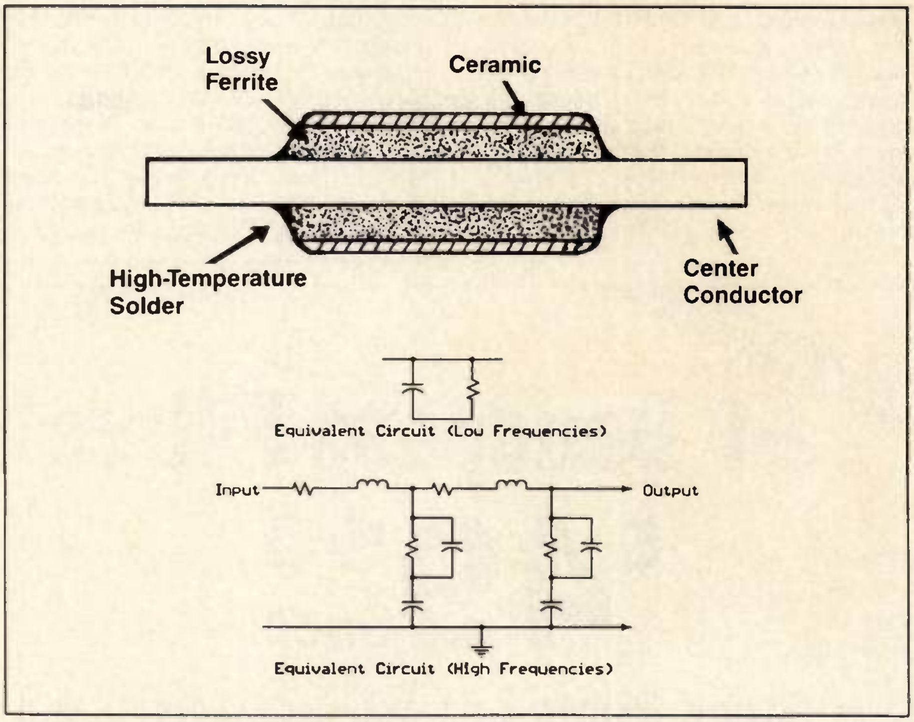


Figure 6. Distributed-element filter.

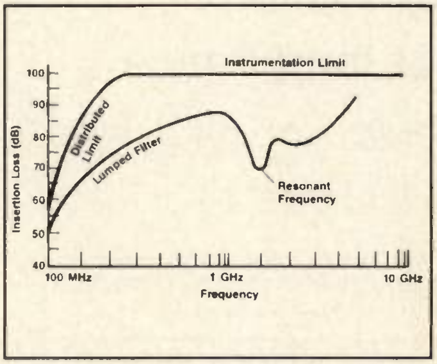


Figure 7. Response curves for distributed-element and lumped-element filters over a frequency range of 100 MHz to 10 GHz.

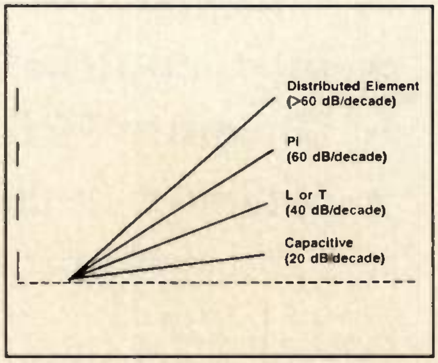


Figure 8. Insertion loss curves of various types of filters with the same capacitance (or the same break frequency).

Topic	Capacitive		L or T	Pi Type	Distributed Element
	Tubular	Planar			
Maximum Insertion Loss*	50 dB	50 dB	70 dB	>70 dB	>100 dB
Slope*	20 dB/decade	20 dB/decade	35 dB/decade	50 dB/decade	>60 dB/decade
DC Working Voltage	>500 V	500 V	500 V	500 V	250 V
Insulating Resistance	5 GΩ	5 GΩ	5 GΩ	5 GΩ	500 MΩ
Capacitance Range	100 to 10,000 pF	100 pF to 5 μF	100 pF to 5 μF	100 to 10,000 pF	100 to 10,000 pF
Operating Voltage DC	>200 V	200 V	200 V	200 V	100 V
*Testing to MIL-STD-200.					

Table 1. General filter performance in connectors.

ohms, the insertion loss rises as well. Unfortunately, the circuit impedances are often unknown, and they also change with varying loads and temperatures. Large variations in impedance result in apparently erratic filter performance. Nevertheless, if the filter provides a sufficient margin in its loss characteristics, changes in performance may not matter.

The third drawback to reflective filters is a drop in the insertion loss curve because of resonance of the discrete components. In some cases where the filter sees an inductive source or load, the Pi section filter can cancel the original source-to-load mismatch to present a net insertion gain over part of the stopband. The filter acts as a matching network that provides greater voltage to the load than existed before the filter was inserted in the circuit.

Distributed-Element Filters

Distributed-element filters are more effective at high frequencies (Figure 6). The filter sees a sleeve of ferrite surrounded by a barium titanate ceramic. During manufacturing, the ceramic is sintered to the ferrite to form a one-piece sleeve which is then plated inside and out. The ferrite is a compound specially formulated to be both conductive and absorptive or lossy.

The one-piece construction distributes inductance and capacitance along the length of the sleeve. The ferrite provides the inductance and the ceramic provides the dielectric to form a capacitor between the ferrite/center conductor and the ground in which the filter is mounted.

At high frequencies the filter is molded as a lossy coaxial transmission line. The capacitance passes the high frequencies but the influence of the lossy ferrite absorbs the energy in the process. While the filter still reflects some energy (about 12 dB in a 50 ohm system), the remainder of the insertion loss results from absorption and dissipation. The energy is thus eliminated rather than simply directed elsewhere in the circuit. Figure 7 shows response curves for two representative filters.

Summary

A comparison of the insertion loss curves of various types of filters with the same capacitance (or break frequency) is shown in Figure 8. Table 1 provides a general overview of the electrical parameters of the various filter connector technologies currently offered.

It should also be pointed out that the filter technologies can sometimes be intermixed in the same connector. For example, an application might call for mostly low-level logic signals and a few higher voltage power lines. To accomplish this, one could incorporate the maximum filtering via distributed-element filters on the sensitive logic signals and tubular feed-through capacitors on the high voltage lines. Although high insertion loss is not required, higher working voltage levels are critical.

About the Author

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RF Propagation in Buildings

The Results of an Indoor Radio Wave Propagation Study Are Presented

By Dr. T. Koryu Ishii
Marquette University

In spite of growing interest in wireless radio frequency data communications and remote controls inside large buildings (15) and residential homes (1-3), very little literature on the propagation characteristics of radio frequency electromagnetic waves indoors is available. Propagation of radio waves in a building, a thick jungle, or even the congested downtown of a large city is generally different from radiation in free space. For indoor propagation, dissipation, dispersion, scattering and absorption of the electromagnetic waves will occur due to the floor, wall, ceiling, furniture, equipment, utility facilities and people inside the building. The structure of the building, obstacles and objects inside the building determine the propagation characteristics of the radio waves under a given frequency. If the operating frequency is changed, the propagation characteristics change in a way unique to the given building or media. In this article the results of an indoor radio wave propagation study are presented and methods of characterizing a building in terms of RF propagation are introduced.

Norton's propagation equation considers direct waves, reflected waves and surface waves over a smooth sphere (earth) of finite reflection coefficient, and conductivity (4-6). Bullington simplified Norton's equation by separating the direct, reflected and surface waves (7-8). Reudink further simplified the propagation equation for the radio waves propagating along the earth in a direction almost parallel to the surface of the earth (9). This is shown in equation 1:

$$P_r = P_t\, g_t\, g_r \left(\frac{h_t\, h_r}{d^2}\right)^2 \qquad (1)$$

where,

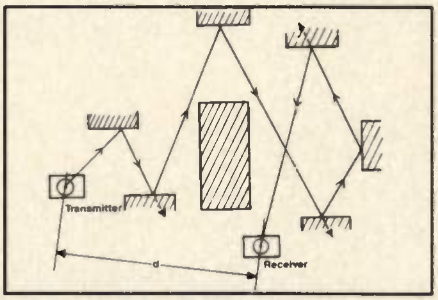

Figure 1. An illustration of electromagnetic wave propagation path and additional dispersion and dissipation for propagation in congested media.

P_t is the transmitted power;
g_t is the power gain of the transmitter antenna;
g_r is the power gain of the receiving antenna;
h_t is the elevation of the mounting location of the transmitter antenna; and,
h_r is the elevation of the mounting location of the receiver antenna.

For a fixed propagation system with given transmitter power, Reudink's equation can be rewritten as:

$$P_r = K_o \cdot \frac{1}{d^4} \qquad (2)$$

where K_o is a proportionality parameter. This relates to equation 3:

$$K_o = P_t\, g_t\, g_r\, (h_t\, h_r)^2 \qquad (3)$$

Reudink's equation was found to be useful for outdoor propagation (9-13), yet additional consideration was needed to accurately represent indoor propagation or propagation in any other congested media. For indoor electromagnetic wave propagation, in addition to the reflections from the ground, there is a loss due to dispersion and dissipation of radio waves through walls, equipment, furniture, curtains, doors, ceilings, windows, wall-and-room decorations and people in buildings. This is illustrated symbolically in Figure 1. This situation resembles that of electrons drifting in solids or of photons interacting with the lattice of solid material (14). The electromagnetic wave dispersion and dissipation along the propagation path are random and non-periodic. Therefore, these are handled statistically. The median power term, F, due to the dispersion and dissipation loss for the propagation distance d, is assumed to be:

$$F = e^{-\alpha d} \qquad (4)$$

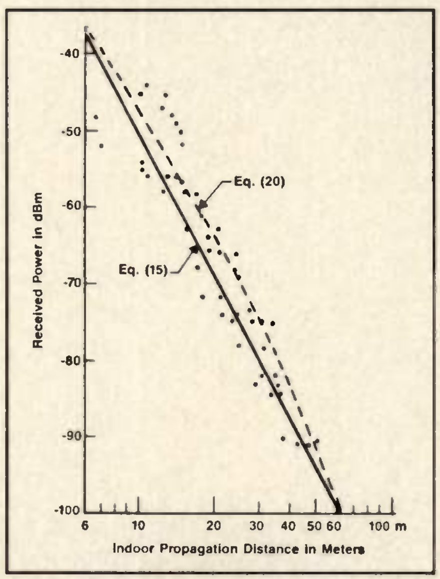

Figure 2. Experimental and theoretical results of indoor electromagnetic wave propagation at 49.83 MHz.

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where α is a combination of statistically averaged dissipation and dispersion constants per unit length of propagation due to random and finite reflections and transmission by random obstacles indoors. Therefore, it is considered to be a function of operating frequencies and the building's configuration and content. By combining equation 2 and equation 4, for the indoor electromagnetic wave propagation:

$$P_r = K \frac{1}{d^4 e^{\alpha d}} \tag{5}$$

where K is a proportionality parameter. When the transmitter and receiver are specified, the locations are fixed, the transmitter power is given, and K is set as a constant, the propagation characteristic is characterized by the loss number α at the operating frequency. When frequency is given, the loss number α is determined by measurement. If the received power at d_o from the transmitter is P_{ro} and the received power at d from the transmitter is P_r, then:

$$\frac{P_r}{P_{ro}} = \left(\frac{d_o}{d}\right)^4 e^{-\alpha d}\left(\frac{d_o}{d} - 1\right) \tag{6}$$

From equation 6:

$$\alpha = \frac{\left[\frac{P_r}{P_{ro}}\right]_{dB} - 40 \log \frac{d_o}{d}}{4343 d \left(\frac{d_o}{d} - 1\right)} \tag{7}$$

For the practice of RF wave propagation study, d_o is taken as the edge of the direct transmission range or near zone range and d is in the far zone. Theoretically, both d_o and d can be any distance in the far zone.

Frequency Characteristics of Loss Number

After numerous measurements, it was found that for RF wave propagation in congested media, the frequency dependency of the loss number can be described by the function shown below.

$$\alpha(f) = \frac{\alpha_c}{1 + (f/f_c)^{N_c}} \tag{8}$$

where, α_c is the loss number at the lowest frequency investigated;
N_c is 1/10 of the dB slope (10 log) decay of $\alpha(f)/\alpha_c$ per decade of operating frequency; and,
f_c is an operating frequency at the intersection of the low- and high-

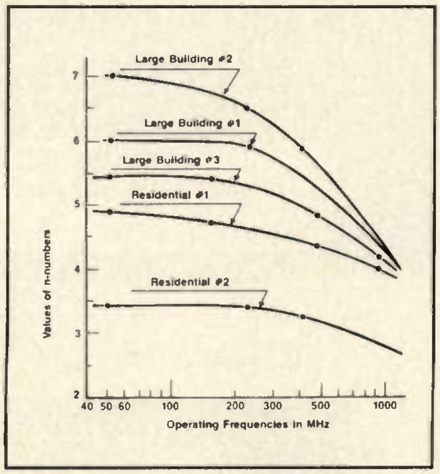

Figure 3. Frequency characteristics of n-numbers.

frequency asymptote on dB plot (10 log) of normalized loss number ($\alpha(f)/\alpha_c$).

Further Approximations

After numerous RF propagation wave measurements indoors, it was found experimentally that equation 5 can be further simplified:

$$P_r = K\frac{1}{d^n} \qquad (9)$$

If equations 5 and 9 are equivalent, then by solving for α,

$$\alpha = \frac{n-4}{0.4343} \cdot \frac{\log d}{d} \qquad (10)$$

or solving for n,

$$n = 0.4343\alpha \frac{d}{\log d} + 4 \qquad (11)$$

Following a similar approach with equations 6 and 7:

$$\frac{P_r}{P_{ro}} = \left(\frac{d_o}{d}\right)^n \qquad (12)$$

$$n = \frac{\left[\frac{P_r}{P_{ro}}\right] dB}{10 \log \frac{d_o}{d}} \qquad (13)$$

Frequency Characteristics of n-Number

Using a similar approach, the frequency characteristics of indoor RF propagation are characterized by:

$$n(f) = \frac{n_c}{1 + (f/f_c)^{N_c}} \qquad (14)$$

where, n_c is the n-number at the lowest operating frequency investigated;
N_c is the slope of 1/10 of dB (10 log) plot of normalized n-number $n(f)/n_c$ per decade of operating frequency; and,
f_c is an operating frequency at the intersection of the low frequency asymptote and the high frequency asymptote of dB plot of normalized n-number $n(f)/n_c$.

Indoor RF Propagation Equations

From equation 13,

$$\left[\frac{P_r}{P_{ro}}\right] dB = 10\, n \log\frac{d}{d_o} \qquad (15)$$

Substituting equation 14 in equation 15:

$$\left[\frac{P_r}{P_{ro}}\right] dB \Big/ \frac{-10\, n_c}{1 + (f/f_c)^{N_c}} = \log\frac{d}{d_o} \qquad (16)$$

Equation 16 describes dB power decay of indoor RF propagation at any frequency at any distance in the far zone.

Experimental Correlations

As an example, propagation test results obtained at 49.83 MHz inside a large building, #1, are shown in Figure 2. Scattering of the data points is caused by the orientation and the location of the receiver antenna. From this data, the medium values of the received data are:

$$\begin{aligned} &\text{at } d_o = 6.1 \text{ m} \quad P_{ro} = -37 \text{ dBm} \\ &\text{at } d = 61 \text{ m} \quad P_r = -79 \text{ dBm} \end{aligned} \qquad (17)$$

Substituting these values in equations 7 and 17:

$$\alpha = 0.0839 \ ^{\text{neper/m}} = 0.364 \ ^{\text{dB/m}} \qquad (18)$$

Substituting equation 17 in equation 13:

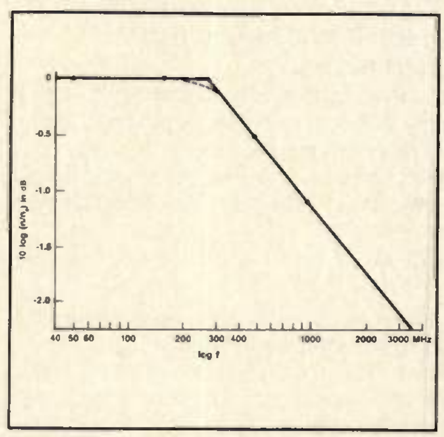

Figure 4. dB plot of normalized n-numbers of Large Building #3.

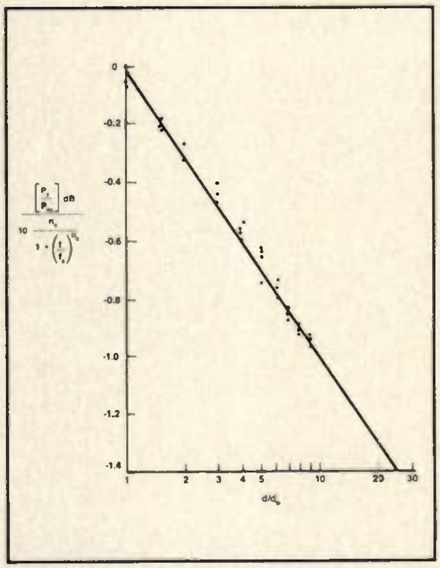

Figure 5. Composite plot of observed median attenuation on calculated median attenuation for indoor electromagnetic wave propagation in various buildings and frequencies.

$$n = 6 \qquad (19)$$

The calculations performed by equation 15 are shown by the solid line in Figure 2. From equation 7,

$$\left[\frac{P_r}{P_{ro}}\right] \text{dB} = 40 \log\frac{d_o}{d} + 4.343\, \alpha d \left(\frac{d_o}{d} - 1\right) \qquad (20)$$

The above is shown by the dashed line in Figure 2. Both curves approximately describe the observed received power levels. The same technique is applied to other buildings to obtain n-numbers and α numbers at 49.83 MHz. The results are tabulated in Table 1. As seen in this table, n-numbers and α numbers can be used for characterizing the indoor RF propagation. The values of n-numbers obtained at various frequencies are plotted in Figure 3. As an example, the 10 log $n(f)/n_c$ plot for building #3 is shown in Figure 4. Figure 4 was constructed from Figure 3. From Figures 3 and 4,

$$n_c = 5.4 \qquad (21)$$
$$N_c = 0.2$$
$$f_c = 219 \times 10^6 \quad \text{(Hz)}$$

Thus, from equations 14 and 21, the n-number of large building #3 is described by:

	n	α
Residential Home #1	4.9	0.0377
Residential Home #2	3.4	0.0252
Large Building #1	6.0	0.0839
Large Building #2	7.0	0.1258
Large Building #3	5.4	0.0587

Table 1. Electromagnetic wave characterization of various buildings at 49.83 MHz.

$$n(f) = \frac{5.4}{1 + \left(\dfrac{f}{265 \times 10^6}\right)^{0.2}} \qquad (22)$$

The same approach is used to obtain the frequency-characterizing parameters n_c, N_c and f_c, and α_c, N_c and f_c. The results are tabulated in Table 2. As seen from this table, different buildings have different characterizing parameters.

Equation 16 has been calculated and graphed in Figure 5. Experimentally obtained median values of normalized dB attenuation to 10 times the n-numbers for various buildings at various frequencies are compositely plotted in Figure 5. As seen from Figure 5, equation 16 describes the observed data.

Conclusion

A theoretical equation for indoor RF propagation was found. Carefully measured median value of the received power level fit this theoretical propagation equation well. A simplified approximate value to describe the propagation characteristic was also found, together with formulas describing frequency characteristics of indoor RF propagation. The derived equations correlate well with experiments, and the equation fits the propagation in any media, at any frequency and at any location in the far zone of propagation.

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	Large Building #3	A Residential Home #1
α_c	0.0587	0.0377
N_c	2.87	7.6
f_c	400×10^6	430×10^6
n_c	5.4	4.9
N_c	0.2	0.0139
f_c	265×10^6	219×10^6

Table 2. Frequency characterizing parameters.

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

MODEL #	FREQ. RESP.	DESCRIPTION
SAS-200/542	20 - 300 MHz	Biconical, Folding
SAS-200/550	.001 - 60 MHz	Active Monopole
SAS-200/560	per MIL-STD-461	Loop - Emission
SAS-200/561	per MIL-STD-461	Loop - Radiating
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FCC Revises Part 15

Changes Offer Opportunities for RF Product Development

By Gary A. Breed
Editor

The First Report and Order in FCC General Docket 87-389 was adopted by the Commission on March 30, 1989, released April 18, and was effective as of June 23, 1989. To quote the opening paragraph of the Report and Order, "The objective of this action is to achieve more effective use of the radio frequency spectrum while providing additional technical and operational flexibility in the design, manufacture and use of non-licensed devices."

Frequency of Emission (MHz)	Field Strength (μV/m)
30 - 88	100
88 - 216	150
216 - 960	200
Above 960	500

Table 1. Limits on field strength of radiated emissions from unintentional radiators (except for Class A digital devices), at a distance of 3 meters.

Frequency of Emission (MHz)	Field Strength (μV/m)
30 - 88	90
88 - 216	150
216 - 960	210
Above 960	300

Table 2. Limits on field strength of radiated emissions from Class A digital devices, as determined at a distance of 10 meters.

Frequency (MHz)	Field Strength (μV/m)	Measurement Distance (meters)
0.009 - 0.490	2400/F (in kHz)	300
0.490 - 1.705	24000/F (in kHz)	30
1.705 - 30.0	30	30
30 - 88	100*	3
88 - 216	150*	3
216 - 960	200*	3
Above 960	500	3

*Except for certain perimeter protection systems, fundamental emissions shall not be located in the bands: 54-72, 76-88, 174-216, or 470-806 MHz.

Table 3. Limits on radiated field strength from intentional radiators. These figures are for unwanted emissions from devices whose fundamental frequencies are covered in specific provisions of Part 15.

The introductory comments for the new Part 15 list four points by which the new rules meet the objective stated above: 1. providing for the production of non-licensed equipment on almost any frequency with minimal restrictions on usage, bandwidth, and modulation technique; 2. establishing more uniform technical standards; 3. clarifying and simplifying administrative requirements; and, 4. retaining currently permitted services. Point number one is easily the most important of the four, although the other points are not minor by any means.

The benefit of the new rules for manufacturers of Part 15 devices is that limits have been lifted on allowable uses and frequencies. According to the "Electromagnetic News Report," published by Liberty Labs, Inc., the new Part 15 paves the way for new technologies, including wireless stereo speaker systems, video cassette recorders, computers and data transmission systems, biomedical telemetry, and factory control and monitoring systems.

Standards and Test Methods

The conducted emissions standards have been simplified to a single 250 μV limit applied to both intentional and unintentional radiators, except for Class A digital devices. This limit would apply over the 450 kHz to 30 MHz frequency range. Measurements of conducted emissions are to be made with the device connected to the power service through a 50 ohm/50 μH line-impedance stabilization network (LISN), rather than the 50 ohm/5 μH LISN previously specified. This value LISN also conforms to ANSI C63.4 and CISPR standards.

Measurement methodology has been changed to use the CISPR quasi-peak detector and bandwidth specifications below 1000 MHz, and peak measurements above 1000 MHz. The revised emissions standards are based on these measurement techniques. A few exceptions are specified, and readers should become familiar with the complete Part 15 provisions.

Tables 1-3 summarize the major emissions limits for unintentional and intentional radiators. For intentional radiators, there are some variations from the limits in Table 3 according to frequency and type of usage. Again, refer to the original FCC documentation. Table 4 is a list of the restricted frequencies, on which intentional radiators may not have their fundamental frequency of operation. Some swept systems and cable locating systems are exempt from these frequency restrictions.

To allow for a smooth transition from the previous Part 15 requirements to these new standards and methods, a generous time period for compliance with the new rules has been established.

90 - 110	kHz	399.9 - 410	MHz	5350 - 5460	MHz
490 - 510		608 - 614		7250 - 7750	
2.1735 - 2.1905	MHz	960 - 1240		8025 - 8500	
8.362 - 8.366		1300 - 1427		9000 - 9200	
13.36 - 13.41		1435 - 1626.5		9.3 - 9.5	GHz
25.5 - 25.67		1660 - 1710		10.6 - 12.7	
37.5 - 38.25		1718.8 - 1722.2		13.25 - 13.4	
73 - 75.4		2200 - 2300		14.47 - 14.5	
108 - 121.94		2310 - 2390		15.35 - 16.2	
123 - 138		2483.5 - 2500		17.7 - 21.4	
149.9 - 150.05		2655 - 2900		22.01 - 23.12	
156.7 - 156.9		3260 - 3267		23.6 - 24	
162.0125 - 167.17		3332 - 3339		31.2 - 31.8	
167.72 - 173.2		3345.8 - 3358		36.43 - 36.5	
240 - 285		3600 - 4400		Above 38.6	
322 - 335.4		4500 - 5250			

Table 4. Restricted bands of operation. Fundamental operation is not permitted, spurious emissions must meet normal limits.

Receivers may be manufactured and imported under the current regulations for ten years. Extensions are not planned, and manufacturers are warned that they are responsible to develop products for authorization under the new rules with a firm ten-year deadline.

For all other Part 15 devices, a five-year transition period has been established. The FCC Chief Engineer has been given the authority to grant extensions of up to two years in situations where the need for additional time can be demonstrated.

Summary

The new Part 15 represents a complete revision of the non-licensed equipment rules. The FCC has gone to great lengths to explain their reasons for adopting each aspect of the new rules. While it is clear that encouraging marketplace-driven product development and competition was a significant factor, they believe that the operation of licensed radio services has been adequately protected.

Additional matters to be addressed in subsequent proceedings may include kits of RF devices, definitions applicable to digital equipment, spread spectrum operation above 900 MHz, expedited proceedings to resolve cases of widespread interference, and labelling of verified devices to indicate the identity of the responsible party.

This report only touches on some highlights of the revised Part 15 rules. Copies of the Report and Order can be obtained from the FCC's copy contractor, International Transcription Service, 2100 M Street N.W., Suite 140, Washington, D.C. 20037, telephone (202) 857-3800. The cost is $31.11. Other commercial FCC update services can also supply copies of the document.

A Simple Low-Cost RF Switch

Economical Design Utilizes Standard Components

By Tad Harris
Universal Data Systems

This RF switch is the result of a project that required a rotary switch to a single feed line. The desire was to develop a switch with no inductors or critical parts, that could be reproduced easily with standard components. Only one section of the "rotary" switch is described here, since it may be built with any number of poles.

The circuit diagram and parts list for this RF switch are given in Figure 1. The performance of the switch can be seen in Figures 2 to 6. At higher frequencies, the stray capacitance begins to take its toll on the ON/OFF isolation performance. Inductors are absent from the design for two reasons: 1) the high ratio of bias resistor to line input/output impedances, and 2) the forward/reverse bias scheme using the CMOS gates. Different bias resistor values will accommodate different PIN

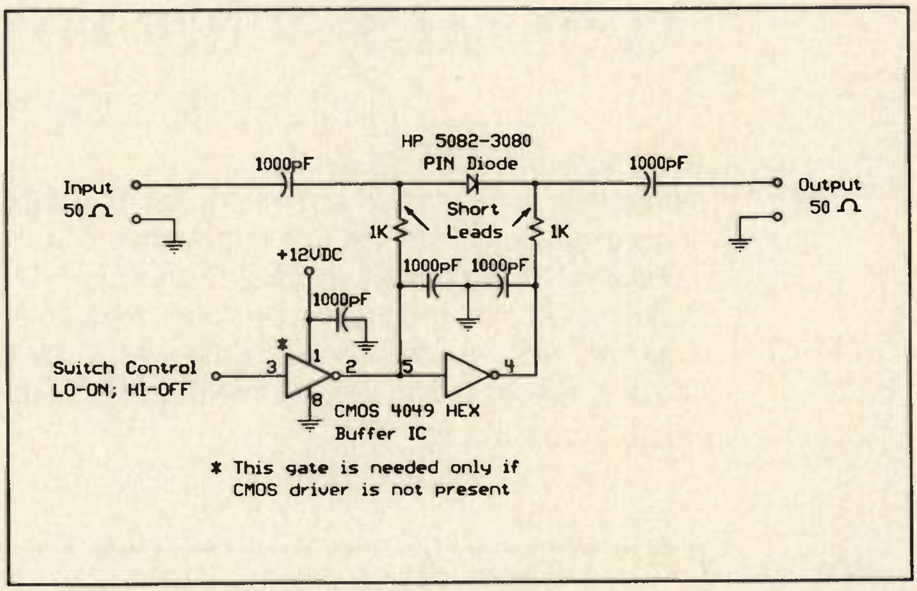

Figure 1. A simple low-cost RF switch.

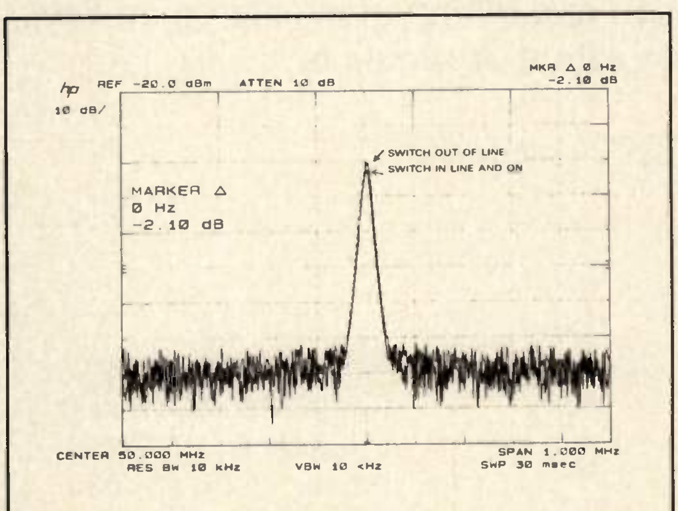

Figure 2. Switch insertion loss at 50MHz (−2.1 dB).

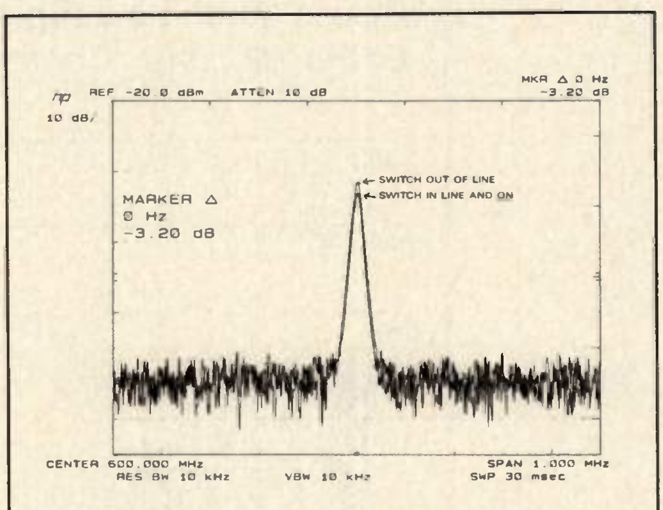

Figure 3. Switch insertion loss at 600 MHz (−3.2 dB).

diodes as well as various different supply voltages.

The switch operates as follows: When the digital logic level at the control input is low, the PIN diode is forward-biased by the CMOS gates. The two 1K bias resistors limit this current to the PIN diode's safe forward current limit. In this state, the switch is ON. When the control input is high, the diode is reverse-biased and the switch is OFF.

This switch is well-suited for electronically steered antenna arrays, multiple path switching, and other applications requiring small, low-cost RF switches. This particular design was used in a four-pole rotary switch for a Doppler shift radio direction-finder operating at 144 MHz. The switch had to have at least 30 dB of ON/OFF isolation. The design performed to specifications as shown in Figure 5. rf

About the Author

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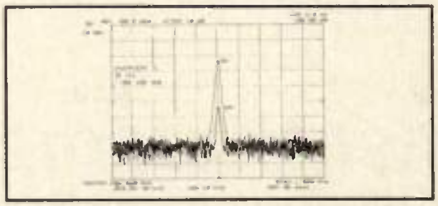

Figure 4. Switch isolation at 50 MHz, ON to OFF =– 39.7 dB.

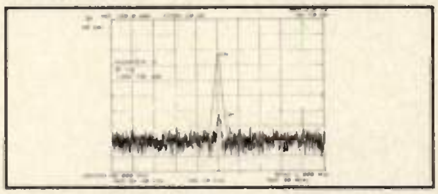

Figure 5. Switch isolation at 200 MHz, ON to OFF = –30.2 dB.

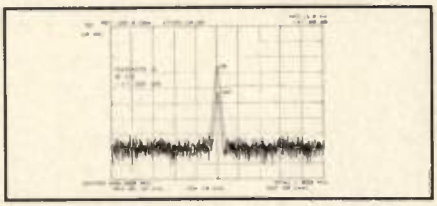

Figure 6. Switch isolation at 600 MHz, ON to OFF =–17 dB.

Lumped-Element Quadrature Couplers

Design Equations and Circuit Examples for Power Combiners/Dividers

By Andre Boulouard
Centre National d'Etudes des Telecommunications (CNET)
Lannion, France

For engineers involved in MMIC design, synthesis equations for matched quadrature couplers and power combiners/dividers are needed as a good starting point for computer optimization. This article presents the basic design equations in compact form, leading to simple synthesis solutions with the help of a computer.

The lumped element quadrature coupler under consideration is shown in Figure 1. The design equations are very similar to those of lumped phase shifters and may be derived by following the even-mode and odd-mode approach used in References 1 and 2.

The equations required to derive the parameters for Figure 1 are:

$B_2 = \sqrt{G_G G_L}/\sin\psi$
$B_4 = -G_G\cot\psi$
$B_5 = -G_L\cot\psi$
$B_1 = -B_4 - B_2$
$B_3 = -B_5 - B_2$

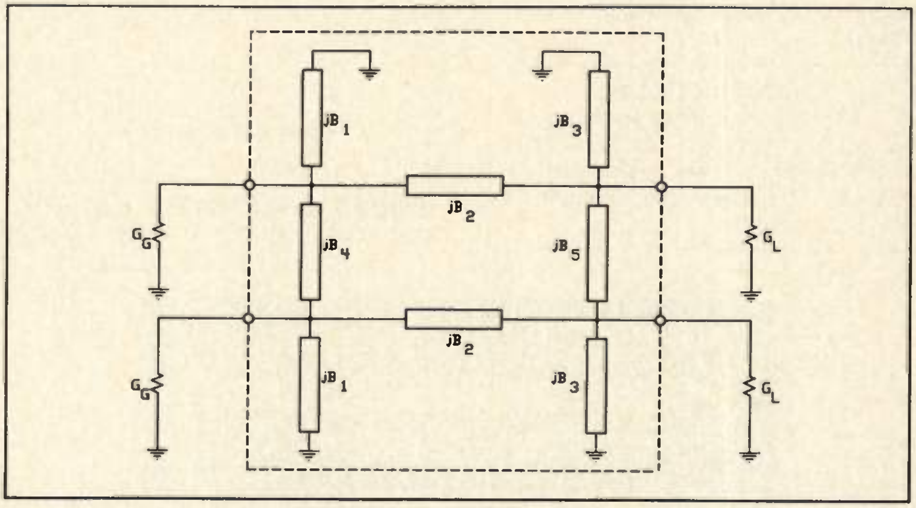

Figure 1. A symmetrical lumped-element quadrature coupler.

Where, G_G is the input admittance, G_L is the output admittance, and B_1 to B_5 are the respective susceptances shown in Figure 1.

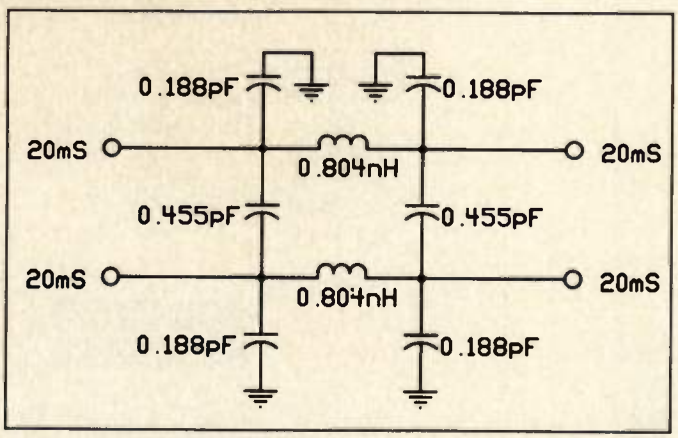

Figure 2(a). A 7 GHz, 50/50 ohm, 3 dB coupler with $\psi = +45°$.

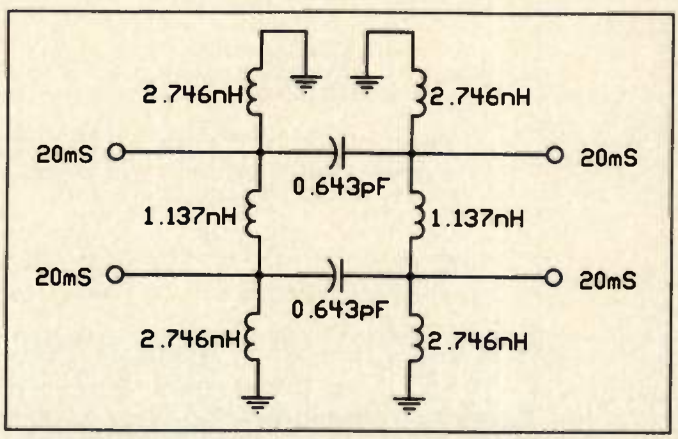

Figure 2(b). A 7 GHz, 50/50 ohm, 3 dB coupler with $\psi = -45°$.

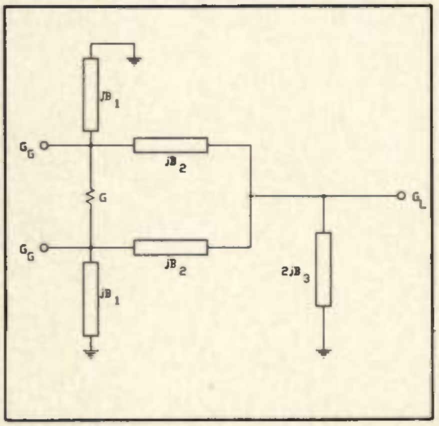

Figure 3(a). A Pi-section Wilkinson power combiner/divider.

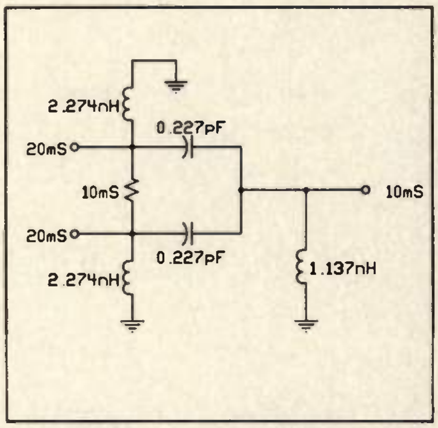

Figure 3(b). A 7 GHz, 50/100 ohm Pi-section Wilkinson power combiner/divider.

An Example

To further explain this process, a 7 GHz, 50/50 ohm, 3 dB coupler is designed. First, for a 3 dB coupler, ψ is ±45°. The impedances are then converted to admittances. Therefore,

$G_G = G_L = 0.02$

Then,

$B_2 = 0.02/\sin 45 = 0.02828$
$X_2 = 1/B_2 = 35.3606$

Since $X_L = \omega L$,

$L_2 = 35.35/(2\pi(7 \times 10^9)) = 0.804$ nH

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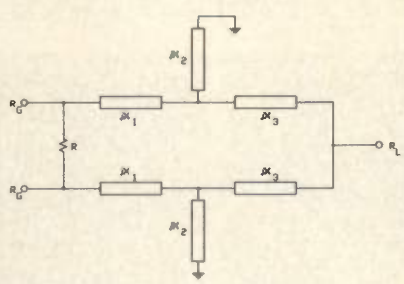

Figure 4(a). A Tee-section Wilkinson power combiner/divider.

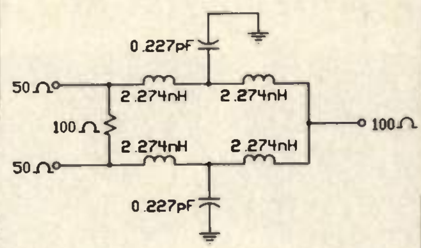

Figure 4(b). A 7 GHz Tee-section, 50/100 ohm Wilkinson power combiner/divider.

From the previous section, $B_4 = -0.02$ and $X_4 = -50$.

Since $X_C = -1/\omega C$,

$C_4 = 1/(2\pi(7 \times 10^9)(50)) = 0.455$ pF

Using the same technique,

$C_5 = 0.455$ pF

Then,

$B_1 = 0.02 - 0.02828 = -0.00828$
$X_1 = -120.77$

Therefore, $C_1 = 0.188$ pF.

C_3 is obtained similarly and found to have a capacitance of 0.188 pF also. Figure 2 shows 7 GHz, 50/50 ohm, 3 dB couplers designed from these formulas. The circuit designed here is shown in Figure 2(a). The circuit in Figure 2(b) was obtained with the same procedure with ψ at $-45°$.

Lumped Power Combiners/Dividers

Pi and Tee-section Wilkinson power combiners/dividers are shown in Figures 3 and 4. The equations are derived from 90-degree, 3-element phase shifter design.

The block diagram in Figure 3(a) shows a Pi-section Wilkinson power combiner/divider where:

$$B_2 = \pm\sqrt{G_G(G_L/2)}$$
$$B_1 = B_3 = -B_2$$
$$G = G_G/2$$
$$\psi = \pm 90°$$

A 7 GHz, 50/100 ohm Wilkinson power divider is illustrated in Figure 3(b).

The circuit parameters were derived from the above equations similarly as illustrated in the previous section.

The following equations relate to Figure 4(a), a Tee-section Wilkinson power combiner/divider. Figure 4(b) is an example of a circuit designed at 7 GHz, with 50 and 100 ohm input and output impedances.

$$X_2 = \pm\sqrt{R_G/(2R_L)}$$
$$X_1 = X_3 = -X_2$$
$$X_3 = -X_2$$
$$R = 2R_G$$
$$\psi = \pm 90°$$

Conclusion

Design equations for synthesizing lumped couplers and power combiners/dividers have been presented and should be a useful design aid to the network designer. These formulas have been programmed in an in-house network analysis computer program and have been used successfully in MMIC design (3). rf

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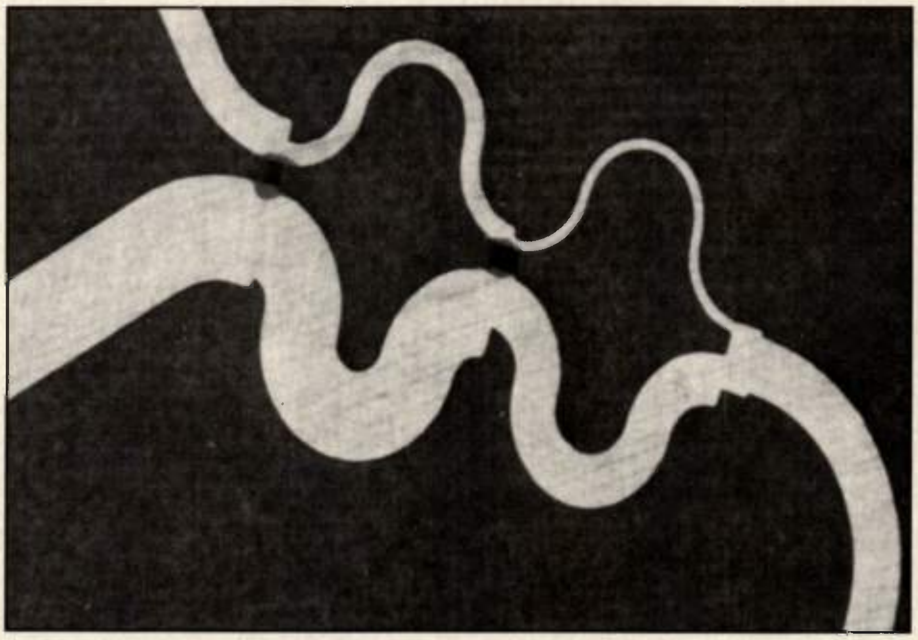
The Ohmega-Ply foil is a bi-metallic material consisting of a layer of copper foil and a thin layer of resistive material. The Ohmega laminates are processed by standard subtractive PC board techniques to produce resistors on single-sided, double-sided or multi-layer microwave circuit boards. The laminates offer several design and processing advantages over traditional resistor mounting techniques. These include savings in space, time and assembly costs, and eliminating the need for relief and access holes.

A large number of resistors can be generated in a process with this material—an advantage in the design of termination resistors. Resistor values are infinitely variable with planar resistors. This helps to eliminate the problem of discontinuities and impedance mismatches.



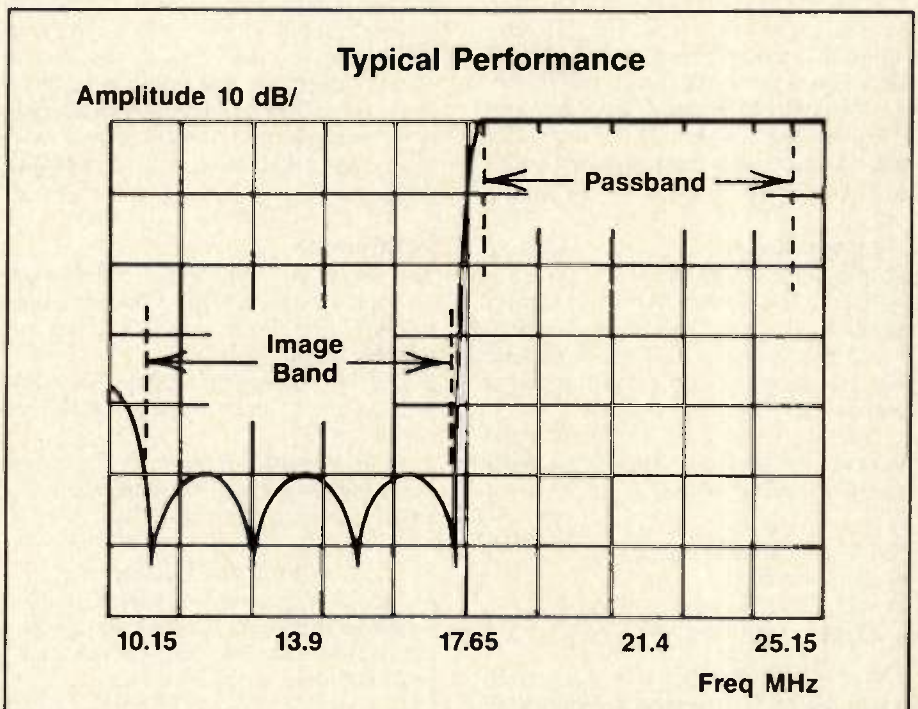
Typical electrical performance characteristics include a sheet resistivity in the X and Y directions of 25 ohms/square at 73°F, an etched resistivity tolerance of 7 percent within a 10 in. X 12 in. panel, and a resistance change of 1 percent max. after 20 seconds at 536°F in solder tip. These resistors are 0.25 microns thick. **Rogers Corp., Rogers, CT. INFO/CARD #230.**

Maxtech Introduces an Imageless Mixer

The IM102DL-ADA is a broadband imageless mixer designed for signal processing applications where it is impractical to use a filter for image rejection. Typical applications include IF to tape conversion, baseband conversion for analog to digital converters, and receiver front-ends.

The IF center frequency is typically 21.4 MHz. The LO input frequency is 17.650 MHz and the LO input level is ±1 dBm. Output frequency range is 0.075 to 7.50 MHz, image rejection is typically 50 dB and conversion loss is 12 dB max. Amplitude ripple is 0.25 dB max p-p, and third-order intercept point is typically +25 dBm. Impedance and VSWR at all ports are 50 ohms and 1.5:1, respectively. The mixer uses SMA female connectors.

Options for the Model IM102DL-ADA include internal crystal LO, phase-locked LO, group delay equalization, passive design, IF or baseband amplifiers for reduced conversion loss, and anti-alias lowpass filters. **Maxtech, Inc., State College, PA. INFO/CARD #229.**



TEM Cell/Anechoic Chamber

EMCO introduces a broadband test device which is a combination TEM cell and anechoic chamber. It can be used for both susceptibility and emissions


EMCO GTEM

testing in either the frequency or time domain mode. Applications include EMI/RFI testing, TEMPEST, RADHAZ, ESD, NEMP, lightning and high-power microwave-related experiments. The GTEM offers an enclosed test environment for controlled repeatable results. **EMCO, Austin, TX. INFO/CARD #228.**

1 kW Amplifier

The Harris Long Range Radio Division introduces an amplifier for the RF-3200 marine HF/SSB radio. The RF-3230 provides 1 kW PEP/average power over the 1.6 to 30 MHz range. It features a redundant modular concept which permits the unit to continue functioning at reduced power in the event of a malfunction in one of the identical final amplifiers or the power supply module. A digital lowpass filter network that tunes in less than 50 ms is standard. **Harris RF Communications, Rochester, NY. INFO/CARD #227.**

Coaxial Test Cable Test Assembly

The Heliax test cable assembly is available with SMA plug connectors. Designated Type 39812-11, the 3 ft. assembly features an impedance of 50 ohms with maximum VSWR of 1.55. **Andrew Corp., Orland Park, IL. Please circle INFO/CARD #226.**

Exponential Pulse Generator

R&B introduces the PGF-5006 exponential pulse generator designed to meet the monopulse waveforms for the performance of lightning susceptibility tests for RTCA D0160B Section 22. Front panel selection allows for testing to either the short or long waveform. A polarity switch selects either positive or negative polarity pulses. **R & B Enterprises, W. Conshohocken, PA. Please circle INFO/CARD #225.**

SMD Test Fixture

This test fixture for surface-mount devices is suited for testing tight-tolerance, low-value passive components at frequencies from 1 to 1000 MHz. It accommodates 0805 and 1812 body sizes. Price ranges from $425 to

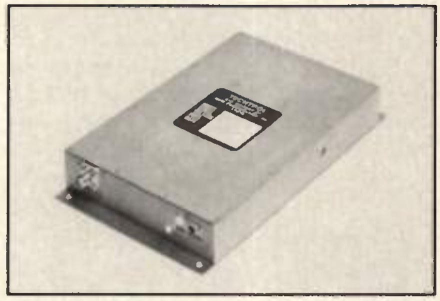
$498 depending on configuration selected. **Coilcraft, Cary, IL. Please circle INFO/CARD #224.**

Voltage-Controlled Filter/Oscillator

The SSM-2044 is a four-pole voltage-controlled filter/oscillator. As an alternative to switched capacitor filters, it offers a bandwidth of 1 MHz. The device is a −24 dB/octave lowpass filter with a 10,000 to 1 variable cutoff frequency. Dynamic range is 90 dB. **Precision Monolithics, Inc., Santa Clara, CA. INFO/CARD #223.**

Frequency Multiplier

This frequency multiplier allows for multiplication of references without degrading the phase noise above 20 log N. Residual system noise floor is −170 dBc/Hz at 1 kHz and −174 dBc/Hz at 10 kHz. It is available for output frequencies up to 1500 MHz and accepts inputs of 5 to 125 MHz. Multiplication factors of X2 to X18 can be specified for ±1 percent bandwidth. Output power is ±10 dBm min with a +24 VDC input. The package measures 6 in. X 4 in. X 1 in.



Techtrol Cyclonetics, Inc., New Cumberland, PA. INFO/CARD #222.

GPIB Interface

Model 488 from JFW Industries accepts commands from a GPIB and outputs a binary coded TTL word to drive programmable attenuators or RF switch matrices. Local operation is available through eight binary coded TTL input lines and a local/remote switch. **JFW Industries, Inc., Indianapolis, IN. Please circle INFO/CARD #221.**

SiO_2 Reference Lines

Kaman introduces SiO_2 reference lines for the HP 8510 vector automatic

In Ferrites for EMI suppression...

WE'RE INNOVATORS — NOT IMITATORS

WE'RE FAIR-RITE

While other ferrite manufacturers exploited existing markets, our R&D looked to the future—and keeps looking—to find better ways to put ferrite to work in the control of Electromagnetic Interference.

In addition to the beads and sleeves which are industry standards, we developed most of the ferrite EMI suppressors widely imitated today, and were the first to offer what has become a series of Engineering Evaluation Kits.

No other source offers faster service and more tooling know-how for unique shapes than Fair-Rite Products.

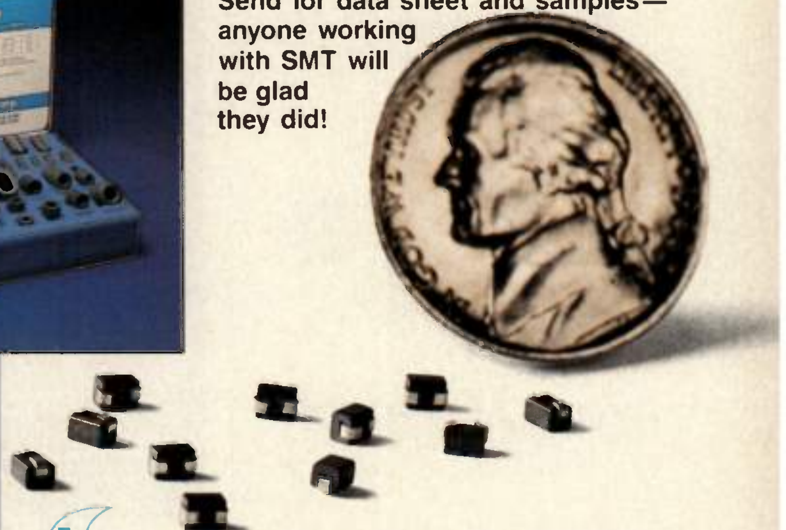
When it comes to EMI ferrites, the leaders come to Source One for quality, service and price from a U.S. manufacturer that supplies the world—FAIR-RITE

SURFACE MOUNT BEADS UPDATE

Z where you need it—
in a still smaller package.

Part number 2743027447 offers a minimum impedance of 20 ohms from 100 MHz thru 1 GHz, and a maximum of .7 milliohm DC; all in a rugged package 1.75mm wide × 1.5mm high × 2.0mm long. The terminal electrodes measure 0.2mm thick × 0.6mm square. Manufactured from application proven 43 material ferrite, stability and reliability far exceed what other manufacturers achieve using multiple turn, high resistivity thru hole plating.

Send for data sheet and samples—anyone working with SMT will be glad they did!



Fair-Rite Products Corp.

PO Box J, One Commercial Row, Wallkill, NY 12589
Phone (914) 895-2055 • TWX 510-249-4819 • FAX (914) 895-2629

INFO/CARD 48

rf products *Continued*

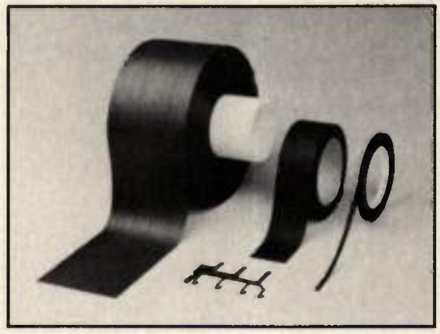
network analyzer. These reference lines are designed to reduce random errors caused by phase drift with temperature, improve measuring accuracy in semi-controlled to uncontrolled environments, allow for measurements over a broader temperature range, and extend calibration life. They are used to match the electrical length of the test cables on the HP 8510. **Kaman Instrumentation, Colorado Springs, CO. INFO/CARD #220.**

Automated Noise Calibration System

HP introduces the HP 8970B-E09 automated noise-source calibration system which is capable of performing 75 to 100 calibration actions per year. The system is configured using a sweeping local oscillator, a noise-figure meter, some standard calibration sources, and a custom instrument with special switching and a downconverter with low-noise amplifier. A software-controlled test procedure is used. It is priced from approximately \$20,000 to \$100,000. **Hewlett-Packard Company, Palo Alto, CA. INFO/CARD #219.**

EMI Shielding Material

W.L. Gore introduces a Gore-Tex EMI shielding material which provides suppression of radiated and conducted EMI and RFI together with atmospheric sealing. It has more than 100 dB of E-field



suppression at 100 kHz and more than 80 dB of suppression at 10 MHz. Planewave shielding is more than 50 dB from 1 to 18 GHz. Volume resistivity is 0.5 ohm-cm and decreases as the material is compressed, causing conductivity to increase. **W.L. Gore & Associates, Inc., Newark, DE. Please circle INFO/CARD #218.**

DIP TCXOs

Series CO-253 offers a frequency stability of $\pm 2 \times 10^{-7}$ over 0°C to +50°C or $\pm 1 \times 10^{-6}$ over −55°C to +85°C in a 24-pin package. Output is HCMOS/TTL compatible from 5 MHz to 50 MHz. This resistance-welded package measures 0.8 in X 1.5 in. X 0.39 in. **Vectron Laboratories, Inc., Norwalk, CT. Please circle INFO/CARD #217.**

GaAs IC Prototyping/Testing Kit

GigaBit unveils a prototyping board to provide an inexpensive way to build and test circuits with LSI GaAs ICs. The 90GKIT-68 high-speed prototyping kit provides all the components required to prototype and test the company's line of PicoLogic™, NanoRam™, and NanoRom™ GaAs ICs. It features two sites for 68-pin packages, six sites for 36- or 40-pin packages, one site for a 0.3 or 0.4 in. DIP IC, and six user-selectable power planes. The kit is designed to support signals with 100 ps rise and fall times and up to 3 GHz clock rates. **GigaBit Logic, Inc., Newbury Park, CA. Please circle INFO/CARD #216.**

Conductive Adhesive Coating

Master Bond AC82 is a one component conductive adhesive coating designed for EMI/RFI shielding applications. It can be applied to adhere to most surfaces including plastics without need for pre-treatment. Application methods include air or airless type spraying equipment, brushing, dipping and silkscreening. Uniform continuous coatings characterized by high electrical conductivity and good electromagnetic shielding effectiveness are easily formed. **Master Bond Inc., Hackensack, NJ. Please circle INFO/CARD #215.**

New Signal Generators

The SMGU and SMHU signals generators from Rohde & Schwarz feature high spectral purity for selective measurements, frequency agility to handle frequency hopping, and broadband modulation. The SMGU covers 100 kHz to 2.16 GHz, while the SMHU covers the 100 kHz to 4.32 GHz range with 0.1 Hz resolution. Up to 200 frequencies and a variety of modulation and level settings can be called up with an external trigger or in a continuous sequence. SSB phase noise at 20 kHz offset is −134 dBc. The FM modulation range is from DC to 1 MHz. **Rohde & Schwarz, Munich, West Germany. INFO/CARD #214.**

Video Mixer

The TMC2249 is a monolithic IC for image-processing applications which eases the filtering and mixing aspects of image manipulation. It is a high-speed digital arithmetic circuit consisting of two 12-bit multipliers, an adder and a cascadable accumulator. The 16-bit accumulated output is available up to 30 MHz. Applications include video switching, image mixing, digital filtering, frequency synthesis, and arithmetic functions. **TRW LSI Products, Inc., La Jolla, CA. INFO/CARD #209.**

1989 EDITORIAL CALENDAR

Issue	Shows & Special Coverage	Featured Technology	Industry Insight
JANUARY	• SMART IV (Surface Mount Expo)	RF Power Amplifiers	Test Equipment
FEBRUARY	• RF Technology Expo 1989 • Aerospace Conference	Small-Signal Amplifiers	GaAs Technology
MARCH		Filter Design	Power Transistors
APRIL	• National Association of Broadcasters (NAB)	The RF Spectrum: I. Low and Medium Frequencies	Packaging Update
MAY	• Frequency Control Symposium • IEEE EMC Conference • Recruitment Issue/College Distribution	The RF Spectrum: II. HF Technologies	Frequency Synthesis
JUNE	• MTT-S • European Microwave Show	The RF Spectrum: III. VHF and UHF	Subsystems
JULY	• EMC Expo	Electromagnetic Compatibility • 1989 Design Contest Results	Filters
AUGUST	• Quartz Devices Conference • Antenna Measurement Techniques Association (AMTA)	Crystal Oscillators and Filters	Attenuators and Switches
SEPTEMBER	• Coil Winding Show	Test and Measurement Techniques	Inductors, magnetic materials (ferrite, iron powder)
DIRECTORY (SEPT.)	All Shows	**DIRECTORY ISSUE** Vendor Listings, Design Notes	
OCTOBER	• Old Crows • Ultrasonics Symposium • Electronica (Europe)	System Design: Build-or-Buy Decision	SAW Update
NOVEMBER	• RF Expo East 1989	Using Passive Components	RF Software
DECEMBER	Article Index	Mixers, Modulators and Demodulators	Cables & Connectors

rf design

Rate Card No. 11, January, 1989

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Englewood, CO 80111
Tel: (303) 220-0600
Fax: (303) 773-9716

FREQUENCY

RF Design is published monthly with a 13th directory issue distributed in Fall. Issued on the 25th of the month preceding the cover-date month. Issues close for advertising orders on or about the 5th of the month preceding cover-date.

MARKET SERVED

RF Design is edited exclusively for engineers and engineering managers who design equipment and systems using electronic frequencies from the audio range up to about 3 GHz, whether for communications or other purposes, in military, data-handling, aerospace, entertainment, instrumentation and other applications.

EDITORIAL OBJECTIVE

RF Design has the dual mission of tutoring electronics engineers in the basic techniques of high frequency circuits (which are not often taught in academic curricula) and of keeping senior engineers abreast of the latest developments in high-frequency technology. In-depth technical design articles provide the main staple of the editorial diet. They are supplemented by new products, news and other forms of topical coverage of interest to the RF engineer.

COMMISSION & TERMS

Agency discount of 15 percent to recognized agencies provided payment is made in accordance with credit terms.

Credit terms: Net 30 days. No cash discounts. Agency accepts full responsibility for payment of insertion orders placed through them.

Display Advertising Rates

*1989 Rates

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1/2 Island	1990	1900	1740	1360
1/2 Page	1780	1690	1550	1220
1/3 Page	1520	1450	1320	1040
1/4 Page	1140	1090	990	790
1/6 Page	840	790	730	570

*1988 Rates

(Available in 1989 to qualifying advertisers)

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1/2 Page	1680	1595	1460	1145
1/3 Page	1435	1365	1250	975
1/4 Page	1080	1025	940	735
1/6 Page	795	755	690	540

*Rates for volume beyond 12 pages, or for multiple fractions in an issue, will be quoted on request.

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Standard colors, per page or fraction $250
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Metallic colors, per page or fraction 350
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Bleed add 10% to space cost only.

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Display and commercial classified space available at regular display rates. Non-commercial classified space available at $125.00 per column inch.

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Directory issue rates are 90% of earned magazine rate, and do not count towards frequency. Bulk page and insert rates for the Directory will be quoted upon request.

INSERTS

Furnished Inserts: Rates and deadlines for carrying furnished inserts available on request. Inserts are commissionable and count one time towards frequency.

Printed By Publisher: Rates and deadlines for multi-page inserts printed by publisher from negatives or camera-ready artwork available on request. The space cost alone is commissionable and counts one time towards frequency.

POSTCARDS

Bound-in cards may be placed in every issue of *RF Design*. Cards are printed on 80-lb. card stock and bound in three per page. Cards measure 3½" × 5½", and when removed leave 3½" × 2¼" stub which can be used for additional ad space. The $795 rate is commissionable and does count towards frequency. Publisher's second color is available at $350 per card.

Postcard Packs: Published in Spring and Fall, cards measure 3½" × 5½" and are priced at $795 each for RF contract advertisers; $895 each for non-advertisers. Card Pack insertions are commissionable, but do not count toward magazine frequency. Second color (red, blue, or green) available at $250 per card. Rates for multiple cards quoted upon request.

rf design

6300 S. Syracuse Way, Suite 650 / Englewood, Colorado 80111 (303) 220-0600

July 18, 1989

Alan Victor
Monicor
2964 N.W. 60th Street
Ft. Lauderdale, FL 33309

Dear Alan:

It's almost that time!

I know that we still have two weeks until we close our special **"test equipment"** issue of RF Design for advertising space, but I wanted to give you the details on this reader favorite.

Test equipment is one of the most exciting and challenging areas for RF design engineers. Today's engineers must meet tighter specifications...in a shorter time...using more complex components. Because of this, test equipment has never been a more important tool or a more important issue. And that's not all! Today's engineers and engineering managers have more vendors, features, and performance levels to choose from...making the buying decision more complex and important than ever.

Because of this, the September issue of RF Design will be dedicated to bringing our 40,000 subscribers and 90,000 readers up to speed in the most significant test equipment areas. Make sure that these exclusive subscribers learn not only about the technology involved...but learn about your products unique features and cost/performance benefits as well.

Your ad in this special issue can market your products into a thriving, expanding industry...the RF industry. While the microwave industry suffers from a declining economy, the RF industry is healthy and growing. Make sure our 40,000 specifying engineers and engineering managers send a "request for quote" your way! Ad closing is July 31...materials deadline is August 7.

I will be calling next week!

Sincerely,

Bill Pettit

Bill Pettit
Account Executive

SOIC Op Amps

Analog Devices introduces a family of small outline operational amplifier with gain bandwidth from 4 to 725 MHz. The AD712 is a dual-channel BiFET and the AD744 is a 13 MHz unit. The AD847 features a 50 MHz unity-gain bandwidth and 300 V/μs slew rate. The AD848 and AD849 feature gain bandwidths of 175 MHz and 725 MHz, respectively. **Analog Devices, Norwood, MA. Please circle INFO/CARD #210.**

Quad SPST Analog Switch

The MAX334 CMOS high-speed quad SPST analog switch has a turn-on time of less than 100 ns and turn-off time of less than 50 ns. The channel ON resistance is 50 ohms max. Applications for the MAX334 include high-speed test equipment, sample-and-hold circuits, and communication equipment. **Maxim Integrated Products, Sunnyvale, CA. INFO/CARD #213.**

Blindmate/Slide-on Connectors

Radiall introduces the BMA blindmate/slide-on connectors which are rated up to 18 GHz. To reduce insertion loss and reflections, interconnect cabling is eliminated. The BMA plug-in connectors are available for semi-rigid and flexible cable applications, floating or rigid mount, stripline and printed circuit attachment, and hermetic and non-hermetic applications. **Radiall, Inc., Stratford, CT. Please circle INFO/CARD #212.**

Integrated Phase-Locked Loop Chip

MB1501 is a single chip prescaler and phase-locked loop synthesizer which operates at a maximum frequency of 1.1 GHz. A built-in analog switch offers fast switching times between channels. The switch temporarily changes the lowpass filter settling time constant when communication channels are switched, reducing lock-up time for frequency steps without decreasing signal-to-noise ratios. It is available in 16-pin DIP and small outline J-lead packages. In quantities of 1000, price is $12 each. **Fujitsu Microelectronics, Inc., San Jose, CA. INFO/CARD #211.**

rf literature

Amplifier Databook

This databook from Avantek features the company's line of military, industrial, commercial and telecommunications microwave and millimeter-wave amplifier products. Product categories include low-noise, gain control, power and limiting amplifiers; downconverter assemblies; and cellular radio base-station amplifiers. **Avantek, Inc., Milpitas, CA. INFO/CARD #200.**

Crystal Oscillator Catalog

Oscillators ranging from 1 Hz to 1.2 GHz with stabilities from ±0.01 percent to 1 $X10^{-10}$ are featured in this catalog. It details clock oscillators (TTL, CMOS, HCMOS and ECL), small sinewave oscillators, low phase noise OCXOs, TCXOs, VCXOs, and VCOs. **Vectron Laboratories, Inc., Norwalk, CT. Please circle INFO/CARD #199.**

PIN Diode Application Note

Distortion in PIN Diode Switches contains four application notes which discuss distortion. The articles include one that is titled, "Distortion in Microwave and RF Switches By Reverse Biased PIN Diodes," by G. Hiller and R. Caverly. **M/A-COM, Inc., Burlington, MA. Please circle INFO/CARD #198.**

DSO Video

A video discussing digital storage oscilloscope (DSO) technology and the features to look for when selecting a scope is available from John Fluke. The video, "DSOs With a Difference," highlights the PM 3335, PM 3350, and PM 3365 family of medium frequency analog/digital storage oscilloscopes as an example of instruments that combine digital storage capability with analog ease-of-use. **John Fluke Mfg. Co., Inc., Everett, WA. INFO/CARD #197.**

EMC Test Lab and Consultants Directory

This directory is a reference tool used to locate EMC testing and consultant services. The listings include articles and reference information, history and background information, capabilities and services, facilities, major instrumentation, lead times, contacts, and personnel information. This yearly directory together with a supplemental update is priced at $65. Further information can be obtained by circling the reader service number. **ENR/Liberty Labs, Inc., Cedar Rapids, OH. Please circle INFO/CARD #196.**

Brochure Features Chip Resistors, Attenuators and Terminations

TRX, a division of Barry Industries, introduces a brochure that provides specifications on wraparound resistors, QPL chip resistors, flip-style chip resistors, microwave chip terminations, and microwave chip attenuators. Also included is a list of engineering kits for prototype development work or for pilot runs. **TRX, Inc., Attleboro Falls, MA. INFO/CARD #195.**

Japanese Technical Literature Bulletin

Japanese Technical Literature Bulletin is designed to provide English language reviews of Japanese R&D and federal programs. It is produced about every two months by the Technology

Administration's Japanese Technical Literature program and is available at no charge by writing to: Rm. 4833, U.S. Dept. of Commerce, Washington, D.C. 20203. **U.S. Dept. of Commerce, Washington, DC. INFO/CARD #194.**

Diode Brochure

This brochure presents information on FEI's Schottky, planar tunnel, PIN/NIP, and step recovery diode families. A section describes the JAN/JANTX/JANTXV 5711, 5712 and 5719 Series of QPL devices. **FEI Microwave, Inc., Sunnyvale, CA. INFO/CARD #193.**

Connector Catalog

Amphenol introduces a catalog that details 34 connector series comprising over 4,000 coaxial, triaxial and twinaxial cable connectors, receptacles, launchers, terminations, adapters, accessories and related products for RF, microwave and data transmission system interconnections used in commercial and military/aerospace applications. **Amphenol Corp., Danbury, CT. Please circle INFO/CARD #192.**

rf software

RF/Microwave/MMIC Software

TOAST™ is a frequency domain linear simulator to analyze, optimize and perform sensitivity and yield computation of RF and microwave circuits. It includes a nonlinear circuit simulator, based on spectral analysis, to perform power output, distortion, spectral response and AC/DC analysis. The program can be used to design microwave and RF linear and nonlinear circuits like filters, low-noise amplifiers, power amplifiers, mixers and frequency converters. It includes pop-up menus and features a link for schematic capture (Hewlett-Packard or Mentor Graphics). The software is available for most popular workstations. **Caro-Line Habigand Sa., Toulouse, France. Please circle INFO/CARD #190.**

Pre-Processing Program

Intusoft's Pre-Spice provides various components including switched capacitor networks, servos, tubes, pulse width modulators, and analog computer functions. It converts data sheet parameters into Spice and integrates the data with Pre-Spice's library of generic models. Model categories include standard semiconductor devices, power electronic components, ICs, switched capacitor networks and Z-transform models, analog computer functions, phase-lock loops, digital models, thermal models, and ASIC libraries. The program is priced at $200. **Intusoft, San Pedro, CA. Please circle INFO/CARD #189.**

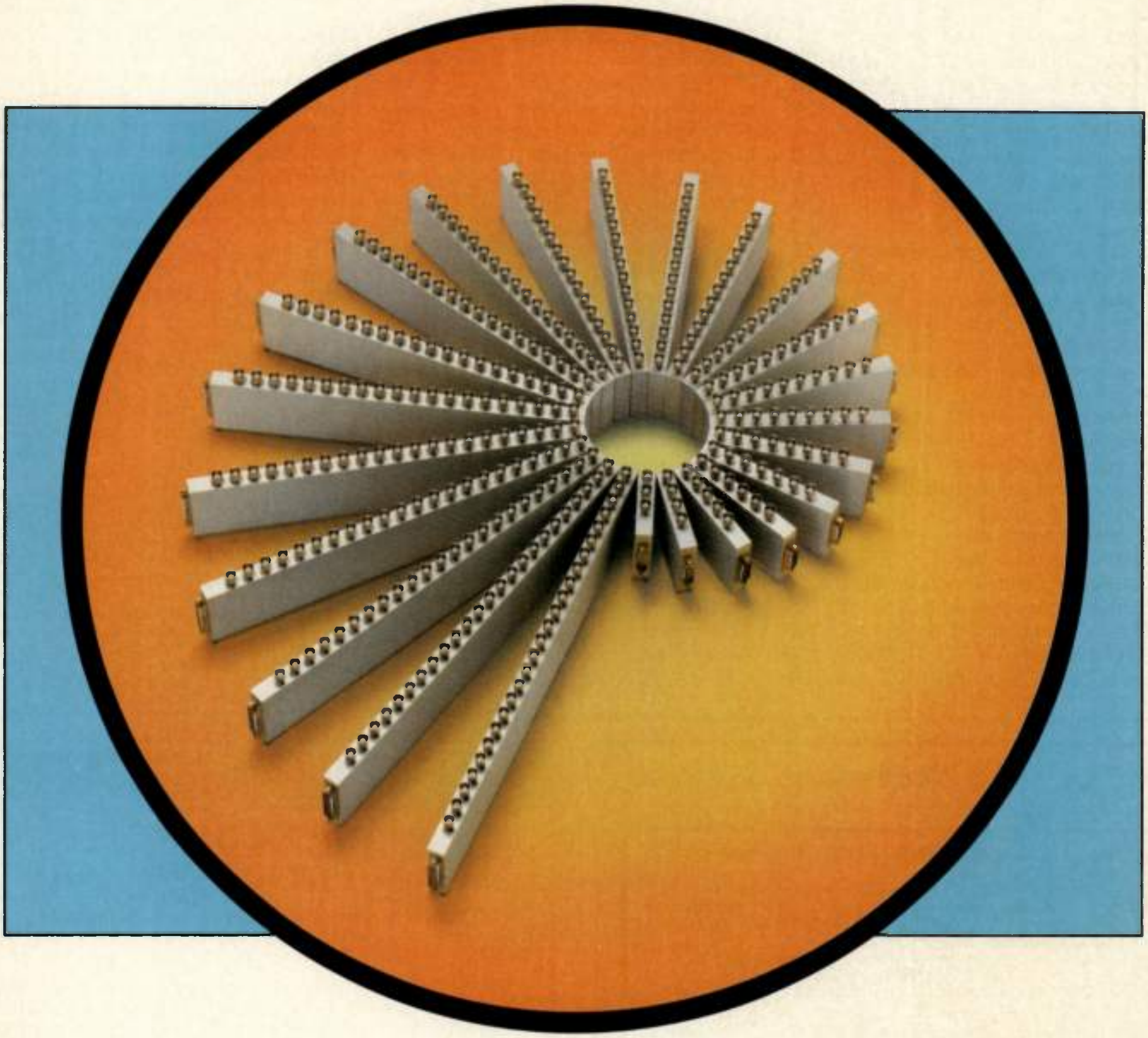
Differential Equations Software

ODELIB solves the initial value problem for systems of first order differential equations. Also, higher order systems can be transformed to the required form by a simple reduction technique. It features interactive and graphics tutorial programs that work in concert with the tutorial documentation. The program sells for $350. **McGraw Hill Publishing Co., New York, NY. Please circle INFO/CARD #188.**

GPIB Test Program Generator

Wavetek introduces WaveTest™ Version 2.5 — a software package that automatically generates a test program, creates the relevant code and provides documentation. It allows the user to create sub-programs and templates only once and copy and merge them into new programs. Version 2.5 costs $3,990. Existing customers can upgrade to version 2.5 for $495. **Wavetek San Diego, Inc., San Diego, CA. Please circle INFO/CARD #187.**

NEED BROAD-BAND COAXIAL RELAYS? FROM 2 TO 24 THROW, MATRIX HAS THE ANSWER



Our versatile 7000 series of coaxial relays have band-widths from DC up to 800 MHz. They're available from 2 to 24 throw. And by using our 9000 series cross-straps, switching matrices of any size can be configured.

Why have Matrix broad-band relays become the industry standard? Because we construct them of precision machined anodized aluminum alloy, all signal shield paths are silver plated, and basic switch elements are hermetically sealed in nitrogen filled gas envelopes with rhodium plated contacts to insure non-stick operation.

The end result is extremely low crosstalk, EMI and VSWR. Another plus, all switchpoints are individually field replaceable.

The units are plug compatible with Matrix 6100A and 1600 Series Logic Modules for compatibility with RS-232, RS-422 and IEEE-488 Interface busses as well as 16 bit parallel.

Non-blocking Matrix configuration may be easily assembled using our self-terminating relays and 5100A series power dividers. Built-in Video/RF amplifiers allow zero insertion loss designs.

So if you're looking for broad-band relays, it pays to deal with Matrix. After all, we've been designing state-of-the-art reed relay and semiconductor switching systems for over 18 years.

Our customers include government agencies, defense contractors, the TV industry, ATE and telecommunications companies—and more.

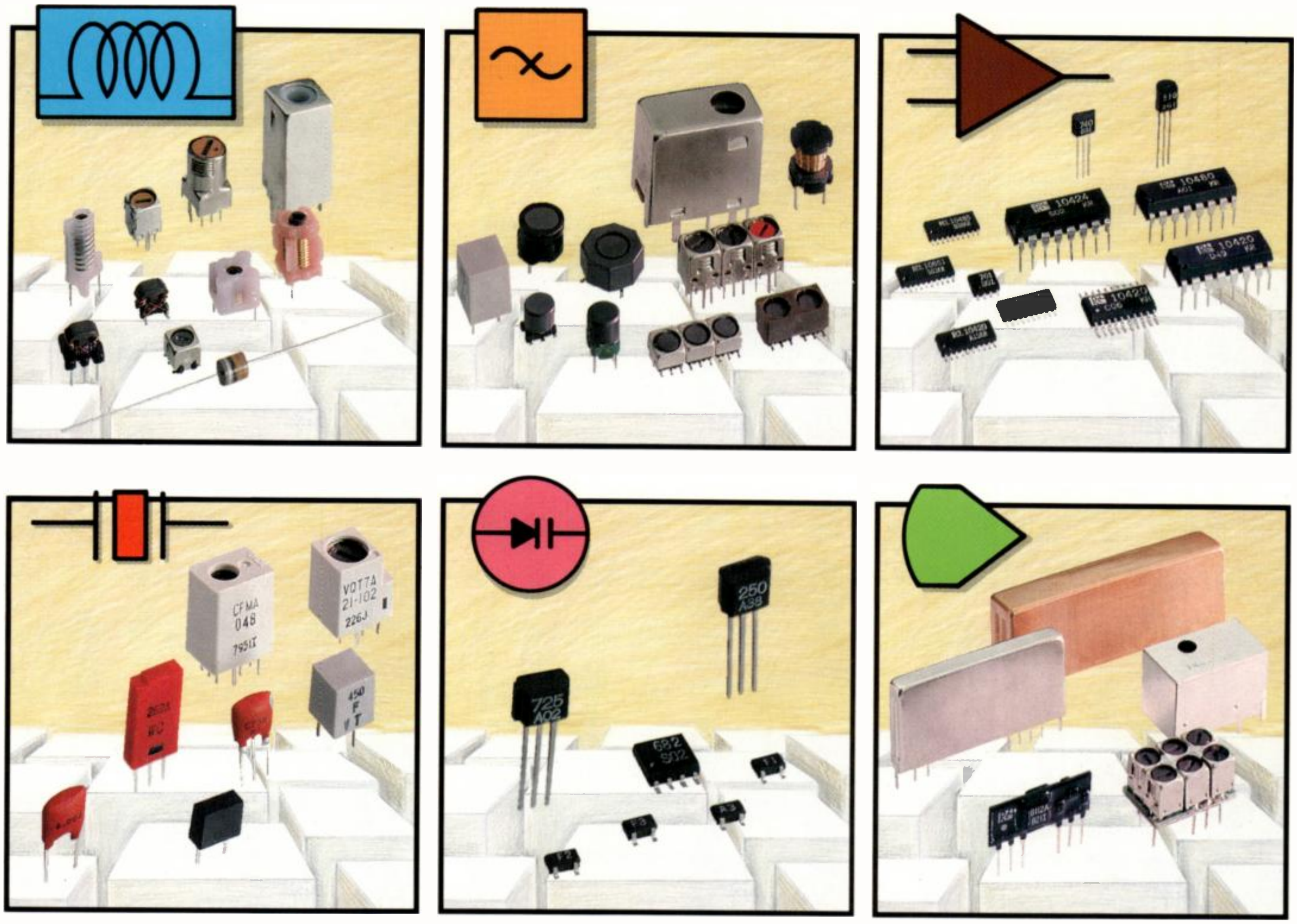
MATRIX
SYSTEMS CORPORATION

5177 NORTH DOUGLAS FIR ROAD
CALABASAS, CALIFORNIA 91302

Phone: 818-992-6776
TWX: 910-494-4975
FAX: 818-992-8521

INFO/CARD 65

Advertiser Index



RF Design Simplified. Use Toko's RF Building Blocks and Stop Going to Pieces!

Now you can reduce your RF design and component sourcing problems with just one phone call. Toko can supply more than 75% of the components required for most RF design projects, with "building blocks" that save time and space.

Toko is the world's largest manufacturer of small coils, with radial and axial leaded-inductors and SMT inductors—adjustable or fixed, in single or multiple coil configurations. We have what you're looking for.

Why design a complex LC or helical filter when Toko already makes it! We offer a wide variety from 4mm SMT styles to helical filters large enough to handle RF power signals—all in single or multiple configurations.

For I-F amplification and detection, Toko offers a variety of ICs to simplify design and reduce product size. Other available types include: compandors for noise reduction, equalizers, analog switches and voltage regulators.

For I-F or oscillator circuits, Toko ceramic filters and oscillating elements are an alternative to more costly technology. These high Q devices are also extremely compact and stable.

With over 36 stock types in a wide range of capacitance values, Toko has the varactor diodes you need, in SMT or SIP packages. An exclusive manufacturing process eliminates tracking errors in matched sets.

The fastest way to design and produce your next RF design project is with Toko RF modules. Utilizing the latest hybrid circuit technology, these circuits will also save space and cost. Custom variations are also available.

Let Toko put the pieces together for you. For a free catalog or information on samples, contact Toko America today.
(312) 297-0070; FAX (312) 699-7864.

TOKO
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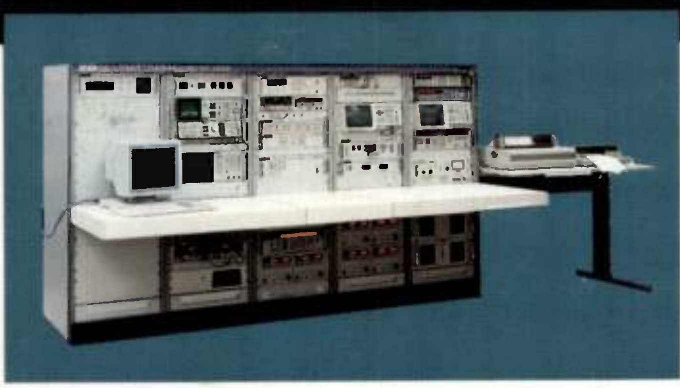
Your strategic partner…
for all the right reasons.

INFO/CARD 66



The WJ-1580 Automatic Countermeasures Test System (ACTS) provides the most comprehensive test capability for complex electronic warfare depot and intermediate-level test problems.

Designed to test both active and passive countermeasures systems in the 0.5 to 18 GHz frequency range, the WJ-1580 possesses advanced features that offer the user fast, accurate testing, maximum hardware flexibility, multiuser and multitask operations, and reduced programming costs. The WJ-1580's modular hardware design permits easy system expansion.

Watkins-Johnson Company's ACTS offers a complete testing capability for ECM assemblies such as:

- High Probability of Intercept Receivers
- High-Power Solid-State and Traveling-Wave-Tube Amplifiers
- Frequency Memory Loops (FMLs)
- Instantaneous Frequency Measurement Systems
- VCOs and Frequency Synthesizers
- RF Switching Assemblies
- RF Processing Assemblies
- Digital Control Assemblies

For full information on Watkins-Johnson Automatic Test Equipment, contact Systems Development Division Applications Engineering, 2525 North First Street, San Jose, CA 95131-1078, (408) 435-1400.

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