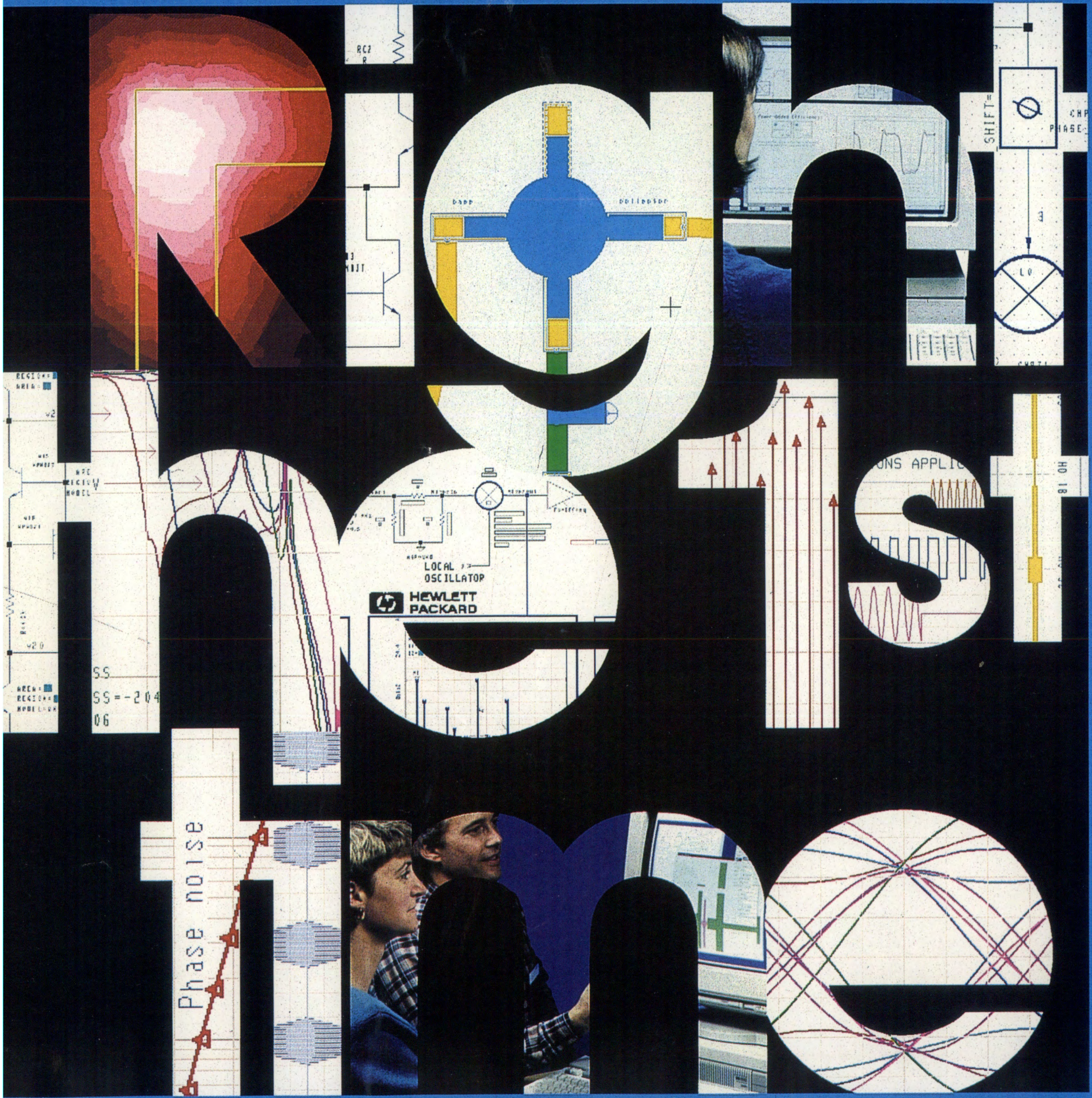
RF design

engineering principles and practices

September 1992



Official Show Issue
RF Expo East

Cover Story

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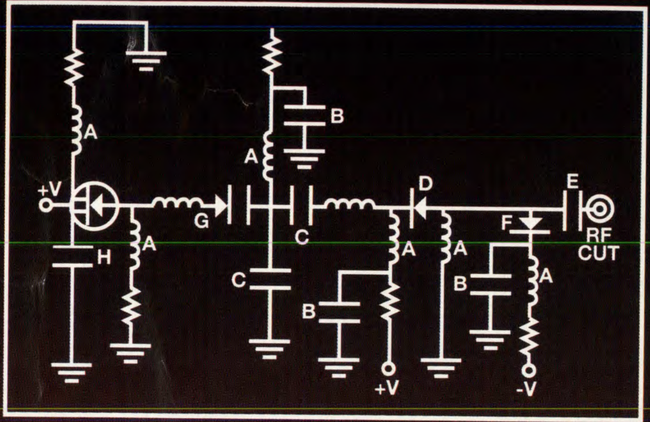

Typical switch with FSI semiconductors. (A) GC9002 Spiral Inductor; (B) GC84020 MNS Chip Capacitor; (C) GC86001 MNS Chip Capacitor; (D) GC4901 Beam Lead; (E) GC84030 MNS Chip Capacitor; (F) GC4221 PIN Diode; (G) GC15004 Linear Tuning Varactor; (H) GC84001 MNS Chip Capacitor.

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INFO/CARD 1

LORAL

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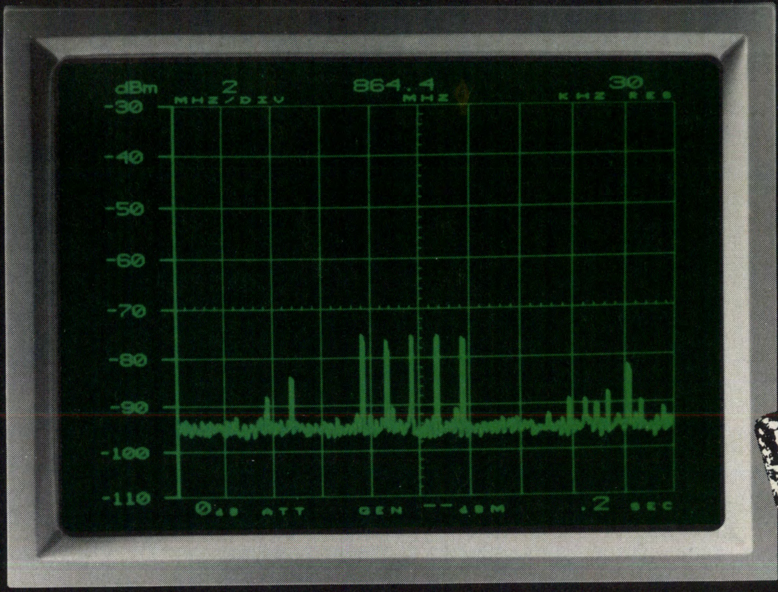

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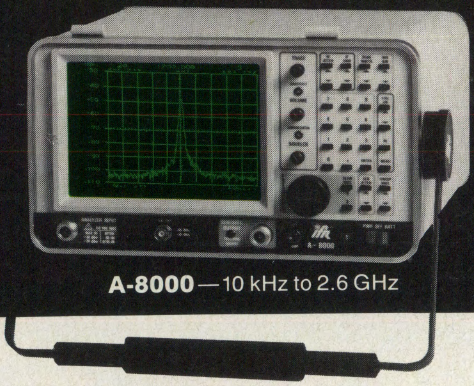
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You can't hear them or see them, but there are probably thousands of electromagnetic signals being transmitted through the air around you at this very moment.

Many of these signals originate from radio and television transmission. However, many other signals are emitted from computers, CRT display terminals, and other electrical devices.

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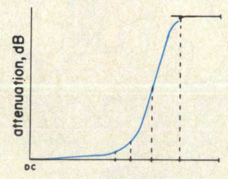
low pass, Plug-in, dc to 1200MHz

LOW PASS



Model No.	Passband MHz loss < 1dB	Stopband, MHz loss > 20dB	Stopband, MHz loss > 40dB
PLP-5	DC-5	8-10	10-200
PLP-10.7	DC-11	19-24	24-200
PLP-21.4	DC-22	32-41	41-200
PLP-30	DC-32	47-61	61-200
PLP-50	DC-48	70-90	90-200
PLP-70	DC-60	90-117	117-300
PLP-90	DC-81	121-137	167-400
PLP-100	DC-98	146-189	189-400
PLP-150	DC-140	210-300	300-600
PLP-200	DC-190	290-390	390-800

Model No.	Passband MHz loss < 1dB	Stopband, MHz loss > 20dB	Stopband, MHz loss > 40dB
PLP-250	DC-225	320-400	400-1200
PLP-300	DC-270	410-550	550-1200
PLP-450	DC-400	580-750	750-1800
PLP-550	DC-520	750-920	920-2000
PLP-600	DC-680	840-1120	1120-2000
PLP-750	DC-700	1000-1300	1300-2000
PLP-800	DC-720	1080-1400	1400-2000
PLP-850	DC-760	1100-1400	1400-2000
PLP-1000	DC-900	1340-1750	1750-2000
PLP-1200	DC-1000	1620-2100	2100-2500

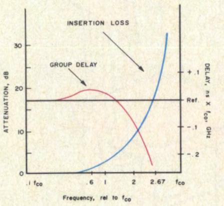
Price, (1-9 qty), all models: plug-in $14.95, BNC $32.95, SMA $34.95, Type N $35.95

Surface-mount, dc to 570MHz

SCLF-21.4	DC-22	32-41	41-200
SCLF-30	DC-30	47-61	61-200
SCLF-45	DC-45	70-90	90-200
SCLF-135	DC-135	210-300	300-600
SCLF-190	DC-190	290-390	390-800
SCLF-380	DC-380	580-750	750-1800
SCLF-420	DC-420	750-920	920-2000

Price, (1-9 qty), all models: $11.45

Flat Time Delay, dc to 1870MHz

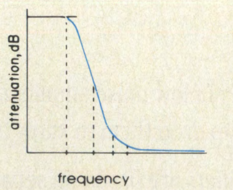


Model No.	Passband MHz loss < 1.2dB	Stopband MHz loss > 10dB	Stopband MHz loss > 20dB	VSWR Freq. Range, DC thru 0.2fco X̄	VSWR Freq. Range, DC thru 0.6fco X̄	Group Delay Variations, ns Freq. Range, DC thru fco X̄	2fco X̄	2.67fco X̄
PBLP-39	DC-23	78-117	117	1.3:1	2.3:1	0.7	4.0	5.0
PBLP-117	DC-65	234-312	312	1.3:1	2.4:1	0.35	1.4	1.9
PBLP-156	DC-94	312-416	416	0.3:1	1.1:1	0.3	1.1	1.5
PBLP-200	DC-120	400-534	534	1.6:1	1.9:1	0.4	1.3	1.6
PBLP-300	DC-180	600-801	801	1.25:1	2.2:1	0.2	0.6	0.8
PBLP-467	DC-280	934-1246	1246	1.25:1	2.2:1	0.15	0.4	0.55
▲BLP-933	DC-560	1866-2490	2490	1.3:1	2.2:1	0.09	0.2	0.28
▲BLP-1870	DC-850	3740-6000	5000	1.45:1	2.9:1	0.05	0.1	0.15

Price, (1-9 qty), all models: plug-in $19.95, BNC $36.95, SMA $38.95, Type N $39.95
NOTE: ▲: -933 and -1870 only with connectors, at additional $2 above other connector models.

high pass, Plug-in, 27.5 to 2200MHz

HIGH PASS



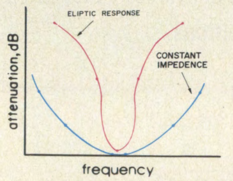
Model No.	Stopband MHz loss < 40dB	Stopband MHz loss < 20dB	Passband MHz loss < 1dB	VSWR Pass-band Typ.
PHP-25	DC-13	13-19	27.5-200	1.8:1
PHP-50	DC-20	20-26	41-200	1.5:1
PHP-100	DC-40	40-55	90-400	1.8:1
PHP-150	DC-70	70-95	133-600	1.8:1
PHP-175	DC-70	70-105	160-800	1.5:1
PHP-200	DC-90	90-116	185-800	1.6:1
PHP-250	DC-100	100-150	225-1200	1.3:1
PHP-300	DC-145	145-170	290-1200	1.7:1

Model No.	Stopband MHz loss < 40dB	Stopband MHz loss < 20dB	Passband MHz loss < 1dB	VSWR Pass-band Typ.
PHP-400	DC-210	210-290	395-1600	1.7:1
PHP-500	DC-280	280-365	500-1600	1.8:1
PHP-600	DC-350	350-440	600-1600	2.0:1
PHP-700	DC-400	400-520	700-1800	1.6:1
PHP-800	DC-445	445-570	780-2000	2.1:1
PHP-900	DC-520	520-660	910-2100	1.8:1
PHP-1000	DC-550	550-720	1000-2200	1.9:1

Price, (1-9 qty), all models: plug-in $14.95, BNC $36.95, SMA $38.95, Type N $39.95

bandpass, Elliptic Response, 10.7 to 70MHz

BANDPASS



Model No.	Center Freq. (MHz)	Passband I.L. 1.5 dB Max. (MHz)	3 dB Bandwidth Typ. (MHz)	Stopbands I.L. > 20dB at MHz	Stopbands I.L. > 35dB at MHz
PBP-10.7	10.7	9.6-11.5	8.9-12.7	7.5 & 15	0.6 & 50-1000
PBP-21.4	21.4	19.2-23.6	17.9-25.3	15.5 & 29	3.0 & 80-1000
PBP-30	30.0	27.0-33.0	25-35	22 & 40	3.2 & 99-1000
PBP-60	60.0	55.0-67.0	49.5-70.5	44 & 79	4.6 & 190-1000
PBP-70	70.0	63.0-77.0	68.0-82.0	51 & 94	6.0 & 193-1000

Price, (1-9 qty), all models: plug-in $18.95, BNC $40.95, SMA $42.95, Type N $43.95

Constant Impedance, 21.4 to 70MHz

Model No.	Center Freq. MHz	Passband MHz loss < 1dB	Stopband loss > 20dB at MHz	VSWR 1.3:1 Total Band MHz
PIF-21.4	21.4	18-25	1.3 & 150	DC-220
PIF-30	30	25-35	1.9 & 210	DC-330
PIF-40	42	35-49	2.6 & 300	DC-400
PIF-50	50	41-58	3.1 & 350	DC-440
PIF-60	60	50-70	3.8 & 400	DC-500
PIF-70	70	58-82	4.4 & 490	DC-550

Price, (1-9 qty), all models: plug-in $14.95, BNC $36.95, SMA $38.95, Type N $39.95

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Freq./ Min.	GHz Max.	[1] Noise Fig/dB	[2] Gain dB/Min.	[2] Flatness ± dB/Max.	[2] Power 1 db GCP dBm/Min.	DAICO Part No.
2.20	2.40	1.8	27.5	0.50	10	DAML6074
2.20	2.40	1.8	37.0	0.50	15	DAML6075
2.70	3.10	2.0	37.0	0.50	15	DAML6078
4.40	5.00	2.5	37.0	0.50	18	DAML6083
5.40	5.90	1.8	27.5	0.50	10	DAML6019
2.00	6.00	3.0	37.5	1.00	15	DAML6020
4.00	6.00	2.0	30.5	1.00	15	DAML6022
5.90	6.40	2.0	27.5	0.50	10	DAML6084
7.25	7.75	2.0	27.5	0.50	10	DAML6088
8.00	8.40	1.7	27.5	0.50	10	DAML6024
9.00	10.00	2.0	27.5	0.50	10	DAML6090

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[2] guaranteed -55/+85°C

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INFO/CARD 4

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RF design

contents

September 1992

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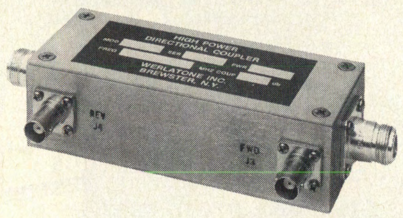

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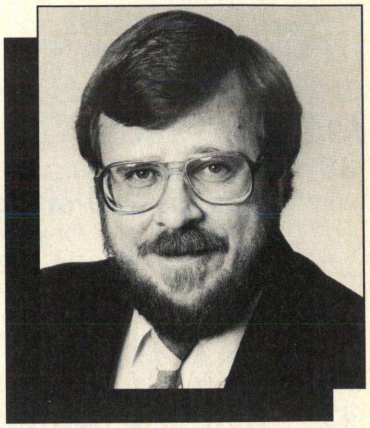
RF editorial

It's That Forest and Trees Problem Again



By Gary A. Breed
Editor

Too often, we find ourselves using catch-all terms as if they were precise descriptions. For example, the "commercial market" is being touted as the way to prosper in the face of a slowing "military market." Of course, we all know that there is no single commercial or military market; there are many different markets with distinctly different needs.

Yet we hear these broad generalities every day. It worries me that so many good companies can't see the forest (the "commercial market"), even though the "trees" of individual product developments are all around. Or, they see a slower "military market" without realizing how much funding remains for avionics, SDI development, and communications systems.

Perhaps the biggest over-generalization is the term "wireless." (I've been guilty of this myself.) Wireless data communications, wireless personal communications, wireless control systems, wireless computer peripherals, wireless alarm systems, wireless factory communications — they are all different markets! What looked like one big family of applications a few years ago has developed into a diverse collection of specific markets. Lumping them together as "wireless" also steers our thinking away from other applications like GPS, HDTV, digital audio broadcasting (DAB), or collision-avoidance systems.

Getting to the Point

When I am asked about the condition of the RF industry, my current answer is, "Some companies are suffering, while others can't keep up with the orders." A too-general view of the marketplace may be part of the difference between them.

Many companies that are doing well have specifically targeted individual applications like cellular equipment, cordless telephones or automotive electronics. Others have simply responded to every opportunity, large or small, with the flexibility and creativity to design and deliver products to customer specs. Rather than being concerned about the "commercial market," they find themselves in the middle of it by dealing with specific applications, one at a time.

Companies that are not doing so well might benefit from an assessment of their approach. Do basic engineering and manufacturing capabilities allow response to varied customer needs, or is everything done the same way it was five years ago? Are sights set too high by going after a few big contracts while smaller orders are discouraged? Is military business ignored, even if it is significant, or growing? Is management wringing its hands with worry, or is it beating the bushes for every market opportunity? At the most fundamental level, are company goals well-defined or nebulous?

Just as there is no single market that will save troubled RF companies, there is no single key to success or failure. These observations are only a small part of the answer, and the discussion needs to continue. There's a lot of business available for RF companies — we want you to get your share.

INFO/CARD 5

1990s Technology

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Noise Com's UFX 7000 Series noise-generating instruments are equipped with new user-friendly dedicated keys. A bright 4 x 20-character LCD display provides easy-to-read information about the instrument's status and operating parameters. The UFX 7000 Series makes testing of BER, S/N, E_b/N_o, CATV, image and spurious rejection, NPR, EMC, and intermodulation easier than ever before.

The UFX 7000 has improved noise output flatness and adds extended programmability such as unattended execution of user-created routines and simplified operation. An IEEE-488 interface is standard, RS-232C and others are optional.

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UFX 7102	10 Hz - 100 kHz	+13	-37	±0.75	3162
UFX 7103	10 Hz - 500 kHz	+13	-44	±0.75	1414
UFX 7104	10 Hz - 1 MHz	+13	-47	±0.75	1000
UFX 7105	10 Hz - 10 MHz	+13	-57	±0.75	316
UFX 7106	100 Hz - 30 MHz	+13	-62	±0.75	183
UFX 7107	100 Hz - 100 MHz	+13	-67	±0.75	100
UFX 7108	100 Hz - 500 MHz	+10	-77	±1.0	31.6
UFX 7109	100 HZ - 1000 MHz	+10	-80	±1.5	22.4
UFX 7110	100 HZ - 1500 MHz	+10	-82	±1.5	18.2
UFX 7111	1 GHz - 2 GHz	+10	-80	±1.5	22.4
UFX 7112	1 MHz - 2 GHz	+0	-93	±2.0	5.0
UFX 7124	2 GHz - 4 GHz	-10	-103	±2.0	1.58
UFX 7218	2 GHz - 18 GHz	-20	-122	±2.0	0.18
UFX 7240	2 GHz - 40 GHz	-20	-126	±4.0	0.11

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INFO/CARD 6

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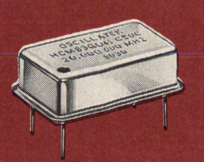

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Voh: Vcc-0.2 V, MIN.
Vol: 0.2 V, MAX.
Control Voltage: 0.5 VDC to 4.5 VDC
TRANSFER SLOPE: Positive

TEMP. RANGE	STABILITY	DEVIATION*
0° to 50°C	±.0025%	±50 ppM
0° to 70°C	±.005%	±100 ppM
-20° to 70°C	±.01%	±150 ppM
-45° to 85°C	*Deviation over full 0.5 to 4.5 V control voltage range.	
Deviation per volt = ± deviation / 2 V		

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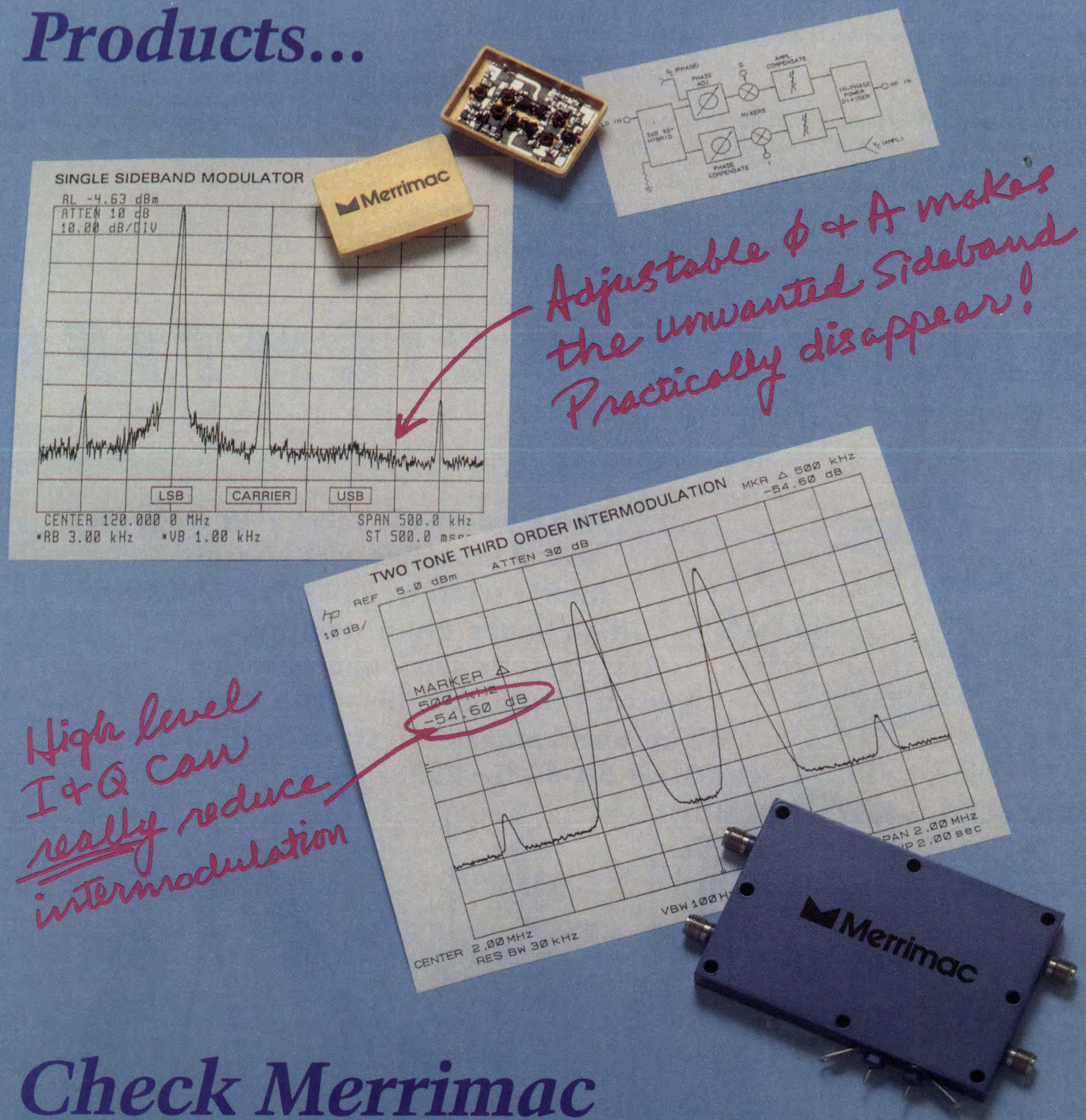

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Letters should be addressed to: Editor, *RF Design*, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111. Published letters may be edited for length or clarity.

Phase Shifters

Editor:

I have been experimenting with 90 degree lattice phase shifters (*RF Design*, April 1989) and have found that the highest frequency element can be replaced by a transmission line with a time delay of 2/p seconds, where p is the value of the appropriate root. For a phase shifter whose outputs are connected by coax to the rest of the circuitry, this gives almost an extra order of complexity for free. It also provides a useful substitute where the lattice components are small.

The transmission line gives only a linear approximation to the lattice phase response. But, at the frequency where the lattice provides a phase shift of 45 degrees, the error of the transmission line approximation is only 2.5 degrees. At this frequency, the linear approximation of phase shift per octave in the other arm of the phase shifter is beginning to run out.

The difficulty with this approach is that it upsets the mathematics. This means that, to achieve equal phase difference ripple, the values have to be "adjusted" slightly using a circuit simulator. I should be interested to hear from anyone who has the exact solution, or is more nimble at negotiating mathematical morasses than I.

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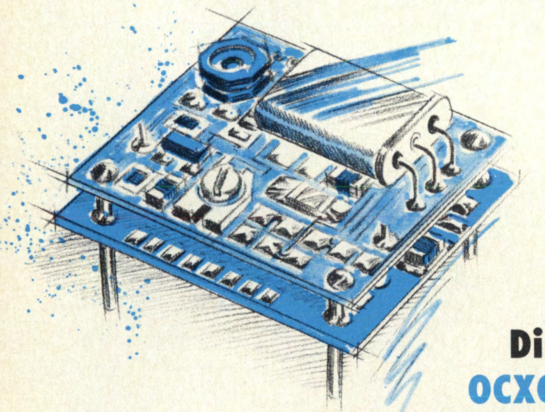

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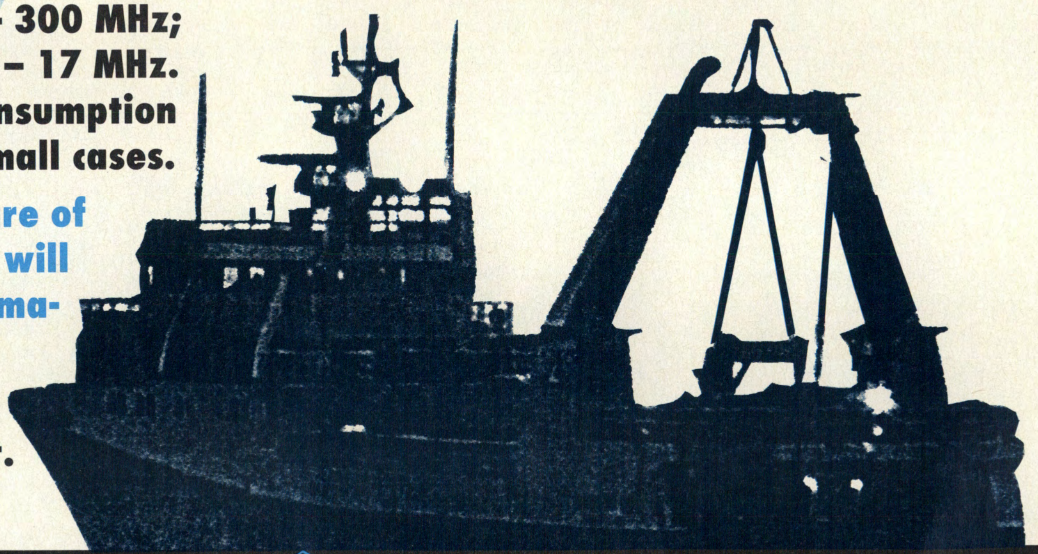

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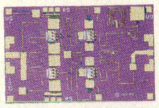
Military

- ☑ MMIC/Foundry and Discrete devices
- ☑ AMPLIFIERS
- ☑ OSCILLATORS
- ☑ SUBSYSTEMS



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- Custom Chip Design Available
- Litton also processes OEIC, CCD, and mixed-signal function chips, **FETs, MMICs and Diodes, for Commercial, Military, and Hi-rel applications.**

Commercial

- ☑ SEMICONDUCTOR DEVICES (Diodes, FETs and MMICs)
- ☑ OSCILLATORS
- ☑ AMPLIFIERS
- ☑ LINEARIZERS

SEMICONDUCTOR DEVICES (Diodes, FETs and MMICs) Litton is recognized as the world leader in GaAs IMPATT diodes, both GaAs and InP GUNN diodes and semiconductor devices for microwave and millimeter wave applications.

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- For sources and power amplifiers
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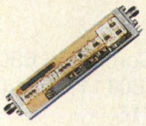
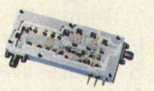


AMPLIFIERS Amplifiers providing state-of-the-art performance with a full spectrum of requirements from low noise through power.

- **Leader in MMIC insertion.**
- Frequency 1 to 100 GHz
- Power output up to three watts
- Narrow band through multi-octave frequency bands
- Single amplifier covers 2 to 18 GHz
- Standard catalog amplifiers available from stock

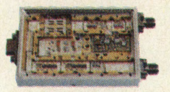
Designs available using FETs, MMICs, and Diode Reflection techniques to customer specific requirements.



OSCILLATORS A broad capability for oscillators in small packaging, and with multiple outputs.

- FET, Bipolar, and GUNN Diode designs.
- DRO, VCO, PLO and Cavity
- Frequency coverage 3 to 140 GHz
- Low Phase Noise
- Wide tuning range available
- Mechanical and Electrical tuning options
- Low DC Power consumption

Designs are available to meet customer's form, fit and function requirements, including Hi-Rel.



SUBSYSTEMS Unique solutions are available to integrate multiple functions into a very small unit/subsystem or **"super component"** to solve specific system requirements.

- Frequencies 1 to 100 GHz
- Integration of RF amplifiers, mixers, oscillators, switches, filters, modulators, and threshold detector functions
- Incorporate all analog and digital controls along with DC power conditioning.
- Full thermal and packaging design capabilities.

A broad range of products including frequency Up/Down Converters, T/R Modules, Multi-oscillator sub-systems, Serrodyne Repeater Amplifiers, 8-channel gain matched frequency block converters and TWTA replacements.

Litton NSL Products

NSL-1050 Series Modular Digital IFM Receivers

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- 80 dBm Dynamic Range
- −70 dBm Sensitivity
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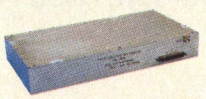
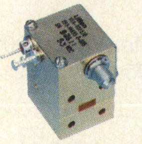
- Weight: 12.5 lbs.



NSL-1070 Digital Frequency Discriminator

- 2–18 GHz Frequency Band
- 5 MHz RMS Frequency Accuracy
- 65 dBm Dynamic Range
- −55 dBm Sensitivity
- Weight: 10.5 lbs.

A Broad Range of applications for Airborne and Shipboard ESM/ECM systems.



OSCILLATORS Proven expertise in oscillator design for high volume, low cost commercial applications featuring:

- Standard bands, 10 GHz, 18 GHz, 23 GHz, 26/28 GHz, 31 GHz, 38 GHz, 57/58 GHz, 63 GHz, 77 GHz, 94 GHz
- Low Phase Noise
- Low DC power consumption
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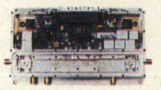
Commercial applications include Traffic control, Wireless LAN, **AUTO-COLLISION AVOIDANCE,** Industrial controls, Tagging, Point-to-point communication, **PCN INFRASTRUCTURE.**



AMPLIFIERS Recognized leader in SATCOM up-link/down-link amplifier drivers for TWTAs and KPAs.

- Frequencies—standard C, X, and Ku, MMW bands, domestic and international
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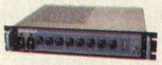

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RF calendar

September

15-17 **The 14th Annual Piezoelectric Devices Conference and Exhibition**
Kansas City Westin Crown Center
Information: Peter J. Walsh, Staff Vice President, Components Group, Electronic Industries Association, 2001 Pennsylvania Avenue, N.W., Washington, DC 20006. Tel: (202) 457-4932.

22-24 **RF Expo East**
Tampa, FL
Information: Barb Binge, Cardiff Publishing Company, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111. Tel: (303) 220-0600, (800) 525-9154. Fax: (303) 773-9716.

29-30 **International Analog Applications Conference & Exhibit**
Santa Clara, CA
Information: CMP Conference & Exhibit Group. Tel: (800) 972-5244, (516) 562-5717. Fax: (516) 562-7272.

29-2 **1st International Conference on Universal Personal Communications**
Dallas, TX
Information: ICUPC '92 Registration, c/o Bob McFadden, NEC America, Inc., 1525 Walnut Hill Lane, Irving, TX 75038. Tel: (214) 518-5341. Fax: (214) 518-5160.

October

4-7 **1992 IEEE GaAs IC Symposium**
Miami Beach, FL
Information: Courtesy Associates, Cathy Coyle, Suite 300, 655 Fifteenth Street, N.W., Washington, DC 20005. Tel: (202) 347-5900. Fax: (202) 347-6109.

13-15 **MM 92: The Civil and Military Applications of Microwave Technology**
Brighton, UK
Information: Microwave Exhibitions and Publishers, 90 Calverley Road, Tunbridge Wells, Kent TN1 2UN, UK. Tel: 44 (0) 892 544027. Fax: 44 (0) 892 541023.

15-19 **7th International Audio, Video, Broadcasting and Telecommunications Show**
Milan, Italy
Information: IBTS, Via Domenichino, 11 (C.P. 15117 - 20150 Milano), 20149 Milano, Italy. Tel: (02) 4815541. Fax: (02) 4980330.

19-21 **The Third International Symposium on Personal Indoor and Mobile Radio Communications**
Boston, MA
Information: Technical Program Chairman, Dr. K. Pahlavan, Electrical Eng. Dept., Worcester Polytechnic Institute, Worcester, MA 01609. Tel: (508) 831-5634. Fax: (508) 831-5491.

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RF courses

Radar Cross Section Reduction
October 13-16, 1992, Atlanta, GA
Infrared/Visible Signature Suppression
October 27-30, 1992, Atlanta, GA
Principles of Modern Radar
November 2-6, 1992, Atlanta, GA
Information: Georgia Institute of Technology, Continuing Education. Tel: (404) 894-2547.

Linear & Nonlinear System Analysis and Identification
September 15-17, 1992, Washington, DC
Modern Microwave Techniques
September 21-24, 1992, Del Mar, CA
Navstar/GPS: Design and Applications
October 14-16, 1992, Washington, DC
Information: University Consortium for Continuing Education. Tel: (818) 995-6335. Fax: (818) 995-2932.

Personal Communications Networks
October 14-16, 1992, Los Angeles, CA
RF and Microwave Circuit Design I
November 16-20, 1992, Los Angeles, CA
Advanced Communication Systems Using Digital Signal Processing
November 16-20, 1992. Los Angeles, CA
Information: UCLA Short Course Program Office. Tel: (310) 825- 1047. Fax: (310) 206-2815.

Digital Signal Processing: Principles, Devices and Applications
September 27-October 2, 1992, Leicester, UK
Information: IEE, Savoy Place, London WC2R 0BL, United Kingdom.

Grounding, Bonding, Shielding and Transient Protection
October 27-30, 1992, Orlando, FL
Analyzing Communications System Performance
October 19-21, 1992, San Diego, CA
November 2-5, 1992, Vienna, Austria
Cellular Radio Telephone Systems
September 21-23, 1992, Washington, DC
Modern Receiver Design
September 14-18, 1992, Washington, DC
October 26-30, 1992, Amsterdam, Netherlands
Personal Communications Systems and Networks (PCS and PCN): A Telecommunications Revolution
October 7-9, 1992, Washington, DC
Mobile Cellular Telecommunications Systems
October 14-16, 1992, Washington, DC
November 16-18, 1992, San Diego, CA
Microwave High Power Tubes and Transmitters
October 19-23, 1992, Washington, DC
Principles of High Frequency Radio Communications
October 20-23, 1992, Washington, DC
Modern Digital Modulation Techniques
October 26-29, 1992, Washington, DC
Satellite Communications Engineering Principles
November 4-6, 1992, Washington, DC
Information: The George Washington University, Continuing Engineering Education, Merril A. Ferber. Tel: (202) 994-8522 or (800) 424-9773.

Spectrum Measurements and Analysis
October 19-22, 1992, Lexington Park, MD
Information: U.S. Department of Commerce, NTIA/ITS.S2, Frank H. Sanders. Tel: (303) 497-5727.

Digital Signal Processing in Modern Communication Systems
September 14-18, 1992, Davos, Switzerland
Modern Digital Modulation Techniques
September 21-24, 1992, Davos, Switzerland
Personal Wireless Communications: Cellular Telephony, Portable Computing, and Broadband Wireless Networks
September 21-25, 1992, Davos, Switzerland
Modern Microwave Techniques: Measurements, Signal and Network Analysis, Microwave Products and Systems Characterization
October 19-23, 1992, Madrid, Spain
RF and Microwave Circuit Design: Linear and Non-Linear
October 19-23, 1992, Madrid, Spain
RF and Microwave Component Modeling
October 21-23, 1992, Davos, Switzerland
Broadband Telecommunication Networks: MAN, ATM, B-ISDN, Self- Routing Switches, and Optical Networks for Voice/Data/Image/HDTV
October 26-30, 1992, Madrid, Spain
Fast Algorithms for Adaptive Signal Processing
November 2-5, 1992, United Kingdom
Information: CEI-Europe/Elsevier, Mrs. Tina Persson. Tel: (46) 122-175-70. Fax: (46) 122-143-47.

EMC Foundation Course
October 21, 1992, Harrogate, Yorkshire, UK
November 4, 1992, Guildford, Surrey, UK
Information: Surrey Conferences. Tel: (44) 0784 461393.

Electronic Design Techniques and Analysis Required to Meet Electromagnetic Compatibility Requirements
September 23-24, 1992, Novi, MI
Advanced EMC Printed Circuit Board Design
September 25, 1992, Novi, MI
Information: JASTECH. Tel: (313) 553-4734.

Physical Security Standards for Sensitive Compartmented Information Facilities (SCIF)
September 18, 1992, San Diego, CA
Electromagnetic Test Facilities: Design Principles, Applications and Examples
September 21-23, 1992, San Diego, CA
High Power Microwaves: Sources, Systems and Effects
October 19-21, 1992, San Diego, CA
Information: Praxis International. Tel: (215) 524-0304.

EC/Design Seminar Series
September 29-30, 1992, San Jose, CA
Basic HIRF Seminar
October 13-15, 1992, Mariposa, CA
EMCad1™ Computer Analysis Workshop
October 16, 1992, Mariposa, CA
October 30, 1992, Mariposa, CA
Advanced HIRF Design
October 27-30, 1992, Mariposa, CA
Advanced HIRF Testing Seminar
November 17-20, 1992, Mariposa, CA
Information: CKC Laboratories. Tel: (209) 966-5240. Fax: (209) 742-6133.

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Hong Kong (852) 848-7070 • Korea (82-2) 784-2666

INFO/CARD 15

FCC Formally Proposes Spectrum for PCS Applications

On July 16 the Federal Communications Commission formally proposed a plan for spectrum allocation, regulation and licensing for Personal Communications Services (PCS). The FCC has proposed a broad definition of PCS to include services such as advanced paging, mobile and portable telephone services, wireless facsimile machines, wireless electronic mail services and more. In order to include space for all these technologies, the commission has proposed allocating space in two portions of the spectrum. The first is for narrowband service in the 900 MHz band for applications such as the advanced pagers. The second band would be in the 2 GHz area but approval is contingent on completion of Docket ET 92-2, which will allocate spectrum for emerging technologies. Protests have already been lodged against the reallocation of space in the 2 GHz range by major utilities and railroads that currently use the band.

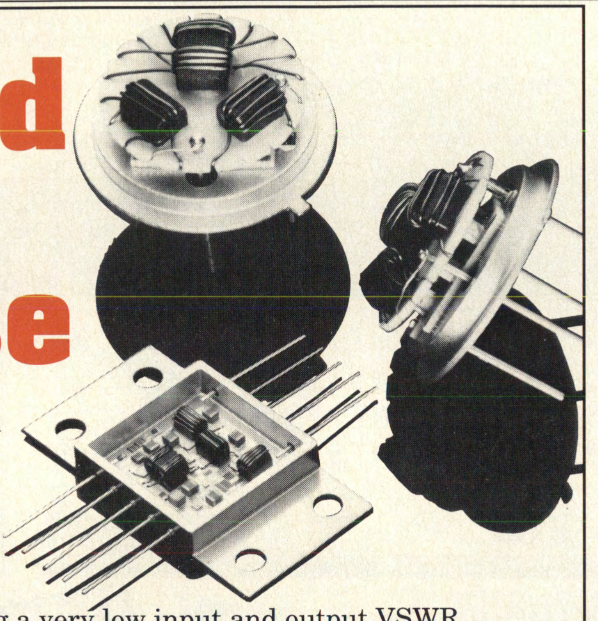
Licensing of space is also undergoing close scrutiny. According to a statement by Commissioner Sherrie P. Marshall, "I hope Congress will move swiftly to give the FCC Auction authority for PCS spectrum. Our experience in the cellular arena confirms that, despite our best efforts to prevent speculators from "gaming" the process, abuses will occur." Comments have been requested for possible reforms of the lottery process and possible competitive bidding rules. Several options for licenses have been offered by the FCC.

In addition, the FCC has proposed that space be allocated in the 2 GHz band for Part 15 compliant PCS devices. No indication was given as to whether this would be to support the proposed wireless computer services now under development by Apple and other PC companies.

The FCC also tentatively awarded a Pioneer's Preference in the 900 MHz narrowband PCS service to Mobile Telecommunications Technologies because of work that company has done in furthering PCS technologies. Comments are now being requested on technical and operational issues such as standards, and industry development.

Frequency and Time Forum Call For Papers — The European Frequency and Time Forum, to be held March 6-18, 1993 in Neuchatel, Switzerland has issued a call for papers. Papers are invited in areas such as: piezoelectric materials, resonators and sensors; quartz crystal oscillators and filters; atomic frequency standards and clocks; receivers for standard frequency and time signals; redundant frequency generating and timing systems; synchronization, acquisition and tracking equipment; and time and frequency in space. A one-page summary clearly outlining the major elements of interest should be sent by October 12, 1992 to the Swiss Foundation for Research in Microtechnology, FSRM, 7th European Frequency and Time Forum, Rue de l'Orangerie 8,

Avoid the Noise



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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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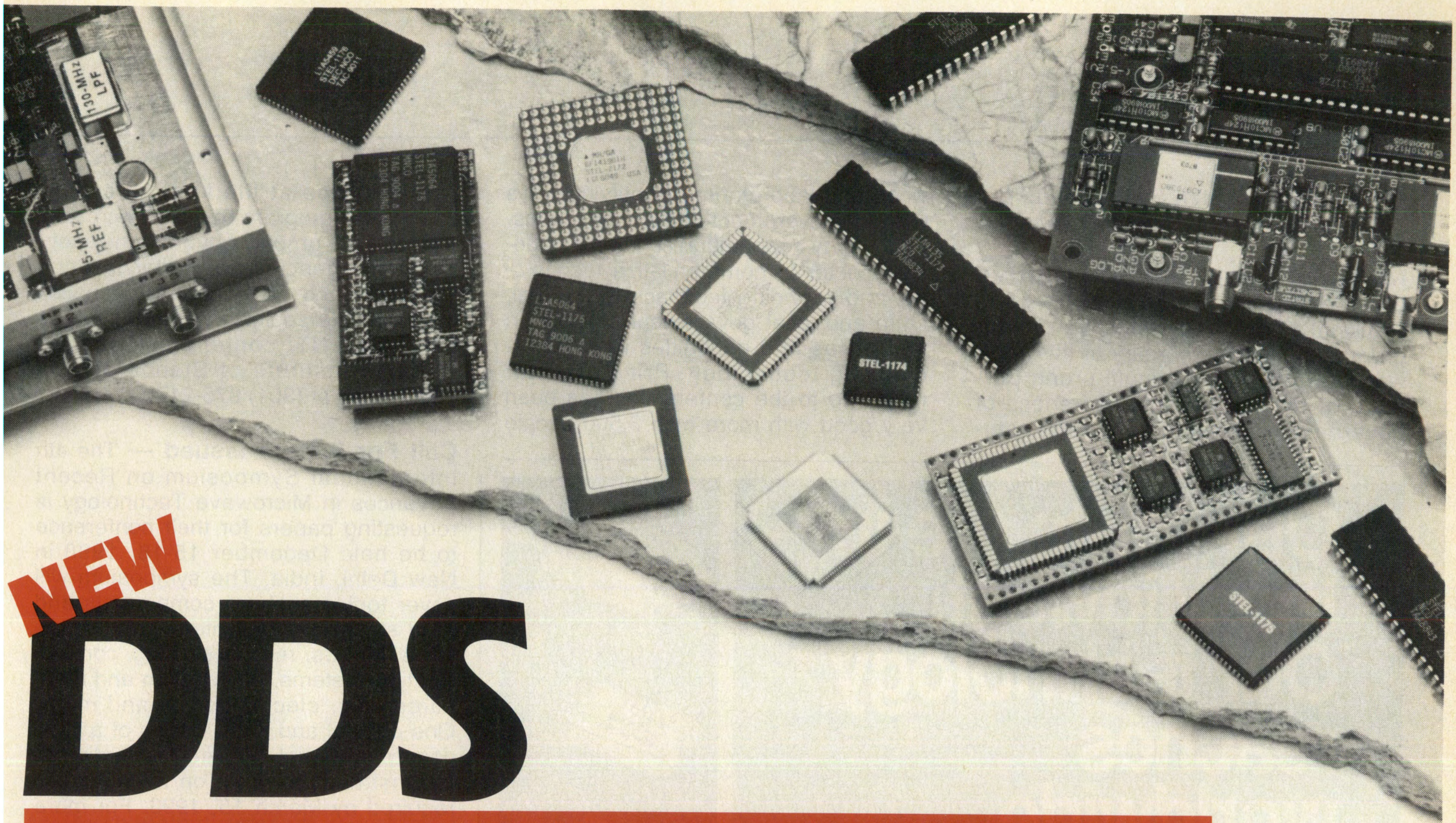

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STEL-1377	miniature assembly based on 1177 MIL Spec version now available NEW
STEL-1378	miniature assembly based on 1178 NEW
STEL-1276	based on 1176, 0-35 MHz
STEL-1277	based on 1177, 0-25 MHz
STEL-2272	based on 2172, 0-130 MHz
STEL-2273A	based on 2173, 0-400 MHz NEW
STEL-2373	based on 2173, 0-400 MHz - miniature hybrid NEW

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STEL-9272	300 MHz Synthesizer based on 2172
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National Technology Initiative Conferences Underway — The National Technology Initiative, a program launched in February of this year, was started with the multi-prong purpose of raising industry awareness of government sponsored programs, services and laboratories, identifying ways in which government-industry and industry-industry cooperation can help the private sector to commercialize technology, and providing feedback to Federal, state and local government officials on successes and difficulties in commercializing technology. Private sector response to the conferences has been very good with more than 3,500 people in attendance at the first ten conferences. The program has also increased cooperation among government agencies and agencies are better tailoring their information to respond to the needs of the private sector. For more information about NTI conferences call the National Institute of Standards and Technology at (301) 975-2170.

INFO/CARD 18

Call For Papers Issued — The 4th International Symposium on Recent Advances in Microwave Technology is requesting papers for their conference to be held December 15-18, 1993 in New Delhi, India. The symposium will cover topics such as components and solid state devices, antenna and radar, MICs, MMICs, remote sensing, communication systems, propagation and measurements, electro-optics, and more. One original and three copies of a four-page manuscript prepared according to the instructions (sent on request) are required by March 15, 1993. For more information contact, Banmali Rawat, Co-Chair, Technical Program Committee, ISRAMT-93, Department of Electrical Engineering, University of Nevada, Reno, Reno, NV 89557-0153. Tel: (702) 784-1457. Fax: (702) 784-6627.

Nepcon Call for Papers — The National Electronic Packaging and Production Conference, to be held June 14-17, 1993 in Boston, Mass., has issued a call for papers. A 200-300 word abstract should offer non-commercial user-oriented solutions in all disciplines of electronic circuit and systems design, packaging, fabrication, production, test and management. Abstracts must be received no later than October 2, 1992 and may be sent to Cahners Exposition Group, Cahner's Plaza, 1350 E. Touhy Ave., Des Plaines, IL 60018, Attention: Director of Conference Nepcon East '93.

New Paper Covers Status of Precision RF and Microwave Measurements — A recent paper from NIST summarizes the principles and present status of microwave measurements in scattering parameters, noise and power. Topics covered include calibration methods for automatic network analyzers, on-wafer MMIC measurements, microcalorimeters and other methods of high-accuracy measurements for power, and various radiometric techniques for noise measurements. The paper also contains an extensive bibliography. For a copy of the paper, which was published in *Metrologia* (May

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1992) for the 1992 Conference on Precision Electromagnetic Measurements, contact Jo Emery, Div. 104, NIST, Boulder, CO 80303. Tel: (303) 497-3237. Ask for paper number 25-92.

Virginia Tech Receives DARPA Grant — the Defense Advanced Research Project Agency recently awarded a $637,000 grant to researchers at Virginia Tech to develop new techniques for prediction of radio signal travel in buildings and urban areas. The project will combine computer graphic techniques and radio wave propagation theory for designing and installing radio systems in cities or inside buildings. The research team will consist of Ted Rappaport, director of the university's Mobile and Portable Radio Research Group, electrical engineering professors Marty Feuerstein and Charles Bostian and industrial systems engineering professor Hanif Sherali.

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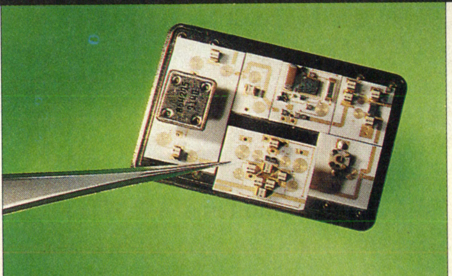

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Part Number	Frequency	Available
852103	300 MHz	May 1992
852104	400 MHz	May 1992
852105	500 MHz	March 1992
852106	600 MHz	March 1992
852107	700 MHz	March 1992
852108	800 MHz	March 1992
852109	900 MHz	April 1992
852110	1,000 MHz	June 1992
852111	1,100 MHz	June 1992
852112	1,200 MHz	June 1992

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MIT/Lincoln Lab Awards DSP IC Contract — Analog Devices, Inc., Electronics Designs, Inc. and Westinghouse Electronic Systems Group have signed a contract with MIT/Lincoln Laboratory to develop a high-performance DSP chip. Under the multimillion dollar contract, the team will produce a single-chip DSP for high-performance radar and IR applications, image analysis, missile guidance, FIR filtering, and high-performance graphics.

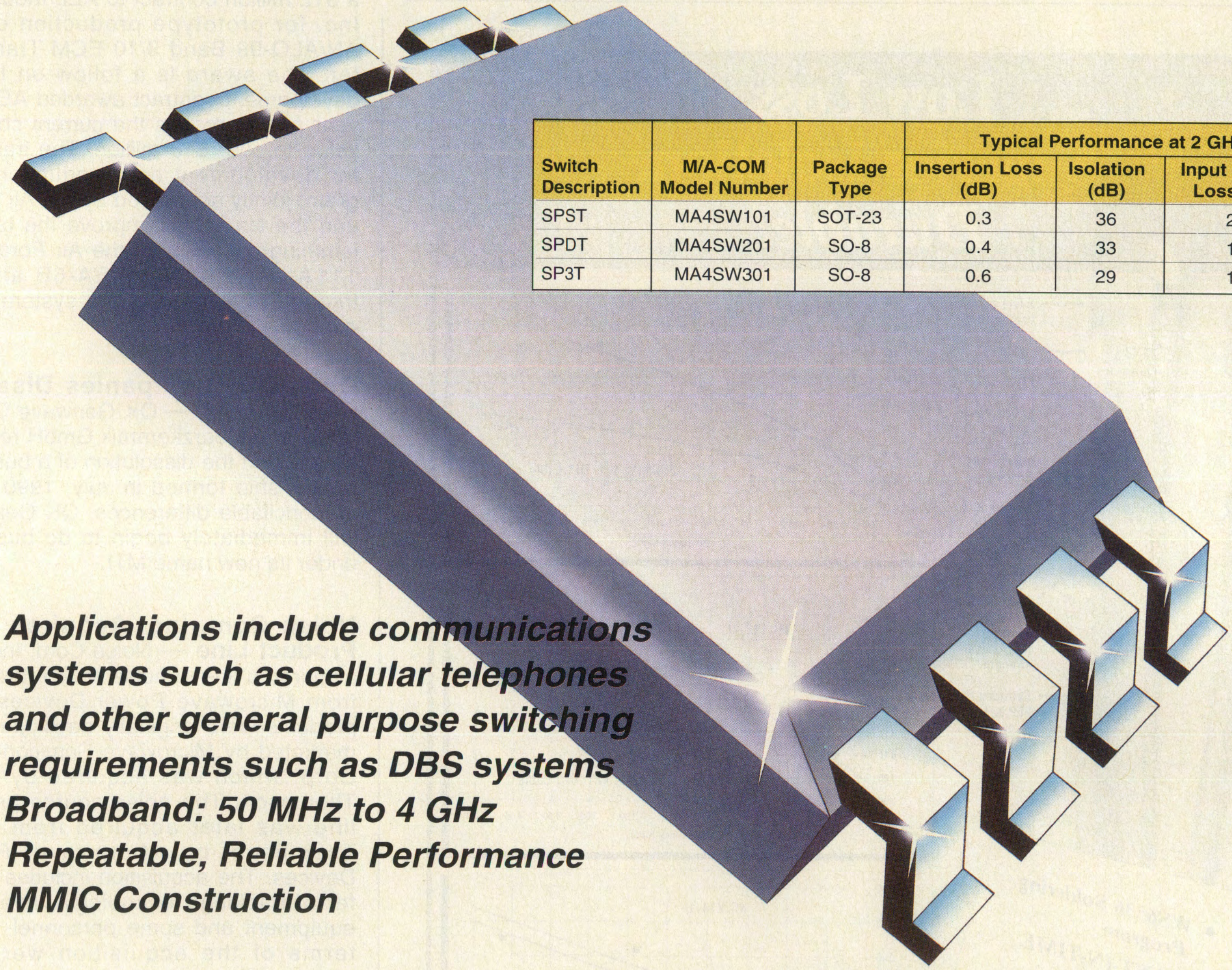
Stanford Telecom Wins Taiwanese PBX/PCS Contract — Stanford Telecom recently announced the award of a spread spectrum cordless telephone system development program from a consortium of five Taiwan companies. Stanford Telecom will design and develop handsets, base stations, remote handset cradles and other key equipment for the Taiwanese consortium who are building a next- generation digital telephone system for PBX and Personal Communications Service applications. A working cordless phone system is expected to be in field testing by the first quarter of 1993.

Power Electronic Components Moves — Power Electronic Components, Inc. has moved to a new location. Their new address is 6333 Old LaGrange Road, Crestwood, KY 40014. Tel: (800) 245-0193. Fax: (502) 241-5970.

Harris Awarded NATO Communications Contract — Harris RF Communications has signed a contract with the Turkish Ministry of National Defense to modernize the communications systems at COMEDNOREAST War Headquarters, a NATO site near Ankara, Turkey. The $7.5 million NATO infrastructure funded program calls for Harris to supply a turnkey 10 kW HF transmitter/receiver system for ship-shore communications. It will include Harris R-2368/URR receivers, RF-755 transmitters, a computer control system and other equipment.

FTS Receives Cesium Beam Tube Contracts — Frequency and Time Systems has been awarded multiple contracts to supply the U.S. Coast Guard, NASA and the U.S. Navy with

Surface Mount 50 MHz to 4 GHz PIN Diode Switches SPST, SPDT, SP3T



Switch Description	M/A-COM Model Number	Package Type	Typical Performance at 2 GHz		
			Insertion Loss (dB)	Isolation (dB)	Input Return Loss (dB)
SPST	MA4SW101	SOT-23	0.3	36	21
SPDT	MA4SW201	SO-8	0.4	33	17
SP3T	MA4SW301	SO-8	0.6	29	13

SO-8

- ***Applications include communications systems such as cellular telephones and other general purpose switching requirements such as DBS systems***
- ***Broadband: 50 MHz to 4 GHz***
- ***Repeatable, Reliable Performance***
- ***MMIC Construction***

For technical and applications assistance contact Jerry Hiller on ext. 2625
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See us at RFEE '92, Booths #704, 706, 708

replacement cesium beam tubes for Hewlett-Packard company cesium standards. The most recent awards follow six years of consecutive awards from the Coast Guard in support of the Loran-C navigation system. Other terms of the contract were not released.

Capax Technologies Relocates — Capax Technologies Inc. has announced the completion of its relocation into new facilities. Their new address is Capax Technologies Inc., 25435 Rye Canyon Road, Valencia, CA 91355. Tel: (805) 257-7666. Fax: (805) 257-4819.

Stanford Research Awarded Navy Contract — Stanford Research Systems has been awarded a three year contract with the U.S. Navy to deliver 180 FFT spectrum analyzers. The analyzers will be used in a variety of applications including sound, vibration, noise and electronic systems analysis. Other terms of the contract were not released.

AEL Receives Air Force Contract — The U.S. Air Force recently awarded a $12 million contract to AEL Industries, Inc. for prototype production of the AN/ALQ-99 Band 9/10 ECM Transmitter. The award is a follow on to the development contract awarded AEL last year and increases the current contract value to over $18 million. The transmitter development and production program, jointly sponsored by the Air Force and the Navy, will improve the tactical jamming capability of the Air Force EF-111A and the Navy EA-6B aircraft through expansion of the system's frequency and power.

Oscillator Companies Dissolve Relationship — QK Genwave Corporation and Quarzkeramik GmbH recently announced the dissolution of a business relationship formed in July, 1990 citing irreconcilable differences. QK Genwave will immediately begin to do business under its new name MTI.

Noise Com, Inc. Acquires MSC Product Line — Noise Com, Inc. has acquired a line of noise source products from Microwave Power Devices. The products were originally developed and marketed by Microwave Semiconductor Corp. which was acquired by SGS-Thompson. The noise source product line was later acquired from SGS-Thompson by Microwave Power Devices. The acquisition includes transfer of backlog, technology, inventory, equipment and some personnel. Other terms of the acquisition were not announced.

AT&T Microelectronics Forms New Business Unit — AT&T Microelectronics has formed a business unit to provide underlying technology for a new class of devices called Personal Communicators — small, mobile devices that accept input from a special pen, and which will be used primarily for communications. The unit, called Personal Communication Systems, will develop semiconductor products, development tools and software based on the Hobbit microprocessor.

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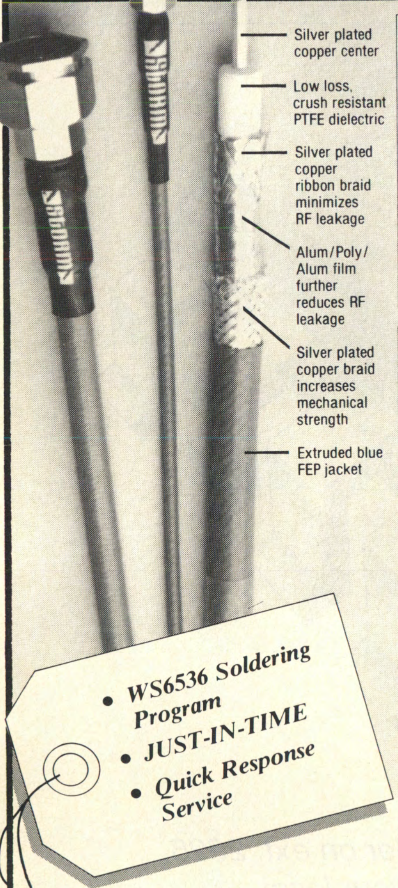
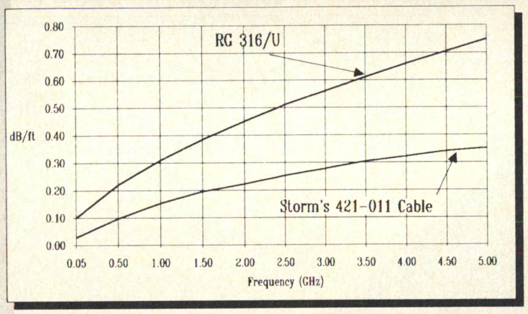
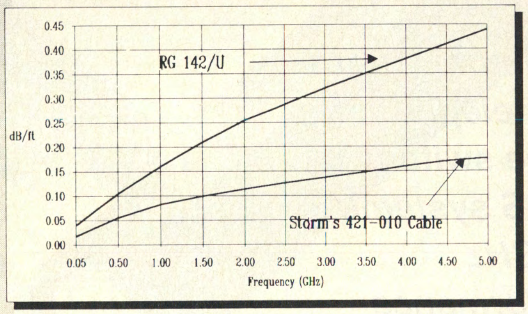
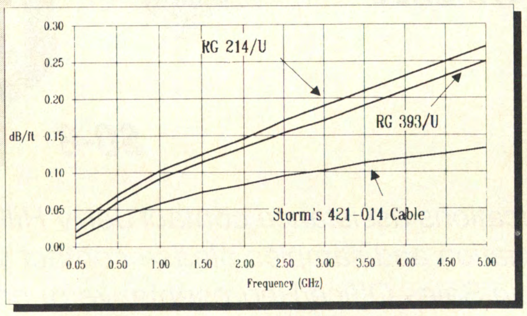
Superior Isolation,
Durable!









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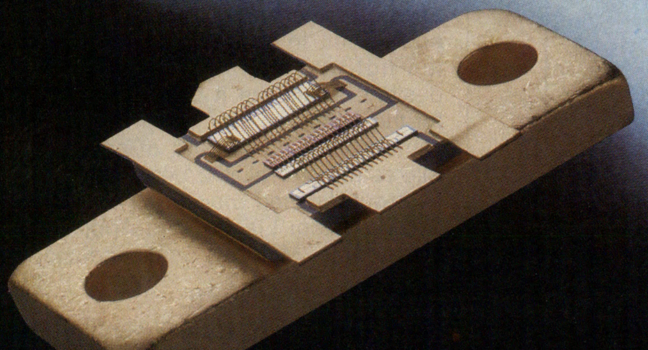
Application	P/N	Freq. (MHz)	P_{out} (w)	Gp Gain (dB)
Cellular	SD1650	860-960	60	7.0
Satcom	SD1898	1650	32	9.0
Avionics	SD1542-04	1090	600	6.0
Radar	AM1214-175	1215-1400	160	7.3
Radar	AM2729-100	2700-2900	105	6.5



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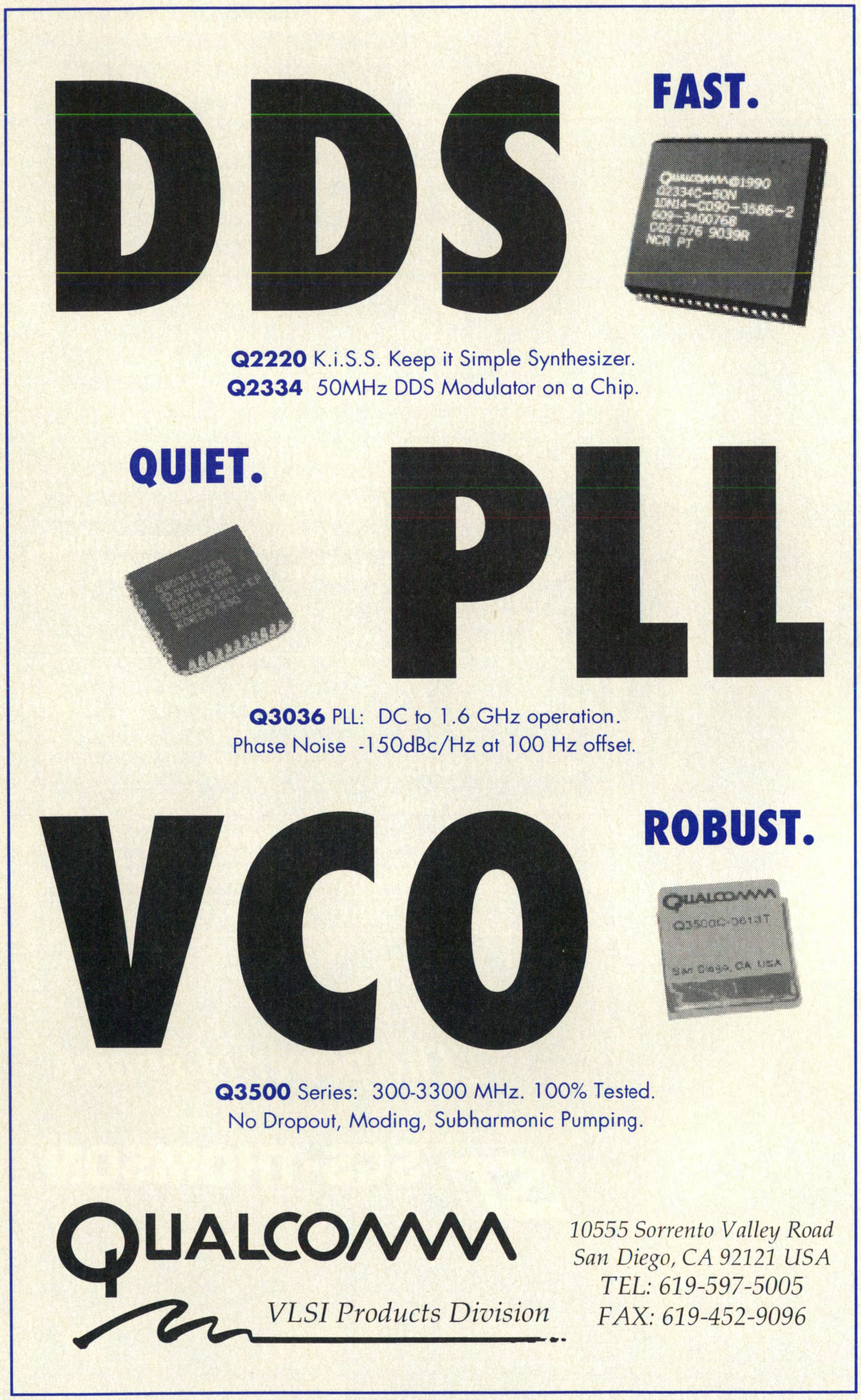
RF Oscillators and Synthesizers Perform on Demand

By Andy Kellett
Technical Editor

RF oscillators and synthesized sources have reached a stage of maturation. With the addition of SAW resonators and direct digital synthesizers to the stable of RF signal producing methods, further development in RF signal sources will take the form of refinement of existing technologies to fit new applications. This report will take a look at selected areas of refinement.



New Applications

Oscillators and synthesizers operating at radio frequencies are finding themselves in products not normally associated with RF. Frank Perkins, Vice President of Marketing for RF Monolithics notes that the SAW oscillators his company makes are found not only in keyless entry systems and avionics systems, but also as the clock base for engineering workstations. Crystal oscillators made by Hybrids International have been used in the same application, according to its President Abdul Ghafoor.

More and more RF products are found in transit. The requirements associated with this new demand for portability are affecting signal sources. RF oscillators and synthesizers are shrinking; according to Hybrids International's Ghafoor, his company produces a 200 MHz oscillator measuring only $0.31 \times 0.31 \times 0.1$ inches. This oscillator is targeted for the camera, lap-top computer and military markets. The X01145C ovenized crystal oscillator by Piezo Technology occupies $1.5 \times 1.5 \times 0.5$ inches. "The market we are seeing for the XO1145C is within portable test instruments; a piece of equipment someone is going to buy and use at a variety of sites, versus setting it down in his lab and using it there forever," says Paul Dechen, Director of Sales for Piezo Technology.

Not only size, but power requirements are shrinking. "The low current which our devices provide is a lot more preferable, of course, in portable battery equipment. In the mobile [applications] it's not quite as important, but the designers still like it because the case for the radio can be made much smaller without vent holes, and of course that makes it more reliable because you can seal out dirt and moisture," says David Babin, Patent Liaison and Applications Engineer for Motorola Semiconductor.

Better Performance

Long term drift is generally the result of climatic changes or component aging. The increased use of portable devices

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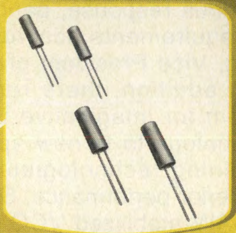


C-2
- LOW FREQUENCY TUNING FORK TYPE
- 32.768 KHz
- 20 KHz - 100 KHz

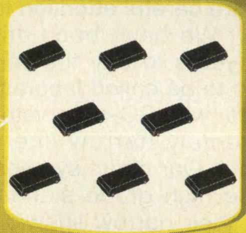
CA-301
- HIGH FREQUENCY AT - STRIP TYPE
- 4.0 MHz - 67.0 MHz

SMD

MC-405/6
- HEAT RESISTIVE C-2H INSIDE
- L10.4xW4.06xH3.56mm
- 20 KHz - 100 KHz

MA-505/6
- HEAT RESISTIVE CA-303 INSIDE
- L13.46xW5.08xH4.57mm
- 4.0 MHz - 67.0 MHz

MINIATURE SMD



MC-306
- NEW TUNING FORK CRYSTAL TECHNOLOGY
- L7.9xW3.8xH2.5mm
- 32.768 KHz

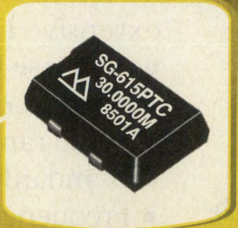
MA-306
- NEW AT - STRIP CRYSTAL TECHNOLOGY
- L7.9xW3.8xH2.5mm
- 17.7 MHz - 40.0 MHz

OSCILLATORS

THRU-HOLE

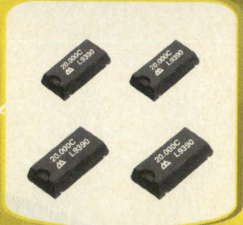
SG-51/531
- COST EFFECTIVE AUTO INSERTABLE FULL SIZE (SG-51) AND HALF SIZE (SG-531) DIP'S
- CA-301 CRYSTAL INSIDE
- 1.025 MHz - 67.0 MHz

SMD



SG-615
- SURFACE MOUNTABLE
- HEAT RESISTIVE CA-303 INSIDE
- L14.0xW9.8xH4.7mm
- 1.025 MHz - 67.0 MHz

MINIATURE SMD



SG-636
- MINIATURE SURFACE MOUNT
- NEW AT - STRIP CRYSTAL TECHNOLOGY
- L10.3xW5.8xH2.8mm
- 2.20 MHz - 40.0 MHz

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INFO/CARD 25

has increased the demand for oscillators that will remain stable when moved from climate to climate. Manufacturers are meeting this demand; Motorola's Babin noted that the MC145000 line of PLL synthesizers has parameters guaranteed over the temperature range of – 40 to +85 degrees C.

Close-in phase noise performance has been achieved in crystal oscillators for years, but some applications, such as nuclear magnetic resonance, still require careful attention to this specification. "We have been driven in the last few years to buy the very best of what used to be called laboratory grade oscillators, with SC-cut crystals to satisfy the extremely narrow line width requirements. Our direct synthesizer processes these high grade standards and transfers their narrow linewidth from MHz to GHz, " says George Lohrer, President of Programmed Test Sources.

The ability to characterize high performance oscillators has been limited in the past by test equipment. "We now have measurement systems with noise floors of –195 dBc/Hz, so we can now measure super quality oscillators which virtually no one could measure before," says Fred Walls, Project Leader for Phase Noise in the Time and Frequency Division of the National Institute of Standards and Technology (NIST).

1/f noise in crystal oscillators has also been investigated by Walls' group at NIST. A model which links this noise to quartz crystal electrode diameter, thickness, Q and frequency will allow crystal manufacturers to optimize their crystals to reduce this flicker noise.

As more RF products are sold to consumers, the prices of the components that go into those products are dropping, including signal sources. High volume manufacturers can produce hundreds of thousands of oscillators for a few dollars apiece. However, even signal sources that go into items with production runs of a few thousand are coming down in price. "Our Q2220 DDS is thirteen dollars in thousand piece quantities; add a DAC for four dollars, and you can have a turnkey DDS system for under twenty dollars," says Jim Madsen, Manager of Business Development for QUALCOMM. Companies that formerly concentrated on high-end products are also working to supply the expanding market. When asked about future goals for their products, Mike Cronin, Director of Sales and Marketing for the Vari-L Company said, "for the VCO line, lower cost is one of the more important items, lower power consumption is another, and maintaining and improving phase noise is still very important."

Although well developed, existing technologies are by no means ideal. DDS manufacturers are working to improve spurious response, bandwidth and power requirements according to Phil Feinberg, Vice President of Sales at Sciteq. In addition, there remains plenty of room for imaginative use of existing technology to fit new applications. Combining technologies often results in superior performance. Sawtek supplies a SAW stabilized VCO which improves the far-out phase noise performance of a crystal referenced phase lock loop. *RF*

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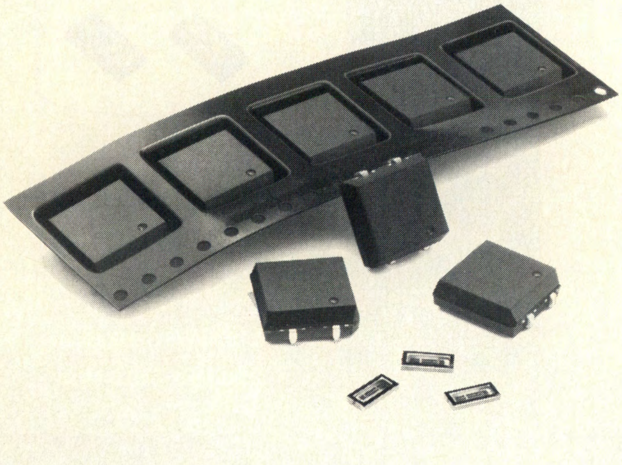

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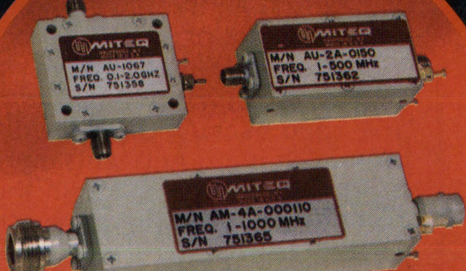
INFO/CARD 26

BIPOLAR AMPLIFIERS TO MEET YOUR SYSTEM REQUIREMENTS

The table below presents a sample from our standard bipolar amplifier design library. Most of these designs are available from stock. The amplifiers use internal voltage regulators and are reverse voltage protected. In addition, the following custom options are available on most models:



- Bias via the output RF port
- Low-loss input bias tee for photo diodes
- Integral input limiter
- Frequency response to below 10 kHz
- Amplitude/phase tracking and matching
- Fast recovery from input overloads
- Type N or BNC connectors

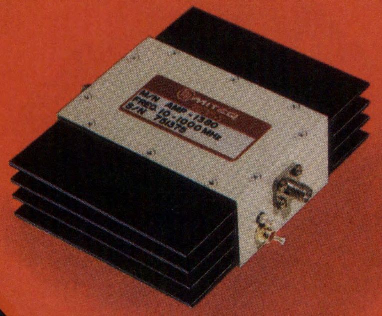
If the above options do not include your requirements, please give us a call. We might be able to satisfy your custom needs.

Guaranteed two-week delivery on most models for small quantity orders. Most models are available screened to tighter tolerances.

LOW NOISE AND GENERAL PURPOSE AMPLIFIERS

Model Number	Frequency (MHz)	Gain (Min.) (dB)	Gain Var. (Max.) (±dB)	Noise Figure (Max. dB) Band Low	Mid	High	VSWR (Max.)	Output 1 dB Gain Comp. (Min., dBm)	DC Power (+V)	(@ mA)
AU-1310	.01 – 500	30	0.50	1.3	1.4	1.5	2:1	8	15	50
AM-1300	.01 – 1000	25	0.75	1.4	1.6	1.8	2:1	6	15	50
AU-1378*	1 – 300	17	0.50	1.9	1.9	1.9	2:1	–2	6	10
AU-1379*	1 – 500	13	0.50	2.2	2.3	2.4	2:1	–2	6	10
AU-2A-0150	1 – 500	30	0.50	1.3	1.4	1.5	2:1	8	15	50
AU-3A-0150	1 – 500	45	0.50	1.3	1.4	1.5	2:1	10	15	75
AM-2A-000110	1 – 1000	25	0.75	1.4	1.6	1.8	2:1	8	15	50
AM-3A-000110	1 – 1000	37	0.75	1.4	1.6	1.8	2:1	9	15	75
AU-1021	5 – 300	24	0.50	2.2	2.4	2.6	2:1	20	15	175
AU-1158	20 – 200	30	0.50	2.7	2.7	2.7	2:1	17	15	125
AMMIC-1318	100 – 2000	6	1.00	4.5	4.0	4.0	2:1	12	15	35
AMMIC-1348	100 – 2000	14	1.00	5.0	5.0	5.0	2:1	14	15	150
AM-2A-0510	500 – 1000	24	0.50	1.4	1.5	1.6	2:1	0	15	50
AM-3A-0510	500 – 1000	38	0.50	1.4	1.5	1.6	2:1	10	15	75
AM-3A-1020	1000 – 2000	30	0.50	1.8	2.1	2.4	2:1	10	15	75

*Designed for low current battery operation



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AUP-1374	10 – 500	30	1.50	4.5	5.0	5.5	2:1	29	21	550
AUP-1383	20 – 300	35	1.50	2.4	2.5	2.6	2:1	29	21	650
AUP-1382	20 – 300	40	1.50	2.4	2.5	2.6	2:1	29	21	630
AMP-1380	10 – 1000	20	1.50	6.0	6.5	7.0	2:1	29	21	590
AMP-1381	20 – 1000	30	1.50	4.2	3.6	3.8	2:1	29	21	670
AMP-1389	10 – 1000	12	1.00	10.0	10.0	10.0	2:1	29	21	500

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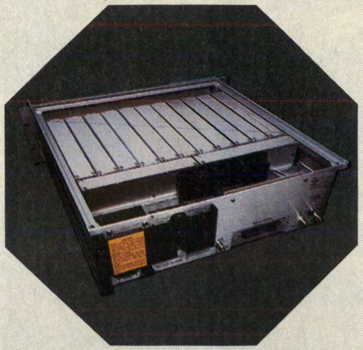

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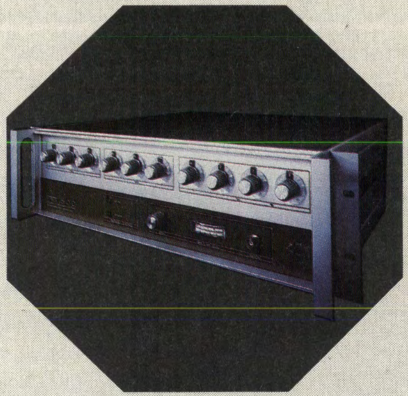

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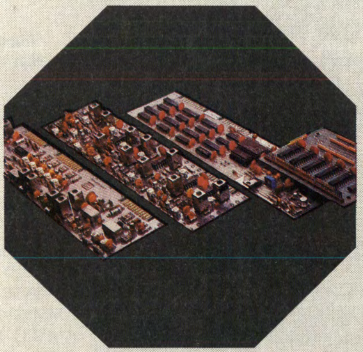

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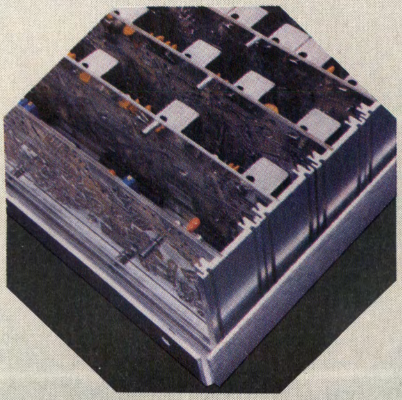

INFO/CARD 28

Fundamentals of Receiver Design for Part 15 Applications

By Bernard Kasmir
ADEMCO

In establishing reliability in communication, it is recognized that reliability is reduced as the signal becomes weaker and approaches threshold sensitivity of the receiver. At the limit, communications cease as the received signal level falls below the detection threshold of the receiver.

In this article, we shall look at all factors affecting this receiver threshold level and what can be done to the receiver to extend the reliability of the system. The purpose of this article is to interpret the basic equations involving sensitivity and noise figure and to explore practical methods of receiver design.

Just how sensitive can we make a receiver? What are the practical limits? In order to answer these questions, it is necessary to go back to the fundamentals.

The theoretical noise floor of a system at bandwidth B and temperature T, expressed as equivalent noise power is:

$$P = kTB \qquad (1)$$

Where:

k = Boltzmann's constant
T = Temperature in Kelvin
B = System bandwidth

This is the theoretical signal level where the signal to noise ratio would be equal to 1 (tangential sensitivity) in an ideal receiver. However, this level is deteriorated by the receiver noise figure by an amount equal to F, described as:

$$F = (S_i/N_i)/(S_o/N_o) \qquad (2)$$

Any network noise figure can be defined as the actual noise output divided by the output noise of an ideal system. All networks have some noise figure greater than 1. The best noise figure is the lowest amount.

Each stage in a receiver has a noise figure which deteriorates the signal to noise ratio. The total contribution of noise figure of several stages would be:

$$F_t = F_1 + (F_2-1)/G_1 + (F_3-1)/G_1G_2 + \ldots (F_N-1)/(G_1G_2 \ldots G_{(N-1)}) \qquad (3)$$

Where:
F_t = overall noise figure
F_1 = noise figure of the first stage, etc.

Stage	Noise Figure	Gain
1	3 DB	10 DB
2	7 DB	5 DB
3	5 DB	20 DB

Figure 1. Parameters for the composite noise figure.

G_1 = amplification of the first stage, etc.

As a numerical example, if the given parameters are shown in Figure 1, then by equation 3, the composite noise figure = 3.92 dB. It can be seen that the composite noise figure is most dependent upon the first stage of the receiver.

An additional input signal equal to the value of F_t is necessary simply to bring the receiver back to tangential sensitivity. But the receiver will not usually decode at tangential sensitivity. A particular signal to noise ratio (determined by the design of the system and circuits) is necessary. Therefore, the received signal level must be increased both by the system noise figure, F_t, and the signal to noise ratio.

For example, assume we have the following information:
System noise figure = 5 dB

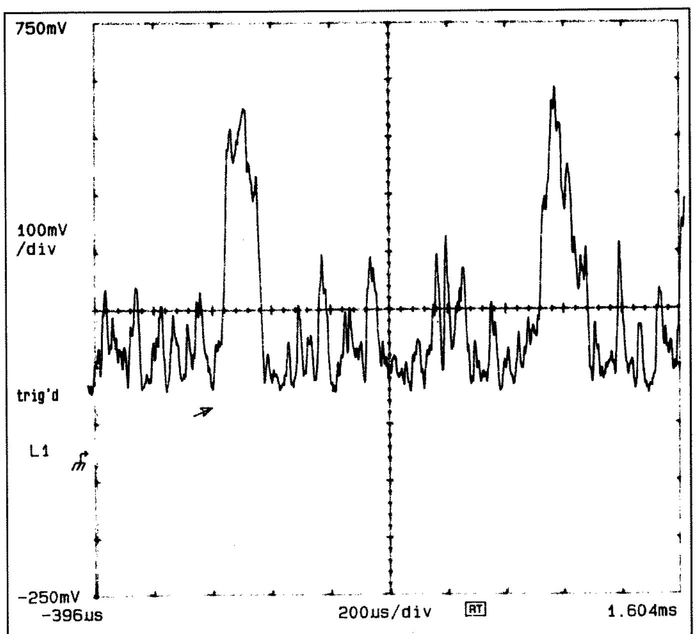

Figure 2. Weak signal.

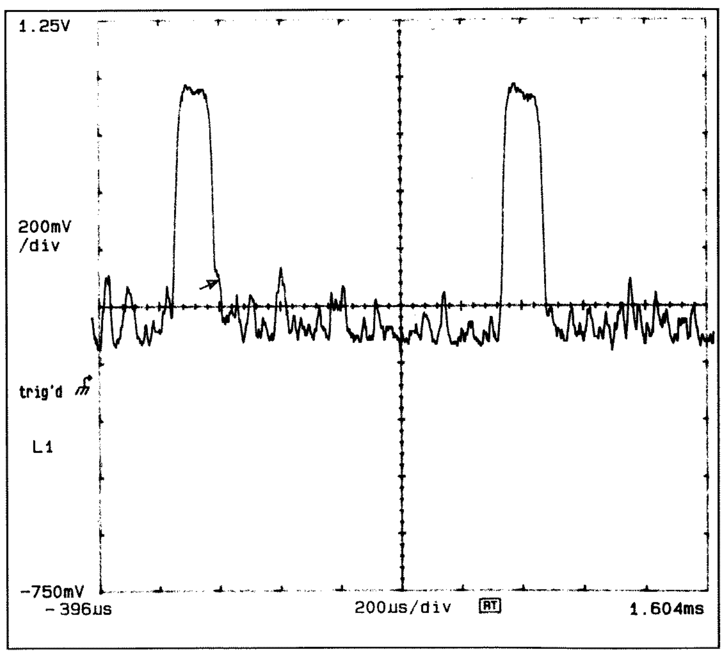

Figure 3. Stronger signal.

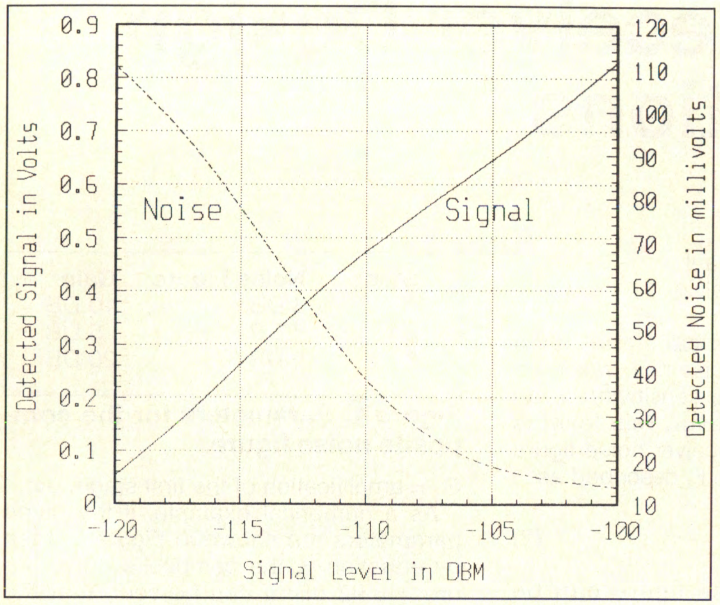

Figure 4. Signal and noise output vs. input.

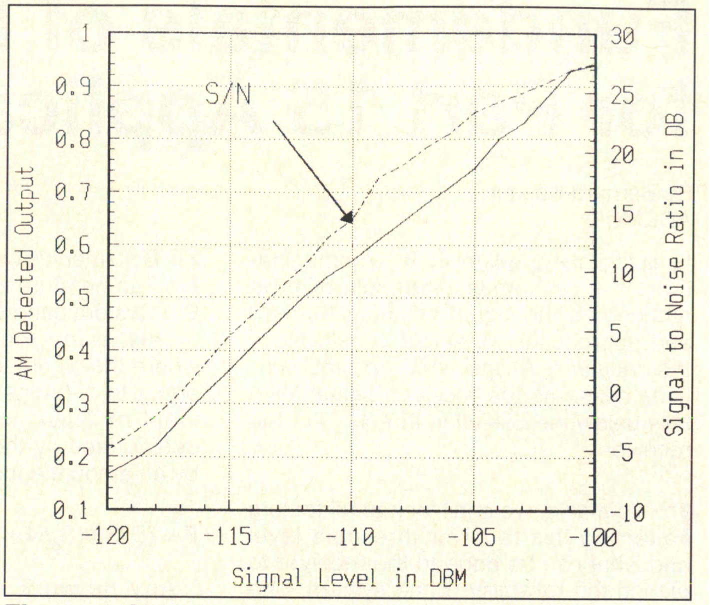

Figure 5. Signal and signal to noise ratio vs. input.

Minimum signal to noise ratio for detection = 10 dB
System bandwidth = 30 kHz

The equivalent noise power from equation 1 would then be: 1.2×10^{-16} watts, or -129.2 dBm. The input signal would have to be greater than this amount by at least the noise figure of 5 dB. Also, since we specified 10 dB minimum signal to noise ratio, the minimum signal has to be 10 dB above the noise level or a total of 15 dB in this example.

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Noise Figure Measurement Below 10 MHz

The simple test set-up shown in Figure 1 can be used to make swept noise figure and gain measurements. It requires only a Noise Com NC 3201Y noise source, low-noise preamplifier, and a spectrum analyzer, and optionally a 3-to-6 dB attenuator and precision attenuator.

The noise figure is defined as:

$$NF(\text{in dB}) = ENR - 10\log(Y - 1) + 10\log A$$

where:

$Y = P_{oON}/P_{oOFF}$
P_{oON} = output power of DUT with the noise source on
P_{oOFF} = output power of DUT with the noise source off
$10\log A$ = a temperature correction factor

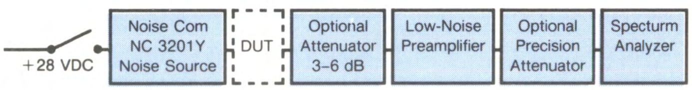

Figure 1. Noise Figure Measurement Set-up Using Spectrum Analyzer and NC 3201Y Noise Source.

The spectrum analyzer memory can be used to store the P_{oOFF} trace (in dBm/Hz). The P_{oOFF} trace is then subtracted from the current P_{oON} trace (in dBm/Hz) and the difference is the Y factor (in dB). Some manufacturers provide resident software in their spectrum analyzers for noise figure measurements using Noise Com's NC 3201Y noise sources.

For ambient temperatures significantly different than 290 K, a correction (10logA) must be applied. The correction is most significant when measuring low noise figures and when a noise source with less than 15 dB ENR is used.

$$A = 1 - [(T_c/290) - 1] \times [Y/10^{(ENR/10)}]$$

Corrections for second-stage effects (the preamplifier plus attenuators plus analyzer noise figure) should be made using the following equation:

$$F_{actual} = F_{measured} - [(F_{preamp} - 1)/G_a]$$

where:

F_{actual} = the actual noise factor of DUT (not in dB)
$F_{measured}$ = the measured noise factor (not in dB)
F_{preamp} = the noise factor of the preamplifier plus analyzer and attenuators (not in dB)
G_a = the available gain of the DUT (not in dB)

$$G_a = \frac{P_{oON} - P_{oOFF}}{k \times B \times (T_h - T_c)}$$

where:

B = the noise bandwidth of the measurement system
k = Boltzmann's constant = 1.380×10^{-23} J/K
$T_h = 290 \times [1 + 10^{(ENR/10)}]$ in degrees K
T_c = room temperature in degrees K

The last effect that must be considered is the impedance mismatch between the DUT, the noise source, and the preamplifier. The mismatch leads to measurement uncertainty. Consequently, impedances should be kept as close to 50 ohms as possible. This can be done by using a well-matched preamplifier, by introducing an attenuator between the preamplifier and the DUT, and by using a well-matched noise source.

Using a Noise Com NC 3201Y noise source and following these guidelines results in a contribution to uncertainty that is less than that contributed by the spectrum analyzer.

Spectrum analyzer uncertainty mainly originates from the linearity of its power meter. Improved measurement accuracy can be obtained by RF substitution. A precision attenuator is placed between the preamplifier and the spectrum analyzer and the attenuation is changed from 0 to Y dB as the noise source is switched from off to on. The RF power entering the spectrum analyzer is kept constant, which eliminates power meter nonlinearity. The Y factor (in dB) is read directly on the precision attenuator.

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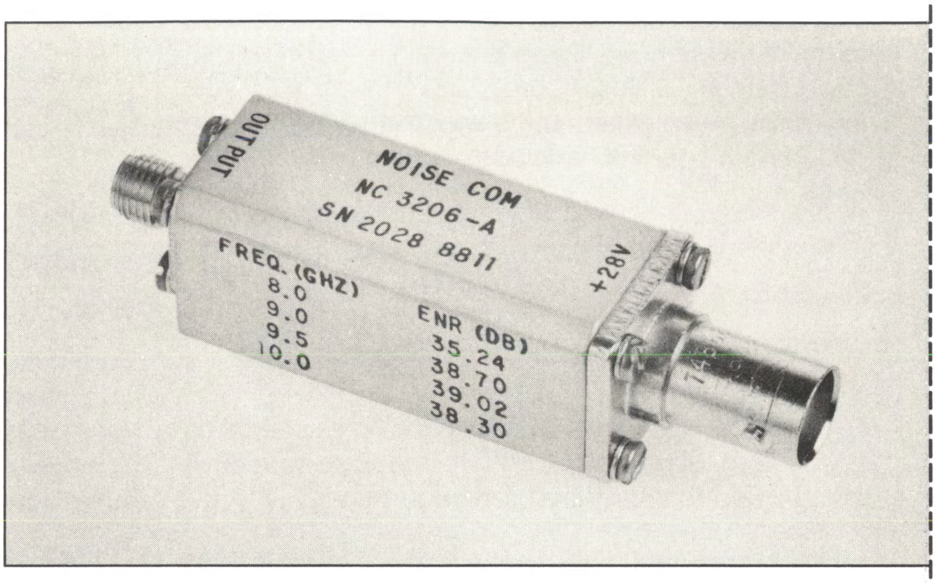

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NC 3101	0.01– 8	15.5±0.5	1.20:1	10 MHz, 100 MHz, 1 GHz & 1 GHz steps
NC 3102	0.01–12.4	15.5±0.5	1.20:1	
NC 3103	1 –12.4	15.5±0.5	1.20:1	1 GHz steps
NC 3104	1 –18	15.5±0.5	1.35:1	
NC 3105	12 –18	15.5±0.5	1.35:1	
NC 3108	0.1 –18	15.5±0.5	1.35:1	
NC 3110	12 –40	15.5±0.5	1.50:1	

HIGH NOISE OUTPUT TYPES

MODEL	FREQUENCY RANGE (GHz)	NOISE OUTPUT		CALIBRATION FREQUENCIES
		ENR (dB)	Flatness (dB)	
NC 3201	10 kHz–1.1	30–35	±1	10, 100, 500 & 1000 MHz
NC 3202	0.001– 0.6	30–35	±1	10, 100 & 500 MHz
NC 3203	1 – 2	30–35	±1	1, 1.5 & 2 GHz
NC 3204	2 – 4	30–35	±1	1 GHz steps
NC 3205	4 – 8	30–35	±1	
NC 3206	8 –12	28–33	±1	
NC 3207	12 –18	26–32	±1	
NC 3208	1 –18	26–32	±1	
NC 3208S	1 –18	30–32	±1	
NC 3209	18 –26	29–32	±1	
NC 3210	12 –40	19–23	±1	

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Therefore, the minimum acceptable signal would be:

Noise power + Noise figure + Signal to Noise Ratio

-129.2 dBm + 5 dB + 10 dB = -114.2 dBm

In this example, the receiver sensitivity would be .43 microvolts into 50 ohms. This is the best the receiver can do given these parameters.

Now that the sensitivity of the receiver has been defined, how much amplification or gain is necessary to complete the design. If we define the minimum amplitude for the detector as 100 mV into 10 Kohms, this is a power level of 200 microwatts. The minimum receiver signal level calculated was -114.2 dBm, or 3.8×10^{-15} watts. The minimum amount of receiver amplification would then be the output of 200 microwatts divided by 3.8×10^{-15} watts or a power gain of 91.2 dB. This particular amplification level will then be at the transition point between a gain limited and noise limited receiver.

A noise figure limited receiver is defined as a receiver with adequate gain so that the decodable signal is limited by the noise figure of the receiver for a defined bandwidth. In other words, the weakest signal that can be received is limited by the signal to noise ratio. Figure 2 shows the detected output of a receiver. If we assume that a minimum of 10 mV signal is required for decoding, the amplitude is adequate but the signal to noise would not allow the signal to decode. Figure 3 shows the results of a stronger signal. In this case, the signal to noise ratio is adequate.

A gain limited receiver is defined as one where the receiver has insufficient amplification. That is, the signal to noise ratio is adequate for decoding, but the detected signal level is too low to be processed. Improving the noise figure of a gain limited receiver would not enhance the sensitivity because the signal to noise ratio is already adequate.

If the gain of a gain limited receiver were increased, there would be a crossover point where the detected amplitude and the signal to noise ratio would be adequate. The signal level for this event would be less than that of the gain limited receiver because more amplification means the detection of a lower level signal (as long as the signal to noise ratio met our minimum requirement).

Based upon these principles, what would happen if a preamplifier were placed ahead of a radio? That depends upon the noise figure of the preamplifier as well as that of the receiver itself. There are various combinations:

a) The receiver is gain limited and the preamplifier enhances the noise figure.

b) The receiver is gain limited and the preamplifier deteriorates the noise figure.

c) The receiver is noise figure limited and the preamplifier enhances the noise figure.

d) The receiver is noise figure limited and the preamplifier deteriorates the noise figure.

For a), if the preamplifier only enhanced the noise figure, the sensitivity would not change, but if the preamplifier has gain, the system would change from gain limited to noise figure limited and could accommodate a lower level signal. Therefore, the sensitivity would be improved under those conditions.

For b), since the receiver is gain limited, the additional gain may bring the system into noise figure limited. However, the preamplifier deteriorates the noise figure. The end result would depend upon the magnitude of the additional gain and deteriorated noise figure and can only be determined if values are assigned.

For c), since the preamplifier enhances the noise figure as well as the gain, the sensitivity will increase.

For d), since the preamplifier deteriorates the noise figure, and since the system is already noise figure limited, the sensitivity will decrease.

To review, if there is an excess of total amplification, the system is noise figure limited. If there is a deficiency of total amplification (for the same noise figure), the system is gain limited. The receiver should be designed to be somewhat noise figure limited because variations in receiver amplification will not affect sensitivity. However, for best receiver dynamic range, the gain should be just enough to make it noise figure limited since excessive gain will not improve the sensitivity but will cause signal compression or distortion at higher levels.

Now that the receiver has been designed, what can we do to make this better? The factors usually under the control of the designer are:

— System bandwidth. This defines the theoretical noise floor. Bandwidth is affected by the amount of information, baud rate and other practical factors as temperature coefficient and frequency stability. The bandwidth must always be increased to accommodate these inaccuracies.

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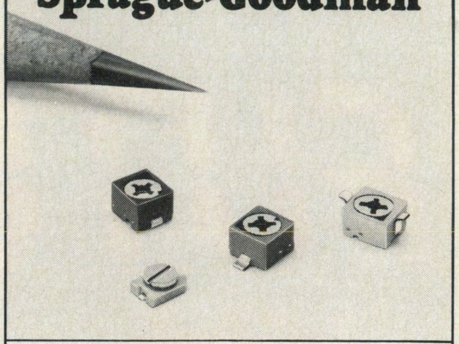

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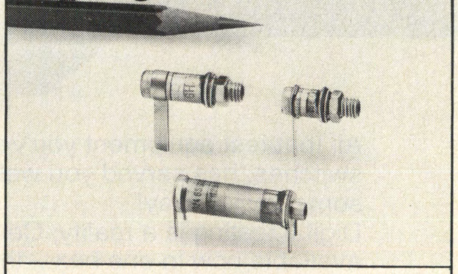

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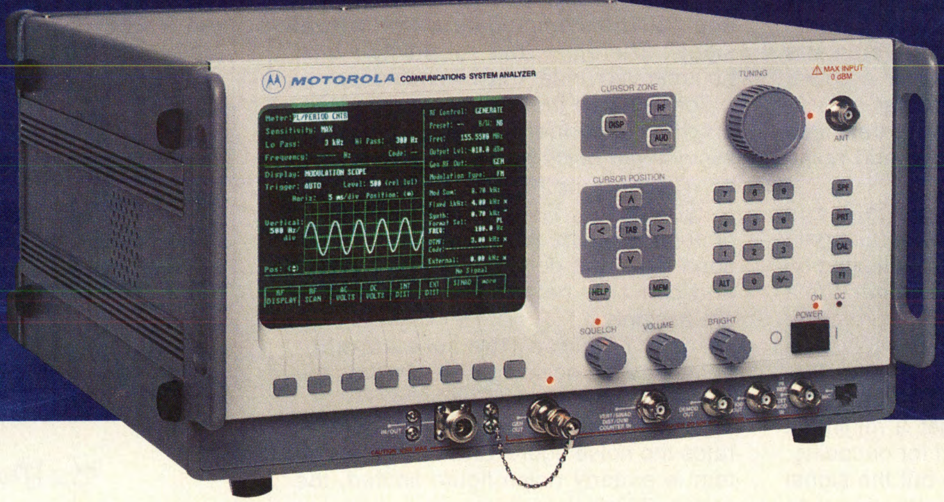

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— System noise figure. This requires the choice of low noise amplifiers and noise matching. There are choices that can be made to some extent on cost/performance tradeoffs.

— Required output signal to noise ratio. This involves the equivalent of a lower noise figure by designing the post detection circuitry to operate in a poorer signal to noise ratio. Again, factors such as baud rate, transmission length, clash probability, etc., must be optimized for any particular design.

The receiver effective sensitivity can also be controlled by the choice of antennas. Field intensity for Part 15 devices is specified and measured in microvolts/meter. The receiver does not end at the antenna input terminal. A receiving system consists of a receiver plus an antenna. The ability of a receiver to pick up a signal is characterized as Sensitivity = Receiver Sensitivity × Antenna Factor.

In our example, the receiver sensitivity was calculated as .43 microvolts into 50 ohms. Now the antenna factor transforms this in a device capable of intercepting electromagnetic energy. Typically, for a dipole at 300 MHz, the antenna factor is 10 or 20 dB. This means that with a .43 microvolt receiver, we can receive and decode a field intensity of 4.3 microvolts/meter. In a previous article, this quantity was related to systems reliability (1). The point we wish to make is that the receiver antenna is very important to actual sensitivity and that is why care must be taken in an installation to provide the correct environment to maintain a good antenna factor.

Other Factors

So far, we have discussed only "natural" noise. In many cases, the limiting factor is man-made noise. Previously, we have equated receiver sensitivity to the equivalent noise of the universe and have calculated how much above this level a signal would have to be in order to decode. Now, suppose that man-made noise was substantially above this natural noise level. The calculations for a minimum received signal would be the same except that the level is above the total ambient noise. It is possible for two receivers to measure different sensitivities under shielded laboratory conditions but have similar performance in a noisy environment.

Deliberate Loss of Sensitivity

A very useful installation procedure is to deliberately desensitize the receiver during initial installation and then test that communications have been reliably established between the receiver and the remote transmitters. During normal operation, the receiver sensitivity is restored to normal levels. This amount of desensitization during the installation process then becomes the additional signal margin for the system. For example, if the desensitization is 10 dB, then the installer knows that the weakest signal is at least 10 dB above the minimum threshold.

Now that we understand the differences between and principles involved in gain and noise figure limited, we can proceed to desensitize the receiver.

The most direct and most predictable method would be to introduce attenuation in the antenna lead. This reduces the signal as well as the overall gain of the radio. It is not really necessary to know which way the receiver is limited since this method affects both.

One could also reduce the gain of the receiver somewhere in the IF where it did not materially affect the noise figure. In this case, sensitivity would not be affected until sufficient gain is reduced to where the receiver becomes gain limited (assuming it was initially noise figure limited).

The sensitivity of the post detection processing circuits can be reduced. Again, desensitization will only occur when the receiver effectively becomes gain limited. Also, care must be taken in the video circuitry of a pulsed system to avoid characteristic pulse distortion. This could make the level of desensitization less reliable.

Figure 4 shows the output of an AM detector for a pulsed signal. At low signal levels, the noise output is greater than the signal. With stronger signals, the noise decreases while the detected signal increases. Figure 5 displays the same information in a different form. This is a plot of detected signal level and signal to noise ratio.

For illustrative purposes, assume that one receiver requires a detected level of 0.6 V and a signal to noise ratio of 5 dB. From Figure 5, at 5 dB S/N, the detected output is roughly 450 millivolts. In this example, at the desired signal to noise ratio, the detected output is insufficient. This is an example of gain limited. Now, let us assume that we require a 10 dB signal to noise ratio for adequate detection and only 200 mV signal level. From the curve at 10 dB signal to noise ratio, the recovered audio is better than 500 mV. The receiver cannot decode at lower levels because the signal to noise ratio is inadequate. This is an example of a noise figure limited receiver.

In this article, we have defined how to determine the sensitivity of a receiver and what factors are under the control of the designer. We have also defined the limiting factors of gain and noise figure. Good receiver design is essential in a low power communication system operating with very limited power for optimization of system reliability. ***RF***

References

1. B. Kasmir, "Communications Range and Reliability of Part 15 Devices," *RF Design*, April 1991, pp. 65-68.

About the Author

Bernard Kasmir is Senior RF Design Engineer at Alarm Devices Manufacturing Company (ADEMCO), 165 Eileen Way, Syosset, NY 11791. He holds BSEE and MSEE degrees and a P.E. license.

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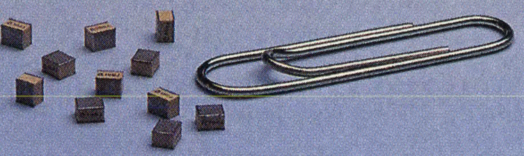

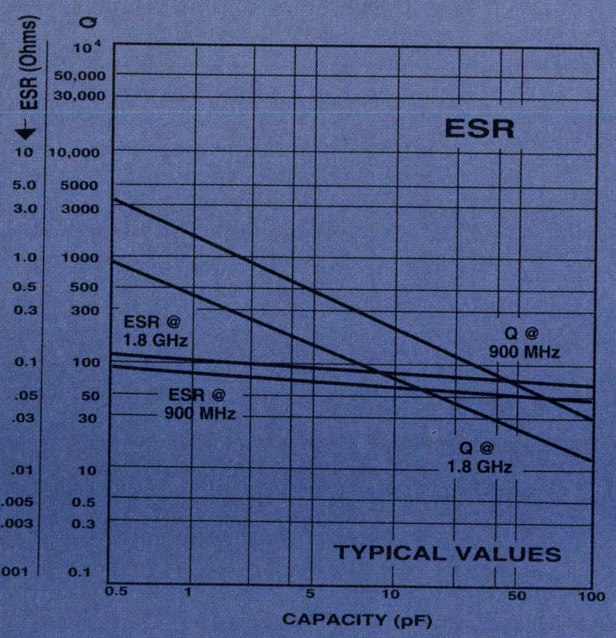

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New CAE Software for Designs That Work the First Time

Daren B. McClearnon
Hewlett-Packard Company

This month, Hewlett-Packard Company will demonstrate the HP RF Design System at RF Expo East in Tampa, Florida. The high-performance software package provides for designers of RF systems, subsystems, and circuits the tools needed to finish designs in less time, with better performance, fewer prototypes, and higher manufacturing yields.

In the past, RF design software usually consisted of simulators and PC-based tools that were not well integrated and supported only specific tasks in the RF-design process. Now, the HP RF Design System addresses the entire process and completely integrates the most advanced schematic-entry system, parts libraries, system and circuit simulation in the time domain and frequency domain, data analyses, test-equipment links, artwork, and documentation capabilities available.

HP RF Design System Overview

The HP RF Design System provides value to the RF communications market because it supplies, for the first time, a complete tool for simulating RF circuits and systems together in either the frequency or time domain. Every feature and module is designed to support accuracy, increase productivity, and eliminate design iterations. The software's ability to simulate noise in nonlinear circuits allows designers to solve a whole new class of problems.

The system consists of high-frequency design modules seamlessly integrated within a common user interface. The modules include frequency- or time-domain simulators, libraries, artwork generation, translators, and accessory products.

At the heart of the HP RF Design System is a graphical, icon-based user interface. Menus and commands are invoked from a mouse-driven interface that provides on-line help, fast keyboard customization for advanced users, macros, and even multi-event "undo" functions. Designs are stored hierarchically in icons within the design environment, resulting in a single external file and vastly simplified design management.

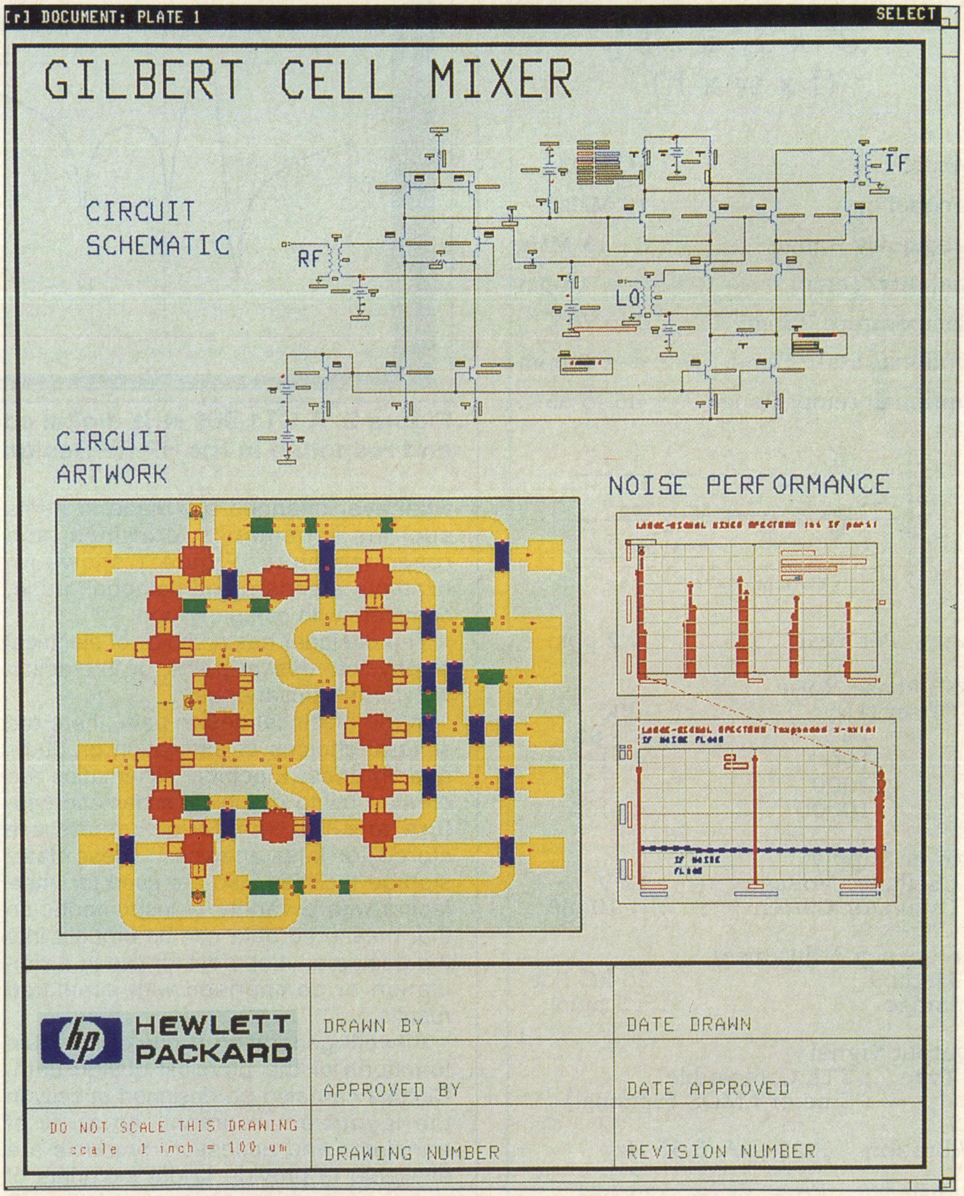

Figure 1. Views of integrated HP RF Design System tools can be collected in a documentation area, with design changes updated in real time.

The system uses an object-oriented database that stores all of the information related to a design in a common location. As a result a notebook can be created and automatically updated

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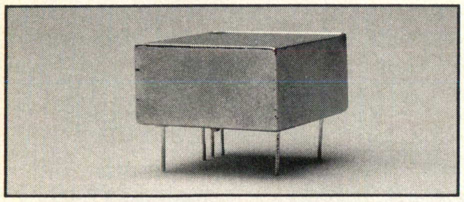
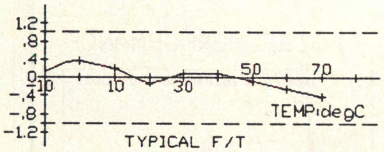


.8" x .8" x .45"
(l x w x h)

Model	: XO3003
Frequency	: 10 MHz
Frequency Range	: 8 - 15 MHz
Stability/Temp.	: < ± 1 ppm
Temperature Range	: 0 to 70°C
Optional Stab/Temp.	: < ± 2 ppm
Optional Temp. Range	: -45 to 85°C



Aging Per Year : < 2 ppm

SSB Phase Noise

Offset (Hz)	dBc
10	: < - 80
100	: < -110
1000	: < -120
10000	: < -135

Power Supply
Oscillator Voltage : 12 V
Oscillator Current : < 10 mA

Frequency Adjustment
Method : 20 kΩ Pot
Range : ±5 ppm

Output Signal
Type : TTL Compatible
: Sine or CMOS (optional)

Vibration : .025" DA, 5-55 Hz

Shock : 50 g, 11 ms, 1/2 sine

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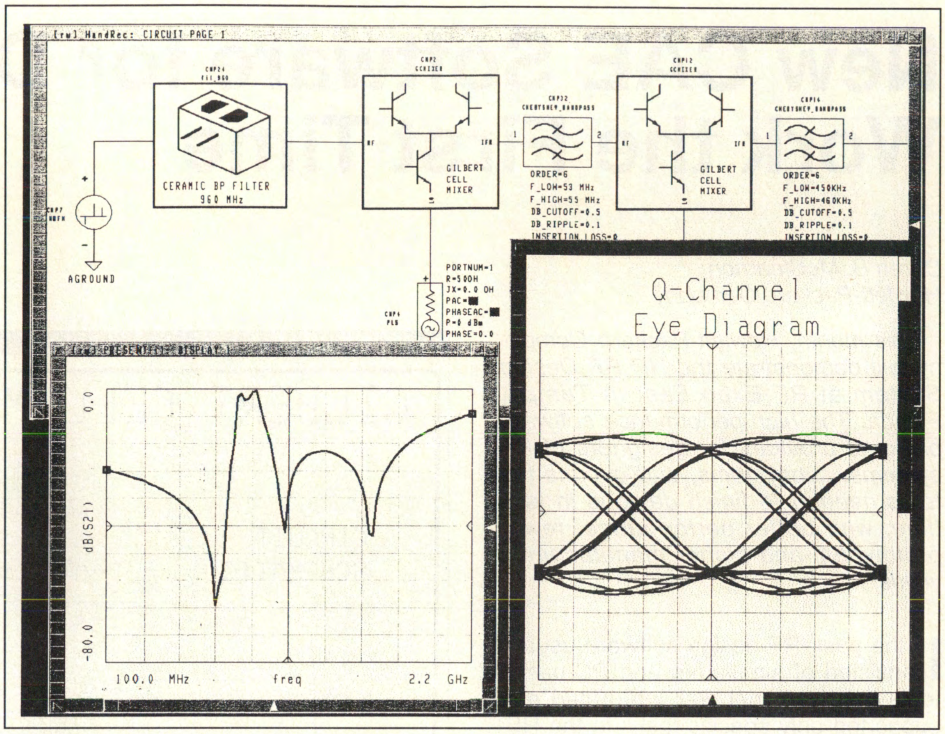

Figure 2. A CT1 900 MHz digital cordless phone example: block diagram and response in the HP RF Design System.

whenever changes are made to items such as schematics, drawings, and response plots. Figure 1 shows an example standardized notebook page, complete with company logo.

The designer has a choice of electrical or physical design paths, or interaction with both simultaneously.

In the electrical design path, there are several choices of simulation products and libraries. Electrical schematics are created using graphical schematic capture and the simulation results are stored for later analyses. These data-storage facilities also are used for interfacing with a variety of instruments, so that measured data can be brought into the design environment for use in a simulation or comparison with simulated results.

Artwork generation is one of the main functions of the physical design path. Circuits can also be designed directly in the layout environment. An array of translator and accessory products are available to provide layout footprints of common devices and connect the HP RF Design System to an existing fabrication system. The graphics editor provides an advantage over other editors because its artwork tools share the same database and interface with other tools in its design environment.

Optional links to the Falcon 8.0 framework (Mentor Graphics) and Design Framework II (Cadence Design Systems) are also available.

Simulation Within the HP System

The HP RF Design System is equipped with linear and nonlinear simulators that operate in either the time or frequency domain. The foundation of the simulation tool set is the HP RF Linear Simulator. This basic capability includes DC and small-signal, frequency-domain AC and S-parameter simulations.

Nonlinear frequency- and time-domain options build upon the HP RF Linear Simulator. The simulator works in the frequency domain and is a proprietary implementation of the harmonic-balance simulation technique. Harmonic balance allows circuit designers to refine the nonlinear performance of amplifiers, mixers, oscillators, and other circuits. The new technology of the time-domain simulator, HP Impulse, is discussed in detail at the end of this article.

The HP System Model Library adds a block-diagram personality to the HP RF Nonlinear Simulator. Data sheet descriptions of commercial parts, such as gain blocks, mixers, and filters can be combined with device-level circuit schematics. This gives designers the choice of managing the design process from the block level downward (adding

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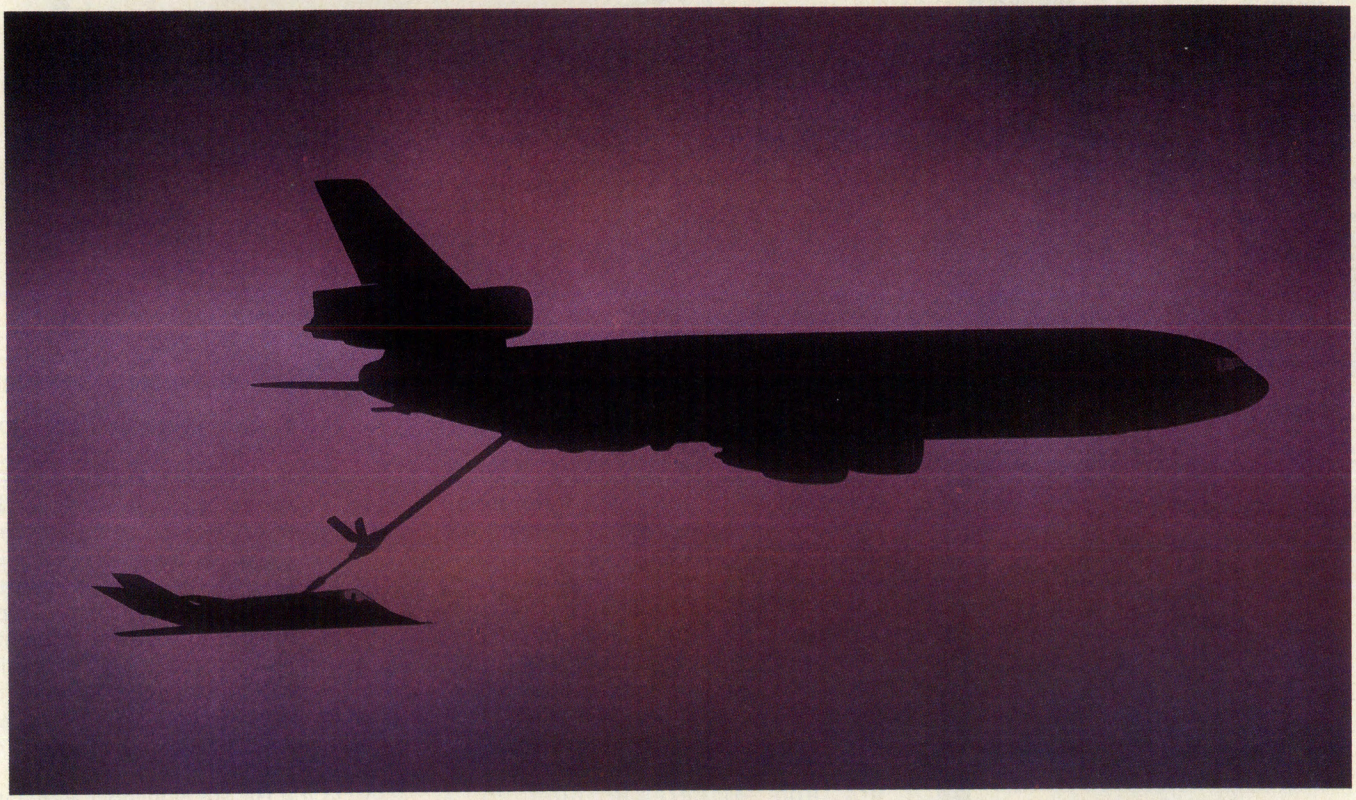

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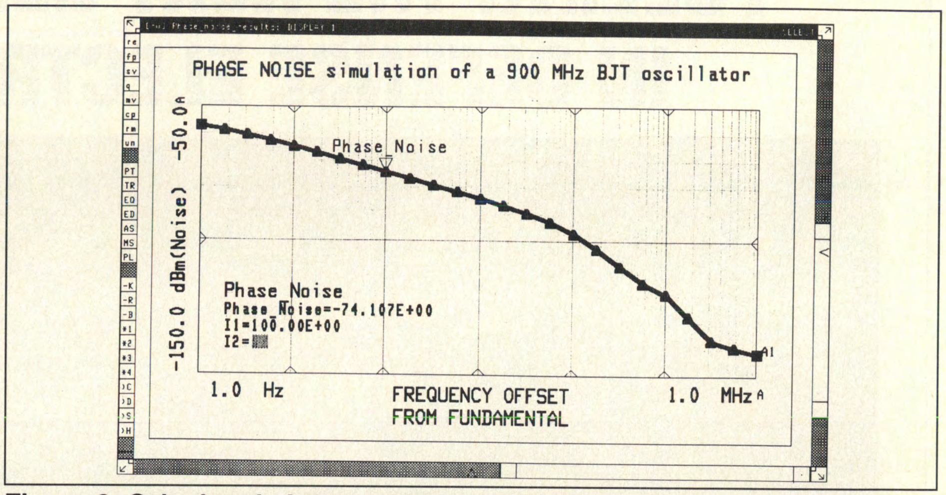


Figure 3. Calculated phase noise for a 900 MHz cellular radio oscillator.

details as they develop), or upward from the transistor level.

Optimization, Monte Carlo/yield analysis, sensitivity analysis, nodal noise analysis, and parameter sweeping are provided with the basic DC, linear, and nonlinear simulation modes. Figure 2 shows a snapshot of the design process within the HP RF Design System for a 900 MHz CT1 digital cordless telephone. In the schematic (upper half of Figure 2), behavioral (i.e., system-level) Chebyshev bandpass filters are connected to subcircuits for Gilbert Cell mixers and a ceramic filter in a stimulus/response test. The response plots in the foreground of Figure 2 show the receiver's response to incoming frequencies from 100 to 2,200 MHz; this is essentially an EMC susceptibility simulation. An "eye" diagram for the Q channel shows demodulated performance given a specific channel condition.

Noise Through Nonlinear Circuits

The HP RF Design System makes a breakthrough with the calculation of frequency-translated noise through nonlinear circuits and systems. This information is available for any node in a circuit or system and is not restricted by topology, subcircuits, or feedback. Noise arises in an RF circuit from many sources, such as thermal noise, flicker (1/f) noise, and noise sources in transistors that depend on dynamic bias point. The HP RF Design System accounts for these noise sources, and also allows voltage and current noise sources to be placed directly in the circuit to model the natural causes of noise, which can often be correlated.

With the HP RF Nonlinear Simulator, it is now possible to quantify the noise figure and IF noise floor of mixers, phase noise in oscillators, and the noise performance of whole systems. Previously, nodal noise information was only available for linear circuits.

Figure 3 shows phase noise as a function of the frequency offset from the fundamental of a 900 MHz SAW oscillator. The large-signal performance of the oscillator (not shown) was determined using the harmonic balance analysis technique. The phase noise result was

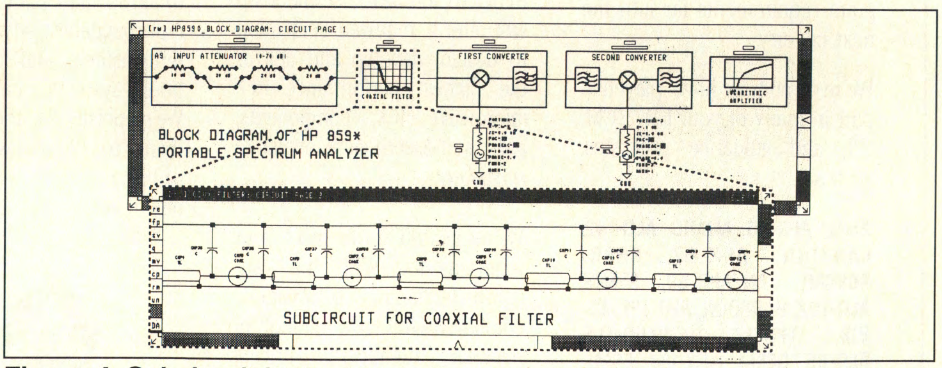


Figure 4. Subcircuit in the block diagram of an HP 8590-series spectrum analyzer, expanded to show its circuit contents.

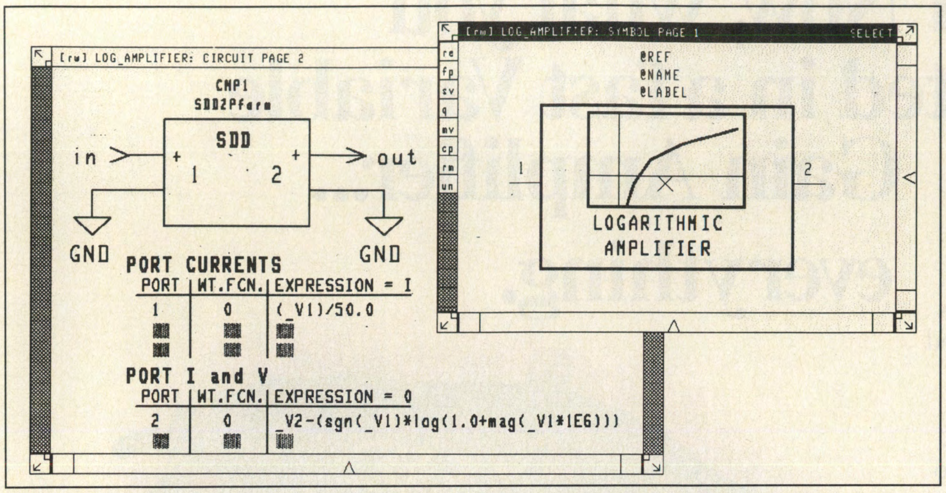


Figure 5. User-defined nonlinear model and schematic symbol for the log amplifier in the upper right corner of Figure 4.

calculated using the results of the large-signal performance, and accounted for all noise power that resulted from both downconverted thermal noise and upconverted 1/f noise. Figure 3 shows that the thermal noise floor is much lower than the upconverted 1/f noise of the transistor itself.

Integrated Personality for System Design

Another contribution of the HP RF Design System is the HP System Model Library, a block-diagram level design personality that extends the capability of the HP RF Nonlinear Simulator.

Because the HP System Model Library is built on the foundation of a powerful, general-purpose circuit simulator, it surpasses the idealized, ad-hoc structure of competing system simulation tools. This strategy also provides a seamless transition between system- and circuit-level design processes because both use the same simulation engine. One advantage is that topology is completely unrestricted. Feedback, reverse isolation, coupling between signal paths, non-ideal sources, subnetworks, user-defined components, and much more are no problem for the HP software. Designers can move beyond idealized, cascaded-only systems into the real world.

The HP System Model Library also allows the designer to combine system-level components with device-level components in the same schematic. Early in a design, the feasibility of the block diagram may be considered. As the design

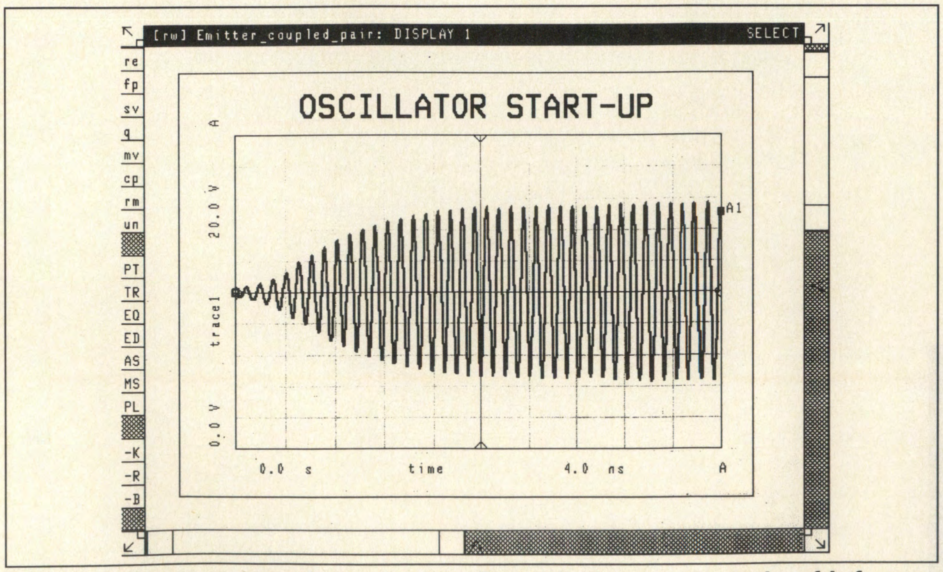
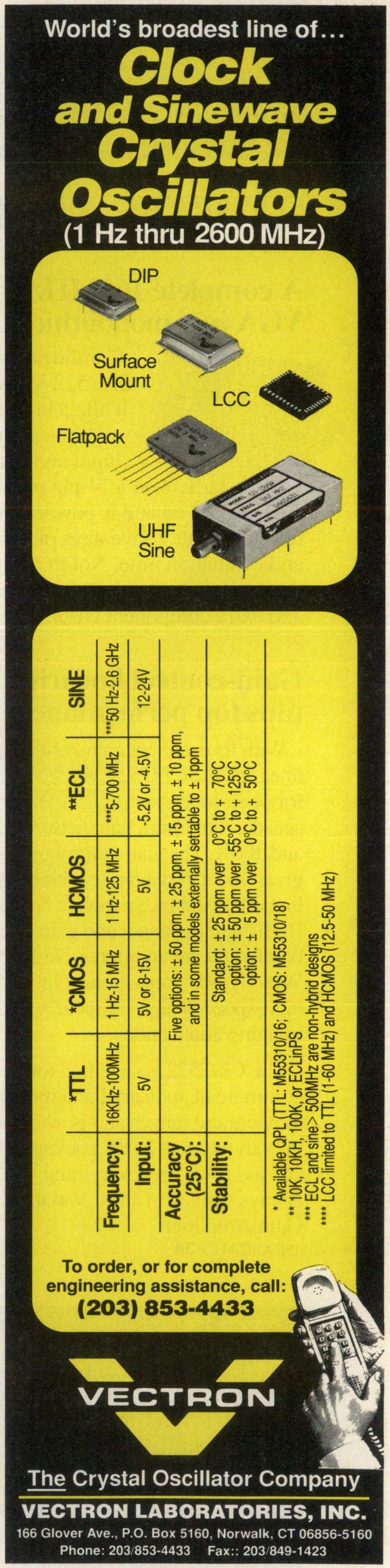


Figure 6. Start-up transient of a low-Q oscillator, an example of information uniquely available from a time-domain simulation.



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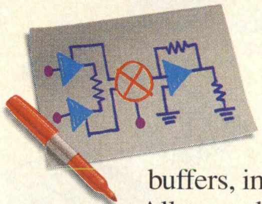

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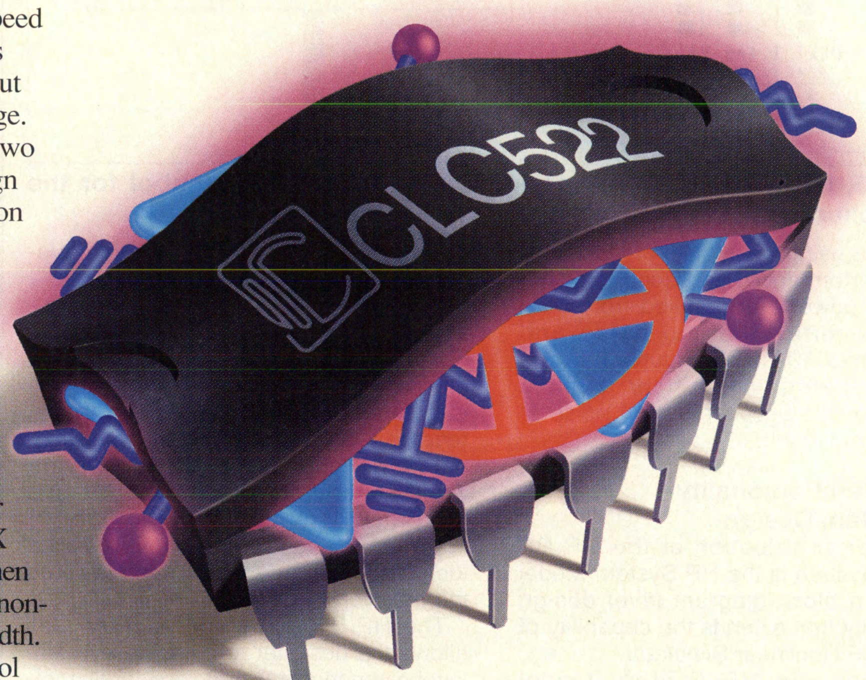

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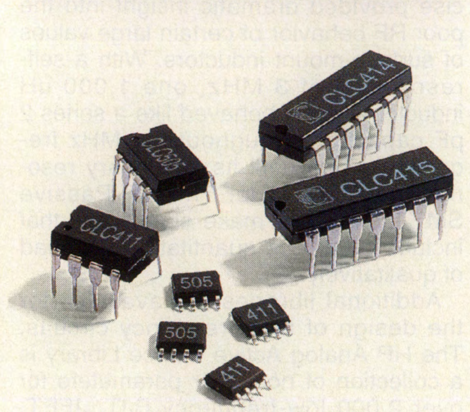

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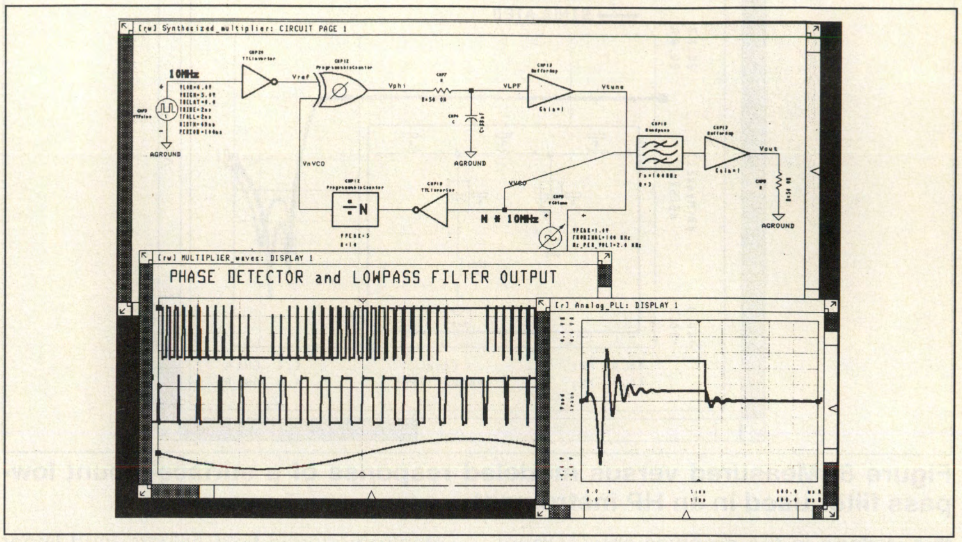


Figure 7. A frequency-multiplying phase-locked loop. Unlike HP Impulse, SPICE would have difficulty simulating the frequency-domain and high-frequency portions of this circuit.

matures, design team members can substitute their completed circuits for idealized equivalents such as amplifiers or filters. The performance of the block diagram can then be verified with a mixture of system-level components and raw circuitry. At all times, a single schematic is used to give the designer a single point of control.

Open System Enhances Modeling

The HP System Model Library is easy for the user to customize. The HP RF Design System provides many facilities for adding components to the system without programming.

The top half of Figure 4 shows part of a block diagram of an HP 8590-series portable spectrum analyzer that has been entered into the HP RF Design System. Each component is actually a subcircuit defined by the creator of the circuit (note the custom schematic symbols).

The lowpass filter is highlighted and expanded to show its contents. An idealized Chebyshev filter specified by order and cutoff frequency could have been used. However, the designer chose to use an equivalent circuit for this precision coaxial lowpass filter. Also shown is a custom schematic symbol that looks like a graph of the filter response which was defined by the designer (see top half of Figure 4).

The equivalent circuit was used because it predicted a secondary passband at a higher frequency that could potentially allow interfering signals to enter the system. This was a deliberate choice of the designer, not one imposed by the simulation tool.

User-Defined Nonlinearities

The designer also modeled the logarithmic amplifier (upper right, Figure 4) with a user-defined nonlinearity. The HP RF Design System has a "symbolically-defined device" (SDD) which allows the designer to describe voltage-current relationships using equations.

Figure 5 shows the details of the logarithmic amplifier model, as well as the schematic symbol that the designer created. Although it is possible to pass parameters into the subcircuit from higher-level schematics, that feature was not used here. Two equations are shown on the device in Figure 5, one for each unknown (i.e., each port). At port 1, the current is simply equal to the voltage at port 1 (_V1) divided by 50. This is equivalent to terminating port 1 in an internal 50 ohm resistor.

The second equation is more complicated. The voltage at port 2, minus an expression involving the logarithm of the voltage at port 1, is set equal to zero. In other words, the output voltage is equal to a function of the log of the input voltage.

It is possible to build sophisticated nonlinear devices using the SDD, such as a complete, customized Gummel-Poon-like transistor model. It is also possible to define math operators and look-up tables (with up to four dimensions of interpolation) for complete generality. Furthermore, it is possible to pass parameters into subcircuits. All of

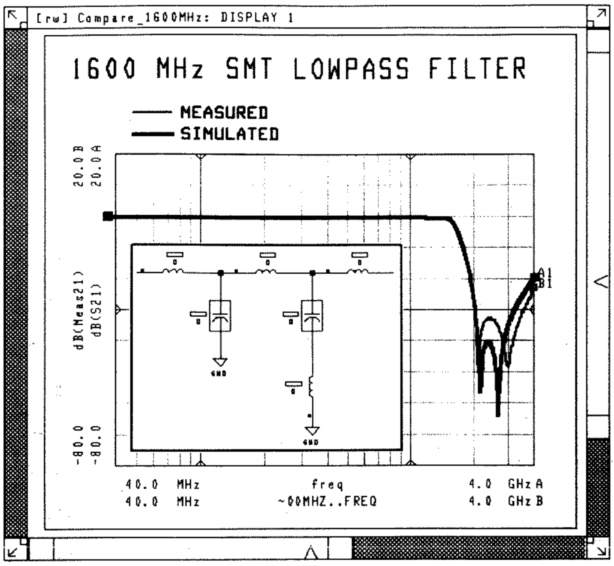


Figure 8. Measured versus modeled response of a surface-mount lowpass filter used in an HP instrument.

this is done in the graphics environment without programming, allowing great potential for custom, user-defined libraries.

The HP Impulse Time-Domain Simulator

The time-domain simulation capabilities of the HP RF Design System represent a significant advance for the high-frequency designer. The new simulator, HP Impulse, overcomes some key limitations of SPICE for high-frequency applications. SPICE has been used successfully for many years to understand the transient behavior of circuits such as the oscillator start-up shown in Figure 6. However, high-frequency designers often must simplify the true engineering problems they face in order to fit them into the assumptions of SPICE.

HP Impulse extends the capabilities of SPICE to include the dispersive effects (frequency-dependent impedance and delay) caused by skin effect and non-TEM transmission structures. Accurate models are also included for distributed microstrip and stripline transmission lines and discontinuities. These models are based on dimensions and other physical constants. Some commercial implementations of SPICE also accept physical dimensions (such as width and length) as input, but internally idealize the true performance with simplified (often lumped) components. HP Impulse retains the full generality and accuracy of all dispersive, distributed models.

HP Impulse also includes the ability to use raw S-parameter data elements directly in nonlinear transient simulations without lumped equivalent networks. Linear S-parameter data is extremely popular because of its simplicity and accuracy. Network analyzers, electromagnetic field solvers, and linear circuit simulators are common sources of S-parameter data.

The simulator is a direct replacement for SPICE for high-frequency designers. Figure 7 shows a frequency-multiplying phase-locked loop circuit. This circuit has both analog and logic components, low and high-frequency effects, and ideal frequency-domain filters and transient behavior, and is likely to have high-frequency parasitics and interconnections.

Accurate Part Libraries

The accurate libraries of commercially available RF parts make the HP RF Design System an invaluable, turnkey system. The HP Packaged BJT Library and the HP RF Passive SMT Library are standard, and several more libraries are available as options.

The HP Packaged BJT Library is a collection of the most popular high-frequency bipolar devices. Gummel-Poon parameters have been extracted to characterize DC and full nonlinear performance to approximately 4,000 MHz.

The HP Passive SMT Library is a collection of approximately 200 commercial surface-mount technology parts, including resistors, capacitors, and inductors. This library includes equivalent circuit models that emulate the true measured response of these devices up to 4,000 MHz. HP has measured and modeled most of the parts quantitatively past their first and second resonances. These model much more than the effective "Q" of the devices.

Hewlett-Packard uses the SMT library with great success in the design of its own test equipment. Figure 8 shows the measured vs. modeled response of a 1,600 MHz lowpass filter using parts from the library (1). This modeling exercise provided dramatic insight into the poor RF behavior of certain large values of surface-mount inductors. With a self-resonance of 3 MHz, one 1,000 uH inductor actually behaved like a series 2 pF capacitor throughout the MHz frequency range until its secondary resonance at 750 MHz. The HP Passive SMT Library can make simulations that include parasitics quantitatively instead of qualitatively.

Additional libraries are available for the design of high-frequency circuits. The HP Analog Active Device Library is a collection of nonlinear parameters for over 2,000 low-frequency BJT, JFET, MOSFET, and diode devices. These are useful for bias circuits and other support circuitry. The HP Murata SMT Capacitor Library is also available for a wide selection of surface-mount capacitor models.

Productivity Improvements

As an RF hardware manufacturer, Hewlett-Packard understands the industry's concerns for bringing leading-edge designs to market in the shortest possible time. HP's dual role as manufacturer and software supplier gives it a strong incentive to provide high-quality, high-performance tools. The development of the HP RF Design System is an extension of HP's 50-year commitment to engineering productivity tools. This manufacturing and design expertise has created new frontiers in simulation and user- interface technology that can now be applied to your design problems.

The HP RF Design System will be available in December, 1992. Configurations start at $28,093 (U.S. list price). The software is supported on Unix-based computer workstations from HP, Apollo, Sun, IBM, DEC, and 386/486 PCs. Readers with technical questions may call Dan Pleasant at Hewlett-Packard, (707) 577-5202. For literature, circle Info/Card #161. ***RF***

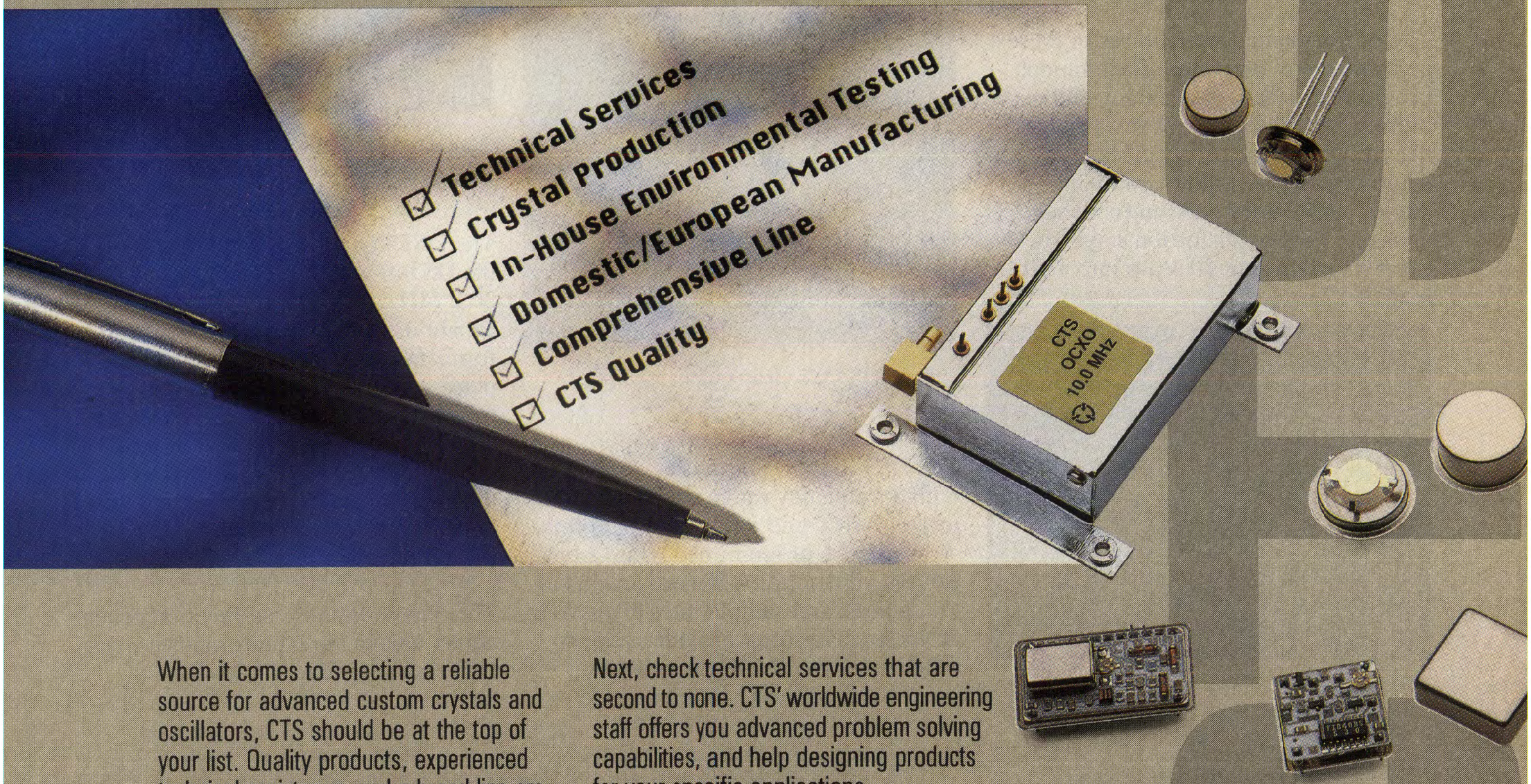
Reference

1. Daren McClearnon, "Modeling Passive Surface-Mount Components," *Proceedings*, RF Expo West, March, 1992.

About the Author

Daren B. McClearnon is an Application Engineer at Hewlett-Packard, 1400 Fountaingrove Parkway, Santa Rosa, CA 95403. He has a BSEE from Case Western Reserve University and has been at Hewlett-Packard since 1985.

If A Standard Frequency Control Product Won't Do, Check With CTS.



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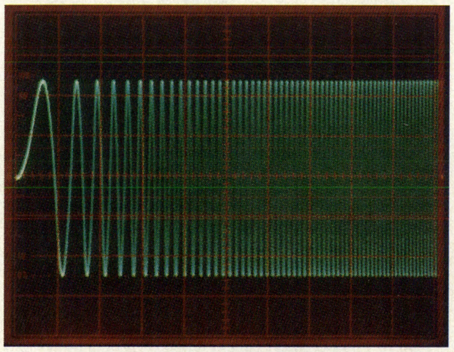
INFO/CARD 40
See us at RFEE '92, Booth #318.

Synthesized Function Generators

Something New

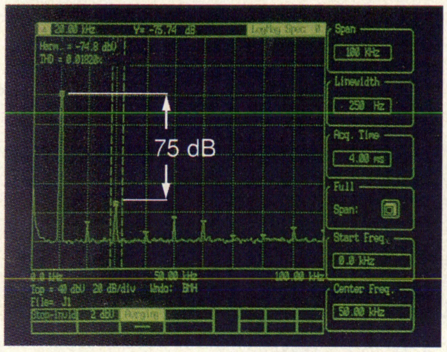
Every so often a powerful concept changes the way things are done. For a long time, function generators have been a jumble of analog circuits – ramp generators, VCOs, mixers, limiters and shapers. The new concept is Direct Digital Synthesis, and function generators will never be the same.

The performance and features of these instruments is unrivaled. Each model provides synthesizer accuracy and resolution, seamless sweeps over their entire frequency range, clean standard waveforms (sine, square, ramp, triangle), and a dirty one too (wideband gaussian noise). Distortion stays low even when driving 10 Vp-p into a 50 Ω load.



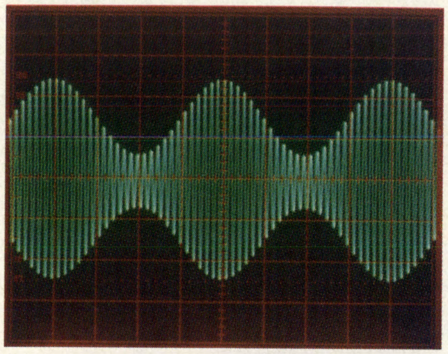
Seamless linear or log sweeps

The DS335's 3.1 MHz frequency range, 1 μHz resolution, and its clean (0.05% THD) and flat (±0.1dB) outputs, establish it as an outstanding value at $995. An optional GPIB/RS-232 interface allows integration into automatic test applications.



DS335's clean output spectra

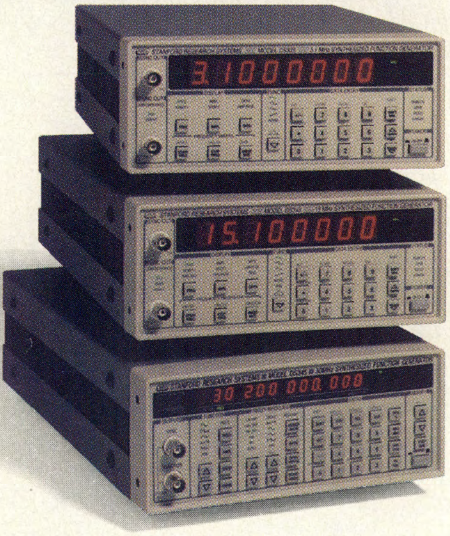
The DS340 is similar to the DS335, with a frequency range which extends to 15.1 MHz, and arbitrary waveforms. Arbs may be programmed with 12 bits of vertical resolution, record lengths to 16k points, and sample rates to 40 MHz. A linear phase filter provides smooth outputs with a 10 MHz bandwidth.



DS345 offers AM, FM, PM and Burst modulation

The DS345 has all of the features of the DS340 with frequencies up to 30.2 MHz, and a rich set of modulation capabilities. Any of the standard waveforms may be amplitude, frequency, or phase modulated by sines, squares, ramps, triangles or arbitrary modulation patterns. Several DS345's may be slaved together via an external clock input, and the phases between their outputs may be adjusted with millidegree resolution.

Three new synthesized function generators. Outstanding performance. Unsurpassed value.



	DS335	**DS340**	**DS345**
Max Freq (Sine/Sq)	3.1 MHz	15.1 MHz	30.2 MHz
Freq Resolution	1μHz	1μHz	1μHz
Standard Timebase	±5 ppm	±5 ppm	±5 ppm
THD (f_o=10 kHz)	< 0.05%	< 0.05%	< 0.10%
Spurs (f_o=1 MHz)	<-65 dBc	<-65 dBc	<-55 dBc
Level Accuracy	±0.1dB	±0.1dB	±0.2dB
Modulation	FSK	FSK	AM,FM,PM, FSK,Burst
Arbitrary Waveforms	none	12 bits to 16k points and 40 Msamples/s	
GPIB/RS232	$395	$495	$495
Price	**$995**	**$1595**	**$2195**

SRS

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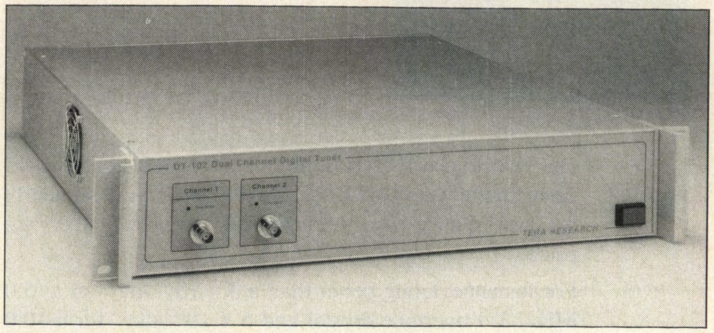
INFO/CARD 41

Dual Channel Digital Tuner

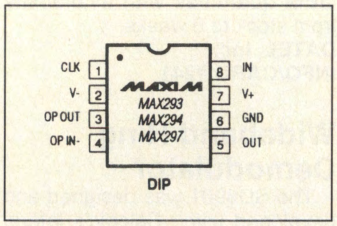
Tera Research's DT-102 dual channel digital tuner is a 64 MSPS all-digital tuner that can be used in a variety of signal processing, analysis and collection systems. The tuner accepts analog or digital inputs, provides complex-valued digital outputs and is remotely controlled. The DT-102 features two 64 MSPS 10 bit A/D converters, two tuner channels and a variety of digital and analog outputs. Each tuner's center frequency is programmable within 0 to 31 MHz with fractional Hz resolution. The two channels are time aligned to better than 1 ns. Filtered tuner outputs are optimally decimated to facilitate processing and storage by downstream signal processing systems. The DT-102 supports a wide range of output filter bandwidths and is controlled from either a RS-232 or IEEE-488 interface.

Tera Research, Inc.
INFO/CARD #250



DIP/SO 8th Order LP Filter

Maxim Integrated Products announces the MAX293/294/297 elliptic, 8th-order, lowpass, switched-capacitor filters. These filters require no external resistors or capacitors and come in 8-pin miniDIP and 16-pin SO packages. The MAX293/MAX297 has stopband/passband transition ratio of 1.5, which yields a steep rolloff of 135 dB/octave. The MAX294 has a transition ratio of



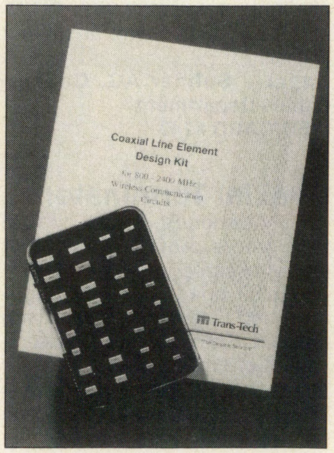
1.2, resulting in an even sharper rolloff of 205 dB/octave. The filter's corner frequency is set by the frequency of a clock signal. The clock to corner frequency ratio for the MAX293/MAX294 is 100:1 with a 0.1 Hz to 25 kHz corner frequency range, and 50:1 for the MAX297 with a 0.1 Hz to 50 kHz corner frequency range. The clock can be externally driven by a CMOS level signal, or an external capacitor can set the device's internal clock frequency. The MAX293/294/297 operate with +5V single or ±5V dual supplies. An uncommitted op-amp is provided to build a continuous-time lowpass filter for post-filtering or anti-aliasing. Prices start at $3.18 (1000pc., FOB USA). Production quantities are available now.

Maxim Integrated Products
INFO/CARD #249

Coaxial Line Element Design Kit

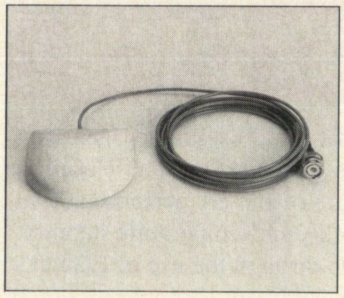
Trans-Tech, Inc., a subsidiary of Alpha Industries, Inc., introduces its Coaxial Line Element Design Kit. The kit contains 32 coaxial line elements of various types to allow design versatility with frequencies selected to cover the majority of wireless communications applications. Trans-Tech's rugged ceramic coaxial line elements exhibit higher Q, superior parallel resonant impedance, and better temperature stability than the classical inductor coils and associated inductor/capacitor lumped element components used in amplifier and oscillator tank circuits. The kit features broad frequency range and allows quick determination of exact frequency requirements. The kit includes standard and low profile units, and complimentary replacements are available.

Trans-Tech, Inc., a subsidiary of Alpha Industries, Inc.
INFO/CARD #248



Miniature Ceramic GPS Antenna

A new miniature ceramic antenna measuring just 2.75 × 2.75 × 0.79 inches (7×7×2 cm) for ground positioning systems (GPS) is now available from Murata Erie North America. Its very low profile and small size make this antenna ideal for automobile and other portable GPS applications. This new antenna system incorporates an integral low noise amplifier (LNA) immediately following the antenna to provide exceptionally low noise figures. The antenna element is a microstrip consisting of a radiating element and a grounding element of thick film sil-



ver. These electrodes are printed on both sides of an extremely high dielectric constant ceramic substrate. Beneath this antenna element is the LNA utilizing proprietary GaAs FETs developed by Murata Erie. A 14 dB return loss is guaranteed over the full 15 MHz bandwidth of this 1575.42 MHz (L1 band) antenna. It boasts a 90 degree Kelvin noise temperature and minimum power at band center of 20 dB.

Murata Erie North America
INFO/CARD #247

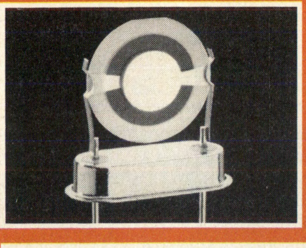
Electrically Conductive Coating

A new air-dry, electrically conductive coating for electroplating, electroforming, shielding, radar cross section modeling, applica-

tion onto tantalum capacitors and circuit board repair is being introduced by Carroll Coatings Company of Providence, RI. C-646 electrically conductive coating can be applied by brush or spray onto nonconductive surfaces and air dries tack free within 20 minutes. Featuring <0.1 ohms/square conductivity, this silver conductive coating produces a very smooth finish and has a service temperature to +250 degrees F. C-646 electrically conductive coating provides 120 cm^2/gram coverage, good abrasion resistance and will act as effective shielding over the 100 MHz to 10 GHz frequency range. It adheres well to primed and un-primed metals, most plastics, various waxes, fabric and leather, and can be soldered using low-temperature solder. Carroll Coatings C-646 is priced from $0.38 per gram; packaged in 25, 50, 100, 250 and 500 gram containers.

Carroll Coatings Company
INFO/CARD #246

Our WORLD CLASS CRYSTALS STAY STABLE



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INFO/CARD 42

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INFO/CARD 120

See us at RFEE '92, Booth #917.

CABLES & CONNECTORS

Low Noise Coaxial Cables

A semiconductive layer placed between the dielectric and shield of Harbour Industries' new series of coaxial cables counteracts triboelectric effects developed when cables bend, twist or undergo other motions. The cables are available in impedances of 50, 75 and 95 ohms with an outer diameter of 0.075 to 0.148 inches. Capacitance is maintained at low levels through careful selection of materials, and attenuation is minimal.

Harbour Industries
INFO/CARD #245

SMA Connector

Andrew Corporation announces the availability of the 41ASWS (SMA male) and 41ASNS (SMA female bulkhead) connectors for use on 0.25 inch HELIAX® cables. These connectors employ a unique spring collet compression design that allows attachment in less than three minutes while providing high retention against pull-off. A pair of connectors has maximum VSWR of 1.55:1 in the .045 to 19.5 GHz band.

Andrew Corporation
INFO/CARD #244

Easy to Bend Semi-Rigid Cable

Huber + Suhner AG, distributor for EZ-Form Cable Corporation, introduces a new generation of aluminum type semi-rigid cable. The cables allow very tight bending radii due to the soft aluminum outer conductor. The cables are available in sizes 0.086, 0.141 and 0.250 inches with or without tin plating.

Huber + Suhner AG, Coaxial Cables Department
INFO/CARD #243

Coaxial Terminators

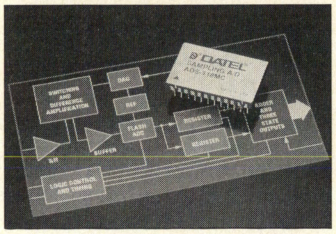
ITT Cannon/Sealectro coaxial terminators provide a low-cost means of cable junction to a printed circuit board where engagement and disengagement are not required. Styles are available for a variety of RG series cable types and cables of similar dimensions. The terminators are two piece parts and do not require that the braid be transferred down to the board.

ITT Cannon/Sealectro
INFO/CARD #242

SEMICONDUCTORS

Small, Low-Power ADC

DATEL's new 12-bit, 5 MHz sampling A/D converters include a sample-and-hold amplifier, an external reference and clock, all on a 24-pin DDIP. The ADS-118



offers three state outputs while the ADS-118A features direct adjustment of offset and gain errors. These devices operate from ±15 Volts and +5 Volts with typical power dissipation of 1.8 Watts. Pricing starts at $203 in OEM quantities with availability from stock to 6 weeks.

DATEL, Inc.
INFO/CARD #241

Wideband Ring Demodulator

The SD8901 was designed and developed using Calogic's lateral DMOS process. The SD8901 is a direct replacement for the Siliconix SI8901 and has maximum drain current of 50 mA, R_{DS}(on) of 50 ohms and third order intercept of +35 dB at 250 MHz. The device is available in TO-78 and SO14 packaging as well as in die form. Available in 100 piece minimum quantities, it sells for $3.15, $2.86 for quantities in the 1000s.

Calogic Corporation
INFO/CARD #240

Power Amplifiers for Handheld Cellular

Mitsubishi's MGF7100 series and the FA01312 are power amplifiers for use in handheld cellular telephones. The MGF7100 series operates in the 900 MHz band and provide 31.5 dBm typical output power with 60 percent efficiency. Samples in a surface mount package are available now

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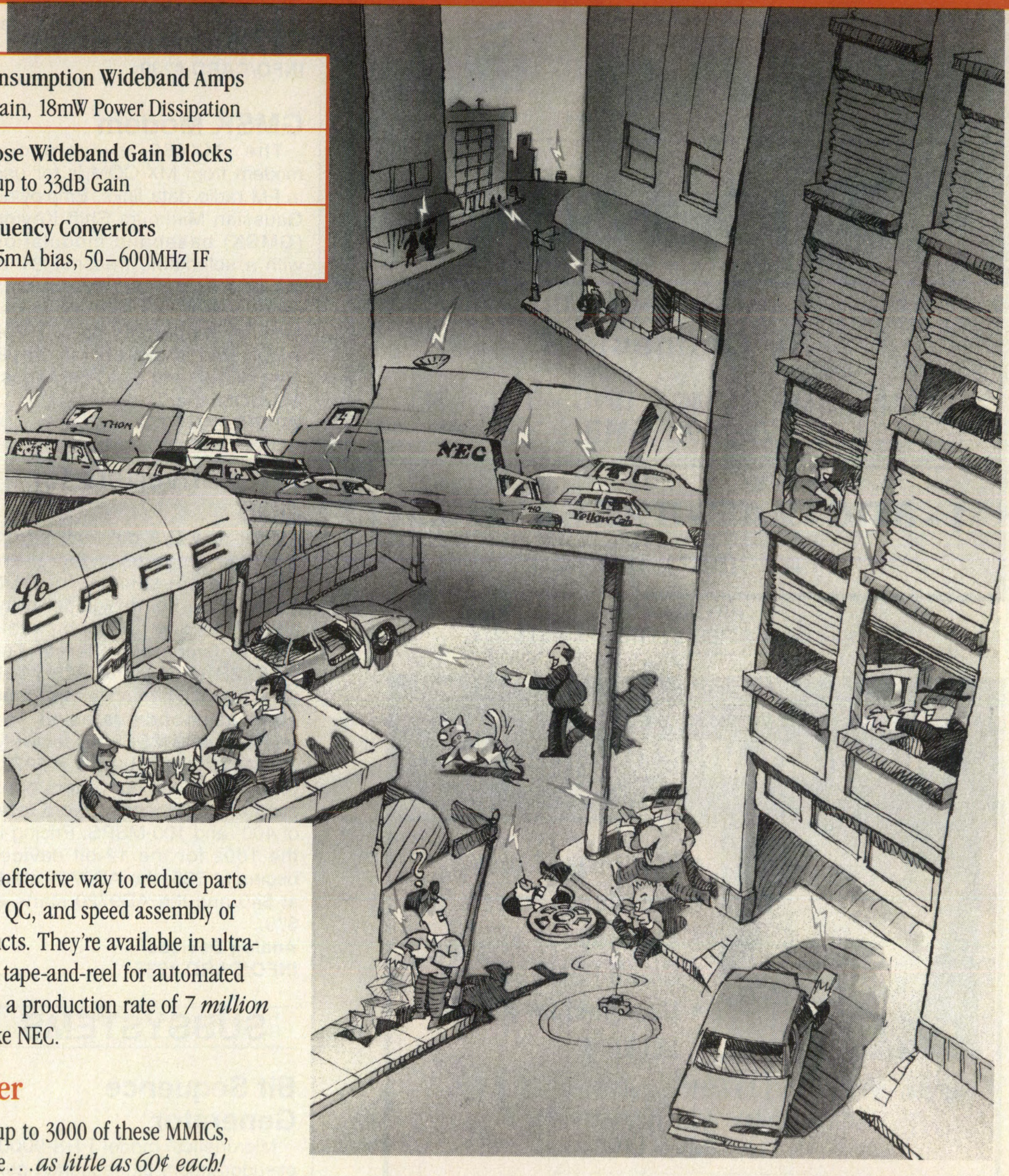
2GHz Operation, 5V/35mA bias, 50–600MHz IF



Our new Silicon MMICs are a cost-effective way to reduce parts count, improve reliability, simplify QC, and speed assembly of the circuits for your wireless products. They're available in ultra-small surface mount packages, on tape-and-reel for automated assembly, and they're proven. With a production rate of *7 million a month,* no one knows MMICs like NEC.

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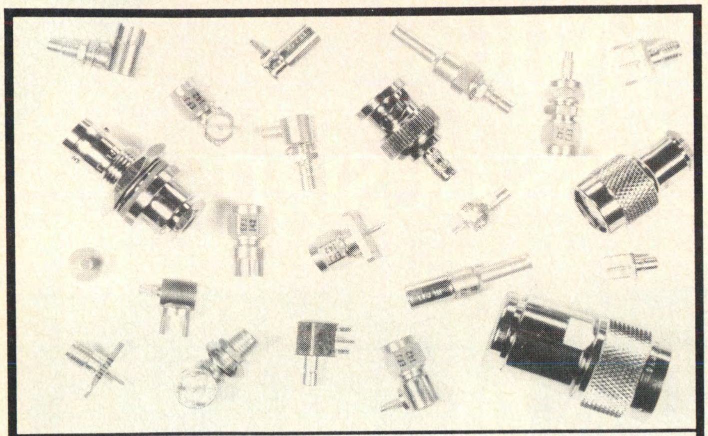

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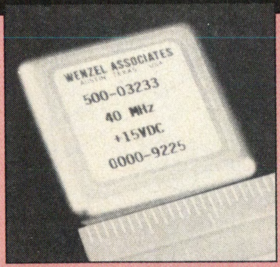
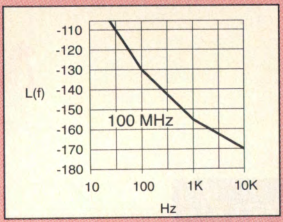
See us at RFEE '92, Booths #211, 213.

Miniature VHF Oscillators



- Low Phase Noise
- Excellent Temperature Stability
- Small Package
- Designed for Phase Locking

Frequency: 12.5 to 130 MHz Package: 1.5x1.5x0.5"
Temp Stability: $\pm 3 \times 10^{-7}$, 0°to+50°C
Phase Noise L(f):



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INFO/CARD 45

for evaluation. The FA01312 operates in the 824 to 849 MHz band and switches between Class-AB and Class-C modes. The FA01312 will be available in the third quarter of 1992.
Mitsubishi Electronics America, Inc. Electronic Device Group
INFO/CARD #239

GMSK Modem

The MX489 synchronous modem from MX-COM is for use in FM radio data links. Employing Gaussian Minimum Shift Keying (GMSK) baseband modulation with a selectable transmit BT of 0.3 or 0.5, it is capable of data rates of 4000 to 19200 bits per second. It operates on a 5 Volt supply and is available in 24-pin ceramic DIP and SOIC packages.
MX-COM, Inc.
INFO/CARD #238

D/A Converters for DDS

Four new D/A converters from Analog Devices, Inc. are designed specifically for DDS applications. For design flexibility, both ECL- and TTL-compatible input versions are offered. The AD912B (ECL) and AD9713B (TTL) are 12-bit converters operating at encode rates of 100 MSPS and 80 MSPS respectively. The AD9720 (ECL) and AD9721 (TTL) are 10-bit converters operating at corresponding word rates of 400- and 100-MSPS. Pricing in the 100s for the 12-bit devices begins at $35; the AD9721 starts at $40 and the AD9720 starts at $79.
Analog Devices
INFO/CARD #237

SUBSYSTEMS

Bit Sequence Generator

Micronetics' model PRS-90000 pseudorandom bit sequence generator (PRBS) is designed for spread spectrum application. The unit offers sequences of 2^8, 2^{10} or 2^{15}. Version A provides TTL output at rates up to 20 Mbits/s and version B provides ECL output at rates up to 100 Mbits/s.
Micronetics
INFO/CARD #236

Radio Modem Series

A line of radio modems from DATARADIO features the ability to interface with popular E.F. Johnson LTR® trunked radio systems to support data transmissions at speeds up to 4800 bps. The line is an enhanced version of the VIS (Vehicular Information Series) mobile data product line. DATARADIO will make available a standard cable to connect between the LTR trunking radio and modem for a "plug and play" solution.
Dataradio Corporation
INFO/CARD #235

SS Generator/SS Demodulator

SIGTEK is now shipping the ST-101 direct sequence spread spectrum signal generator and the ST-102 programmable spread spectrum demodulator for the PC/XT-AT bus. The boards feature DDS and programmable BPSK and QPSK spreading sequences. Software using pull-down menus is included with each board. The ST-101 costs $2250, and the ST-102 costs $3800. Delivery is stock to six weeks.
Sigtek, Inc.
INFO/CARD #234

2.8 oz Double Downconverter

Systron Donner Microwave has developed a double downconverter with a signal gain of 45 dB (min.) that weighs only 2.8 ounces. Frequency conversions are available from 2 to 26.5 GHz in the 2.362 × 1.575 × 0.395 inch unit. Other features include +30 dBm (max.) input signal; 60 dB (min.) spurious response; 60 dB (min.) image rejection; and an AM/FM noise figure of 140 dBc at 700 kHz.
Systron Donner, Microwave Division
INFO/CARD #233

Low Profile HF Antenna

A new HF monitor antenna from Maxim Technologies provides high dynamic range over the 1 to 60 MHz band. The MA-701 has an adjustable whip mounted atop a trans-impedance amplifier. Power is supplied via the antenna's coaxial cable, but a DC isolator prevents DC current from entering the receiver. The MA-107 comes complete with ground stake, stand, and collapsible and fixed whip elements for installation flexibility.
Maxim Technologies, Inc.
INFO/CARD #232

New Lower Prescaler Prices for Acronym-Happy Engineers

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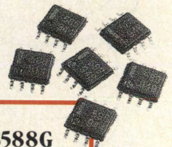

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UPB584G	UPB585G	UPB586G	UPB587G*	UPB588G
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500 MHz to 2.5 GHz	500 MHz to 2.5 GHz	500 MHz to 2.5 GHz	50 MHz to 1.0 GHz	500 MHz to 2.5 GHz

Prices based on volume run rates

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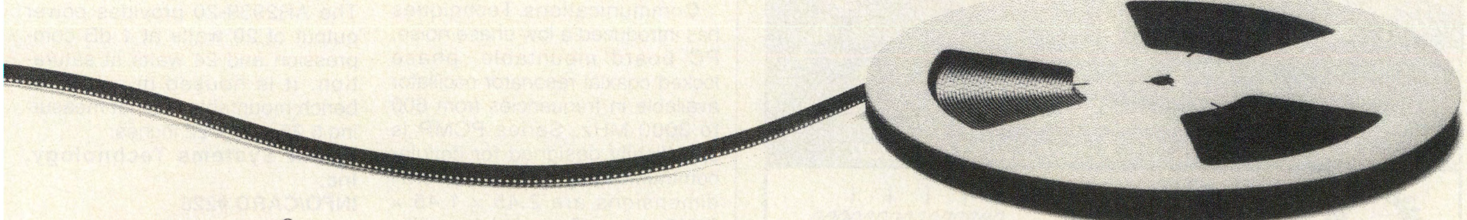
For a free **Product Selection Guide**, call your nearest CEL Sales Office or circle the number below.



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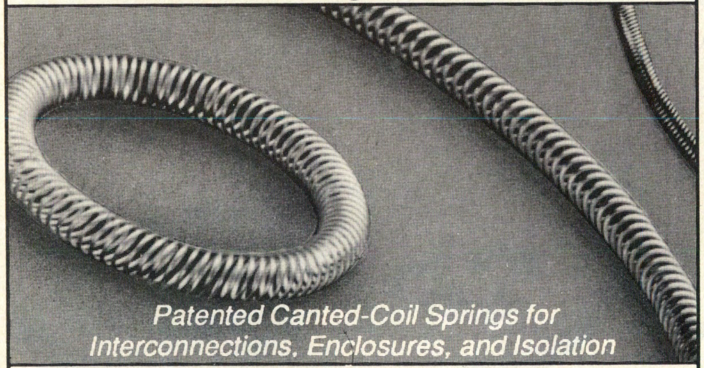
INFO/CARD 47
See us at RFEE '92, Booth #811.

EMI Spring Gaskets



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U.S. patents: 4,655,462; 4,934,666.

INFO/CARD 48
See us at RFEE '92, Booth #914.

SIGNAL SOURCES

Programmable Synthesizer

The model PLL158P-77 synthesizer from RF Prototype Systems can be easily programmed from an IBM PC via an RS-232 interface. The synthesizer covers 500 to 1000 MHz in 10 kHz steps and has on board DIP switches to set the frequency. Output power is +10 dBm min., with flatness of ± 1.5 dB. Spurious is less than – 60 dBc, and SSB phase noise at 100 kHz is less than –117 dBc/Hz.
RF Prototype Systems
INFO/CARD #231

Space Qualified Oscillators

The FTS 9300 series of ultra stable oven-controlled crystal oscillators are designed for spacecraft application in the 4 to 200 MHz range. The series offers fixed tuned, voltage-controlled and phase-locked oscillators. The SMT units meet grade 1, MIL-STD-975 standards. SC-cut crystal resonators are incorporated, providing long term aging (5×10^{-11}/day) and a phase noise floor of –160 dBc.
Frequency and Time Systems, Inc.
INFO/CARD #230

Coaxial Resonator Oscillator

Communications Techniques has introduced a low phase noise, PC board mountable, phase locked coaxial resonator oscillator available in frequencies from 600 to 3000 MHz. Series PCMP is specifically designed for cellular communication systems. Overall dimensions are 2.45 × 1.45 × 0.440 inches. Typical phase noise for a 925 MHz units is –103 dBc at 1 kHz offset and –133 dBc at 100 kHz offset.
Communications Techniques, Inc.
INFO/CARD #229

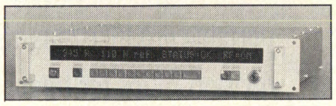
Four Channel Synthesizer

Guide Technology's GT310 frequency synthesizer is a PC/AT/386 plug-in board containing four completely independent PLL sources. Each channel covers the range of 360 kHz to 120 MHz, with one channel covering the extended range of 0.0024 Hz to 120 MHz. The GT310 includes a DOS-based software package which provides instrument-like control and a driver for programming in Microsoft C or QuickBASIC. The GT310 is available from stock and sells for $495.
Guide Technology, Inc.
INFO/CARD #228

AMPLIFIERS

High Efficiency RF Power Generator

Dressler's CESAR RF power generator employs a class-E amplifying scheme to reach an efficiency of up to 90 percent. The high efficiency design of these



500, 600, 1000 and 2000 Watt supplies reduces power consumption and size. The units are water cooled. A RS-232 interface is a standard feature, and an IEEE-488 interface is optional.
Dressler HF Technik GmbH
INFO/CARD #227

Class A, 20 Watt Amplifier

Power Systems Technology announces the introduction of its model AR2939-20 solid state RF amplifier operating over the 2000 to 3000 MHz frequency range. The AR2939-20 provides power output of 20 watts at 1 dB compression and 24 watts at saturation. It is housed in a rack or bench mountable cabinet measuring 5.25 x 19 x 22 inches.
Power Systems Technology, Inc.
INFO/CARD #226

SIGNAL PROCESSING COMPONENTS

Multiplier

MITEQ's model MAXZ0026 provides times twenty multiplication to output frequencies from 2.9 to 3.2 GHz, with 0 dBm minimum output power. The device includes interstage amplification and filtering designed for optimum spurious performance: 65 dBc typical

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UPC1653 To 1300MHz 18dB G_P	**UPC1654** To 1100MHz 19dB G_P	**UPC1655** To 900MHz 18dB G_P	**UPC1656** To 850MHz 19dB G_P	**UPC1658** To 1100MHz 17dB G_P 2.0dB NF	
UPC1659 600MHz to 2300MHz 23dB G_P	**UPC1675** To 2100MHz 12dB G_P	**UPC1676** To 1300MHz 20dB G_P 4.0dB NF	**UPC1677** To 1700MHz 24dB G_P P_{out} = 19.5dBm	**UPC1678** Up to 1900MHz 23dB G_P P_{out} = 18dBm	**UPC1688** Up to 1000MHz 21dB G_P 4.0dB NF

Prescalers – From $2.20 each

UPB581 ÷ 2 500MHz to 2.8GHz	**UPB582** ÷4 500MHz to 2.8GHz	**UPB584** ÷2 500MHz to 2.5GHz	**UPB585** ÷ 4 500MHz to 2.5GHz	**UPB586** ÷ 512/256 500MHz to 2.5GHz	**UPB587** ÷ 2/4/8 50MHz to 1.0GHz V_{CC}=2.2 to 3.5V	**UPB588** ÷64/128 500MHz to 2.5GHz

Freq. Convertors – From $1.58 each

UPC1685 DC to 890MHz 14dB GAIN Double Balanced Mixer Applications	**UPC1686** DC to 890MHz 22dB GAIN	**UPC1687** DC to 890MHz 28dB GAIN I_{CC} of 38mA

IF Amplifiers – From $3.50 each

UPC1668 10 to 170MHz G_L=14.5dB 60dB Isolation	**UPC1669** 10 to 180MHz G_L=10.5dB 55dB Isolation	**UPC1670** 10 to 150MHz IM_3 of 56dBc 60dB Isolation

Transistor Arrays From $2.40 each

UPA101 F_T=9GHz Double Balanced Mixer Applications	**UPA102** F_T=9GHz Differential Amplifier
UPA103 F_T=9GHz Differential Amplifier	**UPA104** F_T=9GHz OR/NOR Functions

Special Function MMICs From $1.08 each

UPC1684 LED DRIVER 150mA Drive Current 300 Mbits NRZ	**UPC1663** VIDEO AMPLIFIER 170MHz @ A_{VOL}=100 1.6ns Propagation Delay

Note: MMIC prices based on 25K quantities

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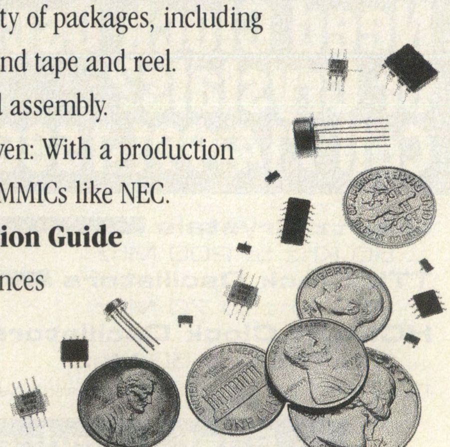
Our **Silicon MMIC Product Selection Guide** lists specifications for dozens of parts. Chances are good it has just what you need. To get a copy, call your nearest CEL Sales Office or circle the number below.



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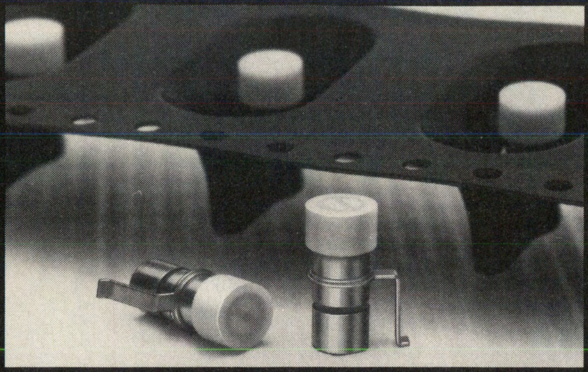
INFO/CARD 49

New 47000 Series
Commercial Giga-Trim® Variable Capacitors



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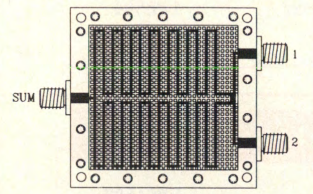
rejection in the region of 2 to 4 GHz and 55 dBc elsewhere. MAXZ0026 is 4.5 × 1.5 × 0.5 inches without SMA hardware.
Miteq
INFO/CARD #225

WAFFLELINE™ Circuits

Lorch Electronics will use WAFFLELINE to build designs that traditionally use stripline. Developed and patented by Harris Corporation, WAFFLELINE is a system whereby wires are routed through channels in a conductive slab, resulting in higher isolation than stripline construction. WAFFLELINE interconnects are equivalent to a semi-rigid coaxial cable to 20 GHz.



400 MHz. shown (Internal wiring)

Lorch Electronics, Division of Vernitron Corp.
INFO/CARD #224

EMI Caps with Transient Protection

Based on the European approved DLT range of feedthrough filter capacitors, the Oxley TVS filter range incorporates a unique transient-voltage suppression element. For example, the DLT/10000/5/TVS combines a 10 nF capacitor with a 5.6 V suppressor element capable of withstanding 500 applications of a 150 A 8/20 us waveform with maximum clamping voltage of 15.5 V.
Oxley, Inc.
INFO/CARD #223

Broadband Power Splitters

These resistive power splitters from Lucas Weinschel provide an output whose source impedance is essentially matched to 50 ohms. They operate from DC to 18.0 GHz and have a maximum SWR of 1.30. Insertion loss is nominally 6 dB between input and either output, and the maximum input power is one watt average, 1 kW peak. Model 1870A uses all female Type-N connectors while model 1872A uses a female Type-N at its input and GPC-7 connectors for outputs.
Lucas Aerospace, Communications & Electronics
INFO/CARD #222

Coaxial Feedthrough Terminations

Elcom Systems announces the availability of a series of coaxial feedthrough terminations. They are available in 50, 75 and 93 ohms impedance, in BNC, TNC, N and SMA male to female connectors, with gold, silver or nickel plating. Model FT-50 operates from DC to 1 GHz, FT-75 from DC to 500 MHz and FT-90 from DC to 150 MHz. Prices start at $17.50 each in small quantities; delivery is from stock to 30 days ARO.
Elcom Systems, Inc.
INFO/CARD #221

Frequency Multipliers

KW Microwave introduces ×13 and ×20 multipliers to its multiplier product line. The ×20 multiplier has an input frequency of 500 Mhz with output at 10.00 GHz. Input power is +17 dBm while output power is +10 dBm. Spectral purity is –50 dBc. The ×13 multiplier has an input frequency range of 660 - 690 MHz and output range of 7.58 - 8.97 GHz. Input power is +10 dBm with output power of +20 dBm. Spectral purity is –65 dBc.
KW Microwave Corporation
INFO/CARD #220

MANUFACTURING, TOOLS AND MATERIALS

Grounding Spring for Medical Electronics

A unique coil spring used in implantable devices as a conductor and a holding mechanism is now available from Bal Seal Engineering. The spring acts as a contact between an electrode and power supply. Advantages include a concentrated load distributed around the periphery of the electrode or lead. The spring load remains relatively constant over a large deflection. The springs can be made with 0.015 inch coil heights and an inside diameter as small as 0.010 inch.
Bal Seal Engineering Co., Inc.
INFO/CARD #219

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INFO/CARD 53

MOBILE RADIO
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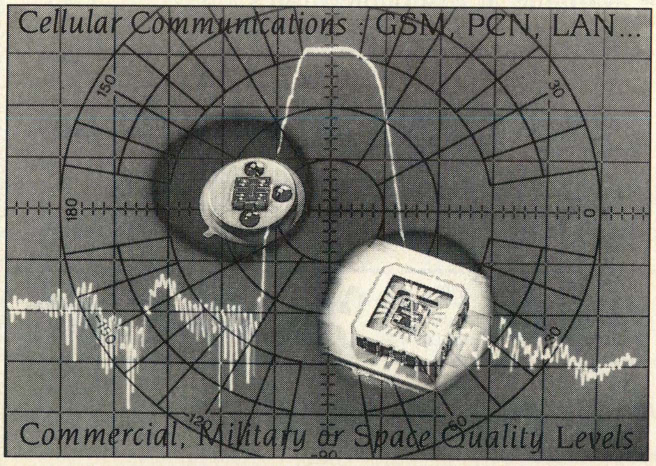

Center frequency:	40 to 1500 MHz
Fractional Bandwidth:	0,2 to 10%
Insertion Loss:	<10 dB

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INFO/CARD 54

See us at RFEE '92, Booth #912.

Circuit Board Cleaner

International Specialty Products' MICROPURE® CDF is an ozone-safe, semi-aqueous CFC alternative. Based on ISP's N-methyl-2-pyrrolidone chemistry, the solvent has high flash point, low vapor pressure and can be used in commercially available cleaning equipment. The solvent earned cleaning efficiency ratings of "Better Than" when compared to CFC-113/methanol.
International Specialty Products, Inc.
INFO/CARD #218

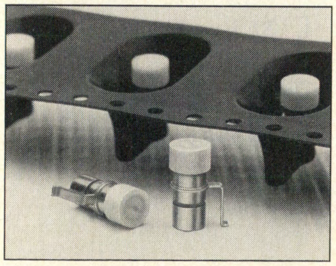
Absorber Tiles

Ferrite absorber tiles measuring 100 × 100 × 6.3 mm for use in anechoic chambers are available from Fair-Rite. The tiles are made of a newly developed nickel zinc ferrite, optimized for noise absorption over the lower end of the EMI spectrum. A bulletin containing specifications and applications is also available.
Fair-Rite Products Corp.
INFO/CARD #217

DISCRETE COMPONENTS

Commercial Trimmer Capacitors

Johanson Manufacturing announces its 47000 series trimmer capacitors. These Giga-Trim capacitors are extremely small



multi-turn trimmers which have been engineered to reduce costs; for example, a ceramic dielectric replaces the traditional sapphire dielectric with no loss in Q or performance range. Selected mounting styles are available on tape and reel for SMD applications. Cost is $6.00 in the 1000s. Delivery is in 6 - 8 weeks ARO.
Johanson Manufacturing Corporation
INFO/CARD #216

High Current Surface Mount Inductors

A new line of high current surface mount inductors from American Precision Industries have maximum current rates of 1640 to 260 mA depending on inductance value. The inductances range from 1.0 uH to 100 uH with a standard tolerance of 10 percent. The line is called series #2512 and has a ferrite core and comes in a molded package with pretinned leads. The inductors cost approximately $0.599 per 1000 pieces with delivery of 4-6 weeks ARO.
American Precision Industries Electronic Components Group, SMD Division
INFO/CARD #215

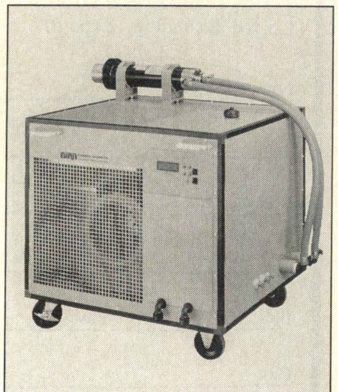
Common Mode EMI Suppression Inductors

MagneTek now offers kits for the new standard line of common mode EMI suppression inductors. Two kits of E-core construction inductors and one kit of toroidal construction inductors are available. The inductors are designed for frequencies from 35 to 150 kHz and may be purchased in production quantities.
MagneTek
INFO/CARD #214

TEST EQUIPMENT

RF Load Resistor Series

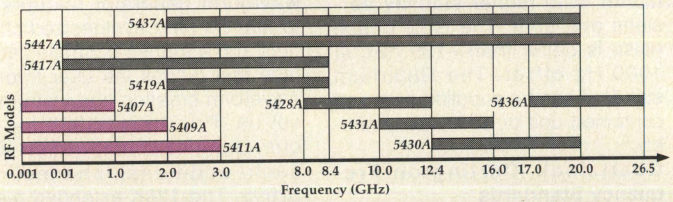
Bird Electronic Corporation announces a series of MODULOAD® RF load resistors for the broadcast equipment market. The first unit in the series is rated at 15 kW. The cooling system uses



The Complete Scalar System With Source and Savings Built-In.

Introducing RF Versions of Wiltron's 5400A Scalar Measurement System

Three RF models provide frequency coverage from 1 to 3000 MHz. The 5400A provides a total system for measuring transmission loss or gain, return loss and RF power. The 5400A provides synthesized sweeper accuracy, ease of use, and 71 dB dynamic range - in a single integrated package - for less than the cost of an ordinary scalar analyzer and sweep generator combination.



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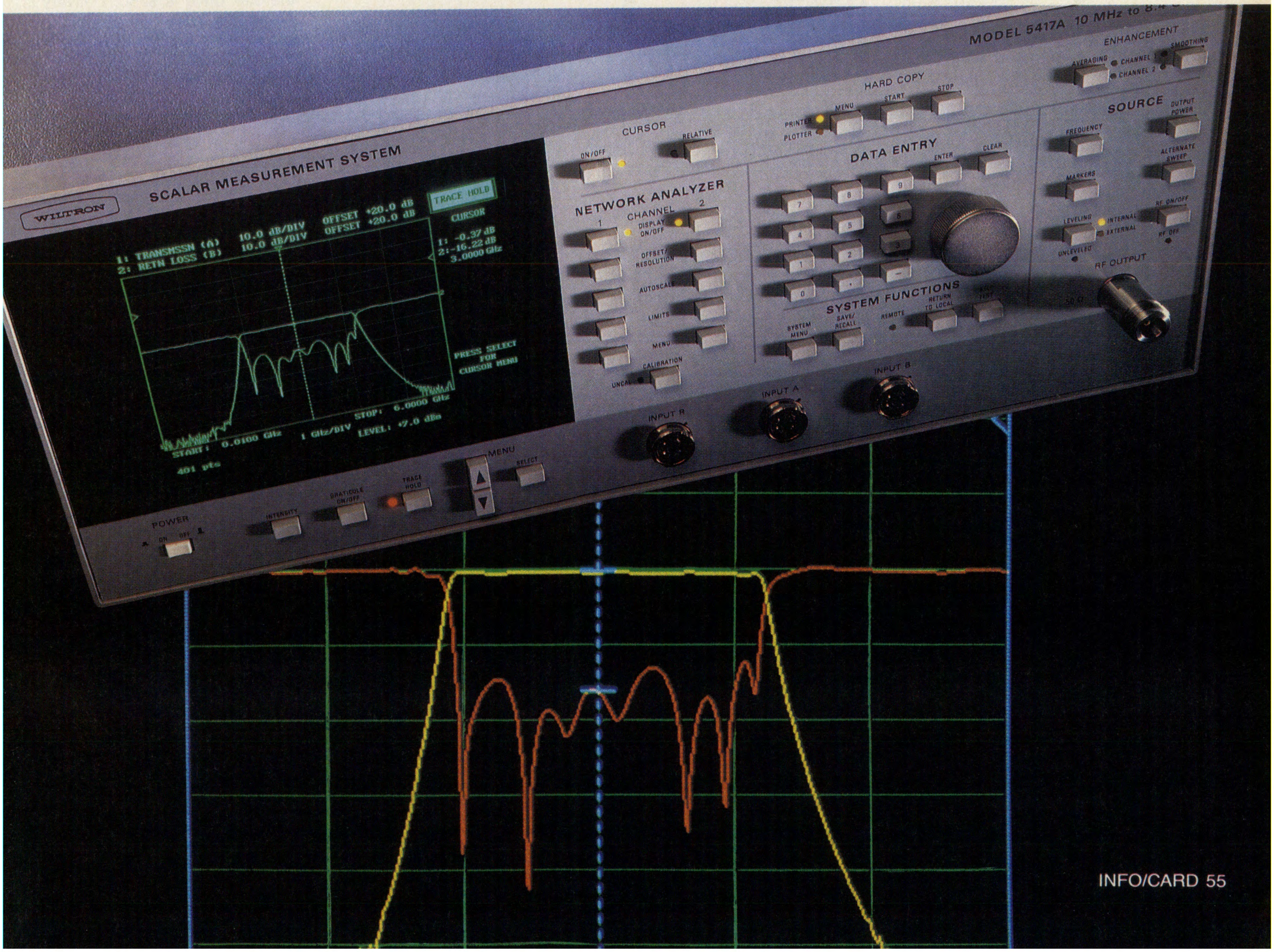

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both air and fluid flow, and the load resistor is protected from cooling failure through coolant flow and temperature interlocks. The load is mounted externally and can be remotely located.

Bird Electronic Corporation
INFO/CARD #213

Rubidium Atomic Frequency Standard

The Quartzlock model 10A-01 is a highly accurate frequency and time calibrator with true portability. Frequency stability is typically better than 4 parts in 10^{-11} per month, with higher stability versions available. The SSB phase noise is better than –135 dBc at 1000 Hz offset. The Rubidium standard can be supplied as a self contained unit or as a subassembly.

Quartzlock, Dartington Frequency Standards
INFO/CARD #212

Frequency Difference Meter

The Tremetrics 527E frequency difference meter can determine the fractional frequency difference between two stable oscillators at a sensitivity of one part in 10^{11}. The 527E may be used to match two oscillators to the same frequency, determine or adjust the offset between two oscillators or measure short or long-range stability.

Tremetrics
INFO/CARD #211

Arbitrary Waveform Generator

Wavetek's model 295 arbitrary waveform generator features up to four 50 MHz synthesized channels and a mouse-controlled interface that guides the user through waveform creation and instrument set-up. Pricing is $5995 for the one-channel configuration with each additional channel for $2995. The 128K extended memory option is $795/channel and the disk drive is $250. Delivery is 4-6 weeks ARO.

Wavetek Corporation, Instruments Division
INFO/CARD #210

O F F T H E S H E L F

K Connectors™

M/A-COM Omni Spectra now offers K Connectors from stock.

M/A-COM Omni Spectra introduces the new OS-2.9 connector family which has been designed for applications requiring superior performance up to 46 GHz and mechanical compatibility with SMA, 3.5mm and K Connectors. Millimeter wave performance of the connectors is achieved through the special 2.92mm outer conductor line size and air dielectric interface. The OS-2.9 features a .032 inch outer conductor, guaranteeing reliability and repeatability during mating by offering superior resistance to overtorquing. OS-2.9 compatibility with SMA and 3.5mm connectors eliminates the need for adapters that change connector type or sex to complete your system.

Rely on M/A-COM Omni Spectra, the industry leader in RF, microwave and millimeter wave coaxial connectors. Call today to receive our new OS-2.9 Interconnect Products Brochure.

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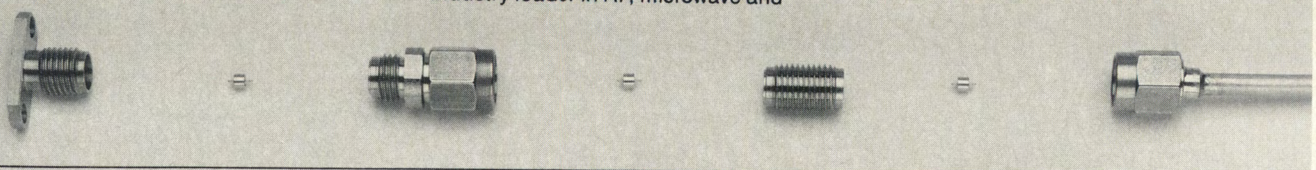

M/A-COM
OMNI SPECTRA

INFO/CARD 57
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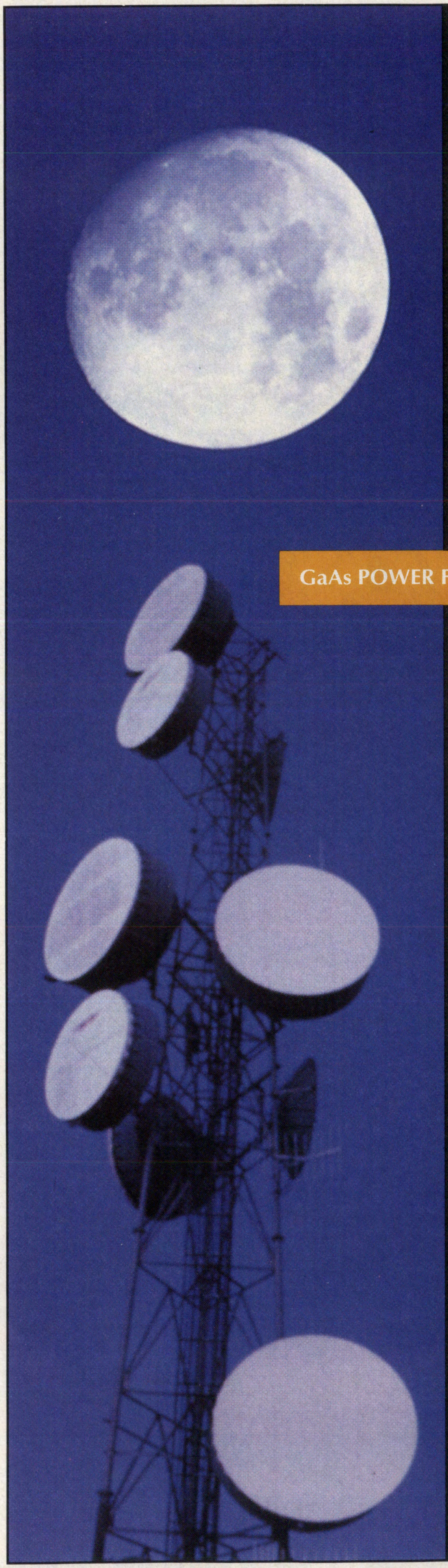
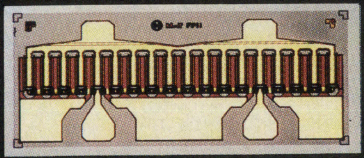
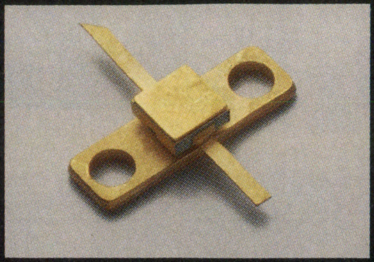


Whether the application is wireless communications, cable distribution, satellite uplink or microwave radio, Microwave Technology has the GaAs FET devices for efficient gain and power from 900 MHz to 23 GHz. A full line of small-signal devices are available to support complete amplifier designs for telecommunications transmitters and receivers.

GaAs POWER FOR TELECOMMUNICATIONS

MwT offers a complete line of low-noise and power GaAs FET devices, hybrid MIC gain modules, and complete RF and microwave amplifiers. Our applications engineers are ready to offer a solution to your telecommunications application.





1 WATT AT 5 VOLTS BIAS

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- Power outputs to +31 dBm
- Third order intercepts >12 dB above P1dB
- Quarter micron gold gates
- Power added efficiencies >30%
- 5 volt capability
- Low phase noise
- Available as chips or in hermetic packages

DEVICE	GATE WIDTH (mm)	RECOMMENDED FREQUENCY RANGE (GHz)[1]	TYPICAL POWER OUTPUT (dBm) 10% BANDWIDTH	AVAILABILITY
MwT-2	0.63	10 to 24	+24[2]	Chip, Hermetic Package, Module
MwT-8	1.2	0.5 to 18	+27[3]	Chip, Hermetic Package
MwT-9	0.75	0.5 to 18	+25[3]	Chip, Hermetic Package
MwT-11	2.4	0.5 to 15	+30[3]	Chip, Hermetic Package
MwT-12	0.9	10 to 24	+27[2]	Chip, Module
MwT-13	1.2	10 to 20	+29[2]	Chip, Module

1 - Frequency range limited to 14 GHz for packaged devices 2 - In balanced circuit 3 - In single-ended circuit

MwT

MICROWAVE TECHNOLOGY

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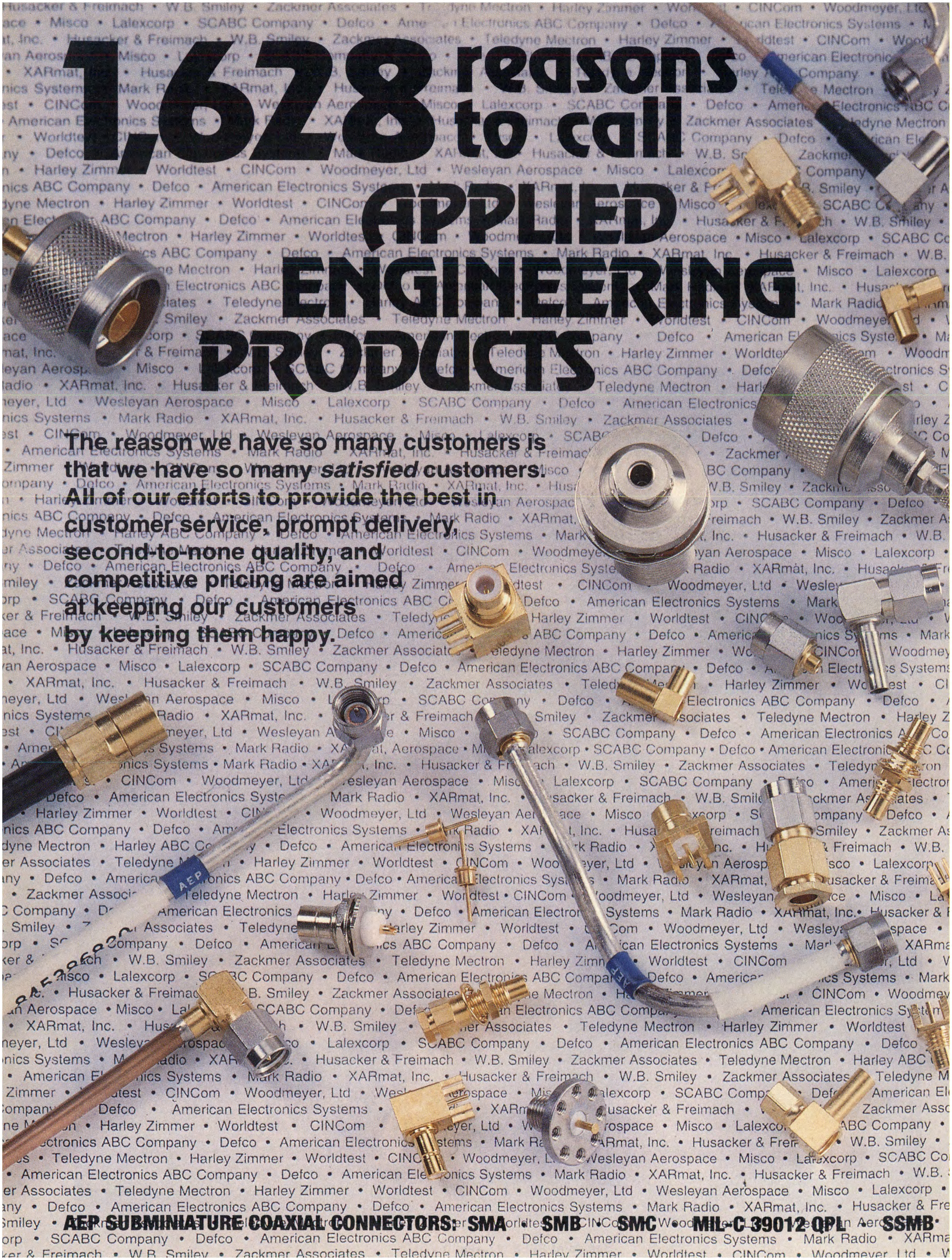
INFO/CARD 58

See us at RFEE '92, Booth #317



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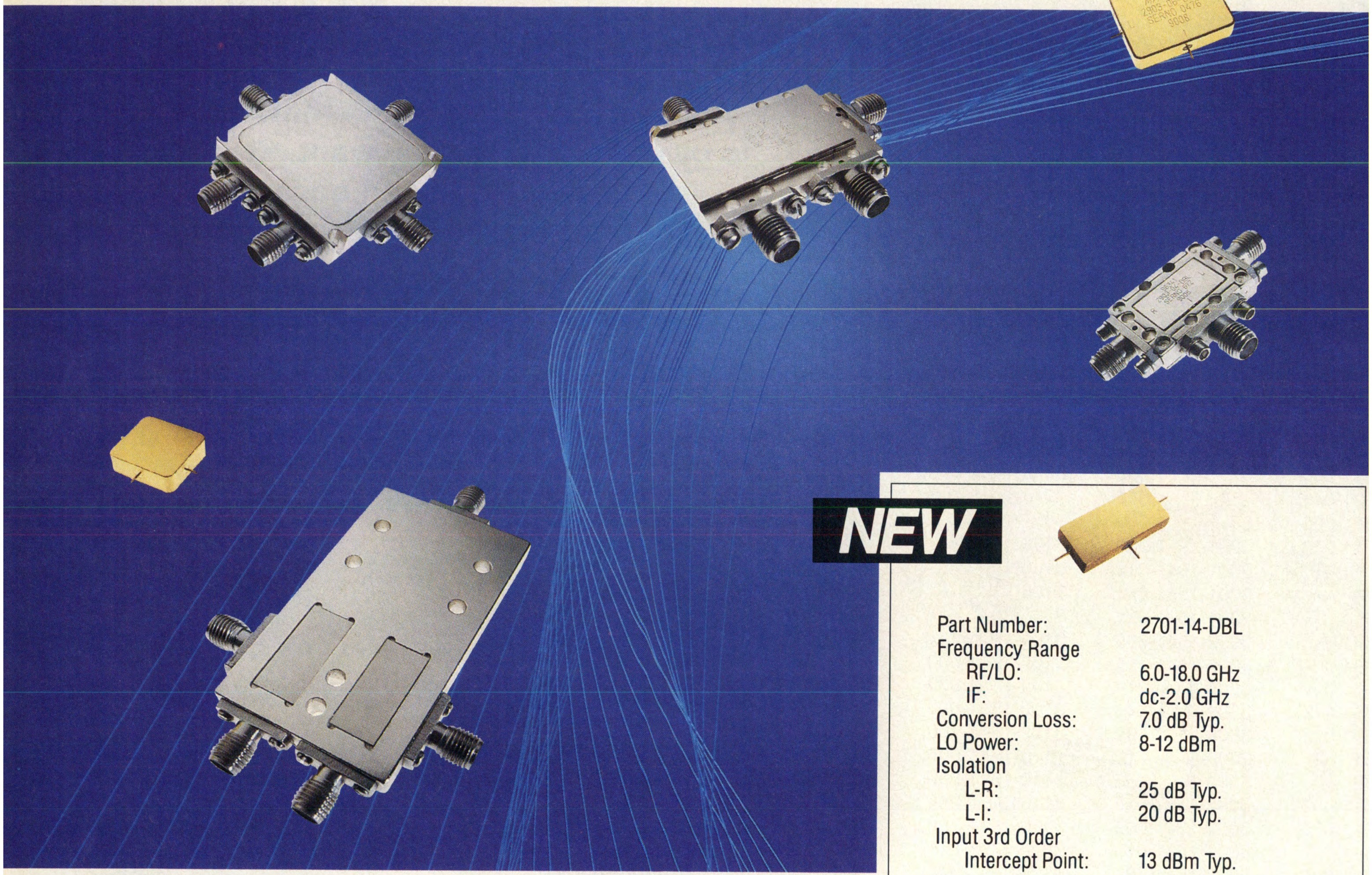

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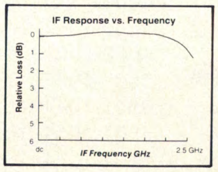

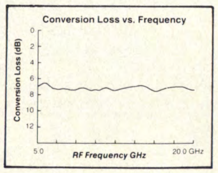

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RF books

Transmission Line Transformers, 2nd Edition

By Jerry Sevick
Published by The American Radio Relay League, Inc., 1990
List Price $20.00

Every ten or more years, a book is published which changes forever the way something is done. If you have not yet had the opportunity of experiencing this infrequent occurrence, you can do so with a recently published book, *Transmission Line Transformers*, 2nd edition, by Dr. Jerry Sevick.

Those familiar with broad-band transformers and their design are no doubt familiar with the frequent references to C.L. Ruthroff's classic paper, "Some Broad-Band Transformers," which appeared in the *IRE Proceedings*, Vol. 47, August 1959, pp. 1337 - 1342. Because of the wide distribution of the *IRE Proceedings*, Ruthroff's design procedures quickly became the standard throughout the industry. However, many years previously, G. Guanella published his paper, "Novel Matching Systems for High Frequencies," in Volume 31, September 1944 of the *Brown-Boveri Review*. Unfortunately, because of the relatively limited distribution in the U.S. of the *Brown-Boveri Review*, Guanella's superior transformer designs never got the exposure that Ruthroff's designs received. However, the improved procedures discussed in Sevick's 2nd edition of *Transmission Line Transformers* will no doubt now become the preferred method for the design and construction of this type of transformer.

As Sevick explains in the preface of his 2nd edition, Ruthroff's approach to the design of wideband transformers was to sum a direct voltage with a delayed voltage which traversed a single transmission line. "By connecting the transmission line such that a negative potential gradient existed, the transformer became a balun; with a positive gradient, it became an unun (unbalanced-to-unbalanced transformer)." Even with transmission lines having optimized characteristic impedances, Ruthroff's transformers had a built-in high-frequency cut-off. In comparison, Guanella's approach was "... to connect transmission lines in a parallel-series arrangement such that in-phase voltages were summed at the high-impedance side. The transmission lines were appropriately coiled to prevent the unwanted (transformer) current." Guanella's short but important statement, "... a frequency independent transformation ...," which appeared in his paper, had been overlooked by Sevick (and probably by others) and that is the reason for the scarcity of Guanella's designs in the first edition of Sevick's book. Sevick corrects this oversight in the second edition where he includes detailed explanations of how the Ruthroff and Guanella design approaches to broad-band transformers are distinctly different and how each approach has its specific application in the field of transmission line transformers.

Contents

Transmission Line Transformers consists of fifteen chapters beginning with the analysis of broad-band, impedance-matching transformers (Chapter 1) and concluding with an 8-page summary (Chapter 14) of important investigative results observed by Sevick during his many experiments. Chapter 15 is a listing of thirty-six references spanning the years from 1944 to 1989. There is no index.

Chapters two through five discuss the low- and high-frequency characteristics of broad-band transformers and the transformer parameters for low- and high-impedance applications. Chapters 6 through 8 discuss unbalanced-to-unbalanced transformer designs with impedance ratios of 1:4, less than 1:4 and greater than 1:4. Chapters 9 and 10 discuss baluns and multimatch transformers. Chapters 11 and 13 discuss core materials and "how-to-do" procedures for selecting the proper ferrites, winding rod and toroidal transformers, constructing low-impedance coaxial cable and handling and taking care of ferrite transformers.

Chapter 12 provides design details of the test equipment Sevick used in making measurements on the many different transformers he constructed. Sevick explains that " ... this chapter is directed to the person who does not have access to sophisticated test equipment and must rely on simple equipment which can be constructed from readily available parts." The instrumentation discussed includes a simple resistive Wheatstone bridge with a current amplifier for measuring characteristic impedance, different types of signal generators, and a discussion of how to measure transformer efficiency and the characteristic impedance of short transmission lines used in the winding of transformers.

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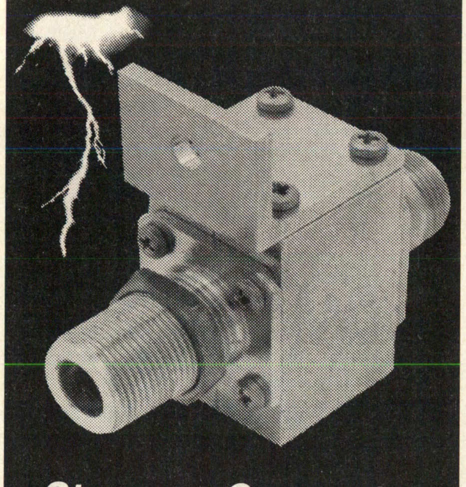

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The extensive effort expended by Sevick in constructing a countless number of broad-band transformer designs and then making hundreds of measurements to obtain data for the plotting of the many frequency response curves is most impressive. For example, Sevick explains that the techniques described evolved over many years of winding hundreds of transformers and it usually took about three attempts in order to arrive at the final design. The transformer constructions are clearly defined both schematically and photographically, and many plots of transformer loss vs. frequency are included. There are fifty-nine photographs of transformers with each photo illustrating between one and four different designs. The photos have good contrast (an improvement over the first edition) which is important in clearly showing the construction details.

Minor errors this reviewer noted in Sevick's book were inconsequential. For example, an editorial decision to use a vertical bar "I" instead of a lowercase script "l" (as was done in the first edition) resulted in the numeral "1" being incorrectly substituted for the vertical bar in the first line on page 7-5. In comparison, on page 7-22 (following equation 7-7), the statement "When l = L ..." is correctly shown. This demonstrates that the lower-case script "l" is preferable to prevent confusing it with the numeral "1." On page 7-31, the text following equation 7-9, "... N = the number of trifilar turns ..." should be "... quadrifilar turns", and the text on page 13-6 "... and hence much shorter windings then their rod counterparts" should be "... than their rod counterparts." The address listing in Table 11-3 for Amidon Associates is their old address.

Unlike most publishers, the ARRL provides a "feedback" form at the back of this book for readers to offer comments and to include any suggested changes or corrections.

The qualifications of the author, Jerry Sevick, are most impressive. He has a BS from Wayne State University in education and a PhD in applied physics from Harvard. In 1956, he joined the staff at AT&T Bell Laboratories in Murray Hill, New Jersey, where he became a supervisor in groups working on high-frequency transistor and integrated-circuit development. He later served as Director of Technical Relations. Although having retired from Bell Labs in 1984, he remains active in many areas, and this book is just one example of his activities. It is unusual that technical authors are as experienced as Dr. Sevick in both the theoretical and practical aspects of their discipline.

If you have any interest in the design or construction of broad-band transformers, this is a book you should have in either your personal or company library. For information, call the American Radio Relay League at (203) 666-1540, or circle **Info/Card #131**

Reviewed by: E.E. Wetherhold, Signal Analysis Center, ALLIANT TECHSYSTEMS INC. P.O. Box 391, Annapolis MD 21404 (410) 266-1769

Spectrum and Network Measurements

By Robert A. Witte
Published by Prentice-Hall, 1991
274 pages

This book is about the theory and practice of frequency domain measurements. It is intended for engineers who use spectrum analyzers and network analyzers to measure and characterize the performance of circuits operating up to about 1 GHz. The basic principles can be applied to higher frequencies, but the examples given are most closely related to lower frequency applications.

Substantial attention is given to the theoretical basis for various measurements in order to help the reader understand what the measurement is trying to achieve. The theory is then related to the use of these instruments, with further explanation of how instrument performance affects the measurement.

Chapter titles reflect this approach: Introduction to Spectrum and Network Measurements, Decibels, Fourier Theory, Fast Fourier Transform Analysis, Swept Spectrum Analyzers, Modulation Measurements, Distortion Measurements, Noise and Noise Measurements, Pulse Measurements, Averaging and Filtering, Transmission Lines, Measurements Connections, Two-Port Networks, Network Analyzers, Transmission Measurements, Reflection Measurements, and Analyzer Performance and Specifications.

This book is recommended for any engineer desiring a fundamental theoretical and practical understanding of instruments for the frequency domain, and the measurements they make.

Author Robert Witte is an engineer at Hewlett-Packard's Colorado Springs Division. For more information, circle **Info/Card #130.** *RF*

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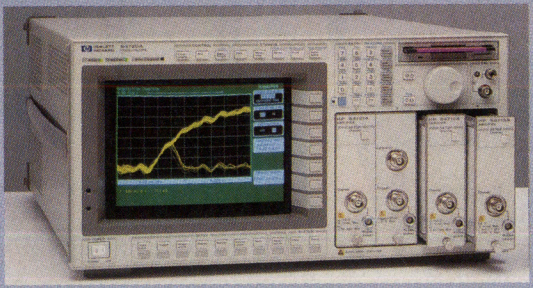

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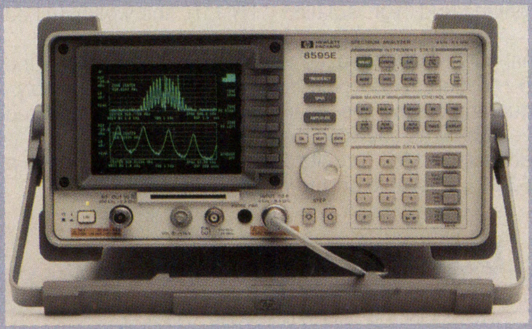

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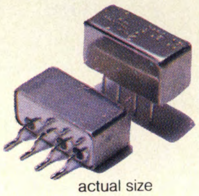

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TUF-5H	17		7.5	0.17	50	13.95
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■ $\bar{X}$ = Average conversion loss at upper end of midband ($f_U/2$)
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Distortion Measurements Using a Spectrum Analyzer

By Robert A. Witte
Hewlett-Packard Company

Many of the circuits that are used in electronic systems are considered to be linear. This means that for a sinusoidal input, the output will also be sinusoidal with perhaps a different amplitude and phase. In the time domain, the user expects to see an output waveform that has the exact same shape as the input waveform. In the frequency domain, we expect to see at the output the same frequency that was at the input (and only that frequency). Any other frequencies that are generated due to the input signal are considered distortion.

Most of the distortion mechanisms measured with spectrum analyzers are low level. That is, the devices producing the distortion are mostly linear and have only a slight nonlinear behavior. Such a weakly nonlinear system can be modeled with a power series.

$$V_{out} = k_0 + k_1 V_{in} + k_2 V_{in}^2 + k_3 V_{in}^3 + k_4 V_{in}^4 + \cdots \quad (1)$$

The first coefficient, k_0, represents the DC offset in the system. The second coefficient, k_1, is the gain of the circuit associated with linear circuit theory. The remaining coefficients, k_2 and above, represent the nonlinear behavior of the circuit. If the circuit were completely linear, all of the coefficients except k_1 would be zero.

The model can be simplified by ignoring the terms that come after the k_3 term. For gradual nonlinearities, the size of k_n decreases rapidly as n gets larger. For many applications the reduced model is sufficient, since the second-order and third-order effects dominate. (Expansion of the model to higher order is discussed later.)

$$V_{out} = k_0 + k_1 V_{in} + k_2 V_{in}^2 + k_3 V_{in}^3 \quad (2)$$

Single-Tone Input

The simplest distortion test of a system is to input a pure sinusoid and measure the frequency content of the output signal:

$$V_{in} = A \cos \omega t \quad (3)$$

The angular frequency, $\omega = 2\pi f$, where f is the frequency in hertz.

Inserting this into the distortion model gives:

$$V_{out} = k_0 + k_1 A \cos \omega t + k_2 A^2 \cos^2 \omega t + k_3 A^3 \cos^3 \omega t \quad (4)$$

$$V_{out} = k_0 + k_1 A \cos \omega t + (k_2 A^2/2)\cdot(1 + \cos 2\omega t) + k_3 A^3 (3/4 \cos \omega t + 1/4 \cos 3\omega t) \quad (5)$$

Collecting terms,

$$V_{out} = k_0 + k_2 A^2/2 + (k_1 A + 3k_3 A^3/4)\cdot \cos \omega t + (k_2 A^2/2) \cos 2\omega t + (k_3 A^3/4) \cos 3\omega t \quad (6)$$

This leaves us with an output voltage containing a DC component, the original (fundamental) frequency, and its second and third harmonics. Had we used a higher-order model, the analysis would have shown even high-order harmonics present at the output. Note that the fundamental amplitude is affected by the nonlinear third-order coefficient of the model, k_3. Similarly, the DC component of the equation is affected by the second-order coefficient. The fundamental is mostly proportional to A, the second harmonic is proportional to A^2, and the third harmonic is proportional to A^3.

The model is somewhat limited since

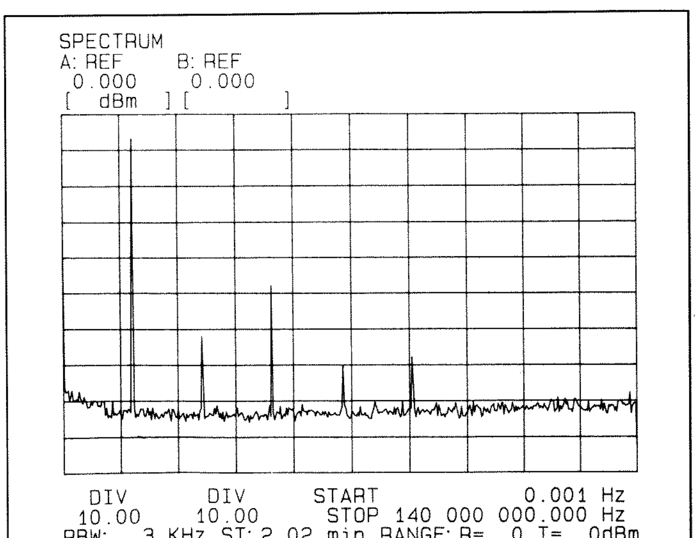

Figure 1. Measurement of harmonic distortion in a signal.

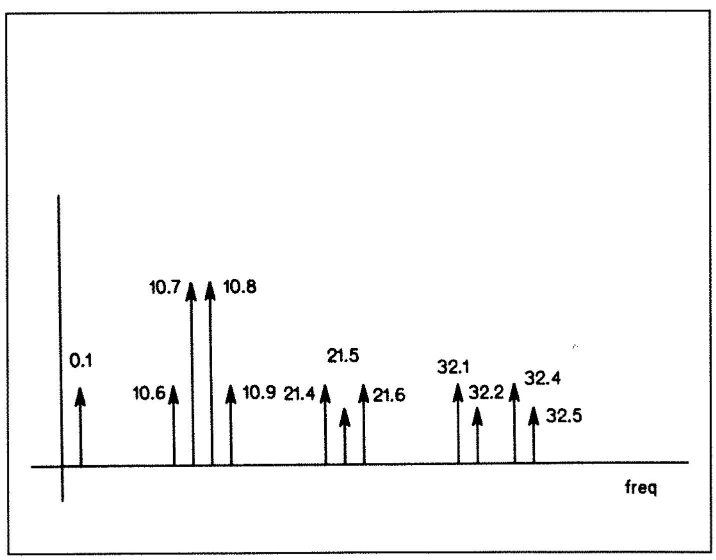

Figure 2. The spectrum a two-tone signal with third-order distortion products.

we do not usually know the values of k_0, k_1, k_2 and k_3 for a particular device. However, we can infer some useful information from the model anyway. Consider what happens when the signal level, A, is reduced. The fundamental will be reduced almost in direct proportion to the signal amplitude. We might say that the fundamental is reduced 1 dB per dB change in the signal level. The second harmonic will go down as the square of A, or converting to dB,

$$20 \log (A^2) = 2(20 \log A) = 2\ A_{dB} \tag{7}$$

This means that the second harmonic will be changed 2 dB per dB of signal level change. Similarly, the third term has an amplitude proportional to A^3. Converting to dB,

$$20 \log (A^3) = 3(20 \log A) = 3\ A_{dB} \tag{8}$$

which means that the third harmonic will be reduced by 3 dB per dB of signal level reduction.

Figure 1 shows the spectrum of a typical signal having harmonic distortion. (Ideally, a pure sine wave would have no harmonics.) Note that the odd harmonics, particularly the third harmonic, is larger than the even harmonics. Distortion that maintains the 50 percent duty cycle of the ideal waveform will create only odd harmonics. Distortion mechanisms that upset the symmetry of the signal produce even harmonics.

When using good-quality spectrum analyzers, one must get accustomed to the fact that there are very few pure sine waves. For example, a good signal or function generator may have a third harmonic that is 30 or 40 dB lower than the fundamental. When viewed on an oscilloscope, this signal will appear to be a pure sine wave since the distortion is not discernible. When measured with even a modest performance spectrum analyzer, the harmonics will be easily visible. This illustrates the advantage of a narrowband receiver (the spectrum analyzer) versus a wideband receiver (the oscilloscope).

Two-Tone Input

Another input signal commonly used for distortion tests is the two-tone signal.

$$V_{in} = A_1 \cos \omega_1 t + A_2 \cos \omega_2 t \tag{9}$$

Using our distortion model,

$$V_{out} = k_0 + k_1 V_{in} + k_2 V_{in}^2 + k_3 V_{in}^3 \tag{10}$$

the result is in the form

$$\begin{aligned} V_{out} = {} & c_0 + c_1 \cos \omega_1 t + c_2 \cos \omega_2 t \\ & + c_3 \cos 2\omega_1 t + c_4 \cos 2\omega_2 t \\ & + c_5 \cos 3\omega_1 t + c_6 \cos 3\omega_2 t \\ & + c_7 \cos (\omega_1 t + \omega_2 t) \\ & + c_8 \cos (\omega_1 t - \omega_2 t) + c_9 \cos (2\omega_1 t \\ & + \omega_2 t) + c_{10} \cos (2\omega_1 t - \omega_2 t) \\ & + c_{11} \cos (2\omega_2 t + \omega_1 t) \\ & + c_{12} \cos(2\omega_2 t - \omega_1 t) \end{aligned} \tag{11}$$

where $c_0, \ldots, c_{12}$ are coefficients determined by $k_0, \ldots, k_3$, A_1, and A_2.

Besides the harmonics of the two tones (as in the single-tone case), there are also sum and difference frequencies. These new frequency components are called intermodulation distortion (IMD), because they result from the two tones modulating together. The frequencies present in the output satisfy the following criteria:

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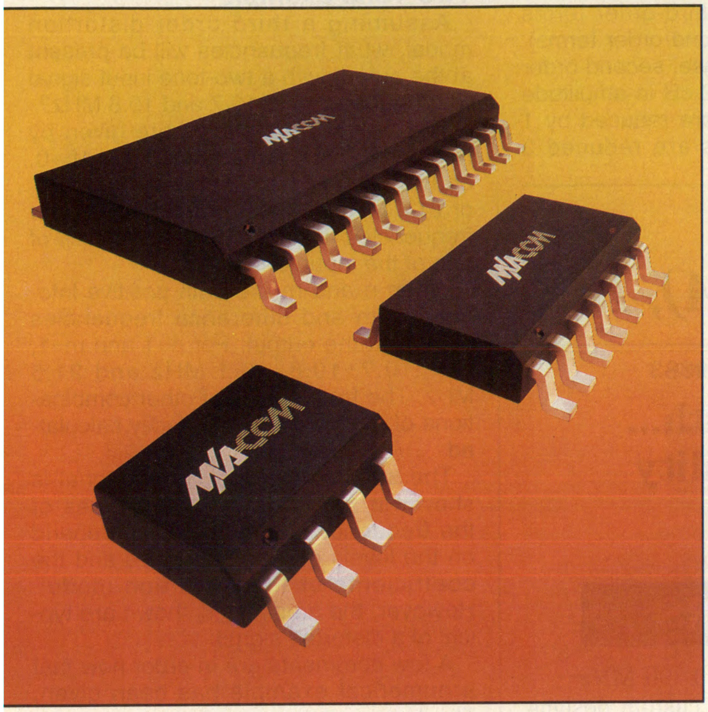
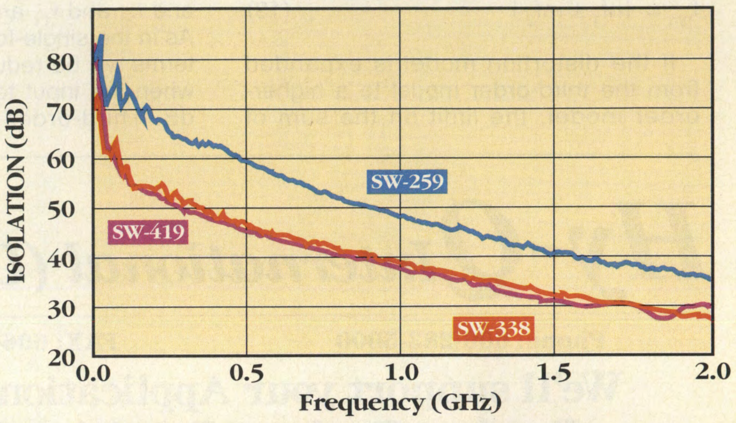
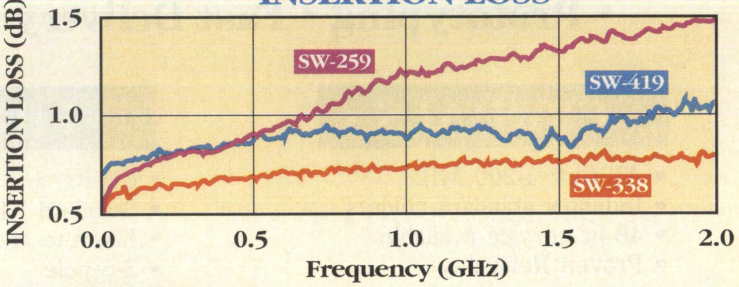


ISOLATION



INSERTION LOSS



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SW-259	SPST Non-Reflective	1.0	47	1.1:1	46	SOIC, 8 lead
SW-289	DPDT	0.5	36	1.1:1	46	SOIC, 14 lead
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
SW-419	SP4T Non-Reflective	0.9	38	1.2:1	46	SOIC, 24 lead

* - All parameters are typical specs @ 1.0 GHz.

RELIABILITY

TEMPERATURE CYCLE	200 Cycles, -65°C to 150°C.
LIFETEST	6,000,000 equivalent device hours @ 85°C.
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$$\omega_{nm} = |n\omega_1 \pm m\omega_2| \quad (12)$$

where n and m are positive integers such that $n+m \leq 3$ or, in units of hertz,

$$f_{nm} = |nf_1 \pm mf_2| \quad (13)$$

If the distortion model is expanded from the third-order model to a higher-order model, the limit on the sum of n+m is raised accordingly.

The order of a particular frequency component is the sum of the n and m values used to obtain that frequency (e.g., f_{12} and f_{21} are third-order terms and f_{20} and f_{11} are second order terms). As in the single-tone case, second order terms will be reduced 2 dB in amplitude when the input tones are reduced by 1 dB. Third-order terms are reduced 3 dB/dB of input signal reduction, and so on for higher-order terms, if present.

Two-Tone Example

Assuming a third-order distortion model, what frequencies will be present at the output with a two-tone input signal with frequencies of 10.7 and 10.8 MHz?

The output frequencies are given by $f = |nf_1 \pm mf_2|$. For n=1 and m=0, $f_{10} = 1f_1 + 0$, or 10.7 MHz. For n=2 and m=0, $f_{20} = 2f_1$, or 21.4 MHz. If either n or m is zero, the frequencies are simply the harmonics of one of the input tones.

When n and m are both positive integers, sum and difference frequencies appear at the output. For n=1 and m=1, $f_{11} = |10.7 \pm 10.8| = 0.1$ MHz and 21.5 MHz. The frequencies of other combinations of n and m can be readily calculated.

The spectrum of the output signal is shown in Figure 2. The amplitudes of the frequency components will depend on the levels of the input tones and the coefficients of the distortion model. However, the amplitudes shown are typical of a distorted signal.

A few comments are in order now that a numerical example has been given. The two input tones were chosen to be close to each other in frequency, as is usually the case for two-tone testing. An examination of Figure 2 will reveal that the spectral lines fall into four groupings. The f_1-f_2 frequency (0.1 MHz) will fall near DC. The other frequencies fall in groups near the fundamentals (10.7 and 10.8 MHz), the second harmonics (near 21.5 MHz), and third harmonics (near 32.4 MHz) of the original tones. Depending on the system involved, some of these distortion products can be neglected since they will be filtered out at some point. For instance, an intermediate frequency (IF) amplifier stage will usually be narrowband, centered on the two input tones. Spectral components at the second and third harmonics will be easily filtered out. The distortion components close to the original tones will be more troublesome since they fall near the desired frequencies. In general, odd-order intermodulation products are of the most concern to RF designers, since the distortion products fall "in-band."

Higher-Order Models

We have chosen to limit the number of terms in the distortion model to produce a third-order behavior. Even with such a simple model, the derivation of the output signal frequency components is

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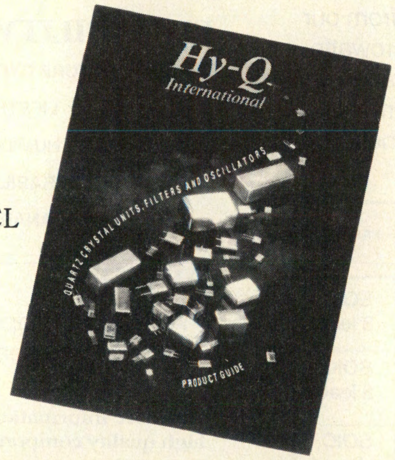

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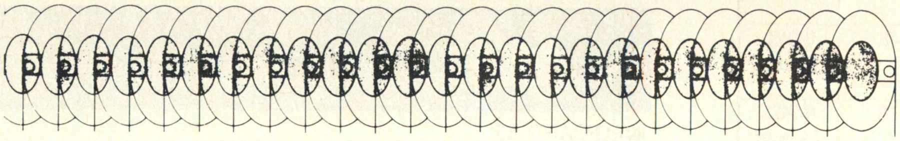

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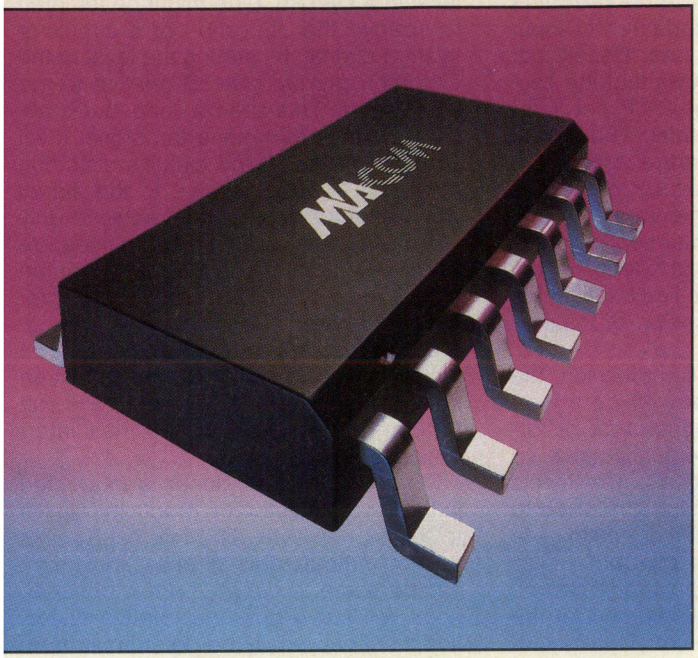

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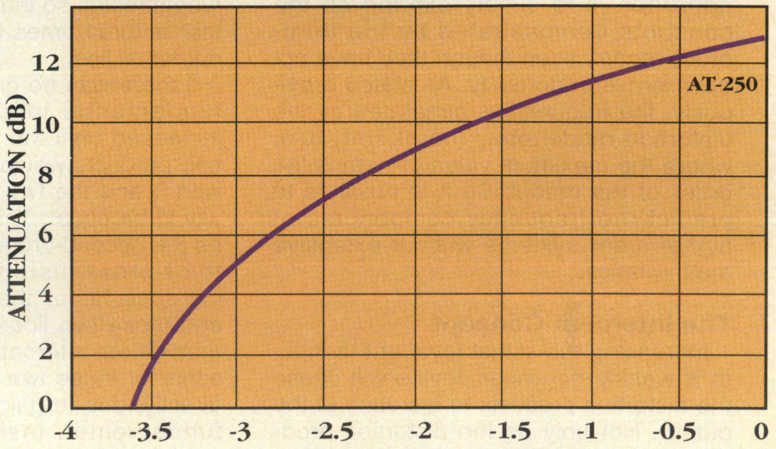

IGITAL ATTENUATORS

IODEL	DESCRIPTION	INSERTION LOSS (dB)	VSWR	Ip3 dBm	PACKAGE
AT-210	4bit, 1dB step to 15 dB	1.8	1.5:1	45	SOIC, 16 lead
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

IGITAL /ANALOG ATTENUATORS

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

OLTAGE VARIABLE ATTENUATORS

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

All parameters are typical specs @ 1.0 GHz.
All V VAs are non-reflective.

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lengthy and expanding the model to higher order only makes the situation worse. Fortunately, for many situations, a third-order model is sufficient.

But what if the third-order model is insufficient? For instance, it is common to have significant energy in the fifth, sixth or seventh harmonic of a single tone, yet the third-order model does not show this effect. The analytical approach used previously can simply be expanded to include the higher-order terms, with the penalty of the mathematics getting more difficult. Another approach is to simply expand on the concepts demonstrated by the third-order model, even though they have not been proven rigorously. As stated previously, the frequencies generated by the distortion model obey the $nf_1 \pm mf_2$ rule, where the maximum value of $n+m$ is the order of the model. So it is possible to predict the frequency components of higher-order systems without extensive mathematics.

The Intercept Concept

Increasing the signal level at the input to a weakly nonlinear device will cause the distortion products to increase at the output. Not only do the distortion products increase in amplitude, that increase is faster than the input signal's. Figure 3 shows a plot of the output power versus the input power for the fundamental, second-order frequency components, and third-order frequency components. For increasing fundamental power, the fundamental output power increases in a linear manner, according to the gain or loss of the device. At some point, gain compression occurs and the fundamental output power no longer increases with input power. The output power of the second-order distortion products also increases with fundamental input power, but at a faster rate. Recall that the distortion model shows that the second-order terms change 2 dB per 1 dB change in the fundamental. Thus, on a decibel plot, the line representing the second order output power has twice the slope of the fundamental line. Similarly, the third-order distortion products change 3 dB per 1 dB of change in the fundamental, so that the line has a slope that is three times the slope of the fundamental line.

If there was no gain compression, the fundamental input power could be increased until the second-order distortion products would eventually catch up with it and the two output power levels would be equal. This point is referred to as the second-order intercept point. The third-order distortion products also increase faster than the fundamental, and those two lines will intersect at the third-order intercept point. Rarely can either of these two points be measured directly, due to gain compression of the fundamental. Instead, the intercept points are extrapolated from measurements of the fundamental and distortion products at power levels below the point where gain compression occurs. The intercept points are usually specified in dBm and may refer either to the output or input. (It is important to always specify whether the intercept point refers to the output power or the input power. The two points will differ by the gain of the device.)

The utility of the intercept concept is in specifying and predicting the distortion level in a system. One might be tempted to specify the distortion of a circuit or system directly by stating the level of the distortion products in dB relative to the signal level. This can be done, but is not very meaningful unless the signal level is also specified. One circuit's distortion might be –80 dB relative to the signal while another circuit might achieve only –40 dB. However, these two values are not a fair comparison unless the same signal level is used. the second-order and third-order intercept points are figures of merit which are independent of signal level. Therefore, the distortion performance of two different circuits can be compared quite easily if their intercept points are known.

Most often, an engineer is interested in the level of distortion products relative to the signal level. The intercept points do not indicate this directly and seem cumbersome to use, but a few observations will show how the relative distortion level can easily be determined from the intercept point. The difference between the level of the second-order distortion products and the fundamental signal level is the same as the difference between the fundamental signal and the intercept point. Suppose the second-order intercept point is +15 dBm and the fundamental signal level is –10 dBm (both referred to the output of the device). The difference between these two values is 25 dB. Therefore, the second-order distortion products will be 25 dB below the fundamental, or –35 dBm.

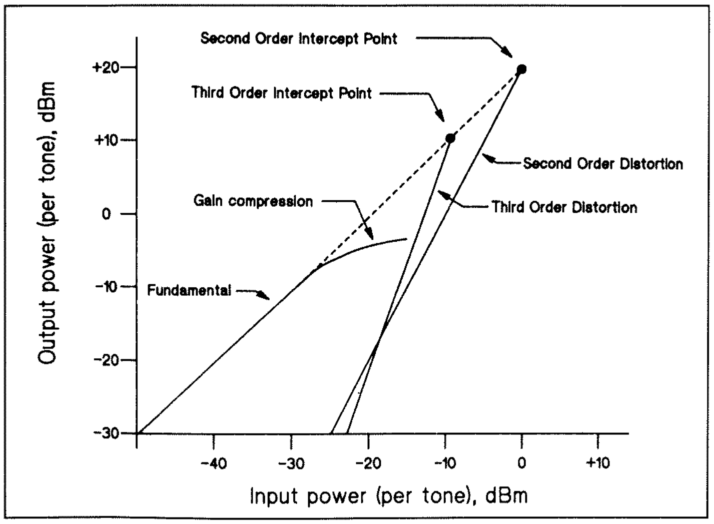

Figure 3. Plot of fundamental, second-order and third order product distortion levels, to illustrate the intercept point concept.

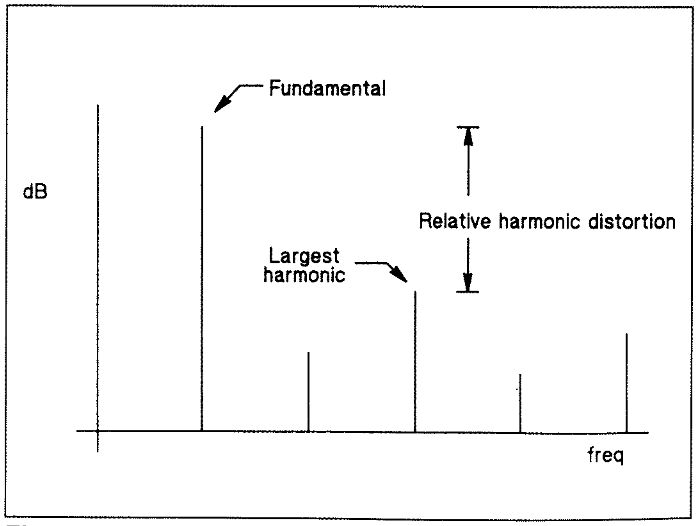

Figure 4. The harmonic distortion of a signal is often specified y stating the largest harmonic relative to the fundamental.

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SPDT Switches	ASCO2R2-12 ASCO2S2-12 AHOO1R2-12	The $2.75 Switch Single Positive Bias Low Loss, High Power	0.7/30 dB 1.1/28 dB 0.4/25 dB	SOIC 8 SOIC 8 SOIC 8	General Purpose Battery Powered Handsets GSM Portable Handsets
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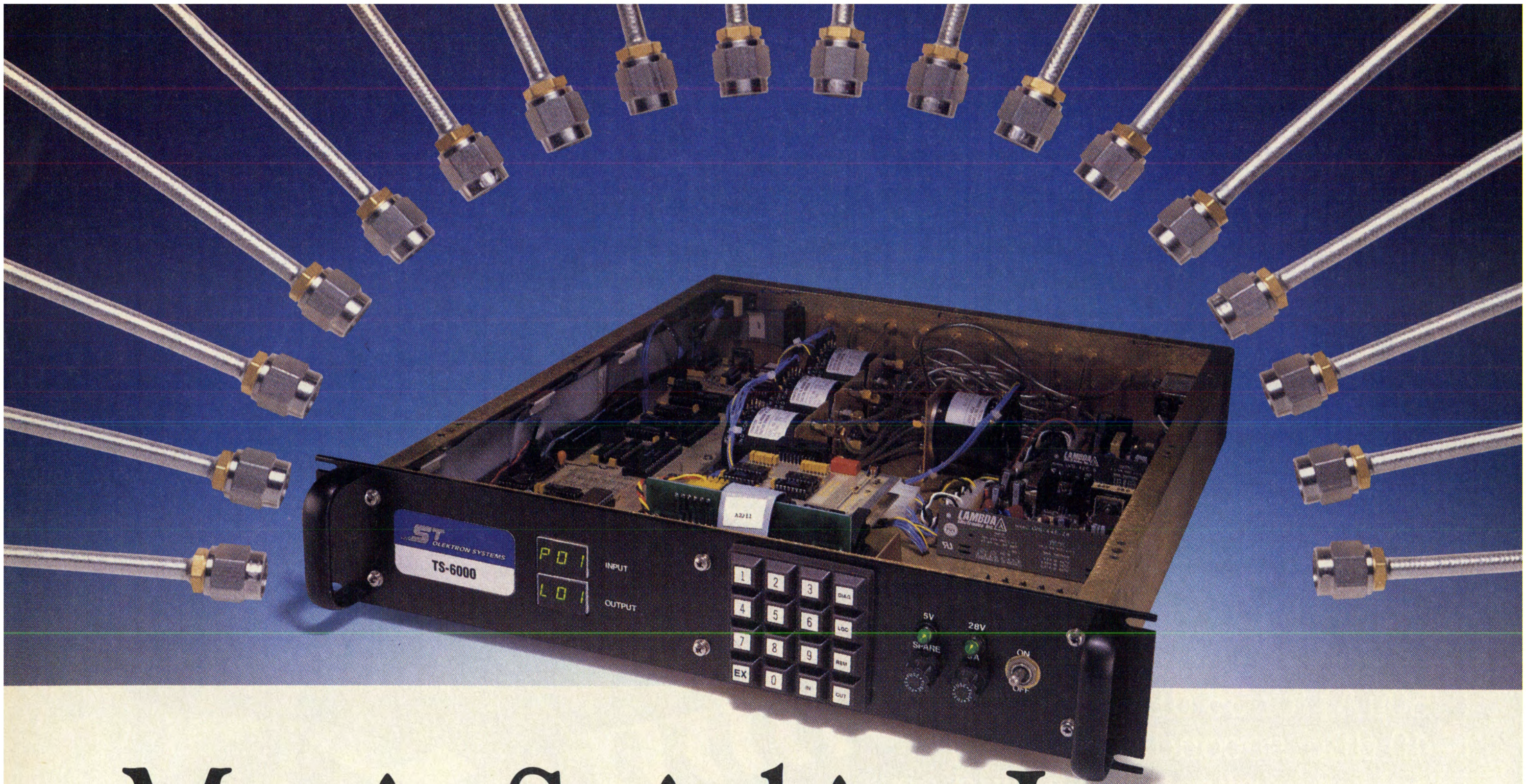

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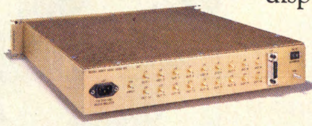

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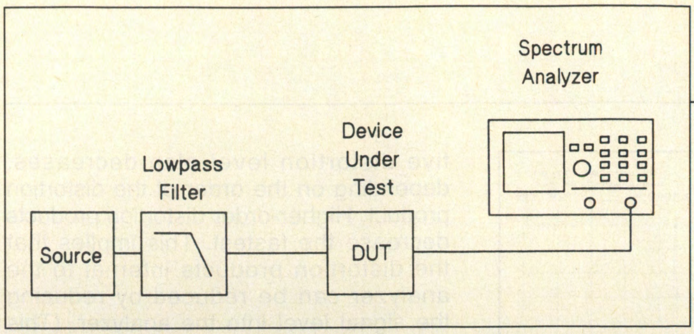

Figure 5. Harmonic distortion of a signal source can be improved by installing a lowpass filter.

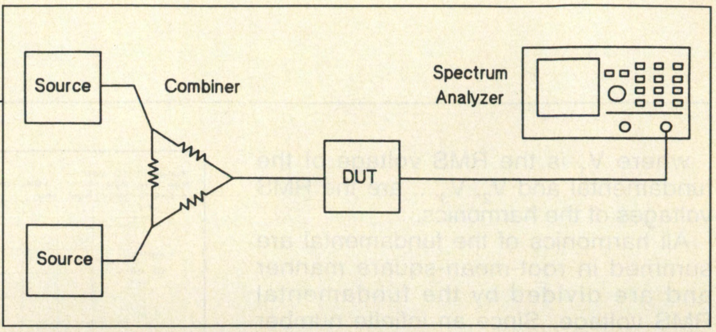

Figure 6. Two signals combined to make a two-tone signal for IMD testing.

So the intercept point allows easy conversion between fundamental signal level and distortion level. Often the distortion level is specified relative to the fundamental power level, and the conversion to absolute power (dBm) is not necessary.

The difference between the level of the third-order distortion products and the fundamental signal level is twice the difference between the fundamental signal level and the third-order intercept point. (Note that the second-order intercept point is not the same as the third-order intercept point.) Suppose that the third-order intercept point is +5 dBm and the fundamental signal level is –25 dBm, both referred to the output. The difference between the intercept and the fundamental is 30 dB, so the third order distortion products will be two times 30 dB down from the fundamental. The relative distortion level is –60 dB and the absolute power level of the distortion products is –85 dBm.

Harmonic Distortion Measurements

Harmonic distortion measurements can easily be made with a spectrally pure signal source and a spectrum analyzer. The quality of the measurement is limited by the harmonic distortion of both the signal source and spectrum analyzer. The signal source is most often the limiting factor, with harmonic distortion performance often not much better than 40 dB below the fundamental.

The source provides a signal to the device under test and the spectrum analyzer is used to monitor the output. Figure 4 shows a typical harmonic distortion measurement. The distortion level may be specified by expressing the largest harmonic in dB relative to the fundamental, as shown in Figure 4.

Alternatively, the distortion may be specified as total harmonic distortion (THD), usually as a percent of the fundamental. THD takes into account the power in all harmonics:

$$THD = \sqrt{V_2^2 + V_3^2 + \cdots} / V_1 \qquad (14)$$

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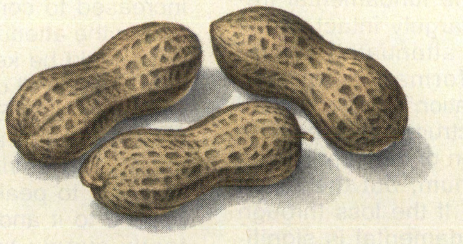

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where V_1 is the RMS voltage of the fundamental and V_2, V_3,... are the RMS voltages of the harmonics.

All harmonics of the fundamental are summed in root-mean-square manner and are divided by the fundamental RMS voltage. Since an infinite number of harmonics cannot be measured, a finite number will have to suffice. Fortunately, the harmonic amplitude tends to decrease with higher harmonic numbers. The calculation is somewhat tedious for a large number of harmonics, but some spectrum analyzers include an automatic THD function. If not, the user must determine each harmonic amplitude and compute THD.

Use of Low-Pass Filter on Source

The signal source is often the limiting factor in harmonic distortion measurement, due to its own harmonic distortion. A typical signal generator has harmonics on the order of –40 dB relative to the fundamental, while a typical spectrum analyzer may have a dynamic range of 70 or 80 dB.

A lowpass filter can be used to improve the source's effective harmonic distortion, as shown in Figure 5. The cutoff frequency of the low-pass filter is chosen such that the fundamental frequency is passed largely intact, while the harmonics are attenuated significantly. The performance of the source/filter combination can be verified directly by the spectrum analyzer. The passband attenuation of the filter should be kept to a minimum, but the exact value is not critical. If the loss through the filter at the fundamental is significant, it should be accounted for when setting the coarse output level. The spectrum analyzer can be used to check directly the fundamental level at the output of the filter.

Intermodulation Distortion Measurements

To test for intermodulation distortion, two stimulus sine waves are required. The test setup shown in Figure 6 has two independent signal sources connected with a power combiner to drive the device under test. The sources are set at the same output level, but at different frequencies. The 6 dB loss of the combiner should be accounted for when setting the output amplitudes of the sources. A typical spectrum analyzer display of the two-tone distortion test is shown in Figure 7. As shown, the third-order products (f_{21} and f_{12}) that fall close to the original two tones are being measured. This is a common measurement since the two distortion products fall close to the original two tones and are difficult to remove by filtering.

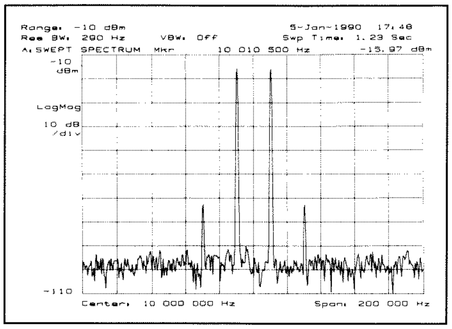

Figure 7. A typical two-tone IMD measurement, which measures third- order products close to the original two tones.

In some cases, the two sources may interact and produce intermodulation distortion. This problem can be detected with the spectrum analyzer and can be cured by inserting attenuators at the outputs of the sources. These attenuators increase the isolation between the sources and prevent internally generated intermodulation distortion. The output levels of the sources should be increased to compensate for the signal loss in the attenuators.

It should be kept in mind that the two sine waves will combine to create a signal which is 6 dB larger than the individual tones (after accounting for combiner loss). The device under test is often sensitive to peak instantaneous voltage applied to it and the user may inadvertently supply twice the desired peak input voltage.

Distortion Internal to the Analyzer

The preceding discussion was oriented toward understanding and measuring distortion in the device under test. However, the internal circuits of the analyzer are imperfect and will also produce distortion products. The distortion performance of the analyzer is specified by the manufacturer, either directly or lumped into a dynamic range specification. The instrument user can stretch the performance of the analyzer by understanding the nature of these distortion products.

As shown in this article, distortion products can be reduced in amplitude by reducing the signal level. Not only do the absolute levels of distortion products decrease, they decrease more rapidly than the decrease in the signal level. So as the signal level decreases, the relative distortion level also decreases, depending on the order of the distortion product. Higher-order distortion products decrease the fastest. This implies that the distortion products internal to the analyzer can be reduced by reducing the signal level into the analyzer. (This may not be true of some analyzers which use a digital IF section, which can have low-level distortion products that are artifacts of the analog-to-digital conversion process, and do not decrease with a decrease in signal level.) The internal input attenuators of the analyzer may be used or an external attenuator may be attached, improving the distortion measurement range of the analyzer. The most obvious disadvantage of reduced signal level is reduced signal-to-noise ratio. The user may find that the low-level distortion products are buried in the noise. Reducing the resolution bandwidth of the analyzer will reduce the residual noise level, but at the expense of a slower sweep rate.

In some measurement situations, the amount of distortion is not of concern, and the signal level at the input of the analyzer can be increased to provide a better signal-to-noise ratio. For many measurements, the distortion products are known to occur at frequencies which are not of interest. For example, a narrowband measurement around the fundamental frequency of a sine wave will not be degraded by the presence of harmonic distortion, since the harmonics will fall far away from the frequency range of interest. The instrument user must always be careful not to apply too large of a signal to the input of an analyzer, so that the damage level is not exceeded. ***RF***

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2. James K. Hardy, *High Frequency Circuit Design*, Reston Publishing Company, Inc., 1979.
3. W.H. Hayward, *Introduction to Radio Frequency Design*, Prentice-Hall, Inc., 1982.

About the Author

Bob Witte is an engineer at Hewlett-Packard Company, Colorado Springs Division. This article is an excerpt from his book *Spectrum and Network Measurements*, published in 1991 by Prentice-Hall. A review of the book is contained on page 70 of this issue.

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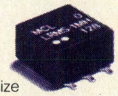

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LRMS-2	5-1000	DC-1000	+7	6.8	38	6.95
LRMS-2D	5-1000	DC-1000	+7	6.8	40	7.25
LRMS-2U	10-1000	10-1000	+7	6.5	46	11.45
LRMS-5	5-1500	DC-1000	+7	6.0	41	13.95
LRMS-11A	1400-1900	40-500	+7	7.0	25	16.95
LRMS-1LH	2.0-500	DC-500	+10	5.8	47	7.95
LRMS-2LH	5-1000	DC-1000	+10	6.6	40	8.95
LRMS-5LH	10-1500	DC-900	+10	5.4	38	14.95
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An Ultra-Low Distortion HF Switched FET Mixer

By Eric Kushnick
LTX Corporation

This entry in the 1992 RF Design Awards Contest is a unique application of switched FET mixer technology. It uses op amp feedback techniques to yield an HF mixer with superlative performance in the area of intermodulation distortion and local oscillator power requirements.

One measure of the useful dynamic range of a mixer is its 3rd order input intercept point. (A definition of 3rd order input intercept point will be provided shortly.) Diode ring double balanced mixers (DBM) and the standard switched FET DBM's are capable of 3rd order input intercept points of +25 dBm to +35 dBm, but this occurs with local oscillator power levels in the +20 dBm to +30 dBm range for the diode ring DBM, and local oscillator power levels in the range of +10 dBm to +15 dBm for the switched FET DBM using resonant drive circuit techniques (1). (Resonant drive circuits are certainly inconvenient when designing a broadband mixer circuit.) Since higher local oscillator (LO) power means more local oscillator design problems, more shielding problems, etc., there exists a need for a mixer that can achieve high dynamic range (as measured by its 3rd order input intercept point) while operating on low local oscillator power.

A Brief Review of Mixer Theory

An ideal mixer is simply a multiplier. Given an input of $X = A\cos 2\pi f_1 t$, the ideal mixer simply multiplies by $B\cos 2\pi f_2 t$ to produce the output, $Y = AB(\cos 2\pi f_1 t)(\cos 2\pi f_2 t)$. Using trigonometric identities,

$$Y = (AB/2)\cos 2\pi(f_1 - f_2)t + (AB/2)\cos 2\pi(f_1 + f_2)t$$

However, constructing an ideal multiplier is not always easy. Another technique is to use an ideal square law device, where $Y = a_2X^2$. A diode or a FET may approximate this characteristic under the proper conditions. If,

$$X = A\cos 2\pi f_1 t + B\cos 2\pi f_2 t$$

then,

$$Y = a_2(A^2\cos^2 2\pi f_1 t + 2AB(\cos 2\pi f_1 t)(\cos 2\pi f_2 t) + B^2\cos^2 2\pi f_2 t)$$

Again, using trig identities,

$$Y = a_2(A^2/2 + B^2/2 + AB\cos 2\pi(f_1 - f_2)t + AB\cos 2\pi(f_1 + f_2)t + (A^2/2)\cos 2\pi 2f_1 t + (B^2/2)\cos 2\pi 2f_2 t)$$

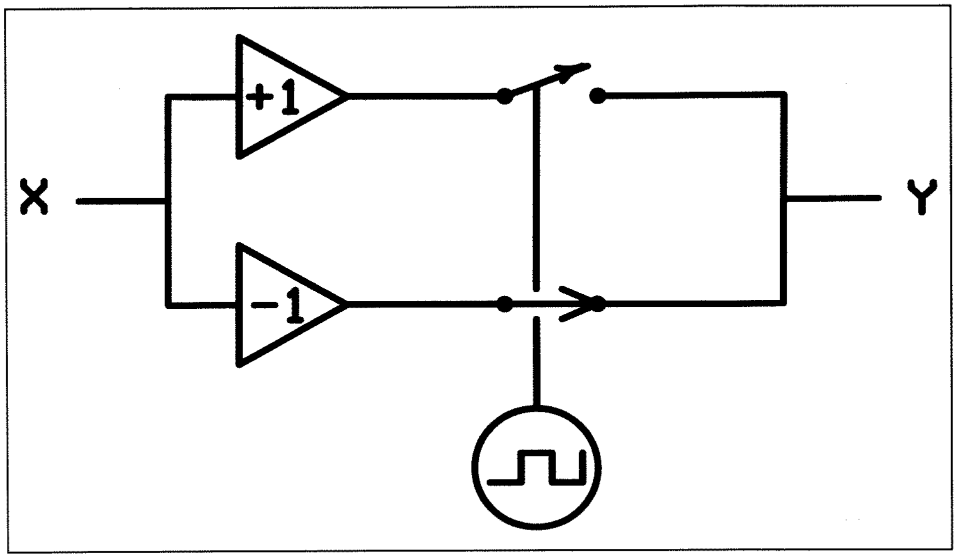

Figure 1. Switched mixer.

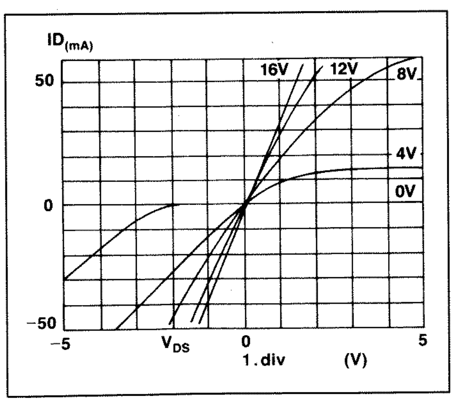

Figure 2. Typical MOSFET V-I characteristics.

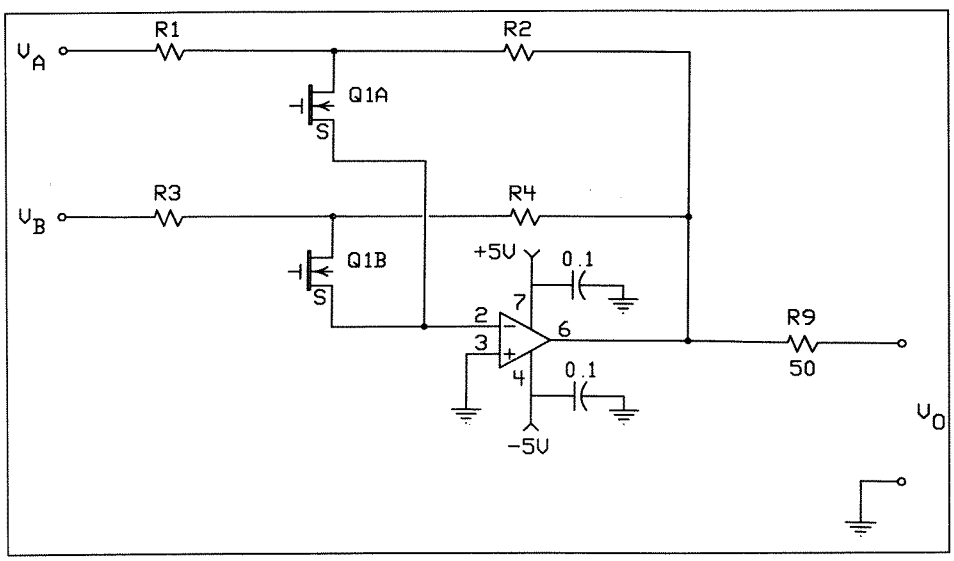

Figure 3. FETs in op amp feedback loop.

Here, in addition to the desired outputs, there are outputs at DC and $2f_1$ and $2f_2$. These are usually not too hard to filter out. Unfortunately, it is impossible to get a real device that behaves in an ideal square law manner. Most devices will also contain 3rd order and higher terms in their input output relationships, such as,

$$Y = a_0 + a_1X + a_2X^2 + a_3X^3 + ...$$

If two desired output signals at f_1 and f_2 encounter a 3rd order term, then outputs at frequencies of $2f_1 - f_2$ and $2f_2 - f_1$ will also be generated. This can be troublesome, because if f_1 and f_2 are near each other in frequency, then $2f_1 - f_2$ and $2f_2 - f_1$ will both be close in frequency to the desired outputs f_1 and f_2, and cannot be removed by filtering.

Another type of mixer takes the input signal and multiplies it alternately first by 1 and then by −1. This is the same as multiplying the input signal by a square wave. This "multiplication" is very easy, however, because it can be done simply with an inverter and two switches (See Figure 1). The Fourier series for a square wave is:

$$B\cos 2\pi f_2 t + (B/3)\cos 2\pi 3f_2 t + (B/5)\cos 2\pi 5f_2 t + ...$$

Therefore,

$$Y = AB(\cos 2\pi f_1 t)(\cos 2\pi f_2 t) + (AB/3)(\cos 2\pi f_1 t)(\cos 2\pi 3f_2 t) + (AB/5)(\cos 2\pi f_1 t)(\cos 2\pi 5f_2 t)$$

Using trig identities, Y has frequency components at $f_2 + f_1$, $f_2 - f_1$, $3f_2 + f_1$, $3f_2 - f_1$, $5f_2 + f_1$, $5f_2 - f_1$, etc. The undesired frequencies are easy to filter out.

Engineering

Why do mixers based on MOSFET switches require such high local oscillator drive power? While the gates of the MOSFETs do not consume, or dissipate power, they do require a large voltage for the mixer to operate properly. The problem is that there is considerable signal voltage across the FET switch, and considerable signal current through the FET switch. Figure 2 shows the V-I characteristics of a typical MOSFET. When the FET is on, the gate voltage must be high enough to keep R(on) as low and as linear as possible. (R(on) is the reciprocal of the slope of the V-I curves.) When the FET is off, the gate voltage must be sufficiently negative to prevent the FET from being turned on by large negative signal voltages on its drain. The solution to these problems, rather than driving the FET gate with higher voltages, is not to allow signal current to pass through the FET, and not to allow signal voltage to appear across the FET.

Figure 3 shows a circuit that accomplishes the first of these two objectives. Two MOSFET switches are placed at the summing junction of a current feedback operational amplifier. When switch Q1A is closed and switch Q1B is open, the feedback network consisting of R1 and R2 is connected to the op amp, and the op amp feedback insures that there is little voltage across switch Q1A and that only the op amp input current flows through switch Q1A. Since the op amp input current is at least an order of magnitude less than the signal current, the effect of the MOSFET R(on) is reduced by at least an order of magnitude.

It is also necessary to ensure that the voltage across switch Q1B, the off

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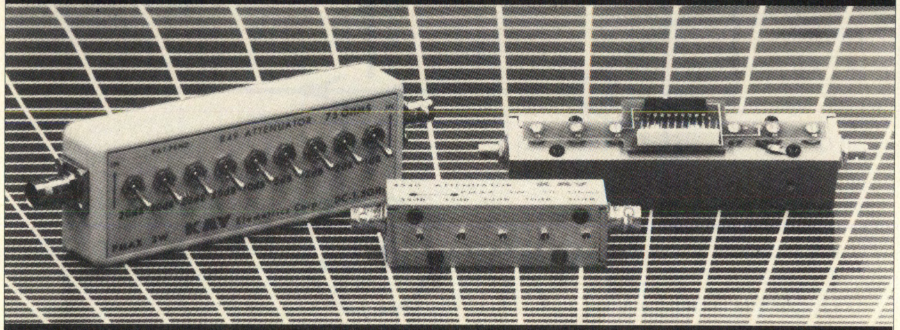

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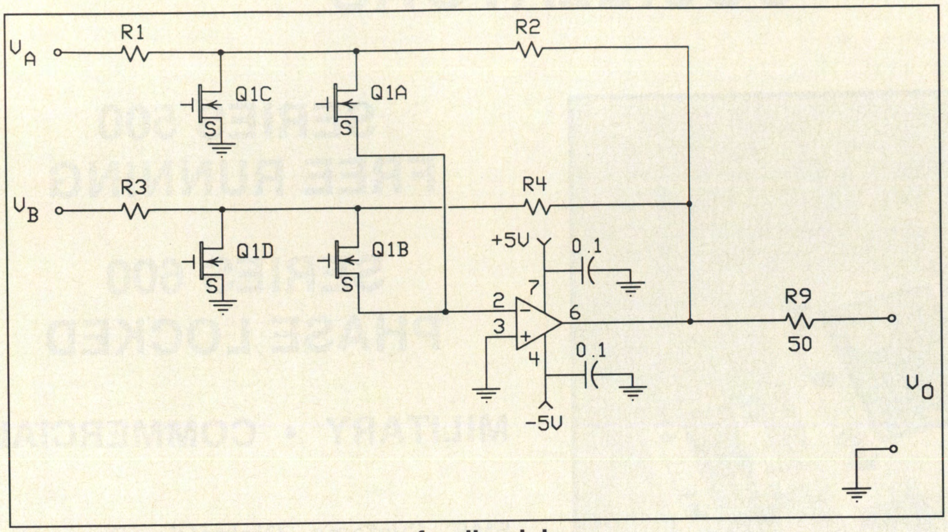

Figure 4. More FETs in op amp feedback loop.

switch, remains low, the second objective, so that a large negative gate voltage will not be required to turn off the FET. This can be accomplished by adding two more MOSFETs, Q1C and Q1D, as shown in Figure 4. When MOSFET Q1B is off, MOSFET Q1D is turned on to hold Q1B's drain near ground potential. (Similarly for MOSFETs Q1A and Q1C in their turn). Although Q1B and Q1D do conduct signal current, the current that they conduct does not contribute to the op amp output. Thus, Figure 4 is a MOSFET switch circuit in which the FETs that contribute to the output do not pass signal current or withstand signal voltage. Considerably less drive voltage is required with this circuit than with other FET switch circuits used in mixers.

How can this circuit be used as a mixer? Referring again to the circuit of Figure 4, suppose that V_A is some signal voltage, V_{signal}. If $V_B = -V_A = -V_{signal}$, and if all the resistors R1 - R4 are equal, then alternately switching on Q1A (and Q1D), and then Q1B (and Q1C) produces an output which is just V_{signal} first multiplied by 1 and then multiplied by −1. This is the same as multiplying V_{signal} by a square wave at the switching frequency of the FETs, which means that the circuit acts as a mixer.

Figure 5 shows the actual circuit that was constructed. R6 is chosen for input matching (approximately 56 ohms for a 50 ohm system). T1 accomplishes the signal inversion from V_A to V_B ($V_A = -V_B$). T1 has a 6 dB loss due to its turns ratio, so R2 and R4 have been made two times the value of R1 and R3 to make up for this. Transformers T3 and

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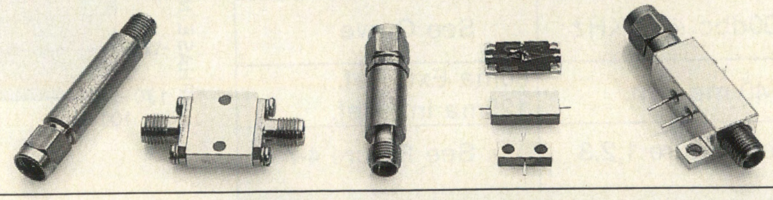

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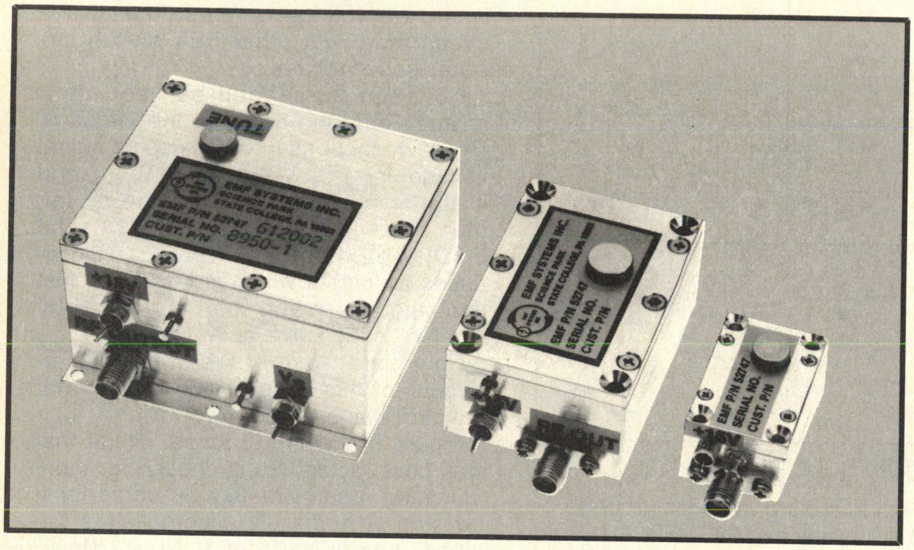

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Package Outline	See Figure 1,2,3	See Figure 4-7

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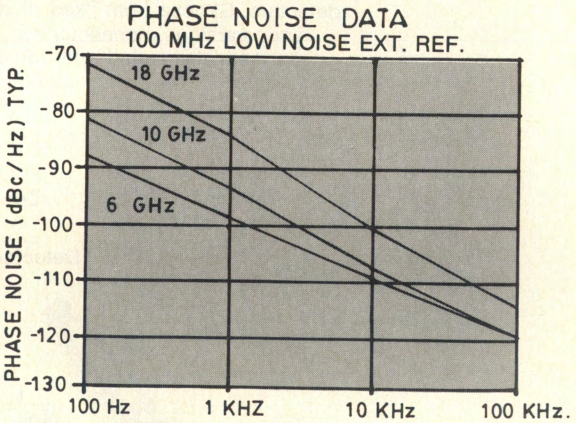

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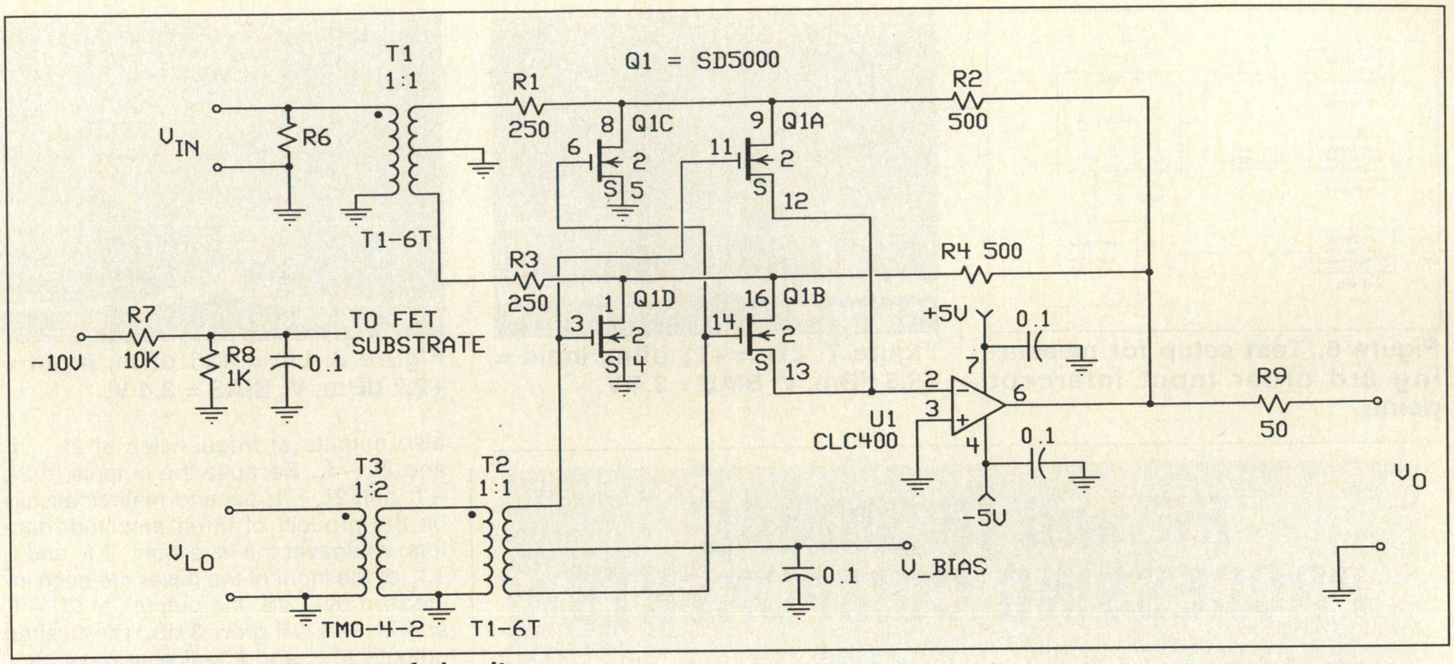

Figure 5. Schematic diagram of circuit.

T2 take the local oscillator signal and apply it in the proper phase to the gates of the FETs. A DC bias voltage is added to the gate drive signals to give them the proper DC level with respect to the FET gate threshold voltage. The output of the circuit is taken through R9, the 50 ohm series terminating resistor. This produces a 6 dB drop when driving a 50 ohm load.

Measurements and Observations

The measurement of most interest to be made on this circuit is, of course, the measurement of two tone 3rd order input intercept point versus local oscillator power. All mixers have some level of 3rd order intermodulation distortion. This means that two tones at the input of the mixer at frequencies of $f_1 + f_{lo}$ and $f_2 + f_{lo}$, will not only produce outputs at f_1 and f_2 (f_{lo} is the local oscillator frequency and an IF at the difference frequency of f(input) – f_{lo} has been assumed), but

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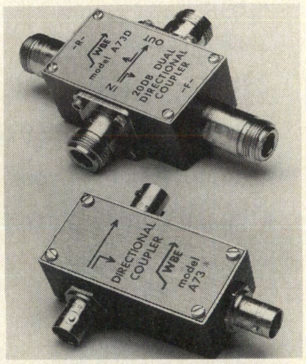

Model	Freq Range MHz	Coupling Level dB	Coupler Type	In Line Power	Minimum Directivity (dB) 1-500 MHz	5-300 MHz	In Line Loss (dB)	Flatness of Coupled Port (dB)	VSWR	Price 50 ohm with BNC conns.
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A73-20GA					30	40			1.5:1 1-500 MHz	131.00
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A73-30P2	1-100	30	single	200W cw 50 ohm	30 dB		.05	±.15	1.05:1 max	312.00

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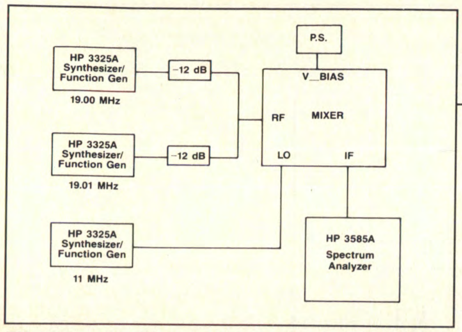

Figure 6. Test setup for measuring 3rd order input intercept point.

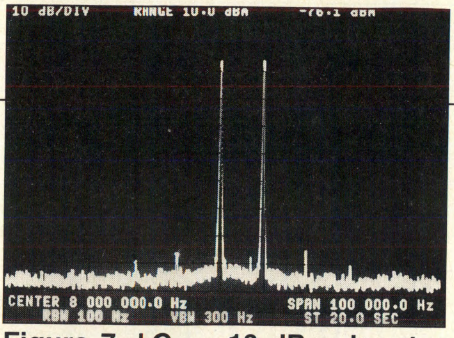

Figure 7. LO = +13 dBm, Input = +8.5 dBm, V_BIAS = 3.4 V.

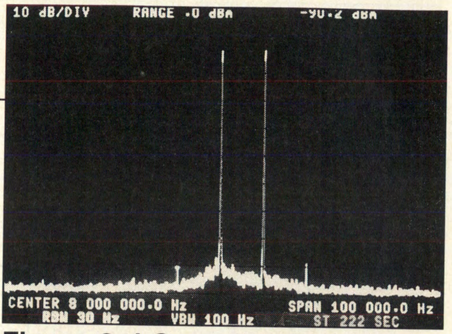

Figure 8. LO = +13 dBm, Input = +2.7 dBm, V_BIAS = 3.4 V.

also outputs at frequencies of $2f_1 - f_2$ and $2f_2 - f_1$. Because the outputs at $2f_1 - f_2$ and $2f_2 - f_1$ depend mathematically on the product of three amplitude factors, whenever the levels of $f_1 + f_{lo}$ and $f_2 + f_{lo}$ at the input of the mixer are each increased by 1 dB, the outputs at $2f_1 - f_2$ and $2f_2 - f_1$ will grow 3 dB. The desired outputs at f_1 and f_2 will only grow by 1 dB each, however, so even though initially the levels of the $2f_1 - f_2$ and $2f_2 - f_1$ outputs are much lower than the levels of the outputs at f_1 and f_2, there is in theory an input level of $f_1 + f_{lo}$ and $f_2 + f_{lo}$ which would cause the $2f_1 - f_2$ and $2f_2 - f_1$ outputs to equal the desired f_1 and f_2 outputs. This input level is called the 3rd order input intercept point, and is generally expressed in dBm. (The mixer, of course, can not actually operate at this input level, it is mathematically determined from measurements made at lower levels.)

Figure 6 shows a test setup for measuring the 3rd order input intercept point. All measurements were made at a local oscillator frequency of 11 MHz, an $f_1 + f_{lo}$ frequency of 19.00 MHz and an $f_2 + f_{lo}$ frequency of 19.01 MHz. This produces desired outputs at 8.00 MHz and 8.01 MHz, and 3rd order products at 7.99 MHz and 8.02 MHz. Signal levels at various points in the test setup were measured without disturbing the test setup using a ×10 scope probe connected to the 1 Megohm input of the spectrum analyzer.

The adjustment of V_BIAS is fairly critical. It can be adjusted either for maximum gain, producing a gain close

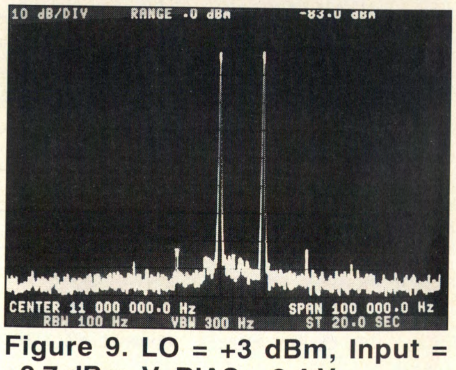

Figure 9. LO = +3 dBm, Input = +2.7 dBm, V_BIAS = 2.4 V.

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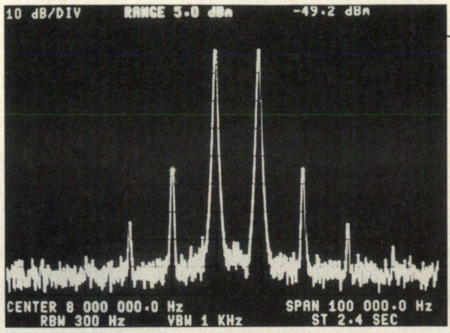


Figure 10. LO = –3 dBm, Input = +8.5 dBm, V_BIAS = 2.2 V.

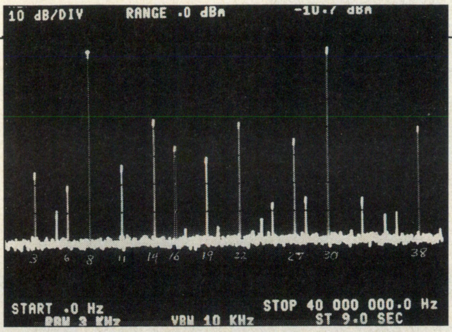


Figure 11. Broadband spectrum of mixer. Table 1. Performance results.

to the theoretical value, or for minimum 3rd order intermodulation distortion, usually at a gain several dB less than theoretical. At the higher (+13 dBm) LO power levels, adjusting V_BIAS to approximately the FET gate threshold of 2 V produces near maximum gain, less sensitivity to V_BIAS and LO power variations, and a 3rd order input intercept point of +34 dBm. This is not bad at all. Adjusting V_BIAS for minimum 3rd order intermodulation products gives a 3rd order input intercept point of +44 dBm with +13 dBm of LO power, and +38.7 dBm with only +3 dBm of LO power. This is very good! Even with a LO drive level of –3 dBm and an input signal level of +8.5 dBm for each tone, the 3rd order input intercept point is still +25 dBm (try that one with your diode ring mixers). Table 1 summarizes these results, while Figures 7 - 11 show some of the measurements from which the 3rd order input intercept points were calculated. *RF*

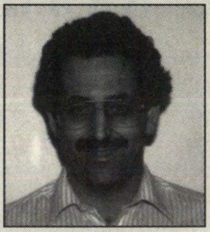
TABLE 1

LO dBm	V_BIAS volts	Input, each tone dBm	Output each tone dBm	3rd order products dBm	3rd order input intercept dBm
+13	3.4	8.5	–5.2	–76	44
+13	3.4	2.7	–10.8	–90.2	40
+13	2.2	2.7	–6.8	–67.4	33
+3	2.4	2.7	–10.9	–83	38.7
–3	2.2	2.7	–11.8	–67.7	30.7
–3	2.2	8.5	–5.7	–49.2	25

References

1. *Siliconix Low Power Discretes Data Book*, 1989, Siliconix Incorporated, LPD-14, pp. 9-134, 9-140 to 9-143.

About the Author



Eric Kushnick has been with LTX Corporation since 1984, where he has been involved in the design of analog and digital circuits in the DC to several hundred MHz frequency range. He attended the Massachusetts Institute of Technology, where he participated in a work-study program at Bell Laboratories and received SB and SM degrees in Electrical Engineering and Computer Science in 1976. He is currently enrolled in a Ph.D. EE program at Northeastern University. He may be reached at LTX Corp., University Ave., Westwood, MA 02090. Tel: (617) 461-1000 ext. 5079.

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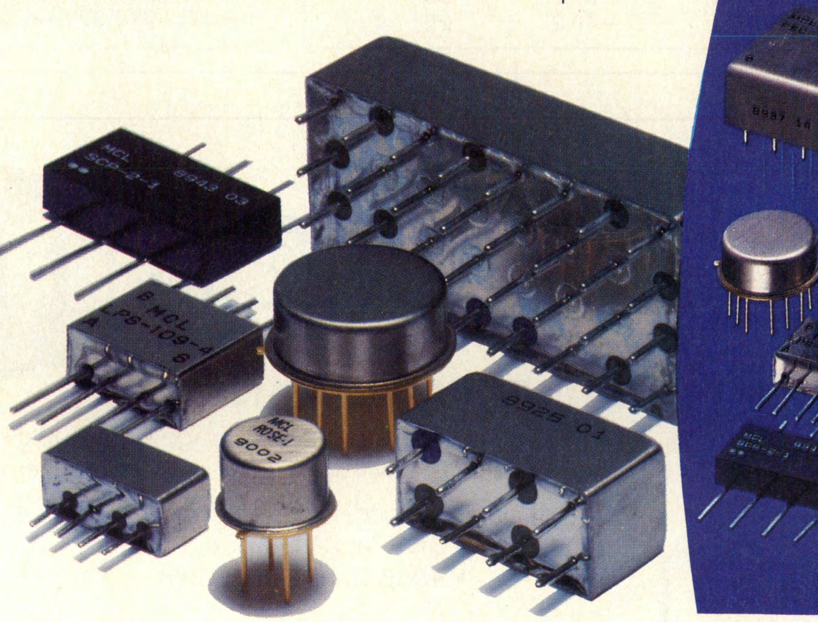

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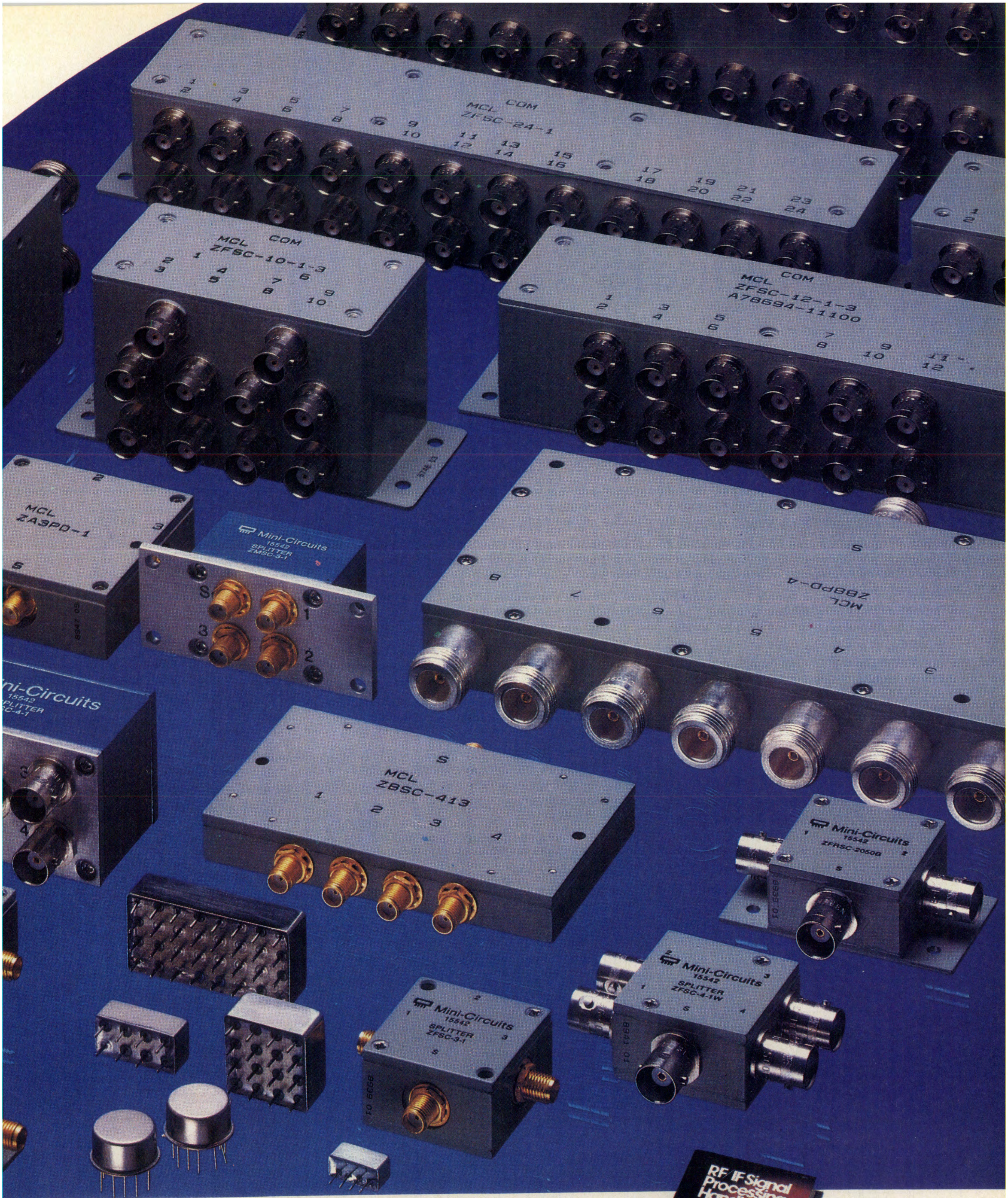

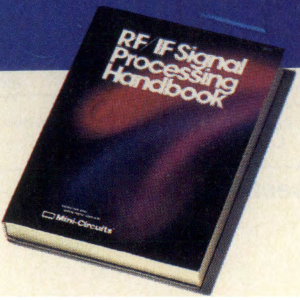

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Program Designs Active Elliptic Filters

By Jack Porter
Cubic Corporation

This program performs the calculations required for designing lowpass, highpass and bandpass active elliptic filters using cascaded single-amplifier biquad circuits. It also calculates the theoretical response of these filters. The program, written in True Basic, will run on any IBM-compatible computer. A VGA, EGA, CGA or Hercules monochrome graphics card is required to display schematics of the various circuits and to display filter response curves. The executable program is file AEF.EXE; the file AEF.TRU contains the source code.

An elliptic filter, sometimes called a Cauer or Zolatarev filter, has finite stopband zeros, and thus requires fewer poles and active filter sections than an all-pole filter with equally sharp cut-off. However, because of these finite zeros it requires more components than a Chebyshev filter with the same number of biquadratic sections.

An elliptic low-pass prototype filter is completely specified by four parameters: N, the number of poles; A_{max}, the maximum passband ripple in dB; A_{min}, the minimum stopband attenuation in dB; and F_S/F_B, the ratio of minimum stopband frequency to maximum passband frequency. The last three of these parameters are depicted in Figure 1. Any three of them can be chosen independently; the fourth is then calculated. These parameters and others used in filter design are defined in Table 1.

As shown in Figure 1, the elliptic filter response is equiripple in both the passband and the stopband. The response of a related filter type, the inverse Chebyshev filter, is maximally flat in the passband, as is a Butterworth filter, and equiripple in the stopband. Its cut-off is not as sharp, but the passband group delay is more constant than that of an elliptic filter. If a value of zero is chosen for A_{max}, the prototype values for an inverse Chebyshev filter are calculated. The specified filter bandwidth is the 3 dB bandwidth for this filter; for an elliptic filter it is the ripple bandwidth. Elliptic filter prototype element values and lowpass to highpass and lowpass to bandpass transformations are calculated using

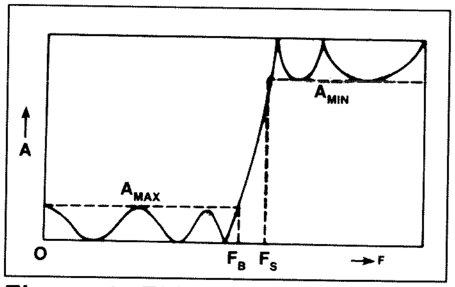

Figure 1. Elliptic filter response and some defining parameters.

methods described in Reference 1. Inverse Chebyshev prototype calculations are based on Reference 2.

This program can be used to design filters composed of either of two different types of single-amplifier biquads, one using negative feedback, described in References 3 and 4, and the other positive feedback, described in Reference 5. Both are relatively insensitive to op-amp parameters and produce acceptable results when used with real, finite gain and bandwidth amplifiers. The program will design lowpass, highpass and bandpass filters, but narrow band-

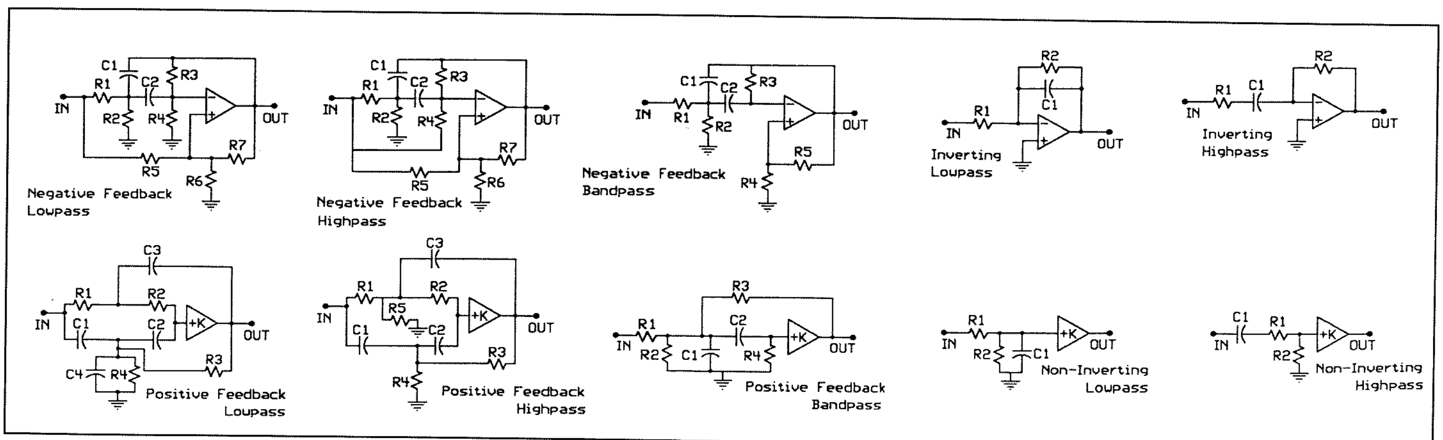

Figure 2. Filter section types.

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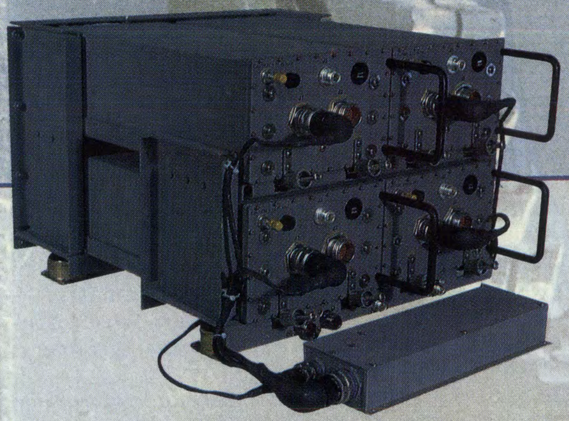

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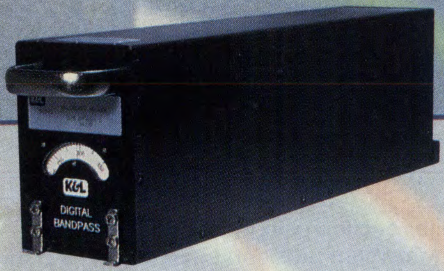
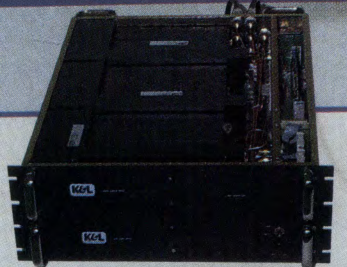


DIGITALLY CONTROLLED BANDPASS FILTERS:

K&L's digitally controlled bandpass filters are high Q devices covering the frequency spectrum from 24MHz to 18GHz, 1 octave per filter. Each model has its own built-in microprocessor which controls a precision stepping motor. This filter series gives the system designer the options of control logic, drive voltages and packages.



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Low-pass Elliptic Filter 01/30/1992			
N= 4	Amax= .5 dB	Amin= 40 dB	FS/FB= 1.62842
Fc= 5000000 Hz	FH= 8.1421e+6 Hz		
Section 1	A0= 1		
Positive feedback	Low-pass		
K= 1			
Fo= 5.15115e+6 Hz	Q= 3.78876	Fz= 8.67332e+6 Hz	
C1= 35.2726 pF	C2= 100 pF	C3= 200 pF	C4= 64.7274 pF
R1= 270.874	R2= 352.423	R3= 176.212	R4= 1170.61
Section 2	A0= 1		
Positive feedback	Low-pass		
K= 1			
Fo= 3.34575e+6 Hz	Q= .737716	Fz= 1.93069e+7 Hz	
C1= 3.00305 pF	C2= 100 pF	C3= 200 pF	C4= 96.997 pF
R1= 252.248	R2= 897.067	R3= 448.534	R4= 350.926

Figure 3. Lowpass filter design.

Bandpass Elliptic Filter 01/30/1992			
N= 2	Amax= .5 dB	Amin= 30 dB	FS/FB= 4.80871
Fc= 50000 Hz	BW= 10000 Hz	Fa= 49749.4 Hz	
FL= 31211.3 Hz	FH= 79298.3 Hz		
Section 1	A0= 1		
Negative feedback	High-pass		
Op amp DC gain= 200000		Op amp GBW= 15 MHz	
Fo= 44896.5 Hz	Q= 7.13931	Fz= 26338. Hz	
C1= 1000 pF	C2= 1000 pF	a= 10	b= .155706
R1= 3791.77	R2= 1591.53	R3= 10763.2	R4= 42039.9
R5= 2128.21	R6= 1000	R7= 4369.31	
Section 2	A0= 1		
Negative feedback	Low-pass		
Op amp DC gain= 200000		Op amp GBW= 15 MHz	
Fo= 55126.9 Hz	Q= 7.13931	Fz= 93970.7 Hz	
C1= 1000 pF	C2= 1000 pF	a= 10	b= .155706
R1= 6371.86	R2= 1065.66	R3= 7264.49	R4= 5536.54
R5= 8548.5	R6= 1000	R7= 5749.75	

Figure 4. Bandpass filter design.

pass filters aren't very practical because of the high Q's they require. Sensitivity to component value variations and to passband gain is proportional to Q. The positive feedback circuit, like all circuits based on parallel-T networks, is more sensitive than the negative feedback circuit to passive component value variations. The values of "a" and "b" in the negative feedback circuit are adjusted to provide a tradeoff between active and passive sensitivities, depending on the op-amp gain and bandwidth. Parameter "b" is the same as that defined in Reference 3; "a" is the same as that in Reference 4 multiplied by $(\omega_n/\omega_0)^2$. The parameter "K" in Reference 4 is equal to b+1.

Theoretical filter response tables and plots can also be generated using this program. Since there are so many possible combinations of elliptic filter parameters, little of this data is available in handbooks. To facilitate comparing theoretical data with that actually measured or calculated using circuit analysis programs, the theoretical frequency response of selected sections of the filter and the entire filter can be calculated, with the passband gain corresponding to that of the actual filter.

The positive feedback filters are capable of operation at higher frequencies than the negative feedback, since they can be implemented using wideband current feedback op-amps. The lowpass and highpass positive feedback circuits both require unity-gain amplifiers and have unity gain in the passband. Passive RC circuits followed by non-inverting amplifiers are used for the real poles. The passband gain for these, if greater than or equal to one, is the same as that of the amplifier.

The lowpass to bandpass transformation requires one biquad section for each lowpass prototype pole. Each complex conjugate pole pair results in one lowpass and one highpass biquad section. The schematics for these are the same as for lowpass and highpass filters, but the Q's are higher. The gain of the filter section is also higher; the gain value at F_c, which varies, is calculated for each section. Real poles result in a bandpass section with center frequency F_a. The positive feedback bandpass filter requires an amplifier gain of 1.2; maximum passband gain is unity at F_a. Gain of this section, which is also calculated at F_c, is always less than one at that frequency.

The maximum gain for lowpass negative feedback sections varies, depending on Q and F_z/F_0. This value is calculated for each section. Highpass negative feedback circuit maximum gain also varies. Exceeding the calculated recommended maximum value will result in increased sensitivity to component value variations. The maximum possible gain for the highpass section is less than one in any case. An inverting integrator or differentiator with arbitrary gain is used for the real pole.

The maximum passband gain of bandpass filter sections using negative feedback amplifiers is much higher than that of the positive feedback type. Their maximum gain is approximately 4Q.

The program contains a number of requests for input data. Those which end with a double question mark are branch points. In these, entering values of zero for all data items causes a return to an earlier branch point; entering a value of −1 for the first item terminates the program.

Schematics of the various filter section types are shown in Figure 2. Each of these is displayed on the screen when

Symbol Table

A_0: Filter section gain (ratio).
A_{max}: Maximum passband attenuation = 0 for inverse Chebyshev filters.
A_{min}: Minimum stopband attenuation.
A_s: Section gain factor - bandpass filters.
BW: Bandwidth - Bandpass filters.
F_a: Adjusted bandpass filter center frequency = $\sqrt{(F_c-BW/2)\times(F_c+BW/2)}$
F_B: Lowpass prototype passband limit.
F_c: Center frequency - bandpass filters.
F_c: Cut-off frequency (passband limit) - lowpass and highpass filters.
F_H: Stopband limit - Lowpass and bandpass filters.
F_L: Stopband limit - Highpass and bandpass filters.
F_0: Pole frequency in Hz.
F_S: Lowpass prototype stopband limit.
F_z: Zero (transmission null) frequency in Hz.
K: Amplifier gain (ratio) - positive feedback filter sections.
N: Number of lowpass prototype filter poles. N=12 maximum.
Q: Filter pole Q.
Q_p: Lowpass prototype pole Q.
S_p, W_p: Real and imaginary parts of lowpass prototype pole.
ω_{op}: Lowpass prototype pole frequency in radians/sec.
ω_z: Lowpass prototype zero frequency in radians/sec.

Table 1.

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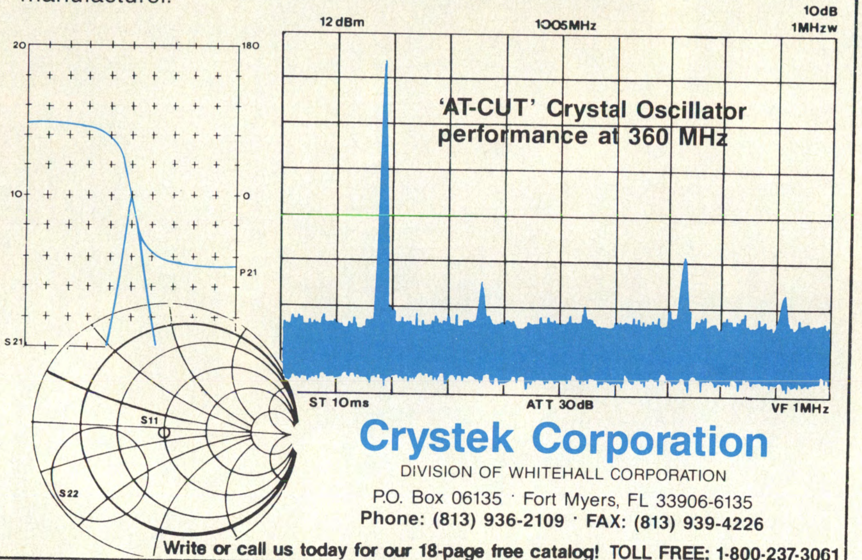

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component values for it are calculated. Figure 3 is an example of a lowpass filter design; Figure 4 shows a bandpass filter design.

The program's theoretical response routine calculates values for gain (reciprocal of attenuation), phase shift, and group delay for a range of frequencies. These values are tabulated and the gain is plotted. The gain plot can be printed using the GRAPHICS command. The minimum frequency (which may be zero), maximum frequency, and frequency interval between data points must be specified.

The file AEFSET.PRT contains printer command codes. The data supplied is for an IBM Proprinter, but it can be edited for other printers. The code numbers can be separated by commas, spaces or any other characters which aren't numbers.

This program is available on disk through the RF Design Software Service. See page 100 for ordering information. *RF*

References

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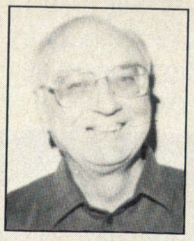
5. A.S. Sedra and P.O. Brackett, *Filter Theory and Design: Active and Passive*, Section 9.5, Matrix Publishers, 1978.

About the Author

Until recently, Jack Porter was a Senior Staff Engineer at Cubic Defense Systems, and this program was submitted via that company. Cubic Defense Systems can be reached at 9333 Balboa Ave., P.O. Box 85587, San Diego, CA 92186-5587.



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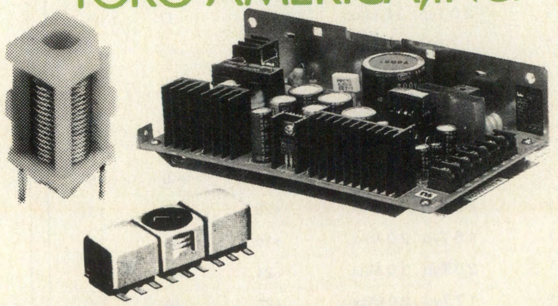

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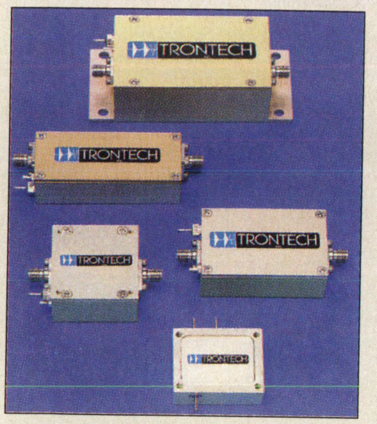

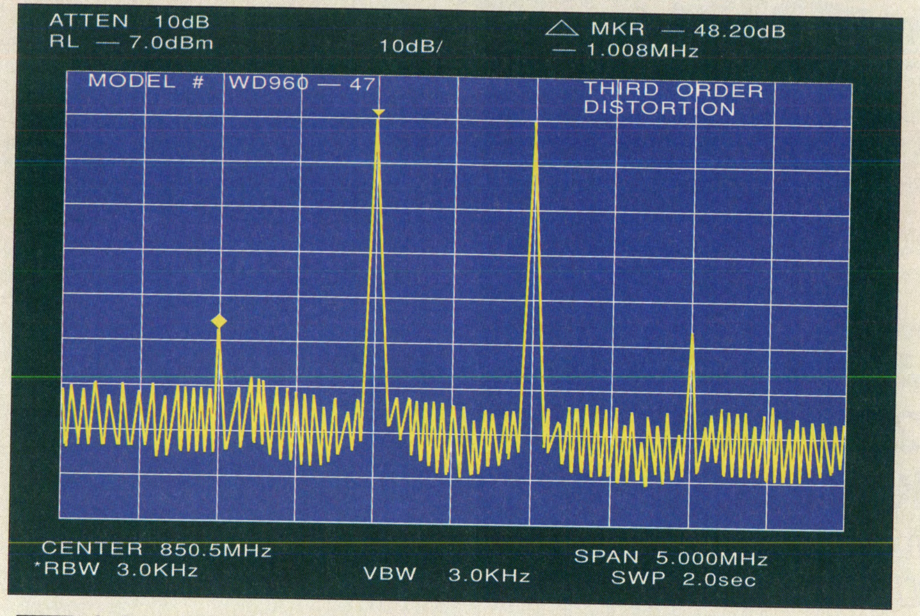

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WD400-40	225-400MHz	17	1	1.6 Typ., 2.0 Max.	+23	+40	15
WD400-47	225-400MHz	28	1	1.6 Typ., 2.0 Max.	+17	+45	15
WD470M	450-470MHz	18	1	1.3 Typ., 2.0 Max.	+22	+40	15
WD500	50-500MHz	17	1	2.2 Typ., 3.0 Max.	+28	+45	15
WD625-45	400-625MHz	28	1	1.5 Typ., 2.0 Max.	+16	+44	24
WD625-47	400-625MHz	21	1	2.0 Typ., 3.0 Max.	+24	+45	24
WD851-40	849-851MHz	17	1	1.7 Typ., 2.0 Max.	+23	+40	15
WD902-43	890-915MHZ	34	1	1.5 Typ., 2.0 Max.	+9	+43	15
WD960-40	800-960MHz	15	1	1.7 Typ., 2.0 Max.	+25	+40	15
WD960-47	900-960MHz	27	1	1.7 Typ., 2.0 Max.	+18	+45	15
WD1150M	1025-1150MHz	30	1	1.8 Typ., 2.0 Max.	+8	+38	15

*Patent Pending

A Quasi-Complementary Class-D HF Power Amplifier

By Frederick H. Raab, Ph.D.
and Daniel J. Rupp
Green Mountain Radio Research Company

Class-D power amplifiers (PAs) employ two transistors in a push-pull configuration. The transistors are driven to act as switches and generate a square wave voltage. The fundamental frequency component of the squarewave is passed to the load through a filter. The class-D PA is ideally 100 percent efficient. In practice, it is significantly more efficient than a similar class-B PA, especially for lower amplitudes and reactive loads.

Virtually all modern broadband RF PAs (with two exceptions (1, 2)) are based upon transformer coupled topologies. In contrast, virtually all modern audio frequency PAs are based upon complementary (transformerless) topologies. The potential advantages of a transformerless RF PA include:

- Greater bandwidth (limited only by the active devices),
- Elimination of power loss in the output transformer,
- Reduced size and weight, and
- Elimination of IMD due to core saturation.

Such a PA might be fabricated in integrated circuit form for considerably less than is currently possible for RF PAs of the same capability.

A true complementary configuration offers great simplicity but requires both n- and p-channel MOSFETs. While p-channel devices are available for power switching applications, none are currently made for RF applications. Experiments with complementary pairs of power switching MOSFETs (e.g., IRF610 and SMP2P20) show that they can be used at frequencies no higher than 3.5 or 7 MHz.

The quasi-complementary power amplifier (Figure 1) directly connects a pair of n-channel MOSFETs in a totem-pole configuration. Drive is supplied through a specially designed input transformer. An important advantage of this topology is that it can be implemented with currently available RF devices.

The voltage and current ratings of currently available devices may not be adequate for delivery of the desired power output directly to a 50 ohm load, hence the use of an output transformer may be required. However, this transformer is external to the PA itself and can therefore be a single-ended, equal-delay configuration. As a result, it is smaller and less complex than the push-pull transformer normally required for class-B or D operation, and also has a larger bandwidth.

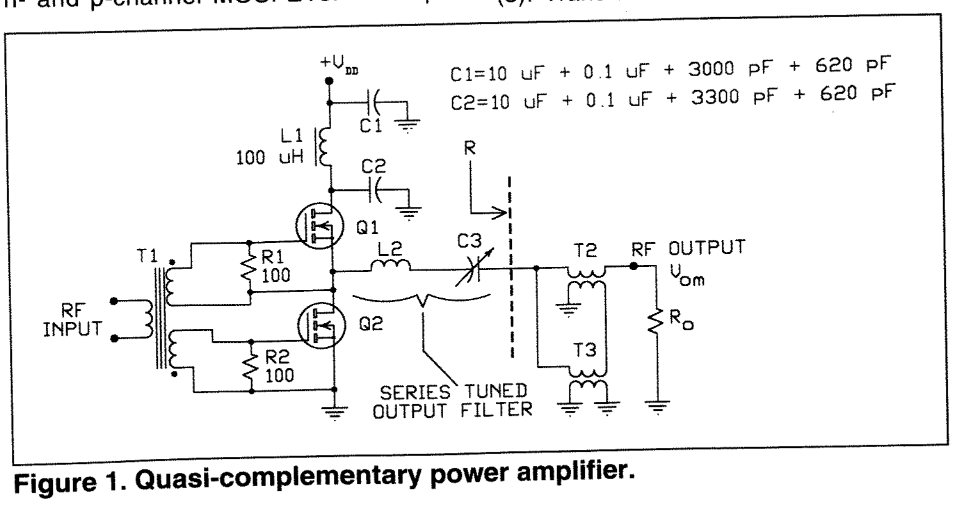

Figure 1. Quasi-complementary power amplifier.

Principles of Operation

The complementary voltage switching configuration (Figure 1) is the least complicated class-D power amplifier (PA) (3). Transformer T1 causes the driving signals applied to the gates of Q1 and Q2 to be of opposite phases. The driving signal is more than sufficient to support the drain current. Consequently, Q1 is on when Q2 is off and vice versa. The pair of FETs therefore forms the equivalent of an SPDT switch and generates the rectangular voltage waveform $v_{D2}(\theta)$ shown in Figure 2.

Maximum fundamental frequency output is obtained with a 50 percent duty ratio. Expansion of the square wave drain voltage into a Fourier series yields:

$$V_{D_2}(\theta) = V_{DD}\left(\frac{1}{2}+\frac{2}{\pi}\sin\theta + \frac{2}{3\pi}\sin 3\theta + \frac{2}{5\pi}\sin 5\theta + \dots\right), \quad (1)$$

where $\theta = \omega t$.

The output of the PA is determined by the response of the load network to each of the components in the applied voltage waveform $v_{D2}(\theta)$. The load network consists of the load resistance R and a series-tuned tank circuit (L2-C3). The tank circuit is assumed to have a reasonable loaded Q (i.e. ≥ 3), to have negligible loss, and to be resonant at the fundamental (switching) frequency. Under such conditions, the input impedance of the load network at DC and at the harmonics of the switching frequency is very high. The amplitudes of the harmonic currents are therefore very low and their effect upon PA operation is negligible. (Additional filtering may be necessary for other purposes, however, such as prevention of radiation of the harmonics by an antenna.)

If the tank circuit is tuned to resonance at the fundamental frequency, its net reactance is zero at that frequency. The fundamental frequency component of the switching waveform is then applied directly to load R, producing the output voltage and current waveforms:

$$v_o(\theta) = V_{om}\sin\theta = \frac{2}{\pi}V_{DD}\sin\theta, \quad (2)$$

and

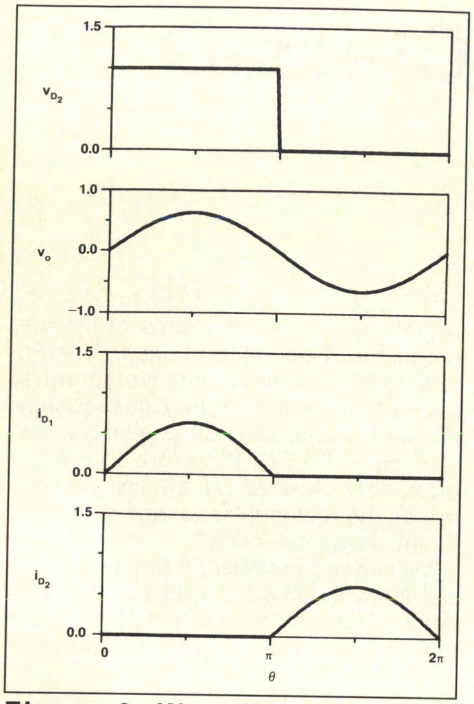

Figure 2. Waveforms of ideal complementary class-D PA.

$$i_o(\theta) = v_o(\theta)/R = I_{om}\sin\theta. \qquad (3)$$

The corresponding power output is:

$$P_0 = \frac{V_{om}^2}{2R} = \frac{2}{\pi^2}\frac{V_{DD}^2}{R} \cong 0.203\frac{V_{DD}^2}{R} \qquad (4)$$

The sinusoidal output current flowing into the load network is drawn through whichever FET is on. When the tank circuit is tuned to have zero reactance at the fundamental frequency, both drain current waveforms are half sinusoids, as shown in Figure 2 ($V_{DD} = R = 1$).

The DC input current is obtained by averaging $i_{D_1}(\theta)$ over one RF cycle, thus,

$$I_{dc} = \frac{1}{\pi}I_{Dm} = \frac{1}{\pi}I_{om} = \frac{2}{\pi^2}\frac{V_{DD}}{R} \qquad (5)$$

The DC input power is therefore:

$$P_i = V_{DD}I_{dc} = \frac{2}{\pi^2}\frac{V_{DD}^2}{R} = P_0 \qquad (6)$$

Since the DC input power and RF output power of this PA are equal, its efficiency is 100 percent ($\eta = 1$). The 100 percent efficiency is consistent with zero dissipation in the FETs; for both FETs, either the drain voltage or the drain current is zero at all times.

The efficiency of a class-D PA is less than unity because of three principal factors:

- On-resistance of the MOSFETs,
- Switching time of MOSFETs, and
- Drain capacitance of the MOSFETs.

The effects of on resistance, R_{on}, can be incorporated into the equations of an ideal amplifier by using the effective supply voltage,

$$V_{eff} = \frac{R}{R+R_{on}}V_{DD}, \qquad (7)$$

in place of V_{DD} except in the calculation of P_i.

Finite switching time reduces the output voltage by the factor $\tau^2/6$, where τ is the rise/fall time converted into radians (Figure 14-6, (3)). Its effects are incorporated into design calculations by including this factor in Equation 7.

The DC input current required to deliver power to the output is given by Equa-

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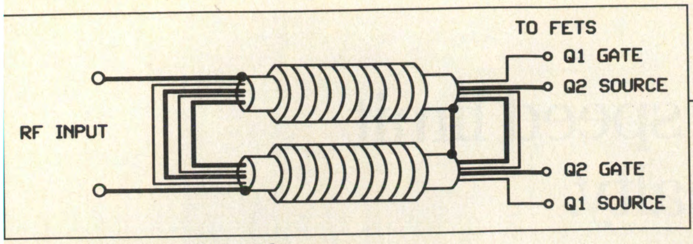


Figure 3. Input transformer.

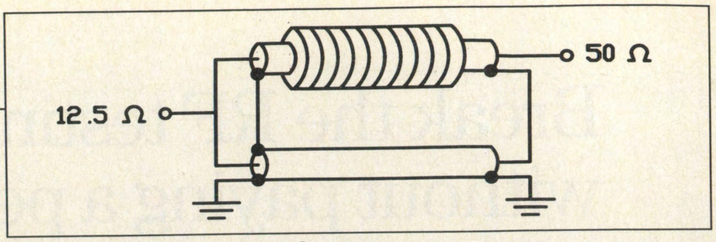


Figure 4. Output transformer.

tion 5. The DC input current required to charge a fixed total drain capacitance C_F is:

$$I_{dc_1} = f\,C_F\,V_{DD}^2 \qquad (8)$$

The DC input current required to charge the total voltage-variable drain capacitance (4) is:

$$I_{dc_2} = 2fC_{V_0}\,\Phi\left[\left(1+2V_{DD}/\Phi\right)^{1/2} - 1\right] \qquad (9)$$

where $C_{V0} + C_F$ is the capacitance at $v_{DS} = 0$ and $\Phi \approx 1$ is the barrier potential.

It is important to note that load reactance ideally does not reduce efficiency in class-D operation. In practice, the larger peak currents that circulate between the power supply and the load (Figure 14-4 of (3)) cause larger losses in the FET on resistances. These losses are, however, a second-order effect and are generally considerably smaller than those in a class-B PA.

The bandwidth of a class-D PA is limited by the bandwidth of its output filter. Since the filter must reject the second harmonic of the lowest frequency while passing the fundamental of the highest frequency, it can have a bandwidth ratio no larger than 2. In practice, filters with bandwidth ratios of up to 1.5 or so can be synthesized.

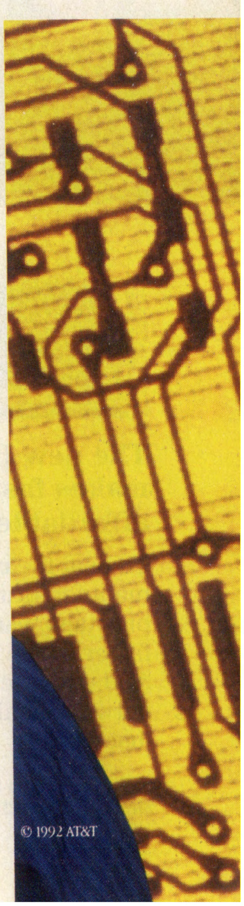
In class-D operation, the FETs are overdriven and linear amplification of the driving signal is impossible. However, class-D amplifiers have excellent amplitude modulation linearity and can be used as part of a linear amplifier system through Envelope Elimination and Restoration (EER) (5). In EER, the hard-limited carrier, which contains phase information, is amplified by highly efficient but nonlinear RF power amplifiers. The signal envelope (amplitude information) is amplified by a highly efficient class-S audio amplifier (3) and remodulated onto the carrier at the final power amplifier.

Circuit

Field effect transistors are preferred to bipolar junction transistors in a class-D PA because they can pass current in the reverse direction, which is essential for operation with a reactive load. In practice, this characteristic means that FETs are reliable in HF class-D PAs, whereas BJTs are not. MOSFETs have been used successfully in a number of class-D PAs operating at HF and above (6 - 9).

Equation 5 shows that delivery of 25 W to a 12.5 ohm load requires FETs with ratings of 39.3 V and 2.0 A. Delivery of 100 W similarly requires ratings of 78.5 V and 4.0 A. Allowance for loss due to on-resistance (R_{on} = 2 ohms) in-

iming recovery unit.



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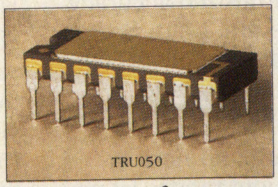

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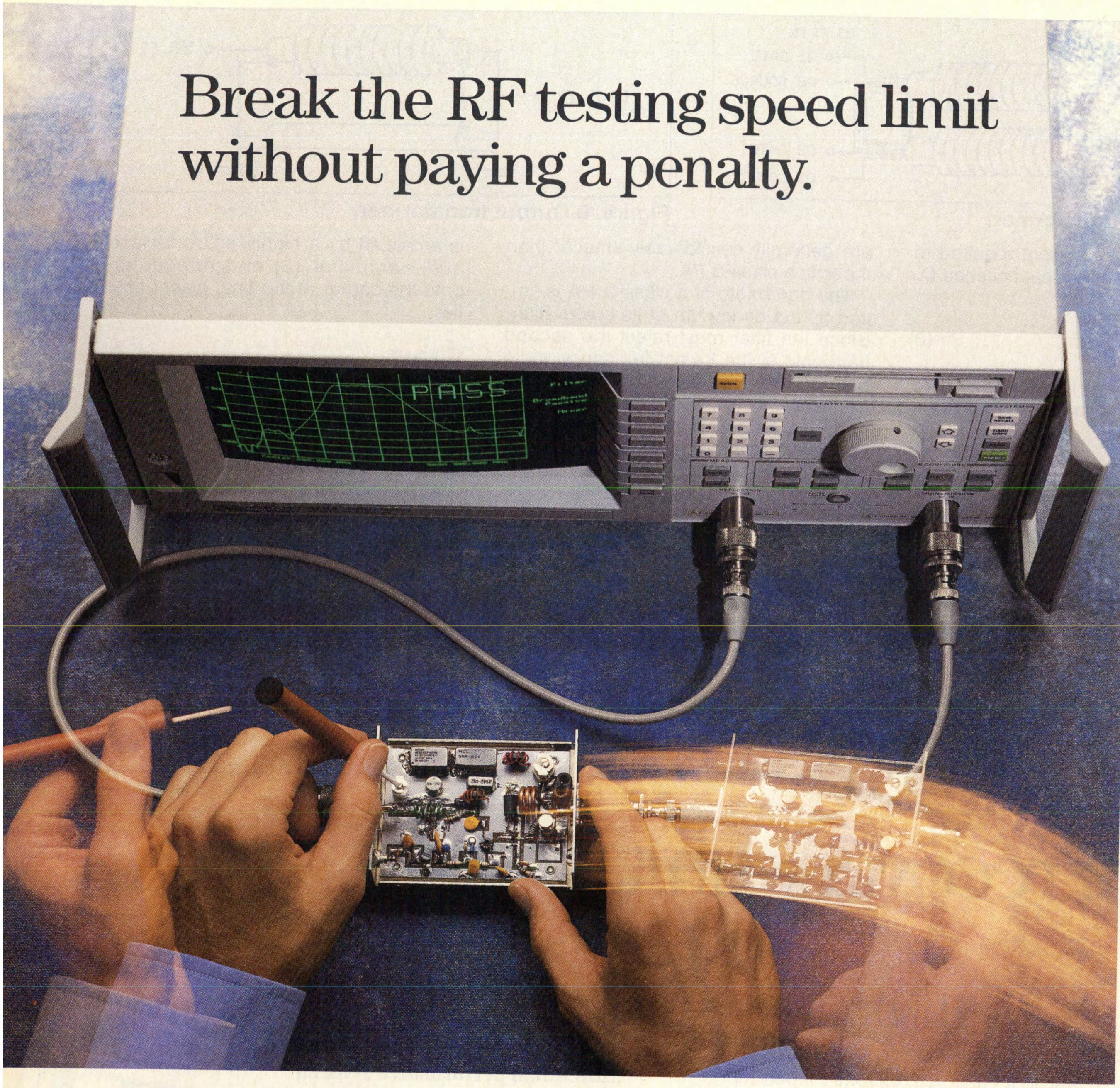

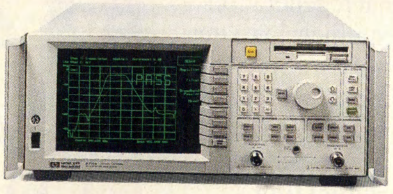

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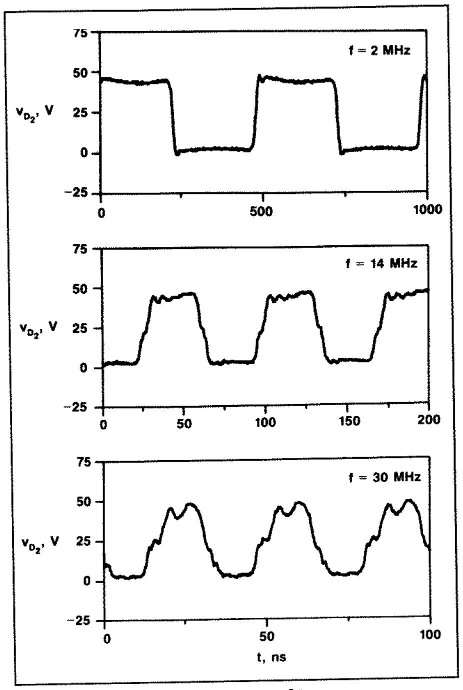

Figure 5. Drain voltage waveforms.

creases the voltage ratings by about 16 percent to 45.6 and 91.1 V, respectively. Since increased voltage ratings are associated with larger values of R_{on} and increased current ratings are associated with larger values of drain capacitance, it is desirable to select FETs with ratings only slightly larger than necessary.

The MRF136 and MRF148 RF power MOSFETs are therefore used for the 25-W and 100-W amplifiers, respectively. The standard SOE package is intended for transformer coupled, grounded source applications and is therefore somewhat awkward in this circuit. Interconnection is accomplished by mounting one MOSFET at a 45 degree angle with respect to the other MOSFET so that the source lead of Q1 meets the drain lead of Q2.

The MOSFET gate is a rather inhospitable load. As the MOSFET approaches saturation, v_D drops below v_G and the drain-gate capacitance abruptly inflates by a factor of eight or so. Resistors R1 and R2 swamp this effect and provide the driver with a reasonably well behaved load. They are also essential for maximum bandwidth of the input transformer.

The most difficult and critical design problem in the implementation of the quasi-complementary PA is the input transformer. The reasons include:

- Large bandwidth,
- Unusual arrangement of secondary-windings,
- Capacitance to ground must be charged during switching, and
- Common mode reactance diverts drain current from output.

Four approaches to the input transformer are:

- Twisted windings on a single toroid,
- Three baluns configured to drive equal current into each MOSFET,
- Three baluns configured to apply equal voltage to each MOSFET, and
- Brass tube transformer with two secondary windings.

The brass tube transformer (Figure 3) provides better overall performance. It achieves a relatively large bandwidth by choking common mode current. The use of separate windings provides a higher common mode impedance than is possible in a balun with cores of the same size.

The brass-tube transformer is built from two 0.48-cm (3/16-in) brass tubes connected in series through ten Ceramic Magnetics CMD5005 cores (9.525 × 4.763 × 4.763 mm) with two 2-turn windings of 24-AWG enamel-coated wire. Its common mode impedance of about 1000 ohms at 3.5 MHz results in a one percent increase in supply current. The driving impedance is about 12.5 ohms when used with the gate swamping resistors shown in Figure 1. It keeps gate-to-gate phase error below 20 degrees for frequencies from 1 to 60 MHz. (A comparable transformer with a 50 ohm input impedance can be fabricated by using only a single turn for each secondary winding and adding one or two more cores to each stack).

Power supply decoupling is achieved by C1, C2, and L1. The bandwidth of this filter determines the maximum amplitude modulation bandwidth. It is essential that C2 provide a good RF ground. It must also be able to store enough charge to feed the output current for half of a cycle (at the lowest operating frequency) without creating voltage drop on the drain of Q1.

The output filters are simple series-tuned circuits with Q = 5. Capacitor C3 is a combination of an air variable and mica capacitors. The inductors (L2) are wound on various toroids and have unloaded Q in the range of 70 to 200.

The output transformer (Figure 4) is an equal delay configuration (10) of two baluns that provides a 4:1 reduction of the load impedance. Both T2 and T3 employ 25 ohm transmission line formed by connecting two RG-196/U 50 ohm lines in parallel. Ten Ceramic Magnetics

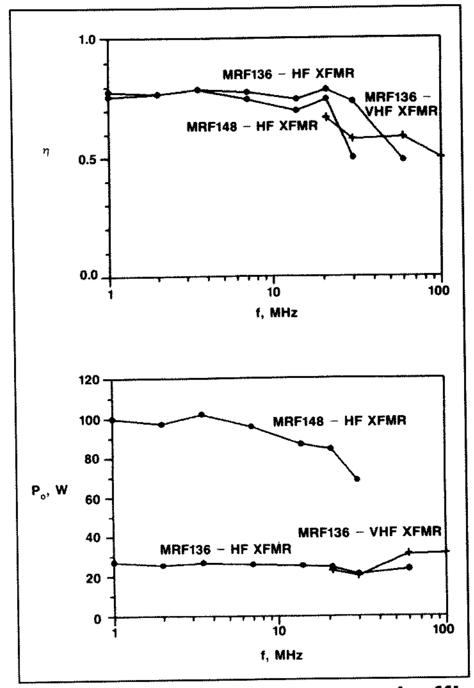

Figure 6. Output power and efficiency.

CMD5005 beads are placed over the transmission line of T2. The bandwidth of the T2-T3 combination is sufficiently large that it can be located either before or after the series-tuned filter.

Performance

The drain-voltage waveforms (MRF136) at 2, 14, and 30 MHz are shown in Figure 5. The waveform is quite rectangular at the lower frequencies, but its top contains significant ringing at the upper frequencies. The ringing is the result of interaction between the drain capacitance and the input transformer inductance, and varies with the load reactance.

The output power and efficiency are shown in Figure 6. With MRF136 MOSFETs, the PA delivers 25 W from 1 to 60 MHz. Its efficiency is 75 percent or better from 1 to 30 MHz and then drops to about 50 percent at 60 MHz. With MRF148 MOSFETs, the PA delivers 85 to 100 W from 1 to 20 MHz and about 75 W at 30 MHz. Its efficiency is 70 percent or better until 20 MHz and then drops to about 50 percent at 30 MHz. (Power output measurements are corrected for the losses in the output filter and transformer).

When the MRF148 is used, the maximum operating frequency is limited by the FETs. However, when the MRF136 is used, the maximum operating frequency is limited by the input trans-

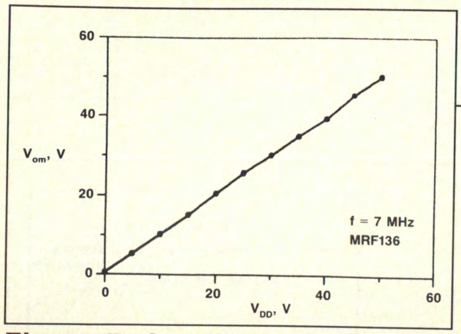

Figure 7. Amplitude modulation linearity.

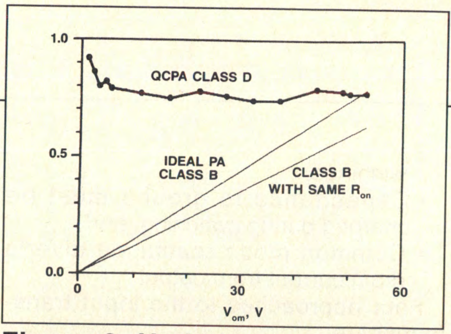

Figure 8. Variation of efficiency with output voltage.

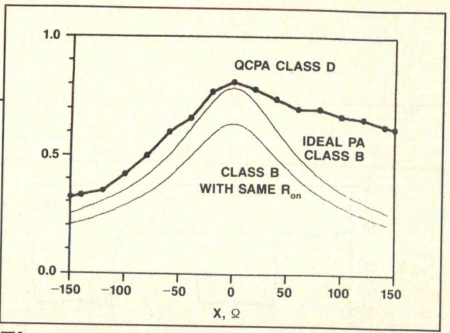

Figure 9. Variation of efficiency with series load reactance.

former. When the input transformer is replaced with one tailored for operation in the VHF range, operation can be extended to 100 MHz (Figure 6).

The amplitude modulation linearity of the 25 W PA at 7 MHz is shown in Figure 7. The rms deviation of the transfer function from a straight line is only 0.7 percent, which is roughly equivalent to IMD products at –43 dBc. With $V_{DD} = 0$, drive feedthrough produces an output voltage of 0.5 V (–40 dBc).

The variation of efficiency with output voltage for 7 MHz is shown in Figure 8. Comparison of the measurements to the curve for an ideal class-B PA shows the advantage of class-D for signals with large peak-to-average ratios. The apparent increase in efficiency near zero output is believed to be due to drive feedthrough.

The variation of efficiency with load series reactance at 7 MHz is shown in Figure 9. For these measurements, MRF136 MOSFETs are used and V_{DD} is readjusted to avoid exceeding the voltage ratings ($P_o = 10$ W for $-100 \le X \le +80$ ohms and $P_o = 5$ W for $-150 \le X \le +150$ ohms). The benefit of class-D operation over class-B operation into reactive loads is apparent.

The experimental PA is driven by an ENI 525L laboratory power amplifier. The drive level is set to produce maximum efficiency and the most square waveforms; no attempt is made to minimize the driving power. For MRF136s, the drive power varies from 5 W at 1 MHz to 7 W at 30 MHz. The resultant sine-waves of 20-V peak on the gates cause dissipation of 4 W of the drive power in the gate swamping resistors. Little additional power is required to produce a 100 W output with the MRF148 MOSFETs.

Conclusions and Recommendations

The experimental quasi-complementary class-D PA described here demonstrates the ability of a MOSFET class-D PA and the quasi-complementary configuration in particular to deliver useful power with high efficiency across the entire HF and lower VHF bands. The modulation linearity is more than adequate for implementation of a linear amplifier system via envelope elimination and restoration.

An obvious area for future development is packaging of the MOSFETs for quasi-complementary operation. Such a package would eliminate the somewhat cumbersome interconnection of standard SOE packages. The resultant reduction in lead lengths should also improve the VHF characteristics of the PA.

Improving the input transformer could also lead to an increased maximum frequency of operation. Development of MOSFETs with higher voltage ratings would eliminate the need for the output transformer, resulting in further simplification and higher efficiency.

A most interesting area for further development is a p-channel MOSFET for true-complementary operation. The use of a true complementary pair would eliminate the need for the input transformer, since the same driving signal could be coupled to both gates through capacitors. This configuration would also allow the use of gate bias, which would reduce the drive power by about 30 percent. Such a PA might be driven by logic-type circuits and fabricated in a single package. *RF*

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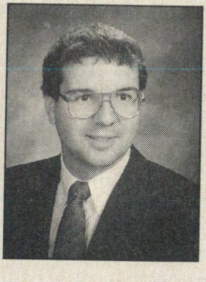
About the Authors



Fred Raab is President and Owner of Green Mountain Radio Research (GMRR). He received BS, MS, and PhD EE degrees from Iowa State University. Dr. Raab is coauthor of *Solid State Radio Engineering* and is the author of over 50 technical papers.

Dan Rupp is an electronics engineer with GMRR. He received his BSEE degree from Iowa State University in 1990, where he received the William L. Everitt Award for Excellence.

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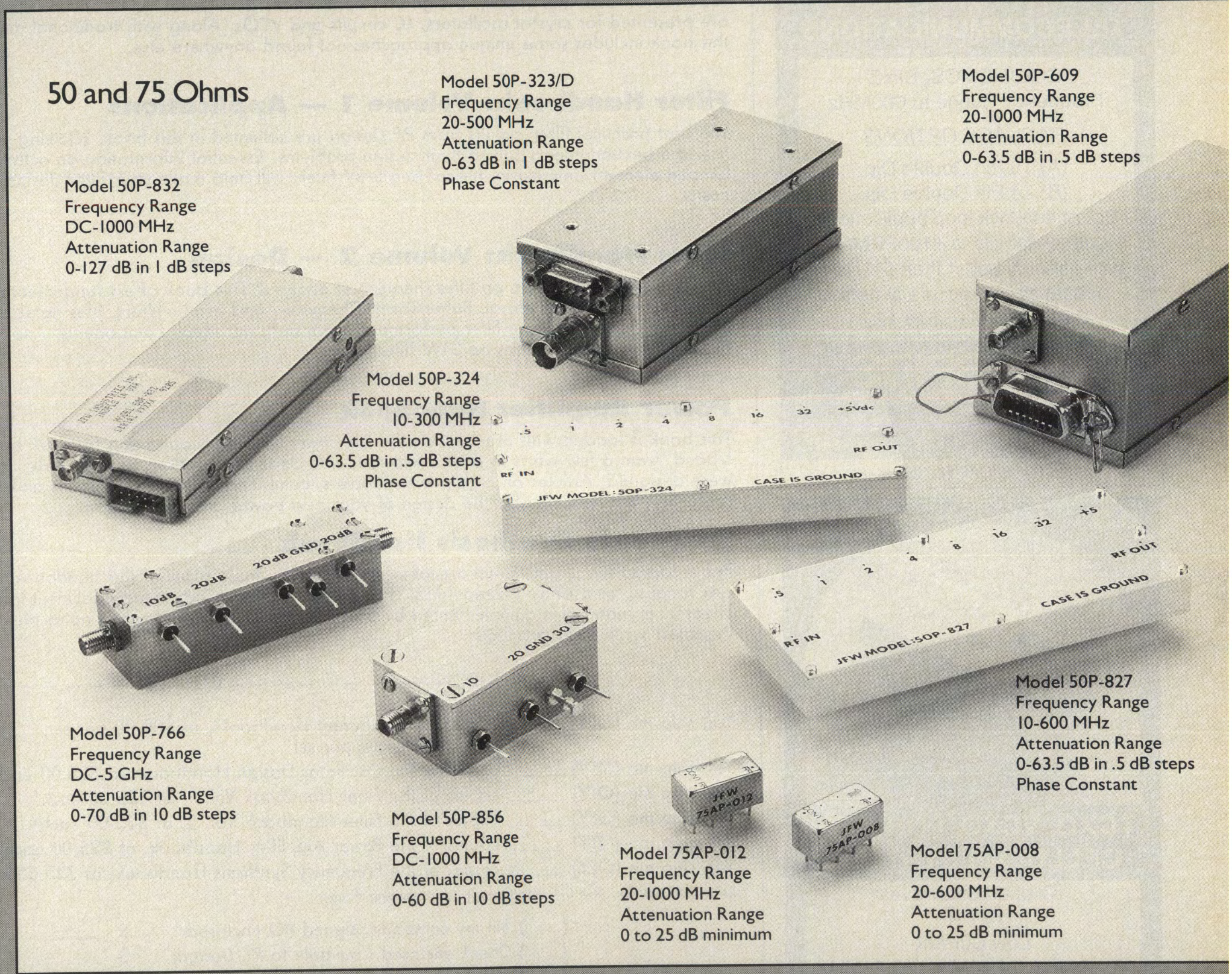

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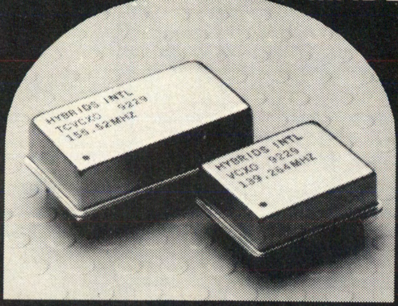

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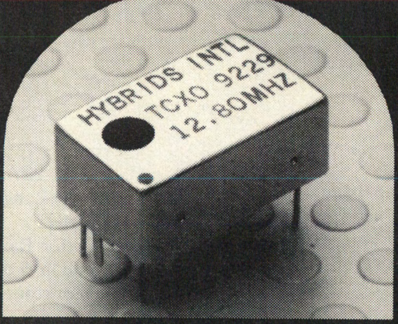

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Do you need to brush up on filter theory and analysis? This book offers fundamental and advanced material on classic Butterworth, Chebyshev and elliptic filters, plus notes on filter implementation, including filter performance with real, not ideal, components. Another highlight is a tutorial series on SAW filter basics.

Power Amplifier Handbook

This book is loaded with practical circuits for power amplifiers operating from HF through L-band, from a few watts to over a kilowatt, with clear explanations of how these circuits were designed. Articles on high power couplers, combiners, biasing techniques and VSWR protection will help simplify the design of your next power amplifier system.

Frequency Synthesis Handbook

Phase locked loops and direct digital synthesis are the main focus of this handbook, with articles ranging from Andy Przedpelski's ''PLL Primer'' series to advanced analytical techniques. Theoretical material is complemented by practical circuits and application notes on some of the latest synthesizer products.

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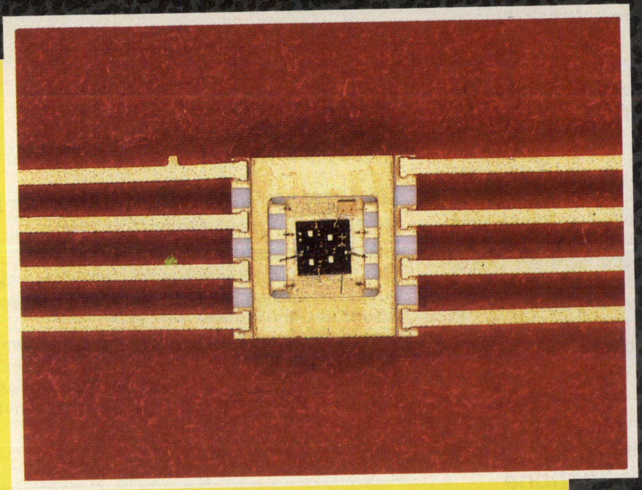

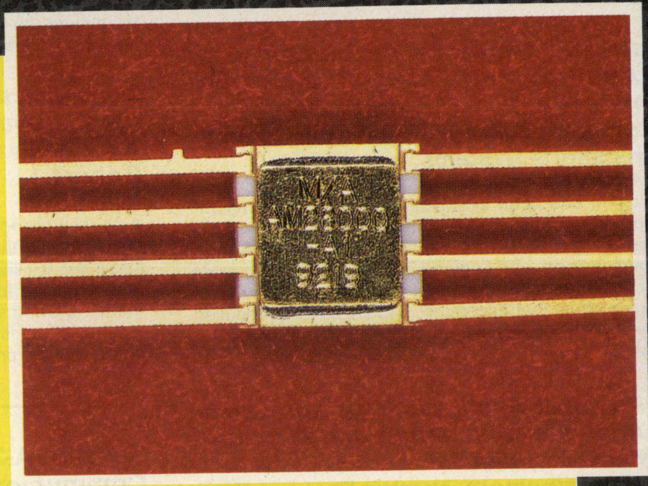

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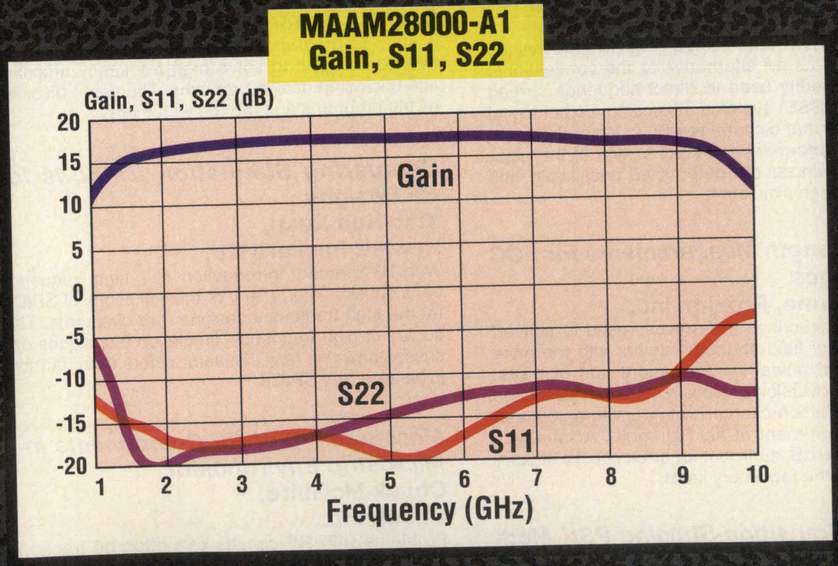

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TUESDAY, SEPTEMBER 22
8:30-11:30 A.M.

Session A-1: CAD Techniques I

RF and Microwave Circuit Design by Addition and Removal of Components
Kiomars Anvari, Teknekron Communications Systems, Inc.
A new technique for optimization of electrical networks has been developed in which the computer program removes components or adds components and nodes in a systematic manner. Although assessment of a large number of alternative circuits consumes a large amount of computer time, examples show that acceptable circuits can be produced with little operator time and reasonable computer time.

Practical Applications of Non-Linear Analysis, Simulation and Measurement Techniques
Joel Dunsmore, Hewlett Packard Co.
Modeling the linear response of RF and microwave amplifiers, such as gain or input match, has been common practice for many years. Similarly, vector network analyzers (VNAs) are used to measure S parameters, allowing displays such as the Smith chart. Further integration of these systems is more important than ever.

Practical Object Oriented CAE for RF Engineers
Dale A. Teets, P.E.
Object oriented techniques simplify the creation for powerful custom CAE in Macintosh Windows with or without C++ programming. This paper explains the basics of object oriented CAE (OOCAE) and the basic components of a OOCAE framework.

Session A-2: Medical and Scientific Applications

ATS-3: Celebrating 25 Years of Service in Space
Michael A. Cauley,
NASA Lewis Research Center
NASA's Application Technology Satellite III was launched on November 5, 1967 and remains active today as a communications satellite. This paper outlines the technical characteristics of the ATS-3 satellite and describes the current ATS Experimenters Program.

Signal Synthesis for Controlled Particle Extraction from High-Energy Accelerators
H. Meuth, G. Heinrichs, H. Halling, Forschungszentrum Julich
In circular accelerators, particle revolution frequencies fall in the range of radio frequencies. Imposing suitable RF signals on the particles is one of the principal schemes for interacting with them. This paper describes sophisticated frequency synthesis techniques for generating unique signals for controlled particle extraction.

Noncontacting Coal and Rock Thickness Measurement with a Vector Network Analyzer
Robert L. Chufo, U.S. Bureau of Mines
A noncontacting electromagnetic sensor has been developed by the U.S. Bureau of Mines that measures the thickness and dielectric constant of coal and rock and locates the interface in multilayer media. Moving the sensor antenna accomplishes the spatial modulation that, through signal processing, solves the problem of dispersion.

Session A-3: Wireless Communications Applications

A Rapid Acquisition Technique of Spread Spectrum Signals Embedded in an ASIC Costas Phase Locked Loop
G.G. Koller, M.A. Belkerdid, University of Central Florida
G.S. Rawlins, Signal Technologies, Inc.
A relatively new rapid coarse acquisition technique is proposed as an alternative to the conventional models currently used in direct sequence spread spectrum (DSSS) systems. It is comprised of digital components that conform readily to integrated technologies. Performance characteristics of the algebraic synchronizer are determined analytically and verified through simulation.

Field Strength Measurements for FCC Compliance
Earl McCune, Proxim, Inc.
This paper describes the process used to convert the regulatory field strength number into the more conventional power measurement unit of watts. From this conversion process a method is described where any person can calibrate a given antenna to allow measurement of field strength. Accuracy is good to a few dB, sufficient for quick checks of compliance with the regulatory limits.

Cosine Transition-Shaping PSK Modulators Using Direct Digital Synthesizers
Roy Greeff and Bruce Williams, Paramax Systems Corp.
This paper discusses an easy implementation for cosine transition-shaping BPSK and QPSK modulators using direct digital synthesizers (DDS). A common major constraint placed on these signals is bandwidth, and these modulation techniques need filtering. DDS offers additional techniques to simplify the design of band-limited modulators.

Session A-4: A/D Converter Tutorial

Understanding Data Converter Frequency Domain Specifications
Robert E. Leonard, Jr., Datel Inc.
This three-hour tutorial session is intended to provide an understanding of frequency domain specifications, as applied primarily to precision A/D converters. Numerous applications have performance requirements that are heavily dependent on data converter performance, and engineers must be able to differentiate between various converter products.

TUESDAY, SEPTEMBER 22
1:30-4:30 P.M.

Session B-1: CAD Techniques II

RF Design with the HP-48
David J. Brunell
A suite of programs for the HP-48 calculator has been developed to solve high frequency amplifier and microstrip design problems. The paper discusses the programs and the relevant theory.

Comparing Simulation Methods for RF Designs
Xiao Hua Xuan,
Hewlett Packard Co.
With HP's recent introduction of a high frequency transient simulator, many of the limitations of SPICE for the high frequency designer are overcome. This survey of high frequency simulation techniques discusses how this new simulator differs from harmonic balance and SPICE.

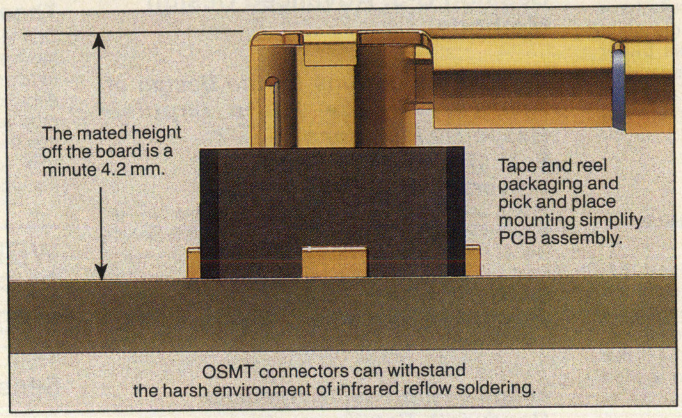
Measurement of RF Components in a Microstrip Environment
Chuck McGuire,
EEsof, Inc.
Problems with RF circuits can often be traced to stray parasitics. It may impossible to measure these effects without proper calibration of the test equipment. This paper describes a method for building and using LRL calibration in the microstrip medium of the engineer's choice.

Until now, no one has produced surface mount connectors with greater frequency.

Now all of the advantages of surface mount technology are available for applications that require microwave coaxial connectors with frequencies of up to 6 GHz.

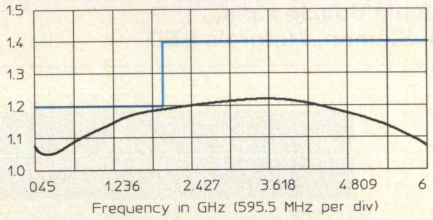
Surface mount technology offers a number of advantages over traditional through-hole connectors. It reduces the required amount of PCB real estate, allowing for greater component density. The smaller, lighter connectors resist shock and vibration better. And surface mount technology gives OEMs the benefit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

But until now, true surface mount connector technology did not cover the broad spectrum of microwave frequencies. Now we have an interconnect solution for every application, in every market.



Their mated height off the board is a minute 4.2 millimeters. And they use substantially less PCB real estate than standard through-hole connectors, which allows for denser packaging and results in smaller, lighter PCBs.

Despite all of their technological advantages, OSMT connectors don't require costly new equipment or placement procedures. In fact, they can be installed using standard surface mount processes. The connectors are packaged to be ESD safe, are available in tape and reel packaging, and are pick and place mountable. If you're equipped for SMT, OSMT won't increase your expenses at all. In fact, the smaller size, higher quality and improved performance of these connectors will ultimately lower your installed costs.



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Session B-2: Complex Modulation

A Network Theoretic Approach to the Reduction of Distortion in Quadrature Detectors

Thomas G. Xydis, Ph.D., Consultant

Many low cost FM receiver integrated circuits have multiple functions integrated onto a single chip. Most use quadrature detectors, and this paper examines the linearity of the frequency-dependent tank circuit and applies network synthesis and optimization techniques to the design of new tank circuits.

The Vector Discriminator

Billy P. Ficklin, Marilyn F. Williams, SRI International

A vector discriminator has been developed that allows accurate demodulation of FM signals and allows detection of small frequency changes without long integration times, DC drifts, or FFT methods. The simplicity of the processing circuits in the digital realm makes the technique applicable to a wide range of problems.

Third Generation Spread Spectrum S-Band Transponder

M. C. Comparini, F. Marchetti, Alenia Spazio s.p.a.
M. Siemon, Motorola SED

A new generation S-band transponder has been developed to support the specific modulation schemes required to operate with the European Data Relay Satellite (DRSS) and to assure compatibility with the existing Tracking Data Relay Satellite System (TDRSS).

Session B-3: Advanced Design Methods

Iterative Algorithms for the Design of Arbitrary Phase Finite Impulse Response Functions

S. M. Ritchie, R. P. Burke, University of Central Florida

Phase compensation and delay equalization in RF systems requires the introduction of arbitrary phase filters. For digital or analog transversal filters such as SAW devices, an arbitrary phase finite impulse response (FIR) function must be synthesized. Several algorithms are discussed and an example filter is presented.

Image Compression Reduces Bandwidth Requirements

Michael Ellis, U.S. Army Corps. of Engineers

This paper reviews the standard lossless and lossy techniques that comprise image compression algorithms. The most important components of the Joint Photographic Experts Group (JPEG) algorithm are targeted including Huffman encoding and the Cosine Transform. LZW, Singular Value Decomposition, fractal compression and other techniques are also discussed.

High Temperature Superconductivity (HTS) and its Role in Electronics Warfare

Thomas N. Tuma, U.S. Army CECOM

During the past eight years, the quest for materials having the property of superconductivity at "high" temperatures has yielded astonishing results. The debut of HTS components for RF was in the design of a variety of microwave equipment. It is evident that this phenomenon has to be exploited at HF and VHF, for sensitive receivers and high power, smaller transmitters.

WEDNESDAY, SEPTEMBER 23 8:30-11:30 A.M.

Session C-1: RF Systems I

Choosing the Optimum Synthesizer Architecture for your Receiver Application

Larry Kimbrough, Harris GCSD

Synthesizer design is limited by the fixed parameters of frequency range and step size, and by four

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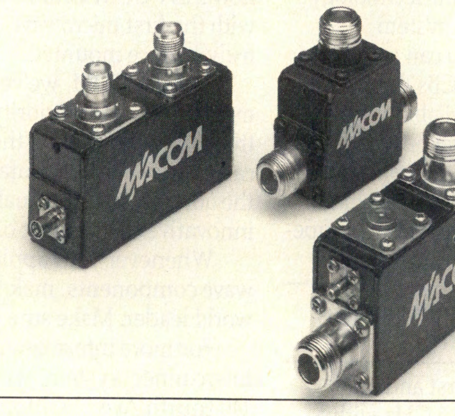

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more flexible parameters. these are phase noise, output spurious, settling time, and size/power. This paper examines the performance of various designs for a given receiver specification.

CW Rejection
Sherman R. Vincent, Raytheon Co.

In receivers adapted to process received RF pulses, strong CW signals may interfere with reception of the desired signal, especially when the desired and interfering signals are close in frequency.

Optimizing Performance of Collocated UHF and VHF Ground to Air Radios
David S. Penney, MITRE Corporation

This paper addresses the problem of collocated VHF and UHF transmitters and receivers in the same ground to air communications facility. Each interference mechanism is defined and the effects of these mechanisms on system performance are described.

Session C-2: Filter Design

IF Transversal Filtering
Richard D. Roberts, Harris Corp.

IF transversal filtering, consisting of a periodically tapped delay line, has a wide variety of uses such as gain slope compensation or adaptive equalization of digital signals. This paper looks at applications and implementations of IF transversal filters up to 6 GHz with baud rates in excess of 1 Gbaud.

Catalog of Rhodes Filter Transfer Functions and Element Values
William B. Lurie, Consultant

Filters based on the Rhodes linear-phase polynomial have received some small attention in the literature. They are interesting because they offer as a compromise a somewhat flat amplitude and delay characteristic over 75-80 percent of the passband. Practical considerations are included in the discussion.

Synthesis of Low-Pass Filters Containing Quads of Zeros
Mark Mell, Microsonics, Inc.

This paper details the calculations required to synthesize a special type of lowpass filter, one that contains quads of zeros. Including a quad of zeros with any existing set of pole locations has the effect of modifying only the amplitude response. Absolute delay is increased while the selectivity of the filter is improved.

Session C-3: Oscillators and Synthesizers

Methods of Frequency Control,
David J. Brunell

This paper describes the basic methods of frequency control, with the goal of maintaining 30-50 parts per billion accuracy compatible with portable paging products. Oscillators and frequency synthesis methods are covered, with attention to an AFC method developed by the author.

Self-Adaptive VCO
Wojciech Klimkiewicz, SSDD Research Corporation

In this paper, an FM VCO design is presented that utilizes a digital signal processor (DSP) to achieve high quality modulation over changing environmental conditions and aging. Practical and theoretical explanations of its performance are presented.

A CAD and Optimization Program for Microwave Oscillators
A. Heitbrink, B. Roth, A. Beyer, Duisburg University

A CAD method is presented which is able to calculate all the important details such as operating frequency and power of harmonics for integrated microwave oscillators. Calculation schemes and important characteristics are discussed.

WEDNESDAY, SEPTEMBER 23 1:30-4:30 P.M.

Session D-1: RF Systems II

A Simple Road Information Transmitter
Thomas Hack, Comlinear Corporation

The 1992 RF Design Awards Contest design category winner, this circuit uses DDS principles with a specialized waveform map to generate modulated signals at 10 kHz intervals over the 540-1600 kHz AM broadcast band. Methods for extending this principle to FM broadcasting are also discussed.

Waveguide and Coax Components for High Power Broadcasting
Jim Stenberg, Passive Power Products

High power transmitters present a unique set of requirements for combiners, diplexers, filters and loads. These components must carry large average power and withstand even larger peak power. The design, manufacture and test of these components is discussed in this paper.

An Image Improvement of Microwave Diffraction Tomography for 2-Dimensional Inhomogeneous Dielectric Cylinder Based on Projection Function
K.W. Suh, D.Y. Lee, Y.C. Han, C.Y. Park, J.W. Ra, SAMSUNG Electronics

An alternative to the conventional Fourier diffraction tomography, the diffraction slice-projection algorithm has been developed. The results of computer simulations illustrate the validity and usefulness of the presented algorithm based on a projection function under the Born approximation.

Session D-2: Crystal and SAW Filters

Crystal Filters Having Superior Intermodulation Characteristics and the Measurement of Low Level Intermodulation in an Swept Frequency Mode
M.D. Howard and W.F. van den Akker, Piezo Technology, Inc.

Many modern receivers, and some radar systems in particular, require levels of intermodulation performance that exceed that of previously available products. This paper describes filters that have been designed and are being manufactured for these requirements. A factor contributing to success is the development of an intermodulation test system capable of detecting IM products for out of band signals below -140 dBm.

Extensions of the Holt & Gray Crystal Filter Design Technique
William B. Lurie, Consultant

The Holt and Gray transformation was developed to make bandpass crystal filters as cascaded lattices from a lowpass prototype. But without modification, the element values tend to have a wide spread. A technique to narrow that spread is presented, with an example.

Modular Implementation of the Coupling of Modes Analysis for Surface Acoustic Wave Resonators
S.M. Ritchie, R.P. Burke, University of Central Florida

SAW resonators can be used to form high Q, extremely narrowband resonators and resonator filters for application in the 20 to 2000 MHz range. Recently the application of coupling of modes (COM) formalism has been applied to these devices, which is discussed via its application to the SAW reflective grating and transducer to obtain a mixed-matrix representation.

Session D-3: Components for New Applications

A Versatile Mixed-Signal Cell-Based Array for Wireless Receiver ASICs
I. Bezzam, C. Robbins, C. Vinn, Raytheon-ASIC Div.

RF/IF applications are growing tremendously in many areas of wireless communications. All of these applications require significant improvements in cost, size, and power consumption to meet high-volume commercial market needs. A high performance ASIC designed to meet these needs, a configurable RF/IF channel for a "receiver on a chip," is described in this paper.

Multiple Receiver Configurations Realizable With One Set of Standard Stanford Telecom ASICs
Herman A. Bustamante, Stanford Telecom

The use of digital signal processing has become the preferred method of modern receiver and modem design. First, digital circuits are exact and predictable, and once designed and proven, can be counted on to perform in the same manner. This note describes some basic receiver building blocks and how they can be used to design a very broad range of receiver systems.

Pushing Low Quiescent Power Op-amps to Greater Than 55dBm 2-Tone Intercept, and an Automated Very Wide Dynamic Range Measurement Approach to Assess These Excep-

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tionally Low Spurious Levels
Michael Steffes, Comlinear Corporation
This paper will outline the intrinsic benefits offered by off-the- shelf current feedbeck operational amplifiers. Low power, high intercept point and gain performance and tradeoffs will be discussed.

THURSDAY, SEPTEMBER 24 8:30-11:30 A.M.

Session E-1: High Power Applications

A 1KW RF Amplifier for UHF 470-860 MHZ Television Applications
M.J. Culling, Harris TVT
The potential benefits of all solid-state transmitters have been well considered and documented. But until recently, such transmitters at UHF were not considered economically viable. This paper describes a new 1 kW television amplifier using bipolar transistors in Class AB, with modular construction.

Design Considerations for Solid State High Power Amplifiers Used in Radar
Mahesh Kumar,
Paramax Systems Corporation
Current and future radar systems require performance beyond that which can be provided by Klystron or other tube-type transmitters. This paper will present design techniques for high power solid state amplifiers with output power around 300 watts in S-band, which are building blocks for a 10-20 kW transmitter.

Quasi-Complimentary Class-D HF Power Amplifier
Frederick H. Rabb and
Daniel J. Rupp,
Green Mountain Radio Research Co.
This quasi-complimentary power amplifier directly connects a pair of n-channel MOSFETs in a totem pole configuration, eliminating the need for a push-pull output transformer. Operating in Class-D, efficiency is 74 percent at 30 MHz for a 25-watt version. A 100-watt version has an efficiency of 70 percent at 20 MHz.

Session E-2: Small-Signal Design

An S-Parameter Based Amplifier Design Program
Donald Miller, MCC Panasonic
This paper is the winner of the software category of the 1992 RF Design Awards Contest. The program performance parameter conversions, noise, gain, and stability analysis, as well as other essential functions for the design of RF amplifiers.

Noise Measurement on Microwave Devices Using a Scalar Analyzer
Stuart A. Fox, Giga-tronics, Inc.
Noise measurements on microwave devices such as amplifiers are generally made using a noise figure meter and downconverter. This paper will discuss a Noise Figure Test Set for use with a Scalar Analyzer that utilizes the Y-Factor technique for the faster measurements needed in a production environment.

Predicting Noise Contributions in RF Circuits
Dan Pleasant,
Hewlett Packard Company
Recent improvements to state-of-the-art RF and microwave CAD tools include the capability to calculate noise effects in harmonic balance analyses. This paper discusses the general mathematical techniques used to calculate noise effects in nonlinear circuits, including practical examples of noise calculations for mixers and phase noise in oscillators.

Session E-3: Test and Measurement

What Ever Happened to the Q Meter?
Albert Helfrick,
Embry-Riddle Aeronautical University
This paper describes Q, Q measurement, a short history of the Q-meter and modern Q-meter design. Many RF engineers have used a Q-meter, and if a unit is available, still use it to verify coil designs before the inductors are placed in a circuit. This is in spite of the fact that a Q-meter hasn't been manufactured in 25 years.

Selecting the Right Frequency Domain Measuring Instrument
Robert Witte, Hewlett Packard Co.
Frequency domain measurements have been used by engineers for decades. With the increased use of the Fast Fourier Transform (FFT) in spectrum analyzers and digitizing oscilloscopes, choosing the right frequency domain measuring instrument is more difficult than before. The performance capabilities of each possible choice are examined in this paper.

Design and Optimization of Lumped-Element LCM Directional Couplers
Edmund (Joe) Tillo II
A new technique for designing directional couplers has been developed, producing couplers that are made entirely from inductances, capacitances and mutual inductances (LCM) and closely approximate the quarter-wavelength coupled transmission line directional coupler in the sinusoidal steady state.

Four Special Courses Cover Key RF Topics

RF Expo East features four full-day courses that feature experienced instructors, complete notes and reference texts, plenty of visual aids, and opportunities for questions and discussions.

Filter and Matching Network Design (Monday, September 21) — Subtitled "L-C and Distributed Circuits — HF to Microwaves," this course is intended to prove practical information on these passive circuits. Course work begins with a review of fundamental filter responses and classic topologies for filters and matching networks, followed by thorough coverage of design methods. The course is accessible to engineers with little experience in this area, yet contains useful information for more senior designers. Instructor: Randy Rhea of Eagleware, Inc.

Fundamentals of RF Circuit Design, Part I (Tuesday, September 22) — The first part of this two-day course covers basic RF concepts and passive circuits. Component models, transmission line fundamentals, impedance transformations, losses and parasitics are covered from both analytical and practical viewpoints. Extensive use of computer analysis aids in visualization of concepts. Although this is a two-day course, the organization of the material allows each to be taken independently, if desired. Instructor: Les Besser of Besser Associates Inc.

Fundamentals of RF Circuit Design, Part II (Wednesday, September 23) — The second part of this course emphasizes active components and circuits. Diode models and circuits are followed by transistor fundamentals, including gain, stability, matching, small-signal amplifiers and large-signal considerations. Physical circuit layout effects are also covered. Lots of examples are used to illustrate the essential RF information being presented. Instructor: Les Besser.

Oscillator Design Principles (Thursday, September 24) — The fundamentals of oscillator design are presented in a unified manner. Theoretical and practical aspects of design allow the engineer to select appropriate circuits or products. L-C, SAW and crystal oscillators are covered, including key performance parameters, such as phase noise, starting time, stability and harmonic levels. Instructor: Randy Rhea.

A Program for the Design of Coupled Resonator Bandpass Filters

By John G. Freed
Teletec Corporation

BANDPASS is a computer program that designs coupled resonator bandpass filters of either the Butterworth or Chebyshev type. The program is written in BASIC and is menu driven. Filters of two or more poles may be designed with either capacitive or inductive coupling between resonators and with either an L network or a capacitive tap input and output. This article describes the sequence of operations in the program, followed by an explanation of the design techniques used in BANDPASS.

To start, the program prompts the user for the type of filter, Butterworth or Chebyshev. After choosing the filter type, the program asks the user to specify the center frequency, bandwidth, ripple (if applicable), and either the number of poles required or a stopband frequency and attenuation.

If the stopband frequency and attenuation have been specified, the program computes the number of poles necessary to achieve the desired response. This is done by calculating a ratio between the center frequency and the stopband frequency and doing an iterative calculation of the filter attenuation as the number of poles is varied. When the attenuation calculated by the iterative procedure is greater than or equal to the desired attenuation, the correct number of poles has been determined.

Once the number of poles is known, the normalized lowpass elements are calculated. Since the values are calculated rather than contained in a table, any value of ripple may be chosen for the Chebyshev filter.

The user is asked for the value of inductor to be used in the resonators. Since a wide range of capacitor values is easier to obtain than a large selection of inductors, the user is able to tell the program what inductor value is available. The program calculates the capacitance value that resonates with the inductor value selected and asks the user if it is a reasonable value. If it is not, the user may enter a new value.

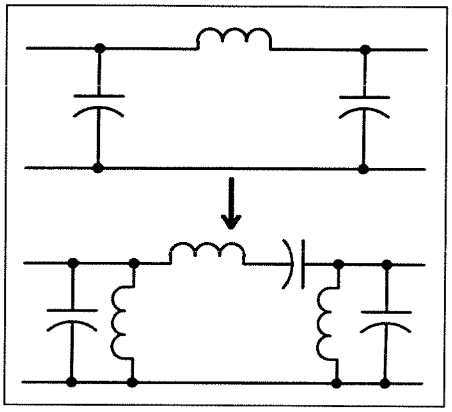

Figure 1. Lowpass filter transformed to bandpass filter.

Next, the source and load impedances are specified. The source and load impedances may be different if desired. The program checks to see if it is possible to obtain the proper end resonator loading given the inductor and load impedance choices. If everything seems reasonable, calculations continue; however, if it is not possible to achieve the correct end resonator Q, the program suggests trying a different inductor to obtain the correct Q.

Finally, the user specifies the coupling method between resonators internal to the filter and the coupling into and out of the filter. The resonators may be coupled with either a capacitor or an inductor. Input and output coupling may be done with either a series capacitor or a capacitive tap.

After the user has entered all the necessary information and the program has computed the normalized lowpass values, a lowpass to bandpass transformation is performed. A simple transformation is performed for fractional bandwidths less than seven percent while a somewhat more complex approximate transformation is performed for fractional bandwidths up to 25 percent. For fractional bandwidths greater than 25 percent the user is warned that the lowpass to bandpass mappings may not be

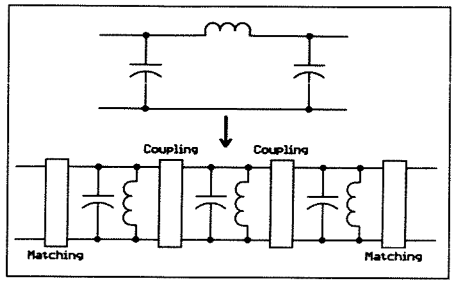

Figure 2. Lowpass filter transformed to coupled parallel resonators.

accurate and to proceed at his own risk.

Next the actual computation of filter values occurs. The series input and output coupling capacitors are calculated first even if capacitive taps have been selected. The coupling elements between resonators are calculated next. Capacitors are always calculated even if inductors have been selected. Subsequent routines convert the coupling capacitors to coupling inductors and the series input and output capacitors to capacitive taps if necessary. Finally the resonator capacitance is calculated for each resonator and the results of the filter design are displayed.

Design Techniques

The design of bandpass filters using the lowpass to bandpass transformation technique is well known and documented (1). The transformation involves series resonating each series element in the lowpass prototype and parallel resonating each shunt element. This technique is illustrated in Figure 1. In other words, each series element becomes a series tuned circuit and each shunt element becomes a parallel tuned circuit.

Less well known is the design technique for coupled resonator bandpass filters. In this method the lowpass prototype is transformed to an equivalent number of parallel resonators and a suitable network is used to couple the resonators together. Finally, an impedance transforming network may be used at

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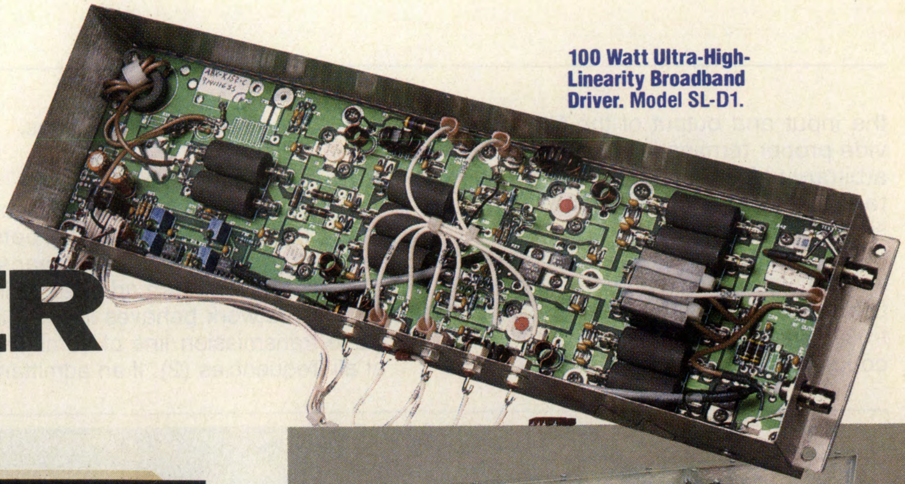

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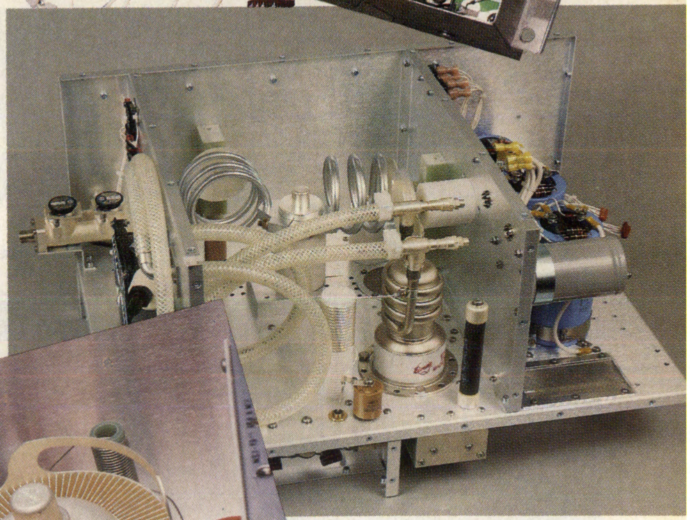

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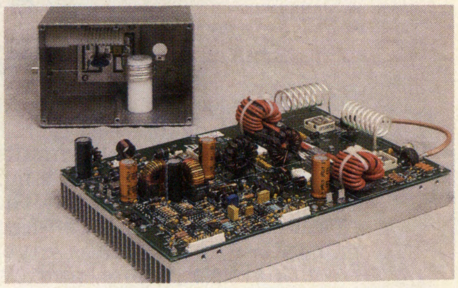

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the input and output of the filter to provide proper termination of the filter when arbitrary values of inductance or capacitance are used in the filter. This is shown in Figure 2.

The configuration shown in Figure 2 is intuitively pleasing; it is easy to visualize the response of the filter being controlled by the number of resonators, the coupling between them, and the terminations seen by the end sections.

Admittance Inverters

In the coupled resonator bandpass filter of Figure 2, the coupling between resonators is achieved by means of a network known as an admittance inverter. This network behaves like a quarter-wave transmission line of admittance J_0 at all frequencies (2). If an admittance of Y_L is connected to the output of the admittance inverter as in Figure 3, the admittance seen at the input of the admittance inverter becomes:

In practice it is difficult to realize a sim-

$$Y_{IN} = \frac{J_0^2}{Y_L} \tag{1}$$

ple circuit which has the property of an admittance inverter at all frequencies, but there are approximations that are valid over a reasonable bandwidth that may be used for narrowband designs. One such device is indeed a quarter wavelength transmission line, but this usually isn't a practical solution if the circuit must be kept to a reasonable size. Two much smaller components that exhibit the property of an admittance inverter over a narrow bandwidth are a series inductor and a series capacitor. These are well suited for use in the coupled resonator filter because they are easy to implement. In fact, in some applications, series capacitor coupling is achieved merely by the physical proximity of one resonator to another.

Input and Output Matching Networks

The matching networks used at the input and output of the filter may be of any convenient type. Two of the simplest networks are the L-match and the capacitive tap. These circuits are well known and described in References 3 and 4. Figures 4 and 5 show a simple implementation of each network and the equations needed to calculate the components required.

Determining the Number of Poles

Now that the basic circuit topology has been defined, let's look at a summary of the theory behind the design of a coupled resonator filter. The first step in the design is to determine whether a Butterworth or Chebyshev filter is required. The relative merits of each filter are well known and are covered in many references (5,6). Once the filter type has

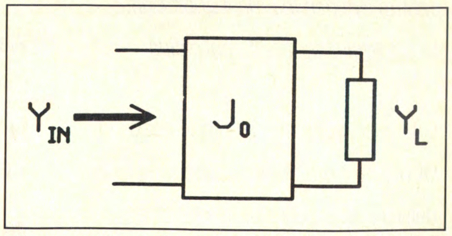

Figure 3. Admittance Y_L connected to output of admittance inverter.

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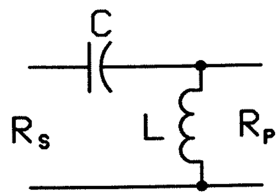


$$X_C = \sqrt{R_P R_S - R_S^2}$$

$$C = \frac{1}{2\pi f X_C}$$

$$X_L = \sqrt{\frac{R_P^2 R_S}{R_P - R_S}}$$

$$L = \frac{X_L}{2\pi f}$$

Figure 4. L-match matching network.

been established, the center frequency, bandwidth, stopband frequency, stopband attenuation, and number of poles are specified. Generally, not all these parameters are known. The number of poles is usually determined from the stopband performance required. To determine the number of poles, a normalized frequency is calculated from the specified center frequency, bandwidth, and stopband frequency. A useful equation relating these quantities in a bandpass filter is

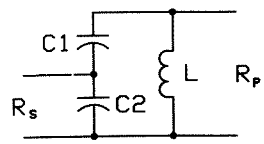
$$f_n = \frac{\left| f_s - \frac{f_0^2}{f_s} \right|}{BW} \qquad (2)$$

where:
f_n is normalized frequency
f_s is stopband frequency
f_0 is center frequency
BW is bandwidth.

The normalized frequency is then substituted into the amplitude response equation for the appropriate filter type and an iterative calculation is done by varying the number of poles until the calculated attenuation is greater than or equal to the specified stopband attenuation. For the case of the Butterworth filter, the attenuation function

$$\alpha = 10 \log\left(1 + f_n^{2N}\right) \qquad (3)$$

is evaluated for N (the number of poles) beginning with N=1 to the value of N that satisfies the stopband attenuation requirement.



$$Q = \frac{R_P}{X_L},\ X_L = 2\pi f L$$

$$X_{C2} = \frac{R_S}{\sqrt{\frac{R_S(1+Q^2)}{R_P} - 1}}$$

$$X_{C1} = \frac{R_P Q}{Q^2+1}\left[1 - \frac{R_S}{Q X_{C2}}\right]$$

$$C2 = \frac{1}{2\pi f X_{C2}}$$

$$C1 = \frac{1}{2\pi f X_{C1}}$$

Figure 5. Capacitive-tap matching network.

Calculating the number of poles for the Chebyshev filter follows the same logic as the Butterworth but the attenuation function is a bit more complicated:

$$\alpha = 10 \log\left(1 + E^2 C_n^2 \left(f_n\right)\right) \qquad (4)$$

In the Chebyshev attenuation function, C_n is a Chebyshev polynomial of order n and E is the ripple factor defined as

$$E = 10^{\mathrm{RIPPLE(dB)}/10} - 1 \qquad (5)$$

Tables of Chebyshev polynomials of various order are available in several references (7,8) and a short table for orders zero through six is presented in Table 1.

It is also possible to calculate the Chebyshev polynomial of any order by the following recursion relation:

$$C_0(f) = 1 \qquad (6)$$

$$C_1(f) = f \qquad (7)$$

$$C_{N+1}(f) = 2f \times C_N(f) - C_{N-1}(f) \qquad (8)$$

For each value of N beginning with N=1, the appropriate Chebyshev polynomial is selected, evaluated, and the result of the evaluation is substituted into the Chebyshev attenuation function which is then evaluated to see if the calculated attenuation equals or exceeds the required attenuation.

Normalized Lowpass Values

Once the number of poles needed has been determined, the normalized lowpass elements are either calculated or looked up in one of the many published tables (9). The only problem with tabular values is that in the case of Chebyshev

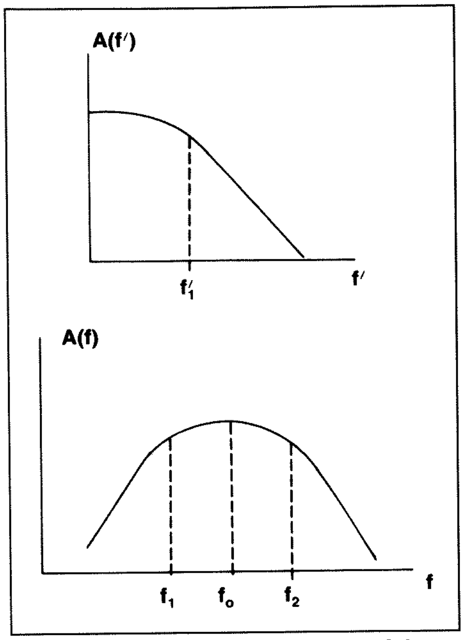


Figure 6. Transformation of lowpass to bandpass filter in the frequency domain.

filters, element values are presented only for certain ripple values; however, in most cases choosing the closest published value to one's requirement will be good enough. If it is necessary to calculate the normalized lowpass elements, there are formulas that can be solved with a calculator or programmed on a personal computer.

The normalized lowpass elements for a Butterworth filter are calculated with the following formulas:

$$g_0 = 1 \qquad (9)$$

$$g_K = 2\sin\left[\frac{(2K-1)\pi}{2N}\right] \qquad (10)$$

for K = 1 to N
N = number of poles

$$g_{N+1} = 1 \qquad (11)$$

The normalized lowpass elements for a Chebyshev filter are calculated with the following formulas:

$$\beta = \ln\left[\coth\left(\frac{\mathrm{RIPPLE}}{17.37}\right)\right] \qquad (12)$$

$$\gamma = \sinh\left(\frac{\beta}{2N}\right) \qquad (13)$$

N = number of poles

$$A_K = \sin\left[\frac{(2K-1)\pi}{2N}\right] \qquad (14)$$

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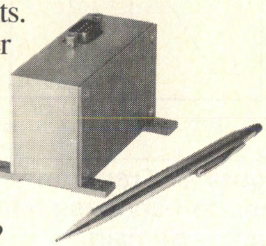

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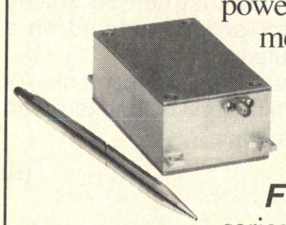

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n	$C_n(f)$
0	1
1	f
2	$2f^2 - 1$
3	$4f^3 - 3f$
4	$8f^4 - 8f^2 + 1$
5	$16f^5 - 20f^3 + 5f$
6	$32f^6 - 48f^4 + 18f^2 - 1$

Table 1. Zeroth through sixth order Chebyshev polynomials.

$$B_K = \gamma^2 + \sin^2\left(\frac{K\pi}{N}\right) \tag{15}$$

$$g_0 = 1 \tag{16}$$

$$g_1 = \frac{2A_1}{\gamma} \tag{17}$$

$$g_K = \frac{4A_{K-1}A_K}{B_{K-1}\,g_{K-1}} \quad \text{for } K = 2 \text{ to } N \tag{18}$$

$$\left.\begin{aligned} g_{N+1} &= 1 \text{ for N odd} \\ g_{N+1} &= \coth^2\left(\frac{\beta}{4}\right) \text{ for N even} \end{aligned}\right\} \tag{19}$$

Lowpass to Bandpass Transformations

This transformation is a mathematical operation that converts the network from a lowpass filter extending from DC to f_1' to a bandpass filter having a center frequency of f_0 and lower and upper passband limits of f_1 and f_2, respectively. The transformation is pictured in Figure 6.

For filters of about 5 percent bandwidth or less the following relationships may be used:

$$f_0 = \sqrt{f_1 f_2} \tag{20}$$

$$B_f = \frac{f_2 - f_1}{f_0} \tag{21}$$

while for filters up to about 20 percent bandwidth these equations are more appropriate (10):

$$f_0 = f_1 + f_2 - \sqrt{(f_2 - f_1)^2 + f_1 f_2} \tag{22}$$

$$B_f = \frac{f_0}{f_1} - \frac{f_0}{f_2} \tag{23}$$

Calculation of Component Values

Now that the preliminary details have been addressed, we can begin the calculation of actual component values to build the filter. The design technique is that of Matthaei, Jones, and Young in Reference 11.

To begin the calculations, choose source and load resistances and the inductance or capacitance of each resonator. Let's assume that the same value of inductor is to be used in each resonator and its inductance is L. Figure 7 shows the schematic of a generic bandpass filter with each component defined.

Once the resonator inductance has been selected, the resonant capacitance is calculated.

$$C_R = \frac{1}{(2\pi f_0)^2 L} \tag{24}$$

Next the admittances of the end coupling capacitors and the admittance inverters are determined using the previously calculated values of resonator capacitance and normalized lowpass elements.

$$J_{01} = \sqrt{\frac{G_A\,\omega_0\,C_R\,B_f}{g_0\,g_1}} \tag{25}$$

$$J_{J,J+1} = \frac{B_f\,\omega_0\,C_R}{\sqrt{g_J\,g_{J+1}}} \quad \text{for } J = 1 \text{ to } N-1 \tag{26}$$

$$J_{N,N+1} = \sqrt{\frac{G_B\,\omega_0\,C_R\,B_f}{g_N\,g_{N+1}}} \tag{27}$$

After the admittance values have been found, the corresponding capacitor values are calculated.

$$C_{01} = \frac{J_{01}}{\omega_0\sqrt{1 - \left(\frac{J_{01}}{G_A}\right)^2}} \tag{28}$$

$$C_{J,J+1} = \frac{J_{J,J+1}}{\omega_0} \quad \text{for } J = 1 \text{ to } N-1 \tag{29}$$

$$C_{N,N+1} = \frac{J_{N,N+1}}{\omega_0\sqrt{1 - \left(\frac{J_{N,N+1}}{G_B}\right)^2}} \tag{30}$$

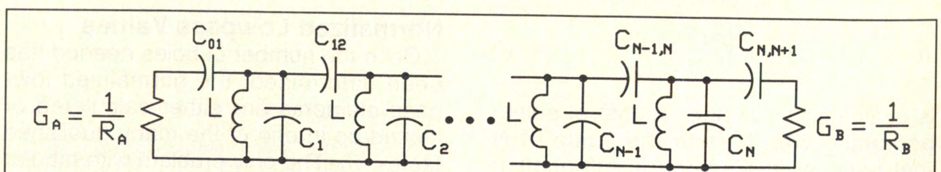


Figure 7. Generic bandpass filter.

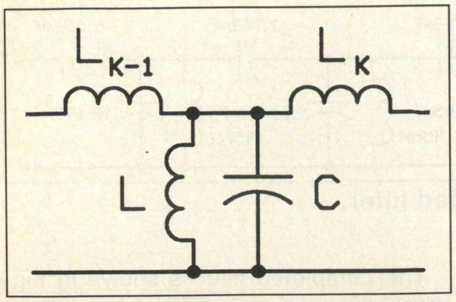


Figure 8. Inductively coupled resonator.

The effective value of the end coupling capacitors is found next. This takes into account that the Q of the end resonators may not be so high that the series and equivalent parallel capacitance are equal.

$$C_{01}^{e} = \frac{C_{01}}{1 + \left(\dfrac{\omega_0 C_{01}}{G_A}\right)^2} \quad (31)$$

$$C_{N,N+1}^{e} = \frac{C_{N,N+1}}{1 + \left(\dfrac{\omega_0 C_{N,N+1}}{G_B}\right)^2} \quad (32)$$

Finally, the node capacitors are calculated as the difference between the resonator capacitance and the two adjacent coupling capacitors.

$$C_1 = C_R - C_{01}^{e} - C_{12} \quad (33)$$

$$C_K = C_R - C_{K-1,K} - C_{K,K+1} \quad (34)$$
for $K = 2$ to $N - 1$

$$C_N = C_R - C_{N-1,N} - C_{N,N+1}^{e} \quad (35)$$

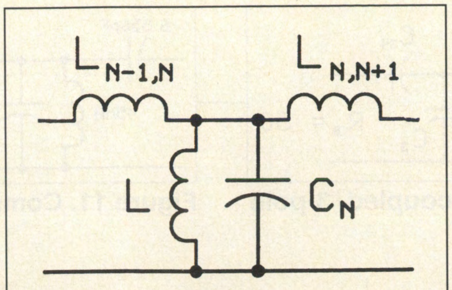
This completes the design of the filter. If desired, the input and output series coupling (basically an L-match) can be replaced with a capacitive tap match and the capacitive coupling between resonators can be replaced with inductive coupling as shown in Figure 8.

Capacitive tap input and output matching can be implemented by first determining the Q of the section and then solving the capacitive tap equations presented earlier in Figure 5 for the same value of Q.

$$Q_{L\text{-MATCH}} = \frac{X_C}{R_S} \quad (36)$$

$$Q_{\text{CAPACITIVE TAP}} = \frac{R_P}{X_L} \quad (37)$$

Inductive coupling between resonators can be implemented by calculating equivalent reactance coupling inductors to replace the coupling capacitors and then recalculating the node capacitors to take into account the effective negative capacitance of the coupling inductor as shown in Figure 9.



$$L_{N-1,N} = \frac{1}{\omega_0^2 C_{N-1,N}}$$

$$L_{N,N+1} = \frac{1}{\omega_0^2 C_{N,N+1}}$$

$$C_N = C_R + 2(C_{N-1,N} + C_{N,N+1})$$

Figure 9. Calculating coupling inductors for Nth resonator.

Filter Design Example

Now that the design procedure has been established, let's design and analyze a filter. Suppose that a filter with the following parameters is needed (see Figure 10):

Center Frequency	145 MHz
Bandwidth	10 MHz
Ripple	0.1 dB
Number of Poles	3
Input/Output Impedance	50 ohms
Resonator Inductance	45 nH
Input/Output Matching	series cap.
Resonator Coupling	capacitive

This would, of course, be a Chebyshev filter. Since the number of poles has been specified, we can proceed directly to the calculation of the normalized lowpass element values. These can be obtained from a table or from the equations presented previously. After calculating or looking up these values we have:

$g_0 = 1.0000$
$g_1 = 1.0316$
$g_2 = 1.1474$
$g_3 = 1.0316$
$g_4 = 1.0000$

The ratio of filter bandwidth to center frequency is greater than five percent so we will use the wideband transformation to find the design center frequency and fractional bandwidth.

$$f_0 = f_1 + f_2 - \sqrt{(f_2 - f_1)^2 + f_1 f_2} \quad (38)$$

$$f_0 = 290 - \sqrt{10^2 + (150)(140)}$$

$$f_0 = 144.742$$

$$B_f = \frac{f_0}{f_1} - \frac{f_0}{f_2} \quad (39)$$

$$B_f = \frac{144.742}{140} - \frac{144.742}{150}$$

$$B_f = 0.06892$$

The resonant capacitance is calculated next as

$$C_R = \frac{1}{\omega_0^2 L} \quad (40)$$

$$C_R = \frac{1}{\left(2\pi \times 144.742 \times 10^6\right)^2 \left(45 \times 10^{-9}\right)}$$

$$C_R = 26.8683 \text{ pF}$$

and the input coupling capacitance is then found:

$$J_{01} = \sqrt{\frac{G_A\, \omega_0\, C_R\, B_f}{g_0\, g_1}} \quad (41)$$

$$= \left(\frac{(.02)(2\pi \times 144.742 \times 10^6)(26.8683 \times 10^{-12})(.06892)}{(1.0)(1.0316)}\right)^{\frac{1}{2}}$$

$$J_{01} = 5.714 \times 10^{-3}$$

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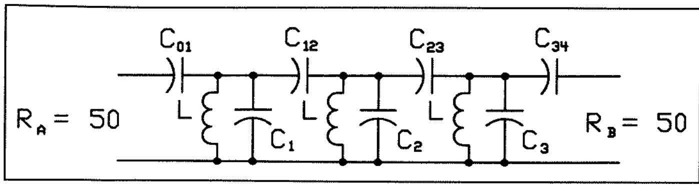

Figure 10. Uncalculated, capacitively coupled, 3-pole bandpass filter.

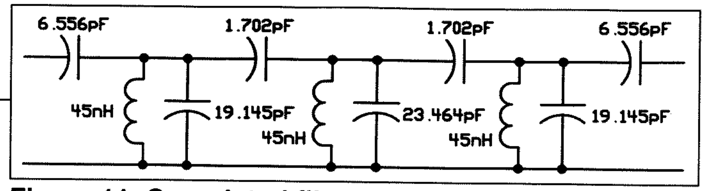

Figure 11. Completed filter.

$$C_{01} = \frac{J_{01}}{\omega_0 \sqrt{1-\left(\frac{J_{01}}{G_A}\right)^2}} \tag{42}$$

$$= \frac{5.714\times10^{-3}}{(2\pi\times144.742\times10^6)\sqrt{1-\left(\frac{5.714\times10^{-3}}{.02}\right)^2}}$$

$$C_{01} = 6.556 \text{ pF}$$

$$C_{01}^e = \frac{C_{01}}{1+\left(\frac{\omega_0 C_{01}}{G_A}\right)^2} = 6.021 \text{ pF} \tag{43}$$

The coupling capacitors between the resonators are calculated next:

$$J_{12} = \frac{B_f \omega_0 C_R}{\sqrt{g_1 g_2}} \tag{44}$$

$$J_{12} = \left((.06892)(2\pi\times144.742\times10^6)\frac{(26.8683\times10^{-12})}{\sqrt{(1.0316)(1.1474)}}\right)$$

$$J_{12} = 1.5479\times10^{-3}$$

$$J_{23} = \frac{B_f \omega_0 C_R}{\sqrt{g_2 g_3}} = 1.5479\times10^{-3} \tag{45}$$

$$C_{12} = \frac{J_{12}}{\omega_0} \tag{46}$$

$$C_{12} = \frac{1.5479\times10^{-3}}{2\pi(144.742\times10^6)} = 1.702 \text{ pF}$$

$$C_{23} = \frac{J_{23}}{\omega_0} = 1.702 \text{ pF} \tag{47}$$

Then the output coupling capacitor is determined:

$$J_{34} = \sqrt{\frac{G_B \omega_0 C_R}{g_3 g_4}} = 5.714\times10^{-3} \tag{48}$$

$$C_{34} = \frac{J_{34}}{\omega_0\sqrt{1-\left(\frac{J_{34}}{G_B}\right)^2}} = 6.556 \text{ pF} \tag{49}$$

$$C_{34}^e = \frac{C_{34}}{1+\left(\frac{\omega_0 C_{34}}{G_B}\right)^2} = 6.021 \text{ pF} \tag{50}$$

Finally, the node capacitors are found by taking the difference between the resonator capacitance and the coupling capacitors.

$$C_1 = C_R - C_{01}^e - C_{12} = 19.145 \text{ pF} \tag{51}$$

$$C_2 = C_R - C_{12} - C_{23} = 23.464 \text{ pF} \tag{52}$$

$$C_3 = C_R - C_{34}^e - C_{23} = 19.145 \text{ pF} \tag{53}$$

The completed filter is shown in Figure 11.

Analysis of the Design

To verify the design, the filter circuit was analyzed using C/NL, an RF analysis program available from Artech House. Figure 12 shows a plot of the filter amplitude and return loss response while Figure 13 shows a close-up of the passband response. Figure 13 verifies that the ripple is 0.1 dB and that the filter bandwidth is 10 MHz as specified.

Summary

This article has presented the basic theory of the coupled resonator bandpass filter implemented using lumped components. A flexible design procedure has been developed which will allow the designer to follow a step-by-step process in designing bandpass filters.

This program is available on disk through the RF Design Software Service and includes step-by-step operating instructions. See page 100 for ordering information. ***RF***

References

1. Bowick, Chris, *RF Circuit Design*, Howard W. Sams and Company, 1982, pp. 59-64.
2. Matthaei, G., Young, L., and Jones, E., *Microwave Filters, Impedance-Matching Networks, and Coupling Structures*, Artech House, 1980, p. 144.
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5. Fink, D., and Christiansen, D., ed., *Electronic Engineer's Handbook*, 3rd edition, McGraw-Hill, 1989, pp. 12-5—12-12.
6. Bowick, pp. 47, 49.
7. Bowick, p. 51.
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9. Matthaei, pp. 98-102.
10. Matthaei, pp. 481, 486.
11. Matthaei, pp. 482-485.

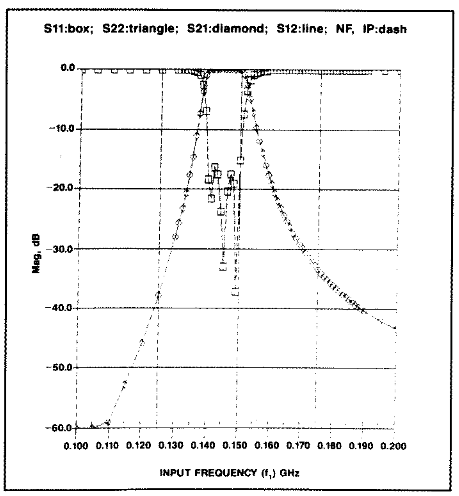

Figure 12. Calculated wideband response of example filter.

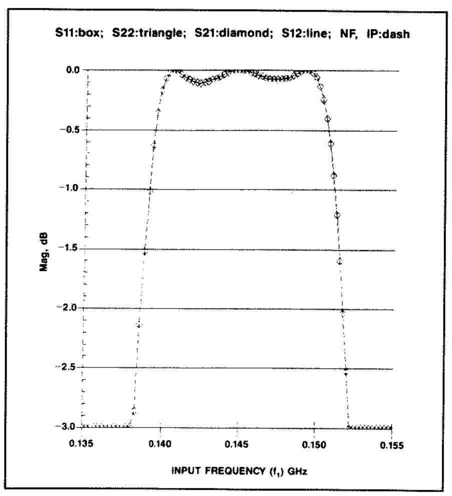

Figure 13. Calculated passband response of example filter.

About the Author

John Freed is a Staff Engineer for Teletec Corporation in Raleigh, NC. His main areas of expertise are frequency synthesizers and receivers. He received a BSEE from Akron University in 1978 and has worked in cellular and land mobile design. He can be reached at PO Box 20405, Raleigh, N.C. 27690-1975, or by phone at (919) 556-7800.

Spectrum Analyzer on a Card

The 9052 is a full-featured, 1600 MHz RF spectrum analyzer that plugs into an IBM AT compatible computer. The 9052 has an amplitude range of –120 dBm to +30 dBm, with a third-order dynamic range of greater than 80 dB. Five resolution bandwidth filters range in bandwidth from 300 Hz to 3 MHz. Other features include marker operations, user defined test sequences, control from a remote PC and various trace math operations.

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Surface Mount VCOs

Magnum Microwave offers two new surface mount voltage controlled oscillators. A hyperabrupt varactor tuned VCO covers 25 to 2000 MHz, and an abrupt varactor tuned VCO covers 400 to 1600 MHz.

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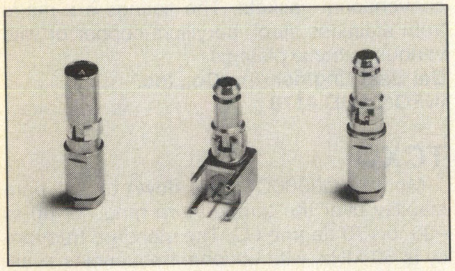


As long as the contacts are engaged, a 50 ohm line is achieved. The design of the PkZ allows it to be scaled up or down in diameter and will allow it to operate from DC to 32 GHz.

The Phoenix Company of Chicago
INFO/CARD #198

90 Degree Hybrids

A new line of 90 degree hybrids with bandwidth ratios of 5:1 is housed in a relay header measuring 0.4 × 0.8 × 0.4 inches. These devices offer low VSWR (less than 1.5:1), low insertion loss (typ. 0.8 dB), and low phase and amplitude imbalances (typ. 3.0 and 0.7 dB, respectively), while offering typical isolation of 23 dB.

Synergy Microwave Corp.
INFO/CARD #197

Broadband Power Amplifiers

Two new solid-state broadband power amplifiers are being introduced at RF Expo East by Amplifier Research. Both cover the frequency range of 400-1000 MHz. Models 500HB at 500 watts and 1000HB at 1000 watts minimum output offer full control functions (leveling, pulse provision, remote control by computer, self diagnostics) and front-panel monitor of forward and reflected power. Both are totally immune to load mismatch, even infinite VSWR. Minimum gain is 60 dB for the 1000MB and 57 dB for the 500HB.

Amplifier Research
INFO/CARD #196

30 Watt Class-A Amplifier

Model 630L is a laboratory grade linear amplifier with 30 watts of class-A power from 400 to 1000 MHz. The amplifier, which is equipped with GPIB and RS-232/422 interfaces, offers nominal gain of 51 dB. An automatic-level-control (ALC) circuit maintains output power over a 30 dB range with ± 0.3 dB flatness.

ENI
INFO/CARD #195

Semicustom Array

The BTA 24 semicustom array is designed to make design of high performance silicon MMICs easy with a PC based computer and the easy to follow instruction manual. Bipolarics' SiMMIC™ process features 10 GHz f_T low noise transistors, thin film resistors, capacitors and multi-layer gold metallization. Circuits with 2 to 96 transistors can be developed quickly and with modest cost.

Bipolarics, Inc.
INFO/CARD #194

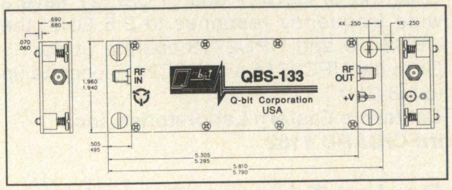
Clock Recovery Module

The AT&T TRU600 is a tiny, surface mount UHF clock recovery and data retiming module designed for high performance datacom and telecom applications, including those at the SONET OC-3 and OC-12 (155 and 622 MB/s) data rates. This product features low power dissipation, fast acquisition time, and low output jitter and low output noise.

AT&T Microelectronics
INFO/CARD #193

Cellular Amplifier

Q-bit Corporation's high dynamic range cellular amplifier is specifically designed for cellular receive stations. The amplifier features a noise figure of less than 0.8 dB at 25 degrees C and greater than +38 dBm third order intercept point. The amplifier operates from 15 volts, with typically less than 220 mA current drawn.



Q-bit Corporation
INFO/CARD #192

Feed Forward Amplifier

The RF-2601A from Locus, Inc. is an HF feed forward amplifier covering 1 to 80 Mhz. The RF-2601A has 16 dB gain with a third order intercept point of +65 dBm and a second order intercept point of +120 dBm. The amplifier has a noise figure of 4.5 dB.

Locus, Inc.
INFO/CARD #191

Programmable Attenuators

Alan Industries has introduced a revolutionary line of GaAs FET programmable attenuators. Utilizing precise gain and equalization circuitry, the GFA series eliminates the high insertion loss typical of GaAs FET devices. Attenuation levels are controlled by direct TTL input with a typical accuracy of ± 0.5 dB and insertion loss of 0.5 dB. Available with attenuation ranges from 63 dB, these 50 ohm devices have an operating frequency of 60-860 MHz.

Alan Industries, Inc.
INFO/CARD #190

Oven Controlled XO

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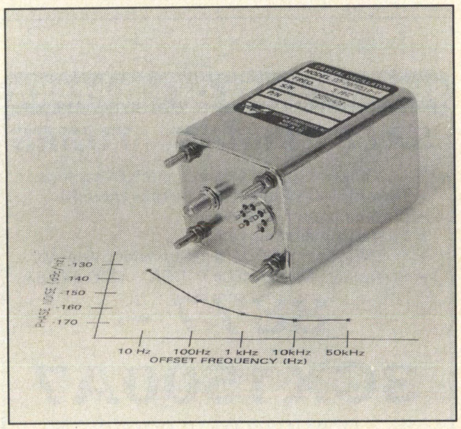

2×10^{-8}/year in a 2 × 2 × 2.75 inch package. Temperature stability is $\pm 5 \times 10^{-10}$ over 0 to +50 degrees C and $\pm 5 \times 10^{-9}$ over –55 to +75 degrees C. Frequencies are available to 75 MHz. Price is $850 at 100 pcs., with delivery in 14 to 16 weeks.

Vectron Laboratories, Inc.
INFO/CARD #189

DS3 Signal Tester

The PF-47 is used for comprehensive in-service or out-of-service testing of DS3 signals. It is a handheld instrument with both transmit and receive capabilities for complete error insertion testing. Its self contained battery pack allows more than eight hours of operation. The auto-configure feature makes it easy to use.

Wandel & Goltermann, Inc.
INFO/CARD #188

SONET SAWs

Sawtek will introduce a series of SAW filters specifically designed for SONET (synchronous optical network) clock recovery applications. Because of their excellent temperature stability and amplitude symmetry, SAW filters are a valuable component in the SONET equipment of many manufacturers. Sawtek offers standard filters for both the OC-3 (155.52 MHz) and OC-12 (622.080 MHz) optical carrier levels with OC-48 (2.488 GHz) devices soon to be available. The OC-3 filters are available with Qs between 260 and 1000, while the Q of the OC-12 device is 600. These devices are priced competitively and are available in both leaded and surface mount packages.
Sawtek, Inc.
INFO/CARD #187

Prescaler IC

Signetics' new low-power prescaler IC has a minimum supply voltage of just 2.7 volts and can reduce battery requirements in portable equipment from five to three NiCad cells. The ICs, available in dual and triple modulus versions, have maximum input signal frequency of 1.2 GHz and modulus set-up time of five nanoseconds. The NE/SA701 chip is a dual modulus, divide by 128/129 or 64/65 circuit, the NE/SA702 chip is a triple modulus, divide by 64/65/72 circuit and the NE/SA703 chip is a triple modulus, divide by 128/129/144 circuit. The ICs are eight-pin, 150-mil-wide surface mount, plastic packages.
Philips Semiconductors-Signetics
INFO/CARD #186

Dual Modulus Divider

The SP8401 is a very low phase noise, 300 MHz, modulus 10/11 divider. Special circuit techniques have been used to reduce the phase noise considerably below that produced by standard dividers. The modulus control input is CMOS or TTL compatible. Phase noise is typically −160 dBc/Hz at 1 kHz offset. The device operates from 5 volts, with typical power consumption of 165 mW.
GEC Plessey Semiconductors
INFO/CARD #185

Switching Subassemblies

JFW is now offering programmable switching systems which consist of numerous switches and power dividers in one rack mountable enclosure. Each unit is also available with operating software. PC control via RS-232 or IEEE-488 are control options. Both DC and AC powered versions are available.
JFW Industries, Inc.
INFO/CARD #184

Signal Processing Components

Hewlett Packard announces several new products: ATF-21185 and ATF-13786 GaAs FETs, MSA-3111 and MSA-2011 silicon monolithic preamplifiers, IAM-82008 silicon bipolar MMIC active mixer, IFD- 53010/53110 1/4 prescalers, IVA-14208/-14228 variable gain amplifiers, AT-41511/86 and AT-60111/211 silicon bipolar transistors.
Hewlett Packard Company
INFO/CARD #183

MMIC Amplifier Series

The UPC2700 series of low cost wideband silicon MMIC amplifiers use NEC's new NESAT III process. This 20 GHz f_t process is optimized to produce devices which have higher operating frequencies, higher efficiency and lower operating voltages. The UPC2708T and UPC2711T operate up to 3 GHz; the UPC2709T and UPC2712T have a wide frequency response to 2.5 GHz; the UPC2710 and UPC2713 operate up to 1.5 GHz; the UPC2714 and UPC2715 operate up to 1.8 GHz.
California Eastern Laboratories, Inc.
INFO/CARD #182

Sealed Trimmer Capacitors

Voltronics is now producing a high reliability line of sealed precision trimmer capacitors using a solid dielectric of PTFE. The SD series comes in four ranges and ten styles to offer the design engineer 40 high reliability parts to choose from. Ranges of 1 to 4 pF through 1 to 23 pF at 1000 piece pricing of $5.00 to $8.00 are available.
Voltronics Corporation
INFO/CARD #181

Circuit Layout Editors

Serenade™ PC and WS layout editors are designed to provide automatic and semi-automatic layout of RF and microwave circuits and subsystems. The Serenade PC and WS layout software interfaces closely to the Serenade PC and WS schematic software which provides extensive schematic entry for both frequency and time-domain simulators.
Compact Software, Inc.
INFO/CARD #180

Wide Dynamic Range Amplifiers

Trontech's new patent pending wide dynamic amplifiers offer the systems designer the widest input dynamic range/low noise amplifiers available today.
Trontech, Inc.
INFO/CARD #179

EMI Spring Gaskets

EMI spring gaskets from Bal Seal offer a patented canted coil design which offers excellent shielding in high and low frequencies and features distinct advantages over conventional gaskets. The springs are made from stainless steel, beryllium copper or with various optional platings.
Bal Seal Engineering Co., Inc.
INFO/CARD #178

TCXO

Model XO3006C offers down to ± 1 ppm stability over the temperature range of up to −30 to +70 degrees C. The standard frequency is 50 MHz, with optional frequencies available between 35.0 and 60.0 MHz. The unit operates from +12 volts and less than 3 mA and features no sub-harmonic spurs. Standard output is 0 dB into 50 ohms. Package dimensions are 1.5 × 1.5 × 0.50 inches.
Piezo Technology, Inc.
INFO/CARD #177

Surface Mount Coaxial Connectors

OSMT surface mount coax connectors are designed to use less PCB space than conventional connectors. The OSMT is ideal for use in telecommunications, GPS, consumer electronics and automotive systems.
M/A-COM Omni Spectra
INFO/CARD #176

Flange Mount Termination

Model 32-1037 is a high performance, 250 W termination which operates over the frequency range of DC to 3 GHz. VSWR is less than 1.1 to 2 GHz and 1.35 to 3 GHz. The unit offers internal impedance matching which eliminates shunt capacitance.
Florida RF Labs, Inc.
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Balo Hermetics Co.
INFO/CARD #174

Spectrum Analyzer

The SNA-7 spectrum analyzer is suitable for a wide range of measurements in the frequency range of 50 Hz to 22 GHz. Low intrinsic distortion figures make it useful in harmonic, intermodulation product, spurious signal and phase noise measurements. The SNA-7 features a high tuning resolution of 0.1 Hz coupled to a 3 Hz resolution bandwidth, with unrestricted use over the entire range.
Wandel & Goltermann, Inc.
INFO/CARD #173

Circuit Board Component Shielding

Style CBS shielding consists of a fence and removable cover that, when assembled and solder mounted, provides effective board-level shielding against EMI. Standard sizes range from 0.5 × 0.5 inches to 6 × 6 inches, including squares, rectangles and odd shapes formed at 0.5 inch increments. Snap-in dimple design provides a positive contact and downward stop, while allowing hand removal of the cover for component accessibility.
Leader Tech, Inc.
INFO/CARD #172

Attenuators

The Super RF series provides attenuation at or above 80 dB at 20 GHz. The series is available off the shelf.
Compac Development Corp.
INFO/CARD #171

CAE Software

Object Engineering is an object oriented software package for the production of custom CAE in Microsoft Windows 3.1. CAE projects can be created visually from the internal objects without programming, and the framework can be extended in C++. Full source code targeted to engineers accelerates the learning of C++ and Windows programming.
Innovation First
INFO/CARD #170

RF Devices

Motorola will show a number of items: MRF10031 long pulse microwave power transistor, MRF899 900 MHz linear power transistor for digital base stations, MHW903 series 900 MHz power modules for GSM digital cellular system, MHW9001 series UHF GaAs power amplifiers for IRIDIUM™ and the MRF947/MRF957 low noise small signal amplifier transistors.
Motorola Semiconductor Products
INFO/CARD #169

Power Amplifiers

Power Systems Technology will show their high power solid state RF and microwave amplifiers from 2 watts to greater than 1000 watts. Frequency ranges from 1.5 MHz to 2 GHz.
Power Systems Technology
INFO/CARD #168

Power Dividers Cover Up to 40 GHz

The range of Merrimac's wideband four way power dividers has been extended to 40 GHz with the introduction of a new series. The PDK-45R series uses a machined housing with K-connectors and includes an ultra-wide 6-40 GHz model. Minimum isolation across this widest band is 14 dB and typically is even better. This series is featured in Merrimac's new 360-page, M-92 catalog.
Merrimac Industries, Inc.
INFO/CARD #167

Discrete Components

Microelectronics Ltd. will show their lines of high and medium power non-magnetic RF capacitors along with their chip inductors and tuning devices.
Microelectronics, Ltd.
INFO/CARD #166

Quartz Materials

P.R. Hoffman Materials Processing Corp. will display their lines of quartz, quartz products, semi-finished blanks and SAW substrates.
P.R. Hoffman Materials Processing Corp.
INFO/CARD #165

Test and Measurement Equipment

The John Fluke Manufacturing Co. will display members of their full line of high purity and general purpose RF signal generators, frequency counters and digital storage oscilloscopes covering the frequency spectrum up to 2 GHz.
John Fluke Mfg. Co., Inc.
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Wideband FM IF IC

The MC13156 has an onboard Colpitts VCO for PLL controlled multichannel operation. The mixer is useful to beyond 200 MHz and may be used in a differential, balanced or single ended configuration. Its precision data shaper has a hold function.
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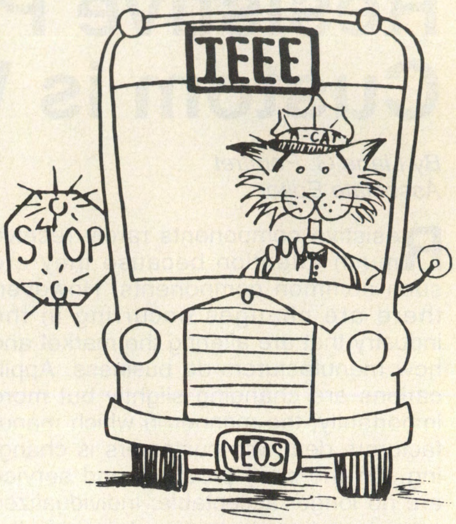

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INFO/CARD 97

RF product report

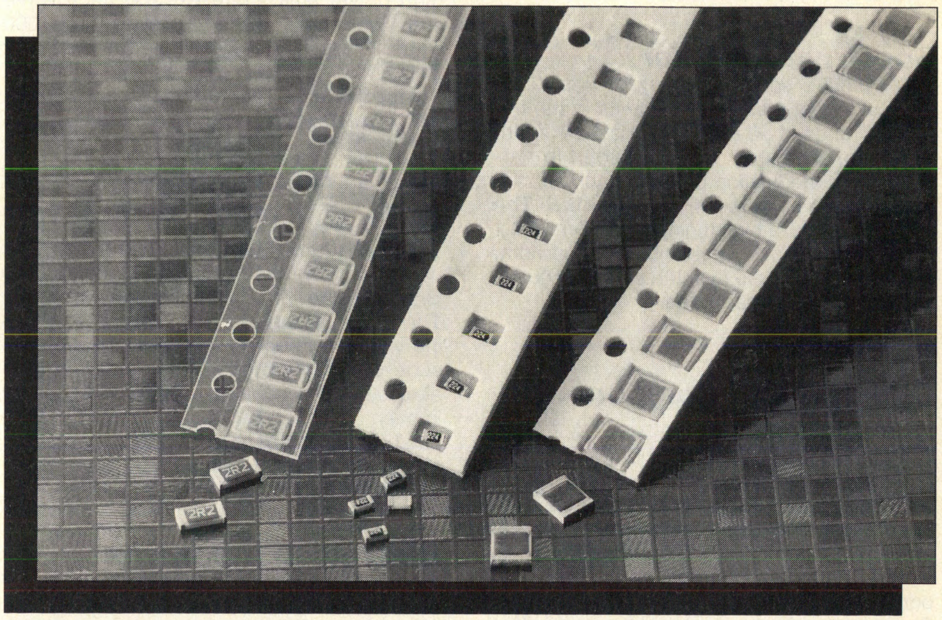
Resistive Products — Custom is Where It's At

By Liane G. Pomfret
Associate Editor

Resistive components rarely receive much attention because they are such common components. However, there are changes occurring in the industry that are altering the market and how manufacturers do business. Applications are changing slightly but more importantly, the manner in which manufacturers deal with customers is changing. Off-the-shelf products and service are no longer acceptable. Individualized service and products now determine the success of manufacturers.

The focus these days for resistive products — resistors, attenuators, terminations and loads — is in the high power/high frequency arena. While typical applications — instrumentation, test bench, standard combiners — are still in demand and products are selling well in those areas, companies appear to be focusing on high power applications. Tim Holt, engineering director at Bird Electronics comments "The main thrust these days is for more reliable resistive devices at higher power levels." These applications include TV and radio broadcast equipment, radar, medical equipment, pulse equipment for particle accelerators, local area networks, mobile radio base stations, and imaging equipment. There's nothing new or startling about these technologies, but they are showing strong growth.

As with many RF components these days, the trend is towards smaller devices, with better performance and lower cost. For resistive products, through-hole devices are being redesigned into surface mount packages. Resistors are now available in rectangular shaped packages and attenuators are becoming more common in TO-8 packaging. Resistive products in MMIC or semiconductors are also appearing on the market. "These new configurations are making products more useable," says Gary Innocenti, component marketing manager for Carborundum Company. Revising the packaging styles can help to lower manufacturing costs because automation can be used instead of manual labor for assembly. In addition, a surface mount or rectangular package can help reduce the required amount of board space, which will also help to reduce cost.



The Customer Comes First

Perhaps the biggest surprise has been the fact that the amount of custom work being requested has increased substantially over the past few years. While no hard statistics were offered, many of the companies contacted for this report commented on the amount of custom work being done. Dave Distler, director of sales and marketing for Trilithic, speculated on possible reasons, "Historically, the commercial market has usually worked with standard parts. Now with more military oriented manufacturers entering the commercial arena they are asking for custom products, because they may not be used to designing in standard parts." Unique designs and in some instances, applications are also driving the demand for custom work. Stock parts are still selling well, but as Geoffrey Smith, director of sales and marketing for Lucas Weinschel notes, "Today's market offers a mature cross section of products, but with increased competition from other manufacturers of standard parts, it makes sense to offer custom services."

These days, customers want to be treated as individuals and not as invoice numbers. Joetta Walker, president of JFW remarks that, "Customers want engineering support both up front and after the sale." For the manufacturer who recognizes and provides this kind of service, business is going well. It is no longer enough to just provide a good product, a manufacturer must also provide strong engineering support in order to be successful.

Resistive products are following the same trends as the rest of the RF industry. As Dave Distler points out, "The increase in the amount of commercial business is making up for the drop in military business, so things are remaining fairly flat." In general, the market for resistive products remains relatively steady. Provided manufacturers are careful in their long-term planning and marketing, the diversity of resistive products and applications will continue to do well. *RF*

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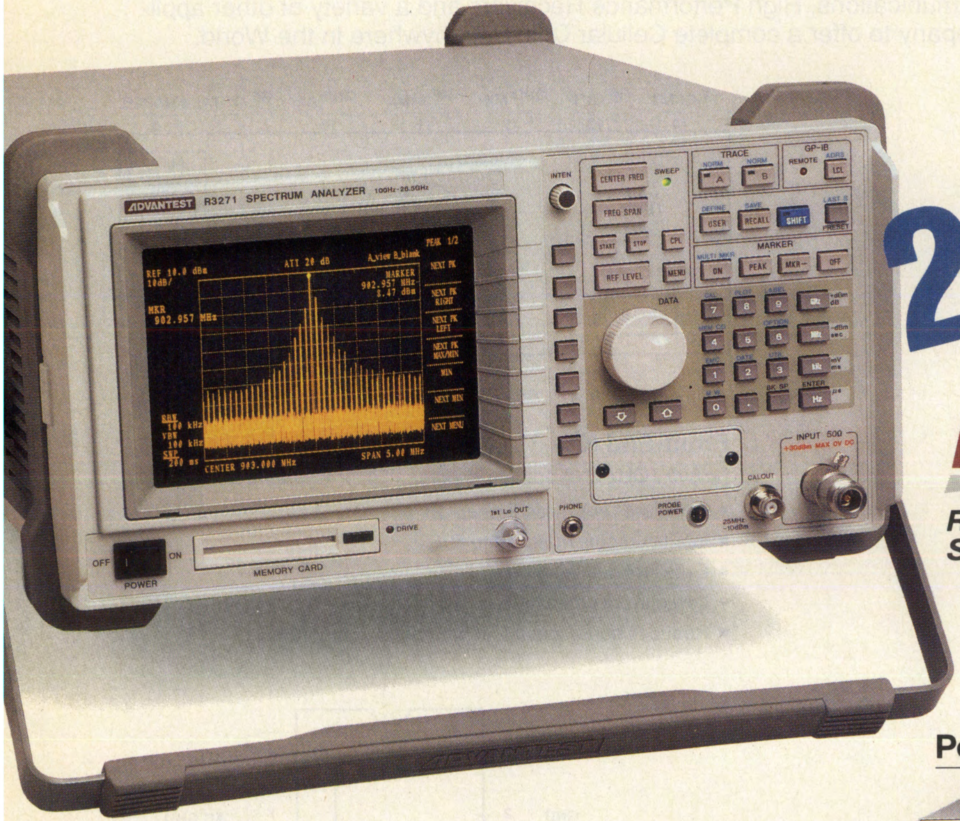

26.5GHz

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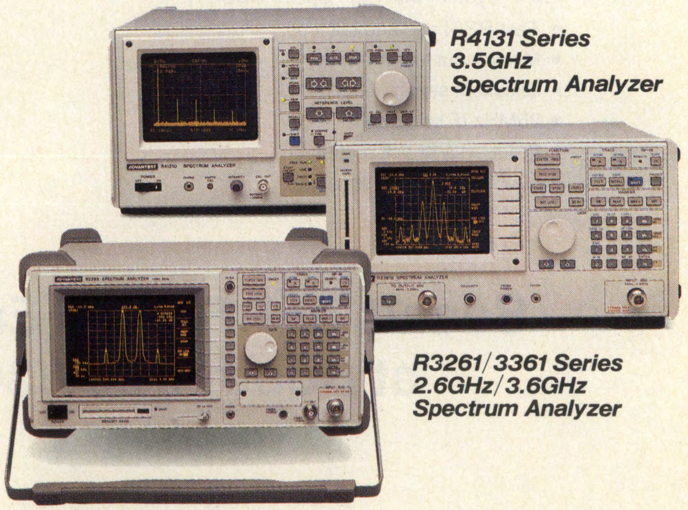

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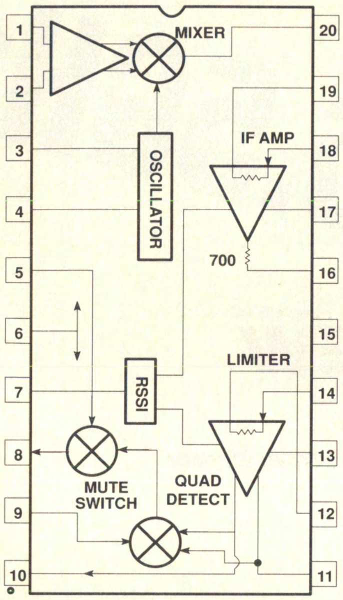
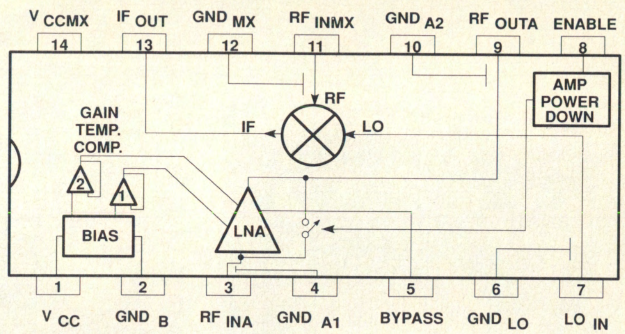


RF Receivers

NE627 High Performance Low-Power Mixer FM IF System with High Speed RSSI

- Fast RSSI rise and fall time
- Mixer input to >500MHz
- Excellent sensitivity
- 25MHz IF bandwidth
- Frequency check output
- 90dB RSSI dynamic range



Front-End ICs

NE600 1GHz LNA and Mixer

- Excellent LNA noise figure 2dB
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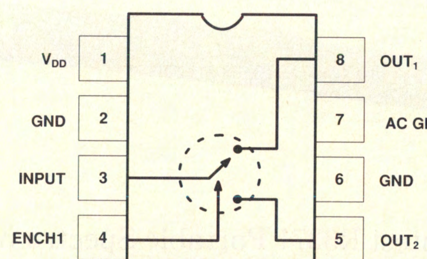
- Excellent gain stability
- Low current consumption
- Reduces boardspace and peripheral components



RF Switch

NE630 Single Pole Double Throw Switch

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1993 RF DESIGN AWARDS CONTEST

CALL FOR ENTRIES

1993 RF Design Awards Contest
Official Rules

The 1993 RF Design Awards Contest provides recognition for innovation and engineering excellence among RF designers. Again this year, there are two separate entry categories. Please note that the rules may not be the same as previous years' contests.

I. The DESIGN Contest

RULES

1) Entries shall be circuits with an RF function, operating in the frequency range below 3 GHz.

2) Circuits entered shall have a complexity equivalent to that of a circuit using 8-10 discrete active devices, or 6-8 integrated circuits. This rule is for ease of judging, to have all entries be of a similar scope. The entry can be a portion of a larger system.

3) Entries may represent design or test methods. Design method entries should include an example circuit. Test method entries should include a description of the device or system under test.

4) The entries shall be the original work of the entrant, not previously published. If developed as part of the entrant's employment, entries must have the employer's approval for submission.

5) Only one entry per person is permitted. An entry may have two or more co-authors.

6) Submission of an entry implies permission for publication by *RF Design*. All prize-winning entries will be published, plus additional entries of merit.

7) Winners are responsible for any taxes, duties, or other assessments which result from the receipt of their prizes.

8) Entries must be postmarked by March 19, 1993 and received no later than March 26, 1993.

9) All entries will remain confidential until the publication of the July 1993 issue of *RF Design*.

JUDGING CRITERIA

Originality — Each design will be evaluated for similarity to work by others, and other judgements of its unique contributions.

Engineering — Entries should clearly identify how the entry was created in response to a need.

Documentation — A complete description of the circuit or technique is required, including sufficient theoretical background, description of circuit operation, and performance data.

II. The PC SOFTWARE Contest

RULES

1) Each entry shall be a computer program which assists in the design, test, or control of RF circuits or systems.

2) Programs must operate on computers compatible with MS-DOS/PC-DOS or Apple Macintosh operating systems. Any special hardware requirements should be noted (memory, graphics, etc.).

3) Programs should be provided in a form that can be run without special support software; programs should be provided in compiled, directly executable form. Programs using the BASICA or GWBASIC interpreters are acceptable. Programs that require spreadsheet or mathematics software packages cannot be accepted.

4) Programs entered must be submitted on disk. Supporting documentation on theory of operation, references, and operating instructions must be supplied in printed form. Source code must be supplied, either on disk or in printed form.

5) The entries shall be the original work of the entrant, not previously published or distributed (including public BBS or shareware). If developed as part of the entrant's employment, entries must have the employer's approval for submission.

6) Only one entry per person is permitted. An entry may have two or more co-authors.

7) Submission of an entry implies permission for publication by RF Design and distribution by the RF Design Software Service. All prize-winning entries will be published, plus additional entries of merit. Some restrictions on publication and distribution of source code may be acceptable.

8) Winners are responsible for any taxes, duties, or other assessments which result from the receipt of their prizes.

9) Entries must be postmarked by March 19, 1993 and received no later than March 26, 1993.

10) All entries will remain confidential until the publication of the July 1993 issue of *RF Design*.

JUDGING CRITERIA

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

Usefulness — The value of the software to the RF engineering community will be evaluated. Both the nature of the computations performed and ease of operation of the program will be considered.

DEADLINE FOR ENTRIES:
POSTMARKED BY MARCH 19, 1993 — RECEIVED BY MARCH 26, 1993

Send entries to:

RF Design Awards Contest
***RF Design* magazine**
6300 S. Syracuse Way, Suite 650
Englewood, CO 80111

BASIC Numeric Compiler

TransERA announces the release of the latest addition to their High Tech Basic product line: the HTBasic DOS 386/486 Numeric Compiler. The numeric compiler produces code that runs in the fast 386/486 processor 32-bit protected mode. It also produces in-line math code to fully utilize the 387/486 math processor. The TransEra HTBasic DOS 386/486 Numeric Compiler, including the complete development and runtime versions, is now available for $1325 in the U.S. or as an upgrade for $450 in the U.S.

TransEra Corporation
INFO/CARD #138

Electronic Packaging Simulation

Arizona Packaging Software announces the release of the AZtec™ System for modeling and simulation of advanced electronic packaging designs. Featured modules in the initial offering of AZtec are package parasitics modeling and rapid digital pulse simulation for multiple coupled transmission lines. Pricing for the AZtec System starts at $35,000. Individual modules are available from $10,000.

Arizona Packaging Software, Inc.
INFO/CARD #137

Amplifier Simulation Program

SW.I.F.T. Enterprises announces the latest release of the design tool ASP (Amplifier Simulation Program) Version 3.11 for interactive development of weak signal, solid state amplifiers. Various auto or manual routines allow the developer to optimize for noise, gain, output VSWR while providing input/output matching circuits. Design by Stern's stability factor, gain and matched conditions are available to the engineer. ASP version 3.11 is IBM compatible and sells for $85 plus $3 shipping.

SW.I.F.T. Enterprises
INFO/CARD #136

Propagation Software

Applied Spectrum Research announces the addition of several new options to their Radio System Design Software family. Coverage calculations accounting for the effects of vegetation and urban area can be made with the Land Cover and Morphology routine. An SMR planning routine includes two site plotting with specified contour levels plus the standard 70 mile spacing circles. The Cellular Carey Contours routine implements new procedures required by FCC Docket 90-6. Also included is a high resolution radial coverage (or "sunburst") routine.

Applied Spectrum Research
INFO/CARD #133

RF Designer Software

The Engineers' Club offers 35 public domain and shareware programs from their library of microwave and RF routines. Included are circuit design solvers, NOVA and Pspice circuit simulation, math and plotting routines, and more. These files are available for MS-DOS based computers on 6 high density 5-1/4 inch floppy disks. Cost is $49.00 plus $10.00 shipping and handling.

The Engineers' Club
INFO/CARD #135

Electromagnetic CAD

MAFIA (MAxwell equation solved by Finite Iterations Algorithm) by AET Associates can provide simulations of almost any electromagnetic device. MAFIA's fully three dimensional finite difference code is based on the most self-consistent formulation of Maxwell's equations, without limiting assumptions commonly found in other programs. Different MAFIA modules solve different classes of problems. MAFIA 3.1 is available for SUN, HP, IBM and VAX workstations as well as Cray supercomputers.

AET Associates, Inc.
INFO/CARD #134

Network Software

Teklogix has announced the availability of TCP/IP (Transmission Control Protocol/Internet Protocol) network software for applications with the company's radio frequency data communications (RF/DC) systems. The software allows Teklogix RF/DC terminals to communicate with host computers using TCP/IP over an RS-232 interface or Ethernet LAN.

Teklogix, Inc.
INFO/CARD #132

Easy TESLA Block Diagram Simulation Runs on Your PC

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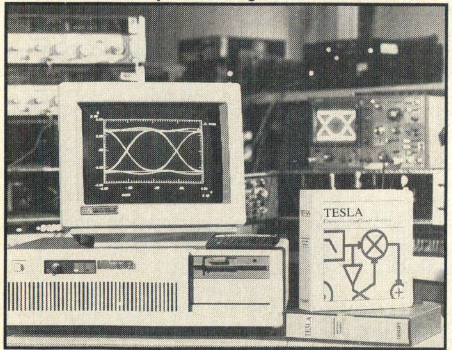
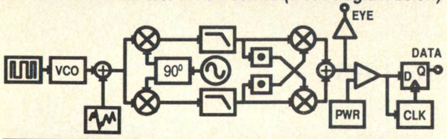
- Use OrCAD® to input block diagrams—runs *under* TESLA



Simulation and lab test of FSK demod (block diagram below)



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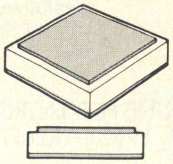
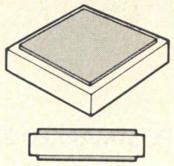


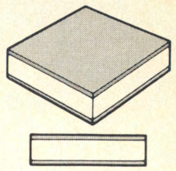
- Safety Margin around top
- Top: TiW/Au
  Bottom: TiW/Pt/Au

Type B
250V



- Safety Margin around top & bottom
- Top: TiW/Au
  Bottom: TiW/Au

Type C
100V



- Solderable
- Top: TiW/Pt/Au
  Bottom: TiW/Pt/Au

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INFO/CARD 83

RF literature

Analog Design Software Catalog
EEsof has released a product guide describing the capabilities and features of their high frequency analog design software. Included is information about the OmniSys system designer, Touchstone, Libra and Microwave SPICE microwave design tools and J-Omega radio frequency circuit designer.
EEsof
INFO/CARD #159

High-Power Coupler Brochure
A six-page brochure from Werlatone describes a series of high power dual directional couplers covering the frequency range of 0.01-1000 MHz. Covered are specifications and performance curves for these broadband devices including coupling, insertion loss, directivity, VSWR and power ratings from 100 watts to 100kW.
Werlatone, Inc.
INFO/CARD #158

Low Noise Components
Techtrol has introduced a technical brochure outlining its Ultra Low Noise products and capabilities. Contained are product specifications, technical descriptions of the products and application notes. Featured in the brochure is the model LNS712A microwave source, designed for calibration of noise measurement equipment and systems.
Techtrol Cyclonetics, Inc.
INFO/CARD #157

CO_2 Laser Exciter Tubes
Seimens' 24-page brochure offers assistance and ideas in the selection of circuitry concepts, tubes and RF cavity resonators for exciting CO_2 lasers. The brochure is divided into four segments: techniques, applications and trends in CO_2 lasers, possible implementation of RF sources for stripline/waveguide lasers and dimensions of high-u triodes and tetrodes.
Siemens Corporation
INFO/CARD #156

Power Transistors and Modules
M/A-COM Power Hybrids Operation announces a catalog describing power transistors and modules for both commercial and military applications. The 162-page catalog offers detailed specifications on RF power MOSFETs, bipolar transistors and power modules as well as design tables, application notes and screening and testing capabilities.
M/A-COM, Inc.
INFO/CARD #155

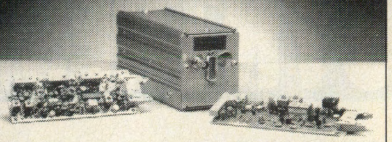
Coil and Transformer Catalog
Renco Electronics' 109-page catalog contains electrical and mechanical specifications for their lines of air core inductors, ferrite core inductors, shielded inductors, high Q flat coils and surface mount components, among others. Also included is information regarding Renco's custom designs.
Renco Electronics, Inc.
INFO/CARD #154

FCC Rules on Disk
Pike & Fischer announces software containing the full text of FCC rules Part 1, Part 2, Part 5, Part 15, Part 18 and Part 68. Also included is the text of all relevant FCC Notices of Proposed Rulemaking and Notices of Inquiry. The software will perform word and phrase searches and can import text to any word processor accepting ASCII files. FCC Equipment Rules on Disk is being sold on a subscription basis with monthly updates for $325 per one year.
Pike & Fischer, Inc.
INFO/CARD #153

GaAs MMIC Products Guide
Alpha Industries announces its 1992 GaAs MMIC products catalog. This 200-page catalog contains extensive product information on a wide range of control products covering a frequency range from RF through millimeter wave. Featured are product selection guides for radar, EW and military telecommunications, along with a complete listing of low cost, high volume components for commercial applications.
Alpha Industries, Inc.
INFO/CARD #152

General Catalog
Wiltron's 1992 catalog describes over 300 microwave components, instruments and systems in the DC to 110 GHz range. Featured are descriptions of Wiltron's new Vector Network Analyzer and new Universal Test Fixture. General information for each major product group, as well as specific dimensional data.
Anritsu/Wiltron Sales Company
INFO/CARD #151

Radio Links



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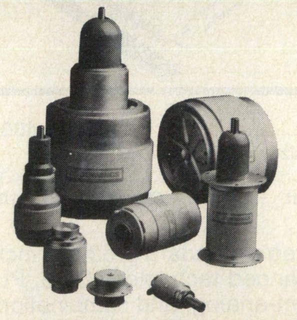

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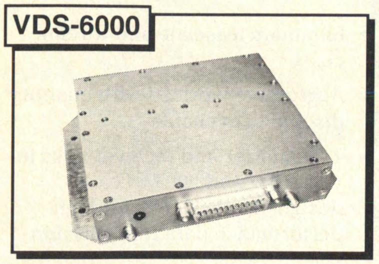
INFO/CARD 102

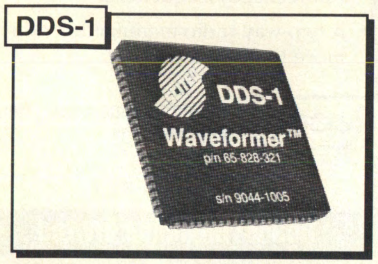
FREQUENCY SYNTHESIS

VDS-6000



BW	**<50% in L-band**
Step size	1.0 MHz
Spurs	<-60 dBc
φ noise @ 1 GHz	-80 dBc/Hz @ 100 Hz
	-90 dBc/Hz @ 10 kHz

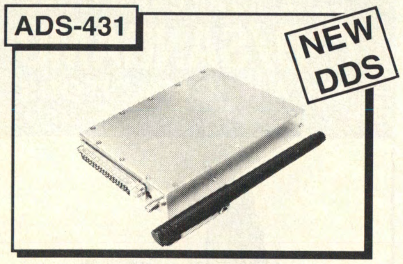
DDS-1



Clock	**up to 25 MHz**
Spurs @ 20 MHz ck	<-60 dBc *after the DAC*
Digital modulation	amplitude, phase, freq
Dimensions	1" x 1", or eval board

ADS-431

NEW DDS



BW	**up to 400 MHz**
Step size	<1 Hz
Spurs	<-45 dBc typical
Output	**Quadrature @ +10 dBm**

Sciteq's standard products include advanced-technology DDS, PLL, direct-analog, and combination synthesizers. Formats include chips, boards, modules, and chassis. Custom designs combine proven modules to minimize risk, cost and time.

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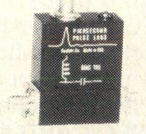
PICOSECOND PULSE LABS.



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Model	Risetime	Low Freq. Cutoff	Insertion Loss	Voltage Max	Current Max
5520A	20 ps	DC	0.4 dB	15 V	30 mA
5530A	28 ps	25 kHz	0.2 dB	200 V	10 mA
5531	28 ps	750 kHz	0.2 dB	1500 V	20 mA
5532A	28 ps	150 kHz	0.2 dB	1000 V	20 mA
5535	32 ps	10 kHz	0.2 dB	50 V	10 mA
5540	8 ps	160 kHz	0.6 dB	50 V	100 mA
5550B	20 ps	100 kHz (<50 mA)	0.9 dB	50 V	500 mA
5555	20 ps	100 kHz	0.9 dB	50 V	500 mA
5575A	32 ps	10 kHz (<20 mA)	0.6 dB	50 V	500 mA
5580	32 ps	10 kHz	1.0 dB	50 V	1 Amp
5590	150 ps	10 kHz	0.1 dB	50 V	500 mA

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California Eastern Laboratories, Inc., (CEL), an employee-owned corporation, and the exclusive North American sales agent for NEC RF and microwave semiconductors, currently has an opening for a Power Bipolar Design Engineer in its Santa Clara office.

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- Security Fault Analysis
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- Related interface hardware

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- Synthesizer Design, Direct Digital
- Power amp and filter design
- MMIC design

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- Parabolic Antenna Design
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- 10 TO 60 GHz

SYSTEMS

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- System Synthesis, System Analysis
- RF Link Budget Analysis
- System Integration/Test
- Customer Interface

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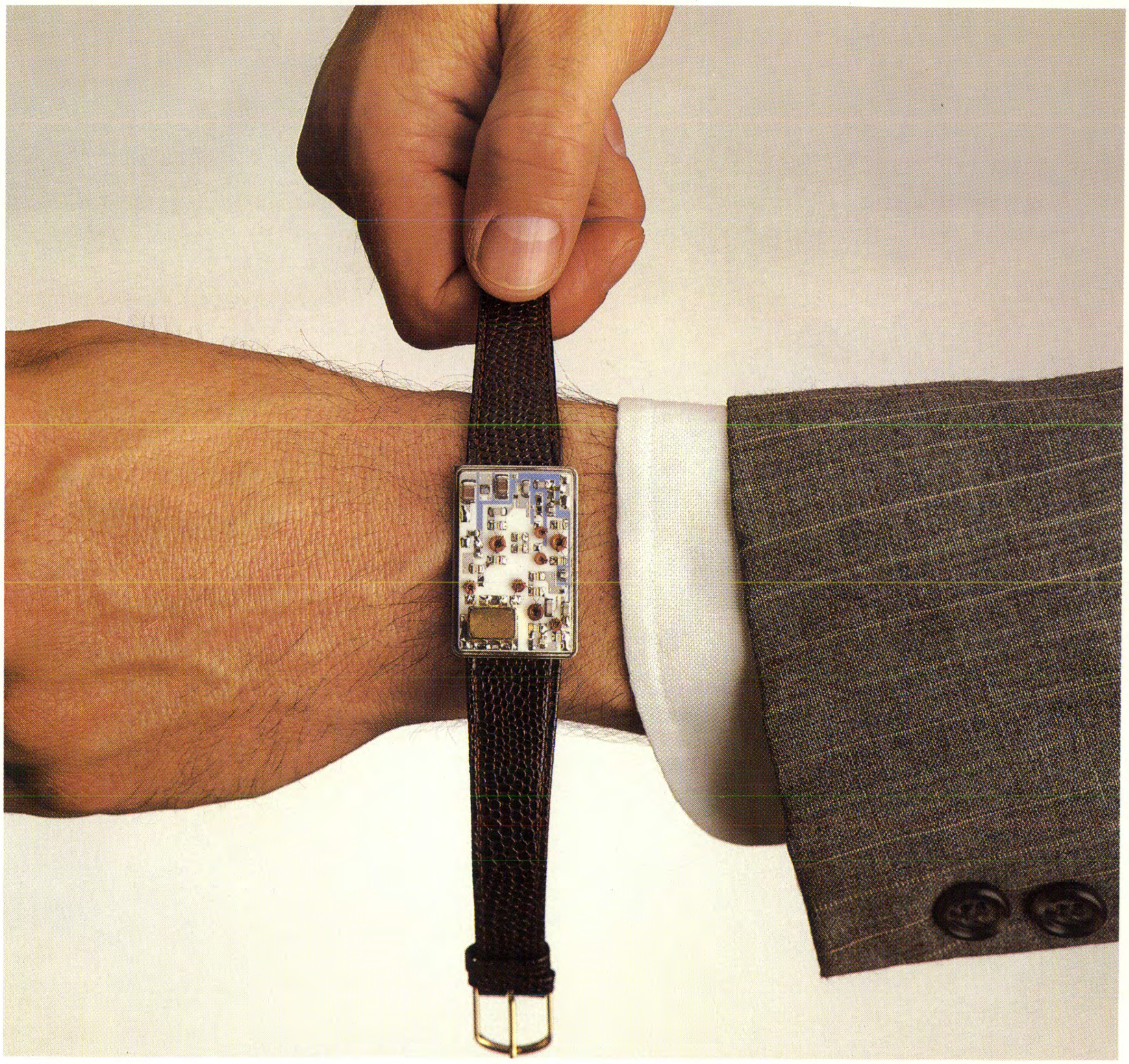

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