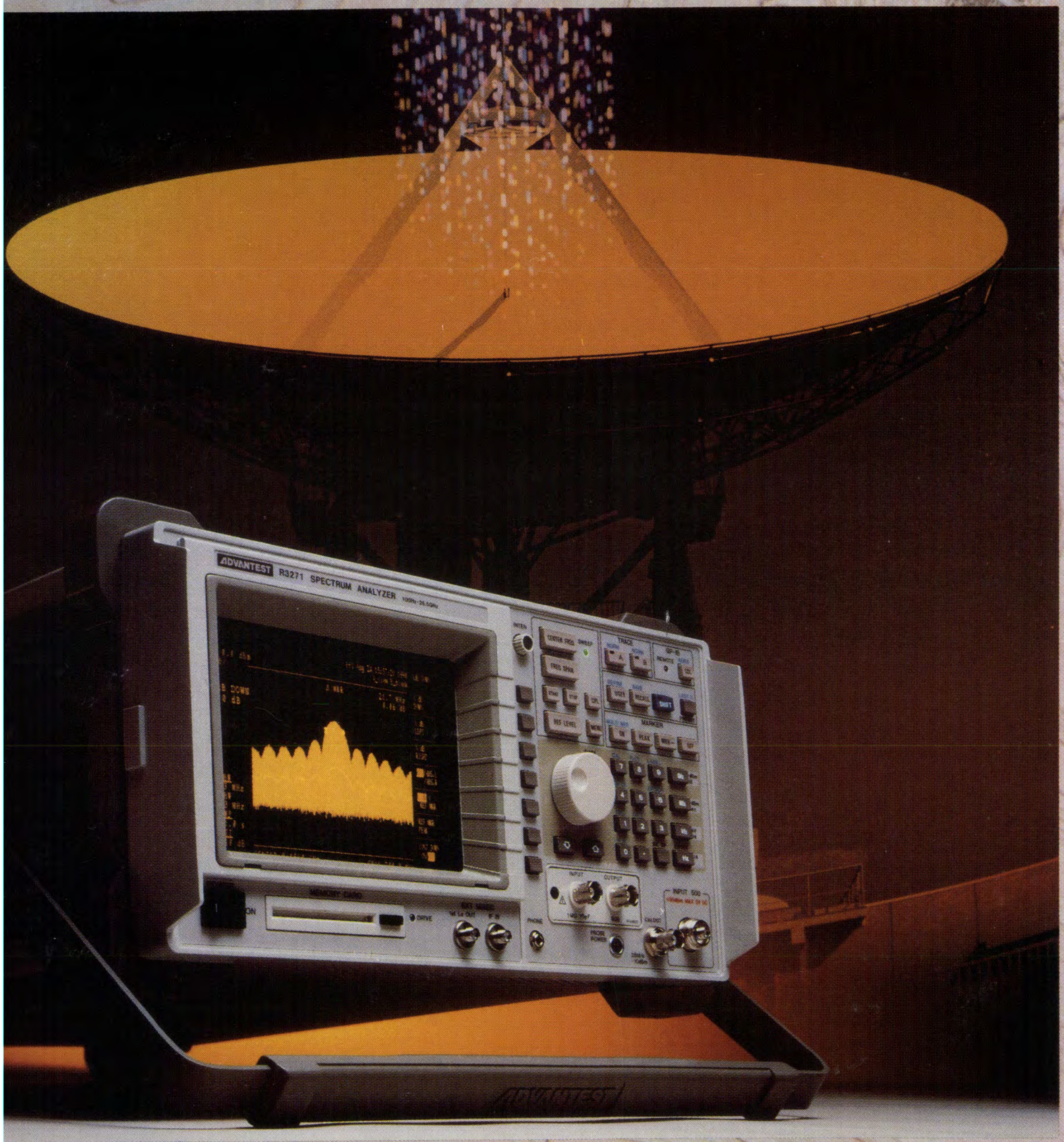
RF design

engineering principles and practices

December 1990



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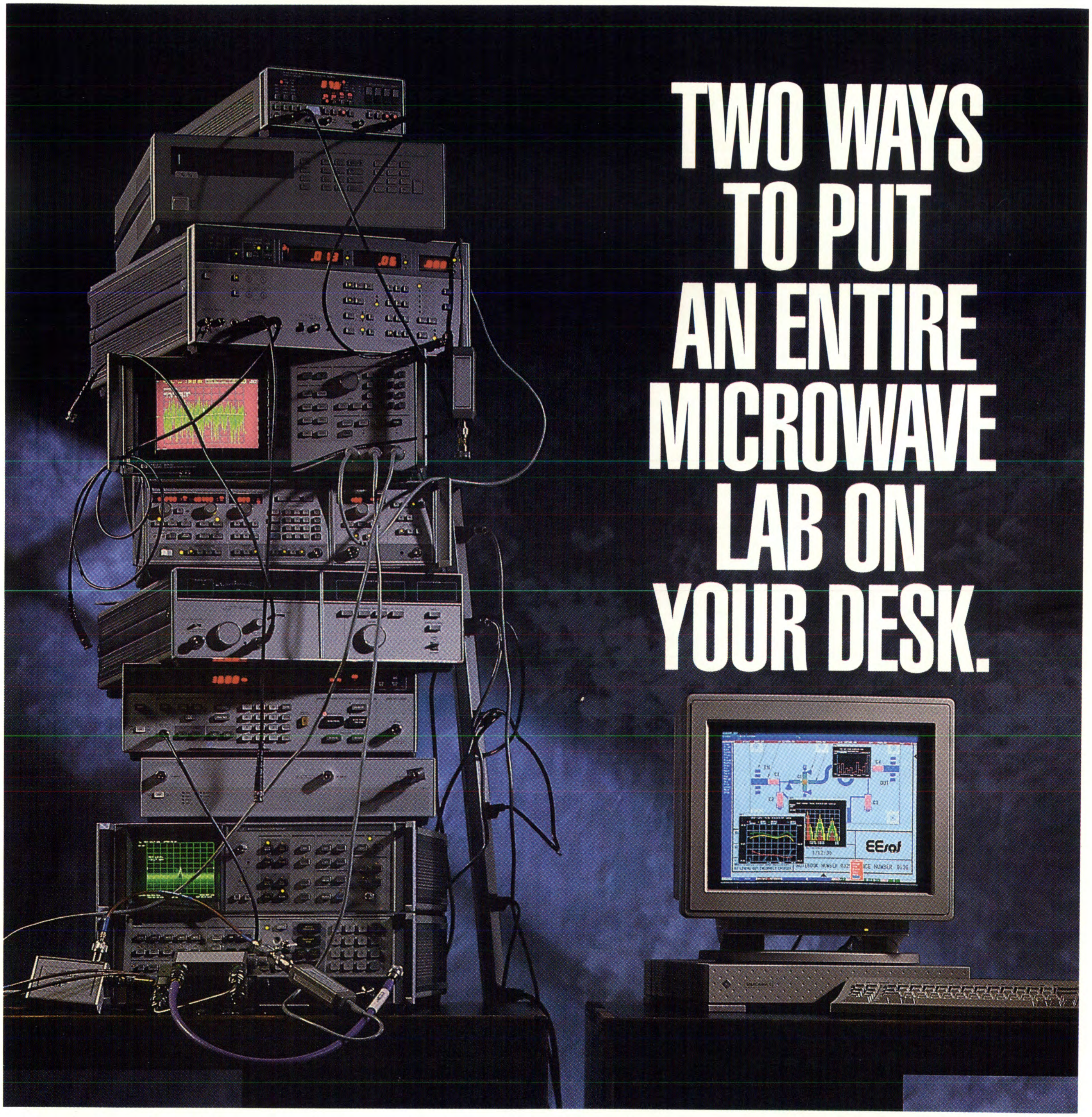

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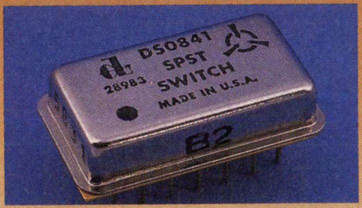

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	50	60		dB	100-200
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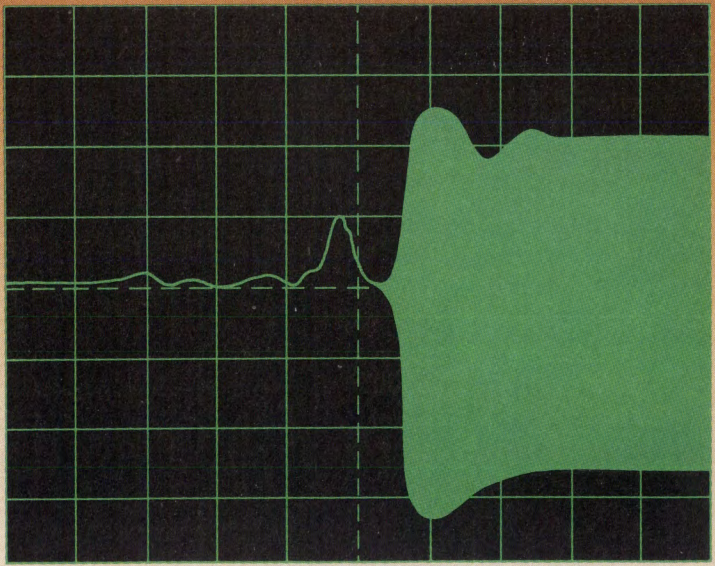

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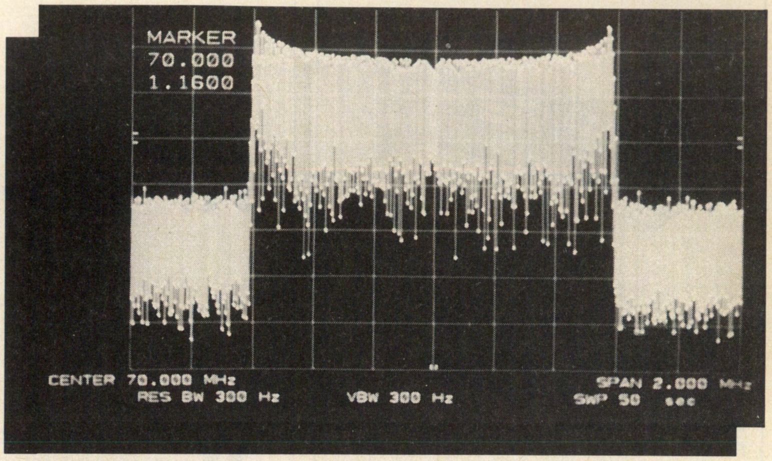

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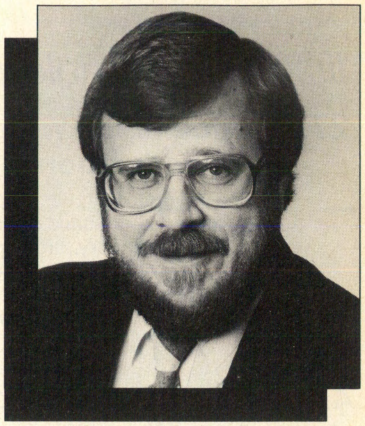

RF-on-a-Chip

By Gary A. Breed
Editor

When did electronics engineers start considering integrated circuit implementation for nearly any product? I'm not talking about mass-produced calculators and watches or development of ICs meant to be sold as commodities. I mean ICs that an engineer can specify to do the particular job he is working on.

Actually, it was nearly 20 years ago that digital product developers began to include programmable array logic (PAL) devices in commercial products. These were custom devices in that they could be either field or mask programmed to perform a specific set of logic instructions. Since then, we have seen plenty of ASICs (application-specific integrated circuits) with mix-and-match logic functions, plus universal gate array devices with thousands of individual gates that can be interconnected to do a huge number of different tasks.

More recently, digital designers have gained the ability to specify start-from-scratch ICs to meet their requirements, using some impressive computer-aided design tools to develop and debug the logic, lay out the circuit on a silicon substrate, and control the processing of the wafer. RF designers have seen some of these custom digital products for phase locked loops, direct digital frequency synthesis, and digital signal processing.

The next step combined analog and digital components, adding mainly op amps and comparators. Low-frequency signal processing, control devices, and data acquisition circuits became prime applications for this technology. At high frequencies, however, such circuits were power-hungry. The development in just the last couple of years of very high F_T PNP transistors has opened up a whole new set of applications, as complementary analog and digital circuitry can now operate into the hundreds of MHz, with its inherently lower power consumption. This type of circuit has only recently reached wide acceptance.

The final step in making full-custom RF ICs accessible is coming together right now. Advanced chip design software tools — like the ones used to design nearly every digital and analog IC — are incorporating true RF characterizations in their models. For analog functions, Spice-based routines are used in these CAD packages. The focus of recent efforts has been the development of Spice-compatible models for transmission lines, S-parameter device characterization, parasitics and discontinuities. With these tools, it is now possible to design ICs for analog or digital operation beyond 1 GHz.

GaAs ICs and MMICs are working their way toward a similar level of CAD modeling and development, albeit from a different starting point. These high performance RF ICs add even more options to an engineer's design arsenal.

Inevitably, technology moves ahead. RF has gone from spark to vacuum tubes, from vacuum tubes to transistors, from individual transistors to functional blocks of ICs and MMICs, and now we have the potential for completely integrated RF systems. We will certainly find some of the new technologies I noted last month implemented completely in silicon or GaAs.

It's hard to believe that some folks still think of RF as "old" technology!

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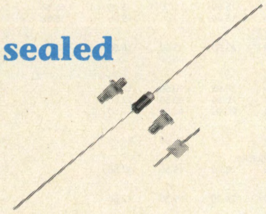

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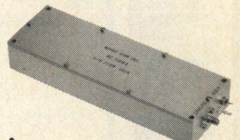

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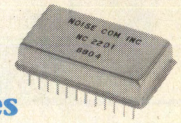

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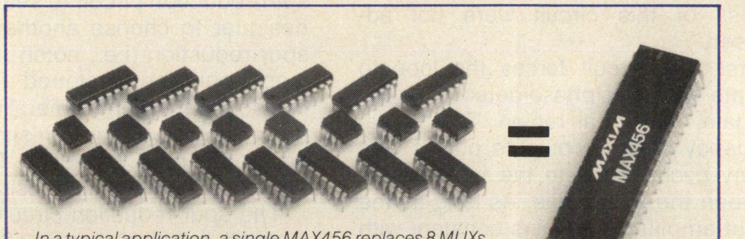

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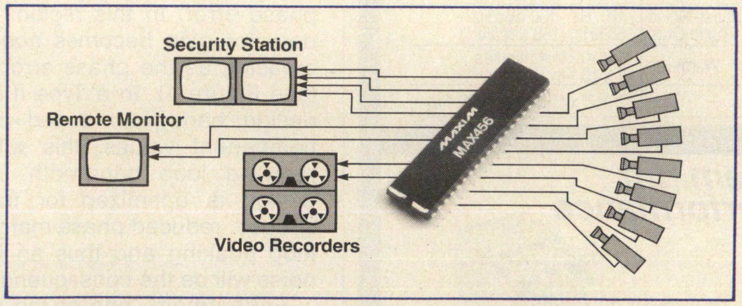

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INFO/CARD 8

RF letters

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Reference Spur Reduction in PLLS

Editor:

The article, "A Feedback Method for Reference Spur Reduction in PLLs," by John W. MacConnell and Dr. Richard W.D. Booth in the September 1990 issue of *RF Design* presents an interesting technique to be applied to phase-locked loop design. However, some significant consequences which may limit the usefulness of this circuit were not addressed.

First, the circuit forces the loop to operate with the phase-detector in the low-gain, non-linear region. The phase-frequency detector outputs pulses with a duty cycle equal to the phase error between the two inputs. As long as the output amplitude remains constant with duty cycle changes, the device provides a linear phase-error to DC-level conversion. For the extremely narrow pulse widths present for low-phase error inputs, internal propagation delay and especially finite slew rates result in a triangular-shaped output pulse, with both the amplitude and width varying with phase error. In this region, the phase-detector gain becomes non-linear, decreasing as the phase error decreases (see Figure 1). In a Type II loop with a design bandwidth based on nominal component values, this will result in reduced loop bandwidth, and if the design is optimized for transient response, reduced phase-margin. Closed-loop peaking and thus an increase in noise will be the consequence.

Furthermore, one of the implied advantages of the circuit is that it allows the more rapid loop settling times associated with wider loop bandwidths. However, this is only partially true. In order for the spur-reduction circuit to not degrade loop stability, its time constant and thus its settling time must be significantly larger than that of the primary loop (I would expect a minimum of ten times larger).

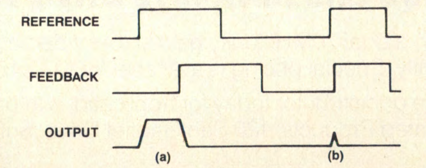

Figure 1. Slew rate resulting in a triangular-shaped output pulse.

Consider the implications: If a designer needs a loop which must settle in one millisecond, with a 60 dBc spurious requirement, his/her design must provide a loop settling time of less than 100 microseconds to allow the spur-reduction circuit to settle. Were the designer to choose another method of spur reduction (i.e., notch filtering), the loop could be designed for the one millisecond settling time. The resulting narrower loop bandwidth should provide additional spur reduction (at least 20 dB with suitable circuit design).

The spur-reduction circuit is interesting and certainly worthy of inclusion in the designer's toolbox. However, the considerations described above may limit its utility to designs where gain variation and long settling times can be accommodated.

Jonathan McLin
Electrical Engineer
Motorola, Inc.

Corrections

In our 1990/91 *RF Design* Directory Issue there were two mistakes in the design guide articles. The first appears in Richard Bain's article, "Noise Bandwidth Calculation," on page 87. Equation 2 should read:

$$B_n = \frac{1}{P_{ref}} \sum_{0}^{k=n} (f_{k+1} - f_k)\left(\frac{P_{k+1} + P_k}{2}\right) \quad (2)$$

Then in W.G. Beauregard's article, "Phase Difference Networks," on page 91 there is a line missing at the end of the page. The sentence should read "An unbalanced bridged tee network can be derived from the lattice network by an application of Bartlett's bisection theorem."

In the October 1990 issue, another mistake was brought to our attention in Tom Cefalo's article, "Microstrip CAD Program." There were some parentheses omitted from equation 9c, on page 38 and the equation should read as follows:

$$a\left(\frac{W}{H}\right) = 1 + \frac{1}{49}\ln\left[\frac{\left(\frac{W}{H}\right)^2\left(\left(\frac{W}{H}\right)^2 + \left(\frac{1}{52}\right)^2\right)}{\left(\frac{W}{H}\right)^4 + 0.432}\right] + \frac{1}{18.7}\ln\left[1 + \left(\frac{W}{18.1H}\right)^3\right] \quad (9c)$$

In Figure 7 on page 42, the inductors should be 3.978 nH and not mH.

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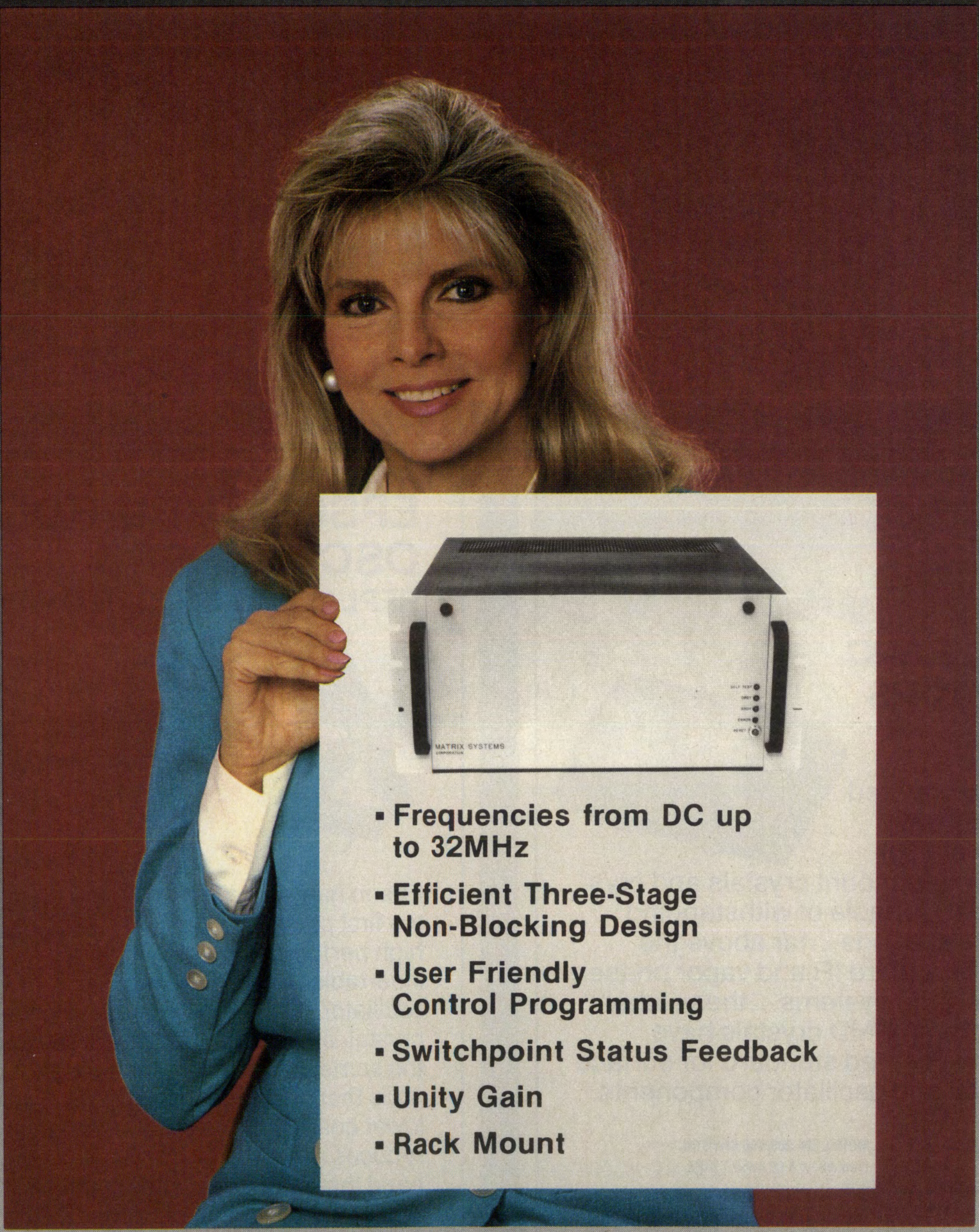

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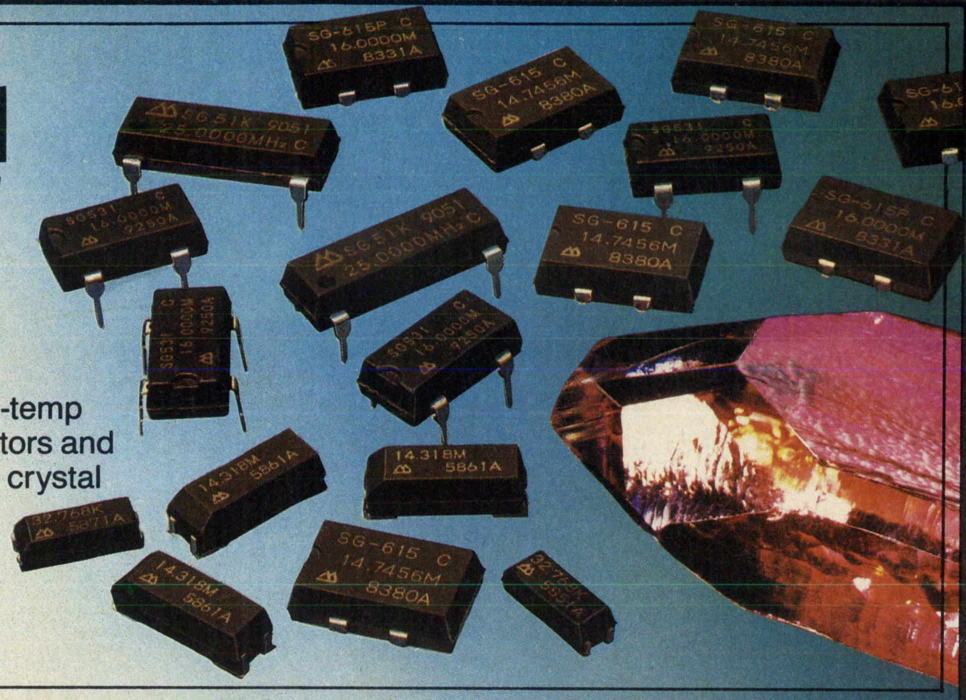

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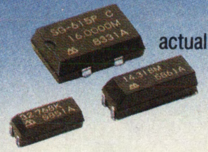

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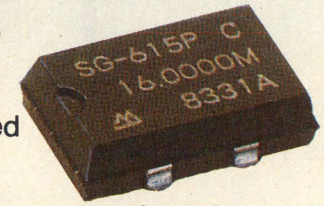

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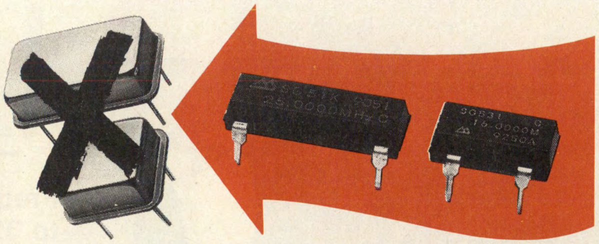

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9-12 **1990 IEEE International Electron Devices Meeting**
San Francisco Hilton, San Francisco, CA
Information: Melissa Widerkehr, IEDM, Suite 300, 655 15th Street, NW, Washington, DC 20005. Tel: (202) 347-5900. Fax: (202) 347-6109.

January

14-16 **4th Annual International Superconductor Applications Convention**
San Diego, CA
Information: SCAA, 27692 Deputy Circle, Laguna Hills, CA 92653. Tel: (800) 854-8263 or (714) 362-9701. Fax: (714) 362-9803.

15-17 **ATE & Instrumentation West**
Disneyland Hotel, Anaheim, CA
Information: Tel: (800) 223-7126 or (617) 232-3976.

22-24 **Hyper 91, Microwave Technology Exhibition and Congress**
Palais des Congres, Paris, France
Information: B.I.R.P., 25 rue d'Astorg, 75008 Paris, France. Tel: 33-(1)-4742-2021. Fax: 33-(1)-4742-7568.

28-31 **Communications Networks '91**
Washington Convention Center, Washington, DC
Information: Michael Sullivan. Tel: (508) 820-8268.

February

5-7 **RF Expo West 91**
Santa Clara Convention Center, Santa Clara, CA
Information: Kristin Hohn, Cardiff Publishing Company, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111. Tel: (303) 220-0600, (800) 525-9154. Fax: (303) 773-9716.

12-14 **4th International Smart Card Exhibition and Conference**
Novotel, Hammersmith, London
Information: Elisabeth Beckett, Marketing Manager, Agestream Ltd., Towermead Business Center, High Street, Old Fletton, Petersborough, U.K. PE2 9DY. Tel: (0733) 60535. Fax: (0733) 45522.

24-28 **NEPCON West '91**
Anaheim Convention Center, Anaheim, CA
Information: Michele Filippi, Cahners Exposition Group, 1350 E. Touhy Ave., Des Plaines, IL 60017-5060. Tel: (708) 299-9311.

March

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

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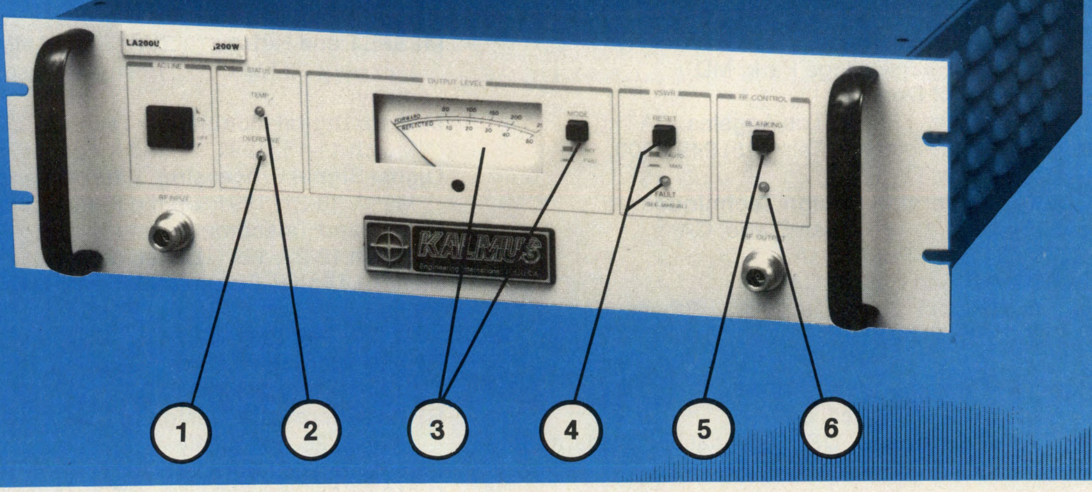

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INFO/CARD 13

RF/Microwave Circuit Design II
December 10-14, 1990, College Park, MD
RF/Microwave Circuit Design
January 28-February 1, 1991, Los Angeles, CA
Information: Besser Associates. Tel: (415) 949-3300.

Modern Power Conversion Design Techniques
February 25-March 1, 1991, San Diego, CA
April 29-May 3, 1991, Phoenix, AZ
May 20-21, 1991, San Rafael, CA
Information: e/j Bloom Associates, Joy Bloom. Tel: (415) 492-8443. Fax: (415) 492-1239.

Digital Signal Processing Workshop
March 12-14, 1991, Campbell, CA
Information: Analog Devices, DSP Applications Department, Maria Butler. Tel: (617) 461-3672.

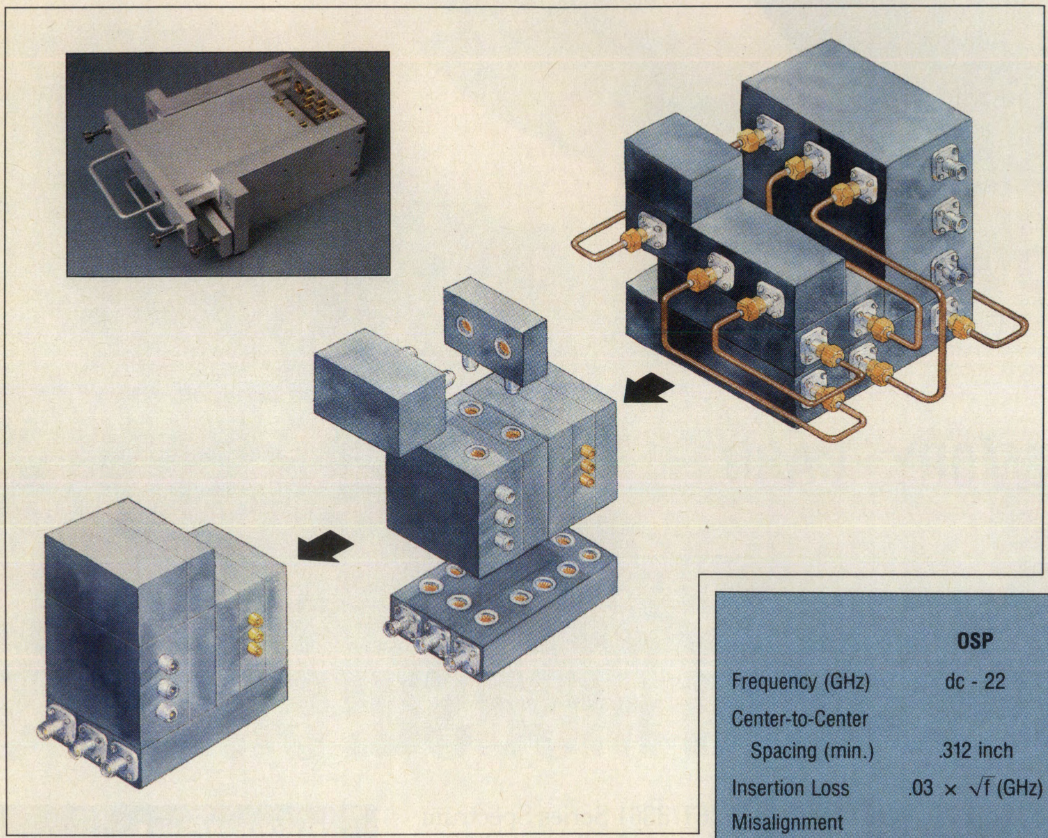
Modern Microwave Techniques
February 25-March 1, 1991, Garmisch-Partenkirchen, Germany
Far-Field, Compact and Near-Field Antenna Measurement Techniques
February 25-March 1, 1991, Garmisch-Partenkirchen, Germany
Aspects of Modern Radar
February 25-March 1, 1991, Garmisch-Partenkirchen, Germany

MESFET and Hetrostructure Based MMICs
February 25-March 1, 1991, Garmisch-Partenkirchen, Germany
Modern Digital Modulation Techniques
March 11-15, 1991, United Kingdom
Digital Signal Processing: Filtering and Estimation
March 18-21, 1991, United Kingdom
Broadband Telecommunications
March 18-22, 1991, United Kingdom
Modern Digital Communications for Space, Satellite and Radio
April 15-18, 1991, Italy
RF and Microwave Circuit Design I: Linear Circuits
April 15-19, 1991, Italy
RF and Microwave Design II: Non-Linear Circuits
April 22-26, 1991, Italy
Information: CEI-Europe/Elsevier, Mrs. Tina Persson, Box 910, S-612 01 Finspong, Sweden. Tel: 46 (0) 122-17570. Fax: 46 (0) 122-14347.

New HF Communications Technology: Advanced Technology
December 17-21, 1990, Washington, DC
Introduction to Radar ECM and ECCM Systems
December 17-21, 1990, Washington, DC
February 20-22, 1991, Washington, DC
Communication and Radar Signals: Detection, Estimation & Geolocation Techniques
January 9-11, 1991, Washington, DC
Hazardous Radio-Frequency Electromagnetic Radiation: Evaluation, Control, Effects, and Standards
January 16-18, 1991, Washington, DC
Mobile Cellular Telecommunications Systems
January 16-18, 1991, Washington, DC
Fiber-Optics System Design
February 4-6, 1991, Washington, DC
Cellular Radio Telephone Systems
February 25-27, 1991, San Diego, CA
Principles of Digital Cellular Telephony
February 25-March 1, 1991, Washington, DC
Microwave High-Power Tubes and Transmitters
February 25-March 1, 1991, Washington, DC
Broadband Communication Systems
March 4-8, 1991, Washington, DC
Satellite Communications: System Planning, Design, and Operation at Ku and Ka Bands
March 4-8, 1991, Washington, DC
Radar Operation and Design: The Fundamentals
March 25-28, 1991, Washington, DC
Information: The George Washington University, Continuing Engineering Education, Merril A. Ferber. Tel: (202) 994-8522 or (800) 424-9773.

Basic Network Measurements Using the HP8510B Network Analyzer
December 17-19, 1990, Boston, MA
January 14-16, 1991, Los Angeles, CA
Microwave Fundamentals
January 8-11, 1991, Los Angeles, CA
HP11776A Waveform Generation Language User Course
January 10-11, 1991, Los Angeles, CA
Programming the HP 8510 Network Analyzer
January 17-18, 1991, Los Angeles, CA
Information: Hewlett-Packard Company. Tel: (800) 472-5277.

Tough Problems...Easy Solutions



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Insertion Loss	$.03 \times \sqrt{f}$ (GHz)	$.04 \times \sqrt{f}$ (GHz)	$.04 \times \sqrt{f}$ (GHz)
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Rigid Mount	±.004 inch	±.0025 inch	±.0015 inch
Float Mount	±.020 inch	±.0200 inch	±.0065 inch
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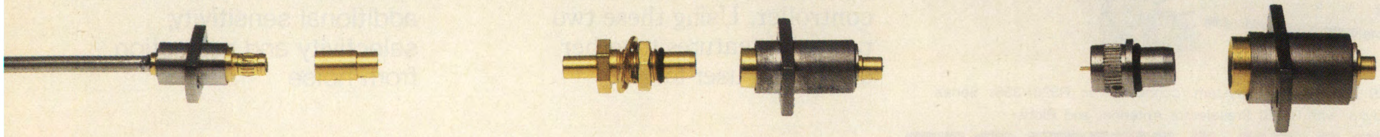

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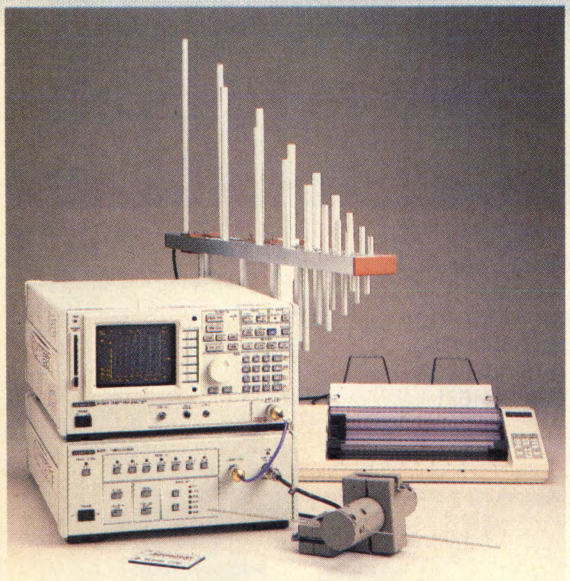


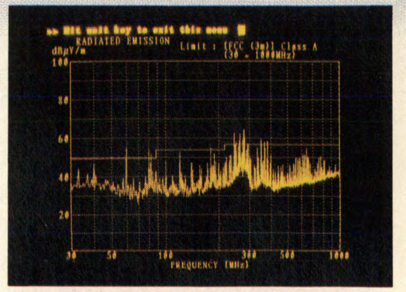
Photo : R2542B EMI Receiver System, connected to R3261/3361 Series Spectrum Analyzer with R3551 Preselector Antenna, and Plotter.

The R3261/3361 Series Spectrum Analyzers are ideally suited for RF signal analysis and offer superior functions and capabilities for EMI measurements.

Featuring the unique Advantest quasi-peak detection scheme, these analyzers offer a full 70dB dynamic range.

The IC memory card function simplifies operation and reduces the time required for routine measurements.

Difficult tasks are made easy by taking full advantage of the IC memory card and instrument controller. Using these two powerful features together allows the user to:



- Enter antenna correction factors
- Display measurements in a log/log format
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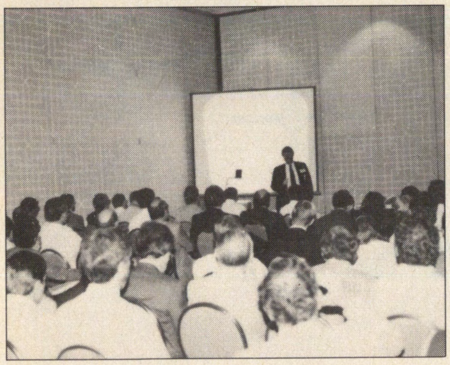
INFO/CARD 16

RF Expo East: The Best Yet

The most successful RF Expo East ever was held in Orlando, Florida at the Marriott Orlando World Center. One hundred twelve exhibiting companies, five full day special courses, and 53 technical sessions at the fifth annual Expo East combined to produce an outstanding show. Attendance was up this year with approximately 870 registered attendees and 523 exhibit personnel for a total attendance of 1393, an increase of 19.1 percent over last year's Expo East. The increase can be attributed to several factors: an excellent technical program, good exhibiting companies, location, and promotion of the show at some local companies.



Steve Russell explains the finer points of DSP Demodulation.

Dr. Frederick H. Raab, this year's technical session chairman, put together a well rounded program. For the first time, a fourth track was added to accommodate the large number of submittals for this year's Expo. The four receiver sessions drew large crowds that left standing room only. The three tutorials as always, drew a large crowd. The second session on PLLs and Synthesizers, drew another large crowd for its papers. "Designing with Direct Digital Frequency Synthesis," "Direct-digital Waveform Generation using Advanced Multi-mode Digital Modulation," and "Optimum PLL design for Low Phase-noise Performance" generated some good questions. Of interest, was the session on RF Systems for Research in Particle Physics. The papers came from three different research facilities — two in the United States and one in Germany. In all cases, the papers were well attended, and response from the attendees was positive.

The design tutorials covered RF circuit design, computer-aided filter design, oscillator design and CAD techniques.

There were a number of new products being exhibited at the show. TTE featured their new line of miniature diplexers that cover the frequency range of 20 kHz to 100 MHz. Phillips Components also received excellent response to their cellular chip set Mark II demo board.

RF Expo West will be held February 5-7, 1991, at the Santa Clara Convention Center in California. Next year's Expo East will be held in Florida again at the Stouffer Orlando Resort from October 29-31. Come join us for the best that the RF industry has to offer in technical presentations and expositions.

SILICON POWER MMICS

1 WATT
Broadband Amplifiers

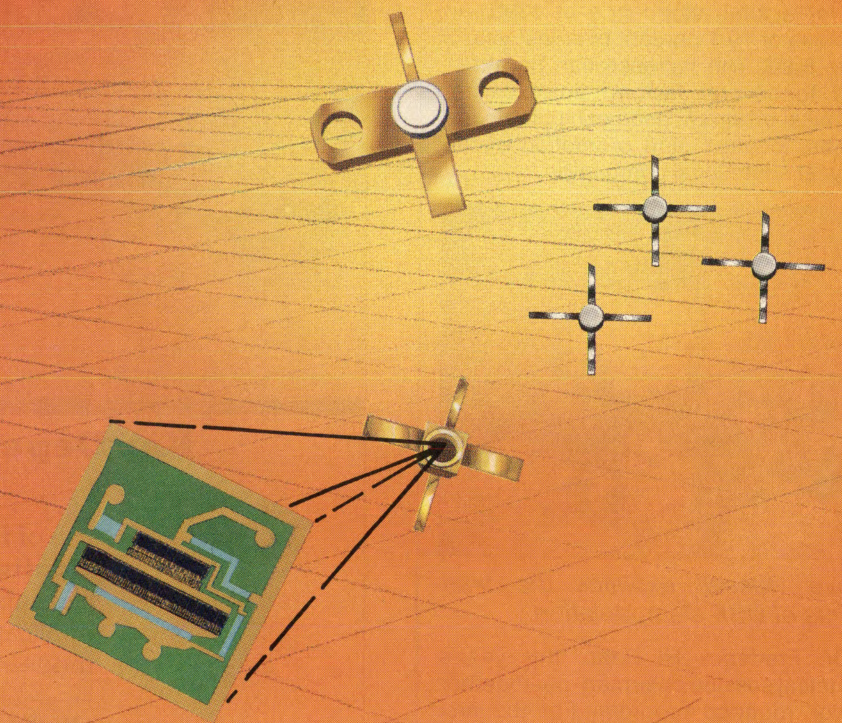
SGS-Thomson Microelectronics introduces the AMP 3020, a +30 dBm, 50 ohm cascadable gain block in the popular 200 mil hermetic, ceramic package.

This unique MMIC amplifier brings a new dimension in power to our silicon MMIC family, a series of broadband amplifiers designed to provide the circuit and system designer with reliable, cost-effective gain-blocks for broadband and narrow-band, commercial and military IF and RF applications.

The "AMP" series cascadable, 50 ohm MMIC amplifiers are manufactured with a proprietary, space-qualified refractory-gold metallization and operate at low junction temperatures for maximum reliability.

A variety of hermetic, surface mount, ceramic package styles are offered to suit individual requirements with chips available for applications demanding the ultimate in miniaturization.

If you are looking for low-cost, reliable solutions to your amplifier needs, come to the Power Source.



"AMP" SERIES SILICON MMIC AMPLIFIERS

Model #	Freq. GHz	Gain (dB)	Gain* (dB)	Pout (dBm)	N.F. (dB)	Vd/Id (v/ma)
AMP3020	**0.5**	**8.0**	**8.5**	**29.5**	**8.0**	**17/300**
	0.2	**8.2**	**8.5**	**31.0**	**7.0**	**17/300**
AMP0135	1.0	19.0	20.0	3.0	4.5	5/17
AMP0235	1.0	12.5	13.0	5.0	5.5	5/25
AMP0335	1.0	11.5	12.5	10.0	6.0	5/35
AMP0435	1.0	8.5	9.1	11.2	6.0	5.3/50
AMP0420	1.0	10.0	11.5	14.0	6.5	6.3/90
AMP0520	1.0	9.7	9.2	23.0	6.5	12/165
AMP0635	1.0	19.0	20.0	4.5	3.0	3.5/16
AMP0735	1.0	13.0	13.7	5.5	4.5	4/22
AMP0835	1.0	19.0	31.0	14.0	3.0	7.8/36
AMP0910	1.0	8.7	8.0	11.3	5.5	7.8/35
AMP1025♦	1.0	6.3	7.6	27.0	9.0	15/325
AMP1120	1.0	11.5	12.2	17.5	3.5	5.5/60

*Gain @ 0.1 GHz

♦Measured in a 50Ω System

SGS-THOMSON
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INFO/CARD 18

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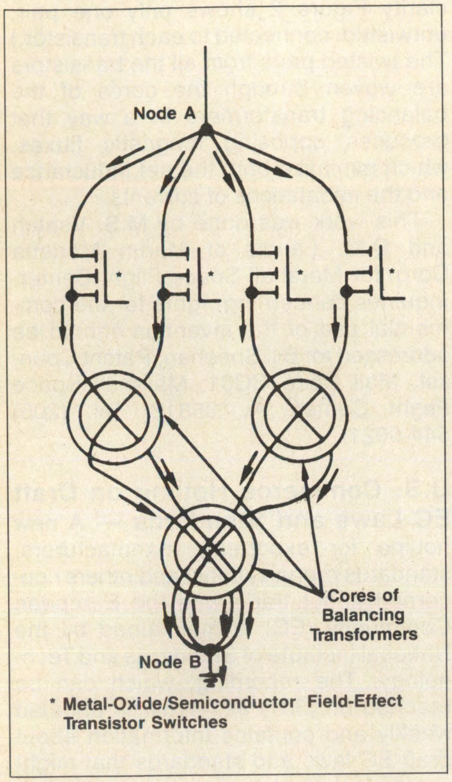


Figure 1. In the old wiring configuration, currents are balanced on single wires, with return current running elsewhere. This configuration has relatively high inductance.

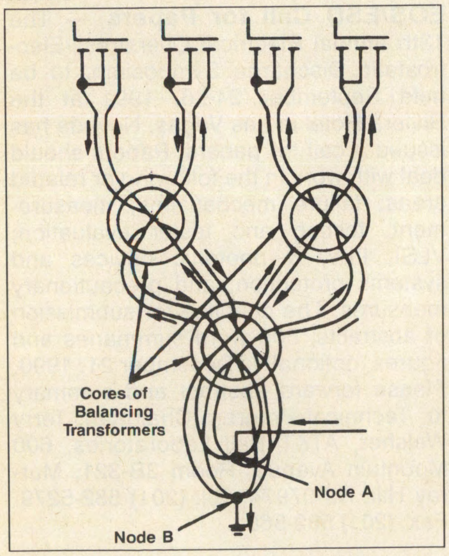


Figure 2. In the new wiring configuration, currents are balanced on twisted pairs of wires. Because the twisted pairs carry both the "hot-side" and return currents, this configuration has relatively low inductance.

Low-Inductance Wiring for Parallel Switching Transistors — This note originally appeared in the *NASA Tech Briefs* October 1990 issue. It describes transistors that share current equally, without sacrifice of switching speed.

A simple configuration for the wiring of multiple parallel-connected switching transistors minimizes the stray wiring inductance while providing for the use of balancing transformers, which equalize the currents in these transistors. The balancing of currents is necessary to prevent overloads in individual transistors, and the minimization of inductance is essential for fast switching of high currents. High-current transistor switches that could benefit from the new configuration are found in controllers for brushless motors, DC-to-DC voltage converters, controllers for electric furnaces, remote controllers for DC power, high-power pulse generators, and driving circuits for traction motors.

Interleaving and multifilar windings can reduce the leakage inductances of transformers to acceptable levels. Previously, however, the use of balancing transformers entailed excessive distributed wiring inductances because the connections to the transistors were made with relatively-high-inductance single wires (see Figure 1).

The new configuration (see Figure 2) is based on the established technique of reducing inductance by laying wires in twisted pairs. The source and drain leads of each transistor are connected to six twisted pairs of wires. (To enhance clarity Figure 2 shows only one pair, untwisted, connected to each transistor.) The twisted pairs from all the transistors are woven through the cores of the balancing transformers in a way that produces opposing magnetic fluxes, which minimize both the net inductance and the imbalances of currents.

This work was done by M.S. Veatch and D.M. Landis of Martin Marietta Corp. for Marshall Space Flight Center. Inquiries concerning rights for the commercial use of the invention should be addressed to: Bill Sheehan, Patent Counsel, Mail Code CC01, Marshall Space Flight Center, AL 35812. Tel: (205) 544-0021.

U.S. Commerce Hotline on Draft EC Laws and Standards — A new hotline for exporters, manufacturers, standards organizations, and others concerned about trade with the European Community (EC) is maintained by the National Institute of Standards and Technology. The recording, which can be reached on (301) 921-4164, is updated weekly and contains information about draft EC laws and standards that might create technical trade barriers. Hotline topics are listed by subject area and product. Information is provided on deadlines for comments and a point of contact for obtaining a review copy of the text.

EOS/ESD Call for Papers — The 13th annual Electrical Overstress/Electrostatic Discharge Symposium, to be held September 24-26, 1991 at the Riviera Hotel in Las Vegas, Nevada has issued a call for papers. Papers should deal with work in the following or related areas: failure mechanisms; measurement, testing and tester evaluation; VLSI, III-V, & photonic devices and systems protection; and precautionary measures. The deadline for submission of abstracts, 500 word summaries and figures (optional) is December 21, 1990. Please forward abstract and summary to: Technical Program Chairman, Terry Welsher, AT&T Bell Laboratories, 600 Mountain Avenue, Room 3B-321, Murray Hill, NJ 07974. Tel: (201) 582-5279. Fax: (201) 582-5661.

Session Proposals Sought for Frequency Control Symposium — Authors are invited to submit papers dealing with recent progress in research, development and applications in the following areas: fundamental properties of piezoelectric crystals, theory and design of piezoelectric resonators, resonator processing techniques, filters, SAW devices, quartz crystal oscillators, sensors and transducers, frequency and time coordination and distribution, and applications of frequency control. Two copies of a summary (500 words) together with the author's name, address and telephone number should be sent to: Dr. Thomas E. Parker, Raytheon Research Division, 131 Spring Street, Lexington, MA 02173. The deadline for submission of summaries is January 14, 1991. The symposium will be held May 29-31, 1991 at the Los Angeles Airport Marriott, Los Angeles, CA.

Ball Corporation Demonstrates Satellite Communication System — Ball Corporation's Airlink™ conformal, aeronautical phased-array antenna was recently demonstrated aboard a Japan Air Lines aircraft. The external portion of the antenna system consists of two conformal antenna arrays, each less than 0.3 inches thick, mounted on opposite sides of the aircraft fuselage. During the inaugural flight from Tokyo to San Francisco, all passengers were provided phone and fax services to anywhere in the world by using the system. Upon successful completion of the Phase II demonstration project, JAL expects to place this service into normal operation aboard this aircraft.

Valpey-Opt Company Formed — Valpey-Fisher Corporation and OPT Industries have joined together to form a company that will offer a wide range of quality crystal and L/C filters. Using CAD, analysis and production techniques, Valpey-Opt will apply standard and non-standard design to customers' requirements in the time and frequency domain. These designs include Butterworth, Chebyshev, Elliptic and linear phase shift filters as well as customized applications demanding phase and amplitude equalization, single-side band rejection or wide-band reject filters. Inquiries are invited. Phone Wim van den Akker, Chief Engineer, tel: (508) 435-6831, ext. 212.

New Divisions Created at Crystal Technology — Crystal Technology, Inc., has been divided into two new divisions, one focusing on crystal growing and wafer fabrication, the other on components utilizing these materials. The Materials Division will supply lithium niobate crystals as well as single crystal material, such as non-linear optical crystals for laser frequency doubling. The Components Division will be responsible for the company's integrated optic, electro-optic, acousto-optic and SAW product lines.

Phoenix Microwave Labs Formed — Phoenix Microwave Labs is a new company formed to provide design, development and manufacture of microwave energy sources. The company's initial product offerings include dielectric resonator oscillators, phase locked DROs, solid state power amplifiers, custom up and down converters and transmitter subsystems. The company is located at 360 Scarlet Blvd., Oldsmar, FL 34677. Tel: (813) 854-4822. Fax: (813) 854-2020.

INFO/CARD 21

VXIbus: A Standard for the Modular Instruments Market

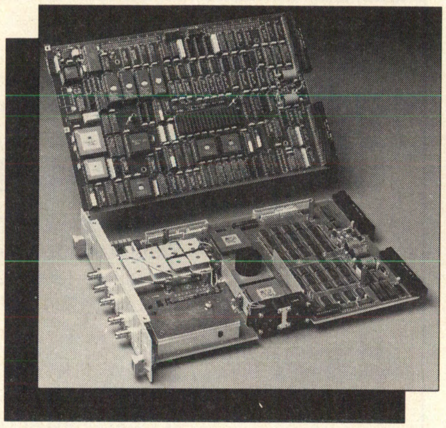
By Charles Howshar and Liane Pomfret, Assistant Editors

Of all the developments in modular RF instruments, probably none has had as much impact as VXIbus. With a VXI-based system, engineers can develop custom test systems for far less money than was feasible only a few years ago. That is not to say that there have not been problems along the way. Initial pricing was too high, integrated software was lacking, and hardware integration had few guidelines. Now companies are expanding their marketing efforts to include all aspects of the test and measurement market. As Malcolm Levy, Business Manager for Racal-Dana remarks, "We're making a big push to demonstrate VXI as an RF and microwave testing system."

There are many companies who are excited about the potential of VXIbus for modular instruments. And, although it is still a young market, some companies have put a great deal of their time and money into the VXIbus design. Levy indicates, "For us VXI is very large. It makes up fifty percent of our business today, and eighty percent of our engineering effort is wrapped up in it." This support seems to be well-placed as interest in VXI is growing. "The market last year was about $20 million; this year it will be about $50 million," agrees John Graff, VXI Product Manager for National Instruments.

Standards have been set for VXI hardware, and using this framework, companies have developed some very good VXI hardware, but the software end of the concept has yet to see a complete industry standard. This can cause a problem for users who wish to integrate modules from different companies. "There is a problem with the software. It is a gray area, and the market doesn't like gray areas," remarks Shalom Kattan, President of Guide Technology.

The problem of developing a language that eases communication between modular instruments regardless of the maker is slowly being solved. The Test and Measurement System Language (TMSL) from Hewlett-Packard was developed as an integrated language for use in any VXI system. Many companies agree that, although a good language for test and measurement, TMSL, like other VXIbus languages, has its limitations. For one, it only supports proprietary instruments. So in order to employ the software, a user first has to have the hardware to run it on. From TMSL and other languages the Standard Commands for Programmable Instruments (SCPI) has been developed. The introduction of SCPI was a significant step. "With its acceptance, the bottleneck that developed because of the software problems was removed. It's been a godsend to VXI users," comments Roger Muller, Marketing Section Manager for Hewlett-Packard.

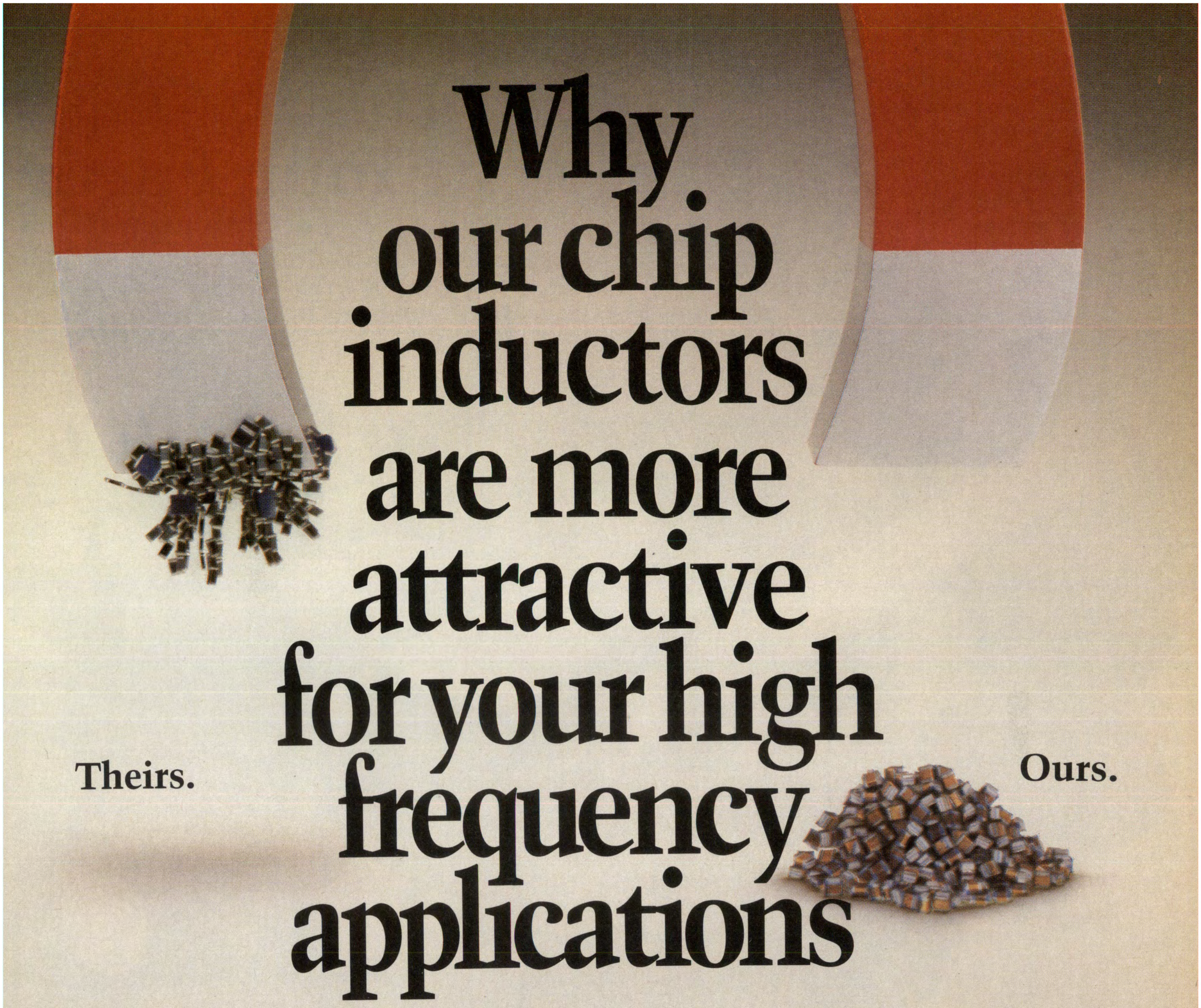


An added benefit from the development of SCPI is the ease with which the system can be programmed. It creates a "user-friendly" interface for the programmer and takes the fear out of using the system. "The user is concerned with programming the instruments. Once he sits down with a VXIbus, he will find out that integrating the VXI system will be much easier than he had thought," observes National Instruments' Graff.

An early problem with the VXIbus concept was that the military pushed for size and weight concerns and ignored price and practicality. When it became available to the public, they found it too expensive. Since then, prices have become more realistic and industry has become an integral part of VXI development. "Military was the early backer but they have scaled back because of budget cuts. General industry has picked up the load," comments Graff. Now improvements in the VXIbus design are appearing on the market. One idea that has just been developed by Rapid Systems is called the PCXI system. The attraction of this design is that it is less expensive than a VXI design. Although PCXI is not applicable for the higher speeds of VXI system uses, Rapid Systems is developing a system with a burst mode of 33 MHz to fit these applications. "For people who are looking for a modular test system, PCXI does most of the operations of a VXI system at a fraction of the cost," remarks Matthew Arksey, Research and Development Engineer for Rapid Systems.

A critical problem for VXI designs, because of the close proximity of many RF circuits is electromagnetic interference (EMI). Most problems with EMI in the VXI design have been successfully dealt with, and improvements are coming onto the market all the time. Donn Mulder, Vice-President of Marketing and Sales for EIP Microwave observes, "Simple attention to grounding and shielding design causes EMI not to be a problem." An interesting observation can be made about EMC concerns for VXI systems: What EMI characteristics will a system have if it incorporates modules from different companies? Mulder notes that the problem is being addressed, "Racal-Dana has come up with a chassis that is more robust and forgiving. It provides a lot of internal shielding."

Test engineers have been waiting for the VXIbus design to become the design standard it was projected to be eighteen months ago. And from the interest it has generated, it seems to be moving into that position. The development of a standard language for VXI will make the test system a mainstay of the modular instruments market in the coming years.



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**.

See our catalog in Vol. A, Section 1800
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Circuits for Wide-band FM Demodulation

By Alvin Wong and Ali Fotowat
Signetics Company

In applications such as telemetry, high-speed data transmission, and FM over fiber-optics, wide-band discriminators are highly desirable. In such cases, large frequency deviations are needed to use the benefits of wide-band FM. For low distortion demodulation of these signals, higher final IF frequencies such as 21.4 MHz or 70 MHz should be used. These frequencies are generally much harder to work with than lower IFs such as 10.7 MHz, or 455 kHz. The circuits described here use the property that the sinusoidal output of a bandwidth-limited IF limiter can be as effective as a square wave in producing low distortion, wide-band demodulated signals, when the sinusoid is effectively clipped by the demodulator mixer. The complete mathematics of the problem are also derived with normalized design curves. A Fortran 77 program is available for computing demodulator performance.

The circuits presented give stable IF amplifier gains at 21.4 MHz and 70 MHz, and wideband discriminators for 300 kHz baseband signals, with 600 kHz frequency deviation and distortion of the order of 1 percent.

Theory

FM demodulation can normally be achieved by multiplying two symmetric square waves with a 90 degree phase difference at the carrier frequency and a linear phase relationship around it (see Figure 1). The required phase shift can be obtained from an LC resonator, or a delay line among others. In the case of an FM demodulator based on an LC resonator, the resulting phase difference is:

$$\phi = \tan^{-1}\left[\frac{\frac{\omega_0}{Q\omega}}{1-\left(\frac{\omega_0}{\omega}\right)^2}\right] \qquad (1)$$

where $Q = R(C_p + C_s)\,\omega_o$, and $\omega_o = 1/\sqrt{L(C_p + C_s)}$. C_p is the quad tank capacitor; C_s is the capacitor coupling to the tank, and R is the total effective loading resistance.

If a delay line is used, the resulting phase difference will be:

$$\phi = \frac{\pi}{2}\left(\frac{\omega}{\omega_0}\right) \qquad (2)$$

where the time delay is $\tau = \pi/2\omega_o$.

In a phase detecting multiplier, the ideal case is that of two square waves. Therefore, the low pass filtered output's DC level will be linearly proportional to the phase difference. In reality however, the waveforms may not be perfect square waves. For a quad tank, the phase shifted signal will be sinusoidal due to the resonator. In cases where the limiter itself may not have enough bandwidth to pass the odd numbered harmonics of the IF, both signals will be sinusoidal no matter what type of phase shifting technique is used. The use of a multiplier like a Gilbert cell which requires a small signal (50-100 mV peak) to switch hard, will then be equivalent to multiplying clipped sinusoids in a perfect multiplier. The multiplier transfer function will then be:

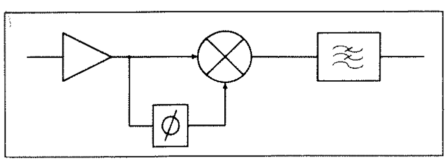

Figure 1. Basic FM demodulator.

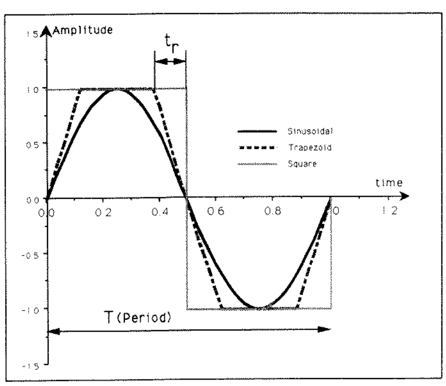

Figure 2. Different wave forms used in the analysis.

$$VDC_{out}(\phi) = \frac{1}{T}\int_0^T Z\left(\frac{2\pi}{T}t\right) Z\left(\frac{2\pi}{T}t - \phi\right) dt \qquad (3)$$

By making various assumptions for the shape of the signal $Z((2\pi/T)t)$, see Figure 2, this integral can be evaluated as shown below. If a unity peak amplitude square wave is assumed, the output DC transfer function of the discriminator will be:

$$VDC_{out} = 1 - \frac{2\phi}{\pi} \qquad (4)$$

Although a sinusoidal wave is the worst case in practice, it is mathematically easy to evaluate. For this example we have assumed an equivalent sinusoid peak amplitude of 1.

$$VDC_{out} = \frac{1}{2}(\cos\phi) \qquad (5)$$

More realistic, however, is an effectively clipped sinusoid which is approximated by a trapezoidal wave-form of unity peak amplitude. Then,

$$VDC_{out}(\phi) = 1 - \frac{8r}{3} - \frac{\phi^2}{2\pi^2 r} + \frac{\phi^3}{24\pi^3 r^2} \qquad (6)$$

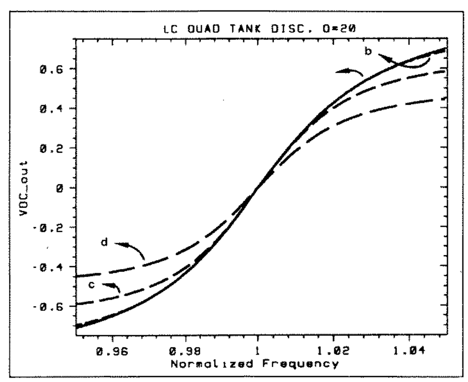

Figure 3. LC quad tank discriminator O=20, a) square wave assumption b) trapezoidal wave assumption $2t_r$ = 0.125T c)Trapezoidal wave assumption $2t_r$ = 0.25T d) Sinusoidal wave assumption.

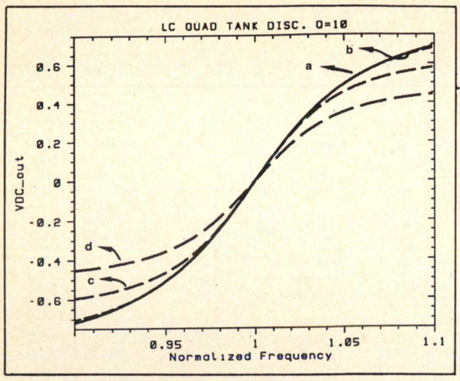

Figure 4. LC quad tank discriminator O=10, a) square wave assumption b) trapezoidal wave assumption $2t_r$ = 0.125T c) Trapezoidal wave assumption $2t_r$ = 0.25T d) Sinusoidal wave assumption.

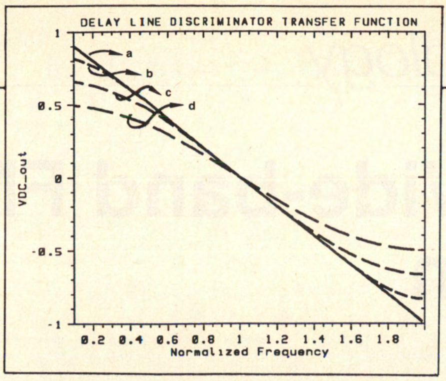

Figure 5. Delay line discriminator transfer function, a) square wave assumption b) trapezoidal wave assumption $2t_r$ = 0.125T c) Trapezoidal wave assumption $2t_r$ = 0.25T d) Sinusoidal wave assumption.

for $0<\phi<4\pi r$

$$VDC_{out}(\phi) = 1 - \frac{2\phi}{\pi} \qquad (7)$$

for $4\pi r<\phi<\pi - 4\pi r$

$$VDC_{out}(\phi) = - VDC_{out}(\pi - \phi) \qquad (8)$$

for $\pi - 4\pi r< \phi<\pi$, where $r = t_r/T$ is the ratio of one half of the trapezoidal waveform's rise time over the period as shown in Figure 2, and $r < 0.125$.

By using the phase difference in the above formulas, the FM demodulator transfer function can be obtained as shown in Figures 3-5. Obviously, a delay line discriminator can result in much wider band discriminators as expected. The interesting point, however, is that a trapezoidal wave form with a total normalized rise or fall time of 25 percent will also result in a perfectly linear discriminator curve in the middle of the transfer function. As the rise and fall times are reduced to 12.5 percent the resulting transfer function approaches that of a perfect square wave.

To get an estimate of distortion, a 3rd order polynomial can be fitted on the part of the transfer function in use. In that way, the 2nd and 3rd harmonic distortion can be estimated.

$$VDC_{out}(f) = a_0 + a_1(f - f_0) + a_2(f - f_0)^2 + a_3(f - f_0)^3 \qquad (9)$$

$$f - f_0 = A\cos(\omega_m t) \qquad (10)$$

where f_0 is the carrier frequency and ω_m is the baseband modulating frequency. The 2nd and 3rd harmonics will then be:

$$H_2 = 20\log\left[\frac{2a_2A}{4a_1 + 3a_3A^2}\right] \qquad (11)$$

$$H_3 = 20\log\left[\frac{a_3A^2}{4a_1 + 3a_3A^2}\right] \qquad (12)$$

Circuits

Figures 6-9 show the circuits for the

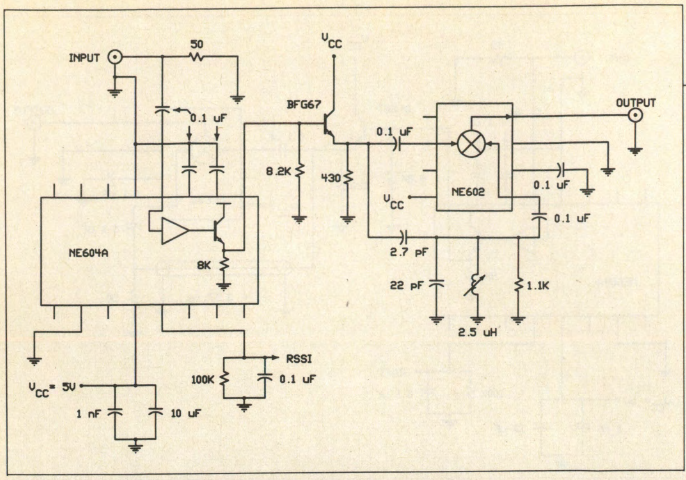

Figure 6a. 21.4 MHz LC discriminator.

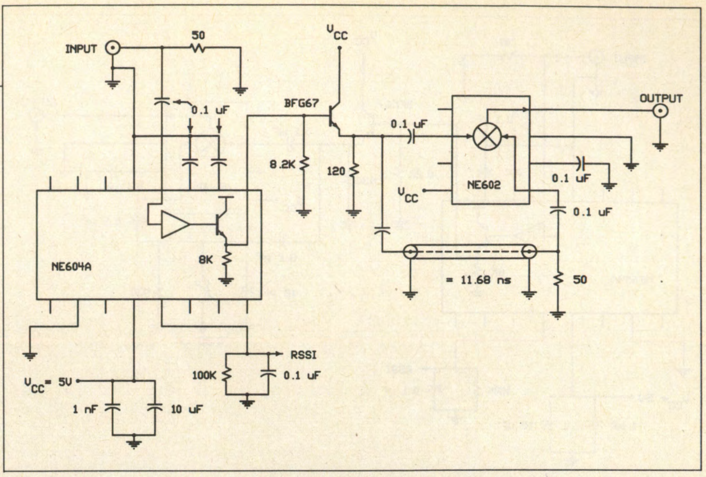

Figure 7a. 21.4 MHz delay line discriminator.

quad tank and delay line discriminators at 21.4 MHz and 70 MHz. For the limiter the second IF amplifier of the NE604A chip is used. This part of the chip is a standard FM IF limiter with about 60 dB of low frequency gain, high input impedances and a standard self biasing scheme as shown in the data sheet. The operation of the device with 1500 ohms input impedance at 21.4 MHz, or 70 MHz will obviously result in severe bandwidth limitations as well as instability. Reduction of the limiter's input impedance by an external 50 ohm termination solves these problems. The limiter generates a square wave (at 21.4 MHz) with 200 mV RMS output from an emitter follower with an 8k resistor to ground as shown in the data sheets. For wideband demodulators at 21.4 and 70 MHz, low impedances should be driven at high speeds which this buffer can not do. As a result, a BFG67 Philips RF transistor is added as a second buffer. An NE602 is used as the demodulator multiplier. The high impedances of the NE602 as well as its high speed make it almost an ideal mixer for this circuit. The Gilbert cell mixer switches hard as long as its inputs are higher than 50-100 mV peak as expected. Therefore, one can assume the limiter signals to be similar to the clipped sinusoid shown in the theoretical analysis. For the delay line, accurately cut lengths of RG-174 cables measured by both TDR and Network Analyzer methods were used. A 50 ohm termination at the end of the delay line is needed for proper operation. The use of 50 ohm delay lines required high currents in the emitter follower. Higher impedance delay lines can also be used to save power.

Test Results

Figure 6a shows the 21.4 MHz LC Circuit. A 2.7 pF coupling capacitor is used for coupling to a quad tank with a

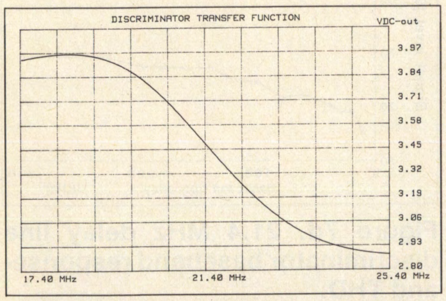

Figure 6b. Demodular DC transfer function.

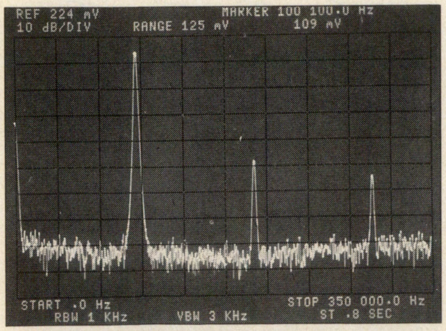

Figure 6c. Baseband output spectrum (600 kHz deviation).

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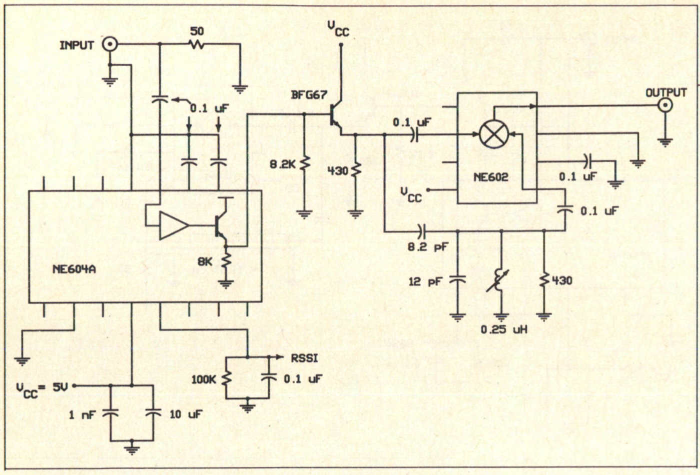

Figure 8a. 70 MHz LC discriminator.

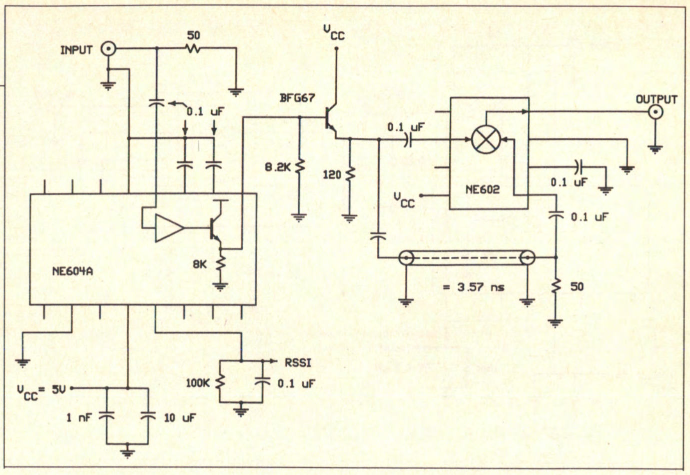

Figure 9a. 70 MHz delay line discriminator.

loaded Q of 3.5. Figure 6b shows the discriminator's DC transfer function. The result agrees with analysis of the previous section. The 180 degree inversion of the curve is due to phase inversion of the Gilbert cell multiplier.

Figure 6c shows the baseband 100 kHz output of 109 mV with a 600 kHz frequency deviation and a THD (Total Harmonic Distortion) of 1 percent. Figure 6d shows the frequency response of the baseband and the THD plotted on the same curve. Note that the THD is a function of baseband frequency. With a maximum baseband frequency of 300 kHz, and 600 kHz frequency deviation, the total IF spectrum increases to 1.8 MHz (Carson's rule) resulting in less linearity. Figure 6e shows the 21.4 MHz limited carrier having 700 mV_{p-p} amplitude and the phase shifted sinusoid (out of the quad tank; lower trace) with 250 mV_{p-p}. When probing the quad tank for this measurement, care must be taken to avoid loading or de-tuning the resonator.

Figure 7a shows the 21.4 MHz delay line discriminator's performance. A 50 ohm termination is essential for proper operation of the delay line. Figure 7b shows the measured discriminator transfer function. The linearity range and the deviation from it agree with the theoretical analysis. Figure 7c shows a 13.3 mV, 100 kHz output level with 1 percent distortion for 600 kHz deviation. Figure 7e shows the limiter output and the phase shifted signal in time domain. Both signals are 600 mV_{p-p} in amplitude and close to a square wave with the 5th harmonic (106 MHz) attenuated due to bandwidth limitations.

Figure 8a shows the 70 MHz LC discriminator's performance. In the quad tank there is a trade-off between the coupling capacitor to the tank and the baseband output and its distortion. With

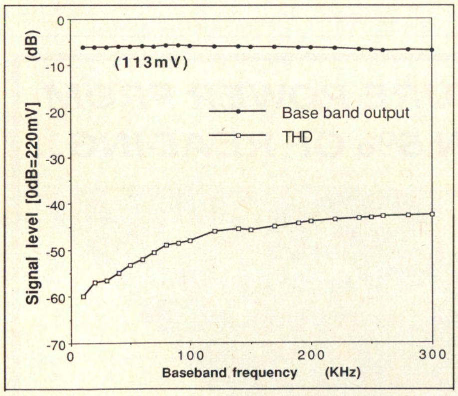

Figure 6d. 21.4 MHz LC discriminator baseband response and THD.

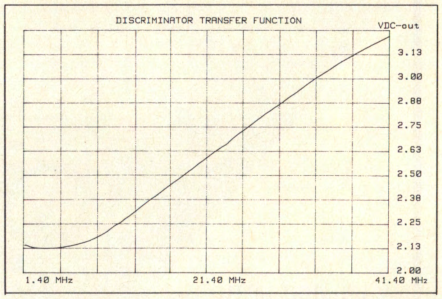

Figure 7b. Demodular DC transfer function.

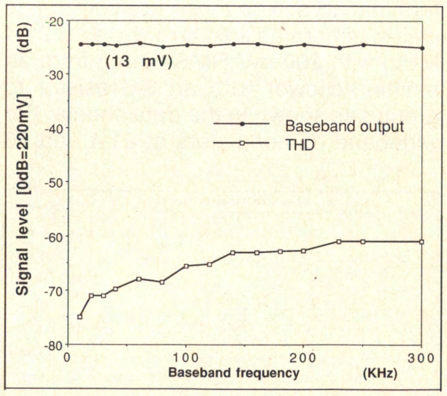

Figure 7d. 21.4 MHz delay line discriminator baseband response and THD.

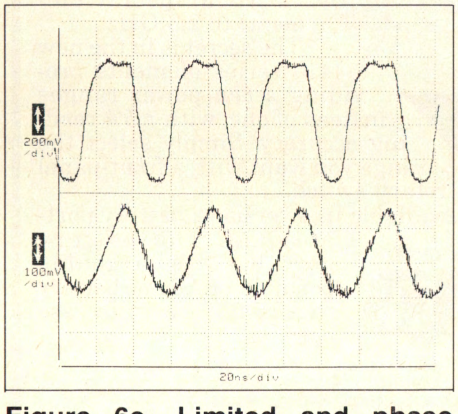

Figure 6e. Limited and phase shifted carrier in the time domain.

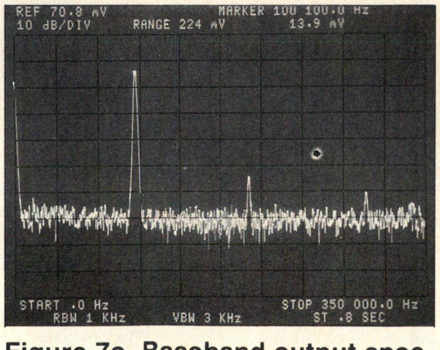

Figure 7c. Baseband output spectrum (600 kHz deviation).

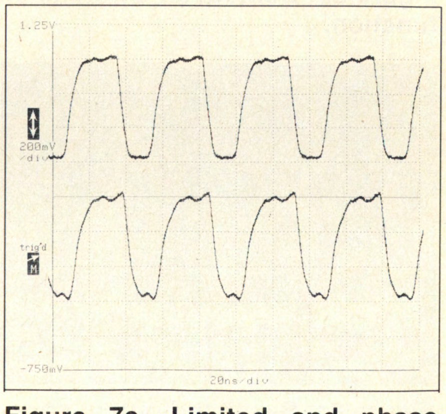

Figure 7e. Limited and phase shifted carrier in the time domain.

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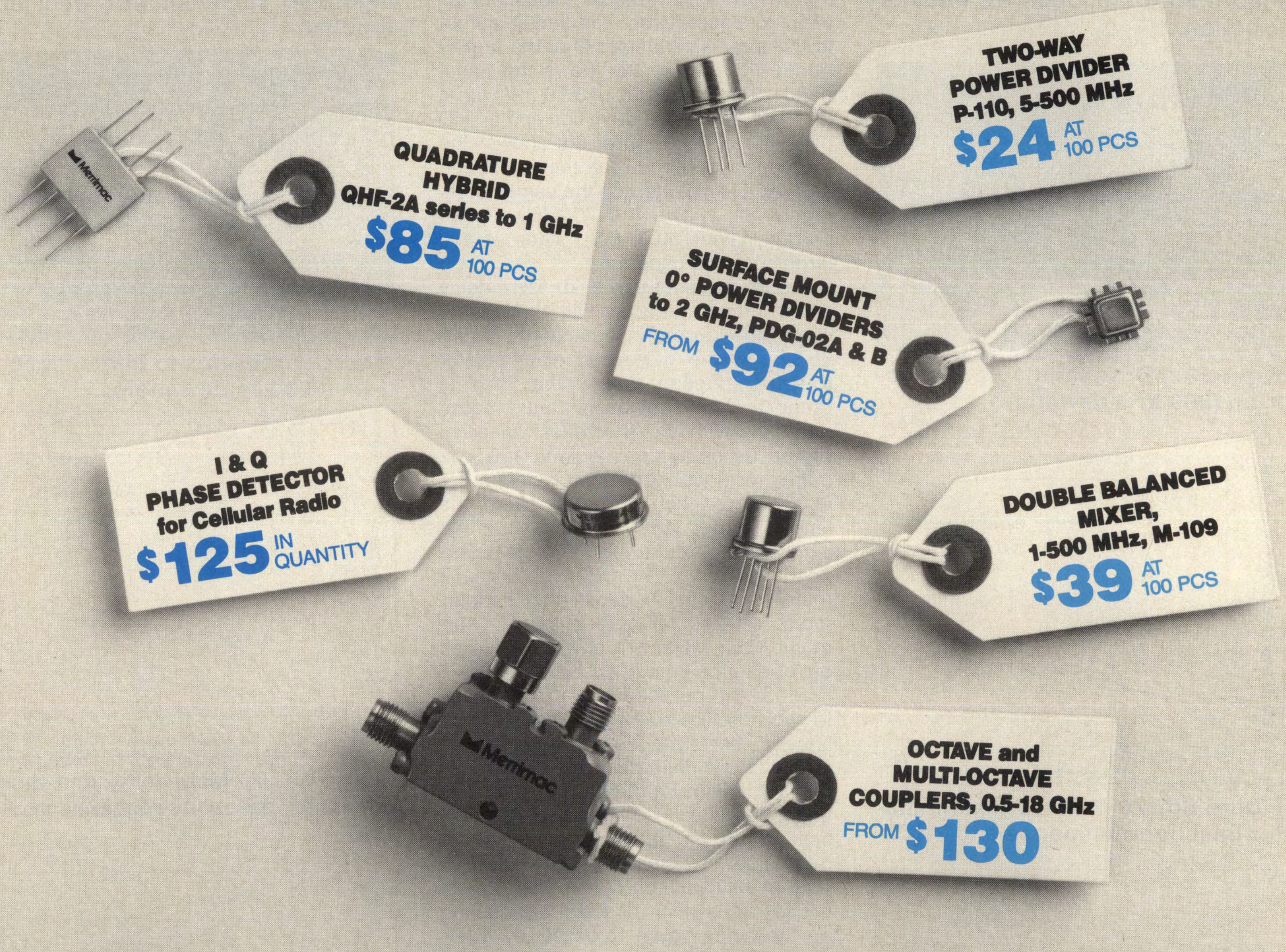

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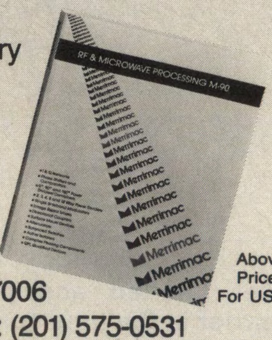

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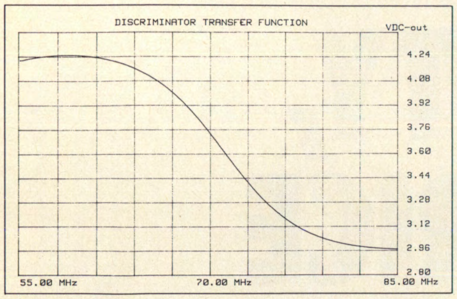

Figure 8b. Demodular DC transfer function.

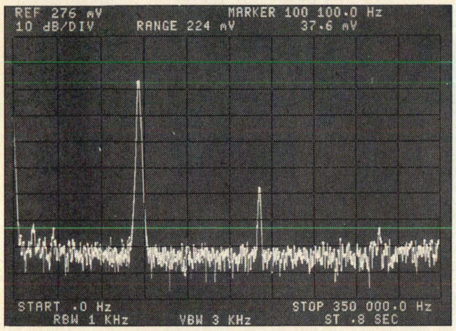

Figure 8c. Baseband output spectrum (600 kHz deviation).

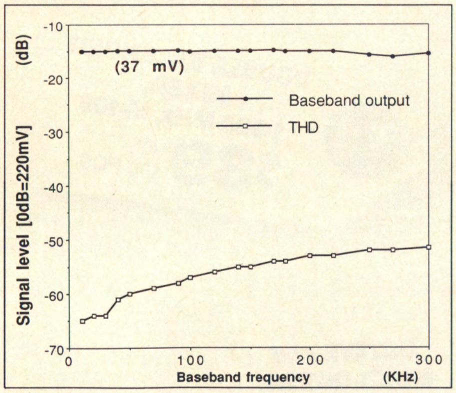

Figure 8d. 70 MHz LC discriminator baseband response and THD.

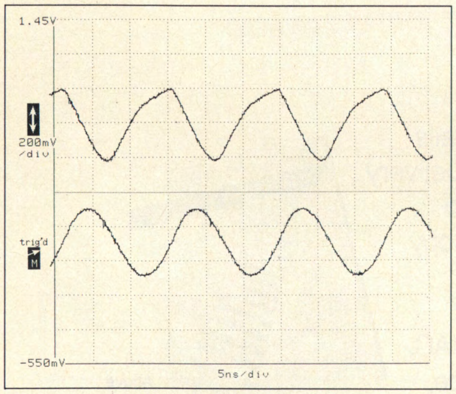

Figure 8e. Limited and phase shifted carrier in the time domain.

a small coupling capacitor a higher Q can be maintained. However, the output of the quad tank will be a low level sinusoid which will degrade the operation of the phase detector multiplier as shown in the theoretical analysis where a true sinusoidal input (effectively unclipped) was shown to degrade linearity very quickly. As a result the value of this capacitor should be optimized for the best combination of the baseband level and distortion. An 8.2 pF capacitor was used to couple from the limiter output to the tank. The loaded Q of the tank is about 4.5. Figure 8c shows the baseband 100 kHz output of 37.6 mV with a 600 kHz frequency deviation and a THD of 1 percent. Figure 8d shows the frequency response of the baseband and the THD plotted on the same curve. Figure 8e shows the 70 MHz limited carrier having 400 mV_{p-p} due to the large coupling capacitor.

Figure 9a shows the 70 MHz delay line discriminator again with the 50 ohm delay line properly terminated. Figure 9b shows the measured discriminator transfer function where the linearity range and the deviation from it agree generally with the theoretical analysis. Figure 9c shows a 5.24 mV, 100 kHz output level with 2 percent distortion for 600 kHz deviation. Figure 9e shows the limiter output and the phase shifted signal in time domain. Both signals are about 380 mV_{p-p} in amplitude but look more sinusoidal because the 3rd harmonic (at 210 MHz), and the 5th harmonic (350 MHz) are obviously filtered out due to the limiter's bandwidth.

Since the purpose of the circuits was to show the limiter and the discriminator linearity performance, no attempt was made to band-limit the input noise by using a minimum bandwidth (1.8 MHz) bandpass filter. As a result, only -3 dB limiting (the RF input level at which the demodulated baseband drops 3 dB from normal fully limited FM level) was measured. Full sensitivity tests will depend on the noise figure to be set by a low noise front-end.

As a final note, the RSSI output (Received Signal Strength Indicator) of the chip was also tested and found to exhibit a 60 dB range (logarithmic) at 21.4 MHz and a 50 dB range at 70 MHz (See Figure 10). Table 1 summarizes the performance of the four circuits.

Circuit Limitations

Figure 11 shows the test set-up used. To achieve wide frequency deviations an HP8640B RF at 550 MHz is mixed

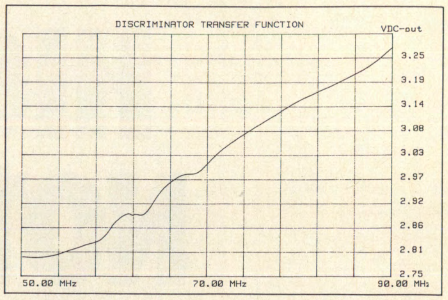

Figure 9b. Demodular DC transfer function.

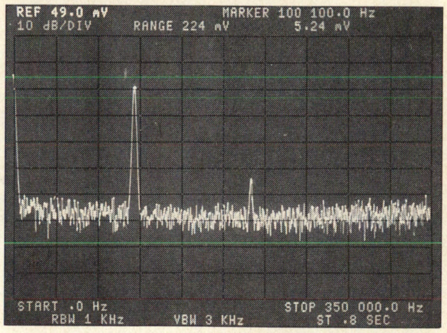

Figure 9c. Baseband output spectrum (600 kHz deviation).

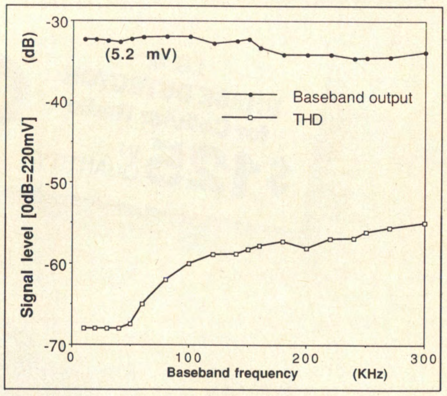

Figure 9d. 70 MHz delay line discriminator baseband response and THD.

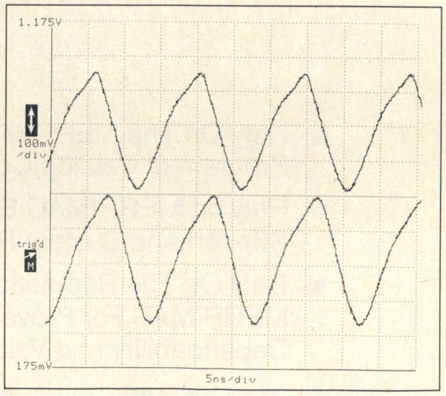

Figure 9e. Limited and phase shifted carrier in the time domain.

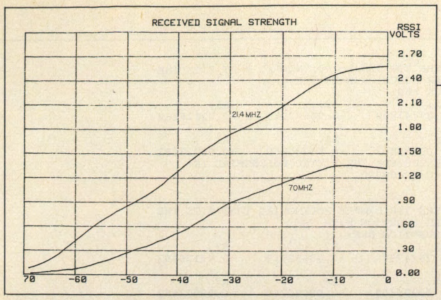

Figure 10. RSSI response for 21.4 MHz and 70 MHz.

down to 21.4 MHz or 70 MHz by using a second generator. The output of the NE602 is DC coupled to a 1 MHz low pass filter. The distortion measurements are limited by the HP8640B 1 percent distortion specification. The instrument

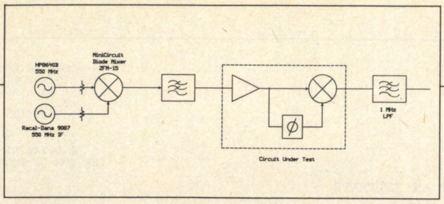

Figure 11. Test set up.

also limits the modulating signal to 300 kHz. The reduction in the baseband level at 300 kHz is due to the generator and not the output bandwidth of the NE602 which is flat up to 15 MHz with a 1500 ohm output impedance.

Figures 6e-9e showed that the reduced input impedance IF limiter has at least 70 MHz bandwidth. Lack of small signal bandwidth in the limiter will cause

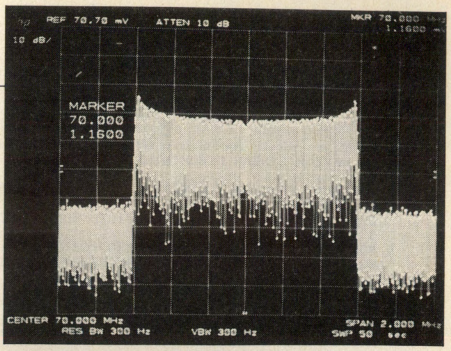

Figure 12. 70 MHz IF response at -3 dB limiting.

the distortion to be signal level dependent because of extra phase delay introduced when the IF limiter runs out of gain. Figure 12 shows the IF spectrum at 70 MHz with a 1 kHz tone and 600 kHz frequency deviation. This measurement has been done at -3 dB limiting. The symmetry of the signal while the limiter has run out of gain indicates a flat response at 70 MHz. In our circuits the distortion is not signal level dependent and the noise eventually captures the IF as expected.

As in any high speed IF stage, extreme care must be taken to avoid regenerative instabilities. These are radiated limiter output signals that are picked up by the limiter's input imped-

Discriminator Type	21.4 MHz IF		70 MHz IF	
	LC	Delay	LC	Delay
Baseband output level (100 KHz)	109 mV	13.3 mV	37.6 mV	5.24 mV
Distortion for 100 KHz baseband and 600 KHz deviation	1%	1%	1%	2%
−3 dB limiting	0.2 mV	0.15 mV	0.17 mV	0.18 mV
AM rejection for 1 KHz tone (30% AM, −40 dBm RF)	−25 dB	−18 dB	−17 dB	−15 dB
RSSI linear dynamic range	60 dB	60 dB	50 dB	50 dB

Table 1. Performance Summary

```
            Example 3    (Telemetry application)

# INPUT PARAMETERS WITH A DECIMAL POINT
# WHAT IS YOUR IF FREQUENCY? (in MHZ)
21.4
# WHAT IS THE IF BANDWIDTH OVER WHICH YOU WANT
# TO HAVE A LINEAR DISCRIMINATOR?  (in MHZ)
2.0
# FOR WHAT FREQUENCY DEVIATION DO YOU WANT TO SEE AN ESTIMATE
  OF DISTORTION?  (in MHz)
  SET EQUAL TO 1/2 OF IF BANDWIDTH FOR WORST CASE RESULTS.
1.0
# HOW MANY POINTS SHOULD BE USED FOR THE ANALYSIS?
 (MORE POINTS RESULT IN MORE ACCURACY, 100 Maximum)
20.
#WHAT IS THE LOADED Q FOR THE LC QUAD TANK?
5.0

          ****************************************

   THE SELECTED RESONATOR  Q   IS=       5.00
   THE RISE TIME FOR SQRE  WV  IS=       0.000 NANO SECONDS
   THE RISE TIME FOR TRPZD WV1 IS=       5.841 NANO SECONDS
   THE RISE TIME FOR TRPZD WV2 IS=      11.682 NANO SECONDS
#N,    FREQ, FREQ-DEV,  PHASE,    SQRE WVE, TRPZD WVE1,  TRPZD WVE2,  SIN WVE
  1   20.400  -1.000  115.5824  -0.2842492  -0.2842492  -0.2765936  -0.2159046
  2   20.500  -0.900  113.2577  -0.2584183  -0.2584183  -0.2526659  -0.1974333
  3   20.600  -0.800  110.8614  -0.2317928  -0.2317928  -0.2276416  -0.1780539
  4   20.700  -0.700  108.3988  -0.2044313  -0.2044313  -0.2015835  -0.1578148
  5   20.800  -0.600  105.8766  -0.1764070  -0.1764070  -0.1745770  -0.1367834
  6   20.900  -0.500  103.3027  -0.1478075  -0.1478075  -0.1467311  -0.1150476
  7   21.000  -0.400  100.6858  -0.1187307  -0.1187307  -0.1181727  -0.0927112
  8   21.100  -0.300   98.0360  -0.0892892  -0.0892892  -0.0890520  -0.0698979
  9   21.200  -0.200   95.3642  -0.0596026  -0.0596026  -0.0595320  -0.0467435
 10   21.300  -0.100   92.6816  -0.0297959  -0.0297959  -0.0297870  -0.0233931
 11   21.400   0.000   89.9999   0.0000008   0.0000008   0.0000008   0.0000006
 12   21.500   0.100   87.3307   0.0296593   0.0296593   0.0296506   0.0232859
 13   21.600   0.200   84.6852   0.0590532   0.0590532   0.0589845   0.0463138
 14   21.700   0.300   82.0742   0.0880639   0.0880639   0.0878363   0.0689448
 15   21.800   0.400   79.5075   0.1165836   0.1165836   0.1160554   0.0910536
 16   21.900   0.500   76.9939   0.1445124   0.1445124   0.1435064   0.1125275
 17   22.000   0.600   74.5410   0.1717663   0.1717663   0.1700771   0.1332740
 18   22.100   0.700   72.1553   0.1982740   0.1982740   0.1956758   0.1532187
 19   22.200   0.800   69.8418   0.2239797   0.2239797   0.2202342   0.1723064
 20   22.300   0.900   67.6045   0.2488384   0.2488384   0.2437023   0.1904985
 21   22.400   1.000   65.4461   0.2728209   0.2728209   0.2660521   0.2077744

THE COEFFICIENTS OF THE 7TH DEGREE POLYNOMIAL THAT FITS THE
DISCRIMINATOR TRANSFER FUNCTION WITH THE
SQUARE WAVE ASSUMPTION ARE:
    A0,            A1,            A2,            A3,
 -0.00000103     0.29747990    -0.00692919    -0.02126262

    A4,            A5,            A6,            A7
  0.00141792     0.00255403    -0.00020260    -0.00023652
THE SUM-OF-SQUARED-ERRORS FOR THE FITTED POLYNOMIAL IS:
                    0.00000000

THE COEFFICIENTS OF THE 7TH DEGREE POLYNOMIAL THAT FITS THE
DISCRIMINATOR TRANSFER FUNCTION WITH THE
TRAPEZOIDAL WAVE1 ASSUMPTION ARE:
    A0,            A1,            A2,            A3,
 -0.00000103     0.29747990    -0.00692919    -0.02126262

    A4,            A5,            A6,            A7
  0.00141792     0.00255403    -0.00020260    -0.00023652
THE SUM-OF-SQUARED-ERRORS FOR THE FITTED POLYNOMIAL IS:
                    0.00000000

THE COEFFICIENTS OF THE 7TH DEGREE POLYNOMIAL THAT FITS THE
DISCRIMINATOR TRANSFER FUNCTION WITH THE
TRAPEZOIDAL WAVE2 ASSUMPTION ARE:
    A0,            A1,            A2,            A3,
 -0.00000141     0.29749090    -0.00691769    -0.03012002

    A4,            A5,            A6,            A7
  0.00197687     0.00455276    -0.00032921    -0.00060205
THE SUM-OF-SQUARED-ERRORS FOR THE FITTED POLYNOMIAL IS:
                    0.00000000

THE COEFFICIENTS OF THE 7TH DEGREE POLYNOMIAL THAT FITS THE
DISCRIMINATOR TRANSFER FUNCTION WITH THE
SINUSOIDAL ASSUMPTION ARE:
    A0,            A1,            A2,            A3,
 -0.00000128     0.23363860    -0.00543066    -0.02518363

    A4,            A5,            A6,            A7
  0.00165373     0.00383577    -0.00028750    -0.00045154
THE SUM-OF-SQUARED-ERRORS FOR THE FITTED POLYNOMIAL IS:
                    0.00000000

THE 2ND HARMONIC DISTORSION, (in dB)
                                SQRE WVE, TRPZD WVE1, TRPZD WVE2,   SIN WVE
                                -38.198     -38.198     -38.005     -37.962

THE 3RD HARMONIC DISTORSION, (in dB)
                                SQRE WVE, TRPZD WVE1, TRPZD WVE2,   SIN WVE
                                -34.479     -34.479     -31.248     -30.658
==================================================
==================================================
Fortran STOP
```

Figure 13. Computer program example: Telemetry application with a 21.4 MHz LC quad discriminator.

ance. The reduction of the input impedance reduces this effect considerably. If more IF gain is needed, physical isolation (shielding) of the IF stages is essential.

Finally, if wider band operation is needed, the Q of the tank can still be reduced. As long as the noise of the system is dominated by the receiver front end and not by the phase detector, the audio output, no matter how small, can be post amplified without loss of the dynamic range.

Computer Program

A program written in standard Fortran 77 has been written to accomplish the above-described performance calculations. Figure 13 is an example using 2 MHz bandwidth and 21.4 MHz IF, as might be found in a telemetry application. Only the LC quad configuration case is shown here, although the program generates performance data for the delay line discriminator, as well. *The program is available on disk from the* RF Design *Software Service. See page 71 for ordering information.* **RF**

About the Authors

Alvin Wong is an RF Applications Engineer at Signetics. He holds a BSEE from San Jose State University. Ali Fotowat is an RF Design Engineer at Signetics. He holds a BS degree from CalTech, and an MS and PhD degrees from Stanford University. They may be reached at: Signetics Company, 811 E. Arques Ave., PO Box 3409, M/S 60, Sunnyvale, CA 94088-3409. Tel: (408) 991-2000.

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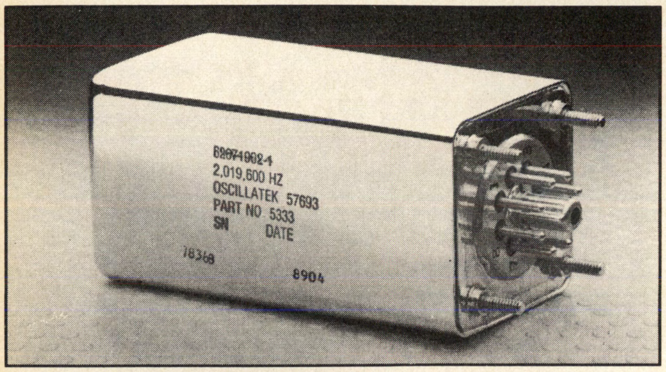

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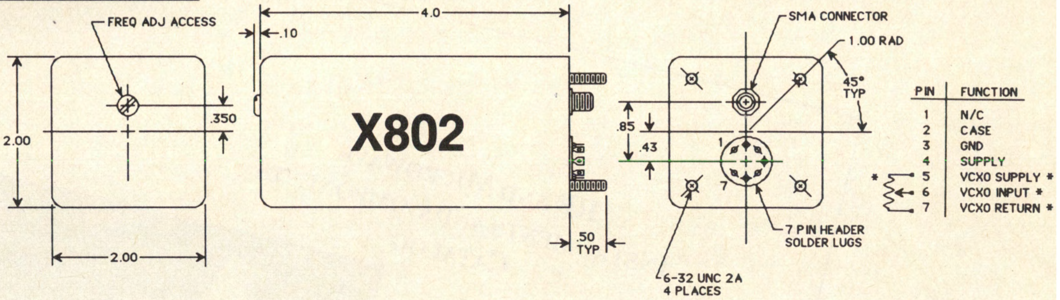
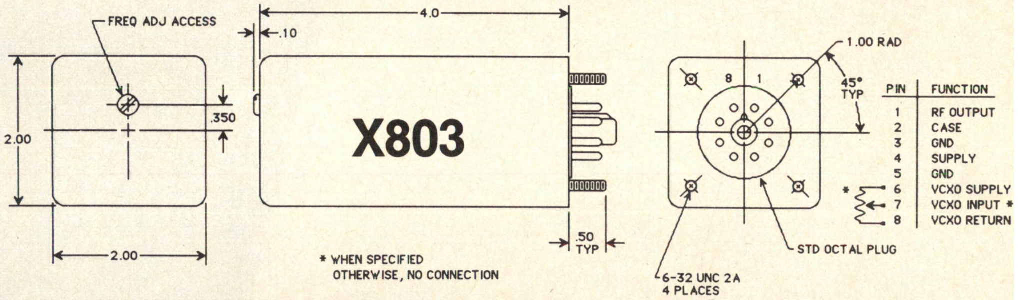
50 KHz	-150dBc/Hz





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ECL: Available with addition of -5.2V supply

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OUTPUT LOGIC	MODEL NUMBER	AGING PER DAY	TEMP. RANGE	FREQUENCY STABILITY OPTION	VOLTAGE CONTROL	FREQUENCY	SUPPLY VOLTAGE
T = TTL	X802	1 = 1 x 10^{-8}	A = 0°C to +50°C	39 = ± 3 x 10^{-9}, 28 = ± 2 x 10^{-8}, 18 = ± 1 x 10^{-8}	E IF VOLTAGE CONTROL OPTION IS DESIRED	TTL = .01Hz to 30MHz	SPECIFIED SUPPLY VOLTAGE: 12VDC TO 28VDC
S = SINE	X803	2 = 5 x 10^{-9}	B = 0°C to +70°C	59 = ± 5 x 10^{-9}, 58 = ± 5 x 10^{-8}, 18 = ± 1 x 10^{-8}		SINE = 100KHz to 30MHz	
C = CMOS		3 = 3 x 10^{-9}	C = -20°C to +70°C	18 = ± 1 x 10^{-8}, 58 = ± 5 x 10^{-8}, 28 = ± 2 x 10^{-8}		CMOS = .01Hz to 15MHz	
HC = HCMOS		4 = 1 x 10^{-9}	D = -40°C to +70°C	28 = ± 2 x 10^{-8}, 17 = ± 1 x 10^{-7}, 58 = ± 5 x 10^{-8}		HCMOS = .01Hz to 30MHz	
			E = -55°C to +70°C	58 = ± 5 x 10^{-8}, 27 = ± 2 x 10^{-7}, 17 = ± 1 x 10^{-7}			

EXAMPLE	S	X802	2	B	18		—	10 MHz	,	24V

NOTE: SX8022B18-10MHz,24V is a catalog number which defines an ovenized crystal oscillator in the X802 package with a 10MHz Sine output, aging of 5 x 10^{-9}/day, stability of ± 1 x 10^{-8} over the temperature range of 0°C to +70°C, operating on a +24VDC supply.

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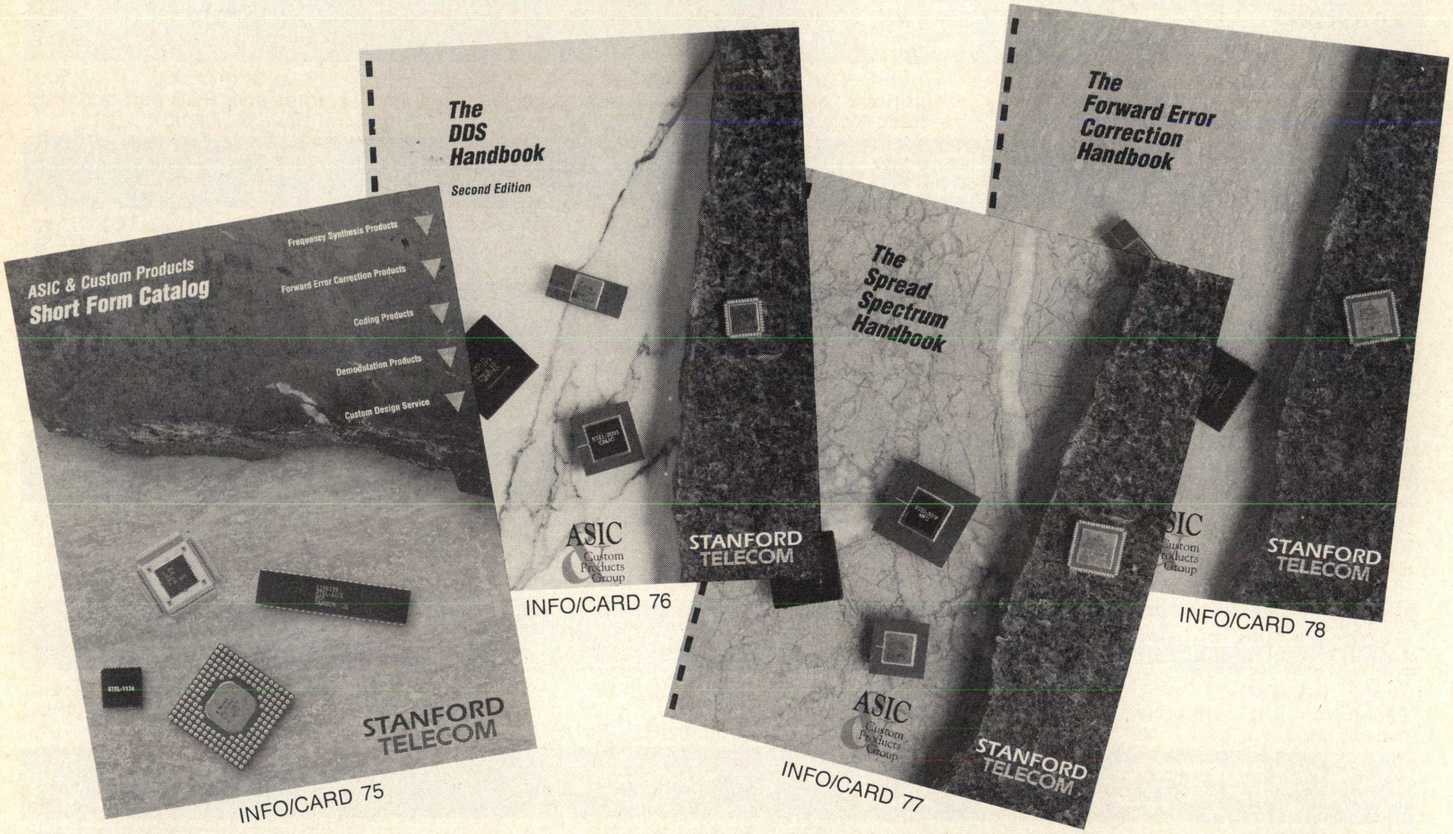

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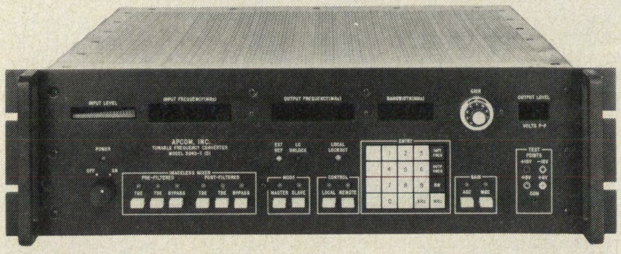

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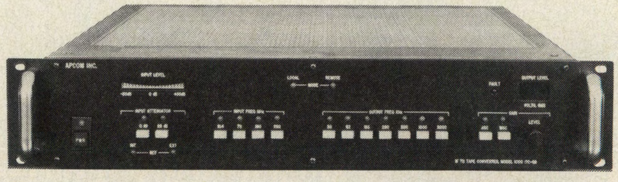

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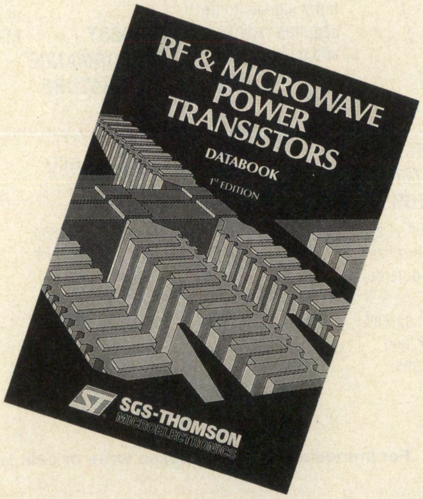

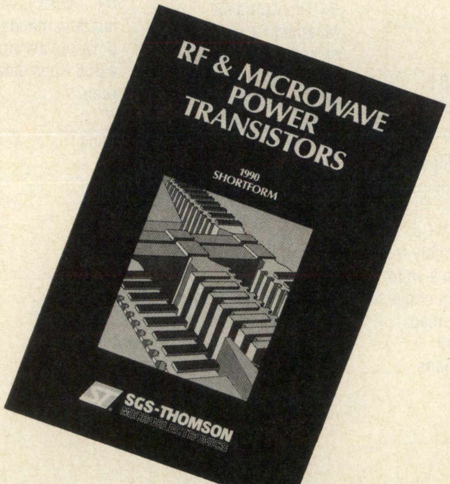

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INFO/CARD 36

Using Current Feedback Amplifiers

By Al Little
Signal Processing Products
Harris Semiconductor

Current feedback amplifiers are becoming increasingly popular for high frequency and RF designs. They are used like ordinary op amps, yet they offer higher speed, nearly constant bandwidth at increased gains, and a number of other benefits. This article describes these benefits as well as the special features and applications of the Harris HA-5004, a 100 MHz current feedback amplifier.

Although conventional op amps have been greatly improved over the past decades, very few are good enough for accurately processing analog signals approaching 100 MHz. At these frequencies, the open loop gain (and AC performance) of op amps usually drops so low that they are unusable. Even those that do work at high frequency often need to be tweaked with external feedback networks to achieve the best performance. For applications that need more than unity gain, even the best high speed op amps may still fall short due to the well known gain-bandwidth tradeoff.

To combat this shortcoming, current feedback amplifiers have now become widely used. Due to their unique circuit topology, current feedback amplifiers side-step the "gain-bandwidth" tradeoff of conventional op amps altogether, delivering nearly equal bandwidth over a wide range of gains. This feature makes them ideal for high frequency applications like video drivers, pulse amplifiers, radar and IF signal processing.

Figures 1 and 2 illustrate this basic difference between conventional op amps and current feedback amplifiers. Figure 1 shows the frequency response of a 100 MHz op amp at several closed loop gains. As shown, the unity gain bandwidth actually exceeds 100 MHz, but some undesired gain peaking also occurs. At higher closed loop gains, the bandwidth is proportionally lower, such that at a gain of 10 (20 dB) the amplifier delivers less than 10 MHz of signal bandwidth.

By contrast, Figure 2 shows the same response for the Harris HA-5004, a high performance monolithic current feedback amplifier. At unity gain, the amplifier also provides a bandwidth of 100 MHz, but without significant gain peaking. At a gain of 10, the HA-5004 rolls off at only 80 MHz, providing 8 times the bandwidth of the comparable op amp.

Extended bandwidth is by no means the only advantage of current feedback amplifiers. Unlike op amps, current feedback amplifiers do not exhibit slew rate limiting. This remarkable property is illustrated in Figure 3, in which the large signal response of the HA-5004 looks virtually identical to the small signal response. With no slew rate limiting, the output rise time and fall time remain constant, independent of signal amplititude. This linear characteristic translates to extremely low distortion, which makes the amplifier ideal for high fidelity video and RF signal processing applications.

Current feedback amplifiers achieve their superior performance by using a slightly different principle than traditional op amps. Figures 4a and 4b illustrate these differences. Inside the current feedback amplifier, there is a unity gain buffer from the non-inverting (+) terminal to the inverting (−) terminal. The inverting terminal is, by definition, a low impedance point at all times. Error currents are sensed at the inverting input and amplified such that a small change in input current produces a large change in output voltage. The ratio of output voltage delta due to input current delta is the transimpedance, Z, of the device. Like voltage gain in a conventional op amp, transimpedance is a function of frequency.

Steady state currents at the inverting input are very small because the transimpedance is large (typically 100 V/mA in the HA-5004). The voltage across the input terminals is nearly zero (typically 1 mV) due to the small offset voltage of the buffer amplifier. The ideal properties of zero input current and zero offset voltage are also true for current feedback amplifiers, and likewise simplify circuit design and analysis. The response to reactive feedback elements, however, is entirely different than to op amps due to this difference in structure, so great care must be used. Fortunately, current feedback amplifiers often eliminate the need for reactive feedback (like feed-forward designs to compensate op amps) in the first place.

Equations 1a through 4a show the relationships for traditional operational amplifiers. In equation 2a, as long as the open loop gain A(s) is much greater than the ideal closed loop gain G, the transfer function closely approximates the ideal. As frequency increases negatively or as G is increased negatively the overall bandwidth is limited by the bandwidth of A(s). These equations demonstrate why voltage feedback amplifiers must always trade bandwidth for gain.

In the case of current feedback,

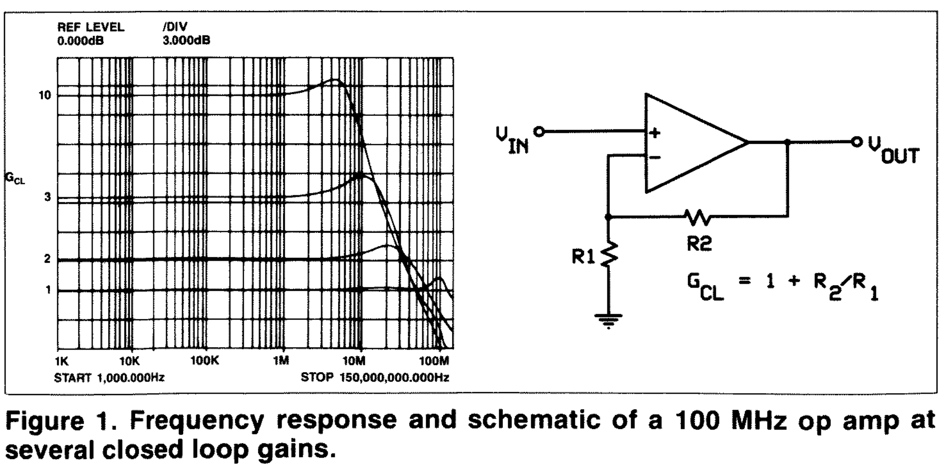


Figure 1. Frequency response and schematic of a 100 MHz op amp at several closed loop gains.

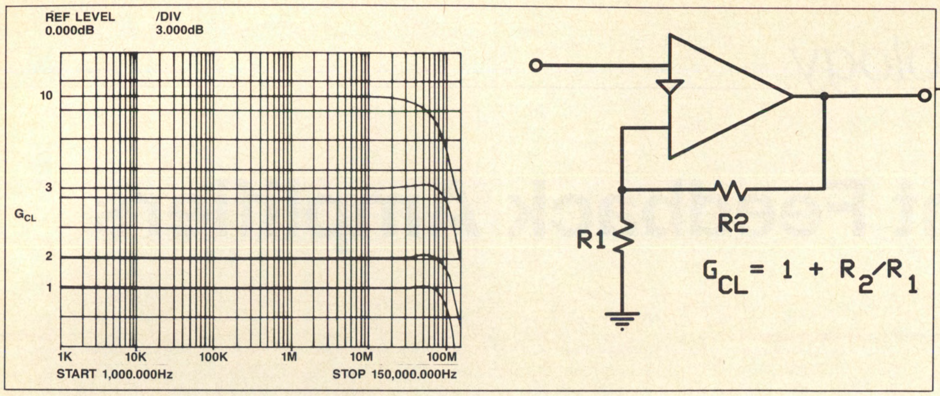
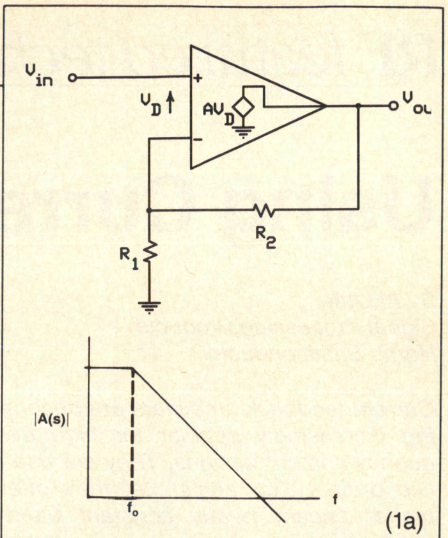


Figure 2. Frequency response and schematic of a 100 MHz current feedback amplifier at several closed loop gains.



$$V_{out} = AV_D = A\left[V_{in} - V_{out}\left(\frac{R_1}{R_1 + R_2}\right)\right] \quad (1a)$$

$$\frac{V_{out}}{V_{in}} = \frac{G}{1 + \frac{G}{A(s)}} \quad (2a)$$

$$G = 1 + \frac{R_2}{R_1} \quad (3a)$$

$$A(s) = \frac{A_o}{1 + j\left(\frac{f}{f_o}\right)} \quad (4a)$$

Figure 4a. Voltage feedback amplifier schematic.

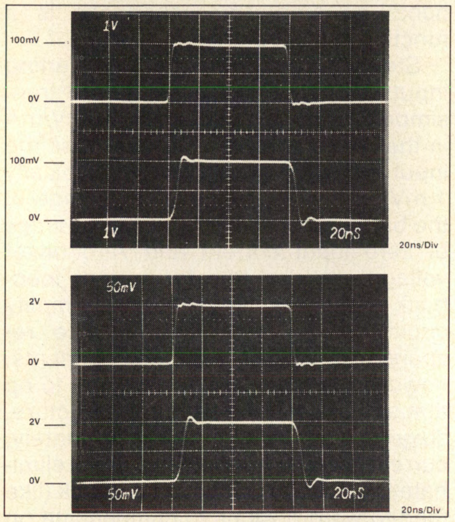


Figure 3. Comparison of slew rate limiting of the large signal response and the small signal response for the Harris HA-5004 current feedback amplifier.

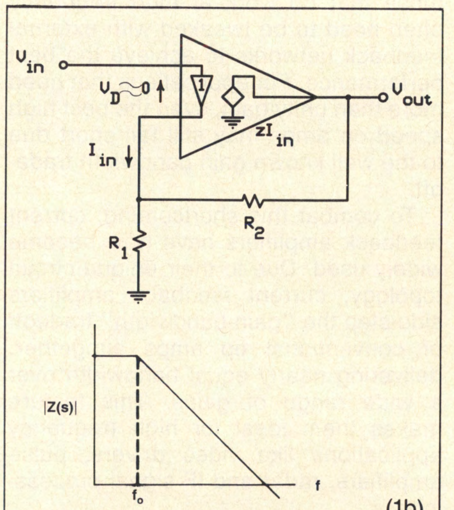
however, while the transfer function 2b is similar, the closed loop bandwidth is no longer a function of gain G. Instead, bandwidth is limited only by the ratio of R_2 to Z(s), independent of the ideal closed loop gain G. Although Z(s) decreases with frequency like A(s) for op amps (equations and graphs 4a and 4b), it does not factor into the closed loop bandwidth the same way. Thus, the closed loop bandwidth for current feedback amplifiers is independently set by R_2. In actual use, bandwidth is slightly reduced at higher gains due to the non-zero output impedance at the inverting terminal. This bandwidth reduction can easily be recovered, however, by simply lowering the value of R_2. It should also be noted that R_2 is always required, even for unity gain configurations.

Current feedback amplifiers are ideal for a wide range of high frequency applications. Figure 5a shows how the HA-5004 can be used as a basic video buffer. Setting the amplifier for a gain of +2 provides unity gain with matching 50 Ohm output and load resistances. Resistors R_{c+} and R_{c-} may be used, if needed, to limit the output current for fault conditions.

Since the HA-5004 typically requires only 12 mA of quiescent current to operate, power dissipation is rarely a problem. But for particularly demanding applications at high temperature and output current, the HA-5004 has the capability to automatically sense overheating and shut down its output stage. Special circuitry triggers when the junction temperature exceeds approximately 180 degrees Celsius. If this occurs, an open collector output signal (TOL) can be used to notify microprocessor circuitry that a thermal overload condition exists. Once disabled, and the chip temperature drops, the HA-5004 will automatically reactivate itself and resume normal operation. If desired, a TTL input (TOI) can override or prevent the disable operation; in this case, thermal overload will not be indicated by the TOL output.

By using the output enable (OE) function manually, the HA-5004 can be configured in parallel to form a multiplexed video amplifier as shown in figure 5b. The TTL-compatible Enable control turns off the amplifier so that outputs can be tied together in common. Only one pair of output protection resistors is needed in this case since only one output is active at any time. When switching takes place from one channel to the next, a small overlap may occur in which one amplifier becomes active before the last has reached a high impedance state. To prevent this condition, the enable command to each amplifier should be skewed by a few microseconds.

Another common application for the HA-5004 is to buffer the input of a flash A/D converter as shown in Figure 6. This function can be deceptively demanding on a buffer amplifier because of the high transient currents associated with CMOS



$$V_{out} = ZI_{in} = Z\left(\frac{V_{in}}{R_1} + \frac{V_{in}}{R_2} - \frac{V_{out}}{R_2}\right) \quad (1b)$$

$$\frac{V_{out}}{V_{in}} = \frac{G}{1 + \frac{R_2}{Z(s)}} \quad (2b)$$

$$G = 1 + \frac{R_2}{R_1} \quad (3b)$$

$$Z(s) = \frac{Z_o}{1 + j\left(\frac{f}{f_o}\right)} \quad (4b)$$

Figure 4b. Current feedback amplifier schematic.

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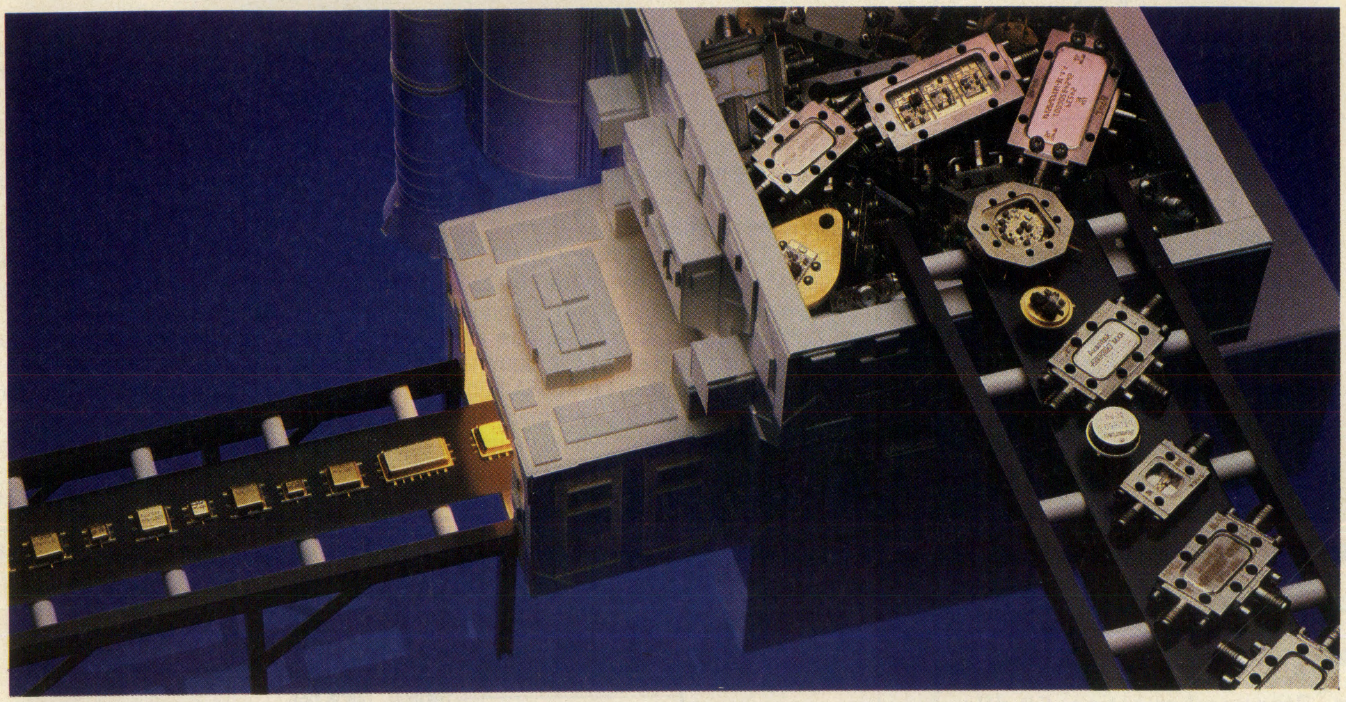

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Your next microwave system needs more functions in less space. You're ready for the advantages of microwave surface mount technology (SMT)... Avantek's PlanarPak™ components. You'll get smaller size†... lighter weight...increased board density...and, circuits on both sides of the board. Your design team can put the system you need together faster...with lower risk and more control, with off-the-shelf PlanarPak components. More than thirty different high performance PlanarPak components provide DC to 18 GHz coverage. All are 50 ohm matched for easy microstrip integration. Avantek offers the widest range of microwave functions in surface mount packages—

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- Detectors—.02 to 6.0 GHz
- Limiting Amplifiers—.01 to 1.0 GHz
- Mixers—.05 to 2.5 GHz
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—with more functions on the way. PlanarPak components make complete surface mount microwave systems a reality.

Designed for High Volume Applications in Demanding Environments

These hermetically sealed, thin-film MIC components are designed for rugged operating environments...missiles...smart munitions...expendables ...and RPVs. Wherever size and performance are critical. PlanarPak products are ideal for automated manufacturing. A proven technology...in volume production today at Avantek.

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Avantek, the world leader in microwave SMT, offers a comprehensive selection of literature, test fixtures, and technical assistance. And, all PlanarPak components are in stock at your local Avantek distributor for immediate delivery. Contact us today to receive PlanarPak products literature and the name and address of your local Avantek distributor.

†Three sizes: .25 x .25 x .11 in, .375 x .375 x .15 in and .4 x .8 x .15 in

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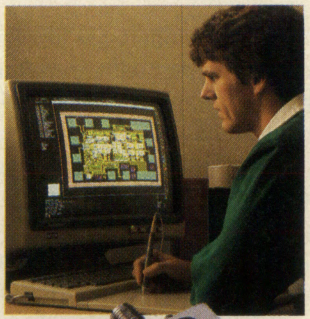
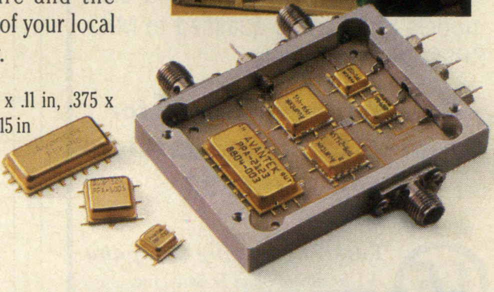




AVANTEK

INFO/CARD 30

flash converters. The buffer must quickly charge the input capacitance at the beginning of each conversion cycle, settling to within 1/2 LSB before the sample time is complete. In the case of the HI-5700 8-bit flash converter operating at 20 MHz with a 50 percent duty cycle clock, full settling must be complete within 25 ns. The 100 mA output current capability of the HA-5004 makes it ideal for fast settling, high current applications like this.

Figure 7 shows how the HA-5004 can be used in conjunction with the HA-2547 high speed multiplier to form a voltage controlled amplifier or mixer. The current output of the HA-2547 is summed at the inverting terminal of the HA-5004, providing current-to-voltage conversion the same way an op amp would be used. Using a 250 Ohm feedback resistor, the full 100 MHz bandwidth of the current feedback amplifier preserves the bandwidth capability of the HA-2547. Alternatively, the internal feedback of the 2547 may be used instead, with some sacrifice to the bandwidth.

With few exceptions, the rules for properly applying current feedback amplifiers are the same as those for high speed op amps:

- Keep all component leads as short as possible, particularly at the inverting input. Always minimize the stray capacitance at this node.
- Separate signal grounds from power grounds and connect them together at only one common (star) ground point.
- Use properly terminated coaxial cable at the input and output if they are located some distance from the amplifier.
- For best performance, use a ground plane for PC mounted devices.
- Make input and feedback resistors as small as possible consistent with the specified feedback resistance, output

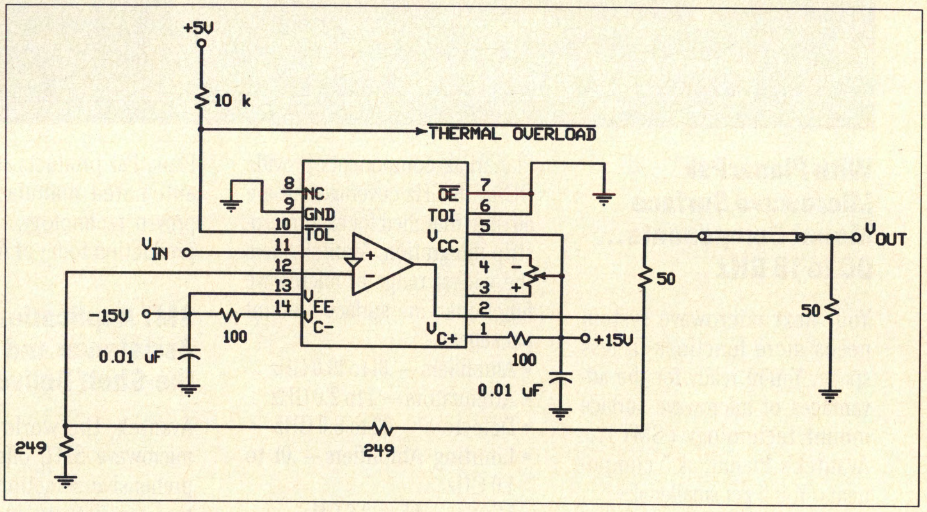


Figure 5a. Current feedback amplifier used as a video buffer.

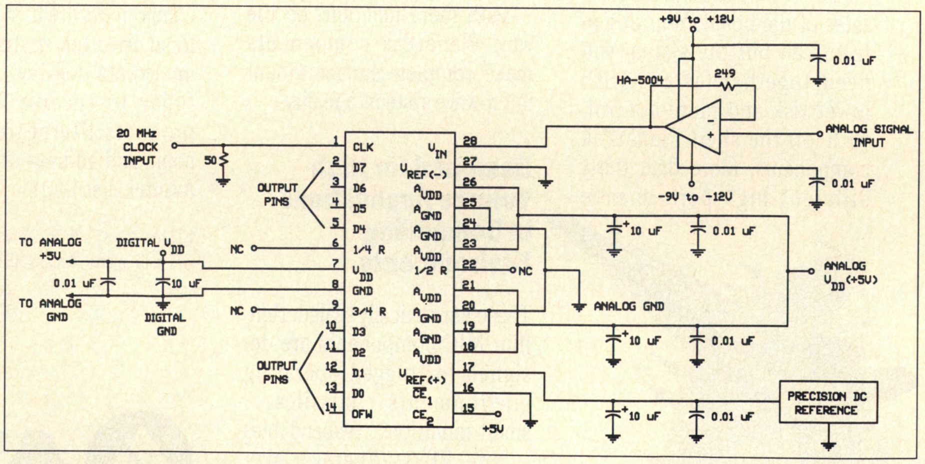


Figure 6. Current feedback amplifier used as a buffer for a flash A/D converter.

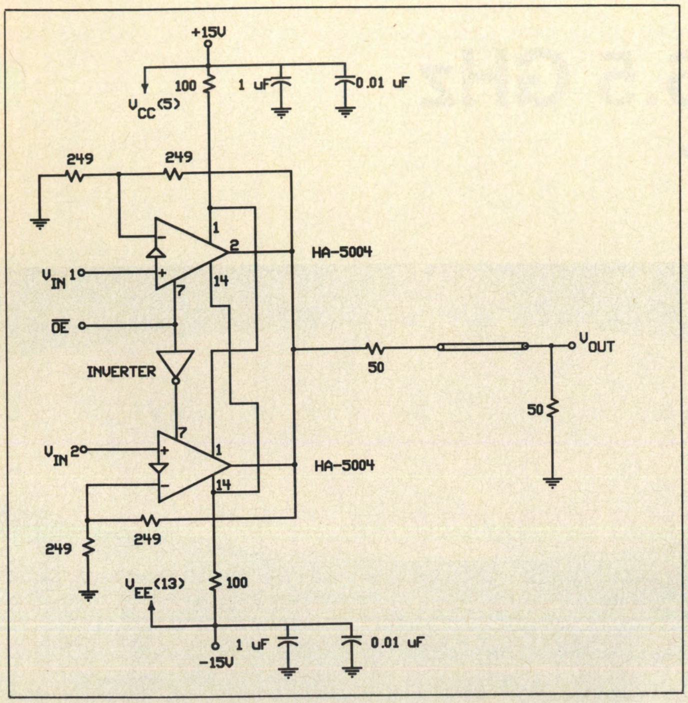


Figure 5b. Current feedback amplifier used as a multiplexed video amplifier.

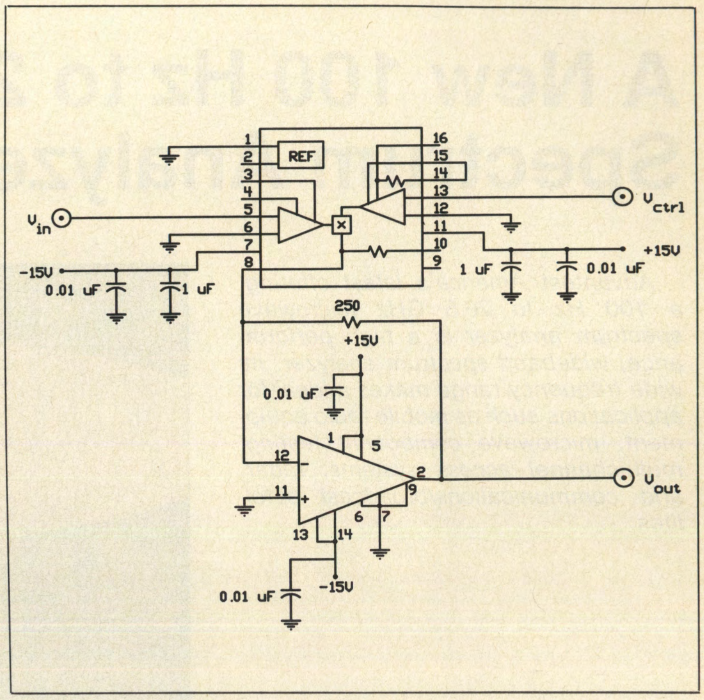


Figure 7. Current feedback amplifier used as a voltage controlled amplifier or mixer.

drive capability, and circuit requirements. In current feedback amplifiers, bandwidth can usually be increased somewhat by using lower value feedback resistors. Values too small, however, may cause stability problems.

• Use good power supply bypass capacitors and connect them right at the power supply pins. A tantalum capacitor in parallel with a ceramic capacitor gives good bypass performance at both low and high frequencies.

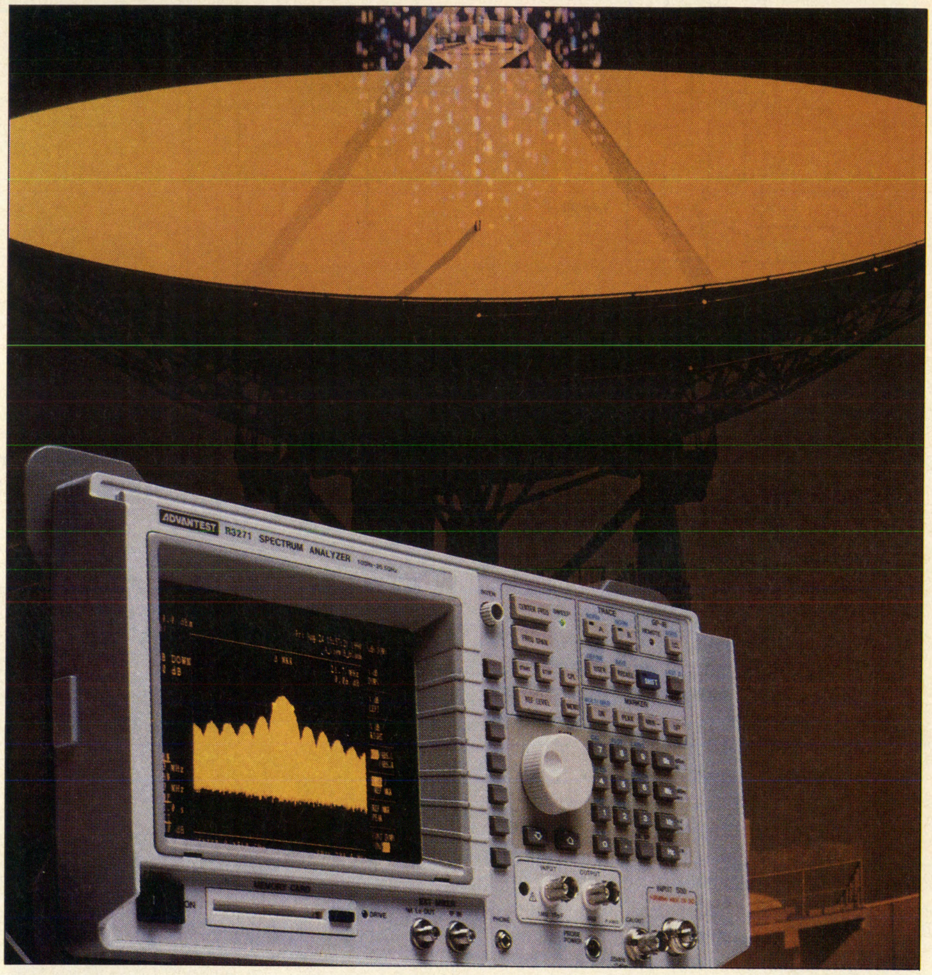
In summary, current feedback amplifiers now give designers new freedom in high frequency designs. Compared to op amps, they provide wider bandwidth at higher gains, lower distortion, better large signal performance and faster settling time with relatively low power. External compensation, often required with high speed op amps, and other kinds of external optimizing adjustments are eliminated as well. The next generation of current feedback amplifiers are certain to provide even higher levels of performance for analog signal processing applications approaching one gigahertz. *RF*

About the Author

Al Little is an Applications Engineering Manager for the Signal Processing Products Division of Harris Semiconductor. He may be reached at (407) 724-3842.

A New 100 Hz to 26.5 GHz Spectrum Analyzer

Advantest America's latest offering, a 100 Hz to 26.5 GHz microwave spectrum analyzer is a high performance, wideband spectrum analyzer. Its wide frequency range makes it ideal for applications such as mobile radio equipment, microwave component testing, multi-channel access systems, radar, and communications/broadcast satellites.



Advantest's R3271 Microwave Spectrum Analyzer.

The R3271 Microwave Spectrum Analyzer makes use of advanced high frequency measurement techniques to cover its frequency range. The frequency range of the analyzer is internally preselected, providing more than 100 dB of true dynamic range and a displayable dynamic range of 100 dB. A synthesized local oscillator provides excellent frequency accuracy. In addition, its counter mode provides accurate frequency measurements of broadcast signals up to 26.5 GHz. Its high input sensitivity, coupled with its built-in counter, make it ideal for detecting individual signals in a complex waveform at extremely low-level signals. Precise measurements of modulated and complex signals are also possible. Key specifications are listed in Table 1.

Advantest's R3271 has a standard GPIB which provides for full remote control making it an excellent choice for use in automated measurement systems. It has a controller option for built-in computing as well as a memory card.

The R3271 measures 7 inches high by 14 inches wide by 18 inches deep and weighs less than 50 lbs. These dimensions and relatively light weight make it an ideal instrument for field applications.

The Advantest R3271 is priced at $32,000 and will be available for delivery by March 1, 1991.

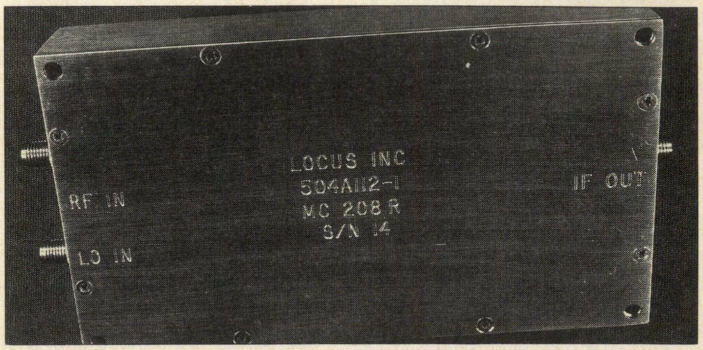
For more information on the R3271 spectrum analyzer from Advantest America, circle Info/Card #170.

Frequency Range:	100 Hz - 26.5 GHz
Residual FM (N=1):	3 Hz
Noise Sidebands:	100 dBc/Hz (with 1 kHz offset frequency)
Resolution Bandwidths:	10 Hz - 3 MHz
Noise Level (N=1):	−135 dBm
Sweep Time:	250 ms/50 μs (zero span) to 1000 s

Table 1. Primary specifications for the R3271 Microwave Spectrum Analyzer.

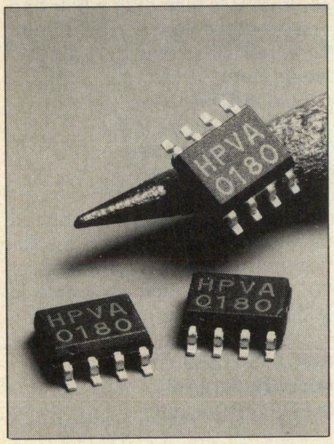
Image Reject Mixers

Locus, Incorporated announces their new lines of image reject mixers. The lines of mixers include image reject down-converters, single-sideband up-converters, and I-Q modulators/detectors. There are twenty-three image reject down-converters with image rejection ranging from 25 to 55 dB, RF-IF conversion gain from −15 to +35 dB typically, and LO drive level ranging from 10 to 17 dBm. The single-sideband up-converters have image rejection ranging from 15 to 60 dB minimum, RF-IF conversion gain ranging from −15 to +54 dB typically, and carrier suppression ranging from 15 to 45 dB minimum. The I-Q modulators/detectors have phase balance ranging from ±0.1 to ±5 degrees, RF-IF conversion gain of −14 to +3.5 dB, and LO drive level of 10 to 13 dBm.
Locus, Inc.
INFO/CARD #211



Monolithic Variable Gain Amplifier

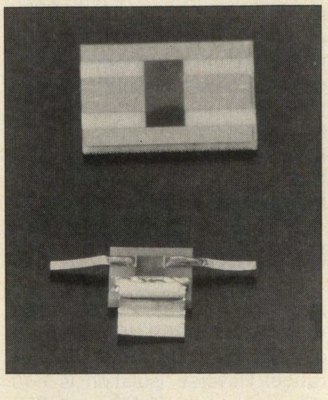
The HPVA-0180 silicon-monolithic, variable gain amplifier from Hewlett-Packard features a 3 dB



bandwidth of 2.5 GHz, provides 20 dB gain with 20 dB gain control over its DC to 2.5 GHz bandwidth, and dissipates only 250 mW from a single 6 volt power supply. It is packaged in a plastic, surface-mount SO-8 package and is suitable for wide or narrow bandwidth applications from DC to 2.5 GHz. The HPVA-0180 is designed for VHF/UHF receivers, RF data links, and broadband local area networks. Pricing for the HPVA-0180 is $14 each in quantities of 1-99, $13.25 each in quantities of 100-499, and $12.25 each in quantities of 500-999. The unit is available from HP sales offices or distributors of HP component products.
Hewlett-Packard Company
INFO/CARD #210

Surface Mount RF Attenuators

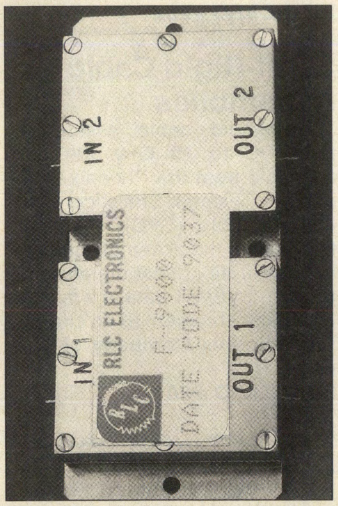
Hokuriku USA, Ltd. has released the PAD/PADU series of attenuators for both small signal and high power applications. The series is based on thin film ceramic substrates and utilizes a coaxial design. PAD types are available for applications from DC through 5 GHz, with PADU types rated at up to 100 watts for transmitting applications up to 4 GHz. Both types offer low-profile surface mounting, and impedance is 50 ohms. The PAD series can be specified for any attenuation between 1 and 32 dB, and the PADU series is available in 3, 6, 10, and 20 dB versions. The PAD series is sized at 8 mm × 18 mm, and the PADU series measures 21 mm × 11.6mm. Applications include communications equipment, CATV equipment, video systems, test instruments, and RF/microwave accessories.



These attenuators are available in OEM quantities with prices starting at less than $8 each.
Hokuriku USA, Ltd.
INFO/CARD #209

Matched Filter Sets

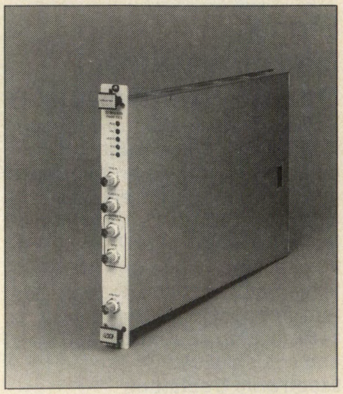
RLC Electronics has introduced a new line of matched filter sets. The filters are available with



phase, amplitude, and delay matching. Typical requirements include phase matching at ±5 degrees, group delay at ±2 ns, or amplitude matching with a ±0.2 dB variation over a portion of the filter's passband. The filter pictured is a surface mount device with two matched filters on a common carrier plate. The device is phase, delay, and amplitude matched over center frequency ±20 MHz. Full Mil-Spec environmental requirements such as vibration, shock, and temperature are applicable. Pricing for the matched filter sets start at $500 per set and are they available for delivery in small quantities in 2 to 4 weeks.
RLC Electronics, Inc.
INFO/CARD #208

VXI Arbitrary Waveform Generator

Wavetek has introduced Model 1375, a VXI arbitrary waveform generator (arb) compatible with the Standard Commands for Programmable Instruments (SCPI) remote programming format. The 20 MHz arb is a single slot C size VXI module that has 12 bit vertical resolution. It is equipped with 32K of volatile memory expandable to 128K. The module's automatic waveform scaling feature provides the ability to expand or contract a waveform into a larger or smaller memory space than originally occupied. A user can divide memory space into blocks of 8K and perform phase continuous switching between memory blocks upon receipt of a valid trigger signal either over the VXI backplane or externally. The price for Model 1375 is $3995, and the expanded memory option costs $1595.



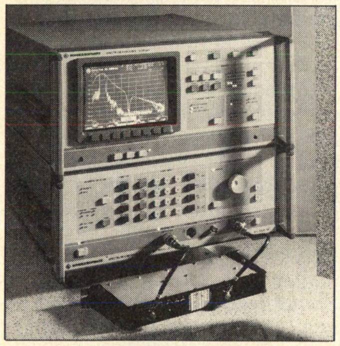
Wavetek Corporation
INFO/CARD #207

UHF Backpack Booster Amplifier

Model M200U-BP contains a 200 Watt CW from 200-400 MHz (broadband) into 50 ohms, when driven by a 3-5 Watt exciter. It is linear and can be used for voice communications and CW. RF power is monitored from an LED bargraph meter, and automatic antenna T/R switching is also built-in. 50, 100, 150, and 200 Watt power output levels can be selected and power consumption is 28 volts and 18 amperes at 200 Watts output power. The unit weighs 19 lbs.
Kalmus Engineering International, Ltd.
INFO/CARD #206

Spectrum and Network Analyzer

The Spectrum & Network Analyzer FSBS from Rohde & Schwarz features a user-selectable frequency offset of its tracking generator and sensitivity of more



than −145 dBm over the range from 100 Hz to 5 GHz. It is designed for scalar frequency-conversion network analysis and spectrum analysis.
Rohde & Schwarz
INFO/CARD #205

Signal Microprocessor Development Station

The Signal MicroProcessor (SMP) Development Station contains the hardware and software needed to program and operate a 128 tap, 100 MHz programmable transversal filter for applications such as waveform equalization, matched filtering, and pattern recognition. The device is self-clocking at a rate of 360 MHz and costs $8500.
Electronic Decisions Incorporated
INFO/CARD #204

Universal Counter

The CDC250 dual-channel counter will count signal frequency of sine, square, and triangle waves from 5 Hz to 175 MHz at input levels from 20 mV to 24 V peak. The instrument also provides period measurements, frequency ratio, time interval, and total measurement functions. It is priced at $595.
Tektronix
INFO/CARD #203

Portable 9600 BPS RF Modem

UDS has announced the DR 96, a 9600 BPS radio frequency modem that uses the 470 MHz frequency band and has a sensitivity of 0.35 μV. The DR 96 can transmit in both synchronous and asynchronous modes, and it has 10 ms RTS and CTS signaling time. Initial pricing is set at $1,295.
UDS
INFO/CARD #202

Semi-Rigid Cable Assemblies

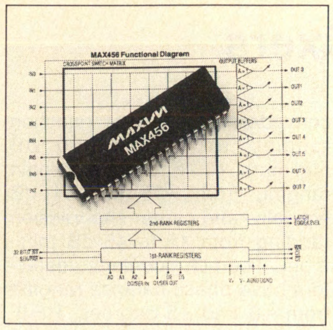
Semi-rigid cable assemblies from DC to 18 GHz are now available from the Phoenix Company of Chicago. Statistical Process Controls (SPC) are used throughout the assembly process to control all critical dimensions. SPC and VSWR data is available with each shipment at no charge.
The Phoenix Company of Chicago
INFO/CARD #201

Toroid Transformer Tester

Atlantic Magnetics has announced a toroid transformer tester for in-process testing and inspection of transformers and inductors wound on toroid cores. Model 3500A handles toroid transformers with inside diameters from 0.060" to 8.0". It completely tests transformers of 10 to 2000 turns in less than 15 seconds with an accuracy of ±0.2 percent or better. Model 3500A is priced at $1995.
Atlantic Magnetics, Inc.
INFO/CARD #200

8 × 8 Video Crosspoint Switch with Buffers

The MAX456 uses a digitally controlled 8 × 8 switch matrix to



connect eight high speed signals to any or all of the eight output channels. It has eight switch-matrix outputs that connect to eight 35 MHz, 250 V/μs video buffer amplifiers that can be disabled under digital control. Channel control logic is also included on chip. Single channel crosstalk is −70 dB at 5 MHz, and prices start at $19.97 for quantities of 1000 and up.
Maxim Integrated Products
INFO/CARD #199

SPDT Switch

Model 62P004, an SPDT switch, offers 80 dB isolation from 410 MHz to 2300 MHz. It is TTL compatible and is available with SMA connectors. Video leakage is less than 10 mV in a 100 MHz bandwidth, insertion loss is 3.5 dB maximum, and VSWR is 2:1 maximum. It is priced at $397 in quantities of 10-24.
ECM Devices, Inc.
INFO/CARD #198

High Speed 250 MHz Counter

The KL-5402D covers the 10 Hz to 250 MHz frequency range and measures 1ppm in 0.02 seconds. Counting accuracy is independent of input test frequency, and the unit is RS-232C compatible. Resolution is 1 ppm at 0.02 seconds gate time and improves to 0.0001 ppm at 200 seconds gate time.
Kolinker Industrial Equipment, Ltd.
INFO/CARD #197

RF Amplifier

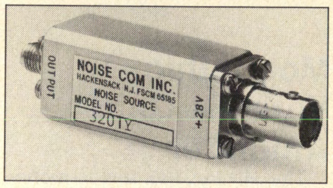
Amplifonix Model TM6181 RF amplifier features 2.5 dB max. noise figure and 9 dB gain max. over the 10 to 400 MHz frequency range. Reverse isolation is −10 dB max. and VSWR is 2.0:1. Screening to the tables of Mil-Std-883 is available.
Amplifonix
INFO/CARD #196

5 GHz SPDT Switch

Mini-Circuits' YSW-2-50DR SPDT switch has a built in driver and operates over the DC to 5 GHz range with 3 ns switching speed. It has greater than 40 dB isolation in the off state and less than 1 dB insertion loss in the on state. It is priced at $19.95 in quantities of 1-9.
Mini-Circuits
INFO/CARD #195

Broadband Noise Source

The NC 3201Y is a broadband coaxial noise source that delivers white Gaussian noise from 10



kHz to 1100 MHz. Noise output rise and fall times are less than 1 μs, VSWR is less than 1.2:1, and noise output variation with temperature is less than 0.01 dB/C.
Noise Com
INFO/CARD #194

8 Bit 100 MHz Flash ADC

The MN5901 is a high speed monolithic ADC with 8 bit resolution and a conversion speed of 100 MHz. Differential linearity is ± 0.95 LSB maximum and SNR is 38 dB minimum. Prices are from $75 in quantities of 100.
Micro Networks
INFO/CARD #193

Alternating Voltage Measuring System

The model 4920 Alternating Voltage Measurement Standard from Datron Instruments measures signals from 1 Hz to 1.25 MHz with uncertainties of ±28 ppm in stand-alone mode. It measures from 100 mV to 1000 V RMS and is priced at $9,995.
Datron Instruments Ltd., A Division of Wavetek Corporation
INFO/CARD #192

Self-Biased Gain Block Amplifier

A GaAs monolithic gain block amplifier for use in the 2 to 10 GHz frequency range has been developed by Texas Instruments. It uses on-chip DC blocking which

allows it to be directly cascaded. Output power of the TGA8810 at 1 dB compression is typically 14 dBm and gain is 15.0 dB nominal at 5 V and 80 mA.
Texas Instruments
INFO/CARD #191

2-Way Power Divider

The model PD032-2 is a 2-way power divider spanning from 300 MHz to 2000 MHz. Insertion loss is 0.8 dB, isolation is 18 dB, VSWR input is 1.35:1, and VSWR output is 1.3:1. Amplitude balance for the PD032-2 is 0.2 dB and phase balance is 3 degrees.
Microwave Research and Development, Inc.
INFO/CARD #190

DSP Array Processing Chip Set

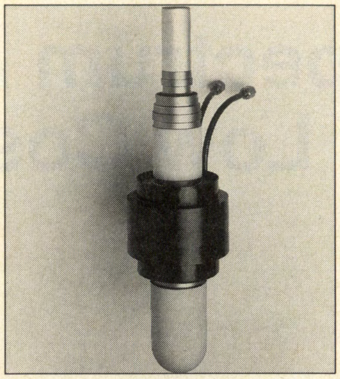
Array Microsystems, Inc. has released a DSP chip set in a 144-pin grid array. The set includes the A66111, digital array signal processor and the A66211, a programmable array controller. The A66111BCG and A66211BCG each sell for $495.
Array Microsystems, Inc.
INFO/CARD #189

1000 MHz Frequency Synthesizer

The PTS 1000 frequency synthesizer covers the 0.1 to 1000 MHz frequency range with resolution from 0.1 Hz to 100 kHz. Harmonics are at −30 dB at full output, phase noise is −60 dBc, and noise floor is at −130 dBc/Hz. The PTS 1000 is priced at $11,500.
Programmed Test Sources, Inc.
INFO/CARD #188

High Power Magnetron

Burle Industries has introduced the C94604E 60 kW CW magnetron. The C94604E generates 60 kW of power at 915 MHz with an anode input of 16.1 kV and 4.4 amperes for a tube efficiency of 85 percent. Tube warranty of the C94604E is within two years of the tube's shipping date.
Burle Industries, Inc.
INFO/CARD #187



Ferrite-Free Divider

Sage Laboratories has released an 8-way reactive power divider which covers the range of 250 to 300 MHz. Typical performance over the band features insertion loss of less than 0.2 dB, input return loss of greater than 20 dB, amplitude unbalance of less than 0.1 dB and phase unbalance of less than 1.2 degrees.
Sage Laboratories, Inc.
INFO/CARD #186

Programmable Pulse Generators

The PSPL Model 10,000A features 40 volts amplitude and 400 ps risetime along with a GPIB, IEEE-488 interface. The amplitude can be adjusted down to 26 mV in 1/8 dB steps, and the pulse duration is adjustable from 1 ns to 100 ns in 25 ps steps. Negative polarity pulses to −40 V are also included.
Picosecond Pulse Labs, Inc.
INFO/CARD #185

Digitally Refreshed Spectrum Display

M/A-COM Government Products has released Model DRD-3572, a Digitally Refreshed Spectrum Display. The display has a wideband sweep of 40 MHz, and IF inputs at 70 MHz or 160 MHz are available. Other features include 70 dB dynamic range and two amplitude-calibrated adjustable markers.
M/A-COM Government Products, Inc.
INFO/CARD #184

Spread Spectrum ASIC Eases Design of Low Cost Part 15 Systems

By Raymond W. Simpson
O'Neill Communications, Inc.

Part 15 of the Federal Communications Commission's Rules governs the operation of RF communications devices without an individual license. Previously, Part 15 devices were limited to extremely low transmitted field strengths (low effective power), which reduced the ability to provide efficient, reliable communications in a mass marketed, no-license device. The power limits were necessary to assure that mass marketed, license-less devices would not cause harmful interference to vital communications services. Recognizing that spread spectrum techniques reduce the potential for harmful interference, in 1985 the FCC allowed the use of power output levels up to 1 watt in three frequency bands which are primarily used for industrial, scientific and medical (ISM) applications (although other services such as government systems, automatic vehicle location, and amateur are also permitted to operate in these bands). The Canadian government has also recognized the potential of unlicensed spread spectrum operation and is now issuing technical acceptances of equipment under rules which are similar (but not identical) to the FCC rules (1).

The FCC Rules (2) permit operation in the bands 902-928 MHz, 2.4 -2.4835 GHz, and 5.725-5.850 GHz using direct sequence and frequency hopping spread spectrum (explained below). Other spreading schemes are not currently permitted under this section of the rules, although hybrid modes have just been added (3). There is no limit on the antenna gain which may be used (until 1994) (4), up to 1 watt power output may be used, and, except for the so called "forbidden bands" (5) the out-of-band emission suppression requirements are not stringent. Operation under Part 15 is on a "sufferance" basis, i.e. the Part 15 system must not cause harmful interference to licensed services and must accept any interference caused by other services. Thus, the only protection a Part 15 device can have from interference is in the cleverness of its design.

O'Neill Communications, Inc. has developed a wireless data communications product, the LAWN[R] for interconnecting personal computers and peripherals and a spread spectrum RF modem, shown in Figure 1. They both use an application specific integrated circuit, the OCI-100 which is being marketed separately for use in low cost Part 15 spread spectrum systems.

Frequency Hopping and Direct Sequence

Frequency hopping (Figure 2a) is the

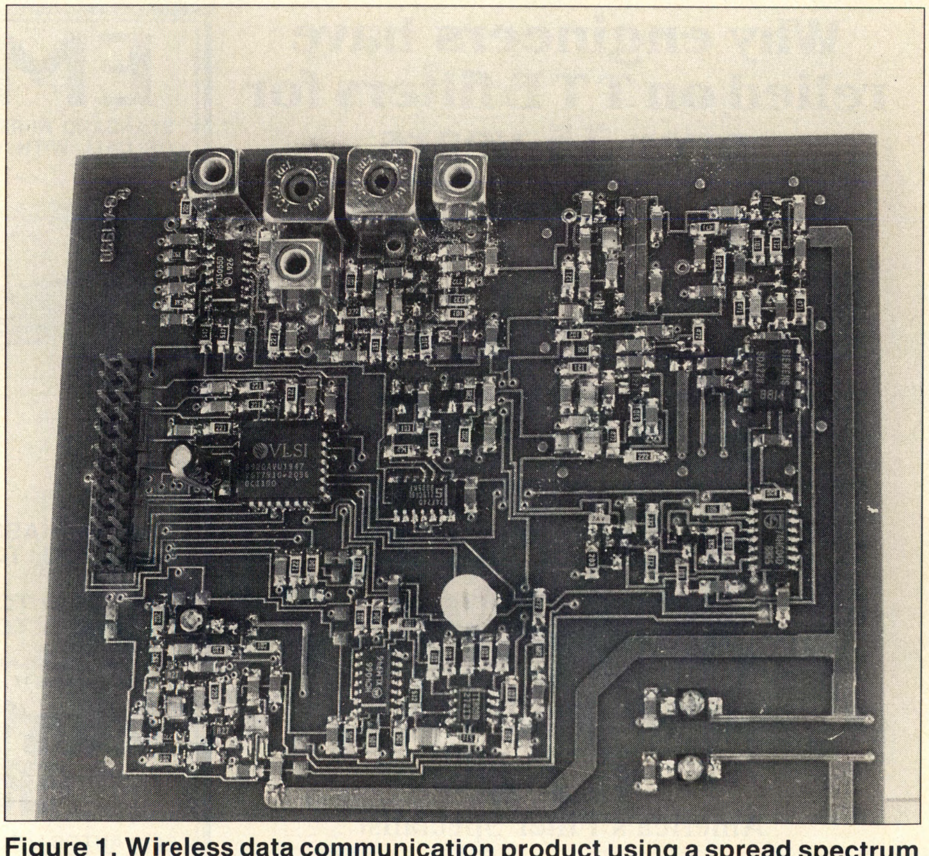


Figure 1. Wireless data communication product using a spread spectrum ASIC for interconnecting PCs, peripherals, and a spread spectrum modem.

easier to understand of the two FCC accepted spreading methods. A frequency hopping radio is just a conventional radio communication system which frequently (at least every 400 ms under the rules) changes its operating frequency, usually in a pseudorandom manner. The details of finding other units, maintaining synchronization in a multi-unit system, and making fast enough synthesizers, make the design of working, low cost systems a challenging undertaking. As the OCI-100 ASIC is not optimized for frequency hopping systems, they will not be discussed further here.

Direct sequence spread spectrum (DSSS) relies on combining a high rate (usually) binary sequence with the signal to be transmitted in such a way that the high rate sequence dominates the modulation bandwidth and directly determines the spread bandwidth (see Figure 2b). The high rate sequence is called a chipping sequence and the high rate bits are called "chips" to distinguish them from the information bits. This can be easily accomplished by adding the high rate sequence modulo-2 to the data to be transmitted, i.e. exclusive OR the two streams, as shown in Figure 3. An alternate way of looking at the process is to consider the chip sequence to be a series of plus or minus one elements, and view the modulo-2 addition as a multiplication (view the XOR as a digital balanced modulator). The two streams don't have to be synchronous, but simplifications in clock recovery can be achieved at the receiver if an integer number of chips are sent for each data bit, as in the OCI-100.

Most of the receiver signal processing in the direct sequence method can be done in the digital domain with relatively simple circuits, thus it is usually less costly to implement simple DSSS than frequency hopping SS.

To receive a DSSS signal coded as above, it is possible to simply multiply the received chips with a locally generated, synchronized copy of the chip sequence. The multiplication can be done at baseband, IF or RF. The output of the multiplication is then low pass filtered (bandpass filtered if RF or IF processing is used) to provide the despread, filtered data. If a baseband implementation is used, the multiplication can be accomplished with an XOR gate.

Synchronizing the local copy of the chip sequence can be accomplished by a variety of techniques (6) such as the delay locked loop. Delay locked loops require an acquisition time to attain synchronization, which can be a problem in asynchronous communication applications.

Another method of despreading a DSSS signal is to use a matched filter, especially a digital matched filter (7). The digital matched filter is just a shift register with taps at each stage. The output of the filter is the weighted sum of the outputs of the taps. Usually, binary weights suffice. The digital matched filter may be viewed as a digital approximation to an analog matched filter using a 1 bit analog to digital converter (i.e. a comparator).

The comparator should make the decision on each chip at the time at which the signal to noise at its input is maximum. This requires either recovery of the chip clock or operation at an oversampled rate. The latter approach was selected in the design of the OCI-100 as it allows a simpler implementation in CMOS and faster sequence acquisition. The received chips are sampled at 4 times the chip rate, and the correlation peak on each bit (16 chips, or 64 samples) is detected and used for

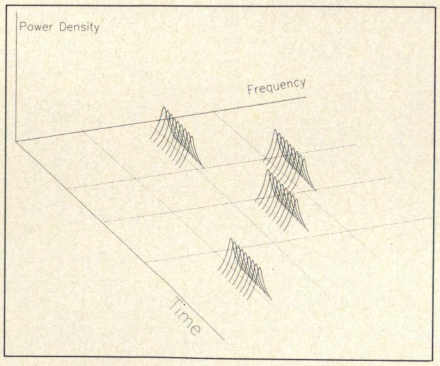


Figure 2a. Graphical representation of frequency hopping spread spectrum.

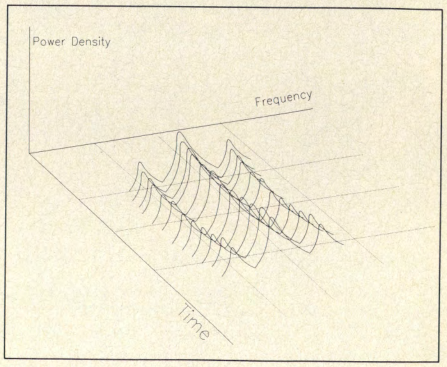


Figure 2b. Graphical representation of direct sequence spread spectrum.

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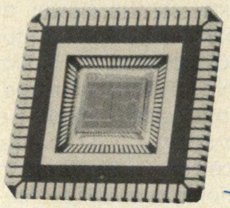

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bit clock recovery.

In theory, this scheme requires only 1 bit to synchronize, but for greater robustness the OCI-100 uses 3 bits to decide it has acquired bit clock. It then tracks the bit clock with a digital phase locked loop and detects loss of synchronization by the absence of 16 consecutive bit clocks.

To achieve a smooth transmitted spectrum, so the power density does not peak at a few frequencies, the OCI-100 includes the option of further randomizing the transmitted chips using a maximal length linear feedback shift register scrambler at the transmitter. At the receiver, a self-synchronizing descrambler is used to reverse the process before the receiving digital matched filter. The 33 stage shift register used provides a high degree of randomness in the transmitted chip pattern.

Part 15 Spread Spectrum Contrasted with Military Systems

Military spread spectrum systems are designed to withstand intentional jamming from a determined, intelligent adversary in a combat situation and/or to provide low probability of detection. The cost of communication system failure is high (loss of aircraft, personnel etc.), so the dollar scale on which the cost of the communication system is minimized is relatively high. Part 15 devices, however, are often subject to stringent cost constraints so a compromise between cost and performance must be struck, which often results in a Part 15 system spread spectrum system being very different from military spread spectrum systems.

For Part 15 devices, the sources of interference are indifferent, i.e. they are not actively seeking to jam or intercept Part 15 communications. Furthermore, the user usually has some form of premises control, i.e. he can choose which RF based system he will operate within his household or business. Most Part 15 operations are relatively short range, at least as far as the individual RF links are concerned, so the desired signal can be dominant by virtue of

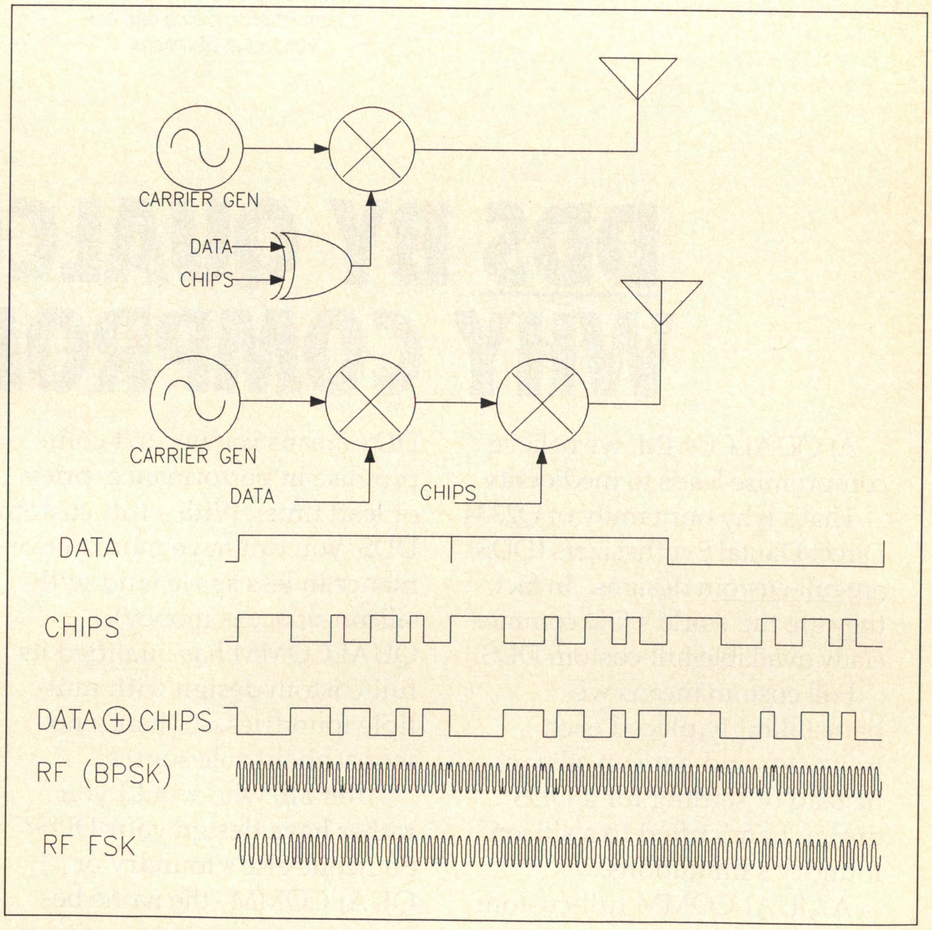

Figure 3. Distinguishing high rate sequence from information bits by using an exclusive OR.

shorter range. Finally, most reported Part 15 spread spectrum systems are digital, so they are operated only in the domain in which reliable digital operation is obtained. Operation under marginal conditions, which is more common in voice radios, is usually not a consideration.

A major consideration for Part 15 spread spectrum systems is the multipath environment, especially within buildings (8, 9). Within buildings, multipath propagation causes a pattern of nulls to be generated when the signal components arriving by the various paths arrive out of phase and nearly cancel. The spacing between nulls is related to the wavelength; in the 915 MHz band the nulls tend to be spaced within 6 to 12 inches of one another. A narrowband system has a non-negligible probability of being in a null for a given path. The wider bandwidth of a spread spectrum system effectively diffuses the nulls, reducing their depth. At OCI, we have seen that even relatively simple spread spectrum techniques can provide significant improvements in the reliability of indoor communication relative to narrowband techniques.

The goal of a Part 15 spread spectrum is to reclaim otherwise wasted spectrum, providing a benefit to the public which would not otherwise be as readily available or affordable. The spread spectrum attributes of the system are designed to reduce the likelihood of harmful interference to other systems and to allow the system to withstand the challenges of its environment at a price the intended market can afford.

Unfortunately, there has been a tendency, in the less technical press, toward simplifications of the type "spread spectrum is interference free". In fact, given a time and bandwidth allocation, there are about

$$D = 2BT \qquad (1)$$

degrees of freedom in which to disperse a spread spectrum signal (6). Thus there is a limit to the number of fully separable signals which can occupy a given time bandwidth product, and beyond this there will be interference, even in an ideal implementation. The spreading process at the transmitter spreads the transmitted power, so the power density (W/Hz) is reduced. Thus there is less power in any single narrowband channel, but there is some transmitted power in more narrowband channels.

The receiver reverses the spreading process or "despreads" the signal. The despreading process actually spreads signals which are not correlated to the spreading function, so that narrowband interferers are spread. The despreader is followed (at least equivalently) by a narrowband filter, which passes the despread signal but rejects the major portion of the uncorrelated interfering signal (which was spread rather than despread). The gain in signal to interference ratio under these conditions is often called the processing gain, G_p:

$$G_p \leqslant \frac{B_{RF}}{B_I} \qquad (2)$$

where B_{RF} is the spread RF bandwidth and B_I is the information bandwidth. In real systems, the inequality always holds. For "successful" communication to occur, the received desired signal power, P_d in dB, must obey

$$P_d > P_I + J - G_p \qquad (3)$$

where P_I is the interfering power and J is the jamming margin, i. e. the required final signal to interference power required by the demodulator to operate at a "successful" level of performance.

A consequence of the above equation and the available bandwidth of 26 MHz in the 902-928 MHz band is that spread spectrum processing gain, (which may be thought of as the spread spectrum analog of selectivity), will be limited unless the data rate is kept low. For instance, a 26 kb/s data system would have at most 30 dB of processing gain. Such a system would occupy the entire 902-928 MHz band; if it caused or received harmful interference to or from a licensed service, there would be little the operator of the Part 15 system could do to resolve the problem, except to cease operating. It may not be in the best interests of the system designer, therefore, to rely only on processing gain to prevent interference. If the example system were instead designed with 6 dB lower processing gain, it could offer the user a choice of 4 sub-bands in which to operate, providing an additional dimension of flexibility for solving interference problems.

Now Just Add a Radio...

A low cost spread spectrum system needs a low cost radio as well as low cost processing. Most of the elements of low cost 915 MHz radios are readily available, if a suitable radio architecture is used. A modulation method and frequency control scheme which help control cost should be selected.

The most common modulation schemes used in spread spectrum applications are binary phase shift keying (BPSK) and frequency shift keying. Binary phase shift keying is the "classical" approach and provides better rejection of inband interferers than FSK (at least in low cost realizations). Its main weakness is in the carrier recovery loop if one is used.

FSK helps in low cost realizations,

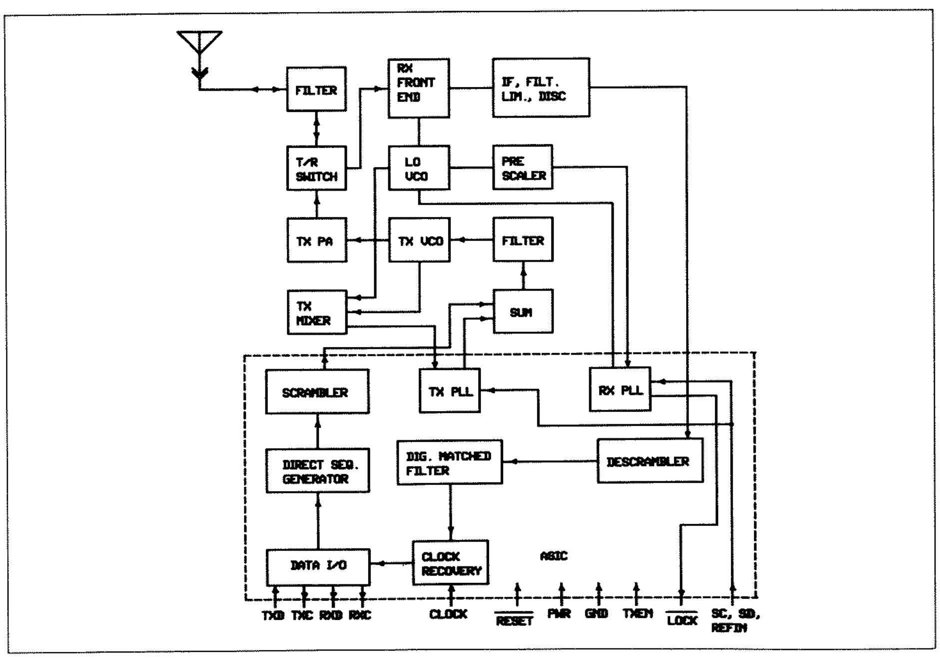

Figure 4. An FSK spread spectrum transceiver.

because complete IF/demodulator systems are available in low cost single chip form, such as the Motorola MC13055 and MC3356 (10). At 16 chips per bit, the FSK approach tends to produce a more uniform spectrum than BPSK. Sidelobe control can be easily obtained by low pass filtering the chip stream between the OCI-100 output and the modulator. This results in a compact, smooth spectrum with most of the transmitted energy in the main lobe. An FSK transmitter can be as simple as a VCO, as shown in Figure 4.

The performance of FSK using non-coherent detection and post-detection despreading is compared with non-coherent detection following ideal despreading, both using 16 chips per bit, in Figures 5 and 6. Figure 5 shows the bit error rate as a function of carrier to noise ratio (in terms of bit energy E_0 and the noise power spectral density, N_0) with a constant signal to jammer ratio of 3 dB. Figure 6 compares the probability of successful packet completion, assuming 256 byte packets with no forward error correction. The figures show that in most practical cases (i.e where a probability of packet success must be better than 90 percent) the difference in jammer rejection is less than 3 dB. It must be considered that the performance of a real system using pre-detection despreading will be degraded from the performance of the ideal despreader by the error (phase noise) in its synchronization of its local chip sequence with the chip sequence of the received signal.

The penalty for using post-detection despreading with FSK in additive white Gaussian noise is about 3 dB relative to ideal despreading. This is also an acceptable tradeoff for most Part 15 applications.

Frequency control at low cost is usually a problem if phase locked loop (PLL) ICs must be added to the design. Most of the standard PLL ICs are designed for the land mobile radio market, and are optimized for closer channel spacings than are needed for spread spectrum systems. Achieving the close channel spacing requires the use of a low reference frequency, with consequent longer loop lock up time. The OCI-100 contains two internal phase locked loops which, with external prescalers, can be used to control the receiver local oscillator and transmitter frequencies.

The receiver PLL uses a single modulus divide by 64 prescaler, such as the

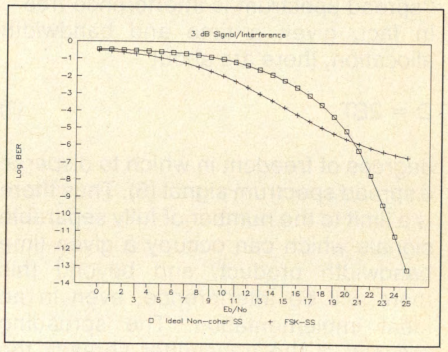

Figure 5. Bit error rate vs. SNR for comparing FSK (using non-coherent detection and post-detection despreading) and non-coherent detection following ideal despreading.

Siemens SDA 2211 or similar. The single modulus prescaler saves cost relative to a dual modulus prescaler. The subsequent dividers and a tri-state frequency/phase detector are contained in the OCI-100. (There is a provision for using external phase detectors with the internal counters of the OCI-100 if desired.) The tri-state phase detector may be used with either passive or active loop filter, depending on the system requirements.

It is usually desirable for the transmitter to have a minimum turn on delay, because the turn on time represents an overhead during which there is no data being transmitted. In carrier sense multiple access (CSMA) packet radio systems, long turn on time increases the probability of collisions. To reduce the turn on time as well as cost, a mixing type of transmitter control loop is used in the OCI-100 design. The transmitter VCO is mixed with the receiver local oscillator (if the two circuits are on one board, it's almost inevitable that this mixing will occur, so you might as well make use of it as fight it). Mixing translates the transmitter frequency down to the radio's intermediate frequency without frequency division.

Depending on the chosen intermediate frequency, the difference frequency may be used directly or divided externally by 2 or 4 before being applied to the OCI-100. The result is a PLL with a much higher sampling frequency, therefore much quicker frequency acquisition than would result if the transmitter VCO frequency was divided down to the reference frequency. Yet, since the transmitter is referenced to the receiver local oscillator, it can have the same tuning resolution as the receiver. This approach does require the receiver local oscillator to be stable enough not to lose lock while the transmitter is keying up, but this has not proved to be much of a problem in practice.

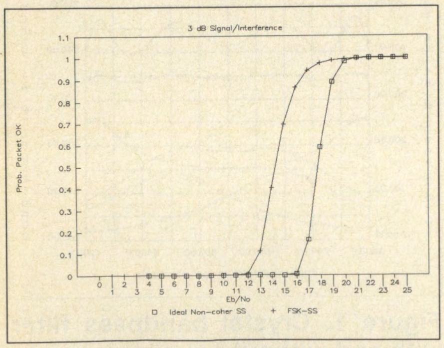

Figure 6. Probability of success.

Operation of the OCI-100 at 50 kb/s with a peak frequency deviation of 200 kHz will result in a 6 dB bandwidth of 640 kHz, comfortably above the FCC minimum of 500 kHz. Operation at rates up to 125 kb/s is supported by the OCI-100.

Conclusion

Direct sequence spread spectrum communications devices can be simple enough to be manufactured at low cost, if small reductions in performance relative to "classical spread spectrum architectures" are acceptable. The radio section needn't be very different from the usual FM radio and the spreading/despreading functions can be handled by a low cost integrated circuit. The resulting package can be low in cost, small and operate at low power consumption. *RF*

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Crystal Delay Equalizers

By William B. Lurie
Consultant

In a previous article (August, 1989), a method of providing envelope delay equalization for crystal bandpass filters was discussed, ending with the design of a single equalizer section in semi-lattice form, containing two crystals. More often than not, the amount of equalization needed necessitates the use of more than one "section" of equalizer. In general, these are, over the narrow band of interest, constant-impedance devices, and they can be cascaded with little interaction.

In practice, as was seen in Figure 7 of my previous article (1), there is a need to provide additional inductors, beyond what the basic filter and equalizer designs actually require. Inductors, in general, are the least desirable components, as compared with crystals and capacitors, because of their size, dissipation factor, and their temperature characteristics. A method has been devised, therefore, for combining a number of equalizers into a single section, putting multiple crystals in parallel in each of the two branches of the half-lattice, thereby minimizing the number of added inductors.

As an example, consider a filter with a bandwidth of 8 kHz at a center frequency of 10.7 MHz. A typical 6-pole filter would have delay versus frequency as shown in Figure 1 (A). Assuming that the delay requirements are such that two equalizer sections are required, it is possible using various available programs to arrive at design constants for the equalizers. Equation 1 shows the equalizer transfer function, and in Figure 1, curves B, C, and D show the delays of equalizer sections alone and together. Figure 1 also shows the composite delay of the filter with both equalizers added (curves E and F, where the scale for curve F is the axis to the right, expanded 5 times compared with E). It is readily seen that the delay variation without equalizers is 105 microseconds across the 3 dB passband, but only 18 microseconds with the equalizers added.

What is sought next is a single all-pass lattice which has the same transfer function as:

$$S_{12} = \frac{(p - p_1)(p - p_2)(p - p_3)(p - p_4)}{(p + p_1)(p + p_2)(p + p_3((p + p_4)}, \qquad (1)$$

where

$p_1 = -2253 + j10698420$
$p_2 = -2253 - j10698420$
$p_3 = -2252 + j10701580$
$p_4 = -2252 - j10701580$

all in Hertz

Each single equalizer section has two frequencies at which the transfer phase is a multiple of 180 degrees (F1 and F2 in Figure 2a), and the combination of two such equalizers, in cascade or combined, must have four such frequencies (F3 through F6 in Figure 2b). The location of these frequencies is not immediately obvious, but a computer program has been written to accomplish this.

A straightforward analysis of the phase of the transfer functions of the two single equalizers, added together, gives the phase versus frequency of the combined equalizer. It is a simple task to examine the phase curve or table, visually or by computer, to find the frequencies at which the phase becomes all multiples of 90 degrees. In the example, these were found to be as shown in the following table.

PHASE	FREQUENCY(Hz)
90	10694194
180	10697250
270	10698698
360	10700001
450	10701305
540	10702753
630	10705810

Now, by a process of mathematical induction, the schematic of the combined half-lattice is as shown in Figure 3, with all element values listed. Note that the shunt capacitance in the "A" branch has been arbitrarily set at 4 pF, and in the "B" branch, −4 pF. These values, knowing the pole and zero locations of the branch reactances, uniquely determine all the element values, simply by using a partial-fraction expansion (program available, of course). The scaling of impedance level to match the filter itself is trivial.

As discussed for the single-section equalizer, the negative capacitance is swamped and made positive by the "borrowing" of capacitance from sources external to the lattice (1). The process of combining equalization properties of several sections into one section does serve to lower the number of inductors and capacitors as well, but it is not entirely without penalty. Although quartz crystals have stability and Q far better than coils, capacitors, or combinations of both, they suffer nevertheless from certain practical limitations. Their motional inductance or capacitance can only be obtained in a rather narrow range of values. The setting tolerance, at 10.7 MHz, is of the order of ± 100 Hz, and the motional parameters, in practical manufacturing terms, can only be held to about 2 percent. None of these factors is any more serious for the combined equalizer than for the individual sections, but it is more difficult, with fewer adjustments possible, to tune up or align a more complex section than two simpler ones. It is beyond the scope

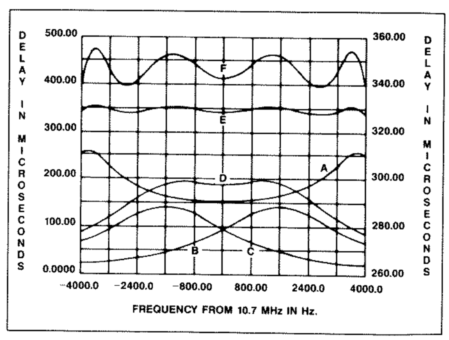


Figure 1. Crystal bandpass filter with and without equalizers.

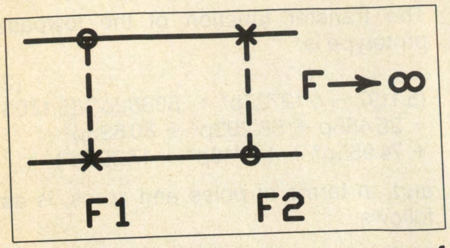


Figure 2a. Branch reactances of single delay equalizer.

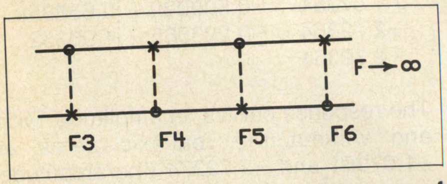
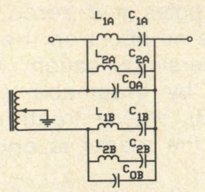


Figure 2b. Branch reactances of combined equalizers.

of this article to go into more details, but it is surely possible to program a computer to perform a tolerance study and predict yields and performance, using statistical data and assumptions regarding component variations. In addition, there are secondary methods of changing the equivalent crystal frequency and motional capacitance using additional tuning capacitors external to the crystals.

This series of articles has, so far, discussed some of the aspects of the creation of band-pass devices with envelope delay made more constant by the addition of one or more constant-impedance all-pass networks in cascade. In many respects, this technique has distinct drawbacks, as pointed out by Watanabe (4). It does seem less than optimum to design for amplitude first, create a certain amount of delay non-uniformity, and then correct the delay back toward uniformity at substantial cost. It should be pointed out, however,



Crystal 1A	0.0402249055 H	5.50616365 fF
Crystal 2A	0.179071127 H	1.23521117 fF
C_{OA}	4 pf	
Crystal 1B	0.178985359 H	1.23640542 fF
Crystal 2B	0.0401877843 H	5.49929652 fF
C_{OB}	−4 pf	

Figure 3. Combined equalizers with element values.

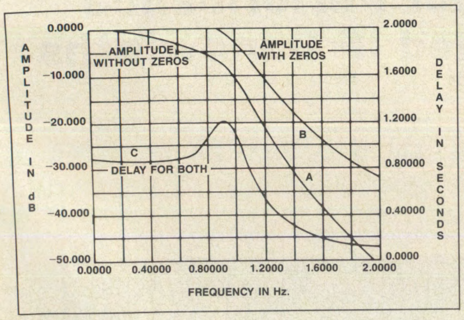


Figure 4. Rhodes N=6 low pass filter.

that the state of the art, as it progressed, allowed no other solution. More modern mathematical and network theory methods have presented several viable alternatives.

In modern network methods, the transfer function of the system (or filter) is created, such that the system's needs are met, and then a network is synthesized by one or another technique, to have that transfer function. In terms of classical filter-plus equalizer methodology, one might say that the transfer function is created by paying attention only to amplitude, and then an all-pass transfer function is added to correct the phase (or delay), and then the two parts are synthesized and combined.

One alternative approach would be to design a transfer function loosely based on consideration only of the delay, which implies that the passband amplitude corners might be severely rounded, rolling off too soon. Then the amplitude could be corrected using corrector poles and zeros, so located as to have no effect on the delay. This is an oversimplification of methods proposed by Watanabe (4), and by Rhodes (3), whose treatments of this approach the reader is encouraged to study.

As an illustration of this technique, consider a bandpass filter described by Herzig and Swanson (2). The transfer function of the lowpass prototype is:

$$(5.1197 - 5.12723p^2 + .59882p^4)/(5.1203 + 25.455p + 58.203p^2 + 80.695p^3 + 74.955p^4 + 44.743p^5 + 17.202p^6) \quad (2)$$

and, in terms of poles and zeros, is as follows:

ZEROS	POLES
+1.07351	−0.567877 ± j0.202613
−1.07351	−0.509092 ± j0.658859
+2.72364	−0.223565 ± j1.06352
−2.72364	

The response curves, in amplitude, with and without the real-axis zeros at ±1.07351 and ±2.72374 (the amplitude correctors) are curves A and B of Figure 4. Correspondingly, the delay curve of Figure 4 is applicable to both cases, since the two pairs of zeros, being symmetrical about the $j\omega$ axis, have no effect on the delay. The bandpass filter was designed from this low pass prototype, based on a polynomial derived from Rhodes' articles, and an entire class of lowpass filters can be (and has been) created from these, as will be discussed in a separate article. In many respects, this class of filters is quite interesting, since, for any degree, a parameter can be selected by the designer, affecting the rate of cutoff in the stopband with relatively minor effect on the passband delay, and even yielding frequencies of infinite attenuation similar to the Elliptic filters. *RF*

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William Lurie has worked as a mathematician, physicist and electronics engineer in a variety of fields, including magnetic compasses, X-ray tubes and measuring equipment, airborne Doppler radar, and filters. He holds Life Senior Member status in the IEEE. He works as an independent consultant and can be reached at 8503 Heather Place, Boynton Beach, FL 33437. Tel.: (407) 369-3218.

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Amplifier Simulation Program

ASP is a functional transistor amplifier simulation program developed by SW.I.F.T. Enterprises that includes design routines, performance data, matching circuits, auto 'Q', and a NF optimizing utility. ASP Version 3.0 operates on greater than DOS 2.11 and requires an EGA or better monitor. A math coprocessor is recommended and the price is $75.
SW.I.F.T. Enterprises
INFO/CARD #218

Schematic Entry System

Vutrax-II GES is an ECA-2/LCA-1 compatible schematic entry system available through Tatum Labs. It can be used as a stand-alone drawing package or as the drawing interface to circuit analysis. Vutrax-II GES runs on IBM or compatible computers with Hercules, CGA, EGA, or VGA monitors. It requires a hard disk with 4M available. The price is $495 and it will interface with dot matrix and laser printers.
Tatum Labs, Inc.
INFO/CARD #217

Optimizing Circuit Simulator

Meta-Software has released an enhanced version of its optimizing analog circuit simulator, HSPICE™. Version H9007 features an enhanced transmission line model with an improved field solver for board/hybrid and LSI applications, a lossy first order skin effect model, and S-parameter calculations. It is available on Sun, MIPS, DEC, and Apollo workstations.
Meta-Software, Inc.
INFO/CARD #216

DSP Code Generation Software Upgrade

Burr-Brown has released an upgraded version of DSPlay XL™. The software allows users to quickly develop DSP code for AT&T's WER DSP32 and DSP32C processors through the creation of simple block diagrams. The upgrade, 3.16, expands the available DSP and related functions to over 100 from the program's original 60. DSPlay XL is priced at $1495.
Burr-Brown Corporation
INFO/CARD #215

Waveform Creation and Modification Interface

The SM 5000 WaveCAD software from John Fluke Mfg. provides a mouse-driven interface for the creation or modification of waveforms. It allows users to sketch a waveform freehand or point-to-point. SM 5000 WaveCAD software is priced at $1095 and is available on a four week delivery schedule.
John Fluke Mfg. Company, Inc.
INFO/CARD #214

IC Design Libraries

EEsof's new SMART (Simulatable Micro-wave ARTwork) libraries are used to translate a design developed from schematic circuit simulation into final physical layout. Each SMART library includes schematic symbols, simulation models, and layout artwork representations most often used in MMIC design. The libraries are priced at $5,000 each.
EEsof, Inc.
INFO/CARD #213

Microstrip Analysis Software

Sonnet Software introduces version 2.1 of Em, an electromagnetic analysis package. This software evaluates any discontinuity, validates large portions of entire designs, and enhances existing design sofware. Em offers a matrix solution speed typically six times faster than version 2.0.
Sonnet Software
INFO/CARD #212

RF literature

GaAs FETs and MMICs

Harris Microwave has released a data book titled GaAs FETs, MMICs & Foundry Service. The data book gives specifications and qualifications of Harris Microwave's GaAs FETs and MMICs and also includes application notes on these items.
Harris Microwave Corporation
INFO/CARD #230

DSP Products

Burr-Brown has announced the availability of a brochure that describes the company's complete line of digital signal processing products. Described are software, DSP processors for both PC and VME platforms, and analog I/O boards and systems.
Burr-Brown Corporation
INFO/CARD #229

Quartz Products Catalog

Tele Quarz has introduced a short form catalog covering their quartz crystals, oscillators, filters, and discriminators. The catalog contains crystal units between 1 and 300 MHz. Fundamental crystals up to 60 MHz and 3rd through 9th overtone crystals up to 300 MHz are described.
Tele Quarz Group
INFO/CARD #227

Mixers and Frequency Doublers Application Note

An application note which discusses mixers and frequency doublers has been published by FEI Microwave. The note includes a section on mixer terminology and mixer applications and a section on frequency doublers and their applications. Also included are product summaries of FEI Microwave's mixers and frequency doublers.
FEI Microwave, Inc.
INFO/CARD #228

IF/RF Solid State Devices

Daico Industries has announced its 1991 catalog, which features Daico's complete line of GaAs, PIN diode, Schottky diode, and relay devices. Key features, typical performance, operating characteristics, and block diagrams are included.
Daico Industries, Inc.
INFO/CARD #226

Optics Control Techniques

A booklet providing information on five optics control techniques has been released by Anritsu. The booklet describes high-speed modulation of lightwaves, control of optical waves, optical effects, development of a hyper-coherent optical sweep generator,

RF engineering opportunities

Advertising Index

phase-conjugate optics, and semiconductor quantum effect devices.
Anritsu America, Inc.
INFO/CARD #225

Quartz Crystal Oscillators

Piezo Crystal has announced their new product brochure covering quartz crystal oscillators for commercial, military, and GPS industries. It includes technical information on part numbers 2900082 and 2890080, Piezo's newest crystal oscillator products.
Piezo Crystal Company
INFO/CARD #224

Filter Catalog

Catalog Fastrap/90 from Microwave Filters Company defines and explains positive and negative trapping in cable systems, suggests strategic rules and illustrates a wide range of channel notch filters and tiering filters.
Microwave Filter Company, Inc.
INFO/CARD #223

Surface Mounted Components Catalog

Murata Erie North America has released a catalog that includes specifications and application information on surface mount capacitors, potentiometers, inductors, EMI/RFI filters, and ceramic filters and resonators.
Murata Erie North America
INFO/CARD #222

Phase Locked Sources

This catalog covers the XDMP series phase locked sources which utilize an internal crystal oscillator in the 100 MHz region which is locked to an external 5 or 10 MHz reference. These sources are for applications where high spectral purity and extremely low FM noise characteristics must be determined by a low noise crystal oscillator.
Communication Techniques, Inc.
INFO/CARD #221

Materials Measurement Package

A four page data sheet from Wiltron describes their materials measurement package which measures permittivity and permeability of a solid, powder, or liquid material in coaxial or waveguide holders at microwave frequencies. Also presented is an overview of the measurement method, sample preparation, and transmission line fixturing.
Wiltron Company
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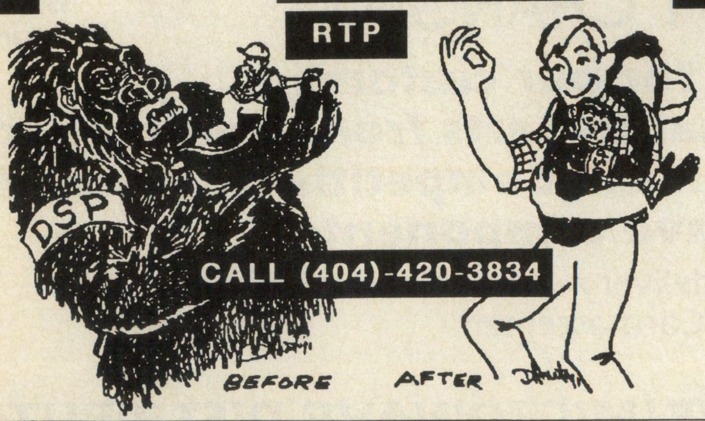

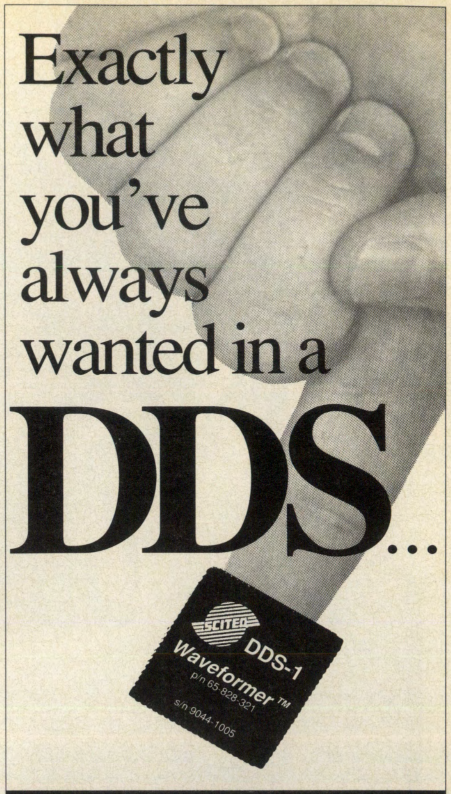

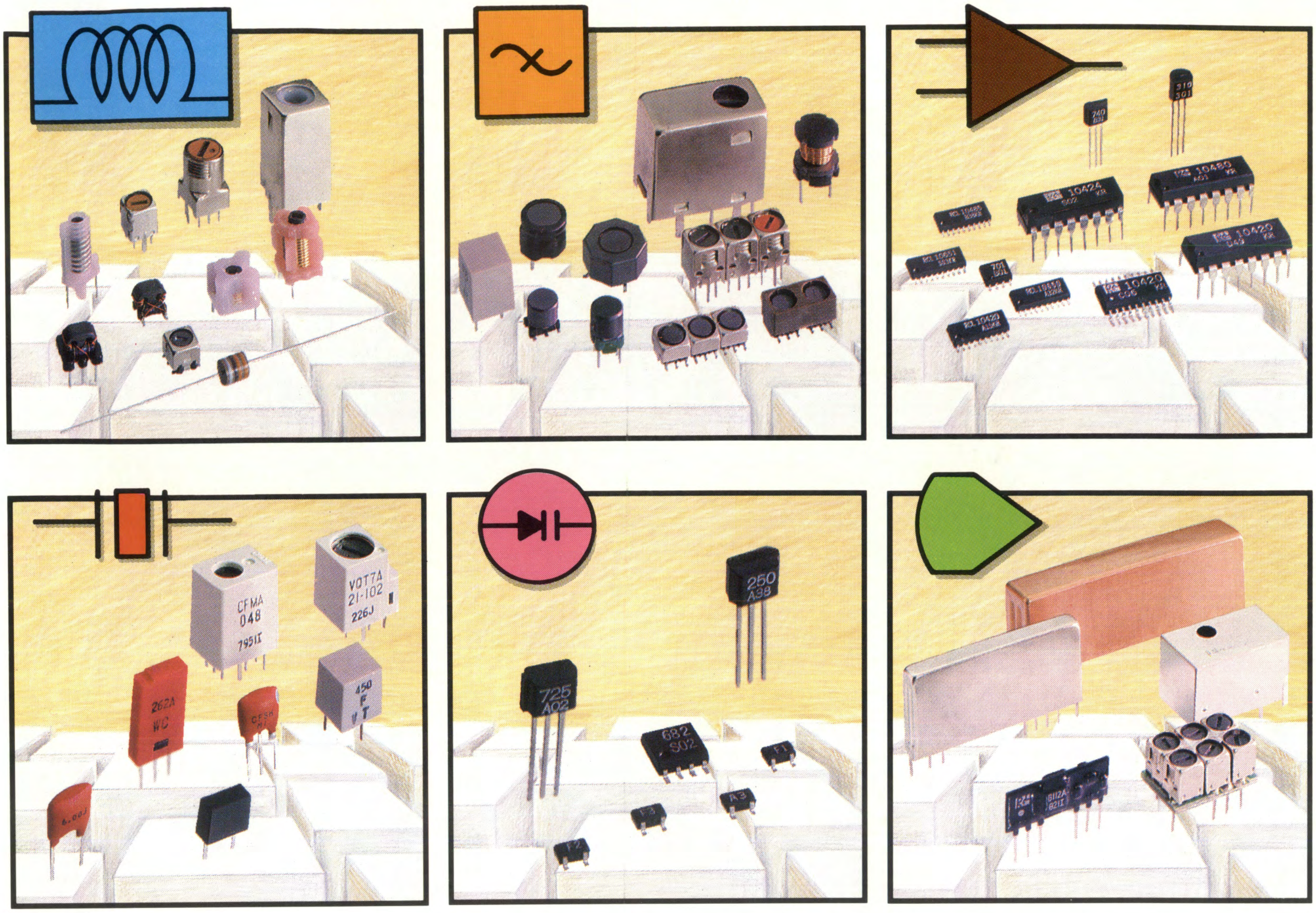

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For I-F amplification and detection, Toko offers a variety of ICs to simplify design and reduce product size. Other available types include: compandors for noise reduction, equalizers, analog switches and voltage regulators.

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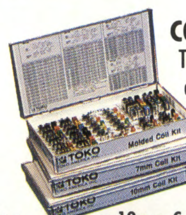
The fastest way to design and produce your next RF design project is with Toko RF modules. Utilizing the latest hybrid circuit technology, these circuits will also save space and cost. Custom variations are also available.



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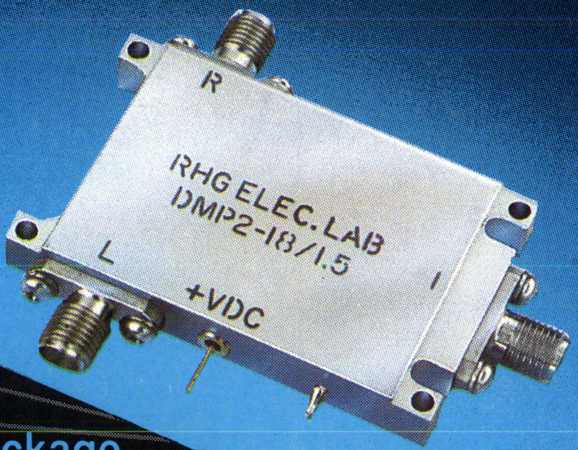
How Low Can It Go?

New Mixer Preamp Provides Ultra-Low Intermodulation Levels For Ultra-Wideband RF Front Ends



- Hermetic drop-in package
- 2-18 GHz RF/LO, 1.5 GHz IF
- Low noise, high ½ IF rejection

OPERATING SPECIFICATIONS, MODEL DMP2-18:

INPUTS:

Frequency, RF/LO (GHz)	2-18
VSWR, RF/LO	2.5:1/2.0:1
Peak power (1 μsec pulse) (W)	150
Power, LO (dBm)	+13
Power, DC (Volts/mA)	+12/75

OUTPUTS:

Frequency, IF (GHz)	1.2-1.8
VSWR, IF	2.5:1
Intermodulation (dBm):	
2nd/3rd order	+20/+30
Power, IF @ 1 dB comp. (dBm)	+12

TRANSFER CHARACTERISTICS:

Conversion gain, min. (dB)	15.5
Isolation, min. (dB)	25
Noise figure (dB)	10
Single tone intermodulation with −10 dBm RF input (dBc):	
1LO-2RF	50
1LO-3RF	60
2LO-1RF	25
2LO-2RF	50
2LO-3RF	60
3LO-2RF	65
3LO-3RF	70

ENVIRONMENTAL: Operating temperature −55°C to +85°C; hermetically sealed package; meets applicable requirements of MIL-E-5400.

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