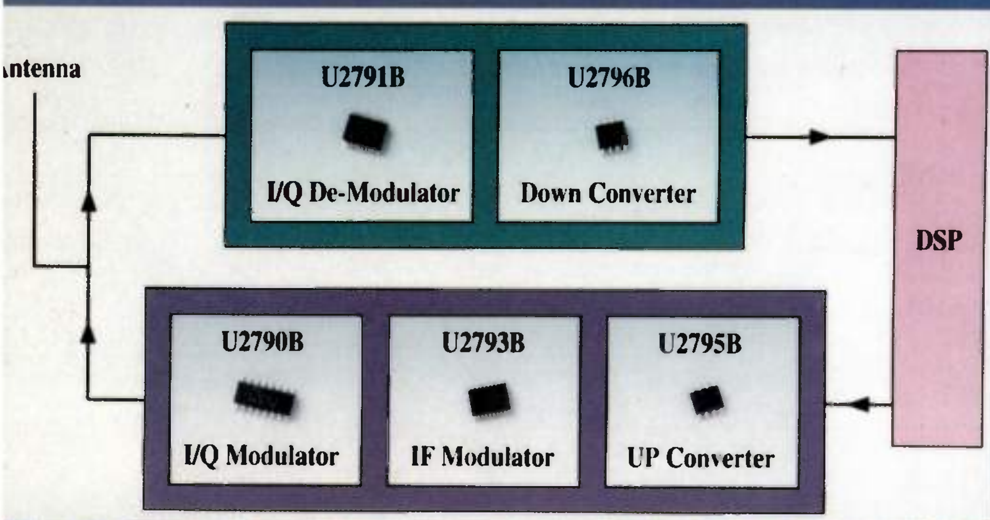
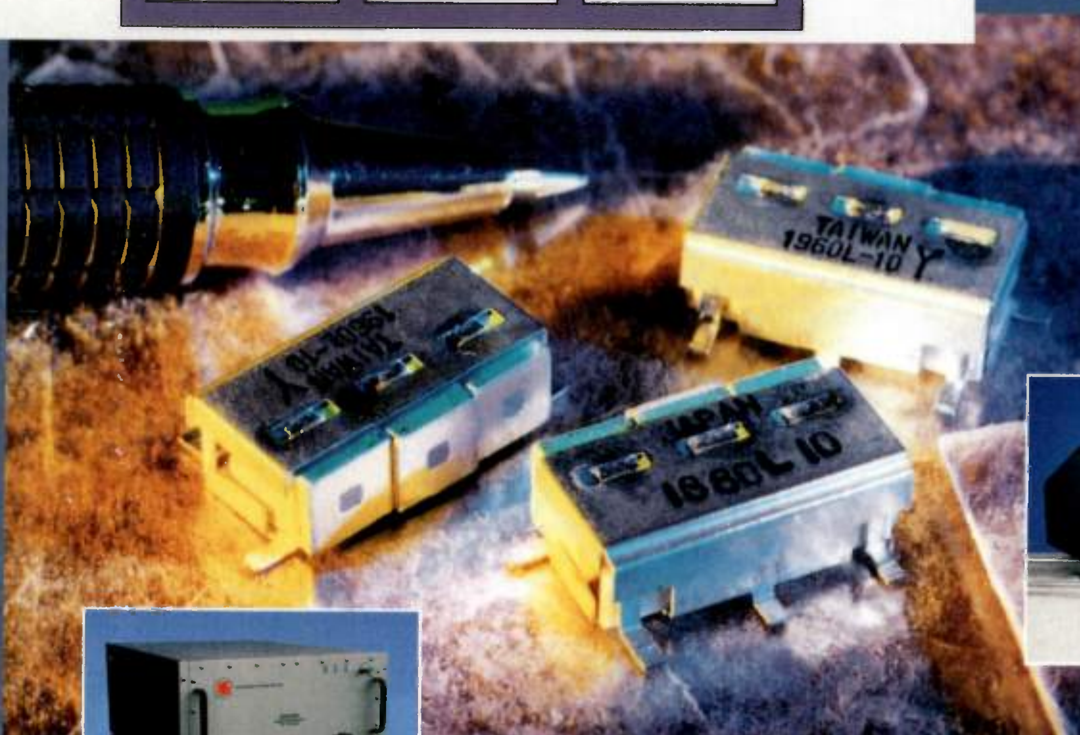
WIRELESS 16 YRS LEADERSHIP

RF design™

engineering principles and practices

June 1995





New RF Products Feature Wireless Applications

Featured Technology —
DSP Implementation of GFSK and GMSK

Wireless International Corp. solves tough test problems.

As test engineers and designers, you understand the importance of efficient and effective testing for development and production testing. Poor test methods and inaccurate measurements can bring the production of the best designed equipment to a crawl. Wireless International Corp. (WIC) has answered these problems with plug-in-and-run test systems which provide stimulus, control, and measurement capabilities at your fingertips. Here are just a few of the test solutions.

Circle #1 for information on WIC.

CDMA Multipath and AWGN test station.

For CDMA testing, WIC's MP-CDMA test system generates multipath and additive noise as specified by IS-98. This system is comprised of the multipath emulator, a precision carrier-to-noise generator, an integrated 486DX/33 computer, and control software with a graphics user interface. The precision carrier-to-noise generator has been designed specifically to operate with extreme accuracy even under the difficult conditions of slow fading for which CDMA systems must be tested. Available frequency ranges are between 800 and 2500 MHz.

Circle #78 for CDMA Test Equipment.

WRH

Multichannel Eb/No versus BER test station.

If you are performing BER versus Eb/No curves for multiple devices simultaneously, then here is your test equipment answer. WIC's Series 800 test stations can generate these curves for up to 8 channels simultaneously, all under the control of a single integrated computer or via the IEEE-488 interface. Each channel contains a precision carrier-to-noise generator and a programmable bit error rate tester. Bit errors and bit error rate data are stored to disk for analysis. Eb/No versus BER curves are generated in both tabular and graphical form for hard-copy output. Data rates up to 8 X 10 MBPS are programmable with error insertion rate, bit slip, and PN code length selections.

Circle #79 for Series 800 Test Station.

A look into the future.

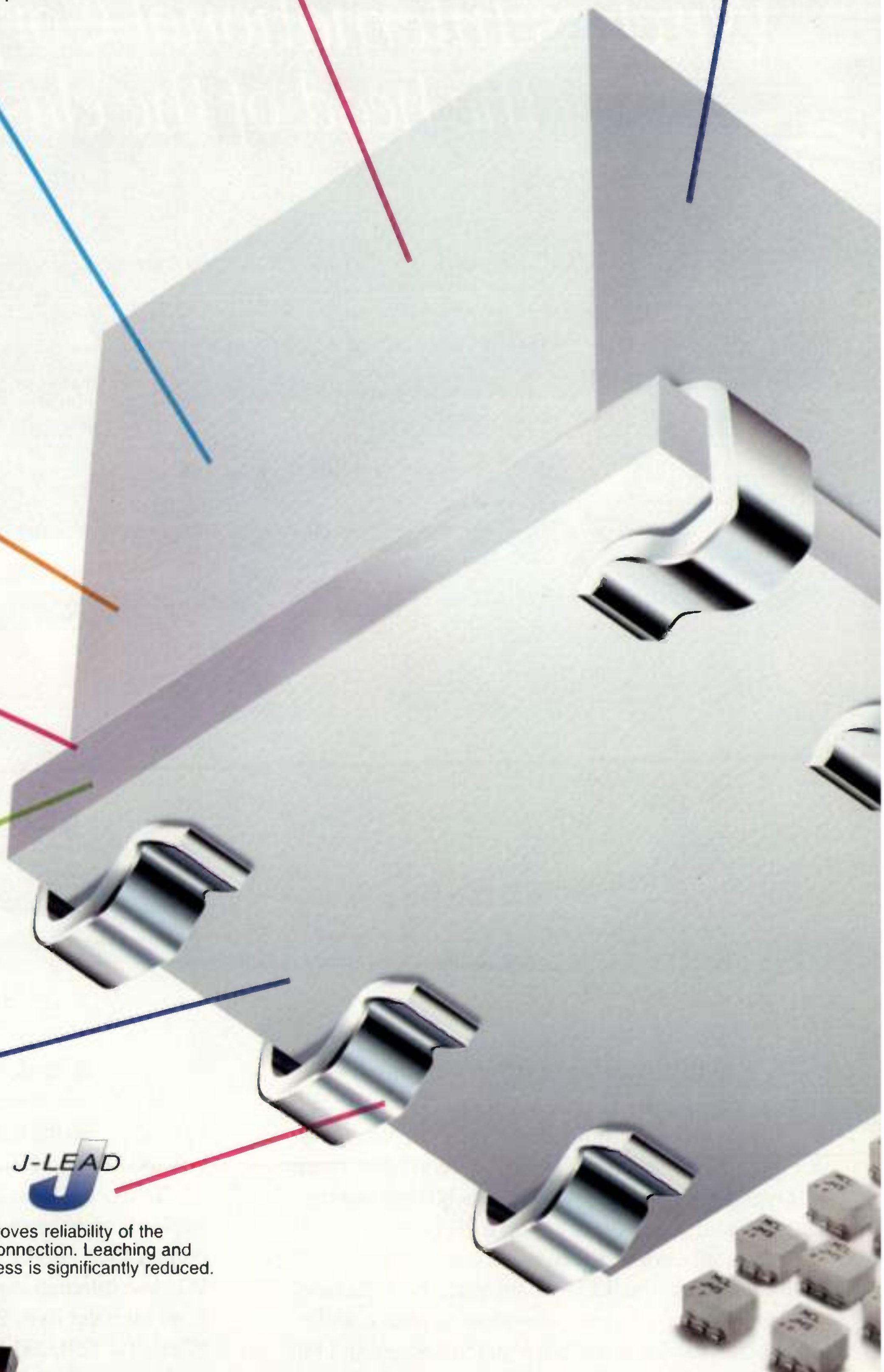
WIC's products are modular to provide easy upgrading. So the instrument that you buy today won't be obsolete tomorrow.

To receive a FREE copy of our application notes and catalog, contact Bent Hessen Schmidt at Wireless International Corp., E. 49 Midland Ave., Paramus, NJ 07652. Tel: (201) 261-8797. Or fax us at (201) 261-8339.

Wireless Telecom Group INC.

NOISE/COM WIC Wireless International Corp.

Circle #80 for General Information.



Mini-Circuits RF/IF Surface Mount Designer's Guide features 48 pages of the most up-to-date and complete product and specification information about Mini-Circuits surface mount components.
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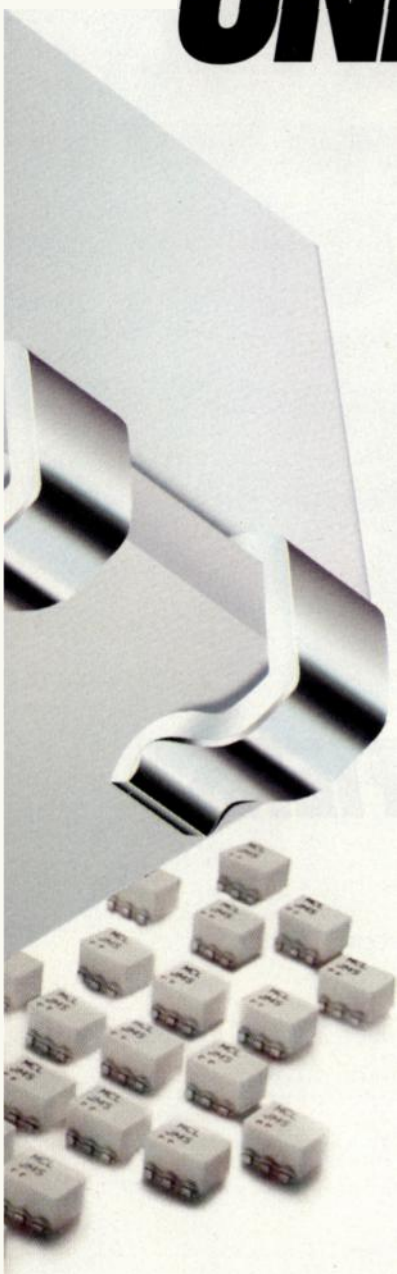
NEW!

J-LEAD J MIXERS UNPRECEDENTED IN VALUE.



All Ceramic 2 to 1900MHz from $4.25 (qty. 10-49)

Mini-Circuits JMS mixers are at the forefront of the industry for performance, reliability and value! With all-welded construction, these low cost ceramic mixers feature the functional J lead designed for strain relief...and set the pace for ruggedness required to go through surface mount reflow soldering and aqueous wash.

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SPECIFICATIONS

Model	LO (dBm)	Freq. (MHz) LO/RF	Freq. (MHz) IF	Midband (dB, Typ.) Conv. Loss	Isol. L-R	Isol. L-I	$ea. (qty. 1-9)
JMS-1	+7	2-500	DC-500	5.75	45	45	4.95
JMS-1LH	+10	2-500	DC-500	5.75	55	45	8.45
JMS-1MH	+13	2-500	DC-500	5.75	60	45	9.45
JMS-1H	+17	2-500	DC-500	5.90	50	50	11.45
JMS-2L	+3	800-1000	DC-200	7.0	24	20	7.45
JMS-2	+7	20-1000	DC-1000	7.0	50	47	7.45
JMS-2LH	+10	20-1000	DC-1000	6.5	48	35	9.45
JMS-2MH	+13	20-1000	DC-1000	7.0	50	47	10.45
JMS-2H	+17	20-1000	DC-1000	7.0	50	47	12.45
JMS-2W	+7	5-1200	DC-500	6.8	60	48	7.95
JMS-11X	+7	5-1900	5-1000	6.7	35	37	4.25*

Note: *10-49 qty.

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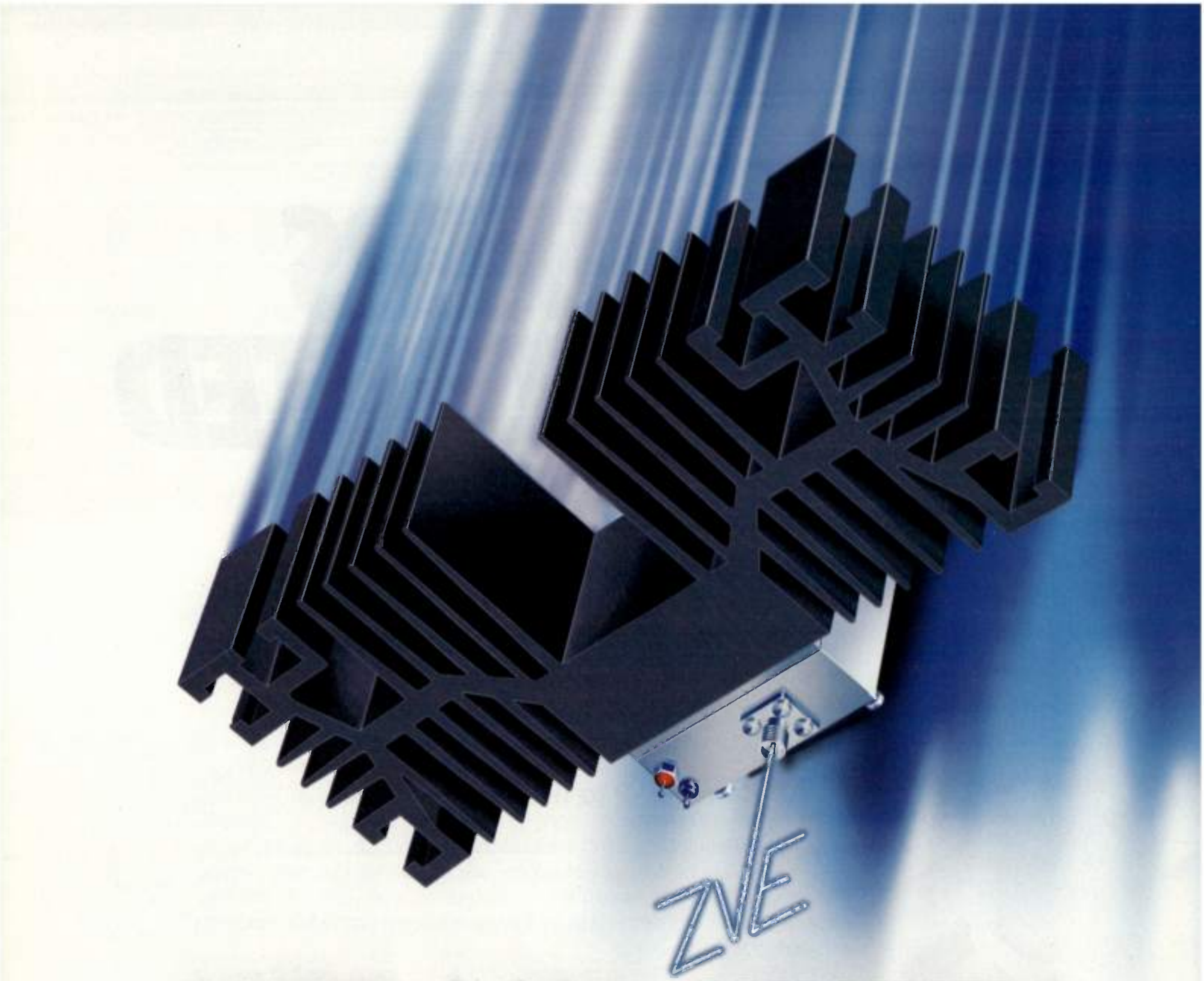
F 181 Rev A

INFO/CARD 2



1 WATT 2-8GHz AMPLIFIERS

34dB Gain, ±1.0dB Flat only $1095 (Qty. 1-9)

Here is the big price/performance breakthrough in Space Age communications...Mini-Circuits ZVE-8G high power microwave amplifiers...high-end performance without the high-end price! Housed in a rugged, hermetically sealed package, this unconditionally stable unit is tough enough to withstand abuse by shorting/unshorting 12V terminal to case with full RF power applied. And while you'll find the high performance of the ZVE-8G amplifier unmatched at the amazing price of just $1095 (qty. 1-9), you'll also find the VSWR input-output extremely *well* matched at typically better than 1.5! The incredibly versatile range of wireless applications include line-of-site transmit and receive, satellite up and down links, S- and C-Band radars and spread-spectrum communications. Additionally, this unit is extremely useful as an optical transmitter driver amplifier.To order from stock with a guarantee to ship within one week, call Mini-Circuits today.

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For system configurations, units may be used with heat sink removed for mounting to chassis and with connectors removed for solder connection to a PC board land.

ZVE-8G SPECIFICATIONS

Freq.Range (GHz)	2-8
Gain (dB)	
Min.	30
Flatness Max.	±2.0
Power (dBm)	
Min. Output (at 1dB Comp.)	+30*
Max. Input (No Damage)	+20**
Dynamic Range	
NF (dB) Typ.	4
Intercept Point (dBm) 3rd Order Typ.	40
VSWR In/Out (Max.)	2.0:1
DC Power	
Voltage	+12
Current A Max.	2.0

* At 25°C; +29dBm typ. at +90°C case.
** With no load, reduce input by 10dB.

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INFO/CARD 3

RF design

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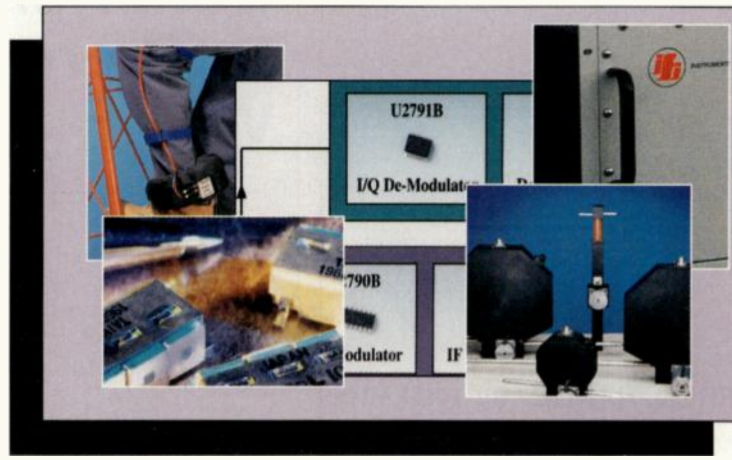

On the cover: Products from – Telefunken Semiconductors (center); Holaday Industries, Inc. (upper left); Instruments for Industry, Inc. (upper right); Electro-Metrics, Inc (lower right); Toko America, Inc. (lower left). These items also appear in the "new products" section (page 46).

departments

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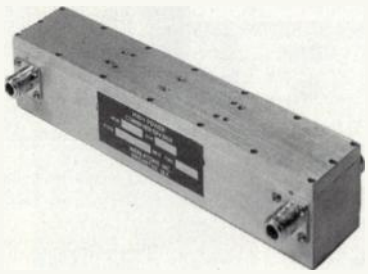
HIGH POWER QUADRATURE HYBRIDS



FREQ. RANGE 100-1000 MHz
POWER 250 WATTS CW

TYPICAL SPECIFICATIONS

MODEL QH3198 100-300 MHZ
MODEL QH3199 300-1000 MHZ

LOSS ....0.5db. max.
AMP.BAL.... ± 0.3db max.
ISOLATION ....20db typ.
PHASE....90 ± 2deg.
VSWR ....1.3:1 max.
CONNECTORS ....N

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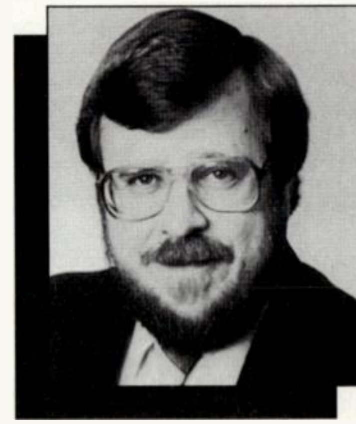
INFO/CARD 4

RF editorial

Getting Data from Here to There via RF



By Gary A. Breed
Editor

Our main feature article in this issue is on the topic of data communications, and we are highlighting new products for this RF application on the first page of our New Products section. We recognize the important role that wireless techniques will play in the continued growth of this part of the communications industry.

If you think about it, RF has played a major role in data communications for many years. The telephone companies have provided dedicated long-distance data lines for major companies since the technology was invented. Even in the early days of computer communications, a large number of links used microwave radio. A bit later, satellites took some of the load, too.

However, local connections using RF are quite new. Sure, there have been some telemetry installations and security monitoring applications, but the numbers have been few. The performance required to adequately and economically replace wired connections is only now being achieved.

The challenges have been varied — from power consumption and small size to data rates and reliable range. The demand created by incredible growth in portable communications and computing fueled the development of RF techniques that finally have reached the necessary maturity for mass-market products. Now, the race is on to bring products to market, to offer exactly the right combination of features, and to do it on the fastest possible schedule.

A major milestone in this race is coming later this year, when the IEEE 802.11 standard for the 2.4 GHz ISM band finally is approved. This technical standard allows interoperability among various companies' equipment, a necessity for large-scale marketing of consumer equipment. 802.11 also sets the stage for future development. The serious technical and commercial sparring that has been part of its development should taper off, making it possible to rapidly address new frequency bands, higher data rates, and improved protocols.

The world depends on computers, and they must communicate with each other to be effective. Adding reliable, cost-effective RF links to the "information infrastructure" will keep computers communicating while their users are on the go!

Reform or Anarchy?

I couldn't resist the temptation to add my two cents to the recent uproar over a suggestion that the Federal Communications Commission be eliminated as part of government downsizing. Although the idea is absurd, plenty of other absurd ideas are being seriously considered on Capitol Hill. I worry that the euphoria of the conservatives, along with the political drift of the President as the election nears, will allow some poor decisions to become law!

Simply put, there is no alternative to the FCC. Its authority cannot be given to the states: Whether radio or wire, communications doesn't happen just within one state's boundaries, much less one country's! In the case of the radio spectrum, it is a valuable and limited natural resource requiring wise management and expert international representation.

To Congress: Look at downsizing; look at efficiency, but don't seriously consider eliminating the FCC!

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GERMANY	EMCO ELEKTRONIK	898562071	898597785
SWEDEN	MAT & TEST TEK.	87926100	87923190

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704FC	4W CW	.5-1000 MHz	33dB	$ 2,095
210LC	10W CW	.008-225 MHz	40dB	$ 2,495
710FC	10W CW	1-1000 MHz	40dB	$ 6,695
*727LC	10W CW	.006-1000 MHz	44dB	$ 7,950
713FC	15W CW	20-1000 MHz	42dB	$ 5,680
225LC	25W CW	.01-225 MHz	40dB	$ 3,295
*737LC	25W CW	.01-1000 MHz	45dB	$ 9,995
712FC	25W CW	200-1000 MHz	45dB	$ 6,950
714FC	30W CW	20-1000 MHz	45dB	$ 9,350
250LC	50W CW	.01-225 MHz	47dB	$ 5,550
715FC	50W CW	200-1000 MHz	47dB	$ 14,990
707FC	50W CW	400-1000 MHz	50dB	$ 10,990
716FC	50W CW	20-1000MHz	47dB	$ 17,950
*747LC	50W CW	.01-1000 MHz	47dB	$ 18,550
116FC	100W CW	.01-225 MHz	50dB	$ 9,500
709FC	100W CW	500-1000 MHz	50dB	$ 16,990
717FC	100W CW	200-1000 MHz	50dB	$ 19,500
718FC	100W CW	20-1000 MHz	50dB	$ 29,800
7100LC	100W CW	80-1000 MHz	50dB	$ 19,500
*757LC	100W CW	.01-1000 MHz	50dB	$ 29,950
122FC	250W CW	.01-225 MHz	55dB	$ 19,950
723FC	300W CW	500-1000 MHz	55dB	$ 29,995
LA500V	500W CW	10-100 MHz	56dB	$ 12,900
LA500UF	500W CW	100-500 MHz	57dB	$ 46,000
LA500G	500W CW	500-1000 MHz	57dB	$ 55,000
LA1000V	1000W CW	10-100 MHz	60dB	$ 22,500
LA1000UF	1000W CW	100-500 MHz	60dB	$ 75,000
LA1000G	1000W CW	500-1000 MHz	60dB	$ 99,000
LS-1000	1000W CW	.01-1000 MHz	60dB	$230,000

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MODEL	RF OUTPUT	FREQUENCY RANGE	GAIN	SPECIAL USA PRICE
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122C	200W CW	.01-220 MHz	53dB	$ 12,950
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140C	2000W CW	.01-220 MHz	64dB	$ 46,500

Warranty: Full 18 months all parts. Vacuum tubes 90 days.
* = Indicates Dual-Band System (coaxial band switching)

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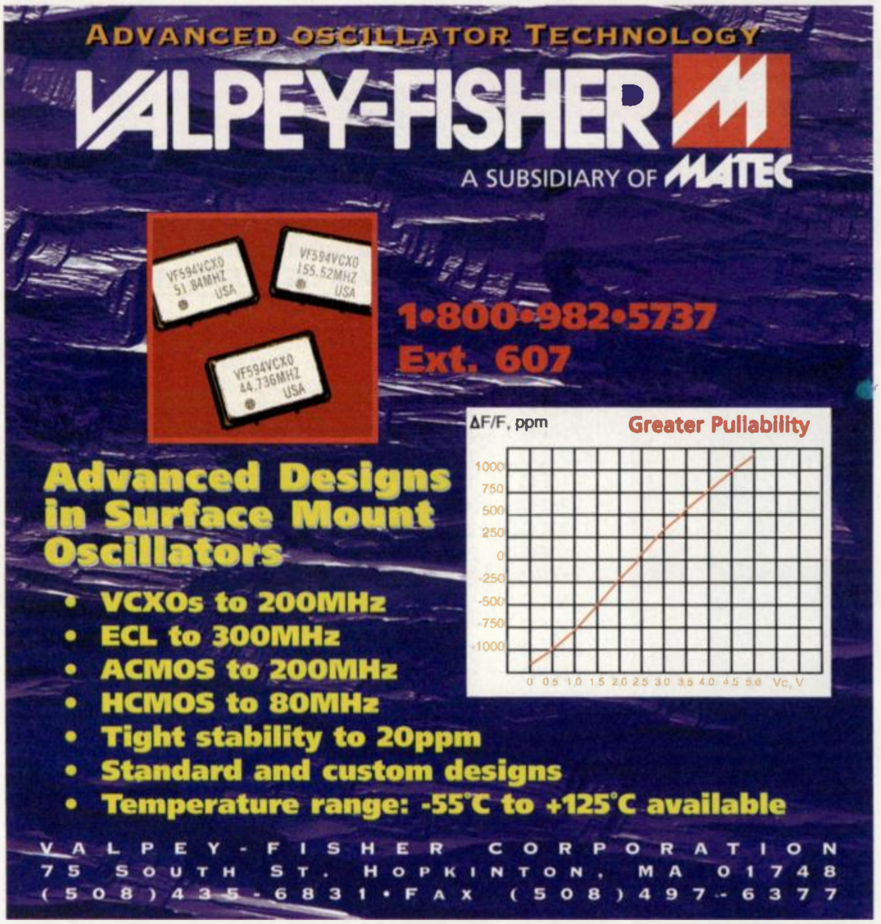

INFO/CARD 6

INFO/CARD 7

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3 dB BW (kHz)

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7.5	2
15	2
30	2

4 pole	qty.
7.5	2
15	2
30	2

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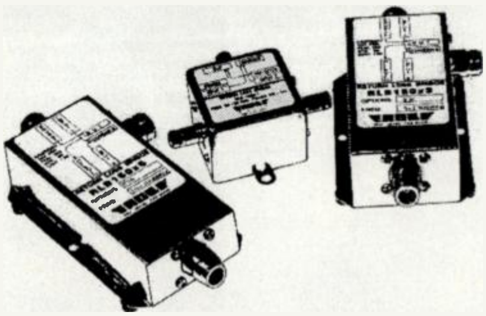
INFO/CARD 8

RETURN LOSS BRIDGES



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RF letters

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

Transmission Line Transformers

Expression 12 in Donald McClure's May 1995 article, "Broadband Transmission Line Transformer Family Matches a Wide Range of Impedances – Part 2" should read:

$$Z_0 + \frac{1}{2Y_0} \qquad (12)$$

Also, capacitors C1 - C4 in Figure 9 are specified as 0.1 µF, 100 VDC, ceramic chip capacitors using X7R dielectric.

Simulating Spread Spectrum

A statement appearing in Stephen Kratzet's May 1995 article, "Simulator Package Models a Spread Spectrum System – Part 2" requires clarification. Towards the top of the first column on page 98, it is stated that "If an integrator type of lowpass filter is required, the 1 in the Laplace denominator can be changed to a zero."

A more complete explanation follows: If an op-amp integrator type of lowpass filter is required, the Laplace denominator can be changed by eliminating the T_2 factor, and changing the 1 to a 0 (zero). Also, because the op-amp inverts, both terms in the numerator are negative.

$$H[s] = \frac{-(T_2)s^1 + (-1)}{(T_2)s^1 + 0}$$

where $T_1 = R_1C_1$, and $T_2 = R_2C_1$

$$H[s] = \frac{-(3.3\ k \cdot 2500\ pF)s^1 + (-1)}{(10\ k \cdot 2500\ pF)s^1 + 0}$$

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Information: Miller Freeman, Kathryn Piersall, 13/6D Noel Road, Suite 500, Dallas, TX 75240. Tel: (214) 419–7969. Fax: (214) 419–7915.

27-29 **Networks '95**
Birmingham, England
Information: IDG World Expo, 111 Speen Street, P.O. Box 9107, Framingham, MA 01701–9107. Tel: (508) 879–6700. Fax: (508) 872–8237.

July

9-12 **30th Annual Microwave Power Symposium**
Denver, CO
Information: IMPI, 10210 Leatherleaf Court, Manassas, VA 22111. Tel: (703) 251–1415.

30-1 **Electrex 95**
Johannesburg, South Africa
Information: Joe Nemchek, International Sales Force, Reed Exhibition Companies, 383 Main Avenue, Norwalk, CT 06851. Tel: (203) 840–5398. Fax: (203) 840–9398.

August

15-17 **Advanced Technology Acquisition, Qualification, and Reliability Workshop**
Newport Beach, CA
Information: Dorthy Connor, General Technical Services, Inc., 3100 Route 138, Wall Township, NJ 07719. Tel: (908) 544–3231. Fax: (908) 389–3992.

27-31 **Surface Mount International '95**
San Jose, CA
Information: Yolanda White. Tel: (415) 905–4994.

September

5-7 **International Conference on 100 Years of Radio**
London, UK
Information: HYR95 Secretariat, Conference Services, IEE, Savoy Place, London WC2R 0BL, UK. Tel: 071 344 5477. Fax: 071 497 3633.

11-14 **17th Annual Electrical Overstress/Electrostatic Discharge Symposium**
Phoenix, AZ
Information: EOS/ESD Symposium, Inc., P.O. Box 913, Rome, NY 13442–0913. Tel: (315) 339–6726. Fax: (315) 339–6793.

12-15 **International Conference on Electromagnetics for Advanced Applications**
Torino, Italy
Information: ICEAA 95, c/o Dipartimento di Elettronica, Politecnico di Torino, Corso Duca degli Abruzzi 24, 10129 Torino, Italy.

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INFO/CARD 11

RF courses

RF/MW Measurement Techniques I
June 12-16, Cambridge, UK
Applied RF Techniques I
June 12-16, 1995, Cambridge, UK
Digital Cellular and PCS Communications – The Radio Interface
June 13-16, 1995, Singapore
RF Component Modeling
June 19-22, 1995, Cambridge, UK
Applied RF Techniques II
June 19-23, 1995, Cambridge, UK
Applied RF Techniques
June 26-30, 1995, Morristown, NJ
Wireless RF System Design
June 26-30, 1995, Morristown, NJ
Information: Besser Associates, 4600 El Camino Real, Suite 210, Los Altos, CA 94022. Tel: (415) 949–3300. Fax: (415) 949–4400.

EMC Workshop - Immunity Measurements
July 4-5, 1995, Surrey, England
EMC Diagnostics Workshop - Emission Measurements
July 6-7, 1995, Surrey, England
MIL–STD–1553B Tutorial
July 12, 1995, Surrey, England
Safe Design of Electrical and Electronic Products
September 7, 1995, Surrey, England
Information: Miss Nikki Hamann, Conference Group, Technical Services Division, ERA Technology Ltd., Cleeve Road, Leatherhead, Surrey, KT22 7SA England. Tel: 44 (0)372–374151 ext. 2595. Fax: 44 (0)372–377927.

DSP Without Tears
September 20-22, 1995, Washington, DC
Information: Z Domain Technologies, Inc., 325 Pine Isle Court, Alpharetta, GA 30202. Tel: (800) 967–5034, (404) 587–4812. Fax: (404) 518–8368.

EMI Design and Product Safety
September 25-28, 1995, San Jose, CA
Information: Compatible Electronics, Inc., 114 Olinda Drive, Brea, CA 92621. Tel: (714) 579–0500. Fax: (714) 579–1850.

Electrical Grounding of Communications Systems
June 21-23, 1995, Madison, WI
Technical Cellular
July 31-August 3, 1995, Madison, WI
Radio System Design for Telecommunications
August 21-24, 1995, Madison, WI
Information: Department of Engineering Professional Development, University of Wisconsin-Madison, 432 North Lake Street, Madison, WI 53706. Tel: (800) 462–0876. Fax: (608) 263–3160.

Kalman Filtering
August 28-31, 1995, Los Angeles, CA
Wavelet Transform Applications to Data, Signal, Image and Video Processing
September 11-15, 1995, Los Angeles, CA
Information: UCLA Extension, Engineering Short Courses, 10995 LeConte Ave., Ste. 542, Los Angeles, CA 90024. Tel: (310) 825–1047. Fax: (310) 206–2815.

Successfully Simulating Circuits with SPICE
June 20-22, 1995, Denver, CO
September 20-22, 1995, Chicago, IL
September 25-27, 1995, Minneapolis, MN
Information: RCG Research, Inc., P.O. Box 509009, Indianapolis, IN 46250–0900. Tel: (800) 442–8272 or (317) 877–2244. Fax: (317) 776–9095.

Microwave Antenna Measurements: Far-Field, Near-Field, Compact Ranges and Anechoic Chambers
June 13-16, 1995, Northridge, CA
Information: Shirley Lang, Center for Research & Services, School of Engineering & Computer Science, California State University-Northridge, 18111 Nordhoff Street, Northridge, CA 91330–8295. Tel: (818) 885–2146. Fax: (818) 885–2140.

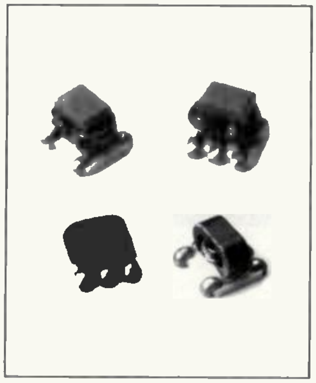
Digital Cellular Radio
July 11-14, 1995, Washington, DC
Electromagnetic Interference and Compatibility (EMI/EMC): A Practical Approach to Testing and Problem Solving
July 24-28, 1995, Washington, DC
Lightning Protection with Emphasis on Electrical and Electronic Equipment
September 7-8, 1995, Washington, DC
Communication Satellite Systems: The Earth Station – A Practical Approach to Implementation
September 11-14, 1995, Washington, DC
Global Positioning System: Principles and Practice
September 11-14, 1995, Washington DC
Information: The George Washington University, Continuing Engineering Education, Academic Center, Room T-308, 801 22nd Street, N.W., Washington, DC 20052. Tel: (202) 994–6106 or (800) 424–9773. Fax: (202) 872–0645.

Practical Aspects of Analog and Mixed-Mode IC Design
July 10-14, 1995, Portland, OR
Information: OCATE, 19500 NW Gibbs Drive, Beaverton, OR 97006. Tel: (503) 690–1460. Fax: (503) 690–1466.

Grounding and Shielding Electronic Systems
July 25-26, 1995, Bloomington, MN
August 9-10, 1995, Suwanee, GA
Circuit Board Layout to Reduce Noise Emission and Susceptibility
July 27, 1995, Bloomington, MN
August 11, 1995, Suwanee, GA
Information: Continuing Education, 119 Mechanical Engineering Annex, University of Missouri - Rolla, Rolla, MO 65401–4992. Tel: (314) 341–4132. Fax: (314) 341–4992.

Microwaves and RF Measurements & Applications
July 10-13, 1995, Monterey, CA
Information: University Consortium for Continuing Education, 16161 Ventura Boulevard, M/S C-752, Encino, CA 91436. Tel: (818) 995–6335. Fax: (818) 995–2932.

Wireless Communication Systems
August 2-4, 1995, Ann Arbor, MI
Information: Engineering Conferences, 400 Chrysler Center, North Campus, The University of Michigan, Ann Arbor, MI 48109–2092. Tel: (313) 764–8490. Fax: (313) 936–0253.



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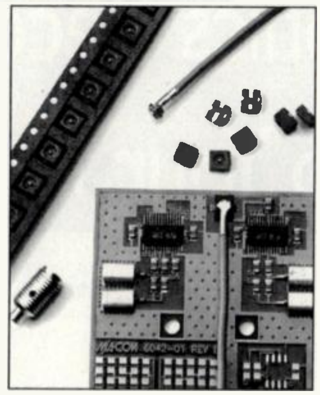

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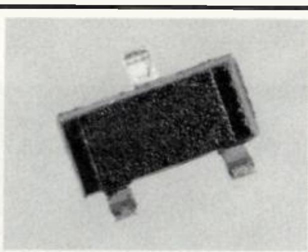

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INFO/CARD 16

Giga-tronics Recommends A Better Path To Your Next Power Meter.

Re-evaluate The Standard

THE STANDARD

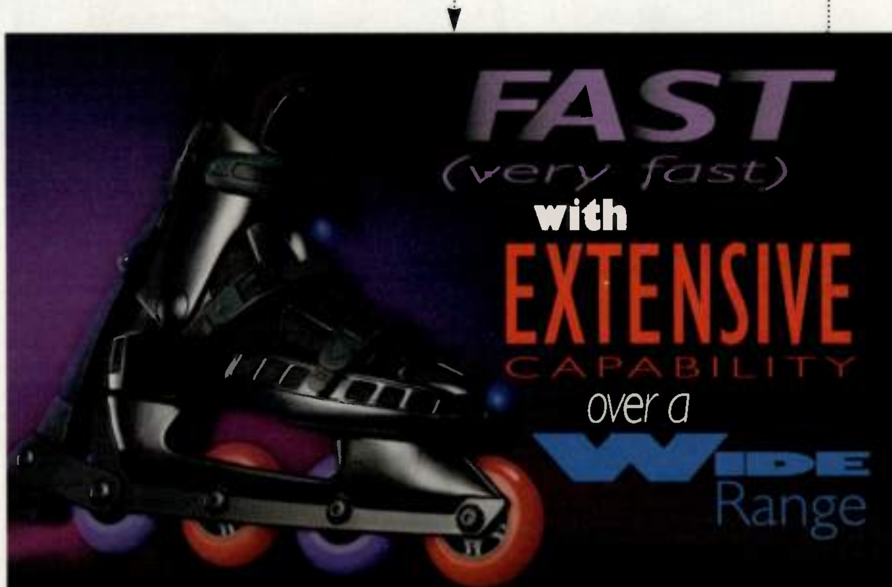
The HP 437B and HP 438A power meters are considered the industry standards.

It's true, there are a lot of them out there, and they've been around a long, long time. But during this time, your needs have changed—significantly.

Imagine the Ideal Alternative

THE ALTERNATIVE

High volume manufacturing requires greater speed and increased dynamic range, and testing today's sophisticated communication devices requires more extensive capability.



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Hewlett-Packard specifications are from the HP 437B and HP 438A technical data sheets dated April 1988 and July 1984, respectively.

Capabilities	Giga-tronics 8541	Hewlett-Packard HP 437B	Giga-tronics 8542	Hewlett-Packard HP 438A
Measurement Speed over GPIB (rdgs/sec)	200	20	200	20
Measurement Speed using Fast Modes (rdgs/sec)	4,000	Not Available	4,000	Not Available
Maximum Dynamic Range with a Single Sensor	90 dB	50 dB	90 dB	50 dB
Direct CW and Peak Power Measurements	Yes	No	Yes	No
Built-in Frequency Cal Factors	Yes	No	Yes	No
Measurement Channels/Display Lines	One/Two	One/One	Two/Two	Two/One
Display Pulse Waveform on External Scope	Yes	No	Yes	No

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Colombia Wandel & Goltermann Andina Ltda. 12-564-001 • **Croatia** Amtest Ltd 54-141-1963 • **Czech Republic** Amtest CA 632-625-34 • **Denmark** Crimp A/S 481-00531 • **England** Sematron UK Ltd 17348-19970
Finland Profelec Oy 805-664-477 • **France** Elexience 1-60-11-94-71 • **Germany** Pro Nova Nachrichtentechnik GMBH 07141-2858-0 • **Greece** American Technical Enterprises 152-406-20 • **Hong Kong** Schmidt Electronics 22507-022-2
Hungary Amtest Ltd 125-155-20 • **India** Technical Trade Links 228-326-719 • **Indonesia** Mecomb Tehnik Division 21-780-8008 • **Israel** Dan-El Technologies Ltd 236-478-770 • **Italy** L.P. Instruments SRL (Milano) 24-84-01-713; (Roma) 68-68-00-119
Japan Midoriya 33-561-5763 • **Korea** Unitek Corporation 255-427-24 • **Lebanon** Mabek 118-644-79 • **Malaysia** CNN Sdn. Bhd. 657-958-4 • **The Netherlands** Air Parts Electronics B.V. 01720-43221
New Zealand John Cutler Instruments 644-234-7203 • **Norway** Vistron Feiring A/S 22-64-90-70 • **Peru** BMP Ingenieros SA 149-012-24 • **Oman** Mustafa & Jawad Trading Co. L.L.C. 870-995-5 • **Poland** Amtest Ltd 264-267-74
Saudi Arabia ARASCO 626-610-006 • **Singapore** Mecomb 469-8833 • **South Africa** PNI Electronics 117-863-647 • **Spain** REMA Leo Haag, SA 1-38-39-01-7 • **Sweden** Ferner Elektronik AB 87-608-360
Switzerland Isatel Emitec AG 42-418-041 • **Taiwan** Schmidt Electronics 25-013-468 • **Turkey** Elektro 2-164-18-1270 • **Venezuela** Equilab C.A. 22-393-233

Giga-tronics Incorporated ▪ 4650 Norris Canyon Road ▪ San Ramon, California 94583 ▪ Telephone: 510-328-4650 ▪ Telefax: 510-328-4700

INFO/CARD 17

FCC Approves New Channels for Cordless Phones

The FCC has made 15 new channel pairs available for use by cordless telephones in order to alleviate congestion on existing channels. Over the years the Commission has considered a variety of alternatives for dealing with congestion on cordless telephone frequencies. The new channels, near 44 and 49 MHz, offer several advantages over other frequency bands. The close proximity of the new channels to existing frequencies will enable continued use of existing technology. Furthermore, new cordless telephones will be able to make use of both the old and new channels to offer twenty-five channel capability. The risk of interference to the public land mobile radio service users that will share spectrum with cordless telephones on the new channels, and to television receivers that use these frequencies internally is very low.

As a corollary to the allocation of additional channels, the Commission also eliminated the rule that permits use of offset channels by cordless telephones. The 15 additional channels being provided now will more than double the number of frequencies available for cordless telephone use, rendering the need for offset moot.

Wireless Communications Behind Growth in GaAs Market

The gallium arsenide and gallium phosphide wafer business is forecast to reach $360 million in 1999, growing from $250 million in 1994, according to a recently published report by Strategies Unlimited. The growth in wireless communications is creating demand for semi-insulating GaAs wafers at a fast pace, while the production of optoelectronic devices is driving semiconducting GaAs and GaP wafer usage.

Call For Papers

The 1995 Wireless Circuits, Interconnection, and Assembly Workshop has sent out a call for Papers for their October 8-11, 1995 show. The Workshop will be held in Tucson, AZ. Suggestions for paper submissions include: new applications and designs; MMIC and hybrid circuits; packaging techniques; advancement in materials; materials testing; design and implementation problems; software tools; processing and quality; and high volume manufacturing methods. Abstracts of proposed papers should be submitted by June 30, 1995 to Sharon Aspden, Wireless Workshop, 100 S. Roosevelt Avenue, Chandler, AZ 85226. Phone (602) 961–1382. Fax (602) 961–4533.

Wide Range RFID and Tracking System Introduced

Savi Technology has developed a wide area radio frequency identification (RFID) and tracking system. Transporters, the military, commercial and industrial companies will be able to locate and identify thousands of assets (items) in transit or stored over a wide area — automatically and simultaneously — in minutes.

The Savi system is comprised of radio tags, hand-held or fixed omnidirectional transmitter/receivers and a central computer. It uses advanced radio technology for true two-way communication that enables assets to be automatically located and identified via accurate, high speed data transfer. All Savi products operate below the FCC Part 15 power level so no site license is required. The proprietary design allows them to operate at 315 MHz and at 433.92 MHz with no interference.

The Savi system uses a variety of transporter ID tags that incorporate a miniature radio transmitter, radio receiver and micro computer in a strong, tamper-proof compact case. The tags have a memory capacity of up to 128,000 bytes and are powered by lithium/manganese dioxide batteries with a five-year battery life.

Savi's technology is capable of locating assets automatically, without human intervention, while simultaneously identifying the assets and communicating with them through a high speed data transfer.

FCC Expands Rules

The FCC's recent decision to reaffirm use of the 902-928 MHz Industrial Medical and Scientific (ISM) band for commercial radio location will boost mobile communications, top Pinpoint Communications officials said. Under Report Order 95-41, the FCC modified its rules to allocate permanently, spectrum for location and monitoring services (LMS) in this band.

The new rules broaden the applications of LMS to any mobile object. The permanent FCC rules explicitly allow for the wide-band forward link pioneered by Pinpoint. The FCC's allocation of spectrum for LMS allows Pinpoint to deploy its 8-MHz broadband networking technology.

Call For Papers

The 1995 IEEE International Electron Devices Meeting will be held December 10-13, 1995 in Washington, DC. Papers in the following areas are requested: integrated circuits; quantum electronics and compound semiconductor devices; device interconnect technology; CMOS devices and reliability; solid state devices; modeling and simulation; vacuum electronics; detectors, sensors and displays. Abstracts must state purpose of work, manner in which it advances the art, and specific results and their significance. Abstracts are due by July 5, 1995 to Melissa Widerkehr, IEDM, Suite 270, 101 Lakeforest Boulevard, Gaithersburg, MD 20877. Phone (202) 986–1137.

IC-EMC Task Force Approves Test Procedures

In March 1991 the SAE EMC committee created an IC-EMC Task Force to investigate EMC test methods that would evaluate IC level emissions and immunity. The task force has representation from automotive, consumer electronics, communication equipment, IC manufacturers, and EMC instrumentation manufacturers.

In October 1994 SAE document J1752/2, *Integrated Circuit Radiated Emissions Diagnostic Procedure, 1 MHz to 1000 MHz, Magnetic Field Loop Probe*, was approved by the SAE EMC committee. This procedure defines a method for evaluating the near field magnetic component of the electromagnetic radiation from an integrated circuit. The method uses a loop probe to measure narrowband magnetic RF emissions from 1 MHz to 1000 MHz at a controlled distance and orientation from an integrated circuit. This method is applicable to measurements from an individual IC mounted on a standardized test fixture or in-situ on a circuit board.

The second SAE document J1752/3, also approved, was *Integrated Circuit Radiated Emissions Test Procedure, 150 kHz to 1000 MHz, TEM Cell*. This

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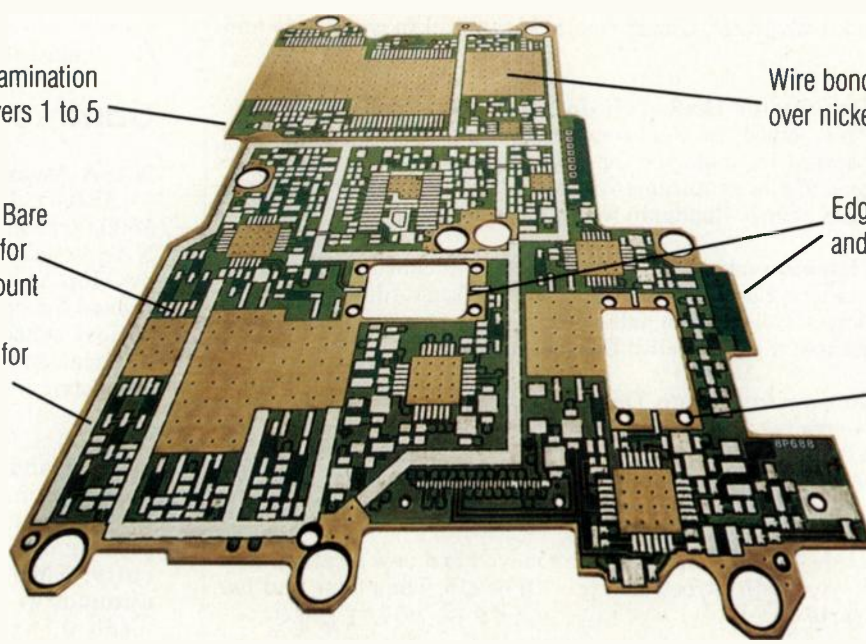

This high frequency microwave printed circuit board meets all MIL, ANSI, and IPC specifications and standards for electrical performance and physical tolerances. The innovative techniques used to fabricate this board put us head and shoulders above ordinary board shops.

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INFO/CARD 18

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Business Briefs

Joint Venture Creates Mitsubishi Wireless Communications Co. — Mitsubishi Electronic Corp. and Mitsubishi Corp. announced the creation of a joint venture establishing a re-aligned U.S. company dedicated to development, marketing and distribution of wireless communications products. The new company will merge and relocate to Braselton, CA where it will be adjacent to the manufacturing facility of Mitsubishi Consumer Electronics America Inc.

Instrument Specialties Acquires Equity Interest in TCS — Instrument Specialties has acquired an equity interest in T.C. Shielding Limited (TCS) based in Ross-on-Wye, Herefordshire, United Kingdom. Terms of the purchase include a technology agreement between the two companies. In addition, they will participate in joint manufacturing and worldwide sales and marketing.

European Compliance Available at Compatible Electronics — Compatible Electronics can now perform CE mark testing to the European Union's EMC directive for both emissions and immunity. They have added an eighth open field test site and their third semi-anechoic chamber has been expanded with ferrite tiles, providing field uniformity from 27 to 1000 MHz.

Tokin Announces Distribution Network — Tokin America, Inc. has signed agreements with six distributors covering the United States and Canada to market Tokin's line of passive components. Tokin's products include surface mount EMI filters, inductors, coils, transformers, ceramic chip capacitors and ferrite cores.

Cadence and Semiconductor Research Join for Research — Cadence Design Systems, Inc. has joined the Semiconductor Research Corp., targeting the research of new design technologies for deep submicron processes, electronic systems design and the requirements of the Semiconductor Industry Association National Technology Roadmap for Semiconductors.

Richardson Opens Headquarters in Amsterdam — Richardson Electronics, Ltd. has opened its new European headquarters in Amsterdam. Formerly operating out of its Argenteuil, France sales office, the new corporate facility in The Netherlands is the company's tenth European location.

Comlinear Offers Access to Design Data — Comlinear Corporation now offers a 24-hour, automatic fax-back service that gives callers immediate access to detailed design data on the company's high-performance components. Included in the FaxCOM database are complete data sheets and application notes for Comlinear's products. Designers can access FaxCOM at (800) 970–0102.

AeroComm, Inc. Moves — AeroComm, Inc. has moved to a new location. The new address is 13228 West 99th Street, Lenexa, KS 66215. The phone and fax numbers remain the same.

Proxim and Fujitsu Expand Strategic Relationship — Proxim, Inc. has announced an expanded strategic relationship with Fujitsu Personal Systems, Inc. (FPSI). The relationship is based on FPSI's integration of Proxim's RangeLAN2™ technology into Fujitsu's Stylistic 500, and Intel 486DX2-50 Windows-based tabled computer.

Chesapeake and Loral Form Alliance — Chesapeake Microwave Technologies Inc. and Loral Microwave-Narda have formed a strategic alliance that covers the development and manufacture of RF products for the wireless industry. The initial focus will be on cellular, PCS and related products in the U.S. and international marketplaces.

Lindgren Opens Office in Southeast Asia — Lindgren RF Enclosures, Inc. has opened a sales office in Southeast Asia in order to increase their global EMI/RFI shielding product and service offerings.

procedure defines a method for evaluating the electromagnetic radiation from an integrated circuit. The method uses a standardized IC test board containing the IC being evaluated mounted to a mating port cut in the floor of a TEM cell. This standardized test board controls the geometry and orientation of the operating IC relative to the TEM cell and eliminates any connecting leads within the cell (these are on the back side of the board which is outside the cell). One of the EMC cell ports is terminated with a 50 ohm load and the other one is connected to the input of a spectrum analyzer which measures the narrowband RF emissions over the frequency range of 1 MHz to 1000 MHz emanating from the integrated circuit.

These IC-Radiated-Emissions documents will help the various IC users make comparisons and decisions based on their EMC needs early in the electronic circuit design process. The documents can be obtained from SAE Headquarters, Phone (412) 776–4841.

Contracts

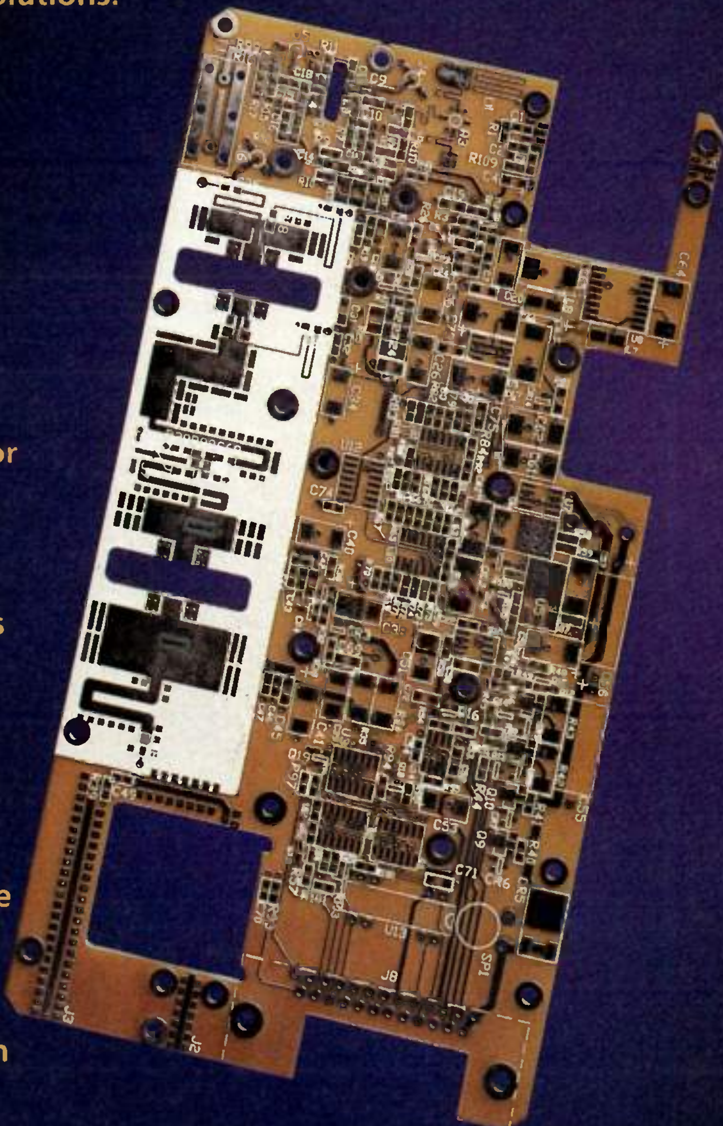
NASA Awards Contract to SRICO — SRICO, Inc. has been awarded a $600,000 Phase II SBIR contract from NASA Lewis Research Center to develop an integrated optic voltage sensor for space power systems. The optical sensor has many commercial applications including measurement of electric field and voltage in electric power systems and hazardous environments; lightning detection in avionics and mining; high speed electro-optic modulators in fiber optic communication systems; non-contact probing of high speed integrated circuits; EMI/EMC measurements in automotive electronics testing; biomedical engineering and instrumentation; and charge measurement and control in photocopiers and ion neutralization systems.

Scientific-Atlanta Wins Optus Vision Contract — Australia's Optus Vision has awarded Scientific-Atlanta a major contract to supply equipment for the transmission of video, data and telephone signals to homes linked by the Optus Vision broadband digital network. Along with the telephone service and standard broadcast channels, Optus Vision's network will provide interactive and multi-media applications.

Maximize Your Competitive Advantage

"Mixed Dielectric" Multilayer Circuits

Poly Circuits commitment toward improving their customers' ability to compete in today's Microwave and Communications Markets has led to significant accomplishments with cost effective Printed Circuit Board solutions.



The concept of the "Mixed Dielectric" Multilayer Circuit is to combine high frequency PTFE (high cost) and low frequency (low cost) circuits into a single cost effective circuit assembly with the following advantages:

Reduced Cost:

- Minimize use of costly high frequency high performance PTFE circuit materials.
- Maximize use of inexpensive material for majority of assembly.
- Allows surface mount pick and place operation for a single assembly.
- Eliminates cost of external interconnects between separate circuits.

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- Permits high performance PTFE material in critical signal path.
- Eliminates performance loss from external interconnects between separate circuits.

Our engineers not only helped make the "Mixed Dielectric" Multilayer Circuit into today's reality, they have perfected it to an art form.

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The Technology Leaders for Microwave Circuits

INFO/CARD 19

Manufacturers Track Markets with Agile Manufacturing Techniques

By Andy Kellett
Technical Editor

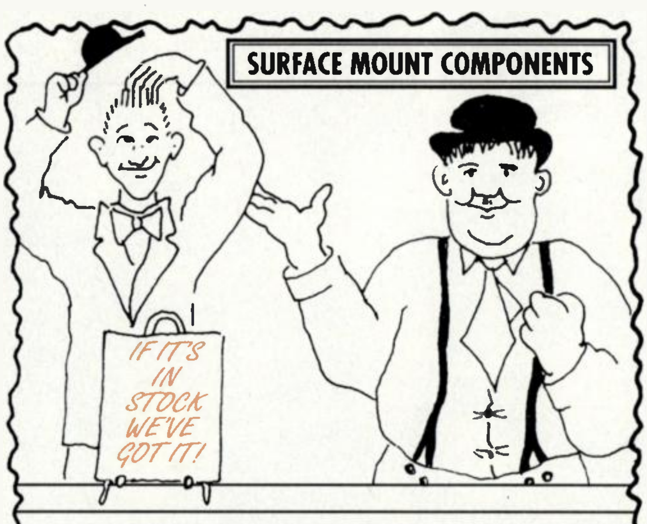
The way RF manufacturers make money in the RF market has changed, and so has the way they make their products.

Getting the Parts

Outsourcing, or contracting with other companies to manufacture part, or even all of a product, has become common practice for many RF manufacturers. "In the eighties, vertical integration was popular, in the nineties, it's outsourcing," says Steve Ulett, Marketing Communications Manger for Penstock.

Outsourcing has gained popularity for two reasons. First, many companies try to get into markets in which they have incomplete expertise. Outsourcing allows companies to enter those markets without developing the expertise on their own. Second, companies are wary of letting their corporate structure sprawl. "We think our organizations are already as big as we like to manage," says Mark Testa, Director of Manufacturing Technology for Motorola's Paging Products Group.

According to Penstock's Ulett, there are very few suppliers left that are "just component distributors". Ulett says most component distributors offer "value-added" services to go with the components they sell. Penstock has even become something of a contract manufacturer, having produced such assemblies as military transceivers and control boards for rides at Disney World.



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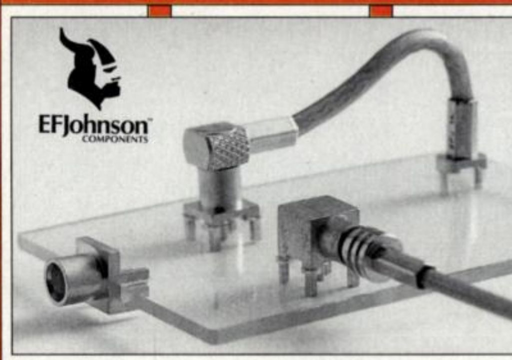

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With the shortage of real estate - why give up valuable board space to larger RF connectors? Our MCX is 30% smaller than SMB connectors and an excellent choice for high density packaging requirements. With a rugged snap-on mating design, these crimpable contact connectors allow for fast and dependable crimp-crimp assembly. Our MCXs employ .020" standoffs for improved board cleaning and are available with .100" leg spacing. The original E.F. Johnson Company end launch connector is also part of the MCX family.

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INFO/CARD 21

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Available with LO drive levels from +7 to +17dBm, performance features include very low conversion loss flat over the entire band, high isolation (L-R, L-I), and well-matched VSWR at all ports.

All-welded internal and external construction is used to assemble and package the TUF-unit in its tiny 0.5 by 0.2 by 0.25 in. metal case, for plug-in or surface-mount* assembly.

TUF-Ultra-Rel™ mixers are guaranteed for five years and boast unprecedented "skinny" sigma (δ) unit-to-unit repeatability as shown in the Table.

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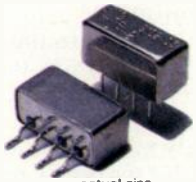

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TUF-3MH	13		5.0	0.33	46	8.95
TUF-3H	17		5.0	0.33	50	10.95
TUF-1	7	2-600	5.82	0.19	42	3.95
TUF-1LH	10		6.0	0.17	50	5.95
TUF-1MH	13		6.3	0.12	50	6.95
TUF-1H	17		5.9	0.18	50	8.95
TUF-2	7	50-1000	5.73	0.30	47	4.95
TUF-2LH	10		5.2	0.3	44	6.95
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TUF-5	7	20-1500	6.58	0.40	42	8.95
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■ $\overline{X}$ = Average conversion loss at upper end of midband ($f_U/2$)

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Putting Them Together

It is possible to run a high speed production line where every unit going through the line is different – thanks to computer integrated manufacturing (CIM). As each unit moves from workstation to workstation, the computer keeps track of where the unit is and what needs to be done to each unit. For example, in the manufacture of pagers, one pager can be programmed to give prompts in English, while the next one on the line can be programmed to give prompts in Spanish, says Motorola's Testa.

Devices which are easily programmed, trimmed, or otherwise altered during the manufacturing process make this type of automation possible. Testa points out his company's increased use of laser trimmable resistors and capacitors, and increased use of software configurable products.

Depending on the product, all or some of the products coming off the end of a production line are then fed into some sort of automatic test equipment. Improved automated handlers and faster test equipment makes it increasingly feasible to test every unit.

Quality Assurance

Getting ISO 9000 certification has become an important stamp of approval for a company's manufacturing process. How important has ISO 9000 become to selling an RF product? "ISO 9000 is not recognized by the consuming public," says Tony Mirabelli, Vice President of Marketing for Uniden, "however, it does speak to a standard of product, and for our industrial or commercial customers ISO 9000 is significant."

Other standards of quality have importance in the market as well. Many companies are including statistical data in their ads and catalogs (for example Mini Circuits' inclusion of mean and standard deviation figures for many of their device parameters).

Environmental Testing

"With ISO 9000 and various minimum quality standards that need to be met, a lot of people in the private sector are waking up and realizing they have to do more environmental testing," says David Russell, Sales Manger for temperature test equipment maker Sigma Systems. Both Uniden's Mirabelli and Motorola's Testa say their companies' accelerated life testing programs are important for finding problems in their manufacturing processes.

Hot/cold plates have made temperature testing of devices, especially semiconductor devices, much easier says Sigma Systems' Russell. Using a hot/cold plate, devices can be maintained at a constant temperature, and can be left accessible to test probes, something that can be difficult using temperature chambers.

Conclusion

In the past, manufacturers bid on high priced contracts to make a small number of individually crafted devices. Today, manufacturers have to identify markets, design a device to satisfy the market's need, and then produce the device at a lower cost and in less time than their competitors. In order to do this, companies will continue to develop new manufacturing techniques and technologies. *RF*

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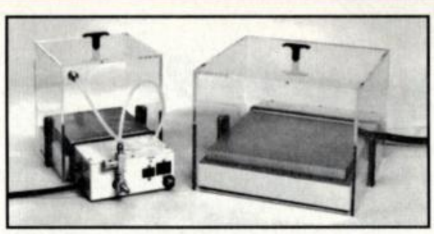

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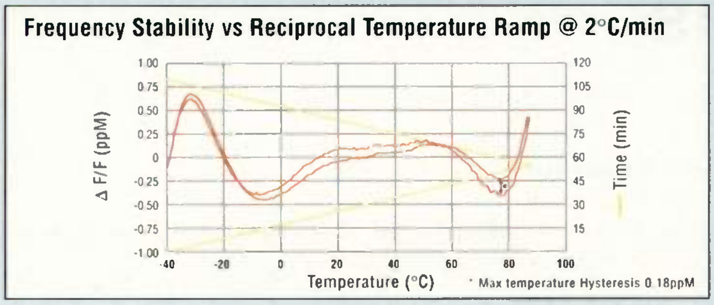

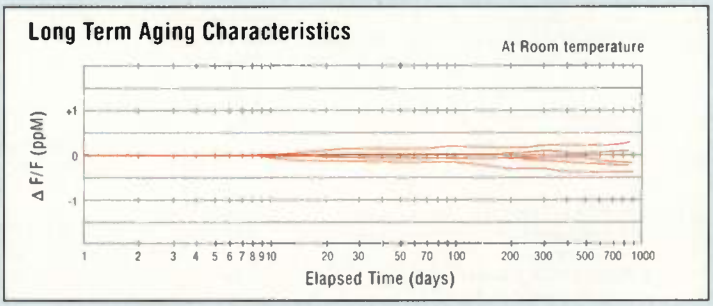

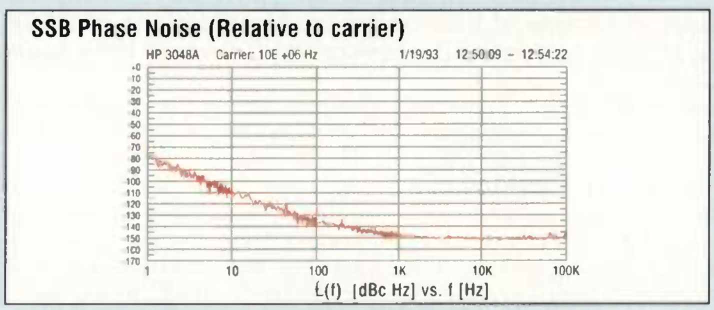

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DSP Implementation of GFSK, GMSK and FQPSK Modulated Wireless Systems

By Hongying Yan, Michael Soderstrand, Jörg Borowski and Kamilo Feher
University of California at Davis

Digital Signal Processing (DSP) and filtering inventions of Dr. Feher & Associates led to implementations of new generations of efficient Gaussian filtered Frequency Shift Keyed (GFSK) and Guassian Minimum Shift Keyed (GMSK) digitally modulated wireless systems. Design examples of these standardized modulation systems for IEEE 802.11, DECT, GSM and CDPD specified applications illustrate the cost-saving benefits of these inventions. Standards compatible, simpler and more power/spectral-efficient FQPSK (Feher's QPSK) implementations are also described. With Read Only Memory (ROM) look up table implementation of an FQPSK processor, the hardware gate count of the filter can be reduced by a factor greater than eight compared to a traditional Gaussian Finite Impulse Response (FIR) filter design. While the number of packed configurable logic blocks (CLBs) needed for the novel filter design in a XILINX FPGA is reduced from 67 to 8, the modulated GFSK signal still meets the standardized system requirements.

Illustrative examples described in this article assume noncoherent Voltage Controlled Oscillator (VCO) structures for generating GFSK modulated signals. Figure 1 shows the block diagrams of both GFSK and GMSK which were used for simulations and hardware RF spectral measurements. For GMSK signal generations a modulations index m = 0.5 (exact value is required). For this reason, recent GMSK implementations such as GSM standardized systems use crosscorrelated quadrature modulator structures, such as patented by Kato/Feher [4] and described in Feher's recent book [1] and in reference [2].

In Figure 1(a) a simpler FQPSK-1 baseband filter is used as a premodulation filter for GFSK, while in Figure 1(b) a conventional structure with standard Gaussian filter implementations is illustrated. The FQPSK-1 baseband filter was designed at the University of California, Davis (UC Davis) using a XILINX FPGA which is described in the following section. We use another UC Davis designed FIR filter design for gate count comparison purposes and the Gaussian baseband filter of National Semiconductor's LMX2411 DECT chip for comparative system performance measurements.

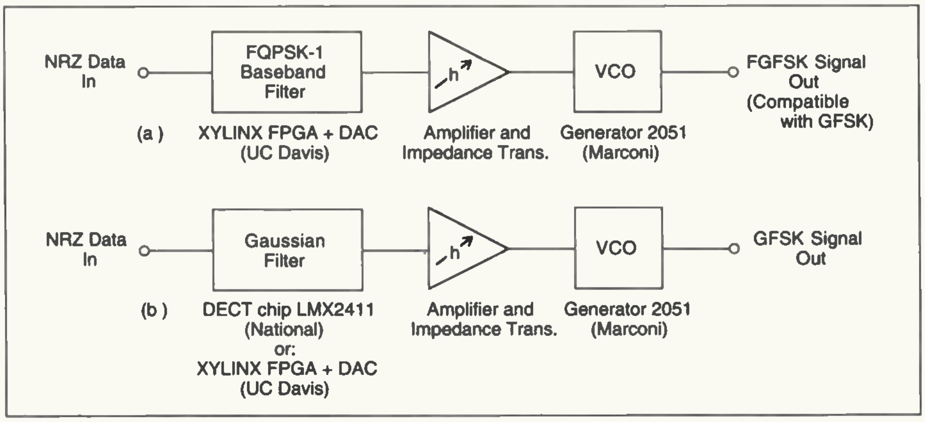


Figure 1. Block diagrams of VCO-based implementations of GFSK radio systems: (a) FQPSK baseband processor, (b) Gaussian filter example.

Filter Implementations

Until the late 1980s, designers devoted their attention to "linear" filter implementation means, including analog (passive and active) and DSP filters. The DSP filters are based on Infinite Impulse Response (IIR) and Finite Impulse Response (FIR) designs. These "linear transversal filter structures" are in extensive use. They are implemented based on certain transfer function, H(z), specifications. H(z) is only an approximation of the analog transfer function H(s), with an acceptable approximation error.

Figure 2 shows the frequency response of an FIR-based Gaussian filter with, $BT_b = 0.5$, where B is the 3 dB filter bandwidth and T_b is the bit duration. From an implementation point of view, the increase of steps means that the number of delays, adders and multipliers is greatly increased, and each multiplier requires many gates. For high speeds such DSP-based filters could be very "power hungry" and could require far too many gates.

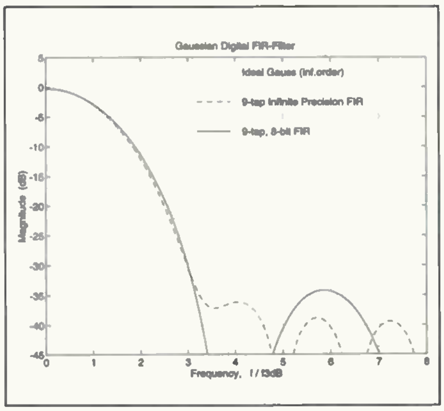
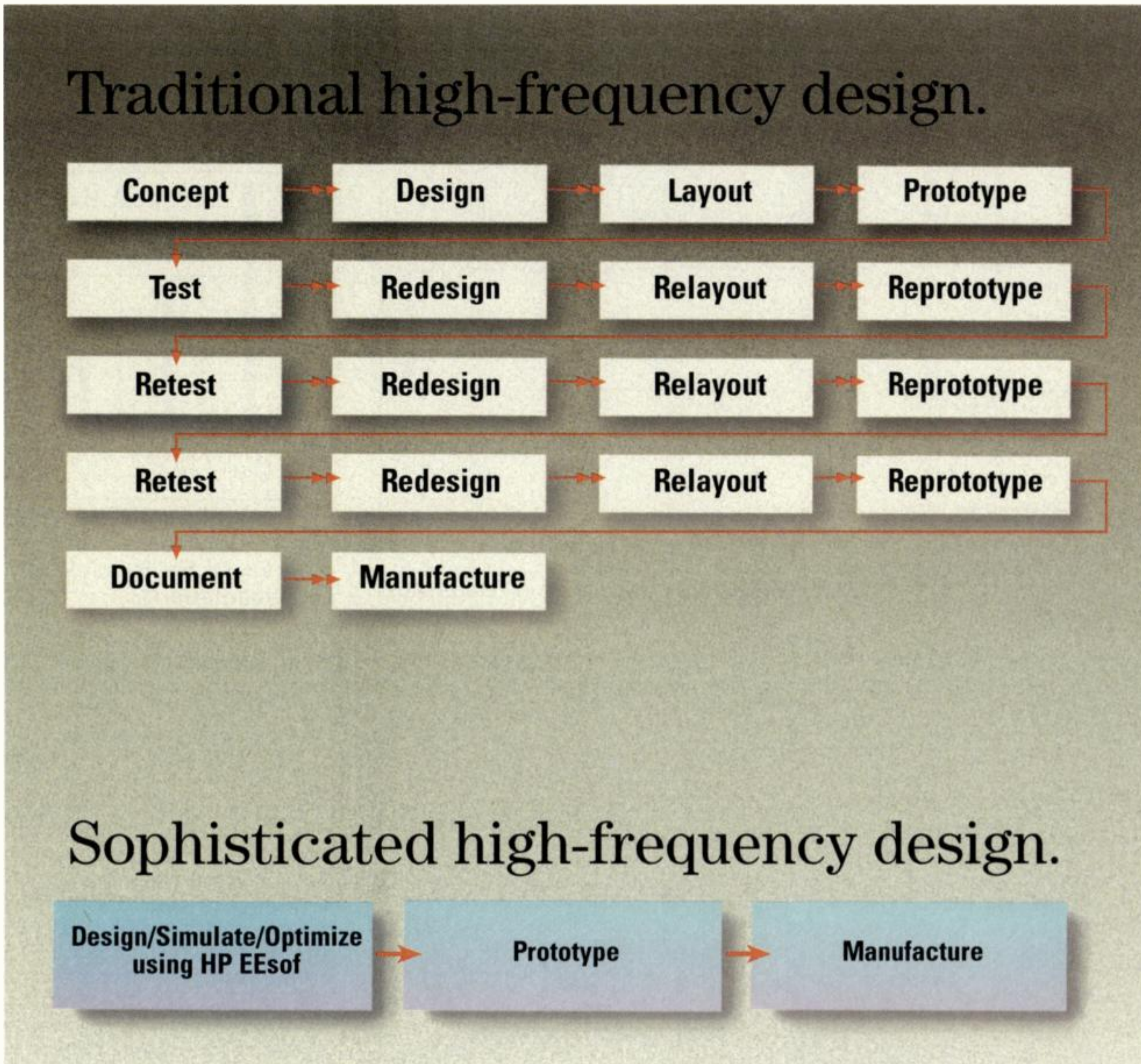


Figure 2. Frequency response of a Gaussian, $BT_b = 0.5$ FIR filter.



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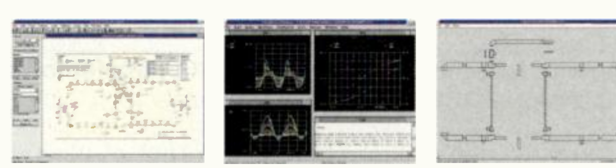

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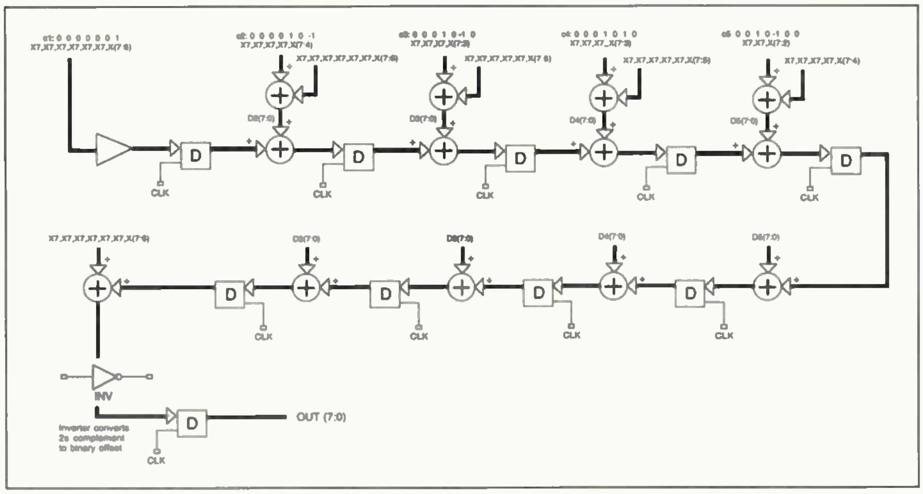


Figure 3. Schematic for a XILINX FPGA implementation of Gaussian, $BT_b = 0.5$ FIR filter.

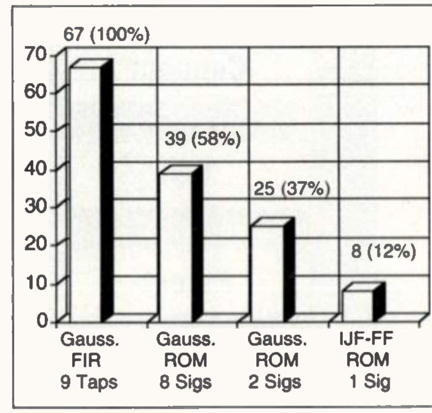


Figure 5. XILINX FPGA gate count comparison in numbers of packed CLBs for several filter implementations.

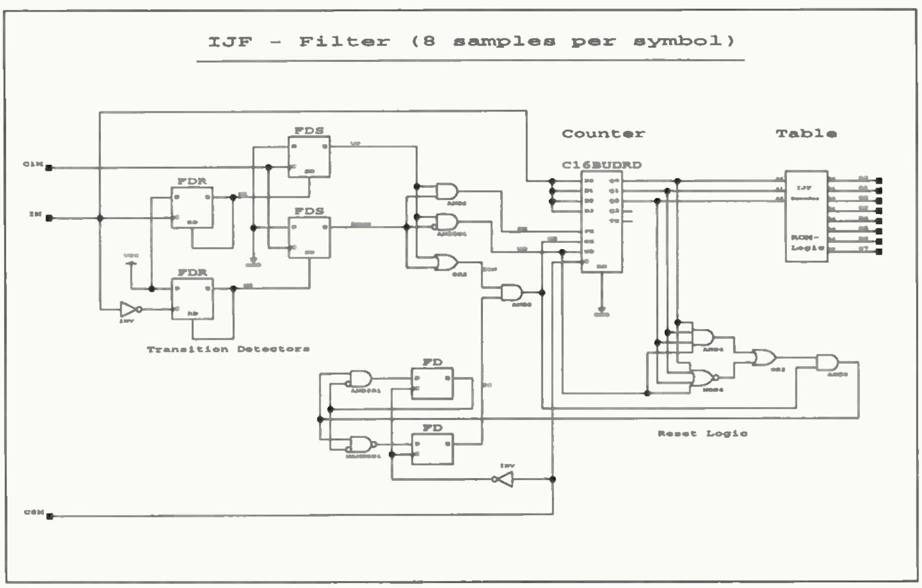


Figure 4. Schematic for a XILINX FPGA implementation of ROM look-up table based FQPSK-1 baseband filter.

It is a difficult task to reduce the complexity of FIR or IIR filters, since simpler, reduced complexity filters do not lead to accurate complex high-order filter implementations.

Figure 3 shows a schematic for a XILINX FPGA implementation of Gaussian, $BT_b = 0.5$ FIR filter with 8-samples per data bit, 9-tap impulse response length, and 8-bit signal resolution.

Figure 4 is a schematic for a XILINX FPGA implementation of ROM look-up table based FQPSK-1 baseband filter with 8-samples per data bit, 1 bit signal length and 8-bit signal resolution. This design uses a patented filter structure [1-7] and reduces the number of packed CLBs by a factor of about 8 compared to the design illustrated in Figure 3.

The principles and implementations of more efficient ROM based look-up table and a large class of nonlinear systems were patented and described by Feher [1-7]. A brief review of some of the applications of these technologies follows:

Feher's Filter (FF) family products have been implemented with ROM-based DSP architectures. In the ROM various signal shapes, e.g., $s_1(t)$ ..., $s_8(t)$ are stored or it is arranged that the ROM is used as a waveform selector/generator/switch. Depending on bit by bit or multibit comparison between data patterns, these stored waveforms are "read out" or switched to the D/A converter, converted into an analog waveform and transmitted.

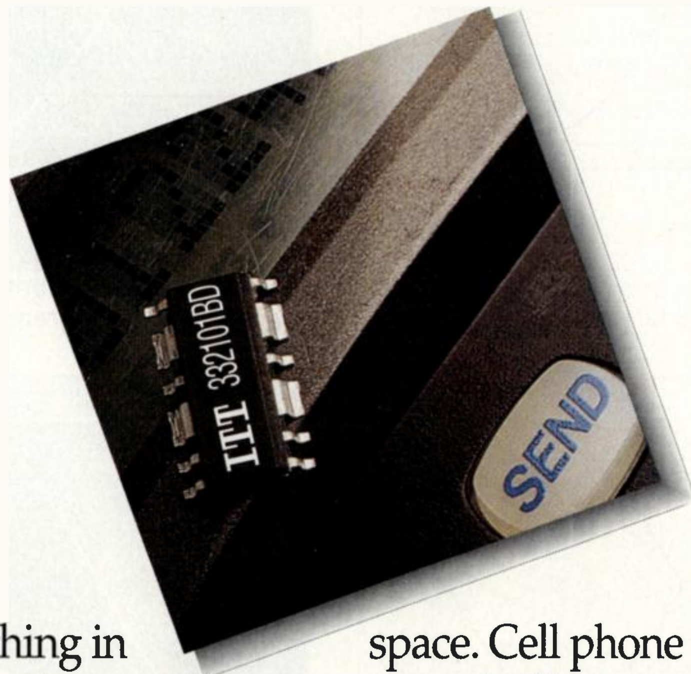
Using this method to implement a linear filter, first truncate the pulse response of the filter to certain length, and generate all possible combinations, then oversample, quantize and map them with the input signals into a table. As the tail of the pulse response with small magnitude has little impact on the quantified magnitude value, the inaccuracy in this approach is mainly due to the quantization noise which exists in any digital process. This implementation requires a shift register, a counter and a ROM.

The advantage of the look-up table method is its simplicity and reduced power consumption, which is critical for mobile communication units. Another advantage of the ROM look up table implementation is that it could be further simplified by reducing the memory bits required. The way to achieve this is by further truncation of the response. If it causes a larger degradation than permitted, you can modify the wave shape a little bit to reduce the impact without affecting the complexity.

67 packed CLBs are needed for the conventional Gaussian FIR filter (Figure 3) and only 8 for the Gaussian compatible FQPSK-1 baseband filter design (Figure 4).

The numbers of packed configurable logic blocks (CLBs) needed in the FPGA implementations for the Gaussian FIR and the FQPSK-1 baseband intersymbol interference and jitter-free (IJF) patented filter

Small talk.



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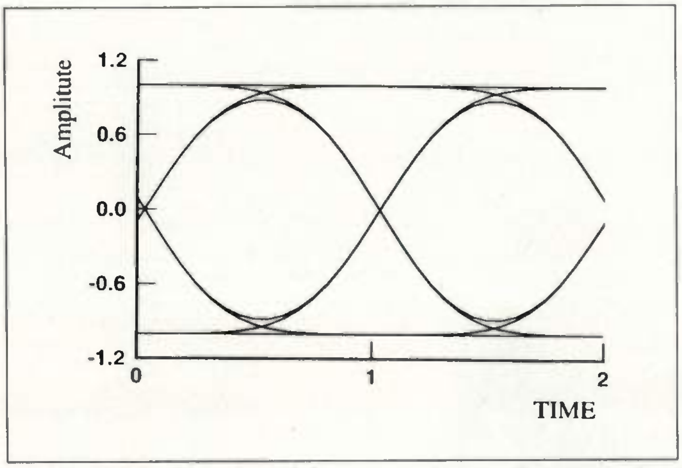


Figure 6(a). Simulated baseband eye diagram after an ideal Gaussian, $BT_b = 0.5$ filter.

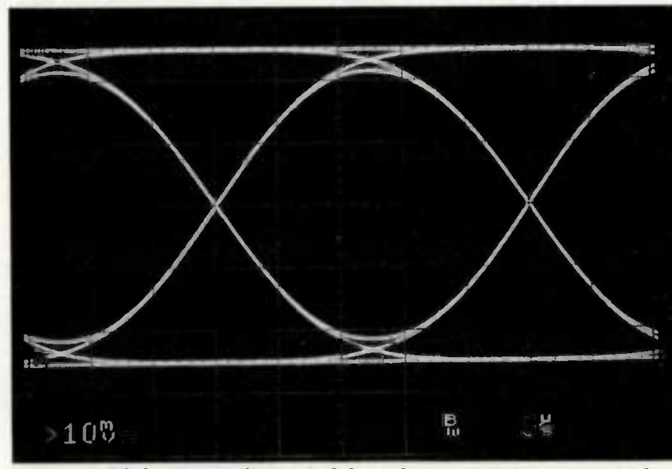


Figure 6(c). Experimental hardware measurement — baseband eye diagram after a Gaussian, $BT_b = 0.5$ filter.

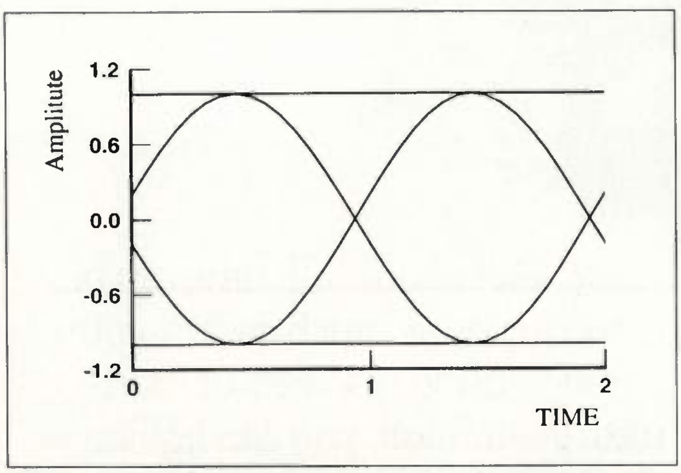


Figure 6(b). Simulated baseband eye diagram after an ideal FQPSK-1 baseband filter.

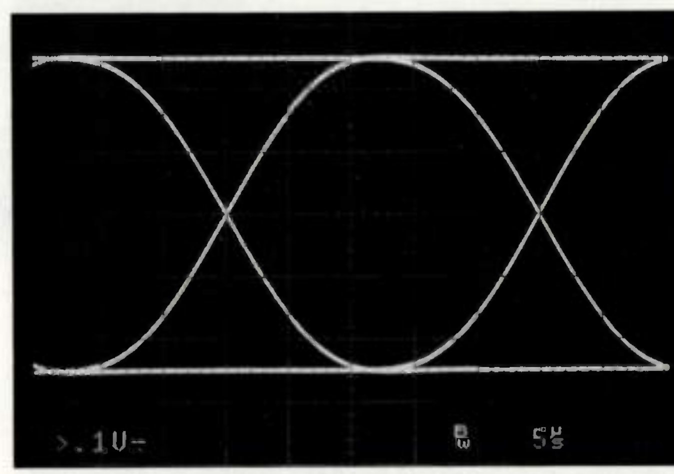


Figure 6(d). Experimental hardware result — eye diagram after a FQPSK-1 baseband filter.

designs are shown in Figure 5. The left column belongs to the Gaussian FIR filter of Figure 3 and the right column to the FQPSK-1 baseband filter implementation of Figure 4. The columns in the middle of the graph represent two ROM-based Gaussian filter designs [15].

The comparison charts of Figure 5 illustrate the tremendous hardware complexity reduction when the methods of [3,4] are used — by a factor 8.4 when using the FQPSK-1 baseband filter instead of a standard Gaussian FIR filter implementation. Only eight packed CLBs are needed (FQPSK-1) instead of 67 (Gaussian FIR). If compared to the best optimized ROM-based Gaussian filter implementation the hardware reduction for the proposed FQPSK-1 baseband filter implementation remains still a factor of 3.1 (25 CLBs vs. 8). An extraordinary hardware saving with a new FQPSK-1 filter design, which is compatible with GMSK specified standards.

Transmitter Signal Comparison

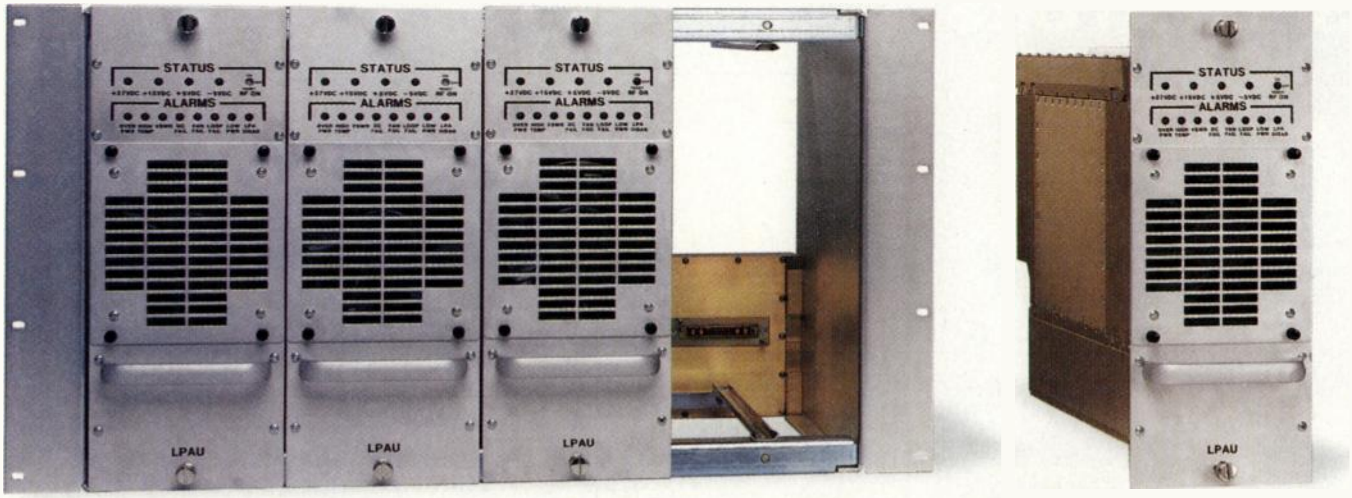
The Gaussian compatible FQPSK-1 baseband filter can potentially be used instead of the Gaussian, $BT_b = 0.5$ filters in GFSK and GMSK wireless applications. The eye diagrams and the power spectral densities are very similar, practically the same in both cases.

The simulated and measured eye diagrams of the filter signal outputs are shown in Figure 6 for both Gaussian, $BT_b = 0.5$ filter and the simpler, Gaussian FQPSK-1 baseband filter. Both signals are jitter-free at the bit interval transitions. The Gaussian, $BT_b = 0.5$ filtered signal has some intersymbol interference (ISI) in the center of the bit interval. The FQPSK-1 filtered signal is intersymbol-interference (ISI) free in the center of the bit interval. Neglecting the unessential ISI of the Gaussian, $BT_b = 0.5$ filtered signal the eye diagrams are the same in both cases.

Figure 6(c) shows the experimental hardware measurement of a baseband eye diagram after Gaussian, $BT_b = 0.5$ filter at normalized scaled down bit rate of 40 kb/s (measurements were also done at 1 Mb/s), using the LMX2411 (National Semiconductor); 9 samples/bit; 8 bits resolution.

The experimental result of Figure 6(d) is after a FQPSK-1 baseband filter at normalized scaled down bit rate of 40 kb/s (measurements were also done at 1 Mb/s), using the XILINX-FPGA (UC Davis); 8 samples/bit; 8 bits resolution.

Figure 7 compares computer simulation and hardware experimental



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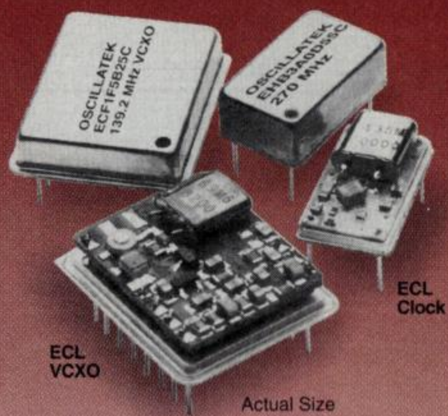

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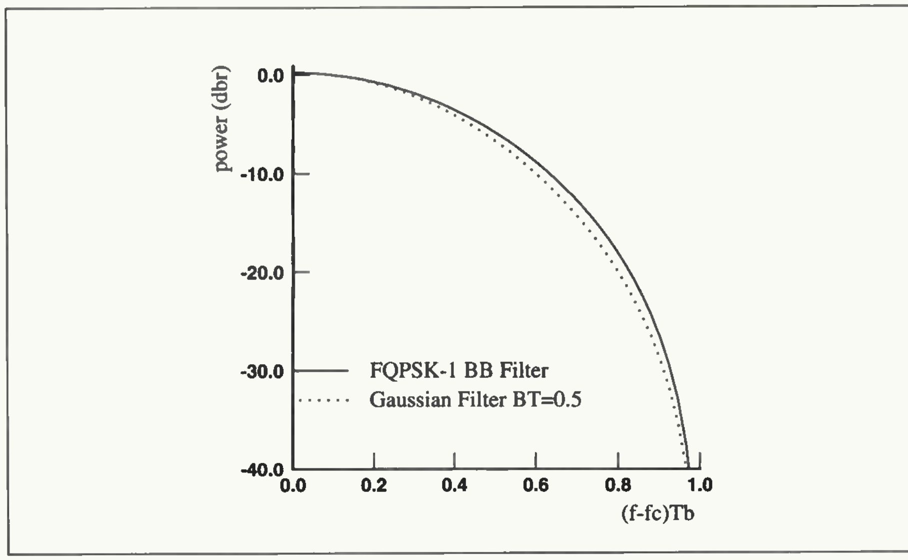

Figure 7(a). Simulated baseband power spectral densities after ideal Gaussian, BT_b = 0.5 Filter (dotted) and FQPSK-1 baseband filter (solid).

results of the baseband power spectral densities of the Gaussian, BT_b = 0.5 and the FQPSK-1 baseband filtered signals. Despite some small differences in the baseband power spectral density the power spectral densities after frequency modulation become practically the same as shown in Figure 8.

The RF power spectral densities of Gaussian, BT_b = 0.5 filtered FSK (GFSK) and FQPSK-1 baseband filtered FSK (FGFSK) in Figure 8(a) were simulated for modulation indexes of 0.32, 0.35, 0.38. Figure 8(b) shows a picture of the RF power spectral densities for a modulation index 0.32 at 915 MHz carrier frequency.

Measurements were made at a normalized scaled down bit rate of 40 kb/s (measurements were also done at 1 Mb/s).

Figure 8 demonstrates that both conventional and Gaussian filtered GFSK and the compatible simpler FQFSK-1 filtered GFSK modulated signals meet the –20dB spectral requirement of the Frequency Hopped Spread Spectrum

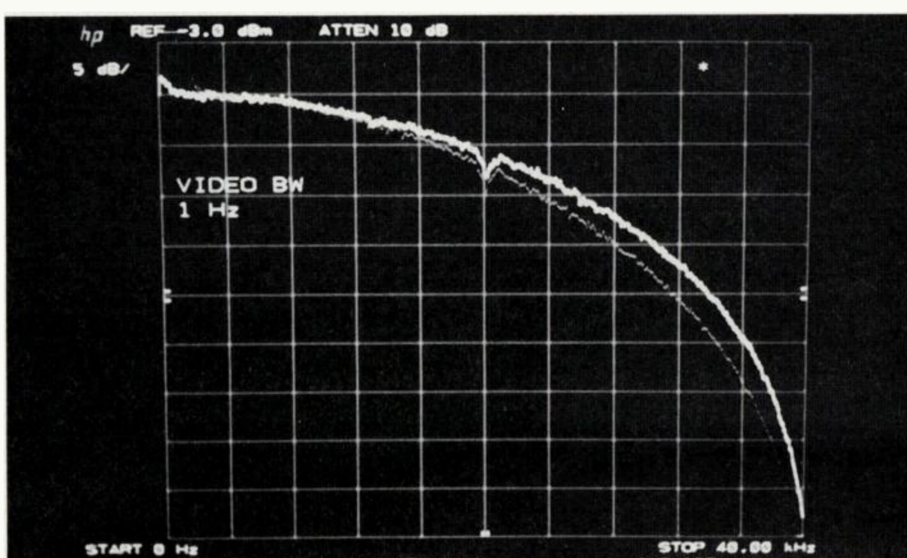

Figure 7(b). Photograph of baseband power spectral density after Gaussian, BT_b = 0.5 filter (thin curve, LMX2411; 9 samples/bit; 8 bits resolution) and FQPSK-1 baseband filter (light curve, XILINX FPGA; 8 samples/bit; 8 bits resolution).

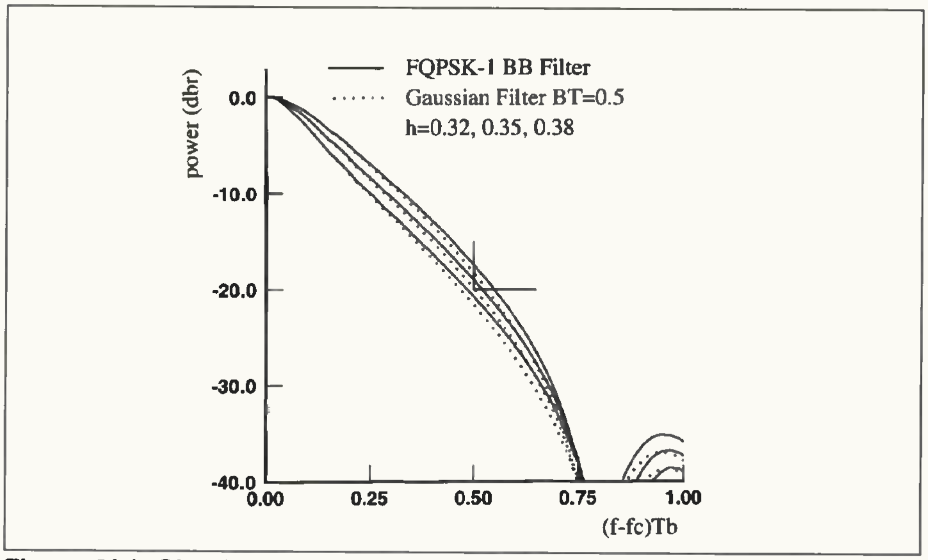

Figure 8(a). Simulated RF power spectral densities with ideal Gaussian, BT_b = 0.5 filter: GFSK (dotted) and FQPSK-1 baseband filter: FGFSK (line) for modulation index h = 0.32, 0.35, 0.38.

IEEE 802.11 Wireless LAN standard for modulation index 0.32 (see $(f-f_c)T_b$ = 0.5 in simulation and 2 divisions from the center frequency in picture). The integrated power within the transmission band ±0.5 $(f-f_c)T_b$ was calculated to be 0.9949 times the total power of the proposed GFSK with modulation index 0.32 and is greater than the specified 99 percent value (corresponding to –20 dBr).

ROM-based filter implementations are more cost and power efficient than the conventional IIR and FIR "transversal" DSP structures and lead to significant reduction in gate count because they do not require multipliers and are based solely on ROM-driven waveform synthesis. Our survey of some leading DSP and filter IC products indicates the trend for higher speed systems is towards the implementation of nonlinearly switched waveform-synthesized filters which are implemented by ROM technology, disclosed in [3] and [4]. *RF*

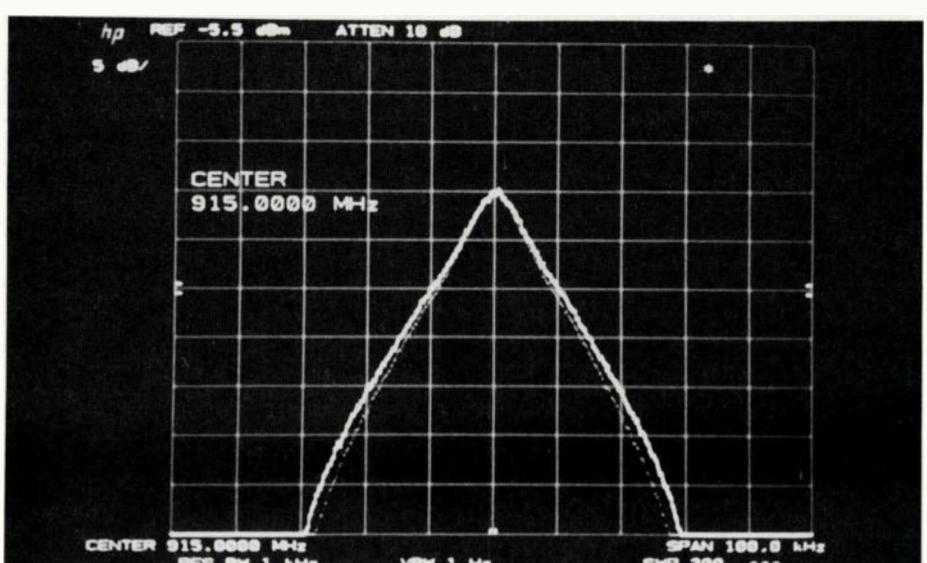

Figure 8(b). Experimental RF (915 MHz) power spectral densities with Gaussian, BTb = 0.5 Filter: GFSK (thin curve, LMX2411; 9 samples/bit; 8 bits resolution) and FQPSK-1 baseband filter: FGFSK (light curve, XILINX FPGA; 8 samples/bit; 8 bits resolution) with h = 0.32.

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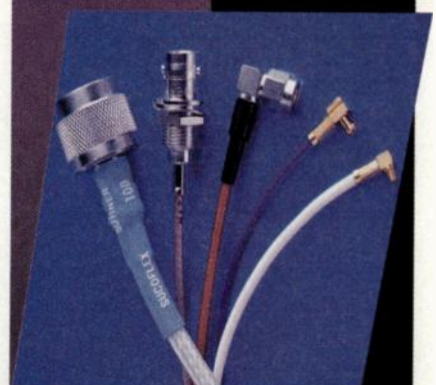

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Notes

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The authors are all affiliated with the Digital Communications Research Laboratory, Department of Electrical and Computer Engineeringat the University of California at Davis, Davis CA 95616; Tel: (916) 752-0583; Fax: (916) 752-8428. Project Director, Dr. Kamilo Feher is Professor of ECE and UC Davis, which has one of the world's most productive wireless research laboratories. Dr. Feher's also operates the company Digcom, Inc., a consulting and development firm.

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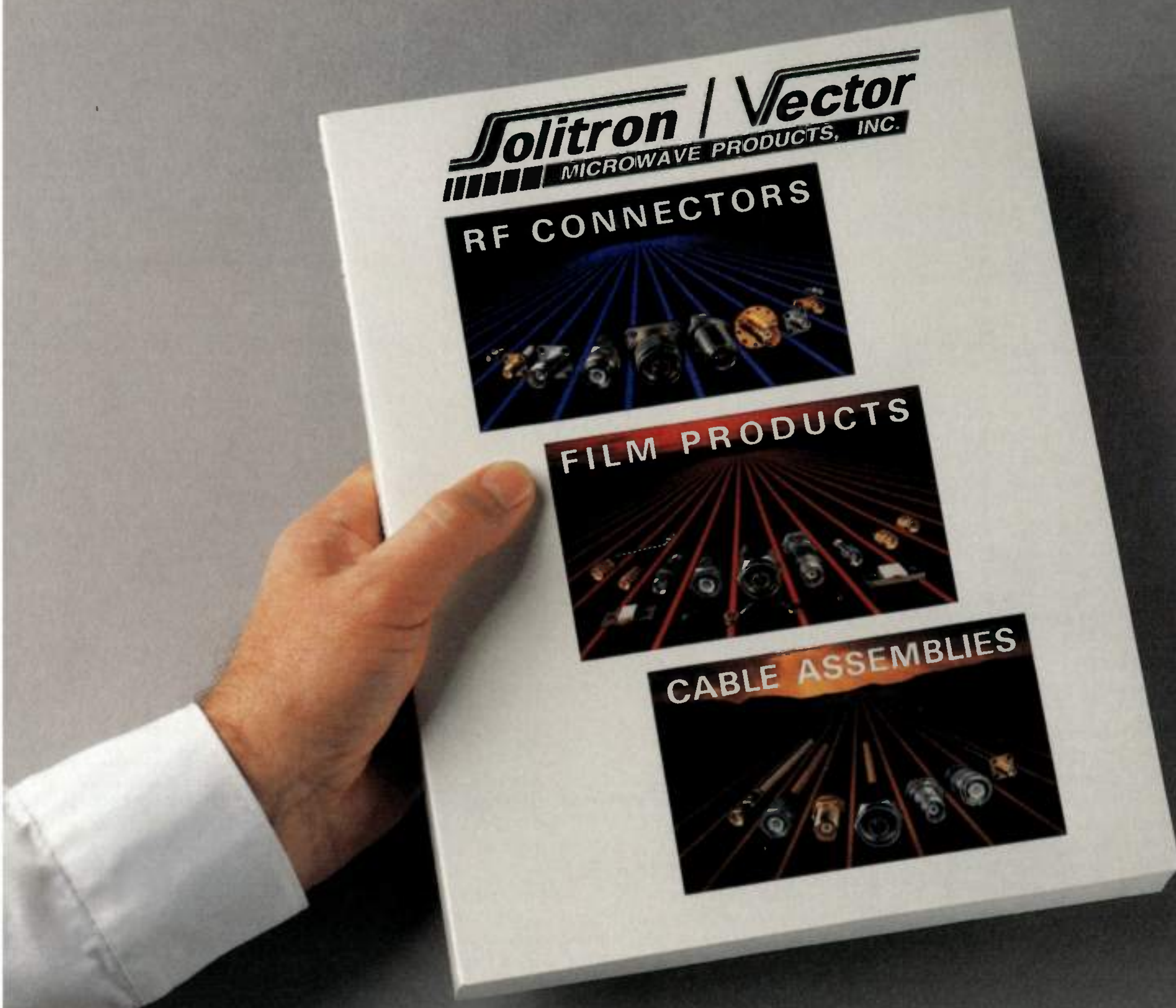

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RF Transformers Part 1: The Windings

By Nic Hamilton
RAF Signals Engineering Establishment

This article is about the use of wide-band ferrite loaded transmission line transformers for RF. For brevity's sake, these will be referred to as transformers.

There are many sorts of transformers. It is a measure of the confusion surrounding these useful devices that no standard terminology has yet evolved. Dr. Sevick wrote an excellent book on the subject [1]. It is widely used by amateurs and professionals, so his terminology will be used here. In addition, the 'conventional transformer', where coupling between the windings is solely in the form of magnetic flux, will be referred to as the Faraday transformer.

Introduction

The standard text on using ferrite in components by Snelling [2], contains data that cannot be found elsewhere. It dismisses the (ferrite loaded transmission line) transformer in a single paragraph, stating that optimizing the design at the lower transmission frequency can be even more effective than using the transmission line technique.

In contrast, Sevick has written a 150 page book called *Transmission Line Transformers*. It has an extensive bibliography. This disagreement will be addressed, and the conclusion will be drawn that both are correct.

Part 1 of this article describes the Faraday and Guanella transformers. It then provides a very swift review of some of Sevick's work, and explains some transformers not discussed in his book. Sevick's book concentrates on transformers for antenna matching but does not discuss many of the possible transformer configurations. As a result, the emphasis of this article is on low power transformers.

Part 2 considers the selection of transformer cores, and presents some mathematics required to predict a transformer's performance.

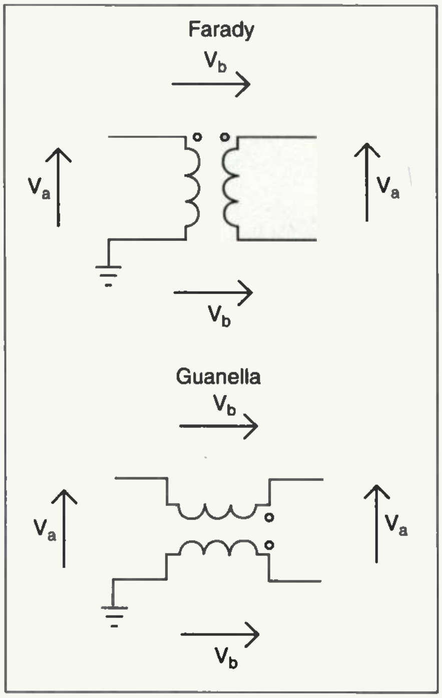

Figure 1. Comparison of transformer types.

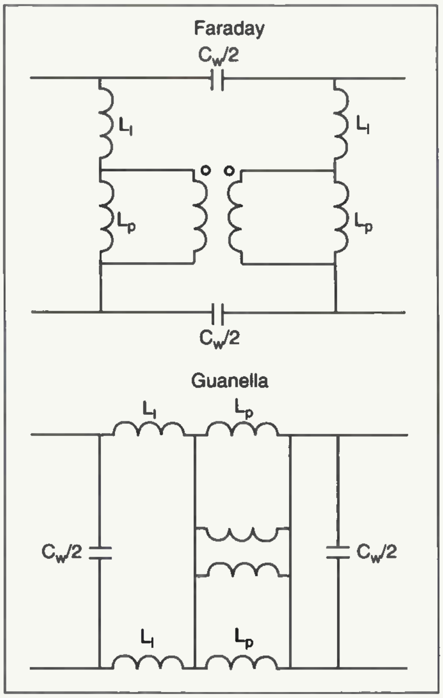

Figure 2. Equivalent circuits of the transformers.

Faraday and Guanella

The standard 1:1 Faraday transformer is shown in Figure 1. Current through the primary winding causes a magnetic flux in the core, which causes an emf to be developed at the secondary. In this way, the voltage V_a is conveyed from the primary winding to the secondary. The function of the 1:1 transformer is to isolate the primary from the secondary. The potential difference across the isolation is V_b.

As the working frequency of the transformer is increased, the flux linkage between the input and output coils becomes progressively poorer. This manifests itself as a leakage inductance. In order to reduce the leakage inductance, the primary and secondary windings have to be physically close to each other. The leakage inductance is reduced to a minimum when bi-filar wire is used. However, the capacitance of this wire is about 100 pF/m, and this appears as inter-winding capacitance. As the frequency is increased, the reactance of the inter-winding capacitance becomes very small, and the voltage V_b, which was presented with an infinite impedance at DC, is shorted out at RF.

To remedy this problem, the transformer connections can be rotated through 90° to give the Guanella transformer shown in Figure 1. The isolation between the input and output is now provided by the inductance of the windings.

Figure 2 shows simplified equivalent circuits for the two transformers. In the Faraday transformer, the high frequency performance is limited by two factors: the low reactance of the inter-winding capacitance, C_w, and also by

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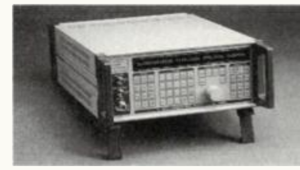

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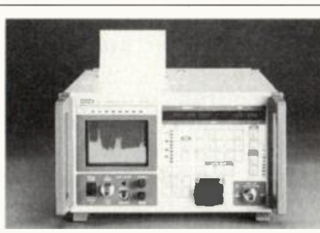

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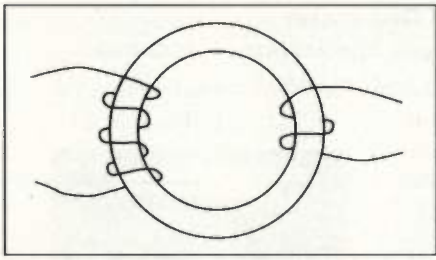

Figure 3. The Faraday transformer consisting of a few wire turns on a ferrite ring. Don't try it at RF: it won't work.

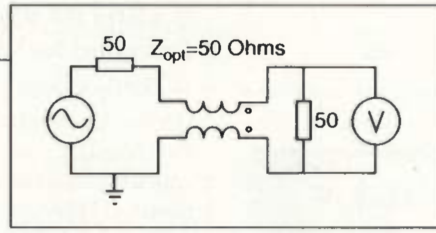

Figure 4. The Guanella prototypes.

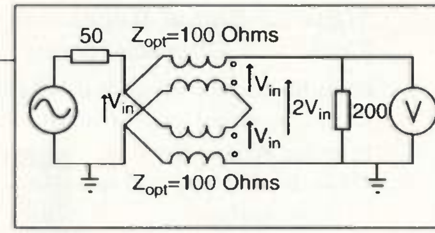

Figure 5. The Guanella 1:4 transformer.

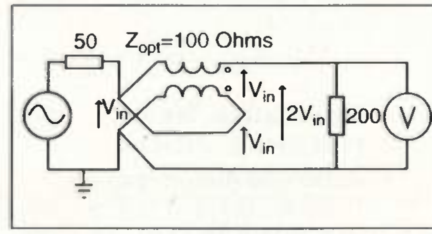

Figure 6. The Ruthroff 1:4 transformer.

the high series reactance of the leakage inductance L_I. The winding inductance is represented by a parallel component L_p.

The Guanella transformer in Figure 2 has no such high frequency limit, because the path from the input to the output is now a transmission line composed of the distributed C_w and L_p with a small contribution due to L_I. The perfect transformer at the heart of the circuit could be removed without changing the transformer's response.

In deciding whether a Faraday or a Guanella transformer will give the lowest loss, consideration must be given to the ratio V_a:V_b and the ratio of the impedances presented to these two voltages.

In the examples considered so far, the Faraday and Guanella transformers are quite distinct. However, most transformers are capable of working in both modes at the same time. These will give the widest bandwidths of all.

In an attempt to reduce the interwinding capacitance, some textbooks that discuss RF transformers show the Faraday transformer shown in Figure 3. As a broadband RF transformer for impedances of 50Ω or more it is useless: the reasons for this will be discussed in the second part of this article. This type of Faraday transformer does have a use as a narrow-band transformer with a coupling coefficient less than 1. An example of this application is the inductive coupling of a tuned circuit. For this, use a dust iron core to give a higher Q.

Sevick's Work

Figure 4 shows the prototype of most transformers. The windings form a transmission line which is generally of either coaxial cable or enameled copper wires twisted together. Any other low pass type of transmission line can be used. The RF signal travels from the input to the output between the transmission line's conductors; the fact that the line is coiled on the core makes no difference. However, each winding, when measured from end to end, will have such a large RF impedance as to approximate to an open circuit. So the input is connected to the output at DC, but not at RF.

It is obvious that the voltage across the output of the transformer in Figure 4 is the same as the voltage across its input. This does not seem very useful, until it is realized that this is still true even if one of the transformer's output wires is connected to any arbitrary RF voltage. Note that the output voltage is phase shifted with respect to the input voltage due to the delay of the transmission line.

Figure 5 shows the Guanella 4:1 transformer. In 1944, Guanella published a paper on this transformer. Two of the transformers from Figure 4 are used; the transformer inputs are wired in parallel, and their outputs in series. It is easy to see that the output voltage must be twice the input voltage. If the voltage is doubled on the output, the current must be halved, so the impedance must have gone up by a factor of four. In this case 50Ω has been transformed to 200Ω. The upper frequency limit of this transformer is theoretically infinite, because the transmission lines are matched both in phase delay and in impedance. (For this reason, the Guanella transformer has also been called an equal delay transformer.) The two 100Ω lines in parallel give a 50Ω input impedance, and the two 100Ω lines in series give a 200Ω output impedance. Note that the two bi-filar windings must be on different cores to avoid a shorted turn. This is a general characteristic of Guanella unbalanced to unbalanced (unun) transformers.

Figure 6 shows the Ruthroff transformer (in 1959, Ruthroff published a paper which included this transformer). One of the transmission lines has been replaced by a straight DC connection. The low frequency response of the Ruthroff transformer is similar to the Guanella transformer. However, the two paths between the input and output now have different delays, and, at the point where the transmission line is λ/2 long, the voltages are in anti-phase, and the output will be zero.

Figure 7 shows the development of the Ruthroff transformer. A little thought will show that Figures 7a and 7b are electrically identical. Figure 7a emphasizes the transmission line qualities of the transformer. Figure 7b emphasizes the fact that at LF, the Ruthroff transformer is capable of operating in the 'conventional' Faraday mode. Notice that if the transformer is wound as in Figure 7c, the length of the transmission line is halved, however, the same length of wire is used, so the frequency limits of this transformer will be the same as those of Figure 7b.

The next logical step from Figure 7c is to achieve different impedance transformation ratios by selecting different tapping points on the Ruthroff transformer. Sevick also shows that tapping points can be made part of the way along a winding. This gives a pseudo-transmission-line transformer with a greater number of available turns ratios.

These transformers can be quite fiddly to make, and the problem gets rapidly worse as the number of windings is increased. 6-filar transformers represent the limit of my personal patience. (For additional information on impedance transformers and hybrid combiners in general, see references [3,4,5].)

Conventional Transformer?

The winding of Figure 7b has 8 bifilar turns. These were split in half to obtain the 4-turn, 4-filar winding of Figure 7c. This winding may, by a similar process, be split into two, and then into two again. This gives a 1 turn 16-filar transmission line. The length of the line is now an eighth of the origi-

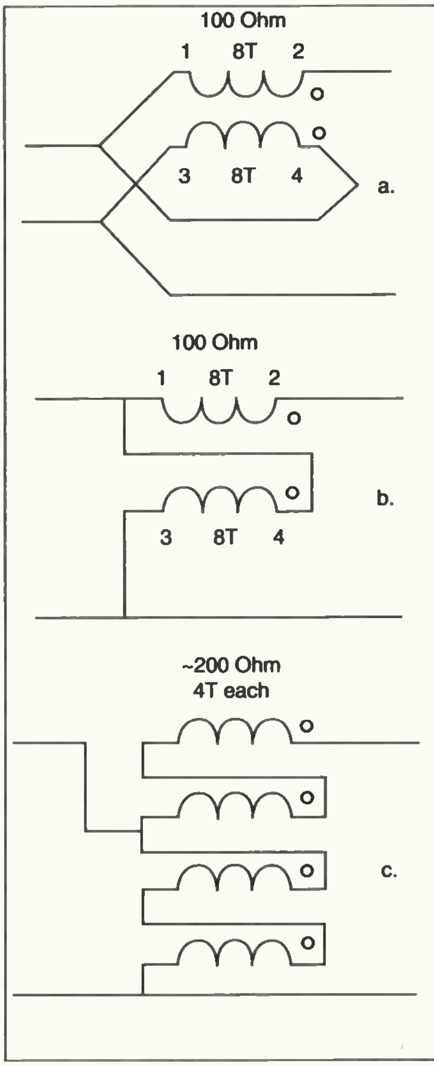

Figure 7. The development of the Ruthroff 1:4 transformer.

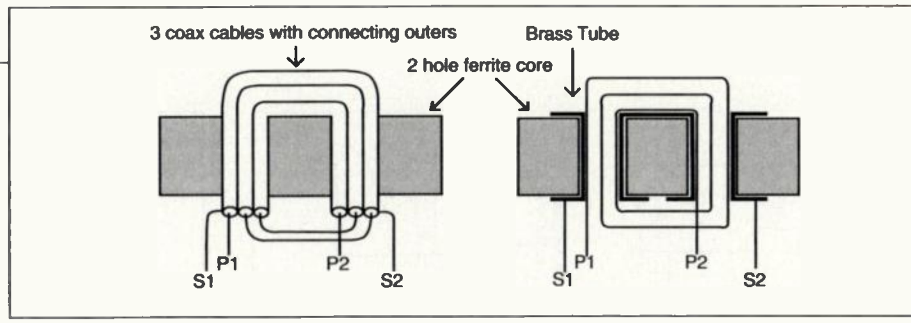

Figure 8. Evolution of the "Brass Tube" transformer.

nal length. Obviously, a 16-filar transformer would be very difficult to wind and connect correctly.

However, if this 16-filar transformer is wound using a twin-hole core, the resulting transformer is identical to a "conventional" autotransformer wound on the same core. Stated more clearly: an auto-transformer with a total of N turns wound on a twin-hole core can be viewed as an N-filar transmission-line transformer with a transmission-line length of 1 turn.

This resolves the main difference between Snelling's approach and Sevick's. There is a residual difference: Snelling views the inter-turn capacitance as always harmful, whereas Sevick views it as part of the transformer's N-filar transmission line impedance. For RF transformers, I view leakage inductance as a crude LF approximation to the changing equivalent impedance of a combination of

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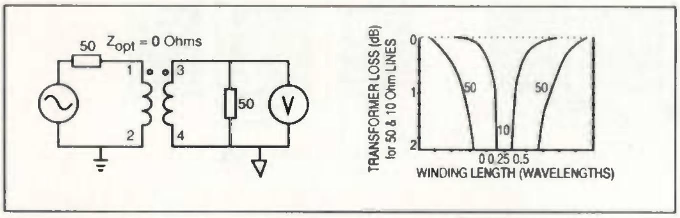

Figure 9. The 1:1 DC isolating transformer.

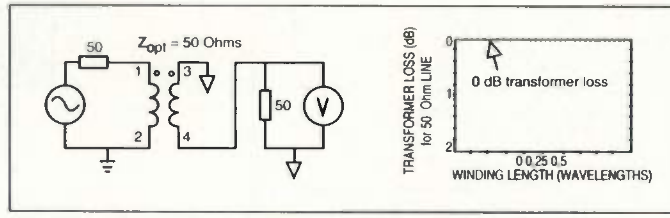

Figure 10. The 1:–1 DC isolating transformer.

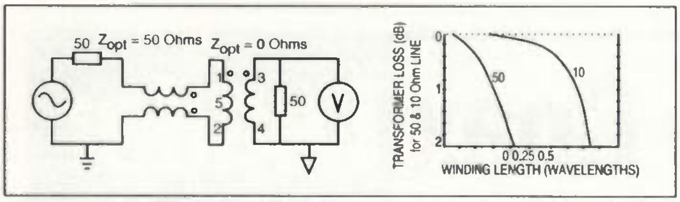

Figure 11. The 1:1 DC isolating transformer with balun feed.

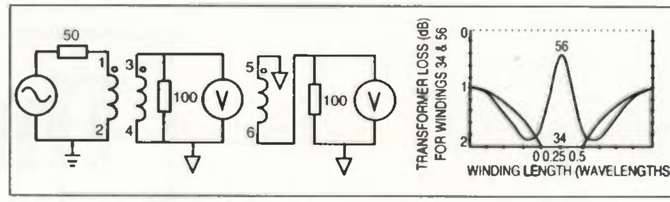

Figure 12. The 1:1:–1 ring mixer transformer.

transmission line stubs formed by the transformer's turns.

Brass Tube Transformer

This transformer is usually wound on a twin hole core, or on two single hole cores side by side. A metal tube is passed through each hole. These are connected at one end to form a single turn primary. To achieve sufficient winding inductance using just one turn, designers use MnZn ferrites because they give a very high inductance per turn. The secondary has many turns.

This sort of transformer is widely used for matching 50Ω to the very low impedances found in high power transistor push-pull amplifiers for RF. It provides DC isolation, and the transformer's balance is good.

Having established that the autotransformer can be analyzed in transmission-line terms, it will come as no surprise that the brass tube transformer can be analyzed in the same way [6]. The evolution of the brass tube transformer from a Ruthroff type is shown in Figure 8.

This type of transformer is used because the solid metal tube provides a low winding loss for the primary, and screens the high impedance (= high voltage = high loss) windings from the ferrite. This ensures that dielectric losses in the core are kept to a minimum and enhances the transmission-line properties of the transformer a little.

However, the brass tube transformer's loss is only a little lower than the same core with a conventional wire primary.

The one turn primary of the brass tube may be extended to any number of turns by connecting extra thick wire looped through the tubes. This will give any turns ratio required. These are best calculated with a sliderule [5].

DC Isolating Prototypes

DC isolation is important for diode ring mixers, for the input and output matching of transistors, and for certain hybrid transformers. Figures 8-12 show circuits with various earth connections. These represent RF earths, which do not preclude a DC potential existing between the transformer's input and output. To emphasize this, the input and output sides of the transformers have different symbols for the earth connections.

Each figure is in two parts, showing first the transformer connections, and then a computer simulated 'frequency response'. Instead of the X axis being calibrated in terms of the frequency, the electrical length of the transmission line in the transformer is used. The measured frequency responses correspond well with the simulations, however the transformer's low frequency response is not at issue here, and is not modeled. The circuit simulator program used was Berkeley SPICE 2G.6. Beware! For Ruthroff transformers, this simulator gives wrong results if the simple transmission line model (T) is used. I do not know if other SPICE derivatives suffer from the same fault.

Figure 9 shows the 1:1 transformer. In this circuit, points 2 and 4 are both at RF ground, so the impedance across the transformer is due to the impedance of a short circuit as modified by the transmission line length. Where the winding is $\lambda/4$ long, the 'short circuit' between 2 and 4 is transformed to an open circuit between 1 and 3, so the transformer attenuation reaches a maximum. Use of a low winding Z_0 increases the transmission line capacitance, and so improves the high frequency response. This transformer has zero phase shift at all frequencies.

Figure 10 shows an inverting Guanella transformer derived from Figure 4 which provides DC isolation. Because the signal is applied between 1 and 3, and travels to 2 and 4, the upper frequency response is theoretically infinite. The phase shift of this transformer increases with frequency.

Figure 11 shows the transformer of Figure 9 with a Guanella balun on the input. At low frequencies, point 2 is close to earth potential, and the whole transformer works in the same way as Figure 9: V_{24} (the small potential between points 2 and 4) equals V_{13}. However, when the DC isolating transmission line is $\lambda/2$ long, $V_{13}=-V_{24}=V_{42}$. At the same time, point five is at earth potential. So $V_{25}=V_{24}=V_{31}=V_{51}$. Summing these voltages round the loop (Kirchoff's Law) gives the resultant output $V_{43}=0$. Thinking about this circuit is a good way of burning mental CPU time.

Note that the Guanella balun shown gives only an approximately balanced output, due to the winding inductance and resistance. For true balance, an inductor must be connected from point 1 to ground. The inductor should have the same type of core and num-

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In a move to increase its presence in Europe, Richardson recently announced the addition of five new lines now available to its European customers. The Company has added Watkins-Johnson and RF Prime to its distribution coverage and expanded its distribution agreements with W.L. Gore, Amphenol and Ericsson to include Europe.

Richardson's European operations is continuing to add RF and microwave sales specialists and expanding sales offices to provide better coverage and increased technical support. Currently, Richardson has offices in Lincoln, England; Paris, Munich, Rome, Florence, Milan, Madrid, Barcelona. The Company will be expanding operations into London, Amsterdam, Hamburg and adding an office in Scandinavia.

While the Company's presence has been felt in North America, the addition of these lines in Europe provide further proof of Richardson's dedication to be THE choice for each type of RF, microwave and power semiconductor and related component it carries.

"Richardson Electronics has made a commitment to the RF marketplace on a global basis," said Peter Saxby, Product Marketing Manager for Richardson's Solid State and Components Strategic Business Unit and former technical sales person for M/A-COM and BFI. "Due to a number of customer-focused programs, we have built ourselves into one of the largest RF distribution specialists in the world."

Growth

Richardson's growing success can be attributed to a few key areas.

Inventory

Currently an $10 million RF semiconductor inventory housed in 14 worldwide stocking locations. In most cases, Richardson can ship the RF, microwave or power semiconductors its customers need the same day they call. This extensive inventory also enables the Company to handle a customer's midstream production increases.

Technical Sales Support

The Company has brought together sales and marketing people from leading manufacturers in the RF and microwave industry. Richardson is capable of assisting OEMs with new designs as easily as suggesting the proper replacement to service dealers and end users. Value-added services: Stocking programs, electronic data interchange, special testing, selecting and matching, bar coding. Richardson routinely provides the extra services its customers demand.

"There are many "broad-line" distributors; however, the markets addressed by Richardson are unique and (they) require specialized service," said Joel Levine, Vice President and Strategic Business Unit Manager for Richardson's Solid State and Components group. "Concentrating on our specific niche, Richardson Electronics has reached an unparalleled position in the electron tube and power semiconductor industry."

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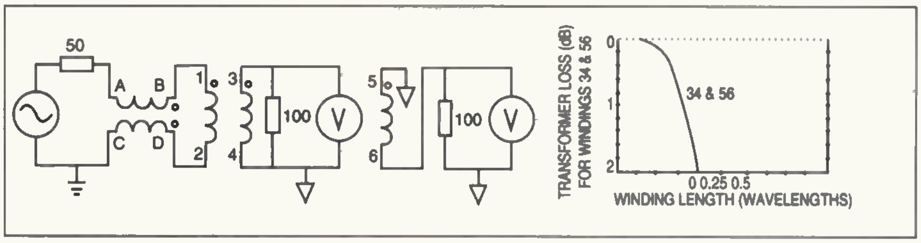

Figure 13. The 1:1:–1 ring mixer transformer with balun.

INFO/CARD 41

ber of turns as the Guanella balun. Winding inductance and resistance will be discussed in the second part of this article.

Mixer Transformers

These must provide DC isolation, low insertion loss and also good balance between the two output voltages. These are also desirable attributes for the transformers used in RF bridges.

The transformer in Figure 12 is typical of the type used in diode ring mixers. The transmission line consists of a tri-filar winding with the input winding 12 coupling to windings 34 and 56. At first sight, V_{34} will be equal to V_{56}. This is not the case. It can be seen that the transformer 1234 is similar to Figure 9, which has zero phase shift and varying attenuation. The transformer 1256 is similar to Figure 10 which has zero attenuation and varying phase shift. This approach is only approximate because of the existence of the third transmission line 3456. This has the effect of balancing the responses a little.

The imbalance in the high frequency response inherent in this transformer design means that the transmission lines must be very short. Take a typical 0.5 to 500 MHz Schottky diode ring mixer. Its transmission lines have a maximum electrical length of 1 inch ($\lambda/20$). To achieve this, the transformer consists of 4 turns on a ferrite ring core with a diameter of 0.1 inches and an AL of 4000 nH/turn^2. The small number of turns will give rise to a high insertion loss.

Figure 13 shows a dual output version of the transformer in Figure 11. This gives perfect balance between the two outputs, and an attenuation maximum at $\lambda/2$. This circuit [7] is widely used to improve high frequency mixer balance where the transformer has electrically long windings. This will allow a greater number of turns which will reduce the insertion loss.

DC Isolating Impedance Transformers

Figure 14 shows a non-inverting, DC isolating Guanella transformer [8]. This splits V_{in} into two equal parts V_a and V_b. These two voltages can be traced through the transformer to the output, where they are re-combined in phase to give $V_{out} = V_{in}$, but with DC isolation. This transformer can also be viewed as two of the inverting transformers in Figure 10 connected with

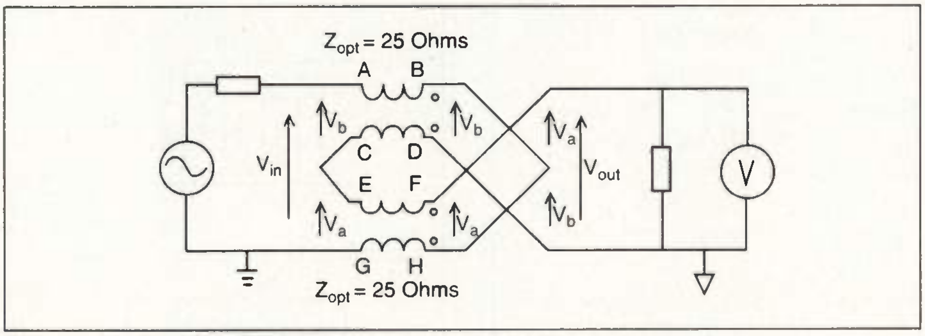

Figure 14. The DC isolating 1:1 Guanella transformer.

their outputs crossed in order to give zero phase shift at LF.

This transformer can be used as a building block: multiple transformers of this sort can have their inputs wired in series and the outputs in parallel in order to give 1:4 1:9 1:16 etc. impedance transformation ratios. The standard Guanella transformer gives the same ratios, but without DC isolation.

Summary of Unun Transformers

Figure 15 shows the family progression of the Ruthroff and Guanella types. For the Ruthroff 9:1 tri-filar transformer, SPICE modeling was used to estimate the theoretical optimum line impedance. The impedance shown is that between each of the three wires. So the model of the tri-filar transformer consists of 3 transmission lines of equal impedance: Z_{12}, Z_{13} and Z_{23}. This result has not been verified practically and should, like all models, be treated with caution. What it does show, however, is that the required impedances are generally quite high, and that a reasonable approximation to this high impedance could be achieved by winding an auto transformer on a twin hole ferrite core. (These are also known as "Balun Cores" or, rather charmingly, as "Ferrite Binoculars" in the English edition of DUBUS info magazine.)

Fractional Ratio Transformers

The final (and most complex) transformers of this sort are shown in Figure 16. These examples are intended to provide an approximate 50 to 75Ω match. The Ruthroff transformer is comparatively simple (even if its SPICE model has 10 transmission lines).

The Guanella transformer was developed according to the principles of reference [9], which shows that the Guanella transformer is not limited to the 1:1 1:4 1:9 etc. ratios, as many have stated. However, this transformer is complex, as it consists of five sub-transformers.

Assume that an RF signal of 4 V pd is applied to the 50Ω input. The four lower sub-transformers are connected in series on the 50Ω side, so each has an input pd of 1 V. On the 75Ω side these are connected in parallel, so the output pd of this part of the transformer is 1 V. The other sub-transformer is a Guanella prototype, as in Figure 4. Its output is connected to the 1 V pd of the other sub-transformers. The final output on the 75Ω side is thus 5 V pd. This, by a similar argument to that used for Figure 5, implies that, for a 50Ω source, the output impedance is $(5/4)^2 \times 50 = 78.125\Omega$.

HF Limits of Guanella Transformers

The HF limit of the Guanella transformer is usually due to an imperfection of the windings for example:

1) The HF loss of the line in each sub-transformer;

2) Impedance inaccuracy of the sub-transformers;

3) Group delay mismatch of the different sub-transformers;

4) Start to finish capacitance of the windings (this will be discussed in part 2);

5) Impedance mismatch due to the connections to the lines (especially at low Z_0).

Notice that 1 and 5 work in opposite directions. In order to keep the connection mismatch small, the diameter of the line must be small, and this implies high line loss.

HF Limits of N-Filar Ruthroff Transformers

Measurements and SPICE models show that N-filar Ruthroff transformers have peaks in their high frequency insertion loss. These are caused by resonances of the length of the winding. Assuming that the line impedance is correct, it is the lowest frequency resonance which controls the upper frequency limit of the transformer. This resonance occurs where the total length of wire is between .5 and 1 λ. For a given wire length, transformers with high turns ratios have lower frequency resonances. N-filar Ruthroff transformers where N=2,3,4 have repeatable high frequency characteristics. SPICE models of these transformers are useful. As N increases, the impedance between each winding and its fellows becomes more indeterminate, so the high frequency performance becomes more variable; the variability might be reduced by using a ribbon cable or a finely platted line. I have not tried this.

HF Limits of Auto-Transformers

Auto-transformers also have peaks in their high frequency insertion loss. Resonance occurs where the total length of wire is between 0.25 and 0.5 λ. Again, lower frequency resonances tend to occur with high turns ratios. However, identical looking transformers can have quite different high frequency characteristics, depending on how they are wound. This is because the impedances between the turns are much less well defined. It is particularly important to keep apart turns near the start and finish of the winding.

The velocity factor of the transmission lines in auto- and Ruthroff transformers can be as low as 0.5. This, combined with the high frequency resonances provides a practical basis for the old adage that the total length of wire used in a transformer should be no longer than an eighth (some authors say a twentieth) of a wavelength at the highest frequency of operation.

A Choice of Transformers

Guanella transformers can be much more complicated to build than their Ruthroff equivalents, as Figure 16 shows. The only justification for using the Guanella type is that its upper frequency is not limited by the transmission-line length. However, it can be difficult to find the transmission lines of the correct impedance that are needed in order to take full advantage of this. Sevick [1] gives some useful tips on this subject. It is worthwhile experimenting. For instance, pulling out the dielectric and centre conductor from MCX 50Ω sub-miniature coaxial

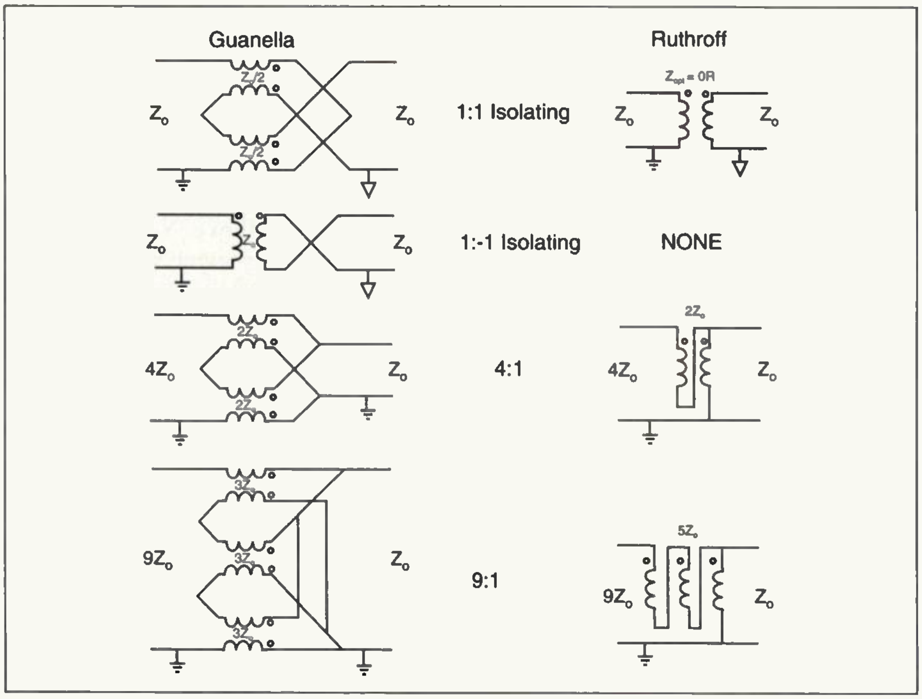

Figure 15. Unun transformers.

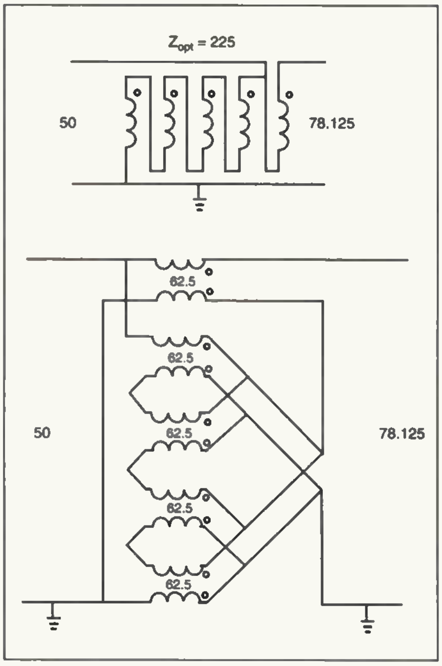

Figure 16. 50 to 78.125 Ohm transformers.

cable, and replacing it with 0.5 mm enameled copper wire gives a cable with a characteristic impedance close to 11Ω.

In high power transformers, the sheer bulk of ferrite required implies large line lengths. In a Ruthroff transformer, this gives low bandwidth, so this is where the Guanella transformer should be used.

Conclusion

From Figure 4 onwards, the transformers discussed in this article are all transmission-line types. In fact, it is quite difficult to design the transmission-line mode out of an RF transformer. In particular, the twin-hole core, so widely used at RF, forces the designer to wind a transmission-line transformer.

It is my contention that, if it's intended to be a broadband RF transformer, and consideration has not been given to its transmission line properties, then it has not been designed properly. In some cases it may be hard to achieve the correct line impedance. In some cases, it has not been fully established what the correct line impedance should be.

To say that the high frequency roll-off of an RF transformer is due to leakage inductance is misleading, as most of the energy is coupled through the transmission lines formed by the turns, and very little through the core. As a result, flux leakage from the core is unimportant. However, it is sometimes useful to model the high frequency roll-off of a transformer as being due to a series inductance. Has anyone a more appropriate name for it?

The high frequency performance of a transformer depends on its windings, and using a circuit simulator like SPICE can help with the design process. The second part of this article will discuss the core, which is largely responsible for determining the low and mid-band performance of the transformer. *RF*

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[8] C. Norman Winningstad, *Nanosecond Pulse Transformers*, IRE Trans. on Nucl. Sci., Vol. NS-6, No. 1, March 1959, pp. 26-31.

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About the Author

Nic Hamilton is the section leader of the EMC and RF RADHAZ laboratory at the Royal Air Force Signals Engineering Establishment (RAFSEE). Before joining RAFSEE, he worked at the Royal Signals and Radar Establishment where he worked on HF radio research. He can be reached at The Officers' Mess, RAFSEE, RAF Henlow, Beds., SG16 6DN, UK.

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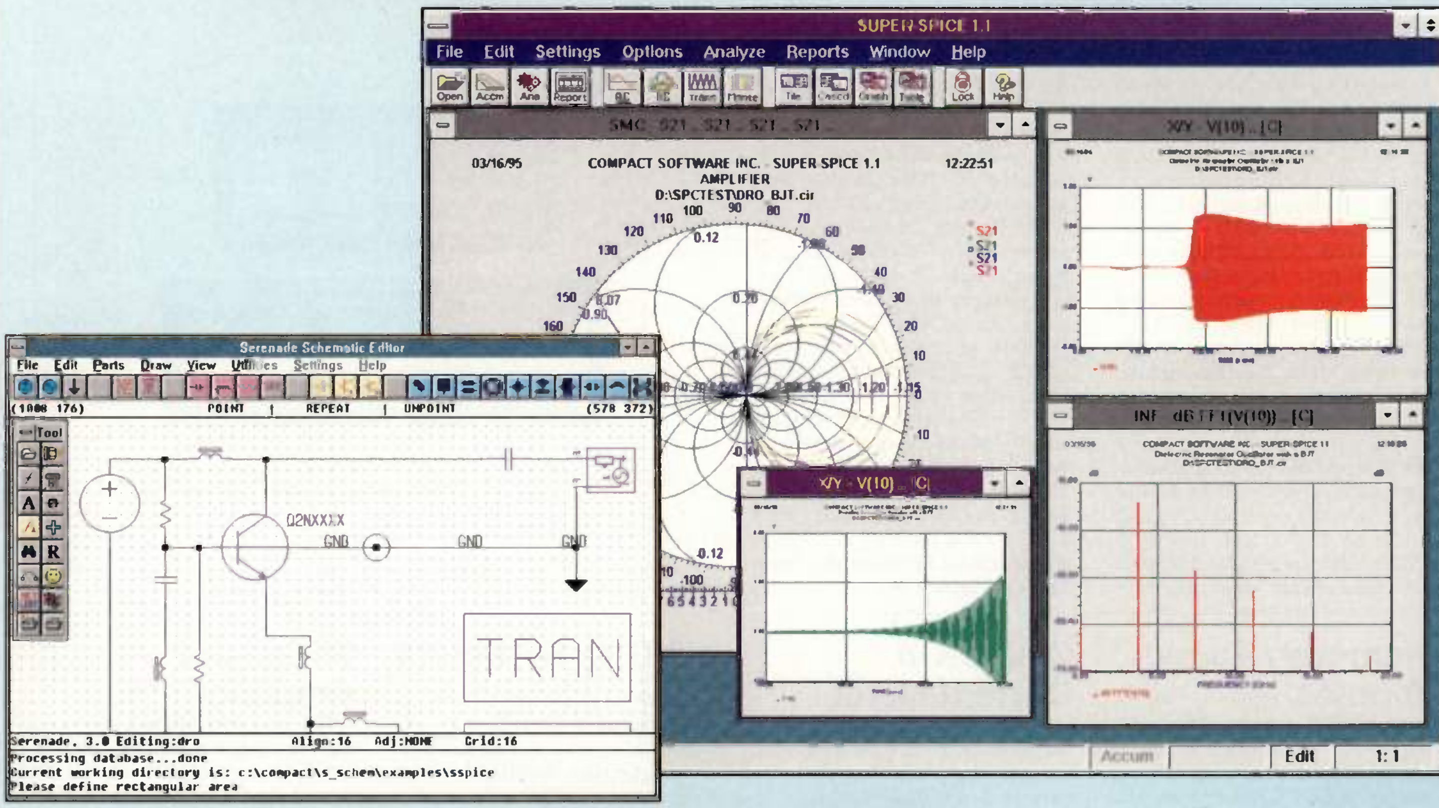

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RF Tranceiver ICs

A family of five RF ICs for wireless applications has been released by Telefunken Semiconductors. The small-outline, surface-mount products include three ICs for the transmit path and two ICs for the receive path. Packaged in a 16-pin SOIC, the U2790B is a direct conversion I/Q modulator that operates from 100 to 1000 MHz. Its companion direct conversion demodulator, the U2791B, offers a frequency range of 100 to 1000 MHz. Neither device requires external components for phase shifting, and both feature a patented on-board phase splitter and control loop, providing phase error less than 1° and an amplitude error of less than ±0.25 dB. For outputs in the 30 - 300 MHz range, the U2793B I/Q modulator provides the same level of accuracy. Both the U2793B and U2791B are packaged in the 20-pin SSO. The U2795B is a double-balanced mixer that can be used as an up- or down-converter from 50 MHz to 2.5 GHz. Inputs and outputs are single-ended. The U2796B is a single balanced mixer with a frequency range of 0 to 2.5 GHz. A duty cycle control loop generates balanced signals. Both the U2795 and U2796B are packaged in 8-pin SOICs. Samples and production quantities of all five new RF ICs are available now. Pricing in 10,000-piece quantities is \$5.75 for the U2790B and U2791B; \$5.40 for the U2793B; \$2.50 for the U2795B; and \$2.25 for the U2796B.
Telefunken Semiconductors
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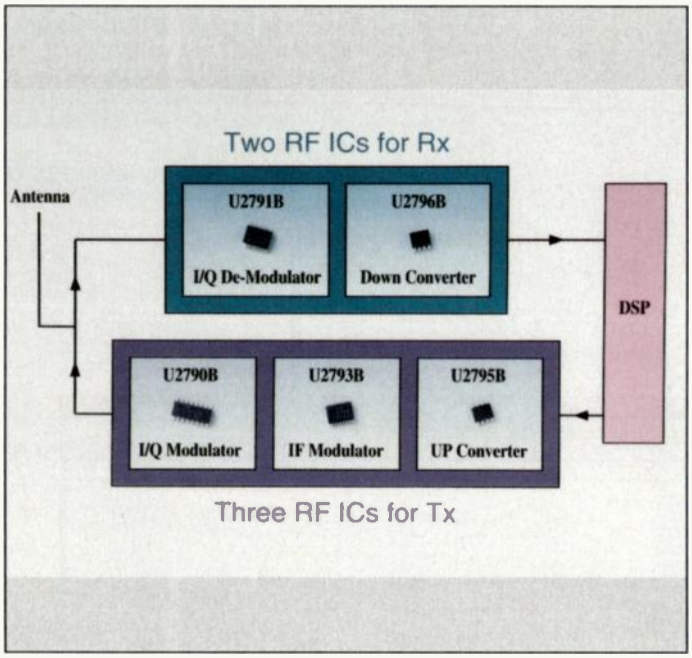
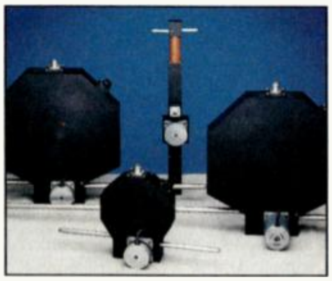


Remotely Tuned Dipoles

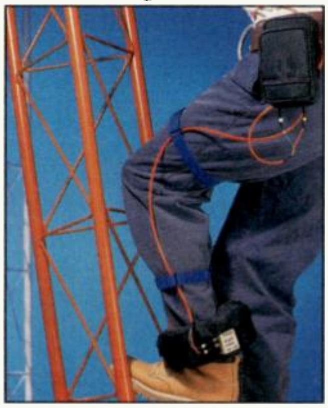
Electro-Metrics has announced the introduction of their new remotely tuned dipole set. The EM-6929 covers the frequency range of 30 MHz to 1 GHz. Eliminating the requirement for manually adjusting the element length at each frequency, the EM-6929 utilizes



precise remotely controlled element length adjustment. The EM-6929 remotely tunable dipole set consists of four antennas and a single control module. The remote control unit is digitally controlled via the IEEE-488 interface. The EM-6929-1 covers 30-65 MHz; the 6929-2, 65-180 MHz; the 6929-3, 180-400 MHz; and the 6929-4, 400-1000 MHz. Typical VSWR is less than 1.6:1, and maximum balun loss is 0.5 dB. Maximum continuous power handled by the set is 20 W.
Electro-Metrics, Inc.
INFO/CARD #183

Clamp-On Current Meter

Holaday Industries introduces the Model HI-3702 clamp-on RF induced current meter. To measure induced body currents in accordance with ANSI/IEEE C95.1, the HI-3702 uses current transformer sensing to measure more directly body currents flowing through arms or legs. With the light weight HI-3702 in its comfortable, cusioned case, induced current measurements are now possible even while walking or climbing. Frequency response is from 3 kHz to 110 MHz, covering the ANSI/IEEE



C95.1 induced current requirements and the full FM broadcast band. The meter has 60 dB dynamic range (1 - 1000 mA).
Holaday Industries, Inc.
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1.9 GHz RF Filters

4DBF ceramic dielectric filters designed for U.S. 1.9 GHz personal communications systems are now available from Toko America. These three pole, low insertion loss, high selectivity filters are specifically designed to cover the new two-way PCS band from 1850 to 1910 and 1930 to 1990 MHz. Part number 4DBF-1880L-10 is centered at 1880 MHz and part number 4DBF 1960L-10 is centered at 1960 MHz. Both filters have a bandwidth of 60 MHz,

insertion loss of less than 2.5 dB, and offer greater than 24 dB of rejection at ±110 MHz from center. These filters are included in Toko's standard 4DBF series and are available in surface mount packages with profile less than 5mm and a footprint of 17.5 × 11mm. They are reflow solderable and available on tape and reel.
Toko America, Inc.
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Wideband Amplifier

IFI has introduced a bench top (or rack mountable) 800 W solid state amplifier that is based upon a new modular MOSFET design. The MOSFET

design portion results in reliability and efficiency, while the modular aspect of the design allows for rapid heat dissapation and easy servicing. Covering an instantaneous frequency range of 100 to 500 MHz, this amplifier delivers 500 W linear power at 1 dB compression and 800 W at saturation. Maximum harmonics are –20 dBc and remote electronic gain control has a range of 40 dB or better. Standard features include a backlit LCD screen; self-diagnotic circuitry that provides thermal, power supply, high VSWR and module fault indication; and high load VSWR protection.
Instruments for Industry, Inc.
INFO/CARD #186

SUBSYSTEMS

GPS Antenna Protector

PolyPhaser introduces a specially designed coaxial protector for GPS receivers with active antennas. The IS-MR50LNZ+6 protects in-line amplifiers and GPS receivers while providing protection for both DC and RF paths on coax. Measured throuput energy is 0.4 microjoules at 2 kilovolts (3/250 ns waveform). Providing both positive and reverse voltage protection, the IS-MR50LNZ+6 exhibits low loss as well as low VSWR over a large bandwidth.
PolyPhaser Corp.
INFO/CARD #187

Remote Cell

The cellExtender from Penninsula Wireless Communications is a high-power, microwave-linked remote cell which helps cellular providers extend coverage to low-traffic, light-density areas. Power output is 10 to 15 W per carrier, with up to 16 carriers, resulting in 160 to 240 W of composite power.
Penninsula Wireless Communications
INFO/CARD #188

Rugged Radio Modem

The RFM96W from Pacific Crest is a high speed, compact, water proof wireless modem designed for portable and basestation applications in DGPS, AVL and data telemetry. The modem is available in 2 W, 15 W, and 35 W versions across the UHF frequency band. Intelligent firmware features forward error correction and data scrambling for reliable data communication in transparent, packet switched and digipeating operating modes.
Pacific Crest Corp.
INFO/CARD #189

Wireless Base Station Antenna

Andrew Corporation announces a line of base station antennas for cellular, trunking, ESMR, PCS/PCN and rural telephony applications. The antennas are offered in a variety of patterns, gain and/or other electrical characteristics such as null fill and beamtilt. These antennas offer long-term performance, low intermodulation generation, and excelent system flexibility.
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Fiber-Optic Analog Detector

Lasertron offers an impedance-matched receiver module for use in high-preformance CATV receivers. The QDIV is a low-cost, pigtailed pin photodetector featuring a 75 Ω impedance-matching transformer housed in a heremetically sealed 14-pin DIL package.
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High Gain Antenna

The Digital Dominator™ from the Decibel Products Division of Allen Telecom Group is designed specifically for digital cellular systems. Operating with 13 dB gain at 820 to 900 MHz, the Digital Dominator is designed for time division multiple access systems. The antenna provides 80 MHz bandwidth, sharp patterns and a 34 dB front-to-back ratio. The antenna pattern is 14° vertical and 70° horizontal.
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UHF Video/Telemetry Transmitter

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tivities. The Digi-Field uses a 3.5 digit display and is powered with a 9 V battery. The meter can be used with its own telescoping antenna or an external antenna of the user's choice. Typical calibration curves in dBm and volts/power conversion are available. Price for the model-C Digi-Field is $185.00.
IC Engineering
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Enhanced Synthesizers

Hewlett-Packard has added options to the HP 8370 family of synthesized microwave sources and improved the standard instrument performance at no increase in price. New standard features include linear AM modulation, lower specified phase noise, and seven FM sensitivity ranges for more accurate FM deviation settings. New options include, analog phase modulation, −110 dB step attenuator, and an internal modulation source. The HP 83710B and HP 83730B microwave sources have code compatibility with older HP 8673 generators. U.S. prices range from $21,400 for the HP 83711B to $33,150 for the HP 83732B
Hewlett-Packard Co.
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EMI Receiver

Messelektronik introduces a low cost, portable EMI receiver with a built-in memory card and LCD display. The superheterodyne receiver has intermediate frequency bandwidths of 0.2, 9.0, and 120 kHz and uses a synthesized LO and automated HF/IF attenuators. The receiver includes pre-selectors and has CISPR quasi peak demodulation. All functions are controllable via the front panel keyboard, IEEE, or optically decoupled RS-232 interfaces.
MEB Messelektronik Berlin
INFO/CARD #196

Radio Data Logging

The Radio Monitor System (RMS) from Microwave Communications Consultants incorporates a 12-bit analog to digital conversion board with inputs for up to 32 analog inputs. The software included with the system allows the user to set triggering levels to activate alarms, record statistical data, and view plots of stored data. RMS is available as a turnkey system, or it can be installed in an existing 286 to 486 PC.
Microwave Communications Consultants
INFO/CARD #197

Inductive Coupler

The PT-4000-200 RFI Unidapt® inductive coupler allows you to perform test measurements on cellular units with fixed or retractable antennasThe operation is based on the indutive signal transfer between the coupler and the original factory installed antenna. When used with any Unidapt® or Celludapt® adapter, the coupler allows you to perform power measurements on those cellular units whose antenna port you cannot reach.
RF Industries, Ltd.
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Spectrum Analyzers

Tektronix has introduced the Rhode & Schwarz FSEA series of spectrum analyzers. The FSEA has a fast minimum sweep time of 5 ms with fully synchronized sweep. The FSEA also has a zero span sweep time of 1 μs, full sweep. The FSEA also offers 25 screen updates per second. The FSEA 30 resolution bandwidth is 1 Hz to 10 MHz. Intermodulation free dynamic range is 113 dB, noise floor is −145 (at 10 Hz bandwidth), and phase noise is −123 dBc/Hz at 10 kHz offset. The

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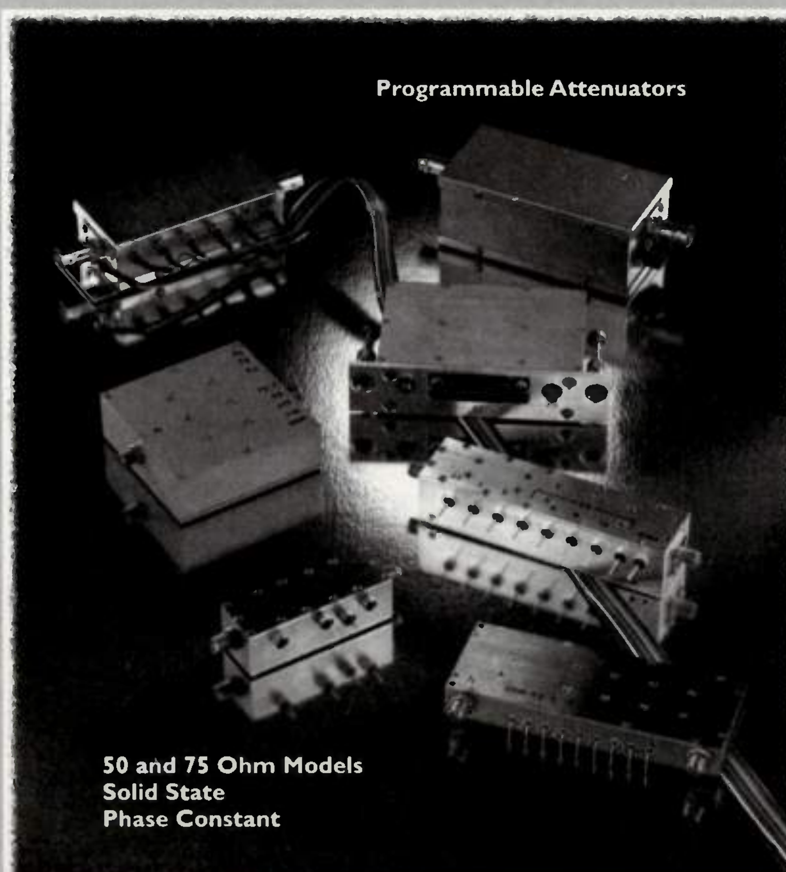

INFO/CARD 46

FSEA 20 is U.S. priced at $29,995 and the FSEA 30 is U.S. priced at $43,895.
Tektronix
INFO/CARD #199

Digital Modulation Generator

Anritsu Wiltron introduces the MG3670B, a digital modulation generator designed for testing mobile communications systems utilizing GSM, DCS1800,

DECT, PDC, PHS, NADC, and CT2 digital modulation techniques. The generator covers 300 kHz to 2.25 GHz, and has enhanced abilities for testing DECT systems operating at 1152 Kbits/sec. Typical adjacent-channel leakage power is -75dB at 600 kHz offset. Output level is variable from –143 to +13 dBm. The MG3670B has a suggested U.S. price of $45,500 without modules. The modules range in price from $3,550 to $4,050.
Anritsu Wiltron
INFO/CARD #200

DISCRETE COMPONENTS

Xtals for GSM

Tele Quarz offers 13 MHz quartz crystals which have been developed specifically for use in GSM handsets. These crystals have been designed to overcome the known problem of discontinuities of frequency (frequency dips). Resonance resistance is less than 60 Ω, and frequency stability is 10 ppm for the whole operating temperature range. The crystals are produced in several versions of the HC-52/U enclosure, including two SMD types.
Tele Quarz GmbH
INFO/CARD #202

Chip Caps

Philips Components' X7R, 16 V, surface-mount ceramic multilayer capacitors offer greatly increased capacitance per unit volume compared with 63 V and 25 V devices. The capacitors are offered in 0603 size, with capacitances from 22 nF to 47 nF, 0805 size, with capcitances from 47 nF to 220 nF; and 1206 size, with capacitances from 220 nF to 680 nF.
Philips Components
INFO/CARD #201

Resistor/Capacitors

KOA Speer Electronics introduces a new line of capacitor/resistor chips that provide impedance and noise filtering functionality. The CR73 capacitor/resistor chips are available in a 1206 package that reduces mounting space and component count by 50 percent.
KOA Speer Electronics, Inc.
INFO/CARD #203

Chip Inductors

Sprague-Goodman has expanded its SURFCOIL® product line to include the GLX series of very small, nonmagnetic SMT chip inductors. These epoxy encapsulated inductors are 2.0 × 1.25 × 1.25 mm. Inductance ranges from 10 nH to 1.0 µH with standard tolerances of ±10%. Q is typically 40 at 200 MHz, and up to 90 at 1 GHz. Self resonant frequency ranges up to 2500 MHz. Price is $0.63 each in quantities of 2,000.
Sprague-Goodman Electronics, Inc.
INFO/CARD #204

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Series 9401 Thin-Trim® capacitors operate to 2 GHz and high-

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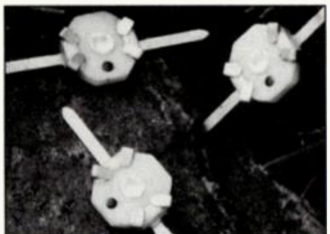


er. Mechanically, these capacitors measure 0.140 × 0.125 × 0.040 inches. The series has a capacitance range of 0.25 - 0.7 pF to 1.5 - 4.0 pF, with rated voltage of 250 VDC. Q is better than 100 at 2 GHz, and operating temperature range is –55 to +125 °C. Price is $1.02 in 25k quantity.
Johanson Mnfg. Corp.
INFO/CARD #205

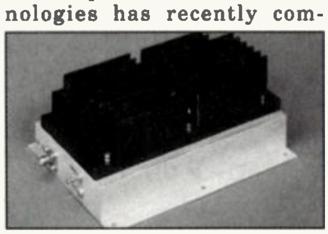
AMPLIFIERS

High Efficiency LNA

The AFS amplifier series offers low noise figure and moderate gain, with typical current drain of only 16 mA at +6 V. The AFS2-04000800-08-HE-2 is a 4 to 8 GHz design with 20 dB minimum gain, 0.8 dB maximum noise figure, and 0 dBm output power. All AFS series amplifiers are available in hermetically sealed chassis with military screening.
Miteq
INFO/CARD #206

Cellular Receive Amplifier

Chesapeake Microwave Technologies has recently com-



menced shipments of an 824-849 MHz amplifier for digital base stations. Specifications include 1.4 dB noise figure, +39 dBm IP3, +19 to +31 VDC at 900 mA input supply, and 38 dB gain.
Chesapeake Microwave Technologies, Inc.
INFO/CARD #207

Wideband Amplifiers

The UWLA 1000 series of linear amplifiers are rugged, low-power MOSFET amplifiers for the 9 kHz to 1000 MHz range. Different models fulfill all power needs from 1.5 to 10 W. Flatness is ±1.5 dB, and the harmonic suppression is is better than –20 dBc. Drive requirement is only 0 dBm.
Dressler Hochfrequenztechnik GmbH
INFO/CARD #208

Broadband 45 dB DLVA

American Microwave offers a truly DC-coupled 40/45 dB detector log video amplifier with a frequency range of 100 MHz to 18 GHz. Frequency flatness is better than ±1.0 dB (at –20 dBm, 0.1 to 18.0 GHz). Typical recovery time is 200 ns, and typical settling is 45 ns. The DLVA measures 1.5 × 2.2 × 0.4 inches.
American Microwave Corp.
INFO/CARD #209

CDMA Power Amp

Model SSPA8689-32/15191 from Microwave Power Devices provides low distortion amplification of CDMA modulated carriers used in 869 to 894 MHz cellular systems. This amplifier delivers 40 W total average output power with 32 dBc spectrum degradation in a 1.25 MHz CDMA channel. Gain is 60 dB ±1 dB. The amplifier uses +18 to +30 VDC, and consumes 20 A at 28 V. The unit is air cooled and measures 17 × 5.4 × 5.7 inches.
Microwave Power Devices, Inc.
INFO/CARD #210

CABLES & CONNECTORS

Reverse Polarity Connectors

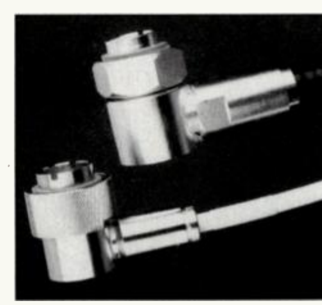
A family of reverse polarity connectors from Amphenol is designed for use in wireless applications where a non-standard interface has been mandated by FCC publication 47CFR PT 0 19. These connectors are available in popular styles such as TNC, BNC, and SMA.
Amphenol Corp.
INFO/CARD #211

Hermetically Sealed Connectors

Three hermetically sealed twinax/triax connectors from Trompeter Electronics prvent leakage through the inside of the connectors. The three bulkhead mount models include the feed through BJ78HS, front mount BJ77HS, and rear mount BJ79HS. The trio of new connectors joins the BJ28HS hermetically sealed coax bulkhead jack. The connectors meet MIL-C-49142 specifications. Cost of the BJ77HS is from $98 in quantities of 25.
Trompeter Electronics, Inc.
INFO/CARD #212

7/16 DIN Connectors

A series of 7/16 din connectors that can be made with a new low-profile design is being introduced by Tru-Connector. The new connectors use a knurled nut and accomodate cables from 0.141 to 7/8 inches in diameter.



They provide 50 impedance. Tru-Connector 7/16 series RF coaxial connectors are priced according to configuration.
Tru-Connector Corp.
INFO/CARD #213

SIGNAL PROCESSING COMPONENTS

Crystal Filter

Piezo Technology has introduced a 71.0 MHz fundamental crystal filter. Model 7996C features a minimum 3 dB bandwidth of ±92 kHz and a stop-

INFO/CARD 82

band attenuation of 60 dB at ±540 kHz maximum. Ultimate attenuation is 70 dB, ripple is 0.5 dB, and input and output impedances are 50 Ω. The unit is packaged in a 2.29 × 0.68 × 0.40 inch enclosure.
Piezo Technology, Inc.
INFO/CARD #214

Single Sideband Modulator

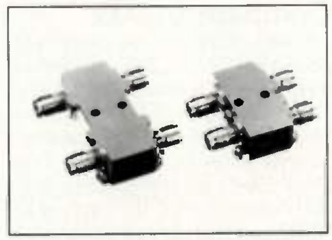
TRM's single sideband modulator, model SSM 352, features low conversion loss and high sideband suppression for single sideband modulation of an RF signal at a frequency of 352 ±15 MHz. I and Q baseband modulation inputs are DC to 10 MHz. Maximum conversion loss is 8 dB, with signal input of +10 dBm and modulatio input level of 0 dBm. Carrier and sideband suppression is specified as –25 dBc minimum.
Technical Research and Manufacturing, Inc.
INFO/CARD #215

Dual Directional Coupler

Sage Laboratories' FC5546 dual directional coupler covers the frequency range of 1030 to 1090 MHz and has a maximum VSWR of 1.15. The coupling is 20 ±1 dB, with a minimum directivity of 30 dB. Power rating is 10 kW (peak), 500 W average. The unit uses a conservative air trough line construction to reliably handle the specified power. Package size is 3.5 × 2.4 × 1.2 inches, not including connectors and mounting feet.
Sage Laboratories, Inc.
INFO/CARD #216

Directional Couplers

A series of 90° hybrid directional couplers from Lucas Weinschel have been designed to achieve high isolation and low SWR in a miniature stripline configuration. Model



1541R2 operates from 0.5 to 1.0 Mhz, model 1541R3 from 1.0 to 2.0 GHz, and model 1541R4 from 2.0 to 4.0 GHz. All have coupling loss of 3.1 ±0.6 dB, and all have frequency sensitivity of ±0.5 dB. Power handling for the series is 50 W (average), 3 kW (peak). Stainless steel female SMA connectors are standard.
Lucas Weinschel
INFO/CARD #217

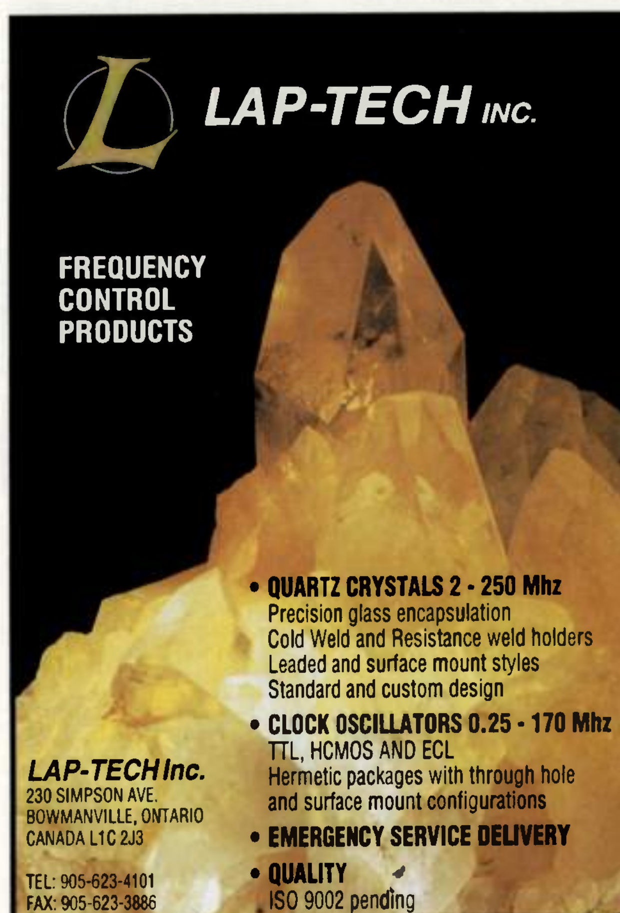
PCS Diplexer

The PT1800SA from Penny Technologies is a high isolation diplexer designed for PCS applications. The diplexer is a three-port device using combline technology. The receive side operates from 1710-1785 MHz, and the transmit side operates from 1805-1800 MHz. Maximum passband insertion loss is 2.0 dB, and transmit to receive isolation surpasses 85 dB. The VSWR for all ports is 1.5:1. Power handling is 100 W. The PT1800SA is priced at $250.00 for 1 to 10 pieces.
Penny Technologies
INFO/CARD #218

CompactBias Tee

Wide band coverage and low insertion loss are features of the PBTC-1G bias tee from Mini-Circuits. Operating from 10 to 1000 MHz, this bias tee has typical insertion loss of 0.6 dB. It accepts a maximum input current of 500 mA, and maximum RF power of 30 dBm. The bias tee measures 0.8 × 0.8 × 0.4 inches. Price is $25.95 each.
Mini-Circuits
INFO/CARD #219

INFO/CARD 47



INFO/CARD 48

SIGNAL SOURCES

VCOs

Synergy Microwave announces the availability of narrow band, coaxial resonator voltage controlled oscillators (VCOs). These units cover narrow bandwidths from 450 to 1900 MHz and offer superior phase noise characteristics. Typical bandwidths are 40 MHz

around center nominal frequencies. At 10 kHz offset from the carrier, phase noise is better than –100 dBc/Hz for a 450 MHz carrier and –120 dBs/Hz for the 950 MHz carrier. Output signal level is typically +10 dBm with a bias voltage of +12 VDC.

Synergy Microwave Corp.
INFO/CARD #220

OCXO

MTI - Milliren Technologies introduces the 230-series OCXO. The 230-series makes use of both AT and SC-cut resonators. A stability of 2.5×10^{-8} over –30 to +70 °C is specified for models using the SC-cut resonators. AT-cut performance is guaranteed at 2.0×10^{-7} over the same temperature range. The 1.42 × 1.07 × 0.76 package is an industry standard dimension. The 1.00 × 0.70 inch footprint is pin-compatible with many current applications in wireless, cellular, and test equipment applications.

MTI- Milliren Technologies, Inc.
INFO/CARD #221

Clock Oscillator

Model 222 from Reeves-Hoffman is a low cost clock oscillator with tristate output. Encased in a 14-pin DIP package, the model 222 is able to drive TTL and CMOS loads and provides a reduced parts count for better reliability.

Reeves-Hoffman
INFO/CARD #222

Compact OCXO

Oak Frequency Control Group's 4597 OCXO, available from 12 to 30 MHz, features a TCXO-sized footprint of just 30.3mm × 30.3mm and a seated height of only 10.2mm. The 4597 meets a temperature stability spec of $\pm 3 \times 10^{-8}$ over 0 to 70 °C and features aging of ±0.5 ppm per year. The compact OCXO operates from a +5 V supply with a +4 V reference. Typical pricing is $120/1000, and samples at 16.384 are available for quick delivery.

OAK Frequency Control Group
INFO/CARD #223

SEMI-CONDUCTORS

2.7 V, 2.4 GHz Transistors

Hewlett-Packard has introduced additions to its series of silicon bipolar transistors that are optimized for 3 V operation. The AT-32011 provides 14 dB gain and 1 dB noise figure at 900 MHz when biased at 2.7 V, 2 mA. The AT-32033 combines 12.5 dB gain and a 1.0 dB noise figure at 900 MHz. These transistors may be used between 1 and 5 V with as little as 1.0 mA of current. The AT-32011 comes

in a four-lead SOT-143 package, the AT-32033 in a three-lead SOT-23 package. U.S. pricing in quantities of 1 to 999 are $0.58 and $0.55 for the AT32011 and AT32033, respectively.

Hewlett-Packard Co.
INFO/CARD #224

10 W, UHF Power Transistor

Philips' new 10 W UHF power transistor for common emitter class AB operation in the 820 - 960 MHz range is now available from Richardson Electronics. The BLV910 has a minimum gain of 11 dB and 55% efficiency at 10 W. Other features include emitter ballasting resistors, gold metalization, and internal matching. The BLV910 has a six-lead SOT171 flange envelope with a ceramic cap.

Richardson Electronics, Ltd.
INFO/CARD #225

14-bit, 10 MHz ADC

Datel's ADS-945 is a 14-bit, 10 MHz sampling A/D converter with no missing codes to the 14-bit level over the full military temperature range of –55 to +125 °C. It has signal to noise ratio of 78 dB and a total harmonic distortion of –80 dB. Requiring ±15 V, +5 V and –5.2 V supplies, the ADS-945 typically dissapates 4.2 W. The ADS-945 contains a track-and-hold amplifier, reference, timing/control logic, and error correction circuitry, and is priced at $866 in 100-piece quantities.

Datel, Inc.
INFO/CARD #226

Log Amp

The AD641 from Analog Devices is a high-speed, logarithmic demodulating amplifier designed to operate at frequencies up to 250 MHz while typically drawing only 35 mA. The fully differential signal path allows for easy cascading of two AD641s to achieve 58 dB of dynamic range with ±2.5 dB accuracy. The amplifier costs $19.95 in lots of 1000.

Analog Devices, Inc.
INFO/CARD #227

1.8 GHz PA Driver

The Motorola MRFIC1806 driver amplifier and ramp circuit is designed primarily for use in DECT, PHS and other wireless PCS applications. The device includes a two-stage driver amplifier and transmit waveform shaping circuitry in a low-cost SOIC-16 package. The ramping circuit controls the burst-mode transmit rise and fall time and is adjustable through external components. The MRFIC1806 has typical gain of 22 dB, and operates from 3.0 to 5.0 Volts.

Motorola Semiconductor
INFO/CARD #228

10-Bit, 40 Msps ADC

The HI5703 is a 10-bit, 40 Msps analog to digital coverter with 3.3 V-compatible digital outputs and on-chip sample-and-hold. The ADC has a minimum 8.3 ENOB (effective number of bits) at f_{in} = 10 MHz with 450 mW maximum power dissapation at 40 Msps. Spurious-free dynamic range at 10 MHz is 54 dB minimum and 63 dB typical for all spurious components across the Nyquest band (0-20 MHz) input. The HI5703 is available from stock for approximately $35.00 in 1000-piece quantities.

Harris Semiconductor
INFO/CARD #229

Quad, Wideband Op Amp

Burr-Brown's OPA4658 is a quad, ultra-wideband, lowpower current feedback operational amplifier featuring a high slew rate, and low differential gain/phase errors. Its low 50 mW power dissapation combines with a high bandwidth of 450 MHz at a gain of 2. Differential gain and phase errors are 0.015% and 0.02°, respectively. OPA4658 is priced from $5.76 in 1000 piece quantities.

Burr-Brown Corp.
INFO/CARD #230

600 kbps, 12-bit ADC

Linear Technology's LTC®1279 is a complete 600 kbps analog to digital converter that features true 12-bit performance from DC to Nyquist. Dynamic performance is is 70 dB signal-to-noise plus distortion and 74 dB total harmonic distortion. Typical power consumption is 60 mW - 8.5 mW in its powerdown mode. Pricing in 1,000 and up quantities is $13.00 in 24-pin SOIC packages.

Linear Technology Corp.
INFO/CARD #231

750 MHz Voltage Feedback Op Amp

Comlinear has announced the CLC440 voltage feedback op amp, with 1500 V/us slew rate and 750 MHz unity-gain bandwidth. Differential gain and pahse errors are 0.015% and 0.025°, respectively. Input noise is $2.5\ nV/\sqrt{Hz}$. The CLC440 is available in 8-pin plastic DIPs or SOICs for $4.20 in 1000-piece quantities.

Comlinear Corp.
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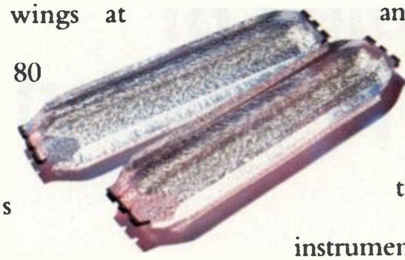
PERFECT STABILITY

The Hummingbird is acclaimed as the epitome of stability. This it achieves by beating its wings at frequencies up to 80 beats per second.



C-MAC's frequency control products are equally renowned for their perfect stability, even at frequencies up to 2.6 GHz.

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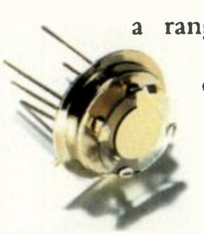
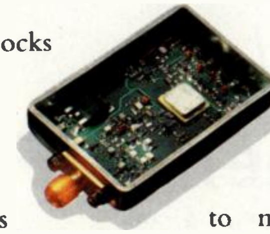


C-MAC oscillators are used worldwide in SONET / SDH systems and now in the latest ATM applications.

They are widely used in master and slave clocks in telecom networks and as precision references for frequency and time measurement instruments.



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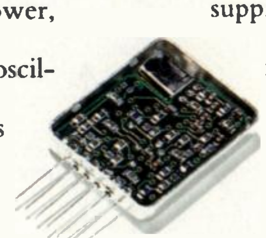

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Essex CM20 2DE England
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C-MAC

INFO/CARD 49

Fundamental Principles of Class AB Linear Amplification

By Gary A. Breed
Editor

This tutorial is intended to provide a less-experienced RF engineer with the basic concepts that define the operation of Class AB linear (or more accurately, approximately linear) RF power amplifiers. This class of operation has increasing importance as complex-modulated digital communications systems become more commonplace.

As a group, the various classes of amplification define the operating conditions, bias requirements, output waveforms and, usually, the circuit topology of power amplifiers. The linearity performance of each class is quite different: from being a meaningless specification for a hard-limited switch-mode amplifier, to perfect linearity for an ideal Class A amplifier.

The Class AB amplifier is a compromise, intended to provide linearity that is sufficient to meet the requirements of a particular system, usually traded against power consumption and corresponding thermal concerns.

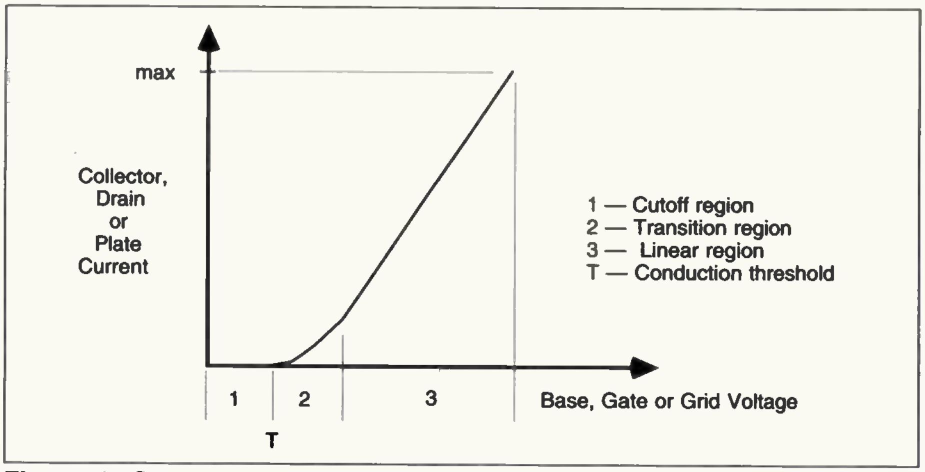

Figure 1. Generalized characteristic input/output curve for amplifying devices. The actual voltages and currents, and the amount of curvature will vary for different bipolar, FET or vacuum tube devices.

Class A and Class B

The transfer characteristic, collector current versus base voltage, of a typical transistor is shown in Figure 1. The general shape of this curve is similar for bipolar transistors, FETs, and even vacuum tubes. There are significant differences, however. The actual base, gate or grid voltages vary widely, as does the range of collector, drain or plate current. The curve is generally shallower for FETs and vacuum tubes than for bipolars, with the smaller deviation from an ideal straight line a major factor in the greater popularity of these devices over bipolars in linear amplifier applications. The curve has three distinct regions: (1) The cutoff region where no current flows; (2) The

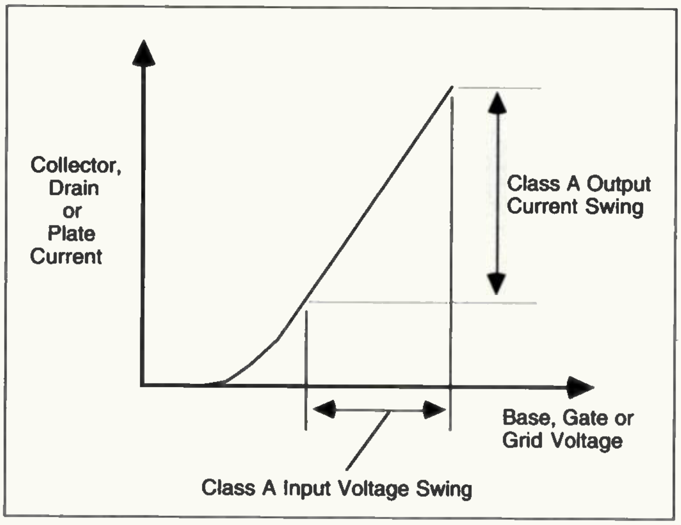

Figure 2. Class A operating region on the characteristic curve for a practical device (an ideal device would start at zero output port current).

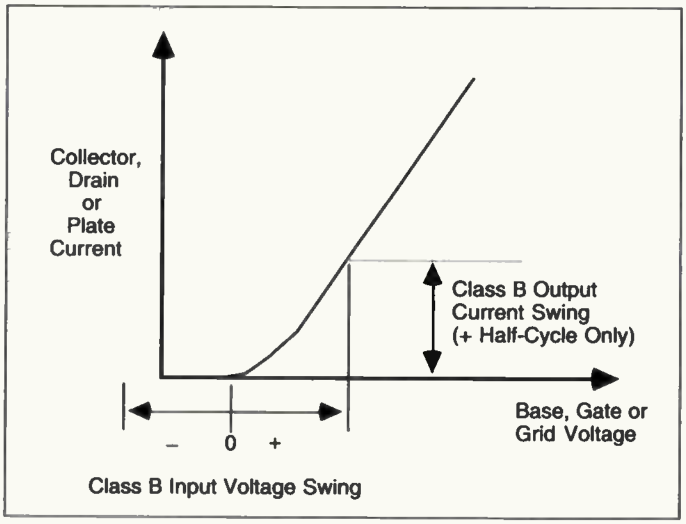

Figure 3.The Class B operating region places the center of the input waveform at the conduction threshold.

2.5KHz-8GHz AMPLIFIERS

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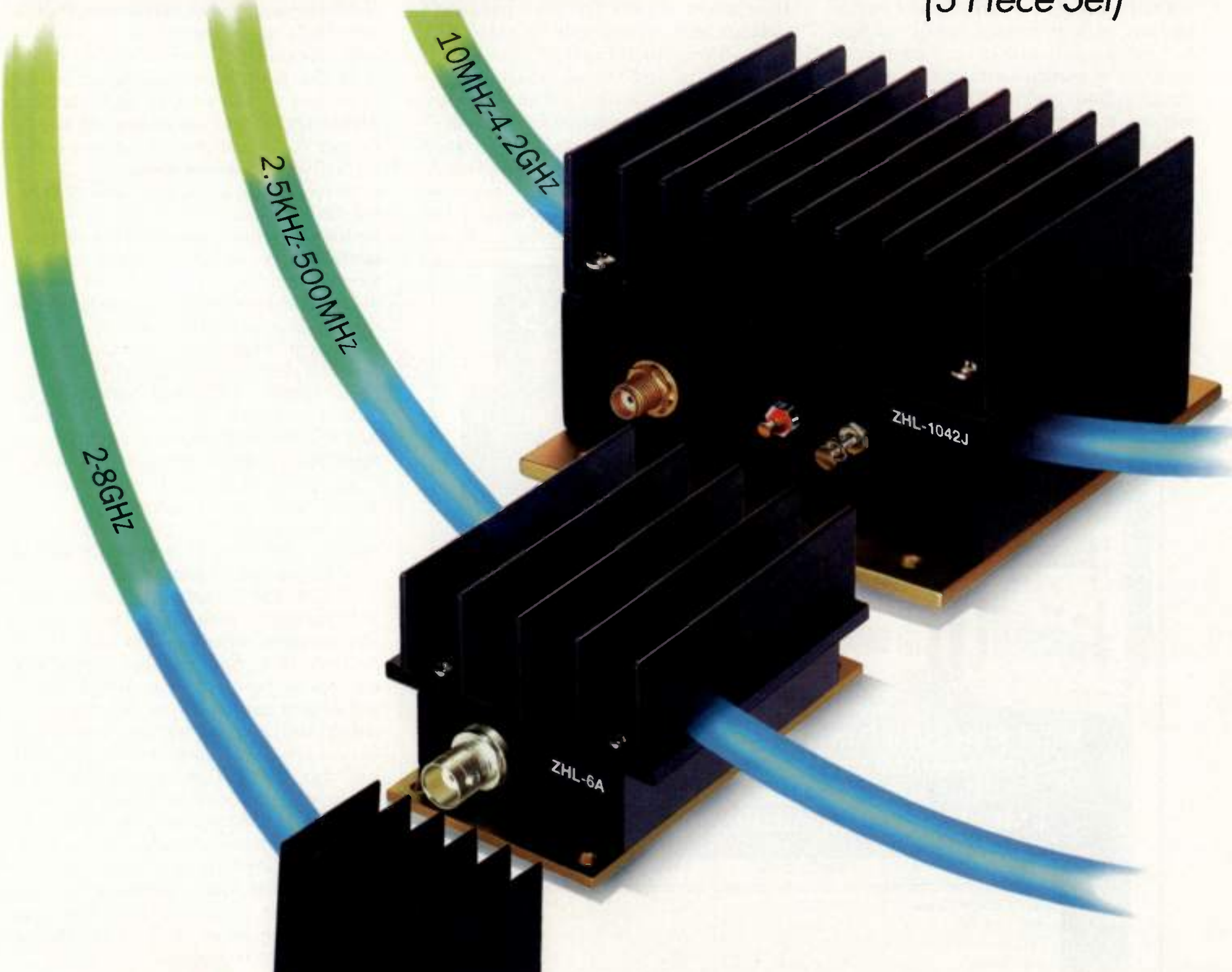

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ZHL-6A	.0025-500	21	+23	+24	350	BNC	199
ZHL-1042J	10-4200	25	+20	+15	330	SMA	495
ZRON-8G	2000-8000	20	+20	+15	310	SMA	495

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Expand laboratory capabilities and put a full spectrum of power at your fingertips with Mini-Circuits 2.5KHz to 8GHz medium power amplifier set. Each ultra-wideband set contains three individual heat sinked RF amplifiers with at least +20dBm output and overlapping frequency response range capabilities; 2.5KHz to 500MHz, 10MHz to 4.2GHz and 2GHz to 8GHz. Applications for these amplifiers include increasing the signal levels to power meters, spectrum analyzers, frequency counters and network analyzers as well as boosting signal generator outputs.

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transition region where current begins to flow with increased input voltage, but not linearly due to turn-on characteristics of transistor junctions or tube electron flow; and (3) The linear region where a change in input voltage results in an exactly proportional change in output current.

Manufacturer's data provide such specifications as turn-on threshold and maximum values for base/gate/grid voltage and collector/drain/plate current. Along with current gain (beta) for bipolars and transconductance for tubes and FETs, the parameters are the basis for linear amplifier design.

Figure 2 shows the operating region on the characteristic curve for Class A. True Class A maintains all operation within the most linear portion of the characteristic curve. Note that this is very high on the input voltage and collector/drain/plate current scales. Class A is the most linear operation available, but at the cost of high current consumption and its attendant inefficiency (greatest power consumption for a given RF power output).

In receiver applications, and in critical transmitter circuits, Class A may be the best choice to obtain the desired performance. In small signal applications, Class A is used almost exclusively, because power consumption is not an issue at such low levels.

Class B (Figure 3) is defined as conduction during one-half of the input cycle. Ideally, a Class B device will be biased to exactly its turn-on threshold, and will conduct when the input waveform goes positive beyond that point. Class B has little value with a single device, and is used almost exclusively with two devices in a push-pull configuration (driven with signals having a 180° phase difference).

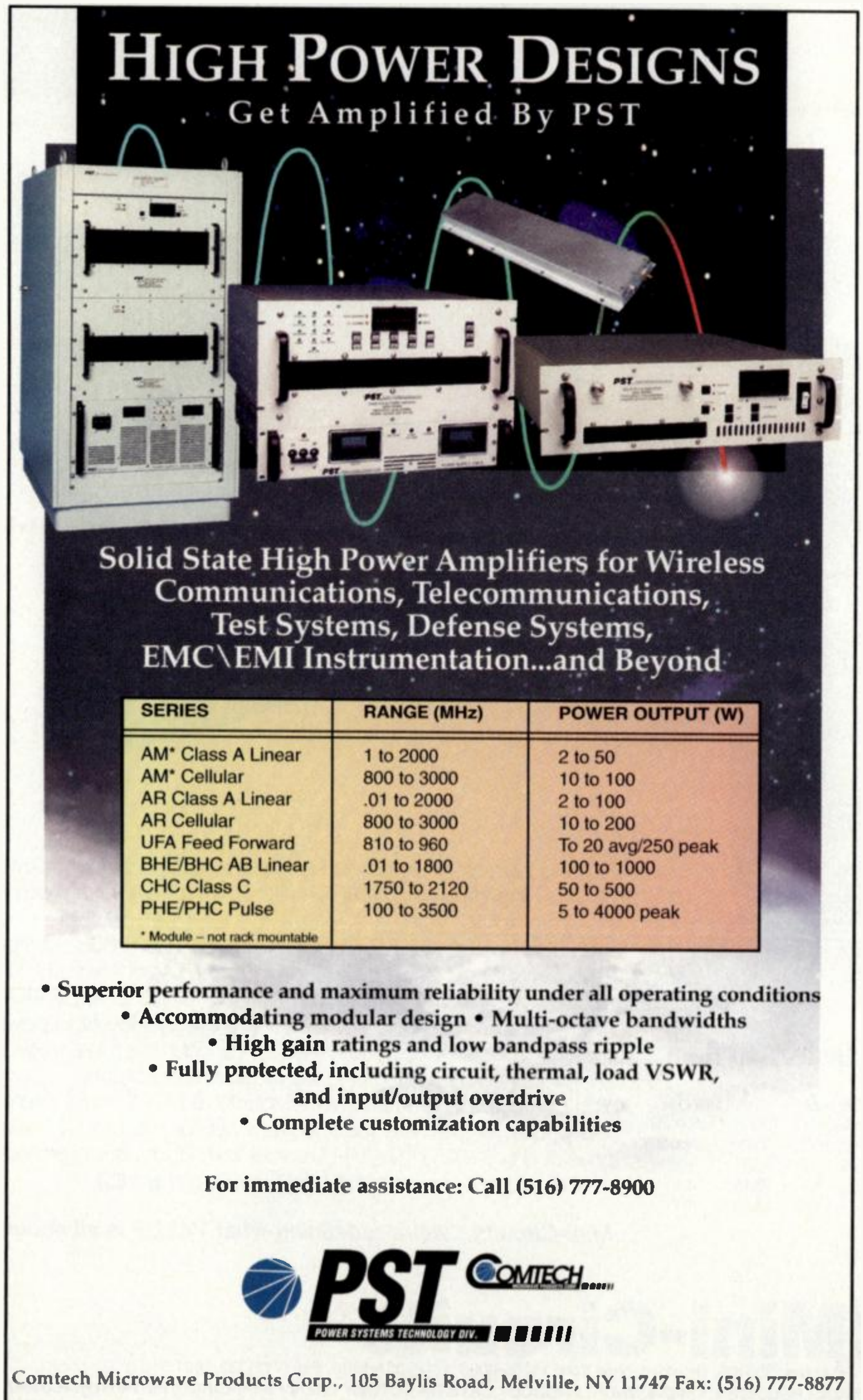
Class B has the advantage of reasonable linearity, with much lower power consumption requirements than Class A. But, it is not "up the curve" far enough for best linearity. Also, Class B requires a push-pull configuration to obtain its linearity performance, each device contributing power to the output during half the time. Such an arrangement requires either matched devices or compensating circuitry, both of which add complexity and cost. Push-pull operation in any class of amplifier has the advantage of suppressing the even-numbered harmonics that are generated by whatever non-linearity is present.

A Compromise — Class AB

As you might expect, Class AB places the bias point somewhere between Class A and Class B on the characteristic curve, as shown in Figure 3. The point to be chosen requires consideration of several factors:

Desired Linearity — How much performance is necessary? Does two-tone testing need to have 3rd order distortion products 30 dB below either tone? Or, is 20 dB acceptable; or 40 dB?

Power Consumption — (including thermal considerations, which are directly related). Is this a high-power amplifier where heat dissipation makes a big difference is size and cost? Is there a limit on the power supply available for the required output? Does the required package or enclo-



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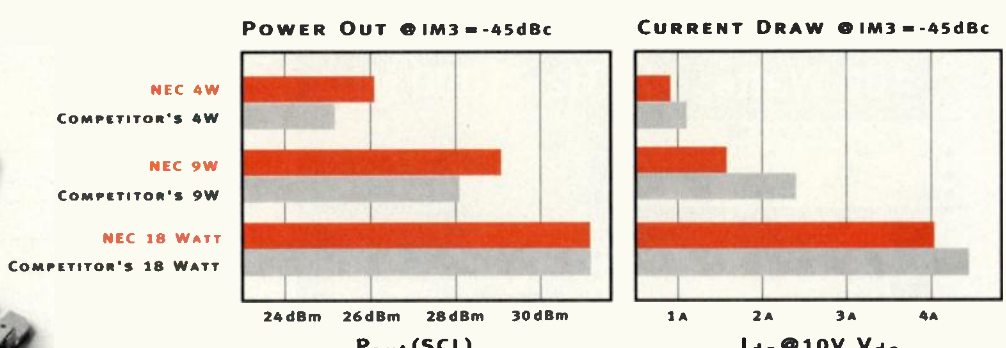

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Figure 4 shows two methods for biasing bipolar transistors. The first is the most simple. The voltage drop across a diode junction provides a relatively constant bias voltage. The voltage is a fair approximation of the turn-on threshold of bipolars, since the base-emitter junction is essentially a diode. Also, the changes versus temperature for the bias diode and the transistor will be similar. This method has an additional necessary characteristic for bipolar biasing — low impedance. Bipolar inputs are low RF impedances, and the bias circuit must be low enough to avoid changes due to self-biasing as RF input currents are rectified by the base-emitter junction diode.

A low-impedance, temperature-compensated voltage regulator is a more flexible biasing method for bipolars. The output impedance must be very

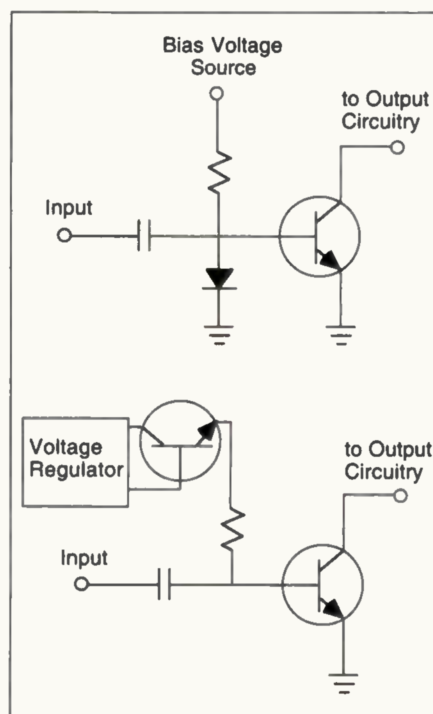

Figure 4. Typical biasing methods for bipolar transistors: Simple diode-voltage-drop biasing (top); low-impedance series voltage regulator (bottom).

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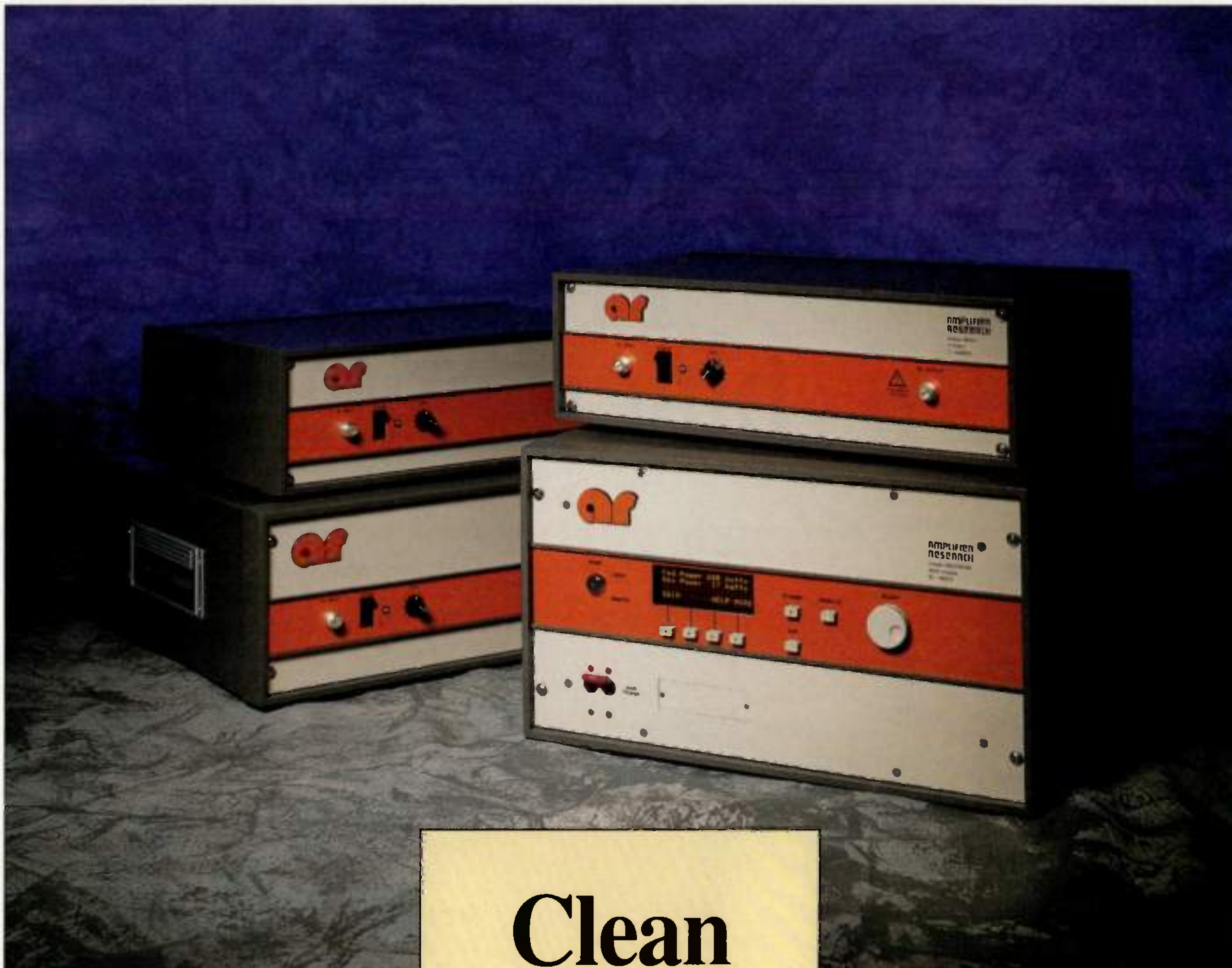

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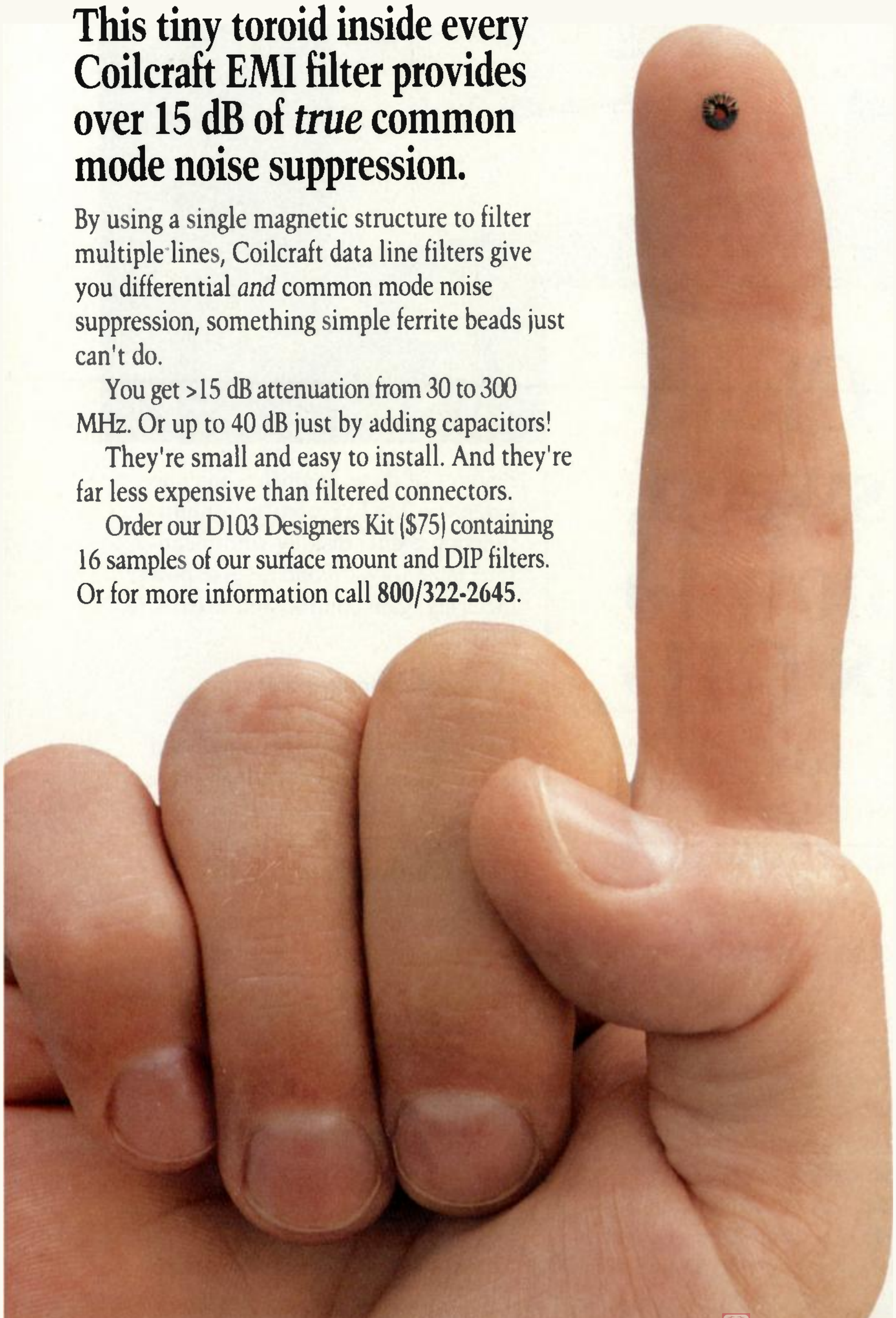

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low, as noted above. Also, the performance of the regulator at the low voltages (on the order of 1 volt) is important, as well.

The physical structure of FETs gives them high resistive impedance, but very low capacitive reactance (large value of input capacitance). The capacitance can represent an extremely low impedance at RF. To present a more practical impedance to the driving circuitry, series resistance is sometimes added to the gate circuit. Also, DC continuity to ground is needed to protect the gate from an unwanted charge buildup (it *is* a big capacitor).

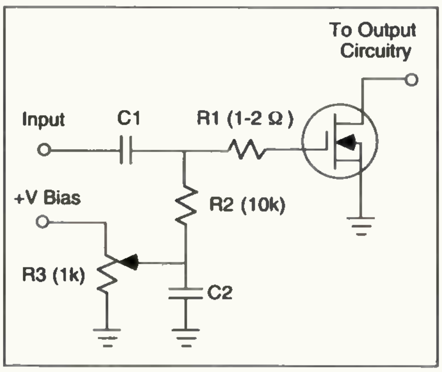


Figure 5. Example of a FET input circuit that accomplishes several circuit objectives.

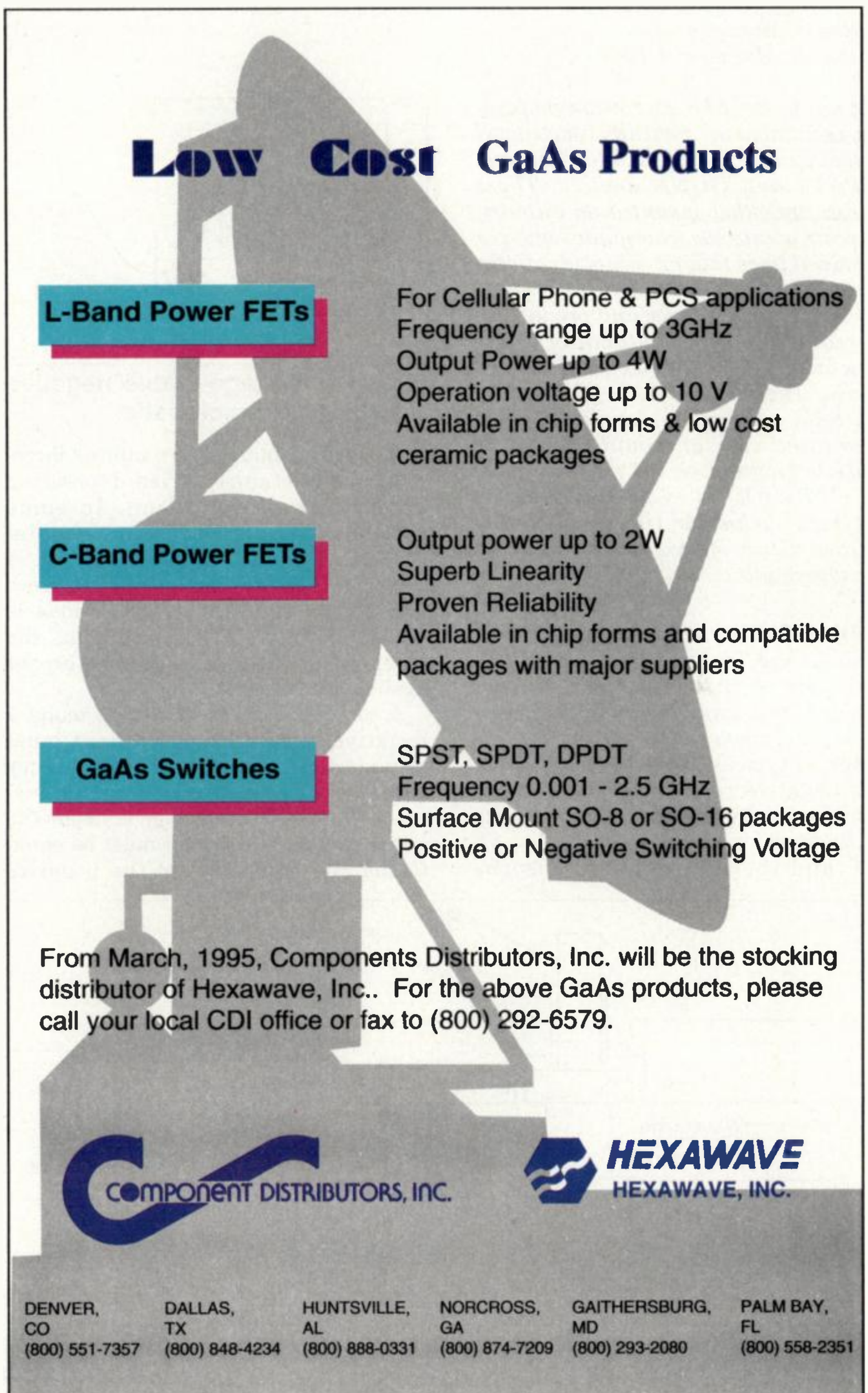
Figure 5 shows just one way to accomplish those design goals while providing a fixed DC bias to the gate. C1 isolates the RF input matching circuit, and should be an appropriately low reactance. R1 raises the gate impedance to a practical value for a broadband input transformer. R2 fixes the bias circuit impedance at a high value, keeping RF energy out, aided by bypass capacitor C2. R1 completes the DC path from gate to ground, and allows adjustment of bias voltage. D1 is simply a protection for the bias source against any reverse voltage that might result from a transistor failure. This is probably the minimum-component solution to proper FET input. However, other configurations are also possible, which may simplify another portion of the circuit, offer greater reliability, or better accommodate feedback.

Conclusions

Class AB offers a designer the trade-off of power consumption, efficiency and heat for modestly lower linearity than the maximum afforded by Class A. The choice of exact bias point depends on the linearity requirements, which are usually specified as a maximum level of distortion products in a two-tone or multi-tone test. The linearity requirements may also be augmented by linearizing circuitry — feedback and feed-forward techniques.

In general, some degree of linearity is required for any modulation type that includes amplitude control, including amplitude modulation, single-sideband and the various forms of quadrature amplitude modulation.

Applications using these modulation types are seeing increased usage, especially in digital communications systems, and RF amplifier engineers must design to specifications that require consideration of Class AB. *RF*



INFO/CARD 56

Program Aids Design of Coaxial and Waveguide Oscillators

By Roy C. Monzello
California Microwave, Inc.

At millimeter and microwave frequencies, oscillators are typically constructed in waveguide using diodes such as IMPATT and GUNN devices. These devices are either mounted on cylindrical posts within the waveguide region or in coaxial lines that are coupled into the waveguide via a coaxial to waveguide transition. Performance and producibility requires an understanding of the structural effects on electrical performance. The work of A.G. Williamson has been utilized to develop a post-mount and coaxial mount waveguide analysis program on the personal computer. WGMOUNT is used to study the electrical behavior (i.e., impedance characteristics) of a coaxial line coupled into waveguide.

IMPATT (IMPact-ionization Avalanche and Transit Time) and GUNN diodes are negative resistance devices. Negative resistance devices are components that convert DC power into AC power. A typical current-voltage curve of a negative resistance diode might look like Figure 1. The static resistance for the diode is always positive (i.e., v/i > 0) and therefore always absorbs power at DC, but the dynamic or incremental resistance varies depending upon the bias condition. In some regions of the curve the incremental resistance (dv/di) is positive, but in region a-b the incremental resistance has a negative value. If the diode is biased at point Q the circuit has the potential to oscillate under the proper loading conditions.

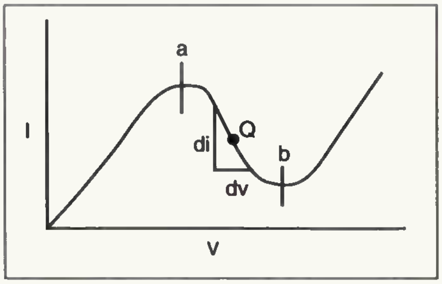


Figure 1. Voltage-stable negative resistance characteristic.

A complete oscillator design using a negative resistance component must provide for loading with a frequency selective circuit of the proper resistance. For optimum power transfer the resistive part of the load must be equal to the absolute value of the negative source resistance. If the resistive load is greater than the source resistance then the voltage will increase across the source driving the diode farther into the regions around points a and b where $|r_n|$ becomes very large. As $|r_n|$ increases the available power will decrease and eventually $|r_n|$ will be equal to the load resistance at which point circuit equilibrium will be reached.

Oscillator Structures

There are two popular methods of constructing a waveguide oscillator employing diodes as the active element: 1) a coaxial mount coupled into waveguide via a coaxial-to-waveguide transition, and 2) a post-mounted diode positioned within the waveguide cavity. Both of these methods are shown in Figures 2 and 3.

The waveguide mounted configuration is used in place of the coaxial mounted version when, due to package parasitics, the negative resistance of the diode is much closer to that of the waveguide mount impedance. If the resistance of the diode is very low (on the order of an ohm) then the coaxial

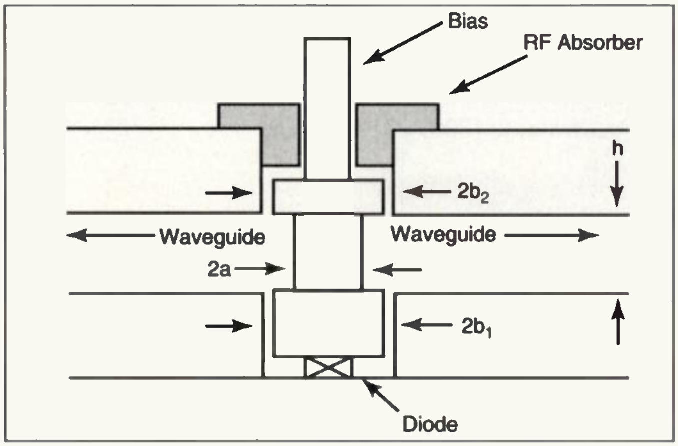


Figure 2. Diode oscillator using a coaxial mount coupled into waveguide.

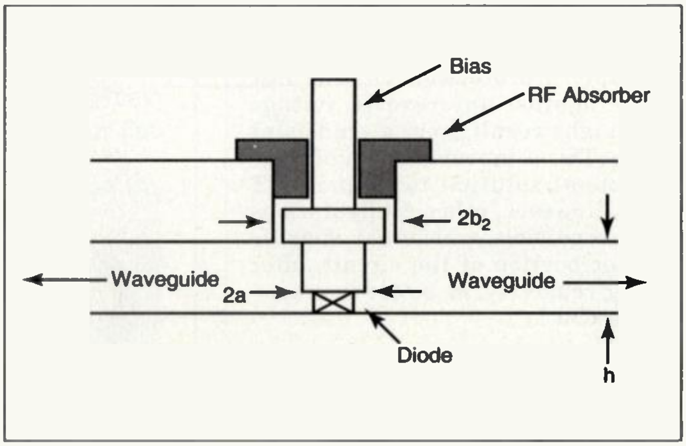


Figure 3. Diode oscillator using a coaxial entry with a post gap structure.

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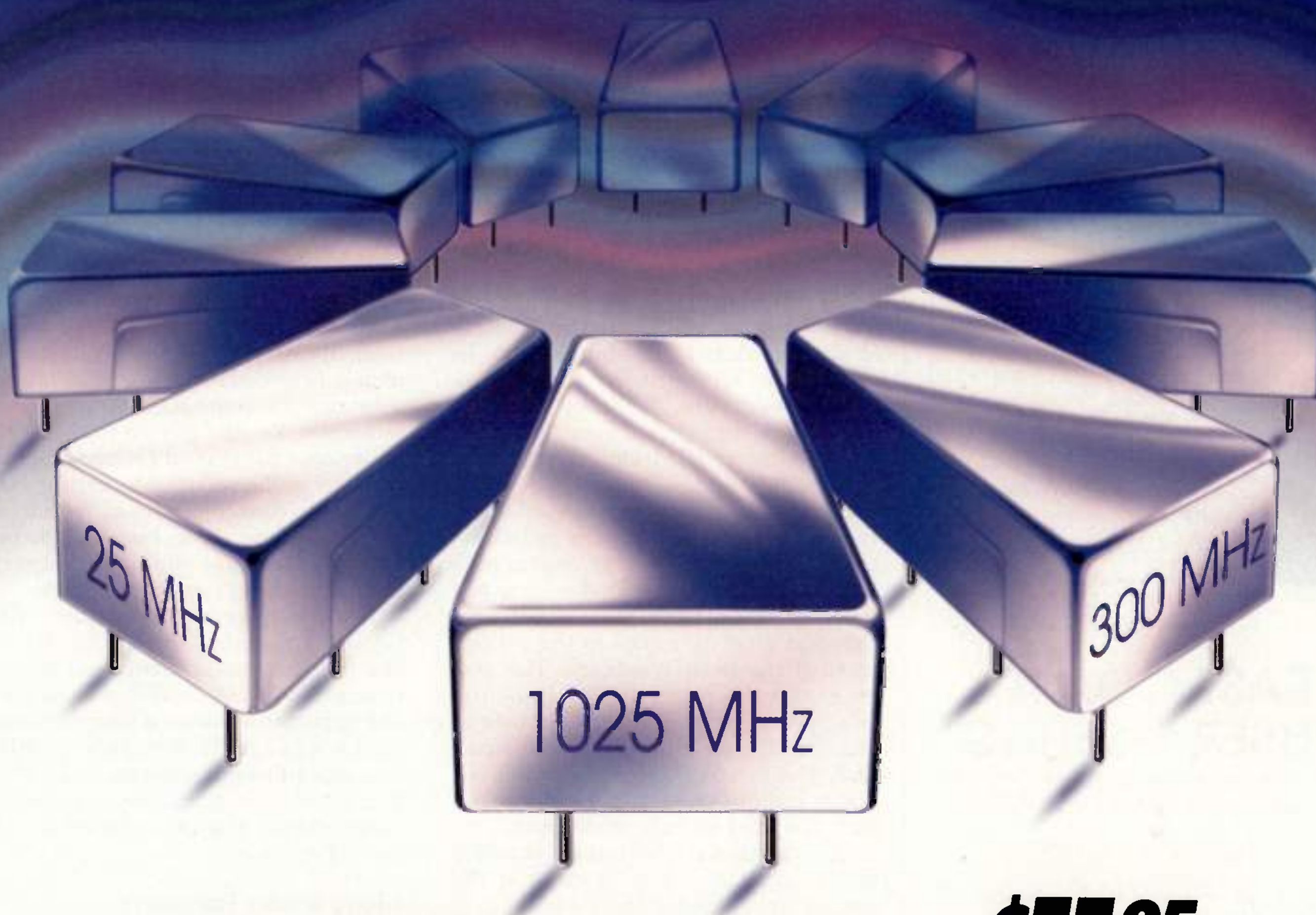

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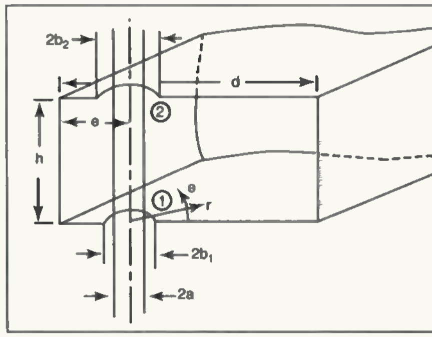

Figure 4. Double coaxial entry structure as analyzed by A.G. Williamson [2].

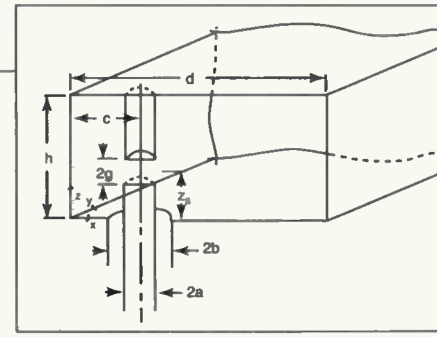

Figure 5. Coaxial entry with post-gap structure as analyzed by A.G. Williamson [3].

mount is preferred due to the transforming characteristics of the quarter-wave transformer that connects the coaxial mounted diode to the waveguide entry point.

In both cases DC bias is applied to the diode through the coaxial opening at the top. A low impedance choke in addition to RF absorber material is used to isolate the RF circuit from the bias circuit and prevent any substantial amount of RF from leaking back onto the bias line.

These structures can be used for a variety of applications such as coaxial to waveguide transitions, waveguide filters, and oscillators. To model any of these circuits thoroughly, additional elements must be added to the various ports of the basic structure. The program has the capability of modeling two-port networks of any number of elements for all four ports of the structure. The four ports mentioned are the upper coaxial port, the lower coaxial port, and the two waveguide ports.

The impedance calculations provided by the analysis are referenced at the coaxial aperture for the coaxial entry case and at the post circumference for the post gap case. These impedance calculations are then transformed (optionally) through an arbitrary two-port network to derive the final impedance characteristics of the structure. A circuit file, called the upper coax file, is used for placing component information that transforms a load impedance in the upper coaxial line to the coaxial-waveguide interface. After the coaxial-waveguide analysis has been completed to derive the impedance at the lower coaxial-waveguide interface, the resulting impedance is transformed (optionally) through another two-port circuit using the lower coax file for component information.

The preceding description is true for the following structures; double coaxial entry, single coaxial entry, and the single coaxial entry with a gap in the center conductor. The gap can be loaded with an arbitrary load to simulate a variety of conditions. For the post-gap structure the circuit file for the upper coax file can be used to provide a circuit transformation between the gap load impedance and the circumference of the post. For the case of the "Gap with a Coaxial Entry" (see Figure 5) the coaxial port is loaded and the impedance calculations are referenced at the circumference of the post gap. In this case, the upper coax file is used for load transformations to the waveguide aperture and the lower coax file is used for impedance transformations from the circumference of the post gap to the output of the user defined two-port network.

The two waveguide structures illustrated in Figures 4 and 5 are only two of many waveguide structures rigorously analyzed by A.G. Williamson[1]-[5]. His approach involves a transformation of the waveguide structure problem into an equivalent antenna array which is excited by a magnetic current frill representing the excitation from the coaxial aperture.

Diode Mount Parasitics

The post coupling analysis of Williamson provides the impedance characteristics at different reference planes depending on the problem at hand. For the coaxial entry problem the reference plane is located at the entry point into waveguide; for the post-gap problem the reference plane is at the edge of the gap (post circumference). To achieve an accurate understanding of the impedance characteristics presented to the diode, and hence oscillator performance, it is important to include the mounting parasitics of the diode. The mounting parasitics can have a major influence on impedance behavior and can limit diode performance if not properly accounted for.

Of the two structures previously described, the double coaxial entry structure contains the greater number of parasitic elements as can be seen by comparing Figures 6 and 7 . In analyz-

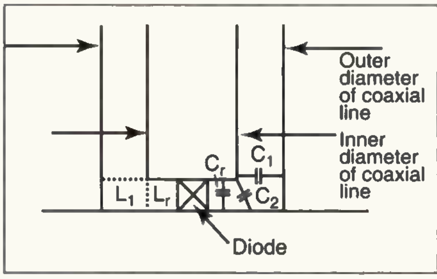

Figure 6. Mount parasitics for a diode mounted at the end of a coaxial line.

ing the oscillator structure the impedance calculated by the post coupling analysis must be properly transformed to the diode chip. What becomes evident after a few structures have been analyzed is the major, often dominating effect mount parasitics have on overall circuit performance. To a large extent, diode package parasitics also play a major role in contributing to mount parasitics and must therefore be included in any meaningful study of circuit behavior. Figures 8 and 9 show the circuit models for the two cases.

Using WGMOUNT

The Upper, Lower, and Waveguide files are circuit files that define two-port circuits which are connected to the respective ports of the structure. A representative circuit file might look like this:

```
ind ser l = 12
cap par c = 130
trl zo = 20  l= 0.25  er= 2.05
```

A simple editor and a number of component types are supplied to allow the input and editing of these circuits.

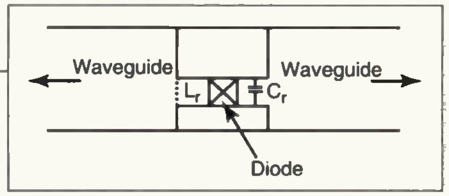

Figure 7. Mount parasitics for a diode mounted on a waveguide post.

The parameters of the waveguide junction are shown on the main screen along with frequency and variational parameters. These values may be entered or edited from the edit menu.

The analysis routine calculates the impedance characteristics of the structure as a function of either frequency or a chosen waveguide junction parameter. The program default is frequency, but it can be changed to any of the structural parameters.

Graphing Data

Once data has been calculated the results may be displayed graphically by pressing "G" from the main menu. Both the reactive and real parts of the impedance/admittance can be graphed, with both manual and auto scaling. The results of calculation can also be saved to an ASCII file and imported into a spreadsheet or graphing program.

Electrical Characteristics

As an example of how the program would be employed to analyze the electrical sensitivity to structural parameters, assume the following conditions; diode height = 0.035″

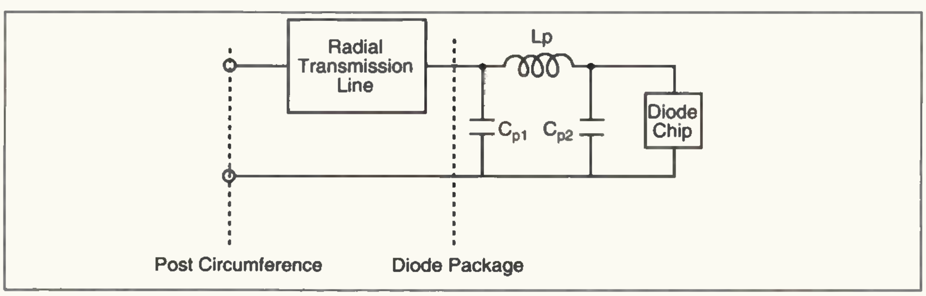

Figure 8. Circuit model for impedance transformation from circumference of waveguide post to diode chip.

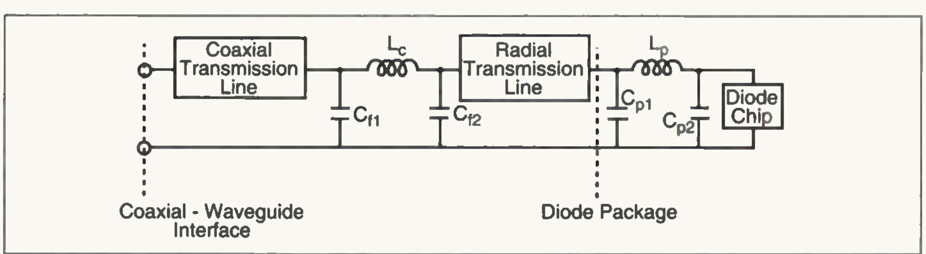

Figure 9. Circuit model for impedance transformation from coaxial-waveguide aperture to diode chip.

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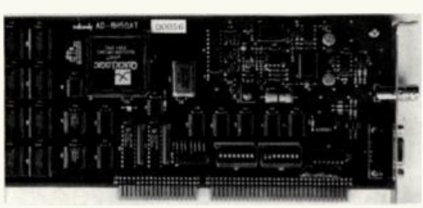

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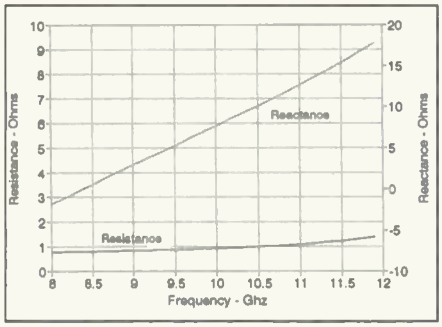

Figure 10. Circuit impedance at the diode chip for the coaxial mounted diode oscillator.

diode diameter =0.050″
diode chip impedance = –0.9 – j 6.0 at 10 GHz
waveguide size = X-Band (0.9″ x 0.4″)
coaxial hole diameter = 0.101″

After varying various parameters using the computer simulation, a configuration was determined which provides the proper impedance (approx. 0.9 + j6.0 ohms) at the reference plane of the diode chip. These parameters are as follows:

waveguide width = 0.9″
waveguide height = 0.4″
coax outer diameter = 0.1015″
post diameter = 0.075″
Upper coax: RF choke with RF absorber
Lower coax: 20 ohm 1/4 wave transformer

The impedance plot is shown in Figure 10 for a frequency variation across X-Band. With the above parameters used as nominal values, variations of each parameter was studied to observe the sensitivity of the impedance presented to the diode chip. The effects of structural variations on the electrical performance of the oscillator are summarized in Table 1.

Summary

A method of designing a solid state microwave/millimeter wave oscillator has been explained. A computer program has been developed using the analysis of A.G. Williamson which enables the user to easily manipulate the various parameters of the oscillator. This analysis capability offers an insight to the changes in electrical performance as effected by structural variations. Using this information, sensitive structural parameters can be identified and tightly controlled in a production environment in order to provide consistent performance.

Parameter	Resistance	Reactance
Wg Ht	–0.053%/mil	0.28%/mil
Sliding Short	0.14%/mil	0.026%/mil
Lwr Coax Dia.	0.1%/mil	–0.5%/mil
Post Dia.	0.75%/mil	1.03%/mil
Side Wall Pos	0.02%/mil	0.0013%/mil
Xfmr Imp	5.7%/mil	–0.64%/mil
Xfmr Length	–0.08%/mil	1.55%/mil

Table 1. Sensitivity of resistance and reactance to various waveguide parameters.

WGMOUNT is available through Argus Direct Marketing. To order, see the ad on page 109. *RF*

References

1. A.G. Williamson, "Analysis and Modeling of 'Two-Gap' Coaxial Line Rectangular Waveguide Junctions," IEEE Trans. on Microwave Theory and Techniques, Vol. MTT-31, No. 3, March 1983, pp. 295-302.
2. A.G. Williamson, "The General Cross-Coupled Coaxial-Line/Rectangular Waveguide Junction - Theory and Computer Programs," School of Engineering Report No. 352, May 1984, Dept. of Electrical and Electronic Engineering, University of Auckland, Private Bag, Auckland, New Zealand.
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4. A.G. Williamson, "Analysis of Various Coaxial Line - Rectangular Waveguide Junctions Including Double Coax Entry and/or Waveguide Short Circuit Cases," School of Engineering Report No. 240, Nov. 1980, Dept. of Electrical Engineering, School of Engineering, University of Auckland, Private Bag, Auckland, New Zealand.
5. A.G. Williamson, "Analysis of a Coaxial Line - Rectangular Waveguide Junction," School of Engineering Report No. 236, Dept. of Electrical Engineering, School of Engineering, University of Auckland, Private Bag, Auckland, New Zealand.

About the Author

Roy Monzello received his BS in applied mathematics and his MSEE degrees from California State University, Northridge in 1979 and 1982, respectively. He has worked at Hughes Aircraft, working on IMPATT diode transmitters, and at American Nucleonics, where he was involved in the development of interference cancellation systems. Presently he is employed at California Microwave, Inc., where he is involoved in the development of airborne communication systems. He is a member of the IEEE and a licensed professional engineer in the state of California. He can be reached at 5423 N. Alfonso Dr., Agoura Hills, CA 91301.

ENGINEER'S NOTEBOOK
From the Editors of RF design
Practical Ideas for RF, Analog, Digital and DSP Circuits
Volume 1, Number 1 • June 1995

A Low-Cost, High Performance Mixer for HF Applications

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Republic of South Africa

The performance of a high frequency receiver is mainly determined by the first mixer. For strong-signal handling capabilities, the double balanced diode ring mixer is nearly universally used in high performance receivers. For applications requiring very high dynamic range from the receiver, diode ring mixers driven with up to 0.5 watt (+27 dBm) LO power are used. Power consumption and LO leakage at these LO levels becomes a problem. This article describes a high-performance, low cost double balanced mixer suitable for input frequencies up to and above 20 MHz, with low local oscillator and DC power requirements. This mixer was developed after an idea by J.M. Grundlingh.

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Passive double balanced mixers usually consist of a diode quad with the LO, RF and IF signals coupled to the diodes via two RF transformers (see Figure 1). In the diode ring mixer, the local oscillator signal is split by the RF transformer and fed with a 180° phase difference to the diodes, causing the two pairs of diodes to alternately conduct on the positive and negative half-cycles of the LO signal. The diodes are thus being used as switches. The RF input signal as well as the IF output is coupled to the mixer by another RF transformer. For an ideal double balanced mixer, the only frequency components at the IF port would be ($F_{RF} + F_{LO}$) and ($F_{RF} - F_{LO}$), for the case where F_{RF} is higher than F_{LO}. This switching type of mixer is also known as a chopper modulator [1].

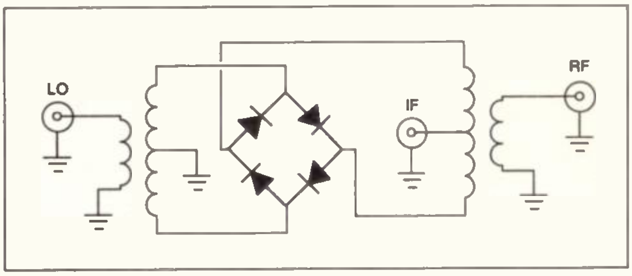

Figure 1. Basic diode ring mixer.

For a diode to work satisfactorily as a switch, the switching signal needs to be much more powerful than the signal being switched. This is the reason for a +27 dBm LO level for high level received signals. There is approximately 0.3 volt drop across the diodes (typically Schottky barrier type) and the transfer function is not very linear. This causes the unwanted mixing products that become a big problem when strong signals are present at the input.

Another Way to Make a Switching Mixer

The CD4066 quad CMOS switch has been used as a chopper modulator for low frequency applications for many years. With the development of High Speed CMOS it is now possible to use these devices up to more than 20 MHz. By replacing the diodes in the ring with 74HC4066 CMOS switches, the performance of the mixer can be improved by the linear transfer function and low ON resistance of the switches at the lower HF frequencies. A schematic diagram of a mixer circuit using the 74HC4066 is shown in Figure 2.

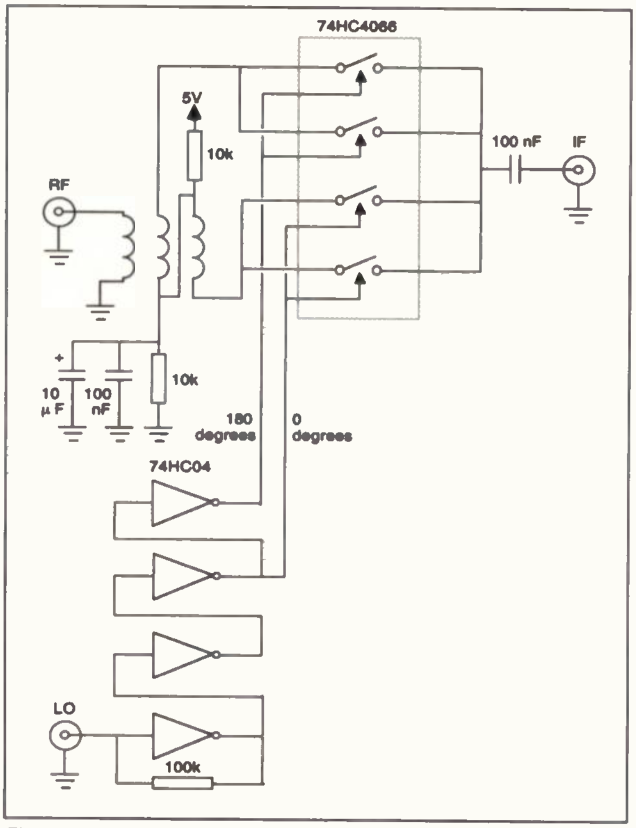

Figure 2. Schematic diagram of the 74HC4066 switching mixer.

Conversion Test Results

F_{LO}(MHz)	F_{IF}(MHz)	F_{RF}(MHz)	Insertion Loss (dB)	P_{1dB}(input) (dBm)	P_{LO} at RF port (dBm)
5.5	9	3.5	7	+15	-32
5.5	9	14.5	6.5	+15	-32
30.7	10.7	20.0	6.5	+15	-20
40.7	10.7	30.0	7.5	+9	-17

Two Tone Test Results

Tone 1 (MHz)	Tone 2 (MHz)	3rd Order Product (MHz)	3rd Order Intercept Point (dBm)
3.6	3.7	3.5, 3.8	+29
14.6	14.7	14.5, 14.8	+31
20.1	20.2	20.0, 20.3	+29
30.1	30.2	30.0, 30.3	+19

7 MHz Direct-Conversion Receiver

FRX(MHz)	RX BW (kHz)	MDS (dBm)	Tone 1 (MHz)	Tone 2 (MHz)	Dynamic Range (dB)
7.020	2.4	-128	7.040	7.060	105

Table 1. Performance test results.

The required 180° phase shift of the LO signal is done in the 74HC04 hex inverter, which also converts the sine wave LO to the square wave required to control the switches in the 74HC4066. One of the RF transformers required by the diode ring mixers is thus replaced with an inexpensive CMOS IC.

The RF input signal is coupled into the mixer with the aid of an RF transformer. A Mini-Circuit T4-1 was used in the example, but a "home made" unit would work as well. Two switches are used in parallel to reduce the ON resistance, with $V_{CC}/2$ DC bias applied via the transformer.

It is very important to terminate the IF port in a wideband 50 ohm load to realize the full performance of the mixer. This is applicable to all diode ring mixers, as noted in [2] and [3].

Performance Measurement Results

The insertion loss, signal handling capability and LO leakage at the RF port of the prototype mixer were evaluated. With the high input impedance presented by the CMOS gate to the oscillator, the mixer is driven with a voltage of a sufficient amplitude to switch the biased gate. For the tests, the output level of the signal generator driving the LO port was set at +7 dBm, which corresponds to a 1.4 V_{pp} signal into 50 ohms.

A direct-conversion receiver covering the 40-meter (7 MHz) amateur band was also constructed to evaluate the noise figure and dynamic range for down conversion to baseband applications. The results of all tests are summarized in Table 1.

It is quite a problem to accurately measure 3rd order intercept point of the mixer. The spurious outputs and wideband noise at high output levels from a synthesized signal generator can easily cause faulty measurements. Ideally, the test signals should be generated by high performance crystal-controlled oscillators.

Application Hints

It is important to note that the mixer is not suitable for transmitting applications without the necessary filtering, due to the presence of harmonics from the square wave LO signal at the IF port.

When using the mixer in a receiver application, the above mentioned point can be utilized for harmonic down conversion (with a reduction in performance compared to fundamental operation). With the use of an appropriate bandpass filter on the RF input, it is possible to build a multi-band receiver without having to switch the LO frequency. This technique is frequently used by microwave engineers. As with all CMOS logic, the DC power requirement of this mixer increases with frequency.

Finally, it is exciting to think of the possibilities of using a 74AC4066, if it became available! (Manufacturers take note.)

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Convert a Low Phase Noise Source to a Low Jitter ADC Clock

Michael Steffes
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Jitter on the convert clock into a high performance analog-to-digital converter (ADC) can quickly degrade the signal-to-noise ratio below the converter's intrinsic performance capabilities. Low phase noise crystal oscillators are often used to control this possible source of performance degradation. To maintain the highest possible level of performance, some means of converting the low level sinusoid out of the crystal oscillator into a square wave with minimal additional jitter is required. Several very low jitter ECL line driver components developed for the serial digital video market work extremely well in this application. For example, the Comlinear CLC006 offers an adjustable output swing, single supply operation with only 2.2 ps noise induced jitter. This note describes a converted circuit using the CLC006.

Operating the CLC006 as a CW to TTL square wave converter simply requires that the common mode voltages be controlled on both the input and output of the device. The input operates best at about 1.6 V below the positive supply rail while the output operates best swinging symmetrically between the two supply pins. Figure 1 shows an example circuit.

R1 and R2 use the positive supply to develop the DC common mode operating voltage at the input. R3 and R4 are used to slightly offset the DC differential input voltage to ensure a fixed output when no input is present. The output swing is maximized and centered between the supplies by R5 and R6. Both input and output are AC coupled with R7 and R8 providing a slight downward level shift at the output to produce the required TTL clock voltage levels. Figure 2 shows tested performance for a 50 MHz, –10dBm (0.2 V_{pp}) input signal.

The scale is for the TTL output swing while the input has been magnified and included to show the actual delay from zero crossing to output transition. Very little change in performance is observed with larger input levels. The output was measured at <1.5 nsec propagation delay and 1.6 nsec rise/fall times for these relatively large TTL levels. Faster performance is possible when producing ECL swings (differential output is available). Designers should note that the minimum required input power increases as the frequency increases due to the rolloff in the open loop gain of the comparator.

Note: The author was employed by Comlinear Corporation at the time this article was written. Comlinear can be reached at: (800) 776-0500.

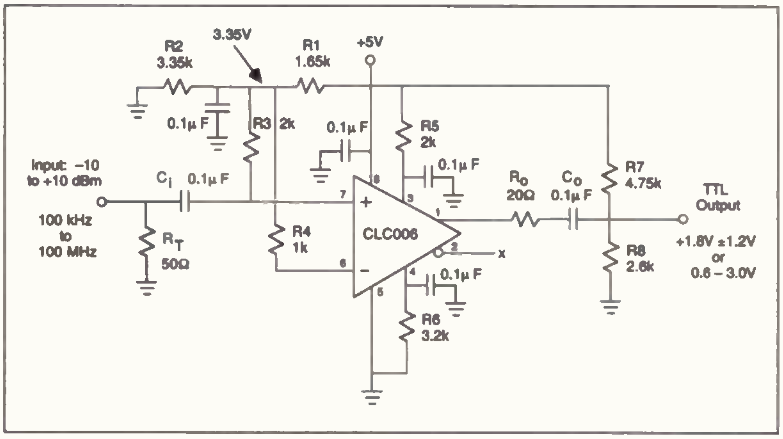

Figure 1. Sinusoid to TTL clock buffer using the CLC006 ECL line driver.

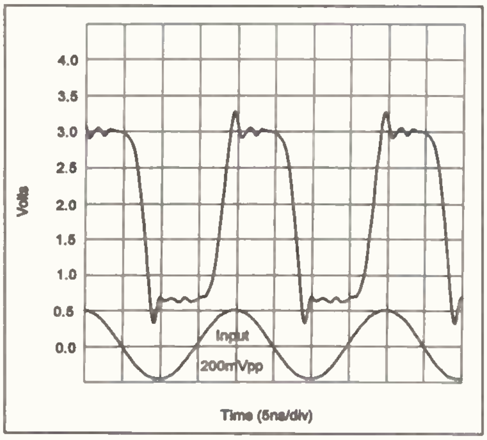

Figure 2. TTL output for a –10dBm, 50 MHz input.

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BOOK REVIEWS

Introduction to Radio Frequency Design

By Wes Hayward

This book is a reprinting of Hayward's 1982 textbook, previously published by Prentice-Hall. This new edition has few revisions to the text, but it now includes a disk with utility software to aid in the design and analysis of LC filters, crystal filters, RF system dynamic range, feedback amplifiers and phase-locked loops. An instruction manual for the software is included as a text file on the disk.

Introduction to Radio Frequency Design has a thorough, yet practical approach. Chapter 1 introduces low-frequency transistor models for bipolars and FETs. Chapters 2 and 3 cover filter basics and coupled-resonator filters, respectively. Transmission line fundamentals are presented in Chapter 4, followed by two-port networks in Chapter 5. These first 5 chapters provide the background that is required before real RF design can be accomplished. The only criticism that might be made is in the completeness of the coverage. A lot of ground is covered in relatively few pages (201 pages for Chapters 1-5). However, the author manages to present a great deal of material clearly and efficiently. Extensive references lead the reader to additional sources when greater depth is needed.

Chapters 6, 7 and 8 cover practical RF circuits and systems. Amplifiers, mixers, oscillators and frequency synthesizers are described thoroughly, and the final chapter uses a receiver as the example of a large RF system, where the elements of design, testing and integration of circuit functions come together.

The best feature of the book is its emphasis on *design*, rather than theory. Every topic is presented as if the immediate goal is to build that circuit. It is clear that the book was written by an engineer (not an academic) for use by other engineers.

383 pages, softbound, with disk. Published by the American Radio Relay League, 1994. ISBN: 0-87259-492-0. List price: $30.00.

Spectrum Guide (Second Edition)

Bennet Z. Kobb

Subtitled: *Radio Frequency Allocations in the United States from 30 MHz to 300 GHz*, this book compiles all Federal Communications Commission (FCC) frequency assignments in a single volume. Author Kobb obtains his information directly from the FCC offices, where he also researches pending actions and rulemakings. With this information, he adds comments to each allocation listing that describe its history, justifications for current usage, and possible future actions.

The second edition will be available in mid-June, updating a successful first edition which was only first published in November 1994! Whether this rapid updating will continue in the future isn't entirely certain, but the author indicates that he intends to revise the book as often as necessary to reflect current spectrum allocation activity.

Est. 240 pages, softbound. Published by New Signals Press, 1995. List price: $35.00.

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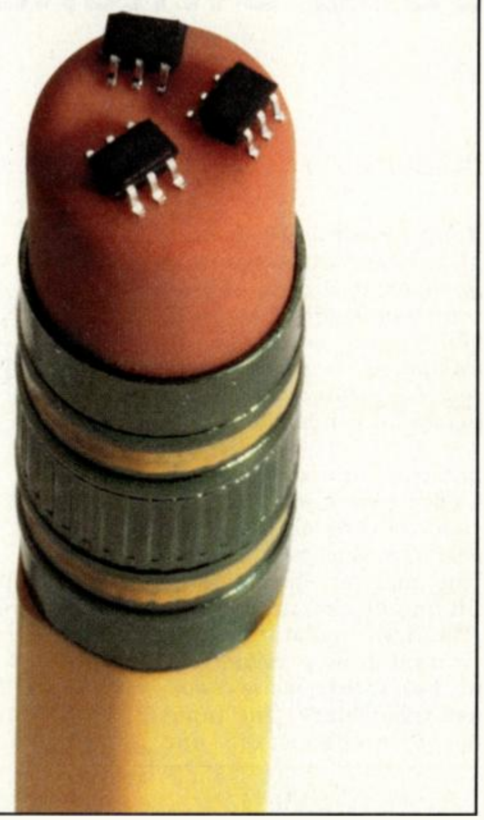

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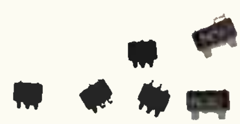
INA-52063	DC-1600	5	30	3.5	20	+17	SOT-363
MGA-86563	500-6000	5	15	1.6	20	+15	SOT-363
MGA-87563	500-4000	3	4.5	1.6	14	+ 8	SOT-363



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Inexpensive Spectrum Analyzer IF Uses Switchable Filtering

By David Tomanek

The goal of this project was to build a simple, high dynamic range spectrum analyzer for home and quasi-professional use, with acceptable frequency range, resolution and good logarithmic/linear amplitude response. Nearly all performance expectations were met with the IF circuit described here.

The main factors driving the design of this spectrum analyzer were simplicity and the availability of key semiconductor components. The architecture of the analyzer, including the various building blocks, is shown in Figure 1. The first oscillator is not a YIG type as used in most commercial instruments, but rather, it uses two 5 GHz bipolar transistors. The input stage uses a diode mixer and microstrip filters on the classic cuprex-card board.

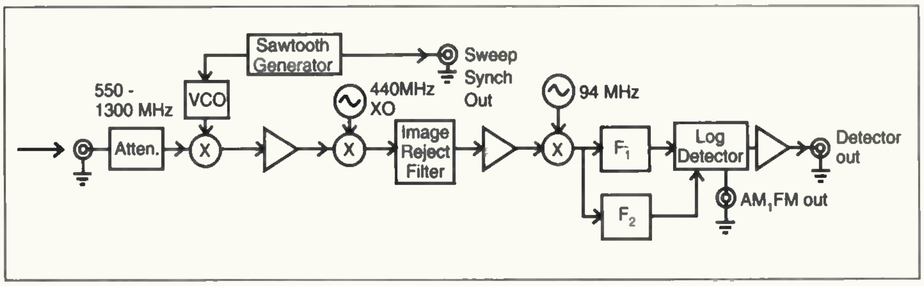

Figure 1. Overall block diagram of the complete spectrum analyzer.

The circuit of the IF is shown in Figure 2, and is the subject of this article. The most interesting and problematic task seemed to be the design of a lin/log detector stage that would have an acceptable dynamic range (not taking noise and dynamic range of previous stages into consideration). The NE614AN in Figure 3 incorporates a log detector with 80 dB range in full bandwidth mode, with just 2 dB error ripple inside this

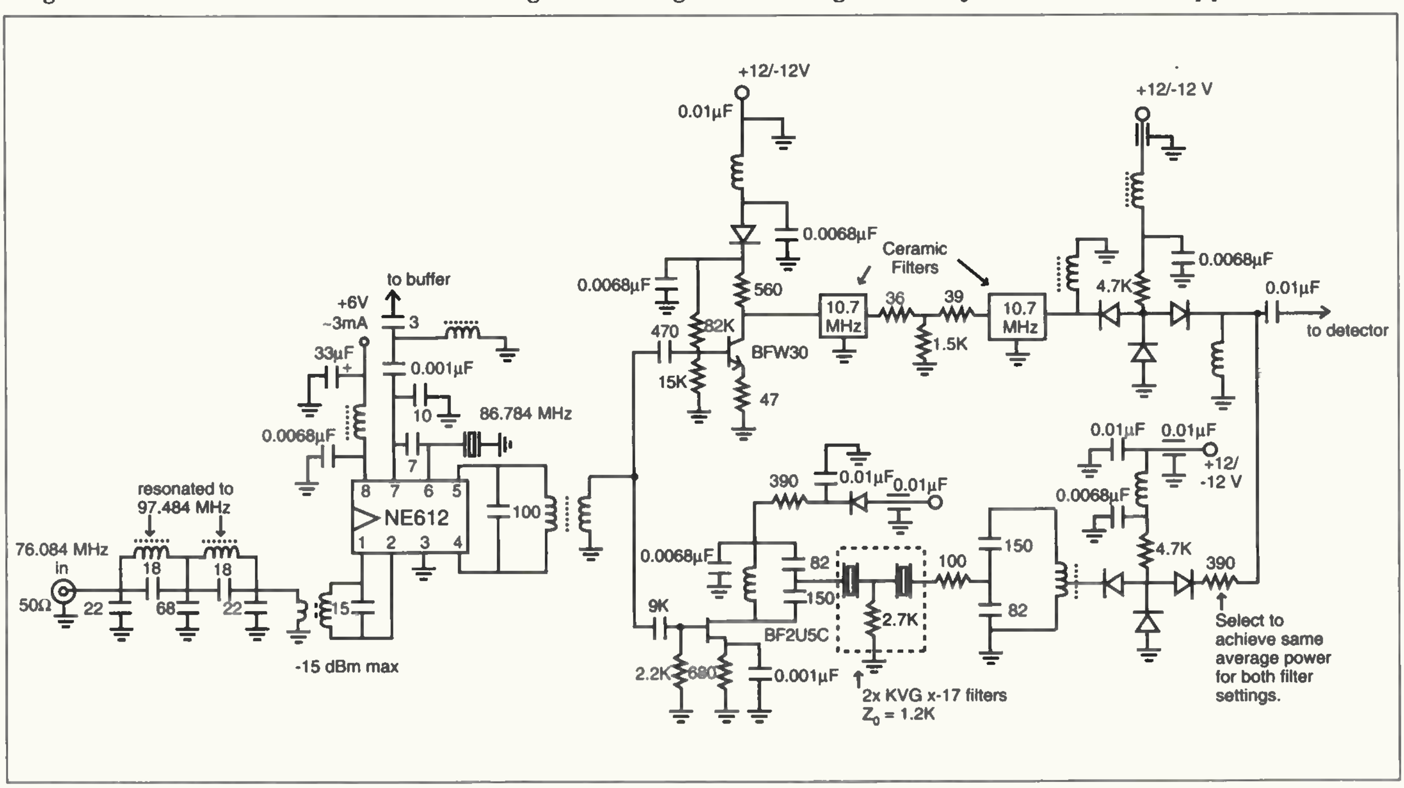

Figure 2. Circuit diagram of the second mixer, IF amplifiers and switchable filters.

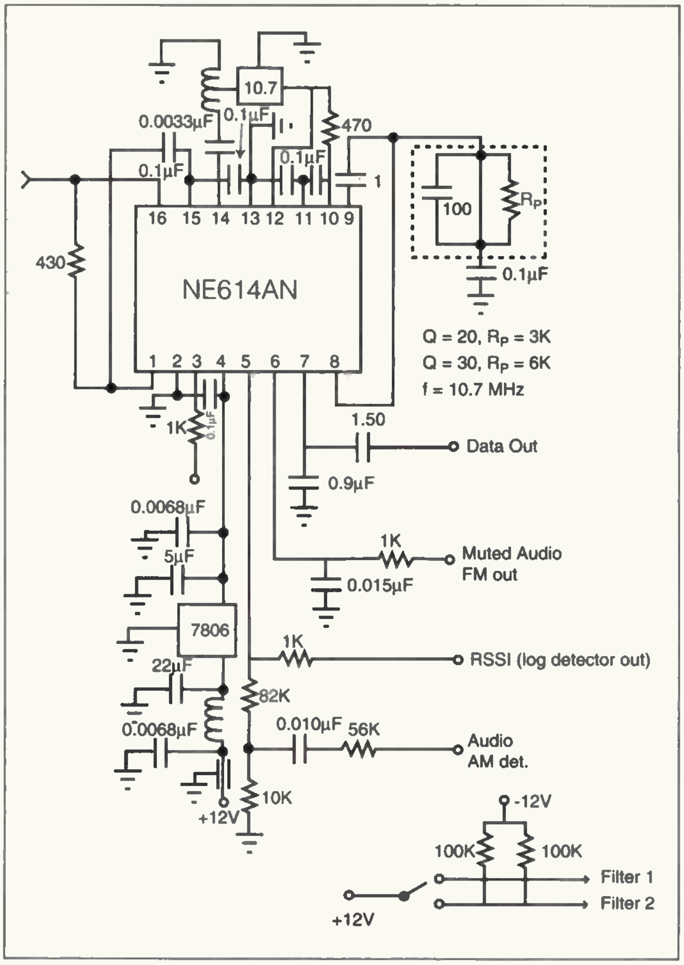

Figure 3. Diagram of the lin/log detector stage, using the Philips NE614AN's RSSI output.

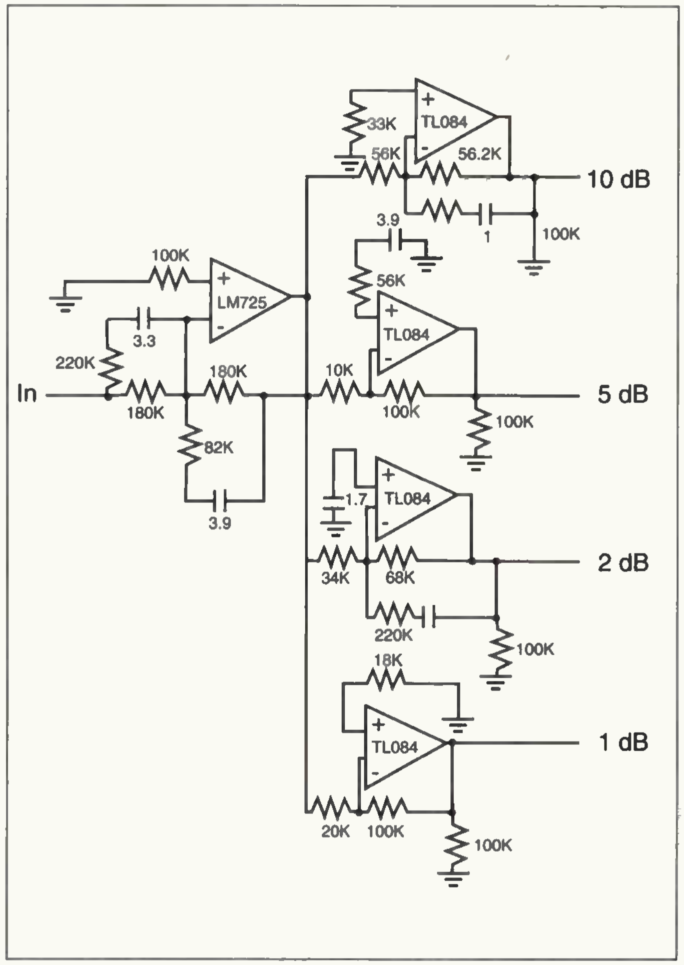

Figure 4. DC amplifiers that provide display scaling and buffering for the high impedance RSSI voltage.

range. Driving a 91 kΩ resistor, and with a 150 kHz bandwidth, the output log sensitivity is 0.5 V/10 dB. Therefore the detector needs high impedance input DC amplifiers to get the 10, 5, and 2 dB scales on the oscilloscope used as a monitor. The log detector in the NE614AN contains a series of limiters with summed output currents. There are two log-limiting stages, with the first stage containing two amplifiers and the second containing three. The first stage has a small signal bandwidth of 41 MHz and the second has a 28 MHz bandwidth. The output of the log detector is independent of frequency, so a filter is required to prevent spurious products and regenerated signals from affecting the logarithmic output. AM and FM audio output are also available from the detector.

Figure 3 shows the DC amplifiers, preceded by an impedance follower, which are used to get the proper voltage ranges for 10, 5, 2 and optional 1 dB per division outputs.

The IF frequency was chosen to be 10.7 MHz to use readily available filters. Preceding the detector is an IF filtering stage with switchable IF bandwidths, as shown in Figure 2. A crystal or ceramic IF filter can be selected, with bandwidths of 9 kHz and 150 kHz, respectively. The pi resistive network near the filters makes the attenuation of the input signal equal for both signal paths. The filters need perfect impedance matching and shielding to eliminate any signal feedthrough at stopband frequencies. On the same board, a mixer such as the NE612 or NE602, both of which use Gilbert cell circuitry and a local oscillator (a third harmonic crystal oscillator) is used as a down converter. Output impedance of the mixer coil is appropriately low.

Using the 9 kHz filter the detector can reach dynamic ranges approaching 90 dB. The noise values at the input port of the logarithmic detector are calculated in the following way:

$$V_{noise} = \sqrt{(4kTBR)}$$

where k is Boltzman's constant, T is temperature in Kelvins, B is the bandwidth of the filter, and R is the source resistance.

$$F(dB) = 10 \log F(kT_0)$$

150 kHz filter: V_{noise1} =

$$4k(290)(150\times10^3)(390) = 0.9679\ \mu V$$

9 kHz filter: V_{noise2} =

$$4k(290)(9\times10^3)(390) = 0.2371\ \mu V$$

The noise figure for the 9 kHz filter is 12.2 dB lower than the 150 kHz filter. It means potential extension of dynamic scale. The input noise figure is approximately 6 dB. Using F (noise factor) we get:

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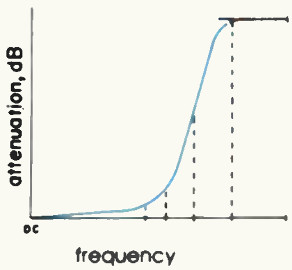

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★LP-10.7	DC-11	19-24	24-200
★LP-21.4	DC-22	32-41	41-200
★LP-30	DC-32	47-61	61-200
★LP-50	DC-48	70-90	90-200
★LP-70	DC-60	90-117	117-300
★LP-90	DC-81	121-157	157-400
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★LP-300	DC-270	410-550	550-1200
★LP-450	DC-400	580-750	750-1800
★LP-550	DC-520	750-920	920-2000
★LP-600	DC-680	840-1120	1120-2000
★LP-750	DC-700	1000-1300	1300-2000
★LP-800	DC-720	1080-1400	1400-2000
★LP-850	DC-780	1100-1400	1400-2000
★LP-1000	DC-900	1340-1750	1750-2000
★LP-1200	DC-1000	1620-2100	2100-2500

All models priced qty. 1-9 ($ea.), Conn. Type P = 11.45, B = 32.95, S = 34.95, N = 35.95
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SCLF-21.4	DC-22	32-41	41-200
SCLF-25	DC-25	36-47	47-200
SCLF-30	DC-30	47-61	61-200
SCLF-45	DC-45	70-90	90-200
SCLF-95	DC-95	146-189	189-400

dc to 1200MHz

SCLF-135	DC-135	210-300	300-600
SCLF-190	DC-190	290-390	390-800
SCLF-225	DC-225	340-440	440-1200
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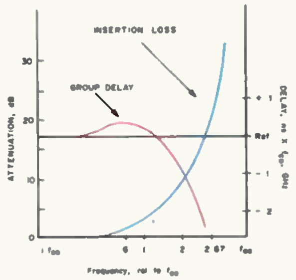
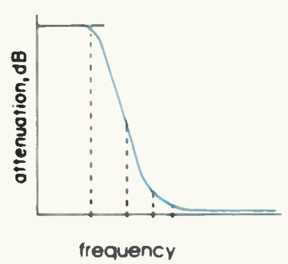


Flat Time Delay, dc to 1870MHz

Model No.	Passband MHz loss < 1.2dB	Stopband MHz loss >10dB	Stopband MHz loss >20dB	VSWR Freq. Range, DC thru 0.2fco X	VSWR Freq. Range, DC thru 0.6fco X	Group Delay Variations, ns Freq. Range, DC thru fco X	Group Delay Variations, ns Freq. Range, DC thru 2fco X	Group Delay Variations, ns Freq. Range, DC thru 2.67fco X
★BLP-39	DC-23	78-117	117	1.3:1	2.3:1	0.70	4.0	5.00
★BLP-117	DC-65	234-312	312	1.3:1	2.4:1	0.35	1.4	1.90
★BLP-156	DC-94	312-416	416	1.3:1	1.1:1	0.30	1.1	1.50
★BLP-200	DC-120	400-534	534	1.6:1	1.9:1	0.40	1.3	1.60
★BLP-300	DC-180	600-801	801	1.25:1	2.2:1	0.20	0.6	0.80
★BLP-467	DC-280	934-1246	1246	1.25:1	2.2:1	0.15	0.4	0.55
▲BLP-933	DC-560	1866-2490	2490	1.3:1	2.2:1	0.09	0.2	0.28
▲BLP-1870	DC-850	3740-5000	5000	1.45:1	2.9:1	0.05	0.1	0.15

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

HIGH PASS



high pass, Plug-in, 13 to 1200MHz

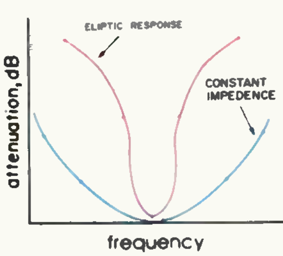
Model No.	Stopband MHz loss > 40dB	Stopband MHz loss > 20dB	Passband, MHz loss < 1dB	VSWR Pass-band Typ.
★HP-25	DC-13	13-19	27.5-200	1.7:1
★HP-50	DC-20	20-26	41-200	1.5:1
★HP-100	DC-40	40-55	90-400	1.5:1
★HP-150	DC-70	70-95	133-600	1.8:1
★HP-175	DC-70	70-105	160-800	1.5:1
★HP-200	DC-90	90-116	185-800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

210 to 2200MHz

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

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

BANDPASS



bandpass, Elliptic Response, 10.7 to 70MHz

Model No.	Center Freq. (MHz)	Passband I.L. 1.5 dB Max. (MHz)	3 dB Bandwidth Typ. (MHz)	Stopbands I.L. > 20dB at MHz	Stopbands I.L. > 35dB at MHz
★BP-10.7	10.7	9.5-11.5	8.9-12.7	7.5 & 15	0.6 & 50-1000
★BP-21.4	21.4	19.2-23.6	17.9-25.3	15.5 & 29	3.0 & 80-1000
★BP-30	30.0	27.0-33.0	25-35	22 & 40	3.2 & 99-1000
★BP-60	60.0	55.0-67.0	49.8-70.5	44 & 79	4.6 & 190-1000
★BP-70	70.0	63.0-77.0	58.0-82.0	51 & 94	6.0 & 193-1000

Price, (1-9 qty), all models: plug-in $18.95, BNC $40.95, SMA $42.95, Type N $43.95

Constant Impedance, 21.4 to 70MHz

Model No.	Center Freq. MHz	Passband MHz loss < 1dB	Stopband loss > 20dB at MHz	VSWR 1:3:1 Total Band MHz
★IF-21.4	21.4	18-25	1.3 & 150	DC-220
★IF-30	30.0	25-35	1.9 & 210	DC-330
★IF-40	42.0	35-49	2.6 & 300	DC-400
★IF-50	50.0	41-58	3.1 & 350	DC-440
★IF-60	60.0	50-70	3.8 & 400	DC-500
★IF-70	70.0	58-82	4.4 & 490	DC-550

Price, (1-9 qty), all models: plug-in $14.95, BNC $36.95, SMA $38.95, Type N $39.95

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

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$V_{noise} = \sqrt{(4kTBR_n)}$

therefore,

$V_{noise1} = 1.7212\ \mu V$
$= -111.2\text{ dBm}/390\text{ ohm}$

$V_{noise2} = 0.4216\ \mu V = -123.4\text{ dBm}$

Summary

The circuits used in the IF of a low-cost spectrum analyzer have been described, the key element being the log detector. The detector used is a multi-purpose unit with good repeatability. With the simple IC downconversion mixers and switchable filters, good dynamic range and good accuracy have been obtained. It is hoped that such a simple unit can enhance every bench. *RF*

About the Author

David Tomanek can be reached at: Bubenecska 27, Praha 6, 160 00, Czech Republic. This circuit was an entry in the 1994 RF Design Awards contest.

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Phase Noise: 10 Hz	<-120 dBc/Hz	<-125 dBc/Hz	<-130 dBc/Hz
100 Hz	<-150 dBc/Hz	<-150 dBc/Hz	<-150 dBc/Hz
1 kHz	<-155 dBc/Hz	<-155 dBc/Hz	<-155 dBc/Hz
Operating Range	0° to 50°C	-10° to +60°C	-20° to +70°C
Allan Variance (1 second)	$<1\times10^{-11}$	$<5\times10^{-12}$	$<2\times10^{-12}$

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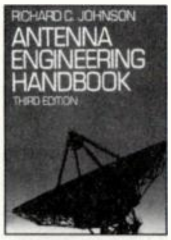

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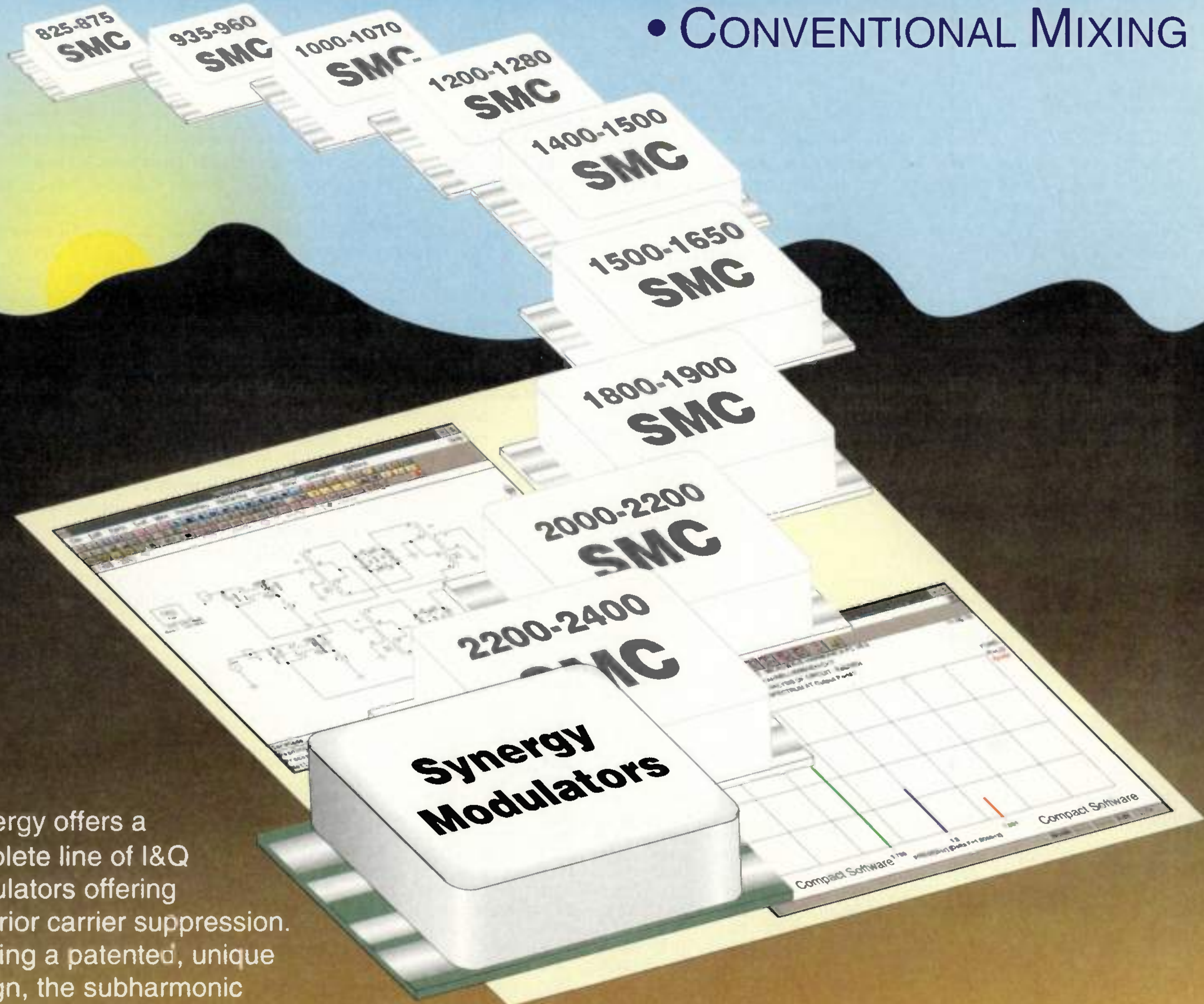

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INFO/CARD 66

Intermodulation Distortion in a Multi-Signal Environment

By Michael Leffel
Motorola, Inc.

This article provides a discussion on intermodulation caused by multiple tones. The discussion focuses on predicting the behavior of the device with multiple tones at its input by using the well known two-tone intermodulation intercept parameters. Two-tone intermodulation distortion is a well documented and easily understood phenomena. Using the equations that relate intercept point to intermodulation rejection (IMR) for two-tones is easily done.

Example One

A device with an input intercept point of 20 dBm and two input tones at 800 and 801 MHz will generate third-order intermodulation at 799 and 802 MHz. If the two input tones have an amplitude of –20 dBm, the resulting IMR is 80 dB. This follows from the equation:

$$\text{IMR} = \frac{2}{3}(\text{IPi} - \text{TonePwr} + \text{IMR}) \qquad (1)$$

$$\frac{1}{3}\text{IMR} = \frac{2}{3}(\text{IPi} - \text{TonePwr})$$

$$\text{IMR} = 2(\text{IPi} - \text{TonePwr}) = 2(20 - (-20)) = 80\text{dB}$$

A more interesting case involves multiple tones, and the effects they have on the apparent intercept point of the receiver. Consider evenly spaced tones (spaced by Δ), since their IM products can be additive at some frequencies and will be the theoretical worse case. These tones will generate third-order intermodulation (as well as higher order IM products – but assumed to be lower in amplitude than any of the third-order products, and therefore ignored for the purposes of this discussion). Third-order intermodulation products will be present at the carrier tone frequencies, as well as evenly spaced above and below the tones. The highest amplitude intermodulation products will be the products one Δ above or below the tones for a set of evenly spaced tones, or the product that is in between the tones if one of the tones is missing, ignoring products that are coincident with a tone. The worst possible IM product will be one that is centered between tones with the same number of evenly spaced tones above and below it in frequency.

To handle cases with multiple tones while still using the normal intercept point equations, this article derives the expected increase in IM power due to additional tones over

M\N	0	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18
0 (in dB)			0	6.99	10	12.6	14.3	15.9	17.2	18.3	19.3	20.2	21	21.8	22.4	23.1	23.7	24.2	24.7
1			6.99	11.1	13.4	15.3	16.7	18	19	20	20.8	21.6	22.3	23	23.6	24.1	24.7	25.2	25.6
2	0	6.99	**10**	13.4	15.4	17.1	18.3	19.4	20.4	21.2	22	22.7	23.3	23.9	24.5	25	25.5	25.9	26.4
3	6.99	11.1	13.4	**15.3**	17.1	18.5	19.6	20.6	21.5	22.3	23	23.6	24.2	24.8	25.3	25.8	26.2	26.6	27
4	10	13.4	15.4	17.1	**18.3**	19.6	20.7	21.6	22.4	23.1	23.8	24.4	24.9	25.5	25.9	26.4	26.8	27.2	27.6
5	12.6	15.3	17.1	18.5	19.6	**20.6**	21.6	22.5	23.2	23.9	24.5	25.1	25.6	26.1	26.6	27	27.4	27.8	28.2
6	14.3	16.7	18.3	19.6	20.7	21.6	**22.4**	23.2	23.9	24.6	25.2	25.7	26.2	26.7	27.1	27.5	27.9	28.3	28.6
7	15.9	18	19.4	20.6	21.6	22.5	23.2	**23.9**	24.6	25.2	25.8	26.3	26.7	27.2	27.6	28	28.4	28.8	29.1
8	17.2	19	20.4	21.5	22.4	23.2	23.9	24.6	**25.2**	25.8	26.3	26.8	27.2	27.7	28.1	28.5	28.8	29.2	29.5
9	18.3	20	21.2	22.3	23.1	23.9	24.6	25.2	25.8	**26.3**	26.8	27.3	27.7	28.1	28.5	28.9	29.2	29.6	29.9
10	19.3	20.8	22	23	23.8	24.5	25.2	25.8	26.3	26.8	**27.2**	27.7	28.1	28.5	28.9	29.3	29.6	30	30.3
11	20.2	21.6	22.7	23.6	24.4	25.1	25.7	26.3	26.8	27.3	27.7	**28.1**	28.5	28.9	29.3	29.7	30	30.3	30.6
12	21	22.3	23.3	24.2	24.9	25.6	26.2	26.7	27.2	27.7	28.1	28.5	**28.9**	29.3	29.7	30	30.3	30.6	30.9
13	21.8	23	23.9	24.8	25.5	26.1	26.7	27.2	27.7	28.1	28.5	28.9	29.3	**29.7**	30	30.3	30.7	31	31.2
14	22.4	23.6	24.5	25.3	25.9	26.6	27.1	27.6	28.1	28.5	28.9	29.3	29.7	30	**30.3**	30.7	31	31.3	31.5
15	23.1	24.1	25	25.8	26.4	27	27.5	28	28.5	28.9	29.3	29.7	30	30.3	30.7	**31**	31.3	31.5	31.8
16	23.7	24.7	25.5	26.2	26.8	27.4	27.9	28.4	28.8	29.2	29.6	30	30.3	30.7	31	31.3	**31.5**	31.8	32.1
17	24.2	25.2	25.9	26.6	27.2	27.8	28.3	28.8	29.2	29.6	30	30.3	30.6	31	31.3	31.5	31.8	**32.1**	32.3
18	24.7	25.6	26.4	27	27.6	28.2	28.6	29.1	29.5	29.9	30.3	30.6	30.9	31.2	31.5	31.8	32.1	32.3	**32.6**
19	25.2	26.1	26.8	27.4	28	28.5	29	29.4	29.8	30.2	30.6	30.9	31.2	31.5	31.8	32.1	32.3	32.6	32.8
20	25.7	26.5	27.2	27.8	28.4	28.9	29.3	29.7	30.1	30.5	30.8	31.2	31.5	31.8	32.1	32.3	32.6	32.8	33.1
21	26.1	26.9	27.6	28.2	28.7	29.2	29.6	30	30.4	30.8	31.1	31.4	31.7	32	32.3	32.6	32.8	33.1	33.3
22	26.5	27.3	27.9	28.5	29	29.5	29.9	30.3	30.7	31.1	31.4	31.7	32	32.3	32.6	32.8	33.1	33.3	33.5
23	26.9	27.7	28.3	28.8	29.3	29.8	30.2	30.6	31	31.3	31.6	32	32.2	32.5	32.8	33	33.3	33.5	33.8
24	27.3	28	28.6	29.1	29.6	30.1	30.5	30.9	31.2	31.6	31.9	32.2	32.5	32.8	33	33.3	33.5	33.7	34
25	27.7	28.4	28.9	29.4	29.9	30.4	30.8	31.1	31.5	31.8	32.1	32.4	32.7	33	33.2	33.5	33.7	34	34.2
26	28	28.7	29.2	29.7	30.2	30.6	31	31.4	31.7	32.1	32.4	32.7	32.9	33.2	33.4	33.7	33.9	34.2	34.4
27	28.4	29	29.5	30	30.5	30.9	31.3	31.6	32	32.3	32.6	32.9	33.1	33.4	33.7	33.9	34.1	34.3	34.6
28	28.7	29.3	29.8	30.3	30.7	31.1	31.5	31.9	32.2	32.5	32.8	33.1	33.4	33.6	33.9	34.1	34.3	34.5	34.7
29	29	29.6	30.1	30.6	31	31.4	31.8	32.1	32.4	32.7	33	33.3	33.6	33.8	34.1	34.3	34.5	34.7	34.9
30	29.3	29.9	30.4	30.8	31.2	31.6	32	32.3	32.6	32.9	33.2	33.5	33.8	34	34.2	34.5	34.7	34.9	35.1
31	29.6	30.1	30.6	31.1	31.5	31.9	32.2	32.5	32.8	33.1	33.4	33.7	33.9	34.2	34.4	34.7	34.9	35.1	35.3
32	29.9	30.4	30.9	31.3	31.7	32.1	32.4	32.7	33.1	33.3	33.6	33.9	34.1	34.4	34.6	34.8	35	35.2	35.4

Table 2. Power increase (in dB) relative to two-tone IM for M evenly spaced tones on one side of the gap, and

the two tone case. The results of this calculation will allow the use of the commonly specified third-order intercept point to calculate the "expected" intermodulation rejection for any number of tones.

There are multiple ways that IM is generated at a particular frequency under the conditions of more than two-tones present at the input to a system, and more than one IM product will be produced at some frequencies. When multiple IM products are present at a particular frequency, they will combine. The resultant is a function of the amplitude and phase of the individual IM products. If the IM products add in phase, the resultant will be maximized. If they add out of phase, they will partially cancel, minimizing the resultant. Example two will serve to illustrate this point.

Example Two

Calculate the increase in power (in dB with respect to $A\cos(\omega t+\phi_1)$ at the frequency ω due to the addition of $2A\cos(\omega t+\phi_2)$ to $A\cos(\omega t+\phi_1)$.

For $\phi_1 = \phi_2 - \pi$ the RMS power of $A\cos(wt+f_1)$ into 1 ohm is the reference power level:

$$\left(\frac{A}{\sqrt{2}}\right)^2 = \frac{A^2}{2} \qquad (2)$$

The RMS power of $A\cos(\omega t+\phi_1)$ $+2A\cos(\omega t+\phi_2) = A\cos(\omega t+\phi_2)$ is:

$$\left(\frac{A}{\sqrt{2}}\right)^2 = \frac{A^2}{2} \qquad (3)$$

and the power level is the same, for a 0 dB increase in power.

For $\phi_1 = \phi_2$ the RMS power of $A\cos(\omega t+\phi_1)$ + $2A\cos(\omega t+\phi_2)$ = $3A\cos(\omega t+\phi_1)$ is:

$$\left(\frac{3A}{\sqrt{2}}\right)^2 = \frac{9A^2}{2} \qquad (4)$$

and the power level increased by 9 times, or by 9.5 dB.

For any other relationship such that $0<|\phi_1-\phi_2|<\pi$, the power increase will be somewhere between 0 and 9.5 dB.

For ϕ_1 and ϕ_2 identically and independently distributed, (i.i.d.) with uniform distribution over 0 to 2π, the expected RMS power of $A\cos(\omega t+\phi_1)$ is:

$$E\{A\cos(\omega t+\phi)\} = \frac{A^2}{2} \qquad (5)$$

the expected RMS power of $2A\cos(\omega t+\phi_2)$ is:

$$E\{2A\cos(\omega t+\phi)\} = \frac{4A^2}{2} \qquad (6)$$

and the expected power of $A\cos(\omega t+\phi_1)$ $+2A\cos(\omega t+\phi_2)$ is:

$$\begin{aligned} &E\{A\cos(\omega t+\phi_1)+2A\cos(\omega t+\phi_2)\} \qquad (7)\\ &= E\{A\cos(\omega t+\phi_1)\}\\ &+E\{2A\cos(\omega t+\phi_2)\}\end{aligned}$$

Since ϕ_1 and ϕ_2 are independent, this yields an expected power of:

$$\frac{A^2}{2}+\frac{4A^2}{2}=\frac{5A^2}{2} \qquad (8)$$

which is five times the power of $A\cos(\omega t+\phi_1)$.

When converted to dB this power increase of five times equals 10log(5) = 7 dB.This will be denoted the expected power of the summation of two cosines.

Note that this is the expected increase in power (above $A\cos(\omega t+\phi_1)$ as a reference) for the summation of two cosine signals with independent, random phase. Also note that the peak power level will actually be 9.5 dB higher than $A\cos(\omega t+\phi_1)$ when the cosines add in phase, and the minimum power level can be shown to be the same power level as the original $A\cos(\omega t+\phi_1)$ signal.

Throughout the rest of the article, the assumption of random phase will be made, but the explicit notation (ϕ) will be dropped.

The above technique can be used to calculate the expected power level of the combination of two IM products at a single frequency. This can be used to calculate the expected IM level of multiple tones applied to a linear system. The next logical case to analyze will be the IM caused by three equally spaced, equal amplitude tones. This case will be compared to the two-tone case to see how IM is affected by the addition of a third tone. First the two-tone case. Using the MacLaurin series expansion:

$$P_o = K_0 + K_1P_i + K_2P_i^2 + K_3P_i^3 + \ldots \qquad (9)$$

and assuming that the K_n coefficients are non-zero and decreasing for increasing n when n > 1, the output signal P_o can be calculated for an input signal P_i.

For two tones:

19	20	21	22	23	24	25	26	27	28	29	30	31	32
25.2	25.7	26.1	26.5	26.9	27.3	27.7	28	28.4	28.7	29	29.3	29.6	29.9
26.1	26.5	26.9	27.3	27.7	28	28.4	28.7	29	29.3	29.6	29.9	30.1	30.4
26.8	27.2	27.6	27.9	28.3	28.6	28.9	29.2	29.5	29.8	30.1	30.4	30.6	30.9
27.4	27.8	28.2	28.5	28.8	29.1	29.4	29.7	30	30.3	30.6	30.8	31.1	31.3
28	28.4	28.7	29	29.3	29.6	29.9	30.2	30.5	30.7	31	31.2	31.5	31.7
28.5	28.9	29.2	29.5	29.8	30.1	30.4	30.6	30.9	31.1	31.4	31.6	31.8	32.1
29	29.3	29.6	29.9	30.2	30.5	30.8	31	31.3	31.5	31.8	32	32.2	32.4
29.4	29.7	30	30.3	30.6	30.9	31.1	31.4	31.6	31.9	32.1	32.3	32.5	32.7
29.8	30.1	30.4	30.7	31	31.2	31.5	31.7	32	32.2	32.4	32.6	32.8	33.1
30.2	30.5	30.8	31.1	31.3	31.6	31.8	32.1	32.3	32.5	32.7	32.9	33	33.3
30.6	30.8	31.1	31.4	31.6	31.9	32.1	32.4	32.6	32.8	33	33.2	33.3	33.6
30.9	31.2	31.4	31.7	32	32.2	32.4	32.7	32.9	33.1	33.3	33.5	33.5	33.9
31.2	31.5	31.7	32	32.2	32.5	32.7	32.9	33.1	33.4	33.6	33.8	33.8	34.1
31.5	31.8	32	32.3	32.5	32.8	33	33.2	33.4	33.6	33.8	34	34	34.4
31.8	32.1	32.3	32.6	32.8	33	33.2	33.4	33.7	33.9	34.1	34.2	34.2	34.6
32.1	32.3	32.6	32.8	33	33.3	33.5	33.7	33.9	34.1	34.3	34.5	34.4	34.8
32.3	32.6	32.8	33.1	33.3	33.5	33.7	33.9	34.1	34.3	34.5	34.7	34.6	35
32.6	32.8	33.1	33.3	33.5	33.7	34	34.2	34.3	34.5	34.7	34.9	34.7	35.2
32.8	33.1	33.3	33.5	33.8	34	34.2	34.4	34.6	34.7	34.9	35.1	34.9	35.4
33.1	33.3	33.5	33.8	34	34.2	34.4	34.6	34.8	35	35.1	35.3	35.1	35.6
33.3	**33.5**	33.8	34	34.2	34.4	34.6	34.8	35	35.1	35.3	35.5	35.2	35.8
33.5	33.8	**34**	34.2	34.4	34.6	34.8	35	35.2	35.3	35.5	35.7	35.4	36
33.8	34	34.2	**34.4**	34.6	34.8	35	35.2	35.4	35.5	35.7	35.9	35.5	36.2
34	34.2	34.4	34.6	**34.8**	35	35.2	35.4	35.5	35.7	35.9	36	35.7	36.4
34.2	34.4	34.6	34.8	35	**35.2**	35.4	35.5	35.7	35.9	36	36.2	35.8	36.5
34.4	34.6	34.8	35	35.2	35.4	**35.5**	35.7	35.9	36.1	36.2	36.4	36	36.7
34.6	34.8	35	35.2	35.4	35.5	35.7	**35.9**	36.1	36.2	36.4	36.5	36.1	36.8
34.8	35	35.2	35.4	35.5	35.7	35.9	36.1	**36.2**	36.4	36.5	36.7	36.2	37
35	35.1	35.3	35.5	35.7	35.9	36.1	36.2	36.4	**36.5**	36.7	36.9	36.3	37.2
35.1	35.3	35.5	35.7	35.9	36	36.2	36.4	36.5	36.7	**36.9**	37	36.5	37.3
35.3	35.5	35.7	35.9	36	36.2	36.4	36.5	36.7	36.9	37	**37.2**	36.6	37.4
35.5	35.7	35.8	36	36.2	36.4	36.5	36.7	36.9	37	37.2	37.3	**37.4**	37.6
35.6	35.8	36	36.2	36.4	36.5	36.7	36.8	37	37.2	37.3	37.4	37.6	**37.7**

N evenly spaced tones on the other side of the gap.

$$P_i = A_1 \cos(\omega_1 t + \phi_1) + A_2 \cos(\omega_2 t + \phi_2) \tag{10}$$

For the worst case, several assumptions must be made: $A_1 = A_2 = A$, equal amplitude tones, and ϕ_1 and ϕ_2 are independent and identically distributed with a uniform distribution over 0 to 2π.

Substituting (10) into (9) yields the output signal:

$$\begin{aligned} P_0 &= K_0 + K_1A(\cos\omega_1 + \cos\omega_2) \\ &+K_2A^2(\cos\omega_1 + \cos\omega_2)^2 \\ &+K_3A^3(\cos\omega_1 + \cos\omega_2)^3 + \cdots \end{aligned} \tag{11a}$$

$$\begin{aligned} P_0 &= K_0 + K_1A(\cos\omega_1 + \cos\omega_2) \\ &+K_2A^2\left(\cos^2\omega_1 + 2\cos\omega_1\cos\omega_2 + \cos^2\omega_2\right) \\ &+K_3A^3\left(\cos^3\omega_1 + 3\cos^2\omega_1\cos\omega_2 \right.\\ &\left.+3\cos\omega_1\cos^2\omega_2 + \cos^3\omega_2\right) + \cdots \end{aligned} \tag{11b}$$

Then using the identity:

$$\begin{aligned} \cos(a)\cos(b) &= \frac{1}{2}\cos(a+b) + \frac{1}{2}\cos(a-b) = \\ &\frac{1}{2}\cos(b+a) + \frac{1}{2}\cos(b-a) \end{aligned} \tag{12}$$

and expanding:

$$\begin{aligned} P_0 &= K_0 + K_1A(\cos\omega_1 + \cos\omega_2) \\ &+K_2A^2\left(1 + \frac{1}{2}\cos 2\omega_1 + \cos(\omega_2 - \omega_1) \right.\\ &\left.+\cos(\omega_2 + \omega_1) + \frac{1}{2}\cos 2\omega_2\right) \\ &+K_3A^3\left(\frac{3}{4}\cos\omega_1 + \frac{1}{4}\cos 3\omega_1 + \frac{1}{4}\cos 3\omega_2 + \frac{3}{4}\cos\omega_2 \right.\\ &+\frac{3}{2}\cos\omega_1 + \frac{3}{4}\cos(2\omega_1 - \omega_2) + \frac{3}{4}\cos(2\omega_2 - \omega_1) \\ &\left.+\frac{3}{4}\cos(2\omega_1 + \omega_2) + \frac{3}{4}\cos(2\omega_2 + \omega_1) + \frac{3}{2}\cos\omega_2\right) + \cdots \end{aligned} \tag{13}$$

Examining the third-order case and ignoring the input tones, second-order IM, third-order IM around the tonal harmonics, and IM higher than third-order leaves:

$$\begin{aligned} P_{IM} &= K_3A^3\left(\frac{3}{4}\cos(2\omega_1 - \omega_2) + \frac{3}{4}\cos(2\omega_2 - \omega_1) \right.\\ &\left.+\frac{9}{4}\cos\omega_1 + \frac{9}{4}\cos\omega_2\right) \end{aligned} \tag{14}$$

and by defining $w_2 - w_1 = \Delta$, the equation simplifies to:

$$\begin{aligned} P_{IM} &= K_3A^3\left(\frac{3}{4}\cos(\omega_1 - \Delta) + \frac{3}{4}\cos(\omega_2 + \Delta) \right.\\ &\left.+\frac{9}{4}\cos\omega_1 + \frac{9}{4}\cos\omega_2\right) \end{aligned} \tag{15}$$

This equation shows the source of the two third-order IM products that are always seen above and below two-tones. Because the amplitude of these products is proportional to the cube of the amplitude of the input tones, a doubling of the amplitude of the input tones results in an eight-times increase in the amplitude of the IM products. In terms of dB, a 3 dB increase in the level of the input tones results in a 9 dB increase in the level of the IM products, and of course, the point where these lines intercept is called the third-order intercept point. These are the well understood properties of two-tone third-order IM, and will be used as a reference point for the rest of the derivations.

Now the three-tone case. Repeating the above calculations for

$$P_i = A(\cos\omega_1 t + \cos\omega_2 t + \cos\omega_3 t) \tag{16}$$

substituting into (9):

$$\begin{aligned} P_0 &= K_0 + K_1A(\cos\omega_1 + \cos\omega_2 + \cos\omega_3) \\ &+K_2A^2(\cos\omega_1 + \cos\omega_2 + \cos\omega_3)^2 \\ &+K_3A^3(\cos\omega_1 + \cos\omega_2 + \cos\omega_3)^3 + \cdots \end{aligned} \tag{17}$$

$$\begin{aligned} P_0 &= K_0 + K_1A(\cos\omega_1 + \cos\omega_2 + \cos\omega_3) \\ &+K_2A^2\left(\cos^2\omega_1 + 2\cos\omega_1\cos\omega_2 + \cos^2\omega_2 + 2\cos\omega_2\cos\omega_3 \right.\\ &\left.+\cos^2\omega_3 + 2\cos\omega_1\cos\omega_3\right) \\ &+K_3A^3\left(\cos^3\omega_1 + 3\cos^2\omega_1\cos\omega_2 + 3\cos\omega_1\cos^2\omega_2\right. \end{aligned} \tag{18}$$

and expanding:

$$\begin{aligned} P_0 &= K_0 + K_1A(\cos\omega_1 + \cos\omega_2 + \cos\omega_3) \\ &+K_2A^2\left(\frac{3}{2} + \cos 2\omega_1 + \cos(\omega_2 - \omega_1) + \cos(\omega_1 + \omega_2) \right.\\ &+\cos 2\omega_2 + \cos(\omega_3 - \omega_2) + \cos(\omega_2 + \omega_3 \\ &\left.+\cos 2\omega_3 + \cos(\omega_3 - \omega_1)\cos\omega_1\omega_3 - \right) \\ &+K_3A^3\left(\frac{3}{4}\cos\omega_1 + \frac{1}{4}\cos 3\omega_1 + \frac{3}{2}\cos\omega_2 + \frac{3}{4}\cos(2\omega_1 - \omega_2)\right. \\ &+\frac{3}{4}\cos(2\omega_1 + \omega_2) + \frac{3}{2}\cos\omega_1 + \frac{3}{4}\cos(2\omega_2 - \omega_1) \\ &+\frac{3}{4}\cos(\omega_1 + 2\omega_2) + \frac{3}{4}\cos\omega_2 + \frac{1}{4}\cos 3\omega_2 \\ &+\frac{3}{2}\cos\omega_3 + \frac{3}{4}\cos(2\omega_2 - \omega_3) + \frac{3}{4}\cos(2\omega_2 + \omega_3) \\ &+\frac{3}{2}\cos\omega_2 + \frac{3}{4}\cos(2\omega_3 - \omega_2) + \frac{3}{4}\cos(\omega_2 + 2\omega_3) \\ &+\frac{3}{4}\cos\omega_3 + \frac{1}{4}\cos 3\omega_3 + \frac{3}{2}\cos\omega_1 + \frac{3}{4}\cos(2\omega_3 - \omega_1) \\ &+\frac{3}{4}\cos(2\omega_3 + \omega_1) + \frac{3}{2}\cos\omega_3 + \frac{3}{4}\cos(2\omega_1 - \omega_3) \\ &+\frac{3}{4}\cos(\omega_3 + 2\omega_1) + \frac{3}{2}\cos(\omega_1 + \omega_2 + \omega_3) \end{aligned} \tag{19}$$

$$+\frac{3}{2}\cos(\omega_1+\omega_2-\omega_3)$$

$$+\frac{3}{2}\cos(\omega_1-\omega_2+\omega_3)$$

$$+\frac{3}{2}\cos(-\omega_1+\omega_2+\omega_3)\Bigg)$$

Again, examining only the third-order case and ignoring the input tones, second-order IM, third-order IM around the tonal harmonics, and IM higher than third-order leaves:

$$P_{IM}=K_3A^3\Bigg(\frac{15}{4}\cos\omega_1+\frac{15}{4}\cos\omega_2 \qquad (20)$$

$$+\frac{15}{4}\cos\omega_3+\frac{3}{4}\cos(2\omega_1-\omega_2)$$

$$+\frac{3}{4}\cos(2\omega_2-\omega_1)+\frac{3}{4}\cos(2\omega_1+\omega_3)$$

$$+\frac{3}{4}\cos(2\omega_3-\omega_1)+\frac{3}{4}\cos(2\omega_3-\omega_2)$$

$$+\frac{3}{4}\cos(2\omega_2-\omega_3)$$

$$+\frac{3}{2}\cos(\omega_1+\omega_2-\omega_3)$$

$$+\frac{3}{2}\cos(\omega_1-\omega_2+\omega_3)$$

$$+\frac{3}{2}\cos(\omega_2+\omega_3-\omega_1)\Bigg)$$

To simplify the analysis further, the worst case situation will be analyzed, restricting the tones to even spacing: $(\omega_3-\omega_2)=(\omega_2-\omega_1)=\Delta$.

$$P_{IM}=K_3A^3\Bigg(\frac{15}{4}\cos\omega_1 \qquad (21)$$

$$+\frac{15}{4}\cos\omega_2+\frac{15}{4}\cos\omega_3$$

$$+\frac{3}{4}\cos(\omega_1-\Delta)+\frac{3}{4}\cos(\omega_3+\Delta)$$

$$+\frac{3}{4}\cos(\omega_1-2\Delta)+\frac{3}{4}\cos(\omega_3+2\Delta)$$

$$+\frac{3}{4}\cos\omega_1+\frac{3}{4}\cos\omega_3$$

$$+\frac{3}{2}\cos(\omega_1-\Delta)+\frac{3}{2}\cos(\omega_3+\Delta)$$

$$+\frac{3}{2}\cos\omega_2\Bigg)$$

Collecting all two- and three-tone, third-order IM products yields:

$$\frac{18}{4}\cos\omega_1+\frac{21}{4}\cos\omega_2+\frac{18}{4}\cos\omega_3 \qquad (22)$$

$$+\frac{3}{4}\cos(\omega_1-2\Delta)+\frac{9}{4}\cos(\omega_1-\Delta)$$

$$+\frac{9}{4}\cos(\omega_3-\Delta)+\frac{3}{4}\cos(\omega_3+2\Delta)$$

By comparing equations (15) and (22), it can be shown that the three-tone third-order IM products are two times the amplitude of the two-tone third-order IM products, or four times the power; 6 dB higher. However, these new three-tone products are also proportional to the cube of the tone power, so a 3 dB drop in tone power will still result in a 9 dB drop in IM power. A second note is that there is a single-tone, third-order IM product generated by each tone at its own frequency. This IM product will always exist, even if a single tone is fed into a linear system. It is usually ignored, since it is masked by the input tone itself, but in some cases it may prove to be significant. It is also important to note that the equation can be bro-

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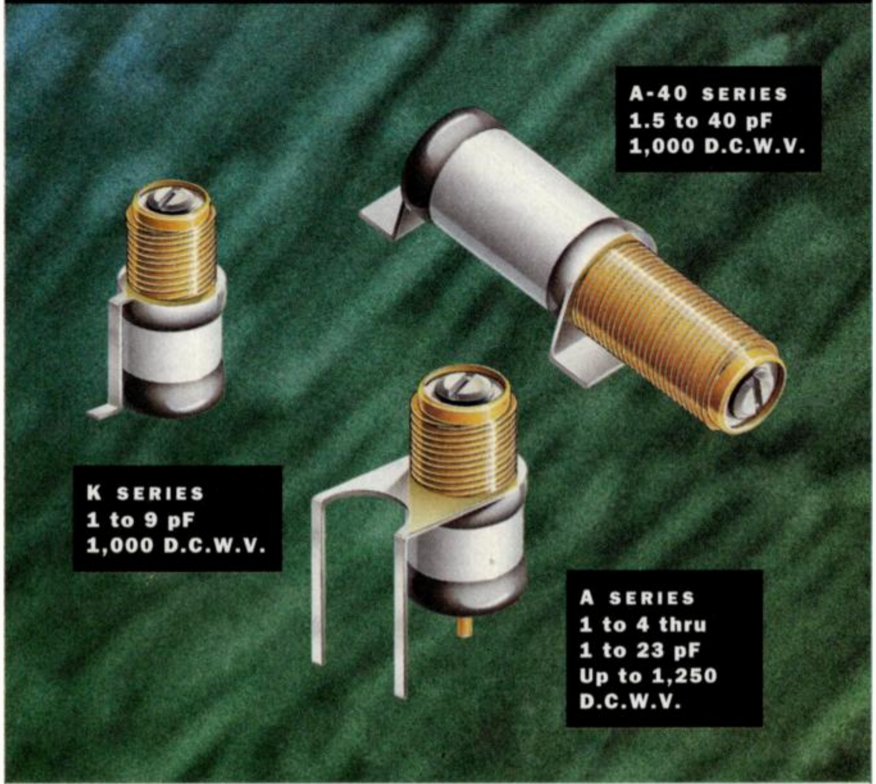

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ken down into three parts, the single tone third-order IM products, two-tone third-order IM products as produced in the original two-tone case, and three-tone third-order IM products introduced by the third tone.

A worst case frequency, where third-order IM products exist, will be Δ above or below the three tones evenly spaced by Δ, where a two-tone IM product and a three-tone IM product will combine for the worst case IM power. Since the IM product is now the combination of the two independent IM terms (independent because the phases of the tones are independent), the expected powers of the two IM terms add as in example two. This expected power level of the IM product Δ below ω_1 or Δ above ω_3 is five times (or 7 dB) higher than the original two-tone IM product. This is the expected increase and not the peak increase caused by the IM products adding in phase. The peak increase would be 9.5 dB, and would occur only if the three tones had equivalent phase offsets.

The addition of another tone (or tone N+1) will result in the addition of new products that will add with the products that have already been generated by the previous (N) tones. It can also be shown that the worst position for IM products will be one Δ above or below the tones evenly spaced by Δ. Lastly, any new IM products produced by two-tones will increase the expected IM power one Δ above or below the evenly spaced (by Δ) tones by one times the reference two-tone level, and any new IM products produced by three tones will increase the expected IM power one Δ above or below the evenly spaced (by Δ) tones by four times the reference two-tone level. (Single tone IM products are only produced at the frequency of the tones themselves, and will never appear Δ above or below the tones.) Adding a fourth tone, one new two-tone product and one new three tone product are generated Δ above and Δ below the tones. The power increase over the three tones will be five times the two-tone reference, for a total power increase of ten times the two-tone reference. In dB, the increase will be 10 log(10) = 10 dB. This procedure was repeated for a fifth tone, sixth tone, etc., until a pattern appeared. Then that pattern was generalized to the following equation:

In general, the addition of tone N+1 to N existing tones will yield the following increase in IM power level over the N tone IM power level: (Note: this equation is valid for $N \geq 2$)

$$4\left\{\frac{N+1}{2} - \frac{\mathrm{mod}\left(\frac{N+1}{2}\right)}{2} - 1\right\} + \mathrm{mod}\left\{\frac{N+1}{2}\right\} \qquad (23)$$

where mod(a/b) is the remainder of a/b.

equally spaced carriers	power multiplier due to added IM products (dB)	avg. power incr. (dB)	equally spaced carriers	power multiplier due to added IM products (dB)	avg. power incr. (dB)
1	--	--	33	1040	30.170
2	1	reference	34	1105	30.434
3	5	6.990	35	1173	30.693
4	10	10.000	36	1242	30.941
5	18	12.553	37	1314	31.186
6	27	14.314	38	1387	31.421
7	39	15.911	39	1463	31.652
8	52	17.160	40	1540	31.875
9	68	18.325	41	1620	32.095
10	85	19.294	42	1701	32.307
11	105	20.212	43	1785	32.516
12	126	21.004	44	1870	32.718
13	150	21.761	45	1958	32.918
14	175	22.430	46	2047	33.111
15	203	23.075	47	2139	33.302
16	232	23.655	48	2232	33.487
17	264	24.216	49	2328	33.670
18	297	24.728	50	2425	33.847
19	333	25.224	51	2525	34.023
20	370	25.682	52	2626	34.193
21	410	26.128	53	2730	34.362
22	451	26.542	54	2835	34.526
23	495	26.946	55	2943	34.688
24	540	27.324	56	3052	34.846
25	588	27.694	57	3164	35.002
26	637	28.041	58	3277	35.155
27	689	28.382	59	3393	35.306
28	742	28.704	60	3510	35.453
29	798	29.020	61	3630	35.599
30	855	29.320	62	3751	35.741
31	915	29.614	63	3875	35.883
32	976	29.894	64	4000	36.021

Table 1. Power increase at worst IM frequency due to multiple evenly spaced tones without a gap.

This recursive equation has been summarized in the following table. This table shows the increase in IM power level versus the number of evenly spaced tones, and can be used to predict the IM increase for N evenly spaced tones over the reference two-tone IM measurement.

Evenly spacing a given number of tones is not the worst possible arrangement of N tones. If the tones are evenly spaced with a gap in the center (skipping one channel), the IM at this center channel (frequency) will be even worse than the IM one Δ above or below the evenly spaced tones. This additional degradation is caused by three-tone third-order IM that can be produced by combinations of tones above and below the location of the IM product. Three-tone third-order IM will always generate three IM products around the three tones. One will be below all three tones in frequency, one will be above all three tones, and one will be between the lowest and highest tones. Evenly spaced three-tone third-order IM will always generate an IM product Δ above and Δ below the outside tones, and there will also be a product at the frequency of the center tone. However, three-tone, third-order IM can also be generated by tones that are not evenly spaced. If the tones are located at the frequencies of ω_1, $\omega_1 + 2\Delta$, and $\omega_1 + 3\Delta$, for instance, IM products would be produced at Δ below ω_1, Δ above ω_1, and 5Δ above ω_1. The products Δ above ω_1 that make the center location the worst case position for third-order IM when more than two tones are present.

To analyze this case, the problem can be broken down into three parts. First, the M tones below the gap can be analyzed to produce an IM product Δ above them. Use the recursive equation (23) to solve for this IM product.

Then the N tones above the gap will produce an IM product Δ below them. Once again, use the recursive equation (23) to find the IM product's level. Lastly, the combinations of one tone above and two-tones below the gap along with two-tones above and one tone below the gap can be calculated to produce additional IM in the gap of the tones. Equation 24 below can be used to find the IM due to these middle three-tone IM products. This equation was derived by trial and error.

$$4\{MN-\min(M, N)\} \qquad (24)$$

These three expected power levels can be summed, and the resultant IM power is the worse case third-order IM product. The general equations for the case of evenly spaced tones with or without a gap follow:

If M is equal to 0 and N ≥ 2,

$$1+\sum_{j=3}^{N}\left(4\left\{\frac{j}{2}-\frac{\mathrm{mod}\,\frac{j}{2}}{2}-1\right\}+\mathrm{mod}\,\frac{j}{2}\right) \qquad (33)$$

If N is equal to 0 and M ≥ 2,

$$1+\sum_{k=3}^{M}\left(4\left\{\frac{k}{2}-\frac{\mathrm{mod}\,\frac{k}{2}}{2}-1\right\}+\mathrm{mod}\,\frac{k}{2}\right) \qquad (34)$$

If M is equal to 1 and N ≥ 2,

$$1+\sum_{j=3}^{N}\left(4\left\{\frac{j}{2}-\frac{\mathrm{mod}\,\frac{j}{2}}{2}-1\right\}+\mathrm{mod}\,\frac{j}{2}\right) \qquad (35)$$

$$+4\{MN-\min(M,N)\}$$

If N is equal to 1 and M ≥ 2,

$$1+\sum_{k=3}^{M}\left(4\left\{\frac{k}{2}-\frac{\mathrm{mod}\,\frac{k}{2}}{2}-1\right\}+\mathrm{mod}\,\frac{k}{2}\right) \qquad (36)$$

$$+4\{MN-\min(M,N)\}$$

If N and M are both ≥ 2:

$$1+\sum_{j=3}^{N}\left(4\left\{\frac{j}{2}-\frac{\mathrm{mod}\,\frac{j}{2}}{2}-1\right\}+\mathrm{mod}\,\frac{j}{2}\right) \qquad (37)$$

$$+1+\sum_{k=3}^{M}\left(4\left\{\frac{k}{2}-\frac{\mathrm{mod}\,\frac{k}{2}}{2}-1\right\}+\mathrm{mod}\,\frac{k}{2}\right)$$

$$+4\{MN-\min(M,N)\}$$

where min(a,b) equals the smaller of the two numbers, and mod(a/b) equals the remainder of a/b.

Since the result is the IM power increase over the two-tone case, there must be at least two tones.

These formulas have been evaluated and put into tabular form in Table 2. Note that this table also shows the IM level for evenly spaced tones without a gap. Simply set either M or N equal to zero. The table also shows the symmetry of the IM, and you can quickly see that M and N can be interchanged yielding the same result.

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Example Three

Repeat example one, but now calculate the IMR with 32 tones (equal amplitude, random phase) instead of two tones.

A device with an input intercept point of 20 dBm and 32 input tones at 800 through 832 MHz, spaced by 1 MHz, but without a tone at 816 MHz, will generate third-order intermodulation. If the 32 input tones have an amplitude of –20 dBm, what is the expected value of the IM at the worst point?

The two-tone IMR level was already calculated to be 80 dB. Looking up 16 tones by 16 tones in Table 2 shows that the expected value of the IM will be degraded by 31.5 dB over the two-tone case, so the IMR is expected to degrade to 48.5 dBc. This assumes random phase for the 32 tones.

Example Four

How much would the power per tone need to be backed off to restore 80 dBc IMR?

Since this IM is third-order, it still follows the 3:1 intercept equation. To improve the IMR by 31.5 dB, the power per tone will need to be decreased by (1/2) of 31.5, or 15.75 dB.

This yields a two-tone IMR level of 2(20–(–35.75)) = 111.5 dB. When the number of tones is increased to 32, the IMR will degrade to 80 dBc.

Notice that the total power input to the device is higher for the two-tone case than it is for the 32-tone case. Two tones at –20 dBm yields an input power of 0.02 mW with 80 dB IMR. 32 tones at –35.75 dBm yields an input of 0.0085 mW and 80 dB IMR. The 32-tone total input power is 3.7 dB lower in power than the two-tone input power.

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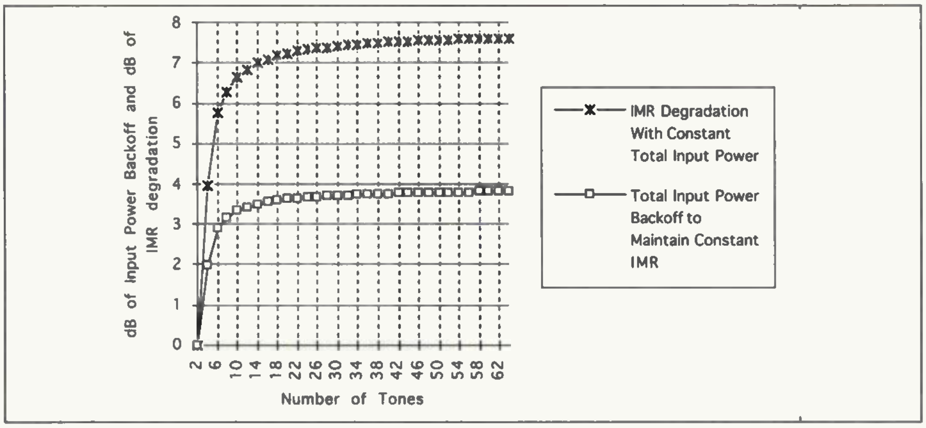

Figure 2. Power backoff required to maintain constant IMR for 2 to 64 tones.

Example Five

How much would the power per tone need to be for 60 dBc IMR with 32 tones?

32 tones will degrade the IMR by 31.5 dB, so two-tone IMR will only need to be 91.5 dBc in order for the 32-tone IMR to be 60 dBc. To get a two-tone IMR level of 91.5 dBc, the power per tone will need to be 20 dBm – (91.5 dB/2) = –25.75 dBm / tone. 32 tones at –25.75 dBm is 0.085 mW of total input power with 60 dBc IMR performance. 60 dBc performance for two tones would require a total input power level of –10 dBm per tone. This yields a total power of 0.2 mW. Once again, in order to maintain constant IMR, the total input power for 32 tones needs to be backed off by 3.7 dB from the two-tone case. (0.2 mW – 0.085 mW = 3.7 dB)

Example Six

Generate a graph that shows how much the total input power needs to be backed off to maintain a constant IMR level from two tones to 64 tones. Also graph how much the expected IMR level degrades if a constant total input power level is maintained. Results are shown in Figure 2.

Conclusion

Multi-tone intermodulation performance can be predicted from the ordinary two-tone intercept specification commonly available for linear components. In the general case where the phases of the carriers aren't controlled or predictable, this method allows the engineer to calculate the expected level of IM in a linear device that is subjected to more than two tones. Keep in mind that this calculation is not a worse case, or even a conservative calculation. To the contrary, it is a calculation that yields what could be called the average level of IM that would be measured if a device were tested hundreds of times with a fixed number of input tones, but with each tone having a randomphase offset. *RF*

About the Author

Michael Leffel is a Lead Engineer in Motorola's Advanced Product Development Group, where he works on the design of high power, multi-carrier linear amplifiers. He holds an MSEE degree from the Illinois Institute of Technology, and a BSEE from the University of Wisconsin - Madison. He is currently working on an MS degree in Computer Science from the National Technical University. He can be reached at: leffel_mike@macmail2.cig.mot.com or by mail at Motorola, Inc., Room 400, 1475 West Shure Dr., Arlington Heights, IL 60004.

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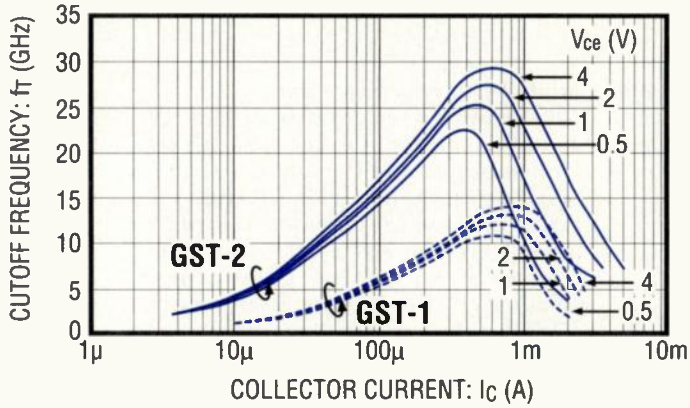

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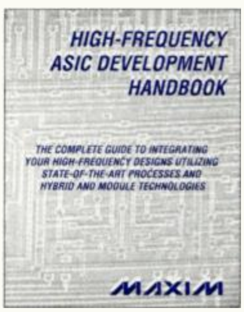

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INFO/CARD 63

Market Growth Calls for Versatile and Customized RF Signal Generators

This month's Product Forum looks at the market changes affecting RF signal generators. Several manufacturers offer their opinions regarding today's marketplace and trends in the industry.

Tektronix

Growth in the cellular and two-way radio markets, coupled with the conversion from analog to digital technology, is fueling an increased demand for RF signal generators. Correspondingly, the shift from analog to digital technologies is driving a need for signal generators with a wide variety of digital modulation capabilities, including modulation schemes for testing CDMA. The constantly changing market requires products that have flexibility. For example, an RF signal generator can address a variety of standards when used with an arbitrary waveform generator.

Tektronix participates in the RF signal generator market through its United States and Canadian distribution agreement with Rohde & Schwarz based in Germany. Tektronix also offers the AWG 2000 series of arbitrary waveform generators. Recently, the Rohde & Schwarz IQSIM software for generating I/Q signals was made compatible with Tektronix' AWG product line.

Tektronix AWG 2040 arbitrary waveform generator.

Programmed Test Sources, Inc.

For commercial manufacturers targeting test and measurement OEM markets, RF signal generator requirements today are being set largely by the expanding wireless communications sectors. To meet the needs of these OEMs, generators must either support the variety of modulation formats employed, or provide low-noise, fast-frequency switching signals required for testing wireless communications components or systems. PTS serves this market through a complete product line of low-noise, fast-switching single and dual channel generators, and will soon announce a product designed to support the 3.2 GHz spectrum.

The technologies employed in this sector will be a mixture of digital and analog — fine frequency resolution and modulation generation are largely digital, but for the foreseeable future, analog technologies will continue to form the basis for high-frequency carrier generation. However, the increasing availability of RFICs in the multi-GHz frequency bands will lead to improved generator performance and lower costs in the near future.

Wiltron Company, Microwave Measurements Division

How do I test my device at my frequency with my modulation technique? Easy question, tough answer. Today's producers insist on solutions to their problems, not just products. Additionally, the over-burden of the RF frequency spectrum is leading communications to ever higher frequencies and causing a virtual plethora of modulation techniques.

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Multichip Module Assembly Process

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Surface Mount	Solder-mounting of packaged components	Low cost; uses pretested components
Chip-and-Wire	Epoxy-attach of chip components; e.g., ICs, transistors, capacitors, crystals, and wire-bonding to substrate	Minimum interconnect parasitics; compatible with hermetic sealing
Mixed-Tech	Combines chip-and-wire with surface mount	Maximum flexibility
Electro-Optic	Combines chip-and-wire with fiber alignment to 0.25μM and >80% coupling	Minimizes interconnect parasitics between active circuits and electro-optic components
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2. Work with our customers to provide semi-custom solutions to their problems.

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Hewlett-Packard Company

Hewlett-Packard is experiencing substantial worldwide growth in RF signal generator sales, especially in the Asia Pacific region. As the communications market develops at breakneck speed, so does the need for test equipment. Manufacturers of pagers, two-way radios, and cellular telephones are rapidly adding and expanding their production lines, and time-to-market is critical.

Although there is a trend toward digitally-based architectures, analog performance is still in demand and will be for some time to come. However, as digital signal processing becomes more efficient, manufacturers will move toward the more flexible capabilities of digitally-based architectures.

Hewlett-Packard has a broad offering of RF signal generators, and is continuing to invest in new products. Their focus is not only on new technology, but also on developing flexible products that offer measurement accuracy, ease of use, upgradability and economical prices.

Berkeley Varitronics Systems

Berkeley Varitronics Systems is a small market-driven company that serves the communications industry. Our products are of a customized nature. As an example, we are now in the process of finalizing the design of the "Spyder" portable synthesized signal source for use by PCS users.

The market for specialized instrumentation has always been present, however with the rapid growth in the wireless industry, many small companies are finding an environment rife with opportunities. The next six months to two years is expected to be a period of substantial growth for companies that are able to offer solutions to problems, not just off the shelf products. This is the philosophy of Berkeley Varitronics Systems.

As an example, the "Spyder" is a self-contained, battery powered, signal generator that provides a cw or modulated signal for use with the companion "Champ" portable signal strength meter. This pair of instruments is used for field mapping and propagation studies in local areas.

Wayne Kerr

Two fundamental specifications for RF signal generators are phase noise and level accuracy. Direct digital synthesis techniques are used for modulation generation and can replace some phase locked loops, but this method is always limited by the generation of non-harmonic spurs. For this reason, high Q analog oscillators locked to a crystal reference by indirect techniques is still the preferred method for lowest phase noise without spurs. Level accuracy on signal sources can be improved by frequency error compensation in software. However, a well designed analog switching attenuator will deliver the same performance and will be superior when in sweep mode. Software correction should be limited to no more than 3 dB, it can easily be "overdone". *RF*

For more information on RF signal generator companies circle the INFO/CARD numbers below:

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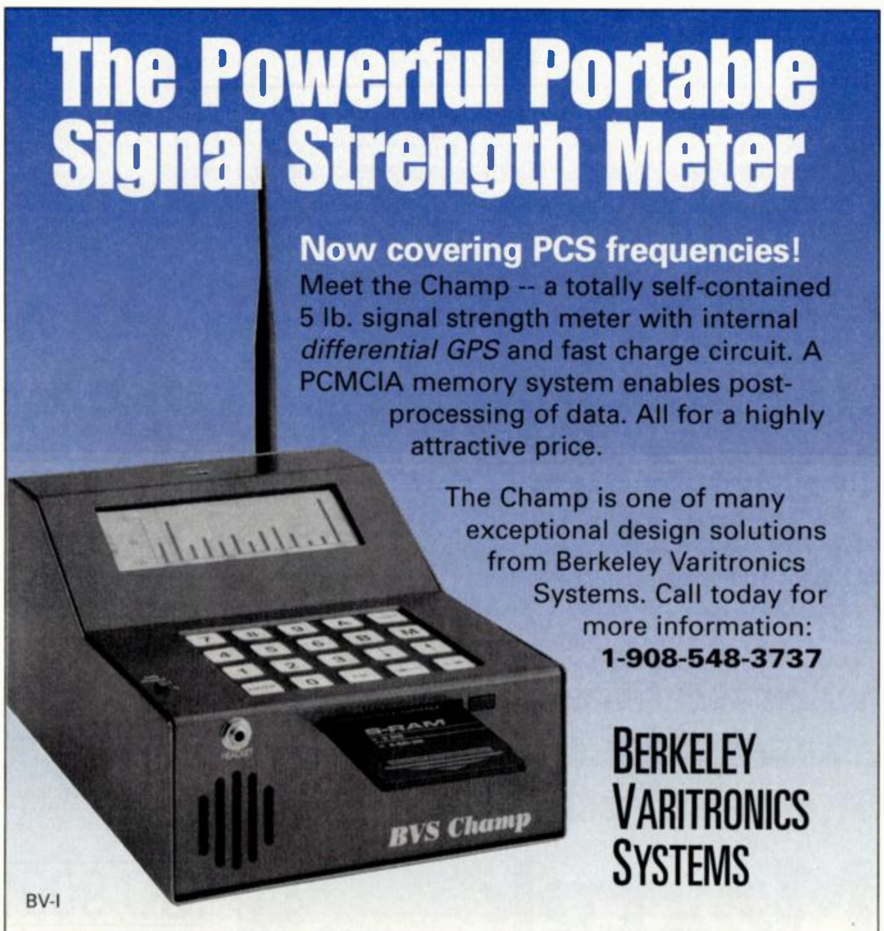

INFO/CARD 71

RF guide to editorial coverage

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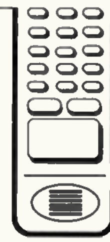

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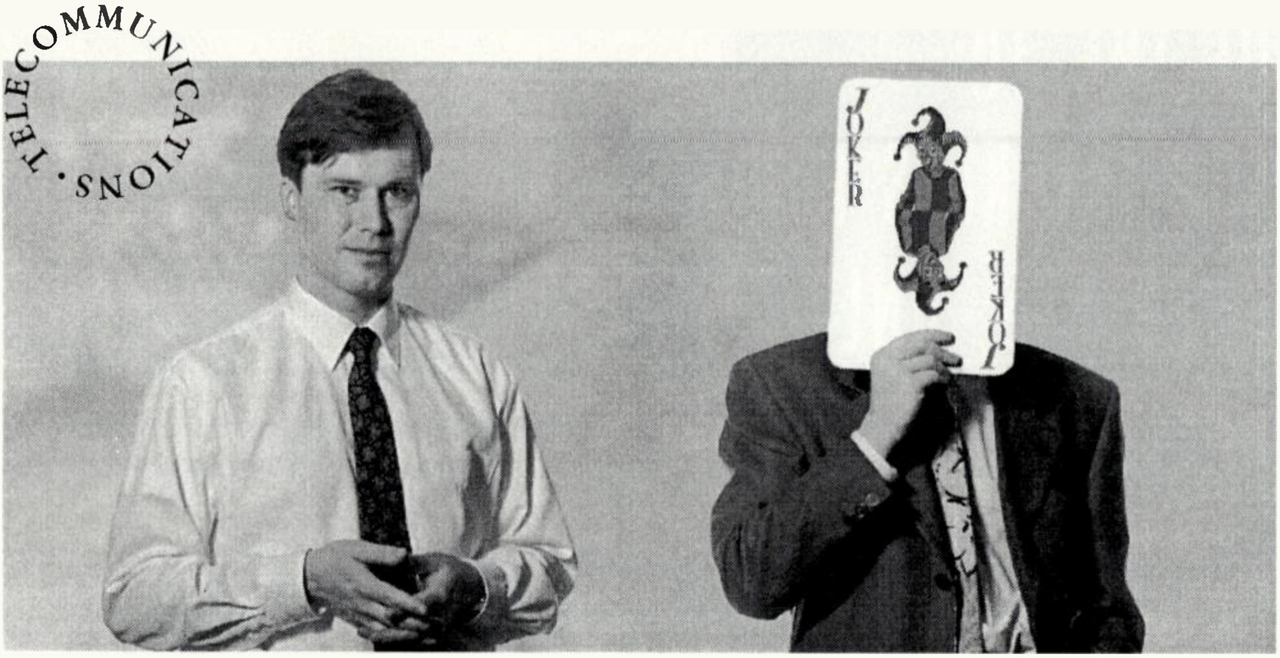

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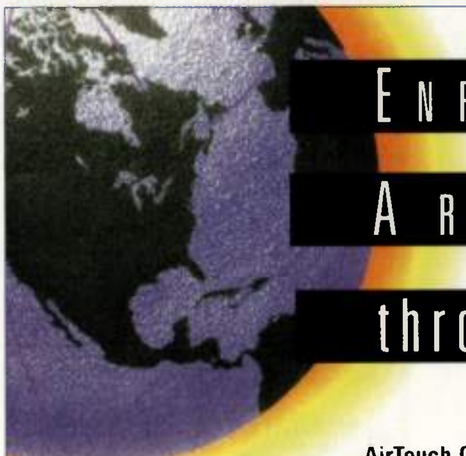

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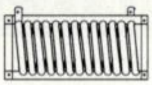

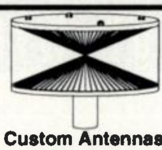

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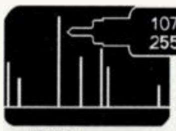

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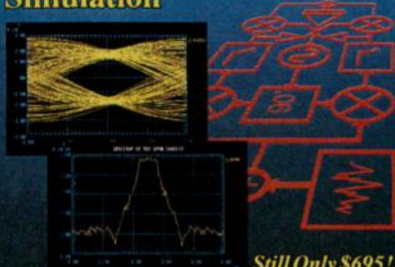

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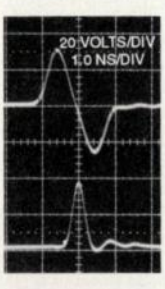

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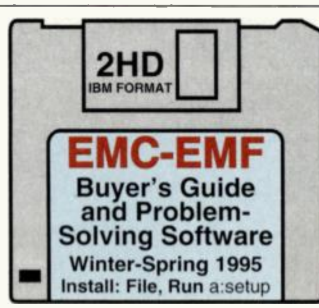

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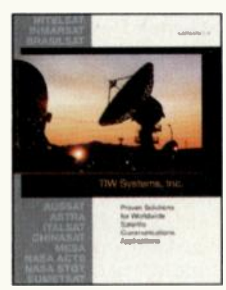

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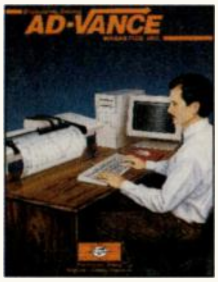

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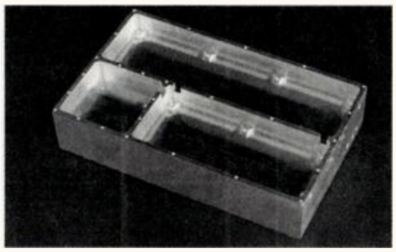

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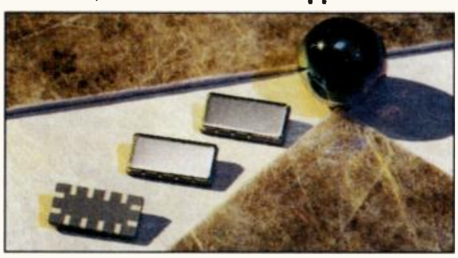

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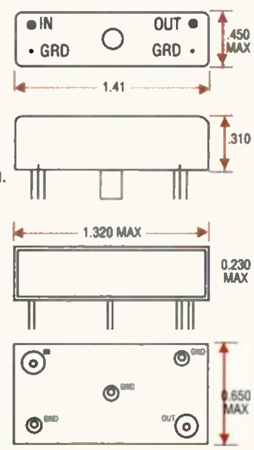

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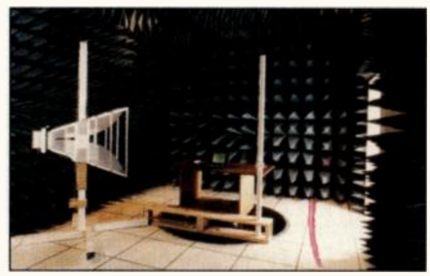

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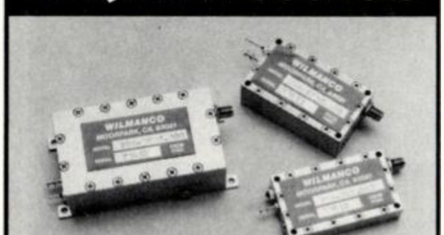

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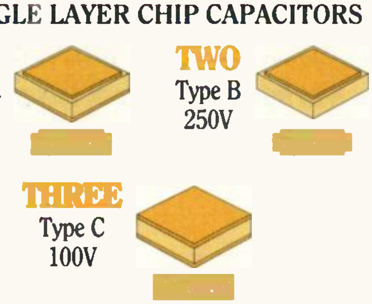

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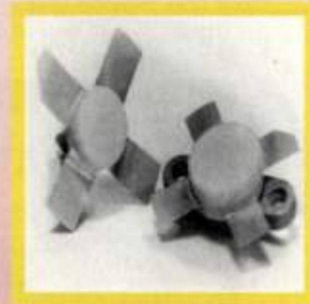

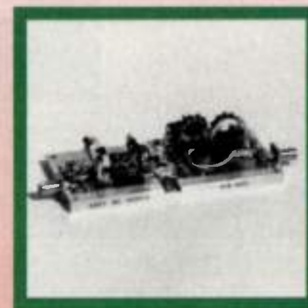

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ADC Telecommunications, 4900 W. 78th St., Minneapolis, MN 55435 ....(800) 366-3891 X3000
Phoenix Contact Inc., PO Box 4100, Harrisburg, PA 1711-0100..................................(717) 944-1300
Optical Fibers And Connectors
ADC Telecommunications, 4900 W. 78th St., Minneapolis, MN 55435 ....(800) 366-3891 X3000

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Absorbers
Advanced Electromagnetics, Inc., P.O. Box 711719, Santee, CA. 92072- 1719............(619) 449-9492

IBEX Group, Inc., 23 Markham Dr., Long Valley , NJ 07853 ....(800) 403-3930
Rantec, 24003 Ventura Blvd., Calabasas, CA. 9130 ....(818) 591-8189
Anechoic Chambers
Advanced Electromagnetics, Inc., P.O. Box 711719, Santee, CA. 92072- 1719............(619) 449-9492
IBEX Group, Inc., 23 Markham Dr., Long Valley , NJ 07853 ....(800) 403-3930
Rantec, 24003 Ventura Blvd., Calabasas, CA. 9130 ....(818) 591-8189
Antennas
Antenna Research Associates, Inc., 11317 Fredrick Ave., Beltsville, MD 20705....(310) 937-8888
Antennas Above 30 MHz
A.H. Systems Inc., 9710 Cozy Croft Ave, Chatsworth, CA 91311..(818) 998-0223 Fax (818) 998-6892
EMCO, P.O. Box 1546, Austin, TX. 78767...(512) 835-4684
IBEX Group, Inc., 23 Markham Dr., Long Valley , NJ 07853 ....(800) 403-3930
Antennas Masts
Antenna Research Associates, Inc., 11317 Fredrick Ave., Beltsville, MD 20705....(310) 937-8888
EMCO, P.O. Box 1546, Austin, TX. 78767...(512) 835-4684
IBEX Group, Inc., 23 Markham Dr., Long Valley , NJ 07853 ....(800) 403-3930
Current Probes
Fischer Custom Communications, 2905 W. Lomita Blvd, Torrence, CA 90505.............(310) 891-0635
IBEX Group, Inc., 23 Markham Dr., Long Valley , NJ 07853 ....(800) 403-3930
Ion Physics Corp., 11 Industrial Way, Atkinson NH 03811 .... (603) 893-6687
Ferrite Absorber Tiles
IBEX Group, Inc., 23 Markham Dr., Long Valley , NJ 07853 ....(800) 403-3930
GTEM Cells
Antenna Research Associates, Inc., 11317 Fredrick Ave., Beltsville, MD 20705....(310) 937-8888

ELECTRONIC COMPONENTS AND EQUIPMENT

EMC TEST EQUIPMENT - EMISSIONS

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IBEX Group, Inc., 23 Markham Dr., Long Valley , NJ 07853 .... (800) 403-3930

Near Field Probes
Antenna Research Associates, Inc., 11317 Fredrick Ave., Beltsville, MD 20705 .... (310) 937-8888
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Preamplifiers
IBEX Group, Inc., 23 Markham Dr., Long Valley , NJ 07853 .... (800) 403-3930

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

Solar Preamplifiers
Picosecond Pulse Labs, PO Box 44, Boulder, CO 80306 .... (303) 443-1249

EMC TEST EQUIPMENT - SUSCEPTIBILITY

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

Power Amplifiers
Applied Systems Engineering, Inc., 8623 Hwy. 377 S., Fort Worth, TX 76126 .... (819) 249-4180
Varian Microwave Equip. Prods., 3200 Patrick Henry Dr, Santa Clara, CA 95054 .... (408) 496-6273

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

Surge Generators
Velonex Corp., 560 Robert Avenue, Santa Clara, CA 95050 .... (408) 727-7370
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Monroe Electronics, Inc., 100 Housel Avenue, Lyndonville, NY 14098 .... (800) 821-6001
TREK INC., 3932 Salt Works Rd., P.O. Box 728, Medina, NY. 14103 .... (800) FOR-TREK

Surface & Volume Resistivity Meters
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TREK INC., 3932 Salt Works Rd., P.O. Box 728, Medina, NY. 14103 .... (800)FOR-TREK

EMC TEST EQUIPMENT - ADDITIONAL

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Combinova/Ergonomics, Inc., PO Box 964, Southhampton, PA 18966 .... (215) 357-5124

MATERIALS, HARDWARE AND PACKAGING

SHIELDING MATERIALS

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

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

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

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

Electromagnetic Shielding
RFI Controls Co., 320 N. Santa Cruz Ave., Los Gatos, CA 95030-7243 .... (408) 399-7007

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

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

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

Open Area Test Site
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Non-Compliance Investigation
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SHIELDING EMI/RFI SERVICES

Vacuum Platers, Inc., 115 S. Union Street, Mauston, WI 53948 .... (608) 847-5644

SHIELDED ENCLOSURES - EQUIPMENT

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

Tempest Enclosures
Technical Environment Control, Inc., 7950 Cessna Ave., Gaithersburg, MD 20879 .. (301) 948-5911

SHIELD ROOMS AND CHAMBERS

Shielded Rooms EMI/RFI/Magnetic
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TUV Product Service Inc 1775 Old Hwy 8 New Brighton MN 55112 .... (800) 472-7999

EMI & Safety
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Compatible Electronics - 8 open field test sites in So. Calif .... (714) 579-0500
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Ultratech Eng. Labs, Inc.
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TEST CAPABILITIES

Military EMC Standards
Northrop Grumman,
B-2 Division, 8900 E. Washington Blvd., T623/XE, Pico Rivera, CA 90660 .... (310) 942-3895

Shielding Effectiveness
Shielding Integrity Services, 1905 Hercules Dr., Colorado Springs, CO 80906 .... (719) 635-7719

Susceptibility/Immunity
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ADDITIONAL SERVICES

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Optotek Ltd.
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Filter Design Software

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Tatum Labs, Inc.
INFO/CARD #179

Tek 271x Utilities

A collection of utilities from Ramshorn at Yamhill runs on any 386 or better computer running Windows™ 3.1 and provides a number of functions for Tektronix 271x spectrum analyzers. Among the utilities included are those that: retrieve waveform data from any register, download saved waveforms to any register, display multiple tabular or graphical results on the computer and then paste them directly into other Windows applications, create and download arbitrary limit lines, control the analyzer remotely from the computer, and save entire analyzer settings groups. Price is $299.
Ramshorn at Yamhill
INFO/CARD #178

Simulators

Compact Software has begun shipments of its 6.5 series of frequency-domain simulators for PCs running Microsoft Windows™. The Serenade 6.5 family includes the Super-Compact linear simulator, the Microwave Harmonica nonlinear simulator, and the Microwave Scope opto-electronic nonlinear simulator. New features include improved integration between schematic editors and simulators, enhanced transmission line synthesis and analysis, EEsof to Compact file converter, new models, enhanced Windows functions, and and optional nonlinear device library.
Compact Software
INFO/CARD #177

Tuner Driver for HP-Vee®

Focus Microwaves has introduced CCMT-VEE, which operates under Windows 3.0 and permits the integration of Focus' CCMT programmable tuners in a VEE-defined setup as fully calibrated instruments. It allows VEE to calibrate the tuners, integrate them in the setup, and use the de-embedded data to synthesize any (interpolated or extrapolated) impedance at any calibrated frequency at the DUT's reference plane.
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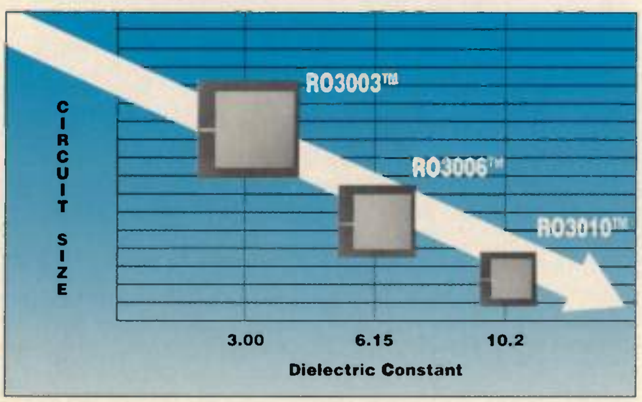

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Power Meter Data

Boonton Electronics has released a two-page data sheet on its Model 4300 microprocessor-based power meter. The black-and-white data sheet describes all the pertinent features and functions of the meter, including frequency range, measurement accuracy, and automatic test capability. Ordering information is also provided.
Boonton Electronics Corp.
INFO/CARD #175

Antenna Catalog

A new commercial antenna catalog, covering wireless frequencies from 800 MHz to 10.5 GHz, is now available from Tecom Industries. The 16-page book features a wide spectrum of standard and custom antennas, including Tecom's new family of embedded designs.
Tecom Industries, Inc.
INFO/CARD #174

Trimmer Catalog

Voltronics has a new 34-page catalog of its precision trimmer capacitors. The catalog has complete specifications of Voltronics' glass, quartz, air, sapphire and PTFE precision trimmer capacitiors. Engineering sample kits, tuning tools, and special design capabilities are described, and four-pages specify many types of DRO and microwave cavity tuners.
Voltronics Corp.
INFO/CARD #173

SPICE Newsletter

The latest edition of Intusoft's free publication, the Intusoft Newsletter, contains information on simulating digital filters and current mode designs. The issue also includes information on a new multi-threaded, multi-processor version of IsSpice4, as well as their simulation products for Macintosh platforms.
Intusoft
INFO/CARD #172

Test Probe Catalog

A 14-page, full-color catalog from Test Probes Incorporated (TPI) includes information on TPI's lines of oscilloscope probes (including repairable, differential, and active FET types), test leads, high voltage and RF probes, connector kits, and attenuator and termination kits.
Test Probes, Inc.
INFO/CARD #171

Connector Catalog

High reliability, rugged design, low VSWR and high resistance to corrosion are key attributes of the Spinner GmbH coaxial connetors detailed in a 160-page catalog from Precision Tube Company. The Spinner connector assortment ranges in sizes from 3.5 mm to 16 mm dia. (for the outer conductor) and includes 7/16 connectors for cellular and mobile communications systems, with low IM performance.
Precision Tube Co., Inc.
INFO/CARD #170

PCB Material Selection Guide

Rogers offers a six-page selector guide for its microwave circuit board materials. The guide describes the properties of RT/duroid and TMM® temperature stable laminates and RT/duroid 6002 circuit materials, which have a very low thermal coefficient of dielectric constant. The selector guide also describes the cladding thicknesses available.
Rogers Corp.
INFO/CARD #169

Amplifier Catalog

Comtech Microwave Products Corp., Power Systems Technology Div. has issued a 16-page product catalog for its wide selection of high quality, solid state high power amplifiers. Product categories encompass: AM and AR cellular, UFA feedforward, wideband BHE/BHC class AB linear, class A general purpose, and custom designs, including CHC class C and PHE/PHE pulse amplifiers.
Microwave Products Corp.
Power Systems Technology Div.
INFO/CARD #168

Analog Component Selection

Comlinear has announced the availability of its 1995 Product Selection Guide. This 8-page guide is organized by product type and provides key specifications on the company's large family of high-performance components. Included in the guide are Comlinear's high-speed, wide bandwidth op amps; variable-gain, buffer, high-power, track-and-hold, and modular amplifiers; high-performance A/D converters; and new serial-digital interface components.
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Frequency Control Data

Motorola offers a 204-page data book presenting technical data on a broad line of Motorola integrated circuits useful in a wide variety of PLL applications. The Hipercomm High Performance Frequency Control Products book contains data sheets for prescalers, VCM and VCOs, phase/frequency detectors, and frequency synthesizers, as well as application information.
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INFO/CARD 74

Component Modeling

The Handbook of Component Modeling, a two-volume set by Colin May, stresses knowledgable component modeling for analog circuit simulation. Compoents range from BJTs and FETs to thyristors and TRIACs - passive filters to logarithmic amplifiers. The set is published by Tatum Labs and costs $95.
Tatum Labs, Inc.
INFO/CARD #165

9th EMC Conference Proceedings

ERA Technology has published the proceedings of its ninth Electromagnetic Compatibility (EMC) conference. The conference theme was, "EMC: The Clock is Ticking - Countdown to Compliance."The report costs £80.00.
ERA Technology Ltd.
INFO/CARD #164

Linear Design Reference

A 700-page linear circuit design reference book published by Analog Devices is now available. The 11-chapter book is tailored to the entry-level engineer, advanced college student, and engineers who are not familiar with the fundamentals of linear circuits. The retail price is $25 for single copies.
Analog Devices, Inc.
INFO/CARD #163

Telecomm Test Equipment

Telecom Analysis Systems' 12-page short form catalog contains descriptions and four-color photographs of their communication analyzers, network emulators, RF channel emulators, modem and terminal emulators, ISDN emulator, and loop emulators.
Telecom Analysis Systems, Inc.
INFO/CARD #162

RF Cable and Connector Catalog

Solitron/Vector Microwave Products (SV Microwave) has released their new catalog of RF connectors, film products, and cable assemblies.
SV Microwave
INFO/CARD #161

IC Guide

Hearst Business Publishing announces the publication of the 1995 IC MASTER catalog. The catalog covers 108,000 commercially available ICs, including more than 20,000 ICs listed for the first time. The U.S. price for the three-volume, 3,000-page set is $185.00 plus tax and $10 shipping and handling.
Hearst Business Publishing
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RF Design Software

Programs from RF Design provided on disk for your convenience

June Disk — RFD-0695
"Program Aids Design of Coaxial and Waveguide Oscillators" by Roy Monzello. This program applies the work of A.G. Williamson to analyze the impedance characteristics of coaxial line coupled into waveguide, to help design GUNN and IMPATT diode microwave oscillators. (Directly executable, with example files and manual that can be printed from the disk. For acceptable calculation speed, math coprocessor and '386 or higher CPU are recommended.)

May Disk — RFD-0595
"Program Synthesizes Antenna Matching Networks for Maximum Bandwidth" by Robert Dehoney. This program determines component values for networks of multiple elements that maintain a feedpoint VSWR under 2:1 for up to 40% bandwidth. (Written in GWBASIC for operation on any MS-DOS PC. GWBASIC is bundled with DOS versions 3.x and lower. Minor changes are required to run under QBASIC or other BASIC varieties)

Index of *RF Design* Articles: 1978-1994 — Disk RFD-INDEX
The RFD-INDEX disk has been updated to include all articles published in *RF Design* from its first issue (November/December 1978) through December 1994. Data is provided as ASCII text, which can be loaded into your favorite word processing program for searching and printing.

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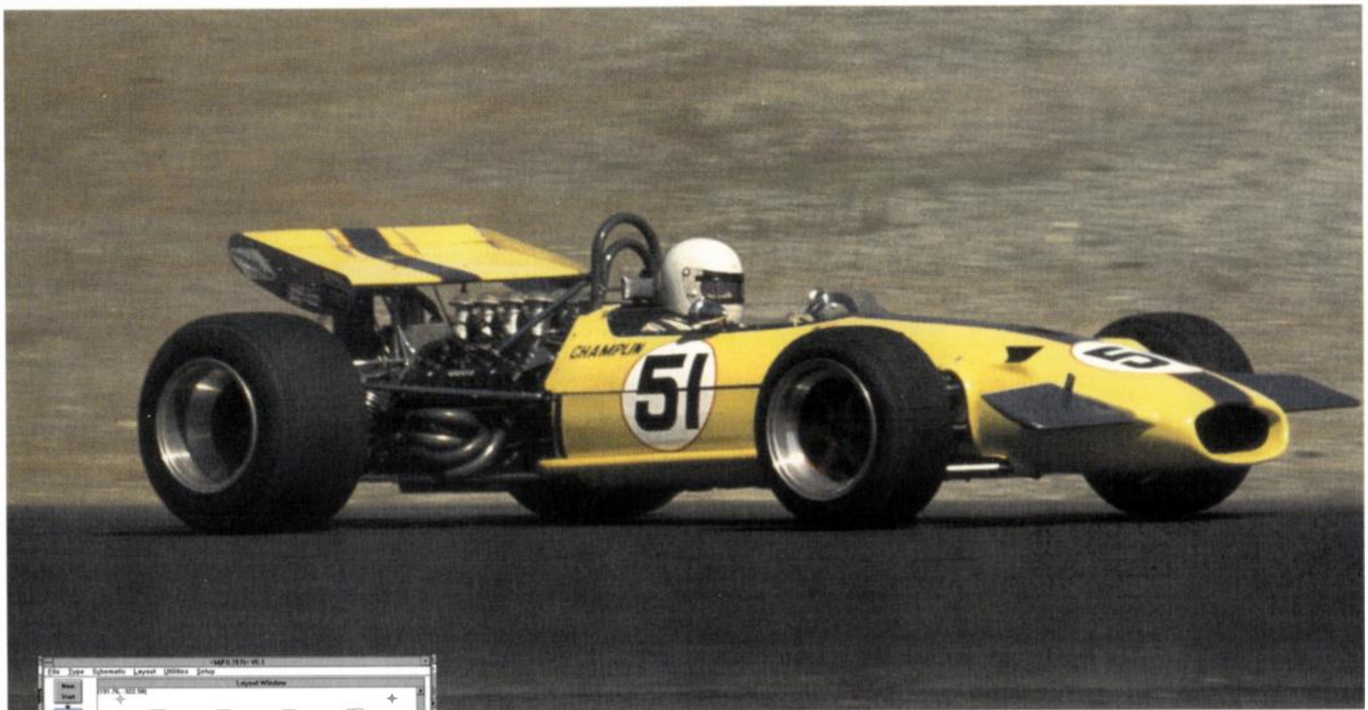

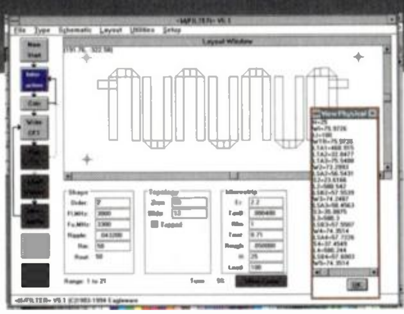

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L-C Colpitts
L-C Clapp
T-line and L-C VCO
VCO with xformer
Cavity bipolar and hybrid
Dielectric resonator
Terminal SAW bipolar
Port SAW hybrid
Port SAW MOSFET
Pierce and Colpitts crystal
Driscoll crystal
Butler overtone
Overtone with multiplier

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

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

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

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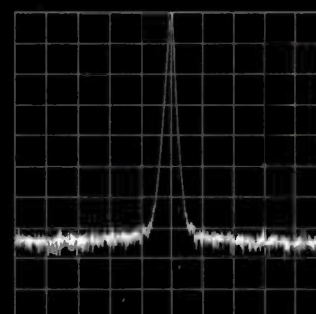

Ultra-low phase noise floor says it all.

	LMX1501A	LMX1511	LMX2314/15	LMX2320	LMX2325	LMX2330	LMX2331	LMX2332
RF Input-Main PLL	1.1GHz	1.1GHz	1.2GHz	2.0GHz	2.5GHz	2.5GHz	2.0GHz	1.2GHz
RF Input-Aux PLL						510MHz	510MHz	510MHz
I_{CC} (typ) @3V	6mA	6mA	6mA	12mA	14mA	15mA	14mA	8mA
Powerdown (typ)	N/A	N/A	30µA	30µA	30µA	1µA	1µA	1µA

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