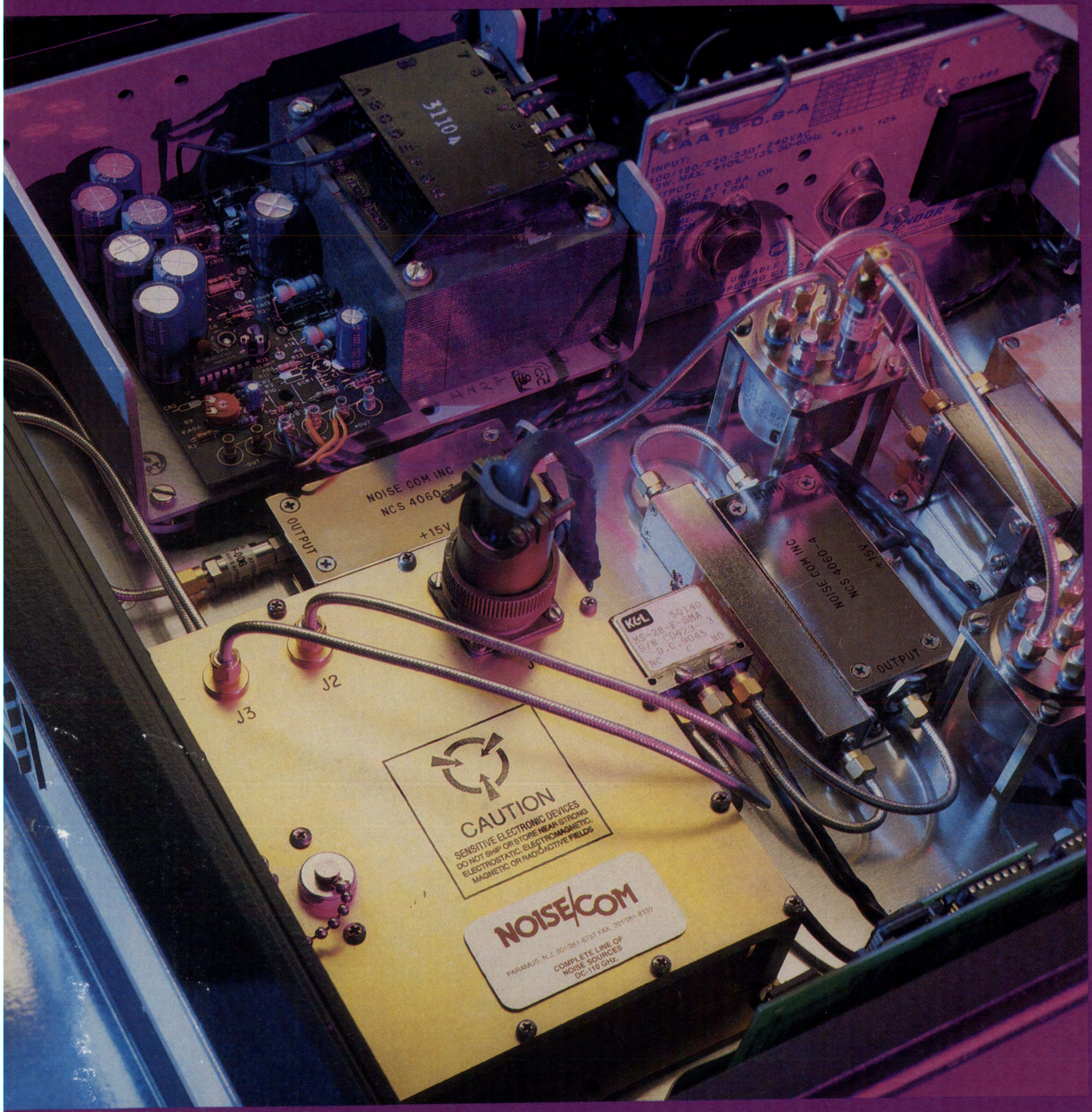
RF design

engineering principles and practices

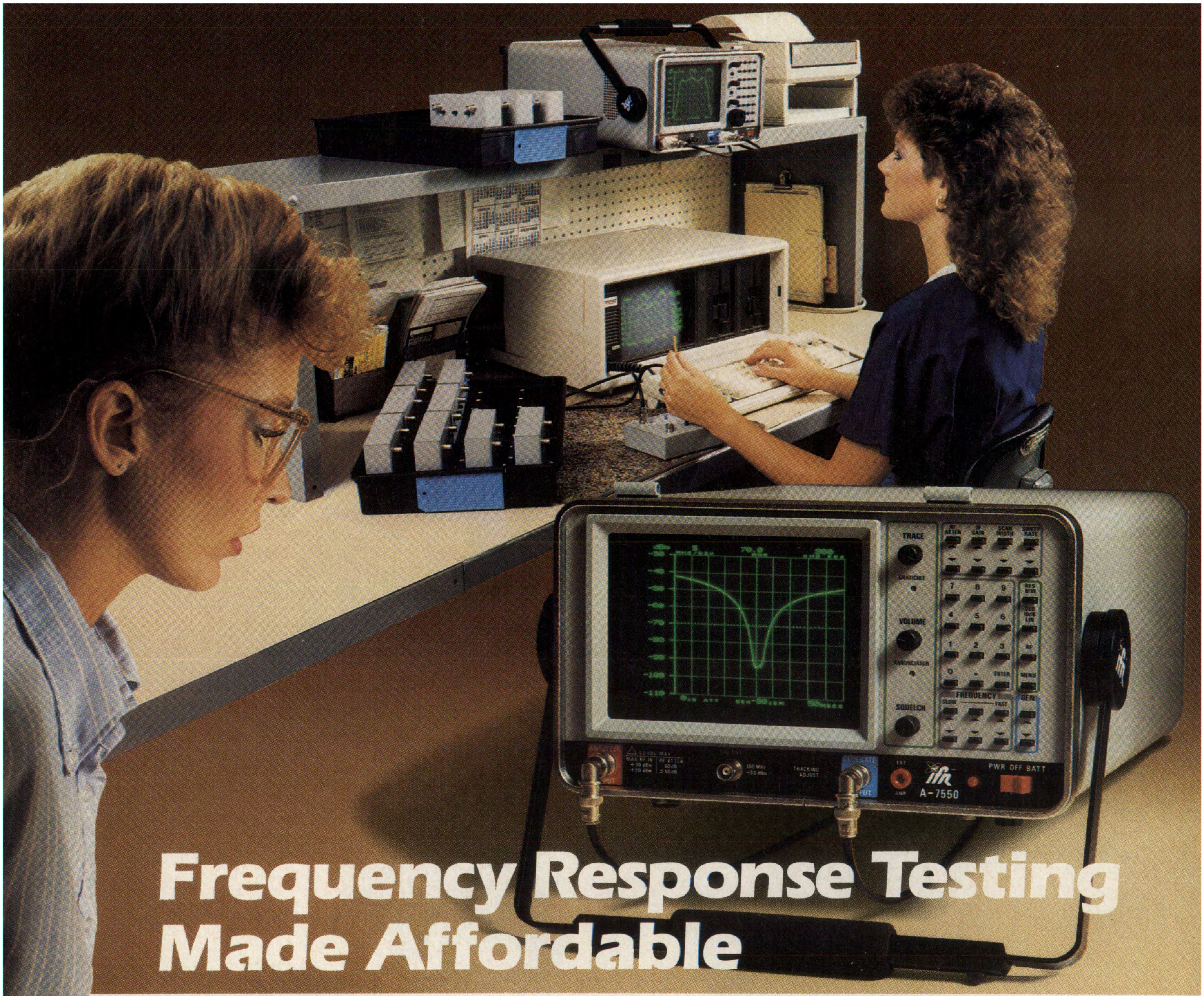
March 1991



Cover Story
New Tunable Noise Source

Featured Technology
Analog Signal Processing

Product Report
RF Mixers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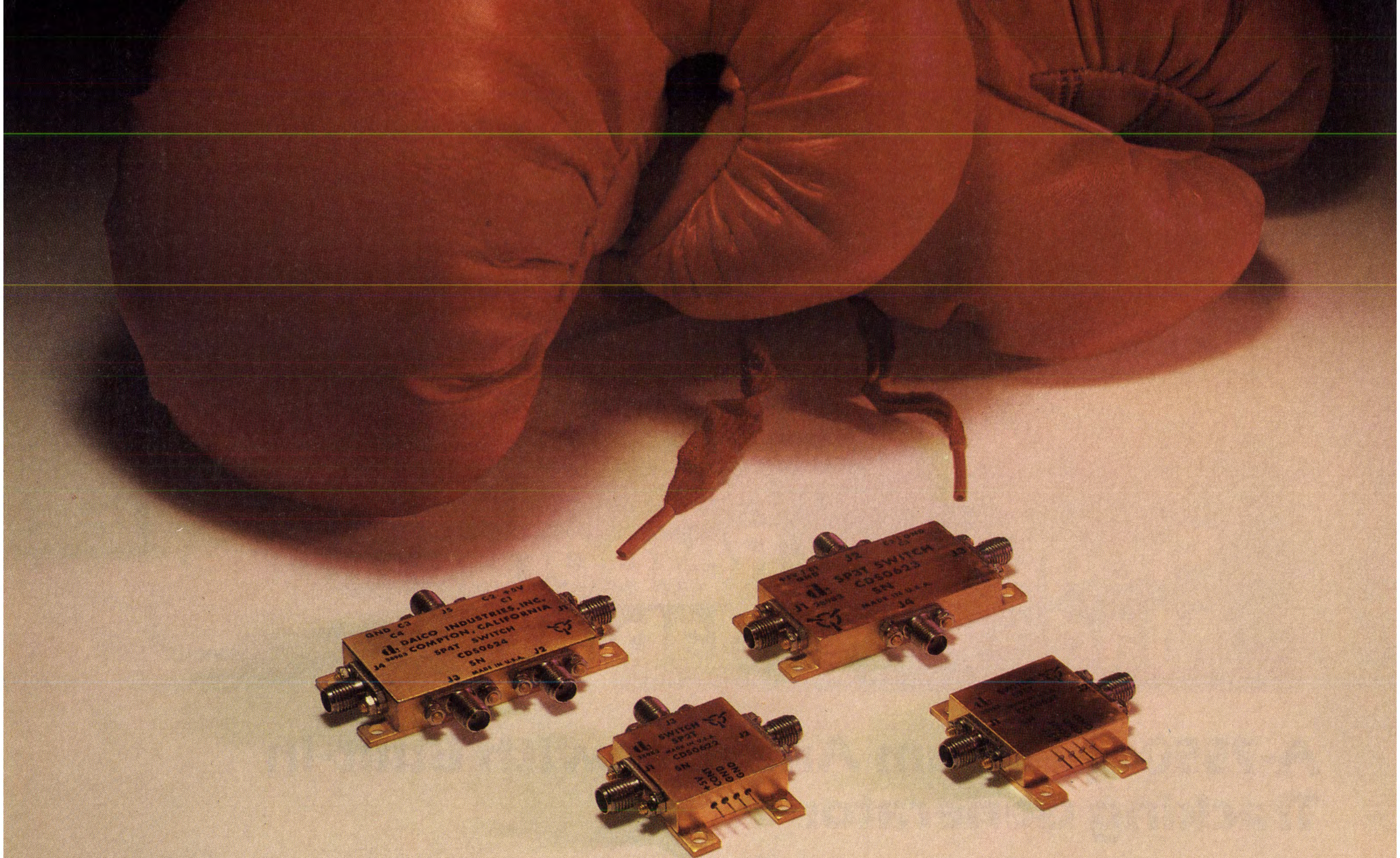
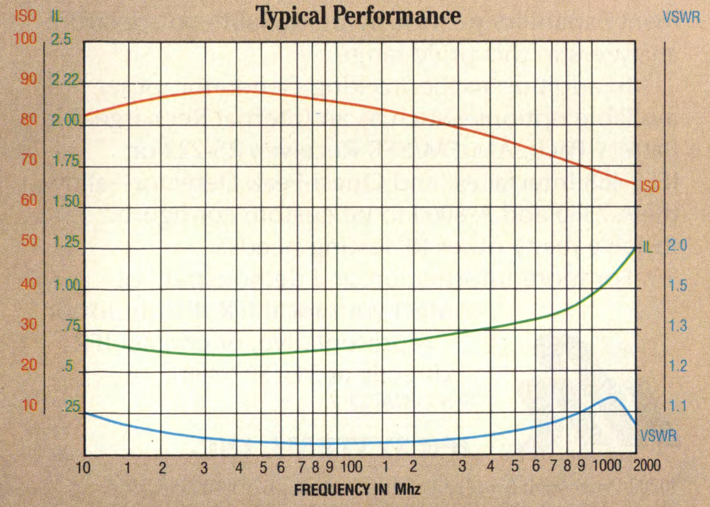
Miniature Connectorized Switches





Typical Characteristics

Frequency:	20 - 2000MHz		Insertion Loss:	0.9dB	20 - 1000MHz
Isolation:	75dB	20 - 500MHz		1.3dB	1000 - 2000MHz
	70dB	500 - 1000MHz	DC Power:	12mA at +5VDC	
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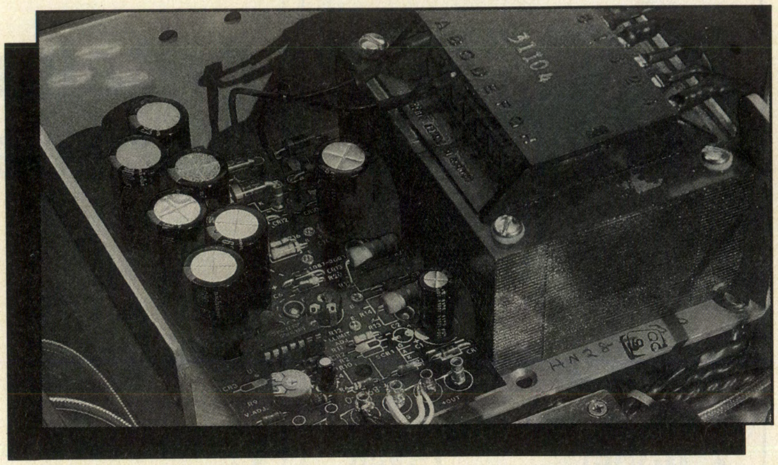

featured technology

cover story

emc corner

design awards

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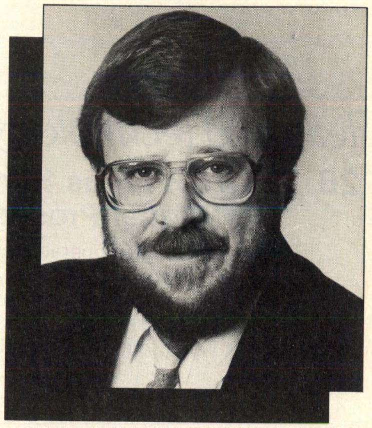
RF editorial

A Balanced Approach



By Gary A. Breed
Editor

I get lots of input about how engineers should go about their jobs. To say I've heard a hundred ways to get the job done may not be exaggerating. Every once in a while, an individual comment really irritates me; usually from an engineer insisting that his way is the only acceptable alternative. So, I feel compelled to respond to the critique of a tiny minority on behalf of the great number of reasonable engineers in our business.

A complaint I hear fairly often is directed toward engineering that is less than "perfect." For reasons unknown to me, a few engineers feel that exact numerical results and a precise implementation of those results in hardware is the only acceptable technique. Many of these practitioners are convinced that, by using the most advanced software tools and a shelf full of textbooks, an RF circuit can now be completely and accurately characterized. They, of course, forget how many approximations, algorithms, and round-offs have been used in the software and in the textbook analyses, but it doesn't matter to them anyway.

In one sense I welcome the appearance of this new group of engineers — they definitely don't see RF as mysterious, but as something that can be well understood. But they carry things too far. To them, any engineering decision that is based on experience or a rule-of-thumb is necessarily inappropriate. If a circuit doesn't work as designed, the blame goes to the technician who constructed it (forget tolerances and realistic component values).

A few years ago, I would have complained about the other side of engineering; the quick-and-dirty guys. Too much engineering was done without taking advantage of the resources at hand. Designs copied from the work of others were prevalent, with performance achieved by manual adjustment, not through proper design. I suppose it only makes sense that an over-reaction would take place.

I was told of an experiment that was done at a major military communications equipment plant: A power amplifier was to be designed, and two engineers were assigned to the project. One was a young hotshot who knew his way around the latest theory and software design tools. The other engineer was the old timer who was regarded with awe as the master of RF Black Magic. The young engineer banged away on his computer, optimizing and refining his design, while the other punched some numbers into his calculator, checked his notebook, and drew on years of experience.

They designed nearly identical circuits! You see, both were good engineers who knew how to make the best use of the tools they were familiar with. Neither the new nor the old way was superior. It was the individual who mattered.

That's my simple comment this month — it's not the process of engineering that leads to success or failure, but the ability of the engineer. A good engineer can get results without all the latest hardware and software toys, just like a mediocre engineer can create a mediocre design despite having the best tools at his or her disposal.

So, next time you are getting ready to make a critical comment to a fellow engineer because a little math was skipped or a shortcut was used, think about it. Instead of pointing out the apparent errors, ask why they chose to make such simplifications. You might find a better route to a successful design.

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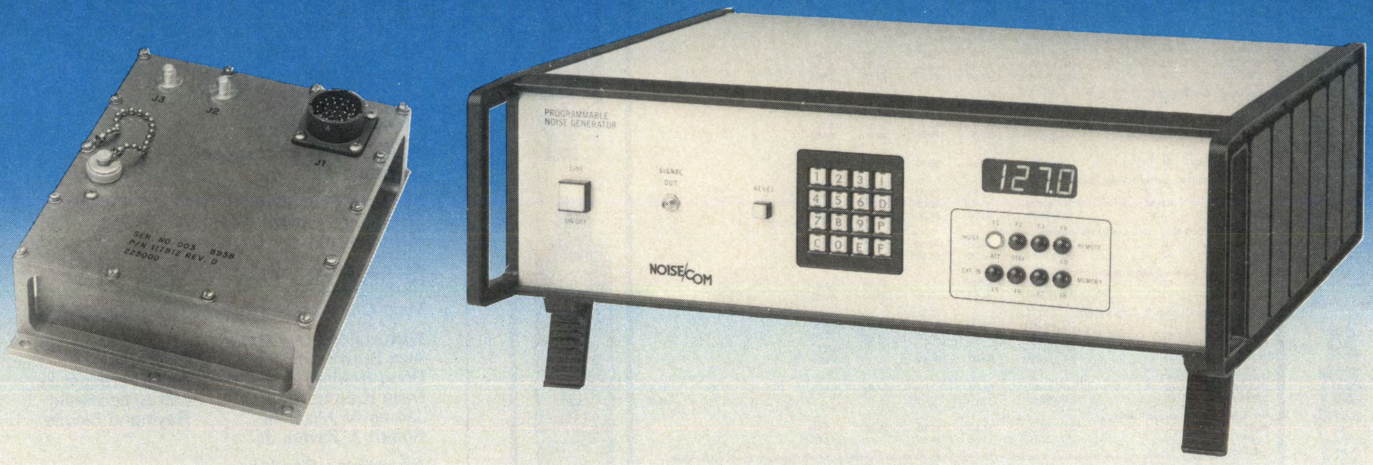

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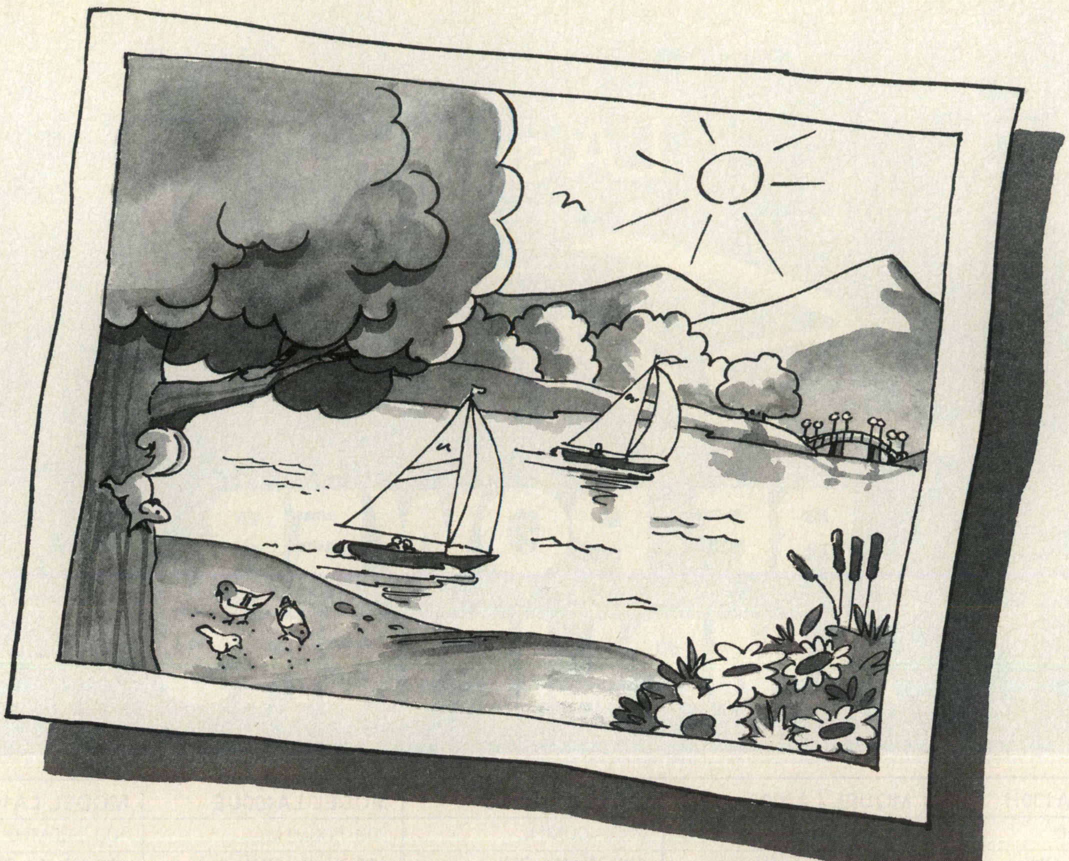

For Most People, It Was Just Another Warm September Day.

For design engineers, it was the day mixed analog/digital design came of age.
The event was the mixed-signal design demonstration at the IEEE Bipolar Circuits and Technology Meeting (BCTM). The goal was to give credence to mixed-signal simulation and to benchmark companies in the marketplace. The results were conclusive.

Viewlogic came up with the right answer first.
But more to the point, what we did at BCTM in September, we can do for you now. We're the only company with a proven technology and a three year track record of success. The only one that integrates design capture, simulation and analysis.

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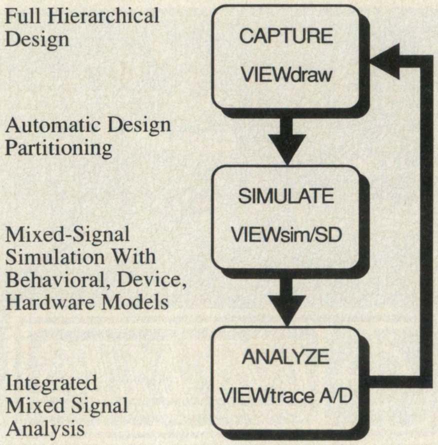
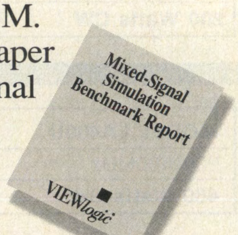


Our white paper "Mixed-Signal Simulation Benchmark Report" proves the point. **Call us at 1-800-422-4660, Ext. 103.** You'll like the climate we're creating for mixed-signal design.



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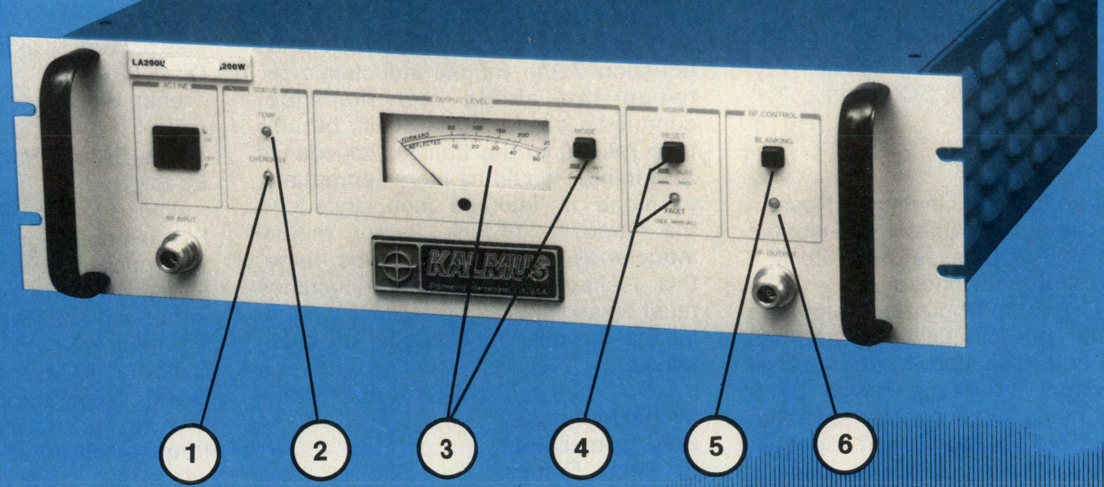

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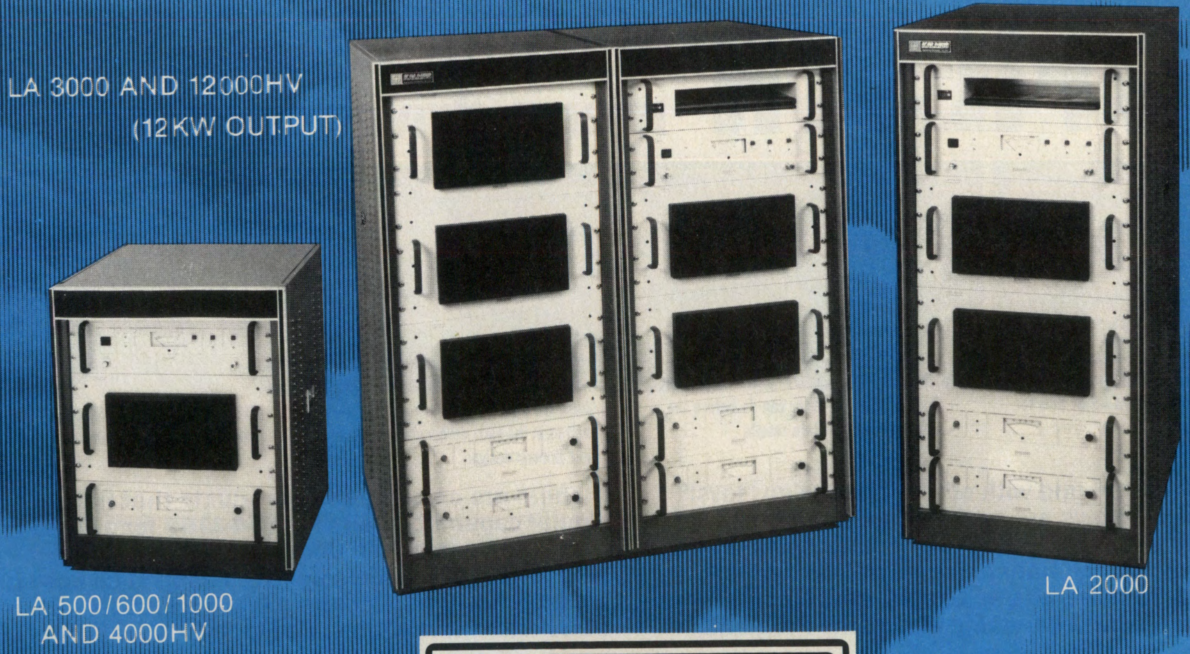

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Letters should be addressed to Editor, *RF Design*, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111.

Shielded Cables

Editor:

The interesting article in your November 1990 issue, "Application of Shielded Cables," by Thomas Jerse exhibits, I think, two slight misconceptions.

1) On Figure 1a, the circuit, although a workable one, is anything but balanced. A balanced load, to get the best CM rejection should be driven by a balanced pair, which can be shielded, provided its two wires exhibit a decent symmetry in their resistances, capacitances and inductances. As soon as a load is driven by a coaxial cable as shown in Figure 1, it becomes primarily an unbalanced circuit, with either the load floated from the chassis ground, but yet not balanced (Figure 1a) or connected to it (Figure 1b).

2) Before equation 1, the mutual inductance between the shield and the center conductor is said to be equal to the shield self-inductance L_s (i.e. both are large) which is correct. Then on equation 7 the mutual inductance between the shield and the inner conductor is referred to again, but this time related to the transfer impedance Z_T. I think it would be more appropriate, to name it 'leakage inductance.' To the opposite of the former mutual inductance, which had to be large, this leakage (or transfer) inductance must be as small as possible. Congratulations for this good tutorial on cable shields.

Michel Mardiguian
EMC Consultant
St. Remy, France

Passive Filter Software

Editor:

It would be appreciated if you published a review of passive filter (LC) software that specifies return loss.

Standards ANSI C93.41, IEC 481 specify return loss for power filters. Return loss is always more difficult to achieve than to get accepted insertion loss according to the above standards.

Dr. J.J. Hajek
Trench Electric

Wide-Band FM Demodulation

Editor:

The article "Circuits for Wide-Band FM Demodulation" December 1990, was very good. I think your readers who are interested in this topic would like to know about an application note I wrote for Frequency Sources, Inc. entitled "Low Distortion FM Generation and Detection Using Hyperabrupt Tuning Diodes." Many of the circuits operate at the same frequencies as those in the article but use a PLL circuit using the same types of ICs and an external VCO.

Jon GrosJean
Woodstock Engineering

For those readers interested in obtaining the application note, AN-110 is available from Frequency Sources, Inc., Semiconductor Division, 16 Maple Road, Chelmsford, MA 01824. Tel: (617) 256-8101. - Editor

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For applications assistance contact Jerry Hiller on ext. 2625.

M/A-COM

RF calendar

March

26-28 **International Mobile Communications Expo**
Anaheim Convention Center, Anaheim, CA
Information: April DeBaker, Cardiff Publishing Company, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111. Tel: (303) 220-0600, (800) 525-9154. Fax: (303) 773-9716.

April

7-10 **1991 US Conference on Gallium Arsenide Manufacturing Technology**
Reno, Nevada
Information: 1991 GaAs MANTECH Conference, Suite 300, 655 15th Street, N.W., Washington, DC 20005.

15-18 **36th International SAMPE Symposium and Exhibition**
San Diego Convention Center, San Diego, CA
Information: SAMPE, P.O. Box 2459, Covina, CA 91722. Tel: (818) 331-0616.

15-18 **NAB '91**
Las Vegas Convention Center, Las Vegas, NV
Information: NAB, 1771 N Street, N.W., Washington, DC 20036-2891. Tel: (202) 429-5350.

16-18 **Electro/International**
Jacob K. Javits Convention Center, New York, NY
Information: Electro, 8110 Airport Blvd., Los Angeles, CA 90045. Tel: (800) 877-2668.

28-May 1 **The 1991 MIC Workshop**
Kingsmill Resort and Conference Center, Williamsburg, VA
Information: Arlon, Microwave Materials Division, 1100 Governor Lea Road, Bear, DE 19701. Tel: (800) 635-9333.

May

13-15 **41st Electronic Components and Technology Conference**
Westin Peachtree Plaza, Atlanta, GA
Information: Electronic Industries Association. Tel: (202) 457-4930.

14-16 **IEEE Instrumentation and Measurement Technology Conference**
Omni Hotel, Atlanta, GA
Information: Robert Myers, 3685 Motor Avenue, Ste. 240, Los Angeles, CA 90034. Tel: (213)-1463. Fax (213) 287-1851.

22-25 **The 3rd International Symposium on Recent Advances in Microwave Technology**
Reno, Nevada
Information: Banmali Rawat, Chairman, Technical Program Committee, Electrical Engineering and Computer Science Department, University of Reno, Nevada, Reno, NV 89557-0030. Tel: (702) 784-6927. Fax: (702) 784-1300.

29-31 **45th Annual Symposium on Frequency Control**
Los Angeles Airport Marriott, Los Angeles, CA
Information: Annual Symposium on Frequency Control, PO Box 826, Belmar, NJ 07719.

Fundamentals of Radar Cross Section
May 20-24, 1991, San Diego, CA
Information: Kelly Brown, SCEEE, 1101 Massachusetts Ave., St. Cloud, FL 34769. Tel: (407) 892-6146.

Cellular Radio
March 25-28, 1991, Madison, WI
ESD - Electrostatic Discharge
April 17-19, 1991, Madison, WI
Information: University of Wisconsin - Madison, Department of Engineering Professional Development. Tel: (608) 262-2061. Fax: (608) 263-3160.

Radar Operation and Design: The Fundamentals
March 25-28, 1991, Washington, DC
Grounding, Bonding, Shielding and Transient Protection
April 8-11, 1991, Orlando, FL
Spread Spectrum Communications Systems
April 8-12, 1991, London, England
June 10-14, 1991, Washington, DC
Lightning Protection
April 11-12, 1991, Orlando, FL
Analog/RF Fiber-Optic Communications
April 24-26, 1991, Washington, DC
Microwave Systems Engineering
April 29-May 3, 1991, Washington, DC
Vulnerability of Spread Spectrum AJ and LPE Communications Systems
May 6-9, 1991, Washington, DC
Electromagnetic Interference and Control
May 6-10, 1991, Washington, DC
Nonlinear Digital Signal Processing and Applications
May 6-10, 1991, Washington, DC
Electromagnetic Interference and Control
May 6-10, 1991, Washington, DC
Cordless and Microcellular Business Telephony
May 13-15, 1991, Washington, DC
Frequency Hopping Signals and Systems
May 20-22, 1991, Washington, DC
Principles of High Frequency Radio Communications: Applications for Operators and Managers
May 20-23, 1991, Washington, DC
Information: The George Washington University, Continuing Engineering Education, Merril A. Ferber. Tel: (202) 994-8522 or (800) 424-9773.

Digital Signal Processing Workshop
April 10-12, 1991, Norwood, MA
June 12-14, 1991, Norwood, MA
Information: Analog Devices, DSP Applications Department, Maria Butler. Tel: (617) 461-3672.

Digital Signal Processing: Filtering and Estimation
March 18-21, 1991, United Kingdom
Broadband Telecommunications
March 18-22, 1991, United Kingdom
Modern Digital Communications for Space, Satellite and Radio
April 15-18, 1991, Italy
RF and Microwave Circuit Design I: Linear Circuits
April 15-19, 1991, Pisa, Italy
RF and Microwave Design II: Non-Linear Circuits
April 22-26, 1991, Pisa, Italy
Digital Microwave Systems: Theory and Applications
April 22-25, 1991, Pisa, Italy
Satellite Communication and Broadcasting
April 22-26, 1991, Italy
Information: CEI-Europe/Elsevier, Mrs. Tina Persson, Box 910, S-612 01 Finspong, Sweden. Tel: +46 (0) 122-17570. Fax: +46 (0) 122-14347.

DSP Without Tears™
March 20-22, 1991, San Jose, CA
Information: Right Brain Technologies. Tel: (404) 664-6738. Fax: (404) 420-3834.

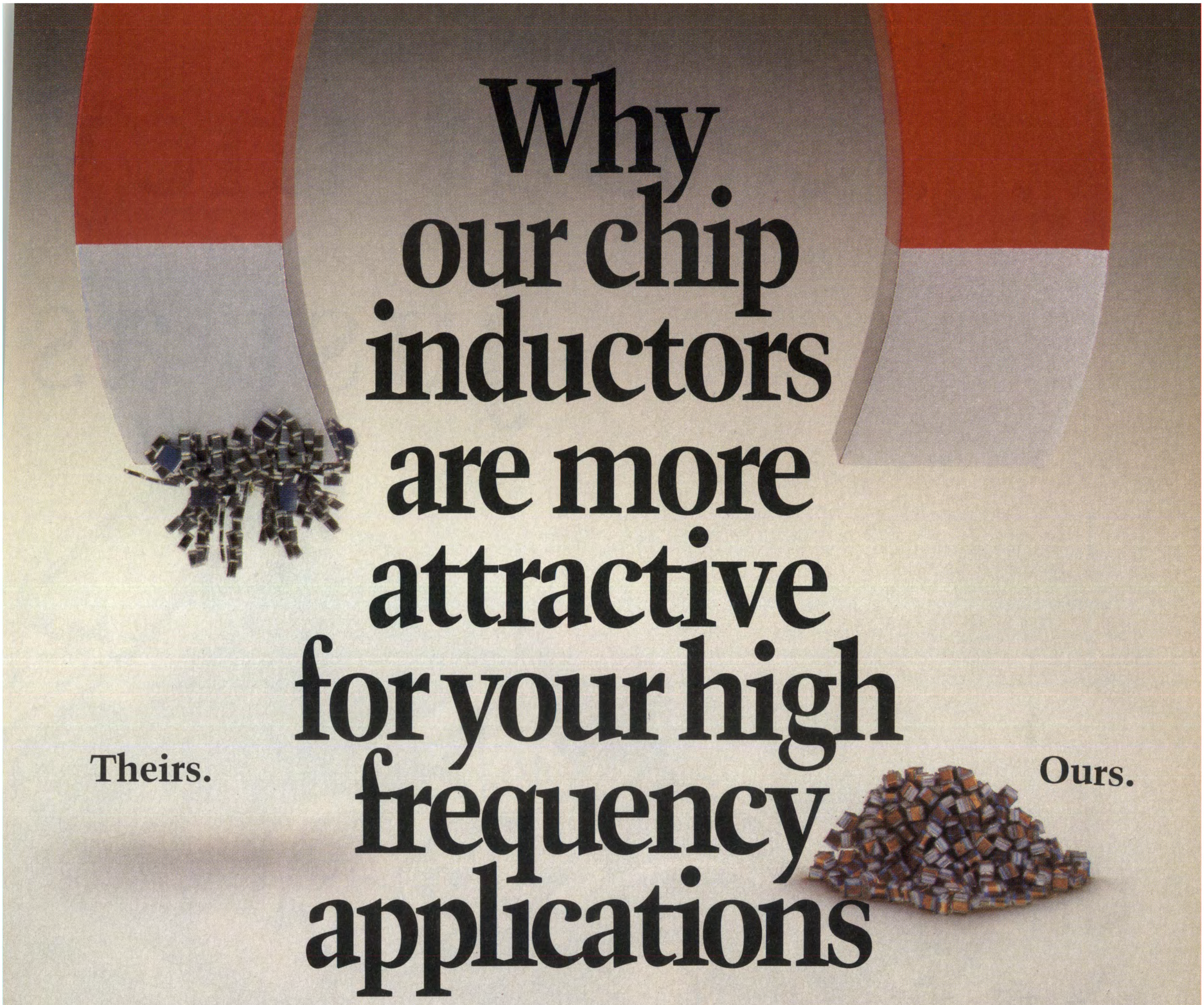
RF/MW Linear/Nonlinear Circuits and Applications
March 14-20, 1991, Palo Alto, CA
RF Component Modeling: CAE and Measurements
March 25-27, 1991, Santa Clara, CA
Information: Besser Associates. Tel: (415) 949-3300, Fax: (415) 949-4400.

Modern Power Conversion Design Techniques
April 29-May 3, 1991, Phoenix, AZ
May 20-21, 1991, San Rafael, CA
Information: e/j Bloom Associates, Joy Bloom. Tel: (415) 492-8443. Fax: (415) 492-1239.

EW Receivers
May 7-9, 1991, Syracuse, NY
ELINT Analysis
May 7-9, 1991, Syracuse, NY
ELINT Interception
May 14-16, 1991, Syracuse, NY
ELINT/EW Applications of Digital Signal Processing
May 14-16, 1991, Syracuse, NY
Integrated EW
May 21-22, 1991, Syracuse, NY
ELINT/EW Data Bases
May 21-23, 1991, Syracuse, NY
Information: Research Associates of Syracuse. Tel: (315) 455-7157.

Introduction to EMI
April 15-16, 1991, Philadelphia, PA
Understanding and Applying MIL-STD-461C
April 17-19, 1991, Philadelphia, PA
Design for EMC
April 22-23, 1991, Philadelphia, PA
Practical EMC Retrofits
April 24-26, 1991, Philadelphia, PA
Information: R&B Enterprises. Tel: (215) 825-1960.

Introduction to EMI/RFI/EMC
March 19-21, 1991, Anaheim, CA
Grounding and Shielding
April 2-5, 1991, Washington, DC
Practical EMI Fixes
April 22-26, 1991, Boston, MA
High Speed Digital Design for EMC
April 9-12, 1991, Seattle, WA
EMI Control Methodology and Procedures
March 19-22, 1991, San Francisco, CA
System Integration and Design for EMC
April 16-19, 1991, Boston, MA
Information: Interference Control Technologies, Registrar. Tel: (703) 347-0030.



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See our catalog in Vol. A, Section 1800
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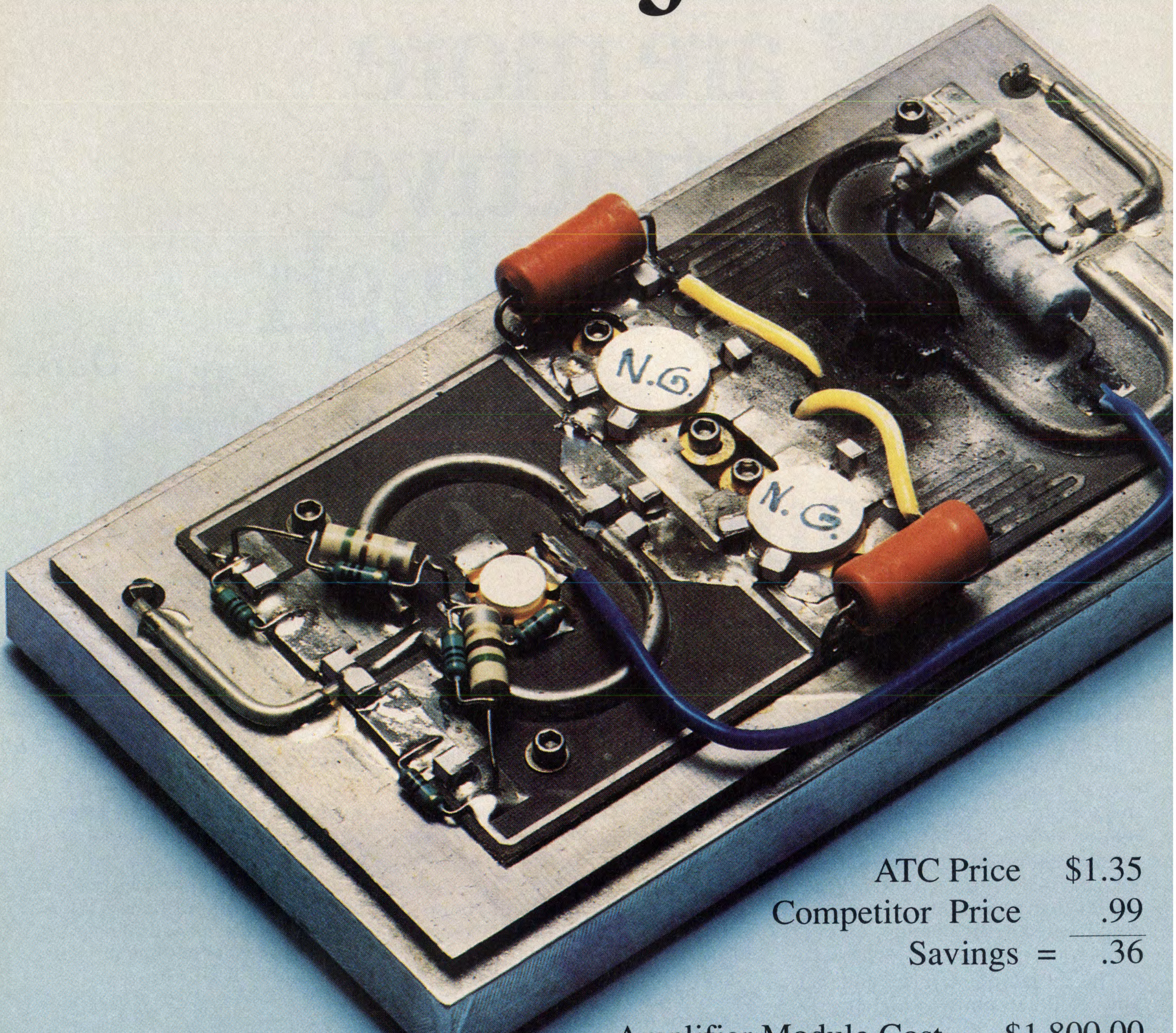

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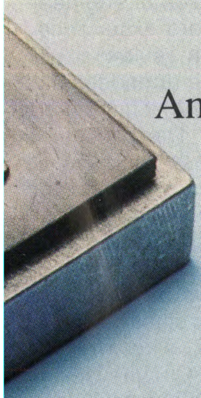
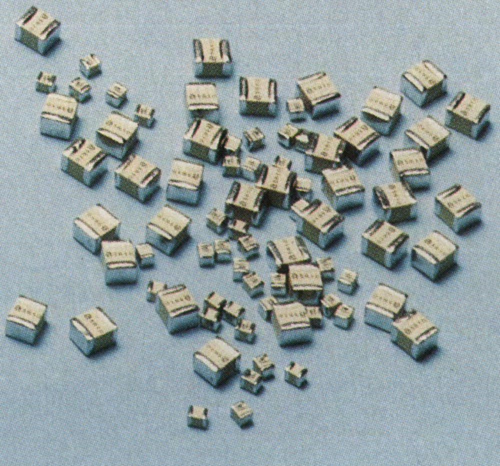




3700 Attend RF Expo West 1991

Over 2300 visiting engineers and nearly 1400 exhibitor personnel filled the Santa Clara Convention Center during February 5-7. New products announced in the February issue of *RF Design* and shown to the public for the first time were examined as engineers took time to investigate solutions to their current design tasks.

Other design solutions were presented in 55 technical papers, covering topics starting with a tutorial on PIN diodes, and ending with a method for tuning microwave filters. Sessions on RF Diodes, Power Amplifiers, Frequency Synthesis, GaAs MMIC Applications, Receivers, Oscillators, and Mobile Radio Systems saw a broad range of interest and good attendance. Many other papers were extremely well-received by their audiences. Unfortunately, four speakers were unable to attend due to travel restrictions. These papers are all contained in the RF Expo West Proceedings, giving interested engineers an opportunity to read them and contact the authors with questions and comments.

News coverage was given to RF Expo West by KGO television in San Francisco. That station broadcast a story including interviews with representatives of several companies whose components are part of the military hardware being used with success in the Persian Gulf war to drive Iraq from Kuwait.

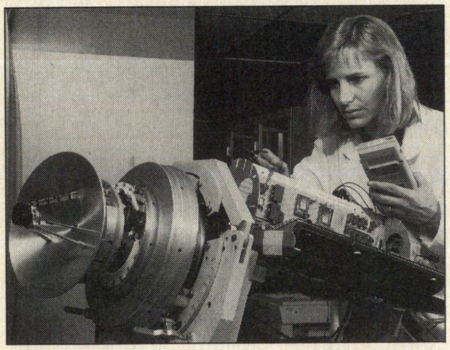
The four all-day courses again saw strong attendance by interested engineers. A quick survey indicated that many of these engineers were either young RF engineers or were changing job tracks from other areas of engineering. Many of the rest were experienced engineers looking for a review of RF basics. Only a few of the attendees had been at RF Expo the last time it was held in Santa Clara in 1989. The continuing strong interest in these courses is evidence that RF remains a growing and vital engineering discipline.

55 engineers attended the Oscillator Design Principles course taught by Randy Rhea of Eagleware, continuing the good attendance this class saw in its first two presentations. The two-part Fundamentals of RF Circuit Design taught by Les Besser of Besser Associates had well over 100 students, like it has at nearly every previous RF Expo. Also seeing good interest despite being presented concurrent with Part I of the RF Fundamentals class was Randy Rhea's Computer-Aided Filter Design course.

Exhibitors were nearly unanimous in their observation that Silicon Valley has many engineers working in RF and related applications, in addition to the more publicized digital and chip design engineering areas. Those companies with a direct involvement in non-traditional RF applications noted that the attending crowd included many potential customers. The comment was overheard many times that RF-like engineering for digital, medical, or control systems is increasing rapidly.

RF Expo East 1991 will be held October 29-31 at Stouffers Orlando Resort in Orlando, Florida. RF Expo West 1992 is scheduled for March 18-20, 1992 at the San Diego Convention Center, San Diego, California.

Smart Munitions, Countermeasures Featured in Gulf War — The war against Iraq is the first use of many U.S. weapons systems outside the test range, and the results have been generally impressive. Infrared, laser, and radar-guided bombs have been used to keep the U.S. attacks as precise as possible, as military commanders seek to destroy the ability of Iraq to wage war without causing much "collateral damage" (civilian casualties). Among the various systems, the Tomahawk cruise missiles appear to have been a major success, often delivering their explosives within a few feet of the target after a low-level flight of several hundred miles. HARM and SLAM missiles have struck radar and tactical targets, with their exploits vividly recorded by military strike cameras.



It will be weeks or months, however, before we know how effective this "surgical" strategy has worked. There are indications that some very primitive means have been used successfully by Saddam Hussein's forces to minimize the impact of U.S. attack. For example, it is reported that thousands of miles of buried cable was installed to back up radio and microwave links between Baghdad and the Kuwait front lines. The now-infamous SCUD missiles have clearly demonstrated the ability of mobile systems to avoid detection. Also, the real-time communications offered by radio, television and satellite have altered the way both propaganda and news are handled by both sides. Without the benefit of complete information, an analysis is not yet possible, but in a few months we will see the first combat performance review of recently developed weapons, communications and countermeasures systems.

Gauging MMIC Measurement Needs — NIST has launched a government/industry consortium to establish needed national MMIC standards.

Before MMICs can be successfully developed commercially, there are numerous obstacles, largely measurement related, to overcome. MMIC devices need standardized, precision-engineered components to ensure that devices from different companies can be integrated into working systems. The group is also examining problems such as how to reliably gauge temperature in a MMIC circuit and how to best measure the characteristics that promote quality control in MMIC device production.

Antenna Specialists Acquires Majority Interest in Grayson Electronics — The Antenna Specialists recently announced that it has acquired a majority interest in Grayson Electronics Co. Grayson is a designer and manufacturer of high-technology electronic communications products and will operate as an independent subsidiary.

Electronics Sales for 1990 Reaches $282 Billion — The EIA announced that the 1990 U.S. factory sales of electronic products reached nearly $282 billion, representing a 4.2 percent increase over last year's $270 billion. The communications equipment sector reached $71.3 billion or 6.1 percent over last year's $67 billion. Electronic components hit $56.6 billion and the computer and industrial sector reached $91.9 billion.

In-Flight Phone Given FCC Approval for Air-to-Ground — The Federal Communications Commission granted the applications of In-Flight Phone Corporation to provide nationwide air-to-ground radio telephone service. IFPC has spent more than $8 million over the last 18 months designing a new digital air-to-ground system that will reduce the airlines' operating costs and will provide high quality voice communications and data services for passengers.

Anadigics and Thomson Sign Second-Source Agreement — Thomson Composants Microondes (TCM) has signed an agreement with Anadigics to be a worldwide second source for Anadigics' Ku-band direct broadcast satellite downconverter gallium arsenide integrated circuit chip set. Under the agreement, TCM will manufacture the DBS products by its own GaAs process but will maintain full functional interchangeability with Anadigics' DBS devices.

Hewlett Packard and Telogy Win Lockheed Contract — A partnership between Hewlett Packard and Telogy, Inc. has been awarded an exclusive contract to provide all instrument leasing services to Lockheed Missiles & Space Company. Inc. The 3 year contract, estimated at $20 million, will provide Lockheed with test equipment on an "as needed" basis and builds in a level of equipment management that minimizes the internal resources needed to track the equipment. HP will be the lessor within the partnership, owning and servicing the test equipment. Telogy will provide equipment management services.

North American Van Lines Signs Order With QUALCOMM for Mobile Satellite Communications System — North American has signed a $12 million contract with QUALCOMM, Inc. to outfit their Commercial Transport employee fleet as well as part of its High Value Products division with OmniTRACS[R] two-way satellite communications units. The contract includes OmniTRACS hardware, software and initial messaging services.

Teradyne Sells Test Systems to Silicon Systems — Teradyne, Inc. announced the receipt of orders from Silicon Systems for an A520 Analog VLSI Test System and three A520C Communications Test Systems. The systems are being used in both the Tustin and Singapore facilities to expand capacity for production testing of microperipheral and communications devices. The orders, the values of which were not disclosed, were delivered in the third and fourth quarters of 1990.

Siliconix Becomes AEG Subsidiary — Under a reorganization plan, AEG Capital Corp., has become an 80.1 percent shareholder of Siliconix. As a result, Siliconix, Inc., is no longer under Chapter 11 bankruptcy protection. According to the January 14 edition of *Electronic News*, AEG will pay International Rectifier $12.3 million plus interest as well as $5.8 million immediately in exchange for a royalty-bearing license for various MOSFET patents,

Ball Antenna System Receives FAA Type Certification — Ball Corporation's Airlink™ low-gain antenna system, along with Rockwell Collins' Avionics, was recently FAA type certified as a result of a Boeing flight test conducted aboard a United Airlines 747-400 aircraft. Type certification means that the Ball antenna is included in the airplane's FAA certification, and it has been approved for use on that aircraft with the specified avionics.

New Address for Inter-Continental Microwave — Inter-Continental Microwave recently announced their move to larger facilities. Their new address is 1515 Wyatt Drive, Santa Clara, CA 95054. Their phone and fax numbers remain the same.

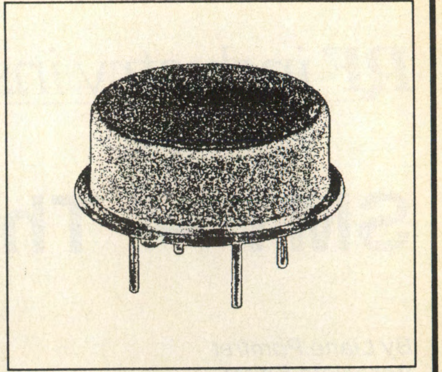
EMV Expands in Europe — Founded by Amplifier Research and EMV GmbH, EMV S.A.R.L. will provide engineering assistance, sales, and service throughout France to the growing number of manufacturers concerned with protecting their products against electromagnetic vulnerability.

Kings Electronics Company Incorporates Kings Electronics (Canada) Limited — Kings Electronics Company recently announced that Kings Electronics (Canada) Limited has ceased to exist as an active company. Kings is making this change because of the new trade regulations between the United States and Canada that will enable to establishment of closer working relationships with their Canadian customers. All correspondence should be addressed to Kings Electronics Company, Inc., 40 Marbledale Road, Tuckahoe, NY 10707.

RF Products, Inc. Formed — RF Products, Inc. has been formed to manufacture high power products for use in RF and microwave applications. The company will be able to provide "drop-in" replacement transistors of the Acrian and CTC types, high power passive resistor, termination, and attenuators for both standard and custom applications, bias devices for temperature compensating high power class A and AB transistors, and broadband 50 ohm transistor unit amplifiers.

EIA Information Systems Forecast Available — The *EIA Information Systems Forecast for the 1990s* is available from the Government Division Requirements Committee. The report includes a five year forecast of the Information Systems market as well as other market research data. The price of the report is $125 for EIA Government Division Member Companies and $250 for non-member companies. Contact Mary Lamb at (202) 457-4942 for ordering information.

Call for Papers — The International Institute of Connector and Interconnection Technology is accepting abstracts for technical papers for presentation during its 24th annual Connectors and Interconnection Technology Symposium, to be held October 6-9 in San Diego, Calif. Papers are encouraged in the following areas: printed wiring applications, data and documentation requirements, standardization and reliability, termination and connector techniques, fiber optics, backplane interconnection wiring and international technology. Abstracts, due by March 16, should contain a minimum of 200 words, the author's name, company and phone number. They may be sent to: Papers Chairman, 24th Annual Connectors and Interconnection Technology Symposium, 104 Wilmot Road, Suite 201, Deerfield, IL 60015-5195.



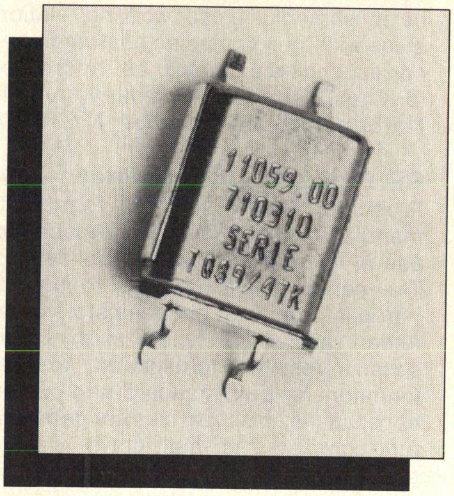
SMT — The Footprint of the Future

By Liane Pomfret
Associate Editor

Surface mount technology is relatively new in the RF industry. It has only been within the past few years that RF designers have started looking at surface mount as a viable alternative to through-hole design. With RF applications appearing in new places, surface mount technology can help solve problems caused by high frequency or lack of space. The attractions to surface mount are varied, as Steve Skiest, RF Product Manager at Toko notes, "It's not only size, it's manufacturability and automation as well." However, the lack of standards and compliance to them is causing a great deal of confusion and frustration among manufacturers and buyers alike.

The issues surrounding the switchover from through-hole to surface mount are varied. There are several reasons why a manufacturer might make the switch. Marty Markson, Vice President of Sales and Marketing at Sprague-Goodman observes, "There are basically two advantages, the first being increased circuit density or smaller size of the completed product. The other aspect of surface mount technology is the capability for automatic placement or assembly techniques." The space savings can be enormous as we have seen for portable cellular phones, pagers and laptop computers. In addition, surface mount offers better performance at high frequencies than through-hole design. Cost is also a consideration. With surface mount products, there's no need to drill holes or line up component leads with the holes. These items add to the cost of through-hole design and in many cases, surface mount design comes out costing less. In addition, manufacturers can sometimes use the same equipment they have been using for through-hole assembly, thereby saving even more money.

Given these reasons for switching to surface mount, it is only a matter of time before most manufacturing is done this way. Sandy Cohen, Engineering Sales Manager for Raltron predicts, "We think that in the long term the way to go is surface mount, especially with the types of companies manufacturing surface mount." As the technology becomes available, manufacturers are making the switch. Some have already completed their conversion to surface mount manufacturing while others are just starting. Larry Barbary, International Sales and Marketing Director for M-tron observes, "We've started to see more surface mount applications than through-hole applications in the past year." While there will always be a place for through-hole devices, surface mount technology's advantages will make it the method of choice in coming years.



Standards

Despite the advantages, there are some problems that still need to be resolved. Perhaps one of the largest is the lack of standardization within certain sectors of the surface mount industry. Parts such as capacitors and resistors, that have been around for years, have well established standards. However, for crystals, oscillators, filters and the like there is no industry standard as of yet and it is apparent in the variety of surface mount packages available. "The industry hasn't settled down on a standard footprint," says Kory Stone, Product Marketing Manager for Crystal and Clock Oscillator Products at CTS. As a result, manufacturers will set their own standard instead of waiting for an industry consortium, which makes for some inconsistent standards. Among the simpler surface mount parts, standards appear to have settled. "It seems that standards for monolithic chips that are larger than 1206 are pretty good. Anything smaller and there's not a lot of consistent literature. Also reliability information is sketchy," says Mark Nicholson, Manufacturing Engineer at John Fluke Manufacturing. While there are some extremely small packages out there, they are few and far between. Signetics manufactures an SSOP - shrink small outline package which is 1/3 the size of standard surface mount packages, and is "the worlds smallest commercially available 20-pin package," according to Michael Sera, RF Marketing Manager at Signetics. If sizes or footprints vary between manufacturers, it can be very difficult for a buyer, especially if one manufacturer discontinues a product line.

Even with this uncertainty, surface mount is becoming more and more popular. For example, the computer industry is just beginning to use RF surface mount technology. Over the past few years, computer speeds have increased and are now running around 50 MHz. For the computer designer this brings up some special RF problems that may not have existed at lower speeds. These designers are finding that RF surface mount products are a good way to avoid problems with stray inductance and capacitance brought on by long leads used in through-hole devices. Bruce Malcolm, President of Trilithic Corporation notes that "By 1993-94 time frame, clock speeds will be around 75-100 MHz. You really have to start paying attention to your RF design or you will start seeing unacceptable digital pulses." In the case of computers, surface mount offers not only a reduction in RF caused problems, it also allows for extremely small packages.

Surface mount will be the manufacturing method of the future. There will always be a place for through-hole, but with the advances being made in surface mount - in size, performance, manufacturability and materials - there is no reason why the changeover shouldn't occur in the near future. **RF**

SWOP Amps Simplify RF Signal Processing

By Tom Anderson
Burr Brown Corporation

If you design systems for RF and video applications requiring amplifiers, multiplexers, or complex functions like high performance programmable gain amplifiers, you have probably noticed that finding the right components to do the job is not simple. Using operational amplifiers and switches is fine for slow signals, but as signal frequencies increase, so do the problems that must be addressed.

A pair of new SWOP[R] Amps from Burr Brown, the OPA675 and OPA676, open the door to high speed, high accuracy applications without the headaches. These devices are true operational amplifiers with a built-in switchable front end (hence the name SWOP Amp, for SWitchable input OP amp).

The SWOP Amp has two identical input stages that share an active load and output stage. One input stage at a time is active and controls the output of the amplifier. The unselected output stage has a very high input impedance and very little crosstalk to the active input stage. The resulting integrated circuit functions as an operational amplifier with a DPDT switch on the two input pins (see block diagram and specifications in Figure 1). Not only is the SWOP Amp an ideal component for simplifying the design of an analog signal processing system, the performance is outstanding. With gain-bandwidth product greater than 2 GHz, channel select time less than 5 nS, low distortion, and operational amplifier flexibility, the SWOP Amp can improve the dynamic and DC characteristics of many high speed analog systems while reducing complexity and cost. These devices have been optimized for driving flash convertors, and will handle 75 and 50 ohm cables as well.

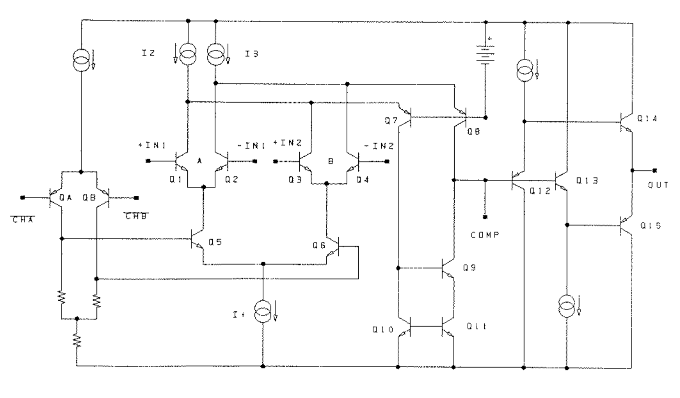
Fast Switching Multiplexer With Gain

One of the most widely used circuits for modern systems is the input signal multiplexer, and multiplexing RF or video signals requires special attention. Using a SWOP Amp makes it easy to switch from one input signal to the other in just a few nanoseconds, and even include signal gain in the bargain.

Figure 2 shows two of the many possible multiplexer configurations. The OPA676 provides two input channels that are multiplexed into the single output. The gain for each channel is set by external feedback resistors according to standard operational design techniques. In Figure 2a, R1 and R2 set the voltage gain:

$$A_v = 1+(R2/R1) = 8$$

The inverting inputs for both input stages are connected to the same feedback network, guaranteeing that the gains will be identical for both channels without requiring precision matched resistors. This principle can be especially useful when designing active filters, since the bandwidths of both channels may be made identical by design without



Channel switching delay	4.6 nS (ECL select, OPA675) 6.0 nS (TTL select, OPA676)
Slew Rate comp cap = 0	1000 V/uS
comp cap = 5pF	400 V/us
Settling time 1.25 V step, Gain=10	1% 9 nS .1% 15 nS .01% 25 nS
Small signal bandwidth	
A_v = 50, comp cap = 0	50 MHz
A_v = 10, comp cap = 5pF	185 MHz
A_v = 2, comp cap = 35pF	100 MHz
Noise fb = 10 Hz to 10 MHz	8 uV rms
Vos, each channel	500 uV typical, 2 mV max
Input bias current	20 uA
CMRR Vcm +1.25V to −1.25V	100 dB
Output swing 150 ohm load	+1.25V to −1.25V min
50 ohm LOAD	+1V to −1V typ
Open loop gain 150 ohm load	70 dB
PSRR +V = +5.5 to +4.5, −V = −5.7 to −4.5	86 dB
Power dissipation	200 mW typical

Figure 1. Simplified circuit diagram of the OPA675 and measured performance statistics.

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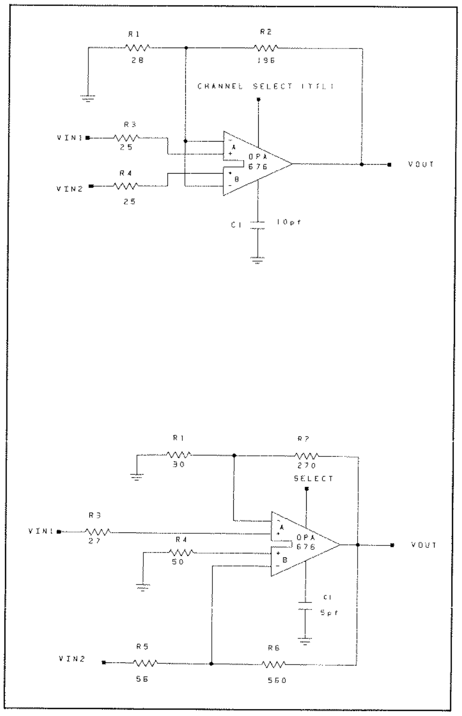
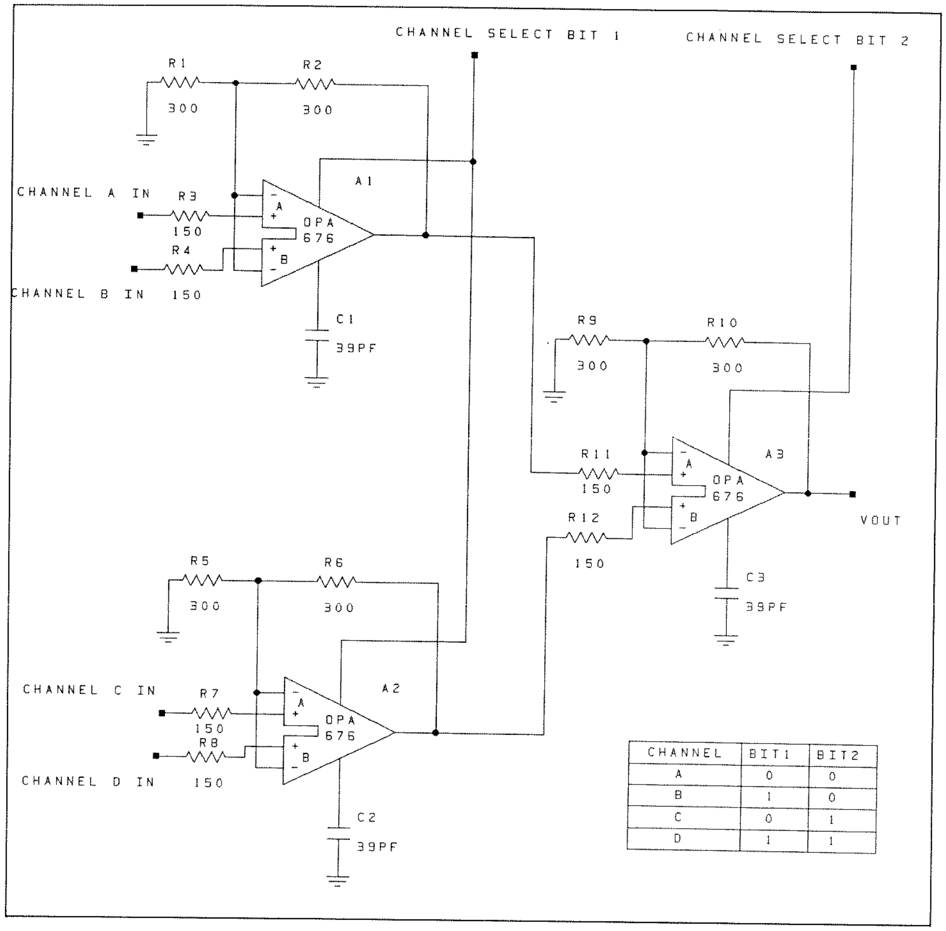


Figure 2. Two input multiplexer circuits, (a) both channels have identical gains of 8 without precision resistors, (b) separate feedback networks give different gains for each channel. Gain is +10 for channel A and −10 for channel B, with bandwidth greater than 100 MHz.



CHANNEL	BIT1	BIT2
A	0	0
B	1	0
C	0	1
D	1	1

Figure 3. A fast four input MUX using three SWOP Amps $V_{out} = 4V_{in}$ for the selected channel. Bandwidth is 70 MHz.

precision components or manual adjustment. Filtering may be added to any of the circuits shown here using standard operational amplifier filter designs.

R3 and R4 are used to balance the impedance for the two non-inverting inputs, canceling the contribution to offset error from input bias current flowing in R1 and R2. The value for R3 and R4 is selected to be equal to the parallel combination of R1 and R2. R3 and R4 could be deleted, but this will add an extra term to the offset equal to

(25 ohms x 20uA) = 500 uV.

For many RF systems offset is unimportant, and the values for R3 and R4 are selected to correctly terminate the source impedance. If different gains are desired for the two channels, separate feedback networks may be used as shown in Figure 3. Now channel B has a separate feedback network, and is configured for an inverting gain of 10:

$A_v = -(R6/R5) = -10.$

R4 is connected between the non-inverting input for channel B and ground to cancel Ib x R errors as described above, but may be deleted if offset is not critical.

Note that the compensation capacitor is connected to a node that is common to both channels. The capacitor must be large enough to make the channel with the lowest gain stable.

These multiplexer circuits allow channel selection in less than 5 nS, with settling to 0.1 percent accuracy only requiring about 9 nS. Small signal bandwidth is over 60 MHz in either of the configurations shown above.

Besides providing two input channels for a signal processor, these multiplexer or selectable input channel circuits have many other applications. The selectable input amplifier can also function as a gated amplifier, where one input is connected to the signal and the other is connected to ground. During "normal" operation, the signal is processed by the amplifier circuit, but the output may be driven to ground by selecting the grounded input channel. This technique can be used for noise blanking with RF signals, or in other applications to prevent overload of circuits following the amplifier. The same idea can be used to make a fast settling return to zero deglitcher for high speed digital to analog convertors, where the grounded channel can be selected while the DAC is glitching and settling.

Pulse Generator/Pin Driver

The ability of the SWOP Amp to drive 50 ohm cables with speed and precision, combined with fast switching between signals and fast, smooth settling make the selectable input circuits shown in Figure 2 work well as pulse generators for test systems. The SWOP Amp is configured as described in Figure 2a, 2b or with both channels inverting. DC voltages are applied to the two inputs via voltage dividers, a DAC, or any other source. When the channel select is driven with a square wave between high and low logic levels, the output will switch between the two

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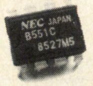
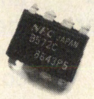
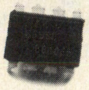
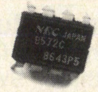
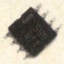
÷ 2  UPG502 UPB565 UPG504 UPB581 UPB558 UPB584	÷ 4  UPG501 UPB565 UPG503 UPB582 UPB585	÷ 8  UPG506 UPB565 UPB567	÷ 2/4/8 ÷ 8/9  UPB555 UPB587	÷ 10/11  UPB551 UPB552 UPB554
÷ 16/17  UPB553 UPB556 UPB555 UPB571	÷ 20/21  UPB572	÷ 20/22  UPB551 UPB552 UPB554	÷ 32/33  UPB555 UPG569 UPB571	÷ 40/41  UPB572
÷ 40/44  UPB551 UPB552 UPB554	÷ 64  UPB564 UPB565	÷ 64/65  UPB566 UPB569 UPB571	÷ 64/68  UPB562 UPB568	÷ 80/81  UPB572
÷ 128 ÷ 128/129  UPB564 UPB566	÷ 128/64  UPB588	÷ 128/136  UPB562 UPB568	÷ 256  UPB564	÷ 256/512  UPB586

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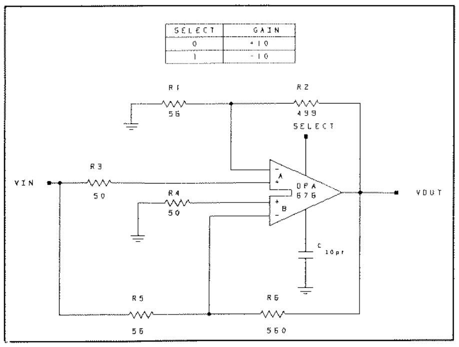

Figure 4a. Using different gains for each channel and connecting both multiplexer inputs together makes a simple high performance programmable gain amplifier.

output levels determined by the DC input voltages and the gain of each channel. With the inherent 50 ohm drive capability, this circuit could also be described as a precision programmable pin driver.

A Four Channel Multiplexer

Figure 3 shows a four channel MUX made with three SWOP Amps. The circuit for each amplifier is as shown in Figure 2a, with SWOP Amps A1 and A2 used to MUX the four channels into two, and A3 performing the final multiplexing down to one output channel. The circuit use two bits for channel select, as shown in the chart in Figure 4. The first bit selects channel A or B on amplifiers A1 and A2, and the second bit selects channel A or channel B on amplifier A3 (which selects the output of amplifier A1 or amplifier A2). Channel selection still requires less than 5 nS, while settling time to 0.1 percent at the output increases to approximately the RMS sum of the settling time for each stage. In this example, all amplifiers are configured for a gain of +10, giving a net gain of 100 for each of the four channels (settling time is 12 nS to 1 percent, 21 nS to 0.1 percent, 35 nS to 0.01 percent). Gains could be different for any of the stages, giving each channel a different gain.

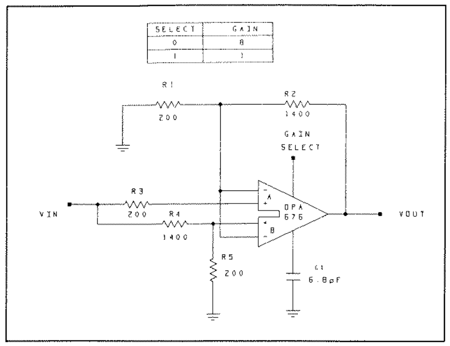

Figure 4b. This PGA circuit gives both channels optimum bandwidth when gains are not similar in magnitude.

Programmable Gain Amplifiers

A simple way to construct a programmable gain amplifier is to take the Figure 2b circuit and connect the two inputs together, as shown in Figure 4a. Now the channel select functions as a gain select pin. Both inputs are connected to the signal source, and the channel select signal now determines which input stage is active. The feedback attenuator to the active input stage sets the gain. The feedback attenuator for the unselected output is simply one component of the load resistance and is driven by the output.

This particular configuration may be used as a gain of 10 absolute value circuit, by driving the gain/channel select pin with a fast comparator that compares the input signal with ground. This circuit can select a positive gain for one input signal polarity, and a negative gain for the other input signal polarity, resulting in a unipolar output for bipolar inputs. Since the SWOP Amp changes channels when the output is at ground, the glitches around zero will be small if a fast, high resolution comparator is used because no slewing is required when channel switching occurs.

Yet another application of the Figure 4a circuit is in the field of communications. Driving the channel select pin with a carrier frequency square wave produces an amplitude modulated output. Reversing the gain polarity on every half cycle of the carrier yields a balanced

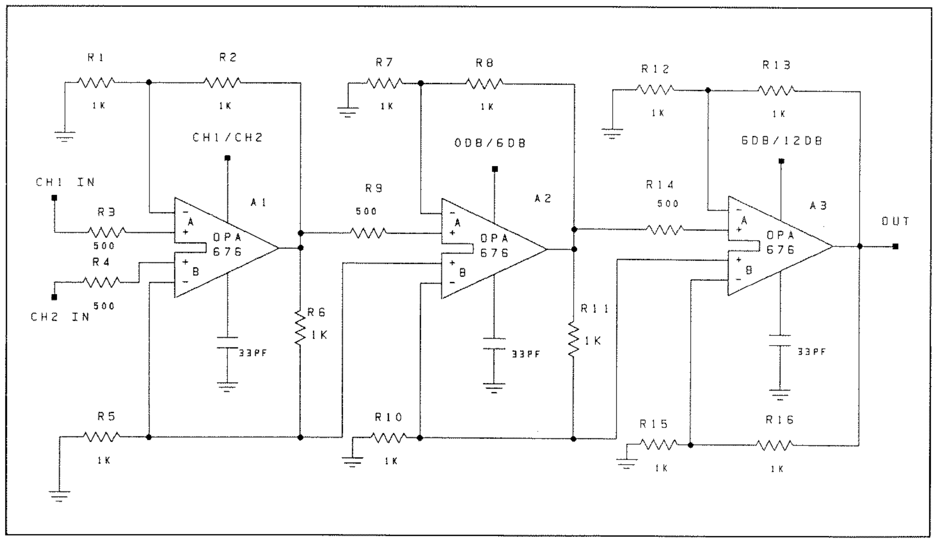

Figure 5. Two channel programmable gain amplifier with gains of 6, 12 and 18 dB.

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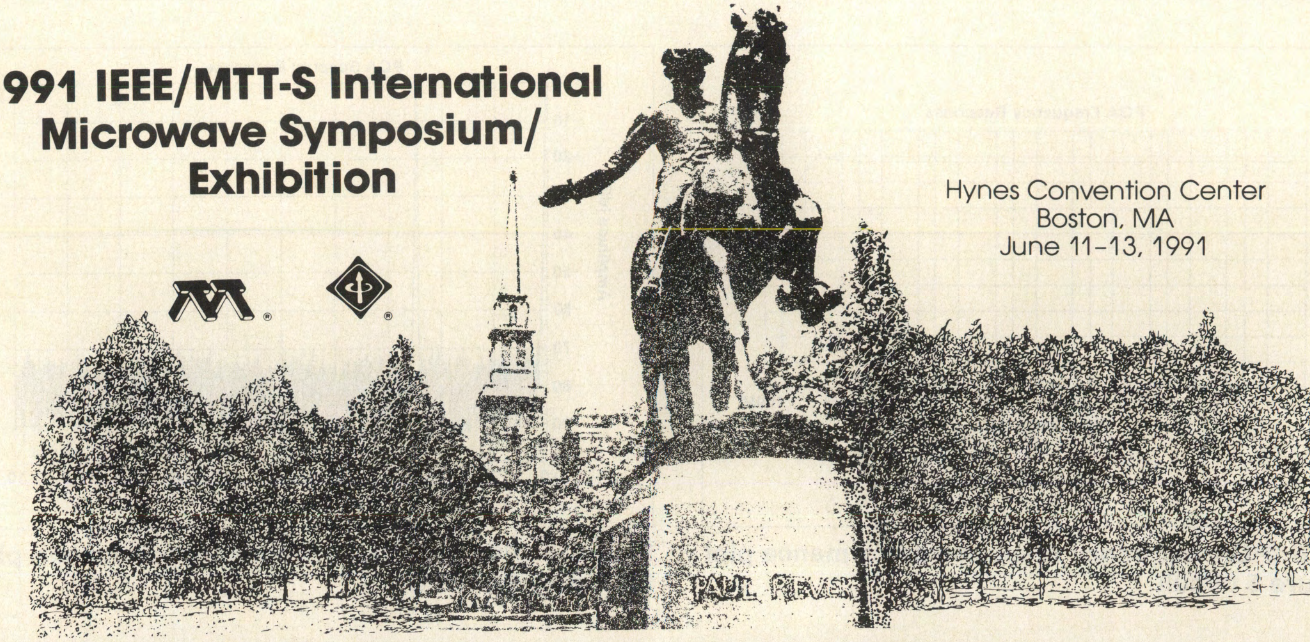

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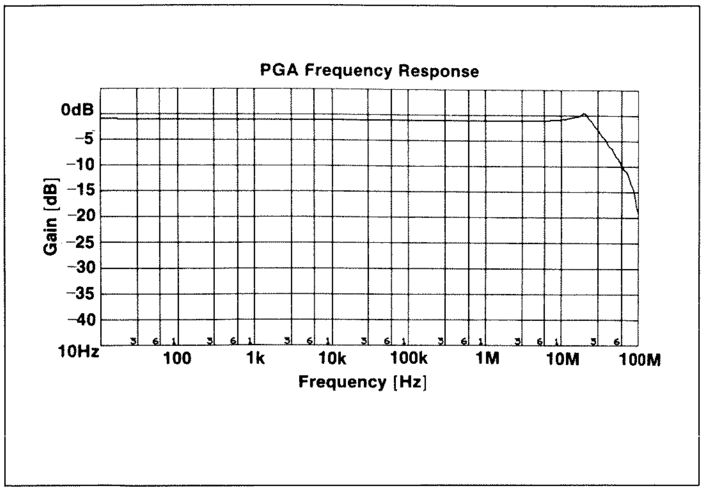

Figure 6. PGA frequency response performance plot for Figure 5 circuit.

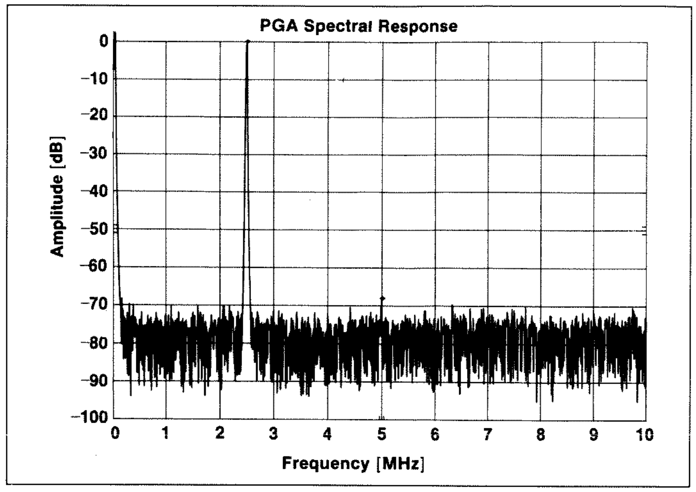

Figure 7. PGA spectral response performance plot for Figure 5 circuit.

modulator, and various filters can be used at the output to produce single sideband, double sideband, or suppressed carrier modulation.

Optimizing Bandwidth for Two Different Gains

Since the compensation capacitor is connected to the circuitry driven by both amplifiers, the programmable gain amplifier must be compensated for the lower of the two gain settings. For large differences in the closed loop gain, the bandwidth for the higher gain channel will be less optimum. This is not a problem in every application, but a second way to build a programmable gain amplifier easily eliminates this limitation.

Figure 4b shows a programmable gain amplifier with optimized bandwidth for both channels. A common feedback network is shared by both channels, like the multiplexed input amplifier from Figure 2a, so both channels have the same loop gain. The input signal drives a voltage divider, so the channel A amplifies the full input signal, while the input signal to channel B is attenuated. In Figure 4b the gain from input to output is 8 when channel A is selected, and 1 when channel B is selected. Since the feedback factor for each amplifier channel is the same, the compensation may be optimized to take advantage of the full gain bandwidth product available from the amplifier. Since the SWOP Amp is optimized for the best AC performance in higher gains, this approach is especially useful when one or more channels must operate at a gain of unity.

Putting It All Together

Figure 5 shows 3 SWOP Amps used to make a two channel multiplexer with selectable gains of 6 dB, 12 dB, or 18 dB. The channel 1 or channel 2 input signal is selected by the channel select signal to A1, which has a gain of 2. Amplifiers A2 and A3 provide selectable gains of 1 or 2 in the following manner. Each amplifier has a gain of 2, set by the two feedback attenuator resistors being equal, but channel A picks off the output of the previous stage directly, while channel B picks off the output of the previous stage at the midpoint of an equal valued resistor voltage divider. Thus the gain set resistors for A1 also provide the "selectable" attenuator for A2 to use. A3 uses the gain set resistors for channel B of A2 (R11 and R10) for its selectable input attenuator.

Even though there are 3 amplifiers cascaded, the dynamic performance is outstanding as shown by the spectral (Figure 6) and frequency (Figure 7) analysis plots. Since there are always three amplifiers in the signal path, bandwidth and distortion are virtually constant for all gains. The plot shows the measured distortion for this circuit, with the 2F harmonic less than 65 dB.

This circuit can be simplified considerably by deleting the top row of 1kohm resistors (R1/R2, R7/R8, and R12/R13) and then connecting the inverting input for channel A of each amplifier to the channel B inverting input. This lets both channels for each stage share a common feedback network, an idea that was mentioned earlier. There is a trade-off; when this simpler circuit is used, harmonic distortion increases a few dB due to the extra capacitance at each summing node. One additional benefit of the simpler circuit is that gains for the two channels are absolutely identical.

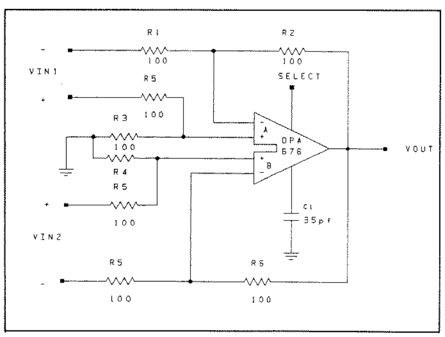

Figure 8. Multiplexer with two differential inputs and a gain of 1. Bandwidth is 100 MHz. CMRR is 80 dB at 50 MHz.

Differential Circuits

Since each channel of the SWOP Amp is a true operational amplifier with outstanding high frequency common mode rejection characteristics, any of the above circuits can be implemented as differential amplifier circuits. This allows the system designer to separate signals from noise even when both are in the RF range. Figure 8 shows a simple differential MUX with a gain of 1 for the differential input, 100 MHz bandwidth,and CMRR of 80 dB at 10 MHz.

About the Author

Tom Anderson is a Design Project Manager at Burr-Brown Corporation. He may be reached at P.O. Box 11400, Tucson, AZ 85734. Tel: (602) 746-7611.

Multi-Stage Amplifier Bandwidths

By Marlin Greer, Associate Professor, Murray State University

This article examines the technique of determining the bandwidth of a system of amplifiers when the individual stage bandwidths are known. The exact method is developed and then three special cases are defined which permit simplified calculations for most situations.

An otherwise ideal amplifier with a single lag network (either internal or external) can be represented as an ideal amplifier driving an RC network (Figure 1). $A_{v(mid)}$ represents the gain at low frequency where the response is flat, that is, not dependant on frequency.

Mathematically, the voltage gain as a function of frequency for the above model can be expressed as a complex algebra expression. The frequency of interest, f, is the frequency at which the gain is to be determined. The critical frequency, f_c, is also called the break point or 3 dB frequency. The letter "j" is the imaginary number designator used in engineering for which "i" is used in mathematics.

$$A_v(f) = \frac{A_{v(mid)}}{(1 + jf + f_c)} \qquad (1)$$

$$f_c = \frac{1}{2\pi RC} \qquad (2)$$

When $f = f_c$:

$$A_v(f) = \frac{A_{v(mid)}}{(1 + jf_c/f_c)} = \frac{A_{v(mid)}}{(1 + j1)} \qquad (3)$$

therefore:

$$|A_v(f_c)| = \frac{A_{v(mid)}}{\sqrt{2}} = A_{v(mid)}(dB) - 3dB \qquad (4)$$

That is, when $f = f_c$, there is a 3 dB loss of gain.

Since the gain is fairly flat (within 3 dB) from zero Hertz to f_c, the frequency distance from the origin to the corner is called the 3 dB bandwidth (Figure 2).

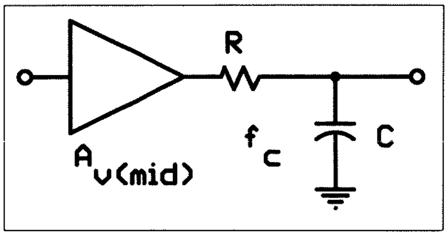

Figure 1. An ideal amplifier driving an RC network.

This bandwidth is the usable frequency range.

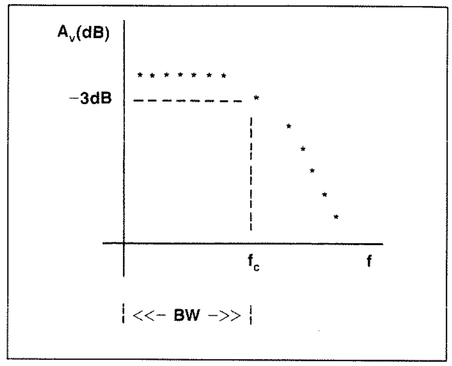

Figure 2. The usable frequency range.

Multistage Systems

This concept can be extended to multistage systems of the type shown in Figures 3a and 3b. The amplifier symbol represents an ideal, that is, infinite bandwidth, infinite input impedance, zero output impedance amplifier. Therefore it assumes no interaction between the RC stages.

The voltage gain of the system is just the product of the gain of the individual stages:

$$A_v(f) = \left(\frac{A_{v(mid1)}}{(1 + jf/f_{c1})}\right)\left(\frac{A_{v(mid2)}}{(1 + jf/f_{c2})}\right) \qquad (5)$$

$$A_v(f) = \frac{A_{v(mid)}}{(1 + jf/f_{c1})(1 + jf/f_{c2})} \qquad (6)$$

where:

$$A_{v(mid)} = A_{v(mid1)}A_{v(mid2)} \qquad (7)$$

Since no low frequency losses are assumed because of the absence of lead networks, the 3 dB bandwidth is equal to the frequency distance from zero Hertz to the upper 3 dB point. Because 3 dB corresponds in magnitude to $1/\sqrt{2}$, the composite or system corner (f_o) can be described as occurring as follows:

$$|A_v(f_o)| = \frac{A_{v(mid)}}{\sqrt{2}} = \frac{A_{v(mid)}}{|(1 + jf_o/f_{c1})(1 + jf_o/f_{c2})|} \qquad (8)$$

or in simplifying to the essentials by leaving out the common $A_{v(mid)}$

$$|1 + jf_o/f_{c1}| \ |1 + jf_o/f_{c2}| = \sqrt{2} \qquad (9)$$

$$\left(\sqrt{[1 + (f_o/f_{c1})^2]}\right)\left(\sqrt{[1 + (f_o/f_{c2})^2]}\right) = \sqrt{2} \qquad (10)$$

$$[1 + (f_o/f_{c1})^2][1 + f_o/f_{c2})^2] = 2 \qquad (11)$$

This formula can be used to solve for the composite 3 dB corner (f_o) for any f_{c1} and f_{c2}. However, special cases can be defined that simplify calculations for most situations.

Equal Corner Frequencies (Case I)

When the two corner frequencies are equal in an amplifier system, equation 11 leads to a simplification since the two bracketed terms become equal to each other.

$$f_{c1} = f_{c2} = f_c \qquad (12)$$

$$[1 + (f_o/f_c)^2]^2 = 2 \qquad (13)$$

$$1 + (f_o/f_c)^2 = \sqrt{2} \qquad (14)$$

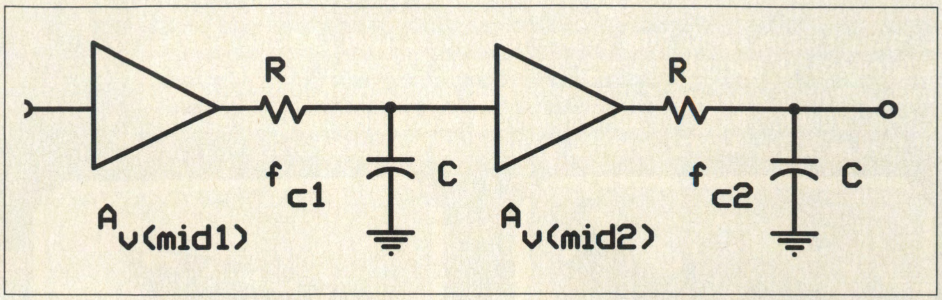

Figure 3. A multistage system.

$$(f_o/f_c)^2 = \sqrt{2} - 1 \tag{15}$$

$$f_o = f_c\sqrt{(2^{1/2} - 1)} \tag{16}$$

or the more general formula:

Case I, if $f_{c1} = f_{c2} = \ldots = f_{cn} = f_c$

$$f_o = f_c\sqrt{(2^{1/n} - 1)} \tag{17}$$

where n = the number of equal corners

Example:

Three cascaded amplifier stages each have 3 dB corners at 1.5 MHz. The composite corner (f_o) can be found as follows:

$$\begin{aligned} f_o &= f_c\sqrt{(2^{1/n} - 1)} \\ &= 1.5\text{M}\sqrt{(2^{1/3} - 1)} \end{aligned} \tag{18}$$

$f_o = 765$ kHz

Widely Separated Corner Frequencies (Case II)

When one corner is at much higher frequency, the two corners are considered widely spaced. ($f_{c2} >> f_{c1}$). Since f_{c2} is greatly removed from f_{c1} it will have very little effect on the gain back at f_{c1}, allowing f_{c1} to be considered as the approximate system corner frequency.

It will be shown later that corners can be considered widely separated if they are at least a factor of 3 apart. At this separation there is about 10 percent error, with the error diminishing as the separation increases.

Case II, if $f_{c2} \geqslant 3f_{c1}$

$f_o \approx f_{c1}$

Example:

$f_{c2} = 300$ MHz, $f_{c1} = 60$ MHz

Since the factor of three separation requirement is met, $f_o = f_{c1} = 60$ MHz.

Closely Spaced Corners (Case III)

When corners are in the vicinity of each other, (the second frequency is greater than a multiple of 2 but less than a multiple of 3), then the original, universal equation (equation 11) must be used to solve for the 3 dB point of the system.

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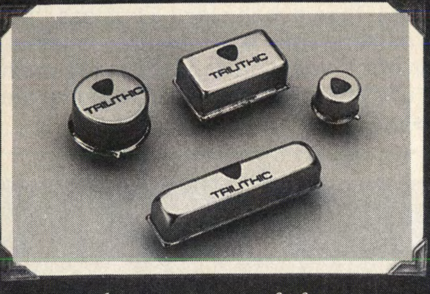

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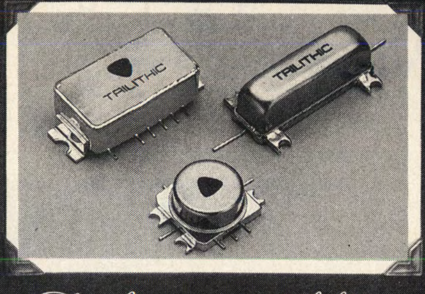

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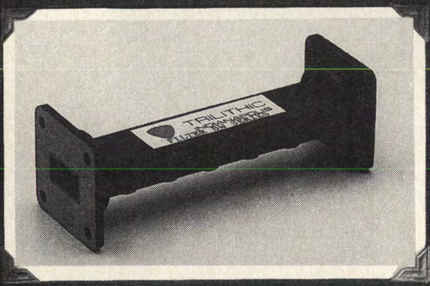

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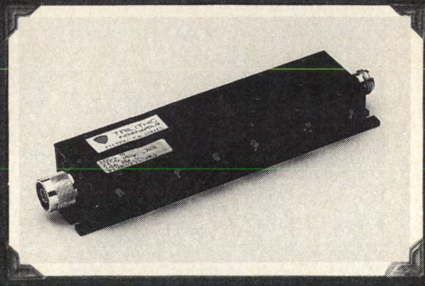

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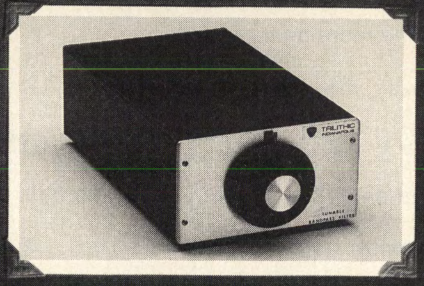

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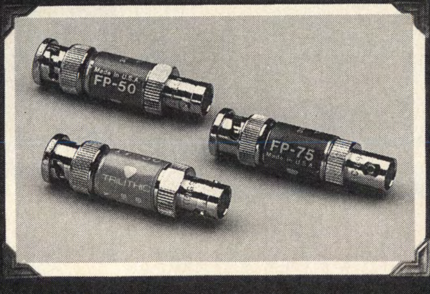

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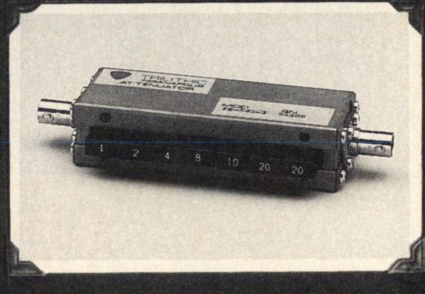

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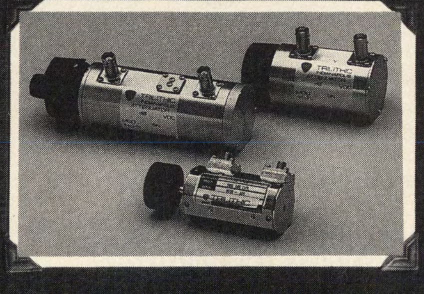

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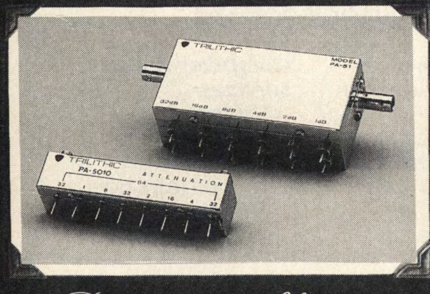

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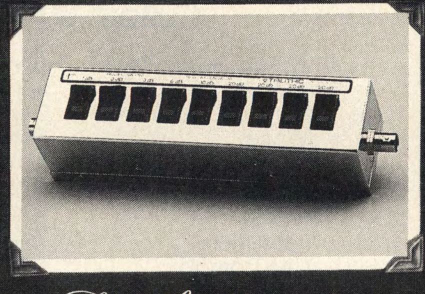

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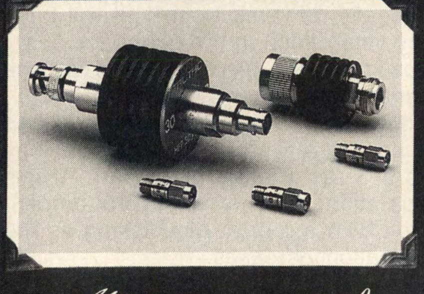

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CASE	DEFINITION	FORMULAS
I	$f_{c2} = f_{c1}$	$f_o = f_c\sqrt{(2^{1/n}-1)}$ n = # equal corners $f_c = f_{c1} = f_{c2} = ...f_{cn}$
II	$f_{c2} > 3f_{c1}$	$f_o \approx f_{c1}$
III	$2f_{c1} < f_{c2} < 3f_{c1}$ (or any situation for more accurate answers)	$a = 1/(f_{c1}^2 f_{c2}^2)$ $b = 1/f_{c1}^2 + 1/f_{c2}^2$ $c = -1$ $x = [-b + b2 - 4ac]\ /2a$ $f_o = x$
IV	$f_{c2} \leqslant 2\ f_{c1}$	$f_{ave} = \sqrt{(f_{c1}\ f_{c2})}$ $f_o \approx f_{ave}\sqrt{(2^{1/2}-1)}$

Table 1. Formulas and categories for calculating system corner frequencies.

$$[1+(f_o/f_{c1})^2]\ [1+(f_o/f_{c2})^2] = 2 \quad (19)$$

When multiplied out, this results in an equation of quadratic form.

$$[1/(f_{c1}^2\ f_{c2}^2)]f_o^4 + [1/f_{c1}^2 + 1/f_{c2}^2]f_o^2 - 1 = 0 \quad (20)$$

In order to simplify the appearance of the equation and to temporarily reduce its order, the following definitions are made:

$$x = f_o^2 \quad (21)$$

$$a = 1/f_{c1}^2 f_{c2}^2 \quad (22)$$

$$b = 1/f_{c1}^2 + 1/f_{c2}^2 \quad (23)$$

$$c = -1$$

These definitions make comparison with the standard form quadratic formula easy since the equation can now be rewritten as follows:

$$ax^2 + bx + c = 0 \quad (24)$$

Solve for the positive root of x using the quadratic equation to obtain a physically realizable frequency. The positive root of this x value is the 3 dB frequency, f_o.

The exact formula for Case III is:

$$x = [-b + \sqrt{(b^2 - 4ac)}]/2a \quad (25)$$

$$f_o = \sqrt{x} \quad (26)$$

The following is an examination of the formula at the multiple-of-3 point, mentioned above:

Let $f_{c1} = f_c$, $f_{c2} = 3f_c$, where f_{c1} equals the lower corner. The exact system corner (f_o) can be found by substituting into equation 20:

$$(1/9f_c^4)f_o^4 + (1/f_c^2 + 1/9f_c^2)f_o^2 - 1 = 0 \quad (27)$$

$$f_o^4 + (9f_c^2 + f_c^2)f_o^2 - 9f_c^4 = 0 \quad (28)$$

$$f_o^4 + 10f_c^2 f_o^2 - 9f_c^4 = 0 \quad (29)$$

$$x^2 + (10f_c^2)x - 9f_c^4 = 0 \quad (30)$$

$$x = 0.8310f_c^2 \quad (31)$$

(using the quadratic equation)

$$f_o = \sqrt{x} = 0.912f_c \quad (32)$$

$$f_o = 0.912 f_{c1} \quad (33)$$

Note that at a frequency separation of a multiple of three, taking f_{c1} as the composite corner, as in the widely separated case, would correspond to a nine percent error. This multiple of three point is used as the beginning of being widely separated (Case II).

Definition: Case III exists if f_{c2} is less than $3f_{c1}$, that is, if less than a multiple of three separation exists. Otherwise, approximate the system corner by the widely spaced assumption.

Very Closely Spaced Corners (Case IV)

If corners are within a factor of two of each other, the tedious Case III approach can still be avoided by taking the average of the two corners and assuming both corners to exist at that average. Since the lower corner contributes more to the location of the system corner than the higher one it would be better to use a definition of average that skews the result towards the lower value. Geometric averaging has that effect. The individual corners, taken as each being at the average value, is then modified by the multiplier, from the Case I formula. This method is especially useful for situations that have theoretically identical corners, but the actual laboratory values are somewhat separated due to slight differences in components or in amplifiers.

Note that the skewing created by geometric average would be incorrect for a lead network caused by a coupling or bypass capacitor.

Example:

$$f_{c1} = 100 \text{ MHz}, f_{c2} = 200 \text{ MHz}$$

$$f_{ave} = \sqrt{(100M \times 200M)} = 141.4\text{MHz} \quad (34)$$

$$f_o = f_{ave}\sqrt{(2^{1/2}-1)} = 141.4M(.6436) \quad (35)$$

$$f_o = 91.02 \text{ MHz}$$

The actual 3 dB frequency is 83.76 MHz which is an error of about 9 percent.

Case IV, if $f_{c2} \leqslant 2f_{c1}$

$$f_{ave} = \sqrt{(f_{c1}\, f_{c2})} \quad (36)$$

$$f_o = f_{ave}\sqrt{(2^{1/n}-1)} \quad (37)$$

Summary

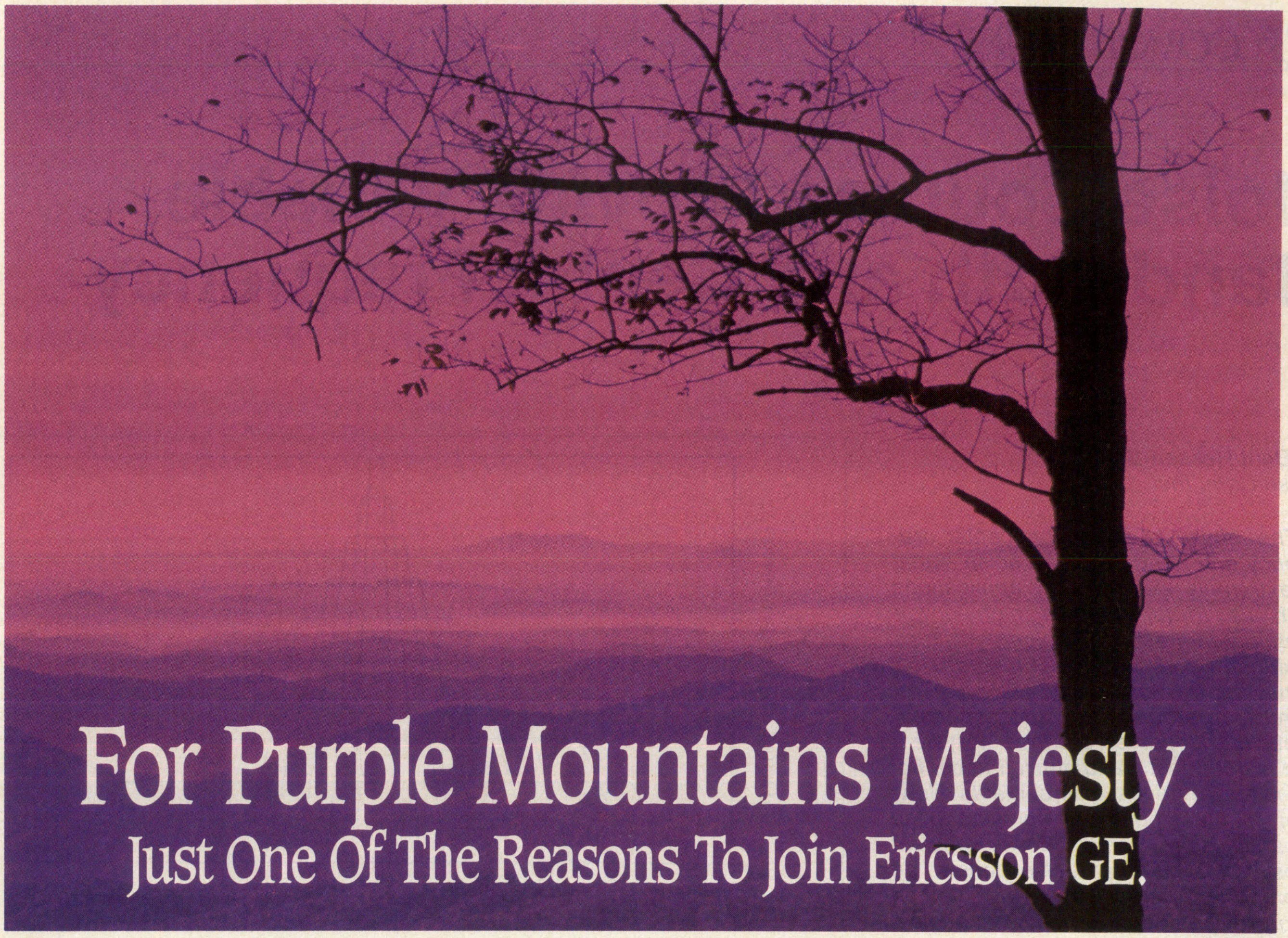
The various categories and the formulas for calculating their system corner frequencies are summarized in Table 1.

Case II & IV are accurate to better than 10 percent; for 3 percent accuracy, use factors of 1.5 and 6 instead of 2 and 3 used above.

Although it is difficult to calculate the exact corner frequency when multiple phase-shifting networks are involved, most situations can be resolved to relatively easy to determine approximations. *RF*

About the Author

Marlin Greer is an associate professor in the Department of Electrical Engineering at Murray State University. He may be reached at College of Industry and Technology, Department of Engineering Technology, Murray, KY 42071-3307. Tel: (502) 762-3393.



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Noise Sources Provide Variable Bandwidth and Center Frequency

By Bent Hessen-Schmidt
Noise Com, Inc.

The ability to vary the center frequency and bandwidth of a noise signal is extremely desirable in applications such as frequency-agile high power jamming and testing of communications equipment. The traditional methods for achieving this capability generally require a large number of custom components that together produce a system that is both costly and difficult to configure in space-critical environments. However, it is now possible to provide these capabilities in a small, power efficient subsystem that is far less expensive an "component intensive." This method also provides other benefits not achievable before.

An efficient way to provide a noise signal that can be varied in both center frequency and bandwidth has eluded designers for many years. The technique relies on amplified noise, which is passed through a bank of bandpass filters that are switched to provide the desired noise bandwidths. While this solution accomplishes the task, it does so with a penalty in size, cost and efficiency. And, even when implemented effectively, the sheer number of filters needed by such a system all but precludes many bandwidths from being employed since each bandwidth requires its own bandpass filter.

A better solution is achieved by using a noise source with an extremely wide bandwidth along with a wideband frequency modulator. The roll-off characteristics of the signal are equivalent to those of a 12-pole Chebyshev filter, but no filters are required. The resulting bandwidth, variable frequency noise source is extremely space efficient, requiring only a lightweight rack-mount system measuring 12 × 1.5 × 19 in. In subsystem form it is even smaller. The same capabilities produced by conventional means would require at least 2 × 3 × 4 ft. just for the filters. The additional

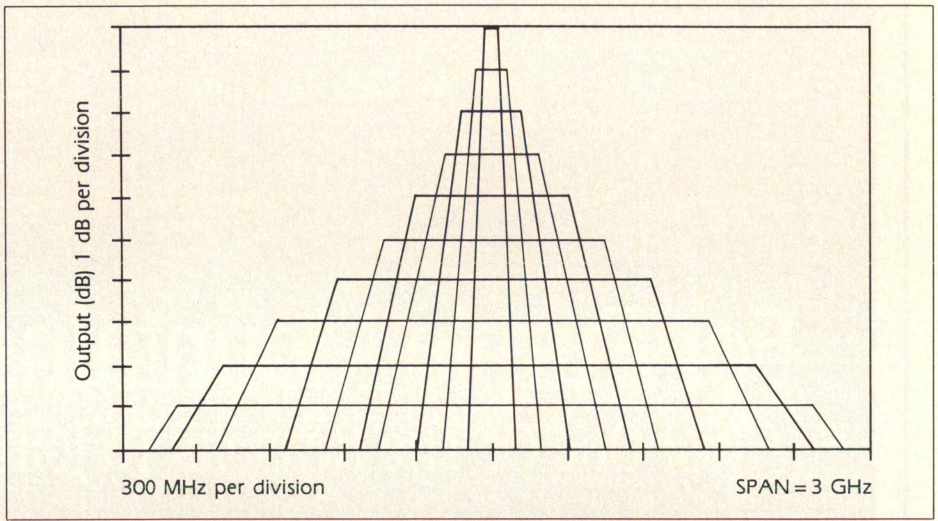

Figure 1. the power spectral density of the NCS-4100 increases with a given output power by 1 dB for each 20 percent reduction in bandwidth.

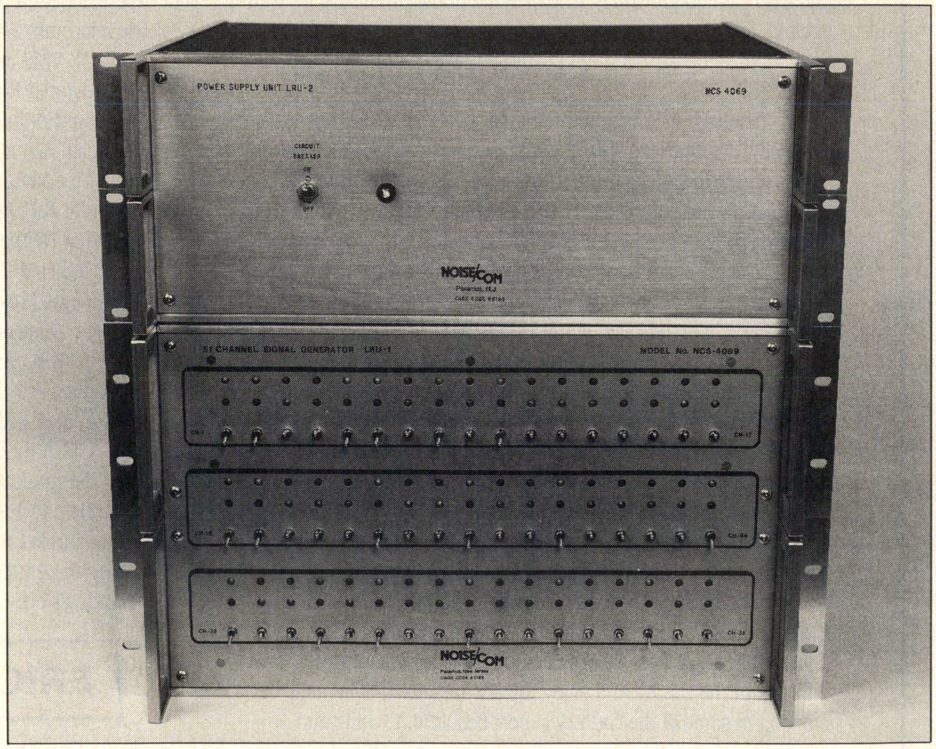

Figure 2. The complete NCS-4100 based multi frequency jammer is smaller than just the filters of a conventional system.

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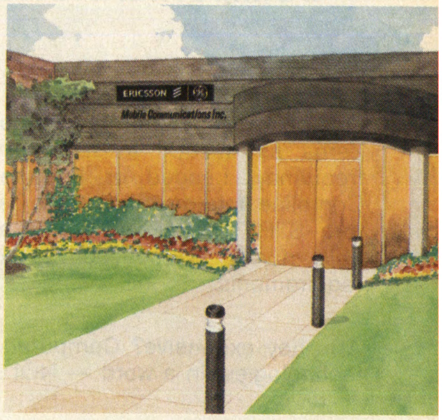

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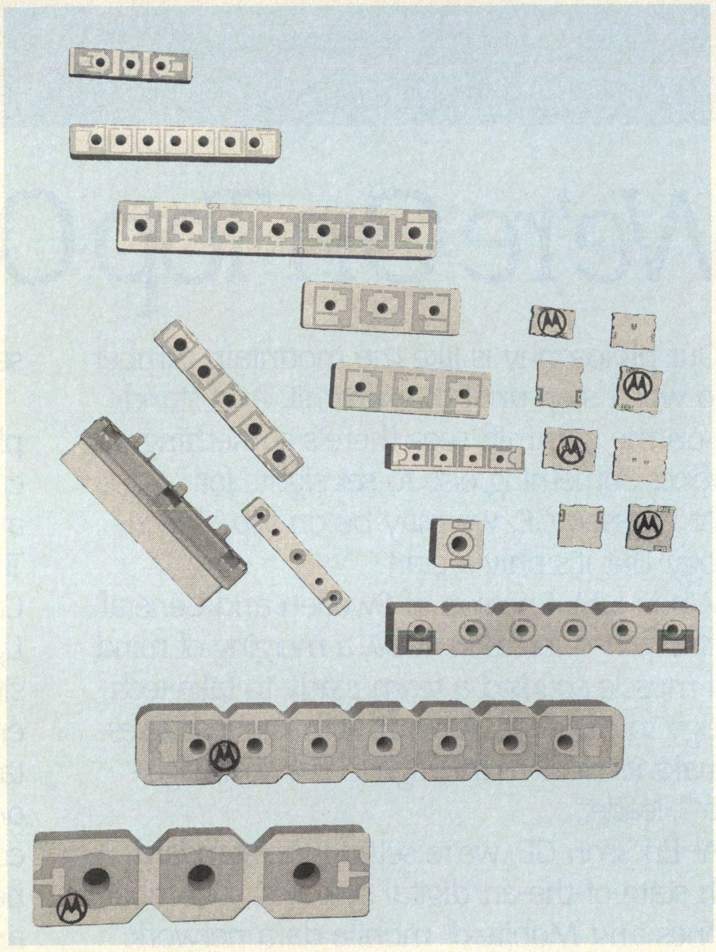

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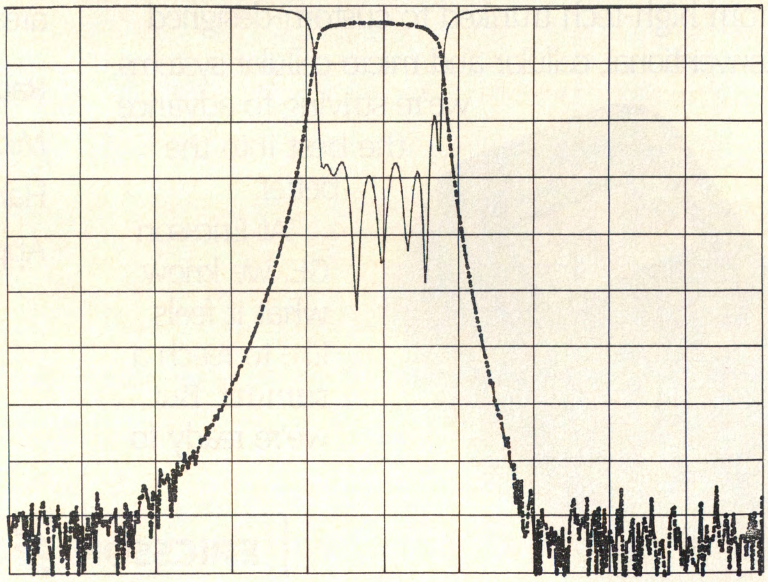

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Components Division

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hardware — cabling, switches and associated electronics — would make the subsystem still larger.

Applications

Applications for the Noise Com NCS-4100 are broad. It can be used as the fundamental component in high power noise jammers, for testing of communications channels, and in intermodulation and spurious testing of frequency converters. There are obviously many applications in which the ability to vary both frequency and bandwidth of a noise signal are desirable.

Uses in High Power Jammers

The impetus for the development of the NCS-4100 subsystem came from military jamming applications, in which noise has long been used as a jamming tool. Noise is generally regarded as the easy way to render an enemy communications or radar system ineffective. It is especially useful when the precise characteristics of the offending signal are not known, and when the transmitter employs techniques such as frequency-hopping, spread-spectrum modulation, or multiphase modulation.

Because of its random nature, noise can blanket a wide range of frequencies simultaneously. Since the noise signal is nearly identical to that of thermal noise, the enemy receiver will not detect any identifiable signals to indicate it is being jammed. It therefore will not employ countermeasures to thwart the offending noise jammer.

The effectiveness of noise jamming depends on the ability to overwhelm the enemy receiver. However, since the power of a noise signal is distributed over a very wide bandwidth, a noise jammer requires extremely powerful amplifiers to radiate enough power at any given frequency to be effective.

Since radiated power density is inversely proportional to bandwidth, any reduction in bandwidth of the transmitted noise signal will produce an increase in effective radiated power density. The NCS-4100 is designed to exploit this relationship, since it can tune to a particular frequency and simultaneously reduce the bandwidth of the noise at that frequency, concentrating radiated power within a smaller range of frequencies. Specifically, the NCS-4100's power spectral density increases with a given output power by 1 dB for each 20 percent reduction in bandwidth (Figure 1).

Bandwidth can be up to 20 percent of the center frequency. Combined with a switching speed of less than 200 ns, the ability to selectively modify bandwidth while changing frequency makes the NCS-4100 an extremely effective jammer subsystem that can address fixed and agile threats of many types.

The NCS-4100 has been designed into a multi-frequency jammer (Figure 2) that utilizes individual units feeding a low-loss combiner which in turn feeds step attenuators.

Test Applications

The NCS-4100 is also well suited to testing of communications receivers as well as amplifiers, mixers and multipliers. In two-tone intermodulation testing, the NCS-4100 can be used as one of the test tone generators. As this unit is

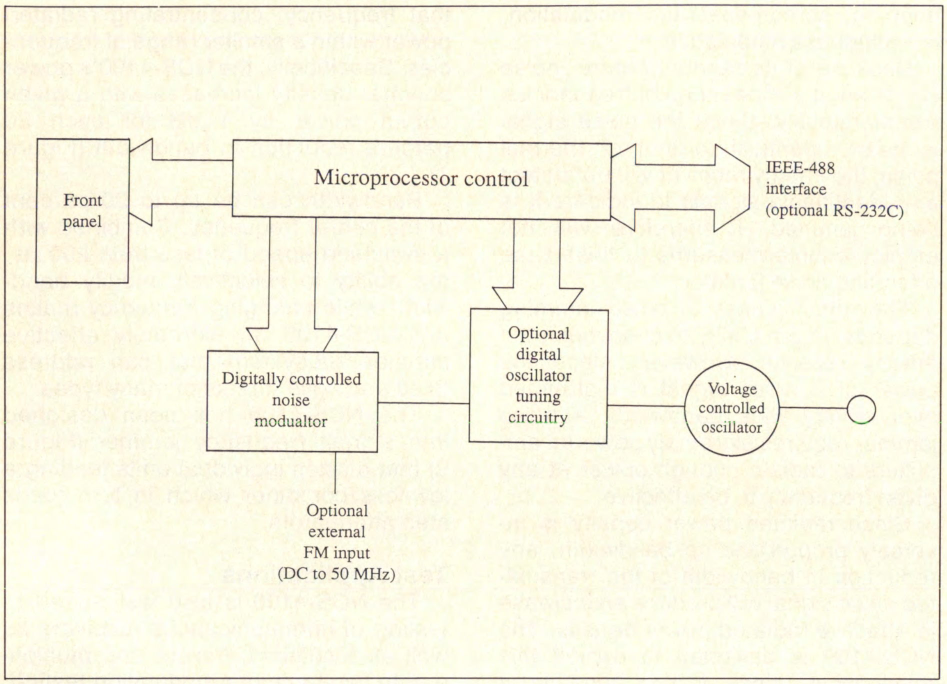

Figure 3. Block diagram of the basic NCS-4100 variable bandwidth, variable frequency noise subsystem.

capable of randomly generating frequencies within a programmable band, testing time can be reduced and intermodulation products can be measured directly by a spectrum analyzer.

EW Simulators

The NCS-4100 can also be used in electronic warfare simulators, in which its ability to accept a wideband external FM input signal allows many modulation characteristics to be produced. An almost infinite number of threat scenarios can be produced to simulate a known signal type or a hypothetical threat.

Technical Description

The NCS-4100 noise subsystem contains a digitally controlled noise modulator with a large memory for different functions (Figure 3). The modulator is temperature compensated and vibration stabilized to ensure good performance in rugged environments. A microprocessor orchestrates all system RF functions as well as control of front panel functions in the rack mount unit. External control is provided via a TTL digital line. Output is at least +10 dBm with flatness of ±0.5 dB over bandwidths up to 10 percent of the center frequency and ±1.0 dB over 20 percent of center frequency.

The NCS-4100 was designed to be configured as both a subsystem for MIL-E-5400 grade airborne applications (Figure 4), or as a programmable bench instrument, In either configuration, control via IEEE-488 bus can be specified, as can precision attenuators and switched outputs. The unit can be custom configured to meet the needs of many specific applications, as well. The standard unit provides variable bandwidth operation over any customer-specified range of center frequencies from 500 MHz to 18 GHz.

Summary

The NCS-4100 allows both frequency and bandwidth of a noise signal to be varied in an efficient manner without the use of filters. The resulting subsystem is much smaller that can be achieved using other methods. It is designed for rugged environments such as aircraft, ships and ground-based military equip-

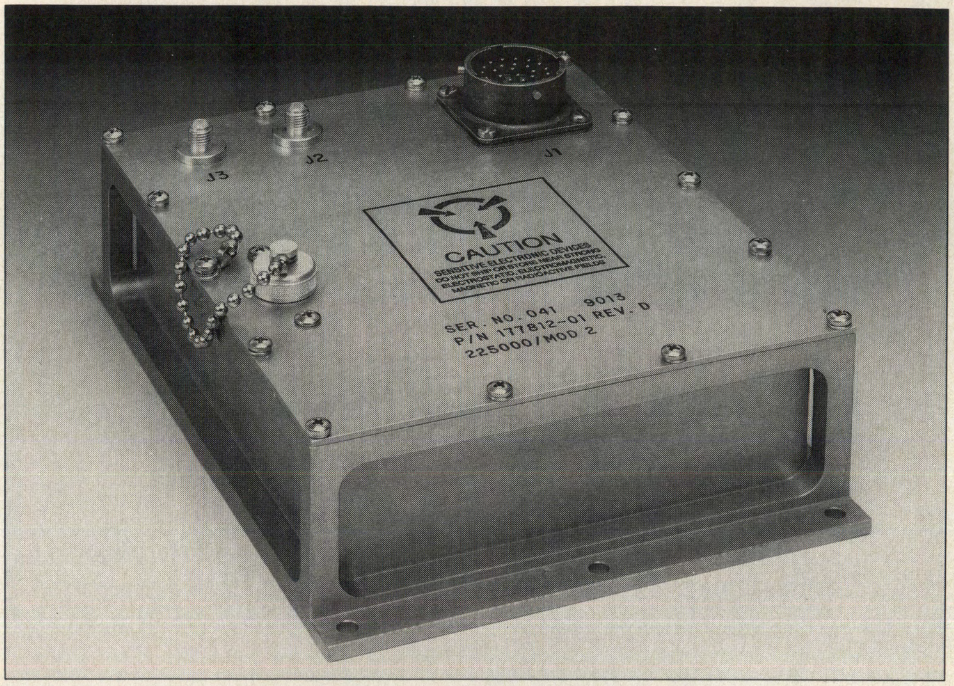

Figure 4. The NCS-4100 can be supplied in subsystem form like this one for MIL-E-5400 airborne applications.

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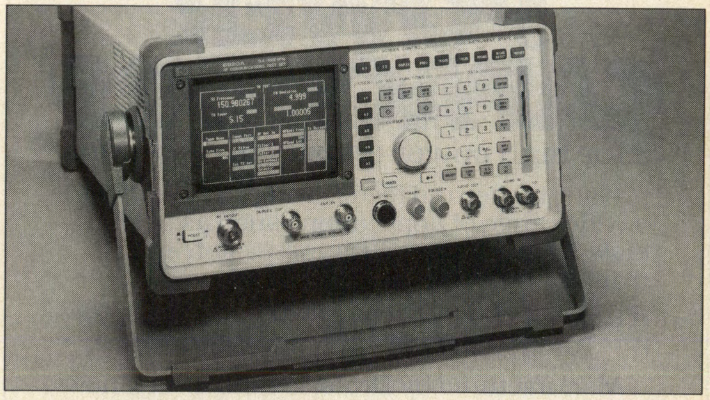
ment, and is well suited to jamming applications in which the unique characteristics of noise combined with the frequency control of a digitally tuned VCO produce an effective frequency-agile noise jamming subsystem. It can also be used to simplify the testing of amplifiers, mixers, multipliers and any other component that produces IM products. Additional uses can be found in testing of communications equipment and in EW simulators.

For more information on the NCS-4100, contact Noise Com at the address below, or circle Info/Card #200.

About the Author
Bent Hessen-Schmidt is Director of Engineering at Noise Com, Inc., East 64 Midland Avenue, Paramus, NJ 07652, telephone (201) 261-8797.

AM/FM/SSB Communications Test Set

Hewlett Packard Company has entered the service market with the new HP 8920A test set for radios, pagers, repeaters, cellular phones, filters, duplexers and combiners. Test functions include signal generator, modulation analyzer, power meter, frequency counter, audio analyzer, spectrum analyzer, tracking generator, full duplex generator, signaling encoder and decoder, digital oscilloscope, function generator, two audio sources, and a AC/DC digital voltmeter. A built-in computer and supplied software can be used to automate routine measurements. For field service, the unit combines a battery pack, MIL-STD ruggedization and environmental specifications, and 35 lb. weight. Maintenance of the unit is enhanced with memory card diagnostic routines and 24-hour delivery of replacement modules (in the U.S.). The HP 8920A is priced at $13,500. Option 002 for the spectrum analyzer and tracking generator is $1,500; Option 004 for signaling is $600.
Hewlett-Packard Company
INFO/CARD #230



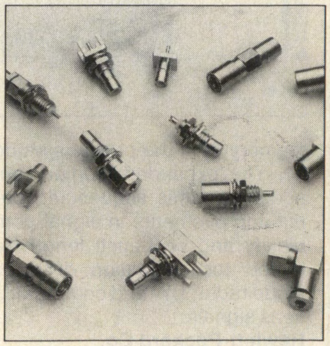
"Ovenless" OCXO

A directly-heated crystal is the heart of the smallest OCXO (1.46 cu. in.), developed by Raltron for VSAT terminals, navigation systems, instrumentation, synthesizers, cellular systems and other precision applications. 2×10^{-7} accuracy is available for −20 C to +70 C ambient with just two minutes warm-up time. The

steady state power requirement is 3 watts. The OCXO dimensions are 35.3 mm × 27 mm × 25.4 mm. The oscillator uses an AT-cut crystal which is wrapped with resistance wire for no-hysteresis reaction to temperature changes. Available frequencies are 1-20 MHz, with a TTL compatible square waveform standard. Worst-case phase noise (at 10 kHz) is specified at −140 dBc. Pricing is $65 each in 10,000 quantities.
Raltron Electronics Corp.
INFO/CARD #229

Microminiature Coaxial Connectors

Applied Engineering Products' Series 7000 are being used for military and civilian avionics and airborne communications systems. These connectors feature VSWR less than 1.2:1 for straight connectors to 12.4 GHz, and less than 1.4:1 to 18 GHz. Although design is similar to SMB connectors, the Series 7000 is two-thirds their size. Testing includes selecting random connectors and subjecting them to 1000 connect-disconnect cycles. Contacts are beryllium copper, bodies are machined brass with PTFE insulators, and all metal parts are gold plated. Additional applications include medical electronics, microwave test and communications systems. AEP Series 7000 are



used in military systems including: GRC-206, GRC-XXX, MSC-64 and SINCGARS.
Applied Engineering Products
INFO/CARD #228

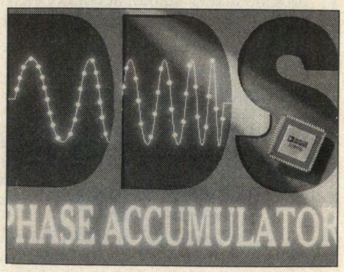
7-Bit GaAs Attenuator

A thin-film hybrid attenuator for DC-1000 MHz using GaAs MMICs has been introduced by Daico Industries. The DA0897 is packaged in a 38-pin hermetic DIP, and provides 0 to 63.5 dB attenuation in 0.5 dB steps. Typical

insertion loss is 6.0 dB from DC-200 MHz and 7.5 dB from 200-1000 MHz. Switching speed is 100 nsec, comparable to Schottky devices. A unique driver and MMIC attenuator design result in a 7-bit attenuator that will operate from DC to 1 GHz. While it operates down to DC it still provides switching speed of 100 ns. The DA0897 has a current drain of less than 15 mA from a +5, −15 VDC supply. Operating temperature covers the entire military range of −55 to +125 C. Availability is from stock. Military screening is available. The unit is pin-compatible with Daico PIN diode 7-Bit attenuators.
Daico Industries, Inc.
INFO/CARD #227

Phase Accumulator for DDS

Designed for direct digital synthesis of frequencies from 30 to 130 MHz, Analog Devices' AD9950 phase accumulator offers low cost and low power. Direct digital synthesis is used in radio communication, radar, frequency-hopping systems, and instrumentation. When combined with an external lookup table and a 10-bit DAC, a DDS system based on this 32-bit accumulator yields frequency resolution to 0.7 Hz (300 MHz clock). Spurious-free dynamic range is 76 dB below full scale, and a complete DDS system containing the accumulator, memory and DAC consumes only 6.9 watts. Built-in quadrature logic reduce the external memory requirement. Control lines are TTL compatible, while output phase data lines are ECL compatible. The AD9950 is available in commercial (0 + 70°C) and military versions. Pricing begins at $99.95 for commercial grade versions, in 100s.
Analog Devices
INFO/CARD #226



Surface-Mount Chip Attenuator

A thin-film attenuator featuring symmetrical-T configuration with wrap-around ground has been introduced by Mini-Systems, Inc. Made with tantalum nitride on alumina, with gold metallization, the units offer 1 to 20 dB attenuation (within 0.2 dB) over 0-20 GHz in a 0.05 × 0.05 in. chip size.

Mini-Systems, Inc.
INFO/CARD #225

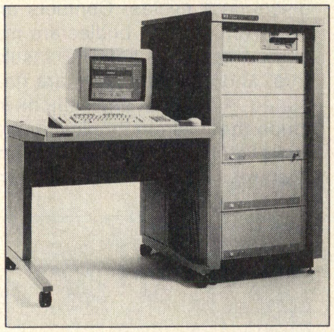
Voltage Feedback Amplifier

For applications where voltage feedback is preferred, Comlinear has introduced the CLC240, a 270 MHz amplifier with unity gain stability and settling to 0.01 percent in 18 ns. Low noise current and 80 dB common mode rejection ratio are featured, along with 40 MHz full-power bandwidth and 1000 V/us slew rate. Applications include active filters, differential amplifiers and transimpedance amplifiers.

Comlinear Corp.
INFO/CARD #224

Signal Simulator System

A 3 GHz frequency-agile signal simulator has been introduced by Hewlett-Packard. The Model HP 8791 Model 11 features enhanced waveform synthesis capability with four times larger



memory than the previous Model 10. The expanded memory and added dynamic sequencing permit improvements in signal complexity and sequence length required for simulation of multimode radar, EW and communications signals.

Hewlett-Packard Co.
INFO/CARD #223

225-400 MHz Amplifier

Pacific Amplifier announces the PS-061, featuring 45 dB gain, 4.0 dB noise figure, and 10 watts output power for the 225-400 MHz communications band. The amplifier is contained in a 3 × 2 × 0.5 inch housing, powered by +12 and +24 VDC. Overall efficiency is 33 percent.

Pacific Amplifier Corp.
INFO/CARD #222

Decoupling Capacitors

Micro/Q^{R} 1000 decoupling capacitors which fit under 32 pin SRAM chips are available from Rogers Corp. Both initial designs and retrofit can benefit from the noise reduction offered by these capacitors. Available values range from 0.002 to 0.1 uF, in ceramic dielectrics of Z5V, X7R or P3J. Their configuration results in a low inductance.

Rogers Corporation
INFO/CARD #221

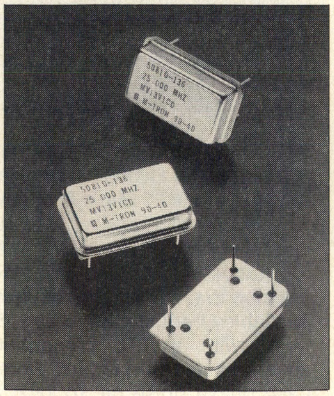
51-Channel Filter Bank

A TACAN band switched filter bank has been developed by K&L Microwave, covering 969 to 1206 MHz with 5 MHz channels. Typical switching speed is 250 ns, insertion loss is −10 dB maximum, channel-to-channel variation is ± 0.5 dB, isolation is −60 dBc, and VSWR is 1.5:1 maximum over 2.5 MHz of the channel bandwidth. Size of the unit is 16 × 8 × 1.36 inches.

K&L Microwave
INFO/CARD #220

HCMOS VCXO

M-tron Industries announces the MV series general purpose VCXO, available with center frequencies from 3 to 25 MHz. Two



pullability ranges are offered: ±50 to 120 ppm and ± 120 to 200 ppm, with 0.5 to 4.5 V control voltage range. Stability at fixed

control voltage is 100 ppm over 0 to 70 C, with higher stabilities available. The MV series is HCMOS and TTL compatible.
M-tron Industries
INFO/CARD #219

PLL+DDS Synthesizer

Sciteq introduces the VDS-1650 synthesizer, combining fine steps and fast switching with spectral purity. The PLL section uses pretuning and other speed enhancing circuitry, while the DDS element contributes fine frequency steps. The frequency range is approx. 30 percent bandwidth below 450 MHz maximum frequency. Non-harmonic spurs are 65 dB below carrier. Pricing is $4000, typical, for 10 units.
Sciteq Electronics, Inc.
INFO/CARD #218

Flat Thick-Film Resistors

A new 1/16 watt flat chip resistor is available from Hokuriku. The CR series is an ultraminiature resistor for high density, high reliability circuitry, using automated assembly. CR series resistors measure 1.6 mm × 0.8 mm × 0.35 mm, and are constructed of oxidized ruthenium film on an alumina substrate.
Hokuriku USA, Ltd.
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Ferrite Materials

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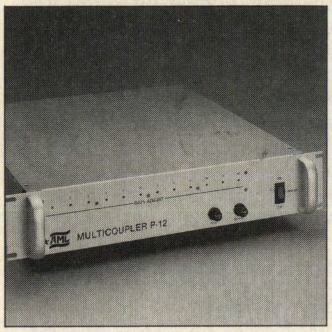
Ceramic Magnetics, Inc.
INFO/CARD #216

Low Cost Synthesizer

RF Prototype Systems offers a low cost synthesizer covering 800-1000 MHz in 1 kHz steps. Output power is +3 dBm, +18 dBm optional, with spurs better than −60 dBc. Switching speed is less than one second, and SSB phase noise at 10 kHz offset is −97 dBc/Hz. Optional frequency coverage in bands from 25 MHz to over 1 GHz is available.
RF Prototype Systems
INFO/CARD #215

Receiving Multicouplers

AML, Inc. introduces a new family of receiving multicouplers for operation at 215-320 MHz (P Band) and 2.2-2.3 GHz (S Band). One input and twelve individually gain adjustable outputs are provided on the rear panel. Applications include monopulse tracking and RF distribution systems. The units are packaged in standard 19 inch rack mounts.



AML, Inc.
INFO/CARD #214

Surface Mount Crystal

Tokyo Denpa has developed the TG-3 surface mount crystal in a ceramic package, 12 × 5 × 2.5 mm. The package allows SMT assembly with IR/VPS reflow soldering. Available frequencies are 3.58 - 24 MHz for automotive, communications, and digital applications. Stability over −40 to +125 C operating temperature is ± 50 ppm. Depending on frequency, prices are $1.00 to $1.50 per unit, 1000 units per reel.
TEW North America
INFO/CARD #213

100 Hz-5.2 GHz Spectrum Analyzer

Model FSB Spectrum Analyzer from Rohde & Schwarz offers high resolution, high stability measurements over a wide range of 170 dB (105 dB on-screen). Resolution bandwidths of 6 Hz to 3 MHz are available, with frequency spans from 10 Hz to 2 GHz. Phase noise is less than 100 dBc/Hz at 1 kHz offset, with performance further enhanced by a wide-range mixer and low-noise GaAs FET preamplifier. Single-function keys, menu-driven soft keys, and automatic test routines keep operation simple.
Rohde & Schwarz, Inc.
INFO/CARD #212

L-Band Amplifier

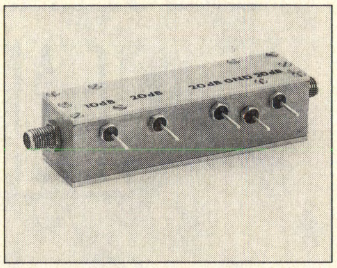
Phoenix Microwave Labs' Model SA 4001 is a solid state power amplifier delivering 40 watts CW power over the 1660-1665 MHz band. Input power is +20 dBm, harmonics are −25 dBc and spurious products are −80 dBc. The DC power requirement is +28 V at 5.0 A, and the unit is designed to operate in a shipboard environment over the −40 to +85 C temperature range. Options include forward and reverse power monitors, variable power control, choice of connectors, and higher power output.
Phoenix Microwave Labs, Inc.
INFO/CARD #211

EMI Shielding Tape

Designed to replace conventional electroplating, CHO-MASK[R] II is a conductive foil shielding tape which can be applied to enclosure flanges and other areas requiring a conductive surface. The tape is solvent resistant, and withstands cleaning, priming, painting and curing operations. The protective nylon film can then be removed to reveal a clean metal surface.
Chomerics, Inc.
INFO/CARD #210

Programmable Attenuator

JFW Industries announces a new attenuator with extended frequency range. The 50P-766 SMA covers DC-5 GHz with 0.5 watt average/100 watt peak



power handling. this unit is 0-70 dB in 10 dB steps, with 10 ms switching speed. VSWR up to 2 GHz is 1.3:1, 1.5:1 to 5 GHz. RF connectors are type SMA female, and switching power terminals are 0.040 diameter solder pins.
JFW Industries, Inc.
INFO/CARD #209

Power Transistor Bias Devices

RF Products manufactures a series of bias devices to replace the Acrian/CTC Byistor and ZO devices. These devices provide low impedance voltage/current capability compatible with the resting collector current of the amplifier, and compensate for the temperature variation in the Vbe of the transistor. The BYI series are exact replacements for the former Acrian/CTC products, and the ZO series are designed to work with high power bipolar transistors operating in Class A or AB.
RF Products, Inc.
INFO/CARD #208

Silicon Bipolar MMIC Amplifier

Bipolarics introduces the BA-08 cascadable MMIC darlington amplifier for use in 50 ohm systems. The unit features 30 dB gain at 100 MHz; 20 dB at 1 GHz, with 10 dBm output power at 1 dB compression. The device is usable over the 100 kHz to 5.0 GHz range. Operating voltage is 7.5-8.5 volts at 35 mA typical current draw. The device is available in 85 mil, 145 plastic Micro-X, 85 mil hermetic Micro-X, and 70 mil stripline packages, as well as in chip form.
Bipolarics
INFO/CARD #207

50 Watt Cellular Base Amplifier

Model STA-1000-50L from Semetex is a 50 watt, 50 dB gain linear amplifier specifically designed for multichannel cellular base station applications in the 860-910 MHz band. Intermodulation performance is rated at 35 dB below peak power with 10 or more equal amplitude carriers applied, and harmonic and spurious products are more than −60 dBc The output stage combines four silicon FETs capable of over 100 watts. The unit is specified at 24 VDC, with enhanced performance if a 28 VDC power supply is used.
Semetex Corp.
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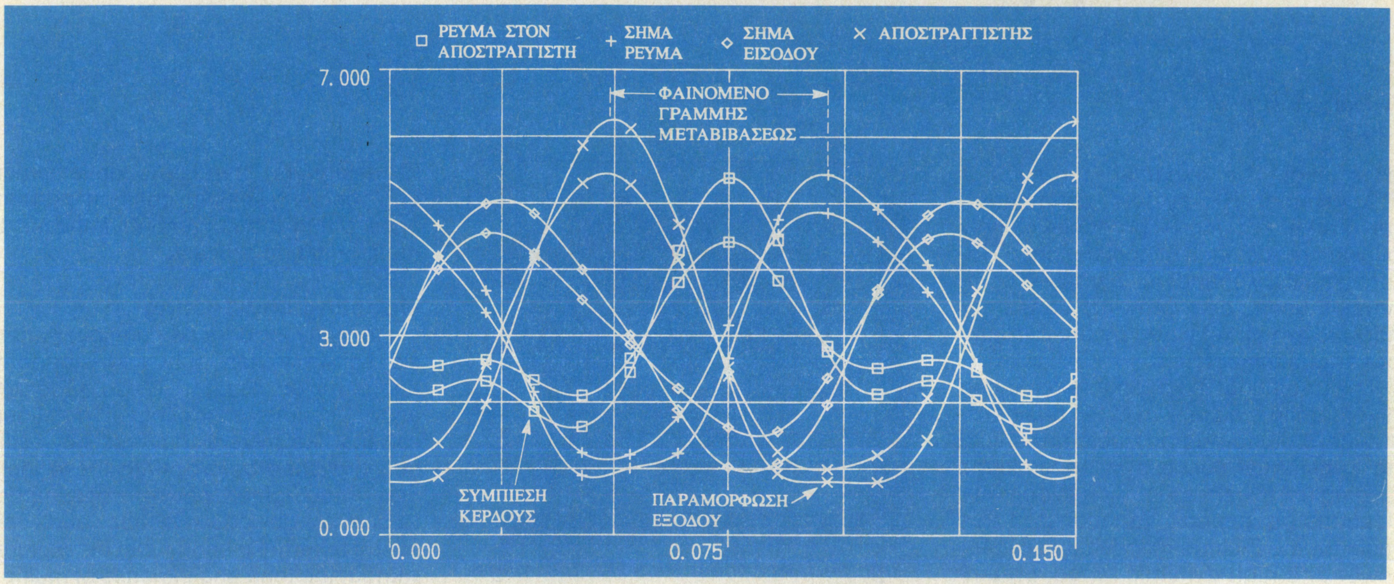
If modern circuit design looks like this to you,



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Mixers Remain a Central Product for RF Applications

By Gary A. Breed
Editor

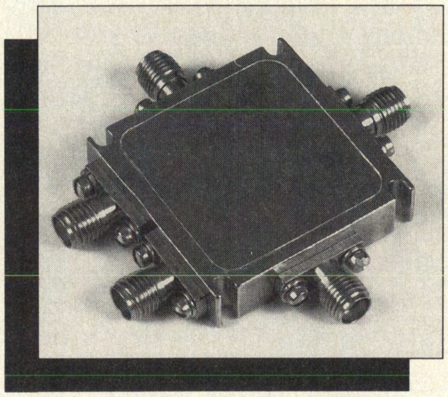
As might be expected for such a universal RF system component, the market for mixer products closely follows trends in RF applications. Reduced military spending, increased commercial uses, and growing consumer markets are all reflected by the demand for various mixer types.

The traditional diode double-balanced mixer family is a stalwart of RF business. Standard diode-ring products are enjoying substantial success in applications ranging from instrumentation to communications equipment. Variations on this basic theme include high local oscillator power and high dynamic range, preamplified local oscillator input, and low threshold diodes. These types have been developed primarily to meet a customer's request for specific performance. Since they can be implemented relatively easily in production, they are viable options.

The sales of other specialized mixer configurations varies considerably. For example, the termination-insensitive type of mixer has seen a noticeable drop in popularity, as engineers often choose to apply lower cost standard mixers with optimum LO levels and output terminations. I and Q demodulators and image-reject mixers closely follow demand for the radar systems, instrumentation, and digital communications systems that are their major end users. Space-qualified components also are tied closely to the development of new satellite systems.

IC Mixers — Silicon and GaAs

Low cost integrated circuits for commercial and consumer products are in demand for these growing markets. Cordless telephones, cellular equipment, radio and television receivers, and wireless control devices have unique demands for low pricing, low power consumption, and adequate RF performance. These are currently best met with silicon technology. Survival in an overburdened spectrum will increase the performance demands without changing the pricing requirements. Silicon IC mixer designers are hard at work refining fabrication processes and topologies, and applications engineers are optimizing circuit implementation of the devices.



High performance mixers like this image-reject model from M/A-COM closely follow demand for high performance systems.

Mixers of this family do not have the signal-handling and distortion performance of diode mixers, although they have the advantages of low cost and low power consumption, usually with conversion gain that can reduce system complexity. Low local oscillator drive, or on-chip oscillator circuitry also serves to simplify system design.

Many of the coming RF applications are attractive applications for gallium arsenide MMIC technology. Cellular systems at 900 MHz, plus the low GHz L-Band and S-Band navigation and communications applications are sizeable markets for integrated front-end subsystems including mixers and other functions. At these frequencies, GaAs performance makes it the technology of choice for many designers.

Like the consumer products targeted by silicon IC makers, many of these applications are quite cost-sensitive. Because GaAs is a more expensive fabrication medium than silicon, high levels of integration are being pursued. New GaAs MMICs combine preamplifiers, buffers, mixers, local oscillators and IF amplifiers on a single die. Like silicon IC designers, the GaAs development engineers are seeking optimum performance within the constraints imposed by the fabrication process.

Modeling and Testing

Recently introduced non-linear analysis software has only just started to affect mixer design. These programs allow accurate modeling of non-linear circuits like mixers that was not easily accomplished before. With these tools, new topologies, power consumption tradeoffs, device selection and optimized external circuitry can be explored without the time and expense of multiple iterations of hardware prototypes.

Testing techniques have also seen recent advances, allowing manufacturers to provide extensive testing using automated systems. Frequency-shifted test sets that work with existing network analyzers allow simplified test setup compared to the test systems made up of several pieces of test equipment that were required only a short time ago. Instead of a controller, two or more signal generators, programmable attenuators, and a spectrum analyzer, new mixer test systems can perform all operations in a single unit with internal software control and a single accessory unit.

One noticeable result of this increased testing capability is a higher level of screening, either provided by the mixer manufacturers or at incoming test at the users' plant. The increased reliability and narrower range of variation in mixer specifications that comes from such testing allows tighter control of design parameters and less over-design.

In summary, mixers represent a central product that is required in virtually all RF products. Although standard products are often utilized in design, the cost and performance requirements of a specific application often dictates a change from the norm. As a result, these demands are driving the development of mixer technology both in building block components and integrated products.

EMI Signal Analysis

By Roger Southwick
EMC Consulting

Although descriptions of methods for measuring EMI signals abound in the literature and in various regulations, methods for analyzing EMI signal data have received far less attention. This may be because the regulations specify only that such data be reported in some specific format. The rationale for analyzing EMI signal data is based on the needs of prequalification testing to determine the causes of noncompliance. Prequalification test methodology is not defined by the regulations, but it is a fundamental requirement for product development. The causes of possible noncompliance found as a result of prequalification testing must be remedied. This requires first that the offending signals internal to the item under test, EUT, be identified, and second that there be methods to evaluate modifications to the EUT. Other analyses include methods for combining sets of data so that the maximum signals of several data sets can be determined.

The most effective way to implement EMI signal analysis is to have the necessary data in some format that can be loaded into a computer. The EMI signal analysis to be discussed here is a part of the EMI Commercial Measurement Program (1). This program measures the EMI signals, which can then be stored on a disk and can then be retrieved for future analysis.

The type of data included in a data set depends on options selected during the actual measurement. The complete, all-option data set consists of a number of identifying parameters, a comment string, the trace data and a data array containing the peak, average and Q-P amplitudes plus the frequency of each individual signal. Most of the analysis is performed on the data contained in the data array. The data analysis menu of the program is shown in Figure 1.

Each data analysis subprogram of the Data Analysis Menu has options to select types of data: peak, average or Q-P. Measurements are made in either of three measurement subprograms: E Field, H Field or Conducted. Each measurement subprogram is divided into measurement bands: E Field has three bands covering 30 to 1000 MHz, H Field has two bands covering .01 to 30 MHz and Conducted has a number of bands that are used in pairs and cover from .01, .15, .45 to 30 MHz. Provision is included to prevent the mixing of data from different measurement subprograms. Also included, when appropriate, are options to select data sets of a specific measurement subprogram taken on a specific phase, for conducted or for mixed or specific polarization for E Field measurements.

The first selection from Figure 1 to be described is "Printout; List & File". This option has two parts: a file listing and a file printout. The file listing will print out the identifying parameters of the files, but no data. A sample of a listing of three files is shown in Figure 2. The file listings are in the order the files were recorded. This option is a convenient way to obtain the file numbers, which must be used to retrieve files for analysis by other analysis subprograms. The file comment is included in the file listing to help identify files. An option to list either all the files on a data disk or only those files from a specific measurement subprogram is provided. The file printout will print out any individual files selected.

As previously mentioned, it is necessary to identify the source of any signals that are in noncompliance. Such signals are mainly the result of intermodulation products of clocks; local oscillators or reference signals internal to the EUT. An intermodulation product is the sum of or difference between the harmonics of these internal signals. The sum of the absolute value of the harmonic numbers is the order of the intermodulation product. Since the level of the harmonic decreases as the harmonic number increases, the level of intermodulation products decreases as the order increases. Third-and fifth-order intermodulation products usually have higher levels.

The Intermodulation subprogram, which is selectable from the Data Analysis Menu of Figure 1, is responsible for identifying the fundamental components of the intermodulation product. A single data set is selected for analysis. The analysis is performed on the frequency data from the data array. The operator then enters into the program the frequency of the internal signals of the EUT, selects the highest intermodula-

*** DATA ANALYSIS MENU ***

Statistical Analysis
Composite Printout
Intermodulation
Composite Plot
Printout; List & File
RETURN

Enter Selection

Statistics	Composite Prt	Intermod
Composite Plot	Return	Printout

Figure 1. Data analysis menu.

File 52	E Field	30 to 200 MHz
Horizontal	Azimuth 0	9: 29: 28: May 30, 1990
Antenna Zero		Limit Level 53
Trial 28/5/90		
File 53	E Field	30 to 200 MHz
Horizontal	Azimuth 0	21: 42: 49: May 31, 1990
Antenna Zero		Limit FCC Class A
Demo of E Field		
File 54	E Field	30 to 200 MHz
Horizontal	Azimuth 0	9: 34: 29: June 2, 1990
Antenna Zero		limit FCC Class A
Testing E Field		

Figure 2. Data disk file list printout.

tion order desired and enters a tolerance (an error term). Since the data is measured in some finite bandwidth, there is usually an error between the actual and the calculated intermodulation product. A tolerance of about half the IF bandwidth used when the data was measured is a reasonable choice to start the analysis. If the results are not satisfactory, the tolerance can be increased. If, however, too large a tolerance is used, the analysis becomes useless, since almost every signal will be part of some IM relationship, and the critical and valuable IM products will be indistinguishable in the total output.

Both two and three signal intermodulation products may be analyzed. The computer does the computations and prints out a list of intermodulation products and a summary of the number of times each internal signal was involved. This information identifies the internal signals that contribute to the intermodulation products that are in noncompliance.

It is often necessary to determine whether differences exist between spectrums. For example, when the EUT is in noncompliance, some modification must be made to bring the EUT into compliance. If a measurement of the EUT spectrum is made before and after the fix, the operator must determine whether the after-fix spectrum is an improvement over the before-fix spectrum.

From the menu of Figure 1, the selection Statistical Analysis offers two statistical hypothesis tests. To implement the calculations, the analysis uses the dBuV signal amplitude values. (Although the use of log values is improper,

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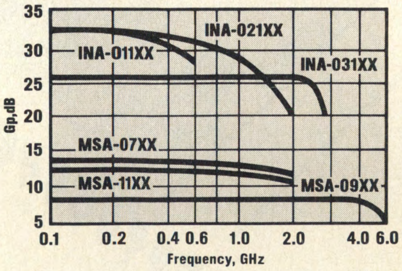
Power Gain vs. Frequency



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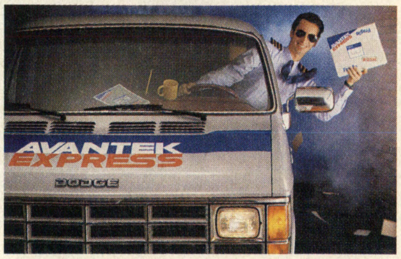

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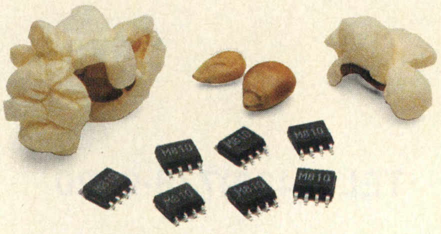

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Statistical Analysis Test

$H_o : u_o = u_1 = u_2 = ... = u_8$

File 21
E Field 30 to 200 MHz
FCC Class A
Azimuth 0 degrees
Polarization Horizontal

Freq. MHz	Peak dBuv/m
35.55	57.80
83.25	57.45
87.75	54.17
92.83	71.82
93.72	73.00
94.84	71.27
96.05	64.05
99.50	69.90

Mean of file 64.93

File 22
E Field 30 to 200 MHz
FCC Class A
Azimuth 0 degrees
Polarization Horizontal

Freq. MHz	Peak dBuv/m
83.25	57.00
87.75	54.05
92.83	71.37
93.72	73.40
94.84	70.47
96.10	63.42
99.52	68.87
181.34	67.45

Mean of file 65.75

Figure 3a. Type #1 statistical analysis.

File 23
E Field 30 to 200 MHz
FCC Class A
Azimuth 0 degrees
Polarization Horizontal

Freq. MHz	Peak dBuv/m
83.26	56.40
87.76	54.25
92.85	71.32
93.72	73.20
94.85	70.90
96.06	64.42
99.50	69.77
181.34	67.42

Mean of file 65.96

CM	103119.33
total SS	1102.54
SST	4.73
SSE	1097.81
MST	2.36
MSE	52.28
F	0.05

$V_1 = 2$ $V_2 = 21$

H_o Accepted no significant difference in means

Figure 3b. Type #1 statistical analysis.

it will not affect the impartiality of the analysis.) The frequency of the signals does not weight the data points. When making this type of analysis it is important to have a sufficient number of data points or signals, since this will increase the statistical confidence of the analysis. It is also necessary to have nearly the same number of data points in each data set to be analyzed. The considerations must be made at the time of the measurements.

The type #1 hypothesis test is (2, 3):

$$H_o : u_o = u_1 \text{ -------- } u_8$$

which states that the means of from 2 to 8 data sets are equal. When this hypothesis is rejected, its statement is not true. This type of test is useful in determining, for example, whether the spectrums of a number of different serial number EUTs are approximately equal. Such a test will determine whether a particular EUT is typical and whether the manufacturing process results in a consistency between serial-numbered EUTs. An example of a type #1 statistical analysis printout using three data sets is shown in Figures 3a and 3b. The input data and the mean of each data set are provided.

The type #2 hypothesis test is (4):

$$H_o : (u_1 - u_2) = 0 \text{ and } H_1 : (u_1 - u_2) \neq 0$$

This test compares the means of two spectrums and is used in the before-fix and after-fix example to determine whether two sets of data are equivalent. A type #2 statistical analysis was performed twice, once with data sets 22 and 23 and again with data sets 21 and 22 (shown in Figure 3a and 3b). The result was that H_o was accepted for data sets 22 and 23 but was rejected for data sets 21 and 22.

There are two types of Composite subprograms. Type #1 is a composite printout of the array data of multiple data sets. The type of data can be selected and up to 20 data sets of different bands from the same measurement subprogram can be combined. The data is ordered by the amplitude of the type of data selected and printed out. An example of a type #1 composite printout for

*** Diagnostic Composite ***

E Field 30 to 200 MHz
mixed polarization
Files 21, 22

Signals Ordered by Amplitude

Freq. MHz	Peak dBuv/m	File	Azimuth	Polarity
93.72	73.40	22	0	H
92.83	71.82	21	0	H
94.84	71.27	21	0	H
99.50	69.90	21	0	H
96.05	64.05	21	0	H
33.55	57.80	21	0	H
83.25	57.45	21	0	H
181.34	67.45	22	0	H
96.10	63.42	22	0	H
83.25	57.00	22	0	H
87.75	54.05	22	0	H
87.75	54.17	21	0	H

Figure 4. Composite printout type #1.

files 21 and 22 is shown in Figure 4. Note that the azimuth and polarity of the data are carried through the analysis and are included with the printout.

The type #2 composite plot is similar to the type #1 composite printout, except that it orders the data based on the magnitude of the difference from a limit. The trace data is also combined in a max hold fashion, and the data is both plotted and printed.

The limit to be used for the type #2 composite plot can be selected by the operator. This permits the analysis of data sets to a number of different limits without having to take additional measurements. It is, however, necessary to measure the initial data using a limit of constant value. The EMI Commercial Measurement Program has five sets of limits per measurement band. The limits can be entered by the operator, so it is a simple matter to enter a limit of constant value for this purpose. Since measurement bands can be combined, it is also possible to plot data over the entire measurement range. An example of a type #2 composite plot from the conducted subprogram is shown in Figure 5. A printout of the individual data sets and the complete composite sorted printout of all the individual signals is also provided by the program.

These types of EMI signal data analysis are extremely useful and very easy to use. The analyses will provide a near-instant diagnostic insight into the causes of noncompliance, as well as providing an impartial evaluation of EMI fixes. Measurement bands of any measurement subprogram may be combined to plot the entire measurement range. Further, it is possible to perform the analysis and plot it at a later date, thus freeing the measurement system for measurements. *RF*

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Conducted Composite Plot
Advantest Model R3361B

Peak Data
Phase mixed
.45 to .30 MHz

Files 1, 2, 3, 4

Freq MHz	dBuz/m	dB	File	Phase
0.55	60.8	3.8	2	1
0.45	60.0	3.0	1	1
4.50	59.0	2.0	3	2
1.29	56.1	-0.9	2	1
5.50	55.8	-1.2	4	1
25.02	55.7	-1.3	4	1

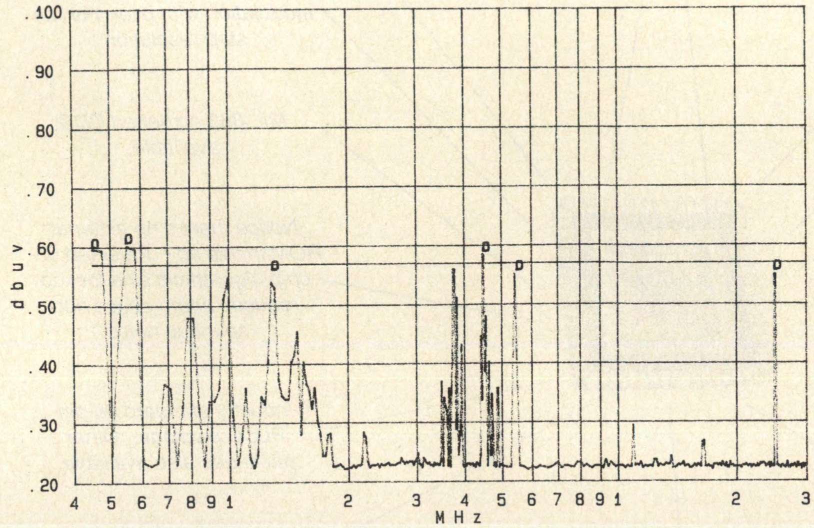
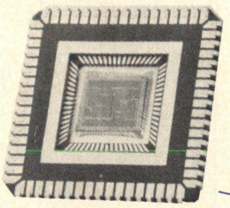


Figure 5. Composite plot type #2.

About the Author
Roger Southwick is owner/president of EMC Consulting, 2716 N. Estrella, Tucson, AZ 85705. He can be reached by telephone at (602) 792-9491.



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INFO/CARD 24

Swept Spectrum Bandpass Filter

By Dr. Roy H. Propst
University of North Carolina, Chapel Hill

As demands have increased manyfold over the past few years for more spectrum availability, we have either gone to higher frequencies or to better utilization of present allocations. New communications modes have arisen, such as spread spectrum systems (1) which transmit voice or data over a swept-frequency carrier and thus time-distribute as well as frequency-distribute spectrum use. This idea has motivated the present work to further develop spread/swept spectrum ideas useful to communications. In the following, we frequency-sweep the center frequency of a bandpass filter in order to achieve two advantages: time multiplexed spectrum for shared use and improved filter bandpass shape characteristics.

VLSI technology has permitted the fabrication of many devices which have changed the outlook of "circuit design", as well as the concept of circuit design itself. We can now think in terms of building blocks rather than single devices, and then assemble these blocks into new structures. One such block is the switched-capacitor-filter (SCF). This device, as usually fabricated, gives lowpass, highpass, or bandpass characteristics at will and the capability to reconfigure externally. This device is based on the "switched-capacitor-resistor", which can be viewed as the simulation of a resistor by a switch and a capacitor in series. An external clock generator controls the opening and closing rate of the switch and therefore the "average" current flow through the switch and into the capacitor, which stores the pulses. Hence, the current flow or "circuit resistance" depends on the switching speed (2). There exists several implementations of the SCF in DIP packages, for example the National MF10 (3). One of the most useful properties of the SCF is the ability to change the center frequency of a bandpass filter implementation by changing the external switching-clock-rate of the filter. This, of course, is quite reasonable since it just changes the equivalent resistance values of the switched-capacitor-resistors in the filter.

In order to achieve the desired results for the swept spectrum bandpass filter, we use a SCF as a narrow-bandpass filter, and sweep this narrow passband over a wide range of frequencies and thereby pass them all - as a function of time. In this manner, we obtain a wider bandwidth than the original SCF, but at the same time keep the original slope at the upper and lower three-dB-break-points, that is the same "ultimate attenuation" as the SCF. This means that we can obtain a high-Q bandpass filter which has a "flatter" passband than normally expected for the same value of Q.

Figure 1 shows the basic design. The "clock" for the SCF is a voltage controlled square-wave oscillator (VCO), this determines the center frequency for the SCF. It should be noted that the center frequency of the SCF is not equal to the frequency of the VCO, but proportional to it by some scale factor -for example in many cases 1/100-th. We now use a sawtooth generator to control the VCO. The output of the VCO is therefore a constant-amplitude variable frequency square wave - a frequency modulated square wave whose frequency varies periodically and linearly with time. The frequency span-range of the VCO thus

TWO POLE SCF WITH A Q-VALUE OF 10

	Lower Bandwidth (Hz)	Lower Points (Hz)	Upper Points (Hz)	Upper Bandwidth (Hz)
3 dB points	31.6	300.0, 331.6	2442.9, 2700.0	257.1
6 dB points	141.9	252.3, 394.3	2054.6, 3210.3	1155.8
60 dB points	8810.9	11.3, 8822.2	91.8, 71837.3	71745.4

SCF SHAPE FACTOR (SF1) = 62.0
SWEPT FILTER (SF2) = 24.3
IMPROVEMENT: RATIO SF1/SF2 = 2.6

Table 1. Shape Factor improvement using swept spectrum filter for a two pole SCF implementation.

THREE POLE SCF WITH A Q-VALUE OF 10

	Lower Bandwidth (Hz)	Lower Points (Hz)	Upper Points (Hz)	Upper Bandwidth (Hz)
3 dB points	31.6	300.0, 331.6	2442.9, 2700.0	257.1
6 dB points	104.9	267.3, 372.2	2176.3, 3030.7	854.3
60 dB points	2921.5	33.7, 2955.2	274.1, 24063.3	23789.2

SCF SHAPE FACTOR (SF1) = 27.8
SWEPT FILTER (SF1) = 8.7
IMPROVEMENT: RATIO SF1/SF2 = 3.2

Table 2. Shape Factor improvement using swept spectrum filter for a three pole SCF implementation.

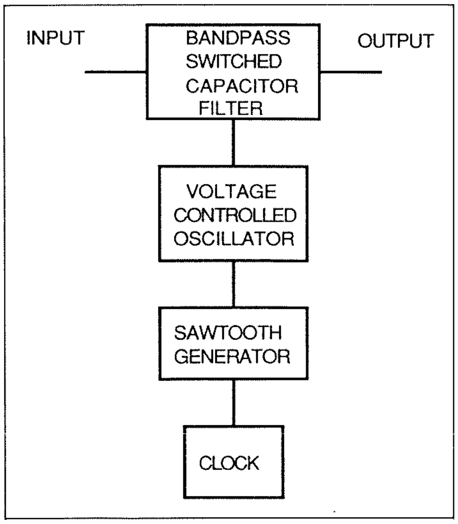

Figure 1. Basic components of the swept spectrum bandpass filter.

sets the bandwidth of the resultant swept spectrum filter. There is, as can be seen in the figure, one more thing needed. A clock is needed to run things, i.e. set the frequency of the sawtooth and thus the period of the frequency sweep of the filter.

Programmable Spectrum

A simple modification of the basic design is given in Figure 2. Here, we add a gate in the output of the filter. We control this gate with a pulse derived from and synchronous with the clock. We generate a control pulse of variable length and delay from the clock output pulse. This control pulse determines the "on" time for the gate, as a percentage of the total sweep cycle. Thus, we switch the output on (or off) for some portion of the sweep cycle. If for example, we use this pulse as a "turn off" pulse and we are sweeping frequency with time,

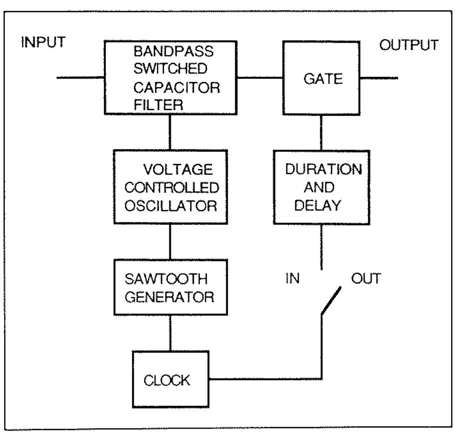

Figure 2. Modification of swept spectrum filter to include notch.

this pulse occurs at some passband frequency and thus means that some part of the frequency spectrum through which we sweep is missing from the filter output signal, i.e. we have created a notch in the filter output. We can use more sophistication and have the output switched off and on for any part of the frequency sweep and have a "fully programmable" output spectrum.

Filter Bandpass Shape

The advantage offered for improvement in the shape of the passband at the upper and lower 3-dB points can be seen by comparison to an "ordinary" bandpass filter. For an ordinary bandpass design with a Q of 10 and two poles, for example. We get a slope of 12 dB per octave or 40 dB per decade at the "ends". Here, we can get a measure for the improvement by comparing the shape factors of the filters. The shape factor (SF) is defined as the ratio:

SF = (Bandwidth at 60 dB down)/(Bandwidth at 6 dB down).

This number is a constant and will be determined by the original SCF filter design. For example, we can choose a two pole SCF and design the overall swept filter for "communications quality" speech, via a 3-dB bandwidth of 2.4 kHz, swept from 300 Hz to 2.7 kHz. For an SCF with a Q of 10, at the lower end of the spectrum, 300 Hz, it has center frequency of 315.79 and 3-dB points of 300 Hz and 331.48 Hz; at the upper end of the spectrum, 2.7 kHz, a center frequency of 2571.43 Hz and upper and lower 3-dB points of 2700 Hz and 2442.86 Hz. Table 1 shows the 6 dB and 60 dB down frequency bandwidth values at the upper (2.7 kHz end) and the lower (300 Hz) end of the swept filter design, as well as the shape factor of the SCF at these points, and the resultant overall shape factor of the swept filter. You can see a 2.6:1 improvement in shape factor for this two-pole filter. Similarly in Table 2 we give the results for a three pole filter, things get better with an improvement of 3.2:1. Thus, for the design, we have independently adjustable: shape factor, center frequency, and bandwidth. The shape factor is determined in hardware, but the center frequency and bandwidth voltage-variable are adjustable parameters.

Results

The swept spectrum filter was implemented using the National MF10; variable, external, sawtooth and clock generators were used to allow maximum flexibility during evaluation. System operating parameters were chosen for communications quality speech, defined above. The SCF Q-value was chosen to be ten, this is a constant fixed by the ratio of two resistor values. Since this has a fixed constant value, it means that the "instantaneous bandwidth" of the filter changes as the center frequency is swept from 300 Hz to 2.7 kHz. A Q of ten is not an extremely "high" Q-value and is easily implemented without any associated ringing problems. Measurements were made with a spectrum analyzer to confirm the performance and the results are in Figures 3 and 4. Figure 3 shows the response of the SCF at the lower and upper ends of the swept filter range, in this case one and four kilohertz. The constant Q-value of the SCF imposes the exhibited different bandwidths of the responses. Now, if the SCF is swept between lower and upper frequency limits, the overall response can be seen in figure 4. The "wiggles" in the response curves are produced by the "sampling" of the spectrum analyzer. The tracking generator of the analyzer was swept across the frequency range at a very low (seconds) rate; whereas the sawtooth swept the filter at a much higher (milliseconds) rate. This method produces an "average spectrum" display; which is, in fact, the result obtained in normal operation of the swept filter.

As a final test for comparison, an audio subjective evaluation of the filter was made with speech and music. The bandpass character was clearly apparent as the minimum and maximum sweep frequency values were changed for several values of sweep rate. Sweep rates are clearly related to the Nyquist sampling rate. However, here we are continuously sampling, at the sweep rate, in the frequency domain rather than digitally sampling in the time domain, as is usually the case. Fidelity was found to be good for a sweep rate equal to the upper passband frequency, and very good for twice this value, which corresponds to the Nyquist rate.

Conclusions

For this design, as is usually the case, a tradeoff is required between added complexity and improved filter characteristics. Although many filter applications may not warrant the described level of sophistication, the increased

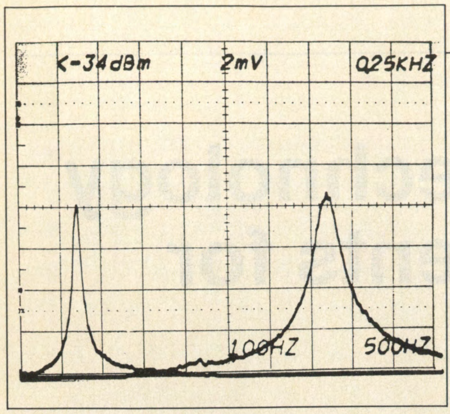

Figure 3. Frequency response of switched capacitor filter at upper and lower extrema of sweep range.

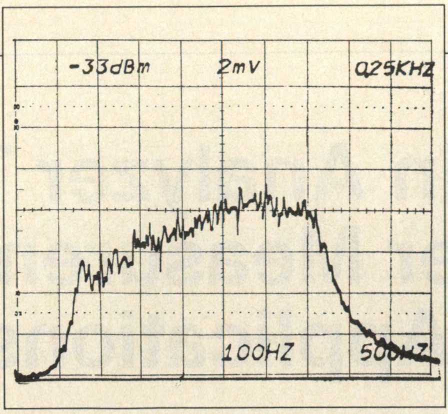

Figure 4. Frequency response of switched capacitor when swept over the same frequency range as in Figure 3.

flexibility may be justified for some. Further, it is expected that enough demand and applications could encourage technology to develop "single-chip" implementations for the swept spectrum filter. The present CMOS fabrication technology places somewhat restrictive frequency limitations on the current generation of SCFs. The limit usually quoted is the product of the center frequency times the Q-value of the implementation. This is now on the order of several hundred thousand Hertz to a megahertz or so, which means we are essentially limited to the audio frequency range when using SCF Q-values on the order of tens. Presumably, future fabrication process evolution will bring this up into the megahertz domain.

RF

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About the Author

Roy Propst is an electrical engineer who teaches solid state devices and communications courses at University of North Carolina, Chapel Hill. He teaches in the biomedical, physics and astronomy departments. He can be reached at Building E, Box 10, CB7003, Chapel Hill, NC 27599. Tel: (919) 966-4578.

New Trends in Analyzer Technology Provide Faster Measurements for Narrowband Applications

By Manfred Bartz
Hewlett-Packard Company

Depending on the nature of the application, the term "narrowband" has grown to include a broad range of measurements. Narrowband measurements traditionally have been exemplified by spectrum or network measurements associated with the bandwidths of the IF stages in super-heterodyne receivers. Examples include determination of narrowband noise, identifying in-band spurious signals, measuring third-order intermodulation distortion, or characterization of bandpass filter shapes. Although, the frequency spans involved may be "narrow" relative to a receiver's front-end, bandwidths may vary considerably from MHz for video applications, to kHz for audio down to even tens of hertz or less for mechanical phenomena.

Other receiver-related examples arise in the design of local oscillators. Critical to providing clean frequency translation in mixers, local oscillators must be characterized for close-in spurious and single sideband phase noise. These measurements typically require finer resolution bandwidths measured in hertz, with spans in tens of hertz. One example involves the measurement of the undesirable although ubiquitous 60 Hz sidebands, which can often plague even the best of oscillator designs.

Applications requiring even narrower bandwidths can arise in solving problems associated with identification and elimination of microphonics. Examples include capacitors whose reactances are affected by near-field dielectric interference or, perhaps more commonly, mechanical mechanisms modulating the signal through a crystal filter. Such phenomena can require sub-hertz measurement resolution and spans measured in hertz.

As a general case, spectrum measurements involving the extraction of small signals from noise, be it for surveillance applications or merely to observe and remove unwanted spurious signals, are often narrowband by nature. In order to achieve adequate signal-to-noise ratio, selection of narrower resolution bandwidths typically improves the resolution and sensitivity of the measurement.

Narrowband Measurements: The Tradeoffs

Narrowband measurements often suffer from slow measurement times. This is a natural consequence of the superheterodyne technology found in most spectrum analyzers today. The local oscillator sweep rates are limited by the settling time of the narrowest IF filter. As the filter bandwidths become narrower, their respective Q's increase, resulting in longer settling time constants. Hence, as smaller resolution bandwidths are chosen, the local oscillator must sweep more slowly to maintain a given accuracy.

On the other hand, as the filter bandwidths become narrower, the total noise power in the filter passband decreases proportionately, thereby increasing receiver sensitivity. This is one of the fundamental advantages of narrowband measurements, but it comes at the expense of slower sweep rates. For every decade drop in resolution bandwidth, the noise floor drops by an additional 10dB, as long as the noise is dominated by the broadband thermal noise in the front-end and not incidental spurious mechanisms in the detector. Hence, the narrowest available resolution bandwidth typically sets the practical lower limit to receiver sensitivity.

A natural way to speed a narrowband

DISCRETE FREQ
110 000 000.000Hz

Discrete Sweep Table

ROW	FREQUENCY (Hz)	RES BW	SETTLING
▷ 1	110 000 000.000	10 Hz	0.500SEC
2	111 000 000.000	10 Hz	0.500SEC
3	112 000 000.000	10 Hz	0.500SEC
4	114 000 000.000	10 Hz	0.500SEC
5	116 000 000.000	10 Hz	0.500SEC
6	120 000 000.000	100 Hz	0.037SEC
7	127 000 000.000	100 Hz	0.037SEC
8	128 000 000.000	100 Hz	0.037SEC
9	129 000 000.000	1 kHz	0.004SEC
10	130 000 000.000	1 kHz	0.004SEC

"FREQ", "RES BW", "SWEEP TIME" SELECT COLUMN, UP/DOWN ARROW KEYS SELECT ROW.

DISCRETE FREQ
FREQ STEP
STEP & INSERT
INSERT ROW
DELETE ROW
CLEAR TABLE
LIN SWP→ DISCRETE
RETURN

Figure 1. User table for entry of discrete frequencies on HP 3577B.

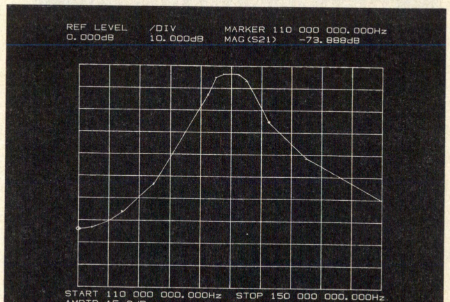


Figure 2. Discrete sweep bandpass filter measurement based on table in Figure 1.

measurement is by selecting a smaller span over which to sweep, since the sweep time is proportional to the span. However, if the measurement goal includes a certain resolution or sensitivity requirement, this may also require very narrow resolution bandwidth filters. The sweep time decreases proportionately to span, but typically increases in inverse proportion to the square of resolution bandwidth:

sweep time ∝ $1/RBW^2$

Therefore, for a given span/resolution ratio, the sweep time increases faster than reductions in span can compensate. At very narrow resolution bandwidths, the sweep times can become unacceptably long.

New Analyzer Trends Provide Faster Solutions

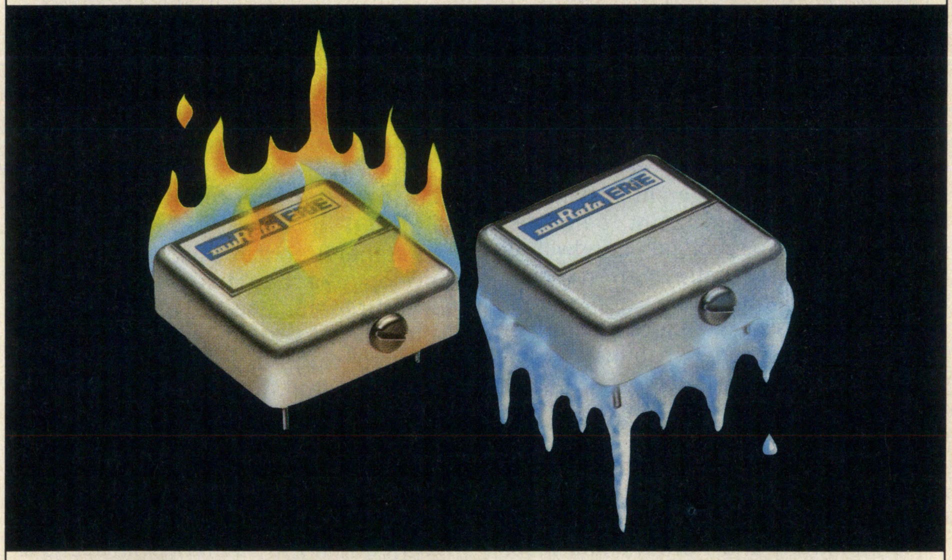
Recent years have seen impressive improvements in the sophistication and performance of frequency domain instruments. These are attributable to improved analog designs and better analog components as well as greater use of digital hardware and microprocessors. In the future, increased utilization of digital hardware and software will continue to be the greatest technological factor behind next generation spectrum/network analyzer improvements.

The following recent measurement improvements exemplify this trend with respect to speeding up measurements for narrowband applications. All three examples are made possible by the benefits of digital techniques in analog measurements. As such they can be classified as feature (software), algorithm (firmware), or block diagram (digital hardware) related.

Feature Improvements: Discrete Sweep

If the frequency of the signal to be measured is precisely known, and the analyzer has sufficient frequency stability, then one method to make faster measurements is by means of a CW approach. The analyzer is tuned directly to the frequency points of interest and a CW measurement is performed, thereby obviating the need to observe the intermediate data over the entire span. This is appropriate when the remaining span contains no useful information and the signal is known to be sufficiently above the noise floor.

This measurement technique has been automated as a feature in newer generation analyzers as "discrete sweep." The HP 3577B network analyzer has elevated this capability to a higher level of sophistication by allowing a user to program a table of up to 51 discrete frequency points for performing a series of measurements in succession, each with an independent resolution bandwidth and settling time. This has particular utility in production environments, by allowing the user to observe the passband and stopband



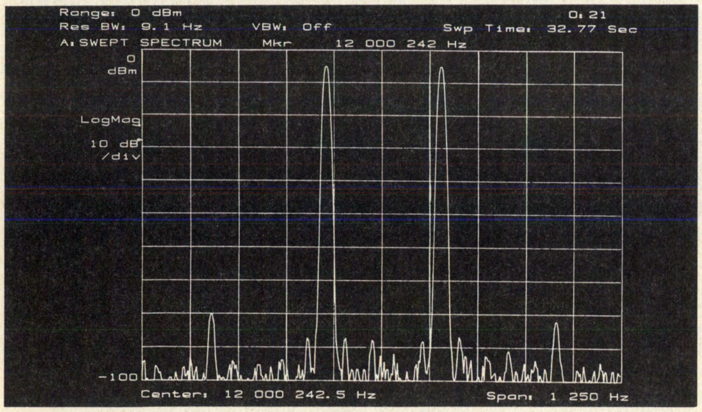


Figure 3a. Two-tone third-order intermodulation distortion measurement using conventional sweep on HP 3588A.

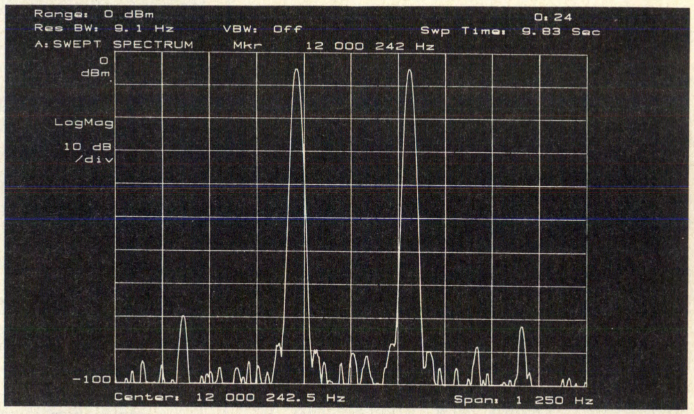


Figure 3b. Same as measurement example of previous figure using oversweep compensation. Note sweep time is 4 times faster.

characteristics of a bandpass filter with varying degrees of resolution. This allows the filter response to be observed over an entire span without unduly compromising measurement speed.

Figure 1 depicts a user generated discrete frequency table for the bandpass filter characterization shown in Figure 2. The measurement shown was tailored for different point densities, settling times and resolution bandwidths across the filter shape. Several user features facilitate table generation. These include multiple stepped entry of frequency points and a softkey which generates a default table from the current linear sweep measurement parameters.

Algorithmic Improvements: Oversweep Compensation

As spectrum and network analyzers have pushed the limits of accuracy, resolution and dynamic range, measurement speed has become ever more important, even for broadband applications. As measurement features have become more sophisticated, algorithmic correction and control have provided improved measurement capabilities.

The HP 3588A spectrum analyzer uses an algorithmic correction approach in firmware to provide previously unprecedented sweep rates for given resolution bandwidths. Because the settling time errors that result from oversweeping too fast are predictable and repeatable for digital resolution bandwidth filters, the errors can be compensated for "on the fly" with each sweep. This is accomplished without compromising measurement accuracy, yielding sweep speeds of up to several times faster than previously attainable. Figure 3 illustrates the dramatic improvement that oversweep compensation provides. The figure compares an audio third-order intermodulation measurement performed on the HP 3588A both in oversweep mode and conventional sweep for comparison. Note that in oversweep mode the sweep time is four times faster.

Diagram Improvements: Digital IF's

Another trend in spectrum analyzer development has been the evolution toward greater use of digital hardware in the block diagram. With IF's becoming increasingly digital, new instruments can benefit from the power of digital signal processing. The HP 3588A, whose block diagram is based on a digital IF, uses FFT calculations to provide a new measurement mode called "narrowband zoom." At any carrier frequency to 150 MHz with spans of 40 kHz or less the instrument can perform measurement updates up to seven times per second and achieve resolution bandwidths measured in millihertz of resolution. Receiver sensitivities are thereby no longer strictly limited by the narrowest available analog resolution bandwidth.

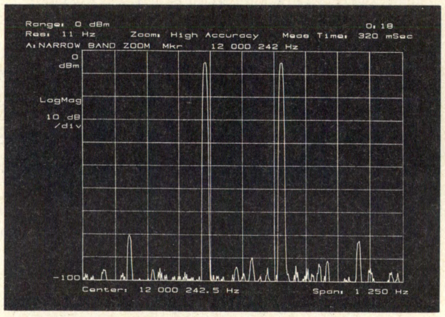


Figure 4. Same as in Figure 3 but using narrowband zoom mode. Measurement time is 100 times faster than conventional sweep shown in Figure 3a.

Figure 4 demonstrates narrowband zoom for the same third-order intermodulation measurement example of Figure 3. For the same effective resolution, the measurement is updated every 320 msec. This is almost 100 times faster than the conventional sweep of Figure 1. Moreover, with response times of up to seven times a second, this mode can provide effective "real-time" observation of carrier modulation in spans of 40 kHz or less.

Summary

Former trends in spectrum/network measurement have focused on providing increasingly better measurement accuracy and dynamic range, often at the expense of measurement time. Current and future trends should provide the same or better measurement capabilities with dramatically faster measurement times. The examples cited have particular significance to narrowband applications. The techniques and technologies being developed in spectrum/network analyzers today provide savvy RF designers with invaluable additions to their measurement tool kits. ***RF***

About the Author

Manfred Bartz is an RF design engineer for Hewlett-Packard. He holds a BSEE from the University of Colorado, Boulder, and MSEE from the University of Washington. He can be reached at Helwett-Packard, Lake Stevens Instrument Division, 8600 Soper Hill Road, Everett, WA 98205-1298. Tel: (206) 335-2647.

Using SAW Double-Mode Resonators for First IF in Cellular Radios

By Sean Broderick
Goldstar Products Company

As cellular technology progresses, engineers are constantly researching and developing new components to improve reliability, decrease cost and make products easier to manufacture. Surface acoustic wave (SAW) devices have been utilized in many diverse electronic applications since their discovery by Lord Rayleigh in 1885. The most common application using SAW devices has been for intermediate frequency (IF) for modern television receivers. Initial SAW filters were of little interest to cellular receiver designers due to their high insertion loss. Now, SAW technology has progressed to allow realization of filters with much lower insertion losses (<5 dB), and circuit Q's that approach 15,000. This opens new doors for applications in cellular telephone products. This paper will discuss the application of a double mode SAW resonator for first IF in cellular radio, and will describe performance results of the receivers tested.

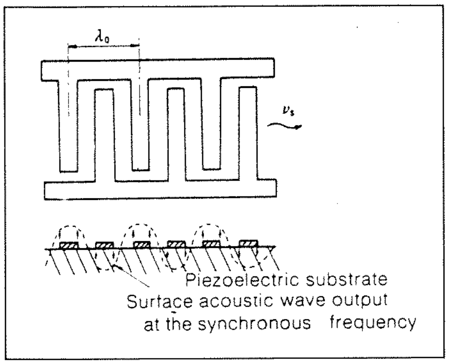


Figure 1. Surface acoustic wave propagating through substrate at desired center frequency.

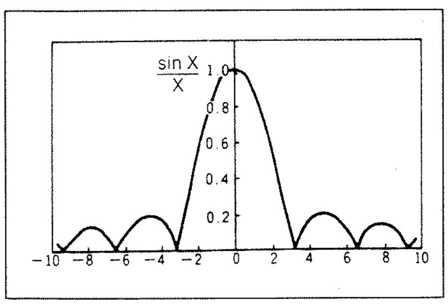


Figure 2. Sin X/X characteristics of SAW filter.

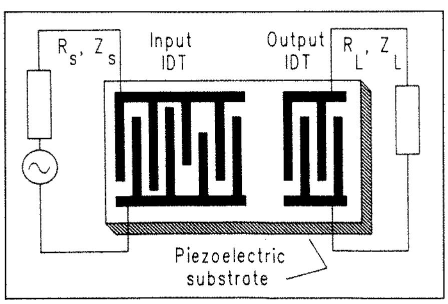


Figure 3. Simple diagram of SAW filter with input/output IDTs.

A SAW device utilizes a single piezoelectric substrate and two voltage excited interdigital transducers (IDTs) to propagate signals. When an impulsive voltage is applied on the interdigital electrodes, strains with opposite phases are generated between adjacent electrodes due to the piezoelectric effect, resulting in the excitation of surface acoustic waves. The interdigital transducer is made up of a photolithographic pattern whose wavelength (λ_o) is that of the desired center frequency (F_o). The center frequency is given by the expression:

$$F_o = \mu S/\lambda_o$$

The bandwidth (δf) of the filter is a function of the proportional relationship of:

$\delta f/F_o$ 1/N where N is the number of electrode pairs.

The maximum output at the center frequency (F_o) coincides with the surface acoustic wave output excited at the spaces between the IDTs, and inversely, the output becomes minimum when the frequencies do not coincide (Figure 1). The frequency selectivity of the surface acoustic wave has characteristics proportional to Sin X/X and therefore can be used as a filter (Figure 2).

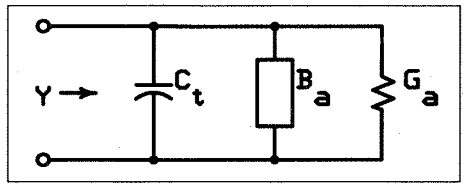


Figure 4. Equivalent circuit for IDT in terms of radiation conductance G_a, susceptance B_a, and transducer capacitance C_o.

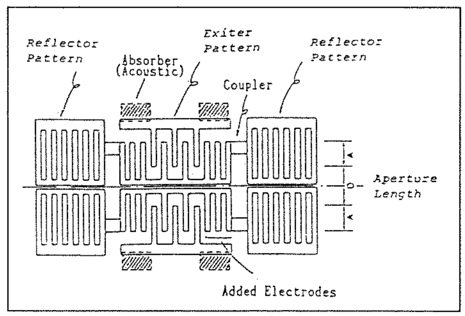


Figure 5. Detailed structural diagram of double mode resonator filter used in test.

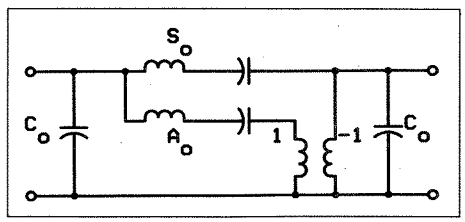


Figure 6. Equivalent circuit of double-mode resonator filter.

An elementary SAW filter with input/output IDTs on a single-crystal piezoelectric substrate is shown in Figure 3, while Figure 4 gives an equivalent circuit for each IDT in terms of radiation conductance G_a, radiation susceptance B_a, and transducer capacitance C_t. It should be noted that the position of the IDTs inside the resonant cavity of the SAW device is critical. This position directly affects the response of the filter by changing the efficiency of the IDTs to launch and receive SAW motion. The insertion loss of a SAW device is a

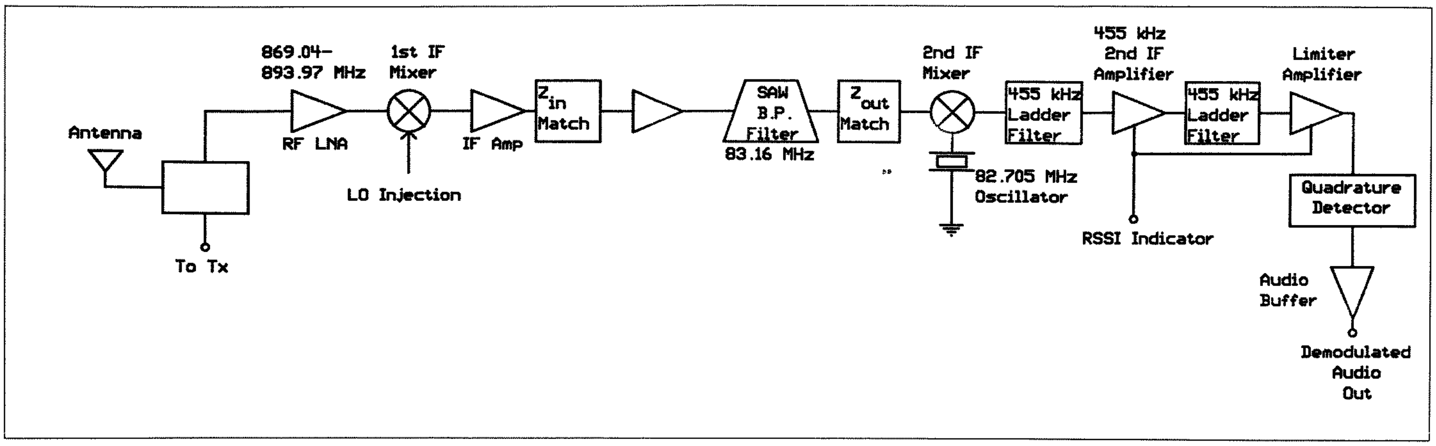


Figure 7. Typical cellular receiver block diagram.

function of the impedance match between the two IDTs and the source/load impedances presented to it.

Until recently, surface acoustic wave resonators were used primarily in oscillator design and not in filter applications. Multimode operation of resonators is achieved by increasing the resonant distance between the IDTs. These high-Q surface wave resonators allow for the realization of cellular first IF filters that would traditionally require multiple crystal filter elements. The most efficient SAW filter architectures for application in cellular IF are double-mode resonators. This newly developed process utilizes laterally-coupled double-mode resonators to achieve circuit Q's of up to 15,000. This high circuit Q is 150 percent greater than conventional SAW filter design technologies. Figure 5 shows a more detailed structure example of a double-mode SAW resonator filter, and its equivalent circuit is shown in Figure 6.

Design Considerations

A block diagram of a typical cellular receiver is shown in Figure 7. A traditional double conversion approach is shown using an IF system integrated chip. The front end of the approach consists of a duplex filter for transmit and receive sections. A bipolar low noise amplifier with 12 dB of gain follows the filter. This design utilized a passible double balanced mixer to reduce harmonics and improve isolation. An active mixer approach could have been used in this design, but would have required extra filtering. Because a passive mixer was used, a first IF amplifier at 83.16 MHz was necessary to overcome the losses incurred. The output of the IF amplifier was impedance matched using a resistive network. Although this is a very lossy way of presenting 50 ohms, it ensures that the SAW device will see a good match even out of its bandwidth. An impedance matching network was also used on the output of the SAW to match to the input of the IF system. The second mixer utilizes a butler overtone crystal oscillator for conversion down to 455 kHz. The signal is then filtered using a two pole ladder filter before it is input to the second IF amplifier. This amplifier has approximately 40 dB of gain and an AC bandwidth of 41 MHz. A second ladder filter is used before the signal is input to the limiter amplifier. The limiter amplifier has approximately 62 dB of gain and an AC bandwidth of 28 MHz. The output of the limiter amplifier is then buffered to an internal quadrature detector. An audio buffer is then used to

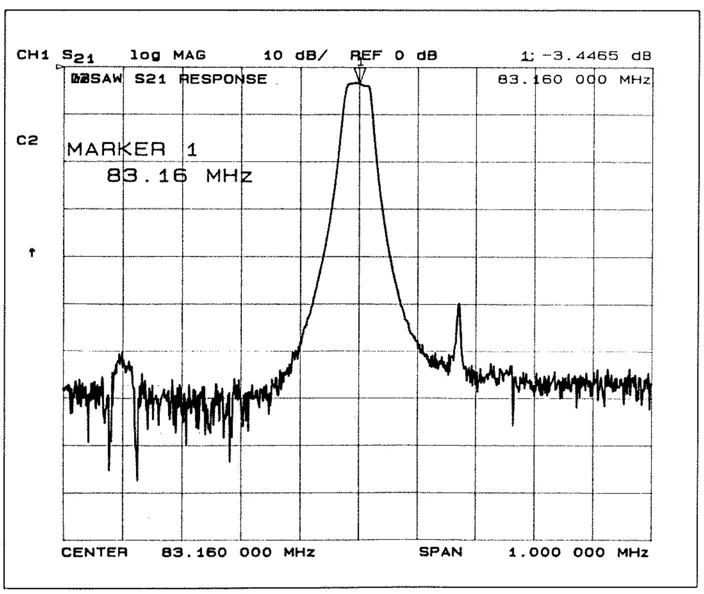


Figure 8. Double-mode resonator S_{21} response.

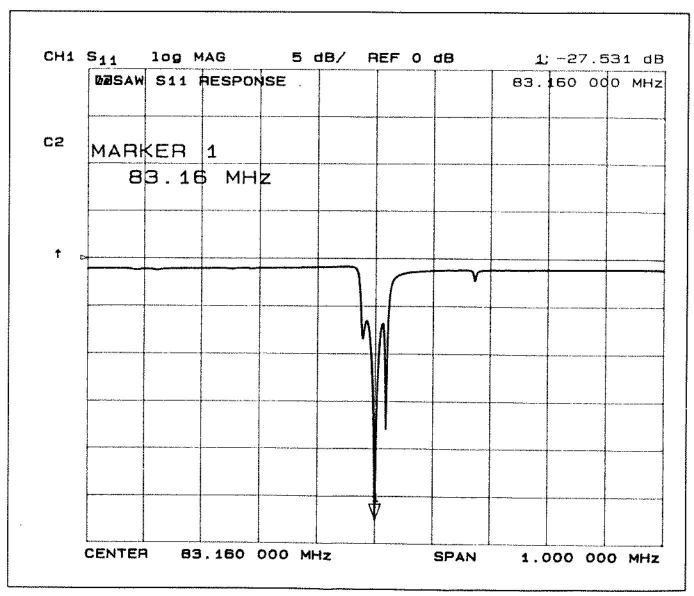


Figure 9. Double-mode resonater S_{11} response.

amplify the demodulated signal. A received signal strength indicator (RSSI) output is derived from the summed stage currents in the limiting amplifiers. The RSSI is used by the logic section of a cellular telephone to determine the strongest channel.

Manfacturability

Because the double-mode SAW device is fabricated using a photolithographic process, the repeatability of device performance is very predictable. This is very important in a manufacturing environment where time for tuning and testing must be kept to a minimum. Compared with a traditional four pole crystal IF filter in cellular radio which requires typically three tuning inductors, the double mode resonator requires no tuning. THe double-mode resonator is also enclosed in a hermetically sealed package which protects the device from moisture and flux. Unlike a crystal filter which is susceptible to fracture if dropped, SAW devices can easily meet tight specifications for acceleration, shock and vibration. This is especially attractive for a hand held cellular telephone customer who accidentally drops an expensive radio. Double-mode resonator filters are available in a variety of packages ranging from through-hole standard packages to custom surface mount devices.

Test Results

The cellular receiver module was constructed on G-10 Fiberglass material and utilized a mixture of surface mount and through hole construction. The SAW filter used in this design was the SWS 83A from Toko. The S_{21} filter response of the SAW filter alone is shown in Figure 8, return loss is shown in Figure 9. Two potential problems spurs were discovered with -56 dB at 161 MHz and -70 dBm at 164 MHz. The manufacturer responded that this was due to impurity in the substrate. The group delay of the unit was measured at 5.5 us typical between 83.15-83.17 MHz. Power was applied to the receiver and with slight tuning the unit began to output demodulated audio. The first test that was run was SINAD (a ratio of signal plus noise and distortion to noise and distortion present at the audio output.) The SINAD for a cellular product must be 12 dB for a -116 dBm (max) input signal with a 1 kHz modulating tone and ±8 kHz deviation. The SINAD for the first model tested was -119 dBm over the temperature range of -30C to 60C. An intermodulation ratio test as describe din EIA IS19C was performed and the unit had a typical intermodulation level of 85.8dB over the specified temperature range. The unit was then tested for first mixer RF image rejection and second mixer image rejection. These results were well within the required levels. The adjacent (alternate) channel selectivity is a measurement of a receiver's ability to receive a modulate input signal on its assigned channel in the presence of a second modulated input frequency spaced 30 kHz (60 kHz) above or below the assigned channel frequency. Alternate (60kHz) and adjacent (30 kHz) channel rejections were measured suing a RSSI test method and were found to be 21 dB minimum for the adjacent channel and 65 dB for the alternate channel over the required temperature range.

Conclusion

Because of their small size, low cost and outstanding performance characteristics, SAW devices are finding more and more applications in consumer products. This article proved that a SAW double-mode resonator filter could be used in an application that traditionally would require a multi-pole crystal filter. Certainly cellular telephone demand offers a strong development incentive for SAW manufacturers and future generations of cellular telephones may rely upon SAW devices for all filtering needs.

RF

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About the Author

Sean M. Broderick is a System Engineer for cellular mobile products and a Project Engineer for portable cellular telephones for Goldstar Products Co., Inc., 7955 East Redfield Road, Scottsdale, AZ 85260. He can be reached at (602) 991-5526 ext. 174.

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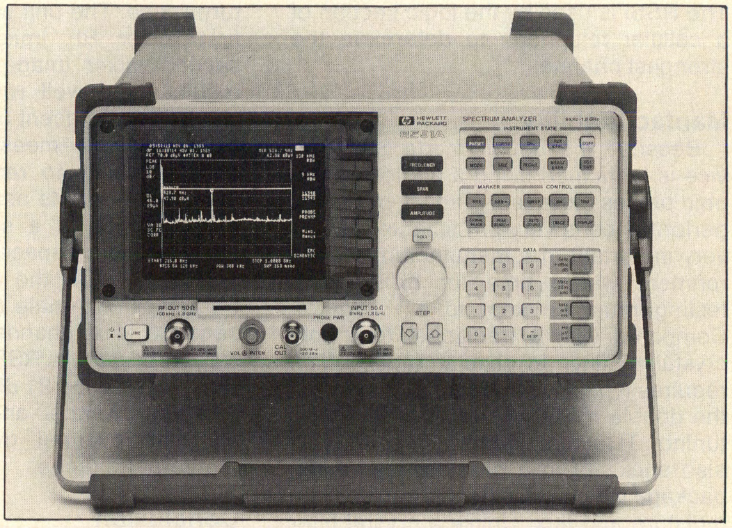

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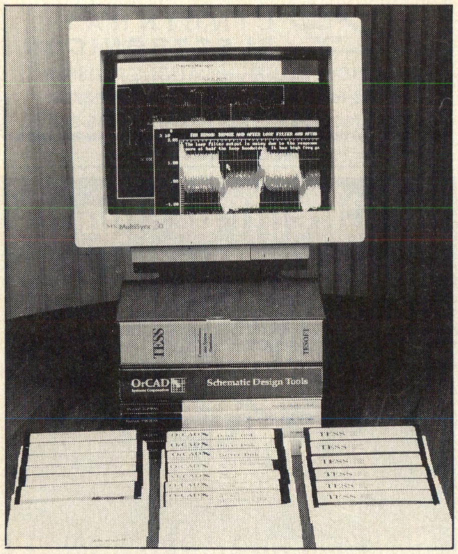

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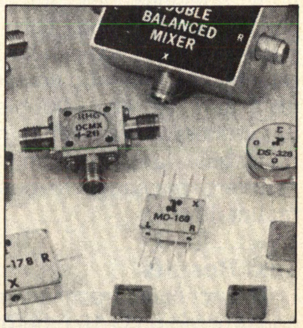

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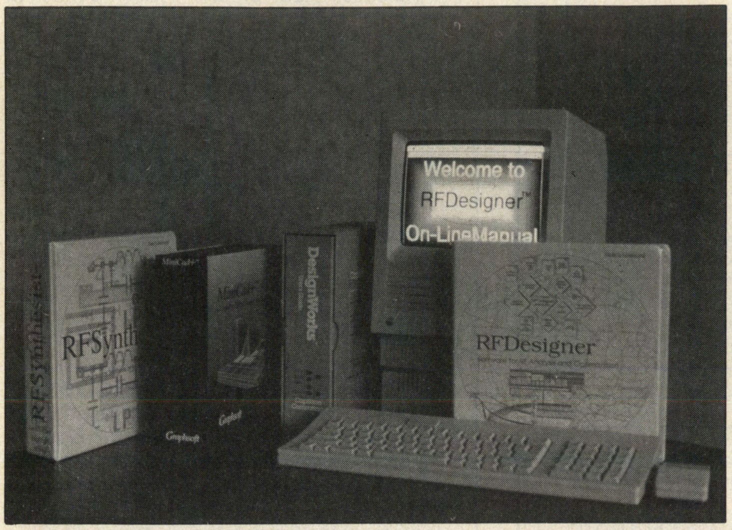

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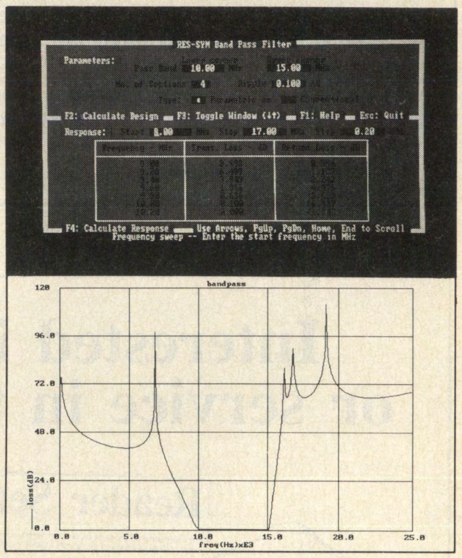

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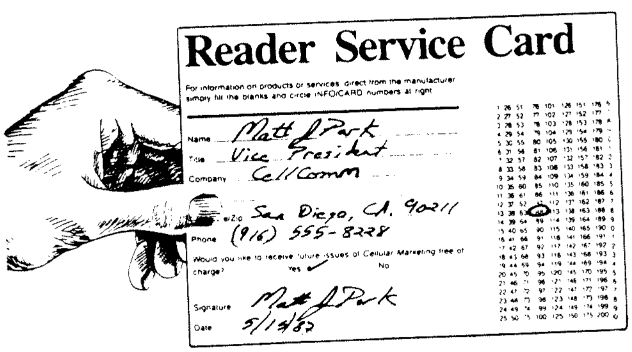

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Circuit Reference Database

Circuit Search announces Version 1.08 of their database of reference articles containing practical electronic circuits. Nearly 12,000 articles and papers from over 250 technical and scientific journals and magazines are contained in the database. These sources cover disciplines where electronics are applied. The database is installed on a hard disk, using an IBM-PC compatible computer, and can use either its own menu-driven front-end to locate circuits using keywords, or it can be used with a dBase^R III/dBase III+ compatible database management program. Pricing is $375.00, including one semi-annual update. Additional updates are $40.00 for one, $65.00 for two.
Circuit Search
INFO/CARD #233

IC Product Database

With the Integrated Circuit Technology Database from Cypress Information Resources, users can access the offerings of over 350 IC companies. A few of the classifications included in the database are ASIC, VMEbus, VXIbus, PSK and FSK modems, linear, HDTV, audio, DSP, op amps, oscillators, DDS, satellite, cellular and others using ECL, GaAs, CMOS, TTL or other fabrication technologies. The license for the database is available for $250, which includes the data diskettes, a printed version of "The Integrated Circuit Product & Market Guide" and an instruction manual.
Cypress Information Resources
INFO/CARD #232

EMI Modeling

GEMS (Graphical EMI Modeling Spreadsheet) is a software tool which allows engineers to lay out the elements of most frequency domain EMI situations as graphic elements in a spreadsheet format. The software contains a library of common EMI sources, coupling paths, victims, and transfer functions. Results are exportable as bitmaps to most word processors and graphics programs. The program requires the Microsoft Windows 3.0 operating system. List price is $3900.
Atkinson Engineering
INFO/CARD #231

RF literature

Catalog On-Disk

Litton Electron Devices announces a new catalog on an IBM PC compatible computer disk, featuring their traveling wave tubes, power supplies and subsystems. Product information is accessible by application, product type, or product number. Applications include ECM, radar, communications and instrumentation. All tube and power supply data may be printed from the screen. Monochrome, EGA and VGA monitors are supported.
Litton Electron Devices
INFO/CARD #250

1991 Catalog

The Standard Product Catalog from ILC Data Device Corp. includes power hybrids, solid-state power controllers, data converters, and 1553 data bus products. A comprehensive listing of applications notes is also included. General technical information augments the detailed product data, along with information on the company's engineering, design and manufacturing capabilities.
ILC Data Device Corp.
INFO/CARD #249

Analog ICs

A new short-form catalog is available from Maxim Integrated Products, with complete listings of commercial and military products. Products included are A/D and D/A converters, DC-DC converters, RS-232 interface devices, amplifiers, buffers, video products, analog switches, multiplexers, active filters, and voltage detectors.
Maxim Integrated Products
INFO/CARD #248

Selection Guide on Disk

Burr-Brown announces the High Performance Electronics Selection Guide disk for IBM PC-compatible computers. The disk contains over 1250 component models with cross-references, pricing, sales and ordering information. Analog circuit functions, A/D and D/A converters, operational amplifiers, multipliers, and instrumentation amplifiers are among the products listed. A new product section is included and updates can be down-loaded from an electronic bulletin board.
Burr-Brown Corp.
INFO/CARD #247

RF Assembly Services

RF and Microwave assembly services are described in a new brochure from Matrix Corp. The company's capabilities for building and testing to customer specifications are described, including MIL-STD-2000, WS-6536, and MIL-I-45208.
Matrix Corp.
INFO/CARD #246

Power Tube Brochure

Burle Industries has released a new catalog describing their power tubes and associated circuit components. Output capabilities range from a few watts to megawatts, and frequencies covered are from DC to over 1400 MHz. Products include broadcast, communications, and special purpose tubes.
BURLE INDUSTRIES, INC.
INFO/CARD #245

Solid-State Temperature Switches

Solid-state temperature switches are listed in a new brochure from Cambridge Aeroflow, Inc. These products, available in SIP or stand-alone packaging, are a reliable alternative to mechanical thermostats, and can be mounted almost anywhere in cooling air flow.
Cambridge Aeroflow, Inc.
INFO/CARD #244

Low-Threshold MOSFETs

Fifteen new MOSFETS with low threshold voltages are described in this new brochure from Siliconix. Designed for instrumentation, power control, telecommunications, and motion control, these devices feature 3.5 V (n-channel) and 4.5 V (p-channel) threshold voltages for compatibility with 5-volt CMOS logic. Low ON resistance and fast switching times and 20-400 V breakdown voltages are additional features.
Siliconix Inc.
INFO/CARD #243

19-inch Card Frames

RITTAL Corp. has published their new Uniset iS/Varioset iS brochure, describing over 44 different card frame kits to meet users' specifications for applications requiring modular construction and flexible layout. This updated brochure includes updates to the line introduced since publication of their Handbook 26, a complete product line catalog.
RITTAL Corporation
INFO/CARD #242

RF/IF Signal Processing Components

Lorch Electronics announces their new RF and IF Signal Processing Components Catalog, with a complete listing of all Lorch products, including couplers, mixers, filters, switches and more. Also outlined are the company's capabilities for custom designs and integration into modules and subassemblies.
Lorch Electronics
INFO/CARD #241

Ceramic Resonators

Two brochures from Trans-Tech, Inc. describe new ceramic resonator product lines. The first is their Coaxial Resonator and Inductor brochure, with detailed information on high Q ceramic coaxial line elements for use in the 310 to 4800 MHz range. The Dielectric Resonator brochure covers both dielectric resonators and temperature compensated ceramic products with design considerations for oscillator, filter and MIC applications.
Trans-Tech, Inc.
INFO/CARD #240

FAX FAX FAX FAX FAX (303) 773-9716 FAX FAX FAX FAX FAX

Reader Survey: Direct Digital Synthesis

Please take a few minutes to complete the following survey. This information will help us determine what you want and need to know about products for Direct Digital Synthesis (DDS).

1) On a scale of 1 through 10 (1=low, 10=high), please rate your knowledge of DDS products and techniques
____

2) Using the same scale, how much do you enjoy reading articles in *RF Design* about applications of DDS?
____

3) Which suppliers do you think of for DDS products?
A)____________ B)____________ C)____________
D)____________ E)____________ F)____________

4) What types of Direct Digital Synthesis products do you use?
Integrated Circuits ____________ Board-Level Subsystems ____________
Modular Synthesizers ____________ Quadrature Outputs ____________
Modulated Waveforms ____________ Other (specify) ____________

5) What are your primary applications?
Consumer ____________ Military ____________
Consumer/Industrial ____________

6) How much does your company spend annually on these products?
Less than $1k ____________ $1k-$25k ____________
$25k-$100k ____________ More than $100k ____________

7) When considering DDS products, do you have the authority to:
Evaluate ____________ Recommend ____________ Specify ____________

8) Using a 1 through 10 scale (1=low, 10=high), please rate the following qualities in DDS products for your RF applications:
Price ____________ Delivery ____________ Performance ____________
Technical Support ____________ Other (specify) ____________

9) In what frequency range is your primary work?
Below 50 MHz ____________ 50-450 MHz ____________
450 MHz-1 GHz ____________ 1-2.5 GHz ____________
Above 2.5 GHz ____________

10) What new features would you like to see in DDS products?
____
____

11) If you would like to be contacted by the manufacturers of DDS products who advertised in this issue, please complete the following:

Name ____________ Company ____________
Address ____________ MS ______
City ____________ State ________ Zip ________
Telephone ____________ FAX ____________

THANK YOU FOR YOUR ASSISTANCE!

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RF design

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RF engineering opportunities

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