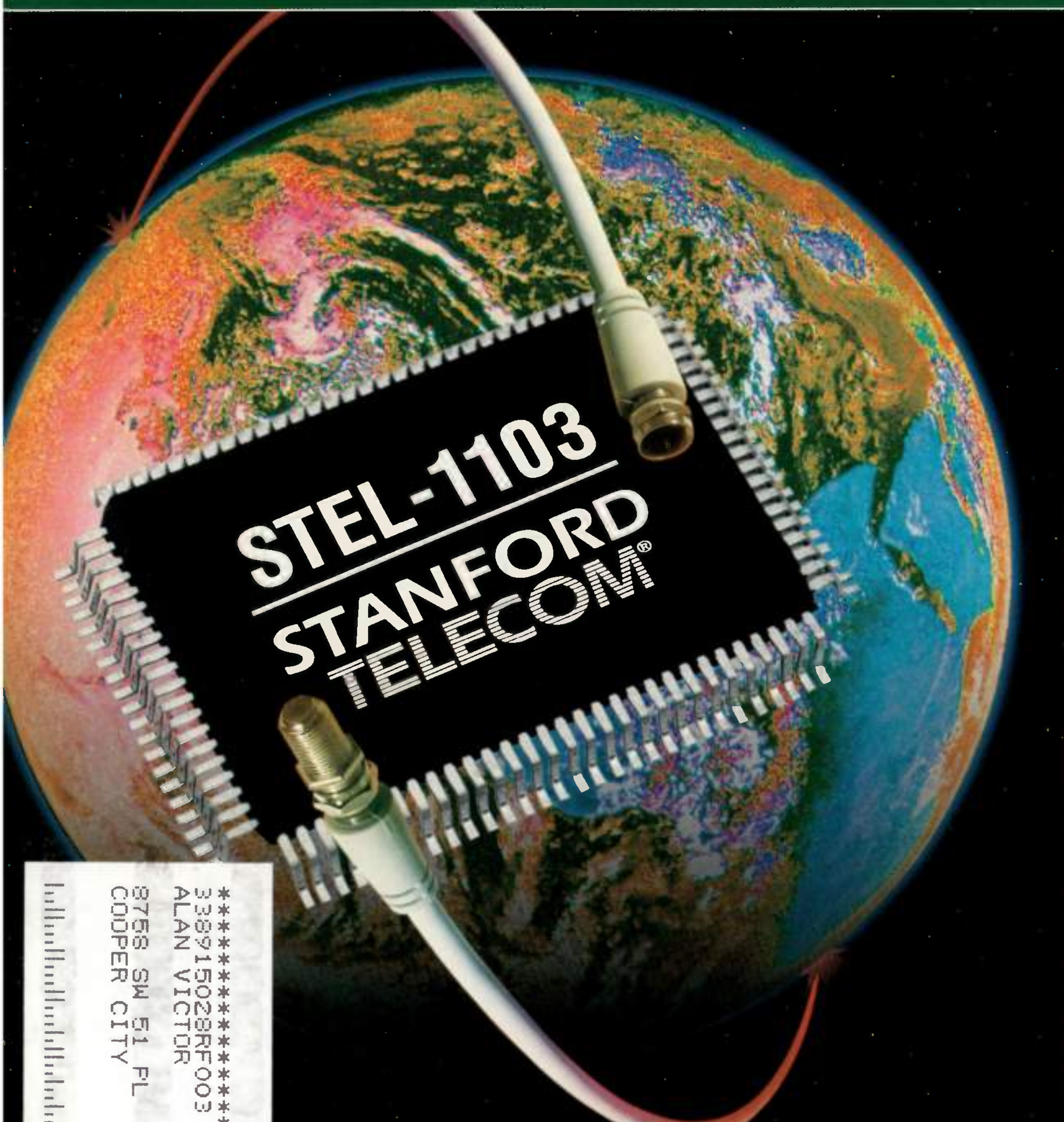
WIRELESS 16 YRS LEADERSHIP

RF design™

engineering principles and practices

March 1995



Cover Story
New ASIC Targets Interactive Cable

Featured Technology
Test Methods

FILTERS

low pass, high pass, bandpass

• less than 1dB insertion loss • greater than 40dB stopband rejection • surface mount • BNC, Type N, SMA available
• 5-section, 30dB/octave rolloff • VSWR less than 1.7(typ.) • rugged hermetically sealed pin models • constant phase
• meets MIL-STD-202 tests • over 100 off-the-shelf models • immediate delivery

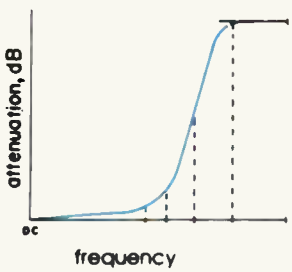
Mini-Circuits RF/IF Surface Mount Designer's Guide features 48 pages of the most up-to-date and complete product and specification information about Mini-Circuits surface mount components. The RF/IF Microwave Handbook is packed with 740 pages of articles, selection guides and detailed specifications for Mini-Circuits components.

Call, write or fax for your free Surface Mount Designer's Guide and Handbook today!

dc to 3GHz from $11^45

low pass, Plug-in, dc to 155MHz

LOW PASS



Model No.	Passband MHz loss < 1dB	Stopband, MHz loss > 20dB	Stopband, MHz loss > 40dB
•★LP-1.9	DC-1.9	3.4-4.7	4.7-200
•★LP-2.5	DC-2.5	3.8-5.0	5.0-200
★LP-5	DC-5	8-10	10-200
★LP-10.7	DC-11	19-24	24-200
★LP-21.4	DC-22	32-41	41-200
★LP-30	DC-32	47-61	61-200
★LP-50	DC-48	70-90	90-200
★LP-70	DC-60	90-117	117-300
★LP-90	DC-81	121-157	157-400
★LP-100	DC-98	146-189	189-400
★LP-150	DC-140	210-300	300-600

dc to 1200MHz

Model No.	Passband MHz loss < 1dB	Stopband, MHz loss > 20dB	Stopband, MHz loss > 40dB
★LP-200	DC-190	290-390	390-800
★LP-250	DC-225	320-400	400-1200
★LP-300	DC-270	410-550	550-1200
★LP-450	DC-400	580-750	750-1800
★LP-550	DC-520	750-920	920-2000
★LP-600	DC-580	840-1120	1120-2000
★LP-750	DC-700	1000-1300	1300-2000
★LP-800	DC-720	1080-1400	1400-2000
★LP-850	DC-780	1100-1400	1400-2000
★LP-1000	DC-900	1340-1750	1750-2000
★LP-1200	DC-1000	1620-2100	2100-2500

All models priced qty. 1-9 ($ea.), Conn. Type P = 11.45, B = 32.95, S = 34.95, N = 35.95
• Exceptions: ★LP-1.9 P = 13.95, B = 34.95, ★LP-2.5 P = 14.95, B = 35.95
On both models, add following to B price: $3.00 for N, $2.00 for S
75 ohm versions available

Surface-mount

dc to 108MHz

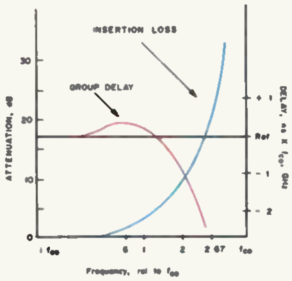
SCLF-5	DC-5.0	8-10	10-200
SCLF-8	DC-8.0	12.5-16.5	16.5-200
SCLF-10.7	DC-11	19-24	24-200
SCLF-21.4	DC-22	32-41	41-200
SCLF-25	DC-25	36-47	47-200
SCLF-30	DC-30	47-61	61-200
SCLF-45	DC-45	70-90	90-200
SCLF-95	DC-95	146-189	189-400

dc to 1200MHz

SCLF-135	DC-135	210-300	300-600
SCLF-190	DC-190	290-390	390-800
SCLF-225	DC-225	340-440	440-1200
SCLF-380	DC-380	580-750	750-1800
SCLF-420	DC-420	750-920	920-2000
SCLF-550	DC-550	800-1050	1050-2000
SCLF-700	DC-700	1000-1300	1300-2000
SCLF-1000	DC-1000	1620-2100	2100-2500

Price: SCLF 21.4~SCLF 420 $11.45 ea. SCLF-8, 10.7, 550, 700, 1000 $12.95 ea. SCLF-5 $14.95 Qty. (1-9)

Flat Time Delay, dc to 1870MHz

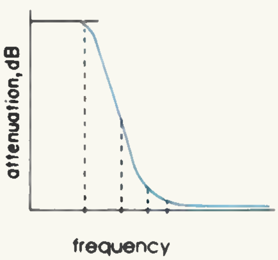


Model No.	Passband MHz loss < 1.2dB	Stopband MHz loss >10dB	Stopband MHz loss >20dB	VSWR Freq. Range, DC thru 0.2fco X	VSWR Freq. Range, DC thru 0.6fco X	Group Delay Variations, ns Freq. Range, DC thru fco X	2fco X	2.67fco X
★BLP-39	DC-23	78-117	117	1.3:1	2.3:1	0.70	4.0	5.00
★BLP-117	DC-65	234-312	312	1.3:1	2.4:1	0.35	1.4	1.90
★BLP-156	DC-94	312-416	416	1.3:1	1.1:1	0.30	1.1	1.50
★BLP-200	DC-120	400-534	534	1.6:1	1.9:1	0.40	1.3	1.60
★BLP-300	DC-180	600-801	801	1.25:1	2.2:1	0.20	0.6	0.80
★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	3740-5000	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 N and SMA connectors.

high pass, Plug-in, 13 to 1200MHz

HIGH PASS



Model No.	Stopband MHz loss > 40dB	Stopband MHz loss > 20dB	Passband, MHz loss < 1dB	VSWR Pass-band Typ.
★HP-25	DC-13	13-19	27.5-200	1.7:1
★HP-50	DC-20	20-26	41-200	1.5:1
★HP-100	DC-40	40-55	90-400	1.5:1
★HP-150	DC-70	70-95	133-600	1.8:1
★HP-175	DC-70	70-105	160-800	1.5:1
★HP-200	DC-90	90-116	185-800	1.6:1
★HP-250	DC-100	100-150	225-1200	1.3:1
★HP-300	DC-145	145-190	†290-1200	1.7:1

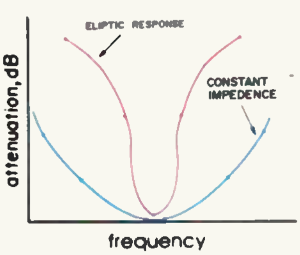
210 to 2200MHz

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

Price, (1-9 qty), all models: plug-in $14.95, BNC $36.95, SMA $38.95, Type N $39.95. For ★HP-25, Add $2 ea. †Loss 1.5 dB max.

bandpass, Elliptic Response, 10.7 to 70MHz

BANDPASS



Model No.	Center Freq. (MHz)	Passband I.L. 1.5 dB Max. (MHz)	3 dB Bandwidth Typ. (MHz)	Stopbands I.L. > 20dB at MHz	Stopbands I.L. > 35dB at MHz
★BP-10.7	10.7	9.5-11.5	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	49.8-70.5	44 & 79	4.6 & 190-1000
★BP-70	70.0	63.0-77.0	58.0-82.0	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	30.0	25-35	1.9 & 210	DC-330
★IF-40	42.0	35-49	2.6 & 300	DC-400
★IF-50	50.0	41-58	3.1 & 350	DC-440
★IF-60	60.0	50-70	3.8 & 400	DC-500
★IF-70	70.0	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.**

Mini-Circuits®

INFO/CARD 1

P.O Box 350166, Brooklyn, New York 11235-0003 (718) 934-4500 Fax (718)332-4661

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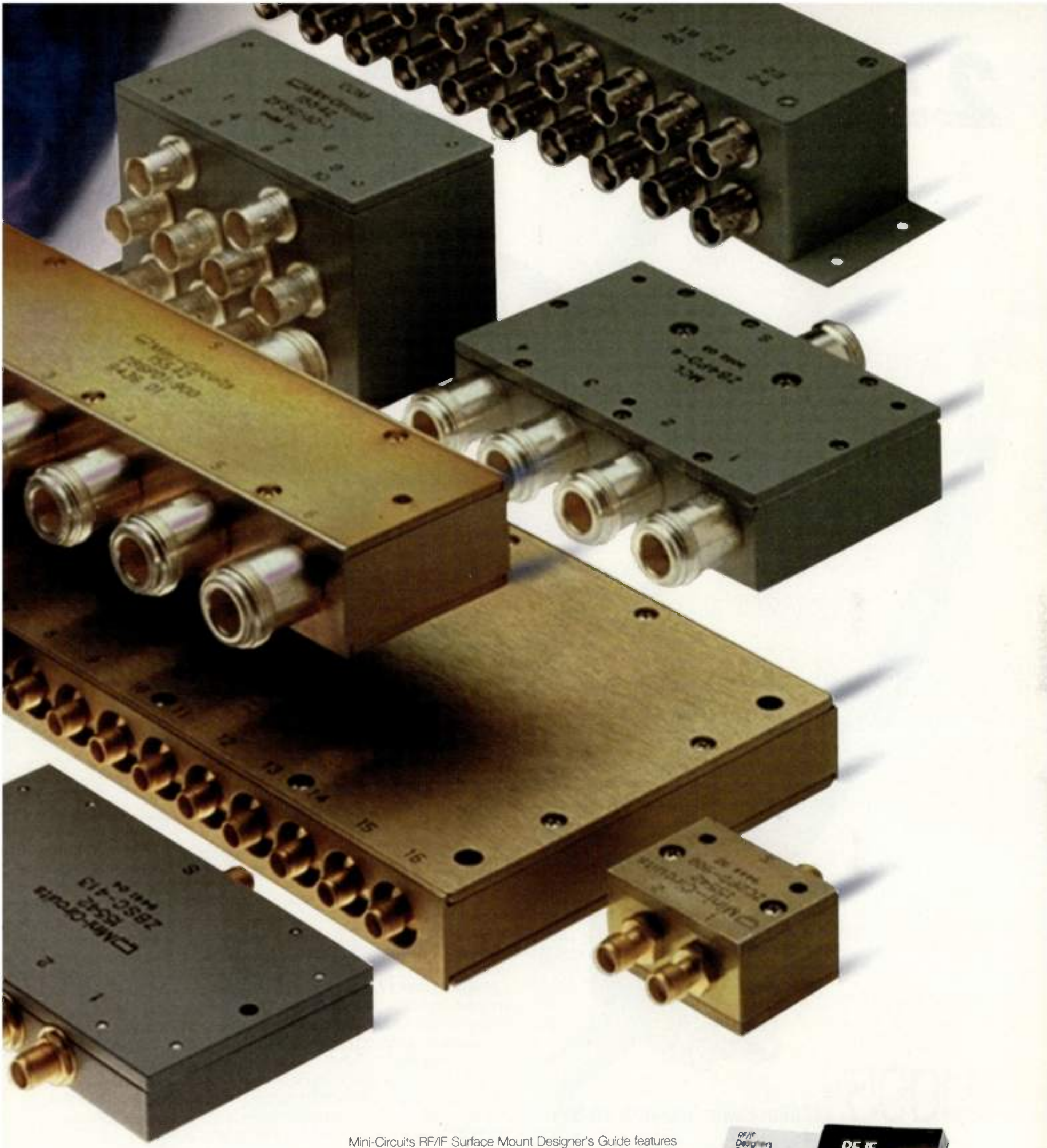
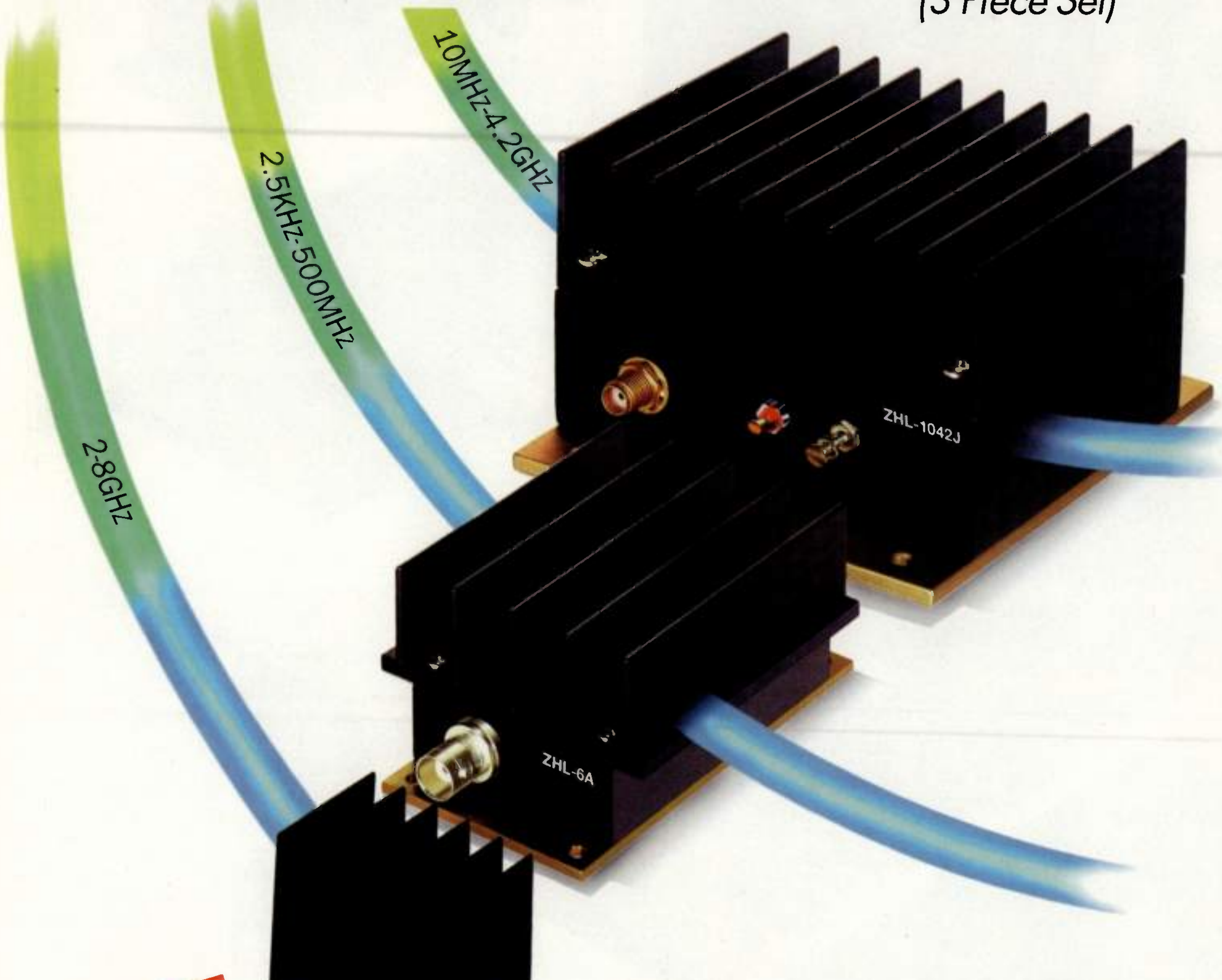
F 200 Rev Orig




RF/IF Designer's Handbook

2.5KHz-8GHz AMPLIFIERS

(3 Piece Set)



WOW!

only $1095 (Set of 3 1-9 qty.)

+20dBm Power Output, 20dB Gain

Model :	Freq. (MHz)	Gain (dB)	Power Output, dBm @ 1dB Compression	DC Power Volt V	DC Power Current mA	Conn. Type	Indiv. Price ($) (1-9 qty.)
ZHL-6A	.0025-500	21	+23	+24	350	BNC	199
ZHL-1042J	10-4200	25	+20	+15	330	SMA	495
ZRON-8G	2000-8000	20	+20	+15	310	SMA	495

Set of 3 Amplifiers # KZHL-318: $1095 (1-9 qty.)

Expand laboratory capabilities and put a full spectrum of power at your fingertips with Mini-Circuits 2.5KHz to 8GHz medium power amplifier set. Each ultra-wideband set contains three individual heat sinked RF amplifiers with at least +20dBm output and overlapping frequency response range capabilities; 2.5KHz to 500MHz, 10MHz to 4.2GHz and 2GHz to 8GHz. Applications for these amplifiers include increasing the signal levels to power meters, spectrum analyzers, frequency counters and network analyzers as well as boosting signal generator outputs.
You can buy these amplifiers individually at Mini-Circuits already low prices, or own the full spectrum set for the money saving price of only $1095 (1-9 qty.) ! To order from stock with a guarantee to ship within one week, call Mini-Circuits today !

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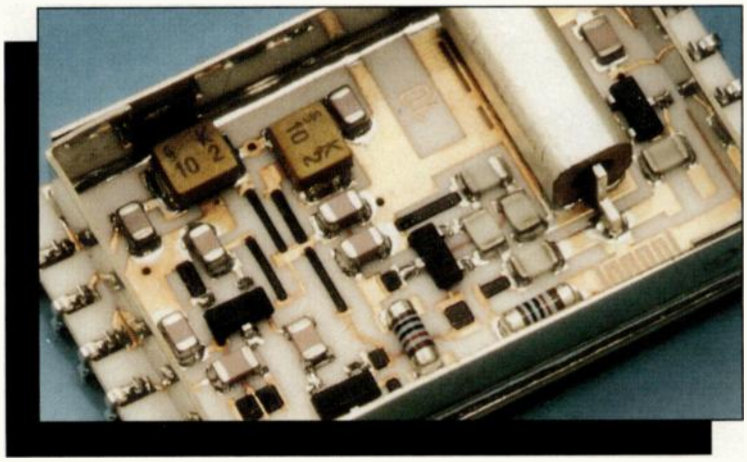
March 1995



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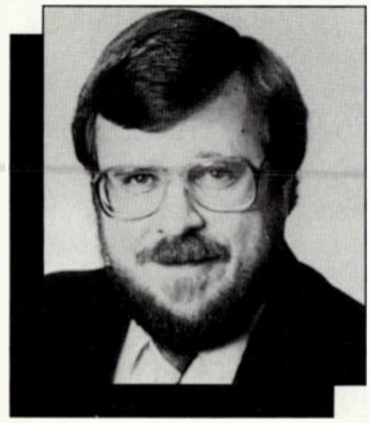
RF editorial

The Changing Role of the RF Engineer



By Gary A. Breed
Editor

At the recent RF Expo West conference in San Diego, our keynote speaker, Dr. Ted Rappaport, gave us a look at the future of wireless. Ted is the founder of the Mobile and Portable Radio Research Group at Virginia Tech. The MPRG is a model of industry/university cooperation, matching research and student education with the present and future needs of the communications industry. His remarks were fascinating, and I'd like to note some of his views in this month's editorial column.

Public demand for wireless services are higher than for any other telecommunications services — As evidence for this, television sets are in nearly every household in the nation, a penetration that was achieved in very few years after its widespread introduction in the late 1940s. More households have televisions than have telephones, and cable TV only reaches about 60 percent of households. Wireless wins!

Cellular growth rate matches that of television's early years — Once cellular service was firmly in place, it really took off. The current growth of cellular is 50 percent a year, an astounding number in any business. New Personal Communications Services (PCS) will simply add new, lower cost choices to fuel further growth in mobile telecommunications.

The challenges of such growth involve a lot more than circuits — Acquiring locations for base stations, connection agreements to the wired telephone infrastructure, and management of spectrum allocations are among the top concerns of the communications industry. These are at least as important as building a new, cheaper handsets. The equipment is essential, of course, but it can't be used if there is no place to put it, physically or in channel space.

The engineer's role is changing dramatically — Low cost, high volume, rapid development and complex technical specifications require new engineering approaches. Today's wireless engineers need to merge design criteria based on propagation, digital signal processing, power management, physical design, reliability, and manufacturability. RF circuit design is but one element of wireless product development.

To add my own comment on this last item: does every engineer need to be a Renaissance Man? No, but those who can achieve this wide range of capabilities will be the most sought after, especially in small-to-mid-size companies. The best alternative (presently used by most companies) is to have close-knit teams whose participants represent all the necessary specialties. Even in a team, each member will be expected to have a familiarity with areas outside his or her area of primary expertise — more than has been expected of engineers in the past. Then we must add an emphasis on cost that makes the personal computer price battles look tame!

New markets bring new opportunities, but they also bring changes. The wireless marketplace is intensely competitive, and that competition extends to the engineers who design wireless products.

GIGA-TRONICS SYNTHESIZED RF SIGNAL GENERATORS

Giga-tronics May Not Be The First Name You Think Of For RF Synthesizers, But It Should Be.

Yes, Giga-tronics is a new name in RF synthesizers.

But this new choice may be your best choice.

Because an RF synthesizer from Giga-tronics is an incredible value. You get proven performance and reliability at a low price, backed by an experienced service organization.

Here are the specifics.

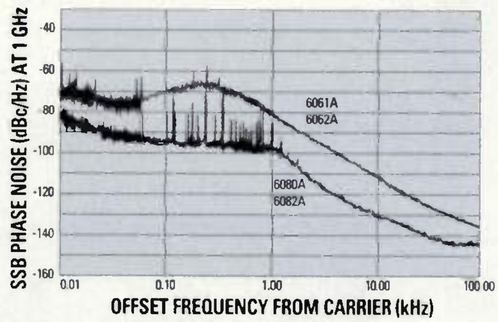
Performance.
Check the charts. Giga-tronics offers four instruments, each with its own unique combination of performance and capability.

The 6061A is ideal for accurate and affordable general purpose RF testing.

The 6062A adds the modulation capability you need to generate complex signals for communications and radar testing.

And the 6080A and 6082A give you the performance and capability to test sophisticated systems accurately and quickly.

It's easy to command the power and capability of each instrument from the front panel. And the low profile, standard rack size and IEEE-488 interface make each instrument ideal for ATE applications.



Reliability.
The John Fluke Manufacturing Company initially developed and introduced this line of synthesized RF signal generators. To date, thousands have performed flawlessly in the field.

The instruments incorporate self-testing,

Giga-tronics offers you a full line of RF synthesizers with proven performance and reliability at low prices.

Specifications	Giga-tronics 6061A	Giga-tronics 6062A	Giga-tronics 6080A	Giga-tronics 6082A
Frequency Range Switching speed	.01 to 1050 MHz <100 ms	.1 to 2100 MHz <100 ms	.01 to 1056 MHz <100 ms	.1 to 2112 MHz <100 ms
Spectral Purity* Spurious Subharmonics	<-60 dBc None	<-54 dBc <-45 dBc	<-100 dBc None	<-94 dBc <-45 dBc
Phase Noise* @ 20 kHz offset	<-117 dBc/Hz	<-110 dBc/Hz	<-131 dBc/Hz	<-125 dBc/Hz
Residual FM* (Bandwidth)	<12 Hz (.5 to 3 kHz)	<24 Hz (.5 to 3 kHz)	<1.5 Hz (.3 to 3 kHz)	<3 Hz (.3 to 3 kHz)
Output Range* Accuracy Reverse Power Protection	+13 to -147 dBm ±1 dB >127 dBm 50 Watts/50 Vdc	+13 to -147 dBm ±1.5 dB >-127 dBm 25 Watts/25 Vdc	+17 to -140 dBm ±1 dB >127 dBm 50 Watts/50 Vdc	+13 to -140 dBm ±1 dB >-127 dBm 25 Watts/25 Vdc
Amplitude Modulation Depth Distortion @ 30%	0–99.9% <3%	0–99.9% <3%	0–99.9% <1.5%	0–99.9% <1.5%
Frequency Modulation Max. Deviation* Distortion	100 kHz <1%	400 kHz <1%	4 MHz <1% @ 50% Dev.	8 MHz <1% @ 50% Dev.
Phase Modulation Max. Deviation*	NA	40 Rad.	40/400 Rad.	80/800 Rad.
Pulse Modulation On/off Rise/fall time Minimum Pulse Width	NA	>80 dB <15 ns <2 µs	>40/60 dB <15 ns (Typ 7.5 ns) <30 ns	>80 dB <15 ns (Typ 7.5 ns) <30 ns
Internal Modulation Source Level Range Waveforms Programmable	400, 1000 Hz NA Sine Yes	100, 1000 Hz NA Sine Yes	0.1 Hz to 200 kHz 0 to 4 Vpk Sine/Sq/Tri/Pulse Yes	0.1 Hz to 200 kHz 0 to 4 Vpk Sine/Sq/Tri/Pulse Yes
Memory Locations (NVM)	50 Full Function	50 Full Function	50 Full Function	50 Full Function

*Specifications for the 6061A and 6080A are at 1 GHz, and specifications for the 6062A and 6082A are at 2 GHz. Phase noise is typical for the 6061A and 6062A.

internal diagnostics and modular design for easy fault isolation.

Service.

If a problem occurs, our technical support staff can often help you find and fix it over the phone.

If you need to return an instrument for service, we can take care of it at our factory in California, or at one of our worldwide service centers.

Experience.

Granted, Giga-tronics may be a new name for RF synthesizers, but it's a trusted name to microwave test professionals.

Giga-tronics has a 14-year history of building test and measurement gear for the most demanding requirements. We've shipped thousands of instruments for use in testing communications, radar and EW systems.

Value.

What makes a Giga-tronics RF Signal Generator such an incredible value is its blend of proven performance and reliability at a low price.

Take a closer look at Giga-tronics. The new choice for RF synthesizers may be your best choice.

Call us toll free at **800 726 GIGA** (4442). We'll send you more information and arrange for a demonstration.

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RF design

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RF calendar

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20-24 **The 11th Annual Review of Progress in Applied Computational Electromagnetics**
Monterey, CA
Information: Ray Luebbers, Department of Electrical Engineering, Pennsylvania State University, University Park, PA 16802. Tel: (814) 865–2362. Fax: (814) 865–7065.

21-23 **Nepcon Electronics**
Birmingham, England
Information: Reed Exhibition Companies, 383 Main Avenue, Norwalk, CT 06851. Tel: (203) 840–5398. Fax: (203) 840–9398.

26-29 **Fifth IEE Conference on Telecommunications**
London, UK
Information: ICT 95 Secretariat, IEE Conference Services, Savoy Place, London WC2R 0BL, UK. Tel: 44–071–344 5478/5477. Fax: 44–071–497 3633.

28-31 **Internepcon/Microelectronics Indonesia**
Jakarta, Indonesia
Information: Reed Exhibition Companies, 383 Main Avenue, Norwalk, CT 06851. Tel: (203) 840–5398. Fax: (203) 840–9398.

28-29 **95 Mid-Lantic Electronics Show and Conference**
King of Prussia, PA
Information: Judith Ginsberg, Show Manager. Tel: (610) 828–2271.

April

3-8 **Hannover Fair 95**
Hannover, Germany
Information: Hannover Fairs USA, Inc. 103 Carnegie Center, Princeton, NJ 08540. Tel: (609) 987–1202. Fax: (609) 987–0092.

11-14 **Internepcon/Semiconductor Korea**
Seoul, Korea
Information: Reed Exhibition Companies, 383 Main Avenue, Norwalk, CT 06851. Tel: (203) 840–5398. Fax: (203) 840–9398.

19-21 **4th International Conference & Exhibition on Multichip Modules**
Denver, CO
Information: International Symposium on Multichip Modules 95, Conference Management, ISHM–The Microelectronics Society, 1850 Centennial Park Drive, Suite 105, Reston, VA 22091. Tel: (800) 535–ISHM or (703) 758–1060. Fax: (703) 758–1066.

23-26 **IEEE Instrumentation/Measurement Technology Conference**
Waltham, MA
Information: Robert Myers, 3685 Motor Ave., Suite 240, Los Angeles, CA 90034. Tel: (310) 287–1463. Fax: (310) 287–1851. or Dan Sheingold, Analog Devices, P.O. Box 280, Norwood, MA 02062. Tel: (617) 461–3294. Fax: (617) 329–1241.

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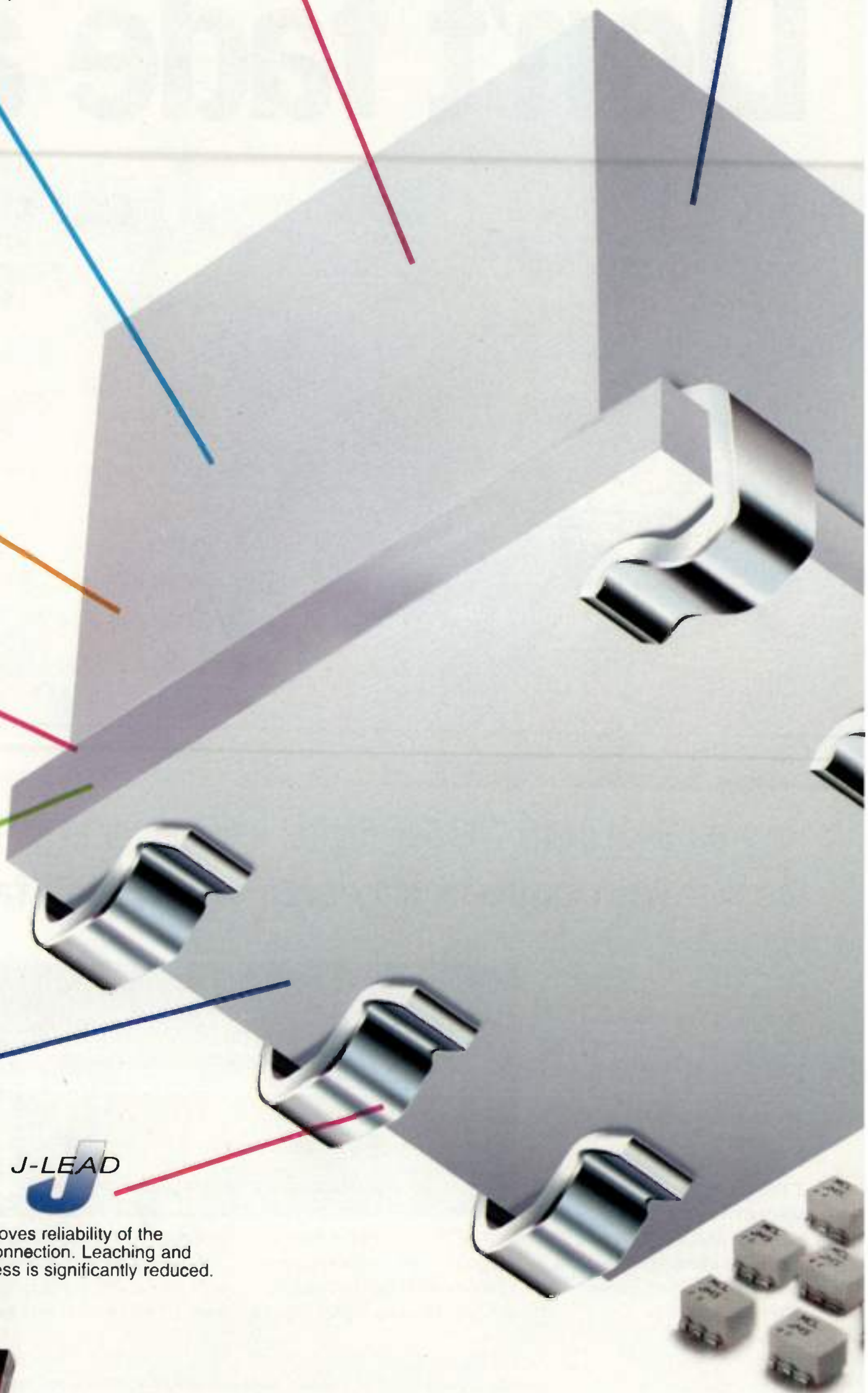

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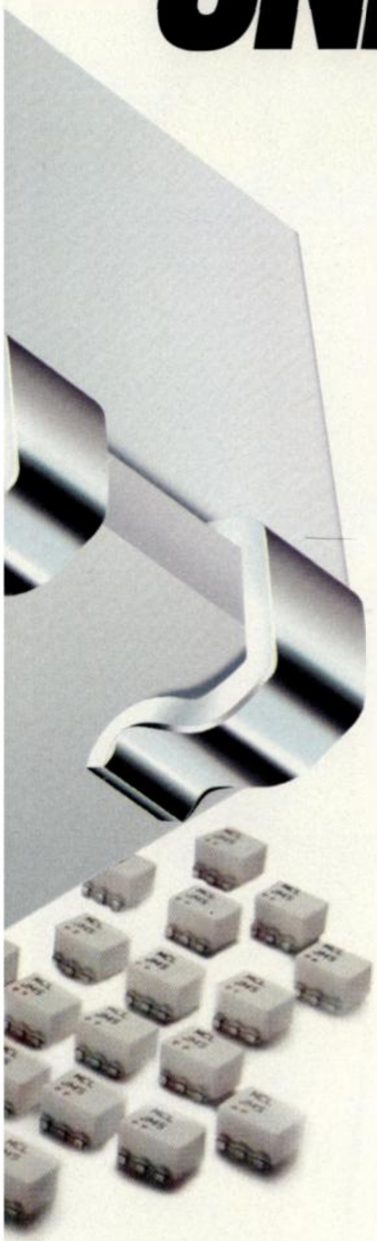

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JMS-1MH	+13	2-500	DC-500	5.75	60	45	9.45
JMS-1H	+17	2-500	DC-500	5.90	50	50	11.45
JMS-2L	+3	800-1000	DC-200	7.0	24	20	7.45
JMS-2	+7	20-1000	DC-1000	7.0	50	47	7.45
JMS-2LH	+10	20-1000	DC-1000	6.5	48	35	9.45
JMS-2MH	+13	20-1000	DC-1000	7.0	50	47	10.45
JMS-2H	+17	20-1000	DC-1000	7.0	50	47	12.45
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RF courses

Power Hybrids
April 3-5, 1995, Los Angeles, CA
Advanced Communication Systems Using Digital Signal Processing
April 3-7, 1995, Los Angeles, CA
Hybrid Microcircuit and Multichip Module Packaging Technologies
April 10-13, 1995, Los Angeles, CA
Multichip Module Design
April 24-26, 1995, Los Angeles, CA
Information: UCLA Extension, Engineering Short Courses, 10995 LeConte Ave., Ste. 542, Los Angeles, CA 90024. Tel: (310) 825–1047. Fax: (310) 206–2815.

Error Correcting Codes/Digital Storage Systems
March 27-30, 1995, Palo Alto, CA
Plasma Etching: Processes & Technologies
April 24-26, 1995, 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.

Wireless Digital Communications
March 20-24, 1995, Switzerland
Information: CEI-Europe/Elsevier, Mrs. Tina Persson. Tel: (46) 122–175–70. Fax: (46) 122–143–47.

Optical Systems Fundamentals
March 20-21, 1995, Washington, DC
Modern Optical Systems Design
March 22-24, 1995, Washington, DC
The Cellular Telephone System
March 27-29, 1995, Washington, DC
Spread Spectrum Communications Systems: Commercial and Government Applications
March 27-31, 1995, Washington, DC
Global Positioning System: Principles and Practice
April 3-6, 1995, London, UK
Video Communications: An Introduction
April 3-6, 1995, Washington, DC
Communication Satellite Systems: The Earth Station-A Practical Approach to Implementation
April 10-13, 1995, Washington, DC
Satellite Communication Engineering Principles
April 18-21, 1995, Washington, DC
Information: The George Washington University, Continuing Engineering Education, Academic Center, Room T-308, 801 22nd Street, N.W., Washington, DC 20052. Tel: (202) 994–6106 or (800) 424–9773. Fax: (202) 872–0645.

Wireless and Mobile Networks: From Theory to Practical Implementation
March 27-28, 1995, Denver, CO
Information: Lori Milhaven, Project Leader, International Institute for Learning, 110 East 59th St., Sixth Floor, New York, NY 10022-1380. Tel: (800) 325–1533 or (212) 758–0103. Fax: (212) 909–0558.

DSP Without Tears
April 5-7, 1995, San Jose, CA
April 24-26, 1995, Chicago, IL
Information: Z Domain Technologies, Inc., 325 Pine Isle Court, Alpharetta, GA 30202. Tel: (800) 967–5034, (404) 587–4812. Fax: (404) 518–8368.

New Research Center to Enliven US Share of Electronics Market

Partners in a new $40 million engineering research center expect to help rebuild the US share of the world electronics market by making dramatic breakthroughs in ultra low-cost, smaller and more powerful electronics. Established at the Georgia Institute of Technology and supported by the National Science Foundation, the Low-Cost Electronics Packaging Research Center will also blaze a trail for enhanced collaboration between industry and universities. Thirty-five companies from all segments of the U.S. electronics industry have expressed interest in participating in the research and education efforts.

The United States leads the world in developing new technologies, but companies in other nations often turn these developments into profitable consumer products and dominate the resulting markets. To maintain or expand their business, US manufacturers will have to compete in the low-cost electronics product market, now dominated by other nations. Development of new products and the restoration of lost markets should mean more jobs in the United States. If US companies can increase their market share from 33 to 40 percent by the year 2004, two million new jobs could be added in electronics manufacturing and another two million in support industries.

Companies participating in the center's work will pay to share the knowledge generated in some or all of the technology areas. Support for the center will come over the next ten years in matching amounts from the National Science Foundation, the State of Georgia, the industry partners, the Department of Defense and Sematech.

1994 US Electronics Factory Sales Total $342.7 Billion

Factory shipments of electronics equipment, components, and related products totaled $342.7 billion for year-end 1994, representing an increase of more than 13 percent over 1993's sales of $303.2 billion, according to preliminary data released by the Electronic Insdustries Association (EIA). Telecommunications manufacturing led industry growth in 1994 with $53.4 billion in sales, up 29 percent over 1993. The electronics components sector totaled $87 billion, up 19.3 percent. Computers and peripherals showed healthy improvement, hitting $70.7 billion for year-end 1994, up 17 percent from year-end 1993. Consumer electronics grew by 7.8 percent, electromedical equipment by 2 percent, and other related products improved slightly, with a growth of 2.9 percent over 1993. Only one sector fell during the reporting period. The defense communications area had total sales of $28.9 billion, a loss of 6.9 percent from the 1993 sales of $31 billion.

Call for Papers

The Advanced Technology Acquisition, Qualification, and Reliability workshop is calling for papers for the August 15-17, 1995 workshop in Newport Beach. Topics of interest include advanced packaging concepts, multichip modules, generic qualification, chip/system reliability, radiation effects, fabrication line certification, testability and self test (BIT/BIST), GaAs space qualification, acquisition changes, known good die, and plastic packaging. Authors should submit five copies of a one-page abstract to the workshop coordinator by May 31, 1995 for review by the program committee. Abstracts should include author's name, complete address, and telephone number. Abstracts will be selected based on technical merit, supporting test results, and overall suitability to advanced silicon and GaAs technology issues. Abstracts should be sent to Dorothy Connor, Workshop Coordinator, General Technical Services, Inc., 3100 Route 138, Wall Township, NJ 07719, tel: (908) 544–3231, fax: (908) 389–3992.

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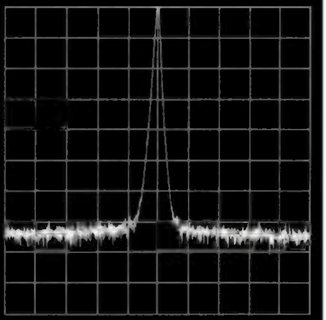

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Department of Commerce's Bureau of Export Administration and the Department of Energy's National Laboratories System, has established a consortium to reduce manufacturing costs and improve the worldwide marketability of I/O pins. The pilot effort hopes to strengthen the competitive position of the US semiconductor industry. The new NIST-led consortium will seek ways to improve energy efficiency; pin cleaning and finishing methods; manufacturing processes; sources of long-term financing; technical training; cooperative buying methods; and supplier relations. Companies who join the consortium are also asked to sign cooperative research and development agreements with NIST.

Call for Papers

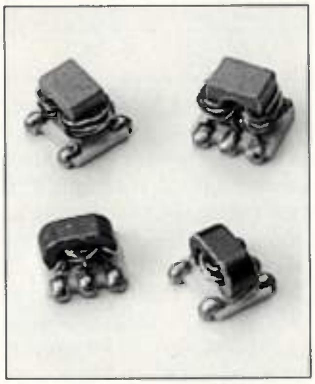
The International Conference on Electromagnetics for Advanced Applications now invites submissions of papers addressing all kinds of advanced applications. Suggested topics are electromagnetic measurements; electromagnetic modeling of devices and circuits; electromagnetic packaging; electromagnetic properties of materials; EMC/EMP; frequency selective surfaces; inverse scattering; mathematical models and simulation; microwave antennas; phased and adaptive arrays; printed and conformal antennas; radar cross section and asymptotic techniques; radar imaging; radomes; and random and non-linear electromagnetics. Authors should send four copies of a 1-2 page summary for consideration by March 31, 1995. Abstracts should be sent to Professor Mario Orefice, Dipartimento di Elettronica, Politecnico di Torino, Corso Duca degli Abruzzi 24, 10129 TORINO, Italy.

CDMA Group Begins International Expansion

The CDMA Development Group has created a new international working group within the organization to address CDMA development issues outside the US. Through this new working group, manufacturers, carriers, and government entities can participate in the CDG efforts. As part of the new international group, the CDG has already formed a Korean working group and is currently discussing membership with carriers and manufacturers in other areas, including countries in Southeast Asia, South America, the Middle East, North America, and Europe. The CDG group will also publish specifications, requirements, documents, and testplans that the group has developed over the last year. The CDG recently completed a specification for higher data rate 13 kbps voice coding.

Business Execs Lead the Pack in Cellular Use

Business leaders under the age of 35 have embraced the power of cellular communications faster than their elders, according to a recent survey by Ameritech. Seven out of ten business executives under 35 own or use cellular phones, compared with six out of ten executives aged 35-54, and four of ten executives aged 55 and older.



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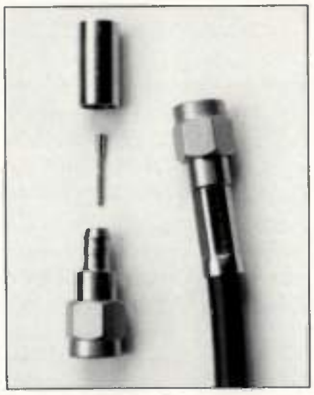

New 3-Piece SMA Connectors

Fewer component piece parts and crimpable center contacts reduce installation costs up to 70%. Stainless steel construction ensures reliable, consistent performance. These connectors are available for flexible RG/U and double braided RD-316 cables .They can be crimped using industry standard tooling.

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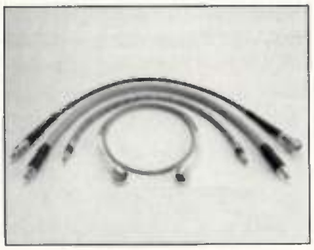

Nomex Braided Test Cables with Interchangeable Connectors

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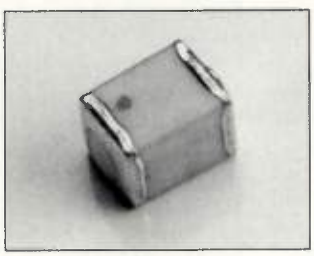

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INFO/CARD 19

Business Briefs

MacNeal-Schwendler and NASA Langley Sign Cooperative Technology Transfer Agreement – The MacNeal-Schwendler Corporation has signed a cooperative agreement with NASA-Langley Research Center for the inclusion of NASA's Finite Element Interface Technology into MSC/NASTRAN. The agreement will lead to the implementation of the technology in MSC/NASTRAN.

Vari-L Announces Results of 1994 Customer Order Analysis – Vari-L Company has published its analysis of 1994 customer orders. Total commercial orders increased 59 percent, while military orders declined 2 percent. 1994 commercial orders represented 63 percent of the total, up from 51 percent from 1993. Domestic orders dropped from $6.66 million to $6.1 million. The overall decrease in domestic orders is attributable to a 24 percent decrease in military orders that was offset by a 14 percent increase in commercial orders.

Jay-El Products Acquires Dynatech Microwave Technology – Jay-El Products, Inc. has acquired the assets of Dynatech Microwave Technology, Inc., a manufacturer of electromechanical switches and microwave/RF components. The company will now be the DMT division of Jay-El Products, and will move its manufacturing operations to Carson, CA.

World's First High Temp Superconductor Transformer to be Built by American Superconductor and ABB – The world's first transformer to use high temperature superconductors will be designed and built by ABB, with technology developed by American Superconductor Corporation. The development is a significant step toward HTS transformers that are expected to constitute a worldwide market of several billion dollars per year.

AEG and Andrew Join Forces to Develop Radar-Based Automatic Train Controls – AEG Transportation systems, Inc, and Andrew Corporation have teamed up to develop a communications-based advanced automatic train control system. When completed, the spread spectrum radio-based system could replace the traditional track circuit approach which employs electrical track shorting for vehicle detection, positioning, and speed code transmission.

Ericsson Unveils Smallest Digital Cellular Phone – The world's smallest digital cellular telephone was introduced by Ericsson, Inc. at the Winter Consumer Electronics Show in Las Vegas, January 6-9. Called the DH338, the digital phone weighs 6.9 ounces and is completely menu-driven. The DH338 takes advantage of TDMA, and operates in both digital and analog modes so that it can be used by all cellular phone customers.

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INFO/CARD 21

Contracts

Xircom and NEC Announce Technology Agreement – Xircom, Inc. and NEC Corp. have signed a joint licensing agreement providing Xircom Netwave wireless networking technology to NEC to develop and manufacture wireless LAN chipsets. Xircom grants NEC rights to develop and market wireless LAN chipsets for customers developing wireless LAN products.

Rohde & Schwarz Sells Trunked Radio System to Fire Brigade of Damascus – The Fire Brigade of Damascus will take delivery of a modern trunked radio system, the first of its kind to be installed in the region. The system will be supplied and installed by R&S BICK Mobilfunk, a German daughter company of Rohde & Schwarz.

Motorola Wins $23 Million Contract for Malaysia's First Digital Cellular System – Motorola's International Cellular Infrastructure Division has been awarded a $23 million contract to install the base station system for Malaysia's nationwide Global System for Mobile Communications Network. It will be the first fully digital cellular network in Malaysia.

Conductus Awarded Navy Contract for Superconductive Geophysical System – Conductus Inc. has been awarded a contract by the Navy to develop a geophysical surveying system based on high-temperature superconducting magnetic sensors. The project spans two years with a total funding of about $.75 million.

PCS Group Signs Agreement with Ericsson for PCS Equipment – The PCS Group has signed an agreement with Ericsson Radio Systems to supply Personal Communications Services network equipment for its clients. The agreement enables the PCS group's clients to obtain timely delivery of Ericsson Equipment.

Northern Telecom Orders $1.4 Million in Test Equipment from Tektronix – Northern Telecom has ordered $1.4 million in test equipment from Tektronix, Inc. Northern Telecom will use the equipment to produce its new comprehensive PCS system.

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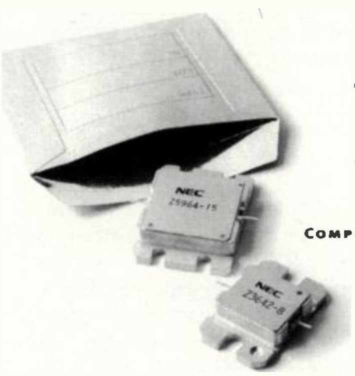

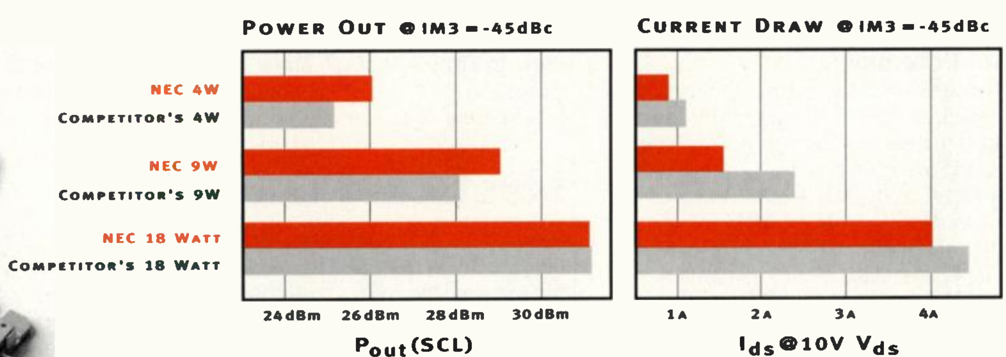

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INFO/CARD 24

WRH

Market Demanding Real Usefulness of RF-Linked Data Devices

By Andy Kellett
Technical Editor

Ten years ago fax machines and cellular phones were technological exotica used only by people who could really see a monetary benefit from using them. Will RF-linked data devices follow the history and fax machines and cellular phones and become ubiquitous, or will they remain the tools of a few specialized users?

According to Brian Button, Vice President of Marketing for RF modem maker Proxim, the markets into which his company sells, "... are all markets where there is a large payback in being mobile and networked." Button cites wireless patient record keeping as an example, "you can measure the time and money saved by not having to re-enter patient information." Wireless meter reading is another area where RF-linking makes economic sense. "They get more data, more accuracy, and it's cheaper than a meter reader," says Morris Engelson, an RF Engineer for American Meter Co. Button notes that wireless technology can even increase revenues by bringing the point of sale to customers. For instance, "You can move the beer cart out to the pool," says Button.

Market Potential

Will more broad markets emerge for wireless data links? "That really is the $64,000 question in our industry," says Proxim's Button.

The key to acceptance of wireless data links, according to Earl McCune Jr, a consultant in the RF communications field, is, "Cost, cost, cost." McCune says that while the price of RF devices is coming down as compared to RF devices in the past, the most important comparison is between RF devices and the wired devices with which they compete.

The expansion of wireless data transmission out of niche markets depends not only on bringing down price, but also on changing user behavior says Proxim's Button. An application often predicted for wireless LANs is networking during meetings. "Right now you really don't see too many people going to meetings and taking notes on their notebook PCs," says Button, "that may be different in two or three years."

Technological Issues

A protocol for wireless data transmission is currently being developed by the IEEE's 802.11 committee. While not yet complete, it is known that the standard covers transmission by infrared, direct-sequence spread spectrum, and frequency-hopping spread spectrum. The protocol also specifies two data rates, 1 Mbps and 2 Mbps. "People are really going to have to stretch to reach that 2 Mbps," says McCune. According to McCune data rates are currently being constrained by the spreading required to get the processing gain required by FCC regulations. "To get the processing gain requires spreading the signal all the way to the band edges," says McCune.

Range versus power consumption is another characteristic that manufacturers are working to improve. Reduction of intermodulation products by increasing the intercept points of the analog portions of data radios, "is happening very nicely," says McCune, "unfortunately most of that comes at the expense of power consumption."

Many of the wireless applications for use with personal computers will take the form of PC Cards (see sidebar). Wireless LAN cards, pagers and other RF devices are being designed to fit in the slim form of a PC Card. The problem some of these manufacturers have is segregating the digital and RF signals. "Interference is a real problem," says Tom Parrish, editor of *IC Card Systems & Design*, "These guys are all trying to figure out how to control all the signals being transmitted inside the computer."

To help the many non-RF engineers designing RF functions into their products, some RF companies are selling transmitting and receiving modules that are almost "plug and play". "That was the main focus for the development of the ASH receiver and HX transmitters," says Frank Perkins, Vice President of Marketing for RF Monolithics. "These have all the RF guts in them, you put data in and get modulated RF out for the HX transmitters, and for the ASH receiver you put RF in and get data out." says Perkins.

RF-linked data devices have yet to break out of niche applications, but the demand for mobile computing, and the increasing ease with which system designers can incorporate RF functions, may change that soon. *RF*

Meet the PC Card

As more RF engineers help design devices to be used with portable computing devices, more come into contact with the PCMCIA (Personal Computer Memory Card International Association) standard. Recently renamed PC Card, this standard specifies an interface whose function is analogous to that of ISA or EISA bus standards. Mechanically, cards come in three sizes, type I (3.3 mm thick), type II (5.0 mm thick), and type III (10.5 mm thick). Most wireless PC Card devices use "extended" cards, which use a "bubble" attached to the end of the card. The "bubble" contains some of the bulkier RF components which will not fit in the actual thin card. The standard also specifies pin assignments for the PC Card's 68-pin connector. Electrically, the PC Card standard allows both 5 V and 3.3 V operation, with a mechanical key to prevent placing a low-voltage card in a high voltage slot. In the past, PC Cards used eight- and sixteen-bit busses, limiting their speed, but the latest revision of the standard features a 32-bit bus called CardBus.

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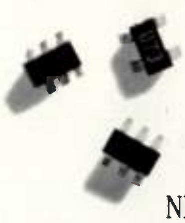

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INFO/CARD 25

Bridge Method for Measuring Amplitude Intermodulation Distortion

By Nina Krikorian
Matrix Test Equipment Inc.

This article describes a distortion measurement method that greatly reduces contributions from the signal source and the spectrum analyzer. This method utilizes a bridge technique to cancel the test signals, leaving the distortion products of the device under test (DUT).

For designing and manufacturing multi-channel communication networks, the distortion products of the components must be measured. Distortion in an RF amplifier can be measured with no more than a signal source and a spectrum analyzer. One or more signals are fed into an amplifier and the amplifier output is displayed on the spectrum analyzer. Those signals not present at the input of the amplifier, but which are a result of the signals, are distortions.

There are two critical limitations that make these measurements difficult. The signal sources may not be completely distortion-free. It is often necessary to design separate combining networks with amplifiers and pads to minimize these products. Second, it is often necessary to amplify the signal source before applying it to the DUT, and amplifying a signal inevitably raises the noise level and degrades the signal-to-noise ratio.

The spectrum analyzer may generate its own distortion. In many applications the distortion from the spectrum analyzer may be overcome by the use of a bandpass filter that passes only the distortion product of interest. This causes some problems in calibration and is impractical in the case of closely-spaced multiple signals.

The bridge method for making distortion measurements solves these problems. This method is similar to that used in the first part of a feedforward amplifier. The circuitry of both the bridge method and feedforward amplifiers extract only the distortion products. The feedforward amplifier design uses these products to effect distortion cancellation at the output, while the bridge method cancels the test signals at the amplifier output, allowing measurements of only the distortion products.

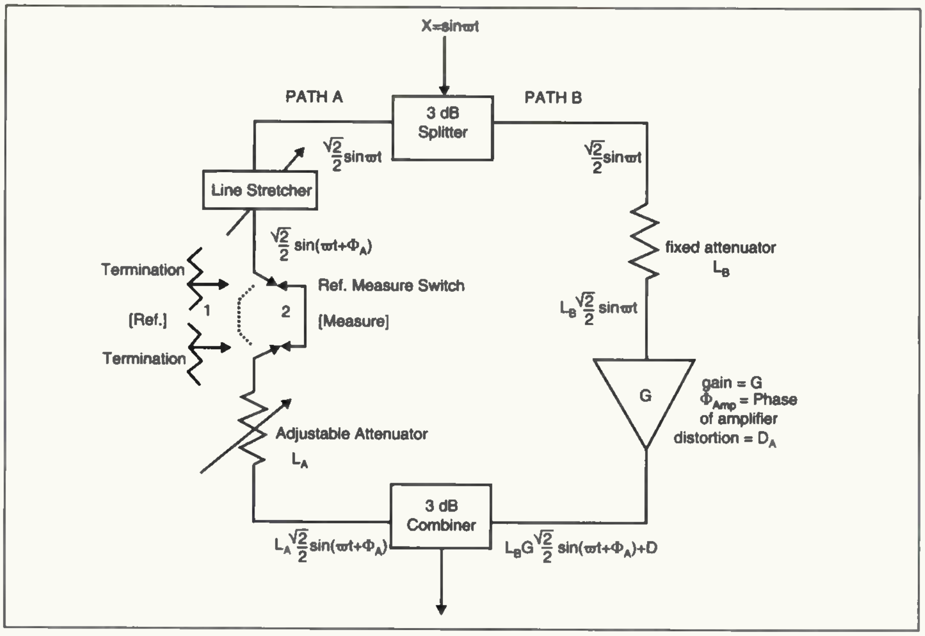

Figure 1. Block diagram of bridge setup to cancel source noise.

Implementation

Referring to Figure 1, the input signal (X) is divided by using a 3 dB power splitter. One half of the signal is fed through path A that contains a continuously variable attenuator (L_B) and a line stretcher (Φ_A). The line stretcher is used as a phase shifter. The second half of the signal is fed through path B, which contains a fixed attenuator (L_B) and the amplifier under test (G). The value of the attenuation should be such that the gain in path B is nominally –5 dB, that is 5 dB loss. Although it may not be immediately apparent, the signal-to-noise ratio of the measurement is better when the attenuation is on the input side of the amplifier rather than the output. When measuring high power amplifiers it may be necessary to add attenuation at the amplifier output. The variable attenuator and phase shifter are now adjusted until the input signals from the source are nulled at the output.

In order to null the signals, a 180° phase shift is required in one path. When testing a non-inverting amplifier, a 180° combiner must be used. Typical nulls of –40 dB can be achieved over 20 percent bandwidth in spite of the fact that the splitters and combiners are not specified to this performance. This is because the variable phase shifter and attenuator compensate for the inherent inaccuracies of the two paths. Actually, a –25 dB null can be attained over decade bandwidths with comparative ease. This suppression not only suppresses the signal, but any other extraneous signals and distortions along with any noise produced by the source.

Although nulls exist whenever the signals are out of phase at the output, the broadest null occurs when the time delays in both paths are equal to each other. It may be necessary to add lengths of cable to one of the paths

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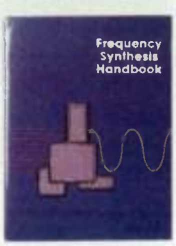

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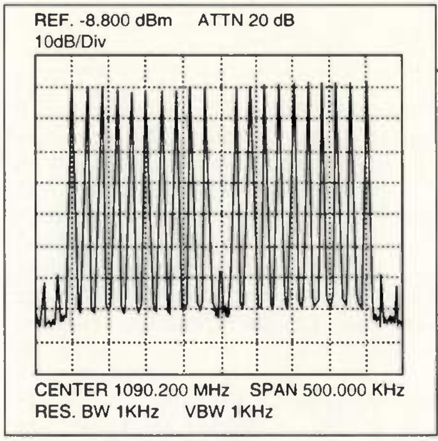

Figure 2. 20 input signals with their inherent distortion.

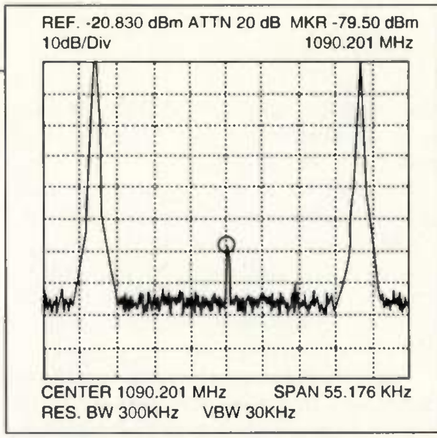

Figure 3. The two innermost signals, with the distortion signal between them.

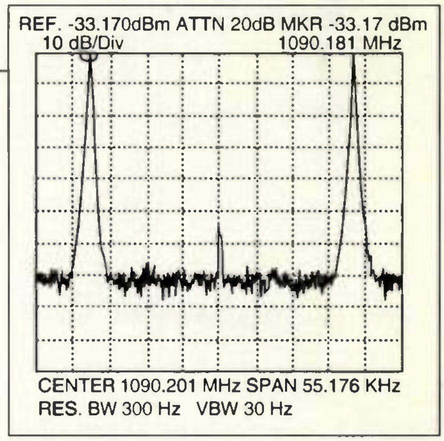

Figure 4. Spectrum of signal through path B.

because the phase shifter range is limited. It may also be difficult to find the null that produces equal time delay in both paths. To simplify the process of adjusting the circuit for equal time delay and hence produce a broadband null, a sweep signal generator or amplified noise source may be substituted for the signal source. This allows the user to adjust the L_A to equal $L_B G$ and therefore, the resulting equation from Figure 1 will be the difference of the two paths:

output = path A – path B

$$\text{output } D_A = L_A\left[\frac{\sqrt{2}}{2}\sin(\omega t + \Phi)\right] - \left[L_B G\left[\frac{\sqrt{2}}{2}\sin\left(\omega t + \Phi_{Amp}\right)\right] + D_A\right] \quad (1)$$

where:
X = input signal = $\sin(\omega t)$
L_A = loss of variable attenuator
Φ_A = variable phase shift
L_B = fixed attenuator
G = amplifier gain
Φ_{Amp} = amplifier phase shift; and
D_A = amplifier distortion

Example

Figure 2 shows the 20 input signals from the signal generator along with its residual distortion. Figure 3 is expanded around the center to show only the two innermost signals and the distortion present in the input signal.

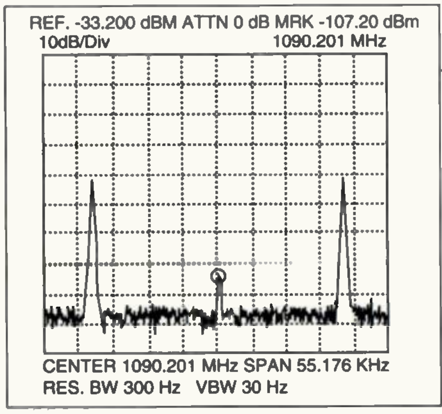

Figure 5. Spectrum of signal after passage through paths A and B and combination.

This distortion is about 60 dB below the signal level.

Figure 4 is a representation of the signals when the transfer switch position is at 1 (referring to Figure 1). The reference level reading of the signal is taken at this point, and is –33.2 dBm. The insertion loss from the amplifier output to the input of the spectrum analyzer is 3.2 dB. This makes the amplifier output level –30 dBm.

Figure 5 is a representation of the signals after a 40 dB suppression with the transfer switch at position 2 (referring to Figure 1). The input signals and their distortions, along with any noise present in the input, have been nulled at this point. This has allowed a 20 dB decrease in the RF attenuation of the spectrum analyzer, therefore improving the noise floor. The amplifier distortion is now visible and is 74 dB below the reference signal level.

The distortion products of the signal generator and the spectrum analyzer are no longer a factor in our measurements because the 40 dB suppression of the input signals and the input distortion level of –60 dB is the equivalent of –100 dB distortion levels at the input. At this particular signal level, the spectrum analyzer is not contributing any significant distortion. This is verified by adding attenuation to the input of the spectrum analyzer and observing that no change occurs in the carrier to distortion ratio.

Conclusion

By using this technique, we have alleviated many of the problems encountered while using other distortion measurement methods, such as signal source distortion and noise from the spectrum analyzer. Although the example chosen was for a low level amplifier, the technique is also useful for high level amplifiers and other active devices. Our experience indicates that the bridge method of distortion measurement eliminates the need for bandpass filters. *RF*

References

1. *RF/IF Designer's Handbook*, Mini-Circuits, 1992.
2. *Multiple Frequency Comb Generator Handbook*, Matrix Test Equipment Inc.
3. *HP 8560 E Spectrum Analyzer Operating and Programming Manual*, Hewlett Packard, 1992.
4. C. Rice, "Techniques to Achieve Linear Amplification at HF", *RF Design*, December 1993, pp. 46-49.

About the Author

Nina Krikorian is employed at Matrix Test Equipment Inc. and presently attends Polytechnic University in New York. She can be reached at Matrix Test Equipment Inc., 90-11 Colin Dr., Holbrook, NY 11741, tel: (516) 472–0153, fax: (516) 472–0624.

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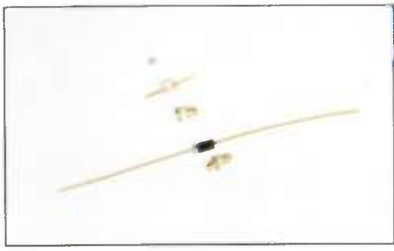

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NC 401	100 MHz – 18 GHz
NC 406	18 GHz – 110 GHz

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NC 513/15	0.2 MHz – 2 GHz	51 dB

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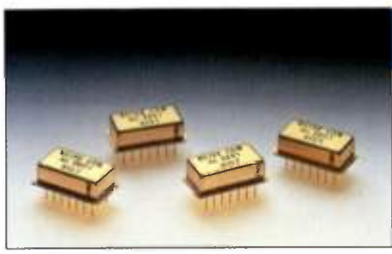
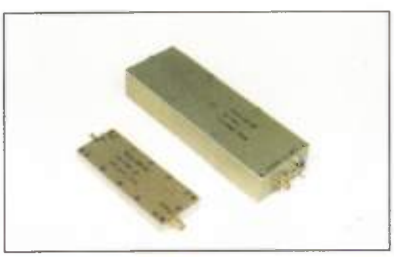


The NC 2000 series amplified noise modules are an excellent choice when a high level noise output is desired and the noise source is to be mounted on a circuit board. 24 pin packages are standard, 14 pins are available.

TYPICAL STANDARD MODELS		
MODEL	FREQUENCY RANGE	OUTPUT
NC 2101	100 Hz – 20 kHz	0.15 Vrms
NC 2105	500 Hz – 10 MHz	0.15 Vrms
NC 2201	1 MHz – 100 MHz	+5 dBm
NC 2601	1 MHz – 2 GHz	-5 dBm



The NC 1000 series amplified noise modules produce white Gaussian noise from -14 dBm to +13 dBm at frequencies up to 6 GHz. They are designed for coaxial test systems, and are available with several bias voltages and connector options.

TYPICAL STANDARD MODELS		
MODEL	FREQUENCY RANGE	OUTPUT
NC 1101A	10 Hz – 20 kHz	+13 dBm
NC 1107A	100 Hz– 100 MHz	+13 dBm
NC 1112B	20 MHz – 2 GHz	0 dBm
NC 1126A	2 GHz – 6 GHz	-14 dBm

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Noise Com's NC 346 series is designed for precision noise figure measurement applications. These products are available with coaxial or waveguide outputs. For OEM applications, the NC 3200 series provides high performance in a small ruggedized package.

TYPICAL STANDARD MODELS		
MODEL	FREQUENCY RANGE	OUTPUT ENR
NC 346A	0.01 GHz – 18 GHz	6 dB
NC 346B	0.01 GHz – 18 GHz	15 dB
NC 346D	0.01 GHz – 18 GHz	25 dB
NC 346Ka	0.1 GHz – 40 GHz	15 dB

Broadband Calibrated Millimeter-wave

The NC 5000 series noise sources feature outstanding stability and convenience in waveguide bands up to 110 GHz.

TYPICAL STANDARD MODELS		
MODEL	FREQUENCY RANGE	WAVEGUIDE
NC 5142	18 GHz – 26.5 GHz	WR-42
NC 5128	26 GHz – 40 GHz	WR-28
NC 5115	50 GHz – 75 GHz	WR-15
NC 5110	75 GHz – 110 GHz	WR-10

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TYPICAL STANDARD MODELS

MODEL	FREQUENCY RANGE	OUTPUT ENR
NC 3404	2 – 4 GHz	30–36
NC 3405	4 – 8 GHz	30–35
NC 3406	8 – 12 GHz	28–33
NC 3407	12 – 18 GHz	26–32

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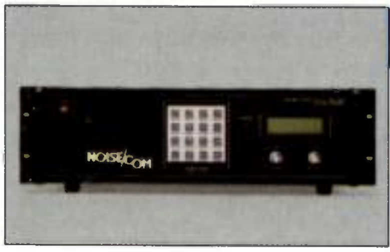
TYPICAL STANDARD MODELS

MODEL	FREQUENCY RANGE	APPLICATION
UFX-BER-70	50 MHz – 90 MHz	General
UFX-BER-IBS/IDR	50 – 90; 100 – 180 MHz	IBS/IDR
UFX-BER-836	824 MHz – 849 MHz	CDMA
UFX-BER-1850	1800 MHz – 1900 MHz	DCS-1800



Amplifier Test Station

The UFX-NPR series instruments perform automatic distortion measurements in mobile telephone (CDMA and FDM) base stations, satellite communications systems, CATV, and other equipment operating in multi-signal environments. Some models are available with tunable measurement frequency or with multiple measurement frequencies.

TYPICAL STANDARD MODELS

MODEL	FREQUENCY RANGE
UFX-NPR-70	50 MHz – 90 MHz
UFX-NPR-CATV	50 MHz – 1.0 GHz
UFX-NPR-1700	1.6 GHz – 1.9 GHz
UFX-NPR-2400	2.2 GHz – 2.6 GHz
UFX-NPR-11900	10.95 GHz – 12.8 GHz

Broadband Noise Generators

The NC 6000 and NC 8000 series noise generating instruments are designed for applications on the test bench or incorporated with other equipment to provide a wide variety of functions. Each instrument contains a precision noise source, amplification, and step attenuators to provide repeatable symmetrical white Gaussian noise with variable output power.

TYPICAL STANDARD MODELS

MODEL	FREQUENCY RANGE	OUTPUT POWER
NC 6107	100 Hz – 100 MHz	+13 dBm
NC 6110	100 Hz – 1500 MHz	+10 dBm
NC 6124	2 GHz – 4 GHz	-10 dBm
NC 8107	250 kHz – 100 MHz	+30 dBm

The new UFX-7000 series noise generating instruments are extremely easy to use, combining dedicated keys for control of operations and programming, with a large 4 x 20 character LCD display. Control of output power, filter settings, and attenuator step size for both the noise and the signal (for units with internal combiners) is performed from the front panel or remotely using the IEEE-488 interface.

TYPICAL STANDARD MODELS

MODEL	FREQUENCY RANGE	OUTPUT POWER
UFX-7107	10 Hz – 100 MHz	+13 dBm
UFX-7108	100 Hz – 500 MHz	+10 dBm
UFX-7110	100 Hz – 1500 MHz	+10 dBm
UFX-7218	2 GHz – 18 GHz	-20 dBm
UFX-7909	1 MHz – 300 MHz	+30 dBm

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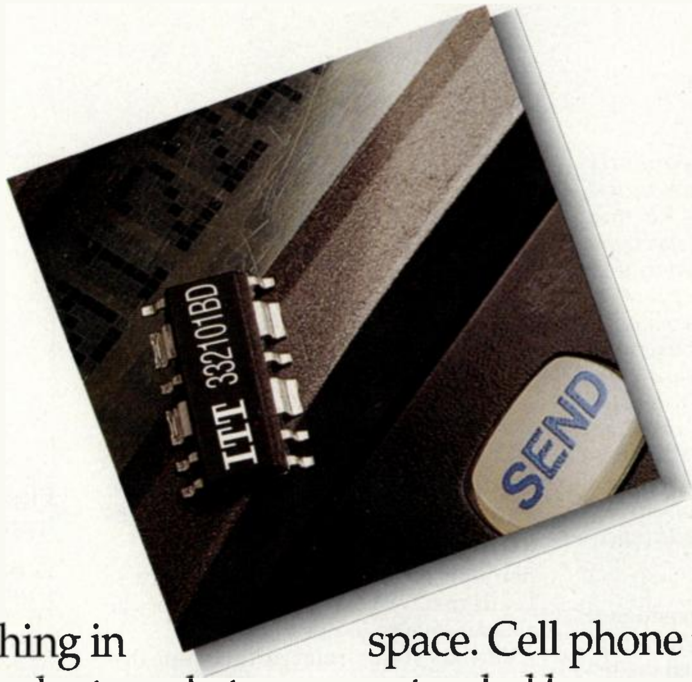
Small talk.



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INFO/CARD 30

Distortion Measurements Using the Spectrum Analyzer

By Morris Engelson
Tektronix, Inc.

The instrument most frequently used to measure, display, adjust or test for distortion, especially at RF and microwave frequencies, is the spectrum analyzer. Much published material is available on the use of the spectrum analyzer in distortion measurements. The reader will find some of this material cited in the references. Usually, an article will deal in some depth with one aspect of distortion, such as intermodulation or circuit gain compression. The intent here is to provide a summary overview of many varieties of distortion items associated with spectrum analysis.

The sinewave is usually considered an ideal signal, and any deviation from a perfect sinewave is called distortion. Usually distortion is not wanted. Sometimes, though, the intent is to create distortion in the sense of having a non-sinusoidal signal, such as a pulse or square wave. While the idea of distortion is usually associated with a signal, it need not always be so. Frequently, distortion is used to determine circuit behavior rather than signal content. Any circuit can cause distortion, and that includes the circuits in the measuring instrument. Therefore, accurate distortion measurement should include consideration, or tests, to establish the validity of the result.

Baseband or RF?

Distortion of a modulated signal, or waveform, can occur at audio or baseband prior to modulation, or at RF after modulation. Usually, the measurement is made at the lower frequency either before modulation or after demodulation. But this does not fully show what happens during the modulation process, for example, what level of distortion the modulator introduces. Today, there is no technical reason for not making the measurement at whatever frequency is of interest, with the availability of high frequency resolution spectrum analyzers such as the Tektronix 2794 (which was used for the measurements in this paper).

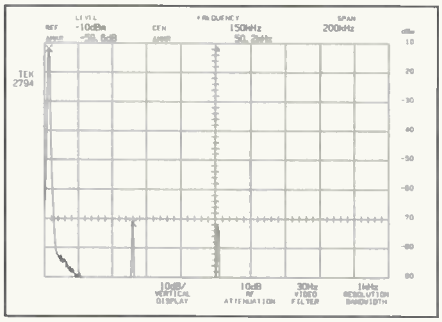


Figure 1. Modulating signal, fundamental and harmonics spectrum.

Various items related to signal distortion measurements with the spectrum analyzer, in general, and differences between pre-modulation and post-modulation, in particular, are illustrated below.

Figure 1 shows the spectrum of a 50 kHz sinewave and its harmonics. The 2794 vertical setting is 10 dB/division for an 80 dB screen. Only two harmonics, the second and third, at about 60 dB down are observed. The 50 kHz signal exhibits about 0.1 percent distortion.

Figure 2 shows the spectrum of a 10 GHz carrier amplitude modulated with the 50 kHz signal discussed in Figure 1. The 50 kHz sideband is 14.8 dB below the 10 GHz carrier amplitude, representative of 36.4 percent AM. We also have spectral lines at the second and third harmonics of the 50 kHz modulating signal. These however, are much greater in amplitude than shown in Figure 1. Here the 100 kHz term is only 20 dB below the 50 kHz line rather than 60 dB as in Figure 1. Clearly, the modulating process has changed the relative harmonic levels.

Not only 50 kHz sidebands, but close-in spectral terms are easily observed and measured at the 10 GHz carrier frequency. Figure 3 shows a 234 Hz tone at 12 dB down, representing 50 percent AM. A closer look using a 10 Hz resolution setting in Figure 4 shows additional modulation terms of 80 Hz.

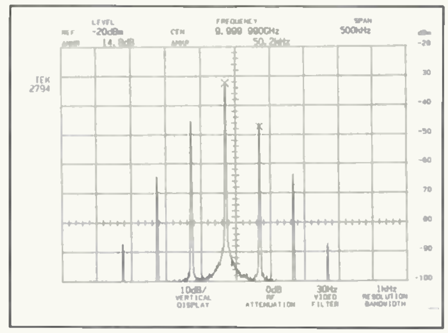


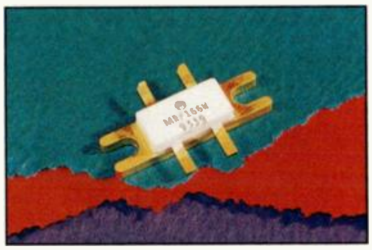
Figure 2. Modulated signal spectrum at 10 GHz.

Harmonic Distortion

Signal harmonics observed on a spectrum analyzer can come from three sources: the original signal, an outside circuit through which the signal passes (such as an amplifier), and the measuring spectrum analyzer. Harmonic distortion caused by a circuit, such as an outside amplifier or the spectrum analyzer input mixer, will be signal-level dependent, while the harmonic content of the original signal will not be amplitude-level dependent. A check for change of harmonic level as a function of signal level is the principal method for distinguishing between circuit-caused versus original-signal harmonic content.

Figure 5 is a dual trace display of the spectrum of a 200 kHz signal (actually 196.6 kHz as measured with the markers). The full screen spectrum analyzer reference level is –20 dBm for both traces. With 10 dB of input RF attenuation, the mixer input level is –20–10 = –30 dBm, and the second harmonic is measured at 68 dB down. With zero dB of input attenuation and a mixer input level of –20 dBm, the upper trace shows a second harmonic almost 10 dB larger. The increase is due to spectrum

MRF166W - Class A, 40 W, 500 MHz, Broadband Power MOSFET



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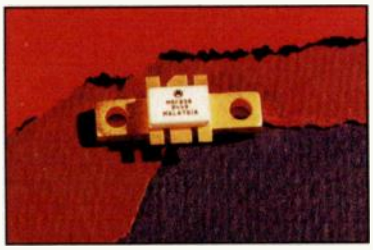

Motorola's 3.6 watt (CW) RF power transistor is designed for 24-volt UHF large-signal common emitter, class A linear amplifier applications in industrial and commercial equipment operating in the 800-960 MHz range. Features include an 11 dB minimum gain, an output capacitance of 6.5 pF (typ) at 24 Vdc, a minimum ITO of +44.5 dBm and typical noise figure of 6 dB. The MRF858 will withstand RF input overdrive of 0.85 W CW. It is in an SOE flange package (case 319).
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MRF860 - Class A, 800-960 MHz, 13.7 W (CW) RF Power Transistor

Motorola's 13.7 watt (CW) RF power transistor is designed for 24-volt UHF large-signal common emitter, class A linear amplifier applications in industrial and commercial equipment operating in the 800-960 MHz range. Features include an 11 dB minimum gain, an output capacitance of 21 pF (typ) at 24 Vdc, a minimum ITO of +51.5 dBm and typical noise figure of 6.5 dB. The MRF860 will withstand RF input overdrive of 3.25 W CW. It is in a push-pull flange package (case 395B).
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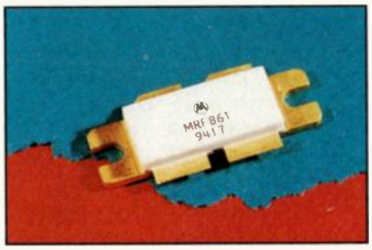
MRF861 - Class A, 800-960 MHz, 27 W (CW) RF Power Transistor



Motorola's 27 watt (CW) RF power transistor is designed for 24-volt UHF large-signal common emitter, class A linear amplifier applications in industrial and commercial equipment operating in the 800-960 MHz range. Features include a 9.5 dB minimum gain, an output capacitance of 45 pF at 24 Vdc, a minimum ITO of +53.5 dBm and typical noise figure of 6.5 dB. The MRF861 will withstand RF input overdrive of 8 W CW. It is in a push-pull flange package (case 375A).
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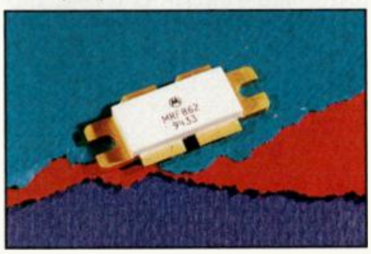
MRF862 - Class A, 800-960 MHz, 36 W (CW) RF Power Transistor



Motorola's 36 watt (CW) RF power transistor is designed for 24-volt UHF large-signal common emitter, class A linear amplifier applications in industrial and commercial equipment operating in the 800-960 MHz range. Features include a 9 dB minimum gain, an output capacitance of 75 pF at 24 Vdc, a minimum ITO of +55 dBm and typical noise figure of 6.5 dB. The MRF862 will withstand RF input overdrive of 13.6 W CW. It is in a push-pull flange package (case 375A).
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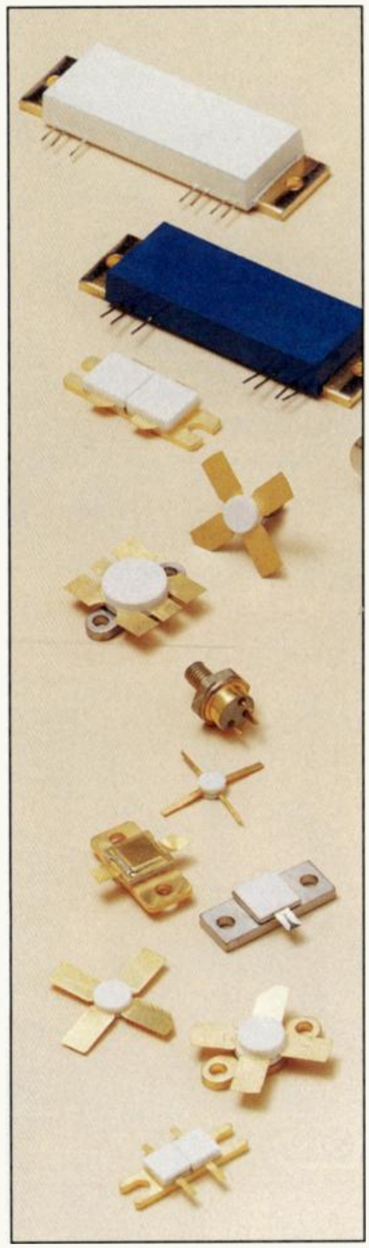

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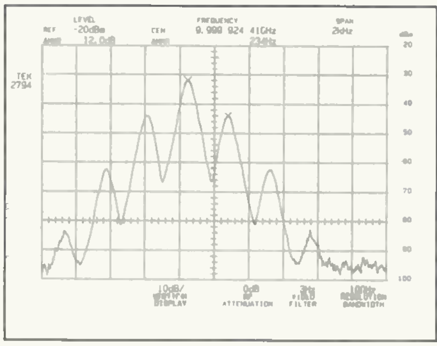

Figure 3. Higher resolution look at 10 GHz signal.

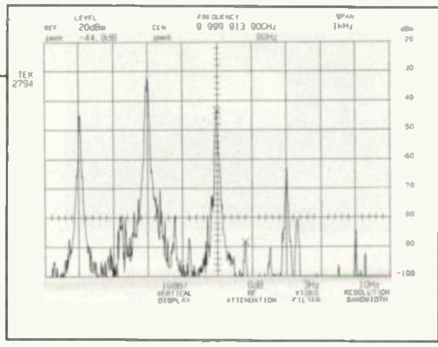

Figure 4. Close-up observation of 10 GHz signal spectrum using 10 Hz resolution bandwidth.

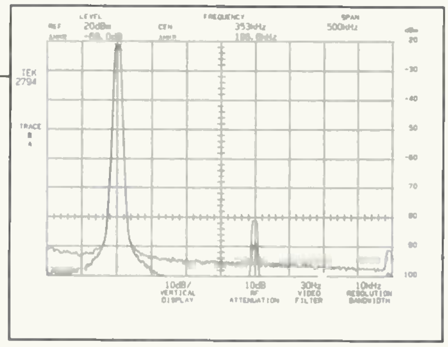

Figure 5. Testing for harmonic distortion caused by the spectrum analyzer.

analyzer generated distortion and the measurement is not correct.

It is not safe to assume that the observed distortion is part of the signal without additional information. Either the user needs to check spectrum analyzer specifications for appropriate safe settings, or the signal input level needs to be changed to check for impact on distortion. Figure 6 shows the spectrum of a 1 GHz carrier amplitude modulated with a 58 kHz tone. The intent is to determine the total harmonic distortion, THD, of the signal. THD is the rms sum, that is, the square root of the sum of the squares, of individual distortion terms.

First we determine the distortion content on a term-by-term basis, then the THD is easily computed. We have the following for Figure 6.

Harmonic	2nd	3rd	4th	5th
dBc	21.6	43.6	59.6	62.0
% Ratio	8.32	0.66	0.10	0.08

The square root of the sum of the squares yields a THD of 8.35 percent.

A distortion meter usually shows THD while a spectrum analyzer usually determines distortion on a term-by-term basis and THD is calculated. However, many spectrum analyzers, including the Tektronix 2794, have macro software available where the dis-

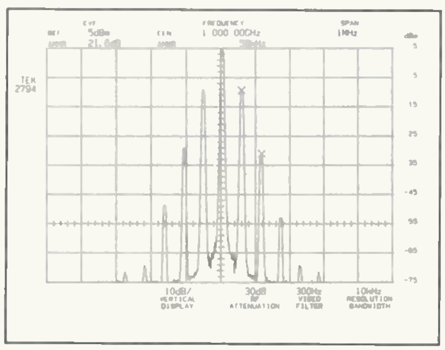

Figure 6. Harmonic distortion measurement for THD computation.

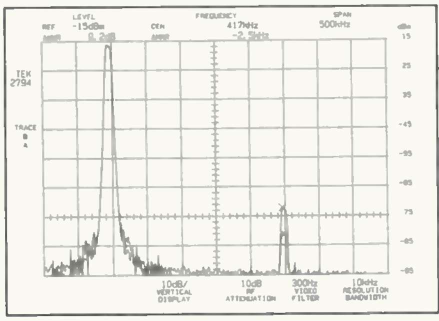

Figure 7. Low harmonic level, test for valid measurement.

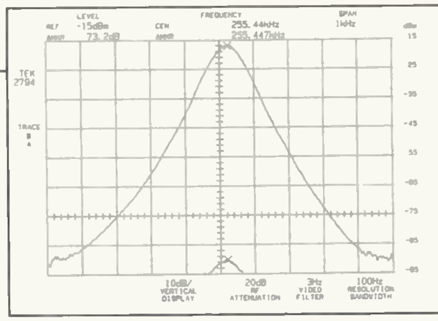

Figure 8. Measuring relative amplitude of low level 2nd harmonic.

tortion measurement and THD calculation is obtained automatically by executing a single measurement command.

Figure 7 shows an attempt to measure the second harmonic of a very clean 250 kHz signal. A change of 10 dB in the spectrum analyzer input attenuator shows an almost 10 dB (9.2 dB as seen in the upper left corner of the readout) change in harmonic level. This is too close for comfort. Some of the displayed harmonic content is likely due to the spectrum analyzer. Therefore, we need to reduce the spectrum analyzer input level by an additional 10 dB to make certain. The result of inserting 20 dB of RF attenuation is shown in Figure 8. The second harmonic is very small now, and we need to improve the signal-to-noise ratio to measure it. This is accomplished by reducing the 10 kHz resolution bandwidth used in Figure 7 to 100 Hz used in Figure 8. Such a narrow resolution setting makes it difficult to span enough spectrum to observe the fundamental and harmonic in one sweep. Hence, Figure 8 shows two traces, obtained in two sweeps, one for the fundamental, and one for the harmonic. The 2794 has "intelligent" markers that will make measurements between different traces. A delta marker measurement shows a second harmonic level at 73.2 dB below the fundamental, or 0.02 percent distortion. A modern spectrum analyzer can measure harmonic distortion down to about 0.01 percent.

Intermodulation

Intermodulation, IM, is a form of distortion where multiple sinewave signals interact to create additional signal components. The simplest and most usual intermodulation tests involve two signals. Thus, if the two signals are at frequencies f_1 and f_2, we can get new signals at frequencies: f_1+f_2, f_2-f_1, $2f_1-f_2$, $2f_2-f_1$, etc. The usual test is for the so called, third order terms consisting of $2f_1-f_2$ and $2f_2-f_1$.

Intermodulation testing is a complicated and extensive topic. This paper does not deal with it in depth. Here are some examples of the factors that need to be considered. These are discussed in more detail in the references.

Test signal spacing – Some circuits are sensitive to test signal frequency spacing. The closer the two signal frequencies at f_1 and f_2, the larger the IM products at $2f_1-f_2$ and $2f_2-f_1$. Be sure to vary the signal frequency spacing if you suspect that you have such a circuit.

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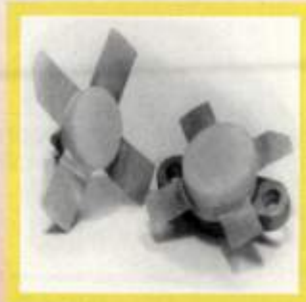

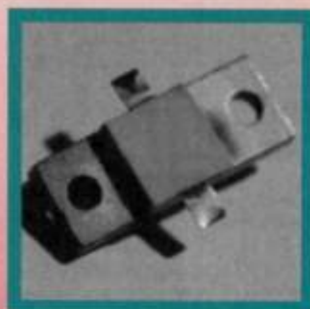

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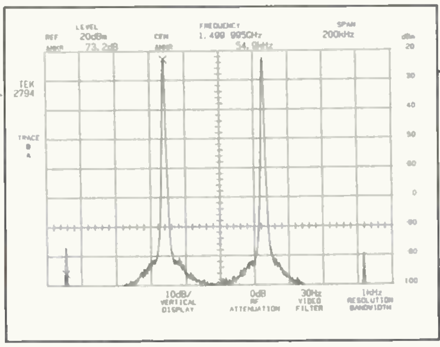

Figure 9. Third order intermodulation spectrum, testing for distortion from spectrum analyzer.

chased. Most people, however, combine the outputs from two independent signal generators to create a two-tone signal. It is very important that the two sources be well isolated from each other. Otherwise the interaction between the test sources will introduce IM terms which can be mistaken for legitimate test results. Isolators, directional couplers or filters and attenuators are primary means for maintaining independence between the test sources. Crosstalk by radiation from cables is also a potential problem when looking for very low levels of IM.

Most circuits are "well behaved" over a large range of signal frequency and amplitude levels. This means that equations involving the intercept point, as discussed later, are applicable. Using such equations makes IM testing much easier, as it means that testing can be performed at one set of conditions and results can be determined by computation for another set of conditions. However, not all circuits are well behaved, and not all well behaved circuits are well behaved all times. Make sure that you are not going beyond the range of your circuit when getting results by computation rather than direct measurement.

Figure 9 shows a typical IM test spectrum. The two signals have a 55 kHz separation (marker shows 54.9 kHz) centered at 1.5 GHz. The display shows two traces, one with 0 dB of input attenuation and one with 10 dB of attenuation. As with harmonic distortion, IM distortion products are circuit input level sensitive. A change in input level to the spectrum analyzer will show whether the IM products are real or being generated by the test instrument. We note in Figure 9 that a change of RF attenuation does change the IM level. Hence, the result at 0 dB attenuation setting is not correct. Measurement of the smaller IM (at 10 dB attenuation) shows 73.2 dBc.

If the test specification were 70 dB, for example, then a 73.2 dB result

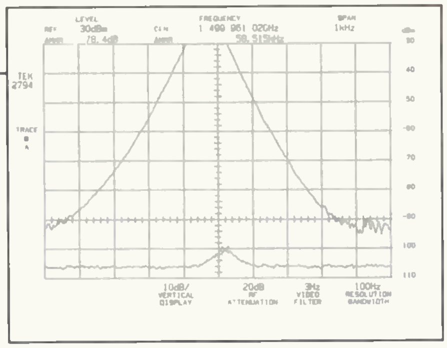

Figure 10. Third order IMD test with main signal off screen.

would be sufficient to show compliance. It would not be important to determine whether the real number is better than that. If, however, we need to know the true result, then it becomes necessary to see what happens at a 20 dB attenuator setting. Possibly, some of the IM at 73.2 dBc is still due to the spectrum analyzer. Figure 10 shows such a measurement. As was the case with Figure 8, the result was obtained in two traces, once for a input test signal and once for an IM component. Here the spectrum analyzer gain setting needed to properly display the IM component is such that the main signal goes off screen. Nevertheless, the "intelligent" marker of the 2794 is able to make the measurement at 78.4 dBc. Indeed, some of the result shown in Figure 9 was due to IM from the spectrum analyzer.

Figure 10 may be acceptable for getting a result, but it is deficient in showing all the information. Most people prefer that all signal parts be on screen. Such a measurement is shown in Figure 11. Here the gain was reduced for the upper trace so as to bring the signal top on screen. The "intelligent" markers keep track of this change and the measurement using the markers again shows 78.4 dBc. However, one must be careful to do this only while making a measurement with the markers. Had the result been determined manually from the traces, with-

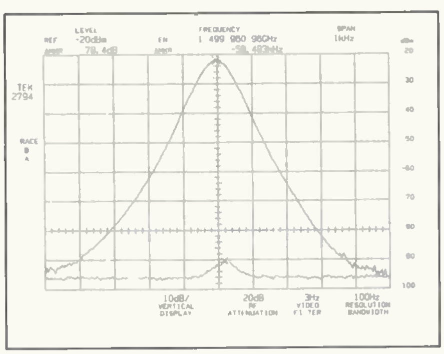

Figure 11. Third order IMD test with main signal fully displayed on screen.

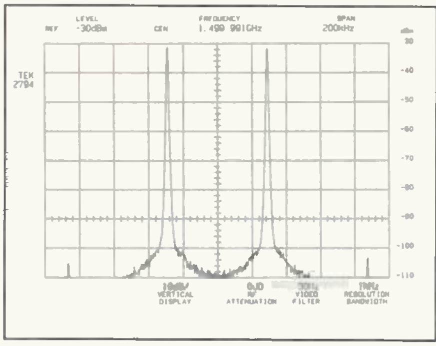

Figure 12. Two tone IMD spectrum.

out the markers, the result would show 70 dB with a seven-division display difference at 10dB/div.

The intermodulation distortion level generated by a well behaved circuit depends on the test signal input level (S in dBm) and the circuit intercept point (I in dBm). The reader will find extensive discussion on the intercept point concept in the references. For now it is sufficient to note that the intercept point is a measure of circuit linearity. The higher the intercept, the more linear the circuit and the lower the distortion products. The relationship is given by I = (dBc/n–1)+S. The quantity dBc is the IM product level compared to the carrier and n is the IM order number. The order number is the sum of the test signal harmonics involved and is equal to three for the $2f_1$–f_2 and $2f_2$–f_1 products previously discussed.

Using the intercept point concept it is possible to make measurements at whatever signal level is convenient and the desired result is computed. Thus, if the IM level is measured at dBc = 70 dB, when each of the test signals is at S = –30 dBm, then: I = (70/3–1)–30 = +5 dBm. From this we can determine that dBc is 110 dB when S = –50 dBm, even though our spectrum analyzer may not be able to make this measurement.

Multi-Tone IM

Multi-tone intermodulation is much more complicated than the two-tone variety. This is because the number of possible distortion terms increases geometrically as the number of tones increases. Some systems, such as baseband telephony, use three tones, while CATV applications can involve hundreds of tones. Choice of tone frequencies will have an impact on results depending upon whether multiple IM terms will fall on top of each other or not. For example, three tones at 200 kHz, 600 kHz and 1000 kHz will have IM clusters at: 2x200=400, 1000–600 = 400, 600–200 = 400; 200, 600–2x200 = 200, 2x600–1000 = 200; etc.

A two tone arrangement has two IM

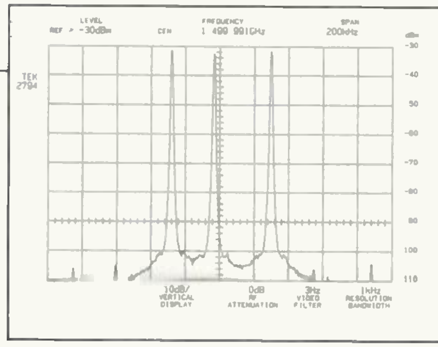

Figure 13. Three tone IMD spectrum.

products in the frequency vicinity of the test tones at $2f_1$–f_2 and $2f_2$–f_1. A three tone arrangement with f_1, f_2 and f_3 near each other can produce nine IM products near the test tones at: $2f_1$–f_2, $2f_2$–f_1, $2f_2$–f_3, $2f_3$–f_2, $2f_1$–f_3, $2f_3$–f_1, f_2+f_3–f_1, f_1+f_2–f_3, f_1+f_3–f_2. In other words, multi-tone IM can get messy. Here is an example of a three-tone IM test.

Figure 12 shows a familiar two-tone IM test. Signal separation is 59.3 kHz. Two intermodulation components are observed, one on each side of the spectrum at a 59.3 kHz offset. Figure 13 shows the same spectrum as before but with a third test signal inserted between the first two. The left side test signals are spaced at 25.7 kHz, the right side at 33.6 kHz and the outer signals are spaced at 59.3 kHz as in Figure 12. We now have several new IM components in addition to the two observed in Figure 12. Some are very obvious, and easy to identify as to cause. Thus, the left most IM component is the same as in Figure 12 and spaced at 59.3 kHz from the nearest test tone. The IM component next in line, at two divisions from the left, is spaced 33.6 kHz below the left test signal. It comes from f_1+f_2–f_3. Some of the IM products are very difficult to recognize unless you know where to look. The little bump at the right side base of the middle test signal is the f_2+f_3–f_1 component, spaced at 33.6–25.7=7.9 kHz above the middle test tone.

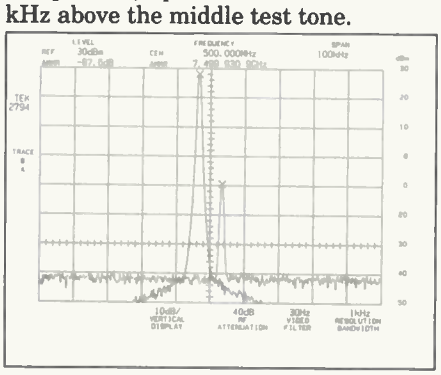

Figure 15. Harmonic level measurement using a preselector.

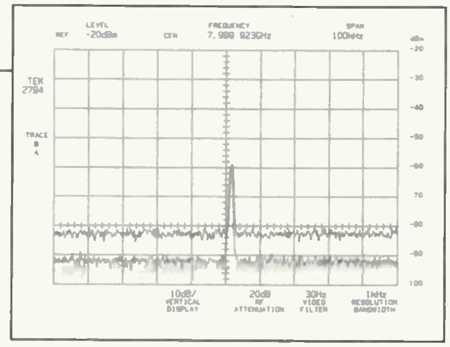

Figure 14. Checking for validity of harmonic test result using a preselected spectrum analyzer.

The Preselector

A preselector is a signal tracking filter at the input of the spectrum analyzer used primarily at microwave frequencies. Signals must be simultaneously present in a circuit to cause distortion. Therefore, signals whose frequency spacing is greater than the preselector bandwidth will not cause distortion in the spectrum analyzer.

Figure 14 shows the sixteenth harmonic of a 500 MHz signal. A change in input attenuation does not change the harmonic amplitude level. This is because the spectrum analyzer is protected from generating harmonic components by an input preselector. Figure 15 shows the same harmonic at 8 GHz as shown in Figure 13, along with the 500 MHz fundamental on a second trace. Gain and attenuation settings have been changed between traces to permit both signals to show on screen. But the spectrum analyzer markers take these changes into account as discussed for Figure 11. The harmonic amplitude level measures at 87.6 dB below fundamental.

Preselected measurements can easily exceed 100 dB, even though the normal display range on screen is only 80 dB. This is illustrated in Figure 16. Here we determine the relative amplitude of the eighteenth harmonic at 9 GHz. The result shows 119.2 dB.

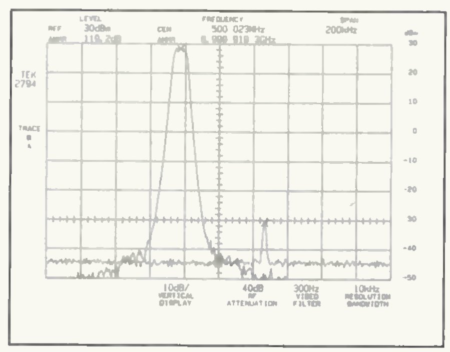

Figure 16. Measuring a very low level harmonic at 119.2 dBc using a preselector.

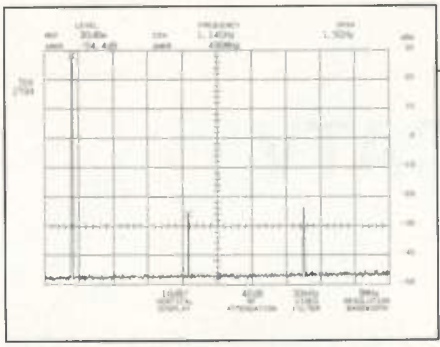

Figure 17. Spectrum of fundamental at 500 MHz and two harmonics.

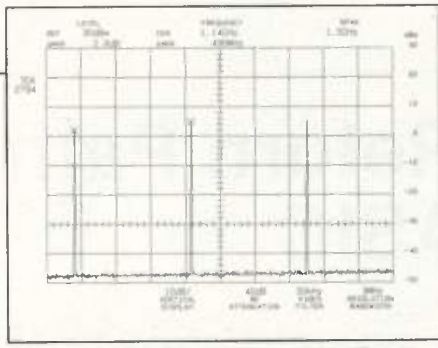

Figure 18. Spectrum of 17 after snap diode multiplier.

Deliberate Distortion

Not all distortion need be avoided. Some signals are intended to be non-sinusoidal and hence, "distorted".

Figure 17 shows the lower frequency spectrum of the 500 MHz signal whose sixteenth and eighteenth harmonic was previously measured. The second harmonic at 1 GHz is 54.4 dB below the 500 MHz fundamental. To provide a stable higher frequency test signal it is desired to enhance the harmonic level at higher frequencies. This is accomplished by means of a snap diode multiplier. Connecting the multiplier results in the spectrum shown in Figure 18. This results in a substantially flat spectrum. The fundamental level has dropped and the second harmonic level increased to 2.8 dB above fundamental. The spectrum analyzer is an excellent tool for testing the performance of the multiplier. *RF*

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About the Author

Morris Engelson is the former chief engineer at Tektronix and is currently associated with Tektronix on a consulting basis. He is a Fellow of the IEEE for contributions to spectrum analysis and is the author of several books on the topic. He is a NARTE-certified EMC engineer. He is currently director at the Portland, Oregon-based consulting firm of JMS. He can be reached at Tektronix at (503) 627-4823, and at JMS at (503) 292-7035.

A Low Cost Return Path CATV Transmitter ASIC

Bill Xenakis
Stanford Telecommunications, Inc.

Cable television's coaxial cable distribution network has traditionally been used for one-way video broadcast only. This high-bandwidth medium, however, provides the clear possibility of 2-way interactivity, enabling such services as video-on-demand, data communications, and telephony. Transmission on the cable return path can be accomplished using a new chip developed by Stanford Telecom, the STEL-1103 Digital PSK Modulator. This chip's novel architecture (patent pending) is ideally suited to the unique challenges associated with upstream cable transmission, and has been selected by several major companies for inclusion in their high volume products. It is specifically designed to perform burst data transmission of bandwidth-efficient signals while meeting cable operator requirements for out-of-band interference. Requiring only the addition of a digital to analog converter and a lowpass filter, the chip produces BPSK or QPSK modulated, spectrally-shaped RF signals at center frequencies of up to 40 MHz. Based on its low cost and unique performance characteristics, the STEL-1103 represents a major breakthrough for the interactive communications market.

A complete transmission system based on the STEL-1103 is shown in Figure 1. Note that 12 bits of output data are provided; however, fewer bits may be used, depending on the performance requirements or cost goals of the particular systems. It has been found that 10 bit DACs are economical and provide very high performance.

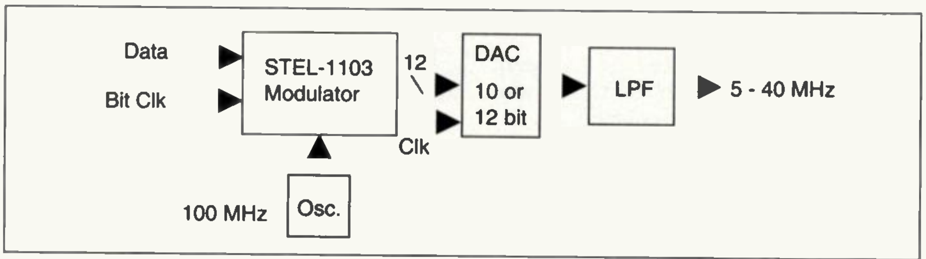

Figure 1. A complete transmitter based on the STEL-1103.

The key features of the STEL-1103 PSK modulator chip are:

- Programmable FIR filters providing:
  - Spectral shaping to minimize occupied bandwidth
  - Excess bandwidths as low as 20 percent
  - Pulse shaping to minimize intersymbol interference
  - Stop band rejection greater than 50 dB
- Precise tuning (24 bit NCO provides 6 Hz steps with a 100 MHz master clock)
- Fast tuning (less than 1 microsecond, compared to many milliseconds for analog PLLs)
- Very low phase noise (equivalent to that of the crystal oscillator time base being used)
- Multi-octave tuning (DC to 40 MHz)
- Programmable bit rates up to 10 Mbps in QPSK, 5 Mbps in BPSK mode
- Burst operation control interface
- Differential encoding option, permitting simple I/Q ambiguity resolution
- Variable-rate interpolation filters eliminating unwanted alias energy
- All-digital modulator design resulting in:
  - Zero phase and amplitude error (the constellation is perfectly centered)
  - Performance constant over entire output frequency range
  - Immunity to process-related performance variations (very high repeatability)
  - No need for factory or field adjustments
  - Possible integration with other digital system functions to save cost
  - Lowest possible overall system parts-count
- Low cost 80 pin PQFP package

CATV Advantages

A number of the characteristics listed above result in the STEL-1103 being suited for upstream transmission of voice and data over CATV cable (or for any application having a direct RF output of under 40 MHz). Note the following advantages:

1) Since the STEL-1103 covers the entire upstream path frequency range, no analog frequency conversion is required. This eliminates: a) nonlinearities associated with mixers, b) phase noise associated with analog PLL-based synthesizers, and c) unnecessary complexity.

2) The output signal can be spectrally shaped to occupy minimum bandwidth, thereby allowing many signals to be closely spaced.

3) The stop band rejection is greater than 50 dB, thereby preventing interference with other upstream digital signals. The required lowpass filter will eliminate spurious energy occurring above 40 MHz.

4) The built-in burst control interface allows short bursts to be transmitted without splatter, i.e. the start up and shut down of the bursts are precision controlled to prevent stray high frequency energy from being sent onto the cable. This can eliminate the need for RF switches.

Applications Details

The STEL-1103 can transmit in either continuous or burst mode. The most common use of this chip in a cable system is as a TDMA transmitter. Each subscriber sends bursts of data upstream during assigned time slots. These bursts are demodulated at the cable head end by a single receiver.

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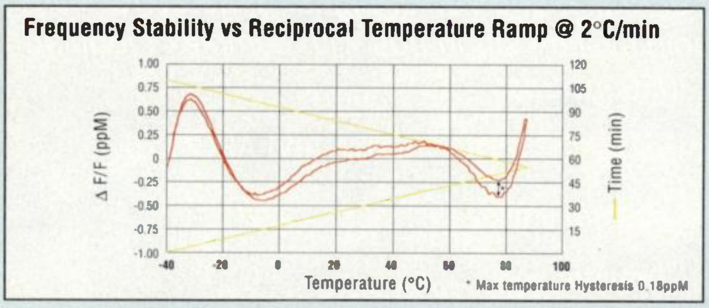

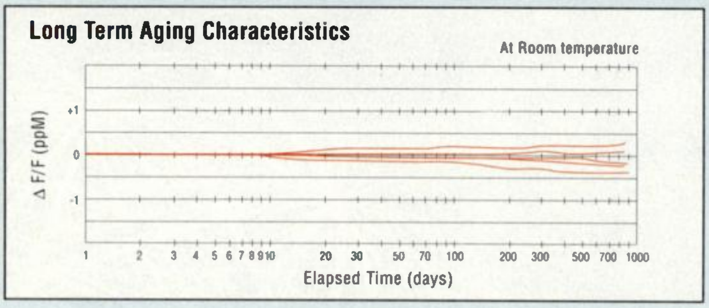

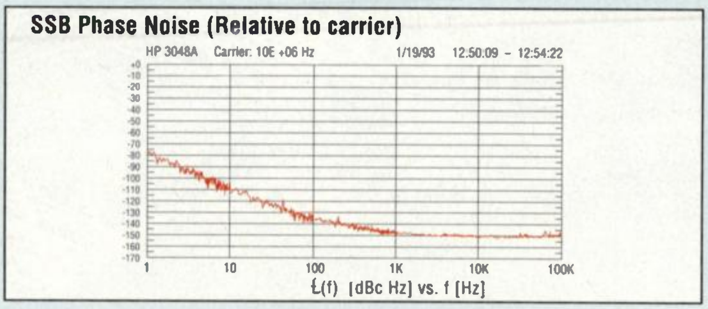

- Miniature size
- SMD Low profile package available
- Excellent temperature stability performance ±1 to 2.5 ppm from -40 to 85°C
- No frequency perturbations
- Low power consumption of only 1.5mA
- Very low Hysteresis
- Low parts in 10^{-10} Alan Variance
- G sensitivity of $2x10^{-10}$ typical
- Excellent vibration performance
- Clipped sinewave output
- Frequency range of 9.6 to 20 Mhz

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INFO/CARD 32

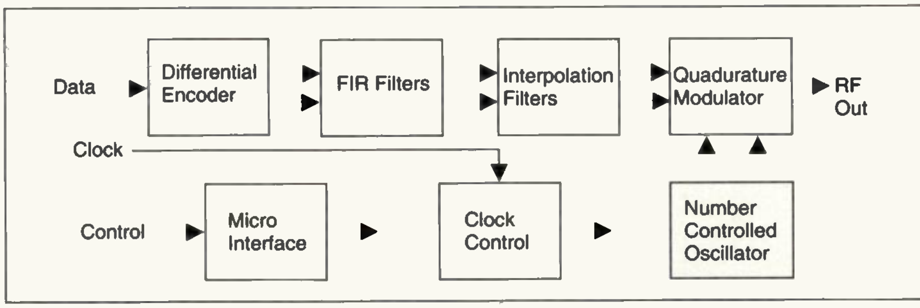


Figure 2. STEL-1103 digital PSK modulator block diagram.

One goal of the network controller is to maximize the system throughput by assigning time slots as close together as possible. The STEL-1103 supports the precision burst timing control required for TDMA systems.

In CATV applications, frequency division multiplexing (FDM) is also used in the 5 to 40 MHz return channel. Data rates in the range of 128 kbps up to 4 Mbps, with excess bandwidths of 20 to 50 percent are common. A one megabaud signal with 20 percent excess bandwidth will occupy 1.2 MHz of the return spectrum, allowing signals to be spaced 1.2 MHz apart. The STEL-1103 provides precise center frequency control and bandwidth shaping to support FDM systems.

Operation

A simplified block diagram of the chip is shown in Figure 2. Data is clocked into the STEL-1103 at the bit rate, which can be as high as 10 Mbps for QPSK modulation, or 5 Mbps for BPSK. The data may then be differentially encoded (optional). Differential encoding provides a simple way for a receiver to determine the correct reference phase (four phases are possible for QPSK) for reassembling the data in the right sequence. The data is next passed to dual I and Q 32 tap FIR filters operating at four times the symbol rate. These are symmetrical filters each having 16 programmable 10-bit coefficients. These filters permit precise spectral shaping to obtain excess bandwidths as low as 20 percent of the symbol rate and can be programmed for any desired shape such as square-root-raised-cosine.

If the filtered baseband data were to be digitally modulated at this point the results would not be desirable. The signal spectrum would fold about the sample rate which is now at four times the symbol rate (See Figure 3). This effect is eliminated by first passing the data through a multi-stage interpolation filter, which interpolates multiple points between those generated by the FIR filter. The effect of this process is to suppress the unwanted alias energy associated with the lower sampling rate. Figure 4 shows an example of a 2.5 Mbps QPSK signal centered at 20 MHz. Note the low spurious content.

Output frequency agility could be achieved without interpolation filtering; however, the FIR filters would have to be very large in order to accommodate the many samples per symbol necessary to run the chip at 100 MHz (1 Megabaud clocked at 100 MHz would require dual 100 tap filters).

Following the interpolation filter is the complex modulator which modulates the signal with a quadrature LO generated by a 24-bit NCO. The STEL-1103 stores three programmable frequency control words, enabling rapid switching of carrier frequencies. The final output of the complex modulator is a modulated carrier signal centered anywhere from DC to 40 MHz. The upper limit is determined by both the 100 MHz maximum clock rate and the requirement to filter out an unwanted alias signal located above one half the NCO clock frequency. It is also required that the master clock rate be selected to be an integer multiple of four times the symbol rate. For example, given a symbol rate of 1.25 Msps and a master clock rate of 100 MHz, the FIR filter will operate at 5 MHz, resulting in the need to interpolate 20 points.

Although the STEL-1103 can be used for continuous data transmission, it was specially designed for burst data transmission. The chip incorporates appropriate control lines to enable each burst to begin and end gracefully. Since the chip contains certain latencies, such as the 8-symbol delay of the FIR filters, the data is flushed at the end of each burst so

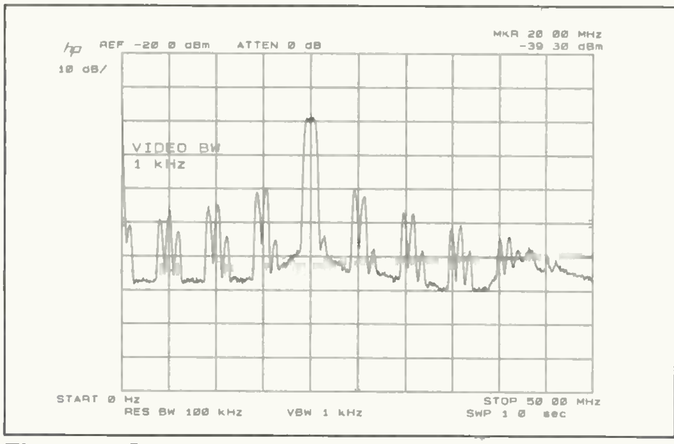


Figure 3. Spectrum of filtered signal without interpolation.

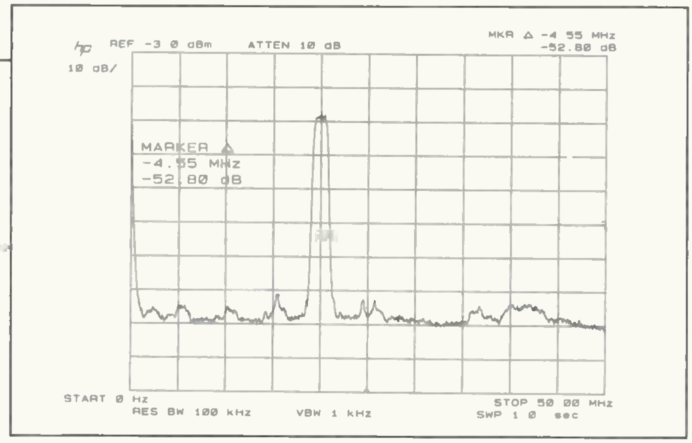


Figure 4. Spectrum of filtered signal with interpolation (1.25 Meg symbol per second, alpha = 0.3).

that each successive burst will contain no residue from the prior burst. Also, the output will trail-off to be exactly zero, representing greater than 100 dB on-off ratio, thereby eliminating the need for an RF transmit switch in burst mode applications.

In addition, the chip includes some other useful functions. It provides a re-synchronized DAC clock which has a programmable delay. This greatly simplifies interfacing with various DACs. It also includes a programmable (divide by 1 through 16) prescaled output of the main clock input. This function simplifies building systems which are synchronously locked to the STEL-1103's clock. This is useful for generating the data clock which must be synchronized with the STEL-1103's clock, or vice versa.

A direct analog modulator can be designed to have minimum phase and amplitude error only at singular frequencies. Tuning the output to other frequencies results in performance loss. More sophisticated analog IF modulation schemes require complex hardware and up or down converters. The STEL-1103 digital modulator has the advantage of high performance that remains constant as the output is varied over frequency, while achieving very low system complexity. It has nearly zero phase and amplitude error (the constellation is perfectly centered) over its entire output frequency range, thus enabling the best possible system performance. A result of this accuracy is excellent sideband rejection. Analog QPSK modulators achieve typical sideband rejection of only 40 dB (50 dB if hand trimmed). Spurious signals, such as unwanted sideband feedthrough, appear to a demodulator as an interferer, thereby reducing the receiver's performance. In contrast, the STEL-1103 has inherent sideband rejection of 75 dB without any need for trimming.

STEL-1103 Analog Filtering Requirements

Since the spectrum generated by the combined STEL-1103 and DAC contains aliases of the desired signal above the Nyquist frequency, a low-pass filter must be included at the modulator output. These aliases occur at integer multiples of the master clock rate plus and minus the center frequency of the desired signal. For example, with a clock rate of 100 MHz and a carrier output at 40 MHz, aliases will exist at 60, 140, 160, 240....MHz. The amplitude of these signals rolls off as a sinc function with nulls at multiples of the master clock rate. For the case of a 40 MHz upstream signal, the first alias, which occurs at 60 MHz, is right next to TV channels 2 and 3, and is only 3 dB down from the desired signal. The required lowpass filter typically must provide enough rejection at this frequency to keep the alias 60 dB lower than the analog TV signal.

Future Versions

The next integration step for this product involves combining the DAC on chip. The mixed-mode technology needed to economically produce complex, high-speed digital functions combined with 100 MHz DACs is just nearing viability. Stanford Telecom is working closely with select foundries on future versions which will incorporate the DAC. For maximum flexibility, however, these two high performance functions are best left as independent chips. The key drivers to further integration will be 1) the advancement of high-speed CMOS DAC technology, and 2) the specific economies associated with combining the two dissimilar functions and technologies.

Stanford Telecom is also presently designing a version of this chip which will generate 24 MHz maximum RF outputs. Many interactive CATV applications do not require the full 40 MHz output. This chip will be lower cost and will be an immediate candidate for having the DAC included on-board since 50 MHz CMOS DAC technology is already mature.

Evaluation Board

The STEL-1203A is an evaluation board based on the STEL-1103. It includes a 10-bit DAC, lowpass filter, programmable output level function, and an output amplifier. Also included is a Windows-based user interface for configuring all of the chips parameters.

Conclusion

The STEL-1103 Digital PSK Modulator chip is a good example of what can be achieved through innovative design and in-depth knowledge of a specific application's requirements. Based on the current selection of this chip by various high volume users, the STEL-1103 may soon become the industry standard choice for many systems. In order to maintain and expand its industry leadership position, Stanford Telecom is on an aggressive path to further advance this product as well as many other specialized communications products. Stanford Telecom is also supplying a number of other products to the interactive CATV market including the STEL-2105, which is an all-in-one digital QPSK receiver chip, as well as a number of custom, low-cost, board-level burst receivers that are being used in the host digital terminal (HDT) of Hybrid Fiber/Coax networks. The STEL-1103 is available in quantities of 10,000 pieces for under $20 each. For more information please contact the author or circle Info/Card #252.

RF

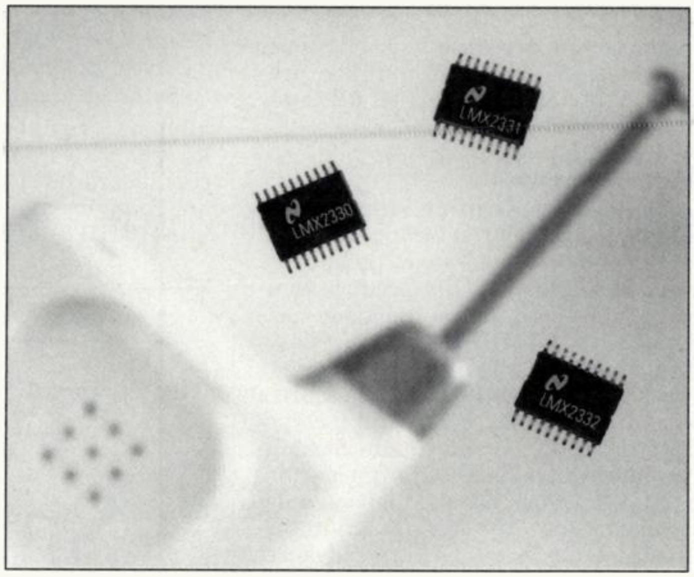
About the Author

Bill Xenakis is the Program Manager for Cable Products in the Telecom Products Group at Stanford Telecom. He has 10 years of experience in developing digital communications products at Stanford Telecom and holds a BSEE degree from California Polytechnic University. He can be reached at 480 Java Drive, Sunnyvale, CA 94089. Telephone: (408) 745-0818, x5548.

Dual Frequency Synthesizers

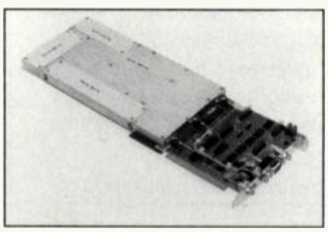
National Semiconductor's Double PLLatinum LMX233x family of monolithic, integrated dual frequency synthesizers, including prescalers, is designed to be used as a local oscillator for the RF and IF of a dual conversion transceiver. It is fabricated using National's ABiC IV silicon BiCMOS process. Members of the family include the LMX2330 (2.5 GHz and 510 MHz), LMX2331 (2.0 GHz and 510 MHz), and LMX2332 (1.2 GHz and 510 MHz). The LMX233x contains two dual modulus prescalers. A 64/65 or a 128/129 (32/33 or 64/65 in the 2.5 GHz LMX2330) can be selected for the RF synthesizer and a 8/9 or a 16/17 can be selected for the IF synthesizer. Using a digital PLL technique, the LMX233x can generate a very stable, low noise signal for the RF and IF local oscillator. Serial data is transferred into the LMX233x via a three wire interface (data, enable, clock). Supply voltage can range from 2.7 to 5.5 V. The LMX233x family features current consumption of 15 mA at 3 V for the LMX2330, 14 mA at 3 V for the LMX2331, and 8 mA at 3 V for the LMX2332. The family is available in a 20-pin TSSOP surface mount plastic package. Thousand unit pricing ranges from $6.50 to 9.95 per unit.

National Semiconductor
INFO/CARD #147



PC-Based Telemetry Receiver

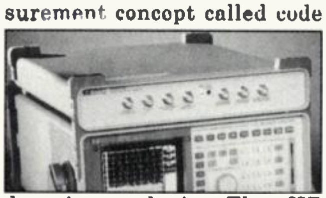
The PCTR-500 is a dual conversion, general purpose, synthesized, frequency selectable telemetry receiver integrated with a single slot, full length ISA bus (IBM-PC AT or compatible) personal computer card. The series PCTR-500 receivers provide reception and FM demodulation for telemetry signals transmitted in the L and S frequency bands. The PCTR-500 receivers employ a modular architecture that enables the user to configure the receiver for a variety of different data rates (up to 20 Mbps), IF bandwidths, FM deviation response and signal output voltage levels. The PCTR-500 receiver offers replaceable plug-in modules for the 2nd IF stage, FM discriminator, and post detection amplifier/filter. The receiver is controlled via a simple menu-driven DOS program that is supplied with the receiver, or can easily be integrated into a custom developed DOS or Windows application program.



Systems Engineering & Management Co.
INFO/CARD #146

CDMA Base Station Test

Hewlett-Packard has announced the availability of a code division multiple access (CDMA) adapter for the HP 8921A cell-site test set. The HP 83203B embodies a new measurement concept called code domain analysis. The HP 83203B, when connected to the HP test-set, measures power in each code channel, as well as measuring code-channel timing and code-channel phase relative to the pilot. The measure of waveform quality, ρ, is the normalized correlated power coefficient. It is a measure of the percentage of transmitted power that correlates to an ideal signal. This new measurement, developed by HP, indicates the overall performance level of the CDMA transmitter. In addition, the HP 83203B makes true-average and channel power-transmitter measurements, and provides a reverse CDMA channel source with data buffer with interim Standard-95-compliant error correction and channel coding for base-station receiver testing. U.S. price for the HP 83203B CDMA cellular adapter is $18,000.



Hewlett-Packard Co.
INFO/CARD #145

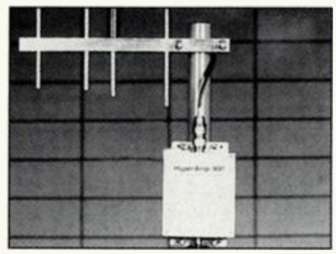
Push-Pull Power MOSFET

Motorola has added the MRF166W RF power MOSFET to its portfolio of UHF broadband components. This second generation N-channel enhancement mode device is characterized at 28 V and operates in the frequency range of DC to 500 MHz. Typical performance at 400 MHz is: output power = 40 W, gain = 13 dB, and efficiency = 50%. Typical performance at 175 MHz is: output power = 40 W, gain = 17 dB, and efficiency = 60%. Designed with a push-pull configuration to reduce even-numbered harmonics, the MRF166W also offers high thermal stability and a typical power gain of 13 dB at 400 MHz. The transistor is designed primarily for wideband large-signal output and driver applications and has the ability to withstand 30:1 load VSWRs of any phase angle. C_{rss} is 4.5 pF at VDS of 28 V. Pricing for the MRF155W is $83.33 in low volumes.

Motorola Semiconductor
INFO/CARD #144

Bilateral Power Amplifiers

HyperLink Technologies, Inc. has released the HyperAmp 900 and HyperAmp 2400 series of spread spectrum amplifiers/ range extenders for 900 MHz and 2.4 GHz wireless LAN adapters and radio modems. The devices amplifies in both transmit and receive modes. HyperAmp is a linear, wideband power amplifier suitable for digital and analog modulation. Consisting of a low-noise receive amplifier and a transmit power amplifier, HyperAmp is designed to be remote mounted at the antenna. A fast RF sensing circuit switches the unit between transmit and receive modes in under 3 μs, ensuring reliable operation with even the fastest spread spectrum radio equipment. HyperAmp provides receive gain of 19 - 24 dB, with models available with maximum output power of 100 mW, 250 mW, 1 W, 2.5 W, 5 W, and 10 W (output is variable for all units). Suggested retail prices range from $595 to $850.



HyperLink Technologies, Inc.
INFO/CARD #143

TEST EQUIPMENT

PCS Test Set

Racal Instruments has launched a low-cost, portable test set, designated Model 6113, for base transceiver stations (BTS). Applicable to E-GSM, PCS-1900 and DCS-1800 systems, the tester is well suited to

development, repair and maintenance as well as production testing. The Model 6113 permits E-GSM, PCS-1900 and DCS-1800 BTSs to be fully tested manually and automatically via both A-bis and RF interfaces.
Racal Instruments
INFO/CARD #142

Analog ATE

Pensar Technologies has released the UAT A-2000, a computer-based system of ATE (automated test equipment) designed for testing analog components on printed circuit boards. The UAT A-2000 is a 2 channel, PC-driven test device and software that can derive and store the I/V curves for parts on known good boards and compare those with the I/V curves of parts on boards to be tested. In addition, the tester automatically creates a database from each board tested. The UAT A-2000 sells for under $7,500.
Pensar Technologies, LLC
INFO/CARD #141

Signal Generators

Tektronix has introduced the SMP03 and SMP04 microwave signal generators. The SMP03 covers 2 to 27 GHz and the SMP04 covers 2 to 40 GHz, both with optional frequency extension down to 10 MHz. These signal sources, developed by Rohde & Schwarz, have comprehensive modulation capabilities, including AM from DC to 250 kHz, PM from DC to 100 kHz, pulse modulation, and FM down to DC.
Tektronix, Inc.
INFO/CARD #140

Mobile Radio Test

OCTAS by Berkeley Varitronics is a portable multichannel mobile radio test and measurement system for land mobile propagation analysis. Eight receivers can measure signal strength to within 0.5 dB, and navigational information is supplied by internal GPS and LORAN receivers as well as an odometer input from the vehicle. OCTAS is available for 935 to 941 MHz, 851 to 870 MHz, and 405 to 512 MHz.
Berkeley Varitronics Systems, Inc.
INFO/CARD #139

SIGNAL PROCESSING COMPONENTS

Bandpass Filter

KEL-Com bandpass filter model number 5MB6-1000/65 has a center frequency of 1000 MHz with a minimum 3 dB bandwidth of 65 MHz. Insertion loss is less than 3.0 dB at 1000 MHz. Maximum group delay deviation over 980 to 1020 MHz is 5 ns. Typical VSWR is 1.5:1, and minimum rejection at 850 and 1165 MHz is 50 dB. Outline dimensions are 0.97" L × 0.50" W × 0.25" H.
KEL-Com
INFO/CARD #138

Directional Coupler

Sage Laboratories announces the availability of a commercial grade 20 dB coupler for use over the 824 to 849 MHz cellular band. Model FC5621 offers insertion loss of 0.2 dB and input VSWR of 1.25:1. Typical directivity is 15 dB and flatness is ±0.20 dB. Average forward and reflected power is 200 W. Type N male and female connectors are utilized on the mainline, and an SMA female connector is used on the coupled port.
Sage Laboratories, Inc.
INFO/CARD #137

70 MHz X-tal Filter

Piezo Technology's model 7822C is a 70.0 Mhz fundamental crystal filter with a minimum 3 dB bandwidth of ±53 kHz and stopband attenuation of 50 dB at ±220 kHz maximum. Additional features include a 50 dB ulti-

mate attenuation, 50 dB spurious attenuation and 50 ohm input and output impedances. The unit is packaged in a 1.95 × 0.59 × 0.40 inch enclosure.
Piezo Technology, Inc.
INFO/CARD #136

TOOLS, MATERIALS & MANUFACTURING

Bonding Film

Polyflon bonding film is well suited for bonding single or multi-layered microwave stripline circuit boards. The bonding film is thermally stabilized, irradiated polyolefin co-polymer, and is recommended for use with circuit board substrates consisting of irradiated cross-linked polyolefin; virgin PTFE, PTFE fiberglass; or ABS. Standard thickness is 0.002 inches, and sheet size is 24 x 36 inches. Pricing is $12.84 per sheet in 100-sheet qty.
Polyflon Co.
INFO/CARD #135

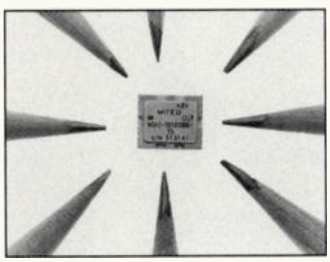
Aluminum Nitride Substrates

Sherritt has added R130 to its ThermicEdge line of aluminum nitride substrates. R130 substrates are available in large volumes and provide high lot-to-lot consistency. Aluminum nitride has thermal conductivity five times that of alumina and is non-toxic. Sherritt maintains an inventory of standard size substrates.
Sherritt, Inc.
INFO/CARD #134

AMPLIFIERS

Low Noise Amplifiers

MITEQ offers the NSH series building block LNAs. The NSH is available in either a solderable or screw-mount chassis and



is suitable for drop-in application with a variety of substrates. Typical performance includes a 2 to 18 GHz design with 14 dB minimum gain, 3.0 maximum noise figure, 2:1 input/output VSWR, and typical current consumption of 60 mA at 6 V.
MITEQ
INFO/CARD #133

10 W UHF Amp

ENI's model 411LA power amplifier produces 10 W of linear class A output over a frequency range of 150 kHz to 300 MHz. With a gain of 40 dB, the 411LA features low harmonic and intermodulation distortion, and complete protection against short and open circuit loads. The amplifier is also unconditionally stable, and has +13 dBm overdrive protection and infinte maximum load VSWR. The 411LA has integral power supply and forced-air cooling.
ENI
INFO/CARD #132

Feedforward Amp

Four, low profile, "laptop" feedforward cellular power amplifiers have been released by AML Communications. These units cover the the four instantaneous 25 MHz, USA and international transmit cellular bands. These amplifiers support 1 W of multicarrier power (10 W PEP) while maintaining intermodulation products at 55 dBc. The amplifiers operate on a single 26 V supply and provide 53 dB gain.
AML Communications
INFO/CARD #131

DISCRETE COMPONENTS

SMT Crystals

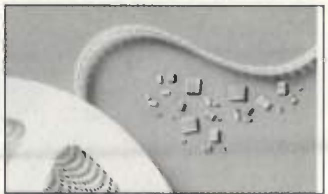
The TSMX-1S is a surface mount crystal measuring 7.5 x 5.0 x 1.3 mm. This low profile crystal from Tellurian Technologies was developed for applications where small size and SMD capability are mandatory. The series frequency range is from 12.0 to 150.0 MHz, with an operating temperature range of −10 to +70 °C. Standard stability is ±100 ppm, with tighter stabilities available on request.
Tellurian Technologies, Inc.
INFO/CARD #130

Surface Mount Inductors

Standex Electronics series 166 and 168 surface mount inductors are now available in ten values from 3.5 nH to 84 nH. These inductors have superior performance, especially in "Q", when compared to similar value "chip" inductors. The flat top allows vacuum pick and place. These components are available tape and reel or bulk packed. Prices are $0.16 each in volume quantities.
Standex Electronics
INFO/CARD #129

Multilayer Chip Capacitors

A line of low cost, general purpose multilayer ceramic chip capacitors, manufactured for American Technical Ceramics, is available for high volume, surface mount applications. The capacitors are offered in NPO, X7R, and Z5U dielectrics, in six



standard EIA case sizes: 0603, 0805, 1206, 1210, 1812, and 2225. Capacitance values range from 1 pF to 4.7 μF. Tolerances are: C, D, G, J, K, M, and Z.
American Technical Ceramics
INFO/CARD #128

CABLES & CONNECTORS

Low Insertion Loss Cable

40 GHz cables from Storm Products feature an insertion loss of 1.14 dB/ft., with a cable O.D. of 0.125 inches. Part number 421-011, an assembly using K plug connectors yields typical VSWR of < 1.30:1 up to 40 GHz. All cables are manufactured using low density PTFE and feature an operating temperature range of –55 to +150 °C.
Storm Products Co.
INFO/CARD #127

Patch Cable Kit

RF Industries' Unicable® Kit allows users to mix and match any combination of connectors or adapters to the ends of a 48-inch cable assembly. The cable is extra flexible RF-58A/U type 50 Ω cable with 95% double shielding. The Unidapt® universal connectors at each end feature machined brass, silver plated bodies, gold plated contacts, and PTFE dielectric insulators.
RF Industries, Ltd.
INFO/CARD #126

90° 7/16 Plug

A series of 7/16 right angle plugs that is compatible with Andrew LDF4-50A cable is being introduced by Tru-Connector. The right angle plug permits smoother cable runs. All mating dimensions conform to DIN 47223, IEC 1694, VG 95250, and CECC 22 190 specs. The 50 Ω plugs have a voltage rating of 2700 Vrms. The Tru-7631 right angle 7/16 plug is priced from a list price of $98.00.
Tru-Connector Corp.
INFO/CARD #125

45° and 90° BNC Connectors

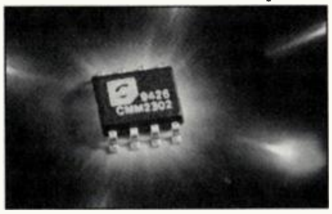
Trompeter's 45- and 90-degree BNCs use the same standard stripping dimensions and standard BNC crimp tools as their straight BNCs. The new BNC connectors feature fully heat-treated beryllium copper outer conductor springs. Cost for the 45-degree BNC, UPLFF220, and the 90-degree BNC, UPLR220, is $7.72 each in quantities of 1,000.
Trompeter Electronics, Inc.
INFO/CARD #124

INFO/CARD 44

SEMI-CONDUCTORS

Driver Amplifier

A low-cost driver amplifier from Celeritek operates from a single +5 V power supply and uses less than 10 mA. Model CMM2302 is designed for use as a low-current gain stage in space constrained PCMCIA applications in the 1.7 to 2.5 GHz. The wideband design of the CMM2302 allows the amplifier



to be used in many applications, including DECT DCS1800, PCS, and PHS. The amplifier is internally matched to 50 ohms and internally biased. The device produces about 12 dB of gain at signal levels up to +7 dBm. Production quantities of 1,000 are priced as low as $3.95 per unit.
Celeritek
INFO/CARD #123

High-Speed, Low-Cost Op Amp

The AD8011 operational amplifier offers 300 MHz bandwidth on a maximum supply current of just 1 mA. The op amp from Analog Devices consumes 5 mW maximum power from a single +5 V supply. The AD8011 offers 0.1 dB gain flatness to 25 MHz, 0.02% differential gain, and 0.06° differential phase error. The current feedback op amp also features worst case harmonic distortion of –62 dB at 20 MHz while driving a 150 Ω load. In 1,000 piece quantities, the AD8011 is priced at $1.95.
Analog Devices, Inc.
INFO/CARD #122

Low-Cost, High-Speed Op Amps

SGS-Thomson Microelectronics has introduced three new op amps that combine high speed (100 MHz - 300 MHz) with low input current, low noise, and low cost. The TSH10/11/31 use an advanced BiCMOS process, and have standard op-amp pinout and packaging. The voltage-feedback op-amps feature low noise (6 nV/Hz for the TSH10) and high input impedance (2 pA for the TSH11/31) Housed in DIP8 and SO8 packages, these devices cost less than $1 each in 10,000 piece quantities.
SGS-Thomson Microelectronics
INFO/CARD #121

SIGNAL SOURCES

VCO for Cellular

The miniature U-package VCO-190 series voltage-controlled oscillators from Vari-L is offered with center frequencies from 773 to 992 MHz and bandwidths of 26 MHz. Units operate with a supply voltage of 2.5 V to 5.0 V and exhibit ultra low current draw (typically 3.3 mA at 3.0 V). The VCOs are packaged in a 0.375 × 0.375 × 0.117 inch shielded surface mount package.
Vari-L Co., Inc.
INFO/CARD #120

Fast Tuning Synthesizer

A fast tuning synthesizer covering the 50 to 100 MHz frequency range has been announced by Communication Solutions. The synthesizer has 1 Hz resolution and <1.0 microsecond lock time. The unit is rack mounted, with a parallel digital input, built-in test, and modular construction for ease in maintenance. Additional frequency ranges are available.
Communication Solutions
INFO/CARD #119

Basestation VCO

Z-Communications has introduced the V502MC03 miniature surface mount VCO for cellular basestation applications operating in the 750 to 850 MHz region. The VCO uses a control voltage of 1 to 8 V and exhibits average gain of 40 MHz/V. The V502MC03 delivers 6 ±2 dBm into a 50 ohm load operating off a nominal +5 VDC bias, while drawing less than 25 mA. Typical phase noise performance is –95 dBc/Hz at 10 kHz offset. The device is packaged in a 0.50 × 0.50 × 0.20 inch package.
Z-Communications, Inc.
INFO/CARD #118

Nonlinear Analysis in RF Design

By Steve Maas, Ph.D.
Nonlinear Technologies, Inc.

Nonlinear circuit analysis is an important part of modern circuit design. Indeed, the relentless advances in the performance of digital circuits would not have been possible without it, and advances in communications circuits, although perhaps not quite as dramatic, owe a great deal to computerized analytical tools. In communications circuits the designer's ability to control phenomena such as nonlinear distortion is critical to the system's performance, so nonlinear circuit analysis is essential.

The importance of nonlinear circuit analysis has never been greater. New types of communications systems are remarkably sensitive to distortion phenomena. Interference – which is often manifest as intermodulation distortion (IM) – presents a fundamental limit to the performance of many types of military and commercial space communications systems, and is already a major concern in wireless applications. Of course, the need for good, "clean" mixers and amplifiers in more conventional RF systems has not diminished.

As the importance of such systems increases, the variety of nonlinear circuit-analysis programs has increased. Only a few years ago, the sole option for nonlinear circuit analysis was SPICE and its variants; now the user can choose between a wide variety of programs (with a wide variety of price tags!) including harmonic balance, time-domain, and Volterra simulators. In this article we will discuss these, describe how they work, and how they're best used.

What's Out There

Since its development at the University of California at Berkeley in the mid-1970s, SPICE [1] has been the circuit-analysis workhorse of the electronics industry. Since SPICE was made available as a public-domain program, a number of software companies have ported it to popular DOS and MacIntosh machines, added graphical and schematic-capture user interfaces, and graphical output. These have the same "engine" underneath the glitzy interfaces: SPICE version 2, or SPICE2 in common parlance. Berkeley has released versions of its next-generation simulator, SPICE3, a completely rewritten program, and it is starting to appear in commercial products. You can buy one of these, or even get your own copy of the source code from the Industrial Liaison office of the Berkeley Computer Science Department.

A more recent addition to the stable is the harmonic-balance (HB) simulator. Several of these are available from mainstream software companies. Again, one of the popular simulators was developed at UC Berkeley [2], but work at Sarnoff Labs and the University of Bologna [3], in Italy, found its way into mainstream programs.

Most of the mainstream HB programs are oriented toward microwave design. Models of RF devices and circuit elements are not widely represented in these simulators. Perhaps for this reason, in the RF community there seems to be some degree of disappointment with these simulators. This is indeed unfortunate, because HB simulators offer the RF user many significant benefits. At the same time, however, there are many things the HB simulators don't do well. We'll examine these issues a little later.

A third product is the Volterra simulator [4]. Volterra-series analysis predates HB, and certainly Volterra simulators existed long before either SPICE or the HB products. However, Volterra analysis never "caught on". This is unfortunate, since Volterra analysis works well for precisely the types of problems where HB is poor. HB and Volterra analysis are nearly perfect complements. At present, only one commercial product is available that uses Volterra analysis, produced by (wouldn't you know it) the author's company [5].

Time Domain Analysis

SPICE performs linear simulations of electronic circuits. It also performs *time-domain* analysis of nonlinear circuits, which in SPICE's documentation is called *transient analysis.* To analyze nonlinear circuits, SPICE sets up the circuit equations as a set of nonlinear differential equations, in matrix form, and integrates them in the time domain (see Figure 1). To do the integration, SPICE must obtain a solution iteratively at each of many successive time intervals; depending on the circuit, this is a long process, and is the reason why transient analysis is a slow process.

This simple description already illustrates the greatest difficulty in using SPICE for RF or microwave circuits: how does one describe lossy, dispersive transmission lines and other distributed circuits in the time domain? A circuit analyzed by SPICE

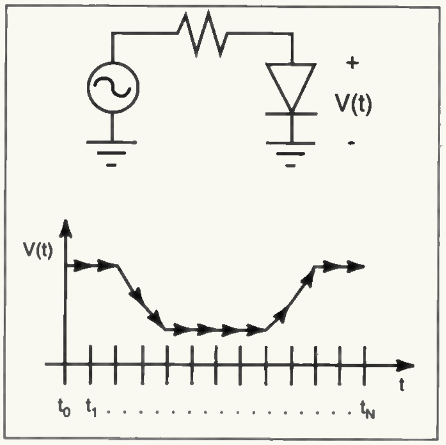
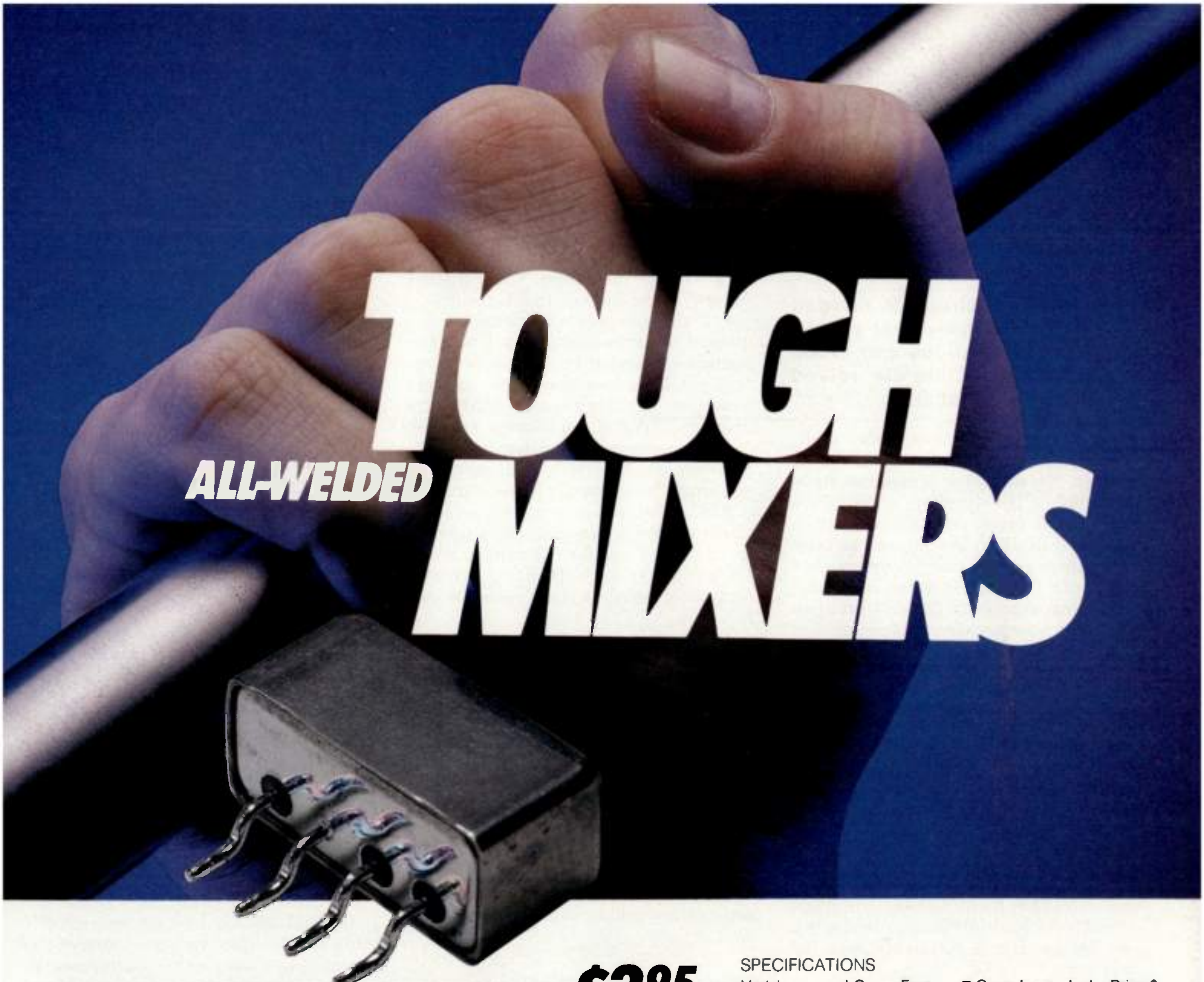


Figure 1. The voltages in the circuit are determined at a series of discrete time points, t_n, by numerically integrating a set of nonlinear differential equations.



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Model	LO Power (dBm)	Freq. LO/RF (MHz)	■ Conv. Loss (dB) $\bar{X}$	Conv. Loss (dB) δ	Isol. L-R (dB)	Price,$ Ea. 10 qty
TUF-3	7	0.15-400	4.98	0.34	46	5.95
TUF-3LH	10		4.8	0.37	51	7.95
TUF-3MH	13		5.0	0.33	46	8.95
TUF-3H	17		5.0	0.33	50	10.95
TUF-1	7	2-600	5.82	0.19	42	3.95
TUF-1LH	10		6.0	0.17	50	5.95
TUF-1MH	13		6.3	0.12	50	6.95
TUF-1H	17		5.9	0.18	50	8.95
TUF-2	7	50-1000	5.73	0.30	47	4.95
TUF-2LH	10		5.2	0.3	44	6.95
TUF-2MH	13		6.0	0.25	47	7.95
TUF-2H	17		6.2	0.22	47	9.95
TUF-5	7	20-1500	6.58	0.40	42	8.95
TUF-5LH	10		6.9	0.27	42	10.95
TUF-5MH	13		7.0	0.25	41	11.95
TUF-5H	17		7.5	0.17	50	13.95
TUF-860	7	860-1050	6.2	0.37	35	8.95
TUF-860LH	10		6.3	0.27	35	10.95
TUF-860MH	13		6.8	0.32	35	11.95
TUF-860H	17		6.8	0.31	38	13.95
TUF-11A	7	1400-1900	6.83	0.30	33	14.95
TUF-11ALH	10		7.0	0.20	36	16.95
TUF-11AMH	13		7.4	0.20	33	17.95
TUF-11AH	17		7.3	0.28	35	19.95

*To specify surface-mount models, add SM after P/N shown.
■ $\bar{X}$ = Average conversion loss at upper end of midband ($f_U/2$)
δ = Sigma or standard deviation

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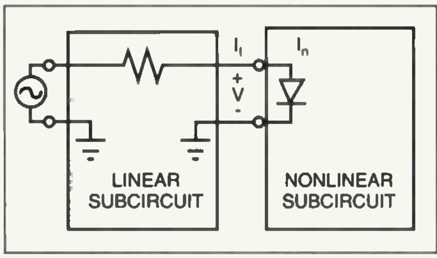

Figure 2. The circuit is divided into linear and non-linear subcircuits, and the nodes and loops between the circuits are solved for Kirchoff's equations.

may include ideal transmission lines, but cannot handle anything more complex (a few software companies have modified SPICE to allow more versatility in the use of transmission lines). A second problem is the need to integrate through long transient responses to obtain the steady-state response; this is an especially difficult problem in high-frequency circuits having long time constants. SPICE3 includes some features that minimize this problem, but SPICE2 doesn't. Finally, SPICE goes into cardiac arrest if a circuit contains loops of capacitors and voltage sources or cutsets of inductors and current sources; many matching and filter circuits have such structures.

SPICE is nearly ideal, however, for switching and logic circuits. It can simulate a wide variety of nonlinear devices, circuits, and phenomena. For example, SPICE can be used (although sometimes with difficulty) to design oscillators; HB is relatively poor for this task.

Harmonic Balance

Harmonic-balance analysis effectively partitions a circuit into two subcircuits: the *linear subcircuit,* which contains all the linear parts, and the *nonlinear subcircuit,* which contains only nonlinear elements. The two subcircuits are connected by a number of ports. Figure 2 illustrates this partition.

The voltages at these interconnecting ports – the DC, fundamental frequency, and its harmonics – are treated as circuit variables. The harmonic-balance process iteratively tries to find a set of these voltages that satisfies (1) the linear circuit equations of the linear subcircuit (usually a multiport admittance matrix), and (2) the nonlinear equations describing the nonlinear subcircuit. The linear equations are easily solved in the frequency domain, but the nonlinear equations must be solved in the time domain. The frequency-and time-domain quantities are related by a Fourier series.

Again, an astute reader might anticipate the well known difficulties:

1. The iterative solution process has no guarantee of success. "Convergence failure" is the bane of all HB users.

2. If there is only one excitation frequency, all is well; however, if two frequencies are used (e.g., in a two-tone test for IM), how does one convert between the frequency and time domains? Fourier transforms for such "noncommensurate" frequencies exist, but they don't work as well as classical, single-tone Fourier transforms.

3. Fourier transforms (especially those used for two-tone IM analysis) have limited numerical range. Weak IM products are often lost in the numerical noise of the Fourier transform.

4. Not obvious from the above description is the fact that each iteration of the solution process requires inversion of a huge matrix. For this reason, HB analysis is inherently very slow; reasonably efficient HB analysis of a large circuit requires a lot of memory and computer power.

In spite of these difficulties, and partly because of aggressive marketing, HB simulators have become the most popular tools for analysis of nonlinear microwave circuits, and they are rapidly taking over RF design as well.

Volterra-Series Analysis

Volterra-series analysis can be described as a perturbation method: the RF signal is viewed as a small perturbation of the DC bias voltage at each node in the circuit. This is not much different from the linear equivalent circuits with which we are already familiar. For example, the conductance g_d of a biased diode junction is:

$$g_d = \left.\frac{dI_d}{dV_d}\right|_{V_d = V_{bias}} \tag{1}$$

where V_d is the junction voltage, I_d is the current, and V_{bias} is the bias voltage. From this, the small-signal junction current under RF excitation i_d is:

$$i_d = g_d v_d = \frac{dI_d}{dV_d} v_d \tag{2}$$

with $V_d = V_{bias}$ and where v_d is the small-signal RF junction voltage. This is simply the linear response of the diode.

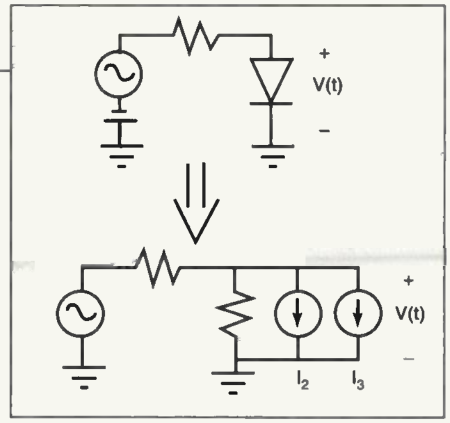

Figure 3. The circuit is represented by linear, second-, and third-order terms.

Equation 2 can be viewed as the first term in the Taylor-series expansion of the nonlinear drain current,

$$i_d = \frac{dI_d}{dV_d} v_d + \frac{1}{2}\frac{d^2I_d}{dV_d^2} v_d^2 + \frac{1}{6}\frac{d^3I_d}{dV_d^3} v_d^3 + \cdots \tag{3}$$

where the obviously nonlinear v_d^2 and v_d terms account for the distortion in the device. Volterra series analysis provides a recursive method for calculating the current represented by these terms directly in the frequency domain (Figure 3). These currents are then treated as excitations of the linear circuit; as a result, only *linear* circuit analysis need be performed! For this reason Volterra series analysis is much faster than either harmonic-balance or time-domain analysis. Furthermore, no Fourier transforms are used, so the numerical range is limited only by machine precision. Very low IM levels can be calculated by Volterra analysis.

This remarkable efficiency has a price, evidenced by equation 3: the Taylor series of equation 3 is accurate only for small voltage deviations near the bias point. If the signal is too large, accuracy suffers. Volterra series analysis is useful only for weakly nonlinear circuits, or circuits operated well below saturation. However, it is very good for analysis of intermodulation distortion, and most RF engineers will agree that calculating IM levels is a major interest.

C/NL2 for Windows from Nonlinear Technologies [5] is the only real Volterra simulator on the market. The distortion analysis in SPICE is, in fact, a type of Volterra analysis; however, it is applicable only to diodes and bipolar transistors, and because of poor device modeling, is not very accurate.

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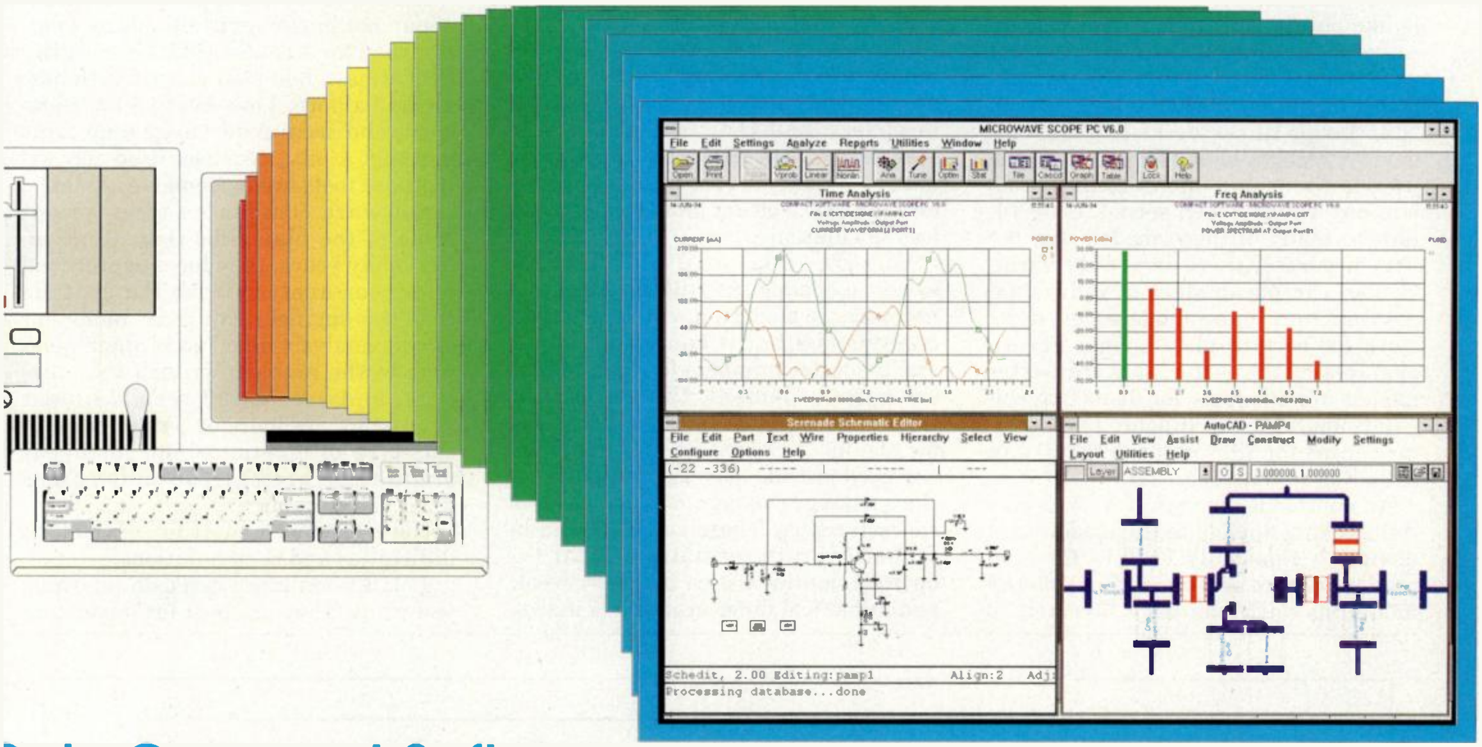

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Device Modeling

Nonlinear analysis is only half the task. Analyzing a circuit without accurately modeling the solid-state device is like painting a house without cleaning and scraping it—the results just won't hold up.

SPICE has been around long enough that large libraries of commercial devices are available. These tempt the user to think that the modeling problem has already been solved. Unfortunately, many of these models are not very accurate; there is a wide variation in the parameters of individual devices, and many models just don't work for certain phenomena. For an example of the latter, few transistor models are adequate for distortion calculations, and the bipolar transistor modeling for distortion analysis in SPICE is seriously flawed.

An enormous amount of work is currently being applied to the modeling of devices – especially FETs – for harmonic-balance analysis. Most of the modeling work has been directed to single-tone power circuits: power amplifiers, frequency multipliers, and similar components. Virtually all the "canned" FET models in HB simulators are inadequate for IM analysis. Bipolar models are better for IM, but still not very accurate. Diode models are OK, but diodes are used primarily in mixers, and HB simulators are so inaccurate that they are useless for mixer IM analysis. HB simulators generally work well for mixer conversion loss calculations.

Volterra-series analysis offers the easiest modeling task [6]-[8]. All that is necessary to model a device for Volterra analysis is a linear equivalent circuit and two more numbers for each nonlinear circuit element: the second and third derivatives in (3). This is true also for nonlinear capacitors and inductors; one need provide only the derivatives of their charge-voltage or flux current characteristics. These can be found by numerical differentiation, or, if the device's nonlinearities are very weak, and numerical differentiation is inaccurate, by RF measurements.

Why Doesn't It Work?

I have observed a general malaise about nonlinear analysis. Many engineers who are comfortable using SPICE have had poor results with harmonic balance, have become disheartened, and seem ready to give up. This is really a pity, because, used correctly, these tools are extremely valuable.

Software companies must accept some of the blame for this situation. For many years, they have approached nonlinear analysis with the attitude that the engineer can treat nonlinear circuit analysis as a "no brainer"; just plug in the problem, crunch the numbers, and accept the result. Linear simulators are more or less tolerant of this type of ignorance, but nonlinear simulators aren't. You must know what you're doing.

Here, then, is a short list of common difficulties and their solutions:

1. HB simulators can't do anything you want. They are best for mixer con-

version loss analysis and single-tone power circuits. They are virtually useless for mixer IM, IM in saturated power amplifiers, or mixer compression levels. The simulator is OK for IM in small-signal amplifiers, but the device models are probably inaccurate. You might want to consider adding a "user-defined" model for FETs.

2. Remember, your device and circuit-element models must be valid at DC and all harmonics as well as the fundamental frequency.

3. Use conversion-matrix analysis if available, not two-tone HB, for mixer calculations. It is much faster and more accurate.

4. Don't use more harmonics than necessary. Four or five is usually enough for FET circuits, unless they are driven into hard saturation; diode circuits, because of their stronger nonlinearities, may require eight.

5. When you have a new problem, start out simple. Use only ideal elements (e.g., an ideal transformer instead of a balun) and leave out complex bias circuits, etc. When you have the basic design established, start replacing the idealized elements, one at a time, with real ones. Trying to stuff all the complex things into the circuit at the outset leaves you with too many variables; then you can't optimize the circuit.

6. Accuracy in HB is usually reasonable only for fundamental-frequency outputs and low harmonics. Accuracy degrades for higher harmonics. An attempt to calculate the fifth-harmonic output level of a power amplifier is probably futile.

7. Many nonlinear phenomena (especially high-order IM, spurious responses, and harmonics) are very sensitive to signal level and circuit parameters. This high sensitivity inevitably degrades accuracy. For example, don't expect to be able to simulate fifth-order IM or a high-order mixer spurious response. It just won't work.

8. In the final analysis, harmonic-balance analysis is a poor choice for IM analysis. Consider using Volterra analysis for IM problems.

Convergence problems

1. Make sure that a node is not isolated at DC. For example, a node connecting two capacitors has an indeterminate DC voltage.

2. The process of partitioning the circuit into linear and nonlinear subcircuits can cause difficulties. Often, this leaves part of the linear circuit disconnected, and numerical problems result. Adding large-value resistors (e.g., 1 meg) between nodes of nonlinear elements and ground often solves this problem.

3. It is easy to enter data for device models that just won't work. Be sure your drain current function really reaches zero at pinchoff, and does not rise again below pinchoff. Check the transconductance and linear gain predicted by the nonlinear model, and be sure it is consistent with measured S or Y parameters.

4. Convergence problems are worse at high power levels. Often it is faster to perform a simulation over a range of input power levels instead of just one high level. HB simulators often use the previous result as the initial estimate of the next. Convergence at many small steps is fast; one big step, from zero to a high level, is much slower.

Conclusions

Nonlinear distortion, and other nonlinear effects, are among the meanest beasts in the RF wilderness. Nonlinear circuit simulation is the thing that can tame them. These tools, however, are not enough; inevitably, success requires some intelligent involvement of the user. It is especially important to choose the right tool for the problem at hand. I hope this article helps in that regard. *RF*

About the Author

Steve Maas is President and Principal Consultant of Nonlinear Technologies, Inc., which specializes in nonlinear circuit analysis, software, and consulting. He received BSEE and MSEE degrees from the University of Pennsylvania in 1971 and 1972, respectively, and a Ph. D. in Electrical Engineering from UCLA in 1984. Steve is the author of *Microwave Mixers*, (Artech House, 1986 and 1992) and *Nonlinear Microwave Circuits* (Artech House, 1988). From 1990 to 1992 he was the editor of the *IEEE Transactions on Microwave Theory and Techniques*. He received the *Microwave Prize* in 1989 for his work on distortion in diode mixers. He is a Fellow of the IEEE. He can be reached at P.O. Box 7284, Long Beach, CA 90807, or by phone/fax at: (310) 426-1639.

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CircleX Program Analyzes S Parameter Data

By Chris Buckingham

CircleX is a simple yet powerful tool for inspecting and investigating transistor S parameter data sheets provided by the industry. With CircleX one can view the stability and gain properties of any given device. By displaying the parameter boundaries on the Smith chart the user is able to visualize the overall properties at a glance. This contrasts dramatically with looking through rows and columns of tabulated data.

CircleX is composed of two blocks: an input data editor, and a stability/gain calculator and display module.

Data Editor

CircleX accepts common-emitter (source) two-port data in S, y, or h formats. The input editor can read from one to ten sets of data entered via the keyboard and can be stored and retrieved from disk. The file format can be viewed and edited with any line/screen text editor, allowing the addition of comments, titles, or any other information to the file. While in the editor, the user selects a single set (frequency) at a time, and once selected the user can jump out of the editor and go to the stability/gain calculator.

Stability/Gain Calculator

The stability and gain calculations start out in text mode and give the user the choice of investigating one of four configurations for the same device:
Common-Emitter (Source)
Common-Base (Gate)
Common-Collector (Drain)
or Cascode

Once a configuration is selected, the user can proceed with the calculations directly in text mode or use the graphics interface by pressing the "F1" key. We will describe the graphics display to explain the overall function of CircleX, the text mode being the same except that it accesses the equations directly, prompting the user for specific inputs.

After pressing the "F1" key, the display will appear as shown in Figure 1. Here we have entered the common emitter data for a bipolar transistor at 1 GHz and 10 V, 10 mA bias (see HXTR3103.S file), and examine the common-base configuration and some noise gain circles for the purpose of illustration.

There are two planes. The left side is the source-plane, the right side the load-plane. The source-plane represents all possible impedances that can attach to the device input port, and the load-plane represents all impedances that can attach to the output port. These impedances are shown on polar Smith charts and we have simultaneous display of the termination reflection coefficient and its corresponding impedance.

The cursor keys can be used to position the cross-hair anywhere on the chart. The cross-hair represents the independent cursor (the cursor we can move around at will). The opposite plane will have a circular cursor which is fixed (a dependent cursor, to be explained shortly).

By default we are placed actively in the source-plane, but we can jump to either plane (input or output) by pressing the "F1" key for the source side or the "F2" key for the load side. When we say the "active" side we mean that we can select any impedance (source or load) with the arrow (cursor) keys. At default the cursors are homed at the center of the chart. This is the 50 Ω, zero reflection coefficient point on the chart, which is the condition where we measured our S parameters.

The control keys at the bottom of the

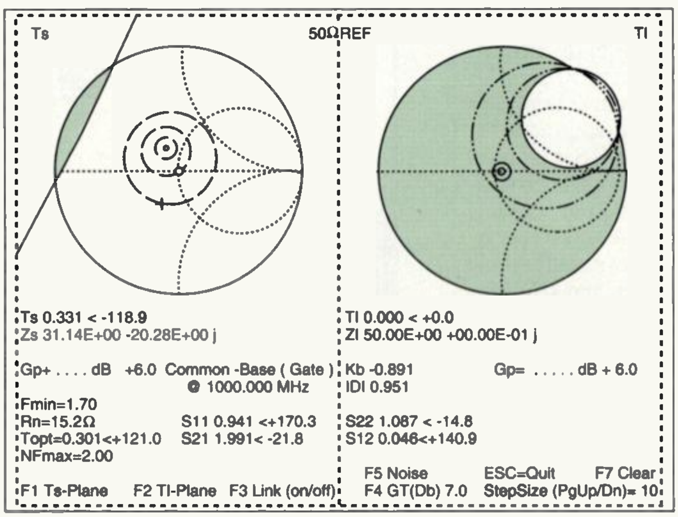

Figure 1. CircleX graphics display.

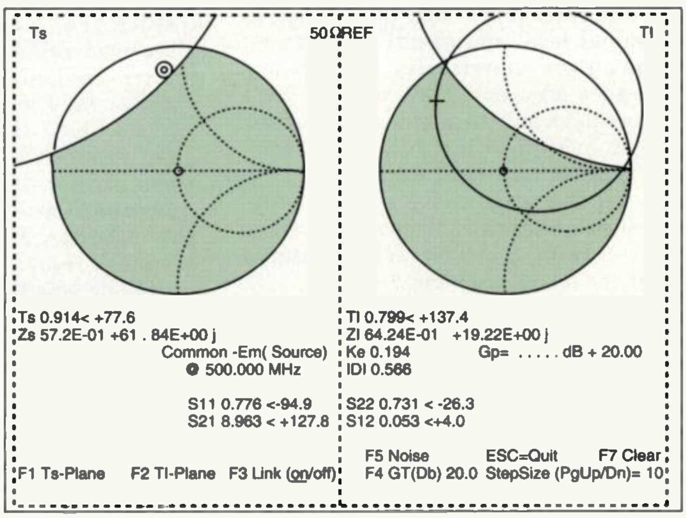

Figure 2. BFR90A design example.

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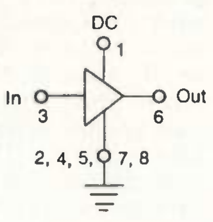

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F 180 Rev Orig

display are shown in Table 1.

Referring back to Figure 1, the stability circles are calculated and plotted for both the source and the load planes and the stable regions are then filled in green. These represent the impedances that produce stable outputs at the opposite port, which can then be conjugate matched. Therefore, any impedance on the Smith chart that is not in the green cannot be conjugate matched, and instability will occur if this is attempted.

If a device configuration is unconditionally stable, then both planes will be completely green, meaning that any combination of source/load terminations can be selected and, therefore, simultaneous conjugate matching at both ports will provide the maximum available gain and the lowest return losses. However, in most cases, particularly at lower frequencies where gain is high, we invariably have a conditionally stable configuration only, which requires careful selection of impedances.

The "F3" Link function

We will describe the "Link" process by illustration. If we select an independent source impedance by moving the cross-hair to some value of choice and then toggle "F3" on, the program will calculate the output impedance of the opposite port (in this case the load side), and will move the circular (dependent) cursor to the conjugate matching impedance and will then calculate the overall transducer gain GT in decibels and display the value next to the "F4" key automatically. This means that the independent cursor can be moved around, and the dependent cursor will track as described. The catch is that if the independent cursor is not in the green area, the dependent conjugate match will not be realizable, and the value of the impedance will fall outside the Smith chart.

When an unstable condition occurs during "Link", the dependent cursor will be placed on the outside periphery of the Smith chart to indicate this condition and will not move off the screen. However, the actual value of the impedance is displayed numerically bellow the respective Smith chart. This is valuable information if one wishes to evaluate the potential of the amplifier as an oscillator! It becomes a sort of reverse logic and we won't delve into it here, but this can be useful in "negative resistance" oscillator design work.

F1	selects the Source-Plane (Ts), as the active cursor
F2	selects the Load-Plane (Tl), as the active cursor
F3	link(on/off) will toggle the Link function on or off
F4	calculates and displays transducer gain value GT in dB
F5	activates noise parameter data entry/editing
F6	not used
F7	clears or resets the display (same as exiting and entering again) use this key to unclutter the screen of too many circles, etc.
"Page Up"	(PgUp) increases cursor step size by ten fold
"Page Down"	(PgDn) decreases cursor step size by ten fold
"Home"	returns the independent cross-hair to home base (center of graph)
"Esc"	when in the F5 noise editor, Esc jumps you back to previous entry fields
"Esc"	quits the graphics display (all data is erased!), returns to text mode
"Enter"	press to enter numeric data or skip to next entry
"Left/Right" arrows	move the cursor horizontally
"Up/Down" arrows	move the cursor vertically
"Shift/Print Screen"	used to dump graphics display (need GRAPHICS. COM)

Table 1. CircleX command keys.

Disabling the "F3" link function will not return the dependent cursor to "home", but will simply leave it where it was last, and GT will not be automatically updated. This leads to the "F4" key, which is used to calculate the transducer gain GT at a pinch, so that it will give us the gain for the existing source and load impedance settings. For example, if we "home" the source and load cursors (both at center of charts) and press "F4", we will get the insertion gain ($10\log10(S_{21})$^2) dB. This is the gain obtained if the device is driven directly with a 50 Ω source and the output is loaded with a 50 Ω resistor. This condition was used to measure our S parameters on the network analyzer.

The "F5" Noise key

This key allows for the display of noise circles in the source (input) plane and is used to enter optimum source reflection data to minimize amplifier noise. Manufacturers such as HP, Avantek, Motorola, and several others provide this data, particularly with gallium arsenide FETs and bipolars used in UHF and microwave applications. Pressing "F5" will activate an entry field on the left bottom side of the screen. Once noise parameters are successfully entered, we can display constant noise circles on the source plane and evaluate the stability and associated power gains for various source impedances.

Finally, all that is left is the numerical entry that we make. If we are in the source-input plane and the prompt cursor is in the Gp field, entering a number (dB) will display the input constant power circle. The last entry is always highlighted and any previous entries are darkened for reference. The same conditions apply to NF (noise figure in dB). To get into the NF field from Gp, hit the "Enter" key without a numerical entry, and do the same to go from NF to Gp.

When we plot constant power circles, the gain in decibels associated with that circle (Gp operating gain) can only be realized for impedances that fall inside the Smith chart and in the green area as well as on the power circle locus.The power gain circles require or assume a conjugate match at the opposite port.

Interestingly enough, CircleX really doesn't need the power circle. It suffices to turn on "Link" and move around until one obtains the gain required. However, power circles give a sense of range and scale, and indicate the boundaries at a glance.

To wrap up we will describe the filing format used in CircleX and complete the discussion with an example tested in Berkeley 2G.6 Spice.

File Management

We start with data taken from manufacturers or measurements and we enter the data manually via the keyboard using CircleX's editor or, via a word processor. CircleX will only read and write S parameter files. The editor will accept y and h data in rectangular format that is entered directly into the editor via keystrokes. These are then internally converted to polar S format which can be saved to disk.

CircleX has a small library of files to

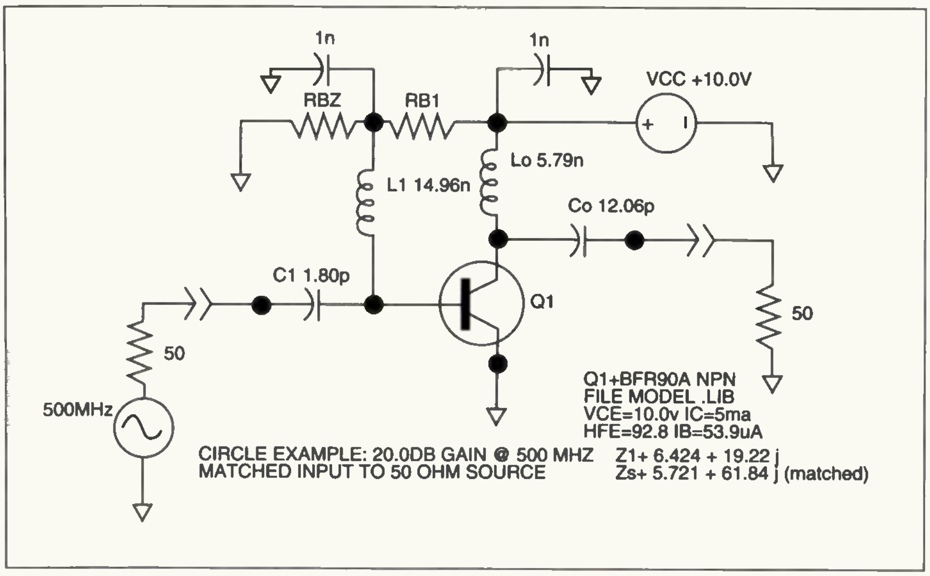

Figure 3. BFR90A amplifier schematic.

illustrate its operation. The directory is as follows:

AT-00572.S	1758
AT-01672.S	1764
AT-41472.S	1757
AT-8111.S	2088
BFR90A.S	377
BFR96.S	694
MRF901.S	593
HXTR3103.S	2099
EMPTY.S	1657

Each user can create an unlimited number of files.

All files have the “S” extension for convenience, but any DOS path\file naming convention can be used. If you don’t remember the file name while in CircleX, then you can “shell” to DOS, by pressing the “F7” key while in the editor window, and then find a “DIR”.

To load a file you must type the file name and the extension exactly as it appears in the disk directory into the “FILE NAME” window, then press the “Enter” key. If the file is located and data is found, then the editor will stop at the highest set# found in the file, or stop at set# 10, whichever is smallest. File “EMPTY.S” has no data, and if loaded into CircleX, it clears all set# buffers.

To save a file you must be in the “FILE NAME” window and press control-S to invoke the file save mode, then you type the file name you wish to use and press the “Enter” key. The data will be saved up to the set# displayed in the editor window, this means that to save all sets you must be at the last (highest) set#. Also, CircleX will not alert you of any existing file names, and does not create file backups. Any overwrites will eradicate any existing comment fields created with a text editor. Save the data first with CircleX, and then add comments and additional information with the text editor later.

If we look at file HXTR3103.S with a text editor we can examine the basic rules used to file the data.

* Salad S-Parameter File *
* HP General Purpose BJT 2N6838 (HXTR-3103) VCE=10 V Ic=10 mA 25C
* Noise Data

* Freq (GHz)	Fmin (dB)	Topt (Mag)	Angle	Rn (Ω)
* 0.5	1.4	0.121	96	114.4
* 1.0	1.7	0.301	121	15.2
* 2.0	2.5	0.461	173	5.2
* 3.0	3.3	0.553	–157	8.4
* 4.0	4.2	0.648	–139	13.4

SET# 1
FREQ1.00000E+08
S11 70.50000E-02 < –5.00000E+01
S21 24.26600E+00 < +1.49000E+02
S22 90.30000E-02 < –2.00000E+01
S12 15.00000E-03 < +6.00000E+01

SET# 2
FREQ 2.00000E+08
S11 64.20000E-02 < –8.60000E+01
S21 19.05500E+00 < +1.29000E+02
S22 74.20000E-02 < –3.10000E+01
S12 25.00000E-03 < +4.90000E+01

SET# 3
FREQ 3.00000E+08
S11 60.60000E-02 < –1.10000E+02
S21 14.96200E+00 < +1.16000E+02
S22 62.40000E-02 < –3.70000E+01
S12 30.00000E-03 < +4.30000E+01

SET# 4
FREQ 4.00000E+08
S11 57.90000E-02 < –1.27000E+02
S21 12.02300E+00 < +1.08000E+02
S22 54.80000E-02 < –3.80000E+01
S12 33.00000E-03 < +4.10000E+01

SET# 5
FREQ 5.00000E+08
S11 56.50000E-02 < –1.39000E+02
S21 10.11600E+00 < +1.02000E+02
S22 49.90000E-02 < –4.00000E+01
S12 36.00000E-03 < +4.10000E+01

SET# 6
FREQ 6.00000E+08
S11 55.20000E-02 < –1.49000E+02
S21 86.10000E-01 < +9.70000E+01
S22 46.70000E-02 < –4.00000E+01
S12 38.00000E-03 < +4.20000E+01

SET# 7
FREQ 7.00000E+08
S11 56.20000E-02 < –1.57000E+02
S21 74.99000E-01 < +9.30000E+01
S22 44.10000E-02 < –4.20000E+01
S12 40.00000E-03 < +4.30000E+01

SET# 8
FREQ 8.00000E+08
S11 55.90000E-02 < –1.62000E+02
S21 66.83000E-01 < +8.90000E+01
S22 43.00000E-02 < –4.20000E+01
S12 43.00000E-03 < +4.40000E+01

SET# 9
FREQ 9.00000E+08
S11 55.80000E-02 < –1.68000E+02
S21 59.57000E-01 < +8.50000E+01
S22 41.20000E-02 < –4.30000E+01
S12 45.00000E-03 < +4.60000E+01

SET# 10
FREQ1.00000E+09
S11 57.10000E-02 < –1.69000E+02
S21 53.30000E-01 < +7.80000E+01
S22 40.80000E-02 < –4.00000E+01
S12 52.00000E-03 < +4.20000E+01

* End of File *

The file shown above was created in CircleX and uses the following key words to define each data field: SET#, FREQ, S11, S21, S22, and S12. The magnitude values are delineated from the angle values by the less than sign “<”. A space “ ”, or comma “,” is also acceptable. Comments can be placed

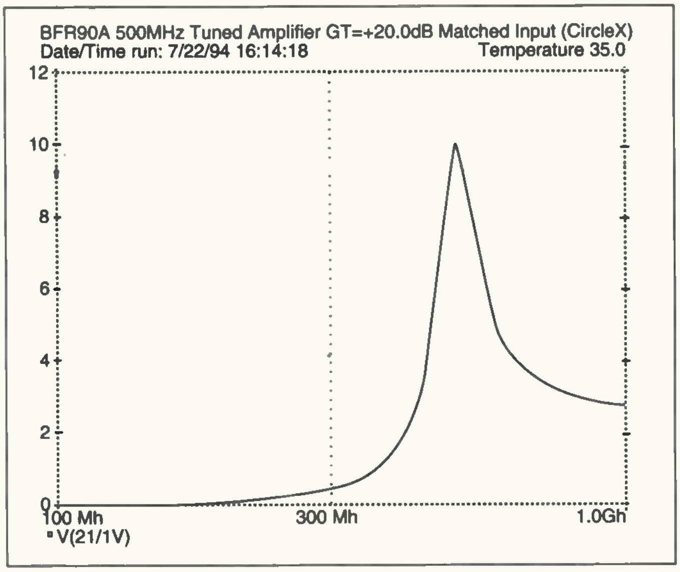

Figure 4. BFR90A insertion gain S_{21} magnitude.

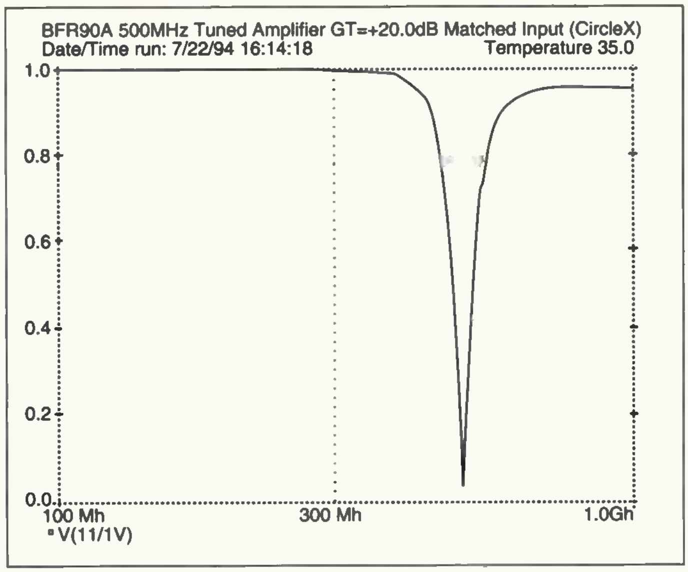

Figure 5. BFR90A input reflection S_{11} magnitude.

anywhere, and should be preceded with an asterisk "*" to avoid confusion with the key data field words. The S values can be entered in any order desired; S_{11}-S_{22}-S_{21}-S_{12}, or S_{21}-S_{12}-S_{11}-S_{22}, etc. Numerical entries can be in any format, as in the following example:

250.450 MHz – 2.50450E+08 = 250450000 = 250450K

In order to demonstrate how CircleX can be used, we will simulate a typical scenario by running a SPICE analysis. Start with a bipolar NPN transistor model. In this case we choose model BFR90A, and bias the device at VCE=10 V, and I_c=5 mA, and also choose to set the temperature at 35° C. The common-emitter S parameters are then measured at 500.000 MHz and entered into CircleX via the editor keyboard.

Once loaded into CircleX, we move to the stability/gain calculator and select Ke (common emitter) configuration, and activate the graphics display. Figure 2 shows the resulting stability regions for this device. The device is conditionally stable, and we choose to evaluate the device for a gain of 20 dB with the input matched to the source. In order to accomplish this, we jump to the load-plane by pressing "F2", and enter 20.0 in the Gp field. We can now move the independent cursor to a point on the 20.0 dB power circle. In this case we move the load impedance to 6.424+19.22j Ω and activate the link "F3" function. The resulting matched source impedance appears in the upper left quadrant of the source-plane with a value of 5.721+61.84j Ω.

Since we are going to use a 50 Ω measurement system to analyze the amplifier, we design simple matching "L" networks to transform 50 Ω into the source and load terminations indicated above by CircleX at 50 MHz:

	Resistor side	Transform side
Source (50 Ω)	C_{series}1.80 pF	L_{shunt}14.96 nH
Load (50 Ω)	C_{series}12.06 pF	L_{shunt}5.79 nH

We can now assemble the amplifier as shown in Figure 3, and submit the network list to SPICE and remeasure the two-port parameters of the BFR90A device with these matching networks attached. The results are shown in Figure 4 and Figure 5, and in this case we get a narrow band response that provides a 20 dB power gain at 500 MHz with a very low return loss at the input port.

The choice of where to pick values on the constant power circle is really a cut and dried process. There are infinite possibilities. However, the characteristics of the amplifier will vary considerably from point to point, and especially from quadrant to quadrant. Experimentation will show that bandwidth, noise, return loss, tunability, stability, and other factors all change along a constant power circle.

Requirements

CircleX will run on 386 or higher PCs with a VGA adapter/display system, platform DOS version 4.0 or higher.

Program CircleX is a subset of program SALAD version 1.08 (Spice Accessory for Linear Amplifier Design). It is available from Argus Direct Marketing. To order, see ad on page 69. *RF*

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9. Power Basic Version 3.0, Copyright 1992 by Robert S. Zale.

About the Author

Chris Buckingham graduated with a BSEE from Northeastern University in Boston, MA. He was employed by Teledyne as a test engineer in discrete semiconductors, and later worked for ten years at the Raytheon Company as a product engineer, specializing in small-signal hybrid thick-film RF modules. He is presently engaged in power hybrids development. He can be reached at 158 Summer St., Somerville, MA 02143, tel: (508) 534–5776.

Surface Mount Guide

The new *RF/IF Surface Mount Designer's Guide* is now available, free of charge, from Mini-Circuits. This 48-page guide features up-to-date and complete product and specification information about Mini-Circuits' lines of RF/IF components for surface mount technology. Products include monolithic amplifiers, fixed attenuators, attenuators, switches, bias-tees, directional couplers, lowpass filters, frequency doublers, frequency mixers, I&Q modulators and demodulators, power splitter/combiners, switches and RF transformers.

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Decoder Application Note

Qualcomm's VLSI Products Group announces the availability of "Peripheral Data Mode Operation of the Q0256/Q1650 Viterbi Decoder", a 24-page application note describing how to implement the peripheral data mode for forward error correction in applications like INMARSAT, VSATs and other low data rate terrestrial or satellite data links.

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Military Cross Reference

Analog Devices offers a free *Motorola-to-Analog Devices Military Product Cross Reference Guide* identifying more than 120 of Analog Devices' products that functionally replace those which Motorola has planned for obsolescence.

Analog Devices
INFO/CARD #156

Modulation Meter Data Sheet

A two-page data sheet on the Model 8211 FM/AM modulation meter has been released by Boonton Electronics. The data sheet lists all the specifications and enhancements to the Model 8211. Model 8211 provides high-performance, self-calibration capability, and true peak detection. All specifications are listed, including carrier frequency range, sensitivity, frequency modulation, deviation accuracy, and modulation bandwidth.

Boonton Electronics Corp.
INFO/CARD #155

Measurement Equipment

Tektronix has announced the availability of its 1995 Measurement Products Catalog. More than 80 new products are detailed. The 596-page, soft-cover catalog's full-color product section provides a synopsis of Tektronix' business focus and features a variety of new form-factor measurement solutions. The catalog includes Tektronix partnerships/alliance solutions offered by Advantest and Rohde & Schwarz.

Tektronix, Inc.
INFO/CARD #154

"Intelligent" Car Report

A new study on the "intelligent" automobile of the future has been released by Strategies Unlimited. The study forecasts a $4.9 billion market for in-vehicle navigation, radar, and wireless toll collection systems within the coming decade. The comprehensive findings of this study are available in a 189-page report. Both system and RF component requirements are reviewed. The report is available for $3,950.

Strategies Unlimited
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Design Software Brochure

Compact Software has a new eight-page, four-color brochure describing the company's PC/Windows-based Serenade 6.5 software for integrated RF, microwave and lightwave design. Serenade software includes schematic capture, linear or nonlinear frequency domain simulation, simulation libraries and optional physical design software tools.

Compact Software
INFO/CARD #152

Finite-Element Software on Alpha AXP

MacNeal-Schwendler has announced that several of its software product lines are available on Digital's AXP 64-bit workstations and servers. These products include MSC/NASTRAN v. 68, MSC/ARIES v. 6.1, and MSC/PATRAN release 1.4. Additional MSC product lines will become available on the AXP platform during 1995.

MacNeal-Schwendler Corp.
INFO/CARD #151

Mathematical Software

Waterloo Maple Software has announced the release of THEORIST® 2 for the Macintosh and Power Macintosh. The program, with algebra and graphing, combines the power of advanced mathematics with a simple yet powerful graphical user interface. The "Click and Solve" technology eliminates the need to learn a programming language. THEORIST 2 handles algebraic equations and expressions, and has two- and three-dimensional graphing. THEORIST 2 is designed to work with 2 MB RAM and 2 MB on a hard disk. US commercial list price is $299.00.

Waterloo Maple Software
INFO/CARD #150

Active Filter Design

Eagleware has released the program =A/FILTER= which designs active electronic filters. =A/FILTER= synthesizes a wide range of structures, including GIC transform, single feedback, multiple feedback, state variable (biquadratic), dual amplifier, voltage controlled voltage source, and low sensitivity types. The program is a member of Eagleware's System 32 family of synthesis and simulation programs for DOS, Windows 3.1 and Windows NT. Active filter synthesis and simulation packages range from $1195 to $2235.

Eagleware Corp.
INFO/CARD #149

Analog Simulation

Optimization System Associates is now shipping OSA90/hope™ v.3.0. OSA90/hope is CAD software for simulation, modeling and optimization of linear an nonlinear circuits. Version 3.0 features 3D graphics, Space Mapping™, and augmented libraries and optimizers. Space Mapping links "course" and "fine" models, such as empirical and electromagnetic models. The expanded library includes HEMT models. One-sided l1 and one-sided Huber optimization implementations are also included in version 3.0. OSA90/hope version 3.0 runs under X windows on HP, Sun, and DEC workstations. Its base price is $20,000.

Optimization Systems Associates Inc.
INFO/CARD #148

Analysis and Optimization of Oscillators for Low Phase Noise and Low Power Consumption

By Ulrich L. Rohde and Chao-Ren Chang
Compact Software, Inc.

This article presents investigations into the design, analysis and performance of practical voltage-controlled or tunable LC oscillators. Modeled versus measured performance of several common oscillator designs are compared, and the steps taken to obtain more accurate models for CAD analysis are noted.

Modern hand-held radios need low phase noise oscillators which show low power consumption. Traditional choices are either bipolar transistors or FETs. Typically, for frequencies up to 500 MHz, FETs have been considered more favorable, compared to bipolar transistors because of the low phase noise in frequencies above the higher f_{max} of bipolar transistors, which makes them more attractive. In this paper. We intend to evaluate the following topics:

I. AGC action via diode clamping

A. Performance of FET oscillators in Class A Operation, self-limiting [1]

B. Performance of Class A oscillators using a clamping diode between gate and source [2]

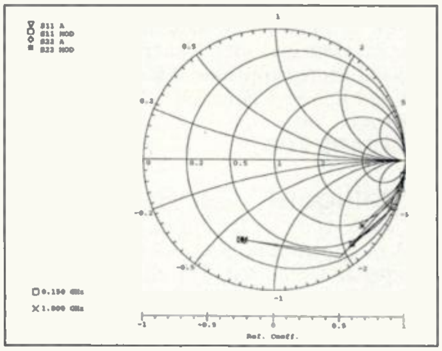

Figure 1. Comparison of measured and predicted S-parameters for the 2N4416, input and output.

C. Class A oscillator with source resistor [3]

II. Differential Oscillators using FETs and different bias points

The following are bipolar transistor oscillators:

A. 800 MHz Motorola VCO for hand-held radios [4]

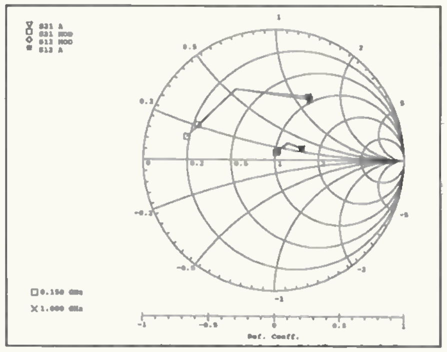

Figure 2. Comparison of measured and predicted S-parameters for the 2N4416 reflection coefficients.

B. Analysis of the Motorola MC 1648M voltage controlled oscillator IC [5]

C. BIP differential limiter VCOs from [6].

Considerable mystery has surrounded oscillator design and evaluating the above-mentioned circuits should help to shed some light in this area.

Another fascinating topic has been the modeling issue as a whole. For the

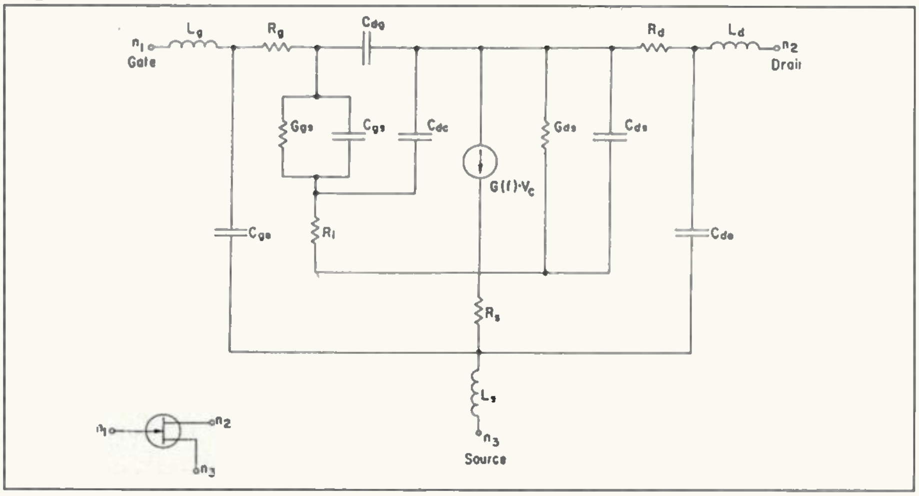

Figure 3. Equivalent circuit for the linear FET. The non-FET model follows the Spice approach by Statz and Curtice.

FET 1 2 0
G=?6.6042E-3?
CGS=?2.94488PF?
F=?4.5638GHZ?
T=?3.2199PS?
CDG=?.657116PF?
CGE=?.046388PF?
CGDE=?.117311PF?
CDE=?.186107PF?
CDS=?.573044FF?
RI=?4.619?
RG=?6.39222?
RD=?2.16345?
RS=?2.09905?
GDG=?11.2376E-6?
LG=?.022313NH?
LD=?3.57586NH?
LS=?2.1183NH?

Table 1. Equivalent linear parameters for the FET equivalent circuit of Figure 3.

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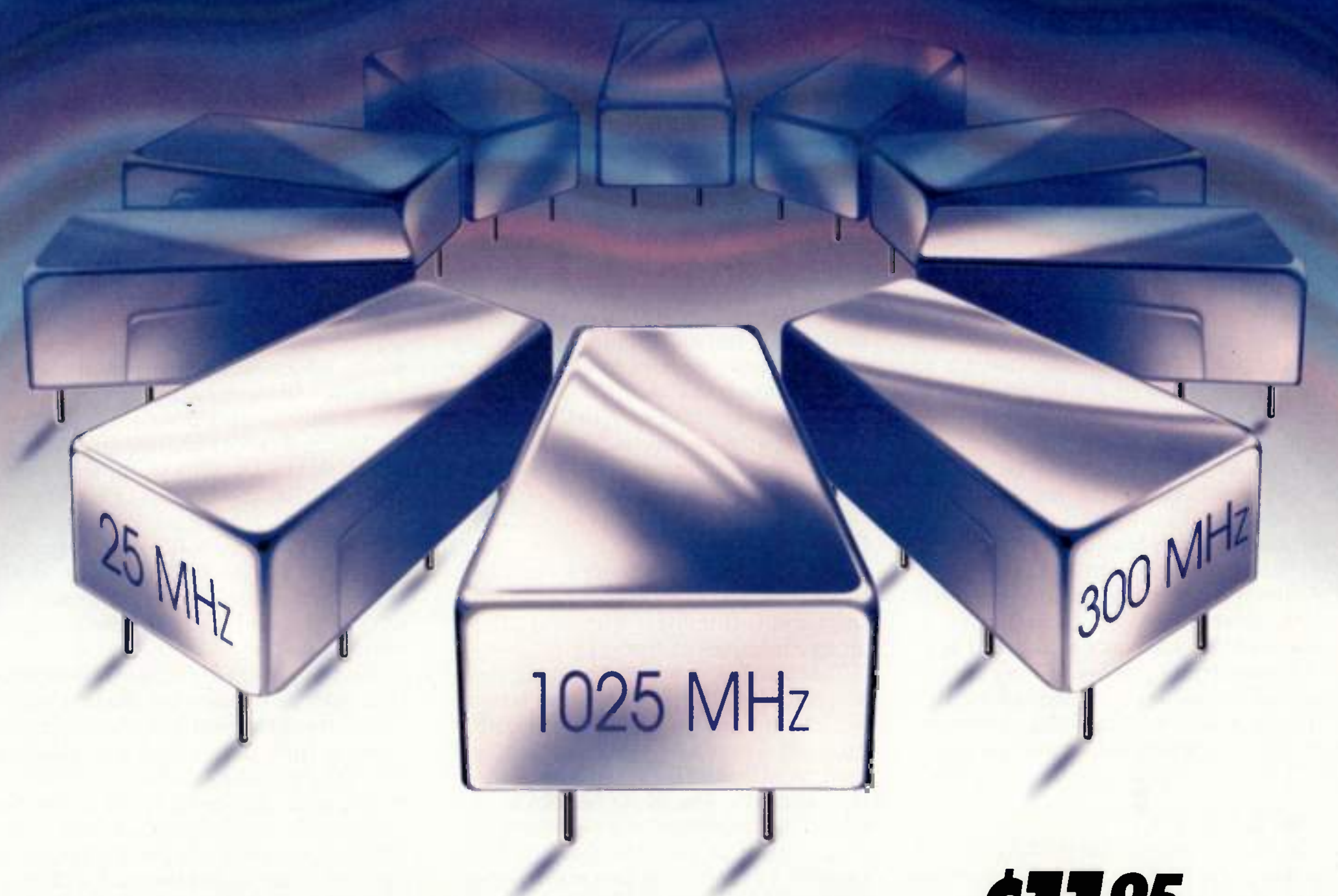

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It's a fact! With Mini-Circuits new POS family of shielded, laser sealed voltage controlled oscillators, you pay less but get more... top notch quality, superior performance and value pricing.

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DESIGNER'S KITS: K-POS1 $124.95 (contains 1ea. all models).
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Model No.	Freq. Range (MHz) Min.	Phase Noise (dBc/Hz) SSB @10kHz Typ.	Harmonics (dBc) Typ.	Power 12V DC Current mA	Price (Qty.5-49) $ ea.
POS-50	25-50	-110	-19	17	11.95
POS-75	37.5-75	-110	-27	17	11.95
POS-100	50-100	-107	-23	18	11.95
POS-150	75-150	-103	-23	18	11.95
POS-200	100-200	-102	-24	18	11.95
POS-300	150-280	-100	-30	18	13.95
POS-400	200-380	-98	-28	18	13.95
POS-535	300-525	-93	-26	18	13.95
POS-765	485-765	-85	-21	22	14.95
POS-1025	685-1025	-84	-23	22	16.95

Notes: Tuning voltage 1 to 16V required to cover freq. range.
Operating temperature range: - 55°C to +85°C.
Current typically 4.5 mA greater at +15V.

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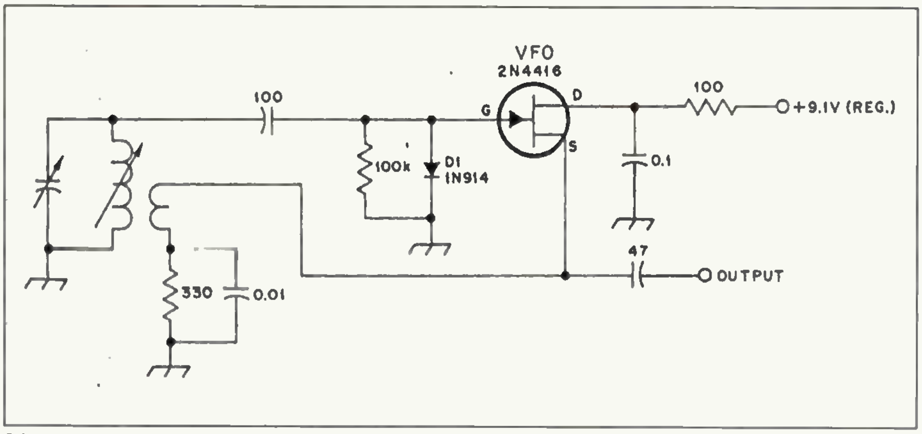

Figure 4. Oscillator test circuit.

purpose of obtaining data, Compact Software, Inc. has been doing its own parameter extraction and has always maintained that the parameters available from Spice libraries are not sufficiently accurate for high frequency applications. In particular, the use of these transistors in oscillators require reasonable assumptions for flicker noise and other related parameters. Also, one of the tests is to obtain agreement between the measured and simulated values for the same transistor at the same bias-point. Specifically, we mean that the S-parameters obtained from modeling and measured data must agree. Our experience in this area is based on a number of subcontracts. Among those are our activities with Siemens Semiconductors in Munich where CSI modeled their new 25 GHz f_t transistors, as well as GaAs FET lines, including new power FETs.

The FET Model

In our opinion, the best known N junction FET is the 2N4416 or members of its family. Table 1 shows the elements circuit file used for optimization. Figures 1 and 2 show the agreement of measured versus predicted S-parameters.

For the purpose of this paper, we took measured Y-parameters covering the frequency range up to 450 MHz and forced an optimization against the complex equivalent circuit for the linear FET.

We will not go into the equivalent procedure for nonlinear parameters but Compact Software's Scout program, using the modified Materka model or the Statz or Curtice models can generate an equivalent nonlinear circuit for junction FETs, which are perfect for this procedure.

FET Oscillator

The first FET oscillators looked fairly similar to those which came from the tube technology. Figure 4 shows an oscillator taken from [1].

This oscillator was both built, measured and modeled. Published reports in [7] and [8] caused considerable interest because no other work had ever mentioned the noise contribution of this diode. The basic oscillator is a modified Colpitts oscillator which has a clamping or rectification diode in parallel at the input. The noise performance and output power of this transistor oscillator depends highly on the nonlinear parameters of the transistor. The purpose of the diode is claimed to provide more uniform performance such as constant output power over tolerances of the transistor or temperature [7, 8]. The diode had been described as increasing the stability, but it only improves the thermal drift, not the SSB noise.

In an effort to better understand this circuit shown in [7], we have done an evaluation of the phase noise for three cases: no diode, a clamping diode and a RF-bypassed source resistor. Starting conditions are essentially the same transistor circuit shown in Figure 4; however, we initially eliminated the diode and the source resistor and then exchanged or added components. We used a specific nonlinear model, as implemented in the nonlinear program Microwave Harmonica with phase noise analysis capabilities, and added the parameters which described the transistor's noise performance.

In the previously mentioned publication, phase noise both with and without the diode was presented and compared to measured data [7]. The diode, which was connected in parallel between the gate and source, is activated only in an RF sense to just look at the noise contribution without affecting the DC bias point. One needs to know that the junction FET already has a diode from gate to source and therefore, the circuit effectively puts two diodes with different threshold levels in parallel. Figure 5 shows the output phase plane of the oscillator for both cases.

Next, we looked at the noise prediction with or without the diode. Considering the fact that the diode at this time is fully active and has changed the DC operating point , the output power will change by 8 dB. Since the phase noise is the ratio of two power levels, the absolute power output of the oscillator is irrelevant for close-in phase noise considerations. By looking at Figure 6, we get the same phase noise as reported in [7] and we also show the flicker noise contribution of the FET is the same when the diode is removed. With the diode, the output power is –6 dBm and without the

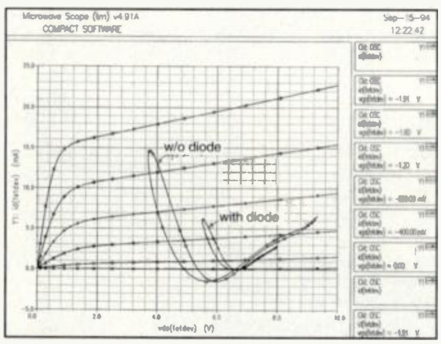

Figure 5. Output DC-IV curves with load line showing case with and without diode.

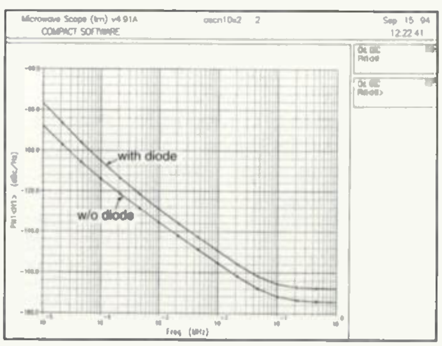

Figure 6. Predicted phase noise of the oscillator of Figure 4 with and without the diode.

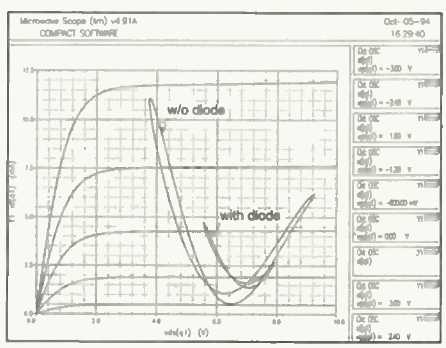

Figure 7. DC-IV output and load lines with and without the diode for a 2N3819 (library model).

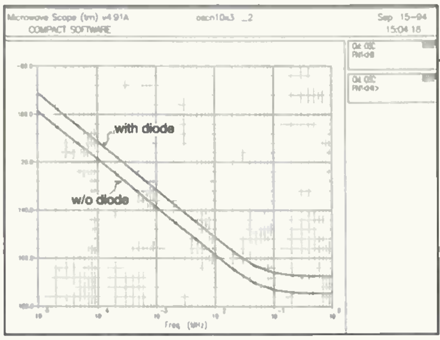

Figure 8. Predicted phase noise with and without the diode for the oscillator of Figure 4, using the 2N3819 library model. Note the absence of flicker noise.

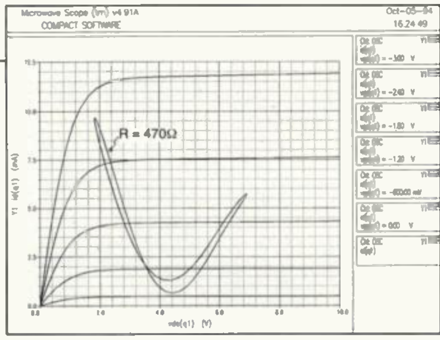

Figure 9. DC-IV output curve with load line for a 470 ohm source resistor instead of the diode.

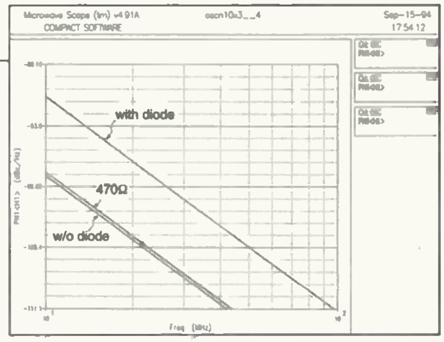

Figure 10. Predicted phase noise of the oscillator of Figure 4 without the diode, with the source resistor, and with the diode.

diode, the output power is +2 dBm.

The next step was to evaluate the quality of the Spice library models. Figure 7 shows the simulation of the load line using the parameters supplied by a third party Spice Library (PSPICE). Close examination of the DC/IV curve shows unrealistic behavior of the FET in the saturation mode. This type of early (simple) parameter extraction method is not sufficient for accurate high frequency applications. There is no provision for parasitic elements. Having exchanged the two transistor models for the same circuit, we re-calculated the noise and it again showed a significant difference (with and without the diode) consistent with the measured data previously reported, shown in Figure 8.

The output power for this oscillator was the same as the simulation with our model. This indicates that the calculation of the output power is less sensitive to the parameters which make the difference in the noise calculation. In the noise calculation, the two most relevant quantities are the loaded Q under operation, which the simulator predicts; and the noise contribution of the semiconductor. It is interesting to note that the Spice Library significantly underestimated the transistor noise contribution and provides unrealistic low close-in phase noise.

The next logical step was to evaluate the circuit without the diode, but with a resistor in the source. Figure 9 shows the output load line/DC-IV curves for the same circuit with 470 Ω resistor. The resulting output in this case is +1 dBm output which compares favorably with the case without the diode and, of course, the exciting question then arises, "What will the phase noise do?" Figure 10 shows all three cases: phase noise without the diode, AGC action with a 470 Ω source resistor, and with the diode. The phase noise with the diode still remains the worse case and the DC feedback is the best case. This is also consistent with reports in [3]. For further reducing the undesired "warm-up" effect of the FET at high bias point, one must go to large emitter resistors as 1 kΩ or much more and may even have to bias the gate positive.

Figure 11 shows the circuit diagram of the 85-119 MHz FET oscillator from the Rohde & Schwarz SMDU signal generator. This signal generator, which is no longer in production, had the same specifications as the HP 8640 and obtained its extremely good phase noise with the use of a helical resonator. It also adheres to the statement made above whereby the DC bias of the transistor was provided by a fairly large source resistor of 1.5 kΩ, and an adjustable resistor at the gate is the optimum bias condition. The large DC feedback compensates for both tolerance in the device and temperature effects. This is the correct method in operating the transistor, rather than using a AGC diode.

Other questions arise in the modeling area. How much does the actual Q of the inductor affect the performance in high-Q cases like the 10 MHz oscillator? Is it permissible to allow the Q to be simulated at Q = 8, whereby Q =

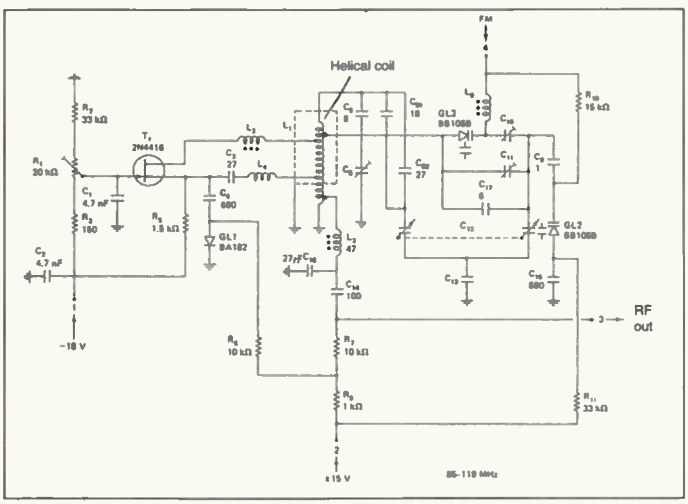

Figure 11. Circuit diagram of the 85-119 MHz FET oscillator used in the Rohde & Schwarz SMDU signal generator.

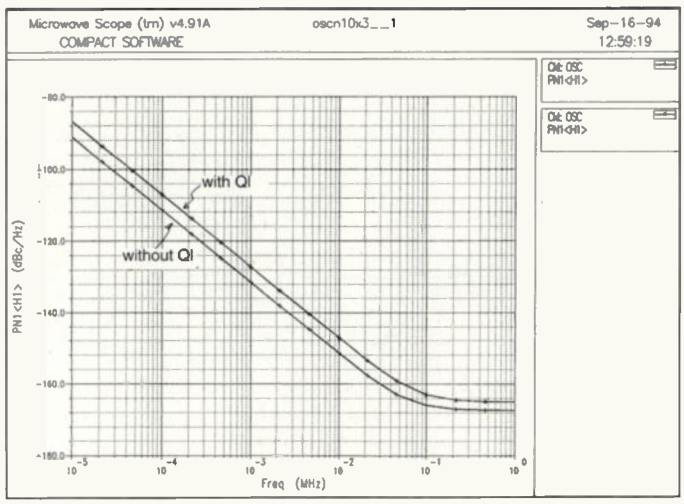

Figure 12. Predicted phase noise of the circuit of Figure 4 with loaded Q of 135 and Q = 500, which is the same as Q = ∞.

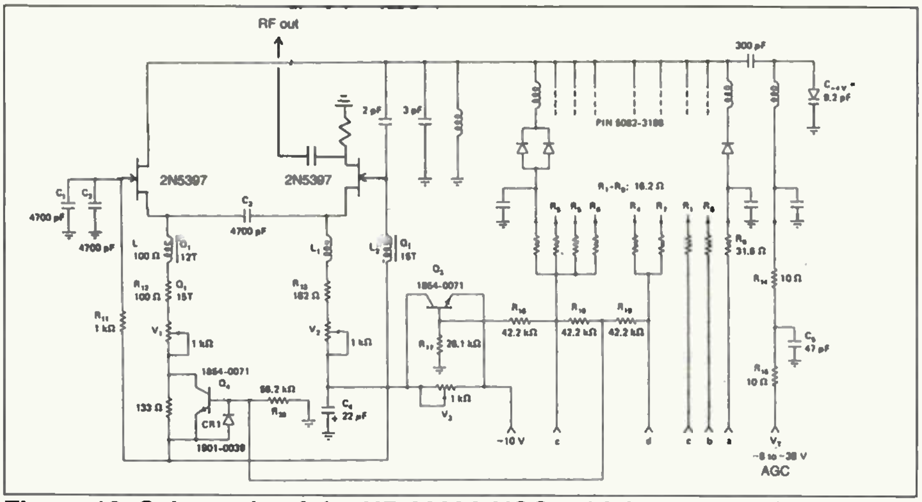

Figure 13. Schematic of the HP 8662A VCO, which operates from 260 to 520 MHz.

500, which is obtainable at these frequencies, is used as a point of reference. If we examine a more modest Q, like 135, we get approximately 4 dB deterioration of phase noise, which shows up both close-in and far out. For high frequency applications above 30 MHz, it is certainly important to properly model the Q of the tuned circuit prior to its being connected to the transistor. Figure 12 shows the noise for different loaded Qs.

Using integrated circuits for reasons of DC coupling, one has to resort to symmetrical circuit for a differential limiter type oscillator circuit as shown in [3]. These circuit's outputs are more distorted; what about phase noise performance? We took the same circuit and, using FETs, modeled a differential oscillator circuit. This circuit is consistent [3], but operating at 10 MHz and having a less elaborate DC circuitry (see Figure 13) which shows the differential FET oscillator which has become the heart of the Hewlett-Packard HP 8662 signal generator.

The initial starting condition had the common source resistor at 470 Ω, which produced the phase noise prediction of the first plot of Figure 14. By varying with the output load, the phase noise far out can be improved slightly. We finally decided to take more drastic steps and reduce the source resistor by a factor of 10. As can be seen, this changed the output phase noise significantly, with an improvement of approximately 12 dB across the board. The phase noise calculation is still somewhat optimistic because the FET model supplied is unrealistic in its flicker noise contribution, but it clearly shows the trend.

Figure 15 shows the output waveform as a function of DC bias, with the absolute output of the differential oscillator being –6 dBm.

The N junction FET model in SPICE does not consider any parasitics, which limits its practical high frequency use. Table 1 indicates that the first criteria for successful modeling has to do with a matching between predicted and measured S parameters.

Figure 16 shows the equivalent circuit for the packaged nonlinear FET. In order to model energy traps in the transistor, a series combination of R and C is identified as parasitic labeled CDSD and RDSD. By setting those two values to zero, the DC-IV curve shifts and now looks as shown in Figure 17. The fly-wheel effect to the right of the transistor has disappeared. This is the case with no diode connected and no emitter resistance. Figure 18 shows the time domain display of the device's drain current. It becomes obvious that this has a fairly narrow duty cycle and, therefore, a high harmonic content. For the sake of completeness, one should now look at a SPICE modeling of the same circuit. One should remember that the SPICE models are less complete for high frequencies. Also, the inherent dynamic range is less than that of our harmonic balance simulator. SPICE typically is lucky to get a 60 dB dynamic range. Our harmonic balance simulator has an 180 dB dynamic range. By inspecting Figure 19, it is apparent that there is no difference in resistance as the drain current is numerically unstable compared to the calculation in harmonic balance.

The big advantage SPICE offers is

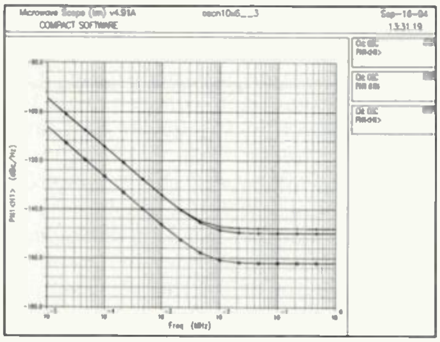

Figure 14. Predicted phase noise of an FET differential limiter oscillator for low current of 10 mA (lower curve).

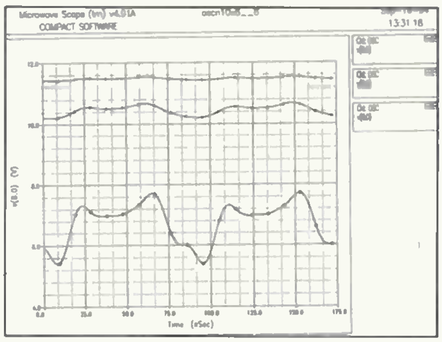

Figure 15. Output waveform of an FET differential oscillator for three different bias points.

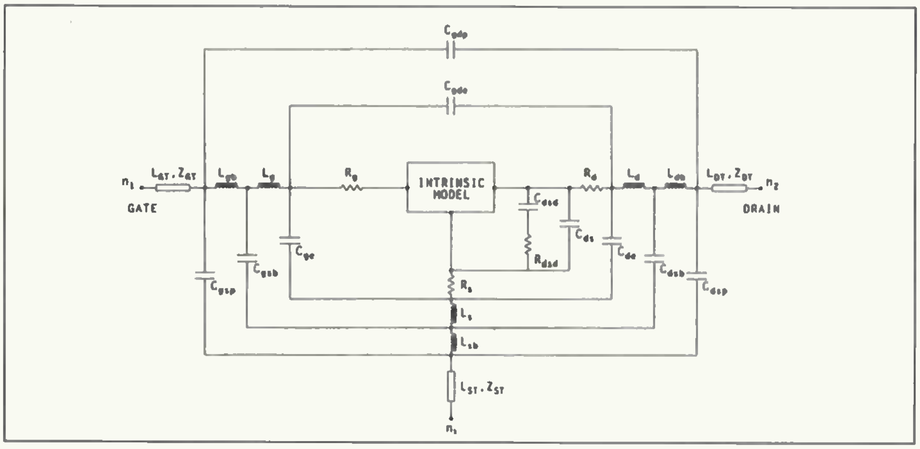

Figure 16. Extrinsic parasitics used to accurately model high frequency effects of FETs.

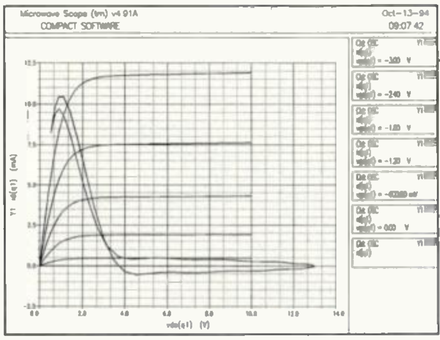

Figure 17. DC-IV curve of the FET oscillator with the external parasitic effects removed.

that it looks at the transient response and not only at the steady state as supplied by the harmonic balance simulation. Figure 20 shows the "start up" condition for the same oscillator. The "start up" condition begins at the lower left-hand comer with current and voltage being at the zero level. By the time the voltage goes up to 3.5 V and higher, the current begins to saturate, and then oscillation starts. This moves the DC bias point to the left and then automatically settles. This curve then becomes similar as the previously shown DC-IV curve (Figure 19). The final interesting thing is the time it takes the oscillation to actually start. This is done by inspecting the load voltage. The oscillation hesitates 10 μs, then the oscillation builds up to reach most of its amplitude after 25 μs, and 50 μs before it has stable amplitude. This is shown in Figure 21.

This ended our interest in FETs and we decided to look at the bipolar transistor circuits. Most hand-held radios for wireless applications will be built with silicon bipolar technology and the noise of those circuits will become an issue. The single transistor oscillator operating at 800 MHz already had been successfully modeled and was reported previously [4]. Figure 22 shows a plot of the predicted and measured phase noise as well as its schematic. Oscillators for highly integrated circuits make intensive use of differential types of transistor pairs. Figure 23 shows the standard two differential transistor pairs as offered by many suppliers. It has become the de facto standard for the pairs in either mixers or oscillators.

The first successful medium scale integrated circuit oscillator was built by Motorola (the MC1648M oscillator chip). A first inspection of the circuit indicates that there is no DC voltage difference between the chip and collector of the transistor on the left. Figure

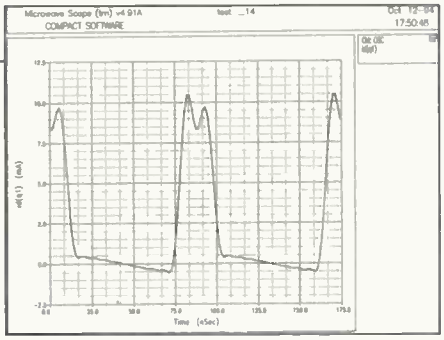

Figure 18. Drain current of the FET oscillator. Please note the short duty cycle.

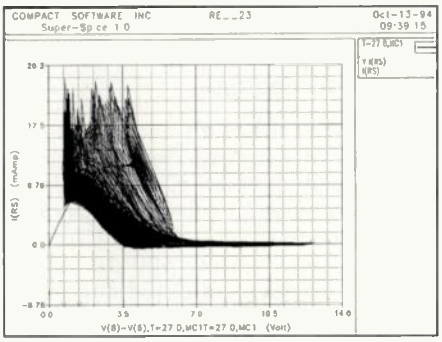

Figure 20. DC-IV curve starting condition of the FET oscillator using SPICE (see comments in text).

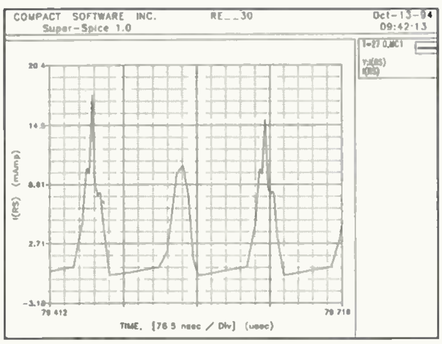

Figure 19. SPICE analysis of the FET oscillator showing the drain current. Note the numerical instability, inherent in SPICE.

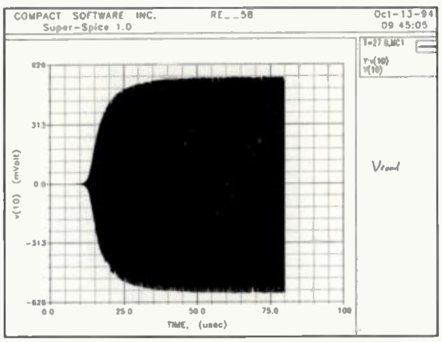

Figure 21. Transient accuracy of the FET oscillator. It takes about 50 μs before achieving a stable amplitude.

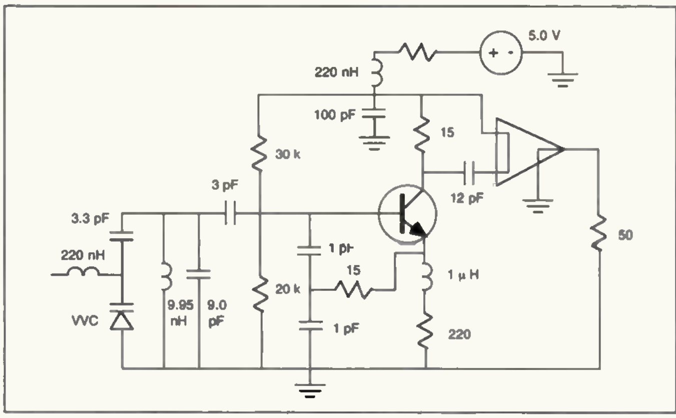

Figure 22(a). Schematic of the Motorola 800 MHz VCO.

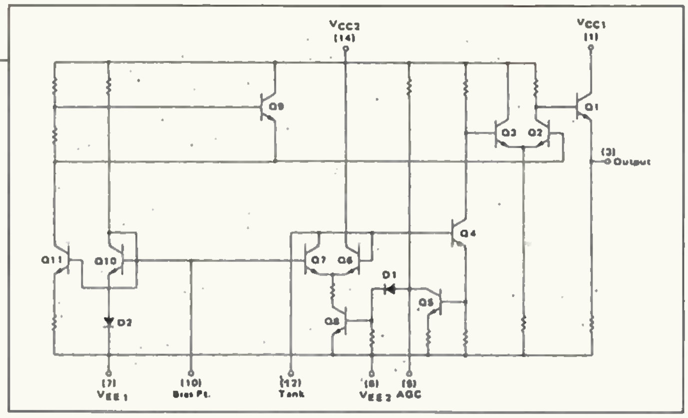

Figure 24. Schematic for the Motorola MC1648 oscillator IC.

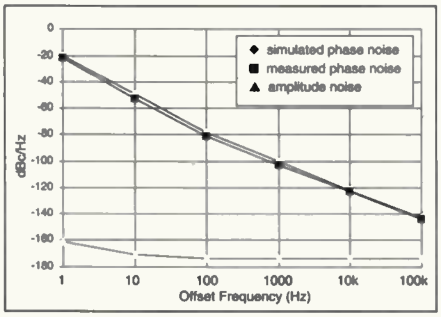

Figure 22(b). Comparison between predicted and measured phase noise. The resistor between emitter and capacitive feedback divider reduces the flicker noise contribution.

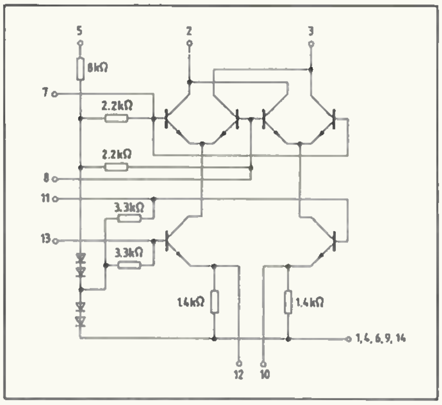

Figure 23. Pair of differential transistor amplifiers found in most modern IC designs.

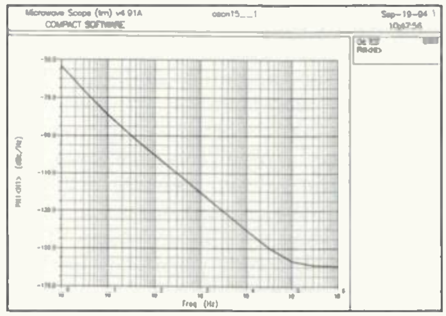
Figure 25. Simulated phase noise for the bipolar transistor oscillator at 10 MHz. By reducing collector currents to 500 μA, flicker noise is improved from 30 dB to 52 dB at 1 Hz.

24 shows the schematic of the internal oscillator as published by Motorola.

We were pleased to successfully model this circuit, but the resulting phase noise was quite surprising. Figure 25 shows the phase noise of the bipolar transistor IC. It is well understood that the phase noise of these oscillators is significantly higher than the phase noise of the FETs. It was quite surprising to find that the flicker noise contribution would worsen the phase noise to such a degree. We thereafter did an experiment wherein we changed the bias of the oscillator and thereby reduced the loop gain. The DC current was reduced from 3 mA to approximately 500 μA. Since the flicker noise is extremely bias dependent, the phase noise improved, but the output power dropped by about 10 dB. The simulation of this circuit also showed a non-sinusoidal waveform at the output, which, due to space constraints, is not being reproduced here. Further, a circuit which typically used for LSI application is shown in Figure 26 and is taken from [1]. These types of oscillators are actually being used for various applications and the interested user should use modern CAD tools to investigate their performance. Due to the heavy feedback in those oscillators, one will see in the simulation that they are not as sensitive to parameter variation of

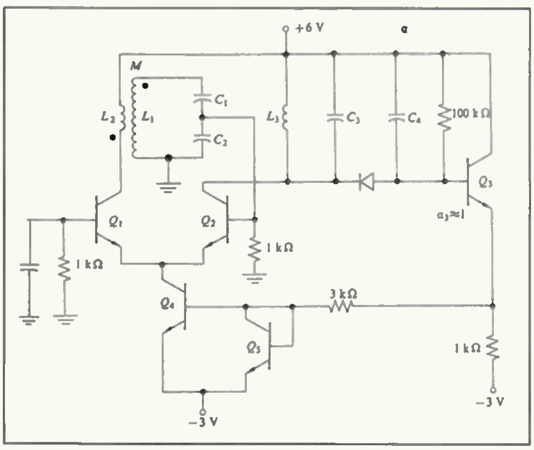

Figure 26. Oscillator using a differential amplifier and a DC biasing scheme.

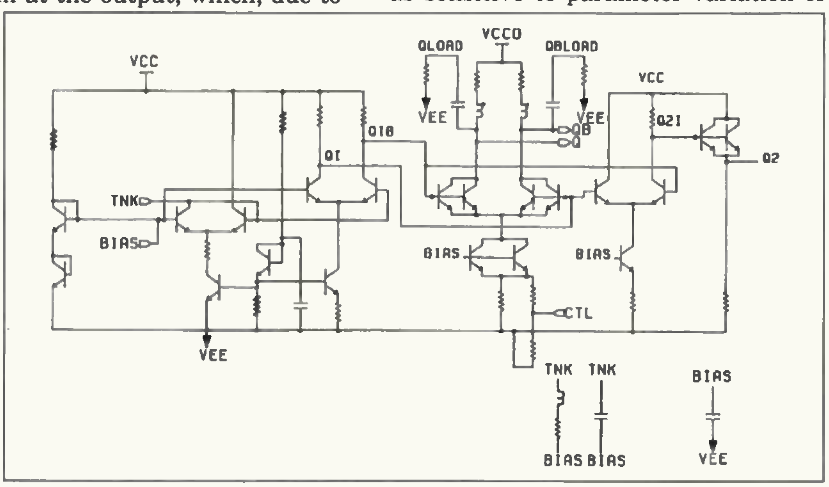

Figure 27. Schematic of the Motorola VCO.

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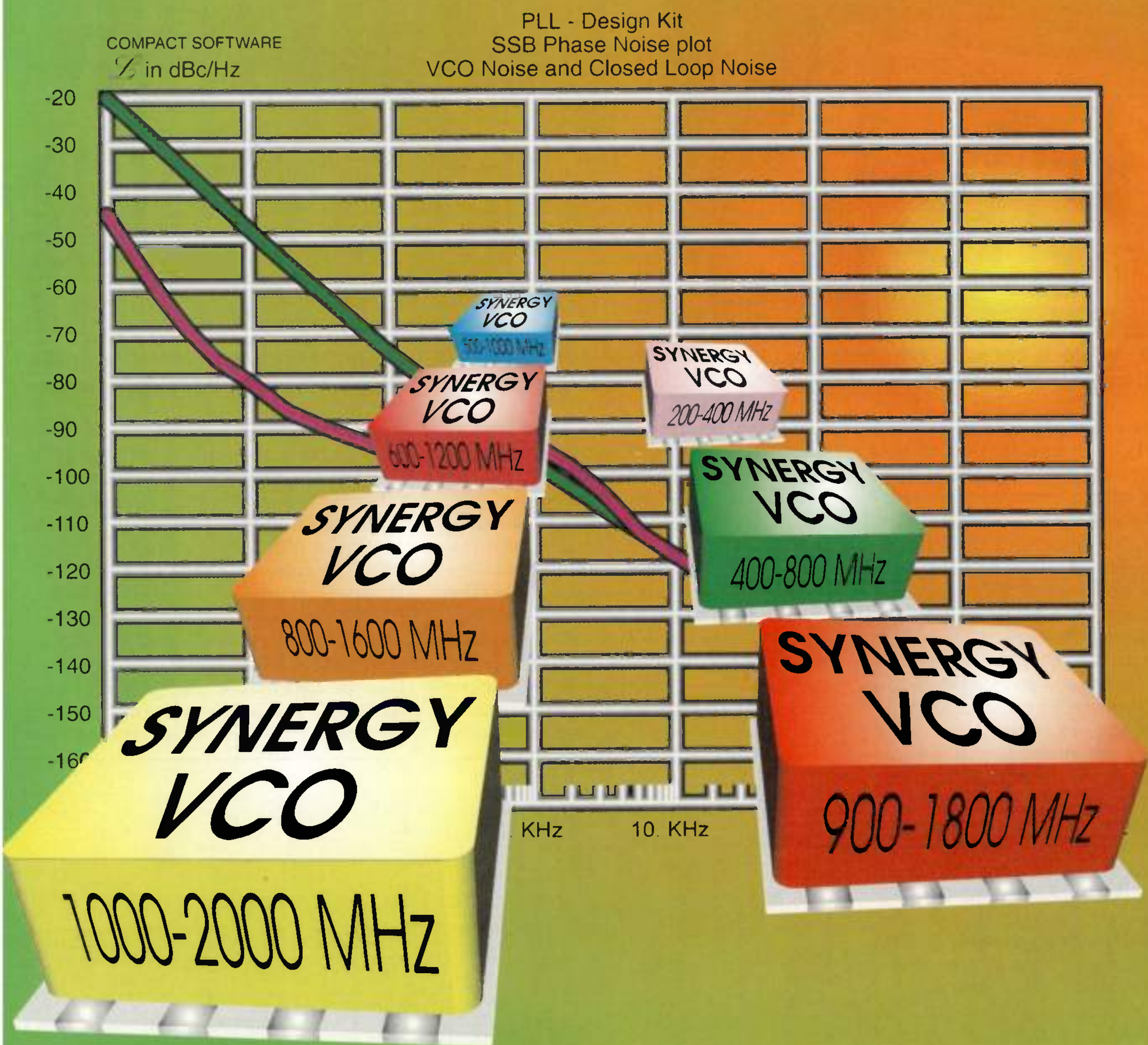

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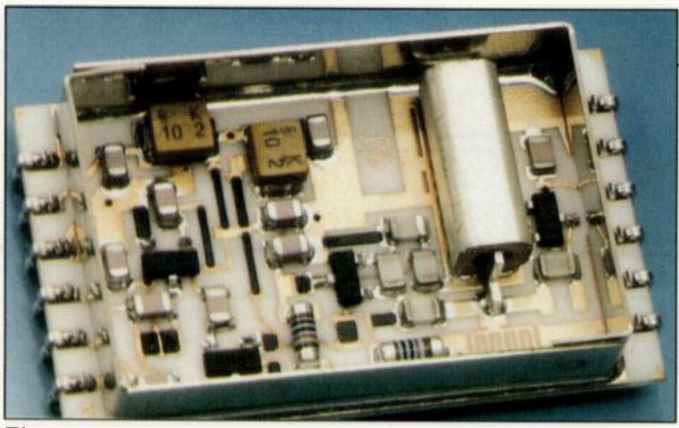

Figure 28. Modern ceramic resonator oscillator for hand-held cellular telephones.

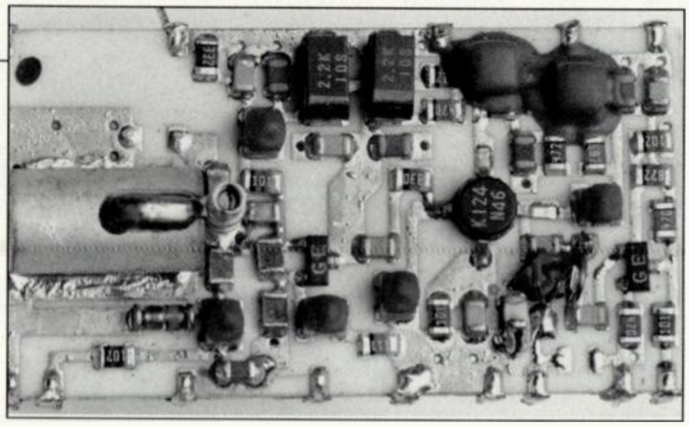

Figure 29. Test oscillator with conventional SMD design for 900 MHz hand-held cellular telephones.

the devices as might be expected. A particularly nice application of this type has been a Motorola VCO done entirely on silicon material. Figure 27 shows its schematic representation. The monolithic inductors used in the VCO were designed using Compact Software's Microwave Explorer program to model the effect of various ways of building those inductors. Measured and predicted Qs of the inductors show extremely good agreement. Figure 28 shows a discrete VCO. Considerable progress has been made here in high integration and first hand layout of these as shown in the older version of Figure 29.

As mentioned, these types of circuits will be found in large integrated circuits. We decided to model the circuit from [6], shown in Figure 30. The difference between this circuit and other previously discussed circuits is that the tuned circuit is coupled very loosely to the transistor. We also made a modification to the circuit whereby we took the feedback from the second transistor rather than the first one and put the second transistor in the loop. Figure 31 shows a proper presentation with modern ICs.

When comparing these two cases with the modified one, we found an improvement in the SSB phase noise of approximately 6 dB with the same output power. Figure 32 shows the predicted phase noise and Figure 33 shows the output wave-form available at the collector. According to our simulation analysis, both circuits only provide +4 dBm output. We were not able to confirm the +17 dBm reported by the authors.

Circuit Optimization

The oscillator circuit itself provides several degrees of freedom over which the designer has control. One of the three key issues to consider is the DC operating point of the transistor. While it varies the output power, it more drastically changes the flicker noise as has been shown several times. Secondly, the feedback circuit as such is responsible for the loop gain and the loaded Q. Thirdly, some negative rather than positive feedback like resistive feedback between the emitter of the bipolar transistor and the capacitive voltage divider is responsible for the phase noise reduction to the flicker noise contribution. The automated optimization procedure for phase noise optimization as implemented in Microwave Harmonica can drastically improve the circuit.

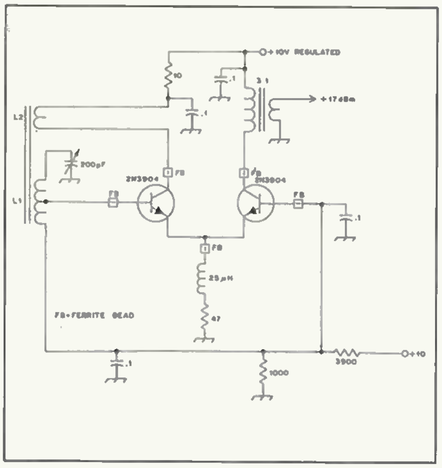

Figure 30 Circuit of the K7HFD low-noise oscillator from [6].

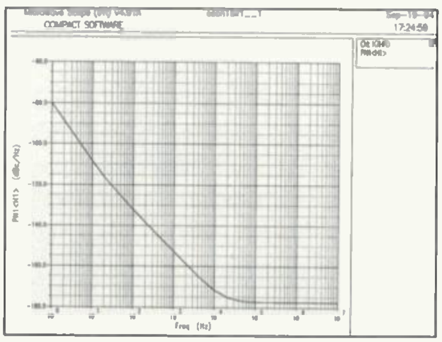

Figure 32. Simulated phase noise of the oscillator of Figure 31.

Figure 34 shows the phase noise – before and after optimization – of the Motorola 800 MHz oscillator (Figure 16). The close-in phase noise at 1 Hz has been improved by approximately 32 dB and even the phase noise from approximately 1-10 MHz off the carrier was improved. Because of the reduction in amplitude and the resistive feedback, the phase noise at 20 MHz and further away is set at 160 dBc per

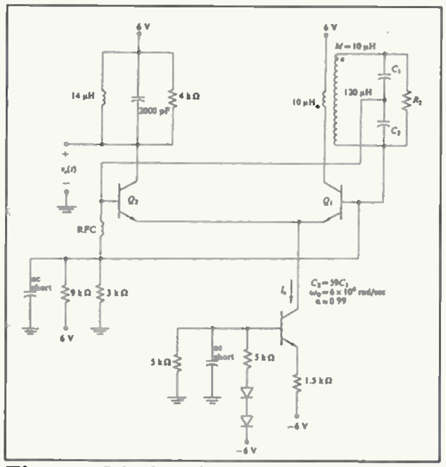

Figure 31. Implementation of the oscillator of Figure 30 using an IC.

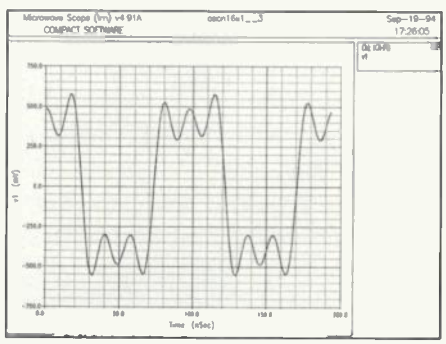

Figure 33. Output waveform of the oscillator of Figure 31.

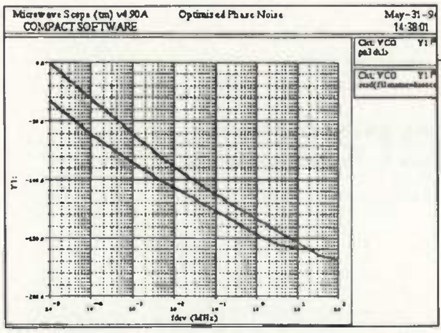

Figure 34. Phase noise of the oscillator of Figure 16 before and after optimization.

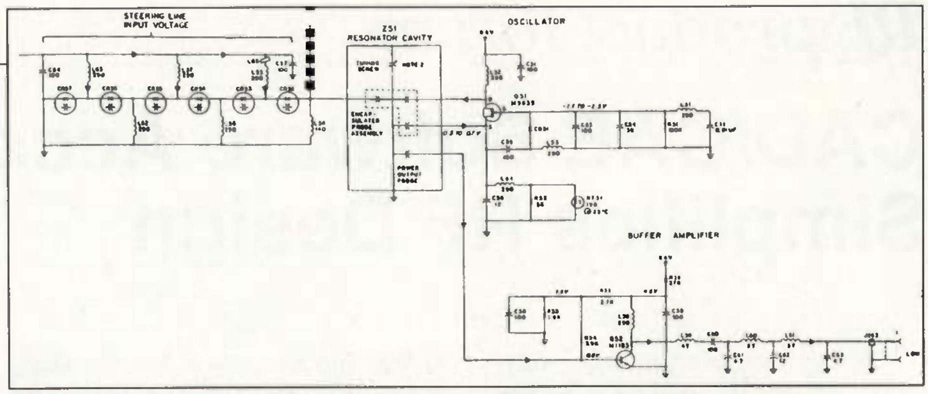

Figure 37. Cavity stabilized VCO for application in UHF repeaters.

Hz while the original circuit predicted slightly better performance. This technique is applicable for arbitrary oscillator topologies and active devices for which we have a good understanding of the relationship between the flicker noise and the bias point.

Other Useful Oscillator Circuits

Figure 35 shows a bipolar transistor oscillator in the 2.75 to 3.75 MHz frequency range. By adjusting L1 and C1, the frequency range can be shifted. A good application is 5 – 5.5 MHz. Inside its oven controlled environment, this bipolar oscillator has a frequency shift of less than 50 Hz per day. Even without the electric heating, it will not drive more than 200 Hz per day after 15 minutes warm up.

For those attempting to build low phase noise VCOs for the 220 MHz amateur band, as shown in Figure 36, here is a unique low phase noise oscillator which should be made part of the synthesizer. Its phase noise at 200 MHz is measured at 150 dB per Hz, 25 kHz off the carrier.

Finally, for those having "fun" with noise in their UHF repeater oscillators, Figure 37 shows a cavity-resonator based low phase noise VCO which may eliminate headaches caused by some noisy designs.

Summary

This paper has analyzed both FET and bipolar oscillators, including circuit combinations which are adapted for high level integration. The general trend seems to be that FETs require more current, but the f_t of the available N junction FETs is not sufficient to build oscillators above 400 MHz, and the flicker noise contribution is very small. On the other hand, bipolar transistors are ideally suited for oscillator circuits for high integration. Their bias point has to be chosen very carefully to avoid too much flicker noise contribution. When these types of oscillators are designed as one terminal oscillator circuit, they do not provide an attractive choice of topologies and a good compromise between performance and complexity has to be sought. Most hand-held two-way radios or radio telephones with digital signal processing do not require very high performance oscillators. This paper has sought to provide some guidelines for obtaining high performance circuits. We have also addressed AGC action by self-limiting the use of DC feedback and diode clamping. There has been some discussion on these three topics within the engineering community and we have used our tools to provide insight on these sensitive circuits. *RF*

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7. Ulrich L. Rohde, "All About Phase Noise in Oscillators", *QEX*, Dec. 1993, Jan. 1994, Feb. 1994.
8. Ulrich L. Rohde, "Key Components of Modem Receiver Design, Parts I, II and III," *QST*, May 1994, June 1994, July 1994, pp. 29-32, pp. 27-30 and pp. 43, 45. respectively.

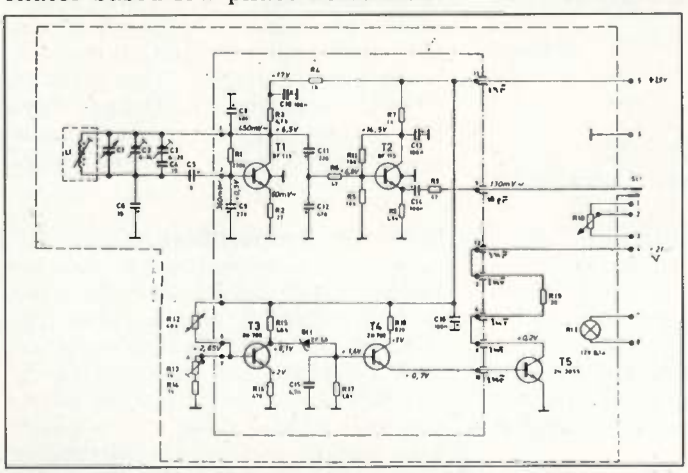


Figure 35. 2.75-3.75 MHz high stability oscillator with proportional controlled oven.

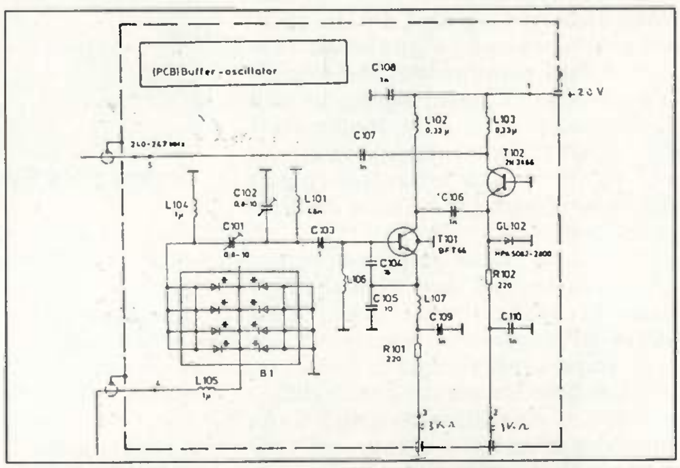


Figure 36. Low phase noise 200 MHz-range VCO from the Rohde & Schwarz XPC synthesizer.

CAD/CAE Software Accelerates, Simplifies RF Design

This month the product forum highlights the market changes affecting CAD/CAE software. Several manufacturers offer opinions regarding today's marketplace and trends in the industry.

Analog and RF Models

Although the rate has slowed in the last year, the market is still growing. This growth is fed by the lower cost of fast PCs and affordable Spice simulation software for those PCs. Key applications are RF and high-speed, where bench work is often limited by probe loading or probe bandwidth. The outlook for the next six months to two years is somewhere between stable and 10 percent growth.

On the business end, shorter development cycles can only happen when simulation and prototyping are combined. On the technology end, the move toward semi-custom array ASICs and MMICs for many high-speed applications with their high NREs force simulation to be a large part of the picture. We are participating by providing Spice compatible models of RF transistors, RF PIN diodes, laser diodes, and other active high-speed devices.

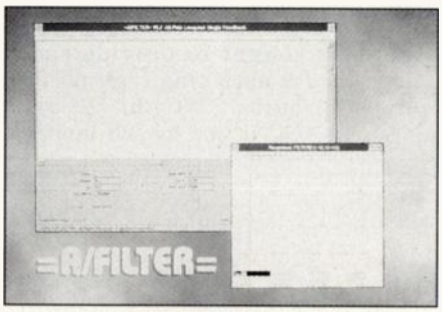
Compact Software

Efficiency and size are key requirements for RF power amplifiers used in cellular radios, two-way radio systems, and DoD applications. New applications are creating new design problems. For example, digital transmission of data requires that the amplifier's response to pulsed signals be well understood. Compact Software's CAE/CAD tools let designers create and optimize new amplifiers using simulation instead of building multiple prototypes.

In order to create amplifier designs which can be built and work the first time, accurate simulation of high power RF/microwave transistors are very important. Compact Software, Inc. has been cooperating with different device manufacturers and GaAs foundries in order to create and validate simulation models for high power devices. Participating vendors include Philips and Siemens in Europe, along with Motorola and several GaAs foundries in the United States. Compact has modeled all of Siemen's new bipolar transistors and GaAs FETs for cellular telephone base stations. In a separate effort sponsored by Motorola, Compact Software has developed a mathematically continuous model for power MOS transistors. This has resulted in a new nonlinear model for both time domain and frequency domain simulation that is much faster than existing models and provides better convergence. Compact's expertise in this area has also resulted in SBIR contracts from ARPA, specifically in Class E operation for power amplifiers. This work allows Compact to provide solutions for commercial customers.

Eagleware Corporation

Desktop access to powerful computers has revolutionized the way engineers design. Regardless of the end application or market, software requirements are universal: accuracy, flexibility, speed, ease-of-use and high-quality support.



Eagleware's System 32 screens.

Actual use is the ultimate test of software quality and value. At Eagleware, we design, build, and test circuits daily in a fully equipped lab. We have as much capital invested in instrumentation as in computers! This enhances both the practical capabilities of Eagleware tools and the quality of our support.

Most CAE developers focus on simulation and manufacturing tools. Eagleware's System 32 also integrates synthesis into the design process. Eagleware System 32 synthesis programs help design L-C filters, oscillators, microwave filters, matching networks, active filters, and amplifiers. With System 32 programs you choose the operating system: DOS, Windows or Windows NT. Windows programs are 32 bit for fast operation and full access to system memory. All System 32 programs share the same user files for mixed department use and painless upgrades.

Nedrud Data Systems

The CAD/CAE software market will continue to grow as programs become more powerful, hardware becomes more affordable, and as engineers become more aware of the benefits of CAD tools: increased productivity through faster virtual prototyping. The time saved is often spent performing more calculations, which means fewer prototypes and smaller, lighter designs. The truth is that engineers can no longer afford to design without software.

Now that more companies offer competent software, engineers are beginning to evaluate programs less on features and more on how well they perform their functions. The length of a design task depends much less on the computer's clock frequency that it does on the program's user interface; e.g. how quickly the user can tell the computer what to display. This gives an edge to programs like DragonWave which allow engineers to create and analyze circuits in a straightforward and intuitive way.

MicroSim Corporation

As frequencies continue to increase for both analog and digital applications, the modeling of transmission line effects becomes more and more important. Therefore, CAE tools which properly simulate transmission line effects and tools which extract the necessary data from physical layout information are increasing in popularity.

MicroSim™ PSpice for Windows supports single and multiple transmission line effects for simulating reflections, loss, dispersion, and crosstalk in both the time and frequency domains. MicroSim's new signal integrity product – MicroSim Polaris for Windows – extracts transmission line data from a physical board layout. Whether the parameter extraction is accomplished using MicroSim Polaris, or the parameters are calculated manually, the ability to include the transmission line effects in a simulation allows the proper examination of issues such as amplifier stability, data transmission, and ground bounce. The types of transmission lines which can be modeled include coaxial cables, twisted pairs, PCB traces, striplines, microstrips, and the IBIS models.

Viewlogic Systems, Inc.

Most RF equipment manufacturers now use electronic design automation software to design the printed circuit boards, modules, and ASIC/custom integrated circuits used in their products. Increasingly that includes sophisticated digital technology as well as RF circuits. The digital part of such products may be extremely complex, as seen in frequency synthesizers. With both design complexity and digital content growing, many RF manufacturers find themselves using design methods similar to those used for such products as FAX/modems, multimedia video/sound, and disk drive interfaces. All must optimize critical analog circuitry while accounting for the effects of digital circuits with thousands of logic gates. Consequently, co-simulation of digital and analog circuits is becoming mandatory to verify total system performance before committing to manufacture. To simplify digital design, the use of VHDL and Verilog Hardware Design Languages is becoming popular. A typical design may therefore use several different simulators, both digital and analog, to promote optimum system design. These "high-level design" approaches also support mixed analog/digital design reuse. Viewlogic Systems provides a suite of EDA tools and graphical design environments that make it possible for customers to build a "best in class" design ensemble. The underlying design environment, available on PCs and Unix workstations, allows the use of both Viewlogic and third-party tools through standard industry interfaces. This enables Viewlogic to provide major RF/wireless manufacturers with advanced EDA software that is targeted to their needs. It will soon include Hewlett-Packard's HP-EEs line of RF/wireless simulators.

Design Automation, Inc.

The explosive growth in portable wireless communications is focusing attention on transmitter power efficiency (increased efficiency, reduced battery drain, reducing battery size and weight and/or lengthening time between battery rechargings). Switching mode RF power amplifiers (Classes E and F) are finding favor because of their increased efficiency and reduced battery drain. Their drawback has been difficulty in optimizing, because of complicated nonlinear and implicit mathematical relationships between the circuit parameters and the resulting efficiency and output power. That drawback has been removed with the availability of fast-computing software developed specifically for designing such amplifiers, able to automatically optimize the design in about a minute; or to simulate the circuit and compute the voltage and current waveforms and the RF output power, DC input power, and power dissipations in each circuit component in 0.14 second (times are for a 486/33 computer). Outlook: More use of Class E and F amplifiers. *RF*

For information on CAD/CAE software, circle the following Info/Card numbers:

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Due to space limitations, our January Product Forum did not include a complete list of companies which manufacuter TCXOs. Below is an expanded listing of these companies. For information, circle the Info/Card numbers below:

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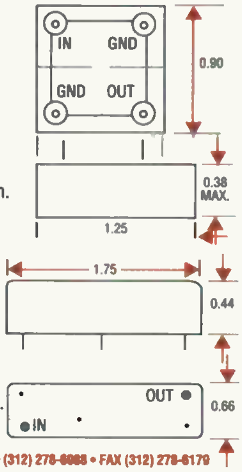

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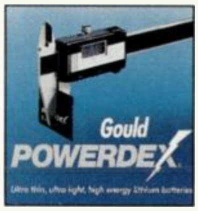

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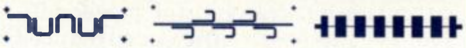

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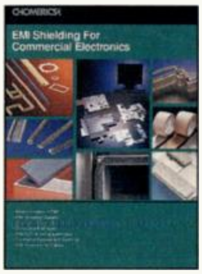

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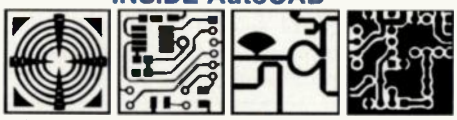

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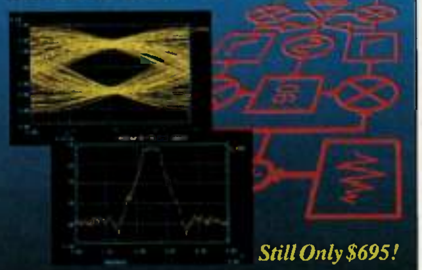

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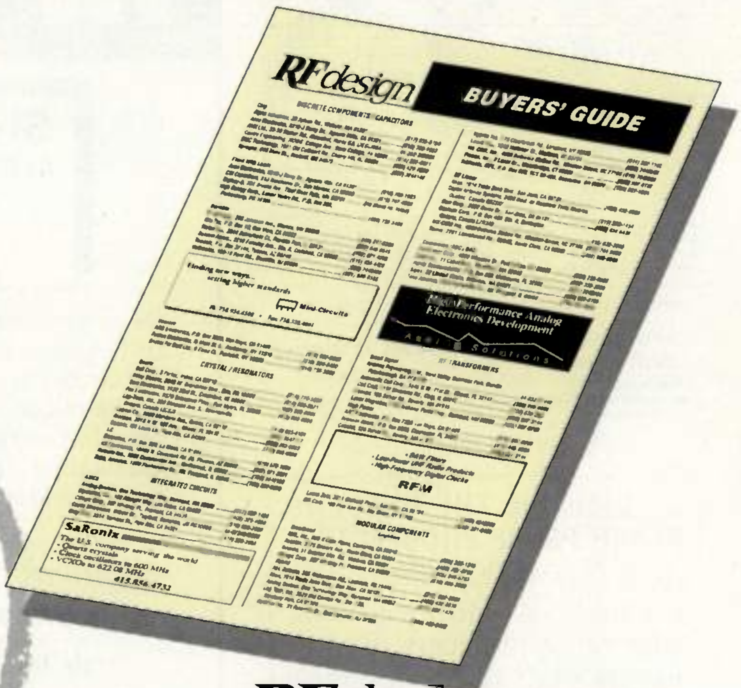

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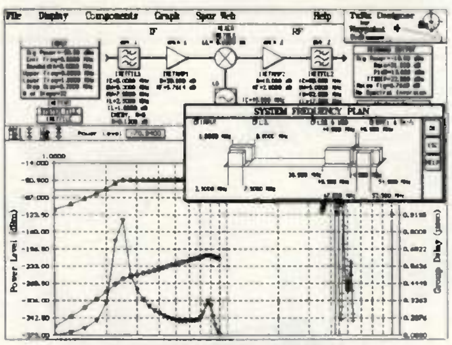

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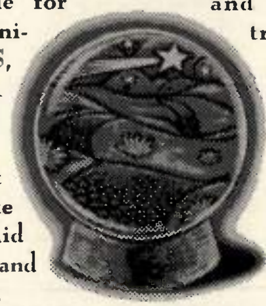

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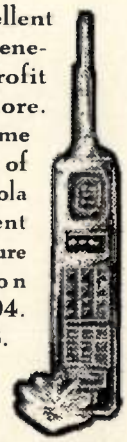

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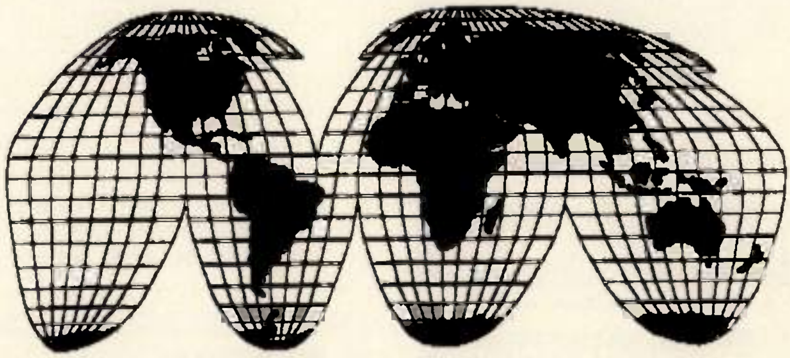

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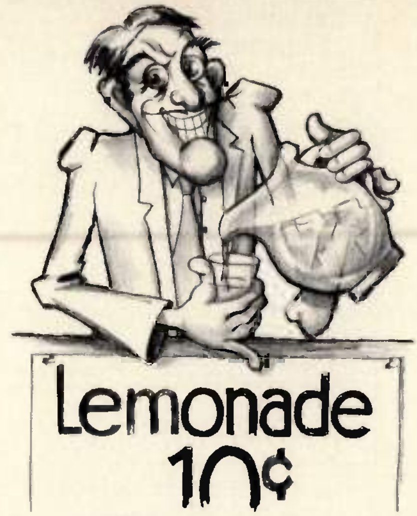

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EMC/EMI TESTING

Liberty BEL EMC SVCS., P.O. Box 5431, MS20, Compton, CA 90224 .... (310) 537-4235
Above MS20 represents RF Design

EQUIPMENT CALIBRATION

Instrument Repair Labs, Inc., 2100 W. 6th Ave., Broomfield, CO 80020 .... (800) 345-6140

INSTRUMENT RENTAL/LEASING

Naptech, 12312 Hwy 175, Cobb, CA 95423 .... (800) 336-7723 Fax (707) 928-1963

OEM PRODUCTS

Locus, Inc., 1842 Hoffman St., Madison WI 53704 .... (608) 244-0500

TELEMETRY SYSTEMS & ENGINEERING

Telemetry Technologies, Inc., 3307 West St., Rosenberg, TX 77471 .... (713) 344-9000

DIGITAL SIGNAL PROCESSING

SYSTEMS DEVELOPMENT REAL TIME

Adaptive Technology, Inc., 309 Curtis St Syracuse NY 13208 (800) 724-6968 Fax (315) 472-3279

EMC PRODUCTS

SURGE & TRANSIENT PROTECTORS

Polyphaser Corp., P.O. Box 9000, Minden, NV 89423 .... (800) 325-7170, (702) 782-2511

RF design INTRODUCES THE YELLOW PAGE DIRECTORY SECTION
CALL
1-800-443-4969
FOR DETAILS ON AN EXCITING INTRODUCTORY OFFER!

RF advertising index

RF guide to editorial coverage

LCF ENTERPRISES RF POWER AMPLIFIERS

(Standard or Custom)

1 MHz - 2 GHz • 1 W - 1 kW

Part No.	Output Power	Gain	MODULES DC Supply	PKG
2 MHz - 30 MHz				
30-2-5-35	5W	35dB	24/28V	E3
30-2-50-35	50W	35dB	24/28V	E3
30-2-100-35	100W	35dB	28V	E3
30-2-200-35	200W	35dB	28V	E3
30 MHz - 100 MHz				
100-30-5-35	5W	35dB	24/28V	E3
100-30-100-35	100W	35dB	28V	E3
50 MHz - 150 MHz				
150-50-2-40	2W	40dB	12/15V	C3
150-50-100-35	100W	35dB	28V	A3
150 MHz - 200 MHz				
200-150-100-10	100W	10dB	28V	A1
200-150-100-18	100W	18dB	28V	A2
200-150-100-35	100W	35dB	28V	A3
200-150-200-35	200W	35dB	28V	A3
50 MHz - 250 MHz				
250-50-200-10	160W	10dB	28V	A1
250-50-200-18	160W	18dB	28V	A2
250-50-200-35	160W	35dB	28V	A3
50 MHz - 400 MHz				
400-50-1-30	1W	30dB	12/15V	C3
225 MHz - 400 MHz				
400-225-1-35	1W	35dB	12/15V	C3
400-225-10-35	10W	35dB	24/28V	A3
400-225-30-10	30W	10dB	24/28V	A1
400-225-30-18	30W	18dB	24/28V	A2
400-225-30-35	30W	35dB	24/28V	A3, B3
400-225-50-10	50W	10dB	24/28V	A1
400-225-50-18	50W	18dB	24/28V	A2
400-225-50-35	50W	35dB	24/28V	A3
400-225-100-10	100W	10dB	28V	A1
400-225-100-18	100W	18dB	28V	A2
400-225-100-35	100W	35dB	28V	A3
225 MHz - 600 MHz				
600-225-30-10	30W	10dB	24/28V	A1
600-225-30-18	30W	18dB	24/28V	A2
600-225-30-30	30W	30dB	24/28V	A3, B3
400 MHz - 600 MHz				
600-400-30-10	30W	10dB	24/28V	A1
600-400-30-18	30W	18dB	24/28V	A2
600-400-30-30	30W	30dB	24/28V	A3, B3
925 MHz				
925-1-8	1W	8dB	12/15V	D1
100 MHz - 500 MHz				
500-100-5-30	5W	30dB	24/28V	A3
500-100-10-30	10W	30dB	24/28V	A3
500-100-100-30	100W	30dB	28V	A3
500 MHz - 1000 MHz				
1000-500-10-8	10W	8dB	24/28V	A1
1000-500-10-16	10W	16dB	24/28V	A2
1000-500-10-30	10W	30dB	24/28V	A3
10 MHz - 1200 MHz				
1200-10-10-30	10W	30dB	24/28V	A3

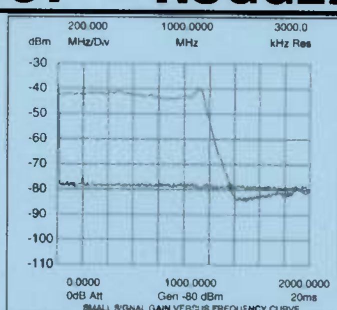
* Test Fixture (Option "B") includes heat sink, fan, thermal shutdown, and electrical fuse protection
** Rack-mount amplifiers: 120 Vac - 60 Hz / 240 Vac - 50 Hz

MODULES - SMALL SIZE • HIGH EFFICIENCY

- 200 watts
- 150 - 200 MHz
- 55% overall efficiency
- 35 dB gain
- 4.84" x 2.0" x 1.0"

- 100 watts
- 225 - 400 MHz
- 45% overall efficiency
- 10 dB gain
- 3.0" x 2.0" x 1.0"

SYSTEMS - LOW COST • RUGGED PROTECTION



- 10 MHz - 1200 MHz
- 10 watts CW
- 40 dB
- 19" x 7" x 18"
- 35 lbs maximum

LCF ENTERPRISES • 651 Via Alondra, # 712 • Camarillo 93012 USA • Phone: 805-388-8454 • FAX: 805-389-5393