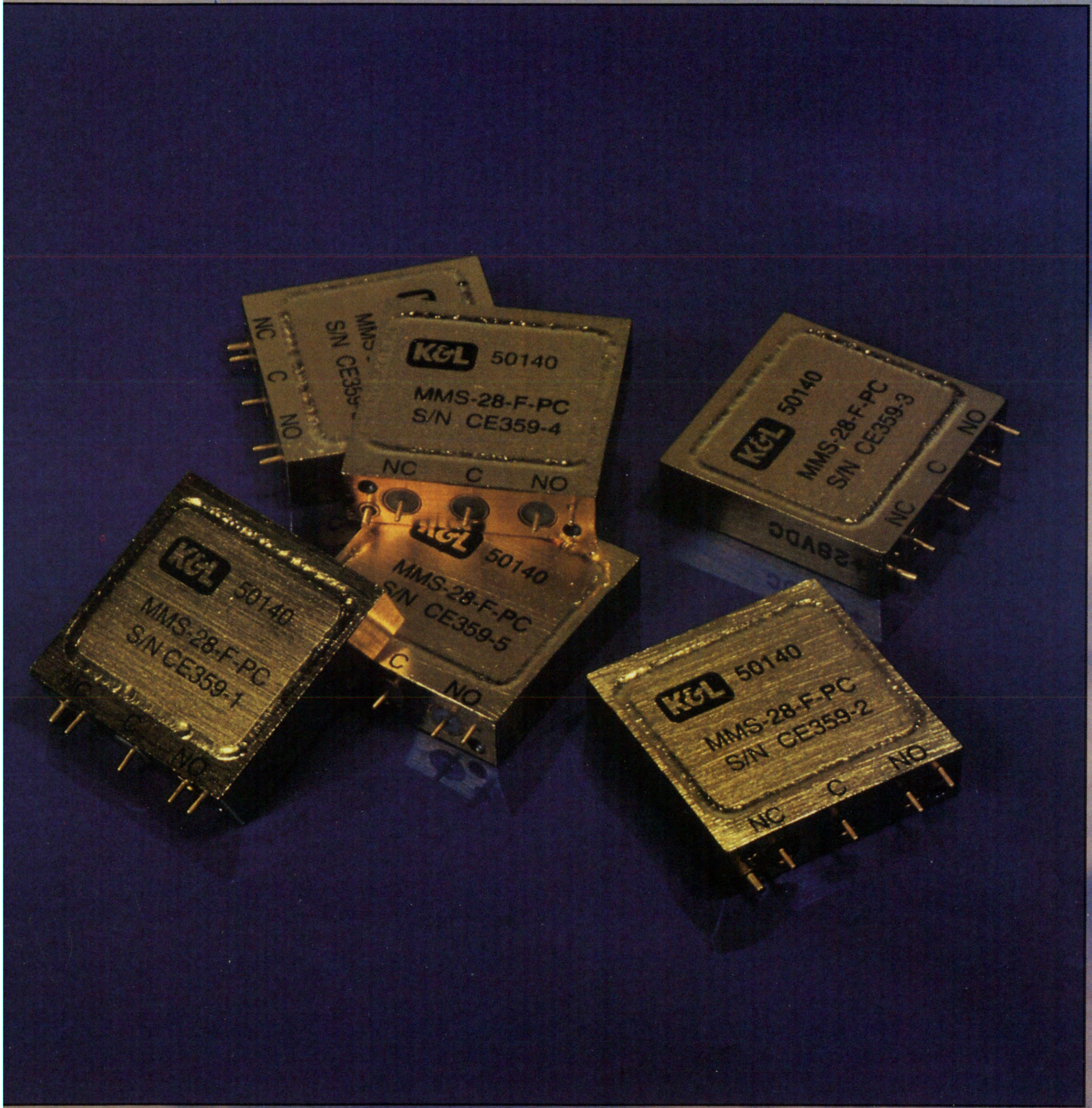
RF design

engineering principles and practices

June 1992



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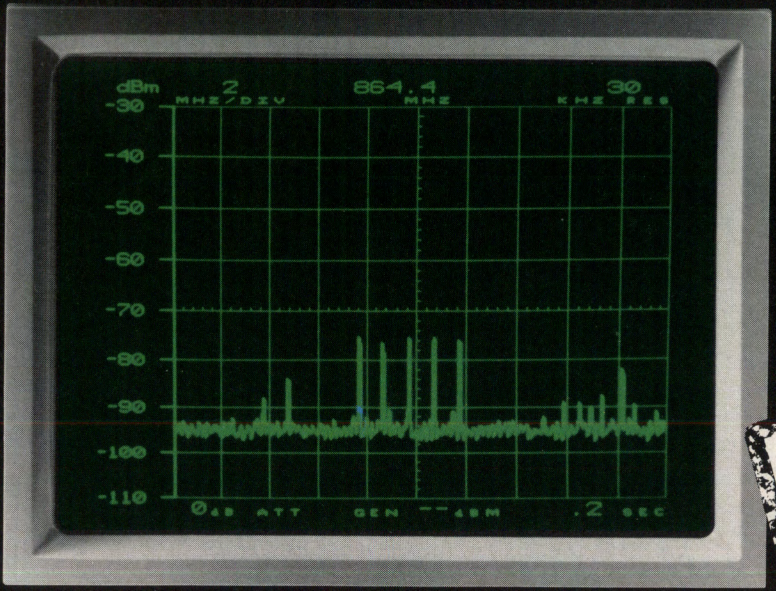

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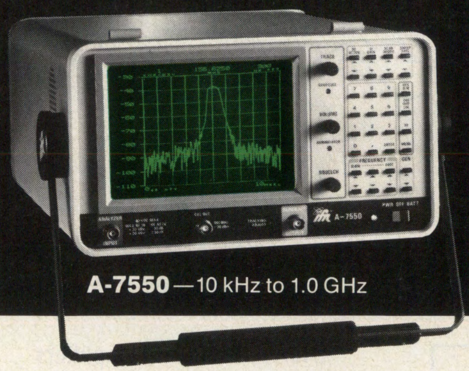

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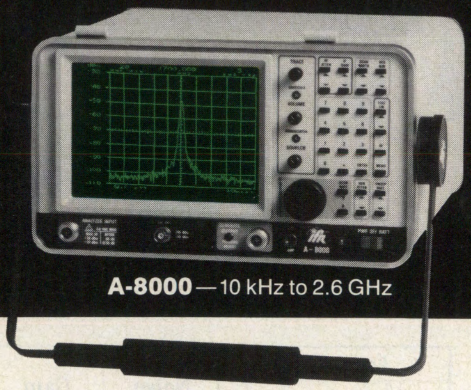

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RF design

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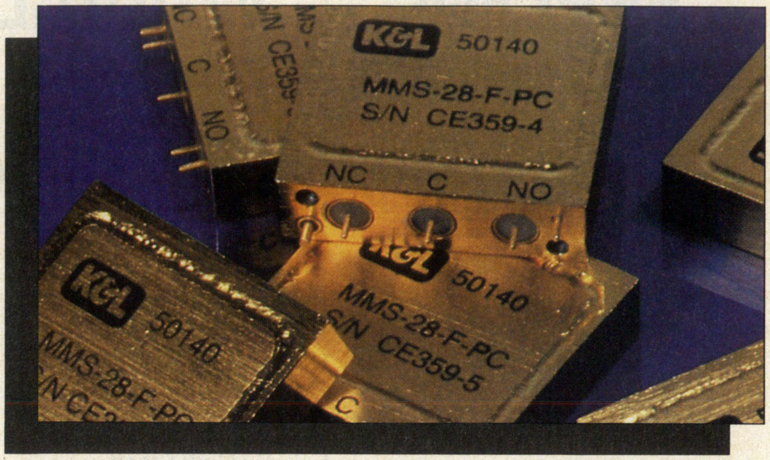

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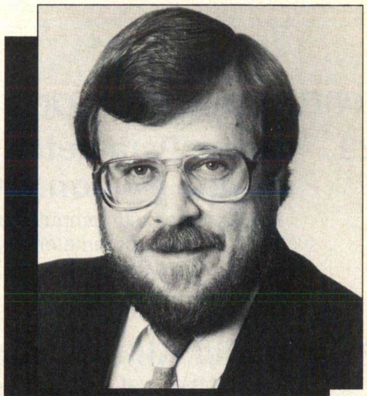
When Does Digital Become RF?



By Gary Breed
Editor

Once in a while I'll get a call or letter from an engineer who is completely anti-digital. This is an understandable position, since digital technology is considered glamorous and gets nearly all the public attention. RF is taken for granted, just getting the job done day after day, and being called on to do even more in the future. It is a bit ironic that this larger role for RF is mainly to carry information in digital form!

But those letters make me stop and think — Is RF completely non-digital? (I mean other than being modulated to carry those infernal ones and zeroes.) Not really. Consider the ubiquitous diode double balanced mixer — to operate, the mixer must be nonlinear, and the best performance comes when that nonlinearity is fast switching, which means square wave drive. Square waves are digital, aren't they? I have used digital ICs to generate the square waves for driving mixers. Does that make me a "bit freak?"

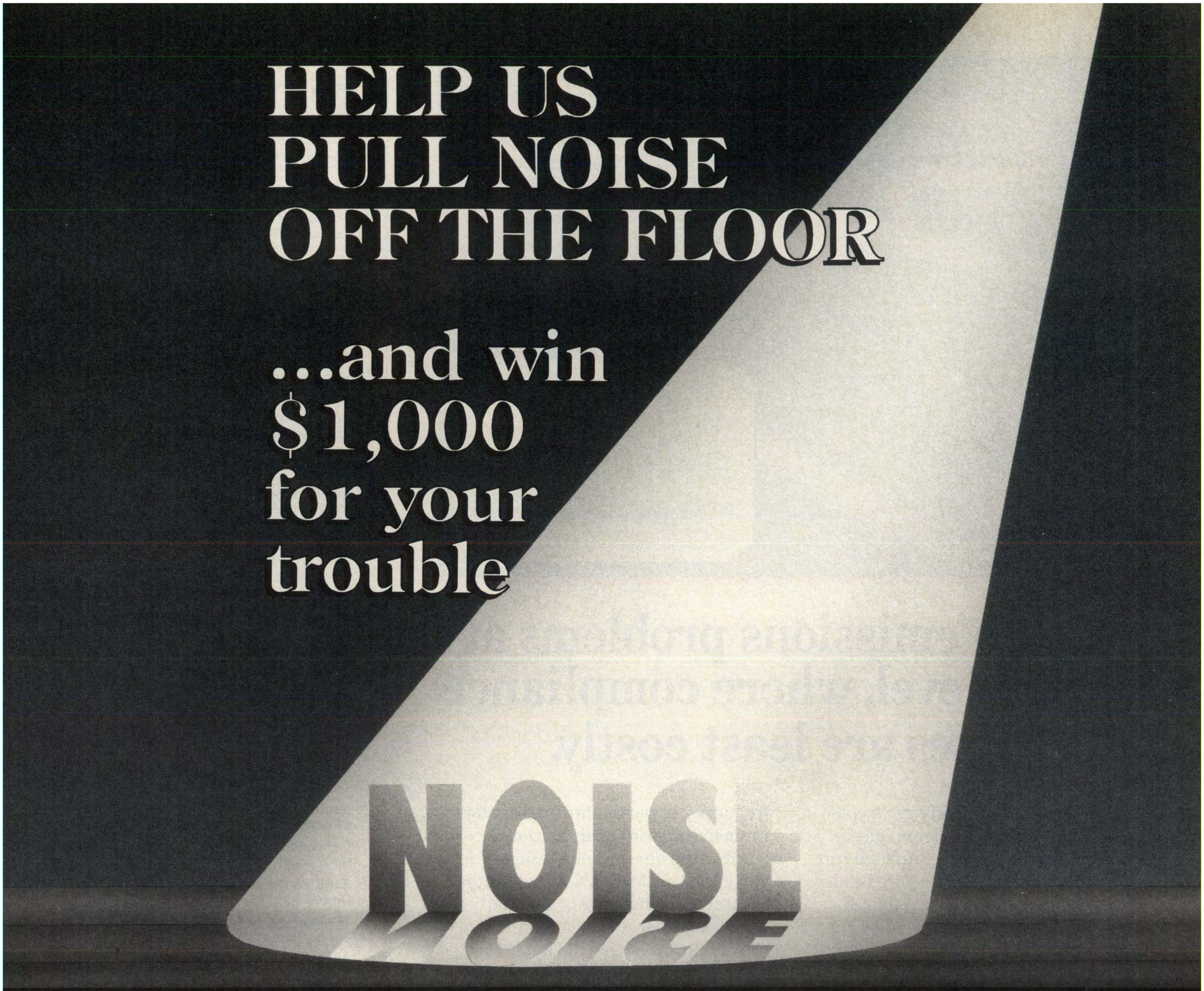
How about a phase-locked loop. Unless you have a 1:1 loop with an analog phase detector, there is digital circuitry involved in the dividers, reference circuitry and phase detector. Of course, all that corrupting digital circuitry is why PLLs never work as well as good crystal oscillators, but they still represent circuits that are digital and RF at the same time.

Now we have a whole host of digital terminology in our RF world: direct digital synthesis (DDS), digital signal processing (DSP), analog-to digital and digital-to-analog conversion (ADC and DAC), fast Fourier transform (FFT), plus all the different modes of modulation for transmission of digital signals.

The time has finally arrived when RF and digital are irreversibly connected.

It is now a viable choice to replace analog IF signal processing and demodulation with a DSP system, if it will meet cost and performance goals. Analog instruments are now adding digitizers and FFT analysis for added utility and precision. These applications aren't limited to low frequencies and baseband, either. DSP at 50 MHz and higher is possible, and new digital downconversion products are being introduced for applications needing greater precision that can only be done at lower frequencies.

You've heard the old saying among RF die-hards — that it is easier for RF engineers to learn digital techniques than it is for digital engineers to pick up RF expertise. If this is true, then RF engineers will be able to learn how to use these exciting new problem-solving digital methods. We still may not get any headlines, but that is OK with me. It's not so bad being considered well-established and reliable enough to be taken for granted.



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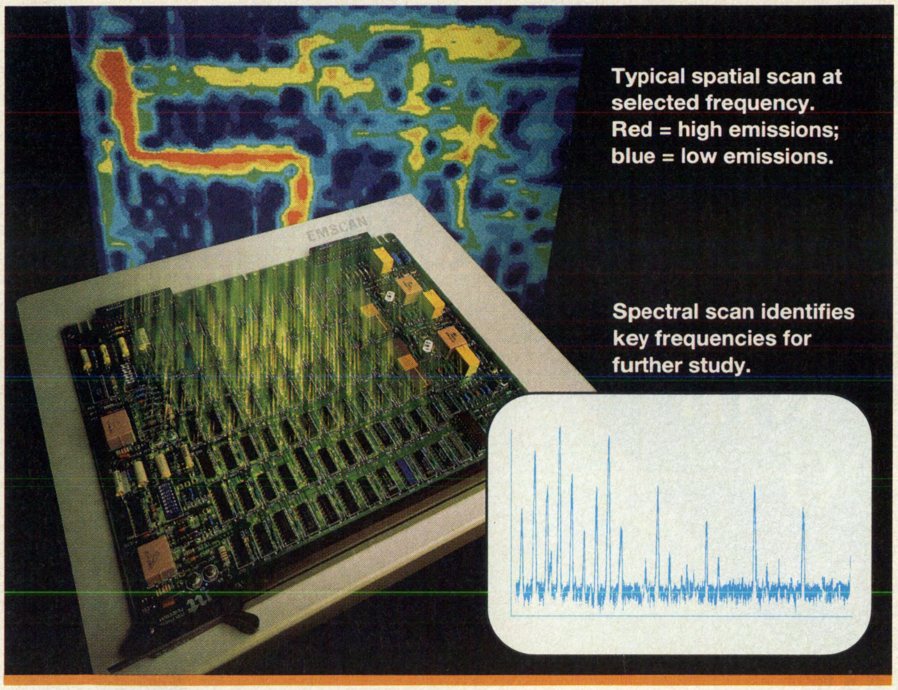
INFO/CARD 5



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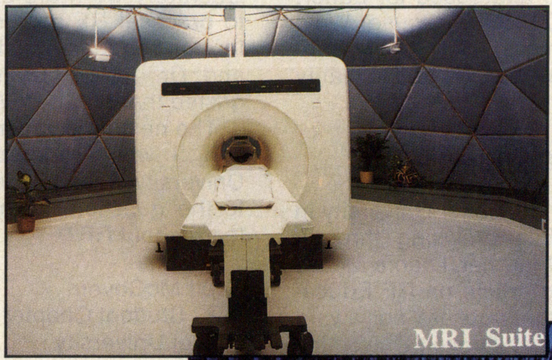
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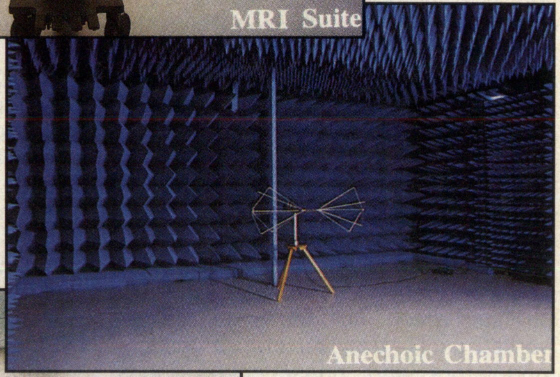
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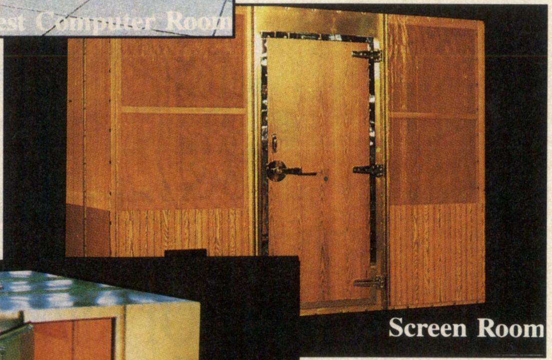
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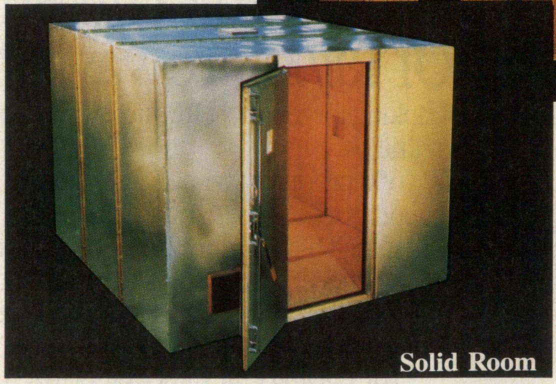
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INFO/CARD 7

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RF Engineering Education

Editor:

I am a student at Drexel University in Philadelphia. While browsing through your August 1991 journal, I came across your editorial and article on RF Education. It appears that you are not aware of RF education at Drexel University. Since Drexel is a cooperative school, its students get exposed to both classroom and practical training. Consequently, professors must keep pace with the current technology, since students returning from their co-op jobs have questions pertaining to current design practices.

As an example of the program, one senior sequence is Advanced Electronic Circuits, which includes matching networks, oscillators, mixers, amplifiers, modulators and phase-locked loops, among other topics. Classroom and laboratory work are specifically geared to practical RF design. The instructors for this sequence are Dr. Bahram Nabet and Dr. Peter Lewin, who also suggested that I read *RF Design*.

The next time you mention RF education, I would appreciate it if Drexel University was included in your answer to the question, "Where do engineers learn about RF?"

Scott McGovern
V.P., Student Chapter of IEEE
Drexel University

Smith Chart Correction

Editor:

Just a short note thanking you for an excellent magazine. Of all the magazines I subscribe to, yours has the most dog-eared pages!

I'd like to point out a small error in the recent article by Neal Silence, "The Smith Chart and its Usage in RF Design." In the first example, illustrated in Figure 6, the calculation of the shunt inductance should have the normalized susceptance of 0.29 divided by Z_0, not multiplied. This results in a susceptance of 0.0058 mhos, or 172.4 ohms, rather than the published value of 14.5 ohms.

This brings up an interesting dilemma in using the Smith chart. While the chart can certainly make one's life easier when transforming impedances, Murphy's Law does not yield to simplicity! It is useful to check the results of any Smith chart calculation with the common series to parallel conversion equations. Had that been done in this example, the matching network would be shown to have an output impedance of 2.19 −j26.7 ohms, not 50 +j0.

Robert J. Fontana, Ph.D.
Multispectral Solutions, Inc.

We take all the blame for this error, not Mr. Silence. In the process of preparing the article for publication, this example was added to his original manuscript. Unfortunately, the required care was not taken, and the checking recommended by Dr. Fontana was not done. The other example, a stub tuner, is correct. — *Editor.*

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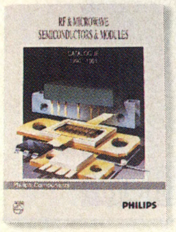

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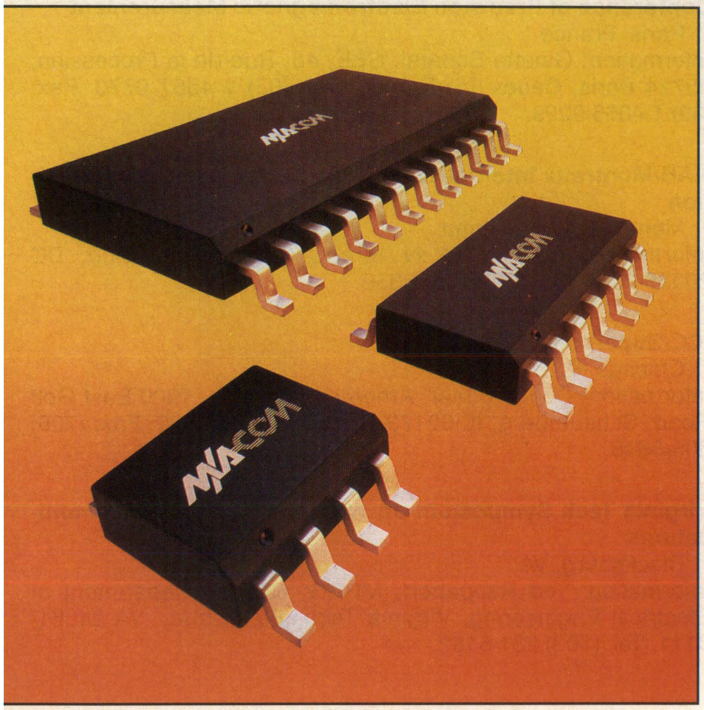

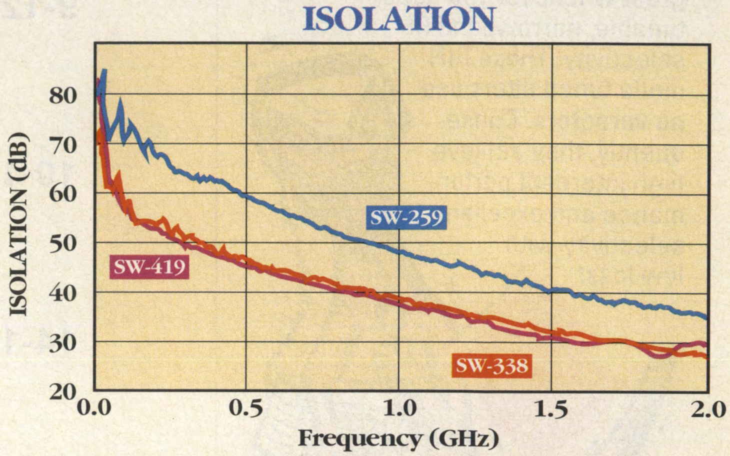

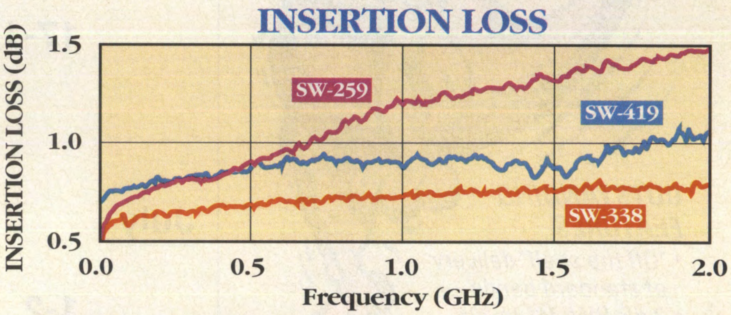

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SW-338	SPDT Non-Reflective	0.7	40	1.2:1	46	SOIC, 8 lead
SW-339	SPDT Non-Reflective	0.7	36	1.2:1	46	SOIC, 8 lead
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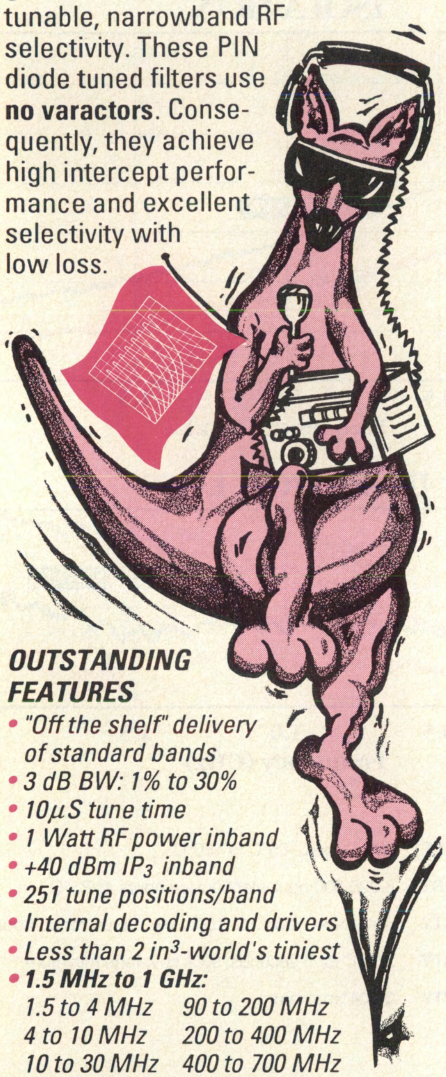

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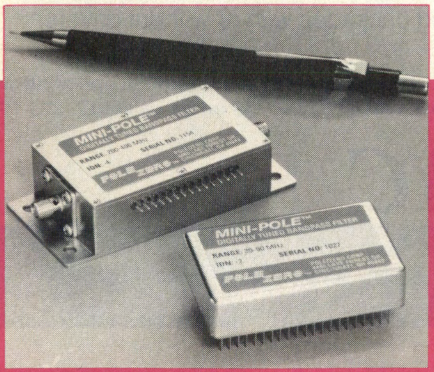

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INFO/CARD 10

RF calendar

June

9-12 **Conference of Precision Electromagnetic Measurement**
Paris, France
Information: Ginette Bonami, SEE, 48, Rue de la Procession, 75724 Paris, Cedex 15, France. Tel: (33) 1 4567 0770. Fax: (33) 1 4065 9229.

10-13 **NAB/Montreux International Radio Symposium and Exhibition**
Montreux, Switzerland
Information: NAB, 1771 N Street, N.W., Washington, DC 20036-2891. Tel: (202) 429-5350.

14-17 **ICC/Supercom '92**
Chicago, IL
Information: Doug Lattner, Ameritech Service, 1900 East Golf Road, Schaumberg, IL 60173. Tel: (708) 605-2500. Fax: (708) 605-3648.

17-19 **Virginia Tech Symposium on Wireless Personal Communications**
Blacksburg, VA
Information: Ted Rappaport, MPRG, Bradley Department of Electrical Engineering, Virginia Tech, Blacksburg, VA 24061-0111. Tel: (703) 231-5182

July

1-2 **JEMIMA Measurement Technology Exhibition**
Nagoya, Japan
Information: Japan Electric Measuring Instruments Manufacturers' Association, 1-9-10 Toranomon, Minato-ku, Tokyo, 105, Japan. Tel: (03) 3502-0601. Fax: (03) 3502-0600.

3-7 **International Broadcasting Convention**
Amsterdam, Holland
Information: Secretary, IBC Convention Office, IEE, Savoy Place, London WC2R 0BL, United Kingdom. Tel: (44) 071 240 1871. Fax: (44) 071 497-3633.

August

11-13 **Eighth Annual Advanced Microelectronic Technology Qualification, Reliability and Logistics Workshop**
Santa Clara, CA
Information: John Farrell, Reliability Analysis Center, PO Box 4700, Rome, NY 13440-8200. Tel: (315) 339-7056.

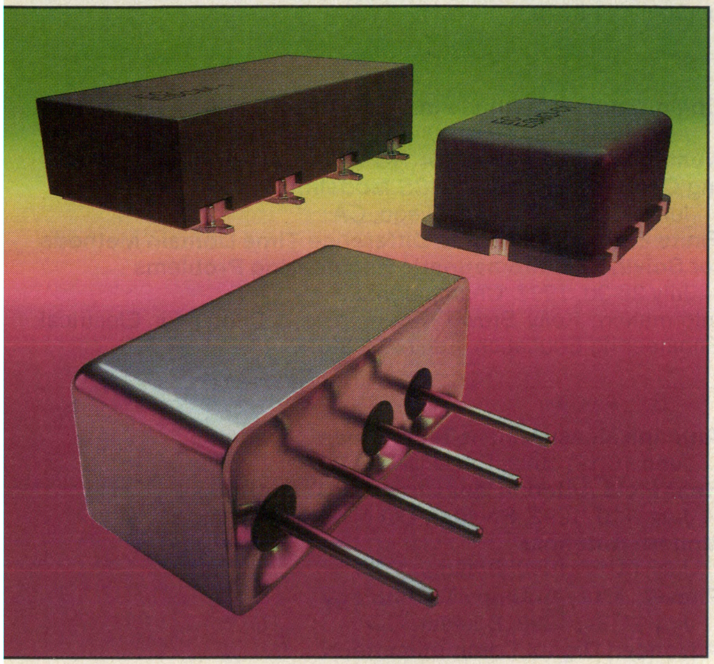
18-20 **IEEE International Symposium on Electromagnetic Compatibility**
Anaheim, CA
Information: Oscar Crawford, Jr., Rockwell. Tel: (213) 922-4091.

24-26 **22nd European Microwave Conference**
Helsinki, Finland
Information: Microwave Exhibitions and Publishers, 90 Calverley Road, Tunbridge Wells, Kent TN1 1BR, England.

RF Mixers

Low Cost and High Quality Do Go Together

See Us At MTTS Booth #805



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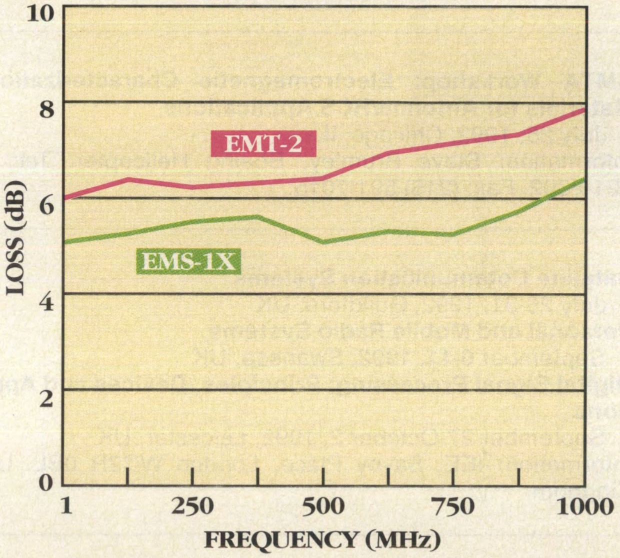
CONVERSION LOSS



DOUBLE BALANCED MIXERS

MODEL	FREQUENCY (MHz) LO/RF	IF	CONVERSION LOSS (dB) (TYP.)	ISOLATION (dB) L-R (TYP.)	L-X (TYP.)	PACKAGE
EMS-1	1 - 500	DC - 500	5.5	45	40	R1
EMA-1	.5 - 500	DC - 500	5.5	45	40	R1
EMS-1X	10 - 1000	5 - 500	6.0	40	40	R1
EMRS-1	.5 - 500	DC - 500	5.5	33	30	SM1
EMA-1H	.5 - 500	DC - 500	5.5	45	40	R1
EMT-2	1.0 - 1000	DC - 1000	6.5	40	35	R3
ESCM-1	1.0 - 500	DC - 500	6.0	45	45	SM3
EMT-1MH	2.0 - 500	DC - 500	6.0	40	35	R3
EMT-4	5.0 - 1250	DC - 1250	6.0	40	35	R3
EMSD - C1	1.0 - 1000	DC - 1000	6.5	45	30	SM2

SM1 = .25" x .31" Surface Mount
SM2 = .37" x .49" Surface Mount
SM3 = .38" x .75" Surface Mount

R1 = 8 Pin Relay Header
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All Mixers available in Surface Mount Packages

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SOLDERABILITY 260°C for 5 seconds; 90% coverage minimum.

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INFO/CARD 11

Technical Cellular
July 6-10, 1992, Madison, WI
Surge Protection for Telecommunications
July 27-29, 1992, Madison, WI
Noise Mitigation
July 29-31, 1992, Madison, WI
Information: University of Wisconsin - Madison, Francis P. Drake. Tel: (608) 262-2061. Fax: (608) 263-3160.

Adaptive Signal Processing
August 10-14, 1992, Los Angeles, CA
RF Component Modeling
August 24-26, 1992, Los Angeles, CA
Information: UCLA Short Course Program Office. Tel: (213) 825-3344. Fax: (213) 206-2815.

AMTA Workshop: Electromagnetic Characterization of Materials for Antenna/RCS Applications
July 25, 1992, Chicago, IL
Information: Steve Brumley, Boeing Helicopter. Tel: (215) 591-2302. Fax: (215) 591-7015.

Satellite Communication Systems
July 26-31, 1992, Guildford, UK
Personal and Mobile Radio Systems
September 6-11, 1992, Swansea, UK
Digital Signal Processing: Principles, Devices and Applications
September 27-October 2, 1992, Leicester, UK
Information: IEE, Savoy Place, London WC2R 0BL, United Kingdom.

Time and Frequency Seminar
June 23-25, 1992, Boulder, CO
Information: Patsy Tomingas, NIST. Tel: (303) 497-3276.

Electromagnetic Interference and Control
June 1-5, 1992, Washington, DC
Grounding, Bonding Shielding and Transient Protection
June 8-11, 1992, Orlando, FL
Spread Spectrum Communications Systems
June 15-19, 1992, Washington, DC
Radio Frequency Spectrum Management
June 15-19, 1992, Washington, DC
Mobile Communications Engineering
July 8-10, 1992, Washington, DC
Trends in Digital Signal Processing
July 20-24, 1992, Washington, DC
Analyzing Communications System Performance
July 22-24, 1992, Washington, DC
Hazardous Radio-Frequency Electromagnetic Radiation
July 28-30, 1992, Washington, DC
Global Positioning System: Principles and Practice
September 9-11, 1992, Washington, DC
Lightning Protection
September 10-11, 1992, Washington, DC
Modern Receiver Design
September 14-18, 1992, Washington, DC
Microwave Radio Systems
September 16-18, 1992, Washington, DC
Information: The George Washington University, Continuing Engineering Education, Merril A. Ferber. Tel: (202) 994-8522 or (800) 424-9773.

Computational Methods in Electromagnetics
June 15-18, 1992, San Diego, CA
Numerical Techniques for RCS Computation and Scattering Center Approach to RCS Modeling
June 15-18, 1992, San Diego, CA
Finite Element and Finite Difference Time Domain Methods for Solving Electromagnetic Engineering Problems
July 28-30, 1992, Champaign, IL
Information: Kelly Brown, Southeastern Center for Electrical Engineering Education. Tel: (407) 892-6146.

Antenna Measurement Techniques
June 16-19, 1992, Rockville, MD
Near-Field Antenna Measurement Techniques
July 7-10, 1992, Boulder, CO
Modern Antennas
July 14-17, 1992, Boulder, CO
Phased Array Antenna Technology
July 21-24, 1992, Boulder, CO
Advanced Digital Communications
August 24-28, 1992, Anaheim, CA
Information: Technology Service Corporation, Lynda S. Epstein, Training Coordinator. Tel: (301) 565-2970. Fax: (301) 565-0673.

The EC Directive on EMC
August 17, 1992, Anaheim, CA
Information: Technology International, Inc. Tel: (804) 644-7735 or (800) 242-8399.

Modern Microwave Techniques: Measurements, Signal and Network Analysis, Microwave Products and Systems Characterization
July 13-17, 1992, Singapore
Digital Signal Processing in Modern Communication Systems
July 20-24, 1992, Singapore
Modern Digital Modulation Techniques
July 27-30, 1992, Singapore
Far-Field, Compact and Near-Field Antenna Measurement Techniques
July 27-31, 1992, Singapore
RF and Microwave Circuit Design: Linear and Non-Linear
July 27-31, 1992, Singapore
Personal Wireless Communications: Cellular Telephony, Portable Computing, and Broadband Wireless Networks
July 27-31, 1992, Singapore
Digital Receivers for Satellite and Mobile Communications
July 27-31, 1992, Singapore
RF and Microwave Component Modeling
July 29-31, 1992, Singapore
Information: CEI-Europe/Elsevier, Mrs. Tina Persson. Tel: (46) 122-175-70. Fax: (46) 122-143-47.

Introducing the premier high speed video op amp – the AD811 from Analog Devices.

What makes the AD811 such a star is that it delivers maximum performance in all the critical specs for video, while costing just $2.85 (in 1000s).

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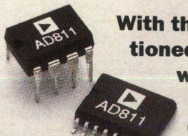

With the specs mentioned above, as well as excellent flatness (0.1 dB to 35 MHz), settling time (50 ns to 0.1% and 65 ns to 0.01%), low noise (1.9 nV/$\sqrt{Hz}$) and low distortion (-74 dB @ 10 MHz), the AD811 will make your video design look great. Also available in an 8-pin SOIC.

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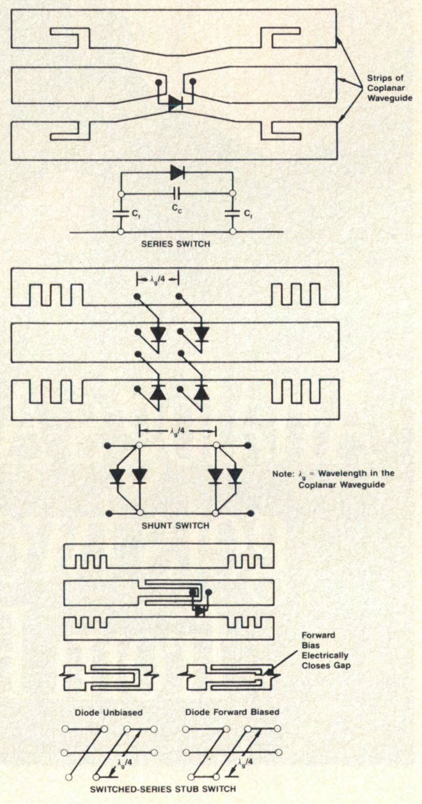
INFO/CARD 12

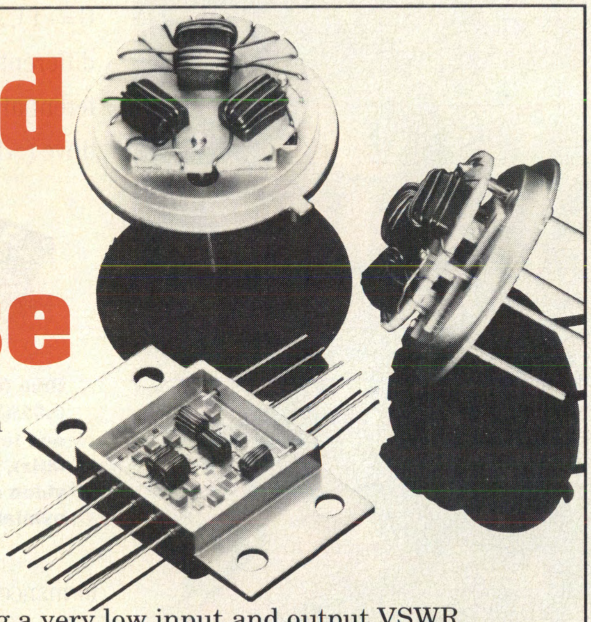
Self-Healing Circuit Technology Developed

Scientists at the GE Research and Development Center have developed technology for designing "self-healing" chips — next generation integrated circuits that will police themselves for errors caused by malfunctioning circuit elements and produce signals that compensate for error the faulted elements would otherwise cause. This new methodology lends itself to the design of both digital and analog fault-tolerant integrated circuits of the type whose behavior is represented by state-variable equations. Such circuits, including a large class of filters and controllers, are widely used for diverse control and signal-processing applications. The circuit works by computing "checksum codes" - specified weighted linear sums of the terms on both sides of the state equations that the primary circuit solves in the course of performing its function. If there is a fault, the checksums do not agree, and an error is signaled. The checking circuit then does an error-check of itself, and, if okay, it computes the error value and automatically feeds this value back to the main circuit for error correction.

New Waveguide Switches — Researchers at NASA have developed a series of three PIN-diode coplanar waveguide switches as alternatives to microstrip for microwave and millimeter wave applications. Depending on how the PIN diodes connect the transmission lines, the transmission line will either act like a transmission line or a filter. When the diode is forward biased, it acts as a conductor and the transmission line is re-evaluated to see what properties it will exhibit. The switches exhibit greater than 20 dB isolation with insertion loss of less than 1 dB. The three versions of the CPW PIN-diode switches are shown in the figure below. This story originally appeared in the April 1992 issue of *Nasa Tech Briefs.* Further information can be



Avoid the Noise



Q-bit Corporation amplifiers are low on noise, high on performance.

These amplifiers utilize our patented Power Feedback™ technology providing a very low input and output VSWR, extremely flat gain response over a wide bandwidth and unconditional stability for any source and load impedance.

Noise figure, 1 dB compression point, reverse isolation, third and second order intercept points are specified and GUARANTEED over the full military temperature range (-55°C to +85°C).

Rather than give up other performance characteristics for low noise figure, specify Q-bit amplifiers in your commercial and military RF designs.

Guaranteed –55°C to +85°C Performance

Model Number	Frequency Range (MHz)	Gain (dB)	Gain Flatness (dB)		1dB Compression (dBm)		Noise Figure (dB)		Reverse Isolation (dB)		Output Intercept 3rd/2nd (dBm)		Power (V/mA)		Price For Quantity 1-9
			Rm	Temp	Rm	Temp	Rm	Temp	Rm	Temp	Rm	Temp	Rm	Temp	
QBH-117	5-100	16.5	0.4	0.8	4.5	3.0	1.5	1.8	35	34	17/24	16/22	15/11	11	$80
QBH-118	3-100	16.3	0.4	0.8	13.0	11.0	1.9	2.1	35	35	27/38	25/35	15/21	22	$80
QBH-120	5-500	14.5	0.6	1.0	2.0	1.0	2.0	2.3	26	26	14/18	13/17	15/11	11	$95
QBH-841	5-100	19.0	0.5	0.7	4.5	3.0	1.5	1.8	35	34	17/24	16/22	15/11	11	$85
QBH-838	50-500	15.0	0.6	1.0	1.0	0.0	1.5	1.8	25	24	14/18	13/17	15/9	9	$95

Q-bit standard product TO-8 designs, like the amplifiers above, are also available in a flatpack with leads formed for surface-mounting as an option.

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Q-bit Corporation

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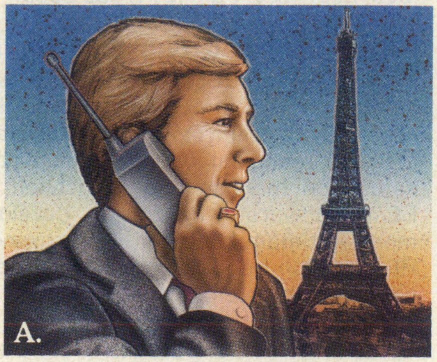
INFO/CARD 13

Please see us at MTT-S, Booth #413.

RF Design News



A.

New quasi-linear modules for GSM digital cellular portable & mobile radios.

Motorola has a quasi-linear module to fit most cellular radios, each with specified performance limits for those critical parameters needed in today's competitive cellular radio marketplace. These compact modules offer small and less complex system designs, as well as savings in time and overall cost. And like all Motorola mobile communications RF products, they meet rigid standards for physical and electrical characteristics even in extreme operating conditions.

B.

Highest power Motorola discrete 900 MHz RF transistor available today.

High gain, good linearity and ruggedness distinguish Motorola's highest RF power, single package, linear transistor for use from 800 to 960 MHz. The silicon bipolar MRF899 delivers 150 watts PEP at 9 dB typical gain, 26 volt, and is housed in a common emitter, push-pull package configuration. The MRF899 is the output part of a lineup comprising a 30 watt driver (MRF897) and 3 watt pre-driver (MRF896) designed for the linear base station requirements of digital cellular radios.

C.

Motorola's highest-power UHF Power Transistor for land mobile radios.

Motorola now offers its highest power UHF power transistor designed specifically for land mobile radios operating from 400 to 520 MHz. Distinguished by its broadband characteristics and extreme ruggedness, the MRF658 delivers 65 watts with a collector efficiency in excess of 50%, and provides over 54 dB gain while operating from a 12.5 volt supply. Packaging is in the popular 6-lead flange.

D.

Low-voltage UHF GaAs FET Power Amplifiers.

Motorola's new MHW9002 Series 900 MHz low-voltage IRIDIUM™ modules offer a wide power range control (30 dB typical*) and operate from a 5.8 volt supply and require only 5 mW of RF input power, thus making them ideal for a wide variety of telecommunications equipment such as portable and cellular applications, satellite cellular applications, gateways and satellite stations. A 1 mW line (MHW9001) is also available. Key features include high efficiency, small size, extreme stability and ruggedness.

E.

Broadband VHF/UHF amplifiers offer low distortion and high output.

Additional broadband VHF/UHF amplifiers with improved distortion characteristics are now available from Motorola. The CA902 operates from 28 volts, provides 17 dB gain over the 40 to 860 MHz frequency range, and is packaged in the conventional CATV style package. The CA912 has identical specifications except for an operating voltage of 15 volts.

*"Typicals" may vary in different applications and are given only to suggest the product's common capability.

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Get more information.

To learn more about any of these Motorola products, contact your local Motorola sales office, complete and return the coupon below to Motorola Semiconductor Products, Literature Distribution Center, P.O. Box 20912, Phoenix, AZ 85036, or call us toll-free any weekday, 8:00 a.m. to 4:30 p.m. (MST), 1-800-441-2447.

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To receive literature on any of these Motorola products, check the applicable box(es) below:

- ☐ A. Quasi-Linear Modules for Portables & Mobiles (SG46/D, Rev. 9)
- ☐ B. MRF899 Linear Power Transistor (SG46/D, Rev. 9)
- ☐ C. MRF658 UHF Power Transistors (SG46/D, Rev. 9)
- ☐ D. MHW9002 Series Power Amplifiers (SG46/D, Rev. 9)
- ☐ E. CA902 VHF/UHF Amplifiers (SG46/D, Rev. 9)

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found in NASA TM-102289 [N90-11943], "Channelized Coplanar Waveguide Pin-Diode Switches." Copies may be purchased (prepayment required) from the National Technical Information Service, Springfield, VA 22161. Tel: (703) 487-4650.

Call for Papers — The Design & Test Expo, formerly known as ATE & Instrumentation Conference, has issued a call for papers for their show to be held January 11-14, 1993 in Anaheim, Calif. Papers must be original, unpublished work or research relating to any and all aspects of test and design for test in the commercial and military environments. A 300-500 word abstract and a biography of the presenter are due by June 15, 1992. Abstracts for papers and tutorial proposals may be sent to Sharon Schifano, Technical Conference Manager, Design & Test Expo, 13760 Noel Road, Suite 500, Dallas, TX 75240. Tel: (800) 223-7126. Fax: (214) 385-9003.

EEsof Users' Group Meeting at MTT-S — The EEsof Users' Group will hold a meeting in conjunction with the MTT-S Symposium in Albuquerque, New Mexico. Technical papers will be presented on the use and application of EEsof software in the design, development and manufacturing of microwave/RF and high-frequency analog devices, circuits and systems.

United Airlines Selects Ball Corporation's Antennas — United Airlines has selected Ball Corporation's AIRLINK® satcom antenna systems for its new Boeing 777 fleet, as well as its other long range fleet. The airplanes will be equipped with satellite communication systems consisting of Ball's AIRLINK high gain antennas; low gain antennas; and high power amplifiers. The systems will be integrated with Collins Avionics SATCOM-906 and will provide multi-channel voice service and data transmission capability for the flight crew and passengers anywhere over land or sea.

Hypres Claims Fastest Digital Logic — Hypres Inc. recently demonstrated a toggle flip-flop operating at 144 GHz, as well as a 4-bit and 32-bit shift register operating at 60 GHz and 45 GHz respectively. Power dissipation of the flip-flop was 1.6 uW, the dissipation for the entire 32-bit shift register was 100 uW. The results were obtained using Rapid Single Flux Quantum superconducting digital logic family. The circuits were implemented in Niobium based technology with 3.5 micron geometries and operated at 4.2 K. Superconducting RSFQ logic will be used for transient digitizers, low-power satellite correlation receivers, digital signal processors, LPI communications and LPI radar.

Relay Conference Call For Papers — A call for papers has been issued for the 41st International Relay Conference to be held in Los Angeles, Calif., April 26-28, 1993. Papers may be tutorial or application-oriented dealing with the following topics: solid state relays, applications, telecommunications, automatic test equipment, or electrical data processing. Abstracts (up to 200 words) in English and on plain paper should

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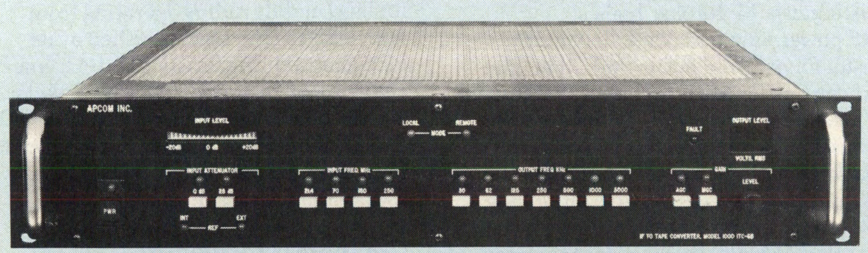

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IEEE-488 control interface simplifies system integration. Just plug it in. A complete software package has been developed for use with the HP-85.

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Model 1000 TIC-6B is the Tape to IF Converter.

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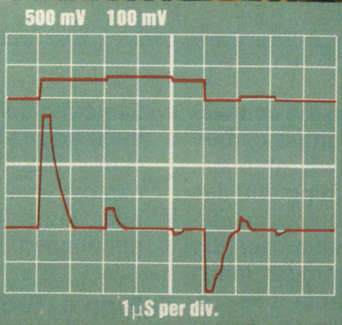
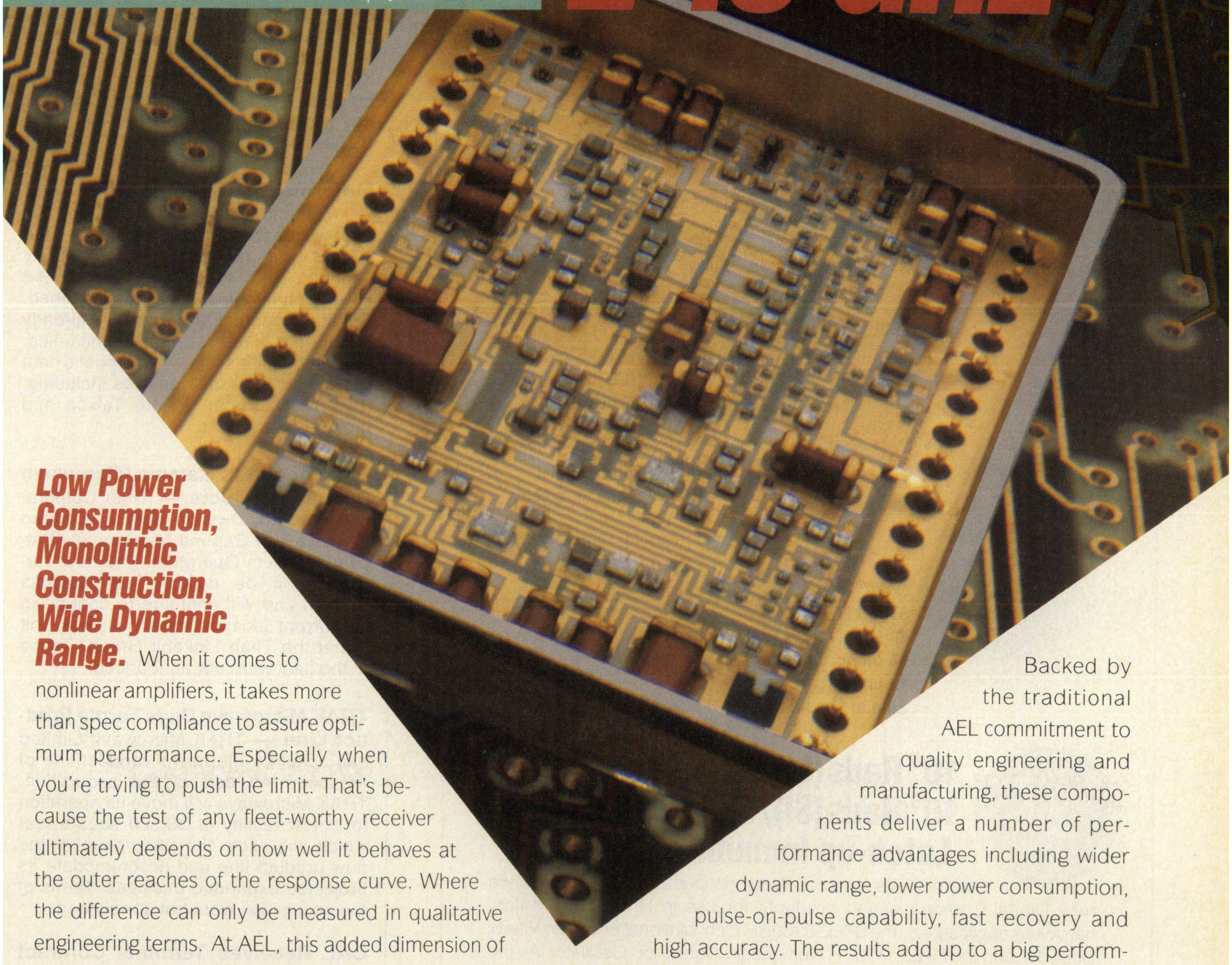
RANGE	− 42 to + 12 dBm
VSWR	1.90 : 1 @ − 20 dBm
OUTPUT RANGE (93Ω)	− 0.33V to − 4.55V peak
FALSE ALARM RATE	< 120 counts/sec, at 170 mV
PULSE WIDTH	200 n Sec nominal
PULSE ON PULSE	± 2 dB
PULSE ON CW	± 1 dB



LOG AMPS 2-18 GHz



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Backed by the traditional AEL commitment to quality engineering and manufacturing, these components deliver a number of performance advantages including wider dynamic range, lower power consumption, pulse-on-pulse capability, fast recovery and high accuracy. The results add up to a big performance edge under difficult ECM and communications conditions. Send for details today!

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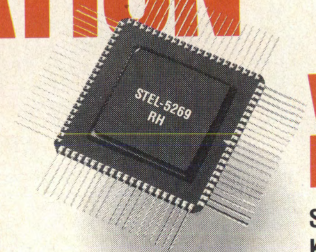
INFO/CARD 15

clearly describe the content, scope and key points of the presentation and should be submitted prior to October 1, 1992 to the International Relay Conference, 512 E.N., Oklahoma State University, Stillwater, OK 74078-0532. Tel: (405) 744-5714. Fax: (405) 744-5033.

Spectrum Control Acquires Murata Erie Product Line — Spectrum Control recently announced the completion of an agreement to purchase certain assets of Murata Erie North America, Ltd. The assets include equipment, jigs and fixtures, manufacturing documentation and customer information for the electronic filter products. Under terms of the agreement, Spectrum has acquired all rights to the following Murata Erie products: bulkhead mounted EMI filters including coax and single series 9000 and 1200, filter circular connector series 38999, filter D connectors and custom manufactured filter networks and plates. Terms of the sale were not announced.

RADIATION HARD ASICs



Viterbi Decoder

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K=7
Rates 1/2, 1/3
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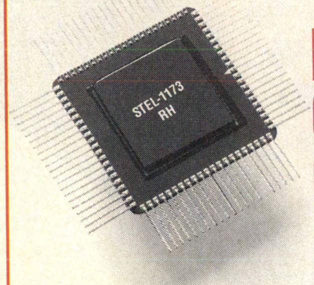
BER Monitor
V.35 Scrambler



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INFO/CARD 16

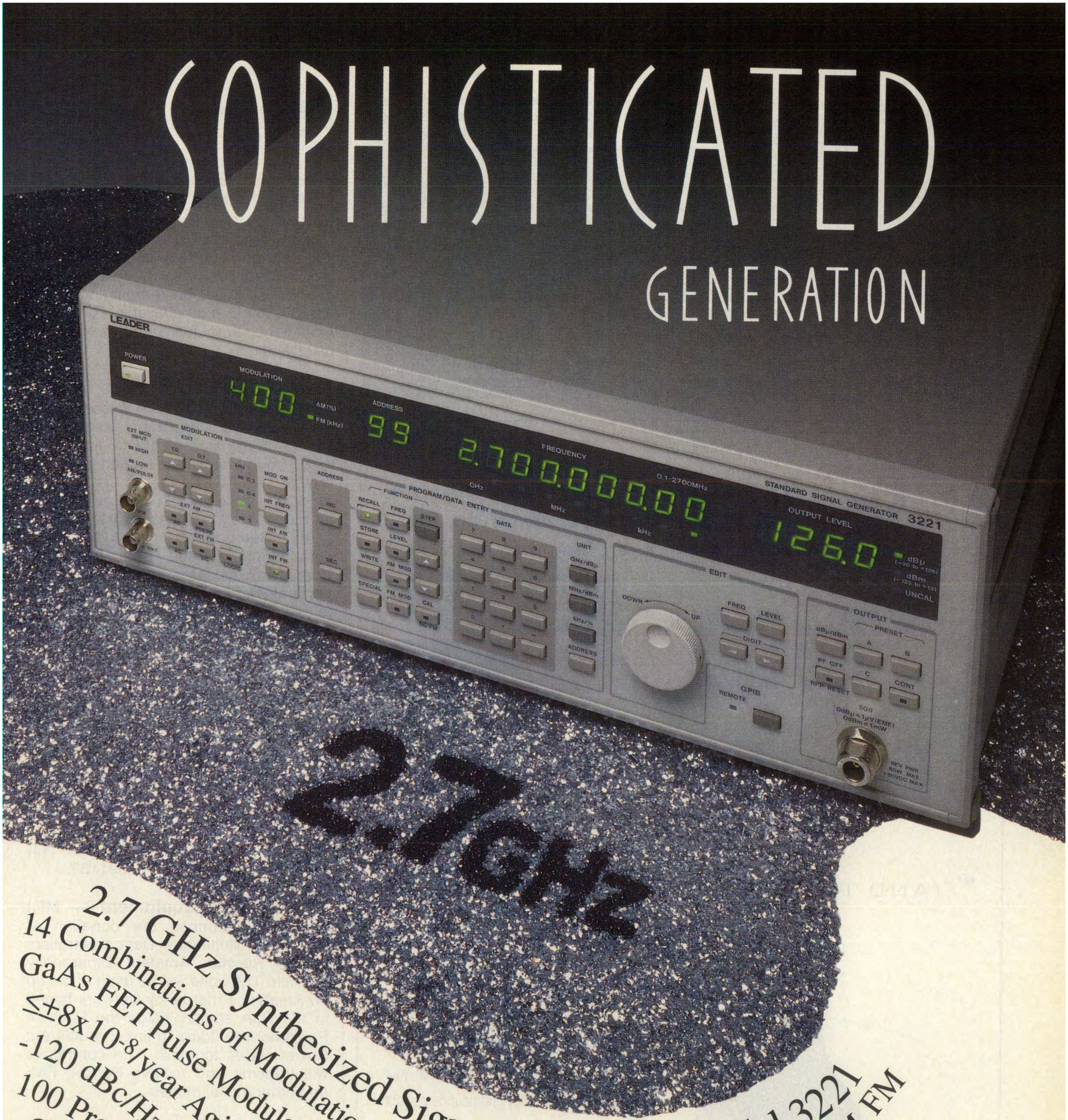
Signal Technology Acquires Keltec Florida — Signal Technology Corporation has acquired the business and substantially all of the assets of Keltec Florida, Inc. The acquisition was made by a subsidiary of Signal Technology which will operate under the name ST Keltec Corp.

Motorola Opens ARDIS Protocol — Motorola recently announced that they will open the protocols used in the ARDIS radio data network to third-party vendors. Both the 4.8 kbps and 19.2 kbps technologies have been opened. Several manufacturers have already integrated wireless data communications capabilities for use with public data networks inside their devices including: IBM, NCR, Poqet, Psion, Telxon and Itron.

Titan Acquires Gamma Microwave Satcom Products — Titan Corporation's Linkabit Division recently acquired the assets of Gamma Microwave's Satcom Products Operation. The new purchase will be called Titan Gamma Satcom and will continue to operate in its current location as a business unit within the Titan Linkabit Group. Terms of the deal were not announced.

TRAK Microwave Buys Ferrite Product Line — TRAK Microwave Limited has acquired the ferrite product line of Albacom. As part of the acquisition, TRAK Microwave will move its operation to a new facility of 38,500 square feet which will allow the absorption of the new product line and accommodate its projected expanded business. Terms of the acquisition were not announced.

CAL Receives Terminal Contract — CAL Corporation has received a $2 million order for GPS-equipped mobile SATCOM terminals, for use in Telesat Mobile Inc.'s satellite based mobile communications and tracking system. With the addition of GPS technology, CAL's terminals will provide worldwide SATCOM coverage, high precision and reliability of transmission, local read-out of GPS coordinates and automatic position reporting.



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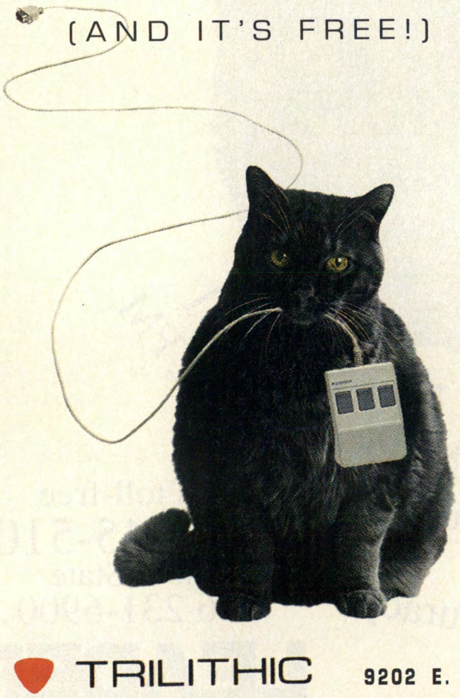

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Broadcast Electronics Licensed to Manufacture Design Automation Circuit — Broadcast Electronics recently received from Design Automation, Inc. an exclusive license to manufacture and sell radio broadcast transmitters using Design Automation's patented Class-E high-efficiency RF power amplifier circuit. License terms were not disclosed.

Electronics Factory Sales Up For First Quarter 1992 — According to the Electronic Industries Association, U.S. factory sales of electronic equipment, components and related products totaled $67.1 billion for the first quarter of 1992, resulting in a 5.6 percent increase over last year's first quarter sales. Components, telecommunications, computers and peripherals, electromedical, consumer, and other related products and services all showed an increase in sales. The only area to show a decline was defense and specialized communications which declined by 2.3 percent.

Trilithic Opens UK Subsidiary — Trilithic Incorporated has announced the opening of Trilithic Limited to serve the UK market for their RF and microwave components and instrumentation including CATV test equipment. Their address is Trilithic Limited, Unit E, Tring Industrial Estate, Upper Icknield Way, Herts HP23 4JX, England. Tel: (44) 442-891139. Fax: (44) 442-891132.

MFJ Donates Equipment — MFJ recently announced the donation of MFJ-1278 Multimode Data Controllers to the Russian Amateur Emergency Service. The units will be used to set up an Amateur Emergency Network based in the R3A station inside the Russian Parliament Building to link the different Soviet states together. The controllers will transmit and receive Packet, FAX and other digital modes.

Philips and Motorola Enter Cross Licensing Agreement — Philips Electronics and Motorola, Inc. have announced a cross-licensing agreement which includes patents in the field of digital cellular radio communications systems. The agreement covers several standards, in particular European GSM, PCN and US-TDMA presently known as IS-54. The removal of barriers associated with patent rights will make it easier for system suppliers to implement an internationally standardized cellular system.

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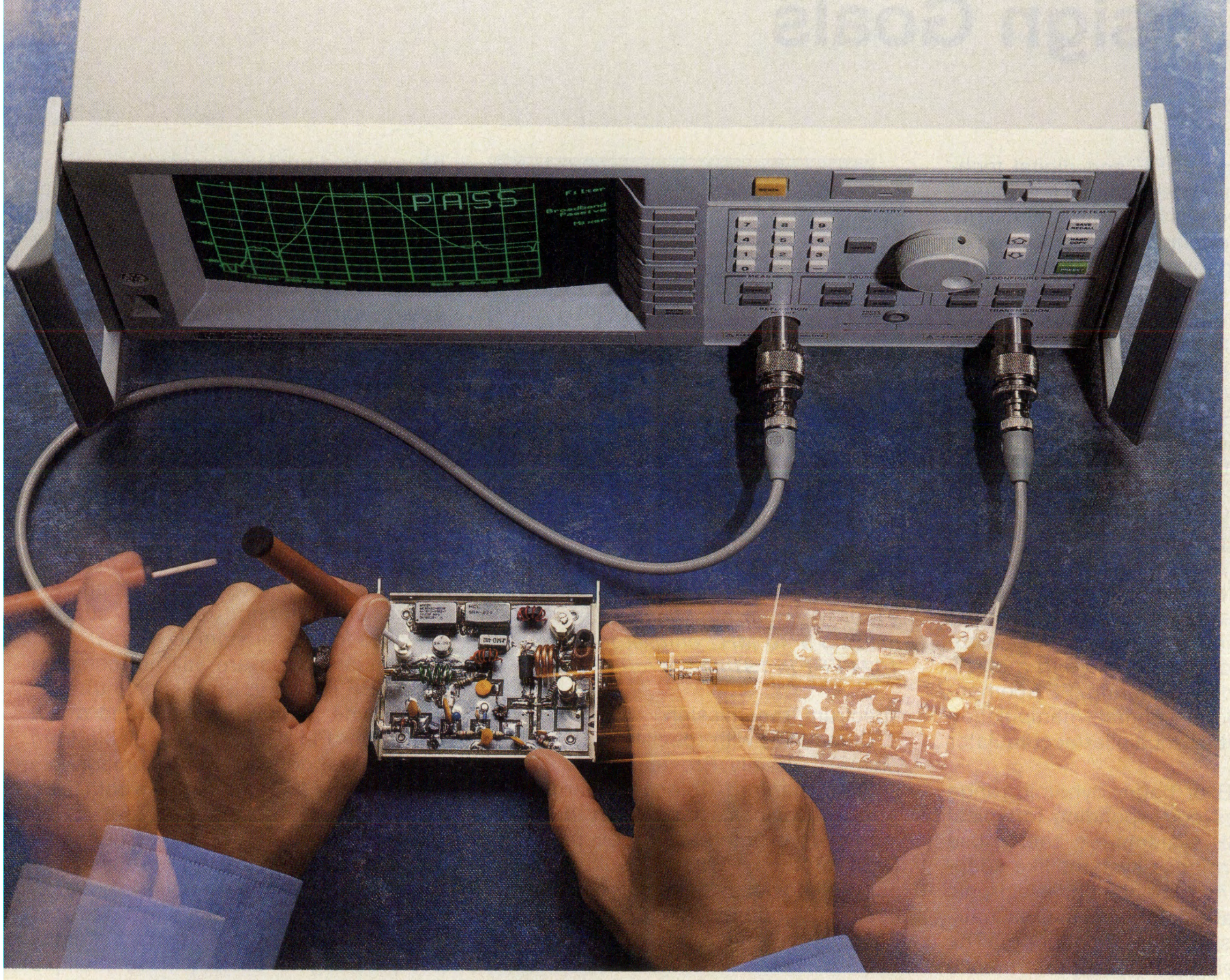

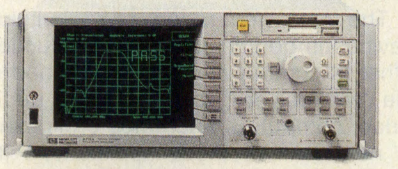

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Digital Techniques Help Achieve RF Design Goals

By Gary A. Breed
Editor

The merging of RF and digital technologies is accelerating. High speed digital components and better familiarity with digital techniques are combining to give RF engineers a new set of tools to apply to their design assignments.

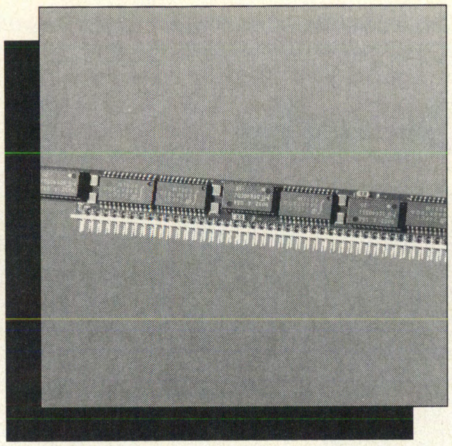
Analog-to-digital conversion (ADC) and digital signal processing (DSP) have reached the point where 12-bit accuracy can be easily obtained into the tens of MHz, with 16-bit accuracy to a couple of MHz. Many RF engineers are taking advantage of low-cost evaluation and development kits from Analog Devices, Burr Brown, Motorola and array Microsystems to learn about DSP techniques.

Recently, these and other companies like Comlinear, TRW, Datel and Harris have dramatically increased their high-speed product lines, each with a different focus — digital filtering, FFT computation, or more general manipulation through flexible multiplier/accumulator architectures.

With 70 dB or better dynamic range at frequencies corresponding to typical IFs, ADC/DSP systems are seeing use in high performance systems like adaptive radar, instrumentation, electronic surveillance and countermeasures, navigation and spectrum monitoring. The lower frequency, higher performance systems are already in place in most major spectrum analyzer manufacturers' equipment, offering extremely fine frequency and amplitude resolution using FFT analysis.

Digital Synthesis

Direct digital frequency synthesis (DDS) has been a significant RF technology for nearly ten years. Recently, acceptance and understanding of DDS has placed it near the top of the list for consideration for any signal generation application. The limitation of finite spurious performance (due to both digital truncation and digital-to-analog converter nonlinearities) is often secondary to the excellent close-in phase noise and completely flexible phase and frequency control. This flexibility has come to the forefront with recent developments in markets that require spread spectrum transmission.



An FFT processing kit is available for $1495 from array Microsystems, Inc., utilizing a 40 MHz chipset to perform a 1024 FFT in 131 us.

Companies with specialized DDS components, mainly Proxim, Sciteq, Qualcomm, Stanford Telecom, Harris and Analog Devices, have seen dramatic increases in interest in their products. Most of these firms also have other support devices to implement specific coding algorithms like Viterbi coding, convolution and forward error correction.

Key to best DDS performance is the construction of the signal in the digital-to-analog converter (DAC). Devices from Sony and TRW have been the most commonly used, and recent products from Analog Devices, Burr-Brown, Comlinear and Tektronix have been targeted specifically for DDS applications.

Digital Techniques

A more subtle influence on RF engineering has been the adoption of design and development techniques used by digital companies. Standardized part numbers, custom integrated circuits, automated layout for both chips and boards, and completely computerized analysis of circuits have been common in digital electronics. RF engineering has been slow to adopt such uniformity, mainly because it could not be achieved until very recently.

It has been only in the past year that application-specific integrated circuits (ASICs) have begun to be seriously considered, although they have seen some limited applications for several years. Various standard cells, tile arrays, and full-custom designs are now offered by Gennum Corp., Harris, Tektronix and Raytheon. High speed digital and/or mixed signal custom ICs have been obtained by RF designers from NCR Microelectronics, AT&T Microelectronics and Oki Semiconductor. To use this option to advantage, designers must shift from a board-level approach to a silicon substrate. In these early stages of development, manufacturers are offering generous support services to their custom IC clients.

Standard parts are still not common in the RF industry. Small-signal transistors, some MMIC amplifiers and a few consumer FM IF ICs are available from multiple vendors. Generic components like resistors and capacitors in both fixed and trimmer designs are pretty well standardized. But most RF components would require a measure of redesign to accommodate a new supplier's product. This is also true at the highest performance levels in digital products, but many complex digital devices have multiple licensed or reverse-engineered competitors.

The next generation of radio-linked devices is getting ready for production. What is coming might be described as the "wireless era" for telephones, audio, video, computers and office equipment. In nearly all of these applications the transmitted data is digital, which means that some digital circuitry will already be on-board. Where that digital circuitry ends and analog RF circuitry begins will be determined primarily by economics. Design managers for these products must decide where this point lies, requiring an understanding of both digital and analog portions of the system. **RF**

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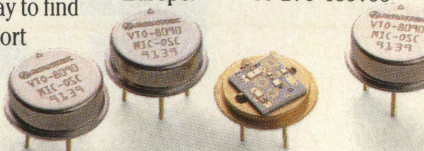

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A Change in the Wind

For years, the reputation of EMI/EMS instrumentation manufactured by Eaton has stood proudly alone in the field. The reliability of these instruments will never be surpassed.

Though, as with most things, there is still change. A change, as with the wind, can be subtle. You will note a change here. The complete line of EMI/EMS instrumentation is now being manufactured by Carnel Labs.

That tradition of quality and reliability will be carried on through Carnel Labs. This line of EMI/RFI instrumentation is not new to the industry nor to the Carnel Labs staff responsible for manufacturing and supporting this strong line of reliable instrumentation.

Carnel Labs is also manufacturing and supporting the former Eaton Model 445 RF power signal source line.

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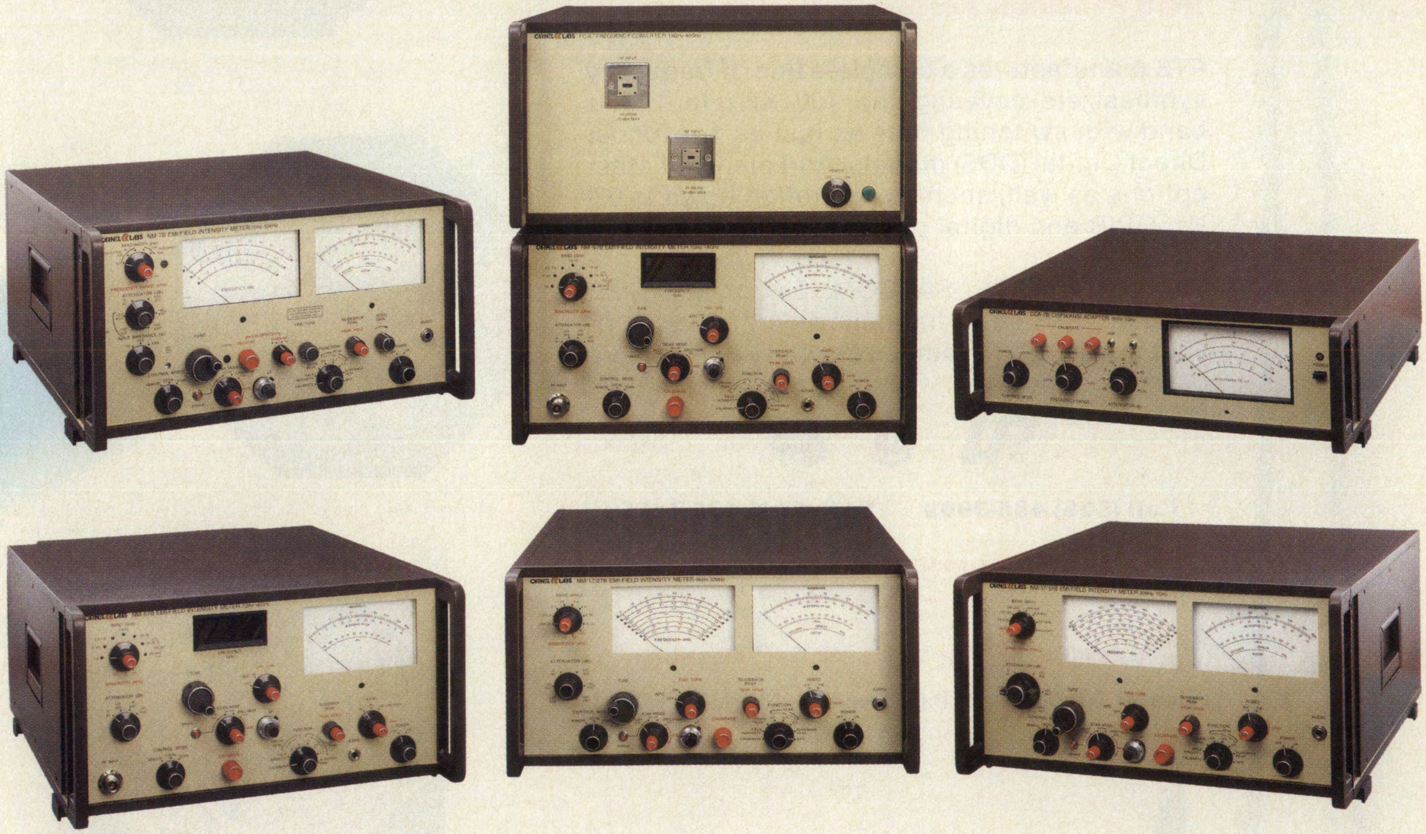

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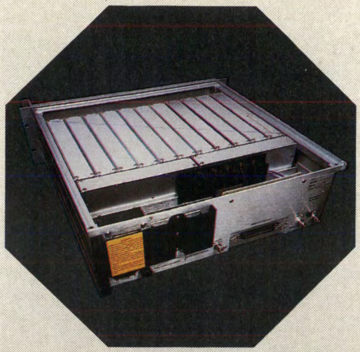
INFO/CARD 21

The Race for Quality . . .



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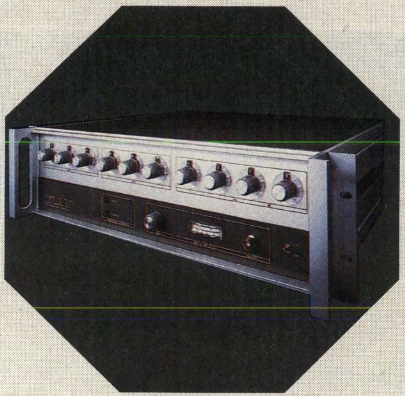

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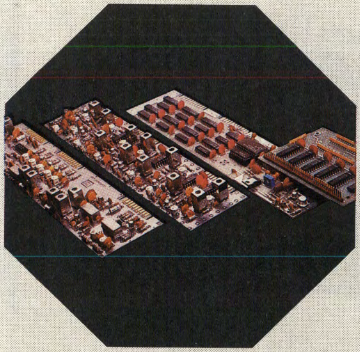

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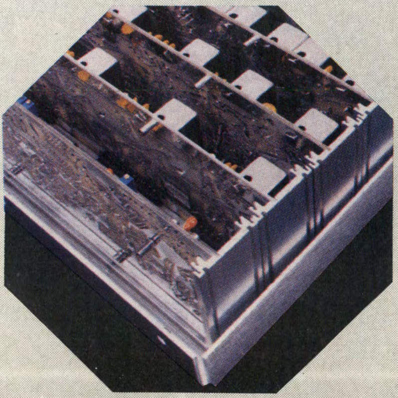

INFO/CARD 22

Noise Parameter Calculations on the Personal Computer

By Malcolm E. Mayercik
SGS-Thomson Microelectronics

The personal computer has eased the burden on the working engineer, allowing calculations, circuit simulations and device performance characterizations to be made rapidly and precisely. Unfortunately, the engineer working with broadband noise sources has not had the opportunity to avail himself of this privilege. However, the computer program described in this article was written with this in mind and calculations can be made quickly and easily. The program was modularized to allow the calculation of several noise parameters and to allow for ease of modification as it continues to evolve in the future.

Calculations regarding noise sources usually fall into several categories, converting ENR to or from power in dBm, to rms voltage across a known load resistor, or to noise spectral density. The basic equation for noise power is:

$$P = k\,T_e\,B \qquad (1)$$

where k is Boltzmann's Constant, 1.380622×10^{-23} Joules/K, T_e is the noise temperature in kelvins, and B is the noise bandwidth in Hertz. P is the noise power in watts. ENR is converted to noise temperature from the equation

$$T_e = T_o[1 + 10^{N_r/10}] \qquad (2)$$

where T_o = 290 K and N_r is the Noise Source ENR in dB.

The ENR of a noise source is defined as the noise in excess of that of a resistor at room temperature. ENR may be calculated if a reference standard's ENR is precisely known, and if the Y factors of that standard and the noise source under test are measured. The Y factor is defined as the ratio of the indicator readings with the noise source switched on to that with the noise source switched off. If the ENR of the unit under test is more than 10 dB higher than that of the standard, and external attenuator will be necessary. Its value is subtracted from the ENR of one unit under test. The system noise figure is calculated from the relation:

$$\text{NF dB} = N_s - 10\log(Y - 1) \qquad (3)$$

where N_s is the standard ENR in dB and Y is the Y factor of the indicator device ("noise figure meter" or otherwise) expressed as a ratio. The ENR of the unit under test is then calculated from the measured Y factor and the system noise figure from equation 4:

$$N_x = NF + 10\log(Y_x - 1) \qquad (4)$$

where N_x is the ENR of the unknown device, NF is the system noise figure in dB and Y_x is the measured Y factor of the unknown device. It is also practical to calculate N_x directly from the value of N_s, which is usually attached to the noise source, and Y, and Y_x, which are measured at the frequency of interest. Combining equations 3 and 4 and rearranging, we have:

$$N_x = N_s - 10\log(Y - 1) + 10\log(Y_x - 1) \qquad (5)$$

where:
N_s = The standard noise source's ENR in dB
Y = The standard's Y factor (the ratio, not in dB)
Y_x = The noise source under test's Y factor (the ratio, not in dB)

```
0 EXIT TO BASIC
1 ENR to dbm
2 ENR to Vrms across load resistor
3 ENR to noise spectral density
4 dbm to ENR
5 Vrms across load resistor to ENR
6 Noise spectral density to ENR
7 ENR calculator
8 db ENR to Kelvins
9 Kelvins to db ENR


Enter Number of choice: ?
```

Figure 1. Main menu for noise program.

Equation 5 may be rewritten if the Y factors are to be used in dB. The rewritten equation 5 becomes:

$$N_x = N_s - 10\log[10^{(y'/10)} - 1] + 10\log[10^{(y_x'/10)} - 1] \qquad (6)$$

where y' is the standard's Y factor and y_x' is the noise source under test's Y factor in dB.

The program is based on equations 1 through 6, and converts the Y factors from dB to ratios. It also makes provision for an attenuator pad, if one must be added to the noise source under test to reduce its level, and to have its contribution corrected. If the pad is to be part of the noise source, "0" is entered. Prompts were inserted to make the program as user-friendly as practical. Tolerances were added to the Y factor readings in the branch beginning on line 1300. A value of 0.03 dB was used as an uncertainty in readings. If a value of

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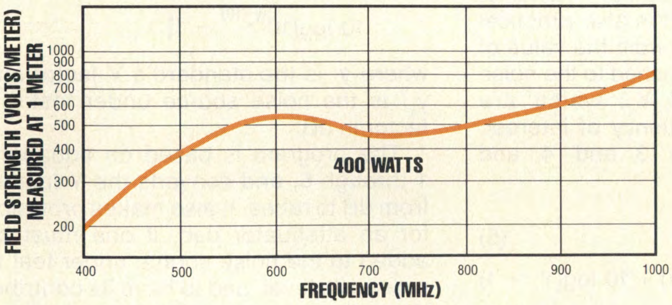

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more or less than 0.03 dB (AA = 0.03 on line 1310) is desired, that value may be used in place of the one in the program.

Using the Program

The program is entered from GWBASIC by loading "NOISE". The main menu options are shown in Figure 1. To

```
6 DIM A$(11)
10 AB=0:CLS
20 REM **** NOISE CALCULATION PROGRAM ****
25 REM **** By M. E. Mayercik ****
30 REM **** SGS - THOMSON MICROELECTRONICS ****
40 REM **** 25 Schoolhouse Road ****
50 REM **** Somerset, NJ 08873 ****
100 REM **** STARTUP AND MENU ****
120 TO=290:K=1.38E-23
130 A$(1)="ENR to dbm"
140 A$(2)="ENR to Vrms across load resistor"
150 A$(3)="ENR to noise spectral density"
160 A$(4)="dbm to ENR"
163 A$(8)="db ENR to Kelvins"
165 A$(9)="Kelvins to db ENR"
170 A$(5)="Vrms across load resistor to ENR"
180 A$(6)="Noise spectral density to ENR"
190 A$(0)="EXIT TO BASIC"
192 A$(7)="ENR calculator"
200 CLS:PRINT:PRINT:PRINT:PRINT
210 FOR I =0 TO 9
220 PRINT I;A$(I):NEXT I:PRINT:PRINT
230 INPUT"Enter Number of choice: ";CN
240 IF CN<0 THEN 230
245 IF CN>9 THEN 230
250 CLS:PRINT A$(I):PRINT:PRINT
253 IF CN=1 THEN GOTO 290
255 IF CN=2 THEN GOTO 360
256 IF CN=3 THEN GOTO 440
257 IF CN=4 THEN GOTO 520
258 IF CN=5 THEN GOTO 620
259 IF CN=6 THEN GOTO 730
260 IF CN=7 THEN GOTO 1000
261 IF CN=0 THEN GOTO 840
262 IF CN=8 THEN GOTO 1700
263 IF CN=9 THEN GOTO 2000
270 GOTO 230
280 REM **** ENR to dbm ****
290 GOSUB 920
300 DBM=10*LOG(PN/.001)/LOG(10)
305 DBM=(.01*INT(100*DBM))
310 PRINT"dbm= ";DBM
320 GOSUB 870
330 GOTO 200
340 :
350 REM **** ENR to Vrms across load ****
360 GOSUB 920
370 INPUT"Rload= ";RL
380 VRMS=SQR(PN*RL)*1000000!
385 VRMS=(.001*INT(1000*VRMS))
390 PRINT"Vrms across load resistor= ";VRMS;" microvolts"
400 GOSUB 870
410 GOTO 200
420 :
430 REM **** ENR to Noise spectral density ****
440 GOSUB 920
450 INPUT"Rload= ";RL
460 V1=SQR(PN*RL)/SQR(BW)*1000000!
465 V1=(.001*INT(1000*V1))
470 PRINT"Vrms per root hertz across ";RL;" ohms = ";V1;" microvolts"
480 GOSUB 870
490 GOTO 200
500 :
510 REM **** dbm to ENR ****
520 INPUT"Noise bandwidth in MHz= ";BW:BW=BW*1000000!
530 INPUT"dbm= ";DBM
540 PN=.001*(10^(DBM/10))
550 TE=PN/K/BW
560 NR=10*LOG(TE/TO-1)/LOG(10)
565 NR=(.01*INT(100*NR))
570 PRINT"ENR= ";NR;" db"
580 GOSUB 870
590 GOTO 200
600 :
610 REM **** Vrms across load to ENR ****
620 INPUT"Vrms across load (uv)= ";VRMS:VRMS=VRMS*.000001
630 INPUT"Rload= ";RL
640 INPUT"Noise bandwidth in MHz= ";BW:BW=BW*1000000!
650 PN=VRMS^2/RL
660 TE=PN/K/BW
670 NR=10*((LOG(TE/TO-1))/(LOG(10)))
675 NR=(.01*INT(100*NR))
680 PRINT"ENR= ";NR;" db"
690 GOSUB 870
700 GOTO 200
710 :
720 REM **** Noise spectral density to ENR ****
730 INPUT"Vrms per root hertz across load (uv)= ";V1
740 INPUT"Rload= ";RL
750 INPUT"Noise bandwidth in MHz= ";BW:BW=BW*1000000!
760 VRMS=V1*SQR(BW)*.000001
770 PN=VRMS^2/RL
780 TE=PN/K/BW
790 NR=((LOG(TE/TO-1))/(LOG(10)))*10
795 NR=(.01*INT(100*NR))
800 PRINT"ENR= ";NR;" db"
810 GOSUB 870
820 GOTO 200
830 :
840 REM **** EXIT TO BASIC ****
850 PRINT"          END OF NOISE PROGRAM"
860 END
870 REM **** WAITING ****
880 INPUT"Hit any key then [enter] to continue ";M$
890 IF M$="" THEN GOTO 880
900 RETURN
910 :
920 REM **** Standard 1-3 input subroutine ****
930 INPUT "ENR= ";NR:NR=NR/10
940 TE=TO*(10^NR+1)
950 INPUT"Noise bandwidth in MHz= ";BW:BW=BW*1000000!
960 PN=K*TE*BW
970 RETURN
990 :
1000 REM **** Noise source "Y" factor to ENR cruncher ****
1010 CLS
1015 PRINT"'Y' factor to ENR calculator. Enter values in db."
1020 INPUT"Pad value: ";A
1025 INPUT"Frequency in GHz: ";F
1030 INPUT"Standard ENR: ";ES
1035 INPUT"Standard 'Y' factor (in db): ";YS
1040 INPUT"Unit under test 'Y' factor (db): ";YX
1050 FS= ES-4.34*LOG((10^(YS/10))-1)
1060 E4= FS+4.34*LOG((10^(YX/10))-1)
1070 EX=E4+A
1080 PRINT:PRINT"Frequency= ";F;" GHz"
1090 FS=(.01*INT(100*FS))
1100 EX=(.01*INT(100*EX))
1190 PRINT"System Noise Figure= ";FS;" db"
1200 PRINT"              Unit under test ENR= ";EX;" DB"
1210 PRINT:GOTO 1300
1220 INPUT"MORE ?  enter [Y] or [N] ";C$
1230 IF C$="y" THEN C$="Y"
1240 :
1250 IF C$="Y" THEN GOTO 1025
1260 GOTO 100
1290 :
1300 REM **** ADD TOLERANCES ****
1310 AA=.03
1320 YS=YS-AA:YX=YX+AA
1330 :
1350 FS=ES-4.34*LOG((10^(YS/10))-1)
1360 E4=FS+4.34*LOG((10^(YX/10))-1)
1370 EX=E4+A
1390 PRINT"Maximum ENR condition"
1400 FS=(.01*INT(100*FA))
1410 EX=(.01*INT(100*EX))
1420 :
1430 PRINT"              Unit under test ENR= ";EX;" DB"
1440 PRINT:YA=YT:YK=YW
1450 AB=AB+1
1460 YS=YS+(2*AA):YX=YX-(2*AA)
1490 :
1500 REM **** ADD TOLERANCES ****
1550 FS=ES-4.34*LOG((10^(YS/10))-1)
1560 E4=FS+4.34*LOG((10^(YX/10))-1)
1570 EX=E4+A
1590 PRINT"Minimum ENR condition"
1600 FS=(.01*INT(100*FA))
1610 EX=(.01*INT(100*EX))
1620 :
1630 PRINT"              Unit under test ENR= ";EX;" DB"
1650 AB=AB+1
1660 GOTO 1220
1700 REM **** db ENR to Kelvins ****
1705 PRINT
1710 INPUT "ENTER ENR in db: ";EX
1720 TK=290*((10^(EX/10))+1)
1730 PRINT "Noise temperature = ";TK;" kelvins"
1740 INPUT "Hit [ENTER] to continue or enter any key to end";A$
1750 IF A$="" THEN GOTO 1700 ELSE GOTO 200
2000 REM **** Kelvins to db ENR ****
2005 PRINT:PRINT"::::::::::::::::::::::::::::::::::::::::::"
2010 INPUT "ENTER noise temperature in kelvins: ";TK
2015 IF TK<=290 THEN 2005:PRINT
2020 EX=4.343*LOG((TK-TO)/TO)
2030 PRINT TK;" Kelvins = ";EX;" db ENR"
2040 PRINT
2050 INPUT "ENTER [C] to continue or any other key to quit ";A$
2060 IF A$="c" THEN A$="C"
2070 IF A$="C" THEN 2000
2080 GOTO 200
```

Figure 2. Program listing.

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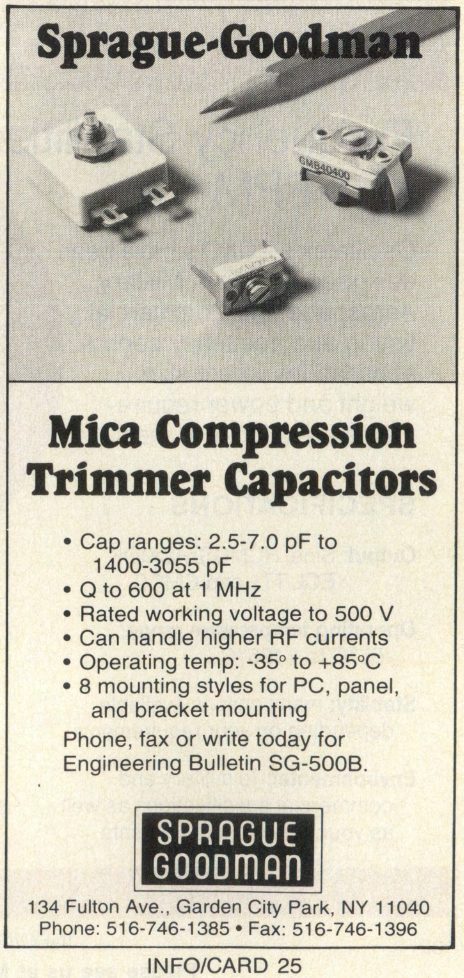

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use the program, select the calculation to be made from the menu by entering a selection, 0 through 9, and hitting the ENTER key. The program will ask for the necessary parameters to make the calculation. Each parameter is entered, followed by ENTER, as it is prompted. After the final parameter is entered, the desired parameter is displayed on the screen. Option 7 allows the ENR of a noise source to be measured, using the ENR of a calibrated standard noise source and Y factor measurements of that standard and the unit under test. Parameters are entered, as requested, in the following order: the attenuation value of an external pad added to the noise source under test, the test frequency in GHz (not used in any of the calculations, but it is useful when trying to remember where you left off), the ENR value of the noise standard (obtained from the manufacturer or NIST), the Y factor of the standard as measured, and the Y factor of the noise source to be tested. The program prints the frequency in GHz, the system noise figure in dB, the ENR of the unit being measured, and the maximum and minimum values of ENR, using an uncertainty of 0.03 dB in the Y factor measurement. The value of 0.03 dB was selected rather arbitrarily, being based on the worst case measured in an actual setup. If another value is desired, the value of AA in line 1310 can be changed to that value, in dB. A listing of the program is shown in Figure 2.

Examples

Several examples are given to illustrate the program's functions. The menu shown in Figure 1 appears on the screen when the program is loaded and the choice to be selected is entered. For example, it is desired to calculate the noise power in a 1 MHz bandwidth, in dBm, produced by a noise source putting out 15.5 dB of ENR. From the main menu, press <1>, ENTER to begin the ENR to noise power calculation. The prompts will come up asking, in turn, for the ENR (in dB) of the noise source and the noise bandwidth in MHz. Enter the values 15.5 ENTER, 1 ENTER. The computer will display the total noise power in the 1 MHz bandwidth as being −98.36 dBm. Hitting any key and pressing ENTER returns to the main menu.

To measure the ENR of any noise source, use the ENR calculator (item 7 on the menu), enter the program by pressing <7>, ENTER. The prompt will ask for the pad value. This is the value of an external pad used on the device under test. It will then ask for the ENR of the calibrated standard noise source (from the manufacturer or NIST), the measured Y factor of the standard and the Y factor of the device under test. The display will then show the inputted data and the calculated ENR of the device under test. The remaining modules convert noise parameters.

The programmable calculator and design curves have been relegated to occasional use in the lab (only when the computer is not available) since this program was written. While it is not intended to be all things to all people, this program has allowed the writer to operate more efficiently, and the algorithms have also allowed the generation of programs for automatically conducting measurements on the Noise Source product line.

This program is available on disk from the RF Design Software Service. See page 87 for ordering information. **RF**

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2. Seymour Schneider, "A Noise Nomograph", *RF Design*, June 1985, pp. 48-50.
3. Microwave Semiconductor Corporation, "System Noise Figure Monitoring", Application Note NM-201, October 1983.
4. Hewlett-Packard Company, "Noise Figure Primer", Application Note 57, June 1962.
5. Ailtech, a Cutler-Hammer Company, "Noise Slide Rule", copyright 1973.

About the Author

Malcolm E. Mayercik is a senior manufacturing engineer for SGS-Thomson Microelectronics. His present responsibilities are product engineer for solid state noise source product line, and project engineer for solid state power amplifiers. He can be reached at 25 Schoolhouse Road, Somerset NJ 08873. Phone (908) 563-6252.

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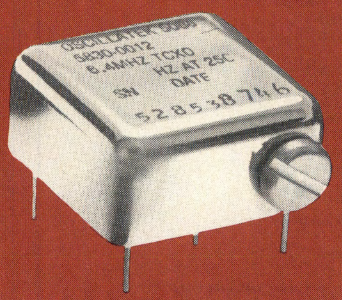

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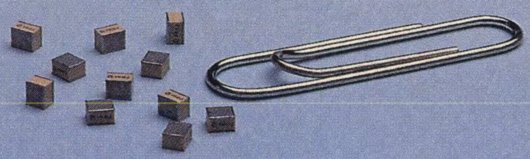

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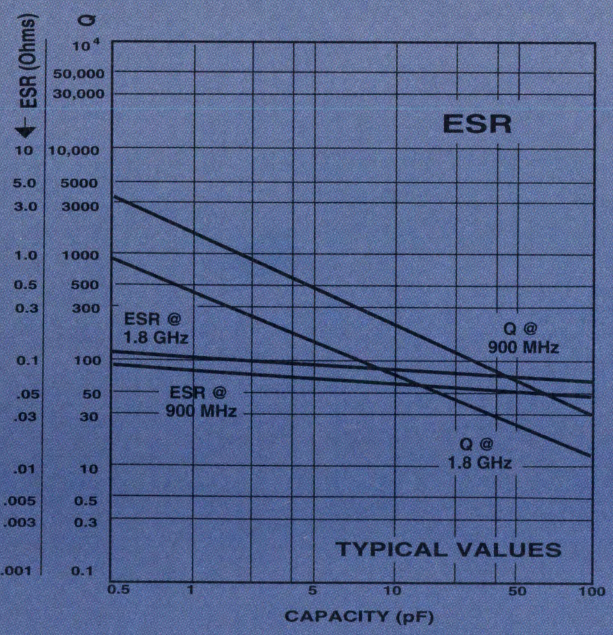

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Avoid Killer Avalanches (and h_{FE} Degradation)

By Don C. Schultz
California Eastern Laboratories

Avalanche conduction of the base-emitter junction is a recognized cause of h_{FE} degradation in silicon bipolar junction transistors (BJT's). Performance degradation of RF power gain, noise figure, oscillator phase noise and frequency stability are all associated with reverse B-E conduction. Unfortunately, these phenomena may not be recognized for many hours after operation has begun. When finally detected, the cause of failure is often judged to be "poor component reliability," while the true cause is improper circuit design.

This paper reviews the causes of h_{FE} degradation and presents a theoretical analysis useful in predicting the maximum RF input that can be tolerated without h_{FE} degradation. $P_{AVALANCHE}$ is described by a combination of circuit and device parameters which are easily determined. Actual test results of +20 and +30 dBm output devices at 1.6 GHz are compared to predictions using this analytical method. A general approach to bias circuit analysis and design that results in the least amount of reverse B-E conduction and the highest reliability for a given amount of RF input power is described.

Effect of Avalanche Conduction

McDonald published results in 1970 showing degradation of h_{FE} due to avalanche conduction of B-E junctions in planar diffused structures. During avalanche breakdown there is sufficient kinetic energy in the ionized carriers to create dislocations in the lattice structure. These sites act as trapping centers for minority carriers and are the direct cause of lowered current gain. These effects are shown in Figure 1.

A BJT undamaged by avalanche conduction will exhibit a constant h_{FE} over a wide range of collector currents. Once avalanche conduction has occurred, the low current h_{FE} is reduced. At progressively higher avalanche currents, the damage sites propagate deeper into the base region and the low current h_{FE} is reduced even further and high current h_{FE} begins to degrade.

It has been demonstrated that h_{FE} degradation has a stronger dependance on the magnitude of avalanche current than on time at a particular current. It has also been demonstrated that a partial reversal of the damage occurs with high temperature storage or operating life, providing the cause of avalanche conduction is removed and the damage level is not extreme.

Causes of Avalanche Conduction

Avalanche conduction occurs whenever the reverse bias potential exceeds the breakdown voltage of the base-emitter junction as shown in Figure 2.

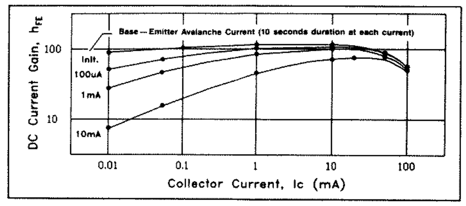


Figure 1. Effect of base-emitter avalanche conduction on h_{FE}.

Three causes of base-emitter avalanche conduction are common: 1) Intentional avalanche conduction during device testing. 2) Unintentional avalanche conduction due to bias circuit design and excessive input power levels. 3) Unintentional avalanche conduction due to improper sequencing of multiple bias supplies and improper design of the bias circuit. In most cases, avalanche conduction can be eliminated or its effects minimized.

Testing BJT's for V_{EBO} is routinely done during the manufacturer's test processes and at the user's incoming inspection facility. In order to minimize damage due to testing, the test current must be held to a minimum level consistent with the specification and pulsed conditions of less than 100 usec should be used.

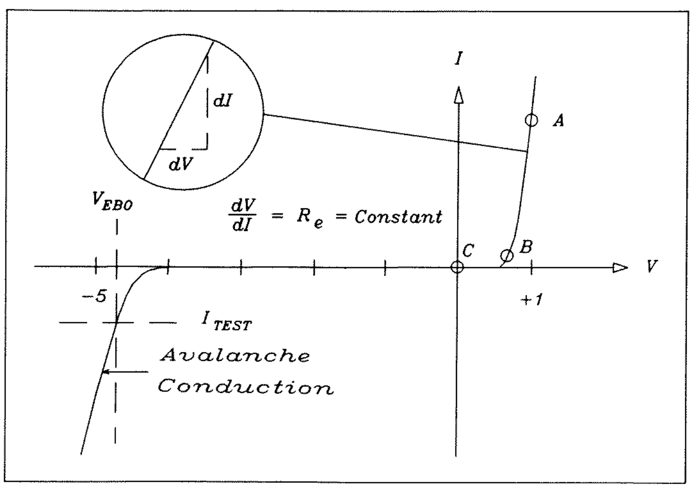


Figure 2. Typical base-emitter IV characteristic.

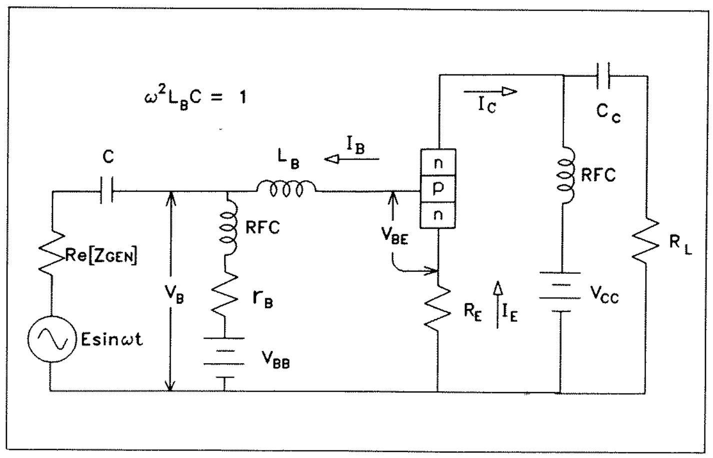


Figure 3. Common emitter amplifier with generalized bias circuit.

Whenever the base-emitter input signal is large compared to the forward threshold of conduction, the B-E junction becomes an effective rectifier. Rectification can cause the B-E junction operating point to be shifted negatively thus setting the stage for the negative peaks of the signal to bias the B-E junction into avalanche conduction. In Figure 2, we see that the B-E bias points are progressively less positive for Class A, B, and C bias respectively. As the base-emitter junction becomes reverse biased, the terminating impedance for the signal source approaches an open-circuit condition and causes the peak reverse B-E bias to nearly double. The implication is that Class B and C amplifiers are predisposed to avalanche conduction and h_{FE} degradation.

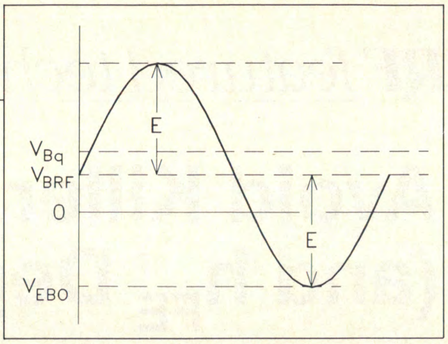
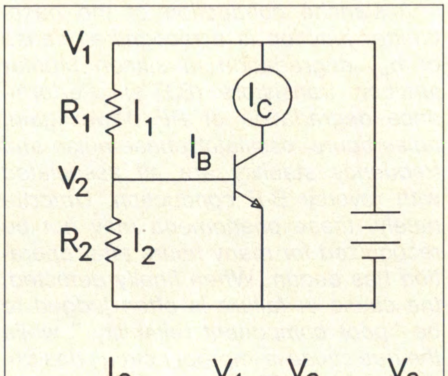


Figure 4. Base-emitter driving voltage under RF conditions.



$$I_B = \frac{I_C}{h_{FE}} = \frac{V_1 - V_2}{R_1} - \frac{V_2}{R_2}$$

Figure 5. Single supply bias circuit.

When positive and negative bias supplies are used together, it sometimes happens that the negative supply (if turned on first) will bias the B-E junction into avalanche conduction until the positive supply is turned on and stabilized.

Theory

By definition: Class A amplifiers never operate in cutoff (i.e the instantaneous collector current is always greater than zero.), while Class B and C amplifiers always operate in cutoff. Class B amplifiers are in cutoff less than 180 degrees while Class C amplifiers are in cutoff more than 180 degrees.

A generalized CE amplifier circuit with a series equivalent input matching section and series equivalent base-emitter bias circuit is shown in Figure 3. Element R_E is the total unbypassed emitter resistance and includes R_e, the internal emitter ballast resistance of the transistor chip (see Figure 2). Note that the circuit is assumed to be conjugately matched and $Re[Z_{GEN}] = Re[Z_{IN}]$, the real part of the transistor input impedance.

V_{BB} and r_B are the series equivalent DC driving voltage and the dynamic source resistance of the base bias

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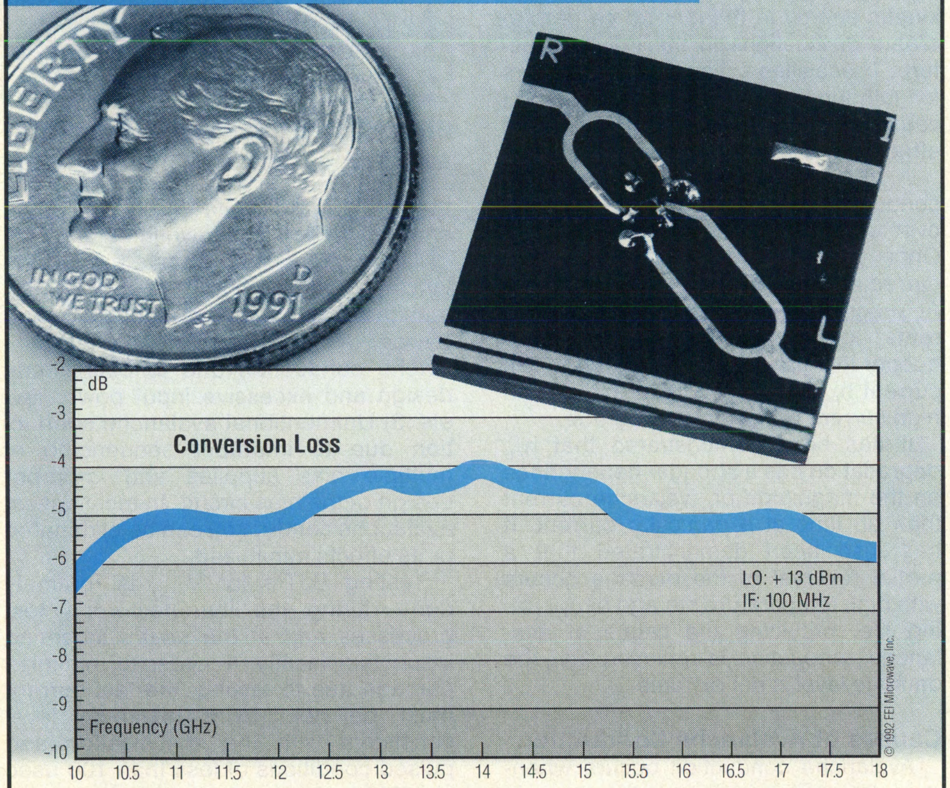

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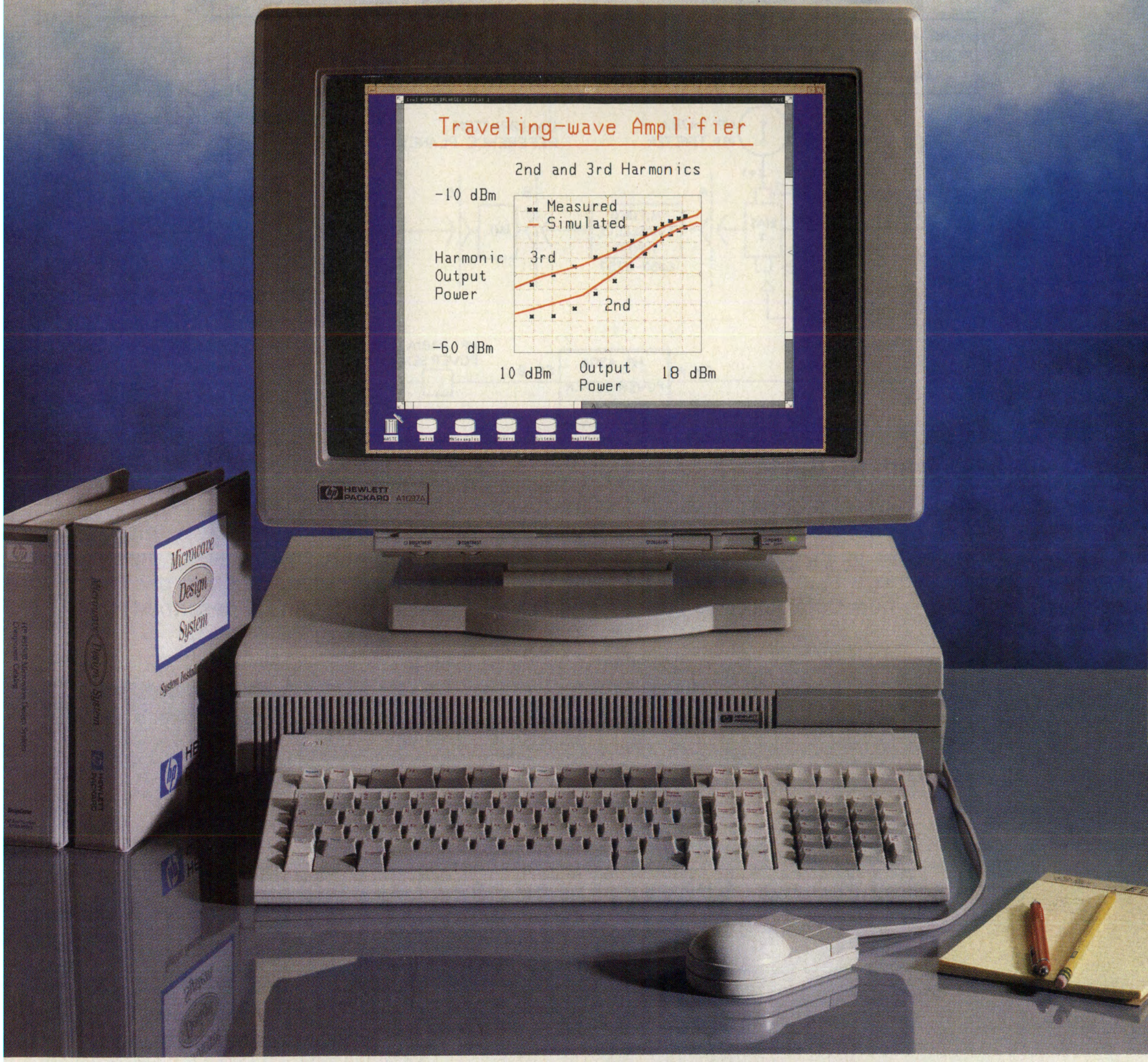

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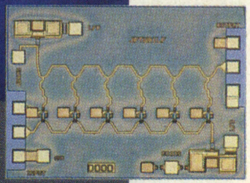

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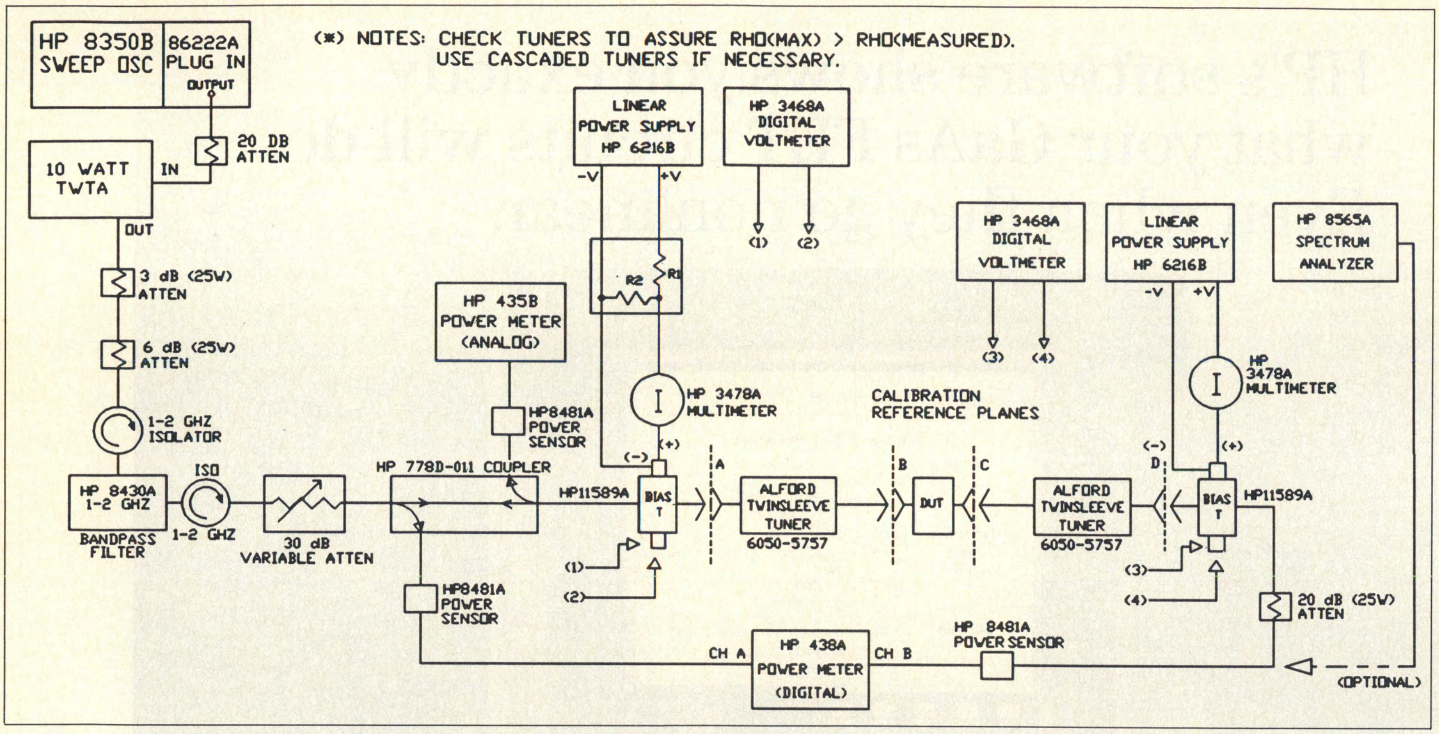

Figure 6. Test setup for h_{FE} variation with RF drive.

network. r_B provides a resistance across which any rectified current will develop a reverse DC component to the B-E bias voltage. V_{BB} is simply the open circuit voltage that would appear across the base-emitter terminals if the transistor were removed. If dual power supplies are used, each supply/divider network has a V_{BB} value associated with it and when both supplies are on and stabilized, $V_{BB}=V_{BB}(+) + V_{BB}(-)$.

The maximum power that can be delivered to the transistor input without avalanche conduction is based on Figures 3 and 4 and is described by:

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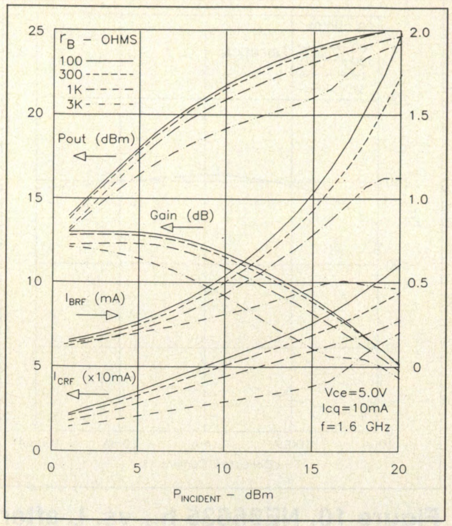

Figure 7. NE85635 P_{OUT}, Gain, I_{CRF} & I_{BRF} vs. $P_{INCIDENT}$ and r_B.

$$P_{AVALANCHE} = \frac{(V_{EBO} + V_{BRF})^2}{8\ \mathrm{Re}\ [Z_{GEN}]} = \frac{(V_{EBO} + V_{Bq} - I_F r_B)^2}{8\ \mathrm{Re}\ [Z_{GEN}]} \quad (1)$$

where V_{EBO} is the emitter-base breakdown voltage (not BV_{EBO}, the emitter-base breakdown voltage rating), V_{BRF} is the base voltage under RF drive conditions, V_{Bq} is the base voltage with zero RF drive and I_F is the forward rectified base current under RF drive conditions. V_{BRF} and V_{Bq} are represented by V_B in Figure 3 and both are easily measured under operating conditions.

Detecting Avalanche Conduction

If a DC current meter is inserted between the base and the DC driving voltage, V_{BB}, it is observed that the DC base current will increase with RF drive. The change in base current is due to base-emitter rectification of the input signal (a large signal effect). If the RF drive is sufficiently large to cause the B-E junction to be reverse biased into avalanche conduction, there is an additional component, I_R, in the measured base current and $I_B = I_F - I_R$. Therefore, an effective means of detecting avalanche breakdown is to measure $h'_{FE} = I_{CRF}/I_{BRF}$, where I_{CRF} and I_{BRF} are the time-averaged DC collector and base currents under RF operating conditions,

The presence of avalanche conduction will manifest itself as an increase of h'_{FE} as the RF drive is increased. This is contrary to the normal variation of h_{FE} with IC at high currents as shown in Figure 1.

Unfortunately, there is no easy method to accurately measure the base current in most designs unless there is

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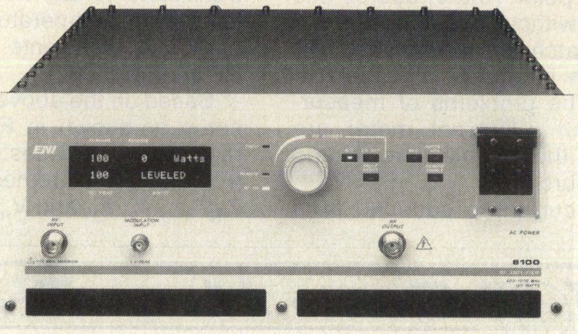

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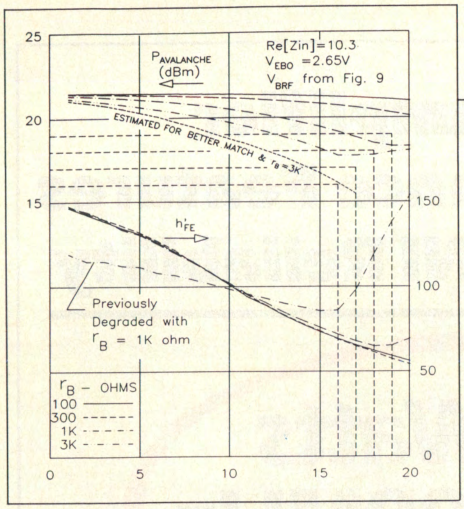

Figure 8. NE85635 h'_{FE} and $P_{AVALANCHE}$ vs. $P_{INCIDENT}$ and r_B.

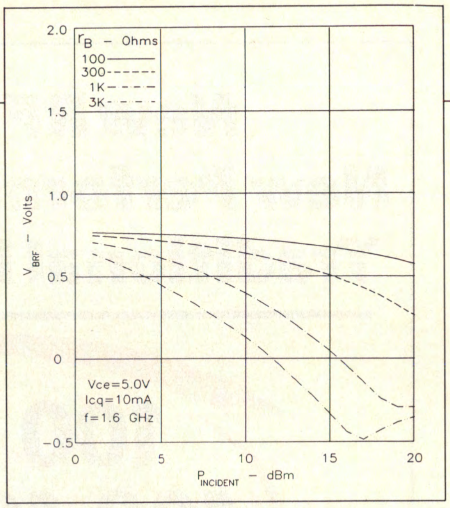

Figure 9. NE85635 V_{BRF} vs. $P_{INCIDENT}$ and r_B.

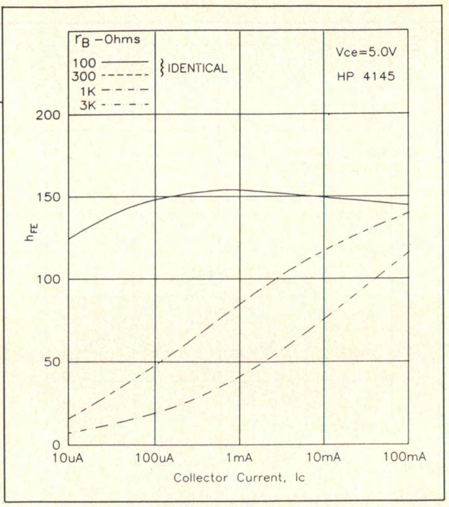

Figure 10. NE85635 h_{FE} vs. I_C after testing with different r_Bs.

a convenient way to insert a low resistance, high resolution, DC current meter (DMM) in series between the base and the DC driving point. And of course this must be done without upsetting the RF impedance match or introducing RF radiation which interferes with the test instruments. The problems of measuring h_{FE} without a current meter are illustrated by the simple circuit and equation in Figure 5.

If the base current is deduced from measurements of V1, V2, R1 and R2, then a measurement accuracy on the order of 5-500 ppm is required for typical transistors to say nothing of the problems with temperature stability, temperature coefficients of resistance and RF interference.

Based on the above discussion, a test setup as shown in Figure 6 is recommended. It provides independent control and measurement of P_{FWD}, P_{REV}, Z_{IN}^*, r_B, I_B, I_C, and V_B while maintaining excellent control of parasitics, radiation, oscillations, etc.

Experimental Procedure

Two NEC bipolar transistor types were selected for evaluation at 1.6 GHz. The NE85635 is intended for applications up to +20 dBm output while the EXP53 is an experimental device intended for outputs up to +30 dBm. Both have practical gains greater than 10 dB.

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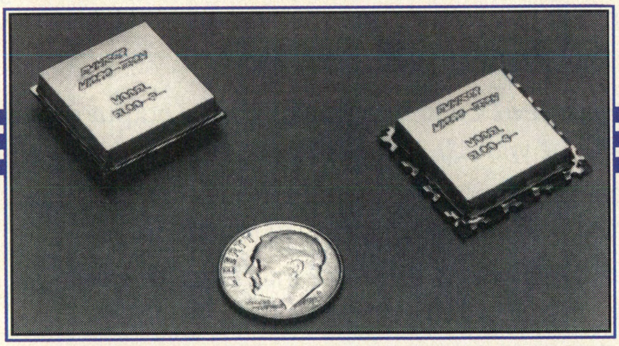

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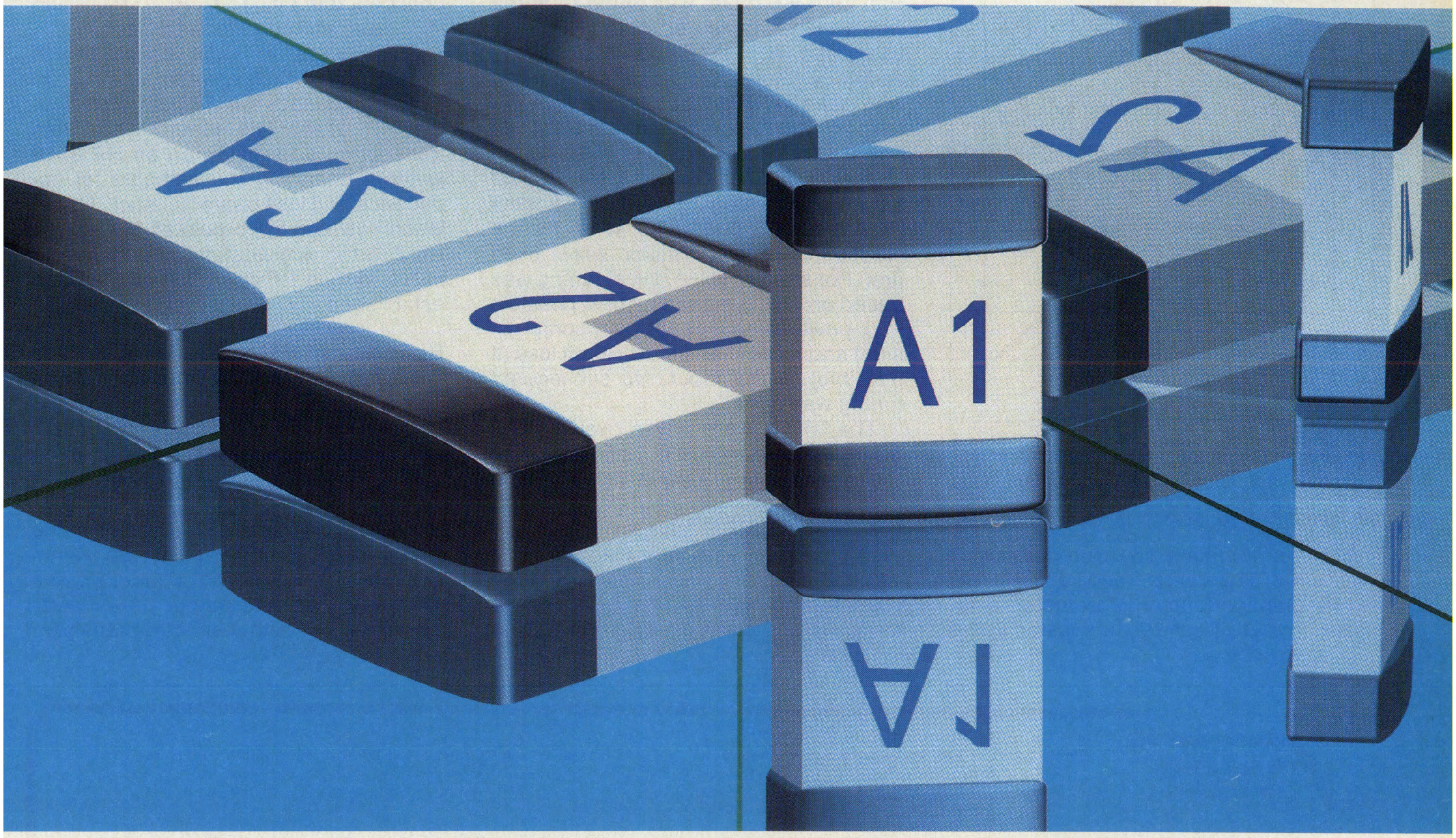

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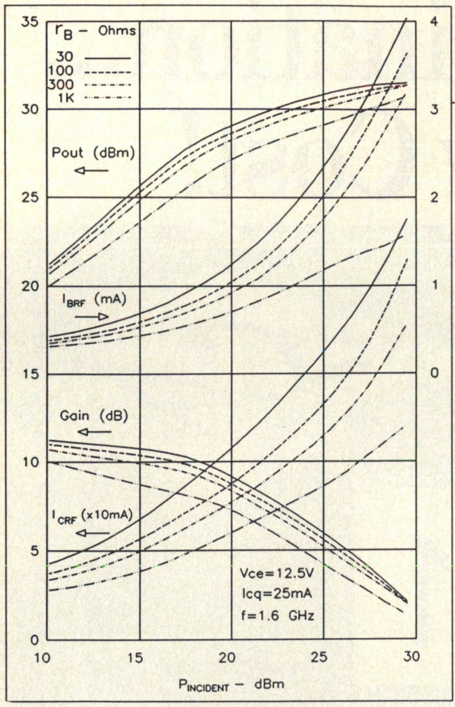

Figure 11. EXP53 P_{OUT}, Gain, I_{CRF} & I_{BRF} vs. $P_{INCIDENT}$ and r_B.

istics were initially measured using a Tektronix 576 curve tracer and an HP 4145 Semiconductor Parameter Analyzer. The DUT's were then tested in the setup of Figure 6.

The h_{FE} measured in the RF setup, with zero RF drive, was found to agree well with that measured for the NE85635 using the HP 4145. To improve the accuracy, however, an offset current was used with the base current meter to make the RF setup and the HP 4145 agree within 1 percent at I_{Cq}.

The DUT's were tested in the RF test setup with r_B ranging from 30-1 kohms for the EXP53 and 100-3 kohm for the NE85635. Lowest values were used first. For each r_B value, initial tuning was based on achieving the highest possible P_{1dB} power output (1 dB gain compression) and more than 20 dB return loss (if possible) at the input. No subsequent tuning was done.

The DUT's were driven over a 15-20 dB input power range in 1 dB steps with the nominal P_{1dB} output power rating being the maximum input (hard saturation). For each RF drive level, V_{CC} was readjusted to the quiescent current value to compensate for changes in the voltage drop due to 4 ohm resistance in the collector bias tee. (Remote power supply sensing was attempted but offset currents in the current meters were unacceptable.) V_B, I_B, I_C, P_{FWD}, P_{REV}, and P_{OUT} were recorded for each step.

Subsequent to each series of RF tests, and for each r_B value, the h_{FE} vs. I_C characteristics were again tested on the HP 4145. The tuners were then transported and tested on an HP 8510 (without changing the settings) for impedance and loss characteristics. Correlation between the predicted $P_{AVALANCHE}$, $Re[Z_{in}]$, h_{FE} degradation, and increases in h'_{FE} during RF operation can then be established.

Experimental Results

NE85635 — For low values of r_B there was no difficulty in achieving an excellent return loss. As r_B was increased, however, it became impossible to achieve 20 dB return loss at the drive levels associated with P_{1dB} output. Furthermore, as the drive was increased the return loss became even worse — not unusual for large signal operation. The reflected input and output power ports were checked with a spectrum analyzer

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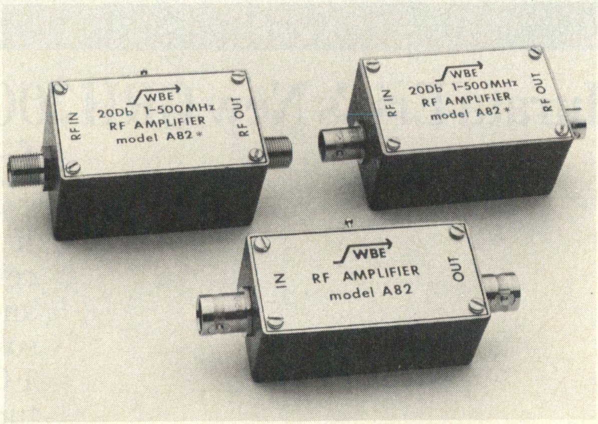

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A82A	1-500	.3-650		±.15	.7	28				3
A82L	.1-50	.050-150		±.5	1.0	50				3
A82LA	.4-30	.3-100		±.5	1.0	50				3

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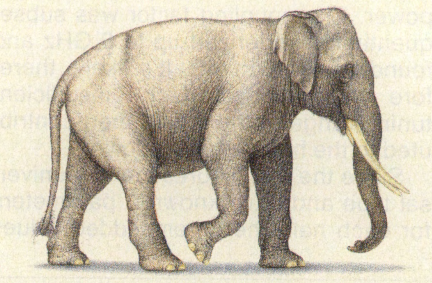

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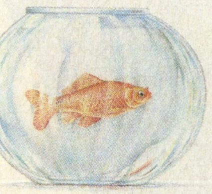

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103M	10 - 50 MHz	TTL	+/- 1.0 to +/- 5.0 PPM	.970" X .795" X ..375"	16-Pin DIP, hybrid construction Class B-Military
544	3 - 25 MHz	TTL or HCMOS	+/- 0.5 to +/- 5.0 PPM	1.00" X 1.00" X .500"	1" square metal package, low phase noise, +5 VDC
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for oscillations, but none were found. It was observed, however, that the second harmonic level at the reflected power port of the input coupler was about the same level as the fundamental. But the coupler is not specified for operation beyond 2 GHz and it was simply assumed that the second harmonic was the major contributor to the reflected power. The coupling factor was subsequently characterized up to 5 GHz and found to be 20±1 dB. It's likely, therefore, that an added effect — insufficient tuning range of the input tuner, contributed to the high reflected power.

Since the test fixtures are of a universal type and have known S parameters for each half, the de-embedded values of Z_{GEN} and Z_{LOAD} at the package boundaries were easily obtained.

The experimental results are shown graphically in Figures 7-10 for the NE85635. In Figure 7, the anomalous power transfer curve (3 kohms) is due to poor input return loss; 6 dB, 3.5 dB and 12 dB at +1, 15 and 20 dBm input respectively. An estimate of how this curve would look if the input were better matched is also shown. Figure 8 clearly shows the correspondence between the minimum h'_{FE} value and the calculated $P_{AVALANCHE}$. The measured Re[Z_{in}] values were 6.6 ohms (large signal) and 10.3 ohms (small signal at V_{ce}=10V, I_c=30 mA). The 10.3 ohm value correlates best.

The measured h'_{FE} and h_{FE} values at low currents compare well in Figures 8 and 10, validating the other data. The variation in V_{BRF} with RF drive, as shown in Figure 9, is indeed interesting. The point of inflection agrees well with P_{IN} associated with the minimum h'_{FE}. Since V_{BRF} is so much easier to measure than h'_{FE}, and if the point of inflection is verified to correlate with the minimum h'_{FE} in all cases, then a truly simple means of detecting avalanche conduction has been found.

The final measure of h_{FE} degradation is shown in Figure 10. While not an immediate threat to functional performance as a gain block, there is little doubt of functional failure with continued operation.

EXP53 — The EXP53 is a flange-mounted unit and required a different, but fully characterized test fixture. The input impedance is lower than a single tuner can accommodate so two tuners were cascaded at the input. There was no difficulty in achieving a 20 dB input return loss. The part is intended for +30 dBm output at 14 VDC but we chose to evaluate it at 12.5V. The lower voltage and the relatively high values of r_B account for the lower output powers achieved.

The data collection was straightforward and varied little from the methods used for the NE85635. The tuner losses were higher though and necessitated de-embedding them from the raw data. The results are shown in Figures 11-14.

Equation 1 predicts $P_{AVALANCHE}$ of 27.5 dBm for V_{BRF} associated with 300 ohms. Although there was not an obvious indication from h'_{FE} (Figure 12) or V_{BRF} (Figure 13) that avalanche conduction had begun, there is clearly h_{FE} degradation shown in Figure 14. There is a strong correlation between h_{FE} degrada-

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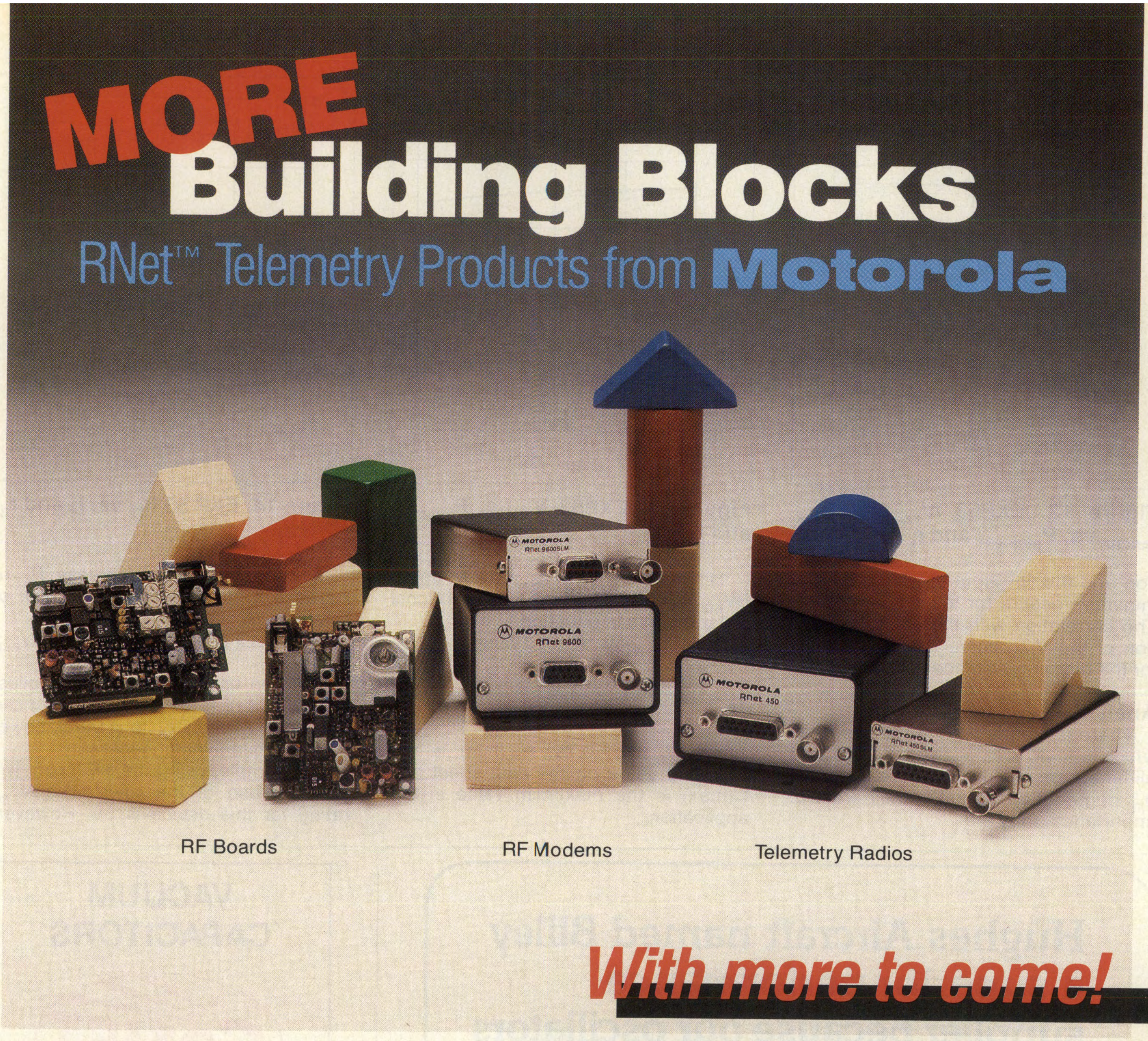

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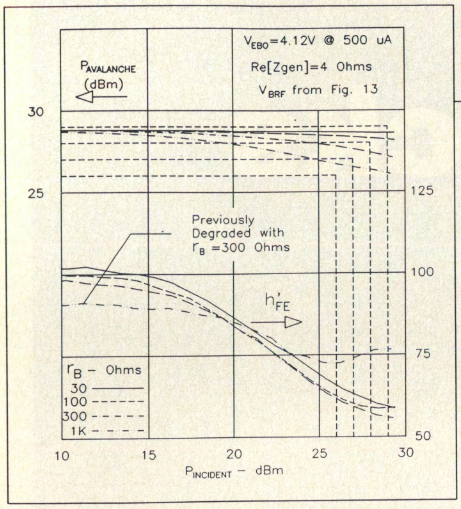

Figure 12. EXP53 h'_{FE} and $P_{AVALANCHE}$ vs. $P_{INCIDENT}$ and r_B.

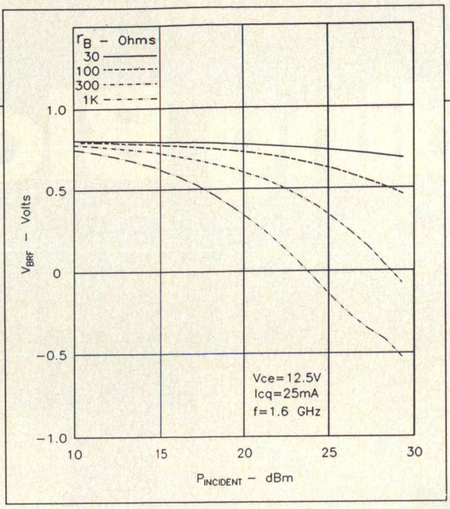

Figure 13. EXP53 V_{BRF} s. $P_{INCIDENT}$ and r_B.

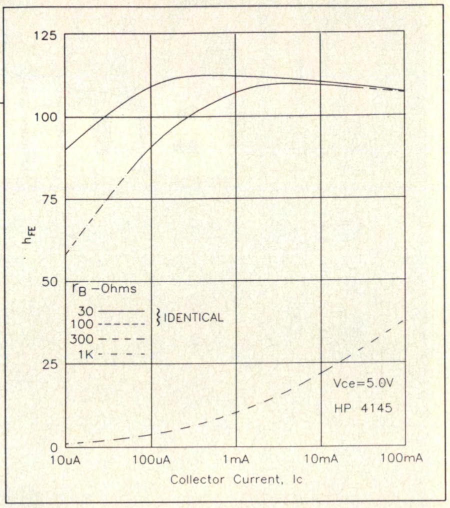

Figure 14. EXP 53 h_{FE} vs. I_C and r_B.

tion and the RF input at which the h'_{FE} minimum occurs for the 1 kohm value. The correlation with the point of inflection of the V_{BRF} curve is also apparent for the higher resistance value.

Analysis

There is clearly a set of design rules which should be followed in order to eliminate or minimize the likelihood of h_{FE} degradation. Listed in their order of importance, they are:

1) Keep $P_{IN} < P_{AVALANCHE}$, using BV_{EBO}, and Re $[Z_{IN}]$ from the small signal S parameters in equation 1

2) Keep r_B low. A good value is

$$r_B \leqslant \frac{V_{be}(typ)h_{FE}(min)}{I_C(max)} \quad (2)$$

where V_{be}(typ)=0.7V, h_{FE}(min) is the specified value on the data sheet and I_C(max) is the maximum value in the application.

3) Keep V_{BRF} high. Increase R_E or increase I_{Bq} to increase the margin of safety.

4) Select a device with a higher BV_{EBO} rating.

5) When using dual power supplies, keep: $|V_{BB}(-)| < BV_{EBO}$. This will make any special circuitry to control power supply sequencing unnecessary.

An example, using the NE85635, will illustrate use of the rules. The V_{EBO} rating for this device is 3V. However,

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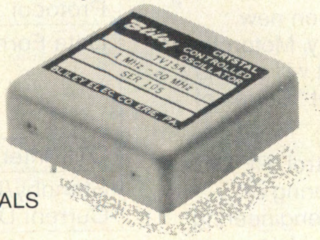

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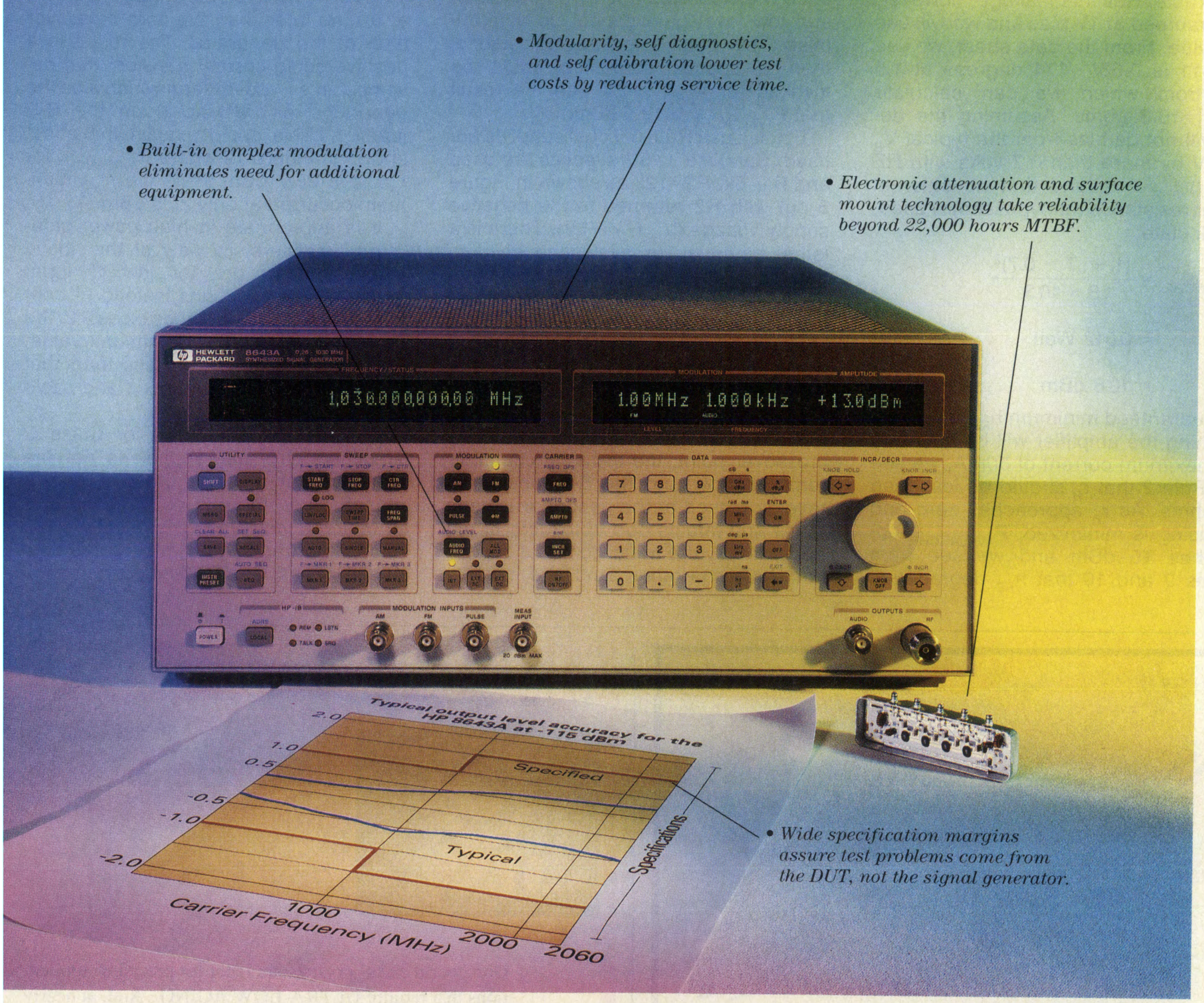

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that value is not supported by a 100 percent test in the Electrical Characteristics Table. Reverse leakage, however, is guaranteed at 1V bias and we will use that value. From the data sheet, we see that S11 is 0.66/ 174 degrees at 1.6 GHz from which we can calculate $Re[Z_{in}]$=10.3 ohms. Assuming the device will not de-bias from the typical V_B of 0.7V by more than 0.7 volts with RF drive (V_{BRF}=0V), and that no external emitter resistance (unbypassed) is used, we calculate:

$$P_{AVALANCHE} = \frac{(1 + .7 - 0.7)^2}{8 \times 10.3}$$

$$= 0.012 \text{ Watt} \qquad (3)$$

$$= 10.8 \text{ dBm}$$

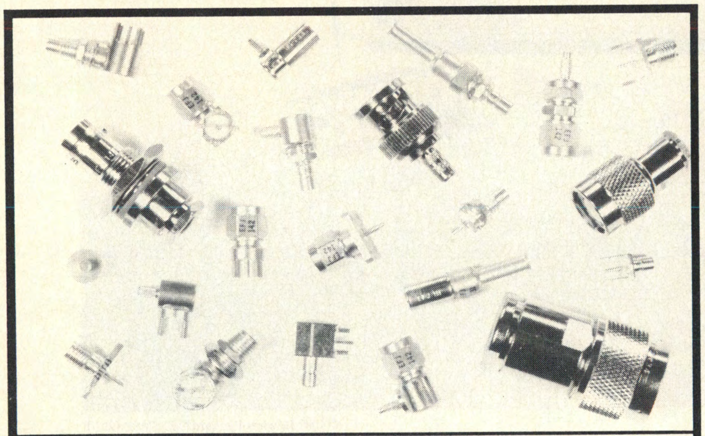
The guaranteed minimum h_{FE} is 20, and assuming the amplifier will be operated at a maximum current of 50mA, we see from Rule 2 that r_B should be less than 280 ohms. As r_B approaches zero and de-biasing is minimized, $P_{AVALANCHE}$ approaches 15 dBm and we see from Figures 9 and 10 that h_{FE} degradation will not occur with either of these conditions. If higher RF drive is required to be withstood without h_{FE} degradation, and low level gain cannot be compromised by adding emitter resistance, then a special screening could be invoked for $V_{EBO} > 2$V with the result that $P_{AVALANCHE} > 19.5$ dBm.

Finally, use Rule 5 if dual supplies are used. Suppose ±15V supplies are used and R1=6k, R2=12k as shown in Figure 5 but with R2 returned to the negative supply voltage ($V_{BB}(-)$=−5V). There is a lethal potential for avalanche conduction with improper power supply sequencing and the bias scheme should be considered anew.

Equation 1 suggests that avalanche conduction is dictated by the generator voltage which is in turn based on the input resistance at lower power levels. If the transistor goes into cutoff near the crossover point, the instantaneous currents are low and the B-E junction is reverse biased at the generator voltage - less any effects due to loading of the generator by the transistor input capacitance. On the other hand, the minority carrier charge storage in the base region becomes large at high drive levels and causes the transistor to switch into cutoff at a time later than the zero crossover time of the generator. The result is a positive-going current transient that develops an e=−Ldi/dt term which aids the generator in breaking down the B-E junction. This could explain the effect shown in Figure 12 where $P_{AVALANCHE}$ for r_B=1k is 26.7 dBm while the h'_{FE} minimum occurred at only 25.5 dBm.

It is appropriate in high power saturated amplifiers, in view of the above discussion, to opt for low Q input matching (two sections instead of one) to minimize the inductance closest to the transistor input. The minimum value, however, can never be less than that associated with the wirebond and package inductances.

It is worth noting that for Class C amplifiers, h_{FE} degradation is not the disaster that it is for Class A and B stages. First, there is no DC base bias to contend with and r_B is usually a DC short in order to maximize gain. Secondly, Class C amplifier performance



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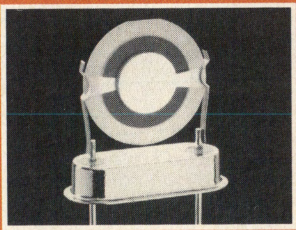

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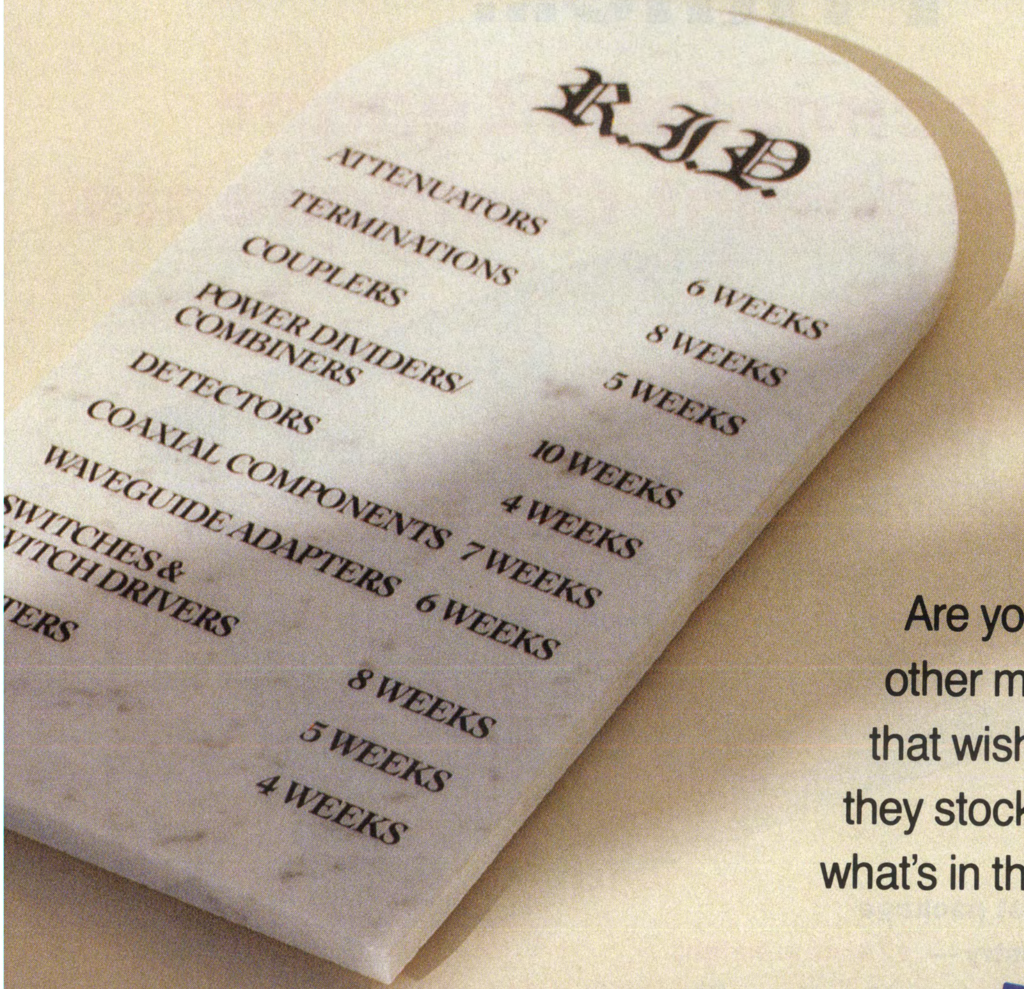

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RFMS-2A	5-1000	10-1000	+7	8.0	5.95
RFMS-4	5-1500	DC-1000	+7	9.0	9.95 ~~10.95~~
RFMS-5	10-2000	10-900	+7	9.5	12.95 ~~15.95~~
RFMS-6	10-2500	10-900	+7	10.0	14.95 ~~24.95~~
RFMS-1A-10	2-500	DC-500	+10	7.7	6.95
RFMS-2-10	5-1000	DC-1000	+10	9.0	7.95
RFMS-5-10	10-1500	DC-1000	+10	9.5	11.95
RFMS-1A-13	2-500	DC-500	+13	7.7	7.95
RFMS-2-13	5-1000	DC-1000	+13	9.0	8.95
RFMS-5-13	10-1500	DC-1000	+13	9.5	10.95
RFMS-1A-17	2-500	DC-500	+17	8.5	9.95
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Note 1: Max. over total range.

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relies heavily on the high current gain; even though the low current h_{FE} may be degraded, the damage is often insufficient to cause functional failure. But where high reliability is required, no avalanche breakdown should be allowed without extensive reliability testing. From an economic viewpoint, it's cheaper to eliminate it.

Conclusions

It is clear from the data and this simple analysis that equation 1 is quite good but not perfect in predicting avalanche conduction. Nevertheless, a judicious choice of which $Re[Z_{IN}]$ to use (large or small signal), based on the best correlation with h_{FE} and h'_{FE} changes, can alleviate these shortcomings. In view of the usually dire results of avalanche conduction, it is probably wise to err on the conservative side and use the small signal value for $Re[Z_{IN}]$.

A strong correlation has been found to exist, on a limited sample size, between minimum h'_{FE} and the point of inflection of the V_{BRF} curves when plotted versus P_{IN}. This is an entirely reasonable, but previously overlooked relationship.

Applying a simple set of design rules shows that $P_{AVALANCHE}$ can be increased by nearly an order of magnitude with proper bias circuit design and transistor specifications.

Acknowledgements

The author is indebted, first and foremost, to the many CEL customers who have experienced h_{FE} degradation problems and called upon us for help. That was the stimulus for this work. I am indebted to Richard Mencik, who made the time and resources available to complete this work and whose constant admonition "Keep It Simple" finally took root; to Richard Q. Lane, our resident expert in all matters electronic, for his stimulating discussions and particularly his focus on using well-defined symbols and terminology to avoid conflict and confusion. And finally, I wish to thank Tuan Nguyen and Khann Luu whose painstaking data collection and attention to detail made the expected results happen. *RF*

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2. B.A. McDonald, "Avalanche Degradation of h_{FE}," *IEEE Trans. Electron Devices*, vol. ED-17, pp. 871-878, October 1970.
3. "h_{FE} and Noise Figure Deterioration Caused by EB Junction Breakdown in Transistors," NEC Application Note TEB-3011, May, 1990.

About the Author

Mr. Schultz is a thirty year veteran in the RF semiconductor industry; doing process and product development, technical market research, and applications engineering for Hoffman Semiconductors and TRW RF Semiconductors. He is currently an Applications Engineer with California Eastern Laboratories. He may be reached at CEL, Inc., 4590 Patrick Henry Drive, Santa Clara, CA 95056-0946. Tel: (408) 988-3500.

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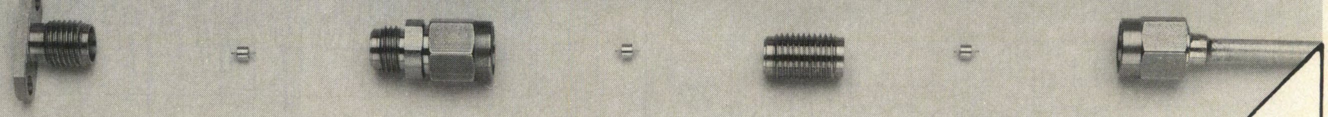

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INFO/CARD 46

Micro Miniature RF Switch Cuts Size, Weight and Power

By James A. Lapke
K&L Microwave Inc.

RF system design engineers are often challenged to provide switching functions at substantially reduced size, weight and power consumption. K&L Microwave has introduced a new series of precision electromechanical switches to meet that challenge. The new Micro Miniature Switch provides a 1P2T failsafe function in less than 1/6 the volume and weight in less than 1/4 the area, and with less than 1/4 the consumed power of traditional electromechanical miniature switches using SMA connectors.

A miniature coaxial switch with a precision bridging contact (not a fixed bladed contact) has typically been 1.34×1.20×.50 in. (w×h×d) less the connectors and terminals. The new Micro Miniature is .75 wide by .75 high by .20 deep (inches) and performs functionally like its larger counterpart. Printed circuit board mounted versions of the traditional miniature switches fit in a .67 square inch area (1.34×.50 in.) whereas the new Micro Miniature fits in .15 square inches. The new switch weighs about 1/2 an ounce, a reduction from 4 ounces or more. Failsafe switches with SMA connectors usually consume 3-1/2

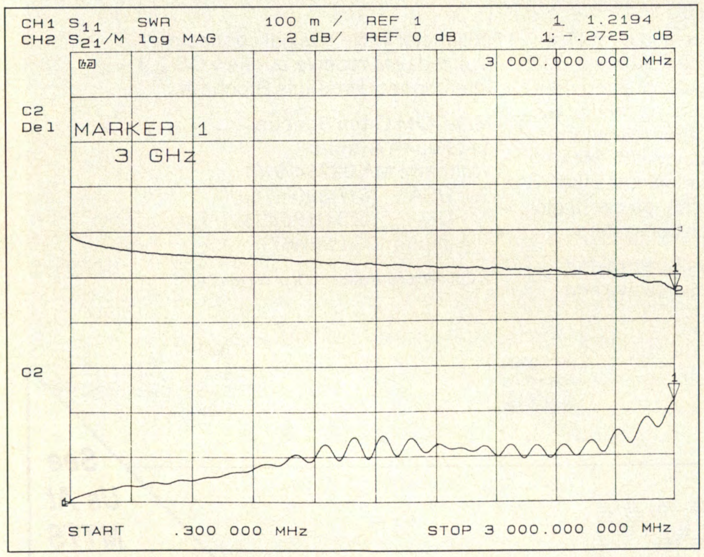

Figure 1. VSWR (upper) and insertion loss (lower) plot for the switch (0.2 units per vertical division).

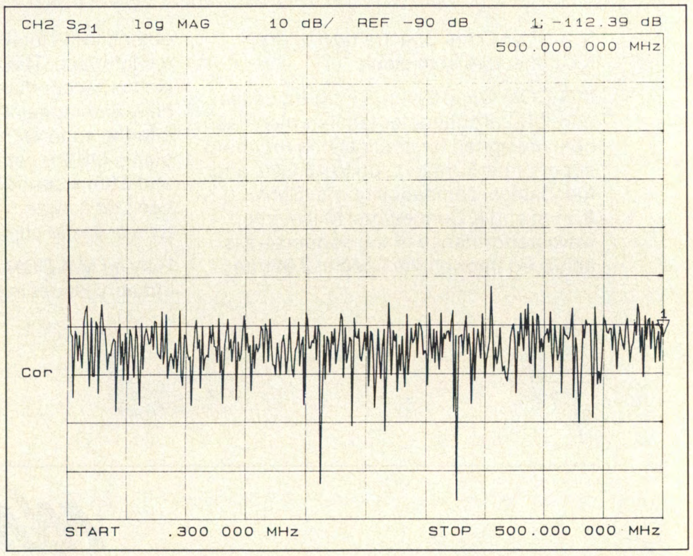

Figure 2. Typical isolation below 500 MHz exceeds 100 dB, and remains better than 80 dB to 3000 MHz.

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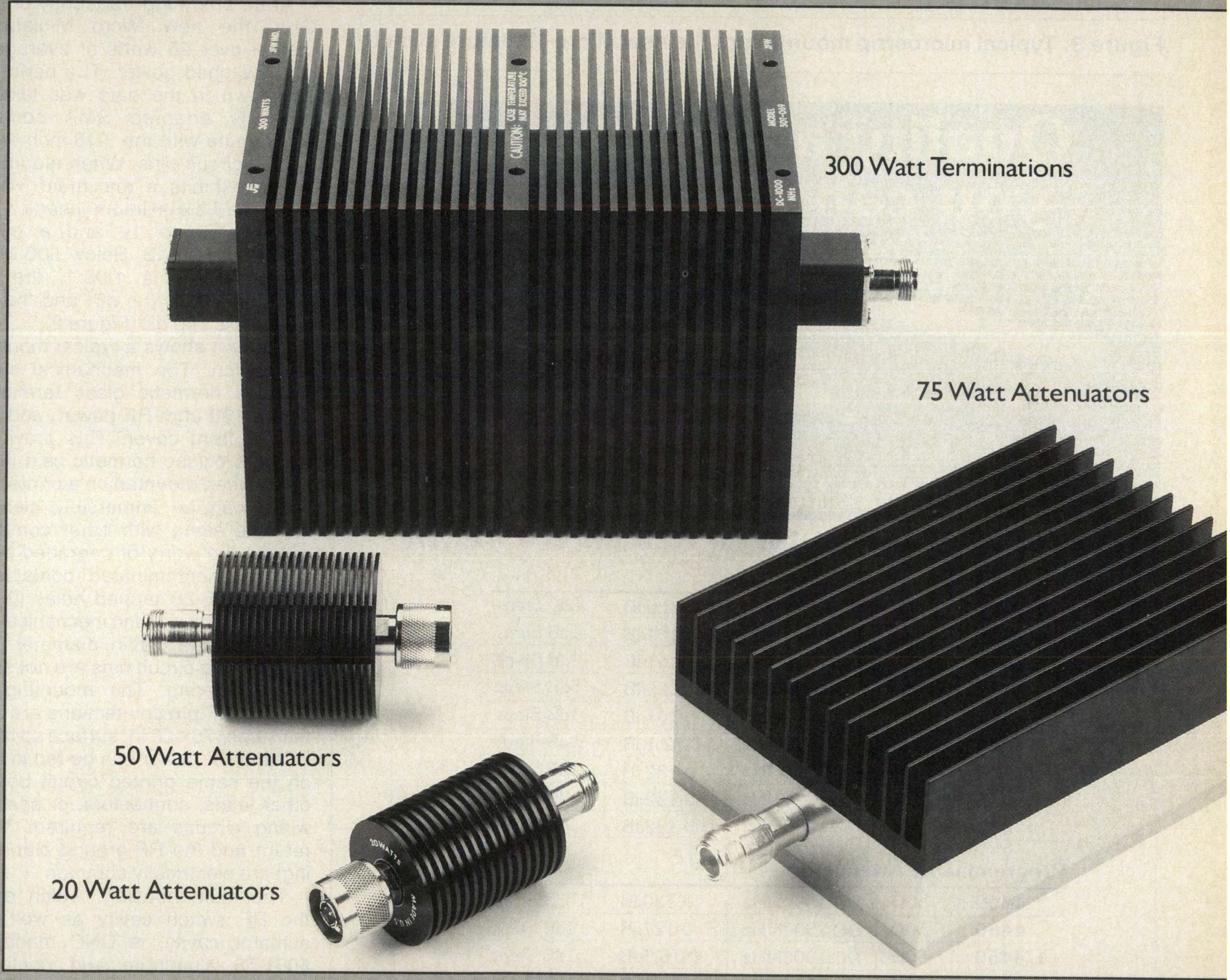

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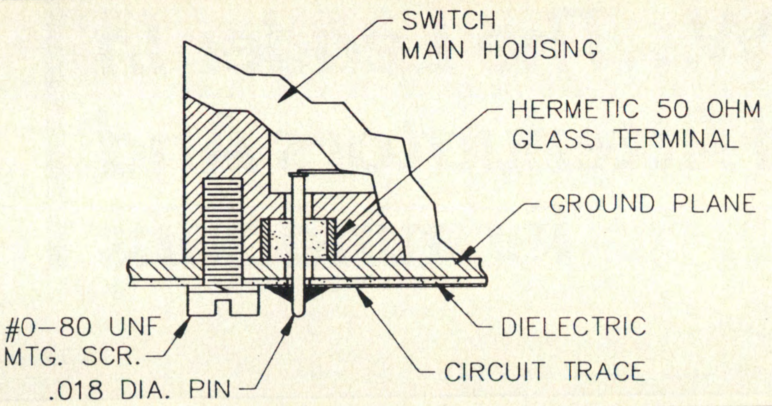

Figure 3. Typical microstrip mounting of the new Micro Miniature switch.

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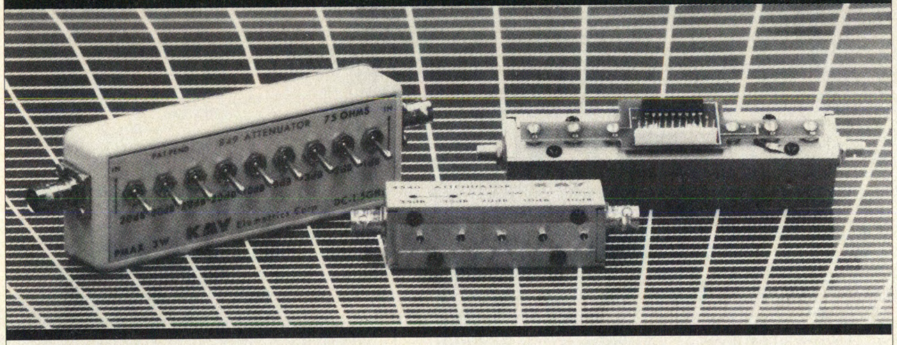

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4550	50Ω	DC-500MHz	0-127dB	1dB Steps
1/4550	50Ω	DC-500MHz	0-16.5dB	.1dB Steps
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to 4 watts of coil power when energized. Because of the reduced contact throw and size, the new Micro Miniature consumes less than 1 watt, but without the expected tradeoff in isolation that is inherent to the smaller, bladed type designs.

At an operating frequency of up to 3 GHz, the new Micro Miniature will control over 20 watts of average (CW) cold switched power. The performance as shown in the data was taken with specially adapted SMA connectors which mate with the .018 inch diameter printed circuit pins. When mounted, the switch exhibits a maximum VSWR of 1.25:1 and a maximum insertion loss of 0.3 dB (Figure 1), and a minimum isolation of 80 dB. Below 500 MHz the typical VSWR is 1.05:1, the typical insertion loss is 0.1 dB, and the typical isolation is 100 dB (Figure 2).

Figure 3 shows a typical mounting of the switch. The mechanical structure includes hermetic glass terminals for DC and 50 ohm RF power, and a laser welded front cover. This provides full 10^{-8} atm cc/sec hermetic seal integrity, which when mounted on a printed circuit board can be immersibly cleaned in solvents along with other components without the worry of degraded performance or contaminated contacts. Two #0-80 UNF-2B tapped holes (0.150 in. deep) provide a sound mechanical mounting so that the .018 in. diameter, .200 in. long printed circuit pins are not stressed after soldering. The mounting holes, DC and RF pin connections are all from the same .75×.20 in. surface so that both DC and RF power can be fed in and out on the same printed circuit board. No other leads, connectors, or special flex wiring circuits are required. The DC return and the RF ground plane (housing) are electrically common.

The main housing, which encloses the RF switch cavity as well as the actuator cavity, is CNC machined of 6061-T6 aluminum and contains no internal threaded fasteners. All parts and assemblies are welded or pocketed in place so that the structure is complete once the front cover is laser welded. The contacts are gold plated beryllium copper, and with the matched contact areas and forces will provide over 1 million life cycles without performance degradation. The actuator is a magnet assisted, failsafe (mechanical spring), balanced armature type. In addition to the low power to actuate, the drop out is below 1/4 of the operating voltage so that external power cut back circuits can reduce the holding power by 1/2 if desired. The make before break switching time is typically less than 10 msec.

Ongoing environmental qualifications are expected to verify operating performance from −55 to +85 degrees C. Mechanical integrity will be to the survival and operational requirements of MIL-STD-202 for vibration (Method 204, Test Condition C, 10 GRMS) and mechanical shock (Method 213, Test Condition G, 50 G/11 msec). In general all other environmental performance will be to the "H" hermetic category of MIL-S-3928. Screened versions are also available.

Projected applications include any type of printed circuit, stripline or microstrip construction where the space and cost to launch a SMA or larger p.c. pin switch is not desirable. The small size and weight are especially attractive for switch matrices and automatic-test systems. This approach can make the interconnect loss closer to that of the switch path. The solid mounting feature and hermeticity will allow application beyond the ground fixed benign category where the MTBF per MIL-HDBK-217E of more than 1 million hours is calculated. Many airborne and missile applications within the general requirements of MIL-S-3928 can use this switch. For applications where frequent switching is required, a quick change surface launch, plug in connector is being developed so that change out down time can be less than 5 minutes depending on p.c. board access.

Some of the switch options expected to grow out of this new series include alternate operating voltages (5, 12, 15 etc.), external finishes (nickel, silver, painted), and higher operating frequencies. A pulse latching actuator and expanded switching functions (2P2T transfer, and up to 1P6T multiposition) are being developed so that this new series can provide the basic building blocks which are needed to achieve the most versatile switching options and with the least number of switch contacts in the series path.

This new Micro Miniature series of switches will provide the RF design engineer with a new choice when his demanding application needs excellent performance from the smallest, lightest, best performing electromechanical coaxial switch available.

Interested readers can contact K&L Microwave at the address or telephone number given below, or by circling Info/Card #180. **RF**

About the Author

Jim Lapke is Product Line Manager for switch products at K&L Microwave Inc., 408 Coles Circle, Salisbury, MD 21801. He received a BSEE from the University of Bridgeport and has 20 years experience in electromechanical design. He can be reached at (410) 749-2424.

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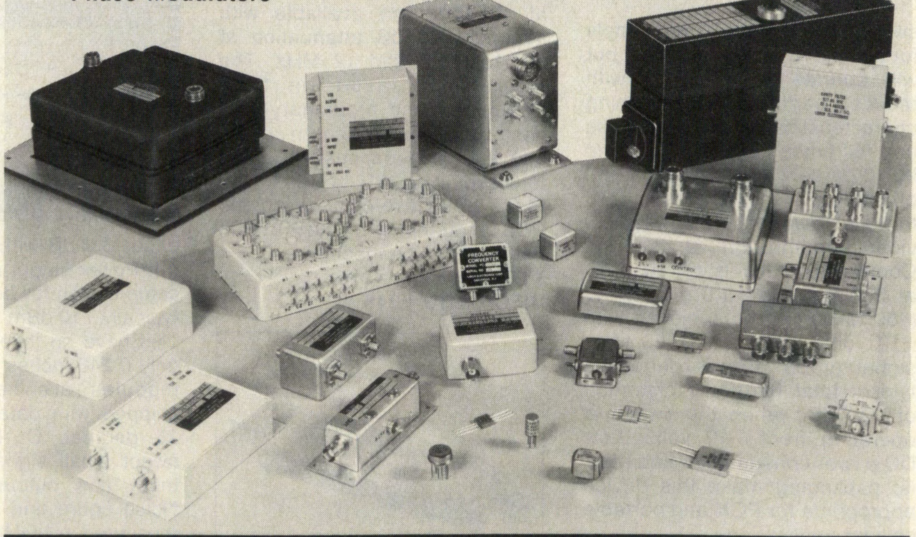

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Inline Components, Inc.
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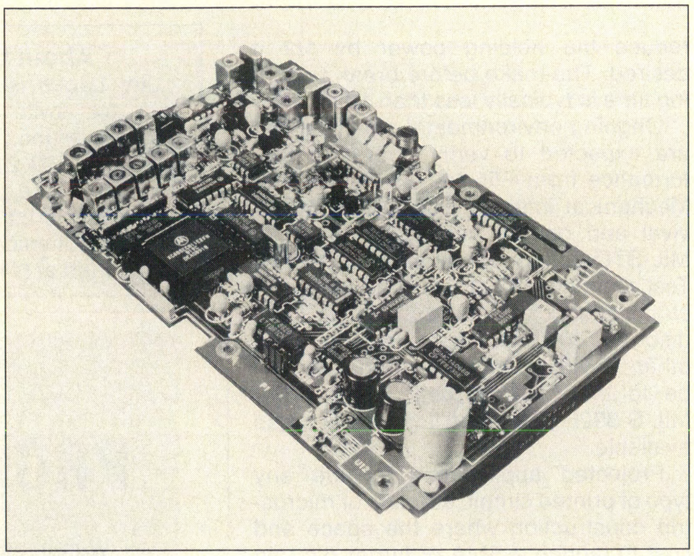
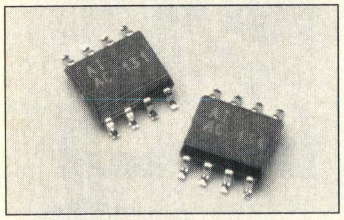


GaAs MMIC Transmit/Receive Switch

Alpha Industries announces its new GaAs 4 Watt SPDT MMIC TR switch, AH001R2-12, which is packaged in a surface mount SOIC-8 plastic package. The AH002R2-12 is a SPDT reflective FET MMIC switch designed for low distortion at high power. The



switch requires a −10 volt supply for maximum power handling, but will operate at −5 volts with reduced performance. Switching characteristics include typical rise and fall times of 6 ns (10/90 and 90/10 percent RF), typical on and off times of 12 ns (50 percent CTL to 90/10 percent RF) and typical video feedthrough of 30 mV. Compression levels of −0.1 dB at 3 Watts and −1.0 dB > 4 W (450 MHz) can be achieved with −10 volts biasing. An 850 MHz signal at 33 dBm experiences typical total harmonic distortion of −70 dBc when the switch is operating from −10 volts. Low DC power consumption and plastic packaging make this device appropriate for PCN and portable cellular applications.
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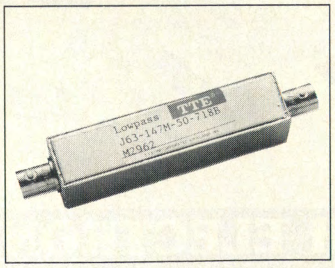
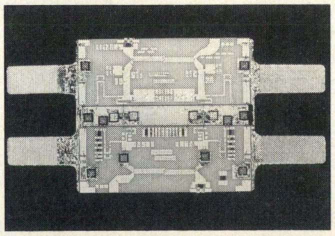


Low Noise Gain Modules

Microwave Technology offers two new hybrid gain modules exhibiting low noise figure and excellent gain from 6 to 18 GHz. These modules use the recently announced MwT-H4 pseudomorphic (PMHEMT) FET which employs MwT's 0.3 micron gate technology and unique diamond like passivation. The balanced gain stages are built on thin gold plated carriers that can easily be inserted into higher assemblies. The MwT-0618S-H4N1 has a typical noise figure under 3.5 dB and a gain of 8.5 dB from 6 to 18 GHz while the MwT-0618S-H4N2 module has a typical noise figure



under 3.0 dB with a gain of 9.0 dB. Power output at one dB gain compression for both modules is typically +10 dBm. The new modules have an insertion length of only 0.246 inches (6.25 mm). All modules are serialized and shipped with data measured at 25 degrees C. Data includes swept small signal gain, swept input and output return loss. Noise figure and P − 1 dB are measured in 1 GHz increments.
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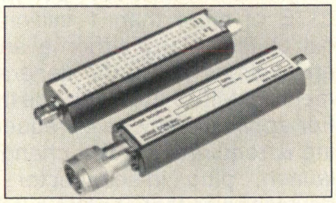

increasing the measurement accuracy of most noise figure setups. Models NC 346C and NC 346E have broadband 10 MHz to 26.5 GHz coverage and extemely good temperature and voltage stability. Outputs of 6, 15.5 and 22 dB ENR are available in the NC 346 series, allowing the units to accurately measure noise figures up to 20, 30 and 36 dB respectively. Other features include: noise output on and off times less than 20 and 80 us respectively in repetitive operation, single shot turn on less than 3 ms, VSWR less than 1.15:1 from 10 MHz to 5 GHz for units with 5 to 7 dB or 14 to 16 dB ENR, temperature coefficient less than 0.009 dB/degree C, noise output variation with voltage less than 0.002 dB/percent change in voltage. The NC 346 series is available with N, APC3.5, APC7 or SMA connectors. Each unit is supplied with calibration data traceable to NIST.
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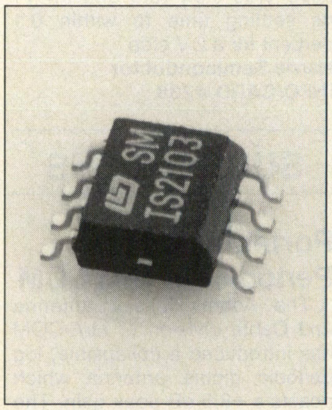
GaAs MMIC SPDT Switch

ST Olektron's low cost GaAs MMIC SPDT switch features ultra fast switching speed (<5 nsec) and high isolation (70 dB at 100 MHz). DC to 2 GHz frequency coverage is offered in a standard



8 lead SOIC package. The device (Model SM-IS-2103) sells for $2.95 (US) in quantities of 25 to 1000.
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U.S. patents: 4,655,462; 4,934,666.

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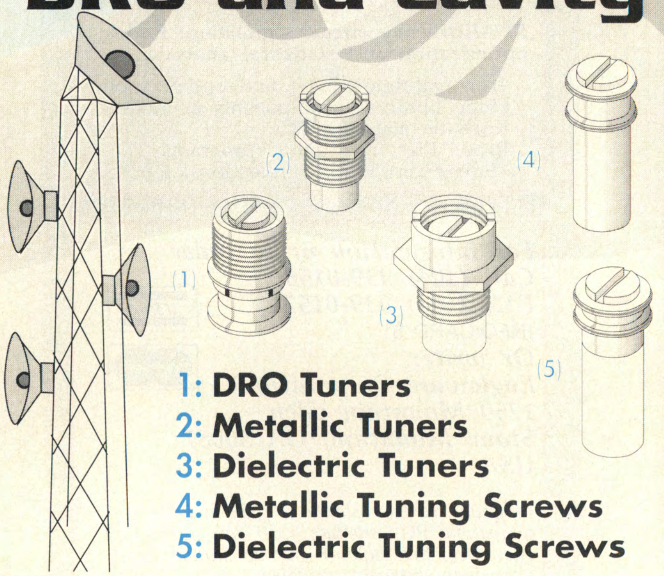
TUNERS DRO and Cavity



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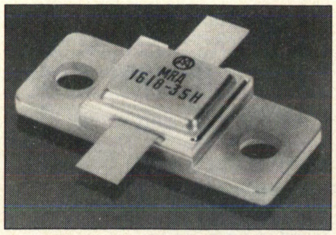

MRA1618-35H, a microwave power transistor designed primarily for wideband, large signal output and driver amplifier stages in the 1.6 to 1.8 GHz frequency range. At 28 V, 1.8 GHz, its power gain is 7.0 dB with a collector efficiency of 40 percent. The device is designed for Class C, common base power amplifiers and features a built-in matching network for broadband operation, gold metalization and diffused ballast resistors.
Motorola
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Demo Board for Fast Op-Amps

A demonstration board from Harris allows easy evaluation of the HFA1100 series of op amps. Harris' HFA1100/1120 and 1130 op amps utilize not only fast NPN transistors but also fast PNP transistors. The NPN transistors used have an F_t of 8 GHz, and the PNP transistors have an F_t of 4 GHz. The devices have a wide bandwidth (870 MHz, 3 dB), a 2500 V/us slew rate and an 11 ns settling time to within 0.1 percent for a 2 V step.
Harris Semiconductor
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SUBSYSTEMS

Portable Log Periodic for SATCOM

The Adams-Russell Antenna and Cable division of M/A-COM has introduced a collapsible, log periodic dipole antenna which boasts a +8.5 dB peak gain. The high gain provides better link margins and lower BIT error rates, and its small size and weight (5.5 lbs. with tripod) make the antenna easily portable and deployable.
M/A-COM
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Cellular Bandpass Combiner/Duplexer

K&L Microwave's 850 MHz/895 MHz cellular bandpass combiner/duplexer allows simultaneous receiver and transmitter operation from a single antenna. This unit combines two transmitters operating in one band (894 to 896 MHz) with one receiver operating in another band (849 to 851 MHz). The combiner/duplexer offers insertion of < 1.4 dB on the receive port and < 5.0 dB on each transmit port. The unit handles 70 W average and 350 W peak.
K&L Microwave, Inc.
INFO/CARD #236

MMDS Antenna/ Downconverter

A 31 channel downconverter and Yagi antenna system is available from California Amplifier. The one piece unit is easily installed on a mast. The housing is a lightweight alloy and the antenna elements are chem-coated aluminum. The antenna has 17 dB gain and the standard operating frequency is 2500 to 2686 MHz with an output of 222 to 408 MHz.
California Amplifier
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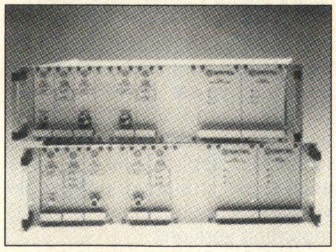
Optical TVRO Transmission Links

High performance SNR figures are achieved in 15 km transmission links using Ortel's new TVRO links. These links, used to transmit satellite downlink signals up



to 25 km at L-band and up to 15 km at C-band, are fiber optic. This eliminates microwave interference and rain and lightning effects. The L-band, C-band and dual band links allow improved performance at shorter distances or maintain performance over distances of up to 40 km.
Ortel Corporation
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Rapidly Deployed HF Antenna

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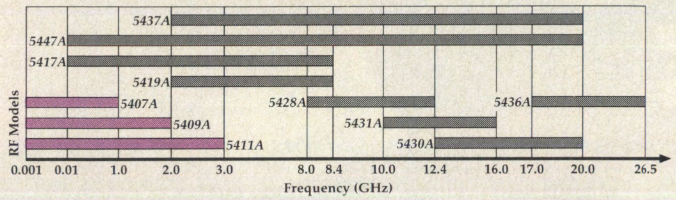

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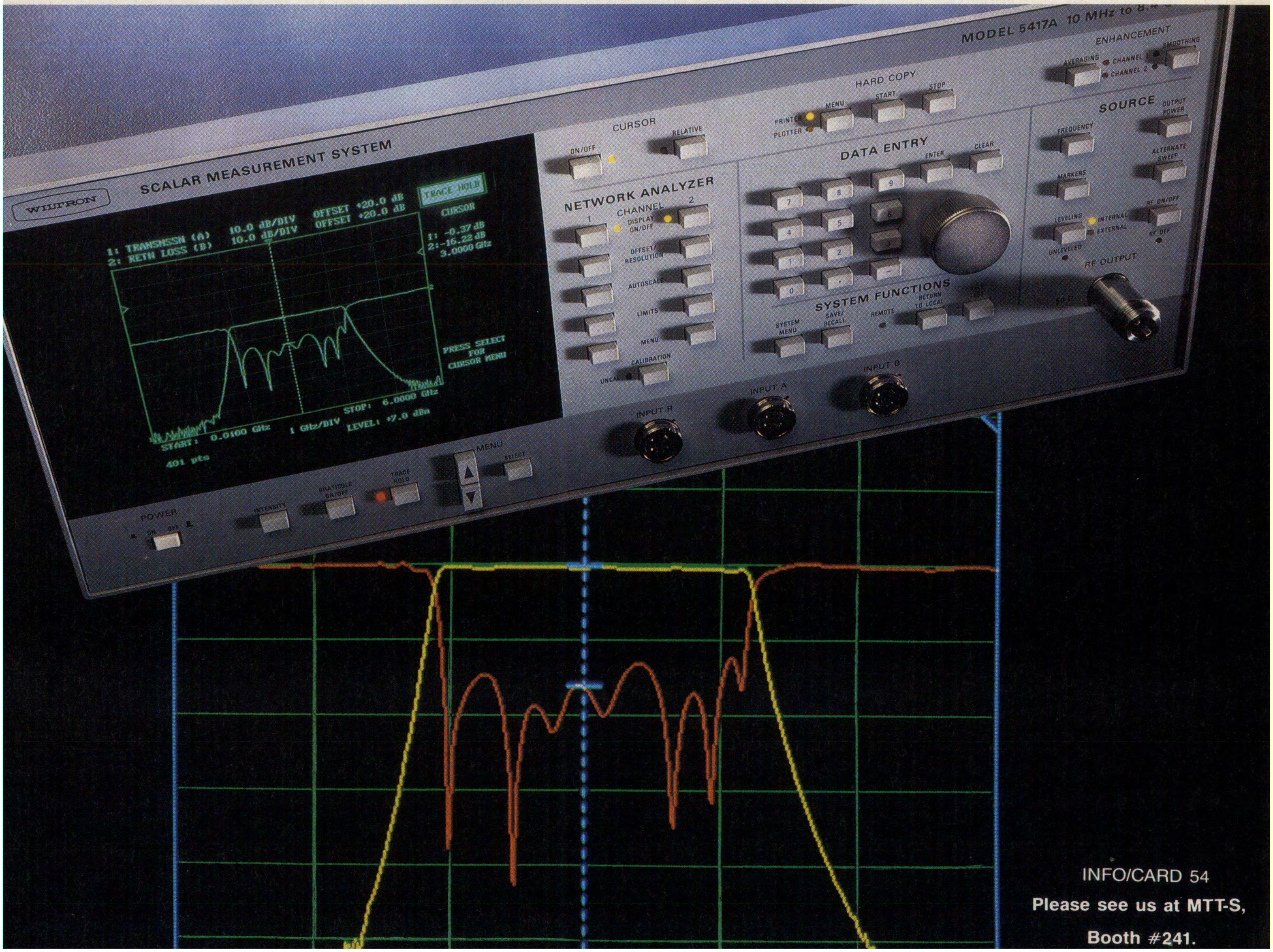

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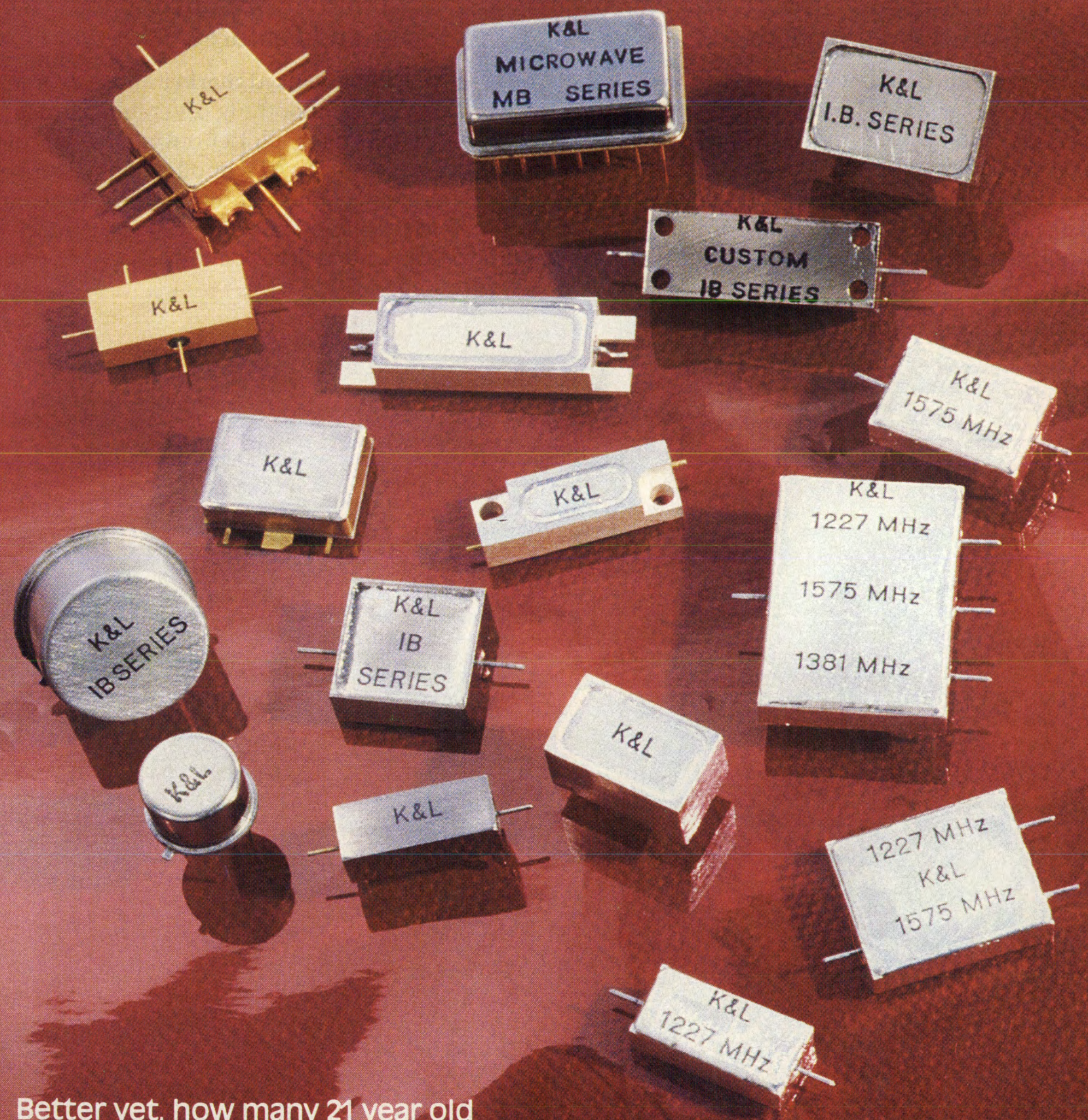

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two people can erect in less than ten minutes. The antenna is available in all length/power combinations of 45, 35, 20 and 16 feet and 150, 550, 1000, 2000 and 10,000 Watts. The VSWR is 2.5:1 (max) over 1.5 to 30 MHz.
Astron Corporation
INFO/CARD #233

DISCRETE COMPONENTS

SMT Power Inductors and Transformers

Surface mount power inductors and transformers are now available from Vanguard Electronics. These devices, available in commercial and mil-spec versions, feature a compact low profile, tin plated phosphor bronze terminations and automatic insertability. Inductance values range from 0.001 mH to 10.0 mH with a frequency range up to 10 MHz. Commercial components are $3.75 to $5.00 each in the thousands, mil-spec are $7.00 to $9.00 in the thousands.
Vanguard Electronics Co., Inc.
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High Voltage Trimmer Caps

PTFE replaces air as the dielectric in Voltronic Corporation's high voltage precision trimmer capacitors. In this internally sealed non-rotation design, a tubular electrode slides into a complementary slot lined with PTFE. Use of PTFE dielectric increases voltage ratings to around 2000 V. Component values range from 23 pF max and 1,500 withstanding voltage (WV) to 4 pF max and 2,500 WV. Non-magnetic versions are available for NMR and MRI applications.
Voltronics Corporation
INFO/CARD #231

Non-Inductive 50 W Resistors

Caddock now offers a 50 W power film resistor in a TO-220 style package. Rated at 50 W at 25 degrees C case temperature and derated to 0 at 150 degrees C, the MP 850's copper heat sink is integral in the molded package. It is available in resistances from 1.0 to 10K ohms and tolerances of ±1, ±2, ±5, or ±10 percent.
Caddock Electronics, Inc.
INFO/CARD #230

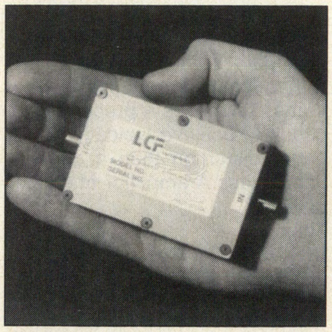
Surface Mount Trim Caps

Miniature trimming capacitors with five different mount configurations and the capability of surface mount placement are available from Murata Erie North America. Available capacitance ranges are from 1.4 pF to 3.0 pF through 7.0 pF to 50 pF. Temperature coefficients range from NPO through N1200. These new capacitors can withstand solder baths to 260 degrees C for five sec. and may be reflow soldered. They are also impervious to solvent cleaners.
Murata Erie North America
INFO/CARD #229

AMPLIFIERS

Broadband Power Amplifiers

LCF Enterprises has recently introduced two small, rugged broadband RF amplifiers. Model 1000-5-15-30 covers the frequency range from 5 MHz to 1 GHz with 15 W output power, and Model 960-5-1-8 covers 5 to 960 MHz with 1 W output power. The 15 W model operates from a 28 V power supply, while the 1 W model requires 12 V, (opening battery powered possibilities).



LCF Entreprises
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High Power VHF Amplifier

R.F. Solutions announces a range of high power, forced air cooled, hybrid VHF amplifiers. The RFP5000 series are semi-custom amplifiers supplying 5000 W of CW power for an input of 0 dBm. Modular construction means custom versions can readily be supplied at any frequency within the 40 - 200 MHz range.

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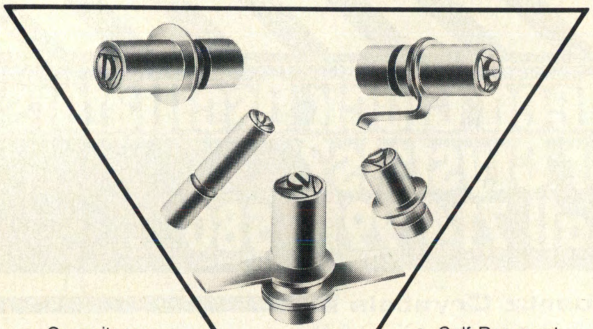

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R.F. Solutions, Inc.
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MOSFET Power Amplifier

Kalmus Engineering announces its new model 500FC MOSFET RF power amplifier. This model produces 50 Watts of linear CW power into 50 Ohms and covers an instantaneous frequency range of 1 to 500 MHz. Gain is 47 dB minimum and is flat typically within 1.5 dB. Harmonics are listed as better than −20 dBc at full power output. Priced at $8990.00, the 500FC is available within six to eight weeks.
Kalmus Engineering, Inc.
INFO/CARD #226

Cellular Repeater Amplifiers

Telia has introduced two amplifiers for AMPS and DAMPS cellular radio repeaters. These amplifiers exhibit a continuously variable gain from 45 dB to 65 dB with a peak output power of 5 Watts. Telia uses a precorrection circuit to achieve typically 10 dB better distortion than a conventional class A amplifier of similar power.
Telia
INFO/CARD #225

Linear UHF Amplifier

ENI has announced the availability of a new computer controllable UHF amplifier for testing, ATE and laboratory use. The Model 630L produces 30 W of class A linear output power over a 400-1000 MHz range with a nominal gain of 51 dB. The 630L features an automatic level control which maintains power output over a 30 dB range with ±0.3 dB flatness. Extensive RFI shielding restricts RF emissions from the amplifier. The 630L is available for 30 day delivery at a price of $14,445.
ENI
INFO/CARD #224

Low Noise Amplifier

A low noise RF amplifier, model ANR 17835 from TRM operates over the cellular frequency range of 825 to 845 MHz. This amplifier is intended for use in AMPS and IS-54 cellular systems. Unit gain is 17 dB ±1.0 dB with a flatness across the band of ±.25 dB. Noise figure is specified at 2.2 dB maximum. The output power at 1 dB gain compression is 0 dBm minimum.
TRM, Inc.
INFO/CARD #223

SIGNAL PROCESSING COMPONENTS

Low Loss Power Splitter

Tailored for cellular radio applications, Mini-Circuits' new ZC16PD-960W 16-way, 0 degree power splitter/combiner exhibits extremely low loss (0.5 dB typ.). The device operates at 700 to 1000 MHz and provides 26 dB isolation, VSWR of 1.06:1, amplitude unbalance within 0.3 dB and a maximum input power rating of 10 W. When used as a power combiner, internal dissipation is 2.4 W max. The ZC16PD-960W is priced at $265 (1-9 qty) and is available for immediate delivery.
Mini-Circuits
INFO/CARD #222

Ultra-Broadband IF Mixers

Narda has introduced a new line of mixers offering exceptional broadband performance for communications and IF signal processing. Model 482039 operates from 10 MHz to 3 GHz. Model 482026 is a double balanced mixer that operates from DC to 1 GHz. Both models offer excellent isolation, low conversion loss and are provided with SMA connectors.
Loral Microwave-Narda
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High Performance Preselector

SMT announces a high performance preselector. Based on CAE optimized combline designs, this unit operates in three bands over the 8 to 12 GHz range. 70 dB rejection is achieved 10 percent from the band edges. Measuring only 1.20 × 2.80 × 0.65 inches, this preselector is laser sealed and screened to MIL-STD-883 Level B.
Sierra Microwave Technology
INFO/CARD #220

RF Transistor Specifications

By Gary A. Breed
Editor

To select the right RF transistor for the application at hand, an RF engineer must rely on manufacturers' data sheets and other performance specifications. It is essential that the engineer understand what those specifications mean and how they are determined by the manufacturer.

All transistors are specified by two groups of specifications: DC and operational. DC characteristics include breakdown voltages, leakage currents, h_{FE} (DC beta), threshold voltages, and maximum currents. Junction capacitances may also be included in DC specifications. Operational specifications include such things as gain, noise figure, F_T, power output and impedance characteristics. The most important additional specifications are for power dissipation and thermal resistance. In a sense, power ratings are a DC specification, but it is most useful to discuss them separately.

A typical data sheet looks like Figure 1, which includes a tabulation of all specifications. In addition, most devices will have one or more charts showing key performance parameters plotted versus frequency, voltage, collector current, or other variable.

DC Parameters

Breakdown voltages are determined by the materials and fabrication of the device. Each junction voltage (collector-base and emitter-base for bipolars, or drain-source and gate-source for FETs) is generally specified at a current level that is well within the safe operating limits of the junction. The methods for deriving these specifications will generally be consistent from one manufacturer to another.

Leakage currents are not uniformly specified, particularly in small-signal devices where testing could destroy the device. When specified, leakage currents have the greatest importance where low power consumption is a critical design issue. The problem in interpreting leakage current specifications is that the cause of the leakage will probably not be known. Leakage due to processing limitations tends to be constant and relatively inconsequential. However, leakage currents due to impurities and mask defects are potential reliability problems. Military burn-in requirements are intended to sort out the latter. In an FET leakage is a function of "off" resistance, and the mechanisms of the material, impurities and fabrication are also involved.

DC current gain (h_{FE}) is relatively unimportant for RF devices, where the gain at a particular frequency is desired. However, at low frequencies the AC gain will usually track DC gain. Devices may have a h_{FE} *range* specified, but an engineer's consideration should probably be limited to an acceptable minimum value, and not too wide (3 to 1) spread in the range. For FETs, the DC gain (transconductance) is not always specified, although a low frequency AC (1 kHz) specification may be given.

Junction capacitance is mainly a function of die area, so it will have much

Absolute Maximum Ratings			
Collector-emitter voltage	V_{CBO}	20	V
Collector-base voltage	V_{CEO}	15	V
Emitter-base voltage	V_{EBO}	2	V
Collector current	I_C	30	V
Total power dissipation *25°C ambient temp*	P_{tot}	200	mW
Junction temperature	T_j	150	°C
Thermal resistance	$R_{\theta ja}$	500	K/W

Electrical Characteristics		MIN	TYP	MAX	
Collector cutoff current V_{CB}=10 V	I_{CBO}	-	-	50	nA
Collector-base breakdown voltage I_C=10 uA	$V_{(BR)CBO}$	20	-	-	V
Collector-emitter breakdown voltage I_C=2 mA	$V_{(BR)CEO}$	15	-	-	V
Emitter-base breakdown voltage I_E=10 uA	$V_{(BR)EBO}$	2	-	-	V
DC forward current transfer ratio V_{CE}=5 V, I_C=30 mA	h_{FE}	25	50	-	

AC Characteristics		MIN	TYP	MAX	
Gain bandwidth product V_{CE}=10 V, I_C=30 mA, f=500 MHz	f_T	-	5	-	GHz
Feedback capacitance V_{CE}=5 V, I_C=2 mA, f=1 MHz	C_{fb}	-	0.4	-	pF
Collector-base capacitance V_{CB}-10 V, f=1 MHz	C_{CBO}	-	0.5	-	pF
Emitter-base capacitance V_{EB}=0.5 V, f=1 MHz	C_{EBO}	-	1.8	-	pF
Noise figure V_{CE}=5 V, I_C=4 mA, R_G=R_{Gopt}, f=800 MHz	NF	-	2.4	-	dB
Power Gain I_C=14 mA, V_{CE}=10 V, f=500 MHz	G_{pb}	-	19.5	-	dB

Figure 1. Typical small-signal transistor data (BFR 91A).

larger values for power transistors than for small-signal types. In both cases, input and output circuitry must allow for the contribution of these capacitances, along with any stray capacitances in the package or circuit. Junction capacitance also varies with voltage, so any comparison of devices must note the voltage at which capacitance is measured. Typically, this will be the recommended operating voltage.

Threshold voltages (FETs) and currents (bipolars) indicate to designers how to overcome junction voltage drops at the base or gate. These tend to have a wide range, so precise design is not possible using published specifications. Together with DC gain specifications, general biasing requirements can be estimated.

Operational Specifications

The most important thing to remember about operational specifications is that they are measured using a specific test circuit. Occasionally, different parameters will be measured in different test circuits, so close attention to the footnotes in the data sheet is essential for accurate evaluation.

Gain, power output and efficiency have little meaning if the test circuit is unknown. Usually, the test circuit is designed for easy mechanical handling of test devices, rather than for optimum electrical performance. This means that the performance achieved in a user's similar circuit may actually be somewhat better than specified. Also, whether the test circuit is broadband or fixed-tuned will alter the specified performance. A broadband circuit will deliver lower gain and efficiency that a fixed-tuned circuit that has been adjusted to various frequencies across the measured band. In order to eliminate a manufacturer's "specsmanship," be sure devices are tested in comparable circuits.

Ruggedness specifications for power transistors also need the test circuit for reference. Hopefully, specifications for tolerance to an overdriven condition and output VSWR are realistic. Shorted or open output, high collector voltage and up to 50 percent excess drive represent extreme, but very possible conditions for most power transistors.

Noise Figure is typically measured in a tuned test fixture, since input match affects performance. The most accurate specifications will give best possible noise figure, noise figure with input matched to 50 ohms, and supply the available gain under both conditions.

Gain is specified based on testing in a standard commercial test fixture, assuring consistency and comparability.

Impedance parameters may be either in the form of S parameters or R ±jX impedance data. This data is obtained by placing the transistor in a test fixture and tuning input and output matching networks to match 50 ohms. By transforming 50 ohms through the known matching network values, the device impedances can be derived. This can be measured with a network analyzer by removing the transistor, terminating the input or output port, and measuring the impedance at the transistor connection port. The user should also find out how testing is applied to production devices, and how consistent the data sheet parameters are for a quantity of transistors.

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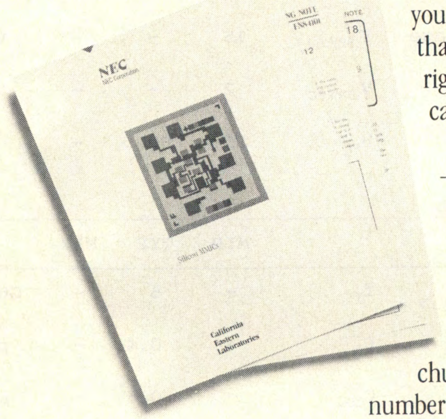
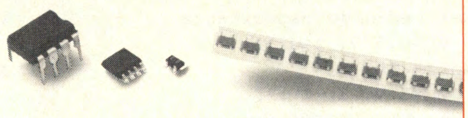

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Thermal Considerations

A brief review of transistor thermal specifications includes some warnings, just like the electrical specifications.

Power dissipation is usually measured with the device case held constant at 25C, which is not a practical situation. However, this is generally consistent among manufacturers, so comparisons are valid.

Thermal resistance (T_{jc}) is a measure of heat flow capability from the transistor junction to the device's case. If the maximum die temperature is known, the power dissipation can be readily calculated for any case temperature. Cautions include the fact that thermal resistance increases with temperature, and that conservative design will increase reliability. With modern high-dynamic range class A designs using significant collector or drain current, both power and small-signal devices must have careful thermal analysis to maintain safe junction temperatures.

General Considerations

Additional information may be required to fully understand test methods and results. This may be published with the data sheets, or it may be obtained directly from the manufacturer.

As noted above, the level of testing in production should be known. Which tests are applied to all devices, DC and operational? What spread of values can be expected in the specifications? What does the statistical distribution look like between the published maximum and minimum specifications (is the *typical* spec really the most likely value)? Answers to these questions are essential for modern computer-aided design. Monte Carlo analysis and manufacturing yield analysis require statistical models for all components.

Are operational parameters obtained by statistical analysis of test results for many devices, or are a few devices (or even a single device) selected for measurement. If so, how are these sample devices chosen — at random, by DC performance, or using other criteria?

Also, variations due to different packaging options should be known. This is generally done, but the designer must be aware that the surface-mount and TO-39 versions of the same device will have slightly different electrical and thermal specifications due to the lead structure, and the mass and surface area of the package.

Finally, take some time to compare the data sheet formats of different transistor manufacturers, including those who do not make many RF-specific devices. By learning what is industry-standard and what is unique to each manufacturer, an RF engineer can more wisely select the right transistor for his or her application. ***RF***

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2. *FET Design Catalog*, Siliconix, Inc.
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4. *Semiconductor Data Book*, Avantek, Inc.
5. *RF Device Data*, Vols. I and II, Motorola, Inc.

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UPB584	0.5–2.5GHz	5V	18mA	2
UPB585	0.5–2.5GHz	5V	26mA	4
UPB586	0.5–2.5GHz	5V	28mA	512/256
UPB587	0.5–1.0GHz	2.2V to 3.5V	5.5mA	2, 4, 8
UPB588	0.5–2.5GHz	5V	26mA	128/64

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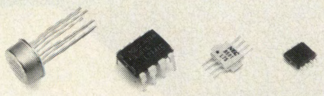

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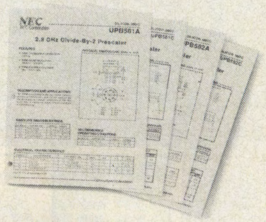

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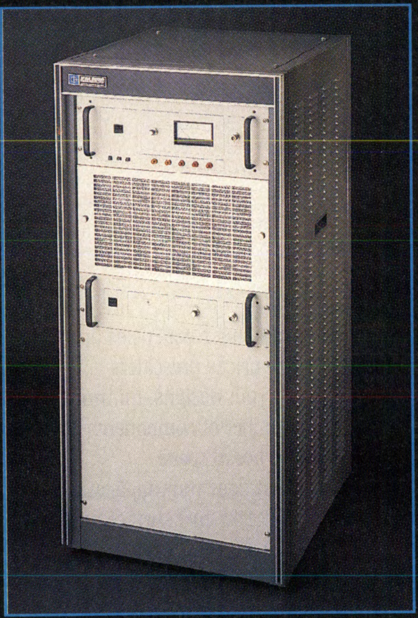

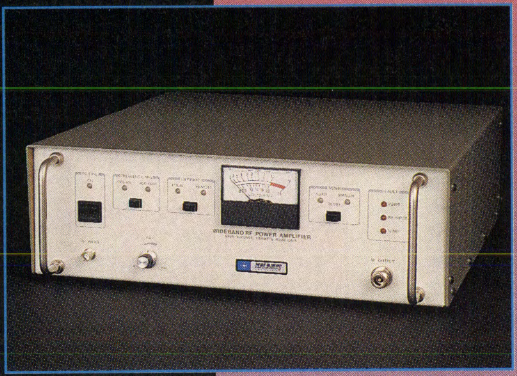

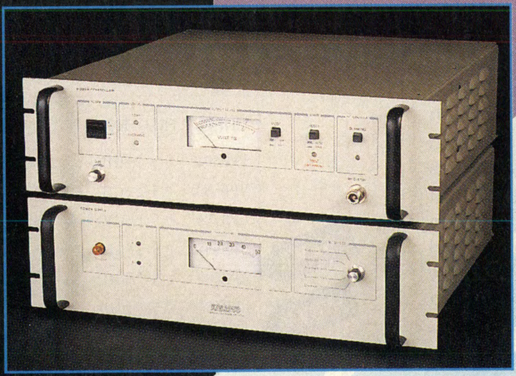

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Noise Cancellation in RF Detectors

By George Kassabian
University of California, San Diego

The amplification of the signal from a crystal detector is hindered by DC (from the carrier) and by very low frequency noise. AC coupling and high pass filtering can remove these unwanted components at the input, but those techniques also introduce phase errors and phase noise. This circuit actively cancels the DC carrier and suppresses the very low frequency noise while preserving phase integrity. This allows maximization of the signal of interest by a subsequent lock-in amplifier.

Low frequency AM signals (30 to 300 Hz) in microwave circuits are plagued by very low frequency (DC to several Hz) noise. The source of the noise can be quite variable. It can be due to microphonics, mechanical vibrations in the waveguide and cabling, noise inherent to the YIG oscillator or Klystron. Also carrier amplitude variation due to power noise or mismatch and ground loops between detector and amplifier can contribute additional noise.

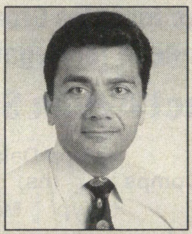
A simple circuit, (see Figure 1), is employed to significantly reduce the effects of very low frequency noises including power line related noises. The circuit is designed around National Semiconductor's LMC669 autozero IC.

The microwave crystal detector is properly matched and DC coupled to a low noise, high gain amplifier (Linear Technologies' LT1028). The noise cancellation IC is coupled to the output through lowpass filters to ensure only DC and very low frequencies are sampled. An error signal is generated at the output of the LMC669 and is fed back to the summing node of the inverting amplifier. A lowpass LC filter is used to suppress any clock noise from the LMC669 and a trimpot is used to set the loop gain.

Additional advantages of this circuit are bias current in the crystal or the power of the carrier can be varied without saturating the high gain amplifier, and no AC coupling capacitor is required. Also, there is reduction in opamp offset errors (typically 5uV) regardless of gain setting.

Relative noise measurements were performed by measuring total AC noise through a 100 Hz bandpass filter using a spectrum analyzer. The RF source is a stabilized Klystron feeding a tuned cavity via five feet of waveguide and a variable attenuator. A microwave AM detector is coupled to the main waveguide through a directional coupler and the power is set to bias the detector to nominal current of 200uA at 500 ohms load. See Figure 2. **RF**

About the Author



Since 1984, George Kassabian has been designing scientific instruments and providing electronic consultation to researchers and graduate students as development engineer at the University of Califonia at San Diego Physics Department. He can be reached at U.C.S.D., Physics Dept. 0319, La Jolla, CA 92093.

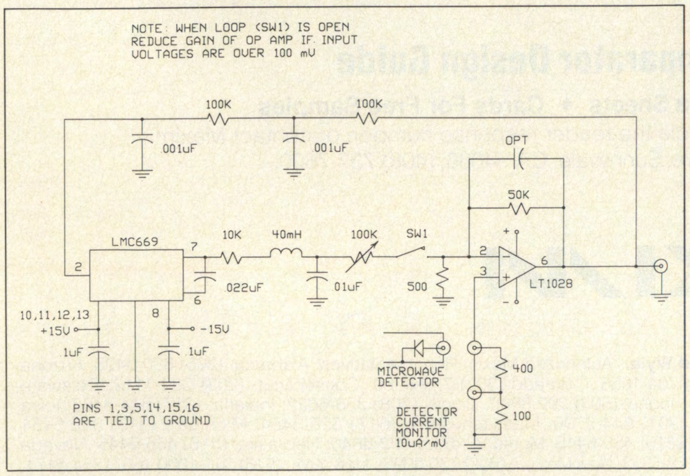


Figure 1. Schematic of the noise canceling circuit.

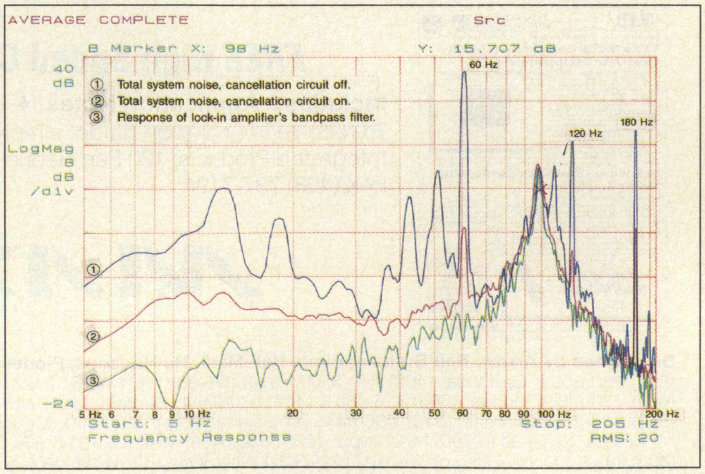


Figure 2. Frequency response curves.

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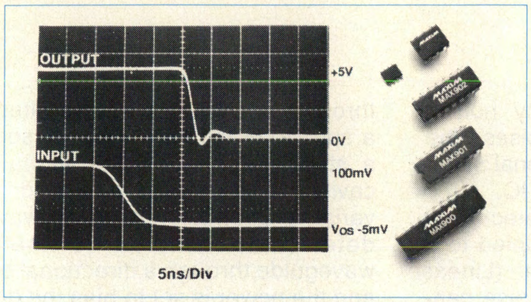

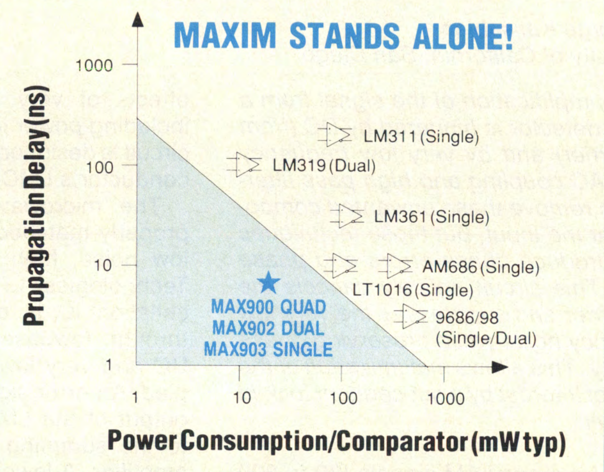

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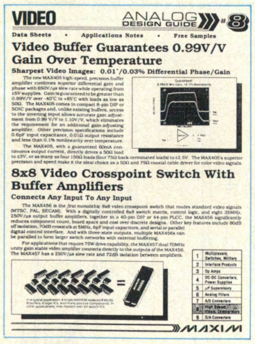

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High Dynamic Range AM Detector

By James E. Raiser
Scientific Atlanta

This circuit was developed in response to a need for a very low cost AM detector to be used on an IF signal with a large dynamic range. Cost considerations prevent the addition of an AGC loop, yet a diode detector requires a signal of about 1 volt peak to peak minimum, (forward biasing the diode can reduce the voltage requirement some). The system's power supply was 5 volts DC which limits the dynamic signal range to a maximum of 14 dB.

The AM detector uses two, readily available low cost ICs, (see Figure 1). U1 is a dual comparator of which only one section is used. U2 is a quad analog switch of which only one section is used. The circuit uses the comparator to turn the analog switch on and off each time the AM signal passes through zero. The analog switch then functions like a diode, only passing the signal when it goes above zero carrier, (assuming the signal would be driving the diode's cathode).

The AM signal is input at C1 and C2. R1 and R2 set a reference voltage for the comparator U1A. The minus input of the comparator (pin 4) is low pass filtered to remove any RF signal and allow only the DC voltage of the R1 and R2 divider to reach the minus input. The AM signal drives the plus input (pin 5) and when the carrier goes low, which is below the R1 and R2 divider voltage and the offset voltage of the comparator, the output of the comparator goes high. The high output of the comparator causes U2A to turn on and pass the AM signal, as a diode would. When the AM signal swings above the R1 and R2 divider voltage and the offset voltage U1A's output will go low and turn off the analog switch U2A again, as a diode would.

The output of the analog switch looks like a diode detected signal; half cycles of the AM signal whose amplitude changes with the modulation content of the signal. A low pass filter formed by R5 and C4 filters out the RF content of the signal and passes only the baseband information that was on the AM signal.

This circuit requires the switching speed of the comparator and the analog switch to be much faster than the cycle time of the AM carrier. To work with a higher IF frequency a faster analog switch and comparator would be required. The offset voltage of the comparator limits the lowest signal level to which the circuit will respond. The LM 339 has an offset of 2 mV zero to peak or 4 mV peak to peak. To work with lower level signals a comparator with a smaller offset voltage could be used.

In its application, the circuit saw an IF of 225 kHz with a dynamic range of over 50 dB. The high dynamic range of the detector eliminates the need for an AGC circuit, and the use of standard parts keeps cost low. ***RF***

About the Author
James E. Raiser is a Senior Staff Electrical Engineer with Scientific Atlanta where he is responsible for RF and analog design of CATV converters. He has a BSEE from Purdue University. He can be reached at 404-903-5329.

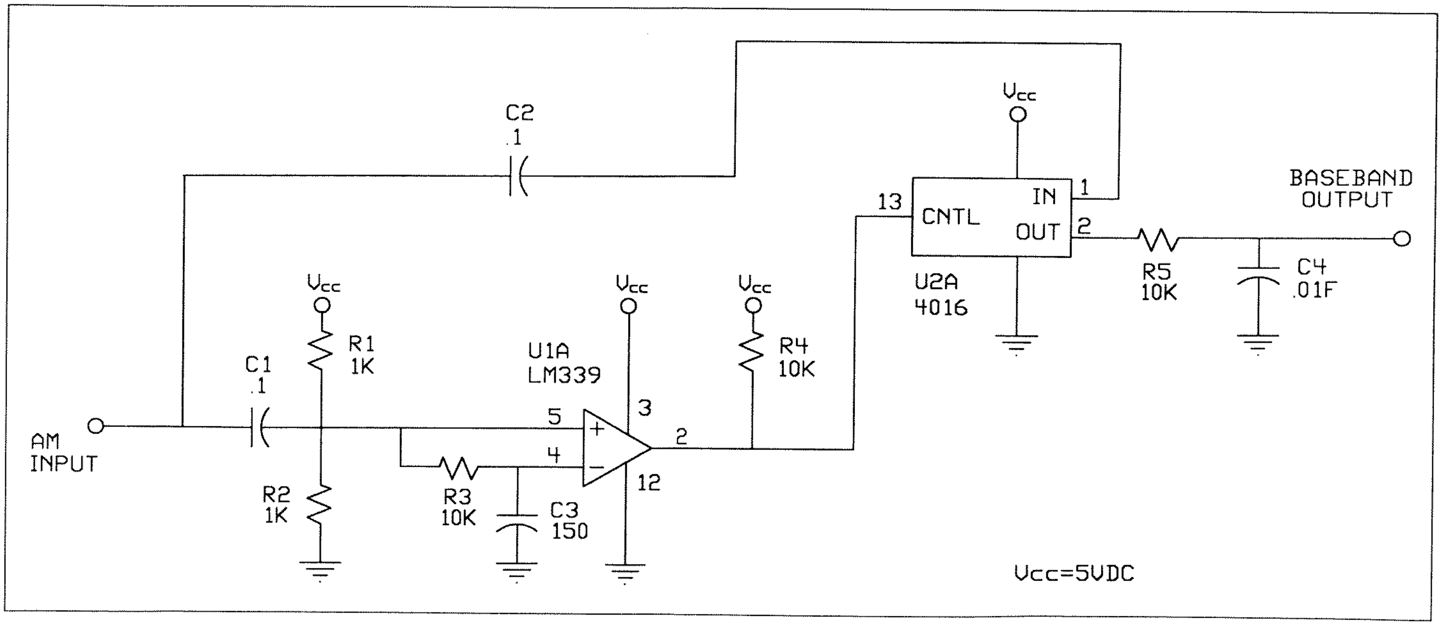

Figure 1. Schematic of the High Dynamic Range AM Detector.

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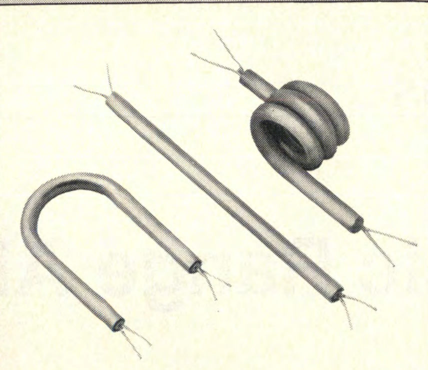

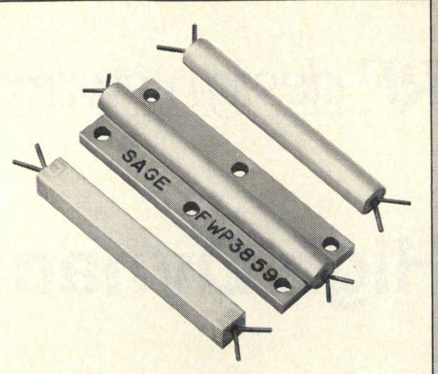

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FilDes: A Filter Design System for the RF Engineer

By Bob Lombardi
Harris GASD

FilDes is a program that was designed to allow the RF designer to complete the majority of LC filter designs without resorting to handbooks, graphs, and tables of values. If you can specify the filter in terms of general class (such as lowpass, highpass, bandpass, or bandreject), specify the frequencies you want to filter, and specify how much attenuation you want, FilDes can design the filter.

FilDes will design ladder networks having Butterworth or Chebychev response transfer functions in lowpass, highpass, wide bandpass (for low Q applications), narrow bandpass (coupled resonators for high Q applications) or band reject filters with up to 15 poles of attenuation and any degree of ripple desired up to 3 dB.

In addition, FilDes will design elliptic lowpass or highpass filters of symmetrical type.

In all cases, FilDes synthesizes the filter using equations in the literature. Butterworth and Chebychev filters all start as their lowpass prototype. No look-up tables are involved. The user can specify the order of the filter, or allow FilDes to calculate it. In the design of Chebychev filters, the user can define whether they would like to specify ripple bandwidth, as is common in some applications, or 3 dB bandwidth as is used in the Butterworth routines. FilDes can match differing input and output impedances for ladder filters with an odd number of poles. Filters are looked at from the source end, and the load specified as desired.

At every point in the program where output is presented to the user, the program pauses and waits for the user to hit any key. This allows you to write down values. Circuit descriptions follow the convention used in commercial analysis programs such as Compact® and SuperCompact® ; i.e., components are listed as either SER or PAR to describe their connection across the input terminals. SLC and PLC represent series LC and parallel LC circuits, respectively. Thus, a PLC SER is a parallel LC circuit connected in series from input to output, and a PLC PAR is the same combination connected in shunt to ground.

FilDes was written in Turbo Pascal, and runs on the IBM PC or clones under MS-DOS. Memory requirements are minimal; it will run on under 512K of memory. A math coprocessor is not required; however, using one will speed up operation of certain sections. All user input/output is in text mode windows, but color graphics of CGA quality or better is required.

Theory of Operation

FilDes calculates element values for Butterworth and Chebychev filters based on techniques found in many places in the literature. It is shown in many sources (1,2,3) that a Butterworth filter terminated in equal resistive impedances has its element values given by the series:

$$g[k] = 2\sin\frac{(2k-1)\pi}{2n} \qquad (1)$$

where k = 1,2,3,...,n. The method used for matching to unequal source and load impedances is discussed below.

A Chebychev filter requires more calculation to determine element values. There are two general methods for the specification of Chebychev filters, using either the 3 dB bandwidth or the ripple bandwidth. FilDes allows the user to specify the filter in terms of either its ripple bandwidth, or its 3 dB bandwidth. Any decimal degree of ripple can be specified, from .005 to 3 dB (3 dB is, admittedly, absurd, but a limit was needed that is likely to never be exceeded).

There are a couple of methods for the calculation of Chebychev filter element values. The most commonly presented method of calculation (1,3) is used in FilDes. It calculates the element values for equally terminated odd order Chebychev filters, or unequally terminated even order filters. Odd order Chebychev filters have zero relative attenuation at DC, while even order filters have a loss equal to the ripple. As calculated by FilDes, even order filters have a load impedance that you are stuck with. This is not unlike designing from tables, where the ratio of source to load (and vice versa) is fixed at several different values for a particular filter. Transforming an even order Chebychev filter to operate between equal impedances degrades the stopband attenuation, which is probably the reason for choosing this class of filter in the first place. Odd order filters allow greater flexibility in selection of unequal terminations.

The element values for Chebychev filters are given by:

$$r = \text{ripple in dB} \qquad (2)$$

$$\varepsilon = \sqrt{10^{\frac{r}{10}} - 1} \qquad (3)$$

$$\phi = \frac{1}{n}\operatorname{arcsinh}\frac{1}{\varepsilon} \qquad (4)$$

$$b = \ln\left(\coth\frac{r}{17.37}\right) \qquad (5)$$

$$p = \sinh\frac{b}{2n} \qquad (6)$$

$$g[1] = \frac{2a[1]}{p} \qquad (7)$$

(3 dB BW specifications multiply this by cosh(b))

$$a[k] = \sin\frac{(2k-1)\pi}{2n} \qquad (8)$$

for k = 1, 2,. . . n

$$b[k] = p^2 + \sin^2\frac{k\pi}{n} \qquad (9)$$

$$g[k] = \frac{4a[k-1]a[k]}{b[k-1]g[k-1]}$$

(multiply by $\sqrt{\cosh b}$ for 3 dB BW)

if n is even,

$$R_{load} = \tanh^2\frac{b}{4} \qquad (11)$$

for a matched filter.

Odd order filters have a symmetry that allows the implementation of

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Bartlett's Bisection Theorem (1,2). Bartlett's theorem states that a network can be bisected at its midpoint and its impedance scaled from the midpoint to the different impedance. In the normalized one ohm, one radian/second lowpass prototypes that are calculated in FilDes, this means that the impedance is changed by simply multiplying or dividing the values in one half of the filter by the new impedance. For example, a three pole Butterworth filter calculated by FilDes has normalized values of one Farad, two Henries and one Farad. The center component, here an inductor, is cut in half yielding two one Henry inductors in series. The half connected to the different impedance is scaled by multiplying by the new load. For capacitors, the new value is obtained by dividing by the new load. This is continued for the rest of the filter.

This routine only works for Butterworth or Chebychev ladder filters in FilDes. The program offers the option of increasing the order of even order filters wherever it calculates an even order for you, or whenever you choose an even order. In the lowpass prototype this amounts to the addition of a capacitor. In a highpass filter, it is an extra inductor. In bandpass or bandreject filters, it requires an extra LC combination. This may or may not be a good engineering compromise, so it is your decision. Of course, the vast majority of RF filters operate between matched 50 ohm impedances, so the transformation routines are likely to go largely unused.

FilDes calculates two different classes of bandpass filters. There are many different topologies of bandpass filters, and they each have their advantages. The two types here are a simple lowpass to bandpass transformation, and top-C coupled resonators. The first class is best suited to situations where low Q is in order. (Q here is defined as center frequency divided by bandwidth.) These are obtained by parallel resonating the shunt capacitors in the LP prototype, and series resonating the series inductor. The result is then impedance, bandwidth and frequency scaled.

For deciding which class is used, let 10 be the dividing line between high and low Q. Q values less than 10 use the first type.

The second class of filter relies on the narrowband approximation, and is best used in high-Q applications. The calculation of the values for these filters also begins with the normalized array of lowpass values, but the filter is described in terms of coupling coefficients

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and Q values. The equations for this type of characterization are found in many old sources, such as the older editions of the *ITT Handbook for Radio Engineers,* and Reference 4 herein. As these filters have an impedance that is dependent on the chosen inductor, a means for scaling the filter to a lower circuit impedance is included, namely a matching series capacitor. This type of matching circuit will not work to match to a circuit impedance higher than the filter's impedance.

Another feature of the bandpass filter routines is a calculation of insertion loss at mid-band. This allows comparison of filters with differing amounts of ripple, or between the two types of filters if the Q is near 10. The calculation depends on the unloaded Q of the components used; FilDes offers 500 as the default Q value, a good component in many applications. The equation is:

$$\text{Loss} = 4.343 \frac{f_o}{BW\ Q_u} \Sigma\, g[k] \qquad (12)$$

The result is most accurate for narrowband filters, but provides reasonable estimates up to an octave bandwidth. It is most accurate for bandwidths under 20 percent.

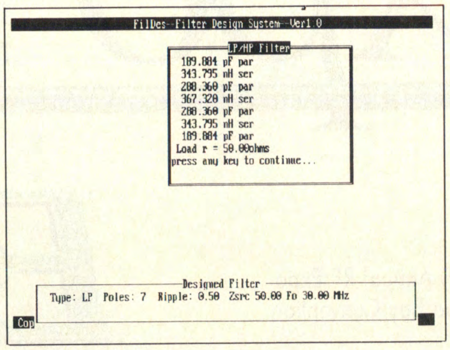

Figure 1. Output screen for the FilDes program.

FilDes will design bandreject filters derived from the lowpass prototype by first transforming to a highpass filter, then resonating the shunt capacitors with series inductors, and the series capacitors with parallel inductors. The resultant is then frequency and impedance scaled.

The elliptic filter algorithms in FilDes were first presented in a seminal paper by Amstutz (6). His program, written in FORTRAN, was translated into a Pascal procedure by the author. Other writers (5,8) have implemented the algorithm in dialects of BASIC; however, the Pascal translation presented here is the only one known by the author, and it is the first translation to eliminate "GOTO" statements. This section calculates filter component values for lowpass or highpass elliptic filters based on input of the number of transmission zeroes, frequencies for the stopband and passband, and desired ultimate attenuation.

Experienced elliptic filter designers may be upset at losing their old familiar form of specification, as in Zverev's classic work. These filters are specified by choosing the number of transmission zeroes and specifying the stopband ultimate rejection. Ripple is something it tells you, not something you specify. Still, it seems easy to converge on a design that performs as required.

To aid in this convergence, FilDes will also estimate the complexity of the required elliptic filter. The algorithm for this was presented first by Darlington in 1939 in the Journal of Mathematical Physics, and updated by Orchard (7). It relies on the calculation of values in a series solution to the elliptic function. The order of an elliptic filter is not the same as the number of transmission zeroes. Consequently, the order that the program estimates is not always suffi-

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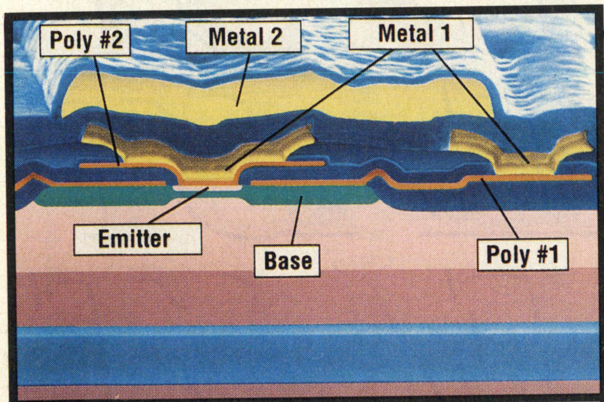

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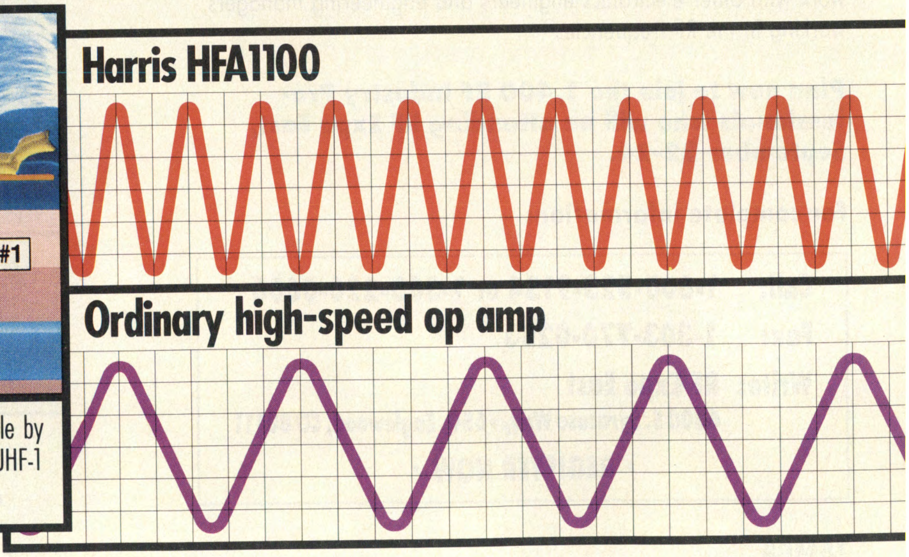

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And your creativity needn't stop with standard products. Because the HFA-1 process is available in semicustom, as part of Harris' industry-leading cient for determining the complexity of the required filter, so the order is included as an estimate for those who are not sure how big a filter is required. It sometimes causes the program to generate filters with ripple many times worse than that desired. In such cases, the user should try the filter with a manual input of at least one more zero than the estimate. It is for this reason that FilDes presents the calculated ripple and the desired ripple. Of course, when the calculated ripple meets requirements the filter will work with the estimated complexity.

Using FilDes

FilDes inputs and outputs are all carried out in windows at various places on the screen. In general, input is on the left and output is on the right. Output never scrolls off the screen; if the results would take up more than the remainder of the screen, output pauses for you to read or copy and gives you a "press any key to continue" prompt.

You are first prompted for the general class of filter, Butterworth/Chebychev ladder filters or elliptic LP/HP filters. This choice is selected by moving the highlighted bar with the cursor up/down keys and then hitting return. After this, the input is all done by keyboard entry. For us fat-fingered typists there is no need to fear, the inputs have been "idiot-proofed" for incorrect inputs. Still, a word of caution is in order: error trapping can't remove something that could make sense, for example, saying you want a 500 ohm filter instead of a 50 ohm filter. FilDes will design what you say you want. In order to allow you to check what you entered, FilDes displays a window along the bottom of the screen containing a summary of the filter's parameters. (Figure 1).

Like all software, FilDes is only 95% done. There are many changes envisioned, including synthesis of filters from provided transfer functions in s, and thereby calculation of component values for filter topologies for which simple closed form calculations do not exist. A chain matrix program for plotting filter responses is also envisioned to enhance ease of use.

This program is available on disk from the RF Design Software Service. See page 87 for ordering information. ***RF***

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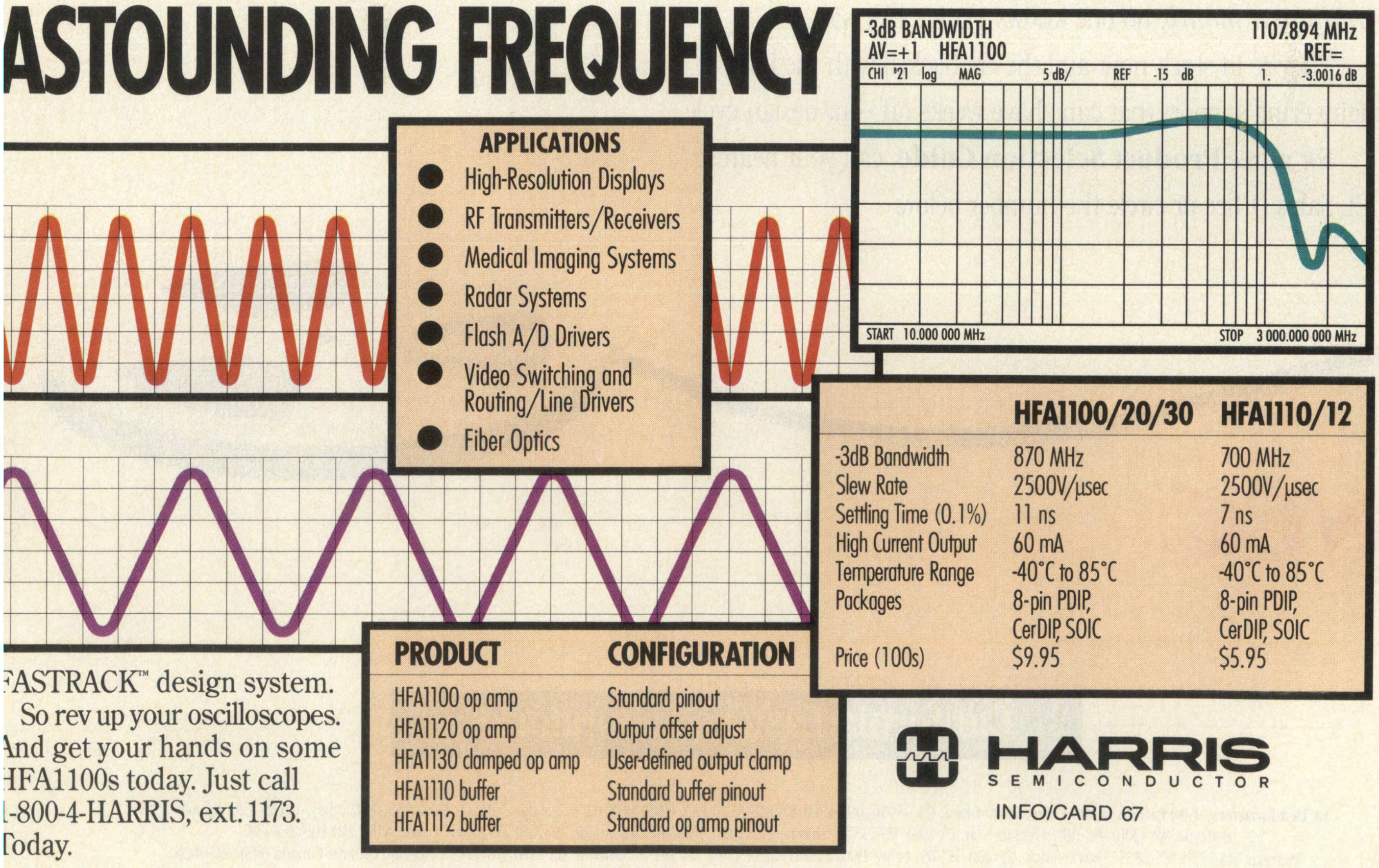
About the Author

Bob Lombardi is a senior engineer at Harris Government Aerospace and Systems Division, where he works on various RF designs including receivers, filters and synthesizers. He can be reached at Harris GASD, P.O. Box 94000 MS 102-4826, Melbourne, FL 32902, or at (407) 727-6360.



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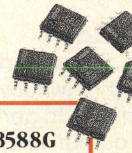

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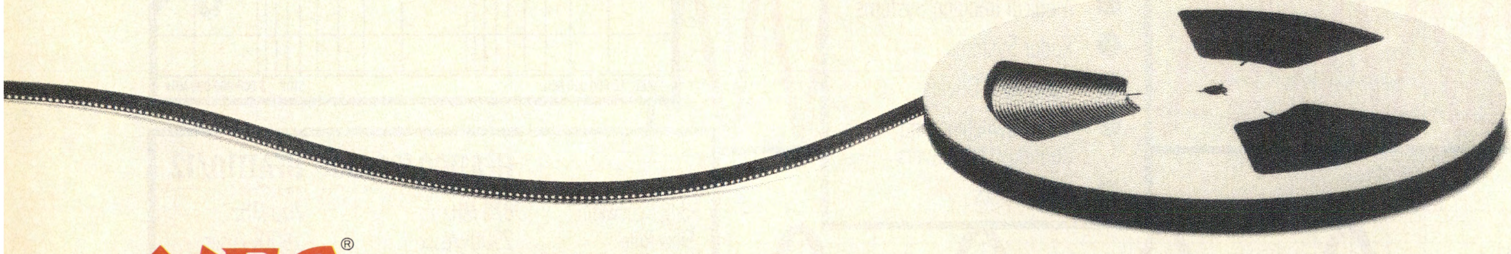

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INFO/CARD 68

NAVSTAR Global Positioning System

By Edward S. Troy
Aerospace Consulting

NAVSTAR GPS (Global Positioning System) is a military sponsored system that is revolutionizing navigation, surveying, and time keeping and transfer. In addition to the far-reaching effects on the fields mentioned above, it allows for many other systems and tools that until now had only been "science fiction", to become reality. For example, GPS provides the needed positional accuracy to allow for useable moving map displays in cars. GPS will undoubtedly spawn other systems and devices that have not even been imagined.

Navstar GPS was developed to become the navigation and time transfer system for the 21st century. It utilizes the latest technology in terms of frequency, stability, tracking capability, modulation techniques, and computer technology to correct the faults of the previous systems and provide a reliable service that will represent the state of the art for years to come. (The Soviet Glonass system is almost identical, and will not be discussed here, except to note that GPS receivers can probably be designed to utilize both GPS and Glonass, further increasing system reliability and accuracy). GPS receivers saw their first combat usage during the recent Persian Gulf War. More than 4000 military units and 5000 civilian units were in use in the Gulf. In addition, SLAM missiles utilized during the war used single channel GPS receivers to generate midcourse corrections (1). Also, a February 7, 1991 article in the *Wall Street Journal* reported that Iraqi forces were using GPS receivers to fix the position of their mobile SCUD missile launchers (2) (This is why the military demands a degraded accuracy for non-authorized users).

GPS consists of several different segments: the ground based control segment, the user segment, and the space segment. The overall system was conceived of and developed under the direction of the U.S. Air Force Systems Command, Space Division/CWN, Navstar GPS Joint Program Office. The space segment consists of the satellites. The user segment consists of all users, which can be based on land, sea, air, or even in space. The control segment consists of a master control station at Falcon Air Force Base in Colorado Springs, Colorado, plus four monitor stations scattered around the world.

Using GPS, a user can derive information on precise latitude, longitude, altitude, and time. This information can then be used for such applications as speed determination, navigation, surveying, time transfer, network synchronization, etc. There are two basic classes of users, authorized and unauthorized. Authorized users consist of the U.S. military, NATO, and selected military forces. These users will experience accuracies of 16 meters or better. All other users, such as civilians, are unauthorized users. These users will experience accuracies that could be as bad as 100 meters (although the exact degree of accuracy degradation will be determined by the GPS Joint Program Office).

GPS is a space-based system that utilizes a constellation of 21 satellites and three spares. These satellites are in orbits with a height of about 10900 miles and an inclination of 55 degrees. This orbit translates into an orbital period of slightly less than 12 hours. The actual orbit is established in such a way that at least four satellites will be visible from any point on the earth at any time.

As of the writing of this article, a total of 24 GPS satellites have been launched. The original 11 Block I satellites were launched between 1978 and 1985, and were intended for system test and evaluation. Block II satellites, intended for general use, were launched starting in 1989. The long time delay between the Block I launches and the Block II launches was due to the *Challenger* disaster. At the present time, there are 17 useable satellites in orbit, although five of these are Block I satellites. This is sufficient for two-dimensional position fixing 24 hours per day and three-dimensional position fixing about 21 hours per day over most of the world (3).

GPS satellites, like all objects that orbit the earth, obey the normal laws of orbital motion first described by Kepler and later expanded upon by Newton. One of the results of Kepler's laws is that the period of a satellite orbit is determined by the mean motion, n, of a satellite.

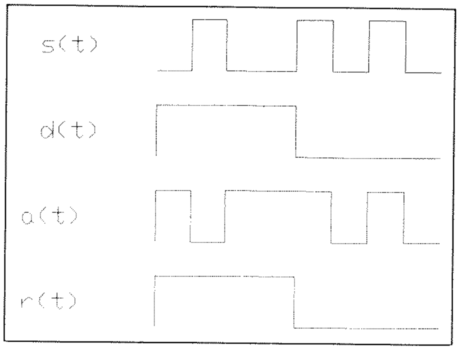

Figure 1. Digital data d(t), spreading signal s(t), and transmitted data a(t). r(t) is de-spread recovered data.

$$P = \frac{2\pi}{n} \quad (1)$$

In the above equation, p is the period of the orbit, in seconds, and n is the mean motion, or the angular velocity of the satellite in radians per second. n is simply a function of the semi-major axis of the orbital ellipse.

$$n = \sqrt{\frac{u}{A^3}} \quad (2)$$

A is the semi-major axis of the ellipse, which in this case is the sum of the height of the orbit, 20,200 km, and the semi-major axis of the earth ellipsoid, which, according to WGS 84 (World Geodetic System 1984) is 6,378,137.0 meters (4). u, the universal gravitational constant, is 3.986005×10^{14}. Using these values,

$$n = \sqrt{\frac{u}{A^3}} = \sqrt{\frac{3.986005 \times 10^{14}}{(2.657814 \times 10^7)^3}}$$

$$= 1.457 \times 10^{-4} \text{ seconds} \quad (3)$$

$$P = \frac{2\pi}{n} = \frac{2\pi}{1.457 \times 10^{-4}} \quad (4)$$

$$= 4.31 \times 10^4 \text{ sec.} = 11.98 \text{ hrs.}$$

Of course, orbital period is only a small amount of the information that is required to determine the precise position of a satellite in space in reference to the earth. Other required information includes inclination, right ascension, and the argument of the perigee. A full description of the orbital data can be

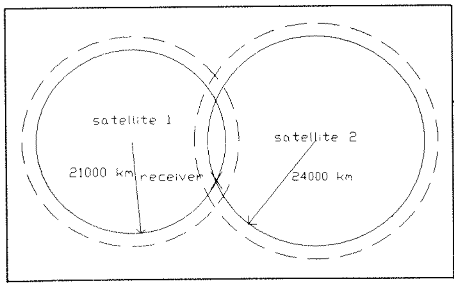

Figure 2. Illustration of distance errors due to clock inaccuracy.

calculated, however, by applying Kepler's basic laws and adding corrections to compensate for such effects as solar wind, friction with the earth's atmosphere, gravitation effects from other bodies such as the sun and moon, the magnetic field of the earth, and changes in the earth's gravitational field primarily due to the tides (5). All of these various orbital elements, as well as correction factors to the orbital elements, are termed ephemerides. These ephemerides are determined by the master control station and associated tracking stations. In order to accurately model and predict actual orbital positions, as of 1981, the earth tracking stations use mathematical models with terms out to the 41st order. (In fact, to model the gravity field of the earth, a 180th degree model is available, and is used for some purposes). The current state-of-the-art in determining a satellite's position from a ground site is on the order of 1 meter (6). Ephemeris data is transmitted to the satellite by the master ground station, and then retransmitted from the satellite to the user as part of a nav message. The nav message contains ephemerides for that satellite for the next four hours, plus approximate orbit information for each of the other satellites. It can take up to three minutes to collect the current ephemeris information on a given satellite, and up to 12.5 minutes to collect all almanac data for the other satellites in the system. The ephemeris information can be degraded by the master control station so as to deny non-authorized users full accuracy. This is often done by encrypting some of the information, and only allowing authorized users access to the decryption key. In addition to ephemeris information, the satellite nav message includes information pertaining to clock correction, system status messages, ionospheric propagation model parameters, etc.

Spread Spectrum Technology

Spread spectrum technology, the enabling force behind the Global Positioning System, allows for all satellites to operate on common frequencies without interference while at the same time allowing for precise time transfer.

In its simplest form, spread spectrum is any communications system that uses a bandwidth that is very wide compared to the information bandwidth to transmit information. This would include, for example, FM communications, since in many FM situations, the modulation factor is four or five, implying that an RF bandwidth of four or five times the modulation bandwidth is required by the channel. Spread spectrum, however, does not refer to such small spreading factors. Generally, in a spread spectrum system, the transmission bandwidth is often hundreds, thousands, or even hundreds of thousands of times greater than the information bandwidth. For example, with GPS, the chipping rate, which determines the required RF bandwidth, is 1.023 MHz on one channel and 10.23 MHz on the other channel, while the information bandwidth is 50 Hz. On the slower chipping rate channel, this yields a spreading factor of

$$\text{spreading factor} = \frac{\text{chipping rate}}{\text{information rate}}$$

$$= \frac{1.023\ \text{MHz}}{50} = \frac{1.023 \times 10^6}{50}$$

$$\text{spreading factor} = 20460 \qquad (5)$$

But why would one want to make the required RF bandwidth any larger than is necessary for normal communications? The answer for this is varied depending upon the reason for using spread spectrum. One of the principle reasons for using spread spectrum is its inherent lack of detectability. That is, as a signal is spread out over a broad bandwidth, the energy density at any individual frequency, or narrow band of frequencies, is greatly reduced. For example, in normal communications techniques, if it is desired to reduce the noise floor of a receiving system, one of the first things that is done is to reduce the bandwidth of the system. This is because the ultimate noise floor of a communications system is determined by the thermal noise of the system. Thermal noise of a system is described by the equation

$$N = kTB \qquad (6)$$

where N is the noise power in watts, k is Boltzmann's constant (1.38×10^{-23} Joules/kelvin), T is the temperature in kelvins, and B is the bandwidth in Hertz. At a room temperature of 300 kelvins and a bandwidth of 1 Hertz, this becomes

$$N = (1.38 \times 10^{-23})(300)(1)$$

$$= 4.14 \times 10^{-21}\ \text{Watts} \qquad (7)$$

which, if you divide both sides by 1 Hertz, and convert from watts to dBm, yields a noise floor of −174 dBm/Hz. If you were then to look at a bandwidth of 1000 Hz, the noise floor would be increased by 30 dB (10Log(Bandwidth)). This would equal a noise floor of −144 dBm. For the spread spectrum systems mentioned above, the required receiver bandwidths would be twice the chipping rates, or 2.046 MHz and 20.46 MHz, respectively. These figures translate to noise floors of −111 dBm and −101 dBm respectively. At the same time, if a transmitted signal is spread out over a broad bandwidth, the energy spectral density per Hertz is reduced by the spreading factor. Thus, for a 1 milliwatt, or 0 dBm signal spread over a 1.023 MHz bandwidth, the energy spectral density would be

$$E = \frac{0\ \text{dBm}}{1.023 \times 10^6\text{Hz}} = -60\ \text{dBm/Hz} \qquad (8)$$

This obviously makes the level of the signal, at any observation bandwidth, much closer to the level of the noise. In fact, in many spread spectrum systems, including GPS, the received signal strength is below the noise floor. Specifically, in the case of Navstar, the minimum received signal strength on the narrow channel varies from −130 dBm to −136 dBm (7). As we noted earlier, the thermal noise floor on this channel bandwidth is −111 dBm, and thus the received signal strength is 19 to 25 dB below the noise floor. This is certainly not a signal that will stand out on a spectrum analyzer.

This is not the primary reason that GPS is a spread spectrum system, however. There are other benefits to communicating with signals that are below the noise floor besides low detectability. One is that many signals can coexist on the same channel without

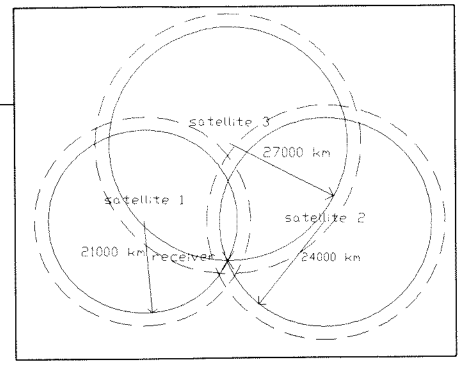

Figure 3. Using a third satellite to pinpoint location of receiver.

interfering with each other. Obviously, if a signal is below the thermal noise floor, it is not likely to interfere with other communications channels trying to use that channel. The question becomes, however, if the signal is below the noise floor, and therefore apparently not detectable, of what use is it to a communications system? The answer lies in the fact that under very specific circumstances, the bandwidth of the channel can be collapsed to a width corresponding to the width of the information bandwidth. Since, in this case, we said that the information bandwidth is 100 Hz, corresponding to the 50 bit per second (bps) information signaling rate, this represents a channel bandwidth reduction of

$$\frac{2.046 \times 10^6}{100} = 20460 \qquad (9)$$

which, as an earlier derivation shows, is equal to 43 dB. If you thus reduce the noise bandwidth to 100 Hz, the noise floor now drops to −154 dBm. This is now 18 to 24 dB below the level of the received spread spectrum signal. Now you have a solid signal with which to work. The way this works is related to communications theory, and will be explained next.

To generate a spread spectrum signal, the information signal is added to a pseudorandom spreading signal with a binary, modulo-2 adder (exclusive or'ed). Thus, if both signals are a 1, or both are a 0, the output is 0, while if one or the other is a 1, the output is 1. Thus, if the data signal is d(t) and the spreading signal is s(t), Figure 1 shows what the transmitted data, a(t), would look like (although only several chips are shown per bit, as opposed to the actual system which uses 20,460 chips per bit, as discussed earlier).

The pseudorandom spreading signals, or PN codes, are generated by modulo-2 additions of several predefined pseudorandom sequences. These sequences are known as Gold codes, and they have very low cross-correlation characteristics. The 1.023 Mbps code is known as the C/A code, or coarse acquisition code. It is the only code that is officially available to non-authorized users. It repeats itself every 1023 bits, or 1 msec, and there are 37 different codes. Each satellite is assigned a unique code. The 10.23 Mbps code is known as the P-code, and it repeats itself every 267 days. This signal may or may not be available to civilian users depending on the whims of the Department of Defense.

The system uses two frequencies, L1=1575.42 MHz and L2=1227.6 MHz. These two frequencies are exact multiples of the system clock frequency, 10.23 MHz, as are the chipping rates and the data rate. Thus, all signals can be derived from a common source and phase coherence can be maintained throughout the system. The P-code is transmitted on both frequencies, while the C/A code is only transmitted on L1. Thus, only L1 is reliably available to the general user.

In order for the system to work, the receiver must know how to generate the various pseudo-random sequences (Pseudo-noise, or PN codes). Then, if the receiver is generating a PN code that is an exact replica of the transmitted code, and that code is lined up in time with the received code, a correlator can be used to detect correlation of the codes, at which time the correlator will exhibit a maximum output. Then, with the two PN signals synchronized, the detected signal must be multiplied (modulo-2 added) with the locally generated PN sequence. The result of this modulo-2 addition of s(t) and a(t) is r(t), which is the same as the original data signal, d(t), as shown in Figure 1. In this way, the received signal is de-spread, and the original data signal is recovered.

In order to explain exactly how GPS determines location and time, it is probably easiest to assume a perfect world, with perfect equipment, and describe how the system performs under those conditions. Then, a simple experiment will be described that tells how the system performs under less than perfect conditions.

First, we will assume that we are in a two-dimensional world. Our model should work fine, except that we must remember to add additional complexity at the end of the discussion to account for a three-dimensional world. Also, in the beginning we will assume that all clocks in the system are perfect and are perfectly synchronized. We will also assume that there are no doppler effects, relativistic effects are non-existent, the distance to satellite 1 is 21,000 km, the distance to satellite 2 is 24,000 km, and the speed of light is 3×10^8 meters per second. Since time equals distance divided by speed, the signal from satellite 1 will reach the receiver in 70 ms (21,000 km divided by 3×10^8). Similarly, the signal from satellite two will reach the satellite in 80 ms. Thus, we can draw two circles (Figure 2) centered on the satellite positions, with the radius of those circles representing the range, as determined by time delay, to each of the two satellites. These two circles intersect at two points. So, we have to be at one of those points. One of those points can usually be eliminated because it is in an unreasonable location, like somewhere between the moon and the earth. Of course, if you were an astronaut, maybe that would not be an unreasonable answer, but then the alternate location would probably be unreasonable to an astronaut. So, with two satellites, we should be able to precisely locate our position, right? Wrong. Let's now say that the clock in our receiver is not perfect. Let's say it runs 10 ms slow compared to the clocks in the satellites. Well, the receiver now thinks that the signal from satellite 1 took 80 ms to reach the receiver, and the signal from satellite 2 took 90 ms. If we redraw the diagram, now with the range to satellite 1 being 24,000 miles (speed of light times 90 ms), while the distance to the second satellite is 27,000 miles. This situation is shown by the dashed circles in Figure 2. Thus, if the clock is not exactly synchronized with GPS time, the range to each of the satellites (called Pseudo-range) will be incorrect. Another way to think of this is to say that when we try to find the range from two satellites when we know time precisely, we are trying to solve two equations with two unknowns. This is fine. However, if we do not know time precisely, then we are trying to solve two equations with three unknowns (range to satellite 1, range to satellite 2, and time).

So, how can this problem be resolved? Referring to Figure 3, let's try to find the range to another satellite, again keeping our knowledge of time perfect. Satellite 3 has a range of 27,000 km. This range translates into a time delay of 90 ms. A quick glance at Figure 3 shows that all three range circles intercept at one point, as they should. This confirms our earlier measurements, and also removes any doubt as to the location of the receiver. Now, referring to the dashed lines, let's assume that the receiver clock is again slow by 10 ms. This gives a range to the third satellite of 30,000 km. As the intersection of the three dashed circles shows, there is no one spot where all three circles intersect, so there is no way of knowing your precise location. However, if we move the receiver clock forward by 10 ms, we now find that all three circles intercept at the same point, thus uniquely identifying the location of the receiver (8). At that time, all of the clocks are synchronized. Thus, if the receiver clock is not perfect, a two-dimensional position fix requires the use of three satellites. In a similar way, a three-dimensional fix requires the use of four satellites.

Another way to think of the process involves a review of the equations that must be solved to determine position. These equations are shown below, where R1 represents the range to satellite 1; X, Y, and Z represent the positions of the satellites; UX, UY, and UZ represent the user positions; and CB represents the offset of the user clock from true GPS time.

$$(X1 - UX)^2 + (Y1 - UY)^2 + (Z1 - UZ)^2 = (R1 - CB)^2 \tag{10}$$

$$(X2 - UX)^2 + (Y2 - UY)^2 + (Z2 - UZ)^2 = (R2 - CB)^2 \tag{11}$$

$$(X3 - UX)^2 + (Y3 - UY)^2 + (Z3 - UZ)^2 = (R3 - CB)^2 \tag{12}$$

$$(X4 - UX)^2 + (Y4 - UY)^2 + (Z4 - UZ)^2 = (R4 - CB)^2 \tag{13}$$

This represents a system of four simultaneous equations with four unknowns. The determination of position and time is simply a matter of finding the solution of these equations, and since there are no more unknowns than equations, a unique solution is possible (9).

Of course, if the receiver has a very accurate system clock (such as an atomic clock), complete three-dimensional solutions can be had with measurements to only three satellites. Or, if any of the other parameters pertaining to the user position are known with a high degree of accuracy, one or more of the pseudorange measurements can be eliminated.

There are many sources of error in this system. However, any that can be controlled by the system are tightly controlled. For example, each satellite has its own atomic clock that is continually corrected by ground control. Also, the actual frequency of the atomic clock is adjusted before launch to a frequency of 10.22999999545 MHz to correct for the relativistic effects of reduced gravity and high orbital velocity. Other error sources are related to the atmosphere and ionosphere, and can be corrected by noting, differentially, the time of travel and the relative phase differences of the L1 and the L2 signals.

The least important relativistic correction is that due to the velocity of the satellite. The Lorentz transformation describing the frequency shift due to special relativity, where C is the speed of light, V is the satellite velocity, ΔF represents the frequency shift, and F represents the frequency, is shown below.

$$\frac{\Delta F}{F} = \frac{1}{2}\frac{3874^2}{(3 \times 10^8)^2} = 8.338 \times 10^{-11} \tag{14}$$

Inserting the speed of light (3×10^8 meters per second) and the speed of the satellite (3,874 meters per second) yields the following results:

$$\Delta\frac{F_g}{F} = \frac{u}{C^2}\left(\frac{1}{R} - \frac{1}{r}\right) \tag{15}$$

This states that the speed of the clock in space slows down by about eight parts in 100 billion.

The most important relativistic correction is that due to general relativity. This is the correction due to reduced gravity in space, and it is described by the equation:

$$\Delta\frac{F_g}{F} = \frac{3.986 \times 10^{14}}{(3 \times 10^8)^2} \tag{16}$$

$$\left(\frac{1}{6378000} - \frac{1}{26580000}\right) = 5.3 \times 10^{-10}$$

where ΔF_g is the frequency correction due to reduced gravity, u is the gravitational constant (given earlier), R is the earth's radius, and r is the radius of the satellite's orbit. Substituting into this equation, we get

$$\frac{\Delta F}{F} = \frac{1}{2}\frac{V^2}{C^2} \tag{17}$$

This states that the clock in space speeds up by about five parts in ten billion. This is the predominant effect, but both of these effects must be taken into account to arrive at the final oscillator frequency offset mentioned above (10).

Another area that has achieved widespread use is differential GPS. In this technique, two receivers are used. One is placed at a known location, and the other is placed at a location to be determined. Then, both receivers receive signals simultaneously from the same satellites. Since the received signals both pass through the same atmosphere and ionosphere (approximately) at the same time, any errors due to these effects cancel out. In a similar way, most of the other errors of the satellite clocks and satellite ephemeris information also cancel (since you are really only measuring a time difference of reception between a known location, the reference receiver, and the unknown location). The signals received at the two receivers are stored over a period of minutes or hours and then post-processed to determine the exact location of the test receiver. This technique eliminates almost all systematic errors, and accuracies of 0.1 ppm have been achieved. This corresponds to centimeter accuracies over a 100 km baseline (11). In fact, recent advertising literature from WM Satellite Survey Company, backed up by a study performed by that company, advertises differential GPS accuracy of 5 mm plus or minus one part per million! Their data shows accuracies, over a precisely measured 3 km baseline, of less than 1 mm with a measurement time of as little as ten minutes (12).

Summary

GPS is a new positioning system that is here to stay. It uses the latest in communications technology, and promises to completely revolutionize the navigation, time transfer, and surveying industries. ***RF***

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About the Author

Edward Troy is an engineering consultant in the RF and microwave field. He can be reached at Aerospace Consulting, PO Box 536, Buckingham, PA 18912-0536. Tel: (215) 345-7184. Fax: (215) 345-1309.

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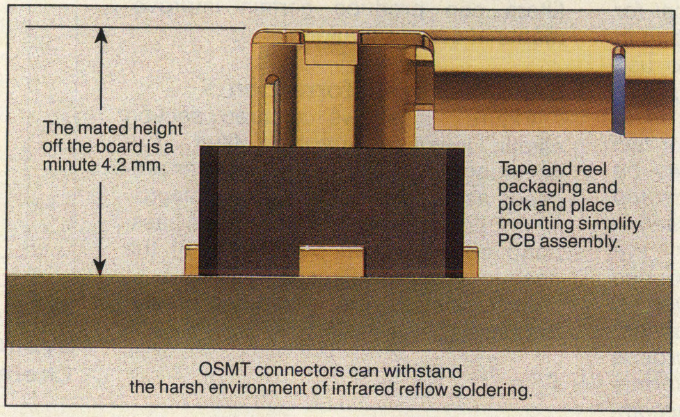

OSMT connectors can withstand the harsh environment of infrared reflow soldering.

Surface mount technology offers a number of advantages over traditional through-hole connectors. It reduces the required amount of PCB real estate, allowing for greater component density. The smaller, lighter connectors resist shock and vibration better. And surface mount technology gives OEMs the benefit of improving quality and producing a smaller end product.

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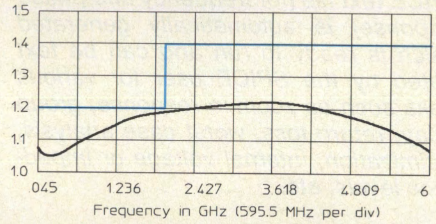

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The Elliptic Bandpass LC Filter

By William E. Sabin
Consulting Engineer

A PC program, written in Pascal for MS-DOS machines, will be described in this article. Its sole purpose is to use catalog listings (1,2,3) of lowpass prototype filters of orders 3, 5, 7, 9 or 11 to design LC wide-bandpass filters of two specific types. The response may be elliptic (ripples in passband and stop band) or maximally flat delay (Bessel) in the passband and ripples in the stopband (2). The designs produced have the following useful properties: 1) The topology can, if desired, absorb stray or pc-board capacitance to ground into the filter, 2) Impedance transformations within the filter are easily accomplished, 3) The designs can be rapidly fine-tuned via the interactive menu program design. In addition, a generic SPICE text file (for frequency and phase response) is automatically generated which is ready to run and can be text edited by the SPICE user for various tasks such as zoom-in response, group delay, return loss, worst case analysis, optimization, internal voltage or impedance levels, etc.

Figure 1 compares a seventh order Chebyshev bandpass filter which has 0.18 dB passband ripple with one particular seventh order elliptic filter which has the same passband ripple. The elliptic filter is characterized by its steeper transitions and by the stopband ripple which, on a dB scale, appears as response "loops". The magnitude of these loops, the steepness of the transitions, the locations of the stopband notches (transmission zeroes), the passband ripple and the input/output reflection coefficients are interrelated in complex ways which cannot be covered here but are discussed in the literature (1,2). The main point is that in many applications the steeper transitions, at the cost of stopband loops, is a desirable tradeoff. For example, in radio communications systems, if adjacent channel signals can be reduced several dB, substantial improvements in such things as desensitization, reciprocal noise mixing and intermodulation may be obtained, especially in collocated strong signal environments (4). In certain digital receiver topologies DSP aliasing problems can be improved considerably. Also, the cascading of low-order elliptic filters can produce certain improvements in some situations.

Bandpass Transformation

Figure 2a shows a three section elliptic lowpass prototype filter. The development of the bandpass filter proceeds according to the following steps:

1) Decide the upper and lower passband edges (ripple bandwidth) of the desired bandpass filter ($BW=F_{Hi}-F_{Lo}$)

2) Calculate the geometric center frequency:

$$F_o = \sqrt{F_{Hi}F_{Lo}}$$

3) Calculate Q_B:

$$Q_B = \frac{F_o}{BW}$$

4) Change the cutoff frequency of the lowpass prototype from 1.0 radian/sec to $1/Q_B$ radian/sec, as shown in Figure 2b ($C' = CQ_B$, $L'=LQ_B$).

5) Resonate each C and L at 1.0 radian per second in the manner shown in Figure 2c. Each L = 1/C and each C = 1/L. This is the bandpass prototype filter.

6) Transform the four components L_x, C_x, L_y, C_y, Figure 2c, into the configuration L'_x, C'_x, L'_y, C'_y as shown in Figure 2d, using the methods described in (1,2). Note that L'_x, C'_x is resonant at the upper notch frequency, 3.0460 radian/sec, and L'_y, C'_y resonate at the reciprocal of 3.0460, which is 0.3283 radian/sec. Note also that $C'_x = 1/L'_y$ and $C'_y = 1/L'_x$.

7) Optionally, install shunt capacitor C_z, and perhaps L_z, as suggested in Figure 2d. This procedure will be fully discussed in the next section.

8) Scale the filter to the desired high frequency band, Figure 2e, using the final value of the generator R_g:

$$L'' = \frac{L'R_g}{2\pi F_o}, \quad C'' = \frac{C'}{2\pi F_o R_g}$$

An examination of the final filter will often reveal some problems concerning unrealistic component values. For example, values of L may be too large or too small for the frequencies involved (the coils may have too much stray C or poor Q). Values of C may be too large

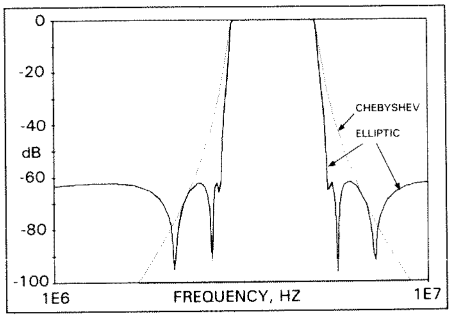

Figure 1. Seven pole elliptic bypass filter vs. seven pole Chebyshev.

(too much stray series L and R in the capacitor) or too small (comparable to stray C values in the coils and the layout). Bandwidth values which are too narrow or too wide may be impractical to realize using this "classical" design technique. The program allows us to evaluate rapidly any adjustments of the generator R_g, the bandwidth and the lowpass prototype to try to solve these problems. For example, increasing R_g makes the L values larger and the C values smaller. Fine tuning of the design can be easily done (5) to adjust for passband rounding due to moderate values of coil Q, as determined from SPICE.

Stray Capacitance

Figure 2d shows the optional shunt capacitor C_z. Its purpose is to solve a problem which is often (but not always) observed in the topology of that figure, especially at higher frequencies and at high values of R_g. The junction of the two tuned circuits L'_x, C'_x and L'_y, C'_y can be at a high impedance level at certain frequencies, and any stray C from that point to ground which is not included in the filter design might affect the response significantly. In higher-order filters the effect is often cumulative at the several junctions. A quick check, using a SPICE simulation with the stray C values added, will show the severity of the problem. One good solution is to use low C standoff insulators at these junctions. But in most layouts, especially a pc board layout, a better solution is to use a fixed capacitor which is several times larger than the capacitance of the circuit board pad, so that a predictable value of total C_z is always present. The other filter component values are then modified so that the desired response is *exactly* restored. The vehicle for modifying the filter is the Norton transformation. The use of this tool, however, introduces another complication into the design which the program deals with very effectively.

Norton Transformation

The modification of the filter to incorporate the additional shunt components proceeds from the generator end toward the load. The circuit L'_x, C'_x, Figure 2d, is repeated in Figure 3a. The Norton transformation of this circuit results in the pi network at the right in Figure 3a. There are three things to notice:

1) The reactances of L and C are multiplied by the factor n.

2) If $n > 1$ then the negative values of inductance and capacitance occur on the right hand side of the pi. If $n < 1$ then the negative values occur on the left side.

3) All reactive elements to the right of the pi, and also the load resistance, are modified by the factor n^2.

Figure 3b shows the complete transformation of the filter of Figure 2d. This is just the first segment of a more extensive filter. In the top part of Figure 3b the transformation is applied to both circuits (i.e., L_b, C_b and L_c, C_c) using the parameters n and m, respectively. The result is at the bottom of Figure 3b, where the transformed circuits are combined with the remaining shunt elements L_a, C_a and L_d', C_d'. The circuits in parallel are lumped together as indicated in the figure to get L1, C1, L0, C0 and L4, C4. Note carefully that *everything* to the right of the first transformation is modified by the factor n^2 and *everything* to the right of the second transformation is *further* modified by the factor m^2.

The next step is to repeat the modification procedure for the second pi section of the filter. In this case the first shunt section of the second pi is L4, C4 as obtained in the exercise just completed. These two components are then *further* modified during the modification of the second pi. The program keeps track of these values so that the "chain" transformation from one pi to the next is properly executed.

As each segment is transformed a cumulative factor, $(n1 \times n2 \times n3 \ldots \times m1 \times m2 \times m3 \ldots)^2$, is calculated. This then finally becomes the value of the filter output load resistance R_L (the generator R_g remains at 1.0). By manipulating this product, the load R_L can be adjusted up or down or kept at 1.0.

Choosing n and m

The program requires interactions by the user. You are the filter designer, not the computer. Initially, we consider a third order (N=3) filter, for simplicity. The following discussion presents the guidelines for selecting values of m and n, which determine C0 and L0 and also the value of the filter load resistance. Figure 4 lists the equations which are used to make the transformations which have been discussed. We wish now to enumerate the facts which these equations bring to light and which will be useful during the design process.

1) From equations 1, 2, 7 and 8 we see that some restrictions apply to m and n in order to prevent L1, C1, L4 and C4 from acquiring negative (unrealizable) values. The program lets you know if this has happened. In practice the values of m and n usually range from 0.9 to 1.1 and negative values of these parts are seldom encountered, except for certain prototypes which may be troublesome.

2) The products L2×C2 and L3×C3 do not change. These determine the stopband "notch" frequencies which must be kept constant. But the L/C ratios do change.

3) If the two terms in the denominator of equation 9 are made nearly equal (within the computer's ability) the value of L0 becomes so large that it can be neglected. In this case the shunt C0 stands alone and its value is given by equation 10.

4) If m = 1.0 and n = 1.0 then C0 = 0.0, L0 is infinite and the network remains unmodified. In the program, these are the default values.

5) If $m \neq 1.0$ then a value of C0 is

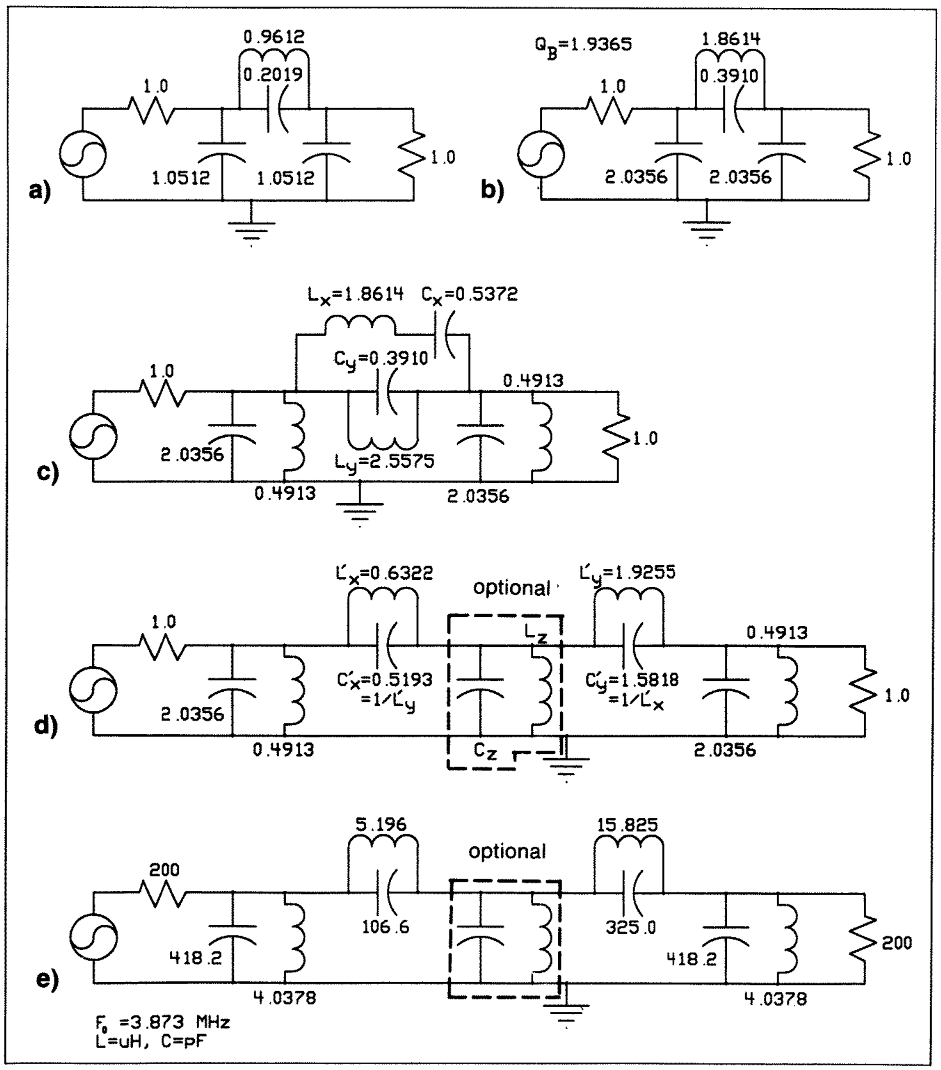

Figure 2. Development of HF elliptic BPF from lowpass prototype.

calculated. For a selected value of m, equation 11 gives that value of n which makes L0 infinite. The program provides this value of n by default. However, you can override the default by entering some other value of n, in which case a finite value of L0 is inserted. A special case is when n = 1/m. The product of n and m is then unity and no impedance transformation occurs. The load $R_L = 1.0$.

6) For a certain value of m, and for the corresponding value of n which makes L0 infinite, equation 10 shows that C0 can be negative, depending on the values of L_b and L_c. If this happens the program automatically interchanges L_b, C_b and L_c, C_c (this can be done with impunity) and recalculates the network. If we insert the value of n from equation 11 into equation 10 we can show that if (m < 1) and ($L_b > L_c$) or (m > 1) and ($L_c > L_b$) then $C0 \geqslant 0$.

7) We also note from equation 11 that if m > 1 then for infinite L0, n > 1 and vice versa.

Impedance Shifting

When a set of "C0s" (as in Figure 2e) are inserted, the various values of m and n can be manipulated to achieve an overall impedance change from input to output, either up or down. The change may be spread out among the various pi sections. Or, if we combine step up and step down the load R_L can be made equal to the generator R_g (in a filter of order 5, 7, 9 or 11). In a third order filter an L0, C0 combination can be used as described in part 5) of the previous section to get equal R_g and R_L. Quite often an impedance change is needed in the equipment design. Also, changing the internal impedance levels or adding "L0s" (as in Figure 2e) at certain locations within the filter may lead to more reasonable values of L and C.

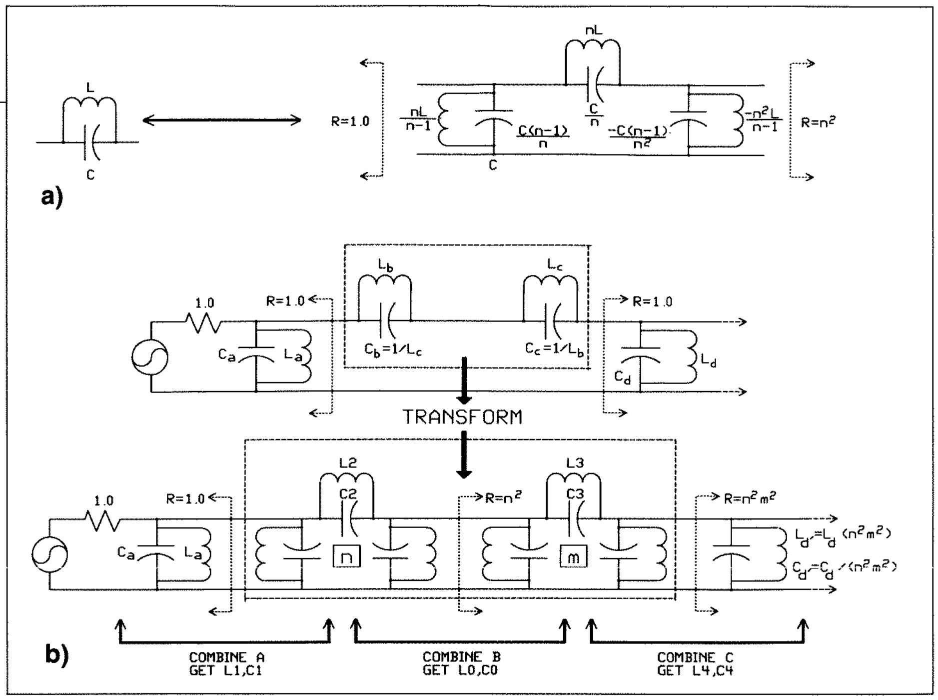

Figure 3. Norton transformation applied to bandpass filter.

Finally, recall that the familiar narrowband L and pi impedance transformation circuits, often used to match a generator or load resistance to a filter impedance, are seldom satisfactory in these filters. However, wideband transformers of one kind or another, optimized for the frequency range, can be used quite well.

High Frequency Filter

After preliminary definition of the prototype bandpass filter, the next step is to calculate the high frequency filter. The value of R_g is requested and the component values are then scaled to the final frequency range. At this point the efficacy of the design becomes apparent. Quite often, unrealistic values, as mentioned previously, are noticed. The value of R_g, the m and n values, and the bandwidth can be changed and the results quickly evaluated. The m and n values usually need to be experimentally tweaked until desirable values of the shunt capacitors (e.g., standard catalog values) are achieved. Another alternative is to start over with a different lowpass prototype which may be more realizable. It may be necessary to use cascaded filters, separated by amplifiers or other isolation devices. When the design is complete, request a SPICE file. At this time the coil Q value (assumed to be >50 or so and valid at the F_0 of the filter) is requested so that the true response of the filter will be calculated. Iteration between SPICE and the program leads to a final design. The generated SPICE file can be text-edited to get other information about the filter.

Physical Considerations

Some comments are in order about inductors. An inductor consists of a so-called true value of inductance in parallel with the coil self-capacitance. Because of self resonance effects, this combination produces a value of effective (or "apparent") inductance at the actual operating frequency. This value is what would be measured on an RF impedance bridge at the operating frequency and can be significantly higher than the true value. Inductor catalogs (also inductance bridges and Q meters) provide an inductance value which is measured at some low frequency where self capacitance effect are very small. This value is usually a close approximation to the true inductance at the filter operating frequency, but variations in core permeability with frequency (2) can sometimes cause significant errors in this assumption. In order to build the circuit properly, the true inductance (at the operating frequency) should be

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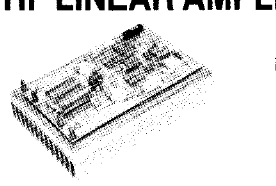

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$$1)\ L1 = \frac{L_a}{1 - \frac{L_a(1-n)}{n L_b}}$$

$$2)\ C1 = C_a - \frac{1-n}{n L_c}$$

$$3)\ L2 = n L_b$$

$$4)\ C2 = \frac{1}{n L_c}$$

$$5)\ L3 = m n^2 L_c$$

$$6)\ C3 = \frac{1}{m n^2 L_b}$$

$$7)\ L4 = \frac{m^2 n^2 L_d}{1 + \frac{L_d(1-m)}{L_c}}$$

$$8)\ C4 = \frac{1-m}{m^2 n^2 L_b} + \frac{1}{m^2 n^2 L_d}$$

$$9)\ L0 = \frac{m n^2}{\frac{(1-n)m}{L_b} - \frac{1-m}{L_c}}$$

$$10)\ C0 = \frac{1-n}{n^2 L_c} - \frac{1-m}{m n^2 L_b}$$

$$11)\ n = 1 - \frac{L_b(1-m)}{L_c m}\ \text{(for L0 = inf)}$$

$$12)\ L2\,C2 = L_b C_b\ ;\ L3\,C3 = L_c C_c$$

$$13)\ R1 = m^2n^2\ ;\ R_g = 1.0$$

Figure 4. Equations for filter modification.

equal to the value called for by the program. The self-capacitance C_s is part of the C value (across L) which the program calls for. Methods of determining the true L and stray C_s, near the operating frequency, are in the metrology literature (2,6).

An excellent way to tune a filter is to temporarily disconnect each parallel-tuned circuit from all the others and to tweak each one to its resonant frequency (provided by the program). I have used the very accurate frequency readout of a network analyzer for this purpose. A small "current sniffer" loop at the end of a coax, held near the coil, detects a very small "blip" of return loss (S11) at the resonant frequency. A fairly slow frequency sweep is needed.

My construction experience is that higher order filters having stopband return loops which can be more than 100 dB down can have "leak through" problems that degrade the stopband. It is best to build the filter in a straight line, input to output (not folded back on itself) and to use a completely enclosed and subdivided box. The pc-board ground plane should be solidly grounded to the bottom of the box. Use coax connectors to improve the ground path integrity. Also, the resistive terminations must be accurate. Small values of stray series L and shunt C in the generator and load can be absorbed into the filter, as required, perhaps using Smith® chart methods (7).

Program Details

The program ELIPBPF.EXE uses a single 80 x 25 text mode screen, VGA or EGA (either one preferred), CGA or Hercules on any MS/PC-DOS machine. The cursor (for data entry) is moved by using the four arrow keys. The various operations are invoked by using the function keys, F1. . .F9 and ESC. The various screen contents can be sent to the printer using the Shift-PrtSc keys. A complete circuit data summary can also be printed using F6. The input data for the lowpass prototypes is read from a text file prepared by the user and the program disk contains several examples. This disk also contains an .EXE file which calculates group delay, and an example is provided. The SPICE program creates the required frequency/phase file. A graph plotting capability is also required. The source code (public domain) is in Quick Pascal (virtually identical to Turbo 5.0) and is on the disk. The Procedure "Spice" (this one only, please!) can be customized by a Pascal programmer. A Touchstone or Compact subcircuit file is also a possibility.

The author is pleased to acknowledge the help of James C. Meier in testing this program, reading the text and offering several valuable suggestions.

This program is available on disk from the RF Design Software Service. See page 87 for ordering information. **RF**

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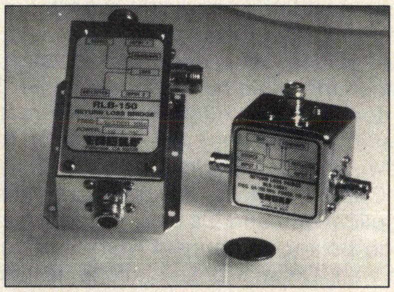
About the Author

William E. Sabin holds BSEE and MSEE degrees from the University of Iowa. He recently retired from Rockwell-Collins in Cedar Rapids, IA. He now performs freelance R&D projects in his personal lab and computer facility. He is co-editor (with E.O. Schoenike) of the book Single-Sideband Systems and Circuits, McGraw-Hill, N.Y., 1987. He can be reached at 1400 Harold Dr., S.E., Cedar Rapids, IA 52403-3735. Tel: (319) 364-8801.



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INFO/CARD 71

High Power Components — An Eye Towards the Future

By Liane G. Pomfret
Associate Editor

An increasing number of new applications and thriving traditional markets are forcing high power component manufacturers to improve their products and their marketing strategies. The market continues to do well despite the recession, with a steady flow of technological developments making incremental improvements in RF power products. Gerald Hiller, manager, semiconductor applications at M/A-COM comments, "There's a healthy mix of small and large orders as well as technologies."

It comes as little surprise that the list of technologies incorporating RF is very long and uses for high power components are no exception. Examples of unique applications using high power components include: wireless credit card verification systems, marine warning or signal buoys, over-the-horizon radar, pipe welding, medical hyperthermia, and sounding devices for mining or well-drilling. These do not always fall under the heading of traditional RF but it doesn't take much thought to realize just how well they are suited for high power devices and components.

Combine those with more traditional applications and it's easy to see why the high power component market is doing so well. Cellular base stations, avionics, satellite uplinks and downlinks and broadcast transmitters for television and radio all use high power components. Most broadcast transmitters still use high reliability, high power tubes because, until recently, reliable solid state components weren't available. For the moment, most solid state components capable of handling broadcast power requirements have a high price tag. But this is changing as more users demand the lower maintenance of solid state equipment. Solid state components capable of handling the necessary power and accompanying heat are difficult to design and manufacture. However, progress is being made and components for medium power applications are readily available.

For cellular communications applications, users are more concerned about lower loss characteristics and improved linearity. Better linearity means a greater ability to transmit multiple channels from the same cell site without interference or distortion. As with broadcast, improvements in performance tend to be relatively small, one percent at a time as opposed to ten or 20 percent leaps.

A related and upcoming technology is the area of personal communications networks or personal communications services (PCN/PCS). Most of the PCN/PCS technology will operate in the low gigahertz range and can build upon existing satellite and cellular technology. However, there will be new problems to deal with because of the combination of high frequency and high power. Here again, users will demand passive devices with low loss and active devices with high linearity. Virtually all of the manufacturers contacted for this article indicated that they are doing a great deal of work for the PCN/PCS market.

Medical applications are also growing. While relatively unheard of a few years ago, they now account for a small, but significant portion of the high power component market. Gary Moore, sales manager at Florida RF Labs, noted a marked increase in inquiries for medical applications just in the past year. Magnetic resonance imaging is the best known and fastest growing of these applications. Pulsed power devices are finding use in applications such as hyperthermia; heating the blood of leukemia patients to kill the cancer cells at a temperature which is not fatal to healthy cells. This is but one example of a medical application.

One thought to keep in mind when reading about all these applications is the definition of high power. For some, it would be anything above one watt while for others, anything less than a kilowatt wouldn't qualify. While there are certainly problems that arise in the kilowatt range that wouldn't in the watt range, high power components are still used for all high power applications. As Robert Portmess, vice president of manufacturing for Microwave Filter Company, points out, "We must design for optimum heat transfer and a minimum temperature creep while still providing RF integrity in relatively small mechanical packages." Engineers are working close to the physical limits so the rate of improvements has slowed considerably. The biggest problem has to do with heat transfer characteristics. For example, in a broadcast transmitter that uses tubes, it is relatively easy to used forced air or put a water jacket around that tube to remove the heat. However, when a solid state device is substituted for the tube, removing the heat from it becomes much more difficult. The inherent nature of the materials used in a solid state device, silicon, beryllia, and metal housings, make heat removal more of a problem. Add to this requests for smaller package sizes and the problem is multiplied.

In speaking with users and manufacturers of high power components, several brought up the issue of high reliability and ruggedization. Companies who used to do or still do a lot of military business, have found another plus to building to MIL-specifications. In the case of Microwave Technology, Al Rosenzweig points out that using their MIL-spec. standards when trying to sell a commercial part, does indeed help. Broadcast equipment must be extremely ruggedized. It must be able to stand up to extremes in temperature, and constant usage over a period of years. Users cannot afford to have equipment malfunction every six months or even once a year. As a result, standards are high. This carries over to cellular applications, satellite and even medical. Not only do repairs cost money, but downtime costs as well.

It could be said that high power components and technology is undergoing a quiet evolution. Advances are not happening in leaps and bounds but rather in small, individual steps. Unique applications and the more traditional applications are pushing the technology even further down the path. **RF**

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RF software

Analysis in the Modulation Domain

Modulation domain analysis is possible for less than $3000 with John Fluke's TimeView™ software package. Designed to be used with Philips' PM 6680 timer/counter, TimeView will measure frequency variations over time, perform statistics on sampled data and will FFT data in the modulation domain to yield a directly demodulated spectra. Cost of the software without the PM 6680 is $245 U.S. list price.
John Fluke Mfg. Co., Inc.
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Radar Cross Section Measurements

Flam & Russell, Inc. is now shipping its new FR959 Automated Radar Cross Section Measurement workstation. The FR959, which includes an Intel 80486 based computer and peripherals, controls both data acquisition and data analysis. Data analysis capabilities are independent of data acquisition and can be run on a different computer while the FR959 hardware is occupied with data gathering.
Flam & Russell, Inc.
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Microwave Power Device Characterization

Focus Microwaves' Power Data Manager is a data processing package able to retrieve power load pull data from a computer controlled microwave tuner. The Power Data Manager's analysis can obtain input/output power saturation characteristics, gain, power added efficiency and DC bias current at any load or source impedance seen by the device under test. Extension to Intermod (C/I) and Intercept (IP3) is possible on request. Price is $2,000 for the basic package, $3,000 including Intermod.
Focus Microwaves, Inc.
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Radar Development Tools

A new library of radar models lets users of Comdisco Systems' Signal Processing WorkSystem (SPW™) simulate, analyze and optimize the design of radar systems before committing to hardware. The Radar Library consists of six groups of model blocks: target and clutter processes, pulse compression waveforms, Doppler processing, automatic detectors, components and subsystems, and radar range equation scale factors.
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This Month's Program: RFD-0692
"FilDes: A Filter Design system for the RF Engineer" by Robert Lombardi. Calculates component values for various filter types in ladder network configuration. (Tubo Pascal, compiled and source code, requires CGA or better color graphics.)
"The Elliptic Bandpass LC Filter" by William Sabin. Uses lowpass prototype data to design wide bandpass elliptic filters, with topologies that allow strays to be incorporated into the design. (Quick Pascal, compiled and source code, CGA or better graphics — EGA/VGA preferred)
"Noise Parameter Calculation on the Personal Computer" by Malcolm Mayercik. Performs conversions between various noise source specifications. (needs BASIC or GWBASIC interpreter).

May's Program: RFD-0592
"QMIX — Mixer Analysis Program" by Kevin McClaning. Offers a graphical analysis of RF, LO, IF and image frequencies for selection of conversion scheme. Further analysis of spurious responses, including computation of total spurious energy is also included. (Pascal, compiled)

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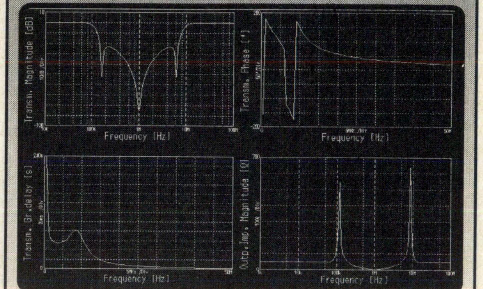

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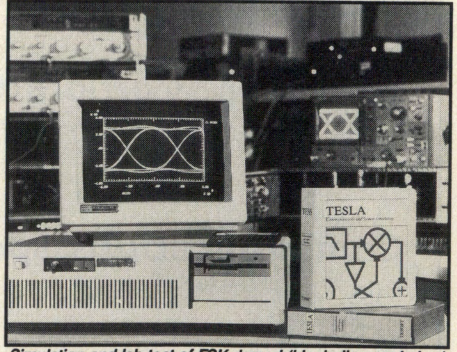

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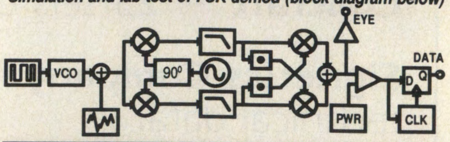

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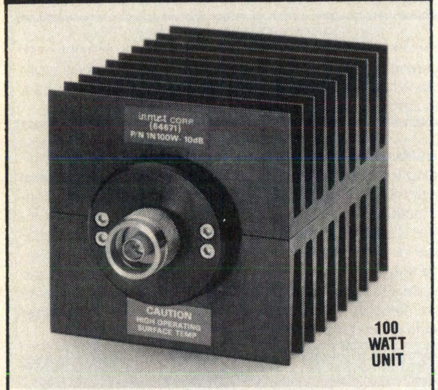

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RF literature

Electronic Materials Catalog

GE Components has released a two page catalog describing their electronic materials and components. The bulletin provides brief descriptions and applications information on tungsten and molybdemum wire, Dumet and Cumet glass sealing wire, encapsulation glass, and more.
GE Components Marketing & Sales Operation
INFO/CARD #210

Broadband IF Data Sheet

Six ultra-broadband IF components are introduced in a data sheet published by Narda. Detailed specifications are given for each model in the IF series including directional couplers, power dividers, and mixers.
Loral Microwave-Narda
INFO/CARD #209

Quartz Oscillator Brochure

Piezo Crystal Company announces the availability of a new product brochure covering quartz crystals and quartz crystal oscillators. The brochure includes Piezo's newest products including a low cost, high performance ovenized oscillator for test equipment.
Piezo Crystal Company
INFO/CARD #208

RF Amplifiers, Mixers and Dividers

Avianka's 16-page catalog contains information on a wide range of RF and Microwave drop-in and connectorized low noise, medium power and high intercept amplifiers. Also included are balanced mixers and a wide range of in-phase and 180 degree dividers.
Avianka, Inc.
INFO/CARD #207

Broadband RF Instrument Brochure

Holaday Industries, Inc. announces the publication of an all new brochure on its line of broadband RF field strength meters. Included is data on recently introduced instruments and probes to measure for compliance with the IEEE/ANSI C95.1-1992 EMF standard.
Holaday Industries, Inc.
INFO/CARD #206

Card Modular Instrument Catalog

The Tektronix Card Modular Instruments Information and Ordering Guide for 1992 is

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Eagleware Corporation — Grand Prize, PC Software Contest — '386 Personal Computer System loaded with Eagleware design and analysis software.

Communications Specialists, Inc. — 25 chip resistor and chip capacitor kits.

Coilcraft — 25 sets of chokes, inductors and EMI filter Designer's Kits.

Signetics Company — 25 NE605 FM IF demonstration boards and technical data.

See the Winning Designs in the July issue of *RF Design!*

now available. The catalog contains a number of new VXIbus products and some changes in the CDSbus product offering. All specifications needed to select card modular instruments are included.
Colorado Data Systems, Subsidiary of Tektronix
INFO/CARD #205

DSP Catalog

Array Microsystem's Spring 1992 short-form catalog describes their line of DSP ICs and board level components. Included is a description of the a66311 reconfigurable array store and the a664xx memory module family.
Array Microsystems, Inc.
INFO/CARD #204

RF Power Measurement Catalog

Bird Electronic Corp. has released its new 56 page, 50th Anniversary full-line catalog of RF power measuring equipment and accessories. High accuracy, peak reading, high and low power, multi-power, and low frequency wattmeters are described, as well as RF power calorimeters, RF loads and a wide range of accessories.
Bird Electronic Corporation
INFO/CARD #200

Power MOSFET Application Note

DEI's eight page application note on the DE-Series Fast Power MOSFET explains the electrical and thermal features of these power devices that enable them to switch 1 kV in less than 5 ns. Also included is a general discussion of high speed gate drive circuits for use with any power MOSFET.
Directed Energy, Inc.
INFO/CARD #203

Data Sheets for Conductive Ink

Electro-Science Laboratories has published data sheets for Au Metallo-organic thick film materials that can be applied using a gravure-offset printing process. The data sheets describe the material specifications which allow a minimum line width of less than 20 microns.
Electro-Science Laboratories
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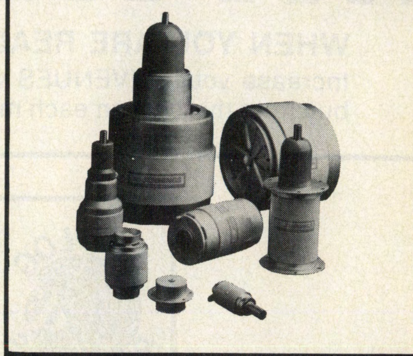

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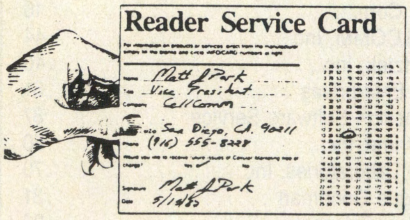

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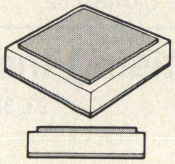
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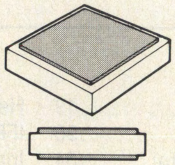
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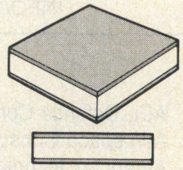
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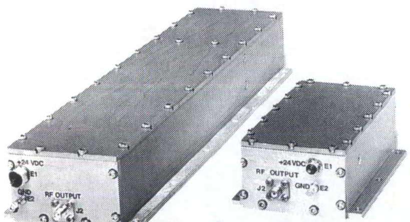

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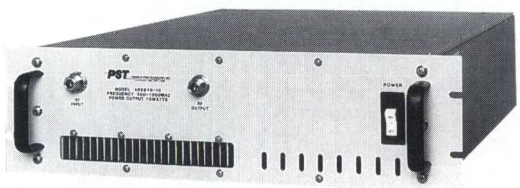

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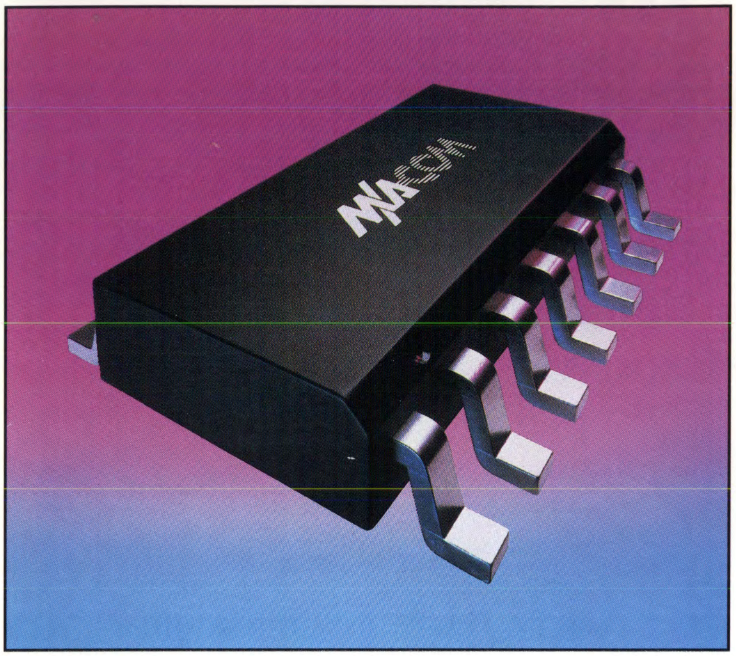

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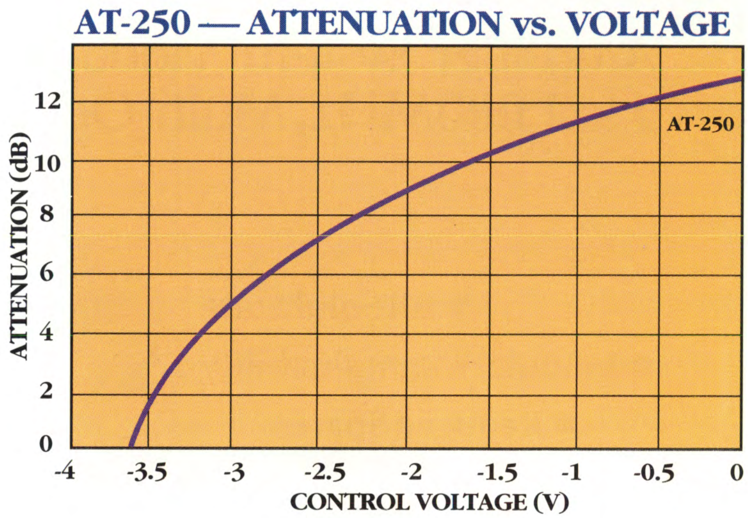

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AT-220	4bit, 2 dB step to 30 dB	1.8	1.5:1	45	SOIC, 16 lead
AT-230	3bit, 4 dB step to 28 dB	1.8	1.5:1	45	SOIC, 14 lead

DIGITAL /ANALOG ATTENUATORS

MODEL	DESCRIPTION	INSERTION LOSS (dB)	VSWR	Ip3 dBm	ANALOG ATT. LINEARITY ATT/CONTROL V.	PACKAGE
AT-240	3bit, 4 dB step to 28 dB + 12 dB VVA	4.5	1.8:1	35	±10%	SOIC, 20 lead

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MODEL	DESCRIPTION	INSERTION LOSS (dB)	ATTENUATION RANGE (dB)	VSWR	LINEARITY ATT/CONTROL V.	Ip3 dBm	PACKAGE
AT-250	Single Control Linear	2.8	13	1.5:1	±10%	35	SOIC, 8 lead
AT-309	Dual Control	1.1	20	1.5:1	N/A	15	SOIC, 8 lead
AT-339	Dual Control	1.1	40	1.5:1	N/A	15	SOIC, 8 lead
AT-635	Single Control Linear	6.0	30	2.0:1	±10%	33	SOIC, 20 lead

* - All parameters are typical specs @ 1.0 GHz.
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