RF design™

engineering principles and practices

May 1994

Featured Technology
Designing With Microstrip
Plus-
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Amplifier And Filter Design

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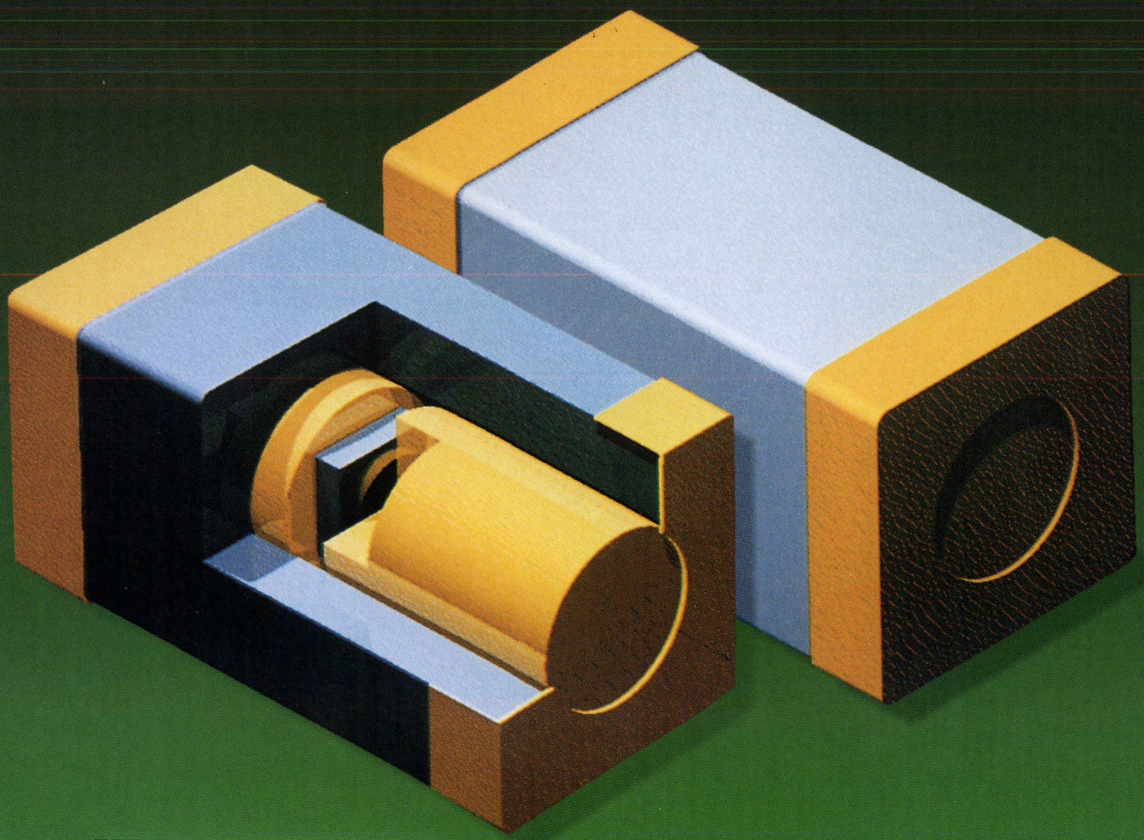
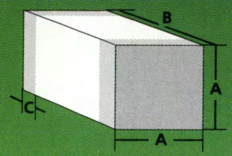
Part No.	Suggested Case Style	Reverse Voltage Rating (Vr) – min	Total Capacitance (pF) – max F=1MHz Vr=50v		Series Resistance (RS)–ohms F=100 MHz		Carrier Lifetime (Tl) µsec (typ)	Thermal Resistance (Θj)
		Ir< 10µA	M1	M2	If=100mA (max)	If=200mA (typ)	If=10mA	C/W
SM0502	M1 or M2	500	0.45	0.50	0.70	0.55	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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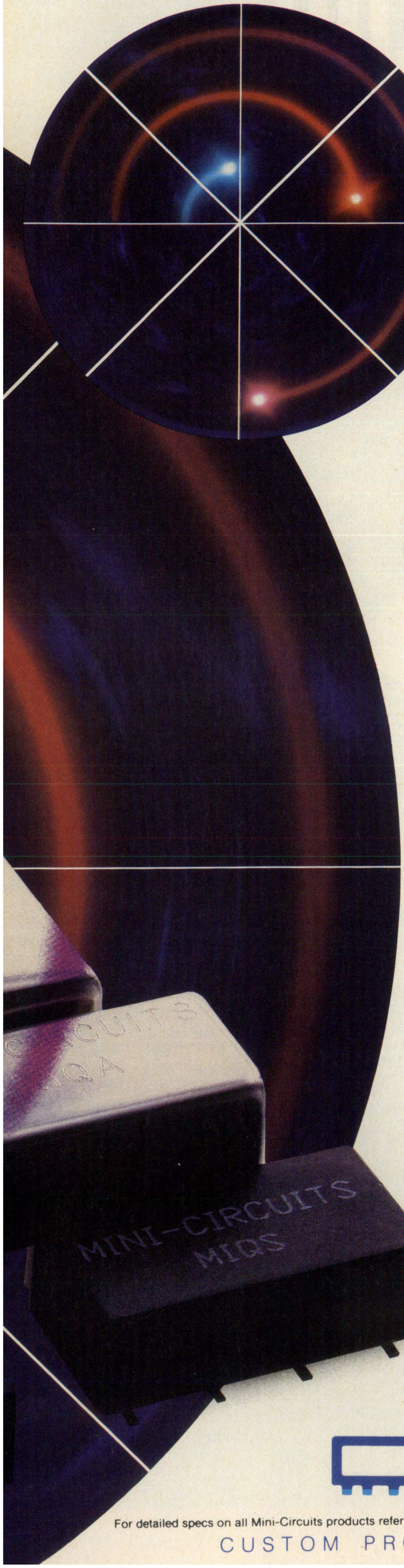

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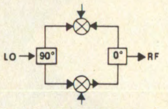
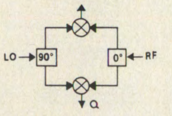


I/Q MODULATORS

MODEL NO.	FREQ. (MHz) f_L	f_U	CONV. LOSS (dB) x	σ	CARRIER REJ. (dBc) Typ.	SIDEBAND REJ. (dBc) Typ.	HARM. SUPPRESS (dBc) Typ. 3xI/Q	5xI/Q	PRICE $ QTY (1-9)
MIQA-10M	9	11	5.8	0.20	41	40	58	68	49.95
MIQA-21M	20	23	6.2	0.14	50	40	48	65	39.95
MIQA-70M	66	73	6.2	0.10	38	38	48	58	39.95
MIQA-70ML	66	73	5.7	0.10	38	38	48	58	49.95
MIQA-91M	86	95	5.5	0.10	38	38	48	58	49.95
MIQA-100M	95	105	5.5	0.10	38	38	48	58	49.95
MIQA-108M	103	113	5.5	0.10	38	38	48	58	49.95
MIQA-195M	185	205	5.6	0.10	38	38	48	58	49.95
MIQC-88M	52	88	5.7	0.10	41	34	52	66	49.95
MIQC-176M	104	176	5.5	0.10	38	36	47	70	54.95
MIQC-895M	868	895	8.0	0.10	40	40	52	58	99.95
MIQC-1785M	1710	1785	9.0	0.30	35	35	40	65	99.95
MIQC-1880M	1805	1880	9.0	0.30	35	35	40	65	99.95
☐ MIQY-70M	67	73	5.8	0.20	40	36	47	60	19.95
☐ MIQY-140M	137	143	5.8	0.20	34	36	45	60	19.95



I/Q DEMODULATORS

MODEL NO	FREQ. (MHz) f_L	f_U	CONV. LOSS (dB) x	σ	AMP. UNBAL. (dB) Typ.	PHASE UNBAL. (Deg.) Typ.	HARM. SUPPRESS (dBc) Typ. 3xI/Q	5xI/Q	PRICE $ QTY (1-9)
MIQA-10D	9	11	6.0	0.10	0.15	1.0	50	65	49.95
MIQA-21D	20	23	6.1	0.15	0.15	0.7	64	67	49.95
MIQC-895D	868	895	8.0	0.20	0.15	1.5	40	55	99.95
☐ MIQY-1.25D	1.15	1.35	5.0	0.10	0.15	1.0	59	67	29.95
☐ MIQY-70D	67	73	5.5	0.25	0.10	0.5	52	66	19.95
☐ MIQY-140D	137	143	5.5	0.25	0.10	0.5	47	70	19.95

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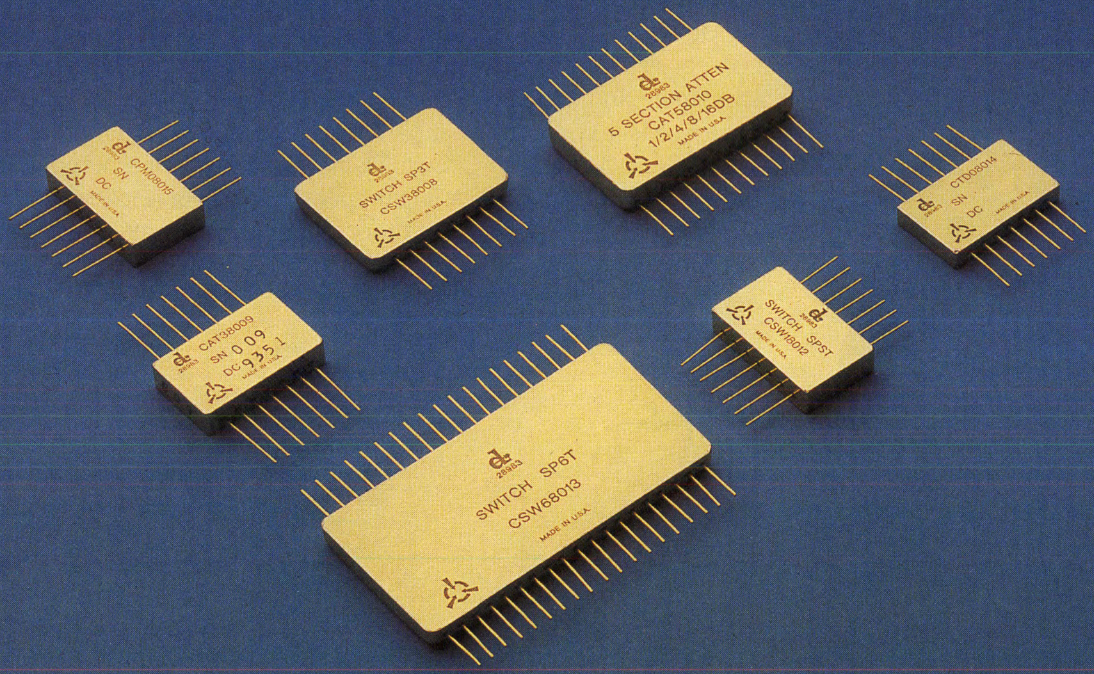

Flatpacks

Config	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

Daico broadens it's product line with a proven series of Flatpacks and almost every part in our catalog can be manufactured in one of these packages! Please call us...we have creative ways to meet all of your price, quality and performance requirements.

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RF design

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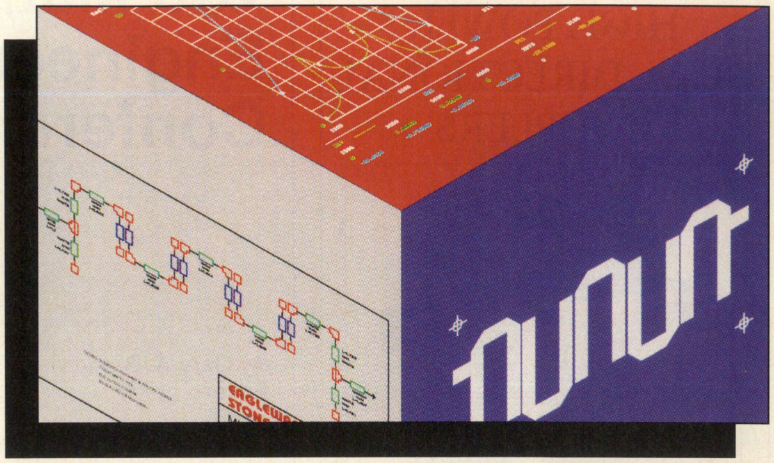
May 1994



featured technology

34 Combined Technology Amplifier Design

Microstrip transmission line stubs are replaced by capacitors in this amplifier design, reducing size, and making designs more flexible.

— Stanley Novak

42 Simulating Coupled Transmission Lines with Super-Spice

Lossy multiple-conductor transmission lines and their various discontinuities are accurately modeled by Compact Software's Super-Spice. A number of examples demonstrating the simulation accuracy are presented.

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Power transfer efficiency, taking into account finite inductor Q, is maximized in this nethod for designing narrowband matching circuits. *— Mihai Albulet*

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This article is an entry-level introduction to the characteristics of frequency hopping spread spectrum, and the methods used for its implementation

— Gary A. Breed

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— Frederick H. Raab, Ph.D. and Daniel J. Rupp

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The program presented in this article calculates the elements of equiripple Chebyshev lowpass filters using special mapping functions.

— Ljubomir Urshev and Antoaneta Stoeva

78 RF Amplifier and Oscillator Design Using the UCFCAD Tools.

The procedures used to design several types of linear RF amplifiers and oscillators using a design tool developed at the University of Central Florida is presented. *— Michael Rothery, Sam Richie, and Madjid Belkerdid*

The cover photo on this issue of RF Design *was provided courtesy of Eagleware Corporation, representing the Featured Technology subject for May: Microstrip Design Techniques. The photo illustrates their M/FILTER program's design capability for microstrip filters.*

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RF editorial

Are There Too Many Engineering Conferences?

By Gary A. Breed
Editor

The recent RF Expo West in San Jose drew nearly 2000 engineers to attend classes, hear technical presentations and look over RF products. This admittedly moderate level of success was achieved despite the fact that the Northern California area was saturated this spring with conferences whose titles contained such hot buzz words as mobile data, wireless, and portable.

In the past three years, at least five brand new trade shows or technical conferences have been started, to address the growing level of engineering and product development in RF-based wireless communications systems. Several other events have added the words wireless or portable, in order to cash in on the marketing frenzy surrounding those terms. I have heard quite a few suggestions that there are now too many shows and conferences.

Well, the rush to capitalize on the rapid growth of the RF market may be slightly premature. This will be a very big market for RF components and systems, but until it achieves the promised growth, there won't be enough new engineers to fill up every one of these conferences.

Having said that, I still don't think there are too many! I have attended some of the new conferences, and know most of the established ones. Each one has its place — some are practically-oriented, others are more theoretical; some emphasize traditional academics, others feature up-and-coming researchers; some are directed toward management, others are aimed at engineers; some are narrowly focused, others have a broad approach. If there is unnecessary duplication, the "marketplace" will surely make its decision about which ones are worth supporting.

In a very short time (one or two years at the most), we will know just how big the RF market is going to become. In this period, many engineers will discover RF, either as a specialty or as part of a team working on some RF-based system. These engineers need conferences and trade exhibitions to speed the learning process.

Which brings the discussion back to the RF Expos. Although we have a long and successful track record, we are seeing the effects of increased competition. To combat this, we have redoubled our efforts to identify their value. To help you decide to attend, and to help you justify that choice to your boss, we will provide far more advance information on future shows, starting with RF Expo East, November 15-17 in Orlando.

You will get a syllabus for each of the short courses we offer in the first informational mailing. Then we will publish paper abstracts and speaker biographies for the scheduled technical presentations. We will provide ongoing updates of the companies who will be exhibiting their products. We will be providing all the good things that RF Expo is known for, but with a renewed effort to do them in the most relevant and useful manner.

In the meantime, we will be participating in other industry forums. Look for me and my RF Design staff colleagues at the MTT-S Symposium, the Frequency Control Symposium, the IEEE EMC Symposium, the Piezoelectric Devices Conference, and other important conferences and trade shows. Like I said before, I don't think there are too many of them, but don't ask my boss' opinion when he reviews my travel budget!.

RF design
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GIGA-TRONICS 6080A & 6082A SYNTHESIZED RF SIGNAL GENERATORS

Giga-tronics Presents The Alternative To $30,000 And $40,000 RF Synthesizers.

Hewlett-Packard makes a couple of very good RF synthesizers. And if you can afford the luxury of paying $30,000 or $40,000 for the name, by all means, call HP right now. They'll be happy to take your order, and your money.

However, if you're looking for an RF synthesizer with outstanding performance and proven reliability for about half the price, you'd better call Giga-tronics.

Here's why:

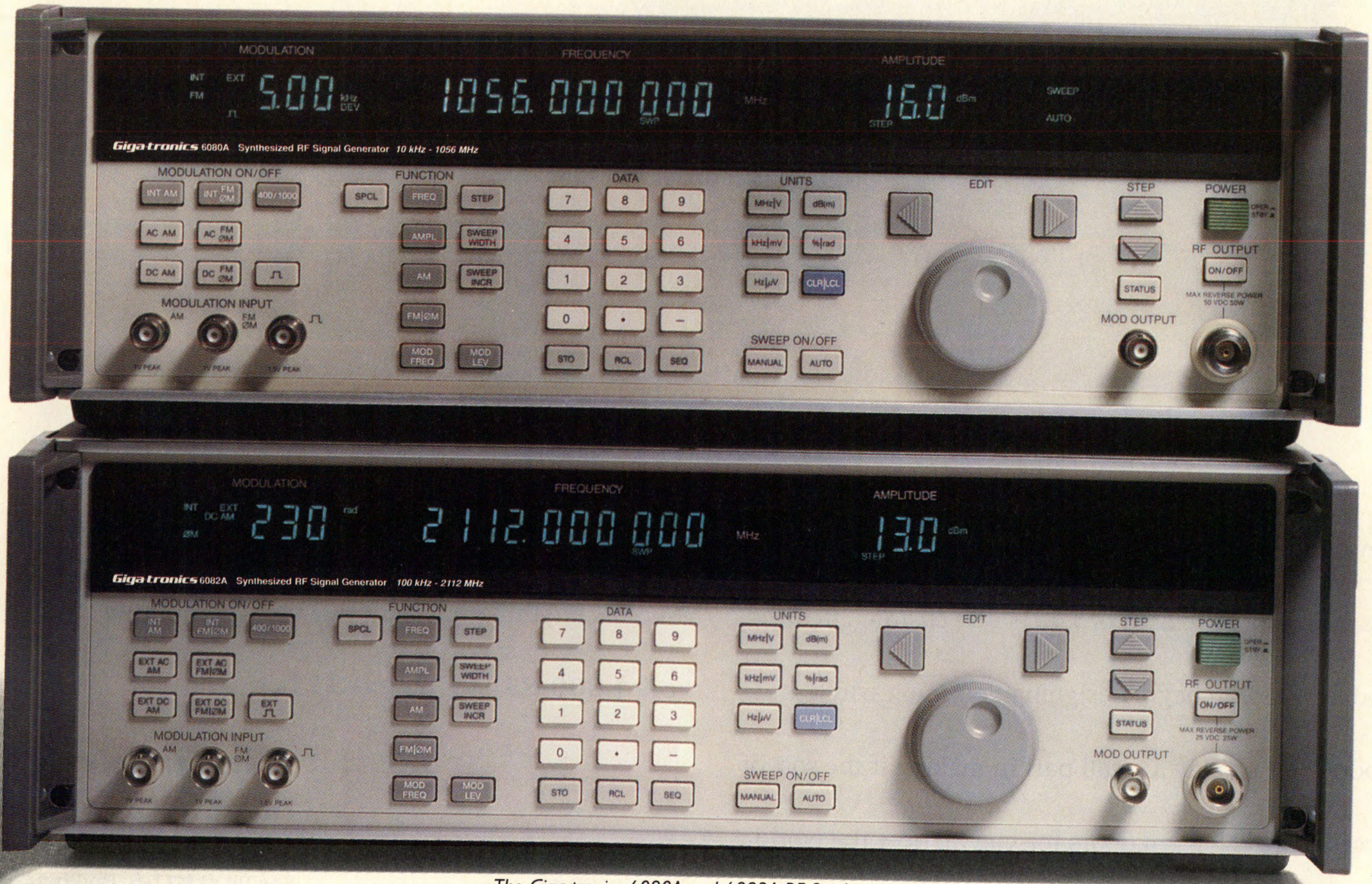
Performance.
Check the charts. In virtually every category, the Giga-tronics 6080A and 6082A RF Synthesizers meet or exceed the specs of the HP machines. And they use the same GPIB command set, for direct replacement without expensive new software.

Experience.
Granted, Hewlett-Packard has been around a long time. But, Giga-tronics is no Johnny-come-lately.

Giga-tronics has a 13-year history of building test and measurement gear for the most demanding requirements. We've shipped thousands of instruments for use in the testing of radar, EW and communications systems.

Reliability.
Making reliable RF synthesizers is usually no fluke.

However, in this case, it is.



The Giga-tronics 6080A and 6082A RF Synthesizers give you great performance and proven reliability for a lot less money.

Specifications	Hewlett-Packard HP 8642A	Giga-tronics 6080A	Hewlett-Packard HP 8642B	Giga-tronics 6082A
Frequency Range Switching speed	.1 to 1057 MHz <85 ms	.01 to 1056 MHz <100 ms	.1 to 2115 MHz <85 ms	.1 to 2112 MHz <100 ms
Spectral Purity* Spurious Subharmonics	<-100 dBc None	<-100 dBc None	<-94 dBc <-45 dBc	<-94 dBc <-45 dBc
Phase Noise* @ 20 kHz offset	<-134 dBc/Hz	<-131 dBc/Hz	<-125 dBc/Hz	<-125 dBc/Hz
Residual FM* (.3 to 3 kHz BW)	<2 Hz	<1.5 Hz	<5 Hz	<3 Hz
Output Range* Accuracy Reverse Power Protection	+16 to -140 dBm ±1 dB >-127 dBm 50 Watts/50 Vdc	+17 to -140 dBm ±1 dB >-127 dBm 50 Watts/50 Vdc	+16 to -140 dBm ±1 dB >-127 dBm 25 Watts/25 Vdc	+13 to -140 dBm ±1 dB >-127 dBm 25 Watts/25 Vdc
Amplitude Modulation Depth Distortion @ 30%	0–99.9% <2%	0–99.9% <1.5%	0–99.9% <2%	0–99.9% <1.5%
Frequency Modulation Max. Deviation* Distortion	3 MHz <2%	4 MHz <1% @ 50% Dev.	3 MHz <2%	8 MHz <1% @ 50% Dev.
Phase Modulation Max. Deviation*	100 Rad.	40/400 Rad.	200 Rad.	80/800 Rad.
Pulse Modulation On/off Rise/fall time Minimum Pulse Width	>40 dB <400 ns <2 μs	>40/60 dB <15 ns (Typ 7.5 ns) <30 ns	>40/80 dB <400 ns <2 μs	>80 dB <15 ns (Typ 7.5 ns) <30 ns
Internal Modulation Source Level Range Waveforms Programmable	20 Hz to 100 kHz 0 to 3 Vpk Sine Yes	0.1 Hz to 200 kHz 0 to 4 Vpk Sine/Sq/Tri/Pulse Yes	20 Hz to 100 kHz 0 to 3 Vpk Sine Yes	0.1 Hz to 200 kHz 0 to 4 Vpk Sine/Sq/Tri/Pulse Yes
Memory Locations (NVM)	51 Full Function	50 Full Function	51 Full Function	50 Full Function
U.S. List Price	$30,340	$16,950	$41,680	$22,950

The question is not why Giga-tronics is so much less,

but rather, why Hewlett-Packard wants so much more.

*Specifications for both the 6080A and the HP 8642A are at 1GHz. Specifications for both the 6082A and the HP 8642B are at 2GHz. Prices and specifications for the HP 8642A and HP 8642B are from the Hewlett-Packard 1993 catalog. Prices for the Giga-tronics 6080A and 6082A are U.S. list prices.

Both the 6080A and 6082A were originally introduced in 1990 by John Fluke Manufacturing Company. To date, thousands have performed flawlessly in the field.

For added confidence, the instruments incorporate self-testing, internal diagnostics and modular design for easy fault isolation and repair.

Service.

If a problem occurs, Giga-tronics technical support staff can often help you find and fix the problem over the phone.

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

But at Giga-tronics, customer service starts even before you become a customer.

Whether you're looking to buy one unit or one hundred, you'll get the same assistance, including a demonstration at your facility.

Price.

Considering all this, the real question is not why Giga-tronics is so much less, but rather, why Hewlett-Packard wants so much more?

So, if you're interested in paying a lot less for great performance and proven reliability, backed by a worldwide network of service and support, call us toll free at **800 726 GIGA** (4442). We'll send you more information and arrange for a demonstration.

Giga-tronics

Giga-tronics Incorporated
2495 Estand Way
Pleasant Hill, California 94523
Telephone 800 726 4442
Telefax 510 680 7736

INFO/CARD 11

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

Feedback Requested

Editor:

My article, "A Program for Design and Analysis of Receivers", ran in the October 1993 issue of RF Design. I am interested in obtaining feedback from those who have ordered the program. I am writing this letter in hopes that those subscribers who ordered the program will respond to me with suggestions on improvements and also inform me of how it is being used in their applications and its degree of usefulness. Any responses would be most appreciated. Please send all correspondence to: Naval Research Laboratory, 4555 Overlook Avenue Code 8131, Washington DC 20375, Attn: John Donohue

John Donohue
Naval Research Laboratory
(202) 574-7331

Biasing Error

Editor:

In Figure 1(a) of your tutorial on transistor biasing (June '93) there is an emitter resistor missing. It would be very unusual to use the voltage-divider-type biasing without this resistor, as the base voltage would essentially be "clamped" to about 0.7 volts by the base emitter diode. In fact, without the emitter resistor, the bias circuits of Fig. 1(a) and 1(b) operate very similarly.

A good rule of thumb is to make the ratio of the base-to-ground bias resistor to the emitter resistor somewhere between about 5 and 10. This gives fairly good independence of operating point from variations in transistor parameters and temperature. In critical cases, this ratio might be as low as 3.

Jack Streater
Mgr. Engineering
Mentor Radio Company

Errata

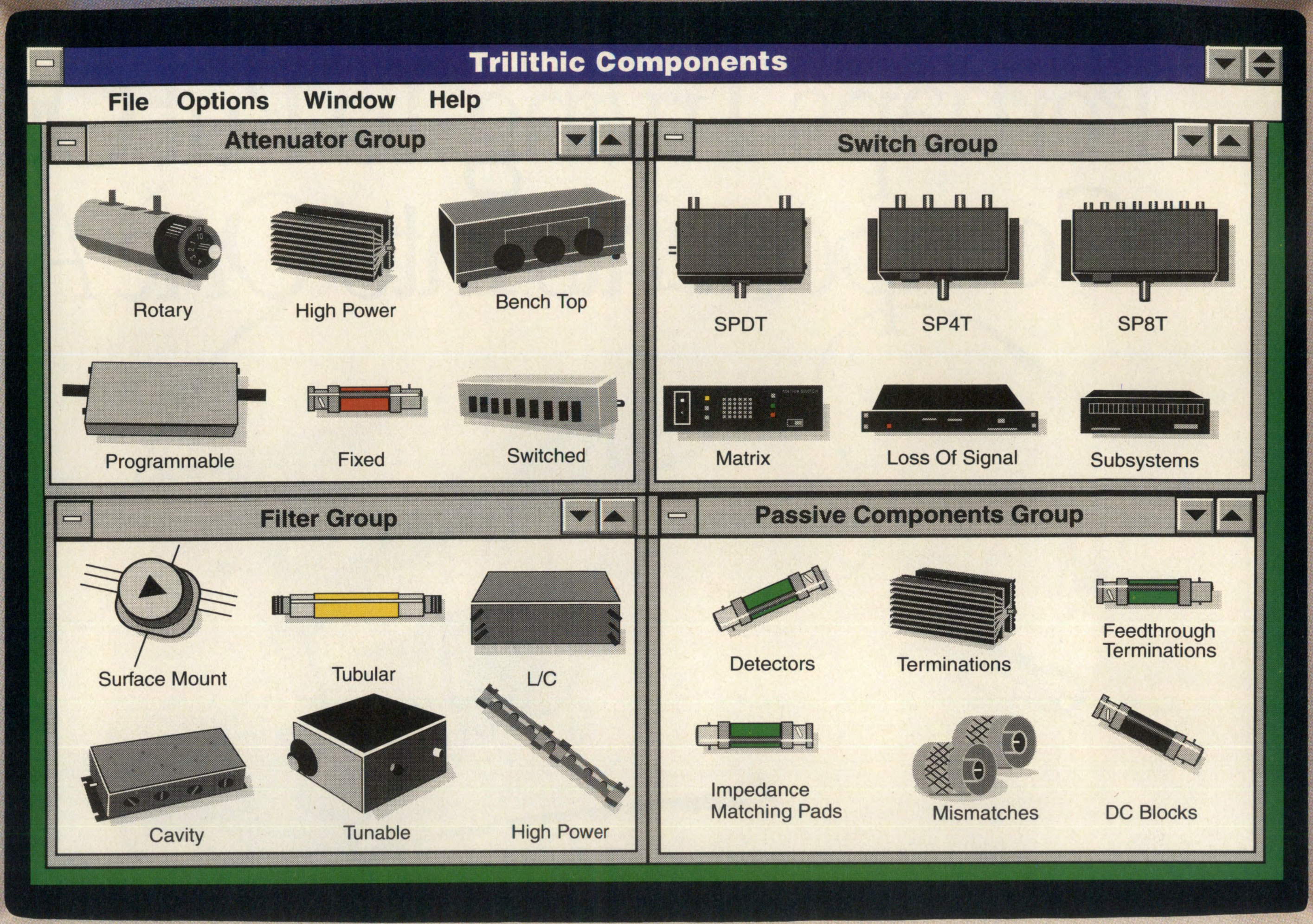
In Albert Klappenberger's March article, "Computer Design of Equal Shunt Value Tubular Bandpass Filters", the expression for W_l under equation 3 should read: $W_l = L(2\pi F_0)^2$, and the expression for C_k' in Figure 4 should be: $C_k' = C_k/n$.

Peter Hoberg was not identified as ECal Product Manager in March's "Industry Insight" column.

In the third paragraph of Carl Zatl's March article, "A Transformed Feedback Attenuator", V_{out} should be replaced with V_0. In the schematic of Figure 5, the inverting and non-inverting inputs to the amplifier should be switched, and the value of resistor R3 was omitted; it should be 30 ohms. In the "results" section, IP3 was measured at gains of 0 dB, –6 dB and –12 dB.

In Table 1 of Don McClure's February article, "Broadband Transmission Line Transformer Family Matches a Wide Range of Impedances", the entry for a 2/3 voltage ratio should read:

Voltage Ratio K	Connections	Minimum No. Lines
$\frac{2}{3}$	2S, 1 > P	3



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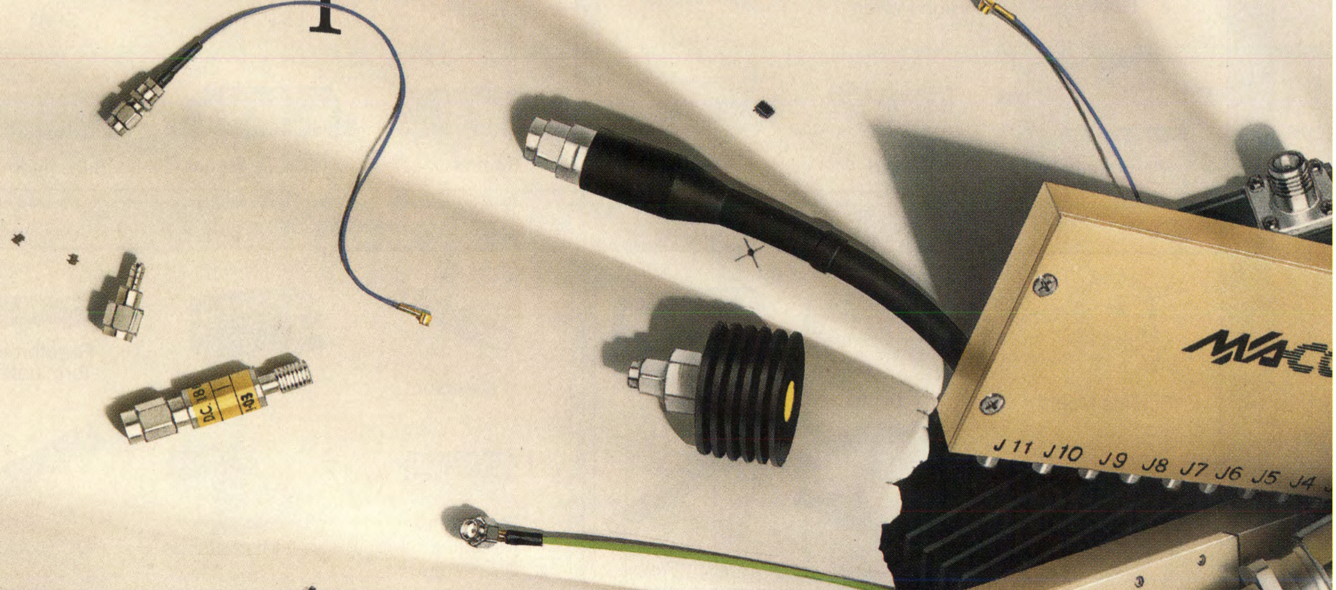

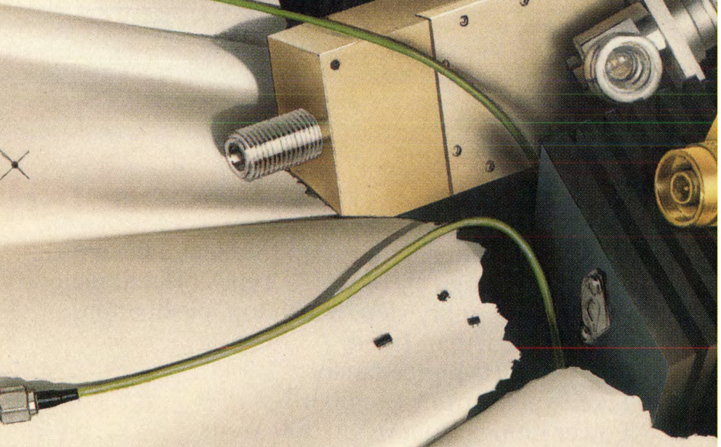

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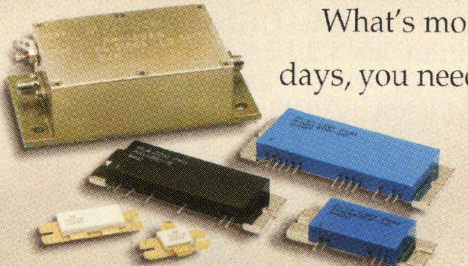

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Design and Processing of Power Hybrids
May 12-13, 1994, Los Angeles, Ca
Microwave/Millimeter-Wave Monolithic Integrated Circuits
May 17-20, 1994, Los Angeles, CA
Microwave/Millimeter-Wave Monolithic Integrated Circuits
May 17-20, 1994, Los Angeles, CA
Information: UCLA Extension, Engineering Short Courses, 10995 Le Conte Ave., Ste. 542, Los Angeles, CA 90024. Tel: (310) 825–1047. Fax: (310) 206–2815.

1994 Worldwide HP-EEsof High-Frequency Modeling and Simulation Seminar
June 6, 1994, Ft. Lauderdale, FL
June 8, 1994, Richardson, TX
June 10, 1994, Chicago, IL
June 13, 1994, Mt. View, CA
June 15, 1994, Los Angeles, CA
June 17, 1994, Phoenix, AZ
Information: HP-EEsof, 5601 Lindere Canyon road, Westlake Village, CA. Tel: (800) 343-3763.

Low Earth Orbit Satellite Systems (LEO's)
May 16-18, 1994, Washington, DC
Global Positioning System: Principles and Practice
June 7-10, 1994, Orlando, FL
Microwave Radio Systems
August 9-12, 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.

RF/MW Circuit Design: Linear/Non-Linear, Theory and Applications
June 6-10, 1994, United Kingdom
June 8-14, 1994, United Kingdom
Active and Passive RF Components: Measurements, Models, and Data Extraction
June 8-14, 1994, United Kingdom
Information: CEI-Europe/Elsevier, Mrs. Tina Persson. Tel: (46) 122–175–70. Fax: (46) 122–143–47.

Introduction to Radar Systems and Signal Processing
May 17-19, 1994, Washington, DC
Information: Research Associates of Syracuse, Incorporated, Hancock Army Complex, 510 Stewart Drive, N. Syracuse, NY 13212. Tel: (315) 455–7157.

Circuit Board Level Microwaves: Issues and Solutions
May 23-24, 1994, San Diego, CA
Information: Tom Laverghetta, TPL Associates, Inc., 516 E. First St., Auburn, IN 46706. Tel: (219) 925–1819.

RF Component Measurements, Models and Data Extraction
June 8-14, 1994, UK
RF/MW Circuit Design II
June 6-10, 1994, UK
Information: Besser Associates, 4600 El Camino Real, Suite 210, Los Altos, CA 94022. Tel: (415) 949–3300. Fax: (415) 949–4400.

Modern Antennas
May 16-18, 1994, Boulder, CO
Antenna Measurement Techniques
May 23-25, 1994, Boulder, CO
Phased Array Antenna Technology
June 13-15, 1994, Scottsdale, AZ
Information: Technology Service Corporation, 962 Wayne Avenue, Suite 800, Silver Spring, MD 20901. Tel: (301) 565–2970. Fax: (301) 565–0673.

DSP Without Tears
May 18-20, 1994, San Jose, CA
May 11-13, 1994, Richardson, TX
May 18-20, 1994, San Jose, CA
June 20-22, 1994, Boston, MA
July 20-22, 1994, Salt Lake City, UT
Information: Z Domain Technologies, Inc., 325 Pine Isle Court, Alpharetta, GA 30202. Tel: (800) 967–5034, (404) 664–6738. Fax: (404) 442–1210.

Component Technology
May 16-17, 1994, Washington, DC
Reliability Predictions
May 18, 1994, Washington, DC
Information: Sav-Soft Products, P.O. Box 360974, Milpitas, CA 95036. Tel: (408) 263–9150.

Electromagnetic Fields - An Overview
May 25, 1994, Detroit, MI
June 15, 1994, Atlanta, GA
Power Quality: Problems, Analysis & Solutions
May 17-18, 1994, Detroit, MI
June 7-9, 1994, Atlanta, GA
Information: PowerCET Corporation, 2700 Augustine Drive, Suite 178, Santa Clara, 95054. Tel: (408) 988–1346. Fax: (408) 988–4869.

Fundamentals of Data Communications & Local Area Networks
May 16-17, 1994, Grand Rapids, MI
May 17-18, 1994, Seattle, WA
May 18-19, 1994, Orlando, FL
May 25-26, 1994, Philadelphia, PA
Understanding Data Communications
May 16-17, 1994, Chicago, IL
May 19-20, 1994, New Orleans, LA
May 23-24, 1994, St. Louis, MO
Understanding LANS
May 16-17, 1994, Anchorage, AK
May 16-17, 1994, Chicago, IL
May 19-20, 1994, Washington, DC
Fundamentals of Telecommunications
May 16-17, 1994, Tampa, FL
May 23-24, 1994, Atlanta, GA
May 25-26, 1994, Chicago, IL
Information: Technology Interchange Group, Inc., Dept. 43, P.O. Box 2107, Clifton, NJ 07015. Tel: (201) 478–5400. Fax: (201) 478–4418.

Measured Equation of Invariance
May 19-21, 1994, San Francisco, CA
Information: MEI Institute, P.O. Box 679, Berkeley, CA 94701. Fax: (510) 254–7639.

RF calendar

May

23-24 **Microwave and Millimeter-Wave Monolithic Circuits Symposium**
San Diego, CA
Information: Richard B. Gold, Pacific Monolithics, 245 Santa Ana Court, Sunnyale,CA 94086–4512. Tel: (408)732–8000. Fax: (408) 732–3413.

23-27 **IEEE/MTT-S International Microwave Symposium**
San Diego, CA
Information: MTT-S Symposium 1994, c/o LRW Associates, 1218 Balfour Drive, Arnold, MD 21012.

27 **Automatic RF Techniques Testing 43rd Conference**
San Diego, CA
Information: Bill Pastori, Maury Microwave Corp. Tel: (909)987–4715. Fax: (909) 987–1112.

May/June

30-2 **IEEE International Symposium of Circuits and Systems**
London, England
Information: Electronic Industries Association, EIA Components Group, 2001 Pennsylvania, Avenue N.W., Washington, DC 20006-1813. Tel: (202) 457-4930.

June

1-3 **1994 IEEE Frequency Control Symposium**
Boston, MA
Information: Michael Mirarchi or Barbara McGivney, Synergistic Management, Inc., 3100 Route 138, Wall Township, NJ 07719. Tel: (908) 280-2024.

1-3 **Virginia Tech Symposium on Wireless Personal Communications**
Blacksburg, VA
Information: Conference Registrar, Donaldson Brown Hotel and Conference Center, Virginia Tech, Blacksburg, VA 24061–0104. Tel: (703) 231–5182. Fax: (703) 231–3746.

19-24 **1994 IEEE AP-S International Symposium and URSI RadioScience Meeting**
Seattle, WA
Information: Jan Kvamme, Conference Manager, UW Engineering Professional Programs, 3201 Fremont Avenue North, Seattle, WA 98103. Tel: (206) 543-5539. Fax: (206) 543-2352.

July

4-7 **HF Radio Systems and Techniques**
York, UK
Information: HF '94 Secretariat, IEE Conference Services, Savoy Place, London WC2R 0BL, UK. Tel: 44-071-344-5478/5477. Fax: 44-071-497 3633.

August

25-28 **IEEE Electromagnetic Compatibility Symposium**
Chicago, IL
Information: Thomas Braxton, Vice-Chair, AT&T Bell Laboratories, Room 2B-217, 2000 N. Naperville Road, Naperville, IL 60566. Tel: (708) 979-1299. Fax: (708) 979-5755.

Boston to Host 1994 IEEE International Frequency Control Symposium

The 48th annual IEEE International Frequency Control Symposium will be held June 1, 2, and 3 at the Westin Hotel, Boston, Massachusetts, with tutorial sessions on May 31. The event is sponsored by the IEEE Ultrasonics, Ferroelectrics, and Frequency Control Society, with technical participation by personnel of the U.S. Army Research Laboratory at Ft. Monmouth, NJ. Topics at the international forum will include: fundamental properties of quartz and other piezoelectric crystals; processing of quartz and other piezoelectric materials; theory and design of piezoelectric resonators and filters (BAW and SAW); stable oscillators and synthesizers; atomic and molecular frequency standards; and frequency and time coordination and distribution. Topics for special sessions include: new piezoelectric materials, cryogenic and whispering gallery mode dielectric resonators, cryogenic hydrogen masers, RF- and laser-excited trapped ion standards, and 1/f noise processes in oscillators. Eight invited and 127 contributed papers are scheduled to be presented. Plenary session speakers will present "1/f Noise Universality in High-Technology Applications" and "Applications of Silicon Micromachining to Resonator Fabrication." Invited speakers will present papers on acoustic wave attenuation, piezoelectric resonator materials, radar frequency requirements, SAW sensors, SAW devices on lithium tetraborate, and limitations on the instabilities of quartz oscillators. For an advance program and for registration information, call or write: Barbara McGivney, Synergistic Management, Inc., 3100 Route 138, Wall Township, NJ 07719, USA. Tel: (908) 280-2024.

System Improves Permittivity Measurements – Recently, NIST researchers have developed and evaluated an improved coaxial air line system for measuring permittivity. The new system employs a 77 mm diameter coaxial air line instead of the conventional 7 mm line. The increased diameter line reduces the ratio of surface area to volume for the sample dielectrics inserted in the line. An air gap between the sample dielectric and center conductor is inevitable, but the improved proportions minimize the effect of the air gap. For example, a commercial ceramic that had nominal real permittivity of 270 (at 50 to 1000 MHz) was measured at 275 after air-gap correction. The researchers hope to extend the method to the low-frequency range of 0.1 MHz to 50 MHz. For technical information, contact Claude Weil, Div. 813.08 NIST, Boulder, CO 80303-3328. Tel: (303) 497-3198.

Testing Lab Agreement Renewed – NIST and the Standards Council of Canada renewed an agreement that provides mutual recognition of testing laboratories located within the territorial U.S. that are accredited by the NIST National Voluntary Laboratory Accreditation Program (NVLAP), and for testing laboratories within Canada that are accredited by SCC's Program for Accreditation of Laboratories, Canada (PALCAN). Both programs meet international standards under ISO/IEC Guides 25 and 58. Officials responsible for administering each system have participated in assessment visits to testing laboratories accredited under the other national program. NVLAP accredits 900 labs in 16 fields of testing, and the Canadian PALCAN program has accredited 150 labs in 14 areas. For PALCAN information contact, Manager Testing Division, SCC, 1200-45 O'Connor St., Ottawa, Ontario, Canada K1P 6N7. (613) 238-3222. Fax: (613) 995-4564. Contact NVLAP at A162 TRF Bldg., NIST, Gaithersburg, MD 20899-0001, (301) 975-4016. Fax: (301) 926-2884.

ITU Reports 1992 Boom Market – The size of the global telecommunication market in 1992 was estimated at $535 billion according to a report released at the World Telecommunication Development Conference in Buenos Aires this March. In 1992, revenues for services increased 8% from the previous year and equipment revenues rose by 9%. The 24 industrial nations of the Organization for Economic Co-operation and Development (OECD) contributed 85% of global telecommunications service revenues although they are home to only 16% of the population. High income countries with only 15% of the population have 71% of the world's telephone lines. The report looks at the successes of selected countries under different organizational set-ups and various stages of economic development.

MCM's Software Tools Contract – The Advanced Research Projects Agency (ARPA) awarded a two-year $2.8 million contract to Tanner Research Inc. to develop affordable software design tools for multi-chip modules (MCMs). Funded by ARPA's Application Specific Electronic Module (ASEM) initiative, it calls for the development of a suite of cost reduced MCM-specific tools for place-and-route, signal integrity, design partitioning, testability, and thermal analysis.

$11 Million Technology Reinvestment – The Communication Electronics Technology Division (CET) of Watkins-Johnson Company, along with Array Comm, Inc. and Spectrian, Inc., have received a research grant in a joint project to develop a Spatial Division Multiple Access (SDMA) digital base station for the Personal Communications Services (PCS) wireless market. The system will address both commercial and military requirements with advanced technologies such as SMART antenna systems, digitizers and FET power amplifiers.

Command Post Aircraft – Boeing Defense and Space Group has given Scientific-Atlanta a $2.2 million contract for test equipment in support of Boeing's work on the E-4B National Emergency Command Post Aircraft. Tested will be the triband radomes that are used with satellite communications equipment carried aboard the modified 747 aircraft. The test system will assure the radome is performing properly by characterizing the unit before it is installed on the aircraft. The triband radome allows the Milstar Extremely High Frequency communications system, a jam resistant satellite satellite communications capability for the strategic and tactical forces of all U.S. military services, to become operational.

HTS Wire Research – A two year, $2 million extended development program between American Superconductor Corporation (ASC) and Inco Alloys International (IAI) continues IAI's funding for ASC's accelerated production of specialized "metallic precursor" powders. These powders are the basis of one of the major methods of producing high temperature superconducting (HTS) wires. The use of HTS wires not only

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UPC1659 600MHz to 2300MHz 23dB G_P	**UPC1675** To 2100MHz 12dB G_P	**UPC1676** To 1300MHz 20dB G_P 4.0dB NF	**UPC1677** To 1700MHz 24dB G_P P_{out} = 19.5dBm	**UPC1678** Up to 1900MHz 23dB G_P P_{out} = 18dBm
UPC1688 Up to 1000MHz 21dB G_P 4.0dB NF				

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Freq. Convertors – From $1.58 each

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Transistor Arrays From $2.40 each

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UPA103 F_T=9GHz Differental Amplifier	**UPA104** F_T=9GHz OR/NOR Functions

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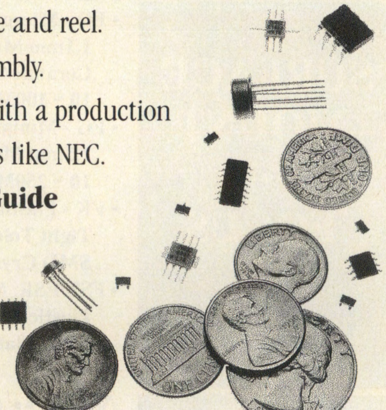

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"Window Curtain" Antenna – An antenna printed on both sides of a flexible fiber-reinforced Teflon sheet can be hung in a window, from a wall, or placed on a roof. Developed at the Georgia Institute of Technology, the antenna can be camouflaged on the flat surfaces of buildings for satellite communications, TV, cellular and links. Mounted on a foam insulating board atop a conducting sheet (to prevent radiation from its backside), the antenna weighs just a few pounds and averages 18 inches wide, 30 inches long and 1 1/2 inches thick (at 2 GHz). When suitably mounted it can be designed to operate from 500 MHz to 4 GHz. The phasing of the dipoles on the antenna array determines the beam aiming direction. An automated software system for the design of these antennas turns user specification into instructions which can be sent electronically to a photolithography house so that within 48 hours an antenna in a mailing tube could be delivered. Forty curtain antennas have been made. Sponsored by the U. S. Department of Defense, Georgia Institute of Technology is continuing research on inexpensive materials that would allow the antenna to be installed in the walls of new structures under thin plastic, stucco or even brick.

Superconductor Base Station Filter Unveiled – Illinois Superconductor showed a prototype cellular base-station filter March 2nd at the CTIA convention in San Diego. The filter uses high-temperature superconductors to improve receiver sensitivity, minimize interference and increase the system's capacity. The filter technology reduces the interference of radio systems that are both physically and electrically close

ABB HAFO to Produce Space-Qualified Components – ASEA Brown Boveri subsidiary, ABB HAFO, will produce radiation-hardened ASIC chips based on Harris Semiconductor designs using its own proprietary SOS5 technology. The technology includes layering silicon onto an insulating substrate of sapphire. These will become standard components for satellites. The first round of products includes the Harris CD4000, HCS/HCTS, ACS/ACTS families and 65643/65647 memories.

More Motorola GSM Equipment for U.K.'s Cellnet – A fourth contract between Motorola's Cellular Infrastructure Group and U.K. cellular operator, Cellnet, brings to $98 million the total investment Cellnet has in Motorola's GSM equipment. Installation of 600 sites across southern Britain, including the area around London, will be completed during the first half of 1994. The new outdoor cellsite, the ExCell™, will be installed in several locations.

Harris, Nat. Semi. Make MCM Agreement – Harris Electronic Design Automation, Inc. (EDA) will become the prime supplier of multi-chip module (MCM) design systems for National Semiconductor. The partnership could exceed $3 million in products and services over the next three years. National

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NEW

	NF	Ga	P1dB	Freq
NE326	0.5	11.5	10.75	12

In chip and low cost ceramic and hermetic packages

"How about a chip that does it all? Good noise figures, good gain, moderate output power, and bandwidth enough for my 6 - 18 GHz amplifiers."

You've got it. Introducing the new Heterojunction version:

NEW

	NF	Ga	P1dB	Freq
NE332	0.8	10.5	12.5	12

In chip and low cost ceramic or hermetic packages

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

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

	NF	Ga	P1dB	Freq
NE760	0.8	15.5	14.5	4

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

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

Meet the ideal second stage device:

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	NF	Ga	P1dB	Freq
NE424	0.8	11.5	11.0	12

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

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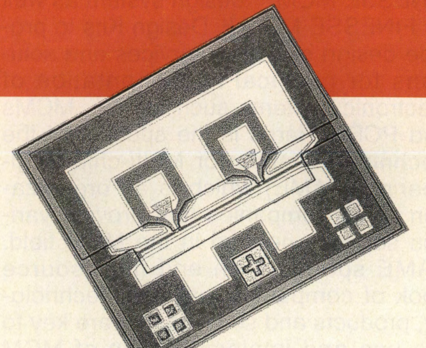
"I need a <u>low</u> noise FET for the LNA in my TVRO system. And, of course, I need it in a high volume, tape and reel package."

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

	NF	Ga	P1dB	Freq
NE324	0.6	11.5	11.0	12

In chip and low cost ceramic and hermetic packages



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	NF	Ga	P1dB	Freq
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Semiconductor purchased EDA's FINESSE MCM™ design system as well as FINESSE MCM™ Design Kits to provide design systems, services and solutions for physical implementation of electronic systems such as ICs, MCMs and PCBs. Harris is the sponsor of the Technical Alliance for Multi-chip Engineering (TAME) which is an organization of 35 companies that provide various services needed in the MCM field. TAME sponsors an end user source book of companies that offer technology, products and services that are key to the use and implementation of MCM technology.

Trunked Radio Contract – Rohde & Schwarz has been given a contract to deliver a nation-wide trunked radio network for the United Arab Emirates by the end of the year. Initially, it will provide communications between civil defense, fire services and the Coast Guard, as well as providing access to the public telephone service. A digital trunking switch, MMX-64, (based on ISDN techniques) will be the central node of the ACCESSNET®, which in the first stages will be covered by 20 radio cells.

Harris Receives Contract – The Royal Netherlands Army awarded Harris RF Communications Group in Rochester, New York a $40 million communication system contract. An additional $40 million is included for options for its next-generation high frequency secure communications systems.

Advanced Technology Program – NIST has opened a new general competition to seek innovative industrial research programs for support under the Advanced Technology Program (ATP). An estimated $20 to $25 million in first year funding will be available in this competition. Proposals from all areas of technology will be accepted, and the deadline for full proposals is 3 p.m. EDT, June 22, 1994. Several program competitions focused on specific technological and business goals will be announced later. For information or application, call (800) ATP-FUND (287-3863). Fax: (301) 926-9524 or (301) 869-1150. Internet: atp@micf.nist.gov. Write: ATP, A430 Administration, NIST, Gaithersburg, MD 20899-0001.

Motorola's GSM Bases Sold – Base station equipment and service contracts have been concluded with Sweden, Morocco and Thailand. The total value exceeds $35 million.

Mid-May Move for FSY – The new address for FSY Microwave Inc.'s new facility is 7125 Riverwood Drive, Columbia, MD 21046. The 22,000 square foot facility was acquired to sustain FSY's continuing growth in supplying filters and multiplexers. Tel: (301) 294-8950. Fax: (301) 294-3007.

RF Power Components Moves to Larger Facilities – RF Power Components, Inc., manufacturers of high power resistors, 90° hybrids, dividers, and dual directional couplers has moved to 125 Wilbur Place, Bohemia, NE 11716-2482, USA. Tel: (516) 563-5050. Fax: (516) 563-4747.

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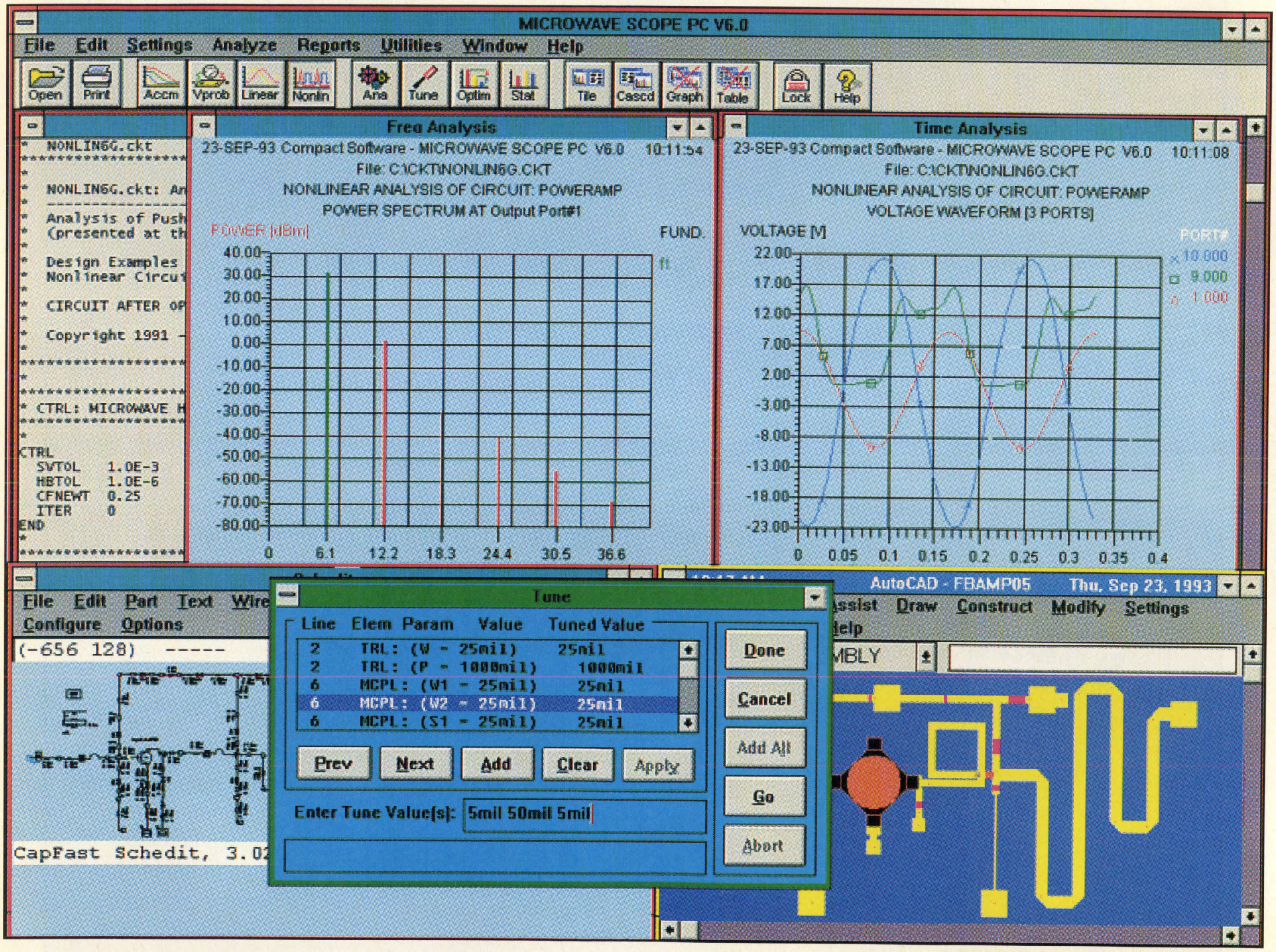

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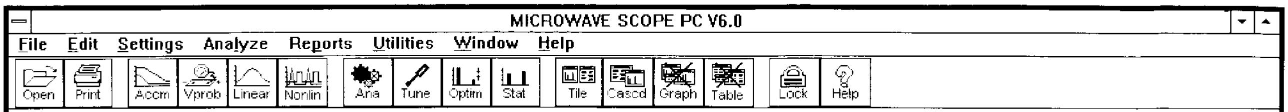

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Miniaturization: Getting More for Less

By Andy Kellett
Technical Editor

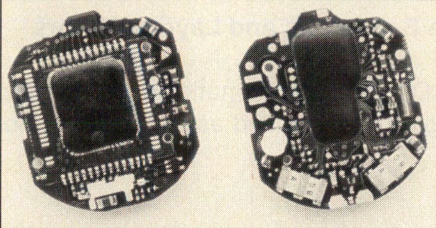
"Smaller is better" is almost an axiom in electronics. However, as axiomatic as that statement may seem, there are reasons for the ever-smaller size of electronic devices and components. This report looks at the factors driving RF miniaturization, and some of the technologies that enable RF miniaturization to take place.

Unquestionably there is more demand for portable RF devices, and that is a major source of the demand for miniaturization. "Portability is a society driven thing," says Tom Rose, Director of Strategic Marketing for M/A-COM's Microelectronics Division, "the idea of having an office and secretary going away, and having people on the road with their phones, faxes and pagers is becoming popular."

Even in military markets, small size has become increasingly important. For example, the manpack radio has largely been replaced with handhelds. "Miniaturization is an enabling technology," says Dr. Paul Hoff, Director of Lockheed-Sander's Microelectronics Center, "for example, the avionics in the F22 could not be implemented without the use of MMIC technology. "

Circuits and components are also shrinking so that increased functionality can be placed in familiar packages. A receiver within Seiko's Message Watches™ can extract time, personal messages, traffic reports and other data from an FM broadcast subcarrier. The Advanced Communications and Timekeeping chipset within the watch comprises a frequency synthesized receiver with digital baseband filtering and data recovery. The receiver chip is sophisticated enough to contain an image cancelling front-end and a power-down feature incorporated into the normal biasing scheme says Seiko Telecom Senior RF Engineer, Rick Suter.

Miniaturization has had another effect on the Message Watch, however, by further intergating the circuits in the watch onto fewer die, the cost and complexity of the watch has been decreased. This third generation of the watch is 40 percent smaller and uses three instead of six chips says Gary Gaskill, Director of Systems Engineering and Business Development at Seiko Telecom.



The three chips inside Seiko's Message Watch™ comprise a frequency synthesized receiver with digital baseband filtering and data recovery

Cost is an important reason for miniaturization. "If you're going to keep up with the cost curve you have to get more and more on a wafer," says Richard Langlais, Director of Alpha Industries' Sales Group. M/A-COM has reduced their chip size four to one in the last three years says Rose, "that's allowed us to take GaAs chips into the competitive commercial marketplace at a competitive price," says Rose. Increased integration reduces an end product's manufacturing and testing expenses, notes Kevin Townsend, Director of Sales and Marketing for Daico Industries, because more integration reduces parts counts

How to Get More for Less

Making things smaller, whether to satisfy price or functional demands, can be done in several ways. For oscillator manufacturer MTI - Milliren Technologies, Inc., it is design work, not special devices, which let them shrink their products. "Right now crystals are about as small as technology permits, so we have to be creative in the use of other components," says Dan Trepanier, Sales Manager for MTI.

Shrinking the size and cost of GaAs hybrid circuits is the goal of M/A-COM's GMIC process. The process bonds GaAs chips to a layer of glass which has been deposited on a silicon wafer. Using standard photolithographic processes, passive devices and interconnects can be deposited on the glass surface, and Si devices can be grown in the substrate. Only after all the components are placed on the substrate is the wafer diced.

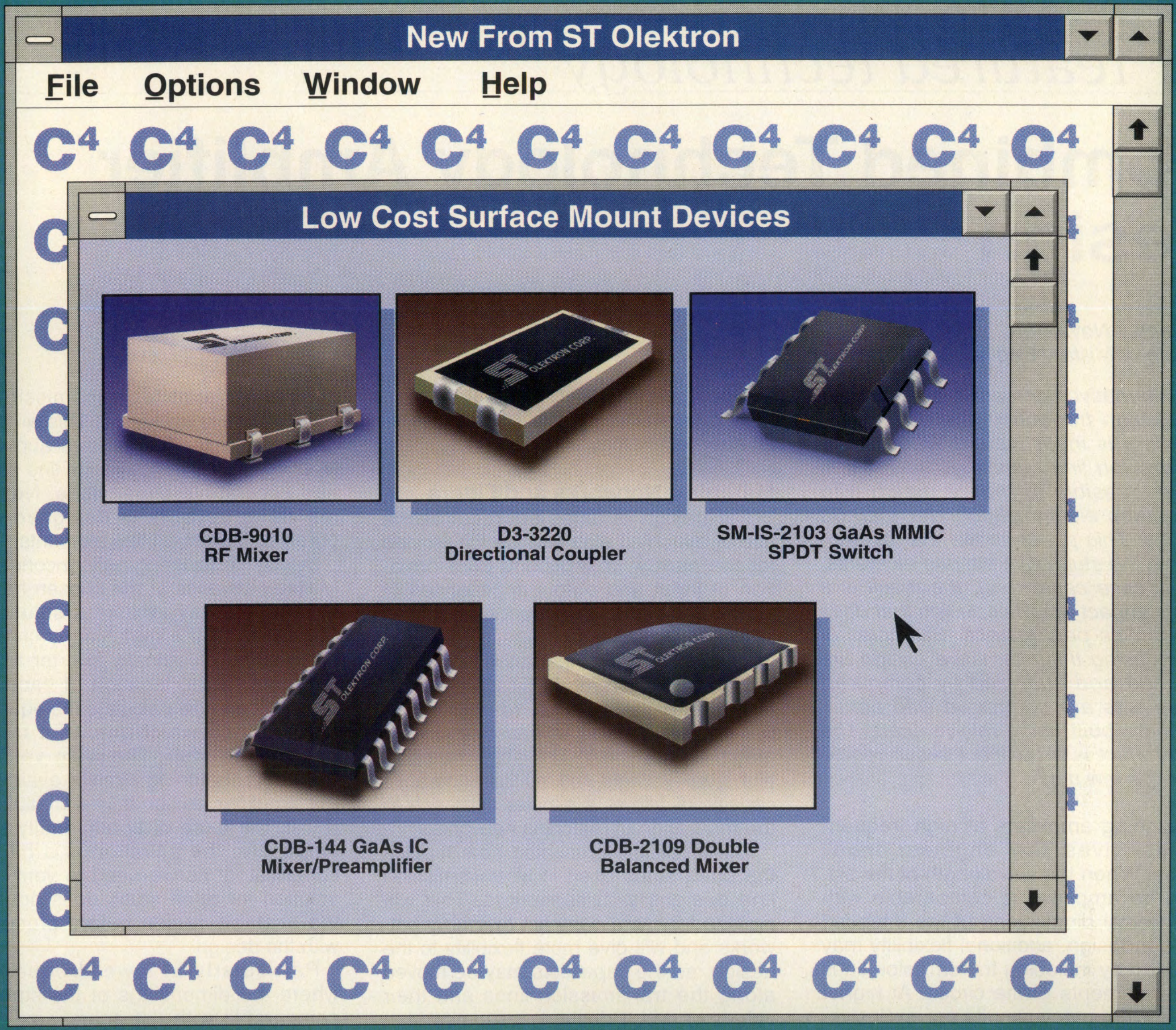
Higher levels of integration have been the key to Teledyne Microwaves' PCMCIA Type 2-compatible "transceiver-on-a-chip", the TFE-1050. According to Teledyne Marketing Manager Mel Lee, both size and cost reduction prompted Teledyne to make theTFE-1050 monolithic. "It's a pretty elegant solution in the sense that it does transmit and receive functions, it has the mixers on it and it even has trap filters on it," says Lee.

The next hurdle for GaAs MMIC manufacturers will be to economically produce 0.25 micron devices says Lockheed-Sanders' Hoff. X-ray lithography will do that, and will also be used in silicon fabrication, says Hoff.

As IC manufacturers work to put more and more functionality into a single package, discrete semiconductor makers are simply trying to make the devices they have smaller and cheaper. "If you look back five years, people were willing to use ceramic packages, glass axial-lead devices, things of that nature," says Alpha's Langlais, "If you look at the devices we make for chip-on-board, you'll see they don't do anything different from the ones made five years ago, but they do have a smaller footprint, and most importantly, they cost less."

Customers are also asking for smaller passive discrete devices. "The smaller you make a coil, the worse it performs in an RF environment," says Paul Liebman, Marketing Manager for Coilcraft. For Coilcraft's new series of 0603-sized wirewound inductors, this tendency is partly offset by the reduced wire lengths needed for the smaller inductances used at higher frequencies. "We've all been working hard at 900 MHz, and now we're going up to 1.2 or 1.8 GHz," says Liebman, "the bulk of the action we're going to experience on this part is under 100 nH."

The progress of miniaturization has always been a benchmark for the progress of electronics, and it will soon be marking the rapid progress of RF technology. *RF*



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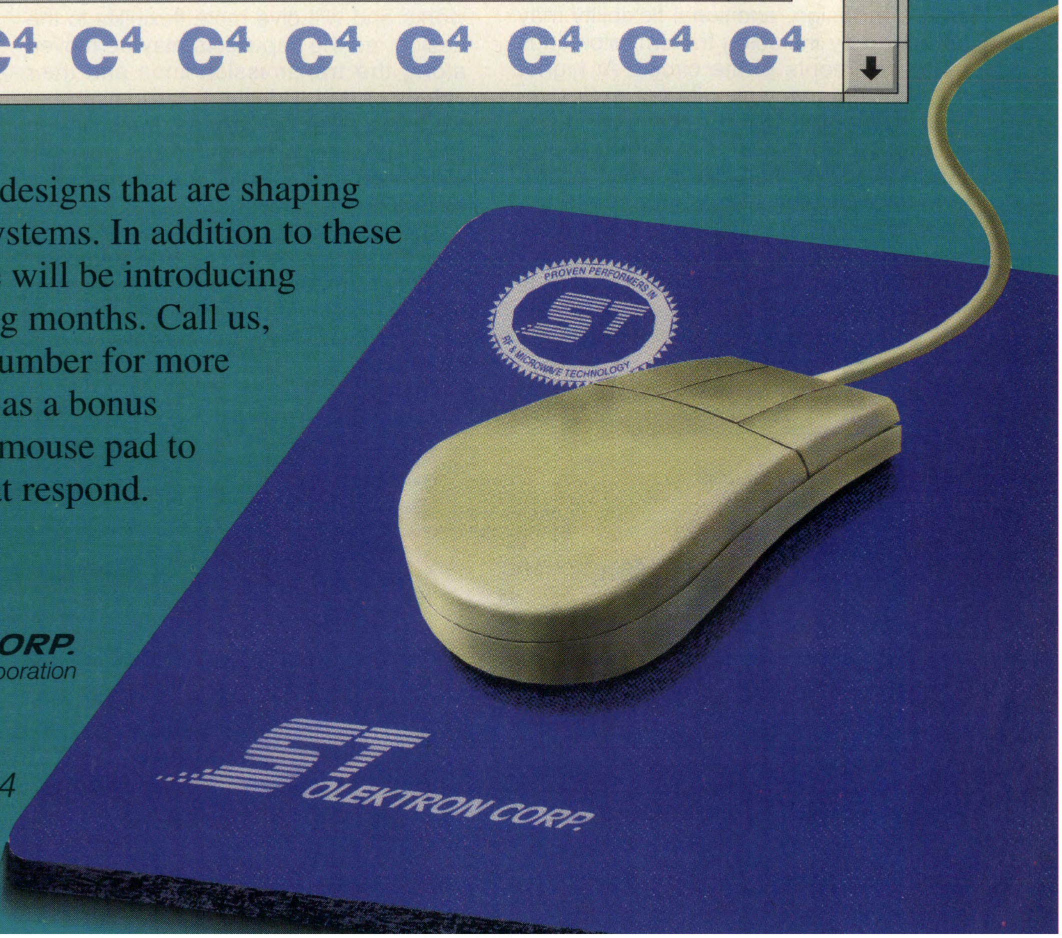

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INFO/CARD 26

Combined Technology Amplifier Design

By Stanley Novak
Institute of Military Engineering, Brazil

An alternative approach for the realization of high frequency and microwave amplifiers is investigated. When using transmission line matching circuits, it is always possible to realize match with open stubs electrically shorter than 90 degrees. This paper shows that by substituting the stubs with lumped elements, (in this case capacitors), the result is a more compact amplifier design that does not sacrifice performance. Examples of circuits using this alternative design are designed and analyzed by computer, and results are compared with actual amplifiers built using this method. The final product is a compact circuit requiring only a few parts.

Designing amplifiers at high frequencies gives the engineer many options. When the wavelength of the signal to be amplified is comparable with the physical dimensions of the intended amplifier design, additional flexibility may be added by including transmission lines as the elements of the circuit. At higher frequencies, many circuits are exclusively implemented using transmission line elements. Alternatively, lumped elements can be used exclusively to provide for matching at the appropriate frequency. This approach is used particularly in some types of MMICs.

In this work the combination of both technologies is investigated, and it is shown that in some cases the combined approach can have considerable advantage over the exclusive use of one or another technology.

Matching Networks and Filters

The design of amplifiers requires the use of matching elements, which provide for the narrow or wideband transformation of input and output impedances of active devices to the impedance of the selected transmission line. Often, the input and output impedances of these bipolar transistors and FETs vary wildly. In the majority of cases, we exclusively use lumped element or transmission line matching networks. The appearance of high quality single and multilayer ceramic capacitors allows a simpler approach to the realization of matching networks.

The amplifiers presented here replace the open stubs used in all-transmission line designs with capacitors. This will provide for more compact matching networks and will give more flexibility to the design, as the capacitors may be moved along the transmission lines and their values varied if desired, (assuming that we use a shielded coplanar transmission line [1]). This is of course not possible with open stubs, which are permanently etched on the substrate.

The actual design proceeds as follows: first the active device for the particular amplifier is selected, and its S-parameters are evaluated for the selected frequency band. Then the suitability of the complete circuit for transmission line construction is evaluated. This will be the case mainly for frequencies above 300 MHz where the transmission line solution has acceptable dimensions. Next, the matching network is designed using purely transmission-line elements.

Initially considering only unconditionally stable devices at the chosen frequency, there always exists a solution for realization of the matching circuits with open stubs electrically shorter than 90 degrees. At the location of each open stub we can now calculate the equivalent capacitance which must be used to replace the stub. The wide variety of realizable matching circuits available to the amplifier designer is well illustrated [2]. In the case of conditionally stable amplifiers, the situation is a bit more complicated, as we need to verify if the solution for open stubs does not fall in the unstable region before proceeding with the design.

Particularly at lower frequencies, where the dimensions of the stubs are comparable with the actual size of the substrate, replacing stubs with capacitors can substantially reduce the amplifier's dimensions.

For wider band operation we could use a negative feedback network in the design [3], or multiple element matching networks designed only with open stubs and calculate the equivalent capacitances to be placed at appropriate dis-

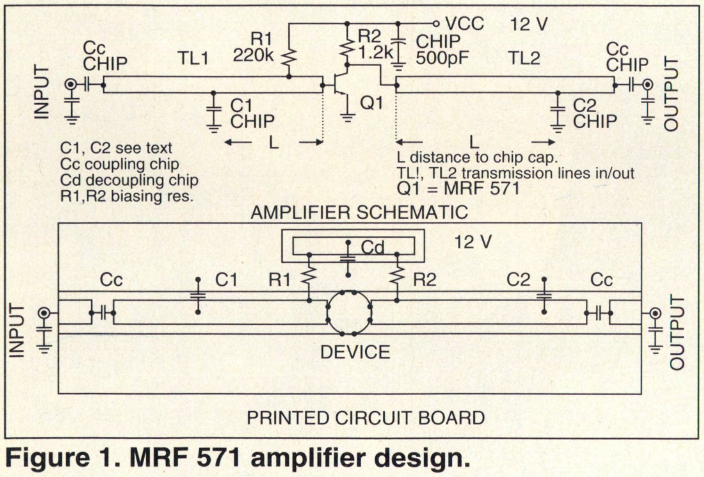


Figure 1. MRF 571 amplifier design.

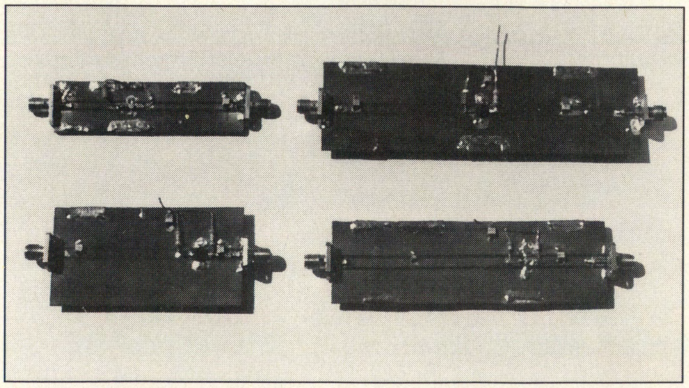


Figure 2. Photographs of maximum-gain and low-noise designs using MRF 571 and NE 02135 transistors.

tances. The final configuration will again be a single transmission line at the input and output of the amplifier, with chip capacitors placed at calculated intervals along the lines. This approach could obviously be extended to the design of filters and other types of ladder networks [4].

Alternatively, in many cases we can try to run an optimization program using elements calculated for narrow band design as the starting values.

Transmission Line Matching

To demonstrate transmission line and stub matching with a two-element lossless matching network, it is best to visualize the situation using the Smith chart. The proper transmission line length transforms the transistor input impedance to the point around the Smith Chart where the real part of the impedance is equal to the characteristic impedance of the transmission line. Note that for the shortest length of the transmission line there are always two such points, one in the inductive part of the chart, another in the capacitive part of the chart. The length of the shorted or open stub then must cancel the imaginary part of the impedance at this point, completing the match. Because for this purpose we use parallel stubs, we complete the solution by using admittances on the chart. A similar approach is then used on the transistor output. Because in our case we are interested only in open stubs and their equivalent capacitances, we could write for admittance of the open stub

$$Y_{os} = jY_o \tan\theta \tag{1}$$

where

$$\theta = \beta L = \frac{2\pi L}{\lambda(\text{rad})} \tag{2}$$

or

$$\theta = \frac{360L}{\lambda(\text{deg})} \tag{3}$$

where Y_{os} is the admittance of the open stub, Y_o the admittance of the transmission line, β the phase constant of the line, L the length of the line, λ the free space wavelength = $3 \times 10^8/f$ (m), and f is the frequency in Hertz.

Because an open-circuited stub shorter than 90 electrical degrees always has positive susceptance, i.e. is always capacitive, we could write

$$Y_{os} = jB_{os} = jY_o \tan\theta \tag{4}$$

or

$$\tan\theta = \frac{B_{os}}{Y_o} \tag{5}$$

substituting for Θ from 2,

$$\beta L_{os} = \tan^{-1}\left(\frac{B_{os}}{Y_o}\right) \tag{6}$$

giving for the open stub length

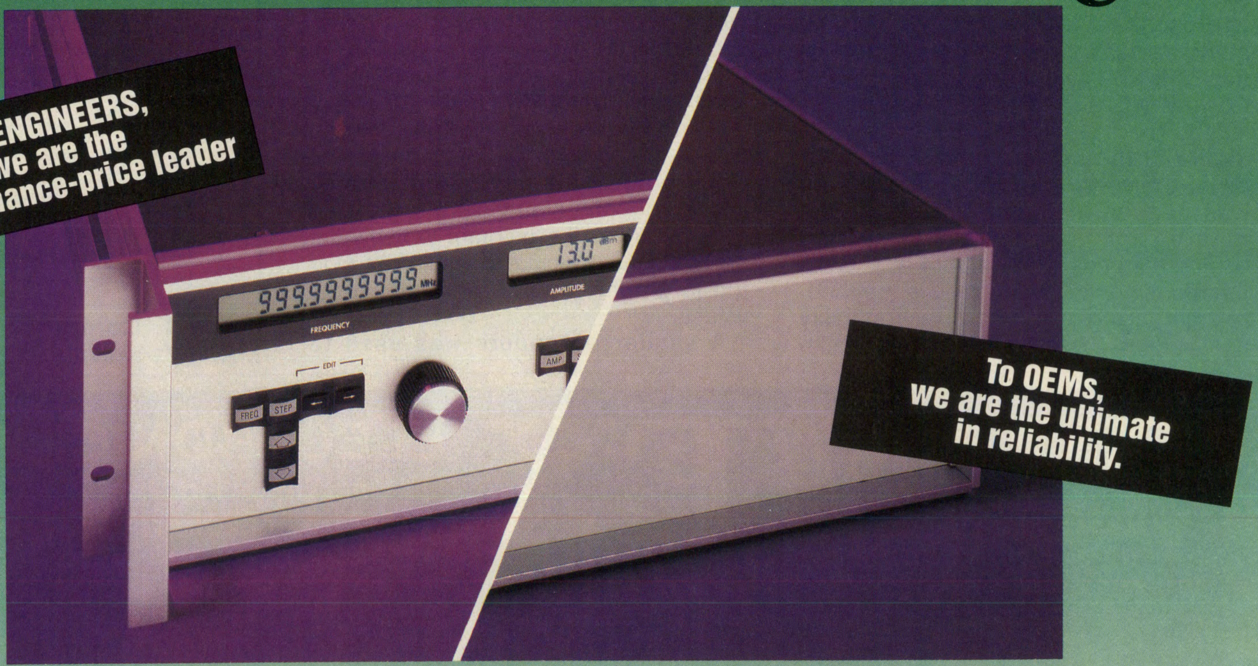
$$L_{os} = \frac{\lambda}{2\pi}\tan^{-1}\left(\frac{B_{os}}{Y_o}\right) \tag{7}$$

Finally, for the open stub equivalent capacitance we have

$$C_{eq} = \frac{B_{os}}{2\pi f} \tag{8}$$

From equation 8 we can calculate the value of the capacitor which can replace the open stub at the particular frequency and distance. A word of caution must be added here. The capacitive susceptance of the stub is a tangential function, there-

Split Personality



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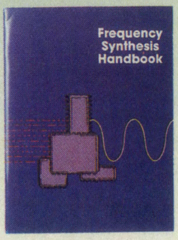

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fore for electrical lengths of the stub more than about 60 degrees, the capacitive variation for the unit length becomes increasingly severe. In such cases we may need to use a variable capacitor or test a few fixed capacitors in the circuit to obtain the exact capacitance value.

Combined Element Amplifier

To verify the validity of this theoretical discussion, an amplifier with a two-element matching network for 1 GHz was designed and analyzed using various types of transistors. The results, as expected, confirm the validity of this approach. As an example, we demonstrate designs using the bipolar transistors MRF 571 (manufactured by Motorola) and NE 02135 (manufactured by NEC). The first device is unconditionally stable at the selected frequency of 1 GHz, therefore no further adjustment was needed, and initially a maximum gain design was used. The second device is conditionally stable at 1 GHz, therefore we decided to reduce the theoretical gain to 16 dB to obtain stability and verified the condition on the Smith chart. The parameters of the input and output transmission line-open stub matching circuits were calculated together with corresponding capacitances for the open stubs.

The amplifiers were designed using computer synthesis programs [5], [6], selecting the solution for the open stubs. An additional feature of the program is the subroutine for calculation of the equivalent capacitance of the open stubs, which is displayed next to the dimensions for the stub parameters.

The resulting computer printout for MRF 571 is shown in Table 1, together with calculations for the chosen substrate material, in this case Duroid D5880, having relative permeability 2.2 and thickness of 0.7874 mm.

To check the accuracy of the solution we used a circuit analysis program. The open stub equivalent capacitances instead of open stub lengths were used as an entry to the program to verify the assumptions. Calculated results are plotted in graphs together with measured values.

A similar procedure was used to

```
TRANSISTOR  MRF571 @ FREQUENCY 1000 MHZ
INPUT–S(11)=              .61              178
FORWARD–S(11)             3                78
REVERSE–S(12)=            9.000001E–02     37
OUTPUT– S(22)=            .28              –69

CHAR. IMPEDANCE OF THE LINE = 50 OHM
STAB. FACTOR K=1.036  **UNCONDITIONALLY STABLE**
MAX.GAIN IN DB=14.054              MAX.NUMERIC GAIN= 25.438

IMPEDANCE TO MATCH SOURCE          RS=2.887       XS= .56
IMPEDANCE TO MATCH LOAD            RL=17.571
XL=–73.95

INPUT REFL.COEFF.         MAGNITUDE= .89
ANGLE=–178.711
OUTPUT REFL.COEFF.        MAGNITUDE= .806
ANGLE=66.097

LINE AND STUB MATCH TRANSISTOR MRF571
@ FREQUENCY 1000 MHZ
****INPUT****
PARAMETER      IMP.(Ω)   LO   EL.DEGREE    EQ.PARAM
LINE LENGTH    50        .035 12.868
OPEN STUB      50        .21  75.691       (CAP=12.48 PF)
****OUTPUT****
OPEN OR SHORTED STUB? [O/S]O

****OUTPUT****
PARAMETER      IMP.(Ω)   LO   EL.DEGREE    EQ.PARAM
LINE LENGTH    50        .208 75.093
OPEN STUB      50        .194 69.842       (CAP= 8.671 PF)

****MICROSTRIP CIRCUIT DESIGN****
PARAMETERS FOR MICROSTRIP LINES MRF571
@ 1000 MHZ CENTER FREQ.
SUBSTRATE REL.DIEL.CONST. = 2.2
SUBSTRATE THICKNESS = 0.7874 [mm]
LINEWIDTH FOR 50 OHMS STUBS AND LINES = 2.397 [mm]
OPEN STUB-IN LENGTH = 45.882 [mm]
OPEN STUB-OUT LENGTH = 45.519 [mm]
LINE LENGTH-IN = 7.8 [mm] / LINE LENGTH-OUT = 45.519 [mm]
LINEWIDTH FOR MAINLINE 50 OHMS = 2.397 [mm]
*******CIRCUIT CONFIGURATION*******

  GEN.* S *** LINE *** TRANS. *** LINE ** S * LOAD
        T                                 T
        U                                 U
        B                                 B
     OPEN                              OPEN
```

Table 1. Computer synthesis for MRF 571 maximum gain amplifier.

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49 shielded, 49 unshielded (2 of each)
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"Unicoil" 5 mm Tuneable Inductors
Inductance: 9 nH - 281 nH
19 Shielded, 19 unshielded (2 of each)
Kit M105 $60

"Slot Ten" 10 mm Tuneable Inductors
Inductance: 0.7 µH - 1143 µH
18 shielded, 18 unshielded (3 of each)
Kit M100 $60

"Slot Seven" 7 mm Tuneable Inductors
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design an amplifier for low noise operation. For this type of design, the synthesis program requires that the minimum noise input reflection coefficient be specified; this quantity is obtained from manufacturer spec sheets.

For the NE 02135, which is only conditionally stable at the design frequency, we made similar calculations for both maximum gain and low-noise amplifiers.

Amplifiers were realized by etching striplines on the chosen substrate, and providing ground along both sides of the stripline, thus forming a shielded coplanar transmission line. (alternatively feedthrough to the ground side of the substrate can be made), The capacitors can be placed at the calculated distances, and for fine tuning, moved a bit along the lines to obtain the best match. This type of design is very flexible because it is possible to etch 50 or 75 ohm line of sufficient length on the substrate, and calculate amplifiers for different frequencies. Then, only the positions of the capacitors of different values will determine the exact frequency and final performance of the amplifier. Therefore, one type of substrate layout can be used for equalization of many types of amplifiers. Such a circuit has considerable advantages over standard stripline designs.

Amplifier Circuit

To keep the circuit as simple as possible, two-resistor biasing was used. The circuit itself and the layout on the substrate is shown in Figure 1. The coupling capacitors Cc at the input and output of the amplifier limit the low frequency response, and are chosen so that the gain at lower frequencies is reduced to the desired value. In our case we chose 22 pF coupling capacitors for both devices. All chip capacitors used in the design were manufactured by ATC. Capacitors C1 and C2 given by the computer for each particular application were placed at calculated distances L along the matching transmission line. Note that in the actual circuit their ground side is placed on the side where the power supply decoupling capacitor is placed. This reduces the chance of unwanted resonances, which are always a danger in high frequency work.

Proper grounding is of utmost importance and the best solution is to provide feedthrough plated holes to the bottom ground plane close to each capacitor. The power supply decoupling capacitor Cd is 500 pF chip. Resistors R1 and R2 are 1/8 W resistors or smaller, soldered directly on the line close to the device with the shortest leads possible. Their values are calculated to provide for the proper biasing conditions. The device itself is placed in the drilled hole in the substrate with ground leads soldered through holes on both sides of the substrate. The top of the substrate and the bottom ground plane then form a coplanar shielded transmission line. A small island is etched on one side to provide for connection of a 12 V power supply in the case of the MRF 571, and 15 V in the case of the NE 02135. To assure that the coplanar planes along the central transmission line provide good ground, wide copper strips are wrapped on the edge of the substrate close to capacitors C1 and C2 and soldered on the top and bottom of the substrate. On the connector side, the ground plane is soldered to the connector on top and bottom with the center connector soldered to the central transmission line. Proper grounding prevents unwanted resonances which can substantially influence the overall performance of the amplifier.

Results

The completed amplifiers were checked and adjusted using the HP 8754A network analyzer and measured using the HP 8790 noise figure meter. In the case of the MRF 571 bipolar transistor, the maximum gain design for a 1 GHz center frequency actually resulted in a center frequency around 920 MHz, as shown in Figure 2. Reducing the input capacitor C1 (which has the biggest influence on the setting of center frequency) from the calculated value of 12.48 pF (13 pF was used) to 8.2 pF moved the center frequency to 1 GHz, as is also shown in Figure 2 (this change also reduced the noise figure).

The reason for the discrepancy in the C1 value may be explained by the fact that the equivalent open stub is over 60 degrees long, i.e. the equivalent reactance is changing rapidly in this area (because it is a tangential function), so the capacitance adjustment is very sensitive.

No attempt was made to move capacitors to compensate for the frequency deviation, and the distances to the capacitances were measured from the center of the device. Also, no correction for stray capacitances was included in the analysis program. In cases where a particular frequency is desired, a variable capacitor should be used at the input to set the frequency value.

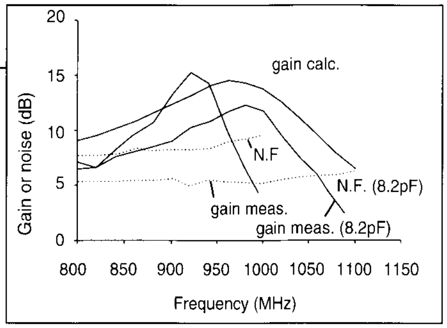

Figure 3. MRF 571 maximum gain amplifier.

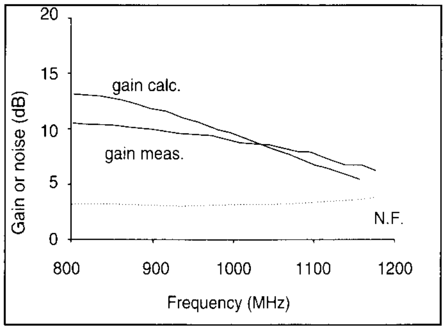

Figure 4. MRF 571 low noise amplifier.

For the next experiment we designed an amplifier for the same center frequency, but optimized it for low noise operation. The result was an amplifier with lower center frequency (using now 2.7 and 3.6 pF chips instead of the calculated 2.67 and 3.48 values), but with much wider bandwidth, and this time the noise figure was, as expected, much lower and centered around 2.2 dB. Analysis of the amplifier actually indicated the frequency shift, shown by comparing calculated and measured values in Figure 3.

This result could be expected, because to achieve the low noise performance we must actually mismatch the input of the amplifier, and this subsequently results in shifted center frequency. Therefore, to actually make an amplifier for minimum noise at some desired frequency, we need to start with the design at higher frequency and then use an analysis program to finely adjust the circuit for proper center frequency. But note that the measured and calculated values are very close for the 1 GHz design. In this case, the reduction of input capacitance did not result in significant change in gain and noise performance, because the amplifier was already mismatched for low noise operation and because the equivalent open stub length was shorter than 60 degrees (actually 47.5 degrees), the circuit was not as sensitive to the change of operating condition as the maximum gain case.

Amplifiers realized with the NE 02135 device were built using fiberglass sub-

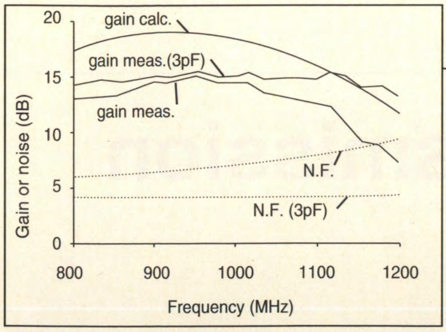

Figure 5. NE 02135 maximum gain amplifier.

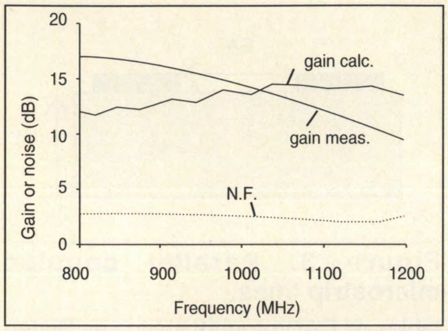

Figure 6. NE 02135 low noise amplifier.

strate with permitivity 4.5 and thickness of 1.6 mm. The results of the measurements for both types of the amplifier are shown in Figure 4 and Figure 5. As in the case of the MRF 571 maximum gain amplifier, the calculated input capacitance of 6.8 pF needed to be decreased to 3 pF for the same reason (stub longer than 60 degrees), resulting once again in reduction of noise figure and wider bandwidth, but not to the same extent as the MRF 571. This is due to the fact that in this case we have a conditionally stable device with K = 0.664, which must be partially mismatched at the output to achieve stable operation, and is therefore desensitized. This is similar to the case of low noise operation, but here we purposefully mismatch the input of the device. Both measurements are included in Figure 4.

In the case of the low noise amplifier, the performance and behavior is similar to that of the MRF 571, but the usable gain extends above 1.2 GHz. Again, the reduction of input capacitance did not result in significant improvement of the performance, as the equivalent open stub was only 33 degrees long.

To compare performance, amplifiers using both shielded coplanar line and conventional stripline were constructed, but the resulting performances were similar. The shielded coplanar line is obviously more suitable for the realization of combined technology amplifiers because it offers more flexibility in adjustments.

Conclusion

For maximum gain amplifiers the actual gain at design frequency was lower than predicted, indicating the presence of losses not included in the theoretical models used for amplifier analysis.

Some loss in gain may also be attributed to the spread in actual device specifications. In all cases the biasing was adjusted to the manufacturer recommended value. All the devices were selected at random from a batch and their S-parameters were not measured, but taken from manufacturers specifications.

In the case of low noise amplifiers, the calculated gain intercepted the measured values close to the design frequency, and were lower at frequencies above the design frequency. Noise figure were close to the manufacturer's values.

Coplanar shielded line, therefore, can be used as an attractive alternative to the conventional stripline for realization of microwave and high frequency amplifiers and other devices. *RF*

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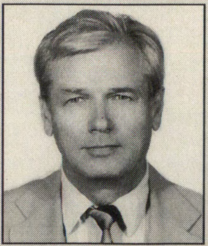

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Simulating Coupled Transmission Lines with Super-Spice

By Krishnamoorthy Kottapalli
Compact Software, Inc.

Compact Software's Super-Spice™ time-domain simulator extends the traditional capabilities of SPICE-based simulation by providing RF and digital designers with the capability to model systems containing lossy coupled transmission lines. The coupled-line model in Super-Spice can represent up to ten coupled transmission lines on single and multi-layered substrates. This coupled-line model makes it possible for designers to quickly identify circuit failures due to cross talk. Figure 1 shows a typical Super-Spice, work session which utilizes a graphics user interface based on X-Windows ™ and OSF/Motif ™ .

Traditionally, Spice simulators have been used extensively for transient and steady state analysis of many analog linear and non-linear circuits. The original Berkeley-Spice™ simulator, SPICE 2G6, required that the circuit topology be specificed using discrete circuit elements. Distributed elements such as uniform lossless transmission lines could be specified by their characteristic parameters, delay time and characteristic impedance. While these parameters were easy to determine for transmission lines in homogenous media, (e.g., coaxial lines), their specification is very complicated in non-homogenous media.

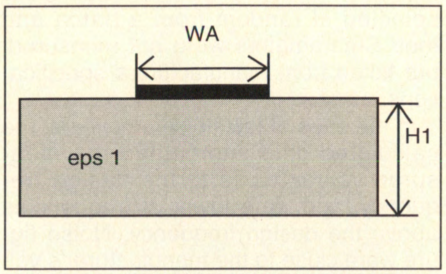


Figure 2. Single microstrip line.

The characterization of today's dominant microstrip-based interconnects requires detailed numerical analysis. Relevant values for the characterization of typical single and coupled line structures are readily accessible through tables or CAD programs such as Super-Compact™ . Instead of specifying interconnect characteristics by their electrical parameters, it is more convenient for designers to specify their physical dimensions.

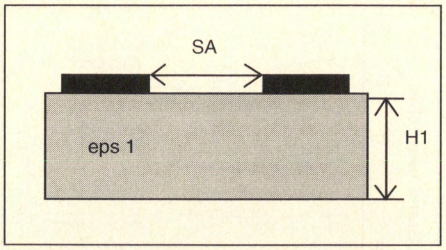


Figure 3. Parallel coupled microstrip lines.

Super-Spice

Super-Spice is an enhanced version of Berkeley-Spice and follows the PSPICE™ conventions. A unique feature of Super-Spice is the integration of an electromagnetics module that characterizes single and coupled lossy microstrip lines on multiple substrate layers [1] accurately. Based on the physical dimensions and substrate parameters of a single or coupled line geometry, the electromagnetics module automatically computes the associated electrical characteristics. These electrical characteristics are then passed to the Spice simulation portion of the program. With this modeling approach the designer can specify the line by the available physical parameters. Also cross talk can be handled by specifying the geometry of closely spaced transmission lines, whether they are parallel or broad side coupled.

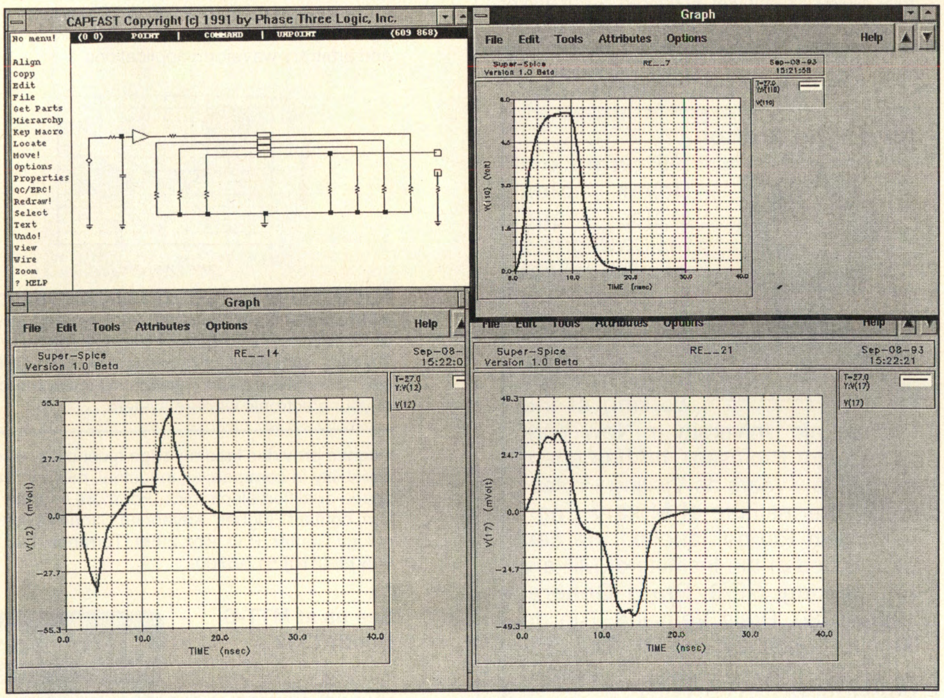


Figure 1. Typical Super-Spice ™ session.

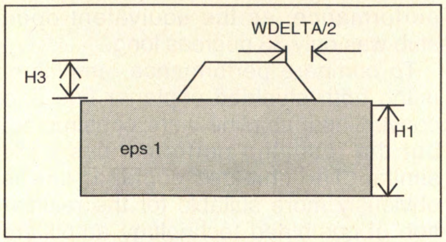


Figure 4. Thick conductor lines analyzed with Super-Spice.

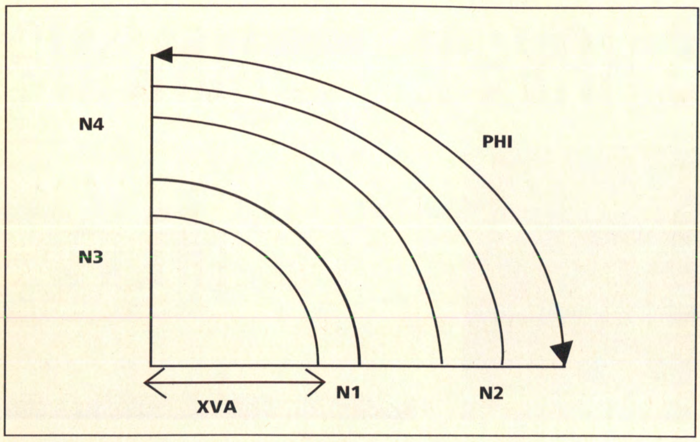


Figure 5. Schematic of radial coupled lines.

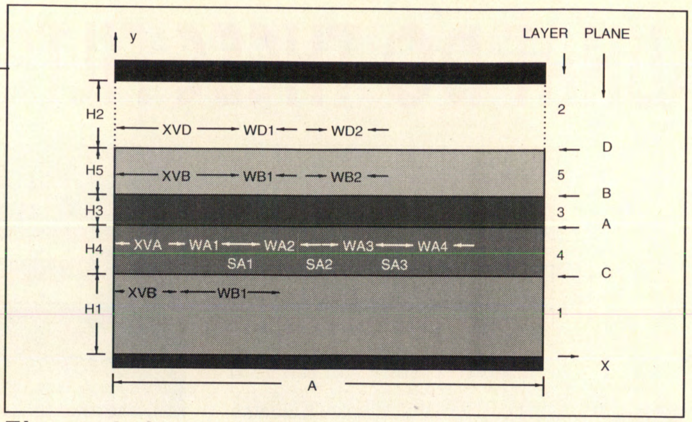


Figure 6. Lossy, multi-conductor, coupled-line model, STRUC, in Super-Spice.

A single or coupled transmission line element is defined by following the native Spice sub-circuit and model convention. Designers familiar with Spice can apply the Super-Spice specific interconnect models. For instance, a single microstrip line as shown in Figure 2 can be referenced in Super-Spice by the following sub-circuit reference:

U_Input 1 2 feed_line L=10mm

"Feed_line", representing a feedline of lenght 10mm, is characterized in a model statement. Single and coupled line models are identified by the keyword STRUC:

model feed_line STRUC
+ WA=0.5mm
+ H1=0.635mm EPS1=9.8
+ RFA=1HMET=0.02mm

The model parameters H1, WA and EPS1 refer to the line width, substrate height and dielectric constant. RFA specifies the metal resistivity normalized with respect to copper and HMET characterizes the metal thickness. Optionally, an additional parameter H2 can be specified to characterize the cover height of a housing.

Parallel coupled microstrip lines with spacing SA (as shown in Figure 3) are specified in a similar way:

U_Input 1 2 3 4 feed_lines L=10mm

.model feed_lines STRUC
+ WA=0.5mm, 0.5mm
+ SA=0.3mm
+ H1=0.635mm EPS1=9.8

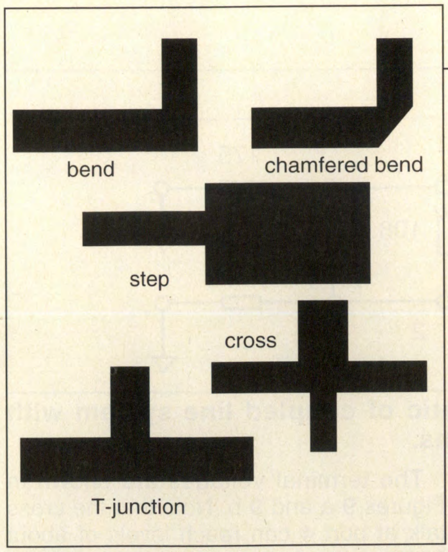


Figure 7. Microstrip discontinuities that can be analyzed using Super-Spice.

+ RFA=1 HMET=0.005mm

The specification of the metal thickness as described by HMET is valid for conductors with a width to thickness ratio of more than ten. However, for many digital and RF integrated circuits, the conductor thickness can be of the order of the conductor width. The cross section of such lines is shown in Figure 4. It is common for the lines' cross section to be trapezoidal due to their manufacturing process. Such thick lines can be modeled in Super-Spice by including two control parameters in the model statements. For the lines shown in Figure 4, the model statement includes the control parameter IDICK and its thickness is specified by H3. The parameter WDELTA defines the undercut for the line:

```
U_Input 1 2 3 4 thick_feed_lines
L=10mm

.model thick_feed_lines STRUC
+ WA=0.5mm, 0.5mm
+ SA=0.3mm
+ H1=0.635mm EPS1=9.8
+ RFA=1 IDICK=1 H3=0.5mm
WDELTA=0.25mm
```

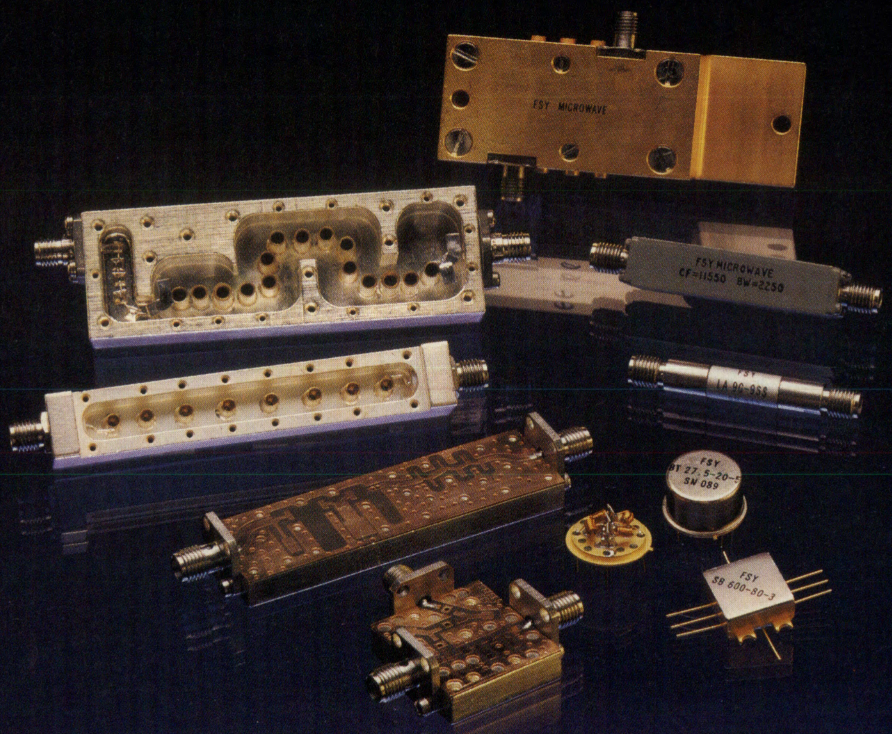
Super-Spice not only handles uniformly coupled lines but also models radial coupled lines. A schematic of radial coupled lines is shown in Figure 5. This coupled transmission line system can be modeled in Super-Spice by replacing the length parameter L with the angle PHI in the Super-Spice element statement, and specifying the inner radius XVA of the line with the shortest radius:

```
U_Input 1 2 3 4 radial_lines PHI=90

.model radial_lines STRUC
+ WA=0.5mm, 0.5mm
+ SA=0.3mm
+ H1=0.635mm EPS1=9.8
+RFA=1 HMET=0.005mm
```

The general STRUC model from which many lossy multi-conductor, coupled line geometries can be derived is shown in Figure 6. The STRUC model in Super-Spice can be used to model transmission lines in various media such as microstrip, buried microstrip, stripline, coplanar and suspended stripline.

Additional features of Super-Spice include models for microstrip discontinuities such as right angle bends, mitered bends, crosses, step junctions and T-junctions. A schematic of such junctions is shown in Figure 7. Junctions are referenced in Super-Spice through a sub circuit element that begins with the letter



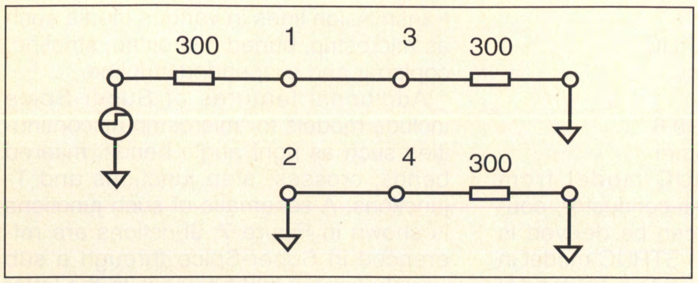


Figure 8. Schematic of a pair of parallel coupled lossy lines with mismatched termination.

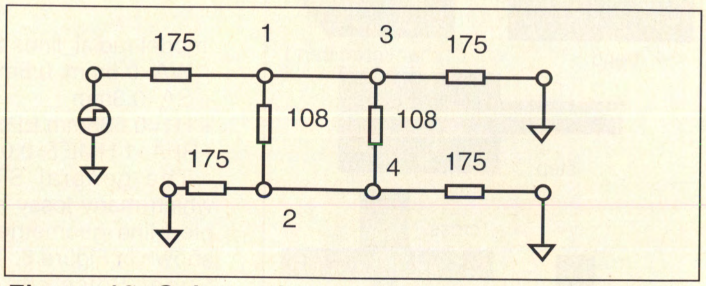


Figure 10. Schematic of coupled line system with matched terminations.

Z. As an example, a right angle bend element is specified by the following statement:

Z_junction 1 2 corner

.model corner LBEND
+ H=0.635mm
+ W1=0.635mm W2=0.635mm
+ ER=9.8

A T-junction is specified by the following statement:

Z_junction 1 2 3 corner

.model corner TJUNC
+ H=0.635mm
+ W1=0.635mm W2=0.635mm W3=0.635mm
+ ER=9.8

Examples

Super-Spice was used to analyze a pair of parallel coupled microstrip lines on alumina substrate. One meter long lines with line width 140 mm and spacing 30 mm were deposited on a 0.56 mm high substrate. A schematic for the coupled line system together with terminations is shown in Figure 8.

The terminal voltages are shown in Figures 9 a and 9 b. Note that the cross talk at port 4 can reach levels of about 50% of the input pulse at port 1. To reduce the coupling while maintaining the spacing between the lines, the coupled line system can be terminated in its characteristic termination impedance network. This network can be derived from the characteristic impedance network [2], which can be computed from Super-Spice's output parameters. Figure 10 shows the schematic of the coupled line system with the matched termination network. Associated results for

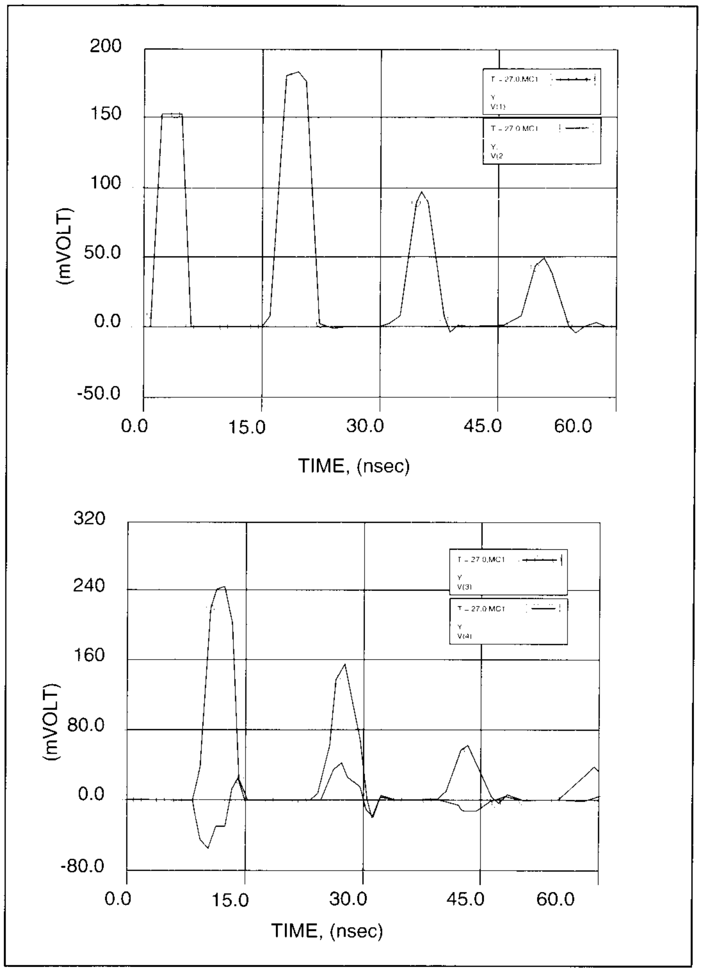


Figure 9. Terminal voltages for the coupled line system from Figure 8, (top-reflection, bottom-crosstalk).

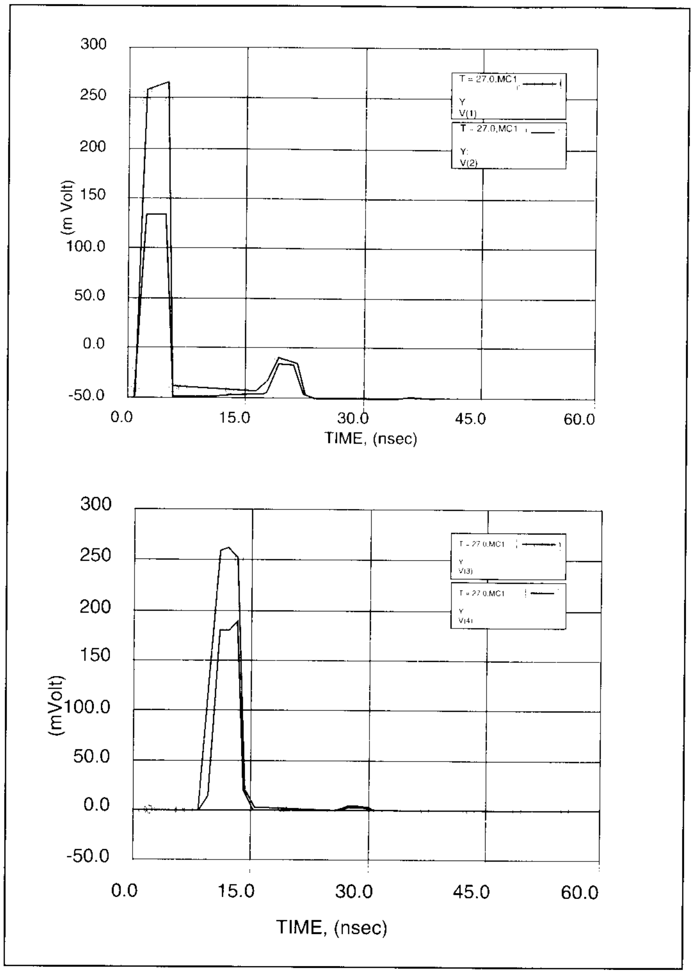


Figure 11. Terminal voltages for the coupled line system from Figure 10. (top-reflection, bottom-crosstalk).

the terminal voltages are shown in Figure 11 a and 11 b. Thanks to the proper termination of the lines, reflections and cross talk have been significantly reduced.

Super-Spice provides the characteristic parameters for the coupled line system. These parameters can be used for design purposes with conventional Spice or other programs. For instance, those parameters can be used to synthesize a matching network for the coupled line system under investigation. Super-Spice outputs the model line parameters of the coupled line system, including the decoupled modal propagation constant (β), the decoupled modal impedance Z_0 and the current eigenvector matrix. The delay values and DC resistance of the lines can also be obtained.

To demonstrate the pulse degradation on a long propagation path, a microstrip meander line was analyzed. This circuit includes ten parallel coupled lines and 18 right angle bends. The circuit uses a 0.6 mm thick alumina substrate and 5 mm thick copper lines. Each line section was chosen to be 500 mm long. A schematic of the line together with terminations is shown in Figure 12. Figure 13 shows a plot of the input pulse and the output voltage waveform when the spacing is equal to the width of the line (s = w). Figure 14 shows a plot of the input pulse and the output voltage waveform when the spacing is one half the width of the line (s = w/2). Note the apparent signal degradation at the output end of

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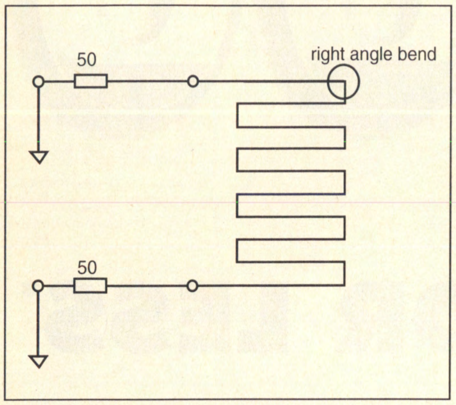

Figure 12. Schematic of a meander line consisting of ten parallel coupled lines and 18 corners.

the line, which is mainly due to cross talk.

As a final example, broad side coupled lines were analyzed, as illustrated in Figure 15. The first line was deposited on a 0.6 mm alumina substrate and the second line was deposited 0.2 mm above the first line on the same substrate. The terminal voltages are displayed in Figure 16a and 16b.

All of the examples listed were analyzed on a Sun SPARC 10 workstation. The CPU time for the first example was

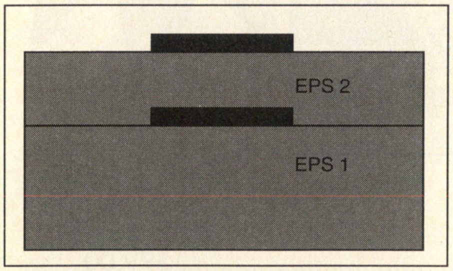

Figure 15. Broad side coupled lines with terminations.

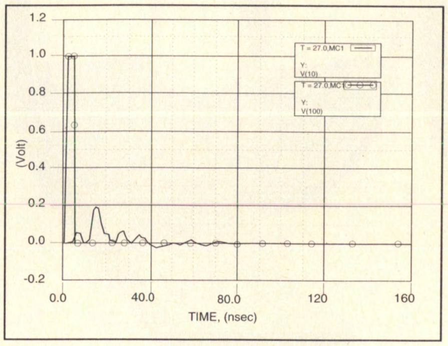

Figure 13. Terminal voltage for the meander line of Figure 11 for s = w.

two seconds for 80 data points. The CPU time for the third example, with 10 coupled lines and 18 bends, was 22 seconds for 220 analysis points.

Summary

Compact Software's Super-Spice includes an electromagnetics module that makes accurate modeling of lossy multi-conductor transmission lines possible. The model for uniform and radial coupled lines accepts geometrical input data and conforms to the Spice syntax. Additionally, Super-Spice provides models for microstrip discontinuities such as tees, bends, steps and crosses. Various applications for the analysis of coupled lines have been presented to demonstrate the usefulness of the program.

Acknowledgments

This paper is based on an earlier work by Achim Hill and Raymond Pengelly. *RF*

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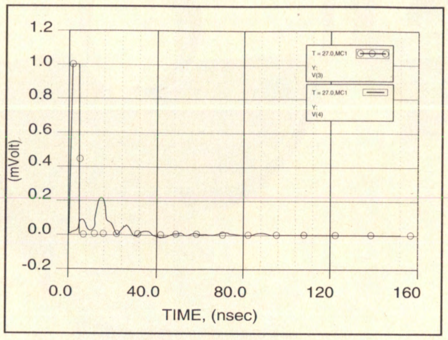

Figure 14. Terminal voltage for the meander line of Figure 11 for s = w/2.

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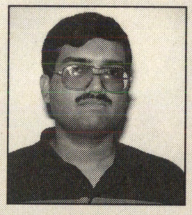

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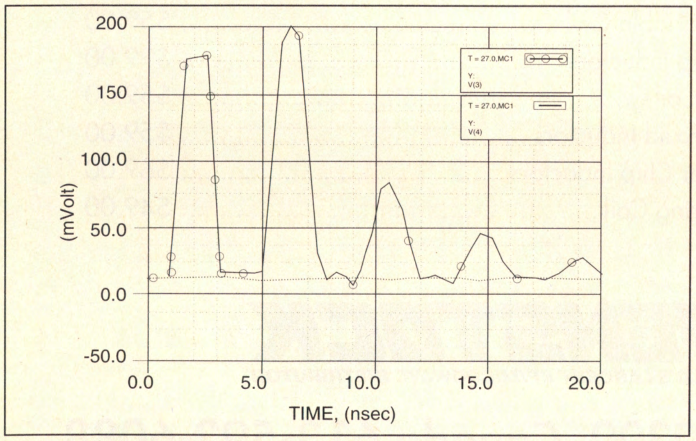


Figure 16a. Terminal voltage of broad side coupled lines from Figure 15.

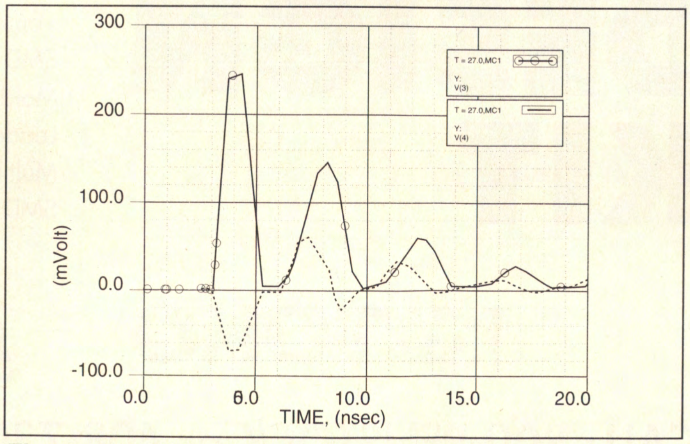


Figure 16b. Terminal voltage of broad side coupled lines from Figure 15.

A New CAD Method For Narrowband Matching Circuits

By Mihai Albulet
Polytechnic Institute of IASSY

The general problem of impedance matching consists of transforming a load impedance to the optimal working impedance of the generator. This article investigates narrowband matching circuits (made up of three lumped elements) used in RF power amplifiers.

We choose this case to study because this solution is the most used in practice, being simple and providing enough design flexibility. Considering the specific problems of RF power amplifiers, the matching circuit connected between the electron device and the load must meet a number of requirements. It should, among other things:

a) Transform the load impedance into an equivalent resistance which is optimal for the electron device. Otherwise, the RF power amplifier would operate under unfavorable conditions, its efficiency might decrease, and the signal being transmitted might be distorted.

b) Maintain the specified amplitude- and phase-frequency response over a certain frequency range according to the type of intelligence being transmitted and in compliance with relevant standards.

c) Attenuate (filter out) the higher harmonics so that any of the harmonic's power delivered to the load (the input to the subsequent stage or the antenna of the final stage) will not exceed the maximum safe value.

d) Have insignificant power loss, that is, maintain a high efficiency. In addition, there are some requirements with regard to cost, size, weight, reliability and practicality.

Since the above requirements cannot be satisfied equally well, in practice they are divided into primary and secondary.

Classical Design of Narrowband Matching Circuits

The classical design method for narrowband matching circuits (with three ideal reactive elements) is based on choosing the circuit quality factor ([1] - [4]). Quality factor is correlated (to some extent) with circuit bandwidth, harmonic attenuation and efficiency of power transfer through the circuit. In order to emphasize the bad points of this method we will analyze some examples.

There are five matching circuits, (a through e, Figure 1), designed for the same source resistance (R_g=10 ohms), at center frequency f_0=100 MHz. Q=3 was chosen for circuits A, C and E, Q=4 for B and D circuits.

Table 1 contains the performances of the five matching circuits (with ideal reactive elements); thus we observe that bandwidth, harmonics attenuation and efficiency depend on both quality factor and (to a large extent) circuit configuration and load and source impedances.

For the same circuit and same load and source impedances, if a higher quality factor is chosen for the same circuit and same load and source impedances, a more selective circuit, with a lower power transfer efficiency, is obtained. Therefore, *the quality factor does not offer an accurate quantitative information about circuit performances.* On the other hand, neglecting losses in reactive elements results in significant errors, especially if the coil quality factor is low, or the frequency is relatively low.

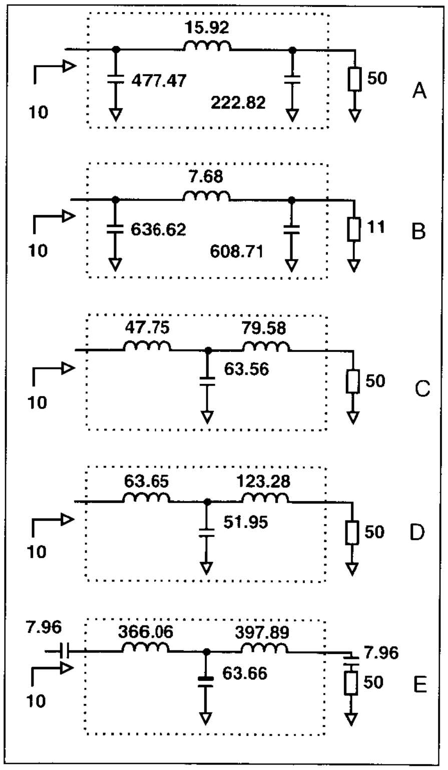


Figure 1. Five matching circuits designed for the same source resistance and center frequency.

Proposed Design Method

The method for narrowband matching circuit design proposed in this paper presents two important advantages:

a) The design parameter used is the efficiency of power transfer through circuit. This is very important in every situation and, especially, when high power is transferred. This parameter has a precise physical significance irrespective of circuit configuration and load and source impedances.

b) It takes into account the finite quality factor of inductances from circuit, reducing (sometimes significant) design errors. We note that because we know the order-of-magnitude of the of the inductance in the circuit, we know the inductors' quality factors with adequate precision.

Taking into account the requirements that narrowband matching circuits used in RF power amplifiers must meet, we analyze here only lowpass and band-

CIRCUIT	A (Q=3)	B (Q=4)	C (Q=3)	D (Q=4)	E (Q=3)
a_2 [dB]	29.04	27.33	18.63	25.85	47.7
a_3 [dB]	41.06	39.37	30.37	37.78	60.96
η [%]	90.9	92.4	96	94.5	73.3
f_s [MHz]	109.2	111.1	123.7	116.6	103.8

Note:
- the second and third harmonics attenuation (a_2, a_3, in dBc), and the 3 dB upper frequency limit of band, have been calculated for source resistance R_g=10 ohms;
- power transfer efficiency through circuit has been calculated neglecting lost power in capacitors; for coils, a quality factor Q_0=100 is assumed;

Table 1. Performance of matching circuits in Figure 1.

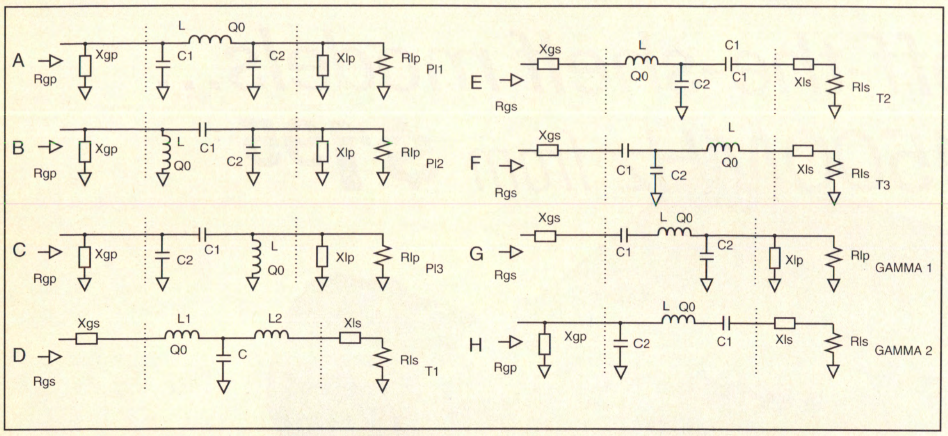


Figure 2. Circuit configurations used in the new design method.

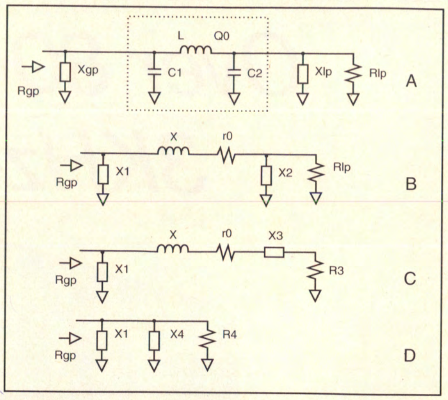


Figure 3. Transformation of PI_1 circuit.

pass matching circuits. Circuit configurations are presented in Figure 2.

Depending on matching circuit configuration, load and source impedances are modelled in the most suitable way: resistive-reactive groups, parallel or series.

The efficiency calculation only takes into account losses in coils, making the assumption that thet have finite quality factor, Q_0.

Calculations involved in obtaining design relations are simple (successive transformations resistance-reactance series parallel and solving nonlinear equation systems) but tedious. The only example presented here will be the way for obtaining the system of equations and solutions for PI_1 circuit, from Figure 2.

The requirements for theoretical realizability impose restrictions (specific to each circuit configuration) concerning load and source impedances, efficiency and quality factor Q_0. Because these requirements are expressed as complicated and irrelevant inequalities (from a practical point of view) we give up presenting them. The program directly computes the values of circuit elements, and when non-real or negative values are computed the decision is that the circuit cannot be practically achieved.

Design example. PI_1 circuit

The original circuit is shown in Figure 3a. In Figure 3b, X_1 and X_2 denote, respectively, the reactances corresponding to the parallel groups (C_1, X_{gp}) and (C_2, X_{lp}), while X corresponds to the coil reactance.

Transforming the parallel group (R_{lp}, X_2) into its series equivalent (R_3, X_3) results in the circuit shown in Figure 3c. The efficiency for the matching circuit is:

$$\eta = \frac{R_3}{R_3 + r_0} = \frac{R_{1p}X_2^2Q_0}{R_{1p}X_2^2Q_0 + X\left(R_{1p}^2 + X_2^2\right)} \quad (1)$$

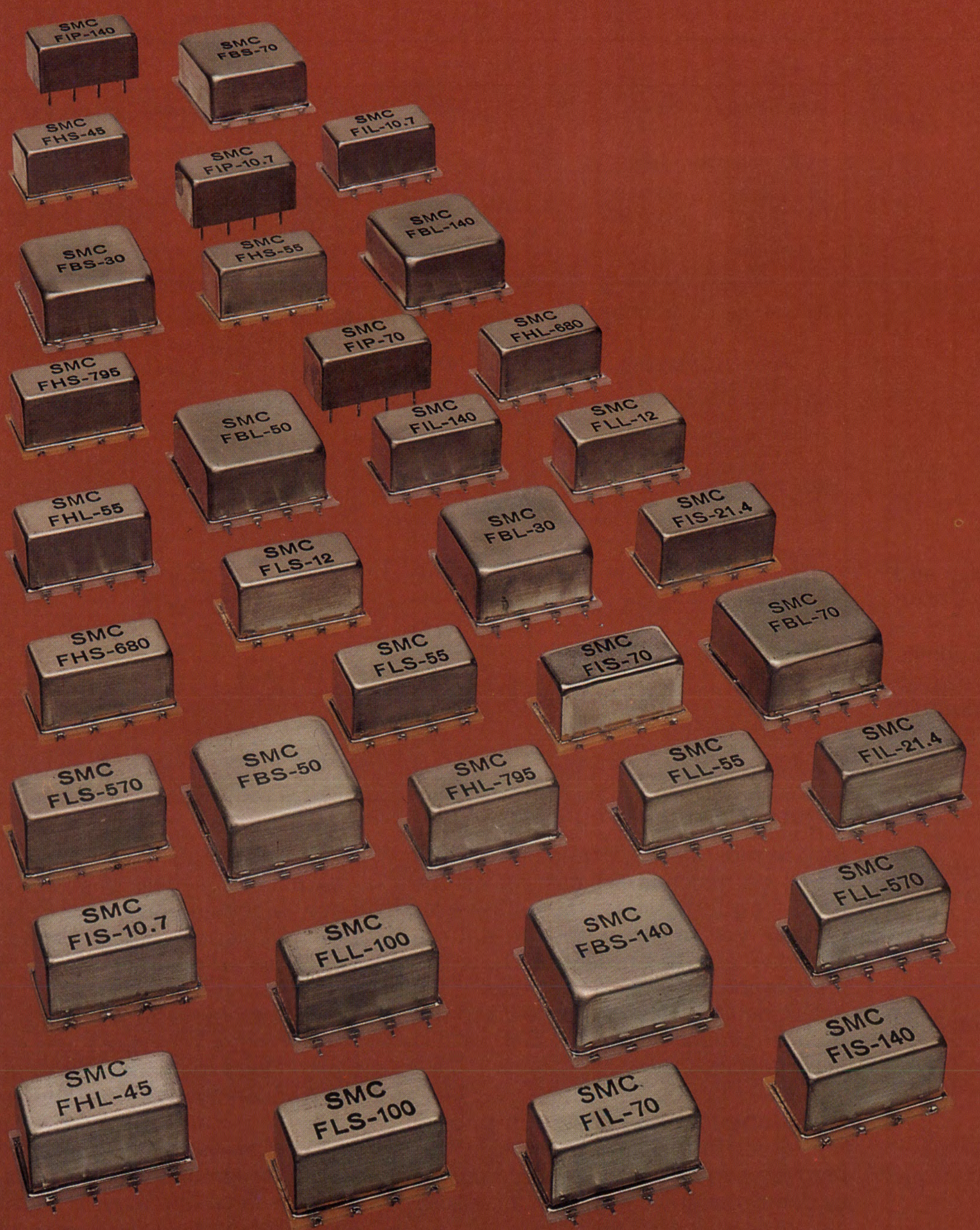
The circuit shown in Figure 3d is derived from the circuit shown in Figure 3c, transforming the series group R_3+r_0, $X+X_3$ into the parallel equivalent (R_4, X_4); for this circuit, we can write the matching conditions:

$$\begin{cases} X_1 + X_4 = 0 \\ R_{gp} = R_4 \end{cases} \quad (2a,b)$$

The solutions for the system of equations (1) and (2) are:

$$X_1 = -\frac{R_{gp}X_2}{Q_0(1-\eta)X_2 + \eta R_{1p}} \quad (3a)$$

FILTERS



Synergy has introduced a selection of high performance, low cost standard catalog and custom highpass, lowpass and bandpass filters covering passbands within the DC to 2000 MHz range. These filters are readily available in standard pin packages as well as the newest Synergy patented metal surface mount package with either leaded or flush mounting configurations.

Most importantly, when you specify Synergy, you benefit from our Sales and Applications team to assist you in the wide choices of available filters.

For a copy of our latest catalog, contact:
SYNERGY MICROWAVE CORPORATION,
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Paterson, NJ 07504.
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INFO/CARD 45

$$X = Q_0 \frac{1-\eta}{\eta} \frac{R_{1p} X_2^2}{R_{1p}^2 + X_2^2} \tag{3b}$$

$$X_2 = R_{1p}\left(\frac{-\eta Q_0 (1-\eta)}{1+Q_0^2(1-\eta)^2 - k\eta} \pm \frac{\sqrt{\eta[kQ_0^2(1-\eta)^2 + (k\eta - 1)(\eta - k)]}}{1+Q_0^2(1-\eta)^2 - k\eta} \right) \tag{3c}$$

where

$$k = \frac{R_{gp}}{R_{1p}} \tag{4}$$

Remarks:

From X_1 and X_2 we extract X_{gp} and X_{lp}, respectively, and then the resulting C_1 and C_2

In some situations the problem admits two solutions. If the load or source impedances don't have an inductive part, the problem has at most one solution.

If

$$Q_0(1-\eta)X_2 + \eta R_{1p} = 0 \tag{5}$$

then the problem has solution only if $X_{gp}>0$; in this case the solution is

$$C_1 = \frac{1}{\omega X_{gp}} \tag{6}$$

X_1 and X_2 are found using equation 3.

If

$$1+Q_0^2(1-\eta)^2 - k\eta = 0 \tag{7}$$

then:

$$X_2 = \frac{(k-\eta)R_{1p}}{2(1-\eta)Q_0} \tag{8}$$

X, X_1 are given in (3); if $X_{gp}>0$ we have, in addition, the solution:

$$X = \frac{Q_0(1-\eta)R_{1p}}{\eta} \tag{9a}$$

$$X_1 = -\frac{R_{gp}}{Q_0(1-\eta)} \tag{9b}$$

$$C_2 = \frac{1}{\omega X_{gp}} \tag{9c}$$

Program Description

Based on the new design method we wrote an IBM PC - GWBASIC program meant for the design of narrowband matching circuits, with the limitation that only resistive or capacitive load and source impedances are accepted. In fact, in most cases, impedances which are to be matched (at least for not too high frequencies), are resistive or capacitive. In addition, the source impedance is given as an RC parallel impedance (usually, the source is an active device which is modeled with a parallel capacitance or possibly a null capacitance).

```
PI NETWORK

F   = 100 MHz
R1, C1 series/parallel - p
R1  = 50 ohms
C1  = 20 pF
Rg  = 25 ohms
Cg  = 10 pF
ETA = 97%
Q0  = 150

              L
Rg   =>  Cg  C1  C2  Cl  Rl

L  = 42.213 nH (26.523 ohms)
C1 = 104.619 pF
C2 = 68.588 pF

FREQUENCY ANALYSIS
- POWER ATTENUATION
(I) = IDEAL CURRENT SOURCE
(U) = IDEAL VOLTAGE SOURCE
- INPUT IMPEDANCE - MODULE [ohms] & PHASE [deg]
-REAL & IMAGINARY PART [ohms]

HARMONIC 2
ATEN (I) = 26.545 dB
ATEN (U) = 16.893 dB
Zin = 8.228 EXP (j* -89.541) = .066 + j* -8.228

HARMONIC 3
ATEN (I) = 38.826 dB
ATEN (U) = 24.74 dB
Zin = 4.939 EXP (j* -89.936) = .006 + j* -4.939

HARMONIC 4
ATEN (I) = 46.854 dB
ATEN (U) = 30.007 dB

Zin = 2.838 EXP (j* -89.989) = .001 +j* -2.838

FREQUENCY = 101 MHz
ATEN (I) = -.269 dB
ATEN (U) = .284 dB
Zin = 26.646 EXP (j* -2.411) = 26.622 +j* -1.121

FREQUENCY = 102 MHz
ATEN (I) = -.516 dB
ATEN (U) = .567 dB
Zin = 28.32 EXP (j* -5.242) = 28.201 +j* -2.588

FREQUENCY = 103 MHz
ATEN (I) = -.73 dB
ATEN (U) = .848 dB
Zin = 29.978 EXP (j* -8.492) = 29.649 +j* -4.427

PI NETWORK

F   = 100 MHz
R1, C1 series/parallel - p
R1  = 50 ohms
C1  = 20 pF
Rg  = 25 ohms
Cg  = 10 pF
ETA = 97%
Q0  = 150

              L
Rg   =>  Cg  C1  C2  Cl  Rl

L  = 17.153 nH (10.778 ohms)
C1 = 1446.693 pF
C2 = 162.213 pF

FREQUENCY ANALYSIS
- POWER ATTENUATION
(I) = IDEAL CURRENT SOURCE
```

Figure 5. Example output.

Frequency Analysis Assumptions

In order to facilitate the design, the program can make an optional frequency domain analysis based on the following observations:

-It is assumed that the coils' quality factor and elements' values are constant versus frequency. Also, load resistance and capacitance may be considered constant with frequency either in the series or the parallel modelling, depending on design data input.

-In parallel modelling, source capacitance is considered constant with frequency. Therefore, output data are useful only as a first cut, because, in practice, load and source reactance are seldom constant with frequency.

-The program computes power attenuation on load (on harmonics or at a certain frequency) based on the fact that either the circuit is driven by a current source (it models a non saturated active device), or that the circuit is driven by a voltage source (it models an active voltage switching device). Power attenuation computed in this way characterizes only the matching circuit (including source reactance); the drive is assumed to be constant with frequency.

-The program is able to compute the input impedance of the circuit both for harmonics and in arbitrary frequency intervals. Input impedance is computed as modulus and phase, real and imaginary part.

```
frequency F [MHz] = ? 100
load impedance: Rl, Cl parallel/series (P/S) ? p
load resistance Rl [ohms] = ? 50
load capacitance Cl [pF] = ? 20
source resistance rg [ohms] = ? 25
source cap. Cg [pF] - (parallel with Rg) = ? 10
efficiency ETA [%] = ? 97
inductances quality factor Q0 = ? 150
harmonics analysis ? (Y/N) ? y
Fmin...Fmax analysis ? (Y/N) ? y
Fmin [MHz] = ? 101
Fmax [MHz] = ? 103
step Df [MHz] = ? 1
```

Figure 4. Example input screen.

Program Operation

To run the program, call GWBASIC MATCH from the prompt and answer the on-screen questions. Note that:

1. Load impedance is given as a RC series or parallel impedance; if $R_l C_l$ is a parallel group, C_l may be zero.

2. Source impedance is given as a RC parallel impedance; C_g may be zero.

(U) = IDEAL VOLTAGE SOURCE
- INPUT IMPEDANCE - MODULE [ohms] & PHASE [deg]
-REAL & IMAGINARY PART [ohms]

HARMONIC 2
ATEN (I) = 16.369 dB
ATEN (U) = 2.549 dB
Zin = 5.093 EXP (j* -83.594) = .568 + j* -5.061

HARMONIC 3
ATEN (I) = 21.342 dB
ATEN (U) = 2.95 dB
Zin = 3.008 EXP (j* -86.572) = .18 + j* -3.003

HARMONIC 4
ATEN (I) = 24.3 dB
ATEN (U) = 3.086 dB
Zin = 2.174 EXP (j* -87.606) = .091 +j* -2.172

HARMONIC 5
ATEN (I) = 26.443 dB
ATEN (U) = 3.149 dB
Zin = 1.711 EXP (j* -88.14601) = .005 +j* -1.71

FREQUENCY = 101 MHz
ATEN (I) = -.047 dB
ATEN (U) = .077 dB
Zin = 25.361 EXP (j* -4.94) = 25.267 +j* -2.184

FREQUENCY = 102 MHz
ATEN (I) = -.028 dB
ATEN (U) = .151 dB
Zin = 25.52 EXP (j* -9.850001) = 25.144 +j* -4.366

FREQUENCY = 103 MHz
ATEN (I) = .057 dB
ATEN (U) = .223 dB
Zin = 25.483 EXP (j* -14.656) = 24.653 +j* -6.448

3. If we choose "y" for "harmonics analysis", the program computes power attenuation on load (in dB under carrier) and the input impedance of the circuit (as module and phase, real and imaginary part), for the first four harmonics. Power attenuation is computed on two assumptions: a) the circuit is driven by a current source, and b) the circuit is driven by a voltage source - see Figure 5.

4. If we choose "y" for "Fmin...Fmax analysis", the program computes power attenuation on load and the input impedance of the circuit, in an arbitrary frequency interval (for Fmin to Fmax with step Df).

Three different networks can be chosen, PI networks, corresponding to configurations PI1, PI2 and PI3; T networks, corresponding to circuit configurations T1, T2 and T3; and GAMMA networks, corresponding to circuit configurations GAMMA1 and GAMMA2;

The calculations are performed for all circuit configurations which can be theoretically achieved.

After the program is run the user is given the option to print the results to a printer, complete anothert run, modify one or more parameters, or chose another network type (using the same parameters).

MATCH is available through the RF Design Software Service. For ordering information, see page 89 **RF**

References

1. F. Davis, "Matching Network Designs with Computer Solutions", Motorola Application Note AN-267
2. Becciolini, "Impedance Matching Networks Applied to RF Power Transistors", Motorola Application Note AN-721
3. S. Novak, "CAD for Lumped Element Matching Circuits", *RF Design*, February 1989
4. Krauss, H.L., Bostian, C.V., Raab, F.H., *Solid State Radio Engineering*, John Wiley & Sons, 1980

About the Author

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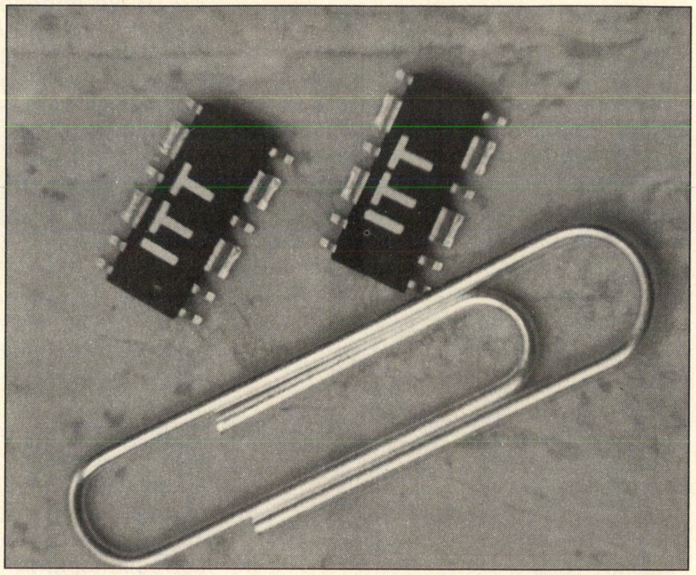
1 W, 3.3 V Power Amplifier IC

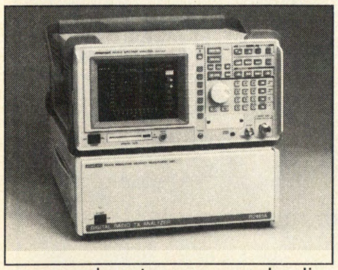


ITT GTC announces the availability of the second of a series of high gain RF power amplifier ICs targeted to meet the low cost, high reliability needs of the wireless telecommunications marketplace. Designated the ITT332 101BD, the new GaAs RF IC two-stage power amplifier was designed for cellular radio power amplifier applications in the 824 to 849 MHz band. The ITT332101BD produces 1 watt of RF power output when driven with 1 milliwatt (30 dB power gain) while operating from a single 3.0 volt battery. The single battery operation not only reduces the overall power requirements of the equipment, but it also eliminates the RFI generated by negative power supply oscillators. The surface mount cellular radio IC is capable of sustaining any mismatch without damage or spurious response. It operates in the "compressed" class A mode which means that the output is driven very hard resulting in increased efficiency. The device is mounted in the industry standard 16-pin SOIC package. Engineering samples are available now, production quantities are scheduled for mid-Summer and will be priced less than $9.00 in quantities of 10,000.

ITT GTC
INFO/CARD #250

Digital Radio Analyzers

Tektronix has introduced the Advantest R2465A and R2471A digital radio transmitter analyzers for North American Dual-mode Standard transmitters. The R2465A consists of the Advantest R3265A spectrum analyzer (100 Hz to 8 GHz) and R3541A modulation accuracy measurement unit. The R2471A consists of the R3271A spectrum analyzer (100 Hz to 25.5 GHz) and the R3541A. Spectrum analysis measurements include frequency accuracy, output power, and harmonic and spurious signal levels. Time-domain and time-gated spectrum analysis functions include burst rise and fall time,

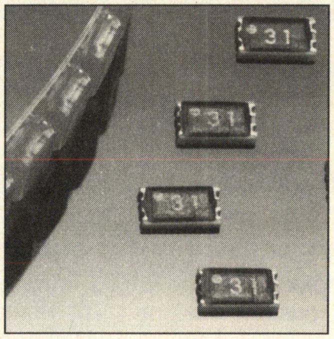


average burst power, and adjacent channel power. The DSP-based R3541A modulation accuracy measurement unit down converts, digitizes, and digitally demodulates $\pi/4$ DQPSK signals. The R2465A digital radio transmitter analyzer is priced at $52,000, and the R2471A is at $58,000.

Tektronix
INFO/CARD #249

Surface Mount Couplers

Two chip couplers, models D3-3120 and 3220, are the latest in a new line of low cost surface mount components from ST Olektron Corp, fit a variety of PCN, GSM and medical applications. These 10 and 20 dB devices can be used as BIT cir-



cuits or as a method of power-monitoring transmit and receive paths. Frequency range is 800 - 2000 MHz with an insertion loss of 0.4 dB and VSWR of 1.2:1. Coupling varies from 21 dB at 800 MHz to 14 dB at 2000 MHz. Guaranteed maximum power handling is 15 W from 800 to 1000 MHz, 5 W from 1400 to 1600 MHz, and 3 W from 1800 to 2000 MHz. Maximum power dissipation for the coupler termination resistance is 200 mW. The ceramic chip device measures 0.18 x 0.10 inches. Price is $2.25 from 25 to 1000 pieces. Delivery is from stock.

ST Olektron Corp.
INFO/CARD #248

ISM Diversity Antenna

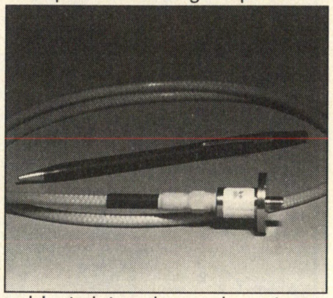
Teledyne Electronic Technologies' model TFE-1015 antenna series provides diversity reception for indoor propagation environments. The antenna covers the 2.400 to 2.483 GHz band with two antennas, one vertically polarized and the other horizontally polarized. An integrated diversity select switch (a SPDT switch) switches in 1 μs with a $\pm$5 V control voltage. Antenna gain is 2 dBi for both the horizontal and vertical antennas. Maximum

VSWR is 2.5:1. Maximum control current is 3 mA, with 1 mA typical. The TFE-1015 comes with an RG-158 cable with SMA male connector. The substrate holding the two antennas measures 3.6 x 3.6 inches.

Teledyne Electronic Technologies
INFO/CARD #247

Rotary Joint Cable

A coaxial cable with built-in rotary joint makes for quicker, more convenient connections in test sets, pedestals and other hard to reach places in microwave systems. The single-channel Kevlin Ro-Cable© should come in especially handy during acceptence testing. It prevents



cable twist or loosening at one end as the other end is tightened. It also eleminates cable twist and rotation due to moving test sets and instrumentation during testing or set-up. It can minimize reading errors caused by such twisting. Usable at up to 26 GHz, Ro-Cable comes in standard 2- and 3-ft lengths, flexible or semi-rigid, with SMA compatible connectors at each end. Typical VSWR is 2.00:1, and insertion loss is 3.10 dB. Other lengths or connectors are available as specials. Specify rotary-joint connector model 21168 for up to 8 GHz, and model 21166 for up to 26 GHz.

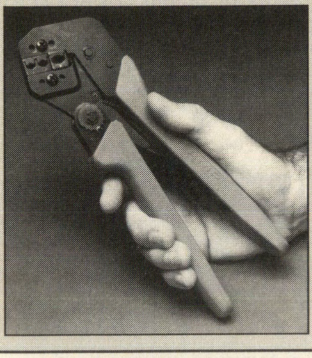
Kevlin Microwave Division
INFO/CARD #246

Product Spotlight: Hand Tools

Low Effort Crimper

The AMP® Pro-Crimper™ II hand tool requires 40 percent less hand force than similar industry-standard tools, while offering high reliability termination of a broad range of products. Its ratchet control ensures a complete crimping cycle. The tool, which uses interchangable dies, will crimp terminals and splices, various coaxial connectors, and many other connector types. Pricing ranges from $65 to $110, for a tool assembly with die for the most popular styles.

AMP, Inc.
INFO/CARD #245



Crimp Tool and Dies

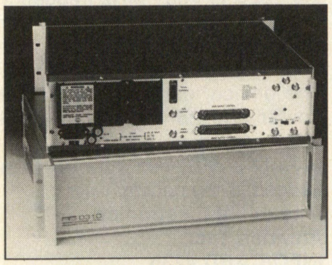
RF Industries' RFA-4005 is supplied with one RFA-4005-020 crimping tool frame and two die sets, the RFA-4005-01 and RFA-4005-02. The RFA-4006 contains two crimping tool frames and the same two die as the RFA-4005. The RFA-4005-01 die crimps RG58/U, RG59/U, RG142/U, RG8X, Proflex and various video cables. The RFA-4005-02 die crimps RG8/U, RG123/U, and Belden 9913 cables.

RF Industries, Ltd.
INFO/CARD #244

TEST EQUIPMENT

310 MHz Dual-Channel Synthesizer

The PTS D310 is a broadband, dual-channel instrument containing two fully independent low phase-noise, low spurious output, fast switching frequency synthesizers, each covering 100 kHz to 310 MHz with 0.1 Hz resolution. Output power is up to +13 dBm. Each channel is controlled through a 50-pin parallel interface which is pin compatible with the interface on current PTS synthesizers.



Programmed Test Sources, Inc.
INFO/CARD #243

12-Digit Counter

The new HP 53132A, a 225 MHz universal counter, provides 12 digit per second measurements. The new counter also speeds computer controlled tests with 200 fully formatted measurements per second. In addition to frequency, period and time interval, the HP 53132A can measure pulse parameters, duty cycle, frequency ratios and more. An optional third channel provides frequency measurements of up to 3 GHz. The HP 53132A universal counter is priced at $2400.

Hewlett-Packard Co.
INFO/CARD #242

Cellular Service Monitor

The µCELL-100 from IFR Systems is a test system for North American Cellular mobile telephones, supporting AMPS, NAMPS and the IS-54 systems, with more systems on the way. Both automatic and manual tests can be performed, and an identification function allows quick checking of MIN, ESN and telephone capabilities. Generator and receiver systems operate from 869.01 to 893.97 MHz and 824.01 to 848.97 MHz, respectively, with 0.01 MHz rsolution.

IFR Systems, Inc.
INFO/CARD #241

Calibration Kit

Maury Microwave announces the availability of the series

8450H precision SC connector calibration kits for vector network analyzers (VANAs). The series 8450 are designed for use with a variety of VANAs which utilize 3.5 mm or 2.92 mm (K) test set or test cable connectors. The series 8450H kits are provided with both male and female calibration standards including sliding and fixed terminations, open circuits, and short circuits, along with calibration constants provided on the storage medium appropriate for the VANA being calibrated.
Maury Microwave Corp.
INFO/CARD #240

CABLES & CONNECTORS

High Shielded Coax

SuperFlex high shielded coax assemblies operate through 18 GHz with insertion loss of 1 dB/ft at 18 GHz and typical VSWR of 1.30:1. The cable has a diameter of 0.125 inches and minimum bend radius of 0.25 inches, with excellent electrical stability versus flexure. Its Quadraform shield design maintains RF leakage below –85 dB through 18 GHz.
Storm Products Co.
INFO/CARD #239

Plug-in Contact SMA

Coaxial Components Corp. has announced the release of a "pre-assembled" SMA plug for RG402 (0.141 dia.) semi-rigid cable. Model 3207A-01-1 is directly soldered to the cable after stripping back the center contact of the cable and plugging it into the female contact at the rear of the connector. The interface is in accordance with MIL-C-39012/SMA.
Coaxial Components Corp.
INFO/CARD #238

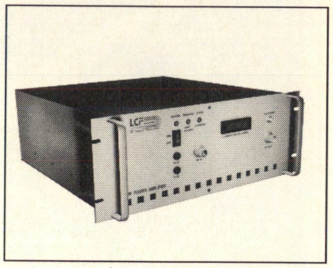
Low Loss Cables

Times Microwave announces the availability of LMR-200 and LMR-240 flexible communications cable. Although LMR-200's outer diameter is 0.195 inches, the same as RG-58, it has more than 30% lower attenuation, even lower than RG-8X. LMR-240's outer diameter is 0.240 inches, yet provides attenuation comparable to RG-8, a 0.405 inch OD cable. LMR cable uses a bonded Al tape and overbraid outer conductor with a foam polyethylene dielectric, providing lower loss and better RF shielding than conventional RG cables. Available from stock, LMR-240 is $0.40/ft, and LMR-200 is $0.32/ft.
Times Microwave Systems
INFO/CARD #238

AMPLIFIERS

10 W Broadband Amp

Model 1200-10-40R is a solid-state amplifier that delivers 10 watts of output power over the frequency range of 10 to 1200 MHz. The unit delivers 40 dB of gain. The amplifier system has



automated current limiting and will continue to operate into high VSWR loads and automatically recover. The amplifier system sells for $5950.
LCF Enterprises
INFO/CARD #237

InmarSat Power

The M67791, from the Electronic Device Group of Mitsubishi Electronics America, operates in the 1626 to 1661 MHz frequency range and functions as the driver stage for InmarSat satellite system mobile telephone uplinks. The device features 7 W output power and 25% minimum total efficiency on a 15 V power supply. The M67791 comes in a flanged, 5-pin package.
Mitsubishi Electronics America, Inc.
INFO/CARD #236

Linearized Multi-Channel Amp

Milcom International is now making available their field proven multi-channel feedforward amplifier. This state of the art amplifier offers up to 1 kW PEP with all intermodulation distortion products well within FCC emission mask requirements. Also available are the individual class A and linearized class AB modules from this amplifier, which feature high 3rd order intercept

points.
Milcom International, Inc.
INFO/CARD #235

Linear Hybrid Amp

Phoenix Microwave's model PA996 offers linear output power of 20 dBm from 50 to 500 MHz with typical third order intercept point of 35 dB. Minimum gain is 18.0 dB, and noise figure is 4.0 dB across the entire frequency range. The amplifier operates from +15 VDC and 160 mA. The unit is offered in a hermetic dual in-line or flatpack package.
Phoenix Microwave Corp.
INFO/CARD #234

AMPS Duplex Amplifier

KW Microwave introduces an AMPS full-duplex amplifier in the frequency range of 869 to 894 MHz for Rx and 824 to 849 for Tx. The Rx gain is 5 dB minimum. Output power at Tx frequencies is 1.5 W minimum for input of 600 mW typical. Input and output VSWR is 2.0:1 maximum and Tx/Rx isolation is 50 dB minimum. DC power consumption is 14 V at 1 A maximum. The KW Microwave part no. is INT-I-891-901-12666.
KW Microwave Corp.
INFO/CARD #233

SEMI-CONDUCTORS

DSP Down-Converters in TAB Form

Harris' HSP45116, 16-bit numerically controlled oscillator, and its HSP43220, linear phase lowpass decimating filter, are now available in tape automated bonding (TAB) packages. TAB-mounted die permit easy handling, with testability comparable to conventionally packaged ICs. The TAB versions of the two devices are available fully tested in commercial and military temperature grades for attachment to printed circuit boards, chip-on-board assemblies, and multichip modules.
Harris Semiconductor
INFO/CARD #232

High Speed Clock Chip Family

The ClockWorks™ family of chips from Synergy Semiconductor consists of a range of frequency synthesizers, clock generators, clock distribution and drivers, programmable delay lines, and phase locked loops. All members of the family interface with each other in PECL, and interface with the rest of the system through TTL. In quantities of 1000, prices range from $7.85 to $19.85.
Synergy Semiconductor
INFO/CARD #231

GSM Baseband Chipset

Analog Devices and the Technology Partnership of Cambridge, England, have announced a GSM chipset, integrating all the circuitry and software required to implement the baseband portion of a Phase 2 handset. The three-device set consists of a digital ASIC (the physical layer processor), an algorithm signal processor, and a dedicated mixed-signal device (the baseband converter). The parts will operate from 3 or 5 V supply voltages and will be available for sampling, along with an evaluation board, in the second quarter 1994.
Analog Devices, Inc.
Technology Partnership
INFO/CARD #230

12-bit A/D

Burr-Brown's ADS605 is a 12-bit sampling analog-to-digital converter that provides a wideband track/hold 12-bit quantizer, timing circuitry, and low drift internal reference, all in an economical 28-pin DIP. Both DC and dynamic AC specifications are guaranteed to Nyquist. The track/hold, with its 32 MHz full-power bandwidth, gives designers –79 dBc spurious-free dynamic range. The ADS605 dissipates 1.4W.
Burr-Brown Corp.
INFO/CARD #229

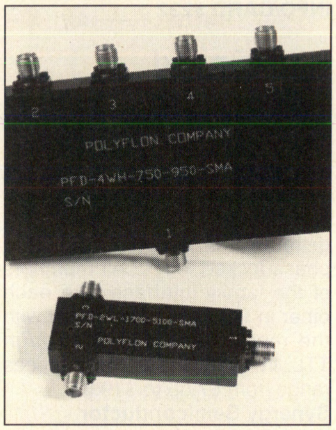
SIGNAL PROCESSING COMPONENTS

SP4T Switch

Mini-Circuits' GSWA-4-30DR is a GaAs SP4T switch covering DC to 3000 MHz, with specification limits 4.5σ typical from mean. Switching time is typically 25 ns, insertion loss is 0.9 dB, and typical isolation is 30 to 40 dB. Housed in a low-cost, 28-pin PLCC package, the GSWA-4-30DR costs $19.95 each.
Mini-Circuits
INFO/CARD #228

Power Dividers

Polyflon now offers two- and four-way power dividers. The power dividers are microstrip designs, computer optimized for lowest insertion loss, input and output mismatches, and highest isolation. Four-way dividers are available covering 824 to 890 MHz and 750 to 950 MHz. Amplitude balance is 0.1 dB. Two-way



dividers covering 1700 to 5100 MHz are available, with 0.1 dB amplitude balance.
Crane Polyflon
INFO/CARD #227

Bandpass Filter

KeL-Com bandpass filter model BP-140/35 has a minimum 3 dB bandwidth of 35 MHz and f_0 insertion loss less than 1.0 dB. Typical group delay variation is 2 ns over 135 to 145 MHz. Ultimate attenuation is 50 dB to 800 MHz. Typical VSWR is 1.5:1. Package dimensions are 0.8 x 0.4 x 0.4 inches, with radial leads for through hole mounting. Cost is $30.00 in production quantities.
KeL-Com
INFO/CARD #226

Terminated/ Reflective Switch

Daico Industries introduces the model DSW48024 SP4T terminated or reflective switch that features an operating frequency of DC to 2000 MHz. The switch provides isolation of 60 dB from DC to 500 MHz, 46 dB from 500 to 1000 MHz, and 33 dB from 1000 to 2000 MHz. VSWR is 1.3:1 with a switching speed of 30 ns (50% control to 10%/90% RF). The switch is controlled by a TTL driver in a 14-pin, 0.38 inch square flatpack.
Daico Industries, Inc.
INFO/CARD #225

Broadband Switches

QEC's SRF series switches are bi-directional and suitable for cable, RF and data applications. Signal routing can be controlled directly by 12 VDC applied via D-9 connector, or TTL interface. The switches' frequency range is DC to 1750 MHz, with typical 0.5 dB insertion loss, 50 dB iolsation and 14 dB return loss.
Quintech Electronics and Communications, Inc.
INFO/CARD #224

SIGNAL SOURCES

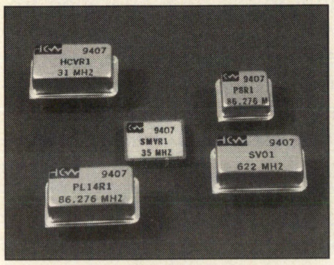
OCXO

Oak Frequency Control Group's 4879 OCXO offers ±5 x 10^{-8} temperature stability from

–20 to +70 °C, in a 1.5 inch square by 0.5 inch package. Aging is only ±2 x 10^{-9} per day. The 4879 is available between 5 and 20 MHz, and offers Sine or HCMOS output. Small quantity orders are available in four to six weeks.
Oak Frequency Control Group
INFO/CARD #223

VCO Oscillators

Conner-Winfield introduces an expanded line of voltage controlled oscillators. These units are available in both through-hole and surface mount packages, and cover a frequency range of 100 kHz through 1.6 GHz. They are available with



either squarewave or sinewave outputs. An example unit, the PL14R1/R2 is priced at $41.50 each in quantities between 10 and 24.
Conner-Winfield Corp.
INFO/CARD #222

OCXO

MTI-Milliren Technologies has released the 230 Series, a family of miniature OCXOs with high performace thermal stability, aging, and phase noise. The typical specs for a 10 MHz, SC-cut OCXO are thermal stability of 2.5 x 10^{-8} (from –30 to +70 ºC), aging of 7 x 10^{-10} per day, and phase noise lower than –85 dBc/Hz at a 1 Hz offset. The series uses either AT- or SC-cut resonators and is available at frequencies from 40 kHz to 40 MHz. Package dimensions are 1.42 x 1.07 x 0.76 inches. Prices range from $160 to $365.
MTI-Millren Technologies, Inc.
INFO/CARD #221

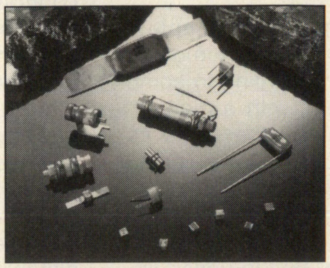
DISCRETE COMPONENTS

X-tals for PCMCIA

Ecliptek has introduced the ECCM1 series of surface-mount crystals in a state-of-the-art ceramic miniature package designed for all Type 2 and Type 3 PCMCIA applications. Standard frequency range is 11.0592 MHz to 40.000 MHz.
Ecliptek Corporation
INFO/CARD #220

Porcelain/Variable Capacitors

North American Capacitor Company introduces the Mallory/Microelectronics lines of multilayer ceramic and porcelain capaci-



tors in chip and leaded styles, and variable capacitors operable at frequencies up to 300 GHz. Q-factors exceed 10,000 at 1 MHz for the porcelain capacitors, while the variable capacitors have Qs ranging from 1200 to 10,000 at 100 MHz.
North American Capacitor Co.
INFO/CARD #219

Trimmer Capacitors

Sprague-Goodman has introduced two new polyimide dielectric trimmer capacitors to its FILTRIM® series; both offer a wide capacitance range of 25.0 to 600.0 pF. The model GZN60100 adjusts from the top, and the model GZP60100 adjusts from the side. Both are leaded, through-hole mounted. Pricing for either model is $7.50 each for 1000 pieces.
Sprague-Goodman Electronics, Inc.
INFO/CARD #218

Low Profile Crystal

Fox Electronics has further reduced the height of their surface mount crystals. The crystals are 1.1 mm high, including standoffs, and are available in the frequency range of 10 to 100 MHz. Standard tolerances are ±50 ppm at 25° C and from –10 to +60° C, with 10 ppm tolerances and other temperature ranges available.
Fox Electronics
INFO/CARD #217

A First Introduction to Frequency Hopping Spread Spectrum

By Gary A. Breed
Editor

Last month, this tutorial column reviewed the fundamentals of direct sequence spread spectrum. This month, the other major spread spectrum technique, frequency hopping, is introduced in a conceptual, block diagram form. This description is intended for readers who have no prior experience with spread spectrum.

Spread spectrum (SS) is a modulation technique that uses a wide bandwidth, transmitting a signal with low energy density per unit frequency. This characteristic can be used to improve reliability in low signal-to-noise conditions, to reduce interference to and from other users of the same region of the RF spectrum, and to introduce a measure of security through the complexity of the modulation scheme.

Last month's review of direct sequence SS showed that a modulated data signal can be spread over a very wide bandwidth using a pseudo-random chipping code. The random nature of the resulting signal creates sidebands which spread over a wide range of frequencies on either side of the carrier frequency. Therefore, in direct sequence SS, the signal has a specific carrier frequency, with information that is "hidden" in the data stream that has been altered by the PN code.

Frequency hopping, in its simplest form, creates an SS signal by moving the carrier frequency according to some pseudo-random pattern, with a modulation method that may be the same as direct sequence SS before the chipping code is introduced (BPSK or QPSK), or some other digital modulation method (FSK, MSK, OOK). In this case, the information is "hidden" by the uncertainty of the next transmitted frequency.

The advantage of frequency hopping is that the signal, although changing frequency, remains narrowband, with the excellent signal-to-noise performance and interference rejection afforded by narrow filters. The disadvantage is the complexity of the RF circuitry required to support very fast frequency changes. This can increase cost and complexity of the hardware, as well as limit the data rate that can be transmitted.

Generating Frequency Hopping Spread Spectrum

Figure 1 is the block diagram of a simple frequency hopping SS transmitter. The digital input goes to a conventional modulator, which generates an IF signal. The PN code is applied to logic that controls a fast-switching frequency synthesizer. The output of this synthesizer becomes the local oscillator of an upconversion stage that takes the modulated signal to the final operating frequency, with hopping that follows the synthesizer frequency changes.

In a current frequency hopping system, the hopping rate might be 1000 frequencies per second. To minimize predictability of the hopping sequence, the pseudo-random number (PN) code should be updated at perhaps 256 or 512 kchips per second. This increases the randomness of the code, making it far more difficult to determine the PN code by analyzing the frequency hopping pattern.

Data rates may be transmitted up to 1 Mbit/s using current commercial hopping technology. This means that up to 1000 bits will be transmitted during one hop (less pause time during the required settling time for a new frequency). Synchronization is required at the data rate in frequency hopping SS, rather than the much faster chipping rate in direct sequence SS.

The key element in frequency hopping

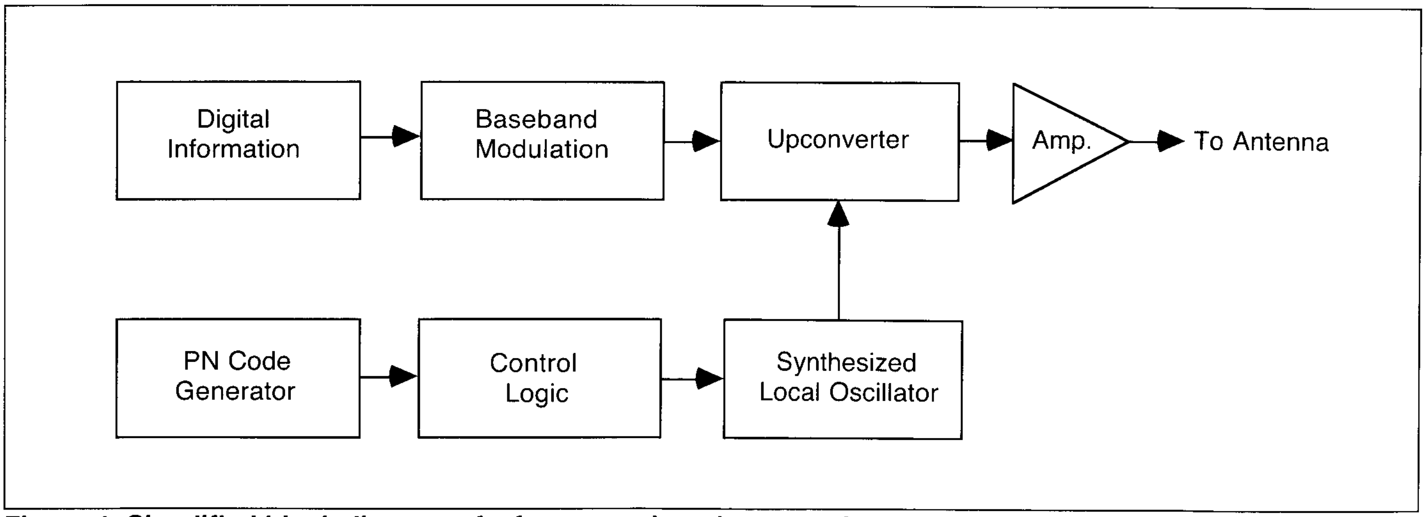


Figure 1. Simplified block diagram of a frequency hopping spread spectrum transmitter.

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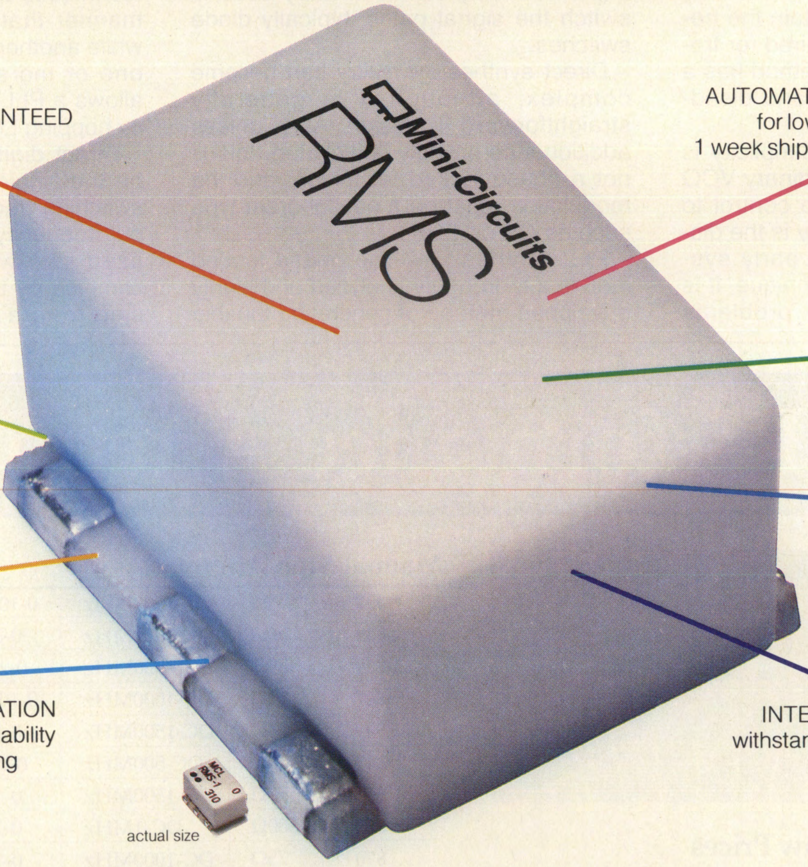

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SS systems is clearly the local oscillator, typically a frequency synthesizer, which must switch rapidly across the band of frequencies being used, with consistent settling time and low spurious content (especially in the receiver). These features may result in compromises in circuit complexity, power consumption and cost, which are major factors in current SS applications.

Fast-Switching Synthesizer Technologies

Direct synthesis, phase locked loop (PLL), direct digital synthesis (DDS), and even unlocked voltage controlled oscillators can be used to obtain the frequency switching speed required for frequency hopping SS. Each method has a different set of advantages and disadvantages:

Unlocked oscillator(s) — Simplicity is the advantage of using an ordinary VCO with some kind of open-loop control to change frequencies. Accuracy is the disadvantage. Although some early systems may have used this technique, it is not very practical. Obvious problems include repeatable tuning, drift due to environmental effects and calibration.

Direct synthesis — A bank of crystal oscillators, each one tuned to a hopping frequency, is the simplest form of a synthesizer for SS. More likely, such a direct synthesizer uses a combination of stable reference crystal oscillators which are multiplied, divided, mixed and filtered to create a programmable frequency output. Fast switching speed is achieved because there is no time constant inherent to direct synthesis — there is no feedback path to add a delay, making the switching speed dependent only on the circuitry used to switch the signal paths, typically diode switches.

Direct synthesis circuitry can become complex, although it is generally straightforward in design. Also, unless additional complexity is included, this is not a phase locked technique, and the modulation must be a noncoherent type such as AM or FSK.

PLL synthesis — The phase locked loop is a thoroughly analyzed and highly developed method of generating tunable signals. However, in frequency hopping, there is a necessary tradeoff between spectral purity and switching speed; they have an inverse relationship. However, techniques have been developed that minimize the performance degradation of high-speed PLLs. Such techniques include pre-steering of the VCO to minimize settling time; fractional-n techniques that allow a higher frequency reference and, thus, a wider loop bandwidth; and multiple-loop designs that achieve high resolution frequency steps from two or more fast synthesizers. The author is also aware of "ping-pong" techniques that use multiple PLLs in a manner that allows one PLL to settle while another is on-line as the LO. Using one or more of these enhancements allows a PLL to be an effective frequency hopping LO.

Direct digital synthesis — DDS may be the single greatest advance in fast-switching frequency synthesis. The output frequency is constructed from a digitized sine wave look-up table that is sampled by the digital circuitry. Conversion from a digital word to a voltage

5359

level is accomplished with a digital-to-analog converter (DAC). Frequency switching speed matches the sampling frequency of the system, and the architecture assures that switching is phase-continuous, minimizing transient behavior as a new frequency is selected.

The disadvantages of DDS are frequency range and spurious outputs. The output frequency is limited to a maximum of 1/2 the clock frequency. As a result, DDS systems are limited by the performance of the logic family used. CMOS allows economical DDS up to about 50 MHz, ECL provides a factor of two or three above that, and GaAs has been shown to operate up to 500-700 MHz. Of course, lower frequency DDS signals can mixed or multiplied to achieve higher frequencies.

Spurious outputs are more difficult to handle. Some systems can accept the presence of spurs that are 40-60 dB below the carrier, but this is unacceptable for many others. Selection of an optimum clock frequency and output band can also help. Another method is to use the DDS as a reference for a PLL. The purity of the DDS carrier makes it an excellent low noise reference, and the PLL loop filter can eliminate spurious energy.

Summary

This and last month's tutorials are intended only to start the reader on the way to understanding spread spectrum techniques. The references given, and other information sources are needed to complete the learning process.

The two main advantages of spread spectrum transmission are security and spectrum utilization, both of which are extremely important in today's common use of RF-linked voice and data paths.

Frequency hopping and direct sequence use different methods, and have different advantages and disadvantages. Hopping is a more difficult engineering task, but it has performance advantages in exchange. Direct sequence can be implemented at low cost, and is an excellent solution where the specific advantages of hopping are not important. Both are currently being developed for mass markets. *RF*

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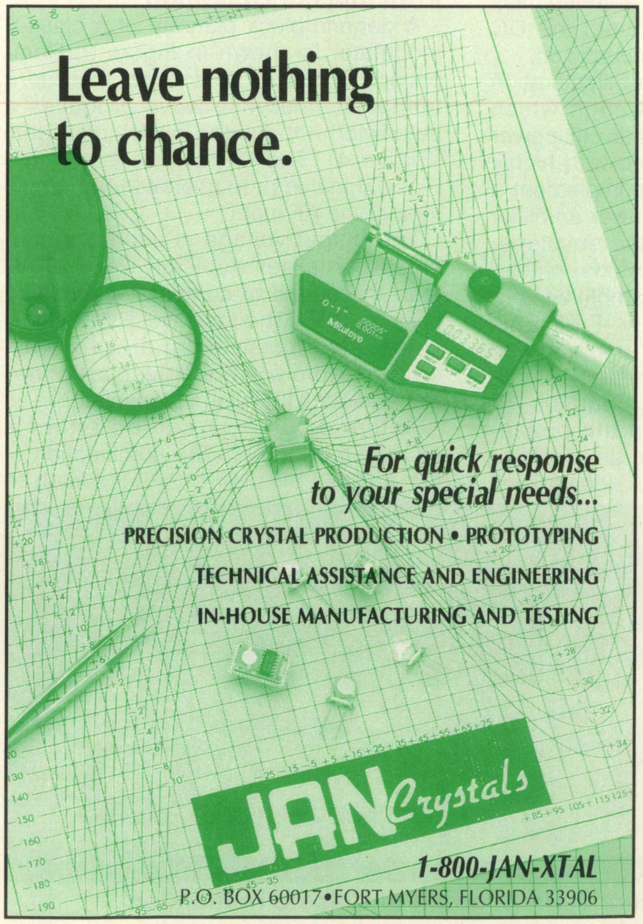

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Class-S High-Efficiency Amplitude Modulator

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

A class-S modulator is a high-efficiency audio-frequency (AF) power amplifier (PA) based upon pulsewidth modulation (PWM). Its output is a baseband AF signal that includes DC. Class-S modulators provide high-level amplitude modulation in transmitters for applications such as full-carrier AM, single sideband (by envelope elimination and restoration), and envelope tracking. This paper describes a 100-W class-S modulator with an envelope bandwidth of 57 kHz and an efficiency of about ninety percent.

A class-S modulator can be used for high level amplitude modulation (AM) of radio-frequency (RF) PAs in applications such as

- Full-carrier amplitude modulation (AM),
- Single sideband (SSB), and
- Envelope tracking.

This paper describes a 100-W class-S modulator with an envelope bandwidth of 57 kHz and an efficiency of about ninety percent.

Full-carrier amplitude modulation is used in applications such as non-directional beacons, AM broadcast, citizens' band, and aircraft communication. Amplitude-modulation techniques [1] include transformer-coupling of a high level AF signal onto the power-supply voltage, series-pass modulation of the supply voltage, and gate-bias modulation of the gain of the RF PA. The class-S modulator (Figure 1) is a series-pass device that can produce both the DC (carrier) level and the AF modulation.

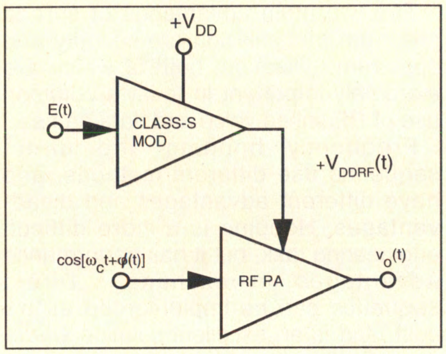

Figure 1. High-level amplitude modulation.

Envelope Elimination and Restoration (EER) is a technique through which highly efficient but nonlinear RF power amplifiers can be combined with highly efficient audio amplifiers to implement a high-efficiency linear RF power amplifier [2]. A limiter eliminates the envelope, allowing the constant-amplitude, phase-modulated carrier to be amplified efficiently by class-C, -D, -E, or -F RF PAs. The detected envelope is amplified efficiently by a class-S modulator. Amplitude modulation of the final RF PA restores the envelope to the phase-modulated carrier, creating an amplified replica of the input signal. Signals such as SSB can be produced efficiently by this technique.

Envelope tracking [3] uses a class-S modulator to control the power supply voltage of a linear RF PA. The efficiency of a linear RF PA increases with the amplitude of its output [1]. Envelope tracking keeps the power supply voltage just large enough to avoid saturation of the RF PA. The efficiency is therefore maintained just below the efficiency of the RF PA at peak-envelope-power (PEP) output. This can increase the efficiency of the transmitter by a factor of two or three for voice signals with large peak to-average ratios [4].

Principles of Operation

A generic block diagram of a class-S modulator is shown in Figure 2. Comparison of the AF (envelope) input to a triangular reference wave produces a pulsewidth-modulated switching signal (Figure 4). The DC component of the PWM signal is proportional to the amplitude of the AF input.

The output voltage is generally considerably larger (typically 28 - 100V) than the output of comparator A1 (typically 0 - 3 V). The low-level PWM signal is therefore translated to a higher voltage suitable for the driver by either Zener diodes [1] or an opto-isolator.

Active device Q1 can be either a MOSFET or a BJT. The combination of Q1 and D1 acts as a SPDT switch to

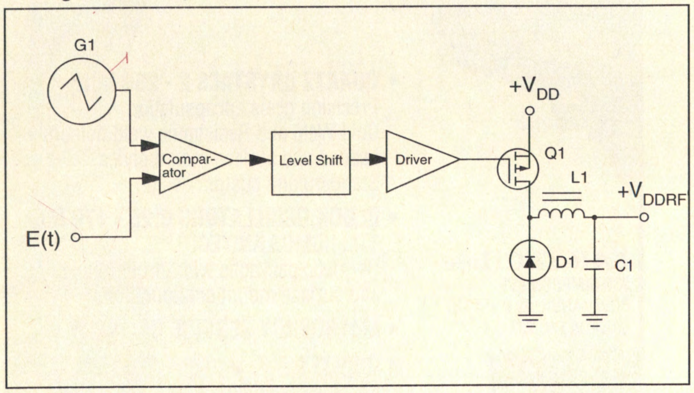

Figure 2. Class-S modulator.

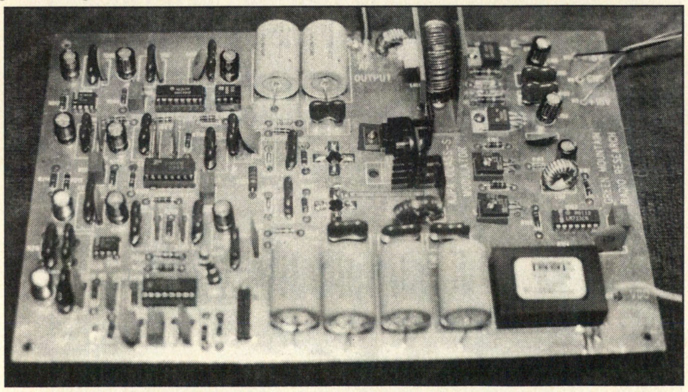

Figure 3. Photograph of Class-S modulator.

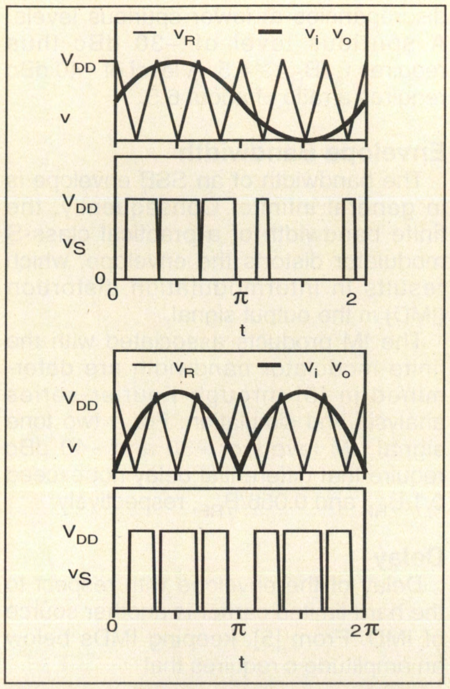

Figure 4. Waveforms for single-tone AM (top pair) and two-tone SSB (bottom pair)

produce a high-level PWM signal. Current flows through Q1 when it is on, and through D1 when Q1 is off (Figure 14-12 of [1]). To achieve fast switching, it is generally necessary to employ pull-up/pull-down circuits in the driver.

The high-level PWM signal is converted into the desired analog voltage by the low-pass output filter (L1-C1). The desired DC and AF components of the switching waveform pass through the filter to the load with minimal alteration. However, the filter presents a high impedance to the switching frequency and its harmonics. This prevents them from reaching the load and also from generating significant currents through the switch.

The linearity of a class-S modulator depends primarily upon the linearity of the triangular reference wave and the switching speed. The switching speed per se does not degrade linearity at intermediate amplitudes. However, operation near zero- and PEP-output levels requires duty ratios near 0 and 100 percent. Maintaining the proper shape of the relatively narrow pulses requires fast switching.

Requirements for Full-Carrier AM

Sampling theory requires that the switching frequency f_s of a class-S PA be at least twice the highest frequency to be amplified. However, spurious prod-

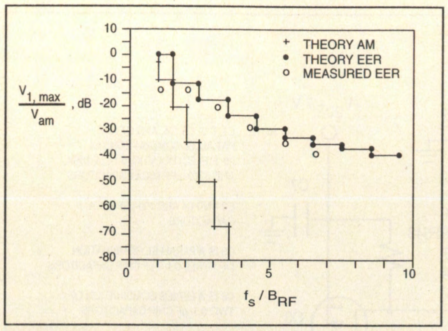

Figure 5. Maximum spurious products.

ucts inherent in pulsewidth modulation generally necessitate a somewhat higher switching frequency. In most applications, the higher switching frequency is desirable as it results in a smaller output filter with less delay.

The signal produced by two-sided pulsewidth modulation with voltage levels of 0 and V_{DD} is (from Appendix 14-3 of [1])

$$v_{D2}(\theta) = \frac{2V_{DD}}{\pi}\left[\frac{y}{2} + \sum_{k=1}^{\infty} z_k(\theta_m)\left\{\cos\frac{k\pi}{2}\sin k\theta + \sin\frac{k\pi}{2}\cos k\theta\right\}\right] \quad (1)$$

where $\theta = \omega_s t$ represents angular time in terms of the switching period and y is the instantaneous half-pulsewidth ($0 \le y \le \pi/2$).

The DC term varies linearly with pulsewidth and produces the desired output signal. However, the amplitudes of the switching frequency and its harmonics vary as

$$z_k(\theta_m) = \frac{1}{k}\sin[ky(\theta_m)] \quad (2)$$

where $\theta_m = \omega_m t$. The sine-function nonlinearity produces the spurious products that are inherent in PWM.

For 100-percent modulation by a sinusoid of frequency f_s, the instantaneous half pulsewidth is

$$y = \frac{\pi}{2}(1+\cos\omega_m t) \quad (3)$$

The spectrum is obtained by insertion of (3) into (2) followed by trigonometric substitution. From (14-39) of [1], the amplitude of the nth spurious product associated with the kth harmonic of the switching frequency is

$$|V_{k,n}| = \frac{2V_{DD}}{k\pi} J_n(k\pi) \quad (4)$$

The amplitudes of various spurious products are shown in Figure 14-15 of

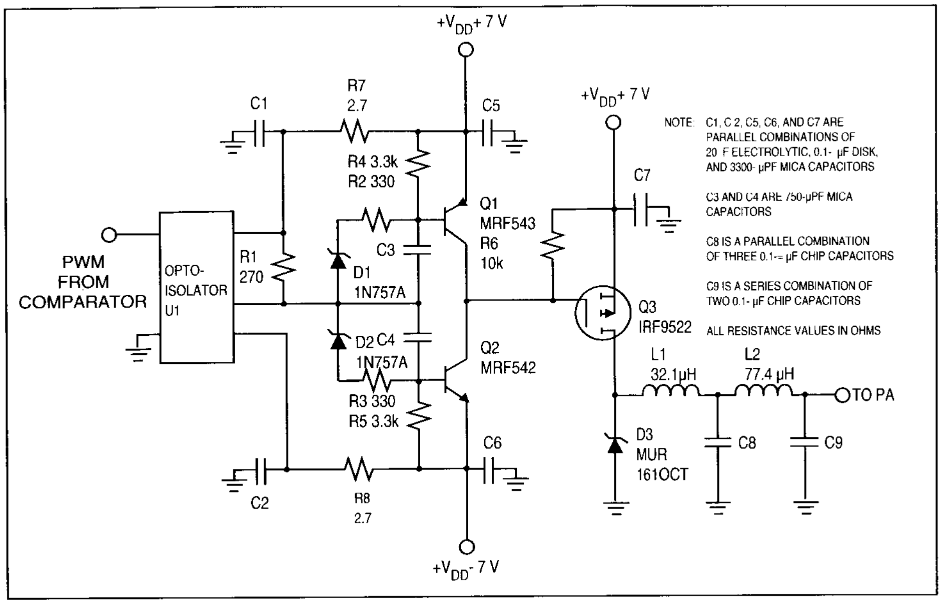

Figure 6. Output circuit schematic.

[1] as functions of the modulation index.

Those spurious products that fall within the passband of the output filter reach the load and are become spurious products in the RF output. While there are a large number of spurious products, those of the lowest order tend to have the largest amplitudes. It is therefore convenient to set f_m to the maximum-output (cut-off) frequency of the output filter.

Figure 5 shows the variation of the largest in-band spurious product with the ratio of the switching frequency to the RF bandwidth, which is

$$B_{RF} = 2f_m \tag{5}$$

Keeping the spurious products at least 30- or 40-dB below the carrier requires the switching frequency to be four or five times the highest modulation frequency, respectively (2 or 2.5 times the RF bandwidth).

Requirements for SSB

The design of a class-S modulator for an SSB transmitter requires attention to be given to

- Spurious products produced by PWM,
- Envelope bandwidth, and
- Delay introduced by the output filter.

Spurious Products

For EER with a two-tone signal,

$$y(\theta_m) = aE(\theta_m) = ac(\theta)\cos\omega_m t \tag{6}$$

where $c(\theta)$ is a cosinusoidal +1 switching function and a is the PEP amplitude. The general approach to analysis of the PWM/EER spectrum is similar to that of the previous section. However, the spectrum now has three sets of components:

- The switching frequency and its harmonics,
- An FM-type modulation spectrum on each harmonic, and
- A switching-function spectrum upon each component of the FM-type spectrum.

The analysis is somewhat tedious. Fortunately, only odd-order Bessel functions are present, for which

$$J_{2n+1}(kac(\theta_m)) = c(\theta_m)J_{2n+1}(ka) \tag{7}$$

After a number of substitutions and rearrangements, the magnitudes of spurious products of the kth harmonic are given by

$$V_{k,2n+1+b} = \frac{4V_{DD}}{\pi^2 k} \bullet \tag{8}$$

$$\sum_{n=0}^{\infty}\left[(-1)J_{2n+1}(ka)\sum_{b=1,3,\cdots}^{\infty}\frac{1}{b}(-1)^{\frac{b-1}{2}}\right]$$

The order of these products is 2n+1+b, hence the resultant frequencies are $kf_s \pm (2n+1+b)f_m$. Index b is the order of the harmonic components of the switching waveform.

The maximum in-band spurious product (Figure 4) is found by computer. The theoretical predictions and laboratory measurements are in excellent agreement for spurious levels of –40 dBc or more. It is thought that errors in the calculation of Bessel functions of large arguments are responsible for the small discrepancies at lower spurious levels. A spurious level of –30 dBc thus requires $f_s/B_{RF} \geq 4.5$. A level of –40 dBc requires a ratio of about 6.5.

Envelope Bandwidth

The bandwidth of an SSB envelope is in general infinite. Consequently, the finite bandwidth of a practical class-S modulator distorts the envelope, which results in intermodulation distortion (IMD) in the output signal.

The IM products associated with the finite modulator bandwidth are determined in [5] through Fourier-series analysis and simulation. For a two tone signal, IM levels of –30 and –40 dBc require that differential delay not exceed $0.1/B_{RF}$ and $0.056/B_{RF}$, respectively.

Delay

Delay of the envelope with respect to the hard-limited carrier is another source of IMD. From [5], keeping IMDs below an amplitude c requires that

$$\tau^2 \leq \frac{\pi}{2}c \tag{9}$$

where τ is the differential delay normalized to a cycle of the modulation frequency. The time delay is then

$$\Delta t = \frac{\tau}{2\pi}\frac{1}{f_m} = \frac{\tau}{2\pi}\frac{2}{B_{RF}} = \frac{\tau}{\pi B_{RF}} \tag{10}$$

For two-tone SSB with a 3-kHz bandwidth, IMDs of –30 dBc require $\tau \leq 23.7$ μs, while IMDs of –40 dBc require $\tau \leq 13.3$ μs.

Unfortunately, delay is inherent in the low-pass output filter of the class-S modulator. The amount of delay depends upon the type of filter, the number of poles, and the bandwidth. The four-pole Butterworth filter used in the subsequently discussed circuit can be regarded as typical. It has a cut-off frequency of 57 kHz and a delay of about 9 μs.

Circuit Implementation

The PC-board implementation of the 100-W class-S modulator is shown in Figure 3.

The triangle-wave generator is an Exar XR-205 integrated circuit. PWM is generated by a MC3431 high-speed comparator. The input audio signal has a peak amplitude of 1.2 V. The DC offset is adjusted so that zero pulsewidth corresponds precisely to zero envelope.

The output circuit is shown in Figure 6. Voltage translation is accomplished by a Harris H11N1 opto-isolator. Com-

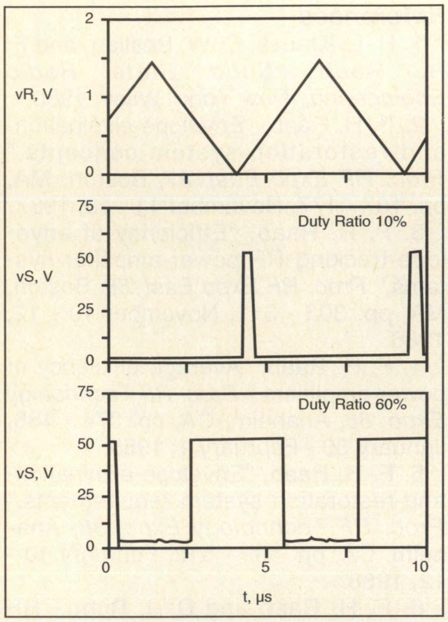

Figure 7. Observed waveforms.

plementary UHF BJTs (Motorola MRF542 and MRF543) ensure rapid switching of the IRF9522 p-channel FET (100 V, 5 A). The commutating diode is a Motorola MUR1610CT (100 V, 16 A).

A small auxiliary power supply (Burr-Brown PWR1726) produces V_{DD} + 7 V and V_{DD} −7 V (about 75 mA each) for gate drive. These voltages are just sufficient to turn the MOSFET on and off solidly and reliably, hence no more power than necessary is expended in charging the gate capacitance. The combination of these floating gate-drive voltages and the opto-isolator allow the supply voltage to vary from zero to the 100-V breakdown voltage of Q1 and D1.

The output filter is a four-pole Butterworth design based upon a 0 ohm input impedance and a 15 ohm load impedance. (The RF PA [6] is roughly equivalent to a 15 ohm load as seen by the modulator). This filter attenuates the switching frequency by about 70 dB and delays the output signal by about 9 µs.

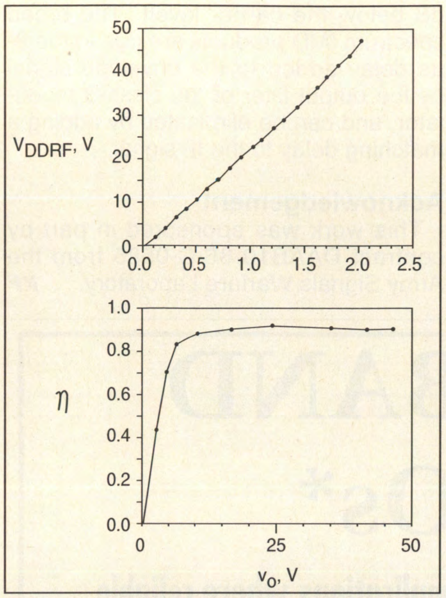

Figure 8. Linearity and efficiency.

Performance

The triangular reference wave and switching waveforms for both 10- and 50-percent duty ratios are shown in Figure 7. The waveforms show little ringing and there is little pulse deterioration at low and high duty ratios. Avoiding pulse deterioration near zero pulsewidth is especially important, as it directly affects the linearity for signals of small amplitudes.

The DC-linearity and efficiency characteristics are shown in Figure 8. The linearity is excellent (1.1 % rms deviation from a straight line). The efficiency remains at 90 percent or better for outputs above 20 percent of peak-output voltage (4 percent of peak-output power).

The harmonic spectrum of the modulator output is shown in Figure 9 for 100-percent modulation (full-carrier AM) by a

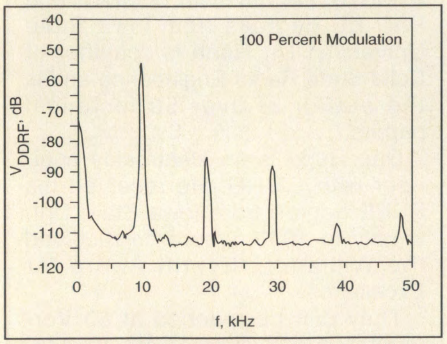

Figure 9. Spectrum of modulator output for single-tone AM.

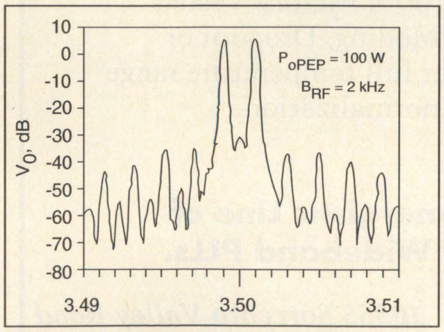

Figure 10. Spectrum of PA output for two-tone SSB.

single 10-kHz tone. The maximum harmonic is 40 dB below the signal for 50-percent modulation and 30 dB below the signal for 100-percent modulation.

Figure 10 shows the RF-output spectrum produced by an EER system[7] consisting of this class-S modulator and the class-D RF PA from [6]. The IMD produced by nonlinearities in the class-S modulator, class-D RF PA, SSB modulator, and EER envelope detector are 40 dB below the carrier level! The broad spectrum IMD products are due to the 9-µs delay added to the envelope signal by the output filter of the class-S modulator, and can be eliminated by adding a matching delay to the IF signal.

Acknowledgement

This work was sponsored in part by contract DAAB10-86-C-0585 from the Army Signals Warfare Laboratory. *RF*

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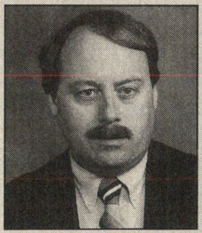
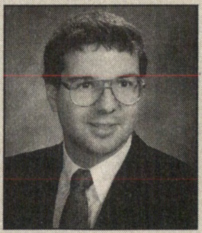
About the Authors





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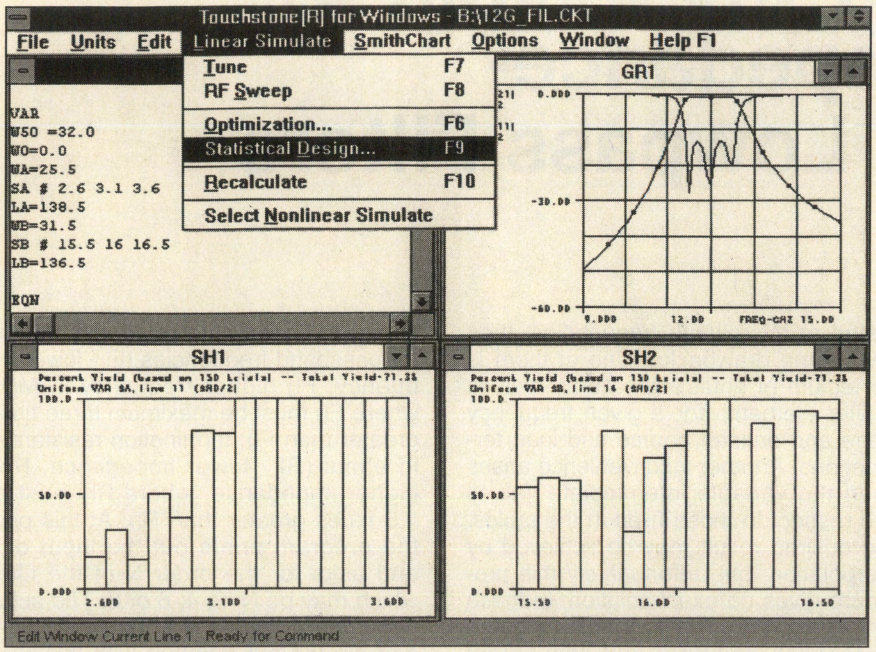
RF Design Awards Contest

GRAND PRIZE! — Touchstone/Libra for Windows



The 1994 Grand Prize provided by **Hewlett-Packard Company**, through their MDS group and their new EEsof operation.

Touchstone linear analysis and Libra non-linear analysis provide valuable tools for RF designers. Our winner will be able to spend even more time being creative, letting this software accurately simulate a circuit, taking away most of the trial-and-error associated with prototyping and design revision.

The personal computer version of Touchstone and Libra will be awarded — allowing the winner to use these powerful programs on a 386/486 PC platform, operating under the popular Microsoft Windows operating system.

Official Rules

1. Entries shall represent RF functions operating at frequencies from tens of kHz to 3 GHz.
2. Entries may be circuits, circuit design methods, test procedures, or design software programs.
3. If the entry is a circuit, it shall have a complexity equivalent to that of a circuit using 8-10 discrete active devices or 6-8 integrated circuits. The circuit may be a portion of a larger system.
4. If the entry is a design method, it must include an example of a circuit designed using the method described.
5. If the entry is a test method, it must include actual results of the measurement described.
6. If the entry is a computer program, it must operate on either an MS-DOS or Apple Macintosh system. It must be provided in a form that can be operated directly, without additional software (e.g., compiled). Programs must be submitted on disk, with supporting documentation provided in printed form.
7. Entries shall be the original work of the entrant, not previously published or publicly distributed. If developed as part of the entrant's employment, entries must have the approval of the entrant's employer.
8. Only one entry per person is permitted. An entry may have two or more co-authors.
9. Submission of an entry implies permission for publication in *RF Design*, and distribution of software submissions by the RF Design Software Service.
10. Winners are responsible for any taxes, duties, or other assessments which result from the receipt of their prizes.
11. Entries must be postmarked no later than July 29, 1994 and received no later than August 8, 1994.
12. All entries will remain confidential until publication of contest results in the November 1994 issue of *RF Design*.

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The single objective of the judging is do determine which entry makes the most significant contribution to the advancement of the art and science of RF engineering. Specific judging criteria include:

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Send entries to:

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A Program For the Design Of Chebyshev Impedance-Transforming Lowpass Filters

By Ljubomir Urshev
and Antoaneta Stoeva,
Elco Star Ltd.

This article describes a computer program which determines the values of the elements of Chebyshev low-pass matching filters. By using this program impedance transforming networks can be realized for n=2,4,6 and 8 reactive elements. A design example is provided together with the simulated response of the filter.

Low-pass matching filters are widely used in broadband transistor amplifiers and frequency multipliers. The procedure for designing such filters in most cases relies on tables given by Mathaei [1]. Although these tables of normalized element values for impedance transforming networks are useful for the design at microwave frequencies, they have some drawbacks. One of them is the tedium of obtaining accurate values of filter elements for a given frequency range and defined source and load terminations. Another inconvenience arises when the variable intermediate values with respect to those fixed in the tables. The desired result may be achieved by interpolation, but unfortunately this procedure gives no exact solution. In these cases the program described here offers a practical, accurate and convenient method for determining the circuit elements. The program is based on the results presented in [2] and calculates the elements of Chebyshev low-pass filters with an equal-ripple in the passband using special mapping functions.

```
Lower Frequency Fl in MHz
? 600
Higher frequency Fh in MHz must be greater
than 600 and less than 1800
? 1400
Lower impedance Rl in ohms
? 50
Higher impedance Rh in ohms must be greater
than 75
? 150

Lower frequency Fl = 600 MHz
Higher frequency Fh =  1400 MHz
Lower impedance Rl = 50 ohms
Higher impedance Rh = 150 ohms
Sure y/n? y

Network order (it may be N=2 or N=4 or N=6 or
N=8)
? 6

Network order N= 6
Fl          = 600 MHz
Fh          =  1400 MHz
Ripple Lar = 2.346127E-02 dB
Rs          = 150 ohms
Cp          = .9244648 pF
Ls          = 19.66351 nH
Cp          = 2.086634 pF
Ls          = 15.64968 nH
Cp          = 2.621813 pF
Ls          = 6.933454 nH
Rl          = 50 ohms
Another calculation with new N y/n? n
Ok
```

Table 1. Example run of filter design program.

Program Description

The program has been written in Microsoft Basic. When started it prompts for passband frequencies (Fl - lower frequency, Fh higher frequency in MHz, where Fh must be maximum three times greater than Fl), termination resistances in ohms (Rl - lower impedance, Rh - higher impedance, where Rh must be 1.5 times greater than Rl). At this point the program prints out the input data and asks for the order N of the filter, which may be N=2, 4, 6 or 8. The actual component values are then calculated and printed out together with the passband ripple Lar. Table 1 is an example of program operation for a 6th order filter with Rl=50 ohms and Rh=150 ohms, Fl=600 MHz and Fh=1400 MHz. For this example the calculated ripple is Lar=0.0234 dB. The computed component values are as follows: Cp=0.924 pF, Ls=19.6635 nH, Cp=2.086 pF, Ls=15.64968 nH, Cp=2.621 pF, Ls=6.93345 nH. After that, the computation may be repeated for a new value of N or terminated. In Figure 1 the filter equivalent circuit and the computed frequency response are given.

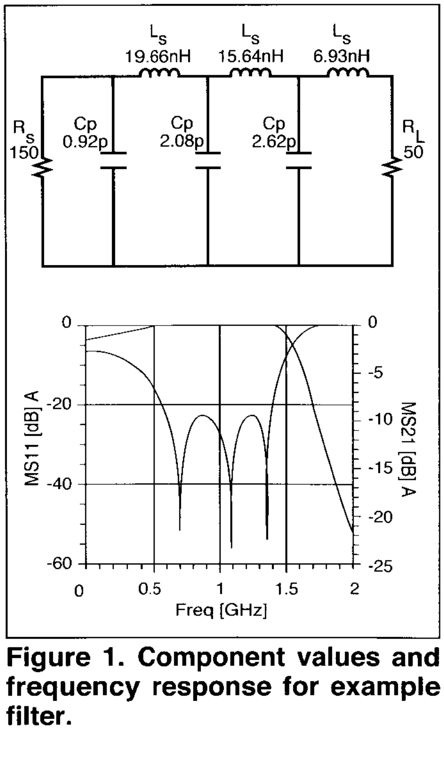

Figure 1. Component values and frequency response for example filter.

This program is available from the RF Design Software Service. For ordering information, please see page 89. *RF*

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RF Amplifier and Oscillator Design Using the UCFCAD Tools

Michael Rothery, Sam Richie, and Madjid Belkerdid
University of Central Florida

This article presents a comprehensive design for small signal RF amplifiers and oscillators in the VHF and UHF bands. The development of these systems is based on the small signal S parameters and noise parameters which can supply information on stability, gain, and noise. The negative resistance model is used for RF oscillator design. Colpitts and Pierce topologies are first reduced to the generic negative resistance model before the oscillation criterion is applied.

A PC based computer aided design tool (UCFCAD) has been developed to provide for a user friendly, interactive, fast, and inexpensive program to be used in the design, analysis, and synthesis of the above RF circuits. The input to the CAD system consists of the operating frequency, the S parameters, the noise figure parameters, and the system resistance, UCFCAD then plots the stability circles, constant gain circles, and constant noise figure circles on a graphical display of the Smith chart. The CAD tool also plots the frequency response, and generates the circuit diagrams for the wideband amplifier design. The program was written in C and runs on an MS-DOS PC.

The design of RF amplifiers and oscillators in the VHF and UHF band using lumped elements boils down to an impedance matching problem. Matching networks are normally done graphically, in the reflection coefficient domain, using a Smith chart. For a given reflection coefficient on the Smith chart, an intersection of circles technique is used to determine the amount of reactance or susceptance required from each component in the matching network. Then, given the frequency of operation and the system resistance, the value of the lumped inductor or capacitor are then be determined.

This article describes procedures used to design several types of linear RF amplifiers and oscillators. The steps for developing a set of matching networks to meet design specifications for a narrowband amplifier are first introduced, followed by an automated iterative design for single stage broadband amplifiers. A design procedure to develop oscillation conditions for a negative resistance oscillator are also presented. Colpitts and Pierce oscillator topologies, using their respective AC equivalent circuits, are used in the negative resistance model. Instead of matching the input and output ports of the active device, a resonant circuit is chosen for one port, while the other is matched under loaded conditions. These design techniques are based primarily upon the common emitter scattering (S) parameters of the biased transistor at the operating frequency.

Narrowband Amplifier Design

The design of a narrowband transistor amplifier is simply a matter of finding the proper matching networks for the input and output ports of a biased transistor circuit. Using the Smith chart, the stability, gain, and noise figure of the circuit are graphically analyzed with the information provided by the S parameters and noise parameters of the unmatched device. The proper source and load reflection coefficients may then be determined from the Smith chart. These reflection coefficients translate directly to the required matching networks which are easily obtained using the Smith chart. The matching networks determine the stability, gain, and noise figure of the amplifier. Therefore, the challenge is to find the source and load reflection coefficients on the Smith chart which will allow the designer to satisfy a given set of design specifications.

Narrowband design follows a flow

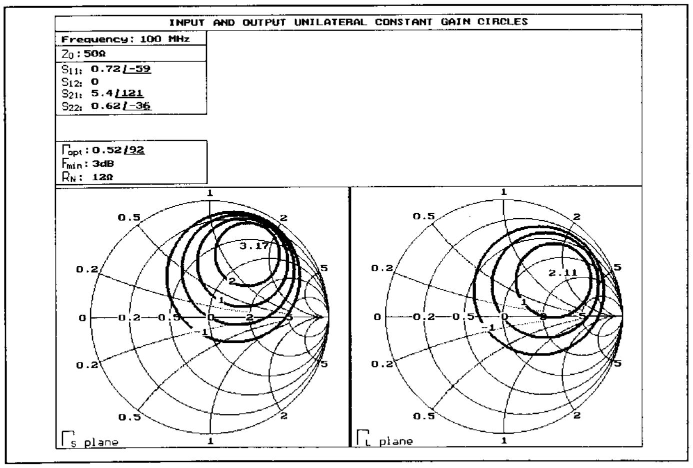


Figure 1(a). Input and output constant gain circles.

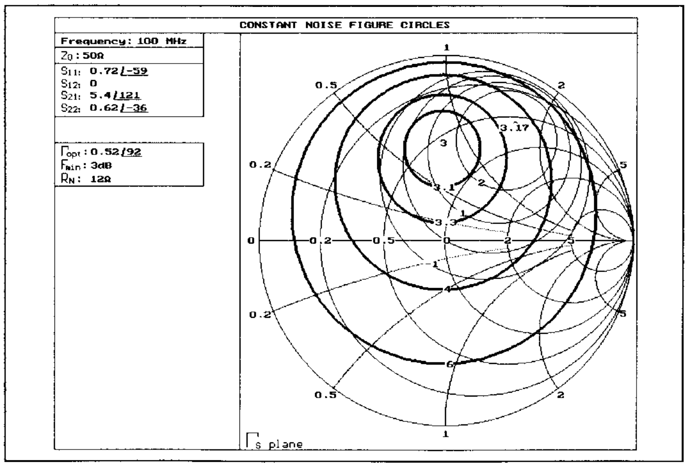


Figure 1(b). Constant noise figure circles.

chart type procedure that provides the steps necessary for the graphical development of stability circles, constant gain circles, and constant noise figure circles on the Smith chart for both unilateral and bilateral conditions. For a more thorough description of the analysis for narrowband amplifiers, references [1], [2], [3], and [4] provide in-depth information on this subject.

This comprehensive approach handles a very wide range of amplifier design specifications and requirements, and lends itself to computer automation. Unilateral and a bilateral narrowband amplifier design examples are described in detail to highlight the attributes of this flow chart design approach.

For a unilateral narrowband amplifier $S_{12} = 0$ is assumed and the design is simplified because there is no interdependence between the input and output matching networks. The following parameters of a transistor at 100 MHz are used as input:

$S_{11} = 0.72\angle{-59°}$	$F_{min} = 3\text{dB}$
$S_{12} = 0$	$\Gamma_{opt} = 0.52\angle 92°$
$S_{21} = 5.4\angle 121°$	$R_N = 12\ \Omega$
$S_{22} = 0.62\angle{-36°}$	

The maximum gain produced by the input and output matching networks is found to be $\Gamma_{S,max} = 3.17$ dB and $\Gamma_{L,max} = 2.11$ dB. A set of constant-gain circles is drawn on the Γ_S and Γ_L planes as shown in Figure 1a. Displaying the Γ_S-plane individually, a set of constant noise figure circles are added to the Smith chart (Figure 1b). If the design requires that the noise figure be no greater than 4 dB with the highest gain possible, for example, the source and load reflection coefficients are selected by moving a marker around the Γ_S and Γ_L-plane Smith charts. The fact that $\Gamma_S=S^*_{11}$ lies within the 4dB noise figure circle enables the selection of both source and load reflection coefficients at the points for maximum input and output gain, respectively. In other words $\Gamma_S= S^*_{11} = 0.72\angle{-59°}$ and $\Gamma_L= S^*_{22} = 0.62\angle 36°$. Figure 2c displays the possible input matching networks for Γ_S.

A bilateral 100 MHz narrowband amplifier design example, where unilateral conditions cannot be assumed, is also performed using a transistor with the S parameters given by:

$S_{11} = 0.7\angle{-68°}$
$S_{12} = 0.04\angle 60°$
$S_{21} = 8.5\angle 140°$
$S_{22} = 0.62\angle{-25°}$

The CAD system computes and outputs the stability factors K=0.462 and $|\Delta|=0.434$. Since K<1 the transistor is potentially unstable. The stability circles and a family of constant available power gain circles are computed and drawn as depicted in Figure 2a using the procedures outlined in the flow chart. When the source reflection coefficient is selected using an available power gain circle, it is assumed that the load reflection coefficient will be selected such that the output port is conjugately matched. The source reflection coefficient is selected by moving a marker around the Γ_S-plane Smith chart. The conjugate match for Γ_S is calculated and displayed in the Γ_L-plane. In addition, the possible matching networks for the locations of Γ_S and Γ_L are displayed.

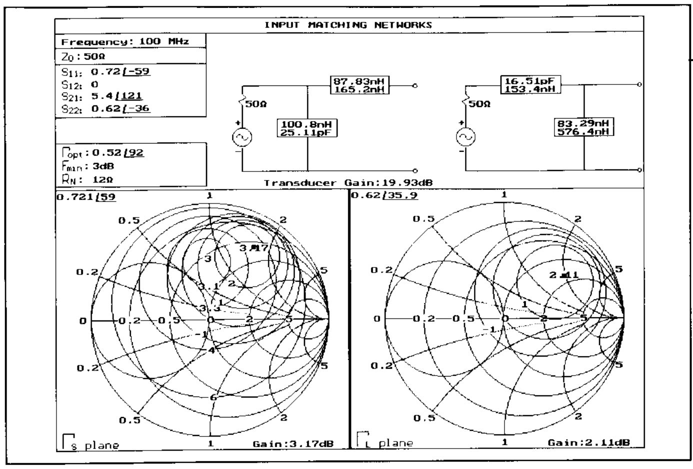


Figure 1(c). Input and output matching networks.

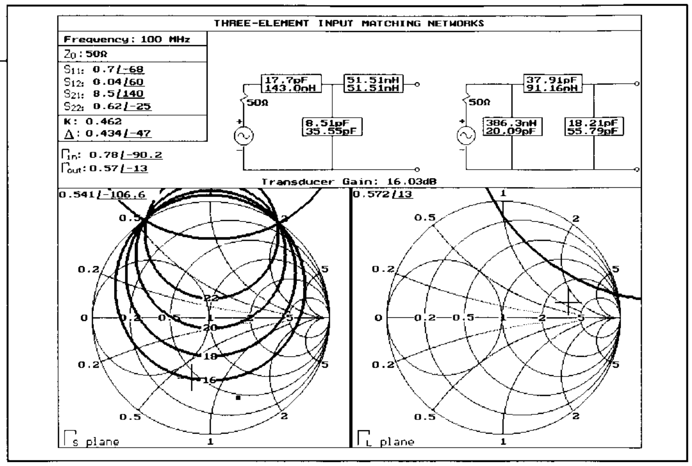


Figure 2(b). Three element input matching network.

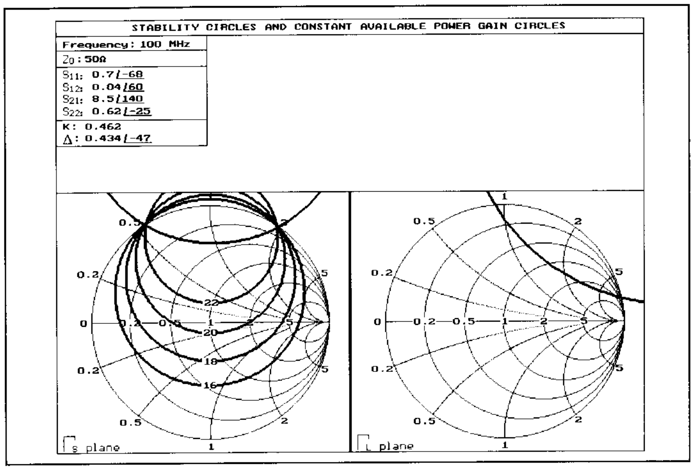


Figure 2(a) Stability and constant power gain circles.

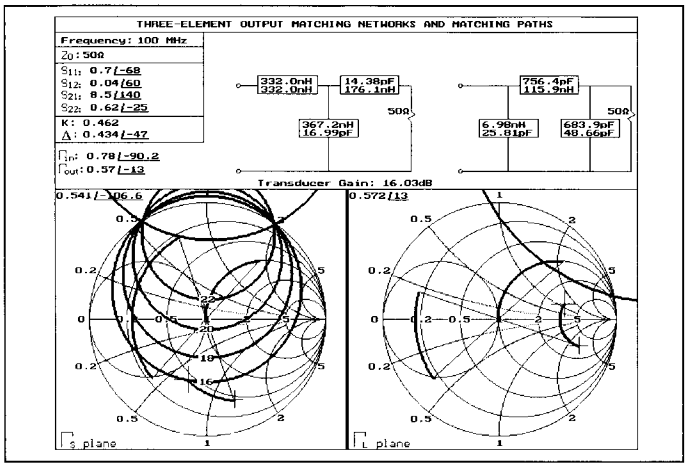


Figure 2(c). Three element output matching network and matching paths.

If, for example, a gain of 16 dB is desired from the amplifier, a source reflection coefficient on the 16 dB gain circle in the stable region of the Γ_S-plane is selected, in this case Γ_S = .0.541 ∠–106.6°. Figure 2b displays four possible three-element input matching networks for Γ_S. The conjugate match in the Γ_L-plane is shown to be Γ_L = .0.572 ∠13°. Figure 2c displays four possible output matching networks for Γ_L.

Wideband Amplifier Design

The technique used for the design of single stage RF amplifiers operating over a wide range of frequencies consists of three principal elements — a feedback network to flatten the gain response for constant gain specifications, a shunt resistor across the output port to stabilize the device, and matching networks to minimize the input and output VSWR [1]. A schematic of this RF broadband amplifier design circuit is shown in Figure 3.

One of the difficulties in the design of wideband amplifiers is caused from the variation of $|S_{21}|$ as a function of frequency. The resistor-inductor feedback network is used to compensate for these variations and assist in flattening the gain response, $|S_{21}|^2$. The CAD system operates by first tuning the active network of the device to provide the desired gain across the frequencies of interest by varying the values of R_f and L_f. The active network of the device includes the transistor, R_f , L_f , and R_S. For each change in R_f and L_f a new set of S parameters is calculated for the active network. This iterative optimization technique is the heart of this wideband amplifier design technique and is explained below.

To calculate this new set of S parameters, the common emitter S parameters provided by the user are first transformed to y parameters. The impedance of the feedback network is represented by $Z_f = Z_{fr} + jZ_{fx}$, where $Z_{fr} = R_f/Z_0$ and $Z_{fx} = 2\pi f L_f/Z_0$ are the resistance and reactance parts respectively, and f is the frequency for the set of S parameters under transformation. $Y_f = 1/Z_f$ is then added to y_{11} and y_{22}, and subtracted from y_{12} and y_{21}. If R_S is present in the network, this new set of y parameters is transformed to h parameters. To account for the shunt resistor, Z_0/R_S is added to the output admittance parameter, h_{22}. These h parameters are then transformed to the new set of S parameters for the active network. If R_S is not present in the circuit then the y parameters are transformed to S parameters. R_S is required for the circuit if the tuned active network is potentially unstable.

After tuning the active network, a check for potential instability (K<1 or |Δ|>1) is performed at each frequency for which a set of S parameters is given. K and Δ are stability factors and are defined as:

$$K = \frac{1 - |S_{11}|^2 - |S_{22}|^2 + |\Delta|^2}{2|S_{12}S_{21}|}$$

$$\Delta = S_{11}S_{22} - S_{12}S_{21}$$

If potential instability is found, a value of R_S is determined which will provide unconditional stability at each of these frequencies. This is done so that no passive matching network will yield an unstable device that could produce oscillations, and therefore, complete

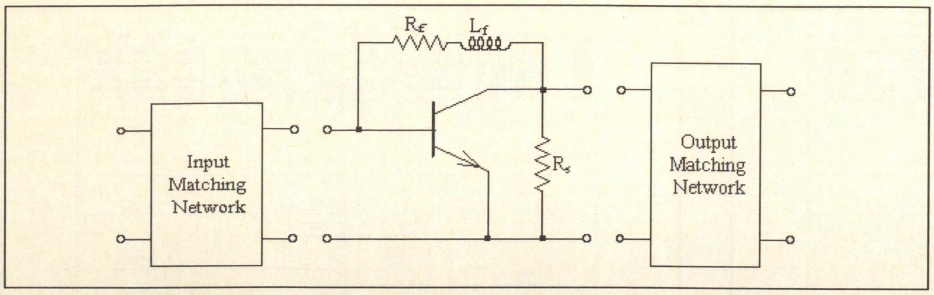


Figure 3. Wideband amplifier design configuration.

freedom in selecting the input and output matching networks is available. Of course, this stabilization is made at the expense of transducer power gain; and it should be noted that conditions do exist where this technique will not produce a stable network and other stabilization methods must be found.

The calculations for a value of R_S are performed in the CAD system by using the current values of R_f and L_f, and by starting with a small value of R_S. R_S is increased by 20 ohms until instability is detected, then decreased by 1 ohm until unconditional stability is achieved. For each change in the value of R_S, a new set of S parameters is calculated and the check for potential instability is made.

Further improvements are made in the wideband amplifier performance with the addition of input and output matching networks. The matching networks are designed to minimize the input and output VSWR while still maintaining a flat gain response over the desired frequencies. Using the CAD system, a set of matching networks is selected which serve as the initial values for automatically tuning the device.

Once the active network of the device is made unconditionally stable, the input and output reflection coefficients for a simultaneous conjugate match [1] are calculated at each frequency for which a set of S parameters is given. These input and output reflection coefficients are displayed on a set of Smith charts representing the Γ_S and Γ_L planes. As the user scans through each frequency point on the Smith charts, the possible lumped element matching networks for a simultaneous conjugate match at the highlighted frequency are displayed. The user then selects the input and output matching networks. Note that not all matching network configurations for a simultaneous conjugate match at a single frequency are acceptable across the band. For instance, suppose a an amplifier with constant gain specifications in the frequency band ranging from 10 MHz to 900 MHz is desired and a simultaneous conjugate match is selected at 10 MHz which includes a shunt capacitor. At the higher frequencies this shunt capacitor begins to appear more and more as a short and no power will be transmitted across the device. The CAD system used here warns the user of these unacceptable matching networks.

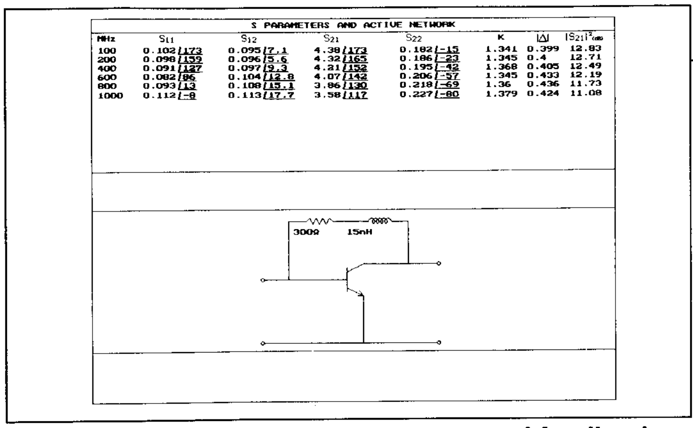
After a set of matching networks has been selected, the device is ready to be tuned. As each component in the device is varied, a set of S parameters is calculated to include the active network (transistor, R_f, L_f, and R_S) and the matching networks. The S parameters for the active network are first calculated as described previously. These S parameters are transformed to chain (ABCD) parameters and assigned to a matrix $[M_A]$. Next, the ABCD parameters are found for the input and output matching networks and assigned to matrices $[M_{in}]$ and $[M_{out}]$ respectively. The matrix for the ABCD parameters of the entire circuit, $[M_{tot}]$, are found by multiplying the previously defined matrices as follows, $[M_{tot}]=[M_{in}][M_A][M_{out}]$. The ABCD parameters of $[M_{tot}]$ are then transformed to the S parameters which now describe the entire wideband amplifier circuit.

The CAD system's optimization routine attempts to tune the device for a flat frequency response at the desired gain with a minimum input and output VSWR by varying R_f, L_f, and the matching network elements. The optimization process tries to minimize an error function defined as:

$$EF = \frac{1}{N}\sum_{F_i}^{F_f} |S_{11}|^2 + |S_{22}|^2 + \left(|S_{21}|^2_{(db)} - G_{(db)}\right)^2$$

where N is the number of S parameter sets given, F_i and F_f are the initial and final frequencies in the band, and G is the desired gain in decibels.

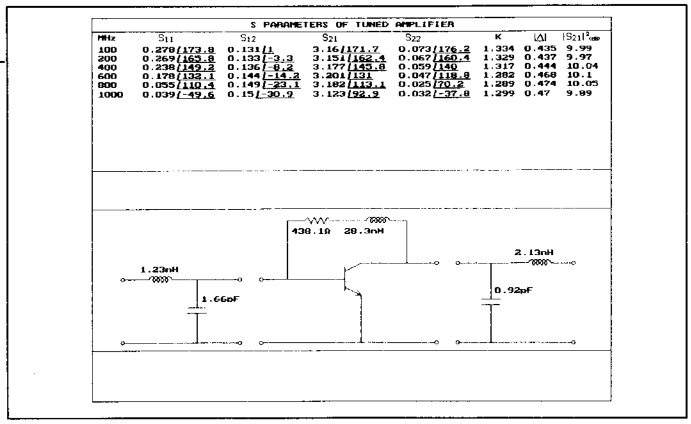
For example, suppose a design calls for a constant gain of 10 dB across a band from 100 MHz to 1000 MHz. The transistor common emitter S parameters and the initial default values for the feedback elements are shown in Figure 4a. The values of R_f and L_f and the resulting S parameters obtained after tuning the feedback network for the desired gain are shown in Figure 4b.



S PARAMETERS AND ACTIVE NETWORK

MHz	S_{11}	S_{12}	S_{21}	S_{22}	K	$\lvert\Delta\rvert$	$\lvert S_{21}\rvert^2$(dB)
100	$0.102\angle 173$	$0.095\angle 7.1$	$4.38\angle 173$	$0.182\angle -15$	1.341	0.399	12.83
200	$0.098\angle 159$	$0.096\angle 5.6$	$4.32\angle 165$	$0.186\angle -23$	1.345	0.4	12.71
400	$0.091\angle 127$	$0.097\angle 9.3$	$4.21\angle 152$	$0.195\angle -42$	1.368	0.405	12.49
600	$0.082\angle 86$	$0.104\angle 12.8$	$4.07\angle 142$	$0.206\angle -57$	1.345	0.433	12.19
800	$0.093\angle 13$	$0.108\angle 15.1$	$3.86\angle 130$	$0.218\angle -69$	1.36	0.436	11.73
1000	$0.112\angle -8$	$0.113\angle 17.7$	$3.58\angle 117$	$0.227\angle -80$	1.379	0.424	11.08

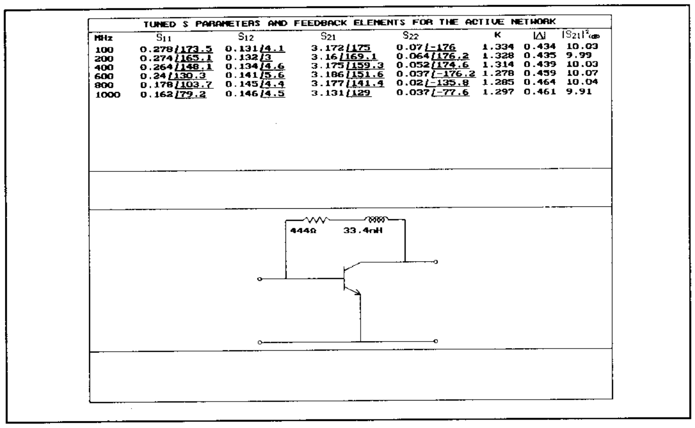
Figure 4a. Wideband amplifier setup and feedback.



S PARAMETERS OF TUNED AMPLIFIER

MHz	S_{11}	S_{12}	S_{21}	S_{22}	K	$\lvert\Delta\rvert$	$\lvert S_{21}\rvert^2$(dB)
100	$0.278\angle 173.8$	$0.131\angle 1$	$3.16\angle 171.7$	$0.073\angle 176.2$	1.334	0.435	9.99
200	$0.269\angle 165.8$	$0.133\angle -3.3$	$3.151\angle 162.4$	$0.067\angle 160.4$	1.329	0.437	9.97
400	$0.238\angle 149.2$	$0.136\angle -8.2$	$3.177\angle 145.8$	$0.059\angle 140$	1.317	0.444	10.04
600	$0.178\angle 132.1$	$0.144\angle -14.2$	$3.201\angle 131$	$0.047\angle 118.8$	1.282	0.468	10.1
800	$0.055\angle 110.4$	$0.149\angle -23.1$	$3.182\angle 113.1$	$0.025\angle 70.2$	1.289	0.474	10.05
1000	$0.039\angle -49.6$	$0.15\angle -30.9$	$3.123\angle 92.9$	$0.032\angle -37.8$	1.299	0.47	9.89

Figure 4c. Final optimized wideband design.



TUNED S PARAMETERS AND FEEDBACK ELEMENTS FOR THE ACTIVE NETWORK

MHz	S_{11}	S_{12}	S_{21}	S_{22}	K	$\lvert\Delta\rvert$	$\lvert S_{21}\rvert^2$(dB)
100	$0.278\angle 173.5$	$0.131\angle 4.1$	$3.172\angle 175$	$0.07\angle -176$	1.334	0.434	10.03
200	$0.274\angle 165.1$	$0.132\angle 3$	$3.16\angle 169.1$	$0.064\angle 176.2$	1.328	0.435	9.99
400	$0.264\angle 148.1$	$0.134\angle 4.6$	$3.175\angle 159.3$	$0.052\angle 174.6$	1.314	0.439	10.03
600	$0.24\angle 130.3$	$0.141\angle 5.6$	$3.186\angle 151.6$	$0.037\angle -176.2$	1.278	0.459	10.07
800	$0.178\angle 103.7$	$0.145\angle 4.4$	$3.177\angle 141.4$	$0.02\angle -135.8$	1.285	0.464	10.04
1000	$0.162\angle 79.2$	$0.146\angle 4.5$	$3.131\angle 129$	$0.037\angle -77.6$	1.297	0.461	9.91

Figure 4b. Tuned feedback elements.

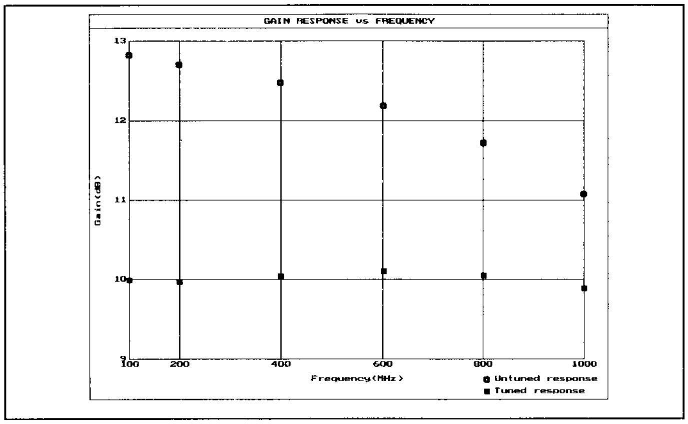


Figure 5. Tuned and untuned wideband responses.

Since this network is unconditionally stable over the band, R_S will not be required for this design.

To improve the input and output VSWR (minimize $|S_{11}|$ and $|S_{22}|$) a set of matching networks is added to the device and the amplifier circuit can then be tuned. The optimized amplifier design and resulting S parameters are shown in Figure 4c, while the untuned and optimized flat gain curves as a function of frequency are shown in Figure 5.

Oscillator Design

The same S parameter design techniques used in the design of narrowband amplifiers are applied directly to the design of the negative resistance oscillator model. The principal difference, however, is that the oscillator ports are passively terminated such that the transistor circuit will operate under unstable conditions. The Colpitts and Pierce oscillators are reconfigured to the negative resistance oscillator model using their respective AC equivalent circuit topologies.

The conditions of oscillation require an unstable active device coupled with refection coefficient conditions given by [4]:

$$\Gamma_{in}\Gamma_S = 1 \tag{1}$$

$$\Gamma_{out}\Gamma_L = 1 \tag{2}$$

where,

$$\Gamma_{in} = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L} \tag{3}$$

and,

$$\Gamma_{out} = S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1 - S_{11}\Gamma_S} \tag{4}$$

It can be shown [1] that if (1) is satisfied then (2) is also satisfied, and vice versa.

The negative resistance oscillator model — Given a transistor that is potentially unstable (K<1 or $|\Delta|$>1) at the desired frequency of oscillation, passive terminations can be found which satisfy the oscillation criteria given in (1) and (2). First, using the computerized Smith chart of the CAD tools, the input and output stability circles are calculated and drawn. The source and load reflection coefficients, Γ_S and Γ_L, which satisfy the oscillation criteria are then determined.

Γ_S and Γ_L are found by first selecting Γ_S from the unstable region of the Γ_S-plane, calculating Γ_{out} from (4), and finally computing $\Gamma_L = 1/\Gamma_{out}$ from (2); or select Γ_L from the unstable region of the Γ_L-plane, calculate Γ_{in} from (3), and find $\Gamma_S = 1/\Gamma_{in}$ from (1). Either method yields the same result [5]. The input and output matching networks corresponding to Γ_S and Γ_L are then determined.

An example of a 100 MHz oscillator design is discussed, the transistor has S parameters at 100 MHz given by:

$S_{11} = 0.85\angle{-50°}$
$S_{12} = 0.06\angle 54°$
$S_{21} = 9.7\angle 148°$
$S_{22} = 0.79\angle{-18°}$

The stability factors are found to be K= 0.381 and $|\Delta|$= 0.889, and since K<1 the transistor is potentially unstable. The source reflection coefficient is selected by moving a marker to the desired location on the Γ_S-plane Smith chart. As each value of Γ_S is highlighted, a corresponding value of Γ_L is calculated to satisfy the oscillation criteria. These values of Γ_L are graphically displayed on the Γ_L-plane Smith chart along with the possible matching networks for the current values of Γ_S or Γ_L.

Colpitts Oscillator Design

The standard topology for the Colpitts oscillator is shown in Figure 6a. In order to use the criteria for oscillation as defined in (1) and (2), this topology must

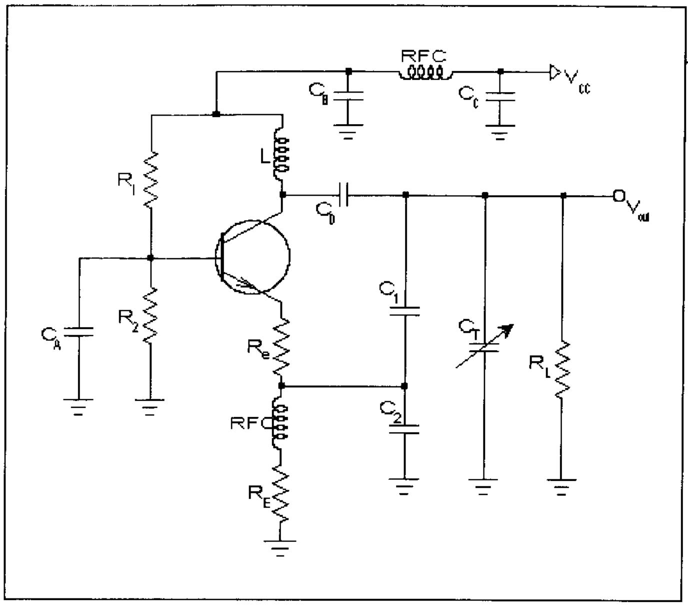


Figure 6a. Colpitts oscillator topology.

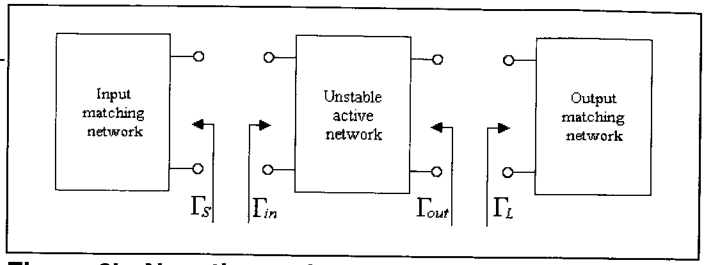


Figure 6b. Negative resistance oscillator model.

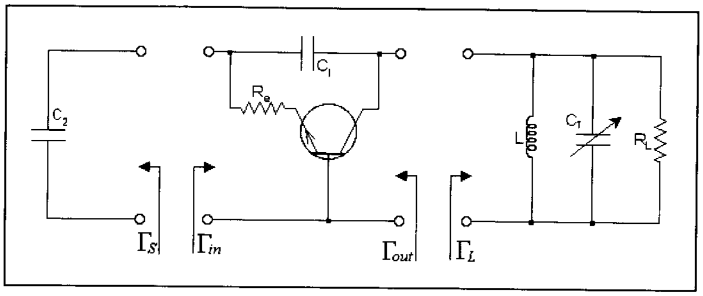


Figure 6c. Colpitts negative resistance model.

be made to conform to the two-port configuration as seen in Figure 6b [6], and the resulting AC equivalent circuit shown in Figure 6c satisfies this requirement. Here, the potentially unstable active network consists of the transistor, feedback capacitor (C_1) and resistor (R_e). The output matching tank network is made up of three parallel elements — an inductor (L), a tuning capacitor (C_t), and the load resistor (R_L). A single capacitor (C_2) makes up the input matching tank network.

The S parameters of the potentially unstable active network in this new configuration is determined from the common-emitter S parameters of the transistor and from the values of C_1 and R_e. The common emitter S parameters must first be transformed to common emitter h parameters and then to common base h parameters. R_e/Z_0 is added to the real part of h_{11} and the resulting h parameters are transformed to y parameters. $2\pi fC_1Z_0$ is then added to the susceptance portions of y_{11} and y_{22}, and subtracted from the susceptance portions of y_{12} and y_{21}. These y parameters may now be transformed to the S parameters of the active network and it can be confirmed that the device is potentially unstable (K<1 or $|\Delta|>1$).

Next, given the load resistance and a value for C_t, an incremental search is made for an acceptable inductor value that satisfies the condition $|\Gamma_{in}| = 1$. This condition, necessary so that the input match can be accomplished with a single capacitor to satisfy the Colpitts topology, is performed by the CAD tools. The iterative calculations begin with an inductor value of 1 nH. Γ_{in} is calculated using equation (3), where $\Gamma_L = (Z_L-1)/(Z_L+1)$, and $Z_L = (L||C_t||R_L)/Z_0$. The value of L is increased by 1 nH until a value is found which yields $|\Gamma_{in}| = 1$. When certain combinations of R_e, C_1, C_t, and R_L are used with a particular transistor, it is possible that no reasonable inductor value exists that will result in $|\Gamma_{in}| = 1$. Therefore, a maximum value of L should be supplied to prevent unnecessary iterations. There is an additional requirement to ensure that the input match can be made with a capacitor (rather than an inductor). If $-180° < \Gamma_S < 0°$, then using the oscillation criteria given in (1) results in $0° < \Gamma_S < 180°$ and the input cannot resonate with a single capacitor, and the Colpitts topology cannot be satisfied using the particular transistor at the desired frequency of oscillation which is, of course, inferred by the given S parameters.

If an acceptable inductor value is found and assuming that $0° < \Gamma_{in} < 180°$, a value will be determined for C_2 which satisfies the oscillation criteria. Using the relationship $\Gamma_S = 1/\Gamma_{in}$ from (1), the source impedance defined by $Z_S = (1+\Gamma_S)/(1-\Gamma_S)$ is found. C_2 is then given by $C_2 = -1/2\pi fZ_SZ_0$ where Z_S is always negative.

Using the same set of S parameters from the previous example, the following element values were supplied to the CAD system:

$R_e = 90\ \Omega$
$C_1 = 20$ pF
$C_t = 15$ pF
$R_L = 1000\ \Omega$

The active device S parameters were calculated for these element values and found from Figure 7 and are given by:

$S_{11} = 0.270\angle{-39.9°}$
$S_{12} = 0.632\angle{48.70°}$
$S_{21} = 0.794\angle{8.40°}$
$S_{22} = 0.774\angle{-43.6°}$

Figure 7 also indicates that an inductor value of 76.3 nH was found to satisfy the condition $|\Gamma_{in}| = 1$ and a capacitor value of 126 pF was found to satisfy the oscillation criteria of (1) to complete the design. Also displayed are a set of Smith charts which contain the input and output stability circles and locations of Γ_S and Γ_L. The element values can be changed directly on the circuit with an on screen editor. With each element value change a new set of S parameters is calculated, new values for L and C_2 determined, stability circles redrawn, and reflection coefficients relocated on the Smith charts.

Pierce Oscillator Design

The Pierce oscillator design is very similar to that for the Colpitts. The Pierce oscillator circuit is depicted in Figure 8 must be made to conform to the same two-port configuration shown in Figure 6b for the Colpitts. The AC equivalent circuit of the form shown at the top of Figure 9 satisfies this topology for the Pierce oscillator. The potentially unstable active network includes the transistor and feedback inductor (L). The output matching network is a parallel network made up of a capacitor (C_1) and load resistor (R_L), and the input matching network is once again a single capacitor (C_2).

The S parameters of the potentially unstable active network will be determined from the common emitter S parameters of the transistor and the feed-

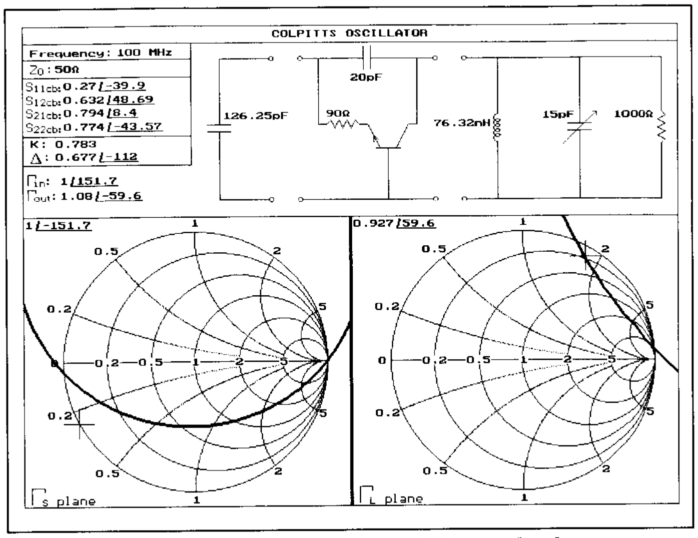


Figure 7. Optimized Colpitts oscillator design.

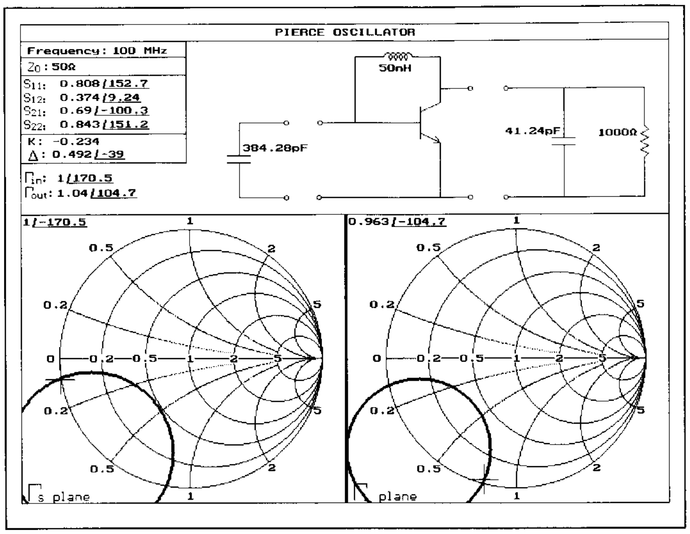


Figure 9. Optimized Pierce oscillator design.

back inductor, L. The common emitter S parameters given by the user must first be transformed to y parameters. $Z_0/2\pi fL$ is then subtracted from the susceptance portions of y_{11} and y_{22}, and added to the susceptance portions of y_{12} and y_{21}. These y parameters are then transformed to the S parameters of the active network, and it can be confirmed that the device is potentially unstable.

Given the load resistance, an incremental search is made for a value of C_1 that satisfies the condition $|\Gamma_{in}| = 1$ so that the input match can be accomplished with a single capacitor. If the search is successful and assuming that $0° < \Gamma_{in} < 180°$, C_2 will be calculated to satisfy the oscillation criteria given in (1).

Using the following S parameters and element values for a transistor at 100 MHz:

$S_{11} = 0.3\angle{-39°}$
$S_{12} = 0.02\angle 32°$
$S_{21} = 3.4\angle 121°$

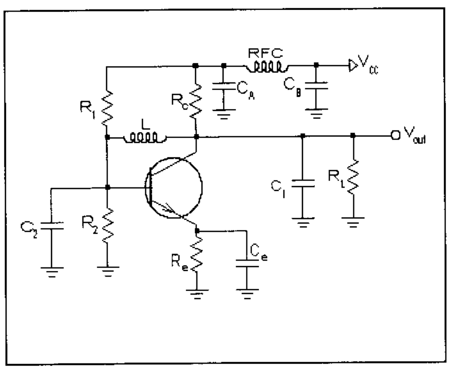


Figure 8. Pierce oscillator circuit topology.

$S_{22} = 0.43\angle{-24°}$
L = 50 nH
$R_L = 1000\ \Omega$

The active network S parameters were calculated and found from Figure 9, and are given by:

$S_{11} = 0.808\angle 153°$
$S_{12} = 0.374\angle 9.2°$
$S_{21} = 0.69\angle{-100°}$
$S_{22} = 0.843\angle 151°$

A value of C_1 = 41.2 pF was found to satisfy the condition and a single capacitor value of C_2 = 384.3 pF was found for the input matching network to satisfy the oscillation criteria. As with the Colpitts design, the input and output stability circles and locations of Γ_S and Γ_L are displayed.

Summary

This paper has presented CAD tools for RF amplifier and oscillator design developed at the University of Central Florida. The UCFCAD tools are based on a flow chart design approach to narrow band, and a custom optimizer for wide band amplifier design, as well as an iterative design approach to the negative resistance oscillator model. The UCFCAD tools, written in C for the IBM PC or compatible, are pull-down menu driven and can be used for design and analysis. These tools include custom graphical displays such as a computerized Smith Chart and interactive design capability, and also offer some basic schematic capture capability.

The UCFCAD program will be available in June, 1994, and will be distributed by the Argus Inc. Direct Marketing Department, 6151 Powers Ferry Road, Atlanta, GA 30339; (404) 618-0219. See the June issue of RF Design *or call for pricing and availability.* ***RF***

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Communications Test Equipment Copes with Myriad of Standards

By Ann Marie Trudeau
Assistant Editor

Communications test equipment is in the same flux as the communication industry itself. R&D needs flexible applications for designing. Manufacturers and service providers want fully automated equipment that can be used by an unskilled tester. But, the biggest concern is the standards race. With this turmoil in mind what follows is an overview of how some companies are interpreting the race and focusing their products.

"The tremendous growth and development in the wireless communications market is driving the needs of the test equipment market," says Carla Slater, Product Development Engineer at Giga-tronics Corp. They are targeting manufacturing requirements. Slater said that Giga-tronics is not pursuing any wireless standard, waiting to see what the market will settle on.

Sharing the same view, Jack Tiley, World Wide Sales Manager for Hewlett Packard's Spokane, Washington Division, said that they are involved in RF communications, service, and manufacturing. As a result they have responded to the many available standards, he said, by taking a wait-and-see mode and by building customizable general purpose platforms.

"It's a heck of a market to keep chasing," said Doug Mach, Test Engineering Manager of Motorola's Government and Systems Technology Group. "The biggest change I see is the constant evolution of communications." Their testing platforms for test radios, pagers and cellular sites have room for upgrades.

International Responses

Deputy Managing Director Sandy Warwick, of Marconi Instruments Inc. pointed out that the traditional main players are changing and looking toward other markets because of military cut backs. "In Europe the same thing is happening and everyone wants something from test manufacturers," Warwick said. As a result, Marconi also has decided to leave room in their radio test set for features to deal with different standards.

Ken Harrison, the National Sales Manager for Wayne Kerr, Inc., a British manufacturer, said that they are already into the Far East but are not well known in the U.S. The markets that they are into are wireless PCS and LAN. They don't provide any dedicated systems but modulation sources can be used.

Another company that doesn't provide standard dedicated systems but does combine the needs of R&D and communications is the Lake Stevens Instrument Division of Hewlett Packard. Rod Nemitz, Product Marketing Engineer, said that their R&D equipment has been driven by diverse and complex applications and tighter restrictions in a restricted spectrum. Nemitz thought that CDMA is making strong moves in communications networks and PCN or PCS will soon be in the same position that CDMA is in now.

With instruments for R&D, manufacturing and maintenance, Noise Com tries to bring the real world into the test lab with their multi-path fading emulator, Marketing Manager Bent Hessen-Schmidt said. Noise Com's focus continues in these three test equipment market areas.

Joining Forces

Rohde & Schwarz, Tektronix, and Advantest Corporation, a Japanese company, have taken the tactic of working together so that whatever way the standards race goes they are protected because one of the three companies will be able to jump into the market with the support of the others. Gerhard Sonnde, Director of Marketing and Sales for Rohde & Schwarz said they are the only Type Approval Test System for testing European GSM phones. If a vendor wants to make phones for the European GSM standards they have to be tested, by law, on a Rhode & Schwarz tester before they can be sold. They are targeting R&D, manufacturing and servicing, with the biggest focus on the manufacturing side. They don't expect analog to grow any more because everything is going digital, Sonnde said.

A Resourcful Solution

Dr. Herman C. Okean, Vice President of Government Systems Division for LNR Communications, Inc., said that their niche has been in satellite communications. But, because they had to develop flexible spread spectrum equipment for their own use, going into the spread spectrum test generator and receiver business was a natural extension for LNR

Views on Fallout Time

EIP Microwave, Inc., is still focusing on the shrinking military market as they expand into a communications niche serving sparsely settled, remote areas. They service dish antennas, cell sites, and microwave links. "I believe the consumer will get upset with the government and want a single standard," Ray Beers, Product Marketing Manager said. Also he said the standard fallout may be decided in one to two years.

Then there's the long term approach to the standards fallout. "There's talk in the industry that sometime down the road, it may be in as many as 15 years, we will have a world wide standard," Craig Hendricks, Product Marketing Engineer for Anritsu Wiltron Sales Company said. Hendricks thinks that the future standard will be in the spread spectrum area. Anritsu Wiltron focuses on a narrow niche of cellular and PCS design, manufacturing, including the components market, by offering a general piece of equipment that has dedicated standards accessible by push buttons.

Summary

One person jokingly pointed out that they were offering dedicated general equipment. That observation certainly exemplifies the dilemma facing the industry and how test equipment providers are trying to cope with the changes and still stay marketable. *RF*

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INFO/CARD 82

SPECTRUM PROBE®

107 255

EMC RFI

This low circuit loading ***scope probe*** produces a 60 dB **spectrum analyzer** display on **your** inexpensive scope

AC Power line - FCC limited	***computers***
Field Strength - E field	***medical***
Conducted Radiation - current	***avionic***
Radiation Sources - voltage	***automotive***
EMI Sources,and leak sniff	***ind.process***
Ultrasonics Infrared	***TV remotes***
Remote (line) Controllers	***household***

Ask for these application notes,specifications, sales & warranty information.

255 Spectrum Probe 30 KHz to 2.5 MHz $279
107 Probe 1 to 100 MHz $249 ($279 extra shielding)
Accessory Current Adapters: IA5 $19 IA7 $29

Technical tips show how scope amplitude allows estimating field strength 300 µV/M to100 V/M, to find and reduce RF radiation per FCC part 15,18;VDE. You can sniff and find that radiation!

Other tips show current adapters (40 µA to 100 mA) with typical calibration curves. Useful for RTCA/ DO-160, MIL-STD461 requirements. Good also for capacitor, board, and IC lead currents.Computer switching power supply servicing is speeded by current measurements.

An unusual tip shows how you can connect standard low cost components within an outlet strip. These outlets then allow operation of your equipment well isolated from power line RF noise.Included is the FCC line (LISN) network, allowing you to measure conducted EMI from your equipment per FCC part 15,18;VDE.
--- Fix conducted EMI !

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RF software

ASIC Design Kit on CD-ROM

Toshiba America has released their first CD-ROM software of Toshiba's application-specific-integrated-circuit (ASIC) design kits which will be updated quarterly. The design kits include gate array, embedded array, and standard cell libraries for the 0.5 micron (μm) TC180 Series, 0.8μm TC160 and TC163 Series, 1.0μm TC140G Series and others such as the 0.8μm TC26SC/TC25SC standard cell families and TC14L Series. Also included are necessary interface and application software for various third party EDA tools.
Toshiba America Electronic Components
INFO/CARD #210

Component Databases

Information Handling Services® offers two new databases, one for ICs and discrete semiconductors, the other for passive components. The IC/Discrete Parameter Database contains information on more than 1.5 million parts, which can be searched by part number, function, keyword, characteristic, generic number, etc. The ReCal/z™ has information on millions of passive components from over 360 manufacturers. Both databases also include manufacturer's information such as data-sheets and cross-referencing guides.
Information Handling Services
INFO/CARD #209

High-Speed Design Instruction

An interactive application disk, available from Tektronix teaches designers working with high-speed integrated circuits techniques to solve problems like metastability, ground bounce, crosstalk and more. The disk allows the users to set up measurements, change parameters, and view the results on simulated instruments screens.The disk is available free by writing on company letterhead to Tektronix, Test & Measurement Group, P.O. Box 1520, Pittsfield, MA 01202
Tektronix
INFO/CARD #208

ASIC Place and Route Tool

Tanner Research has released L-Edit/SPR™ v.4. The software provides placement and routing for ASIC standard cell design, automatic padframe generation and routing. The new version handles 100% more circuit elements than version 3.0 and includes user control of the maximum time to be used by the placement optimizer. Other features are a row evener and a wiring density maximizer. L-Edit/SPR 4.0 supports PC compatibles, Macintosh, SUN/SPARCstations and HP/Unix stations. Pricing for the UNIX systems begins at $2995.
Tanner Research, Inc.
INFO/CARD #207

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Data Acquisition Data Book

Data Acquisition 1994 is the 1,104-page data book from Harris Semiconductor. It provides long-form data sheets on hundreds of products for commercial and industrial applications and includes 39 new products. Separate sections covering A/D converters include: display, integrating, successive approximation, flash, and subranging. Other sections include D/A converters, switches, multiplexer, RS-232 communications interface ICs, display drivers, counters with display drivers/time base generators, and special-purpose ICs. Application note abstracts are included.

Harris Semiconductor
INFO/CARD #205

New Microwave Brochure

A four page brochure from K & L Microwave, Inc. uses plots and block diagrams to describe their manufacturing capabilities for switched filter banks, multi-function systems, PIN diode switches and a comb generator assembly.

K & L Microwave Incorporated
INFO/CARD # 204

Rohde & Schwarz Journal

The latest edition of Rhode & Schwarz's News from Rohde & Schwarz includes a 16 page brochure that reviews the company's history. The rest of the in-house technical journal includes application notes on: DECT RF measurements, type appropriate tests on GSM/PCN base stations, audio analyzer UPD, and calibration of test antennas to measure signal and interference field strengths.

Rohde & Schwarz
INFO/CARD # 203

System Applications Guide

The System Applications Guide for Analog Devices is a 19-chapter handbook of techniques, topologies, and methods covering all aspects of the analog-to digital signal path and environment. Topics covered include: sensors interfacing, multiplexing, amplifiers, high-accuracy performance, audio applications, high-speed systems, interfacing to A/D converters, verifying performance, D/A converters and direct digital synthesis.

Analog Devices
INFO/CARD # 202

Crystal Oscillator Catalog

Vectron's 1994 16-page catalog summarizes a complete line of crystal oscillators available in the 0.01 Hz to 2 GHz plus range and introduces new products. Areas of crystal oscillators covered are in moderate stability clocks, temperature compensated, subminiature oven controlled, and voltage controlled situations. Also listed are the specifications and features of each product line.

Vectron Laboratories, Inc.
INFO/CARD #201

Coaxial Connector Catalog

Delta Electronics Manufacturing Corporation offers an updated catalog which presents their line of N coaxial connectors in a logical format. Choices of plating, cable clamping types, receptacle configuration and MIL specifications, (including MIL-C-39012 and MIL-A-5539 QPL connectors), are presented for each connector type. Line drawings are proportionally accurate and can be easily scaled and reproduced for use on internal specification drawings.

Delta Electronics Manufacturing Corp.
INFO/CARD #200

1994 SME Catalog

The Society of Manufacturing Engineers (SME) is offering a free 100-page catalog that features more than 300 manufacturing technology publications and video tapes. It shows products and services that focus on both the traditional and changing technologies. The catalog also gives information on SME's certification program, INTIME (Information on Technology in Manufacturing Engineering) data bank services and upcoming educational events.

Society of Manufacturing Engineers
INFO/CARD #199

ADC Application Note

Datel, Inc. offers application note AN-6, a four-page note that describes how data converter specifications relate to an actual image and undesirable artifacts. The note explains the advantages of using an A/D converter designed for time-domain applications such as imaging. Special attention is paid to the effects of differential nonlinearity (DNL) on an image.

DATEL, Inc
INFO/CARD #198

EMI Filter Connector Catalog

Spectrum Control, Inc. has released a 44-page catalog that provides extensive information structured to aid in determining the most efficient method to filter a connector. Technical data covers connectors meeting MIL specifications 38999 for compact shell and extended shell classifications 3899, 83723, 26482, and 24308 (D-Subminiature).

Spectrum Control, Inc.
INFO/CARD #197

Test & Measurement Catalog

Analogic Corporation announces a new test and measurement catalog featuring high precision instrumentation products. Featured in the 72-page catalog are waveform generators, synthesizers and analyzers. General purpose instrumentation is also included. Ten waveform generators have been added and include standard and programmable generators, arbitrary function generators, high bandwidth arbitrary wave form generators, and a flexible pulse generator for comparator and device testing, telecommunications, and ATE.

ANALOGIC
INFO/CARD #196

SURFACE-MOUNT or PLUG-IN FROM $3.95 (10 qty)

Expose Mini-Circuits' TUF-mixers to 250°C for five minutes, or to the extreme shock and vibration stresses of MIL-M-28837, or to 200 cycles of thermal shock from -55° to +100°C...they'll survive without any change in specs. They are mighty tough mixers!

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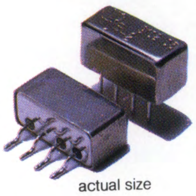

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SPECIFICATIONS

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

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

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