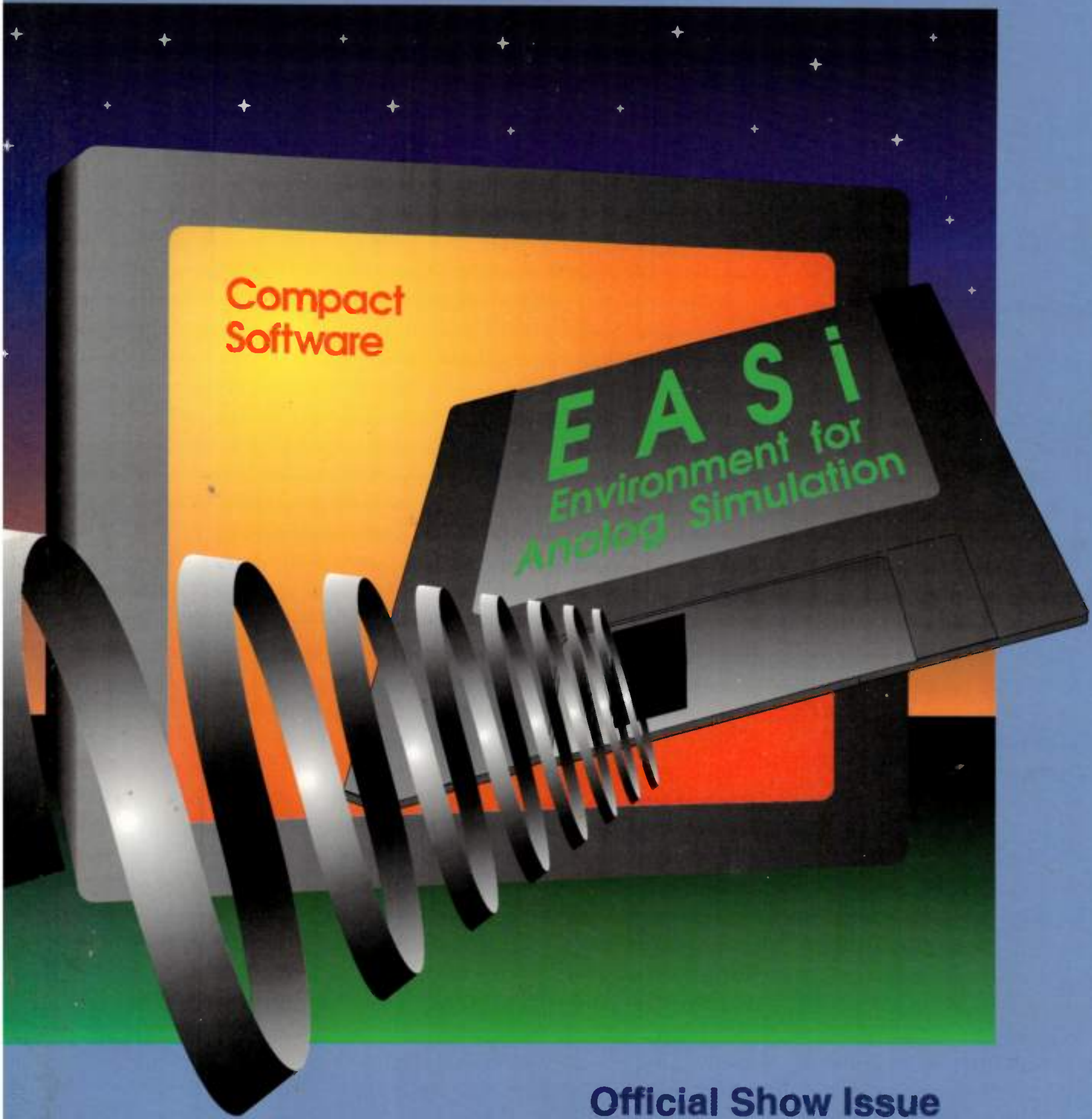
RF design

engineering principles and practices

March 1993

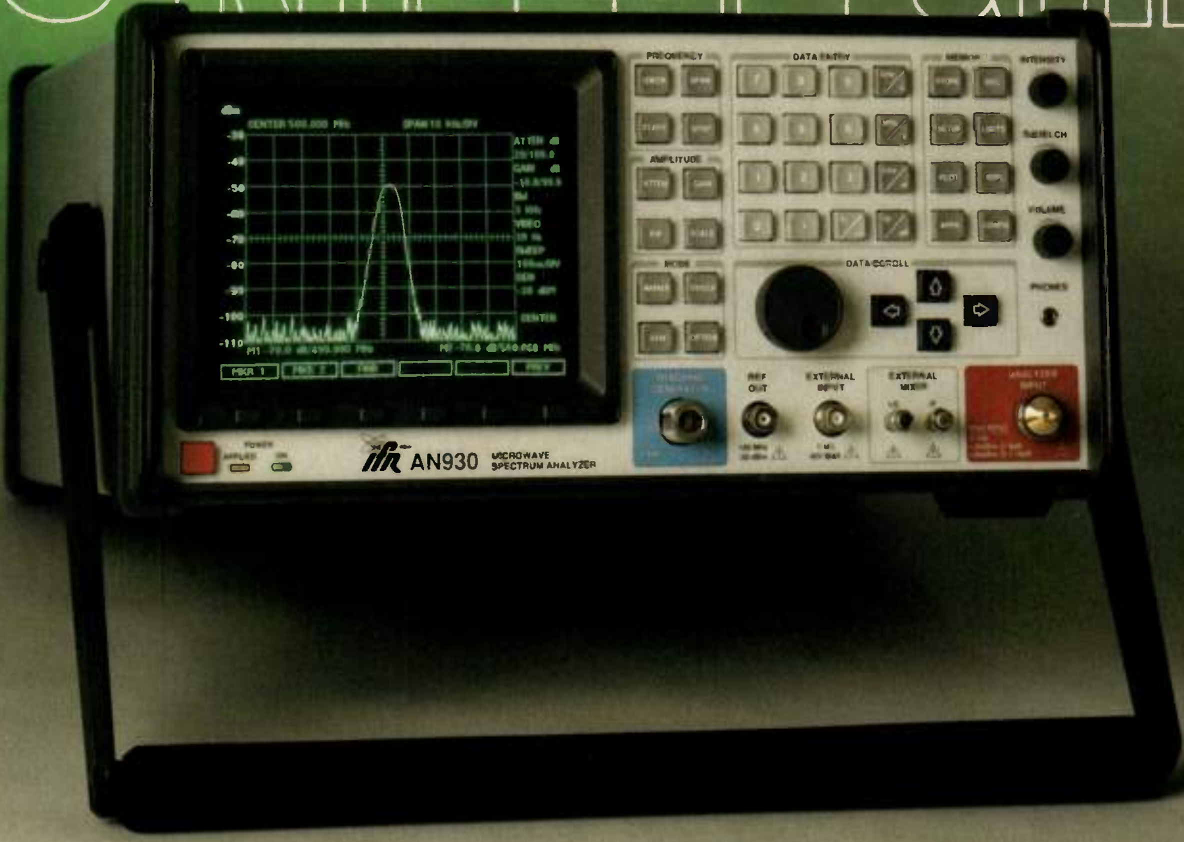


Official Show Issue
RF Expo West

Cover Story
New Workstation Interface
Simplifies Analog Simulation

Performance and value beyond anything in its class...

9 kHz-22 GHz



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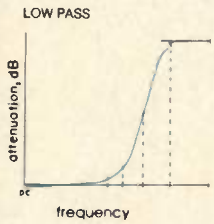


low pass, Plug-in, dc to 1200MHz

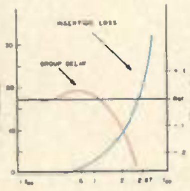
Model No.	Passband MHz loss < 1dB	Stopband, MHz loss > 20dB	Stopband, MHz loss > 40dB	Model No.	Passband MHz loss < 1dB	Stopband, MHz loss > 20dB	Stopband, MHz loss > 40dB
★LP-5	DC-5	8-10	10-200	★LP-250	DC-225	320-400	400-1200
★LP-10.7	DC-11	19-24	24-200	★LP-300	DC-270	410-550	550-1200
★LP-21.4	DC-22	32-41	41-200	★LP-450	DC-400	580-750	750-1800
★LP-30	DC-32	47-61	61-200	★LP-550	DC-520	750-920	920-2000
★LP-50	DC-48	70-90	90-200	★LP-600	DC-680	840-1120	1120-2000
★LP-70	DC-60	90-117	117-300	★LP-750	DC-700	1000-1300	1300-2000
★P-90	DC-81	121-137	167-400	★LP-800	DC-720	1080-1400	1400-2000
★LP-100	DC-98	146-189	189-400	★LP-850	DC-760	1100-1400	1400-2000
★LP-150	DC-140	210-300	300-600	★LP-1000	DC-900	1340-1750	1750-2000
★LP-200	DC-190	290-390	390-800	★LP-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-190	DC-190	290-390	390-800
SCLF-30	DC-30	47-61	61-200	SCLF-380	DC-380	580-750	750-1800
SCLF-45	DC-45	70-90	90-200	SCLF-420	DC-420	750-920	920-2000
SCLF-135	DC-135	210-300	300-600				

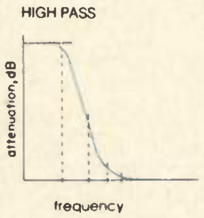
Price, (1-9 qty), all models: $11.45



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
★BLP-39	DC-23	78-117	117	1.3:1	2.3:1	0.7	4.0	5.0
★BLP-117	DC-65	234-312	312	1.3:1	2.4:1	0.35	1.4	1.9
★BLP-156	DC-94	312-416	416	0.3:1	1.1:1	0.3	1.1	1.5
★BLP-200	DC-120	400-534	534	1.6:1	1.9:1	0.4	1.3	1.6
★BLP-300	DC-180	600-801	801	1.25:1	2.2:1	0.2	0.6	0.8
★BLP-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

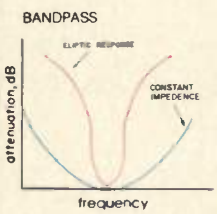
Price, (1-9 qty), all models: plug-in $19.95, BNC $36.95, SMA $38.95, Type N $39.95
NOTE ▲: -933 and -1870 only with connectors, at additional $2 above other connector models



high pass, Plug-in, 27.5 to 2200MHz

Model No.	Stopband MHz loss < 40dB	Stopband MHz loss < 20dB	Passband, MHz loss < 1dB	VSWR Passband Typ	Model No	Stopband MHz loss < 40dB	Stopband MHz loss < 20dB	Passband, MHz loss < 1dB	VSWR Passband Typ
★HP-25	DC-13	13-19	27.5-200	1.8:1	★HP-400	DC-210	210-290	395-1600	1.7:1
★HP-50	DC-20	20-26	41-200	1.5:1	★HP-500	DC-280	280-365	500-1600	1.8:1
★HP-100	DC-40	40-55	90-400	1.8:1	★HP-600	DC-350	350-440	600-1600	2.0:1
★HP-150	DC-70	70-95	133-600	1.8:1	★HP-700	DC-400	400-520	700-1800	1.6:1
★HP-175	DC-70	70-105	160-800	1.5:1	★HP-800	DC-445	445-570	780-2000	2.1:1
★HP-200	DC-90	90-116	185-800	1.6:1	★HP-900	DC-520	520-660	910-2100	1.8:1
★HP-250	DC-100	100-150	225-1200	1.3:1	★HP-1000	DC-550	550-720	1000-2200	1.9:1
★HP-300	DC-145	145-170	290-1200	1.7:1					

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

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
★BP-10.7	10.7	9.6-11.5	8.9-12.7	7.5 & 15	0.6 & 50-1000
★BP-21.4	21.4	19.2-23.6	17.9-25.3	15.5 & 29	3.0 & 80-1000
★BP-30	30.0	27.0-33.0	25-35	22 & 40	3.2 & 99-1000
★BP-60	60.0	55.0-67.0	49.5-70.5	44 & 79	4.6 & 190-1000
★BP-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
★IF-21.4	21.4	18-25	1.3 & 150	DC-220
★IF-30	30	25-35	1.9 & 210	DC-330
★IF-40	42	35-49	2.6 & 300	DC-400
★IF-50	50	41-58	3.1 & 350	DC-440
★IF-60	60	50-70	3.8 & 400	DC-500
★IF-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

NOTE: **★Add Prefix P, B, N, or S for Pin, BNC, N, or SMA connector requirement.**

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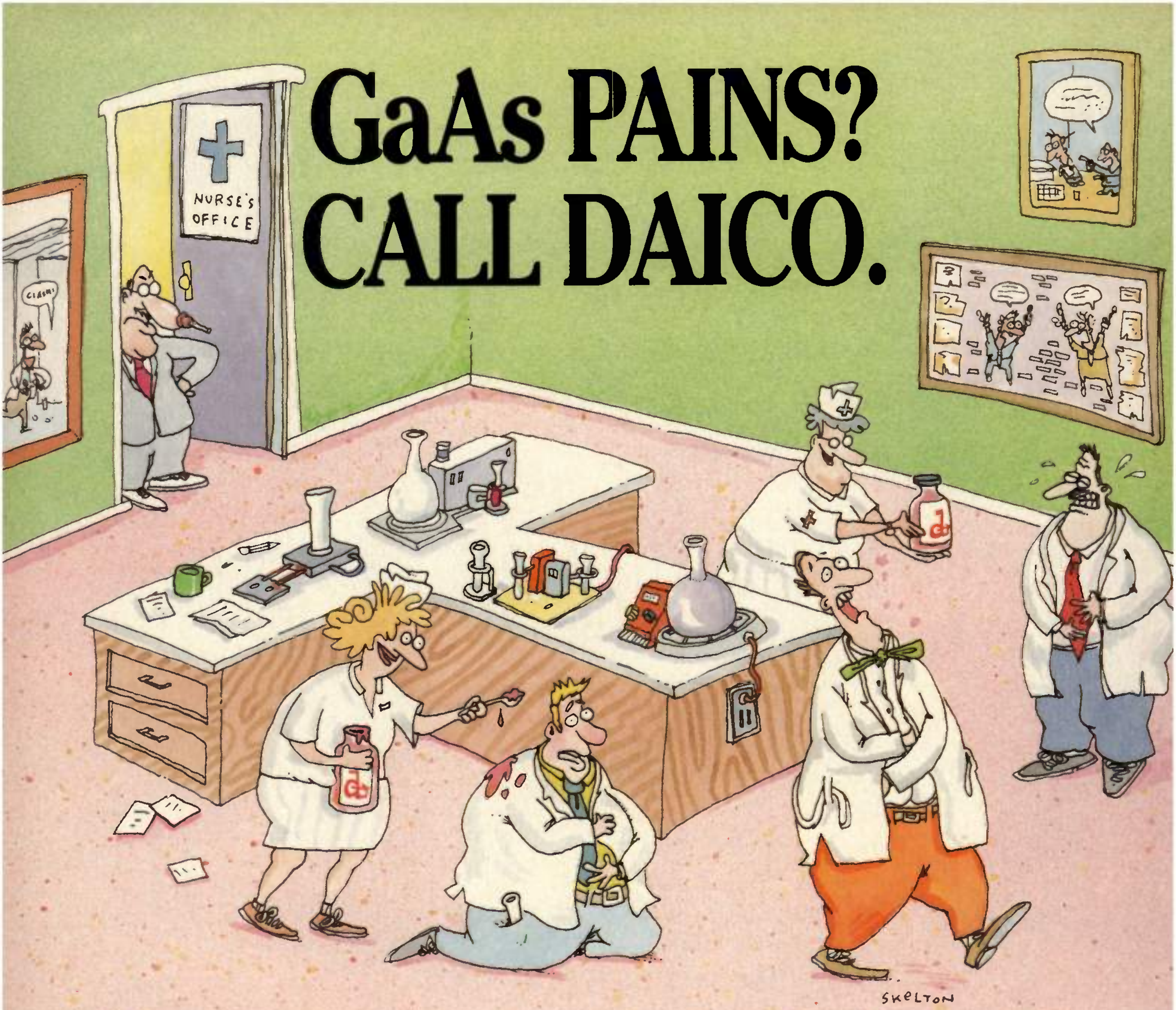

Stock GaAs Bi-Phase Modulators

Freq	IL dB	Switch Speed nSEC	Phase Accuracy Degree	Con-trol	Package	Part No.
10-500 MHz	1.0	50	+/-1	TTL	TO8	DBP0938
10-500 MHz	1.0	50	+/-1	TTL	SMP	DBP0738

Stock Packaged GaAs Switches

Con-fig	Freq MHz	IL dB	Iso dB	Switch Speed nSEC	Con-trol	Package	Part No.
SPST	5-1500	1.0	48	25	TTL	TO-8	DSO699
SPST	10-2000	1.8	67	35	TTL	14 Pin SMP	DSO790
SP2T	DC-2000	0.5	35	3	–	8 Pin SOIC	DSO702R
SP2T	DC-2000	0.6	30	3	–	8 Pin SOIC	DSO702T
SP2T	DC-2000	0.4	68	3	–	TO-5	DSO850
SP2T	DC-2000	0.7	50	200	TTL	TO-5	DSO813
SP2T	5-2000	1.15	55	35	TTL	14 Pin DIP	DSO602
SP2T	5-4000	1.0	79	35	TTL	SMA	CDSO882
SP4T	DC-2000	1.7	70	75	TTL	14 Pin DIP	DSO874

Stock GaAs MMIC Amplifiers

Freq Range/GHz	Typ Gain/dB	Typ Noise Figure	Typ Power /dBm	Package	Comments	Part No.
0.05-3.5	10	6.0	22	Chip		P35-4100-0
0.5-3.5	9	4.5	22	Chip	Self-Biased	P35-4101-0
0.05-3.0	18	6.0	13	Chip	Low VSWRs	P35-4104-0
0.8-1.8	21	3.5	8	Chip		P35-4105-0
1-6	7.5	4.6	20	Chip		P35-4110-0
6-18	5.5	5.5	15	Chip	Pos.Gain Slope	P35-4140-0
2-18	6.0	7.5	15	Chip	AGC	P35-4150-0
3-6	2.0	2.8	14	Chip	Low VSWRs	P35-4160-0

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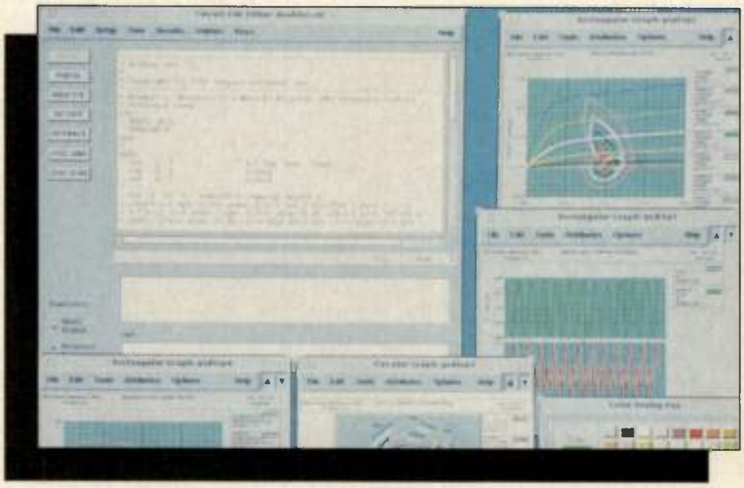

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RF editorial

A Message to Washington

By Gary A. Breed
Editor

Now that all the new legislators and political appointees have had time to outfit their offices, hire their staffs, and locate the best restaurants in central Washington, it's time to get to work. Part of the reason that the public elected a new President is because Mr. Clinton promised change.

One change needed by the RF industry is quick action on new technologies. Senators, Representatives and agencies: it's time to admit that technology moves faster than you are accustomed to responding! In today's global marketplace, we can no longer afford to deal with these issues in a leisurely manner.

To Senators and Representatives — It's time to wake up to the fact that dividing up the pork barrel district-by-district may help you get reelected, but it's bad for the country as a whole. Your leadership is needed to get the U.S. financial house in order, to establish a good business and trade climate, and to solve many other problems that make life difficult. Get to work on the budget deficit, the absurd level of national debt, the cost of health care and the banking crisis. Find ways to encourage businesses to invest in research and in advanced manufacturing equipment. Help their employees, too, by addressing modern issues of family leave, job training and workplace safety, and by reducing the tax burden of the working middle-class. Give us tools for the future by seeking answers to our educational decline and by keeping a strong base of scientific and medical research.

To the Federal Communications Commission — You are governing the radio spectrum, where the next electronic revolution is underway. Advanced technologies that represent huge markets for U.S. companies are in your hands. Take a leadership role in developing or encouraging technical standards and making responsible frequency allocations. Change your approach from *de*-regulation to *appropriate* regulation — the past approach of leaving the fate of technology to the marketplace has failed too many times. With so many new ideas being tossed about, chaos will rule the marketplace. The FCC *must* provide guidance and leadership.

To President Clinton and close advisors — You promised to work hard to help American business and the American people get back on the road to prosperity. We hope you can maintain the right balance between the needs of owners and workers. Trickle-down economics doesn't work; so let's try your approach to simultaneously help the top, middle and bottom of the ladder.

To all of us — It is too easy to blame Washington or our statehouses for the problems that confront us. But remember, *we* elect the officials that set policy and *we* need to hold them accountable. We also need to keep our own houses in order by controlling our personal finances, by influencing our employers to be responsible businesses, and by teaching the next generation of Americans the lessons that will help them succeed.

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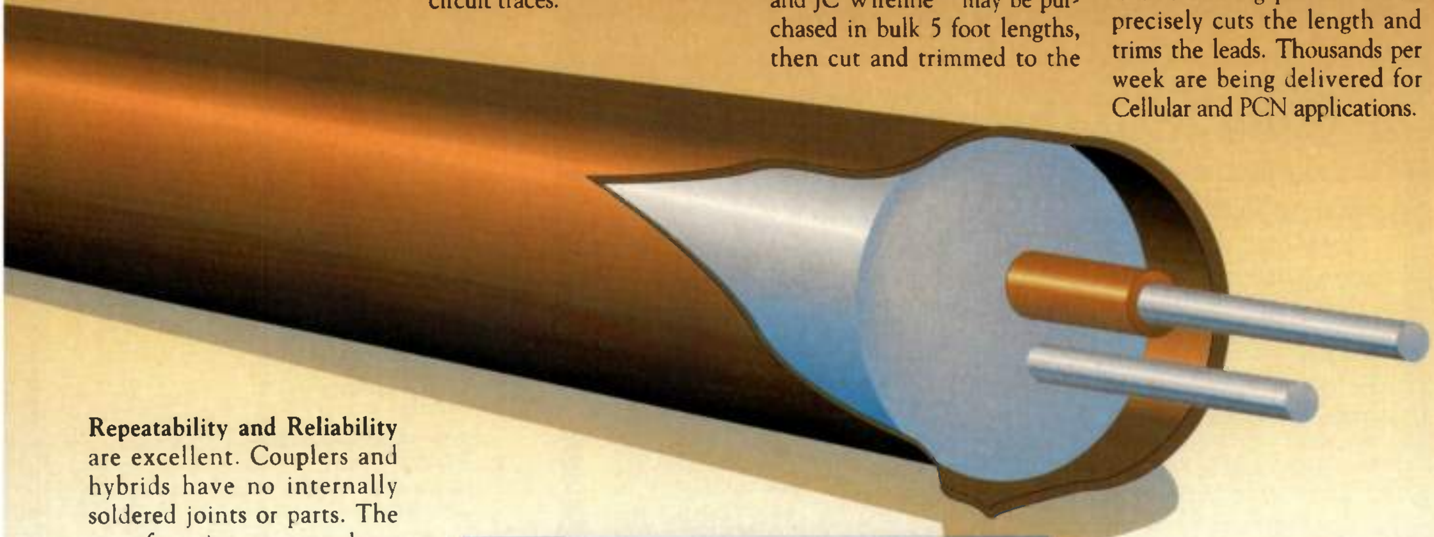
From 25 MHz to 18,000 MHz

Miniature 0.050 Dia. Wireline™ Hybrids · New 10 dB Wirepac™ Couplers

Wireline™ and Wirepac™ are 90° hybrid and coupler product lines which use Sage Laboratories, Inc. patented, balanced, twin conductor technology. These caseless products provide unequaled design and layout flexibility at an affordable price. Coil it to reduce coupling length, shape it to fit around layout obstacles and even loop it to jump over circuit traces.

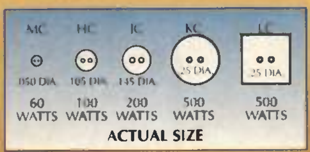
Couplers by the foot, a great design tool for engineers. HC and JC Wireline™ may be purchased in bulk 5 foot lengths, then cut and trimmed to the correct quarter wavelength for optimum frequency response. Need a second iteration, just cut another length. The coupling curves are virtually text book.

100,000 unit Production quantities, no problem. Sage has developed a state of the art laser trimming process which precisely cuts the length and trims the leads. Thousands per week are being delivered for Cellular and PCN applications.



Repeatability and Reliability are excellent. Couplers and hybrids have no internally soldered joints or parts. The manufacturing processes have been qualified for space environments. Wireline™ and Wirepac™ are performing demanding hybrid functions on such space programs as SCS Superbird and TIROS. The quality is designed in, so commercial programs run without delays or failures.

Average Power ratings from 60 to 500 WATTS, strictly a function of outline diameter.



DROP-IN QUADRATURE COUPLERS AND HYBRIDS

PART NUMBER	FREQ MHz	MID BAND COUPLING	INSERTION LOSS	ISOL.	VSWR
HCB2-5.93	225-400	2.7 dB	0.3 dB	20 dB	1.2:1
JCB2-2.80	440-880	2.7 dB	0.3 dB	20 dB	1.2:1
LCB2-2.63	500-1000	2.7 dB	0.2 dB	30 dB	1.1:1
MCB3-2.06	810-990	2.8 dB	0.3 dB	20 dB	1.2:1
GCB6-2.06	810-990	10.25 dB	0.1 dB	30 dB	1.2:1
KCB2-1.31	1000-2000	2.7 dB	0.2 dB	30 dB	1.1:1
HCB1-1.00	1710-1990	3.0 dB	0.2 dB	20 dB	1.2:1
GCB6-1.00	1710-1990	10.25 dB	0.1 dB	30 dB	1.2:1
MCB3-0.14	9000-18000	2.8 dB	0.3 dB	20 dB	1.2:1

Many Wireline™ family standard products are available. Octave and narrow band products with three form/power factors. The new miniature MC Wireline™ hybrids offer significant size reduction without sacrifice of reliability or performance. The new GCB6 10 dB coupler provides octave performance with 20 dB directivity. Above are just a few of our product offerings.

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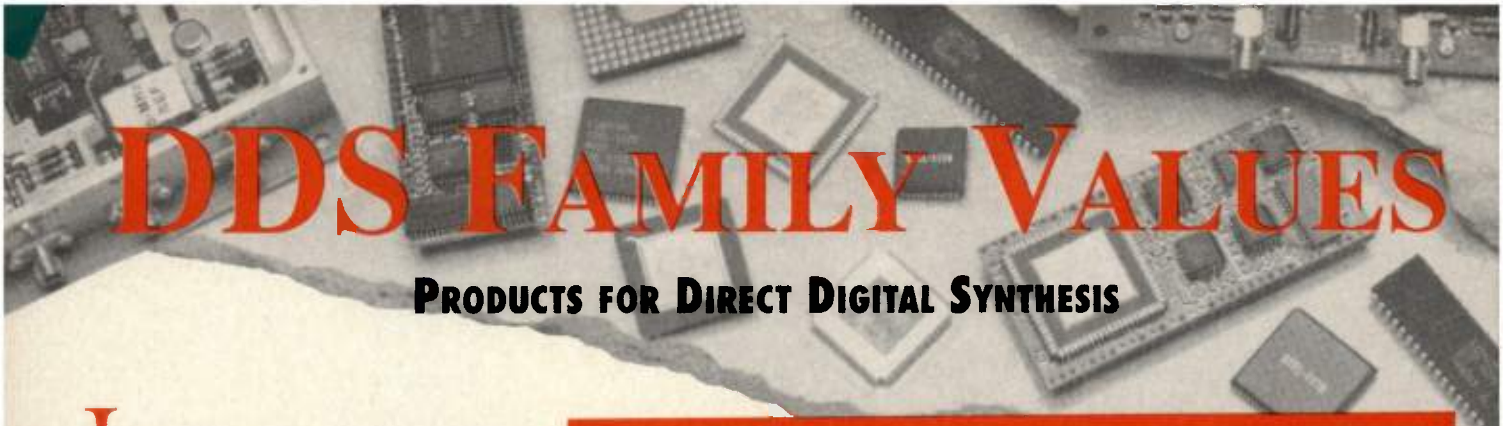

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Monolithic NCOs

Product	Clock Frequency	Description
STEL-1130	60 MHz	Linear Quadrature Amplitude Modulator
STEL-1172B	50 MHz	32-bit NCO, Quadrature Outputs
STEL-1173	50 MHz	48-bit NCO, Sub-MicroHz Resolution
STEL-1173RH	40 MHz	Rad Hard with 1 MRad Tolerance
STEL-1174	50 MHz	16-bit NCO, Low Price
STEL-1175	80 MHz	32-bit NCO with Linear PM
STEL-1176	80 MHz	83/4 Decade Decimal NCO with BCD Control
STEL-1177	60 MHz	32-bit NCO with Linear PM and FM ports
STEL-1178A	80 MHz	32-bit NCO Dual with PSK
STEL-1179	25 MHz	24-bit NCO, PSK and Low Price
STEL-1180	60 MHz	32-bit Chirp Generating NCO
STEL-2172	300 MHz	28-bit ECL NCO
STEL-2173	1 GHz	GaAs NCO with PSK

Board Level DDS Products

Product	Clock Frequency	Description
STEL-1272	Quadrature output 0 to 22 MHz	DDS
STEL-1273	0 to 22 MHz	DDS with Sub- MicroHz Resolution
STEL-1275	0 to 35 MHz	DDS with Linear PM
STEL-1276	0 to 35 MHz	DDS with 0.1 Hz Resolution and BCD Control
STEL-1277	0 to 35 MHz	DDS with Linear PM and FM
STEL-1375A	0 to 35 MHz	Miniature DDS Module with Linear PM
STEL-1376	0 to 35 MHz	Miniature DDS Module with BCD Control
STEL-1377	0 to 35 MHz	Miniature DDS Module with Linear PM and FM
STEL-1378A	Dual 0 to 35 MHz	Miniature DDS Module with PSK
STEL-1479	0 to 12 MHz	Hybrid DDS, 1.5" by 0.8", Low price
STEL-2272	0 to 130 MHz	DDS
STEL-2273A	0 to 400 MHz	DDS with PSK
STEL-2373	0 to 400 MHz	DDS Hybrid with PSK, 2" by 1.1"

Chassis Level DDS Products

Product	Clock Frequency	Description
STEL-9272	0 to 130 MHz	Complete DDS with Phase Lockable Clock
STEL-9275	0 to 400 MHz	Complete DDS with Phase Lockable Clock

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For sales support outside the continental United States, please contact your International Sales Representatives.

United Kingdom: BFI - IBEXSA Electronics Ltd. Mr. Mick Mercer Phone: +44 622 882467 FAX: +44 622 882469

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INFO/CARD 9

Please see us at RF Expo West, Booths #721, 723.

RF calendar

March

15-18 **IEEE Multi-Chip Module Conference**
Santa Cruz, CA
Information: MCMC-93, Attn: Jean McKnight, Computer Engineering, University of California, Santa Cruz, CA 95064. Tel: (408) 459-2303. Fax: (408) 459-4829.

17-19 **RF Expo West**
San Jose, CA
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.

23-25 **Nepcon Electronics**
Birmingham, UK
Information: Reed Exhibition Companies, 999 Summer St., PO Box 3833, Stamford, CT 06913-0130.

24-31 **CeBIT '93**
Hannover, Germany
Information: Hannover Fairs USA Inc., 103 Carnegie Center, Princeton, NJ 08540. Tel: (609) 987-1202. Fax: (609) 987-0092.

30-31 **The 1993 Mid-Lantic Electronics Show**
King of Prussia, PA
Information: Mid-Lantic Electronics Show '93, Judith Ginsberg, 4113 Barberry Drive, Lafayette Hill, PA 19444. Tel: (215) 828-2271. Fax: (215) 941-6773.

30-2 **8th International Conference on Antennas and Propagation**
Edinburgh, UK
Information: Conference Services, IEE, Savoy Place, London WC2R 0BL, UK. Tel: (44) 071 240 1871. Fax: (44) 071 497 3633.

April

18-21 **The 4th IEE Conference on Telecommunications**
Manchester, UK
Information: ICT 93 Secretariat, Conference Services, IEE, Savoy Place, London, WC2R 0BL, UK. Tel: (44) 071 240 1871. Fax: (44) 071 497 3633.

18-22 **Symposium on Ceramics for Wireless Communication**
Cincinnati, OH
Information: Henry O'Bryan, AT&T Bell Laboratories. Tel: (908) 582-6980. Fax: (908) 582-2521.

18-22 **The NAB Multimedia World Conference and Exhibition**
Las Vegas, NV
Information: NAB, 1771 N Street, NW, Washington, DC 20036-2891. Tel: (202) 429-5350. Fax: (301) 216-1847.

28-30 **EMC/ESD International 1993**
Denver, CO
Information: Renae Fierros, Cardiff Publishing Company, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111. Tel: (303) 220-0600, (800) 525-9154. Fax: (303) 773-9716.

RF courses

Cellular Radio
May 11-14, 1993, Madison, WI
Information: The University of Wisconsin-Madison, Engineering Information. Tel: (800) 462-0876.

Infrared Technology and Applications
April 6-8, 1993, Atlanta, GA
Principles of Electronic Counter-Countermeasures
April 6-8, 1993, Atlanta, GA
Phased-Array Radar System Design
April 13-16, 1993, Atlanta, GA
Information: Georgia Institute of Technology, Continuing Education. Tel: (404) 894-2547.

Microwave/Millimeter-Wave Monolithic Integrated Circuits
May 11-14, 1993, Los Angeles, CA
Hybrid Microcircuit and Multichip Module Packaging Technologies
May 24-26, 1993, Los Angeles, CA
Radiation Hardening of Electronic Systems
June 7-11, 1993, Los Angeles, CA
Information: UCLA Short Course Program Office. Tel: (310) 825-1047. Fax: (310) 206-2815.

Modern Receiver Design
March 29-April 2, 1993, London, England
Telecommunication Traffic Engineering
March 15-17, 1993, Washington, DC
Modern Digital Modulation Techniques
March 29-April 1, 1993, Washington, DC
Modern Digital Signal Processing: Analysis, Design and Applications
March 29-April 2, 1993, London, England
Microwave System Engineering
March 29-April 2, 1993, London, England
April 26-30, 1993, Washington, DC
Mobile Cellular Telecommunication Systems
April 7-9, 1993, Washington, DC
Grounding, Bonding, Shielding and Transient Protection
April 20-23, 1993, Washington, DC
Ionospheric Radio Propagation: Principles and Application
April 20-23, 1993, Washington, DC
Communication Satellite Engineering
April 26-30, 1993, Washington, DC
Analog/RF Fiber Optic Communications
May 26-28, 1993, Washington, DC
Anomalous Microwave and RF Propagation
June 2-3, 1993, Washington, DC
Electromagnetic Interference and Control
June 7-11, 1993, Washington, DC
Spread-Spectrum Communication Systems
June 7-11, 1993, Washington, DC
Information: The George Washington University, Continuing Engineering Education, Merril A. Ferber. Tel: (202) 994-8522 or (800) 424-9773.

Modern RF & Microwave Techniques
April 20-23, 1993, Phoenix, AZ
Navstar/GPS
June 2-4, 1993, Monterey, CA
Information: University Consortium for Continuing Education, 16161 Ventura Boulevard, M/S C-752, Encino, CA 91436. Tel: (818) 995-6335. Fax: (818) 995-2932.

Electromagnetic Compatibility and Interference
May 4-7, 1993, St. Cloud, FL
September 14-17. 1993, San Diego, CA
Information: SCEEE, Central Florida Facility - Management Office, 1101 Massachusetts Ave, St. Cloud, FL 94769, Kelly Brown - Registrar. Tel: (407) 892-6146. Fax: (407) 957-4535.

Satellite Communication Systems: Techniques and Technology for Communications and Broadcasting
April 19-23, 1993, Cambridge, United Kingdom
Combined Coding and Modulation Techniques
April 20-21, 1993, Cambridge, United Kingdom
Low Sidelobe Antennas
May 10-14, 1993, Cambridge, United Kingdom
Modern Microwave Techniques: Measurements, Signal and Network Analysis, Microwave Products and Systems Characterization
May 10-14, 1993, Cambridge, United Kingdom
Error Correcting Codes and Trellis-Coded Modulation with Application to Communication Systems
June 7-10, 1993, Stockholm, Sweden
Frequency-Time Signal Processing: Applications and Algorithms for High Resolution Spectral Analysis and Time Series Analysis
June 7-11, 1993, Stockholm, Sweden
Far-Field, Compact & Near-Field Antenna Measurement Techniques
June 7-11, 1993, Stockholm, Sweden
Information: CEI-Europe/Elsevier, Mrs. Tina Persson. Tel: (46) 122-175-70. Fax: (46) 122-143-47.

Electronic Design Techniques and Analysis Required to Meet Electromagnetic Compatibility Requirements
May 5-6, 1993, Novi, MI
Advanced EMC Printed Circuit Board Design
May 7, 1993, Novi, MI
Information: JASTECH, James P. Muccioli. Tel: (313) 553-4734

1993 High-Speed Digital Symposium
April 20, 1993, San Francisco, CA
April 22, 1993, Los Angeles, CA
April 27, 1993, San Diego, CA
April 29, 1993, Phoenix, AZ
May 25, 1993, Dallas, TX
May 27, 1993, Austin, TX
June 2, 1993, Burlington, MA
June 4, 1993, Washington, DC
June 8, 1993, Boca Raton, FL
Information: Hewlett-Packard Company, Microwave Instruments Division (MID). Tel: (800) 765-9200.

Inherently Conductive Polymers, an Emerging Technology
March 15-17, 1993, Burlington, VT
Information: M. Aldissi, Champlain Cable, 12 Hercules Drive, Colchester, VT 05446. Tel: (802) 655-2121. Fax: (802) 655-2025.

Collision Avoidance (IVHS) Topical, Technical Workshop
April 13, 1993, Washington, DC
Information: SAE International (The Engineering Society for Advanced Mobility Land Sea Air and Space), 400 Commonwealth Drive, Warrendale, PA 15096-0001. Tel: (412) 776-4841. Fax: (412) 776-5760.

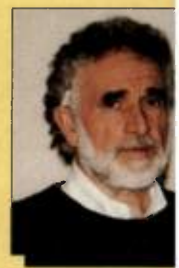
WHITE NOISE NEWS

NEWS AND TECHNICAL INSIGHT FROM NOISE COM, INC./VOLUME 1 NUMBER 1

ABOVE THE NOISE — BY GARY SIMONYAN

What's *your* application?

The more you know about white noise sources, the more useful they become. You can design them into a system for built-in test, use them to evaluate the bit-error rate of satellite and terrestrial digital communications systems, test electronic components, reduce spurious in A/D converters...the list is long.



Gary Simonyan

The problem is, many more people need to know how much white noise can do. If you wonder how white noise can serve your application, please call me personally at (201) 261-8797.

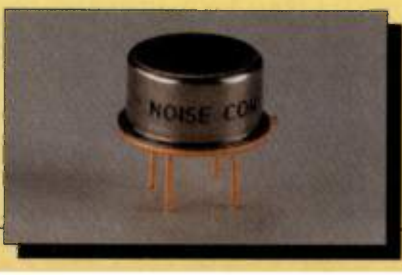
NOISE IN THE FIELD

CATV instruments use built-in noise sources

A leading manufacturer of cable TV test equipment is integrating Noise Com's NC 502/15 TO-8 noise sources into its briefcase and hand-held CATV signal strength meters.

The NC 502/15 is the built-in reference in the instruments, and was selected because of its guaranteed output power flatness of less than +/-1.0 dB and frequency coverage up to 1000 MHz. The NC 502/15's high efficiency, -53 dBm output with typically only +15 VDC at 2.5 mA, was another criteria.

The NC 502/15 is part of Noise Com's NC 500 Series of noise sources mounted in TO-8 packages for built-in test. They range in frequency to 5 GHz and have flatness from +/-0.5 dB. They all feature small size (0.5 in. dia.) and low cost.



NEW FROM NOISE COM

UFX Series redefines noise-based testing

Noise Com's new NC 7000 Series white noise-generating instruments earned a fine reputation for quality and performance. Now our new UFX Series instruments extend this capability, with dedicated function keys, simplified programming, a 4 x 20 character LCD display, and an even greater range of options. You can even create and run extensive test routines automatically under program control. They're simply the most advanced noise-generating instruments available.

Typical Specifications

Output	Up to +13 dBm
Frequency bands available	Up to 40 GHz
VSWR	1.5:1
Attenuators	0 to 127 dB in 1-dB steps or 0 to 128 dB in 0.1 dB steps
*Combiner	Combines input signal with noise
*Filter banks	Act on noise, signal, or both. Up to four of any type

**Optional*

INFO/CARD 16

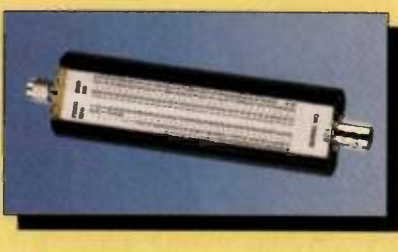
Please see us at RF Expo West, Booth #445.

Precision Noise Sources

NC 346: A quality alternative for noise figure measurement

Noise Com's new NC 346Ka is the only noise figure meter-compatible noise source that delivers flat, broadband white noise from 100 MHz to 40 GHz. It's compatible with the HP 8970 and other popular noise figure meters, and has low VSWR across the entire band.

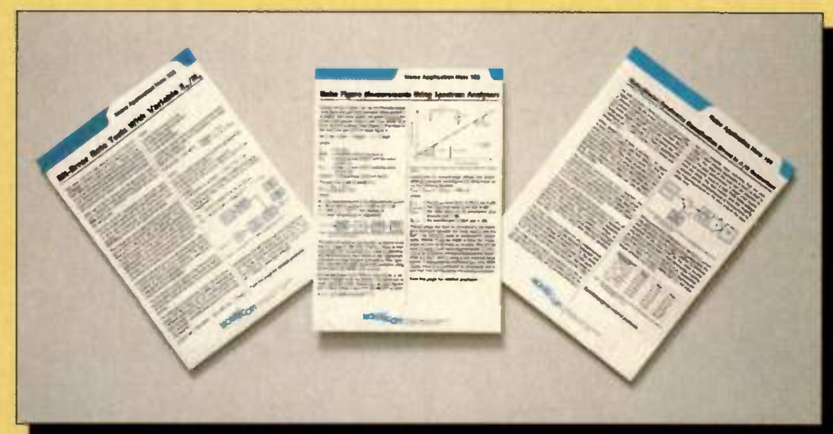
The NC 346Ka has a built-in regulator that makes it insensitive to input voltage variations. The units are fully calibrated at frequency points throughout their operating range...and they're available from stock.



Specifications

Frequency range (GHz)	Output VSWR
0.1 to 5	1.25:1
5 to 18	1.30:1
18 to 26.5	1.40:1
26.5 to 40	1.50:1

ENR	10 to 17 dB
Input power	+28 VDC +/-2 V at 30 mA maximum. Built-in regulator



New literature

Free ap notes describe white noise applications

The latest in Noise Com's expanding series of application notes describe three uses for white-noise generating devices and instruments. Application Note 103 covers bit-error rate (BER) testing with variable E_b/N_o, Application Note 105 covers noise figure measurement using a spectrum analyzer, and Application Note 109 covers a technique for reducing harmonic distortion in A/D converters. Each application note fully describes a technique and the steps required to implement it. The application notes are absolutely free and you can get them by calling Noise Com at (201) 261-8797, or by circling the reader service number below.

The white noise adviser

Q: *What is ENR?*
A: Excess Noise Ratio (ENR) expresses how many times the effective noise temperature (T_e) of a noise source exceeds that of a load at the reference temperature, 290 K.

$$ENR = 10\log\left(\frac{T_e - 290}{290}\right)$$

ENR can be conveniently applied in noise figure measurements (see Noise Application Note 115). ENR and T_e are parameters calibrated with traceability to the National Institute of Standards and Technology (NIST).

Q: *How do we convert from ENR in dB to the effective noise power spectral density, N_o, in dBm/Hz?*
A: For ENR values up to 30 dB, take the effective noise temperature (T_e) of the noise source and insert it in the following equation:

$$N_o \text{ (in dBm/Hz)} = 10\log(kT_e) \text{ dBm/Hz}$$

where $k = 1.38 \times 10^{-23}$ J/K is Boltzmann's constant. A table for these conversions is provided on page 17 of Noise Com's catalog (call us for a free copy). The following approximation may be used for ENR values greater than 30 dB:

$$N_o \text{ (in dBm/Hz)} = \text{ENR (in dB)} - 174 \text{ dBm/Hz}$$

Q: *What is the difference between dBm and dBm/Hz?*
A: The power of white noise is evenly distributed over the frequency of interest, and is commonly specified as power spectral density in dBm/Hz (which refers to power in a 1-Hz bandwidth). To obtain the total noise output power in dBm, multiply the power density by the noise bandwidth or integrate over the frequency range. This is equivalent in dB to adding 10log(NBW) where NBW is the noise bandwidth in Hz. The total noise power in dBm is therefore equal to the noise power spectral density in dBm/Hz plus 10log(NBW).

NOISE/COM

E. 49 Midland Avenue, Paramus, New Jersey 07652
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PCS Trials in Pittsburgh

Tests are currently underway in Pittsburgh, Pennsylvania for a cellular-based single number, single phone communications service. The trials are being conducted by Bell Atlantic Mobile in conjunction with faculty, staff and graduate students at Carnegie Mellon University. The personal line service is based on Motorola's PPS-800™ technology and takes advantage of the wired and wireless networks already in place to give customers total accessibility and user-defined features. A "smart" pocket-sized Personal Communicator telephone manufactured by Motorola is being used in the service trial. It operates as a high quality cordless phone in the home; as an extension of a PBX or Centrex system in the office; and as a cellular phone when the user is mobile.

The marketing trial will involve more than 500 customers and last approximately nine months. The company expects to learn how customers — ranging from large corporate users to individual entrepreneurs — respond to the expanded communications opportunities delivered by the service, including accessibility, call screening, and other features. Because most trial participants will be paying for their equipment and service, the company also expects to learn about "real world" market demands for this type of service.

MMS Consortium Releases Specifications — At a recent meeting, the Modular Measurement System Consortium approved the release of the specifications for the Modular Measurement System. The Consortium is composed of nine prime contractors, instrument manufacturers, and system integrators. The MMS specifications define the technical details and requirements of the MMS architecture, allowing instrument manufacturers to completely evaluate the MMS and to build equipment that is fully compatible with that from other MMS manufacturers. The two-volume set of MMS Specifications contains full descriptions and drawings of the electrical and mechanical interfaces, including communication protocols and EMC aspects. Copies are available at $50 per set from Cal Central Press, PO Box 551, Sacramento, CA 95812-0551; phone 1-800-821-3481, extension 3434.

DoD Further Clarifies BeO Status — The Defense Electronic Supply Center has further clarified its position on the use of beryllium oxide (BeO) in military electronic devices. According to its most recent announcement, DESC is requiring that military devices containing the material be labeled with the chemical symbol "BeO". DESC has written the labeling requirement for BeO into MIL Spec S-19500 covering semiconductor devices, MIL Spec M-38150 covering microcircuits and MIL Spec H-87111 covering heat sinks. Since the electronics industry uses only solid forms of the material, the action clarified beryllia's role for continued use in substrates and insulators.

Tektronix Acquires Analytek Limited — Tektronix, Inc. recently announced the acquisition of Analytek Limited. Analytek is a five year old, privately held company specializing in high

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INFO/CARD 18

Please see us at RF Expo West, Booths #430, 432, 444, 446, 529, 531, 543, 545, 547.

performance VME card modular digitizers for single-shot and multi-event signal analysis. Terms of the acquisition were not released.

Spectrum Allocated for LEOSATS — The Federal Communications Commission recently allocated spectrum space for satellite based data communications services. At least three companies are expected to be granted licenses by late 1993 or early 1994. The 137-138, 148-150.05, 399.9-400.05 and 400.15-401 MHz bands have been set aside for tracking, messaging and document distribution applications. Volunteers in Technical Assistance (VITA) currently uses the technology under an FCC Pioneer's Preference license to transmit documents and messages to developing countries. Three other companies, LEOSAT, Starsys Global Positioning, and Orbcomm have expressed interest in developing products for the newly allocated spectrum.

Digital Radio Proposals — Five technical proposals were recently submitted to the Electronic Industries Association for a digital audio radio system. The EIA is planning on starting tests on the systems on April 15. The companies submitting proposals are AT&T Bell Laboratories; AT&T and Amati Communications Corp.; Thomson, for the European Eureka-147 digital audio broadcasting consortium; the Jerrold division of General Instrument; and NASA and Voice of America.

US Tech Moves Headquarters — US Tech has announced the move of their headquarters and primary testing facility from Roswell, Georgia to 3505 Francis Circle, Alpharetta, GA 30201. An open house will be held on March 12th from noon to 4:00 p.m. for those interested in touring the new facility. Call (404) 998-9454 for more information.

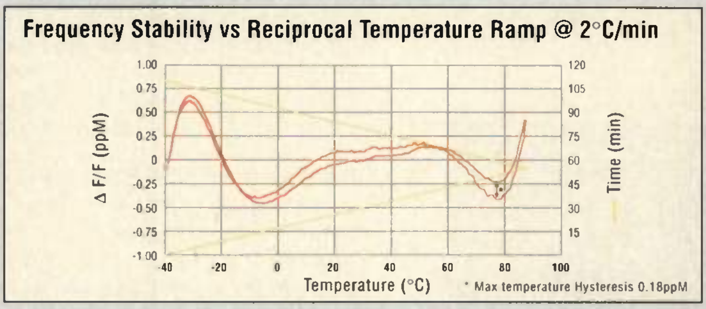
Dale Electronics Awarded ISO 9001 Registration — The first ISO 9001 registration granted by Underwriter's Laboratories to a U.S. resistor manufacturer has been awarded to the resistor facility of Dale Electronics, a business unit of Vishay Electronic Components.

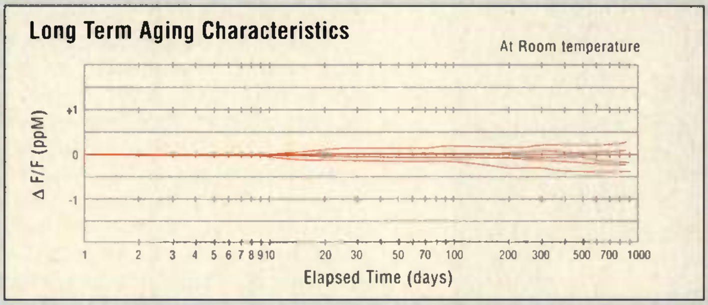
Haefely Undergoes Name Change — Haefely Test Systems, Inc. recently announced a name change to Haefely, Inc. This change was necessary in order to include the new responsibility of selling Haefely's utility and power compensation products, in addition to high voltage test systems for power equipment and EMC test systems for electronics and aerospace applications.

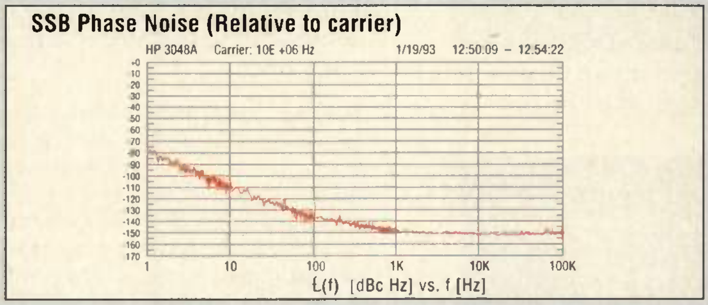
ISO 9001 Registration Awarded to Okaya — Okaya Electric America has announced that its parent company and manufacturing facilities have received certification to ISO 9001. This quality certification outlines a model for quality assurance in design, development, production, installation and servicing. The ISO 9001 Certification of Okaya includes the manufacturing facilities for EMI/RFI Power Line Noise Filters, AC Noise Suppression Capacitors and Spark Quenching R-C Networks.

RAKON TCXO's

OUTSTANDING QUALITY & PERFORMANCE







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INFO/CARD 20

Unitrek Acquires Cable Operation — Unitrek Corp. has announced the acquisition of the cable manufacturing operation of Tektronix, Inc. Unitrek was formerly an operation of Tektronix and spun off in 1990. The acquisition of the cable manufacturing unit assures Unitrek of a stable supply of specialty cables for its custom products. Terms of the sale were not disclosed.

Genisco Relocates Facilities — Genisco Electronics has relocated its final assembly, engineering, test operations, and administrative offices from San Diego, Calif. to its parent company facilities in Anaheim, Calif. Their new address is 1230 South Lewis Street, Anaheim, CA 92805. Tel: (714) 563-4300. Fax: (714) 563- 4355.

Motorola Receives Chinese Cellular System Order — Motorola recently announced that its Cellular Infrastructure Group received its largest China cellular system order. The Shanghai Post and Telecommunication Administration awarded Motorola an expansion contract to add to its existing system. The contract will bring the system total to more than thirty cell sites extending coverage in the city of Shanghai. Shanghai was one of the first cities in China to use cellular communications.

Radian and WELS Research Sign Joint Agreement — Radian Corporation recently completed an agreement with WELS Research Corporation to produce an integrated weather decision support system for environmentally sensitive industries and activities. For the new weather information decision support system, Radian will provide sensor data to support a WELS-developed, object oriented forecasting system. A key source of sensor data will come from Radian's family of high performance remote sensing instruments, including the LAP™-3000 radar wind and temperature profiler and the Echosonde Doppler acoustic sounder. WELS will support the development of the new system with its local area weather forecasting system called Weather Pro.

ASTeX Awarded DARPA Contract — Applied Sciences and Technology, Inc. announced that it has been awarded a contract from Defense Advanced Research Projects Agency for the development of equipment and processing technologies for diamond substrate material to be used in multichip modules. The contract for the first year is $600,000 with optional second year funding of $930,000. The ASTeX award is aimed at the rapid commercialization of cost effective techniques for large-area diamond substrates to be produced at high growth rates and to be used for thermal management of electronic packages.

US West and Two Russian Groups Selected to Coordinate Cellular in Russia — US West International, Intertelecom and VART have been selected to coordinate a consortium to construct an integrated 900 MHz GSM digital cellular telephone service in Russia. US West and its partners were awarded the rights to build and operate cellular systems in eight of the 12 regions available for development.

NEW!

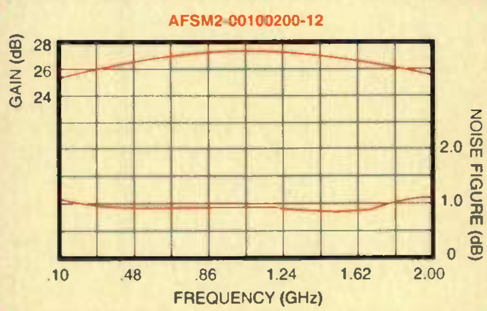
SURFACE-MOUNT AMPLIFIERS

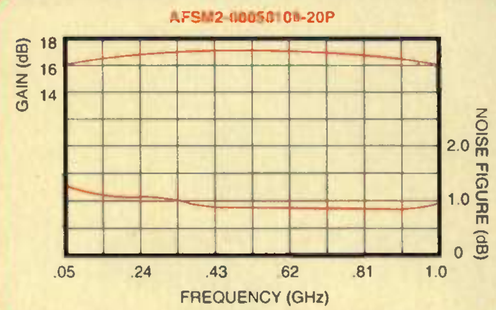
Finally, AFS low-noise performance in surface-mountable packages.

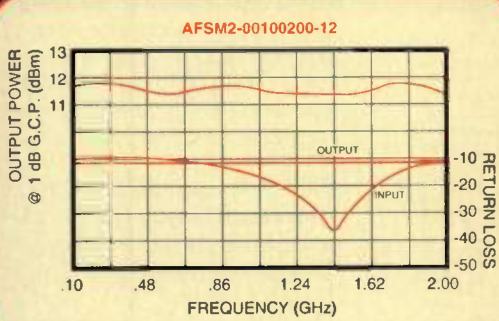
Commercially available low-noise amplifiers in surface-mountable chassis have lagged in noise figure and dynamic range performance when compared to amplifiers in removable connector AFS chassis. A few months ago MITEQ started a development program to study the feasibility of achieving comparable performance with the commercial SMTO-8 chassis.

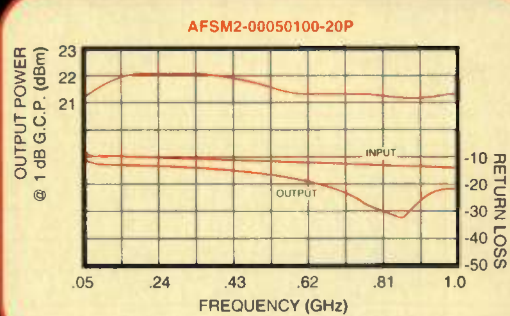
The successful results of this development are summarized below:

MODEL NUMBER	FREQUENCY (GHz)	GAIN (MIN.) (dB)	GAIN VAR. (MAX.) (±dB)	NOISE FIGURE (MAX.) (dB)	VSWR (MAX.) IN	VSWR (MAX.) OUT	1 dB GAIN COMP. OUTPUT (MIN., dBm)	NOM. DC POWER (+15V, mA)
AFSM2-00100200-12	.1-2	22	±1	1.2	2:1	2:1	+8	80
AFSM2-00100300-14	.1-3	20	±1	1.4	2:1	2:1	+8	80
AFSM2-00100400-16	.1-4	20	±1	1.6	2:1	2:1	+8	80
AFSM2-00100600-18	.1-6	18	±1.5	1.8	2:1	2:1	+8	80
AFSM2-00500100-11	.5-1	25	±1	1.1	2:1	2:1	+8	80
AFSM2-01000200-10	1-2	25	±1	1.0	2:1	2:1	+8	80
AFSM2-02000400-13	2-4	21	±1	1.3	2:1	2:1	+8	80
AFSM2-02000600-15	2-6	20	±1	1.5	2:1	2:1	+8	80
AFSM2-00050100-20P	.05-1	13	±1	1.5	2:1	2:1	+20	150
AFSM2-01000200-20P	1-2	13	±1	1.6	2:1	2:1	+20	150









For additional information please call Hillar Kiiss on extension 259.

ACTUAL SIZE

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INFO/CARD 22

ARFTG Call for Papers — The Automatic RF Techniques Group has issued a call for papers for its 41st conference. The conference will be held on June 18, 1993 in Atlanta, Georgia in conjunction with the Microwave Theory and Techniques Symposium. The conference theme is "Managing Information and Automated Testing". Two copies of the abstract and summary should be sent to Richard Henle, Technical Program Chairman, and should be received no later than April 8, 1993. His address is: Mr. R.A. Henle, JHU/APL, 13-S389, Johns Hopkins Road, Laurel, MD 20723-6099. Tel: (301) 953-6000 ext. 7819.

Scientific-Atlanta to Provide VSAT Network to Thailand — Scientific-Atlantic recently announced that it will provide a $4 million very small aperture terminal network to Thai Skycom Ltd., a common carrier based in Bangkok, Thailand. The VSAT network will enable TSL to offer satellite based data, voice, fax, and video communications services to businesses and governments throughout the Asia-Pacific region. In its initial stage, the VSAT network will consist of a turnkey 11-meter satellite hub, a 4.5 meter mini-hub, 50 VSAT terminals, and 10 digital MCPC terminals. More VSAT terminals will be added as TSL customer demand grows. The network is expected to be fully operational within six months.

Harris Awarded Radio Contract — The Sacramento Air Logistics Center has awarded a multi-million dollar contract to Harris RF Communications for AN/URC-119(V) radios to be used by the U.S. Air Force and the U.S. Navy under the PACER BOUNCE program. The initial contract is for $3.44 million, with options for an additional two fiscal years, which could bring the total to $9.95 million. The AN/URC-119(V) is a high-performance HF radio designed to provide reliable, easily maintained long-range voice and data communications for fixed plant, transportable, shelter and mobile stations.

Illinois Superconductor Wins U.S. Commerce Dept. Contract — Illinois Superconductor has won a multi-million dollar federal Advanced Technology Program contract for a joint project with AT&T and Ameritech to use ISC's high temperature superconductor technology to dramatically improve wireless communications. The three-year, $3.5 million ISC program will develop a prototype cellular telephone base station that will use high temperature superconductor technology to increase the number of channels, improve signal reception and stabilize radio frequencies.

Delfin Acquires Maxim Technologies and Major Contract — Shortly after announcing the merger of Delfin Systems and Maxim Technologies, Delfin Systems' new Signal Products Division was awarded a contract by the U.S. Department of Defense. While specific details of the DoD award were not made available, it is the first multiple order for the division's new SIGINT Manpack System (SMS). SMS is a line of HF/VHF/UHF signal intelligence and direction finding equipment.

Where Does HDTV/ATV Stand?

By Gary A. Breed
Editor

The present status of high definition television HDTV (also called advanced television, ATV) is still something of a guess. The search for a terrestrial broadcast standard is proceeding slowly in the U.S., the largest television market and one of the few countries where the delivery of programming is not provided by the government. While sweeping changes in television technology can be readily implemented in many countries by simple mandate, HDTV in the U.S. must be phased in without disturbing the huge installed base of transmitting and receiving equipment.

HDTV Background

High definition television is based on a few fundamental improvements over the current worldwide television standards. First, both horizontal and vertical are increased approximately fourfold. Together with other video processing techniques, the result is a picture with roughly the same detail as a standard 35 mm photograph. Next, the proportions of the picture will be 5:3, horizontal to vertical, which matches the proportions of a motion picture screen (current television is 4:3). In side-by-side demonstrations of studio-quality video, the HDTV systems are an obvious improvement — especially when viewed on a large, wall-size screen.

At the video baseband, HDTV requires a minimum of 25 MHz bandwidth, compared to 4.2 MHz for the current U.S. NTSC standard. The principal problem if HDTV then arises — it requires the bandwidth of more than four 6 MHz television channels. If the principal medium of television transmission were coaxial cable or microwave satellite links, this bandwidth could be accommodated. However, terrestrial television transmission is "over the airwaves" and already occupies a significant portion of the radio spectrum. The principal problem, then, is the development of a narrower bandwidth transmission mode that allows transmission of as much of the improved quality as possible, without excessive spectrum occupancy.

The Federal Communications Commission (FCC) recently concluded testing of five systems competing to become the ATV broadcast standard in the U.S. Performance factors included in the tests were both subjective and objective, based on viewer response and measurements of the received signals. Perceived quality, transmission range, interference, and other factors have been evaluated for the five systems. The approval of a standard was to have been completed by mid-year.

- General Instruments — DigiCipher
- David Sarnoff Labs, Thomson Consumer Electronics, Philips Consumer Electronics, National Broadcasting Company — Advanced Digital HDTV
- Zenith Corporation, AT&T — Digital HDTV
- Massachusetts Institute of Technology, General Signal — Channel Compression
- NHK (Japanese Group) — Narrow MUSE

The five competing systems for the U.S. HDTV broadcast standard.

However, testing revealed greater interference to existing NTSC transmissions than expected. By the time this report is published, the FCC is expected to have announced re-testing of the systems with modifications to reduce the interference level. If this proceeds as expected, a standard will not likely be determined until early 1994.

The issue of re-testing has arisen before, with system developers expressing concern about matters ranging from quantitative evaluation of actual picture resolution to noisy video sources and improper equipment setup. Re-testing for interference may allow some of these other testing matters to be addressed at the same time.

In the meantime, other portions of the HDTV arena are being addressed. Scientific Atlanta has been pursuing a digital compression technique that requires smaller bandwidth for satellite or microwave transmission of full-quality HDTV video, with a minimum of loss in quality. Chip sets for the major systems are in development laboratories at major semiconductor manufacturers, each hoping that they are betting on the right system. Limited HDTV broadcasting in Japan, both terrestrial and via satellite, is a reminder that HDTV was initially developed in that country.

The Cost of Viewing

The FCC's Advanced TV Systems Advisory Committee recently released an estimate of the initial cost of receivers for HDTV. Very small differences were found among the five different systems. Consumers purchasing the first HDTV sets can expect to pay around $2500 for a 34-inch direct-view (picture tube) set, or $3800 for a 56-inch projection system. The prices are surprisingly comparable to other high-end consumer products.

Prices are expected to drop rapidly as programming becomes available in the HDTV format, and as accessories such as HDTV VCRs (already developed and waiting for the market) become widely available. Another major factor is the degree of quality improvement perceived by the viewers, not just experts and television professionals. On the broadcast end, studio equipment is already developed, with the major networks and many broadcast stations budgeting for upgraded facilities to originate HDTV programming.

At present, the estimates of a phase-in time range from 7 to 20 years following the first U.S. HDTV broadcasts. Concurrent NTSC and HDTV broadcasts will be continued as long as necessary to assure that existing television sets are not rendered obsolete. **RF**

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	225LC	25W CW	.01-225 MHz	40dB	$ 3,295
	*737LC	25W CW	.01-1000 MHz	45dB	$ 9,995
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NEW	718FC	100W CW	10-1000 MHz	50dB	$ 26,950
	*757LC	100W CW	.01-1000 MHz	50dB	$ 29,950
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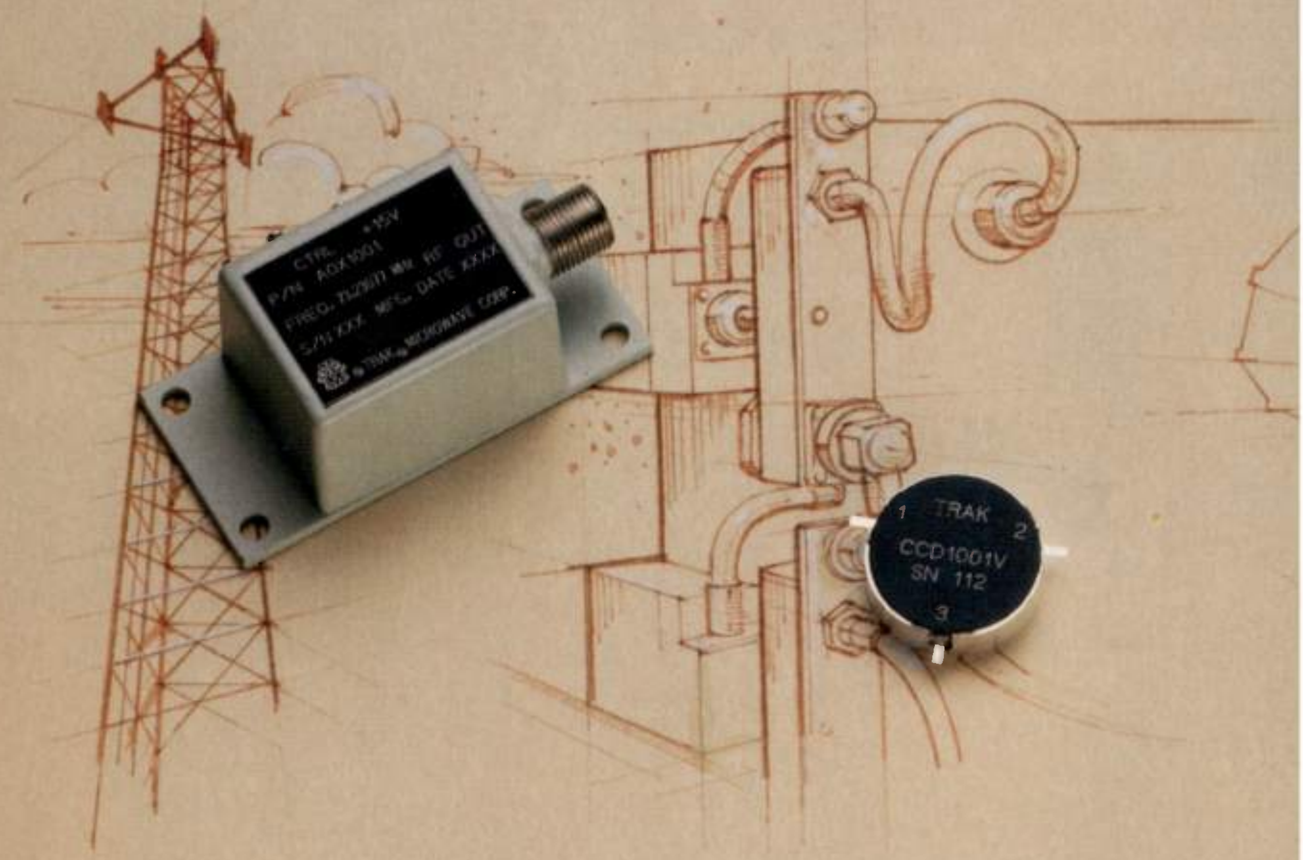

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INFO/CARD 27

An Introduction to RF/Microwave Filter Design

By David R. Loker
The Pennsylvania State University at Erie

Filter design is a common and useful means of attenuating or rejecting unwanted signals such as out-of-band signals and noise in receivers. Often it is necessary to remove additional signals which are a result of processing by non-linear circuits such as harmonic generators and saturating amplifiers. Most high frequency filter designs or RF/microwave filter designs use transmission line stubs as the basic elements of the filter, but lumped-element filters are occasionally used. In this paper, both types of filters will be discussed.

The purposes of this paper are to present a review of lumped-element filter design, specify the relationship between lumped-elements and distributed-elements, and design RF/microwave filters using transmission lines.

Filters are classified by their frequency selection characteristics. The principal categories are lowpass, highpass, bandpass, and bandstop. Also, various parameters are used to describe filter performance. For the lowpass and highpass filters, the cutoff frequency is the frequency at which the passband ends and is defined by some dB of attenuation. For bandpass and bandstop filters, two such cutoff frequencies or band edge frequencies exist which are defined by some dB of attenuation. The difference in frequency between these two band edge frequencies is the bandwidth. In addition, some types of filters exhibit small changes in the attenuation characteristic of the passband. This is called the ripple and is the maximum allowable attenuation in the passband. Other performance parameters will be discussed later in the paper.

Filter design consists of finding the number of elements needed and the specific values for each element to give the desired characteristics. One attack on this problem is known as Darlington Synthesis (1). Rather than carry out this synthesis procedure, normalized tables (2) exist which permit the determination of each element of the filter. It simply becomes necessary to scale the element values for the desired cutoff frequency and system impedance. This will form the basis of filter design in this paper.

Lumped-Element Filter Design

There are many types of filter response characteristics. These are listed in Reference 3 and are the Butterworth filter, the Chebyshev filter, the elliptical filter, and the Bessel filter. This paper will concentrate on the design of Chebyshev filters since they provide the maximum possible stopband insertion loss for a specified maximum passband insertion loss.

The Chebyshev lumped-element lowpass filters consist of a ladder network of series inductors and shunt capacitors. As explained earlier, the values for the inductors and capacitors and the filter order are based upon the desired filter characteristics.

The equation governing the Chebyshev response is shown below.

$$|S_{21}|^2 = \frac{1}{1+\varepsilon^2 T_n^2(\omega)} \qquad (1)$$

where:

$T_n^2(\omega) \equiv \cos^2(N\cos^{-1}\omega)$ for $\omega \le 1$

$\equiv \cosh^2(N\cos^{-1}\omega)$ for $\omega \ge 1$

$\varepsilon^2 \equiv 10^{r/10} - 1$

$r \equiv$ ripple in dB

$N \equiv$ filter order as an integer

$|S_{21}|^2 \equiv$ forward transmission coefficient using S-parameters

Thus, insertion loss (IL) is defined as shown below.

$$IL = +10LOG\,[1 + \varepsilon^2(\cos^2(N\cos^{-1}\omega))] \quad \omega \le 1 \qquad (2)$$

$$= +10LOG\,[1 + \varepsilon^2(\cosh^2(N\cosh^{-1}\omega))] \quad \omega \ge 1 \qquad (3)$$

The filter order can be determined by using the following formula.

$$N \ge \frac{\cosh^{-1}\sqrt{\dfrac{10^{\frac{A}{10}} - 1}{\varepsilon^2}}}{\cosh^{-1}(\omega_s\ \omega_c)} \qquad (4)$$

where:

N = filter order (number of reactive elements)

ε^2 = ripple factor = $\left(10^{\frac{r}{10}} - 1\right)$

r = ripple in dB

ω_c = cutoff frequency

ω_s = stop frequency

A = attenuation in dB at ω_s

The elemental values for the Chebyshev filter are found using the following equations:

$$R = \tanh^2(B/4) \quad \text{for N even} \qquad (5)$$

$$R = 1 \quad \text{for N odd} \qquad (6)$$

$$g_1 = 2a_1/p \qquad (7)$$

$$g_k = \frac{4a_{k-1}a_k}{b_{k-1}g_{k-1}} \quad \text{for } K = 2, 3, \ldots, N \qquad (8)$$

where:

$$B = \ln\left[\coth\frac{r}{17.37}\right]$$

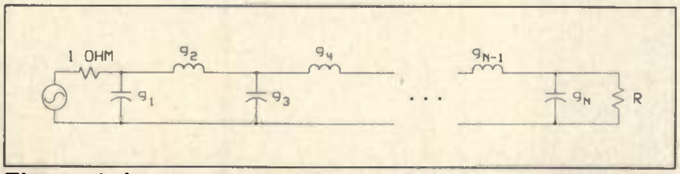


Figure 1. Lowpass prototype network with capacitive input.

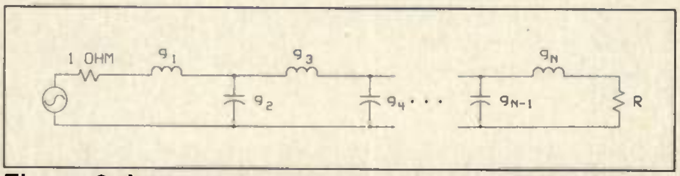


Figure 2. Lowpass prototype network with inductive input.

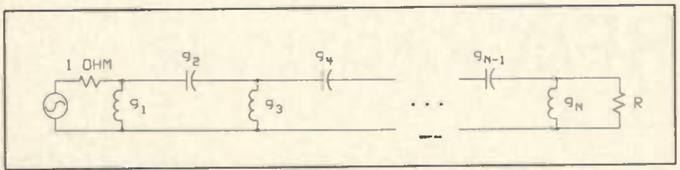


Figure 3. Highpass prototype network.

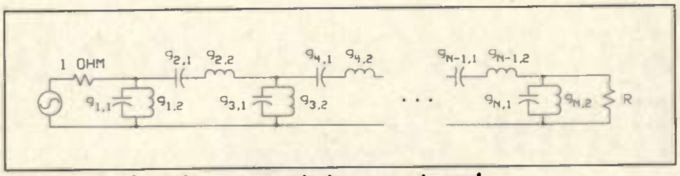


Figure 4. Bandpass prototype network.

$r \equiv$ ripple in dB

$$a_k = \sin\left[\frac{(2k-1)\pi}{2N}\right] \quad \text{for } k = 1,2,\ldots,N$$

$p = \sinh(B/2N)$

$b_k = p^2 + \sin^2(k\pi/N)$ for $k = 1,2,\ldots,N$

$R \equiv$ terminating resistance

Where it is inconvenient to program these equations on the computer, tables (2) can be used which give the elemental values for a given ripple characteristic and filter order.

There are two possible lowpass prototype networks using the elemental values discussed previously. From Figures 1 and 2, two possible filter configurations are shown: one having a capacitive input and one having an inductive input. The choice between the two designs is based on the availability and cost of the capacitors and inductors, and on any stability considerations. For example, some amplifiers will oscillate with capacitive or inductive loading at high frequencies.

As mentioned previously, the filters shown in Figures 1 and 2 are normalized for a system impedance of 1 ohm and a cutoff frequency of 1 rad/sec. Thus, to find the practical filter, the elemental values must be scaled. In order to scale for a system impedance, R, and a cutoff frequency, ω_c, the following equations are used.

$$L_i \rightarrow \frac{Rg_i}{\omega_c} \tag{9}$$

$$C_i \rightarrow \frac{g_i}{R\omega_c} \tag{10}$$

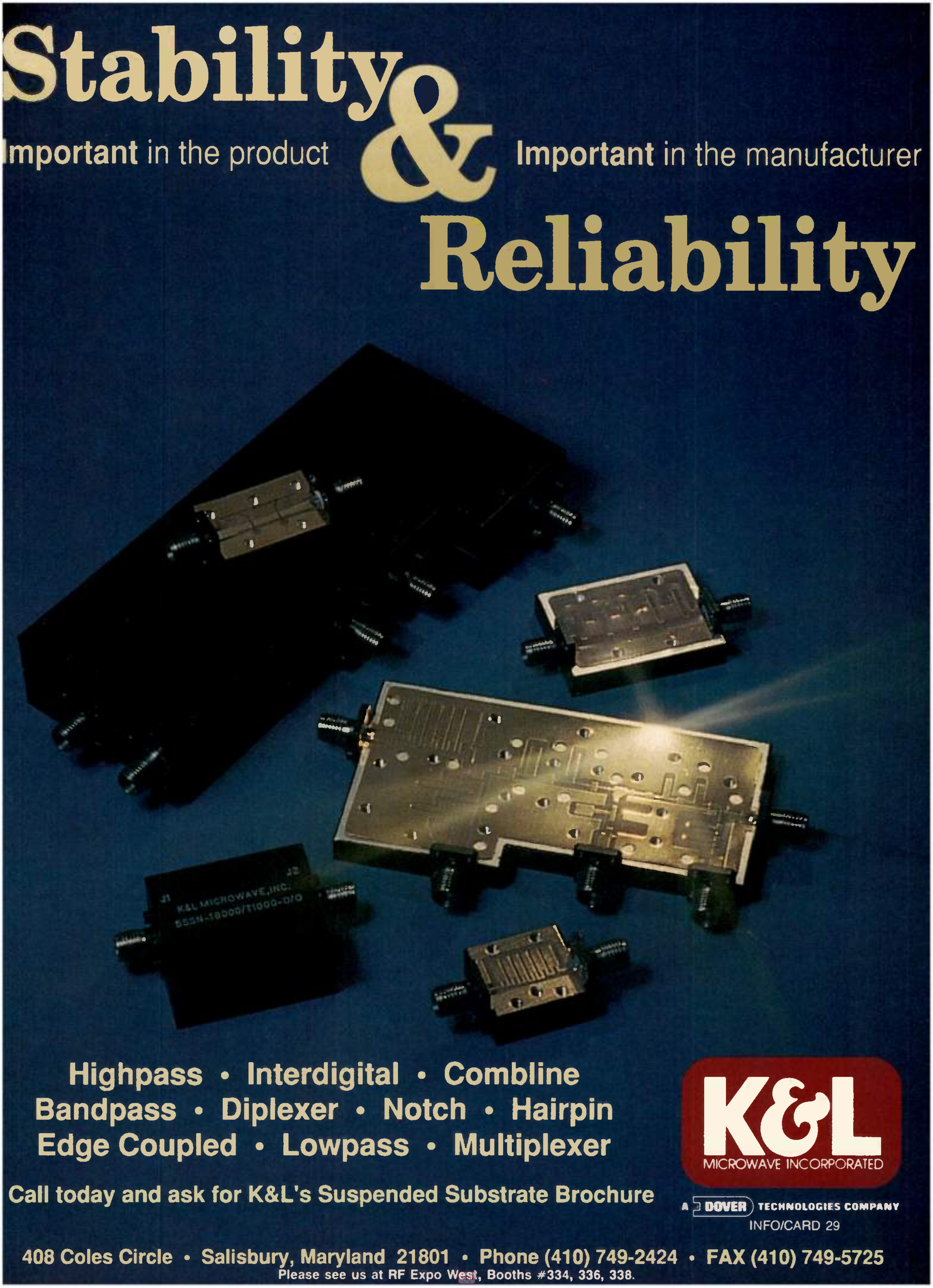
In general, the design procedure consists of determining the filter order based upon a specified attenuation in the stopband, calculating the required normalized elemental values based upon the filter order and amount of specified ripple, and scaling for the desired system impedance and cutoff frequency.

Based upon this general design procedure for lowpass lumped-element filters, one can use simple transformations for the design of highpass, bandpass, and bandstop filters. For the highpass filter case, the following transformations are used.

$$g_i \rightarrow 1/g_i \tag{11}$$

$$\text{series L} \rightarrow \text{series C} \tag{12}$$

$$\text{shunt C} \rightarrow \text{shunt L} \tag{13}$$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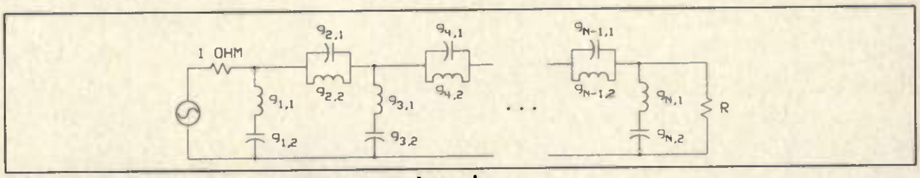

Figure 5. Bandstop prototype network.

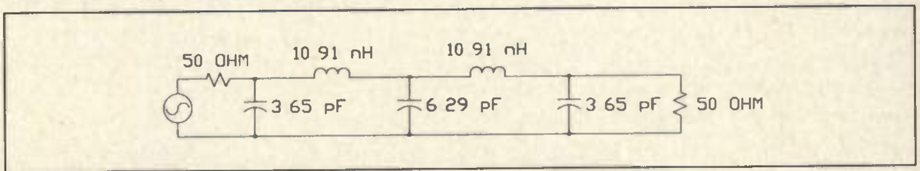

Figure 6. Chebyshev LPF lumped-element network.

The capacitive input lowpass prototype network of Figure 1 is transformed into the highpass network shown in Figure 3.

For the bandpass filter case, the following transformations are used.

$$\text{series } L \rightarrow \text{series } LC \tag{14}$$

$$\text{where } L = \frac{g_i R}{\Delta\omega},\ C = \frac{\Delta\omega}{g_i R \omega_0^2}$$

$$\text{shunt } C \rightarrow \text{shunt } LC \tag{15}$$

$$\text{where } L = \frac{R\Delta\omega}{\omega_0^2 g_i},\ C = \frac{g_i}{R\Delta\omega}$$

$$\omega \rightarrow \frac{\omega_0}{\Delta\omega}\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right) \tag{16}$$

where:

ω_0 = ctr. freq. in rad/sec. $= \sqrt{\omega_1\omega_2}$
ω_1, ω_2 = lower and upper band-edge freq. in rad/sec.
$\Delta\omega$ = bandwidth = $\omega_2 - \omega_1$

Shown in Figure 4 is the bandpass prototype network using the previous transformations. The following transformations are used for the bandstop filter.

$$\text{series } L \rightarrow \text{shunt } LC \text{ (in parallel)} \tag{17}$$

$$\text{where } L = \frac{g_i R}{\Delta\omega},\ C = \frac{\Delta\omega}{g_i R \omega_0^2}$$

$$\text{shunt } C \rightarrow \text{series } LC \tag{18}$$

$$\text{whre } L = \frac{R\Delta\omega}{\omega_0^2 g_i},\ C = \frac{g_i}{R\Delta\omega}$$

where:

$$\omega \rightarrow \frac{\omega_0}{\Delta\omega}\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right) \tag{19}$$

where:

ω_0 = ctr. freq. in rad/sec. $= \sqrt{\omega_1\omega_2}$
ω_1, ω_2 = lower and upper band-edge freq. in rad/sec.
$\Delta\omega$ = bandwidth = $\omega_2 - \omega_1$

The bandstop prototype network is given in Figure 5.

As an example, consider a lowpass filter with the following requirements.

1. r = 0.1 dB
2. f_c = 1.0 GHz
3. f_s = 2.0 GHz
4. A = 30 dB
5. R = 50

In accordance with these filter requirements, a fifth-order filter is used. After scaling for impedance and frequency, the resulting network is shown in Figure 6. This circuit is simulated on SPICE with the results shown in Figure 7.

Distributed-elements

Apart from the job of carrying signal energy from one part of the circuit to another, transmission lines or distributed elements may be used as filtering elements just like inductors and capacitors. The equation for the input impedance of a transmission line has been developed in Reference 4 and is shown below.

$$Z_{IN} = Z_0 \frac{Z_R \cos\beta l + j Z_0 \sin\beta l}{Z_0 \cos\beta l + j Z_R \sin\beta l} \tag{20}$$

where:

$Z_0 \equiv$ XMS line characteristic impedance
$Z_R \equiv$ load resistance
$\beta \equiv$ wave number $= \frac{2\pi}{\lambda}$

(expressed as the phase change of a wave traveling down a transmission line)

$l \equiv$ length of transmission line

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For a short-circuited transmission line, the following holds true:

$$Z_R = 0 \tag{21}$$

$$\therefore Z_{IN} = Z_0 \frac{jZ_0 \sin\beta l}{Z_0 \cos\beta l} = jZ_0 \tan\beta l \tag{22}$$

$$= jZ_0 \tan\frac{2\pi l}{\lambda}$$

From this equation, it is observed that the impedance of the transmission line is purely reactive and that it is a function of its length. The shorted transmission line exhibits a positive reactance (or inductive) from $\beta l=0$ to $\beta l=\pi/2$. At the latter condition the following holds true:

$$\beta l = \frac{2\pi l}{\lambda} = \frac{\pi}{2} \tag{23}$$

$$\therefore l = \frac{\lambda}{4} \tag{24}$$

Since the length of the transmission line is a quarter-wavelength long, the transmission line is called a quarter-wavelength shorted stub. As βl increases from $\pi/2$ to π, the reactance is negative or capacitive.

The quarter-wavelength frequency can be defined as shown below.

$$l = \frac{\lambda}{4} = \frac{v_c}{4_f} \tag{25}$$

$$\therefore f = \frac{v_c}{4_l} = f_0 \tag{26}$$

Then:

$$Z_{IN} = jZ_0 \tan\frac{2\pi}{\lambda}\left(\frac{v_c}{4f_0}\right) \tag{27}$$

$$\therefore Z_{IN} = jZ_0 \tan\frac{\pi}{2}\left(\frac{f}{f_0}\right) \tag{28}$$

It can be seen that the stub will exhibit a periodic resonant behavior, alternating between inductive and capacitive reactances as the frequency f varies. Such a stub can be used as a filtering element just like a parallel LC tank circuit. A similar result occurs for the open-circuited transmission line.

$$Z_{IN} = Z_0 \frac{Z_R \cos\beta l}{jZ_R \sin\beta l} = -jZ_R \cot\beta l \tag{29}$$

For the quarter-wavelength transmission line:

$$Z_{IN} = -jZ_0 \cot\frac{\pi}{2}\left(\frac{f}{f_0}\right) \tag{30}$$

Again, this stub will exhibit a periodic resonant behavior, alternating between capacitive and inductive reactances. This stub can be used as a filtering element just like a series LC circuit.

Thus, it has been shown that for frequencies less than the quarter-wavelength frequency, the short-circuited stub acts inductive and the open-circuited stub acts capacitive. The reverse holds true for frequencies greater than the quarter-wavelength frequency. This is the relationship that exists between distributed-elements and lumped-elements.

Transmission-Line Filter Design

In the previous section, it has been shown that transmission-line stubs may be used as filtering elements just like inductors and capacitors. Open-circuited stubs are treated as capacitors and short-circuited stubs are treated as in-

Model	Frequency Range MHz	Gain dB Typ.	Gain dB Min.	Noise Figure dB Typ.	Noise Figure dB Max.	Power Output dBm Typ.	Power Output dBm Min.	I.P.3 dBm Typ.	D.C. Volts Nom.	D.C. mA Typ.
Guaranteed Specifications: 0 to 50°C										
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AC262	30-250	8.0	6.8	1.3	2.3	19.3	16.5	35.0	15	23
AC305	10-250	11.8	10.5	2.1	3.1	15.5	14.5	31.0	5	33
AC386	10-250	28.0	26.5	2.6	3.8	8.0	7.0	20.0	5	21
AC552	30-500	22.0	20.5	2.4	3.5	10.0	8.0	22.0	5	16
AC1035	10-1000	25.0	24.0	2.5	3.5	5.0	2.5	15.0	5	18
AC2056	10-2000	20.0	18.0	4.0	5.5	9.5	7.0	21.0	5	34
AC4083	1000-4000	18.5	17.0	3.3	4.5	12.5	11.5	23.5	8	27
AP388	10-250	14.0	12.5	5.0	6.0	23.0	21.5	37.0	15	65
AR2584	20-2500	18.5	17.5	3.0	4.0	15.0	13.5	26.0	8	41

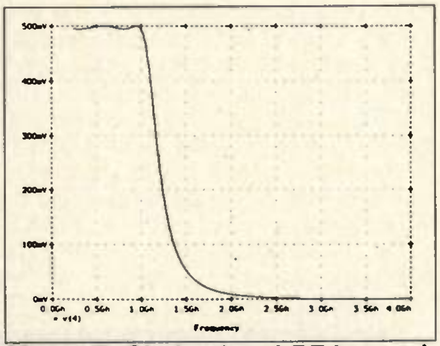

Figure 7. Chebyshev LPF lumped-element response.

ductors. This makes it possible to design filters using transmission-line stubs.

As with the lumped-element filters, first let's consider the transmission line lowpass filter. The design procedure is very similar compared to the lumped-element case. The filter order is found by using equation 4 with frequencies ω replaced with the variable shown below.

$$\omega \rightarrow \Omega = \tan\left(\frac{\pi}{2} * \frac{f}{f_0}\right) \tag{31}$$

where:
$f \equiv$ either cutoff or stop freq.
$f_0 \equiv$ quarter-wavelength freq.

Based upon this filter order, the elemental values are determined as discussed previously.

The elemental values are scaled for the proper impedance and cutoff frequency as shown below.

$$L_i \rightarrow \frac{Rg_i}{\Omega_c} \tag{32}$$

$$C_i \rightarrow \frac{g_i}{R\Omega_c} \tag{33}$$

Now, the lumped-elements are replaced by transmission lines. The series inductor is replaced by a short-circuited stub and the shunt capacitor is replaced by an open-circuited stub. The impedances for the transmission lines are found as shown below.

$L_i \rightarrow Z_0$ (shorted stub)

$C_i \rightarrow \frac{1}{Z_0}$ (open stub)

Following this general procedure for the lowpass design, the same transformations discussed previously are used for the design of highpass filters.

Since the transmission lines have a periodic nature, these filters exhibit repeating passbands and stopbands. It follows that true lowpass and highpass

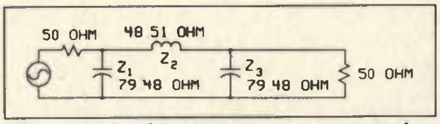

Figure 8. Chebyshev transmission line network.

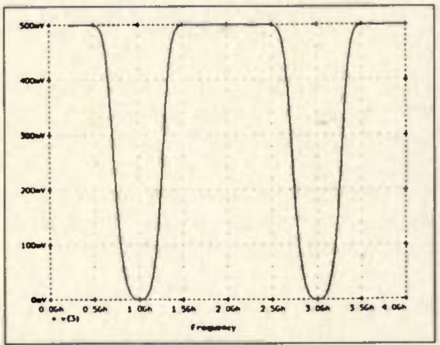

Figure 9. Chebyshev transmission line response.

filters can't be built using equal-length stubs.

Based upon the previous designs, the lowpass filter is really a bandstop filter and the highpass filter is really a bandpass filter.

As an example, assume a lowpass filter is needed with the following requirements:

1. r = 0.01 dB
2. f_0 = 1.0 GHz
3. f_c = 0.5 GHz
4. f_s = 0.8 GHz which extends to 1.1 GHz minimum
5. A = 10 dB
6. R = 50

These requirements result in a third-order filter. The filter is scaled for both impedance and frequency and is shown in Figure 8. The circuit is simulated on SPICE and the results are shown in Figure 9.

Conclusions

In this paper, an introduction to RF/microwave filter design has been given. The design of RF/microwave filters has been based upon the design of lumped-parameter filters. The lumped-elements can then be replaced with distributed-elements such as transmission lines. The same design techniques used for lumped-element filters can then be extended to distributed-element filters using transmission lines.

As was previously shown, transmission lines have a periodic nature. As a result, true lowpass and highpass filters can't be realized using constant line

length transmission lines. Instead, only bandstop and bandpass filters can be realized. *RF*

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About the Author

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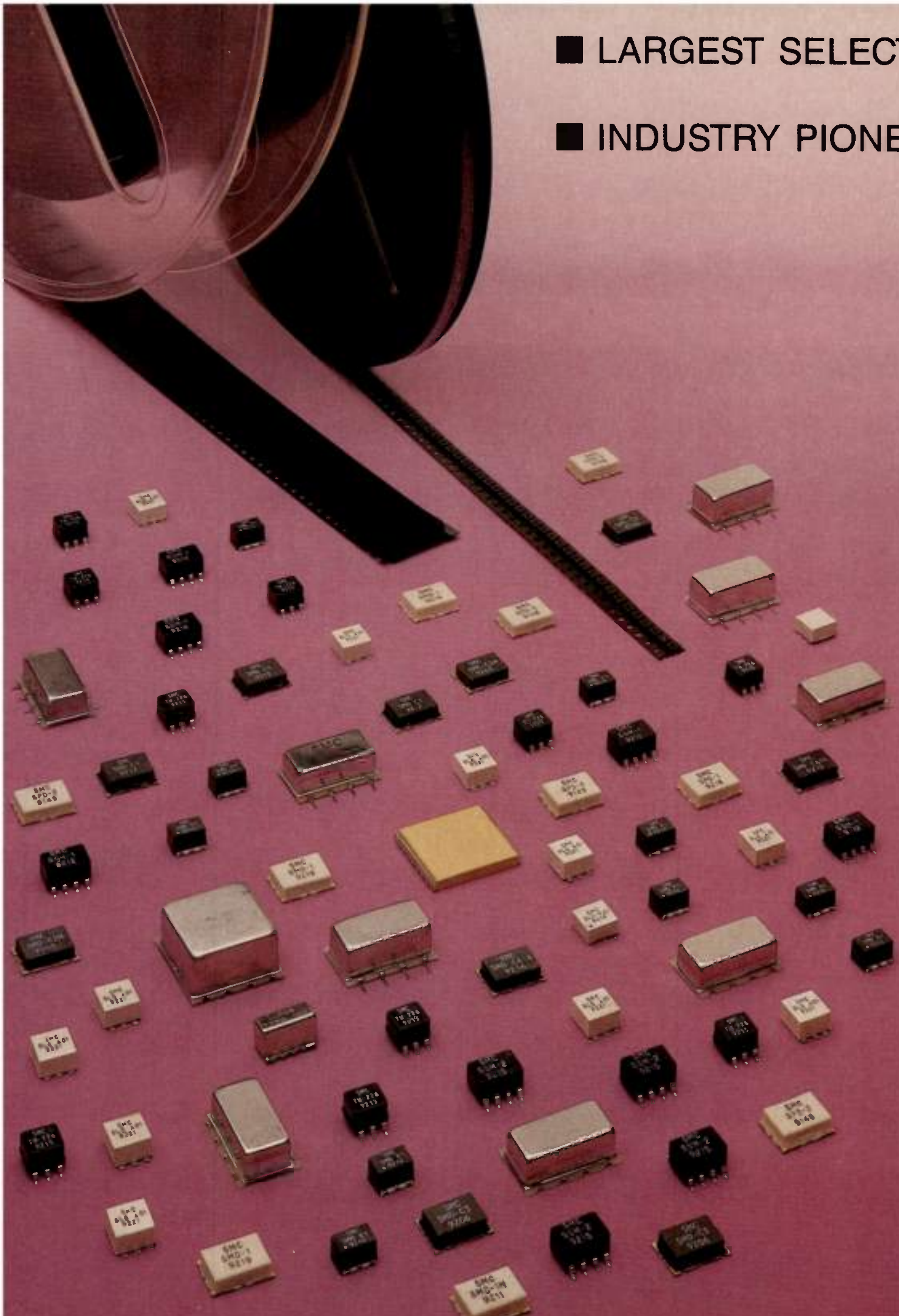

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The Mechanical Filter in HF Receiver Design

By William E. Sabin
Consulting Engineer

The high quality HF communications receiver for AM, SSB, CW and narrowband data modes presents many interesting design problems in terms of sensitivity, strong signal immunity, frequency control/readout and selectivity. A major goal is to achieve a high ratio of performance to cost. This article discusses the circuit design aspects of a new generation of Collins mechanical filters in the IF section as a way to help realize these goals. The present discussion focuses on the analog receiver signal path.

The example filters to be discussed here are the Collins Low Cost Series (526-8634/5/6-010), made by the Filter Products Division of Rockwell International, Costa Mesa, CA. These filters are centered at 455.0 kHz and come in 3 dB/60 dB bandwidths of 0.5/2.0, 2.5/5.2 and 5.5/11 kHz. They have a maximum insertion loss of 6 dB. They use the newly perfected Torsional Mode design (1) which leads to performance/cost and performance/size ratios which are unprecedented in the long history of the IF mechanical filter. Figure 1 conveys their impressive dimensions (1.25 x 0.5 x 0.25 inches). A wide range of other miniature Torsional Mode filters are also in the product line. The Collins Radio Company (now part of Rockwell International) started manufacturing mechanical filters in 1952.

Receiver Topology

A brief review of the analog receiver system architecture will help to put the mechanical filter into perspective as a receiver component.

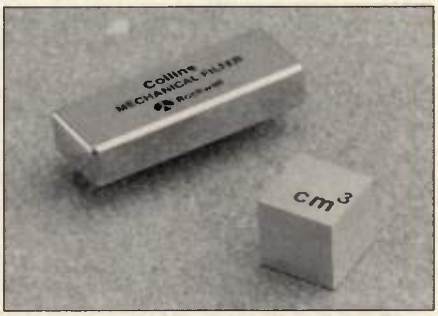

Figure 1. Collins mechanical filter with size comparison.

Figure 2a is a block diagram of an *archaic* HF general coverage receiver which uses a 455 kHz IF frequency. In order to avoid spurious responses which are from signals 910 kHz away (the image response) and others, notably the one only 227.5 kHz away (the fourth order "IF/2" response), multiple tuned circuits at signal frequency are needed. Even-order responses like the IF/2 may be somewhat reduced if *balanced* mixer circuits of some kind are used. But still, the gang-tuned and gang-switched preselector and local oscillator assembly is much too expensive and too inadequate for today's performance and cost requirements. Instead, the "upconversion" approach shown in Figure 2b is an improvement. The implication and the cost burden of upconversion is that the frequency stability requirements for the very high frequency local oscillators (LO) mandate a synthesized, low phase-noise first local oscillator which is now digitally switched and tuned, and an equally stable second mixer fixed, or vernier tunable, LO. The state-of-the-art in synthesizer design and cost has advanced so dramatically that upconversion is now the accepted approach, even for low cost receivers.

The tightly controlled and stable center frequency and passband of the mechanical filters make them especially desirable where synthesized, accurate local oscillators are used.

The same image and IF/2 responses mentioned above must now be rejected by the first IF filter. Its stopband requirements are therefore very stringent; 90 dB rejection minimum, is needed in today's world of really big HF signals, coming from large, high gain receiving antennas. An additional 100 MHz IF filter can be used to achieve the required rejection. More commonly, a more desirable IF filter is used by including a third, lower IF in the 10 MHz region. The spurious responses of the first mixer are controlled by a 1.6-30 MHz highpass/lowpass filter or possibly by PIN diode-switched half-octave filters. The first IF filter helps to minimize the spurious responses of the second mixer. Figure 2c is the method usually found in practice today in high quality, narrowband HF receivers.

A low cost alternative for the approach in Figure 2b, occasionally used, would be an image reducing second mixer. This can reliably add an additional 20 dB or perhaps 30 dB of image and IF/2

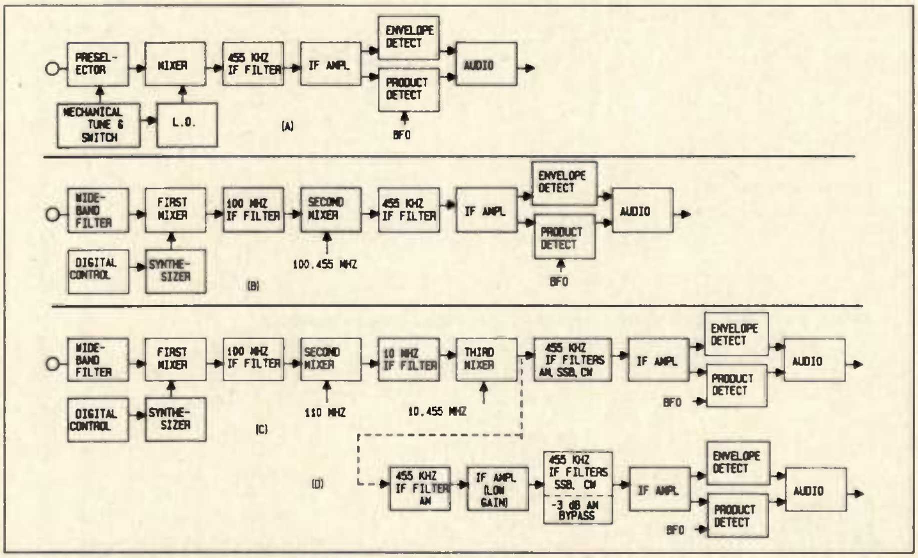

Figure 2. The evolution of the modern HF receiver.

rejection, which may not be enough, at the relatively low cost of some additional mixer and LO circuitry.

And of course the various LO and IF frequencies in the multiple conversion receiver must be optimal in terms of harmonic intermodulation (IM) products (harmonics of the LO's combining with harmonics of the RF/IF to produce spurs within the IF passbands (2,3,4)). For this reason, the first IF is preferably greater than three times the highest signal frequency. But if a sufficiently high level first mixer is employed, a ratio of greater than two to one may also be an acceptable approach (it does a better job of rejecting harmonic IM products). Also, internal "tweets" (various LO frequencies leaking into places where they don't belong) must also be minimized, therefore internal shielding and lead filtering are critical aspects of the design. An interesting example is when the 110 MHz LO, Figure 2c, leaks into the first mixer. A vulnerability to a 10 MHz antenna signal occurs (IF feedthrough). A very important consideration is the distribution of gain and noise figure along the signal path. The high frequency IF filters preceding the mechanical filters are usually substantially wider (but see later discussion), and since AGC is usually produced only by signals which pass *through* the mechanical filter, the early stages are vulnerable to strong signals which are inside the wider band but outside the mechanical filter band. Therefore the early stages must have low gain, low noise figure and high intermodulation performance. These conflicting goals lead to some difficult tradeoffs. In other words, the desire to put the "knothole" close to the antenna is not realized in the upconversion receiver which is intended for *narrowband* modes.

A consequence of low front end gain is that the gain *after* the mechanical filters must now be *increased,* and this can cause high levels of wideband noise to be generated which degrades receiver sensitivity (5). In a narrowband CW/FSK receiver, in particular, a second narrow filter downstream is often used to establish the required narrow noise bandwidth. This in turn can cause AGC loop design problems because of the time delays that the cascaded filters introduce. The second filter may be kept outside the AGC loop. Alternatively, an audio bandpass filter in conjunction with an image reducing product detector is an acceptable solution (5), except that wideband IF noise may affect the performance of the AGC circuitry.

A compromise approach is shown in Figure 2d. The wide AM filter, with its lower group delay, is always in place and an IF amplifier drives the switched, narrower filters at a higher signal level (say 20 dB or so). The AM filter provides at least some 6 kHz wide "roofing" protection for the circuitry preceding the other filters.

Another approach, often used, is to use the design in Figure 2c and let the 10 MHz filter be the roofing filter (a crystal filter) with a bandwidth of, say 10 kHz. Its shape factor is not critical since the mechanical filters do the "hard work". This is an excellent method, but involves a relatively low cost crystal filter (4 poles).

A further elegant escalation is to use a

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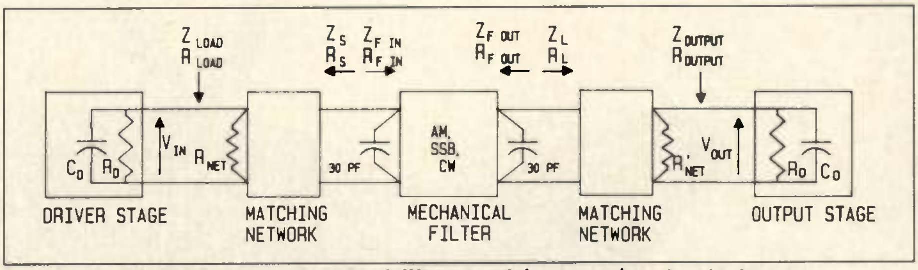

Figure 3. Matching mechanical filter to driver and output stages.

more elaborate 10 MHz crystal filter and then manipulate the various local oscillators in such a way that the intersection of the crystal filter passband and the mechanical filter passband produces a variable bandwidth and a variable center frequency. As these are changed the net tune-frequency remains constant. Pairs of AM, SSB or CW filters with matched frequency responses are most effective. This method creates a valuable "operator's aid" feature.

Impedance Matching the Mechanical Filter

The termination requirements for the mechanical filters are shown in Figure 3. The source Z_S and load Z_L should be 2000 ohms, within 5 percent and as purely resistive as possible across the 6 dB bandwidth, and should not become highly reactive in the transition bands (see later discussion). The matching networks establish the correct value of load impedance, R_{LOAD}, at the output terminal of the driver stage (to establish the desired gain, stability and maximum output excursion for that stage) and the correct value of source impedance, R_{OUTPUT}, for the output stage (to establish the desired input excursion, noise figure and stability for that stage). They also determine the ratio V_{OUT}/V_{IN} which is needed for correct gain distribution. The physical resistors, R_{NET}, are usually required in order to help meet *all* of the above requirements simultaneously (see text below for a further word of clarification of this point). Some signal power is lost in them, but at this point in the signal path some resistive loss is tolerable. Some of this resistive loading resides within the QX product of the matching network inductors. Also, in the interest of accurate filter termination, the resistive loading supplements the loading by the active devices, which often show considerable variation over temperature and production tolerances.

Variations in the filter input and output impedances $Z_{F\ IN}$ and $Z_{F\ OUT}$ also affect Z_{LOAD} and Z_{OUTPUT} (Figure 3). Figures 4 and 5 show typical AM filter impedance

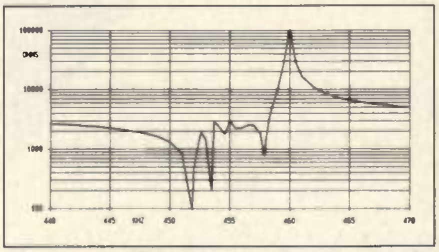

Figure 4. Filter input-Z magnitude vs. frequency.

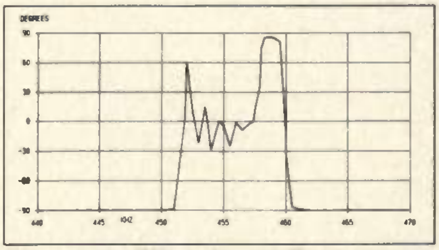

Figure 5. Filter input-Z angle vs. frequency.

magnitude and phase plots. In particular, the out-of-band overloading tendency at the output terminal of the driving stage could be degraded by a large increase in Z_{LOAD} outside the filter passband. The matching networks should be designed with this potential problem in mind. Equally important is that the filter will work properly if it "sees" the correct Z_S and Z_L at each end.

The L Network

The method shown in Figure 6 is a simple network which provides the necessary impedance transformation from some high value down to 2000 ohms without "coloring" the frequency response of the mechanical filter. The idea is to let the mechanical filter provide the selectivity. This method provides a very easy way to achieve resonance and impedance matching simultaneously. The behavior of this network can best be visualized with the Smith Chart diagram of Figure 6, in which the center point is normalized at 2000 ohms. Starting at some desired high value of R at 455 kHz, the shunt coil has an inductive sus-

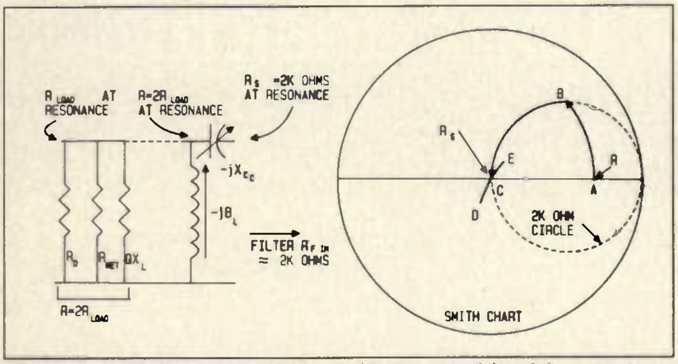

Figure 6. Matching network (L-network) without stray capacitance.

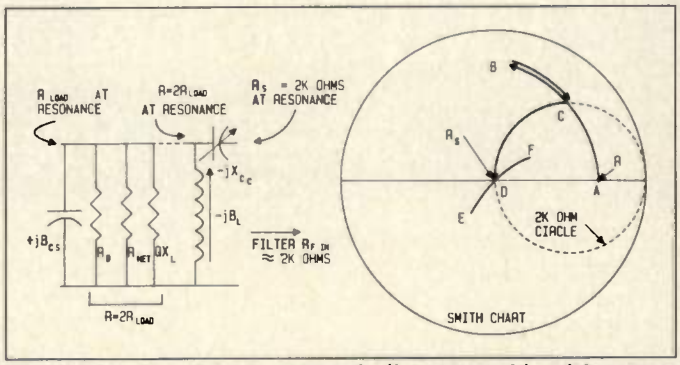

Figure 7. Matching network (L-network) with stray capacitance.

ceptance which traverses the path AB. It arrives at the 2000 ohm constant resistance circle at point B. A series capacitive reactance then traverses the path BC, where C is 2000 ohms. As the frequency varies from 450 kHz to 460 kHz the value of Z_S (or Z_L) varies along the path DCE.

A complication occurs because of the stray capacitance, C_S, shown in Figure 7. This consists of the capacitances of the coil, the wiring and the active device involved. The Smith Chart of Figure 7 shows that now the inductive susceptance must be significantly greater, path AB. The shunt capacitive susceptance of C_S lies along the return path BC, as shown, and point C is on the 2000 ohm circle, just as before. In other words, the "effective" inductive susceptance is the path AC. Series capacitive reactance then traverses CD to the 2000 ohm point. But when we look at the spread (EDF) of Z_S (or Z_L) from 450 kHz to 460 kHz we see that it is significantly greater than before. The stray C_S sharpens the selectivity of the network.

Too much LC selectivity, especially if it is repeated at input and ouput, produces excessive rolloff at the passband edges. It will also cause impedance mismatch, causing excessive ripples at the passband edges. This is based on my experience with both actual circuits and accurate SPICE simulations. The "passband" involved might be that of a single AM filter or a set of USB/LSB, or even four channel multiplex, SSB filters which are switch selected as discussed in a later section. The L network is an excellent answer to this problem.

For the best results over the 450 to 460 kHz band, especially for the wider filters, the coil and the sum of its self-capacitance and other stray C_S should be resonant as far as possible above 460 kHz. Coil catalogs usually give the value of inductance at some low frequency, where capacitive effects are small, and also the self-resonant frequency from which C_{COIL} and L_{TRUE} can then be estimated.

The coil and capacitor C_C must be high quality components which are stable over temperature and time. C_C should be an NPO trimmer in parallel with a stable fixed capacitor. Other parameters such as stray C_S and active device I/O impedances should be as constant as reasonably possible. These stabilities become more necessary as R in Figure 6 or 7 becomes larger because then, for some small fractional change of L_{EFF} or C_C, the mismatch and the passband response degrade more rapidly.

Equations 1 and 2 give values of L_{EFF} and C_C (coupling capacitor) for a given value of R, where R (as shown in Figures 6 and 7) is equal to twice the desired value of R_{LOAD} (or R_{OUTPUT}) indicated in Figure 3. The variation in Z_S (or Z_L) over frequency can then best be determined either from a Smith Chart program (6) or by SPICE simulation. A further word of explanation: R is the total shunt resistance across the high impedance terminals, *not* including the contribution of the mechanical filter. If it is twice R_{LOAD} or R_{OUTPUT} then the filter is properly terminated and, simultaneously, the load impedance of the driver or the source impedance of the output stage, in the passband, is the desired value. In other words, the filter and its matching network provide the other R which, in parallel with the R mentioned above, gives the desired R_{LOAD} (or R_{OUTPUT}) within the filter passband.

$$L_{EFF} = \frac{1.564\times10^{-5}\times R}{\sqrt{R-2000}} \qquad (1)$$

$$C_c = \frac{1.224\times10^{-13}}{L_{EFF}} + \frac{L_{EFF}}{R^2} \qquad (2)$$

$$L_{TRUE} = L_{EFF}\left[\frac{\sqrt{1+4KL_{EFF}^2}-1}{2KL_{EFF}^2}\right]; \qquad (3)$$

$$K = 8.173\times10^{12} C_S^2$$

where in equation 3, L_{TRUE} is the catalog low frequency value of the inductor.

Figure 8 shows the L network as it is used to transform from a low impedance line R_{LINE} up to 2000 ohms. In this case the $Q'X_L'$ product of the coil is part of the mechanical filter's source or load resistance (R_S or R_L) and the L network transforms R_{LINE} up to a value R′ which, in parallel with $Q'X_L'$, is the desired 2000 ohms. A word of clarification: The *series* combination R_{LINE} and C'_C in Figure 6 is equivalent to a *parallel* combination which consists of the resistive component R′ and a capacitive component X_C' which is "tuned out" by L'_{EFF}. Equations 4, 5 and 6 can be solved recursively for L'_{EFF} and C'_C, starting with a value of R′ = 2000 ohms. This process converges rapidly because $Q'X_L'$ is usually much greater than 2000 ohms. Be sure, also, that the value of R_{LINE} does not vary.

$$C'_C = \frac{1}{2.859\times10^6\times\sqrt{R_{LINE}\left(R'-R_{LINE}\right)}} \qquad (4)$$

$$L'_{EFF} = \frac{1+8.173\times10^{12}\times C_C'^2\ R_{LINE}^2}{8.173\times10^{12}\times C'_C} \qquad (5)$$

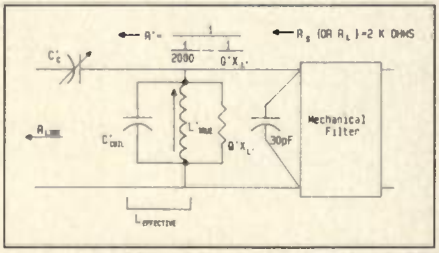

Figure 8. Matching mechanical filter to lower impedance R_{line}.

$$R' = \frac{1}{\dfrac{1}{2000} - \dfrac{1}{Q'X'_L}} \tag{6}$$

$$L'_{TRUE} = L'_{EFF}\left[\frac{\sqrt{1+4KL'^{2}_{EFF}}-1}{2KL'^{2}_{EFF}}\right]; \tag{7}$$

$$K = 8.173 \times 10^{12} C'^{2}_{COIL} \tag{8}$$

Other Design Factors

A receiver which has perhaps 100 dB or more of gain compression due to AGC has stringent stopband attenuation requirements. In order not to spoil the stopband rejection of strong adjacent channel signals, my experience has been that the inductors must be well shielded, both electrostatically and magnetically. Circuit layout, shielding and ground management are critical. Procedure: temporarily disconnect the mechanical filter input and output leads and fix all the problems. A later section on filter switching contains further discussion of leakage. I have measured better than 100 dB of ultimate attenuation for these mechanical filters, therefore the stray paths around the filters must be at least that good. I have also noticed that small amounts of stray coupling can produce some anomalous and undesirable frequency response distortions both inside and outside the passband. These leakage problems are a lot more manageable at the 455 kHz frequency than at the much higher IF frequencies that are often used.

The inductor core material should be correct for 455 kHz. For high impedance levels, ferrite with a μ_r of 125, using type 61, Q1 or 4C4 material provides small coils. For lower impedance levels, powdered iron assemblies with a μ_r of 20 (Carbonyl C) or 35 (Carbonyl HP) are preferred. Since the coil is heavily loaded (low operating Q) the coil Q need not be greater than 50 or so as long as its QX product is included in the impedance transformer calculations.

When designing matching circuits, it is necessary to be careful about intermodulation (IM) in the core material (the mechanical filters are excellent (1) in this respect). It is natural to want to use inductors whose dimensions are appropriate for these small filters, but troubles can easily occur. During the receiver design process, determine the maximum two-tone signal level that the mechanical filters and the inductors are expected to encounter and for which inband IM performance is guaranteed. Using a two-tone signal of the appropriate level, check to see that the IM generated by the cores is at least 20 dB better than the requirement. Changes in core material (powdered iron is preferred) and size and perhaps a reduction in gain ahead of the filters may be needed.

Another good reason for using the L network is that it makes the inductance value large and therefore the volts per turn small, compared to other approaches which use smaller values of inductance, thereby improving IM performance. The inductor may, however, be slightly larger physically. C_{COIL} also tends to increase (try to minimize).

A special case is when the filter input network is at the output of the last mixer. This core is vulnerable to wideband IM distortion which can ruin the front end intermodulation performance of the receiver. My experience with solid state receiver circuitry is that small (but not too small), fully shielded powdered iron core coils at this location cause very little trouble if the front-end amplification is minimized.

Quite often we desire to drive the mechanical filter from a diode double balanced mixer which "likes to see" a pure resistance output load (50 ohms) over a wide frequency range, in order to maximize the mixer's dynamic range. One excellent approach is to insert a low gain, low noise buffer amplifier whose resistive input impedance is determined by lossless feedback (4) and which is invariant with respect to the amplifier's output load variations (i.e., unilateralized). Some kind of diplexing (LP/HP or BP/BR) between mixer and amplifier, to improve wideband resistive loading of the mixer, is highly recommended. A simple (read "crude") compromise is a 3 dB pad. Also, mixers which are called T.I.M. (Termination Insensitive Mixers) are commercially available and should be considered. Quite often, some reduction in mixer performance can be traded for cost reductions. Maximum intermodulation performance is not always *the* prime consideration.

Balanced active FET mixers, as used in medium performance receivers, tend to be more tolerant of output load impedance variations when driving a mechanical filter because a buffering action similar to that described above is inherent in their design.

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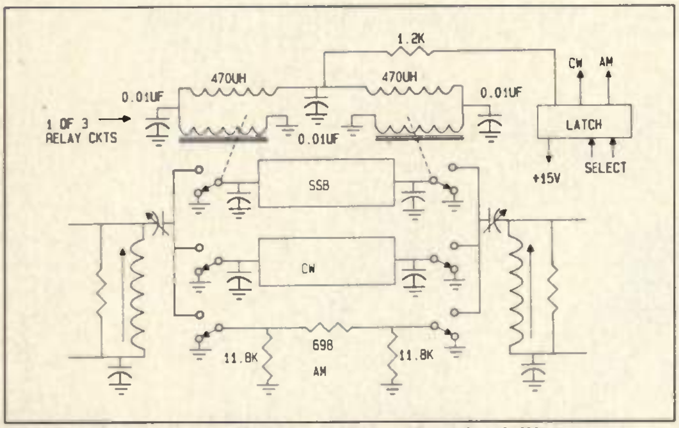

Figure 9. Relay switching of mechanical filters.

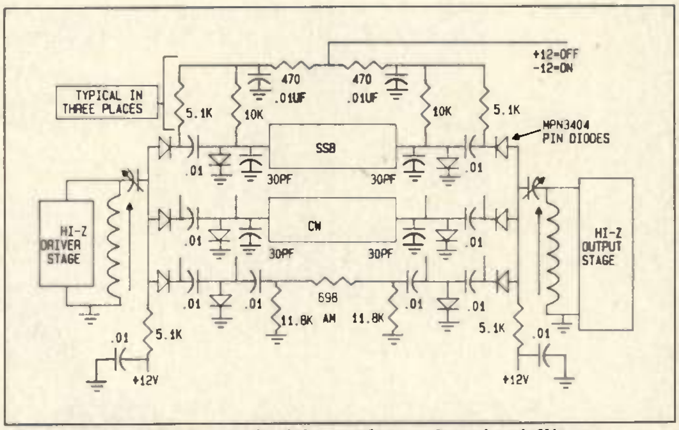

Figure 10. Diode switching of mechanical filters.

Finally, a very desirable approach to circuit design with a mechanical filter, as (1) points out, is to design its driver (or load) amplifier stage to provide 2000 ohms output (or input) impedance, thus minimizing the number of tuned LC networks. This may be easy to do, using recently perfected wideband opamps (7) from many vendors as IF amplifiers. Some other examples are the NE602 mixer and the NE605 mixer/FM discriminator devices which have input/output resistances in the 1000 to 2000 ohm range.

It is necessary, though, to restrict somehow the noise bandwidth of the later amplifier stages, as mentioned before, so some tuned circuits will be needed. The IF properties of any IC, such as noise figure and two-tone intermodulation distortion, at maximum expected signal level and especially if used before the filter, should be checked out. Check also for crossover distortion, which can degrade the IM ratio at lower signal levels. If a gain controllable amplifier is used, check the IM over the expected range of gain control. The gain control function should have, ideally, a constant dB per volt slope characteristic so that the AGC loop will be stable.

Switching Circuitry

Two methods of switching mechanical filters will be considered: relays and diodes. The use of rotary gang switches has been pretty well obsoleted in recent times because of size restraints, packaging problems, the use of automated assembly and the desire to have software control of the radio. Relay switching is an excellent and simple method, but involves electromechanical hardware which may be objectionable in some cases. Diode switching is more complicated and may be subject to intermodulation in the diodes but may have greater reliability in difficult environments. It is also an excellent method.

Figure 9 shows a relay switching method. In AM mode (see Figure 2d) a 3 dB, 2000 ohm pad is switched in. The relays are ultra miniature SPDT types, Potter & Brumfield type T81 or similar (for example, Radio Shack 275-241), with *separate* relays at input and output. The filter I/O terminals are grounded when the filter is "deselected", and better than 100 dB of isolation is obtainable if the layout is carefully done. One cause of serious response degradation, especially in the stopband, is due to capacitance from the center arm to the relay coil forming a sneak path *around* a "selected" filter. In one case, this capacitance was 6 pF. A net value of just 0.1 pF from input to output causes very serious problems. The decoupling circuitry for the relay coils helps to prevent problems of this kind. The relay method is

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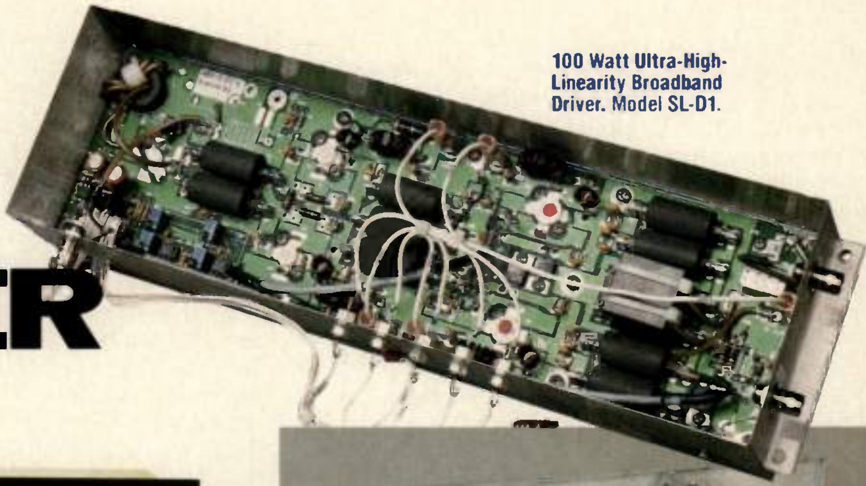

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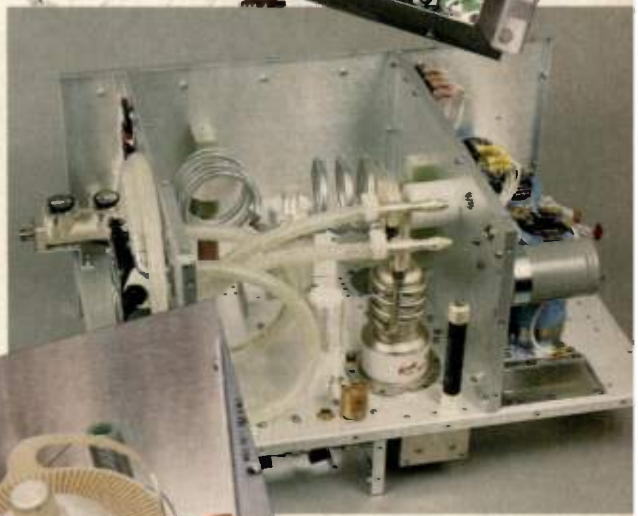

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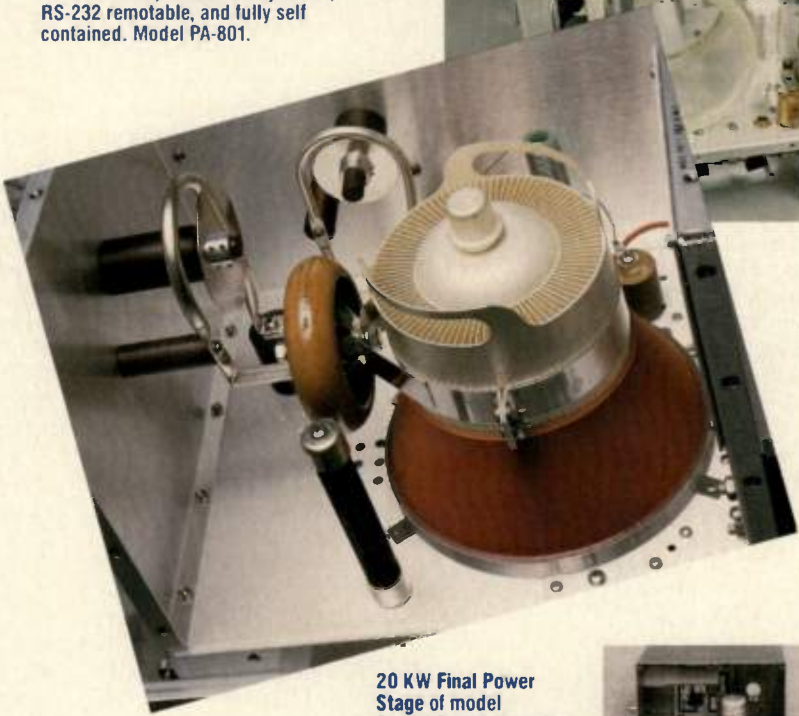

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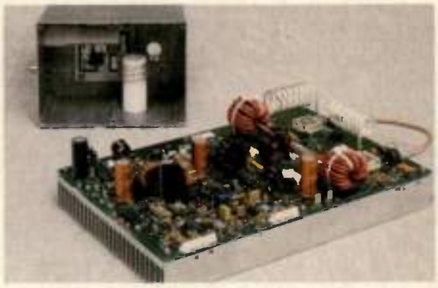

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very straightforward and low loss; the relays are cheap (about $3 each).

The diode switching method is shown in Figure 10. The diodes are Motorola MPN3404 PIN bandswitching types with 1.5 pF reverse biased (12V) and 1.5 ohms or so forward biased (2.5mA). Similar types of diodes are available in surface mount packages. Note the absence of RF chokes in the signal path which can spoil the 100 dB isolation by spurious inductive coupling effects.

In this circuit we take a slightly different approach to the L network in order to minimize the signal loss due to the biasing resistors. This loss could become quite large (in one particular case I had about 4.8 dB per switch) if we use the approach in Figures 6 and 7. In Figure 10 the combined value of the diode biasing resistors and the QX product of the inductor (after it has been transformed down) is the 2000 ohm pure resistance loading which the filter demands. In order to make this work efficiently the driver stage should be a current source (ideally) and the output stage should have a very high input impedance, both of which are easy to get at 455 kHz. The values for the L networks are then chosen to get the desired load impedance for the driver stage and the source impedance for the output stage. The values given in Figure 10 can be considered "starters" and can be tweaked as required (especially the 10K resistors). SPICE simulations are very helpful if accurate Q values are used.

In my prototype setup, using the AM filter, two tone third order inband IM products (at 453.5/456.5 kHz) are better than 70 dB below each tone (454.5/455.5 kHz) at an input level of 0.0 dBm per tone. The total isolation for a pair of turned off switches is better than 100 dB, if the layout is properly done (100 dB is a lot of dBs in such a very small volume but, as mentioned before, it is very desirable).

The commercially available GaAs SPDT switches, designed for 50 ohm circuits, are less desirable because they would require a pair of tuned matching transformers for each mechanical filter.

A possible source of signal leak-through concerns the grounding of the input/output pins of a deselected (switched-out) *wideband* filter or an attenuator. If the points at which these pins are grounded are at a different signal potential than the case of the filter a small signal can be coupled into the filter and can induce a wideband output which may be down only 80 dB or so. This can degrade the ultimate attenuation of a switched-in *narrowband* filter. I have seen this kind of problem.

There is another leakage problem which I have found troublesome. Because of the large amount of amplification at the 455 kHz IF frequency, especially at low signal levels when the AGC is not operating, or just barely operating, it is possible for the high level IF output to find its way back into the input of either the mechanical filter or the 455 kHz amplifiers preceding the filter. The symptom is that for low input signal levels the ripple in the passband increases, due to the feedback signal and the input signal adding or subtracting at various frequencies in the passband. As the input signal level increases it overrides the feedback leakage, which tends to remain constant, and this rippling effect is reduced somewhat, but not adequately. The lead filtering methods shown in Figures 9 and 10 help to eliminate this effect but other sneak paths, especially on pc boards, can cause the problem. Shielding, lead filtering and circuit layout improvements are usually indicated.

IF Circuit Alignment

The tuneup of the mechanical filter networks and all the other IF circuitry involves simultaneous adjustments for resonance and correct passband response (low ripple and flat response). The use of a sweep generator and logarithmic amplitude display, for example a spectrum analyzer-tracking generator pair or scalar network analyzer with 455 kHz capability, should be considered indispensable test equipment.

The Signetics NE604 log video detector chip ($5) makes an excellent lab-built 70 dB input device (total parts cost $50) for a low frequency ($300) oscilloscope which can be used in conjunction with a low cost ($150) function generator with sweep capability to produce a very low cost alignment tool. I have built and used an instrument using this approach; it works quite well and does not tie up a very expensive ($10000 minimum) piece of test gear at a test station. A high impedance input probe can be attached at various points for signal tracing and troubleshooting operations.

Conclusions

The small size, low cost and high quality performance of the torsional mode mechanical filters contribute to the design of a superior HF receiver in a small package. Layout of the filter circuitry in a small volume is a critical item. The filters must be accurately terminated and impedance matched to the driver and output stages. The receiver block diagram, as discussed briefly in connection with Figure 2, is some acceptable compromise between cost and performance, with low intermodulation and high narrowband sensitivity as conflicting but resolvable goals.

The numerous contributions and critical review of this article by Bob Johnson (8), Bill Domino and Peter Ysais, at Rockwell Filter Products are greatly appreciated. The encouragement of Lee Cornet is also appreciated. Their many years of experience with mechanical filter design, usage and marketing are prominent throughout this article. **RF**

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About the Author

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

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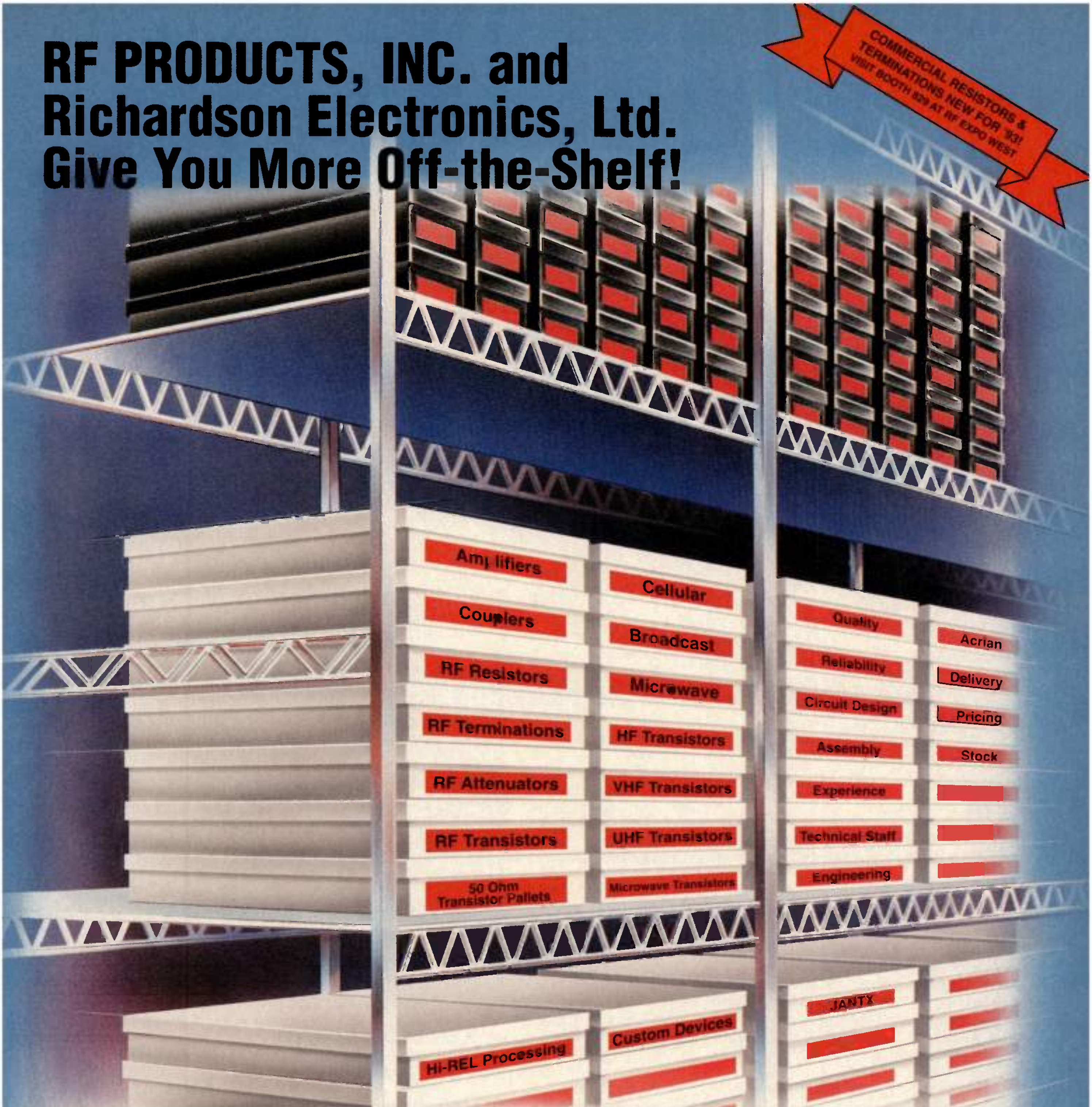

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New User Interface Makes RF CAD *EASi*

By Raymond S. Pengelly
Compact Software

Compact Software's Super-Compact and Microwave Harmonica linear and nonlinear circuit simulators have been on the market for many years — 20 years in the case of Super-Compact! For much of this time these powerful CAD tools were used by engineers sitting at monochrome or gray-scale terminals linked to mainframe computers. Graphics capabilities, the use of mice and user-friendly menus were limited by the data rate of the links between terminals and CPU. More recently, users of Compact's PC products have enjoyed superior graphics and user interfaces thanks to OS/2 and Windows 3.x.

Compact Software has now introduced totally revamped versions of its Super-Compact and Microwave Harmonica programs for workstations. Using X-Windows and OSF-Motif standards, Versions 3.0 of these simulators employ a very flexible interface called EASi.

The EASi (*E*nvironment for *A*nalog *Si*mulation) interface not only provides state-of-the art Motif-compliant menus, dialogue boxes, display and event managers, it also provides a powerful and extensive "command level" language allowing the software to interact with other third-party toolsets. The command menus, dialogue boxes, etc. have the "look and feel" of Microsoft Windows for PCs. The interface provides a modular program structure that keeps the simulator "engine", graphics and editor modules separated, so a number of different engines have access to the graphics and editor modules. Many resizable windows containing different information can be open concurrently.

Upon invoking the EASi interface the user is presented with an Editor window (Figure 1) which provides an immediate indication of the Compact simulators available on that workstation or node of a network. The desired simulator is chosen by clicking on the appropriate button (top left window of Figure 1). Reflecting designers' needs, the interface allows full editing and manipulation of circuit

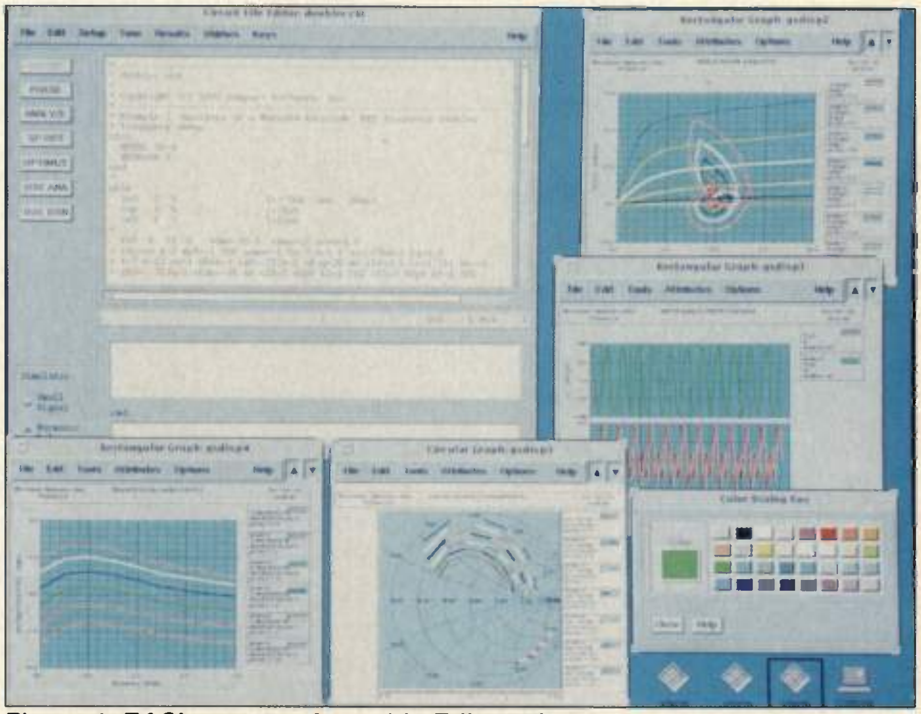

Figure 1. EASi operates from this Editor window.

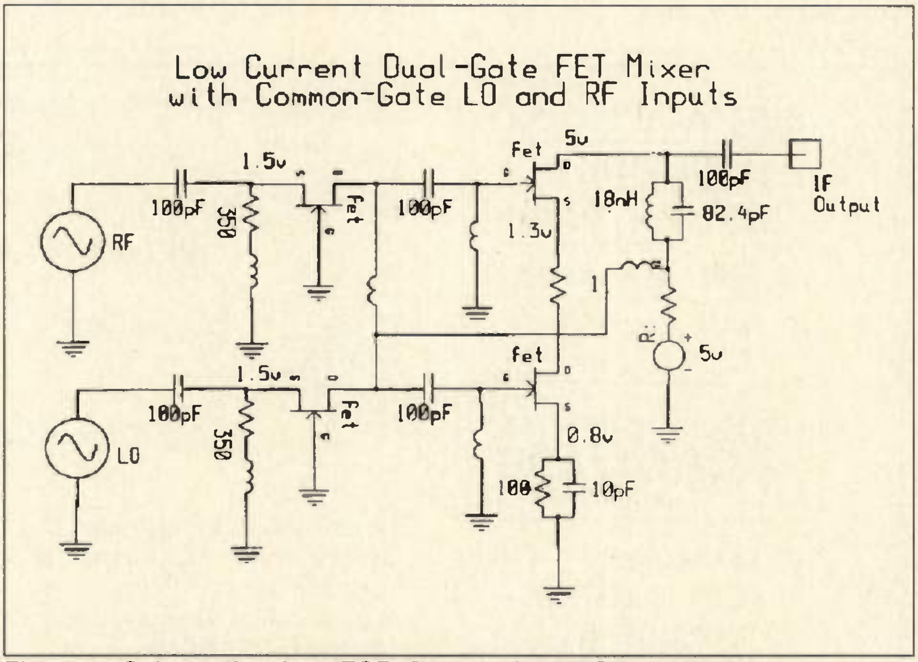

Figure 2. Schematic of a MESFET mixer for DECT application.

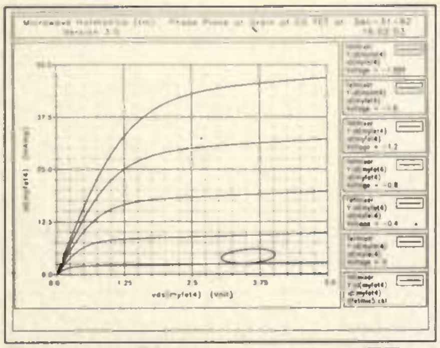

Figure 3. Phase plane of FET at LO port.

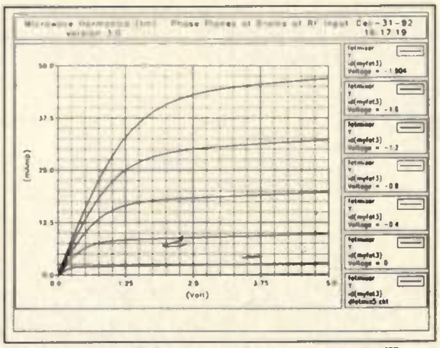

Figure 4. Phase plane of FET at RF port.

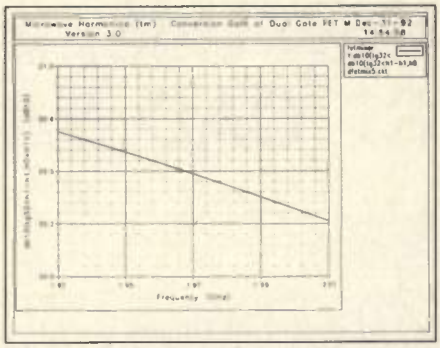

Figure 5. Conversion gain versus frequency.

netlists whether they be prepared manually or via Serenade schematic extraction.

Productive Circuit Simulations

The new Version 3.0 simulators and the EASi interface are best illustrated with examples. An application of current interest is new portable digital cellular telephone equipment at signal frequencies around 1800 MHz. Figure 2 shows the schematic of a dual-gate MESFET, low current consumption, IC mixer for the Digital European Cordless Telephone (DECT) application. The schematic is prepared using the X-Windows Serenade Schematic Editor, the netlist auto-extracted and then viewed in the EASi interface Edit window. The mixer consists of four FETs — two used in cascode configuration for the mixer itself and two used to provide active matching to the RF and LO ports of the mixer. All the FETs are self-biased. The gate-to-source voltages and drain-to-source voltages are shown in Figure 2. FET1 (the "top" FET of the mixer) has a V_{DS} of 3.7 volts with a V_{GS} of –1.3 volts. FET2 (the "bottom" FET of the mixer) has a V_{DS} of 0.5 volts and a V_{GS} of –0.8 volts. Both FETs are driven by common-gate connected FETs operating with 3.5 volts drain-to-source voltages and –1.5 volt gate-to-source voltages. FET2 will be driven with LO signal in this configuration. This FET actually operates well below the "knee" voltage. The common-gate FETs are biased to achieve approximately 20 mS transconductance and hence match to 50 ohms.

The RF signal frequency is in the new digital European cellular communications band of 1805 to 1880 MHz with an LO frequency variable between 1935 and 2010 MHz resulting in an IF frequency of 130 MHz. The mixer is matched through a simple parallel inductor/capacitor network at 130 MHz. The LO level available is –10 dBm. The required conversion gain of the mixer is 10 dB minimum.

Let us first investigate the characteristics of the 4 FETs. After harmonic balance simulation, various data displays can be viewed by using the Display Manager to define the parameters required.

The drain of the common-gate FET attached to the LO port has the phase plane shown in Figure 3. The current through this device is 3.5 mA with a voltage swing of approximately 1 volt into the gate of the mixer cascode FET. The average current in the mixer FET is 8 mA with an LO modulation of ±2 mA. The phase plane plot at the drain of the common-gate FET connected to the RF port is shown in Figure 4. Again the mean current is 3 mA. Figure 4 also shows the phase-plane plot at the drain of FET1 in the cascode.

Figure 5 plots the conversion gain of

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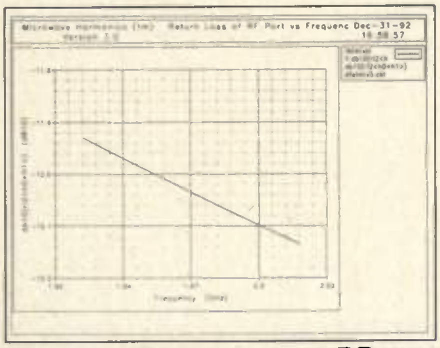

Figure 6. Return loss at RF port vs. frequency.

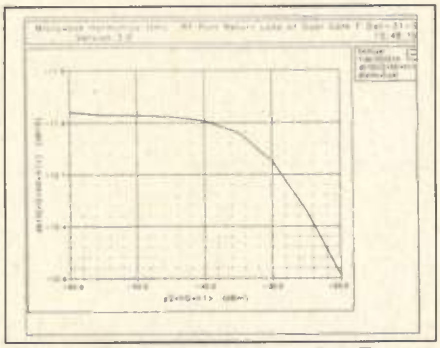

Figure 7. Return loss at RF port vs. power level.

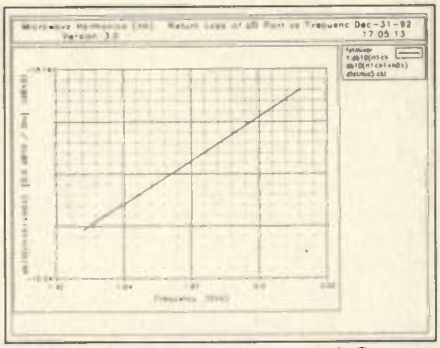

Figure 8. Return loss at LO port vs. frequency.

the mixer as a function of RF signal frequency where the LO frequency is also varied to maintain a constant IF of 130 MHz. The LO signal level was –10 dBm. The RF signal level for this test was –20 dBm. The conversion gain is approximately 21 dB over most of the frequency range. Changes in conversion gain over the band are due to the interaction between the common-gate output impedance and the input impedance of the dual-gate FET at the RF port. The total current taken by the mixer is 15 mA. The small-signal conversion gain of the mixer at the RF center frequency is 25.8 dB. 1 dB compression occurs at –30 dBm RF input power corresponding to a –4.2 dBm output 1 dB compression point.

Figures 6 and 7 show the return losses at the RF port of the mixer as a function of frequency and power. In the first case the RF input power level is –30 dBm. In the latter case we are sweeping the power level over a 40 dB dynamic range. The return loss is about 12 to 13 dB in both cases.

Figure 8 shows the return loss at the LO port of the mixer as a function of LO frequency at an RF input power level of –30 dBm. The return loss is constant at 15 dB and is acceptable.

Figures 9 and 10 show the spectra at the IF port corresponding to two different RF signal levels — one at –60 dBm and one at –30 dBm, the latter corresponding to the 1 dB compression point. The rejection of LO and RF signals at the IF port for the first case is greater than 50 dB, while in the second case it is 30 dB; the LO signal being the highest "breakthrough" signal at the IF port. The RF signal frequency for this test was 1850 MHz.

Finally, Figure 11 shows the RF signal and IF signal waveforms at the RF and IF ports of the mixer respectively at a signal level of –20 dBm corresponding to 5 dB compression.

The EASi interface allows the user to specify an almost limitless set of displays including user-defined equations. For example, Figure 12 shows the AM to PM conversion, or (IF output phase) – (RF input phase) as a function of RF input signal level over –60 dBm to –20 dBm. Note that the phase difference is constant up to input signal levels of –45 dBm, with a significant change as the RF signal level goes through and beyond the input 1 dB compression point.

Electromagnetic Simulation

In many circuit designs today where space is at a premium, conventional nodal circuit analysis using pre-determined component models cannot always accurately describe the structures that a design engineer wants to use, or they may give no indication of potential problems due to the close physical proximity of components. In order to overcome this situation Compact Software supplies design engineers with a very easy-to-use 3D electromagnetic simulator called Explorer. Explorer is particularly useful for multi-layer metal/dielectric structures. The dielectric, enclosure

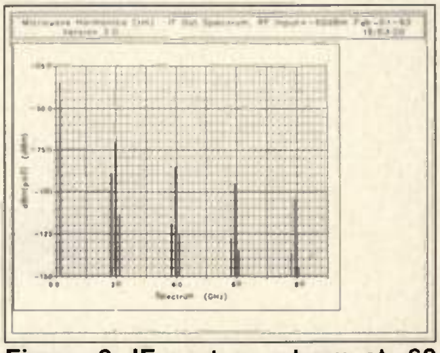

Figure 9. IF port spectrum at -60 dBm input.

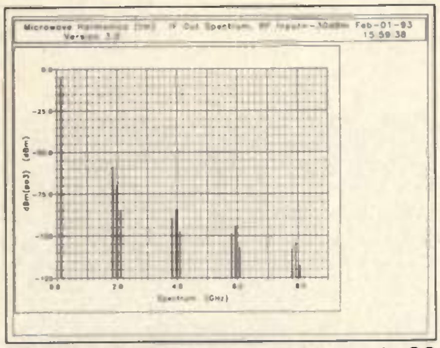

Figure 10. IF port spectrum at -30 dBm input.

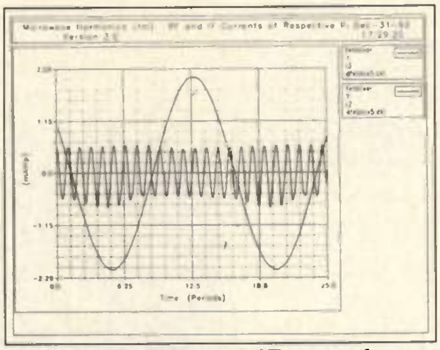

Figure 11. RF and IF waveforms (at respective ports).

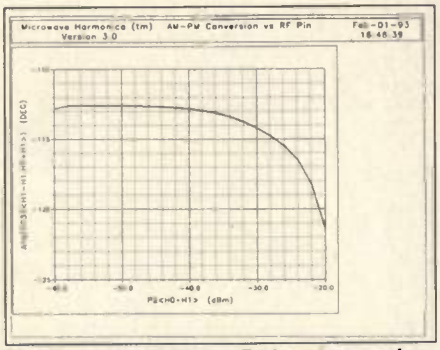

Figure 12. AM to PM conversion vs. power level.

and other parameters are input using Motif dialogue boxes. After simulation, data is displayed in a series of concurrent windows — displays include color s-parameter plots and current density distributions. S-parameter data is transferred in Super-Compact/Microwave Harmonica compatible format to files that can be used directly by the circuit simulators.

Circuit Layouts

Besides allowing designers to define circuits using the comprehensive Serenade Schematic Editor, Compact Software has also developed an extensive Layout Editor. This editor is an application overlay to the powerful AutoCAD® software from Autodesk. It provides design engineers with auto-generation of multi-layer hybrid and integrated circuit layouts from the Serenade schematics. Circuit elements can be dynamically replaced within the layout and modified layout attributes back-annotated to schematics and netlists. Layouts can be produced from hierarchical schematics maintaining hierarchical information. To aid in producing customized component libraries there are a number of example footprints for transistor packages, chip components, etc.

The design of a typical circuit starts by drawing a schematic of that circuit using the Serenade Schematic Editor. This editor has multiple pop-up menus, electrical rule-checking as well as user-configurable menus and extensive libraries of Compact Software simulator elements. All the circuit elements as well as the information required to control the simulator are contained in this schematic. When the schematic is completed an Electrical Rule Check (ERC) is performed to ensure, for example, that all nodes are connected. The netlist for the simulator is then automatically extracted for simulation by one of Compact's simulators, or by other third-party simulators such as P-Spice.

Once the basic design has been analyzed and optimized, basic layout-linked information is added to the netlist — for example, the orientation of bends, tees etc. which will make the auto-layout generation more efficient. Within the property pages of the schematic symbols company part codes can be defined. This information can be used to automatically select particular footprints such as transistor packages or bondpads and preferred wire-bonding information.

Take, for example, the VHF/UHF BJT amplifier schematic shown in Figure 13. The schematic representing the first layout iteration is prepared and the circuit re-analyzed to include all this information. Figure 14 shows the layout of the complete single stage amplifier. The layout is multiple layer with the attributes of different layers being defined in a user library. Typical layers for hybrid circuits include first and second level metallization, dielectric, thin film resistor, hole drilling, alignment and mask copyright. Figure 15 shows the differences between the simulated performance of the amplifier from schematic alone compared with the performance achieved after layout including all layout artifacts such as bends, tees etc. and features required because of layout and technology constraints.

Super-Spice Uses EASi Interface

Super-Spice uses the EASi Interface to provide extensive menu- driven control and output facilities. In microwave ICs distributed elements are used frequently as circuit components in such circuits as amplifiers, oscillators, phase shifters and switches. The majority of circuits in the microwave arena are simulated in the frequency domain. Howev-

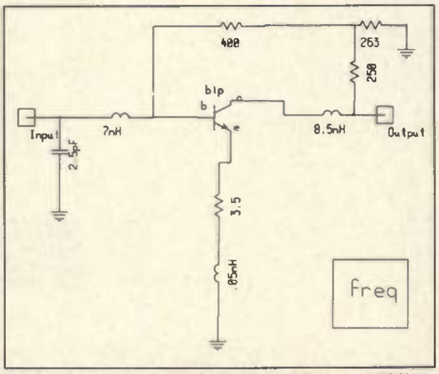

Figure 13. VHF/UHF BJT amplifier schematic.

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NEW

	NF	Ga	P1dB	Freq
NE326	0.5	11.5	10.75	12

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NEW

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NE332	0.8	10.5	12.5	12

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"How about a low cost plastic version? One I can order in volume for applications to 14 GHz."

Our high-yield Ion Implant process produces a consistant, uniform MESFET at a new low cost.

	NF	Ga	P1dB	Freq
NE760	0.8	15.5	14.5	4

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NOTE:
Noise Figures (NF) in dB
Gain (Ga) in dB
Power Out (P1dB) in dBM
Frequency in GHz

"How about a part with Heterojunction-style performance for my DBS, MMDS and commercial communications applications—at a MESFET price."

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NEW

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NE424	0.8	11.5	11.0	12

In low cost ceramic packages

"I'm building mixers at 900 MHz and need dual gate MESFETs by the thousands. Can you fix me up—at under 50¢ per part?"

We can—two different ways:

	NF	Ga	P1dB	Freq
NE251	0.8	12.0	12.0	0.9
NE253	0.8	12.0	15.0	0.9

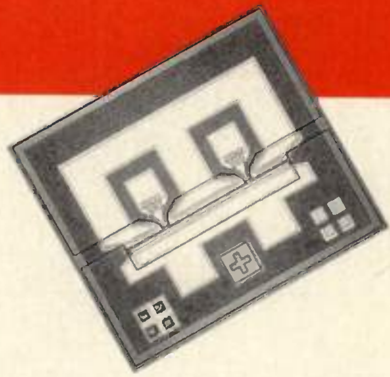
In low cost plastic packages

"I need a low noise FET for the LNA in my TVRO system. And, of course, I need it in a high volume, tape and reel package."

How about a 0.6 dB noise figure at 12 GHz? Introducing the new:

	NF	Ga	P1dB	Freq
NE324	0.6	11.5	11.0	12

In chip and low cost ceramic and hermetic packages



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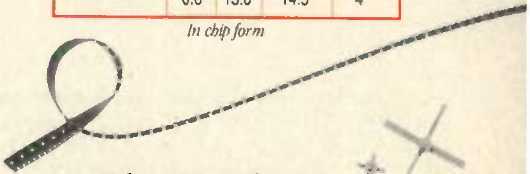
Here you go:

	NF	Ga	P1dB	Freq
NE710	0.6	13.0	14.5	4

In chip form



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INFO/CARD 52

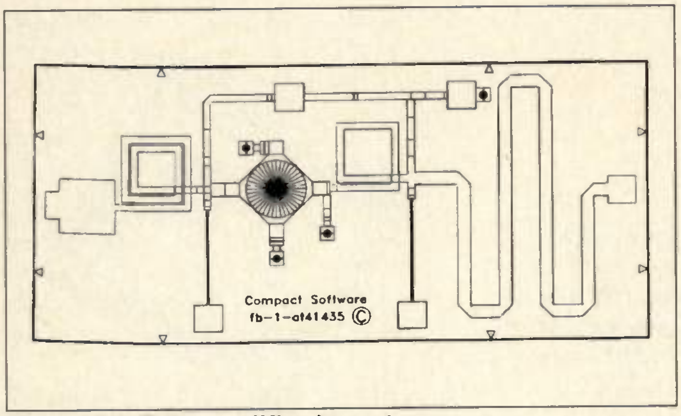


Figure 14. BJT amplifier layout.

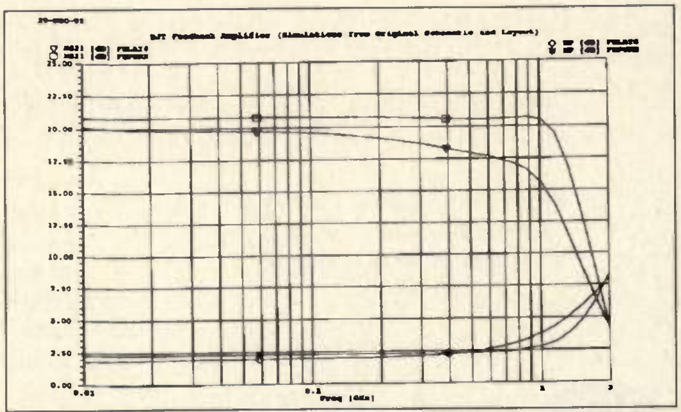
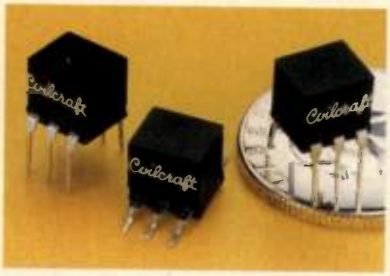


Figure 15. Schematic-only simulation compared to full layout simulation of the BJT amplifier.

er, in high speed digital circuits, distributed elements such as coupled signal lines cause unwanted parasitic effects producing crosstalk, reflections and loss.

Microwave designers have used a combination of frequency and time domain simulation techniques for a number of years. The so-called harmonic balance techniques are used in the nonlinear simulator Microwave Harmonica. However, the analysis of true transient effects in circuits is impractical using such a technique. Time domain simulators such as Spice are widely used, but they generally lack the capability of providing accurate models for discontinuities and dispersive, lossy distributed elements.

These design problems are solved using Super-Spice. This simulator contains a two-dimensional EM solver for distributed elements and discontinuities where quasi-TEM wave propagation is used, leading to fast computation times. The program has been tested for accuracy for PCB applications with risetimes down to 100 picoseconds. Super-Spice automatically generates equivalent circuits for the interconnects and imple-



Coilcraft offers low-cost wideband RF transformers

Coilcraft offers a family of low-cost wideband RF transformers that cover the frequency range from .005 to 600 MHz, with impedance ratios from 1:1 to 4:1.

The transformers are offered in tapped or untapped configurations and are packaged in a low-profile DIP-style plastic case. All parts are available in either a through- hole or surface mount version.

Applications include impedance matching, voltage or current transformation, DC isolation, balanced/unbalanced mixing, matching, power splitting, coupling, and signal inversion.

For more information, contact **Coilcraft**, 1102 Silver Lake Road, Cary IL 60013. 708/639-6400.

INFO/CARD 134

Precision test fixtures for surface mount devices

Coilcraft offers a series of test fixtures that assure faster, more accurate measurements of surface mount components. They're especially well-suited for testing tight-tolerance, low value passive components at frequencies from 1 to 1000 MHz.

The fixtures accommodate body sizes 0805 to 1812 and have exceptionally low stray capacitance and residual inductance. Several versions are offered for use with network analyzers, spectrum analyzers, and RF impedance analyzers such as the Hewlett Packard 4191A. Prices range from $425 to $498 depending on configuration.

For more information, contact **Coilcraft**, 1102 Silver Lake Road, Cary IL 60013. 708/639-6400.

INFO/CARD 135

ments them directly into the matrix equation. This results in fast set-up times.

The Wirth technique (1) is used for the analysis of lossy multiple coupled lines in Super-Spice. This is a fast technique which, when coupled to the "method of lines" used for the calculation of the characteristic transmission parameters of finite metal thickness lines, results in unique and superior simulations.

Coupled line systems are modeled as lossless coupled systems with augmented RLC networks (1). Fast simulation is provided by implementing directly the model in a modified nodal analysis matrix (2). Super-Spice contains a library of discontinuities which are typically encountered in high-speed board layouts. These include steps, right-angle bends (with and without chamfers), tee junctions and cross junctions.

Summary

The EASi Interface has been developed to provide engineers and managers with an intuitive, efficient and comprehensive environment for RF and microwave design. EASi is designed so that it can be extended to accommodate future simulator developments — a good example of this being in the extension of its display capabilities to the outputs required from the new Super-Spice program. This article has not only described the EASi interface but has also given a number of circuit examples to illustrate the functionality of the tool. *RF*

Readers desiring more information can contact the author at the address below, or they can circle Info/Card #99.

References

1. K.-H. Wirth, "New Model for Time Domain Simulation of Lossy Coupled Lines" *Electronics Letters*, Vol. 26, No. 20, pp. 1723-1724, September 1990.
2. K.-H. Wirth and J. Siegl, "Zur Schaltkreissimulation mit verkoppelten Mehrleitersystemen im Zeitbereich", *Frequenz* Vol. 42, pp. 305-313, Nov/Dec 1988.

About the Author

Ray Pengelly is Vice President of Sales and Marketing at Compact Software, 483 McLean Blvd, Paterson, NJ 07504. He can be reached by telephone at (201) 881-1200.

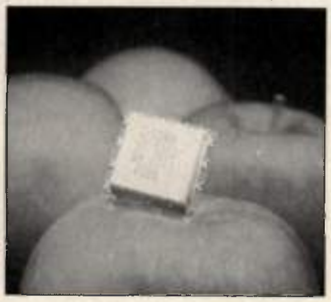
Spread Spectrum Generator

The LRS-200 generates pseudorandom sequences that are useful for developing and testing spread spectrum and conventional data communication systems. These sequences are generated by a configurable 32-stage shift register using linear feedback. The register length, feedback pattern, initial register contents, and length of the linear recursive sequence (LRS) can be set manually via a soft menu on the display, or automatically via an optional GPIB interface. Different modes of operation include BPSK, QS, Gold/JPL, Offset and a burst mode for radar applications, with a maximum bit rate of 25 MHz. External inputs include clock, data, sequence strobes, I and Q direct sequences, and a parallel interface for driving a frequency hopper. Options include longer generators (up to 192 stages), multiple generators (up to six R32 generators) and custom non-linear codes. Synchronous operation of two or more LRS-200s is possible. Semi-custom configurations are available. Both the chassis and keyboard are designed for rack mounting. A 16-stage shift register generator, the LSB-100, is a benchtop, non-GPIB programmable version of the LSB-200. Both the LSB-200 and LSB-100 are available off-the-shelf.

New Wave Instruments
INFO/CARD #250

S-Band VCO

Z-Communications announces the model C-810 VCO, with a frequency coverage of 2.6 to 3.0 GHz within a tuning voltage of 0 to 12 volts. This rugged VCO operates over the temperature range of –30 to +70 degrees C, making it suitable for many out-



door and indoor environments. The exceptional linearity, specified at greater than 90 percent over the 2.6 to 3.0 GHz range, is another feature of this VCO targeted for broadband applications requiring large frequency coverage and improved linearity. The C-810 requires a 15 V supply and draws less than 50 mA. It outputs 15 ±2 dBm into a 50 ohm load, which is strong enough for most low level mixers. Phase noise is typically –95 dBc/Hz 10 kHz from the carrier. The C-810 is available in the "Z-COMM industry standard", 0.91 × 0.91 × 0.29 inch surface mount package. Samples are available from stock, with production quantities delivered within six weeks.

Z-Communications, Inc.
INFO/CARD #249

UHF, "All-Band", Panel Antennas

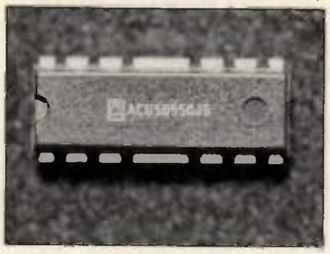
Micro-Communications introduces a new line of UHF "All-Band" multi-channel panel antennas, designed for HDTV. With the upcoming simulcast period quickly approaching, broadcasters are starting to plan their HDTV RF system strategies. A recent survey of stations found that 95 percent of all stations want their new antenna to be located on their present tower. In some installations, it will be possible to diplex the new HDTV signal and the existing NTSC signal

into one multi-channel panel antenna. VSWR is less than 1.10 from 470 to 800 MHz (channels 14-69). Power handling capability is 2.5 kW per panel. Omnidirectional; peanut; wide, narrow and very narrow cardiod gain patterns can be produced with different panel configurations. Channel combiners and coaxial cable are also avialable.

Micro-Communications, Inc.
INFO/CARD #248

GaAs CATV Converter/TV Tuner MMIC

A GaAs upconverter MMIC which provides all the front-end functions for double conversion cable television set-top converters and broadcast television tuners has been developed by



Anadigics. The ACU50550 incorporates a high degree of integration, replacing more than 30 discrete components in a typical converter.The MMIC has three main sections: a Gilbert cell comprising a low-noise amplifier with automatic gain control and a double balanced mixer, a phase splitter, and a voltage controlled oscillator. The chip features AGC of 0 to 8 dB, typical voltage tuning range for the VCO of 1.85 to 18 V, and typical phase noise of –83 dBc/Hz 10 kHz from the carrier. Typically, 84 channels in the 50 to 550 MHz range are available at the RF input. The ACU50550 is supplied in a plastic DIP and is priced at $14 in quantities of 100.

Anadigics, Inc.
INFO/CARD #247

SAW Filters for Cellular Phones

Hitachi America expanded its SAW (Surface Acoustic Wave) filter product line with a complete family of extremely compact 820 to 950 MHz devices precisely tailored for most major cellular phone markets worldwide. These high stability, adjustment-free, low insertion loss filters use 0.8 micron line and space geometry. They feature sharp, wide passbands and deep stopband attenuation for use in AMPS, ETACS and GSM/NMT equipment. The HWC Series input and output impedances are designed to fit pure resistive (50 ohm) circuits, eliminating external matching net-

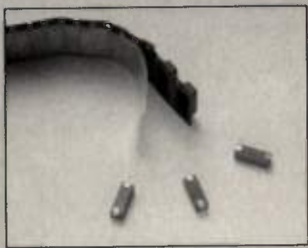
works. The HWC Series devices measure 3.5 × 3.5 × 1mm in their surface mount packages. In 10,000 piece quantities, the HWCA605 and HWCB605 are $3.65.

Hitachi America, Ltd.
INFO/CARD #246

Product Spotlight: Ferrites

SMT Ferrite Beads

Philips Components Discrete Products Division has introduced two surface mount ferrite beads for EMI/RFI suppression



applications. The beads are available in 0.335 and 0.160 inch size categories and are manufactured of 4S2 ferrite material. A flat piece of tinned copper wire is passed through and crimped on a rectangular ferrite bead.
Philips Components
INFO/CARD #245

Flat Cable Cores

Flat cable cores from Intermark slip over flat cables and offer 65 to 100 ohms impedance at 100 MHz. The cores are flat to conserve space and come is sizes accommodating 10 to 50 conductor cables.
Intermark (USA), Inc.
INFO/CARD #244

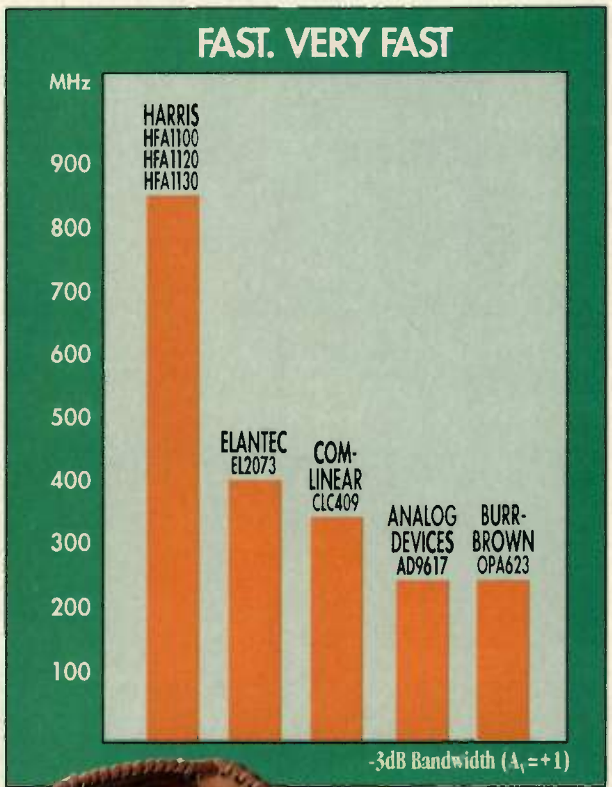
Beads On Interconnects

Meritec's Single Signal Interconnect (SSI™) assemblies are now available with ferrite beads. The assemblies meet the requirements of electrically sensitive applications using high speed CMOS, ECL or GaAs logic. Current rating is 1.0 amp/contact, and standard impedances from 50 to 100 ohms are available.
Meritec, Div. of Associated Enterprises
INFO/CARD #243

Absolutely The World's Fastest Op Amps.

Twice The Speed Of The Previous Champ.

The new ultra-high speed HFA1100 is absolutely the fastest monolithic silicon op amp you can get. With remarkable gain flatness of 0.14 dB to 100 MHz. Slew rate of 2300 V/µS. And bandwidth of 850 MHz, thanks to our absolutely amazing HFA-1 process for high-speed linear devices. You can tap into this speed through a whole family of convenient op amp configurations. Or as a semicustom design. Call us today. You'll be absolutely satisfied.



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- 2300 V/µs slew rate
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- Excellent video specifications
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- User programmable output voltage clamp (HFA1130)
- 60 mA output current
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HARRIS SEMICONDUCTOR

INFO/CARD 59

TEST EQUIPMENT

Direct Digital Synthesizer

The 2910 synthesizer is a bench instrument with an 11 digit liquid crystal display. It produces a very stable and low phase noise sine and TTL signals that are programmable from 0.000001 Hz to 12 MHz. An external reference input is standard. An optional RS-232 interface is available. Base price for the 2910 is $995; the optional RS-232 interface is $200.

Novatech Instruments
INFO/CARD #242

Low Price Signal Generator

The HP 8647A is a synthesized signal generator for RF communications testing from 250 kHz to 1000 MHz. The generator has specified amplitude accuracy of 1.5 dB (typically < ±1 dB) and modular repair capability which does not require recalibration after module replacement. Cost of the HP 8647A is $4965.

Hewlett-Packard Co.
INFO/CARD #421

Fading Simulator

The RFS-1 signature measurement set is now available with improved software which provides more powerful facilities for precise measurement of fading-related parameters. Software improvements mainly concern the documentation of results.

Wandel & Goltermann
INFO/CARD #240

Frequency Synthesizer

Model SI-102 provides high performance over the frequency range of 0.1 Hz to 16 MHz. The synthesized source features 5 1/2 digits of resolution from 0.1 Hz to 16 MHz, providing 0.00001 Hz frequency resolution from 0.1 Hz to 1.0 Hz and 100 Hz resolution from 1 to 16 MHz. Price of the SI-102 frequency synthesizer is $779, or a module only can be purchased for $554.

Syntest Corp.
INFO/CARD #239

Switch Matrix System

The model 7002, a ten-slot high density switch system from Keithley Instruments, is designed to support up to 400 two-pole

multiplexer channels or 400 matrix crosspoints. Due to high density design, the 7002 takes up just seven vertical inches in a standard rack. An optional light pen and LED channel status display allow quick programming. Price for the model 7002 is $3595.

Keithley Instruments
INFO/CARD #238

SUBSYSTEMS

Guyed Tower

The S-36 guyed tower is an all welded design meant for high strength, durability and minimal wind resistance. The tower fea-

tures a 36-inch face with 20 foot sections and high strength, solid round legs and bracing. Climbing ladders and 12 cable supports are built-in.

Andrew Corporation
INFO/CARD #237

Paging Antennas

Decibel Products has introduced the DB809K-YP and DB810K-YP, 928-932 MHz paging antennas with 9 or 10 dB gain. Each uses Decibel's Center Fed Transposed Tube™ design to provide a VSWR of 1.2:1 or

RF products

better. DB809K-YP is 3.5 meters long, and the DB810K-YP is 4 meters long. Each antenna is tested for power rating compliance and the absence of intermodulation generators.
Decibel Products
INFO/CARD #236

SIGNAL PROCESSING COMPONENTS

Harmonic Generator Switchable Filters

KW Microwave introduces new a new series of integrated generator switchable filters, covering 31.25 to 343.75 MHz. It consists of a 31.25 MHz harmonic generator and a six channel switched filter selecting harmonics between 31.25 and 343.75 MHz with a 62.5 MHz separation. Maximum switching speed is 200 ns.
KW Microwave Corp.
INFO/CARD #235

Five Port Hybrid-Divider

Sage Laboratories has developed a device which combines a power divider and a magic tee hybrid. The input at one port is divided into two equal amplitude outputs and a third which is 3 dB above the other two; the phase unbalance of these three output ports is less than 7 degrees. A second input is evenly split between two output ports, but with 180 degree phase difference.
Sage Laboratories, Inc.
INFO/CARD #234

Isolators and Circulators

Densitron Microwave has developed a range of isolators and circulators specifically for PCN systems. These devices have wide bandwidth (1710-1880 MHz) or standard bandwidth

RF products

(1805-1880 MHz). A standard 4-port offers 50 dB isolation with 0.4 dB insertion loss. 3-ports provide 25 dB isolation and 0.25 db insertion loss.
Densitron Microwave
INFO/CARD #233

Continuously Variable Delay Line

Dynetics has introduced a continuously variable delay line, model CVDL-40-10-5, that produces any delay from 150 ns to 5 us, with better than 1 ns precision. The line can be custom built with intermediate frequencies from 30 to 90 MHz and 25 percent fractional bandwidths. The CVDL achieves maximum insertion loss of 17 dB with less than 4 dB variation over the delay range.
Dynetics, Inc.
INFO/CARD #232

Cavity Filters

RLC Electronics has expanded its cavity filter line to include designs optimized for cellular, digital and other wireless communications applications. Coaxial, interdigital and combline types are available, with models in the 500 MHz to 40 GHz range.
RLC Electronics, Inc.
INFO/CARD #231

SIGNAL SOURCES

High Stability Ovenized Oscillator

A gold plated, stress-compensated-cut crystal, gold plated crystal enclosure and environmental compensation result in an oscillator with aging stability of down to 5×10^{-11}/day or better, and 1×10^{-8}/year stability. Temperature stability is 2×10^{-9} over –50 to +70 degrees C. Phase noise is 145 dBc/Hz 1 kHz from the carrier.
Quartzlock Instruments
INFO/CARD #230

964 MHz VCO

Model VCO-91010 voltage controlled oscillator, is ideally suited for cellular receiver applications, delivering a 964 ±12 MHz signal from a 1 to 4 V tuning voltage. The unit operates from +5V DC and provides 0 dBm minimum output power. The VCO is available in pin, surface mount and seven pin hermetically sealed DIP packages.
Micronetics
INFO/CARD #229

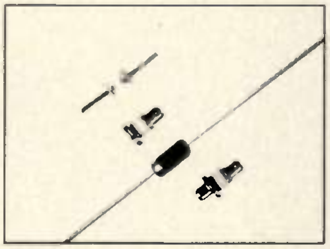
SEMI-CONDUCTORS

High Output Amplifier

Amplifonix announces the model TM9725, a linear high intercept amplifier. The power output at 1.0 dB compression is typically +26.5 dBm while maintaining a NF of 4.5 dB. Gain is typically 11.0 dB. The amplifier operates over –55 to +85 degrees C and comes in a TO-8 package, though various surface mount packages are available.
Amplifonix, Inc.
INFO/CARD #228

Noise Diodes

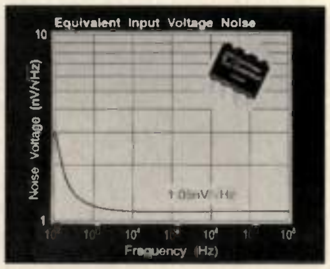
Noise Com has added two diodes to its NC 400 series of white Gaussian noise diodes. The two diodes (NC 405 and NC



406) cover 50 to 75 GHz and 75 to 110 GHz respectively. The diodes produce symmetrical Gaussian white noise with a crest factor much larger than 5:1. The diodes are available in a number of packaging styles.
Noise Com, Inc.
INFO/CARD #227

Low Noise, Wide-band Op Amp

The CLC425 combines a wide gain-bandwidth (1.7 GHz) with an ultra-low input noise (1.05nV/$\sqrt{Hz}$, 1.6 pA/$\sqrt{Hz}$) and excellent DC characteristics (100 uV vos, 2 uV/°C drift) to provide a very precise wide dynamic range



op amp offering closed loop gains of ±10 V/V. The price for the CLC425 is $4.75 in 1000 piece quantities.
Comlinear Corporation
INFO/CARD #226

VCG Amplifier

Burr-Brown's VCA610 is a stable, wideband, continuously variable voltage controlled gain amplifier. It has gain range of 80 dB (–40 dB to +40 dB), 30 MHz bandwidth, 2.2 nV/$\sqrt{Hz}$ voltage noise and 300 dB/us gain control slew rate. The VCA610 is available in 8-pin DIP and SOIC packages and is priced from $8.95 in 100s.
Burr-Brown Corp.
INFO/CARD #225

AMPLIFIERS

Market Specific Amplifiers

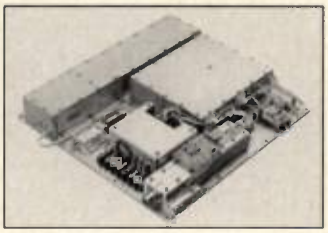
The Commercial Band-Select Amplifiers (CBA™) are a series of SMA connectorized amplifiers with 40 percent bandwidth, +20 dBm power output, 20 dB gain, 12 V operating voltage and a 3.25 x 1.25 x 0.625 inch aluminum case. Members of the series specifically support several satellite systems, HDTV broadcast, wireless data services and several other application areas.
Frequency Products, Inc.
INFO/CARD #224

2 kW, Wideband Amplifier

The IFI M810 is an air cooled amplifier with 65 dB minimum gain for the 0.01 to 220 MHz range. Bandpass flatness is ±2 dB. The M810 dimensions are 69 inches high, 27 inches wide and 30 inches deep. Standard EIMAC tubes are used.
Instruments for Industry
INFO/CARD #223

100 W Amplifier

Model BME5819-100 is a solid state, AB linear amplifier operating from 500 to 1000 MHz in one broadband module. Power output is 100 W with overall gain of 50



dB minimum. The amplifier has input/output overdrive as well as VSWR electronic protection.
Power Systems Technology, Inc.
INFO/CARD #222

CABLES AND CONNECTORS

Filtered Terminal Blocks

A new terminal block providing excellent EMI/RFI filtering of low voltage DC power lines and control lines has been introduced by Spectrum Control. EMI/RFI filtering characteristics include capacitance values of 2500 pF +100 percent, –0 percent, and DC resistance of 0.01 ohms max.
Spectrum Control, Inc.
INFO/CARD #221

High Power Connectors

Tru-Connector LC/LT series connectors handle up to 10,000 Vrms and provide 50 ohms constant impedance. The connectors are available as plugs, jacks, receptacles and adapters. They accommodate cables from 0.400

to 1.200 inch O.D. The LC/LT series is manufactured to MIL-C-3650 specifications.
Tru-Connector Corp.
INFO/CARD #220

RF tutorial

An Introduction to Noise Figure

By Jonathan Bird
Raytheon Company

This article discusses the origin of several types of noise and the use of Noise Figure to describe the noise contribution of circuit elements to a signal.

In all things on earth there exists a degree of randomness. In the world of electronics, the random nature of electric current has been known for over a hundred years. It is a fact that no electronic system is completely free of random noise. Small voltage fluctuations due to noise are always occuring in electronic circuits. One of the most frequently discussed forms of noise is known as thermal noise. Thermal noise is a random fluctuation in voltage caused by the random motion of charge carriers in any conducting medium at a temperature above absolute zero. This noise cannot exist at absolute zero because charge carriers cannot move at absolute zero. As the name implies, the amount of thermal noise in a circuit is proportional to the temperature of the conducting devices in the circuit. The best visualization of thermal noise is to imagine a simple resistor at a temperature above absolute zero. If one were to place a very sensitive oscilloscope probe across the resistor and examine the voltage in the time domain, he would find that there is a very tiny AC noise being generated by the resistor!

If one thinks of the resistor as a small signal generator putting out a tiny noise voltage, it is easy to model how much noise is created. The RMS noise voltage is proportional to the temperature of the resistor and how resistive it is. Larger resistances and higher temperatures generate more noise. The expression for the RMS thermal noise voltage of a resistor is:

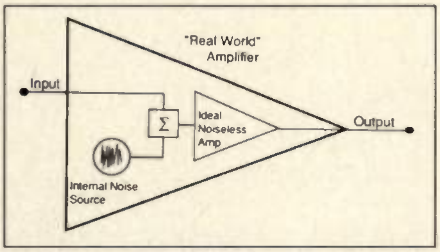


Figure 1. Block diagram of internal workings of a "real world" amplifier.

$$V^2 = 4kTRB \text{ [or } V = (4kTRB)^{1/2}] \qquad (1)$$

where:
k= Boltzman's Constant (1.38×10^{-23} Joules/Kelvin)
T= Temperature in degrees Kelvin (K=273+Celsius)
R= Resistance in ohms
B= Bandwidth in Hertz in which the noise is observed

How does bandwidth figure into this? This basically comes down to noise frequency. We know that the noise is AC, but what frequency is it? We don't know. The frequency is just as random as the voltage! Therefore, there is equal probability of noise at any frequency, occurring at any given time. If we examined the noise on a very sensitive and accurate spectrum analyzer, we would find that the noise has equal amplitude at all frequencies (this type of noise is referred to as white noise). The more frequencies allowed into the measurement (i.e. the larger the bandwidth of the measurement) the larger the measured noise voltage will be. This means that the RMS noise voltage measured across a resistor is also a function of the bandwidth in which the measurement is made.

Example: We wish to find the RMS noise voltage produced in a bandwidth of 1 MHz by a 1Mohm resistor at room temperature (room temp = 17 C or 290 K).

Solution: A shortcut is to remember that 4kT at room temp is 1.60×10^{-20} Joules.

$$V = (4kTRB)^{1/2} \qquad (2)$$
$$= [(1.60 \times 10^{-20})(1 \times 10^6)(1 \times 10^6)]^{1/2}(3) \qquad (3)$$
$$= [1.60 \times 10^{-8}]^{1/2}(4) \qquad (4)$$
$$= 126 \text{ microvolts RMS} \qquad (5)$$

From the above example, we can see that if the input resistance of an AC voltmeter was 1 Mohm and the meter had a 1 MHz bandwidth, the noise floor of the meter would be 126 µV regardless of how much gain the meter had. Signals below 126 µV would be lost in the noise. A 50 ohm resistor under the same conditions would produce only about 0.9 µV. It is easy to see why low impedances are desirable in low noise circuits.

To make things easier when dealing with noise, we frequently use the "lowest common denominator" to compare the noise of different devices. Although we may actually measure the noise of a device in a 1 MHz bandwidth because it is easier, we usually convert that number to a 1 Hz bandwidth (the lowest common denominator) in order to compare it to other noise sources. In other words, the noise measurement is taken in a bandwidth which is convenient and then converted to the value which would've been determined by measuring in a 1 Hz bandwidth. To do this conversion, the measured noise is divided by a scale factor which is the square root of the measured bandwidth. For example, if we had measured a noise source to have 1 mV RMS noise in a 1 MHz bandwidth, we divide that noise voltage by $B^{1/2}$ or $(1 \text{ MHz})^{1/2}$ which is 1000. The noise in a 1 Hz bandwidth would be 1 µV RMS. An examination of the units of this procedure will reveal that we should express this noise voltage as 1 $\mu V/\sqrt{Hz}$. It is important to note that noise voltage is not usually expressed in V/Hz.

Available Noise Power

In RF applications, we usually deal with circuits having matched input and output impedances, and are therefore more concerned with the power available from a device than the voltage. In this case, it is common to express the noise of a device in terms of the available noise power. The Maximum Power Transfer Theorem predicts that the noise power delivered from a source to a matched load can be derived:

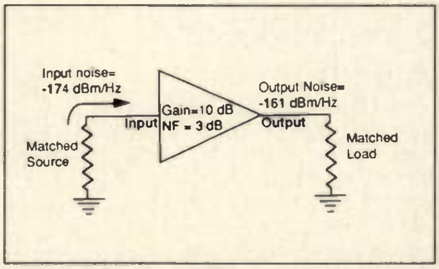


Figure 2. Amplifier at room temperature and with matched input and output.

800 to 2400MHz, to +26dBm from **$199**

In a broadband amplifier, excellent low-noise performance is generally synonymous with low power output because of design compromises. Not so with Mini-Circuits' TO-, ZEL-, and ZHL- low-noise amplifier series, where a noise level less than 1.5dB is accompanied with up to +26dBm power output. In front-end applications, it is undesirable for the amplified output to appear back at the input; these amplifiers effectively isolate the output signal from the input by as much as 35dB and use the shortest possible lead lengths to minimize parasitics and optimize NF performance. Detailed performance specs are included in our 740-pg RF/IF Designer's Handbook.

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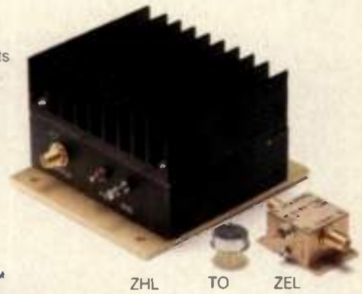
SPECIFICATIONS

Pin Model	TO 0812LN	TO 1217LN	TO 1724LN			
Connector Version	ZEL 0812LN	ZEL 1217LN	ZEL 1724LN	ZHL 0812HLN	ZHL 1217HLN	ZHL 1724HLN
Freq. (GHz)	0.8-1.2	1.2-1.7	1.7-2.4	0.8-1.2	1.2-1.7	1.7-2.4
NF, db, max*	1.6 1.5	1.6 1.5	1.6 1.5	1.5	1.5	1.5
Gain dB, min	20	20	20	30	30	30
Output Pwr., dBm 1dB Comp	+8	+10	+10	+26	+26	+26
Intercept Pt 3rd order, dBm typ	18	25	22	36	36	36
Price $ (Qty. 1-9)	TO $199.00		ZEL $274.00		ZHL $349.95	

NOTES
1. NF max. at room temperature increases to 2 dB typ. at +85°C
2. SMA connectors only, ZEL and ZHL units
3. Operating temperature -54°C to +85°C
4. DC power, 15V, 70mA for TO and ZEL, 15V, 725mA for ZHL
5. VSWR 2.5:1

*TO-8 includes test fixture loss



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(V_{oc} is the open circuit voltage of the source)

$$P = (V_{oc}/2)^2/R \tag{6}$$

$$= [((4kTRB)^{1/2})/2]^2/R \tag{7}$$

$$= (kTRB)/R \tag{8}$$

$$= kTB \tag{9}$$

It is important to notice that the resistance (R) has dropped out of the equation. Therefore, the available noise power from all finite, non-zero resistances is the same.

So, in a 1 Hz bandwidth at 290 degrees K (room temperature), $kTB=4.0x10^{-21}$ watts. The watt however, is not a convenient unit for measuring RF noise or signals. The dBm is preferred. (0 dBm = 1 milliwatt). Therefore, kTB at 290 K in a 1 Hz bandwidth is:

$$= 10LOG[4.0\times10^{-21} \text{ watts}/.001\text{watt}] \tag{10}$$

$$= -174dBm \tag{11}$$

Since this is in a 1 Hz bandwidth (B=1), this is a lowest common denominator, and is expressed as –174 dBm/Hz. This means that at 290 K, the absolute lowest noise power which a circuit can provide is –174 dBm/Hz. The only way to get noise power lower than this is to lower the temperature. Frequently, engineers will say that something has a noise floor of kTB. When kTB is used as a reference noise level, the temperature is generally assumed to be 290 K such that the noise level in question is –174 dBm/Hz.

In addition to thermal noise, amplifiers and other devices with semiconductors in them also contribute other forms of noise to a signal. Shot noise is a type of noise similar in spectral content to thermal noise. This noise is created in diode junctions (both in diodes and transistors). It comes about due to the way charge carriers are randomly emitted from the cathode (or emitter) during forward conduction. Flicker or 1/f (pronounced "one over f") noise also occurs in semiconductors. It is caused by the random recombination of minority carriers in the depletion region of the base-emitter junction of bipolar transistors. This noise is not of equal magnitude at all frequencies like thermal or shot noise, but is greatest at low frequencies and drops off at 10 dB/decade, as the observed frequency is increased. All of these different types of physical processes, unfortunately, lead to noisy devices.

Noise Figure

When designing circuits for use with extremely weak signals, noise is an important consideration. The noise contribution of each device in the signal path must be low enough that it will not significantly degrade the signal to noise ratio.

Noise figure is used to describe the noise contribution of a device. An ideal amplifier would have no noise of its own, but would simply amplify what went into it. For example, a 10 dB amplifier would amplify the signal (and the noise) at its input by 10 dB. Therefore, although the noise floor at the output of the amplifier would be 10 dB higher than at the input, the amplifier would not change the signal to noise ratio. A "real-world" amplifier will not only amplify the noise at its input, but will contribute its own noise to a signal. This reduces the signal to noise ratio at the output of the amplifier. Noise figure (NF) is a measure of how much a device (such as an amplifier) degrades the signal to noise ratio.

In order to predict how noise will affect our signals, we can model noisy amplifiers with ideal amplifiers and noise sources. It is convenient to model a "real-world" amplifier as in Figure 1. This diagram shows the "real-world" amplifer to have two major internal components: an "ideal noiseless" amplifier and a noise source. The noise source adds noise to any signal which enters the amplifier and then the ideal amplifier amplifies the whole thing by an amount equal to its gain, with no noise contribution of its own.

The model makes certain assumptions. The first is that the noise contribution of the internal noise source is fixed. That is, the noise voltage which it supplies does not change with the input signal level. It is a function of the design of the amplifier and cannot change. The second is that the input and output impedances of the amplifier are matched, such that the noise is purely based on the available noise power mathematics which were derived previously.

As an example, let's assume that we have an amplifier at room temperature with 10 dB of gain which has only a matched resistor at its input and output (see Figure 2). The noise at the input of the amplifier must be –174 dBm/Hz. If the amplifier is known to have a 3 dB NF, the internal noise source adds an equal noise to the input noise before amplification (i.e. doubles the input noise, therefore 10LOG(2) = 3 dB). Then, 10 dB of gain increases the noise by 10 dB. Therefore, the noise at the

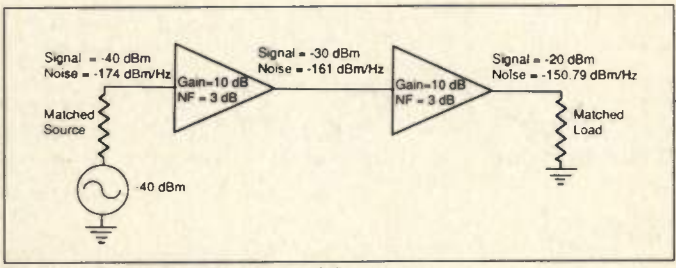

Figure 3. Cascaded amplifiers.

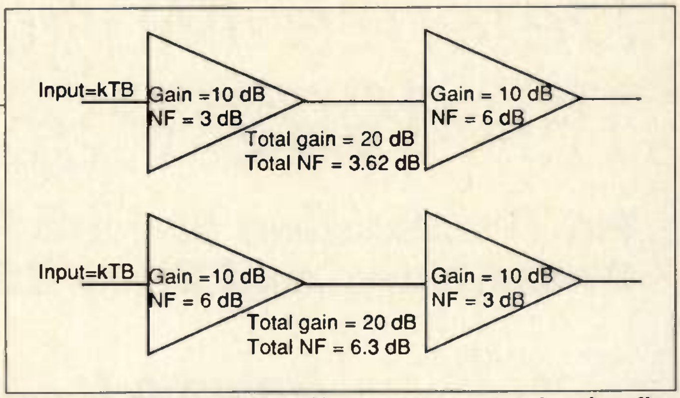

Figure 4. Effect of amplifier order on total noise figure.

output of the amplifier is 13 dB higher than at the input, or (–174 dBm/Hz) + (10 dB gain) + (3 dB NF) = –161 dBm/Hz.

The definition of NF is that it is calculated with the noise floor at kTB and T=290 K such that the floor is –174 dBm/Hz.

Another way to visualize the process is to think in terms of kTB instead of dBm. Using Figure 2 as an example, if the amplifier has a matched input resistance and its noise power is –174 dBm/Hz, then the input resistor is supplying 1kTB of noise to the amplifier. If the amplifier is known to have a noise figure of 3 dB, that tells us that the internal noise source will double the noise before amplification. Therefore, the internal noise source must supply an additional 1kTB of noise, to yield 2kTB, or twice the noise power. By thinking in this way, we can easily calculate how much actual power the noise source is contributing. This will prove helpful a little later.

What if we used the same amplifier as in the previous example (10 dB gain, 3 dB NF) except that the input noise floor was now somehow 6 dB higher than kTB (–168 dBm/Hz)? At first, one might think that we would add 13 dB to the input noise (10 dB of gain and 3 dB NF) and that would be the output noise. However, remember that the noise contribution of the amplifier's noise source is fixed and does not change with input signal. Therefore, when more noise is present at the amplifier input, the contribution of the internal noise source is less significant in comparison. So, when the noise into an amplifier is higher than kTB, the amplifier's noise figure plays a smaller role in the amplifier's noise contribution. We must take this into account when calculating the noise contribution of the amplifier in these cases.

This is quite obvious when we think in terms of kTB. If the input noise is 6 dB higher than kTB, it is 4 times kTB, or 4kTB. We already determined that the contribution of the internal noise source in an amplifier with a noise figure of 3 dB is 1kTB. Clearly, in the case where the input noise floor is higher than kTB, the contribution of the internal noise source (in this case, only 1kTB) is less significant than when the noise floor is at or near kTB.

We can directly measure the noise figure of a device using equation 12 if we adhere to the rule that the noise figure of a device is only calculated with the input noise level at kTB.

NF = [S/N Ratio at device input (in dB)]
– [S/N Ratio at device ouput (in dB)] (12)

If a measurement of NF is attempted using equation 12 when the noise at the input of the device is not at kTB, the measurement will be wrong.

Noise figure need not be expressed in dB. Noise Figure is the logarithm of Noise Factor, which is a power ratio. Frequently (as you will soon see) it is necessary to convert to a power ratio in order to make a calculation. Noise Figure (NF) and Noise Factor are easily converted between using the following formulae:

$$\text{NF} = 10\,\text{LOG}\,[\text{Noise Factor}] \qquad (13)$$

$$\text{Noise Factor} = 10^{\text{NF}/10} \qquad (14)$$

Consider the circuit of Figure 3 which uses two identical amplifiers in cascade. Each amplifier has 10 dB of gain and NF=3 dB. The signal goes in at –40 dBm with a noise floor at kTB. We can calculate that the signal at the output of the first amp is –30 dBm and the noise is (–174 dBm/Hz input noise)+(10 dB of gain)+(3 dB NF) = –161 dBm/Hz. We cannot use this method to compute the noise contribution of the second amplifier because we know that the noise entering the second amp is now significantly higher than kTB. It is fairly staightforward to go through the calculations to compute the noise contribution of the second amp. Generally, one would use the kTB method:

First, figure out how many kTBs are entering the second amp. (–161 dBm/Hz is 13 dB higher than kTB. 13 dB is a power ratio of 20x. So the noise floor at the second amp is 20 times kTB or 20kTB.)

Next, calculate how many kTBs are added by the noise source of the second amp (in this case, 1kTB because the NF=3 dB, derived previously).

Finally, calculate the increase in noise floor at the second amp as a ratio and convert to dB. [Ratio of (input noise floor) + (added noise) to (input noise floor) is (20kTB+1kTB) / (20kTB) = 20/21.] This in dB = 10LOG(21/20) = .21 dB. Therefore, the second amplifier only increases the noise floor by .21 dB even though it has a noise figure of 3 dB, simply because the noise floor at its input is significantly higher than kTB.

An equation can be derived (see Appendix) which computes the noise contribution of an amplifier given the input conditions (i.e. noise level) and noise figure. This equation (15) is a "lumped" version of the previous kTB calculation and does not account for the noise enhancement due to the amplifier gain. That must be added on afterwards.

$$\begin{array}{c}\text{Noise}\\ \text{Contribution}\\ \text{of amp (in dB)}\end{array} = 10\text{LOG}\left[1 + \frac{(\text{NFAC}) - 1}{10^{\left(\frac{\text{Nfloor}+174}{10}\right)}}\right] \qquad (15)$$

To demonstrate this equation, we can see how the second amplifier in Figure 3 affects the noise floor. The input noise is at –161 dBm/Hz and the NF = 3 (a noise factor of 2 from equation 14). From equation 15, the noise contribution of the second amp is found again to be .21 dB. Then, the amplifier's gain adds another 10 dB. The noise floor at the output of the chain is (–161 dBm/Hz) + (0.21 dB) + (10 dB) = –150.79 dBm/Hz.

The first amplifier of Figure 3 degrades the signal to noise ratio by 3 dB, while the second amplifier degrades it by only 0.21 dB. When amplifiers are cascaded together in order to amplify very weak signals, it is generally the first amplifier in the chain which will have the greatest influence upon the signal to noise ratio because the noise floor is lowest at that point in the chain.

Frequently we are interested in determining the total noise figure of a chain of amplifiers (or other devices) but are not particularly interested in the individual contribution of each device. This can easily be approached with a common equation which takes into account the increasing noise floor due to amplification at the input to each amp.

$$NFAC_{total} = NFAC_1 + \frac{NFAC_2 - 1}{G_1} + \frac{NFAC_3 - 1}{G_1 G_2} + \ldots + \frac{NFAC_3 - 1}{G_1 G_2 \ldots G_{n-1}} \qquad (16)$$

where:
NFAC = Noise factor of each stage
G = Gain of each stage as a ratio, not dB (i.e. 4 ×, not 6 dB)

Using this equation, we can solve for the total NF of the amplifier pair in Figure 3.

$NFAC_{total}$ = 2 + (2-1)/10 = 2+1/10 = 2.1, using equation 13, covert this to dB:

NF = 10log(2.1) = 3.22

This answer is almost the same as was calculated by summing the individual contributions of each amplifier. (There are round off errors involved.)

Equation 16 is an excellent tool to show that the first amplifier in a chain has the most significant effect on the total noise figure of the chain than any other amplifier in the chain. To prove that the lower noise figure amplifier should usually go first in a line of amplifiers (assuming all else is equal), imagine two amplifiers with equal gain, but with different noise figures (see Figure 4). Assume 10 dB of gain in each amplifier. One amp is NF=3 dB and the other NF=6 dB. When the 3 dB NF amp is first in cascade, the total NF is 3.62 dB. When the 6 dB amp is first, the total NF is 6.3 dB! This also applies to gain. If two amplifiers have the same noise figure but different gains, the higher gain amplifier should precede the lower gain amplifier to acheive the best overall noise figure.

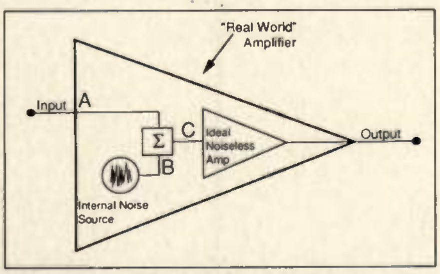

Figure 5. Model for derivation of Equation 15.

Noise Figure of Other Devices

Up to this point, the discussion of noise figure has been confined to amplifiers. This is because amplifiers are commonly measured for noise figure and they make convenient examples. However, all devices which process a signal contribute noise and thus have a noise figure. Mixers, transistors, diodes, isolators, etc. all have noise figures. For example, RF attenuators ("pads") have a noise figure which is equal to their attenuation value. A 10 dB pad has a 10 dB NF. How? If a signal enters a pad and the noise floor is at –174 dBm/Hz, the signal is attenuated by 10 dB while the noise floor remains constant (it cannot get any lower than –174 dBm/Hz at room temperature). Therefore the signal to noise ratio through the pad is degraded by 10 dB. Just like amplifiers, if the noise floor is above kTB, the signal to noise ratio degredation of the pad will be less than its noise figure.

Conclusion

Noise is created by many physical processes which cannot be avoided. Living with noise means we must be able to measure and predict it. One of the ways this is accomplished is through the noise figure, which is a means by which engineers describe the noise contribution of electrical devices to a signal. It takes a bit of experience to develop an intuitive "feel" for noise figure, but knowing what noise figure is and why it is important are the keys to building such experience.

Appendix

Given the model in Figure 5 and the following notation:
kTB@A = Number of kTB's at point A,
kTB@B = Number of kTB's at point B, etc.

The contribution of the "real-world" amplifier to the noise level is a function of the noise floor at the input of the "ideal amp" compared to the noise level at the input of the "real world" amp, or:

$$\text{Contribution of amp} = 10\,\text{LOG}\left[\frac{kTB@C}{kTB@A}\right] \qquad (17)$$

$$\text{But } kTB@C = kTB@A + kTB@B \qquad (18)$$

From Equations 17 and 18:

$$10\text{LOG}\left[\frac{kTB@C}{kTB@A}\right] = 10\text{LOG}\frac{kTB@A + kTB@B}{kTB@A} \qquad (19)$$

$$= 10\text{LOG}\left[1 + \frac{kTB@B}{kTB@A}\right] \qquad (20)$$

$$\text{And } kTB@B = NFAC\text{-}1 \qquad (21)$$

where NFAC is the Noise Factor of the amplifier.

$$kTB@A = 10^{X/10} \qquad (22)$$

where X = the number of dB the noise floor is above kTB. So X = NFLOOR +174 (assuming room temperature, and where NFLOOR is the level of the noise floor in dBm/Hz.)

Substituting:

$$\begin{array}{c}\text{Noise}\\ \text{Contribution}\\ \text{of amp (in dB)}\end{array} = 10\text{LOG}\left[1 + \frac{(NFAC) - 1}{10^{\left(\frac{Nfloor+174}{10}\right)}}\right] \qquad (23)$$

where:
NFAC = Noise Factor of amp
Nfloor = input noise floor in dBm/Hz. *RF*

Acknowledgements

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2. Course notes from (Basic RADAR) (Raytheon Course), January 1985 by M. B. Fishwick, Raytheon file number EM85-0076.

About the Author

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A Narrow Band FM Discriminator

By George Kassabian
University of California, San Diego

This entry to the 1992 RF Design Awards was produced for an instrumentation application which required the ability to detect minute frequency deviations in piezoelectric transducers within milliseconds to several seconds. The operating frequency is 30 kHz and the FM varies from ±0.1 Hz to ±2 Hz in 100 milliseconds or less.

The circuit presented here is an effective method of achieving extremely narrow bandwidth, low noise FM demodulation and/or a tracking bandpass filter. A Q multiplying technique and the phase shifting properties of a series-tuned LC circuit are the main components of this frequency discriminator.

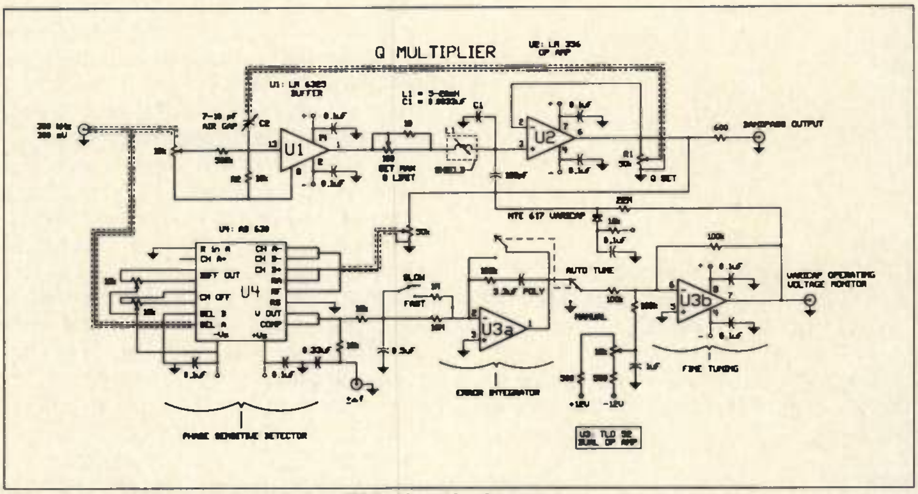

Figure 1. Schematic of the FM discriminator.

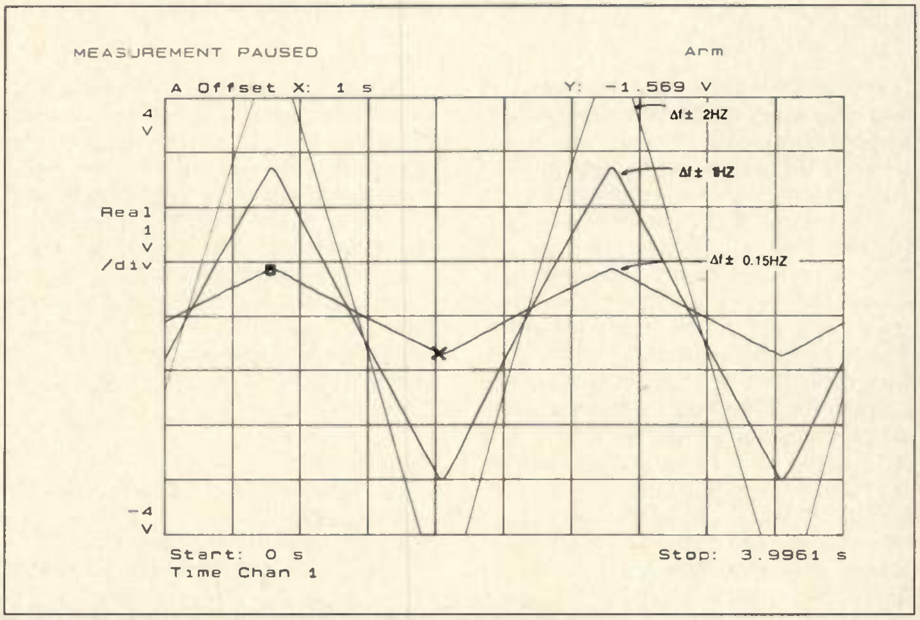

Figure 2. Tracking response of the discriminator.

Circuit Description

The heart of the Q multiplier consists of a low source-impedance buffer (LM6325, U1), that drives a series-tuned LC circuit. At resonance, a maximum voltage is developed at –90 degrees, and is coupled to the high input impedance op amp (LM356, U2), working as a unity gain non-inverting buffer. This provides the driving voltage to a +90 degree phase shifter consisting of C1 (7 to 10 pF) and R1 (10k). This network will provide the necessary positive feedback signal to be summed with the input signal. The initial amplification of the incoming signal is accomplished exclusively by the natural Q of the LC resonant circuit (which is very low, approximately 50, in this circuit). As the positive feedback is applied, the major losses caused by the inductor, and to a lesser degree, the tuning capacitor, are compensated. This results in a large increase in Q or gain without increasing the noise due to the inherent passive amplification of the resonant LC circuit. A Q up to 3000 is possible without any instability.

The Q set potentiometer, R1, will provide a feedback level adjustment to set the width of the passband or the Q factor. The lower the Q, the broader the frequency deviation that can be measured. Electronic fine-tuning is provided by the TLO 52, U3A, and the varicap, D1. The FM signal is obtained by a phase sensitive detector, the AD 630, U4. The output is linear and proportional to the deviation as long as the frequency deviations are within the narrow flat top portion of the tuned LC response curve.

An AFC circuit was added to provide long term frequency stability and/or to be used as a tracking bandpass filter. The AFC consists of TLO 52, U3B. An error voltage integrator with a 3 or 30 second time constant (fast or slow track) provides the summing voltage necessary to fine tune and phase-lock the LC tuned circuit to the incoming signal. *RF*

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

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LRFMS-4L	400-1500	DC-900	+3	8.0	12.95
LRFMS-1	0.5-500	DC-500	+7	6.0	4.95
LRFMS-2	5-1000	DC-1000	+7	7.0	5.95
LRFMS-2A	5-1000	10-1000	+7	6.0	5.95
LRFMS-4	5-1500	DC-1000	+7	7.5	9.95
LRFMS-5	10-2000	10-900	+7	8.0	12.95
LRFMS-2T	1150-1900	DC-300	+7	6.0	10.95
LRFMN-25	2000-3000	DC-500	+7	8.0	11.95
LRFMS-1A-10	2-500	DC-500	+10	6.0	6.95
LRFMS-2-10	5-1000	DC-1000	+10	7.0	7.95
LRFMS-5-10	10-1500	10-1000	+10	7.5	11.95
LRFMS-1A-13	2-500	DC-500	+13	6.0	7.95
LRFMS-2-13	5-1000	DC-1000	+13	7.0	8.95
LRFMS-5-13	10-1500	10-1000	+13	7.5	13.95
LRFMS-1A-17	2-500	DC-500	+17	7.0	9.95
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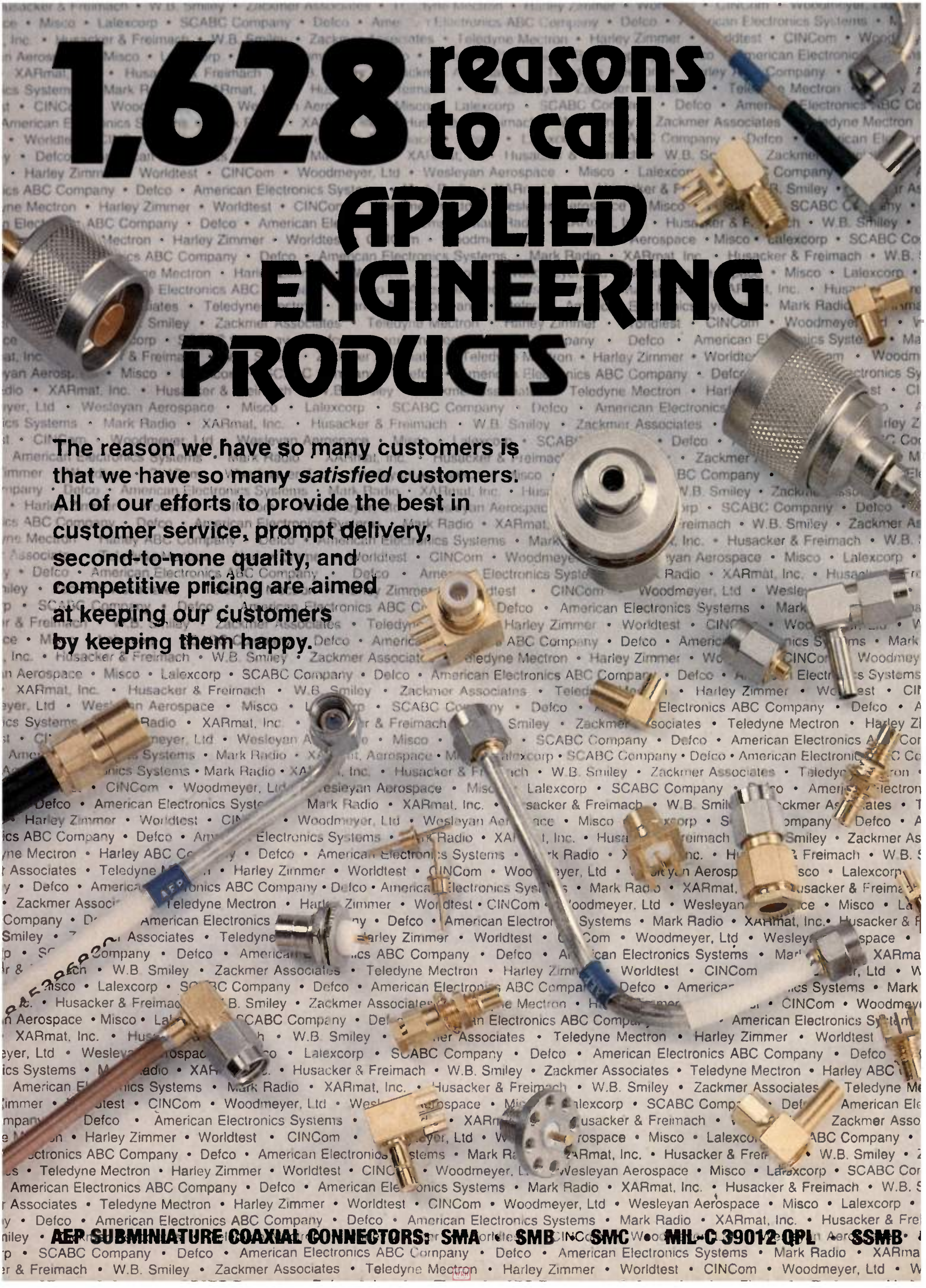
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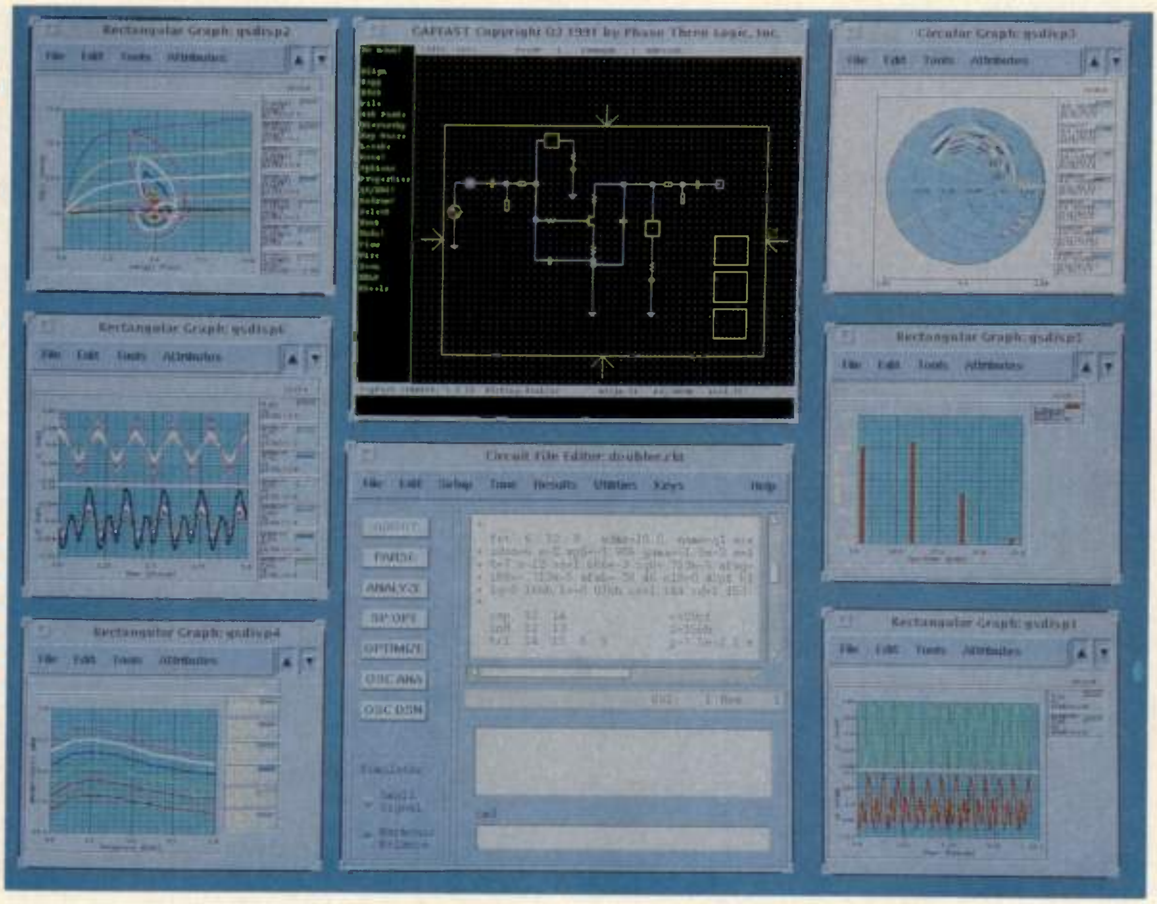

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A Nodal Network Analysis Program

By John A. Eisenberg
John A. Eisenberg and Associates

MNAP9 is a full nodal circuit analysis program capable of analyzing any arbitrary connection of components including resistors, capacitors, inductors, transmission lines, voltage controlled current sources and user supplied S-parameter blocks. In addition, the program will permit inclusion of microstrip lines on an alumina substrate of any reasonable thickness.

The program was first developed to run on Z80 class computers under CP/M. I have converted my original code written in an early (1982 vintage) Microsoft Basic and ported it to a compiled BASIC (Microway) running on the IBM PC and its clones.

I wrote MNAP9 well before the era of commercial network analysis programs such as EEsof's Touchstone. Having it on my desk-top PC saves me many trips to a community workstation on which Touchstone resides. MNAP9 has plenty of power to try out nearly any idea that you will dream up without the need to compete with your colleagues for use of the workstation.

The program is command driven, and after first asking if a printer is available, solicits a command. A rich assortment of commands is provided for entering the circuit description, editing it or listing its contents. Input and output port node number assignments may be changed, as may the source and load resistances. Commands for reading or writing circuit files are available, as is a means of saving an entire circuit as an S-parameter file. The saved S-parameter file may later be read in as a XSTR block, making the analysis of quite large circuits feasible. Commands for analyzing the circuit and printing the results are also present. Finally, a HELP command is provided which allows the user to browse page by page through the instructions provided in the file SELP.INS. It should be noted that all entries to MNAP9 must be in upper case. You will get no help from "help", only from "HELP".

As to theory, MNAP9 is a nodal circuit analysis program based on the reduction of the indefinite admittance matrix (IAM) by a Gauss-Jordan elimination. As each element is entered, its IAM is computed at each specified frequency and is summed into the overall circuit IAM. This matrix is then reduced to a two-port admittance matrix between the specified generator and load nodes. The two-port Y-matrix is converted to an S-matrix for output. The majority of work in this process is bookkeeping. First the user enters node numbers, which can range from 0 (ground) to 99, are internally (and transparently to the user) reordered so that node 1 is the input node, node 2 the output node and the rest of the nodes appear in sequence. This makes it possible to use a common set of routines to reduce the IAM. Additional bookkeeping is needed to always provide the user with their original set of nodes. The list of references used to build MNAP9 would be too long to print here, however the book by Gupta, Garg and Chadha (1) does contain nearly all the theory underlying MNAP9.

MNAP9 will run on any PC (8088, 80286 80386 or 80486), as long as the system has an 80X87 co-processor. The program has passed the test of time in that most of the bigger bugs have been slain over the years since 1983, when the program was written. MNAP9 contains no optimizer nor does it do noise or non-linear analysis. However, the program does contain extensive error checking that helps you out if you make an input or run time error. The easiest way to learn to use MNAP9 is to use the H command or print the file SELP.INS which contains instructions on how to use the program as well as a number of examples. Two sample circuit files, AMP618.DAT (a wideband reactively matched amplifier stage), and FETMDL.DAT (computes GaAs FET S-parameters from a simple equivalent circuit), are included as additional examples and test cases. AMP618.DAT uses the transistor data file LN700.XST. Between the examples and the H command you should have no trouble figuring out how to use the program.

```
CIRCUIT TOPOLOGY AND ELEMENT VALUES

LINE NUMBER  1         TYPE STLN
CONNECTING NODES   1          2
ZO=  80  OHMS   LEN=  .01  IN   EPSILON =  9.7
LINE WIDTH=  0  MILS ON  50  MIL AL203

LINE NUMBER  2         TYPE SC
CONNECTING NODES   2          3
CAPACITANCE IS  .54  PF

LINE NUMBER  3         TYPE STLN
CONNECTING NODES   3          0
ZO=  95  OHMS   LEN=  .102  IN   EPSILON =  9.7
LINE WIDTH=  0  MILS ON  50  MIL AL203

LINE NUMBER  4         TYPE STLN
CONNECTING NODES   3          8
ZO=  40  OHMS   LEN=  .024  IN   EPSILON =  9.7
LINE WIDTH=  0  MILS ON  50  MIL AL203

LINE NUMBER  5         TYPE SL
CONNECTING NODES   3          4
INDUCTANCE IS   .24  NH

LINE NUMBER  6         TYPE XSTR
CONNECTING NODES   4          5          0
FILE NAME IS LN700

LINE NUMBER  7         TYPE SL
CONNECTING NODES   5          6
INDUCTANCE IS   .695  NH

LINE NUMBER  8         TYPE STLN
CONNECTING NODES   6          0
ZO=  70  OHMS   LEN=  .091  IN   EPSILON =  9.7
LINE WIDTH=  0  MILS ON  50  MIL AL203

LINE NUMBER  9         TYPE STLN
CONNECTING NODES   6          7
ZO=  72  OHMS   LEN=  .099  IN   EPSILON =  9.7
LINE WIDTH=  0  MILS ON  50  MIL AL203


GENERATOR NODE IS  1         LOAD NODE IS  7
      SOURCE RESISTANCE IS  50  OHMS
      LOAD RESISTANCE IS     50  OHMS
```

Figure 1. Circuit listing.

```
*******************************
*                             *
*  CIRCUIT ANALYSIS BY MNAP9  *
*                             *
*******************************
```

FREQ	GAIN-DB	T-PHASE	VSWR-IN	VSWR-OUT	ISOLATION-DB
6	6.766122	110.1718	-87.10943	3.849658	-24.02161
8	6.996293	45.14663	26.7601	1.713233	-22.50981
10	6.467228	-14.9333	8.46581	1.747014	-21.74826
12	5.423945	-67.57278	5.440884	2.673651	-21.42445
14	5.188361	-119.1384	3.489869	2.370625	-21.60345
16	5.869666	178.4853	2.408189	1.543109	-22.88529
18	5.675028	86.66737	1.339383	1.618449	-24.22191

Figure 2. Analysis of circuit listed in Figure 1, (short form).

```
*******************************
*                             *
*  CIRCUIT ANALYSIS BY MNAP9  *
*                             *
*******************************
```

FREQ-MHZ	[S11]	<S11	[S12]	<S12	[S21]	<S21	[S22]	<S22	GT-DB
6000.0	1.023	-60.8	0.063	38.5	2.179	110.2	0.588	0.9	6.8
8000.0	0.928	-101.5	0.075	-14.0	2.238	45.1	0.263	-19.1	7.0
10000.0	0.789	-134.8	0.082	-64.3	2.106	-14.9	0.272	17.6	6.5
12000.0	0.689	-160.1	0.085	-111.5	1.867	-67.6	0.458	-2.6	5.4
14000.0	0.555	172.4	0.083	-159.5	1.817	-119.1	0.407	-34.9	5.2
16000.0	0.413	134.3	0.072	151.8	1.966	178.5	0.214	-77.5	5.9
18000.0	0.145	-42.1	0.062	84.1	1.922	86.7	0.236	30.0	5.7

Figure 3. Analysis of circuit listed in Figure 1, (S-parameter form).

MNAP9 is not a substitute for a full featured network analysis program, but it certainly is a handy tool to have on your desk at work or on your home computer. If you do not have a math coprocessor, the MSDOS BASIC compatible file MNAP9.BAS is included; it is slow but it gets the job done. Please enjoy and provide whatever feedback that you wish.

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

References

1. Gupta, Garg and Chadha, *Computer Aided Design of Microwave Circuits*, Artech House, 1981.

About the Author

John A. Eisenberg received BSEE and MEE degrees from Cornell University. He was a co-founder, in 1975, of Narda Microwave, Western Operations, now a subsidiary of Loral Corp., where he later became Vice President and Technical Director. In 1984 he formed John A. Eisenberg and Associates, where his current efforts are largely directed at the design and application of GaAs MMICs and development of ultra linear UHF high power amplifiers. He can be reached at 25 Parsons Way, Los Altos, CA 94022, or by phone at (415) 941-7426.

Educational Project for Electronic Engineering Students

By Yvon Roy
Scientel Ltd.

The object of this paper is to describe an apparatus and method of use, which demonstrate in a simple way, various modulation phenomena. It has been demonstrated to a number of students in electronics and it was noted that greater learning interest and motivation was generated. This paper also describes the quadradyne demodulators used in this project.

The apparatus can generate standard AM with double sideband modulation, suppressed carrier AM with double sideband modulation and FM modulation. Special attention is given to the phenomena of the FM first carrier null and its effects on various waveforms.

The apparatus consists of two modulators; the first one is the AM modulator operating at a frequency of 455 kHz which can generate either standard AM or DSB suppressed carrier. The second one is the FM modulator operating at 100 MHz.

The following test equipment is required to make a full demonstration of each phenomena: an audio function generator with sinusoidal and triangular waves; an oscilloscope; and a spectrum analyzer with log and linear detectors (Our analyzer is an Anritsu model MS610J).

The spectrum analyzer is used for certain measurements as a selective narrow-band demodulator (zero dispersion). It must be capable of selecting any of the FM sidebands separated by a 455 kHz bandwidth, demodulate in the linear mode, and feed the results to the oscilloscope for waveform observation.

Description

Figure 1 shows the circuit diagram of the 455 kHz modulator. It is built around a TV modulator IC, the Motorola MC1374P. Construction is not critical because of the low frequency. The buffer amplifier, a National LH-0002C, provides an impedance transformation. The 100 ohm potentiometer is used to vary the level fed to the FM modulator. The two 5000 ohm potentiometers are used to bias the MC1374P. One is used to bias for minimum carrier for the double sideband suppressed carrier mode; the other is used to bias for standard AM. The final output frequency is not critical and can be 455 kHz ±10 percent. A switch is used to select the modulation mode.

Figure 2 shows the FM modulator built with a Motorola MC1648P oscillator. Two varactors, Motorola MV-209, are used to frequency modulate the oscillator. The 5000 ohm potentiometer is used to bias the varactors for best linearity. The output frequency of approximately 100 MHz can be adjusted by changing the length of the inductor. The output frequency can vary as much as 10 MHz on each side since the spectrum analyzer can follow these frequencies. The output level of the modulator is –5 dBm. In our prototype, the varactor

IC-1 - Motorola MC-1374P
IC-2 - National LH0002C
L1, L2 - Mouser Electronics #421IF200
C1, C2 - silver mica, 300 pF
All 0.1 uF cap., ceramic disc, 16V
C3 - 100 uF, 10 V, electrolytic
All resistors 5%, 0.25 W
ADJUSTMENT: Adjust R1 for minimum carrier; signal from function generator should generate 500 mVpp at pin 8 of IC-2 (suppressed carrier double sideband). Adjust R2 and function generator level to achieve 800 mVpp at pin 8 of IC-2 (standard AM with carrier). Adjust L2 for 455 kHz. Adjust L1 for maximum output.

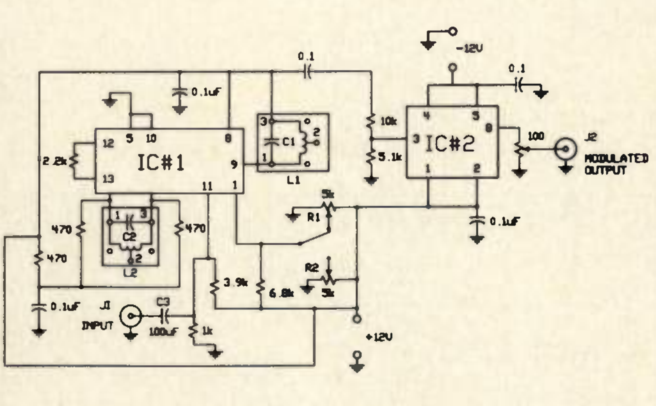

Figure 1. 455 kHz modulator.

IC-1 - Motorola MC-1648P
All resistors 5%, 0.25 W
Zener - 5.1 V, 0.5 W
All 0.1 and 0.01 uF capacitors ceramic, 16V
C4 - 100 uF, 16 V electrolytic
D1, D2 - Motorola MV-209
L1 - airwound, 6 turns, 3/16" dia., 5/16" long, 20 AWG
C3 - 500 pF, silver mica

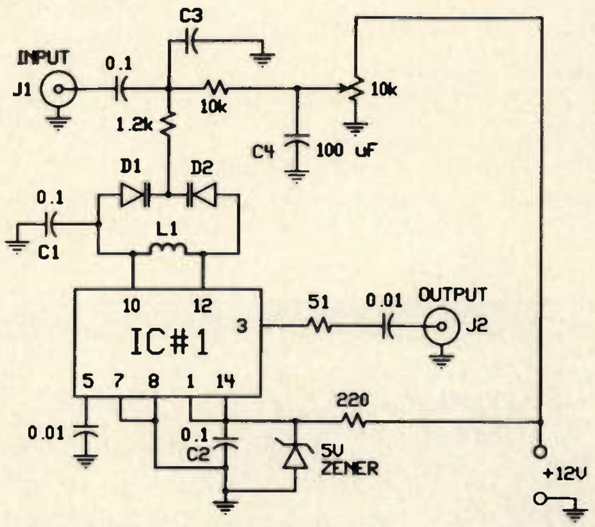

Figure 2. FM modulator.

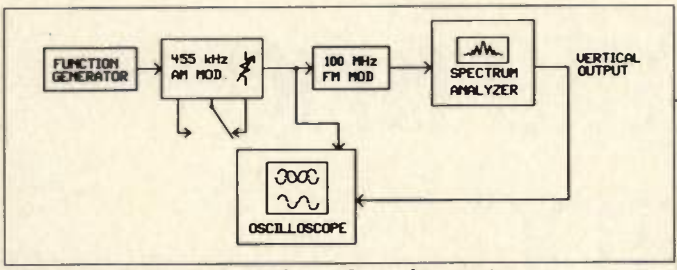

Figure 3. Interconnection of equipment.

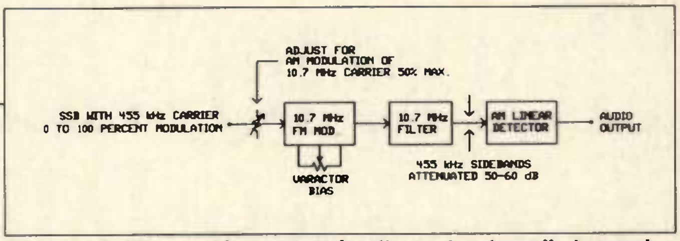

Figure 4. Block diagram of a "quadradyne" demodulator.

bias was 7.5 volts for best linearity; but this could be different with other varactors. Figure 3 shows the interconnections of both modulators and the test instruments.

Observation of Modulation Phenomena

AM: Standard AM and suppressed carrier AM can be observed on the oscilloscope and the spectrum analyzer. Modulation can be applied from the function generator. The spectrum analyzer should be tuned to 455 kHz and sideband analysis can be done. Envelope analysis can be observed on the oscilloscope.

FM: It is a very well known fact that in FM modulation, it is possible with sufficient modulation capability to reach a level where the carrier is highly attenuated; this is called "first carrier null." If the 100 MHz FM modulator is fed with the 455 kHz signal (no modulation, just carrier) it is possible to adjust the modulation level until the first carrier null occurs, Figure 5. If the level is increased further, the first 455 kHz sideband null occurs and so on. If modulation is applied to the AM modulator from the function generator at 5 kHz, the FM modulator will be fed with a signal which varies from zero level (100 percent) to maximum peak envelope, Figure 6. By observing the carrier of the FM modulator on the spectrum analyzer in the linear mode, it will be evident that the latter is AM modulated by a 5 kHz signal. The 455 kHz sidebands of the FM modulator are also modulated by a 5 kHz signal. It is possible to adjust the modulation in such a way as to reach the first carrier null coinciding with the peak of the 455 kHz modulated signal, therefore the carrier of the FM modulator is 100 percent modulated. We now have a condition where the 5 kHz signal has been transferred from the 455 kHz modulated carrier to the carrier of the 100 MHz modulator and its sidebands.

If the spectrum analyzer is operated in a narrow bandwidth in the linear detection mode, it is possible to observe the waveform of the modulated FM carrier. Figure 7 shows the distorted 5 kHz wave at the bottom; the top signal is the envelope of the 455 kHz signal fed to the 100 MHz FM modulator. The waveform is distorted because the transfer function in a FM modulator from no modulation of the carrier to first carrier null is not a linear one. This can be demonstrated by switching the function generator to a 5 kHz triangular wave as shown Figure 8. The response has the shape of a parabola, very similar to the

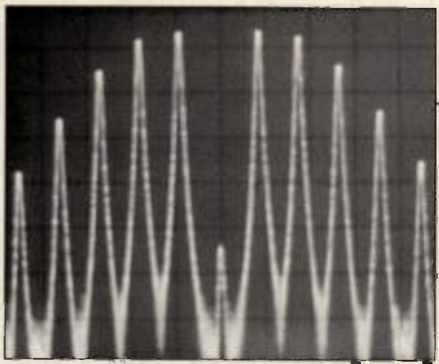

Figure 5. FM carrier modulated by 455 kHz carrier at first carrier null. Log scale: 10 dB division.

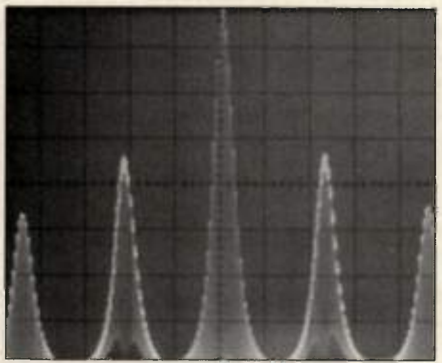

Figure 6. FM carrier modulated by 455 kHz modulated signal. The 455 kHz carrier is 100 percent modulated in standard AM by a 5 kHz wave from the function generator. Linear scale.

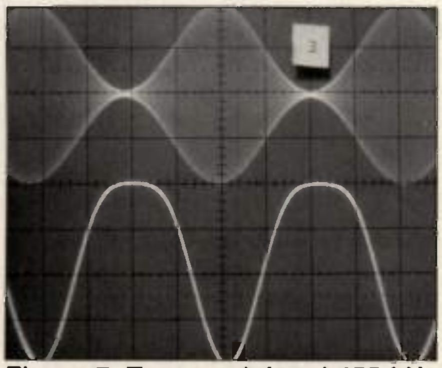

Figure 7. Top: modulated 455 kHz carrier by 5 kHz wave. Bottom: detected signal from spectrum analyzer. Linear scale.

shape of a two tone test from a suppressed carrier AM generator. If such a two tone test signal from the 455 kHz AM modulator is fed to the 100 MHz FM modulator, it turns out that the quadrature component of the two tone test signal has been almost canceled by the non-linear transfer function of the FM modulator as stated earlier, see Figure 9. The demodulated wave is for a 100 percent AM modulation of the 100 MHz carrier as shown in Figure 6. In Figure 10, the modulation of the 100 MHz carrier has been reduced to 50 percent and the detected wave is much more sinusoidal. If the single sideband amplitude of the 455 kHz AM modulator is reduced while the carrier remains the same, the detected wave remains sinusoidal. It appears from this demonstration that such a circuit can demodulate a single sideband signal with carrier and cancel out the quadrature distortion inherent in the single sideband system. This type of detector has been named the quadradyne detector. Conventional methods to eliminate the quadrature component require a synchronous detector, but the carrier regeneration is difficult to achieve because of the presence of both AM and PM in the single sideband with carrier signal.

It has also been demonstrated that if the FM modulator lacks linearity, the 100 MHz carrier of the FM modulator has incidental FM as well as AM modu-

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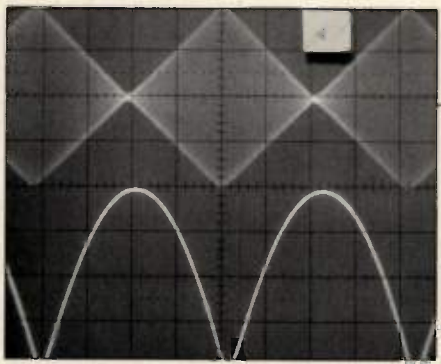


Figure 8. Top: modulated 455 kHz carrier by 5 kHz wave. Bottom: detected signal from spectrum analyzer. Linear scale. FM carrier is 100 percent AM modulated.

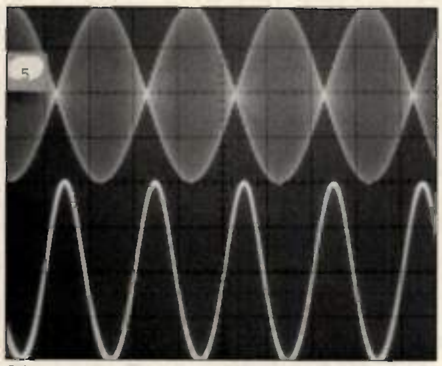


Figure 9. Top: modulated 455 kHz carrier in suppressed carrier mode. Bottom: detected signal from spectrum analyzer. Linear scale. FM carrier is 100 percent AM modulated.

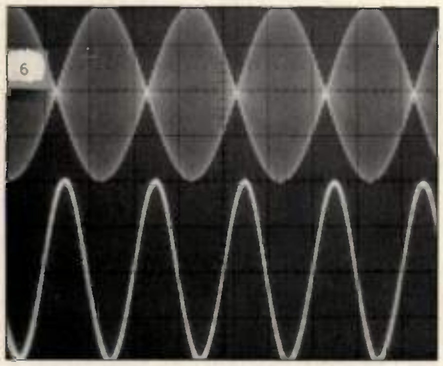


Figure 10. Top: modulated 455 kHz carrier in suppressed carrier mode. Bottom: detected signal from spectrum analyzer. FM carrier is 50 percent AM modulated.

lation. The varactor bias was changed to perform this test and the results are shown in Figures 11 and 12. Because of the limits imposed by our spectrum analyzer (dispersion 100 kHz minimum and IF filter bandwidth 1 kHz minimum), it was not possible to verify the incidental FM with modulating frequencies below 5 kHz. This test has demonstrated that the above modulation scheme can reveal

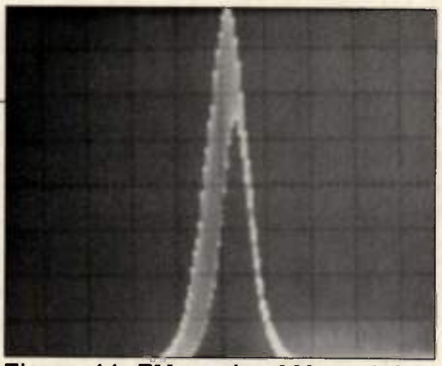


Figure 11. FM carrier AM modulated with incidental FM.

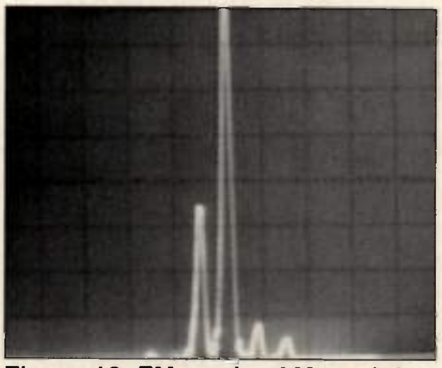


Figure 12. FM carrier AM modulated with incidental FM.

the non-linearity of FM modulators by measurement of the incidental FM on the carrier while using low modulation frequencies from the function generator.

Figure 4 shows the block diagram of a quadradyne detector for a practical application.

The author encourages students and teachers to build this project and use it in the lab and/or classroom. Many more tests too long to be enumerated in this paper can be done. For example, waveforms of the detected 455 kHz sidebands of the FM carrier are different from one sideband to another. Spectrum analysis reveals that some sidebands are AM modulated and others have suppressed carrier. Some sidebands have phase reversal of the 5 kHz demodulated wave, etc. *RF*

About the Author

Yvon Roy is president of Scientel Ltd. and Rextel Communications Inc. since 1985. He left the Canadian Broadcasting Corp. in 1985 after 31 years of service as manager of Transmission and Technical Resources Consultant. Mr. Roy can be reached at 1958 Le Royer St., Laval City, Quebec H7M 2S7 Canada. Tel: (514) 945-6234.

Program Aids Series-Feedback Transistor Oscillator Design

By Theodore Grosch

This program can be helpful in designing series-feedback oscillators of the form shown in Figures 1 and 2. The series-feedback oscillator utilizes a feedback circuit on the common terminal and a resonant terminating circuit on the input (port 1). This program will help to determine what the common terminal should be and the feedback circuit reflection Γ_3 and the termination circuit reflection Γ_t. The program will work on any Apple Macintosh computer with a copy of Microsoft BASIC.

First, enter the transistor S-parameters by using the "Enter" menu. There will be two active options in this menu. The S-parameters can be entered manually, or the S-parameters of a preset device can be selected. Select the "Enter Manually" menu item and answer the questions to enter the two-port, common- emitter S-parameters at the frequency of interest.

Currently, the device used in the option "Preset Device" is a NE02139B at 1.5 GHz, but it can be changed when the program is stopped. To change the device, open the list window and edit the section near the top of the listing to any desired transistor. The preset option is valuable if the same transistor is analyzed frequently.

After the two-port, common-emitter S-parameters are entered, the program calculates the three-port S-parameters. Then the "Config" menu will become active. The transistor can be analyzed in six configurations: common-emitter, common-base or common-collector and the output can be either of the two remaining terminals. The common terminal is where the feedback Z_3 is connected. The six options are listed under the "Config" menu in shorthand form. The CC OE menu item refers to a common-collector, output-emitter configuration.

There is one other menu item under "Config", "All Configs". When this menu item is picked, a table is produced which is shown in Figure 3. The table is designed to help choose a configuration that has the potential to satisfy the oscillation condition $\Gamma_0\Gamma_l=1$, by creating an infinite output reflection coefficient, Γ_0. When the "All Configs" menu item is chosen, the user will be asked what the magnitude of Γ_3 will be. The program will then find the angle of Γ_3 that will minimize the magnitude of Γ_t that forces Γ_0 to be theoretically infinite. A smaller magnitude of Γ_t implies a higher loaded Q of the resonant termination circuit, and a lower phase noise from the oscillator.

The table shows values of feedback reflection coefficient, "GAMMA- 3" and termination reflection coefficient, "GAMMA-T", that will simultaneously minimize the magnitude of Γ_t and create an infinite output reflection coefficient Γ_0. Figure 3 shows a table constructed with the NE02139B transistor in all six possible configurations. Three other columns are shown: the feedback circuit reflection coefficient, the optimum value for the termination circuit reflection coefficient and S_{22} of the transistor in the indicated common configuration and the feedback circuit Z_3 inserted between the common terminal and ground. The first row for each configuration is the magnitude and the second is the phase.

For example, Figure 3 shows the table for the NE02139B when the $|\Gamma_3|$ is equal to one. A NE0219B transistor in the common-emitter, output-collector configuration has an optimum $|\Gamma_t|$ when the angle of Γ_3 is 142 degrees. This is the lowest value for $|\Gamma_t|$ that can be achieved that produces an infinite Γ_0 with any angle of Γ_3 when $|\Gamma_3|$ is equal to one. It can be seen that the lowest value for $|\Gamma_t|$ can be achieved in any configuration (with the $|\Gamma_3|=1$) is the

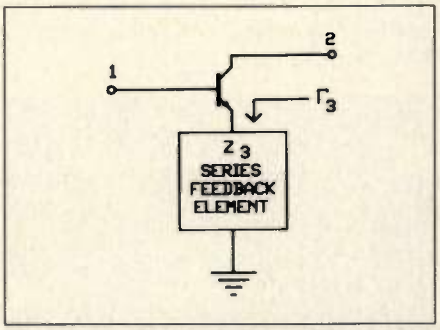


Figure 1. Transistor with series feedback element.

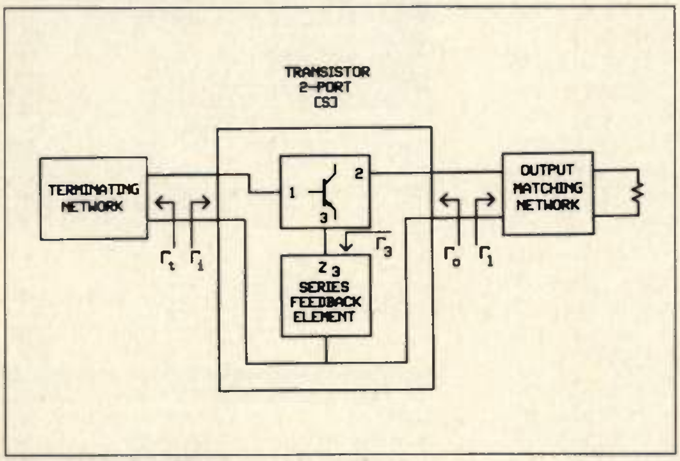
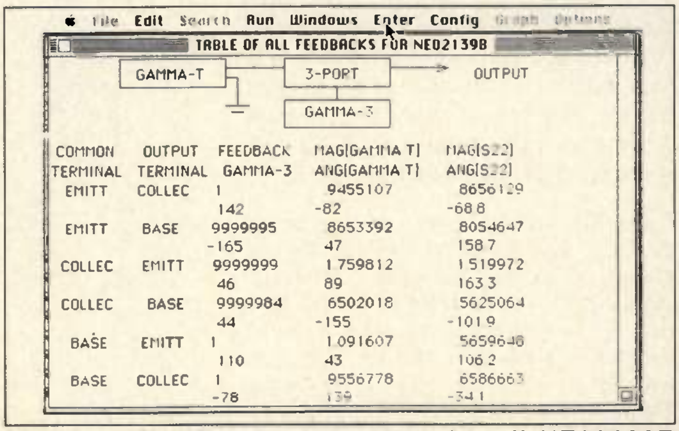


Figure 2. Oscillator schematic.



COMMON TERMINAL	OUTPUT TERMINAL	FEEDBACK GAMMA-3	MAG[GAMMA T] ANG[GAMMA T]	MAG[S22] ANG[S22]
EMITT	COLLEC	1	9455107	8656129
		142	-82	-68 8
EMITT	BASE	9999995	8653392	8054647
		-165	47	158 7
COLLEC	EMITT	9999999	1 759812	1 519972
		46	89	163 3
COLLEC	BASE	9999984	6502018	5625064
		44	-155	-101 9
BASE	EMITT	1	1 091607	5659648
		110	43	106 2
BASE	COLLEC	1	9556778	6586663
		-78	139	-34 1

Figure 3. Feedback parameters for all NE02139B configurations.

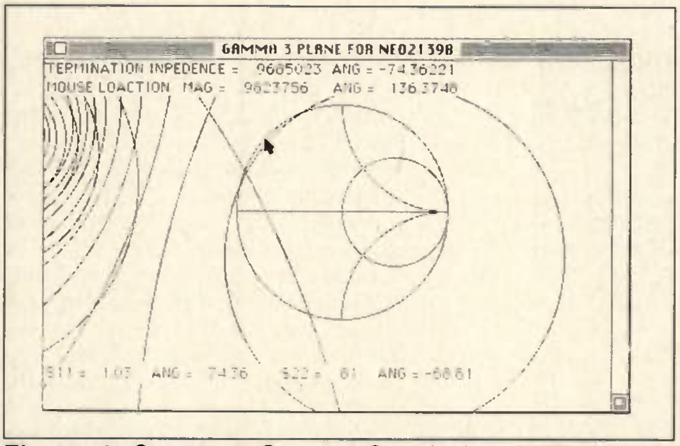


Figure 4. Constant S_{11} and S_{22} circles on the feedback Γ_3 plane.

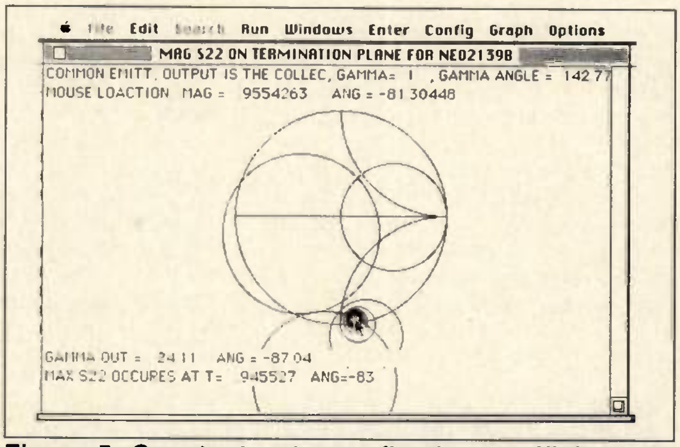


Figure 5. Constant output reflection coefficient circles on the termination plane.

common-collector, output-base configuration ($|\Gamma_t|$=0.65). Also shown in the table is S_{22} of the two-port created by the transistor in the indicated configuration and the feedback Γ_3 attached. This is useful to check if the circuit may oscillate on its own if the termination network were replaced by a matched load. A value of one or greater shows the possibility of spurious oscillation.

This table has four potential benefits:

1. Configurations can be identified that may not oscillate at all, (for example; the common-base, output-emitter and common-collector, output-emitter configurations in Figure 3 if $|\Gamma_3|$=1, Z_1=50 ohms).
2. The configuration with the minimum $|\Gamma_t|$ can be easily identified (to maximize the loaded Q of the termination circuit).
3. Identify possible spurious instabilities produced by the feedback circuit when $|S_{22}|>1$.
4. Quickly determine the phase of Γ_3 to see if the feedback network can be easily synthesized.

Note: When using the "All Configs" option, a particular configuration must still be selected from the "Config" menu to go on with the program.

Once the configuration of the oscillator circuit has been chosen, the "Graph" menu will be activated. The first item is "Stability". This will calculate the stability factor, K, of the transistor in the selected configuration without any feedback.

The next "Graph" option is "FB Plane". This plots the locus of constant magnitude S_{11} and S_{22} circles on the Γ_3 plane. An example of the graphical output is shown in Figure 4. This is a plot of the Γ_3 plane for the NE02139B in the common-emitter, output-collector configuration. The circles are the locus of constant $|S_{11}|$ and $|S_{22}|$=1,2,3,...,10 on the Γ_3 plane. This plot shows the values of S_{11} and S_{22} at any Γ_3 when a feedback circuit is added to the common terminal and the resultant two-port S-parameters calculated.

The mouse can be positioned anywhere in this plane and depressed. The mouse position on the Γ_3 plane will be displayed. At the bottom of the window, the values of S_{11} and S_{22} corresponding to this Γ_3 for the feedback network are displayed. And finally, the reflection coefficient (Γ_t) that forces $\Gamma_0 \rightarrow \infty$ is displayed, (this satisfies the oscillation condition $\Gamma_t\Gamma_i$=1 when $\Gamma_0 \rightarrow \infty$).

The next two options under the "Graph" menu are very similar, "Plot" and "Resistance". The user will be asked to supply the magnitude and angle of Γ_3 that is to be used for the calculation. The default value (in brackets) will be the Γ_3 with a magnitude of one and the optimum phase to minimize Γ_t.

The "Plot" option will map the locus of constant output reflection coefficient circles on the termination impedance plane. Included on this graph is a circle that displays the locus of output resistance of a negative 50 ohms.

The "Resistance" option will plot the locus of constant output resistance circles on the termination plane. Included on this plot is the point where if Γ_t equals this impedance, the output reflection coefficient will theoretically be infinite.

These two options, "Plot" and "Resistance" are used to select the termination impedance to yield a desired output reflection coefficient or output negative resistance. Figure 5 and Figure 6 show an example of the "Plot" and "Resistance" windows for the NE02139B in the common-emitter, output-collector configuration with a feedback $\Gamma_t=1\angle 142.8$. These are plots of the termination plane and

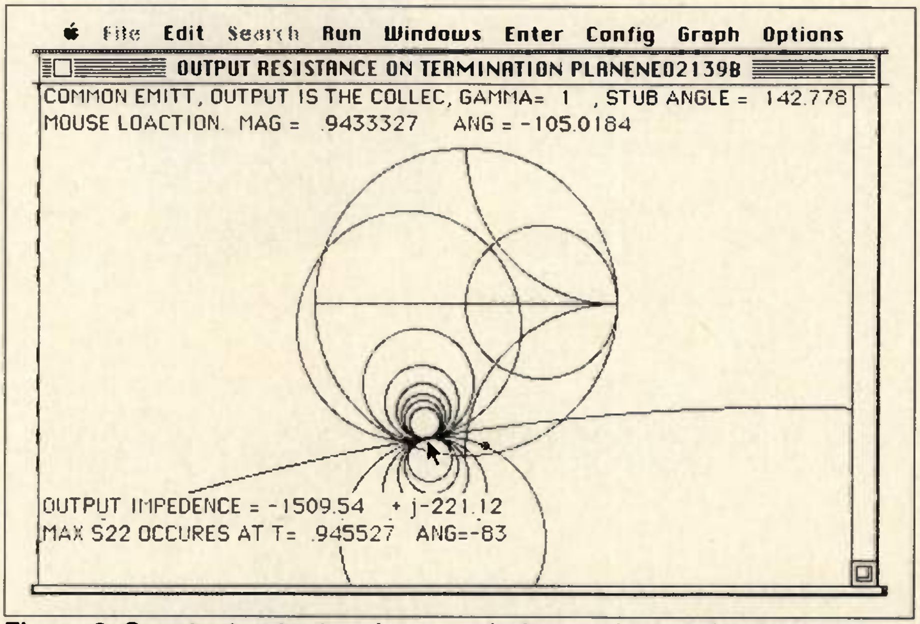


Figure 6. Constant output resistance circles on the termination plane.

how the termination impedance effects the output reflection or resistance. The "Plot" window shows the locus of constant $|\Gamma_0|=1,2,3,\ldots,10$ on the Γ_t plane. The "Resistance" window will display the locus of constant $Re(Z_0)=250$, 200, 150, 50, 0, –50, –100, –150, –200 and –250 ohms on the Γ_t plane. The mouse can be clicked in any region of the Γ_t plane; the reflection coefficient at that mouse location is printed near the top of the screen, and the corresponding output reflection coefficient or output impedance is displayed near the bottom of the screen.

Figure 5 shows the location of the mouse (arrow) where the circles of constant output reflection coefficient of 1,2,3,...10 have been plotted. The center of these concentric circles ($0.955\angle-81.3$) is the point where the combination of feedback and termination reflection coefficients will theoretically create an infinite output reflection coefficient.

Figure 6 shows the constant, real output-resistance circles of 250, 200, 150, 100, 50, 0, –50, ..., –250 ohms. The mouse is at $\Gamma_t=0.943\angle-105$. With this termination impedance and the feedback chosen, the output should have a negative resistance of 1500 ohms.

The "Options" menu becomes active when in the "Plot" or "Resistance" windows. This menu allows the user to add circles and change Γ_3. To change the feedback, choose the "Gamma" option. To add circles of a certain constant output reflection coefficient, choose the "S22" option. To add a specific constant output resistance circle to the display, choose the "Res" option. The "All Phases" option displays a circle on the Γ_t plane that is the locus of $\Gamma_0\rightarrow\infty$ at all phases of Γ_3. *RF*

This program is available on disk from the RF Design Software Service. *See page 90 for ordering information.*

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About the Author

Theodore Grosch received BSEE and MSEE degrees from Pennsylvania State University. He has worked at Hughes Aircraft, General Electric Astrospace, and was president of SSDD Research Corporation. He expects to receive his Ph.D. from Penn State this May. He can be reached at 1014 S. Garner St., State College, PA, 16801, or by phone at (814) 237-1325.

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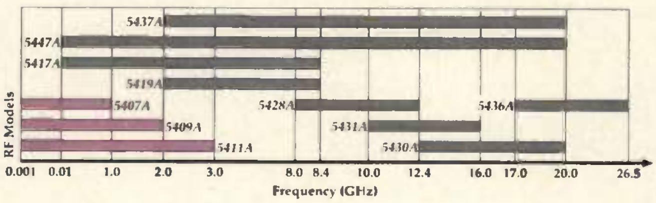

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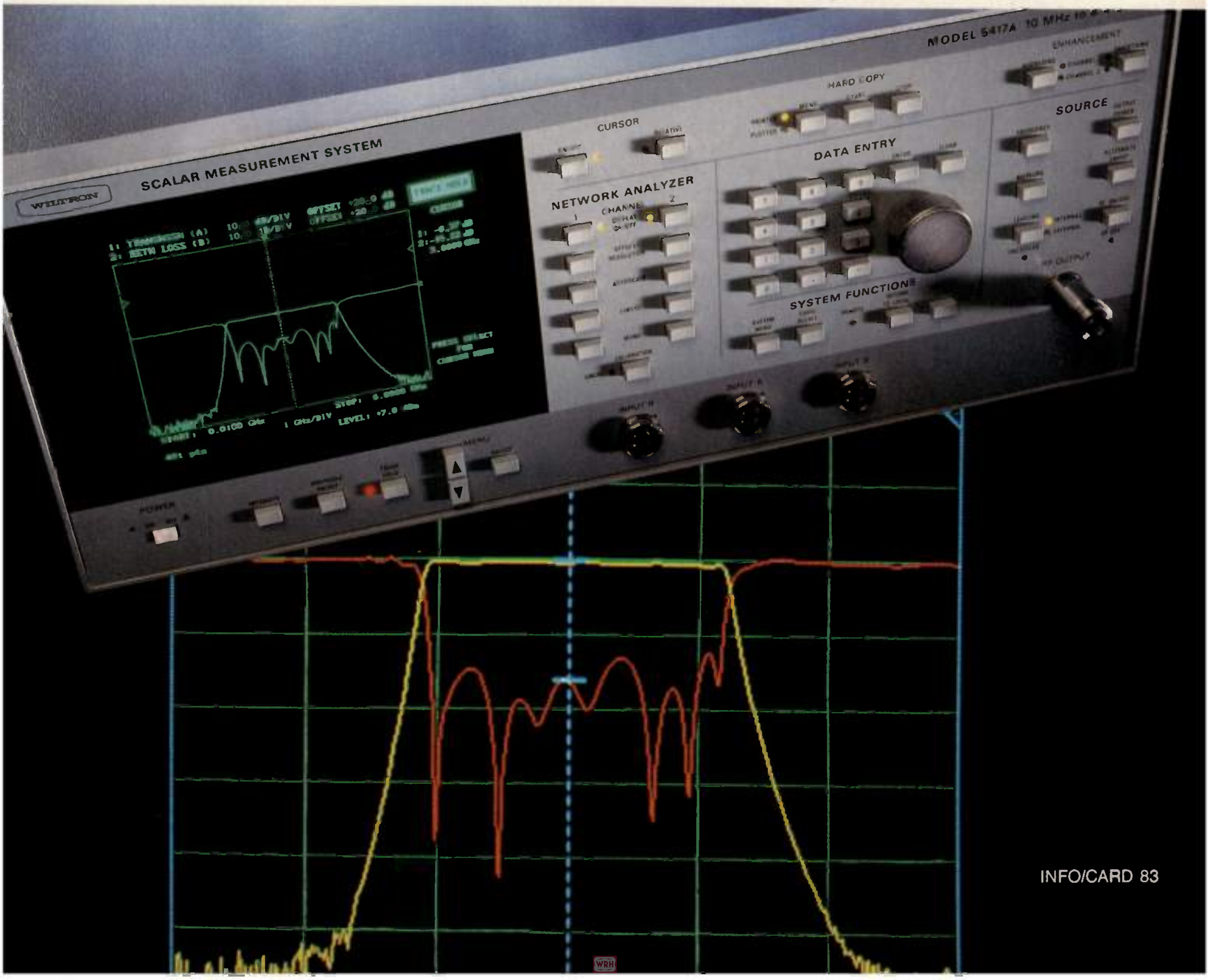

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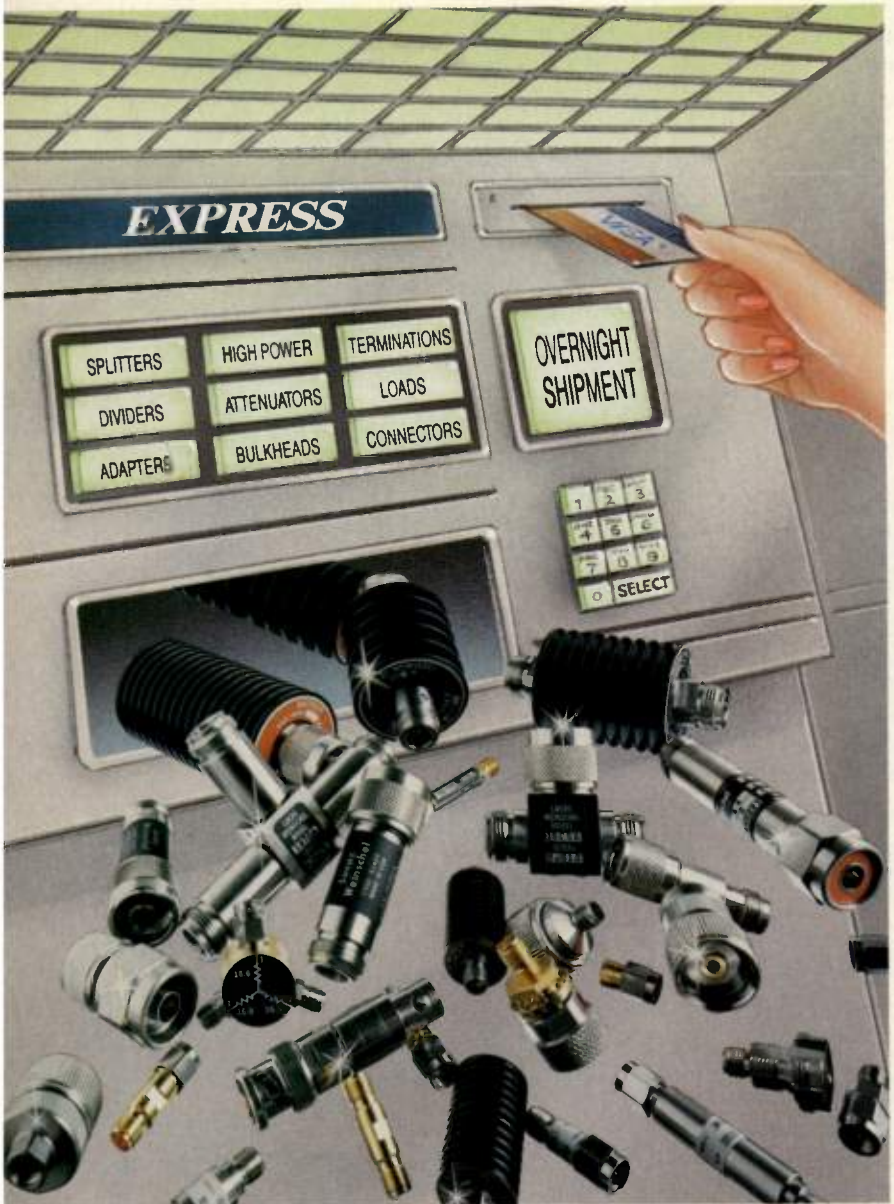

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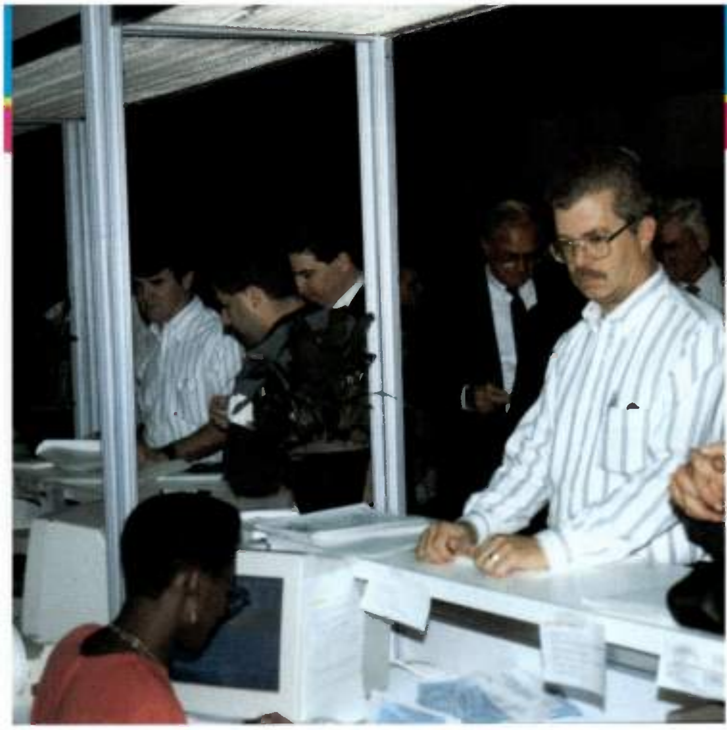

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8:30-11:30 AM	CELLULAR RADIO "A Single Chip Radio Transceiver for 900 MHz GSM Handsets, Mobile Terminals and Base Stations" "Cellular Repeaters and/or Cellular Microcells" "Use of Sigma-Delta Technology to ease Adjacent Channel Rejection for Mobile Radio Systems"	ANTENNAS AND TRANSMISSION LINES "Antenna Field Design for (VHF/UHF) Hand Held and Body Worn Applications" "Ferrite Rod Antennas for AM Broadcast Receivers" "New Generation of RF Transmission Line"	CIRCUIT ANALYSIS TUTORIAL - Part I "Driving Point Impedance Circuit Analysis Techniques"
1:30-4:30 PM	POWER AMPLIFIER DESIGN "Use EMTP (ATP) to Understand RF Power Amplifiers" "Tradeoffs in Practical Design of Class E High-Efficiency RF Power Amplifiers" "1 kW PWM RF Power Generator for the HF Band"	RF SYSTEMS "Wireless Communications using Infrared Techniques" "A Medical Radio Frequency Network for the Aged" "Miniaturization Techniques in Receiver Design"	CIRCUIT ANALYSIS TUTORIAL - Part II "Driving Point Impedance Circuit Analysis Techniques" (continued)

THURSDAY, MARCH 18 — Exhibit Hall Open 10am - 6pm

8:30-11:30 AM	HF/VHF POWER AMPLIFIERS "HF power amplifier operates in both class B and class D" "Class-E Power Amplifier Delivers 24 W at 27 MHz at 89-92% Efficiency, Using One $1.05 Transistor" "A New Method of Input/Output Impedance Measurement for Class C/D Amplifiers Using a Spectrum Analyzer"	OSCILLATORS "Modulating Saw Oscillators" "Practical Methods of Computer Simulation of Very High Q Oscillators Using SPICE and LIBRA"	ELECTROMAGNETIC MODELING "Simplified Calculation of Planar Antenna Field Perturbation by Integrated Active Feed"
1:30-4:30 PM	UHF AND L-BAND POWER AMPLIFIERS "Transistors for High Power Linear Amplifiers in the 1.5 - 2.0 GHz" "100-450 MHz 250 Watt Power Amplifier"	DIGITAL TRANSMISSION SYSTEMS "The Development of X-Band SSPAs and QPSK Modulators for Remote Sensing Satellite Applications" "Simulating Common Impairments Found in RF Digital Communications Systems"	COMPONENT APPLICATIONS "RF Circuits - The ASIC Way" "Tiny Transceiver, Wireless Microphone, and Superhet Receiver"

FRIDAY, MARCH 19 — Exhibit Hall Open 10am - 1pm

8:30-11:30 AM	PERSONAL COMMUNICATIONS "A Fully Integrated Modulator/Demodulator for DECT" "Vector Modulator IC's for use in Wireless Communication"	FREQUENCY SYNTHESIS "Wideband Synthesizer Design" "Operating Principle and Applications of a Frequency Independent Phase Tracking Circuit"	COMPUTER METHODS "Knowledge Systems: The Fast Track to RF Design and Testing Solutions" "Nonlinear Noise Analysis for Wireless Communication Systems" "The Design of Software for Automated RF Testing"

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Part I: Fundamental Concepts (CEUs Available) **March 17**
This new course presents the fundamental concepts of RF systems, components, transmission lines and impedance matching. RF system concepts of gain, bandwidth, linear and nonlinear circuits and amplifier classes are followed by the network concepts of resonance, bandwidth, Q and maximum power transfer. Practical components, their models, characterization and specification are then discussed. Transmission line theory, traveling wave behavior, the Smith chart, and practical realization of transmission lines are discussed, covering the concepts of impedance, loss, dispersion and physical construction. The final topic is impedance transformation networks, including lossless and lossy elements, graphical and analytical design, phase shift and bandwidth considerations, balanced networks and attenuators. Instructors are Dr. Robert K .Feeney and Dr. David R. Hertling of the Georgia Tech School of Electrical Engineering.

Introduction to RF Circuit Design
Part II: Active Circuit Design (CEUs Available) **March 18**
Active circuits are covered in this course, from basic models to advanced computer-aided design methods, from simple two-port concepts to large signal amplifiers and oscillator design. The two- port model includes defi nition of power gain, noise, cascaded elements and termination: Scattering parameters used to describe two-ports are introduced, includ ing their definition and physical meanings. Computer-aided design an analysis using SPICE and commercial software is described, noting mod els, analysis techniques and optimization methods. S parameter base design of amplifiers is presented, followed by large signal amplifie design and the large signal concepts of dynamic range, intermodulatic distortion and load pull. Non-reciprocal networks (couplers, combine and hybrids) are introduced, and the course concludes with oscillat design fundamentals. Instructors are Dr. Robert K .Feeney and Dr. Dav R. Hertling of the Georgia Tech School of Electrical Engineering.

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Learn the fundamentals of oscillator design. Historically, oscillator desig has been obscured with pages of equations for particular configuration In this course, basic concepts are applied to design various oscillato using a unified approach. Attendees learn how to evaluate oscillat designs accurately. L-C distributed element, SAW and crystal oscillato are studied. Also considered are output level, starting time, harmonic le els and phase noise performance. Instructor: Randy Rhea, Eagleware.

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Log Mac: Frequency and Time Domain Analysis

By Jeff Crawford
The Aerospace Corporation

Frequently during the process of designing a circuit, the circuit's frequency response is of importance, and occasionally, the respective impulse or unit step time response is also of interest. Log Mac calculates these responses for any network which can be described as a transfer function in the Laplace domain (s domain). Log Mac can be used to investigate the frequency response of passive or active filters and phase locked loops, and additionally, determine settling times and transient responses of these same kind of systems. Perhaps the most satisfying feature of the program is its full compliance with the MacIntosh Toolbox, featuring multiple resizable windows, scrollable text output, and many user-friendly dialogue boxes to make program operation a breeze. The minimum requirement for use of the program is a later generation MacPlus. A compiled version of the program, supporting an on-board mathematics coprocessor, is also available.

Log Mac performs the traditional Bode calculations, determining magnitude and phase of the rational s polynomial entered by the user. The data calculated is available in a scrollable text window and can be plotted simultaneously in another graphics window. Automatic graph scaling is incorporated to give clean plots in linear or logarithmic format. The number of significant digits on the X and Y axes as well as font size are automatically adjusted to maintain a clearly readable output. A digital readout of cursor position is also selectable by the user to allow quick reading from the graph in numeric form. A robust root solver based upon Moore's technique is resident within Log Mac. At any time it is possible to execute this function and determine the numerator and denominator roots, accompanied by the residual error obtained upon resubstitution of each root into the respective numerator or denominator. The Moore method implemented here has been modified for improved accuracy in cases of multiple identical roots; more will be said about this later.

Log Mac also performs the more interesting and difficult part of evaluating the inverse Laplace transform of the entered T(s) function. While a number of different techniques are available to perform this calculation, the method incorporated within Log Mac is that of partial fraction decomposition. A user may select from the appropriate dialogue box Impulse or Unit Step response analysis. Once again, tabular as well as graphical output of results is available to the user. Text and graphics data in both the frequency and time domain calculations can be saved to disk for later use.

Example of a Frequency Response Calculation

By way of illustration, let us consider the transfer function of a bandpass filter. Consider the transfer function of the bandpass filter in equation 1 below.

$$T(s) = \frac{S^4 + 2.12\times10^8 S^2 + 3.89\times10^{14}}{S^4 + 4260 S^3 + 5.48\times10^7 S^2 + 8.43\times10^{10} S + 3.89\times10^{14}} \qquad (1)$$

Figure 1 illustrates the coefficient entry dialogue in completed form, ready to execute the frequency domain analysis. Before execution, various attributes such as automatic scaling, number of X or Y divisions, graph titles, etc. may be selected by the user. Figure 2 graphically displays the frequency response of T(s), while Figure 3 is the tabulated test output which can be routed to a scrollable text window. Figure 4 shows the results of the root calculations for both numerator and denominator with their associated residual errors to indicate a measure of goodness for each root determined.

The Frequency Domain Calculations

The calculations performed in this part of Log Mac are rather straightforward and involve only normal complex number operations. At each frequency point, the magnitude and respective angle of T(s) are determined. A further description of the root calculation is in order due to its importance to the partial fraction decomposition method incorporated in the time response calculation to be discussed shortly.

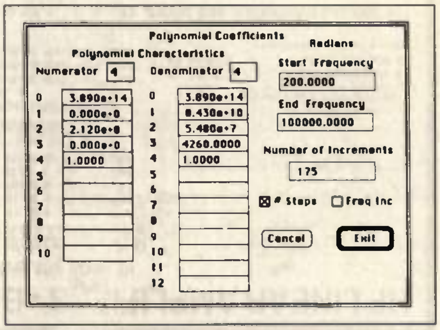

Figure 1. Coefficient entry.

Log Mac's Moore method is a derivative-based algorithm with enhancements for multiple roots. This algorithm, stated mathematically, works to determine the n values of z which make the following polynomial identically equal to zero.

$$f(z) = \sum_{k=0}^{n} (a_k + j b_k) z^k = 0 \qquad (2)$$

$$z = x + jy; \quad f(z) = u + jv$$

The Moore method adjusts the components of z = x + jy until the squared magnitude of f(z) = u + jv is zero at $z = z_i$. The root factor $(z - z_i)$ is then removed from the polynomial by synthetic division and the process repeated until completion. The method is contingent upon the information available through the use of the Cauchy-Riemann equations in equation 3. Incorporating equation 3 into the partial derivative expressions $\partial F/\partial x$ and $\partial F/\partial y$ gives the next

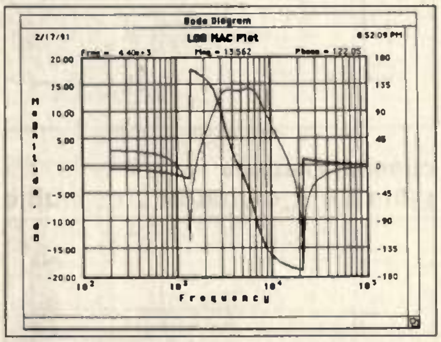

Figure 2. Frequency response.

Log Mac Tabulation		
Log Magnitude / Phase Response Results		
ProgramMACtice™ Vers. 1.0		
Frequency, rad/sec	Magnitude, dB	Angle
200.000	8.8810	3.5097
272.884	8.8886	2.0128
372.329	8.8953	0.7079
508.014	8.9036	-0.5323
693.145	8.9170	-1.8327
945.742	8.9408	-3.3322
1290.390	8.9848	-5.2095
1760.636	9.0672	-7.7348
2402.249	9.2231	-11.3967
3277.680	9.5215	-17.2809
4472.136	10.0887	-28.4595
6101.878	10.9247	-55.8070
8325.532	9.6746	-112.3138
11359.534	5.2244	-146.9424
15499.190	-0.2976	-160.4150
21147.425	-12.3915	12.9241
28853.998	-2.8495	8.9788
39369.010	-1.2588	6.3995
53715.918	-0.6211	4.6215
73291.144	-0.3200	3.3606
100000.000	-0.1683	2.4527

Figure 3. Text output after analysis.

step in the algorithm. The expressions in equation 5 are the same as the steps taken in the familiar Newton-Raphson technique. Further details of the method are available in references 1 and 2.

An enhanced Moore method was

$$\frac{\partial u}{\partial x}=\frac{\partial v}{\partial y},\quad \frac{\partial v}{\partial x}=-\frac{\partial u}{\partial y} \qquad (3)$$

$$F=u^2+v^2 \qquad (4)$$

$$\frac{\partial F}{\partial x}=2\left(u\frac{\partial u}{\partial x}+v\frac{\partial v}{\partial y}\right)$$

and

$$\frac{\partial F}{\partial y}=2\left(u\frac{\partial u}{\partial y}+v\frac{\partial v}{\partial y}\right)=2\left(-u\frac{\partial v}{\partial x}+v\frac{\partial u}{\partial x}\right)$$

$$\Delta x=-0.5\frac{\partial F/\partial x}{|f'|^2};\quad \Delta y=-0.5\frac{\partial F/\partial y}{|f'|^2} \qquad (5)$$

$$\text{where } |f'|^2=\left(\frac{\partial u}{\partial x}\right)^2+\left(\frac{\partial v}{\partial x}\right)^2$$

utilized in Log Mac that enables more accurate root extraction for cases involving multiple roots. In cases of multiple roots, the accuracy of the root location can be significantly improved by an averaging process among those roots appearing to be the same, however corrupted by numerical roundoff and associated effects. Consider the case below in which the following roots have been determined:

$R1 = 0.9999 + j0.9995$
$R2 = 1.000021 - j10.00076$

Making the required adjustments to these roots requires an averaging process among each set of roots which are seen to be identical. The operation

Log Mac Tabulation			
Zeros of the Numerator		Residual Errors: Real, Imag	
0.000000	-0.503070	1.6263033e-19	0.0000000e+0
0.000000	0.503070	1.6263033e-19	0.0000000e+0
-0.000000	-1.087234	1.0842022e-19	1.6444769e-18
-0.000000	1.087234	1.0842022e-19	-1.6444769e-18
Poles of the Denominator		Residual Errors: Real, Imag	
-0.000000	0.000000	-1.2686547e-16	5.5019129e-35
0.000000	-1.421741	2.0160348e-16	-1.7780916e-16
0.000000	1.421741	2.0160348e-16	1.7780916e-16
-0.000000	-0.881676	-4.7378266e-19	-6.5052130e-19
-0.000000	0.881676	-4.7378266e-19	6.5052130e-19

Figure 4. Pole and zero locations.

depicted in expression 6 is performed automatically each time multiple root locations of multiplicity N are detected.

$$\sum_{n=1}^{N}\frac{1}{N}\left(Re_n+jIm_n\right) \qquad (6)$$

Calculation of the Time Response

Before Log Mac can be launched the user must ensure that the T(s) is entered and in proper form; i.e. the numerator order is less than the order of the denominator. Internally, Log Mac takes the inverse Laplace transform of T(s) directly when the impulse response is requested. However, when the unit step response is selected, Log Mac automatically (hidden from the user) increases the order of each s-term by one in the denominator. For the polynomial T(s) entered earlier in the frequency response example, simple long division must be performed before doing an impulse response calculation; this is shown in equation 7. Should only the unit step response be desired, it is seen that multiplying T(s) by 1/s (which is done automatically by the program internally) puts it into proper form for the time domain calculation, therefore the side operation of long division is not necessary. This subtle point can be confusing in cases where T(s) is bordering on proper form.

The coefficients of the rational function are entered as seen earlier.

In the particular example considered

$$T(s)=1+\frac{-4260\,S^3+1.572\times10^8\,S^2-8.43\times10^{10}\,S}{S^4+4260\,S^3+5.48\times10^7\,S^2+8.43\times10^{10}\,S+3.89\times10^{14}} \qquad (7)$$

here, the great distance between pole locations requires the user to enter an allowable error in the roots of approximately 1×10^{-4}, otherwise the Moore root finder will give a nonconvergance error. After launching the impulse response calculation the following results are found as depicted in Figure 5.

A feature which is especially attractive to students of electrical engineering is the fact that Log Mac will write out the time domain expression. The impulse time response for the bandpass filter is given by equation 6. An identical analysis of the filter for its unit step response can also be performed.

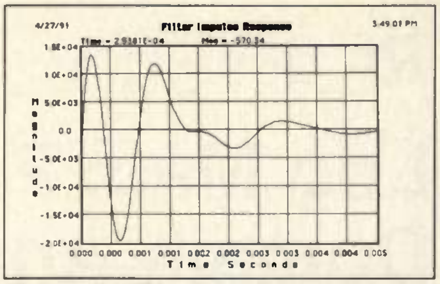

Figure 5. Filter impulse response.

Details of the Time Domain Calculation

As mentioned earlier, a number of methods are available in the literature for the time domain calculation. Some of these methods include the partial fraction technique, Pade'(3), Gaver- Stehfest(4) and Corrington(5) techniques.

The Pade' method makes an approximation of e^{st} which is in the Laplace inversion integral of equation 8.

The approximation is made only once and is represented by a truncated power

$$x(t)=\frac{1}{j2\pi}\int_{c-j\infty}^{c+j\infty}X(s)\,e^{st}ds \qquad (8)$$

series in which the first M + N + 1 terms are identical to the Taylor expansion of e^{st}, with the following terms slightly modified. The accuracy of the entire method is contingent upon the complexity of the approximating function $R_{N,M}$ (N = order of denominator, M = order of numerator), and the accuracy of the pole calculations used to formulate R. The Pade' method is suitable for the calculation of responses to nonperiodic excitations, but fails to satisfactorily solve those problems in which a periodic excitation is used. The validity for increasingly larger time intervals can be increased by using higher M and N, however, the accuracy is increasingly diminished by the integrity of the poles used in the Pade' approximation as M and N grow large.

The Gaver-Stehfest technique relies on the evaluation of the expected value of f(t) through the use of a three parameter exponential probability density function. Further details of this algorithm will not be presented here, primarily because functions which have oscillatory inverses cause considerable problems for this method. Another method which will only be mentioned is the Corrington method. This technique is able not only to calculate the inverse Laplace transform, but furthermore, one of its intermediate steps gives the Z-transform coefficients for the rational s polynomial. This method is not dependent on accurate pole determination, however, greater accuracy requires SVD (singular value decomposition) methods or a similar method to solve the Toeplitz matrix involved in the algorithm.

The partial fraction method was se-

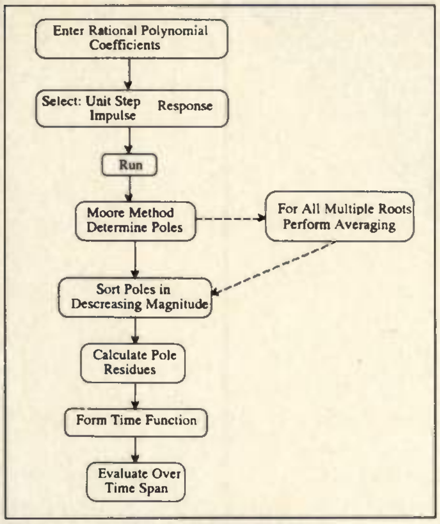

Figure 6. Algorithm flow.

lected for Log Mac because of its ability to give the user an explicit time function representation of a network described in the frequency domain by T(s). Log Mac, by virtue of its ability to write out the time domain function, is a great learning tool for the student as well as the practicing engineer. The overall algorithm flow is shown in Figure 6. The enhancements added to the Moore method for multiple roots are necessary to maintain accuracy for the Laplace inversion method since any repeated roots contribute additional terms of the form t^{n-1}. Summarizing the partial fraction method, the types of terms which can appear include the following:

Single real roots:

$$\frac{K_1}{s+\alpha} \Rightarrow K_i e^{-\alpha t}$$

Multiple real roots:

$$K_1 \frac{t^2}{2} e^{-\alpha t} + K_2 t e^{-\alpha t} + K_3 e^{-\alpha t}$$

Pair of complex roots:

$$K = u + jv$$

$$\frac{K}{s+(\alpha+j\beta)} + \frac{K^t}{s+(\alpha-j\beta)}$$

$$\Rightarrow 2e^{-\alpha t}\left[u\cos(\beta t) + v\sin(\beta t)\right]$$

The case of multiple identical complex roots is not handled in Log Mac.

Log Mac is available for the Macintosh computer with or without the mathematics coprocessor support. The version of the program described here has been successfully used on the later generation Macintosh Plus and the Mac Classic, as well as Macintosh IIx, IIci, and IIsi. A version of the program for Windows may be made available sometime in the future; contact the author for further details. ***RF***

This program is available on disk from the RF Design Software Service. See page 90 for ordering information.

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About the Author

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Antenna and Processor Techniques for Detection of Manmade RF Noise

By F.N. Eddy
Independent Consultant

Present RFI detection gear is inadequate for assessing the vulnerability of high sensitivity receiving systems to local manmade noise. As a result, excess expense is incurred by over-cautious designers of sensitive RF receiving systems who, being unable to measure noise at levels that might interfere with their systems, locate their equipment farther away from interference sources than may actually be necessary.

Radiometric measurement techniques enable 20-40 dB greater man-made noise detection sensitivity than do more conventional RFI measurement procedures. This enhanced sensitivity can be used for finding quiet sites, for localizing interference sources or for providing warning of proximity to energized power lines.

Tuned-loop, electrically-small antenna arrays have been developed which yield a highly directional response without degraded noise performance, or which suppress distant noise sources. Such antennas are complementary to the radiometric processing scheme; they reduce the time required to perform an RFI sweep compared to more conventional procedures (of the general type outlined by Harris (1)).

Radiometric Processing

Most man-made RFI is envelope modulated at power-line or other fixed frequency rates. Computer-generated noise is generally observed at subharmonics of the system clock frequency. Television, radar and high-speed digital communication signals likewise exhibit periodic envelope modulation. Exploitation of this envelope modulation frequently enables greater detection sensitivity than conventional power spectral density measurements. It also provides a means for differentiating and localizing noise sources whose spectral density is even less than that of the irreducible receiver noise background, kTF (k = Boltzmann's constant, T = receiver temperature, and F = receiver noise figure).

To make this point clear, consider the specific example of HF noise radiated by AC power lines. Corona, insulator-breakdown, triac-switching, commutator-arcing and other noise sources of AC power-line noise are all envelope-modulated at harmonics of the fundamental AC line frequency (2). While the underlying physics of corona and other discharges is extremely complex (3), it is self-evident that there can be little RF current when the instantaneous line voltage is near zero.

The observed RF noise current is highly nonlinear; thus even three-phase power lines exhibit substantial modulation at the first several harmonics of the AC line frequency. The strongest frequency component is generally at the fundamental and second harmonic; typically at 180 or 240 Hz in a three-phase system within the United States or 150/200 Hz in Europe (see Reference 2).

The key element in the development of high-sensitivity radio-astronomy sky-mapping systems was R.H. Dicke's switching radiometer (4). By sequentially comparing the output of an antenna directed to some point on the celestial sphere with the output of a resistor at known temperature, it became possible to measure antenna noise temperatures much smaller than the noise temperature of the amplifier. Detection of stellar

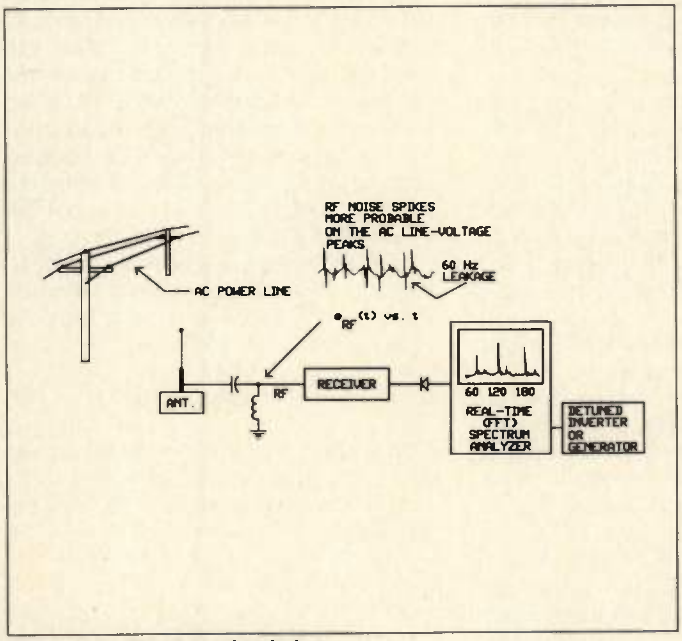

Figure 1. Basic principles.

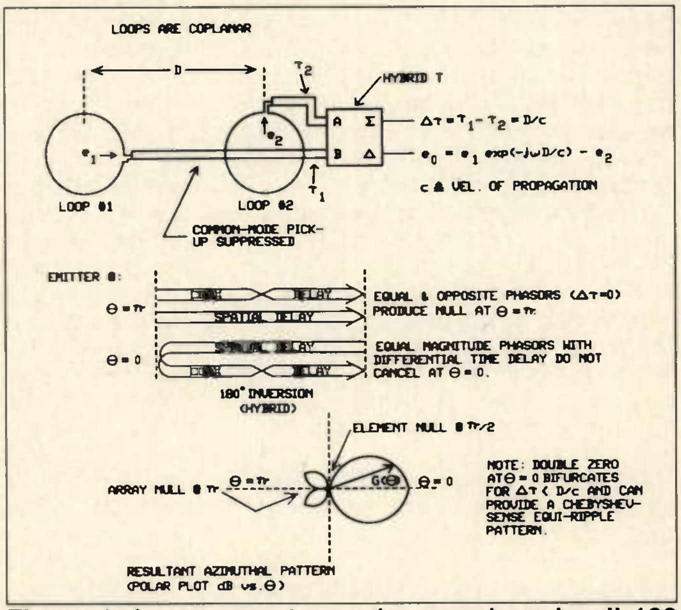

Figure 2. Loop array to produce peak and null 180 degrees apart.

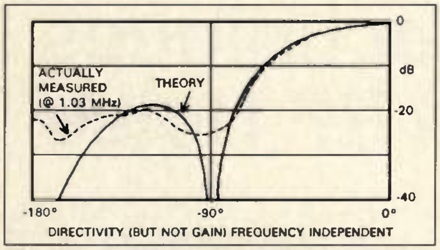

Figure 3. Representative two element response.

objects down to the primordial background temperature of approximately 3 Kelvins became feasible. Intuitively, this process is seen to work because the amplifier noise, unlike the desired signal, is uncorrelated with the switching frequency, and hence, non-coherent processing gains are obtained on the order of:

$$(WT)^{1/2},\ SNR \ll 1 \qquad (1)$$

where W is the RF bandwidth and T is the post-detection integration period. Variants on Dicke's radiometric scheme have been subsequently used to increase sensitivity and reduce DC drift in diverse applications.

The processing stratagem to be described here is essentially that of a Dicke radiometer in which the switching (modulation) is performed externally (and for free) by the power company. While the phase of the interference may sometimes not be known, its fundamental frequency is usually within a small fraction of a Hertz of the nominal line frequency. Thus, by comparing the noise energy in a given post-detection bandwidth at one or more modulation harmonics with that observed at other frequencies it is possible to provide an adaptive noise threshold for detecting envelope modulated interference 30 dB or more below ambient noise.

This general process is shown in Figure 1. If W is on the order of 10 MHz (practical in the low VHF spectrum) and T in the order of 1 sec, then the processing gain approaches 35 dB. At low HF frequencies, W is more typically limited to the order of 5-10 kHz, yielding processing gains approaching 20 dB, again for a nominal 1 second integration.

This large processing gain enables site noise surveys to be carried out using compact, low-gain, moderately directive antennas. For example, it was possible to determine the probable impact of the Soviet Over-the-Horizon Backscatter radar (known as the "woodpecker") on a proposed HF system in Alaska by using a simple Beverage antenna (nominal 17-dB directivity) to simulate the proposed 27 dBi receive array.

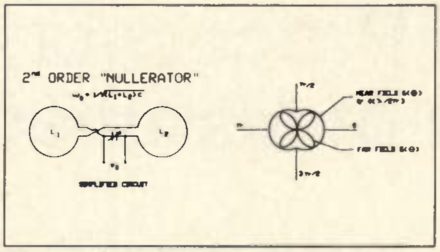

Figure 4. Nullerator response.

The increased radiometric processing gain can be traded off for greater line-of-sight detection range. Measurements carried out at over two-dozen U.S. sites suggest that under good line-of-sight propagation conditions, HF noise from 35 kV distribution circuits can be detected at 2 miles in open, flat terrain, using simple, electrically-short whip antennas. With 750 kV transmission lines, excess noise has been detected at 7-mile range.

Out-of-Phase Antenna Array Weighting

In the far-field an electrically-small loop provides peak response for sources in the plane of the loop as shown in Figure 2. The dual-lobed response shown can be converted into a single lobe response by summing loop outputs with a time-delay matched to the propagation delay in one direction. Use of two overlapping loops, spaced to minimize mutual coupling, results in what we have called the "Master-Card array."

As shown in Figure 2, a time delay between isolated elements $\Delta\tau = D/c$, introduces a perfect null in one direction along the array. Since the differential time delay is maximized in the reverse direction, a response peak is obtained 180 degrees away from the array null. Use of loops aligned with the array axis, in place of whips or dipoles, yields additional broadside nulls.

The basic technique can be generalized with multiple monopoles, dipoles, or loops having oscillating binomial Chebyshev, or similar weighting. Single capacitor tuning of the tuned multiloop array is feasible by appropriately interconnecting the loops. Modest structure sizes provide adequate receive-only SNR from VLF up through low VHF frequencies.

In single loop receive only applications, required loop size can be estimated using the Fano bound (least voltage reflection coefficient for infinite complexity matching network) for a series RL circuit (5).

$$\Gamma = \exp[-RQ/(2\,f_0\,L)] \qquad (2a)$$

$$\cong \exp[(-1.1E{-}5\,f_{MHz}\,b)^3\,Q] \qquad (2b)$$

Where $Q \gg 1$ ($Q = f/B$, B = bandwidth), R is the radiation resistance and L the inductance of the loop (using standard expressions for single turn loop inductance and resistance for loop radius b in meters, assuming wire diameter to be in the order b/100). In equation 2 it is presumed that loop loss resistance is small compared with radiation resistance.

The allowable mismatch loss is determined by two factors: the external noise factor, and the array loss. External noise below roughly 100 MHz is established by background galactic noise,

$$F = 10\log_{10}(10^5/f_{MHz}^{\,2.2}) \qquad (3)$$

Utilizing the first term in the Taylor-series expansions, $|\Gamma| = (1 - 1/F)^{.5} = 1 - 1/2F$, and $\ln(1-dx) = -dx$, it is possible to estimate minimum loop radius b as a function of operating frequency, and single-loop loaded Q; specifically, for the case of a single element, equations 2 and 3 yield

$$b = [4.66/(f_{MHz}^{\,.8}\,Q)]^{1/3})](m) \qquad (4)$$

For B = 10 kHz and f = 5 MHz, for example, the loop radius must exceed roughly .135 m (5.3 inches) for an external noise limited loop (i.e. tuned, receive only applications). As the HF noise seldom approaches the galactic limit even in quiet rural locations below 15 MHz, the bound is reasonably conservative. Note that greatest power-line interference/background noise is generally observed between 2 and 10 MHz, but dependent upon time of day, season, etc.

The far field array factor for the limiting case of an N-element oscillating binomial array (John Stone's sidelobe free array) is readily shown to be

$$A = 2N\sin^{(N-1)}\{(\pi D/\lambda)(1-\cos(\theta))\} \qquad (5)$$

Where θ is the angle between source and array axis. Thus, along the array axis in the mainlobe direction, and assuming $D \ll \lambda$,

$$A(0) \cong 2N\,(2\pi D/\lambda)^{(N-1)} \qquad (6)$$

Similar expressions are readily de-

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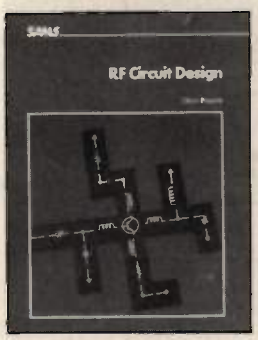

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rived for Chebyshev, Taylor and similar weights. Equation 6 or its dual for different weighting, is readily substituted into equation 3 to establish minimum loop size in receive only arrays. For N = 2 (the Master-Card array), the theoretical minimum loop spacing is given approximately by

$$D \cong 11/(f^{1.7}_{MHz}\, b^3\, Q)\ (m) \qquad (7)$$

Since D is approximately equal to 1.7 b, to minimize mutual coupling and resultant backlobe degradation,

$$b \cong 1.6/(f^{1.7}_{MHz}\, Q)^{1/4}\ (m) \qquad (8)$$

Actually measured and difficult to predict loop losses arising from transmission lines, tuning capacitors, etc., suggest that as much as a twenty percent margin should be added to this theoretical result.

If the loop element itself is presumed to have a cos(θ) response in the plane containing the loop axes, the N-element array, from equation 4, then has angular response

$$e = 2N\{\sin^{N-1}[(\pi D/\lambda)(1-\cos(\theta))]\cos(\theta)\} \qquad (9)$$

Representative two element response is shown in Figure 3 along with an experimentally observed MF pattern.

Mutual coupling and other error sources degrade attainable patterns. Using both brute force numerical solution and local expansion of elliptic integral functions, the mutual coupling was computed for coplanar loops having negligible thickness. Both the theoretical and actually observed mutual coupling between these coplanar loops exhibits a perfect null (> 60 dB) with partial overlap of about 1.7 b, mildly dependent upon loop thickness and lateral overlap.

While Figure 3 shows only the N = 2 case, similar techniques have been developed for minimizing mutual-coupling induced pattern errors in higher order arrays. The experimental measurements shown in Figure 3, were obtained with b = 0.4 and D = .68 m. Pattern measurements were made using local AM broadcast band sources. These antennas have been also been tested in the low VHF frequency range to verify the underlying theory.

The "Nullerator" Near-Field Array

A more careful analysis of the residual voltage difference between a single pair of anti-phased isotropic elements includes near field effects involving a common, range dependent factor

$$k\, e_D = [(2\pi\, j/\lambda) + (1/r)] \qquad (10)$$

where use has been made of the small angle approximations, sin(x) = x, and cos(x) = 1. Here, r is the range to a point source emitter, and j is the imaginary operator. The first term inside the square brackets is that due to far-field phase differential; the second is that due to amplitude difference. It is seen that the near field effects are large compared with the far field term for $r << \lambda/2\pi$.

This relationship is maintained for higher order arrays but only the second order (two loop) case is discussed here. Using the zero in the array factor to cancel the far-field mainlobe, it is then possible to form uniformly weighted arrays with minimum far field response, but with maximum near field response. These apparently novel "near field" tuned arrays are effective in detecting local noise sources at ranges less than one wavelength while simultaneously minimizing response from more distant sources as sketched in Figure 4.

Practical Implementation

While alternative configurations are feasible and in some applications may have advantages, the series connected second order "Nullerator" and "Master Card" magnetic loop configurations shown in Figure 5 provide a reasonable basis for initial design. Note that a Faraday shield is desirable to minimize frequency dependent electric field pickup in large loops, but that this shield must be arranged so as not to support parasitic currents. Winding sense can be reversed one loop to the next to provide the required sign alternation. Note that it is necessary to terminate both even and odd modes to achieve desired null depth.

Receiver/Processor

Receiver/processor design is straightforward, with a representative block diagram is shown in Figure 6. Elaborate IFs with wide dynamic range linear detectors have been designed, but in practice they have not provided any particular advantage over standard low cost, commercial IC IFs with AGC and simple detector circuits. As the result of its inherent nature this detector, no matter how linear, will generate in-band IMD products; n × f1 ± m × f2 beat notes of signals received.

Note further that the SNR response is nonlinear: for input SNR << 1, the output SNR is reduced 20 dB for each 10 dB reduction in SNR at the detector input. (In the region near SNR = 1, the response is more complex and depends upon whether the detector is operated in its linear or square law regime).

Both real time (FFT) digital spectrum analyzers and specially designed circuits have been used for synthesizing "envelope matched" filters. When attempting to detect modulation harmonically related to the power line frequency, it is almost mandatory that the receiver be operated from a DC power source or a generator with detuned governor, so that the system can distinguish local generator and power supply noise from the power line noise of interest.

For the detection of 60 Hz harmonics, crystal controlled divider chains have been used to provide variable post detection bandwidths ranging from 0.05 Hz up to 2 Hz. Bandwidth selection is governed primarily by required data rates. (It is probable that phase lock loops capturing the ambient 60 Hz E-field would provide at least comparable performance.) The outputs of two filters, typically tuned to 360 Hz (for three phase circuits) and 377 Hz are separately en-

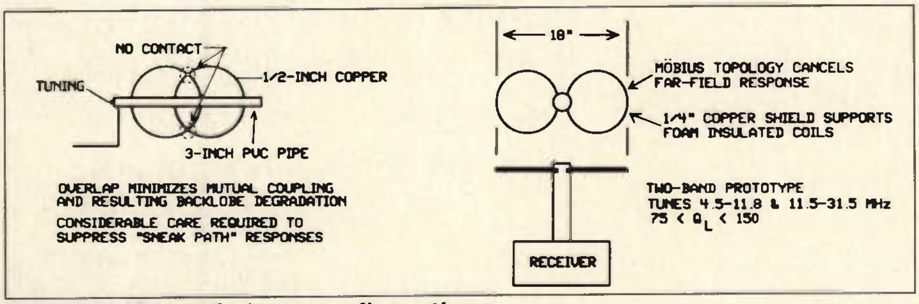


Figure 5. Magnetic loop configurations.

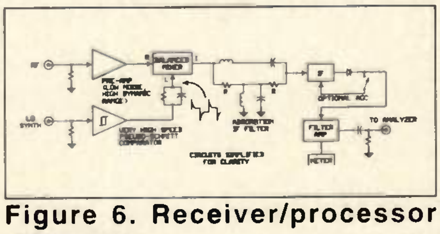
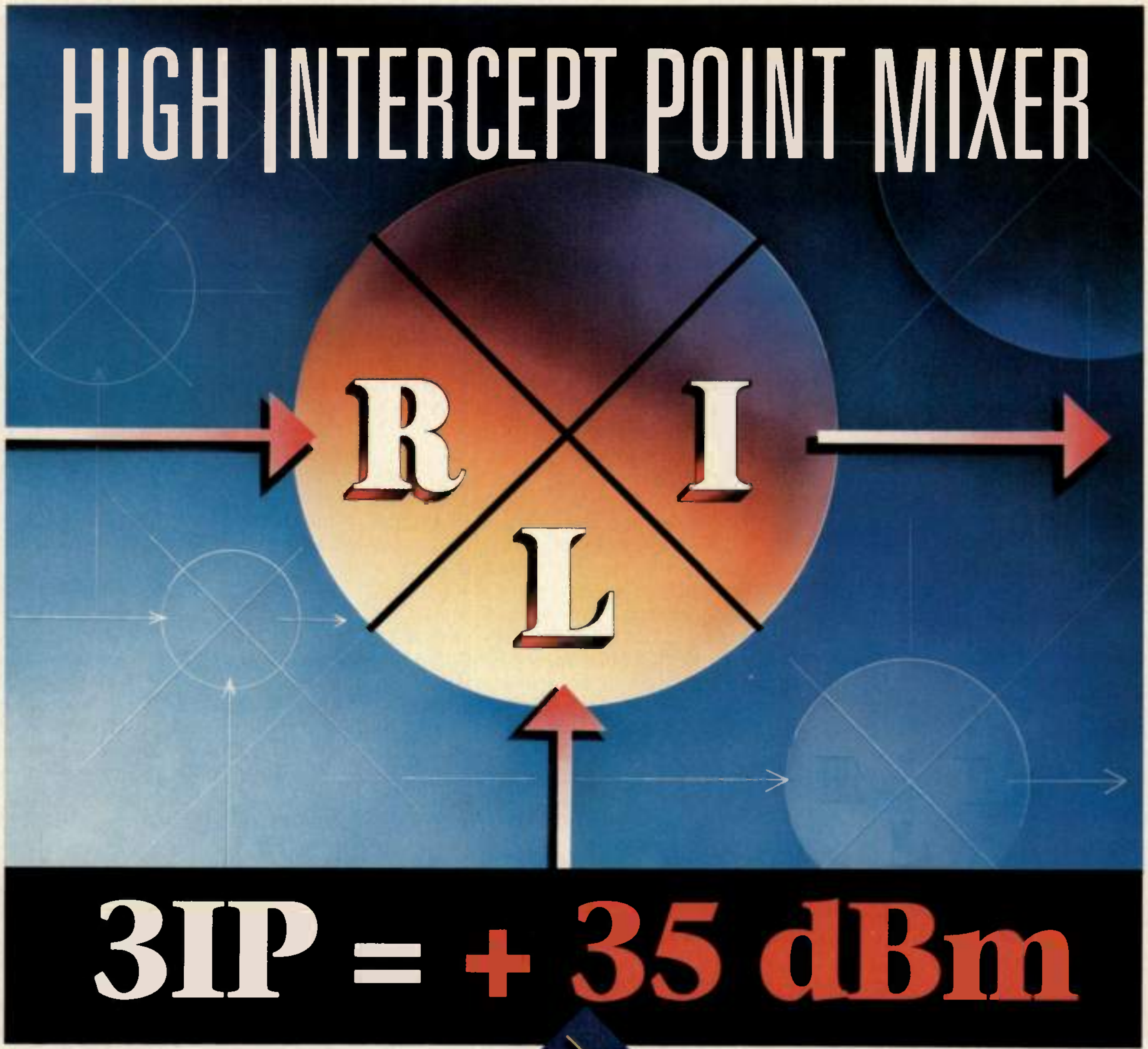
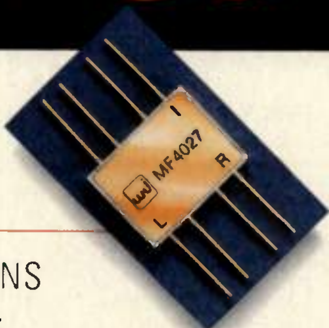


Figure 6. Receiver/processor schematic.





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velope detected. The difference can be applied to a logarithmic difference amplifier whose output is then proportional to the logarithm of the AC line harmonic to background noise level ratio.

Summary

Hand-held, directional HF antennas and high-sensitivity radiometric processing techniques have been described which enable the rapid, unambiguous detection of power line and other modulated noise sources. These techniques have been employed in a variety of applications over the past decade at frequencies ranging from roughly 100 kHz up above 100 MHz. These techniques have greatly increased attainable sensitivity in detecting manmade noise; they may additionally provide a practicable means for detecting and avoiding AC power lines. ***RF***

Acknowledgements

At MITRE, B.A. Dawson and R. Davis were instrumental in the early test and evaluation of the basic radiometric technique whose development was encouraged by R.B. Stevenson and Dr. G.L. Guttrich. "Woodpecker" interference measurements were made possible by Dr. J. Buchau (USAF AFGL) and Prof. R. Hunsucker (University of Alaska) and supported by R. Bush and N. Doherty of MITRE. S. Estes and his UAK team, B.A. Dawson and J.R. Walker (MITRE), and others too numerous to mention helped make it all work. G. Hagn (SRI International) has been extremely generous in sharing results of his own highly relevant work with us.

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2. W.R. Vincent, *Reference Data for Engineers*, 7th ed., 1985, E.C. Jordan, ed., Ch. 34, and Figure 10 in particular. Our own heretofore unpublished data extending back to 1983 exhibit considerable variability in the detailed noise structure. The corona noise (lines 35 kV and above) is highly impulsive — with pulse widths typically in the order of 10 usec. The noise envelope is quite "bursty" with discharges apparently regeneratively triggered by other nearby discharges. Below 35 kV the noise is more typically generated by the load and tends to be much less bursty. Highly regular switching spikes (likely from triac dimmers) are observed with comparable rise times, perhaps the result of bandwidth limitations of the distribution transformers.

3. See for example, Olsen and Wu, "Interference from Power Lines," *Radio Science*, 24 (3), pp. 340-250.

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7. It is often fallaciously presumed that loops are pure magnetic dipoles for diameters less than a fraction of a wavelength. In fact, for E/H response to be suppressed 20 dB, 2b/lambda must be less than approximately 0.01 (in the absence of shielding). See F. Horner, "Properties of Loop Aerials," *Wireless Eng.*, August 1948, pp. 254-259. Or Whiteside and King, PGAP, May 1964, pp. 291-297. If this E field response is not suppressed, the element pattern null will tend to be filled in.

About the Author

F.N. Eddy studied physics at Harvard while working full time repairing early television sets. After nearly ten years at LFE (Senior Scientist, Research Division), he spent over twenty-five years at MITRE (Division Scientist). He currently does independent consulting in a variety of fields. He can be reached at (617) 235- 8366.

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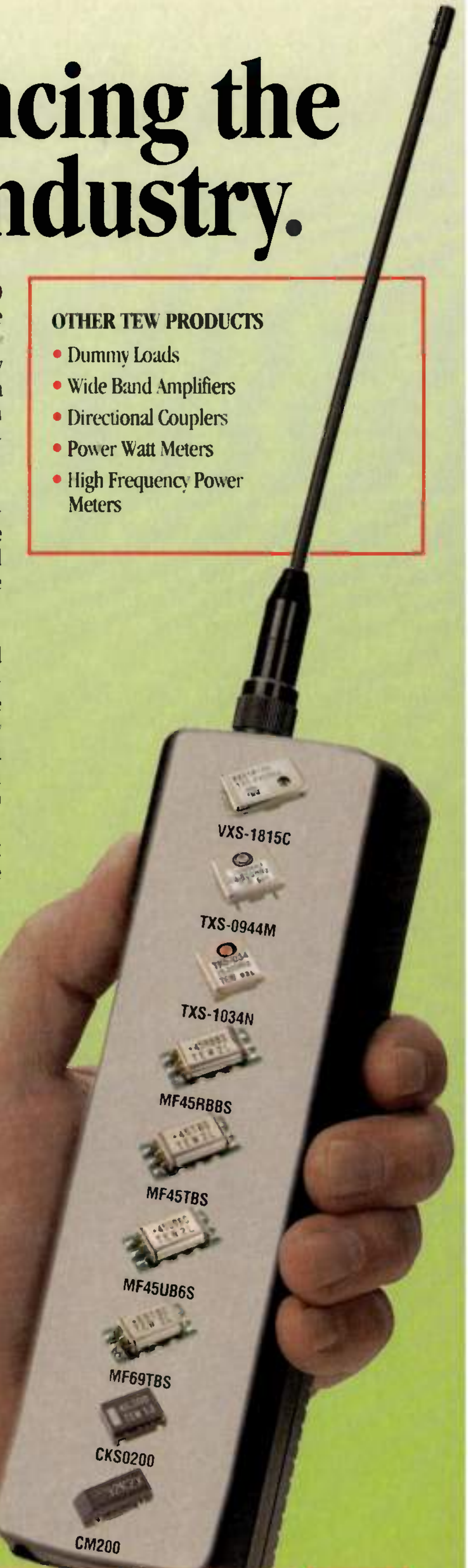

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Cross-Correlation Phase Noise Measurements

By Warren F. Walls
Femtosecond Systems

The typical method used to measure phase noise involves a single mixer that is driven by sources on its RF and LO ports. The IF port is then low pass filtered and amplified before it is connected to a spectrum analyzer or FFT (Figure 1). The noise floor of this technique is highly dependent upon the performance of the mixer, filter, and amplifiers (3-5).

The cross-correlation measurement technique uses two duplicate systems to drive a cross-correlation analyzer. The oscillator signals are split with reactive splitters to provide a total of 2 pairs of input signals. These signals are connected to the LO and RF ports of both double balanced mixers. The output of each mixer has its own low pass filter followed by a set of amplifiers. The output of both channels is then fed into the cross-correlation FFT analyzer (Figure 2). The random noise contributed by the individual mixer-filter-amplifier strings is uncorrelated. The correlated noise is the actual noise of the input oscillators plus some leakage due to the imperfect isolation characteristics of the power splitters. The cross-correlation measurement technique increases the time required to perform a measurement, but the noise floor is lowered due to the cancellation of the incoherent noise of the individual detector systems.

Equipment Used

A variety of off-the-shelf equipment was used to perform the tests discussed in this paper. Two low noise oscillators, one of which had a voltage tune port, were used to drive the mixers for these tests. Three power splitters from Mini-Circuits (ZFSC2-4) were used to split the oscillator signals. The system was calibrated (K_d measured) using a typical analog Tektronics oscilloscope. The actual noise measurements were made on an HP cross-correlation FFT analyzer (35665A) and hard copies were made using an HP plotter. Finally, some care was used in selecting the particular cables and connectors appropriate for low noise measurements.

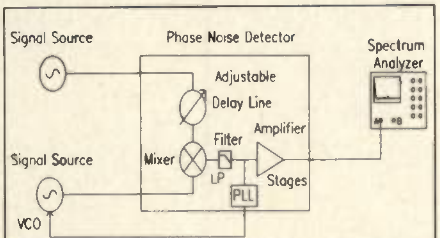

Figure 1. Single channel phase noise test set.

Single Mixer Phase Noise Measurement

The single channel method is similar to a DSB receiver. The outputs of two oscillators are fed into the RF and LO ports of a double balanced mixer. The IF port of the mixer is then low pass filtered and amplified. After the low pass filter there is also a phase locked loop which drives the VCO (voltage controlled oscillator) port of one of the oscillators. This is required in order to have the two oscillators track one another 90 degrees out of phase. The output of the amplifiers then drives a spectrum analyzer (Figure 1). The noise floor of this technique is highly dependent upon the performance of the mixer, filter, and amplifiers (3-5). Initially the oscillators are tuned so that they are within a few thousand hertz of each other. The beat frequency that results is displayed on an oscilloscope. On the oscilloscope one can get a curve that relates a change in voltage to a change in radians. This is used to calibrate the sensitivity of the mixer. The calibration constant, K_d, is expressed in equation 1.

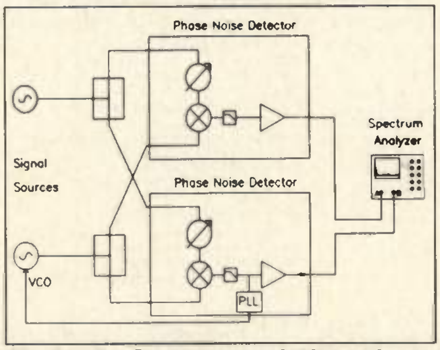

Figure 2. Cross-correlation phase noise test set.

$$K_d = \frac{\Delta V}{\Delta T} \frac{\text{Period}}{2\pi} \frac{1}{\text{Gain}} \tag{1}$$

where:

ΔV = Vertical voltage change on scope
ΔT = Change in time on scope to cause ΔV
Period = Number of horizontal divisions for 1 period
Gain = Gain of amplifiers on output of mixer

Once the sensitivity of the mixer is determined, the two oscillators are matched in frequency by activating the phase locked loop. The mixer now has two signals 90 degrees out of phase from each other. Once the oscillators are locked, a particular frequency range of interest can be viewed on the spectrum analyzer. In order to report the actual value of the phase noise for some offset from the carrier, several parameters need to be taken into account. One typically reports phase noise as spectral density of phase noise, $S\phi(f)$, at some offset from the carrier as shown in equation 2.

$$S_\phi(f) = \left(\frac{V_n^2(f)}{BW}\right)\left(\frac{1}{K_d^2(f)}\right)\left(\frac{1}{\text{Gain}^2(f)}\right) \tag{2}$$

where:

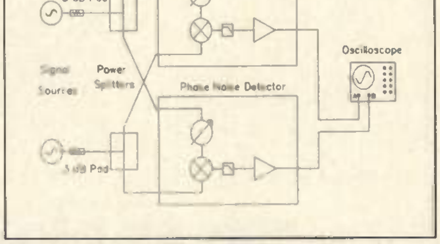

Figure 3. Calibrating cross-correlation.

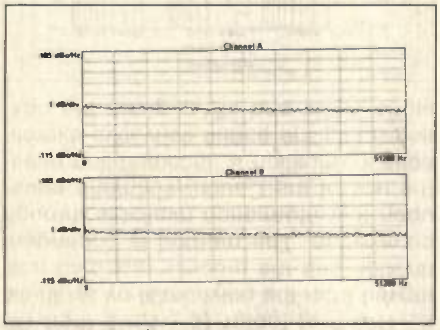

Figure 4. Comparison of single channel systems.

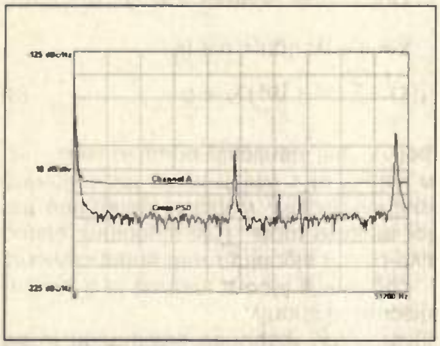

Figure 6. Phase noise floor (0-51.2 kHz).

$V_n(f)$	=	Noise voltage viewed on analyzer
BW	=	Noise bandwidth correction to 1 Hz BW
$K_d(f)$	=	Mixer sensitivity
Gain(f)	=	Gain of amplifiers between mixer and analyzer

This reports the double sideband phase noise of both oscillators. If one assumes equal contribution of noise from both sources, just subtract 3 dB to get the phase noise of one oscillator. The single sideband phase noise of one oscillator is achieved when another 3 dB

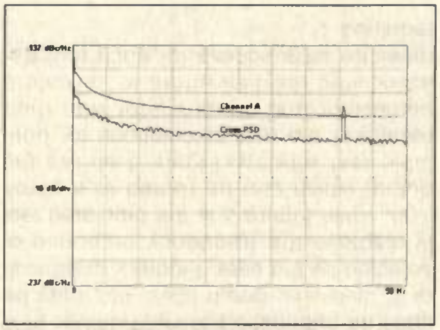

Figure 5. Phase noise floor (0-50 Hz).

is subtracted.

Cross-Correlation System (Dual Mixer Phase Noise Measurement)

The cross-correlation measurement technique uses two duplicate single channel systems like the one in Figure 1. The output of the two detectors drive the 'A' and 'B' inputs of a cross-correlation spectrum analyzer. The oscillator signals are split with reactive splitters to provide a total of 2 pairs of input signals. These signals are connected to the LO and RF ports of both double balanced mixers. The output of each mixer has its own low pass filter followed by a set of amplifiers. The output of both channels is then fed into the cross-correlation FFT analyzer (Figure 2). The random noise contributed by the individual mixer-filter-amplifier strings is uncorrelated. The correlated noise is the actual noise of the input oscillators plus some leakage due to the imperfect isolation characteristics of the power splitters.

Just as with the single channel system, the mixer sensitivity must be calibrated. This is accomplished by allowing the two oscillators to beat at a couple of thousand hertz or less (Figure 3). The difference in frequency between the two signals produces a "beat note" or "beat frequency". The beat frequency is viewed on the oscilloscope and the horizontal change in radians is measured for a specific voltage change. The ratio in volts/radians is the value of K_d (3). This calibration must be conducted on both of the single channel systems and then the combined mixer sensitivity is calculated in equation 3.

$$K_d = \sqrt{K_1 K_2} \tag{3}$$

The two single channel systems that were used for these tests were compared by modulating one oscillator with noise and viewing the individual phase noise detector outputs on the FFT. Notice the relative flatness and levels of the two systems in Figure 4.

The actual measurement of phase noise is very similar to the procedure for the single channel system. The oscillators need to be locked using the phase locked loop and then the analyzer is set to the frequency range of interest. The numbers from the analyzer need to be plugged into equation 2 along with the combined mixer sensitivity which was calculated using equation 1 and equation 3. The big difference between the two systems shows up in the improvement of noise floor due to the cancellation of the non-coherent noise of the individual phase noise detectors. The noise floor improvement of the cross correlation system versus the single channel system is illustrated in Figure 5 and Figure 6.

Measurement Uncertainty

The improved noise floor of the cross-correlation phase noise measurement technique does not come without a price. Many more samples are required in order to average out the uncorrelated noise. The confidence interval of a single channel phase noise detector is (6):

$$S^s_\phi(f) = S^m_\phi(f)\left(1 \pm \frac{1}{\sqrt{n}}\right) \tag{4}$$

The confidence interval of a dual channel (cross-correlation measurement system) is:

$$S^x_\phi(f) = S^m_\phi(f)\left(1 \pm \frac{2S^s_\phi}{S^m_\phi(f)\sqrt{n}}\right) \tag{5}$$

where indices:

x = cross correlation

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m = measured (noise)
s = single channel
and,
n = number of samples

Equation 4 shows that for a single channel the confidence interval is ±10 percent for 100 samples. Equation 5 shows that to obtain the same confidence interval for a phase noise measurement 10 dB below the single channel noise floor, 20,000 samples are required.

Conclusion

The dual channel or cross-correlation method of phase noise results in a lower floor than the standard single channel method. This does, however, come at a cost of measurement speed. A great deal more averages are required to achieve the same level of confidence in a measurement. The lower noise floor that is now achievable using the cross-correlation method provides a level of characterization of extremely good oscillators that was not available using the single channel method. *RF*

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About the Author

Warren F. Walls is the president of Femtosecond Systems in Columbia, Maryland. Over the past nine years, his company has manufactured and marketed phase noise, amplitude noise, and Allan variance measurement equipment. He also had the opportunity to work for an RF solutions firm, Erbtec Engineering in Boulder, Colorado, for a year and a half. He can be reached at Femtosecond Systems; P.O. Box 6005; Columbia, MD 21045-8005; or by phone at (410) 740- 1427.

A Wideband Monolithic, Low-noise, Front-end Amplifier

By Scott L. Williams
Tektronix, Microelectronics Product Line

In order to reduce the cost, size and power consumption of telecommunication systems, it is becoming desirable to integrate discrete RF circuitry on high performance ASICs. Circuit blocks such as low-noise amplifiers (LNAs), mixers, voltage-controlled oscillators (VCOs), automatic gain control (AGC) and base band amplifiers have all been successfully integrated in silicon bipolar processes such as Tektronix' SHPi. SHPi features NPN transistors with an f_{max} of 9 GHz, P-channel JFETs with $g_m/2\pi C_{iss}$ of 600 MHz, vanadium Schottky diodes, two types of implanted resistors (82 and 1350 W/sq), and 50 W/sq thin film nichrome resistors.

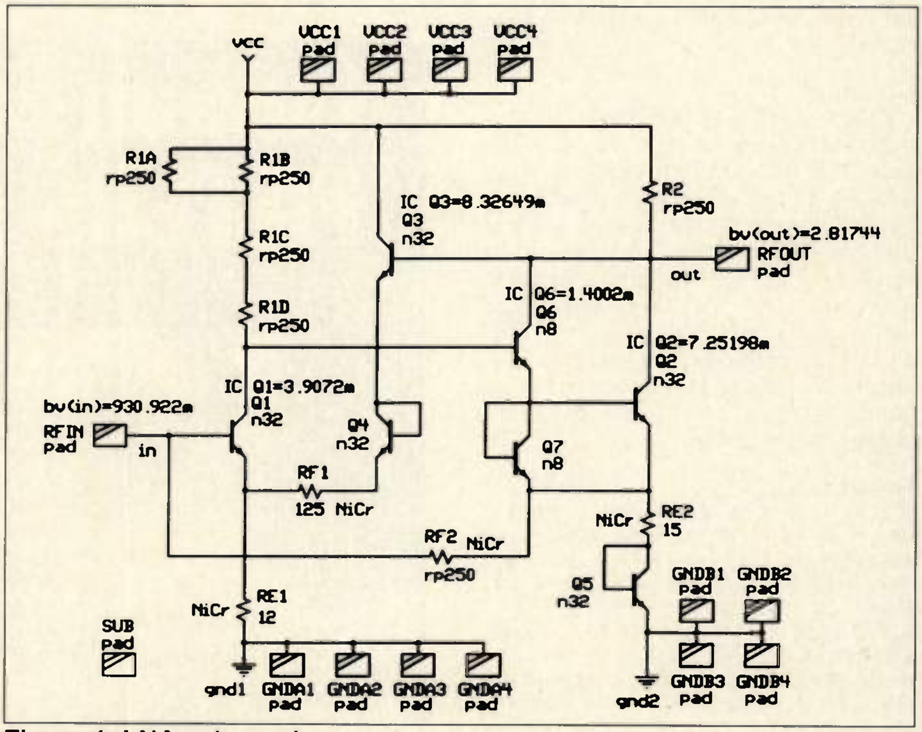


Figure 1. LNA schematic.

As a demonstration of SHPi's capabilities, an LNA was designed and fabricated on a QC6-40 analog array, packaged in a hybrid, and tested. The circuit has a 4.6 dB NF and 20 dB gain at 1 GHz, which will be useful in applications such as GPS and cellular telephone receivers.

LNA Circuit Design

In a receiver the gain and noise figure (NF) of the front-end LNA usually determine the overall system NF. NFs of the later stages are not as critical, provided the LNA gain is sufficient.

There are basically two kinds of noise contributors which limit circuit performance at high frequencies; resistor thermal noise and active device (diode or transistor) shot noise. In order to reduce the number of noise contributors, LNA circuits usually contain very few active and passive devices.

For example, most discrete LNA designs use only one active device which is specially fabricated to have low series base resistance. Reactive components are often used to set narrow-band gain and terminal impedances, further reducing noise figure.

The planar topology of a typical IC process makes it difficult to minimize base resistance and the lack of suitable value on-chip reactive components makes attaining equivalent specifications with an integrated LNA challenging.

Although some work is now being done in the area of IC inductors, IC de-

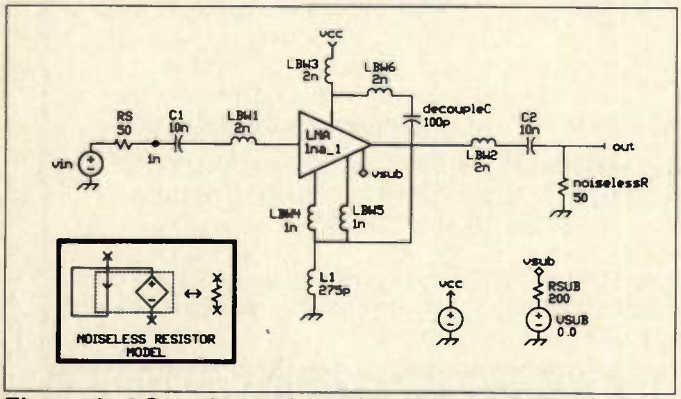
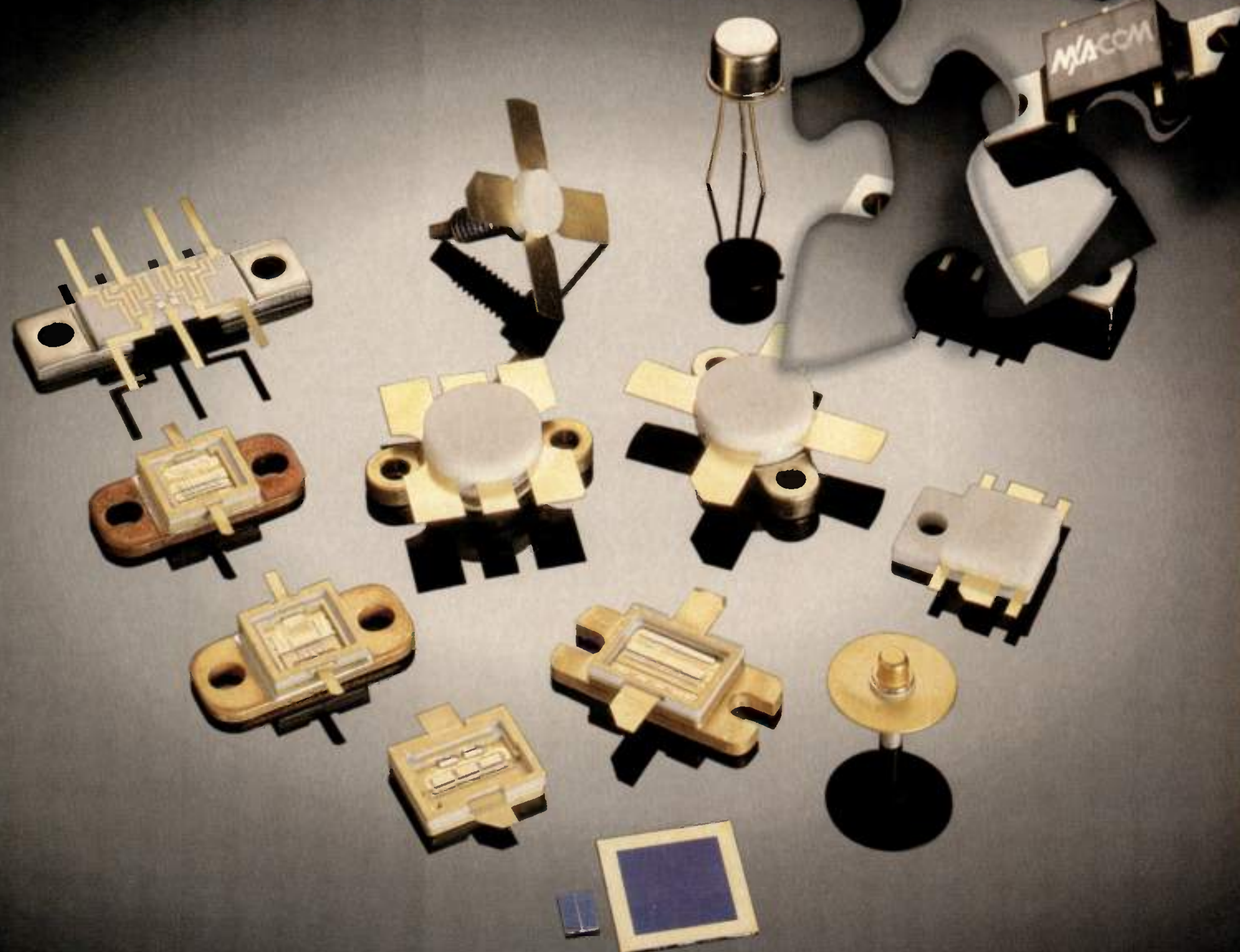


Figure 2. AC and transient simulation schematic.

```
;SIMULATE 20 POINTS PER DECADE FROM 10MHZ TO 10GHZ
ac analysis freq: 10meg 10g 20 type=dec
  ;SAVE NOISE DATA
  noise v(out) vin
endac
;PROBE INOISE (INOISE IS THE EQUIVALENT INPUT NOISE
   VOLTAGE)
probe inoise=inoise
;CALCULATE NOISE FACTOR, "F" (4KT=16.57576e-21V @ T=25C)
F=mag(inoise)^2/(16.57576e-21*50)
;CALCULATE AND PLOT NOISE FIGURE, "NF"
NF=10*log10(F)
plot NF
```

Figure 3. NF extraction.

The Final Piece



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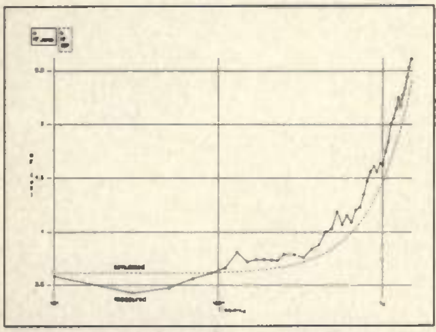

Figure 4. Noise figure vs. frequency.

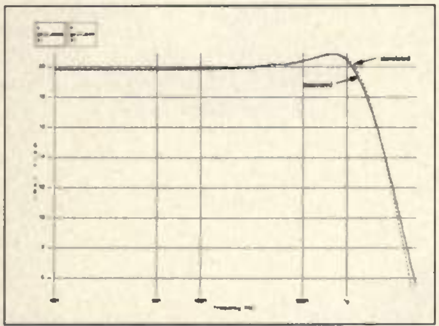

Figure 5. Power gain ($|S_{21}|^2$) vs. frequency.

signs usually depend on the use of active devices to effectively accomplish functions previously performed with capacitors and inductors in discrete design. Since active devices generate shot noise due to collector current and thermal noise due to base resistance (R_B), they must be used judiciously.

In order to achieve a 50 ohm input impedance without external reactive components, the design could use a common-emitter (CE) configuration with an internal 50 ohm termination resistor. CE amplifiers are generally less noisy than common-base and common-collector stages, but the termination resistor adds 3 dB to the NF. A better method is to employ shunt feedback around a CE stage. Not only does the feedback set the input impedance, but it also stabilizes the gain.

Figure 1 shows the LNA schematic. It is similar in topology to the Signetics 5205, employing two feedback loops around a CE stage (Q1) in order to achieve good linearity and NF while maintaining well matched impedances at the input and output. The series-shunt feedback loop containing Q3, Q4, RF1 and RE1 sets the overall gain to approximately (RF1+RE1)/RE1. RE2 and RF2 are a part of the shunt-series feedback which provides 50 ohm terminal impedances over frequency. A coupling capacitor is required along with a 50 ohm output termination in order to maintain proper biasing and 50 ohm input impedance (see Figure 2).

Q1 is a large contributor to the amplifier noise. A lot of design time was spent trying to reduce thermal noise due to R_B and shot noise generated from collector current (I_C). Lowering R_B helps to minimize noise; however, this requires increasing device size which lowers bandwidth due to larger C_{jc} and C_{je}. Reducing I_C will lower shot noise, but also at the expense of bandwidth.

Design goals were a bandwidth of 1.5 GHz, 20 dB gain and NF < 5 dB. This led to the choice of an N32, the largest QC6 NPN, for Q1. The N32 has a nominal R_B of 18 ohms and is the major noise contributor to the circuit. Table 1 lists the top noise contributors at 100 MHz and 1 GHz.

I_{C1} (I_C of Q1) is dependent upon the value of R1. Neglecting the small voltage drop across RE1 leads to the following equation:

$$I_{C1} \cong (V_{CC} - 3V_{BE}) / R1$$

For V_{CC} = 5 V and V_{BE} = 800 mV, choosing R1 to be 650 ohms will set I_{C1} to 4 mA. The additional drop across RE1 was accounted for by lowering the final value to 625 W.

As mentioned previously, RF1 and RE1 determine the gain of the amplifier. RE1 turns out to be a top noise contributor and needs to be minimized. A 12 ohm nichrome resistor was found to be a reasonable choice. The gain of 10 (20

FREQUENCY: 100 MHz Total noise voltage is 4.40e-17 V^2/Hz.

4 noise sources (3 resistors and 1 transistor) representing 95.6% of the total noise.

RESISTOR SQUARED NOISE VOLTAGES (V^2/Hz)

	RS	RE1	RF2
total	1.91e-17	5.41e-18	3.21e-18

TRANSISTOR SQUARED NOISE VOLTAGES (V^2/Hz)

transistor	total	rb	rc	re	ib	ic
Q1	1.43e-17	9.56e-18	8.54e-22	1.93e-19	2.00e-18	2.57e-18

FREQUENCY: 1 GHz Total noise voltage is 6.32e-17 V^2/Hz.

7 noise sources (3 resistors and 4 transistors) representing 95.0% of the total noise.

RESISTOR SQUARED NOISE VOLTAGES (V^2/Hz)

	RS	RE1	RF2
total	2.27e-17	6.35e-18	4.61e-18

TRANSISTOR SQUARED NOISE VOLTAGES (V^2/Hz)

transistor	total	rb	rc	re	ib	ic
Q1	1.99e-17	1.18e-17	1.08e-19	2.26e-19	2.54e-18	5.23e-18
Q7	2.39e-18	1.93e-18	2.18e-21	3.63e-20	1.90e-20	4.01e-19
Q6	2.08e-18	1.13e-18	1.31e-20	2.47e-20	2.07e-19	7.04e-19
Q2	2.05e-18	1.55e-18	3.47e-21	3.14e-20	1.34e-19	3.33e-19

Table 1. Top noise contributors.

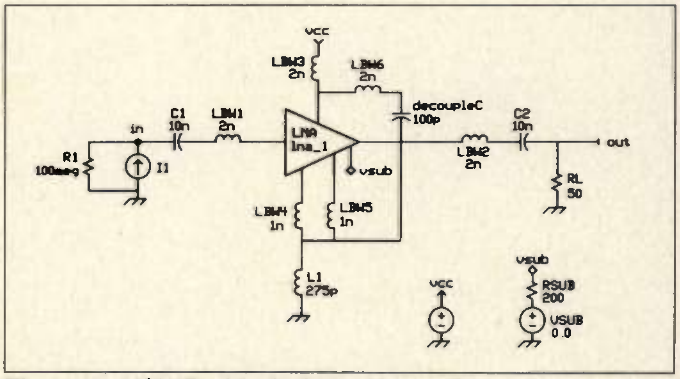

Figure 6. VSWR simulation schematic.

```
;SIMULATE 20 POINTS PER DECADE FROM 10MHZ TO
   10GHZ, SAVE DATA
ac analysis freq: 10meg 10g 10 type=dec
endac
probe v(in)=zin
;CALCULATE THE REFLECTION COEFFICIENT; S11 (GAMMA)
gamma=(zin–50)/(zin+50)
vswr=(1+mag(gamma))/(1–mag(gamma))
plot vswr
```

Figure 7. Code for calculation and plotting input VSWR vs. frequency.

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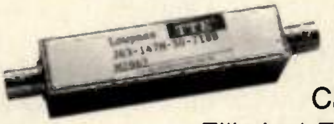

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	1dB Max@	3dB Nom@				
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DL-10.7	11MHz	14MHz	24MHz	DK-21.4-5P	21.4MHz	5%F_o
DL-21.4	22MHz	24.5MHz	41MHz	DK-30-5P	30.0MHz	5%F_o
DL-30	32MHz	35MHz	61MHz	DK-40-5P	40.0MHz	5%F_o
DL-50	48MHz	55MHz	90MHz	DK-50-5P	50.0MHz	5%F_o
DL-70	60MHz	67MHz	117MHz	DK-60-5P	60.0MHz	5%F_o
DL-90	81MHz	90MHz	157MHz	DK-70-5P	70.0MHz	5%F_o
DL-100	98MHz	108MHz	189MHz	DK-100-5P	100.0MHz	5%F_o
DL-150	140MHz	155MHz	300MHz	DK-150-5P	150.0MHz	5%F_o
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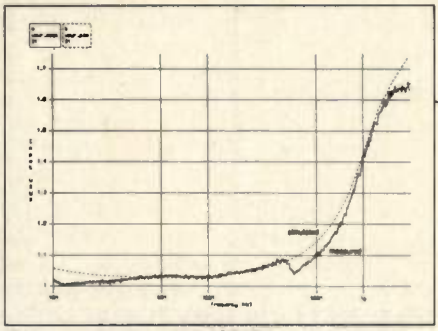

Figure 8. Input VSWR vs. frequency.

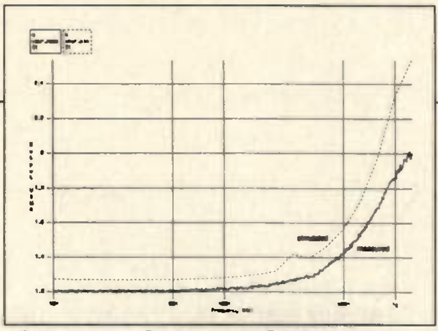

Figure 9. Output VSWR vs. frequency.

```
;PROBE THE OUPUT VOLTAGE
probe v(out)
;EXTRACT THE OUTPUT VOLTAGE IN THE 400NS WINDOW
vout=extract(v(out),5n,404.6n)
;PERFORM THE FFT AND ZOOM IN ON THE FREQUENCIES OF INTEREST
specout=extract(db(fft(vout,per)),90meg,115meg)
;16DB IS ADDED TO THE SPECTRUM TO TRANSLATE INTO UNITS OF DBM
; 6DB (OF THE 16DB) ACCOUNTS FOR THE MIRROR IMAGE OF THE OUTPUT
; VOLTAGE SPECTRUM WHICH LIES ALONG THE NEGATIVE FREQUENCY AXIS.
; THE REST (10DB) CONVERTS FROM VOLTAGE AMPLITUDE TO RMS POWER
; (DISSIPATED IN THE 50 Ω LOAD).

pout=specout+16
plot pout
```

Figure 12. TEKSPICE time domain output, FFT and plot functions.

dB) is met by choosing RF1 = 125 W. RF1 is also made from nichrome in order to match RE1.

An f_T doubler, consisting of Q6, Q7 and Q2, provides level shifting for biasing purposes and improves bandwidth. Its bias current ($I_{C6}+I_{C2}$) is set by RE2. The voltage drop across RE2 is approximately equal to that of RE1. The calculations of RE2 and R2 follow the selection of all the transistor bias currents (for V_{CC}=5V, V_{BE}=800mV).

$$RE2 = (I_{C1}+I_{C3})(RE1)/(I_{C2}+I_{C6})$$

$$R2 = [5 - 2V_{BE} - (I_{C3})(RF1) - (I_{C1}+I_{C3})(RE1)] / (I_{C6}+I_{C2})$$

RF2 determines the input impedance, R_{in}. It should be noted that, to a point, noise is decreased as RF2 is increased. After choosing RF2, the output impedance may need adjustment by tweaking the values of RE2 and R2.

The following equations (derived in Reference 2) give reasonable estimates for the input and output impedances:

$$R_{in} \cong [(RF2 + RE2)(RE1)R]/[(RE1)R + (RE2)(RF1 + RE1 + R)]$$

$$R_{out} \cong [(RF1 + RE1)(RE2)R]/[(RE1)(RE2 + RF2 + R)] \parallel (RF1 + RE1)$$

where:

R = Rs = Rl (Rs and Rl are the input and output termination resistors shown in Figure 2). Table 2 shows a list of simulated and measured LNA specifications.

Simulation Results

TEKSPICE along with the QC6 model libraries (Q6ATALL.3b0b) were used to simulate the LNA circuit performance. TEKSPICE is Tektronix' enhanced SPICE engine with powerful post-processing tools including the FFT.

Alternate process state simulation libraries [referred to as "Up" (U) and "Down" (D)] were used to simulate circuit sensitivity to a "fast" and "slow" process. These libraries are the result of pushing process parameters to their 3 sigma limits. Specifications such as NF and bandwidth are likely to have less variation than indicated by simulation with the U and D libraries.

Some of the simulation techniques are described in this section. Sections of TEKSPICE code are shown for instructive purposes. Note that comments in TEKSPICE code are preceded by the semi-colon (;) character.

Noise Figure (NF)

NF is simulated by running a small signal "AC analysis" with a "noise" statement included in the analysis block. Figure 2 shows the circuit setup used for "AC" simulations. The input voltage source, v_{in}, has an AC value of 1 V. The noise from the output termination resistor should not be included in the NF calculation. In order to "fool" TEKSPICE, a "noiseless" 50 ohm resistor terminates the LNA output. The noiseless resistor model, shown in Figure 2 (inset), is merely a "Current Controlled Voltage Source" (CCVS) with its transfer function set equal to R (in this case 50 ohms).

The TEKSPICE code shown in Figure 3 initiates the simulation and extracts NF.

Figure 4 shows a plot of simulated and measured NF versus frequencies. NF increases at high frequencies because the transistor gains are rolling off and noise contributors following these devices are showing up as bigger terms at the input. As shown in Table 1, the top three noise contributors at 1 GHz are Q1, RE1 and RF2.

Power Gain

The same analysis used for simulating NF also provides power gain ($|S_{21}|^2$) information. The TEKSPICE code shown below calculates and plots gain vs frequency.

```
gain_mag = mag(v(out)/v(in))
gain = 20*log10(gain_mag)
plot gain
```

Figure 5 shows a plot of simulated and measured gain vs frequency. The peaking is due to a common inductance in series with the two grounds (See Figure 2).

```
;SINT(OFFSET,AMPLITUDE,FREQUENCY)
tran=sint(0.0,sqrt(8e-3*rs*10^(pavin/10)),100meg)+
      sint(0.0,sqrt(8e-3*rs*10^(pavin/10)),105meg)
```

Figure 10. TEKSPICE sum of two sinusoids.

```
rs=50
pavin=-30
tran analysis time: 0 405n 400n/1024 reltol=1e-7 chgtol=1e-16
endtran
```

Figure 11. Transient analysis results.

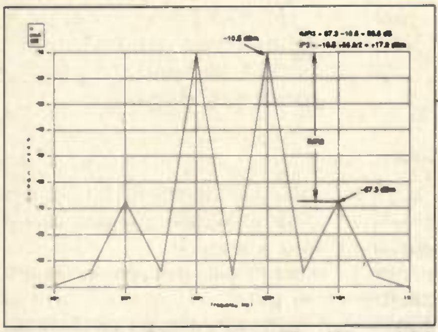

Figure 13. Simulated IP_3 (f1 = 100 MHz, f2 = 105 MHz, P_{avin} = –30 dBm).

```
;RUN A TRANSIENT ANALYSIS VARYING PAVIN FROM -40DBM TO -2DBM
var sweep pavin: -40 -2 2
 tran analysis time: 0 105n 100n/128 reltol=1e-7 chgtol=1e-16it15=1meg
 endtran
endvar
;PROBE THE OUPUT VOLTAGE
probe v(out)
;EXTRACT THE OUPUT WITHIN THE 100NS WINDOW
vout=extract(v(out),5n,104.2n)
;PERFORM THE FFT AND ZOOM IN ON THE FREQUENCY OF INTEREST
specout=extract(fft(vout,per),90meg,110meg))
;EXTRACT THE OUTPUT VOLTAGE AT THE FUNDAMENTAL FREQUENCY
vout_db=db(imag(peak(specout)))
;PLOT THE OUTPUT POWER
pout=vout_db+16
plot pout
```

Figure 14. Transient analysis run over a swept input power.

Voltage Standing Wave Ratio (VSWR)

The circuit in Figure 6 was used for simulating VSWR at the input of the LNA. Small signal AC analysis is needed once again; a test signal, I1 (R1 provides a DC path to ground to satisfy simulator convergence algorithm), of magnitude 1 is fed into the input of the amplifier. The resulting input voltage is therefore equal to the input impedance, Z_{in}. In Figure 7, the TEKSPICE code runs the simulation and then calculates and plots input VSWR versus frequency.

Figure 8 is a plot of input VSWR. Achieving good VSWR over frequency in a feedback circuit requires good overall bandwidth. At frequencies greater than 1 GHz, gain is starting to roll off and, as a result, VSWR is greater than its low frequency value.

The output VSWR, shown in Figure 9, was simulated and plotted in a similar manner, terminating the input in 50 ohms and replacing RL with the test signal.

3rd Order Intercept (IP_3)

IP_3 is a measure of signal distortion due to third order products. Refer to Reference 4 for a detailed definition. The measurement of IP_3 is sometimes referred to as a "two tone test" because the input signal is comprised of two equal amplitude sinusoids. The two sinusoids are usually close together in frequency and thus produce output components due to third order terms fairly close in frequency to the fundamental components. For example, if f1 and f2 are chosen to be 100 MHz and 105 MHz, respectively, there will be output frequency components at 95 MHz (2(f1–f2)), 100 MHz (f1), 105 MHz (f2) and 110 MHz (2(f2–f1)) (See Figure 13). The 95 MHz and 110 MHz distortion components are due to third order terms. IP_3 can be calculated from the output power levels of these components. IP_3 is defined as follows:

$$IP_3 = P_{out} + IMR_3/2$$

where P_{out} is the output power of a fundamental component (in dBm) and IMR_3 is the third order intermodulation ratio. IMR_3 is defined as the difference in output power (dB) between the fundamental components and the third order components.

In TEKSPICE a transient analysis (large signal simulation) is needed to generate the two tones at the input. The simulated output will be in the time domain, but with the use of the FFT function, frequency domain data is obtained. Continuing the example, the text in Figure 10 shows how the input voltage source in Figure 2 is set up in TEKSPICE to deliver the sum of two sinusoids. Each sine wave is defined in terms of offset, amplitude and frequency. Notice that the amplitude is written as a function of source resistance "rs" and available input power "pavin" (in dBm).

In this example the nonlinearities generated by the amplifier will lead to frequency components occurring in 5 MHz increments (5 MHz is the greatest common divisor of the two tones). In order for the FFT algorithm to extract these points, it must operate on a window of time equal to n/(5 MHz), where n is an integer greater than or equal to 1. Choosing n equal to 2 means that the FFT will extract frequency data every 2.5 MHz (1/time window). The extra points show where the noise floor of the simulator resides; a good reality check. Too high a noise floor may dictate increasing the convergence parameters of the simulator. The TEKSPICE command block in Figure 11 shows the transient analysis of the 100 MHz/105 MHz two tone test:

In this case, 2/(5 MHz) suggests a 400 ns time window for the FFT. An extra 5 ns is added to insure the amplifier reaches a steady state. 1024 points within the 400 ns window gives an effective sample rate of 2.56 Gs/s. (Nyquist frequency is 2.56 GHz/2 or 1.28GHz; aliasing should not be a problem). The TEKSPICE FFT must operate on 2^n points, where n is an integer between 4 and 15.

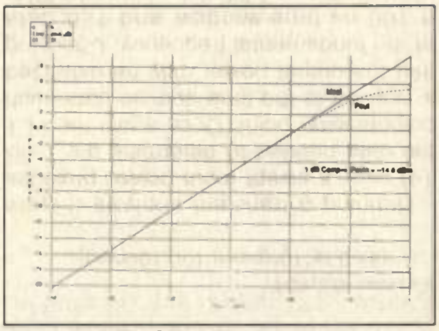

Figure 15. Simulated 1 dB compression at 100 MHz.

After the simulation is run, the TEKSPICE commands process the time domain output, perform the FFT and plot the results (Figure 12).

Figure 13 shows the simulated output spectrum of the two tone test where f1=100 MHz and f2=105 MHz. IMR_3 is designated on the plot. As shown in Figure 13, IP_3 at frequencies near 100 MHz is +18 dBm.

1 dB Compression

The 1 dB Compression point is defined as the output power level at which the gain has decreased 1 dB from its low power value. The decrease is caused by nonlinearities inherent in the amplifier; an indication of the transition between small signal and large signal operation. Since the 1 dB compression is a large signal phenomenon, a transient analysis is required for the simulation. The input voltage source, v_{in}, shown in Figure 2 is programmed to be a sinusoid with a given offset, amplitude

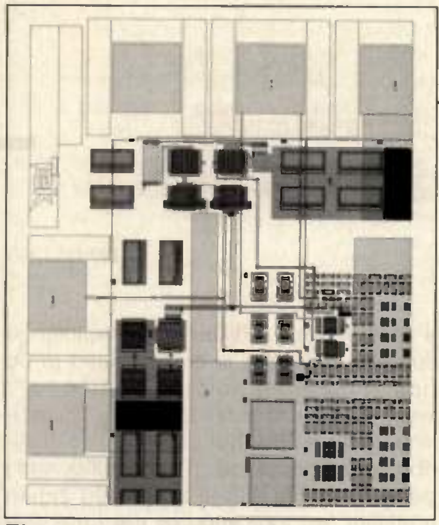

Figure 16. LNA layout.

and frequency. As in the IP3 example above, the amplitude is written as a function of source resistance "rs" and available input power "pavin". Note, for this example the frequency is chosen to be 100 MHz:

```
;SINT(OFFSET,AMPLITUDE,FREQUENCY)

tran=sint(0.0,sqrt
    (8e-3*rs*10^(pavin/10)),100meg)
```

Running a transient analysis (Figure 14) over a swept input power provides the data needed to determine the 1 dB compression point. Once again the FFT is applied to the time domain data such that the output power may be extracted at the fundamental frequency. Choosing a 100 ns time window and 128 data points will give FFT data every 10 MHz out to 1.28 GHz. Aliasing is not a problem since 1.28 GHz is greater than the 12th harmonic of the 100 MHz input frequency.

Figure 15 is a plot of output power versus input power. Notice the 1 dB compression point is marked where the output power is 1 dB less than the straight line.

Measured Results

The measured specifications are listed along with the simulated numbers in Table 2. In addition, the plots in Figures 4, 5, 8 and 9 contain measured data as well as the simulated nominal curves. The agreement is quite good. Some of the differences may be due to stray components not accounted for in the simulations or measurements, or mismatch between nominal simulation device models and the devices on this particular process lot.

Layout and Packaging

QuicKic™, Tektronix' schematic capture and layout tool, was used to layout the LNA on the QC6-40 analog array (see Figure 16). QuicKic provides netlist driven layout with built-in design rule checking, guaranteeing one-to-one correspondence between schematics and layout with no design rule violations. Because the devices are preplaced on the QC6-40 (with the exception of nichrome resistors), layout time is significantly reduced compared to a full custom design. In addition fab turn-around time and NRE are much less.

Special attention was given to minimizing metal run distances in order to keep parasitic capacitances down. The use of multiple bond pads, and thus multiple bond wires, in the power supplies reduces the bond wire inductances. Also, simulations dictated that the two ground connections shown in Figure 1 be separated in order to cut a potential positive feedback loop from the output back to the input. The package used to characterize the LNA is a standard hybrid package built by Tektronix for high frequency circuits (up to 18 GHz, Reference 5).

RF

		Simulated (PIATALL.3b0b)			Test Conditions
Specification	Measured	Nom.	Up	Down	(unless otherwise noted Tj-25°C, Vcc=5V)
3dB Bandwidth	1.51 GHz	1.53 GHz	1.59GHz	1.21 GHz	
Noise Figure	3.6 dB 4.6 dB	3.6 dB 4.5 dB	3.2 dB 3.7 dB	4.5 dB 5.9 dB	f=100 MHz f=1 GHz
Power Gain $\lvert S_{21}\rvert 2$	19.9 dB 20.5 dB	19.8 dB 20.6 dB	19.5 dB 19.3 dB	19.9 dB 19.6 dB	f=100 MHz f=1 GHz
Gain Variation w/Temperature		±0.05dB ±0.68 dB	±0.06 dB ±0.31 dB	±0.05 dB ±1.64 dB	f=100 MHz, T_j=25-125°C f=1 GHz, T_j=25-125°C
Gain Variation w/Frequency	±0.02 dB ±0.13 dB	±0.02 dB ±0.10 dB	±0.02 dB ±0.10 dB	±0.03 dB ±0.38 dB	f=10 MHz-100 MHz f=900 MHz-1 GHz
Gain Variation w/Supply	±0.22 dB ±0.94 dB	±0.18 dB ±0.67 dB	±0.19 dB ±0.54 dB	±0.18 dB ±0.98 dB	f=100 MHz, V_{cc}+4.5=5.5 V f=1 GHz, V_{cc}=4.5-5.5 V
Rev Isolation $\lvert s_{12}\rvert$	-25.6 dB -24.5 dB	-23.9 dB -24.7 dB	-23.9 dB -23.0 dB	-24.3 dB -27.2 dB	f=100 MHz f=1 GHz
Input VSWR Z0=50Ω	1.03:1 1.40:1	1.03:1 1.42:1	1.17:1 1.49:1	1.09:1 1.56:1	f=100 MHz f=1 GHz
Output VSWR Z0=50Ω	1.21:1 1.85:1	1.28:1 2.33:1	1.28:1 1.85:1	1.23:1 3.17:1	f=100 MHz f=1 GHz
1 dB Compression	-13.0 dBm -18.7 dBm	-14.8 dBm -20.9 dBm	-13.2 dBm -16.7 dBm	-16.5 dBm -22.9 dBm	f=100 MHz f-1 GHz
Output IP3	16.7 dBm 5.6 dBm	17.9 dBm 6.4 dBm	19.1 dBm 11.5 dBm	16.3 dBm 3.3 dBm	f_1-100 MHz, f_2=105 MHz, P_{in}=-30 dBm f_1=990 MHz, f_2=1 GHz, P_{in}=-30 dBm
Power Dissipation	102 mW	105 mW	119 mW	90 mW	V_{cc}=5 V, T_j=25°C

Table 2. Low noise amp measured and simulated results.

Acknowledgments

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CAD of a Broad-Band Class-C 65 Watt UHF Power Amplifier

By Robert Baeten
Motorola Inc.

With the RF CAD tools available today, RF power amplifier design is becoming less of a cut and try design methodology. Harmonic balance software, which simultaneously solves a network in the time and frequency domains, is the state-of-the-art in nonlinear RF CAD technology. Nonlinear models of bipolar power transistors for harmonic balance software are spinoffs of the Gummel-Poon model. Unfortunately, these models are not readily available from most device manufacturers which forces RF power amplifier designers to continue using conventional large-signal impedance design techniques. Even though large-signal impedance design isn't ideal, power amplifiers can be designed and constructed with minimal tuning after circuit fabrication.

Hejhall (1) presented a systematic approach to designing power amplifiers using large-signal impedance data. This paper presents a power amplifier designed using Hejhall's large-signal impedance methodology and commercially available CAD software. Simulated and measured results are shown for a 65 Watt Class-C amplifier incorporating a Motorola MRF658 bipolar transistor operating over the frequency range from 470-512 MHz and a supply voltage of 12.5 volts. Minimum goals for gain, collector efficiency and IRL were 5 dB, 55 percent and 15dB, respectively, across the frequency band.

Design Methodology

The first step in any power amplifier design is to choose an appropriate device. Device selection is based on output power levels, power gain, collector efficiencies, supply voltage and frequency band of operation. This paper assumes a device has already been identified.

After a suitable device has been selected, the designer proceeds to the second step which is to obtain the optimum large-signal source and load impedances. Wood and Davidson (2) presented techniques for measuring large-signal impedances of a device which are defined as the optimum input impedance, Z_{IN}, and the conjugate of the load impedance, Z_{OL}^*, at a specific output power, supply voltage and frequency. The success of any power amplifier design using this technique depends on the accuracy of the large-signal impedance data. The designer must be aware that large-signal impedances typically vary when the operating conditions change. It is a misconception that power amplifier circuits present a conjugate match to the output of the transistor (3). In actuality, the outputs are typically mismatched in Class-C power amplifiers in order to achieve a specific output power at a high collector efficiency.

For broad-band designs, optimum impedance data at several frequencies across the band is desirable in order to determine impedance characteristics versus frequency. Ideally, one would prefer to have load-pull data (4) showing output power, power gain, and collector efficiency versus load impedance. This measurement technique is very time consuming and is efficiently accomplished with an automated set-up. Load-pull data gives the designer insight into what happens to the output power if the load impedance deviates from the optimum value. This is especially useful in broad-band designs where it is difficult to track the optimum impedances at all frequencies.

Table 1 shows typical large-signal impedance data for a Motorola MRF658 NPN common-emitter bipolar transistor from 400 MHz to 520 MHz. It is important to realize that this data is referenced to the edge of the 1/2" CQ package and the output power and supply voltage are 65 Watts and 12.5 Vdc, respectively.

The third step in a power amplifier design is to determine input and output matching networks using the large-signal impedance data in Table 1. The simplest design technique is to first find a lumped element topology which matches the input impedance, Z_{IN}, and conjugate of the load impedance, Z_{OL}^*. At UHF frequencies, matching networks typically consist of chip or tuning capacitors, wire-wound inductors, series transmission lines and shunt transmission line stubs. The most practical matching topology in broadband power amplifier designs is the lowpass L section consisting of a series inductor and shunt capacitor.

An initial matching network topology is obtained by trial and error on a Smith Chart™ applying basic impedance matching techniques. A computer program by Moline (5), Motorola's Impedance Matching Program (MIMP), is an excellent tool for determining an ideal lossless matching network using the Smith Chart. This program includes lumped elements, microstrip transmission lines and a unique distributed capacitance element. MIMP allows the designer to see the effect of each matching element on the overall impedance transformation displayed on the Smith Chart. As element values are tuned, the results are continuously modified on the

FREQ [MHz]	Z_{IN} [Ohms]	Z_{OL}^* [Ohms]
400	0.62+j2.8	1.2+j2.5
440	0.72+j3.1	1.1+j2.8
470	0.79+j3.3	0.98+j3.0
490	0.84+j3.4	0.91+j3.2
512	0.88+j3.5	0.84+j3.3
520	0.90+j3.6	0.80+j3.4

Table 1. Typical optimum input and conjugate of load impedances for MRF658. Pout = 65 Watts, Vdc = 12.5 volts.

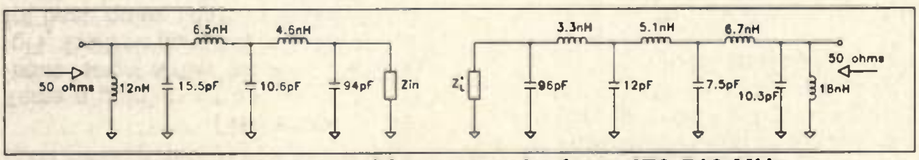
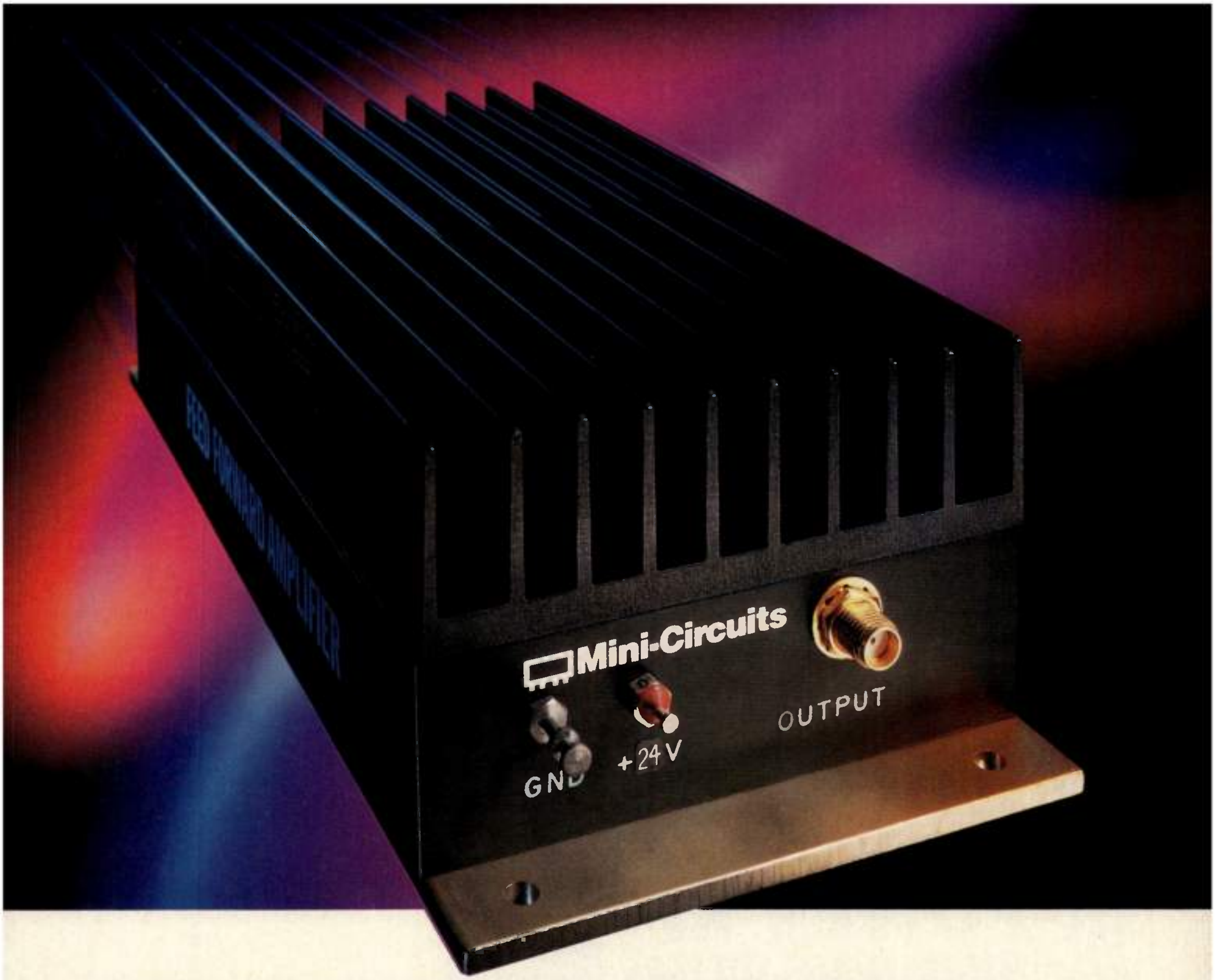


Figure 1. Lumped element matching networks from 470-512 MHz.



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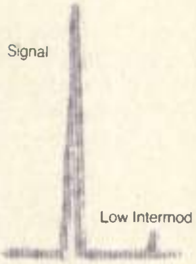

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Intercept Point, dBm		
IP2	70	63
IP3	50	43
DC Power		
Volts	24	24
Current, Amps	1.3	0.65
VSWR, 50 ohms, In/Out	1.5:1	2:1
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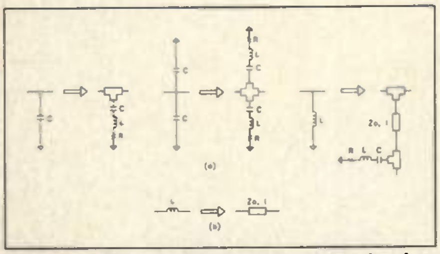


Figure 2. a) Shunt lumped element conversions to microstrip, b) Series inductor conversion to microstrip.

Smith Chart display.

Using the large-signal impedance data in Table 1, broadband lumped element input and output matching networks were designed with MIMP. In order to maintain a low Q, the first element of the matching network must be a shunt capacitor which typically brings the impedance in the vicinity of the real axis on the Smith Chart. This capacitor is followed by lowpass L networks. For broadband designs, the number of lowpass L sections is chosen to maintain a low loaded Q, achieve practical element values and flat return loss.

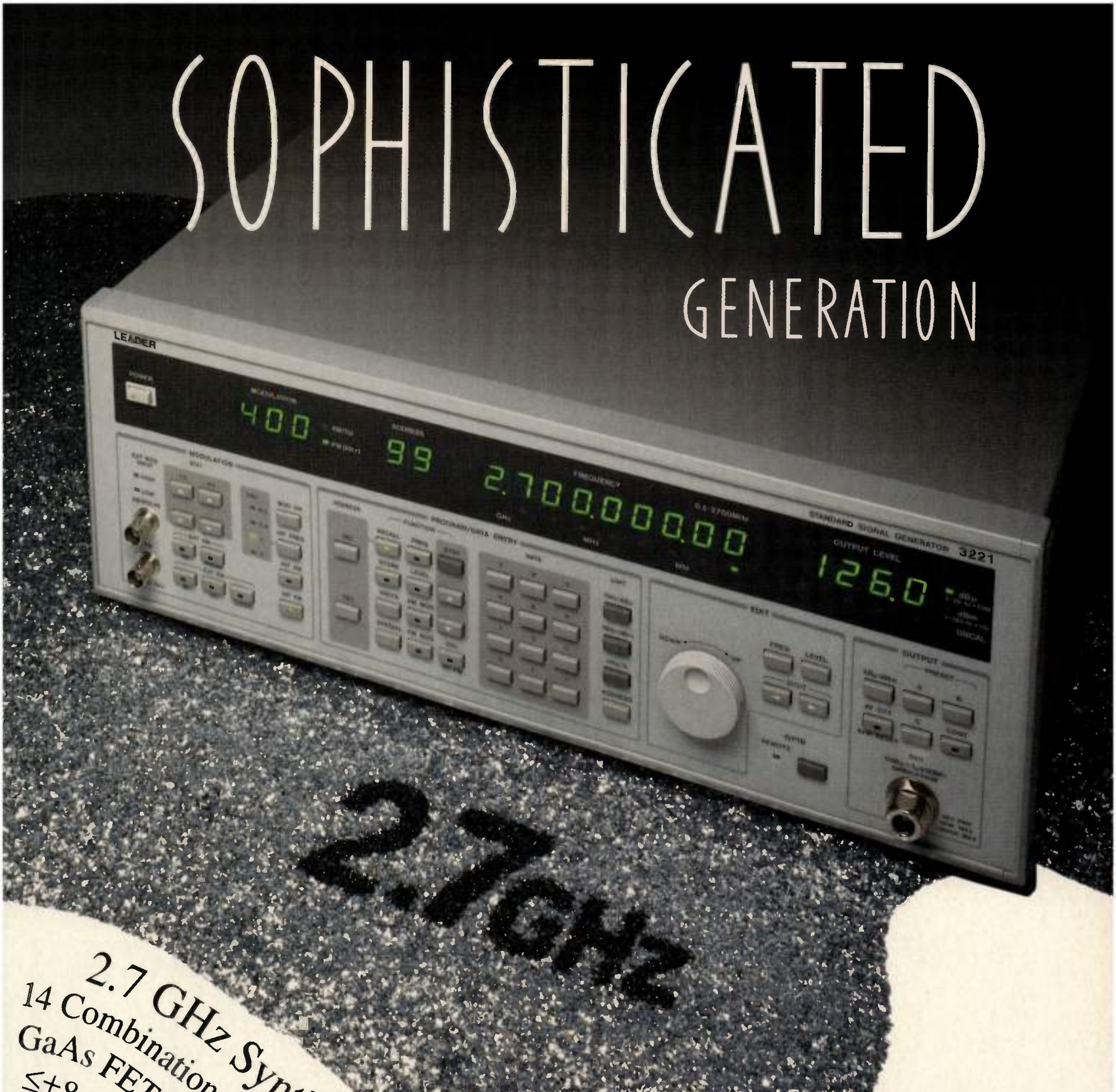
The resulting lumped element networks are shown in Figure 1. The input matching network consists of a shunt capacitor at the device interface and two lowpass L networks. The second L network has an inductor in parallel with the shunt capacitor. This resonant circuit provides the appropriate phase rotation versus frequency for optimal tracking of the device impedance. The output matching network is the same topology as the input with the addition of a third lowpass L network. The desired bandwidth could theoretically be achieved with fewer lowpass L sections at the expense of circuit tunability and sensitivity to parameter variation. In MIMP, the lumped element matching networks shown in Figure 1 yield an input return loss greater than 21 dB and an output return loss greater than 19 dB across the frequency band from 470 to 512 MHz.

A vital factor which is important in achieving optimum performance of a power amplifier is the matching network close to the device. In this design, the 94 pF capacitor on the input and 96 pF capacitor on the output are implemented with two chip capacitors to ground, mounted directly on the package leads. This maintains even current distribution in the ground paths of the coplanar type structure of the 1/2" CQ transistor package. By providing the appropriate matching elements close to the device, the harmonic termination is improved yielding higher gain and collector efficiency.

An important point about the large-signal impedance model is that the input and output are considered as separate entities. In practice this is not true. Any change in the output load impedance changes the device input impedance because of internal feedback. The designer has no feel for input match degradation due to a load impedance which is not optimum over all in-band frequencies. This inadequacy could be overcome with an automated load pull system where the input is tuned for each load condition.

The next step in the design process is to convert this lumped element design to a planar transmission line design such as microstrip and add in the appropriate bias feed circuitry. For this step in the design process, Academy™ (6) was used which generates a layout from a schematic diagram and provides a netlist for Touchstone™ or Libra™ simulation. Standard planar transmission line elements are added to the layout automatically, but the designer manually adds RF grounds where desired and some of the bias feed circuitry using the layout editor within Academy.

In this amplifier design, shunt capacitors to ground were implemented with chip capacitors in the microstrip configuration. A chip capacitor connected in shunt between a microstrip line and ground, was modeled with a microstrip tee for single capacitors or microstrip cross junctions for two capacitors in parallel. Chip capacitor values are adjusted to account for the effect of parasitic inductance. Shunt inductors were converted to high impedance microstrip lines shorted to ground through a low impedance chip capacitor and series inductors were converted to series microstrip lines. It should be noted that microstrip line models as well as open and short circuit stubs exist within MIMP. This software also has a distributed capacitance model which is useful when the operating frequency is high enough where chip capacitor widths are influential. The length of a short-circuited line representing a lumped shunt inductor is given by the following equation: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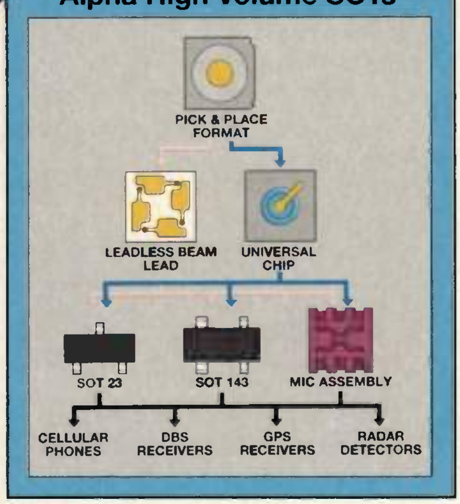

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$$l = \tan^{-1}\left(\frac{X_L}{Z_0}\right)\frac{\lambda}{2\pi} \qquad (1)$$

where
X_L = Reactance of shunt inductor
Z_0 = characteristic impedance of microstrip line
λ = wavelength of microstrip line

The length of a series microstrip line representing a lumped series inductor can be approximated by the following equation:

$$l = \sin^{-1}\left(\frac{X_L}{Z_0}\right)\frac{\lambda}{2\pi} \qquad (2)$$

where
X_L = Reactance of series inductor
Z_0 and λ are given above.

For low impedance lines and very long transmission lines this equation becomes less accurate because of the distributed capacitance inherent in a transmission line. The error produced in this conversion is corrected in circuit optimization. Figure 2 summarizes all of the lumped element conversions used in the transition from the lumped element design to a microstrip configuration. The parasitic elements inherent in a chip capacitor are also shown.

The converted microstrip designs for the input and output matching networks are shown in Figures 3a and 3b. It should be noted that 50 ohm microstrip lines were used throughout and line lengths were calculated using reactance and wavelength values at 490 MHz. The circuit board used in this design was 1/16" thick FR4 with 2 ounce copper. Simulated results of the input and output reflection coefficients in this initial design using Touchstone and the lumped element design using MIMP are shown in Figures 3c and 3d. As shown in these figures, the converted microstrip design has reduced input and output return losses. The input return loss is greater than 8.5 dB and the output return loss is greater than 10 dB over the frequency band. At this point in the design, modification to the value and location of some of the matching elements, microstrip line lengths and characteristic impedances is necessary. This task is best accomplished with an optimizer found on RF CAD programs such as Touchstone, MMICAD™, and Super Compact™.

The first section of the DC feed structures were included in Figures 3a and 3b. It is important to connect the DC feed structures at a low impedance point in the RF circuit for good isolation and minimal loading on the amplifier. For power transistors this is achieved by connecting both feed structures as close as possible to the device. In the output matching network, only the first choke coil and bypass capacitor are shown since this is all that is typically needed to accurately simulate in-band performance. In the final amplifier circuit, three stages of bypassing are employed in the collector feed structure to provide the appropriate terminations to the device at low frequencies to achieve stability. The feed structure on the input matching network does not include the DC return coil necessary for Class-C operation. This will be added later since it will not affect in-band performance. Kwitkowski (7) discusses feed structure design techniques in order to eliminate low frequency instabilities. Many times the first design will oscillate under a certain VSWR, drive condition, and/or supply voltage requiring changes to be made to the feed structure design.

The final step in power amplifier design is to optimize the microstrip design and generate a layout. This step usually requires several iterations between cir-

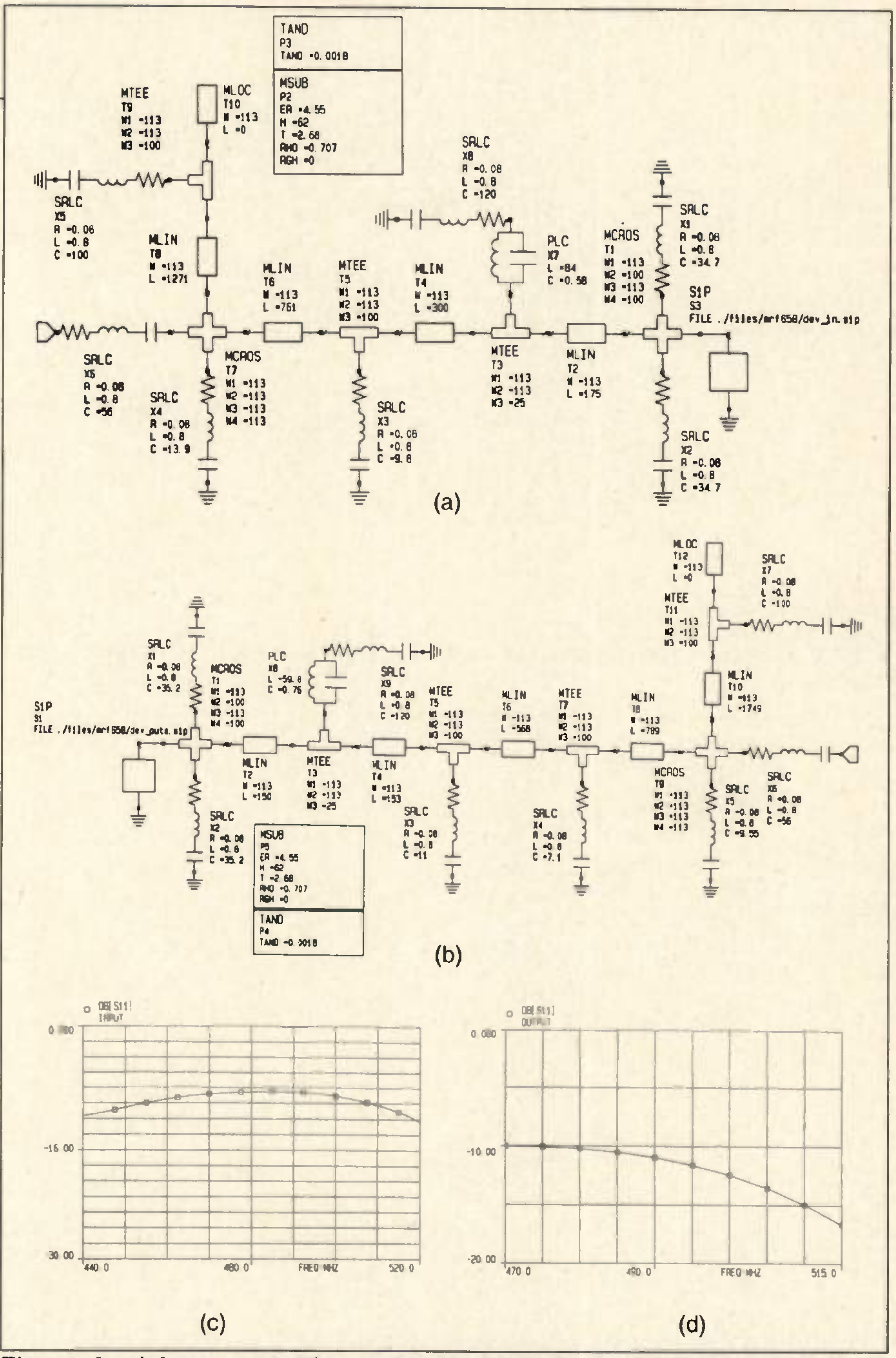

Figure 3. a) Input matching network, b) Output matching network, c) input reflection coefficient, and d) output reflection coefficient.

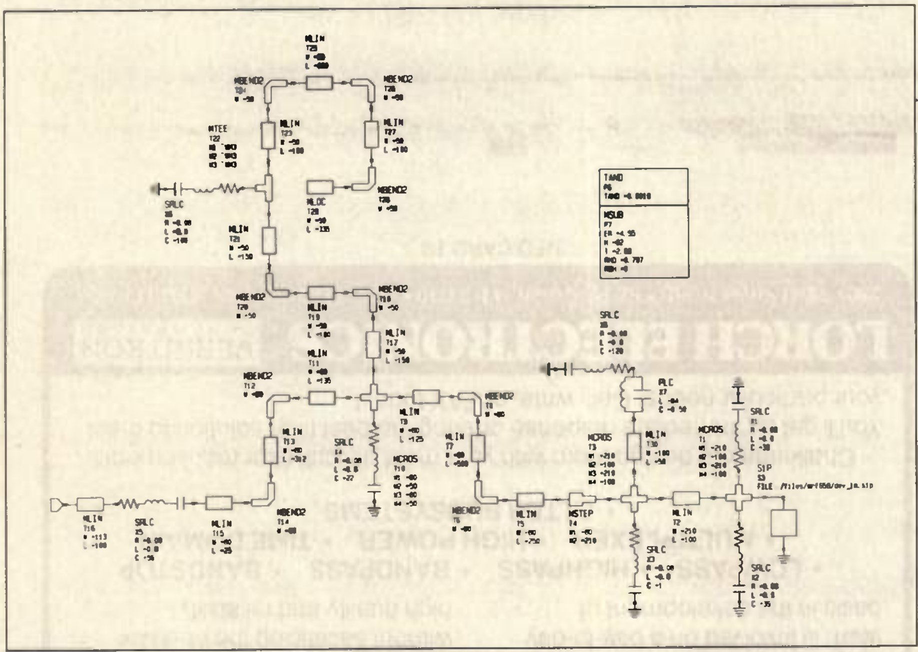

Figure 4. Schematic diagram of optimized input matching network from Academy.

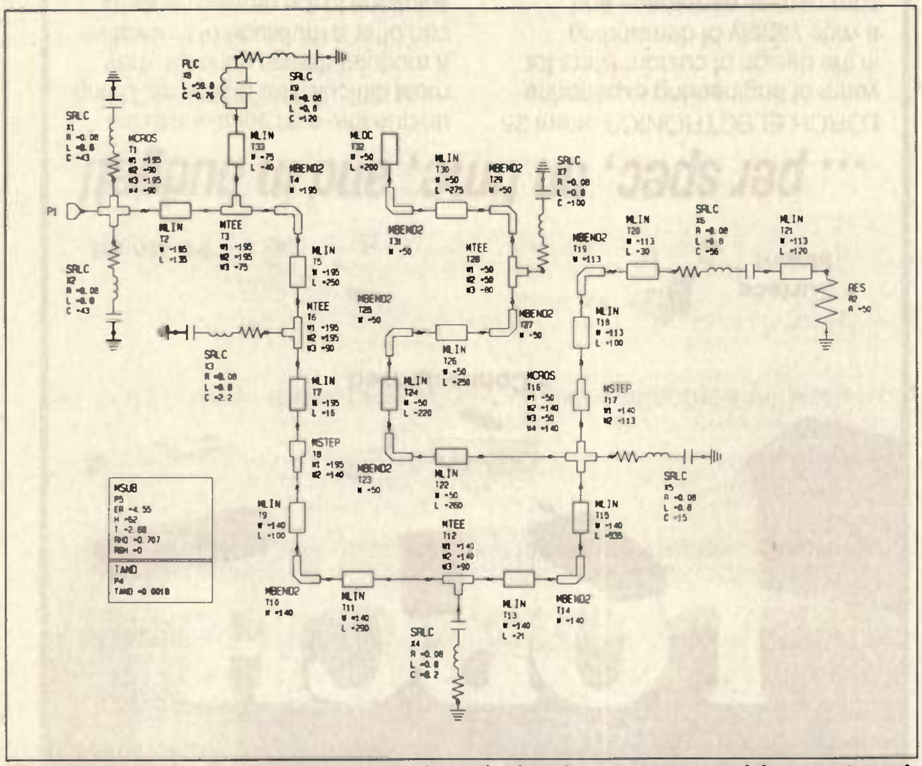

Figure 5. Schematic diagram of optimized output matching network from Academy.

cuit optimization and layout because amplifier designers typically have a given area factor. In this design, the power amplifier was intended to fit on a 3" x 2.5" circuit board. In order to fit within this specified area, microstrip bends were implemented to meander the series and shunt microstrip lines. Models are contained within Touchstone to simulate the discontinuity and phase length of the bends.

With this design technique, the input and output matching networks are best optimized separately. Optimization requires the designer to set specific goals, such as greater than 15 dB return loss over a specified frequency range. Then selected element values are allowed to vary until the error function is minimized. Variables in the optimization were chip capacitor value and location along with microstrip line dimensions. When a change in line width occurred, the microstrip step model was implemented to account for the discontinuity. The optimization goal on the input was a return loss greater than 18 dB from 470 to 512 MHz. Figure 4 shows the optimized input matching network and Figure 6a shows the input reflection coefficient predicted from Touchstone which is greater than 19 dB from 470 to 512 MHz.

Optimization of the output matching network is not as straightforward as the input matching network since the large-signal impedance data does not reveal how fast the gain drops off as the load impedance deviates from the optimum value. There are basically two ways to optimize the output matching network. The first method is to simply specify a return loss looking into the matching network from the 50 ohm load. This technique was used on the input matching network and also used in MIMP to obtain the initial lumped element output matching network. The disadvantage with this technique is that the output power and efficiency are not known for a finite return loss.

The second technique is to look into the output matching network from the device interface and optimize this impedance. This method is especially desired with load pull data because the output power and efficiency are known as the load impedance varies. Once an allowable gain and efficiency variation is known, the range of desirable load impedances are known from the load pull data. This range is then used as optimization goals.

In this design load pull data was not implemented, but the second optimization technique was still employed on the output matching network. As in the input optimization, variables in the output network optimization were chip capacitor value and location along with microstrip line dimensions. Figure 5 shows the optimized output matching network and Figure 6b shows the load impedance seen by the device predicted by Touchstone. The predicted load impedance best matches the desired impedance around 490 MHz with the largest deviation occurring at the high end of the frequency band.

Measured Results

After constructing an amplifier, it functioned, but performance was not optimum. Some tuning was necessary in order to achieve the design goals across the frequency band of interest. The

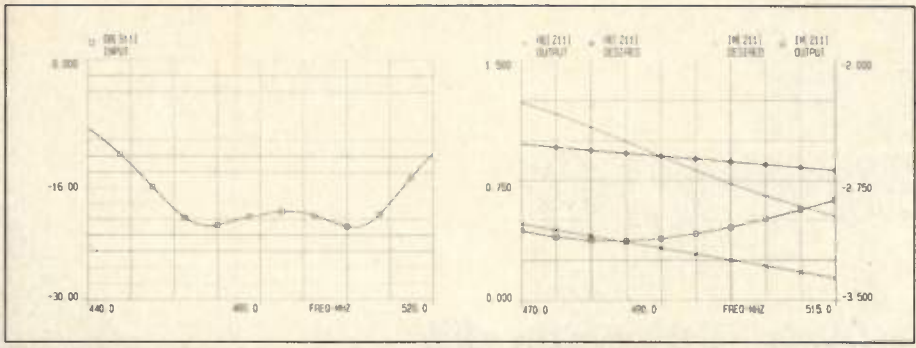

Figure 6. Predicted results for optimized microstrip design from Touchstone, a) input reflection coefficient, b) load impedance.

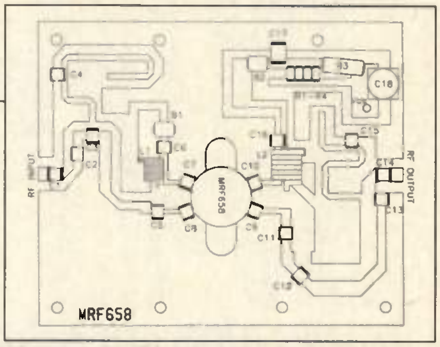

Figure 8. Diagram of assembled RF power amplifier.

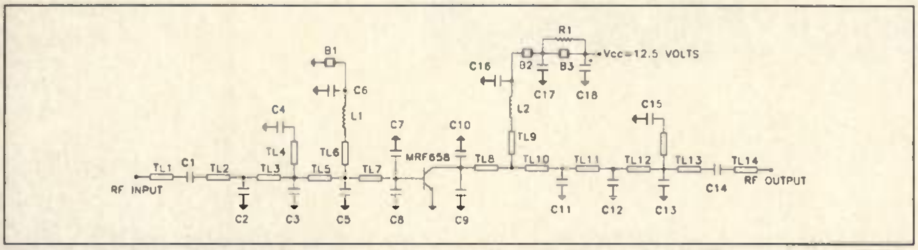

Figure 7. Schematic diagram of completed Class-C UHF power amplifier.

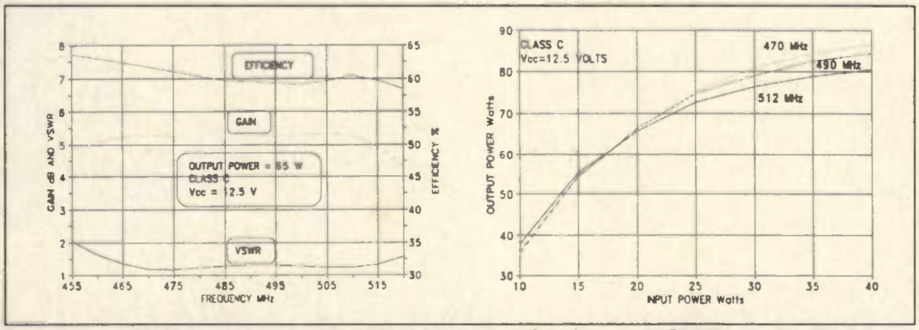

Figure 9. Measured results a) average gain, input VSWR and efficiency for six amplifiers, b) output power versus input power.

value and/or location of a few chip capacitors were changed and an additional capacitor was added to optimize performance of the amplifier. A schematic diagram of the finished power amplifier is shown in Figure 7 and an assembly drawing is shown in Figure 8. It should be noted that the assembly drawing of the amplifier does not include the plated through holes in the ground planes. On the input matching network, a 1.5 pF chip capacitor, denoted by C_2, was added to achieve good IRL performance across the frequency band. The values of capacitors C_3 and C_5 were changed from the optimized values in Touchstone. In the output matching network, the location and value of capacitors C_{11} and C_{12} were adjusted to optimize gain and efficiency.

Six amplifiers were constructed to determine repeatability of the amplifier design. Figure 9a shows the average gain, VSWR, and efficiency for these amplifiers at 65 Watts output power. Across the frequency band from 470 MHz to 512 MHz, the average gain was over 5 dB at 65 watts output power, the worst case VSWR was under 1.3 (IRL > 17 dB), and the collector efficiency was greater than 59 percent. The input VSWR agrees well with predicted results from Touchstone. Figure 9b shows the output power versus input power at 470 MHz, 490 MHz, and 512 MHz. The average saturated output power of the six amplifiers at 470 MHz was over 86 Watts while at 512 MHz it was over 80 Watts.

Summary

A 65 Watt Class-C UHF power amplifier was successfully designed and fabricated using large-signal impedance data and commercially available CAD software. This design technique minimized traditional cut and try methodology. Some tuning of the amplifier was necessary after circuit fabrication, but this was minimized by computer optimization in the design process. The success of this design technique relies heavily on the accuracy of the large-signal impedance data and accuracy of the models incorporated in the CAD program. *RF*

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About the Author

Robert Baeten has an MSEE degree from Marquette University in 1986 and a BSEET degree from the Milwaukee School of Engineering in 1984. He is currently an Applications Engineer in Motorola's RF Land Mobile Group. Since joining Motorola in 1991, he has been responsible for development of high power semiconductor devices. Prior to joinging Motorola, he was a Technical Staff Engineer for 5 years at Rockwell International, Collins Division, in the Advanced Technology & Engineering Group. He can be reached at Motorola Inc., Mail Drop E109, 5005 East McDowell Road, Phoenix, AZ 85008, or by phone at (602) 244-3330.

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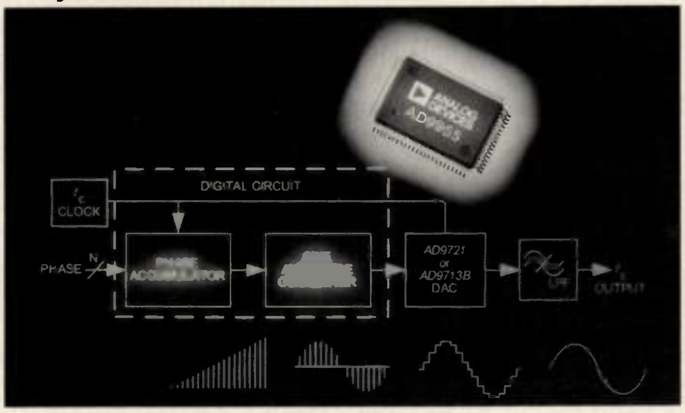

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Alan Industries, Inc.
INFO/CARD #168

Single Layer Capacitor
Compex is pleased to announce the single layer, parallel plate margin capacitor. Margin capacitors have the topside electrode withdrawn from the edges in order to increase the distance between electrodes and dramatically decrease the possibility of shorting when epoxy die mounting.
Compex Corporation
INFO/CARD #167

Filters
RS Microwave will show the DR series of dielectric resonator filters with wide stopbands, and a compact notch filter for Tacan rejection in JTIDS systems.
RS Microwave
INFO/CARD #166

Signal Generators
Models PSG1000B and PSG-2400A are synthesized signal generators covering 10 kHz to 1 GHz and 100 kHz to 2.4 GHz, respectively. Both have AM, FM and PM capability, along with an automatic SINAD meter. The 2400A has two internal modulation sources and supports CTCSS, DTMF and SELCALL signaling systems. The 1000B supports CTCSS signalling systems.
Wayne Kerr/Farnell
INFO/CARD #165

Circuit Design For Windows
EEsof introduces a trio of linear and nonlinear circuit design tools that fully utilize the Microsoft Windows 3.1 graphical user interface. The family includes three programs — Touchstone, Libra and LineCalc for Windows, along with extensive models for high-frequency circuit elements and libraries of popular RF and microwave transistors.
EEsof
INFO/CARD #164

Limiter, Amplifier MMIC
Model A91301A4, a drop-in microwave limiter covering 500 MHz to 18 GHz, handles 1 W CW, 10 W pulsed. The connectorless design is 0.33 inches long and 0.25 inches wide. Model AL-H102C is a distributed HEMT amplifier MMIC covering 2 to 20 GHz and featuring 10 dB gain with 2.5 dB NF.
FEI Microwave, Inc.
INFO/CARD #163

PCN/PCS Products
The VDS 9000 series from Sciteq Electronics includes the VDS-9000, a low-cost modular synthesizer; the VDS-9002 synthesized downconverter, the VDS-9003 AGC amplifier; and the VDS-9004 power module, with 25 dB gain and +30 dBm power output.
Sciteq Electronics, Inc.
INFO/CARD #162

Crystal Ovens
Isotemp Research has introduced a series of miniature crystal ovens for stabilizing the temperature of quartz crystals. These ovens are available at any operating voltage in the range of 5 to 28 Vdc, with set temperatures from +35 to +95 degrees C. Prices range from $11.00 to $15.00 in 1000 piece quantities.
Isotemp Research, Inc.
INFO/CARD #161

Capacitors
Microelectronics will display their high power, non-magnetic ceramic capacitors for medical applications.
Microelectonics
INFO/CARD #160

Cable Assemblies
Standard cable assemblies for instrumentation, commercial, conformable and semi-rigid applications are offered by Penstock. Lengths from three inches to six feet are in stock and ready for immediate delivery. SMA male to SMA male connectors are standard; other configurations are available.
Penstock Engineering Labs
INFO/CARD #159

GSM LNA
The GSM low noise amplifier is specially designed for applications in GSM base station receivers and operates in a frequency range from 880 to 915 MHz. It exhibits exceptionally low noise figure and high third order intercept. While specifically tailored for the GSM band, it provides similar performance in the 824 to 849 MHz AMPS band.
AT&T Microelectronics
INFO/CARD #158

Chip Inductors
The 5000 series of chip inductors are encapsulated surface mounted inductors that are less than 0.046 inches in height. This new series is ideal for use in high frequency hybrid and microwave applications when a high Q factor is required. Inductance values range from 8 nH to 10 uH. Typical Q factor is 45.
Vanguard Electronics Co.
INFO/CARD #157

Substrate Test Fixture
The series WK-3000 adjustable universal substrate test fixture provides a highly flexible test platform for microstrip and coplanar substrates. It also offers the flexibility of additional RF port additions from DC to 26.5 GHz. Thus multiple RF port units can be tested. Adjustments can be made in the X-, Y- and Z-axes to accommodate a variety of different substrate circuit configurations.
Inter-Continental Microwave
INFO/CARD #156

Diplexers
FSY Microwave announces a complete line of standard suspended substrate diplexers, covering the 2 to 18 GHz frequency range. Standard crossovers are at 2, 4, 8, and 12 GHz. Both connectorized and drop-in versions are available.
FSY Microwave, Inc.
INFO/CARD #155

VXI Synthesizer
Racal-Dana model 3261 is a single slot, C-size VXI synthesizer designed for both military and commercial VXIbus ATE systems requiring high stability and low phase noise. The synthesizers cover the range from 1 GHz to 18 GHz (in bands).
Racal-Dana Instruments
INFO/CARD #154

OCXO
The N26 series of oven-controlled crystal oscillators offers a frequency stability of $\pm 5 \times 10^{-9}$ over 0 to 70 degrees C for a frequency in the range 1 kHz to 20 MHz. It has an aging rate of 5×10^{-10} per day after 30 days. The OCXO comes standard with HCMOS output, requires a +15 Vdc power supply, and has a ±1 ppm frequency adjustment. Available options include TTL and Sine Wave output.
Bliley Electric Co.
INFO/CARD #153

Flexible Cable Assemblies
The HP160S high performance flexible cable assemblies are designed for optimal performance to 40 GHz. SMA, 3.5mm, 2.9mm and 2.4mm assemblies are available. SWRs are 1.35:1 max. for straight 2.9mm and 2.4mm assemblies up to 40 GHz.
Semflex, Inc.
INFO/CARD #152

Ceramic Packaging
Tech-Ceram introduces new capabilities to produce leadless surface mount ceramic packages covering a broad range of applications. Utilization of a cofire multilayer ceramic tape process provides a wide variety of designs including through hole vias, edge wraparound metallization and hermetic configurations.
Tech-Ceram Corp.
INFO/CARD #151

High Rel Trimmer Caps
Voltronics' "SD" series of PTFE solid dielectric trimmers have capacitance ranges up to 1 to 32 pF and are available with either a ceramic case and solder seal or a plain case and epoxy seal. Solid stability under shock and vibration with no electrical path for internal shorts and 40 psi continuous seal make this the highest reliability part available.
Voltronics Corporation
INFO/CARD #150

MMICs
A family of SOIC-8 plastic packaged MMICs, designed for applications in the 1.8, 2.4 and 5.8 GHz ISM bands will be displayed by Harris Microwave.
Harris Microwave Semiconductor
INFO/CARD #149

100 W, Broadband Amplifier
A 0 dBm input provides 100 W output power within Amplifier Research's model 100W1000M1 100 MHz to 1 GHz frequency range. The all solid state design is smaller and more broadband than similar models.
Amplifier Research
INFO/CARD #148

PCX Connector
For applications requiring a smaller size, the PCX series combines the benefits of a lower insertion force, snap-on mating and a size reduction of 30 percent over the SMB series. With a frequency range of 3 GHz, this series is available in straight and right angle cable mounts and with broad mounting configurations.
The Phoenix Company of Chicago, Inc.
INFO/CARD #147

Impedance Matching Software
EEZMatch CAD software, version 2, offers ease of use of various circle capabilities of the Smith Chart[R]: constant-gain (unilateral and bilateral), constant-noise, constant-Q, and stability. Also included is multi-frequency and multi-circuit capabilities for lumped and distributed components.
Besser Associates
INFO/CARD #146

TCXO
Piezo Technology's Model X03016C offers a low profile package with a height of 0.310 inches. The unit has ±4 ppm stability over the temperature range of –45 to 70 degrees C, with aging less than 5 ppm over ten years. The standard frequency is 21.95 MHz with optional frequencies available between 15.0 and 23.5 MHz. The unit features a 1 Vpp sinewave output into 1000 ohms.
Piezo Technology, Inc.
INFO/CARD #145

Simulation Environment
EASi (Environment for Analog Simulation) from Compact Software is an interface operating with MOTIF and X-Windows. The program provides a easy to use interface with SuperCompact, Microwave Harmonica and Super-Spice circuit simulators.
Compact Software
INFO/CARD #144

YIG-Tuned Oscillator
Ferretec has introduced a miniature YIG-tuned oscillator in a surface mount configuration. The package occupies a volume of only 0.7 cubic inches and weighs less than 2 ounces. The oscillator is specifically designed for low-cost, commercial applications. Performance features include fast-switching, full 2-8 GHz coverage and low phase noise.
Ferretec, Inc.
INFO/CARD #143

SMT Glass Trimmer Caps
Sprague-Goodman's Pistoncap[R] trimmer capacitor line has been expanded to include surface mount glass models. Available in vertical and horizontal mount types, the trimmers come in standard and extended capacitance ranges and both sealed and unsealed. Prices start at $2.10 for quantities of 5000.
Sprague-Goodman, Inc.
INFO/CARD #142

Multi-Layer Chip Inductors
Toko America is introducing a new line of multi-layer chip inductors designated the LL2012F series. These inductors feature an ultra-small footprint of 2.00mm x 1.2mm (0805 package) and a height less than 1mm. The new line utilizes Toko's proprietary laminated ceramic material, allowing self-resonant frequencies of over 2 GHz and typical Qs of 29-30 at 400 MHz. The inductors are available from 4.7 to 82 nH with 10 and 20 percent tolerances.
Toko America, Inc.
INFO/CARD #141

Cable Assemblies
The "301" cable assemblies offer low loss at low cost. Insertion loss per 100 ft. of cable is 6.3 dB max. at 0.5 GHz and 41.3 dB max. at 18.0 GHz. The "Workhorse" series of cable assemblies are extremely durable assemblies designed for use at 18 and 26.5 GHz. The cables are designed for strenuous flexing/mating situations and are temperature tested from –55 to +125 degrees C.
Quality Microwave Interconnects, Inc.
INFO/CARD #140

OCXO
The CO-738S series of oven controlled crystal oscillators provide aging rates of 5×10^{-8} per year and temperature stabilities of +/- 1×10^{-9} over 0 to 50 degrees C and $\pm 1 \times 10^{-8}$ over –40 to +75 degrees C. The package measures only 1.5 x 1.25 x 0.86 inches, with frequencies ranging from

32 kHz to 32 MHz (and to 50 MHz with relaxed specifications). Price is $373 each for 100 pieces with delivery in 12-14 weeks.
Vectron Laboratories, Inc.
INFO/CARD #139

Bandpass Filters

Synergy Microwave announces its new line of high performance, low cost, elliptic response bandpass filters. These products are designed to cover the most popular IF frequency bands in use today. The insertion loss is 1.5 dB maximum and VSWR is 1.2 dB typical in the passband. Prices start at $17.85 in quantities of 1-9 for standard relay header packaging and $19.85 in leaded and non-leaded surface mount configurations.
Synergy Microwave Corp.
INFO/CARD #138

Hybrid for Wireless and Cellular

A miniature "drop-in" hybrid, series MCB3, is intended for wireless and cellular applications. The patented Sage "Wireline™" technology was used to develop a bendable, 0.50 inch diameter quadrature hybrid. The hybrid features 60 W power handling and extremely low loss.
Sage Laboratories, Inc.
INFO/CARD #137

Circulator and Isolators

A new line of cellular circulators and isolators operating from 150 to 2000 MHz is available from Ditom Microwave.
Ditom Microwave, Inc.
INFO/CARD #136

Coaxial Elements, Bandpass Filters

Miniature and sub-miniature coaxial transmission line elements from Trans-Tech are meant for those circuits where small size is a major design factor. Ceramic bandpass filters for 200 MHz to 2.5 GHz provide low insertion loss, compact design and mechanical and frequency stability.
Trans-Tech
INFO/CARD #135

Monolithic Log Amplifier

The successive detection logarithmic amplifier from AEL Defense covers the 200 to 6000 MHz range with only three band breaks. The logarithmic amplifiers have input dynamic range of 70 dBm and video linearity of ±1.5 dB. The devices feature small size, low DC power and low cost.
AEL Defense Corp.
INFO/CARD #134

Resistors on Board

Goguen Industries offers printed circuit boards with resistive printed components.
Goguen Industries, Syrtek Div.
INFO/CARD #133

SMT AT-Cut Xtals

An AT-cut crystal in surface mount packaging which possesses all AT-cut motional parameters is being offered by EG&G Frequency Products.
EG&G Frequency Products
INFO/CARD #132

Chip Capacitors

Republic Electronics offers temperature compensating multilayer chip capacitors. Available in 0805 and 1206 sizes, the capacitors have temperature coefficient ratings of NPO through N5600. The terminations are surface mountable.
Republic Electronics Corp.
INFO/CARD #131

Amplifiers

The QBS-141/142 were specifically designed for digital cellular receive base stations. Both devices have less than 1 dB noise at 25 degrees C, two redundant RF signal paths, 90 degree hybrid combiners for soft-fail, reverse voltage protection to 100 V, and a DC power input fuse protected to 2.0 A.
Q-bit Corporation
INFO/CARD #130

SMT VCXO for VHF

TEW North America has developed a sine wave VCXO for VHF use in a surface mount package, the VXS1815C. A nominal frequency is available from 60 MHz to 150 MHz. The 1cc package has a profile of only 4.7mm, enabling mounting in tight places. The components come packaged in a plastic tray and are priced at $20 each in quantities of 10,000.
TEW North America
INFO/CARD #129

SAW Filters

Two new SAW filters are designed for North American Digital Cellular (NADC) IF applications. The PX1001 has a center frequency of 83.16 MHz, and the PX1002 has a center frequency of 86.85 MHz. Both filters have a minimum 3 dB bandwidth of 30 kHz and feature low group delay ripple to minimize intersymbol interference. The filters are available in 13.3 x 6.5mm SMT packages.
RF Monolithics, Inc.
INFO/CARD #128

Glass to Metal Feedthroughs

Hermetic RF, DC and capacitive feedthroughs are designed for use in aluminum housings. They provide highly reliable hermetic sealing in cyclic military and processing temperature environments. Capacitance values are available up to 30,000 pF.
Special Hermetic Products, Inc.
INFO/CARD #127

Emission Microscope

The EDO/Barnes 1630 EMMI emission microscope nondestructively detects current leakage caused by dielectric failures. EMMI can monitor process control during fabrication, verify designs and solve reliability problems.
Barnes Engineering Div., EDO Corp.
INFO/CARD #126

RF Mechanical Switches

A complete line of RF mechanical switches are available from stock from Loral Microwave/Narda. The SEM series of switches are available in SPST through SP6T and Transfer types.
Loral Microwave/Narda
INFO/CARD #125

Traveling Wave Tubes

LogiMetrics' series EPA and CA HP TWTAs feature superior spectral purity and pulsed fidelity, which make then particularly suitable for use in radar, EW simulators, radar cross section systems, communications amplifiers and susceptibility test systems. Reduced size, weight and modular design result in ease of maintenance and handling, along with increased reliability.
LogiMetrics, Inc.
INFO/CARD #124

Sweep/Scalar Analyzer

A sweep/scalar analyzer from Wavetek features a 14 inch color monitor, 2-1100 MHz sweep range and a dual channel scalar network analyzer.
Wavetek Communications Div.
INFO/CARD #123

GaAs FET Switch

A 0.5 to 26.5 GHz, high-speed, ECL, GaAs FET switch provides a minimum 60 dB modulation depth. Switching speed is less than 9 ns (a 6 ns option is available). Transition times are less than 3 ns. VSWR is a maximum of 2.0:1 in both the "ON" and "OFF" states. Control is via balanced ECL.
Custom Microwave Components
INFO/CARD #122

Modulator

A high speed, bi-phase modulator is available in frequency ranges from 10 MHz to 1 GHz. The FP-BPM-1047 is packaged in a standard 0.51 x 0.39 inch flatpack. Amplitude balance of ±0.2 dB and phase balance of ±0.2 degrees are achieved.
ST Olektron Corp.
INFO/CARD #121

DDS

Qualcomm introduces the new Q2230C high-speed direct digital synthesizer with a clock rate to 85 MHz. Four new DACs — the 10-bit Q2510I and Q2515I and the 12-bit Q2520I and Q2525I — compliment Qualcomm's family of DDS ICs.
Qualcomm, Inc.
INFO/CARD #120

Temperature Variable Attenuator

Thermopac[R] from EMC Technology is a temperature variable attenuator available with positive and negative temperature coefficients. Typical temperature coefficient magnitudes are 0.01 dB per degree C over an operating temerature of −55 to +125 degrees C. The attneuator measure 0.030 x 0.122 x 0.145 inches.
EMC Technology, Inc.
INFO/CARD #119

Inductors and Capacitors: L and C for Changing Markets

By Andy Kellett
Technical Editor

Electronic eras are defined in terms of their active devices. First there was the tube era, then the transistor age and now we live in the days of the IC. However, even if you removed every active device from some RF device, you still could tell what era that circuit belonged to simply by looking at the inductors and capacitors left behind. Though their function is the same, today's inductors and capacitors have evolved to work with today's applications.

The switch to surface mount devices has been the most visible change. Two requirements have driven this change, the need for smaller circuits and the need for automated parts placement. Standards for chip capacitor dimensions have emerged, but surface mount inductors are still competing to become de facto standards. "We have a basic inductor with an epoxy enclosure which we are offering, however marketing it is more difficult simply because there's no real standard and everybody is out there promoting their own manufacturing capabilities," says Dick Kolster, Senior Marketing Manager for Siemens Components, Special Products Division. While surface mount components are finding their way deeper into the market, leaded components still find plenty of use. "There are still people utilizing the through-hole devices for those places where the high volume isn't there," says Bill Grajek, Sales Engineer for American Precision Industries, Delevan Division.

Other changes which accommodate today's applications are not visible. Several interviewed manufactures echoed each other when they said, "quality is a given." "The market takes quality as a given, either you are pretty good to very good or you aren't doing any business," says Mark Sullivan, Director of Marketing and R&D for Toko America. "Quality is a given, it used to be a selling point, now its a given," says Tom Rowan, Vice-President of Sales and Marketing for Johanson Manufacturing Corporation.

Manufacturers of components are aware of the trend to reduce the number of final adjustments needed for a product. "An area that our customers focus on is the consistency of the devices, within the lot and from lot to lot. Consistency eliminates the need for tunability," says Dennis O'Toole, Product Marketing Manager for thin film devices at AVX. Makers of trimmer capacitors are aware that their products are not among a production engineer's favorites. "If prices are not to an engineer's liking they will work very hard not to design you into a new design, or to design you out of an existing design," says Johanson's Rowan.

For consumer applications, adjustments are to be avoided, but for precise applications, high quality trimmers are still in high demand. "Our capacitors are stable, precise, multiturn capacitors," says Voltronics Chairman Richard Newman, "they're used in communications equipment, aircraft, missiles, MRI, all sorts of instruments — professional applications." Likewise Johanson Manufacturing concentrates on "the middle and the high end," according to Rowan.

What does an inductor or capacitor's design cycle look like? "The decision to make a new component can be driven by individual customers as well as our own understanding of the market," says Jeff Bartlett, AVX's Director of Product Marketing. The next step may require nothing more than the substitution of one material for another, or the creation of an entirely original design. Voltronic's Newman described how long he expects it will take to get a planned multiturn, surface mount trimmer capacitor into production, "It will take us about four months to finish the design, another couple of months to make some simple samples and another six months to evaluate it — make model runs; it will probably take a year and a half to two years to get it into production."

Many inductor and capacitor manufacturers saw minuscule growth at best over the past three years. The growing wireless market only partly offset setbacks caused by reductions in military spending. For those companies that were not heavily invested in the military market, the market actually showed healthy growth. "We've seen a lot of growth in the last three years and a lot of other people haven't. I think that's because we've come out with a lot of new products and we're in a lot of high growth markets," says Toko's Sullivan. "Over the next year I don't expect to see a bull market, but certainly we have bottomed out," says Jerry Gordon, president of Compex Corporation, "and the commercial market is beginning to make itself felt more now than in the past."

Inductor and capacitor makers will undoubtedly have buyers long into the future. Success depends on selling L and C in forms that fit thriving applications.

RF

For reprints of this report, contact Cardiff Publishing Company at (303) 220-0600. Ask for the Circulation Department

RF software

SPICE Model for IGBTs

Intusoft has produced a SPICE model for IGBTs. The model, represented by a SPICE subcircuit, is the first ever available in a form that can be used for a variety of IGBT devices and for virtually all SPICE versions. The new model accurately simulates: switching losses, on-voltage, turn-on/turn-off delay, rise time/fall tail, active output impedance and other characteristics.

Intusoft
INFO/CARD #210

CAD Program

Version 2.02 of Decade-UX has been released by Cadshare Resources. Decade-UX includes a macro language which can use all of Decade-UX's drafting, grouping and database functions and can incorporate arithmetic, algebraic and trigonometric functions. Decade-UX is priced at $995, and currently operates on System 5 UNIX-based hardware platforms.

Cadshare Resources, Inc.
INFO/CARD #209

Smith Chart® Simulation

ARRL MicroSmith v2.00 is available for IBM PC and compatible computers. MicroSmith is a tool for designing matching networks with fixed or variable LC components, stub matching sections with transmission lines, etc. Version 2.00 supports frequency-dependent terminations. The program requires 272k RAM and DOS 2.0 or higher. Retail price is $39.00 plus $3.00 shipping and handling.

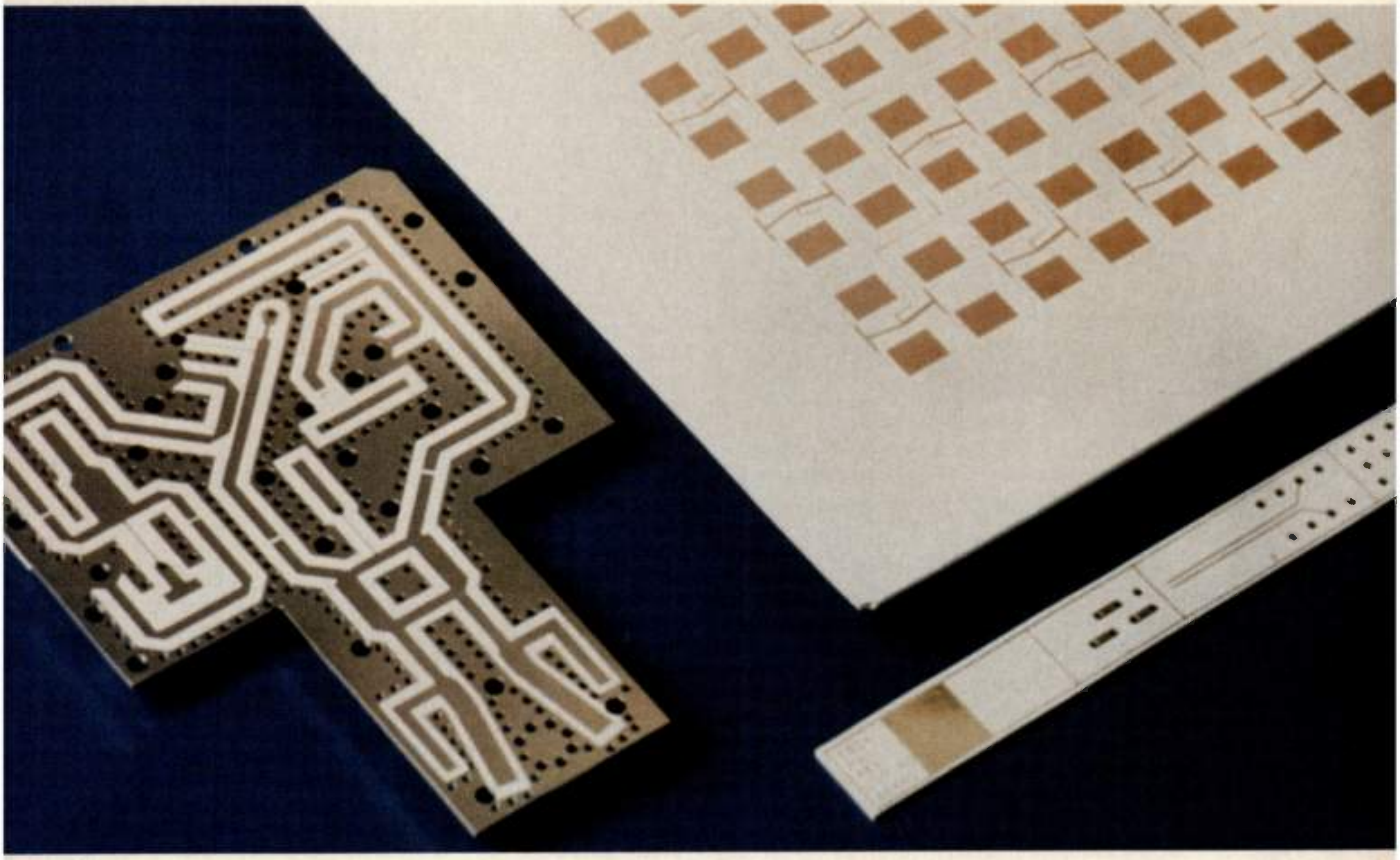
American Radio Relay League
INFO/CARD #208

Measurement Software

EMC Consulting's EMI Commercial Measurement program has been revised to meet the requirements of ANSIC63.4-1991. The program is available in a standard version for the Anritsu MS2601 spectrum analyzer and in custom versions for the HP8568/8566 and the Advantest R3361/3261 A&B spectrum analyzers.

EMC Consulting
INFO/CARD #207

TMM® Microwave Materials. More temperature stable than alumina. But not as temperamental.



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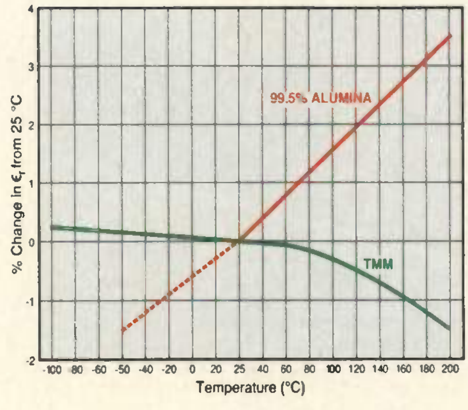

Alumina data based on AFML-TR-72-39. Data below 25°C is extrapolated. TMM data based on Rogers Corp. TR #5049. Contact us for more details.

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INFO/CARD 6

RF literature

Analog Design Software

A 16-page color brochure describing the features and capabilities of the Series IV high-frequency analog design system is available from EEsof. Series IV comprises the OmniSys system simulator, Touchstone and Libra linear circuit simulators, and J-Omega non-linear circuit simulator. The software also includes a file management scheme, "test bench" measurement functions and a layout editor.

EEsof, Inc.
INFO/CARD #200

Semiconductor Data Base

D.A.T.A. is introducing the D.A.T.A. Parametric Access Library (D.A.T.A./P/A/L/™), a CD-ROM version of the D.A.T.A. Digests. Ten component families are available: diodes, optoelectronics, thyristors, transistors, digital, interface, linear, memory, microprocessors and programmable logic. Subscriptions are available for any or all categories. Cost is $295 per category, or $1930 for all.

D.A.T.A. Business Publishing
INFO/CARD #199

Quartz Crystal Catalog

An illustrated, eight-page booklet from JAN Crystals contains descriptions, specifications, and prices on crystals for frequency control, micro-processors, data transmission, telemetry & telecommunication voice. The catalog is available free of charge.

JAN Crystals
INFO/CARD #198

Built-In-Test Note

A one-page application note from Noise/Com describes the use of a noise diode in a balanced amplifier to provide remote measurement of the amplifier's noise figure and gain. Information about Noise/Com's line of noise diodes is also included.

Noise/Com
INFO/CARD #197

Quartz Crystal and Oscillator Guide

The Valpey-Fisher *Expanded User's Guide to Quartz Crystals and Oscillators* contains a whole new section on oscillator terms and definitions in addition to its glossary of crystal terms and definitions. Products described in the guide include high frequency fundamental crystals for filter and VCO applications.

Valpey-Fisher
INFO/CARD #196

GaAs Upconverter MMIC

A detailed description of the use of its GaAs upconverter MMIC in double conversion cable television converters and television tuners is available without charge from Anadigics. The technical bulletin provides a functional description of the ACU50550 and discusses the key parts and functions.

Anadigics, Inc.
INFO/CARD #195

Quartz Crystals and Oscillators

Bliley Electric Co. has published a newly designed catalog of its crystal oscillator and quartz crystal components, Dimensional drawings and specification tables are included for each device type. The catalog is free on request.

Bliley Electric Company
INFO/CARD #194

RF design

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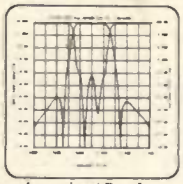
ANALYZE



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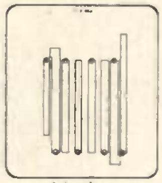

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INFO/CARD 115

RF engineering opportunities

SR. RF/MW DESIGN ENGINEER

Trans-Tech, a leader in electronic ceramics, seeks Sr. Engineer to design Filters and related products, develop testing for materials and circuitry. Experience required: 5-10 yrs. RF/MW Filter Design, Touchstone/Super-Compact, and EE Degree or equivalent. Send resume to: Human Resources, TRANS-TECH, INC., 5520 Adamstown Rd., Adamstown, MD 21710. Fax: (301) 695-7065.

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RF engineering opportunities

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Step size .... <1 Hz
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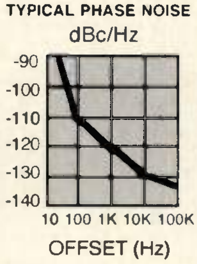
Spurs .... <-45 dBc typ
Size .... 3.75" x 4.78" x 0.89"
Output .... **Quadrature**



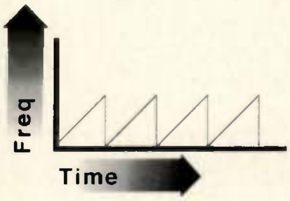
ADS-431
ADVANCED DDS

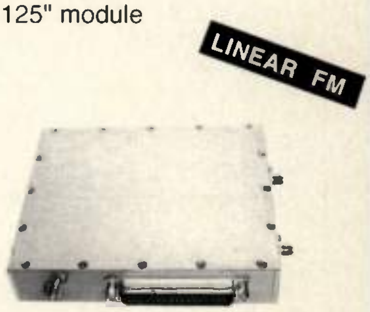
This **LOW COST** module is perfect for ATE, simulation, EW, and all other applications where speed and phase continuity are important.

The VDS-6030 is a low cost C-band solution, with excellent phase noise and low power dissipation. It's also available in L-band.

Bandwidth .... **1 to 230 MHz**
Step size .... <30 Hz granularity
Phase control 12-bit
Switching .... 2 nanosecond update
Spurs .... -55 dBc typicai (CW)

5.25" chassis or 5" x 7" x 1.125" module





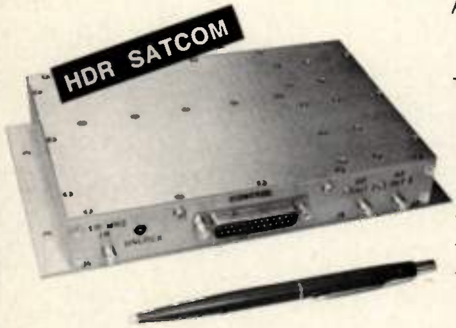
DCP-1
GaAs DDS

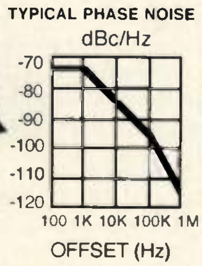
For Synthetic Aperture Radar and other systems requiring state of the art LINEAR FM. Sciteq's DCP-1 provides extended bandwidth, high linearity, phase control, and excellent spectral purity.

This one square inch waveform generator includes digital phase, frequency, and amplitude control, yet it only dissipates 1.5W at f_{ck} max.

SCITEQ
specializes in advanced technology
frequency synthesis,
using direct digital, phase-locked-loop, Arithmetically Locked Loop, and mix/filter designs, plus unique combination architectures that combine the advantages of multiple underlying technologies.

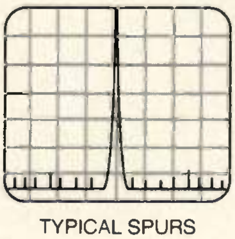
VDS-6030
ARITHMETICALLY LOCKED LOOP





Bandwidth .... **4.65 to 5.25 GHz**
Step size .... 2.5 MHz
Outputs .... 2-channels @ +17 dBm
Total power .... **<5 W**
BIT .... TTL lock indicator
Spurs .... -60 dBc
Reference .... 10 MHz, ext (internal optional)

DDS-1
SYNTHESIZER & MODULATOR



Maximum clock .... **up to 25 MHz**
Frequency control 32-bit
Switching speed .... <1 μsec
Output .... SINE, 1 VP-P
Phase noise .... per the clock
Gating .... >>100 dB ON/OFF, 1-bit toggle
Modulation ...... **digital amplitude, φ, FSK**

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It's compact. It's rugged. It operates in temperatures up to 100°C. It can be tuned up to 1.5 MHz or phase-locked to a reference. Plus, its low mass and low profile make it ideal for surface-mount technology, DILS or flatpaks.

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Course Date: March 17, 1993 8a.m. - 5 p.m.

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Course Date: March 18, 1993 8a.m. - 5 p.m.

☐ Oscillator Design Principles
Course Date: March 19, 1993 8a.m. - 5 p.m.

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