RF design

engineering principles and practices

August 1990

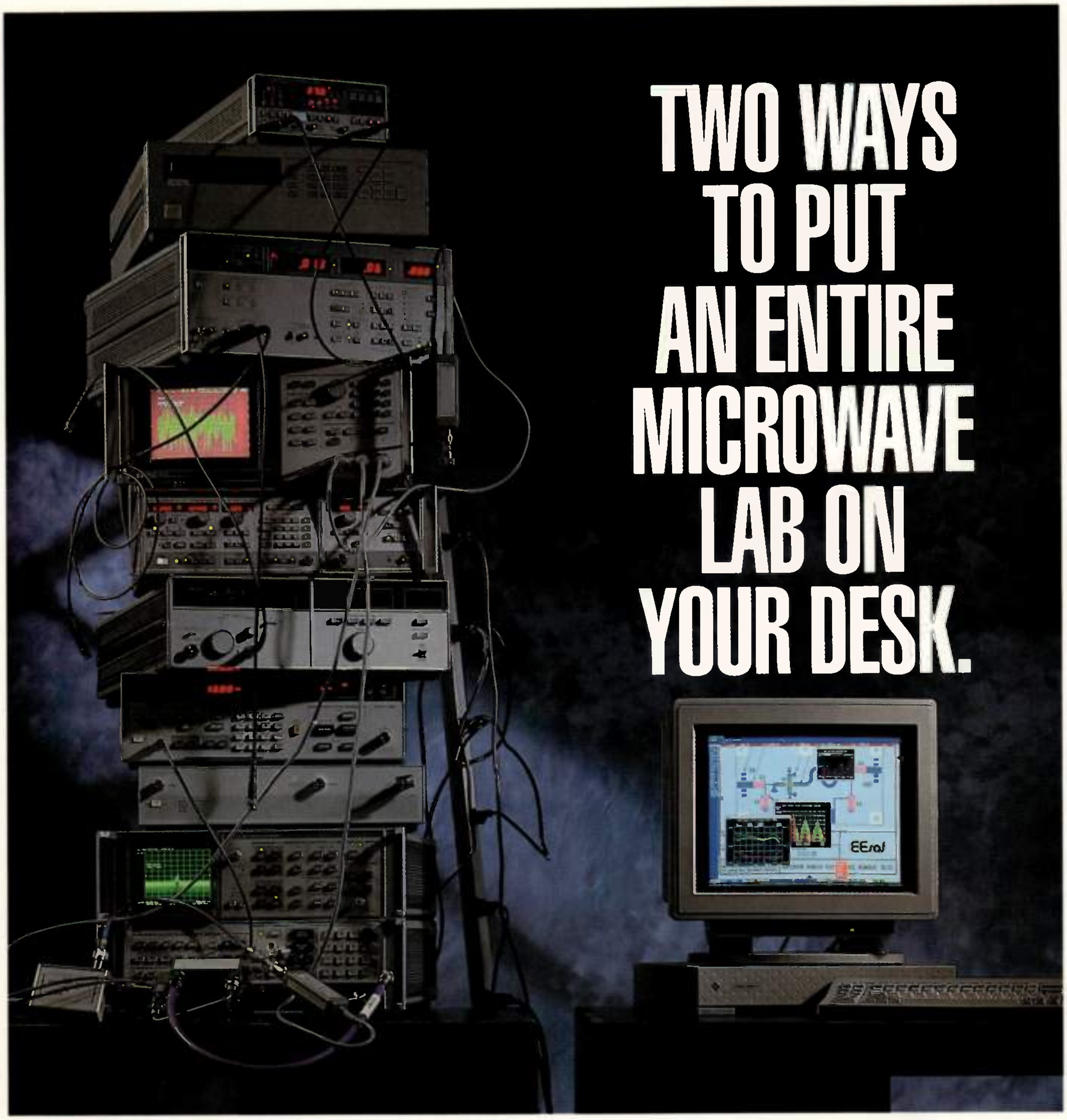
New Test Cell Handles Emissions *and* Susceptibility

featured technology

Electromagnetic Compatibility

industry insight

RF Capacitor Market Report



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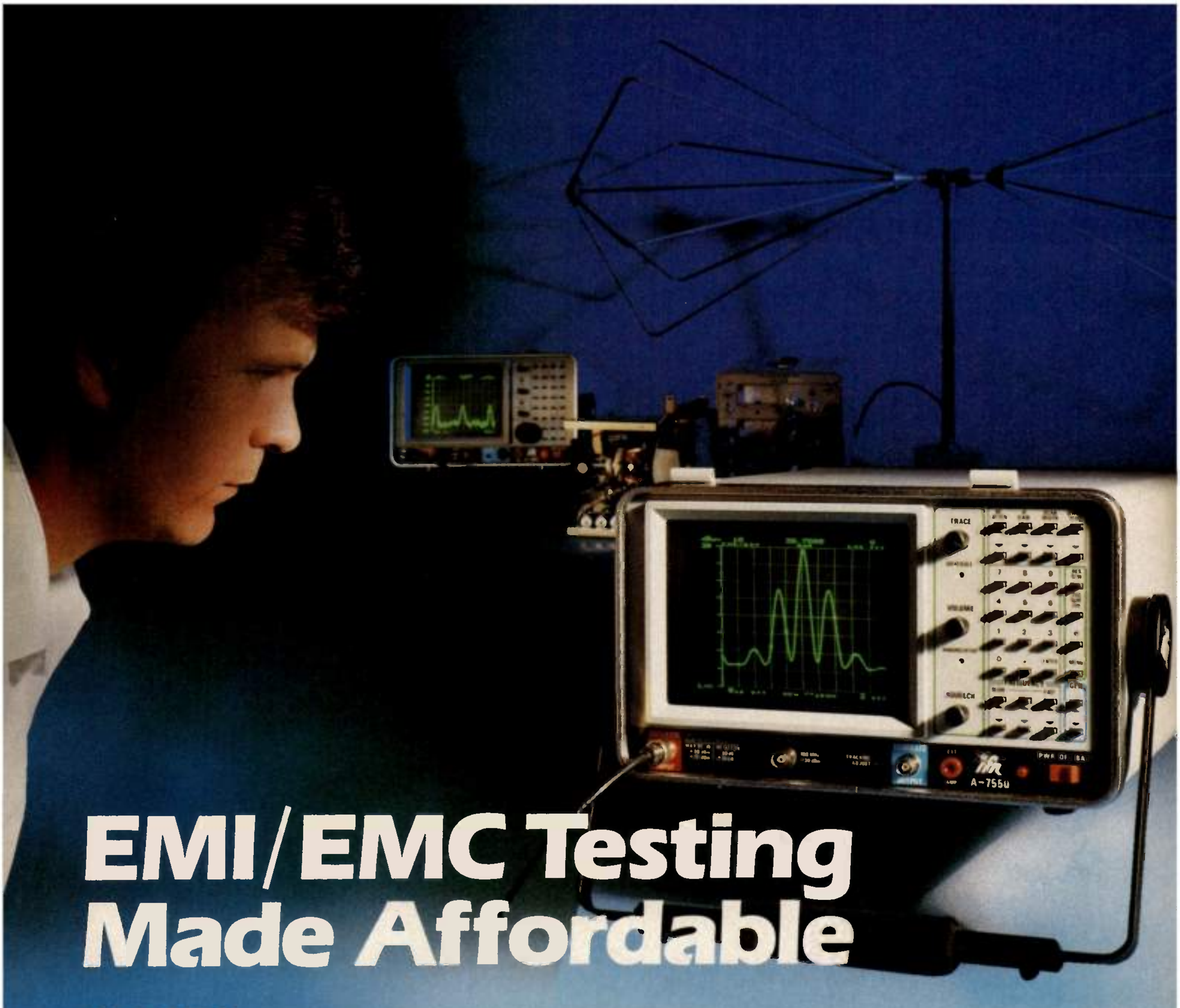

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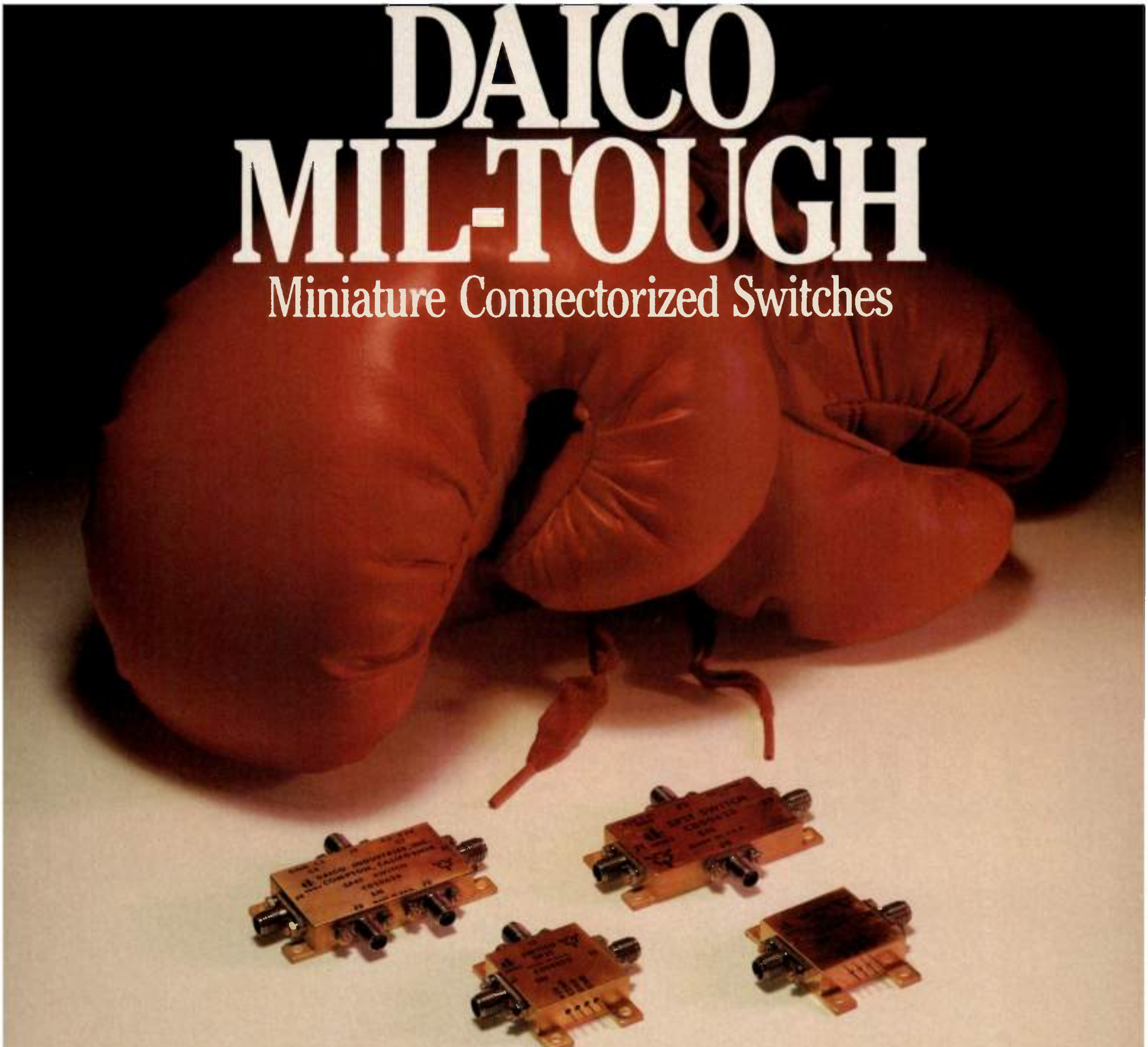

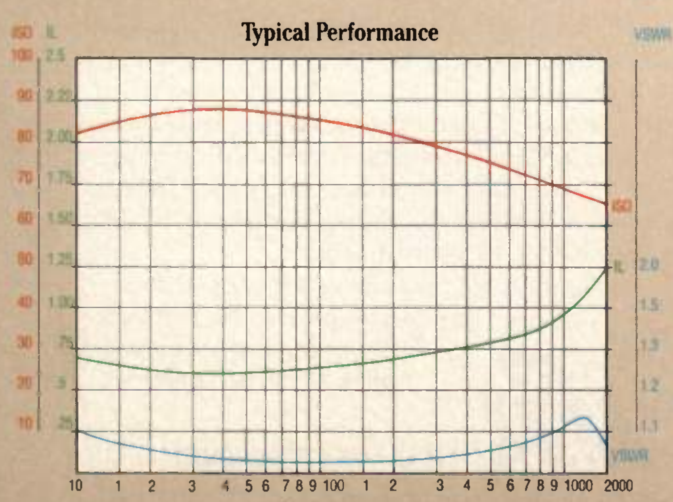

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INFO/CARD 3

RF design

contents

August 1990

featured technology

cover story

design awards

departments

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An Unsolved Mystery

By Gary A. Breed
Editor

Help, I need an answer! There is an exciting trend I'd like to know more about: Why have so many RF engineers recently taken up writing and public speaking?

At *RF Design*, we are now receiving many more manuscripts and article proposals than we can publish. Authors who wrote for us in the past are certainly noticing the longer wait between the acceptance of an article and its eventual publication. This increase in the number of article submissions has been developing over the past two or three years, and if this phenomenon was limited to articles for our magazine I probably wouldn't notice, but it is showing up in other places.

This November's RF Expo East had more proposals for papers than any previous RF Expo, east or west. We have always had a good response, but after the initial Call for Papers, we usually have to recruit a few papers to complete the program. This year, we have had to reject a dozen or so — and that's after increasing the number of sessions to accommodate some of the extras.

Another indicator of increased activity is our annual contest, which has grown over its five-year history and shows no signs of slowing. Also, there have been quite a few new books on RF topics written and released lately. The significant increase in both continuing education and product application courses might be related, as well.

This trend must be widespread, not limited to just *RF Design*. We have maintained the same approach toward engineering for nearly twelve years, and have always gotten good contributions from our readers. Steady growth would be expected, but the big recent rush indicates some thing more. A few authors have commented about the lack of appropriate outlets for "getting published," but that concern has been expressed in one way or another for at least 20 years.

These are things that I see from the perspective of a magazine editor. Maybe you can let me know what is happening "on the inside." Some of the other branches of engineering have become quite competitive. Is this causing a change in the RF engineering profession that encourages more participation and communication than before? Is the increased activity due to management's encouragement, or individual initiative?

Whatever the reason, I am definitely in favor of greater engineering communications. It is good for the engineering community to have a visible, active dialogue among colleagues. One engineer's simple comment can spark the imagination of another. Of course, wasted time "reinventing the wheel" can be avoided, too. The exchange of information in public or published forums serves to increase the stature of the profession. For example, medical researchers and physicists are always in the news after their work has been published.

I keep a close watch on the RF industry and the engineering profession, but the root of this trend has so far escaped me. Let me know how this has affected you and your colleagues.

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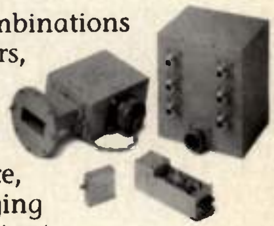

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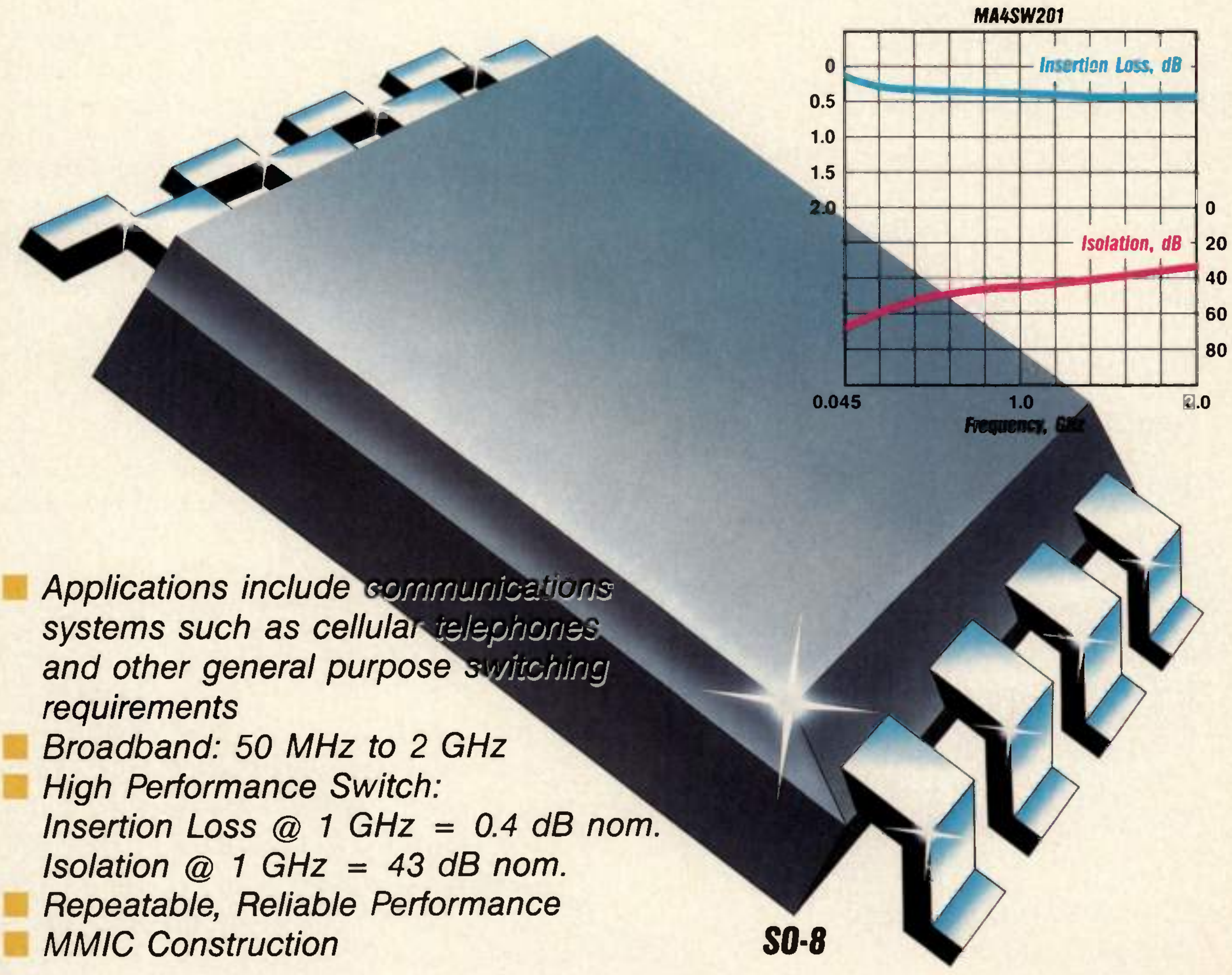

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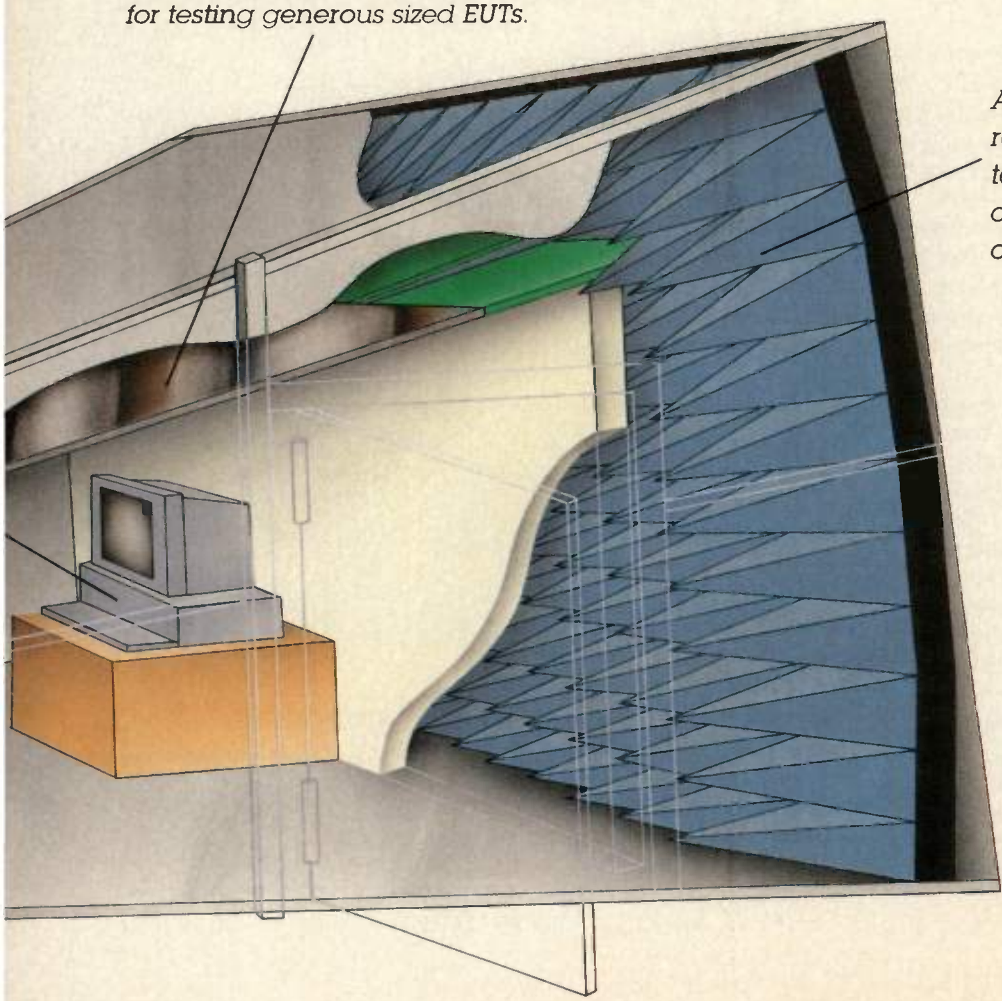

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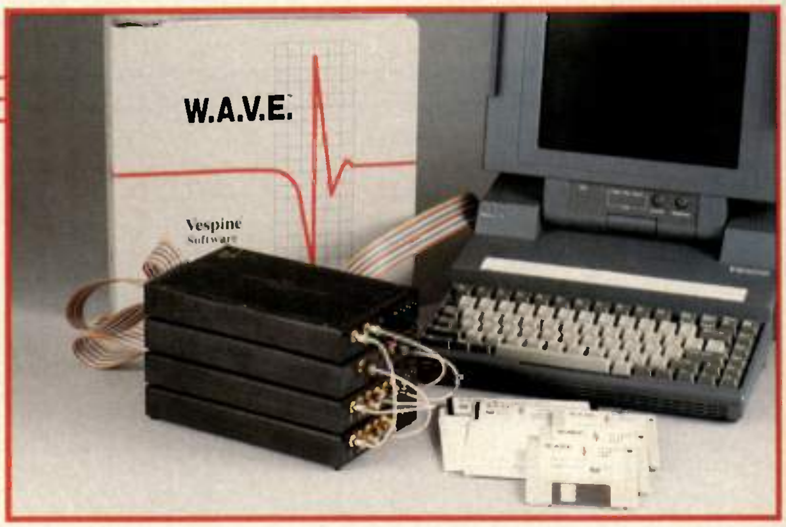

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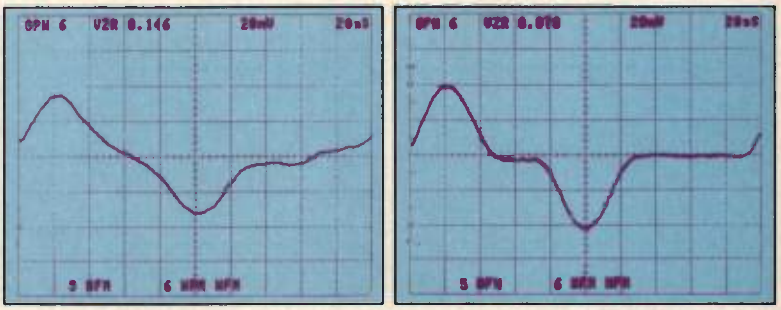

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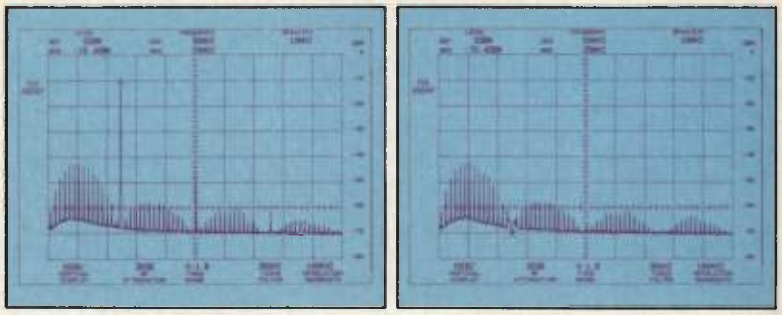

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Complex Signals

Editor:

I'd like to thank *RF Design* for publishing such a wonderful magazine. Without magazines such as yours it would be virtually impossible to keep up with new products and technologies.

I have especially appreciated Mr. Boutins's series on complex signals. However, in Part 3 [March, 1990, *RF Design*], p. 110, there appears to be an error in regenerating the complex signal. Using cos $[(\omega_c + B/2)t]$ and $-\sin(\omega_c t)$ shifted the spectrum by B/2.

It should be noted that the up- and down-conversion processes mirror each other. In fact, as long as the processes mirror and orthogonal waveforms such as sine and cosine are used, the complex input will be correctly recovered. Changing signs and exchanging sine and cosine simply shift phase and negate frequencies in the real data channel.

Brian S. Little
Department of Defense
Laurel, MD

Author's Reply

I would like to thank Mr. Little for his feedback. That gives me the opportunity to clarify one point of my presentation.

Mr. Little's system effectively permits one to up-convert a given baseband complex signal to a real high frequency one and then, regenerate back the same baseband complex signal.

In my paper, I have tried to be more general. The down-convertor to which Mr. Little refers in his letter (p. 110, Figure 4) does exactly what the accompanying text mentions, that is it generates a complex baseband signal having its Fourier Transform located right in the middle of the zero frequency axis. Any other complex baseband signal could have been generated by properly adjusting the frequency of the quadrature local oscillators. What could have mislead the readers is the fact that I started my discussion on the down-conversion process with a real high frequency signal whose Fourier Transform was the same one used previously to explain the up-conversion process of another particular complex baseband signal. It would have been preferable to pick any other Fourier Transform and included Mr. Little's example of a particular system where the cascaded up- and down-conversion processes are transparent.

Noel Boutin
University of Sherbrooke
Sherbrooke, Quebec

Odd Ratio Frequency Dividers

Editor:

This letter is in reference to the article by D. Korn and C. Deierling "A 1.8 GHz Odd Ratio Frequency Divider" in the May 1990 issue of *RF Design*.

I would like to point out that odd-ratio frequency dividers may be designed digitally to perform at the same frequencies as 2^n counters with all the advantages of digital counters such as opera-

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505FC	5W CW	.5-525 MHz	40dB	25x28x13	100-240V	1,655.00
706FC	6W CW	.5-1000 MHz	38dB	25x28x13	100-240V	2,695.00
110C	10W CW	.2-60 MHz	40dB	19x32x15	100-240V	1,450.00
110LC	10W CW	.01-100 MHz	40dB	25x37x15	100-240V	1,950.00
210LC	10W CW	.008-225 MHz	40dB	25x37x15	100-240V	2,495.00
310FC	10W CW	.5-300 MHz	40dB	25x28x13	100-240V	2,250.00
510FC	10W CW	.5-525 MHz	43dB	25x28x13	100-240V	2,595.00
710FC	10W CW	1-1000 MHz	40dB	25x28x13	100-240V	5,995.00
115LC	15W CW	.001-100 MHz	43dB	38x32x13	100-240V	2,255.00
225LC	20W CW	.01-225 MHz	45dB	25x37x15	100-240V	3,295.00
520FC	20W CW	1-525 MHz	45dB	25x28x13	100-240V	3,895.00
720 FC	20W CW	500-1000 MHz	43dB	48x46x13	100-240V	6,650.00
125LC	25W CW	50Hz-100 MHz	40dB	48x40x13	100-240V	3,300.00
125C	25W CW	.5-50 MHz	45dB	48x46x13	100-240V	1,495.00
112C	25W CW	1-120 MHz	45dB	25x37x15	100-240V	1,995.00
320FC	25W CW	1-300 MHz	45dB	30x20x13	100-240V	2,895.00
535FC	35W CW	200-512 MHz	45dB	30x20x13	100-240V	3,895.00
150C	50W CW	.2-50 MHz	47dB	38x37x15	100-240V	2,595.00
250LC	50W CW	.01-230 MHz	45dB	43x42x19	100-240V	7,550.00
250FC	50W CW	1-200 MHz	45dB	38x37x15	100-240V	4,995.00
550FC	50W CW	200-512 MHz	45dB	30x20x13	100-240V	4,995.00
161C	100W CW	.5-90 MHz	45dB	38x37x15	100-240V	5,250.00
190LC	100W CW	.01-100 MHz	50dB	48x43x18	100-240V	8,850.00
162LP	100W CW	20-200 MHz	50dB	38x37x15	100-240V	6,250.00
116C	100W CW	.01-220 MHz	50dB	71x56x76	100-240V	8,995.00
155LCR	100W CW	.006-12 MHz	50dB	48x40x13	100-240V	4,150.00
LA100H	100W CW	.3-100 MHz	50dB	48x48x13	100-240V	5,990.00
LA100F	100W CW	100-250 MHz	50dB	48x48x13	100-240V	6,595.00
LA100U	100W CW	200-400 MHz	50dB	48x48x13	100-240V	6,250.00
LA100UE	100W CW	100-500 MHz	50dB	48x48x13	100-240V	6,995.00
LA200H	200W CW	.5-100 MHz	53dB	48x48x13	100-240V	6,995.00
LA200F	200W CW	100-250 MHz	53dB	48x48x13	100-240V	11,500.00
LA200U	200W CW	200-400 MHz	53dB	48x48x13	100-240V	10,995.00
LA200UE	200W CW	250-500 MHz	53dB	48x48x13	100-240V	12,500.00
162LPS	200W Pulse	10-220 MHz	55dB	48x38x18	100-240V	5,995.00
122C	200W CW	.01-220 MHz	50dB	71x56x76	100-240V	10,850.00
247V	200W Peak	118-138 MHz	10dB	48x46x13	100-240V	5,255.00
247U	200W Peak	200-400 MHz	10dB	48x46x13	100-240V	5,500.00
121CA	250W CW	.5-32 MHz	55dB	48x43x18	100-240V	5,950.00
164UP	300W Pulse	200-400 MHz	55dB	48x38x18	100-240V	7,595.00
LP300H	300W Pulse	.3-100 MHz	56dB	48x48x13	100-240V	7,550.00
LA500H	500W CW	.5-50 MHz	57dB	71x56x76	100-240V	12,500.00
LA500V	500W CW	10-100 MHz	57dB	71x56x76	100-240V	13,500.00
134C	500W CW	.01-220 MHz	57dB	58x69x127	3 Phase	19,800.00
134CM	500W CW	1-200 MHz	57dB	58x69x127	3 Phase	18,500.00
166UP	500W Pulse	200-400 MHz	57dB	48x51x18	100-240V	9,500.00
LA600U	600W CW	200-400 MHz	57dB	71x56x76	100-240V	23,500.00
166HP	800W Pulse	100-200 MHz	57dB	48x51x18	100-240V	11,500.00
137C	1000W CW	.01-220 MHz	60dB	69x69x127	3 Phase	28,500.00
137CM	1000W CW	1-200 MHz	60dB	69x69x127	3 Phase	27,250.00
LA1000H	1000W CW	2-32 MHz	60dB	71x56x76	100-240V	22,500.00
135CB	1000W Pulse	2-30 MHz	60dB	48x51x46	100-240V	10,990.00
LA1000V	1000W CW	10-100 MHz	60dB	71x56x76	100-240V	26,900.00
166LP	1000W Pulse	8-100 MHz	60dB	48x51x18	100-240V	9,550.00
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167LP	2000W Pulse	10-100 MHz	63dB	53x65x46	100-240V	17,900.00
140C	2000W CW	.01-220 MHz	60dB	137x69x127	3 Phase	44,990.00
140CM	2000W CW	.5-150 MHz	60dB	137x69x127	3 Phase	41,500.00
LA3000H	3000W CW	5-45 MHz	65dB	137x76x127	180-255V	69,500.00
LP4000H	4000W Pulse	2-50 MHz	66dB	71x56x76	100-240V	27,500.00

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tion from DC to the maximum operating frequency. The frequency is limited only by the clock to output delay and setup time of the flip-flop. With the analog approach (use of a delay line), as the authors point out, the counter is useful only over a very limited frequency range. The delay line should be used only when the frequency requirements exceed the speed of the logic.

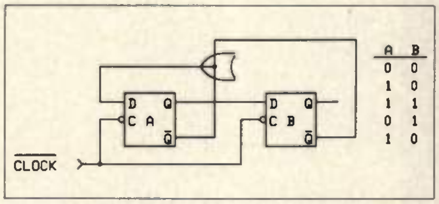
A divide-by-three example usable from DC to the maximum "real" clock frequency, using the same 10G021A circuit, is shown below.

Fourier Analysis of the ideal waveform generated by the digital divider with a 2/3 duty cycle pulse shows that the 2nd harmonic is 6 dB down from the fundamental. This would normally require a better filter, if the fundamental component frequency is to be extracted, compared with a theoretical 1/2 duty cycle waveform with no even harmonics. The authors' data, however, shows the 2nd harmonic down only 9 dB, also indicating a very non-symmetrical waveform. The difference in the fundamental amplitude between the two different (2/3 and 1/2) ideal waveforms is only 1.25 dB.

If GigaBit Logic had designed their D flip-flop with dual D OR-able inputs, similar to the Motorola or Fairchild high speed ECL logic which added no additional delay, the logic could have been performed at these inputs. However, the GigaBit circuits, like ECL, feature output wired-OR capability.

The Q outputs from each flip-flop are close together and connected together before connection to the D-input of circuit A adding negligible delay. Thus, the registers may be operated up to the specified clock frequency limited only by the clock to output delay and the setup time requirements. Using the authors' delay numbers this becomes DC to 1.45 GHz. This type approach, chip and wire bond on a small thick film hybrid, was used in the early 70's to generate a multitude of logic including a pseudo random code generator operating at 800 MHz. The basic circuit was a Fairchild 11CO6 (now Motorola MC12090), all the logic being performed using only the two D-inputs and the Q and not-Q outputs of the several circuits installed within the hybrid. Even the exclusive-or function was developed without any additional gates allowing clocking up to the maximum frequency. This circuit used a half adder or flip-flop adder (F-adder) made out of two circuits both using their two D-inputs. Many of these type circuits have been operating since 1976 in the ALGOR radar clock pulse generator located on Roi-Nemur Island in the Pacific. This system featured range increments of 24 picoseconds.

David D. Freedman
Exetron
Cinnaminson, NJ



Divide-by-three logic shown using "output OR-ing".

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INFO/CARD 15

WRH

RF calendar

August

21-23 **IEEE EMC 90 Symposium**
Washington Hilton Hotel, Washington, DC
Information: Joe Fisher Tel: (703) 521-6336.

28-30 **Surface Mount '90**
Bayside Exposition Center, Boston, MA
Information: MG Expositions Group, 1050 Commonwealth Ave., Boston, MA 02215. Tel: (800) 223-7126 or (617) 232-3976.

28-5 **XXIII General Assembly of the International Union of Radio Science (URSI)**
Prague, Czechoslovakia
Information: Prof. V. Zima, Institute of Radioengineering and Electronics, Czechoslovak Academy of Sciences, 182 51 Praha 8, Czechoslovakia.

September

10-13 **20th European Microwave Conference**
The Intercontinental Hotel, Budapest, Hungary
Information: Microwave Exhibitions and Publishers Limited, 90 Calverley Road, Tunbridge Wells, Kent, TN1 2UN, England. Tel: (0892) 544027.

10-13 **Electrostatic Overstress/Electrostatic Discharge Symposium**
Buena Vista Palace, Orlando, FL
Information: 1990 EOS/ESD Symposium, PO Box 913, Rome, NY 13440. Tel: (315) 339-6937.

12-15 **Radio 90**
Hynes Convention Center, Boston, MA
Information: Andy Peluso, NAB Radio 1990 Registration, 1771 N Street, N.W., Washington, DC 20036-2891. Tel: (800) 342-2460. Fax: (202) 775-2146.

17-18 **1990 IEEE Bipolar Circuits and Technology Meeting**
Marriott Hotel, Minneapolis, MN
Information: Jan Jopke, Conference Coordination Services, 6611 Countryside Drive, Eden Prairie, MN 55346. Tel: (612) 934-5082.

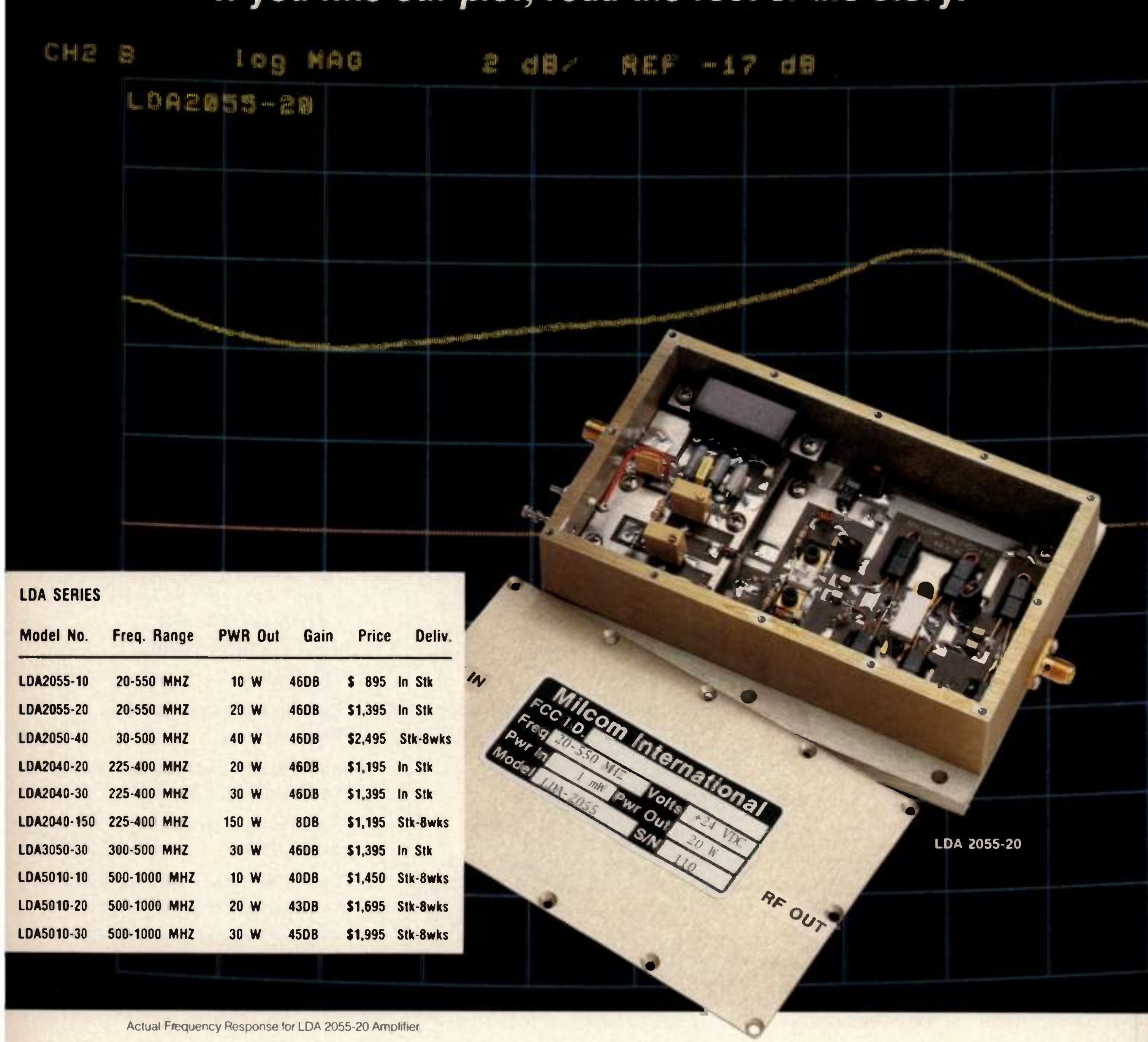
23-27 **Association of Old Crows Meeting**
Hynes Convention Center, Boston, MA
Information: Association of Old Crows, 1000 N. Payne Street, Alexandria, VA 22314. Tel: (703) 549-1600.

25-27 **Piezoelectric Devices Conference**
Kansas City Westin Crown Center, Kansas City, MO
Information: EIA Components Group, 1722 Eye Street, N.W., Suite 300, Washington, DC 20006. Tel: (202) 457-4980.

October

1-4 **SCAN-TECH 90**
Georgia World Congress Center, Atlanta, GA
Information: AIM USA, 1326 Freeport Road, Pittsburgh, PA 15238. Tel: (800) 338-0206, (412) 963-8588.

If you like our plot, read the rest of the story.



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LDA2040-20	225-400 MHZ	20 W	46DB	$1,195	In Stk
LDA2040-30	225-400 MHZ	30 W	46DB	$1,395	In Stk
LDA2040-150	225-400 MHZ	150 W	8DB	$1,195	Stk-8wks
LDA3050-30	300-500 MHZ	30 W	46DB	$1,395	In Stk
LDA5010-10	500-1000 MHZ	10 W	40DB	$1,450	Stk-8wks
LDA5010-20	500-1000 MHZ	20 W	43DB	$1,695	Stk-8wks
LDA5010-30	500-1000 MHZ	30 W	45DB	$1,995	Stk-8wks

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RF courses

Antennas: Radiation and Scattering
August 27-28, 1990, Washington, DC
Grounding, Bonding, Shielding and Transient Protection
August 27-30, 1990, Chicago, IL
Microwave Radio Systems
September 6-7, 1990, Washington, DC
Integrating Fiber Optics and Analog/RF
September 10-12, 1990, Washington, DC
Broadband Communications Systems
September 10-14, 1990, Washington, DC
Modern Receiver Design
September 10-14, 1990, Washington, DC
October 15-19, 1990, London, England
Radar Operation and Design: The Fundamentals
September 18-21, 1990, Washington, DC
Doppler Radar Applications for Severe Weather and Wind-Shear Detection
September 26-28, 1990, Washington, DC
Radio Frequency Spectrum Management
October 1-4, 1990, Washington, DC
Information: George Washington University, Merril Ferber. Tel: (800) 424-9773; (202) 994-6106. Fax: (202) 872-0645.

Basic Telephony
October 22-24, 1990, Madison, WI
Digital Switching
October 25-26, 1990, Madison, WI
Information: University of Wisconsin-Madison, College of Engineering. Tel: (800) 222-3623; (414) 227-3200. Fax: (414) 227-3119.

Analog MOS Integrated Circuits
September 24-28, 1990, Los Angeles, CA
Information: UCLA Extension, Engineering Short Courses. Tel: (213) 825-3344. Fax: (213) 206-2815.

Modern Power Conversion Design Techniques
September 10-14, 1990, San Francisco, CA
Information: E/J Bloom Associates, Inc., Mrs. Joy Bloom. Tel: (415) 492-8443.

RF/MW Amplifier and Oscillator Design: Linear/Nonlinear Considerations
September 17-20, 1990, Budapest, Hungary
Information: Scientific Society for Telecommunications jointly with Besser Associates. Tel: 36-1-153-1027, Fax: 36-1-156-1215.

RF/MW Circuit Design: Linear and Nonlinear Techniques
September 20-26, 1990, Burlington, MA
Information: Besser Associates. Tel: (415) 969-3400. Fax: (415) 965-0800.

Digital Signal Processing: Techniques and Applications
August 14-17, 1990, Washington, DC
August 21-24, 1990, San Diego, CA
September 18-21, 1990 Washington, DC
October 2-5, 1990, Denver, CO
Introduction to Telecommunications
August 21-24, 1990, San Diego, CA
August 28-31, 1990, Washington, DC
September 18-21, 1990, Los Angeles, CA
Information: Learning Tree International, John Valenti. Tel: (800) 421-8166; (213) 417-8888. Fax: (213) 410-2952.

ELINT/EW Data Bases
September 10-11, 1990, Syracuse, NY
ELINT Analysis
September 12-14, 1990, Syracuse, NY
ELINT Interception
September 17-19, 1990, Syracuse, NY
Radar Vulnerability to Jamming
September 20-21, 1990, Syracuse, NY
Integrated EW
September 25-26, 1990, Syracuse, NY
ELINT/EW Applications of Digital Signal Processing
October 2-4, 1990, Syracuse, NY
Information: Research Associates of Syracuse, Inc. Tel: (315) 455-7157. Fax: (315) 455-8037.

Principles of Analog Oscilloscopes
August 21-22, 1990, Los Angeles, CA
Principles of Digital Oscilloscopes
August 23-24, 1990, Los Angeles, CA
Information: John Fluke Mfg. Co., Inc. Tel: (800) 443-5853 ext. 73. In Canada: (416) 890-7600.

Design for ESD and RFI
September 19, 1990, San Jose Hyatt, CA
Information: The Keenan Corporation, Ms. Jean Whitney. Tel: (813) 544-2594. Fax: (813) 544-2597.

Microwave Circuit Design: Linear and Nonlinear
September 24-26, 1990, Los Angeles, CA
October 24-26, 1990, Dallas, TX
Information: Vendelin Engineering. Tel: (408) 867-2291.

Electromagnetic Pulse
August 16-17, 1990, Philadelphia, PA
Information: R&B Enterprises. Tel: (215) 825-1960.

Transient Voltage Suppression Design Seminar
August 16, 1990, Raleigh, NC
September 11, 1990, Toronto, Canada
September 13, 1990, Detroit, MI
October 16, 1990, Boston, MA
Information: GSI Educational Services. Tel: (800) 776-8358.

mwSPICE
August 21-23, 1990, Westlake Village, CA
Academy (Schematic)
September 11-12, 1990, Westlake Village, CA
Academy (Layout)
September 13-14, 1990, Westlake Village, CA
Touchstone
October 2-4, 1990, Westlake Village, CA
Libra
October 9-11, 1990, Westlake Village, CA
Information: EEsof, Ginger Craft. Tel: (818) 991-7530.

Basic Network Measurements Using the HP8510B Network Analyzer
August 28-30, 1990, Los Angeles, CA
September 18-20, 1990, Boston, MA
September 18-20, 1990, Los Angeles, CA
Microwave Fundamentals
August 21-23, 1990, Boston, MA
September 24-27, 1990, Los Angeles, CA
Information: Hewlett-Packard Company. Tel: (800) 472-5277.

NIST Develops New Phase Noise Measurement System

Scientists at NIST have developed a more accurate system for measuring phase noise in oscillators, amplifiers, frequency synthesizers, and other electronic components according to a report in the May 14 issue of *NIST Update*. The system will be of interest to military and civilian calibration laboratories and companies producing high-precision navigation and communications equipment because "the wide bandwidths and the higher accuracy [of phase noise measurement] are necessary to adequately characterize new equipment used in very wide bandwidth communication, navigation, and measurement systems." The system has an accuracy 2 to 4 times that of current commercial equipment and its frequency range is 25 times greater. Its accuracy is reported at 1 dB under most conditions, with analysis bandwidths of approximately 10 percent of the carrier frequency up to a maximum of 1 GHz. The system is able to measure the phase noise at carrier frequencies from 5 MHz to 1.5 GHz, 1.5 to 26 GHz, and 33 to 50 GHz. Additional frequency ranges can be measured using external mixers to convert the signal to one of the above bandwidths. It can be run manually from the front panels of the various instruments or run from an AT compatible microprocessor. In creating this new system, scientists at NIST have attempted to construct a model which accounts for all known contributions to uncertainties in phase noise measurement. For a copy of the paper (paper no. 23-90) describing the system contact Jo Emery, Div. 104, NIST, Boulder, CO 80303. Tel: (303) 497-3237.

Commerce Report Tags 1990 "Emerging Technologies" — Twelve of today's "emerging technologies" will represent a combined world market of about $1 trillion by the year 2000, according to a report issued by the Commerce Department's Technology Administration. *Emerging Technologies: A Survey of Technical and Economic Opportunities* assesses the competitive position of the United States vis-a-vis Japan and Europe and makes 13 recommendations for actions by industry and government to improve U.S. competitiveness. The emerging technologies cited are advanced materials, superconductors, advanced semiconductor devices, digital imaging technology, high-density data storage, high-performance computing, optoelectronics, artificial intelligence, flexible computer-integrated manufacturing, sensor technology, biotechnology, and medical devices and diagnostics.

Georgia Tech Researchers Develop Sensor Fusion Simulator — Scientists at the Georgia Institute of Technology have developed a multi-sensor simulation tool from a database of infrared, satellite and synthetic aperture radar for pilots. While a computer to handle real-time sensor fusion might not be available for another five to ten years, the Georgia Tech Simulator, known as GTSPECS, allows engineers and scientists to work on other portions of the sensor fusion problem while they wait for computing power to catch up. The database represents a different approach toward the difficult problems of simulating sensor fusion. Instead of taking actual sensor data and trying to generalize it to broader scenes, the Georgia Tech researchers began with simulated images and are attempting to apply this "first principles" understanding to model a wide range of real conditions. The researchers are currently validating their database against measured target and scene information to determine how well it would describe real combat situations.

Hughes Scientists Disclose New Photonically Controlled Antenna — A new multi-function antenna technique that uses optical signals to aim the microwave signals from a fixed phased-array antenna has been developed by scientists at Hughes Research Laboratories. The technique uses the large bandwidth and precision time delays provided by a fiber optic network. The use of fiber optics promises to reduce the weight and volume of complex antenna systems while providing more precise beam pointing and improved beam quality at wide scanning angles.

TriQuint, Rockwell Sign GaAs Second Source Agreement — TriQuint Semiconductor has announced an agreement with Rockwell International for the second sourcing of TriQuint's gallium arsenide QLSI standard cell family. Rockwell will use a 1-micron GaAs enhanced/depletion process to manufacture the TriQuint standard cell family and make it available to military and commercial OEMS. Rockwell will use TriQuint's standard cells to develop and manufacture internal products as well as enter the commercial GaAs IC business by providing this second source. A customer may take a workstation based logic design using the TriQuint QLSI library to either Rockwell or TriQuint for manufacturing.

UTMC and Seattle Silicon Sign Joint Agreement — United Technologies Microelectronics Center, Inc. (UTMC) and Seattle Silicon Corporation (SSC) have signed a joint agreement to deliver high-density, rad-hard, application specific integrated circuits for aerospace and military design-ins. Under the terms of the agreement, SSC provides the design methodology and support, and UTMC fabricates and tests the rad-hard ASIC designs.

NAB Asks for Clarification of UHF-TV Channel Usage — The National Association of Broadcasters has asked the FCC to reexamine its rules regarding the use of vacant UHF-TV channels in order to stop possible abuses. The FCC currently allows the use of a vacant UHF-TV channel as a studio-to-transmitter link (STL), provided that it does not create interference. The NAB has asked

INFO/CARD 20

the FCC to freeze acceptance of applications for STLs on vacant UHF-TV channels, because they contend that some low power television stations may be using the vacant UHF-TV channels to transmit programming directly to viewers. The NAB has asked for a rulemaking proceeding that will specifically require: that applicants comply with low power television technical protection criteria; that applicants be required to prove that no interference will result if the application is granted; and that the signals transmitted be scrambled and, therefore, not directly compatible with consumer TV receivers. The NAB says in its filing that "with these modifications, the Commission would substantially clarify its rules and procedures for licensing auxiliary facilities," and "establish reasonable safeguards to protect full service television stations from unwarranted interference."

CCD Chips Record First Images From Hubble Space Telescope —The first test images from the Hubble Space Telescope have been recorded by eight Texas Instruments charge coupled device (CCD) image sensors. The silicon chips, acting as extremely sensitive photographic plates in the telescope's Wide Field/Planetary Camera, recorded Star Field NGC 3532, a 3-billion year old star cluster located 1,500 light years from Earth in the constellation Carina. The CCD image sensors were developed by TI scientists under contract to the Jet Propulsion Laboratory at California Institute of Technology, where the camera was manufactured. The camera will produce images and spectrographic, photometric, and polarimetric measurements, and obtain pictures of the universe on a wider and grander scale than any other instrument to date, according to NASA.

CCIR Assembly Acts on HDTV Standards — The International Radio Consultative Committee (CCIR) of the International Telecommunications Union approved the results of the past four years' efforts of its working party dealing with High Definition Television (HDTV) standards for studio production and international exchange of programs. The CCIR gave tentative approval to 23 of 34 of the parameter values which comprise a television picture. The remaining values relate to the number of lines used to display a picture on the television screen and the number of times per second a TV picture is taken. Those values will continue to be studied during the next four year cycle of the CCIR.

Gigabit Logic Receives $1.1 Million Order — Gigabit Logic announced receipt of a $1.1 million order from Sciteq Electronics. The order is for Gigabit's 1 GHz, 32-bit accumulator and 512 x 8 ROM which form the heart of a DDS that covers up to 200 MHz output bandwidth.

Broadcast Engineering Conference Call for Papers — The 45th Annual Broadcast Engineering Conference is scheduled for April 14-18, 1991 and will provide an opportunity for broadcasters, program producers, research laboratories, and manufacturers to present papers on broadcast engineering technology, system designs, and techniques. The papers presented should contain information on new ideas, technologies and methods that will augment the technical knowledge and skill of attending engineers. A one page proposal should be sent by September 15, 1990 to: Engineering Conference Committee, NAB Science and Technology, 1771 N Street, N.W., Washington, DC 20036. For more information contact the NAB's Department of Science and Technology or call (202) 429-5346.

Session Proposals Sought for Systems/USA — Session proposals are now being requested for the Systems/USA Conference and Exposition to be held February 11-13, 1990. Among some of the topics for proposals are power, EMI and ESD, Government - Defense and General Services, and Display Technology. Proposals must be received by September 1, 1990 and should be sent to Roy Webster, Technology Conference Coordinator, Systems/USA Technology Conference, American Electronics Association, 5201 Great America Parkway, Santa Clara, CA 95054. Tel: (503) 359-5873 or (408) 987-4200.

IEEE Call for Papers — The IEEE has issued a call for papers for their 1991 Aerospace Applications Conference to be held February 3-8, 1991. Papers are requested on topics such as: Communications and Telemetry; Electro-optic Applications; Instrumentation and Management; Millimeter and Microwave Technology; and Energy and Space to name a few. Interested applicants are invited to submit a 500 word summary emphasizing the present and future applications of their topic (2 copies) by September 6, 1990. They should be sent to: Program Chairman, Leo Mallette, 2309 S. Santa Anita, Arcadia, CA 91006.

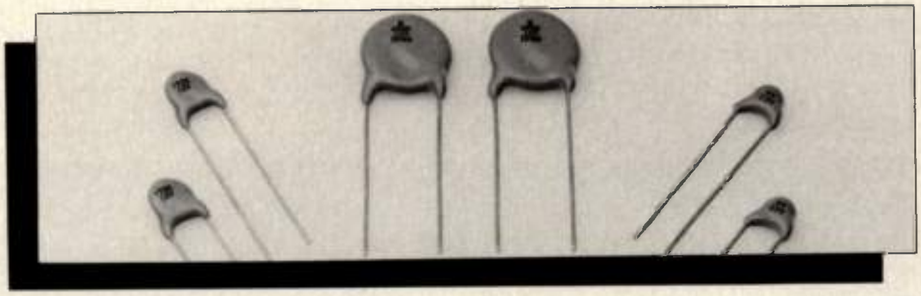
SEMI-THERM Call for Papers — The 7th Annual IEEE Semiconductor Temperature and Thermal Management Symposium has issued a call for papers for their conference to be held February 12-14, 1991 in Phoenix, Arizona. The committee is requesting papers on the following areas: thermal characterization, analytical and computational modeling, measurement techniques including temperature, flow and thermal-mechanical properties, and thermal reliability screening and testing. A 500 word abstract including the name, address and phone number of the principal author must be submitted by August 15 to: Dr. Kaveh Azar, Program Chairman, AT&T Bell Laboratories, 75 Foundation Avenue, Ward Hill, MA 01835. Tel: (508) 374-5508. Fax: (508) 374-5503

EIA Moves to New Location — The Electronic Industries Association has relocated their offices to their permanent headquarters in the James Monroe Building, 2001 Pennsylvania Avenue, N.W., Washington, DC 20006-1813. All telephone numbers and extensions remain unchanged.

Telesync Announces New Facility — Telesync, Inc., has announced that the company will move its manufacturing and administration operations into a new facility at 5555 Oakbrook Parkway, Norcross, GA 30093.

Evolution of the RF Capacitor Market

By Charles Howshar and Liane Pomfret
Assistant Editors



The RF capacitor market has matured to a point where it no longer sees the radical changes that affect younger markets. Rather it is a market evolving under the influences of the markets it sells to. Applications for RF capacitors are increasing in the cellular communications, medical, and aerospace market; as well as in the laser and antenna markets. Many companies are confident of the future of these applications and the role RF capacitors will play in them. On the down side, the reduced number of military projects makes it a difficult market in which to compete. However, potential markets for RF capacitors in magnetic resonance imaging promise to fill in the gap vacated by the military.

Due to the cutbacks in military spending, there are a limited number of military weapons programs available. However, for those with contracts in surveillance and communications applications, the military market is still going strong. "The military weapons-development market is shrinking, but the military listening equipment market is growing," remarks Don Davis, Director of Sales and Marketing for American Technical Ceramics Corporation. "The majority of our market is military, and we have record bookings right now," states Gunther Vorlop, Sales and Marketing Manager for Dielectric Labs, Incorporated. For many manufacturers it is simply a matter of being in the right place at the right time.

Feelings about the commercial market are mixed. Depending on the areas of expertise, some companies feel that the market is down from last year while others feel that it has grown. Companies in cellular communications feel that the market is growing. Dick Phillips, Product Manager for AVX Corporation, comments, " The market for RF capacitors is flat but there is more activity in cellular communications." Scott Newman, President of Voltronics, remarks, "Anything having to do with commercial communications is a good market for capacitors, although there are limited bright spots in the military market."

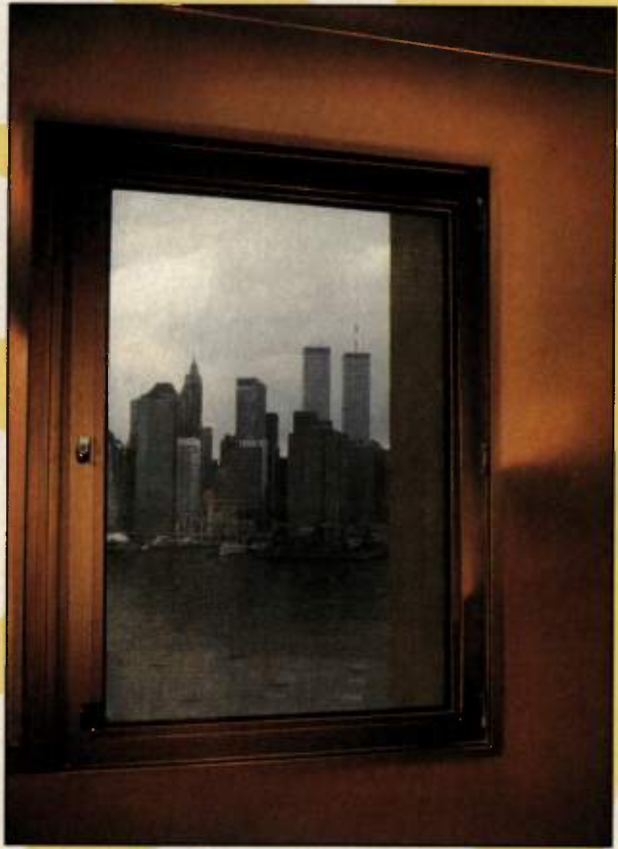
The technological advances that have occurred recently in nuclear magnetic resonance (NMR) and magnetic resonance imaging (MRI) have generated great interest among capacitor manufacturers. Non-magnetic capacitors are required for these applications, generating new developments in capacitor design. "The magnetic imaging market has a very broad applications base, and the future looks very favorable," comments Will LaRusso, General Manager for Polyflon Company. "Non-magnetic capacitors are a must for MRI applications," he adds. Medical applications are also on the rise. "There's a lot of research going on in NMR, some medical areas, and physics. One of our customers was using our capacitors in his research on the heating of body tissue as a cure for cancer," says Doug Gordon, Vice-President of Marketing for Surcom Associates. Some other uses in the medical field include a trimmer used in metered infusion pumps and a transmitter or panic button for elderly or handicapped patients.

New Developments

Since the RF capacitor market is quite mature, customers' questions usually pertain to availability and cost. Customers want the usual improvements that come with any technological advancement. These include increased performance, reliability, lower cost, and smaller size. Some of the new features customers want include feedthrough networks, quick design turnaround, lower insertion loss, and greater power handling capabilities. Surface mount technology is one development that is helping to satisfy these requirements. Thin film capacitor designs are also improving product designs. "Thin film capacitors using a silicon-dioxide and silicon-nitride dielectric in a surface mount package deliver very high Q, very low ESR, and tight tolerance," states Dick Phillips. This makes them ideal for products such as filters. "The integration of capacitors and resistors on substrates will enhance the reliability and performance of our customer's product," says Dielectric Labs' Vorlop.

Surface mount technology (SMT) is currently one of the most desired features in the RF capacitor market. There is a great deal of excitement concerning high dielectric substrate materials. "Surface mount technology is finally beginning to approach the potential that people had talked about five years ago," comments Scott Newman. As with most other areas of electronics, surface mount technology is the wave of the future. "Surface mount products are becoming a very desired commodity. As size comes down because of the trend for SMT, there is an increased demand for consumer type RF transmitters that would use trimmer capacitors," says Marty Markson, Vice President of Sales and Marketing for Sprague-Goodman Electronics Incorporated.

With advances in fields such as MRI, telecommunications, and medicine, new interest has been generated in the RF capacitor market. Ron Nielsen, Director of Capacitor Technology for High Voltage Components, Incorporated, says, "Things are changing so rapidly it seems like everything is surprising." While some markets for RF capacitors are slowing down, companies are investigating other areas where new capacitor designs can make an impact on the development of important technological advancements. **RF**



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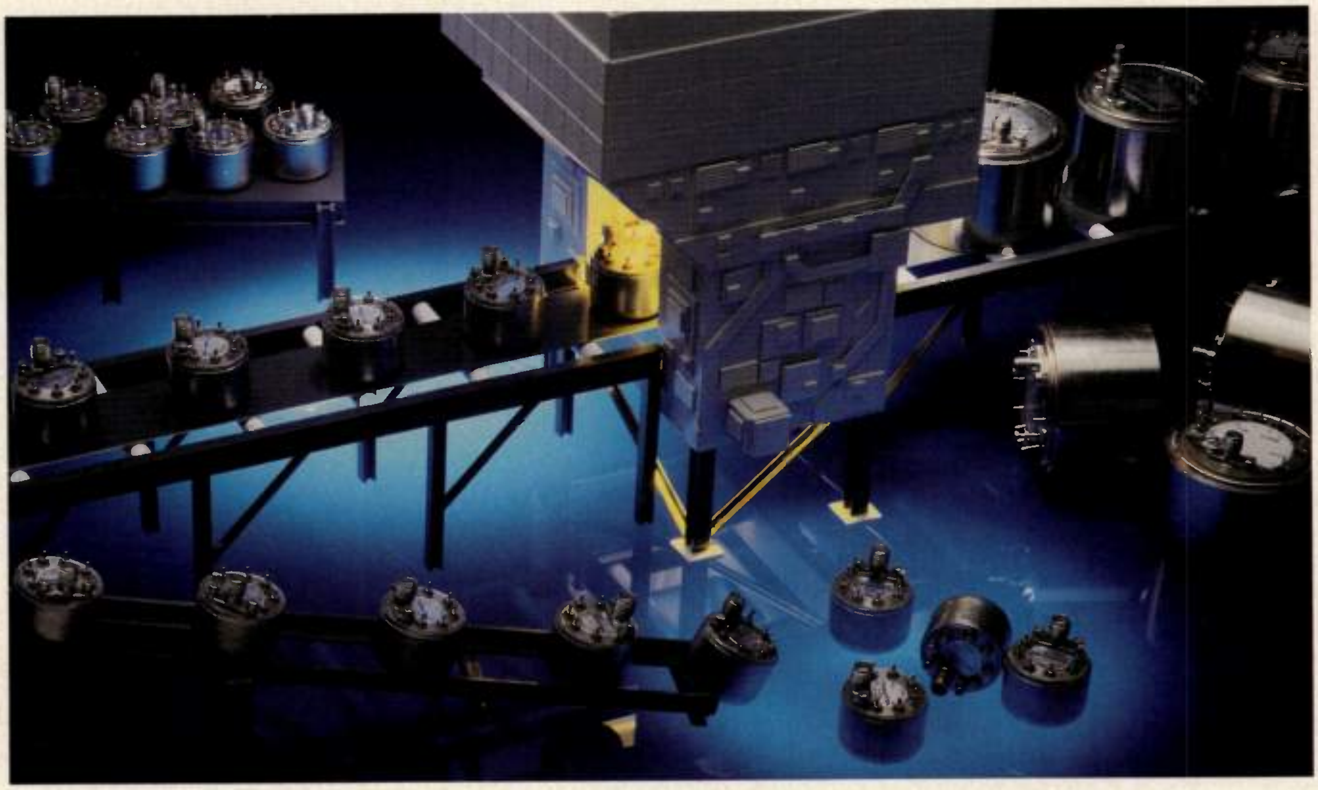

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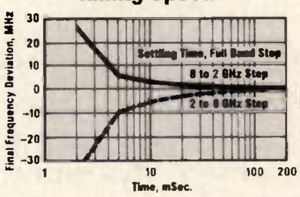

AV-7036 covers the 3.0 to 6.0 GHz range. AV-7028 provides 2.0 to 8.0 GHz frequency coverage. For both models, phase noise is typically −100 dBc/Hz at 20 kHz offset from carrier. And, power output is +13 dBm.

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INFO/CARD 44

Instrumentation Options and Cost Considerations of Compliance Testing

By Roger Southwick
EMC Consulting

Compliance with FCC regulation part 15 subpart J is mandatory for manufacturers of computers and many other electronic devices. This requirement creates the need not only for engineering talents but also for a number of managerial decisions, the principal of which is to determine the most economical way to meet the testing requirements of the regulation.

One option is to have the testing done by an outside testing laboratory. The advantage of this option is the savings in test equipment and test site cost, as these items are provided by the testing laboratory. One disadvantage is the high cost per test. Another possible disadvantage is that scheduling prequalification testing may be more difficult with the test laboratory since testing priorities must be coordinated with other clients.

Another option for meeting the FCC testing requirements is to do the testing in-house. In this case, the initial costs including necessary test instrumentation and a test site are high, and an engineer experienced in the FCC part 15J testing procedures is required. Thus the disadvantage of this option is the high initial cost plus delays in starting.

Yet a third option is to do prequalification testing in-house and have the final testing done by a testing laboratory. This middle-of-the-road approach has a number of advantages: initial instrumentation and site costs are reduced and prequalification testing can be efficiently scheduled. Further, as time passes, in-house experience will increase, making the move to full compliance testing much easier.

A consideration of each of the above options is the necessity for prequalification testing. Prequalification testing is an absolute necessity in the product-development cycle. Those doubting this will learn, as so many have, that the costs of redesign and schedule delays far exceed the cost of prequalification testing. Fortunately, prequalification testing methodology is not specified by the FCC, so test methods can be modified from full compliance testing to reduce both test time and cost. The same is true for the test site, which can be less sophisticated or even temporary. The only criterion is that the test results must provide a reasonable indication of whether or not the product will pass the final compliance test.

One critical factor in the above discussion is the cost of instrumentation. Quite recently a number of less costly spectrum analyzers have been introduced into the EMI measurement market. In the $12,000 to $20,000 range these spectrum analyzers have significant influence over the above decisions.

There has been an age-old argument about the suitability of spectrum analyzers for this type of testing. Those opposed, mostly RFI receiver manufacturers, claim that spectrum analyzers will overload due to the lack of preselection. While it is true that spectrum analyzers can overload more easily than RFI receivers, this does not mean that the spectrum analyzer is unsuitable for this type of testing. In the first place, the situation in which overload will occur is quite rare; and second, there are ways to both detect and prevent overload in most situations. For instance, high and low pass filters will greatly reduce the possibility of overload. If all else fails many spectrum analyzers — including some in this less expensive class — can be purchased with a preselector. My personal recommendation is to test any receiver on the test site where it will be used prior to final purchase to determine if possible overload conditions exist. The advantage of the spectrum analyzer, which is seldom mentioned or understood, is that when used correctly it has a much higher probability of detecting non-steady-state signals than does the RFI receiver (1).

Unfortunately for the potential buyer, instrumentation specification writing has become such an art form that comparing the specifications of any two instruments is nearly impossible. It is, however, possible to discuss some of the options and features offered by this less expensive class of spectrum analyzer.

It is extremely important to have high frequency stability. Some of the less expensive spectrum analyzers come with less than required stability and then offer an enhance stability option. In this situation the option should always be acquired.

Another very useful option is a tracking generator. Tracking generators are very useful for site attenuation and insertion loss measurements; the latter are necessary to calibrate measurement systems. There are two kinds: the internal and the separate tracking generator. The internal version is preferred because everything is in one unit and the alignment is automatic. In the external version the alignment will probably be manual. Tracking generator frequency range is not always the same as in the spectrum analyzer. While most of the spectrum analyzers have a lower frequency range of 9 to 10 kHz, in some cases the tracking generator lower frequency range is only 100 kHz. This is usually not a problem, since applications like insertion loss and site attenuation measurement aren't normally required at such low frequencies.

If final compliance testing is the goal, a Q-P detector is necessary. The FCC will accept peak data, but the difference between peak and Q-P data is often large, thus imposing a penalty on the performance of the item under test. Another necessary option is the IEEE 488 bus interface; without this option automation is impossible.

The following is a list of some of the new class of less expensive spectrum analyzers, along with comments about available options and features. The discussion does not provide all the specifications; for additional information, contact the manufacturers.

Tektronix model 2710 10 kHz to 1.8 GHz. This unit has some very good options, including enhanced frequency stability, additional if band widths, tracking generator, preamplifier, IEEE 488 bus interface, but no Q-P detector.

Without the Q-P detector this unit would not be suitable for final compliance testing.

Hewlett-Packard 8591A 9 kHz to 1.8 GHz. This unit is now available with a Q-P detector option, but only with 9 and 120 kHz band widths. This means that the usable Q-P frequency range starts at 150 kHz. A tracking generator, IEEE 488 bus and a frequency stability enhancement option are offered.

Anritsu model MS2601A 10 kHz to 2.2 GHz. This unit has an IEEE 488 bus interface and a Q-P detector as standard equipment. A 100 kHz to 2 GHz external tracking generator is available. A 100 Hz DC coupled input and a Personal Test Automation (PTA) option is offered. The PTA option is a card input device that, in brief, permits a sort of semi-automation by the use of measurement subroutines, which can be loaded on cards. An external preselector is also available.

Advantest America R3261/3361 A & B. The A version has a frequency range of 9 kHz to 2.6 GHz, and the B version has an extended upper frequency range of 3.6 GHz. A full-range internal tracking generator is available for both the A and B models. Q-P is standard; also it has both a 40 and a 70 dB dynamic range. The IEEE 488 bus is standard and a preselector is available. There is also a PTA option similar to the Anritsu MS2601A.

IFR A-7550/A-8000. The A7550 has a frequency range of 10 kHz to 1 GHz and the A8000 has a frequency range of 10 kHz to 2.6 GHz. Both units are synthesized, and have optional internal tracking generators. Other options include RS-232 or IEEE-488 interfaces, Q-P detector, and an internal battery pack.

The main difference between this class of spectrum analyzers and more expensive models such as the HP 8568/8566 and Tektronix model 2782 is in the frequency stability. Although specifications on frequency stability are difficult to comprehend, the obvious clue is the available band widths offered. In the less expensive spectrum analyzers, 10 Hz and lower band widths will not be found since the frequency stability does not provide sufficient resolution. For FCC part 15J measurements, the narrowest required band width is 9 kHz for Q-P measurements above the 450 kHz frequency range. The 200 Hz Q-P band width requirements are for certain VDE in the 10 to 150 kHz frequency range. The frequency accuracy for EMI-type signals is not well defined, as this class of signal has undefined spectrum shapes that may be unstable. The key to resolving this problem is to measure frequency in a very narrow frequency span, since frequency accuracy is a percentage of frequency span width. By this method the less expensive class of spectrum analyzers perform equally as well as more expensive models.

The most meaningful accuracy specification is total amplitude accuracy. Here it is surprising to find that the less expensive but newer models, such as the Anritsu MS2601A (2) and the Advantest R3261/3361 A & B (3), specify plus or minus 1 dB total amplitude measurement error. The HP 8568/8566 series has a total amplitude error of over ± 2 dB before Q-P detection (4). Again total amplitude error is a difficult figure to determine since manufacturers apparently do not want this type of comparison to be made. (Yet another reason why the buyer should test all relevant specifications before purchase). Amplitude errors can be reduced by other means such as signal substitution, but this adds to the total measurement cost.

There is absolutely no technical reason why the less expensive spectrum analyzers that have Q-P detectors cannot be used to perform those tests specified by the FCC part 15J regulation. To illustrate the capabilities of this class of spectrum analyzers, the author used the Advantest R3361 A. The Advantest unit has a very attractive performance-to-cost ratio. Also included are some features not found on more expensive spectrum analyzers. One of these is single command calibration, which obviates the need to connect cables. Tracking generator calibration is also a single command. The Q-P mode is very simple to initiate; it has a 40 and a 70 dB mode. The 1 dB specification applies to the 40 dB mode only. There are some minor faults in the command set and the manual is much too abbreviated. The Anritsu is slightly less expensive than the Advantest, but also has a less desireable external tracking generator.

In terms of cost efficiency over the long term, labor costs per test must be calculated. The only way to reduce labor costs for testing is to automate the test. Automation also has the advantage of maintaining consistent data. Among the spectrum analyzers mentioned in this less expensive class, only the Advantest model has a third-party software package, the EMI Commercial Measurement Program (5). The reason for this lack of software is that the reduced frequency stability of this class of spectrum analyzers means that the tolerance of the frequency span end points is greater, a fact that greatly complicates the design of a measurement program. As an example of this problem, if the frequency span error is ± 2 percent, a 30 to 200 MHz span would have an error of about ± 3.4 MHz. When a 30 MHz signal is measured in a 30 to 200 MHz span the signal may not appear on the display depending on the sign of the error.

However, there are several solutions to this problem. One solution is to use a large number of narrow measurement bands and put all the bands back together after the measurements are completed. One disadvantage of this solution is that a huge amount of memory is required to store all the traces. Another disadvantage is that end point errors of all the measurement bands will accumulate, and the reconstruction process can in itself create errors.

A second solution is to compensate for the errors by an alignment process. The EMI Commercial Measurement Program, does just that. The alignment compensates for a large part of the frequency span error, and the remainder is compensated for by the size of the initial frequency span in which the selected signal is measured. The software also sorts the trace data in sets of four and saves only the highest of each set of four trace points. This reduces the number of trace points by a four-to-one ratio and greatly reduces the memory requirement, so that there is no need for extra memory. The alignment must be performed before initial use of the program, but the alignment data is saved and the alignment rarely has to be repeated.

The only correct way to measure EMI signals is to measure each signal individually, since this is the only way that the measurement time at specific frequency can be controlled to ensure a high probability of signal detection. To find the signals that need to be measured from the trace data, the frequency of each signal must be calculated. The resolution of this process depends on the number of data points in the trace and the accuracy of the starting point. One characteristic of the less expensive spectrum analyzers is a reduced number of data points in the trace. Because of the reduced number of trace data points, the only option is to decrease the initial span to increase the resolution, and to align the start frequency. In this respect the Advantest model has 700 trace data points, which is more than any of the other models discussed and thus it provides sufficient resolution to locate signals from the trace data without having to resort to a large number of measurement bands. The residual error of the signal frequency is within the initial frequency span in which each signal is measured.

In the EMI Commercial Measurement Program, the signals to be measured individually are selected by an offset from the limit, which is set by the operator. When the desired number of signals has been selected by the offset the program will automatically find each signal and measure it in a very narrow frequency span in the peak and then in the Q-P detection mode. This process automatically passes measured data to succeeding parts of the measurement process to adjust the spectrum analyzer's settings to optimize the measurement process. This type of process ensures the maximum possible accuracy of both frequency and amplitude as well as high probability of detection (6).

In conclusion, considerable cost savings for FCC part 15J compliance testing may be obtained by the careful selection of both a spectrum analyzer and a well-designed measurement program.

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About the Author

Roger Southwick is owner/president of EMC Consulting, 2716 N. Estrella, Tucson, AZ 85705. He has been involved with EMC measurements and compliance testing for the past 29 years. He can be reached by telephone at (602) 792-9491.

Focus On Ferrite Filters

By William D. Kimmel, PE
Kimmel Gerke Associates, Ltd

Ferrites are widely used as electromagnetic interference (EMI) suppression devices. Ferrites are your friend: they eat undesired energy. Ferrites are inductive at lower frequencies, but become lossy at higher frequencies, making them ideal as absorbers of unwanted EMI. But the application of ferrites is subject to misunderstandings. In this article, we will try to show how easy it is to design a filter using a ferrite and capacitor.

The key in ferrite filter design is KISS -Keep it simple, stupid. You don't need a page full of input parameters and a Cray supercomputer to get approximate filter effectiveness. A pencil, paper and calculator to make a quick calculation will get you close enough for many purposes. Depending on the criticality of the function, you may need to cut and try in the lab to fine tune it.

An assumption in EMI suppression is that you are looking for a low pass filter. You are attempting to block the unneeded high frequency components while not degrading the desired signal.

T, L, Or Pi?

A common question is do we use a T or Pi filter? Depending on the source and load impedances, the performance of an L filter will match a T or Pi filter.

We can make a simple statement about filters: Filters are most effective when maximum impedance discontinuity exists between the filter and the source or load. That means that a high impedance source or load should face a low impedance filter element, and a low impedance source or load should face a high impedance filter element.

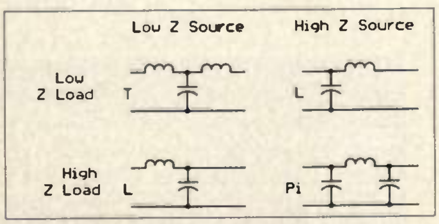

Figure 1. Filter selection

Figure 1 shows the possibilities. Note that a high impedance source or load always faces a low impedance shunt element (capacitor), and that a low impedance source or load always faces a high impedance series element (ferrite, or sometimes a resistor).

Thus, you will find it ineffective to put a ferrite in series with a high impedance input circuit - ferrites need current to work on - so a shunt capacitor is needed.

Similarly, it is ineffective to put a shunt capacitor at the output of a driver, since they are usually low impedance. It is best to insert a small series impedance before the capacitor.

If the source and load impedances are both very high or very low, it is sometimes sufficient to insert only a series inductor or shunt capacitor, eliminating the middle element in a T or Pi filter.

In the real world of electronics we typically see high input impedances and low output impedances. And our filters should be designed accordingly. Strangely enough, commercial power line EMI filter effectiveness is typically specified using a 50 ohm source and load. Where do you ever see 50 ohms, outside of coax and test instruments?

Use Approximations

Use simple algebra to estimate your filter effectiveness. Here are the rules:

1. Design the filter by calculating at the high and low frequencies of interest. Mid-range resonances are not usually a problem if you are using ferrites as lossy elements.
2. Be sure to include parasitic inductances and, in extreme cases, parasitic capacitances.
3. Compute with absolute values of impedances. Use of complex impedances will quickly bog you down.
4. Select impedance discontinuity at each stage. Strive for a minimum of 10:1, and maximum of 100:1. If you need more attenuation, use multiple stages.
5. Ignore small impedance factors.

An example is shown in Figure 2. These approximations are quite good if the inequalities are met by a factor of ten. If the factors drop below that, then the accuracy can be improved with little additional effort.

Ferrite Geometry

What is important in selecting a ferrite? Impedance is primarily a bulk effect. The more pounds the better. The critical parameters are the length and the ratio of outer radius to inner radius.

For a given outer envelope, the effectiveness of a ferrite decreases as the of the inner increases. Thus, it is the inner part of the ferrite that does the most work. Hence, it is to your advantage to select a ferrite with as small an inner radius as possible, provided you don't saturate the core (that's discussed below).

The effectiveness of a ferrite increases linearly with the length. Thus, doubling the length doubles the impedance. Don't get too exuberant with this fact, however. You will find that there is a definite limit to the effectiveness of the ferrites, and it you add more than a few, you will get no further improvement. This is because interference has a nasty way of making an end run via parasitic paths.

Saturating A Ferrite

One problem of using ferrites is that they saturate under large net DC or low

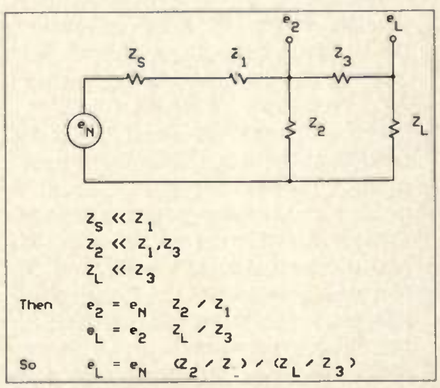

Figure 2. Filter example

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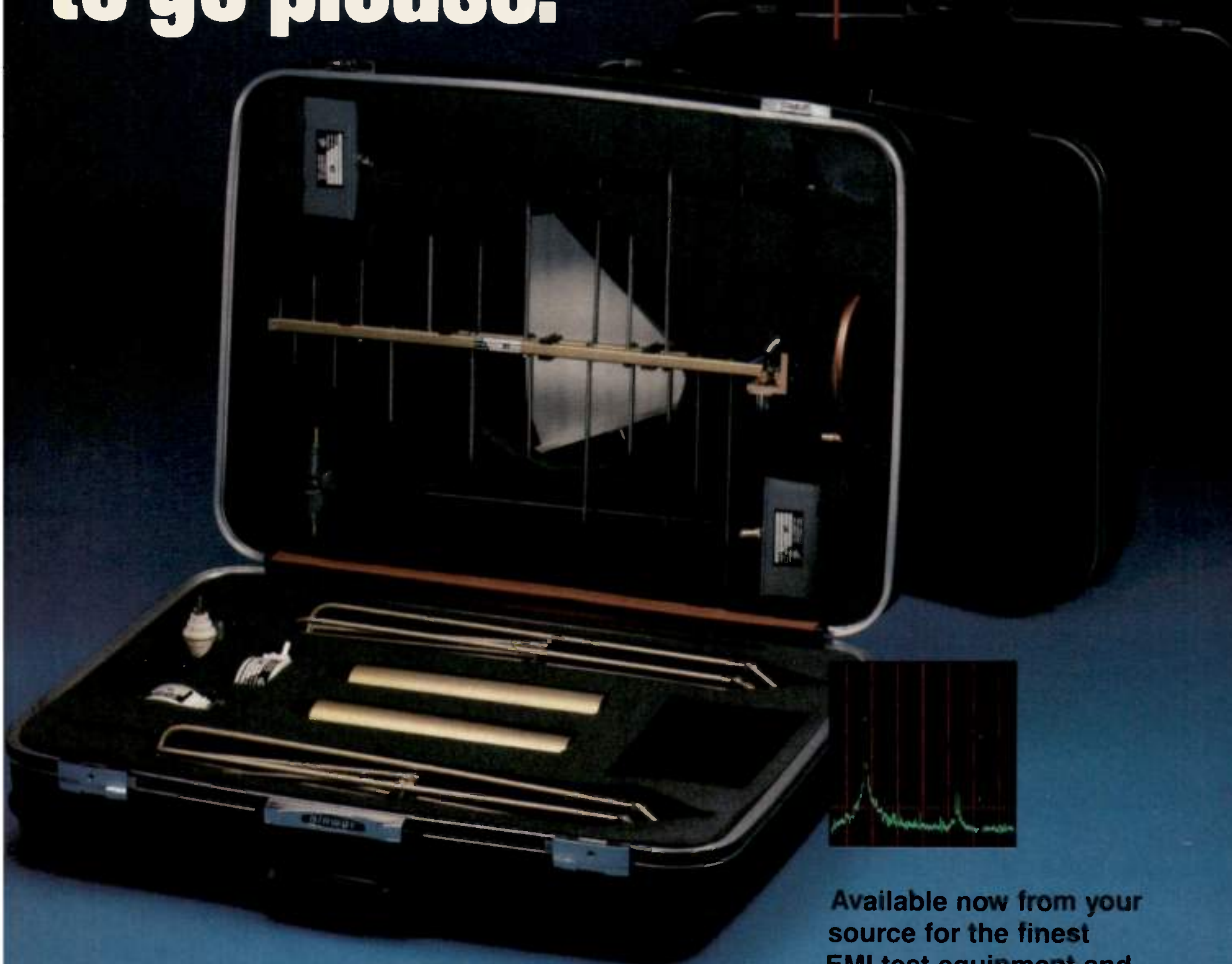

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SAS-200/540	20 - 300 MHz	Biconical
SAS-200/541	20 - 300 MHz	Bicon'l, Collapsible

MODEL #	FREQ. RESP.	DESCRIPTION
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frequency AC currents. Notice we said *net* current. A ferrite clamped over both the DC and return lines will have no net DC current.

But such practice will eliminate only common mode interference. And if you need differential mode filtering, you will need to reckon with DC. If you are filtering signal lines, current is usually negligible. If you are filtering DC or 60 Hz, then you need to know the peak current and design accordingly.

Unless the ferrite is a thin annular ring, they saturate from the inside out. The relationship is:

$$I = 2\pi R H_s$$

where I is total current (multiplied by N turns)

R is radius from center
H_s is saturation magnetic field intensity

In the ferrites most widely used for EMI, H_s is about 1.5 Amp/cm, which gives us a convenient relationship of:

I = 10 R, where R is the inner radius in cm and I is the current at which the ferrite will commence saturation.

This figure is only approximate, due to the nonlinearity of the hysteresis loop.

Again, the core saturates from the inside out, and it is not fatal to have the inner portion in saturation as long as reduced performance can be tolerated. Thus, when selecting a ferrite, the impedance will never decrease by using a small inner diameter.

In fact, saturation is permissible in certain transient power applications. The high frequency spikes and ringing in power turn-on will be suppressed by a ferrite before the current drives it into saturation. Similarly, inductive kick in turn-off of power circuits will drop the ferrite out of saturation in time to suppress most bursts of noise.

Impedance Of A Ferrite

Calculating impedance of a ferrite is not simple, and it is best to refer to the manufacturers' data. They will commonly specify impedance at upper and lower frequency limits, and if those are not in the right range, there are tables for extrapolation.

Even these figures are not real accurate, since the actual impedance will depend on bias current, including any residual magnetism.

The most widely used ferrite formulation is effective at frequencies above about 30 MHz, and is good up to about 1 GHz. Other formulations are available, with the following two tradeoffs:

1. The range of effectiveness shifts with different ferrites. If you want to cover a lower frequency, then you will lose performance at the high end.

2. Permeability decreases with higher frequency ferrites. This means that higher frequency ferrites will have lower average impedances and lower frequency ferrites will saturate at lower currents.

Be sure to use the right formulation. There are many ferrite formulations used for various applications, but which may not be suited for EMI suppression. Power supply designers, especially, need to know that the ferrite they have been using is not formulated for EMI.

Multiple Turns

The impedance of the ferrite increases with the square of the number of turns. This is useful when trying to get effectiveness at frequencies below 30 MHz. When the impedance starts to falter, you can put a couple of extra turns on the core.

The tradeoff is that winding capacitance will take its toll at higher frequencies. If you space the windings around the core, the capacitance will be less significant. Generally, you can get by with three turns (nine times the impedance) without seriously degrading the high frequency impedance.

Don't forget that saturation current increases with the number of turns.

If you need still more impedance, use a multiple hole ferrite, or add a second single turn ferrite in series to block the high frequency capacitive path.

Summary

Ferrites are a most valuable tool in your bag of tricks when dealing with EMI. Just remember some basics:

- Keep your computations simple
- Match your filter configuration to the source and load impedance
- Be aware of saturation effects
- Use the right ferrite formulation *RF*

About the Author

William Kimmel, PE, is a principal at Kimmel Gerke Associates, Ltd. The firm specializes in preventing and solving electromagnetic interference and compatibility (EMI/EMC) problems. Mr. Kimmel can be reached at 1544 N Pascal, St. Paul, MN 55108, Tel: (612) 330-3728.

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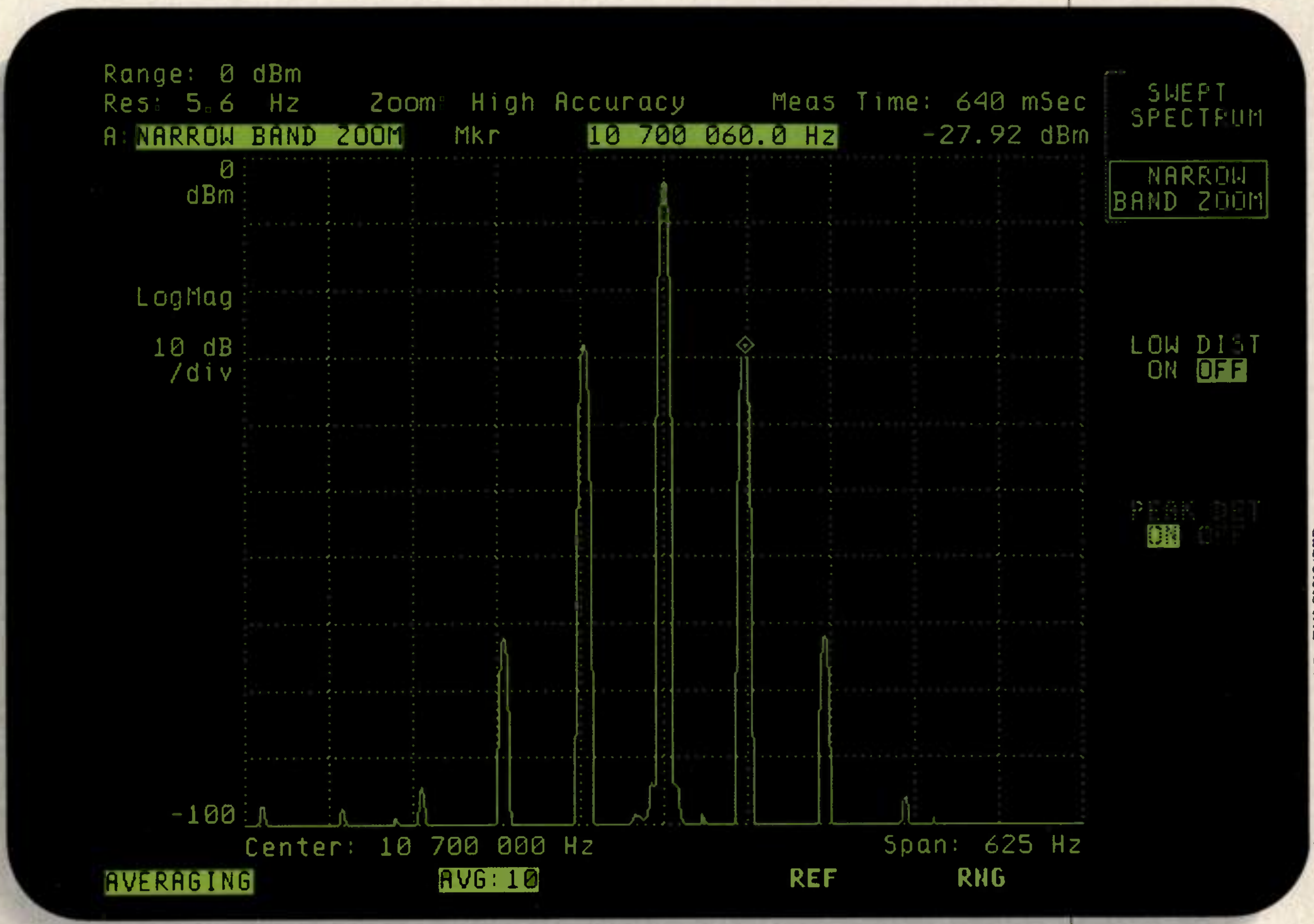

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New Test Cell Offers Both Susceptibility and Radiated Emissions Capabilities

By John Osburn
Electro-Mechanics Company

January 1, 1992 — A day that will be remembered by electromagnetic compatibility (EMC) engineers in the United States for some time to come. It is on this date that the European Common Market EMC Directive becomes effective and EMC immunity testing of all electronic products to be sold in Europe, particularly Information Technology Equipment (ITE), becomes mandatory.

Immunity testing, or susceptibility testing involves exposing electronic equipment and systems to intentionally developed electromagnetic fields. The purpose of this exposure is to determine if the exposed equipment can continue to function in the presence of these fields, without degradation of performance, or even minor malfunction. There have been military electronics which have displayed susceptibility to incident electromagnetic fields and these episodes have had serious, even fatal consequences. The common market regulation authorities feel that susceptibility episodes, in general, are preventable which explains the implementation of immunity requirements.

The application of immunity requirements by the regulatory agencies of the European Common Market is also based on a sober assessment that the electromagnetic environment will continue to become more complex and will be more likely to interact with electronic products. Therefore, compliance with the immunity requirements will go a long way in assuring continuing faultless performance of electronic products sold in Europe. The emission specifications will be standardized throughout the European Common Market, however, they will require immunity testing.

Methods for performing immunity testing are under discussion by several regulatory committees. Compliance with the total set of requirements will mean ease of import and distribution of products. There is then, an increased premium on the rapid, successful conduct of all EMC testing for immunity and emissions. New technologies which increase test performance quality and decrease test time will be in demand. This article describes an EMC radiated testing technology which provides significant technical advantages as well as decreased test time. The new technology allows the conduct of radiated emissions and immunity testing in a rapid, efficient manner at a cost below that for EMC test facilities. This new technology is the GigaHertz Transverse Electromagnetic Cell, the GTEM!™ cell.

Description of the GTEM!™ Cell

The GTEM! cell (Figure 1) is an offshoot of the TEM cell, as developed at the National Institute of Science and Technology (NIST). As shown in Figure 2, it is a section of 50 ohm transmission line with a unique geometry. At the input a normal 50 ohm coaxial transmission line is transformed into a rectangular cross section with a ratio of 2:3 in height to width. The cell is flared along the longitudinal axis to increase the cross sectional dimensions of the transmission line. The septum, or center conductor is transformed from a round cross section to a flat wide conductor, located well above the center of the cell. This maintains the 50 ohm characteristic impedance while increasing the volume of the cell under the septum, allowing a larger test volume. The size of the test volume is proportional to the length of the cell, with larger test volumes requiring a greater length.

The most interesting feature of the cell is the means of termination. A normal transmission line is terminated

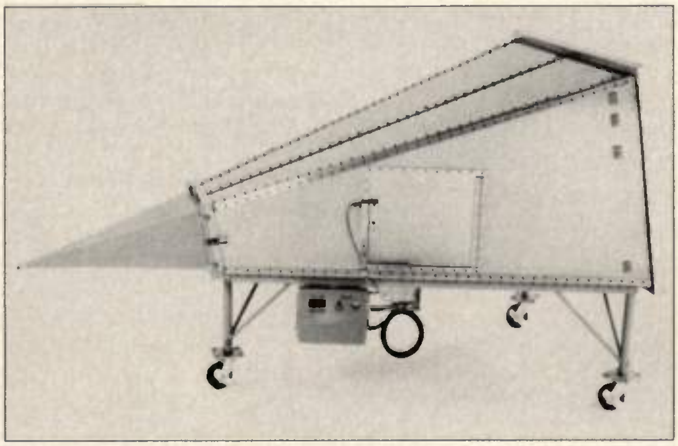

Figure 1. A GTEM! model 5305.

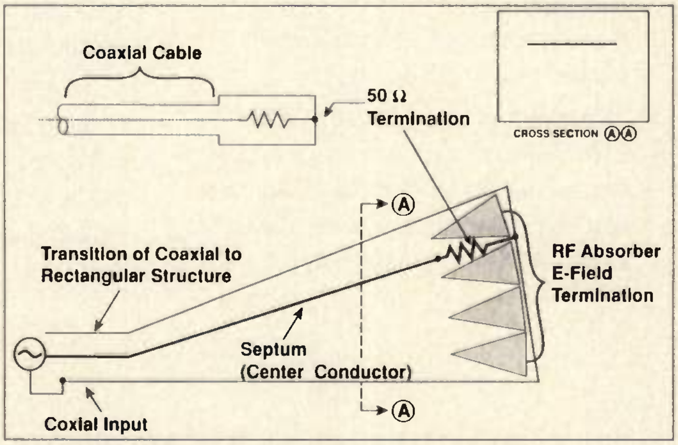

Figure 2. Physical description of a GTEM!

"We came to TRW with our Model 1000 Digital Lincompex, a rack-mounted unit used to suppress noise in strategic voice communication systems. It weighed 22 pounds and drew 115 watts of power. TRW's engineers helped us redesign the unit using just a few off-the-shelf digital signal processing ICs and software-programmable memory chips.

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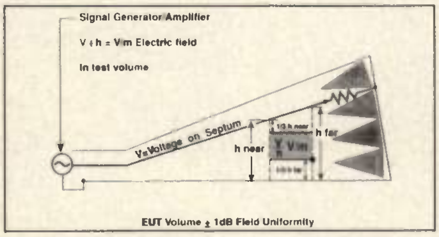

Figure 3. Parameters of a GTEM! Important for radiated immunity (susceptibility) testing.

with a 50 ohm load. A GTEM! is doubly terminated. The septum is terminated in a set of printed circuit boards containing several thousand discrete carbon resistors where total resistance of the boards is 50 ohms. The distribution of

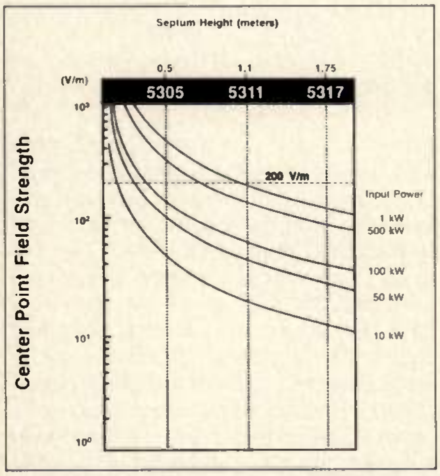

Figure 4. Input power required for GTEM! cells of several septum heights to achieve desired immunity test levels.

the resistive values matches the current distribution in the septum. Fields generated in the GTEM! cell are terminated in an RF absorber whose characteristics are chosen to match the performance of the cell as a function of frequency. Thus, electromagnetic fields generated in the cell by driving the RF connector are terminated in the load boards and the RF absorber. Electromagnetic fields generated in the cell, which could produce standing waves will not do so

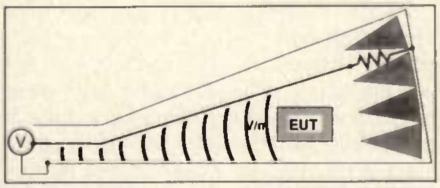

Figure 5. Use of a GTEM! for radiated emissions testing.

Figure 6. A radiated emissions test in progress.

because they are not permitted by the characteristics of the cell. Waves that propagate toward the apex of the cell are damped since the tapering structure will act as a waveguide below cutoff, and will not allow propagation of the wave toward the apex. Waves propagated towards the far end of the cell are absorbed.

Uses of GTEM!

The GTEM! was developed for use for both radiated immunity and radiated emissions testing.

For radiated immunity testing, Figure 3, the Equipment Under Test (EUT) is installed in the GTEM! in the test volume. This test volume occupies the center third of the total volume between the septum and the bottom of the GTEM!. Since the septum is inclined in the interior of the GTEM!, there is the near end height (h near) of the test volume and the far end height (h far) of the test volume. The actual test volume is centered under the septum. It is bound on the top and bottom as follows: the center third of the volume bound vertically by 1/3 of the dimension, h near, down from the septum and 1/3 the of the dimension, h far, up from the bottom of the GTEM!. This volume is almost completely within the ± 1 dB field uniformity area of the GTEM!.

An appropriate signal source is connected to the coaxial connector, and the source is adjusted to the desired frequency. The amplitude level is adjusted to develop the field strength required. The applied field level is swept in frequency, at a rate determined by the time response of the system under test. The sweep rate is limited by the maximum rate that the signal generator and amplifier can be interchanged. This will usually exceed the response time of any susceptibility occurrence of the EUT. Thus, this is not likely to impose a limit on test time. If desired, the EUT may be repositioned to a second or third polarization to maximize coupling in all three swept axes. An estimate of the power required for the development of the desired test field strength level within the ±1 dB volume is shown in Figure 4.

For radiated emissions testing, the situation is the dual of the immunity, as shown in Figure 5. The EUT is installed in the center of the test volume in an initial orientation. A voltage measurement is made covering the entire frequency range desired, usually 30 MHz to 1 GHz. An equipment setup with computer control of the spectrum analyzer as might be used for such a measurement is in Figure 6. Data from such a measurement is shown in Figure 7. The EUT (or EUT array) is then rotated to a second swept axis and the measurement is repeated, followed by a measurement of the EUT in a third swept axis.

At the completion of these measurements, the controlling software program computes the equivalent electric field for comparison to a specification limit, as a function of frequency. The computation is not straightforward and the mathematical development behind it is extensive. It is based on the following steps, at each frequency in the measurement:

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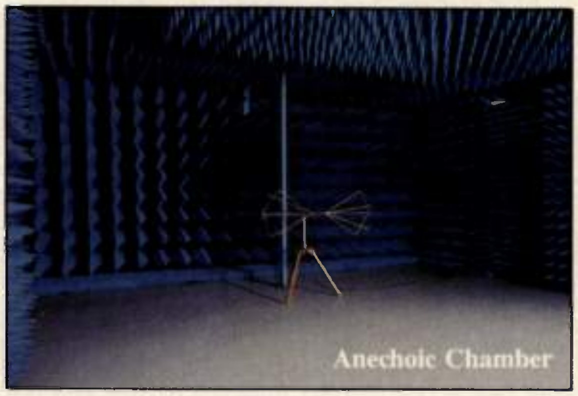

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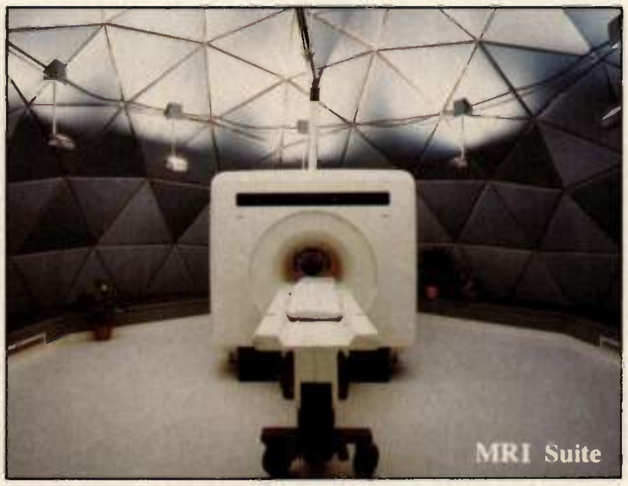
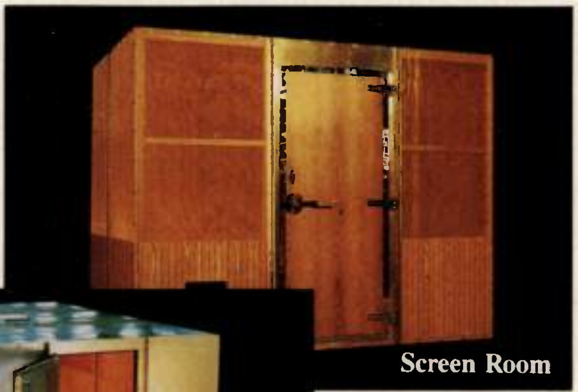


MRI Suite

Tempest Computer Room



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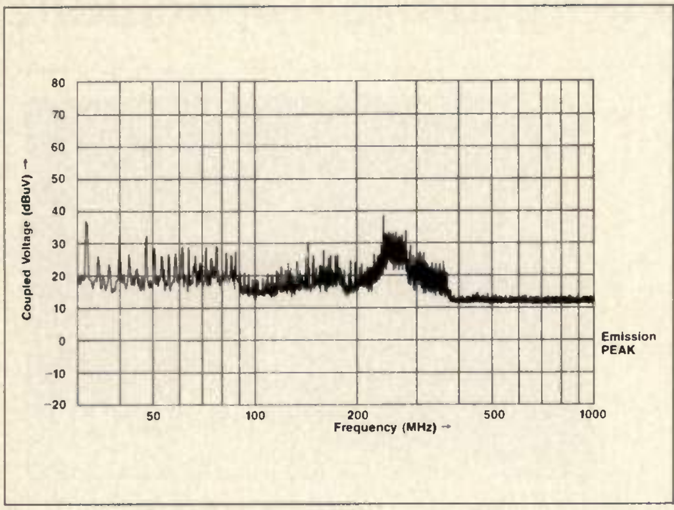

Figure 7. Typical orthogonal axis voltage measurement with a GTEM!

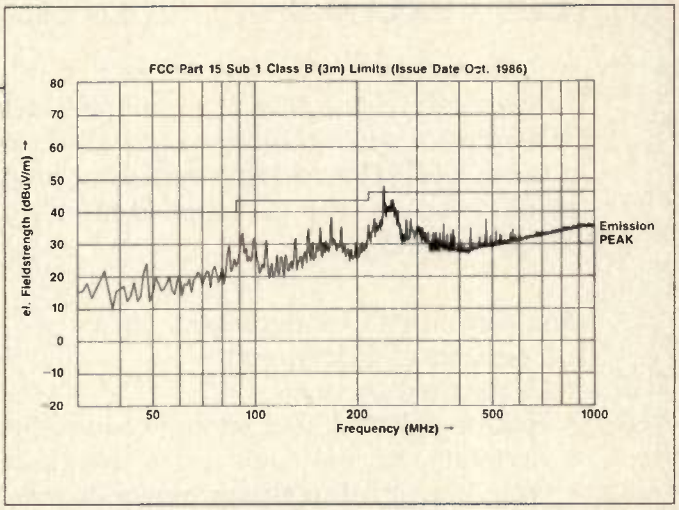

Figure 8. Typical radiated emissions data as taken with a GTEM!

the three orthogonal voltage measurements are summed to produce a resultant voltage; the total resultant voltage measured at the vertex of the GTEM! is used to derive equivalent electric and magnetic dipoles and their assumed linear excitation at the frequency under consideration; these equivalent dipoles, with assumed excitation, are then mathematically placed over a perfect ground plane at a specific height, typically 1 meter; the measurement distance typically 3 or 10 meters is specified; and successive computations are made of the vertically and horizontally polarized electric field strength over the scan height of 1 to 4 meters. The computation is performed at intervals of as little as 0.05 meters; the maximum value of the electric field is selected from the computations of both vertically and horizontally polarized field strengths; and finally, these maximum values are reported, and compared to the appropriate specification limit.

This process is repeated over the measurement frequency range. At the end of the computations, the calculated field strengths are plotted against the specification limit, Figure 8.

Conduct of Actual Testing

While the theory behind the conduct

of both types of radiated testing is easy to describe, the actual testing using GTEM! is only minimally more difficult.

Generally, there are only a few items to be considered when performing testing in a GTEM!. First, the EUT should be kept centered in the test volume, preferably in the ± 1 dB volume to preserve the maximum accuracy. Second, for either immunity or emissions testing, the signal generators for immunity testing, and spectrum analyzer or receiver for emissions testing should not be scanned at rates which exceed the capability of the EUT to respond. This could necessitate the rewriting of the typical emissions test software for the EUT to minimize the necessary dwell time at each observation frequency.

Radiated Immunity Testing. There are two major considerations in the actual conduct of radiated immunity testing in a GTEM!: 1) Whether the test should be based on the computed value of the field strength as the ratio of voltage on the septum to the inside of the cell, or should the GTEM! be augmented with additional instrumentation to measure the field strength during test; and 2) The determination of the maximum acceptable scan speed which allows rapid testing while assuring adequate evaluation of the EUT. Conduct of the actual test is straightforward. It should be noted that the exciting field in the GTEM! is extremely uniform with respect to the fields generated by other, more traditional means.

Radiated Emissions Testing. Radiated emissions testing must follow the procedure of obtaining three swept measurements in an xyz, yzx, zxy sequence. The terms xyz etc, refer to the positive directions of a rectangular coordinate system.

As a practical matter, there may be systems where three swept axes rotation may not be possible as gravity may be necessary for proper equipment operation. In this case, two axis evaluation may be adequate.

As with the case for immunity testing, the scan time of the radiated emission measurement is limited more by the need to adequately exercise the EUT than the test instrumentation. The measurement of the GTEM! voltage over the frequency range of 30 MHz to 1 GHz for a single axis is approximately three minutes with the use of a sophisticated spectrum analyzer. The total measurement in three axes and the computer computations takes approximately 30 minutes, as a function of the number of frequencies where the correlation must take place.

Comparison of GTEM! with Alternate Technical Approaches

It is interesting to compare the GTEM! to other traditional test facilities for EMC testing. The comparison will include both emissions and immunity requirements.

Standard TEM Cell. A standard rectangular TEM cell is limited in the frequency response by the dimensions of the cell. It is normally used for the generation of precise known field levels as standard fields. There has been some effort to develop the use the TEM cell for emissions testing, but since the rectangular structure allows the existence of standing waves, which also limit the upper frequency of use, this has not been adopted by the EMC community as a primary method for emissions measurement.

The GTEM! does not suffer the upper frequency limitations of the TEM cell due to its tapered structure, thus it may be used for both immunity and emissions testing, to frequencies that are well in excess of the geometric limit of the standard TEM cell.

Shielded Enclosure. The shielded enclosure has been used by the military EMC community for many years. Its advantage is that the external ambient environment is removed from the testing environment by the shielding of the enclosure walls. The disadvantage is that the performance of the shielded enclosure, when considered as a conductively walled cavity, with reflection and resonance effects, introduces significant error into both emissions and susceptibility measurements. The GTEM! provides the isolation of the shielded enclosure while avoiding the cavity effects that have plagued them since their popular use began.

Open Area Test Site. The open area test site avoids the problems of the shielded enclosure by not surrounding the EUT with conducting surfaces, but allows the RF ambient at the location of the test site to exist and severely complicate the measurement process. Since the open area test site is outdoors, immunity testing cannot be performed since the generation of test signals would almost certainly interfere with existing RF and broadcast services. The GTEM! prevents the contribution to measurement complexity of the external ambient while preserving the quality of the open field measurement.

Reverberating Chamber. The reverberating chamber is probably the most cost effective structure available for the generation of high levels of electromagnetic fields for high level immunity testing. This statistical method of testing however loses the data traditionally taken by susceptibility test methods. In addition, the reverberating chamber does not function well below 200 MHz. The GTEM! allows immunity testing for frequencies well below this, and as low as DC for special conditions.

The reverberating chamber has also been proposed for emissions testing. Again, this has not been pursued since the measurements possible have not been shown to be directly comparable to the traditional specification limits in dB uV/m.

A review of the material presented in this brief synopsis of the GTEM! indicates that it could be a highly efficient, cost effective, time saving method of performing EMC emissions and immunity testing. Its use would allow the completion of radiated emissions and susceptibility testing in as quickly as one day, provided there are no severe EUT problems encountered and would allow for qualification of products to the new EEC requirements in a timely manner. ***RF***

About the Author

John Osburn is the Director of Engineering at EMCO. He may be reached at Electro-Mechanics Company, PO Box 1546, Austin, TX 78767. Tel: (512) 835-4684.

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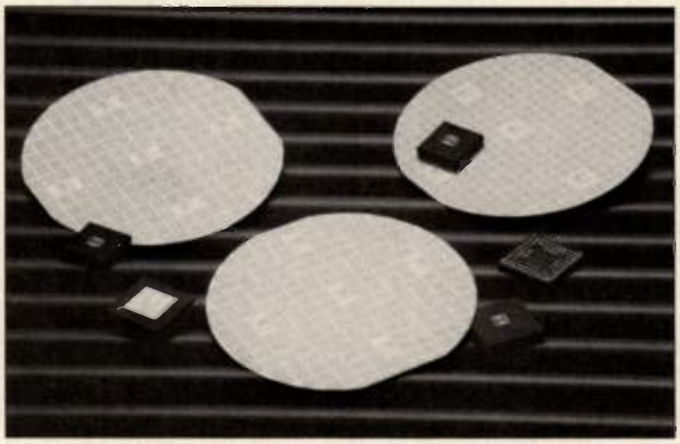
INFO CARD 43

Digital Filters Perform in FIR Applications

A family of three video speed (30 MSPS maximum) digital filters for direct form FIR applications is now available from Harris Semiconductor. The filters can implement adaptive filters, multibit correlators and other filter techniques. They employ a parallel processor architecture. Eight high-speed multiplier accumulator cells are integrated on a single device and are optimized for sum-of-product operations. Some other features of the three filters include 8-bit coefficients and signal data, 26-bit accumulator per stage, expandable coefficient size, data size, and filter length and sample decimation by 2, 3, or 4. These digital filters combine high processing speed and flexibility for a broad range of digital signal processing needs. On-chip decimation registers allow an increase in the effective number of taps per device. For example, an 8 tap 30 MSPS FIR filter can perform a 16 tap FIR filter at 10 MSPS with a 2:1 decimation factor. Applications include higher order 1-D FIR filtering (sharper frequency response), adaptive filtering, and decimation/interpolation.

Harris Semiconductor
INFO/CARD #200



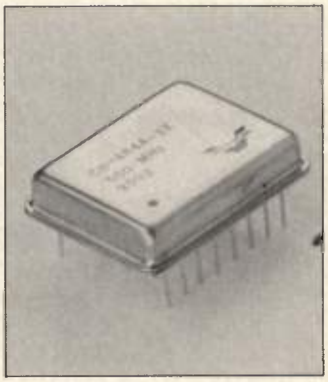
Shielded Enclosure Test Set

ASM Products has introduced its TS-450 portable, shielded enclosure test set. The test set operates at 450 MHz with dynamic range greater than 100 dB and is for measuring the shielding effectiveness of buildings, enclosures, and cabinets. The crystal controlled measuring system of this tester provides attenuation measurements for accurate evaluation of shielded conditions. It can be used as a production tool for building/refurbishing shielded enclosure/cabinets and as a calibration tool to monitor the performance of shielded enclosures. Coupled with appropriate instrumentation, the TS-450 test set can be used to perform a complete certification testing program. The system is a battery operated, lightweight and includes a test transmitter and receiver, and a thirty foot extension cable for leak probing.

ASM Products
INFO/CARD #199

Hybrid Sinewave Crystal Oscillator

Model CO-484 hybrid sinewave crystal oscillator from Vectron Laboratories is available at any frequency in the 4 MHz to 500 MHz range. Initial accuracy options at 25 degrees Celsius range from ±10 ppm to ±50 ppm. Standard temperature stability is ±25 ppm over 0 to 70 degrees Celsius with wider temperature range and higher stability options available. Output level is +7 dBm into 50 ohms with +13 dBm available up to 200 MHz. The crystal is much smaller than traditional 500 MHz crystal oscillators and it is housed in a 16 pin double DIP resistance welded can measuring 0.8 inches × 1.0 inches × 0.2 inches. For military applications it is available screen tested to Class B of MIL-0-55310. Delivery is 10 weeks ARO.



Vectron Laboratories, Inc.
INFO/CARD #198

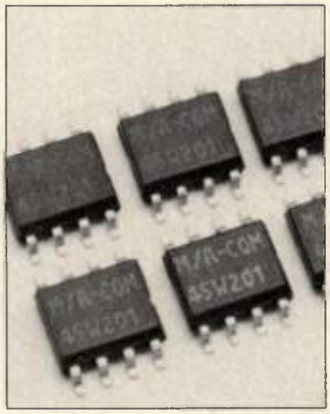
Attenuator Calibration Systems

Wavetek Microwave, Inc. has announced the introduction of two calibration systems for metrology labs and manufacturers of precision microwave components such as coaxial attenuators, filters, couplers and other passive devices. The calibration systems consist of the 8003 precision scalar analyzer, an 8910 series programmable sweep generator, and an attenuator calibration kit. Measurements which previously took several hours with receivers and slotted line techniques now execute in minutes with the same accuracy. The 8910 series programmable sweep generators cover the frequency range from 10 MHz to 26.5 GHz. They use fundamental oscillators for all microwave bands which substantially reduces the false responses and measurement errors due to harmonics and subharmonics with earlier microwave sweepers.

Wavetek Microwave, Inc.
INFO/CARD #197

50 MHz-2GHz SPDT Switch

M/A-COM Semiconductor Products now offers a surface mount 50 MHz - 2 GHz SPDT switch. This switch is ideal for high volume, commercial applications. The MA4SW201 SPDT switch utilizes a single monolithic chip that incorporates four PIN diodes and is housed in an SO-8 package. Characteristics include a nominal insertion loss of 0.3 dB at 50 MHz up to 0.5 dB at 2 GHz and nominal isolation of 65 dB at 50 MHz to 33 dB at 2 GHz. Secondary harmonic distortion at 1 GHz is nominally 75 dBc and third harmonic dis tortion at 1 GHz is nominally 90 dBc. Input return loss is nominally 25 dB at 50 MHz and 18 dB at 2 GHz. Switching speed is 50 nanoseconds nominally.



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3. Circuits shall be the original work of the entrant, not previously published. If developed as part of the entrant's employment, entries must have the employer's approval for submission.
4. Components used must be generally available, not obsolete or proprietary.
5. Submission of an entry implies permission for publication in *RF Design*. All prize winning entries will be published, plus additional entries of merit.
6. Winners are responsible for any taxes, duties, or other assessments which result from the receipt of their prizes.
7. Entries must be postmarked no later than March 30, 1991, and received no later than April 10, 1991.
8. All entries will remain confidential until the publication of the July 1991 issue of *RF Design*.

JUDGING CRITERIA:

Originality — Each design will be evaluated for similarity to work by others, unusual application of a device or technique, and other judgements of its contributions.

Engineering — Since engineering is a problem-solving profession, entries should clearly identify how the circuit was created in response to a need.

Documentation — The judges will require a complete description of the circuit, including a parts list and an explanation of its function. A summary of test data should also be included.

The PC SOFTWARE Contest

RULES:

1. Each entry shall consist of a computer program that assists in the design of RF circuits. Programs for test or control of RF circuits will also be accepted.
2. Programs must operate on computers compatible with either PC/MS-DOS, or Apple Macintosh operating systems. Note any special hardware requirements (memory, graphics, etc.).
3. Programs written in languages other than GWBASIC or BASICA should be supplied in both compiled, executable form, and uncompiled source code.
4. Entries shall be submitted on disk, accompanied by supporting documentation, including theoretical explanations and references, and instructions for operation of the program. A printed copy of the source code is required.
5. Programs must be the original work of the entrant, and must not be previously published or distributed (including distribution by open BBS or shareware). If developed as part of the entrant's employment, entries must have the employer's approval.
6. Submission of an entry implies permission for publication in *RF Design* and distribution by the RF Design Software Service. All prize winning entries will be published, plus other entries of merit. (Some restrictions on publication and distribution of source code may be acceptable.)
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JUDGING CRITERIA:

Technical Merit — Computer programs will be compared to accepted standards for accuracy, and will be judged for their achievement in translating RF design theory into software tools.

Usefulness — Judges will evaluate the software for its contribution to the practice of RF engineering. Criteria include ease of operation, time-saving, and value to designers of the circuitry addressed by the program.

Send entries to: **RF Design Awards Contest**
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Fiber Optic Microwave Delay Line

This fiber optic microwave delay line from Ortel Corporation incorporates a spool of optical

fiber in conjunction with Ortel's lasers and photodiodes to delay the signal. The line has a measurement sensitivity of −75 dBc and an estimated triple transit level of < −100 dBc.
Ortel Corporation
INFO/CARD #195

Broadband SP4T has High Isolation

The model DSO644 PIN diode SP4T switch has 88 dB of isolation from 10 to 100 MHz, 68 dB to 500 MHz, and 58 dB to 1000 MHz. It features only 0.7 dB of insertion loss and intercept point performance of +66 dBm second order, +43 dBm third order. The DSO644 is packaged in a hermetic 24 pin DIP and is available with optional military screening.
Daico Industries, Inc.
INFO/CARD #194

Low Loss SAW Filters

Thomson-CSF has developed a new generation of low loss SAW filters with center frequencies from 70 to 1500 MHz and with 0.8 to 3.8 percent bandwidth at −3 dB. Typical insertion loss is from 3 to 10 dB, shape factor is 2.5 and near-in rejection goes beyond 50 dB. TO8 or surface mounted packages are available.
Thompson-ICS
INFO/CARD #193

High Power GaAs FETs

The MGF-0900A Series of GaAs FETs has been designed for use in the high-power, amplifier stages of GPS systems, and UHF and microwave digital radios. The MGF-0906A, characterized at 2.3 GHz, features typical output power of 37.0 dBm, gain of 11.0 dB, and 40 percent power-added efficiency. The MGF-0907A has output power of 40.0 dBm, gain of 10.0 dB, and power-added efficiency of 37 percent at 2.3 GHz.
Mitsubishi Electronics America, Inc.
INFO/CARD #192

Snag-Free Shielding Gasket

Instrument Specialties Co., Inc. has introduced a snag-free, track-mount beryllium copper shielding gasket designed for heavy-traffic enclosures. Rite-Trak™ incorporates a reverse foldover lip to protect the assembly when the door of an enclosure is open.
Instrument Specialties Co., Inc.
INFO/CARD #191

75 Ohm Snap-Lock Coax Connector

A new 75 ohm SMZ snap-lock coaxial connector designed to eliminate unintentional disconnect has been introduced by Radiall, Inc. The unit mates with standard SMZ connectors and covers a frequency range of 0 -2.4 GHz. To eliminate any discontinuity caused by mechanical

movement, the coupling mechanism is separate from the electrical path.
Radiall, Inc.
INFO/CARD #190

Feed Forward Amplifier

AML, Inc. introduces a high dynamic range, feed forward 50 to 500 MHz amplifier. Typical

gain for this amplifier is 24 dB with a noise figure of 9.0 dB maximum. Its 1 dB gain compression is +28 dBm minimum and VSWR is 1.5:1 in 50 ohm systems. Applications include front end amplifiers, high dynamic range IF amplifiers, instrumentation amplifiers, and multi-carrier signal amplifiers.
AML, Inc.
INFO/CARD #189

High Shock Survivability Crystals

Piezo Crystal Company has announced the development of quartz crystals with high shock survivability to 100,000 gravities. These crystals have many applications in the weapons and telemetry fields. The crystals can be manufactured in the range of 10 to 110 MHz and come in a TO-5 package.
Piezo Crystal Company
INFO/CARD #185

Microminiature Bandpass Filters

Center frequencies for these bandpass filters are 1227, 1381, and 1575 MHz. These filters, designed especially for GPS applications, feature extremely small size. Insertion loss is 5 dB maximum with VSWR of 1.5:1 maximum. 3 dB bandwidth of 50 MHz with a 3 dB to 40 dB shape factor of 5:1.
K&L Microwave Inc.
INFO/CARD #188

Spread Spectrum Generator

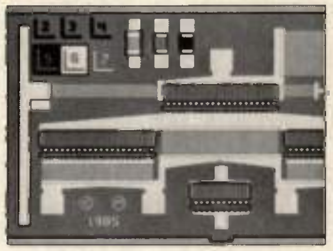
The LRS-200 generates pseudorandom sequences that are useful for developing and testing spread spectrum and conventional data communication systems. These sequences are generated by a configurable 32-stage shift register using linear feedback. Different modes of operation include BPSK, QPSK, GOLD/JPL, offset, and burst, with a maximum bit rate of 25 MHz.
New Wave Instruments
INFO/CARD #187

Semi-Flexible Coaxial Cable

Andrew Corporation announces the availability of a family of coaxial cable for high power HF, MF, and LF broadcasting. Transmitters of 500 kW operating up to about 30 MHz can be accommodated. It is furnished in nominal sizes of 7, 8, and 9 inches.
Andrew Corporation
INFO/CARD #186

Si MMIC Chips

The MSA-0500 and MSA-1000 chips feature high output power and low distortion. Their performance features include −23 dBm and +27 dBm power output at 1 dB gain compression, and +33 dBm and +37 dBm third-order intercept points at 1 GHz, respectively. The MSA-0900 chip provides 8 dB gain with ±0.2 dB



flatness from 0.1 to 4.0 GHz, and a 3 dB bandwidth of 0.1 to 6.0 GHz. The MSA-1100 chip includes a 3 dB bandwidth of 0.05 to 1.6 GHz.
Avantek
INFO/CARD #184

CATV Amplifiers

Power doubler types BGD106, BGD108 and BGD506 each contain two specially modified circuits allowing twice the output power without the need for additional matching transformers. The BGD508 power doubler is designed with three internal circuits, enabling significant user savings in cost and space.
Philips Components Discrete Products Div.
INFO/CARD #183

Dual Video Op Amp

A low cost general purpose dual 50 MHz unity-gain bandwidth op amp, the AD827, is stable driving any capacitive load and features 85 dB channel separation. Differential phase and gain errors are typically 0.19 degrees and 0.04 percent, respectively.
Analog Devices
INFO/CARD #182

Safety Capacitor Design Kit

Murata Erie North America is now offering a design engineering kit for safety capacitors. This kit includes over 375 capacitors ranging in value from 100 pF to 10,000 pF in a variety of voltages and tolerances.

Murata Erie North America
INFO/CARD #181

Couplers and Power Dividers

Surface mount directional couplers and in-phase power dividers have been introduced by Merrimac. The PDG-02B series of power dividers covers 10 to 2,000 MHz and is characterized by high performance with high isolation typically 30 dB or more. The CRG-02B series of directional couplers covers the frequency ranges of 5-500 MHz and 200-1,500 MHz with nominal coupling at 11 or 20 dB.

Merrimac
INFO/CARD #180

RFI Adapter

AVA Electronics has introduced a ground-filtered bulkhead BNC adapter to help deal with EMI and RFI problems on computer network lines. the adapter connects coaxial shields to ground via 10,000 pF of distributed capacitance. The capacitors are rated at 500 volts working and 1,000 volts peak.

AVA Electronics Corporation
INFO/CARD #179

5 to 1000 MHz SMT Limiter

The WJ-SML1 5 to 1000 MHz surface mount voltage variable limiter has been introduced by Watkins-Johnson Company. The limiter offers a phase response of 0.3 degrees/dB, typical, up to 160 MHz. The limiting level can be varied from −10 to 0 dBm, with a typical insertion loss of less than 2 dB.

Watkins-Johnson Company
INFO/CARD #178

Laser Diode Driver

Avtech Electrosystems introduces the Model AVO-9C-C pulsed laser diode driver which features a 0.3 to 2.0 nanosecond variable rise time option and an output current of up to 100 mA. The output pulse width is variable from 0.5 to 10 nanoseconds and the pulse repetition frequency is variable from 0 to 25 MHz.

Avtech Electrosystems
INFO/CARD #177

Data Carrier Filter

Narrow bandpass filter model 6779A selects a data carrier on a LAN network and suppresses nearby interfering carriers. Its center frequency is 74 MHz with a 1.5 MHz bandwidth and insertion loss is 5 dB maximum. 40 dB selectivity is approximately ±3 MHz.

Microwave Filter Company, Inc.
INFO/CARD #176

90 Degree Hybrids

LHQ series of stripline 90 degree hybrids covers the 0.5 to 2.5 GHz frequency range in octave bands. All models exhibit low insertion loss of 0.25 dB maximum, phase deviation from quadrature of 2 degrees maximum, isolation of 24 dB minimum, and VSWR of 1.2:1 maximum.

Norsal Industries, Inc.
INFO/CARD #175

Sampling Phase Detector

Alpha Industries announces a sampling phase detector used in PLL applications where the carrier frequency (1-20 GHz, typically) is locked to a crystal reference (5-500 MHz, typically). The detector is available in all RF drive levels.

Alpha Industries
INFO/CARD #174

Signal Processing For the Macintosh

BV Engineering's Signal Processing Program (SPP) is a general purpose analog signal generation, manipulation, and simulation program. It performs linear and non-linear time-domain waveform analysis, Fast Fourier Transforms, LaPlace transforms, transfer function analysis, statistics, and graphical display of signals, spectra, and transfer function data. It requires 512K of memory and runs on a floppy or hard-disk Mac, Mac Plus, Mac II or Mac SE systems under System 3.2 or later and Finder 5.3 or later. SPP is priced at $349.95.

BV Engineering Professional Software
INFO/CARD #210

High-Efficiency Amplifier Simulator

High Efficiency Power Amplifier Simulator (HEPA-SIM) computes the steady-state periodic time-domain waveforms in a single-ended switching-mode RF power amplifier. It also computes the DC input power, RF output power, all of the power dissipations, and the RF output spectrum. Hardware requirements are IBM PC, XT, AT, PS-2, or compatible with 384K of RAM, and the price is $595 to $995 depending on quantity.

Design Automation, Inc.
INFO/CARD #209

High Performance Circuit Simulator

CONTEC Microelectronics U.S.A. Inc. has released ContecSPICE, an integrated mixed-level, mixed-mode simulator that simulates circuits down to the path delays in PCB AND IC designs with digital clock frequencies of 40 MHz and greater. The model parameter generator, a companion program for ContecSPICE, comes with 2,500 device models containing GaAs MESFET models, power MOSFETs, and voltage-variable capacitance diodes.

CONTEC Microelectronics U.S.A. Inc.
INFO/CARD #208

GPIB Software from Fluke

The version 2.1 of TestTeam software provides enhanced real-time graphics, graphics, print capabilities for 150 different printers, and RS-232 X Modem Protocol. It also gives GPIB instrumentation system programmers and operators a complete environment for the generation and interactive use of application programs.

John Fluke Mfg. Co.
INFO/CARD #207

PMI Spice Diskette

PMI introduces PMISpice, a free diskette which contains models for low-noise and instrumentation amplifiers. Designers can simulate the total output noise and equivalent input noise of their circuits over the entire frequency range of interest. Included are SPICE models for the AMP-01 and AMP-02 instrumentation amplifiers.The PC-compatible diskette contains 54 complete model net-lists of PMI's op amps, instrumentation amps, and matched transistor pairs.

Precision Monolithics Inc.
INFO/CARD #206

Class E RF Power Amplifier Design Program

A new computer program from Design Automation, Inc., called HEPA-DES, now makes it easy to achieve designs quickly for single-ended, switching-mode, wide-band RF power amplifiers. It computes the component values for a user-specified output power and bandwidth, effects of the transistor finite switching times and parasitic resistances of all components. HEPA-DES requires an IBM or compatible computer with 384 KB of RAM.

Design Automation, Inc.
INFO/CARD #205

Harmonic Suppressing Digital Frequency Divider

By Mitch Randall
National Center for Atmospheric Research

A popular and economical technique for generating HF signals is to divide an octave coverage VHF signal by successive factors of two. In this way, octaves of frequency coverage can be added to a frequency synthesizer using simple digital dividers. Filtering is required to convert the digital divider output into a spectrally pure sine wave output. The absence of the second harmonic in the 50 percent duty cycle square wave output of the digital divider eases this filter design. However, the presence of a strong 3rd harmonic in the divider output still leaves a significant challenge for the filter designer.

A newly developed divider greatly simplifies these filtering requirements by suppressing the 2nd, 3rd, and 4th harmonics as shown in Figure 1. The new divider achieves this result by summing two square waves which are specially designed to cancel these harmonics. This divider offers significant advantages for use in successive division schemes.

In order to fully appreciate the benefits of the new divider it is instructive to first review a typical application of successive division as shown in Figure 2. Input to the unit is provided by a frequency synthesizer (not shown) covering an octave of bandwidth in the VHF range. A signal conditioner converts this sine wave input into a logic-compatible square wave. The frequency is then divided by a selectable power of two. The division ratio is determined by the number of divide-by-two stages that are digitally inserted prior to the output stage. A low pass filter removes unwanted harmonics from the 50 percent duty cycle square wave output of the successive division unit. In the example given, one low pass filter is required for each octave of coverage. Other schemes use two or more filters per octave. This eases the requirements for each filter at the expense of requiring more of them. In all, the two choices work out to be almost equal in complexity with the former slightly more attractive due to the less complicated RF switching needed. This successive division technique has many advantages. One relatively inexpensive digital divider, filter and switch is all that is required to gain an additional octave of coverage. Each additional octave has successively improved frequency resolution and phase noise characteristics. In addition, the output is leveled and flat with frequency; influenced almost solely by the flatness of the low pass filters and the RF switch.

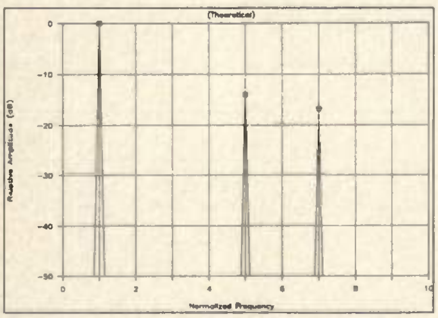


Figure 1. Divider output spectral components.

An Improved Divider Output Waveform

To ease the design of successive division synthesizers, a digital waveform which is easily generated and easily filtered to a sine wave is very desirable. A simple repetitive square wave of arbitrary duty cycle produces an infinite number of harmonics whose amplitudes are given by the well known fourier cosine series result:

$$A_n = \frac{2a}{n\pi} \sin\left(\frac{n\pi\tau}{T}\right) \qquad (1)$$

where n is the harmonic and a, τ, and T are as defined in Figure 3. Only one degree of freedom (τ /T) is available to optimize the simple square wave to ease the filtering problem. Mathematically stated, the desired condition is:

$$A_1 \neq 0, \quad A_n = 0 \quad \{n = 2,3,4,\ldots m\} \qquad (2)$$

Where m is as large as the solution allows. For the simple square wave, the largest m is 2 corresponding to the 50 percent duty cycle square wave (τ/T = 1/2). All solutions satisfying (2) for m = 3 ($A_2 = A_3 = 0$) also require $A_1 = 0$ which is unacceptable. To improve upon the 50 percent duty cycle square wave, an additional degree of freedom is needed. This is achieved by defining the output as the sum of two square waves. The condition for eased filtering is then stated as:

$$A_n = \frac{2a}{n\pi} \sin\left(\frac{n\pi\tau_1}{T}\right) + \frac{2a}{n\pi} \sin\left(\frac{n\pi\tau_2}{T}\right) = 0 \qquad (3)$$

$$\{n = 2,3,4,\ldots m\}$$

where the solution is chosen such that m is again maximized. Equation 3 is transcendental and cannot be solved directly. However, some insight into fourier series empirically leads to a solution for m = 4. Recall that functions with the property:

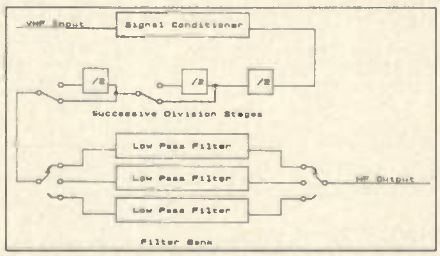


Figure 2. Typical successive division scheme.

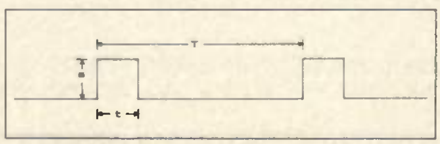


Figure 3. Pulse conventions.

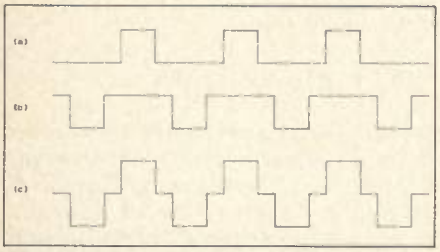


Figure 4. Divider waveforms a) 33 percent duty cycle squarewave, b) 66 percent duty cycle square wave, c) sum of (a) and (b).

$$f(t) = -f\left(t + \frac{T}{2}\right) \qquad (4)$$

contain only odd harmonics in a Fourier series because only odd harmonics satisfy equation 4. Stated another way, even harmonics repeat themselves during the second half of the fundamental cycle, therefore any correlation with the first half cycle would be exactly canceled. Also note that the sum of two equal amplitude digital waveforms can at most have 3 distinct levels. With this in mind, a waveform having equal positive and negative pulses about a centerline (satisfying equation 4) is therefore a likely candidate. This waveform will suppress all even harmonics. The 3rd harmonic can then be suppressed by choosing the pulse width to be T/3. This waveform can be expressed as in equation 3 by choosing $\tau_1/T = 1/3$ and $\tau_2/T = 2/3$ as shown in Figure 4a and 4b respectively. Figure 4c shows the overall waveform resulting from the summation of these two waveforms. The summation can also be expressed in the frequency domain as shown in Figure 5. The frequency domain offers an interesting viewpoint into the origins of the practical limitations as will be discussed later.

Reduced Filtering Complexity

The absence of the 2nd, 3rd, and 4th harmonics in the output spectrum of the new divider greatly simplifies the filter design in a given successive division scheme. To illustrate this, compare the filter requirements for each of two cases. As an example assume a 15 MHz to 30 MHz sine wave is required with harmonic levels less than −45 dBC. In the first case, a standard 50 percent duty cycle square wave from 15 MHz to 30 MHz is available at the divider output. The symmetrical output of the divide-by-two stage naturally suppresses the second harmonic, but the third harmonic amplitude is relatively high at about −10 dBC. The required filter must pass the fundamental which lies from 15 MHz to 30 MHz, but reject 3 × 15 MHz = 45 MHz by >35 dB. A Butterworth low pass filter of 10 poles would be required to meet these specifications (1). Compare this with the filtering requirements using the new divider. The new divider suppresses the 2nd, 3rd, and 4th harmonics leaving the fifth harmonic at about −14 dBc. The required filter must pass 15 MHz to 30 MHz but reject 5 × 15 = 75 MHz by 31 dB. This requires only a 4 pole Butterworth low pass filter which is considerably easier to adjust and reproduce.

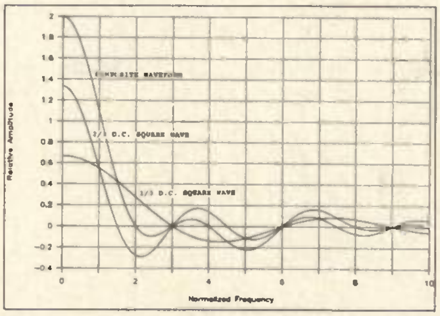


Figure 5. Divider output spectral envelope.

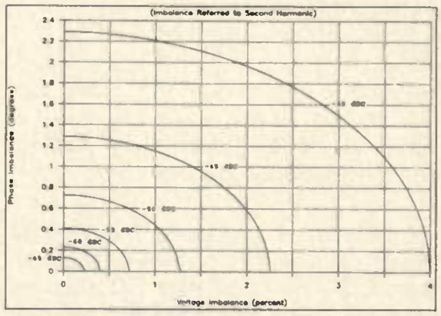


Figure 6. Spectral harmonic level. Imbalance referred to second harmonic.

Practical Considerations

It is implicit from previous discussions that the suppression of the even order harmonics depends on the voltage and phase balance between each of the two summed waveforms of the new divider. The second harmonic is of greatest concern because it is the largest harmonic in each of the constituent waveforms (refer to Figure 5). Therefore, its cancellation after summation is most sensitive to any amplitude and/or phase

	min	typ	max	
dv_{oh}/dT	1.2	1.3	1.5	mV/°C
dv_{oh}/dT	0.0	0.4	0.6	mv/°C

Table 1. ECL 10KH logic family temperature characteristics.

imbalances. In addition, the second harmonic is subject to less attenuation from the low pass filter. This is an especially important fact when generating signals near the lower part of the octave. In this case the second harmonic is adjacent to the band edge and may experience little if any attenuation from the low pass filter. It is worth noting that since the third harmonic is inherently suppressed in each of the constituents, its rejection is not affected by amplitude or phase imbalances. To investigate the effects of amplitude and phase imbalances consider the sum of two sine waves of equal frequency but slightly differing amplitude and phase:

$$\Sigma(t) = \cos(\omega t) + (1 + \delta)\cos(\omega t + \theta) \qquad (5)$$

Since we are only interested in $|\Sigma(t)|$, it

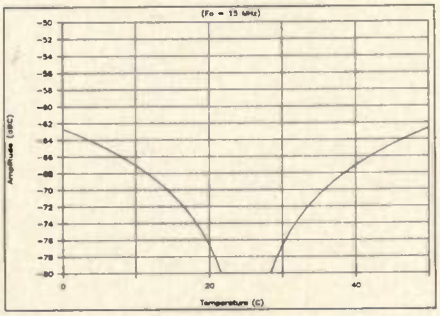


Figure 7. Typical second harmonic level.

is advantageous to express equation 5 in phasor notation as:

$$|\Sigma(t)| = |1 - (1 + \delta)e^{j\theta}| \qquad (6)$$

Taking the square root of the sum of the squares of the real and imaginary parts of equation 6 gives:

$$|\Sigma(t)| = [2(1 + \delta)(1 - \cos\theta) + \delta^2]^{1/2} \qquad (7)$$

To express the amplitude of the second harmonic relative to the fundamental, equation 7 must be scaled by a constant. Multiplying equation 7 by 1/2 accounts for the relative amplitude of the second harmonic in the constituent square waves. Another factor of 1/2 takes into account the doubling of the fundamental in the composite waveform. The expression for second harmonic level is thus:

$$dBC = 20\log(1/4\,[2(1 + \delta)(1 - \cos\theta) + \delta^2]^{1/2}) \qquad (8)$$

Contours of second harmonic level as given by this equation are shown in Figure 6 plotted against amplitude and phase imbalance (δ and Θ respectively).

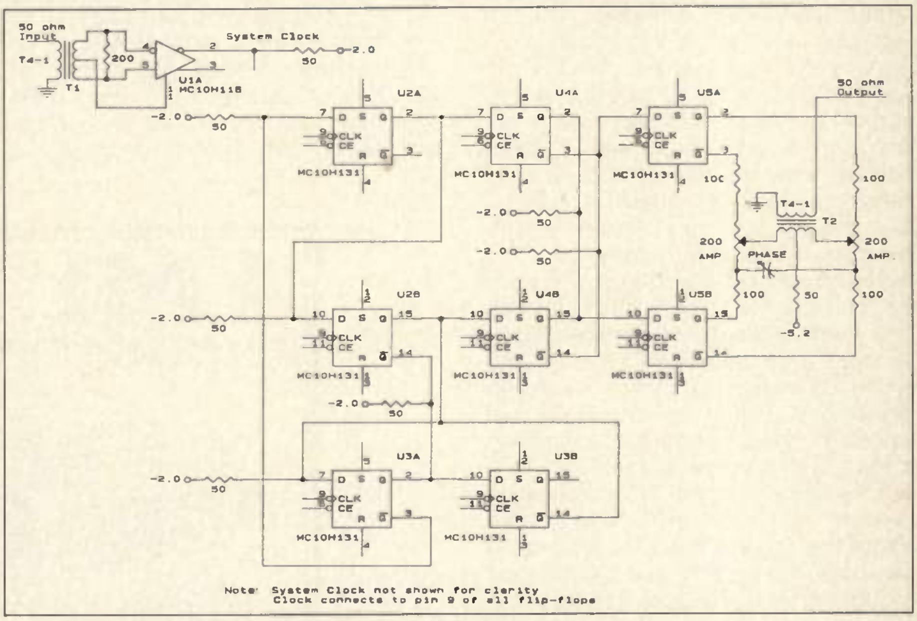


Figure 8. Schematic diagram of divide by 6 counter.

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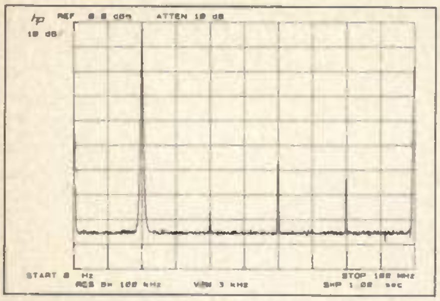


Figure 9. Measured output at F=20MHz.

Note that phase is referred to the second harmonic.

In any practical implementation, care must be taken to assure that proper amplitude and phase balance are maintained. The choice of a suitable logic family is therefore very important. It is desirable that the logic be very fast to allow operation over the greatest number of octaves. The family must also exhibit low phase jitter for good harmonic suppression and phase noise performance. In addition the output of the logic family must have good amplitude and phase delay characteristics over temperature for the suppression of the even order harmonics as discussed above. In considering these factors, the 10KH ECL logic family is an excellent choice. The 10KH ECL logic family is very fast, has excellent phase noise and phase jitter characteristics (2), and is stable and well characterized over temperature.

Amplitude and Phase Temperature Stability

Table 1 gives the temperature characteristics for the 10KH family (3). To analyze the worst case harmonic level temperature sensitivity we start by noting that the peak to peak output voltage of one output stage is given by:

$$V_a = V_{oh} - V_{ol} \tag{9}$$

The output imbalance is the difference between the voltage of each of two output stages, or

$$V_{bal} = V_a - V_b \tag{10}$$

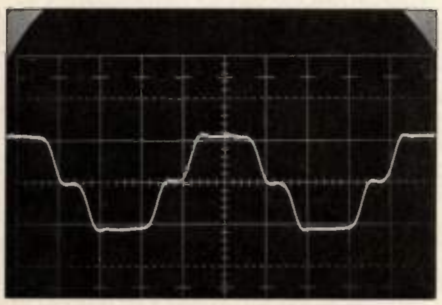


Figure 10. Time domain output (F=7MHz).

Band 1:	1.875 MHz - 3.75 MHz
Band 2:	3.75 MHz - 7.5 MHz
Band 3:	7.5 MHz - 15 MHz
Band 4:	15 MHz - 30 MHz

Table 2. Prototype test frequency bands.

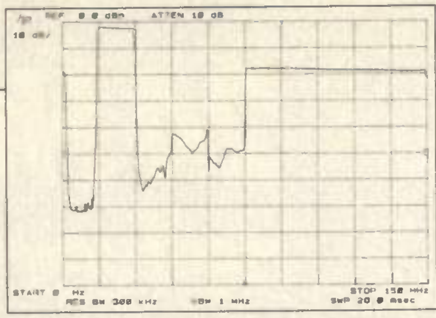


Figure 12. Measured harmonic level sweep.

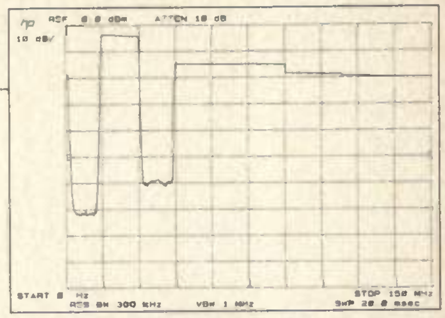


Figure 13. Harmonic level sweep of standard divider by 2.

The worst case imbalance occurs when the difference between the temperature coefficient of V_a and V_b is greatest. Note that all the temperature coefficients in Table 1 are positive. The worst case is thus given by:

$$\left(\frac{dV_{bal}}{dT}\right)_{max} = \left(\frac{dV_a}{dT}\right)_{max} - \left(\frac{dV_b}{dT}\right)_{min} \tag{11}$$

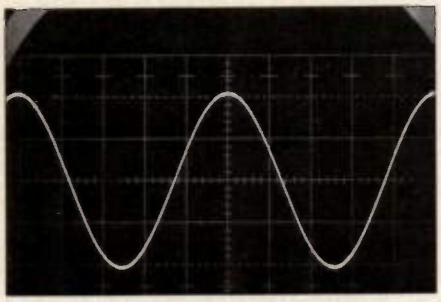


Figure 11. Filtered output waveform (F=1/2Fc)

Where:

$$\left(\frac{dV_a}{dT}\right)_{max} = \left(\frac{dV_{oh}}{dT}\right)_{max} - \left(\frac{dV_{ol}}{dT}\right)_{min} \tag{12}$$

and

$$\left(\frac{dV_b}{dT}\right)_{min} = \left(\frac{dV_{oh}}{dT}\right)_{min} - \left(\frac{dV_{ol}}{dT}\right)_{max} \tag{13}$$

Using the values from Table 1:

$$\left(\frac{dV_a}{dT}\right)_{max} = 1.5 - 0.0 = \frac{1.5\text{mV}}{°\text{C}} \tag{14}$$

$$\left(\frac{dV_b}{dT}\right)_{min} = 1.2 - 0.6 = \frac{0.6\text{mV}}{°\text{C}} \tag{15}$$

so that

$$\left(\frac{dV_{bal}}{dT}\right)_{max} = 1.5 - 0.6 = \frac{0.9\text{mV}}{°\text{C}} \tag{16}$$

or

$$\left(\frac{dV_{bal}}{dT}\right)_{max} = \frac{0.1\%}{°\text{C}} \tag{17}$$

However, this worst case analysis assumes 4 parameters to be at the extremes of their ranges. If we assume that the published minimums and maximums correspond to ± 3 standard deviations on a gaussian distribution of yield, then this value of $(dV_{bal}/dT)_{max}$ represents ± 6.4 standard deviations.

This simply means that this worst case is a very unlikely event. Also if the design uses output stages which are on the same die, the likelihood of this coincidence further diminishes. These facts lead the way to this author's opinion that a worst case value of 0.9 mV/C is grossly conservative (by more than a factor of 5).

Propagation delay variations as a function of temperature must also be considered for even order harmonic suppression since these variations correspond directly to phase imbalance. This information is not explicitly available, but one can deduce from indirect published data (3,4) that it is no greater than 1 ps/°C max. (For reasons similar to those above, the author believes this too is grossly conservative.) At 30 MHz this is equal to 0.011 deg/°C.

Second harmonic levels can be calculated using equation 8 taking into account the effects of both amplitude and phase temperature sensitivities. Figure 7 shows typical expected second harmonic levels using values of temperature stability 1/10 those calculated above to approximate what would likely be a typical temperature dependance.

Circuit Description

The schematic of a divider implemented in 10KH ECL logic is shown in Figure 8. The sinusoidal input signal is passed through the phase splitting transformer T1 into a matched load in parallel with the relatively high impedance of an ECL line receiver input (U1). The line receiver's truly differential inputs act as a zero crossing detector to reduce AM to PM conversion for optimum phase noise performance. The output of U1 provides ECL compatible input to the rest of the circuit. (In a typical successive division application, this ECL signal would first pass through the programmable $\div 2^n$ counter before continuing on to the main divider.)

The heart of the divider is the $\div 6$ counter formed by U2 and U3. Engineers not familiar with ECL circuit design should take note that ECL outputs can be wire-ORed. This feature is used extensively in the divider to avoid unnecessary gate delays that would otherwise limit the maximum clock frequency. Each of the outputs of U2 is a 50 percent duty cycle square wave. However, the two outputs are shifted in phase by 60 degrees with respect to one another. U4 uses wired-ORed outputs to combine these signals into two 33 percent duty cycle signals shifted in phase by 180 degrees.

The output stages are provided by a single chip (U5) to guarantee that amplitude and phase changes with temperature are balanced to minimize harmonic suppression variability. In addition, U5 provides differential outputs to increase the output drive and allow the use of a balanced RF combiner. The outputs are summed by the RF combiner consisting of the 4:1 transformer (T2) and a resistor network. The network presents a 200 ohm matched impedance across the combiner transformer. A 50 ohm resistor at the transformer center tap match-terminates the common mode portion of the combined square waves which does not pass through to the output. Therefore, each of the relatively low impedance ECL outputs sees a constant 267 ohm impedance. Output power from the combiner is defined by equation 2 for a = 0.9/4 V resulting in a predicted value of −2.1 dBm. Two trim pots provide flexibility to null out any amplitude imbalances. Since one of the four possible logic combinations of the two constituent square waves is ruled out by their mutual phase relationship, the adjustment of the two trim pots is redundant. However, two trim pots were deliberately used to maintain electrical balance and layout symmetry. In practice, only one pot needs to be adjusted to balance the output waveform; the other can remain at center scale. A single trim cap allows for cancellation of the second harmonic under the worst case operating condition. Since the capacitor can introduce only a lagging phase shift, it is necessary to first determine which output driver leads. A two capacitor scheme could be designed so that this step would not be necessary. However, the arrangement used was satisfactory for the prototype.

Another possible scheme for trimming the second harmonic could incorporate varactor diodes in a feedback loop servoed on the second harmonic. The advantage of a servo trim circuit such as this could be greater than what is apparent at first glance. A schematic is included in Appendix - of a prototype ÷3 counter which generates the desired waveform yet was verified to operate at clock frequencies of over 270 MHz. This corresponds to an output frequency of 90 MHz. The design was not included in the main text, though, because the second harmonic suppression was found to be too frequency sensitive to be practical.

Prototype Performance

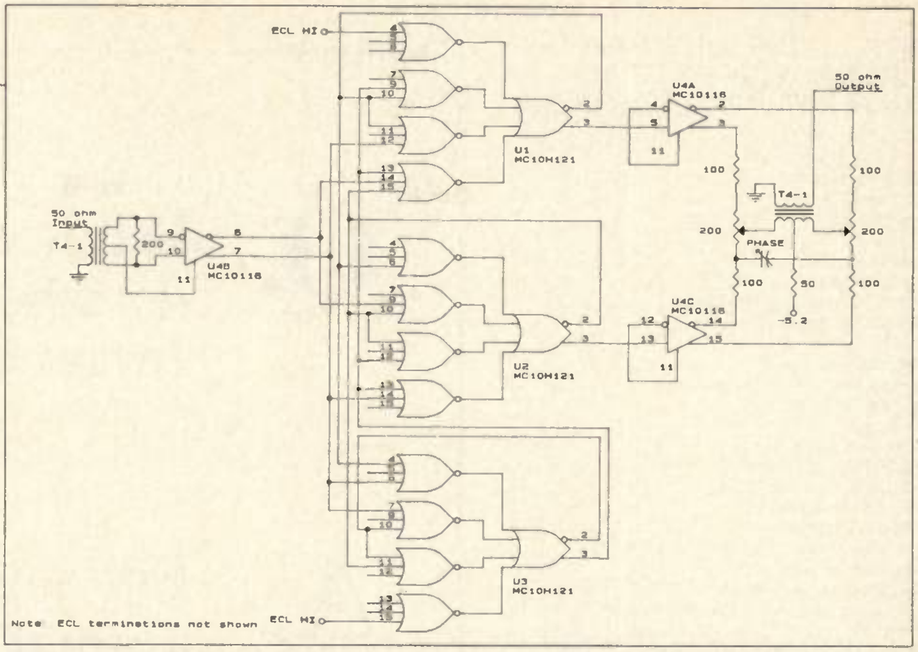
Prototype performance was tested over the four bands in Table 2 to evaluate expected divider performance for a 2 - 30 MHz signal generator. Adjustments were made using the assumption that filtering would be accomplished using 5th order 0.25 dB Chebyshev low pass filters to achieve better than −55 dBC harmonic levels. However, note that no filters were used for any of the following measurements except where specifically noted.

Divider performance matches well with theory as seen in the output spectrum of Figure 9 taken at a frequency of 20 MHz. Amplitude and phase trim adjustments allow nulling of the second harmonic to better than −75 dBC. Harmonic suppression much beyond this becomes very frequency sensitive, particularly when operating at the upper frequencies. Figure 10 shows the time domain waveform at an output frequency of 7 MHz. This figure is useful in showing intuitively why the divider output requires only simple filtering to produce a pure sine wave. Figure 11 shows the waveform after filtering with a four pole Butterworth filter. The test frequency was chosen to be one octave below the filter cutoff for demonstration purposes.

Figure 12 shows typical harmonic suppression over the octave from 15 MHz to 30 MHz which represents the worst case filtering scenario. The figure was generated by sweeping the input over an octave and observing the output with a spectrum analyzer in the peak hold mode. Performance was observed to be consistently better in the lower octaves of operation (see Table 2). This is attributed to the fact that the logic delay mismatches, which are fixed, correspond to smaller and smaller phase imbalances as the frequency decreases. Figure 12 also serves to illustrate the greatly simplified filtering requirements offered by this technique. For comparison, Figure 13 shows a sweep over the same output band created by a standard 50 percent duty cycle square wave divider.

Power output shows good agreement with theory at a nominal value of −2.4 dBm over the entire 2 to 30 MHz range. Fifth harmonic level is also very close to the predicted level of −13.9 dBC.

To achieve the best performance, the amplitude and phase trims were adjusted so that the second harmonic passes through its null somewhere near the highest specified fundamental frequency (2F = 30 MHz). This lessens the worst case condition when the second harmonic is not subject to much attenuation by the low pass filter. It is acceptable for the second harmonic to get larger with frequency for this case so long as it does so with less slope than that of the low pass filter.



Appendix 1. 270 MHz divide by 3 counter with desired waveform output.

Conclusion

A digital ÷6 counter has been introduced, the output of which naturally suppresses the 2nd, 3rd, and 4th harmonic. This offers a great advantage in the design of successive division synthesizers. An analysis of the dependencies on amplitude and phase balance demonstrates that a practical divider can be implemented in 10KH ECL logic with guaranteed harmonic suppression over a reasonable temperature range. A prototype divider has been implemented in ECL and shows excellent agreement with predictions. Custom or semi-custom VLSI technology could also be used to implement an entire successive division unit on a single chip.

Acknowledgment

This work was sponsored by the National Center for Atmospheric Research, Boulder, Colorado. The National Center for Atmospheric Research is supported by the National Science Foundation. *RF*

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3. W. R. Blood, Jr., *MECL System Design Handbook*, Compiled by the Computer Applications Engineering Department, Motorola, Fourth Edition, 1988.
4. MECL DEVICE DATA, Prepared by Technical Information Center, Motorola, 1987.

About the Author

Mitch Randall is an RF design engineer in the Atmospheric Technology Division of the National Center for Atmospheric Research, 1850 Table Mesa Dr., Boulder, CO 80303. He holds a BSEE from Iowa State University and an MSEE from the University of Colorado. Mitch is currently working on the ELDORA project, an airborne x-band dual doppler weather radar. He can be reached at (303) 497-2059.

PROGRAM AT A GLANCE

RF expo EAST
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TUESDAY, NOVEMBER 13 — Exhibits Open 11:00 a.m. - 6:00 p.m.

A-1 Receivers I (Tutorial)	8:30-11:30	• Design of Receivers for Electronic Warfare	S. R. Vincent, *Raytheon*
A-2 Power Amplifiers	8:30-9:30	• VSWR Performance of Transistor RF Power Amplifiers	R. W. Brounley, *Brounley Engineering*
	9:30-10:30	• A Novel Technique for Analyzing High-Efficiency Switched-mode Amplifiers	K. Siwiak, *Motorola*
	10:30-11:30	• RF Power Transistors for 200-W Multicarrier Cellular Base Station	K. Vennema, *Phillips*
A-3 PLLs and Synthesizers	8:30-9:30	• The "Approximation Method" of Frequency Synthesis	A. D. Helfrick, *TIC*
	9:30-10:30	• Digitally Derived Swept-frequency Source for Chirped-pulse Radar Altimeters	G. Whitworth, *Johns Hopkins University APL*
	10:30-11:30	• Ten-bit 300-MHz DAC for Direct Digital Synthesis	P. Jordan, *Analog Devices*
A-4 Microwaves	8:30-9:30	• System Design Considerations for Line-of-Sight Microwave Radio Transmission	G. M. Kizer, *Rockwell* (Dallas)
	9:30-10:30	• Phase Shifter Based Upon Reflectively Terminated Multiport Coupler	M. H. Kori, *Centre for Development of Telematics*
	10:30-11:30	• Instrumentation for Radar Reflectively Measurement	D. J. Kozakov, C. W. Sirles, D. A. Thompson, and R. S. Banks, *Millimeter Wave Technology*
B-1 Receivers II (Spread Spectrum)	1:30-2:30	• Spread Spectrum with Digital Signal Processing	R. J. Zavrel, Jr., *Stanford Telecom*
	2:30-3:30	• Receivers for GPS and GLONASS Spread-spectrum Navigation Systems	J. Danaher, *Structured Systems*
	3:30-4:30	• Detection and Sorting of Frequency-hopped Signals	J. E. Dunn and S. P. Russell, *Iowa State University*
B-2 MMICS	1:30-2:30	• GaAs MMIC Switch Products—Daily Applications	M. D. Smith, *ANZAC*
	2:30-3:30	• Silicon MMICs: 35 dB - 35 dBm - $35	J. Walsh, *SGS-Thomson*
	3:30-4:30	• Some Design Considerations for L-band Power MMICs	R. Weber, *ISU Microelectronics Research Center*
B-3 Filters	1:30-2:30	• Capabilities and Applications of SAW-coupled Resonator Filters	A. Coon, *RF Monolithics*
	2:30-3:30	• At Long Last: Modular, Digitally Tuned RF Filters as Easy as Amplifiers and Mixers	E. A. Janning, *Pole-Zero*
	3:30-4:30	• Narrow-band SAW Filters for IF Applications	B. Horine and S. Gopani, *Sawtek*
B-4 Antennas and Propagation	1:30-2:30	• Large Loop Antennas	R. P. Haviland
	2:30-3:30	• VHF Multipath Propagation: Causes and Cures	D. R. Dorsey, *DocSoft*
	3:30-4:30	• Radio-frequency Identification Systems for Commercial and Industrial Applications	J. Eagleson, *Allen-Bradley*

WEDNESDAY, NOVEMBER 14 — Exhibits Open 11:00 a.m. - 6:00 p.m.

C-1 Receivers III (Digital)	8:30-9:30	• IF Frequency-response Considerations for the Digital Radio Environment	R. Roberts, *Harris*
	9:30-10:30	• DSP Demodulation	S. F. Russell, *Iowa State University*
	10:30-11:30	• A DSP PSK Modem for SATCOM SCPC Voice/Data	Y. S. Rao, R. Asokan, and K. Reeta, *Centre for Development of Telematics* (India)
C-2 Transmitters	8:30-9:30	• Architecture of HF-VHF Radio Transmitters	M. A. Sivers and S. V. Tomashevitch, *Leningrad Electrotechnical Institute of Communications*
	9:30-10:30	• Digital-feedback Techniques for a Pulse-width-modulated Power Supply for RF Power Amplifiers	H. Direen, *ETO*
	10:30-11:30	• Increasing the Efficiency of SSB Transmitters by Envelope-tracking RF Power Amplifier	L. Voronov, *Leningrad Electrotechnical Institute of Communications*
C-3 SAW Tutorial	8:30-11:30	• Surface Acoustic Wave (S.A.W.) Technology	C. A. Erikson, Jr., *Oakmont*
C-4 RF Systems for Research in Particle Physics	8:30-9:30	• RF Systems for the Advanced Photon Source	J. F. Bridges, *Argonne National Laboratory*
	9:30-10:30	• RF Applications in Particle Accelerators	C. Hovater, *Continuous Electron Beam Accelerator Facility*
	10:30-11:30	• A Fully Digital RF-signal Synthesis and Phase Control for Acceleration in COSY	H. Meuth, *Forschungszentrum KFA Julich* (FRG)
D-1 Receivers IV (Applications)	1:30-2:30	• Dynamic Evaluation of High-speed, High-resolution D/A Converters	J. Colotti, *CSD/Telephonics*
	2:30-3:30	• A High-performance Low-cost TV Demodulator	P. R. M. Correa (BRAZIL)
	3:30-4:30	• Digital-signal-processing-based Spectrum-monitoring System for the European Broadcasting Area	I. Novak, *Design Automation, Technical Univ of Budapest*
D-2 Components	1:30-2:30	• Impedance-matching Transformers for RF Power Amplifiers	D. N. Haupt, *Erbtec*
	2:30-3:30	• Temperature-compensating Attenuators	P. F. Hamlyn, *ANZAC*
	3:30-4:30	• Using Current Feedback Amplifiers in High-frequency Active-filter Applications	A. Neves, *PMI*
D-3 Quartz Crystals and Applications	1:30-2:30	• Vibrational Sensitivity and Phase Noise in Crystal Oscillators	G. Kurzenknabe, *Piezo Crystal*
	2:30-3:30	• Quartz-crystal Filters: A Review of Current Issues	M. D. Howard and R. C. Smythe, *Piezo Technology*
	3:30-4:30	• Frequency Correlation of Quartz-crystal Resonators	B. Long, *Piezo Crystal*
D-4 Modulation	1:30-2:30	• New Method of Linear Amplitude Modulation	L. Minggang, G. Liangcai, and Z. Suwen, *Wuhan University* (PRC)
	2:30-3:30	• 4-GHz Multiplied Source for Digital Modulation	D. Balusek, *Rockwell* (Dallas)
	3:30-4:30	• Techniques in Voice Compression and Synthesis	P. G. Beaty, *Erbtec*

THURSDAY, NOVEMBER 15 — Exhibits Open 10:00 a.m. - 2:00 p.m.

E-1 Noise Tutorial	8:30-11:30	• Noise Fundamentals	F. H. Perkins, Jr., and A. Ward, *RF Monolithics, Avantek*
E-2 CAD and Simulation	8:30-9:30	• Spice Modeling and Simulation of an 800-MHz, Class-AB Push-pull Amplifier	R. Y. LaLau, *Mobile Data*
	9:30-10:30	• Computer Programs Design and Optimize High-efficiency RF Power Amplifiers	N. O. Sokal and I. Novak, *Design Automation*
	10:30-11:30	• A Quasi-linear Determination of UHF Power Device Operation from a Spice-simulated Nonlinear BJT Model	P. E. D'Anna, *MMD*
E-3 PLLs and Synthesizers II	8:30-9:30	• Designing with Direct Digital Frequency Synthesizers	F. A. B. Cercas, A. A. Albuquerque, and M. Tomlinson, *Instituto Superior Tecnico* (Portugal)
	9:30-10:30	• Direct-digital Waveform Generation Using Advanced Multi-mode Digital Modulation	B-G. Goldberg, *Sciteq*
	10:30-11:30	• Optimum PLL Design for Low Phase-noise Performance	S. Goldman, *E-Systems*
E-4 Test and Measurement	8:30-9:30	• Design and Development of an RF Data Acquisition System	T. H. Jones and M. A. Belkerdid, *Univ of Central Florida*
	9:30-10:30	• Handheld Probing Techniques for RF-PCB and Hybrid-circuit Characterization	Y. D. Kim, *Hewlett-Packard*
	10:30-11:30	• Using the VXI Bus for RF Test Equipment	M. Levy, *Racal-Dana*

RF literature

Video Graphics System Design for Reduced EMI

This application note from Analog Devices is a guide to the design of a video graphics system in terms of EMC. This note explains EMC and the techniques of reducing radiated emissions and EMI in the design of high-speed video graphics.
Analog Devices
INFO/CARD #230

High Intercept Low Noise Amplifier Handbook

This handbook contains information for effective two-tone intermodulation distortion test systems, cancellation techniques, and suppressing second harmonics, performance plots, tabular data, and outline dimensions. Video, RF and microwave products from AML are included.
AML, Inc.
INFO/CARD #227

RF Components Catalog

This catalog incorporates information on the ULTRAMIN[R] miniature filters along with commercial grade and tunable filters. Also included is information on programmable, fixed, and step attenuators, detectors, matching pads and DC blocks. Formulas, graphs, packaging outlines, and environmental capabilities are also contained.
Wavetek RF Products, Inc.
INFO/CARD #226

Crystal Oscillators

This new catalog includes technical specifications on Murata's DCXOs, VCXOs, TCXOs, OCXOs, and discrete and hybrid crystal oscillators. New products featured include low cost TCXOs and VCXOs and high stability, small size DCXOs.
Murata Erie North America
INFO/CARD #220.

New Devices Catalog

Alpha Industries has released a catalog containing information about mixer and detector diodes, control and generating devices, parametric amplifier varactors, and beam-lead and chip capacitors. Also included is a new line of MMIC devices, including FETs, amplifiers, and switches.
Alpha Industries
INFO/CARD #228

EMC Accessories Catalog

Products described in Hewlett-Packard's EMC Accessories Catalog include transducers and accessories, HP 11966 series antennas, HP 11967 series conducted EMC accessories, and HP 11968 series EMC positioning accessories. An EMC accessory application guide is also included.
Hewlett-Packard Company
INFO/CARD #225

RF/IF Signal Processing Guide

Detailed specifications, computer-automated performance data, and performance curves for Mini-Circuits do to 5 GHz signal-processing components are contained in this guide. Products covered are mixers, power splitter/combiners, monolithic and power amplifiers, RF transformers, GaAs switches, and filters.
Mini-Circuits
INFO/CARD #221

Advertising Index

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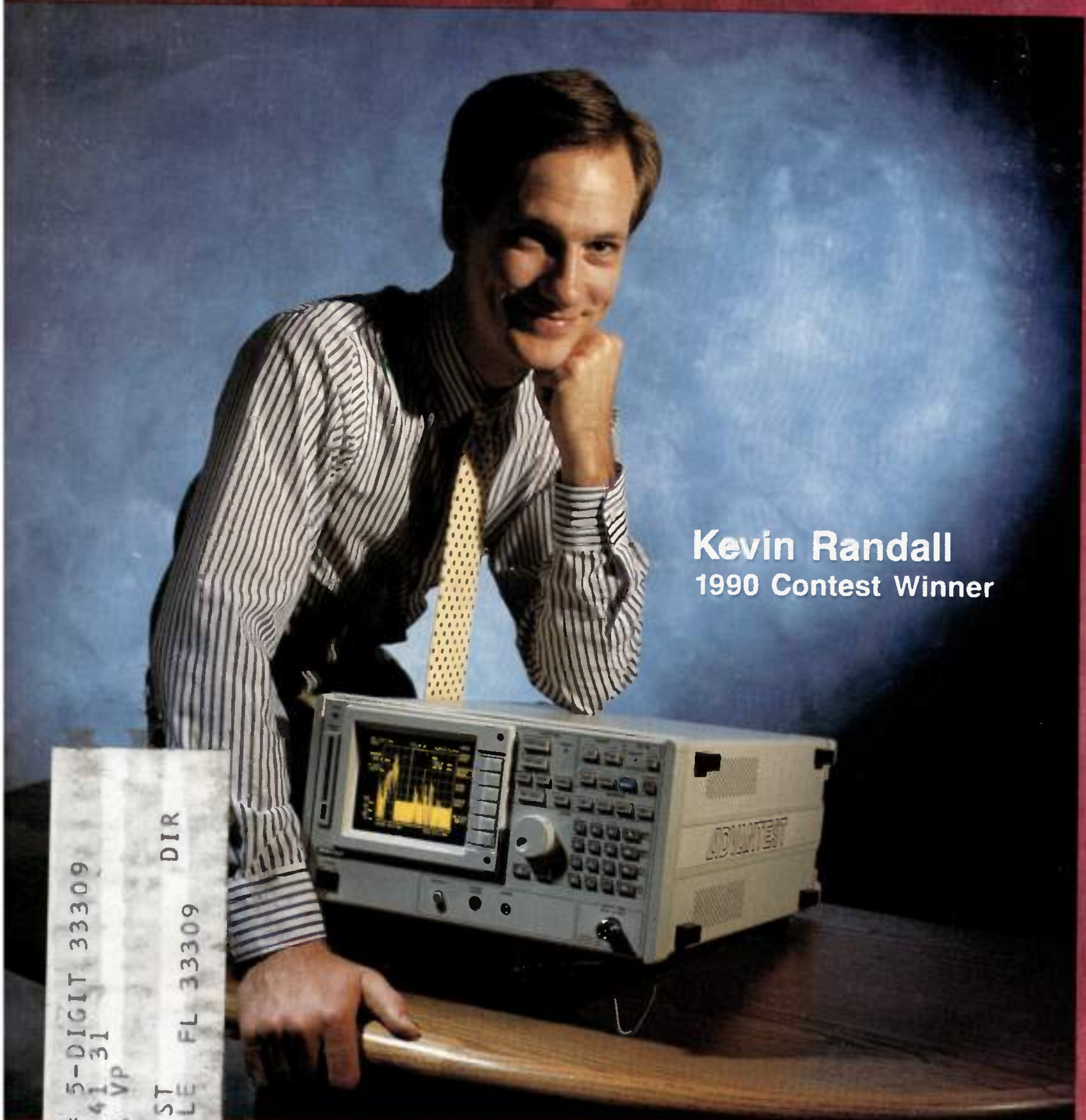
INFO/CARD 62

RF design

engineering principles and practices

July 1990



Kevin Randall
1990 Contest Winner

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The Winning Design

plus
VHF Crystal Oscillators
Power Amplifier Report

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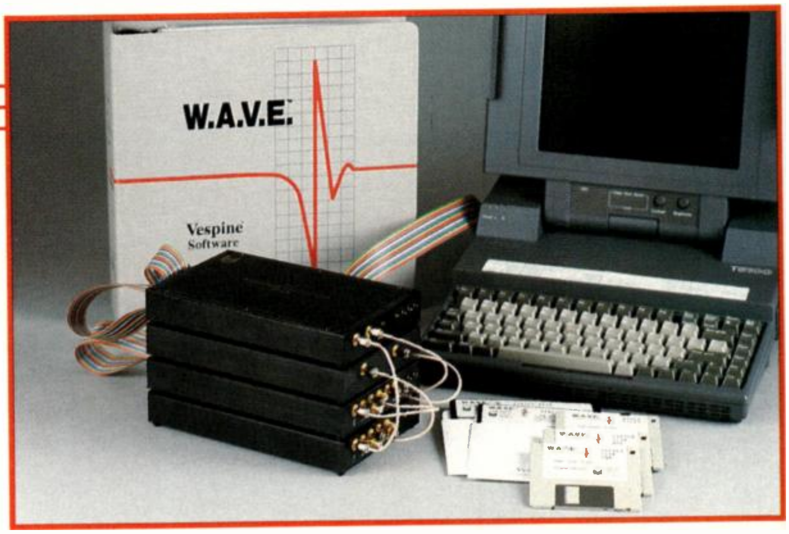

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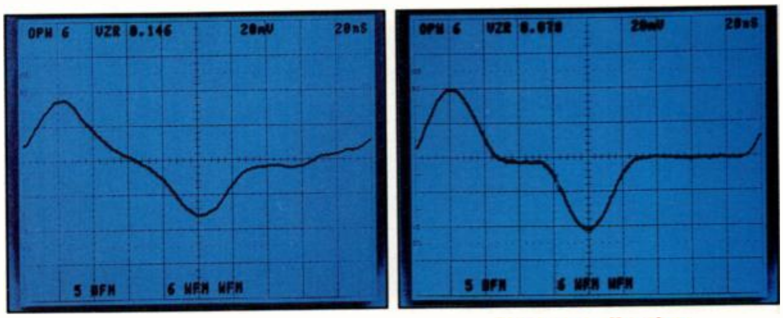

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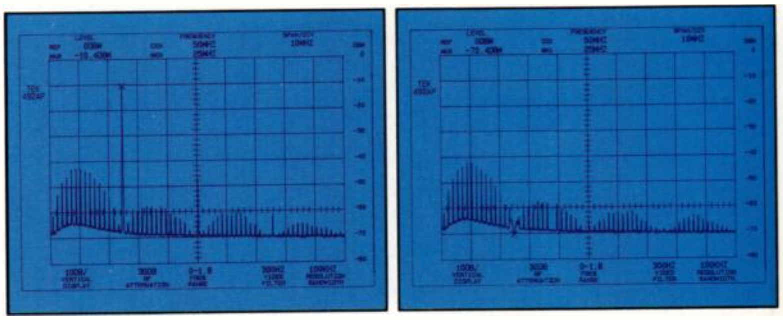

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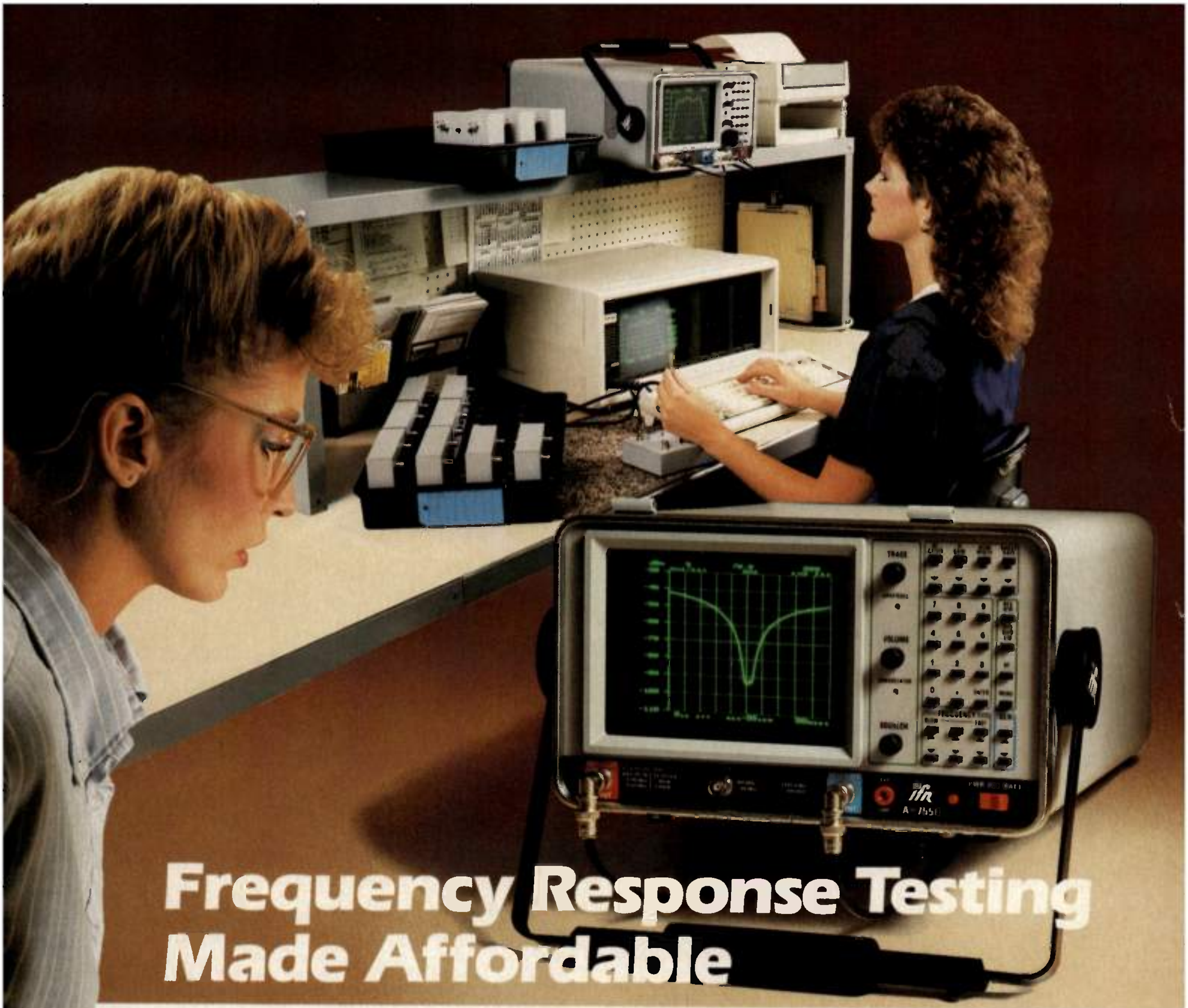

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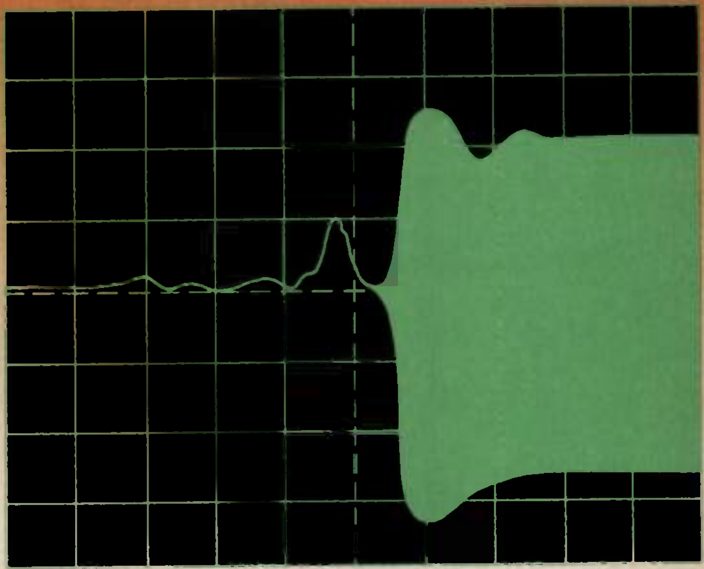
Transition Time		.7	1	nS	90%/10% or 10%/90% RF
Switching Speed		7	10	nS	50% TTL to 90/10% RF
Insertion Loss		1.7	2.3	dB	
Isolation	60	70		dB	10-100
	50	60		dB	100-200
Operating Frequency	10		200	MHz	
TTL Controlled					



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INFO/CARD 3

Congratulations to Mr. Ed Bialek, Section Head, Microwave Design, Amherst Systems, Buffalo, NY, for winning the DAICO MTTS computer giveaway.

RF design

contents

July 1990

design awards

cover story

emc corner

departments

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RF editorial

Lots of Good Ideas

By Gary A. Breed
Editor

Great ideas are in abundance this summer! Of course, the main reason for this observation is the 1990 RF Design Awards Contest. The winners — from Grand Prize to Honorable Mention — are all highlighted in this month's cover story. The top entry, a "quiet switch," submitted by Kevin Randall of Interspec, Inc., is published in this issue, along with one of our Honorable Mention winners, a PLL-DDS hybrid design by Rob Gilmore. Read all about the winners on page 47, then look over these two entries.

Reflecting on the contest, the judges (Consulting Editor Andy Przedpelski, last year's winner Dan Baker, and your Editor) think that this is the best collection of ideas yet. Engineers are a humble sort, and most of the entries didn't really explain how their authors solved difficult problems. After talking to the top prize winners, I soon realized that their entries represented key developments for their companies. We are pleased to have a total of 13 prize packages to present this year, since at least that many entries deserve special notice.

The secretary for the engineering department at Interspec called to make sure we supplied their firm with complete information on Kevin Randall's prizes and publicity. She wanted to be certain that we understood how important his achievement was to them, and she knew that he wouldn't make a big deal of it. I've seen it many times over the five years of our contest — the winners are genuinely humble, but understandably pleased and satisfied that their accomplishments are recognized. So it's our job to make a big deal of it! We hope that our contest provides some well-deserved recognition for excellence in RF design engineering.

Between now and next July's report on the 1991 contest, we will publish all 13 prize winners, and at least two or three other interesting entries. We have gotten plenty of encouragement to continue publishing these "great ideas" every month, and we will certainly keep 'em coming.

Congratulations to all the engineers who entered our contest, and especially to those who were selected as winners. Your contributions have not gone unnoticed.

An RF Expo East Report

Under the guidance of Program Chairman Dr. Fred Raab, the Technical Program for RF Expo East is shaping up extremely well. International participation is high, with proposals from the Soviet Union, People's Republic of China, United Kingdom, Hungary, India, and Korea. Hopefully, all of these engineers will be able to complete their papers and receive approval to travel to Orlando in November.

A list of papers that are likely to be part of the RF Expo East program is contained in the show announcement on page 66. It is very early in the process, so some of these papers may not be able to be presented, but the majority will be included in the best RF Expo Technical Program we have ever had. Make sure Orlando is on your schedule for November 13-15 — it's going to be great!

RF design

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Main Office:
6300 S. Syracuse Way, Suite 650
Englewood, CO 80111 • (303) 220-0600
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Liane Pomfret

Assistant Editor
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News From ATC

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0.5		4.7	C	27	
1.0		5.6		33	J
1.2		6.8		39	
1.5	C	8.2		47	
1.8		10	J	56	
2.2		12		68	K
2.7		15		82	
3.3		18		100	

Tolerance Code: B = ±0.1pF; C = ±0.25pF; J = ±5%; K = ±10%

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RF letters

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

Optimum Lossy Broadband Matching Networks Corrections

In our March issue we made several mistakes in F.J. Witt's article "Optimum Lossy Broadband Matching Networks for Resonant Antennas" and would like to correct these errors. On page 46 there are three errors in column two. In the second paragraph down, the normalized bandwidth should be B_{Nref} and then in the paragraph below that it should read "By setting $dB_N(S_L)/dS_L = 0$, the value of S_L, S_{Lopt}..." Then on page 47, there are corrections to column 3. $\Delta = 0$ is uppercase and not lowercase as shown. At the bottom of that paragraph it should be $dB_N(Z_N)/dZ_N = 0$. In the paragraph below that $B_{Nref} = 1/\sqrt{2}$ and $B_{Nmax} = \sqrt{3}$. Equation 24 should be written as:

$$L_{MNE} = 10 \log \left\{ 1 + \frac{Q_O}{Q_N} \left[\frac{Q_O}{2Q_N} \left(1 - \frac{1}{S_M^2} \right) + 1 \right] \right\} \qquad (24)$$

On page 48 in column one, equation 26 should read as follows

$$S_{Mmin} = ((B_N^2 + 1)^{1/2} + (B_N^2 + 1 + (2Q_O/Q_N)(1 + Q_O/2Q_N))^{1/2})/2(1 + Q_O/2Q_N) \qquad (26)$$

Equation 28 should read as follows

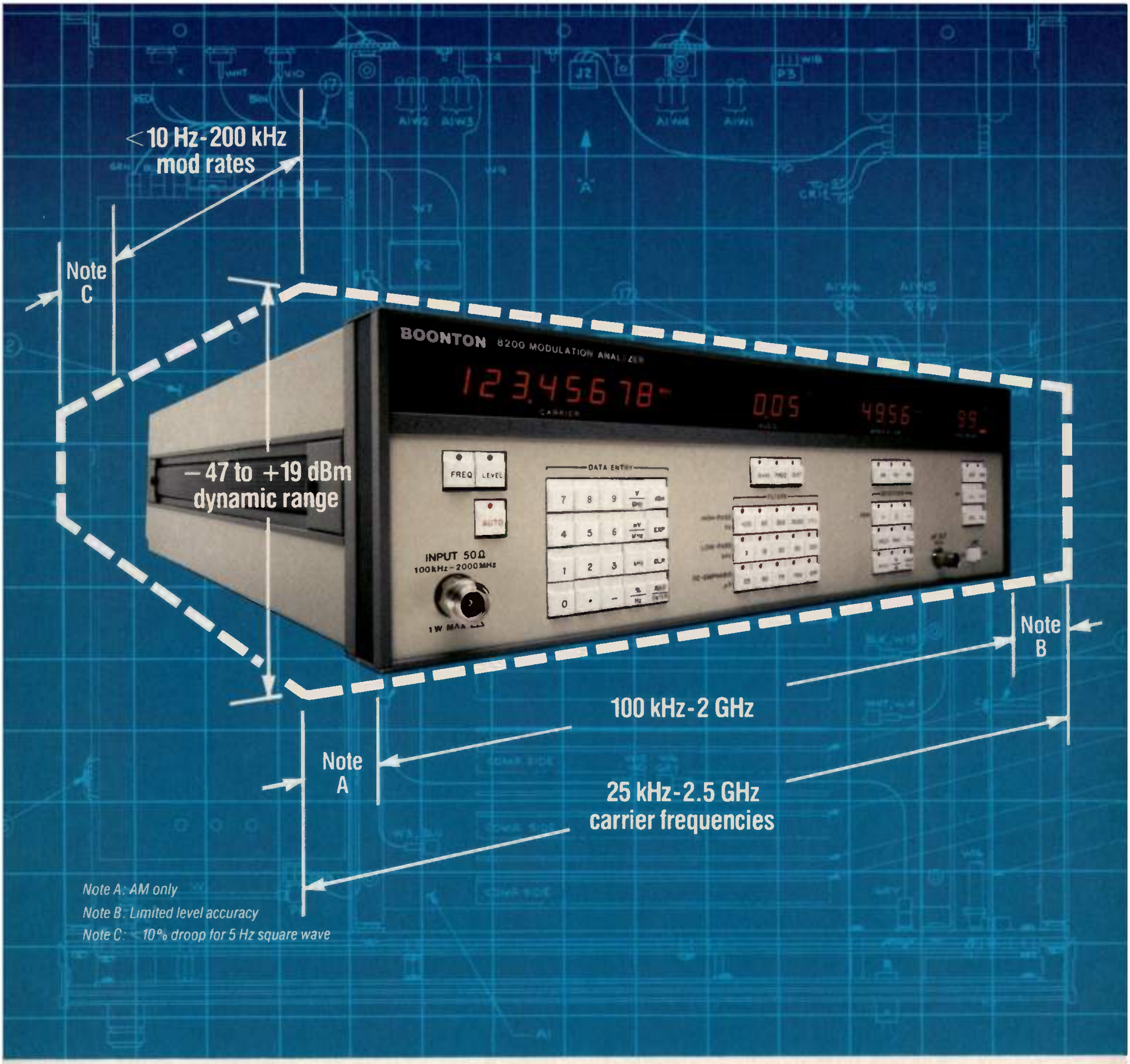
$$B_N = (S_M - 1)^{1/2} \left\{ 2 + \frac{Q_O}{Q_N} \left[2 + \left(1 + \frac{Q_C}{Q_N} \right) \left(1 - \frac{1}{S_M} \right) \right] \right\}^{1/2} \qquad (28)$$

In the second column, about three-quarters of the way down it should read "it is necessary to set $\Delta = 0$."

On page 49, at the bottom of the second column, the sentence should be "Also shown for comparison, (Figure 12(b)), is the SWR..."

Finally, on page 50 there are two more errors. The first is at the bottom of the first column. The sentence should be "Currents flowing on the inside of the resonator..." and then in the middle of the second column it should read "Hence, $Q_N = 42.0$, leading to the following results."

We apologize to Mr. Witt and our readers for the mistakes made in the article. Mr. Witt can be reached at AT&T, Digital Terminals and Microwave Technology Department, 1600 Osgood Street, North Andover, MA 01845 for those readers who wish to reach him by mail.



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INFO/CARD 10

RF calendar

July

26-29 **The 1990 International Tesla Symposium**
Hilton Inn, Colorado Springs, CO
Information: International Tesla Society, 330-A West Uintah Street, Suite 215, Colorado Springs, CO 80905-1095.

August

13-17 **Fifth International Conference on Solid Films and Surfaces**
Brown University, Rhode Island
Information: Dr. Joseph J. Loferski, Tel: (401) 863-2652.

21-23 **IEEE EMC 90 Symposium**
Washington Hilton Hotel, Washington, DC
Information: Joe Fisher, Tel: (703) 521-6336.

28-5 **XXIII General Assembly of the International Union of Radio Science (URSI)**
Prague, Czechoslovakia
Information: Prof. V. Zima, Institute of Radioengineering and Electronics, Czechoslovak Academy of Sciences, 182 51 Praha 8, Czechoslovakia.

September

10-13 **20th European Microwave Conference**
The Intercontinental Hotel, Budapest, Hungary
Information: Microwave Exhibitions and Publishers Limited, 90 Calverley Road, Tunbridge Wells, Kent, TN1 2UN, England. Tel: (0892) 544027.

23-27 **Association of Old Crows Meeting**
Hynes Convention Center, Boston, MA
Information: Association of Old Crows, 1000 N. Payne Street, Alexandria, VA 22314. Tel: (703) 549-1600.

25-27 **Piezoelectric Devices Conference**
Kansas City Westin Crown Center, Kansas City, MO
Information: EIA Components Group, 1722 Eye Street, N.W., Suite 300, Washington, DC 20006. Tel: (202) 457-4980.

October

2-4 **SCAN-TECH 90**
Georgia World Congress Center, Atlanta, GA
Information: AIM USA, 1326 Freeport Road, Pittsburgh, PA 15238. Tel: (800) 338-0206, (412) 963-8588.

8-11 **Antenna Measurement Techniques Association Symposium**
Wyndham Franklin Plaza, Philadelphia, PA
Information: Ms. Jennifer Wentz, 1990 A.M.T.A. Symposium, Flam & Russell, Inc., P.O. Box 999, Horsham, PA 19044. Tel: (215) 674-5100.

17-20 **EMC Expo 90**
San Mateo County Expo Center, San Mateo, CA
Information: EMC Technology, P.O. Box D, State Route 625, Gainesville, VA 22065. Tel: (703) 347-0030.

RF courses

Global Positioning System: Principles and Practice
July 16-18, 1990, Washington, DC
Preparation of Signals for Digital Transmission
August 7-10, 1990, Washington, DC
Antennas: Radiation and Scattering
August 27-28, 1990, Washington, DC
Grounding, Bonding, Shielding and Transient Protection
August 27-30, 1990, Chicago, IL
November 12-15, 1990, Orlando, FL
Automatic Test Systems: Design Programming, Procurement and Utilization
August 27-31, 1990, Washington, DC
Microwave Radio Systems
September 6-7, 1990, Washington, DC
Integrating Fiber Optics and Analog/RF
September 10-12, 1990, Washington, DC
Broadband Communications Systems
September 10-14, 1990, Washington, DC
Modern Receiver Design
September 10-14, 1990, Washington, DC
October 15-19, 1990, London, England
Radar Operation and Design: The Fundamentals
September 18-21, 1990, Washington, DC
Information: George Washington University, Merril Ferber. Tel: (800) 424-9773; (202) 994-6106. Fax: (202) 872-0645.

Designing for Surge and Transient Immunity in Electronic and Computer Systems
July 16-19, 1990, Madison, WI
Information: University of Wisconsin-Madison, College of Engineering. Tel: (800) 222-3623; (414) 227-3200. Fax: (414) 227-3119.

RF and Microwave Circuit Design II
July 16-20, 1990, Los Angeles, CA
Fiber Optic Smart Structures and Skins
July 23-27, 1990, Los Angeles, CA
Analog MOS Integrated Circuits
September 24-28, 1990, Los Angeles, CA
Information: UCLA Extension, Engineering Short Courses. Tel: (213) 825-3344. Fax: (213) 206-2815.

Principles and Applications of Millimeter Wave Radar
July 23-27, 1990, Atlanta, GA
Radar Reflectivity Measurement: Techniques and Applications
July 30-August 2, 1990, Atlanta, GA
Information: Georgia Institute of Technology, Education Extension. Tel: (404) 894-2547.

Modern Power Conversion Design Techniques
July 16-20, 1990, Chicago, IL
September 10-14, 1990, San Francisco, CA
Information: E/J Bloom Associates, Inc., Mrs. Joy Bloom. Tel: (415) 492-8443.

1990 EMI Training Institute I
August 6-17, 1990, Philadelphia, PA
Information: R&B Enterprises. Tel: (215) 825-1960.

Introduction to EMI/RFI/EMC
July 17-19, 1990, Chicago, IL
1992 and the EEC Directive on EMC
July 18-20, 1990, Boston, MA
System Integration and Design for EMC
July 24-27, 1990, Chicago, IL
Information: Interference Control Technologies, Inc., Elizabeth Price. Tel: (703) 347-0030. Fax: (703) 347-5813.

RF/MW Amplifier and Oscillator Design: Linear/Nonlinear Considerations
September 17-20, 1990, Budapest, Hungary
Information: Scientific Society for Telecommunications jointly with Besser Associates. Tel: 36-1-153-1027, Fax: 36-1-156-1215.

ELINT/EW: Data Bases
September 10-11, 1990, Syracuse, NY
ELINT Analysis
September 12-14, 1990, Syracuse, NY
ELINT Interception
September 17-19, 1990, Syracuse, NY
Radar Vulnerability to Jamming
September 20-21, 1990, Syracuse, NY
Information: Research Associates of Syracuse. Tel: (315) 455-7157.

Introduction to Fiber Optic Communications
July 17-20, 1990, Washington, DC
August 21-24, 1990, Los Angeles
September 18-21, 1990, San Francisco
Digital Signal Processing: Techniques and Applications
July 24-27, 1990, Boston, MA
August 14-17, 1990, Washington, DC
August 21-24, 1990, San Diego, CA
September 18-21, 1990 Washington, DC
Introduction to Telecommunications
July 17-20, 1990, San Francisco, CA
July 31-August 2, 1990, Ottawa
August 21-24, 1990, San Diego, CA
August 28-31, 1990, Washington, DC
Information: Learning Tree International, John Valenti. Tel: (800) 421-8166; (213) 417-8888. Fax: (213) 410-2952.

Principles of Analog Oscilloscopes
August 21-22, 1990, Los Angeles, CA
Principles of Digital Oscilloscopes
August 23-24, 1990, Los Angeles, CA
Information: John Fluke Mfg. Co., Inc. Tel: (800) 443-5853 ext. 73. In Canada: (416) 890-7600.

Design for ESD and RFI
September 19, 1990, San Jose Hyatt, CA
Information: The Keenan Corporation, Ms. Jean Whitney. Tel: (813) 544-2594. Fax: (813) 544-2597.

Touchstone
July 24-26, 1990, Westlake Village, CA
OmniSys
August 14-16, 1990, Westlake Village, CA
mwSPICE
August 21-23, 1990, Westlake Village, CA
Academy (Schematic)
September 11-12, 1990, Westlake Village, CA
Academy (Layout)
September 13-14, 1990, Westlake Village, CA
Information: EEsof, Ginger Craft. Tel: (818) 991-7530.

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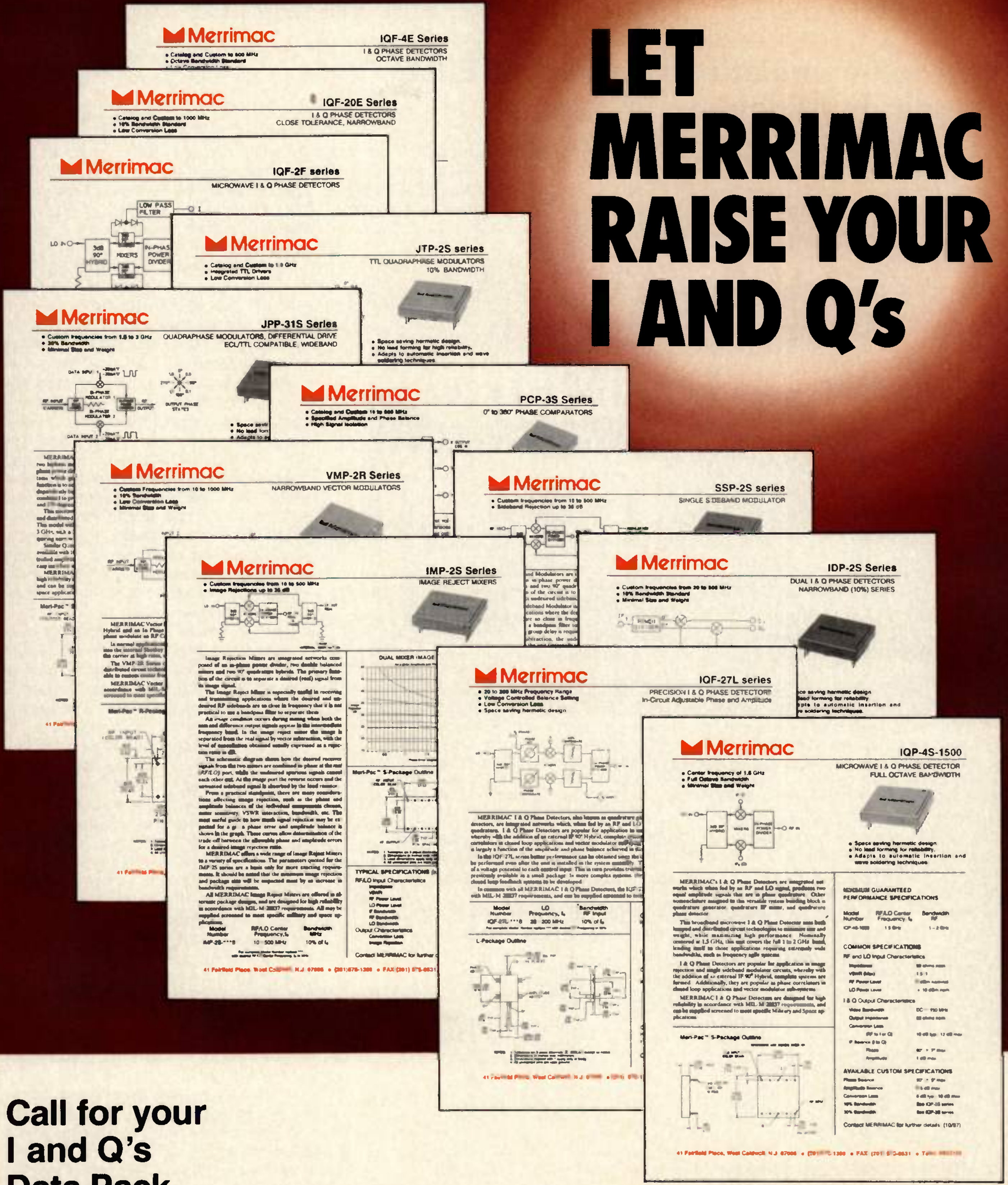

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INFO/CARD 13

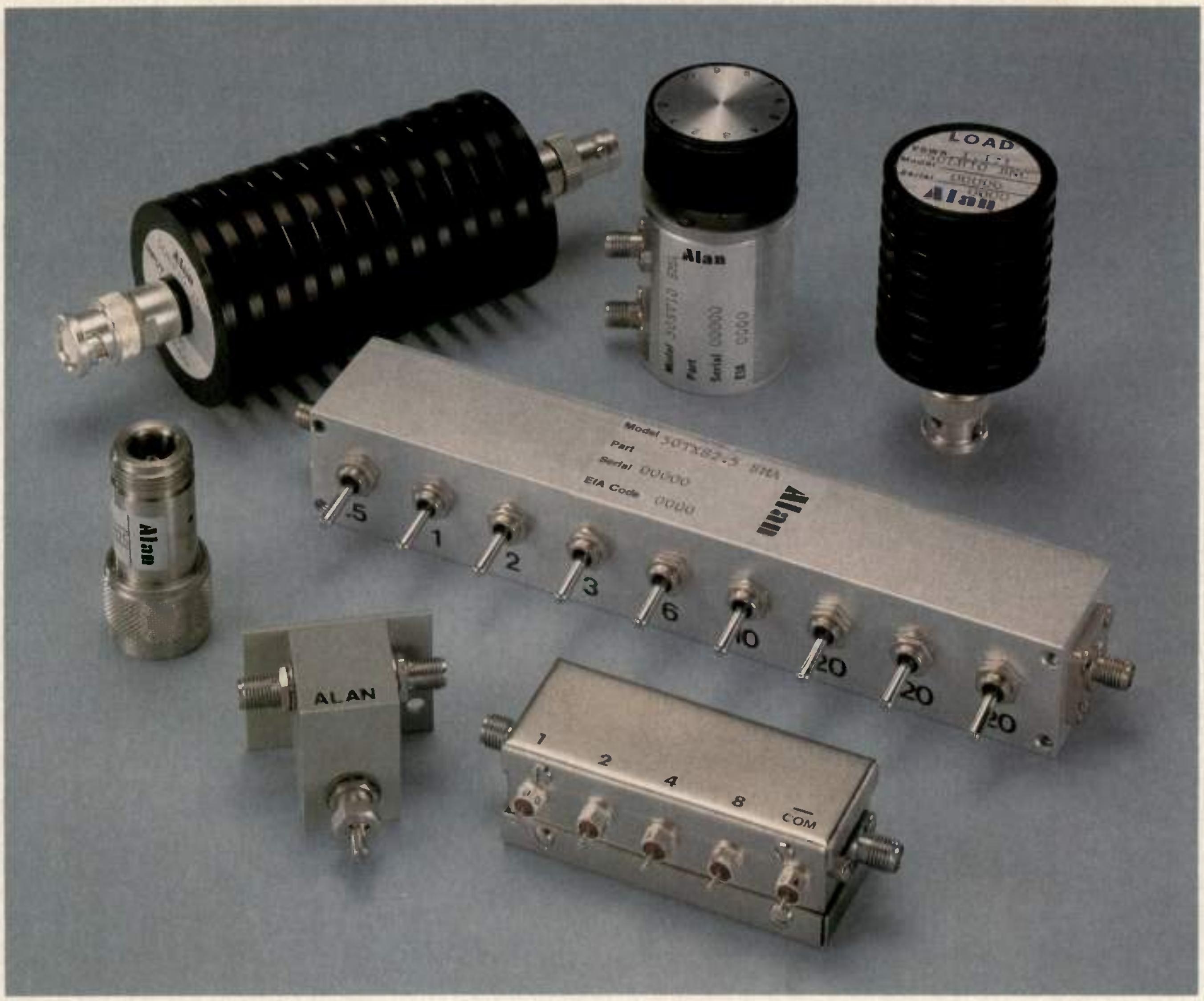
Telecommunications Bill Would Ease Spectrum Allocation Problems

A bill, currently in the House subcommittee on Telecommunications and Commerce, is intended to provide at least a temporary solution to the shortage of available commercial spectrum. The bill, called the Emerging Telecommunications Technologies Act, would reallocate part of the U.S. government's apportioned spectrum to the commercial sector. The government currently holds approximately 40 percent of the spectrum for its own use, much of which the sponsors claim is unused or under developed.

The bill proposes to reassign frequencies that are most likely to have the greatest potential for commercial use, to the commercial sector for new or emerging technologies that do not have spectrum space. This would be a tremendous boon to incustries currently working on projects such as HDTV and new cellular technologies. The specific frequencies would be a span of not less than 200 MHz located below 5 GHz and not critical to the defense of the United States. In some cases, portions of the spectrum would be directly handed over to the commercial sector while others would receive mixed use, i.e., both the commercial sector and the government would share frequencies within separate geographical areas.

The bill would be implemented over a period of up to five years and would include the following provisions: promotion of "the maximum practicable reliance on commercially available substitutes, the sharing of frequencies in geographically separate areas, the development and use of new communications technologies, and the use of nonradiating communications systems where practicable." In addition, the bill would have to "assume reasonable rates of scientific progress and growth of demand for telecommunications services; determine the extent to which the reassignment will relieve actual or potential scarcity of frequencies available for commercial use; and seek to include frequencies which can be used to stimulate the development of new technologies." The Secretary of Commerce will make recommendations as to which frequencies offer the best possibilities for reassignment, a private sector advisory committee will review and report on the Secretary's findings and finally, the frequencies will be reassigned as deemed appropriate. In addition, the President will have the right to rescind any decisions concerning frequency allocation if he thinks it necessary.

The bill was introduced on July 21, 1989 by Representatives John Dingell (D-MI) and Ed Markey (D-MA) and was referred to the Committee on Energy



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and Congress. The House subcommittee is expected to vote on it in June, at which time it is expected to pass. However, there is concern that the bill will run into problems in the Senate. The Department of Defense as well as other agencies within the Federal Government oppose the bill and would prefer to see it defeated. Nevertheless, the bill offers the only current attempt at a solution to the problem of spectrum allocation and would give the commercial sector and the government some time to come up with a long-term solution.

M/A-COM and AT&T Enter Agreement for GaAs MMICs — M/A-COM has announced an agreement with AT&T Microelectronics to offer products and foundry services of GaAs MMICs used in radio frequency and microwave applications for both commercial and defense systems. Under the agreement, M/A-COM will provide customer interface, training, standard and custom product design, test, assembly and product sales services while AT&T Microelectronics will provide, through M/A-COM, molecular beam epitaxy (MBE) GaAs MMIC processing services from its facility in Reading, PA.

Hamilton Engineering to Design and Construct Test Facility in U.S.S.R. — GOSSTANDART, the Soviet Union's national standards bureau, has selected Hamilton Engineering Inc. to design and construct a state-of-the-art electromagnetic test facility in Moscow by February of 1992. The 10,000 square foot facility will enable the Soviet Union to accurately measure electronic emissions from various types of electromagnetic equipment produced in the U.S.S.R. or imported into the country. Among other functions, such as preventing health hazards, tests performed at the new lab will help combat the growing worldwide problem of electromagnetic incompatibility and interference. The test facility will include an anechoic chamber and a transverse electromagnetic cell to measure, respectively, high- and low-frequency emissions from large objects. The project will also include other test labs, offices and a control room, forming a complete metrology complex.

Jet Propulsion Laboratory Reports on Sampling Downconvertor for RF Signals — Scientists at Caltech in conjunction with NASA's Jet Propulsion Laboratory have proposed a circuit with both digital and analog components which would act as a sampling downconvertor for RF signals. The circuit, which was reported in the May 1990 issue of *NASA Tech Briefs,* "would sample an incoming signal in phase and in quadrature, digitize it, and down-convert it to baseband in a single step." According to the report, "The design eliminates the need for several components found in conventional analog designs, including mixers, postmixer filters, and a 90 degree phase shifter." The signal would be processed digitally which would help reduce instrumental phase and delay errors common in all-analog versions.

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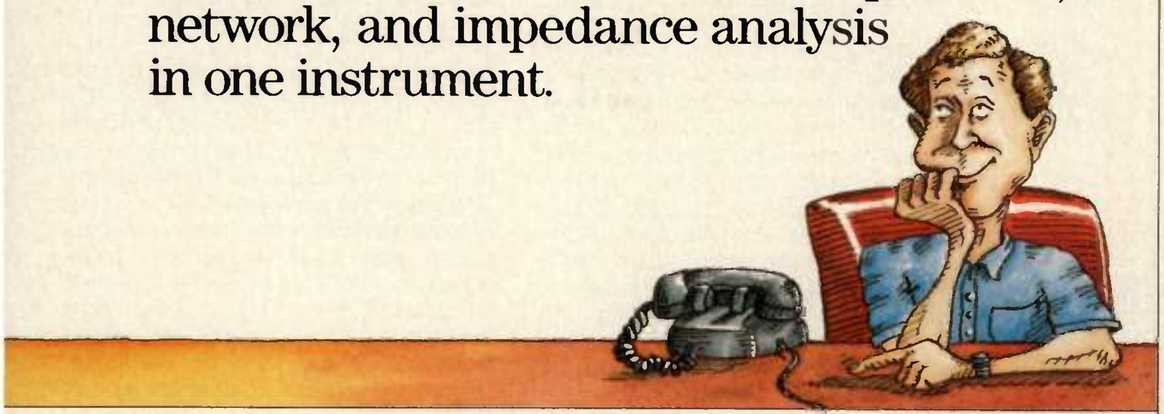

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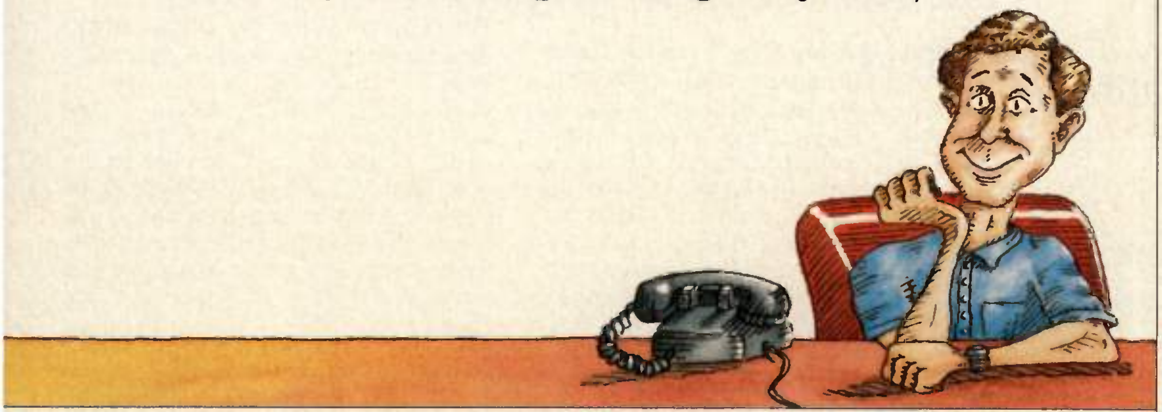

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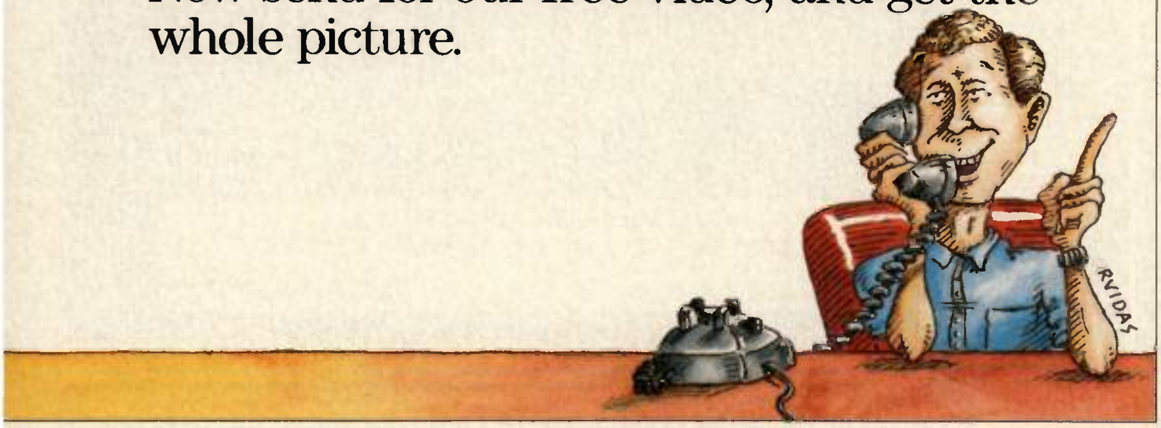

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INFO/CARD 17

Electric-Field Meter Developed — NIST researchers have developed a new isotropic, photonic electric-field meter capable of measuring continuous wave electric fields from 10 to 15,000 volts-per-meter over a frequency range of 10 kHz to beyond 1 GHz. Potential applications include electromagnetic pulse measurement, the precise measurement of any pulsed field of suitable intensity, and the measurement of fields with multiple frequency components. For a copy of paper no. 63-69 which describes the probe, contact Jo Emery, Division 104, NIST, Boulder, CO 80303. Tel: (303) 497-3237.

Campaign for EPA Federal Guidance on Environmental Radio Frequency Radiation — The Electromagnetic Energy Policy Alliance (EEPA) has been urging Congress to examine the possibility of completion of Federal Guidance on environmental radio frequency radiation (200 kHz to 300 GHz) by the EPA. As reported in the Winter 1990 EEPA Newsletter the need for federal guidance has spread to include the whole spectrum from DC to 300 GHz. While Congress did not convince the EPA that guidelines need to be completed they did issue this statement: "Although the conferees have not specifically recommended funds for continuing work on radio frequency radiation guidelines, the conferees believe it is important to the Federal government to issue a rational, scientifically-based standard in order to alleviate the uncertainty, concern and confusion that currently exists as to what levels of radio frequency radiation are safe."

Recruiting Minority Engineering and Science Students — The Southeastern Consortium for Minorities in Engineering (SECME), based at the Georgia Institute of Technology, appears to have hit on at least one part of a solution for leading minority students towards science and engineering careers. Teachers at secondary schools work to relate the curriculum to the real world and career choices, in effect, acting as guidance counselors. The thirteen year old program has seen the number of minority students planning to attend college increase dramatically and their SAT scores have increased as well. In addition, one third of all U.S. black engineers who graduated in 1989 came from SECME member universities. SECME receives funding from corporations, universities and foundations such as The Carnegie Foundation and the National Science Foundation.

FM Radio Service Threatened by Interference —The National Association of Broadcasters (NAB) has filed a joint complaint saying that growing interference on the FM band, aggravated by directional antennas, threatens FM radio service. The NAB is urging the FCC to reexamine their 1988 ruling on the use of FM directional antennas and review their standards for allocation and assignment of FM stations.

Honeywell to Sell Test Instruments Division — Honeywell Inc. recently added their Test Instruments Division to its list of operations it's planning to divest. The division employs approximately 700 people at its plant in Denver and another 300 in test facilities, service centers, and sales offices around the world.

Electronic Measurements Acquires A.L.E. Systems — Electronic Measurements has purchased A.L.E., a manufacturer of high-voltage power supplies for laser-related applications. The company will operate as a subsidiary of Electronic Measurements. Terms of the acquisition were not announced.

Westinghouse Acquires Park Air Electronics — Westinghouse Electric Corporation has announced the acquisition of Park Air Electronics (PAE) Limited of England. PAE becomes a wholly owned subsidiary of Westinghouse Electronic Systems Group. Terms of the agreement were not disclosed. PAE manufactures VHF/UHF radio communications equipment and systems, mainly for the air traffic control market.

Wilcox Electric Awarded Contract by FAA — Wilcox Electric, Inc., has been awarded a $2.09 million Federal Aviation Administration contract for two Microwave Landing Systems (MLS). The two systems, one to be installed at Chicago's Midway Airport and the other at New York City's John F. Kennedy International Airport, are part of the FAA's MLS Demonstration Program. Both MLS delivered under this contract include three major units: azimuth subsystem, elevation subsystem and precision distance measuring subsystem. The Wilcox MLS has been in service at airports in Hartford, CT; Stratford, CT; Hailey, ID; and Wichita, KS.

Piezoelectric Devices Conference set for Kansas City — Formerly the Quartz Devices Conference, this meeting will be held September 25-27 at the Westin Crown Center in Kansas City. Users and manufacturers of piezoelectric devices, and suppliers to the industry will attend. Manufacturing and engineering papers will be presented.

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Model Number	Frequency Response (MHz) Min.	PO @ 1dBc (dBm) Min.	Intercept Point 3rd Order (dBm)	SS Gain (dB) Min.	Flatness (dB) Max.	VSWR (In)	VSWR (Out)	VDC
P200A-40	100–200	+ 40, 10 w	50	33	± .5	2:1	2:1	+ 24
P02500-37	2–500	+ 37, 5 w	47	37	± 1.5	2:1	2:1	+ 24
P02500-43	2–500	+ 43, 20 w	53	46	± 1.5	2:1	3:1	+ 24
P560A	462–562	+ 43, 20 w	53	33	± 1	2:1	2:1	+ 24
P700S	650–750	+ 40, 10 w	50	34	± .5	2:1	2:1	+ 24
P880-46	860–900	+ 46.5, 45 w	Class C	16	± 1	2:1	3:1	+ 24
P1GA-36	500–1000	+ 36, 4 w	46	40	± 1	2:1	2.5:1	+ 24
P1GA-40	500–1000	+ 40, 10 w	50	44	± 1	2:1	2:1	+ 24
P1575-44	1560–1590	+ 44, 25 w	Class C	44	± 1	2:1	2:1	+ 24
P1645-43	1630–1660	+ 43, 20 w	Class C	33	± 1	2:1	3:1	+ 15
P19GA-5	1850–1970	+ 37, 5 w	47	40	± .5	2:1	2:1	+ 24
P2GS-15	500–2000	+ 30, 1 w	40	35	± 1	2:1	2:1	+ 20
P2GS-28	500–2000	+ 32, 1.5 w	42	45	± 1	2:1	2:1	+ 20
P1020-37	1000–2000	+ 37, 5 w	47	27	± 1	2:1	2:1	+ 24
P1020-40	1000–2000	+ 40, 10 w	50	27	± 1	2:1	2:1	+ 24
P23GA-36	2100–2300	+ 36, 4 w	46	40	± 1.0	2:1	2:1	+ 24
P2250-36	2180–2320	+ 36, 4 w	46	44	± 1.5	1.5:1	1.5:1	+ 24
P25GA-40	2200–2500	+ 40, 10 w	50	40	± 1.0	2:1	2:1	+ 24
P42GA	500–4200	+ 30, 1 w	40	40	± 1.5	2:1	2:1	+ 20

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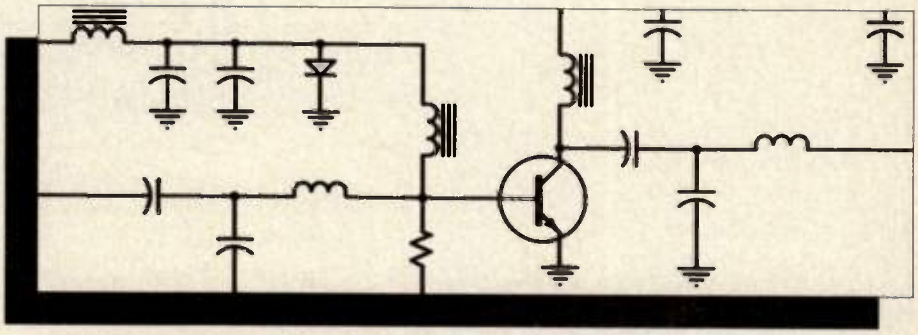
SW-449

INFO/CARD 20

Changes in the Power Amplifier Market

By Charles Howshar, and
Liane Pomfret, Assistant Editors

Like many other RF markets these days, the power amplifier market is in the midst of change — from a military to a commercial focus. As a result, companies are adjusting their marketing and production plans to reduce prices while maintaining high standards. Requirements are no longer DoD and MIL-Spec standards but the "performance, price and reliability" requirements of commercial customers. Companies can use their experience in manufacturing to military standards in their approach towards the commercial sector which makes it easier for them to provide a high quality product. Lance Wilson, Principal Staff Engineer, RF Products Division, Motorola Semiconductor, states, "We are starting to see the first large volume high power commercial applications, that beforehand had been pretty much military."



Trends in Communications

The communications market is one of the fastest growing markets for electronics in the nineties and power amplifiers are quickly finding their niche. According to Al Arbuckle of Trontech, Cellular radio is currently their greatest market. And with the heavy concentration of research in communications and the public's great desire for cellular communications in particular, the demand for power amplifiers has increased proportionately. The use of power amplifiers in satellites to expand communication abilities is also on the upswing. "Commercial satellites for subscriber-oriented global positioning are a huge market, and sales will be in the millions of units over the next decade," remarks Motorola's Wilson. Frank Morgan, Vice-President of Marketing for MCL, notes that PANAMSAT uses are expanding rapidly. "We are working a lot in the commercial satellite frequency bands (6 and 14 GHz)." As satellite communications grow, users will undoubtedly require smaller, more powerful amplifiers for their equipment.

Another trend in the market is continued replacement of vacuum tube designs with solid state MOSFET designs. New designs for solid state amplifiers are reaching higher and higher power capabilities, though still not as high as vacuum tube designs. "You can not get the highest powers out of solid state, but we're moving ahead and as new devices come out, we're incorporating them," states Don Shepherd, President of Amplifier Research. "The customers want the latest solid state MOSFET replacements for older EMC systems with vacuum tubes," says Frank Kalmus, President of Kalmus Engineering International, Ltd. Replacing vacuum tubes with MOSFET designs is desireable because of the reduced size and greater reliability of the solid state design. Ray Boyd, Marketing and Sales Manager for Erbtec Engineering agrees, saying, "Currently, vacuum tubes are used but the trend is toward solid state, which is more stable, with a lower failure rate." However, as Wilson comments, "The change from vacuum tubes to solid state is not a technological revolution in itself. There will be changes in other areas which will cause an erosion in vacuum tube usage. However, there will also be technological improvements in the microwave travelling wave tube area." In general, customers want the power capabilities of vacuum tubes with the reliability of solid state MOSFET designs, and power amplifier companies are struggling to meet this demand by making more reliable vacuum tubes and more powerful solid state amplifiers.

Performance and Reliability

With the military market at a standstill, companies must achieve growth through new markets for their products. "We have seen a greater demand for high power class A amplifiers — almost all commercial. There is no new demand from military markets," remarks Arbuckle, director of Sales and Marketing for Trontech. Conservative designs and high reliability are desired in cases where the customer needs a product that is operating well under its maximum capability, to provide full rated performance without degradation of the product.

Other applications for power amplifiers that are generating interest include magnetic resonance imaging (MRI), EMC, plasma etch, laser excitation, and nuclear magnetic resonance (NMR). "We build a subsystem that goes into high field MRI which amounts to about 40 percent of the total market," comments Ray Boyd. "As the technology for MRI gets better, we can work to improve bandwidths and phase stability." "Most of our sales in the medical market are for general testing purposes," comments Lance Wilson. Reliability requirements have led companies to EMC susceptibility testing for their products. As Don Shepherd states, "Despite a declining total electronics market, manufacturers are finding the need to test their equipment for susceptibility to improve their product's reliability, so our market is expanding." Research applications include large and small particle accelerators, atmospheric sounding radar, and high power lasers.

There is great interest in new technology and expanding the markets for power amplifiers. Third world countries and eastern Europe continue to show potential for future markets. "About 40 percent of our sales are overseas," notes Frank Morgan. New military uses for power amplifiers may not be in demand, but the commercial market is good. Although there is no explosive growth expected in the power amplifier market, most companies are positive about the future. ***RF***

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700LC	1.5W CW	.003-1000 MHz	33dB	25x28x13	100-240V	1,695.00
102LC	2W CW	25Hz-100 MHz	33dB	19x31x15	100-240V	1,349.00
502LC	2W CW	.01-525 MHz	36dB	19x32x15	100-240V	1,695.00
704FC	4W CW	.5-1000 MHz	33dB	23x18x09	100-240V	2,195.00
505LC	5W CW	.05-400 MHz	37dB	19x32x15	100-240V	1,895.00
505FC	5W CW	.5-525 MHz	40dB	25x28x13	100-240V	1,655.00
706FC	6W CW	.5-1000 MHz	38dB	25x28x13	100-240V	2,695.00
110C	10W CW	.2-60 MHz	40dB	19x32x15	100-240V	1,450.00
110LC	10W CW	.01-100 MHz	40dB	25x37x15	100-240V	1,950.00
210LC	10W CW	.008-225 MHz	40dB	25x37x15	100-240V	2,495.00
310FC	10W CW	.5-300 MHz	40dB	25x28x13	100-240V	2,250.00
510FC	10W CW	.5-525 MHz	43dB	25x28x13	100-240V	2,595.00
710FC	10W CW	1-1000 MHz	40dB	25x28x13	100-240V	5,995.00
115LC	15W CW	.001-100 MHz	43dB	38x32x13	100-240V	2,255.00
225LC	20W CW	.01-225 MHz	45dB	25x37x15	100-240V	3,295.00
520FC	20W CW	1-525 MHz	45dB	25x28x13	100-240V	3,895.00
720 FC	20W CW	500-1000 MHz	43dB	48x46x13	100-240V	6,650.00
125LC	25W CW	50Hz-100 MHz	40dB	48x40x13	100-240V	3,300.00
125C	25W CW	.5-50 MHz	45dB	48x46x13	100-240V	1,495.00
112C	25W CW	1-120 MHz	45dB	25x37x15	100-240V	1,995.00
320FC	25W CW	1-300 MHz	45dB	30x20x13	100-240V	2,895.00
535FC	35W CW	200-512 MHz	45dB	30x20x13	100-240V	3,895.00
150C	50W CW	.2-50 MHz	47dB	38x37x15	100-240V	2,595.00
250LC	50W CW	.01-230 MHz	45dB	43x42x19	100-240V	7,550.00
250FC	50W CW	1-200 MHz	45dB	38x37x15	100-240V	4,995.00
550FC	50W CW	200-512 MHz	45dB	30x20x13	100-240V	4,995.00
161C	100W CW	.5-90 MHz	45dB	38x37x15	100-240V	5,250.00
190LC	100W CW	.01-100 MHz	50dB	48x43x18	100-240V	8,850.00
162LP	100W CW	20-200 MHz	50dB	38x37x15	100-240V	6,250.00
116C	100W CW	.01-220 MHz	50dB	71x56x76	100-240V	8,995.00
155LCR	100W CW	.006-12 MHz	50dB	48x40x13	100-240V	4,150.00
LA100H	100W CW	.3-100 MHz	50dB	48x48x13	100-240V	5,990.00
LA100F	100W CW	100-250 MHz	50dB	48x48x13	100-240V	6,595.00
LA100U	100W CW	200-400 MHz	50dB	48x48x13	100-240V	6,250.00
LA100UE	100W CW	100-500 MHz	50dB	48x48x13	100-240V	6,995.00
LA200H	200W CW	.5-100 MHz	53dB	48x48x13	100-240V	6,995.00
LA200F	200W CW	100-250 MHz	53dB	48x48x13	100-240V	11,500.00
LA200U	200W CW	200-400 MHz	53dB	48x48x13	100-240V	10,995.00
LA200UE	200W CW	250-500 MHz	53dB	48x48x13	100-240V	12,500.00
162LPS	200W Pulse	10-220 MHz	55dB	48x38x18	100-240V	5,995.00
122C	200W CW	.01-220 MHz	50dB	71x56x76	100-240V	10,850.00
247V	200W Peak	118-138 MHz	10dB	48x46x13	100-240V	5,255.00
247U	200W Peak	200-400 MHz	10dB	48x46x13	100-240V	5,500.00
121CA	250W CW	.5-32 MHz	55dB	48x43x18	100-240V	5,950.00
164UP	300W Pulse	200-400 MHz	55dB	48x38x18	100-240V	7,595.00
LP300H	300W Pulse	.3-100 MHz	56dB	48x48x13	100-240V	7,550.00
LA500H	500W CW	.5-50 MHz	57dB	71x56x76	100-240V	12,500.00
LA500V	500W CW	10-100 MHz	57dB	71x56x76	100-240V	13,500.00
134C	500W CW	.01-220 MHz	57dB	58x69x127	3 Phase	19,800.00
134CM	500W CW	1-200 MHz	57dB	58x69x127	3 Phase	18,500.00
166UP	500W Pulse	200-400 MHz	57dB	48x51x18	100-240V	9,500.00
LA600U	600W CW	200-400 MHz	57dB	71x56x76	100-240V	23,500.00
166HP	800W Pulse	100-200 MHz	57dB	48x51x18	100-240V	11,500.00
137C	1000W CW	.01-220 MHz	60dB	69x69x127	3 Phase	28,500.00
137CM	1000W CW	1-200 MHz	60dB	69x69x127	3 Phase	27,250.00
LA1000H	1000W CW	2-32 MHz	60dB	71x56x76	100-240V	22,500.00
135CB	1000W Pulse	2-30 MHz	60dB	48x51x46	100-240V	10,990.00
LA1000V	1000W CW	10-100 MHz	60dB	71x56x76	100-240V	26,900.00
166LP	1000W Pulse	8-100 MHz	60dB	48x51x18	100-240V	9,550.00
LP1000	1000W Pulse	10-120 MHz	60dB	48x51x18	100-240V	11,500.00
167HP	1500W Pulse	120-200 MHz	60dB	53x56x46	100-240V	17,950.00
167LP	2000W Pulse	10-100 MHz	63dB	53x65x46	100-240V	17,900.00
140C	2000W CW	.01-220 MHz	60dB	137x69x127	3 Phase	44,990.00
140CM	2000W CW	.5-150 MHz	60dB	137x69x127	3 Phase	41,500.00
LA3000H	3000W CW	5-45 MHz	65dB	137x76x127	180-255V	69,500.00
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INFO/CARD 22

A Lowband RF Quiet Switch

By Kevin S. Randall
Interspec, Inc.

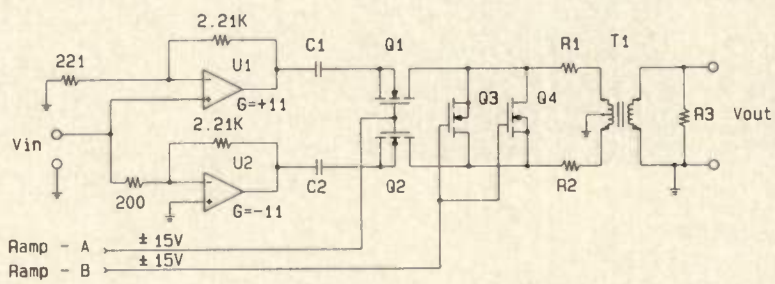
This article is the winning entry in the 1990 RF Design Awards Contest. Judged the best of many good entries, this one represents a combination of design originality, engineering problem-solving, and complete documentation. The engineering problem also shows how medical electronics have developed in recent years. This design was developed for an ultrasound imaging unit with a required signal bandwidth of 10 MHz. Discussions with the author indicate that 20 MHz systems are currently under development.

It is sometimes desirable in RF system design to be able to switch a source on and off or to switch between sources quickly, cleanly and quietly, i.e. without adding any switching residue. In narrowband and high frequency applications this is less of a problem since a reasonably fast switch (say 50-100 ns) can be implemented that will inject very little signal into the band of interest. For broadband, low RF frequency (100 kHz to 50 MHz) applications, this is a special problem, since fast switch times will create interference in the band of interest.

There don't seem to be any off-the-shelf solutions to this problem, with the possible exception of double balanced mixers. Inexpensive mixers, however, can achieve switching signal isolation and off- isolation of only 40-60 dBc. This problem is complicated by the fact that the layout of the printed circuit board on which the circuitry resides can easily degrade this further. This can be illustrated by pointing out that a 1 pF parasitic capacitance between two 50 ohm traces degrades isolation to about 50 dB at 10 MHz. Of course, as frequency increases, isolation degrades proportionately. If the switching signal is larger than the signal of interest (as is frequently the case), the amount of switching residue can quickly get out of hand. Also signal impedances are frequently higher than 50 ohms.

To consistently achieve off-isolation and switching noise isolation in excess of 60 dB without sensitive adjustments for nulling, requires very careful circuit board layout and a novel switching function that can effectively reject switching noise. Figure 1 shows a circuit that achieves this by incorporating several design features that minimize switching signal injection. In addition, off-isolation is typically in excess of 60 dB at 10 MHz.

The circuit of Figure 1 is a differentially configured series-shunt switch. When the switch is "on," Ramp-A is at +15 V and Ramp-B is at −15 V. Therefore Q1 and Q2 conduct a differential signal to the secondary of the transformer, and the input signal appears at V_{out}. When the switch is "off," Ramp-A is at −15 V and Ramp-B is at +15 V. Q1 and Q2 are open and Q3 and Q4 are activated, shunting the secondary of the transformer. The on resistance of the FETs is typically 23 ohms for +15 V V_{gs}. Thus approximately 12 ohms are shorting whatever parasitic capacitance appears across Q1 and Q2 when they are off. This is generally less than 1 pF, largely dependent on layout parasitics. Capacitors C1 and C2 are DC blocking capacitors. Without them, offset voltages of the two amplifiers would cause a square wave at the switching rate to appear at V_{out}. Resistors R1, R2, and R3 and transformer T1 are used to set impedance levels for the



Q1 - Q4 : SD5000 (Siliconix)
C1, C2 : 1uF
U1, U2 : EHA2539 (Elantec)
T1 : MCL T2.5-6T 2.5:1 impedance ratio (Mini-Circuits)
R1, R2 : 470
R3 : 200

Figure 1. A differentially configured series shunt switch.

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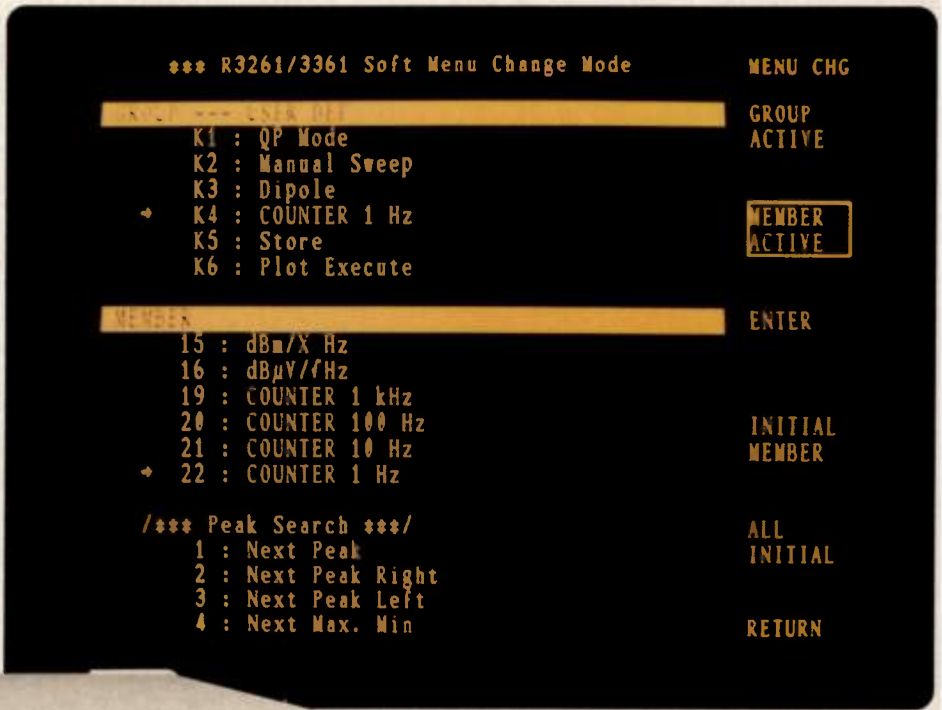

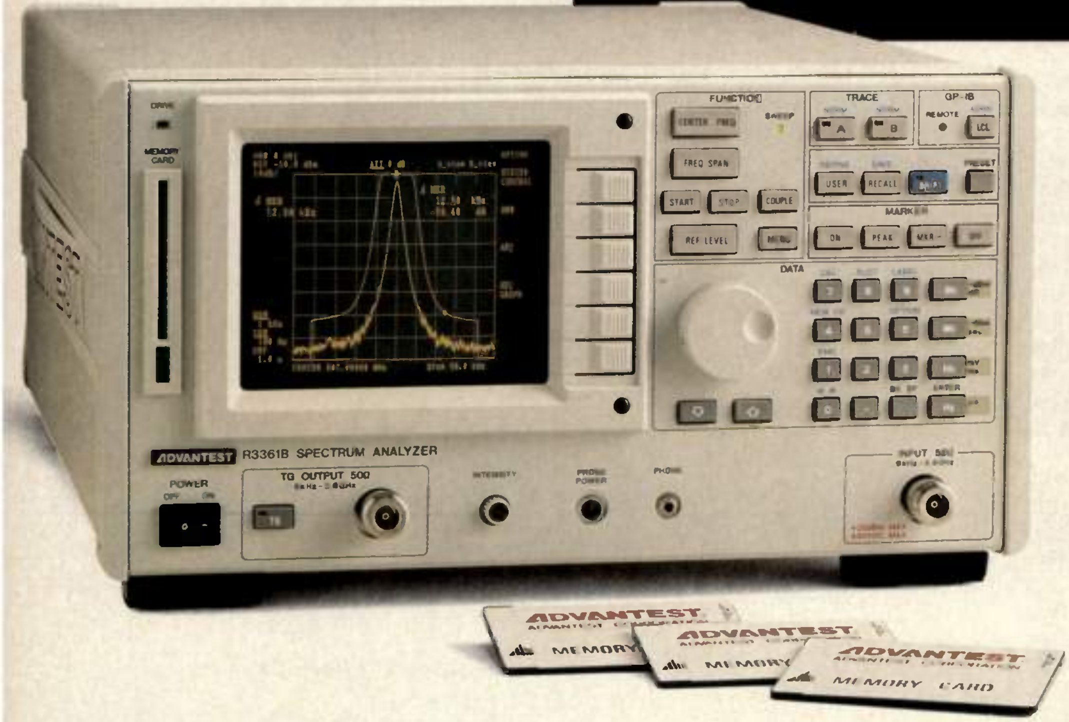

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INFO/CARD 23

amplifiers and set the output impedance. Note that output impedance will in general be different when the switch is closed as compared to when it is open.

Ramp-A and Ramp-B are two switching signals that swing between ±15 V. Each of these signals is generated by a circuit such as that of Figure 2. After a rising edge of the input TTL level, Q5 turns on, pulling one lead of R6 to ground. This provides drive for the current mirror consisting of Q7 and Q8. The constant current out of the collector of Q8 charges capacitor C4, creating a ramping signal until the output ramp reaches such a voltage that CR1 removes base drive from Q8. This keeps Q8 from entering saturation. R6 and R7 will therefore control the switching edge rates (to minimize high frequency content).

Figure 3 shows Ramp-A and Ramp-B signals at a transition. Predictable edge rates allow selection of the crossover gate voltage. This crossover is chosen to be a voltage around 3-6 V, which is slightly higher than the maximum V_{gs} threshold voltage of 2 V. Thus all four FETs are weakly "on" when the switchover occurs. A lower switchover voltage would leave all four FETs off for a short time, increasing the sensitivity of the switch to parasitics during this period. A higher switchover voltage would effectively short the amplifiers' outputs together for a brief period.

The ramps shown in Figure 3 indicate a total switchover time of slightly more than 200 ns. The ramps are, however, somewhat extended because of the extra capacitance of the scope probes that were used to monitor them. Without this loading, the longer of the two is about 150 ns. This is technically also the switch-on/switch-off time, however it only applies for maximum input levels, since the gate voltage is set by these ramped signals and the source voltage is set by the input signal level. For a V_{gs} of only 5 V a FET is essentially "on," and for a V_{gs} of 0 V the FET is "off." Thus for a ± 2 V analog signal level at the outputs of the amplifiers, an essentially complete switchover occurs when the gate drive swings from +7 V to −2 V and the other swings from −2 V to +7 V. This takes approximately 60 ns. Note, however, that best off-isolation is not achieved until the gate drives reach ± 15 V.

The amplifiers U1 and U2 are configured so one is inverting and the other is non-inverting. The actual gains should be similar for best dynamic range. (For highest output signal, both amplifiers

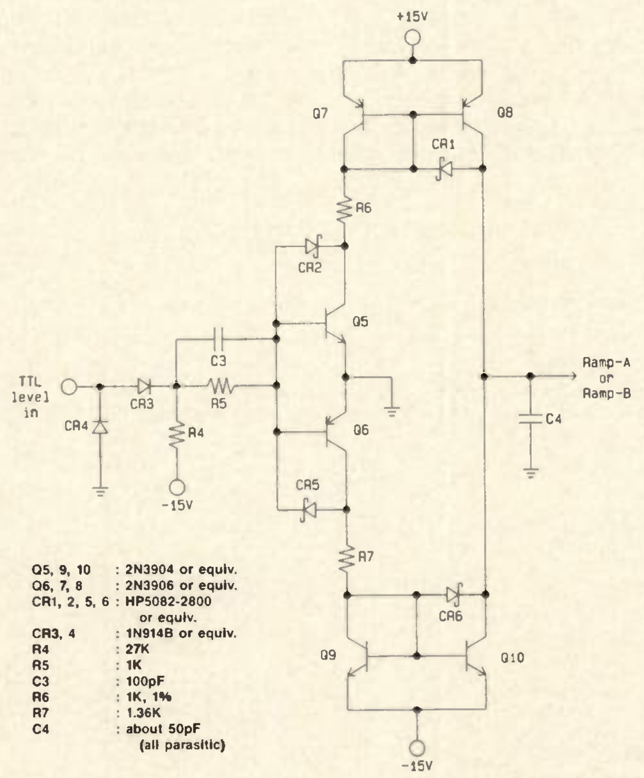

Figure 2. Circuit to generate Ramp-A and Ramp-B.

should be just short of clipping. If one is almost clipping while the other still has some headroom, the maximum output signal will be somewhat lower than it could be, compromising dynamic range.) Interestingly enough, the output impedances do not have to be very closely matched. A considerable mismatch is required to have a noticeable effect on the switching signal feedthrough. Also the absolute value of the input impedances has no noticeable effect on the switching signal feedthrough. An open circuit was substituted for U1 and U2 with no significant change in feedthrough.

The manner in which the circuit achieves very low switching signal injection is by first keeping parasitics to an absolute minimum and second by precisely matching the residual parasitics between the switching gate drives and the two differential signal lines. Figure 4 shows the parasitics between the gate drives and the signal nodes. If the parasitic capacitances are exactly matched, then identical noise injection will occur to each half of the transformer secondary. If the transformer is well balanced and has sufficient frequency response, then it will exactly cancel the switching signal feedthrough.

In practice, it has been found to be rather straightforward to achieve both these objectives with a standard multilayer printed circuit layout. With a ground plane as one of the middle layers of the board, all analog signals are run on layers that are on the component side of the ground plane. Each corresponding differential signal should run in a way that mirrors its counterpart with respect to the gate pins of the IC. The gate drive signals should run on the solder side of the ground plane, to shield these traces from the body of the IC and the analog signal nodes, and should run so that their residual parasitic capacitance balances to each half of the switch. The only parasitic capacitance that the gate drives traces will have will be from the traces on the solder side of the board to the pins of the IC that protrude through the board. Although for best performance the protruding pins of the IC should be clipped off flush with the solder side of the board, I have found that this makes little difference in the performance of the circuit. I found in implementation that the gate drive traces don't need to maintain their symmetry once they are more than a few tenths of an inch away from the IC pins. Also, the ramp generating circuitry is only about 1.5 inches from the switch and is not shielded in any way.

In addition to layout capacitances, of course, the FETs themselves have parasitic capacitances. These capacitances are particularly low for the SD5000 series of FET arrays being about 0.3 pF C_{gd} and about 2 pF C_{gs}. Although it is not explicitly stated in the data sheet, the match of these capacitances within one package appears to be excellent. This is not surprising since they are specified to have a typical on resistance match of 1 ohm. As can be seen from the

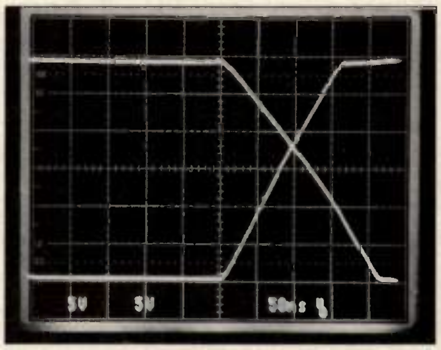

Figure 3. Ramp-A and Ramp-B signals at a transition.

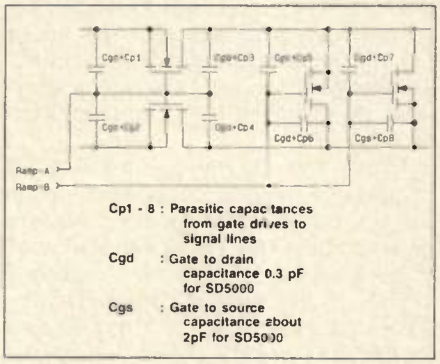

Figure 4. Parasitics between the gate drives and the signal nodes.

schematic, there is one C_{gd} from Ramp-A to each of the transformer halves (one each from Q1 and Q2). Note that C_{ts}, being larger than C_{gd}, for Q1 and Q2 is oriented to discharge intc the low output impedance of amplifiers U1 and U2.

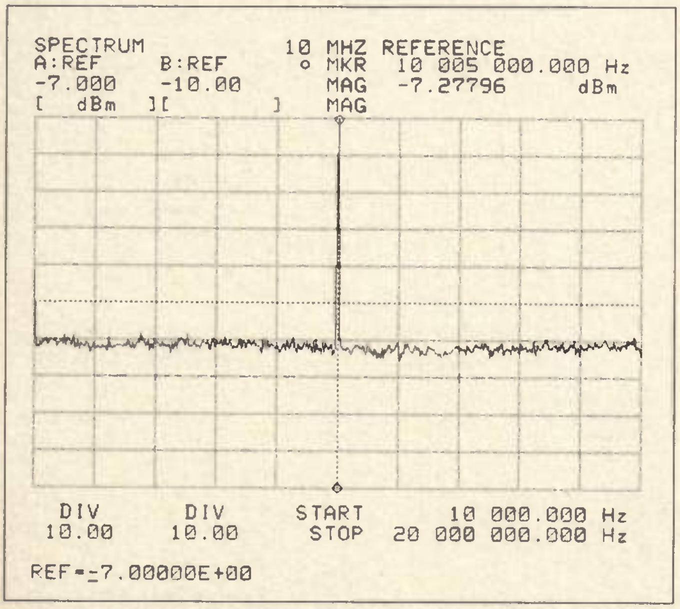

Figure 5. Reference tone of 10 MHz passing through statically-on switch.

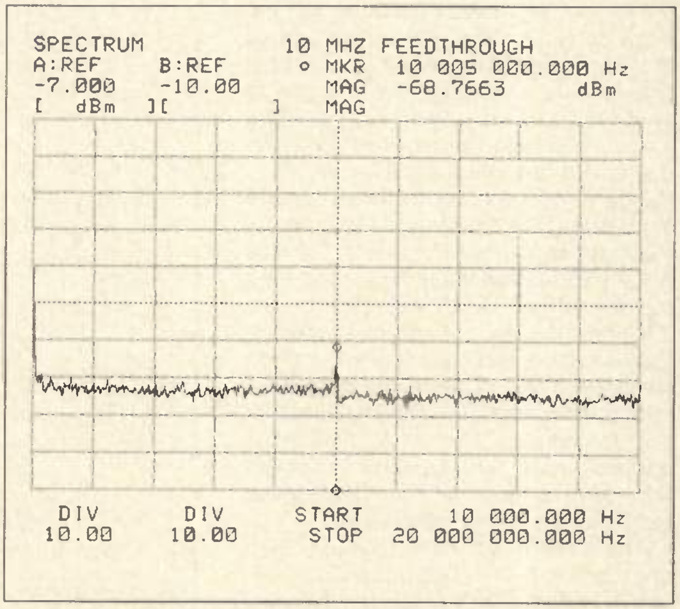

Figure 6. Reference tone of 10 MHz passing through statically-off switch.

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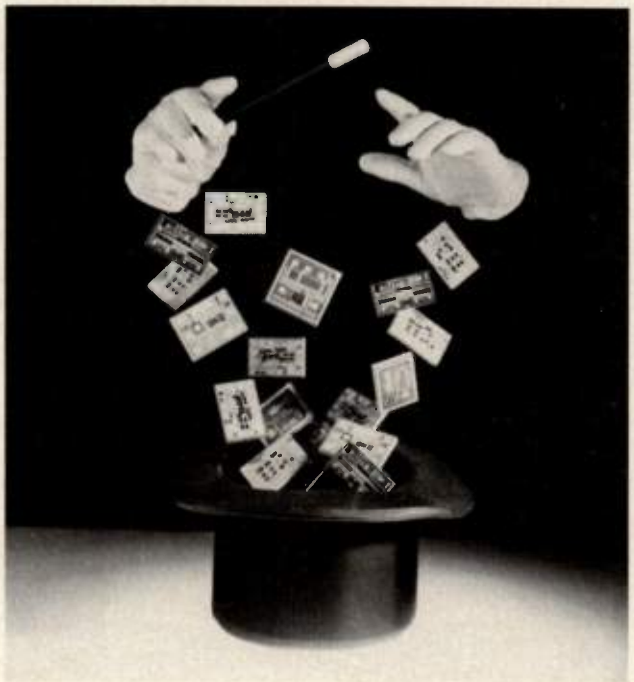

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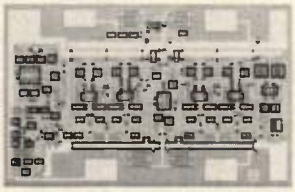

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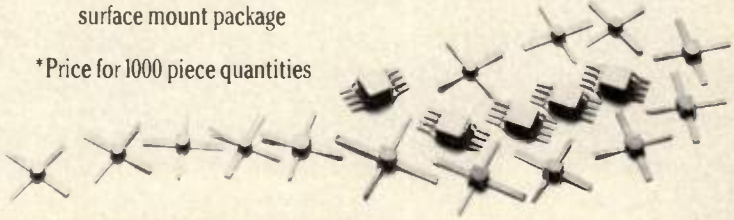

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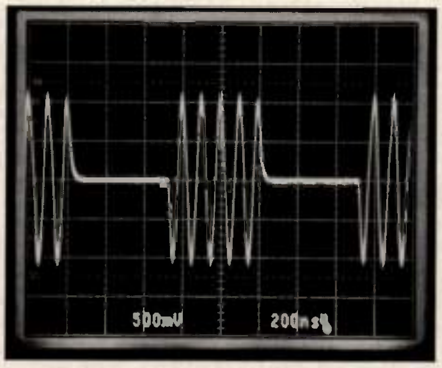

Figure 7. 10 MHz tone being modulated by asynchronous 1 MHz gating function.

Because Q3 and Q4 are connected drain to source and source to drain, there is exactly one C_{gd} and one C_{gs} from Ramp-B to each of the output transformer halves. Thus the predictable FET capacitances of Q1-Q4 should precisely match at each half of the transformer.

One important consideration for best dynamic range is to take full advantage of the signal level that the FETs can handle. For gate drives of $\pm$ 15 V, this is a $\pm$ 10 V_p signal level at a rated current of about 40 mA (50 mA is absolute maximum rating). This was not implemented fully in the circuit in Figure 1. All switching noise criterion were net even with a non-optimal choice of signal impedances, so for convenience I selected amplifiers with maximum output signal level of $\pm$ 10 V at $\pm$ 10 mA.

Figures 5-9 show the measured performance of this circuit in the configuration that is used in an ultrasound medical imaging system. Figure 5 is a reference tone of 10 MHz passing through the switch while it is statically on. A scope probe monitors V_{out}, and the scope channel output is fed into a spectrum analyzer. When the gain of the scope channel is included, 0 dBc = −7.3 dBm. 0 dBc is chosen to be 2 V_{pp}, even though the switch is capable of driving 6 V_{pp}. Figure 6 is the same tone with the switch statically off. It shows an off-isolation in excess of 60 dBc. Figure 7 is the 10 MHz tone being modulated by a synchronous 1 MHz gating function. Note the slightly smaller half-cycle at the beginning and end of the burst, and the slightly distorted last half-cycle. These are due to the ramping nature of the gate drives causing the resistances of the FETs to vary continuously rather than abruptly. Figure 8 is the spectrum of the signal in Figure 7. Figure 9 is the spectrum of the switching noise that

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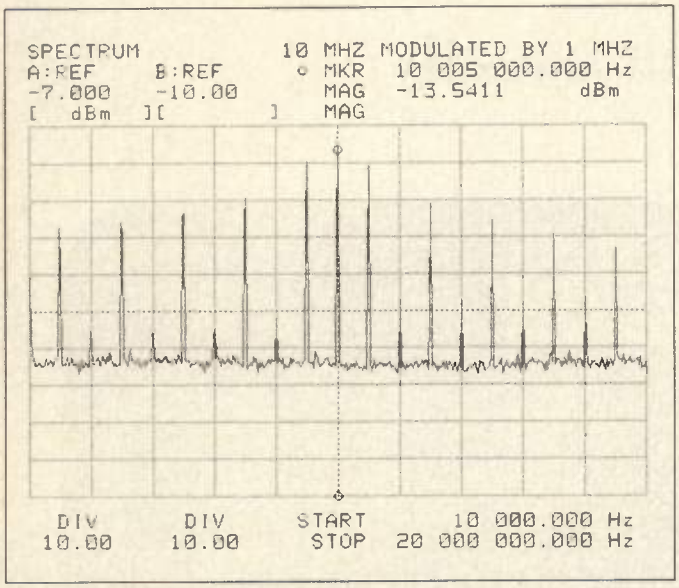

Figure 8. Spectrum of the signal in Figure 7.

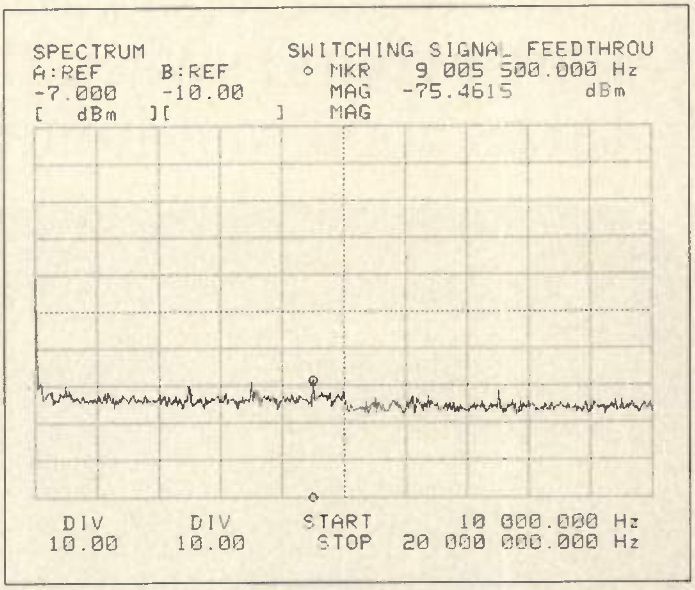

Figure 9. Spectrum of the switching noise that feeds through the switch from the 1 MHz gating function.

feeds through the switch from the 1 MHz gating function. Note that this shows almost 70 dBc of isolation. (In actuality the isolation is almost 80 dBc of the true maximum output signal level of 6 V_{pp}.)

Conclusion

A novel circuit design for quietly switching low frequency RF signals was presented as well as the means by which such a circuit should be implemented. The bandwidth of the actual circuit as presented is from 30 kHz to 20 MHz, but with judicious selection of alternate components, the circuit could readily achieve suitable operation to perhaps 100 MHz. By substituting a grounded center tap transformer secondary for the input amplifiers, the circuit operation could perhaps extend beyond 100 MHz. As monolithic operational amplifiers improve it may be possible to substitute a true differential amplifier for the output transformer, extending performance down to DC. At present none were found that had inverting and non-inverting responses that matched closely enough over sufficient bandwidth to achieve anything near 60 dB of switching isolation. It could be possible to implement some hybrid approach with a relatively low frequency operational amplifier to extend response down to DC. *RF*

About the Author

Kevin Randall is this year's Grand Prize winner in the RF Design Awards Contest. Kevin is a Senior Design Engineer at Interspec, Inc., manufacturer of medical ultrasound imaging equipment. Kevin's design, dubbed the "quiet switch," was developed to solve a significant performance problem in ultrasound units. To achieve proper resolution of the image, the units must continuously change the gain and delay characteristics of the returning echoes. Switching between the various delay lines can add considerable noise unless a very quiet switch is used.

Kevin received his BSEE degree from Lehigh University and his MSEE degree from Drexel University. He has been with Interspec for over three years. He can be reached at Interspec, Inc., 110 W. Butler Avenue, Ambler, PA 19002.

Hybrid PLL/DDS Frequency Synthesizers

Robert Gilmore and Richard Kornfeld
QUALCOMM, Inc.

Increasing demands are being placed upon the frequency synthesizer designer to achieve greater bandwidths of operation, finer frequency resolution, improved phase noise and spurious characteristics, faster settling times, and all of these requirements with reduced physical size and lower power consumption. Single loop phase lock loop (PLL) synthesizers, while small in size, provide poor noise, spurious, and settling performance, and exhibit microphonic behavior when used at high frequencies with small step size. The multiple loop PLL synthesizer provides additional phaselock loops to achieve finer frequency resolution, at the expense of increased physical size, higher power consumption, and a relatively high switching time between output frequencies.

Techniques are described to combine the Direct Digital Synthesizer (DDS) and the PLL to create powerful hybrid synthesizers which can meet difficult performance specifications. Using the QUALCOMM Q2334 Dual DDS and the QUALCOMM Q3036 single chip PLL synthesizer, this hybrid synthesizer can be implemented using a minimal amount of circuitry and power.

Direct Digital Synthesizer/Direct Analog Synthesizer Hybrid

The first hybrid synthesizer presented incorporates a phaselock loop which provides a fixed frequency output. Yet this synthesizer provides both a bandwidth expansion and an upconversion of the DDS output. The authors call it a DDS/DAS hybrid, and an example is given in Figure 1.

Figure 1 depicts a synthesizer required to output 187-227 MHz with extremely fine frequency steps and fast switching speed. The frequency resolution is that of the DDS and the switching speed is determined by the RF switches used to select the various output frequency ranges. The switches in Figure 1 select one of the four output ranges 187-197 MHz, 197-207 MHz, 207-217 MHz, or 217-227 MHz. Therefore, 40 MHz of output bandwidth is achieved using a 10 MHz bandwidth DDS without the use of frequency multiplication techniques. Additional mixing stages, each multiplying with a selection of two or more local oscillators, can be used to provide increased bandwidth expansion and higher output frequencies.

Performance analysis of the DDS/DAS hybrid is straightforward. A conventional intermodulation product analysis must be performed to determine the spurious content of the output, and the frequency plan selected accordingly. Phase noise performance is excellent due to the inherently high performance of the DDS. Since the PLL is not required to switch between output frequencies, optimization of its noise performance is a relatively easy task.

PLL with DDS-Generated Frequency Offset

Figure 2 shows a phaselock loop with an internal offset frequency generated by a DDS. This approach is typical of multi-loop synthesizer designs, but in this case the fine frequency step PLLs have been replaced by a single DDS. Due to the fine resolution of the DDS, this synthesizer is capable of providing better frequency resolution than a PLL approach containing many loops. The local oscillator shown in Figure 2 is optional; the DDS fundamental output can be applied to the mixer directly. DDS image responses may also be used, eliminating the need for the upconverting LO. When the optional "÷P" divider is not used (P=1), the synthesizer output frequency is given by

$$F_{out} = NF_{Ref} + F_{DDS} + F_{LO} \qquad (1)$$

The PLL provides the coarse steps in units of F_{ref}. The DDS provides the fine resolution to fill the gaps between the coarse steps. For continuous frequency coverage, the DDS output bandwidth must be greater than or equal to the reference frequency:

$$BW_{DDS} \geqslant F_{Ref} \qquad (2)$$

The synthesizer step size is that of the DDS, usually << 1 Hz. A design option is to include a fixed divider, "÷P," be-

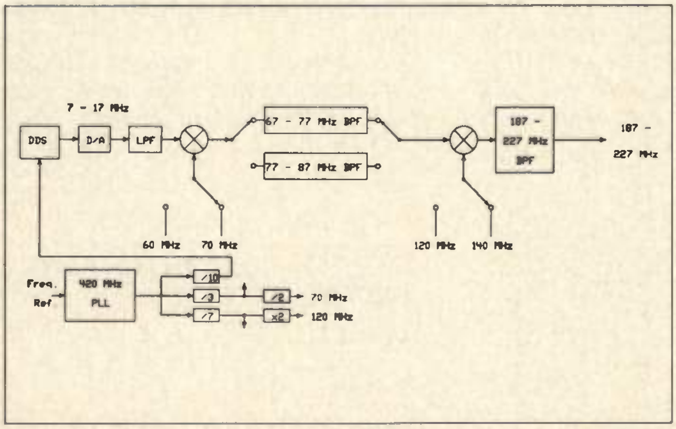

Figure 1. Example of DDS/DAS hybrid.

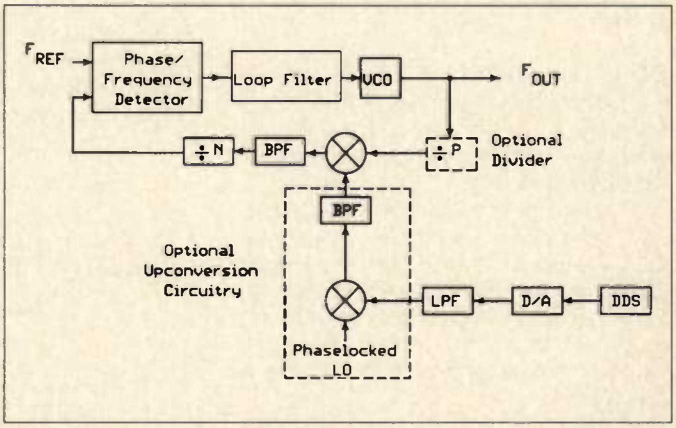

Figure 2. PLL with DDS-generated frequency offset.

tween the synthesizer output and the mixer. In this case the output frequency is given by:

$$F_{out} = NPF_{Ref} + P[F_{DDS} + F_{LO}] \qquad (3)$$

As in the previous situation, the DDS bandwidth must be $BW_{DDS} \geqslant F_{ref}$. The frequency step size of this synthesizer is:

$$\text{Step Size} = (\text{DDS Step Size}) \times P \qquad (4)$$

This DDS/PLL hybrid permits the reference frequency to the PLL to be relatively large, while still providing extremely fine frequency steps. A large reference frequency provides important benefits:

1. The loop division ratio N is a low value. Since the output phase noise within the loop bandwidth is the reference phase noise + 20log(N) dB, a small value of N minimizes this noise.

2. The loop bandwidth is typically $\leqslant$ 10 percent of the reference frequency (this is a rule of thumb: the authors recommend that a detailed noise and stability analysis be performed for all PLL designs). Therefore, a large reference frequency permits a wide loop bandwidth. Since the VCO noise is suppressed by 12 dB/octave beginning at ω_3dB and headed toward DC, a wide loop bandwidth yields a low phase noise output despite a noisy VCO.

3. The wide loop bandwidth made possible by this synthesizer provides a loop with a proportionally faster settling time.

Since this topology is similar to that of multi-loop designs, the performance analysis is that of a conventional PLL synthesizer. Of course, the DDS provides its inherent benefits of fine frequency steps, fast settling, and excellent phase noise and spurious performance.

DDS - Driven PLL

This application addresses the use of the DDS as the reference for a PLL. It is a powerful technique for a number of reasons, including:

1. It utilizes a minimum of hardware and DC power. Using the Qualcomm Q2334 Dual DDS and the Qualcomm Q3036 single chip PLL synthesizer, an octave band fine resolution synthesizer to greater than 1500 MHz can be built using three integrated circuits plus a bandpass filter and VCO.

2. Since no mixers are used, spurious performance is excellent, predictable, and reproducible.

3. The frequency resolution is typically on the order of 1 Hz or less.

4. The design permits rapid switching times and excellent noise performance.

A block diagram of the DDS-driven PLL is given in Figure 3. A filtered, limited DDS output is used as the reference to the PLL. An optional divider may be used to divide the DDS output to provide an appropriate center frequency for a particular filter technology (such as a crystal filter, where the range of center frequencies may be limited), or simply to improve its noise and spurious characteristics.

The operation of the DDS-driven PLL is as follows. The PLL is designed to output a range of frequencies with a frequency resolution equal to its reference frequency. For example, the PLL could output 200 to 400 MHz, and if the reference frequency applied to the PLL is 10 MHz, the possible outputs from the PLL are 200 MHz, 210, 220, 230, ..., 390, 400 MHz. Each PLL output is the division ratio of the loop (the counter "N" in Figure 3 times the reference frequency). By using a DDS as the reference to the loop, the reference frequency may be made to vary in extremely small steps. By selecting the bandwidth of the DDS output appropriately, the synthesizer may be made to have continuous frequency coverage with a resolution of N times the DDS frequency resolution. For continuous coverage, the DDS bandwidth must be

$$BW_{DDS} \geqslant \frac{\text{DDS Center Frequency}}{N_{min}} \qquad (5)$$

where N_{min} is the minimum PLL division ratio. Note that the step size of the DDS-driven PLL varies with N, and therefore is not constant throughout the range of the output frequencies.

To understand the noise and spurious performance of the DDS-Driven PLL the designer must begin with the DDS output characteristics and analyze the contribution of the PLL to this DDS reference. The DDS has excellent phase noise characteristics, but its spurious content is not as good as a high-grade reference oscillator. The DDS output has PM spurs due to phase truncation, primarily AM spurs due to amplitude quantization, and miscellaneous additive spurs due to intermodulation products caused by D/A non-linearities and clock feed-through. The performance of a PLL with AM, PM, and FM spurious tones plus additive noise on its frequency reference will be discussed as an introduction to the performance analysis of the DDS-Driven PLL.

A PLL with a division ratio N acts as a frequency multiplier for the process applied to its reference input. The loop acts as a 1st order bandpass filter of one-sided bandwidth equal to the closed loop bandwidth for this multiplier reference process. First we will discuss the performance of a ×N frequency multiplier with a noisy input. The 1st order response of the PLL to the reference input is superimposed upon these results.

Analysis of Frequency Multiplier with Noisy Input

For a frequency reference with AM spurious tones, consider an AM spectrum input to a frequency multiplier:

$$x(t) = A(1 + m(t))\cos\omega_1 t \qquad (6)$$

If the frequency is multiplied by $n = \omega_2/\omega_1$, the output of the frequency multiplier is:

$$y(t) = A(1 + m(t))\cos(n\omega_1 t) \qquad (7)$$

The modulation index is not changed: the AM spurs are no larger at the output of the multiplier than at the input. In fact, AM spurs may be suppressed by a limiter before or after frequency multiplication.

For a frequency reference with PM spurious tones, consider a carrier phase which is modulated by a sinusoid:

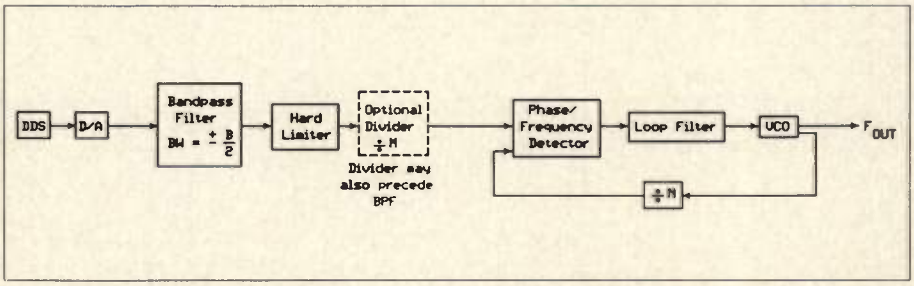

Figure 3. DDS-driven PLL.

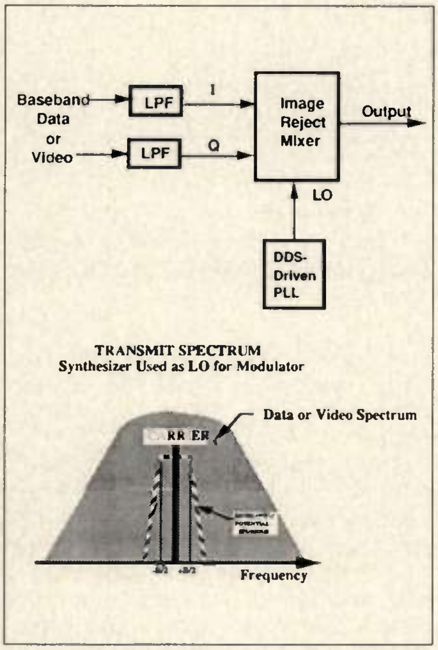

Figure 4. Fast hopping DDS-driven PLL, ideal for upconverting a wide data or video spectrum.

$$y_1(t) = A\cos(\omega_1 t + \beta\sin\omega_m t) \tag{8}$$

where β is the modulation index (β = maximum value of the phase deviation). The phase of $y_1(t)$ is:

$$\phi(y_1(t)) = \omega_1 t + \beta\sin\omega_m t \tag{9}$$

and the frequency is therefore

$$f_1(t) = \frac{1}{2\pi}\frac{d\phi}{dt} = \frac{1}{2\pi}(\omega_1 + \beta\omega_m\cos\omega_m t) \tag{10}$$

After we multiply this signal in frequency by $n = \omega_2/\omega_1$:

$$f_2(t) = nf_1(t) = \frac{1}{2\pi}(n\omega_1 + n\beta\omega_m\cos\omega_m t)$$
$$= \frac{1}{2\pi}(\omega_2 + n\beta\omega_m\cos\omega_m t) \tag{11}$$

We integrate this to determine the phase:

$$\phi(y_1(t)) = \omega_2 t + n\beta\sin\omega_m t \tag{12}$$

Therefore

$$y_2(t) = A\cos(\omega_2 t + n\beta\sin\omega_m t) \tag{13}$$

The modulation index is now $n\beta$ instead of β, but the sideband offset frequency f_m, which is equivalent to the modulation frequency, is not changed. The spectrum of the PM signal $y_1(t)$ is given by inspection from the equivalent relationship:

$$y_1(t) = A\cos(\omega_1 t + \beta\sin\omega_m t)$$
$$= A\sum_{i=-\infty}^{\infty} J_i(\beta)\cos(\omega_1 t - i\omega_m)t \tag{14}$$

where $J_i(\beta)$ are Bessel functions of the first kind. Therefore, the spectrum of $y_2(t)$ is determined by inspection from:

$$y_2(t) = A\sum_{i=-\infty}^{\infty} J_i(n\beta)\cos(\omega_2 t + i\omega_m)t \tag{15}$$

PM spurs centered about the carrier input to a frequency multiplier are increased in amplitude only: their offset from the carrier remains unchanged. For sinusoidal modulation, frequency multiplication by n increases the power of the PM spur at the offset $i \times f_m$ from the carrier by the ratio

$$\text{spur power ratio} = \left(\frac{J_i(n\beta)}{J_i(\beta)}\right)^2 \tag{16}$$

Now consider the DDS-Generated PM process, where the modulating waveform is a sawtooth of peak-to-peak amplitude $2\pi/2^P$ radians. In this case the result of frequency multiplication by n is to increase the amplitude of each spurious tone by n, with no effect upon the offset frequency. Therefore, frequency multiplication by n increases the power of DDS phase truncation spurs by n^2 (20log(n) dB), but does not change the frequency offset of these spurs from the carrier.

For a frequency reference with FM spurious tones, consider the FM modulated signal

$$Z_1(t) = A\cos(\omega t + 2\pi f_d \int^t m(x)dx) \tag{17}$$

where f_d is the frequency deviation constant. For $m(t) = \cos\omega_m t$,

$$Z_1(t) = A\cos(\omega_1 t + \frac{f_d}{f_m}\sin\omega_m t) \tag{18}$$

This is the identical analysis as for the PM case with sinusoidal modulation. Note that for FM, unlike PM, the modulation index is a function of the modulation frequency. With this in mind, the same situation occurs when an FM modulated signal is applied to a frequency multiplier as with a PM modulated signal. The modulation sidebands are increased in voltage as $J_i(n\beta)/J_i(\beta)$; the frequency offset from the carrier is not changed.

This section addresses discrete spurs which are caused by discrete spurious tones such as: harmonics of the reference frequency, and intermodulation products between these harmonics and other tones, clock feedthrough, power supply effects stray coupling, and so on. Symmetrical pairs of sidebands about the reference may be AM or PM or a mixture of both. The phase jitter contribution from a symmetrical pair of sidebands is zero if they are AM. From Reference 1, the phase jitter variance due to each pair of symmetrical PM side bands is

$$\phi^2_{total} = \phi_1^2 + \phi_2^2 + \phi_3^2 + \ldots \tag{19}$$

The phase jitter variance from each single (not part of a symmetrical pair) discrete line is given by

$$\phi^2_{total} = \phi_1^2 + \phi_2^2 + \phi_3^2 + \ldots \tag{20}$$

This makes the conservative assumption that the tones are PM in origin. The phase jitter due to discrete additive spurious tones is increased n times at the output of a ×n frequency multiplier. The power of each spurious tone is increased by n^2(20log(n)) dB. Therefore, discrete spurious tones are increased in power by n^2 after frequency multiplication by n; their offset from the reference frequency is not changed. The effect of frequency multiplication by n on a reference with phase noise is to increase the phase noise power by n^2(20log(n)) dB.

Reference 1 performs a detailed analysis of the effect of frequency multiplication on additive SSB and DSB white thermal noise. The results will be summarized here.

Superposed SSB white Gaussian noise of noise density N_O added to a clean carrier of power C can be separated equally into an AM component and a PM component. After frequency multiplication by n, the modulation index of the PM component is multiplied by n while the AM modulation index is not effected. For n >> 1, frequency multiplication turns additive SSB noise into primarily DSB phase noise. Multiplying frequency f_1 by n to give frequency f_2 yields the following carrier-to-noise density ratio:

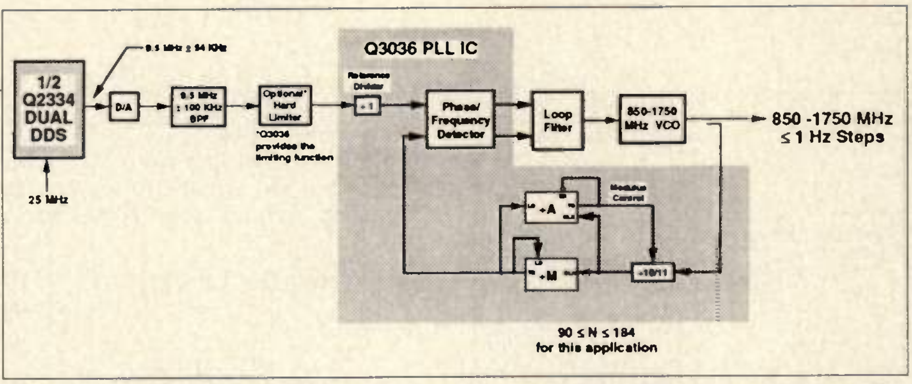

Figure 5. Block diagram of synthesizer using QUALCOMM Q2334 DDS and Q3036 PPL.

Additive SSB Noise

$$\left(\frac{C}{N_0}\right)_2 = \frac{4}{n^2}\left(\frac{C}{N_0}\right)_1 \quad \text{for } n >> 1 \qquad (21)$$

In the case of superposed DSB white Gaussian noise, the carrier-to-noise density ratios after frequency multiplication by n are:

Additive DSB Noise

$$\left(\frac{C}{N_0}\right)_2 = \frac{2}{n^2}\left(\frac{C}{N_0}\right)_1 \qquad (22)$$

Therefore, frequency multiplication by n multiplies phase noise power by n^2, and additive DSB white Gaussian noise by $n^2/2$.

DDS-Driven PLL Performance

We are now in a position to analyze the performance of the DDS-Driven PLL show in Figure 3. The performance is a function of the DDS characteristics, the BPF bandwidth, and the PLL parameters. Note that if switching speed is not important, the PLL loop bandwidth can be made extremely narrow. In this case, the phase noise and spurious performance of the output is essentially that of the VCO. The bandpass filter in Figure 3 can be replaced with a lowpass filter since the reference will have a minimal effect on the output spectrum. If the VCO is clean, this may be the simplest approach to a synthesizer with broad bandwidth, fine resolution, good spurious and phase noise performance (that of the VCO), small size, and low power, but with a slow switching speed between output frequencies.

If fast switching speed is important, the PLL loop bandwidth must be made correspondingly larger. In this case the Bandpass Filter, Hard Limiter, and Optional Divider in Figure 3 become important. In the absence of these three functions, the unmodified DDS output (with anti-aliasing filter) is multiplied by N in frequency by the PLL. Within the loop bandwidth, the following happens to noise processes at the DDS output:

1. Spurs due to Phase Truncation: These spurs are multiplied in power by 20log(N) dB: their frequency offsets from the carrier are not changed. They are filtered by 6 dB/octave by the PLL as a lowpass process from the closed loop bandwidth. Note that DDS PM spurs are increased in power by precisely 20log(N) dB, while a PM process with sinusoidal modulation would be increased in amplitude by a ratio of Bessel functions.

2. Spurs due to Amplitude Quantization: As discussed previously, these tones are AM spurs. Non-linearities may convert a portion of this process to PM. Since frequency multiplication by N does not effect the modulation index of AM, the nonlinear PLL circuitry will tend to suppress the AM spurs. The best thing to do is to hard limit the filtered DDS output to suppress the AM spurs, or the AM component of the noise process. The limiter is located prior to the PLL.

3. Spurs due to D/A Non-Linearities: These are additive discrete spurs, and therefore are increased in power by 20log(n) dB; their offsets from the carrier are not changed. They are filtered by 6 dB/octave by the PLL lowpass characteristic.

4. Phase Noise: The DDS phase noise is that of its reference improved by $20\log(f_{clock}/f_{output})$ dB. This improvement is limited by the noise floor of the DDS circuitry. The PLL frequency multiplication increases this DDS phase noise by 20log(N) dB within the closed loop BW, and filters it at 6 dB/octave outside the loop BW.

5. Additive Thermal Noise: Additive thermal noise generated by the DDS circuitry is multiplied by the PLL by $n^2/2$ or +[20log(N)-3] dB.

It is important to note that the PLL increases the amplitude of the spurious tones and not their frequency offsets from the reference. The bandpass filter in Figure 3 ensures that the DDS-generated spurious tones and noise input to the PLL will be confined to its bandwidth of ±B/2. After frequency multiplication by N, the noise and spurious tones will be increased by 20log(N) dB, but only within ±B/2 of the output frequency. Beyond this, the spurs and noise are suppressed by the skirts of the BPF. In essence, the BPF serves as a tuneable high-frequency tracking filter, but it is actually a fixed, low-frequency design. The output spectrum is a clean tone surrounded by a pedestal of noise and spurs. The pedestal width is ±B/2. The noise and spurs fall off rapidly beyond this point due to the BPF and the first order response of the loop. If a narrow bandwidth crystal filter is used, the pedestal can be made extremely narrow.

The selection of the BPF bandwidth and center frequency is a tradeoff between switching speed, noise performance, and the need for continuous frequency coverage. Good phase noise and anti-microphonic design practice dictate that N be small, and consequently that the reference frequency be high. A small N requires a relatively large BPF bandwidth to provide continuous frequency coverage (refer to equation 5). Certain filter technologies may not be available at the desired reference frequency. Also note that a narrow BPF bandwidth also adversely effects switching speed. Therefore, the selection of the BPF center frequency and bandwidth is a balancing act between these issues.

The resulting spectrum can be made nearly spur-free when observed over large frequency spans, but has the noise pedestal when observed close-in. This may not be acceptable for instrument-grade applications. For these applications an extremely narrow loop BW can be used to eliminate the pedestal (at the expense of switching speed), or another synthesizer topology can by used. This topology may be ideal, however, in a communications application in which the spectrum being transmitted is wider than the synthesizer noise pedestal.

As shown in Figure 4, the DDS-Driven PLL appears spur-free when used to upconvert a relatively wide data or video spectrum. If the demodulation circuitry has no problem demodulation a signal with close-in spurious tones (they are typically 20-35 dBc), this synthesizer approach may be ideal.

The hard limiter in Figure 3 suppresses the AM spurious components from the DDS. These components are the AM amplitude quantization process, as well as the AM portion of the DDS thermal noise. As AM spurs have an unpredictable effect upon PLL output

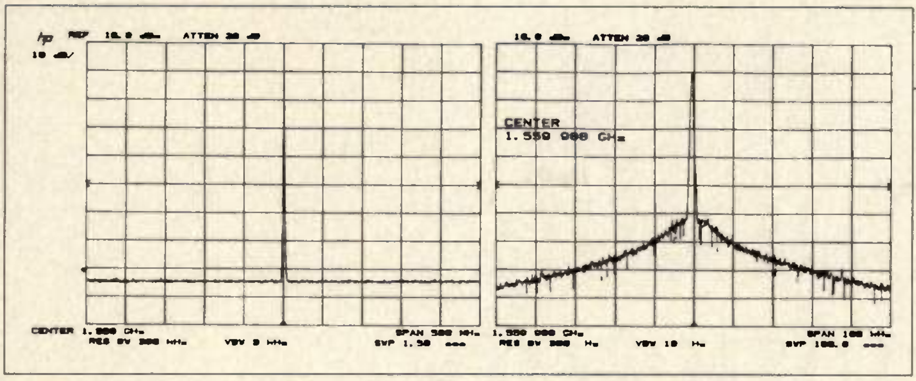

Figure 6. Measured performance of DDS-driven PLL in Figure 5.

performance, it is best to eliminate them. In practice, the hard limiter provides a noticeable improvement in synthesizer performance.

The optional divider in Figure 3 performs two functions. First, it improves the spurious and noise performance of the DDS, in general by 20log(M) dB. Second, it provides flexibility in selecting the reference frequency to the PLL and the center frequency of the crystal filter. This is because the DDS output and PLL reference input are at different frequencies. The resulting extra degree of freedom may help in the specification of the BPF response.

Design Example

A design example using the QUALCOMM Q2334 Dual DDS and Q3036 PLL integrated circuits is shown in Figure 5. The Q2334 DDS is ideal for use in the DDS-Driven PLL because of the following properties:

1. The algorithmic sine lookup function provides phase truncation spurs no larger than -84 dBc (14 equivalent bits of phase). No external lookup memory is required. No other commercially available DDS has comparable performance.
2. The proprietary noise reduction circuit provides excellent quantization performance using an 8-bit D/A converter. Since the AM component of quantization noise is suppressed by the hard limiter anyway, there is no reason to use more than an 8-bit D/A (given sufficient D/A linearity). This results in lower cost, smaller size, and lower power.
3. The Q2334 accepts a clock up to 50 MHz, permitting the use of high reference frequencies.
4. The Hop Clock input permits synchronous frequency hopping.
5. The device uses low power CMOS technology and operates from a single +5V supply.

A block diagram of the synthesizer which operates between 850 and 1750 MHz is shown in Figure 5. The frequency resolution is 1.0 Hz max. (it varies between 0.52 Hz and 1 Hz across the band). The DDS output frequency is 9.5 MHz ±54 kHz, and N in the PLL varies between 90 and 184. The PLL loop bandwidth is $\omega_n \approx 3$ kHz. The design parameters for this synthesizer are: Max. phase truncation spur within BPF and loop

BW = -84 + 20log(184) dBc = -39dBc

Amplitude quantization spurs: Suppressed by hard limiter. Max. spurs due to D/A non-linearities within BPF and loop

BW = -68 +20log(184) dBc = -23 dBc

Note: Careful choice of DDS center frequency can reduce this result.

Actual output spectra from this synthesizer are shown in Figure 6. Note that this synthesizer totals 4 integrated circuits (including op amp), a VCO, plus several discrete components.

Conclusion

Direct digital and phaselock loop frequency synthesis are two very different approaches which can be combined to make powerful hybrid synthesizers. Three DDS/PLL hybrid synthesizers have been described: the DDS/Direct Analog Synthesizer hybrid, the PLL with DDS-Generated Offset, and the DDS-Driven PLL. Each of these hybrids offers fine-resolution frequency synthesis over broad bandwidths.

Perhaps the most important conclusion is that frequency synthesizers should not be designed by rules of thumb. PLL stability must be guaranteed by analysis, not by placing desired extra poles at "n" times the natural loop frequency. The DDS does have spurious responses, but the brute force approach to minimizing these need not always be applied. For example, a hard limiter suppressing quantization noise may eliminate the need for a D/A converter of greater than 8 bits. Finally, the PLL does not treat all spurious and noise processes at its reference input equally. In fact, of the 7 noise processes discussed- AM spurs, DDS quantization noise, sinusoidal PM and FM, DDS phase truncation spurs, phase noise, and additive thermal noise- only 3, DDS phase truncation spurs, phase noise, and discrete PM spurs strictly follow the +20log(N) rule commonly cited.

The hybrid approaches presented can be implemented using a minimum number of components (using VLSI DDS and PLL circuits), and therefore can minimize the effects of printed circuit layout on synthesizer performance. This enhances the correspondence between predicted and actual performance and therefore requires less performance optimization on the bench. *RF*

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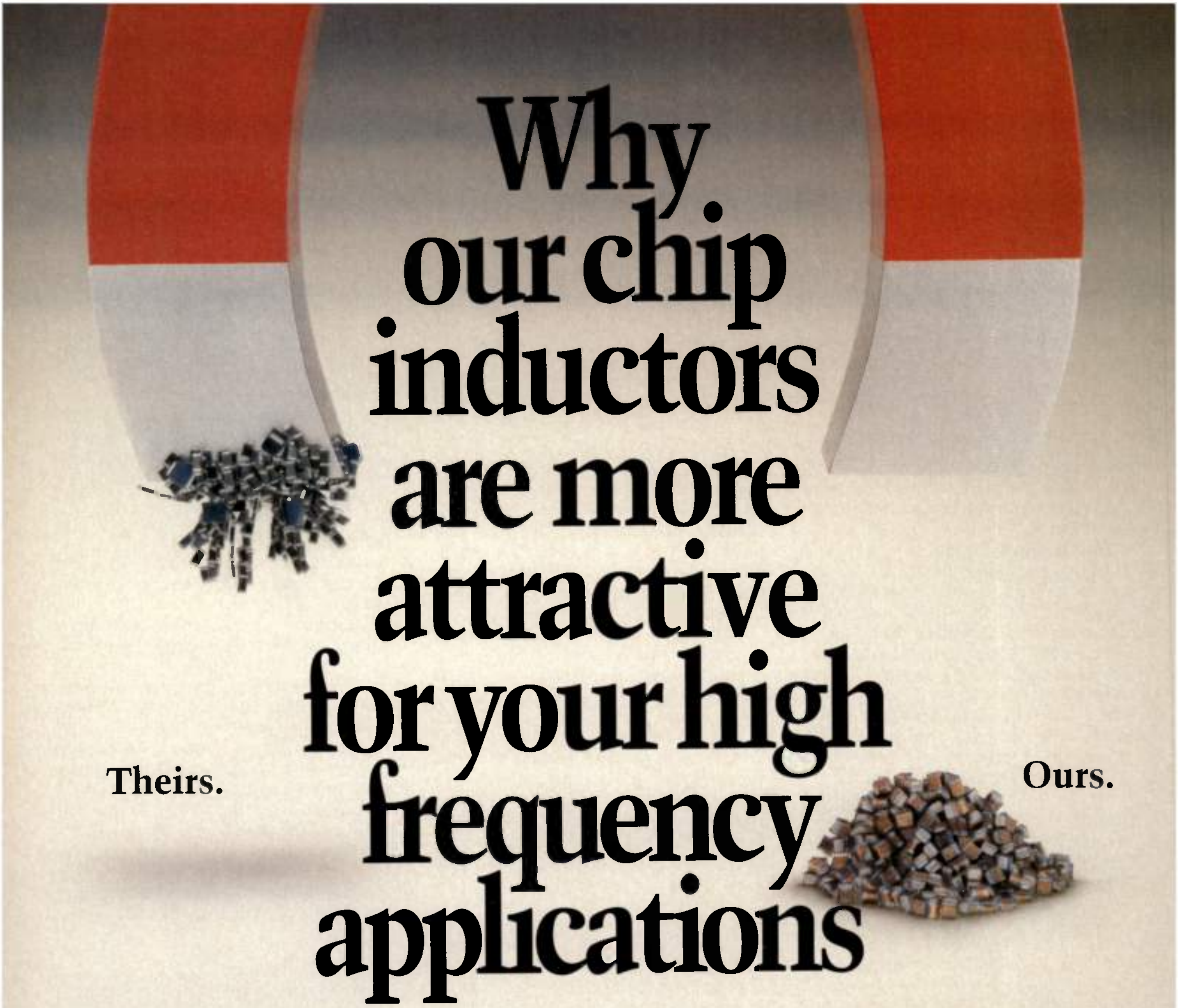
Author Information

Rob Gilmore (above) is the Engineering Director for QUALCOMM, Inc., 10555 Sorrento Valley Road, San Diego, CA 92121. Richard Kornfeld is an Engineering Manager at QUALCOMM. They can be reached at (619) 587-1121.



The point of this little demonstration is that Coilcraft surface mount inductors are made of ceramic. A decidedly non-magnetic material.

Most other chip inductors are made of ferrite. Which is great for demonstrating the principles of magnetism, but not so hot for high frequency magnetics.

Take self resonance, for example. SRFs on our coils are up to 3 times higher than equivalent ferrite chips. And located a safe distance away from your operating frequency.

The actual inductance you'll get with Coilcraft chips at higher frequencies is very predictable and consistent. Not so with ferrites. Beyond the test frequency, their inductance curves rise steeply and vary significantly from part to part.

Coilcraft ceramic chips also have a low temperature coefficient of inductance: +25 to +125 ppm/°C, depending on inductance. TCLs on ferrite chips are often two to four times higher!

And if you need close tolerance parts, we offer even more advantages. Thanks to our computer-controlled manufacturing and ceramic's neutral properties, it's easier for us to make 5% or 2% parts. We can even production-test at your operating frequency! Other chip makers have to cope with ferrite's permeability variations, so their yields are lower. Which means delivery can be unpredictable.

So next time you're selecting surface mount inductors, forget the ferrite and stick with Coilcraft ceramic chips.

For complete specifications and information on our handy Designer's Kits of sample parts, circle the reader service number. Or call **800/322-COIL**.

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INFO/CARD 30

Results of the 1990 RF Design Awards Contest

By Gary A. Breed
Editor

Once again, we have a group of engineers who have earned our recognition for their design achievements. This year's entries required an extra effort on the part of the judges, due to the overall excellent level of quality. Thanks to the hard work of Consulting Editor Andy Przedpelski and the 1989 winner, Dan Baker, the task of judging was completed without any problems. With the results in hand, we are pleased to present the winners of the Fifth Annual RF Design Awards Contest!

Just north of Philadelphia in Ambler, Pennsylvania is Interspec, Inc., a manufacturer of medical ultrasound imaging equipment. In order to achieve the best possible image quality, their latest unit required a fast, quiet switch in its analog front-end circuitry. Senior Engineer Kevin Randall developed that switch, and entered it in this year's contest. As a result, Kevin has won the Grand Prize, an Advantest R3261A spectrum analyzer, provided through the courtesy of Advantest America, Lincolnshire, Illinois. The winning design is featured on page 29, and should prove interesting to many *RF Design* readers.

Second Place was awarded to Mitch Randall (no relation to Kevin!), an RF engineer working on radar systems in the Atmospheric Technology Division of the National Center for Atmospheric Research in Boulder, Colorado. Mitch's design, a digital frequency divider with reduced harmonic content, was developed primarily as a laboratory experiment, but with potential applications in mind. Frequency dividers certainly fit with his current work on an X-band frequency synthesizer. Mitch wins a complete set of software for oscillators, filters, transmission lines and circuit analysis from Eagleware, Inc. (formerly Circuit Busters). This design will appear in the August issue of *RF Design*.

A circuit for cancellation of PLL reference sidebands earned co-authors John MacConnell and Richard Booth the Third Prize in this year's contest. Their circuit, which provides a relatively simple way to achieve up to 40 dB reduction in reference spurs, has won them a Coilcraft SMT test fixture for interfacing SMT devices to test instrumentation, a set of specialized trimmer adjustment tools from Sprague-Goodman Electronics, and a set of five Designers' Kits of various inductors from Coilcraft. The September issue will feature this circuit.

John Heitman of Advantest presents the Grand Prize to winner Kevin Randall.

Honorable Mention Winners

Eleven more entries have been recognized with Honorable Mention awards. They are listed in the sidebar to this report. One of this group, Rob Gilmore's PLL/DDS combination, has been published this month on page 41. These winners receive a set of five Designers' Kits from Coilcraft and a set of adjustment tools from Sprague-Goodman.

We will continue to publish a Design Awards entry every month. All prize winners will be published between now and next July, when the next winners will be announced. We may also publish some of the entries that just missed winning one of the prizes, but represent interesting design ideas.

The Prizes

Special thanks is due to the companies who provided the prizes awarded to this year's contest winners! Advantest America, Inc. generously provided their model R3261A spectrum analyzer, a highly desirable prize for an RF designer! Thanks go to John Heitman, George Neeno, and Vice President F. Kamei for their support of the RF engineering community.

Eagleware provided a complete package of software for the Second Place


2nd place winner Mitch Randall will enjoy a complete software package.

prize winner, including their Oscillator, Filter, TLine and Super Star programs. President Randy Rhea has earned the appreciation of *RF Design* for his contribution.

Coilcraft provided a total of 60 Design Kits, plus an SMT device test fixture for our Third Place and Honorable Mention winners. Paul Liebman can accept our thanks for Coilcraft's encouragement and support for the Design Awards Contest. Also supporting this group of winners is Sprague-Goodman Electronics, who offered a dozen sets of eight specialized trimmer adjustment tools. Thanks to Martin Markson and Jack Goodman for their support.

With the support of RF companies like these, the RF Design Awards Contest continues to be a success.

The 1991 Contest

The announcement of the 1991 RF Design Awards Contest will come next month. Prizes are guaranteed to be exciting, and we might find a few new ideas to make the contest even more exciting. ***RF***

The Winning Entries

Grand Prize
A Lowband Quiet Switch
Kevin S. Randall
Interspec, Inc.
Ambler, PA

Second Prize
Harmonic Suppressing Digital Frequency Divider
Mitch Randall
National Center for Atmospheric Research
Boulder, CO

Third Prize (co-authors)
A Feedback Method for Reference Spur Reduction in Phase Locked Loops
John W. MacConnell
Santa Clara, CA, and
Dr. Richard W.D. Booth
Lawndale, CA

Honorable Mention
Hybrid PLL/DDS Frequency Synthesizers
Rob Gilmore
QUALCOMM, Inc.
San Diego, CA

Constant Reactance Voltage Controlled Oscillator
Raymond Page
Wenzel Associates, Inc.
Austin, TX

Log Fidelity Test Fixture
James D. English
ABC Circuit Design
Aloha, OR

A Linear Driftless VCO
Luis Cupido, Eng.
C & TC
Aveiro, Portugal

A Novel, Wide Band, Crystal Controlled FM Transmitter
Thomas Xydis
Ann Arbor, MI

SSF — Swept Spectrum Bandpass Filter
Roy H. Probst
University of North Carolina
Chapel Hill, NC

Parasitic Positive Feedback Frequency Acquisition in PLL
Jonathon Y.C. Cheah
Hughes Network Systems
San Diego, CA

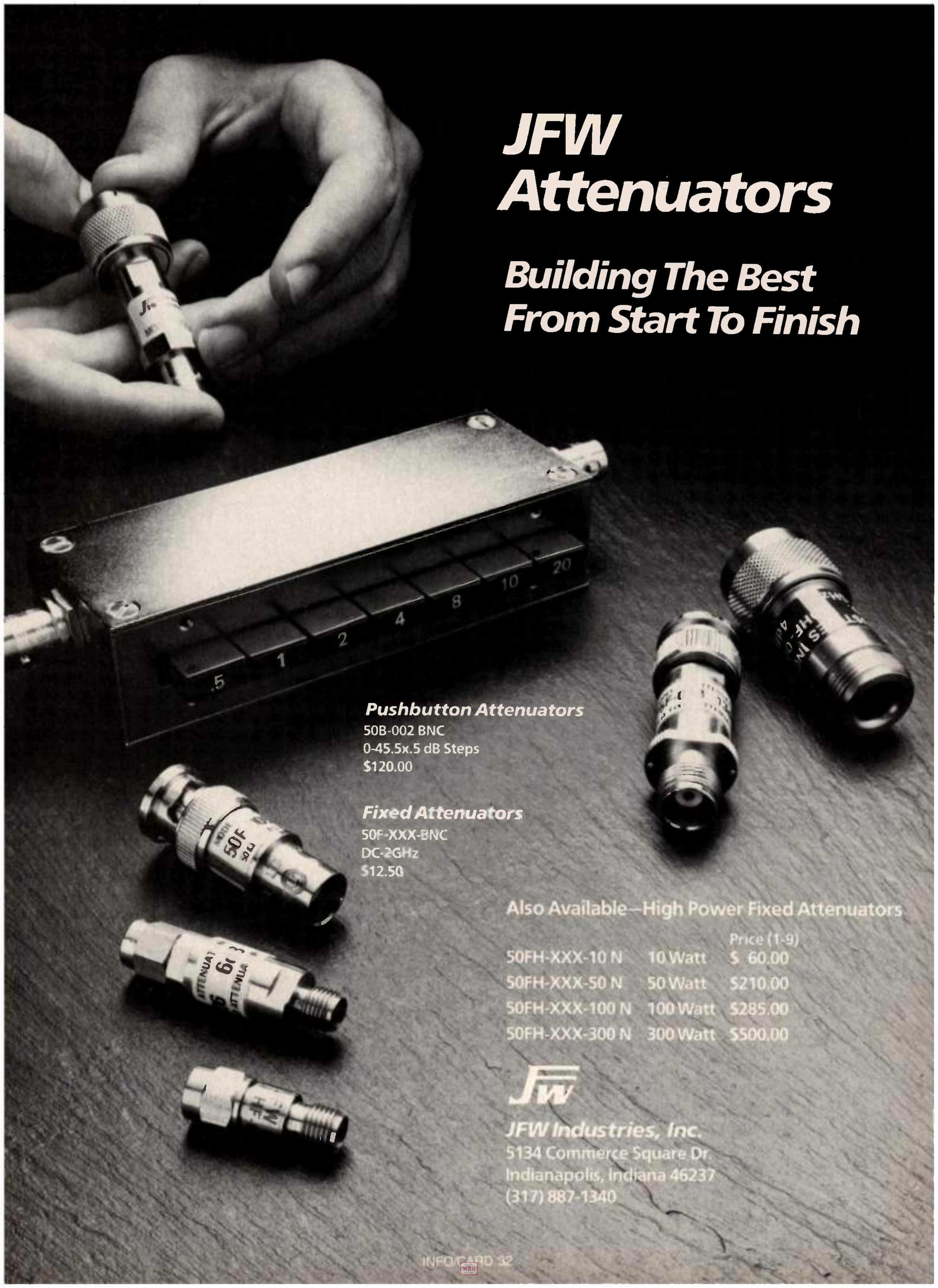
Simple Clock and PSK Carrier Recovery
Francois Methot
Consortel Limited
Beaconsfield, Quebec

Crystal/Transformer Networks for Filtering
Gerald Malizewski
SAIC Range Systems
San Diego, CA

Add Negative Capacitors to Your Component Repertory
Gary Thomas
GE Mobile Communications
Lynchburg, VA

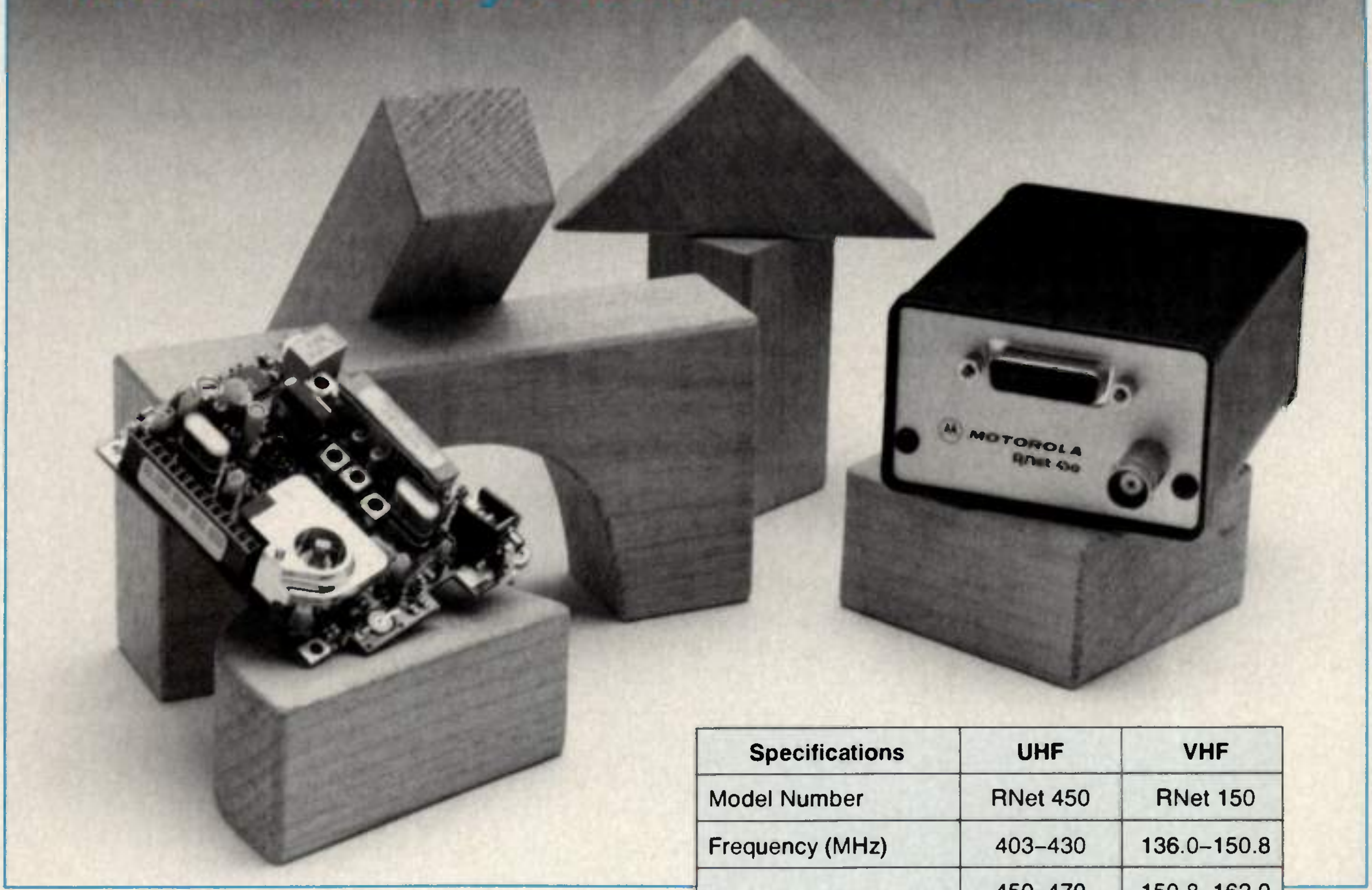
Low Cost UHF AM Receiver
Bob Friday
Metricom
Campbell, CA

INFO/CARD 31


JFW
Attenuators
Building The Best
From Start To Finish
.5
1
2
4
8
10
20
Pushbutton Attenuators
50B-002 BNC
0-45.5x.5 dB Steps
$120.00
Fixed Attenuators
50F-XXX-BNC
DC-2GHz
$12.50
Also Available—High Power Fixed Attenuators
Price (1-9)
50FH-XXX-10 N 10 Watt $ 60.00
50FH-XXX-50 N 50 Watt $210.00
50FH-XXX-100 N 100 Watt $285.00
50FH-XXX-300 N 300 Watt $500.00
JFW Industries, Inc.
5134 Commerce Square Dr.
Indianapolis, Indiana 46237
(317) 887-1340
INFO/CARD 32

Building Blocks

RNet™ Telemetry Radios from Motorola



Specifications	UHF	VHF
Model Number	RNet 450	RNet 150
Frequency (MHz)	403–430	136.0–150.8
	450–470	150.8–162.0
		162.0–174.0
Nominal Power Supply	10 VDC to 17 VDC	
Current Drain		
Standby:	16mA	20mA
Receive:	20mA	24mA
Transmit:	865mA (2W)	870mA (2W)
	1120mA (4W)	1275mA (4W)
Dimensions	3.3″ x 2.7″ x 1.52″	
RF Output-TX	2W or 4W	2W or 5W

For years, design engineers seeking solutions have been turning to Motorola for product support. Now, as you envision new opportunities within the context of a wireless society, Motorola is there again with building block RF components that will enable you to move from design to production . . . the RNet™ Series of telemetry transceivers and OEM radio modules.

It doesn't matter if you are dealing with remote control, RF Data Collection, alarm reporting, status monitoring, metering or SCADA applications, Motorola's heritage of quality engineering and world class products provides you with voice and data capable telemetry radios that will fit your project.

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- Made in the U.S.A. for faster delivery
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Envisioning . . . that's your job.
Enabling . . . that's ours.

For more information and detailed specifications on RNet™ telemetry radios and OEM radio modules,

call 1/800-624-8999
Ext. 105

MOTOROLA
Radius Division

1301 East Algonquin Road • Schaumburg, IL 60196

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Specifications subject to change without notice.
*Telemetry radio models only.

INFO/CARD 33

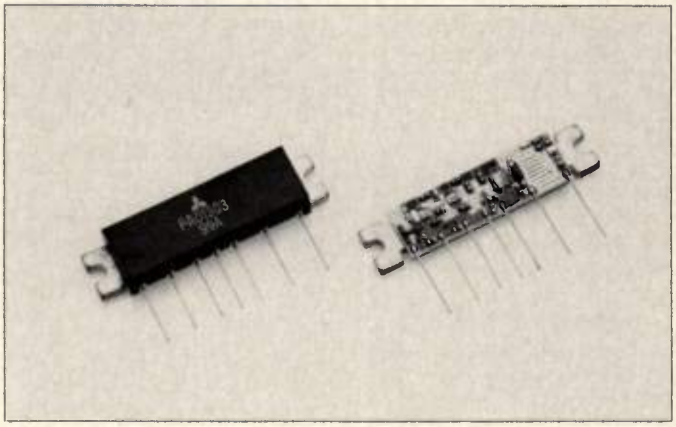
GaAs FET RF Power Module for Handheld Cellular Telephones

Mitsubishi now offers a GaAs FET RF power module specifically designed for transportable and handheld cellular phones. The FA01303 output module provides the small size (45 mm by 12 mm by 7 mm) and high-efficiency operation needed for transportable and handheld cellular phones. The GaAs FET structure of the device results in 50 percent efficiency, thereby extending battery life in these applications. With an input signal of 0.0 dBm (1mW), the module has a rated output power of 32 dBm (minimum), and operates in the U.S. cellular phone market frequency range of 824 to 849 MHz.

For customers who manufacture and market transportable and handheld phones to foreign markets, Mitsubishi offers the device in the ETACS (European Total Access Communication System) frequency range of 872-905 MHz, as well as the Nordic Mobile Telephone System (NMT) frequency range of 890-915.

The Mitsubishi FA01303 GaAs FET RF power module is priced at $40.67 in quantities of 1,000.

Mitsubishi Electronics Semiconductor Division
INFO/CARD #200



VXI Power Meter

The new HP E1416A VXI power meter from Hewlett-Packard Company offers measuring power equivalent to the HP 437B in a single-slot, C-size VXI configuration. The meter is compatible with the HP 8480 family of 17 power sensors that covers 100 kHz to 50 GHz and −70 to +44 dBm. The meter accepts ASCII Test & Measurement Systems Language (TMSL) commands. It has a specified ± 0.02 dB or ± 0.5 percent instrumentation accuracy and uses a 1 mW, 50 MHz power reference on the front panel for sensor calibration. The HP E1416A has selectable resolution from 0.1 to 0.001 dB and automatic filter mode, duty-cycle computation and user-entered, sensor-calibration factor tables. The ATE user can configure it to detect power at more than one point in the system under test. The HP E1416A meter is priced at $2,500. Delivery is estimated at 12 weeks ARO.

Hewlett-Packard Company
INFO/CARD #199

High Power Variable Capacitors

Kilo-Tec announces the availability of three new high power variable capacitors. The new capacitors are designed and manufactured in England and use high quality NS4 Aluminum, Brass and Military quality moldings in their construction. New models include: the TC-150, a 150 pf, 9.8 KV breakdown voltage capacitor with a 3mm air gap and weight of 350 grams; the TC-250, a 250 pf, 14.7 KV breakdown voltage capacitor with a 4.5mm air gap and weight of 620 grams; and the TC-750, which is rated at 750 pf with 7.8 KV voltage breakdown, 2mm air gap. These high quality capacitors are suited for construction of RF amplifiers, RF tuning units, and RF transmitters. For commercial applications, a version using ceramic end plates is available. Prices will be announced soon.

Kilo-Tec
INFO/CARD #198

Wide Dynamic Range Amplifier

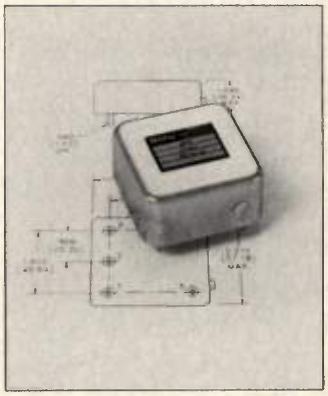
Microwave Solutions introduces a series of wide dynamic range amplifiers that offer low noise figure and medium output power over popular communications bands. Model MSD-1142501 operates in the frequency range of 750-1,000 MHz with 12 dB minimum gain, mode MSD-2142501 operates in the 950-1,250 MHz range with 12 dB gain, MSD-2132502 operates in the frequency range of 1.4-1.7 GHz with 12 dB gain, MSD-3142501 operates in the frequency range of 1.7-2.3 GHz with 10 dB gain, and MSD-3243201 operates in the 2.0-2.6 GHz range with minimum gain of 15 dB. They draw 100 mA at +15 VDC. Input/output VSWR is 1.5/1.5. Maximum noise figure is from 1.8 dB to 2.2 dB depending on the model used.

Minimum power output is +20 dBm. DC power applied through the output port may be specified and internal voltage regulation to protect against over-voltage and reverse voltage is provided.

Microwave Solutions, Inc.
INFO/CARD #197

Compact OCXO

The N60B OCXO features a compact, low profile case (2 1/4" by 2 1/4" by 1" nominal) designed to permit application in restricted mounting space. Mounting is by standard 0.040" through-



hole terminals. The new oscillator is available at any frequency in the range of 5 MHz to 15 MHz, and has standard stability of ± 5 × 10^{-9} at 0 to 50 degrees Celsius, or optionally ± 1 × 10^{-8} at −20 to 70 degrees Celsius. Frequency can be adjusted over a minimum range of ± 1 × 10^{-6} and is settable to 1 × 10^{-8} by means of an external access control. Aging is 1 × 10^{-9}/day.

Bliley N60B has standard minimum sine wave output of 0 dBm with a 50 ohm load. Sine wave output as high as +7 dBm at 50 ohms are optional, as are TTL and HCMOS. Required power is 6 watts at turn-on, and 1.5 watts at 25 degrees Celsius.

Bliley Electric
INFO/CARD #196

Compact AM and FM Receivers

The ML521, ML522, and ML524 portable measurement receivers from Anritsu are compact and battery operated, making them well suited to measuring AM and FM signals in a wide variety of field strength applications. The receivers measure 2.4" x 8.3" x 10" and weigh 6.6 lbs. The ML521 operates from 25 to

300 MHz, the ML522 from 300 to 1,000 MHz, and the ML524 from 25 to 1,000 MHz. The ML524 can cover up to 3 GHz through use of Anritsu's MH669B frequency converter. The double superheterodyne receivers use synthesized local oscillators to produce stability of $\pm 1 \times 10^{-6}$, and dynamic range is between 80 and 100 dB, depending on the model. Measurement accuracy is up to ± 2 dB and resolution is 12.5 kHz in selectable bandwidths of 15 to 120 kHz.
Anritsu America, Inc.
INFO/CARD #195

Magnetic Field Probes

Narda has introduced two new magnetic field probes for the 8700 Series of meters and electric field probes. Models 8731 and 8733 probes measure the magnetic field strength of electromagnetic fields in the frequency range of 10 to 300 MHz. Unlike traditional magnetic field probes, the 8730 series measure only the magnetic field components and do not respond when utilized in fields with higher than 300 MHz emitters. Previously available magnetic field probes could overestimate the magnetic field strength by as much as 10 dB when used at higher than specified frequencies.
Loral Microwave-Narda
INFO/CARD #194

Fixed Frequency SAW Oscillators

These SAW oscillators come in the frequency range of 200-1200 MHz, higher frequencies are possible by using a SAW resonator to multiply the oscillator frequency. The oscillators have an RF output power of +10 dBm and SSB Phase Noise < −105 dBc/Hz at 1 kHz offset. The oscillators come in TO-39 packages and are available in commercial and military versions.
Crystal Technology, Inc.
INFO/CARD #193

1.5 to 30 MHz Linear Power Amplifier

Power Systems Technology, Inc. announces the development of model BHE1637-500, 1.5 to 30 MHz frequency range at an RF power output of 500 watts. Class AB operation with a dynamic range of greater than 40 dB can be AM modulated with less than 10 percent distortion. The power amplifier has an instantaneous bandwidth of 28.5 MHz and a minimum RF gain of 57 dB. The amplifier comes with a 208 VAC ± 10 percent 50/60 Hz 3-phase power supply.
Power Systems Technology, Inc.
INFO/CARD #192

Crystal Clock Oscillators

The MS series VCXOs are available with center frequencies from 183 Hz to 32.768 MHz. Lead configurations for auto-insertion and surface mount assembly are available. Long term reliability is better than 20 million hours MTBF. Control voltage range is 0.5 to 4.5 volts DC, and control voltage bandwidth is 5 kHz minimum. All MS oscillators feature on board tri-state output and

symmetry optimization for TTL or HCMOS loads.
M-tron Industries, Inc.
INFO/CARD #191

Miniature GaAs SP4T

The model DSO864 GaAs SP4T switch has an integrated TTL driver and terminations, and operates from 5 to 2000 MHz. This device has insertion loss of 1.3 dB up to 1 GHz and 1.8 dB up to 2 GHz. Isolation is 65 dB up to 100 MHz, 50 dB up to 1 GHz, and 40 dB up to 2 GHz. Transients are typically less than 28 mV peak. The DSO864 is packaged in a 16 pin DIP and uses only 0.3 mA from a +5 VDC supply.
Daico Industries, Inc.
INFO/CARD #190

Quartz Crystals

Anderson Electronics, Inc. has a wide variety of crystals for the computer and communications markets. Fundamental, third overtone and fifth overtone mode AT cut crystals are available in package styles to fit many applications. Crystals made to MIL-specs are available. Temperature ranges available are 0 to 70, −30 to +60, −40 to +85, and −55 to +125 degrees Celsius.
Anderson Electronics, Inc.
INFO/CARD #189

New Design Method for Analog Bipolar ASICs

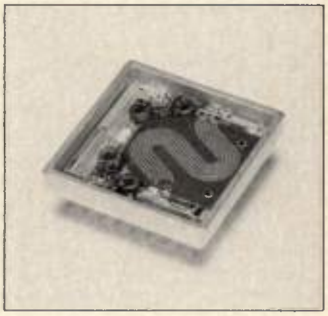
QuickTile™, a new design method for ASICs, is being introduced by Tektronix, Inc. The QuickTile method is an extension of the QuickChip™ 6 family of semicustom analog arrays. It features NPN transistor speeds (f_T) of up to 8.5 GHz, P-channel JFETs, and Vanadium Schottky diodes. Each tile consists of a fixed set of unconnected devices that are arranged for convenient interconnection to provide a functional increment to a larger IC design.
Tektronix, Inc.
INFO/CARD #188

Radar and GPS Oscillator

The model 2870052 from Piezo Crystal Company is designed for military radar and GPS markets and has a frequency range from 30 to 100 MHz. The aging rate is 3×10^{-9}/day. It has a frequency stability of $\pm 1 \times 10^{-8}$ over a temperature range of −45 to +85 degrees Celsius. The approximate price for this model is $500 to $600 in quantities.
Piezo Crystal Company
INFO/CARD #187

Miniature L-Band Quadraphase Demodulators

The IQF-4F-1500 and IQP-4S-1500 are able to cover a full octave from 1 to 2 GHz with typical balance figures of ± 4 degrees and 0.5 dB, and are considerably better over narrower bands. These demodulators are available in SMA connector pack-



ages, or in a one inch square, standard low profile package. This includes the Hi-Rel designs for space and military environments.
Merrimac
INFO/CARD #186

Ultrasonic Transducer Crystals

A new line of standard ultrasonic transduce crystals, called STAN-TRAN, is now available from Valpey-Fisher Corporation. The piezoelectric crystals are available in three materials, five diameters, and in frequencies from 1 to 30 MHz.
Valpey-Fisher Corporation
INFO/CARD #185

GPS/CDS-20 Clock System

Versitron's GPS/CDS-20 is a CDS-20 system with a GPS receiver. It improves 2σ clock jitter to less than 0.2 percent to 1.544 MHz (T1 rate) and 1.0 percent to 2.048 MHz (European T1 rate). It receives from the network of U.S. government global positioning satellites.
Versitron
INFO/CARD #184

Synthesized Signal Generator

The model 5230A synthesized signal generator from Eaton Corporation operates over the frequency range of 10 to 2560 MHz and offers fast switching with spectral purity. Some features are absolute phase noise of −125 dBc at 640 MHz, 1 kHz offset, spurious lower than −90 dBc, harmonics of less than −55 dBc, and fixed output level of 16 dBm.
Eaton Electronic Instrumentation Division
INFO/CARD #183

Broadband GaAs MMIC Switches

The HMMC-2006 chip is a broadband (DC-6 GHz) SPDT reflective switch that has low insertion loss and input/output match to 50 ohms over the full dc to 6 GHz band. The device can be used primarily in EW, ECM, radar, instrumentation, and communications systems. Prices range from $53 for 10 to 24 to $34 for 1,000 to 2,499.
Hewlett-Packard Company
INFO/CARD #182

70 MHz SAW Bandpass Filter

The model FB70-20.5 from Phonon Corporation has a bandwidth of 20 MHz with a center frequency of 70 MHz. It has a minimum rejection of 45 dB and insertion loss of 27 dB. It comes in a TO-8 package. Price $25 in quantities of 20,000.
Phonon Corporation
INFO/CARD #181

70-700 MHz, 100 MHz BW Synthesizers

The Mini-synthesizer, S-100 series offers 100 MHz bandwidth coverage from 70 to 700 MHz in a 25 kHz step size with a serial programming format. Output power is +3 dBm across the band and spurious outputs are better than −60 dB below the carrier. A lock detect signal is available on the S-100 series, which is TTL compatible with a logic 1 indicating a locked condition..
Z-Communications, Inc.
INFO/CARD #180

Modular Integrated Step Attenuator

Model PS604 miniature digital attenuator operates over the frequency range of 50 to 1000 MHz with an attenuation range of 2 to 30 dB in four steps. Insertion loss measures 3.0 dB maximum and switching speed is 2.0 microseconds. It operates over the −55 to +125 degree Celsius temperature range.
Phoenix Microwave Corporation
INFO/CARD #179

RF Design Software Service

Computer programs from *RF Design*, provided on disk for your convenience. All disks are MS-DOS/PC-DOS compatible, unless otherwise noted.

This Month's Disk:

Disk RFD-0790: July 1990
BASIC programs for opamp active filters: Sallen-Key highpass and lowpass filters, multiple-feedback bandpass filters, and first-order allpass phase shift networks. Not from published articles. References and circuits are provided on disk.

From Last Month:

Disk RFD-0690: June 1990
1. "Curve Fitting Made Easy," by Brian Miller. BASIC program used as example in his article. [Text file only—program is for HP9836, not MS-DOS]
2. "Twisted Wire Transmission Lines," by Douglas Linkhart. Performs calculations for line impedance determination. [BASICA or GW-BASIC]

Still Available: RFD-1989-SET
All 11 disks from 1989 issues of *RF Design* — $74.00 (5¼") or $82.00 (3½"). Price includes postage. Foreign orders add $8.00 (outside U.S. and Canada)

Send for a complete listing of available programs, or circle Info/Card below.

Disks are $9.00 each (5¼ in.) or $10.00 (3½ in.). Outside U.S. and Canada, add $8.00 per order. Foreign checks must be in U.S. funds, and must be payable in the U.S. Prices include postage and handling.

Annual subscriptions are available, providing 13 disks for $90.00 (5¼ in.) or $100.00 (3½ in.). Specify starting date. For subscribers outside the U.S. and Canada: add $50.00.

Payment must accompany order...specify disks wanted...send check or money order to:

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

INFO/CARD 65

Semiconductors made easy.

The definitive semiconductor catalog from Frequency Sources.

Over 100 pages of information on state-of-the-art devices.

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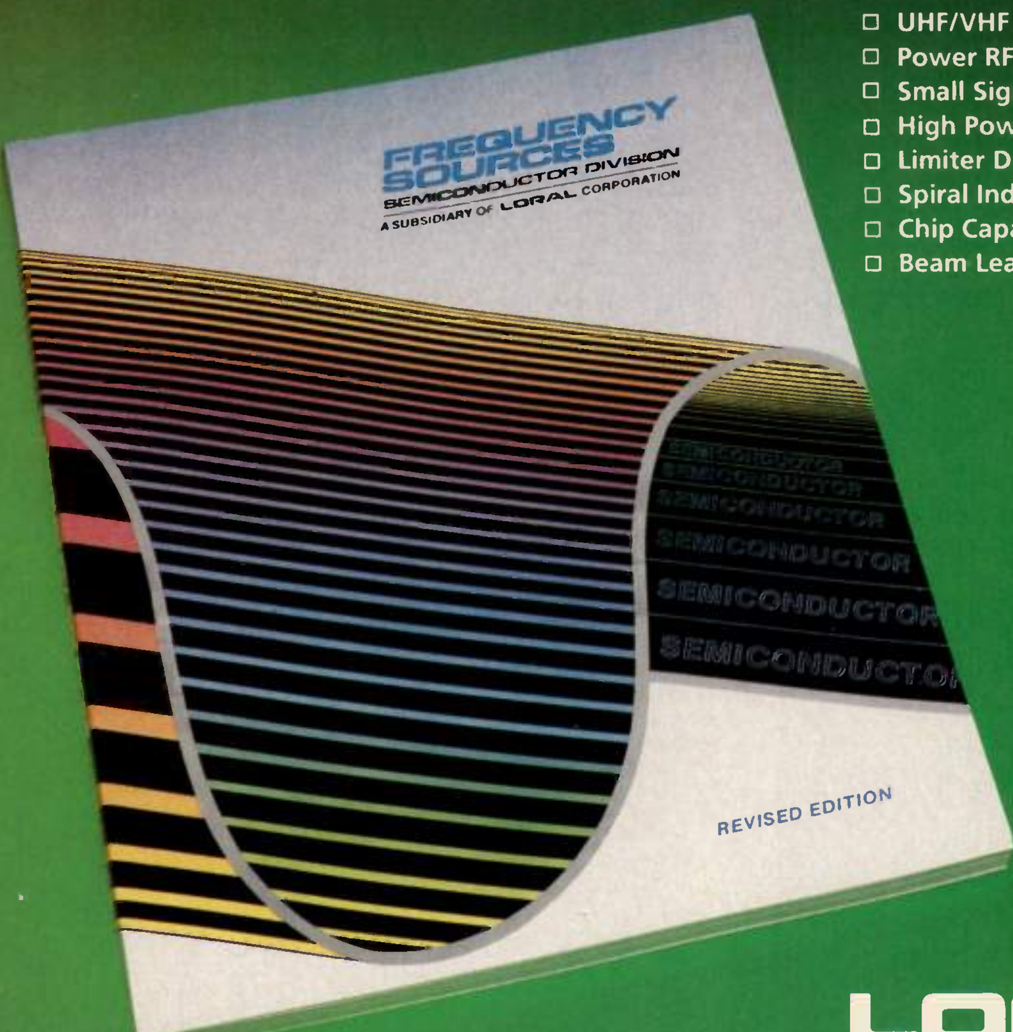
For more information, contact Donna Langan, Semiconductor Sales Manager, Loral Microwave-FSI, 16 Maple Road, Chelmsford, MA 01824. Tel: (508) 256-4113. Fax: (508) 937-3748.

- ☐ Silicon Abrupt Junction
- ☐ Tuning Varactors, 30 to 90V
- ☐ FLTVAR – Frequency Linear Tuning Varactors
- ☐ UHF/VHF Hyperabrupt Tuning Varactors
- ☐ GaAs Tuning Varactors 15 to 90V
- ☐ Power Generation – Multiplier Varactors
- ☐ Step Recovery Diodes
- ☐ Harmonic Generator Varactors
- ☐ Multi-Chip High Power Generator Varactors
- ☐ Noise Diodes
- ☐ Parametric Amplifier Varactors
- ☐ PIN and NIP Diodes
- ☐ UHF/VHF PIN Diodes
- ☐ Power RF Switching
- ☐ Small Signal RF Diodes
- ☐ High Power PIN Diodes
- ☐ Limiter Diodes
- ☐ Spiral Inductors
- ☐ Chip Capacitors
- ☐ Beam Lead PIN Diodes



LORAL
Microwave - FSI

VLF-HF Receiver EK 890

The EK 890 developed by Rohde & Schwarz is a VLF-HF receiver adapted to all shortwave applications. It has an MTBF ≥ 14,000 and has remote-control capability using modems. A maximum of 99 receivers can be remotely controlled using the addressable bus interface. Also, the range of IF filters in the EK 890 allows it to enable up to six bandwidth combinations.
Rohde & Schwarz
INFO/CARD #178

High Stability BVA Oscillator

The model 8600B BVA operating at 5 MHz offers excellent phase noise characteristics, as low as −123 dBc 1Hz from the carrier and −155 dBc at 1 kHz from the carrier. It changes frequency by no more than 1×10^{-11} per day and nominal lifetime is 20 years. The BVA is available at frequencies of 5, 5.12, 10, and 10.24 MHz.
Oscilloquartz
INFO/CARD #177

Tunnel Diode Pulse Driver

The TD1110C tunnel diode pulse driver, when used with the TD1107 tunnel diode pulse head, produces pulses with 25 ps rise time and 240 mV amplitude. The fixed 65 ns delay has a typical drift coefficient of ±1.5 ps/degree Celsius. The trigger delay jitter is typically 2 ps rms with a guaranteed maximum of 3.3 ps rms. The price is $995.

Picosecond Pulse Labs, Inc.
INFO/CARD #176

RF software

Microwave Design System

The Microwave Group of Cadence Design Systems, Inc. introduces the Cadence Microwave Musician Design System. This design software provides front-to-back design capability for both MMIC and hybrid devices. The system includes a schematic capture package, libraries, simulator interfaces, and a set of layout and verification tools and allows designers to begin their designs from either schematic entry or layout and move on to simulation. Microwave Design Systems is offered on UNIX-based platforms.
Cadence Design Systems, Inc.
INFO/CARD #210

Radiated Emissions Measurement Software

Hewlett-Packard announces the HP 85879A radiated-emissions measurment software package, which calculates product compliance with commercial EMI regulations. Features include precompliance scanning capability, and user-defined frequency band measurements. The software allows testing in the 9 kHz to 22 GHz range and is compatible with personal computers that have MSDOS and the HP 8230C BASIC language processor Release II. It sells for $8,000.
Hewlett-Packard Company
INFO/CARD #209

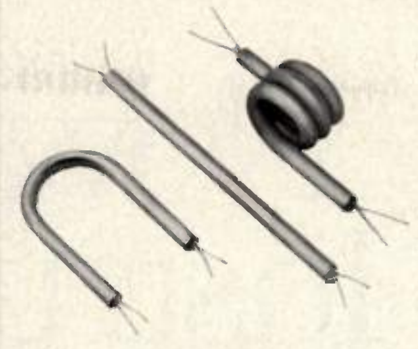
Antenna Analysis Program

Roy Lewallen announces ELNEC, a new antenna modeling and analysis program based on MININEC. This program is used for modeling and analyzing an antenna in its actual environment. ELNEC adds plotting, printing, simplified input and about ten times speed increase to that of MININEC when using the coprocessor version. The program requires an IBM-PC compatible with at least 360K RAM. CGA, EGA.
Roy Lewallen
INFO/CARD #208

ACADEMY Version 3.0

Version 3.0 links all of EEsof's microwave/RF simulators under the ACADEMY™ design framework. Simulators included are Touchstone®, Libra™, OmniSys™, and Microwave SPICE®. ANACAT™ anc Xtract™, supply linear and nonlinear device files for the simulators. Cost: $5,000 to $34,200.
EEsof, Inc.
INFO/CARD #207

WIRELINE® and WIREPAC® give you more design flexibility, greater power handling capability and less loss. You get the handling characteristics of wire with the performance of a machined hybrid.



		WIRELINE	WIREPAC
WHAT IT IS	An octave band hybrid	HC2 & JC2 3 dB + .72 − .6 f2/f1 = 2.0 L(inches) = 1850/FMHz	KC2 & LC2 3 dB + .53 − .45 f2/f1 = 2.0 L(inches) = 1970/FMHz
	A narrow band hybrid	HC1 & JC1 3 dB + .32 − .25 f2/f2 = 1.25 L(inches) = 1850/FMHz	KC1 & JC1 3 dB + .22 − .19 f2/f1 = 1.25 L(inches) = 1970/FMHz
	A directional coupler	A 2″ length is a 10 ± 1 dB coupler from 175 to 229 MHz	A 2″ length is a 10 ± 1 dB coupler from 186 to 244 MHz
OFFERS THESE KEY FEATURES	Low VSWR	1.2 MAX	1.1 MAX
	Good isolation	20 dB MIN	30 dB MIN
	Low insertion loss	0.3 dB MAX	0.2 dB MAX
	Handles average power	HC 100 Watts / JC 200 Watts	500 Watts
	Handles peak power	2000 Watts	2500 Watts
	Frequency range	100 MHz to 5 GHz	246 MHz to 6.5 GHz
	Outside dimensions	HC = .105″ JC = .145″	KC = .25″ dia round LC = .25″ square
	High reliability product	Used in communication satellites	Used in high power amplifiers
PLUS THESE ADVANTAGES	It's easy to use	To trim, use a razor blade and thermal strippers	Comes cut to length
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INFO/CARD 48

EMC News Update

European Community to Include Susceptibility Limits in 1992 — Electronic equipment sold in the European Economic Community in 1992 must not only meet electromagnetic interference (EMI) emissions standards, but must be sufficiently immune from EMI and electrostatic discharge (ESD) for proper operation. Directive 89/336/EEC makes it clear that equipment marketed in europe must meet immunity standards. While there is currently a lack of standards in this area, CISPR has recently released a draft standard, CISPR/G (Secretariat) 22, containing proposed test methodology and immunity standards. The proposed standard covers 9 kHz to 40 GHz, with emphasis on the widely-used 30-1000 MHz range.

Amateurs Include EMC References in Comments to NTIA — In response to the National Telecommunications and Information Administration (NTIA) study of electromagnetic spectrum usage, the American Radio Relay League (ARRL) has filed comments relating to the interests of amateur radio operators. As reported in *QST*, these comments include the statement that services provided to the public by amateurs "will remain available only if public policy continues to recognize the modest requirements of radio amateurs for reasonable access to the radio spectrum, reasonable antenna installations, compatibility of home-electronic devices with transmitting and sensitive receiving equipment nearby, and relief from unjustified state and local regulation." These comments indicate an increasing concern about the proliferation of electronic devices that can cause, or are susceptible to, EMI.

FCC OKs Ferrite Bead Add-on Kits — Personal computers and video cards may now use customer-installed ferrite bead kits to meet FCC Part 15 emission requirements. This approval does not extend to other computer peripherals, and includes a specific list of procedures. First, the computer (or video card) must be properly tested with the ferrite bead kit and a stock video monitor. Also, clear instructions must be provided to the user for installation. Finally, a prominent notice must be made in the instructions that the device is not in compliance with regulations unless the bead kit is installed as instructed. This procedure eliminates the need for a ferrite-loaded cable assembly, which, under previous rules could result in unnecessary duplication by both the monitor and video card suppliers.

International Standards Organization Sets Transient Standards — A draft of ISO 7637 from the International Standards Organization addresses problems with interference from internally-generated signals in vehicles. This document defines the tests for supply line transients in 12 and 24 volt automotive electrical systems that can disturb other electronic circuitry. The types of transients that can be generated in different switching situations are described, allowing test simulators to create waveforms for testing of automotive electronic modules. The standard is especially important in light of the planned European cooperation to take place in 1992, where internationally recognized test procedures will be required.

EMC Gatherings Planned — The 1990 International IEEE EMC Symposium will be held August 21-23 in Washington D.C. at the Washington Hilton Hotel. Interested persons should contact Joe Fisher at (703) 521-6336. EMC Expo 1990 is scheduled for October 17-19 at San Mateo County Convention Center, San Mateo, Calif. This event is sponsored by Interference Control Technologies and *EMC Technology* magazine. Information can be obtained by calling (703) 347-0300. Both of these events include technical papers on EMC topics, and an exhibit area for EMC supplier companies.

Canada Brings Rules Closer to FCC Part 15 — Canada's Department of Communications (DOC) is in the process of amending its specification TRC51, regulating low power transmitting devices in the 300-400 MHz range. New regulations under development will bring the Canadian standards more nearly in line with current FCC requirements. For example, forbidden-band limits on emissions (at three meters) will be: 100 uV/m at 73-75.2 MHz, 150 uV/m at 108-136 MHz, 200 uV/m at 242.8-243.4, 328.6-335.4, 328.6-335.4, 406-410 and 608-614 MHz, and 500 uV/m at 960-1215 MHz. Also the current standard will be extended from its present 310-320 MHz down to 299 MHz. Until the new regulations are completed, applications for approval of these type devices under the proposed standards will be considered case-by-case.

Call for EMC Articles — Because of our ongoing commitment to EMC, *RF Design* has a need for many different types of articles relating to matters of compatibility and compliance. Of particular interest are:

Notes on *regulatory standards*, either comprehensive or addressing a single aspect of performance or testing.

Design techniques to help control unwanted emissions, or to reduce susceptibility to interference. Examples include multi-layer board design, digital pulse shaping techniques, physical arrangement of circuitry, filtering, and shielding.

Theoretical and measured performance of various radiated EMI and susceptibility mechanisms, such as cables, enclosures, interconnections, co-located systems, digital clock circuitry, and ESD/EMP.

Case histories of circuit or system development and their relationship to EMC matters.

Contact the editors of *RF Design* with proposals for articles or questions. Address, telephone and FAX information is on page 8 of this issue. **RF**

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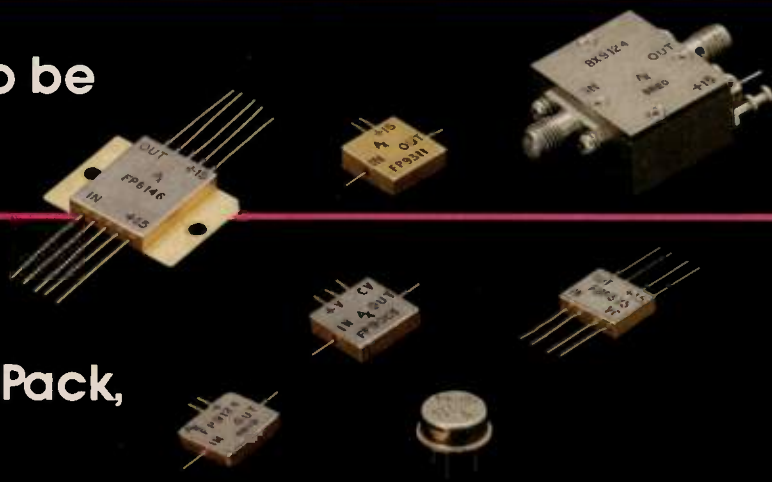

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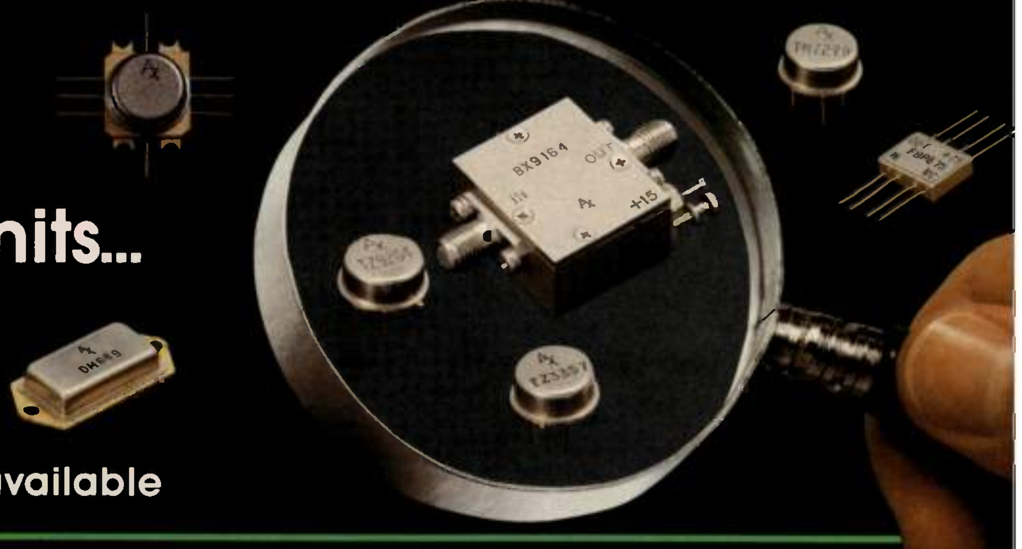

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INFO/CARD 50

VHF and UHF Crystal Oscillators

By Andrzej B. Przedpelski
A.R.F. Products

When you are running out of PC board real estate and you want to increase your MTBF and reduce spuriouses, it may be time to look at some simple high frequency crystal oscillators for your receiver LO design.

Fixed frequency receivers operating at high frequencies usually need a stable high frequency local oscillator. In most cases this LO frequency is obtained by multiplying a crystal oscillator frequency up to the desired level or by using a phaselock loop. Both require a large amount of board space and reduce overall MTBF. What is worse is that in some cases both approaches introduce unneeded frequencies within the receiver and, with them, unwanted spuriouses.

These problems can be considerably reduced by using a crystal oscillator operating directly at the desired LO frequency. Several techniques were tried in connection with a required crystal-controlled LO, where space was very restricted. While discrete transistors gave good results, the use of integrated circuit amplifiers provided additional space requirement reductions. In addition, most of the circuits did not require the use of inductors, which are difficult to obtain with resonant frequencies high above the operating frequency. The inductors are usually required in overtone circuits to prevent operation at the fundamental and/or the undesired overtones.

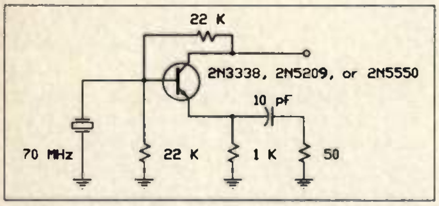

Figure 1. Kleinberg's circuit.

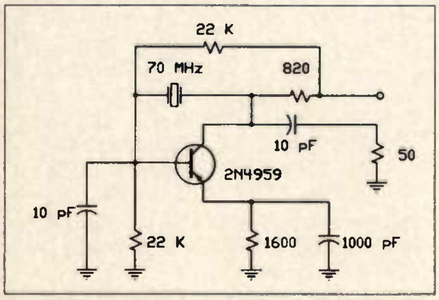

Figure 2. Pierce fundamental oscillator.

Two crystals were used for all the tests:

1. A 330 MHz fundamental crystal obtained through the courtesy of Leonard Kleinberg of NASA.
2. A 350 MHz fifth overtone crystal provided by Innovative Frequency Control Products, Inc.

The fundamental crystal was an experimental unit, not presently available in production quantities. The overtone crystal is a production item.

Circuit #1 (70 MHz fundamental). The first circuit tried was Leonard Kleinberg's (Figure 1). This circuit behaved as described (only "low" frequency transistors worked) (1), was very reliable, and provided strong harmonics up to above 1000 MHz. The output, at the fundamental frequency of the overtone crystal (70 MHz), was about 7-8 dBm with the circuit shown.

Circuit #2 (70 MHz fundamental). For comparison, a standard Pierce circuit (Figure 2) was also tried. Now, a higher frequency transistor had to be used to obtain good performance. The circuit provided the same output as circuit #1, but used more parts, which is not desirable.

Circuit #3 (210 MHz third overtone). Circuit #2 was modified by replacing the 820 ohm collector resistor with an LC circuit tuned to 210 MHz. An output of about 7 dBm at 210 MHz was obtained. Harmonics were strong up to about 1500 MHz. This circuit could not be made to operate at the fifth harmonic with the available parts and using the rather "messy" breadboard construction.

Circuit #4 (70 MHz fundamental). While circuits #1 and #2 performed satisfactorily, they did use many components. Most are needed to obtain the proper bias conditions to operate the active device. The obvious solution was to reduce the component count (thus improving reliability) by using an integrated circuit RF amplifier. Here some caution has to be exercised, since the common devices are either of the inverting or non-inverting type. This can be determined by examining the internal schematic to see if there is a phase inversion or, better yet, by checking the S21 parameter, if available.

Figure 3 shows probably the simplest possible fundamental circuit. It uses a non-inverting RF amplifier. This NEC UPC1651G type is a 50 ohm/50 ohm device with all the needed biases built-in, but does not have any input or output coupling capacitors. Since the crystal does not have a DC path, only an output coupling capacitor is needed. This circuit operated reliably and provided an output of +4 dBm at 70 MHz.

Circuit #5 (210 MHz third overtone). Next, a series trap at the fundamental was placed in the feedback, as shown in Figure 4. This circuit provided a reliable output of +1 dBm at 210 MHz.

Circuit #6 (330 MHz fundamental). The 330 MHz fundamental crystal was next tried in circuit #5 configuration. The circuit had to be modified, as shown in Figure 5, to provide a reliable 0 dBm at 330 MHz. The circuit was not critical and the capacitor was varied from 56 to 270 pF without noticeable effect.

Circuit #7 (210 MHz third overtone). To prevent operation at the fundamental frequency, some means have to be used to provide the required feedback (greater than unity and the needed phase shift) at the desired overtone frequency only. This is usually done with an L/C tuned circuit. However, "pure" inductors are difficult to obtain at the higher frequen-

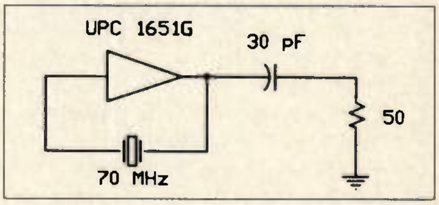

Figure 3. Simple fundamental oscillator.

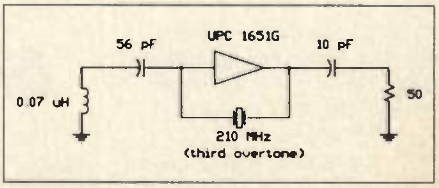

Figure 4. Third overtone circuit.

cies. First, equivalent tuned circuits, using transmission lines, were tried. This did not prove to be very successful, since the circuit exhibited resonances at other frequencies.

The obvious solution is to use an inverting amplifier and supply the needed phase shift in the feedback by means of a transmission line. This approach is shown in Figure 6. The transmission line, together with the two capacitors, cancels the phase shift in the inverting amplifier and provides an in-phase feedback at the desired frequency. The amplifier used was the Avantek GPD-401. One advantage of this type is that it has input and output coupling capacitors. Thus, no outside DC isolation is needed. Unless you know the amplifier phase shift characteristics versus frequency, it is easier to try different line lengths (giving slightly less than 180 degrees phase shift) until optimum operation is obtained. Less than 180 degrees is needed, since the amplifier phase shift decreases with frequency. The circuit shown provided a reliable output of +1 dBm at 210 MHz.

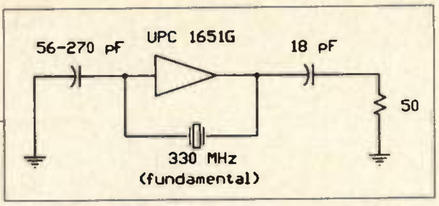

Figure 5. Fundamental 330 MHz .

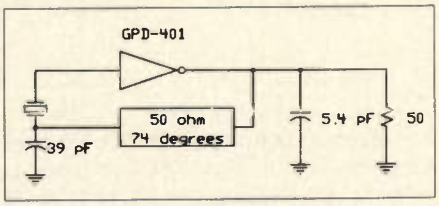

Figure 6. 210 MHz third overtone .

Circuit #8 (350 MHz fifth overtone). The same basic approach was used in circuit #8, shown in Figure 7. This circuit provided reliable -3 dBm at 210 MHz.

Circuit #9 (490 MHz seventh overtone). Since circuit #8 behaved so reliably, the same crystal was tried at its seventh overtone by decreasing the line length to provide about 89.5 degrees of phase shift at 490 MHz. The output of the oscillator was +3 dBm at 490 MHz.

Circuit #10 (overtone circuit using L/C phase shift). If you are using a low dielectric constant PC board material, the required phase shifting transmission lines may become too long at the lower frequencies. The required phase shift may then be obtained using L/C phase shifting networks. Even then, bulky (and expensive) inductors are not necessary.

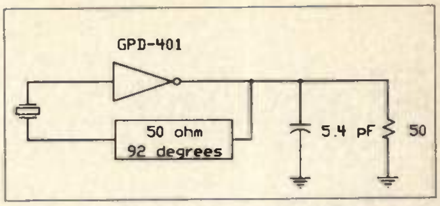

Figure 7. 350 MHz fifth overtone .

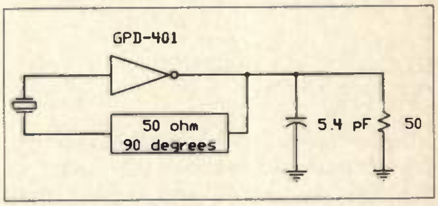

Figure 8. 490 MHz seventh overtone oscillator.

Figure 9 shows a circuit using a pi configured phase shifting network. This circuit operated reliably at the third, fifth and seventh overtones by merely re-

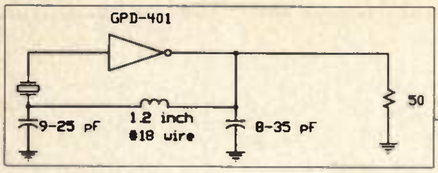

Figure 9. Overtone oscillator with L/C phase shift.

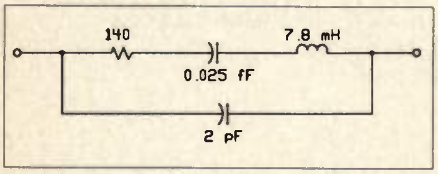

Figure 10. Approximate crystal equivalent circuit at fifth overtone.

tuning the two phase shifting capacitors. The same "inductor" (a 1.2 inch length of #18 wire) was used in all three cases. Once its value is established, this inductor can be part of the PC circuit, and thus is "for free".

Comments

Reliable VHF and UHF crystal oscillators can be built using fundamental and overtone crystal operation with a minimum of components and no discrete inductors. The transmission lines used were 50 ohm coax with a velocity of propagation of about 70 percent. These can be reduced in size considerably by printing them on a high dielectric PC board material. The length is shown in degrees at the operating frequency.

While most of the circuit development and optimizing can be done experimentally on a breadboard, the use of a network analysis program (2) gives a better insight into the circuit operation and provides ideals for optimization which can be then tried on the breadboard.

The above circuits are by no means optimum, but show what can be done with a minimum amount of parts and design effort. They are essentially not critical and slight changes in component values did not seem to affect performance. No adjustable components were used, except in circuit #10. In all cases the output was observed on a broadband spectrum analyzer to ensure that the signal was a true overtone and not a harmonic of the fundamental. No effort was made to optimize or measure phase noise, which may be a consideration in some cases. However, the signal was checked on a spectrum analyzer to make sure that no spurious sidebands were present.

A 350 MHz fifth overtone crystal was used for all the circuits except circuit #6. Its equivalent circuit, at the fifth overtone, is shown in Figure 10. A 50 ohm load was used in all the tests. *RF*

References

1. Leonard Kleinberg, "Reflection Oscillators Containing Series Resonant Crystals", *NASA Tech Briefs*, August 1989, vol. 13, No. 8, and personal communications.

2. Bert Erickson, "Network Analysis on the Personal Computer", *RF Design*, December 1986.

About the Author

Andrzej B. Przedpelski is the consulting editor for RF Design and vice president of development for A.R.F. Products, Inc., 2559 75th St., Boulder, CO 80301. Tel. (303) 443-4844.

Program Chairman Dr. Frederick H. Raab is approaching completion of the Technical Program for RF Expo East. The response of engineers and scientists to the Call for Papers, and Dr. Raab's personal recruitment of potential speakers has resulted in an outstanding collection of technical papers.

The following papers have been proposed for presentation in Orlando. This list is subject to change, since the confirmation and scheduling process is not yet complete. Based on past RF Expos, the majority of these proposals will be completed and presented at RF Expo East, November 13-15, 1990.

Extended Tutorial Papers

Noise Fundamentals
Frank Perkins, RF Monolithics, and Al Ward, Avantek

Receiver Design Tutorial
Sherman Vincent, Raytheon

Surface Acoustic Wave (SAW) Technology
Carl Erikson, Jr., Oakmont Enterprises

International

New Method of Linear Amplitude Modulation
Zhang Suwen, Wuhan University
People's Republic of China

Increasing the Efficiency of SSB Transmitters with Envelope Tracking RF Power Amplifiers
L. Voronov, Leningrad Electrotechical Institute of Communications
USSR

A DSP PSK Modem for Satcom SCPC Voice/Data
Y.S. Rao, A. Asokan, K. Reeta, C-DOT
Telecom Research Centre
Bangalore, India

Reflectively Terminated Multiport Coupler Analysis
M.H. Kori, Center for Development of Telematics
India

An Advanced Monolithic 70 dB Dynamic Range Log Amplifier
P. Chadwick, Plessey Microelectronics
United Kingdom

Other Proposals

GLONASS: The Soviet L-Band Spread-Spectrum Navigation System
J. Danaher
Structured Systems

Detection and Sorting of Frequency Hopped Signals
J.E. Dunn, S.P. Russell
Iowa State University

Large Loop Antenna Variations
R.P. Haviland
Radar Reflectivity Measurement Instrumentation
D.J. Kozakov, C.E. Sirles, D.A. Thompson
Millimeter Wave Technology

RF Systems for the Advanced Photon Source
J.F. Bridges
Argonne National Laboratory

Radio Frequency Identification Systems for Commercial and Industrial Applications
J. Eagleson, Allen-Bradley

RF Applications in Particle Accelerators
C. Hovator
Continuous Electron Beam Accelerator Facility

System Considerations for Line-of-Sight Microwave Radio Transmissions
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Rockwell International

Spread Spectrum With Digital Signal Processing
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P. Jordan
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Dynamic Evaluation of High Speed, High Resolution D/A Converters
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CSD/Telephonics

SAW Coupled Resonator Filter Capabilities and Applications
A. Coon
RF Monolithics

Quartz Crystal Filters: A Review of Current Issues
M.D. Howard, R.C. Smythe
Piezo Technology, Inc.

Narrow Band SAW Filters for IF Applications
B. Horine, S. Gopani
Sawtek

At Long Last: Modular, Digitally-Tuned RF Filters as easy as Amplifiers and Mixers
E.A. Janning
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Linear Phase High Pass Digital Filter Using the Matched Delay Subtractive Approach
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A Practical Guide to S-level Quartz Crystals and Crystal Oscillators
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Piezo Crystal

Quartz Crystal Resonator Frequency Correlation
B. Long
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Direct-Digital Waveform Generation Using Advanced Multi-Mode Digital Modulation
B.G. Goldberg
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Optimum PLL Design for Low Phase Noise Performance
S. Goldman
E-Systems

The 'Approximation Method' of Frequency Synthesis
A.D. Helfrick
Tel-Instrument Corp.

Base-n Direct Frequency Synthesizer Designs Employing Thin-Film Resonator Filters
M.Z. Komodromos, S.F. Russell
Iowa State University

Digitally Derived Swept Frequency Source for Chirped Pulse Radar Altimeters
G. Whitworth
Johns Hopkins University, Applied Physics Laboratory

High-Frequency High-Power Operation of Static Induction Transistors at Cryogenic Temperatures
M. Abdollahian, R.J. Regan, S.J. Butler, R.J. Gage
GTE

VSWR Performance of Transistor RF Power Amplifiers
R. Brounley
Brounley Engineering

A Quasi-Linear Determination of Microwave Power Device Operation from a Spice Simulated Nonlinear BJT Model
P.E. D'Anna
Microwave Modules and Devices

Digital Feedback Techniques for a Pulse Width Modulated RF Power Supply for RF Power Amplifier
H. Direen
Ehrhorn Technological Operations

Pulsed Response of a TWT
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NASA

Development and Application of RF Power Transistors for a 200-watt Multicarrier Cellular Base Station
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VHF Multipath Propagation: Causes and Cures
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RF literature

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Thirty-four magnetrons are listed from 10 watts to 75 kilowatts in the 896, 915, and 2450 MHz FCC authorized CW heating frequencies in this brochure from Richardson Electronics. Technical data is provided for each magnetron as well as warranty information.
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D.A.T.A. Business Publishing
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Electronics Testing Catalog

This catalog includes SENCORE's instruments for testing video, audio, components analyzer, and cable systems; plus waveform analyzers, IEEE/RS232 instruments, and instrument accessories. Specifications are given for all products, along with application information on each product.
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Military Components

A brochure from MuRata Erie describes this company's variety of electronic components for various military applications. Included are capacitors, resistor networks, filter connectors, crystal oscillators, power supplies, and more.
MuRata Erie North America, Inc.
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Radome Design Brochure

Microdyne Corporation has released a brochure describing the company's designs of radomes used for the environmental protection of radar, satellite and other antenna systems. The brochure discusses the technical considerations involved in radome design and describes the differences between dielectric space frame radomes and sandwich foam core radomes.
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Custom Crystal Catalog

Crystek has released a custom crystal catalog that contains their selection of crystals and background information on quartz crystals. Included in their crystal selection are crystal clock oscillators test circuits, voltage controlled crystal oscillators, surface mounted clock crystals, and other crystal products.
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