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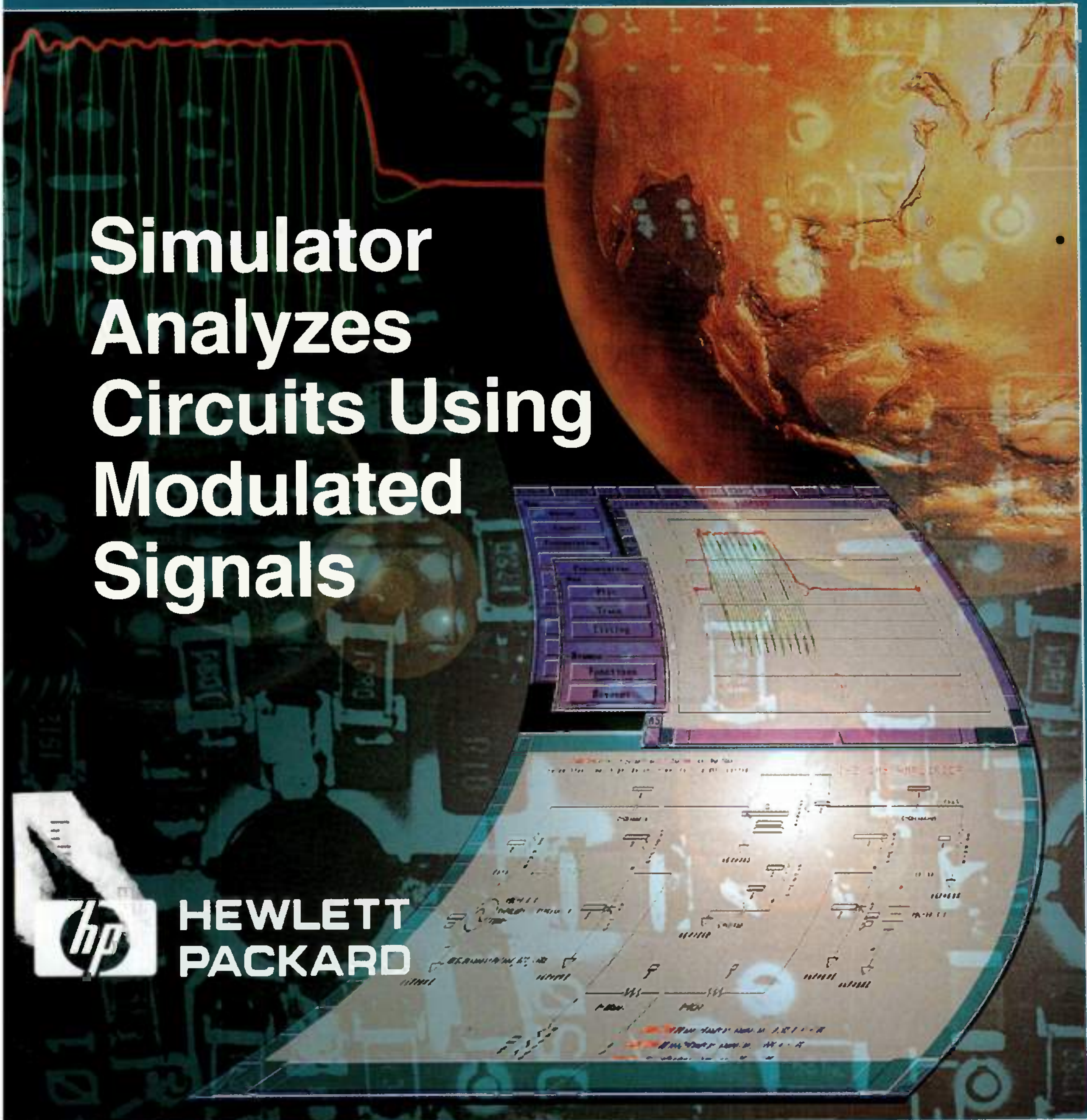
RF design™

engineering principles and practices

September 1995

Simulator Analyzes Circuits Using Modulated Signals



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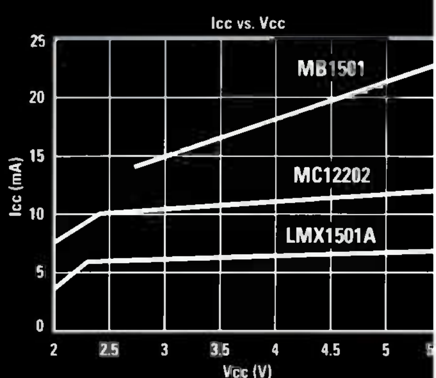
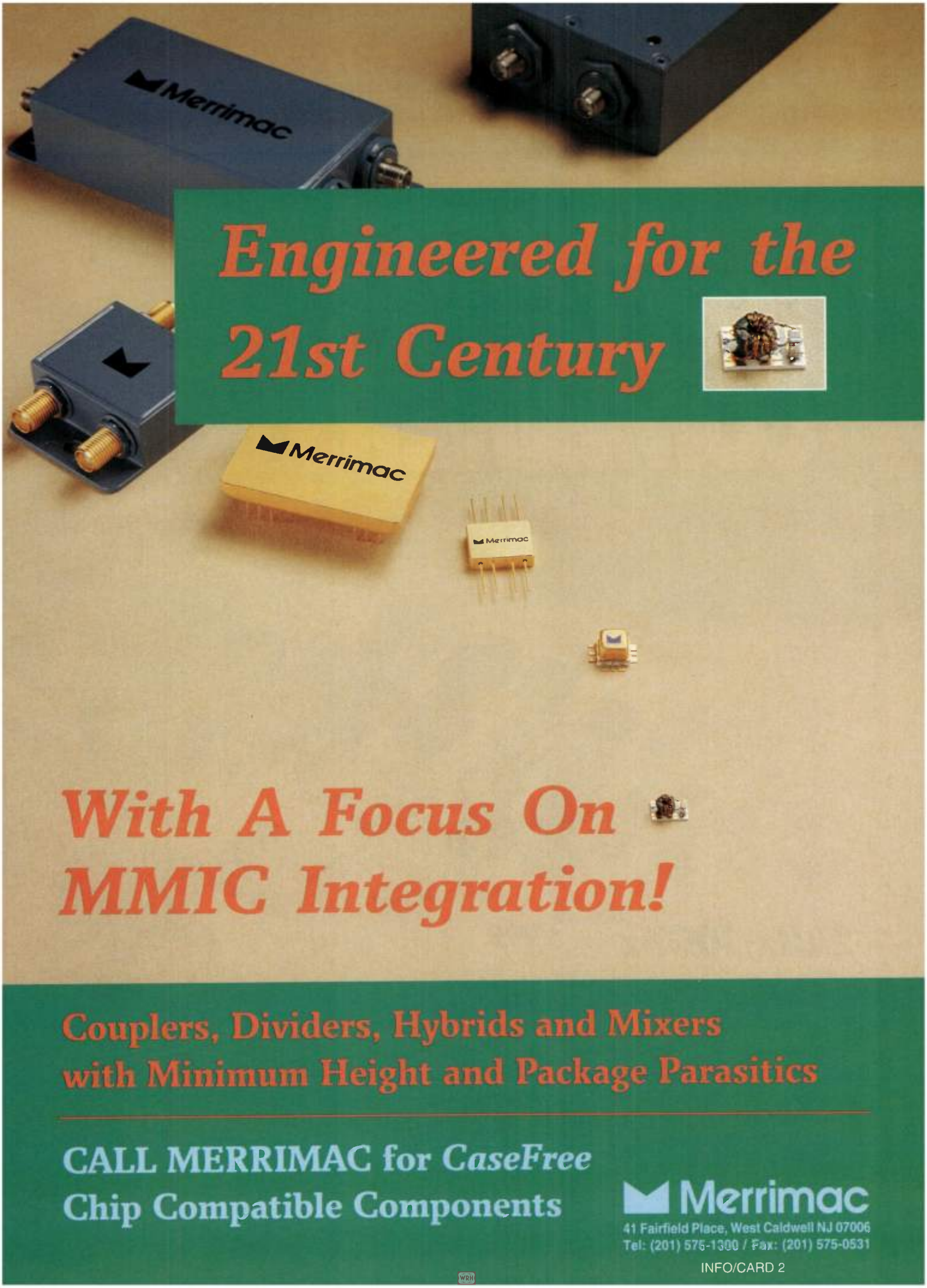


	LMX1501A	LMX1511	LMX2314/15	LMX2320	LMX2325	LMX2330A	LMX2331A	LMX2332A	LMX[illegible]	LMX2336	LMX2337
F Input-Main PLL F Input-Aux PLL	1.1GHz	1.1GHz	1.2GHz	2.0GHz	2.5GHz	2.5GHz 510MHz	[illegible] 510MHz	1.2GHz [illegible]MHz	1.1GHz 1.1GHz	[illegible]	550MHz [illegible]MHz
cc (typ) @3V owerdown (typ)	6mA N/A	6mA N/A	6mA 30μA	10mA 30μA	11mA 30μA	15mA 1μA	14mA [illegible]	8mA 1μA	9mA 1μA	11mA 1μA	9mA 1μA

National Semiconductor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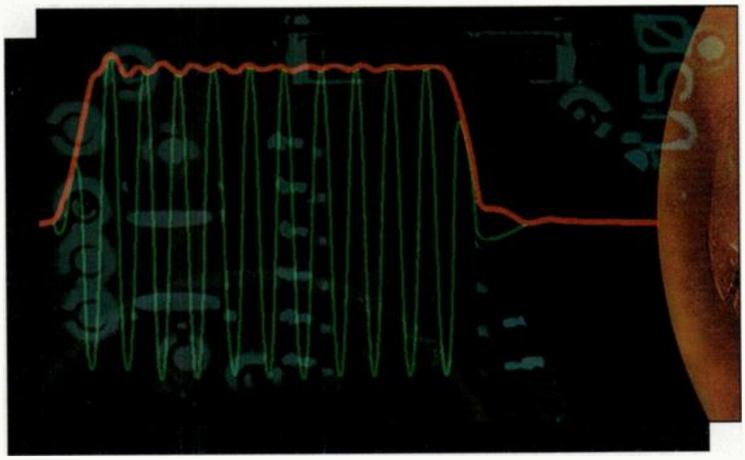
contents

September 1995

featured technology



cover story

tutorial

departments

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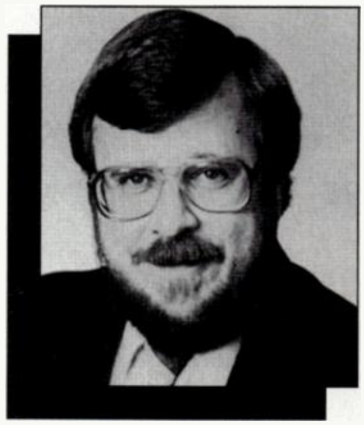
Announcing the New *RF Design Seminar Series*



By Gary A. Breed
Editor

At the center of *RF Design*'s mission is the delivery of timely and useful information. For ten years, part of that mission has been filled by the technical papers program and the full-day short courses offered at the RF Expo trade shows.

Now, we are taking that core educational program to new places. The *RF Design Seminar Series* has been developed to meet the instructional needs of engineers all around the country, not just on the East or West coast. In 1996, we are holding three events featuring an educational program that is much more comprehensive than we've had at the RF Expos.

The first of the series will be held in Dallas, Texas, January 17-19, 1996. We are excited to have an opportunity to bring *RF Design* to this important technology center. To our knowledge, no RF event like this has ever been held in Texas.

Next, we go to Las Vegas, April 24-26, 1996, where our seminars are part of the International Wireless Communications Expo (IWCE), sponsored by magazines of the Communications Group at Argus Integrated Media. Besides the hundred-plus radio communications suppliers, the RF Technology Pavilion at IWCE already has signed up more than 70 RF component, instrument and software companies to exhibit their wares.

The third venue for 1996 is greater Boston, in early October when the weather is great and fall colors are starting to show. We'll be out in the Route 128 corridor so engineers from high-tech companies in that area will be able to get to our seminars quickly and conveniently.

Now for the program — At the basic level, the *Introduction to RF Circuit Design* taught by Dave Hertling and Bob Feeney of Georgia Tech has been expanded to three days in order to provide a more thorough treatment of RF basics. The need for engineers to develop new wireless products has created a huge demand for training (and re-training) of new engineers and those adding new RF expertise.

For instruction in specific circuit design areas, the *Practical High Frequency Filter Design* and *Oscillator Design Principles* courses taught by Randy Rhea will be offered. At least two of our 1996 events will also include *Classes of Power Amplifers and When to Use Them*, taught by Nathan Sokol.

We have a large number of additional courses available to complete an advanced track, as well. We will select from topics like: *Antennas for Wireless*, *Digital Communications*, *Wireless Local Area Networks*, *Advanced Frequency Synthesis Techniques*, and *Low-Noise Amplifiers*. Watch for our announcements of specific seminar events to see which ones are offered.

Note that some of our events will also offer *Wireless Communications for Non-Engineers*, to help sales, marketing and purchasing staff understand new RF-based technologies.

Finally, we will have a technical papers program, too. A day of interesting contributions from you or your colleagues is planned, along with an informal poster session that will include a social reception sponsored by companies who have chosen our forum to exhibit their products and support our RF educational efforts.

We're excited about the new *RF Design Seminar Series* These three events are just the beginning. Different cities in different corners of the U.S. are on our future "hit list." Your neighborhood may be next!

"Over the last couple of years, we've seen EMC issues evolve from technical "what" questions to regulatory "when" statements. EMC is now, and at Kalmus we're addressing it head on."

"Our new Model 7100LC 100 watt RF power amplifier is a case in point. It has been designed specifically for EMC test applications. And, true to the Kalmus tradition of manufacturing highly functional, yet extremely cost competitive amplifiers, the 7100LC, we believe, is the smallest, lightest and least expensive 100 watt RF amplifier for EMC testing on the market."

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Frank Kalmus
Technical Director

Frank Kalmus is the founder of Kalmus (formerly Kalmus Engineering Inc.), a division of Thermo Voltek Corporation. As Kalmus' principal design engineer, he has designed over 200 RF amplifiers used for EMC test, medical/MRI, general laboratory, and communications applications.

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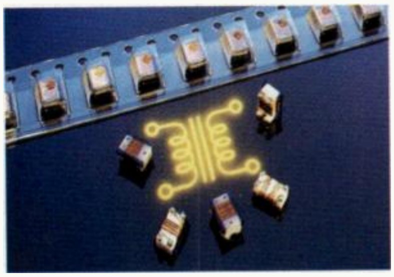

Wideband RF transformers are smallest available

These surface mount wideband RF transformers feature exceptionally low insertion loss and high current ratings. Built on a standard 1812 ferrite/ceramic body, their tops are encapsulated to protect the winding and provide a flat surface for pick and place equipment.

These 1:1 and 2:1 transformers come in 9 models covering the frequency range from 350 kHz to 680 MHz. Insertion loss is less than 1.5 dB and DC current handling ranges from 200 up to 700 mA.

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

INFO/CARD 9

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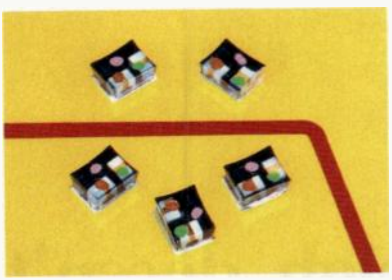

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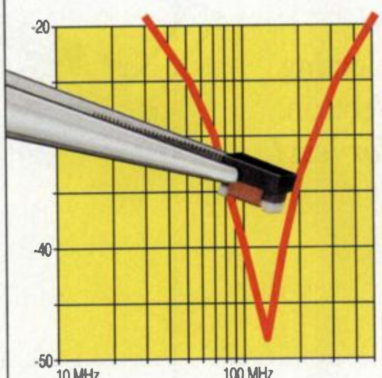

designs, this filter is much smaller and less expensive.

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INFO/CARD 12

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

What Ever Happened to ...

Editor,

I write to ask you or your staff for information concerning a fine old tube manufacturer – the Machlett Co. No information is to be found as to their address in the material that I have, so if you know what happened to them (absorbed/sold to another company??) I'd appreciate being able to contact them, particularly their chief engineer.

Charles Ratcliffe
Box 1068
Friday Harbor, WA 98250

Editor,

Does anyone know what happened to the Radiometer Electronics Co., which I believe was located near Copenhagen, Denmark? They made high quality RF test equipment in the 70's and 80's. Is there a successor company or any enterprise supplying info and spare parts?

Charles King
P.O. Box 116
E. Berlin, CT 06023

Don't Oversimplify

Editor,

When I saw the tutorial on filter responses in *RF Design*'s July issue, (pp. 68 - 73), I felt somewhat stimulated to read it. I was somewhat disappointed, however, when I read your introduction to the article. You justified your grossly oversimplified presentation (not one actual formula for a filter response!) by referring to your, "...philosophy of presenting concepts that assume a BSEE and no experience..." If your idea that a person with a BSEE and no experience cannot handle simple equations for filter responses is accurate, I have some bad news for these budding engineers: they're not likely to get any experience! It is ironic that you could have this opinion, yet, in another article in the same issue, you have no problem showing circuit diagrams and Smith charts! Oversimplified tutorials that simply rehash what graduating engineers have likely already been exposed to in their classwork (to much greater, i.e. useful, detail) don't help anyone. I would suggest that you ditch your philosophy of assuming that your readers are technically illiterate.

Glenn Stumpff

For the most part I agree with Mr. Stumpff. Perhaps we chose too broad a subject to cover in the allotted two pages of space. In general, however, reader feedback has been strongly in favor of these simplified tutorials. This approach is confirmed by the kinds of questions asked during classes held at our RF Expo conferences – many engineers are starting with virtually no RF concepts. – Editor

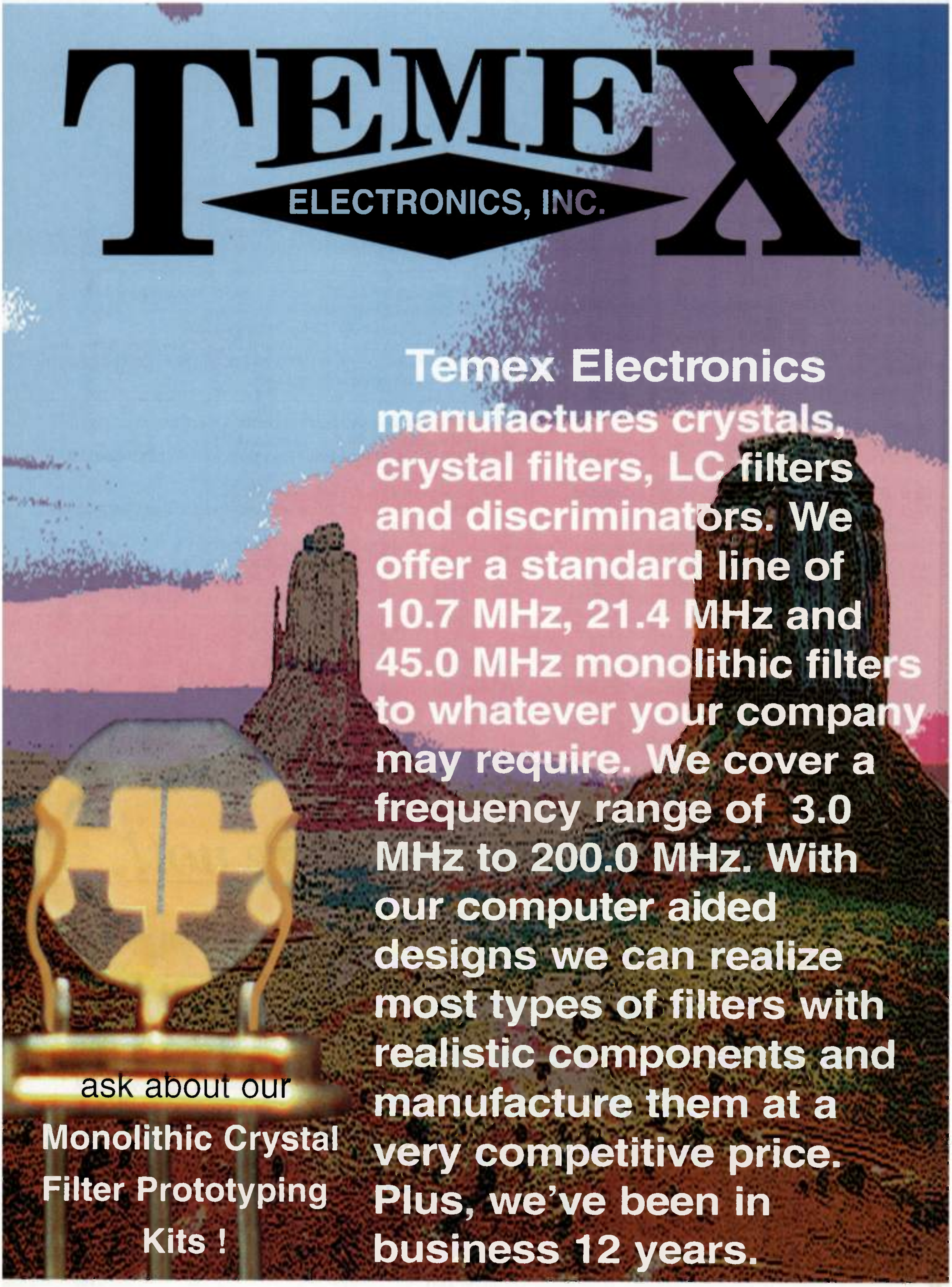
Introduction to RF & Wireless Engineering
October 10-13, 1995, Atlanta, GA
Principles of Modern Radar
November 6-10, 1995, Atlanta, GA
Far-Field, Anechoic Chamber, Compact and Near-Field Antenna Measurements
November 28-December 1, 1995
Information: Continuing Education, Georgia Institute of Technology, Atlanta, GA 30332-0385. Tel: (404) 894-2547.

Antennas and Antenna Systems: Practical Design, Implementation, and Testing
October 23-26, 1995, Washington, DC
Analog and Digital Cellular Networks: CDMA versus TDMA
October 30-November 1, 1995, Washington, DC
Wireless Telecommunications: An Introduction
November 13-15, 1995, Washington, DC
Digital Transmission Systems
November 13-17, 1995, San Diego, CA
Video Transmission and Broadcasting Via Satellite
November 20-21, 1995, Washington, DC
New HF Communications Technology: Advanced Techniques
December 4-8, 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.

Antenna Theory Simplified
September 14, 1995, Longmont, CO
Common-Mode Filter Design
September 15, 1995, Longmont, CO
Electromagnetic Compatibility Engineering
October 4-6, 1995, Palo Alto, CA
Information: Henry Ott Consultants, 48 Baker Road, Livingston, NJ 07039. Tel: (201) 992-1793, Fax: (201) 533-1442.

Short Course on Microwave and RF Measurements for Wireless Communications
November 28-29, 1995, Scottsdale, AZ
Information: Robert M. Judish, NIST, 325 Boradway, Boulder, CO 80303. Tel: (303) 497-3380, Fax: (303) 497-3970, email: judish@boulder.nist.gov.

RF & Microwave Measurements & Applications
October 2-5, 1995, Washington, DC
MPEG-2 & Alternative Video Compression Standards & Techniques
November 6-8, 1995, Palo Alto, CA
Error Correcting Codes with Applications to Communications Systems
December 11-14, 1995, Washington, DC
Information: University Consortium for Continuing Education, 16161 Ventura Boulevard, M/S C-752, Encino, CA 91436. Tel: (818) 995–6335. Fax: (818) 995–2932.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10.7 MHz

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2	TE5000	3	3.75	20	18.0	-	-	2	1.0	50	1800//+4
4	TE5010	3	3.75	30	14.0	-	-	3	2.0	60	1500//+3
6	TE5020	6	3.75	60	12.5	-	-	4	2.0	70	1500//+3
8	TE5030	6	3.75	60	10.0	90	12.5	5	2.0	80	1500//+3
2	TE5040	3	6.50	20	30.0	-	-	1	1.0	50	2700//0
4	TE5050	3	6.50	30	15.0	-	-	2	2.0	75	3100//0
6	TE5060	6	6.50	60	19.5	-	-	3	2.0	90	3100//0
8	TE5070	6	6.50	60	13.0	80	17.5	4	2.0	100	3100//0
2	TE5080	3	7.50	20	35.0	-	-	1	1.0	50	3000//0
4	TE5090	3	7.50	30	17.5	-	-	2	2.0	75	3300//0
6	TE5100	6	7.50	60	22.5	-	-	3	2.0	90	3300//0
8	TE5110	6	7.50	60	15.0	80	20.0	3	2.0	100	3300//0
2	TE5120	3	15.0	20	70.0	-	-	1	1.0	35	5000//-1
4	TE5130	3	15.0	30	35.0	-	-	2	2.0	60	5000//-1
6	TE5140	6	15.0	60	45.0	-	-	2	2.0	90	5000//-1
8	TE5150	6	15.0	60	30.0	80	40.0	3	2.0	100	5000//-1

21.4 MHz

NO. POLES	TEMEX P/N	PASSBAND dB	PASSBAND ±KHz	STOPBAND dB	STOPBAND ±KHz	STOPBAND dB	STOPBAND ±KHz	LOSS dB	RIPPLE dB-MAX	ULT. REJ. dB-MIN.	TERM.(Rp//Cp) OHM/PF
2	TE5180	3	3.75	15	12.5	-	-	2	1.0	50	850//+6
4	TE5190	3	3.75	30	12.5	-	-	3	2.0	70	850//+5
6	TE5200	6	3.75	60	12.5	-	-	4	2.0	90	850//+5
8	TE5210	6	3.75	60	10.0	80	12.5	5	2.0	100	850//+5
2	TE5220	3	6.50	15	20.0	-	-	2	1.0	50	1300//+2
4	TE5230	3	6.50	30	22.5	-	-	3	2.0	70	1400//0
6	TE5240	6	6.50	60	22.5	-	-	4	2.0	90	1400//0
8	TE5250	6	6.50	60	17.5	80	22.5	4	2.0	100	1400//0
2	TE5260	3	7.50	15	25.0	-	-	2	1.0	50	1500//0
4	TE5270	3	7.50	30	25.0	-	-	3	2.0	70	1600//0
6	TE5280	6	7.50	60	25.0	-	-	4	2.0	90	1600//0
8	TE5290	6	7.50	60	20.0	80	25.0	4	2.0	100	1600//0
2	TE5300	3	15.0	15	50.0	-	-	2	1.0	45	3000//0
4	TE5310	3	15.0	30	45.0	-	-	3	2.0	60	3000//-1
6	TE5320	6	15.0	60	45.0	-	-	3	2.0	90	3000//-1
8	TE5330	6	15.0	60	33.0	80	45.0	4	2.0	100	3000//-1

45.0 MHz

NO. POLES	TEMEX P/N	MODE	PASSBAND dB	PASSBAND ±KHz	STOPBAND dB	STOPBAND ±KHz	LOSS dB	RIPPLE dB-MAX	ULT. REJ. dB-MIN.	TERM.(Rp//Cp) OHM/PF
2	TE9420	3-OT	3	3.75	18	16.0	3	1	40	2000//-1.0
4	TE9310	3-OT	3	3.75	30	12.5	3	1	70	2000//-1.0
2	TE7420	3-OT	3	7.50	18	28.0	2	1	40	3000//-1.0
4	TE7430	3-OT	3	7.50	40	30.0	3	1	70	3000//-1.0
2	TE7440	3-OT	3	15.0	15	47.0	2	1	40	8000//-1.5
4	TE7450	3-OT	3	15.0	30	50.0	3	1	70	8000//-1.5
2	TE7730	FUND	3	15.0	15	50.0	2	1	40	1100//+1.5
4	TE7740	FUND	3	15.0	40	60.0	3	1	70	800//+1.0

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RF calendar

September

20-22 **17th Annual Piezoelectric Devices Conference**
Kansas City, MO
Information: Components Group, Electronic Industries Association, 2500 Wilson Boulevard, Arlington, VA 22201–3834. Tel: (703) 907–7500. Fax: (703) 907–7501.

26-28 **Sixth International Conference on Radio Receivers and Associated Systems**
University of Bath, UK
Information: RRAS'95 Secretariat, IEE Conference Services, Savoy Place, London WC2R 0BL UK.
Tel: +44 (0) 71 344 5477. Fax: +44 (0) 71 497 3633.

27-29 **6th IEEE International Symposium on Personal, Indoor and Mobile Communications**
Toronto, Canada
Information: University of Toronto, Dept. of Electrical and Computer Engineering, 10 Kings College Rd., Toronto, Ontario M5S 1A4, Canada; Tel: (416) 978-3652.

October

3-11 **TELECOM 95**
Geneva, Switzerland
Information: International Telecommunication Union, Place des Nations, CH-1211 Geneva 20, Switzerland. Tel: 41 22 730 5111. Fax: 41 22 733 7256.

8-11 **1995 Wireless Circuits, Interconnection, and Assembly Workshop**
Tucson, AZ
Information: Wireless Workshop, 100 S. Roosevelt Avenue, Chandler, AZ 85226.
Tel: (602) 961–1382. Fax: (602) 961–4533.

22-26 **ISHM International Symposium on Microelectronics**
Los Angeles, CA
Information: ISHM, 1850 Centennial Park Dr., Suite 105, Reston, VA 22091. Tel: (800) 535-4746 or (703) 758-1060, Fax: (703) 758-1066, email: ISHM@aol.com.

November

30-1 **45th Conference, Automated RF Techniques Group**
Scottsdale, AZ
Information: Harmon W. Banning, W.L. Gore & Assoc., 1901 Barksdale Road, Newark, DE 19711. Tel: (302) 368-3700, Fax: (302) 292-4607

January

17-19 **RF Design Seminar Series**
Dallas, Texas
Information: Argus Trade Shows, 6151 Powers Ferry Road, N.W., Atlanta, GA 30339. Tel: (800) 828-0420.

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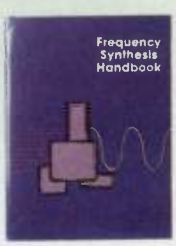

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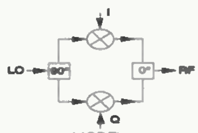

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MIQA-21M	20	23	6.2	0.14	50	40	48	65	39.95
MIQA-70M	66	73	6.2	0.10	38	38	48	58	39.95
MIQA-70ML	66	73	5.7	0.10	38	38	48	58	49.95
MIQA-91M	86	95	5.5	0.10	38	38	48	58	49.95
MIQA-100M	95	105	5.5	0.10	38	38	48	58	49.95
MIQA-108M	103	113	5.5	0.10	38	38	48	58	49.95
MIQA-195M	185	205	5.6	0.10	38	38	48	58	49.95
MIQC-38M	34	38	5.6	0.10	48	37	54	65	49.95
MIQC-88M	52	88	5.7	0.10	41	34	52	66	49.95
MIQC-176M	104	176	5.5	0.10	38	36	47	70	54.95
MIQC-895M	868	895	8.0	0.10	40	40	52	58	99.95
MIQC-1785M	1710	1785	9.0	0.30	35	35	40	65	99.95
MIQC-1880M	1805	1880	9.0	0.30	35	35	40	65	99.95
☐ MIQY-70M	67	73	5.8	0.20	40	36	47	60	19.95
☐ MIQY-140M	137	143	5.8	0.20	34	36	45	60	19.95
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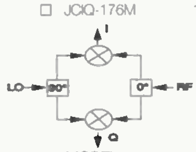
☐ JCIQ-88M	52	88	5.6	0.1	40	35	45	65	49.95
☐ JCIQ-176M	104	176	5.6	0.1	35	35	45	65	54.95



I/Q DEMODULATORS

MODEL	FREQ. (MHz) f_L	f_U	CONV LOSS (dB) $\bar{x}$	σ	AMP. UNBAL. (dB) Typ.	PHASE UNBAL. (Deg.) Typ	HARM. SUPPRESS (-dBc) Typ. 3xI/Q	5xI/Q	PRICE $ Qty. (1-9)
MIQA-10D	9	11	6.0	0.10	0.15	1.0	50	65	49.95
MIQA-21D	20	23	6.1	0.15	0.15	0.7	64	67	49.95
MIQA-70D	66	73	6.2	0.10	0.15	0.7	56	58	49.95
MIQC-38D	34	38	5.5	0.10	0.10	0.5	60	65	49.95
MIQC-60WD	20	60	5.3	0.10	0.15	1.0	55	67	79.95
MIQC-895D	868	895	8.0	0.20	0.15	1.5	40	55	99.95
☐ MIQY-1.25D	1.15	1.35	5.0	0 0	0.15	1.0	59	67	29.95
☐ MIQY-70D	67	73	5.5	0.25	0.10	0.5	52	66	19.95
☐ MIQY-140D	137	143	5.5	0.25	0.10	0.5	47	70	19.95
Surface Mount Models									
☐ JCIQ-176D	104	176	5.5	0.1	0.15	2	52	65	54.95
☐ JCIQ-895D	868	895	8.6	0.1	0.2	1	45	65	99.95
☐ JCIQ-1785D	1710	1785	8	0.2	0.2	2	50	65	99.95
☐ JCIQ-1880D	1805	1880	8	0.2	0.2	2	50	65	99.95

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INFO/CARD 17

FAA, E-Systems Demonstrate GPS-Based Landing System

E-Systems reports that it has successfully completed a series of flight tests at NASA's Crows Landing facility showing that the Global Positioning System (GPS), combined with Kinematic Carrier Phase Tracking (KCPT) technology, is a potential replacement for the existing Instrument Landing System (ILS) for aircraft. The FAA is expected to have a final analysis and report on the flight data by late summer. The avionics equipment included a modified GPS receiver with KCPT software, an E-Systems D8PSK VHF data link transceiver, a Pentium-based processor, and a Litton IRU for attitude data. Software on a 486-based computer calculated the improved GPS position and ILS-like guidance at a 15 Hz update rate. The ground system consisted of a ground reference station (GRS) and an independent ground integrity monitor (GIM). The ground-based GPS receiver was integrated with the GRS to provide pseudorange and carrier phase information which was then uplinked to the approaching aircraft using the D8PSK link. The system achieved accuracies of 0.31 meter vertical and 1.6 meters laterally, well within the requirement of 1.2 meters and 4 meters, respectively.

Conference News

RF Design Seminar Series Expands Educational Program

RF Design magazine announces the new RF Design Seminar Series, with three educational conferences planned in 1996. The program of short courses and technical papers will expand on the educational program developed for the RF Expo trade shows. Three tracks of formal instruction will offer RF design fundamentals, specific design techniques, and advanced instruction. A one-day technical papers program will allow engineers to present recent work, and a late afternoon poster session will be combined with a social reception for informal technical discussions. The first seminar will be held in late January in Dallas, Texas. The second will be held in conjunction with the International Wireless Communications Expo in Las Vegas in April, and the third is being planned for early fall 1996 in Boston, Mass. Engineers interested in presenting papers should contact *RF Design* magazine at 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111. Persons wishing to receive more information on the seminars may contact Argus Trade Shows at (800) 828-0420.

Call for Papers — Multichip Modules

The Fifth International Conference on Multichip Modules has issued a call for papers. Topics sought include MCM applications, testing, assembly and interconnection, MCM/PCB interface, management and cost strategies, electrical the thermal performance, telephone and cellular applications. 150-word abstracts should be prepared by October 16, 1995 and submitted to MCM '96 Conference Services, ISHM, 1850 Centennial Park Drive, Suite 105, Reston, VA 22091; Tel: (703) 758-1060, Fax: (703) 758-1066.

BIS Reports Increase in European Consumer Spending

BIS Strategic Decisions reports a predicted growth in home entertainment in Western Europe to $33 billion through the end of 1995, a nine percent increase over 1994. Interactive multimedia, continued sales of CDs, pay TV and interactive TV play a major role in the increase, according to BIS, who is sponsoring their Third Annual Interactive Conference in Montreaux, Switzerland, September 26-28, 1995. Information on the conference can be obtained from BIS' U.K. office at: +44 1582 405678.

SETI League Relocates

The SETI League, Inc., a non-profit leader in a scientific Search to Extra-Terrestrial Intelligence has moved to new quarters at 433 Liberty Street, Little Ferry, NJ 07643. Offices, laboratory and a SETI library have been provided at no cost to membership through the generosity of building occupants Eventide Inc., a longtime supporter of SETI. Since Congress terminated all NASA funding for SETI in 1993, private organizations have attempted to continue experiments. The SETI League, Inc. maintains a membership hotline at (800) TAU-SETI. Dr. H. Paul Shuch is full-time Executive Director of the SETI League, on a one-year leave of absence from the Pennsylvania State University system.

Magnetic Materials Scholarships Offered

The Magnetic Materials Producers Association (MMPA) announces that applications are now being accepted for the Association's Scholars Program. Two $5000 awards will be made to individuals doing graduate level studies in the field of magnetics. The purpose of the program is to encourage students entering graduate school to pursue a concentration in the area of magnetics. Written applications should be submitted to the MMPA by October 16, 1995. Awards will be announced by February 15, 1996. For information, contact the MMPA at 11 South LaSalle Street, Suite 100, Chicago, IL 60603; tel: (312) 201-0101, fax: (312) 201-0214.

Colorado Scientists Achieve Bose-Einstein Condensation

Researchers at JILA, a joint program of the National Institute of Standards and Technology and the University of colorado at Boulder, have announced that they achieved a temperature far lower than any previously produced, creating an entirely new state of matter predicted 70 years ago by Albert Einstein and Satyendra Nath Bose. The team cooled rubidium atoms to 170 nanokelvin (billionths of a degree Celsius) above absolute zero, where individual atoms condensed into a "superatom" that behaved as a single entity. The scientists further cooled the atoms to 20 nanokelvin above absolute zero, the lowest temperature ever achieved. Physicists worldwide have been searching for 15 years to achieve this condensation.

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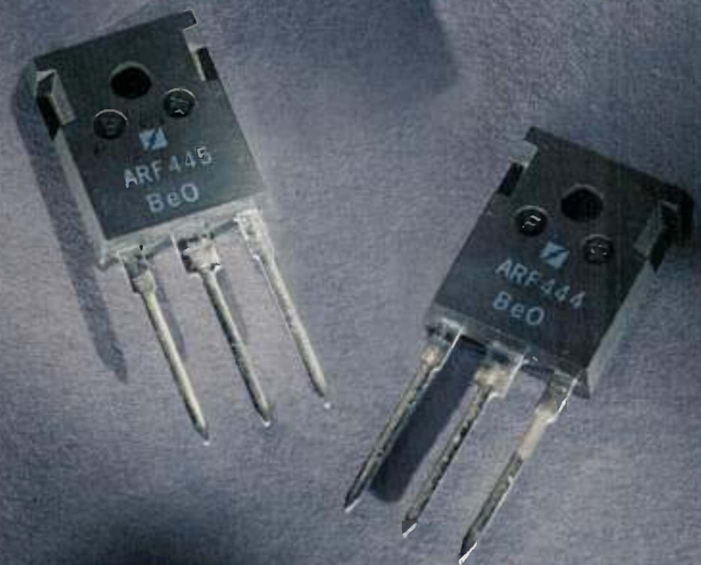

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Business Briefs

KVG and MORION Join Forces — KVG announces completion of a cooperation agreement with MORION of St. Petersburg, Russia and Staudte Engineering of Cedar City, Utah. MORION, a privatized crystal and oscillator company, is majority owned by KVG and Staudte and specializes in high performance oscillators.

BCP Awarded Encoding Patent — Broadband Communications Products, Inc. has been awarded a U.S. patent for a pre-scramble encoding method and device for digital communication. The purpose of the invention is to ensure that a scrambled data transmission has adequate signal balance and frequent data transitions, maintaining proper randomness.

Stanford Telecom Joins Cable Group — Stanford Telecommunications has announced that it has joined the Cable/Information Technology Convergence Forum, an organization sponsored by Cable Television Laboratories, Inc. (CableLabs). the Forum was created to smooth the exchange of information between high technology companies and cable operators.

TAS in Joint Effort with AT&T Paradyne — Telecom Analysis Systems (TAS) and AT&T Paradyne have initiated a joint effort to develop a new standard for testing cellular modem products. The target is a set of laboratory-based testing procedures that can be used in conjunction with field tests to ensure modem quality. TAS recently introduced GT-Cellular test system is being used as the laboratory test bed for cellular modem evaluation.

Harris Opens Singapore Center — A new Center of Excellence is announced by Harris Corporation, located in the Cencon 1 Building in Singapore. The facility combines functions serving the Asian market, including field engineering, technical support, staff specialists, plus marketing and sales.

BTG Licenses Technology to E.F. Johnson — BTG USA Inc. has licensed its narrowband linear modulation technology to E.F. Johnson Company. The modulation technique, called transparent-tone-in-band (TTIB), can significantly increase the capacity of mobile communications systems. E.F. Johnson will use the technology in products for the 220 MHz band.

Scientific-Atlanta Gets IRIDIUM Contract — Motorola's Satellite Communications Division has placed an order for four additional System Control Segment earth terminals for the IRIDIUM system. With this order, fourteen telemetry and command terminals are being produced by Scientific-Atlanta.

Watkins-Johnson International Moves — Watkins-Johnson International has moved to new offices in Windsor, England. The new address is: St. Leonards House, St. Leonards Road, Windsor, Berkshire SL4 3DG, U.K.; Tel: +44 1753 751300; Fax: +44 1753 621794.

Lasertron Forms Wireless Business Unit — Lasertron has formed a new Wireless Business Unit as a separate division to meet requirements of the cellular and PCS markets for fiber-optic subsystem products, primarily for linking remote antennas with base station equipment.

Rogers Opens Hong Kong Office — Rogers Corporation has established a regional subsidiary, Rogers Southeast Asia, in Hong Kong. The office supports growing activity for Rogers and Durel Corporation in Southeast Asia, providing sales and technical support.

Applied Signal Technology has New Address — Although the company has not moved, the address for Applied Signal Technology, Inc. has been changed. Their mailing address is 400 West California Avenue, Sunnyvale, CA 94086, and the shipping address has become 600 West California Avenue.

Contracts

Anritsu Wiltron Gets Marine Corps Order — Anritsu Wiltron announces a $3 Million order from the Marine Corps for the 68347M 20 GHz Synthesized Signal Generator. This order is a follow-on to an initial $2 Million order placed in August 1993, and is part of a five-year contract.

Svetlana Receives British Defence Contract — Svetlana Electron Devices, Inc., the American joint venture partner with Svetlana Electron Devices of St. Petersburg, Russia, announces a major contract for more than 10,000 power electron tubes from the British Ministry of Defence. The power tubes will be used in military shipboard radio equipment. The contract was won in competition with manufacturers in the U.S., France and China.

RFID Runs Underground Traffic System — The Huogito Coal Mine in China has installed a traffic monitoring and control system using Texas Instruments' TIRIS RFID hardware. The underground mine uses operator-driven vehicles rather than the usual rail car system, and the traffic control system helps reduce collisions and traffic jams. The system also tracks each vehicle's location, allowing dispatchers to control their movement efficiently.

W-J Receives $16M AMRAAM Contract — Watkins-Johnson Company announces a contract for more than $16 Million from Hughes Aircraft Company, Missile Systems Group. Under the contract, W-J will continue production of electronics subsystems for Lot 9 of the AIM-120 AMRAAM air-to-air missile.

ManTech Gets Test Contract — ManTech Test System has been issued a four-year GSA contract for their family of Aurora test systems, models VTS-1000/7, VTS-1000/39, VTS-1000/99, VTS-2000 and VTS-3000. The systems are used by the Department of Defense, other government agencies and commercial customers for the test and repair of digital, hybrid and RF electronic circuit boards and black boxes. The VTS-1000/7 has been certified as the replacement for the GenRad 2225/2235 (AN/USM-465) system.

Labs Focus on Commercial Applications

By Andy Kellett
Technical Editor

Electromagnetic fields have been fundamentally understood since 1864, when James Clerk Maxwell unified the equations describing the behavior of electric and magnetic fields. However, more than Maxwell's equations are required to produce the advanced communications devices being built today, and researchers are working on developing new devices and new types of systems to exploit the fields Maxwell described in the 19th century.

Semiconductor Devices

Much of the attention given to advances in RF technology is given to new semiconductor devices. Most of the research being done on new RF semiconductors is being done in commercial labs. Most semiconductor device manufacturers are working on ways to make devices that operate at lower voltages, use less power, and incorporate more functionality.

However, some recent advances have their roots in more "pure research" labs. Dr. Bernard Meyerson, leading a team at the IBM Thomas J. Watson Research Center, developed a new semiconductor process that uses a Si/Ge alloy to produce very fast devices. Not only are the devices fabricated in this process fast, they can be manufactured using essentially the same fab lines being used to make silicon circuits. This technology made the leap to the commercial area in the form of a 1 GHz, 12-bit D-to-A converter designed by Analog Devices and fabricated by IBM.

Superconductors

High temperature superconductors garnered a lot of attention when they appeared on the scene in 1987. Since then, commercial products using this technology have been trickling out of the labs and into the market. An example of one of the emerging superconducting devices are the very sharp, low loss cellular basestation filters developed by Conductus, Inc.

Most of the research done by Conductus has focused on ways to make fabrication of superconducting devices as predictable and routine as the fabrication of semiconductor devices. "A lot of the work is centered on trying to really understand the correlations between material growth characteristics and material properties, and between material properties and device properties," says Randy Simon, Vice President of Technology Programs for Conductus, "Eventually we would like to control material growth parameters so as to produce ideal device parameters and basically not even have to worry about the material's parameters."

However, work on filter design is also being done. "We are working at the problem from both directions," says Simon, "designing devices so they are more tolerant of variations in material properties, and improving materials so that the devices are less constrained. "

Electromagnetic Modeling

Both the materials and geometries used to package RF devices are the subject of research. Reducing package cost and device degradation are the goals of this research.

Refining device geometries to optimize performance requires electromagnetic modeling. Many researchers are working to make electromagnetic modeling algorithms faster and device and material models more accurate. This type of research is particularly attractive to smaller universities, where the cost of cutting edge semiconductor fabrication equipment is out of reach, but a fast powerful computer can be purchased for a small fraction of an annual budget.

System Research

Not all RF research is concerned with devices. Predicting the behavior of RF communications systems and finding ways to make them more reliable is also getting research attention. Not only are service providers doing this kind of research, but there is plenty of this kind of research being done at universities. Members of the Mobile & Portable Radio Research Group at Virginia Tech have done work in propagation prediction, wireless communications simulation and other areas important to system operation.

Lincoln Laboratories is working on ground terminals for military communication systems, says Roger Sudbury, Executive Officer and Director of External Affairs for Lincoln Labs. While this is not a commercial project, "... a lot of things that the Laboratory has done for the Department of Defense ultimately find themselves in commercial applications," says Sudbury.

Laboratory Funding Trends

Whether they produce technology that ultimately becomes commercial or not, many DOD-funded labs have been pared back in recent years. According to Lincoln Labs' Sudbury, funding for Lincoln Labs has shrunk by about 30 percent in the last five tears, and in the same time, the number of people employed at the lab has shrunk from just over 2,800 to about 2,250.

Research groups with weaker ties to DOD funds, such as the Georgia Tech Research Institute, have been able to partially replace lost funds.

Finally, the companies producing devices for the new wireless communications services are investing a lot of money in their R&D departments. If those departments are successful in crafting devices and systems that the marketplace finds useful, they may also share in the return on investment. *RF*

SPICE Techniques for Analyzing Quartz Crystal Oscillators

By T. Kien Truong
Boeing Commercial Airplane Group

We can readily find several articles on oscillator analyses published in trade journals but rarely can one find articles on crystal oscillator simulation. In fact, the author knows of no material published on transient simulation. The fact that a quartz crystal is an extremely high Q circuit and the oscillator is free-running with no driving signal makes crystal oscillator simulation difficult. This tutorial introduces new techniques for obtaining the startup transient by inserting power supply noise, and techniques for finding the open-loop response by inserting a pseudo voltage source at the CMOS amplifier input. We will also examine the frequency pulling, resonant and anti-resonant behaviors of the crystal oscillator.

Often in designing ASICs the convenience, controllability, and economy of an embedded crystal oscillator offers a major advantage. Together with the off-chip passive components, the on-chip active circuitry can give a reliable and stable oscillator. In a Pierce oscillator, the on-chip circuitry is an inverting amplifier that provides the gain necessary to build up and maintain the oscillation through a crystal controlled feedback network. In order to oscillate, the circuit must satisfy the Barkhausen criteria that the phase shift around the loop is n360° and the loop gain exceeds unity at the resonant frequency.

Standard SPICE analyses such as transfer function, fourier transform, signal-to-noise plot and closed loop analyses are straight forward, often described in literature and we will not discuss them here. The tutorial will show that we can obtain a wealth of other information such as startup transient, open-loop response and Nyquist plots using the evaluation version of the PSpice simulator from MicroSim. This free version of MicroSim limits the complexity of the design to about 10 transistors, which is enough to simulate our Pierce oscillator circuit. The circuit in this tutorial is a simplified version of the real circuit used in the ASIC. The ASIC circuit includes additional transistors as loading transistors for the gain stage and as high-value feedback resistor for the inverting amplifier. Omitting these elements allows the use of the evaluation software and does not affect the analyses of oscillator behaviors.

This work is part of the design and development of a highly reliable clock ASIC. The ASIC employs redundant oscillators with automatic fault detection, identification and reconfiguration. The digital/analog ASIC was simulated in a mixed mode environment with a full-blown version of SPICE and was successfully fabricated with a dual-poly NWell CMOS technology.

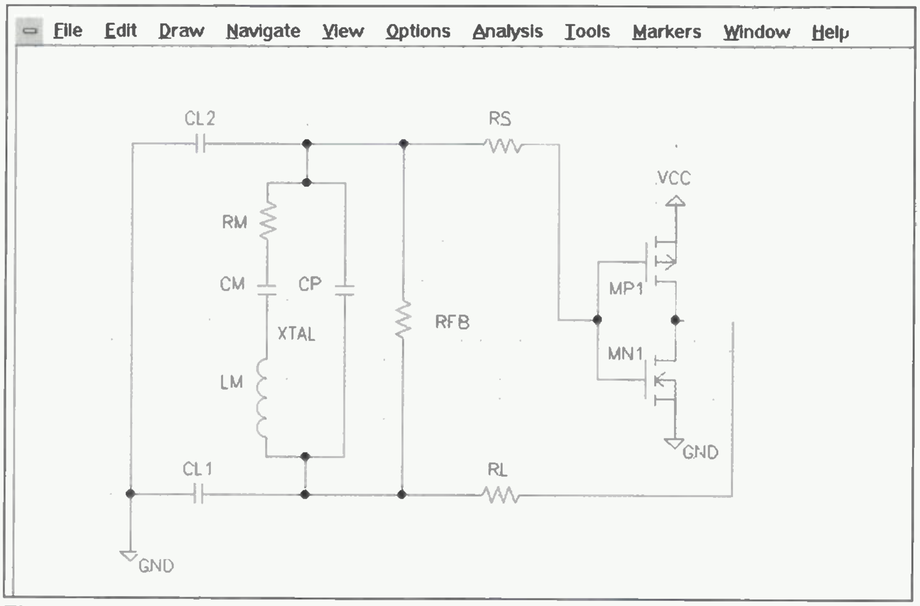

Figure 1. Basic crystal oscillator schematic.

Circuit Description

We can model the quartz crystal as having a motional resonance arm R_x, L_x and C_x in parallel with the crystal holding capacitance C_P (the mounting electrodes). Most crystal manufacturers can provide the motional parameters. Figure 1 illustrates the Pierce oscillator circuit using a CMOS inverting amplifier and a 24 MHz crystal. All circuit parasitic elements, on-chip or off-chip, are lumped into the bulk capacitors. The on-chip amplifier influences the frequency through its input and output (pin-to-ground) capacitances, and its pin-to-pin capacitance. We can also model the amplifier internal delay as additional output capacitance. Figure 2 shows the circuit listing in PSpice format. The level-2 NMOS and PMOS transistor parameters are from a 2-micron NWell CMOS technology.

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The amplifier for the crystal oscillator is a high-gain, single-stage CMOS inverter that is biased at its mid range to act as a linear amplifier for the clock signal. The biasing resistor R_f produces a negative feedback signal that competes with the more favorable positive feedback signal through the crystal network. The biasing resistor therefore must be large enough (22 MΩ) not to load down the low impedance crystal feedback loop and yet is small compared to the CMOS amplifier input impedance. The amplifier provides a nominal 180° phase shift in addition to the 180° phase shift in the feedback loop. The amplifier has to provide enough gain to offset the loss in the feedback loop. Too much gain, however, will facilitate unwanted crystal overtone and spurious effects or even make the oscillator startup unreliable.

```
* Single inverting amplifier
MNl NVOUT NVTN 0 0 NMOS L=2U W=480U
MPl NVOUT NVIN 99 99 PMOS L=2U W=1130U

* 24 MHz crystal
.SUBCKT XTAL24 1 2
*CP 1 2 4.5E-12
RM24 10.2
CM 3 4 20.937E-15
LM 1 3 2.1E-3
.ENDS

* Power supply noise to kick start the oscillator
V99 99 0 PWL ON 5V .lUS SV .105US 0V .11US
5V 100MS 5V

* Feedback circuit
XTAL NXIN NXOUT XTAL24
RL NVOUT NXIN 200
CLl NXTN 0 10PF
CL2 NXOUT 0 10PF
RFB NVOUT NVTN 22E6
RS NXOUT NVIN 200

* Transient analysis
.TRAN lNS 20US 0 10ns
.PROBE

* N-Well transistor model
.MODEL NMOS NMOS
+ LEVEL=2 VTO=0.825 UO=608.3 TPG=l
+ TOX=4.0E-8 NSUB=7.755E15 XJ=4.50E-7
+ LD=1.121E-7 DELTA=3.714 VMAX=49.89E+3
+ NFS=.105E12 CJ=323.1E-6 CJSW=929.9E-12
+ MJ=461.5E-3 MJSW=268.3E-3 PB=.44
+CGSO=96.77E-12 CGDO=96.77E-12
CGBO=40.0E- 12
+ UCRIT=50E3 UEXP=78.26E-3 NEFF=3.36

.MODEL PMOS PMOS
+ LEVEL=2 VTO=-0.70 UO=205.1 TPG=-l
+ TOX=4.0E-8 NSUB=1.486E16 XJ=450E-9
+ LD=230.5E-9 DELTA=1.843 VMAX=40.76E3
+ NFS=0.OlE12 CJ=804.9E-6 CJSW=749.1E-12
+ MJ=525.0E-3 MJSW=495.4E-3 PB=.958
+ CGSO=l99.OE-12 CGDO=l99.OE-12
CGBO=101.5E-12
+ UCRIT=70E3 UEXP=184.2E-3 NEFF=0.69

.END
```

Figure 2. Circuit listing for transient analysis.

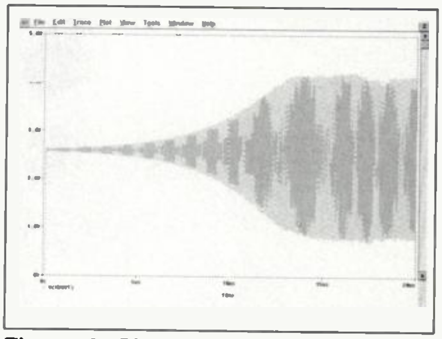

Figure 3. Simulated startup transient waveform.

The crystal in the feedback loop has an extremely sharp phase versus frequency response at the resonant frequency. The series resistor RL helps set the phase in the feedback loop and also isolate the crystal from the amplifier circuit. RS and RL also include the ESD protection resistors in the I/O pads. The bulk capacitors parallel the

```
* Single inverting amplifier
MNl NVOUT NVIN 0 0 NMOS L=2U W=480U
MP1 NVOUT NVIN VCC VCC PMOS L=2U
W=1130U

* Pseudo voltage source inserted in feedback loop
VPS NVIN 0 DC 2.557 AC 1

* 24MHz crystal
.SUBCKT XTAL24 1 2
CP 1 2 4.5E-12
RM24 10.2
CM 3 4 20.937E-15
LM 1 3 2.1E-3
.ENDS

* Feedback circuit
XTAL NXIN NXOUT XTAL24
VC9VCC0DC5V
RL NVOUT NXIN 200
CLl NXIN 0 10PF
CL2 NXOUT 0 10PF
RFB NVOUT NVIN 22E6
RLD NXOUT 0 lE9

*Broad band analysis
.AC DEC 1000 lE6 300E6

*Narrow band analysis for Bode & Nyquist plots
*.AC LIN 1000 23.8E6 24.2E6
.PR OBE
```

Figure 4. Circuit listing for open loop analysis.

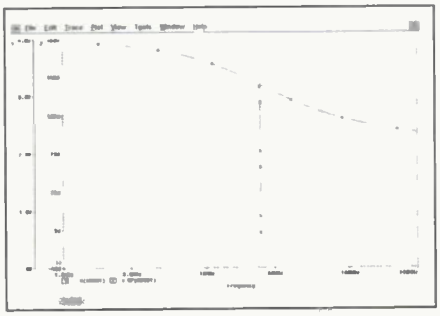

Figure 5. Broadband response.

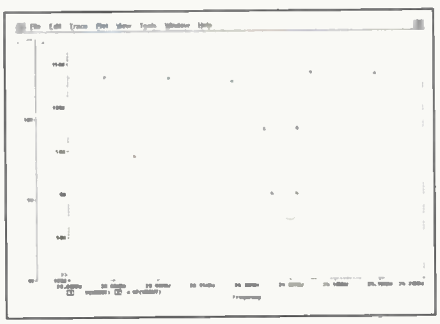

Figure 6. Narrowband response.

internal capacitance of the amplifier and help reduce the effect of amplifier and circuit variation. The feedback loop controls all frequency characteristics. The amplifier influences this frequency only by its gain, bandwidth, and parasitic capacitances, which are added to the load capacitors. We can control the loop gain by sizing the transistors and by selecting the proper ratio between the crystal load capacitors. Since the crystal inductance varies by a wide range near resonance, the exact value of the load capacitors are not critical. They just have to be small (less than 30 pF in a typical circuit) so as not to load down the crystal.

The crystal in this parallel resonant circuit operates in the inductive region between the resonant and anti-resonant frequencies (the crystal bandwidth). Outside this narrow frequency range the crystal looks like a small capacitor (C_P) electrically. Within this bandwidth, the crystal phase varies by a large range with very small change in frequency. In other words, extremely small frequency shift is necessary to change the crystal's impedance to compensate for phase deviation around the loop. The frequency pulling characteristic of the crystal is such that

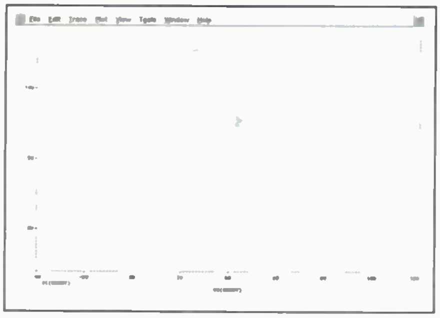

Figure 7. Nyquist plot of open loop response.

additional phase variation incurred in the circuit can be tolerated and is translated into a small frequency shift.

Startup Transient

Transient simulation of extremely high-Q quartz crystal circuits is a real challenge. Most designers rely on the frequency responses (small signal analysis) to judge the ability of their circuits to oscillate. Transient

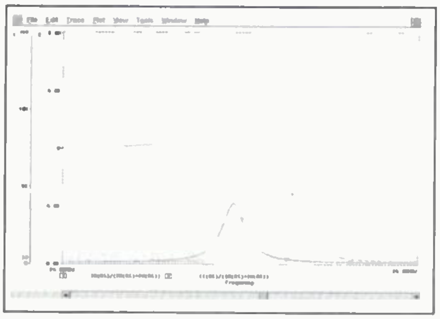

Figure 9. Frequency pulling.

response uses large signal analysis which involves nonlinearity operation, truncation errors and time step estimation which often results in divergence. However, a few tricks can make the oscillation waveform come alive. Figure 3 shows the transient response for the oscillator startup. Here the command, V99 99 0 PWL 0NS 5V .1US 5V .105US 0V .111US 5V 100MS 5V, introduces a voltage spike to the power supply to kick start the oscillator. Note that in a real circuit, oscilla-

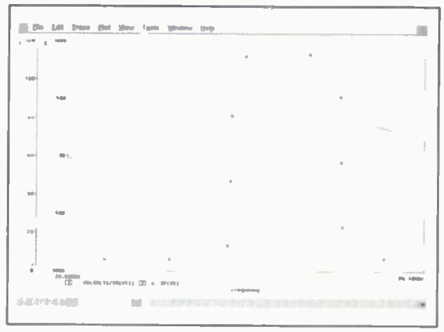

Figure 10 Crystal impedance.

tion is started by random thermal noise in the circuit elements. This noise is amplified by the inverter and is positively fed back through the crystal circuit. The amplitude of the signal increases exponentially as it goes through the loop again and again until saturation is reached. The frequency of the signal is controlled by the crystal. The crystal acts as a narrow band filter that passes only the resonant frequency. We should limit the fourth parameter of the .TRAN command

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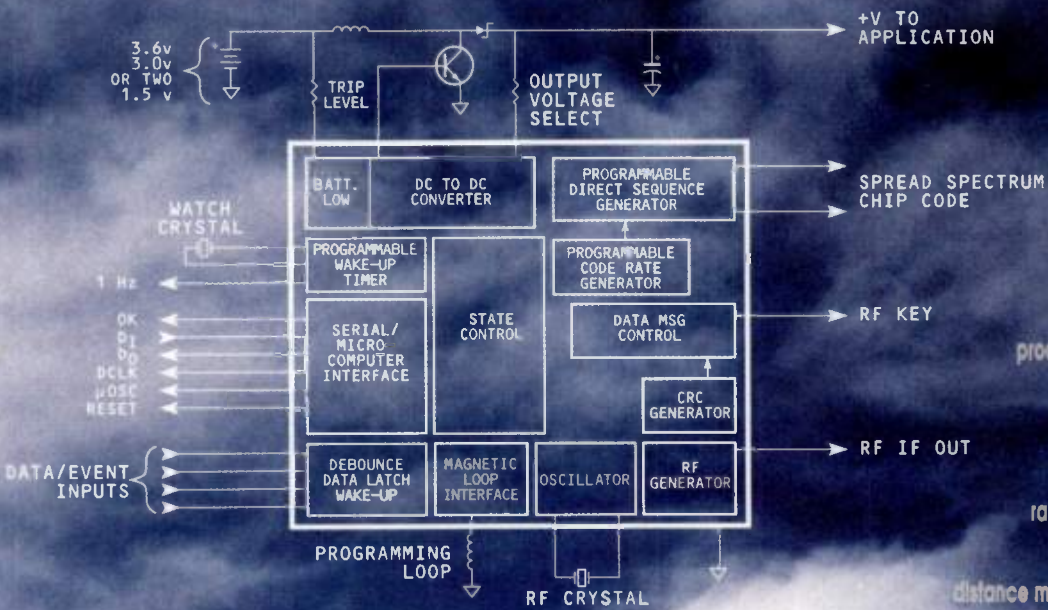

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```
Feedback Analysis

VIN IN 0 AC 1

RS IN 1 100
RM 1 11 10.2
LM 11 12 2. lE-3
CM 12 13 20.937E-15
CP 1 13 4.5E-12
CT 1 0 CMOD 1
CL 13 0 10E-12
RL 13 0 lE9

.MODEL CMOD CAP(C=lOPF)
.STEP CAP CMOD(C) 10PF, 20PF, 5PF

.AC LIN 1000 24.01E6 24.04E6
.PROBE

* Plotting xtal reactance (imaginary part)
* -II(RS) / (IM(RS)*IM(RS))
* Plotting xtal resistance (real part)
* IR(RS) / (IM(RS)*IM(RS))

.END
```

Figure 8. Circuit listing for feedback impedance.

(the maximum stepping time) to one-fourth of the oscillator period. This parameter would prevent the internal step of PSpice to overshoot the oscillator period and results in divergence. To avoid aliasing, we will choose the print interval at 1 ns.

The crystal mounting capacitance C_P also plays an important role in realizing the transient simulation. C_P is in parallel with the motional arm of the crystal and diverts the energy going through the crystal. This reduces loop gain greatly and makes startup much slower. For a typical 24 MHz crystal C_P is about 4.5 pF.

A 40 µs simulation would take about ten minutes on an 486 based personal computer. Examining the transient waveform would reveal a straight line, i.e. no oscillation. Zooming in the signal would reveal a very small (0.01 V) sine wave that does not grow exponentially as expected. As it turned out, at least 200 µs simulation time is needed to see the startup waveform. However, we would have to suppress the first 160 µs of data because the data buffer will be full after 40 µs. To completely show the transient startup from time zero to saturation, we can temporarily comment out C_P (or artificially reduce its value). Good startup waveform results with only 20 µs of simulation.

Open-Loop Response

It's relatively easy to obtain closed-loop Bode plots showing the magnitude and phase responses of the circuit with PSpice. All we have to do is to plot the waveform at the output of the oscillator. However, it's more tricky to obtain the open loop response. It's the open loop response that shows the gain margin and phase margin critical to oscillation.

One common method is to use a large-value capacitor and inductor to break the loop. Another method is to open the feedback loop and reflect the impedances at the break points. A voltage generator can then excite the circuit for open loop measurement.

Whereas the first method can introduce improper loading to the circuit, the second method significantly increases the circuit size and run time. In this circuit we will attempt to open the loop by inserting a 1V AC pseudo voltage source between the output node of the crystal and the input node of the CMOS amplifier. Since the amplifier has almost infinite input impedance and since the ideal voltage source has zero series impedance and infinite parallel impedance, the addition should not disturb the loading of the circuit. The DC level of the voltage source is the DC operating point of the circuit obtained from the transient analysis output file. The command, VPS NVIN 0 DC 2.557 AC 1, accomplishes this task. Figure 4 lists the modified circuit for open loop analysis.

To be certain that no unwanted oscillation modes can occur in the frequency range of interest, we have to consider both broadband and narrowband analyses Figure 5 illustrates the broadband response of the circuit with the command, .AC DEC 1000 lE6 300E6. Zooming in, the spike in the waveform that indicates the frequency of oscillation will not reveal the detail of the response. Instead, we have to sweep the frequency linearly in the immediate vicinity of the resonant frequency to show the amplitude peaking and the zero-degree phase-crossing at the series (resonant) frequency and parallel (anti-resonant) frequency.

Figure 6 illustrates the narrow band response of the circuit with the command, .AC LIN 1000 23.8E6 24.2E6. As expected, the loop gain peaks at f_1 where the loop phase shift crosses through zero degree. The crystal resonating with C_P and the load capacitors creates a complex pole pair that causes this behavior. At f_2 the loop gain drops below unity and the loop phase once again crosses through zero degrees. f_2 is the parallel resonant frequency formed by the crystal resonating with C_P to form a complex zero pair and a high impedance in the feedback circuit.

The oscillator's phase margin is the maximum angle past zero degrees that the loop phase shift has between f_1 and f_2. The phase margin gives a measure of how much loading the oscillator can withstand and still start-up. If the loading on the oscillator is too much, the phase will not pass through zero degrees and the oscillator will not start. Similarly, the oscillator's gain margin is the maximum magnitude past unity that the loop gain has at the zero-degree crossing phase. The larger the gain margin, the more loss the feedback loop the oscillator can withstand.

Using Probe, we can also obtain a Nyquist plot of the combined amplitude and phase responses by plotting the imaginary part of the crystal output voltage versus its real part using linear frequency scaling. Figure 7 shows that the Nyquist plot encircles the point (1,0) which indicates oscillation for a positive feedback circuit. The Nyquist plot uses the same data file as that of the narrowband plot.

The mathematical signal processing ability of Probe is very useful when it comes to plotting the impedance of the feedback loop. Figure 8 is the new circuit file used for this purpose. The circuit consists of only the feedback loop which contains the series resistor, the crystal and the load capacitors. A large value resistor parallel to the load capacitor provides the DC path to ground. To simplify the analysis, we use a 1V AC voltage source for excitation. From the loop voltage and current expressions and after some complex algebra manipulation, the feedback impedance expressions in Probe become:

mag(Z) = 1 / (IM(RS))
phase(Z) = –IP(RS)
real(Z) = IR(RS) / (IM(RS)·IM(RS))
imag(Z) = –II(RS) / (IM(RS)·IM(RS))

In a series-resonant oscillator, the crystal operates at its natural series frequency. In a parallel resonant oscillator, the feedback circuit introduces a load capacitance to the crystal. In this case, the crystal operates at the frequency where the crystal reactance cancels the load capacitor reactance. Therefore when the load capacitance changes the resonant frequency also changes. We can examine the frequency pulling characteristic of the crystal oscillator by performing parametric

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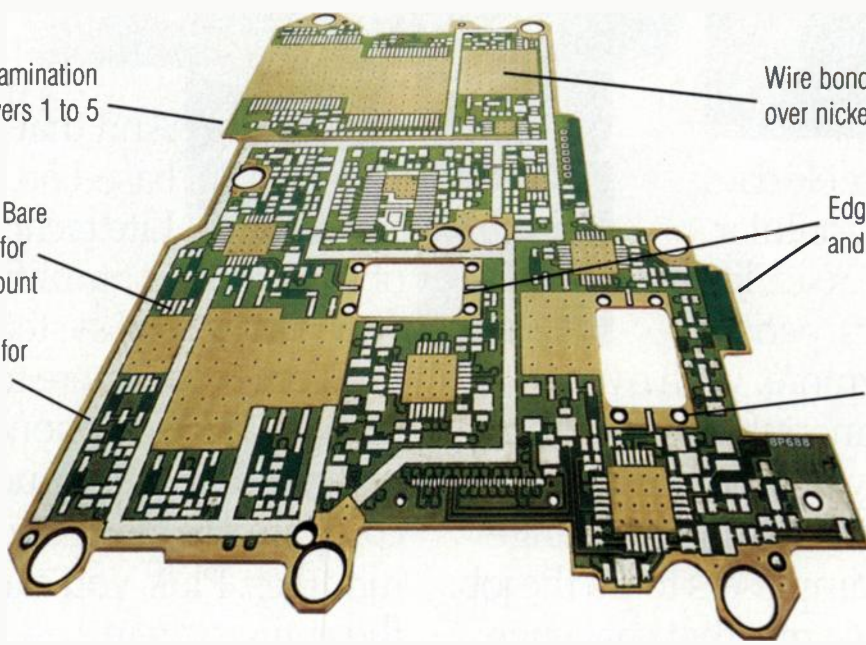

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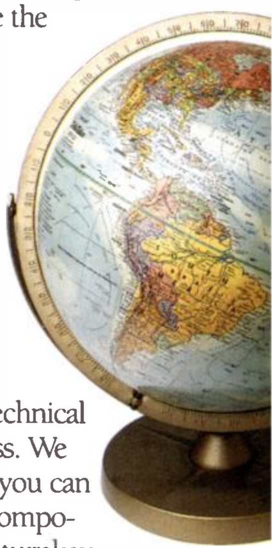

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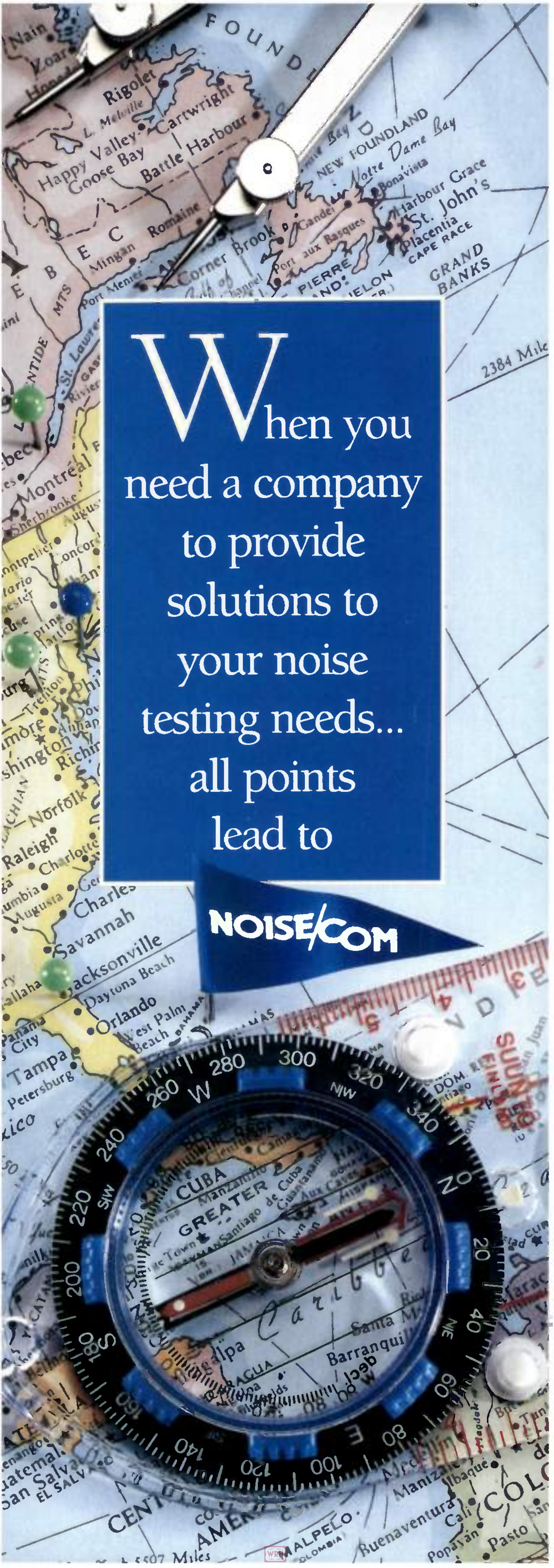
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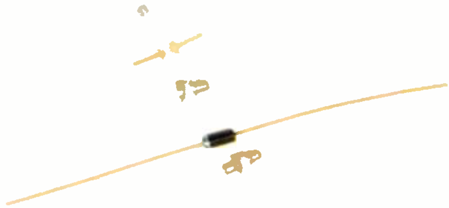

TYPICAL STANDARD MODELS	
MODEL	**FREQUENCY RANGE**
NC 302	10 Hz – 3 GHz
NC 305	10 MHz - 11 GHz
NC 401	100 MHz – 18 GHz
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MODEL	**FREQUENCY RANGE**	**OUTPUT ENR**
NC 501/15	0.2 MHz – 500 MHz	31 dB
NC 502/15	0.2 MHz – 1000 MHz	31 dB
NC 503/15	0.2 MHz – 2000 MHz	31 dB
NC 506/15	0.2 MHz – 5 GHz	31 dB
NC 511/15	0.2 MHz – 500 MHz	51 dB
NC 513/15	0.2 MHz – 2 GHz	51 dB

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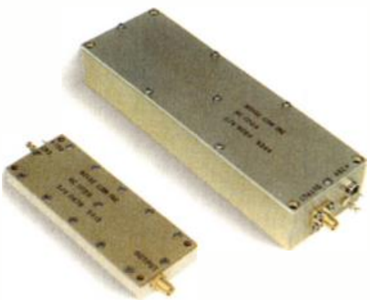
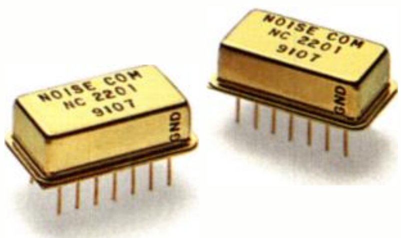
Broadband Amplified Modules



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

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 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 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 also available.

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Broadband Precision, Calibrated Coaxial

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 346C	0.01 GHz – 26.5 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 5122	33 GHz – 50 GHz	WR-22
NC 5115	50 GHz – 75 GHz	WR-15
NC 5110	75 GHz – 110 GHz	WR-10

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 6108	100 Hz - 500 MHz	+10 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
NC 8111	5 MHz - 1 GHz	+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 by remotely using the IEEE-488 interface.

TYPICAL STANDARD MODELS		
MODEL	FREQUENCY RANGE	OUTPUT POWER
UFX-7107	100 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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analysis on the input load capacitance over a wide range. Figure 9 shows that the range of frequency the crystal can pull to either side of its nominal value is extremely narrow. Note that the reactance curve crosses zero at two frequencies which represent the series resonant and parallel resonant. Between the two frequencies, the reactance is inductive. Below and beyond that range, the reactance is capacitive. The oscillator operates in the inductive region between the two resonant frequencies.

The impedance of the crystal itself can also be plotted by commenting out the series resistor and the load capacitors.

Inserting a zero-volt AC voltage source as a current meter in series with the crystal and again use a 1V AC source for excitation, the magnitude and phase of the crystal impedance are expressed as follow:

mag(Z) = DB (VM(1) / IM(VS))
phase(Z) = –IP(VS)

Figure 10 shows the resulting plot. As expected, the response is similar to that of the whole feedback network. The phase of the crystal impedance crosses zero at the series and parallel frequencies. Note that the frequencies of minimum and maximum impedance differ slightly with the series and parallel frequencies. These frequencies would be the same if the motional resistance of the crystal $R_M = 0$. The crystal controls the frequency response of the feedback and the load capacitors only influence this response slightly

Conclusion

We have shown that using transient response, gain and phase margin analysis from the open loop response, the feedback network reactance plot, and the Nyquist plot, we can be confident that the embedded oscillator can reliably start-up and oscillate. We can simulate the crystal operation at an overtone by adding the harmonic motional arm in parallel with the fundamental motional arm of the crystal. A note about temperature analysis is that the cutting angle of the crystal blank governs the performance of the crystal over the temperature range. Most AT-cut quartz crystals have a quadratic temperature dependence. We can model this dependency by giving the motional inductor a quadratic temperature coefficient TC2. Also note that the FET transistors already have their temperature dependencies built into the models of these devices. *RF*

About the Author

Kien Truong graduated from Iowa State University in 1985. he is now a design engineer in Integrated Avionics Research at Boeing. Hie experience ranges from antenna and radar to fiber optic sensors and fault tolerant computing. Kien enjoys sailing, telemarking and family outings. He can be reached at Boeing Commercial Airplane Group, P.O. Box 3707 M/S 6H-EC, Seattle, WA 98124.

Circuit Envelope Simulator Analyzes High-Frequency Modulated Signals

By Andy Howard
HP EEsof

Modern communication systems handle RF signals with complex digital modulation. The most commonly used simulators for RF circuit design, SPICE and harmonic balance, have serious shortcomings when simulating such modulated signals. This article describes the new circuit envelope simulation technology and its advantages over SPICE and harmonic balance.

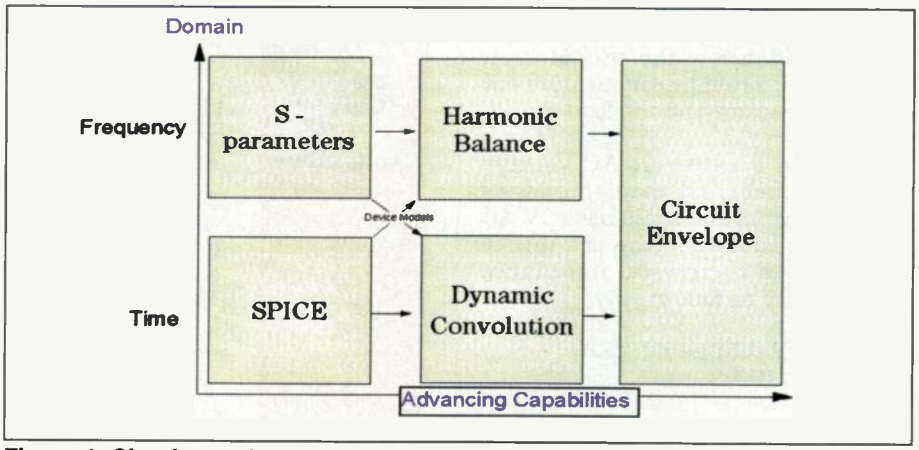

Figure 1. Circuit envelope technology extends HP EEsof's simulation solutions, allowing simultaneous use of time- and frequency-domain techniques.

The HP 85148A Circuit Envelope simulator uses a patent-pending hybrid technology, which combines the advantages of time-domain and frequency-domain techniques. Unlike SPICE or harmonic balance, this technology is best suited for circuits with transient or modulated RF or high-frequency carriers. This new technology allows signals to be represented as a few carriers with time-varying modulation "envelope(s)." The Circuit Envelope simulator allows a designer to see how a circuit affects the modulation directly, and see such things as the spectrum of a modulated signal around a carrier, as well as amplitude, phase, and frequency versus time waveforms. Typical applications include analyzing the spectral regrowth or adjacent-channel power leakage generated by an amplifier or mixer circuit, oscillator turn-on amplitude and frequency versus time, automatic level control loop transients, phase-locked loop transients, and subsystem simulation with digitally modulated carriers.

SPICE

SPICE is the circuit simulator most commonly used by analog designers today, but it has some serious drawbacks when simulating transient and modulated RF signals. Because it is a time-domain simulator, it samples user-defined input waveforms and then solves the circuit at successive time steps. For SPICE simulations, three parameters determine the overall simulation time: the time step, the time required for any initial transient response to die out, and the number of periods of the lowest frequency signal that must be simulated. In most communication systems, the RF signal frequency (1-2 GHz) is much higher than the modulation frequency bandwidth (often <200 kHz). Unfortunately, SPICE treats a modulated RF signal the same as any time-domain waveform, with step size determined by the RF carrier frequency (or its harmonics), and the stop time determined by the number of periods of the modulating signal. This can lead to a very large number of simulation steps (see sidebar example), as well as a large amount of memory for storing the results and post-processing them for display. Another drawback of SPICE simulation is the significant post-processing required to extract the modulation envelopes (or I and Q data) usually desired when simulating digitally-modulated RF waveforms.

Dynamic Convolution

Another SPICE limitation is that it does not handle the frequency-domain models commonly used in high-frequency circuit analysis, instead using lumped element models or transmission lines with a pure time delay. To overcome this problem, dynamic convolution simulators were invented. They allow complex, frequency-domain models and measured data to be used in nonlinear, time-domain circuit simulations. However, the modulated RF signal is treated the same way as SPICE, without distinguishing between the modulation and the RF carrier. Therefore, simulation times for RF modulated signals are as slow as with SPICE.

Harmonic Balance

Harmonic balance analysis is best suited for simulating the steady-state behavior of nonlinear circuits. However, most communication systems now use digital modulation, which is pseudo-random and cannot be considered steady-state. These modulation waveforms can be modeled with harmonic balance, but only if they are modeled as periodic bit sequences of limited (32 bits or less) length. Harmonic balance models these modulated waveforms with a fundamental tone and a very large number of its harmonics, therefore requiring an excessive amount of memory (and time) to simulate even a short bit sequence.

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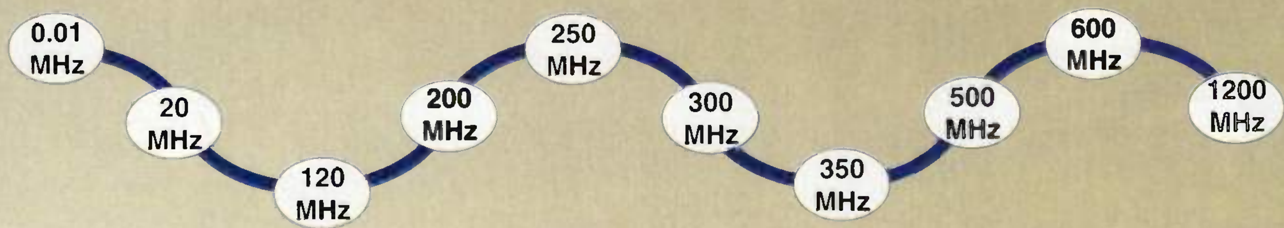

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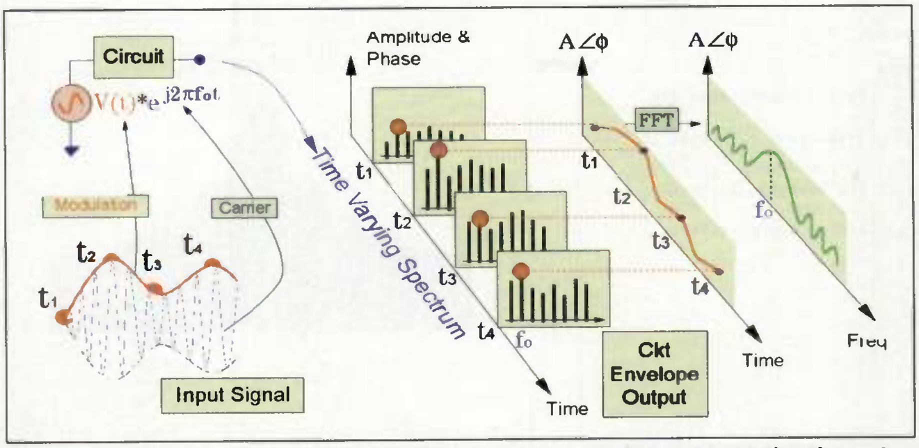

Figure 2. Circuit envelope technology allows the designer to set the time step just small enough to track the modulation envelope.

While harmonic balance has some problems analyzing complex modulated signals, it does have some advantages over SPICE. Harmonic balance is significantly more efficient for the analysis of circuits with widely varying frequency components, such as mixers, as long as a large number of harmonics is not required. Linear devices with arbitrary frequency responses can be easily modeled. At RF and microwave frequencies, many devices cannot be modeled with simple differential equations, and often must be characterized with measured, frequency-domain data. Harmonic balance directly provides the steady-state solution, without having to wait until the transient response dies out.

S-Parameter Analysis

S-parameter analysis is commonly used for RF and microwave circuit design, but it is only applicable for linear analysis. It is not applicable for simulating transient or modulated RF signals in nonlinear circuits.

Circuit Envelope Technology

The Circuit Envelope simulator overcomes the limitations described above by using a combined time- and frequency-domain technology. With this technique, the input waveforms are represented as RF carriers with modulation "envelopes" that are described in the time domain. The solution involves finding the Fourier coefficients, $V_k(t)$, in the following equation:

$$v(t) = \text{real}\left[\sum_{k=0}^{N} V_k(t) e^{j\omega_k t}\right]$$

Here, v(t) is a voltage at any node in the circuit, including the input. In standard harmonic balance, the Fourier coefficients are complex constants, whereas with the Circuit Envelope simulator, they are allowed to vary with time. The may represent an arbitrary modulation of each of the carriers. This spectrum may represent transient signals with continuous spectra (such as a digital modulation envelope), or periodic signals with discrete spectral lines (i.e., to generate two RF tones required for intermodulation distortion analysis).

Figure 2 shows how the envelope is represented in the simulator, for the simple case of amplitude modulation. The simulator computes the instantaneous envelope versus time of each spectral component. This allows the computation and display of the amplitude, phase, I, Q, or frequency versus time for any component. Computing the Fourier transform of a time-varying spectral component allows the designer to see the spectrum around a spectral component, similar to a spectrum analyzer display.

Improvements Over SPICE

Simulation efficiency is much better, because the time step is determined by the bandwidth of the modulation, which is relatively narrow, not by the frequency of the RF carrier or its harmonics.

Simulation accuracy is improved. With simulators such as SPICE, the accuracy of the numerical integration routine degrades as the signal frequency approaches the sampling frequency. With the Circuit Envelope simulator, multiple integrations are being performed, each one centered on a carrier, so the highest accuracy is achieved at each signal frequency of interest. The accuracies at the carrier frequencies are as good as those of the basic harmonic balance simulator.

The instantaneous amplitude and phase information (or I-Q signals) are

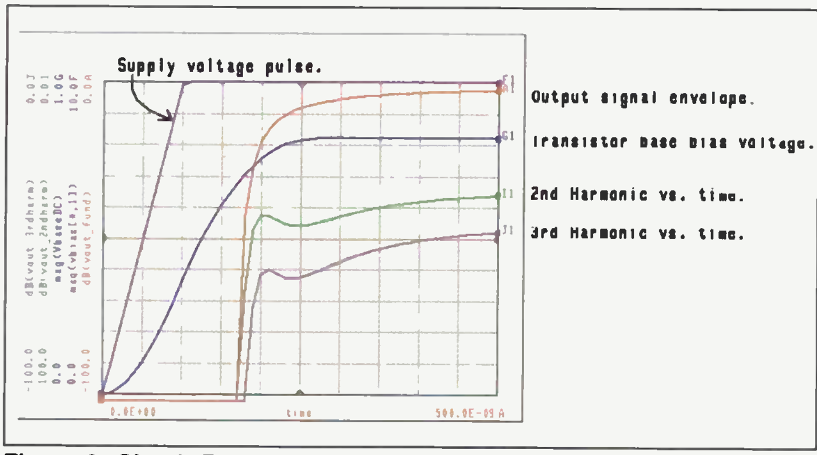

Figure 3. Circuit Envelope simulator predicts the output harmonics of an amplifier plotted versus time. This allows designers to make trade-offs between conserving battery life and output distortion.

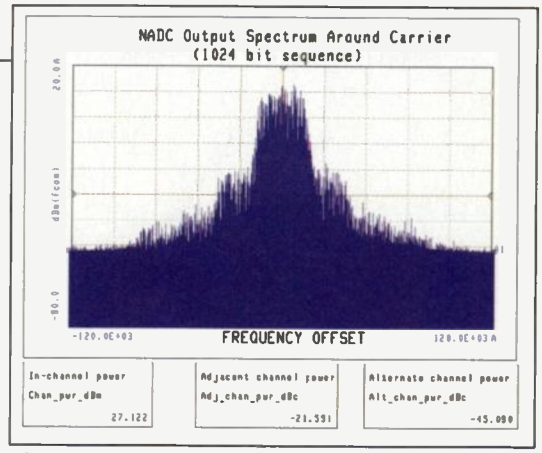

Figure 4. Simulation output showing spectral regrowth around the RF transmission carrier of an NADC power amplifier.

immediately available. This gives the Circuit Envelope simulator the ability to output and/or use such information as signal frequency versus time.

Frequency-domain models are more effectively and accurately incorporated. Circuit Envelope has the advantages of harmonic balance for modeling devices and circuit elements in the frequency domain.

With the Circuit Envelope simulator, the user has the option of starting a simulation from the steady-state solution, or examining turn-on transient responses.

Advantages Over Harmonic Balance

The Circuit Envelope simulator is much more efficient than harmonic balance for simulating digitally-modulated carriers because the modulation waveform is described in the time domain rather than via a fundamental tone and harmonics. This leads to much faster simulations requiring only a fraction of the computer memory.

Applications

The Circuit Envelope simulator is applicable for simulating high-frequency amplifiers, mixers, oscillators, and sub-systems with transient or modulated RF signals. The smaller the modulation bandwidth relative to the RF carrier frequency, the greater the improvement in simulation efficiency over SPICE. The Circuit Envelope simulator is also an excellent tool for simulating transient responses of circuits such as AGC loops and phase-locked loops, where a low-frequency transient signal must be simulated along with a high-frequency RF signal. These applications include:

- Amplifier spectral regrowth and adjacent-channel power leakage, with digitally-modulated RF signals as the input.
- Oscillator turn-on transients and frequency output versus time, in response to a transient control voltage.
- Phase-locked loop transient responses.
- AGC and ALC transient responses.
- Circuit effects on signals with transient amplitude, phase, or frequency modulation.
- Amplifier harmonics versus time.
- Sub-system analysis with modulation signals such as multi-level FSK, CDMA, TDMA.
- Much more efficient analysis of mixer and amplifier third and higher-order intercept points.
- Time-domain optimization of transient responses.

Amplifier Harmonics Versus Time

A designer of amplifiers for mobile communications gear might be interested in plotting amplifier harmonics versus time. To maximize battery life, it is necessary to switch on the power supply to the amplifier just before transmitting. There must be a delay between switching on the power supply and transmitting information, because if the amplifier is not completely on, it will generate excessive harmonics. The Circuit Envelope simulator allows designers to simulate the transient bias supply voltage(s) and see how long it takes the voltages at the device to reach steady-state. After an appropriate delay, the signal can be pulsed on, and the envelope of the output harmonics versus time may be plotted. One such simulation result is shown in Figure 3. The time step can be set just small enough to adequately model the variation in the envelopes of each spectral component.

Amplifier Spectral Regrowth with an NADC1 Input Signal

Spectral regrowth and adjacent channel power leakage are critical

Comparing SPICE and Circuit Envelope Technology

The following compares the time step size and the number of time steps required to simulate an amplifier with an input signal consisting of an 890 MHz RF carrier modulated with a 24.3 kHz symbol rate (NADC modulation). Assume that 512 symbols must be simulated. The following table compares the difference in time step size and the number of time steps.

	Time Step Size	Number of Time Steps
SPICE	62.4 picoseconds	168.8 million
Circuit Envelope	4.1 microseconds	5,139

This assumes the circuit generates significant 3rd harmonics, which means that to satisfy the Nyquist sampling criterion, a sampling rate of at least 5.34 GHz must be used. To obtain reasonable accuracy from SPICE's integration algorithms, a sampling frequency of 16 GHz (time step = 1/(16 GHz) = 62.4 ps) is a more realistic minimum. The simulation stop time is 512/(24.3 kHz) = 21.07 ms. Circuit Envelope does require a vector of complex voltages to be computed at each time point, rather than just a single real value, but the simulation time will still be orders of magnitude faster than SPICE.

specifications for digital communication system designers. Digital modulation is characterized with a power spectral density, whereas analog modulation is characterized with discrete spectra [1]. Until now, designers have characterized amplifier nonlinearities via simulations and measurements of third-order (and higher) intercept points, generated from two unmodulated sinewave input signals. Unfortunately, these signals do not represent digitally-modulated waveforms very well. Consequently, these simulations and measurements do not always predict spectral regrowth and adjacent channel power performance [2]. The Circuit Envelope simulator overcomes the limitations of SPICE and harmonic balance when simulating such waveforms, as described above.

Figure 4 shows the simulated output spectrum resulting when a signal with NADC modulation (1024-bit sequence) is sent through a power amplifier. The output signal's adjacent and alternate channel power ratios are calculated by the post-processor, and eye diagrams and constellation plots may be generated, as well. This simulation required less than 14 minutes and less than 4 MBytes of virtual memory on an HP 9000 Series 715/100 workstation. The simulation may be run with different input signal power levels.

Mixer Nonlinearity Simulations

Mixer nonlinearities are often characterized via a third-order intercept (TOI) simulation, in which two closely-spaced sinewaves (tones) are input at the RF port. These simulations are very time-consuming in SPICE because of the large difference between the beat frequency (the spacing between the two RF tones) and the RF frequencies. Harmonic balance simulations can be very time and memory intensive because three large-signal tones must be modeled. The Circuit Envelope simulator allows a much faster simulation with only a fraction of the memory requirement of harmonic balance, by modeling the two input tones as a single RF tone, with cosine modulation. The cosine modulation splits the RF tone into two signals spaced at twice the modulation frequency:

$$V_{rf}(t) = \mathrm{real}\left[2\cos(2\pi f_m t)e^{j2\pi f_{rf} t}\right]$$

$$= \mathrm{real}\left[e^{j2\pi(f_{rf}+f_m)t} + e^{j2\pi(f_{rf}-f_m)t}\right]$$

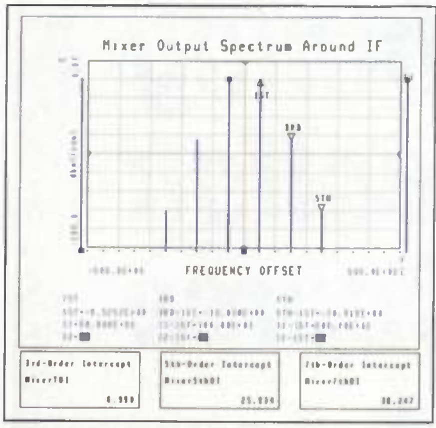

Figure 5. Mixer output spectrum around IF, with computed 3rd, 5th and 7th order intercept points.

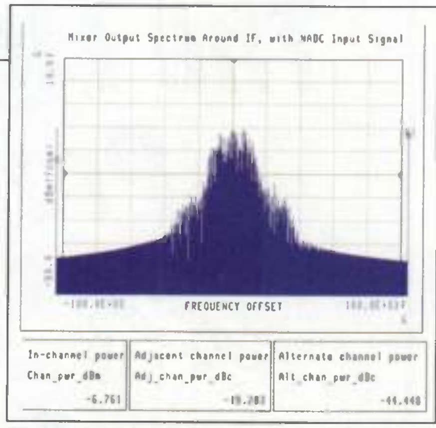

Figure 6. Gilbert cell mixer output spectrum around IF, with NADC modulation on input signal.

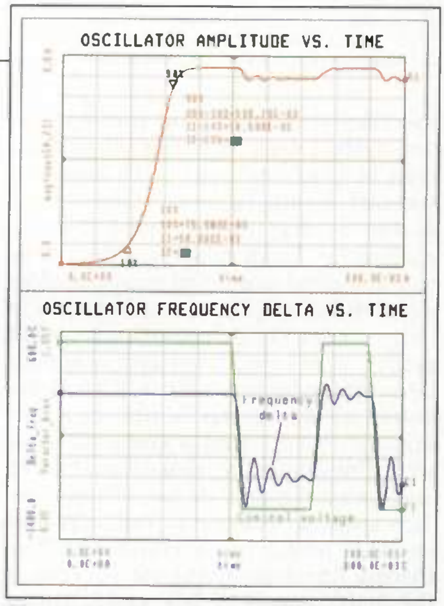

Figure 7. A crystal oscillator's turn-on transient, and its relative frequency in response to a control voltage waveform are shown.

The simulation is run over a full period of the modulation signal, and the IF spectral component is computed as a function of time. Taking the discrete Fourier transform of the IF spectral component versus time gives the intermodulation distortion components, as shown in Figure 5. This simulation required 64.5 seconds and 18.4 MBytes of virtual memory, on an HP 9000 Series 715/100 workstation. The mixer analyzed has 14 BJTs.

Of course, using two sinusoids at the input may not adequately characterize the mixer's behavior when it converts digitally-modulated waveforms. The Circuit Envelope simulator allows a designer to simulate a mixer's contribution to adjacent-channel power leakage and spectral regrowth (Figure 6).

Oscillator Simulations

The Circuit Envelope simulator allows very efficient simulation of oscillator turn-on time, amplitude versus time, and frequency versus time. In the crystal oscillator example shown in Figure 7, the 10-90% turn-on

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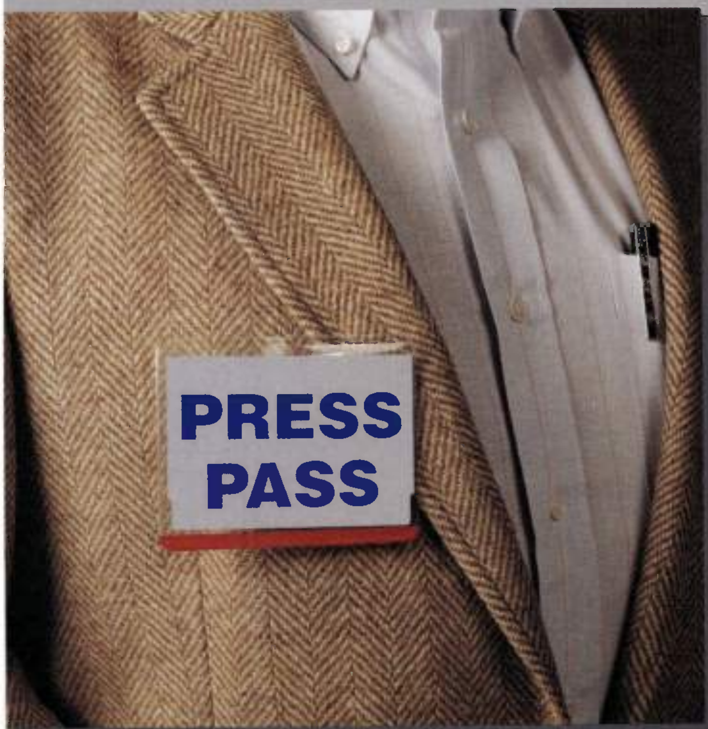

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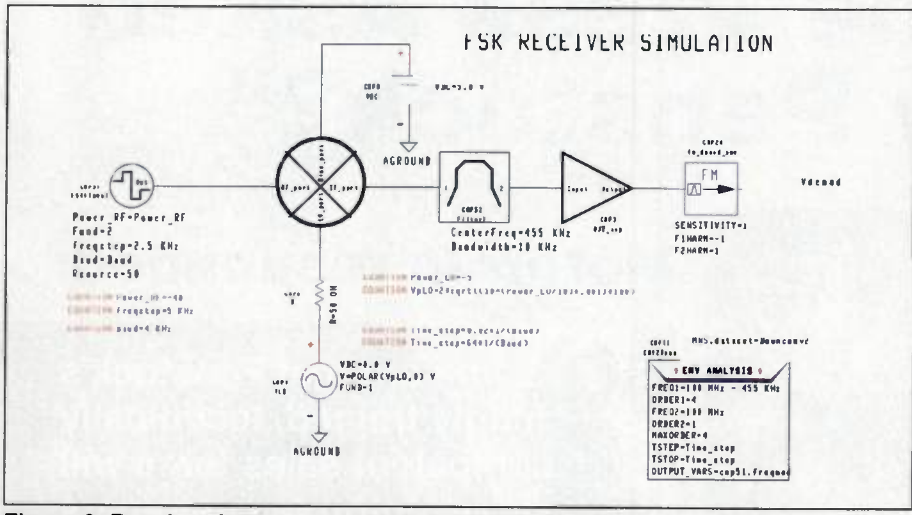

Figure 8. Receiver for four-level FSK demodulation using the mixer of Figure 5.

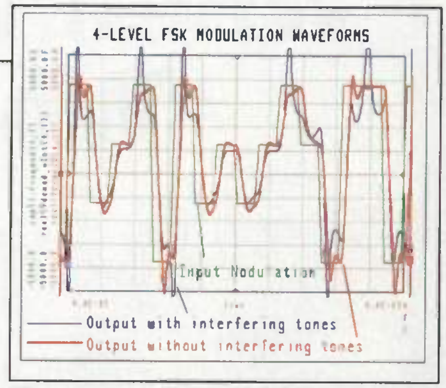

Figure 9. Four-level FSK input and output frequency deviation versus time modulation waveforms are shown.

time is approximately 38.5 milliseconds. It is necessary to simulate from 0 to 150 ms to see the full turn-on transient. The Circuit Envelope simulator can generate the turn-on envelope even with a 5 ms time step (a 0.5 ms time step was used in Figure 7.) This means that only 31 time points are required. SPICE would require approximately 150 million time points, based on an oscillation frequency of 100 MHz and a time step equal to one tenth the period of oscillation.

It is also possible to simulate the oscillator's frequency versus time, in response to a control voltage input, also shown in Figure 7.

Frequency-Shift Keying (FSK) Simulation

The Circuit Envelope simulator can be used to model FSK (and other modulation format) receivers. The designer can see how the demodulated waveform depends on filter bandwidths, modulation waveform shape, frequency step size, interfering signal strength, and noise. Figure 8 shows a simple receiver, and Figure 9 shows the input modulation and the demodulated waveform, with and without adjacent channel interfering tones.

ALC Loop Simulation and Optimization

In ALC circuits, there is usually a high frequency RF signal and slowly-varying DC voltage that is proportional to the power or voltage of the output RF signal. In this example, shown in Figure 10, an amplitude modulator is used to adjust the output signal power. A portion of the output signal is coupled off and converted to a DC voltage by a diode detector. The difference between this DC voltage and a reference voltage is integrated and becomes the input to the amplitude modulator. The output power can be set by changing the reference voltage. The designer is interested in seeing how the output power envelope changes versus time in response to a change in the reference voltage.

In this case, the reference voltage is varied in a piecewise linear way to step the ouput power from 4 dBm to 8, 12 and 16 dBm. One of the traces in Figure 11 shows significant overshoot. The other trace is the output power versus time waveform after time-domain optimization of the detector output capacitor and resistor, integrator resistor and capacitor, and the reference voltage.

Phase-Locked Loop Simulation

Phase-locked loops (PLLs) can be simulated with the Circuit Envelope simulator.This example uses a PLL that could be used to generate the LO frequencies required for the Digital European Cordless Telephone (DECT) standard [3]. The VCO, divide-by-N block, and phase/frequency detector are all simulated using behavioral models. The simulated frequency-versus-time waveform in Figure 12 shows how quickly the VCO frequency settles when the value for N is increased from 1025 to 1026. Figure 13 shows the noise spectrum of a VCO with the PLL open and closed. This PLL did not have a divide-by-N block.

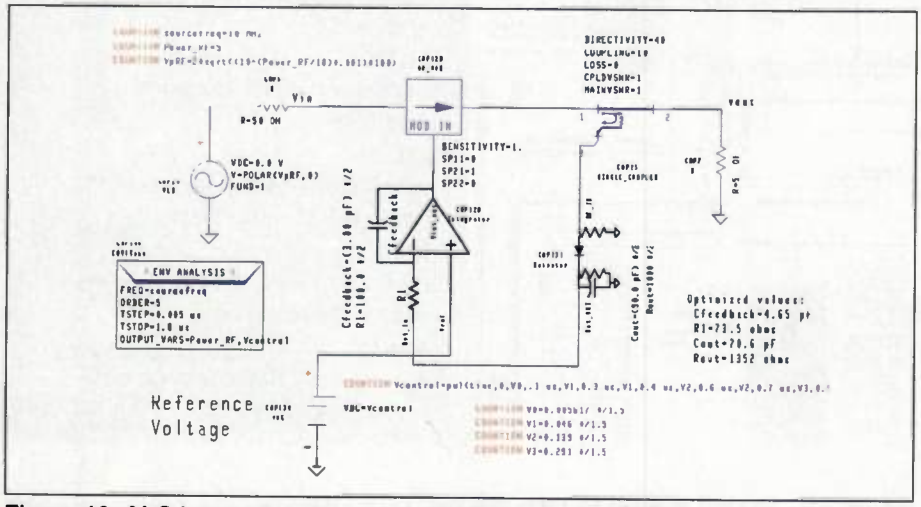

Figure 10. ALC loop setup for transient response simulation.

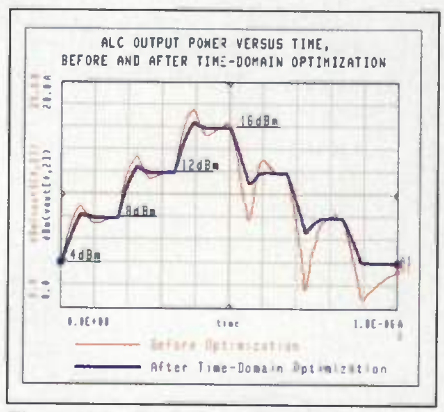

Figure 11. ALC output power versus time, before and after time-domain optimization to minimize overshoot.

Summary

The Circuit Envelope simulator uses a hybrid technology that combines the advantages of time-domain and frequency-domain techniques. This technology overcomes the limitations of SPICE and harmonic balance when simulating circuits with transient or modulated RF or high-frequency carriers. Many applications showing the new simulation capabilities (amplifier and mixer spectral regrowth, oscillator turn-on transient and frequency versus time, ALC transient response, time-domain optimization, efficient mixer TOI, FSK demodulation with and without interfering tones, PLL simulation, etc.) have been included in this article.

Readers desiring more information can contact HP EEsof at (818) 879-6440, or circle Info/Card #250. **RF**

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1. John Sevic, "An Investigation of Nonlinear Analysis Methods for Simulation of Digital Wireless Communication Systems," CAD Design Methodology for Commercial Applications Workshop, 1995 IEEE MTT-S International Microwave Symposium.
2. Dan Pleasant, "Designing Circuits for Wireless Applications," Hewlett Packard Wireless Communications Design Seminar, 1995.
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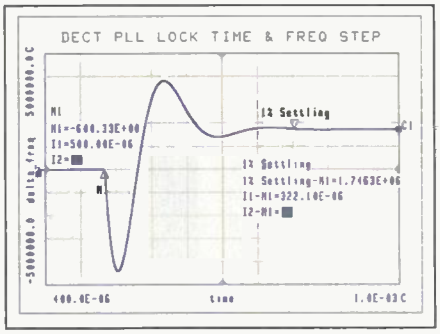

Figure 12. Simulation of the PLL's frequency-versus-time in response to one step in the divide-by-N block's value of N.

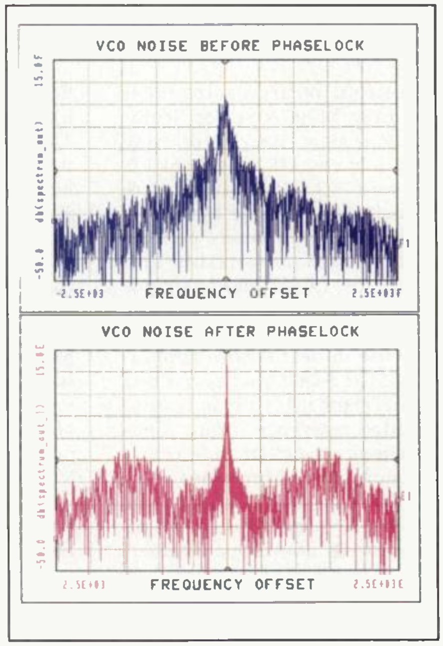

Figure 13. Noise spectrum of a VCO with and without phase lock (using a different PLL without a divide-by-N block).

Network Analyzers

Anritsu Wiltron introduces the new 54000A Series Scalar Measurement Systems offering very low cost RF models to 1.0, 2.0 and 3.0 GHz. Microwave models are available to 8.6 and 20 GHz. The analyzers perform traditional transmission and return loss (SWR) measurements for manufacturing applications. A rugged housing allows reliable use in field service applications. The 54000A series has two inputs standard, with an optional third reference input. Measurement modes include transmission (dB), return loss (dB), SWR (linear SWR), and power (dBm). Dynamic range is –55 dBm to +16 dBm. Source resolution is 10 kHz for RF models, 100 kHz for microwave models. Thirteen sets of front-panel setups and thirteen sets of trace memory can be stored in non-volatile instrument memory. The instrument can perform min/max hold, various cursor functions, and an automated compression test routine. A GPIB interface is standard. The 54000A offers: low harmonic sources, linearized YIG tuning for stable sweeps, 50 ohm and 75 ohm measurement systems, portable package, 10,000 hour MTBF. Starting price for the series is $9,870

Anritsu Wiltron Co.
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Surface Mount 900 MHz Filter

Kel-Com bandpass filter model number SB-900/200 has an insertion loss of less than 2.0 dB and VSWR of 1.5:1 over the passband of 800 to 1000 MHz. A 1.0 dB per MHz flatness is also specified over the passband. Minimum

stopband rejection is 20 dB at 1170 MHz and 45 dB from DC to 605 MHz. Ultimate rejection is 70 dB from 1969 to 2184 MHz. Outline dimensions are 1.10 x 0.60 x 0.27 with plated through holes for surface mounting.

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A new line of quick-disconnect RF connectors for sputtering, test, and communications equipment requiring a high integrity connection with fast on/off capability is being introduced by Tru-Connec-

tor. The connectors mate reliably without bayonets, screws or tools. They are available in three types: QDL corresponding to an LC connector, QDS which has the same electrical properties as C and SC types, and QDM which is compatible with BNC and TNC connectors. Featuring positive locking mechanisms employing spring-loaded sleeves on the male that are drawn back to let self-contained balls "click" into grooves on the corresponding females and then slide forward, Tru-Connector Quick Disconnect RF Connectors can be supplied in straight-through and right angle designs. In-between series adapter are also available.

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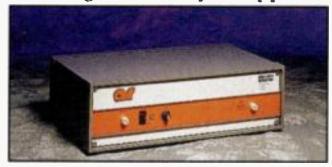
5 Watt Broadband Amp

Model 5S1G4 from Amplifier Research is an all solid-state broadband microwave amplifier providing 5 watts minimum power over two octaves, from 1 to 4.2 GHz. The full bandwidth is instantly available without bandswitching or tuning and permits sweep tests that would typically require two or more amplifiers to cover the same frequency range. The 5S1G4 is rated 5 W (minimum), measured at the output connector. It can be used in test applications which demand no less than 5 watts of RF power over the full bandwidth. An input attenuator with a linearizing circuit results in uniform gain adjustment over a 10 dB (minimum) range. Typical flatness is ±1.0 dB. Designed to operate Class A, the Model 5S1G4 uses GaAs-FET amplification stages, resulting in delivery of approxi-



mately 99 percent of output power at the fundamental or carrier frequency and very little output power in the form of harmonic or other noise content. Maximum harmonic distortion is –20 dBc at 3.0 W IP3 is typically 43 dBm. Price is $7,500.

Amplifier Research
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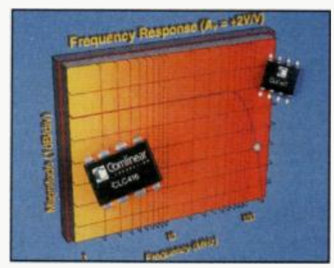
Dual Current-Feedback Op Amps

Comlinear Corporation has announced the CLC416 and CLC417 dual high-speed current-feedback operational amplifiers. Priced at $2.39 in 1000 piece quantities, the CLC416 and CLC417 have been designed to complement their single-version siblings (the CLC405 and



CLC407). The dual amplifiers offer 110 MHz small-signal bandwidths, very high 6 MΩ, very low 100 nA non-inverting input single bias currents, high 60 mA output current drive and very low quiescent current of 3.9 mA per amplifier. Differential phase is 0.03° for both amplifiers; differential gain is 0.01% for the CLC416, and 0.03% for the CLC417. The CLC417 provides internal feedback and gain setting resistors for ±1 and ±2 V/V gain selections. The dual amplifiers come in an eight-pin package with standard dual pinouts.

Comlinear Corporation
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EMI Filter & Seal

EESeal™ EMI filters and transient suppressors are installed simply by pushing the EESeal over the tips of the pins of the connector to be protected. Made of resilient silicone rubber, these filters easily conform to connectors, even if they have misaligned pins. Filters can be made to custom specs, and seals are available for virtually every connector type.
Metatech Corp.
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Surface Mount EMI Filters

The NFM60R series is an EIA standard 1206 size version of the NFM61R. Its large rated current (6A) and low voltage drop due to small DC resistance (typ. 3 - 4 mΩ) are suitable for DC power line use. The feedthrough capacitor provides excellent high-frequency characteristics. Both flow and reflow soldering methods can be used with the series.
Murata North America
INFO/CARD #244

EMI Filters for Video

The ZJSM series of surface-mount EMI filters from TDK are matched to 75 ohms and have flat passband characteristics, making them well suited to filtering applications for video lines. The series uses high-Q choke coils, obtaining sharp cutoff characteristics and high attenuation through 1000 MHz. Cutoff frequencies of 15, 25, 100, 150, and 220 MHz are available.
TDK Corp. of America
INFO/CARD #243

1500 V_{rms} Isolation

Even with their small footprint and low profile, Coilcraft's new CMB Series of surface mount filters are rated at 1500 V_{rms} isolation between windings. Four models with current ratings from 20 to 80 mA and inductances from 28 to 450 µH are available. Common mode noise attenuation for the 80 mA parts is greater than 20 dB from 6 to 50 MHz. The CMB filters measure just 7.0 × 9.5 × 5.3 mm and cost $2.95 each in 1k quantities.
Coilcraft
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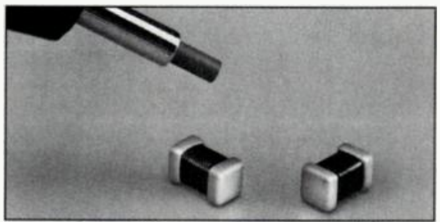
DISCRETE COMPONENTS

Helical Chip Inductors

The GLY series is composed of inductors measuring 1.6 × 0.8 × 0.8 mm (0603 size). The inductor is made by laser cutting a helix in



copper deposited on an alumina core. The terminals are nickel plated with a solder coat. The 17 models in the GLY series range in inductance from 2.2 nH to 47 nH. Standard tolerance is ±5 percent. Self resonant frequency is as high as 3200 MHz. Unit price is $0.41 for a single reel.
Sprague-Goodman Electronics, Inc.
INFO/CARD #241

Chip Capacitors

M-pulse Microwave announces a new family of chip capacitors packaged in Micro SMT™ packages. The devices exhibit low loss through 18 GHz. They are available in capacitances of 0.5 to 600 pF and standoff voltages ranging from 10 to 100 V. The M-pulse capacitors are available in two sizes: Ms - 0.012 × 0.028 inches (0301) and M24 - 0.04 x 0.02 inches (0402).
M-pulse Microwave
INFO/CARD #240

SEMICONDUCTORS

Dual ADC for DBS

The AD9066 is a dual high-speed, low-cost ($4.59 in thousands) analog-to-digital converter (ADC) expressly designed for the direct broadcast satellite (DBS) market. It integrates a pair of matched 6-bit converters, each with a 60 Msps sampling speed, required for digitizing high data-rate I and Q signals. The AD9066 draws 400 mW from a single +5 V supply, and is packaged in a 28 lead SOIC.
Analog Devices
INFO/CARD #239

Transmitter Encoder

EXEL Microelectronics has announced their "code hopping", or "rolling code" transmitter encoder, the XL106. The XL106 Transmitter Encoder incorporates a 56-bit transmission code, with 24 fixed and 32 code-hopping bits. The XL106 can be used to control up to five different receivers and provide seven separate functions to a single receiver. The chip operates from 2.7 to 6.3 V, with maximum standby current of 50 nA. EXEL's XL106 Keeloq Transmitter Encoder is priced at $1.20 in 10k quantities.
EXEL Microelectronics
INFO/CARD #238

I/Q Modulator & Upconverter

The UPC8104GR I/Q modulator with upconverter was developed for use in 900 MHz to 1.9 GHz applications. The chip can be driven by supply voltages of 2.7 to 5.5 V. The chip operates with I/Q signal from DC to 10

MHz modulating a 100 to 400 MHz signal which is then upconverted to frequencies in the range of 900 to 2,500 MHz. Typical power consumption is 28 mA at 3 V. The UPC8104GR includes an internal 90° phase shifter and ports for an external IF filter.
California Eastern Laboratories
INFO/CARD #237

900 MHz SOT-23 Bipolar Transistor

Hewlett-Packard has introduced its lowest-cost, 900 MHz silicon bipolar transistor, the AT-415 series, in the SOT-23 surface mount package. The AT-41533 features a 1 dB noise figure at 900 MHz and power output at 1 dB gain compression of +14.5 dBm (at 5 V, 25 mA). The transistor is also characterized for operation at 2.7 and 8 V. The AT-41533 is $0.27 in quantities of 50,000 to 99,999.
Hewlett-Packard Co.
INFO/CARD #236

200 W Power Transistors

Advanced Power Technology's new 200 W RF power transistors for narrowband push-pull RF power amplifiers are now available from Richardson Electronics. The ARF442 and ARF443 comprise a symmetric pair of RF power transistors with a combined output power of 400 W when operating at 100 V and 13.56 MHz. Each device has minimum gain of 20 dB and 65 percent efficiency.
Richardson Electronics, Ltd.
INFO/CARD #235

SIGNAL SOURCES

DDS Module

Novatech Instruments introduces the Model DDS4m direct digital synthesizer (DDS) module. The DDS4m is a low cost, 34 MHz signal source packaged on a 3.5 × 4.5 inch circuit board. The DDS4m simultaneously outputs a precise Sinewave and an accurate TTL clock signal. The output frequency is programmable from 1 Hz up to a maximum of 34 MHz in steps as small as 0.02 Hz. The DDS4m is priced at $395 each.
Novatech Instruments, Inc.
INFO/CARD #234

C Band Synthesizer

Communication Techniques' ISS-2000 series is a C band synthesizer designed for satellite communications. The synthesizers cover the C band (in bands) with standard tuning ranges

available from 12 percent of center frequency to 800 MHz maximum. Price is $995 each in 100 piece quantities.
Communications Techniques, Inc.
INFO/CARD #233

4-Pin VCXO Line

Oak Frequency Control Group's 794 VCXO, available from 1 to 50 MHz, is housed in an industry standard 4-pin DIP. The internal construction consists of minimal parts count and an OFC-manufactured quartz crystal. Tuning range is greater than ±100 ppm (0.5 to 4.5 VDC) with temperature stability of ±24 ppm from 0 to +70 °C. Pricing for the 794 at the 1000-piece level is $15.00 to $25.00.
Oak Frequency Control Group
INFO/CARD #232

EMI-Reducing Clock Generators

A series of products that can reduce electromagnetic interference by up to 20 dB has been introduced by International Microcircuits. There are three product groups in the series: spectrum spread clock generators (SSCG), spectrum spread modulator generators (SSMG), and spectrum spread oscillator replacements (SSOR). IMI's SSCG and SSMG are $3.15 at the 1,000 piece quantity, and the SSOR is $12 in small quantities.
International Microcircuits, Inc.
INFO/CARD #231

SIGNAL PROCESSING COMPONENTS

Filters

The DMT Division of Jay-El Products has developed a band-reject filter providing out-of-band insertion loss of less than 1 dB and out of band VSWR of 1.5:1. Center frequencies from 500 to 900 MHz are available. Also offered are bandpass filters with center frequencies from 450 to 650 MHz (LB version) and 650 to 860 MHz (HB version). These filters have passband widths typically from 15 to 20 MHz and insertion loss of 3.5 to 4.5 dB.
DMT
INFO/CARD #230

0° Power Splitter

Mini-Circuits has introduced a low cost, 2-way, 0°, 50 ohm power splitter, model ZN2PD - 1900, with SMA connectors, designed for wireless applications in the 1600 - 1900 MHz PCN band. The unit performs with typical isolation of 30 dB, typical phase unbalance of 0.2° and typical amplitude unbalance of 0.02 dB. Typical insertion loss is 0.18 dB, and typical VSWR is 1.08:1. Price is $69.95 each in quantities of 1 to 9.
Mini-Circuits
INFO/CARD #229

High Level Mixer

MITEQ's TIM0206HC2 is a high level (IP3 = +26 dBm) Schottky diode mixer developed to provide source- and load-VSWR-independent performance over the RF and LO band of 2 to 6 GHz with IF coverage from DC to 2 GHz. The mixer typically has RF and LO input VSWR of 1.25:1, and therefore is compatible with termination of any system filters without phase or amplitude ripples.
MITEQ
INFO/CARD #228

4-Way Power Combiner/Divider

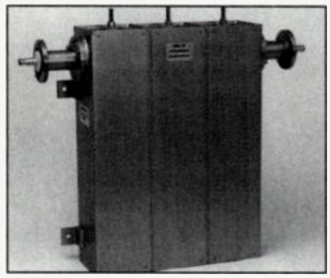
RF Power Components has developed a drop-in 4-way power divider/combiner that provides high isolation and low insertion loss from 180 to 300 MHz. Model RFP-4060 can easily handle 200 W CW with less than 0.5 dB total insertion loss. Isolation is ≥ 20 dB and the device measures 2.5 × 0.5 × 0.02 inches. Model RFP-4060 is priced under $99.00 each in 1,000 piece quantities.
RF Power Components, Inc.
INFO/CARD #227

8x8 Butler Matrix

Anaren's model 560014 is an 8x8 Butler matrix operating from 750 to 950 MHz. Minimum isolation is 20 dB, maximum insertion loss is 1.10 dB and maximum VSWR is 1.25:1. Maximum amplitude balance and phase balance is specified at ±0.80 db and 7 degrees, respectively. Safe power handling in 100 W average CW. The caseless laminated package measures 5.50 × 3.30 × 0.19 inches, not including standard SMA connectors.
Anaren Microwave, Inc.
INFO/CARD #226

5 kW UHF Bandpass Filter

Bandpass filter, model 11050-26 is used at the UHF transmitter output, UHF channel 26, to filter out-of-band spurious emissions. This model is available for any channel in the UHF band. The

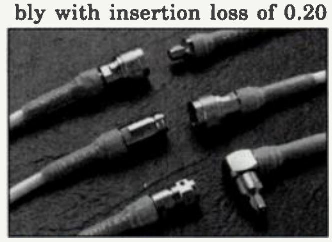


filter handles 5 kW of video power and 500 W audio power. Passband insertion loss is less than 0.35 dB across the 6 MHz NTSC channel.
Microwave Filter Company
INFO/CARD #225

CABLES & CONNECTORS

Cable Assemblies

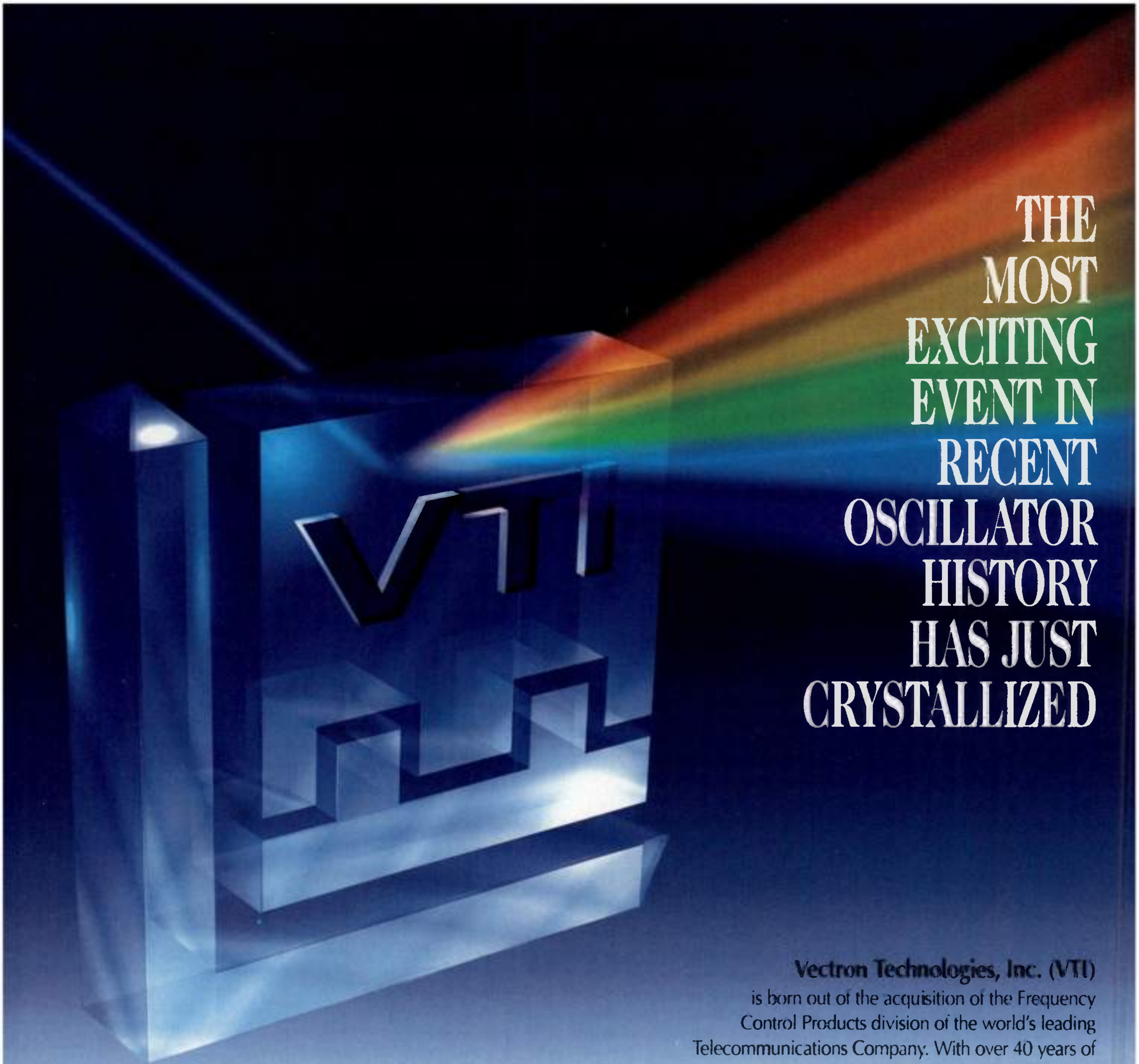
MICRO-COAX has introduced the UFB-311A, a UTiFLEX™ flexible microwave cable assembly with insertion loss of 0.20



dB/ft at 18 GHz. The cable assemblies have a power handling capacity of 490 W CW at 5 GHz, a dynamic bend radius of 4.75 inches, RF leakage of –100 dB at 1 GHz, and an operating temperature range of –65 to +165 °C.
MICRO-COAX
INFO/CARD #224

Jumper Cables

Times Microwave Systems announces jumpers offering both low loss and high flexibility. These cable assemblies provide greater than 100 dB shielding effectiveness and have a flame retardant and abrasion resistant jacket. The connectors used on the assemblies are silver plated, with gold plated center pins.



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Connector types available are type N plug, right angle plug and bulkhead jack (7/16 DIN connectors available soon). A typical NM-NM three-foot assembly has a price of $121.00.
Times Microwave Systems
INFO/CARD #223

F Connectors

The AMP® sealed F-connector offers enhanced performance up to 1 GHz and is designed to meet the requirements of Bellcore TA-NWT-001503 (4.3 and 4.4) and SCTE specifications. Nominal connector impedance is 75 Ω with an operating temperature of –40 to +115 °F.
AMP Incorporated
INFO/CARD #222

TOOLS, MATERIALS & MANUFACTURING

Space Simulation Chamber

The Space Simulation Thermal Chamber (SSTC) from Environmental Stress Systems allows stress testing of RF, microwave and wireless components in pressures ranging from 1 atm to 10^{-6} torr and temperatures from –99 to +200 °C. The system can accommodate components up to 12 × 24 × 8 inches. The system is priced from $19,950.
Environmental Stress Systems, Inc.
INFO/CARD #221

Radar Absorber

Cuming Corporation's flat sheet radar absorbers are now available with a pressure sensitive adhesive backing for fast and easy installation. The C-RAM product line covers the entire microwave frequency range.
Cuming Corp.
INFO/CARD #220

AMPLIFIERS

Receive Amplifier

AML announces the availability of model R100, a low noise dual diversity receive amplifier assembly. Integral bandpass filters are provided in each of the diversity receive paths. Gain is adjustable from –10 to +16 dB. Maximum noise figure is 3 dB 16 dB gain. The unit operates on a 26 VDC nominal supply and features a self check function.
AML Communications
INFO/CARD #219

SDLVA

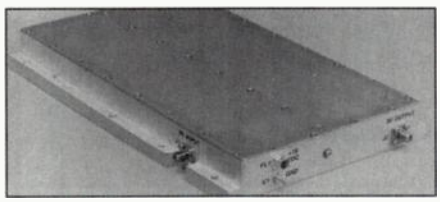
Model SDLVA-0120-70 (0.1 to 2.0 GHz) and SDLVA-06135 (600 MHz to 1.35 GHz) are DC-coupled successive detection log video amplifiers (SDLVAs). Specifications include 65/70 dB dynamic range, TSS ≤ –67 (20 MHz video BW), 25 mV/dB log slope, ±1.0 dB log accuracy, rise time ≤ 20 nsec, fall time ≤ 25 nsec, and delay times of 8 nsec at –10 dBm limited IF output. These SDLVAs measure 2 x 1.75 x 0.4 inches.
Planar Monolithics Industries, Inc.
INFO/CARD #218

Class A Amplifier

Densitron Microwave has created a new range of Class A linear amplifiers, specifically designed for intermodulation testing applications at PCN/DCS/PCS frequencies. The amplifiers are available in either rack-mounted or modular configurations with output powers of 50 or 80 W covering the 1800 to 1900 MHz band.
Densitron Microwave
INFO/CARD #217

800 - 2450 MHz Amp

Comtech Microwave products has introduced the AM88258-10 a Class A linear

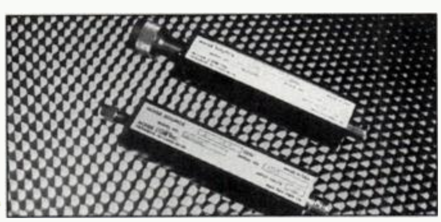


amplifier operating over the full 800 to 2450 MHz frequency range. Power output is 10 W at the 1 dB compression point. Operating voltage is +15 VDC at 7 A, and size is 5.9 x 9.4 x 1.1 inches.
Comtech Microwave Products Corp.
Power Systems Technology Div.
INFO/CARD #216

TEST EQUIPMENT

Noise Source

Noise Com introduces the NC 346 V, a broadband noise source covering the entire 100 MHz to 65 GHz range. The NC 346 V uses V-male connectors and operates from



standard +28 V noise figure meter power supply. It has low VSWR, good temperature, and voltage stability. Output ENR is from 10 to 17 dB.
Noise Com, Inc.
INFO/CARD #215

Cellular Diagnostics

The CES-1000 cellular system activity moni-

tor from Cellular Evaluation Systems displays and updates overhead message information in real time for an AMPS connection. The CES-1000 is a serial data interface used to communicate between a computer's 9-pin serial port to an OKI (900, 910, 1145 or 1150) or AT&T (3710, 3730 or 3760) cellular phone. The included software displays and logs RSSI, follows handoffs and decodes DTMF tones as well as controlling the handheld phone.
Cellular Evaluation Systems
INFO/CARD #214

Communications Test Set

A new hardware option is available for the Wavetek STABILOCK 4032 communications test set. The new option allows testing of DCS

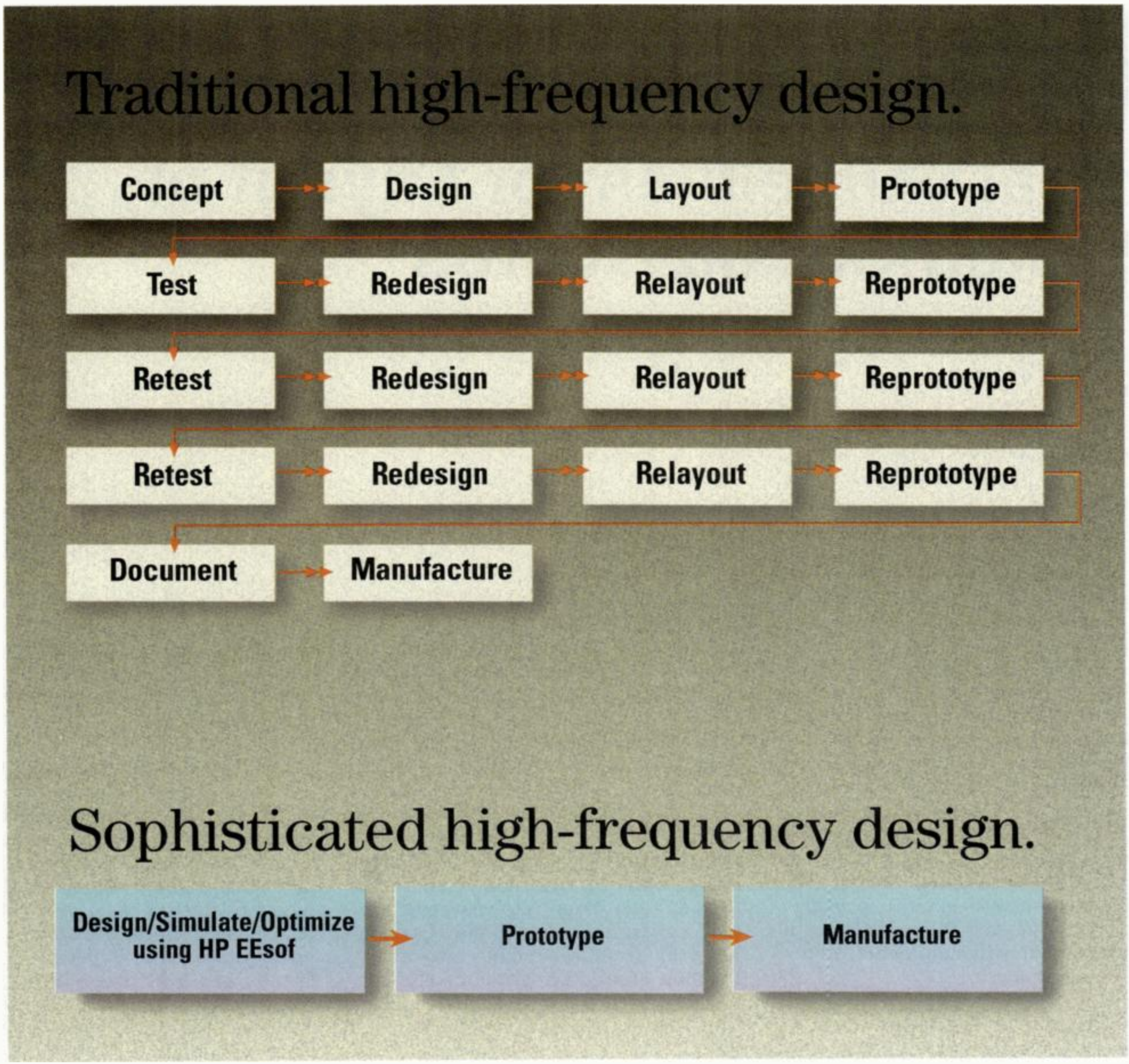
1800 and DCS 1900 radiophones or measurements on cordless DECT telephones. The 2.3 GHz option is a fast mounted plug-in module. In addition to Go/NoGo testing, the STABILOCK 4032 can perform more complex tasks such as repair on defective mobiles.
Wavetek Corp.
INFO/CARD #213

Dielectric Constant Measurements

The Damaskos model 0501 cavity measures dielectric constant of low loss materials over the approximate octave of 2 to 4.5 GHz in a non-destructive manner. It is ideal for thin dielectric sheets ranging in thickness from a few mils to the order of 100 mils. The measurement is made on common analyzers under the instrument control and data processing of Damaskos' "CAVITY" software. Other bands are available.
Damaskos, Inc.
INFO/CARD #212

Digital-TV Signal Analyzer

A new option for the HP 89400 vector signal analyzer family allows the analyzer to characterize digital RF modulation. Targeted at designers of HDTV and other advanced TV systems, the analyzer's demodulator can be configured with a few keystrokes for 16QAM, 32QAM, 64QAM and 256QAM formats as well as the 8VSB (broadcast) and 16VSB (cable) formats. QPSK, BPSK, FSK and MSK formats allow analysis of other digital services. Option AYH for the HP 89400 series VNAs is available for $2,500.
Hewlett-Packard Co.
INFO/CARD #211



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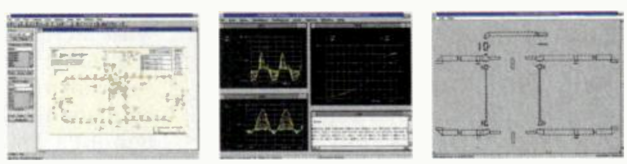

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Initial Guidelines for Layout of Printed Circuit Boards

By Gary A. Breed
Editor

Although design engineers typically do not do final layout of p.c. boards, many do prototypes and most work with the CAD department to develop the best circuit implementation. This month's tutorial reviews some basic rules for layout of RF circuits on printed circuit boards, including considerations for high frequency performance, EMI, crosstalk, and grounding.

Printed circuit board assembly of electronic products has the purpose of achieving low cost, automated, repeatable manufacturing. The board also provides mechanical support for components. In RF circuits, the board can also become an integral part of the circuit's operation, using microstrip and stripline techniques. Accomplishing all of these tasks simultaneously is an engineering art. We will review a few rules that will help achieve successful designs.

General Layout Planning

Here a few rules of thumb for planning the initial layout of p.c. boards. They are not absolute rules, but they will provide a starting point for grouping the various functional blocks in a manner that avoids major problems.

Keep high power circuits away from sensitive circuits — Yes, this is as obvious as it seems, but there are some subtleties to watch for. The power amplifier of a transmitter will normally be positioned away from low level oscillator, amplifier and modulator stages, but we need to be sure that the DC lines follow the same rule, since they can be a path for RF to follow. Sometimes there are conflicting requirements. For example, a power amplifier stage may have an associated VSWR monitoring circuit. We might be able to put sampling circuits by the amplifier and locate the remaining circuitry elsewhere, but then we have the extra length of the interconnections to deal with. Being aware of this kind of problem at the beginning lets the engineer think out the problem before the first trial layouts are made.

Avoid crosstalk problems — Crosstalk can take many forms. It can be a control signal getting into an RF signal and modulating it. It can be an outside effect coupled into a phase locked loop, causing unwanted modulation or instability. It can be an RF signal causing errors in monitoring circuits. The easiest solution is physical separation. More complex solutions include routing signals in shielded cables, through layers of a multi-layer board, localized shielding of affected circuits, or locating problem circuits off the board entirely. Whole books could be devoted to this problem area.

Avoid radiating layouts — If RF is not properly contained, it radiates (and any "antenna" receives signals as well as transmitting them). Certain physical structures radiate more readily than others. Long microstrip line sections, even if matched, will radiate. This is because the RF field is not perfectly contained between the conductor and the ground plane. Loops radiate with an efficiency proportional to the area enclosed. A loop can be formed with a cross-over connection in a path that otherwise is a properly-terminated transmission line. A loop can also be made by the combination of a component's leads and body. An often-overlooked area is the loop made between +DC and ground, which can enclose a very large area. Even if bypassed properly for RF, lower frequency radiation and pickup can occur if this rule is not followed.

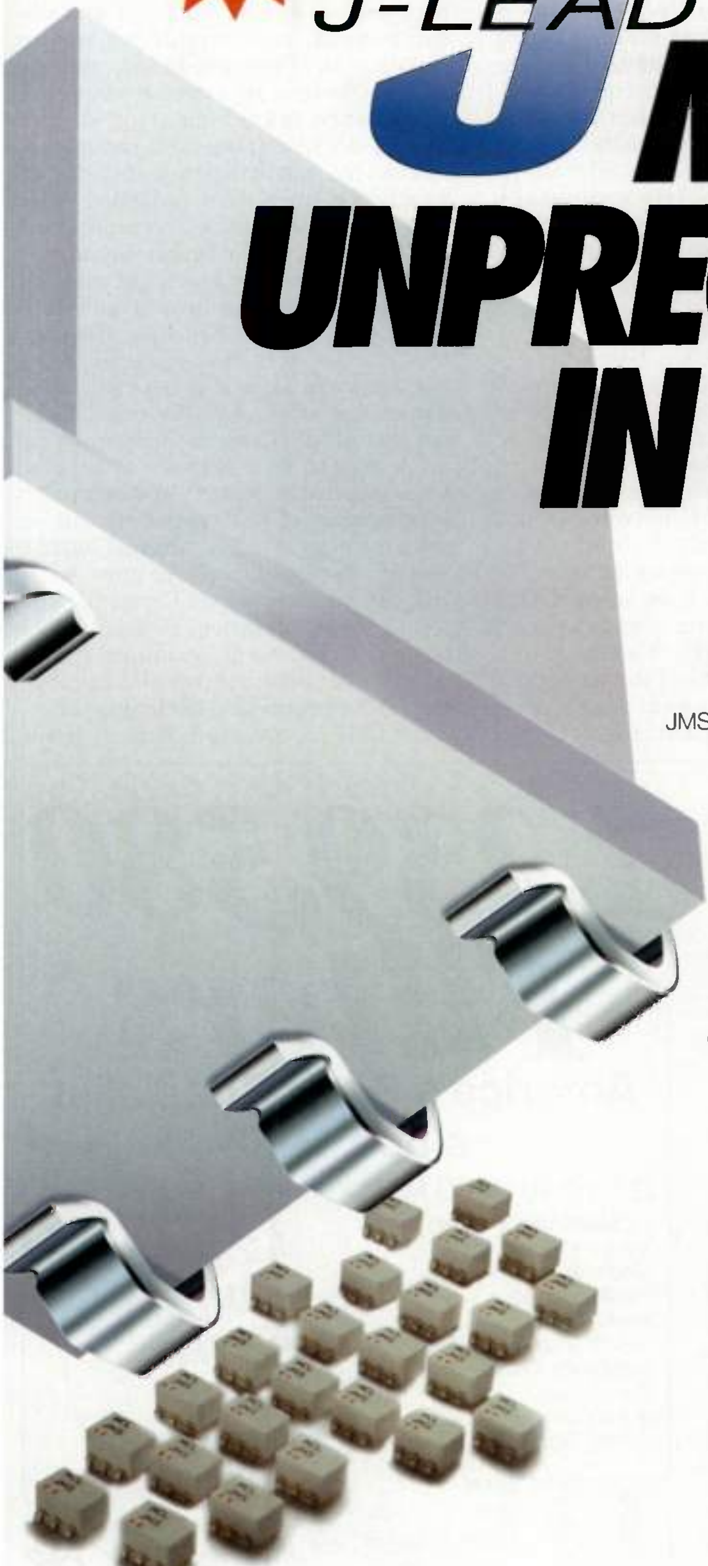
Design compact circuits with short connections — "RF is where every component is simultaneously an R, L and C and there is no ground." This definition always gets a chuckle because it is true. One way to minimize its effect is to use short, direct connections that have the lowest losses, the least chance for mismatches, and low inductance. This is a rule that always has conflicts. The shortest path might violate one of the previous rules, or might be physically impossible within the limits of p.c. board construction. When compromises are required, the additional path length needs compensation with transmission line techniques, or with extra care in its routing.

These few rules provide a starting point for board layout. Other engineers will have their favorite additional rules, based on experience with various types of circuits and performance demands.

Transmission Line Behavior

At radio frequencies, all printed circuit board traces have an effect on the behavior of the circuit. As engineers, we all know this, but in the pressure of getting a design finalized, this fundamental fact is often overlooked. As transmission lines, wide traces have a low characteristic impedance, narrow traces are high impedance. Short transmission line sections are inductive when they have higher impedance than the nominal circuit impedance, and they are capacitive when their impedance is lower. These are the characteristics that make microstrip circuits possible.

The magnitude of these effects is proportional to the dielectric constant ε_r of the board material. The phase velocity V_p of a signal is: $(c/\sqrt{\varepsilon_{eff}})$, where c is the speed of light. Common G10/FR-4 has an ε_r of 4.8, which means that a wavelength (λ) is $1/\sqrt{4.8}$ or 0.46 as long as free space. At 1 GHz, λ is 30 cm in free space and just 13.8 cm on FR-4. A 2.5 cm (1-inch) trace on such a printed circuit board is 0.18 λ, or 65 degrees, a very "long" distance in transmission line terms. Any variation in impedance between the trace and the impedance of the components it is connecting will result in a significant impedance mismatch, often resulting in reduced performance. Even at 100 MHz, the above 1-inch trace is still 0.018 λ, which is 6.5 degrees electrical length, enough to degrade some circuits.



Microstrip is intentionally used to make filters and matching networks, taking advantage of its transmission line properties. However, when we want "perfect" conductors, transmission lines are always present. For example, in high speed digital circuits the clock rate may be 66 MHz. To keep a good square wave shape, the harmonics (Fourier coefficients) should be undistorted up to at least five times that frequency, or 330 MHz. Our 1-inch p.c. board trace may be no problem at 66 MHz, but at 330 MHz, it will have a significant effect. Traces that make up a digital bus need to be matched in physical length to avoid timing problems, and they must be properly terminated. Digital signals have reflected waves just like RF signals, and those reflections can alter the shape of the desired square wave.

The behavior of microstrip line is well-documented (see the Bibliography), and this article is not meant to be a microstrip tutorial. However, microstrip techniques are integral to higher-frequency designs. These methods are so important that specialized printed circuit board materials are constantly being refined to enhance the performance of microstrip circuits. Low ε_r substrates with thin dielectric layers improve the dispersion and radiation performance of microstrip circuits. High dielectric constant ceramic substrates reduce the size of circuits, since signal wavelength is much shorter in these materials.

Power Circuits

The portions of the p.c. board circuitry that distribute DC power to the various components can be the source of problems if not properly designed. Bypass capacitors must be placed very close to the components' power pins to avoid making an L-C network of the wrong type. We want the shunt capacitance close to the component and the inductance on the side toward the power supply, since we always assume that the power supply has the lower impedance of the two. If the p.c. board trace or the component leads contribute a significant inductance between the bypass capacitor and the component, the impedance of the capacitor is effectively raised, making it less effective. In extreme cases, the impedance transformation of the bypass circuit can render it completely ineffective, resulting in instability or oscillation, increased radiation from that part of the circuit, or conduction of RF into the power supply circuitry.

The power supply bus itself must be maintained at the lowest possible impedance. Wide conductors, frequent bypass capacitors, and embedding the bus between layers of a multi-layer board are all commonly-used techniques. All of these methods must be implemented in a manner consistent with microstrip or stripline behavior. Impedances of the traces should be calculated (or at least approximated), and the bypass capacitors must be of the optimum value and type for the frequency of operation. It should also be noted that most techniques that improve RF behavior will also improve the EMI performance of circuits. However, EMI can be a problem at lower

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frequencies where RF concerns are minimal, and additional techniques brought to bear. In a circuit where both RF and EMI performance are being addressed, special attention must be given to both, as it is possible that methods for one can adversely affect the other (e.g., poor RF performance of a "standard" EMI filter above resonance).

Connections to the Board

Our final topic is another often-overlooked area: interconnections that bring signals, power and grounding to and from the board.

First, RF "ground" is somewhat indeterminate. An equipment case or the ground plane of a circuit board is not an infinite current sink of zero impedance! At DC this condition is approximated, but certainly not at RF. Significant RF currents flow in all conducting surfaces considered to be ground, creating potential differences. Ground loops are the result of two separate ground paths creating a signal that is the difference in their currents. Depending on the nature of the equipment, these can be 50/60 Hz or harmonics, switching power supply frequencies, or RF.

Two ways to minimize the effects of ground currents are: 1) Minimize the number of different ground paths (ideally one — central grounding); and 2) Create the lowest impedance ground possible, which usually means a large surface area of high-conductivity material (copper or aluminum).

Besides the ground loop problems noted above, wires are often unshielded, making them more susceptible to pickup of RF. Routing these wires and cables close to the walls of the cabinet lowers impedances and reduces the area enclosed by the loops they create. Locating them away from cables or circuits with high level signals will further reduce the potential for problems. Proper bypassing and EMI filtering will normally eliminate any remaining problems.

Summary

Layout of a printed circuit board for reliable RF performance is no simple task. The number of potential problems is large, but by starting with these basic rules, you can develop a circuit that maintains good RF performance while meeting the requirements for large-scale manufacturing. *RF*

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Phase Noise Measurement for Under $250

By W.A. Suter
Amplifonix, Inc.

It is difficult, if not impossible, to measure the close-in phase noise of wide-tuning, free-running, voltage controlled oscillators using direct spectrum analysis. The signal does not remain steady long enough to obtain a clean trace on the spectrum analyzer, with the result that phase noise closer than 100 kHz cannot be accurately measured. Even if external influences such as temperature changes, or electrical and mechanical noise are eliminated, the carrier drifts too fast.

One way to steady the carrier is to lock the oscillator in some way. It is possible to phase-lock one of two identical VCOs to the other with a crystal in the loop as in Figure 1. The signal which drives the "reference" oscillator is a baseband signal which is related to the frequency noise of the device under test (DUT). The locking system is equivalent to a black box with a volts-out vs. frequency-in transfer function.

A system as in Figure 1 was assembled. The signal from the DUT was mixed with the signal from the second oscillator (the "reference"), and an IF of 10 MHz was generated. This 10 MHz signal was then mixed again with a 10 MHz crystal oscillator and the resulting baseband signal used to drive the reference VCO so as to lock the system. The baseband signal was then amplified, digitized, and analyzed to derive the phase noise of the DUT.

The phase-locked loop bandwidth was made wide enough to create a stable phase-locked system for a large-modulation-sensitivity VCO. The method initially seemed to work well, but a number of problems became clear. The transfer function is offset-frequency dependent because of the filtering effect of the phase-locked loop – within the loop bandwidth the phase noise is suppressed by an amount which changes with frequency – and accurately correcting for this was not easy. White noise could be injected into the loop and the loop suppression characteristic measured. However, other problems with this method – such as preventing the system from locking at some harmonic of the crystal, preventing the reference from jumping from 10 MHz below to 10 MHz above the DUT, and the system's overall complexity – led to the setting aside of this method in favor of the delay line discriminator method, which proved to be very simple, stable, sufficient to the task, and inexpensive.

The delay line discriminator method is a wideband approach, the frequency being limited only by the frequency response of the power divider and/or the mixer. A block diagram of the setup is shown in Figure 2.

A single VCO is required. Its signal is split with an in-phase power divider; one signal drives the LO port of the mixer directly, and the other drives the RF port after passing through a delay line. The two signals produce a voltage at the IF port which depends on the phase difference between them. If the mixer is perfectly balanced and has infinite isolation between the LO and IF ports, the voltage will be negative V volts when they are in phase, zero when they are 90 degrees out of phase, and positive V volts when they are 180 degrees out of phase.

If the delay line length is set so that the two signals are 90 degrees apart (i.e. in "quadrature"), the signal at the IF port will be 0 volts, with an AC signal riding on it proportional to the frequency noise of the oscillator. This will be true regardless of the oscillator frequency. Hence the problem of the drift-

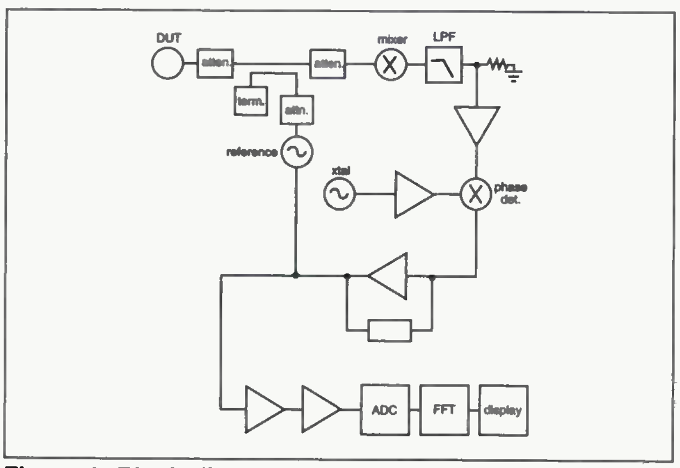

Figure 1. Block diagram of reference oscillator/downconversion phase noise measurement system.

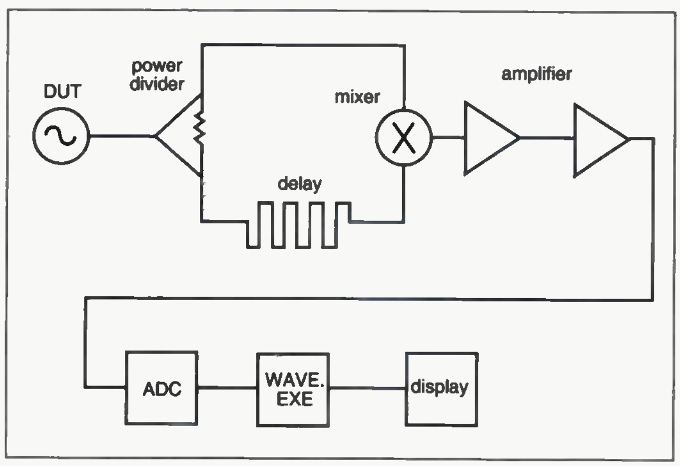

Figure 2. Block diagram of discriminator/delay line set-up for phase noise measurement.

ing carrier is eliminated.

This AC signal is amplified, digitized, and analyzed for its volts vs. frequency content, and the phase noise of the DUT calculated from this. The single side band phase noise is calculated from:

$$L(f_m)\ \frac{\text{dBc}}{\text{Hz}} = 20\log(V_{rms}) - 20\log(K_d) - 20\log(f_m) - 10\log(F_{res}) - 20\log(\text{Vgain}) - 3\text{dB} \qquad (1)$$

where

V_{rms} = rms voltage at offset frequency f_m

f_m = the offset freq in Hz

$K_d = 2\pi(T_d)(K_\phi)$ volts/radian

T_d = delay line time delay in seconds

K_ϕ = volts/Hz constant of the discriminator

Vgain = the voltage gain of the base band amplifier

F_{res} = the frequency resolution of the FFT routine

The ability of this system to convert frequency noise to voltage noise is dependent primarily on the length of the delay line. If the delay line is long, the rate of change of output voltage with change in input frequency will be large at the zero crossing point, and the system will be very sensitive to input frequency noise. Increasing the length of the delay line, however, also increases its loss, which reduces the RF port drive, which reduces the voltage available at the IF port. For a given mixer there is an optimum delay line loss which will give the correct signal levels at the mixer, and the most IF output for linear mixer operation.

The VCO

Since the oscillator's carrier frequency is eliminated at the mixer, the only limitation on the frequency of the VCO is that it be within the bandwidth of the power divider and/or the mixer. The delay line discriminator method is applicable to any oscillator frequency. The VCOs of interest here are those in the 500 MHz to 3500 MHz range.

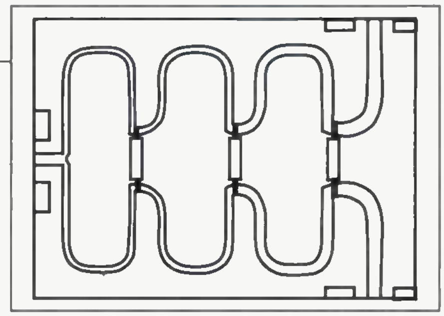

Figure 3. The power divider.

The power requirement of the VCO is that it drive the mixer sufficiently, although an RF amplifier may be used to boost the signal level. If an RF amplifier is used, it must be placed before the power divider and must be operated in its linear range. In this way, the amplifier's residual noise will be correlated out of the measurement at the mixer, and its own 1/f noise will not be modulated up onto the carrier. If it was placed in the delay line path (to allow longer delay lines), the amplifier's frequency noise would be indistinguishable from the frequency noise of the oscillator.

Since the VCOs of interest have high modulation sensitivities (5 MHz/V to 150 MHz/V), great care must be taken to prevent external voltage noise from getting onto the VCO tune line. Often this means a battery supply will be needed for the tune voltage, and a low noise potentiometer must be used for varying the tune voltage. Also, if the VCO has significant pushing figure, external noise may also modulate the carrier via the supply line, so a battery may be useful here too. External noise in the supply lines appear as frequency lines on the phase noise plot.

The Power Divider

A Wilkinson power divider was chosen over a resistive divider for its lower insertion loss and better output port isolation. The Wilkinson divider can easily be made to cover more than an octave. A three-section power divider was designed which covers the 500 to 1500 MHz band. The design, shown in Figure 3, was etched on G-10 and uses axial leaded 1/4 W resistors. Over 500 to 1500 MHz, its insertion loss is 3.5 to 4.5 dB, isolation is greater than 14 dB, and port return loss is greater than 10 dB.

A power divider could easily be made for 1500 - 3500 MHz, though it would be better if this were made on a lower loss circuit board to keep the insertion loss down.

The choice here is between a fixed or a variable delay line. For a fixed delay

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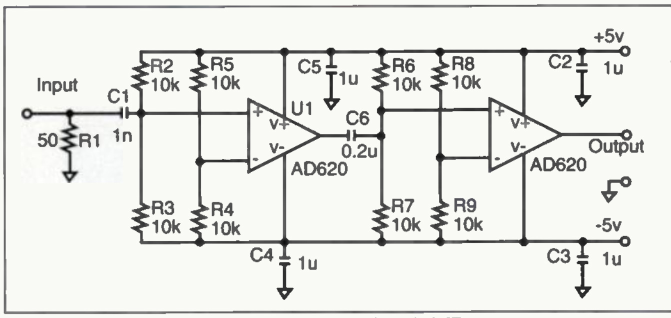

Figure 4. Low noise amplifier to feed A/D converter.

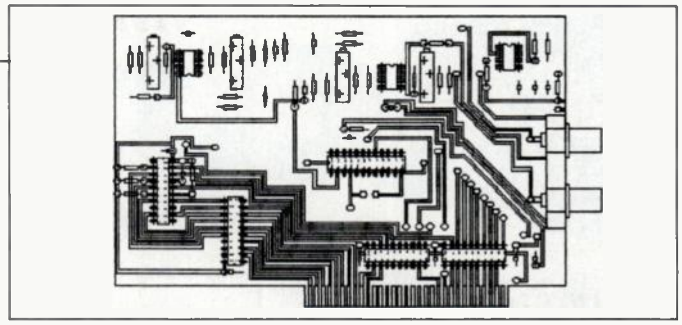

Figure 6. PCB layout for the analog to digital converter.

line, one would need to vary the frequency to achieve quadrature by varying the tune voltage on the VCO. With a variable delay line, one could set the tune voltage and adjust the delay.

A fixed delay was chosen for several reasons. For a fixed delay of the length typically required, there are sufficient nulls within the VCO's tuning band width to adequately characterize a VCO's phase noise performance. If a fixed delay is used, then the value of T_d is known and constant.

Twenty feet of RG-58 coax has a measured delay of 31.2 nsec, and a loss of 2.5 dB at 500 MHz, increasing linearly to 7.5 dB at 3 GHz. The maximum offset frequency which can be measured with a 31.2 nsec delay line is 5.1 MHz. 100 kHz is the maximum offset of interest here, so this is acceptable. For 40 feet of RG-58 coax, the delay is 62.4 nsec and the maximum offset frequency is 2.55 MHz. The maximum offset restriction comes from the desire to avoid having to include the sin x/x characteristic of the discriminator in the equation for phase noise.

As mentioned above, a longer delay line increases the slope of the volts vs. frequency curve at the zero crossing, increasing the discriminator constant, but also increasing the loss. If the mixer requires an LO drive of +7 dBm and no more than +1 dBm RF level, it is desirable to have a delay line with a loss of about 7 dB. This means that, for 500 to 1500 MHz, 40 feet of RG-58 could be used, while, for 1500 to 3000 MHz, 20 feet of RG-58 would work well.

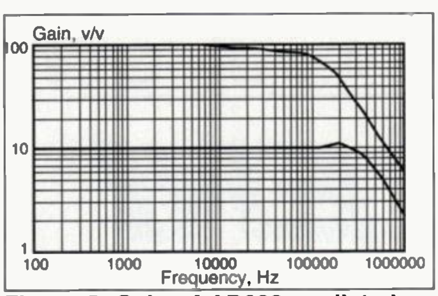

Figure 5. Gain of AD620, as listed on data sheet.

The Mixer

The mixer must be double balanced. This reduces the harmonics present at the IF port, and typically means better isolation between ports. Good isolation and balance will minimize DC offset at the IF port, which must be DC-coupled. To get a large output voltage level a high level mixer could be used, but there must be sufficient signal level to drive it. The RF port must be driven below saturation, (the mixer must operate in its linear region), to minimize the contribution of its own noise.

Given these considerations and the desired operating frequency range, a good choice is the Mini Circuits SRA-3500. This mixer takes a +7 dBm LO drive, has a 500 to 3500 MHz RF and LO bandwidth, and an IF bandwidth of DC to 1000 MHz. It has acceptable LO-IF isolation and is inexpensive.

The Low Noise Amplifier

This is a critical part of the system. The IF port of the mixer is terminated in 50 ohms and the voltage developed across it is amplified before being fed to the analog-to-digital converter. The input voltage range of the A to D converter is set to ±1.25 V. By experiment, it was found that a gain of over 1000 was required to amplify the IF output noise to a level that would make suffi-

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	240L	A	20kHz-10MHz	40W	50dB(±1.5dB)
	2100L	A	10kHz-12MHz	100W	55dB(±1.5dB)
	A25-25PA	A	300kHz-30MHz	25W	51dB(±1.5dB)
	A-150	A	300kHz-35MHz	150W	55dB(±1dB)
	A-300	A	300kHz-35MHz	300W	55dB(±1dB)
	A-500	A	300kHz-35MHz	500W	60dB(±1dB)
	A-1000	A	300kHz-35MHz	1000W	60dB(±1.5dB)
	325LA	A	250kHz-150MHz	25W	50dB(±1.5dB)
	3100LA	A	250kHz-150MHz	100W	55dB(±1.5dB)
	3200L	A	250kHz-120MHz	200W	55dB(±1.5dB)
	400AP	A	150kHz-300MHz	3W	37dB(±1dB)
	403LA	A	150kHz-300MHz	3W	37dB(±1dB)
	411LA	A	150kHz-300MHz	10W	40dB(±1.5dB)
	525LA	A	1MHz-500MHz	25W	50dB(±1.5dB)
	550L	A	1.5MHz-400MHz	50W	50dB(±1.5dB)
	5100L	A	1.5MHz-400MHz	100W	50dB(±1.5dB)
	604L	A	0.5MHz-1000MHz	4W	40dB(±1.5dB)
	607L	A	0.8MHz-1000MHz	7W	43dB(±2dB)
	630L	A	400MHz-1000MHz	30W	51dB(±2dB)
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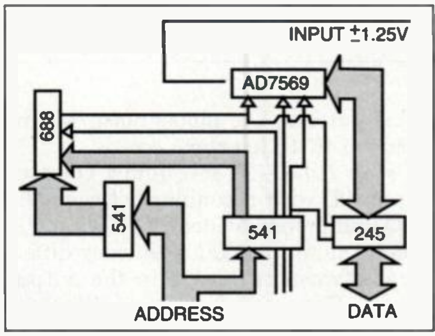

Figure 7. Block diagram of A/D conversion system.

cient use of this ±1.25 V input range. The amplifier's gain needed to be flat to at least 100 kHz. Since no anti-aliasing filter was used, the amplifier's gain can drop steeply after 100 kHz. The amplifiers must generate little noise.

Figure 4 shows the final configuration. The gain of the Analog Devices AD620 is set by a single external resistor. The data sheet gain is shown in Figure 5. Two stages were used, with the gain shared equally between them. This maintained sufficient bandwidth and gain, and made it easier to balance the amplifiers and measure their gain.

A gain of about 40 per stage was used. The data sheet indicates that for this gain setting, the gain starts to roll off at about 100 kHz, and this was confirmed by experiment. The DC blocking capacitor, C6, was used to enable one to set the balance on the second stage. A potentiometer was not used because it is not sensitive enough and is too noisy. Noise generated by the AD620 is low.

The goal in this part of the system was conversion speed, coupled with simplicity and low cost. The Analog Devices AD7569 8-bit ADC/DAC device is an excellent device for this application. It can convert at a maximum rate of 400 ksamples/second and is easily interfaced to the PC bus. Its input and output voltage can be set for 0 - 1.25 V, 0 - 2.5 V, ±1.25 V, or ±2.5 V.

The circuit board which was designed around this device is shown in Figure 6. This board has input and output amplifiers to condition the signal for 0 to 10 V input and output, and a fixed-frequency audio oscillator of known voltage and frequency for the purpose of calibrating the conversion rate. The unipolar, 0 - 10 V input range turned out to be non-ideal for the discriminator method, though it was good for the phase-locked system first used. The circuit was modified for ±1.25 V. The block diagram in Figure 7 shows the final configuration.

The actual sampling rate is determined by the software and the host PC. In a 486DX33 MHz computer, using an assembly language input routine optimized for speed, the sampling rate is about 250 kHz, which is adequate for measuring frequency components to 100 kHz.

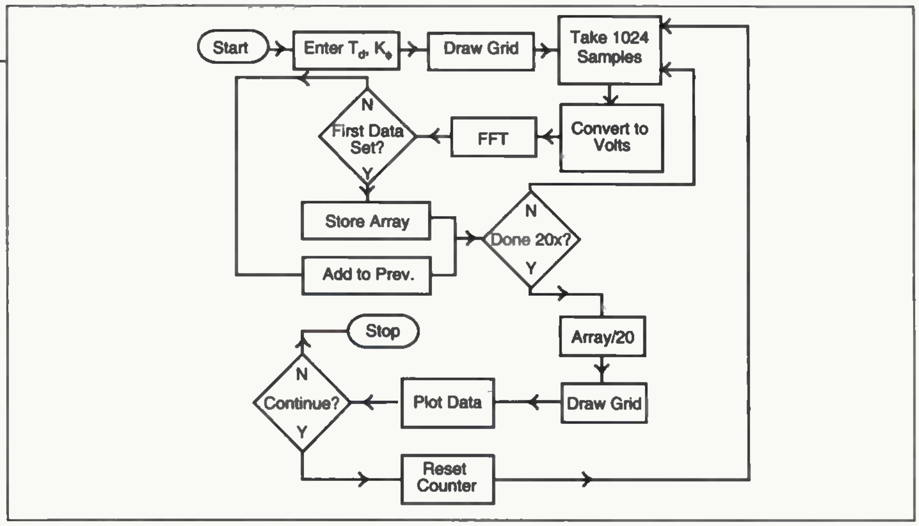

Figure 8. Algorithm used for wave analysis.

Wave Analysis Software

The program used to analyze the signal from the discriminator and calculate and display the phase noise was written using Microsoft QuickC, with an embedded assembly language routine for fast sampling of the signal. A listing of the program is included with the documentation for the the software used in this measurement scheme [*see the end of this article for information on ordering the software, ed.*], and a flow

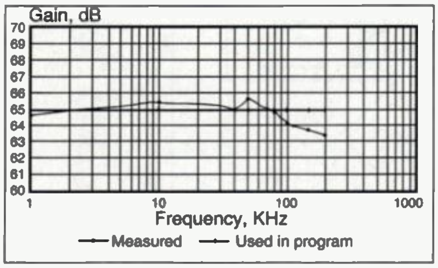

Figure 9. Measured gain of one stage of the low noise amplifier.

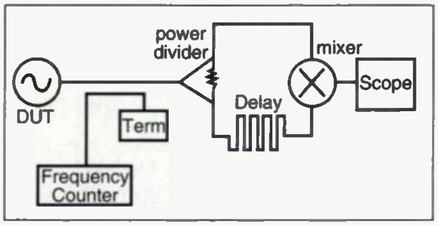

Figure 10. Test setup to measure delay line electrical length.

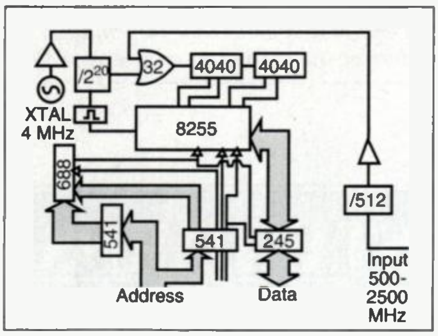

Figure 11. Block diagram of frequency counter card designed to fit in a PC cardslot.

chart is shown in Figure 8.

After sampling has taken place, the data is converted to volts, a Hanning window applied, and the FFT performed. The single sideband phase noise is calculated, and the data stored. After 20 sets of data have been collected and the average calculated, the data is displayed on a 2-cycle log-linear plot. If the wave analysis software is run as a DOS application from Windows, PrintScreen may be used to capture the display, and PaintBrush may be used to print the graph.

The Hanning window is used because, although it may have up to 1.5 dB error in amplitude, it gives good frequency resolution. A flat top window gives good amplitude accuracy, but poor frequency resolution. If no window is used, the amplitude may be up to 4 dB in error. The Hanning window is good for random noise signal analysis.

The number of samples taken is 1024. This gives 512 data points in the frequency domain (each line contains both magnitude and phase information). An rms voltage is calculated from these points and corrections applied for the discriminator constant, etc., and the single sideband phase noise calculated.

Only frequencies from 1 kHz to 100 kHz are displayed. This is the region of valid data based on the bandwidth of the baseband amplifier and the sampling rate.

System Calibration and Accuracy

The key to accurate measurements is good calibration of the various components. Calibration consists of determining the value of the variables in the equation given above. The variables other than T_d and K_ϕ may be determined once, and set in the program. T_d, however, must be known, and K_ϕ measured, prior to making a particular measurement.

The following is a description of how the system was calibrated. An estimate of the effect of various errors was made by examining the effect of small changes in the variables of the single sideband phase noise equation.

Low Noise Amplifier – The gain of one stage was measured using an audio oscillator and a dual-trace oscilloscope. Since the two stages are identical, the overall gain was calculated from the single stage gain. It was too difficult and inaccurate to measure the gain of the two-stage amplifier by this method because of the large gain involved. The gain of the amplifier is shown in Figure 9. The maximum error from gain inaccuracy is less than 1 dB

Frequency Resolution – This was determined by digitizing a known-frequency signal and determining the sampling rate. Knowing the time between samples and the number of samples used in the FFT routine, the frequency resolution is given by the expression: 1/(#samples·time step). This was determined to be 247 Hz. An error of ±2 Hz in this figure would give an error of about ± 0.1 dB in the phase noise plot.

Conversion Accuracy of A to D and FFT Routine – The internal reference voltage of the AD7569 is nominally 1.25 V. The reference voltage was determined by digitizing a known-amplitude signal and determining which reference voltage resulted in the correct digitized amplitude at the output of the ADC. For this particular AD7569, a reference voltage of 1.315 V gave the correct output. If the calculated voltage is in error by ±5 percent, the phase noise will be up to ± 0.46 dB in error.

Delay Line – A frequency counter was used, with a coupler between the VCO and power divider (see Figure 10), to determine the VCO frequency difference between two nulls in the output voltage. From this frequency difference the time delay of the delay line can be calculated from

$$T_d = \frac{1}{2\Delta f} \qquad (2)$$

The frequency counter was a card which was designed to fit in the PC, and uses an NEC divide-by 512 pre-scaler to divide the RF signal down to the input range of the counter. A block diagram of the counter is shown in Figure 11.

For 20 feet of RG-58 cable this method measured a time delay of 30.937 nsec. The cable was also measured on an HP network analyzer; the delay by that method was measured to be 31.2 nsec. Either result could be used. An error of ±5 percent in the delay results in about ±0.45 dB error in the phase noise calculation.

Measurement of K_ϕ

This measurement must be done every time a phase noise measurement is made and involves determining the rate of change of mixer output voltage versus change in input frequency. It is desirable to determine and use the point of maximum sensitivity.

Several factors affect the value of K_ϕ. For a sine shaped curve, the slope at the zero crossing is equal to the peak value of the sinusoid. The slope can thus be measured by measuring the positive and negative peak voltages on either side of the point at which the phase noise measurement will be done, and then using the average of the two peaks. (The VCO tune voltage is used to set the mixer output to these peaks). This measurement can be made on the mixer IF output using either an oscilloscope or voltmeter.

As frequency increases the delay-line loss increases, which causes the signal level at the mixer's RF port to decrease, which causes the output voltage to be a sine wave of slowly decreasing amplitude. This, along with the fact that real mixers do not have perfect balance or infinite LO-IF isolation, means these two voltages will not be the same, and that the mixer will have some DC offset at the point of maximum sensitivity. The average of the two voltage peaks

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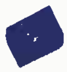

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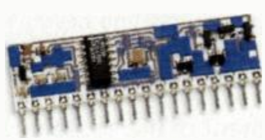

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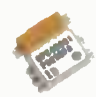

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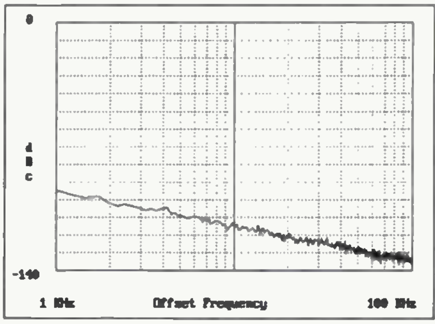

Figure 12. Measured system noise floor.

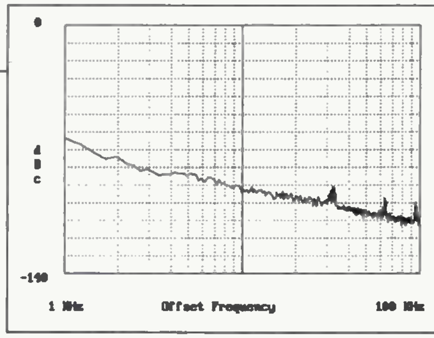

Figure 13. Measured phase noise of an example VCO with center frequency at 1100 MHz..

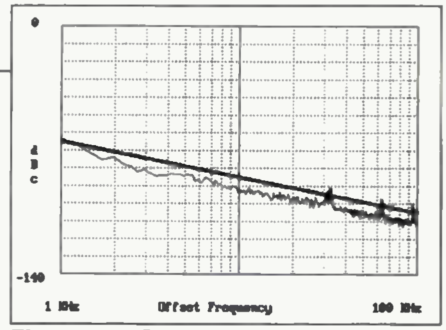

Figure 14. Same measurement as in Figure 13 compared to a measurement done on a Comstron PN-9000

should be used for the value of K_ϕ, and the output voltage should be set to half way between them. An error of ±10% in V_ϕ will give an error of about ±0.92 dB.

It is important that the output voltage remains at the set point during a measurement as departure from this point will give a phase noise that is lower than the true value. How far the output voltage may drift from this set point, given a maximum error allowable in dB, can be found from:

$$\Delta V = K_d \sqrt{1 - 10^{\frac{error}{5}}} \qquad (3)$$

Overall measurement uncertainty

It can be seen that error in K_ϕ is a major contributor to error in the single-side-band phase-noise. If all errors are additive but the output voltage is kept at the set point during measurement, the actual phase noise may be about 3 dB higher or 4.5 dB lower than that displayed. The extra 1.5 dB on the low side is from the Hanning window amplitude inaccuracy. This would be the worst case.

Making a Measurement

A phase noise measurement is simple: set up the equipment as shown in Figure 2; add gain or attenuation as required before the power divider to achieve +7 dBm at the LO port of the mixer, and enough delay line to get about 0 dBm at the RF port; measure K_ϕ and then set the mixer output at the point of maximum sensitivity as explained above. Run the program, entering the delay line length in nsec, and K_ϕ in volts when requested. After the plot is displayed, it may be printed.

As mentioned above, external noise may easily enter the VCO via the power supply or the tune line and this will appear as discrete lines on the phase noise plot. The sources of this noise include computer displays, line frequency hum, power supply ripple, switching power supplies, oscillators in calculators and digital multimeters, demodulated radio station audio, radar pulses, and mechanical vibration. The VCO should be shielded from this noise pickup – this is almost a science in itself! This is especially true for VCOs with high modulation sensitivity and non-zero pushing figure.

System Noise Floor

The noise floor is an important measurement because it shows the "signal to noise" ratio of the system and indicates the lowest phase noise that can be measured with a given delay line and discriminator constant. This measurement is easy to make. Simply set up to make a measurement as normal. Once K_ϕ is known, break the delay line path and terminate the power divider port in 50 ohms. Then run the measurement using the measured value of K_ϕ.

Figure 12 shows the system noise floor, and Figure 13 shows the phase noise of an oscillator. It can be seen that the VCO noise is more than 20 dB higher than the noise floor in this case. The lines at 34 kHz and its harmonics are noise picked up from the PC's monitor – they can be eliminated by making the measurement with the monitor off.

The phase noise of the VCO shown in Figure 13 was also measured by the same discriminator method on the Comstron PN-9000 and the HP 3048 system. The results are compared in Figure 14. The data from the other systems are shown as a solid line drawn between the data at 1 kHz and 100 kHz. The center frequency of thsi VCO is about 1100 MHz. Figure 15 shows a plot for another approximately 1 GHz VCO with the phase noise as measured on the Comstron PN-9000 plotted in the same way as above. Figure 16 is a plot for an octave tuning VCO.

Usefulness of the system

This phase noise measuring system is useful for measuring free running VCOs with moderately low phase noise. Only one VCO is needed, the problem of its carrier drifting is eliminated, and its input frequency range is virtually unlimited. The system is accurate, and is quick and easy to use. Offset frequency is limited to 1 kHz on the low side (by the DC blocks needed in the low noise amplifier), and to 100 kHz of the high side (by the sample rate of the analog to digital conversion).

The various components of the system, including the frequency counter, cost less man $250. All printed circuit boards were home made. This cost does not include the analog oscilloscope, clean power supplies, and 486DX33 computer. Such a phase noise measurement system is easily within the budget of most involved in the design and manufacture of oscillators for use in personal communication networks, and can make a significant contribution to lowering of capital equipment costs.

Improvement of the system

A 12-bit A to D converter would improve the voltage resolution. Such a converter would require a board which has its own RAM for quick read and store of the data, with transfer and processing of the data by the program after an end-of-conversion signal is received. To enable better frequency resolution in the 1 kHz to 10 kHz region, the sample rate could be slowed down, but this would necessitate the use of a good anti-aliasing filter to remove frequency components above half the sampling rate. These hardware improvements would involve an increase in the complexity of the system, and the cost

There are several software improvements which could be made. A routine involving the use of the digital to analog section of the AD7569 and the frequency counter could be used to automatically measure T_d and K_ϕ. Initial experiments with this indicate that two problems will have to be solved. Digital noise must be prevented from getting to the sensitive tune port of the VCO. The other problem is measuring a voltage of about ±0.13 V accurately with a

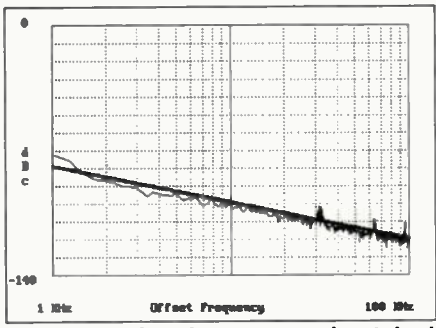

Figure 15. Another approcximately 1 GHz VCO with phase noise measured and displayed as in Figure 14.

±1.25 V input range on the A to D. A 12 bit A to D would be useful for this task. Another software improvement would be the ability to print the phase noise plot directly from the program.

These improvements would add "bells and whistles" to the system. However, as it stands, the system as presented works well and accurately.

Conclusion

The delay line discriminator method is not new. A simple, low cost, implementation has been developed here which enables one to measure the phase noise of oscillators whose carrier drifts too fast, or whose phase noise is too low, to be measured by direct spectrum analysis.

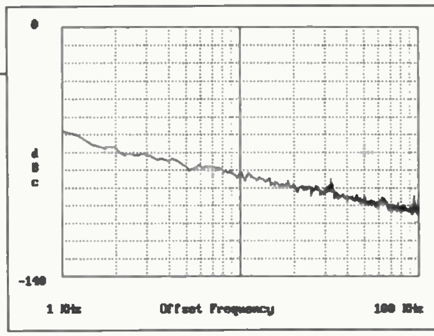

Figure 16. Phase noise plot for an octave-tuning VCO.

This system is valuable to the designer and manufacturer of inexpensive VCOs.

The wave analysis program described in this article, is available from Argus Direct Marketing. For ordering information, see page 96. RF

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About the Author

Bill Suter received his BSEE from Cape Town University in South Africa in 1976. After emigrating to the USA at the end of that year, he has been involved in many aspects of circuit design and fabrication from audio through RF to 20 GHz; from amplifiers, to oscillators, to mixers, to couplers. His hobbies include writing software, electronics, and especially airplanes. He is currently Senior Design Engineer at Amplifonix, designing primarily oscillators. He may be reached at (610) 559-7178, or (215) 464-4000.

Design Parameters for 1- Through 7-Pole Input-Matched Lowpass/Highpass Filters

By Chase P. Hearn
NASA Langley Research Center

The transfer loss mechanism in conventional LC filters is reflective; attenuation is through mismatching of the driving source in the stopband. Filters having constant resistive input impedance in addition to attenuation are desirable for some applications, such as terminating diode mixers. Filters having those properties have been described as branching filters [1], diplexers [2], and constant-resistance filters [3]. Such filters are physically realizable but are not widely used. The work described here was undertaken because the prior work was either limited in scope and/or did not produce the desired performance. These results yield branching filters with low or highpass Butterworth transfer responses through order seven and constant-resistance performance superior to previous designs. With few exceptions, the element values for these filters differ from standard Butterworth values.

A branching filter is illustrated in Figure 1; a low-pass and a high-pass filter are connected either in parallel or series. Power not dissipated by the load on port 2 is dissipated by the load on port 3. Port 2 has a low-pass and port 3 a high-pass transfer response. Ideally, the imaginary components of Y_{in} or Z_{in} of the combined networks sums to zero and the real parts sum exactly to 1/R, or R, respectively [3]; the loss mechanism in a branching filter is therefore absorptive, rather than reflective.

It is possible to achieve n-pole Butterworth lowpass or highpass transfer

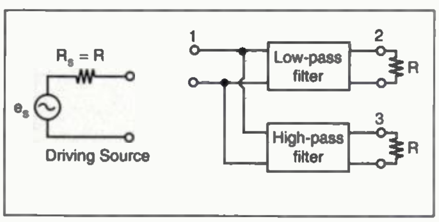

Figure 1a. Shunt-connected branching filter.

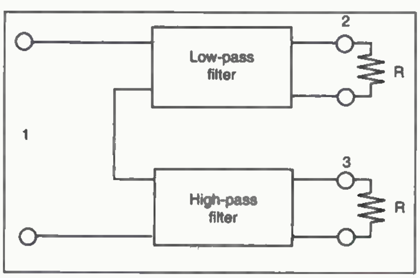

Figure 1b. Series-connected branching filter.

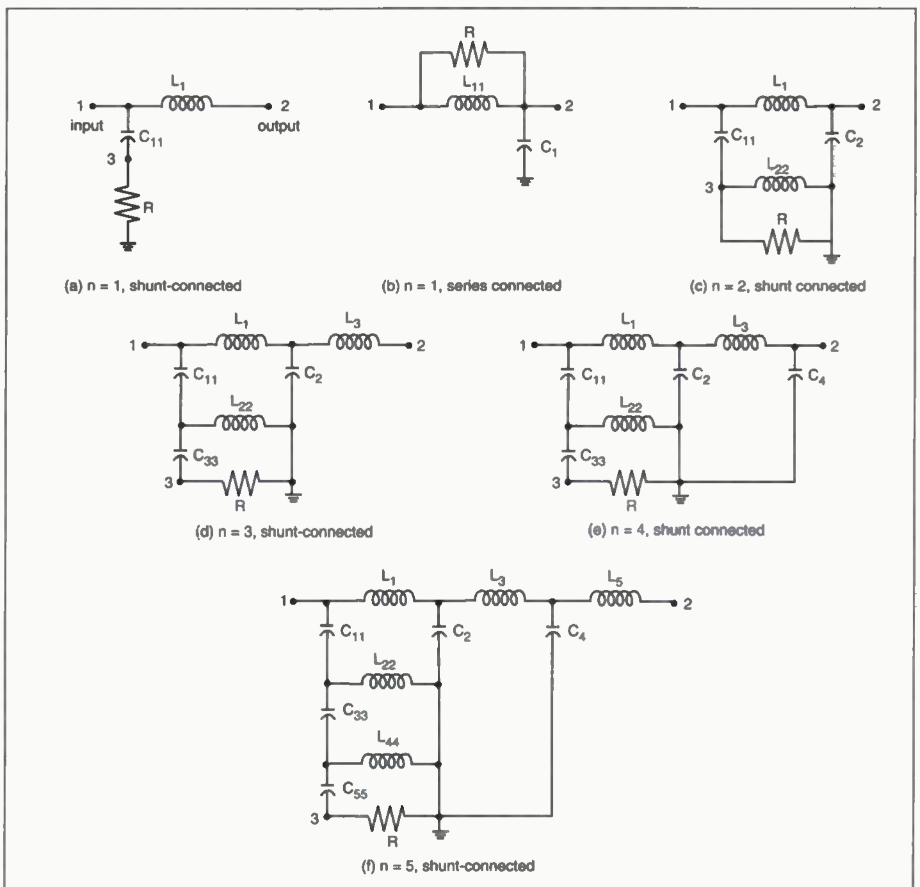

Figure 2. One to 5-pole lowpass branching filters of characteristic impedance R.

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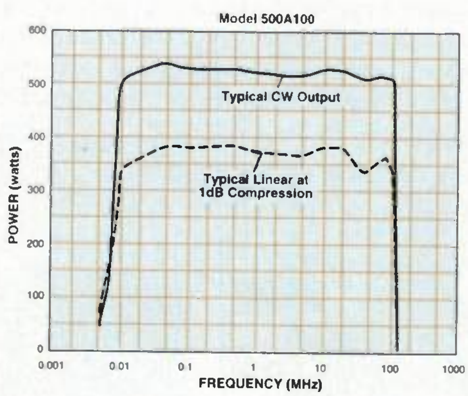

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responses and purely resistive input impedances with 2n+1 circuit elements – without introducing additional transfer loss in the passband. However, filters with ripple in the passband (Tchebycheff) or the stopband (m-derived) cannot be made to have constant input resistance with the additional n+1 elements [3]. The (n+1) element impedance-correcting network could be applied to the output rather than the input of a filter to achieve a resistive output impedance. This would be desirable if the signal or local oscillator port of a diode mixer was driven from a filter.

Consider the 1-pole shunt-connected branching filter shown in Figure 2(a). With ports 2 and 3 terminated in R, the impedance at port 1 can be expressed as:

$$Z_{in} = R\frac{R + \frac{L}{RC} + j\left(\omega L - \frac{1}{\omega C}\right)}{2R + j\left(\omega L - \frac{1}{\omega C}\right)} \qquad (1)$$

If L and C are related by:

$$\frac{L}{C} = R^2 \qquad (2)$$

Z_{in} (but not Z_{out}) will equal R; either L or C can be set arbitrarily to control the cut off frequency. If $L = R/\omega_c$ and $C = 1/\omega_c R$, to satisfy equation 2, the power delivered to the loads on ports 2 and 3 is reduced by 50 percent at $\omega=\omega_c$ and $Z_{in}(\omega)$ equals R for all values of ω.

A series-connected configuration as shown in Figure 2(b) would be used if the filter port to be compensated had a shunt input arm. Symbolic derivations similar to that become increasingly complex when the order of the filter exceeds two. Element-value relationships from Reference 1 for 1- through 3-pole branching filters are shown in Table I. A numerical approach was used to find the element values for 4- through 7-pole shunt-connected branching filters and also to verify the results of Reference 1 for lower-order branching filters.

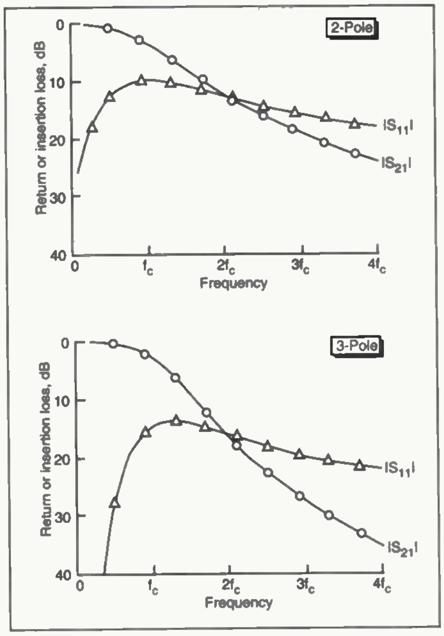


Figure 3. Input impedance, return and insertion loss of branching filters using element values from Reference 1.

Element values for branching filters were derived with a CAD gradient optimization procedure and tabulated in Table II; these values are the same as those in Reference 1 only for the 1-pole case; the shaded entries for 2- and 3 pole filters in Table I differ significantly from those in Table II. Figure 3 shows |S11| and |S21| for 2- and 3-pole filters using the element values from Table I. The minimum input return losses (r.l.) are only 9 dB for the 2-pole filter and 12 dB for the 3-pole filter between $f_c/10$ and $4f_c$, while the values in Table II produce input return losses exceeding 70 dB. Note that there are the same inverse relationships between corresponding L and C elements in the low and high-pass sections of branching filters that are found in conventional filters [4].

Table III contains standard Butterworth element normalized low and high-pass values [4] which were recommended for input-matched diplexers in Reference 2 and the minimum return losses which result when the

n	Lowpass Section					Highpass Section					Minimum return loss, dB
	L1	C2	L3	C4	L5	C11	L22	C33	L44	C55	
1	1	–	–	–	–	1	–	–	–	–	00
2	1	1	–	–	–	1	1	–	–	–	9
3	1.5	1.33	0.5	–	–	0.67	1.33	2	–	–	12

Table I. Element values for first through third-order branching filters from Reference 1; questionable values are shaded. Multiply by (R/ω_c) for inductors and ($1/\omega_c R$) for capacitors.

n	Lowpass Section							Highpass Section							Minimum return loss, dB
	L1	C2	L3	C4	L5	C6	L7	C11	L22	C33	L44	C55	L66	C77	
1	.9979	–	–	–	–	–	–	.9979	–	–	–	–	–	–	99
2	1.413	0.7063	–	–	–	–	–	0.7063	1.413	–	–	–	–	–	126
3	1.499	1.333	0.500	–	–	–	–	0.6661	0.7493	1.998	–	–	–	–	96
4	1.531	1.578	1.085	0.3841	–	–	–	0.6532	0.6335	0.9219	2.603	–	–	–	99
5	1.543	1.692	1.383	0.9013	0.3152	–	–	0.6464	0.5899	0.7245	1.123	3.253	–	–	77
6	1.553	1.760	1.555	1.202	0.7549	0.2562	–	0.6437	0.5675	0.6408	0.8207	1.279	3.675	–	79
7	1.556	1.800	1.662	1.398	1.052	0.6366	0.2239	0.6412	0.5540	0.5976	0.7029	0.9274	1.525	4.699	70

Table II. Normalized element values for first through seventh-order branching filters derived numerically. Multiply by (R/ω_c) for inductors and ($1/\omega_c R$) for capacitors.

input ports are paralleled to form branching filters. The values in Tables I and III are totally different; seven entries in Table I and only two in Table III agree with Table II.

Conventional Butterworth low-pass filters have input and output return losses which degrade from a good match when $f \ll f_c$ to 3 dB at f_c. A reasonably good match (r.l.> 20 dB) is had over an increasingly larger fraction of the passband as n is increased, illustrated in Figure 4 for 3- and 5-pole filters. Adding the n+1 additional circuit elements needed to produce an input-matched filter increases the output r.l. from 3 to 6 dB at f_c.

This is not a theoretical development, so no claim of absolute superiority can be made for the values in Table II. Other combinations might produce equal or better results; however, the values in Table II produce nearly perfect input matches and transfer responses identical to an n-pole Butterworth response at $f_c/10$, f_c, $2f_c$, and $4f_c$.

The gradient optimization process used to derive the element values became increasingly difficult when n exceeded 3. Sub-optimum minima were encountered and the results of each optmization were highly dependent on the optimization statement and the initial values. The values in Table II for n > 4 were obtained only after an extrapolation scheme was devised for selecting the initial values. The element values from the first four successful optimizations were plotted as shown in Figure 6 and extrapolated as indicated by the dotted lines. The n = 5 optimization was constrained arbitrarily to a ±10 percent range around

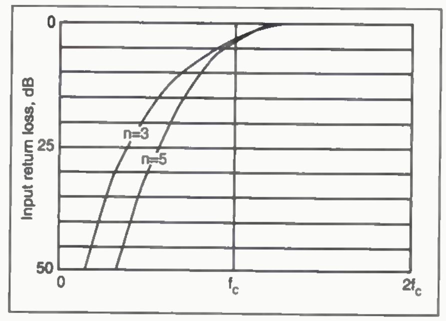


Figure 4. Input return loss in 3 and 5-pole Butterworth filters.

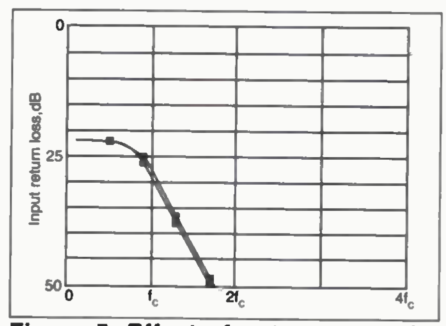


Figure 5. Effect of output port mismatch on a 3-pole branching filter (Γ_L = ±0.082).

n	Lowpass Section							Highpass Section							Minimum return loss, dB
	L1	C2	L3	C4	L5	C6	L7	C11	L22	C33	L44	C55	L66	C77	
1	2.000	–	–	–	–	–	–	0.500	–	–	–	–	–	–	7
2	1.413	1.413	–	–	–	–	–	0.7077	0.7077	–	–	–	–	–	24
3	1.000	2.000	1.000	–	–	–	–	1.000	0.5	1.000	–	–	–	–	7
4	0.7649	1.847	1.847	0.7649	–	–	–	1.307	0.5414	0.5414	1.307	–	–	–	2
5	0.6177	1.617	2	1.617	0.6177	–	–	1.617	0.6177	0.5	0.6177	1.617	–	–	1
6	0.5147	1.414	1.931	1.931	1.414	0.5174	–	1.933	0.707	0.5179	0.5179	0.707	1.931	–	1
7	0.4449	1.247	1.801	2.000	1.801	1.247	0.449	2.248	0.8019	0.5552	0.5000	0.5552	0.8019	2.248	1

Table III. Normalized element values for first through seventh-order branching Butterworth filters from Reference 4 and minimum return loss when connected as branching filters.

the extrapolated initial values. Only the optimized value for L3 (n=5) proved to be out of that range, which became obvious during the optimization process, and the lower limit on L3 (n=5) was reduced. The estimated values for the n = 6 and 7 optimizations obtained with this procedure were within ±5 and ±3 percent of the final values, respectively.

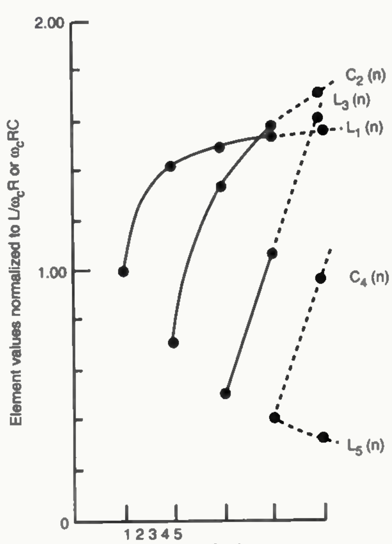
It was observed that sub-optimum minima produced element values inconsistent with the curves in Figure 6.

The constrained parameters in the optimization were input impedance from $f_c/10$ to $4f_c$ and transfer response, $|S21|$, at f_c, $2f_c$, and $4f_c$. The only specification imposed on the highpass section was indirect, through its effect on $Z_{1.c.}$ and the low-pass transfer response. The optimization statement for an n-pole low-pass branching filter (BRANCHn) in Touchstone [5] notation is shown in Table IV.

The heavy weight on the –3 dB point was necessary to make the cut-off frequency precisely f_c. The high-pass transfer responses, $|S31|$, after optimization based primarily on the low-pass response were observed to be

```
SWEEP fc/10 4fc fc/10
BRANCHn RE[Z1] = 0 10
BRANCHn IM[Z1] = 0 10
RANGE fc fc
BRANCHn DB[S21] = -3 50
RANGE 2fc 2fc
BRANCHn DB[S21} = (theoretical attenuation at 2fc)
RANGE 4fc 4fc
BRANCHn DB[S21] = (theoretical attenuation at 4fc)
```

Table IV. Optimization for an n-pole lowpass branching filter in Touchstone notation.



	Estimate	Final Value
C_2 (n)	1.71	1.692
L_3 (n)	1.60	1.383
L_1 (n)	1.56	1.543
C_4 (n)	0.96	.091
L_5 (n)	0.32	0.3152

Figure 6. Determination of initial element values by extrapolation for n = 5 from n = 1 through 4 values.

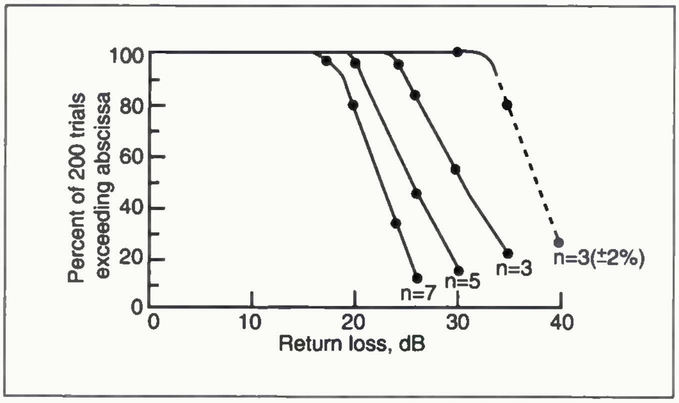


Figure 7. Sensitivity of input match to random ±5 percent element value variations.

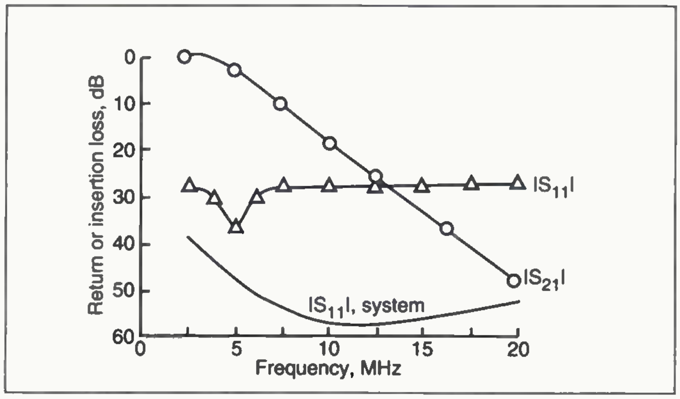


Figure 8. Measured performance of 3-pole branching filter using pre-set components.

nearly ideal Butterworth.

A practical consideration is the extent to which a mismatched load on the output port ($R_{load} \neq R$) affects the input-port match of an otherwise ideal branching filter. It was determined by trial that the input r.l. is essentially that of the mismatched load resistance in the lower portion of the passband and improves as frequency approaches and exceeds f_c. This is shown in Figure 5 for a reflection coefficient of ±0.082, a r.l. of 21.7 dB.

Another practical concern is the effect of element value tolerances. It was observed that changing a single element value one percent from ideal decreased the minimum r.l. of a 3-pole branching to about 40 dB. Monte-Carlo analyses [5] were run for the n = 3, 5, and 7 cases to determine the effect of random element deviations within ± 5 percent of ideal. The solid curves in Figure 7 show that element errors more seriously degrade the performance of higher order filters; tightening the toler-

ances greatly improves the input match. The dashed curve in Figure 7 corresponds to tolerance of ±2 percent, which increased the r.l. for a 90 percent yield to 34 dB, compared to 25 dB with a ±5 percent tolerance.

A 3-pole branching filter with Z_0 = 50 Ohms and f_c = 5 MHz was designed from Table II and fabricated for evaluation. The inductors and capacitors were measured and preset to an accuracy of about ± 3 percent prior to being installed; no further adjustments were made. Return and transfer losses were measured with a spectrum analyzer/tracking generator (HP 8553B/8443A) and a high-directivity return-loss bridge and are plotted in Figure 8. The results demonstrate that a good input match and a Butterworth transfer response can be achieved simultaneously in practice.

Conclusions

Normalized element values for first through seventh order shunt-connected branching filters were determined by numerical methods and are presented here in tabular form. These filters have Butterworth low-pass or high-pass transfer responses but are matched to a resistive source at all frequencies (r.l. > 70 dB) without adding excess transfer loss in the passband. This is accomplished at a cost of additional circuit elements; an n-pole response requires 2n+1 elements. Branching filters based on these results are superior to those based on previously published design procedures. A 3-pole branching filter was fabricated and evaluated, with good results. *RF*

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About the Author

Chase P. Hearn was a NASA research engineer from 1961 until his retirement in 1995. His work there involved analysis, design and development of circuits and systems spanning VLF to mm-waves and microwave measurement theory and practice. He is now pursuing independent R&D. He has been a licensed radio amateur since 1954 and enjoys building HF and VHF amateur equipment and restoring vacuum-tube and discrete component equipment of the forties and fifties.

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Robust Power Transistor Market Riding Wave of Emerging Communications Systems

This month's Product Forum covers the RF power transistor market. Several manufacturers offer their opinions regarding the market and technology trends affecting this market segment.

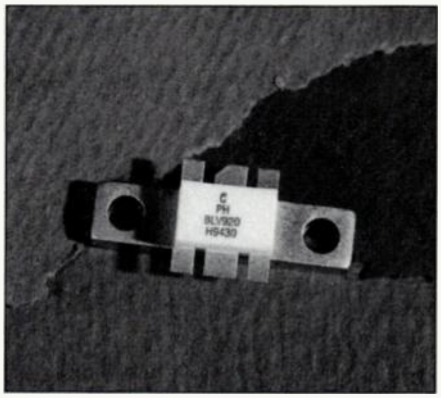
Polyfet RF Devices

Over the past five years Polyfet RF Devices experienced substantial growth and was recognized this year as one of the 50 fastest growing high tech companies in Southern California. Growth came in the areas of cellular base stations, paging base stations and military radios. Polyfet has been offering only gold metallized VDMOS RF transistors. Recently Polyfet introduced a new line of Lateral DMOS transistors. These transistors have the source on the bottom side of the transistor die allowing the die to be directly attached to the heat sink, eliminating the requirement of BeO. Additionally, without having to use wire bonds for the source leads, RF gain is increased by 2-3 dB over the older VDMOS technology. Polyfet sees the trend over the next two years of LDMOS replacing VDMOS and Polyfet is well positioned to support the marketplace with this new technology.

Richardson Electronics

The market continues to grow as applications become more defined and numerous new companies try to grab their niche in this market.

In addition to the PCN and Cellular market, one interesting application that has become very active involves reverse cable amplifiers (5 - 100 MHz). For a large variety of reasons, people want to send information through already existing cable TV lines. By not using the standard telephone line, the data rate is 10x faster through the cable TV line. This has introduced a variety of new business opportunities.



Digital cellular and PCS are driving the production of transistors designed for linear operation in UHF bands. This 20 W transistor, fabricated by Philips and distributed by Richardson Electronics, is designed for class AB amplifiers operating from 820 to 960 MHz.

Time to market is another outstanding variable currently involved in the power transistor market. The customer base that addresses all the different applications are constantly striving to keep themselves up-to-date on the latest products available on the market.

In the last six months we have seen 1.5 - 2.0 GHz PCS applications finally explode as manufacturers have introduced pallets and modules to address them.

Motorola Communications Semiconductor Products Div.

The emergence of PCS has spurred explosive growth in the RF power transistor market. This growth has been driven by two dominant factors, high technology and low cost. The increasing price pressures on the PCS infrastructure market and the need for high performance microwave devices are now guiding Motorola. We are introducing a new lineup of high performance 2 GHz sub-micron silicon bipolar devices led by the upcoming MRF20060, a 60 W, 9 dB gain base station service device. In addition, we are looking at LDMOS FETs for the same frequency range which could utilize very low cost plastic packaging.

The market should shift into high gear over the next year and be extremely strong for the next three to five years. Motorola is on a program to marshal substantial manufacturing resources, from both wafer fabrication and assembly standpoints, to aggressively service this market need.

SGS-Thomson Microelectronics

The market for RF power transistors is clearly in a growth mode, reflecting the general growth in the total market for RF components, subsystems and systems.

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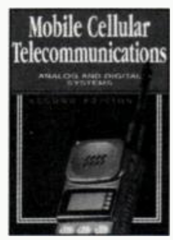

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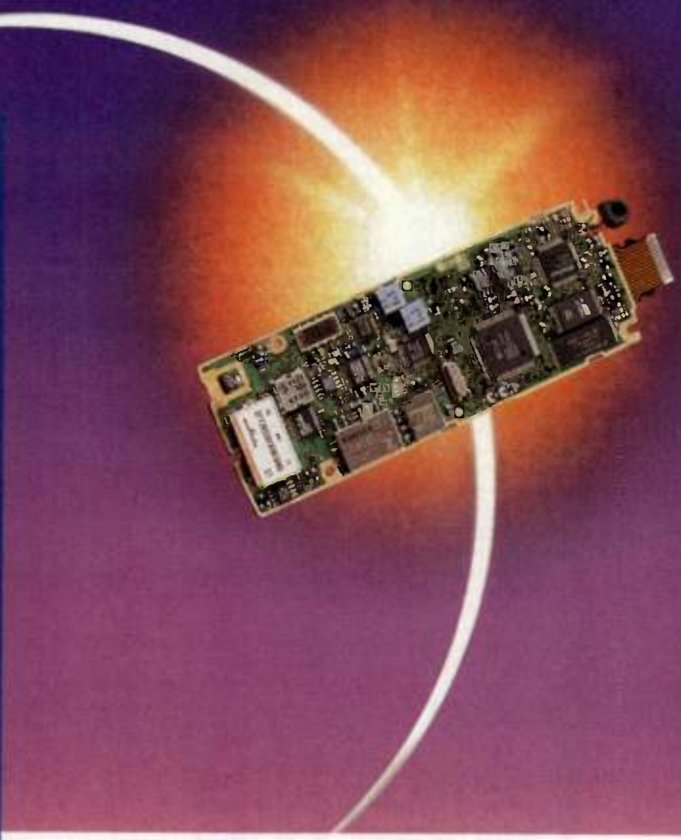

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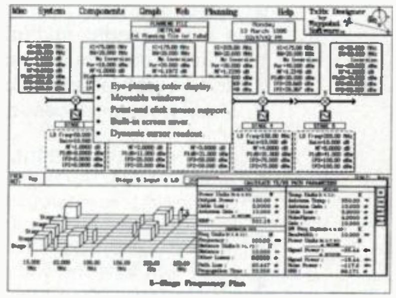

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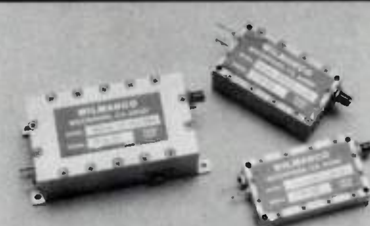

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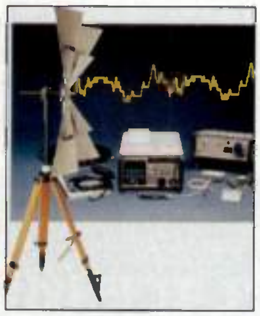

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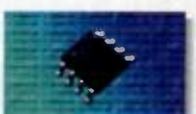

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Varian Microwave Equip. Prods., 3200 Patrick Henry Dr, Santa Clara, CA 95054....(408) 496-6273

Sheilded Fiber Optic Links

Antenna Research Associates, Inc., 11317 Fredrick Ave., Beltsville, MD 20705....(310) 937-8888

Surge Generators

Velonex Corp., 560 Robert Avenue, Santa Clara, CA 95050 ....(408) 727-7370
Pulse, Surge, Transient, ESD, Hybrid Generators

ESD TEST EQUIPMENT

ESD Event Detectors

Monroe Electronics, Inc., 100 Housel Avenue, Lyndonville, NY 14098 ....(800) 821-6001
TREK INC., 3932 Salt Works Rd., P.O. Box 728, Medina, NY. 14103 ....(800) FOR-TREK

Field Strength Meters

Antenna Research Associates, Inc., 11317 Fredrick Ave., Beltsville, MD 20705....(310) 937-8888

Surface & Volume Resistivity Meters

Monroe Electronics, Inc., 100 Housel Avenue, Lyndonville, NY 14098 ....(800) 821-6001
TREK INC., 3932 Salt Works Rd., P.O. Box 728, Medina, NY. 14103 ....(800)FOR-TREK

MATERIALS, HARDWARE AND PACKAGING

SHIELDING MATERIALS

Antenna Towers

EMCO, P.O. Box 1546, Austin, TX. 78767 ....(512) 835-4684

Architectural Shielding

Specialty Technical Components, Inc., P.O. Box 2106, Southeastern, Pa 19399 ....(610) 647-9000

Broadband EMI

Steward, Inc., 1200 E. 36th Street, Chattonooga, TN 37401........(615) 867-4100 Fax (615) 867-4102

Conductive Adhesives

Venture Tape Corp., 30 Commerce Rd., Rockland, MA 02370..........................(617) 331-5900

Conductive Fiber/Fabric

Venture Tape Corp., 30 Commerce Rd., Rockland, MA 02370..........................(617) 331-5900

Ferrite Absorber Tiles

Fair-Rite Products Corp., P.O. Box J, Wallkil, NY 12589..........................(800) 836-0427

Gasketing Materials

Venture Tape Corp., 30 Commerce Rd., Rockland, MA 02370..........................(617) 331-5900

Laminates

Venture Tape Corp., 30 Commerce Rd., Rockland, MA 02370..........................(617) 331-5900

Magnetic Shielding

Ad-Vance Magnetics, Inc., 625 Monroe St., Rochester, IN 46975 ....(219) 223-3158

Positioning Controllers

EMCO. P.O. Box 1546, Austin, TX. 78767 ....(512) 835-4684

Sheilding Foils and Tapes

Venture Tape Corp., 30 Commerce Rd., Rockland, MA 02370..........................(617) 331-5900

MATERIALS, HARDWARE AND PACKAGING

SHIELDING MATERIALS

Tripods

EMCO, P.O. Box 1546, Austin, TX. 78767 ....(512) 835-468

Turntables

EMCO, P.O. Box 1546, Austin, TX. 78767 ....(512) 835-468

SHIELDING EMI/RFI SERVICES

Vacuum Platers, Inc., 115 S. Union Street, Mauston, WI 53948 ....(608) 847-564

SHIELDED ENCLOSURES - EQUIPMENT

Component/Module Cases

Marmin-Hil Plastics, Inc., 101 Roselle St., Linden, NJ 07036..........(908) 925-2940 (800) 886-294

SHIELD ROOMS AND CHAMBERS

Shielded Rooms EMI/RFI/Magnetic

Rantec, P.O. Box 1546, Austin, TX 78767 ....(512) 835-468

ESD PACKAGING

Antistatic and Conductive Polyurethanes

Pad-Tastics, Inc., P.O. Box 50479, Cicero, IL 60650............(708) 780-8402 FAX (708) 780-163

ESD ENVIRONMENTAL CONTROL

Antistatic or Conductive Flooring

Freudenberg Building Systems, Inc./Nora Rubber Flooring
94 Glenn St., Lawrence, MA 01843..........................(508) 689-053

TEST LABORATORIES AND CONSULTANTS

TEST LABORATORIES

Spectrum Control, Inc., 6000 West Ridge Rd., Erie, PA 16506..........................(814) 835-400
TUV Product Service, Inc., 1775 Old Hwy. 8, New Brighton, MN 55112..........................(800) 472-799

EMI/EMC CONSULTANTS

EMC&I, 5700 Corsa Ave., #100 Westlake Village, CA 91362 ....(818)889-224

EMI/EMC TESTING

LambdaMetrics, P.O. Box 1029, Cedar Park, TX 78630-1029 ....(512) 219-821
FCC listed lab, one EMC Engineer, 30 years RF design experience, Prompt Personal Service

CONSULTANTS

Kimmel Gerke Assoc. Ltd., 1544 N. Pascal, St. Paul, MN 55108 ....(612) 330-372
EMC Design/Troubleshooting-Medical, Industrial, ITE, Vehicles & More-50+ years exp.
Regulatory Consultants Network, P.O. Box 344, Liberty Hill, TX 78642..........................(800) 326-633
EMC Design, EMC/EMI/EMS/ESD & Electrical Product Safety

High-Speed/ Wide-Bandwidth Databook

Comlinear's *1995 Databook Supplement* is a 256-page reference book providing full detail of Comlinear's broad line of high-speed, wide-bandwidth components for signal conditioning, data conversion and serial digital interface. It includes complete specifications and performance plots for new and updated products introduced since Comlinear's *1993-1994 Databook*.
Comlinear Corp.
INFO/CARD #210

Fax-on-Demand Service

Andrew Corporation announces the availability of "Answers from Andrew", an automated fax-on-demand system to better serve the information needs of its customers. The service provides product information, technical documents and news publications. By dialing 1-800-861-1700 (708-873-3614 for international callers), anyone with a touch-tone phone and fax machine can access the system.
Andrew Corp.
INFO/CARD #209

SNA Measurement Accuracy

Anritsu Wiltron offers a new technical seminar along with a companiion application note (54100A-7) and free software program. The free seminar identifies common sources of scalar network analyzer measurement uncertainty and provides common sense suggestions for accuracy improvements. An emphasis is placed upon "measurement accuracy" rather than simple "instrumentation accuracy."
Anritsu Wiltron Co.
INFO/CARD #208

Reconditioned Equipment

Catalog no. 35 from Bid Service contains 100 pages of reconditioned electronic manufacturing and scientific/analytical instrumentation. Categories of equipment include: amplifiers, analyzers, filters, component testers, generators, oscilloscopes, power meters, RFI/EMI equipment, telecom/communications testers and others.
Bid Service
INFO/CARD #207

Lightning Protection Newsletter

The latest edition of *Striking News* from PolyPhaser contains articles on the effectiveness of tower down-conductors, lightning current, silicon avalanche diode power line protectors, and information on upcoming trade shows.
PolyPhaser Corp.
INFO/CARD #206

Cable Television Testing

Wavetek's full-color, 28-page Cable Television Selection Guide includes a full range of products designed to meet cable television's specific testing needs. The newly-expanded guide features Wavetek's newest frequency agile MicroStealth Installer/Service Technician Meters which measure channel signal levels with a display similar to the Stealth SAM 4040 and Stealth 3SR sweep receiver.
Wavetek Corp.
INFO/CARD #205

Test & Measurement Equipment

Tucker Electronics announces the release of their full line catalog. This free test and measurement product guide contains over 5000 new and used instruments plus hundreds of components and specifications.
Tucker Electronics
INFO/CARD #204

Waveform Synthesizers

Pragmatic Instruments has released their new Selection Guide for Waveform Synthesizers. Seven different arbitrary waveform generators with sample clock rates ranging from 100 MHz to 2 MHz, vertical waveform resolution from 12-bits to 16-bits, and waveform capacity up to 1 megaword are featured.
Pragmatic Instruments, Inc.
INFO/CARD #203

Communications Videotapes

George Washington University offers the Keiser Communication videotapes, covering high definition television (5 tapes), microwave radio systems (5 tapes), and satellite communication engineering principles (7 tapes). Prices for each series are $1,575 for the High Definition Television series, $1,710 for the Microwave Radio Systems series, and $2,205 for Satellite Communication Engineering Principles series.
The George Washington University
INFO/CARD #202

Used Test Equipment

GE Capital Test Equipment Management Services has released Volume 1 of the 1995 Buyer's Guide. The 13-page guide contains a list of quality used test equipment, grouped by manufacturer, with a brief description of each available model. Information is available by calling 1-800-GE-RENTS.
GE Capital
INFO/CARD #201

MMDS Amplifier Data

A four page data sheet introducing a series of 2.5 - 2.7 GHz solid-state PAs for MMDS (wireless cable) and ITFS (instuctional television) is now available from Chesapeake Microwave Technologies. This data sheet defines critical features and electrical specifications of these solid state, high power amplifiers.
Chesapeake Microwave Technologies, Inc.
INFO/CARD #200

IF/RF Catalog

Daico Industries announces the publication of its 1995-1996 IF/RF catalog. The all new 280-page catalog features DAICO's complete family of IF and RF components. DAICO specializes in the manufacturing of switches attenuators, phase shifters, MMICs, bit detectors, couplers and subassemblies. Over 45 new products are featured in the catalog.
Daico Industries, Inc.
INFO/CARD #199

Test & Measurement Catalog

LeCroy announces publication of the new 1996 Test & Measurement Catalog featuring product information, selected applications information and two tutorials on "Fundamentals of Digital Oscilloscopes" and "Waveform Creation Made Easy". Included is information on LeCroy's 1 GHz bandwidth oscilloscope line and the 9362, which has a 10 Gs/s sampling rate.
LeCroy Corp.
INFO/CARD #198

Connector Catalog

Applied Engineering Products has released a 185-page catalog, called the "Blue Book", describing their series of connectors, adapters, and cable assemblies. Besides showing standard connectors, the "Blue Book" shows mounting and cable attachement options, plating options, cable terminations, adaptors, and other features.
Applied Engineering Products
INFO/CARD #197

Signal Switching Handbook

Keithly Instruments has announced the availability of its 3rd edition of the *Switching Handbook*. The expanded and revised handbook is an application guide to signal switching in automated test systems. It provides over 200 pages of information on the theory and practice of automated switching in low level measurements.
Keithley Instruments, Inc.
INFO/CARD #196

Resistive Devices

EMC Technology has released a 98-page, four-color catalog describing their attenuators, terminations, resistors, connectors and surface mount hybrid couplers. The catalog also contains a special section hilighting new products and eight application notes.
EMC Technology, Inc.
INFO/CARD #195

Component Summary Sheet

Analogic announces a component summary sheet for its complete line of high performance data conversion component product lines.
Analogic Corp.
INFO/CARD #194

Integrated Design Programs

Eagleware has released GENESYS, an integrated set of programs for analog, RF and microwave circuit designers. GENESYS includes a high-speed linear circuit simulator, programs for the synthesis of a wide range of circuits, a schematic interface, and exportation of circuit files for Touchstone, Spice, and em, the electromagnetic simulation program from Sonnet Software. GENESYS is available for DOS, Windows 3.1 and Windows NT. Simulation packages start at $995; synthesis modules range from $599 to $999.

Eagleware Corp.
INFO/CARD #193

Site Management

SoftWright has introduced SiteBase™ for Windows™, a comprehensive communications site management program. A relational database for the user-friendly Windows interface, SiteBase tracks and updates equipment and other management information, as well as creating dozens of customized reports. Using Windows 3.0 or higher, SiteBase runs on 286 (or higher) IBM compatibles. Introductory price is $395.

SoftWright LLC
INFO/CARD #192

Resonator Selection

Trans-Tech offers CARD (Computer Aided Resonator Design) and COAX (Coaxial Element Designer Guide), Windows™-based software design tools that can assist RF designers in the selection of dielectric and coaxial resonators for oscillators and filters from UHF to Ku frequencies. CARD and COAX provide graphical presentation of the frequency response and equivalent circuit values for insight into the interrelationships among circuit components. Both programs relate component material geometry to predicted performance using a wide choice of materials and models based on proven, established development formulas.

Trans-Tech, Inc.
INFO/CARD #191

SPICE for Networks

Intusoft has introduced a keyless version of its popular SPICE 3F based simulator IsSpice4. The software is licensed for a specific number of copies that "float" on a network. The software may be installed on any machine in a network which acts as the server. Any client machine on the network can run the ICAP/4 software, and special provisions have been made for clients to have their own private model libraries, symbols, and preferences. Network simulation performance is as good as that of stand-alone versions. "Dongleless" network versions of the ICAP/4 Windows™, ICAP/4Lite and ICAP/4Lite Xtra packages are priced according to the formula: list price × (N+1), where N is the maximum number of copies to be run at any one time.

Intusoft
INFO/CARD #190

Block-Level Simulator

Visual Systems is shipping VisSim v2.0, a modeling and nonlinear simulation software package. Vis Sim allows users to simulate many types of systems by selecting and connecting icons from a tool bar. RF systems can be simulated using a number of high frequency block models, including Hilbert, Bessel, Chebyshev, Butterworth or inverse Chebyshev filter types. Virtual control panels can be constructed, with user-definable charts, meters, indicator "lamps" and alarms. VisSim operates under Microsoft Windows, Windows NT and UNIX/X operating systems. Prices range from $495 to $4,195 depending upon capability and operating platform.

Visual Solutions Inc.
INFO/CARD #189

RF Design Software

September Disk — RFD-0995
"Phase Noise Measurement for Under $250" by Bill Suter. Wave analysis software used with the signal acquisition hardware described in the article. Takes output of A/D converter, applies a Hanning window and performs a FFT. Displays the system noise from 1 kHz to 100 kHz. (Quick C, source code and compiled, executable version. See notes on program usage in article)

August Disk — RFD-0895
"Microstrip Coupled-Line Bandpass Filter Synthesis and Analysis Program" by Sean Mercer and Eric Mabada. The user enters performance specifications, and the program generates a microstrip filter that meets those requirements. Analysis data is created in tabular form; Touchstone and Super Compact files can be generated for further analysis. (Requires QBASIC interpreter in MS-DOS 5 or higher, VGA or SVGA graphics)

ALSO AVAILABLE:
Index of *RF Design* Articles: 1978-1994 — Disk RFD-INDEX — $25.00 (U.S.)
Amplifier and Oscillator Design Program — RFCAD — $99.00 (U.S.)

Programs from RF Design provided on disk for your convenience
Monthly program disks: $25.00 (U.S.) $30.00 (foreign)

Yearly Disk Sets and Annual Subscriptions Available: Call for Information

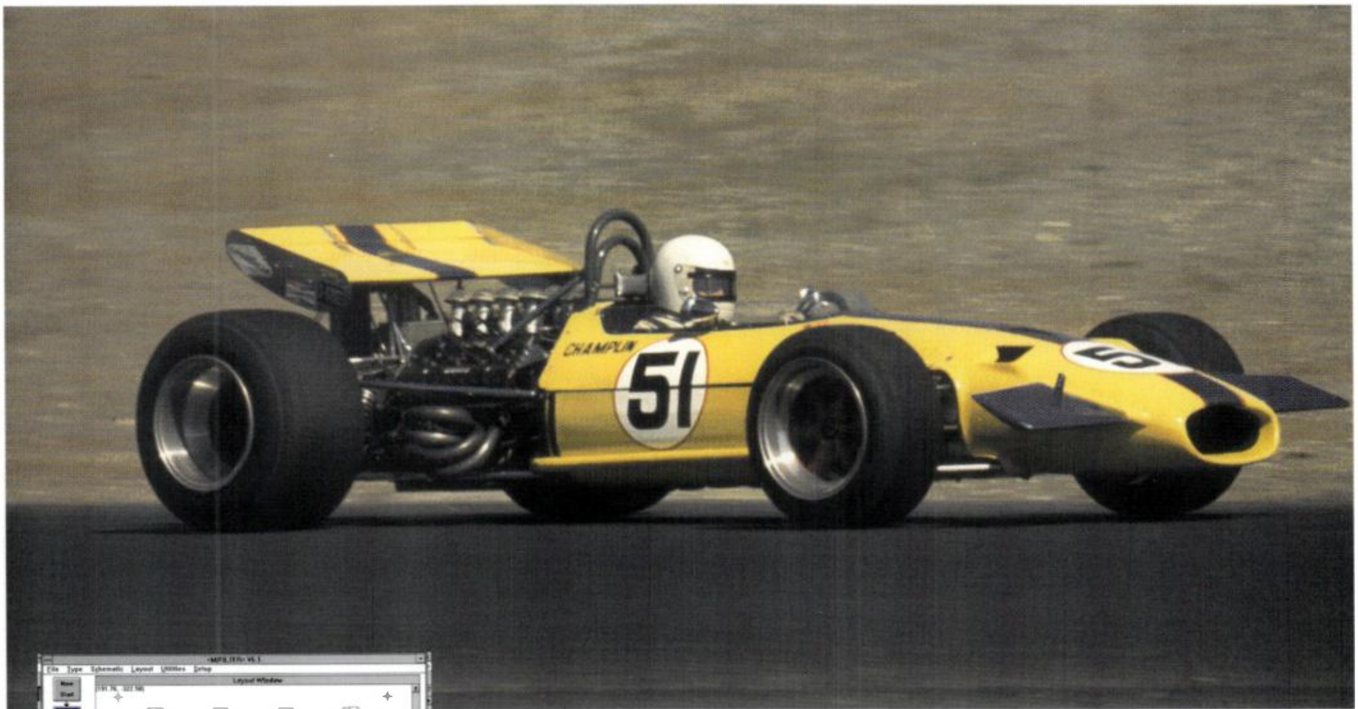
All orders must be pre-paid by check, money order, or major charge card. All checks must be in U.S. dollars and payable to a bank located in the U.S. *Prices include shipping.*

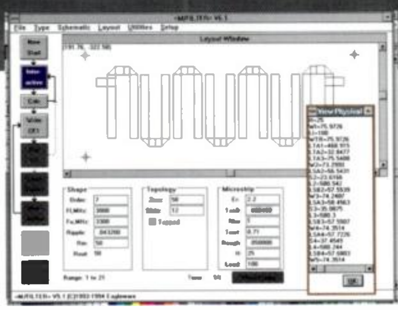
RF Design Software
Argus Direct Marketing Dept.
6151 Powers Ferry Rd., N.W.
Atlanta, GA 30339-2941
Tel: (404) 618-0398
Fax: (404) 618-0347

RF guide to editorial coverage

RF advertising index

TURBO-CHARGE TOUCHSTONE





EXAMPLE GENESYS SCREEN

SUPER-CHARGE SPICE

with GENESYS, a suite of programs for synthesizing microwave filters amplifiers, oscillators, active filters, matching networks and L-C filters for your Touchstone, Spice and =SuperStar= simulator.

MICROWAVE FILTERS
End coupled
Edge coupled
Hairpin
Stepped-Z
Combline
Interdigital
Elliptic lowpass
Elliptic bandpass
Sub lowpass
Stub bandpass
Stub bandstop
Edge bandstop

OSCILLATORS
L-C series mode
L-C Colpitts
L-C Clapp
T-line and L-C VCO
VCO with xformer
Cavity bipolar and hybrid
Dielectric resonator
Terminal SAW bipolar
Port SAW hybrid
Port SAW MOSFET
Pierce and Colpitts crystal
Driscoll crystal
Butler overtone
Overtone with multiplier

MATCHING NETWORKS
L-C pi and L
L-C tee
T-line quarter wave
T-line single/double stub
General order bandpass
L-C pseudo lowpass
T-line pseudo lowpass
T-line stepped-Z
Custom with R' s and xformers

LUMPED FILTERS
Conventional all-pole
Conventional elliptic
Top-C coupled
Top-L coupled
Shunt-C coupled
Tubular
Blinchikoff flat delay
Zig-zag
Eagleware symmetric

ACTIVE FILTERS
GIC transform
Single feedback
Multiple feedback
Low sensitivity
State variable (biquad)
VCVS
Dual amplifier

FOR SIMULATION ENGINES..
GENESYS
TOUCHSTONE
SPICE
SUPERSTAR

Touchstone is a product of HP/EEsof GENESYS and =SuperStar= are products of Eagleware

GENESYS includes free technical support, no annual fees and a money-back guarantee GENESYS is available for

MS DOS
MICROSOFT WINDOWS
MICROSOFT WINDOWS NT

EAGLEWARE
Eagleware Corporation ★ 1750 Mountain Glen ★ Stone Mtn, GA 30087 ★ USA
TEL (404) 939-0156 ★ FAX (404) 939-0157

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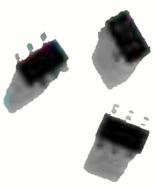
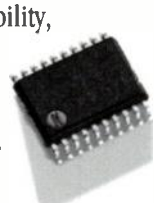
DIRECT SALES & USER SUPPORT
BY FAX, PHONE OR LETTER

INFO/CARD 68

...MORE NEW 3 VOLT RF ICs FOR YOUR HANDHELD WIRELESS PRODUCTS



NEC's 3 Volt ICs not only reduce power consumption in your wireless designs, they help reduce your parts count, improve the reliability, and simplify assembly.



We're committed to 3 Volt Silicon technology. These parts are just a sampling of the wide variety of the semiconductors we stock.

We also stock 5 Volt ICs — plus NEC's huge family of discrete devices. Many of these transistors are characterized at 1 to 3 Volts, and most are available in a number of package styles, including the new ultra-miniature 0.8 X 1.6 mm NEC "19" package.

For a FREE Product Selection Guide, call your nearest CEL Sales Office. Or circle the number below.

Dozens of 3 Volt Amplifiers, Prescalers, Up & Downconverters, IQ Mods & Demods, and Transistor Arrays, including:

UPC8104GR new
IQ Modulator
with Up Converter
800 MHz to 2.4 GHz
2.7 to 5.5 V Supply Voltage

UPC8106T/09T
Up Converters
800 MHz – 2.4 GHz
7 dB CG @ 1.9 GHz
2.7 to 5.5 V Supply Voltage

UPC2758T
Downconverter
100 MHz – 2.4 GHz
17 dB CG @ 2.0 GHz
+6 dBm IP3

UPC2762T/63T
Driver Amplifiers
100 MHz – 1.9 GHz
14 or 20 dB Gain and
+7 dBm P1dB @ 1.9 GHz

UPC2768GR new
Downconverter
DC – 450 MHz
80 dB Conversion Gain
On-chip Oscillator

UPC2771T new
Wideband Amplifier
100 MHz – 2.1 GHz
21 dB Gain and
+11.5 dBm P1dB @ 900 MHz

...plus a wide variety of Silicon Devices for 5 Volt applications, including:

UPC2709T
Wideband Amplifier
100 MHz–2.3 GHz
23 dB Gain
+8 dBm P1dB @ 900 MHz

UPC2766GR new
IQ Demodulator
35 dB AGC Dynamic Range
Flat Response to 1 GHz RF
DC–100 MHz I/Q Output

UPC2721GR
High Compression Point
Downconverter
900 MHz – 2.0 GHz
21 dB Conversion Gain

UPB1505GR new
Prescaler
÷64/128/256
3 GHz Guaranteed RF Input
High Input Sensitivity

...plus dozens of discrete transistors including three new small signal bipolar series:

new

Part*	NF (dB)	GA (dB)	V/IC	fTEST(GHz)
NE68619	1.7	10.0	2V/3mA	2.0
NE68719	1.5	8.0	1V/3mA	2.0
NE68819**	1.6	8.0	3V/7mA	2.0

* Available in six different plastic packages, including the ultraminiature NEC "19" package
** Low Phase Noise Device

CEL California Eastern Laboratories

NEC

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