

# rf design

engineering principles and practices

May 1990

A New Weapon for Design  
1 Watt, 1 GHz TO-3



industry insight  
**Low Noise Transistors**

featured technology  
**Frequency Synthesis**





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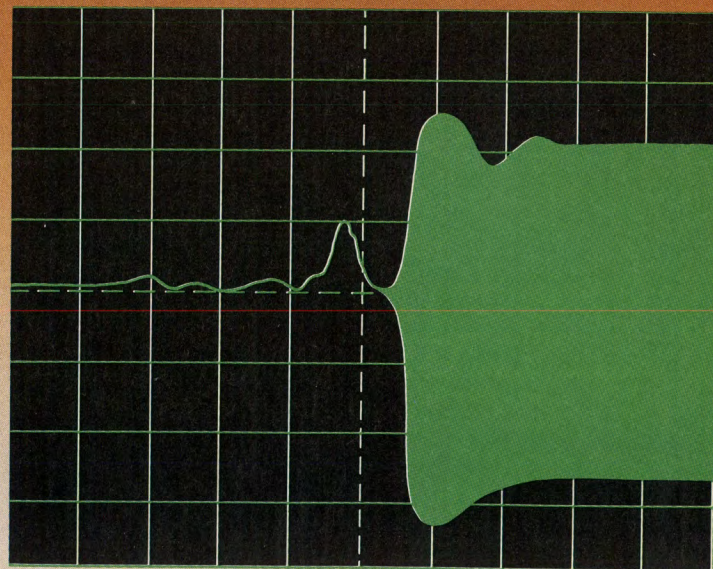
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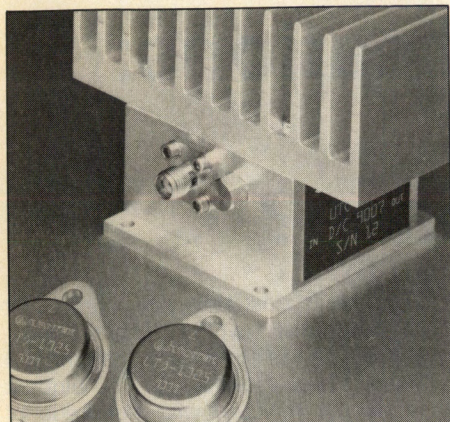


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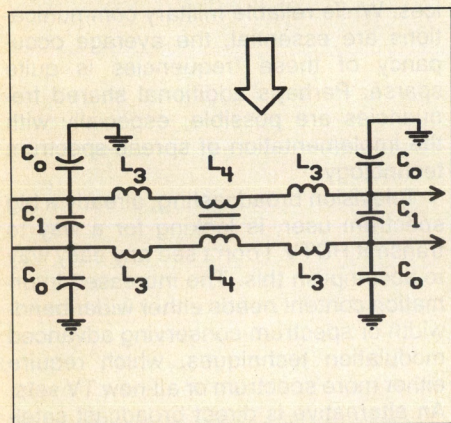
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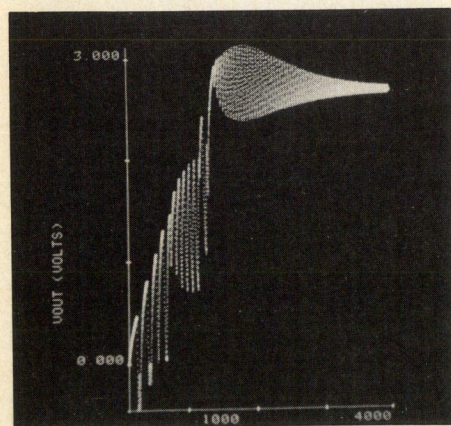




Page 32 — 1-Watt Amplifier



Page 39 — Wideband Phase Shifter



Page 57 — Digital Phase Detector

## industry insight

### 31 The Low Noise Transistor Market

A look at the customer application and technology factors which influence the marketplace for GaAs and silicon low noise transistors. — Gary A. Breed

## cover story

### 32 A 1-Watt Amplifier for 20-1000 MHz

Avantek announces a new 1-watt linear amplifier for 20 MHz - 1 GHz, packaged in a TO-3 case. The amplifier is intended to fill a gap in available components for medium power communications and instrumentation applications.

— Jim Gustafson and Melanie Roberts

## featured technology

### 41 A 1.8 GHz Odd-Ratio Frequency Divider

The author describes a technique for building an odd-ratio frequency divider. This digital IC circuit can divide by non-2<sup>n</sup> ratios at input frequencies above 1 GHz, has many desirable qualities and no known caveats; consequently, it is ideally suited for RF frequency synthesizer applications.

— David Korn and Carl Deierling

### 57 Modeling Discrete-Time Phase Detection in the PLL

The phase locked loop is a feedback control system that is widely applied in the design of electronic communications and controls apparatus. This article examines the classical phase error response forms for the sample phase and frequency steps as well as contrasting transient response and design complications imposed by the digital phase detector.

— C.W. Price

## rf design awards

### 39 A Wideband 70 MHz Phase Shifter

This circuit was an entry in the 1989 Design Awards Contest. While the concept of the phase shifter with a quadrature coupler is well known, this technique improves upon it with better performance and lower costs.

— Francois Methot

## rfi/emc corner

### 64 A Review of Common RFI Sources

RF designers must consider the electromagnetic environment in which their products must operate. Here is a brief review of potential RFI sources in the home and office.

— Gary A. Breed

### 65 Complex Signals: Part IV

The conclusion of this four part tutorial series covers the many applications of complex signals, including AM, FM, SSB, VSB, and QAM systems.

— Noel Boutin

## departments

- 6 Editorial
- 11 Letters
- 15 Calendar
- 16 Courses
- 22 News

- 76 New Products
- 83 New Software
- 84 New Literature
- 85 Advertisers Index
- 87 Info/Card

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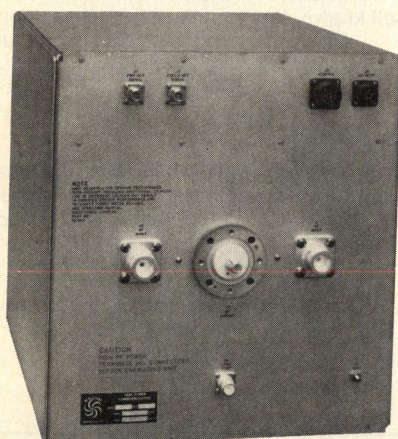
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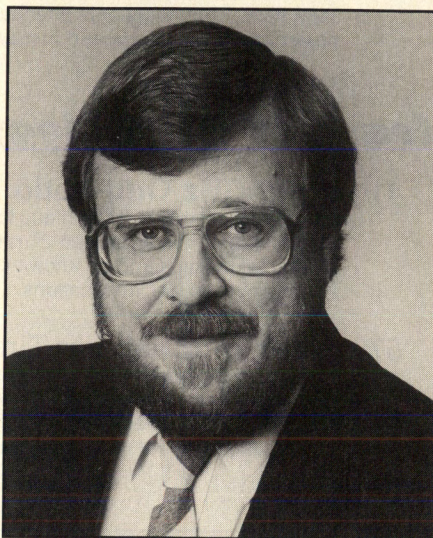
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INFO/CARD 3

## rf editorial

# New RF Applications: Where is the Spectrum Space?



By Gary A. Breed  
Editor

*Wireless everything.*

This is the road to the future of RF technology. Cordless phones, cellular radio/telephones, home security, wireless audio and video distribution, toys, cordless computer peripherals, and data links for anything imaginable. It is an importer's dream, an EMC engineer's nightmare, and soon, the biggest frustration the Federal Communications Commission has seen in a long time.

Add a couple more things to the list: greater bandwidth for high-definition TV, wireless office networks, and still-growing use of mobile business communications. Private and public demand for pieces of the radio spectrum continues to pick up momentum. The futuristic concepts of twenty years ago are becoming reality, and we need to find a place to put them all.

What are the options? Can we really accommodate so many new radio users? Who gets the prime RF real estate and who gets the slums? The answers to these questions are just a guess right now, but, at least for the sake of discussion, there are some obvious spectrum uses that can be re-examined.

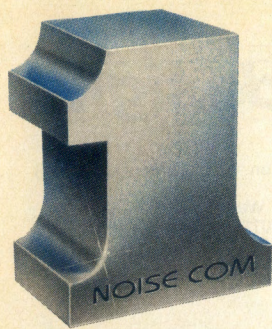
The most desirable spectrum is at VHF and UHF, where there is well-behaved line-of-sight propagation and mature device technology. The biggest user is the military, with hundreds of MHz set aside for their exclusive use, plus frequencies shared with other services. While reliable military communications are essential, the average occupancy of these frequencies is quite sparse. Perhaps additional shared frequencies are possible, especially with the implementation of spread-spectrum technology.

Television broadcasting, already a big spectrum user, is looking for a way to transmit HDTV. I don't see any easy way to accomplish this. The increased information content needs either wider bandwidth or spectrum-conserving advanced modulation techniques, which require either more spectrum or all-new TV sets. An alternative is direct broadcast satellite (DBS) transmission at microwave frequencies; or an expansion of wired communications (coaxial cable or fiber optics).

Some shuffling of a few MHz here and there might keep things going for a few years, but it won't be long until more traumatic measures are needed. Unless an unforeseen technical breakthrough occurs to allow cheap microwave radios or a truly practical bandwidth-reduction method, the accumulated demand for spectrum will force drastic action by the FCC or Congress. Given the government's unwillingness to *seriously* deal with the difficult issues of the deficit, trade imbalance, and the environment, I'm skeptical of their ability to develop a far-sighted frequency allocation solution.

It is up to us, the RF industry, to see that our own future is secure. Our industry has the expertise and experience to be the leaders in proper stewardship of a limited natural resource: the electromagnetic spectrum. Serious discussion must begin outside the government bureaucracy. Can the EIA, AEA, IEEE or some other group provide the forum? Let me (or them) know what you think.





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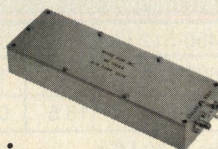
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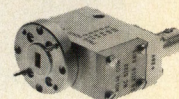
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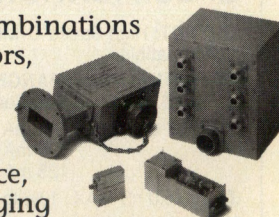


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- Storage temperature -65 to +125° C
- Maximum input power +28 VDC at 20 mA (+15 VDC available)

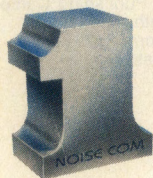
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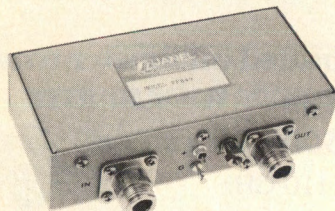
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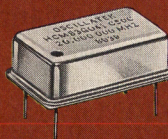
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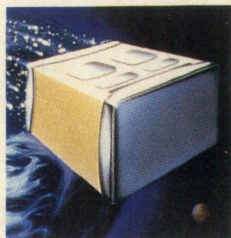
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| 0.5             |      | 4.7             | C    | 27              |      |
| 1.0             |      | 5.6             |      | 33              | J    |
| 1.2             |      | 6.8             |      | 39              |      |
| 1.5             | C    | 8.2             |      | 47              |      |
| 1.8             |      | 10              | J    | 56              |      |
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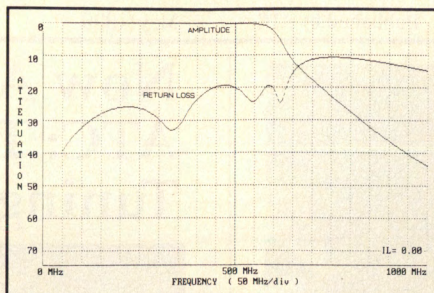
## Terminating Complementary Filters

Editor:

After reviewing the article in the February 1990 issue by David Festing regarding terminating complementary filters, I would like to add the following information.

Although the results obtained were sufficient to improve the harmonic rejection of the amplifier and were somewhat empirically derived, there is an exact procedure to design a complementary low pass/high pass filter.

Classically, for the return loss to be perfect at all frequencies in a complementary filter, we would choose a Butterworth, zero source, low pass prototype. Scale the low pass for frequency and impedance, then derive the high pass components by taking the reciprocal of the normalized low pass coefficients of the same prototype and scale it to the same 3 dB bandwidth and impedance. Connect the two in parallel at the input.



**Figure 1. Response of original filter.**

Figure 1 shows the computed response of the filter in the featured article. Note that the return loss in the stopband is as low as 11 dB. A 6-pole Butterworth design would have a return loss better than 60 dB for any frequency. The reject attenuation of 27 dB at 1 GHz is not as good as the featured article, however.

Generally, a return loss of about 20 dB is sufficient for most practical applications and we can sacrifice the classical design for a more useful approximation. A 0.5 dB Chebyshev, zero source, prototype has return loss degraded to 19 dB (still acceptable) but the attenuation at 1 GHz improves from 27 to 44 dB. At most in-band frequencies, the

return loss is better than 25 dB.

In carrying this approximation even further, a 6-pole elliptic complementary with a design reflection coefficient of 15 percent (0.098 dB) is used. Because elliptic filters are usually normalized to the ripple bandwidth and the complementary design requires the 3 dB bandwidth prototype, you must first scale the prototype so that the normalized bandwidth is 3.01 dB at one radian. Calculated response of this elliptic complementary filter shows better than 27 dB return loss and the 0.098 dB amplitude ripple (16.5 dB) is not reflected in the return loss.

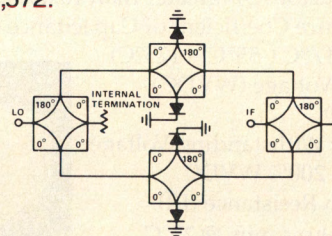
From the promising results of the above filters, and with a computer optimizing program, we have designed and built the 30 MHz low pass shown in Figures 2 and 3. This filter was used in a feed forward 3 kW CW HF transmitter to provide a good termination at all frequencies for stable operation. The low pass portion is a 10 pole elliptic and the high pass is almost a 7 pole elliptic. It is imperative that the reactive frequency traps be put at the crucial frequencies in this design. If the low pass has an attenuation peak at 40

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| MD-160    | 1-1500                         | 1-1000   | 6.5                       | 28                    | 28    | +15                         | +13             | 276                    |
| MD-174    | 1-2800                         | 1-2000   | 6.5                       | 32                    | 38    | +16                         | +20             | 284                    |
| MDC-174   | 1-2800                         | 1-2000   | 6.5                       | 32                    | 38    | +16                         | +20             | 284                    |
| MD-169    | 1-3500                         | 5-1500   | 6.5                       | 30                    | 30    | +7                          | +10             | 282                    |
| MDC-169   | 1-3500                         | 5-1500   | 6.5                       | 30                    | 30    | +7                          | +10             | 282                    |
| MD-179    | 1-4000                         | 5-1500   | 6.5                       | 40                    | 40    | +5                          | +7              | 287                    |
| MDC-179   | 1-4000                         | 5-1500   | 6.5                       | 40                    | 40    | +5                          | +7              | 287                    |
| MD-163    | 1000-6000                      | 10-2000  | 7.0                       | 20                    | 25    | -2                          | 0               | 279                    |
| MDC-163   | 1000-6000                      | 10-2000  | 7.0                       | 20                    | 25    | -2                          | 0               | 279                    |
| MD-162    | 1000-7000                      | 10-2000  | 6.0                       | 25                    | 20    | +8                          | +13             | 278                    |
| MDC-162   | 1000-7000                      | 10-2000  | 6.0                       | 25                    | 20    | +8                          | +13             | 278                    |
| MD-165    | 500-8000                       | 2-2000   | 9.0                       | 25                    | 24    | -2                          | 0               | 281                    |
| MDC-165   | 500-8000                       | 2-2000   | 9.0                       | 25                    | 24    | -2                          | 0               | 281                    |
| MD-164    | 500-9000                       | 10-2000  | 6.5                       | 22                    | 27    | +8                          | +13             | 280                    |
| MDC-164   | 500-9000                       | 10-2000  | 6.5                       | 22                    | 27    | +8                          | +13             | 280                    |

\*\*Specs per Adams-Russell Components Group 1989 RF & Microwave Signal Processing Components catalog.

\*MD-160 5-9 units (higher quantity pricing available).

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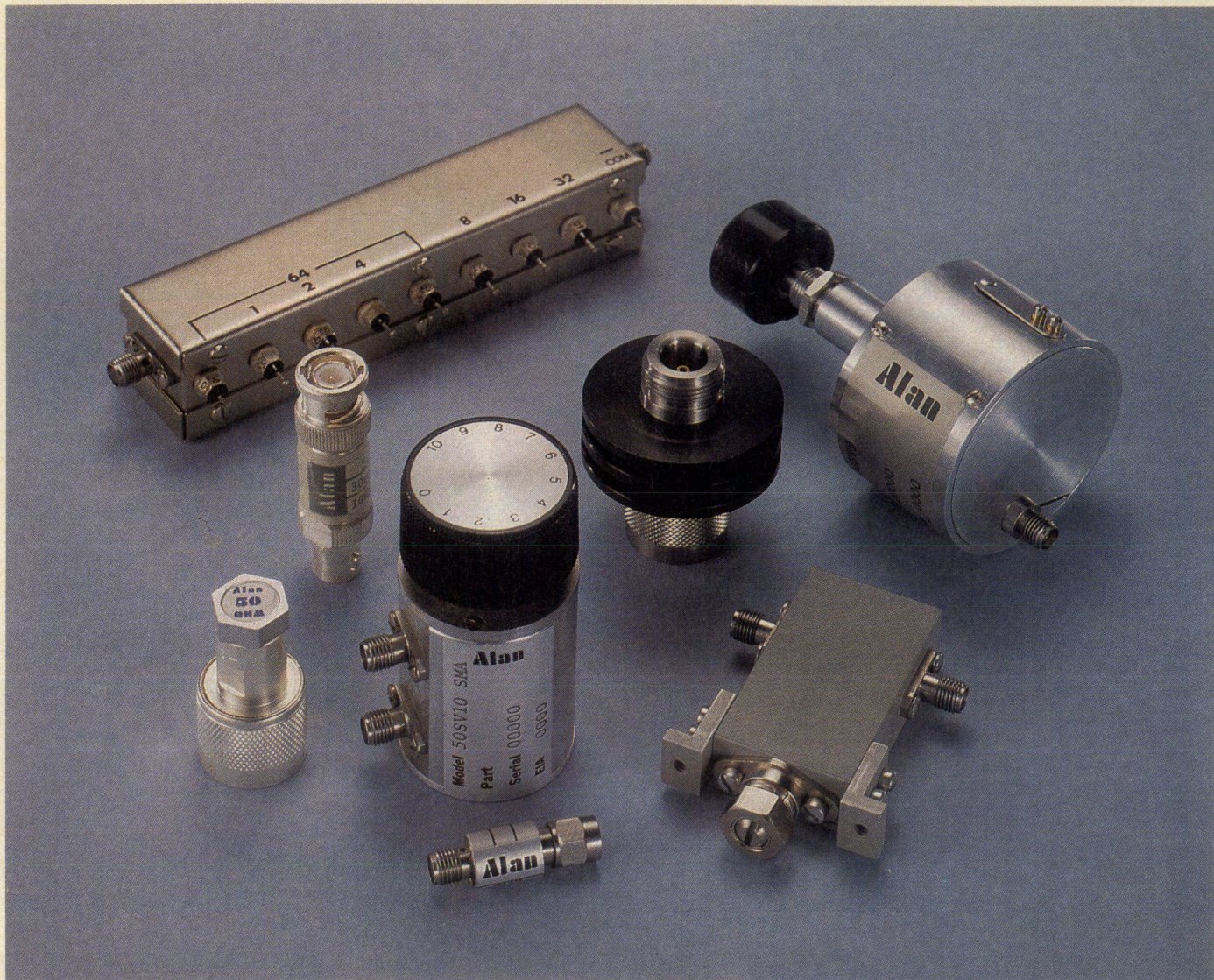
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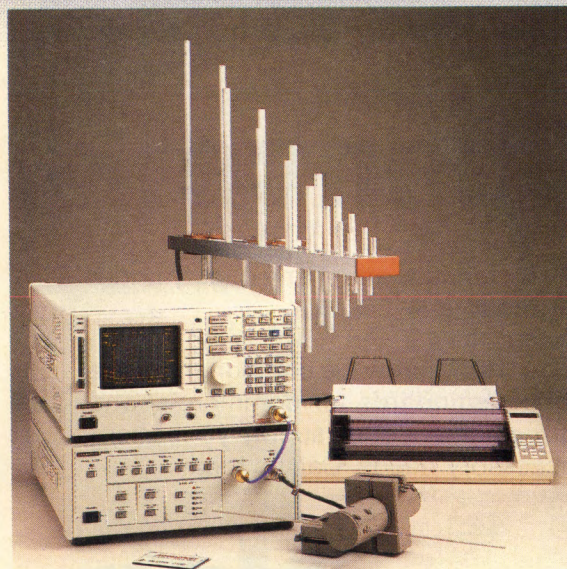


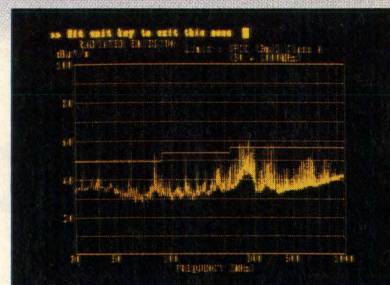
Photo: R2542B EMI Receiver System, connected to R3261/3361 Series Spectrum Analyzer with R3551 Preselector Antenna, and Plotter.

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

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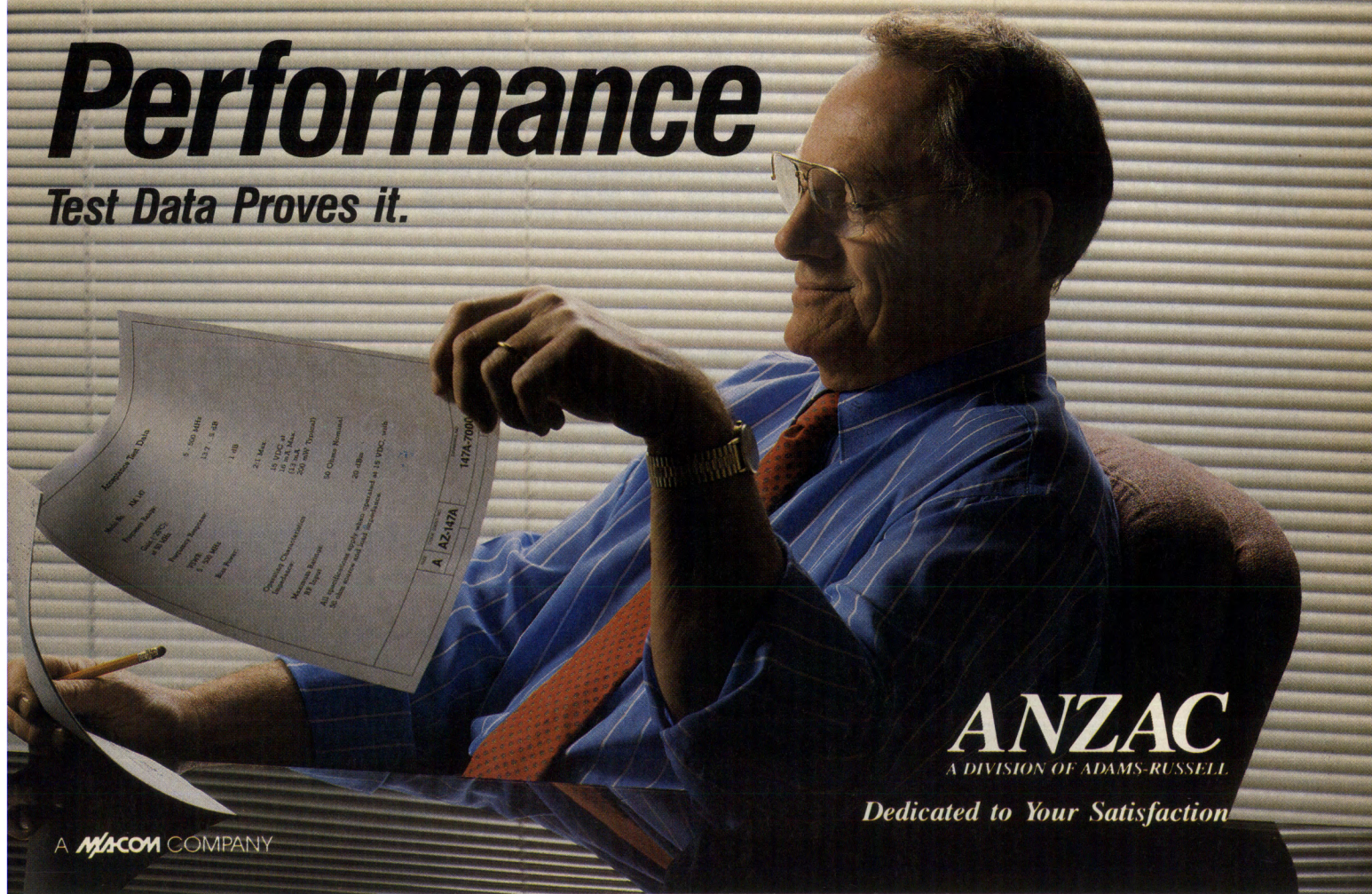
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MHz, the high pass must have a zero of attenuation at 30 MHz for the optimization to have a good VSWR in the transition region. The location of these traps in the schematic is also important for a high power application. Although the input RF current may be small, there are large internal circulating currents that can be as much as ten times the input current near the crossover frequency.

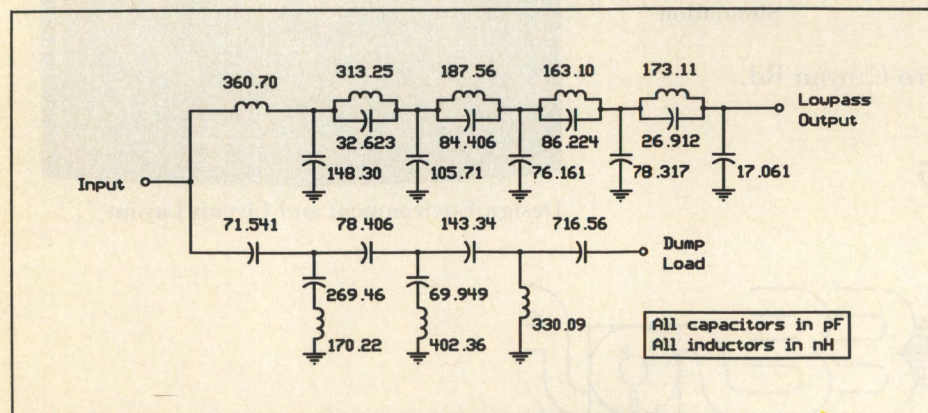
The insertion loss of this complementary quasi-elliptic filter at 30 MHz was less than 0.5 dB and the rejection above 40 MHz was greater than 60 dB. Note

that the worst calculated return loss is 29 dB; when the filter was tweaked up on the bench we achieved 30 dB.

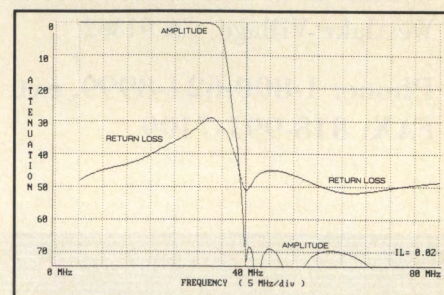
Why must the zero source prototypes be used? The premise of the complementary high pass/low pass filter is for the input to present a constant resistance at all frequencies. Because of this, when driven by any source impedance, the voltage on the input of the filter will remain constant for any frequency. When a particular amplitude response is desired, you must choose a prototype that has the input voltage constant for any frequency. The only way that this

constant voltage can occur and get the same desired response is for the prototype to have a zero source impedance. All of these zero source prototypes start out with an input series element. A shunt starting element would be shorted out by the zero source and have no effect on the response. The last element (at the load) for all of these zero source prototypes usually has a value that looks insignificant but it is necessary to keep it in the circuit to get the n-pole response.

Roy G. Anderson  
Westinghouse Electronic Systems  
Group  
Baltimore, MD



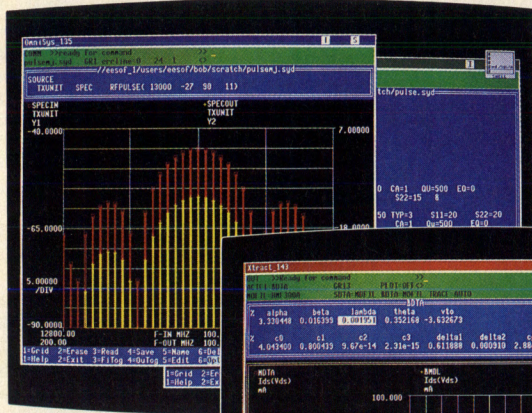
**Figure 2. High performance complementary filter**



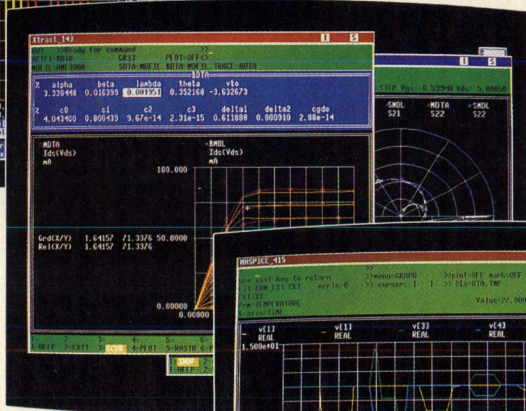
**Figure 3. Response of 30 MHz low-pass filter.**



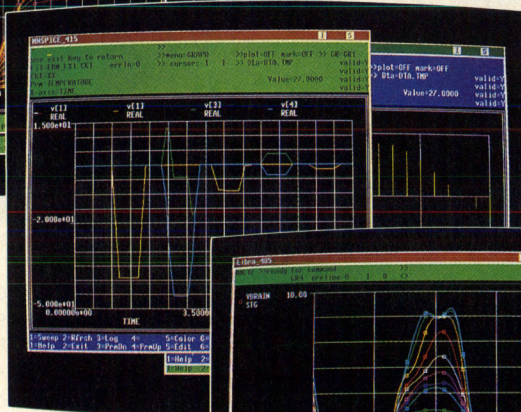
# Analog CAE Solutions— from picoseconds to GHz



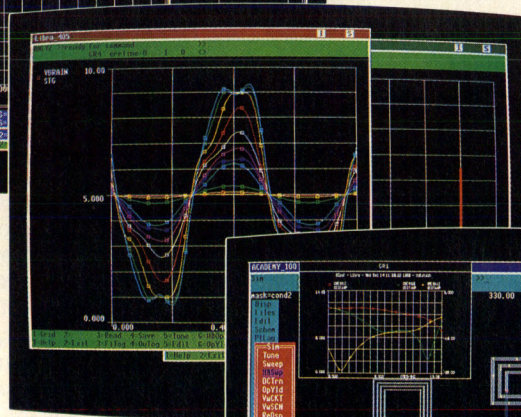
**System  
Simulation**



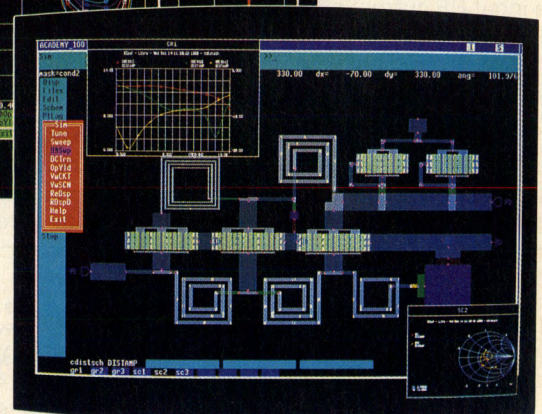
**Device  
Characterization**



**Time-Domain  
Simulation**



**Nonlinear  
Simulation**



**Design Environment and Circuit Layout**

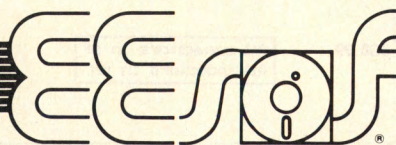
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## rf calendar

**May 23-25, 1990**

**44th Annual Frequency Control Symposium**

Stouffer Harborplace Hotel, Baltimore, MD

Information: Dr. R.L. Filler, U.S. Army Electronics Technology and Devices Laboratory, Attention: SLCET-EQ, Fort Monmouth, NJ 07703-5000. Tel: (201) 544-2467

**June 5-7, 1990**

**ATE & Instrumentation Conference East**

World Trade Center, Boston, MA

Information: MG Expositions Group, 1050 Commonwealth Ave., Boston, MA 02215. Tel: (800) 223-7126

**June 5-7, 1990**

**AFCEA (Armed Forces Communication and Electronics Association)**

Washington D.C. Convention Center, Washington, DC

Information: AFCEA, 4400 Fair Lakes Court, Fairfax, VA 22033. Tel: (800) 336-4583, ext. 6189; (703) 631-6189

**June 11-14, 1990**

**Conference on Precision Electromagnetic Measurements (CPEM '90)**

Westin Hotel, Ottawa, Ontario, Canada

Information: CPEM '90, National Research Council Canada, Building M19, Montreal Road, Ottawa, ON, Canada K1A 0R6.

**July 26-29, 1990**

**The 1990 International Tesla Symposium**

Hilton Inn, Colorado Springs, CO

Information: International Tesla Society, 330-A West Uintah Street, Suite 215, Colorado Springs, CO 80905-1095.

**August 21-23, 1990**

**IEEE EMC 90 Symposium**

Washington Hilton Hotel, Washington, DC

Information: Joe Fisher Tel: (703) 521-6336

**August 28-30, 1990**

**Surface Mount '90**

Bayside Exposition Center, Boston, MA

Information: MG Expositions Group, 1050 Commonwealth Ave., Boston, MA 02215. Tel: (800) 223-7126 or (617) 232-3976

**August 28-September 5, 1990**

**XXIII General Assembly of the International Union of Radio Science (URSI)**

Prague, Czechoslovakia

Information: Prof. V. Zima, Institute of Radioengineering and Electronics, Czechoslovak Academy of Sciences, 182 51 Praha 8, Czechoslovakia.

**September 23-27, 1990**

**Association of Old Crows Meeting**

Hynes Convention Center, Boston, MA

Information: Association of Old Crows, 1000 N. Payne Street, Alexandria, VA 22314. Tel: (703) 549-1600

**September 25-27, 1990**

**Piezoelectric Devices Conference**

Kansas City Westin Crown Center, Kansas City, MO

Information: EIA Components Group, 1722 Eye Street, N.W., Suite 300, Washington, DC 20006.

Tel: (202) 457-4980



## University of Colorado at Boulder

### Computer Aided Design of Microstrip Antennas and Arrays

June 26-29, 1990, Boulder, Colorado

Information: Office of Conference Services, University of Colorado, Campus Box 454, Boulder, Colorado 80310. Tel: (303) 492-5151

## The George Washington University

### Ionospheric Radio Propagation: Principles and Applications

May 29-June 1, 1990, Washington, DC

### Electronic Countermeasures

June 4-8, 1990, Washington, DC

### An Introduction to Numerical Techniques in Electromagnetics

June 5-8, 1990, Washington, DC

### Spread-Spectrum Communications Systems

June 11-15, 1990, Colorado Springs, CO

### Satellite Communications Engineering Principles

June 13-15, 1990, Washington, DC

November 19-21, 1990, Washington, DC

### New HF Communications Technology: Advanced Techniques

June 18-22, 1990, Washington, DC

December 3-7, 1990, San Diego, CA

December 17-21, 1990, Washington, CA

### ELINT: Analyzing Radar Signals

June 19-21, 1990, Washington, DC

### Global Positioning System: Principles and Practice

July 16-18, 1990, Washington, DC

### Antennas: Radiation and Scattering

August 27-28, 1990, Washington, DC

### Grounding, Bonding, Shielding and Transient Protection

August 27-30, 1990, Chicago, IL

November 12-15, 1990, Orlando, FL

### Microwave Radio Systems

September 6-7, 1990, Washington, DC

### Modern Receiver Design

September 10-14, 1990, Washington, DC

### Broadband Communications Systems

September 10-14, 1990, Washington, DC

### Radar Operation and Design: The Fundamentals

September 18-21, 1990, Washington, DC

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

## Hewlett-Packard Co.

### Designing for Electromagnetic Compatibility (EMC)

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May 24-25, 1990, Huntsville, AL

June 5-6, 1990, Seattle, WA

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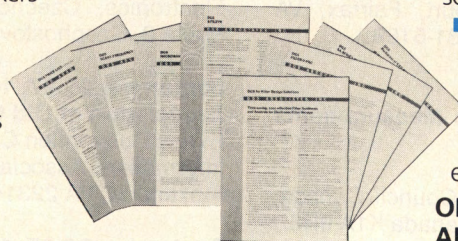
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## **Learning Tree International**

### **Digital Signal Processing: Techniques and Applications**

June 5-8, 1990, Toronto, Ontario, Canada  
June 19-22, 1990, San Francisco, CA  
June 26-29, 1990, Washington, DC

### **Introduction to Telecommunications**

June 5-8, 1990, Los Angeles, CA  
July 10-13, 1990, Washington, DC  
July 17-20, 1990, San Francisco, CA

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

## **Fluke-Philips**

### **Principles of Analog Oscilloscopes**

August 21-22, 1990, Los Angeles, CA

### **Principles of Digital Oscilloscopes**

May 17-18, 1990, Atlanta, GA  
August 23-24, 1990, Los Angeles, CA

Information: John Fluke Mfg. Co., Inc., PO Box 9090, Everett, WA 98206. Tel: 1-800-443-5853 ext. 73. In Canada: (416) 890-7600

## **CEI-Europe/Elsevier**

### **RF and Microwave Circuit Design I: Linear Circuits**

June 18-22, 1990, Cambridge, UK  
July 9-13, 1990, Los Angeles, CA

### **RF and Microwave Circuit Design II: Non-Linear Circuits**

June 18-22, 1990, Cambridge, UK  
July 16-20, 1990, Los Angeles, CA

### **Surface Acoustic Wave Devices for Signal Processing**

June 18-22, 1990, Cambridge, UK

Information: Mrs. Tina Persson, CIE-Europe/Elsevier, Box 910, S-612 01 Finspong, Sweden. Tel: +46 (0) 122-17570. Fax: +46(0)122-14347

## **Amador Corporation**

### **EMC for Digital Designers**

October 23, 1990, Bloomington, Minnesota

### **ESD Immunity for Electronic Equipment**

October 24, 1990, Bloomington, Minnesota

Information: Diane Swenson, Tel: (612) 465-3911

## **Henry Ott Consultants**

### **Electromagnetic Compatibility Engineering**

June 13-15, 1990, East Hanover, New Jersey

Information: Henry Ott Consultants, 48 Baker Road, Livingston, NJ 07039. Tel: (201) 992-1793.

## **Analog Devices**

### **DSP Workshop**

June 6-8, 1990, Massachusetts  
June 12-14, 1990, California

Information: Paula Turco, Analog Devices, Inc., One Technology Way, PO Box 9106, Norwood, MA 02062-9106. Tel: (617) 461-3306.



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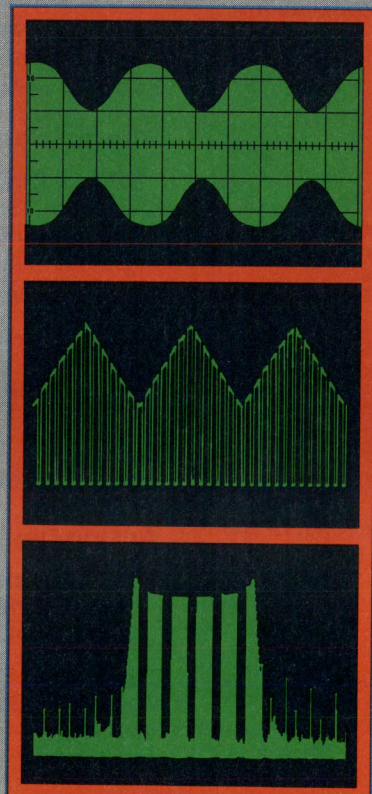
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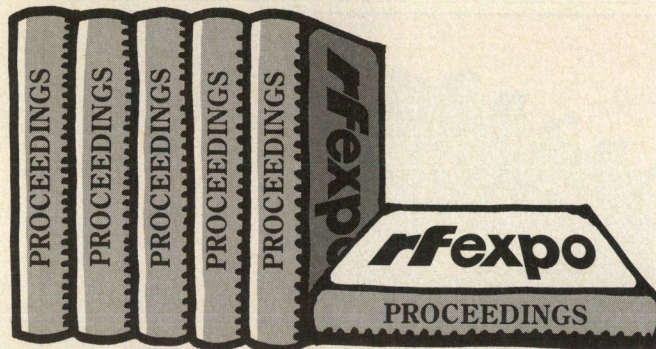
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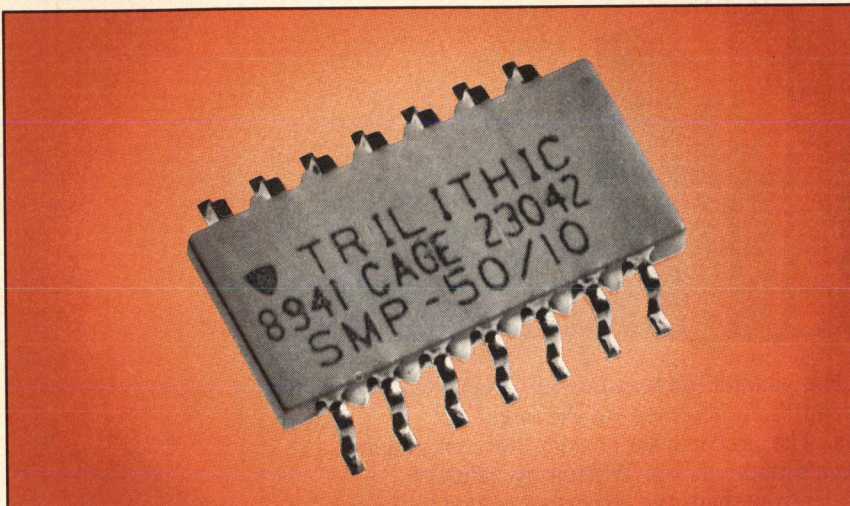
## RF Technology Expo

RF Technology Expo 1990 at the Anaheim Convention Center in Anaheim, California offered something for every engineer this year in both technical programs and exhibits. The technical program offered everything from basic tutorials to highly technical and innovative presentations. On the exhibit floor, exhibitors and attendees

alike were pleased with the quality as well as quantity of exhibits. Attendance matched the last Anaheim RF Expo. A new event this year, a special reception for ham radio operators, also proved to be a big hit and some great door prizes were given away.

Once again the technical program received rave reviews from engineers. Technical chairmen Brian Rose and Ted Dudziak did an excellent job of organiz-

ing the technical program which showed in the quality and diversity of the papers presented. The four special courses that were offered this year received good marks. Fundamentals of Circuit Design — Parts I and II taught by Les Besser and Computer Aided Filter Design and, a new offering this year, Oscillator Design Principles taught by Randy Rhea gave engineers a chance to brush up on their basics or learn something new. All sessions were well attended, but there were a few that quite literally packed them in. All three Receiver Design sessions attracted large crowds of engineers. In particular, Sherman Vincent's tutorial on Receiver Design attracted more people than anticipated and the tutorial got excellent reviews. The session on Phase Locked Loops also drew a large number of engineers. The papers presented in the session were "A Phase Lock Loop in a Non-Cooperative Environment: CAD Verifica-

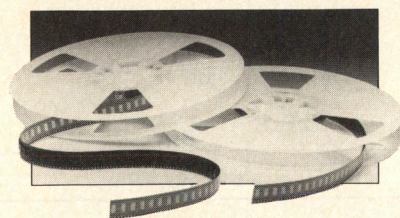


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RF Design's Bill Pettit, digs deep for the winner of a drawing sponsored by Alan Carr, Inc.

tion" by Michael Black and "A Reference-Cancelling Phase/Frequency Detector" by Dan Baker, the 1989 RF Design Awards Winner. Some of the comments from attendees on this session were "Dan Baker was excellent." and "Mr. Black always gives a good presentation. I enjoy his hands-on techniques and direct examples." Many of the other speakers received favorable comments as well: "Excellent speaker with a clear, intriguing presentation." and "Very good coverage of topic especially, the effect of phase noise on dynamic range," were some of the comments from attendees.

Even the papers on highly specialized topics drew a good number of people. Paul Beaty's paper "Introduction to the Maximum Entropy Method (MEM), Linear Prediction (LP), and Linear Predictive Coding (LPC)" and "Analog Integrated Circuit Design on a High Fre-



quency Bipolar Process" by Bob Sparkes, although both were narrow in focus, they were well attended and the questions generated showed a high level of interest.

In addition to the technical program, a marketing conference was also presented. The conference was coordinated by Barry Manz, President of Manz Communications and John McCrea, President of McCrea Management Consultants. The sessions dealt with relevant issues such as "Europe 1992 Redefined: The Impact on Business" and "Maquiladora: Manufacturing in Mexico". There was also a productivity improvement seminar sponsored by *Microwave Systems News* entitled "Productivity Oriented Design & Manufacturing: A Necessity for the 90's."

In addition to hearing about the latest in design and technology, attendees were able to get a good look at the latest products offered by more than 200



*One of the many, fine technical sessions in progress at RF Expo.*

exhibiting companies. One of the new products introduced at the show was Avantek's low noise, cascable silicon bipolar MMIC Amplifier. The INA-10386 has a cascable 50 ohm gain block, a 3 dB bandwidth, 26 dB typical gain at 1.5 GHz, 10 dBm typical P1dB at 1.5 GHz in a surface mount plastic package.

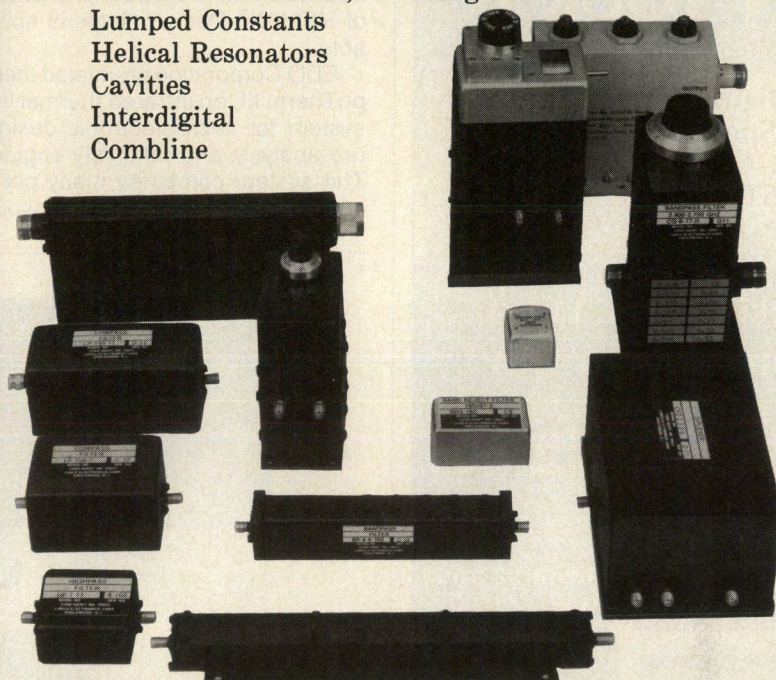
Philips Components-Signetics announced a comprehensive development system called the Cellular Chip Set Design Kit. The kit is aimed help reduce overall research and development costs for designers and manufacturer's using cellular radio technology for traditional mobile and portable telephone applications. Signetics also introduced a wide-band variable gain amplifier, the NE5209. It features an 800 MHz bandwidth, 50 ohm differential output, single 5V power supply, 0-30 dB gain, 7 dB noise figure and full ESD protection.

HELIAX<sup>®</sup>, Type FSJ1-50A, an improved coaxial cable from Andrew Corporation was introduced. It is a 1/4"

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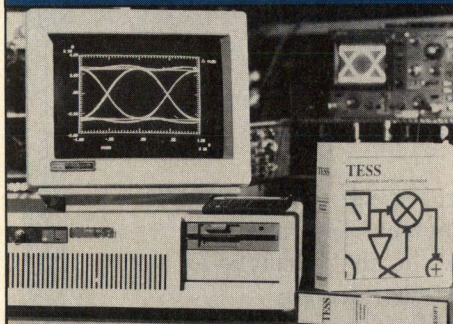
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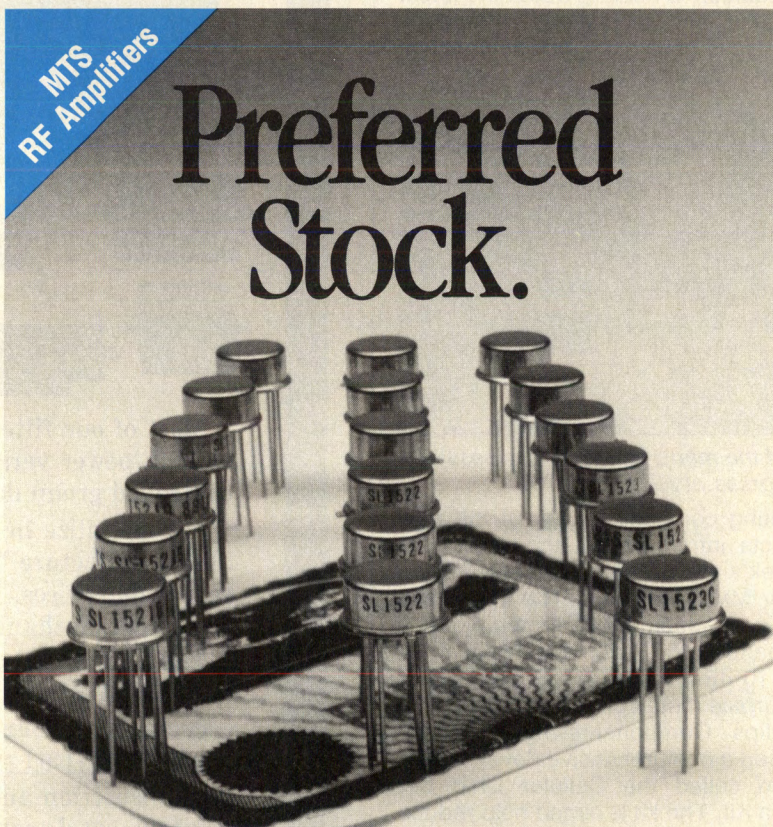
## rf news *Continued*

superflexible copper cable for high performance situations such as antenna and equipment room jumpers for land mobile radio, cellular and earth stations and RF test equipment. Some of its electrical characteristics are: maximum frequency 20.4 GHz, peak power rating 3.5 kW, and an inductance of 0.061 uH/ft. Andrew Corporation also announced that their quarterly publication of HELIAX<sup>®</sup> Design Notes is now available.

EDO Corporation premiered their CompuTherm III, an infrared thermal imaging system for microelectronic design, failure analysis and reliability engineering. The system combines many peripheral control devices into one console. The

system contains an 8 MIPS array processor and analog RGB RS-170 image display format, a high resolution monitor, a computer operating system display, an IEEE-488 interface card and software package.

RF Technology Expo 90 offered a diverse group of papers for the technical program as well as the latest in product and service offerings for everyone in the RF industry. We look forward to seeing everyone next year at RF Technology Expo, February 5-7, 1991 at the Santa Clara Convention Center, Santa Clara, California. In the meantime, don't forget RF Expo East, November 13-15, 1990 at Marriott's Orlando World Center, Orlando, Florida.



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### Frequency Stability:

$\pm 1 \times 10^{-7}$  in temp range

### Operating Temp. Range:

0° to +50°C  
(optional -20°C to +70°C)

### Aging Short Term Stability:

$8 \times 10^{-10}$  at 1 Sec

### Long Term Stability:

$< 1 \times 10^{-6}$ /year

### Warm Up:

$< 20$  seconds to  $\pm 1 \times 10^{-7}$

**Input Voltage:** 15 V  $\pm 5\%$   
5 V (TTL)

### Input Power:

$< 0.5$  W During Warm-up  
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**Size:** 1.26"  $\times$  1.26"  $\times$  0.7"

### Output Waveform:

TTL (optionally Sine)



## Ocxo

*Oven Controlled  
Crystal Oscillator*

### Frequency Range:

to 50 MHz

### Short Term Stabilities:

up to  $5 \times 10^{-12}$  (1 sec)

### Warm-Up Time:

As low as 1 min

### Temperature Stability:

$\pm 5 \times 10^{-10}$   
(0° to +50°C)

### Low Aging Rate:

$< 5 \times 10^{-11}$ /Day

### Low Noise:

$< -157$  dBc@  
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### Low Vibration

**Sensitivity:**

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### Temperature Range:

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*Complete line of quartz crystals are also available.  
All oscillators can be manufactured to meet MIL-O-55310,  
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## Tcxo

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### Frequency Range:

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### Frequency Stability:

$\pm 0.8$  PPM  
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**Aging:**  $\pm 1.0$  PPM/yr  
typ.

### Supply Voltage:

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### Supply Current:

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### Size:

Standard:

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As small as:

0.960"  $\times$  0.5"  $\times$  0.2"



## Cxo

*Crystal Clock  
Oscillator*

### Frequency Range:

TTL: 10 Hz to 100 MHz

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ECL: 5 MHz to 500 MHz

Sinewave: 1 Hz-1 GHz

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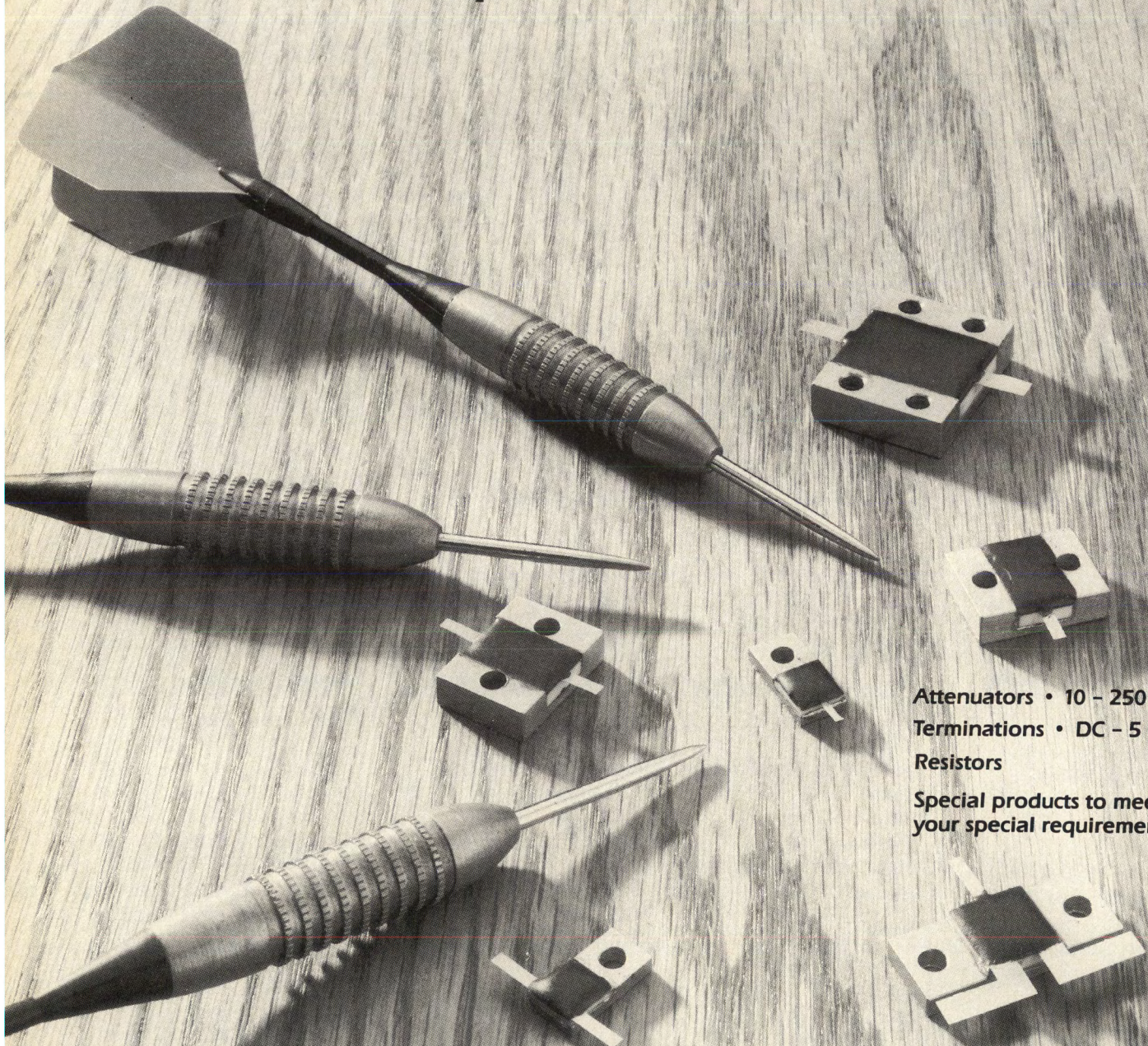
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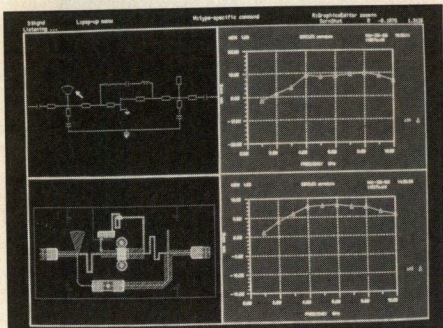
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INFO/CARD 20



## Cadence Introduces Microwave Design System



*Cadence's Microwave Musician*

The Microwave Group of Cadence Design Systems, announced at the 1990 MTT Symposium the release of the Cadence Microwave Musician Design System. The software is the first commercial microwave design solution that provides engineers with a complete and comprehensive suite of highly integrated design automation tools for the development of microwave circuits and systems. Microwave Musician Design System was developed to shorten the design time for microwave components for commercial and military applications.

The system provides full front-to-back design capability for both monolithic microwave integrated circuits (MMIC) and hybrid devices. It includes a schematic capture package, libraries, interfaces to most simulators, and a set of microwave-specific layout and verification tools and allows designers to begin their designs from either schematic entry or layout and move on to simulation.

The design system simulation interface provides simulator control, back annotation of optimized results and graphical display of analyses. Microwave Musician allows multiple simultaneous simulation runs. It supports complete integration of industry standard microwave simulators into the system. Microwave Design's first supported simulator is Microwave Harmonica™, from Compact Software, additional interfaces are expected in the third quarter 1990.

**Circuit Busters Announces New Name** — Circuit Busters is changing its name to Eagleware Corporation. The offices of Eagleware will remain at their Stone Mountain, Georgia address.

## Mt. Everest Team to Carry Satellite Communications Equipment

— A joint Soviet-US-Chinese expedition will carry an advanced transportable satellite communications terminal on the 1990 Peace Climb. The Magnavox system will provide instant telephone, telex and fax links with international switched telecommunications networks, through the INMARSAT network of satellites and earth stations. The team leaders hope to transmit a live message from the peak of Mt. Everest.

## GE Resin Used in 3D Molded Circuit Board

— GE has announced a new, two-way hand-held FM radio built with a three dimensional circuit board. The board features multiple layers of circuitry and has a reduced number of parts. It is manufactured by the Electronic Components Division of Allen-Bradley Co., and replaces several different components typical of most two-way radios, providing faster, easier assembly and simplifying repair work for the end user.

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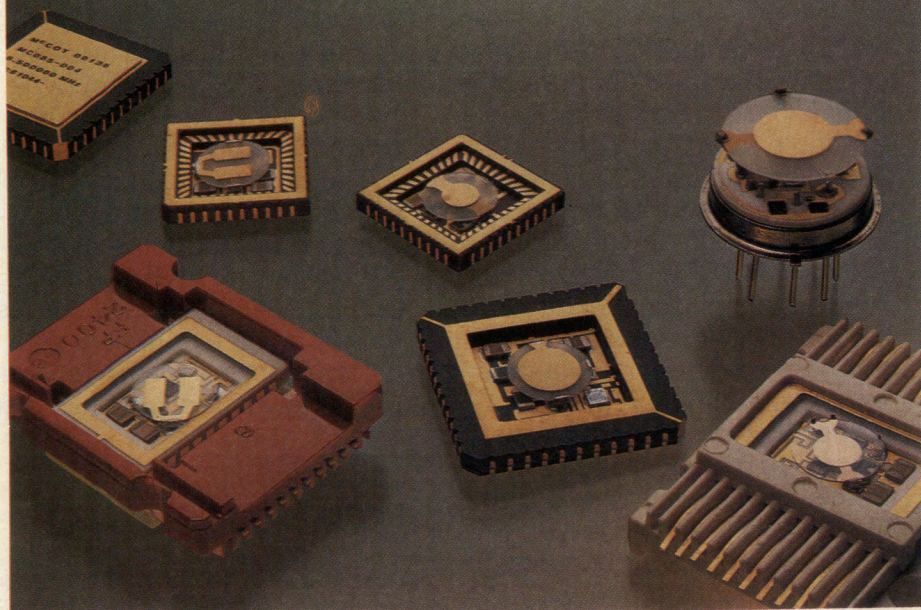
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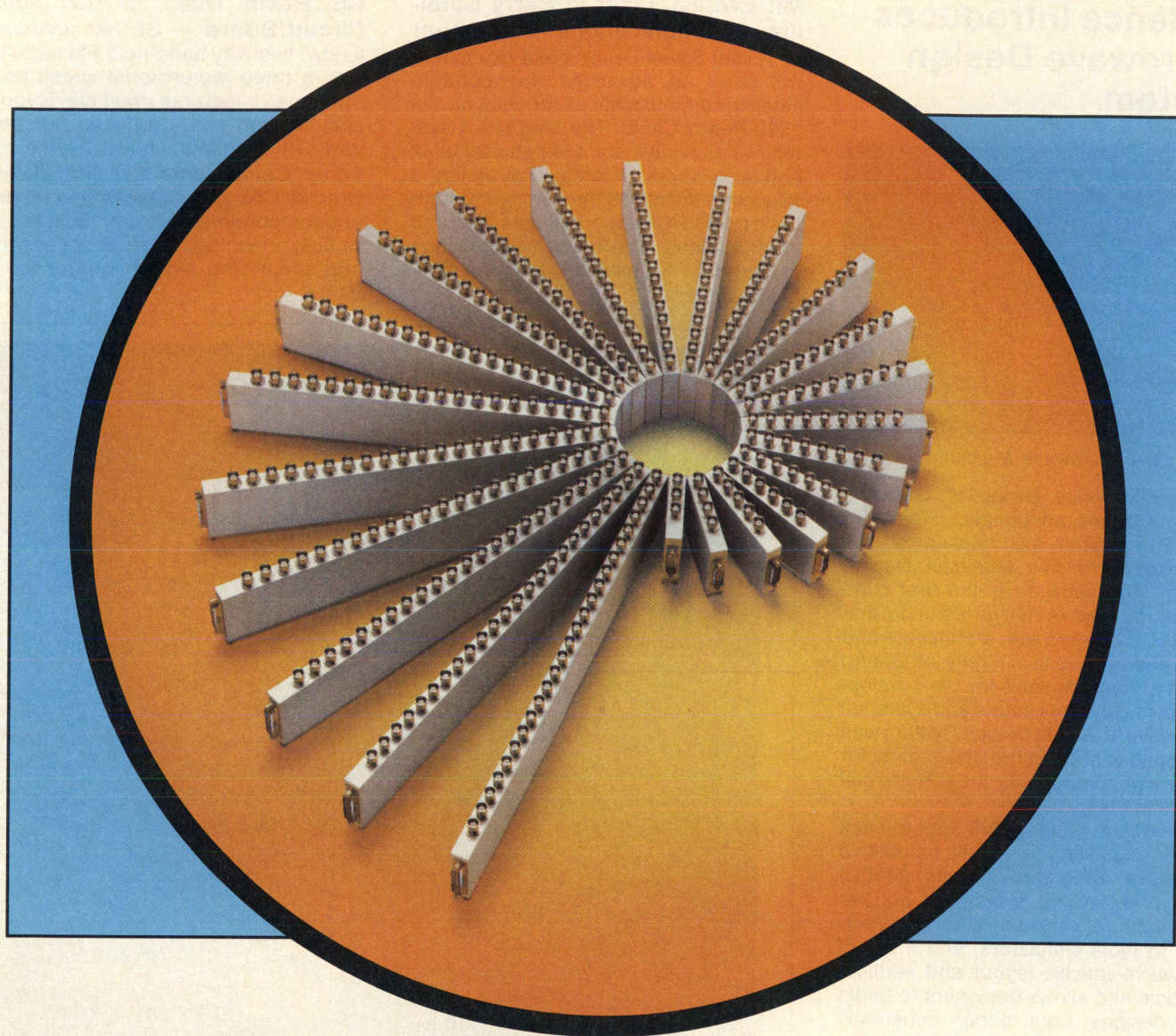
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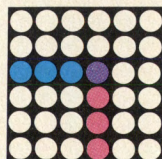
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**Call for Papers Announced** — The International Telecommunications Union has issued a call for papers for their symposium "Integration, Interoperation and Interconnection: The Way to Global Services." The symposium will be held in conjunction with TELECOM 91, October 7-15, 1991. Papers are invited that deal with aspects of present and future technologies, networks and services related to telecommunications. For information write to: FORUM 91 Secretariat, International Telecommunication Union, Place des Nations, CH-1211, Geneva, Switzerland, 20. Tel: 41 22-7305680.

**New Director Confirmed to NIST**

— The U.S. Senate has confirmed the appointment of Dr. John W. Lyons, a chemist and senior manager at NIST, as the ninth director of that agency. In his confirmation hearing, Lyons declared that the agency should continue its strong research program in measurement technology necessary to support quality assurance programs in industry, but should increase its emphasis on "generic" research in newly emerging technologies.

**Sawtek Awarded Contract by Norden Systems**

— Sawtek Inc. has been awarded a contract for the production of state-of-the-art surface acoustic wave (SAW) components and subsystems at a value of approximately seven million dollars. The two and one half year production program for Norden Systems Inc. includes the delivery of advanced SAW expansion and compression radar subsystems as well as high-performance SAW bandpass filters.

**Harris Receives RF Test Bed Contract**

— The Advanced Television Test Center announced it has awarded Harris Corporation, Broadcast Division, a major contract to develop the specialized RF Test Bed to help evaluate advanced television (ATV) transmission systems. The combination of equipment in the RF test bed will permit simulation of over-the-air conditions to be encountered by broadcast ATV signals, and thus help the test center ascertain the likely performance of different kinds of new television systems.

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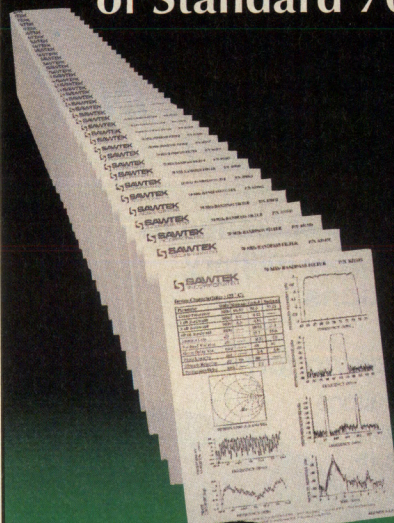
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# The Low Noise Transistor Market

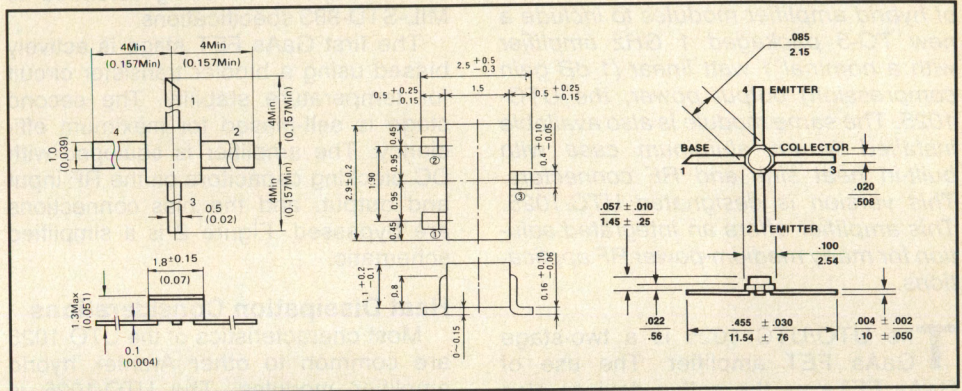
By Gary A. Breed  
Editor

Low noise transistors represent a relatively mature market and a technology that is steadily, but not dramatically growing. However, like other parts of the RF marketplace, there are application areas that are literally exploding, and technologies that will create more excitement in the future.

It seems appropriate to mention the status of the military market, then move on to other matters. Military programs use plenty of GaAs and silicon low noise transistors for communications, radar, avionics, EW and ECM, navigation, and satellite systems. This is a substantial part of the present market, but it is not the most exciting part. The single greatest concern among transistor makers is how to increase their sales to non-military customers. Thus, applications in consumer and commercial RF products are getting much more attention right now.

RF and microwave transistor companies are pleased to comment on the development of commercial applications for their products. Direct Broadcast Satellite (DBS), cellular radio, 800 MHz land mobile, Global Positioning System (GPS), spread spectrum data links, and L-band Personal Communications Network (PCN) are mentioned as areas that are generating business now, with the promise of bigger things to come. Much of the work is in traditional communications. "There's quite a lot of market activity for transceivers," notes Gary LaBelle, Director of Marketing, Semiconductor Products, at Avantek, Inc., adding the observation, "The frequency range of these applications applies well to our technology."

Other companies echo this last comment. Fujitsu sees Ku-band DBS as a major source of customers for their high performance, ultra low noise HEMT transistors. Michel Pradille of Philips Components adds that the medical market is growing. As an example, sensitive detectors for magnetic resonance imaging (MRI) systems require quality VHF front-end transistors. Although companies such as H-P and Motorola have significant internal sales, they are equally interested in other opportunities, particularly those that fit



their technological capabilities.

The diversity of applications means flexibility in product availability, too. "We manufacture quite a variety of devices in different packaging styles," says Gary Franklin of Hewlett-Packard's Microwave Semiconductor Division. He confirms that surface mount packaging is particularly important, as more manufacturers use automated manufacturing with pick-and-place assembly equipment.

## More Integration, Fewer Discrete Devices

The cost of the finished product, particularly at the traditionally more expensive UHF and microwave frequencies, is driving the marketplace. "The market trend is moving from discrete components to high level integration," according to Gary McCollum of Hewlett-Packard. Indeed, products such as Ku-band receivers will have to compete with inexpensive video sources like VCRs and CATV and standard TV. Products can have a much lower overall cost with major portions implemented in silicon or GaAs.

It won't be very long before integrated amplifiers, mixers, and gain control amplifiers can combine to reach the unofficial benchmark of 1.0 dB noise figure at 1 GHz, but discrete devices may never give up their performance edge. In the time it takes for a technology to develop from lab to discrete devices to integrated circuits, another new advance will be underway.

Another aspect of the cost equation that may prove difficult for some semi-

conductor manufacturers is that high volume production is necessary for low cost components for the consumer market. In shifting from a military/aerospace dominated market to a commercial/consumer market, the cost per unit must come down dramatically. Much of the military cost is incurred in testing and documentation; many more consumer devices have to be shipped to earn the same revenue. Although this conclusion is obvious, the methods needed to implement it are not.

All of the transistor manufacturers have significant experience in commercial and consumer devices. Thus, they do not have to cope with a complete change in their own operating procedures. The principal challenge is being prepared to respond when a significant application reaches the production line. To take advantage of the upcoming RF product developments, companies will need flexibility, rapid response, and access to the appropriate technology.

## Conclusion

There are many new opportunities for the low noise transistor market; opportunities that match up well with available technologies, especially in the UHF and low microwave frequency range. Some of these are strong now, such as cellular radio and DBS. Others are in various stages of development, like GPS and consumer spread spectrum. In all, the opportunities add up to make a substantial market for discrete low noise transistors, allowing a relatively smooth transition from military emphasis to commercial applications.



# A 1-Watt Amplifier for 20-1000 MHz

By Jim Gustafson and Melanie Roberts  
Avantek, Inc.

*Avantek, Inc. has expanded its line of hybrid amplifier modules to include a new TO-3 packaged 1 GHz amplifier with a nominal 1 watt linear (1 dB gain compression) output power, the UTO-1025. The same module is also available installed in an aluminum case with built-in heat sink and RF connectors. This version is designated UTC-1025. This amplifier offers an integrated solution for many medium-power RF applications.*

The UTO/UTC 1025 is a two-stage GaAs FET amplifier. The use of GaAs FETs as the active devices also gives the amplifier the very low noise figure of 3.0 dB (typical, at 25C) from 200-1000 MHz. This, combined with the amplifier's power handling capability, makes it a possible choice as a "clobber-proof" front end for receivers and instruments which require a dynamic range up to approximately 85 dB.

The 20-1000 MHz frequency range is a home to a wide range of services using transmitters at the 1-watt power level, including both civil and military aeronautical VHF, maritime VHF, and TACAN. Among the uses of this new amplifier are low-band VHF radio through 800 MHz trunked and cellular systems. Several Avantek test setups use the connectorized amplifier for increasing the output of signal generators.

The UTO/UTC 1025 amplifier provides a guaranteed minimum +28.5 dBm power output over the 20-1000 MHz frequency and -55 to +85C temperature ranges (+30.5 dBm typical at 25C). The amplifier module also features guaranteed 22 dB minimum gain with  $\pm 1.0$  dB full-band flatness, 4.5 dB noise figure from 20-1000 MHz, maximum 2.0:1 input and 2.2:1 output VSWR, and a +36 dBm two-tone, third order intercept point.

Table 1 shows the typical and guaranteed performance characteristics over the 0 to 50C and -55 to +85C temperature ranges, and at 25C. Figure 1 is a set of performance curves for power output, noise figure and VSWR. Compared to other wideband power amplifiers, the UTO-1025 offers higher efficiency, operates from +15 VDC bias instead of the more common +24 VDC,

and is capable of being screened to MIL-STD-883 specifications.

The first GaAs FET stage is actively biased using a bipolar transistor circuit for temperature stability. The second stage is self-biased for maximum efficiency. The amplifier is equipped with DC blocking capacitors on the RF input and output, and the bias connections are bypassed. Figure 2 is a simplified schematic.

## Heat Dissipation Considerations

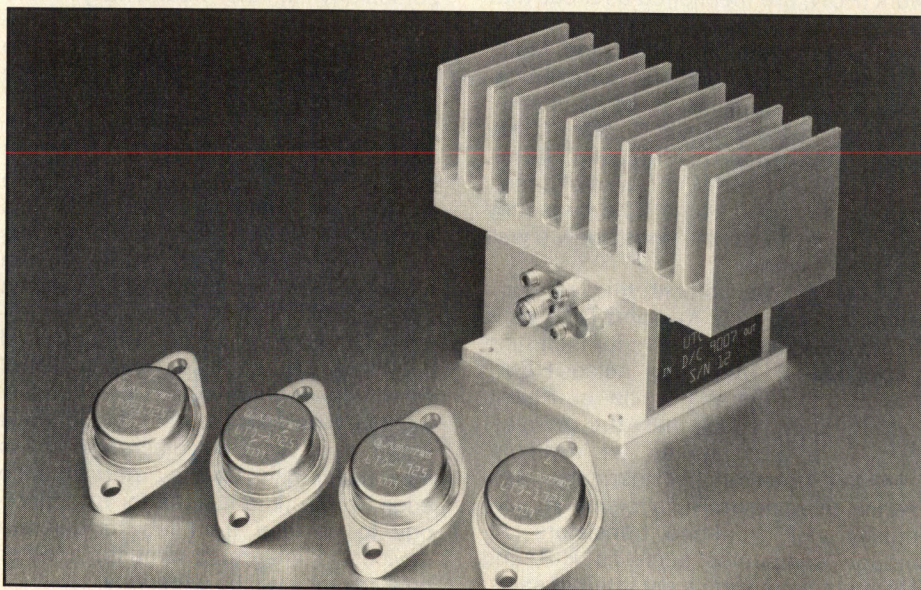
Most characteristics of the UTO-1025 are common to other Avantek hybrid amplifier modules. The UTO-1025 is unconditionally stable and can easily be cascaded with other Avantek UTO-series modules in a 50 ohm microstrip environment with no degradation of performance.

However, this is a class A amplifier which dissipates 7.5 watts of heat whether or not a signal is present. The GaAs FETs are rated for a maximum continuous operating channel temperature of 150C, with an absolute maximum of 175C under any conditions. MTBF drops linearly between these two temperatures, therefore, it is critical to remove heat from the device to maintain a safe operating temperature.

To help the designer in providing sufficient cooling, Avantek provides several parameters which aid in tracking the channel temperature: junction (channel)-to-case thermal impedance ( $\theta_{jc}$ ), individual transistor power dissipation ( $P_d$ ), and junction (channel) temperature rise ( $T_j$ ).

It must be emphasized that the ( $T_j$ ) specification is the actual rise of the active channel temperature above the device case temperature, and not the rise above ambient temperature. Channel temperature rise is calculated as the product of the measured channel-to-case thermal impedance ( $\theta_{jc}$ ) and the known transistor power dissipation. This calculated value is not the same as the overall device power dissipation of 7.5 watts.

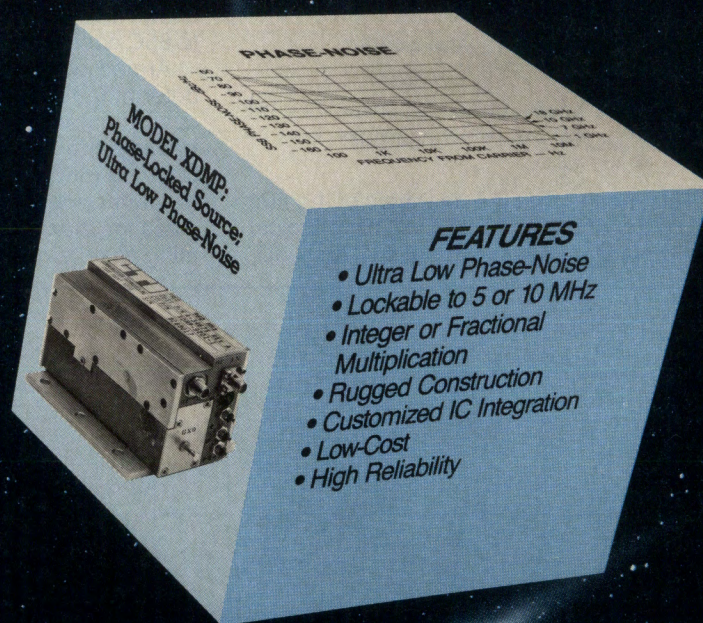
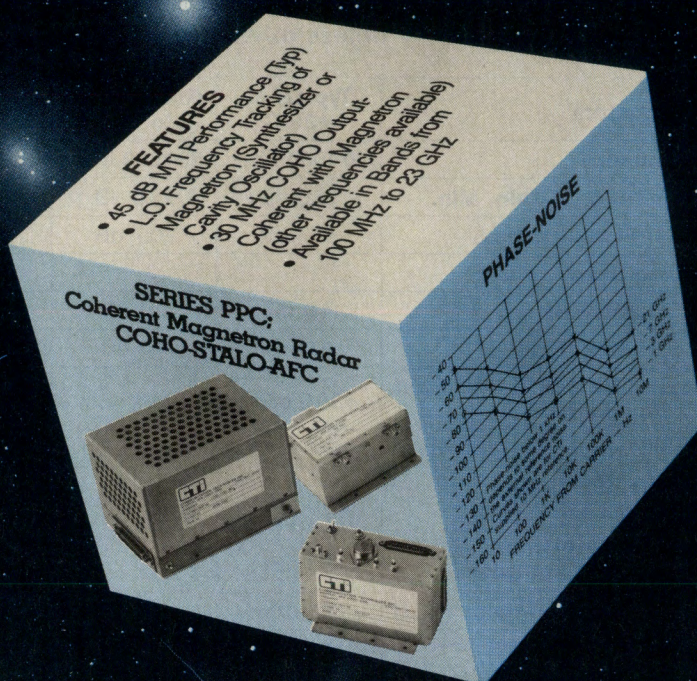
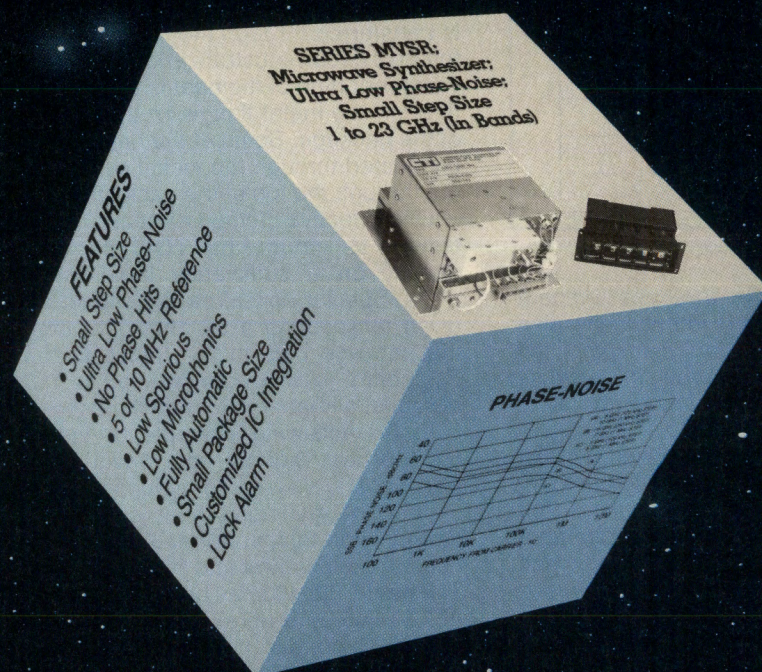
The specific GaAs FETs used in the amplifier were selected for their thermal characteristics, to provide a wide safety margin between the maximum rated power dissipation and the actual dissipation in the circuit. In the UTO-1025, the actual power dissipation of the FETs is 1.15 W and 3.42 W, respectively, compared to their 4.0 and 8.3 W maxima. In addition to careful selection of the devices and bias, the thermal design was improved by using a BeO substrate between the case and transistors. The



The Avantek UTO-1025, in a hermetic TO-3 package, and the UTC-1025, in a connectorized assembly.



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| Parameters                                  | Typical at +25°C | Guaranteed, 0 to +50°C | Guaranteed, -55° to -85°C |
|---------------------------------------------|------------------|------------------------|---------------------------|
| Frequency                                   | 20-1000 MHz      | 20-1000 MHz            | 20-1000 MHz               |
| Small Signal Gain, Min.                     | 25.0 dB          | 23.0 dB                | 22.0 dB                   |
| Gain Flatness, Max.                         | $\pm 0.7$ dB     | $\pm 0.7$ dB           | $\pm 1.0$ dB              |
| Noise Figure, Max.:                         |                  |                        |                           |
| 20-200 MHz                                  | 4.0 dB           | 7.0 dB                 | 8.0 dB                    |
| 200-1000 MHz                                | 3.0 dB           | 4.0 dB                 | 4.5 dB                    |
| Power Output at 1 dB Gain Compression, Min. | +30.5 dBm        | +29.0 dBm              | +28.5 dBm                 |
| VSWR, Input, Output, Max.                   | 1.6:1, 1.7:1     | 2.0:1, 2.0:1           | 2.0:1, 2.2:1              |
| 2-Tone, 3rd Order Intercept Point (IP3)     | +40 dBm          | +37 dBm                | +36 dBm                   |
| 2-Tone, 2nd Order Intercept Point (IP2)     | +53 dBm          |                        |                           |
| 1-Tone, 2nd Harmonic (Hp2)                  | +60 dBm          |                        |                           |
| Input Voltage ( $\pm 1\%$ )                 | +15 VDC          |                        |                           |
| DC Current, Typ.                            | 500 mA           |                        |                           |

Table 1. Specifications for the UTO-1025 1 Watt amplifier.

temperature rise of each channel above case temperature may be calculated as follows:

$$T_{r1} = \theta_{jc1} \times P_{d1} \quad (1)$$

$$= 39\text{C/W} \times 1.15 \text{ W} = 44.85\text{C}$$

$$T_{r2} = \theta_{jc2} \times P_{d2} \quad (2)$$

$$= 19\text{C/W} \times 3.42 \text{ W} = 64.98\text{C}$$

This means that the channels of the first and second stage FETs will have typical temperature rises of about 45C and 65C,

respectively, above the case temperature. Note that this is intentionally lower than is commonly found in amplifiers with this output power level, which allows the UTO-1025 to operate at up to 85C case temperature.

Armed with this information, we can determine the maximum safe operating case temperature for the device. Subtracting the highest channel temperature rise (65C) from the 150C maximum recommended channel temperature gives the maximum recommended operating case temperature of 85C.

The majority of the remaining power in the amplifier (2.93 W) is dissipated in the source and drain resistors, with less than 5 percent of the total power dissipated in the balance of the active circuitry. The temperature rise on these elements does not appreciably affect the operation of the circuit. The thermal network shown in Figure 3 illustrates the flow and dissipation of heat in the UTO-1025.

### An Example of Heat Removal

The heat from the transistor channels is transferred through the transistor die, the BeO circuit substrate and, finally, the vast proportion of the heat is transferred through the metal flange on the bottom of the package. Some of the heat is also conducted through the nickel cover and out through the top of the package.

One effective method of heat dissipation is to mount the TO-3 package directly on a sufficiently large metal plate. Special consideration should be given to ensuring a 50 ohm path from the package pins to the traces of the PC board. If this configuration is not possible, a heat radiator may be used.

In the following example, a model 500403 heat radiator from Aavid Engineering (1) is used to demonstrate a method of managing heat dissipation and junction temperatures under conditions of both natural and forced air flow. The model 500403 has base dimensions of 1.81 inches square, a base thickness of 0.093 inch, and from the base to the top of the radiating fins is 1.25 inches. It is drilled with its own mounting holes, but will need to be drilled for the leads and mounting screws of the Avantek 4-pin TO-3 package. Figure 4 is a photo of the heat radiator, showing how the TO-3 package fits, and Figure 5 is a performance curve.

With natural air flow, the Aavid radiator will limit the ambient-to-heat sink temperature rise to approximately 39C. Using forced air can reduce the sink-to-ambient thermal resistance by at least a factor of four, reducing the temperature rise to about 8C.

Heat flow is analogous to current flow in an electrical circuit, with temperatures the dual of voltages and thermal impedance the dual of electrical resistance. The thermal network in Figure 6 shows the temperature potentials and thermal resistances. For this example with natural air flow, a rise of 39C from ambient to heat sink and a 1 degree rise from heat sink to case would allow a maximum ambient temperature of 45C.

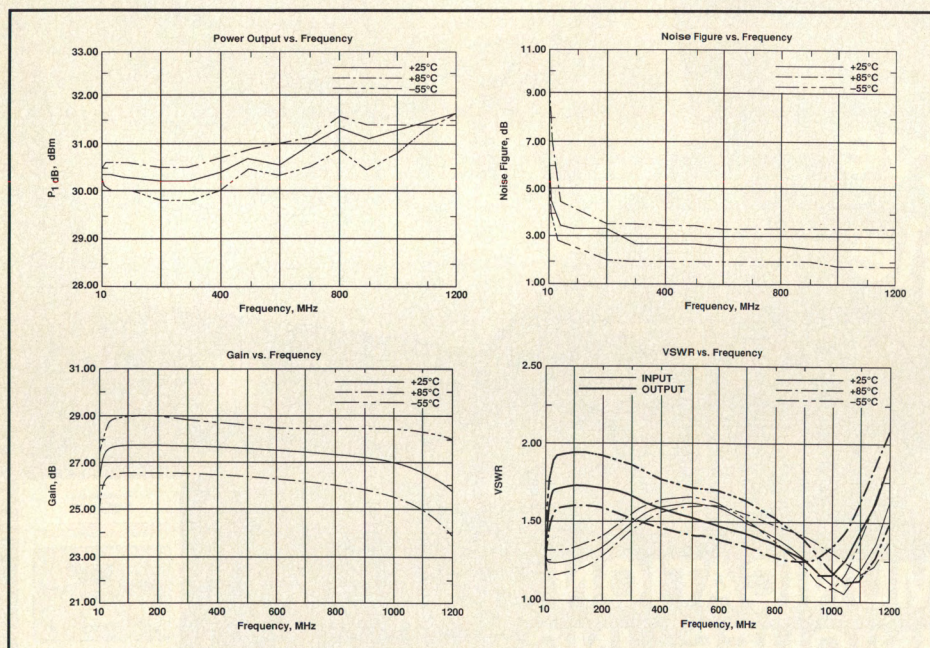
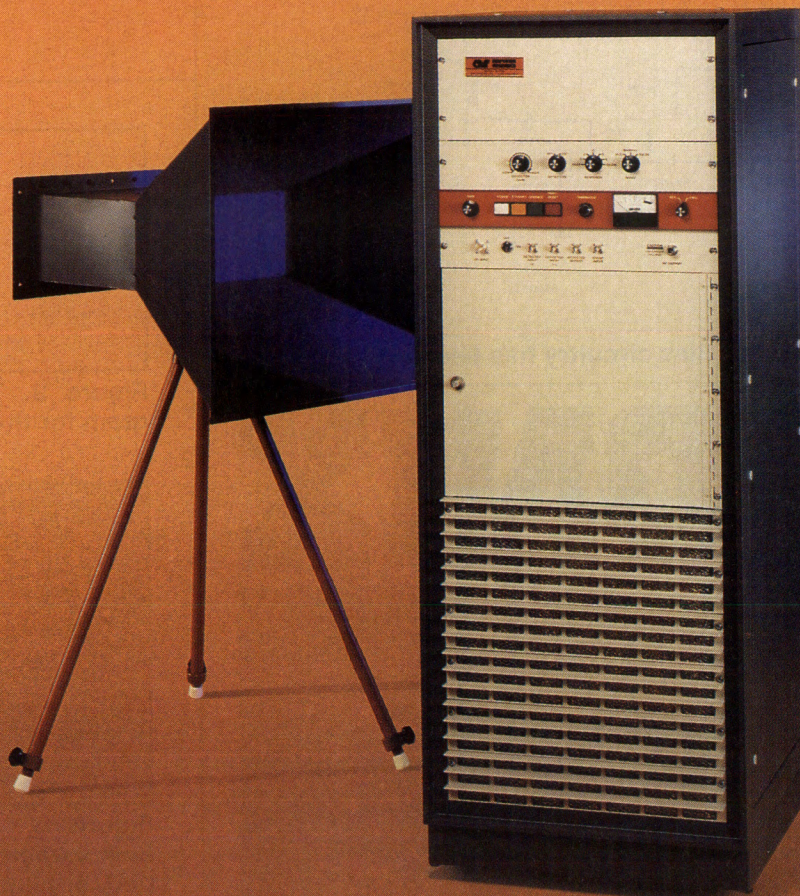


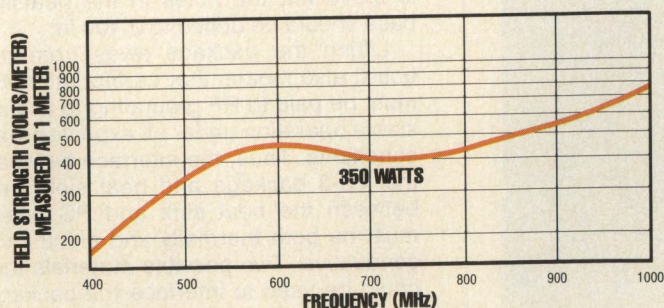
Figure 1. Measured performance at extreme and room temperatures.





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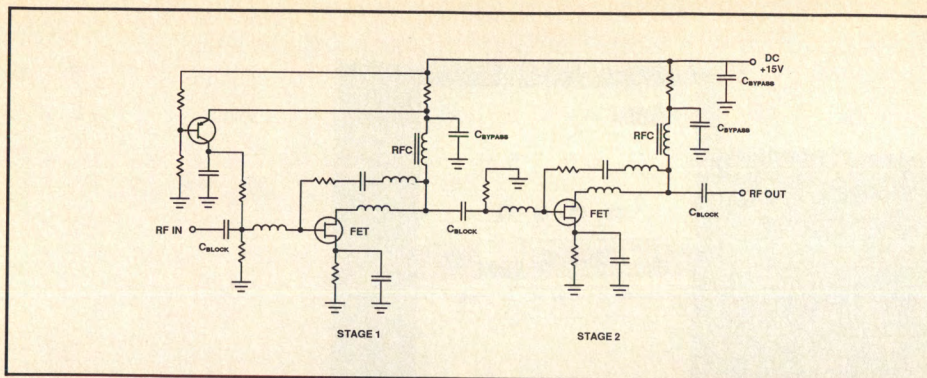


Figure 2. Simplified schematic. Bias circuitry has been omitted.

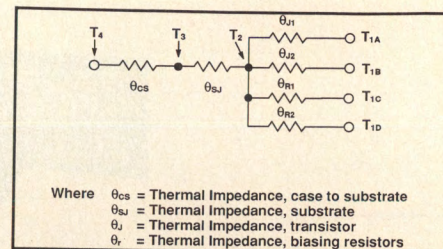
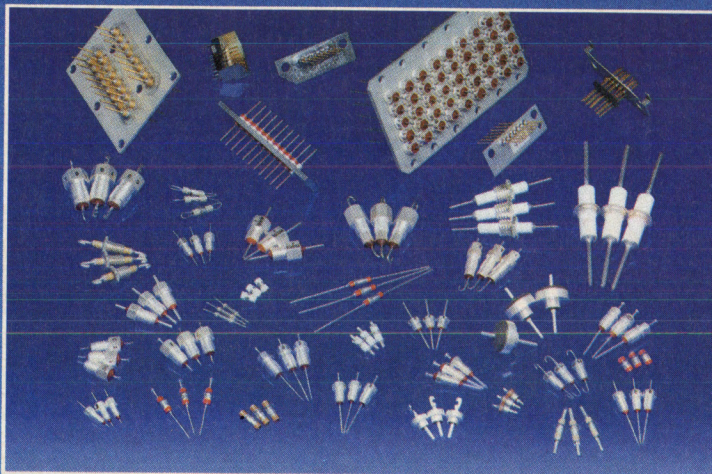


Figure 3. Thermal network diagram for the UTO-1025.

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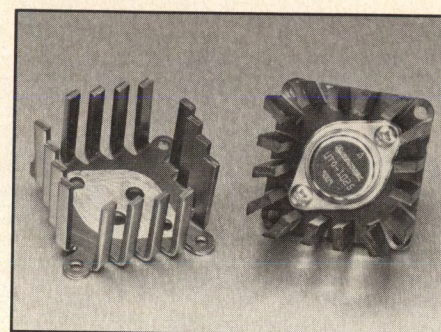


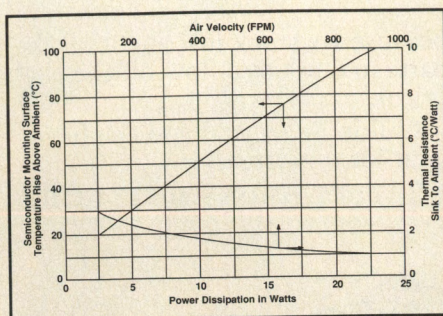
Figure 4. Aavid model 500403 heat radiator (left) with Kon-Dux pad applied. On the right is a UTO-1025 in a model 568403 radiator.

Forced air with a velocity of 800 FPM over the same heat sink could increase the maximum ambient temperature to approximately 75°C.

Since this type of heat sink lifts the TO-3 package flange away from the PC board by 0.093 in., higher-frequency operation will require a continuation of the 50 ohm impedance from the microstrip PC board to the input and output pins. Using air as the dielectric, a pin-to-hole ratio of 2.5 will give a transmission line impedance of about 55 ohms. Since the package pin diameter is 0.040 in., the holes in the heatsink base should be drilled to 0.100 in.

Lifting the package away from the board also means that careful attention must be paid to RF grounding to assure stable operation under all expected load conditions. Thus, the interface between the TO-3 package and heat sink, and between the heat sink and PC board must be both thermally and electrically conductive. Two possible materials that could be used to interface the package to the heat sink are Kon-Dux™, a graphite based pad provided and factory-applied by Aavid, and Ablefilm® (2), a silver-loaded epoxy available as preform or sheets. Beryllium copper sheets may also be used, and provide excellent RF and heat transfer characteristics.





**Figure 5. Performance of the Aavid 500403 heat radiator.**

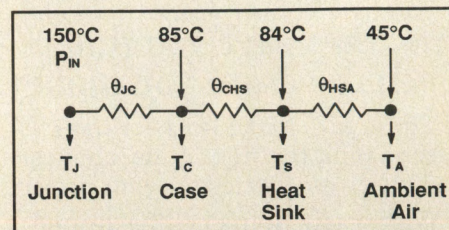
Many other mounting possibilities exist. For example, a heat radiator may be attached to the top surface of the UTO-1025 and the package mounted flush with the PC board ground plane. In the UTO-1025 version, the finned heat radiator on the case will provide adequate heat dissipation in free air for ambient temperatures up to 50°C. Operational temperature tests on the UTO header and channel temperature scans have confirmed these values to be very accurate.

If an installation precludes the use of a heat radiator in natural air flow, forced air or some form of conductive cooling will be necessary. In this example, forced air was shown to increase the maximum ambient performance by 50 percent over natural air flow. By using liquid cooling, the maximum ambient could be raised to within 1 degree of the maximum case temperature of 85°C.

#### Availability

The UTO-1025 is priced at \$595.00 and the UTC-1025 is \$695.00. Both units are available from any Avantek stocking distributor. Avantek is also developing a complementary silicon bipolar transistor TO-3 module providing 1 W minimum output from 5 to 500 MHz, and a 0.5 to 2 GHz FET version, 1 W output.

For a copy of the data sheet, and a list of distributors, contact Avantek, Inc., 481 Cottonwood Drive, Milpitas, CA 95035, (408) 943-3038, or circle INFO/CARD #174. For technical information, call (408) 942-4296.



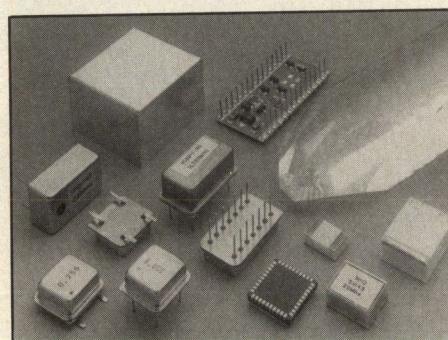
**Figure 6. Thermal model from transistor channel to free air using the 500403 heat sink.**

#### References

1. Aavid Engineering, Inc., One Kool Path, P.O. Box 400, Laconia, NH 03247, (603) 528-3400.
2. Ablestik Laboratories, 833 W. 182nd St., Gardena, CA 90248, (213) 532-9341.

#### About the Authors

Jim Gustafson is Applications Engineer, and Melanie Roberts is Senior Design Engineer at Avantek, Inc., 481 Cottonwood Drive, Milpitas, CA 95035.



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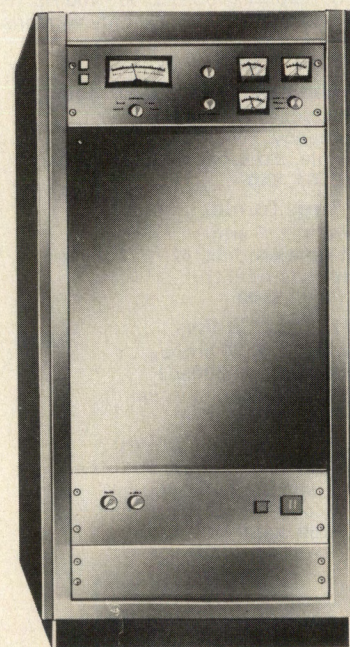
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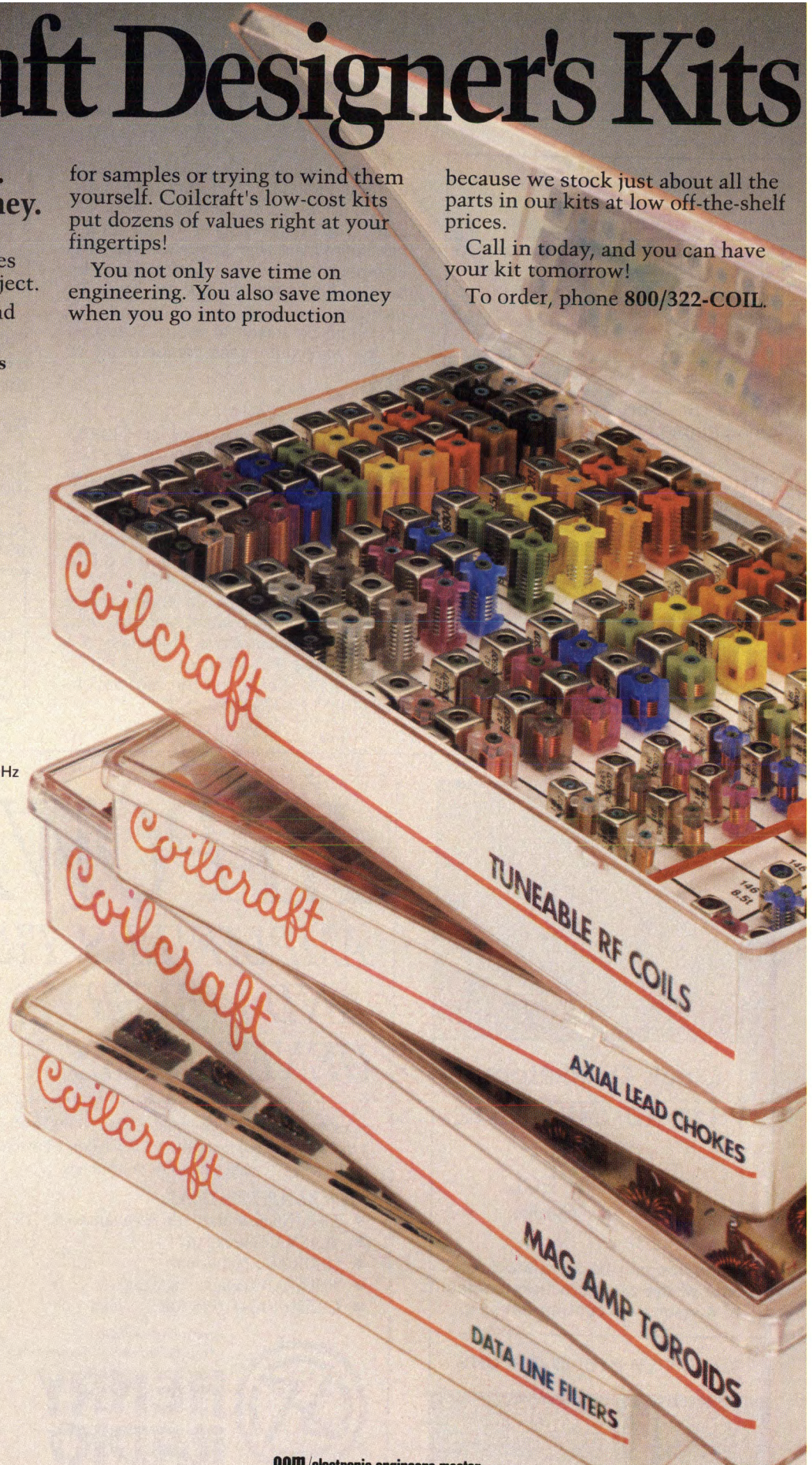
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INFO/CARD 31



# A 70 MHz Wideband Phase Shifter

By Francois Methot  
Consortel Registered

This circuit, a realization of a wideband phase shifter at 70 MHz, was an entry in the 1989 RF Design Awards Contest. Although the concept of using quadrature couplers as phase shifters is well known, this design was developed to improve performance and reduce costs.

The circuit is constructed using three quadrature coupler sections in cascade, using reactive terminations to achieve adjustable phase shifts. Figure 1 shows both the standard quadrature coupler phase shifter, and its modification for use in this design.

The calculation of each element is given by the following equations:

$$K = 10^{\frac{(\text{COUPLING (dB)})}{20}} \quad (1)$$

$$M = KL_1L_2 = KL_1 \quad (2)$$

$$Z_{oe} = Z_o \frac{1 + K}{1 - K} \quad (3)$$

$$X_{co} = Z_{oe}, \text{ therefore, } C_o = \frac{1}{2\pi F_o Z_{oe}} \quad (4)$$

$$X_{c1} = \frac{(1 - K)}{K} Z_{oe}, \text{ therefore, } C_1 = \frac{K}{1 - K} C_o \quad (5)$$

$$X_{L1} = X_{L2} = 2Z_{oe}(1 - K), \text{ therefore, } L_1 = L_2 = \frac{2Z_{oe}(1 - K)}{2\pi F_o} \quad (6)$$

In implementing equation 6, a K of 0.707 is not easy to achieve. However, a K of approximately 1 is a more practical value.

To realize an equivalent of K=0.707,  $L_1$  and  $L_2$  are transformed to a smaller total inductance, with  $K \approx 1$ . The remaining inductance is added by the four  $L_3$  elements.

$$L_4 = 0.707KL_1 \quad (7)$$

$$L_3 = \frac{(L_1 - L_4)}{2} \quad (8)$$

A twisted pair inductor ( $K \approx 1$ ) of the value of  $L_4$  can readily be built. Adjustable reactive termination is done using a varactor diode, as shown in Figure 2. Three sections are sufficient to achieve a total phase shift greater than 360 degrees, over about 30 percent bandwidth.

Figure 3 shows a complete schematic diagram of the phase shifter. Figures 4 and 5 are plots of the phase and amplitude response of the unit versus DC control voltage.

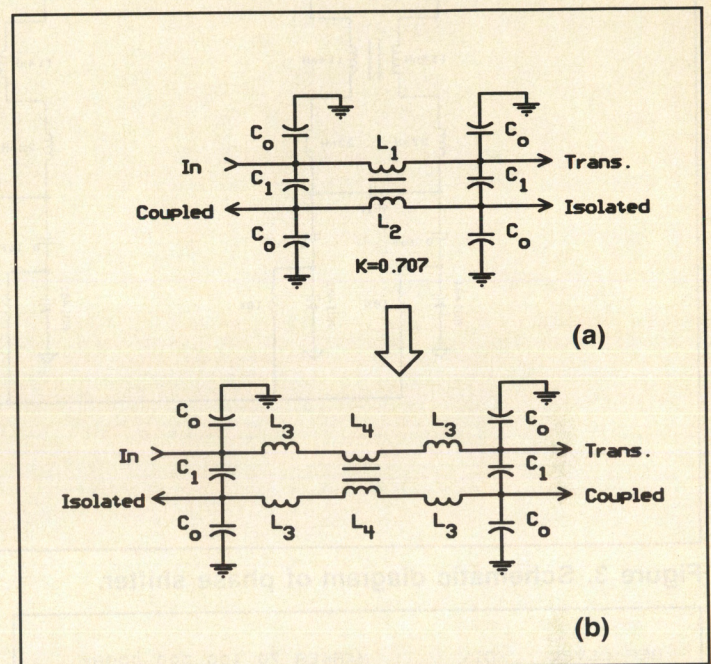


Figure 1. Standard quadrature coupler and modification.

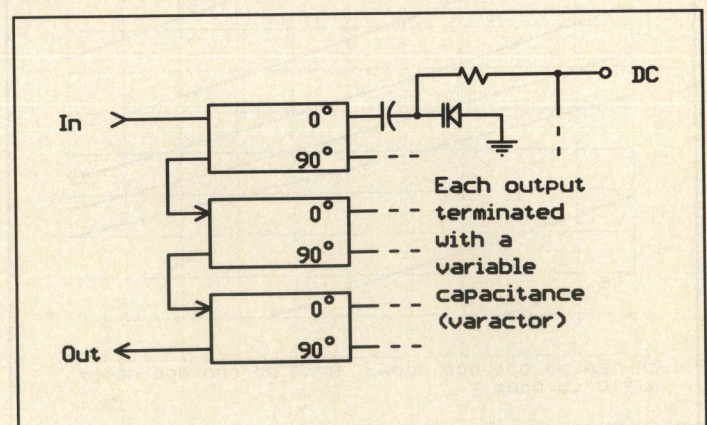


Figure 2. Varactor diode used for adjustable reactive termination.

## About the Author

Francois Methot is a group leader in the analog design division of Consortel. He can be reached at Consortel, 6900 Trans Canada, Point Claire, Quebec, Canada, H9R 1C2.



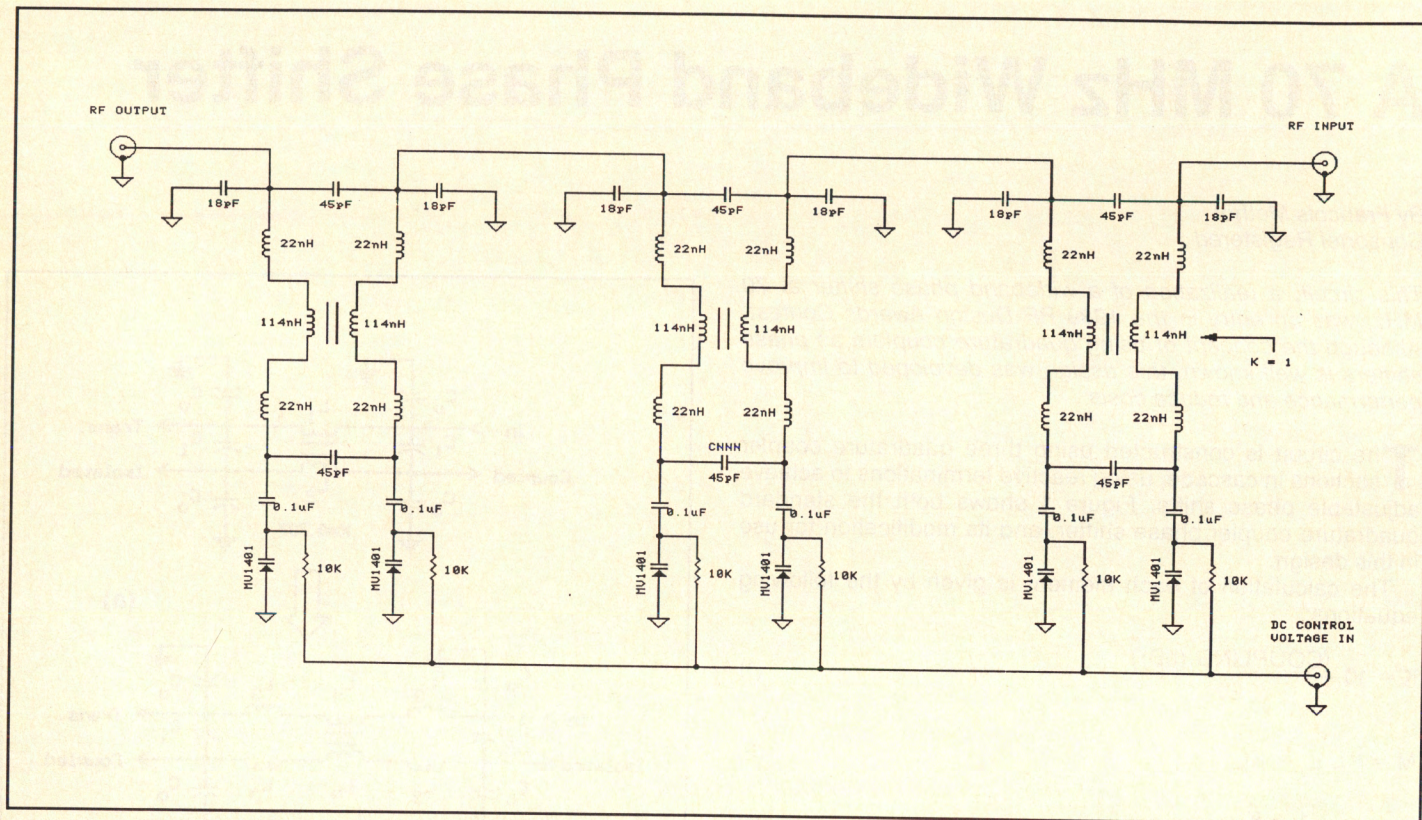


Figure 3. Schematic diagram of phase shifter.

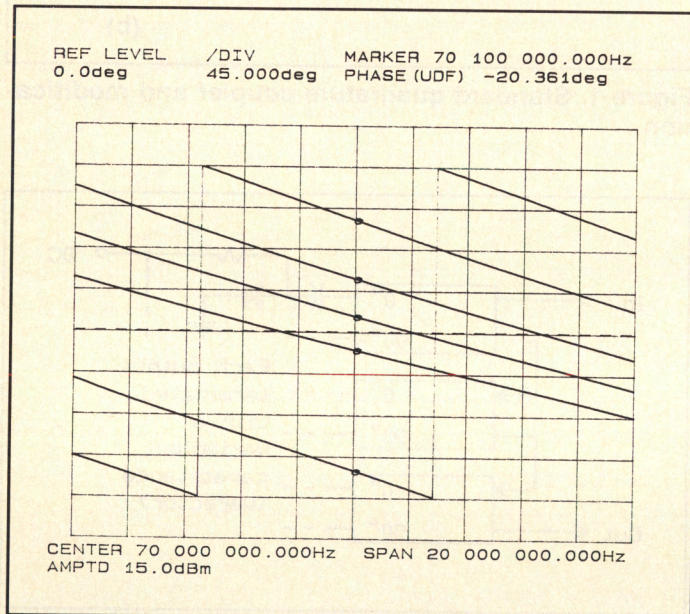


Figure 4. Plot of phase and amplitude response vs. DC control voltage.

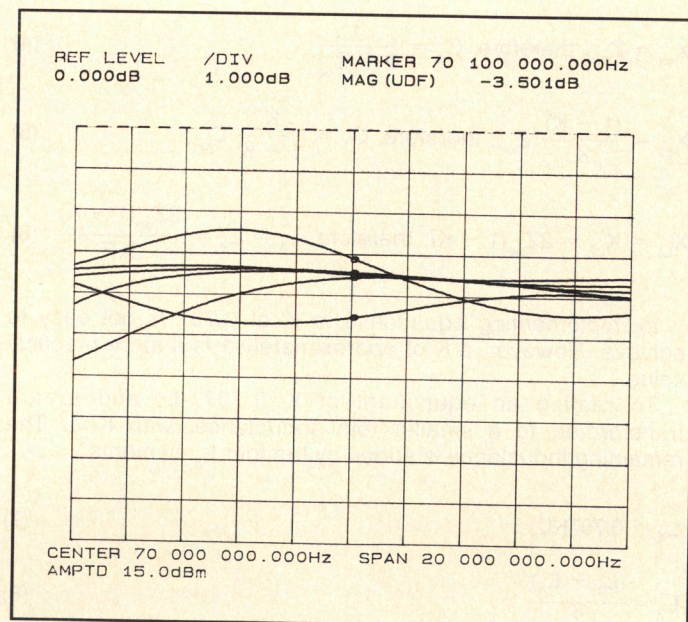


Figure 5. Plot of phase and amplitude response vs. DC control voltage.



# A 1.8 GHz Odd Ratio Frequency Divider

By David Korn, Naval Research Lab and  
Carl Deierling, Select Electronics

*Presented is a new technique for building an odd-ratio frequency divider at a demonstrated 1.8 GHz maximum input frequency which was conceived, designed, tested and built by the authors. In this article, the word "odd" is used as a mnemonic for a non- $2^n$  integer rather than a non-even number. This divider, which was necessitated by, and is intended for RF frequency synthesizer applications, can divide by any integer; however, its novelty is most useful when the integer chosen is a non- $2^n$  number, namely 3, 5, 6, 7 etc., as there are many multi-GHz input  $2^n$  (2,4,8,16, etc.) dividers presently available. Both the circuit itself and its intended function—non- $2^n$  ratio multi-GHz input frequency division—appear to be novel and most certainly useful. Consequently, it gives the frequency synthesizer designer another design option not currently available.*

The circuit concept is based on a D flip-flop with a delay  $\tau$  fed back from the inverted Q output to the flip-flop clear input tied high; the divide ratio N is controlled by selecting  $\tau$ . The input frequency is applied to the clock input and the output frequency is the Q output. Since this RF circuit is based on a digital IC, its features are directly coupled to the anticipated advances in high-speed digital ICs, namely increasing maximum input clock frequency and minimizing the residual (additive) phase noise, physical size and DC power consumption. The most important measured inherent characteristics of this divider technique from a frequency synthesis point of view are: lack of measurable spurious oscillations as the input RF frequency, RF power, DC voltage and ambient temperature are changed, a scarcity of undesired subharmonics, very small size, low DC consumption, low input RF drive and moderate RF power output delivery.

Frequency division is an integral part of RF frequency synthesizer techniques.

The majority of frequency dividers are realized digitally via edge-triggered flip-flops and the remainder are realized as analog circuits based on varactor diodes and resonant circuits.

Edge-triggered digital divider circuits are designed by configuring a flip-flop to change state on a rising or a falling edge, but not on both—this in effect is a divide-by-2. A divide-by- $2^n$  (2,4,8 etc.) is made possible by cascading the desired number of flip-flops. If a non- $2^n$  divide ratio is the goal, a feedback circuit between flip-flops can accomplish this. First, consider the pre-designed option. Plessey Semiconductor has the most comprehensive selection of pre-designed dividers, all built with silicon. Presented here are the highest input frequency odd-ratio dividers for  $N < 10$ . The SP8720 divides by 3 at  $F_{\max} = 300$  MHz, the SP8620 divides by 5 at  $F_{\max} = 400$  MHz, the SP 8740 divides by 6 or 7 at  $F_{\max} = 300$  MHz, and the SP8743 divides by 9 at  $F_{\max} = 500$  MHz. GigaBit Logic has the highest non- $2^n$  divider—the 10G070—a 2.0 GHz dual-modulus divider; however, it divides by 5 or 6 only. If one wants to divide by 3,5,6,7 or 9, these are the options. It is worth noting that a number of companies make multi-GHz input  $2^n$  dividers. AvanteK makes the IFD-01110, a silicon 4.5 GHz divide-by-4, Plessey makes the SP8808A, a 3.3 GHz silicon divide-by-8 and NEC makes the UPG501B, a GaAs 5 GHz divide-by-4 in addition to the UPG506B, a 14 GHz GaAs divide-by-8 to name just a few. These dividers are built with flip-flops as described previously, however, they do not have asynchronous clear inputs, thereby making them ineligible for use in the proposed divider circuit. Nevertheless, as dividers and D flip-flops with asynchronous clear inputs increase in input frequency, so will the input frequency of non- $2^n$  dividers built with this technique. If these dividers and flip-flops can be made of silicon instead of GaAs, the improvement of residual phase noise (specifi-

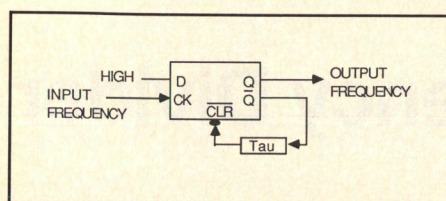
cally approximately 20 dB decrease of the  $1/f$  point) will make this technique better suited for lowest-possible phase noise synthesizers. At this time, the chosen D flip-flop—GigaBit Logic's 10G021A-2L—a GaAs device, is the highest-input frequency D flip-flop with asynchronous clear that is commonly available.

Next, the design-it-yourself option. The numerous 4-bit programmable counters readily lend themselves to  $2^n$  and non- $2^n$  frequency division. This device is available in CMOS, TTL, ECL and gallium arsenide (GaAs) circuit technologies. CMOS and TTL counters, manufactured by many companies, have maximum input frequencies at about a 100 MHz. GigaBit Logic makes a GaAs counter—the 10G061—with a 1.3 GHz maximum input frequency. This is the only 1 GHz-plus input frequency 4-bit counter currently available. However, due to propagation delays, feedback circuits inherently reduce maximum input frequency, often by a multiplicative factor of 0.60 to 0.80, compared to the same flip-flop configured for straight  $2^n$  divide ratios with no feedback.

## Concept and Analysis

The concept of this novel divider technique is based on a D flip-flop and a delay  $\tau$  fed back from the inverted Q output to the asynchronous clear (CLR) input with the D input tied high as shown in Figure 1. The maximum input frequency is limited by the maximum clock (CK) frequency of D flip-flops on the market and the divide ratio is controlled by setting the propagation delay,  $\tau$ , of the feedback circuit. The D flip-flop must have an asynchronous CLR input. The circuit analysis, including device-specific propagation delays, is most clearly visualized by referring to the timing diagram (Figure 2) of the built divider. The input frequency is 1.4 GHz, 714 picoseconds, the flip-flop is GigaBit Logic's 10G021A, and the desired divide ratio, N, is three. The cause and conse-





**Figure 1. Divider block diagram.**

quence of each change of state in either CK, Q,  $\bar{Q}$  or CLR with respect to time is marked by a number in the diagram and is explained as follows.

1) On the clock falling edge, the D input is strobed into the flip-flop. D is hard-wired logic high and the particular flip-flop is negative edge triggered. Whether the particular flip-flop used is negative or positive edge triggered is inconsequential. The state of Q prior to event 1 is irrelevant.

2) As a result of event 1, the Q and  $\bar{Q}$  outputs go high and low respectively after the CK-to-Q delay, in this case 525 picoseconds.

3) Since the  $\bar{Q}$  output is fed back via a delay  $\tau$  to the CLR input, a time delay after event 2  $\bar{Q}$  low appears at the CLR input. There are no changes in Q due to any clock falling edges that might occur between events 2 and 3 since the D input is hard-wired high and Q is already high.

4) As a result of event 3, the Q and  $\bar{Q}$  outputs go low and high respectively after the CLR to Q and  $\bar{Q}$  delay, in this case 625 picoseconds.

5) Since the CLR input is merely the delayed  $\bar{Q}$  output, a time delay  $\tau$  after event 4, the CLR port goes high (deactivated). Remember that between events 3 and 5 when the CLR input is low, the clock input has no effect on the state of either outputs.

6) Event 6 is a repeat of event 1, at that time the D input is strobed into the flip-flop. Q is low at this time.

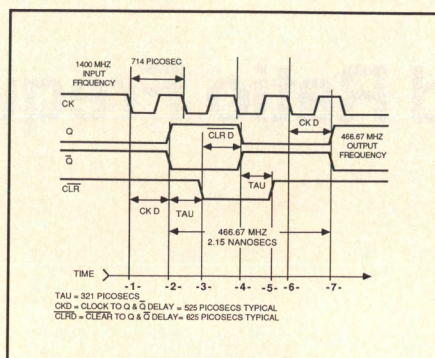
7) As a result of event 6, Q changes state and goes high after the CK-to-Q delay, thus completing one frequency division cycle.

Given the preceding explanation, one can realize the following: The value of  $\tau$  is chosen such that  $2\tau$  plus both CK-to-Q and CLR-to-Q delays is greater than N-1 input clock cycles, and less than N clock cycles where N is the desired divide ratio; this would be stated in equation form as:

$$(1 \text{ CK cycle})(N-1) < 2\tau + \text{CK-to-Q}$$

$$+ \text{CLR-to-Q} < (1 \text{ CK cycle})(N) \text{ sec}$$

The objective is to solve for  $\tau$  given a known (desired) value of N. This is the



**Figure 2. Timing diagram.**

only pertinent design equation.

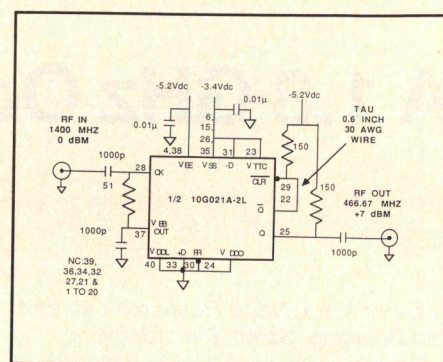
Note that Figure 2 reveals that the input to output time delay of this circuit is the D flip-flop internal CK-to-Q delay, in this case 525 picoseconds.

This technique can also be used to build a high-speed dual modulus prescaler. The only change in the circuit would be to have 2 delay lines, one for N, and the other for N + 1, connected from  $\bar{Q}$  to a 2:1 MUX input, and to connect the MUX output to the CLR input. The MUX control signal would select which delay line is in the circuit, thereby selecting the divide ratio.

## General Circuit Design Procedure

The circuit design procedure is fairly straightforward. Select the D flip-flop with an asynchronous clear input. Synchronous Q and  $\bar{Q}$  outputs are preferred but not essential. The only relevant AC parameters of the chosen flip-flop are the CK-to-Q and CLR-to-Q delays. The maximum divider (flip-flop plus required feedback) operating frequency is limited by the above mentioned delays, the flip-flop maximum input frequency and the external delay that comprise  $\tau$ . The divider maximum frequency has been measured to be about 0.6 to 0.8 times the flip-flop maximum frequency for N = 2 or 3. As expected, the maximum divider input frequency will decrease as the external delay  $\tau$  is increased; this must be considered when selecting the flip-flop.

The next step is to determine the value of the external delay,  $\tau$ , that must be introduced to achieve the desired divide ratio N. The above stated equation yields the value of  $\tau$ . How  $\tau$  is implemented depends on what it physically is. For values of N < 10,  $\tau$  can conveniently be realized passively through a coaxial cable, microstrip transmission line or as in this application, a piece of wire. The method used was to consider the passive delay line as a



**Figure 3. Schematic diagram**

characteristic impedance in series with the shunt CLR input capacitance. These two quantities determine a time constant which in turn determines how quickly the CLR input voltage reaches the threshold voltage. The design example goes through this in detail (Figure 3).

The last step is merely to determine the circuit design practicalities dictated by the already chosen flip-flop and delay line. No "special" circuits such as tuned or diode clamping circuits are required. Refer to the detailed circuit schematic and photos of the design example built (Figures 3, 4 and 5). The salient features summarized. Even though this circuit concept is digital at heart (based on and realized with an edge-triggered device) consider this an RF circuit and design accordingly. Make sure to use chip capacitors and resistors for RF line terminations, flip-flop output voltage pull downs, AC decoupling and DC blocking. Leaded DC-blocking capacitors at the output (Q port of flip-flop) have been determined to cause spurious frequencies near the carrier—the cause of death of many frequency synthesizer circuits.

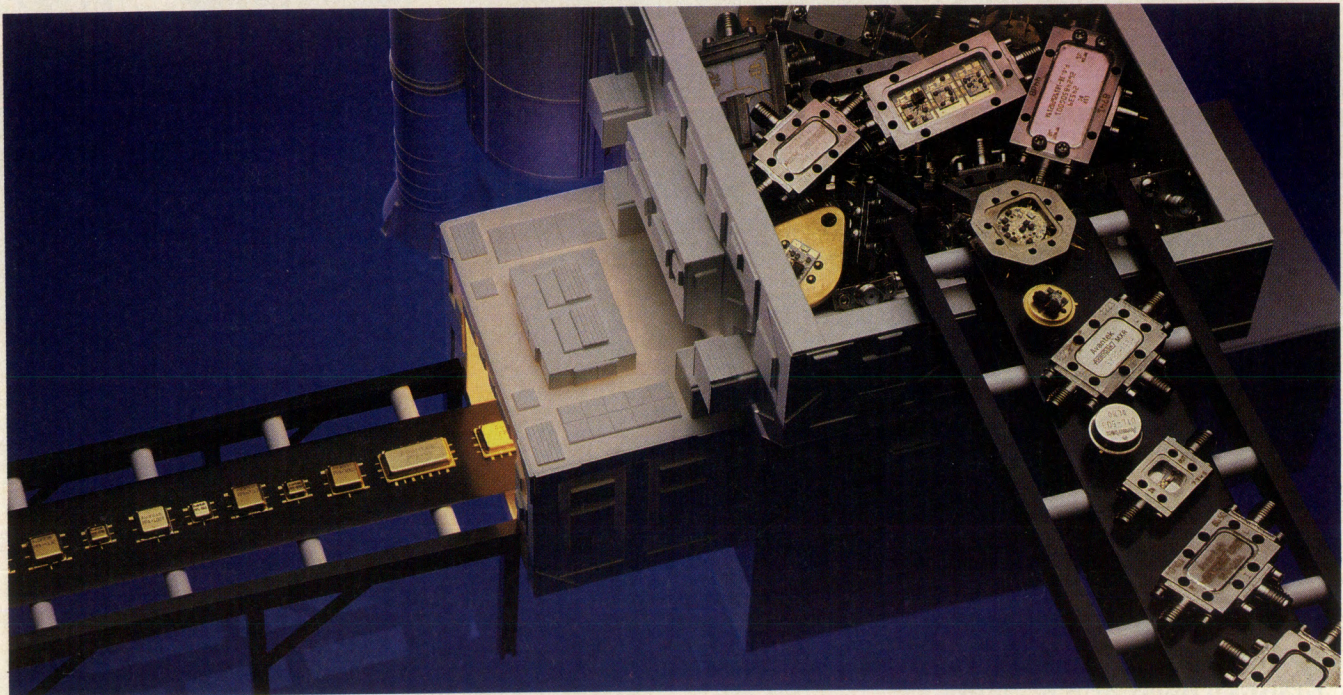
## Design Example

The system-imposed specification that necessitated this circuit dictated a frequency between 450 MHz and 550 MHz with spurious content less than -80 dBc, generated from a 1400 MHz reference. The authors conceived this circuit after confirming that dividing 1.4 GHz by three would be the best solution and that there were no pre-designed divide-by-three's available at 1.4 GHz.

The divider was designed with the GigaBit Logic 10G021A-2L in mind, a dual D flip-flop with a 2.7 GHz maximum clock frequency. Only one flip-flop of the two in the 40-pin IC was used. The required delay  $\tau$  for N = 3 is calculated using the preceding equation. 1400 MHz is 714 picoseconds, therefore the sum of all three delays must be between:



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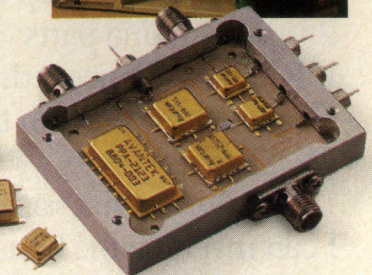
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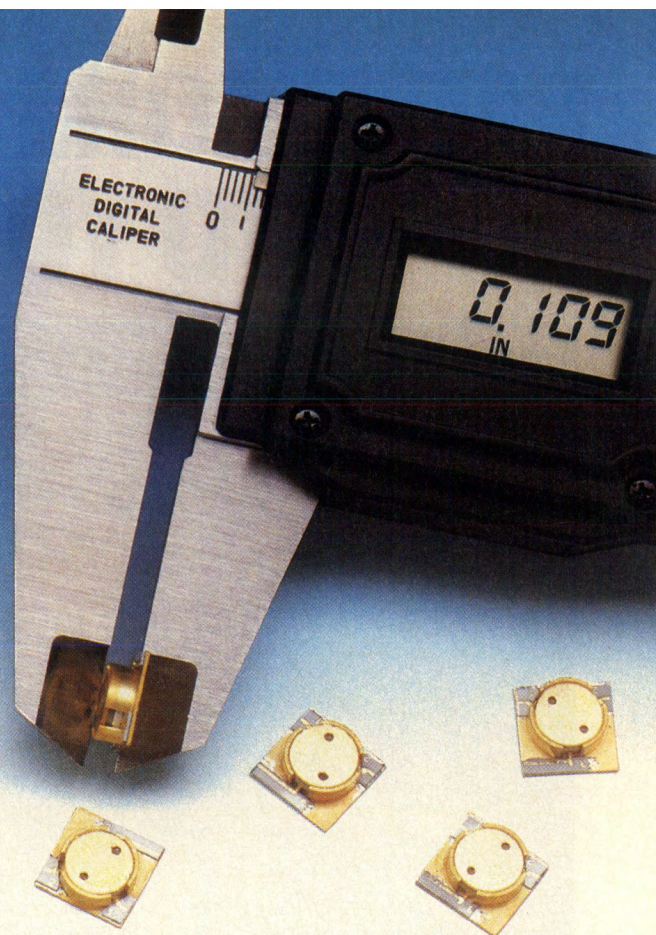
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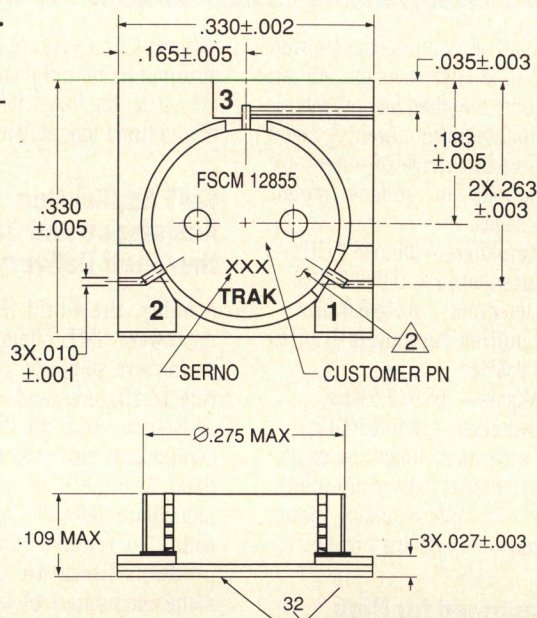
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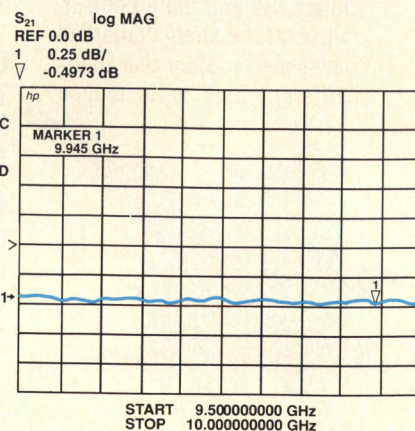
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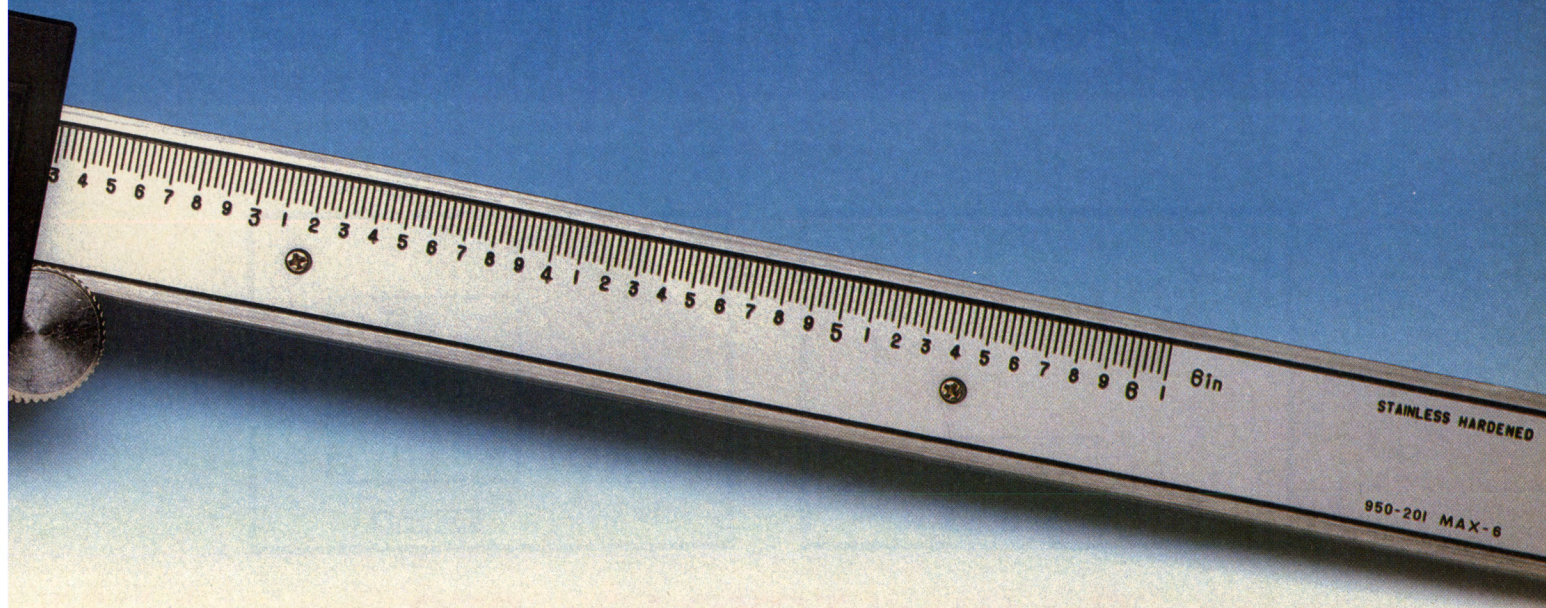
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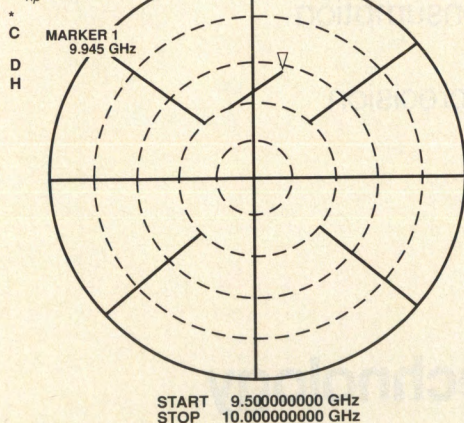






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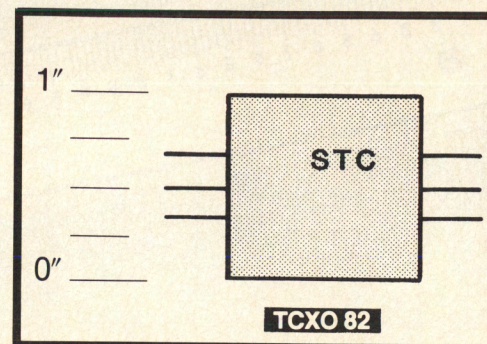
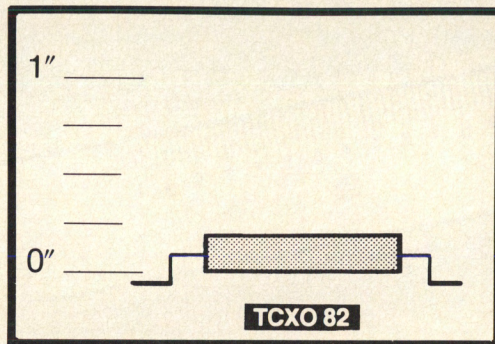
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$1428 \text{ picoseconds} < 2\tau + \text{CK-to-Q} + \text{CLR-to-Q} < 2142 \text{ picoseconds}$

clock-to-Q delay = 525 picoseconds

CLR-to-Q delay = 625 picoseconds

thus:  $140 \text{ picoseconds} < \tau < 496 \text{ picoseconds}$ .

The next step was to try to account for all delays that comprise  $\tau$ ; in order to do this a detailed description of the layout was necessitated. For extra clarity, refer to the circuit photos (Figures 4 and 5) and  $\tau$  equivalent circuit (Figure 6).

The 10G021A-2L dual D flip-flop is

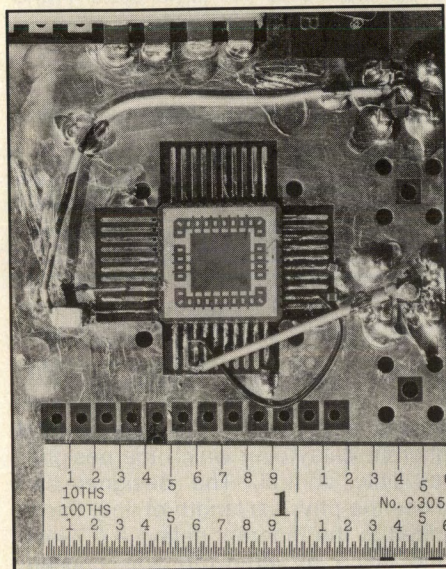


Figure 4. Photograph of top of circuit board.

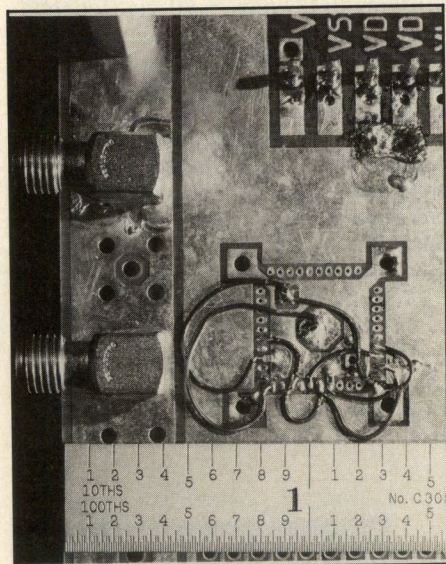


Figure 5. Photograph of bottom of circuit board.

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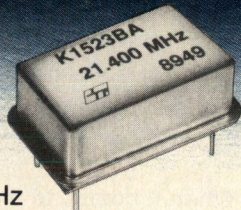
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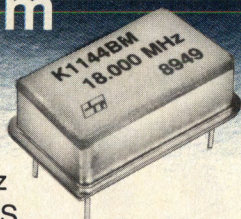


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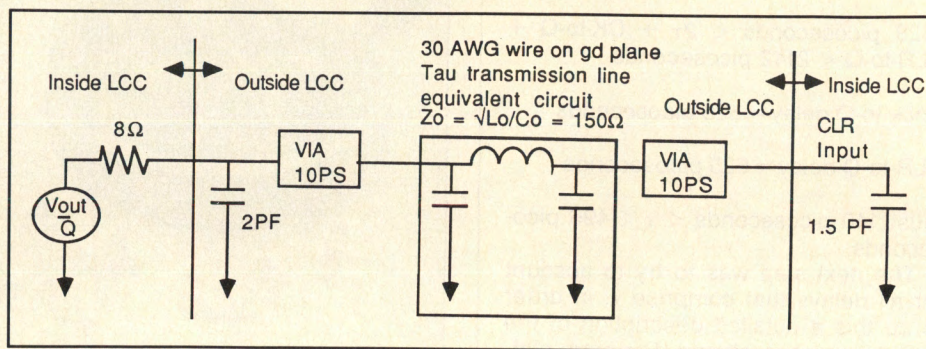


Figure 6. Circuit representation of  $\tau$ .

packaged in a 40-pin LCC that is 0.480 x 0.480 x 0.105 inches. It is directly soldered to the top of a GigaBit Logic prototype board, the 90GUPB—a four-layer, 70-mil thick board made of G-10 material. The top side of the circuit board has 200 mil long, 50 ohm microstrip lines and the bottom side has plated through vias, both corresponding to each of the 40 "pinouts". The two middle layers are used for DC voltage distribution only and are not part of the delay  $\tau$ . For the sake of soldering convenience,  $\tau$  was implemented by a piece of wire-wrap wire, 30 AWG, and was soldered on the bottom side of the board between the Q output (pin 22) and CLR input (pin 29) plated-through holes. Refer to Figures 4 and 5. The board was subsequently re-built on a 1 oz. copper double-sided 63-mil thick epoxy (FR4) circuit board with no change in performance.

First consider the inherent time delays, the significant ones being summarized as follows

1) Assume 10 picosecond delay at each end of the wire due to the path from the LCC through the via to plated-through hole on board bottom. This is 20 picoseconds of delay.

2) The 200-mil long, 50 ohm line on top of the board is not used or terminated and thus is a capacitive open stub driven by a low impedance source. Assuming 2 pF and 8 ohms:

One time constant = (8 ohms)  $(2 \times 10^{-12} \text{ F})$  = 16 picoseconds

3) The largest inherent delay is due to a combination of the shunt capacitance of the CLR input port and the transmission line that is  $\tau$ . A piece of 30 AWG wire laying flush to the board bottom has a characteristic impedance of about 150 ohms and the CLR port input capacitance is estimated at 1.5 pF. One time constant = (150 ohms)  $(1.5 \times 10^{-12} \text{ F})$  = 225 picoseconds.

The following pertains to (2) and (3). One RC time constant is 0.63 of final

voltage. Considering that the voltage levels are -0.6V for logic state 1 and -1.7 for logic state 0:

$$-0.6\text{V} - (0.63)(1.1\text{V}) = -1.29\text{V}$$

This is close to the threshold point, -1.3 V, of an ECL circuit.

Therefore, one time constant is equal to the time it takes an ECL level voltage swing to go from a logic 1 state to the threshold voltage. The CLR is active low, therefore the clear command is initiated at threshold. Summing the time delays yields:

$$20 + 16 + 225 = 261 \text{ picoseconds.}$$

This is the calculated value of total inherent delay over which the designer has little control. The designer can choose the physical length of line, this is added to the inherent delay. A 0.7-inch long piece of wire-wrap was chosen, which causes the following delay:

$$(0.7 \text{ in.}) / (39.3 \text{ in./m}) = 0.018 \text{ m}$$

$$(0.018 \text{ m}) / (3 \times 10^8 \text{ m/s}) = 60 \text{ picoseconds}$$

Therefore, the total calculated delay  $\tau$  is 261 + 60 = 321 picoseconds. This value is close to the midpoint of the calculated 140 to 496 picosecond range required for N = 3. Calculating  $\tau$  is the only design task other than choosing pull-down resistors and decoupling capacitors.

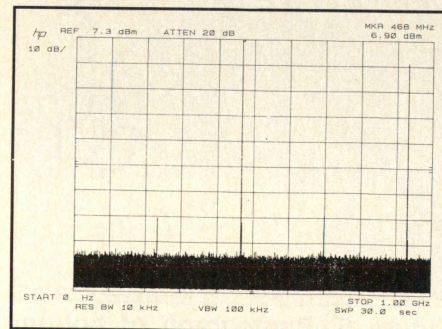


Figure 7. DC to 1 GHz spectrum.



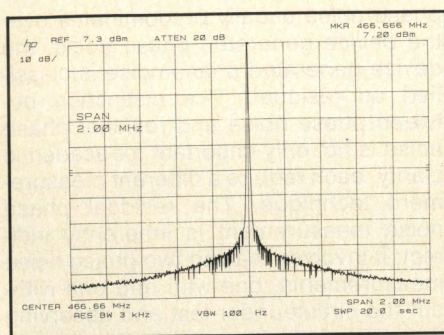


Figure 8. 466.7  $\pm$  1 MHz spectrum.

### Measured Results

From a circuit design perspective, the circuit was an instant success. The delay  $\tau$  did not need to be adjusted. The preceding design procedure was followed and the only required rework was the replacement of a leaded 330 pF chip cap with a non-leaded one functioning as the Q output DC block. The leaded capacitor induced low-level spurious oscillations close to the output frequency. The features and measured results are summarized as follows, refer to Figures 7 and 8. The 1.4 GHz source was an HP 8642B, a spectrally pure synthesized signal source, the spectrum analyzer was an HP 8566B, and the phase noise was measured on a HP 3047 system:

Input frequency (F) of interest: 1.4 GHz  
Desired divide ratio:  $N = 3$   
Input frequency range:

(for  $\tau$  fixed for  $N = 3$ )  
1210 MHz  $< F_{in} < 1810$  MHz  $N = 3$   
 $F_{in} > 1810$  MHz No output frequency  
1180 MHz  $< F < 1200$  MHz  
 $N = 5$  and 7 (unstable)  
590 MHz  $< F < 1170$  MHz  $N = 2$

Note that the maximum operating frequency, 1810 MHz, is about 0.67 of specified maximum clock frequency. Also, note that as the input frequency is decreased, the divide ratio decreases, as predicted by the design equation.

### Output harmonics:

|      |                   |         |
|------|-------------------|---------|
| 2d   | (933.33 MHz)      | -9 dBc  |
| 3d   | (1400 MHz, input) | -22 dBc |
| 4th  | (1866.67 MHz)     | -16 dBc |
| 5th  | (2333.33 MHz)     | -20 dBc |
| 19th | (8.86 GHz)        | -70 dBc |

( $c = +7$  dBm at 466.67 MHz)

Note that the 10G021A-2L puts out an ECL compatible waveform with high odd harmonic content due to the 150 picosecond rise and fall times.

Input (F) subharmonics (other than integer multiples of F/3):  
F/6 (233.33 MHz) -71 dBc

F/2 (700 MHz) -69 dBc  
No others greater than the spectrum analyzer noise floor, about -85 dBc.

### Spurious:

No others greater than the spectrum analyzer noise floor, about -85 dBc. The phase noise measurement revealed a spur -120 dBc at 30 MHz offset, however, its origin is uncertain.

### Residual Phase Noise:

In this article, residual phase noise is defined as the phase noise generated by any circuit that has a frequency input. These circuits include frequency synthesizers, dividers and multipliers, amplifiers and RF switches. The output phase noise of these devices is the logarithmic sum of the RF source input phase noise and the device generated (residual) phase noise. Since the RF source input

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phase noise usually predominates over the device generated phase noise, the device generated phase noise is classified as residual. This distinction between phase noise and residual phase noise is not only important for academic clarity, each require a different measurement technique. The residual phase noise measurement is inherently indirect; it involves making two phase noise measurements, one with and one without the circuit-under-test and observing the degradation caused by the circuit-under-test.

The residual phase noise of the divide-by-three's was surprisingly low considering the D flip-flop is made with GaAs. The phase noise with the dividers is actually a few dB lower than the measurement without the dividers. While this is theoretically impossible, it is practically possible for two reasons.

The phase detector sensitivity was only checked on the measurement without the dividers. This fact does not invalidate the results as long as the input power at the R and L ports does not change from one measurement to the next. Each divider output power was measured into 50 ohms and the RF source set accordingly for the measurement without the dividers.

The dividers directly drove the L and R inputs, there was no impedance buffer. This could result in a deceptively optimum "phase noise impedance match" at 467 MHz between the 50 ohm input 3047 mixer and the uncharacterized flip-flop output impedance. The D flip-flop is capable of driving 50 ohms.

Since phase noise was only measured in the interest of divider characterization, the authors could not devote the large amount of time rightly required for complete phase noise characterization. Presented are some observations of the measurement results (Figures 7, 8, 9, and 10).

1. Presence of 60 Hz harmonics and 20 KHz harmonics in both measurements.

2. Presence of a 30 MHz offset spurious at -120 dBc. The measurements were not made in a RF shielded room.

3. Increase in phase noise beyond 2 MHz in the measurement without the dividers. This is a discriminating effect due to a large time differential between the L and R signal paths. The differential is a result of using a line stretcher with little phase ( $\pm 2$  degrees) shift at 467 MHz; consequently, an excessive amount of cable was required to achieve phase quadrature.



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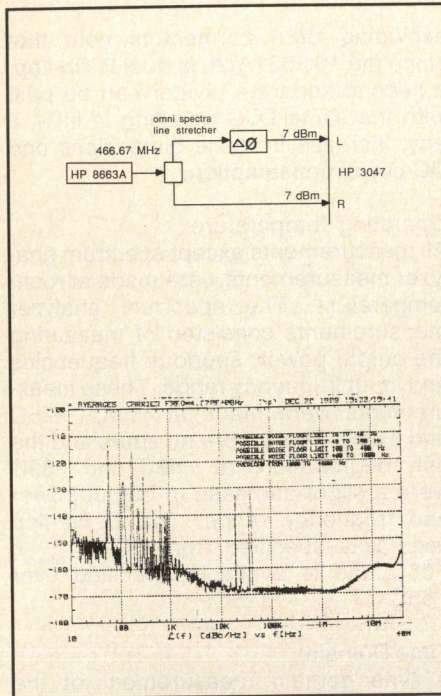


Figure 9. Measurement system setup and phase noise.

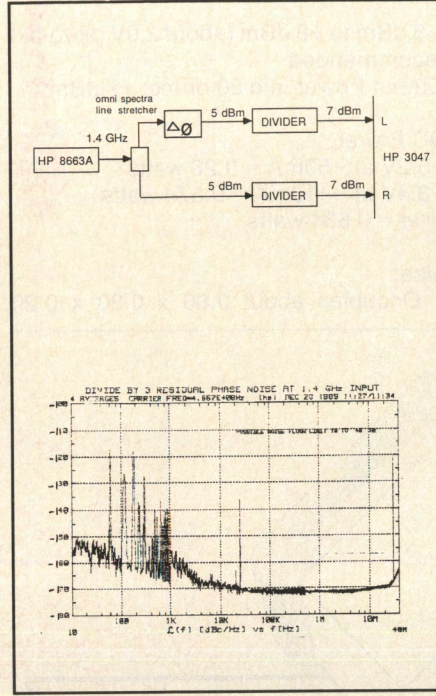
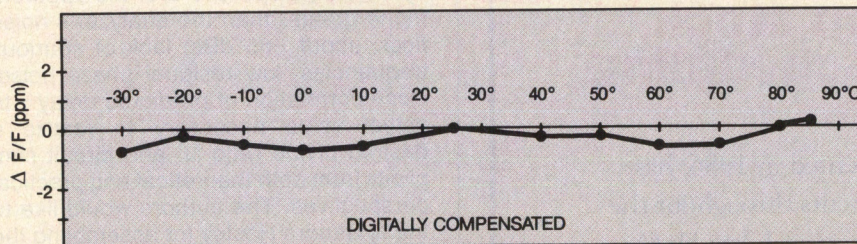


Figure 10. Measurement system with dividers setup and measurement.

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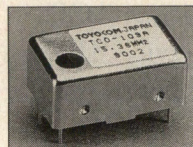


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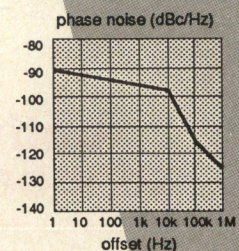
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Since the phase noise with and without the dividers is about the same, one can conclude that the divider residual phase noise at 1.4 GHz input is significantly (at least 6 dB) lower than the measurement system phase noise.

RF Input Power:  
0 dBm minimum  
+17 dBm maximum (-4.0 to +0.5V at 50 ohms is 10G021A max.)

+5 dBm to +8 dBm (about 2.0V pk-to-pk) recommended  
Output Power into 50 ohms: +7 dBm

DC Power:  
-5.2V at -50mA = 0.26 watts  
-3.4V at -110mA = 0.374 watts  
total = 0.634 watts

Size:  
Occupies about 0.80 x 0.80 x 0.20

excluding SMA connectors note that since the 10G021A-2L is dual D flip-flop, a second separate divider can be built with the same LCC resulting in little, if any, increase in these dimensions and DC power consumption.

#### Operating Temperature:

All measurements except spectrum analyzer measurements were made at room temperature. The spectrum analyzer measurements consisted of measuring the output power, spurious frequencies and input frequency range. These measurements were made at -55C, +25C and +85C. There were no spurs and the only degradations at -55C and +85C were a slight decrease in output power and frequency range. The D flip-flop used was specified from  $0C < T_{case} < +85C$  and is available specified over  $-55C < T_{case} < +125C$ .

#### Time Domain:

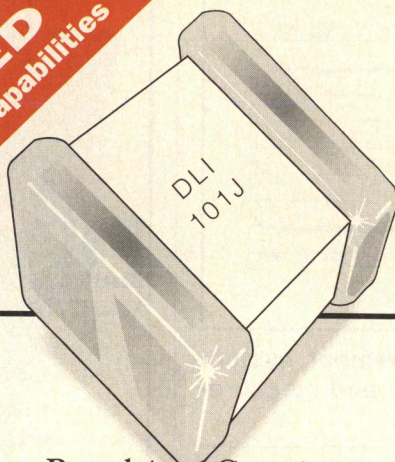
A time domain measurement of the output using the HP 54111D digitizing oscilloscope revealed a periodic non-sinusoidal waveform that had the characteristics of a wave rich in harmonics with a 3.4V peak-to-peak at 1 Mohm impedance and 1.4V peak-to-peak at 50 ohm impedance. It is omitted from this paper since it reveals very little useful information.

In summary, the most attractive features of this circuit from a frequency synthesis perspective are: the apparent (none greater than the analyzer's noise floor, about -85 dBc) lack of spurious frequencies, low residual phase noise, close correlation of predicted delay  $\tau$  to actual value necessary to realize a desired divide ratio N, and circuit simplicity from both theoretical and practical perspectives. The authors would like to thank Sandra Hosley for assembling the circuit, Ike Mathis for generating the drawings and Bill Webster and Jerry Phillips for general support and encouragement of this effort.

#### About the Authors

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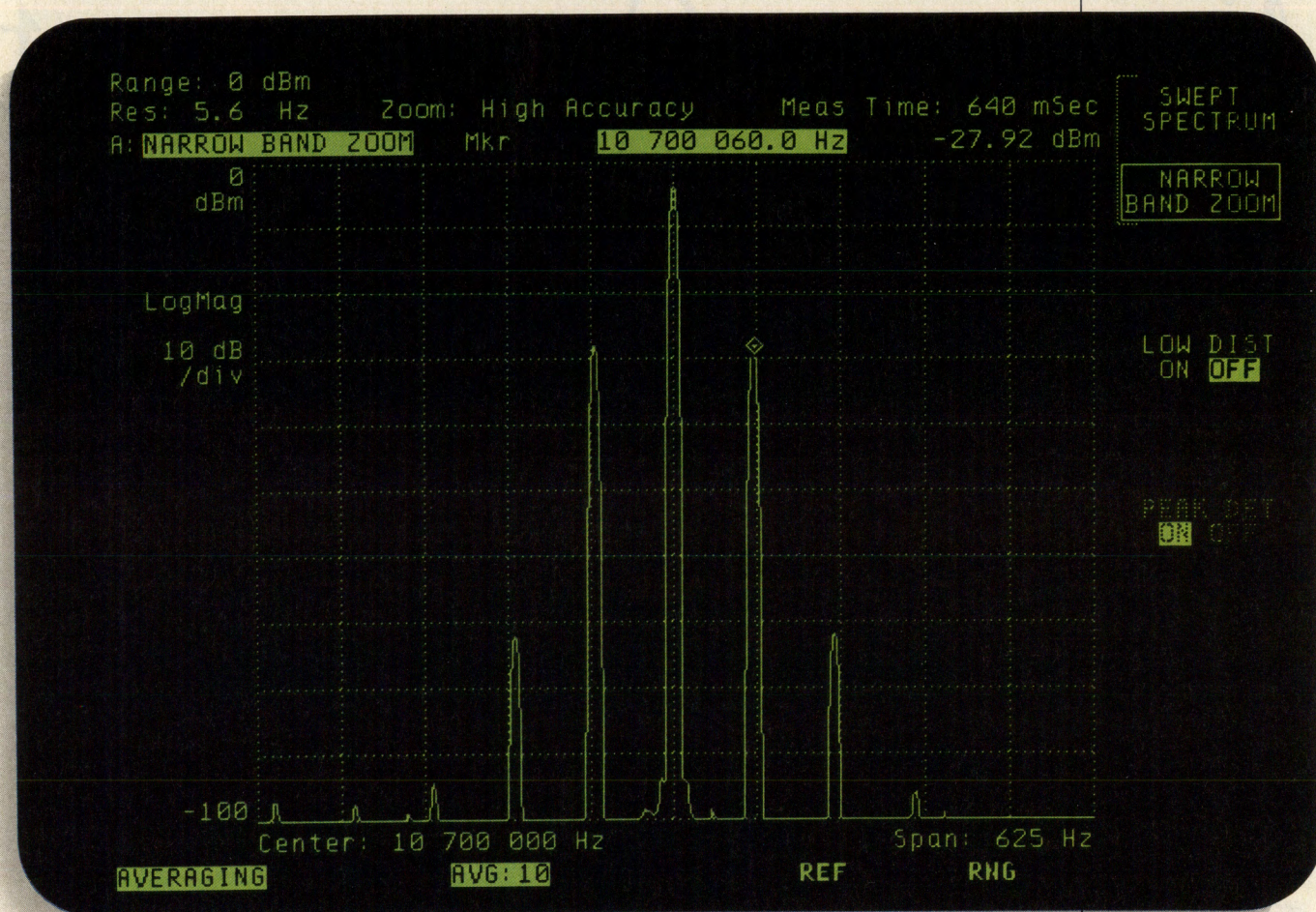


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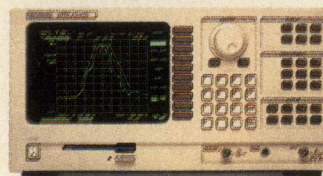
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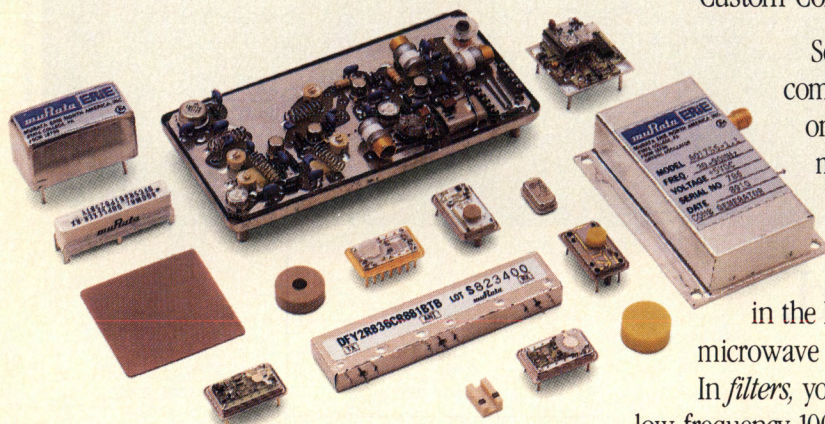
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# Modeling Discrete-Time Phase Detection in the Phase Locked Loop

By C.W. Price, II  
IBM Corporation

The Phase Locked Loop, or PLL, is a feedback control system that is widely applied in the design of electronic communications and controls apparatus. In the PLL, the phase of a local oscillator is matched to that of an external reference making it especially attractive for tracking purposes. Other applications include frequency synthesis, demodulation, timing extraction, and mechanism control.

The majority of phase locked loop analysis is performed using Laplace transform techniques. The Laplace transform, which maps the time response of a system into the complex frequency domain, reduces the problem of solving linear differential equations to that of mere algebraic manipulation.

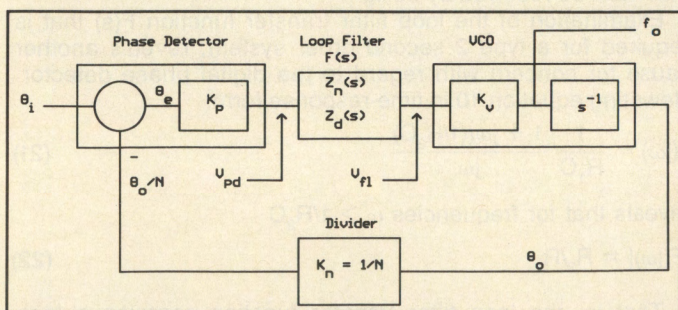
The PLL using a digital, or discrete-time, phase detector can not be directly evaluated using this technique. Behavioral approximations to continuous-time phase detection can be used that allow the Laplace analysis to proceed. In many designs, however, the consequences of digital phase detection have a strong bearing on the design outcome and must be evaluated by computer simulation and laboratory measurement.

In this report, the phase error transient of a type 2 second order phase locked loop is derived using the Laplace transform technique. The classical phase error response forms for the simple phase and frequency steps are presented. The contrasting transient response and design complications imposed by the digital phase detector are discussed.

A computer simulation of loop operation is presented as a general purpose tool in the digital PLL design process. By way of example, the simulation model is used to determine the effectiveness of a pulse suppression filter.

## PLL Fundamentals

Figure 1 defines the terms which are used in this text. The phase detector output  $V_{pd}$ , which reflects the phase error of the loop ( $\theta_e = \theta_i - \theta_o/N$ ) is processed by loop filter  $F(s)$ . The filter output voltage ( $V_{fl}$ ) controls the frequency of a Voltage Controlled Oscillator (VCO). A frequency divider ( $K_n = 1/N$ ) is connected between the VCO and the phase detector, thereby completing the loop.



**Figure 1. Phase locked loop diagram.**

In this report a phase detector is assumed with a voltage output that is linear with phase error, that is:

$$V_{pd} = K_p(\theta_i - \theta_o/N) \quad (1)$$

$K_p$  [volts/radian]

In this case, loop filter  $F(s)$  is an impedance ratio,  $Z_n(s)/Z_d(s)$ . Various loop configurations can be constructed by selecting suitable forms for  $F(s)$ .

The VCO output frequency ( $f_o$ ) is determined by the filter voltage  $V_{fl}$  and a constant  $K_v$ , also called the VCO conversion gain.

$$f_o = K_v(V_{fl}) \text{ [radians/sec]} \quad (2)$$

$K_v$  [radians/volt/second]

## System Type and Order

$G(s)$  and  $H(s)$  are the loop feedforward and feedback transfer functions respectively. The terminology "system-type" and "system-order" are often used to express the general form of the loop transfer function  $G(s)H(s)$ . System type is defined as the number of poles of the transfer function,  $G(s)H(s)$ , located at the origin ( $s=0$ ) and system order is defined as the highest degree of the polynomial:

$$1 + G(s)H(s) \quad (3)$$

Polynomial  $1+G(s)H(s)$  is called the Characteristic Equation of the loop and it contains the closed loop poles of the system.

In this text the behavior of the type 2 second order loop is considered. The interested reader is referred to the literature for discussions of other loop configurations (3).  $G(s)H(s)$  of the type 2 second order loop has the following form:

$$G(s)H(s) = \frac{K(s+a)}{s^2} \quad (\text{type 2}) \quad (4)$$

$$1 + G(s)H(s) = s^2 + Ks + Ka = s^2 + 2\zeta\omega_n s + \omega_n^2 \quad (\text{order 2}) \quad (5)$$

The rightmost expression in equation 5 is recognizable as the normalized second order form. From Figure 1, the feedforward and feedback transfer functions are:

$$G(s) = \frac{K_p F(s) K_v}{s} \quad (6)$$

$$H(s) = K_n = 1/N \quad (7)$$

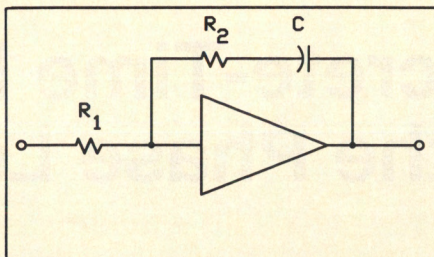
Substituting for  $G(s)$  and  $H(s)$  in equation 4, the transfer function for the loop filter is found.

$$\frac{K_p F(s) K_v}{Ns} = G(s)H(s) = \frac{K(s+a)}{s^2} \quad (8)$$

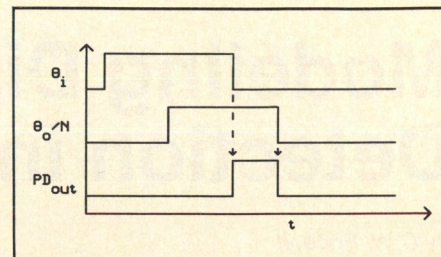


|                | Type 1   | Type 2   | Type 3 |
|----------------|----------|----------|--------|
| Phase Step     | zero     | zero     | zero   |
| Frequency Step | constant | zero     | zero   |
| Frequency Ramp | ramp     | constant | zero   |

**Table 1. Steady state phase errors.**



**Figure 2. Type 2 second order filter.**



**Figure 3. Digital phase detector operation.**

$$F(s) = \frac{K'(s + a)}{s} \quad (9)$$

The circuit of Figure 2 yields the appropriate form for  $F(s)$ .

$$F(s) = \frac{R_2(s + 1/R_2C)}{R_1s} \quad (10)$$

## Loop Response

The response of the type 2 second order system to the step phase and frequency inputs are evaluated in this section using the Laplace transform. Applying control theory to Figure 1 the following relationships are obtained.

$$\theta_e = \theta_i - \theta_o/N \quad (11)$$

$$\theta_e(s) = \theta_i(s)/(1 + G(s)H(s)) \quad (12)$$

Phase error is determined by substitution for the loop input  $\theta_i(s)$ . The final value theorem, equation 13, is applied to the resulting expression for  $\theta_e(s)$  to determine the steady state phase error.

$$\lim_{t \rightarrow \infty} [\theta_e(t)] = \lim_{s \rightarrow 0} [s\theta_e(s)] \quad (13)$$

## Step Phase

$\phi$  radian step phase applied at  $t=0$ :

$$\theta_i(s) = \phi/s \quad (14)$$

$$\theta_e(s) = \frac{\phi s}{s^2 + 2\xi\omega_n s + \omega_n^2} \quad (15)$$

$$\lim_{s \rightarrow 0} [s\theta_e(s)] = 0$$

taking the inverse transformation of equation 15,

$$\theta_e(t) = \frac{\phi e^{-\xi\omega_n t}}{\omega_n \sqrt{1-\xi^2}} [\omega_n \sqrt{1-\xi^2} \cos \omega_n \sqrt{1-\xi^2} t - (\xi\omega_n) \sin \omega_n \sqrt{1-\xi^2} t] \quad (16)$$

## Step Frequency

$\Omega$  rad/sec step frequency applied at  $t=0$

$$\theta_i(s) = \Omega/s^2 \quad (17)$$

$$\theta_e(s) = \frac{\Omega}{s^2 + 2\xi\omega_n s + \omega_n^2} \quad (18)$$

$$\lim_{s \rightarrow 0} [s\theta_e(s)] = 0$$

taking the inverse transformation of equation 18

$$\theta_e(t) = \frac{\Omega e^{-\xi\omega_n t}}{\omega_n \sqrt{1-\xi^2}} \sin \omega_n \sqrt{1-\xi^2} t \quad (19)$$

Steady state phase error for the type 1 and type 3 systems appear in table 1 along with results obtained for a frequency ramp input  $\theta_i(s) = \psi/s^3$ . A similar derivation for the sinusoidal phase detector is contained in the literature (1,3).

## Digital Phase Detection

The transient response forms of the previous section assume a phase detector output that is time-wise continuous. The digital detector output does not meet this description and, although the steady state error is in agreement with Table 1, the error transient of the digital loop is quite different from that derived above.

Phase error in the digital detector is determined from transitions in the detector inputs and is typically represented by the duty cycle of one or more rectangular pulse trains at the phase detector output. (see Figure 3) The period of the pulse train is dictated by the transition rate of the phase detector reference input:

$$f_i = d\theta_i(t)/dt \quad (\text{radians/second}) \quad (20)$$

Exact spectral analysis of the digital phase detector output is quite complex. Suffice to say that there is a strong dependency on the transition rate of the phase detector reference input and a general behavior that follows a  $(\sin x)/x$  characteristic. This raises the concern that energy at the reference frequency passing through the loop filter will frequency modulate the VCO giving rise to undesirable spectral sidebands at its output. This is especially troublesome in the design of phase locked loop synthesizers for communications equipment where spectral purity is closely regulated.

If the reference frequency falls within the bandwidth of the loop it is impossible to configure a filter that will remove this unwanted energy without affecting the characteristics of the loop. A fast loop lock-up time, which is achieved by increasing system bandwidth, and fine frequency resolution, which requires a low reference frequency are inconsistent objectives that must be resolved by design.

Examination of the loop filter transfer function  $F(s)$  that is required for a type 2 second order system, reveals another cause for concern with regard to the digital phase detector. Rewriting equation 10 in time-response form:

$$F(j\omega) = \frac{1}{R_1 C} \frac{1 + j\omega/(1/R_2 C)}{j\omega} \quad (21)$$

reveals that for frequencies  $\omega > 1/R_2 C$

$$|F(j\omega)| \approx R_2/R_1 \quad (22)$$

That is, the loop filter transfer function contains a high frequency gain term  $(R_2/R_1)$  for all frequencies beyond the



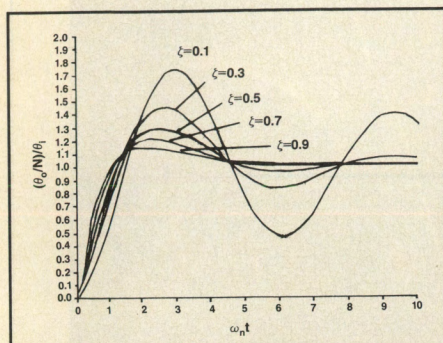


Figure 4. Normalized second order responses.

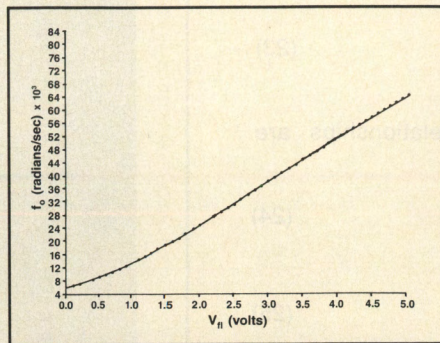


Figure 5. VCM V-F characteristic.

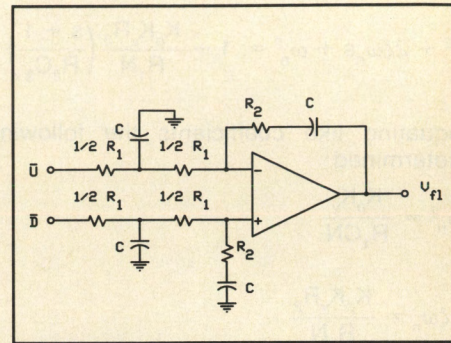


Figure 6. Filter for use with digital phase detector.

$1/R_2C$  cutoff of the integrator section. In most cases the power spectrum of the phase detector output will exhibit non-zero signal energy for this frequency range. If  $R_2/R_1$  is large, pulsewise saturation of the buffer amplifier can result. Clipping of  $V_H$ , which can be translated into a reduction in loop gain, can lead to increased loop settling time and overshoot.

A device commonly used in the design of digital PLLs is the 4044 phase-frequency detector. The 4044 consists of a pair of phase detection circuits, one of which contains a special logic that ensures zero-frequency error at lock. It differs from the classical "phase-only" detector which will allow the loop to lock at a frequency ( $f_o/N$ ) that is harmonically related to the reference. Desirable in some instances, harmonic locking is a topic of considerable interest (1,3).

The 4044 generates rectangular pulses U and D on negative transitions at either its reference ( $R=f_i$ ) or feedback ( $V=f_o/N$ ) inputs. The width of the U and D outputs are proportional to the R-V phase/frequency error. The negative active signal at U is asserted by the 4044 when the feedback phase  $V = \theta_o/N$  lags that of the reference  $R = \theta_i$ , causing the VCO control voltage to increase. The 4044 D output is activated for the case of leading phase. At lock, narrow pulses are generated at U and D that maintain the VCO control voltage. A detailed operating description for leading phase, lagging phase, and frequency error is contained in the manufacturer's specification (6).

### The DPLL Simulator

The design of any system for which there are no established analytic tools is precarious at best. Complete dependence on empirical techniques often results in a system that is over-designed and/or poorly characterized. This, in many instances, is the case of the digital phase locked loop.

In many cases the digital loop can be safely analyzed by making approximations to the continuous-time system. The predictions of the Laplace method can then be compared with hardware measurements to assess the need for a more detailed analysis. In cases where the Laplace analysis is not sufficient to complete the design process then the consequences of digital phase detection must be analyzed directly. This is especially true for systems in which the spectral purity of the VCO is important.

In an effort to reduce the risk associated with the design of the digital PLL a computer model was developed to address the idiosyncrasies of discrete-time phase detection. The simulator includes descriptions of a 4044 phase-frequency detector, loop filter, voltage controlled multivibrator, and frequency divider. With the simulator, signal generators of

arbitrary complexity can be defined and applied to the PLL and the output response can be observed. The simulator is written in the Dynamic Simulation Language, or DSL, a VS Fortran based IBM programming product.

The simulator was constructed by breaking the PLL into a number of interconnected subsystems according to Figure 1. Each of the subsystems was modeled separately and completely characterized prior to incorporation into the larger model.

The modular design of the simulator is receptive to subsystem replacement making it easy to experiment with new design concepts prior to committing hardware. For instance, by substituting an analog phase detector for the 4044 the effects on sideband suppression can be evaluated. Likewise, a sinusoidal oscillator can be constructed to replace the voltage controlled multivibrator, higher and lower order loop filters can be configured, etc.

An advantage over empirical methods, signals of arbitrary order can be modelled during simulation. Static phase offset, amplifier leakage current, offset bias and numerous other parameters can be controlled as well. Following a general discussion of component selection for the second-order loop, the simulator is applied in the design of a pulse suppression filter for the digital phase detector.

### Selecting Filter Components for the Second Order Loop

The process of selecting loop filter components for the digital PLL begins by specifying the maximum overshoot and settling time of the loop. Normalized response curves, Figure 4, are then used to translate these requirements into the second order parameters for loop damping ( $\zeta$ ) and undamped natural frequency ( $\omega_n$ ).

Component values are found by equating the loop transfer function with the normalized second order form:

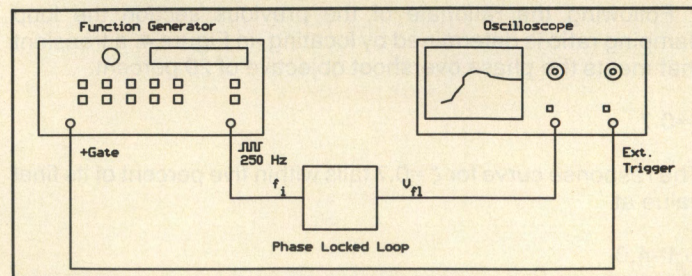


Figure 7. Laboratory measurement setup.



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$$s^2 + 2\xi\omega_n s + \omega_n^2 = 1 + \frac{K_p K_v R_2}{R_1 N} \left( \frac{s+1}{R_2 C_2} \right) \quad (23)$$

equating like coefficients the following relationships are determined

$$\omega_n^2 = \frac{K_p K_v}{R_1 C N} \quad (24)$$

$$2\xi\omega_n = \frac{K_p K_v R_2}{R_1 N} \quad (25)$$

The discrete-time phase detector must be approximated to continuous-time behavior ( $K_p$ ), to apply the results of equation 22.  $K_p$  is given by the average differential phase detector output voltage for a 1 radian phase error,

$$K_p \approx \frac{(V_{hi} - V_{lo})}{2\pi} \text{ [volts/radian]} \quad (26)$$

In the case of the 4044,  $V_{hi}$  is the inactive level and  $V_{lo}$  is the active level for either the U or D output. If U and D are not equal an average level should be taken.

$K_v$  is given by the slope of the VCO voltage-frequency gain characteristic at the nominal operating voltage of  $V_{fl}$ . With this information the selection of loop filter components can be completed.

## Design Example

The procedure outlined in the previous section is illustrated in the following example. A type 2 second order loop with the following characteristics is considered.

$$f_i = d\theta_i(t)/dt = 250 \text{ Hz} \quad (1571 \text{ radians/second})$$

$$K_n = 1/N = 1/20 \quad (f_o = 5.0 \text{ kHz})$$

overshoot  $\leq 20$  percent

$$t_s = 0.2 \text{ seconds} \quad (5 \text{ percent settling time})$$

$$V_{fl} = 2.5 \text{ volts}_{\text{nominal}}$$

The 4044 detector is approximate to the continuous-time detector by:

$$K_p \approx 0.1 \text{ volts/radian}$$

The VCO V-F characteristic is measured in the laboratory and shown in Figure 5.

$$K_v = dF/dV \approx 10^4 \text{ radians/volt/second @ } V_{fl} = 2.5 \text{ volts}$$

Following the rationale of the previous section the loop damping ratio is determined by locating, in Figure 4, a transient that meets the phase overshoot objective of 20 percent.

$$\xi = 0.7$$

The response curve for  $\xi = 0.7$  falls within five percent of its final value at

$$\omega_n t \approx 4.0$$

$$\omega_n = 4/t_s = 4/0.2 = 20 \text{ radians/second} \quad (27)$$

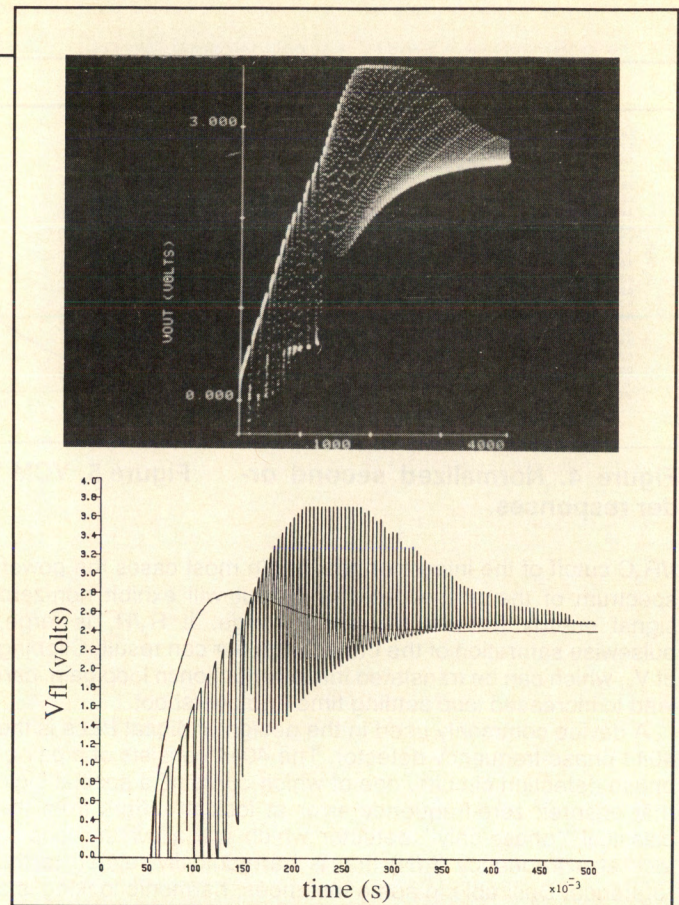


Figure 8. Pulse suppression cutoff at 2000 rad/sec.

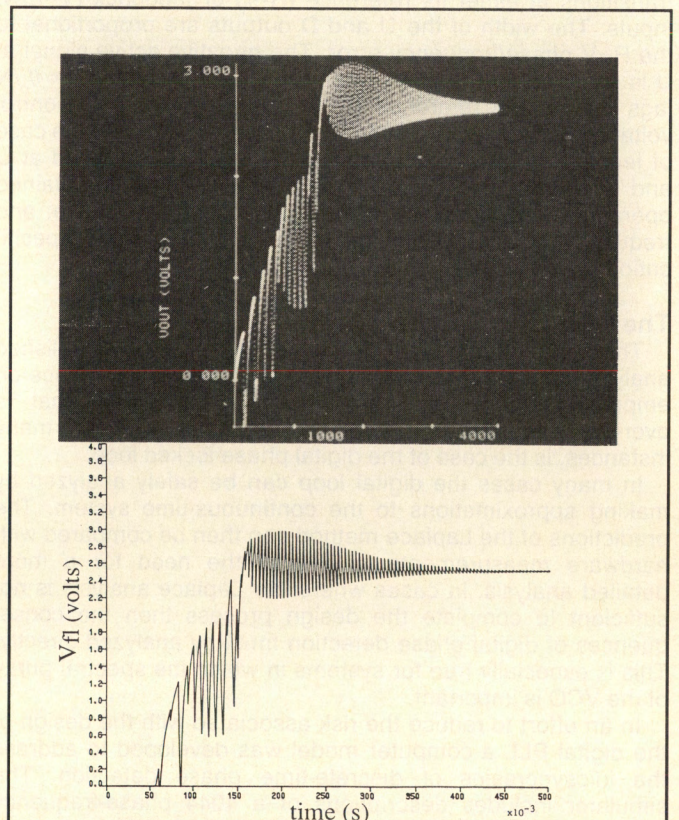


Figure 9. Pulse suppression cutoff at 400 rad/sec.

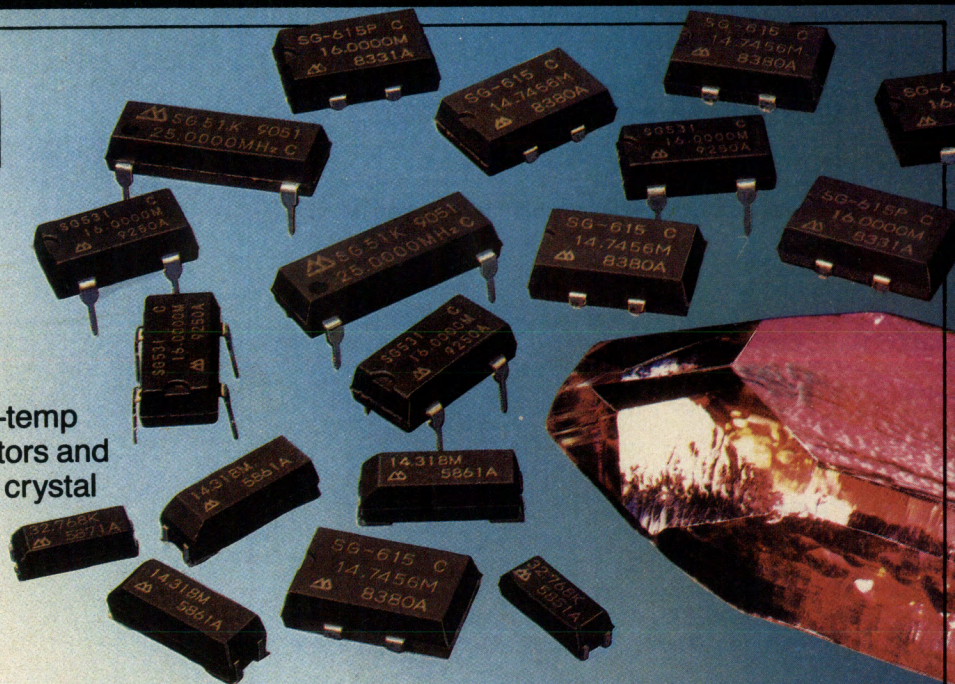


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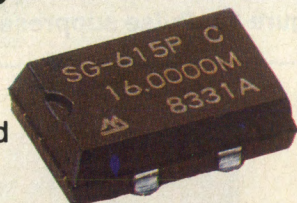
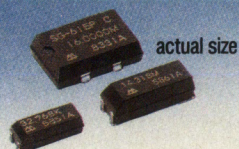
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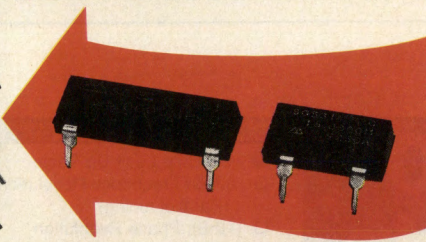
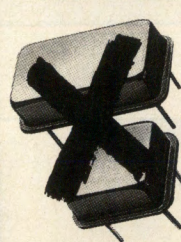
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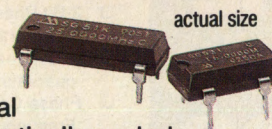
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Filter components are selected by rearranging equation 24,

$$R_1 C = \frac{K_p K_v}{N \omega_n^2} = \frac{(10^4)(0.1)}{(20)(20^2)} = 0.125 \quad (28)$$

Letting  $C=10\mu F$ , then,

$$R_1 = 12.5 k\Omega$$

$$R_2 = 2 \xi / C \omega_n = 7 k\Omega$$

### Locating a Pulse Suppression Filter

The potential for pulsewise saturation of the buffer amplifier and feedthrough of sampling energy can be reduced by adding a simple RC low-pass section to the loop filter as shown in Figure 6. The cutoff frequency of this pulse suppression filter is given by

$$\omega_c = 4/R_1 C^* \text{ [radians/second]} \quad (29)$$

The procedure for locating the cutoff frequency of the pulse suppression filter is not formalized and will be determined iteratively by using the DPLL simulator previously described.

Prototype hardware was constructed as a benchmark for simulator evaluation. To increase the likelihood for correlation the initial conditions for both the hardware and simulator must

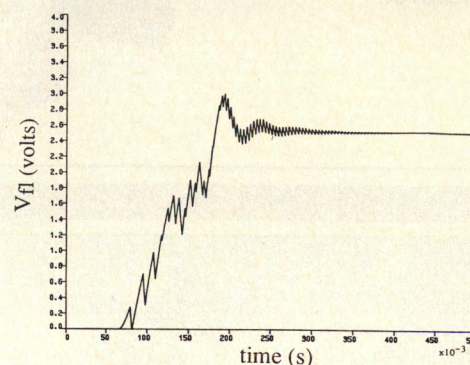
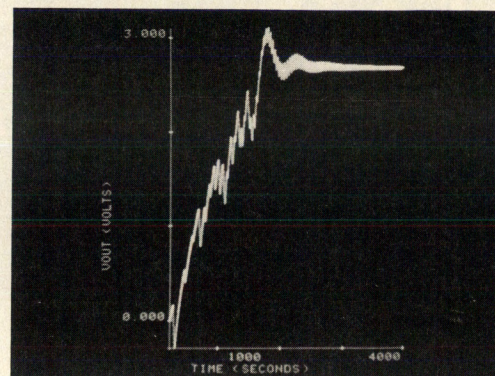


Figure 10. Pulse suppression cutoff at 100 rad/sec.

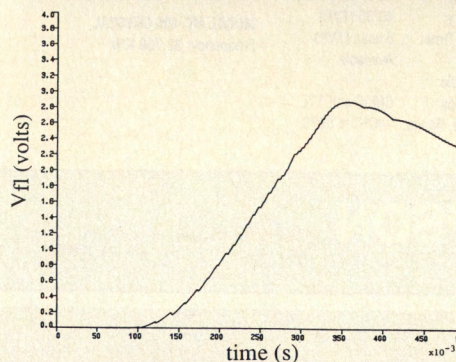
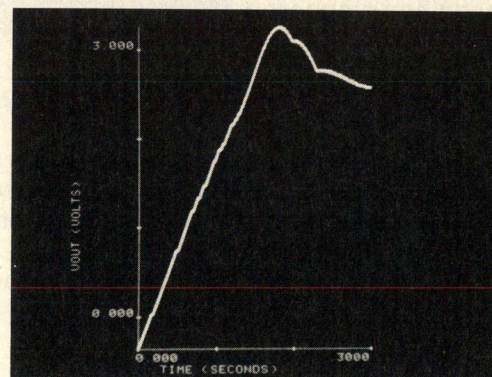
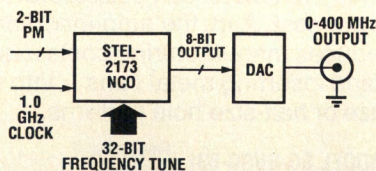


Figure 11. Pulse suppression cutoff at 10 rad/sec.

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be the same. The following procedure was used to calibrate the experiment.

The start-up state of the loop is established by holding the reference ( $f_i$ ) at a logical "1" level while allowing the feedback signal ( $f_o/N$ ) to cycle the phase detector. The reference frequency being zero in this case the 4044 will activate its D output attempting to lower the VCO frequency. As expected from the V-F curve for the VCO, the loop filter voltage is driven to zero volts.

The initial conditions of the loop having been established, the 250 Hz square wave is gated to the reference input. As  $f_i$  now exceeds the feedback frequency  $f_o/N$  the 4044 activates U in an effort to increase the loop filter voltage and correct the frequency error. The loop filter voltage  $V_{fi}$  is an easily accessible point and a good barometer of loop behavior and is monitored in hardware and simulation.

## Experimental Results

The results of locating the pulse suppression filter at  $\omega_c$  equal to 100, 20, 5 and 0.5  $\omega_n$  are contained in Figures 8-11. The transient response of the Laplace method is superimposed in Figure 8 for reference.

In Figure 8, the effects of locating  $\omega_c$  beyond the reference frequency ( $\omega_c > 1571$  rad/sec) is apparent. During a portion of the transient time the buffer amplifier is driven into saturation. An accompanying reduction in loop gain is evidenced in the delayed settling time as compared with predictions made by the Laplace method.

The pulse compensation pole is relocated to  $\omega_c = 400$  rad/sec. The results of Figure 9 demonstrate the effectiveness of the filter in eliminating the saturation problem. The overall transient resembles that of a second order system.

Placing the pole at  $\omega_c = 100$  radians/second shows further improvement in suppressing the effect of the phase detector pulses in the loop filter output. The response is still representative of second order behavior and perhaps is a good choice for locating the filter.

Figure 11 is included for completeness in demonstrating the effect of overdesigning the pulse compensation filter. In this case  $\omega_c = 10$  rad/sec ( $0.5\omega_n$ ) and the response curve departs radically from the second order characteristic. The system is perhaps unstable as well and the smoothness of  $V_{fi}$  should be an indication of underlying problems.

## Conclusion

The contrasting behavior of digital phase detectors when compared to Laplace transform predictions has been illustrated. The Laplace transform, which assumes continuous-time phase detection, is not suited to the evaluation of designs using digital phase detection, although approximations to continuous-time behavior are justified in some instances. In other applications, however, the consequences of digital phase detection have a direct bearing on design outcome and must be evaluated directly.

In particular, reference energy feedthrough is a concern for which there are no established analytical methods for resolution. In cases such as this an inordinate amount of risk is assumed due to the empirical basis of the design. To restore balance to the design process a computer simulator has been developed as a general purpose tool for phase locked loop analysis. In this report the computer simulator was used to demonstrate the effect of a pulse compensation network on the operation of a second-order loop.

Simulator predictions, which demonstrate remarkable correlation with actual hardware measurements, were used to locate the cutoff frequency of a simple RC pulse suppression network, the extremes of which exhibit the effects of pulse saturation and response degradation as predicted.

With this new tool the inherent risk of compromise analysis is averted without resorting to the cut-and-try method. Parameterization of the loop, which begins by estimating component values from classical Laplace derivations, can be concluded with good confidence in the expectations of system performance.  $\square$

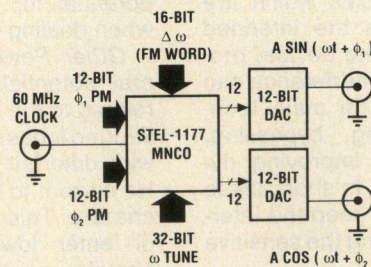
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## About the Author

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# A Review of Common RFI Sources

By Gary A. Breed  
Editor

*This month's column is a review of the RFI hazards that electronic devices face in their operating environment. Some of these interference sources can be dealt with in the design of a product, but others represent either malfunctions or electrical phenomena that can only be corrected at the source.*

Unforeseen problems often arise when electronic equipment is put into use by the customer. In the case of RF products, the potential for problems due to interference is considerable because they deal with electromagnetic energy. Unintentional, potentially disruptive sources of electromagnetic energy are plentiful in today's homes and businesses. Some of the common RFI sources are listed here, along with suggestions for identifying and dealing with them.

**Licensed Radio Facilities.** The most common source of RFI is nearby radio operation, usually the result of either front-end overload, or intrusion into the IF or audio stages. Broadcast transmitting facilities, two-way radio users, amateur and citizen band operators, aviation communications and navigation stations, and even low-power devices like cordless telephones and remote control units can cause interference — not because of improper operation, but simply because of their close proximity.

Since this type of interference involves discrete frequencies which are often far removed from the intended operating frequency of the device, protection from this kind of interference can often be designed-in. The main techniques involve shielding, bypassing, bandpass filtering, and improving dynamic range capabilities. In short, these techniques are meant to keep the interfering signals from entering the sensitive portions of the circuitry.

The principle caution is that these radio facilities operate from kHz through UHF, and the RFI suppression techniques must allow for this fact. For example, effective bypassing at UHF may do nothing to reduce AM broadcast interference. Or, the configuration of a simple front-end filter might allow frequencies far above or below the design frequency to pass unattenuated. Multiple values of bypass capacitors or RF

chokes may be required.

**Electrical Power Equipment.** Whenever a substantial amount of power is drawn in an electrical device, the  $di/dt$  or  $dv/dt$  swing can generate substantial electromagnetic fields. These fields may contain RF fundamental or, more likely, harmonic energy that can cause interference. Adjacent nonlinear devices (diodes or transistors) may be excited by these fields to generate additional energy in the RF spectrum. If a designer has to deal with both RF circuitry and power devices in a product, attention is necessary to avoid self-interference.

A more significant problem lies in the area of minor malfunctions. Power devices like motors, lighting, thermostats, heaters, etc., may continue to operate despite problems with corroded contacts, shorted windings, worn connections, or other partial failures. Many of these problems cause an arc to develop, creating an excellent wideband noise source. The principal characteristic of this kind of interference is modulation at 60 Hz (or 50 Hz) and harmonics.

Some notorious interference sources in the power category include SCR or triac power controllers, bimetal thermostats or thermal protection devices, standard fluorescent lighting, high frequency inverter driven lighting, and power tools of all types. Service personnel need to be fully aware of the potential for interference of this type when dealing with customer complaints.

**Other Power-Related Sources.** Several potential areas for problems are related to the primary AC power source. Ground loops (multiple circuits to ground with different potential drops) allow 60 Hz current to flow through cabinets and chassis. This can modulate RF signals or enter low level audio or control circuitry, usually with a good sine wave. This is typically a customer-created problem. Internal single-point grounding can minimize the effects of ground loops.

Corona discharge and arcing across installation hardware are typical problems that can occur on power transmission lines. Corona noise can be very wideband, while arcing will be like the power-related noise described earlier. Both can be very frequency-specific,

"tuned" by the physical dimensions of the installation. Precise identification of the location of a power line interference source can be difficult, since signals can propagate along the wires and can be coupled from one circuit to another.

**Unintentional Radiators of RF Energy.** Many products generate RF energy in their normal operation — all radio and video equipment, computers and other digital devices, CATV systems, and unlicensed radio control and communications devices. These are all subject to FCC regulations regarding interference, but even if in compliance when manufactured, other circumstances may compromise their performance and cause interference.

Damage, tampering or modification of equipment is the main culprit. For example, illegal cable television connections can cause significant radiation in the VHF spectrum used for transmission. Also, the cable system is subject to increased outside interference since its shielding has been compromised. It is also possible for extended wear or weather damage to result in CATV problems.

Removing cabinets from digital, radio, or video equipment is another common problem. When adding internal computer boards, such as modems or memory, be certain that the unit is reassembled properly. Almost all electronic equipment subject to Part 15 radiation standards relies on the enclosure material to help meet performance requirements. It should be noted that units designated for business use meet less restrictive standards than consumer equipment. An electronic cash register in a business, for example, is more likely to be an interference source than a consumer product.

## Conclusion

These are just a few reminders of the RFI hazards that products may encounter in their users' locations. As business and home use of RF devices continues to grow, problems with RFI will increase. Awareness of potential interference sources will certainly help service personnel, but some consideration for these hazards can also be applied at the design stage.

□



# Complex Signals: Part IV

By Noel Boutin  
University of Sherbrooke

This is the final installment of a four-part series on complex signals. The three previous installments attempted to demystify complex signals. Part 1 illustrated what complex signals look like and why they are used. Parts 2 and 3 showed how to filter and frequency shift complex signals. With such a background, the reader is now ready to probe the various applications of complex signals.

Applications presented in this article rely heavily on material disclosed in the previous installments. Explanations will therefore refer to previous installments.

Single-sideband (SSB) and vestigial-sideband (VSB) are among the oldest communication systems based on complex signals. For sure, SSB and VSB signals can be generated without having recourse to complex signals. Using an appropriate bandpass filter having a real impulse response, one can partially or completely eliminate one sideband of an amplitude-modulated double-sideband suppressed-carrier

(AM DSB-SC). However, this method of generation of an SSB or VSB RF signal lacks flexibility because the bandpass filter must be custom made for the specific carrier frequency used. Greater frequency agility can be obtained if the signal processing is done at baseband level rather than at RF level. Furthermore, with the advent of low-cost powerful digital signal processors, such an approach appears very promising. There exist two baseband processing methods to get an SSB or VSB signal: the Weaver method and the Norgaard method, also known as the phase shift method.

## The Weaver Method

With this method, originally proposed by Weaver in 1956 (1) for SSB (extended here to include VSB), the real baseband message  $r(t)$  is first converted into a complex baseband message  $c(t)$ . The Fourier Transform  $C(j\omega)$  of the complex baseband message is a frequency shifted version of the Fourier Transform  $R(j\omega)$  of the real baseband message  $r(t)$ .  $R(j\omega)$ , which extends from 0 to  $\pm \omega_{\max}$ ,

is respectively shifted to the left or to the right by an amount equal to half its upper frequency  $\omega_{\max}$  according to the desire to preserve the whole upper-sideband (USB) or the whole lower-sideband (LSB). Such a frequency shift is accomplished, as discussed in Part 3, by the frequency shifter shown in Figure 1(a). Next, the resultant complex baseband signal  $c(t)$  is bandlimited with a low-pass filter  $H(j\omega)$  having a real impulse response  $h_R(t)$ . The cut-off frequency of that filter is set equal to  $(\omega_{\max}/2)$  in SSB systems and a fraction larger in VSB systems. Such a filtering operation on the complex signal is accomplished as discussed in Part 2 and illustrated in Figure 1(b). Last, the filtered complex baseband signal  $c_F(t)$  is shifted in frequency by an amount  $\omega_c$  in such a way that the resultant signal is the desired RF real signal. As discussed in Part 3, such a frequency shift is accomplished with the QAM modulator shown in Figure 1(c). The complete Weaver SSB or VSB generator is illustrated in Figure 1 for the case where the LSB is fully transmitted

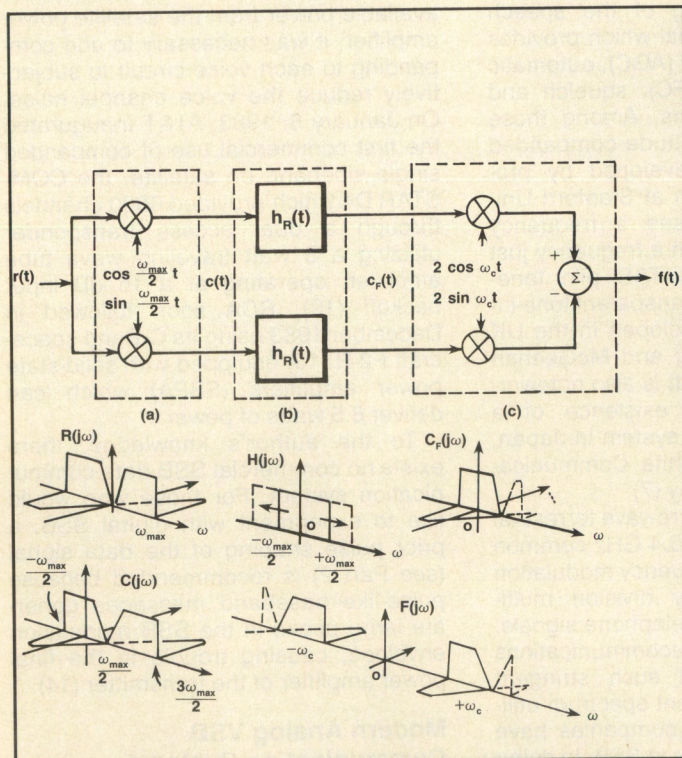


Figure 1. The Weaver method of generating an SSB/VSB RF signal.

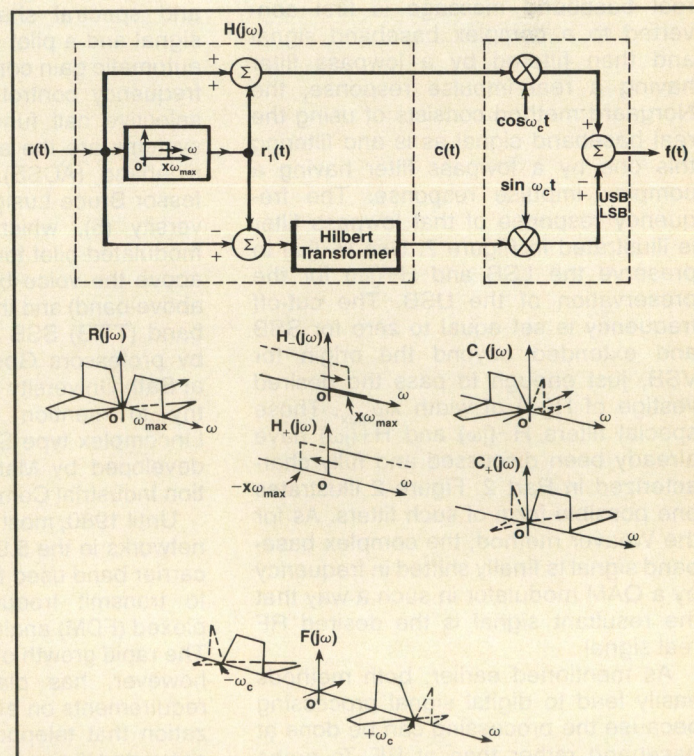
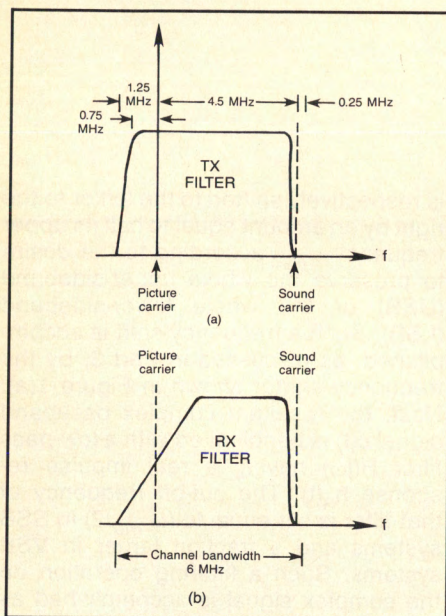


Figure 2. The Norgaard method of generating an SSB/VSB RF signal.





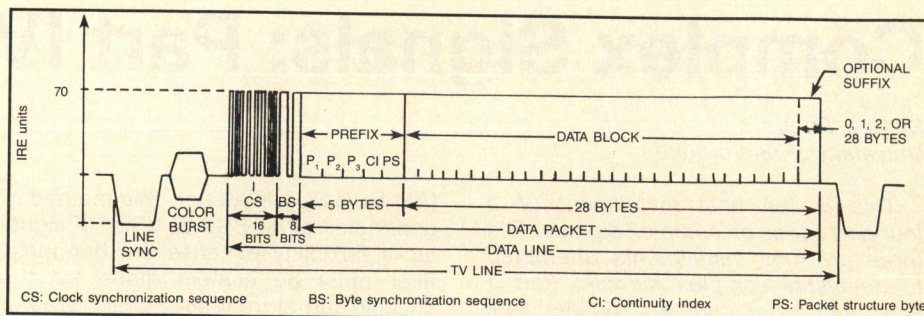
**Figure 3. Idealized transmit and receive filters of VSB TV signal.**

and the lowpass filter has an ideal cutoff.

## The Norgaard Method

Originally proposed by Norgaard in 1956 (2) for SSB, the method will be extended here to also cover VSB. As opposed to Weaver's method where the real baseband message is first converted to a complex baseband signal and then filtered by a lowpass filter having a real impulse response, the Norgaard method consists of using the real baseband signal as is and filtering this one by a lowpass filter having a complex impulse response. The frequency response of that lowpass filter is illustrated in Figure 2. Use  $H-(j\omega)$  to preserve the LSB and  $H+(j\omega)$  for the preservation of the USB. The cut-off frequency is set equal to zero for SSB and extended beyond the origin for VSB, just enough to pass the desired vestige of  $R(j\omega)$  of width  $\omega_{\max}$ . Those special filters  $H-(j\omega)$  and  $H+(j\omega)$  have already been discussed and fully characterized in Part 2. Figure 2 illustrates one possible form of such filters. As for the Weaver method, the complex baseband signal is finally shifted in frequency by a QAM modulator in such a way that the resultant signal is the desired RF real signal.

As mentioned earlier, both methods easily lead to digital signal processing because the processing can be done at baseband rather than at RF. To probe that subject further, refer to Part 2 and reference 3, chapter 7. Please note



**Figure 4. Data line structure for teletext in North America.**

however that this last reference is solely dedicated to SSB and mostly concerned with audio type messages. Blind transposition of those described techniques to data transmission may be very deceptive.

## Modern Analog SSB Communication Systems

Analog SSB communication systems are currently used in many applications. For more than fifty years, radio amateurs have taken advantage of its power efficiency. Over the past ten years, SSB has also been suggested as a bandwidth saver over narrow-band frequency modulation (NBFM) in mobile communications (4). Such systems would use SSB modulation along with baseband processing in the form of amplitude companding and spectral shaping of the speech signal and a pilot signal which provides automatic gain control (AGC), automatic frequency control (AFC), squelch and selective call functions. Among those systems are the amplitude companded sideband (ACSB), developed by professor Bruce Lusignan at Stanford University (5), which uses a frequency modulated pilot tone at a frequency just above the voice band (TAB: pilot tone-above-band) and the transparent tone-in-band (TTIB) SSB developed in the UK by professors Gosling and McGeehan at Bath University (6). It is also noteworthy to mention the existence of a Lincomplex type SSB system in Japan, developed by Matsushita Communication Industrial Company (7).

Until 1980, most microwave terrestrial networks in the 5.9 to 6.4 GHz common carrier band used frequency division multiplexed (FDM) analog telephone signals. The rapid growth of telecommunications however, has placed such stringent requirements on efficient spectrum utilization that telephone companies have converted their systems to SSB. In doing so, they tripled the capacity of their networks and reduced the power by

twenty (8). One of the first companies to do such a conversion was AT&T with its AR6A systems (9) designed to carry 6000 voice circuits on a 30 MHz-wide radio channel. Since no carrier is present with SSB-AM, pilot tones must be added to provide for radio line regulation, signal recovery, and protection switching. To achieve the required very high frequency accuracy of those pilots, some manufacturers phase locked their common terminal reference oscillator to the reference frequency of LORAN C navigation signal (8, 10) while others used a highly stable reference at each station (11).

It was soon realized that the same significant increase in channel capacity could also be obtained with satellite communication. However, due to limited available power from the satellite power amplifier, it was necessary to add companding to each voice circuit to subjectively reduce the voice channel noise. On January 8, 1983, AT&T inaugurated the first commercial use of companded single sideband on satellite, the COMSTAR D4 which provided 6000 channels through a dual access transponder utilizing a 5 watt traveling wave tube amplifier, operating at a 15 dB input backoff (12). RCA soon followed in December 1983 using its C-Band spacecraft F2-R (13) equipped with solid-state power amplifiers (SSPA) which can deliver 8.5 watts of power.

To the author's knowledge, there exists no commercial SSB data communication system. For those who would like to experiment with digital SSB, a prior pulse shaping of the data signal (see Part 2) is recommended because pulse-like baseband messages generate large peaks in the SSB modulation envelope, causing trouble in the final power amplifier of the transmitter (14).

## Modern Analog VSB Communication Systems

One of the most well known analog VSB communication system is surely



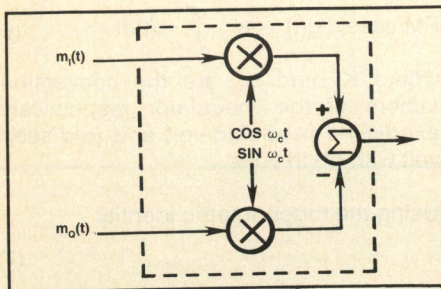


Figure 5. QAM modulator.

the North American TV broadcast system (NTSC). In such a VSB signal, the whole USB, and only 25 percent of the LSB of the video signal is transmitted. As shown in Figure 3, the present standard asks for a brick-wall type transmit filter complemented by a receive filter exhibiting odd symmetry inside the transition interval centered on the carrier frequency (in practice, one must remember that the IF spectrum is a "flipped" version of the transmitted spectrum and that the receive IF filter must follow accordingly). Despite the fact that a synchronous carrier can be regenerated from a VSB-SC signal, it was decided early in the development of the NTSC system to transmit a strong carrier along with the VSB video signal in order to demodulate the signal without too much distortion, with a low cost envelope detector. According to Figures 2 and 3, the signal present at the output of the receive filter is therefore given by:

$$s(t) = \cos \omega_c t + \frac{1}{2} v(t) \cos \omega_c t + \frac{1}{2} v_Q(t) \sin \omega_c t \quad (1)$$

where the first term is the added carrier,  $v(t)$  is the desired video signal and  $v_Q(t)$  is the Hilbert transform of the highpass portion of the video signal lying outside the frequency transition interval. Demodulating  $s(t)$  with an envelope detector, gives

$$m(t) = \sqrt{\left[1 + \frac{1}{2} v(t)\right]^2 + \left[\frac{1}{2} v_Q(t)\right]^2} \quad (2)$$

$$= \left[1 + \frac{1}{2} v(t)\right] \left\{ 1 + \left[ \frac{\frac{1}{2} v_Q(t)}{1 + \frac{1}{2} v(t)} \right]^2 \right\}^{1/2}$$

Equation 2 shows that it is the presence of  $v_Q(t)$  which contributes to the distortion. Increasing the width of the vestigial sideband would reduce  $v_Q(t)$  and the distortion accordingly. It has been determined empirically that the vestigial sideband width can be set as wide as 1.25 MHz without inferring too much distortion.

It is however admitted today that envelope detectors and quasi synchronous detectors no longer suffice and that coherent detection is a must in order to get high quality video signals (15) such as in teletext.

### Modern Digital VSB Communication Systems

A good example of a digital VSB communication system is the North American Basic Teletext System (NABTS) (16). The idea of teletext is to utilize the television channel for data transmission when picture transmission is suspended periodically during the vertical blanking interval. Furthermore, for full-field transmission, all but the synchronization lines of the vertical blanking interval can be used. Figure 4 shows the data line structure. The data signal is a two level NRZ signal transmitted at a bit rate equal to 8/5 of the 3.579545 MHz color burst subcarrier, that is 5.727272 Mbits/sec. It has been found experimentally that in order to avoid visibility during vertical retrace and a "buzz" on the sound channel, the level of that NRZ signal should be restricted to 70 IRE units. As pictured in Figure 4, the NRZ signal is not filtered. However, in practice, it should be filtered because the NTSC TV system limits the baseband composite video bandwidth to 4.2 MHz. Although not optimum in terms of signal-to-noise ratio (SNR) performance, the entire pulse shaping is done at the transmitter and the receiver low-pass is simply used to remove as much noise as possible without further shaping of the received filtered NRZ signal. The data transmit filter must satisfy the zero-intersymbol interference conditions discussed in Part 2 and also restrict the bandwidth occupancy to about 4.2 MHz, that is, 73 percent of the bit rate. Some classic pulse shaping filters have been suggested for that particular application (17, 18).

Quadrature distortion due to envelope detection increases the bit-error-rate (BER). Coherent detection is therefore mandatory. Furthermore, great care must also be exercised in the design of the IF VSB bandpass receive filter with respect to its phase linearity (19). SAW filters seem to be very good candidates for such an application. Many companies already offer SAW filters dedicated to VSB television transmission and reception. A teletext signal is also severely distorted by multipath propagation or echoes in CATV cables which appear as ghosts on regular television pictures. Since a ghost is a delayed signal and

regarded as a linear distortion, it can be cancelled by a linear equalizer. Such equalizers utilize a transversal filter structure such as the one discussed in Part 2 except that the tap gains  $a_i$  are now programmed according to a real time decision algorithm which uses samples of the input and output signals.

### Quadrature Amplitude Modulation Systems

As defined in most communication textbooks, a quadrature amplitude modulation (QAM) system is a scheme which enables two DSBSC modulated waves to occupy the same transmission bandwidth, and yet allows for separation of the two baseband messages at the receiver output (20). This scheme is also known as quadrature-carrier multiplexing. Figure 5 illustrates the basic block diagram of such a system proposed in 1926 by R.K. Potter (21) for broadcasting binaural sound. At the receiver, coherent demodulation is mandatory for total separation of the two messages  $m_I(t)$  and  $m_Q(t)$ . As an unmodulated carrier is not transmitted along with the QAM signal it may be impossible to regenerate such a synchronous carrier at the receiver. As early as 1932, A.V.T. Day recognized this and proposed to add an unmodulated carrier to the transmitted QAM signal for purpose of carrier synchronization (22). There exist today many modern communication systems based on his idea.

The most well known analog QAM system is the NTSC color television system. The signals  $m_I(t)$  and  $m_Q(t)$  are the chrominance signals respectively given by

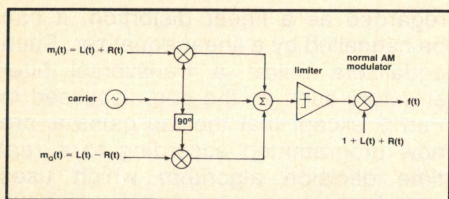
$$m_I(t) = 0.60m_R(t) - 0.28m_G(t) - 0.32m_B(t) \quad (3)$$

$$m_Q(t) = 0.21m_R(t) - 0.52m_G(t) + 0.31m_B(t)$$

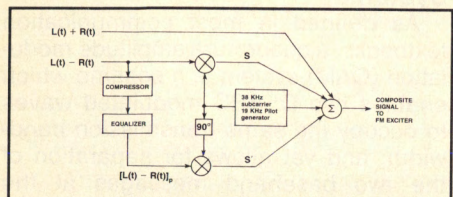
where  $m_R(t)$ ,  $m_G(t)$  and  $m_B(t)$  are the red, green and blue signals generated by the color camera. In the QAM system, the added "unmodulated" carrier is gated on and off by a low duty cycle square wave and appears as a burst of 8 cycles located on the trailing portion of each horizontal line synchronization pulse (see Figure 4). The frequency spectrum of such a gated color carrier is discrete and has a sufficient strength component at the carrier frequency to ensure positive locking of the demodulator local oscillator at the receiver.

Another modern analog QAM system is the C-QAM AM stereo broadcasting system proposed by Motorola (23) which gradually established itself as the lead-





**Figure 6. C-QAM AM stereo generator.**



**Figure 7. FMX FM stereo generator.**

ing system in many US cities as well as Australia and South America. As shown in Figure 6, the in-phase baseband message  $m_1(t)$  is the sum signal  $L(t) + R(t)$  whereas the quadrature baseband message  $m_2(t)$  is the difference signal  $L(t) - R(t)$ . An unmodulated carrier is added with the QAM signal. In order to provide compatibility with monophonic AM receivers, the resultant signal is hard limited and then amplitude modulated as usual by the monophonic compatible sum signal  $L(t) + R(t)$ .

One of the most recent analog communication systems within which QAM is involved is FMX, a new extended range FM stereo system developed at the CBS Technology Center by Emil L. Torick in conjunction with Thomas Keller of the National Association of Broadcasters (24, 25). As shown in Figure 7, a quadrature version of the 38 KHz subcarrier is amplitude modulated by a compressed and equalized version of the difference signal  $L(t) - R(t)$  which still modulates the in-phase 38 KHz subcarrier. Both modulated signals are then added together to form a QAM signal. An unmodulated 38 KHz subcarrier is not added to that QAM signal. Instead, a frequency divided by two versions of the 38 KHz subcarrier was retained. The 19 KHz pilot tone serves the same purpose of locking a local oscillator at the receiver for synchronous demodulation. The baseband spectrum of the composite signal of FMX is generally pictured as shown in Figure 8. Unfortunately, that representation is misleading. As discussed in Part 2, the Fourier transform of a real signal does not have to be real. The same affirmation also applies for an imaginary

signal. If the two Fourier transforms were orthogonal as pictured, it would be possible to separate them at the receiver with two bandpass filters having orthogonal frequency responses! The signals  $S$  and  $S'$  are not orthogonal in the frequency domain but in the time domain. In fact, as discussed in the Part 3, the signals  $S$  and  $S'$  forming the QAM signals are nothing but a frequency shifted version of the complex baseband signal  $c(t)$  given by

$$c(t) = [L(t) - R(t)] + j[L(t) - R(t)]_p \quad (4)$$

where  $[L(t) - R(t)]_p$  is a compressed and equalized version of  $[L(t) - R(t)]$ .

Complex signals are more severely perturbed than real signals by channel imperfections such as non-linearities and multipath propagation or echoes. For example, consider the bi-path propagation situation illustrated in Figure 9. The received signal is formed of two terms. The first one is the original transmitted complex signal  $c(t)$ , whereas the second one is a time-delayed and phase-shifted version of  $c(t)$ . From Part 2, it is known that phase shifting a complex signal means rotating it around the time axis by a given angle. In so doing, the real and imaginary parts of the rotated complex signal no more lie on the same planes than those of the original complex signal. When the two complex signals are added together another complex signal is obtained whose real and imaginary parts are mixed up versions of the real and imaginary parts of the original complex signal. Some of the in-phase channel signal  $m_1(t)$  leaks into the quadrature channel signal  $m_2(t)$  and vice versa. This is the channel crosstalk which is responsible for ghosts and incorrect color tint in NTSC color television receivers, poor channel separation in AM stereo, and a lot of contention (26, 27) concerning FMX performance.

As a final example of analog communication systems within which complex signals are involved, consider the following constant envelope phase or frequency modulated (PM or FM) sinusoidal carrier of nominal frequency  $\omega_c$

$$f(t) = A \cos[\omega_c t + g(t)] \quad (5)$$

The modulating baseband signal  $g(t)$  is related to the baseband bandlimited real message  $r(t)$  to be transmitted in the following way:

$$\text{PM case: } g(t) = K_p r(t)$$

$$\text{FM case: } g(t) = K_\omega \int_0^t r(t) dt \quad (6)$$

where  $K_p$  and  $K_\omega$  are the conversion factors of the modulator respectively expressed in radian/volt and (rad/sec)/volt with  $r(t)$  in volts.

Using the trigonometric identity

$$\cos(a + b) = \cos a \cos b - \sin a \sin b \quad (7)$$

$$\cos(a - b) = \cos a \cos b + \sin a \sin b$$

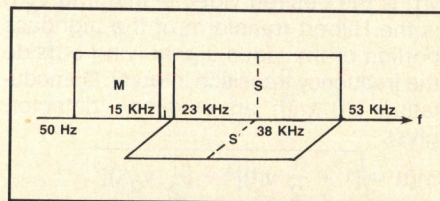
the PM or FM signal given in equation 5 can also be expressed as

$$f(t) = \cos[g(t)] A \cos \omega_c t - \sin[g(t)] A \sin \omega_c t \quad (8)$$

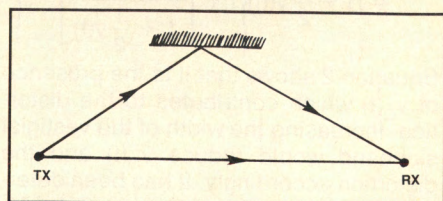
This last expression leads itself to the block diagram illustrated in Figure 10. A QAM modulator whose sole purpose is to shift in frequency the complex baseband signal  $c(t)$  given by

$$c(t) = \cos[g(t)] + j \sin[g(t)] = e^{jg(t)} \quad (9)$$

The complex baseband signal  $c(t)$  is the original real baseband bandlimited signal  $g(t)$  which has been rolled on a unity radius cylinder. As an illustrative example, let  $g(t)$  be a sinusoidal tone of amplitude equal to  $3\pi$  radians. The corresponding complex baseband signal appears in Figure 11. Such a processing of a real signal was discussed in Part 1 where the message was digital. Looking at a PM or FM signal this way elucidates the behavior of such signals with respect to their bandwidth



**Figure 8. A misleading representation of the baseband spectrum of the composite signal of FMX.**



**Figure 9. Example of a bi-path wave propagation.**



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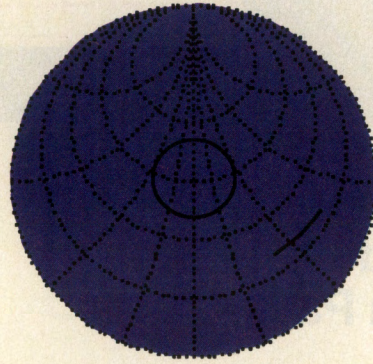
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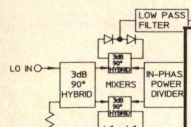
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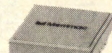
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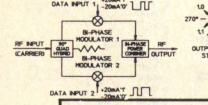


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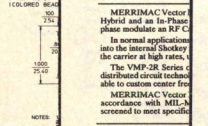
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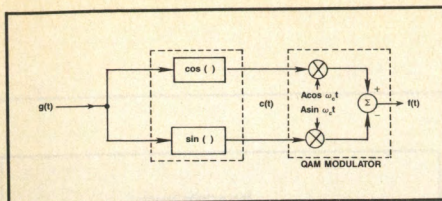
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**Figure 10. Block diagram of a PM/FM modulator.**

occupancy, their performances in the presence of noise and their weaknesses to those channel imperfections discussed earlier, that is non-linearities, multipath propagation or echoes.

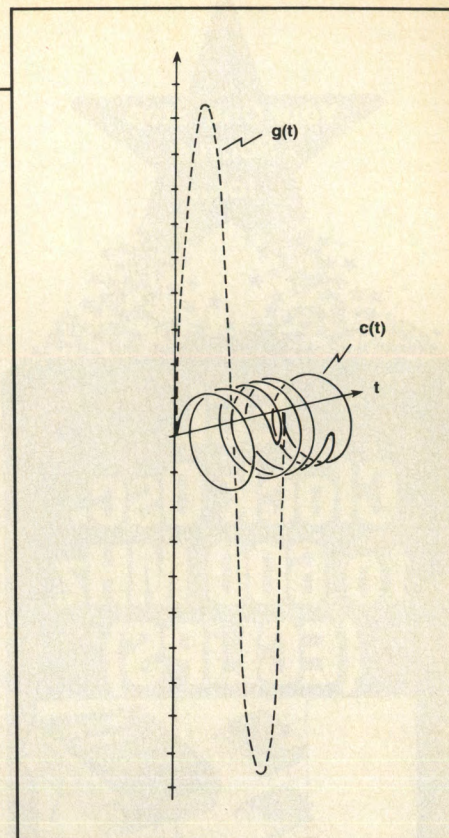
### Modern Digital QAM Systems

Digital QAM systems refer to situations where the in-phase and quadrature signals  $m_I(t)$  and  $m_Q(t)$  injected into the QAM modulator shown in Figure 5 are both digital signals. A priori, those two digital signals may be independent, have different rates and an unequal number of levels. In practice however, as discussed in Part 1, they almost always come from the same binary source. Figures 12 and 13 show two basic digital QAM generators. In the first one, simultaneous state changes may occur on  $m_I(t)$  and  $m_Q(t)$  whereas in offset-QAM (OQAM), the transitions on

$m_I(t)$  and  $m_Q(t)$  are offset by half a baud.

When  $m_I(t)$  and  $m_Q(t)$  are both binary signals, the resultant modulated signals are referred to as quadrature phase shift keying (QPSK) and offset quadrature phase shift keying (OQPSK). Those appellations are misleading because in both cases, when transitions on  $m_I(t)$  and  $m_Q(t)$  are not instantaneous, the envelope of the modulated carriers is not constant as it should be with pure phase modulation. QPSK was first applied in voice-band data communications in the early 1960's. The first high density terrestrial microwave digital radios placed in series in the USA in the mid to late 1970's also utilized QPSK signals. Today, QPSK is still in use in most digital satellite communication systems, including major satellite systems such as that operated by the International Telecommunications Satellite Organization (INTELSAT).

Other bandwidth efficient constant envelope modulation schemes have been derived from OQPSK. Among them are the minimum shift keying-types (MSK-types) (28) and tamed frequency modulation (TFM) (29). In all of those



**Figure 11. One cycle of a rolled sinewave.**

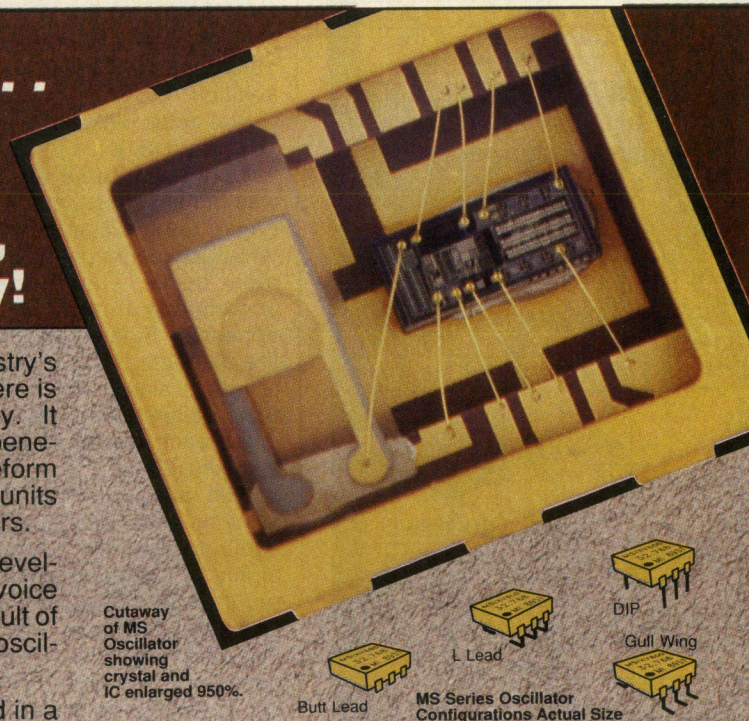
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## rf signals

systems, the original four-state, "square-box" quasi-instantaneous transition complex signal  $c(t) = m_I(t) + jm_Q(t)$  is converted, with linear or non-linear circuits, into a constant envelope complex signal having smooth transitions from one state to another during the full baud period. The resultant constellation display confines itself on a closed circle.

The growing demand for the transmission of digital telephone channels and data signals along with significant technological advances in low noise and high power microwave amplification, predistortion and the application of adaptive techniques led to the substitution of the binary signals  $m_I(t)$  and  $m_Q(t)$  by multilevel signals. 4-level signals result in a baseband complex signal having 16 states which, once shifted in frequency by the QAM modulator, gives a modulated carrier called a 16QAM signal. The typical resultant constellation display appears in Figure 14(a). Operation in the 4 and 5 GHz bands at 200 Mbts/sec., means that digital radio has a transmission capacity of 3600 voice circuits. Incidentally, it was recently demonstrated that such a 6 GHz 16QAM radio signal can also be transmitted over an analog fiber-optic channel (30). Siemens has also demonstrated that 16QAM is a flexible and economical solution for short-haul, high capacity digital radio links operating in the 18 GHz band (31).

Only about three years after 16QAM was put in service, many manufacturers began to offer commercial 64QAM systems for common carriers or private radio bands. The constellation display associated with such an unfiltered signal appears in Figure 14(b). Table I presents the characteristics of some commercial products along with numbers referring in detail each one of those listed 64QAM systems (33-43).

Just after the feasibility of 64QAM systems was confirmed, research on 256QAM systems began. However, a highly accurate modem technology is required for a 256QAM system because, as shown in Figure 14(c), each one of the 256 states is very close to its neighbor. The degradation factors arising at various parts in the system, such as channel amplitude and delay dispersion, modulation amplitude and phase errors, threshold drift, recovered carrier jitter and clock phase error, etc., have been characterized and their allowable maximum values specified (32). This led to the introduction of a variety of novel

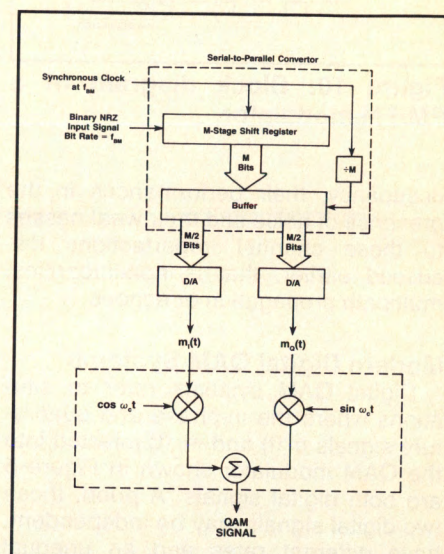


Figure 12. Digital QAM generator.

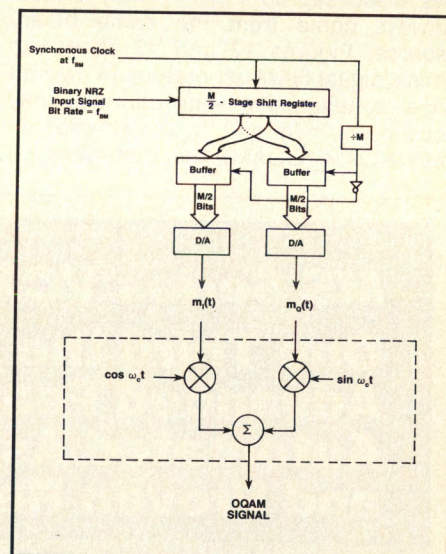
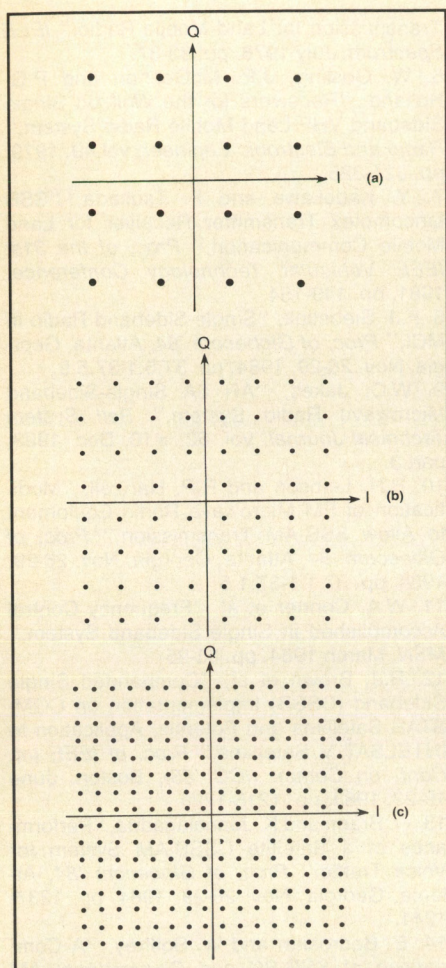


Figure 13. Digital OQAM generator.

techniques to achieve performances similar, in terms of bit error rate, to those of a 64QAM system. For example, the development of a new monolithic multiplier IC having a baseband input voltage linearity of more than 1.5 V and a modulation phase error less than 0.2 degrees is reported in (44). Likewise, to improve existing fading countermeasures, a minimum inband dispersion space diversity combiner controlled by external sensing, a transversal equalizer using a maximum level error (MLE) algorithm, and a transversal type cross-polarization interference canceler was proposed by NTT (45). In order to reduce the recovered carrier phase jitter, a pilot injection method together with feedback

May 1990





**Figure 14. Constellation display of a) 16QAM, b) 64QAM, c) 256QAM.**

balance coding which reduces spectral power near the carrier was adopted by Fujitsu (46). The effects of nonlinearity on RF power amplifiers can be mitigated by reducing the peak-to-average power ratio of the QAM signal. This can be

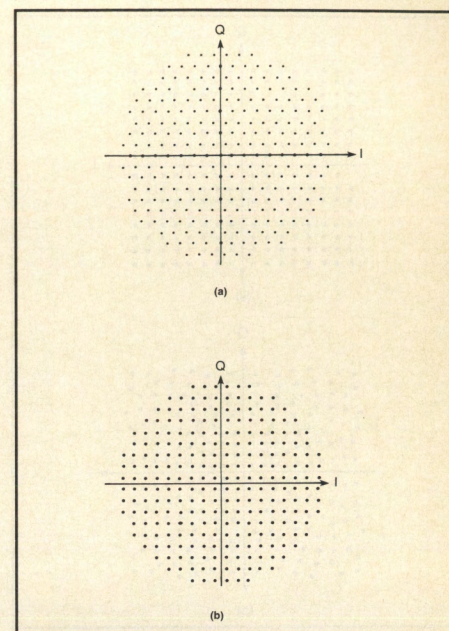
realized with constellations other than the basic square one shown in Figure 14(c). For example, Fujitsu proposed the honeycomb one (47) illustrated in Figure 15(a) whereas NEC developed the stepped-square (SS) one (48) of Figure 15(b).

The transmission of T1 (1.544 Mbit/sec) data streams over the existing analog FM and FDM-SSB microwave radio by means of 256QAM systems is also a reality. Figure 16(a) illustrates the particular constellation used for FM links whereas Figure 16(b) shows the one retained for AM-SSB links where phase jitter is a major impairment. As mentioned in reference 49, hardware improvement is no longer sufficient to get high quality service with those high-level QAM systems operating through FM and SSB radio; one must have recourse to coding. Trellis coded modulation (TCM) is capable of providing a significant coding gain of 2 dB against fading down to 28 dB SNR for 256QAM modulation. Furthermore, even at this low operating SNR, that coding can be followed by conventional forward error correcting codes such as interleaved Reed-Solomon codes used for burst noise protection which still provide over 2 dB of additional coding gain.

There has also been significant progress made in voice-band telephone data transmission since the introduction of the first QPSK modem in the early 1960's, operating at 2400 bits/sec. Due to space limitation and the non-RF nature of those modems, it will not be covered here. It is worth noting however, that many techniques used at low frequencies are directly applicable at RF.

#### Instrumentation

This section is a partial review of currently available test equipment which



**Figure 15. Constellation display of a) honeycomb 256QAM, b) stepped-square 256QAM.**

can serve to generate baseband complex signals and QAM signals as well as those that can display them or their constellations.

**Complex Signal Generators:** A complex test signal  $c(t) = I(t) + jQ(t)$  can be generated by two independent but synchronizable arbitrary waveform synthesizers; one outputting  $I(t)$ , the other  $Q(t)$ . Those synthesizers are available through at least three manufacturers: Wavetek, Hewlett-Packard and LeCroy. The single channel model 75 from Wavetek is the least expensive one. It can generate a 2048 point waveform with 12 bit resolution at a speed up to 2 Mpoints/sec. The dual channel HP8175A

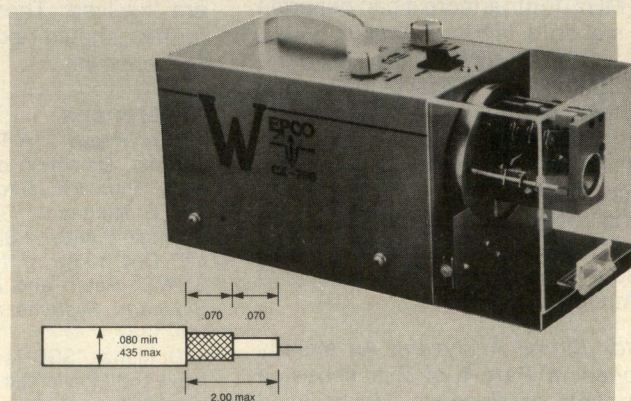
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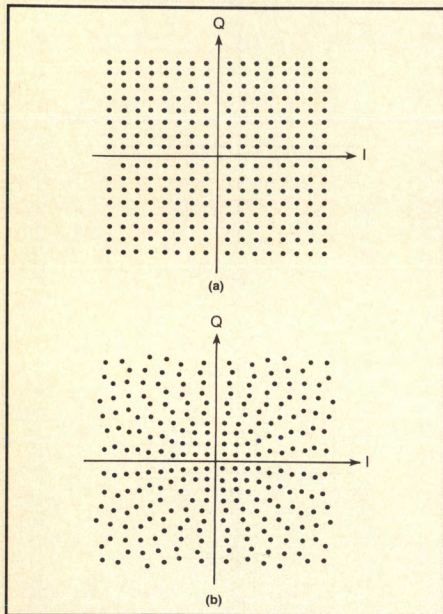
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**Figure 16. Custom constellation display for a) FM Links, b) AM-SSB Links.**

opt 002 from Hewlett Packard outputs two 1024 point waveforms with 10 bit resolution at a speed up to 50 Mpoints/sec. For those who need much more performance, look for the Hewlett Packard single channel HP 8770A. This 512 Kpoints, 12 bit resolution generator can run at speeds of up to 125 Mpoints/sec. Similar speed performance is also available with LeCroy 9100 series offered as single channel (model 9101) or dual channels (model 9100).

**QAM Signal Generators:** A QAM test signal can be generated by feeding a baseband complex test signal into a QAM modulator. Such a QAM modulator is available from Hewlett Packard. The HP8780A can generate a 10 MHz to 3 GHz QAM signal from an external baseband complex signal extending up to 350 MHz. Furthermore it can generate many standard digital modulations like BPSK, QPSK, 8PSK, 16QAM and 64QAM. Although the cost of the second generation HP8782A vector signal generator is substantially lower than the HP8780A, the former lacks of flexibility because, having no analog inputs, it supports only built-in digital modulation formats. Direct digital synthesizers (DDS) are also a viable alternative to traditional QAM modulators (54).

**Complex Signal Analyzers:** As already mentioned in Part 1 of this series, a conventional oscilloscope can be used to visualize the lissajous pattern or the

constellation display of a complex signal. Specialized oscilloscopes are however available which, in addition to displaying constellation displays, can measure many other important parameters. The Hewlett Packard HP3709B constellation analyzer can also measure the linear and non-linear distortions revealed by the constellation displays, and can provide a formatted report containing the constellation and measurement results on a printer. For those who require a more complete diagnostic, consider the HP8980A Vector Analyzer. With this 350 MHz bandwidth instrument you can visualize the 3D complex signal, its I and Q orthogonal projections, its vector display at or around the sampling instants. Furthermore, a built in analysis software quantitatively measures quadrature error or non-linear distortions.

**QAM Signal Demodulators and Analyzers:** The HP8981A Vector modulation analyzer is without question the most complete QAM signal analyzer. It includes the same baseband analyzer than the HP8980A, but adds a precise QAM signal demodulator for carrier frequencies ranging from 50 MHz to 200 MHz. For those who only require a high precision QAM signal demodulator without any further capacity, consider the Mirage Systems model 850A Digital Quadrature Detector. Operating on a fixed IF of 70 MHz the model 850A provides analog outputs as well as 16 bits digital versions of both in-phase and quadrature components of the complex baseband signal.

## Conclusion

This four part series on complex signals should be considered as a primer. Due to space limitations it was impossible to go into detail concerning every system in which complex signals are involved. Those desiring to probe further should examine the following reference list.

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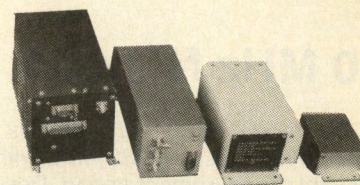
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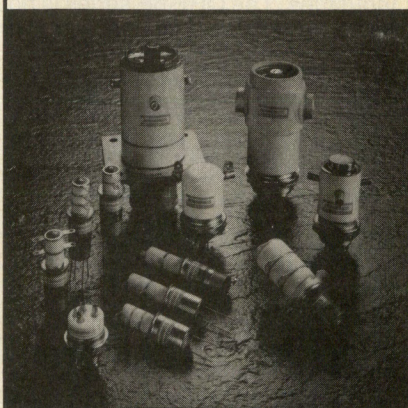
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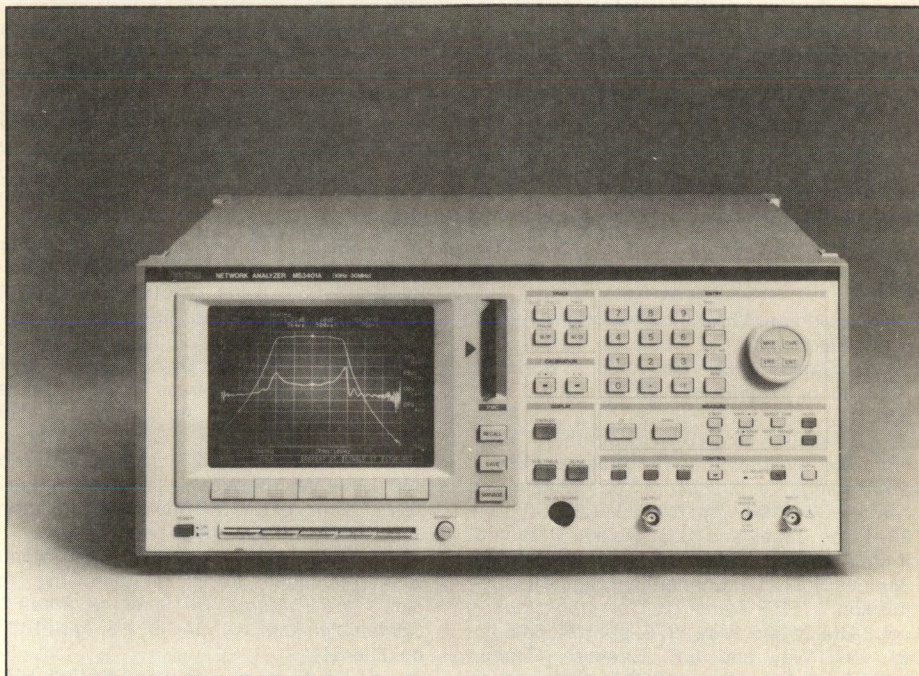


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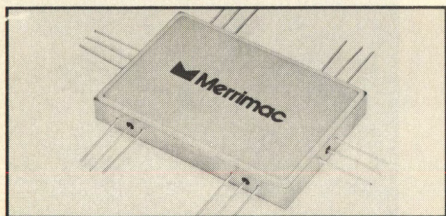
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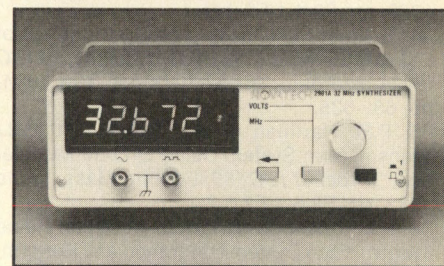
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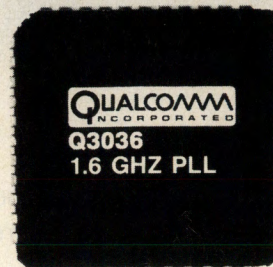
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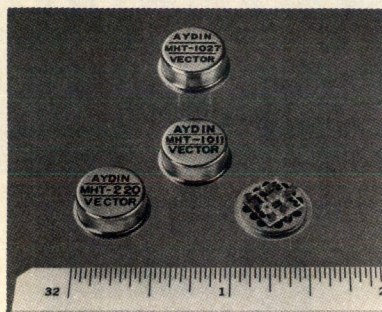
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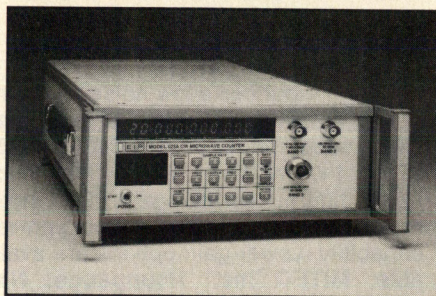
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## rf products *Continued*



half rack enclosure. The small size is suitable for applications in field test, while its performance allows R&D and automated test applications. EIP's YIG preselector technology is incorporated in the unit, along with GPIB programmability. Also included is a 12 volt input to maintain the oven time base while in transit. Price is \$4200 with 6-week delivery. **EIP Microwave, Inc., San Jose, CA. INFO/CARD #192.**

### 200 MHz Op Amp With Disable

CLC410 is a high speed op amp with a disable feature for video switching and distribution applications. Designed for

low gain ( $A_v = \pm 1$  to  $\pm 8$ ), the unit consumes 160 mW and settles to 0.05 percent in 12 ns. The disable feature has 100 ns turn-on and 200 ns turn-off times. Disabled output impedance is greater than 100 kohms. Pricing (100s) ranges from \$9.15 to \$39.30, depending on packaging and temperature range. **Comlinear Corp., Fort Collins, CO. INFO/CARD #191.**

### 1.2-1.4 GHz Amplifier

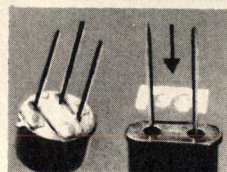
A 10 watt class C amplifier, model PA1214-10, is available from Frequency Products. Designed for either pulse or CW applications, the unit has input and output isolators for operation into any load without damage. Typical gain is 25 dB, VSWR is 1.4:1, and gain flatness is  $\pm 1$  dB. Power consumption is 3.0 amps at +28 V. **Frequency Products, Inc., San Carlos, CA. INFO/CARD #190.**

### Laser Radar Processor

Model ASH803 has been designed to process the doppler information out of laser velocimeters (Lidar). The 3 dB

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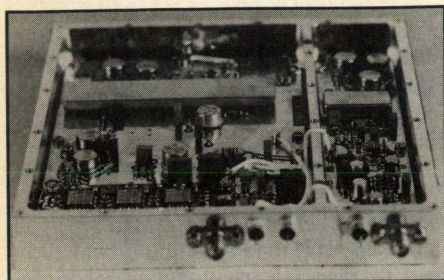


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INFO/CARD 59





resolution of 1 MHz corresponds to 5 m/s velocity with CO<sub>2</sub> lasers, and 0.5 m/s with YAG lasers. The output is log detected with 50 dB dynamic range. A post-processor can be added which automatically measures amplitude and frequency of the spectrum maximum and deliver an 18-bit word. **Thomson-ICS Corp., Southwick, MA. Circle INFO/CARD #189.**

#### High Power Pulse Generators

Directed Energy's SV-M series of pulse generators features all solid state design using high performance MOSFET arrays. Available in 2 kV, 4 kV, and 6 kV models (at 40 A), the series offers

## RF Design Software Service

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

This Month's Disk:

#### Disk RFD-0590: the May 1990 disk

Delay Equalizer Program — Author Tom Hajjar has revised the program by Robert Kane from the April 1989 issue of *RF Design*. The program has been rewritten in Microsoft QuickBasic for MS-DOS computers. Corrections to the computations have also been made. [Compiled (requires EGA), with source code]

From Last Month:

#### Disk RFD-0490: April 1990

Two programs from RF Expo presentations by Douglas Linkhart of Micon Inc.

1. "Unequal Splitter Design" based on quarter wave transformer techniques.
2. "Disk-Rod Filter Design" for distributed coaxial lowpass filters.

Program documentation includes brief operating instructions and references.

\*\* Reprints of these RF Expo Papers are available for \$5.00 \*\*

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INFO/CARD 64

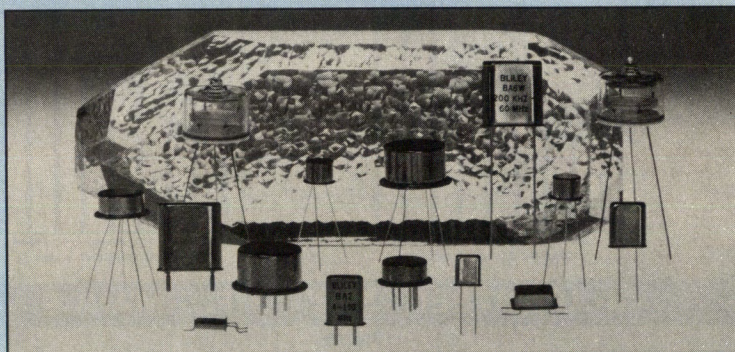
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variable rise time from 40 ns to 120 ns, CW operation to 100 kHz (1 MHz burst), and variable pulse width from 100 ns to >50  $\mu$ s. These generators are designed for driving magnetrons, high power tubes, and other applications. Pricing ranges from \$6480 (2 kV) to \$11,160 (6 kV). **Directed Energy, Inc., Fort Collins, CO. Please circle INFO/CARD #188.**

## Low Cost Precision Reference

Guide Technology announces the GT300 stand-alone reference frequency source, utilizing an ovenized 10 MHz oscillator and driver circuitry that can deliver the signal to as many as 20 instruments. Stability is  $\pm 1 \times 10^{-8}$  over 0 to +40C, with aging  $< 5 \times 10^{-10}$  per day or  $< 1.5 \times 10^{-7}$  per year. Two +7 dBm outputs are provided. Price of the GT300

is \$1195 in single quantities. **Guide Technology, Los Altos, CA. Circle INFO/CARD #187.**

## 75 ohm Attenuators

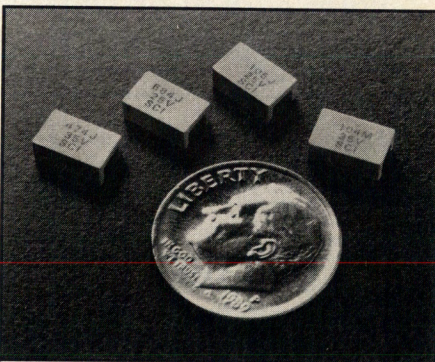
The new MDA line of 75 ohm programmable attenuators has a frequency range of DC to 1 GHz, step size as small as 0.1 dB, and total attenuation as much as 127 dB. Electrical construction includes EMI/RFI filtering on each bit, plus overall filtering. Standard control voltage is 26.5 VDC, with other voltages available. List prices are \$220 to \$410. **Alan Industries, Columbus, IN. INFO/CARD #186.**

## Feedthrough Terminations

Feedthrough terminations designed for connection of transmission lines to oscilloscopes, counters, RF voltmeters and other instruments have been introduced by Elcom Systems. They are available in 50 ohm (DC-1 GHz), 75 ohm (DC-500 MHz), and 93 ohm (DC-150 MHz) versions. Connector options include BNC, TNC, N, and SMA with gold, silver or nickel plating. Prices start at \$17.50. **Elcom Systems, Inc., Boca Raton, FL. INFO/CARD #185.**

## Polymer SMT Capacitors

SpecPML polymer multilayer capacitors have been introduced, offering temperature stability, frequency stability, non-



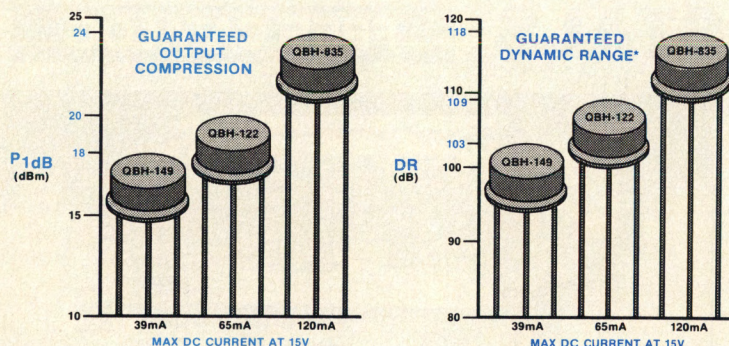
polarity, self-clearing, and standard SMT mounting sizes. Vacuum manufacturing techniques and a thermally stable dielectric assure specified performance. Capacitance values are available from 0.22  $\mu$ F to 1.0  $\mu$ F in 5, 10 and 20 percent tolerances. **Spectrum Control, Inc., Tucson, AZ. INFO/CARD #184.**

## DC-2 GHz Attenuator

A programmable attenuator with 7-bit binary control, model PA-501-R, is announced by RLC Electronics. The unit has 127 dB maximum attenuation, in 1

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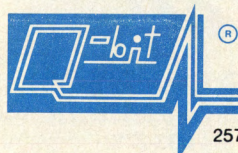


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| QBH-149 | 10-150 MHz | 23.2 dB | 2.8 dB       | 1.5:1 | 28/42 dBm | \$100                  |
| QBH-835 | 10-500 MHz | 11.0 dB | 6.0 dB       | 1.6:1 | 35/45 dBm | \$105                  |

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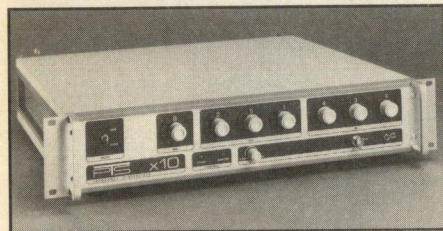
dB steps. Accuracy is 0.2 dB or 1/2 percent at the lowest frequencies, rising to 0.3 dB or 2 percent at 1-2 GHz. Standard control voltage is 12 VDC. Options are available for other voltages or other attenuation cell values. Prices start at \$575. **RLC Electronics, Inc., Mount Kisco, NY. INFO/CARD #183.**

### 20-100 MHz 500 Watt Amplifier

A class AB linear amplifier featuring 500 watts output, 20-100 MHz frequency coverage, 57 dB minimum overall gain (0 dBm input), and IEEE-488 computer control is available from PST. Designated BHE 2718-500/1364, the amplifier includes overdrive, temperature and output VSWR protection, 15 dB of output power control, internal AM modulation, and remote filter switching control. **Power Systems Technology, Inc., Hauppauge, NY. INFO/CARD #182.**

### Low Noise Synthesizer

The PTS x10 is a low noise, fast switching slot synthesizer configured to



cover any 10 MHz span between 0.1 and 100 MHz. The unit features 1 Hz resolution and 1-5  $\mu$ s phase-continuous switching. Over a 2 MHz bandwidth (selectable by the user), ultra low phase noise ( $-132$  dB/Hz at 1 kHz offset) and spurious ( $-60$  dB) is available. Manual or remote controls, and BCD or GPIB programming are available. Prices start at \$1650 for a remote-only, BCD pro-

grammed unit. **Programmed Test Sources, Inc., Littleton, MA. INFO/CARD #181.**

### 14 GHz Prescaler

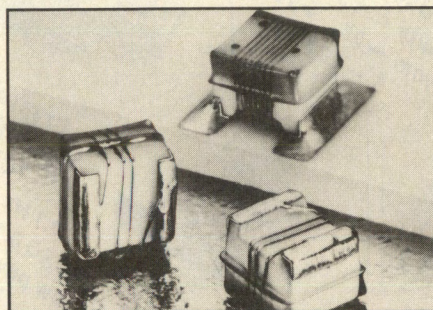
A divide-by-eight prescaler for 8-14 GHz, the NEC UPG506 is announced by California Eastern Laboratories. Available as a chip or in a hermetically sealed ceramic package, the device has a guaranteed operating temperature range of  $-25$  to  $+75^\circ\text{C}$ . Prices start at \$82 (100s). **California Eastern Laboratories, Santa Clara, CA. Please circle INFO/CARD #180.**

### Replacement Oscillator

The Model 2880067 5.000 MHz oscillator from Piezo Crystal Company is designed to be a direct replacement for the Hewlett Packard 10543 oscillator, to fit into the HP 01695 A/U Frequency Standard. The unit has been approved by the Federal Government. **Piezo Crystal Company, Carlisle, PA. INFO/CARD #179.**

### Compact Chip Inductors

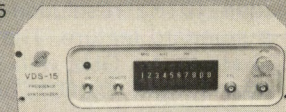
Coilcraft's 0805 series of SMT inductors is designed for high frequency applications where space is at a premium. The



inductors feature a non-magnetic ceramic form for higher Qs and higher

# DIRECT-DIGITAL FREQUENCY SYNTHESIS for ATE

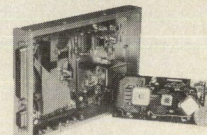
## SYNTHESIZER & PHASE GENERATOR VDS-15



- DC-16 MHz in 0.1Hz steps
- 500 nanosecond switching speed
- $<-50$  dBc spurs
- phase control in  $0.18^\circ$  steps
- ext'l or int'l 10MHz 1ppm source
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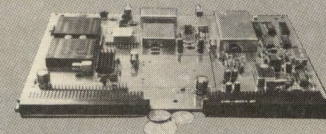
## ALL-GaAs SYNTHESIZER

ADS-232  
ADS-332



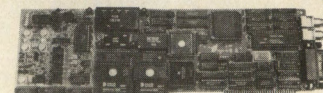
- up to 300MHz coverage,  $<1$ Hz steps
- **$<45$  nanosecond switching speed**
- $<-40$  dBc to  $<-45$  dBc spurs
- phase continuous
- phase noise of reference
- small board or module

## VME FORMAT SYNTHESIZERS



- VDS-3A: DC to 3MHz, 1.0mHz steps
- VDS-3000A: DC to 15MHz,  $\approx 1.0$ mHz steps
- VDS-3001A: DC to 30MHz, 1/16Hz steps
- VDS-3050: DC to 35MHz, 5Hz steps
- VDS-3125: DC to 110MHz, 1.0Hz steps

## MULTIGEN for IBM-PC



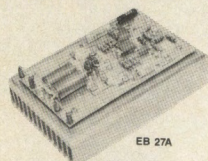
- DC-8MHz in 0.1Hz steps (SINE)
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| AN779L 20W \$ 83.70 | EB104 600W \$448.15 |
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self-resonant frequencies than magnetic core devices. 19 inductance values are available covering 3 to 220 nH with average Qs of 40-50. Price is 25¢ in 10,000 quantities. **Coilcraft, Cary, IL. INFO/CARD #178.**

## 10 Hz Resolution Capability

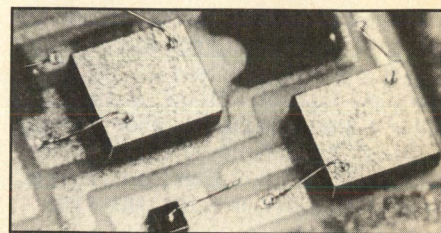
The Tektronix 497P Spectrum Analyzer now includes 10 Hz resolution bandwidth as a standard feature. The 497P covers 100 Hz to 7.1 GHz with 90 dB dynamic range, and 0.0001 percent frequency measurement capability. The price is \$26,250. **Tektronix, Inc., Beaverton, OR. INFO/CARD #177.**

## 50-90 MHz IF Band Synthesizer

Covering the 50-90 MHz communications IF band, the Syntek SY50JW40-TIT1 is designed for OEM applications requiring low noise, low spurious signals. Spurious content is below -80 dBc at the output, and typical phase noise is -116 dBc/Hz at 300 Hz offset, -122 dBc/Hz at 10 kHz offset. Frequency selection is by either external computer or thumbwheel switch assembly. **Syntek Corp., Bohemia, NY. Please circle INFO/CARD #176.**

## Gold Terminated Capacitors

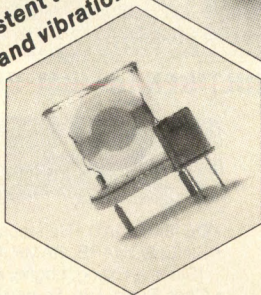
Gold terminated chip capacitors for MIC applications have been introduced, us-



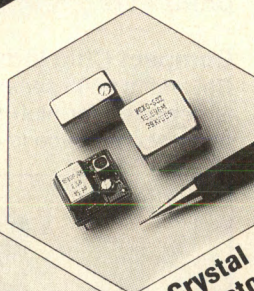
ing high temperature ceramic dielectrics and sputtered or plated platinum-palladium-gold alloy terminations. Various dielectrics are available, as is MIL-STD-883C testing. Pricing ranges from \$0.35 to \$2.50, depending on value, size and quantity. **Johanson Dielectrics, Inc., Burbank, CA. INFO/CARD #175.**

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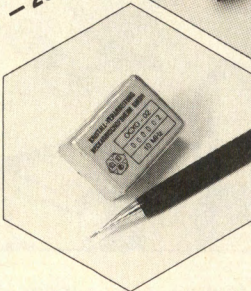
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200 KHz — 200 Mhz high  
temperature stability  
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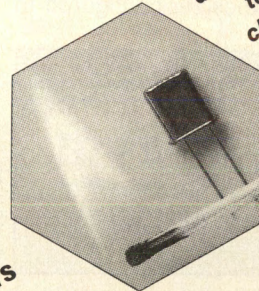
**Ovenized Crystal  
Oscillators**  
stability  $< \pm 5 \cdot 10^{-10}/^{\circ}\text{C}$   
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5 — 20 Mhz



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Quartz Crystals INFO/CARD 76

Crystal Filters INFO/CARD 77

Ovenized Crystal INFO/CARD 78

Crystal Oscillators INFO/CARD 79

Crystal Temperature INFO/CARD 80



## Macintosh Software Update

Ingsoft Limited has introduced their new version of RFDesigner™ for use with Macintosh computers. Version 1.2 now includes a complete "on-line manual" which displays the exact help and instructions needed for all tasks. The software is available with options such as a schematic capture module to allow direct graphical entry of circuits, a Network Analyzer Interface to allow for S-parameter data acquisition through IEEE488-to-SCSI hardware, and an S-Parameter Library which contains data of all major manufacturer's devices. It is priced at \$1500 with certain options extra. **Ingsoft Limited, Ontario, Canada. INFO/CARD #210.**

## Software for Electromagnetic Analysis

Sonnet Software has introduced two new interactive software products, Em, an electromagnetic analysis and Xgeom, a user interface. The combination of the two products creates a design tool with the following capabilities: evaluates any discontinuity (including all interactions) with precision, validates large portions of entire designs, with or without package effects, and enhances existing de-

sign software, no matter who the software is from. **Sonnet Software, Liverpool, NY. INFO/CARD #209.**

## Compact Upgrades

Updated versions of Supercompact™ and Microwave Harmonica™ are now available. For PC based simulators the versions are 4.05 and 1.05 respectively. For workstations and mainframes, Supercompact and Microwave Harmonica have been upgraded to version 2.05. The PC version has a number of enhancements including more accurate models for transformers and diodes; graphics improvements and a sparse solver for noise analysis. The workstation versions now include more accurate models for microstrip tees, spiral inductors, multi-coupled lines and transformers as well as the sparse solver for noise analysis. **Compact Software, Paterson, NJ. INFO/CARD #208.**

## Generic OpAmp Model

Prespice is a new generic subcircuit model that accepts user parameters to allow the circuit designer to simulate hundreds of different OpAmps from the same model library. This allows the designer to convert data sheet param-

eters directly into Spice parameters without complex calculation or parameter tweaking. Parameters can be automatically passed from Prespice to Intusoft's Spice compatible IsSpice package for simulation. The package is priced at \$200. **Intusoft, Santa Monica, CA. INFO/CARD #207.**

## Professional Extension

Eagleware has released =SuperStar= Professional Extension, a general purpose circuit simulation, tuning and optimization program for IBM PCs and compatibles. It contains all the features of =SuperStar= version 3.3 plus: Monte Carlo analysis, yield prediction, microstrip, stripline, waveguide and coax models and equation definition of component values.

Version 3.3 of =SuperStar= is now available with some new features. There are now over 1000 S-parameter data device files, tunable markers with digital readout, additional element models, additional output data formats, built-in support for 192 printers, improved optimizer and an improved editor. It is priced at \$695. **Eagleware Corporation, Stone Mountain, Georgia. INFO/CARD #206.**

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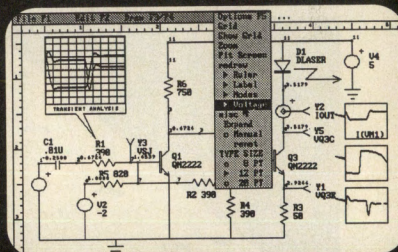
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## Oscillator Catalog

New 1990 catalog features TCXO, VCXO and TCVCXO products. A section on high stability oscillators is also included. **Monitor Products Co., Engineered Products Group, Ocean-side, CA. INFO/CARD #230.**

## Amplifier Handbook

AML, Inc. announces its expanded *High Intercept Low Noise Amplifier Handbook*. Video, RF and microwave products covering DC to 8 GHz are listed, along with reference information, performance plots and package dimensions. **AML, Inc., Camarillo, CA. INFO/CARD #229.**

## Used Test Equipment Listing

Monthly updates highlight new arrivals, best values, and clearance items in RAG Electronics' inventory. Test equipment from many manufacturers is included: oscilloscopes, signal generators, spectrum analyzers, DMMs, and more. All used equipment is calibrated, with a 6-month parts and labor warranty. **RAG Electronics, Inc., Canoga Park, CA. Please circle INFO/CARD #228.**

## Microwave Components Catalog

A new catalog includes technical and applications data on products available from Anaren Microwave. Over 300 models of hybrid couplers, power dividers, mixers, modulators, PIN attenuators and phase discriminators are listed. Subsystem products include DRFMs, ESM and tracking receivers, and digital frequency discriminators. **Anaren Microwave, Inc., E. Syracuse, NY. INFO/CARD #227.**

## Chip and Leaded Inductors

Siemens announces a new short form catalog listing their chip and leaded inductors for EMI suppression, with applications in low and high frequency decoupling, supply filtering, and filters. Major electrical and mechanical specifications are included. **Siemens Components, Inc., Iselin, NJ. Please circle INFO/CARD #226.**

## Designers Guide from Sage

A short form catalog with applications and specifications data on Sage Laboratories' Wireline and Wirepac product lines is now available to assist engineers interested in this balanced twin-conductor material. **Sage Laboratories, Inc., Natick, MA. INFO/CARD #225.**

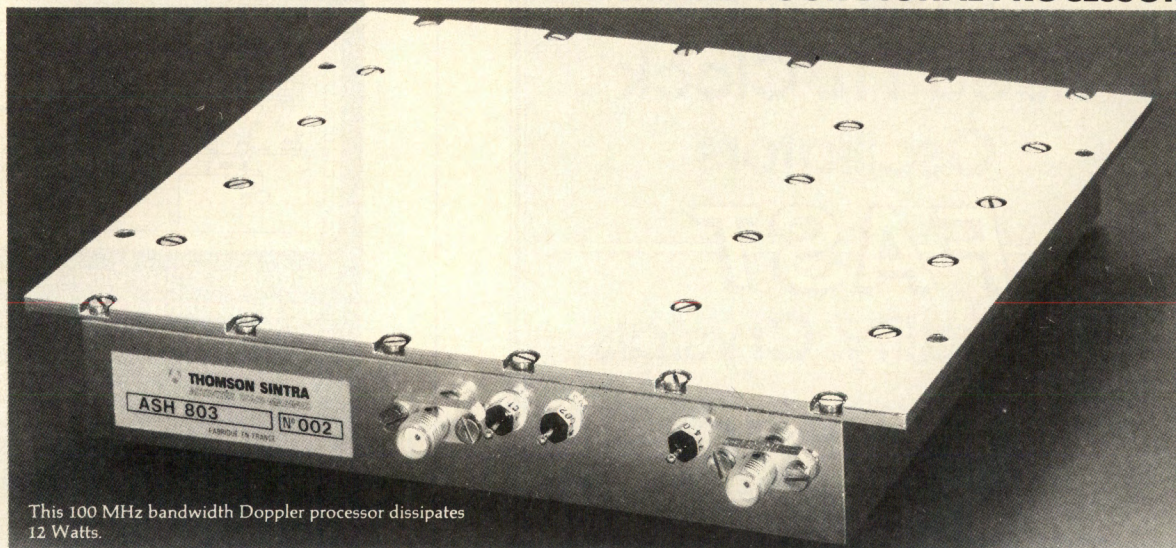
## Single Plate Capacitors

Murata Erie has a brochure available, describing the firm's CLA line of single plate monolithic capacitors for MIC applications. In addition to individual devices, multiple element units with 3, 4, or 5 elements are listed, used primarily for trimming applications. **Murata Erie North America, Smyrna, GA. INFO/CARD #224.**

## RF Selector Guide

Power FETs, power bipolar, and small signal transistors, plus modules are listed in a new selector guide from Motorola. Over 75 new products are included in this cross-referenced guide. **Motorola, Inc., Phoenix, AZ. For more information please circle INFO/CARD #223.**

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### Airborne Signal Conditioning Note

An application note describing recent changes in signal conditioning techniques for data acquisition in flight test programs has been released by Aydin Vector Division. Components, subsystems, and overall system considerations are addressed. **Aydin Vector Division, Newtown, PA. INFO/CARD #222.**

### The Transmission Line News

A quarterly publication from Compact Software updates readers on the latest developments and applications of the company's software products. Included is training seminar information, user applications, and technical notes. **Compact Software, Inc., Paterson, NJ. INFO/CARD #221.**

### Zero Defects Note

Developed from its own experience, M-tron has prepared a tutorial to assist manufacturers in implementing quality control programs. The note explains where statistical process control and other techniques are applied in the sequential crystal production process. **M-tron Industries Inc., Yankton, SD. INFO/CARD #220.**

### High Frequency Clock Data

Data sheets are now available for the AV9101 and AV9104 programmable CMOS Video Frequency Generators, which can generate up to 90 MHz and 65 MHz clock signals. Applications include high resolution graphics and automatic test systems which require up to 16 different pre-programmed frequencies. **Avasem Corp., San Jose, CA. INFO/CARD #219.**

### Brochure Describes Capabilities

A six-page brochure describes the key products, areas of strength, and corporate goals of Erbtec Engineering. RF technology, digital control, systems integration, service and support, OEM support, and quality are areas of note. **Erbtec Engineering, Inc., Boulder, CO. For more information please circle INFO/CARD #218.**

### Electronic Enclosure Catalog

Modular racks, sloped-front consoles, electronic desks and instrument cabinets are described in a new catalog. Included are EMI/RFI shielded cabinets for FCC and TEMPEST requirements. **Equipto Electronics Corp., Aurora, IL. Please circle INFO/CARD #217.**

### MM-Wave Products

New and reconditioned millimeter wave products are offered by Lectronic Research Labs (LRL). A new catalog lists 600 items from 21 manufacturers of waveguide components in the 26.5-220 GHz range. **Lectronic Research Labs, Camden, NJ. For more information please circle INFO/CARD #216.**

### Microwave ARC Sensors

A new data sheet describes a line of microwave arc sensors, solid-state optical devices to detect infrared and visible arcs in high power microwave waveguides. **Gamma Microwave, Santa Clara, CA. For more information please circle INFO/CARD #215.**

### Advertising Index

|                                                        |                |                                                |        |
|--------------------------------------------------------|----------------|------------------------------------------------|--------|
| Adams Russell Electronics/<br>Anzac Division . . . . . | 10, 13, 15, 17 | Lorch Electronics . . . . .                    | 23     |
| Advantest America, Inc. . . . .                        | 12             | Matrix Systems Corp. . . . .                   | 28     |
| Alan Industries, Inc. . . . .                          | 11             | McCoy Electronics<br>Company . . . . .         | 27     |
| American Technical<br>Ceramics Corp. . . . .           | 9              | Merrimac Industries, Inc. . . . .              | 70     |
| Amplifier Research . . . . .                           | 35             | Microwave Software . . . . .                   | 69     |
| Amplifonix . . . . .                                   | 30             | M-Tron Industries, Inc. . . . .                | 71     |
| Andersen Laboratories . . . . .                        | 92             | MTS Microelectronics,<br>Inc. . . . .          | 24     |
| Avantek . . . . .                                      | 43             | Murata Erie . . . . .                          | 56     |
| Aydin Vector . . . . .                                 | 78             | Noise Com, Inc. . . . .                        | 7      |
| Bliley Electric Company . . . . .                      | 79             | Oscillatek . . . . .                           | 8      |
| Cal Crystal Lab., Inc. . . . .                         | 83             | Penstock Engineering, Inc. . . . .             | 86     |
| Champion Technologies,<br>Inc. . . . .                 | 48             | Polaris . . . . .                              | 78     |
| Coilcraft . . . . .                                    | 38             | Q-Bit Corporation . . . . .                    | 80     |
| Communication Concepts,<br>Inc. . . . .                | 81             | QUALCOMM, INC. . . . .                         | 77     |
| Communication<br>Techniques, Inc. . . . .              | 33             | Quartztek . . . . .                            | 77     |
| CTS Corp. . . . .                                      | 91             | Sawtek Inc. . . . .                            | 29     |
| Daico Industries, Inc. . . . .                         | 4              | SCITEQ Electronics,<br>Inc. . . . .            | 51, 81 |
| DGS Associates, Inc. . . . .                           | 16             | Sprague-Goodman<br>Electronics, Inc. . . . .   | 47     |
| Dielectric Laboratories Inc. . . . .                   | 52             | Stanford<br>Telecommunications . . . . .       | 62, 63 |
| EEsof . . . . .                                        | 14             | STC Components . . . . .                       | 46     |
| Electro Abrasives Corp. . . . .                        | 47             | Surcom Associates, Inc. . . . .                | 75     |
| Epson America, Inc. . . . .                            | 61             | Temex Electronics, Inc. . . . .                | 29     |
| Frequency Electronics,<br>Inc. . . . .                 | 49             | Tesoft . . . . .                               | 24     |
| Frequency & Time<br>Systems, Inc. . . . .              | 75             | Thomson Sintra Asm . . . . .                   | 84     |
| Giga-tronics Incorporated . . . . .                    | 18             | Time & Frequency, Ltd. . . . .                 | 25     |
| Henry Radio . . . . .                                  | 37             | Toyocom . . . . .                              | 51     |
| Hewlett-Packard<br>Co. . . . .                         | 19-21, 53-55   | Trak Microwave<br>Corporation . . . . .        | 44-45  |
| Hybrids International . . . . .                        | 37             | Trilithic, Inc. . . . .                        | 22     |
| IFR Systems, Inc. . . . .                              | 2-3            | Tusonix . . . . .                              | 36     |
| Intusoft . . . . .                                     | 83             | Vectron Laboratories, Inc. . . . .             | 72     |
| Janel Laboratories, Inc. . . . .                       | 8              | Verlatone Inc. . . . .                         | 6      |
| JFW Industries, Inc. . . . .                           | 26             | Western Electronic<br>Products Co. . . . .     | 73     |
| KVG . . . . .                                          | 82             | Wide Band Engineering<br>Company, Inc. . . . . | 77     |

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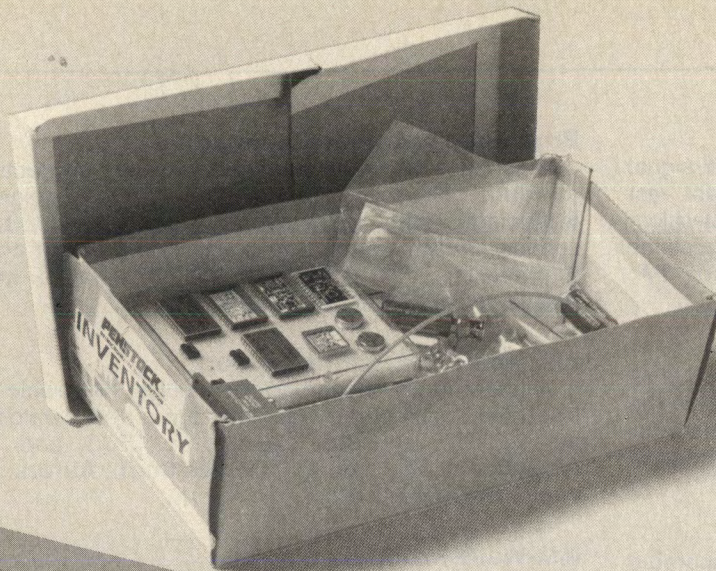
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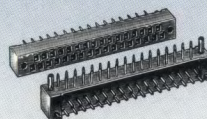
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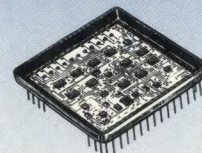
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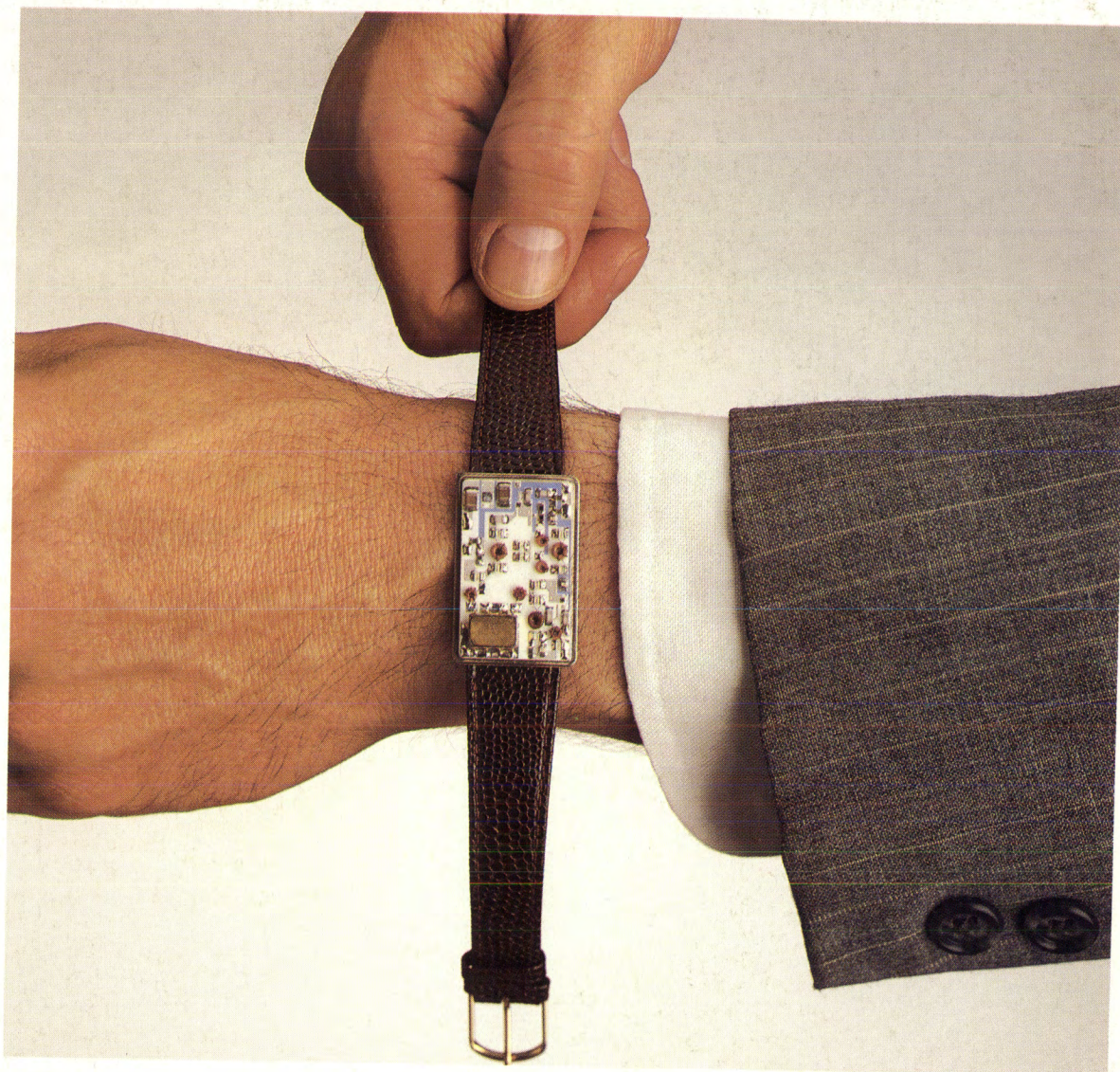


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