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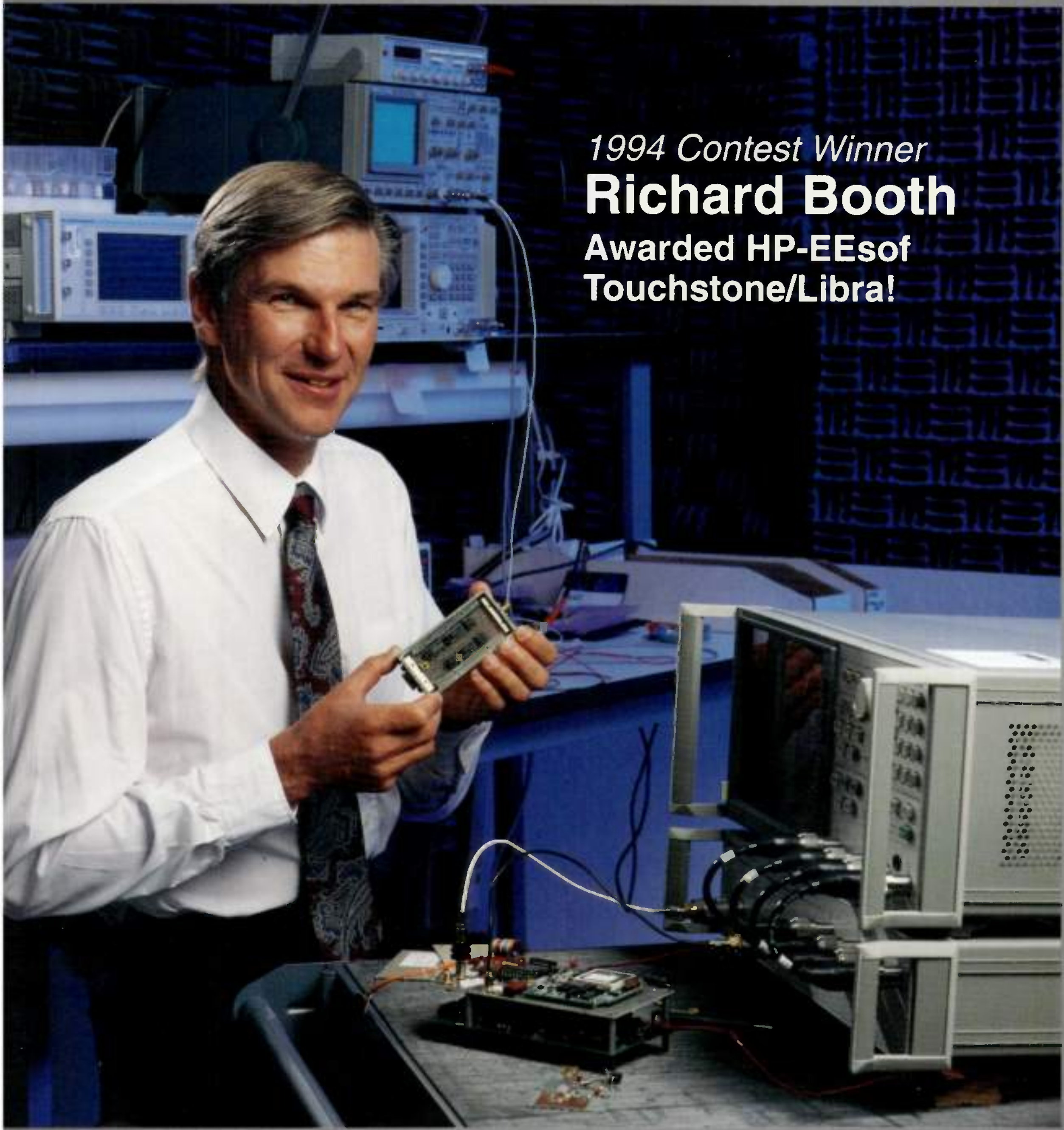
engineering principles and practices

November 1994



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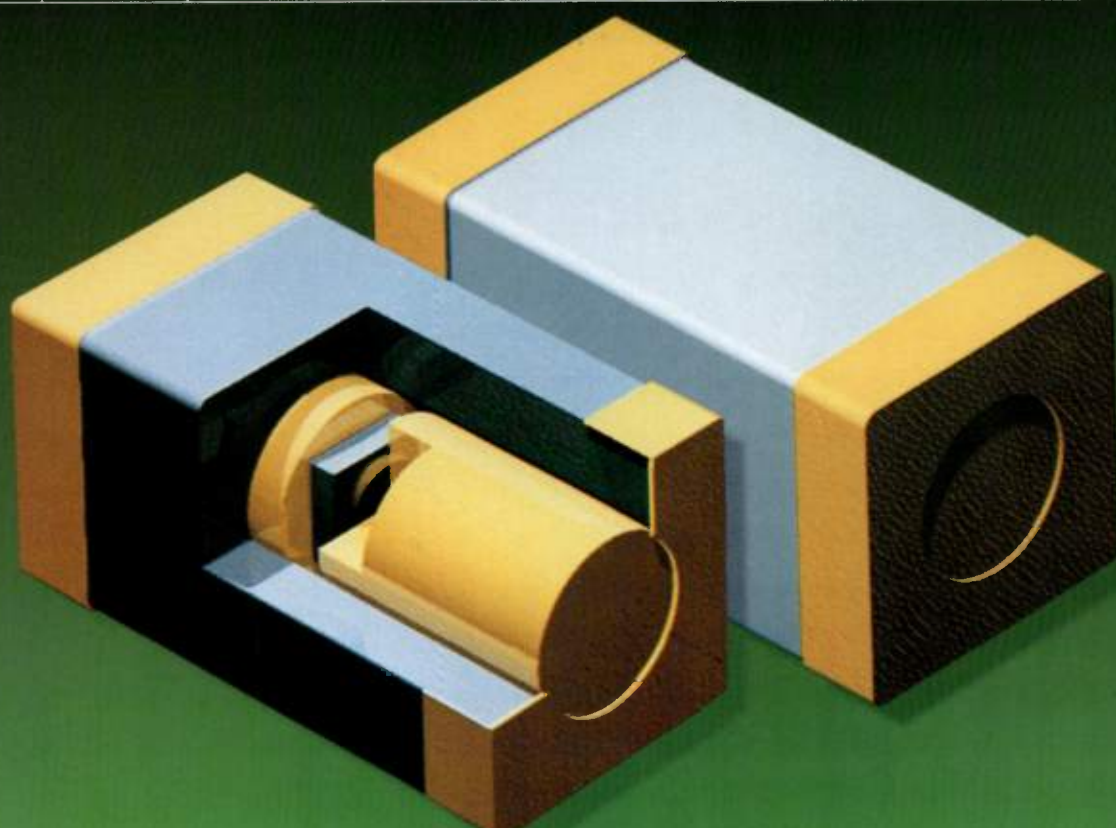
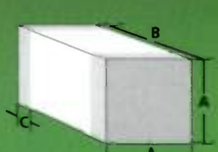
Part No.	Suggested Case Style	Reverse Voltage Rating (Vr) – min	Total Capacitance (pF) – max F=1MHz Vr=50v		Series Resistance (RS)–ohms F=100 MHz		Carrier Lifetime (Tl) μsec (typ)	Thermal Resistance (Θj)
		Ir< 10μA	M1	M2	If=100mA (max)	If=200mA (typ)	If=10mA	C/W
SM0502	M1 or M2	500	0.45	0.50	0.70	0.55	1.00	35
SM0504	M1 or M2	500	0.55	0.60	0.60	0.45	1.50	20
SM0508	M1 or M2	500	0.85	0.95	0.40	0.25	2.00	15
SM0509	M1 or M2	500	1.00	1.25	0.35	0.20	2.50	12
SM0511	M1 or M2	500	1.30	1.45	0.30	0.15	3.00	9
SM0512	M2	500	N/A	1.50	0.25	0.12	3.50	7
SM0812	M2	800	N/A	1.30	0.40	0.25	4.00	7
SM1001	M2	1000	N/A	1.30	0.35	0.20	4.50	7





Case style	M1 (inches)		M2 (inches)	
Dimension	Min	Max	Min	Max
A	0.08	0.10	0.10	0.12
B	0.12	0.14	0.19	0.21
C	0.01	0.03	0.01	0.03

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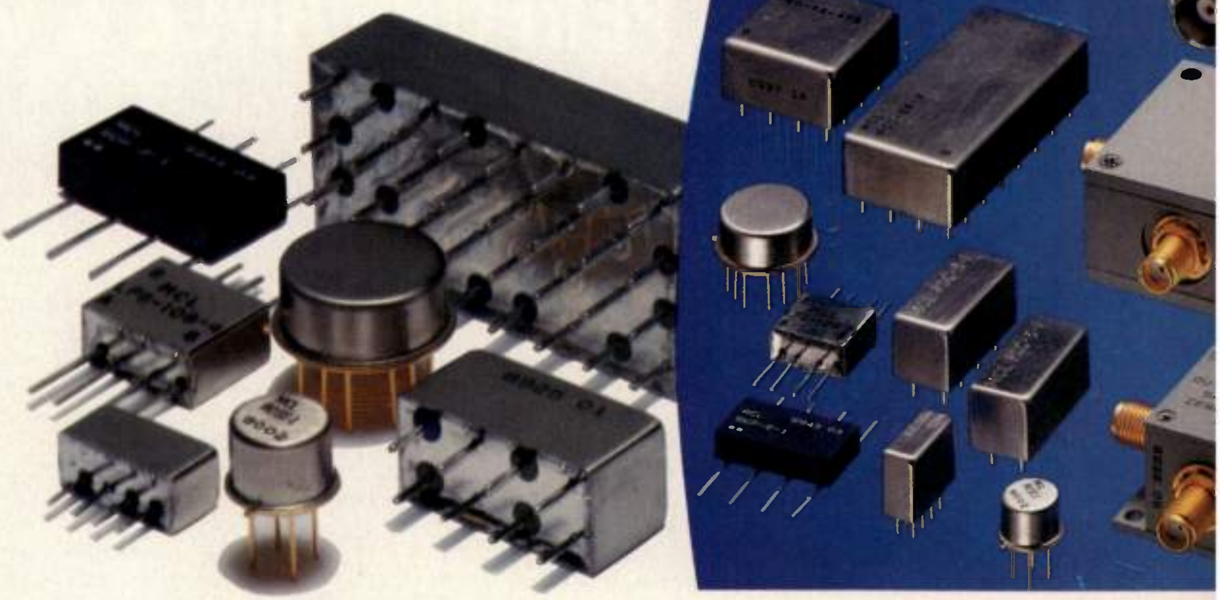

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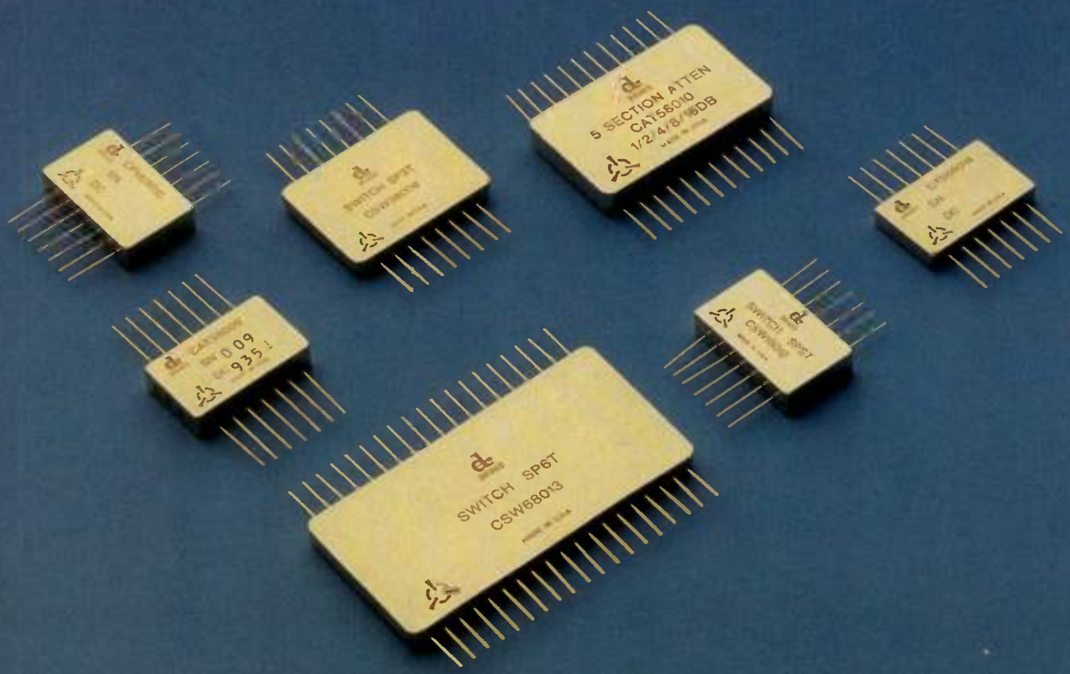

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Con-fig	Freq MHz	IL dB	Iso dB	Switch Speed n SEC	Part No.
SPST	10-1000	1.25	70	35	8012
SPST	400-2000	0.9	67	185	DS0551
SP2T	10-1000	0.7	55	40	DS0860
SP2T	5-2000	1.0	57	24	DS02800
SP2T	5-1000	0.5	70	72	DS0962
SP3T	10-1000	1.2	40	100	8008
SP4T	DC-2000	1.5	50	30	CS048024
SP6T	10-1000	0.75	50	1000	8013
1 Sect Atten	5-1000	0.9	10	30	DA0944-10
1 Sect Atten	800-1200	0.9	50	80	DA0879
3 Sect Atten	30-300	1.0	.25, .5, 1.0	200	8009
5 Sect Atten	10-400	1.7	1, 2, 4, 8, 16	100	8010
Pulse Modulator	10-1000	2	80	300	8015
Threshold Det	10-3000	Detects -25 to -10 dBm		—	8014

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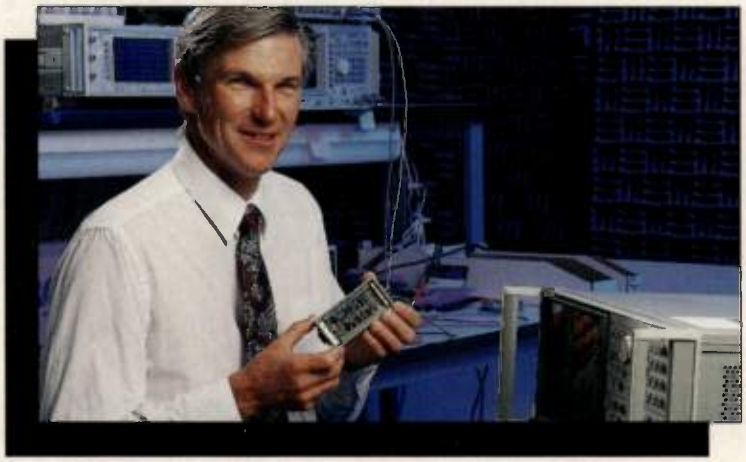
contents

November 1994

cover story



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RF editorial

Can You Explain RF to Non-Engineers?

By Gary A. Breed
Editor

I recently had to explain RF to several non-technical colleagues. The lesson started out all right; it wasn't hard to list a lot of common places where RF is used — broadcasting, radar, cordless and cellular phones, cable TV, etc. However, I didn't have much success with my explanations of the electromagnetic spectrum, radiated waves, or circuit operation. I'm still working on the right way to represent RF to a lay audience, because too few of them grasp basic concepts.

You might ask why they need know. After all, most RF applications are intended to be invisible to the user. Television is just moving pictures; broadcast, cable or VCR sources for those pictures are supposed to be interchangeable. Cellular phones are just one of many ways to get your voice from here to there. The telephone user doesn't care whether his or her conversation is transmitted by copper wire, fiber optic cable, microwave, satellite, 900 MHz radio — or two soup cans and a string.

That invisibility is exactly what I am worried about! If folks don't know how unique and challenging RF technology can be, we'll be like Rodney Dangerfield and "won't get no respect." Right now, a lot of engineers and investors who don't understand RF are getting into the RF business. They come from the computer industry, from traditional wireline telephone, transportation, medicine or other backgrounds. All they want is a wireless replacement for previously hard-wired interconnections.

If the RF industry delivers an invisible replacement, I want the customer to know that it wasn't as easy as it looks. Innovative modulation techniques, diversity antennas, digital signal processing, and years of propagation research could be behind that $20 two-chip wireless design solution.

And, if the wireless version of their product comes up against real-world limitations on data rate, transmission distance or interference, I want the customer to understand that the laws of nature can't be repealed. The Directors running a high-risk startup company need to know when they are asking too much of RF technology (or of the government regulations that control its use)!

Your Ideas Are Needed

Help me out here! Every one of us has an analogy or a comparison to more familiar technology that seems to get through to people. In my case, comparisons to sound and light seem to help when explaining how radio waves travel through the ether. After all, sound and light can be heard and seen, unlike RF. But I really lose an audience when trying to explain how RF signals travel through components in a circuit and how they are modulated and detected.

Send in your best layman's explanation of an RF concept. I'll collect them for a month or so and publish the best ones in the February issue. Mail or fax them to me by December 15 (use the address at the top of page 10). Just to thank you for helping out, we'll put all the names in a hat and draw one out for a prize. That person will get a complete set of RF Design Handbooks, including the new Low-Noise Design Handbook coming out this month. These collections of past articles are excellent references on key design areas.

In the meantime, you can do your part to increase awareness of RF by patiently answering questions from the management, marketing, and other non-engineers that work with you.

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RF letters

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Terms not so Obfuscatory

Editor:

Without doubt we need more clear and simple language, but I don't see that clear enumeration of ideas and principles would help. It is clearer expression or enunciation we need, rather than clearer lists.

I think I know what "national information infrastructure" and "portable" and "personal communications" mean. These terms are formed from words precisely defined in any dictionary, and the last two are internationally agreed upon in organizations like CCIR. "Wireless" is another matter, meaning different things in different places, but it is not an obfuscatory word, is it! "Information superhighway" is a clear and simple simile, and when people say they don't understand what it means, they probably really mean they don't quite see how it will be built. I think it is a brilliant expression, and it has certainly captured the imagination of many people.

I once had a boss who said, "People who cannot express themselves clearly in writing cannot be thinking very clearly either." But not many people are intentionally vague, and I don't think they are trying to cover up anything. They just fall into bad habits. Even the erudite commentators these days are using the word "characterize" when they mean "describe".

Bob Eldridge
Pemberton, B.C.

Unequally Terminated Filters

Editor:

The September issue arrived today, and it did not take me long to zero in on Dr. Ellis' filter article starting on pg. 78.

Analysis of the schematic in Figure 2 reveals that it is indeed the filter he says it is, a 6th order Cauer Elliptic with 1 dB ripple and 50 dB stopband returns. I have tried analyzing the schematic of Figure 3, reputed to have the same characteristics as that of Figure 2, but now terminated in 2 ohms. I tried my own analysis program, and I tried George Szentirmai's S/FILSYN, and I get identically the same results from both tries, but unfortunately I find that it is no longer a Cauer Elliptic filter.

Notice in Figures 2 and 3 that there are shunt branches consisting of series resonant circuits. Notice that in both figures, an identical capacitor of 1.14298 F appears in the branches similarly located. The resonance frequency in both cases is shown as 1.273134, but it is physically impossible for a 1.14298 F capacitor to resonate at that same frequency with two different inductors.

[The capacitor value in question turns out to be a misprint. The correct value should be 0.484. -Ed]

William B. Lurie
Boynton Beach, FL

Editor:

The article "A Design Method For Unequally Terminated Elliptic Filters" by Michael G. Ellis showed a closed form method of changing an evenly terminated prototype elliptic filter into an unevenly terminated filter with the same characteristics. It was stated that optimization routines should be avoided because they almost always converge into a sub-optimal solution. I disagree with this conclusion.

I ran an analysis of the evenly terminated prototype, using the circuit simulator NOVA-586. The filter behaved as stated, the ripple was 1.00 dB to 0.159 Hz and the stopband attenuation was 50 dB from 0.199 Hz and above.

I changed the termination to 2 ohms, set the optimization goals to the above performance, and ran the optimization. The optimization took about 60 seconds. The results were: 0.86 dB ripple to 0.159 Hz, stopband attenuation was 50.0 dB from 0.199 Hz and above. The final optimized circuit file and tabular results are included. In this case the unevenly terminated (optimized) filter exceeds the performance of the prototype.

Optimizers have the advantage that they can work with simulated "real" components which have finite Q, lead inductance, and package capacitance. Optimizers can compensate for small response changes caused by real components, giving a final result that is closer that can be obtained from design formulas alone.

Robert Stanton
Hatboro, PA

My copy of the article "A Design Method for Unequally Terminated Elliptic Filters", as submitted to *RF Design*, shows the correct value of 0.484 farads, instead of the 1.14298 farads, for the capacitor in question in Figure 3.

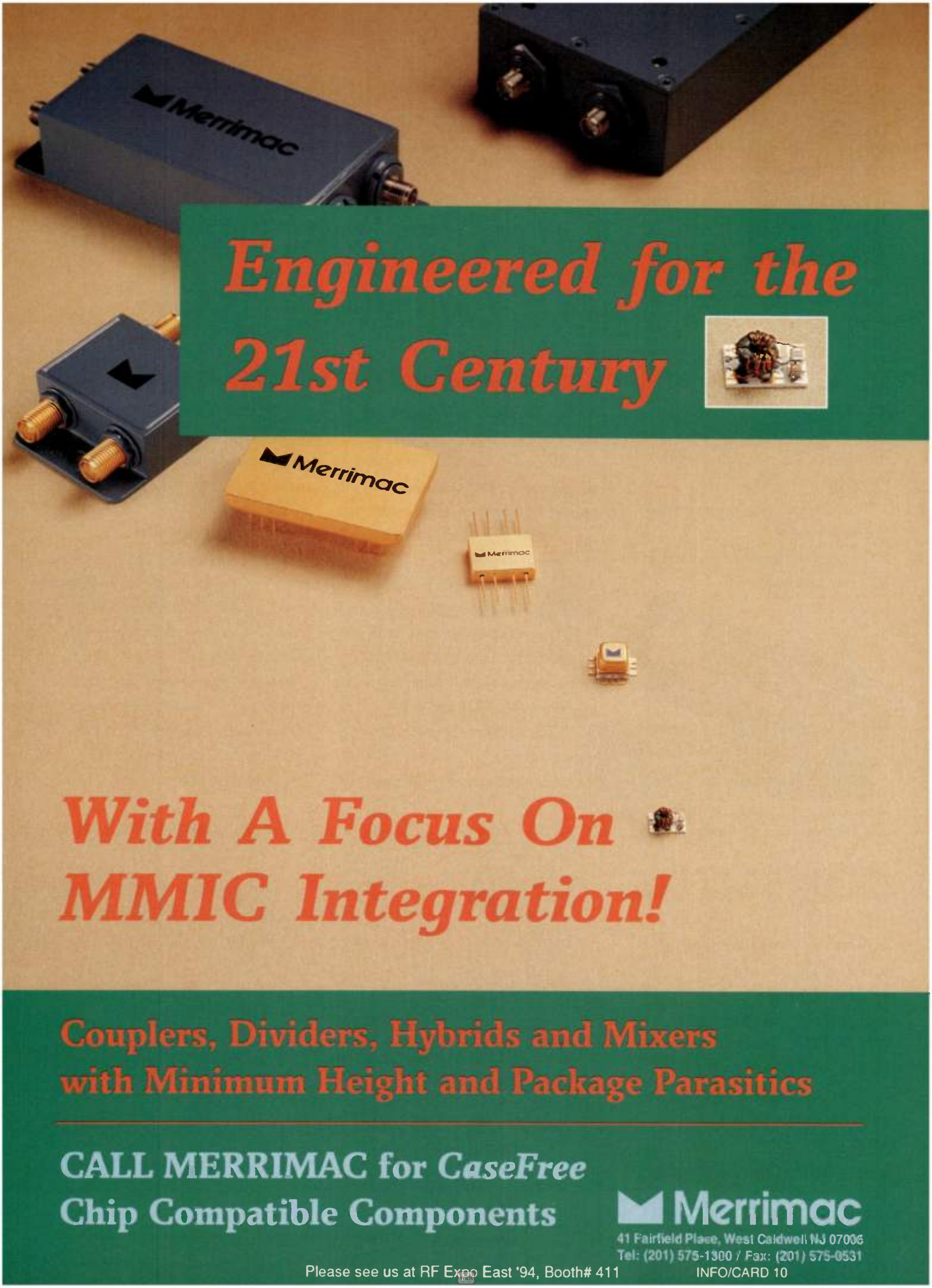
Mr. Stanton got essentially the same results with his optimizer. His circuit is slightly better in the passband, and slightly worse in the stopband. Since iterative techniques cannot be guaranteed to converge to a global minimum, it is always safer to use a closed form method, assuming one exists. Optimization can still be applied as a final step.

Michael Ellis
Research Electrical Engineer

Comp.	Type	Value		Node	Node	Node	Node	Optm.
1	Generator	1.00000	Ohms	0	1			
2	Inductor	1.18370	H	1	2			x
3	Capacitor	1.20622	F	2	3			x
4	Inductor	0.31353	H	3	0			x
5	Inductor	1.28926	H	2	4			x
6	Capacitor	1.09709	F	4	5			x
7	Inductor	0.56283	H	5	0			x
8	Inductor	1.55146	H	4	6			x
9	Capacitor	1.16464	F	6	0			x
10	Resistor	2.00000		6	0			

Frequency	S11		S21		S12		S22	
	dB	(deg)	dB	(deg)	dB	(deg)	dB	(deg)
0.0100 Hz	-9.8	-24.0	-0.5	-13.2	-0.5	-13.2	-9.8	177.6
0.0300 Hz	-11.6	277.3	-0.3	-41.0	-0.3	-41.0	-11.6	180.8
0.0500 Hz	-11.7	194.8	-0.3	-71.2	-0.3	-71.2	-11.7	202.8
0.0700 Hz	-8.7	128.0	-0.6	-102.3	-0.6	-102.3	-8.7	207.3
0.0900 Hz	-7.5	82.1	-0.8	-133.5	-0.8	-133.5	-7.5	190.9
0.1100 Hz	-11.2	41.1	-0.3	-170.5	-0.3	-170.5	-11.2	158.0
0.1300 Hz	-14.2	141.6	-0.2	135.7	-0.2	135.7	-14.2	-50.3
0.1500 Hz	-9.0	73.0	-0.6	67.9	-0.6	67.9	-9.0	242.8
0.1700 Hz	-0.2	126.6	-13.0	-70.3	-13.0	-70.3	-0.2	272.9
0.1900 Hz	-0.0	89.9	-36.2	-103.7	-36.2	-103.7	-0.0	242.8
0.2100 Hz	-0.0	74.4	-52.0	62.5	-52.0	62.5	-0.0	230.5
0.2300 Hz	-0.0	64.6	-51.6	53.8	-51.6	53.8	-0.0	223.1
0.2500 Hz	-0.0	57.5	-62.4	47.7	-52.4	47.7	-0.0	2i7.8
0.2700 Hz	-0.0	52.0	-61.7	-137.1	-61.7	-137.1	-0.0	213.9
0.2g00 Hz	-0.0	47.6	-54.4	-140.8	-54.4	-140.8	-0.0	210.8
0.3100 Hz	-0.0	43.9	-51.8	-143.9	-51.8	-143.9	-0.0	208.3
0.3300 Hz	-0.0	40.8	-50.7	-146.5	-50.7	-146.5	-0.0	206.2
0.3500 Hz	-0.0	38.2	-50.1	-148.7	-50.1	-148.7	-0.0	204.4
0.3700 Hz	-0.0	35.9	-50.0	-150.7	-50.0	-150.7	-0.0	202.8
0.3900 Hz	-0.0	33.8	-50.0	-152.3	-50.0	-152.3	-0.0	201.5

Table 1. Netlist and simulation results for Stanton's filter.



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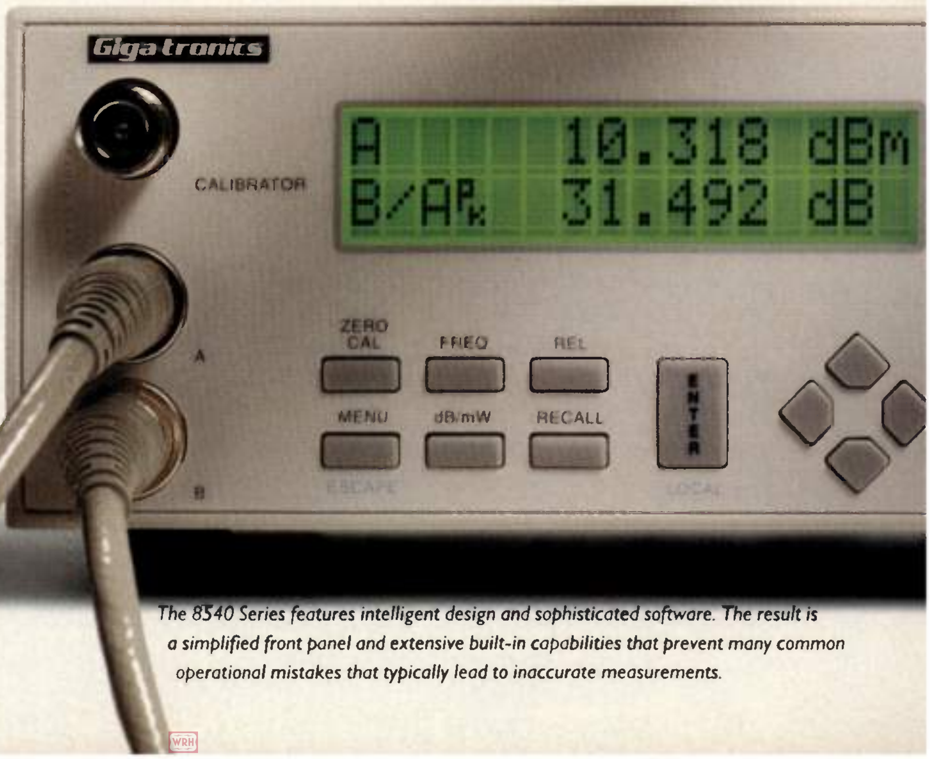
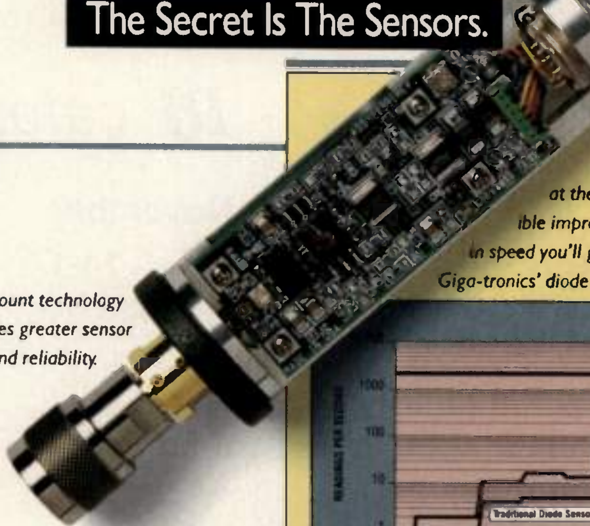
View the pulsed signal's amplitude profile on a scope and see the exact power measurement point on the pulse. Measure the overshoot. Measure the droop.

The two-line display also lets you set the desired resolution and select either Lin or Log readout for each line.



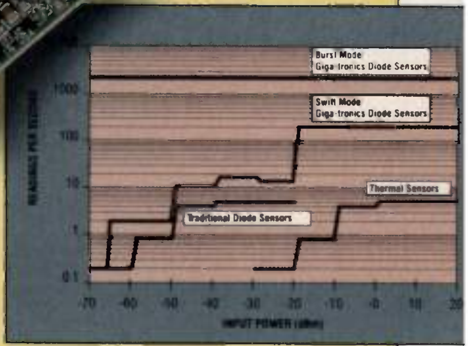
The 8540 Series features intelligent design and sophisticated software. The result is a simplified front panel and extensive built-in capabilities that prevent many common operational mistakes that typically lead to inaccurate measurements.

The Secret Is The Sensors.



Surface Mount technology assures greater sensor accuracy and reliability.

Just look at the incredible improvement in speed you'll get with Giga-tronics' diode sensors.



You'll be confident of your peak power readings, and still have all the benefits of an incredibly fast CW power meter.

ONE OR TRUE TWO CHANNEL OPERATION

If a single-channel meter is what you need, the Model 8541 is the meter for you. But if you need two-channel capability, the Model 8542 lets you see readings from both channels *simultaneously.*

SIMPLE, INTELLIGENT OPERATION

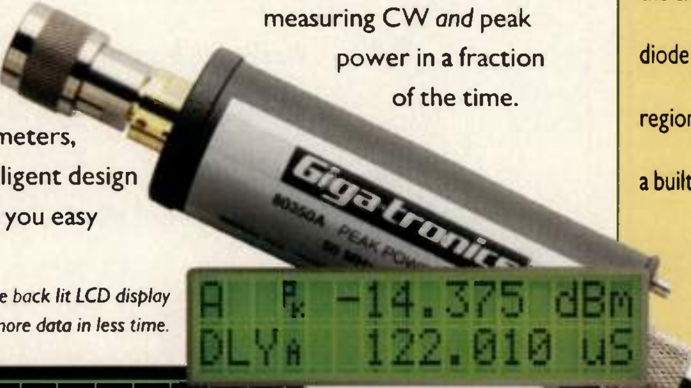
The 8540 Series has only half as many controls as other power meters, but don't let that fool you. Intelligent design and sophisticated software give you easy access to extensive built-in capabilities.

For example, you use the same key to zero and calibrate the power sensors. The meter automatically determines the function you want by detecting whether a sensor is connected to the calibrator.

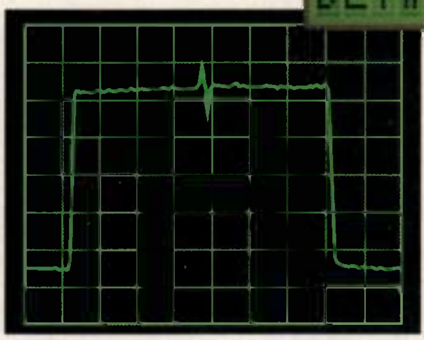
Imagine all this power and performance. But why just imagine? Get the truly incredible Giga-tronics 8540 Series Universal Power Meter, and start measuring CW *and* peak power in a fraction of the time.

The Giga-tronics 8540 Series delivers incredible performance by taking full advantage of the speed and dynamic range of diode sensors.

What's more, Giga-tronics has solved the challenge that previously limited diode sensors to the "square law" region—below −20 dBm—by utilizing a built-in power sweep calibration system. So you get speed and a full 90 dB dynamic range without sacrificing accuracy.



A two-line back lit LCD display provides you more data in less time.



The peak sensor adds a marker on a monitor output for setting an exact measurement point on pulse signals.

Call us toll-free at **1 800 726 GIGA.** Outside the U.S. call your local Giga-tronics representative.

We'll send you more information or arrange for an incredible hands-on demonstration.

Giga-tronics

Giga-tronics Incorporated
4650 Norris Canyon Road
San Ramon, California 94583
Telephone: 800 726 4442 or
510 328 4650
Telefax: 510 328 4700

Please see us at RF Expo East '94, Booth# 702

INFO/CARD 11

RF calendar

November

14-17 **RF Expo East**
Orlando, FL
Information: RF Expo East, Registration Coordinator, 6151 Powers Ferry Rd. NW, Atlanta, GA 30339. Tel: (800) 828–0420. Fax: (404) 618–0441.

15-17 **International Society for Hybrid Microelectronics Symposium**
Boston, MA
Information: ISHM - The Microelectronics Society, 1850 Centennial Park Drive, Suite 105, Reston, VA 22091. Tel: (703) 758–1060. Fax: (703) 758–1066.

December

6-8 **Electronic Contracting Exposition and Conference**
Las Vegas, NV
Information: Contract Manufacturers Association, 3310 W. Big Beaver Rd., Suite 403, Troy, MI 48084. Tel: (313) 643–6807. Fax: (313) 643–0856.

11-14 **1994 IEEE International Electron Devices Meeting**
San Francisco, CA
Information; Melissa Widerkehr, IEDM, 1545 18th Street NW, Suite 610, Washington, DC 20036. Tel: (202) 986–1137.

January

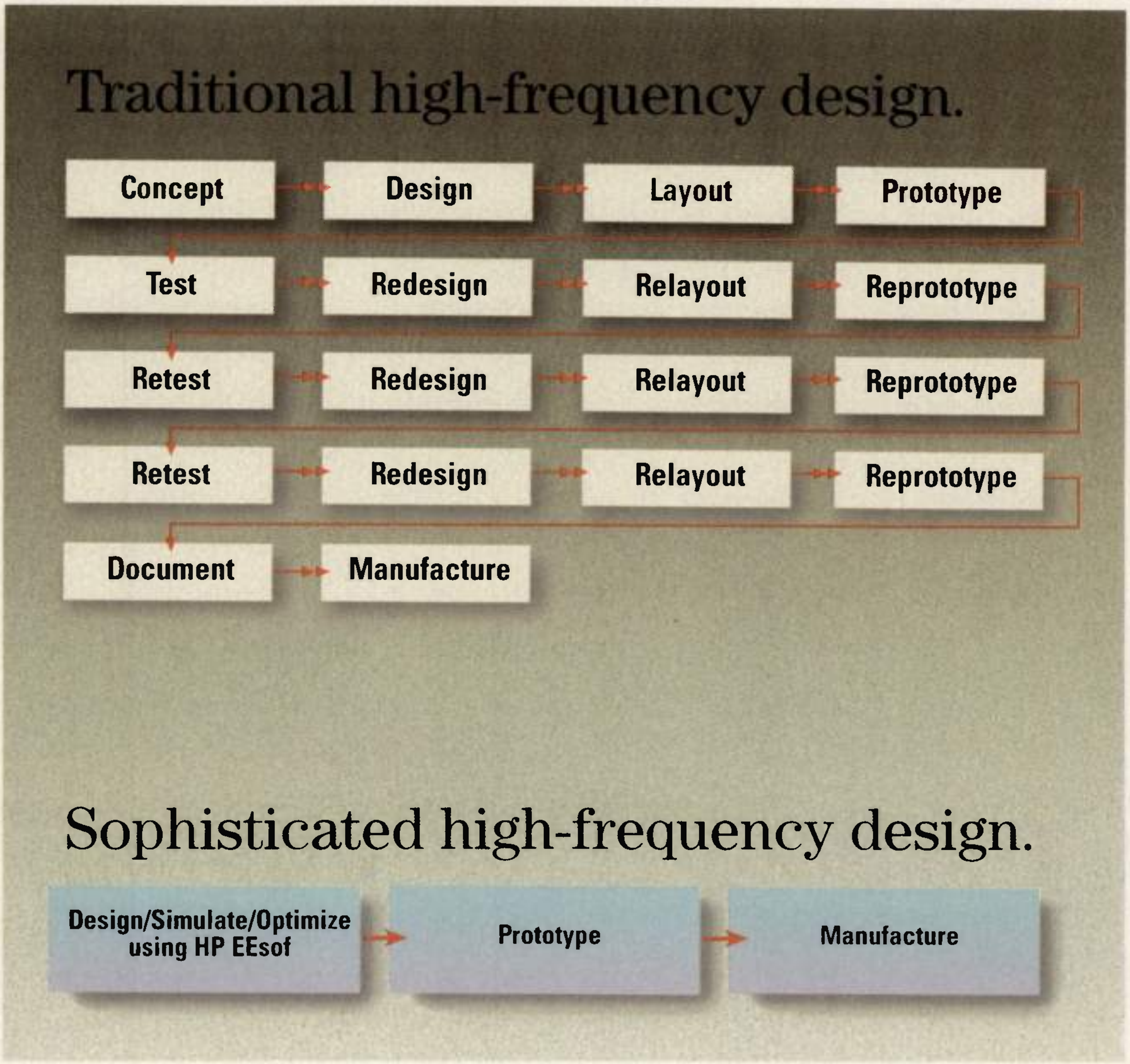
5-6 **Plastics in Portable Electronics**
Las Vegas, NV
Information: Ms. Deborah Cawley, SPE, 14 Fairfield Drive, Brookfield, CT 06804. Tel: (203) 775–0471. Fax: (203) 775–8490.

16-18 **Second Annual Mobile Communications '95 Conference**
Dallas, TX
Information: Frost & Sullivan Conference Division, 26524 Golden Valley Rd. Suite 401, Santa Clarita, CA 91350. Tel: (800) 256–-1076.

23-26 **ComNet '95**
Washington DC
Information: IDG World Expo. Tel: (800) 225–4695 or (508) 879–6700.

29-1 **RF Expo West**
San Diego, CA
Information: RF Expo West, Registration Coordinator, 6151 Powers Ferry Rd. NW, Atlanta, GA 30339. Tel: (800) 828–0420. Fax: (404) 618–0441.

29-1 **EMC/ESD International**
San Diego, CA
Information: RF Expo West, Registration Coordinator, 6151 Powers Ferry Rd. NW, Atlanta, GA 30339. Tel: (800) 828–0420. Fax: (404) 618–0441.



Simple, isn't it?

Designing RF and microwave circuits? Now you can increase the sophistication of these high-frequency designs. And make the process easier at the same time. How? It's simple, really.

MICROSOFT® WINDOWS™ COMPATIBLE

HP EEsof's Windows-compatible software lets you develop, simulate, and optimize circuit performance on your PC. You'll spend less time testing, redesigning, and retesting your prototype.

You'll reduce design cost, and decrease your time to market.

Our integrated design solution offers schematic capture and links to layout, powerful linear and nonlinear frequency domain circuit simulators, and unparalleled model accuracy.

With HP EEsof, easily analyze S-parameters, intermodulation distortion, mixer conversion efficiency—and much more—all on your PC.

To find out more about HP EEsof's complete line of PC- and workstation-based high-frequency simulation software, call **1-800-343-3763**. We'll make your life a whole lot easier.

There is a better way.

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INFO/CARD 13

Please see us at RF Expo East '94, Booth# 508

With the fierce competition in today's wireless markets, you need efficient tests. Every lost minute costs you market share and profit. That's why it's so important that you investigate the test solutions from Noise Com.

Real Time Testing for CDMA, GSM, and DCS1800.

Our Dual-Channel Multi-path Emulator

applies fading to your wireless products. With its built-in peak power meter (AGC) and synthesizer option, it's powerful enough to handle the wide dynamic range for forward and reverse link testing of CDMA, plus up to 12 paths for GSM and DCS1800 TDMA systems.

For amplifier evaluation, our Amplifier Test Station measures

IM distortion, noise, and gain up to 40 GHz with exceptional accuracy using the NPR technique.

We're Better at BER.

BER testing can consume huge amounts of time, so Noise Com designed an instrument to let you do it faster. Our UFX-BER Series Precision Carrier to Noise Generators set

RF courses

Active Circuit Design for Wireless Systems: Principles and Applications
November 28 - December 2, 1994, Los Angeles, CA
Information: UCLA Extension, Engineering Short Courses, 10995 LeConte Ave., Ste. 542, Los Angeles, CA 90024. Tel: (310) 825–1047. Fax: (310) 206–2815.

Equipment Sensitivity & Protection with Lab
November 30-December 2, 1994, Santa Clara, CA
Electromagnetic Fields - An Overview
December 5, 1994, Santa Clara, CA
Power Quality: Problems, Analysis & Solutions
December 6-8, 1994, Santa Clara, CA
Information: PowerCET Corporation, 2700 Augustine Drive, Suite 178, Santa Clara, 95054. Tel: (408) 988–1346. Fax: (408) 988–4869.

HALT and HASS Product Reliability Testing
December 5, 1994, Denver, CO
December 15, 1994, Seattle, WA
Information; Betty Meyers or Virginia Hobbs, Hobbs Engineering Corporation. Tel: (303) 465–5988.

European Compliance Seminar - Safety & EMC
November 29, 1994, St. Louis, MO
European Compliance Seminar - Medical
November 30, 1994, St. Louis, MO
Information: European Compliance Seminars, ICC D/FW Headquarters. Tel: (817) 491–3696.

European Compliance Seminar - Safety & EMC
November 29, 1994, St. Louis, MO
European Compliance Seminar - Medical
November 30, 1994, St. Louis, MO
Information: European Compliance Seminars, ICC D/FW Headquarters. Tel: (817) 491–3696.

Active Circuit Design for Wireless Systems
November 28-December 2, 1994, Los Angeles, CA.
RF Circuit Fundamentals Part I
November 30-December 2, 1994, Los Altos, CA
RF Circuit Fundamentals Part II
December 5-7, 1994, Los Altos, CA
Wireless System Design
January 16-20, 1995, Los Altos, CA
RF/MW Circuit Design I
March 6-10, 1995, Switzerland
RF/MW Circuit Design II
March 13-17, 1995, Switzerland
Information: Besser Associates, 4600 El Camino Real, Suite 210, Los Altos, CA 94022. Tel: (415) 949–3300. Fax: (415) 949–4400.

Outdoor Insulator Technology
January 9-12, 1995, Tempe, AZ
Information: Arizona State University, Center for Professional Development, Box 877506, Tempe, AZ 85287–7506. Tel: (602) 965–1740. Fax: (602) 965–8653.

EMI/EMC Metrology Challenges for Industry: A Workshop on Measurements, Standards, Calibrations, and Accreditation
January 25-26, 1995, Boulder, CO
Information: Ann Bradford, NIST, 813.07, 325 Broadway, Boulder, CO 80303. Tel: (303) 497–3321. Fax: (303) 497–6665.

Microwave System Engineering
December 5-9, 1994, San Diego, CA
Digital Transmission Systems
December 5-9, 1994, Washington, DC
New HF Communication Technology: Advanced Techniques
December 5-9, 1994, Washington, DC
Satellite Electrical Power Systems: Energy Conversion, Storage, and Electronic Power Processing
December 12-15, 1994, Washington, DC
Modern Digital Modulation Techniques
December 12-16, 1994, 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.

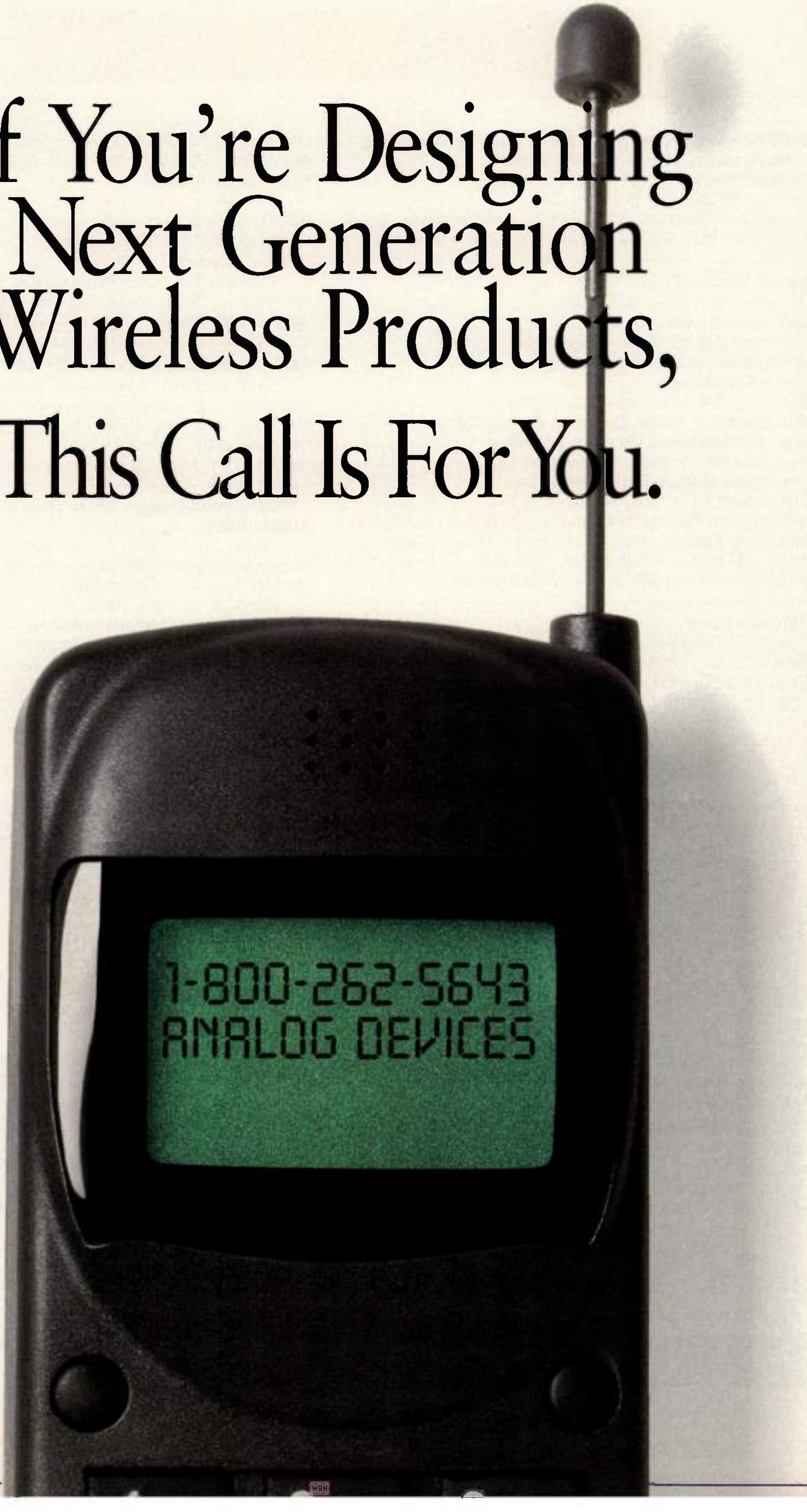
DSP Without Tears
December 7-9, 1994, Atlanta, GA
January 25-27, 1995, Long Beach, CA
February 8-10, 1995, Denver, CO
Information: Z Domain Technologies, Inc., 325 Pine Isle Court, Alpharetta, GA 30202. Tel: (800) 967–5034, (404) 587–4812. Fax: (404) 518–8368.

HIRF Design & Testing Seminar
December 6-9, 1994, Mariposa, CA
Medical Seminar
February 4-7, 1995, Fremont, CA
Information: CKC Laboratories Inc., 5473A Clouds Rest, Mariposa, CA 95338. Tel: (800) 500–4EMC. Fax: (209) 742–6133.

EMC Diagnostics Workshop Emission Measurements
December 6-7, 1994, Surrey, England
Information: Miss Nikki Hamann, Conference Group, Technical Services Division, ERA Technology Ltd., Cleeve Road, Leatherhead, Surrey, KT22 7SA England. Tel: 44 (0)372–374151 ext. 2595. Fax: 44 (0)372–377927.

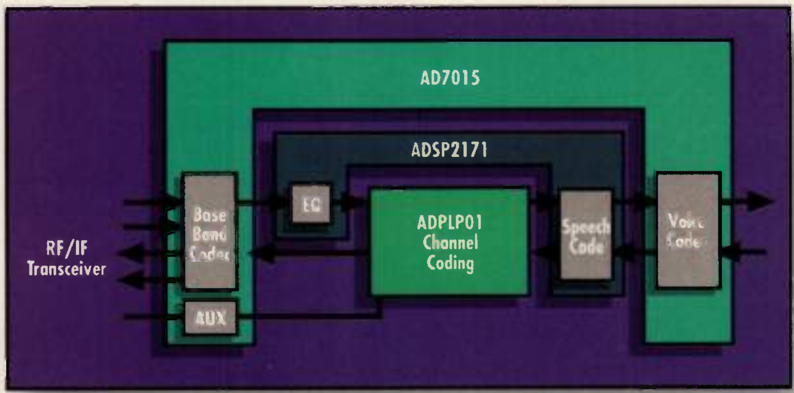
RF/MW Circuit Design: Linear/Non-Linear, Theory and Applications
March 13-17, 1995, Switzerland
Active and Passive RF Components: Measurements, Models, and Data Extraction
March 9-14, 1995, Switzerland
Wireless Digital Communications
March 20-24, 1995, Switzerland
Information: CEI-Europe/Elsevier, Mrs. Tina Persson. Tel: (46) 122–175–70. Fax: (46) 122–143–47.

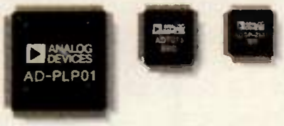
If You're Designing Next Generation Wireless Products, This Call Is For You.



While you're hard at work on your next generation wireless products, so are we. We're Analog Devices, a Fortune 500 company delivering millions of signal processing ICs to leading communications OEMs around the world. Here's what we've been doing for them and why you should be talking to us, too.

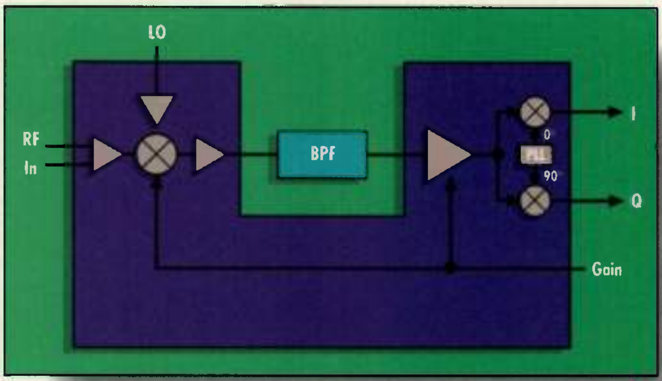
GSM Handsets. Our digital cellular solutions deliver cost, power, and time to market advantages second to none. For example, we've integrated the entire baseband portion of a GSM handset, including hardware and software, into a 3-chip set. Our analog, digital and mixed signal components actually *think* like a handset, so they know when to turn off unused circuits, conserving battery life.





Our AD20msp410 integrates DSP, coding, converters, microprocessor and protocol software into a 3-chip GSM solution.

Other Communications Solutions. Leading TDMA, DECT and PMR manufacturers rely on our mixed signal devices. Basestation manufacturers are reducing part count and cost with our wideband 12-bit ADCs, the first monolithic converters capable of digitizing a complete cellular band.

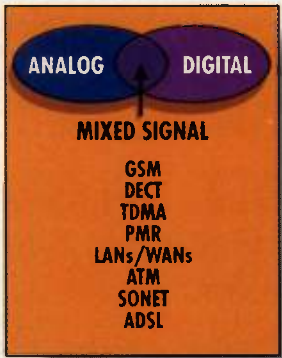


The AD607 is a complete linear IF strip that provides 500 MHz bandwidth and draws just 25 mW from supplies down to 2.7V.

And for the heart of their all-digital designs, they're specifying the ADSP-21060, the highest performance DSP in the industry.

When Inmarsat needed a 3V DSP for their global pagers, they called us. When ATM LAN suppliers need clock recovery/data retiming devices for data rates up to 155 Mbps, they talk to us. For advanced communications ICs, such as 3 volt RF/IF components or integrated ADSL chip sets, OEMs are talking and designing with Analog Devices.

If your wireless team is talking about advanced signal processing ICs today, reducing part costs tomorrow, and working with a state-of-the-art silicon partner, they're talking about us. To get our wireless team talking to yours, call 1-800-ANALOGD. (262-5643)†. Or E-mail comms.div@analog.com.



Emerging communications products require Analog Devices' core signal processing technologies.

ANALOG DEVICES

Analog. Digital. Solutions.

Analog Devices, Inc., One Technology Way, P.O. Box 9106, Norwood, MA 02062-9106.
Distribution, offices and application support available worldwide.
† Outside U.S., please call 617-461-3392.

INFO/CARD 19

NTL and Philips Display MPEG-2 TV at International Broadcasting Convention

NTL and Philips Consumer Electronics both displayed their digital broadcast systems at the International Broadcasting Convention. NTL's VCS 4000's 19-inch rack-mounting encoder compresses video signal to the MPEG-2 main profile/main level standard, suitable for transmitting 625 line PAL or 525 line NTSC signals. Star TV of Hong Kong plans to use NTL's VCS 4000 as the basis of a 32-channel digital satellite TV service scheduled to launch in mid-1995.

Philips' end-to-end system includes a prototype set-top decoder, satellite and modulation systems, video and audio encoders, and a packet switching multiplexer. The Philips broadcasting video system received digital signal from a Telecom France research center in Riennes, France, via the Telecom 1C satellite and a dish in the convention parking lot. The Philips system will be used in trials by the end of 1994, and commercial applications are expected in 1995. Both systems comply with an emerging set of standards promulgated by Digital Video Broadcasting, which operates under the European Broadcast Union. DVB standards for satellite and cable digital transmissions are under review.

HDTV Closer to Reality?

High-definition television (HDTV) has successfully undergone three months of field testing, and results indicate that HDTV broadcasts could be more powerful than the current NTSC broadcasts. In field testing, HDTV broadcasts consistently outperformed the existant NTSC. Within ten miles of broadcast origin, 96% of HDTV test sites received the broadcast as opposed to 67% of NTSC sites receiving acceptable picture. In areas more than 50 miles from the transmission site, 20% of HDTV sites received picture, where none of the NTSC sites received acceptable picture.

Even though HDTV outperforms NTSC broadcasts, implementation of a HDTV standard could be delayed by as much as a year. Broadcasters are concerned about inter-channel interference from HDTV transmissions, and addressing those concerns may delay the standard creation process. The HDTV system will undergo more laboratory testing beginning in January 1995.

IEEE IFCS Call for Papers

Authors are invited to submit papers for the 1995 IEEE International Frequency Control Symposium. The Symposium will be held May 31–June 2, 1995 at the Faimount Hotel, San Francisco, California. Papers dealing with recent progress in research, development and applications in piezoelectric devices, oscillators, synthesizers, frequency standards, noise phenomena, frequency and time coordination, sensors, and measurements are invited. One copy of a summary in sufficient detail for evaluation of the proposed paper (at least 500 words) should be sent to Dr. Lute Maleki - MS 298-100; Time & Frequency Systems Research Group; Jet Propulsion Laboratory; 4800 Oak Grove Dr.; Pasadena, CA 91109; tel: (818) 354-3688; fax: (818) 393-6773; E-mail: lmaleki@fridge.jpl. nasa.gov

Each summary MUST include the author's name, address, telephone and fax numbers, and E-mail address, if available.

278 Attend Piezoelectric Conference

278 people attended the 16th Piezoelectric Devices Conference and Exhibition at the Westin Crown Center in Kansas City, Missouri. The event, held between September 27 and 29, comprised three technical tracks, an exhibition, and a meeting of the P-11 engineering group of the EIA. At the Conference, Shih Chuang of Statek Corporation was presented with the David P. Larsen Award, and Charles Adams of the Hewlett-Packard Company received the Piezoelectric Devices Man of the Year Award.

NIST and ANSI Sign Agreement to Build Electronic Standards Network

NIST and ANSI have laid the foundation for an electronic standards network that will facilitate technology commercialization and promote economic growth. The National Standards Systems Network will eventually link the databases of hundreds of US organizations involved in the development, production, distribution and use of technical standards. Examples of these standards include criteria for safety in consumer products, compatibility between computer software packages and quality systems requirements for products and services exported (such as the ISO 9000 series of standards). NSSN will provide cataloging, indexing, searching, and routing capabilities to end users, allowing access to the entire range of regional, national, and international standards. Both NIST and ANSI expect that the creation of the NSSN will support US economic growth and enhance competitiveness by facilitating technology transfer and promoting rapid development and deployment of standards. It is expected to reduce standards development time and costs, minimize duplication among government and private sector standards, increase the dissemination of standards information to small businesses (for example, through the centers of NIST's Manufacturing Extension Partnership) improve user access to an involvement in national and international standards activities, and provide accurate and current standards information for training and education.

IEE Call for Papers

The IEE has announced a call for papers for the 1995 International Conference on 100 Years of Radio to be held September 5-7, 1995 in London. The conference will cover all radio systems and techniques relating to marine radio (from spark to satcom), pre-broadcasting developments, broadcasting (excluding television), HF communications, data radio communications, military communications, satellite communications, mobile and cellular radio, antennas and propagation, amateur radio, LF and VLF communications, receiver and transmitter development, developments in components (including miniaturization), and the social origins and impact of radio. Those wishing to contribute should submit a synopsis on up to 1 side of A4 paper, to be received on or before January 23, 1995. For further information contact HYR95 Secretariat; Conference Services; Institution of Electrical Engineers; Savoy Place; London WC2R 0BL UK; tel: 071 344 5477; fax: 071 497 3633; E-mail: conference@iee.org.uk.

Contracts

Motorola Wins Vietnam Cellular Phone Contract – Motorola's Cellular Subscriber Group announced that it has been awarded a contract by Vietnam Mobile Services to supply GSM portable cellular phones to Vietnam. Initial shipments began in October.

Superconductivity SSD System to Improve Power Quality – Superconductivity, Inc. has teamed up with the Carolina Power and Light Company, Raleigh, SC to use superconducting technology to help industry improve power quality. Superconductivity's SSD magnetic storage system has been installed at ITW Angleboard, a packing materials manufacturing plant in Hartsville, SC. The SSD protects sensitive automated and electronically controlled equipment by sensing momentary incoming power disruption and instantly providing supplementary power.

Cubic Gets Automatic Toll Collection Contract – Coviares, S.A., a consortium of Argentina's largest construction companies, has awarded Cubic Automatic Revenue Collection Group a $6.7 million contract to design, manufacture, and install an automatic toll collection system on the La Plata-Buenos Aires Tollway that serves metropolitan Buenos Aires. Cubic will also provide 50 emergency call boxes along the route for motorists' convenience in the event of a breakdown.

Scientific-Atlanta to Provide Europe's Largest VSAT Network-Scientific-Atlanta, Inc. announced it is providing Europe's largest interactive data VSAT (very small aperture terminal) satellite network to Safeway Group, one of Europe's largest retailing companies. Installation of the network has begun and will be completed next year.

Andrew Supplies Antennas to Kokusai Denshin Denwa Co. – Andrew Corporation is supplying 4.6 meter antennas and electronic equipment to KDD, the Japanese international telecommunications carrier. These antennas will be part of KDD's F-1 IBS/VSAT Customer Premised Earth Station (CPES) for a point-to-point INTELSAT open network circuit. The service will be dedicated to Japanese companies in foreign companies as an alternative to local telephone services.

MCI Wins NOAA Satellite Contract – MCI has been chosen to provide a satellite-based aviation weather network for 75 countries in the Atlantic and Pacific Ocean regions, according to the National Oceanic and Aeronautic Administration (NOAA). The network will be provided under a seven-year multi-million dollar contract, and is expected to be operational early next year.

EJ to Test for Satellite Program – EJ Systems received a million dollar contract from Raytheon Company to supply QUBEPAK Environmental Test Systems to perform precision thermal and electrical testing of RF Multi Chip Modules for Motorola's IRIDIUM Satellite Program. The systems will be installed at Raytheon's Microelectronics Center in Andover, MA beginning in October.

INFO/CARD 20

Business Briefs

Motorola Demonstrates Two-Way Paging – Motorola has conducted the industry's first true demonstration of two-way paging technology. The demo is based on the company's two-way paging protocol ReFLEX. The ReFLEX protocol was developed to facilitate products and services for the FCC's Narrowband Personal Communications Services spectrum. Spectrum licenses, which were auctioned in July, were designed to encourage the delivery of advanced messaging services, including one- and two-way messaging, acknowledgment paging, and advanced voice paging.

Spartan Satellite Uses Vector's PCM System – Aydin Vector announces that the Spartan Satellite to be launched on the Space Shuttle Discovery will have its micro miniature Pulse Code Modulation (PCM) system on board as part of the Command and Data handling subsystem. Aydin Vector is known for the design, manufacture, and system integration of flight test certification instrumentation for both commercial and military applications.

Lindgren Acquires Ray Proof Ltd. UK – Lindgren RF Enclosures, Inc. announced that they have concluded an agreement under which Lindgren acquired the Ray Proof Ltd. UK shielding business. The acquisition increases Lindgren's electromagnetic shielding product and service offerings, and provides Lindgren with European manufacturing capability.

Fil-Mag Bought by Technitrol – Fil-Mag was acquired by Technitrol Inc., a Philadelphia-based manufacturer of electrical contacts, electronic components, metal laminates, scales, measuring devices, and document handlers. Fil-Mag is now a member of Technitrol's component group companies.

Intellitag Exceeds 99.9% Accuracy Requirements for Turnpike Authority – Intellitag Products announced that it has successfully completed the testing phase of its IT2000™ wireless system for the Kansas Turnpike Authority, surpassing the 99.9% accuracy required by the KTA. The IT2000 system was examined in lab testing and on highway test lanes for electronic toll collection of Kansas tollroads as the new K-TAG system. The testing phase began in June and recently concluded.

New Address for Tele Quarz – The Tele Quarz USA Branch Office has moved to 3545-H Centre Circle Drive, Ft. Mill, SC 29715; phone (803) 547-0770; fax: (803) 547-0775. A list of all US representatives is available there.

Wakefield Engineering Acquires AhamTor – Wakefield Engineering, Wakefield, MA, has announced the purchase of AhamTor, Temecula, CA. AhamTor has approximately 100 employees and will continue to operate under its current name out of its 50,000 square foot facilities in Temecula.

Anadigics Recognized by Blue Chip Enterprise Initiative – Anadigics has been named as a state designee in the 1994 Blue Chip Enterprise Initiative (BCEI). The BCEI, sponsored by Connecticut Mutual, the US Chamber of Commerce and Nation's Business, honors small businesses that have overcome challenges and emerged stronger.

KVG Under New Ownership – Dr. Stefan Reineck, Managing Director of Kristall-Verabeitung Neckarbischofsheim GmbH has assumed ownership of KVG. Under a management buyout, Dr. Reineck acquired all the company's shares from the heirs of the company's founder, Kurt Klingsporn. KVG develops and manufactures quartz crystals, oscillators, and filters for telecommunication and navigation applications.

New Address for Sprague-Goodman – Sprague-Goodman has relocated as of October 1994. The new 15,000 square foot facility is located at 1700 Shames Drive, Westbury, NY 11590; phone: (516) 334-8700; fax: (516) 334-8771.

Contracts *continued*

RIT Receives Grant to Study EM Emissions–The Rochester Institute of Technology received a $20,000 NSF Instrumentation and Laboratory Improvement grant to teach ways to measure and suppress electromagnetic interference (EMI). The money will be used to purchase an EMI measurement set for use by students and faculty in RIT's electrical engineering technology program.

Scientific-Atlanta to Provide VSAT Network – Scientific-Atlanta is providing one of Europe's largest interactive data VSAT satellite networks to Groupe Azur, France's largest insurance company. The Groupe Azur network represents one of the first installations of Scientific-Atlanta's SkyRelay, the VSAT industry's first system designed specifically for companies moving to distributed computing environments that require high-capacity communications support.

START Licenses Drexel's Frequency Converter Technology – START Technology Partnership announced the successful completion of a license agreement awarding Phoenix Microwave Corp. the rights to develop new wireless communications components and systems based on an innovative frequency converter circuit design from Drexel University. This invention helps reduce chip size power consumption and costs, while improving other performance parameters for wireless electronic devices.

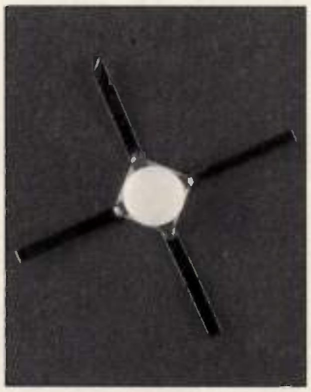
Stanford Telecom and TIW Team up for Wireless VSAT – The ASIC Custom Products Division of Stanford Telecommunications, Inc. has announced its agreement with TIW Systems to develop and produce a wireless VSAT communications subsystem. The subsystem, for use in rural wireless telephone systems in developing nations, will include a Stanford Telecom modulator/demodulator along with TIW's voice compression and telephone interface circuitry.

Wireless "Cellular Digital Packet Data" Network Available – Bell Atlantic Mobile customers in Philadelphia and northern New Jersey can now "test drive" a new mobile digital data service that can revolutionize the way they work and live. The first "Cellular Digital Packet Data" systems in those markets have been turned on and made available for customer use. Commercial service is planned for later this year.

New Wireless Power Transistor

This common emitter class AB bipolar transistor is designed to operate from 850 - 960 MHz at 150 watts PEP. Power gain is rated at 10 dB min. @ 26 volts Vcc with IMD3 of -32 dBc typ. This transistor is designed for wireless base station amplifiers.
P/N PH0810-150

Richardson Electronics 800-348-5580
CIRCLE READER SERVICE NO.124



New High f_T, Low Voltage, Medium Power, Silicon Transistors

Designed for battery operated systems with 3-5 volt collector bias. Ideal for use as low phase noise oscillators through 6 GHz and as low noise moderate power drivers or class C amplifiers through 3 GHz. Available in chip form or ceramic package.
P/N MA4T3243

Richardson Electronics 800-348-5580
CIRCLE READER SERVICE NO.21

New GaAs MMIC Digital Attenuator

3 bit, 4 dB step in a low cost, SOIC 14 lead, surface mount, plastic package. This digital attenuator covers DC - 2 GHz and has superior attenuation accuracy (0.30 dB ± 3% of attenuation setting). $IP_3 \geq 50$ dBm at 0.5-2 GHz. Low loss measures 1.5 dB typ.
P/N AT-230

Richardson Electronics 800-348-5580
CIRCLE READER SERVICE NO.125

New Low Noise Amplifier Provides High Third Order Intercept

The AM-216 amplifier (821-851 MHz) has a third order intercept of +36 dBm typ., which makes it ideal for cellular base station receiver applications. The AM-217 covers the European GSM (880-930 MHz) band. Noise figures are 1.1 dB and 1.2 dB typ. respectively. P/N AM-216 / AM-217

Richardson Electronics 800-348-5580
CIRCLE READER SERVICE NO.126

INFO/CARD 127

Please see us at RF Expo East '94, Booth# 218

RF at Home on Cable

by Andy Kellett
Technical Editor

Despite a lot of talk about a "wireless world", cabled connections are still an important part of many communications systems. Transmission by cable has several properties which makes it indispensible as a transmission medium. Because cables (ideally) don't radiate, signals can be sent without regard to spectrum use and without broadcasting information to everyone within "radio-shot" of the transmitter. This report looks at what services use cable transmission (including fiber optic cable) and how these cabled links are accomplished.

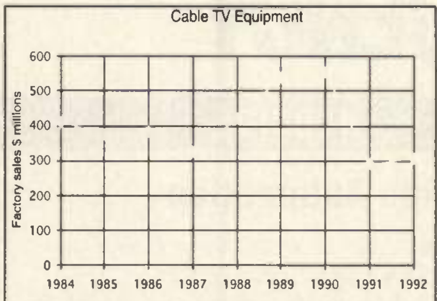
Cable Television

One of the first applications to spring to mind when cable transmission is mentioned is cable television. According to Gary Kim, Senior Vice President for Probe Research, a market research firm concentrating on the telecommunications market, about 58 million homes subscribe to cable television services, which is about 62 percent of the homes that cable lines pass. Cable television lines already pass 93 to 98 percent of the homes in the U.S., and much of the conversion of main trunk lines to fiber optic cable is complete, so the surge in cable infrastructure building is tailing off.

However, the number of proposed cable services is booming. Virtually every kind of service imaginable is being considered says Kim, from interactive video games on demand to POTS (plain old telephone service), and these new services will require wideband cable connections to the home.

New trunk lines have bandwidths around 1 GHz, while most cable drops, (the line from the pole to the subscriber's house), are designed for the older 350 MHz bandwidth technology, says John Valentine, Director of Marketing for Belden Cable. Before new services requiring wider bandwidths can be brought into subscribers homes, nearly all the drops installed before the early nineties must be replaced, says Valentine. While the new drops will be copper coax, they will use cable designed to reduce SRL (structural return loss).

While cable television is perhaps the best example of RF transmission by cable, there are a number of other applications that transmit RF via cable.



From: *1994 Electronic Market Data Book*, EIA Marketing Services Dept., Electronic Industries Association.

Table 1. Cable television equipment sales, (1984-1992).

Extensions, Links and Networks

RF signals are also sent for appreciable distances over cable in cellular base station extensions and studio links. Base station extensions, like those manufactured by the Ortel Corporation, can effectively extend the coverage of a cell site by linking a smaller transmitter/receiver pair to the base station. Ortel's devices use a fiber optic link to reach from the remote site to the base station. An entire cellular band can be linked to the base station this way. Similar links can be used to connect television and radio studios to transmitters.

Multiplying the number of computer networks in existence by the length of the cable used in each of those networks would probably reveal that digital data communication uses a *lot* of cable, and as data rates go higher, they appear more and more "RF-like." For example, many of the packet network protocols are nearly identical to those used in radio packet networks.

Medical imaging systems provide another example. According to Dr. Arturo Gamboa, Director of Engineering for medical image processors Cemax, a state-of-the-art x-ray image is 4000 pixels by 5000 pixels with 2^{12} gradations of grey. These images use JPEG compression techniques (like those used in satellite communications), and transmit at data rates of up to 150 Mbps (megabits per second).

Data communications engineers have managed to increase data rates while still transmitting that data over twisted pair. However, faster data rates and increased EMI requirements may drive more data communications into cable – both fiber optic and copper. Copper cable does show a disadvantage in this application, however, because of its size, says Peter Hanen Manager of Marketing for Amphenol, "there is only so much room in plenums and under floors."

Fiber Optics

Fiber optic cable has several advantages over its copper relatives. It has lower loss over distance, typically about 0.4 dB/km, and it is smaller, more flexible, and more easily laid. Of course light is the transmission medium for fiber optic transmission, but that light is normally modulated at RF frequencies.

Ortel engineers characterize laser diodes in terms of RF performance – noise, linearity, flatness, etc. – in order to design the RF circuits that drive those diodes, says Dr. Hal Zarem, Business Manager for Wireless Communications at Ortel. And the end products are specified in RF terms, says Zarem, "so you don't have to know about the fiber optics."

Clearly, RF signals will be launched into more than antennas. Copper coax and fiber optic cables both play a role which complements the expanding use of "wireless" technology. ***RF***

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SHPi	8	9	9	0.1	YES	3K	OXIDE	2	YES	YES
C-Pi	9.5	9	10.5	5.5	YES	3K	OXIDE	2	YES	YES
GST-1	5.5	13	—	0.1	NO	20K	TRENCH	3	YES	YES
GST-2	4.5	27	—	0.1	NO	60K	TRENCH	3	YES	YES

- **C-Pi** is a recessed-oxide-isolated high-speed complementary bipolar process optimized for analog signal acquisition, amplification, and sourcing. Without the vertical PNP option, C-Pi is designated as **SHPi**.
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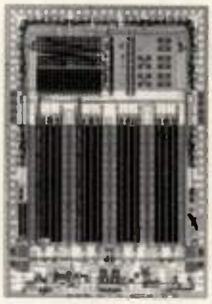

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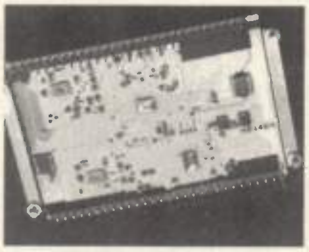

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Case History #5: *The Tektronix TDS 544A Four-Channel 500MHz Color Digitizing Oscilloscope received EDN's Test and Measurement Product of the Year Award. This product contains 13 custom ICs and four custom hybrids manufactured using Maxim's SHPi and GST-1 processes. Maxim-fabricated circuits include four 250 MS/sec A/D converters for signal acquisition, circuits that provide normal and specialized trigger functions, a 500MHz buffer with precision delay adjustment, wide band input amplifiers with variable gain and position control, and hybrid precision attenuators with Maxim die. Maxim-fabricated circuits make possible this instrument and many others in the Tektronix Test and Measurement family of products.*

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An Analog Phase/Frequency Comparator and BPSK Carrier Demodulator

By Richard W.D. Booth
Itron, Inc.

Here is the Grand Prize winning entry in the 1994 RF Design Awards Contest! Of the many outstanding circuit design and software entries submitted for judging, this was selected as the winner. The author has been awarded a software package of HP-EEsof"s Touchstone and Libra for Windows, generously provided by HP-EEsof in recognition and support of RF engineering creativity.

This note describes a phase locked BPSK carrier recovery circuit which contains an intrinsic AFC function that allows the loop to acquire frequency offsets well outside the loop bandwidth. This circuit was originally developed to provide a BPSK carrier recovery circuit that uses only RF type multipliers. The circuit described here uses inexpensive NE602A multipliers.

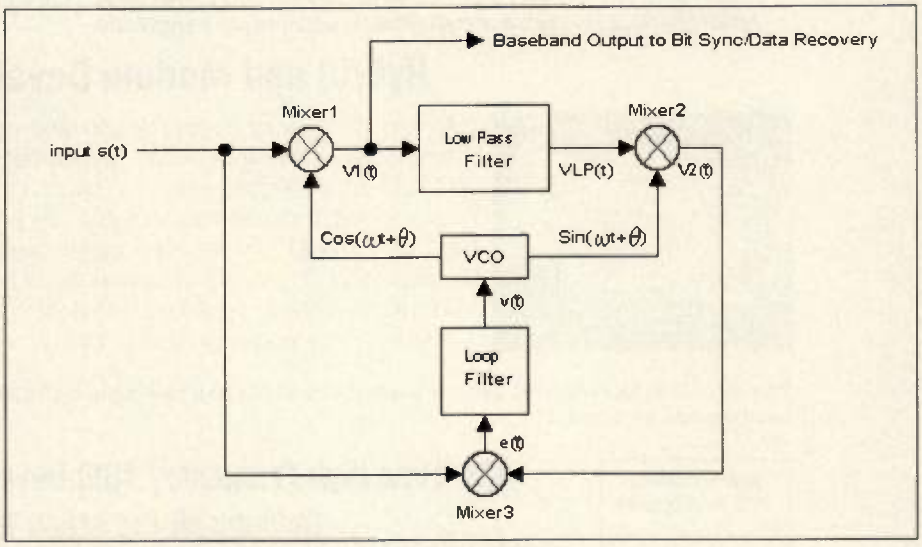


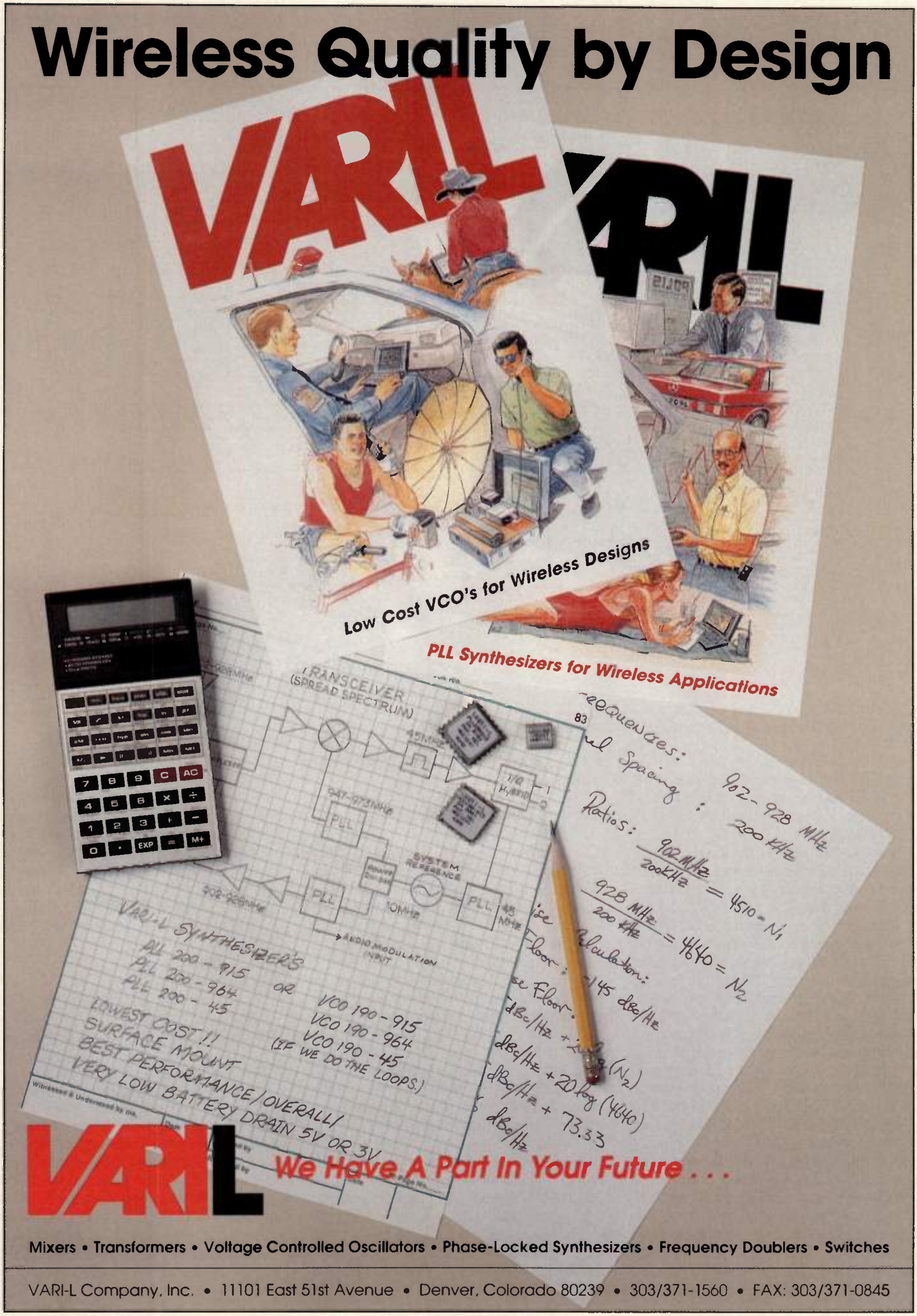
Figure 1. BPSK demodulator block diagram.

The recovery of the carrier for the demodulation of BPSK is usually done with either a Costas loop or a doubling loop. The Costas loop is popular because the base band data is available at the output of one of the mixers. However, the Costas loop requires a baseband multiplier, and the loop must be aided during the frequency acquisition phase by adding circuitry to either implement a sweep of the VCO or an AFC for systems where the frequency accuracy of the carrier is poor. The loop described in this note was designed with the intention of removing the baseband multiplier from the Costas loop, and the result is a design that is implemented with three RF-type multipliers or mixers. During the development of the mathematical model for the loop, it was discovered that the control voltage at the VCO input contained an AFC term that forces the loop to lock for frequency offsets outside the loop bandwidth. This feature has obvious utility in the BPSK carrier recovery loop, since the additional sweep or AFC need not be added for systems requiring frequency acquisition. The circuit may have application in synthesizers, where it may be used as an analog phase and frequency comparator.

Circuit Description

A block diagram of the loop is illustrated in Figure 1. The incoming signal is split and coupled to two of the three mixers. Mixer 1 is driven at the LO port with the in phase (I) version of the VCO. The output of this mixer is the base band recovered data signal. This baseband data signal is sent to a bit synchronizer and/or integrate and dump in order to recover the data. The baseband signal is also filtered by the low pass filter at the output of mixer 1, and the output of the filter is then upconverted in mixer 2. The LO input to mixer 2 is the quadrature (Q) version of the VCO output. The output of mixer 2 is then connected to the RF port of mixer 3. The LO port of mixer 3 is driven with the original input signal. The baseband output of mixer 3 contains the phase and frequency error information. This is filtered using a standard loop filter, and the output of the loop filter drives the VCO input.

The circuit works as follows.: assume a small phase offset between the VCO and the incoming carrier phase. Assume also that data is present on the carrier. Now the output of the first mixer, mixer 1, should just be the baseband data. Since there is an assumed small offset in phase between the incoming carrier and the VCO, the amplitude of this data voltage is reduced somewhat relative to the voltage present, assuming perfect phase lock. This data output is now upconverted in mixer 2, which is used as a bi-phase modulator. The output of mixer 2 or the bi-phase modulator is close to being a copy of the original input signal except the carrier is in quadrature with the original input signal. When this signal is multiplied with the original input signal in mixer 3, several things happen. The modulation is stripped off. This occurs because the BPSK wave form is NRZ and multiplying the data by itself results in +1. The second result is the creation of a term that is proportional to sin(2θ). This is the desired error term and results from multiplying the original signal times the output of mixer 2, which contains the data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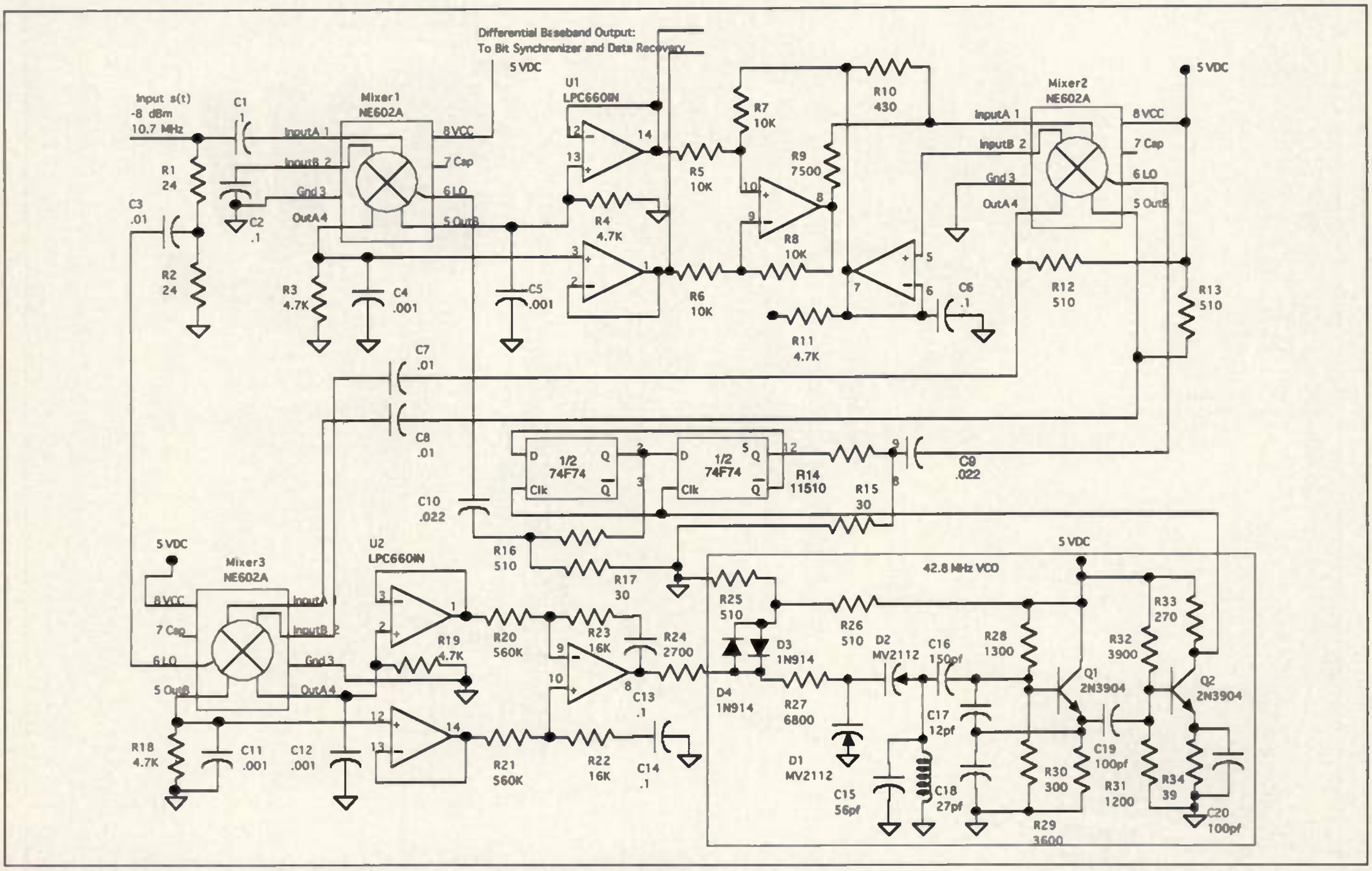


Figure 2. BPSK demodulator/phase frequency comparator schematic.

modulated onto a carrier that is in quadrature with the original input signal. The error signal is filtered in a standard loop filter and is used to drive the control port on the VCO.

For frequency errors, the procedure is a little more complicated. Assuming both a phase and frequency error between the incoming signal and the VCO, then the output of mixer 1 contains the difference signal between the incoming signal and the VCO. When this signal is passed through the low pass filter at the output of mixer 1, the sinusoidal component undergoes a phase shift that is proportional to the frequency. Assuming that the low pass filter is wide enough that sufficient distortion of the data signal does not occur, then this phase-shifted signal is then modulated back up into the up converting mixer, mixer 2. The output of mixer 2 drives mixer 3 and the modulation is removed again at this point. The base band signal, however, contains a term that is proportional to the phase-offset induced by the phase shift of the sinusoid through the low pass filter. This is similar to a discriminator. This term drives the VCO towards frequency lock. A mathematical description of the loop is contained in Appendix A.

Hardware Implementation

A hardware implementation of this circuit is described in Figure 2. This schematic illustrates the use of three NE602A mixers in the loop. The VCO uses two transistors and a dual D flip-flop, a 74F74, in order to generate the quadrature VCO signals. The baseband filters and amplifiers are built around a low-ccst CMOS quad operational amplifier. The filtering in the low pass filter at the output of mixer 1 determines the loop pull-in range. At frequencies much beyond this low pass filter cut off, the loop fails to acquire since the signal is rolling off in amplitude and eventually disappears into the noise. Several sections of the quad op amp were needed in order to perform the DC level shift between the output of the first NE602A, mixer 1, and the input to the second NE602A. This path must be DC coupled.

The low pass filter function that provides the phase shift of the base band sinusoid is provided by the combination of the 1.5 K resistors intrinsic to the output stages of the NE602A at mixer 1 and the external components R3, R4, C4 and C5. Additional filtering could be added at the summing stage that drives the baseband input to mixer 2. This could be provided by adding capacitors to R8 and R7. The values chosen give a low pass cut off of about 140 KHz, which restricts the AFC range to about ± 200 KHz.

The parameters for this loop are as follows:

Center Frequency:	10.7 MHz
Loop Bandwidth:	500 Hz
Loop Pull-in Range:	± 100 KHz
Max. Data Rate:	20 KBPS (apx)

Final Notes

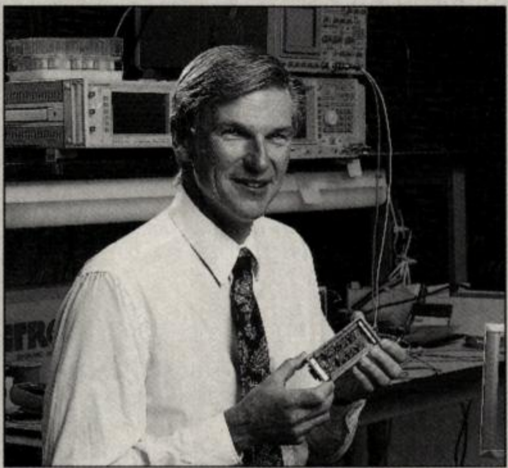
The phase-locking behavior and the frequency-locking behavior are clearly very closely interrelated, just as they are with a phase and frequency comparator. This means that they cannot be individually specified. Usually the phase locking behavior is specified and then the frequency acquisition behavior is an artifact of the phase locking performance. Smaller loop bandwidths do indeed result in longer frequency acquisition

times for a given offset.

The loop exhibits false locking when attempting to acquire data bearing signals and the frequency offsets exceed the data rate. This is a peculiarity of all BPSK type loops and is a result of the distortion of the wave form by the base-band filters used in this loop and in the Costas Loop. Other style loops exhibit similar tendencies. This quirk is usually handled by using acquisition sequences that are preceded by sending a pure carrier for a sufficient length of time to allow the carrier loop to lock. *RF*

About the Winner

Richard Booth is a System Architect at Itron, Inc., a manufacturer of wireless communications networks, mainly for public utility companies' monitoring and metering needs. Rick has worked at Itron for 4 years, with previous experience at TRW. He received a BSEE from MIT in 1969 and a Ph.D. from USC in 1974.



The winning design was derived from earlier work in modulators and demodulators for digital communications. Additional experience in frequency synthesis is clearly evident in the design.

Rick enjoys outdoor activities, and is an active backpacker and rock climber. It was intended that a formal presentation of the Grand Prize would be made at RF Expo East later this month in Orlando, but Rick will be in Nepal, combining a mountain trek and honeymoon.

He can be reached at Itron, Inc., 20520 Prospect Road, Saratoga, CA.

A Contest Summary

Judging this year's contest were the 1993 winners, Terry Hock and Ray Page, along with *RF Design* Consulting Editor Andy Przedpelski and Editor Gary Breed. It was a clear consensus that the entries were the most consistent high quality of *any* past contest. *RF Design* will publish most of the contest entries over the next year. Readers can look for one or two of them in December, and in each issue during 1995.

Once again, the RF Design Awards contest has proven that RF engineering is a creative profession. Congratulations to everyone who entered!

Specal Thanks to HP-EEsof, Our Prize Donor!

Hewlett-Packard Company has developed a strong affinity for the RF Design Awards contest, a result of their interest in encouraging creativity and innovation among RF engineers. This year, their HP-EEsof division has provided a package of their Windows-based design and analysis software, including the well-known Libra and Touchstone programs. HP has repeatedly expressed their interest in supporting the contest with major prize donations. Our thanks to them for the generous donation of the 1994 Grand Prize!

Appendix A-Mathematical Model

Input Signal:
$s(t) = d(t)*\cos(\omega t+\theta)$ where $d(t)$ = NRZ data sequence

In-phase VCO signal:
$R_I(t) = \cos(\omega_0 t+\theta_0))$

Quadrature VCO signal:
$R_Q(t) = \sin((\omega_0 t+\theta_0))$

Mixer 1 Output:
$V1(t) = R_I(t)*s(t) = d(t)*\cos(\Delta\omega t+\Delta\theta)$
where $\Delta\omega = \omega_0-\omega$ and $\Delta\theta = \theta_0-\theta$ and it is assumed the double frequency terms have been removed by the low pass filter.

Low Pass Filter Output:
$VLP(t) = d(t)\ |H(\Delta\omega))|\ *\cos(\Delta\omega t + \Delta\theta + \phi(\Delta\omega))$
where $\phi(\Delta\omega)$ = phase shift introduced by low pass filter and $|H(\Delta\omega)|$ = amplitude response of low pass filter. It has been assumed that the low pass filter does not significantly distort the data wave form d(t).

Mixer 2 Output:
$V2(t) = VLP(t)*R_Q(t)$

Mixer 3 Output:
$e(t) = d(t)*\cos(\omega t+\theta)*VLP(t)*R_Q(t)$

or:
$e(t) =$
$d(t)*\cos(\omega t+\theta)*d(t)|H(\Delta\omega)|*\cos(\Delta\omega t+\Delta\theta+\phi(\Delta\omega))*\sin(\omega_0 t+\theta_0)$

or:
$e(t) = |H(\Delta\omega))|*[\sin(2\Delta\omega t + 2\Delta\theta + \phi(\Delta\omega)) + \sin(-\phi(\Delta\omega))]$

For zero frequency offset e(t) simplifies to $e(t) = \sin(2\Delta\theta)$ This is the same form as the phase error for a Costas loop.

For non-zero frequency offsets that are larger than the loop bandwidth, the error term e(t) is approximately $e(t) = |H(\Delta\omega)|\sin(-\phi(\Delta\omega))$. This is because the first term contains the sinusoidal varying component and this tends to average to zero, leaving the AFC term by itself.

Note that the negative sign is correct since the low pass filter will shift the phase in the negative direction as a function of frequency. This means the phase correction and frequency correction both work in the same direction, which is indeed fortunate. For example the phase shift as a function of frequency for a simple one pole low pass is as represented by $\phi(\Delta\omega)=-\tan^{-1}(\Delta\omega+\omega_c)$ where ω_c is the cut off frequency for the low pass filter.

Note that more complicated low pass filters will provide a larger correction voltage as a function of frequency offset. However, more than two poles can give rise to phase shifts of more than 180 degrees for sufficiently large frequency offsets. This can lead to the disconcerting problem of frequency pushing where the loop is driven away from the signal!

Broadband Impedance Matching – Fast and Simple

by Thomas R. Cuthbert, Jr., Ph.D.
Consultant

The method presented here is an efficient and practical way to automate the selection of both a matching network topology and its component values that are likely to provide a nearly optimal match between a complex load and source. This new technique avoids the labor of graphical design, the complexity of real frequency methods, and the many inferior solutions found by optimizers. The computer program GRABIM, that implements this new GRid Approach to Broadband Impedance Matching for LC networks, is described. The new method also requires a bounded, constrained network optimizer in the last design step; an already published program fulfills this role.

Broadband impedance matching networks are often designed by adjusting element values in a chosen network topology using an optimizer program that finds only the nearest solution. The best solution is often missed because the wrong network is used and/or the starting set of element values is not near enough to the optimal solution. However, selection of approximate element values is difficult, especially in the practical case where the load and/or source impedances are available only in tabular data sets, instead of subnetwork models. More sophisticated methods require extensive mathematical, theoretical, and computational knowledge, and also embody optimization, as reviewed in reference [1].

Today's fast PCs and ultra-efficient analysis of lossless networks allow exhaustive trials for all usable combinations of network branch values (three to eleven values for each branch) at the rate of about 75,000 trials per minute on a 486DX2-66 PC. This article describes a new *non-adaptive* grid search method, outlines the key features required for the companion optimizer, and gives an example using the new search procedure in the DOS program GRABIM.

Adding more network elements not only produces little improvement, it also exponentially-increases the time required to exhaustively search for the optimum solution. Designers can now find a suitable network topology by automatically trying combinations of bandpass, lowpass, highpass, pi, or any other LC ladder network topology having two to eight elements. The associated grid searches also find approximate element values for a nearly-optimum solution. Finally, a minimax optimizer that maintains positive element values will drive out any remaining unnecessary elements, coming close to the best network with the best element values.

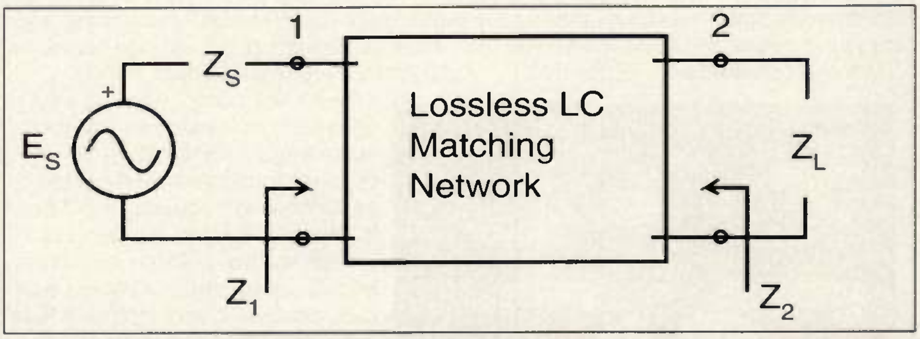


Figure 1. A network for broadband matching sets of terminating impedances.

The two-port matching problem as described in reference [1] is briefly restated here. Figure 1 shows a lossless network inserted between a source and load to improve or control the power transfer over a frequency band. When both Z_S and Z_L are resistances, the network in Figure 1 usually is called a filter or an impedance transformer. Complex impedance terminations require the more difficult broadband matching network. For example, radian frequency samples between 0.3 and 1.0 and corresponding goals of S_{21}=0 dB are contained in the computer file shown in Table 1. The $Z_L = R_L+jX_L$ and $Z_S = R_S+jX_S$ impedances are the rectangular complex number sets shown in Tables 2 and 3, respectively, and correspond to the radian frequencies in Table 1. Impedance matching data is usually normalized to 1 radian/second and one ohm as in Tables 1-3, and this is also required by GRABIM. Normalization is also convenient, since at 1 radian/second, henrys are equal to ohms and farads are equal to mhos. When un-normalizing, actual L 's and C 's are *inversely* proportional to radian frequency. Actual L 's are *directly* proportional to terminal impedance level, while actual C 's are *inversely* proportional to terminal impedance level [2].

The Grid Approach

The labor of graphical matching design, the complexity of the real-frequency methods, and the skill level required for both approaches demand a fast and simple way to design matching networks, especially for the double match case (both Z_L and Z_S complex). The speed of the 486DX2-66 PC used to develop the grid approach has already been exceeded at least 50 per cent (the DX4). So, computing speed, bare-bones network analysis, and efficient bypass of useless trials are the basis of a better matching technique.

The strategy has two related steps: (1) use an exhaustive grid search of element values in multidimensional logarithmic space to find a promising matching network topology and near-optimal element values, and (2) continue with a constrained network optimizer to exceed matching goals and drive out unneeded network elements. The source and load impedance sets may comprise a single fixed resistance or as many as 40 complex impedances and their corresponding frequencies. Both the load and source impedance information are stored in

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Penstock Acquired by Avnet, Inc.

Leading RF/Microwave distributor joins Electronics Marketing Group, retains name, independence, market focus.

By DOUGLAS R. DREW

GREAT NECK, N.Y. — Solidifying its position in the rapidly growing wireless communications market, Avnet, Inc. announced today that it has completed the acquisition of Penstock, Inc.

Penstock, the 39th largest industrial distributor in North America, is the United States' leading technical specialist distributor of microwave and radio frequency products and related value-added services.

Under the terms of the acquisition, the Sunnyvale, California-based Penstock will retain its name and remain a separate company, as part of Avnet's Electronics Marketing Group (EMG), currently the world's largest electronics distributor, with sales of more than $2.3 billion.

Sustained Sales Growth

Penstock's sales reached $45 million in the fiscal year ending this past March, an increase of 32% over the previous year, and were projected to exceed $57 million this year. But the privately-held company "wouldn't be able to reach the next level of growth" on its own, according to Bruce White, Penstock's founder and president.

"Avnet has presented Penstock with a golden opportunity," White said. "We'll be able to maintain our individual presence in the industry while having access to additional financial resources to grow the company in such a way as to benefit our customers, suppliers and employees.

"This alliance will allow us to stay focused in our niche communications market," he added, "using strong technical field sales engineers and providing significant inventory levels of quality products to our customers."

White remains president of Penstock, reporting to Roy Vallee, Avnet's president and chief operating officer.

Opportunities For Penstock

"The RF/Microwave market continues to grow dramatically," Vallee said. "We are excited about the potential opportunities for Penstock. They're as committed to quality as we are, so we're especially pleased to welcome them to the Avnet family."

Prior to the acquisition, Avnet EMG was comprised of five sales and marketing divisions; Allied Electronics, Time Electronics, Avnet International, Avnet Computer Group and Hamilton Hallmark.

> The alliance represents a major opportunity for both distributors.

By significantly expanding the support network of Penstock, the acquisition will enable both companies to offer more extensive services to their respective clients, according to industry analysts.

A Shared Priority

"Both Penstock and Avnet have always made their customers the number one priority and credit much of their success to this," Vallee said. "The proposed structure of this merger strongly reinforces that philosophy and neither company anticipates any 'shakedown' period since the fit is a perfectly logical one on all levels."

Sources familiar with the deal confirmed that no management changes are planned.

Traditionally a military supplier, Penstock entered the commercial arena six years ago and has increased its revenue by at least 30% every year since then. Although the company still generates about $11 million worth of military business annually, approximately 75% of its business now comes from commercial sales.

$2 Billion Commercial Market

In recent years, the military market has remained essentially static, while analysts estimate that the mushrooming commercial market for RF/microwave components has reached $2 billion.

Founded in 1975 and incorporated in 1984, Penstock's specialized product lines include such principal suppliers as Avantek, Comlinear, Hewlett-Packard, M/A-Com, QMI, Sawtek, SGS-Thomson, Siemens, Star Micronics and Toko America.

Financial terms of the acquisition, which was completed in just over four weeks, were not disclosed.

Stock Prices A

Tradi

Canada And United Stat

"CAS 2/90 EX#3. RAD/SEC & GOAL SET"	
8	
.3	0
.4	0
.5	0
.6	0
.7	0
.8	0
.9	0
1.	0

Table 1. Radians/second and S_{21} dB goals.

"CAS 2/90 EX#3. .3-1., 8 PNTS LINEAR ZL"	
8	
0.59016	0.71680
0.71910	0.74944
0.80000	0.77500
0.85207	0.80503
0.88683	0.84174
0.91103	0.88470
0.92837	0.93288
.94118	0.98529

Table 2. R_L and X_L load impedance data.

"CAS 2/90 EX#3. .3-1., 8 PNTS LINEAR ZL"	
8	
0.71313	–0.45230
0.56081	–0.49629
0.41945	–0.49347
0.29915	–0.45789
0.20274	–0.40204
0.12909	–0.33530
0.07539	–0.26403
0.03846	–0.19231

Table 3. R_S and X_S load impedance data.

ASCII files. The load resistance, if used, is one ohm, and the source resistance, if used, can be set to any value. A typical trial run of GRABIM requires about one minute for 75,000 combinations on an IBM PC clone 486DX2-66 computer. This article describes how the grid search works and provides an example with the results from both steps.

Two to eight L 's and C 's can be tried in any simple ladder network topology, and 12 typical topologies are available from a menu. There may be as few as three and as many as 11 trial values for each L and C about a base point, which starts at L=C=1. The grid search may be restarted about a new base point, which is the best 'better-than' or 'equal- to' goal last achieved over all trial combinations. Progress and intermediate results appear on the screen during searches.

Matching Network Topologies

The 12 preprogrammed ladder network topologies are shown in Figure 2. The branches are numbered from 1 to 8 under each network, starting from the load toward the source. GRABIM allows the user to select from two to eight branches from any topology, including those provided by the user. Inductors are labeled type 2 and capacitors are labeled type 3 (this naming convention is a leftover from programming network problems into a handheld calculator, where memory is especially precious). *Minus signs indicate that the L or C is connected like the preceding branch.* For example, the topology at the top of Figure 2(a) (BPP: BandPass Parallel to load) starts with a capacitor in parallel (*the load is considered series*) so that the branch 1 topology is 3. The BPS topology has a series connection to the load, so its branch 1 code is –3. Initial topology choices should use only four to six branches instead of all eight, since excess branches tend to remain in approximately-optimal solutions. Branches can always be added to attempt improvement of a marginally satisfactory topology.

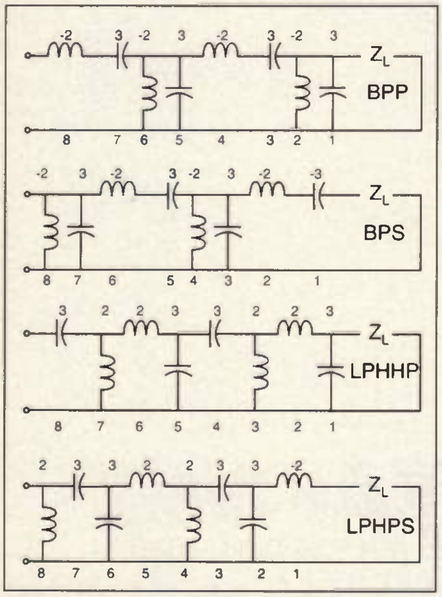

Figure 2. Bandpass, lowpass and highpass topologies programmed in GRABIM. Topology codes appear above and branch numbers below each ladder network.

Exhaustive Grid Search

Walsh [3] has remarked that for a given number of function evaluations, a full search over a courser grid is more efficient than a random search. Consider the matching system in Figure 1 with frequency samples, corresponding S_{21} dB goals, and corresponding Z_S and Z_L data sets as in Tables 1-3. For some candidate LC matching network topology, such as the LPS lowpass network in Figure 2(b), exhaustive combinations of trial values for the L's and C's are tried, each combination being analyzed for the maximum (worst) S_{21} dB found over all frequencies. If that worst S_{21} versus ω is less (better) than the worst previously found, then that combination of LC values and the S_{21} dB are reported and saved until that case can be surpassed.

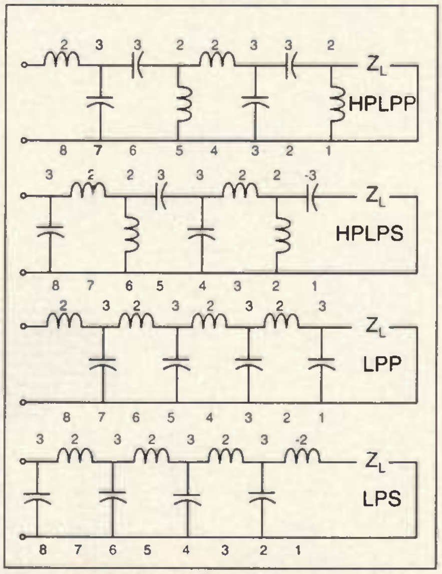

Figure 2b. Topologies continued.

A diagram of the FOR-NEXT loops used in GRABIM for the exhaustive grid search is shown in Figure 3. Use of only the first five branches of the chosen topology is illustrated. As listed below, a set of 3 to 11 trial values is chosen, e.g., 1.0, 0.4, and 2.5. From the start, branches 5, 4, 3, 2, and 1 are set to L or C equals 1.0. Then the S_{21} dB is computed for each sample ω and the branch values and greatest dB are saved. Then branch 1 is set to 0.4, the ω loop is run, and the greatest dB, if less than the prior saved dB, causes that topology to be reported and saved. Otherwise it is ignored.

After branch loop 1 has completed its three trial values, the second trial value

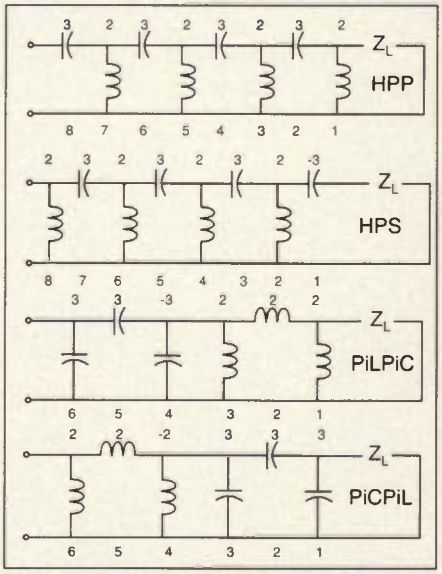

Figure 2c. Topologies continued.

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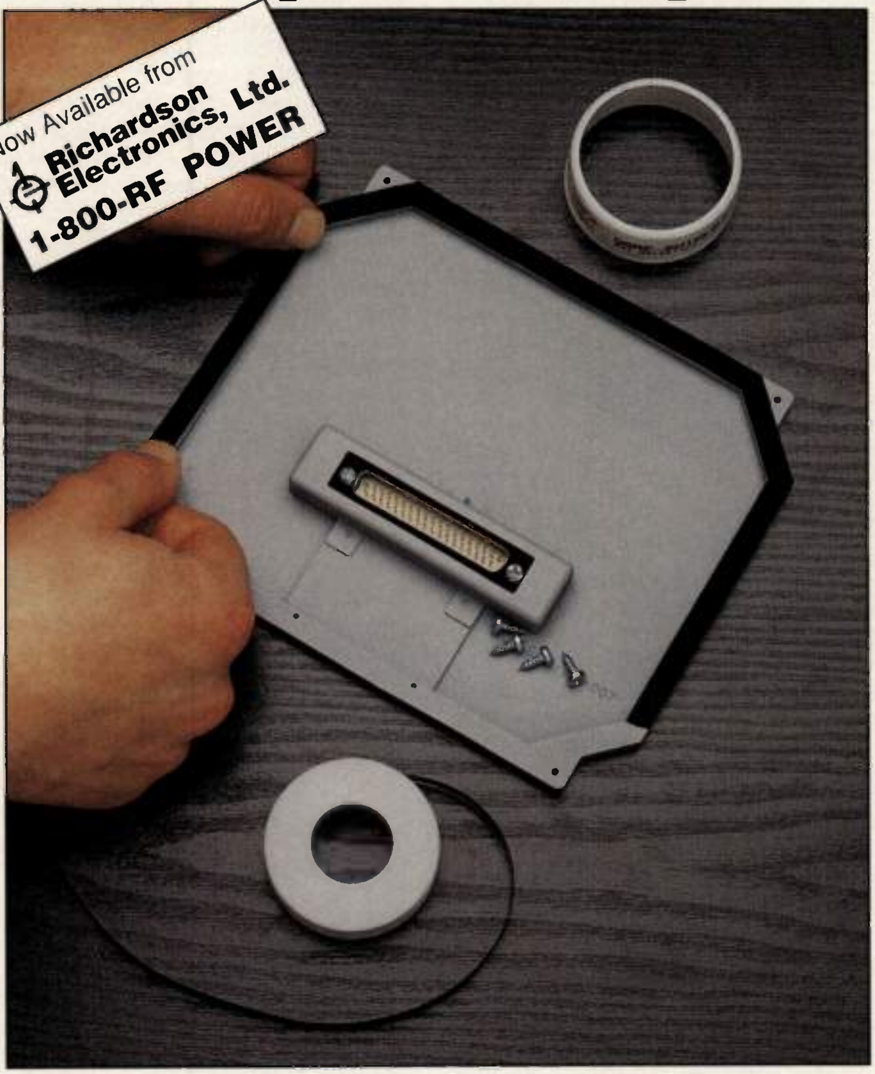

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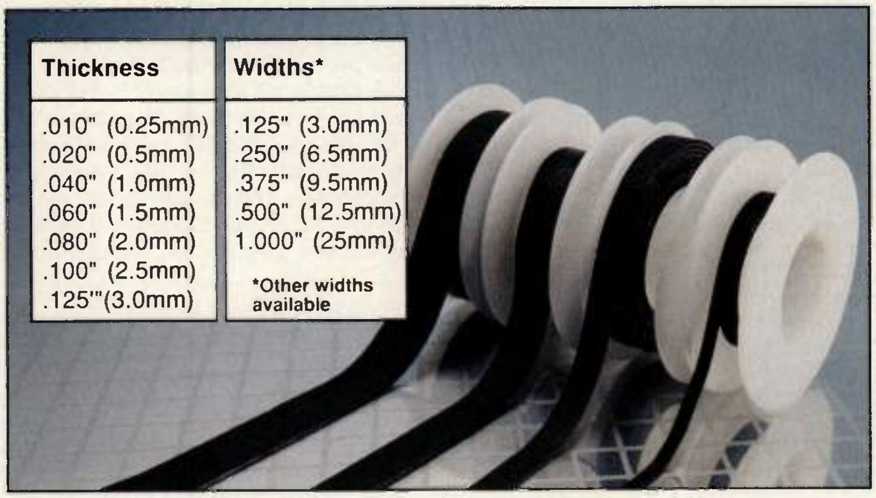

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.060" (1.5mm)	.500" (12.5mm)
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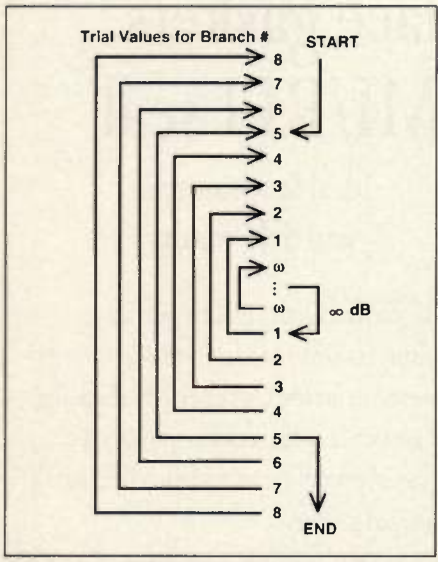

Figure 3. FOR-NEXT loops that try all combinations of values in all branches. 5-branch case is shown.

(0.4) is set in branch loop 2, and branch loop 1 is run again using the three trial values. Eventually, branch loop 5 will have run through all trial values and the exhaustive test is over. During some search modes, the series branch reactances (ohms) or parallel branch susceptances (mhos) may exceed 25. That will be considered a zero of transmission (approaching ∞ dB) in the passband, so the w loop will terminate prematurely as illustrated in Figure 3, since that combination only wastes time.

The sets of trial values are shown in Table 4 and the NT=5 set is plotted in the log-log plane in Figure 4 for two branches. Three branches form a sampled cube, etc. Each trial value set starts with 1.0, which usually produces a reasonable match. That is also the very significant *equal-element filter* [4] which has both nominal element values and nominal performance. Therefore, the very first combination is likely to produce a result that will not be easily improved, so the number of reported cases is minimized. The choice of (nearly) geometrically-related pairs or trial values (e.g., 0.4 & 2.5) is based on the inverted branch reactance values of wL and 1/ωC. The distribution of pairs in a particular trial set is heuristic and reflects the predominance of lower loaded Q values in broadband matching networks. The initial base point sets all branch values to 1.0. However, the Table 4 trial values are actually *factors* that multiply each branch value. Therefore, any subsequent searches starting from a newly discovered base point retain the same pattern in log space, e.g. Figure 4, *no matter where the base point is located.*

The Grow set in Table 4 allows the user to try a fully connected (but not shorted or open-circuited) topology with combinations of *added* LC branch elements having small tuning effects. Growing trials will show that these 'neighborhood' or local tests will not predict global solutions, (this has been confirmed by researchers [5]). Similarly, the Prune set in Table 4 is designed for a full set of unit-valued branches with combinations of *removed* branches to improve the match. Improvements will be noted using both Grow and Prune, but neither case indicates the best global solution. This phenomenon proves the superiority of exhaustive grid search as opposed to any adaptive search method, even adaptive pattern searches such as the popular Nelder-Mead simplex method [3]. Trial sets 3N (Near) and 3F (Far) are provided for search over a large number of branches where computer run time would be excessive for 5 or more trial values. The Tune set in Table 4 is for final converging local searches.

Computer Run Time

A 486DX2-66 PC running a QuickBASIC® v4.5 executable program FOR-NEXT loop multiplies, divides, adds, or subtracts in 9.5 μs. GRABIM estimates run time in minutes as:

$$T = 1.45\,(9.5\times10^{-6})(NS)(NT^{NB})\,(24 + 7.7(NB))/60 \qquad (1)$$

where NS is the number of frequency samples NT is the number of trials, and NB is the number of branches. The parenthetical term represents the number of mathematical operations required to calculate S_{21} as discussed later. The 1.45 factor is the overhead due to program execution. The exponential term is the number of combinations shown in the FOR-NEXT loops in Figure 3.

From interpolation theory and optimization experience, the number of frequency samples should never be less than the number of branches, NB, which is equivalent to the number of optimization variables. Table 5 shows the number of combinations per frequency and Table 6 shows the estimated time for NS=8. Trial sets 3F and 3N (Table 4) are clearly appropriate when more than six branches are explored. The run time in equation 1 is conservative, since topology codes may be all positive, and there may be bypassed trials due to zeros of transmission.

Efficient Network Analysis

GRABIM was developed on the basis of the fewest possible calculations per trial. Of the many ways to calculate S_{21} for a ladder network in the double match arrangement shown in Figure 1, there are at least two that are very efficient. The recursive update method [2] assumes a current in Z_L (see Figure 2) and computes voltages and currents for the lossless branches from load to source. Another method independent of Z_L has been described by Orchard [6] and was chosen for this application. The ABCD (chain) matrix of the lossless network is constructed by adding one branch at a time, starting with branch 1 (next to the load). Both methods require only five operations (multiplications and additions) per lossless branch; however, there is some overhead in either method to increment the branch number and determine the series or parallel connection. Therefore, 7.7, not 5, is used in timing equation (1).

A significant time saver is the pre-calculation of all immittances used within the FOR-NEXT loops in Figure 3. Therefore, *all* combinations of ωL, ωC, their negative reciprocals, and certain terms involving ZS and ZL are calculated and stored for all sample frequencies. Values needed

NT	Trial#1	2	3	4	5	6	7	8	9	10	11
G	0.01	100	0.1	10							
P	1	0.01	100								
3N	1	0.8	1.25								
3F	1	0.4	2.5								
T	1	0.95	1.05								
5	1	0.6	1.7	0.3	3						
7	1	0.7	1.4	0.4	2.5	0.25	4				
9	1	0.8	1.2	0.6	1.7	0.4	2.5	0.25	4		
11	1	0.8	1.2	0.6	1.7	0.4	2.5	0.3	3	0 2	5

Table 4. Trial values (factors) for search about base points in log space.

NB	NT=3	5	7	9	11
2	9	25	49	81	121
3	27	125	343	729	1,331
4	81	625	2,401	6,561	14,641
5	243	3,125	16,807	59,049	161,051
6	729	15,625	117,649	531,441	1,771,561
7	2,187	78,125	823,543	4,782,969	19,487,171
8	6,561	390,625	5,674,801	43,046,721	214,358,881

Table 5. The number of combinations for NT trial values in NB branches.

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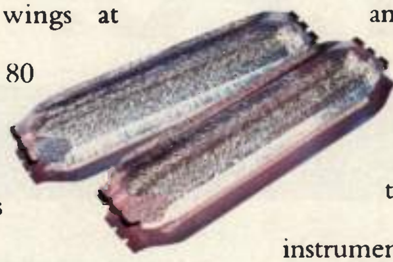

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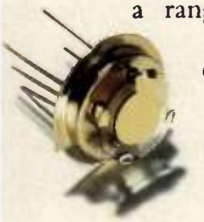

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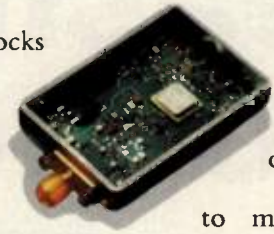

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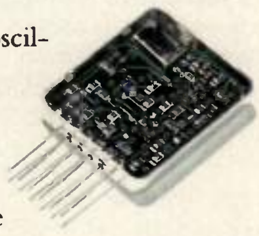

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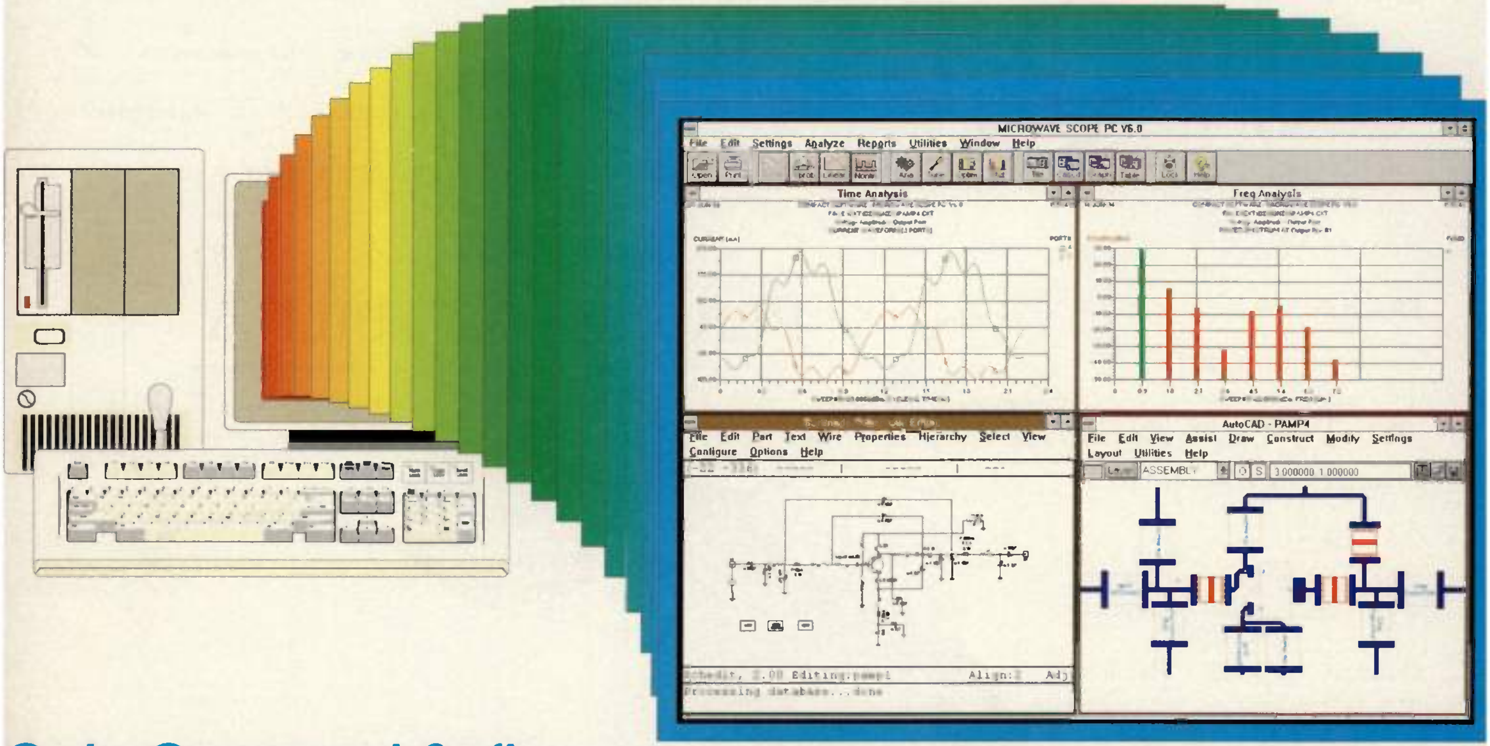

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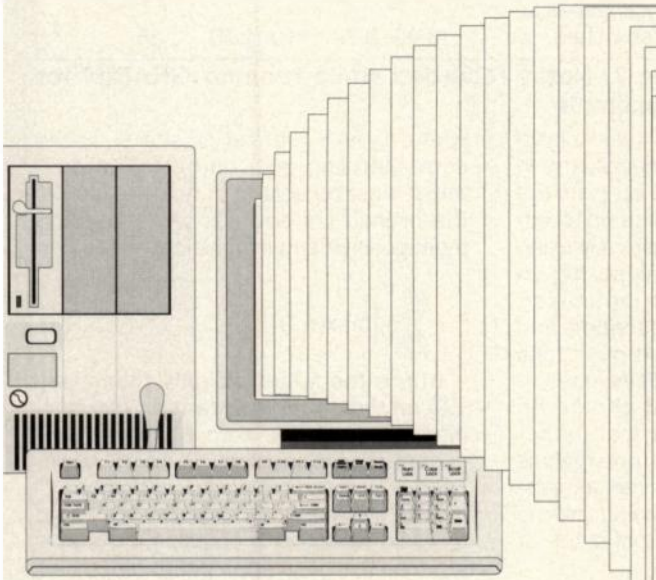

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NB	NT=3	5	7	9	11
2	0	0	0	0.01	0.01
3	0	0.01	0.03	0.06	0.12
4	0.01	0.06	0.24	0.66	1.47
5	0.03	0.36	1.93	6.78	18.5
6	0.09	2.01	15.2	68.5	228
7	0.31	11.2	118	684	2,788
8	1.03	61.4	906	6,768	33,701

Table 6. Minutes required for combinations in Table 5 with eight frequencies.

S21 dB and numbers of trial values: 21.11(5)3 2.11 (T)3 1.84						
Using topology codes:	2	2	-2	3	3	3
For branch numbers:	6	5	4	3	2	1
Branch values after grid searches:	2.48	0.63	2.90	2.85	1.62	0.02
Using topology codes:	2	2	-2	3	-3	
For branch numbers:	5	4	3	2	1	
Branch values after grid searches:	0.85	0.72	110	2.37	2.26	

Table 7. Notes recorded while running GRABIM for the example.

for any new base point, not just unity, are also calculated outside the FOR-NEXT loops. Calculation of the insertion loss using real ABCD variables is [7]:

$$S_{21} = \frac{2\sqrt{R_S R_L}}{AZ_L + jB + jC(Z_S Z_L) + DZ_S} \qquad (2)$$

Use of pre-calculated source and load terms in the real and imaginary parts of (2) reduces the work to only 24 operations, as indicated in timing equation (1).

Constrained Optimization

Program GRABIM helps identify promising matching network topologies and their approximately optimal L and C values. Extreme values (e.g., 0.1 or 10) obtained for some branches may indicate that those branches are unnecessary. In any case, a vital second step in the matching procedure is to force unnecessary branches out of the solution by using a gradient optimizer while imposing an optimal match. One such optimizer, TWEAKNET, has been described and made available [8], although provisions for sets of Z_L and Z_S were not included.

The results of GRABIM should be transferred to an optimizer that further minimizes S_{21} dB to satisfy goal values specified at a set of frequencies as illustrated in Figure 5. The residual arrows, r_i, shown in Figure 5 are the errors at the ith frequency:

$$r_i = L_i - G_i \qquad (3)$$

where L_i is S_{21} dB at ω_i and G_i is the corresponding goal (target). The optimizer must be able to vary the values of the branch L's and C's while keeping them positive and minimizing:

$$F = \sum_{i=1}^{NS} s_i \left[\max(r_i, 0)\right]^2 \qquad (4)$$

where the s_i are weights (nominally 1.0) on the ith residual (error). The minimum value of F is zero when all r_i are either zero or negative. Figure 5 represents a non-optimal choice of branch LC values, showing that residuals r_1, r_2, and r_3 would be active in equation 4. As written, equation 4 is appropriate for $L_i < G_i$ for all i. To force $L_i = G_i$ for all i (as opposed to forcing $L_i \geq G_i$), equation 4 is

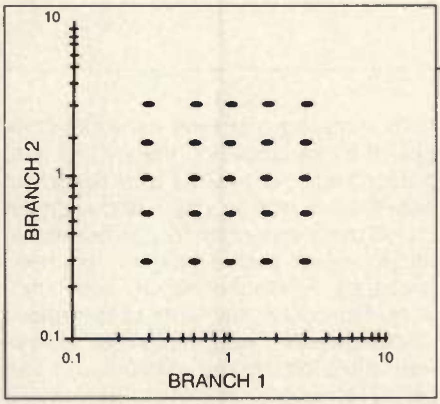

Figure 4. The combinations of five values for two branches.

changed to a simple sum of weighted, squared residuals. In either case, the weights, s_i in equation 4, may have to be drastically increased in order to force some residuals to zero, assuming that is even possible. Simple weighted least-squares minimization is seldom good enough.

A method that minimizes equation 4 or its $L_i = G_i$ form without having some weights tend to infinity employs a *multiplier penalty function* in the *augmented Lagrangian method* [3] [8]. A description of this method is beyond the scope of this article, but the weighted squares form of (4) is still used during a *sequence* of unconstrained optimizations. Between optimizations, the weights (s_i) and also 'slacks' for the goals are adjusted according to a well-grounded theory. Fortunately, the squared-error form of (4) makes available the Gauss-Newton unconstrained minimization technique [8] as opposed to other more general but less powerful methods (e.g., Fletcher-Powell 's quasi-Newton method). The Gauss-Newton method converges quadratically while most other minimization methods converge linearly, i.e. slowly.

An Example Using GRABIM

This problem was introduced by Yarman as example 3 in both [9] and [10]. The eight data samples in Tables 1-3 describe this problem, which was constructed from a load consisting of a series L and parallel LR and a source consisting of a series resistance bypassed by a series LC branch. The latter branch is resonant at 1.29 radians/second, which accounts for the decreasing source resistance versus frequency in Table 3. Yarman states that this example is quite unusual in that the terminating networks have zeros of transmission at dc, infinity, *and* at a real frequency. Knowledge of the termination circuit models is merely convenient for subsequent frequency sweeps and establishes that analytic matching theory

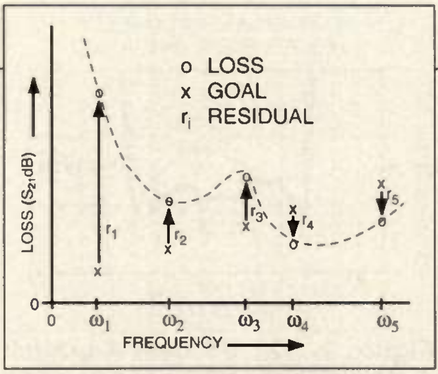

Figure 5. A sampled S21 loss response compared to goals yields residual errors.

is not readily applicable.

GRABIM's main menu is shown in Figure 6. The ASCII frequency sample and goal data file in Table 1 is mandatory and is read by choosing command #1. The optional load and/or source data sets are read using command #3. The grid search menu appears when command #4 is selected and progresses as shown in Figure 7 for this example. Looking at line 7, the question of major importance is what topology should be tried? Obviously, the more that is known about a problem, the more likely an appropriate topology might be found through an intelligent guess.

In this example, study of Tables 2 and 3 indicates that the matching network adjacent to the load should be capacitive and that adjacent to the source should be inductive. Also, it is seen that the resistances are changing substantially with frequency for both the load and the source. These circumstances and Norton transformations [2] suggest that perhaps the PiCPiL topology in Figure 2(c) would be a good first try. Furthermore, if the connection to source and/or load should be series rather than parallel, then the exact Pi-to-T conversion (Figure 6, command #5) would render the decision moot. So, Figure 7 shows that PiCPiL is selected with the Pi of C 's next to the load. Then 7 trial factors are selected.

```
*****THE GRABIM MAIN MENU*****
(USE UPPER CASE ONLY)
1. RECALL SAMPLES AND GOALS FROM
DISK
2. DISPLAY SAMPLES AND GOALS
3. RECALL/REVIEW SAMPLED LOAD AND
SOURCE DATA
4. GRID SEARCH
5. Pi-to-T and T-to-Pi CONVERSIONS
6. QUIT THIS PROGRAM
****************************************
```

Figure 6. GRABIM main menu.

However, GRABIM estimated a run time (on a 486DX2-66) of 15.17 minutes, and the program allows subsequent choices, e.g., 5 trials, which is estimated to run 2.01 minutes. That is accepted (but can be aborted at any time by pressing any key). The initial S_{21} dB for all L 's and C 's set to unity is 21.11 dB, and after 1.91 minutes all possible combinations have been tried. The best found on this search is 3.03 dB with the normalized LC values shown.

Table 7 displays a shorthand notation of what has been tried and what was achieved. The first line in Table 7 shows that the first grid search starting at all unity branches obtained S_{21}=21.11 dB, and, after repeating 5 trial values three times, 2.11 dB was obtained. Then 3 tuning sequences reduced S_{21} dB to 1.84 dB; the decision to change/stop the grid searches was made when improvement was uninspiring. Lines 2 and 3 of Table 7 show the topology codes and branch

```
RESPONSE TYPE SET TO S21 dB,
WANT TO SEE THE SET OF GOALS _/N)? N
SEARCH FOR BEST Equality OR Less-Than CONSTRAINT (E/_)?
  A LOAD DATA SET EXISTS
  A SOURCE DATA SET EXISTS
RETURN TO READ ZL or ZS DATA SET(S) (Y/_)?
ENTER YOUR OWN TOPOLOGY OF +/-2's OR 3's (Y/_)?
IS BRANCH NEXT TO LOAD IN PARALLEL OR SERIES (P/S)? P
BP, LP, HP, LPHP, HPL or PiLC TOPOLOGY (B/L/H/LH/HL,P)? P
  Pi NEXT TO LOAD IS ALL L's or ALL C's )L/C)? C
CHANGE BASEPOINT FROM UNITY Except for GROW (Y/_)?
Prune, Grow, Tune, OR # TRIAL VALUES (P,G,T,or 3N,3F,5,7,9,11) = ? 7
  ESTIMATED 486/66DX2 RUN Minutes (Except GROW)= 15.17
  IS RUN TIME OK (Y/_)?
ABANDON SEARCH CMD (Y/_)?
Prune, Grow, Tune, OR # TRIAL VALUES (P,G,T,or 3N,3F,5,7,9,11) = ? 5
  ESTIMATED 486/66DX2 RUN Minutes (Except GROW)= 2.01
  IS RUN TIME OK (Y/_)? Y
************ CAN ABORT SEARCH BY PRESSING ANY KEY ***********
GRID 1.00 0.60 1.70 0.30 3.00
0.00  0.00  2.00  2.00  -2.00  3.00  3.00  3.00 <—TOPOL CODES
<——TOWARD SOURCE   TOWARD LOAD——> RESID
BR#8 BR#7 BR#6 BR#5 BR#4 BR#3 BR#2 BR#I S21dB MINs
0.00  0.00  1.00  1.00  1.00  1.00  1.00  1.00  21.11  0.00
0.00  0.00  1.00  1.00  1.00  1.00  1.00  1.70  20.72  0.00
0.00  0.00  1.00  1.00  1.00  1.00  1.70  1.00  15.73  0.00
0.00  0.00  1.00  1.00  1.00  1.00  3.00  1.00  12.38  0.00
0.00  0.00  1.00  1.00  1.00  1.00  3.00  0.60  11.67  0.00
0.00  0.00  1.00  1.00  1.00  1.00  3.00  0.30  11.17  0.00
* * * Screen output omitted * * *
0.00  0.00  1.00  0.60  3.00  1.70  3.00  0.30  4.97   0.17
0.00  0.00  1.00  0.60  3.00  3.00  3.00  0.60  3.98   0.18
0.00  0.00  1.00  0.60  3.00  3.00  3.00  0.30  3.55   0.18
0.00  0.00  1.70  0.60  1.70  3.00  3.00  0.30  3.10   1.03
0.00  0.00  1.70  0.60  3.00  3.00  1.70  0.30  3.03   1.07
TOOK 1.91 MINUTES & 125000 TRIALS.
CONTINUE WITH A SUB-SEARCH ABOUT THIS POINT (_/N)?
```

Figure 7. GRABIM search screen for broadband match example.

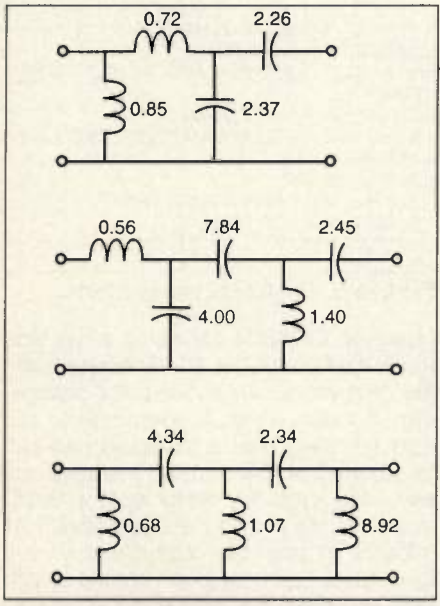

Figure 8. a) Best case found: 0.97 ≤dB≤1.24 (see Figure 9), b) 5-branch BP 1.21≤dB≤1.63, c) 5-branch HP 0.88 ≤ dB≤1.75.

numbers tried first (from the screen output in Figure 7). The element values found by the grid search are shown in the 4th line of Table 7. Clearly, the first branch is unnecessary (less than 0.04), so it was removed, leaving a 5-element network. Lines 5 and 6 record the new network, and the last line of Table 7 shows the element values obtained by starting a *bounded and constrained* optimization from the grid results, thus causing the branch 3 parallel inductor to vanish and leaving the matching network of Figure 8(a).

The S_{21} dB versus frequency responses are shown for the grid and constrained (less-than) optimizations in Figure 9. The grid optimization found a worst-case (over frequency) of 1.84 dB using five branches. Constrained least squares produced a worst case of 1.24 dB over the frequency band, using only four branches. Yarman 's solution [10] obtained nearly the same S_{21} response but with a network consisting of 3 inductors, 2 capacitors, and 1 ideal transformer. It should be added that GRABIM found two other topologies, Figures 8(b) and (c), that produced worst cases of 1.63 and 1.75 dB over the band, respectively.

Of the many test cases run using GRABIM and constrained optimization, it was found that some solutions could have been obtained directly by constrained optimization starting with unit elements *in the nearly-optimal topology*. This was almost never the case with least squares optimization, but starting least squares from the GRABIM solution occasionally gave a nearly optimal solution.

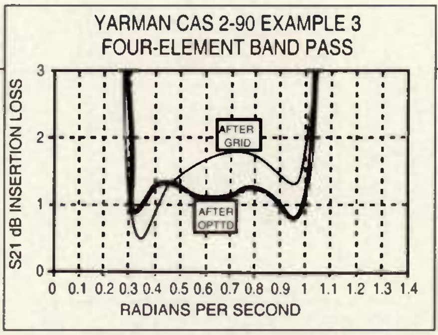

Figure 9. S21 dB best 4-branch matching network from Figure 8a and Table 7.

Other Broadband Matching Methods

A recent article by Dedieu et. al. [12] describes a broadband impedance matching method which uses a strategy similar to GRABIM's at its highest level. The method is called the RSE method (Recursive Stochastic Equalization). It progresses in the same three basic steps as GRABIM: (1) an LC matching network topology is assumed, (2) the space of possible component values is searched for the best match over the sampled frequencies, and (3) the approximate component values are then refined for the best solution.

There are fundamental differences in RSE and GRABIM. In step (2), GRABIM exhaustively searches the component space for a global minimax solution, using sparse computations. RSE uses a stochastic Gauss-Newton least-squares search which requires first derivatives of the transfer loss with respect to each LC component at just one of the frequencies, estimates a matrix of second derivatives, and may converge to a local minimum. In step-3, GRABIM uses a deterministic Gauss-Newton optimizer adapted to refine the minimax solution. RSE uses a random search in a limited region about the step-2 solution to find a minimax solution, hopefully the global solution.

Conclusions

At the very least, GRABIM provides a fast way to evaluate suitable matching network topologies. It also provides a good starting point for LC values used in an optimizer, especially one that is bounded (maintains positive element values). This approach works even better if the optimizer has less-than or equal-to goal constraint capabilities. Including GRABIM as a feature in an optimizer program would be a very convenient arrangement because the grid-to-optimization transfer would be automatic and seamless. This has been accomplished in program CONETOPM (Constrained Network Optimizer for Matching).

There are two possibilities for applications requiring matching networks composed of transmission lines. One is to perform an exhaustive grid search in parameter space for open- and short-circuited stubs separated by cascade transmission lines; such a program has been prepared. A second way to accommodate distributed elements is to replace selected lumped sections by their almost-equivalent distributed network counterparts [11].

 RF

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About the Author

Thomas R. Cuthbert Jr., Ph.D., PE, is a consultant and teacher based in Plano, Texas. He was Director of Advanced Technology at Rockwell International, and Manager of Microwave Technology at Texas Instruments. He studied at M.I.T., Georgia Tech, and S.M.U. His two John Wiley books are: *Circuit Design Using Personal Computers* (1983) and *Optimization Using Personal Computers* (1987). He can be reached at 1709 Hastings Ct., Plano, TX 75023.

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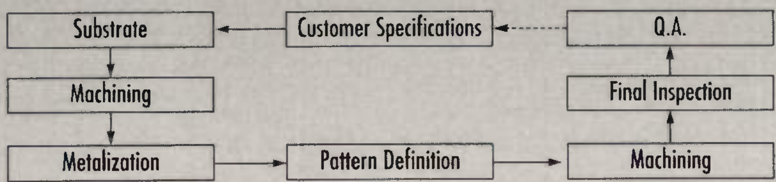

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Designing Accurate Small Inductors for Microwave L-C Filters

By Albert Klappenberger
ALK Engineering

The design of small air-wound inductors for application in lumped component filters has traditionally been done using the well known formula shown below (equation 1). This equation can be found in many texts and has been widely accepted as the "final" answer to the problem of coil winding:

$$L(\mu H) = n^2 \frac{r^2}{9r + 10\,len} \qquad (1)$$

where
n = number of turns
r = radius to wire center (in.)
len = length of coil (in.)

The texts attribute about 1 percent accuracy to this formula. Experience has shown that the error is significantly worse than this when designing the small coils needed for high frequency filter applications. One extra turn often must to be added to obtain the correct inductance on such coils. One attempt to compensate was exaggerating the length of the designed coil so that the coil could be wound with a tighter winding pitch when the winding was actually made, thus increasing the inductance. It was obvious that a better way had to be found. A number of years ago, a book entitled Inductance Calculations Working Formulas and Tables by Frederick W. Grover came to the author's attention. From this book, an equation attributed to Nagaoka is given below:

$$L(\mu H) = 0.00508\,\pi^2\, r \left[\frac{2r}{len}\right] n^2 K \qquad (2)$$

This equation is based on an equation for the inductance of a cylindrical current sheet of finite length. A current sheet is a winding where the current flows around the axis of a cylinder in a layer of infinitesimal radial thickness in the surface of the cylinder. The equation applies a correction factor K to compensate for the ends.

This K factor was presented in Mr. Grover's book in the form of two lengthy tables, arranged as a function of shape ratio (len/2r, or length/diameter). The first table was for short coils with a shape ratio less than 1.0, and the other was for long coils with a shape ratio greater than 1.0. The two correction factor tables were broken into segments, and multiplied by 100 to increase accuracy and curve fit.

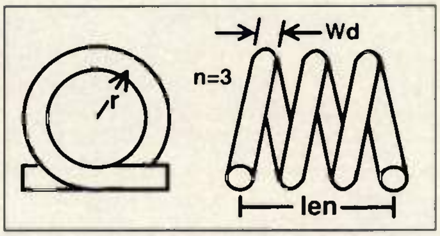

Figure 1. A typical small inductor.

A program incorporating this design equation and these correction factors has been written. The coefficients were written directly to disk by the curve fit program and merged into the program to eliminate the possibility of a typographical error. The curve fits appear in the BASIC program "LX" as the defined functions FNKS() and FNKL() respectively. Because the inductance provided by a coil of round wire separated by an air gap is different from that wound of a cylindrical current sheet, an additional correction is required. This correction Ld, is related to winding pitch:

$$Ld = .01016\pi\, r\, n(G + H) \qquad (4)$$

$$G = 1.25 - \log_e\left[\frac{2\,len}{Wd \cdot n}\right] \qquad (5)$$

Wd is the wire diameter in inches.

The additional correction term H above was also given in the form of a table which was segmented and curve fit. This fit can be found as FNH() in the LX program. The Ld factor must be subtracted from the result of the Nagaoka formula given earlier.

Results

To illustrate the difference between the two methods, the inductance of a typical small inductor as calculated by the LX program is shown below:

(Cr to end) Form dia. (in.)?	0.062
Winding length (in.)?	0.0664
Wire gauge (AWG)?	28
No of turns ?	4
	0.0200 µH.

The inductance of this coil by the standard formula (the diameter of #28 AWG wire is 0.0136 in.) is:

$$\frac{4^2(0.0378^2)}{9(0.0378) + 10(0.0664)} = \frac{0.022861}{1.0042} = .02277\mu H$$

For a disagreement of roughly 13%

For a large coil, as would normally be used at lower frequencies, the two methods agree to better than 0.4%, as illustrated below:

(Cr to end) Form dia. (in.)?	2
Winding length (in.)?	2
Wire gauge (AWG)?	28
No of turns ?	100
	347.4765 µH.

The standard formula yields:

$$\frac{100^2 \cdot 1.0068^2}{9 \cdot 1.0068 + 10 \cdot 2} = \frac{10136.46}{29.061} = 348.8\mu H$$

Effects of Frequency on Inductance

Inductors at microwave frequencies are necessarily physically small. The effect of these miniature dimensions on the time-honored formula for inductance is perhaps the biggest effect high frequency has on inductors for microwave L-C filters. The formulas given here go a long way toward improving the situation, particularly the G factor. This is not to say that frequency does not actually have an effect on the true inductance of any specific coil. The true effect seems to be a reduction in the effective diameter of the winding as frequency increases. Mr. Grover states that the limit of this reduction is to the surface of the wire on the inside of the winding, that is, to the form diameter. Needless to say, the thickness of the wire compared to the form or winding mandrel diameter with coils of this size could be significant. The actual diameter reduction with frequency is almost impossible to cal-

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culate because of the mathematical complexity of the modeling involved.

PCAIRL – A Practical Program

The LX program given here is only useful for finding the inductance of a specified coil, or for verifying the validity of the method. Using this procedure to determine the number of turns required to make a coil of a specific inductance, the equation has to be solved for the number of turns (n) in terms of length and form diameter. This cannot be done because of the curve fits involved, so some sort of reverse function iteration is required. Even when using the simple equation given earlier, which can be easily solved for n, the number of turns must be guessed in advance to determine the length of the coil for the formula. It is therefore necessary to iterate to arrive at the desired winding pitch, especially if a close-wound coil is necessary. In addition, most coils for miniature L-C filters must have an integral number of turns, so that the leads both come out of the bottom of the coil for soldering down to chip capacitors below. This requires a third iteration.

Such a program (PCAIRL) has been developed and is available. PCAIRL will design any inductor with reasonable accuracy for values from 10 nH up. The traditional, approximate formula given earlier is used to seed a reverse function iteration consisting of the "LX" program as the function. Both the form diameter and winding pitch can be set to an automatic mode to begin a coil design. Once the exact number of turns is determined, the program will round the number to the nearest integer and either stretch or compress the length of the coil to return to the desired inductance. Functions have also been included to calculate the approximate unloaded Q and self-resonant frequency of the final coil. A screen-formatted display and menu allows the winding pitch, form diameter and wire gauge to be changed to optimize the Q and final shape factor.

Acknowledgements

The LX program was derived from a program written by my late friend Fredrick J. Radler. I would also like to thank William B. Lurie for encouraging me to make PCAIRL available and to review and expand on the work done by Mr. Radler. The PCAIRL program is being supplied as part of the PCFILT filter design CAD package from ALK Engineering.

The LX (source and executable) and PCAIRL (executable) are available from Argus Inc., Direct Marketing Dept. See page 124 for ordering information. ***RF***

About the Author

Albert Klappenberger is the proprientor of ALK Engineering and has recently worked as a consultant for a major filter manufacturer. He, and the late Fredrick J. Radler, developed PCFILT, a filter design program which designs several classes of filters. Mr. Klappenberger can be reached at 1310 Emerson Ave., Salisbury, MD 21801, or by phone at (410) 546-5573.

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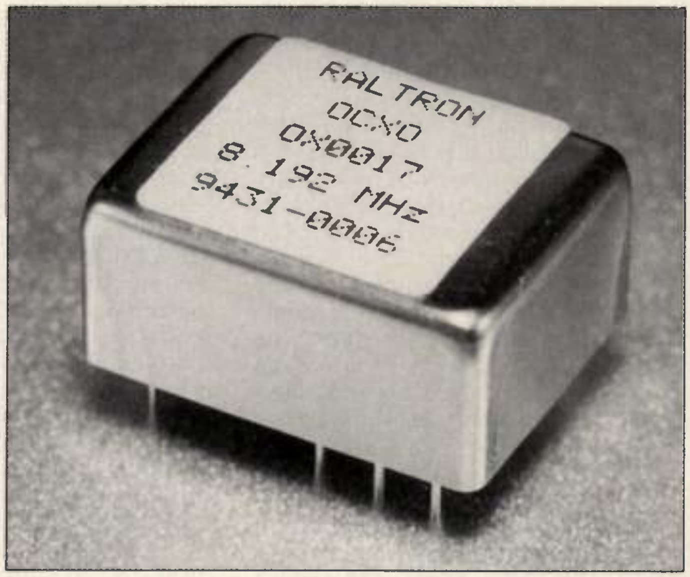

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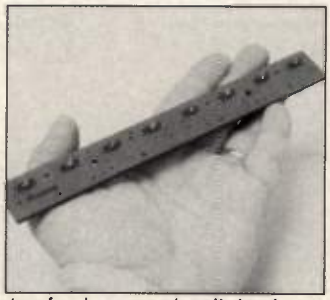

tor for improved cellular base station gain and selectivity. Insertion loss is less that 1.1 dB, VSWR is less than 1.2:1, amplitude balance is typically 0.15 dB rms, and phase balance is typically 2° rms. The package measures only 7.5 inches long, 1.1 inches wide and 0.15 inches thick. The beamformer contains sixteen 90° hybrid couplers and over 20 feet of stripline stacked on three layers. This new caseless, multilayer technology can be connectorized or surface mounted. Anaren's 8x8 beamformer is reliable in extreme environmental conditions, including space.

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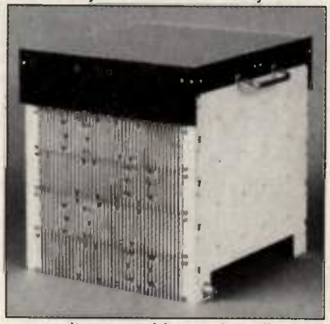
INFO/CARD #212

HF Antenna Tuner

The Act-Fast HF antenna tuner from AEL Industries matches 15 foot or longer whip, wire, loop and tower antennas in less than 50 µsec. Operating in the frequency range of 2 to 30 MHz, the tuner uses four patented "P sections" which are magnetically controlled, varactor-tuned planar circuits and 18 patented high speed/high power FET switching modules. The tuner provides "instant" reactive impedance matching and can be used in HF frequency hopping applications when controlled by AEL's Act-Fast frequency hopping controller. The tuner does not use mechanical parts such as vacuum relays or mechanically tuned



capacitors, and it can handle 500 W CW and 2000 W peak. Input VSWR is typically less than 1.5:1, 2.5:1 max. The tuner measures 16 x 16 x 16 inches and weighs less than 75 pounds. DC power consumption is less than 150 W.

AEL Industries, Inc.
INFO/CARD #211

Product Spotlight: Transmitter and Receiver Modules

Telemetry Transmitters

Digital, video, and telemetry transmitters from Emhiser Research utilize a patented, wideband, DC-coupled, phase-lock loop. Bit rates in excess of 20 Mbps or video bandwidths in excess of 10 MHz are accommodated. For digital transmitters, this development permits wider use of NRZ PCM – while in video links, the baseline hop is eliminated. Units are available in single frequency or channelized – from 100 MHz to 2.4 GHz – and in various powers.

Emhiser Research, Inc.
INFO/CARD #210

2.4 GHz SS Transceiver

GRE America has introduced two new additions to the GINA family of products. The GINA 7000N and 8100N operate in the 2.4 GHz frequency range, incorporating the latest spread spectrum technology, at data rates up to 256k baud. Both units feature 500 mW transit power, four PN code sequences and optional voice communication capability. The model 7000N has 37 selectable channels, while the 8000N can use 19.

GRE America, Inc.
INFO/CARD #209

SEMICONDUCTORS

WLAN Transceiver Chip

Daico Industries has introduced the model P35-4710-1 single chip GaAs MMIC transceiver for wireless LAN applications at 2.4 GHz. The transceiver includes a VCO, upconverter, pre- and post-amplifier, T/R switch, LNA, downconverter and buffer amplifiers. No external matching, coupling or decoupling components are required. DC current consumption is 38 mA in receive mode, 180 mA in transmit mode, and 0.5 mA in standby mode. The transceiver is packaged in a 7 mm square package.

Daico Industries, Inc.
INFO/CARD #208

500 Msps 8-bit ADC

The MAX100/MAX101 analog-to-digital converters include a high-performance track-and-hold plus quantizer with a input bandwidth of 1.2 GHz and aperture jitter of less than 2 ps. This results in 7.1 effective bits at the devices' highest Nyquist frequency. Probability of erroneous codes is one in every 10^{15} clock cycles. The MAX100's maximum sampling rate is 250 Msps, while the MAX101's is 500 Msps. Prices start at $265 for quantities of 100 or more.

Maxim Integrated Products
INFO/CARD #207

Substrate Amplifiers

The new AFSB series of substrate amplifiers are hermetically sealed amplifiers with a 50 ohm line substrate interface for the RF in, RF out, and DC power connections. The series is available in either the 0.1 to 20 GHz range, or optimized in octave or multi-octave

bands. The substrate height can be made to the customer's system requirements.

Miteq
INFO/CARD #206

Burst Processors

Stanford Telecomm's ASIC & Custom Products Division now offers the STEL-2000A+45, at 45 MHz maximum clock frequency; and the STEL-2000A+20, at 20 MHz. Both models provide IF to TX/RX data on a single chip. The STEL-2000A+45 processes data at rates up to 2.048 Mbps with processing gain of up to 18 dB. The STEL-2000A+20, which is available at significantly lower prices, has a maximum data rate of 900 kbps.

Stanford Telecom
ASIC & Custom Products Div.
INFO/CARD #205

1 GHz Op Amp

Burr-Brown's OPA648 ia a high speed current feedback operational amplifier offering a 1 GHz unity-gain bandwidth. The OPA648 achieves this high bandwidth while consuming only 13 mA of quiescent current. The device has a 12-bit settling time of 20 ns to 0.01%, differential gain and phase errors of 0.02%/0.02°, and a 9 dB noise figure. The OPA648 operates from ±5 V and is offered in 8-pin plastic DIP, SOIC and hermetic ceramic packages. Pricing is from $5.95 in 100 piece quantities.

Burr-Brown Corp.
INFO/CARD #204

2.7 V Transistors

Hewlett-Packard Co has introduced a series of silicon bipolar transistors that are optimized for 3 V operation. They are capable of providing exceptional noise figure and gain performance under low bias. They can be used between 1 and 5 V, with as little as 0.5 mA of current. These transistors have f_{max} of 30 GHz. The AT-30533 and AT-31033 transistors are supplied in the SOT-23 surface mount package. The AT-30511 and AT-31011 are offered in the SOT-143 package. pricing starts at $0.43 in quantities of 10k or more.

Hewlett-Packard Co.
INFO/CARD #203

DISCRETE COMPONENTS

Crystals
The XT49S is a low profile (0.152 inch) quartz crystal with a frequency range of 3.5 to 36 MHz. The XT49M has identical operating characteristics, but is configured for surface mounting. Model XT49U is available in both through-hole and surface mount configurations and has an optional third lead. Its frequency range is from 1.00 to 125 MHz.
Dale Electronics, inc.
INFO/CARD #202

Low-Cost SMT SAW Resonators
Hermetic, surface-mount SAW resonators are now being offered by RF Monolithics (RFM). The new RFM package is 0.235 x 0.155 x 0.085 inches. The RFM resonators use the industry standard one-port configuration, and offer nominal 1.5 dB insertion loss (2.0 dB maximum). In quantities of 10,000, the surface mount devices cost $1.34 each, including tape and reel packaging.
RF Monolithics, Inc.
INFO/CARD #201

Surface Mount Crystals
The FOX FG's low-profile package (1.5 mm) incorporates a ceramic pad with flat metal leads extending from the base for precise part placement and easy solder inspection. The crystals cover 10.00 to 250 MHz, with fundamental frequencies available to 50.00 MHz. Stand specifications include frequency tolerance of ±10 ppm at 25 °C and frequency stability of ±4 ppm from –10 to +60 °C.
Fox Electronics
INFO/CARD #200

Low-Cost Crystals
M-tron Industries has introduced a line of low-cost crystals designed for more "forgiving" applications, where cost is the critical factor. The MPG series is currently available in frequencies from 3.579545 to 20.0000 MHz. Standard frequency tolerance is ±50 ppm and stability is ±100 ppm. Standard load capacitance is 18 pF, (series resonant is also available).
M-tron Industries, Inc.
INFO/CARD #199

TOOL, MATERIALS & MANUFACTURING

High Tolerance PCB Material
Rogers' RT/duroid® 5880 high performance microwave material has a dielectric constant with tolerance of ±0.01. The material's nominal dielectric constant is 2.20. 5880 is available on 0.020, 0.031 and 0.062 inch-thick laminates, clad with either electrodeposited cr rolled copper foil.
Rogers Corp.
INFO/CARD #198

Prep Tools for LMR Cable
Cable prep tools for LMR-400, 500, and 600 cables are now available. In a simple two-step process, these tools strip the cable end for clamp-style connector attachment,

allowing for high quality assemblies to be fabricated in the field.
Times Microwave
INFO/CARD #197

SIGNAL PROCESSING COMPONENTS

Linear Phase Bandpass Filters
Two bandpass filters from Eastern Multiplexer feature ±5° phase linearity over their passbands. Model 10FBCX-700-40SFF has a passband from 680 to 720 MHz, while model 10FBCX-700-125SFF has a passband from 637 to 763 MHz. both filters have passband VSWR of 1.25:1 and maximum insertion loss of 2.0 dB. Selectivity is 80 dB over 135 MHz for the –40SFF, and 50 dB over 280 MHz for the –125SFF.
Eastern Multiplexers, Inc.
INFO/CARD #196

6 dB Couplers
Polyflon now offers 6 dB couplers. Model PFC-2WL-800-1000-SMA has been designed for cellular telephone applications, and is based on Wilkinson uneven division design. This mocel has 6.2 ±0.1 coupling flatness over the full band of 800 to 1000 MHz, and input VSWR of 1.15:1, output VSWR of 1.2:1, and isolation of 30 dB. Size is 2.0 x 3.0 x 0.5 inches. Unit pricing is $175 for 100 pieces.
Polyflon Co.
INFO/CARD #195

Dielectric Resonator Bandpass Filter
The model CDF-321-15-3 is a ceramic dielectric resonator filter designed to achieve low loss and temperature stability with a bandpass response at 321 MHz. It features a three-section response with a 3 dB BW of 15 MHz and a maximum 20 dB BW of 40 MHz. The insertion loss at 321 MHz is less than 1.5 dB and the passband VSWR is 1.5:1 maximum. Prices start at $225 in unit quantities.
RLC Electronics, Inc.
INFO/CARD #194

Divider/Combiner
Micro Mart introduces a line of power dividers/combiners directed toward the standard telecommunication bands. The products are low cost with repeatable performance from unit to unit. Typical specs, for example, for model P405 (4-way) covering 3.6 to 4.25 GHz with type N connectors, include 20 dB return loss and isolation, 0.6 dB insertion loss and amplitude balance of ±0.2 dB.
Micro Mart, Inc.
INFO/CARD #193

Lowpass Filter
Model 713L-2075/R1000-XP/XP is a lowpass filter that offers less than 1.0 dB insertion loss from DC to 1600 MHz. The VSWR for this unit is 1.5:1 from DC to 1600 MHz, with an attenuation of 80 dB from 2900 MHz to 10 GHz. The package size is 1.5 x 0.38 x 0.38 inches.
K&L Microwave, Inc.
INFO/CARD #192

DECT Filter
The Sawtek 110.592 MHz DECT filter (part # 854502) is well suited for both the subscriber and base station IF filtering requirements. The substrate material is temperature-stable quartz to ensure excellent adjacent channel rejection at ±1.728 MHz, ±3.456 MHz, and ±5.184 MHz. The DECT filter is hermetically sealed in an industry-standard package.
Sawtek, Inc.
INFO/CARD #191

Power Dividers
The PDNL-80 series of in-line 4-way power dividers/combiners, covers 2 to 1000 MHz, in various bands. Each uses a high performance lumped element design to achieve multi-octave bandwidths. Typical isolation across this widest band is 25 dB and better. Insertion loss is minimized, with the 2 to 100 MHz band typical insertion loss specified at 0.8 dB.
Merrimac
INFO/CARD #190

Crystal Filter
Model 8117C from Piezo Technology features a 3 dB bandwidth of ±25 kHz minimum and a stopband attenuation of 65 dB at ±200 kHz maximum. Additional features include 70 dB ultimate attenuation, 40 dB spurious attenuation and 50 ohms input and output impedance. The unit is packaged in a 2.37 x 1.0 x 0.75 inch enclosure that will survive the military environment.
Piezo Technology, Inc.
INFO/CARD #189

SUBSYSTEMS

GPS Antenna
Antenna Research Associates has introduced another microstrip patch antenna for

New System 32 Software...

verify your designs... find better solutions.

Typical design flow

Initial synthesis in =M/FILTER=

Tune or optimize in =SuperStar= simulator

Layout to plotter, printer, HPGL and DXF files

REAL-TIME CIRCUIT SIMULATION
- Accurate models
- Huge library
- Easy nodal entry
- Schematic entry
- Tune by tapping
- Super-fast optimization
- Unrestricted noise
- Statistical analysis
- 3-D plotting

CIRCUIT SYNTHESIS
=FILTER= (L-C filters)
- Coupled-resonator, zig-zag, symmetric, Blinchikoff, conventional, other structures
- All popular transfer shapes
- Group-delay equalizers

=M/FILTER= (microwave filters)
- End, edge, hairpin, combline, elliptic, interdigital, stepped-Z, lowpass, bandstop, highpass
- Microstrip, rod, stripline, coax, generic
- Layout generation

=OSCILLATOR=
- L-C, SAW, crystal, T-line
- VCOs
- Noise estimation

=MATCH=
- Multiple algorithms
- General synthesis & order
- Active or passive
- Arbitrary terminations

=A/FILTER= (active filters)
- Multiple structures
- All popular transfer shapes
- Non-ideal op-amps

System 32 available for...

MS DOS — MICROSOFT WINDOWS — MICROSOFT WINDOWS NT

Windows versions are 32-bit
Complete packages from $695
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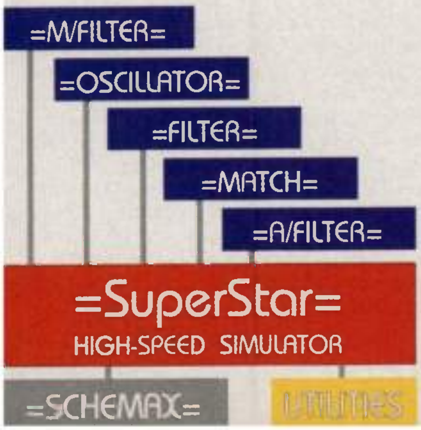
Free technical support
No annual fees



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INFO/CARD 39

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

the GPS marketplace. Model PAS - 1575 is a circularly polarized antenna designed for 1575.4 MHz with less than 1.%;1 VSWR. One output is provided with either SMA or pigtail connection. Typical gain is 5 dBic and typical half power bandwidth is > 80°. The antenna is 3.5 inches in diameter and 0.25 inches thick and weighs less than 3 oz. A flat, hermetically-sealed radome is provided for protection.
Antenna Research Associates Inc.
INFO/CARD #188

Broadband Frequency Translator

Octave-band 4-bit translators from Vectronics Microwave provide high carrier and sideband suppression with up or down translation of up to 0.5 MHz. High-speed TTL drivers step through the unit's 16 phase states, or an optional TTL or voltage-controlled clock can be included. Rugged microstrip construction construction assures stable operation from –55 to +85 °C with RF input power of up to 1 W.
Vectronics Microwave Corp.
INFO/CARD #187

INMARSAT Antenna

Sensor Systems' Aero-L INMARSAT SATCOM antenna (P/N S65-8282-101) conforms with the proposed ARINC 741 specifications for an L-band low gain SATCOM antenna. The antenna is designed to back up ARINC 741 Aero-H communications or initiate low data rate Aero-C communications. The antenna provides 93% sky coverage.
Sensor Systems, Inc.
INFO/CARD #186

TEST EQUIPMENT

DECT Test

An integrated one-box tester delivers fast test times for key DECT manufacturing mea-

surements. The HP 8923A DECT test set is optimized for high-volume DECT manufacturing and can make the accurate, repeatable measurements that are needed to verify correct operation of DECT transceivers. Those measurements include bit-error ratio (sensitivity), carrier power, center frequency accuracy, frequency deviation, power vs. time template, and audio measurements.
Hewlett-Packard Co.
INFO/CARD #185

PC Spectrum Analyzer

The SA2600 turns your PC into a powerful, full-featured spectrum analyzer/receiver. Consuming less than 12 W at 12V, it covers the 100 kHz to 2.4 GHz range. The instrument features fully synthesized tuning using a crystal reference and under 2 Hz frequency resolution provided by a direct digital synthesizer. Signals from –130 dBm to +20 dBm can be measured, and power accuracy is ±1.4 dB. The software supports overlays, trace math, marker functions, autologging and hard copy to a printer. Base price for the SA2600 is $4995.00.
DKD Instruments
INFO/CARD #184

PC Signal Source

Model DDS4 PC from Novatech Instruments is a 34 MHz synthesized signal source on a PC card for use in PC XT and PC AT or later ISA-bus computers. The direct digital synthesizer provides 10 ppm/year stability and excellent spectral purity for only $595. It can generate Sine and TTL clock signals simultaneously from 5 Hz to 34 MHz in 0.02 Hz steps. Phase noise is less than –90 dBc at 1 kHz offset. Output amplitude is 1 Vrms into an open circuit and can be attenuated in 10 dB steps to 70 dB.
Novatech Instruments, Inc.
INFO/CARD #183

Radar Simulator

TSC introduces the 2000 series of low cost radar environment simulators (RES). Configurations range from single channel target and clutter generators to complete multi-channel interactive environment simulators. Each 2000 series simulator includes an IBM compatible PC with one or more RES engine PC boards, two or more radar interface PCBs , software and manuals. A "clutter" engine PCB, scenario generation software, and an expansion chassis to make more slots available are also available.
Technology Service Corp.
INFO/CARD #182

Step Attenuators

Kay Elemetrics' line of manual step attenuators feature a three-watt power rating and an extended frequency range of up to 2 GHz. An example of the series, model 839, has an attenuation range of 0 to 101 dB (50 Ω) with a minimum step of 1 dB. Its frequency range is DC to 2000 MHz, and maximum VSWR is 1.4:1 (found in the 1000 to 2000 MHz range).
Kay Elemetrics Corp.
INFO/CARD #181

Propagation Study Equipment

Berkeley Varitronics Systems announces the newest version of its CHAMP, now kitted with three SPYDER battery-powered rechargeable stick-up transmitters. The kit allows users to place the frequency agile SPYDER transmitters throughout a building or desired area. The data can be collected by

simply walking-about using the CHAMP receiver. The CHAMP's removable PCMCIA memory card stores data, while its internal GPS receiver record latitude and longitude. The complete kit sells for under $10,000.
Berkeley Varitronics Systems, Inc.
INFO/CARD #180

Signal Generator

Signal generator SMY from Rohde & Schwarz can test AM, FM and φM receivers and is useful for component measurements. The SMY comes in two models covering the frequency range from 9 kHz to 1040 MHz or from 9 kHz to 2080 MHz. Its output power range is –140 to +13 dBm. The SMY signal generators have high level-accuracy, low RF leakage and high carrier frequency accuracy. The signal from the integrated AF synthesizer (1 Hz to 500 kHz) can be used externally.
Rohde & Schwarz GmbH
INFO/CARD #179

CABLES & CONNECTORS

Low Intermod Connectors

Intermodulation generated at non-linear passive components, particularly at connection points, can severely detract from the performance of communications systems used in digital cellular applications. Radiall's N Series connectors use a new contact mating concept, selected non-magnetic materials, and a hex coupling nut to reduce this problem. Nominal impedance is 50 Ω, frequency range is DC to 6 GHz, and intermodulation is < –160 dBc.
Radiall Inc.
INFO/CARD #178

75 Ω BNC Connectors

Amphenol RF/Microwave Operations offers a line of 75 Ω connectors, including straight plugs, in-line and mounted jacks, as well as bulkhead receptacles and adapters. Constant 75 Ω performance with low VSWR is provided through 4 GHz. Features include crimp-crimp cable termination for quick and reliable connector assembly and quick connect/disconnect of the BNC two-stud bayonet lock.
Amphenol RF/Microwave Operations
INFO/CARD #177

SIGNAL SOURCES

Low Profile OCXO

Piezo Crystal announces the availability of model 2890080. This oscillator utilizes Piezo's SC cut crystals. The frequency range is from 30 to 110 MHz. Typical phase noise at 100 MHz and 10 Hz offset is –95 dBc/Hz and –125 dBc/Hz at 100 Hz offset. The aging rate at time of shipment is $3x10^{-9}$/day. Frequency stability is $\pm 5x10^{-8}$ over –40 to +70 °C. Model 2890080 measures 2.00 x 2.00 x 0.75 inches and is priced at $400 to $500 in 500-piece quantities.
Piezo Crystal Co.
INFO/CARD #176

TCXOs

Jan Crystals is now offering a line of temperature controlled crystal oscillators (TCXOs). The oscillators operate from 10 to 20 MHz and provide ±2.5 ppm frequency stability from –30 to +75 °C.
Jan Crystals
INFO/CARD #175

AMPLIFIERS

HF Amplifier

Ameritron's ALS-600 includes an AC power supply, 600 W PEP/500 W CW output power, continuous 1.5 to 22 MHz coverage, instant bandswitching, no warm up and is fully SWR protected. The amplifier measures 6 x 9.5 x 12 inches and weighs only 12 pounds. The ALS-600 sells for $1299.
Ameritron
INFO/CARD #174

1.5 - 1.6 GHz LNA

Model VMA 1.6C-126, designed for Inmarsat front-end applications, offers a 0.5 dB noise figure (36 K) across 1.5-1.6 GHz. The LNA has 24 dB gain, ripple of ±0.1 dB and a +16 dBm 3OIP. Size is 1 x 1 x 0.22 inches, and connectors are SMA-F. DC power consumption is 75 mA at +15V.
Veritech Microwave, Inc.
INFO/CARD #173

Wideband Current-Feedback Op Amps for RF Applications

By Michael Steffes
Comlinear Corporation

With operating frequencies pushing 1 GHz, current-feedback operational amplifiers are becoming an attractive alternative to more classical fixed-gain RF amplifiers. However, typical operational amplifier specifications do not include many of the specifications familiar to RF engineers. To help the designer exploit the many advantages these amplifiers can offer, this article will define the RF specifications of most interest to designers, detail what determines each of these particular performance characteristics for current-feedback op amps, and discuss performance optimization techniques where possible.

To apply op amps to RF applications, questions in three areas must be addressed:

- General op amp operating considerations
- Small signal AC performance in an RF context
- Typical limits to RF amplifier dynamic range applied to op amps

The CLC404 will be used as an example in this discussion as a typical, high-speed current-feedback op amp. This particular part is a ±5 V monolithic amplifier intended for use over a linear voltage gain range of ±1 to ±10 V/V. At its midrange gain of +6, the CLC404 offers a DC to 175 MHz frequency range while delivering 12 dBm power into a 50 Ω load. The best amplifier for a particular application will depend upon the desired gain, power output, frequency range, and dynamic range requirements. The higher speed monolithic parts are typically constrained to lower supply voltages, and therefore lower output powers, than hybrid current-feedback amplifiers. The speed and output power of monolithic devices should increase as improved processes become available.

Operating Advantages of Current-Feedback Op Amps

The current-feedback op amp, first commercialized by Comlinear Corporation, is a very wideband, DC-coupled op amp that has the distinct advantage of being relatively gain-bandwidth independent. As with all op amps using a closed-loop negative-feedback structure, the frequency response for current-feedback op amps is set by the loop gain characteristics. The current-feedback amplifier's key development is the de-coupling of the signal-gain and the loop-gain within its transfer function [1]. This de-coupling allows the desired signal gain to be changed without radically impacting the frequency response. If compared to voltage-feedback amplifiers, which are constrained to a gain-bandwidth product operation, the current-feedback topology offers truly impressive equivalent-gain-bandwidth products (e.g. the CLC401 at a gain of 20 V/V yields a flat response with a –3 dB bandwidth of 150 MHz. To match this, a voltage feedback op amp would require a 20*(150) MHz = 3 GHz gain-bandwidth product).

One of the most significant changes in moving from a classical, fixed-gain RF amplifier to an op amp is the exceptional flexibility offered by the op amps. The designer is now charged with setting up the proper operating conditions for the op amp, defining the gain, and determining the I/O impedances with external components. Op amps allow the designer the option of running either a non-inverting or an inverting gain path. For RF applications, the 180° phase shift provided by the inverting mode is often incidental. There are, however, advantages and disadvantages to each mode, depending on the desired performance, and both will be considered at each stage of this discussion.

Most of the information developed here on applying op amps to RF applications applies to any type of op amp. The unique advantages of the current-feedback topology are its higher frequency capabilities and its intrinsically low distortion at low operating currents. If not specifically stated as being unique to the current-feedback topology, the items considered here apply equally to a voltage-feedback op amp.

As a starting point for using op amps

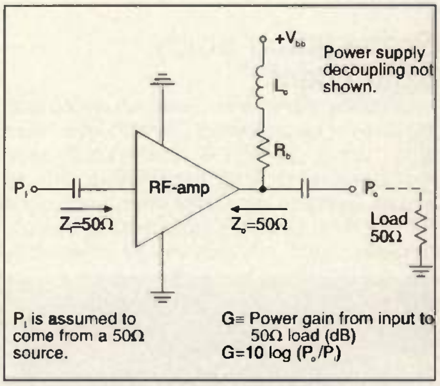


Figure 1a. Typical RF amplifier connection.

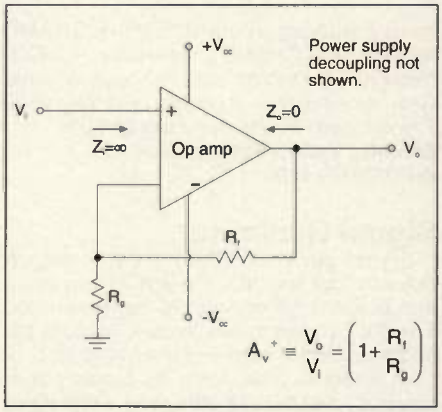


Figure 1b. Ideal non-inverting op amp.

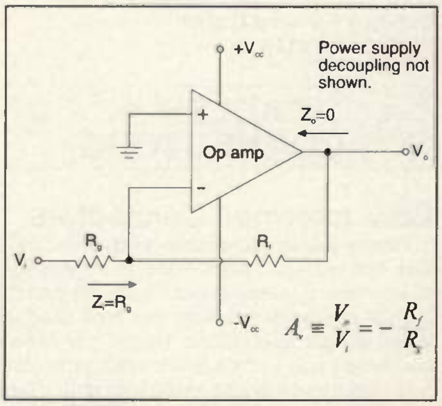
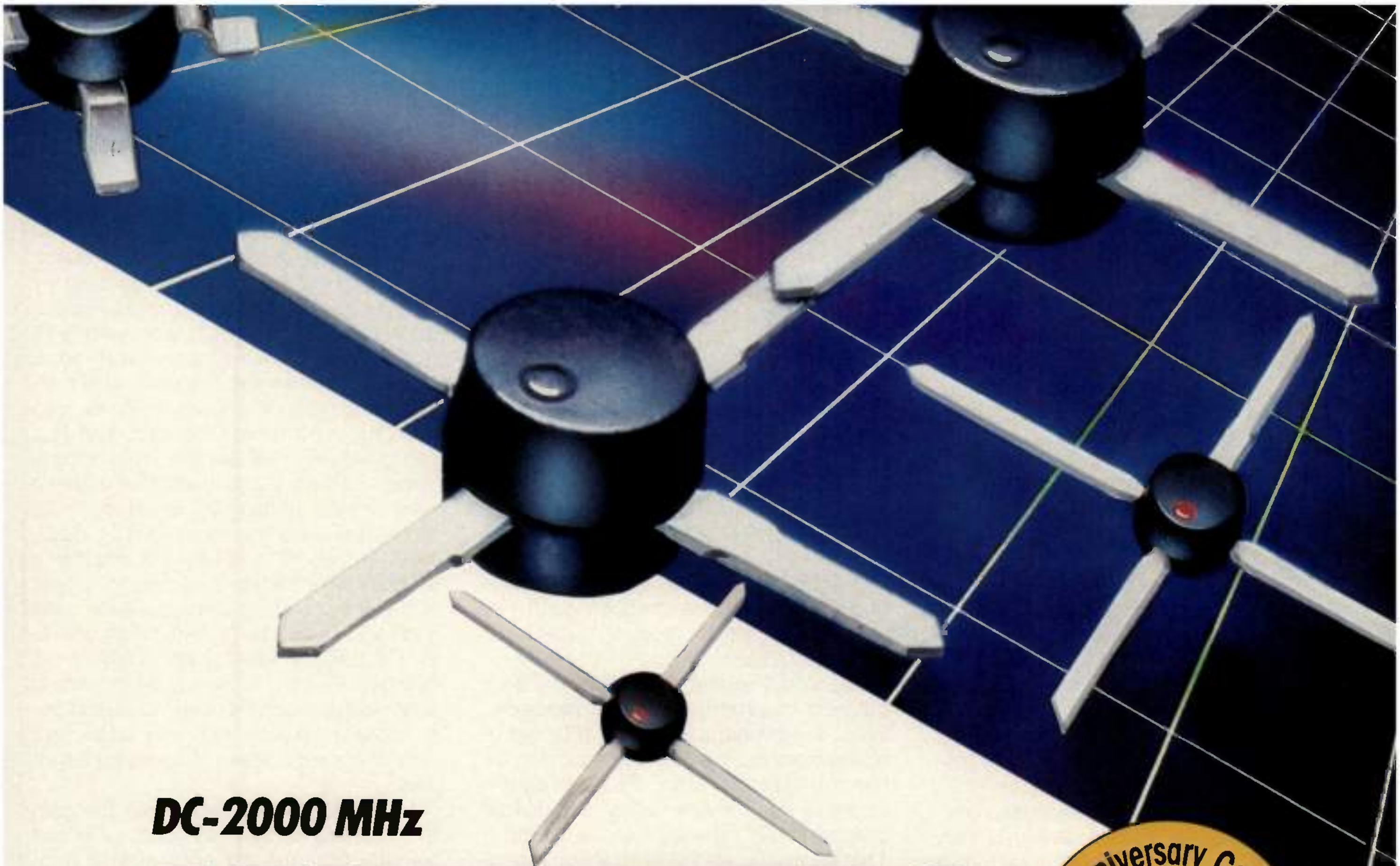


Figure 1c. Ideal inverting op amp.



DC-2000 MHz

AMPLIFIERS

In plastic and ceramic packages, for low-cost solutions to dozens of application requirements, select Mini-Circuits' flatpack or surface-mount wideband monolithic amplifiers. For example, cascade three MAR-2 monolithic amplifiers and end up with a 25dB gain, 0.3 to 2000MHz amplifier for less than $4.50. Design values and circuit board layout available on request.

It's just as easy to create an amplifier that meets other specific needs, whether it be low noise, high gain, or medium power. Select from Mini-Circuits' wide assortment of models (see Chart), sketch a simple interconnect layout, and the design is done. Each model is characterized with S parameter data included in our 740-page RF/IF Designers' Handbook.

All Mini-Circuits' amplifiers feature tight unit-to-unit repeatability, high reliability, a one-year guarantee, tape and reel packaging, off-the-shelf availability, with prices starting at 99 cents.

Mini-Circuits' monolithic amplifiers...for innovative do-it-yourself problem solvers.

Models above shown actual size

from **99¢** Unit price $ (25 qty)

PLASTIC SURFACE-MOUNT			++VAM-3 1.45		+VAM-6 1.29	++VAM-7 1.75		
add suffix SM to model no. (ex. MAR-ISM)	MAR-1 1.04	MAR-2 1.40	MAR-3 1.50	MAR-4 1.60	MAR-6 1.34	MAR-7 1.80	MAR-8 1.75	
	MAV-1 1.15	+MAV-2 1.45	+MAV-3 1.55	MAV-4 1.65				MAV-11 2.15
CERAMIC SURFACE-MOUNT	RAM-1 4.95	RAM-2 4.95	RAM-3 4.95	RAM-4 4.95	RAM-6 4.95	RAM-7 4.95	RAM-8 4.95	
PLASTIC FLAT-PACK	MAV-1 1.10	+MAV-2 1.40	+MAV-3 1.50	+MAV-4 1.60				MAV-11 2.10
	MAR-1 0.99	MAR-2 1.35	MAR-3 1.45	MAR-4 1.55	MAR-6 1.29	MAR-7 1.75	MAR-8 1.70	
Freq.MHz,DC to	1000	2000	2000	1000	2000	2000	1000	1000
Gain, dB at 100MHz	18.5	12.5	12.5	8.3	20	13.5	32.5	12.7
Output Pwr. +dBm	1.5	4.5	10.0	12.5	2.0	5.5	12.5	17.5
NF, dB	5.5	6.5	6.0	6.5	3.0	5.0	3.3	3.6

Notes: + Frequency range DC-1500MHz ++ Gain 1/2 dB less than shown

designer's amplifier kits

DAK-2: 5 of each MAR-model (35 pcs), only $59.95
DAK-2SM: 5 of each MAR-SM model (35 pcs) only $61.95
DAK-3: 3 of each MAR, MAR-SM, MAV-11, MAV-11SM (48 pcs) $74.95

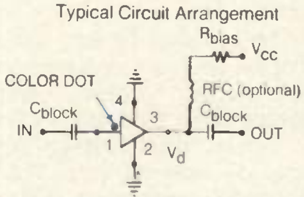
designer's chip capacitor kit

KCAP-1: 50 of 17 values, 10pf to 0.1µf (850 pc), $99.95

chip coupling capacitors at .12¢ each (50 min.)

Size (mils)	Value
80 x 50	10, 22, 47, 68, 100, 220, 470 680 pf
80 x 50	1000, 2200, 4700, 6800, 10,000 pf
120 x 60	.022, .047, .068, .1µf

Typical Circuit Arrangement



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For detailed specs on all Mini-Circuits products refer to • THOMAS REGISTER • MICROWAVE PRODUCT DATA DIRECTORY • EEM • MINI-CIRCUITS' 740- pg. HANDBOOK.

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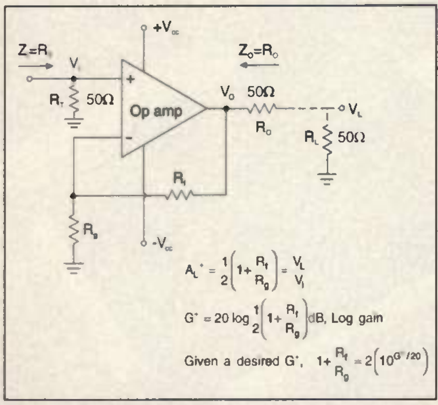

Figure 2a. Non-inverting op amp configured for RF application.

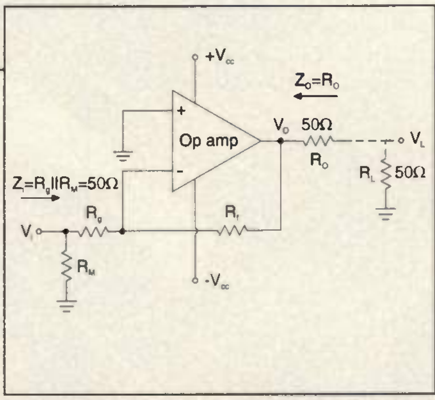

Figure 2b. Inverting op amp configured for RF application.

in RF applications, it is useful to summarize some of the standard operating assumptions for typical RF amplifiers. Although there are certainly exceptions to the typical conditions shown here, RF amplifiers generally have:

1. AC-coupled input and output. A DC voltage generally has little meaning in RF applications.

2. Input and output impedances nominally set to 50 Ω AC over the frequency range of operation. This is seldom a physical 50 Ω resistor, but rather a combination of active element I/O impedances along with passive matching networks.

3. Fixed signal gain operations over a certain band of frequencies. Any particular RF amp is purchased to provide a particular gain and is not user adjustable. A two-decade range of operating frequencies seems typical.

4. Single power supply operation. Since both input and output are AC-coupled, bipolar power supplies, balanced around ground, are not needed. The DC bias point is maintained internally with minimal user adjustment possible.

Figure 1a shows a typical RF amplifier connection, while Figures 1b and 1c show an ideal op amp, utilizing either current- or voltage-feedback, connected for non-inverting and inverting gains, respectively.

For the RF-amp, both input and output are AC-coupled, while a single power supply biases the part through R_b. L_c chokes off the AC output signal from seeing the power supply as a load. Internally, the RF-amp signal gain is specified with the output driving a 50 Ω load and is defined as 10*log (power gain).

The two ideal op amp circuits assume that the source is coming from a ground-referenced, zero-impedance voltage source while their outputs are intended to act as ideal voltage sources to a ground-referenced load. The non-inverting configuration ideally presents an infinite input impedance, a zero Ω output impedance, and a voltage gain, as shown in Figure 1b, from the plus input to the output-pin.

The ideal inverting op amp differs from the non-inverting in several respects. The output voltage is ideally 180° out of phase from the input, which accounts for the signal inversion. The op amp's inverting input ideally presents a virtual ground, while drawing minimal current. This leaves R_g as the ideal input impedance, while the voltage gain from the input of R_g to the output is simply $-R_f/R_g$. This signal inversion is usually of no consequence in an RF application, and most of this discussion will deal only with the magnitude of the inverting gain

When using op amps as RF amplifiers, the I/O impedance matching requirements must be satisfied, the voltage gain must be recast to a power gain (in dB), and the system may be configured for single power supply operation. Figures 2a and 2b show the op amps of Figures 1b and 1c set up to provide I/O impedance matching with the resulting power gain equations, but still using bipolar supplies. Bipolar power supplies allow operation to be maintained all the way down to DC. Single supply operation is also possible.

For the non-inverting case, setting Z_i=50 simply requires a 50 Ω termination resistor to ground on the non-inverting input, R_t. Getting Z_o=50 Ω simply requires a series 50 Ω resistor in the output, R_o.

For the inverting mode, the non-inverting input is ground referenced, and the amplifier input impedance becomes the parallel combination of R_g and R_m. One key difference between current and voltage-feedback op amps is that the feedback resistor determines the frequency response for the current-feedback op amp. With R_f set to obtain the desired frequency response, we cannot simultaneously satisfy both a gain and input impedance constraint with just R_g, which is why R_m is included as an additional degree of design freedom. This is possible with a voltage-feedback op amp by setting R_g to the desired input impedance (not using an R_m), then setting R_f to obtain the desired gain. With each particular current-feedback op amp calling out a particular optimum R_f, R_g can then be used to set the gain and R_m, along with R_g, will set the input impedance. Setting R_g to yield the desired gain and then setting R_m to satisfy Z_i=50 Ω will work until the required R_g < 50 Ω. Having fixed R_f to satisfy the amplifier's stability requirements, going to higher and higher inverting gains will eventually yield an R_g < 50 Ω. Non-inverting operation should be used if this limitation is reached. R_f can, however, be increased beyond the recommended value in order to allow an R_g=50 at higher gains, but only at the expense of decreasing bandwidth.

Note that for both topologies the gain to the matched load has been cut in half (–6 dB), from the earlier ideal case (Figures 1b and 1c), through the voltage divider action of R_{zero} = R_L. It is a simple, but critical conversion from any description of output voltage swing to and from a power (in dBm) defined at the load. These conversions for a purely sinusoidal signal are:

$$P_0 = 10 \log \frac{\left(\frac{V_{lpp}}{2\sqrt{2}}\right)}{50\Omega(1mW)}$$

$$= 10 \log\left(V_{lpp}\right) - 10 \log(8(50\Omega)(.001))$$

$$= 20 \log \quad V_{lpp} + 4dBm$$

Conversely, for a given PO (in dBm):

$$V_{L_{FF}} = 10^{(P_0-4)/20}$$

peak – peak voltage swing at load

$$V_{O_{FF}} = 2 * 10^{(P_0-4)/20}$$

peak – peak voltage swing at output pin

Basically, for any initial description of voltage swing given, we need to convert that into an RMS voltage – square it and divide by the load (R_L=50 Ω normally) – to get the absolute power in watts. This is then divided by 0.001 to reference that power to 1 mW and 10*log of that expression is taken to yield the power in dBm.

Every op amp has a specified maximum output voltage swing that is generally shown as a peak excursion from ground. This type of specification, for balanced bipolar power supplies, is really inferring how close the output may

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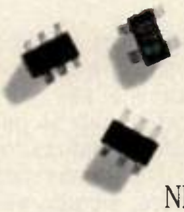

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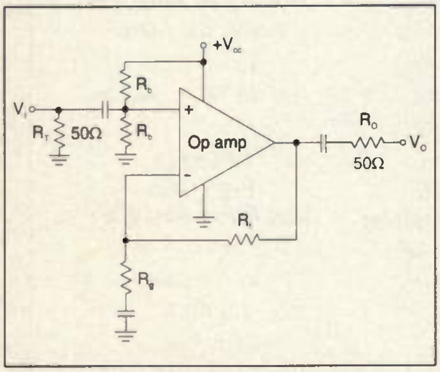

Figure 4a. Single supply, non-inverting op amp operation.

come to the supply voltages before non-linear limiting occurs. For AC-coupled RF applications, it is always best to center the output-pin's DC-level between the two supply-pins in order to provide the maximum output V_{pp}.

Most wideband op amps do not require a ground reference for proper operation and can be operated easily from a single supply. Generally, all that is required is to set the DC voltage on the non-inverting input and the output-pin centered between the voltages appearing on the two supply-pins. For a single supply operation, with one supply-pin held at ground, this translates into the non-inverting input and V_o being held at $V_{cc}/2$. For those amplifiers requiring a ground-pin, that pin should also be driven with a low source impedance voltage midway between the supply-pins.

There are many possible implementations of single-power-supply op amp operation. Figures 4a and 4b show two simple ways to operate non-inverting and inverting op amps as AC-coupled RF amplifiers using a single power supply.

In the non-inverting case, the input termination is still DC-coupled, while the non-inverting input bias is set by the two R_bs to yield $V_{cc}/2$. R_b should be large

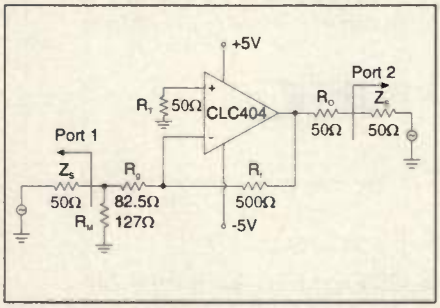

Figure 5a. Non-inverting amplifier S parameter test circuit.

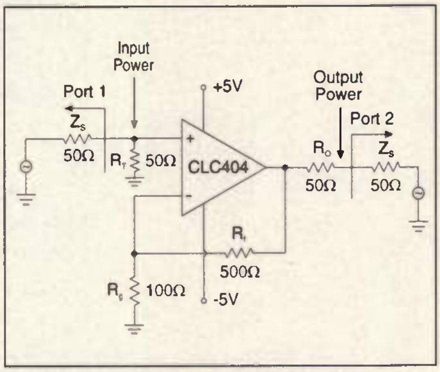

Figure 4b. Single supply, inverting op amp operation.

enough to limit excessive quiescent current in the bias leg, but not so large as to generate excessive DC errors due to the amplifier's input bias current. If this bias path impedance becomes too low, it should be included as part of the input impedance and the termination resistor adjusted accordingly. The gain setting resistor, R_g, is also AC-coupled to limit the DC gain to 1. Hence, the non-inverting input DC bias voltage also appears at the output-pin. The output should be AC-coupled in both circuits to limit the DC current that would be required if a grounded load were driven.

For the single-supply inverting amplifier of Figure 4b, we still require the mid-point reference to be brought in on the non-inverting input. A de-coupling capacitor on that node is also suggested to decrease the AC source impedance for the non-inverting input noise current. The gain for this non-inverting input reference voltage is again AC-coupled to yield a unity DC gain to get $V_{cc}/2$ at the output-pin. The inverting input impedance goes from R_m at DC to 50 Ω at higher frequencies if the parallel combination of R_m and R_g have been set to 50 Ω. R_m, as well as R_t in Figure 4a, could also be AC-coupled to avoid DC

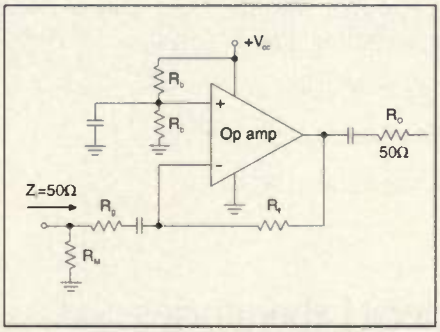

Figure 5b. Inverting amplifier S parameter test circuit input/output VSWR.

loading on the source.

For both of these single supply circuits, we have given up the DC-coupling for the signal path. The low frequency limits to operation will now be set by the AC-coupling capacitors, along with impedances in each part of the circuit. All of the subsequent discussions assume balanced bipolar supplies, but apply equally as well to single supply operation.

Small Signal AC Performance Characteristics

All of the typical small signal AC parameters specified for RF amplifiers are derived from the S parameters for the part [2]. These are:

	Scattering Parameters	RF Amplifier Specifications
S_{11}	Input reflection	Input VSWR
S_{22}	Output reflection	Output VSWR
S_{21}	Forward transmission	Amplifier gain and bandwidth
S_{12}	Reverse transmission	Reverse isolation

These frequency-dependent specifications are measured using a network analyzer and an S parameter test set. A full two-port calibration should be performed prior to any device measurements. The HP8753A, used for the measurements reported here, incorporates full 12-term error correction in its two-port calibration. This basically normalizes all measurement errors due to imperfections in the cabling and test hardware [3].

Figures 5a and 5b show the two configurations for the CLC404 used in demonstrating the small signal AC performance parameters listed above. In each case, the S parameter test set places the device into a 50 Ω input and output environment. Both configurations achieve a linear voltage gain of 6 V/V to the output-pin and 3 V/V to the 50 Ω load. This yields a gain of 20*log(3) = 9.54 dB measured by the network analyzer. Recall that one of the advantages to using op amps in RF applications is the exceptional flexibility in setting the gain. A wide range of gains could have been selected for the test circuits of Figures 5a and 5b. A gain of 6 was selected to allow easy comparisons to the CLC404's data sheet specifications, which are all defined at this gain.

For the inverting-gain configuration,

R_m and R_g set the input impedance to 50 Ω. An R_t of 50 Ω is retained on the non-inverting input to limit the possibility of self-oscillation in the non-inverting input transistors.

The Voltage Standing Wave Ratio (VSWR) is a measure of how well the input and output impedances are matched to the source impedance. (It is assumed throughout that the transmission line characteristic impedance is also equal to the source impedance of both ports—50 Ω in this case). It is desirable for the input and output impedances to be as closely matched to the source as possible for maximum power transfer and minimum reflections.

Measuring the input VSWR is simply a matter of measuring the ratio of the reflected power vs. incident power on port 1 of Figures 5a and 5b (S_{11}). A perfect match will reflect no power. Output VSWR is measured similarly at port 2 (S_{22}).

As described earlier, an op amp's input and output impedances are determined by external components selected by the designer. For this reason, I/O VSWR is never shown on an op amp's data sheet. Excellent VSWR can, nevertheless, be achieved using the components shown in Figures 2a and 2b.

An op amp's gain polarity has minimal effect on the output VSWR. At low frequencies, R_o by itself will determine the output VSWR. Setting this resistor to 50 Ω will yield excellent output VSWR to reasonably high frequencies. As the test frequency increases, however, the op amp's output impedance will begin to increase as the loop gain rolls off [4]. This inductive characteristic can be partially compensated for by a small shunt capacitance across R_o. Figure 6 shows this, for either gain polarity, along with tested VSWR with and without this shunt capacitance. The value of this capacitance will depend on the amplifier and, to some extent, on the gain setting, and was determined empirically for this test by using a small adjustable cap (5-20 pF) directly across R_o.

The marker at 200 MHz indicates an output VSWR of 1.32 when C_t is tuned optimally. Tuning C_t also extends the frequency response (S_{21}) slightly and will be left in place for the remainder of the tests.

The input impedance match of the non-inverting topology (Figure 5a) is principally set by R_t. As the frequency increases, the input capacitance of the op amp will eventually degrade the input VSWR. This effect is so negligible over the expected operating frequency range that no tuning is required.

The input impedance match of the inverting topology (Figure 5b) is, at low frequencies, set by the parallel combination of R_g and R_m. This holds very well as long as the amplifier's inverting input acts like a low impedance over frequency. For current-feedback amplifiers, the inverting input is actually a driven, low-impedance node. Its impedance will, however, increase with frequency. A voltage-feedback amplifier's apparent inverting input impedance will also increase with frequency as its loop gain rolls off. In the voltage-feedback case, the increase in inverting input impedance will be seen at a lower frequency than for a current-feedback amplifier and will depend strongly on the amplifier

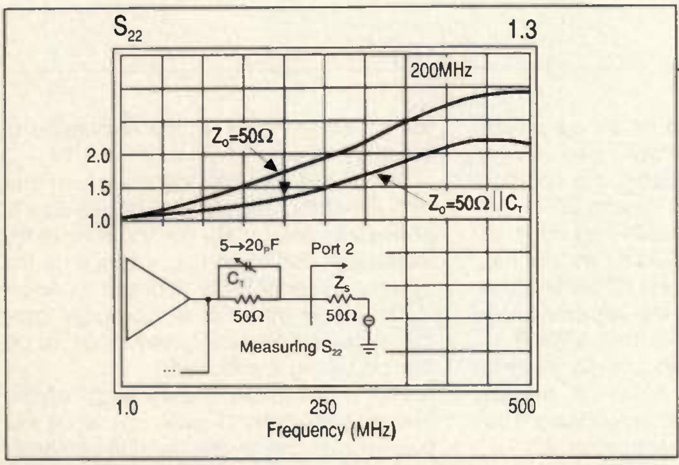

Figure 6. Measuring and tuning CLC404 output VSWR.

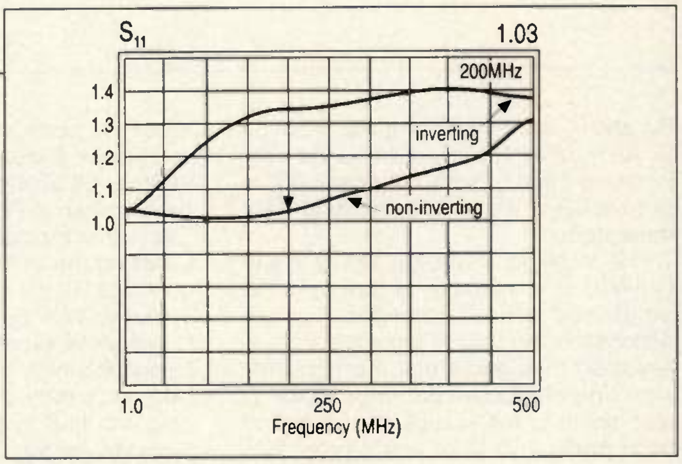

Figure 7. CLC404 input VSWR.

gain setting.

Figure 7 shows the tested input VSWR for the two gain polarities in Figures 5a and 5b. In this case, we are measuring S_{11} and allowing the HP8753A to convert the measurement and display VSWR directly.

Note carefully the change in scale for the input VSWR vs. the output VSWR plot. The marker on the non-inverting test trace shows an exceptional input VSWR of 1.03 at 200 MHz, while the inverting amplifier's VSWR, though higher, remains under 1.4 through this range.

Forward Gain and Bandwidth

Typical RF amplifier specifications show a fixed gain, as defined in Figure 1a, with a specified frequency range for 0.5 dB gain flatness, along with –3 dB cutoff frequencies. For the designer using a current-feedback op amp, a wide range of possible gains is easily obtainable. With the CLC404's specified voltage gain range of ±1 to ±10, and including the additional 6 dB loss from the output to the load, –6 dB to 14 dB gains may be achieved using the CLC404. Higher gains can be achieved with this, or any other current-feedback amplifier, with some sacrifice in bandwidth [1]. For example, the CLC401, specified over a ±7 to ±50 voltage gain range, translates into an 11 dB to 28 dB gain range for RF applications.

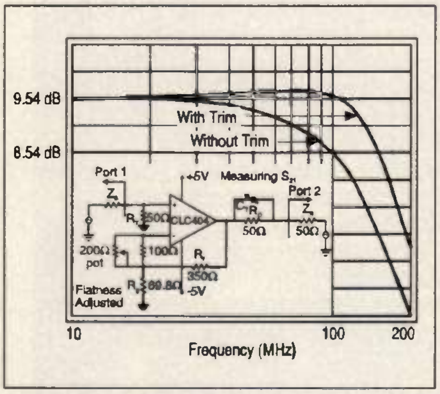

Figure 8. Measuring and adjusting the frequency response 21.

The forward gain over frequency (commonly called the frequency response and measured as S_{21}) is always shown in a current-feedback op amp data sheet—often, over a range of gains. Small signal –3 dB bandwidth and gain-flatness are also typically guaranteed at a particular gain for each amplifier. Rarely does a voltage-feedback op amp show the S_{21} characteristics, since it so strongly depends upon the gain setting. Rather, these amplifiers show an open loop gain and phase plot, and leave it to the designer to predict closed loop gain and phase. Another advantage of the excellent loop gain control of the current-feedback topology is exceptional forward gain phase linearity. This phase is also shown on the frequency response plot. A maximum deviation from linear phase is generally guaranteed at a particular gain setting in the data sheet specifications.

One additional advantage to the current-feedback topology is that a simple frequency response trim is possible by inserting a resistor inside the feedback loop to the inverting node [1]. This frequency response flatness trim has the same effect for either non-inverting or inverting topologies. Figure 8 shows this adjustment added to the circuit of Figure 5a, along with the measured S_{21} with and without this trim. As reference 1 describes, this resistive trim inside the feedback loop has the effect of adjusting the loop gain, and hence the frequency response, without adjusting the signal gain, which would still be set by only R_f and R_g. This particular test achieved a flatness of ± 0.1 dB from DC to 110 MHz at a gain of 9.54 dB for the non-inverting test circuit shown (with identical results for an inverting configuration).

Note that the values for R_f and R_g have been reduced from those used in the circuit of Figure 5, although their ratio, and hence the gain, have remained the same. With the adjustment pot set to zero Ω, this lower R_f value ensures that the frequency response will be peaked for any particular CLC404 used in the circuit. Then, by increasing the resistance into the inverting input, the amplifier can be compensated and S_{21} adjusted to the excellent flatness shown above.

The part-to-part variation in frequency response becomes more pronounced as the desired operating frequencies and signal gains increase. Operation of the CLC404 through 50 MHz at 9.54 dB gain would, for example, have minimal variation relative to operation through 100 MHz and 14 dB gain. Higher frequency operation can be achieved if the degraded flatness and distortion characteristics are acceptable to the application. Newer current-feedback op amp introductions, such as the 1 GHz CLC449, will extend frequency range of RF operation for op amps.

Reverse Isolation

This small-signal AC characteristic is a measure of how much signal injected into the output port makes it back into the input source. The magnitude of S_{12} is the measure of reverse isolation. All current-feedback op amps exhibit excellent reverse isolation relative to most RF amplifiers. This is because both the output and the inverting-input are low impedance nodes. Significant signal attenuation can be expected in taking a signal voltage applied to the output

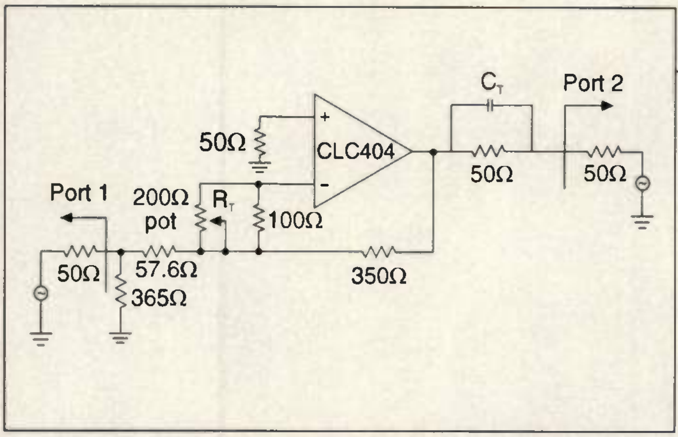

Figure 9a. Inverting reverse Isolation test circuit.

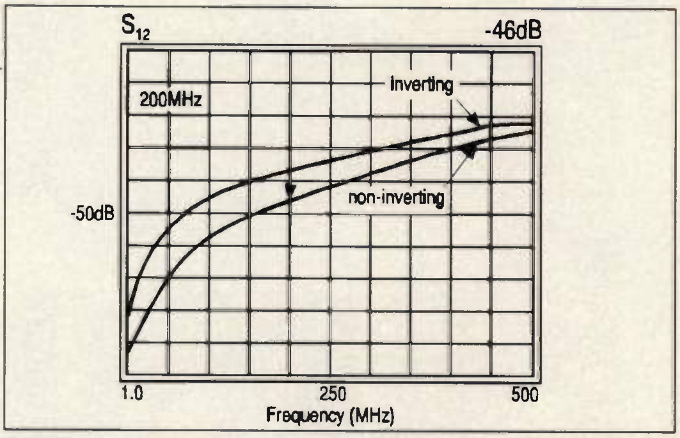

Figure 9b. Reverse gain for the circuits of 8 and 9a.

matching resistor and tracing it back to either an inverting or non-inverting input signal. Slightly more attenuation can be expected for the non-inverting vs. inverting configurations, since the signal must also get from the inverting to non-inverting pin in the non-inverting case.

The circuit of Figure 8, along with the inverting circuit of Figure 9a, were used to measure the reverse isolation for both gain polarities, as shown in Figure 9b. Although reverse isolation is generally specified as a positive number, this is simply the negative of the log gain in going backwards through the amplifier. Hence, the plot of Figure 9a shows a rising "gain" that would be interpreted as a decreasing reverse isolation as we go to higher frequencies. As Figure 9b shows, isolations in excess of 30 dB are easily obtainable through frequencies far higher than the S_{21} operating frequency range, with very high isolations observed at low frequencies.

Dynamic Range Limiting Characteristics

The final area of concern in applying op amps to RF applications are the limits to dynamic range familiar to RF amplifier users. These are generally limited to –1 dB compression point; two-tone, 3rd order, intermodulation intercept; and noise figure.

The –1 dB compression point is a measure of the maximum output power capability of the amplifier. At low frequencies, this is set by the available voltage swing out of the op amp, while at higher frequencies it is limited by the slew rate for the op amp.

The two-tone, 3rd order, intermodulation intercept allows the prediction of spurious signals caused by amplifier non-linearities arising from two input signals closely spaced in frequency. The class AB output stage used in most current-feedback op amps offers considerably higher intercepts for a given level of quiescent power than for the class A outputs used more commonly in RF amplifiers. Although it is often assumed that the intercept is 10 dBm higher than the –1 dB compression point, this is not true for op amps. Both are generally much more frequency dependent than for RF amplifiers, with the intercept typically much higher than would be expected from the –1 dB compression point. Typical numbers for the CLC404 would be an 18 dBm –1 dB compression point and a 45 dBm intercept for frequencies below 10 MHz.

The noise figure is a measure of how much noise is added by the amplifier and will set a limit to the minimum detectable input signal. The noise figure for any op amp can be calculated from the three input referred noise terms for the op amp itself and the noise contributions of the external resistors setting the gain and I/O impedances. Using a physical, noisy, input impedance matching resistor typically increases the noise figure by 6 dB from what it would be without this matching. Op amp noise figure is also gain dependent for any particular device and will decrease as the operating gain is increased. Typical noise figures for the CLC404 range from 12 dB to 16 dB if an input impedance match is maintained. For a complete development of op amp noise figure and further discussion of –1 dB compression and intercepts, refer to reference 5.

Conclusions

High speed current-feedback amplifiers can offer considerable performance advantages when used in IF and RF applications. The flexible gain and I/O impedance capability can be used to the designer's benefit in tailoring the amplifier to the specific requirement. Last minute gain changes or non-standard I/O impedances can be easily handled with simple resistor changes. Exceptional I/O VSWR and reverse isolation come automatically with op amp type devices. Current-feedback op amps that offer bandwidths in excess of 1 GHz are now becoming available. These parts offer reasonable IP3's through about 150 MHz at much lower quiescent power than comparable fixed gain RF amplifiers. And finally, parts from multiple suppliers are bringing the prices down to below $2.00 in volume for these extremely capable op amps. ***RF***

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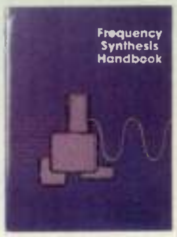

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INFO/CARD 46

Orlando Show Features Timely Presentations, New RF Products

RF Expo East Program

Featuring papers on essential RF topics

Tuesday, November 15, 1994
8:45-10:00 a.m.

8:45	Welcome
9:00	RF Design Awards Grand Prize Presentation
9:15	Opening Address
10:00	Exhibit Hall Grand Opening

Tuesday, November 15, 1994
1:30-4:30 p.m.

High Power RF
Chairman: Adrian Cogan, MicroWave Technology

1:30 — A Low Cost, High Efficiency 13.56 MHz Power Amplifier
Ken Dierberger, Advanced Power Technology
Frederick Raab, Green Mountain Radio Research
Lee Max, Consultant
The design, development, assembly, and performance of a low cost, high efficiency 250 W, 13.56 MHz RF power amplifier are detailed in this paper. Both the theoretical design and physical construction of the input switching network, the output matching circuit, and the DC feed network of the amplifier will be addressed. The paper will also contain a technical description of the RF power transistors including Mean Time To Failure data.

2:30 — Class-S High-Efficiency Amplitude Modulator
Frederick H. Raab, Daniel J. Rupp, Green Mountain Radio Research
This paper describes a 100-W class-S modulator with an envelope bandwidth of 57 kHz and an efficiency of about ninety percent.

3:30 — High Efficiency SSB HF/VHF Transmitter Based on Envelope Elimination and Restoration
Frederick H. Raab, Daniel J. Rupp, Green Mountain Radio Research
An experimental high-efficiency HF/VHF multimode transmitter based upon the Kahn Envelope-Elimination-and-Restoration (EER) technique has been described. The amplifier can produce a wide variety of signals including SSB, AM, and FM, with approximately 60% efficiency for all signal amplitudes, about three times that of a transmitter with a conventional class-B PA.

Antennas for Wireless Applications
Chairman: Andy Kellett, RF Design

1:30-3:30 — Engineering the Characteristics of Small Antennas to Suit the Application (3-hour tutorial session)
Dr. Ian Dilworth, University of Essex
This presentation explores the system requirements, physical and electrical performance requirements and limitations of antennas for portable applications. Various configurations are described, along with the tradeoffs associated with their implementation.

Manufacturing Technology
Chairman: Madjid Belkerdid, University of Central Florida

1:30 — Development of a New Line of Low Cost Microwave Components Using Print and Fire Technologies
Loren E. Ralph, R.F. Prime
A new line of low-cost, surface-mountable microwave components has been developed using a printed technology. The components cover 500 MHz to 6+ GHz, and include mixers, 0° and 90° splitters, 180° splitters, and filters that are packaged in IC-like packages. This new "Blue Cell" technology is compatible with active devices, allowing higher levels of integration to be developed.

2:30 — Implementation of Design for Manufacture in a Semiconductor Fabrication Environment
Adam Williams, Ricardo Borges, John Iwanicki, M/A-COM
This paper describes the ideas behind DFM and its implementation in a semiconductor fabrication facility. The methodology closely follows Motorola's six steps to six sigma program. Examples from the challenges faced by a cross-functional team created to oversee the application of DFM concepts to a silicon hyperabrupt varactor production line are given to demonstrate the transition from theory to practice.

Wednesday, November 16, 1994
8:30-11:30 a.m.

Power Amplifier Tutorial

8:30-11:30 — Classes of RF Power Amplifiers A Through S, How They Operate and When to Use Them (3-hour tutorial session)
Nathan O. Sokal, Design Automation
The electrical characteristics that define classes of power amplifiers are presented in this extended tutorial. Linearity, efficiency, complexity and spectral performance of each type are described. Also featured are applications appropriate for each amplifier type.

Oscillator Design and Analysis
Chairman: Claude A. Sharpe, Texas Instruments

8:30 — Analysis and Optimization of Oscillators for Low Phase Noise and Power Consumption
Dr. Ulrich L. Rohde, Compact Software
Compact Software has solved the complex mathematical problem of analyzing SSB in arbitrary topologies for oscillators, but also solved the ability to optimize such circuits for phase noise, power consumption, and other relevant parameters. After explaining the general concept, examples before and after optimization will be shown.

9:30 — Cryogenic Colpitts Oscillators for High Sensitivity Detection of NMR
Neil S. Sullivan, University of Florida
The author reports on the use of a Colpitts oscillator operating in the 50-100 MHz range at 1-4K as a device for highly sensitive detection of nuclear magnetic resonance (NMR) absorption. The circuit used a GaAs HEMT transistor as the active element and was successfully operated in high magnetic fields up to 4T. The oscillator could be operated at ultralow power dissipation, permitting ultrasensitive low noise detection for samples at low temperatures.

10:30 — Development and Advancements in SC-Cut Crystals
Jim Griffith, Piezo Crystal Company
This discussion concerning the technical aspects of SC-Cut precision crystals includes SSB phase noise, short term stability (Allan Variance), vibrational sensitivity, thermal hystersis, long-term aging, and advancements in miniaturization of SC-Cut crystal technology.

Receiving Systems
Chairman: H. Clark Bell, HF Plus

8:30 — Efficient Satellite Receivers Using TCH Codes
Francisco Cercas, Instituto Superior Tecnico (IST)
This paper describes an efficient satellite receiver using a new type of code specially designed for this purpose. The code properties, using low rate coding, give the system all the advantages of spread spectrum systems with a low degree of complexity and low cost, which makes it suitable for use in VSAT systems or any other kind of mobile communications where a low BER must be achieved in the presence of very weak signals.

9:30 — Wide Dynamic Range HF Receiver Amplifier Uses SST FETs
A.I. Cogan, K. Sooknanan, MicroWave Technology
E. Ressler, NAWC
L.B. Max, Consultant
The design, development, construction, operation and performance of a high dynamic range compact HF amplifier that uses SST FETs to achieve superior linear performance levels is described. The straightforward design presented is an alternative to the use of complex feedforward or feedback circuits which yield larger, less efficient and more expensive amplifiers.

10:30 — CW Rejection in ESM Receivers
Sherman Vincent, Raytheon
This paper presents an automatic CW rejection capability which uses a tunable filter, a detector log video amplifier, and a microprocessor-based controller. Rejection lever greater than 50 dB and set-on times of 10 to 35 msec can be achieved.

Wednesday, November 16, 1994
1:30-4:30 p.m.

Wireless Applications
Chairman: Ulrich Rohde, Compact Software

1:30 — Automatic Vehicle Identification
Claude A. Sharpe, Texas Instruments
This paper describes an Automatic Vehicle Identification (AVI) system under production at Texas Instruments which is compliant with the Title 21 CALTRAN specifications, and describes the circuitry incorporated to deal with multipath, lane discrimination, toll tag sleep and wakeup modes, and a summary of the information obtained from two years of field testing.

2:30 — Wide Area RF Data Systems
Donald J. Marsh, II Morrow
This paper reviews alternative technologies for wide-area RF data transmission with focus on RF systems, demonstrates a case history of an OPS application, and discusses the future of wireless applications, including packet data services and personal communication systems.

3:30 — Millimetre Wave Systems and Propagation
Dr. Ian Dilworth, University of Essex
This paper presents preliminary measured data on millimeter wave aspects such as foliage attenuation and building scatter. The measurements are made using co and cross polarized antenna, and results and method will also be described. Propagation characteristics at millimetre wavelengths (38 GHz) are generally not available in the open literature, yet this information is crucial to successful systems implementation.

Computer-Aided Design
Chairman: Sherman Vincent, Raytheon

1:30 — The "Thin Wire" Program for Moment-Method Antenna Analysis
R. P. Haviland, MiniLab Instruments
This paper describes the capability and limitations of the "Thin Wire" antenna analysis program developed by Richmond at Ohio State for NASA. It shows the basic assumptions of the program, and its structure. A supplementary pro-

gram to prepare data input is described, and a sample problem, with problem parameters, data input, and results output is included.

2:30 — Equivalent Circuit Modeling of Silicon Bipolar Junction Transistors for Nonlinear CAD
Ricardo Borges, M/A-COM
A new parameter extraction and optimization strategy for accurate BJT modeling is presented that emphasizes the bias-dependent base resistance and high frequency parameters. Model verification is done via comparison of the linearized model to measure scattering parameters, power compression, and load-pull measurements. The use of two-dimensional process and device numerical simulations to estimate difficult-to-measure parameters is also illustrated.

3:30 — Linear Phase Filter Synthesis: CAD for Extra Precision Implementation
Alberto Milano, Shlomo Barash, Itzhak Refaeli, ELTA
Linear phase filter tuning is the most consistent difficulty with these filters, and for this reason a CAD synthesis enables an accurate definition of the mechanical dimensions. CAD is able to design both comb line and interdigitate linear phase filters, but so far only the interdigitate filters have been constructed and tested. The tuning has been easy and effective, and the experimental results are in excellent agreement with the theory.

Circuit Design Topics
Chairman: Sigmund W. Mosko, Martin Marietta

1:30 — An Analysis and Implementation of Square Root Raised Cosine Filters for use with PSK Modulated Data Signals
Bruce H. Williams, Roy E. Greef, UNISYS Government Systems Group
This paper discusses the method of using square root raised filters to approach a Nyquist minimum bandwidth filter in satellite communication systems. Pulse shaping filters, such as the square root raised cosine, improve bandwidth efficiency without degrading a system's power efficiency. A practical implementation of these filters will also be discussed.

2:30 — Low Noise Design of a SP8858 Digital PLL Synthesizer
Philip Knights, GEC Plessey Semiconductors
This paper explains both the detailed noise analysis of a digital PLL synthesizer and the design of a suitable loop filter. The main design focus is on minimizing the phase noise power in the spectrum of the output signal given the often conflicting requirements of high resolution, fast settling time, and low-level spurious sidebands. Models have been developed that provide the system/RF engineer with a quick and reliable method of assessing and optimizing the performance of a digital PLL synthesizer.

3:30 — PLL Model Validation
Mike Black, Texas Instruments
This paper describes several unique methods of PLL loop bandwidth measurement. Using these measurements, a more accurate model can be developed. This allows for better statistical optimization and better time and frequency domain simulation.

Thursday, November 17, 1994
8:30-11:30 a.m.

Digital RF Memory
Chairman: Dr. Y.S.N. Murthy, Defence Electronics Research Lab

8:30-11:30 — Application of Digital RF Memory Systems
Dr. Y.S.N. Murthy, Defence Electronics Research Lab.
This presentation outlines the design features and software needed for realization of Digital RF Memory systems (DRFM) for applications including generation of arbitrary waveforms and low probability intercept radar waveforms, local oscillator for frequency agile radar, stable local oscillator for MTI and moving target radar systems, programmable RF emission simulators for evaluation of ESM receivers, estimation of EMI/EMC, and training of radar/EW operators. The paper also outlines other universal applications of DRFM systems.

Low Noise Techniques
Chairman: Joseph Hill, Hill Engineering

8:30 — Ultra Low Noise Signature Characterization for Better Systems Design
Perry Bates, Technitrol Cyclonetics
This paper describes the process of determining the noise signature of components, and presents the techniques to apply that information for design. Covered are: signal dynamics, source noise, active and passive circuit blocks, interconnecting blocks and results.

9:30 — 900 MHz Monolithic Low Noise Amplifier and Mixer
S.-G. Lee, R.D. Schultz, T.D. Brogan, Harris Semiconductor
A 900 MHz LNA/Mixer IC is presented, featuring silicon dielectrically isolated spiral inductor, gain control resistor, LO cross-coupling, and temperature-stable low noise bias. DC-based production RF screening methods are also discussed.

Spread Spectrum
Chairman: Donald J. Marsh, II Morrow

8:30 — A Novel Synchronization Scheme for Spread Spectrum Radio Communications
A.Z. Tirkel, G.A. Rankin, Coded Communications Corp.
L.C. Jenkins, Les Jenkins Assoc.
C.F. Osborne, Monash University
This paper introduces a novel synchronization scheme using minimum hardware and PC based software for spread spectrum Radio communications, which has been successfully implemented in a 900 MHz direct sequence spread spectrum voice radio, operating in an adverse environment.

9:30 — GPS C/A Code Coarse Acquisition Receiver Structure: Design, Simulation and Analysis
J.S. Seybold, G.V. Fountain, The Analytic Sciences Corporation
M.A. Belkerdid, University of Central Florida
This paper describes a coarse acquisition receiver structure suitable for use in acquiring the Global Positioning System Coarse Acquisition code, its simulation model, and an in-depth analysis of the coarse acquisition receiver operation. The required analysis for determining the optimal detection threshold setting is included, as well as additional analysis showing how the simulation run time was reduced by using a reduced integration time and a correspondingly increased input signal to noise ratio.

Blind Mate Connectors
Gilbert announces its new float type blind mate connectors. The jacketed microporous semi-rigid .141" cable.
Gilbert Engineering
J-Squared Marketing, Inc.
Booth 713, 715
INFO/CARD #138

DC - 1 GHz Power Amp
Amplifier Research will introduce the model 40WD1000 broadband RF power amplifier that provides swept capability of DC-1 GHz when controlled through remote connector. It provides a gain of 47 dB minimum, 2 bands, DC-1 MHz and 1 MHz-1000 MHz selectable either local or remote.
Amplifier Research
Booth 401 and 403
INFO/CARD #139

Temperature Compensating Capacitors
Republic Electronics will display its Surface Mount Multilayer Capacitors with negative temperature compensation (NPO through N5600) for the wireless industry.
Republic Electronics
Booth 712
INFO/CARD #140

Connectors
Harting Elektronik, Inc. will have on display their line of connectors including the DIN 41612 for VME and Multibus, Compliant Din, Flat Ribbon Cable, P1396 Telecombus, Industrial power, High density, Metric, SCSI, I/O, And Fiber optic connectors with ISO 9001 Certification.
Harting Elektronik, Inc.
Booth 101
INFO/CARD #141

Vector Network Analyzer
Anritsu Wiltron will exhibit the new price/performance solution for manufacturing and R & D. The 37211A Vector Network Analyzer provides fully reversing S parameter measurements from 22.5 MHz to 3 GHz.
Anritsu Wiltron
Booth 315
INFO/CARD #142

Crystal Filter
Piezo Technology, Inc. will introduce its 70.0 MHz fundamental crystal filter (Model 7823C) which features a 3 dB bandwidth of ± 43 kHz minimum and a stopband attenuation of 45 dB at ± 140 kHz maximum with dimensions of 1.95 x 0.59 x 0.40 inch (LxWxH).
Piezo Technology, Inc.
Booth 214
INFO/CARD #144

Circuit Analysis
HP-EEsof will display its newest release of Libra and Touchstone for Windows. The new releases offer enhanced features for nonlinear/linear circuit design on PCs running under windows.
HP-EEsof
Booth 508, 510, 512
INFO/CARD #146

2.7 GHz Communications Modules
The Motorola Logic Integrated Circuits Division is introducing a set of integrated synthesizer and specialized satellite receiver devices for phase-locked loop communications applications up to 2.7 GHz operation.
Motorola
Logic Integrated Circuits Div.
Booth 601
INFO/CARD #143

Power Transistor
Ericcson Components introduces its new 21 watt 1800-2000 MHz RF Power Transistor for cellular radio. Part TTB 20071 is part of a rugged lineup of products from Ericcson, a world leader in telecommunications.
Ericsson Components
Booth 107
INFO/CARD #150

3 V ICs
For driver applications, CEL offers the UPC2762T amplifier. It delivers 3 dB bandwidth to 2.9 GHz – with 13 dB gain, a 6.5 dB NF, and an 8 dBm P1dB at 900 MHz. For gain, the UPC2763T delivers 20 dB at 900 MHz, with a 5.5 dB NF and a 9.5 dBm P1dB. The new UPC2749T LNA delivers 16 dB gain, a 4 dB NF, with a low 6 mA ICC. CEL will also have a variety of low power down converters and a new transistor array.
California Eastern Laboratories
Booth 302
INFO/CARD #145

Oscillators
Vectron Laboratories, Inc. has introduced a number of new products which include a 622.08 MHz single DIP VXCO, ±20 ppm -40/+85°C temperature stability, ± 100 ppm deviation, ECL output; the 622.08 MHz SONET CDR; Hybrid VCOs, SINE or ECL to 350 MHz in a DIP package and ECL to 1 GHz in a flatpack configuration.
Vectron Laboratories, Inc.
Booth 606
INFO/CARD #147

Circuit Prototype Machines
T-Tech, Inc. will display hardware and software solutions for prototyping circuit boards using windows with its QC7000 system. It will also exhibit automatic fluid dispensing equipment for a wide variety of demanding applications.
T-Tech, Inc.
Booth 701
INFO/CARD #148

1.5 GHz Synthesizer IC
GEC Plessey Semiconductors, Inc. will exhibit the SP8858 -1.5 GHz PLL single chip synthesizer including a dual modulus prescaler, programmable A, M, and R dividers, Digital phase detector, charge pump, and look detect circuits. Low phase noise floor is typically -150 dBc/Hz, with high input sensitivity. Available in 28 pin DIL or leaded chip carrier. Full Mil-STD 883 screening.
GEC Plessey Semiconductors, Inc.
Booth 211
INFO/CARD #149

Trimmer Capacitors
Sprague-Goodman Electronics, Inc. will exhibit the GKM series SURFTRIM® SMD (carrier/reel pack) ceramic dielectric trimmer capacitors that measure only 3.8 x 3.2 x 1.5 mm, with capacitance ranges of 3.5-10.0 pF and 5.0-20.0 pF.
Sprague-Goodman Electronics, Inc.
Booth 102
INFO/CARD #151

Noise Measurement
Techtrol Cyclonetics, Inc. will debut the new TCI 5502A Noise Measurement System which provides -195 dBc noise measurement capability for AM and PM with NIST traceability.
Techtrol Cyclonetics, Inc.
Booth 114
INFO/CARD #152

I&Q Modulators
Synergy Microwave Corporation has introduced a new line of I & Q modulators specifically designed for cellular applications, including PCN. These devices are offered in both surface mount and plug-in and offer superior carrier and unwanted sideband rejection of 35 dB typical. Also introduced is a new line of coaxial resonator based, voltage controlled oscillators which offer a typical phase noise of 120 dBc/Hz at 10 kHz offset from the carrier. A tuning bandwidth of 20 MHz is achieved with a tuning voltage of 1-8 volts DC.
Synergy Microwave Corp.
Booth 406
INFO/CARD #153

Surface Mount Devices
RF Monolithics, Inc. will introduce a number of new surface mount technology (SMT) products at RFEE, including additions to its line of SMT SAW resonators, SMT SAW fil-

Find the Products You Need From These Companies at RF Expo East

ters for SONET clock-recovery applications, and a new line of SMT high-frequency digital clocks.
RF Monolithics, Inc.
Booth 612 and 614
INFO/CARD #154

Spread Spectrum Chipset

The MRFIC2401, MRFIC2403 and MRFIC2404 GaAs RFIC chip set is the Motorola Communications Semiconductor Products Division's latest introduction of an integrated solution for the wireless data market. Designed for the 2.4-2.5 GHz ISM band, these devices may be used in spread spectrum wireless data systems such as frequency hopping or direct sequence.
Motorola – Communications Semiconductor Products Div.
Booth 502
INFO/CARD #155

System 32 Software

Eagleware is demonstrating System 32, an integrated family of synthesis and simulation software for IBM and compatible PCs. System 32 is available form DOS, Windows and Windows NT. The family includes the high-speed simulator =SuperStar= Professional with =SCHEMAX= for schematic entry, and the synthesis programs =FILTER=, =OSCILLATOR=, =M/FILTER=, =MATCH=, and the newest addition, =A/FILTER=.
Eagleware Corporation
Booth 703
INFO/CARD #165

Combline Cavity Filters

TRILITHIC offers a full line of custom combline cavity filters. The filters are a low ripple, CHEBYSHEV design with a frequency range of 400 MHz-20 GHz. Two to twelve section may be specified with percentage bandwidths from 1-50%. Packaging can be as small as 1" x 0.50" x 0.50" in., excluding the typical SMA type connectors. Typical VSWR for the 50 Ω filters is specified as 1.5:1 with 1.2:1 available. Power handling is up to 300 watts CW.
TRILITHIC
Booth 719
INFO/CARD #170

Power Amplifiers

Microwave Power Devices will display several new power amplifiers for cellular, PCS and other uses. Among these amplifiers are a 25 W feed-forward amplifier, a 20 to 60 W class C amplifier, 25 W class AB amplifier, 63 W class A amplifier for 1.88 to 2 GHz, and several other amplifiers.
Microwave Power Devices, Inc.
Booth 609
INFO/CARD #156

Crystal Products

Hy-Q International (USA) manufactures the highest quality precision crystals, clock oscillators, TCXOs, VCXOs and TCVCXOs; and monolithic crystal filters. They meet your exact specifications, all in the most timely delivery schedule possible. Our products withstand the test of time, in an industry where accuracy is so important.
Hy-Q International (USA)
Booth 711
INFO/CARD #158

Thin Film Products

American Technical Ceramics thin film products and services bring a new standard of responsiveness and quality to thin film technology, offering custom metalization and patterned substrates for a broad range of hybrid circuit requirements.
American Technical Ceramics
Booth 616
INFO/CARD #159

Class AB Power Transistor

SGS-Thomson has introduced a new 150 watt, Class AB linear power transistor for 800-960 MHz digital cellular base stations. The SD4590 is available from stock from Richardson Electronics, Ltd., SGS-Thomson's largest distributor. Features include minimum gain of 8 dB at 900 MHz, maximum intermodulation distortion -28 dBc at 150 W PEP, power out 150 watts PEP, common emitter, gold metallization, high linearity.
Richardson Electronics, Ltd.
Booth 407
INFO/CARD #169

Sealed Trim Cap

Voltronics Corporation offers a 40 pF high voltage, 40 p.s.i. sealed, PTFE dielectric multi-turn trimmer capacitor. The patented design has 1000 working volts, Q over 1500 at 100 MHz, and SRF of 500 MHz at 40 pF.
Voltronics Corp.
Booth 710
INFO/CARD #160

VCXOs

Raltron Electronics Corp will exhibit VC-7600 series high speed VCXOs for use over high speed fiber optic networks such as SONET in North America and SOH networks in Europe and Asia. The frequency range is from 40.0 to 155.52 MHz.
Raltron Electronics Corp.
Booth 205
INFO/CARD #161

EM Simulator

Compact Software, Inc. will display its Microwave Explorer Electromagnetic simulator version 3.0, which models both open and packaged environments. New antenna analysis capabilities accurately compute radiation and surface wave losses.
Compact Software, Inc.
Booth 402, 404
INFO/CARD #163

Ceramic Components for GPS Receivers

A selection of bandpass filters with bandwidths of 10 to 45 MHz and patch antennas with dielectric constants of 9.5 to 20 are available at GPS frequencies utilizing ceramics for ruggedness, circuit miniaturization and size reduction.
Trans-Tech
Booth 505
INFO/CARD #162

Synthesizers and Oscillators

TRAK offers synthesizers and other oscillator-related products, specializing in multifunction assemblies. Components include oscillators — XCO, OCXO, TCXO, VCXO, VCO, DSO, DTO and phase-locked designs, frequency multipliers, comb generators, IF amplifiers, isolators and circulators.
TRAK Microwave
Booth 620
INFO/CARD #164

High Frequency Fundamental Crystals

Reeves-Hoffman will introduce its high frequency fundamental quartz crystals at 155.52 MHz in either a resistance weld or cold weld HC-45 package for oscillators and filters.
Reeves-Hoffman
Booth 317
INFO/CARD #166

75 Ohm Attenuator

JFW Industries, Inc. will introduce a DC 3 GHz, 75 Ω fixed attenuator with values of 3-20 dB available. The unit features type N connectors with a VSWR of 1.12:1 max, DC-1.3 GHz, 1.3:3 max from 1.3 to 3.0 GHz.
JFW Industries, Inc.
Booth 818
INFO/CARD #167

Planar Devices

RF Prime introduces its new surface mount Blue Cell™ planar technology for mixers, power splitters, couplers, filters, and modulators/demodulators at frequencies up to 5.6 GHz.
RF Prime
Booth 813, 814, 815
INFO/CARD #168

Components, Software, Instruments, Manufacturing and Materials

RF expo east show guide

Exhibitor Listing	Booth #
American Technical Ceramics Corp	616
Amphenol RF/Microwave Operations	308
Amplifier Research	401
Anadigics, Inc.	216
Anritsu Wiltron Sales Co.	315
Applied Microwave and Wireless	624
Brennan Associates	321
California Eastern Laboratories	302
Compac Development Corp.	707
Compact Software, Inc.	402
Compex Corporation	104
Component Distributors	110
D-Associates Inc.	116
D.C. to Light	215
Eagleware	703
EDSI	224
EF Johnson	723
Ericcson Components	107
Fotofabrication Corp.	209
GEC Plessey SEmiconductors, Inc.	211
Giga-tronics Inc.	702
Harris Semiconductor	201
Hewlett-Packard Company	608
Hewlett-Packard Co.	508
HY-Q International (USA) Inc.	711
Integrated Microwave Corporation	705
Jan Crystals	310
JFW Industries Inc.	117
J Squared Marketing Inc.	713
M/A-COM	501
Marconi Instruments	311
MCLI Microwave Communications Laboratories	304
Merrimac Industries, Inc.	411
Micrometals Inc.	202
Microwave Components Inc.	105
Microwave Journal	112
Microwave Power Devices, Inc.	607
Microwave Product Digest	103
Morrow Technologies	716
Motorola Semiconductor Products	502
National Communications Electronics, Inc.	213
Pacific Coast Technologies	609
Philips Components	301
Piezo Technology Inc.	214
Raltron Electronics	205
Raytheon Company	204
Reeves-Hoffman	317
Republic Electronics Corp.	712
RF Associates, Inc.	101
RF Circuits, Inc.	119
RF Design	509
RF Monolithics, Inc.	612
Richardson Electronics, Ltd.	218
Sawtek, Inc.	222
SGS-Thomson Microelectronics, Inc.	407
Sokol Crystal Products, Inc.	611
Solitron/Vector	721
Sonnet Software, Inc.	706
Sprague-Goodman Electronics, Inc.	102
Statek Corporation	106
Synergy Microwave Corporation	406
Techtrol Cyclonetics	114
Tektronix Inc.	602
TRAK Microwave Corp.	620
Trans-Tech Inc. (A subsidiary of Alpha Industries)	505
TRILITHIC, Inc.	719
Trilogic Sunforce	312
T-Tech, Inc.	701
Vanguard Electronics Company	208
Vectron Laboratories, Inc.	606
Voltronics Corporation	710
Wireless Design and Development	714
W.L. Gore & Associates	212

Richardson Electronics

Richardson Electronics is a leading international supplier of electron tubes, power semiconductors, and related components. The company's Solid State and Components Strategic Business Unit, which markets the SGS-Thomson product lines, focuses on offering design in support for OEMs, leading technical assistance to design engineers and providing RF, microwave, and power semiconductors and related components from stock.

T-Tech

Quick Circuit is a milling, drilling, and contour routing system for making prototype single or double sided circuit boards. The system removes copper by milling instead of chemical etching. Quick Circuit can take files (HPGL, Gerber and Excellon/NC drill) from any CAD package. The system allows for a minimum engraving width of .005". Quick Circuit features an adjustable feed rate for working with a variety of materials.

Reeves-Hoffman

You may know that Reeves-Hoffman is a manufacturer of crystal (1 kHz to 250 MHz), oscillators and hermetic seal packages. Reeves-hoffman continues to expand its capabilities in a rapidly changing market. Do you know that we manufacture high frequency fundamental crystals up to 120 MHz? Are you familiar with Reeves-Hoffman's model 322 DIP VCXO? Do you know that Reeves-Hoffman manufactures custom packages as well as crystal bases and glass-to-metal seal hybrid packages? Stop by the Reeves-Hoffman booth to learn more about our products and capabilities. See you at the show.

Merrimac Industries

Merrimac will be showing the latest in power dividers, quadrature hybrids, hybrid junctions, phase shifters, attenuators, directional couplers extending to 65 GHz, and a wide variety of mixers and I/Q products which now extend up to 18 GHz. Of special note are the new Casefree devices which are miniaturized lumped element components designed for inclusion with MMIC circuits. Design engineers will be present to discuss how to optimize circuit performance.

JFW Industries

JFW Industries is a leader in attenuation and RF switching technologies. Celebrating 15 years of providing the best customer service in the industry, JFW continues to introduce innovative RF components and switching matrices. The JFW catalog features components from each product type. This is an excellent introduction to the over 7000 different products available from JFW. The programmable attenuator line has over 1000 models and continues to expand as JFW responds to customer requirements. For the whole story on RF switches and attenuators visit JFW Industries at booth 117.

Hy-Q International (USA)

Hy-Q International (USA) manufactures the highest quality precision crystals; clock oscillators, TXCOs, VXCOs and TCVCXOs, and monolithic crystal filters. They meet your exact specifications, all in the most timely delivery schedule possible. Our products withstand the test of time, in an industry where accuracy is so important.

Piezo Technology

Piezo Technology, Inc. (PTI) is a leading manufacturer of high performance frequency control products including standard and custom quartz crystal oscillators-TCXOs, OCXOs, VCXOs and clock oscillators-monolithic and discrete crystal filters, LC filters, precision crystal resonators, and RF subassemblies. Among PTI's specialties are VHF crystal filters, UHF and microwave LC filters, SC-cut OCXOs and resonators, and high shock resonators, filters and oscillators. Recognized as a technology leader, PTI's 75,000 sq. ft. facility meets the needs of the OEM community for volume production, quality, and delivery.

Jan Crystals

Founded in 1965, Jan Crystals is focused on advancing frequency control technology, primarily through exploring and expanding the capabilities of quartz crystals. Beginning with your application or design concept, Jan Crystals serves you with a full range of technical services, fro consulting engineering through prototyping, analysis, precision manufacturing, and customer support. To bring your design concepts for reality and create solutions for demanding applications, see the Jan Crystal display at booth 310.

American Technical Ceramics

ATC manufactures high quality RF/microwave capacitors including high Q, low ESR porcelain MLCs, high performance ceramic MLCs, low cost commercial line SLCs and low cost standard size MLCs. ATC also offers special assemblies, dielectric substrates, non-magnetic capacitors, custom SLCs, and Thin Film Technology., This product line is described in the new ATE Surface Mount and Leaded Capacitor brochure.

TRAK Microwave Corp.

TRAK offers synthesizers and other oscillator-related products, specializing in multifunction assemblies. Components include oscillators, XCO, OCXO, TCXO, VCXO, VCO, DSO, DTO, and phase locked designs; frequency multipliers, comb generators, IF amplifiers, isolators, and circulators. RF and IF passive signal processing components include mixers, couplers, splitters, hybrids, I/Q modulators and demodulators, phase shifters, variable attenuators, and matching transformers. Time related products include GPS clocks and Frequency Standards.

Compact Software

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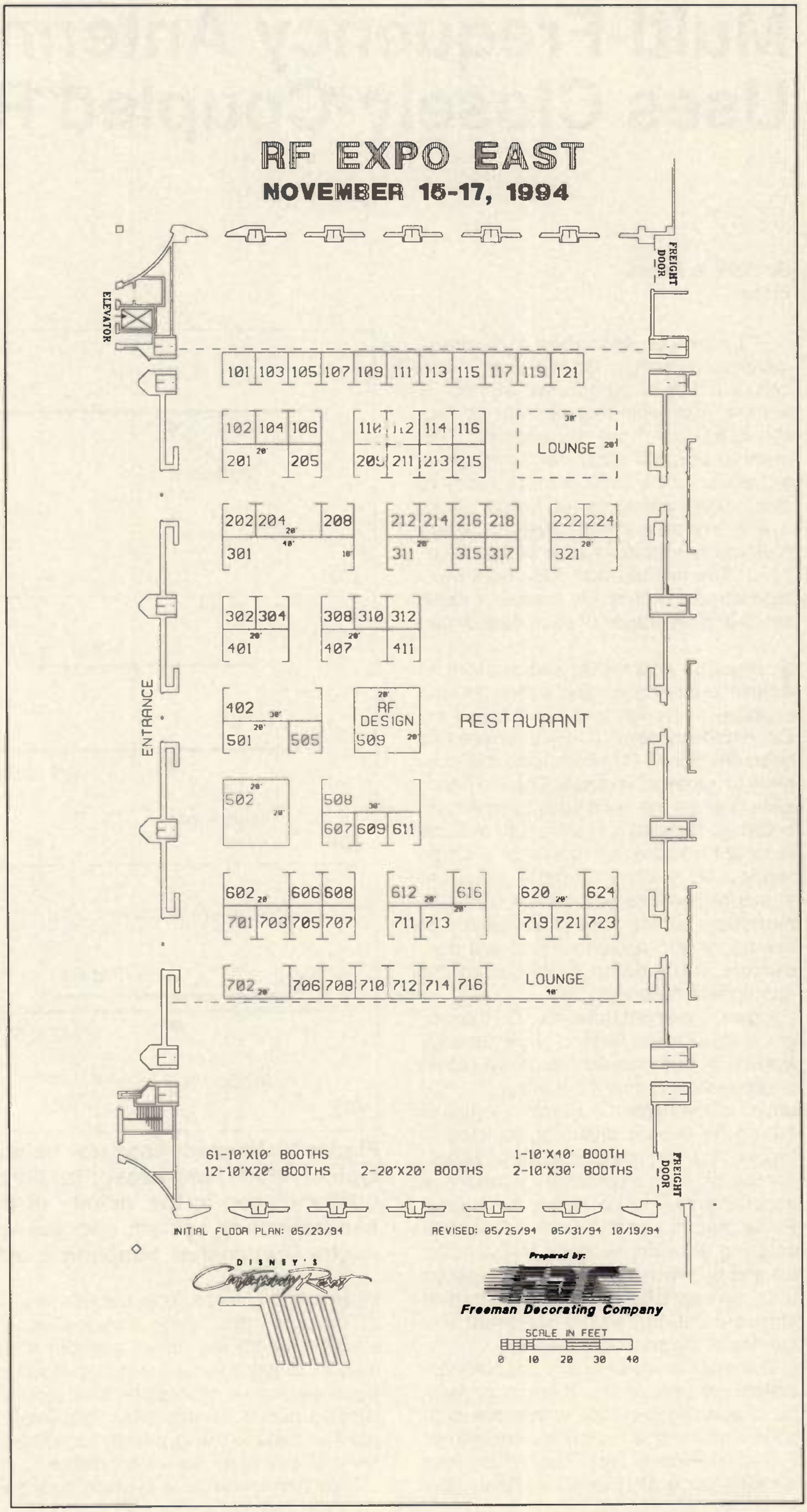


Multi-Frequency Antenna Technique Uses Closely-Coupled Resonators

By Gary A. Breed
Editor

This article describes a recently-developed method for the design and construction of dipole and monopole antenna elements that operate on two, three, four or more frequencies. This method permits such multi-frequency antennas to be built without reactive decoupling networks or tuned stubs, which are commonly used to obtain multiple resonances at a single feedpoint. The method also has the advantage of control over the feedpoint resistance and reactance at each frequency.

It is well known that conductors in close proximity exhibit strong mutual coupling. A design technique called the Coupled-Resonator (C-R) principle has been developed [1] which uses this coupling to great advantage. The C-R principle defines the conditions for optimum coupling, creating a system with multiple resonant frequencies, driven at a single feedpoint. Such a multiple-resonant structure consists of a driven dipole or monopole at the lowest frequency of operation, with additional resonant conductors surrounding it, placed at the appropriate distances.

Figure 1 demonstrates the C-R principle in its simplest form, a two-frequency system. A half-wavelength driven dipole is resonant at some frequency, F_1, and driven at the center. A typical return loss sweep for such a dipole is depicted in Figure 1(a). In Figure 1(b), an additional conductor, half-wavelength resonant at an arbitrarily-chosen higher frequency, F_2, is placed nearby. Some degree of coupling will exist between this conductor and the driven dipole, and the return loss sweep of the dipole feedpoint shows a "bump" at the resonant frequency of the second conductor.

The main premise of the coupled-resonator principle is that there is an optimum spacing distance where the coupling results in a matched condition at F_2, as in Figure 1(c). The return loss remains good at F_1 and, therefore, the system is matched at both frequencies.

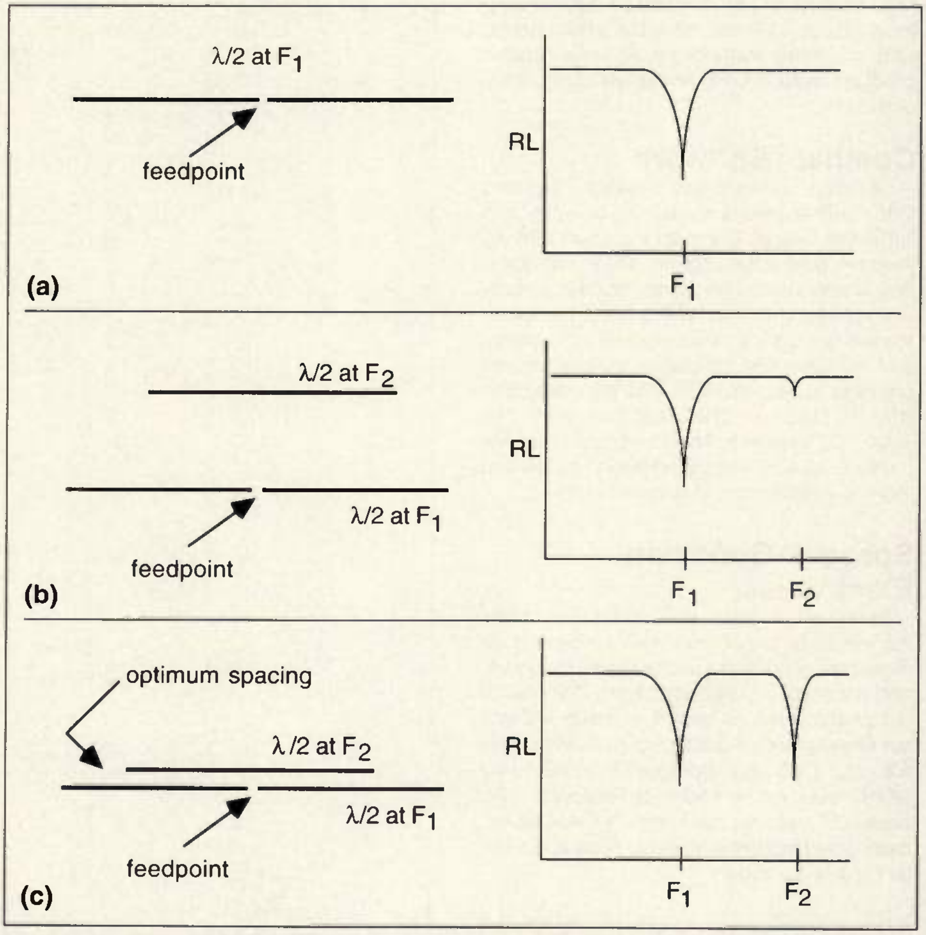


Figure 1. Coupled-resonator behavior: (a) shows a simple dipole and its typical return loss sweep; (b) illustrates the effect of an additional conductor placed in the vicinity of the first dipole, and; (c) shows how a two-frequency system appears when the spacing of the second conductor is such that coupling is optimum.

The above description also applies to systems where the driven element is a monopole fed against ground, given the equivalence of monopole and dipole configurations. In this case, the feedpoint impedance of a monopole will be one-half that of an equivalent dipole.

This two-frequency system can be expanded to three, four, five or more frequencies by adding additional resonators and placing them radially around the fed dipole or monopole, as shown in Figures 2 and 3. A practical upper limit on the number of frequencies this structure will support is reached when the complexity of multiple interactions obscures the desired coupling. The

actual number of frequencies obtainable depends on various factors which determine the degree of coupling, such as harmonic resonances and small frequency separations. Systems up to seven frequencies have been successfully modeled.

Design Equations

The variables involved in the design of antennas using the C-R principle are: conductor diameter, conductor spacing, feedpoint impedance, and the ratio of frequencies. These are all defined from the point of reference of the additional frequency under consideration, F_n.

Conductor spacing follows this general relationship:

$$\frac{\text{Log}(d)}{\text{Log}(D/4)} = .54$$

where d is the distance between conductors and D is the diameter of the conductors, both expressed in wavelengths at F_n. This approximation is normalized for a feedpoint impedance at F_n equal to that of the dipole in free space (72 ohms) or a monopole over perfect ground (36 ohms), and for an F_n/F_1 ratio of 1.3 or greater.

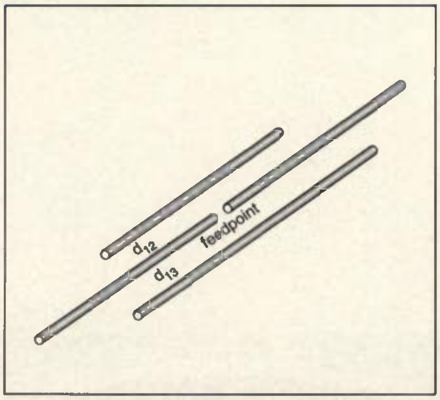


Figure 2. Pictorial representation of a three-frequency C-R dipole.

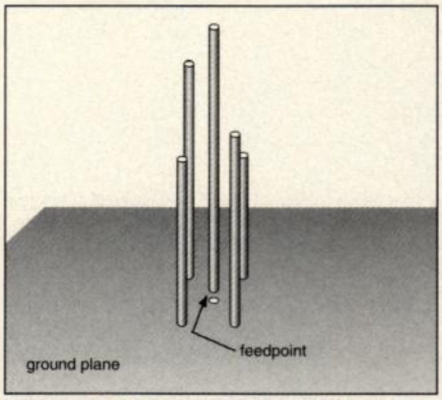


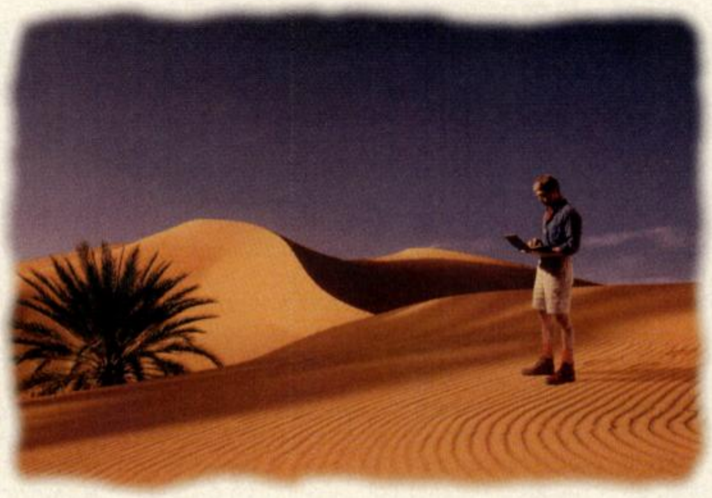
Figure 3. Pictorial of a five-frequency C-R monopole element.

For broader applicability, the equation can be modified to allow for a wider range of impedances and lower F_n/F_1 ratios. Using a straight-line approximation for impedance and a first-order $1/e^{-x}$ curve-fit for frequency ratio correction, the original equation then becomes:

$$d_{1n} = 10^{[0.54\text{Log}(D/4)]} \times \frac{Z_0 + 35.5}{109} \times \left[1 + e^{-[(((F_n/F_1)-1.1)\times 11.3)+0.1]}\right]$$

where,

Z_0 is the desired feedpoint impedance at



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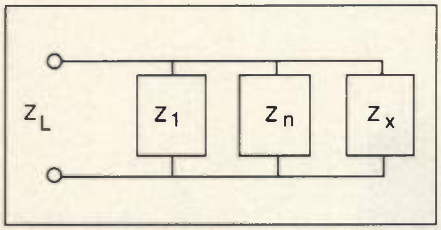

Figure 4. Simplified equivalent circuit at Fn.

F_n, within the range of 25 to 125 ohms.
F_1 is the resonant frequency of the driven dipole
F_n is the resonant frequency of the addional resonator
F_n/F_1 frequency ratio is greater than 1.1:1
d is in the range of 0.01 to 0.00001 wavelength

A significant point is that the impedance can be independently controlled at each frequency, F_2, F_3 ... F_n, by adjustment of the spacing. Combined with the reactance change in an antenna as length is altered, a wide range of adjustment can be obtained.

There are two additional characteristics that can be explained by the simplified equivalent circuit shown in Figure 4. At F_n, the feedpoint impedance is the combination of Z_1, the impedance of the driven dipole or monopole, and Z_n, the coupled-resonator impedance, plus Z_x, which is the total effect of any other resonators in the system (predominantly capacitance). Compensation for Z_x is readily achieved by simply lengthening the resonators to add inductance, typically by 0.25 to 0.5 percent

Another effect is an apparent anomaly that occurs when the ratio of F_n/F_1 is approximately 3, where a significant increase in the spacing is required. This is readily explained by noting that the driven dipole has a relatively low impedance at 3/2 wavelength (3/4 wavelength for a monopole). In this case, Z_n must be higher than normal to achieve the desired parallel combination of Z_1 and Z_n, which corresponds to a greater spacing distance.

Radiation Characteristics

Antennas designed according to the C-R principle are accurately modeled using method-of-moments analysis, including software based on either the Numerical Electromagnetics Code (NEC) or MININEC3 [2]. Figure 5 shows the modeled free-space radiation pattern of a three-frequency C-R dipole, using the program ELNEC [3]. The modeled directivity (gain) is shown at the

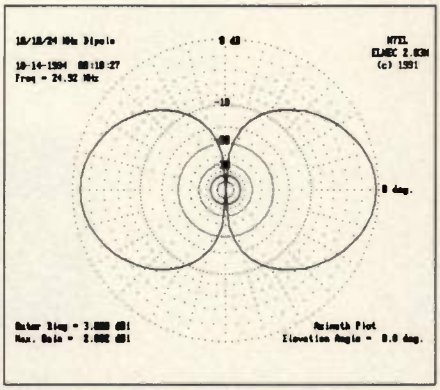

Figure 5. Modeled radiation pattern at the highest frequency of a three-frequency C-R dipole.

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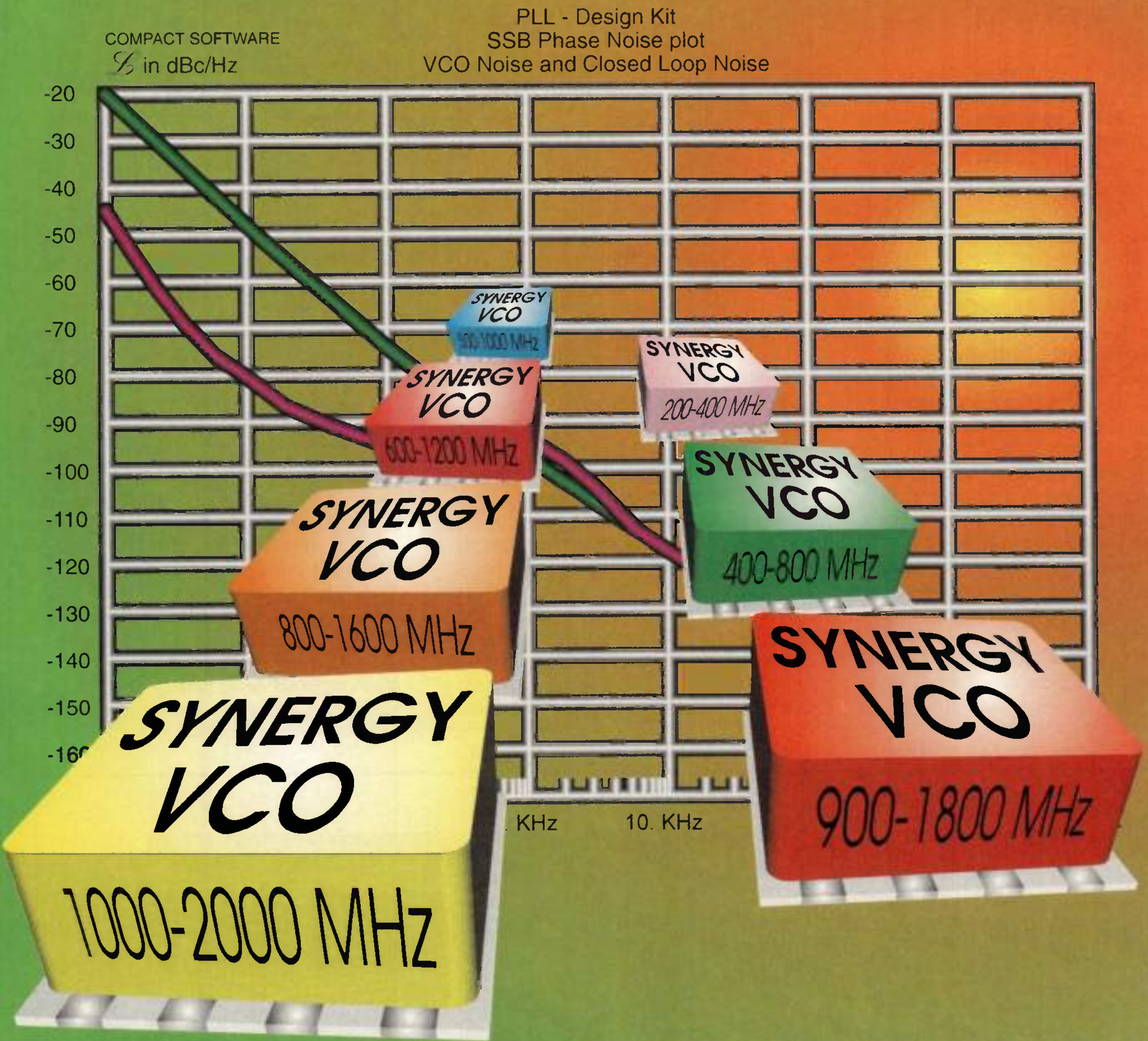

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highest of the three frequencies, and is very close to that of a simple dipole (≈0.5 dB greater), suggesting that radiation is primarily from the resonant conductor. The slight gain over a dipole at this highest frequency also suggests that small in-phase current is present in the portion of the driven dipole which extends beyond the active region. Analysis of the currents in the antenna model verifies these conclusions.

Advantages and Limitations

The principal advantage of this antenna design is the absence of reactive components, such as tuned circuits or capacitively-loaded coaxial stubs, which are often used to achieve multi-frequency operation. These components may introduce losses, or require time-consuming tuning adjustment. The C-R antenna design achieves its performance by controlling the physical dimensions of conductor length, diameter and spacing.

Another significant advantage is that the feedpoint impedance at each additional frequency can be controlled by adjustment of resonator spacing and length. When a C-R antenna element is placed in an array, the mutual impedances can be significantly different at each operating frequency. The C-R principle allows each frequency's resonator to be adjusted over a useful range of resistive and reactive impedance.

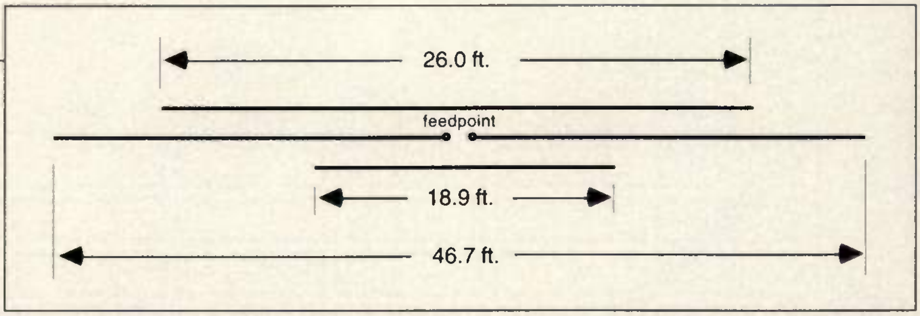

Figure 6. As-built dimensions of a three-frequency C-R antenna for 10.1, 18.1 and 24.9 MHz, using #12 AWG conductors.

Two limitations should be noted. First, the tradeoff for electrical simplicity is a relatively complex mechanical assembly. The structure must support a central dipole or monopole and maintain spacing with the additional resonators with insulators or other means. However, it should be noted that other multi-frequency configurations also have special construction requirements. The other limitation of the C-R method is a reduction in VSWR bandwidth at F_2, F_3 and any higher frequencies of operation, compared to a simple dipole or monopole. This shortcoming can be mitigated by the use of large-diameter conductors, or in extreme cases, additional resonators with overlapping coverage. Again, other common multi-frequency antenna designs also exhibit reduced bandwidth, although some have losses which reduce the system Q and create an apparent increase in bandwidth.

A Practical Antenna

Various antennas were constructed to

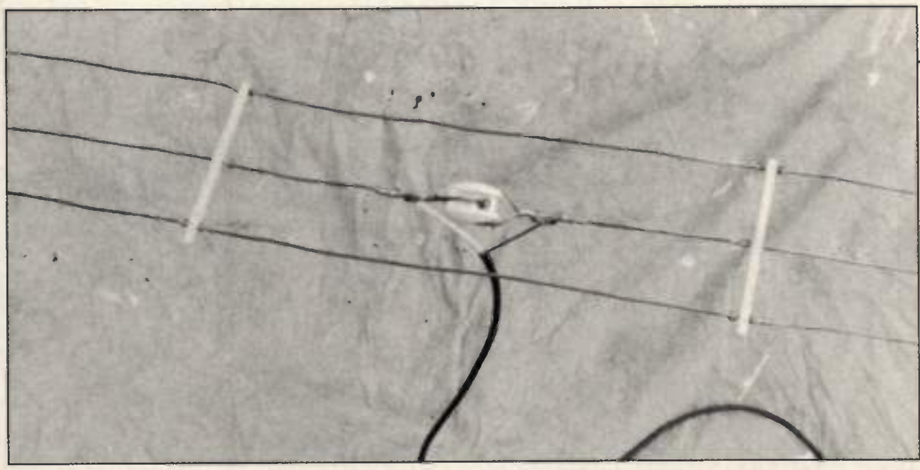

Figure 7. Photo of the feedpoint region of the three-frequency C-R antenna for 10.1, 18.1 and 24.9 MHz.

verify the accuracy of the computer models, and to assure that the concept was valid. The first versions of these antennas were designed for HF amateur radio bands, where they could be evaluated "on the air" and compared with other antennas of known performance. Since the early development of this antenna design was a purely personal endeavor, this arrangement allowed extensive experimentation in conjunction with enjoyment of the hobby.

Figure 6 shows the dimensions of a three-frequency dipole constructed from #12 AWG wire. The driven dipole is resonant at 10.1 MHz, with additional resonators for 18.1 and 24.9 MHz. Using equation (2), and choosing 50 ohms as the design impedance for F_2 and F_3, the required spacing was determined to be approximately 1.75 inches (4.5 cm) for each resonator. After modeling the design in ELNEC, the spacing was increased to 2.0 inches (5 cm), primarily to compensate for installation above real ground at a design height of 45 feet (13.7 m). Figure 7 is a detail photo of the feedpoint region. Figure 8 is a return loss sweep from 8 to 28 MHz. The ripple in the display is caused by small reflections in the long test cables. Return loss is greater than 20 dB at the three design frequencies, exceeding 30 dB at the highest two frequencies (those added by coupled-resonators). 30 dB corresponds to a VSWR of 1.06:1.

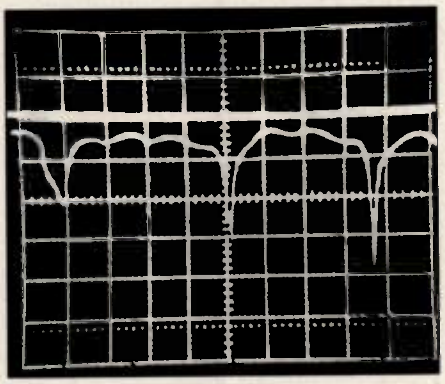

Figure 8. Return loss sweep of the example three-frequency antenna. The scales are 2 MHz/div. horizontal (18 MHz center) and 10 dB/div. vertical.

On-air performance of the three-band antenna proved to be indistinguishable from that of separate dipoles for each band. The radiated performance, the feedpoint impedance, and the variations in impedance with installation height above ground also served to confirm the validity of the MININEC-based ELNEC model.

Applications

Design investigations are proceeding into several specific applications where the ability to cover two or more frequencies in a single antenna would be useful. These applications include combined cellular/PCS bands, 915/2450/5700 MHz ISM bands, 150/450 MHz mobile radio, international shortwave broadcasting, amateur radio and others.

Summary

This article has introduced the reader to the Coupled-Resonator principle, a method for the design and construction of multiple-frequency antenna elements which uses controlled coupling of closely-spaced conductors. Although other methods exist for multi-frequency antennas, this method offers a new option for antenna construction. Using this method, frequencies of operation can be arbitrarily chosen, the impedance at each frequency can be controlled to a significant degree, and operation is obtained without the use of reactive isolating circuits or stubs. This design method is patent-pending. *RF*

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1. Gary A. Breed, "A Method for Constructing Multiple-Frequency Dipole or Monopole Antenna Elements Using Closely-Coupled Resonators," patent-pending, July 1994.
2. J.C. Logan, J.W. Rockaway, "The New MININEC (Version 3): A Mini-Numerical Electromagnetics Code," Naval Ocean Systems Center Technical Document No. 938, September 1986.
3. ELNEC antenna analysis program, available from Roy Lewallen, P.O. Box 6658, Beaverton, OR 97007.

About the Author

Gary Breed is Editor and Associate Publisher of *RF Design*, and Editorial Director of *EMC Test & Design*. He maintains an active interest in the design areas of: antennas, low-cost transmitting and receiving equipment, power amplifiers and zero-IF techniques. Gary can be contacted at *RF Design*, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111, or by telephone at (303) 220-0600.

Designing Punch-Through Varactor Diode Frequency Multipliers

by Paul H. Williams
MIT Lincoln Laboratory

Frequency multipliers are very useful circuits in both the RF and microwave region. For instance, to provide a clean, stable local oscillator (LO) signal at microwave frequencies, a frequency multiplier circuit will multiply the output of a 5-100 MHz oscillator. The same procedure can be used in a transmitter where the signal is multiplied either before or after the power amplifier. These techniques are preferred over a stand-alone UHF or microwave oscillator, because high frequency oscillators are not as quiet nor as stable as a frequency-multiplied low-frequency oscillator.

There are two general types of frequency multipliers: nonlinear-resistive and nonlinear-reactive. Nonlinear-resistive multipliers include comb generators and step-recovery diode (SRD) multipliers without tuning, while nonlinear-reactive multipliers include abrupt-junction varactor diode, punch-through varactor diode, and transistor multipliers. Abrupt-junction varactor diode multipliers were heavily investigated back in the 1950's [1]. Then in the 1960's, punch-through varactor diode multipliers were developed [2,3]. They proved easier to use than abrupt-junction diodes because their operating characteristics are independent of their input power level, they are self-biasing, and they are not hard to stabilize. Transistor multipliers were also developed in the 1960's, and though they provide gain, their operation is dependent on their input power level.

Because punch-through varactor diode multipliers are easy to successfully build, I will show how to design and build lumped-element doublers and triplers. With an understanding of the circuit, these designs can also be used with distributed circuits in the microwave region. While I will not go into a complete theoretical analysis, several sources are available for more information [2,3,4].

How They Work

The key to how a punch-through varactor diode multiplier works is in the characteristics of the diode. To make life easier, we will be referring to the elastance, S, of the diode, which is the inverse of the capacitance. Since the punch-through diode is a varactor, we are naturally interested in its incremental elastance versus charge characteristic (Figure 1), which is a step function. A punch-through diode's incremental elastance versus voltage characteristic is also a step function. This step characteristic is the nonlinearity on which these multipliers are based.

The constant elastance, or capacitance, in the reverse-biased direction ($q > 0$ and $v > 0$ for our sign convention) is due to the depletion region rapidly expanding (or punching-through) to encompass the whole diode. This is the reason for the diode's name. Once punch-through happens, adding any more charge does not change the diode's elastance.

The infinite capacitance or zero elastance in the forward direction is due to the long minority carrier lifetime, τ, of the diode. Typically, we want a diode with a τ at least an order of magnitude larger than the fundamental period of the input signal. Since the charge carriers never have enough time to be swept away, the diode appears to have an infinite capacitance in the forward direction.

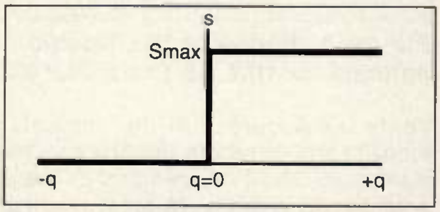

Figure 1. Elastance versus charge characteristic of punch-through diodes.

Keep in mind that the above characteristics are for an ideal punch-through diode, and a real diode will deviate from the above characteristic. For instance, we assumed that the transition from the forward region to the reverse region occurs at zero charge and voltage, when in reality, it occurs at a small forward voltage. Also, the transition is not as sharp as shown and may take a volt or so to reach the diode's minimum capacitance. Finally, the diode's minimum capacitance will decrease slightly as the reverse bias is increased.

Though a punch-through diode has a sharp nonlinearity that produces a lot of harmonic power, it will not work well as a

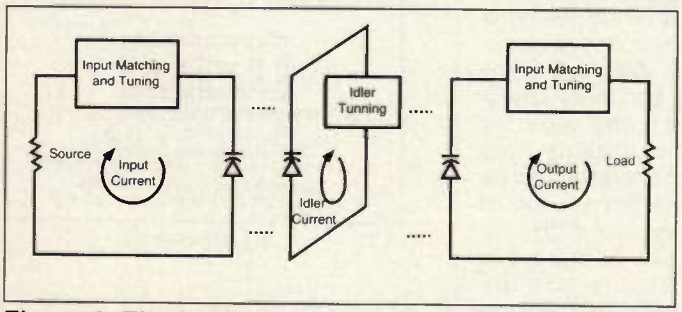

Figure 2. The basic current loops of a punch-through varactor diode multiplier.

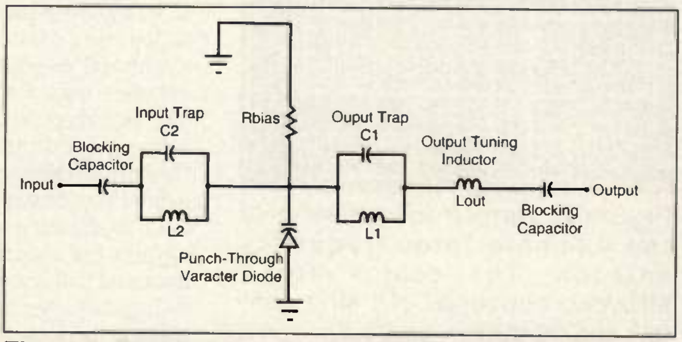

Figure 3. Punch-through varactor diode doubler circuit.

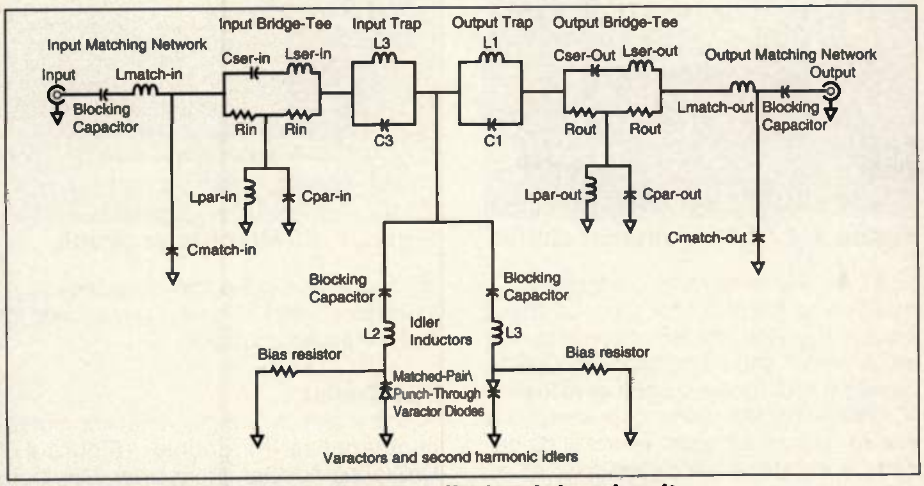

Figure 4. Punch through varactor diode tripler circuit.

multiplier unless it is placed in a properly designed circuit. A properly designed frequency multiplier will consist of two or three different types of current loops (Figure 2): the input loop; the output loop; and for triplers and higher order multipliers, one or more idlers. The input loop tunes out the average capacitance of the diode at the fundamental frequency and prevents any currents except the fundamental current from flowing. Additionally, the input loop will match the source impedance to the diode's nonlinearity. Though the nonlinearity appears resistive to the source, it is not a resistor because it is not linear. Like the input loop, the output loop tunes out the average capacitance of the diode at the output frequency and prevents currents besides the output current from flowing. Also, the output loop will match the load impedance to the diode's nonlinearity. For triplers and higher order multipliers, there are also idlers at one or more intermediate harmonics. The idlers' purpose is to tune out the average elastance at a particular intermediate frequency so as to maximize the current flowing at that frequency. Idlers are required for odd-order multipliers but not even-order multipliers. However, even-order multipliers are more efficient with idlers than without.

For a complete analysis of punch-through varactor diode multiplier operation, see references [4], [2], and [3].

Be aware that no semiconductor manufacturers advertise punch-through diodes. However, they are the same as SRDs that are available from several manufacturers.

The Doubler

The doubler is a simple circuit with just one shunt punch-through diode. In order to divide the circuit into an input and an output loop, we put two traps into it. The input trap is tuned to the output frequency and the output trap is tuned to the input frequency. The input trap elements are chosen so that the input trap tunes out the average capacitance of the diode at the input frequency. However, the output trap can not tune out the average capacitance of the diode at the output frequency because the output trap is capacitive at the output frequency. Consequently, an output tuning inductor is needed. No matching networks are needed since the equivalent input and output resistances are equal.

The pertinent design equations for the doubler are shown in Appendix A. Please note that these equations are valid only if the fundamental radian frequency, ω_0, is far below the diode's radian cutoff frequency, ω_c, or

$$\omega_0 << \omega_c \qquad (1)$$

where

$$\omega_c = \frac{1}{R_s C_{min}} \qquad (2)$$

Here, R_s is the series resistance of the diode, and C_{min} is the reverse-biased capacitance of the diode at the breakdown voltage.

If (1) holds true, we can calculate from (3) the required minimum capacitance for a diode in a system with a characteristic impedance of Z_0 operating at a fundamental input frequency of f_0. Since we have forced the input and output resistances of the doubler to equal our system impedance, we do not need any matching networks.

From (4), we can calculate the mini-

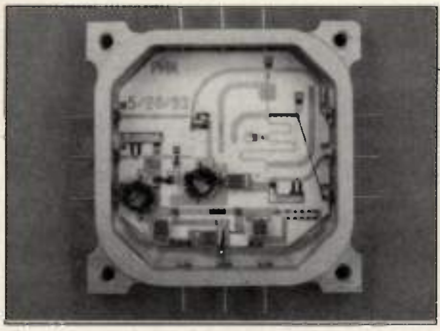

Figure 5. 2.64 GHz doubler circuit.

mum reverse breakdown voltage, V_{max}, needed to handle the desired input power, P_{in}. For efficient operation, the input power should be high enough to require a breakdown voltage of at least 5 V; otherwise, the diode operates more like an abrupt-junction varactor diode, and the equations are not valid.

The bias resistor is calculated from (5) where τ is the minority carrier lifetime of the diode [5]. This is only an approximate expression, so the actual value of R_{bias} should be determined experimentally. As mentioned before, τ must be much longer than the period of the fundamental frequency.

The input trap components, C_2 and L_2, are calculated from (6), and the output trap components, C_1 and L_1, are calculated from (7). Additionally, we can calculate the output tuning inductor's value, L_{out}, from (7).

In order to fully optimize a design, I suggest you use one variable component in each trap and a variable resistor in the bias network. After building the circuit, first tune the input traps using a low level signal (≈ -10 dBm). Once they are nulled properly, connect an input signal (at the designed power level) to the doubler. Adjust the bias resistor and then the traps for the best efficiency, while keeping the doubler stable. Fortunately, doublers are fairly easy to properly tune.

As an example of a doubler, I built a 2.64 GHz doubler that I designed for an input power of 10 dBm in a 50 Ω system (Figure 5). The picture of the doubler also includes a separate amplifier circuit preceding the doubler. I used a M/A-COM MA44621A-C diode with a $C_{min} = 0.25$ pF and a $V_{max} = 20$ V. Because of the high frequency involved, I implemented the traps using microstrip transmission lines. Though the input trap could have been designed with the proper transmission line impedances to resonate with the average elastance of the diode, I instead chose to just use impedances that were realizable in microstrip. This required an extra matching network at the input. The circuit has an input return loss of –13 dB and an efficiency of 32% at an input frequency of 2.64 GHz. I suspect much of the loss is due to the microstrip lines both in the circuit and in the test fixture used to test the packaged circuit.

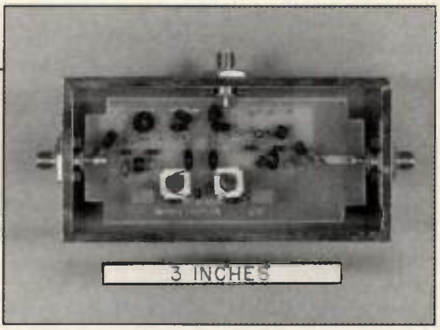

Figure 6. 50 MHz tripler circuit.

The Tripler

The tripler is a more complex circuit compared to the doubler (Figure 4). Unlike the doubler, there is an idler loop in addition to the input and output loops. Though I could have used one shunt diode as I did in the doubler, I would have needed second harmonic traps in both the input and output loops. Instead, I chose to use a push-pull arrangement of two matched diodes that inherently minimizes all even-order outputs. The two diodes in series together with their respective idler inductors form the second harmonic idler loop. Both the input and output loops consist of the parallel combination of the diodes and their respective idler inductors. In addition, both the input and output loops have traps like the doubler; however, unlike the doubler, both traps also tune out the combined reactance of the diodes and their idler inductors. Because the tripler tends to be unstable, I have included bridge-tee circuits in both the input and output circuits to provide known out-of-band resistive terminations for the diodes. Another difference from the doubler is that the input and output impedances of the tripler are different, so I added matching circuits in both the input and output circuits. Finally, there is a bias resistor for each diode.

The pertinent design equations for the doubler are shown in Appendix B. As with the doubler, these equations are valid only if the fundamental frequency, ω_0, is far below the diode's cutoff frequency, ω_c, as mentioned in (1).

Provided that (1) holds true, we can calculate the triplers input and output resistances, R_{in} and R_{out}, from the required characteristic impedance of the system, Z_0, from (8). Using R_{in}, we can calculate the minimum reverse-biased capacitance, C_{min}, of both diodes, along with their idler inductances, L_2, in (9).

From (10), we can calculate the minimum reverse breakdown voltage, V_{max}, needed to handle the desired input

$$C_{min} = 0.03377\frac{1}{Z_0 f_0} \tag{3}$$

$$V_{max} = 1.591\sqrt{\frac{P_{in}}{f_0 C_{min}}} \tag{4}$$

$$R_{bias} = \frac{\tau}{C_{min}} \tag{5}$$

$$C_2 = \frac{2}{3}C_{min} \quad L_2 = \frac{1}{4\omega_0^2 C_2} \tag{6}$$

$$C_1 = \frac{8}{3}C_{min} \quad L_1 = \frac{1}{\omega_0^2 C_1} \quad L_{out} = \frac{2}{3}L_1 \tag{7}$$

Appendix A. Doubler circuit design equations

power, (P_{in}). As in the doubler, the input power should be high enough to require a breakdown voltage of greater than approximately 5 V; otherwise, the diode operates more like an abrupt-junction varactor diode, and the equations are not valid.

The bias resistor, R_{bias}, is calculated from the same equation used for the doubler and is redisplayed in (10) where τ is the minority carrier lifetime of the diode[5]. This is only an approximate expression, so the actual value of R_{bias} should be determined experimentally. As mentioned previously, τ must be much longer than the period of the fundamental frequency.

Before designing the bridge-tee circuits, the loaded Q for each side must be determined. Generally, the higher the loaded Q, the more stable the multiplier, and the lower the loaded Q, the more realizable the components. The exact loaded Q depends on the quality of the inductors and capacitors that you can make or buy. Once the loaded Q is determined, the bridge-tee component values can be calculated from (11)-(14).

Continuing on with the rest of the circuit, the input trap is calculated from (15), the output trap from (16), the input matching circuit from (17), and the output matching circuit from (18)[6].

For the tripler, I suggest you use variable components for as many of the components as possible. The equations will get you close to, but not exactly on, the optimum values. After building the circuit, tune the traps for maximum attenuation and the bridge-tees for minimum loss at their respective frequencies using a low level signal (≈ -10 dBm). Once they are tuned properly, connect an input signal at the designed power level to the tripler and adjust the idlers and bias resistors for the best efficiency, while keeping the tripler stable and the even-order harmonics at a minimum. Occasionally, you may also need to readjust the traps and then return to adjust the idlers and bias resistors again.

As an example of a tripler, I built a 50 MHz tripler that was designed for an input power of 1 watt in a 50 Ω system (Figure 6). I used two Alpha Industries DVB6145-19 diodes with a C_{min} = 8.8 pF, V_b = 75 V, R_{in} = 75 Ω, and R_{out} = 27 Ω. The circuit has an input return loss of 22 dB and an efficiency of 53%.

Conclusion

With the push towards higher operating frequencies and higher performance systems comes the need for very stable and quiet high-frequency LOs. This usually requires some sort of frequency multiplication. I have shown one such frequency multiplying circuit, the punch-through varactor diode multiplier, that offers the advantages of robustness and stability. Included are detailed design equations and examples of circuits already built and operating.

Acknowledgements

I would like to thank Dr. Robert Rafuse for his many helpful discussions on punch-through varactor multipliers as well as on general circuit theory. Additionally, I would like to thank Dr. Donald Steinbrecher for his original thesis and his discussions on the multipliers. This work was sponsored by the Department of the Army under Air Force Contract number F19628-90-C-0002. *RF*

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$$R_{in} = \frac{5}{3}Z_0 \quad R_{out} = \frac{3}{5}Z_0 \tag{8}$$

$$C_{min} = 0.03283\frac{1}{f_0 R_{in}} \quad L_2 = \frac{1}{8\omega_0^2 C_{min}} \tag{9}$$

$$V_{max} = 0.916\sqrt{\frac{P_{in}}{f_0 R_{in}}} \quad R_{bias} = \frac{\tau}{C_{min}} \tag{10}$$

$$L_{ser-in} = \frac{R_{in}Q_{in}}{\omega_0} \quad C_{ser-in} = \frac{1}{\omega_0 R_{in} Q_{in}} \tag{11}$$

$$L_{par-in} = \frac{R_{in}}{Q_{in}\omega_0} \quad C_{par-in} = \frac{Q_{in}}{\omega_0 R_{in}} \tag{12}$$

$$L_{ser-out} = \frac{R_{out}Q_{out}}{3\omega_0} \quad C_{ser-out} = \frac{1}{3\omega_0 R_{out} Q_{out}} \tag{13}$$

$$L_{par-out} = \frac{R_{out}}{Q_{out}3\omega_0} \quad C_{par-out} = \frac{Q_{out}}{3\omega_0 R_{out}} \tag{14}$$

$$L_3 = \frac{4}{3}L_2 \quad C_3 = \frac{2}{3}C_{min} \tag{15}$$

$$L_1 = \frac{20}{9}L_2 \quad C_1 = \frac{18}{5}C_{min} \tag{16}$$

$$L_{match-in} = \frac{Z_0}{\omega_0}\sqrt{\frac{R_{in}}{Z_0} - 1} \quad C_{match-in} = \frac{\sqrt{\frac{R_{in}}{Z_0} - 1}}{\omega_0 R_{in}} \tag{17}$$

$$L_{match-out} = \frac{R_{out}}{3\omega_0}\sqrt{\frac{Z_0}{R_{out}} - 1} \quad C_{match-out} = \frac{\sqrt{\frac{Z_0}{R_{out}} - 1}}{3\omega_0 Z_0} \tag{18}$$

Appendix B. Tripler circuit design equations.

CAE Modeling of a Broadband Magnetic Field Sensor

By Greg B. Gajda, Art Thansandote, and Dave W. Lecuyer
Radiation Protection Bureau, Health Canada

In this paper, we present the use of a linear circuit simulator (SuperStar by Eagleware Corp.) to model a broadband electronic magnetic field sensor. The modeling is carried out in order to assist in the design of the sensor for the measurement of transient pulsed magnetic fields. The sensor consists of a ferrite-loaded solenoidal coil followed by a toroidal current transformer and a low input impedance differential amplifier. The impedance characteristics of the sensor's individual components were measured using a network analyzer (HP 8751A). From these measurements, a lumped-element circuit model was made for each component. After modeling the individual components, the complete sensor circuit was simulated. The results from the simulation were compared with the measurements of the complete magnetic field sensor. A close agreement was found between the simulated and measured results, indicating the usefulness of the modeling procedure. The sensor has a sensitivity of about 10 mW/μT and an operating frequency range from 30 Hz to 2 MHz.

Broadband magnetic field sensors are used to measure continuous wave magnetic fields in a broad range of frequencies. They are also essential for measuring pulsed or transient magnetic fields generated by natural phenomena such as lightning, by man-made actions such as switching high-power industrial devices, and by domestic electrical appliances. The measurement of these fields is important in characterization of sources of electromagnetic interference with electronic systems [1], in studies of natural phenomena [2], and in the investigation of possible harmful interactions of fields with living systems [3,4]. A number of magnetic field sensors have been developed with output voltage either proportional to the field [5], or to the time derivative of the field [2,6]. The latter sensors are passive devices that require an integrator to make the output proportional to the field. These sensors are commercially available. However, they cause phase distortion and are insensitive at low frequencies. Thus, the information obtained may not be sufficient to describe the characteristics of the measured field at these frequencies. Recently developed sensors of the first type [7,8] have been designed to remedy this problem and are suitable for measurement of magnetic fields at frequencies below 50 Hz. In this paper, we discuss the use of a linear circuit simulator to model a broadband magnetic field sensor originally described in reference [8]. The modeling is carried out to assist in the redesign for improvement of the sensor.

Basic Principles

A schematic diagram of the broadband magnetic field sensor is shown in Figure 1. The sensor utilizes a field sensing coil coupled through a toroidal current transformer to two identical current followers (amplifiers). The sensing coil is a multi-turn solenoidal coil wound on a ferrite rod. Both ends of the coil are joined to form the primary winding of the current transformer. The secondary of the current transformer is also a multi-turn coil, but has a center-tap ground. The output terminals of both current amplifiers are connected to a differential amplifier. In addition to the field sensing coil, an additional coil (not shown) is wound on the same ferrite rod to enable test signals to be injected into the sensor. When the sensor is used for measuring fields, the test coil is left open-circuited.

In the presence of a magnetic field, voltage is induced on the terminals of the sensing coil. The current, produced in the primary winding by the induced voltage, is coupled to the electronic circuitry through the transformer, and then amplified. The output voltage of the differential amplifier is proportional to the magnetic field or flux density. The low roll-off frequency demarcating the beginning of the operating response is given by:

$$f_{low} = R/(2\pi L) \qquad (1)$$

where L represents the total inductance and R is the total resistance. The total

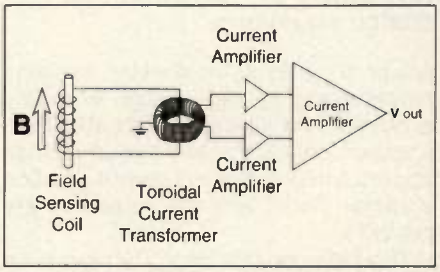

Figure 1. Schematic of a broadband magnetic field sensor.

inductance consists of the sensing-coil inductance and the current-transformer inductance transformed to the terminals of the sensing coil. The sensing-coil inductance depends on the number of turns, the length and the cross-sectional area of the sensing coil, the gap between successive turns, and the relative permeability of the ferrite rod. The current transformer inductance depends on the number of turns of the secondary winding, as well as the type of the material and the dimensions of the core. The current transformer serves two purposes: impedance transformation, which is related to the number of turns, and the attenuation of the common-mode electric-field interference. To obtain a low f_{low}, it is important to maintain a high total inductance.

The total resistance consists of the wire resistance and the load resistance transformed to the sensing coil. To obtain a low f_{low}, it is necessary to keep a small wire resistance and a low load resistance. This implies that a thick wire should be used to make the sensing coil. Also, the amplifier should have a very small input impedance in addition to the requirements for the appropriate transimpedance (V_{out}/I_{in}) and a wide-band characteristic. A high transimpedance is required for measurements associated with the studies of electromagnetic interference, while a low transimpedance is usually preferred in the measurements for health risk assessment. For the sensing coil, the high frequency limit for measurement of pulsed and transient fields tends to be related

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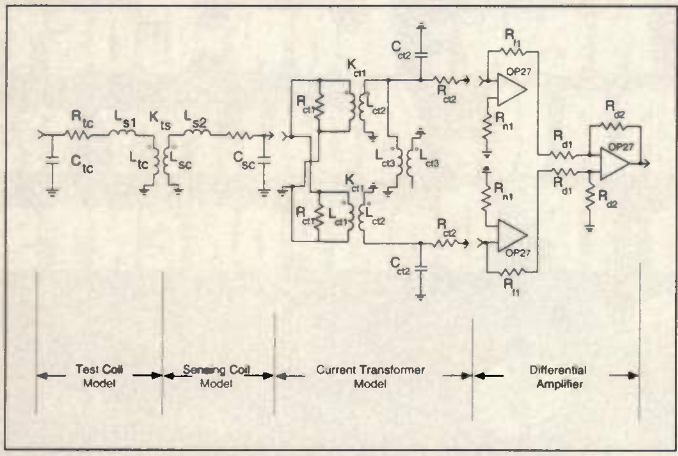


Figure 2. Magnetic field sensor equivalent circuit.

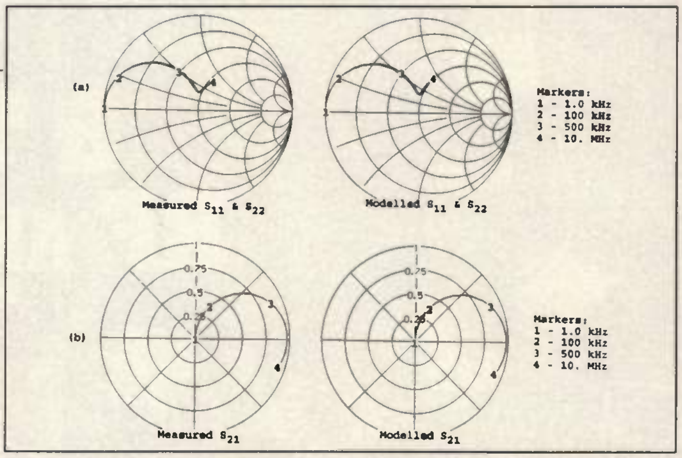


Figure 3. S parameters of the test and sensing coils.

to the transit time on the structure [6], i.e. the electrical length of the solenoidal coil. With this constraint, the high frequency limit may be chosen for the design as:

$$f_{high} < c/(2\pi rN) \quad (2)$$

where c is the speed of light, r is the radius of the solenoidal coil and N is the number of turns. If the sensing coil is electrically small at the highest frequency of interest, f_{high} may be limited by the performance degradation of the ferrite material and the bandwidth of the amplifiers. The transfer function (sensitivity) of the sensing coil is a function of its electrical length and load impedance. To obtain high sensitivity, a large sensing coil and a high-load impedance are required. However, the high-load impedance will cause an increase in f_{low}. Also, the sensing coil cannot be too large, otherwise it will disturb the measured magnetic field [6] and be more sensitive to the unwanted electric field [5]. Therefore, the desired characteristics need to be optimized. High sensitivity is usually a requirement for characterizing sources of electromagnetic interference, while low sensitivity is preferred in human exposure assessment of health risks.

Modeling Procedure

The major components which form the magnetic field sensor were modeled separately. The two-port network characteristics of each component were measured and a lumped-element equivalent circuit was derived to represent the change in behavior with frequency. The values of the elements in the equivalent circuit or model were arrived at by using an optimization procedure. The measurements, in the form of scattering parameters, were made using an HP 8751A network analyzer. The linear circuit simulator used in this procedure was SuperStar by Eagleware Corporation. The circuit simulator possesses two optimization algorithms; one a pattern search and the other a gradient optimizer. Each is manually or automatically selectable by the software. The error function consists of a summation of error terms raised to a power p (p is either 2 or 6 depending on the minimization algorithm) and averaged (weighted if desired) over all the frequencies of interest. The error terms in this work consisted of the difference between the measured and modeled S parameter magnitudes and phases. The optimizer adjusts the circuit-element values until an error function minimum is reached.

Test and Sensing Coil Modeling

The test and sensing coils form a 1:1 ratio transformer. They are wound on a ferrite rod approximately 8.3 mm in diameter and 91 mm long with 211/2 turns of AWG 24 (test coil) and AWG 9 (sensing coil) copper wire. The equivalent circuit of this component forms part of the complete equivalent circuit of the sensor shown in Figure 2. The important elements in the equivalent circuit are the self-inductances of the two coils L_{tc} and L_{sc}, and the ohmic resistance R_{sc} of the sensing coil. The elements $C_{tc,sc}$ represent the parasitic inter-winding capacitances and the elements $L_{s1,s2}$ represent the leakage flux. Originally, two resistances shunting the self-inductances and representing core losses were included in the model; however, they were found to have a negligible effect on the response and were omitted. Because of the extremely small ohmic resistances, R_{tc} and R_{sc}, the modeling procedure is not sensitive enough to determine their values. They were, therefore, computed from tables of DC resistance per unit length. The DC resistances were 0.051 Ω for the test coil and 0.0002 Ω for the sensing coil. Figure 3 shows the measured and modeled S parameters of the test and sensing coils over the frequency range 1 kHz to 10 MHz. The results of the modeling compare favorably to the measured S parameters, as seen in the figure.

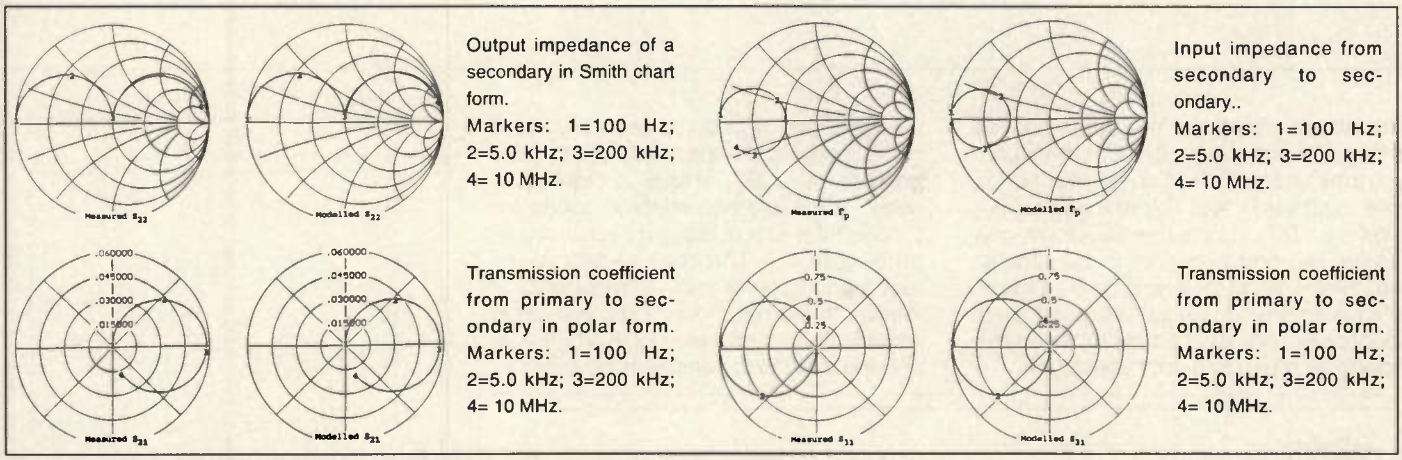
Current Transformer Modeling

The current transformer, shown in Figure 1, consists of a ferrite toroid (16.1 mm O.D. x 9.7 mm I.D. x 6.3 mm wide) with a half-turn of AWG 9 on the primary and 16 turns of AWG 24 on each side of the center-tapped secondary. Being an unbalanced-to-balanced transformer, the two outputs on the secondary are 180 degrees out of phase with respect to each other.

To model a center-tapped, unbalanced-to-balanced transformer, two coupled inductor elements (denoted as MUI in the Superstar library) must be connected as in the model shown in Figure 2. The two primaries are in parallel while the secondaries are in series, with the mid-point being the ground connection or center-tap. This connection will ensure that any signal at the primary will be coupled out-of-phase at each of the secondaries.

An additional MUI element, L_{ct3}, is connected across the secondaries as shown in Figure 2. This element reflects the fact that flux generated by exciting one side of the secondary links the turns making up the other side of the secondary. Since equal numbers of turns are involved, the two secondaries are coupled by a 1:1 ratio inverting transformer as indicated in the model.

Several parasitic elements were omitted from the model because of their negligible effect on the network parameters. The ohmic resistance of the primary is omitted since it is absorbed into the model for the sensing coil. Also, because of the relatively small inductance of the primary (only a half turn), the leakage inductance can be omitted. Leakage inductance was also omitted on the secondary side in order to simplify the model. Finally, the ohmic resistance of the secondary was estimated



Output impedance of a secondary in Smith chart form. Markers: 1=100 Hz; 2=5.0 kHz; 3=200 kHz; 4= 10 MHz.

Input impedance from secondary to secondary.. Markers: 1=100 Hz; 2=5.0 kHz; 3=200 kHz; 4= 10 MHz.

Transmission coefficient from primary to secondary in polar form. Markers: 1=100 Hz; 2=5.0 kHz; 3=200 kHz; 4= 10 MHz.

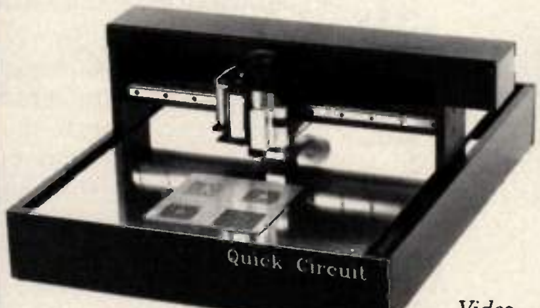
Transmission coefficient from primary to secondary in polar form. Markers: 1=100 Hz; 2=5.0 kHz; 3=200 kHz; 4= 10 MHz.

Figure 4. Measured and modeled S parameters of the current transformer.

from the DC resistance per unit length of AWG 24 wire.

For measurements, BNC connectors were installed on the primary side and on each of the secondaries to ground. Since the component is a three-port network, nine scattering parameters are necessary to define the network. However, because of symmetry and reciprocity, the number of parameters can be reduced significantly. With the primary defined as port 1 and the two secondaries as ports 2 and 3, it was found that the parameters S_{22} (or S_{33}), S_{21} (or S_{31}) and S_{32} were adequate to account for the secondary impedance, primary-to-secondary coupling and secondary-to-secondary coupling, respectively. (For each of these measurements, the unused port is terminated in 50 Ω, by definition of the scattering parameters.)

The remaining parameter S_{11}, necessary to define the impedance of the primary, appeared essentially as a short circuit up to 10 MHz. This is due to the small turns ratio from primary to secondary, transforming the 50 Ω terminations on the secondaries to extremely small resistances across the primary. To overcome this, the two secondaries were open-circuited and a measurement was made of the input reflection coefficient, Γ_p, of the primary. The measured data indicates a parallel R-L-C resonance at the high end of the frequency range (Figure 4), which is useful in determining the values for the primary side elements and, to some extent, the transformed secondary impedances. In



the circuit simulator, the simultaneous fitting of the model to the four measured parameters required three networks; one each for Γ_p and S_{32} and one for S_{22} and S_{21}. This is because Superstar only allows two-port networks to be simulated. Since each network had common elements, the optimizer adjusted them identically until the best fit to the corresponding measured data was found.

From Figure 4 it can be seen that the fit of the model to the measured data is very good for S_{22}, S_{21} and S_{32}, but not as good for the primary reflection coefficient Γ_p when the secondaries are open-circuited (Figure 4c). This may be due to the very small value of primary impedance in relation to the secondary. (The primary is essentially a short piece of AWG 9 wire passing through the center of the toroid.)

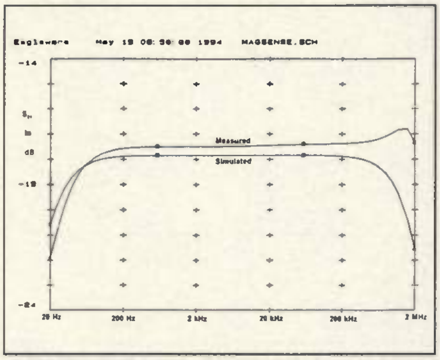


Figure 5. Simulated and measured frequency responses (S_{21}) of the complete magnetic field sensor.

Magnetic Field Sensor Simulation

The complete magnetic field sensor was simulated using the models derived for the test and sensing coils and current transformer, as well as models for the operational amplifiers supplied with the circuit simulator. The results, shown in Figure 5, are compared to a measurement of the actual sensor over the frequency range 20 Hz to 2.0 MHz. The sensor was fabricated using the test and sensing coils and the current transformer described above, along with a small printed circuit board to mount the electronic components. The operational amplifiers were OP-27 op amps manufactured by Burr-Brown Inc. All the components were housed in a 10 cm × 6.5 cm × 3 cm plastic enclosure and were powered by two 12 V rechargeable lead-acid batteries.

The correspondence between the simulated and measured responses is close in terms of the low-frequency roll-off and the mid-band gain. The agreement at the high end of the band is not as good, with the measured response extending beyond the simulated response and possessing a small hump. By adding parasitic capacitances to the op amps in the model, both from the inverting terminal to ground and across the feedback path, the raised "hump" can be reproduced in the simulation. This would indicate that a more detailed model of the op amp may be required at high frequencies, and circuit board parasitics should be taken into account.

Conclusions

All major components of a broadband electronic magnetic field sensor were separately measured for their network characteristics using a network analyz-

er, and subsequently modeled using lumped element circuits. The simulation of the whole sensor circuit was then carried out. The simulated response was found to be in close agreement with the measured results obtained from the test of the whole sensor. This indicates that the modeling procedure is useful in the design of the sensor for desired characteristics. With this procedure, the sensor was improved from the previous one [8] so as to be suitable for measurement of high-intensity transient and pulsed magnetic fields. The improved sensor provides a flat frequency response from 30 Hz to 2 MHz (3 dB roll-off frequencies), has a sensitivity of about 10 mV/μT and can be used to measure fields up to approximately 200 μT. *RF*

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Greg Gajda, Art Thansandote, Ph. D., and Dave Lecuyer are with Health Canada's Radiation Protection Bureau. They are involved in research into the measurements and health effects of electomagnetic fields, including the development of broadband electronic and fiber-optic sensors, the measurement of transient fields from industrial electrical equipment, the investigation of the possible association between exposure to power frequency magnetic fields and cancer in laboratory animals, and investigation into radiofrequency fields in the vicinity of personal communication devices and the resulting energy deposition in humans. They may be reached at the Radiation Protection Bureau, Health Canada, 775 Brookfield Road, Ottowa, Ontario K1A 1C1, Canada. Tel: (613) 954-0309.

Spurious Audio Modulation of VCOs Through RF Coupling

By William F. Egan and Roger A. Lucas
ESL Incorporated, a TRW Company

A challenge in designing frequency synthesizers is safeguarding spectral purity from various corrupting signals, some of which can be very hard to track down. Awareness of the possible signal sources and the processes by which they affect the synthesized output is a valuable asset to the synthesizer designer. This article describes the generation of sidebands that were doubly mysterious because both their source and the process by which they affected the spectral purity were obscure.

We observed an instance in which the VCO in a synthesizer had sidebands that represented modulation at audio frequencies (e.g., 1 kHz) but which were induced by an external RF signal (e.g., 1 GHz). The result was one or more pair of FM sidebands offset from the central spectral line of the VCO by the audio frequencies.

The source of the modulation was not obvious. The sideband amplitude was reduced by increased isolation at the RF port. Yet, providing that port with a high-pass filter, that would greatly attenuate audio frequencies but not RF frequencies, did not reduce the sidebands. The conclusion is that the audio interference was entering the VCO through the output port but at an RF frequency.

There was no apparent source for the audio signal. Yet, by comparing the strength of the audio signal on the VCO's tuning line to the amplitude of the modulation sidebands on its output, we could tell that the interference was entering the loop between these two measurement points, in the VCO. The sidebands appeared on several different VCOs but only when certain devices, an amplifier or a frequency prescaler, were being driven. Yet the sidebands were not observable at the output of these devices when they were driven by a signal generator, rather than by the VCO.

Our Explanation

• The sources of unknown modulating signals are oscillations in GaAs circuits. When we replaced our original GaAs amplifier with a silicon amplifier, the sidebands we had been observing disappeared, but we then noticed weaker sidebands. Their amplitude and frequency changed with the temperature of the GaAs ÷ 2 prescaler that followed the amplifier (which was physically isolated from the prescaler). Low frequency oscillations in GaAs devices have been reported previously [1,2,3]. One paper states: "A subtle but deadly problem too often overlooked by foundries is the low-frequency (1 Hz-10 kHz) substrate oscillation phenomenon." [4]

• High-pass circuits do not block interference because the VCO output is modulated externally and then reflected into the VCO. This is apparent from our inability to decouple the VCO from the modulation source with a filter that strongly attenuates the audio frequency but passes the RF frequency. While we have not studied the process by which the RF signal is modulated, it is easy to conceive of small amounts of phase modulation being induced by an audio voltage at some terminal of an amplifier or frequency divider. For example, an audio voltage could cause the switching threshold in a divider to vary, and thus to vary the time at which the divider switches state. Some of the modulated signal can then be sent back into the VCO, even through an amplifier having finite reverse isolation. It should be apparent that there are potential modulation sources other than the GaAs oscillations that we have implicated here; logic signals or the phase-detector output at the loop reference frequency, for example.

The sidebands are not observable without the VCO because the VCO amplifies them. We have experimental evidence and a theoretical basis to support this and we will present both below.

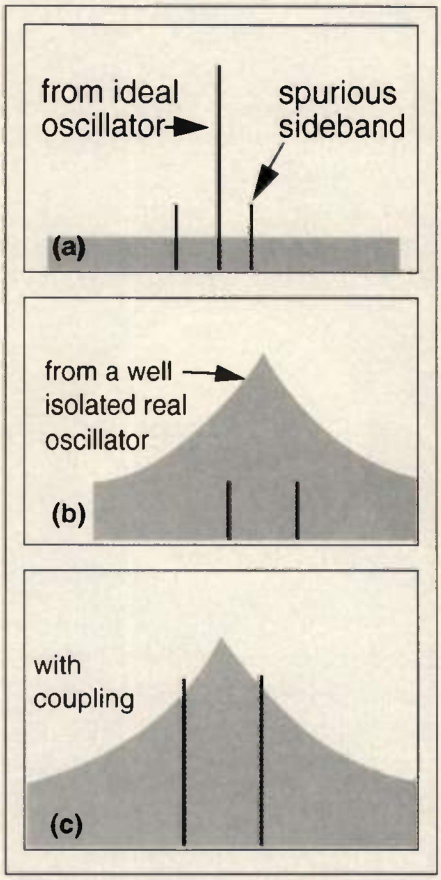

Figure 1. Spurious sidebands, spectral displays of power vs. frequency

They show that the sidebands (Fig. la) are amplified by the same process that increases noise close to the center of an oscillator's spectrum and, as with the noise, the amplification is inversely proportional to the offset of the sideband from spectral center. Thus a weak modulation can be masked (in a given bandwidth) by the noise sidebands of a signal generator whose oscillator is well isolated from the external device (Fig. 1b). However, if that sideband is coupled back into an oscillator it will be amplified and rise with the oscillator's noise sidebands, and may appear above them (Fig. 1c).

Theoretical Basis

Leeson has shown [5] that the noise spectrum of an oscillator with center frequency ω_{osc} and a given loaded Q can be described, at a given modulation frequency ω_m, by the phase deviation $\delta\varphi_{out}$, which is given by [6]:

$$\delta\varphi_{out} = \delta\varphi_{in}\left[\frac{\omega_{osc}}{2Q\omega_m}\right] \tag{1}$$

where $\delta\varphi_{in}$ is the equivalent phase deviation due to noise in some small bandwidth. The process on which this equation is based is one in which ω_{osc} deviates in order to produce a phase shift in the resonator to compensate for the phase shift introduced by the noise. This is needed to maintain 360° phase around the loop, as required for oscillation. The phase deviation $\delta\varphi_{out}$ that corresponds to this frequency deviation may be larger than the original $\delta\varphi_{in}$, as we can see from the equation above. Phase modulation at ω_m causes sidebands at $\omega_{osc} \pm \omega_m$. If $\delta\varphi << 1$ radian, the sidebands will have a level, relative to the carrier, given by [20 dBc $\log\delta\varphi$ $\angle 3$ dB], where $\delta\varphi$ is rms phase deviation in radians. If an injected modulation sideband arrives at the low-power node in the oscillator at a level higher than the effective open-loop phase noise there, the oscillator loop will amplify it by the same factor by which it amplifies the noise. As a result, the sidebands will remain above the noise. However, if the oscillator is better isolated, such that the injected sidebands are buried in the noise (as in the case of a signal generator, for example), the sidebands will not appear above the noise.

Experimental Results

A VCO was locked in a loop with a bandwidth that is narrow enough that the loop does not affect the experimental results. A signal generator was offset slightly from the VCO frequency and injected through a coupler into the output of the VCO (Fig. 2). This single sideband is equivalent to a pair of phase modulation sidebands 6 dB weaker (plus a pair of AM sidebands of the same magnitude) [6]. Fig. 3 indicates the ratio by which the resulting sidebands on the VCO exceeded the injected signal.

As in the noise theory, amplification of the injected level is inversely propor-

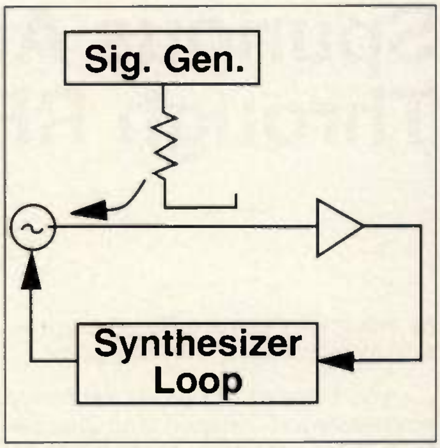

Figure 2. Test setup.

tional to ω_m The oscillator's noise density appears to meet white thermal noise at an offset of about 225 kHz so we would expect the amplification factor, given by the equation, to be 0 dB close to that offset [at 160 kHz when we consider that the phase noise is 3 dB below the thermal noise, so the projection of the oscillator noise slope (all phase noise) has a gain of 3 dB at the apparent intersection with the noise floor. The gain is 6 dB at 160 kHz (= 225 kHz/$\sqrt{2}$). The discrete FM sidebands are -6 dB relative to the injected single sideband so they should have 0 dB apparent gain at an actual gain of 6 dB. If there were no loss, that would occur at 160 kHz.] While the curve shows less than 0 dB at that frequency, this includes loss from the VCO output to the input to the active device. Thus, it is not unlikely that the gain from the low-signal point in the oscillator does go to unity near 225 kHz.

We also ran experiments using a phase modulated carrier. The equivalent gain (the gain from a single sideband 6 dB higher than each of the FM sidebands) from one of these data points is also shown. In many cases the gain limited out because the phase deviation of the VCO became equal to that of the injected signal—the relative sideband level on the output was never greater than that on the injected signal. The gain relative to the injected sideband is apparently identical to that obtained by injecting an offset signal, as long as the injected modulated signal is weak enough.

We believe the first test, wherein a single signal is injected, is more similar to the case where an amplifier, or other device following the VCO, places modulation on the VCO's signal and reflects some of it back to the VCO. Injecting a

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.01-1000	AM-1300	25	0.75	2.0:1	1.4	1.6	1.8	6	$375
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1-100	AU-3A-0110	50	0.5	2.0:1	1.2	1.2	1.3	10	$300
1-500	AU-1A-0150	14	0.5	2.0:1	2.7	2.8	2.9	10	$200
1-500	AU-2A-0150	30	0.5	2.0:1	1.3	1.4	1.5	8	$275
1-500	AU-3A-0150	45	0.5	2.0:1	1.3	1.4	1.5	10	$300
1-500	AU-4A-0150	60	0.5	2.0:1	1.3	1.4	1.5	10	$325
1-1000	AM-2A-000110	25	0.75	2.0:1	1.4	1.6	1.8	8	$300
1-1000	AM-3A-000110	35	0.75	2.0:1	1.4	1.6	1.8	9	$350
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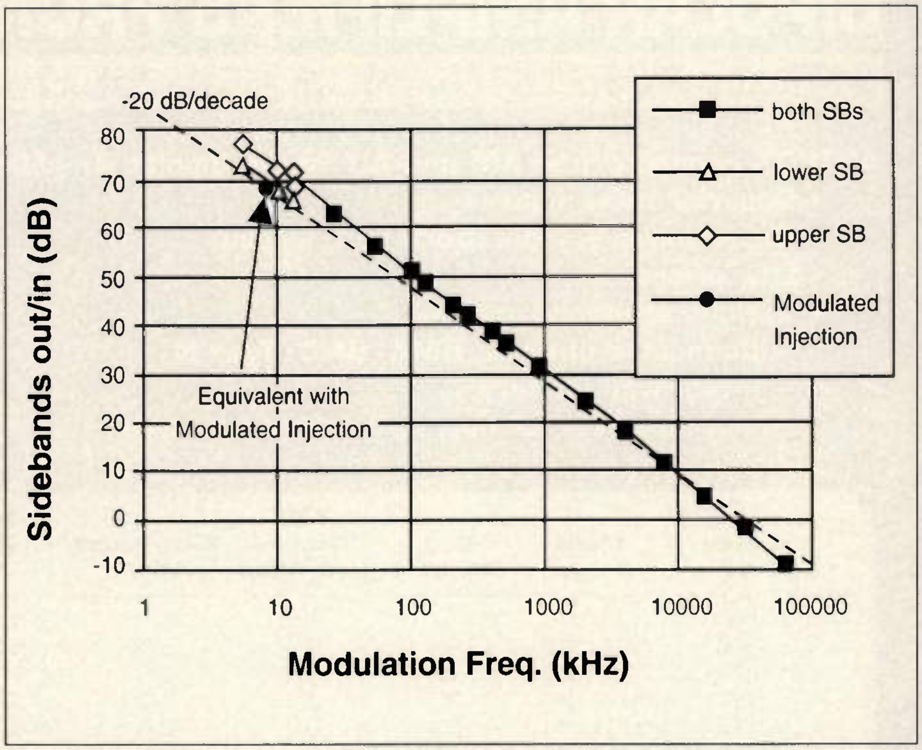

Figure 3. Ratio of output sidebands to level of injected single sideband.

modulated signal whose carrier is at the same frequency as the VCO frequency adds an independent signal, the injected carrier, which is not present in the case of a reflection. A reflected carrier will follow the frequency and phase of the VCO and is not independent.

Relationship to Injection Locking

Mysterious synthesizer sidebands are sometimes explained by the phenomenon of injection locking [6,7]. When the injected signal (possibly some harmonic of a logic signal) does not produce locking, it can cause sidebands [8] that grow larger as their offset from the oscillator frequency is reduced. These phenomena, injection locking, and what has been described here may be related. *RF*

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PC Based Instruments Gaining Ground

This month's product forum highlights PC based instruments. Several manufacturers offer their opinions on today's marketplace and trends in the industry.

National Instruments

We can attribute the slow acceptance of VXI-based RF instrumentation to a fundamental difference between VXI and GPIB-based instruments. VXI defines a small, modular format, a fast

National Instruments' VXI plug and play integrator.

32-bit wide data bus, and special timing and synchronization capability right on the backplane. Although VXI offers many advantages over GPIB controlled systems, VXI instruments do not have front panels. GPIB instruments are bulkier than VXI but users can manually control them using the front panel interface-programming a GPIB instrument into a test system becomes a secondary task. In order to use a VXI instrument right out of the box, users must write a program to control it. However, with the formation of the VXI plug and play systems alliance, using VXI RF instrumentation has become much easier. VXI plug and play standards mandate that executable soft front panels be delivered with the instrument. This means that a software program to control the instrument is available to the end user from day one.

DKD Instruments

The PC instrument market has a lot to offer in terms of price and performance. PC based scopes, ADC/DAC cards, FFT analyzers etc. are all available from various manufacturers. But, RF instruments are rare or simply non-existent.

DKD Instruments offers several PC based spectrum analyzers. Our newest instrument, the SA2600, has a bandwidth of 2.4 GHz. the SA2600 is self contained and fits into a PC drive bay. Communication and control is via the standard parallel interface. This design avoids the pitfall of PC based RF instruments, PC bus and power supply noise. It is very difficult to get enough isolation from computer generated noise using the standard ISA/EISA form factor/interface.

We believe that the future for PC based RF instruments lies with instruments that use external bus standards such as SCSI, PCMCIA and parallel interface, not the ISA/EISA interface. This approach also allows more portability of design and cost to different hardware, i.e. MAC, VME, VXI, etc.

Keithley MetraByte

We see the market growing on the order of 10% a year. Key applications are production/quality test, environmental monitoring, prototype development/ verification, primary R&D, and remote data collection.

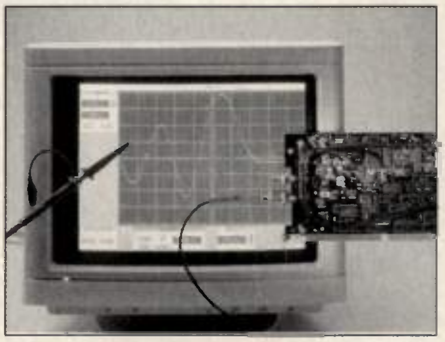
Keithley MetraByte's DacPac.

Next generation PC based instruments should evolve into higher performing feature competitive instruments. In addition to cost, PC based instruments should enjoy a broader packaging advantage as applications move away from traditional 19" rack installations.

The main business thrust is the adoption rate of notebook PCs. This opens up new applications for portable and transportable test instrumentation. Paralleling this is the PCMCIA credit card sized interface which has replaced the traditional expansion bus on portable PCs. Small instruments that are PCMCIA compatible will find adoption in field service and remote testing applications. Keithley MetraByte, through its DacPac high performance portable data acquisition solution for notebook PCs, serves these markets today. PCMCIA data acquisition and instruments products are planned and under development.

PC Instruments

One of the trends in PC-based instrumentation is the "free ride" that the PC-based instrument companies are getting from the computer and communication industries. Driven by the consumer and business markets, these industries are



PC Instruments' oscilliscope.

developing technologies such as the PCI bus, the Pentium, Direct Digital Synthesis, and low-power, high-frequency components.

Unlike their VXI and "rack-and-stack" competitors, PC instrument companies can concentrate their R&D efforts on designing the instrumentation electronics, and not the ancillary technologies. The result of this concentration of efforts has been a rapid increase in the instrumentation performance of PC-based products. Examples of these high performance PC-based products include: 5 1/2 Digit DMMs, scopes with bandwidths of 300 MHz and equivalent sampling rates of 200 Gigasamples/second, transient recorders with real-time sampling rates of 1 Gigasample/second, and 1.6 GHz spectrum analyzers.

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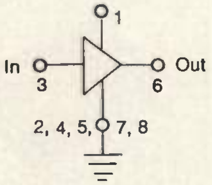

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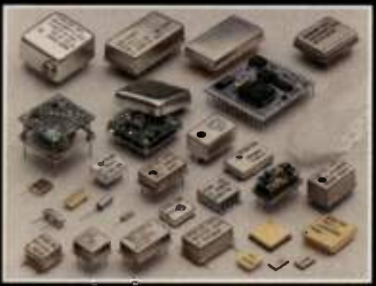

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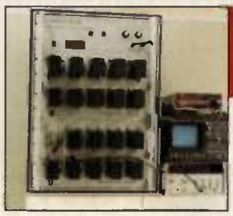

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CEL California Eastern Laboratories

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RF DESIGN is pleased to announce the availability of the Product Showcase section. Your ad in this new monthly advertising section will allow you to:

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- **Achieve Four Color Impact** in this highly visible cost-effective section
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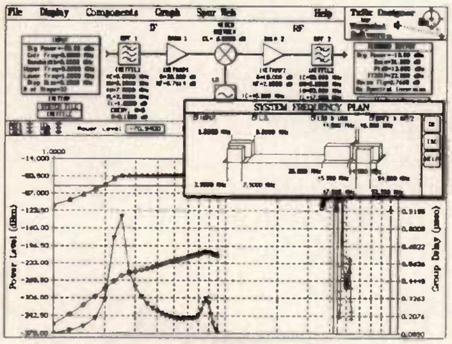

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NEC America, Inc. is a recognized leader in information technology and telecommunication. Due to growth in our Mobile Radio Division, we are seeking an experienced RF Engineer at our Dallas, Texas location.

Position is responsible for the design and development of NEC voice/data subscriber equipment for cellular telecommunication systems and related equipment.

Qualifications include BS degree in EE with 6+ years experience in 900 MHZ RF transmitter and receiver design at both system and circuit level. In-depth experience with analog and digital modulation and demodulation techniques and circuit implementation is required. Such experience includes system architectural design, component selection, hardware debugging and testing. Knowledge of cellular concepts and technologies (AMPS, GSM, TDMA, CDMA, RF Propagation, Antennas, delay spread, channel equalization, A/D conversion, AGC, and synchronization) are desirable.

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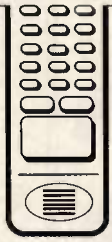

Chief Scientist: Responsible for design of analog and RF systems and circuits for consumer and commercial digital wireless products. Five to ten years experience in RF systems analysis and design. Experience with low-cost design techniques for frequency synthesizers, power amplifiers, up/down converters and baseband circuits for digital communications systems. Must be able to derive RF systems and module requirements to meet overall performance and cost goals. Familiarity with time division duplex or CDMA a plus..

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Systems Engineers: In-depth experience in one or more of the following areas is essential: RF propagation, modem and RF design, wireless protocols and algorithms, low-power, low-cost handheld consumer products. Participation in the development of baseband and RF IC's strongly desired. BSEE

Manager, Commercial Products: Project engineering and management for development of GaAs MMIC chips and MMIC-based products. Products include phase-locked 2.5 GHz television converter for high volume production, monolithic telephone converter and power amplifier, integrated amplifier/switch for data communications; voltage controlled oscillator and converter for satellite communications; Frequencies ranged from 800 MHz to Ku-band.

RF Design Engineer: Design and develop various RF circuits (including but not limited to PA's, LNA's, mixers and filters). Perform subsystem and system level testing. BSEE/MSEE 5-7 years RF circuit design experience with proven track record of developing RF subsystems from concept through manufacturing introduction.

RFIC Design: MS or PhD in Electrical Engineering with minimum 5 years related experience preferred. The candidate should have a good knowledge and experience in Linear Bipolar High Frequency IC design and measurement techniques in design IC's like Amplifiers, Mixers, Oscillators, VCO's, Prescalers, Synthesizers, Limiting Amplifiers, etc. operating up to 2 GHz in Bipolar or BiCMOS technologies.

MMIC Design Engineer: Develop L/S band GaAs MMIC power amplifiers for commercial wireless communications. Requires: M.S. or BSEE, +2 years experience with GaAs MMIC design, simulation, packaging and test.

Filter Design Engineer: B.S. Minimum 3 years experience in the design and development of Broad Band, comb-line strip line, interdigital, low pass and high pass filters, multiplexers, diode switches, (phase shifters), attenuators and microwave sub-systems desirable.

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RF Design Engineer

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We are seeking engineers experienced in linear high frequency integrated circuit design, who have applied their skills to the development of low-noise amplifiers, mixers or oscillators in silicon bipolar or BiCMOS technologies. Dept. RFD/RFIC

DSP Engineer

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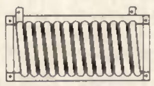

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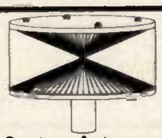

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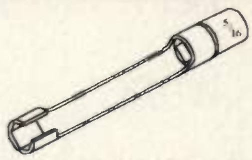

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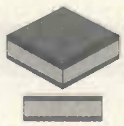

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RF software

Circuit Analysis for Macs

DragonWave 7.0 from Nedrud Data Systems is a circuit analysis program which runs on Macintosh computers. DragonWave features intuitive schematic entry, 1 - 99 external ports, full nodal noise calculation, variables and equations in substrate or element parameters, sweep of any variable, output equations, and optimization. Matrix reduction is used for increased speed. Reports contain text, pictures, schematics, tables and plots. The single copy price for DragonWave 7.0 is $3790.

Nedrud Data Systems
INFO/CARD #220

3-D EM Calculations

HP EEsof has announced HP 85180A HFSS 3.0, a high-frequency structure simulator that computes S-parameters for passive 3-D structures. The enhanced simulator performs complete electromagnetic solutions, but users only need minimal background in electromagnetic field theory or finite-element analysis to operate it. The new version of the structure-simulator offers a fast frequency sweep mode, absorbing boundary conditions, and numerous user interface enhancements. HFSS Release 3.0 is priced at $41,800, and is offered as a free upgrade to supported customers.

HP EEsof
INFO/CARD #221

Math Modules for Programmers

Waterloo Maple Software has announced the release of MathEdge, a suite of component software modules that allows application developers to integrate powerful symbolic and graphic computation into new products. The MathEdge "engine" is derived from Waterloo Maple Software's Maple V® product. MathEdge for Microsoft Windows is available for $1995, and includes a developer's Guide and a copy of Maple V. Versions for SUN OS and Solaris will

Waterloo Maple Software
INFO/CARD #222

Q Factor Measurement

Two programs that calculate Q from data taken by vector or scalar network analyzers has been released by Vector Fields. QZERO utilizes the data from an automatic vector network analyzer to calculate loaded Q, coupling coefficient, unloaded Q and their standard deviations. SCALARQ performs similar calculation with data from a scalar network analyzer. The theory and examples of use are described in the accompanying 180-page book. The diskette and book are available for $69.00.

Vector Fields
INFO/CARD #223

Noise Modeling in FM Systems

FM is a software product to help the system design engineer get the best from wide or narrow band analog FM and GMSK systems. Transmission link and oscillator noise are modeled in detail, giving an accurate simulation of baseband noise distribution. FM deviation, emphasis and noise weighting are also included within the simulation. FM requires an IBM compatible with 640k and DOS 3.1 or better.

Phasor Design
INFO/CARD #224

RF Propagation

The HTZ computer-aided design software package from the French company ATDI is for planning and managing radio networks in the VHF/UHF/SHF bands. This latest version of HTZ integrates multilayered geographic information including altimetric data, reference images, ground occupancy, and building heights. HTZ runs on 486-based PCs with DOS 6.0 or later and a minimum hardware configuration of 16 MB RAM, 400 MB hard disk, and a multi-sync 1024x768 pixel, 256 color screen.

ATDI
INFO/CARD #225

Spectrum Sharing Software

Comsearch has announced their newest release of Spectrum Sharing software (Release 2.0). The new release enables engineers to create automatic cell layouts based on demographics, RF performance parameters, and traffic, including an option for sectorization.

Comsearch
INFO/CARD #226

Dynamic System Simulator

Running under Microsoft Windows, SystemView is a high level conceptual design and analysis engine embedded in intuitive, friendly and completely visual design environment. Its applications include DSP systems, signal processing, communications, control, and general mathematical modeling. SystemView supports mixed mode (ie. analog and digital) multi-rate systems, parallel simultaneous systems, and internal or external data sources and sinks. Optional libraries for communications, DSP and logic are available. SystemView is priced at $2450.

Elanix, Inc.
INFO/CARD #227

Multi-Processor Spice

Intusoft is developing a multi-threaded, multi-processor version of the popular SPICE analog circuit simulation program. When complete the new simulator, called IsSpice5x, will be able to simulate circuits with a linear increase in speed per additional CPU. Initially, Windows NT platforms will be targeted, but multiprocessor versions of Windows95 are also expected to be able to run the new software. The software will be available in the first half of 1995, as will pricing information.

Intusoft
INFO/CARD #228

RF literature

Oscillator Catalog

A 12-page catalog from Wenzel Associates lists their lines of crystal oscillators, which include, low cost, low profile and low noise models. Also listed are frequency standards, including an ultra-low noise standard with phase noise of –135 dBc/Hz 10 Hz from a 1 MHz carrier, and –177 dBc/Hz 20 kHz from a 5 MHz carrier. Other related products are also listed. Also included is general information about crystal oscillators, ordering information, and a list of sales representatives.
Wenzel Associates, Inc.
INFO/CARD #229

Power Amplifier Catalog

Matcom announces a catalog of solid state power amplifiers. These amplifiers, which are manufactured by R&K Laboratories, cover various frequency ranges between 5 MHz and 3 GHz with powers up to 1 kW CW and 2 kW pulsed. They are available in modular or rack mounted form with optional power supplies.
Matcom, Inc.
INFO/CARD #230

Technical Data Sheets

Delta Microwave offers a series of technical data sheets on their latest RF products including: bandpass filters, diplexers and multiplexers. These devices are engineered in either coaxial waveguide or lumped-component configurations. Data sheets cover specific components ranging from ultra-low loss waveguide diplexers to ultra-high rel space-qualified filters. Data on filters/amplifiers, gain equalizers, iso-filters, and integrated assemblies are also available.
Delta Microwave
INFO/CARD #231

Test & Measurement Catalog

LeCroy Corporation, celebrating their 30th year, announces the publication of the 1994 Test & Measurement Catalog. It features product information, selected applications information and a "Fundamentals of Digital Oscilloscopes" tutorial. Copies are available through LeCroy's Customer Care Center at 1-800-4LeCroy (1-800-453-2769)
LeCroy Corp.
INFO/CARD #232

Network Analyzer Accuracy Brochure

Wiltron, a member of the Anritsu Wiltron Measurement Group, announces the release of a free technical brochure identifying measurement inaccuracy sources and optimum network analyzer configuration for complete test systems covering RF, microwave and mm-wave applications.
Wiltron
INFO/CARD #233

Semiconductor Catalog

Loral Microwave - FSI has published a short-form catalog titled Semiconductor Products for the Nineties. FSI covers its new PIN diodes, tuning varactors, and Schottky devices in this new catalog. Complete specifications, including outline drawings and performance charts are provided for more than 65 semiconductor catalog products.
Loral Microwave - FSI
INFO/CARD #234

Components Catalog

Mouser Electronics announces the publication of their latest industrial electronic components catalog. The 276-page catalog contains over 45,000 in-stock, factory authorized product selections from more than 100 quality concious manufacturers.
Mouser Electronics
INFO/CARD #235

Power Transistor Catalog

A catalog from Ericsson describes RF power transistors for 450 MHz to 2000 MHz. Rugged bi-polar transistors are rated to 150 watts. Ion implantation, nitride surface passivation and gold metalization are used to insure excelent uniformity and reliability. Special emphasis is placed on devices for cellular base stations and broadcast TV.
Ericsson Components Inc.
INFO/CARD #236

Single-Supply Op Amps

Analog Devices has published a six-page single-supply operational amplifier selection guide, including key specifications and technical information for over 30 products. Six products are given special attention. They include Analog Devices' newest single, dual and triple op amps with unmatched video specifications, as well as ultralow-power dual and quad rail-to-rail ICs.
Analog Devices
INFO/CARD #237

Filter in Cable Data

K&L microwave is now offering a two-page data sheet describing "Filters in a Cable". Specs for lowpass, highpass, and bandpass filters are given as well as environmental testing information.
K&L Microwave, Inc.
INFO/CARD #238

Software Catalog

The 38th edition of the Dynacomp software catalog conatains over 220 new products. The 228-page catalog contains over 700 products, including software for electrical engineering, data analysis, and graphing.
Dynacomp, Inc.
INFO/CARD #239

Designer's Guide

The latest edition of Mini-Circuits' 128-page Designer's Guide is now available, featuring comprehensive product listings and specifications for over 1000 standard RF/IF and microwave components.
Mini-Circuits
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RF guide to editorial coverage

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RF advertising index

ADVERTISER	PAGE #	READER SVC #
AAT Communications Corp.	109	98
A-Comm Electronics	113	121
Aerotron Repco Sales Inc	116	71
Alpha Components Inc.	112	114
American Tech Ceramics	51	34
Amplifier Research	104	67
Analog & RF Models	113	120
Analog Devices, Inc.	22-23	19
Andersen Laboratories	130	79
Aydin Vector Division	80	48
Brian Robinson	113	122
Broadcast Microwave	117	73
C-Mac Quartz Crystals	45	31
California Eastern Labs	65	43
CCI/USA	106	86
Comatlas S.A.	84	52
Comms People, Inc.	109	101
Communication Concepts	10	7
Compact Software	46-47	32
Comtronix Systems, Inc.	112	116
D.C. to Light	28	23
Daico Industries	6	4
Delphi Components, Inc.	111	110
Digcom	35-36	27
Don Gallagher & Assoc.	109	100
Eagle	88	55
Eagleware	59	39
Eesof, Inc./Hewlett Packard	17	13
EIP Microwaves	20-21	18
EMC Consulting, Inc	113	123
EMF Systems	94	60
Fortune Communications Group	108	97
Fortune Personnel Consultants of Raleigh, Inc	92	90
General Instrument	109	103
Giga-tronics Inc.	14-15	11
Hamilton Hallmark/Avnet	40-41	29
Harris Corp	110	105
Henry Radio	60	40
Hy-Q International	29, 106	24, 88
Hybrids International, Ltd.	106	85, 89
Integrated Component Systems	16	12
IFR Systems, Inc.	3	2
Int'l Crystal Mfg.	85, 123	53, 77
Isotemp Research, Inc	112	119
ITT GTC	39	28
Jan Crystals	28	22
JFW Industries	105, 124	69, 80, 81
Jim Young & Associates	108	95
Kalmus Engineering	9	6
Kay Elemetrics	123	76
KCR Products	112	113
Kintronics Labs, Inc.	111	111
KR Electronics	104	68
KVG North America	10	8
KS Electronics	112	115
Lap-Tech	67	44
LCF Enterprises	88, 106	56, 90
Locus Incorporated	112	112
Loral Microwave	2	1
M/A Com (3 divisions)	11	9
Management Recruiters of Boulder, Inc.	108	93
Maxim Integrated Product	30-31	25
Merrimac	13	10
Micro Communications Executive Search	109	103
Micronetics	67	45
Microunity Systems Engineering, Inc.	111	108
Mini Circuits	4-5, 53, 63, 103	3, 26, 42, 66
Miteq	99	63
Motorola	109	102
MTI Milliren Technologies, Inc.	106	87
Murata Electronics North America, Inc.	81	49
NEC America, Inc.	109	99
Noble Publishing	68A	
Noise/Com Inc.	18-19	14, 15, 16, 17
Nolan Laboratories, Inc.	111	109
Oscillatek	87	54
Penstock, Inc.	79	47
Piezo Technology	55	36
Programmed Test Soucrces	70	46
Q-Bit Corp	25	20
Ramsey Electronics	113	117
Reeves-Hoffman	100	64
Richardson Electronics	27, 43, 101	21, 30, 65
Sanford Associates	108	96
Silicon Valley Power Amps	120	74
Sprague Goodman Electronics	57	37, 38
Storm Products	96	61
Surcom Associates	116	72
Synergy Microwave	83	51
T-Tech	93	59
Tecdia	112	118
Tele-Tech Search	108	92
Temex	48	33
Tesoft, Inc.	98	62
Trak Microwave	91	57
Trilithic	107	70
TTE, Inc.	93	58
Vari-L Company, Inc	33	35
Vectron	61	41
Waypoint Software	108	94
Werlatone, Inc.	8	5
Wide Band Engineering	82, 121	50, 75
Z-Domain Technologies, Inc	110	106

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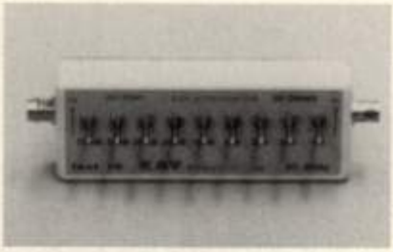

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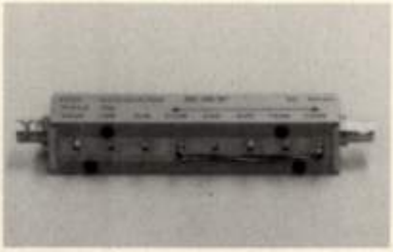

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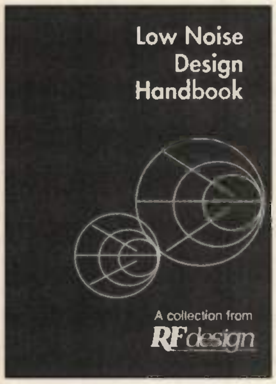

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