Vol. 48 • No. 11
November 2005

Microwave Journal

Passive and Control Components

Passive Components: A Brief History

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MODEL NUMBER	FREQUENCY RANGE (GHz)	GAIN (dB, Min.)	GAIN FLATNESS (±dB, Max.)	* NOISE FIGURE (dB, Max.)	VSWR IN/OUT	* OUTPUT POWER @ 1 dB COMP. (dBm, Min.)
LOW-NOISE, VARIABLE GAIN AMPLIFIERS						
NSP1000-NVG	0.1–10	35	2	2.3	2:1	10
NSP1200-NVG	0.1–12	32	2	2.5	2:1	10
NSP1800-NVG	0.1–18	30	2.5	4	2.5:1	10
NSP2200-NVG	0.1–22	30	2.75	4.5	2.5:1	10
MEDIUM POWER, VARIABLE GAIN AMPLIFIERS						
NSP1000-PVG	0.1–10	35	2	5	2:1	20
NSP1200-PVG	0.1–12	32	2.5	5.5	2:1	20
NSP1800-PVG	0.3–18	30	2.75	6.5	2.5:1	20
NSP2000-PVG	0.3–20	30	3	7	2:1	20
LOW-NOISE, FIXED GAIN AMPLIFIERS						
NSP1000-NFG	0.1–10	28	2	2.3	2:1	10
NSP1200-NFG	0.1–12	28	2	2.5	2:1	10
NSP1800-NFG	0.1–18	20	2.5	3	2.5:1	10
NSP2650-NFG	0.1–26.5	22	2.75	4.5	2.5:1	10
NSP4000-NFG	0.1–40	22	3	5	2.5:1	8
MEDIUM POWER, FIXED GAIN AMPLIFIERS						
NSP1000-PFG	0.1–10	25	2	5	2:1	20
NSP1200-PFG	0.1–12	25	2.25	5.5	2:1	20
NSP1800-PFG	0.3–18	18	2.75	8	2.5:1	20
NSP2000-PFG	0.3–20	18	3	8	2.5:1	20
NSP2200-PFG	0.3–22	18	3	8	2.5:1	20

* Specification applies above 500 MHz.** Split into 2 separate bands (0.1–26 GHz and 26–40 GHz).

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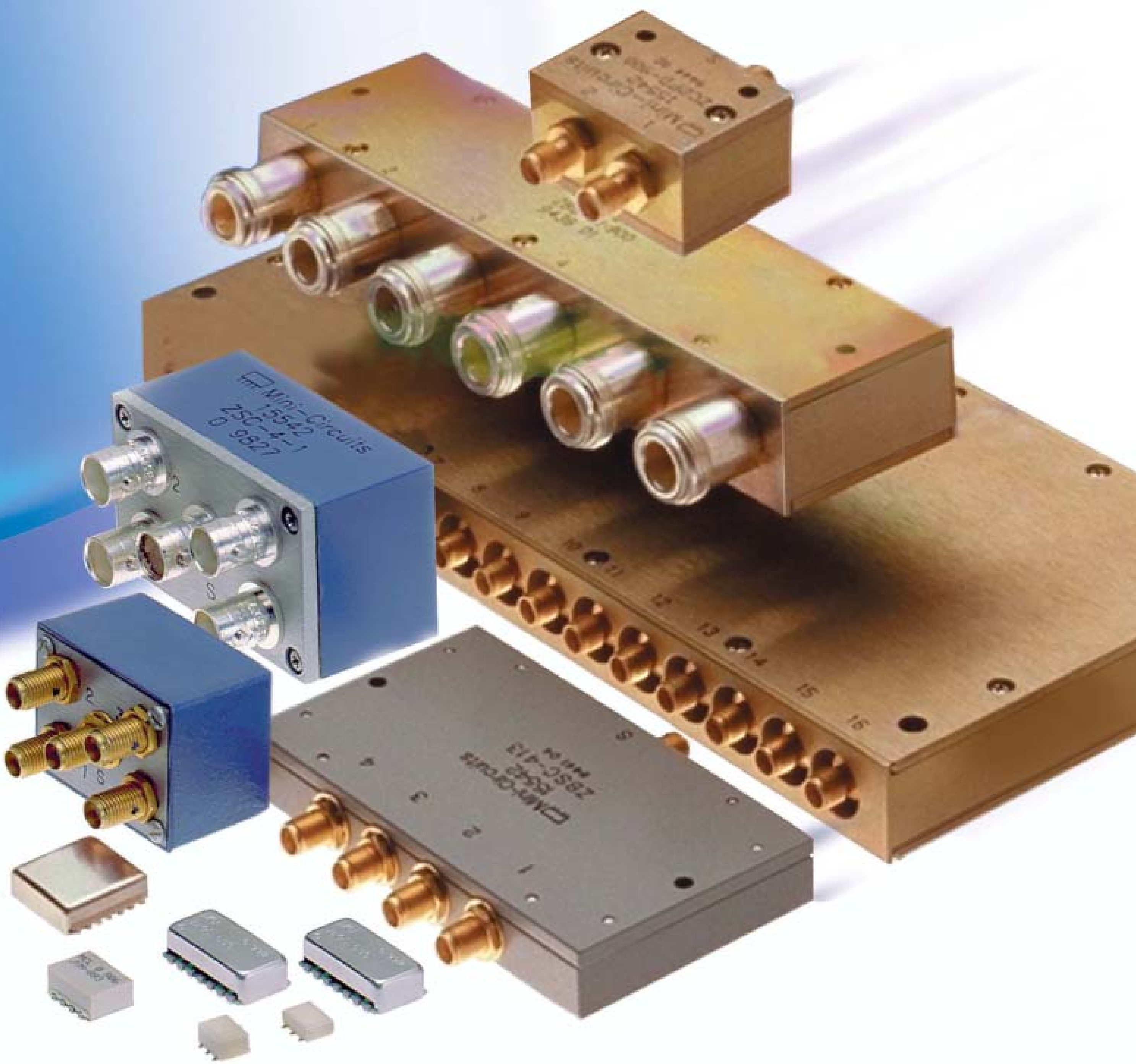

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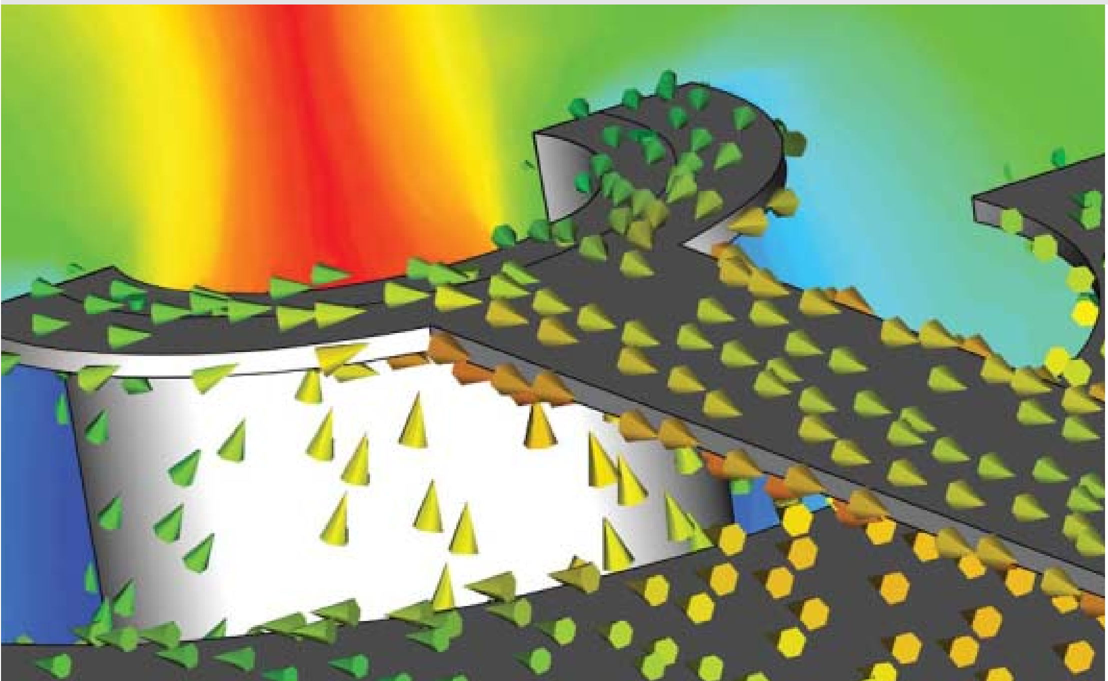

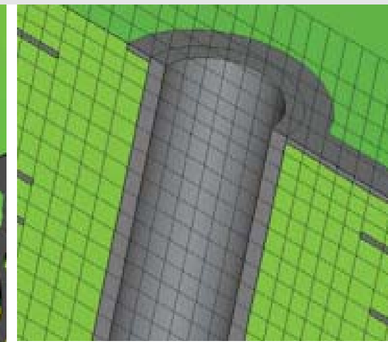

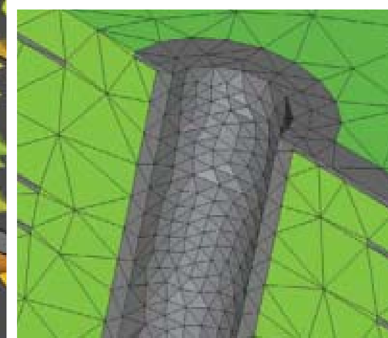

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Coming Events

66th ARFTG Microwave Measurements Conference
December 1–2, 2005
Washington, DC

This conference will explore all aspects of microwave measurements for both the industrial and government-related engineering communities. Topics: test challenges for high power devices (> 10 W), measurements on emerging technologies, measurement accuracy, nonlinear measurement as well as other areas of automated RF measurements. For further information, contact: J. Gregory Burns, conference chair, Northrop Grumman (410) 765-7331, e-mail: john.g.burns@ngc.com or visit www.arftg.org.

Asia-Pacific Microwave Conference
December 4–7, 2005
Suzhou, China

The 17th Asia-Pacific Microwave Conference (APMC 2005) is widely supported by the academia and industries of microwave, wireless and correlative areas. The conference will highlight solid-state devices and circuits, low/high noise devices and techniques, monolithic integrated circuits, passive devices and circuits, packaging, interconnects, MCMs, scattering and propagation. A large-scale exhibition of commercial products related to the conference topics will be held concurrently. For more information, visit www.apmc2005.org.

IEEE Radio and Wireless Symposium
January 17–19, 2006
San Diego, CA

The IEEE Radio and Wireless Symposium (RWS 2006) is a major expansion of the successful Radio and Wireless Conference (RAWCON), most recently held in Atlanta, GA, September 2004. This conference maintains a focus on interdisciplinary aspects of wireless and RF systems and technology with an emphasis on how the elements fit together to shape the latest developments in communications technology and enable the convergence of applications. In addition to oral presentations and posters, RWS includes workshops, panels and a major exhibition. The inaugural RWS 2006 is part of a week-long major technical event – *MTT Wireless*. Also participating in *MTT Wireless* are the *Topical Meeting on Silicon Monolithic Integrated Circuits in RF Systems* (SiRF) and the *IEEE Topical Workshop on Power Amplifiers for Wireless Communications* (PA Workshop). Companies interested in the exhibition or in sponsorships should contact Kristen Dednah at (781) 769-9750 or e-mail: kdednah@mwjournal.com. For additional information, visit www.radiowireless.org.

SATELLITE 2006
February 6–9, 2006
Washington, DC

The annual SATELLITE 2006 conference and exhibition is organized by *Via Satellite* and Access Intelligence. SATELLITE 2006 is a global conference and exhibition where industry executives share marketplace intelligence. This event offers attendees invaluable business solutions and mission-critical knowledge for specific operational requirements. For more information, visit www.satellite2006.com.

RF & Hyper Europe 2006
March 21–23, 2006
Paris, France

Now celebrating its 32nd edition, RF & Hyper Europe 2006 will, for the first time, be organized by Exposium's Industry, Commerce & Technologies department that has taken over from BIRP. Growth, especially on the international front, is expected for the event, which will showcase the technological innovations of some 160 exhibitors and reflect the current activity in the mobile telecommunication, data processing, military and automotive sectors. As always it will be dedicated to the promotion of RF, microwaves, optical fiber and wireless products and services,

not only through the exhibition but also via technical conferences. For additional information, visit www.rfhyper.com or contact Exposium, 1 rue du Parc, 92593 Levallois Perret, Cedex, France +33 1 49 68 51 00, fax: +33 1 49 68 52 31 or e-mail: rfhyper@exposium.fr.

CTIA Wireless 2006
April 5–7, 2006
Las Vegas, NV

This event focuses on bringing together the enterprise industry with the consumer and vertical markets to exchange ideas, create partnerships and collaborate to further advance wireless telecommunications. This global event draws attendees from dozens of different industries in more than 80 countries around the world, serving every aspect of wireless providers, users, developers, buyers and manufacturers. Fore more information, visit www.ctiawireless.com.

AMTA Europe Symposium
May 1–4, 2006
Munich, Germany

The first European Antenna Measurement Techniques Association (AMTA) Symposium aims to build a strong European foundation in the domain of leading edge technologies for antenna and RCS measurements. Topics will address measurement methods relating to theory and application of antennas and radar scattering, advances in near-field, compact range and RCS measurements, automation in RF measurements, diagnostic methods and characterizing smart antennas. For more information, visit www.amta.org/europe or e-mail the technical coordinator at tech-coord-europe@amta.org.

IEEE MTT-S International Microwave Symposium and Exhibition
June 11–16, 2006
San Francisco, CA

The IMS Symposium will serve as the centerpiece of Microwave Week 2006. Topics: research, development and application of RF and microwave theory and techniques. **Call for Papers**: technical papers for this symposium must be submitted via the IMS2006 Web site (www.ims2006.org). Complete information on how to submit a paper or register for the conference, as well as other information can be found on this site. **Deadline for paper submission: December 2, 2005.** In addition to IMS2006, a microwave exhibition, a historical exhibit, the RFIC symposium and the ARFTG conference will be held during Microwave Week 2006. The technical sessions will run Tuesday through Thursday of Microwave Week. Workshops will be held Sunday, Monday and Friday, and the ARFTG Microwave Measurements Conference will be held on Friday. For exhibition information, contact Kristen Dednah, Horizon House Publications, 685 Canton St., Norwood, MA 02062 (781) 769-9750 or e-mail: kdednah@mwjournal.com.

IEEE Radio Frequency Integrated Circuits Symposium
June 11–13, 2006
San Francisco, CA

The 2006 IEEE Radio Frequency Integrated Circuits Symposium (RFIC 2006) will be held in conjunction with the IEEE MTT-S International Microwave Symposium. **Call for Papers:** papers are solicited describing original work in RFIC design, system engineering, system simulation, design methodology, RFIC circuits, fabrication, testing and packaging to support RF applications in areas such as, but not limited to: cellular system, wireless data system, wideband communication system, optical system ICs and architectures, small-signal, large-signal, frequency generation circuits, RFIC and device technology, and RFIC modeling and CAD. Deadline for paper summary submission: **January 2, 2006**. For more information, visit www.rfic2006.org.

Workshops & Courses

Microwave Measurements for High Performance Devices and Applications

■ **Topics:** This 1½ day course covers microwave measurement fundamentals, including vector network analysis, scattering parameters, power and thermal noise. Practical issues such as cables, fixtures, probes and on-wafer measurements are covered. For more information, visit www.arftg.org.
■ **Site:** Washington, DC
■ **Dates:** November 29–30, 2005
■ **Contact:** David Walker, NIST, M.S. 818.01, 325 Broadway, Boulder, CO 80305 (303) 497-5490 or e-mail: dwalker@boulder.nist.gov.

IEEE WirelessMAN 802.16 Broadband Wireless Technical Review

■ **Topics:** This course provides a comprehensive overview of the modes, features and options supported by the IEEE 802.16 both in PHY and MAC layers. It also describes how different modes can be effectively used to target different applications, usage models and wireless network deployment scenarios. The most popular modes of operation are discussed in detail. For more information, visit www.unex.berkeley.edu.
■ **Site:** Redwood City, CA
■ **Date:** November 30, 2005
■ **Contact:** UC Berkeley Extension, 1995 University Avenue, Berkeley, CA 94720 (510) 642-4151 or e-mail: course@unex.berkeley.edu.

Far-field, Anechoic Chamber, Compact and Near-field Antenna Measurements

■ **Topics:** This course presents the state-of-the-art in antenna measurements, including far-field, anechoic chamber, compact and near-field measurements. The course also includes range evaluation and compensation techniques and microwave holography. For more information, visit www.pe.gatech.edu.
■ **Site:** Atlanta, GA
■ **Dates:** December 5–9, 2005
■ **Contact:** Georgia Institute of Technology, Professional Education, PO Box 93686, Atlanta, GA 30377 (404) 385-3500.

OPERA-3d Software – Introductory Training Course

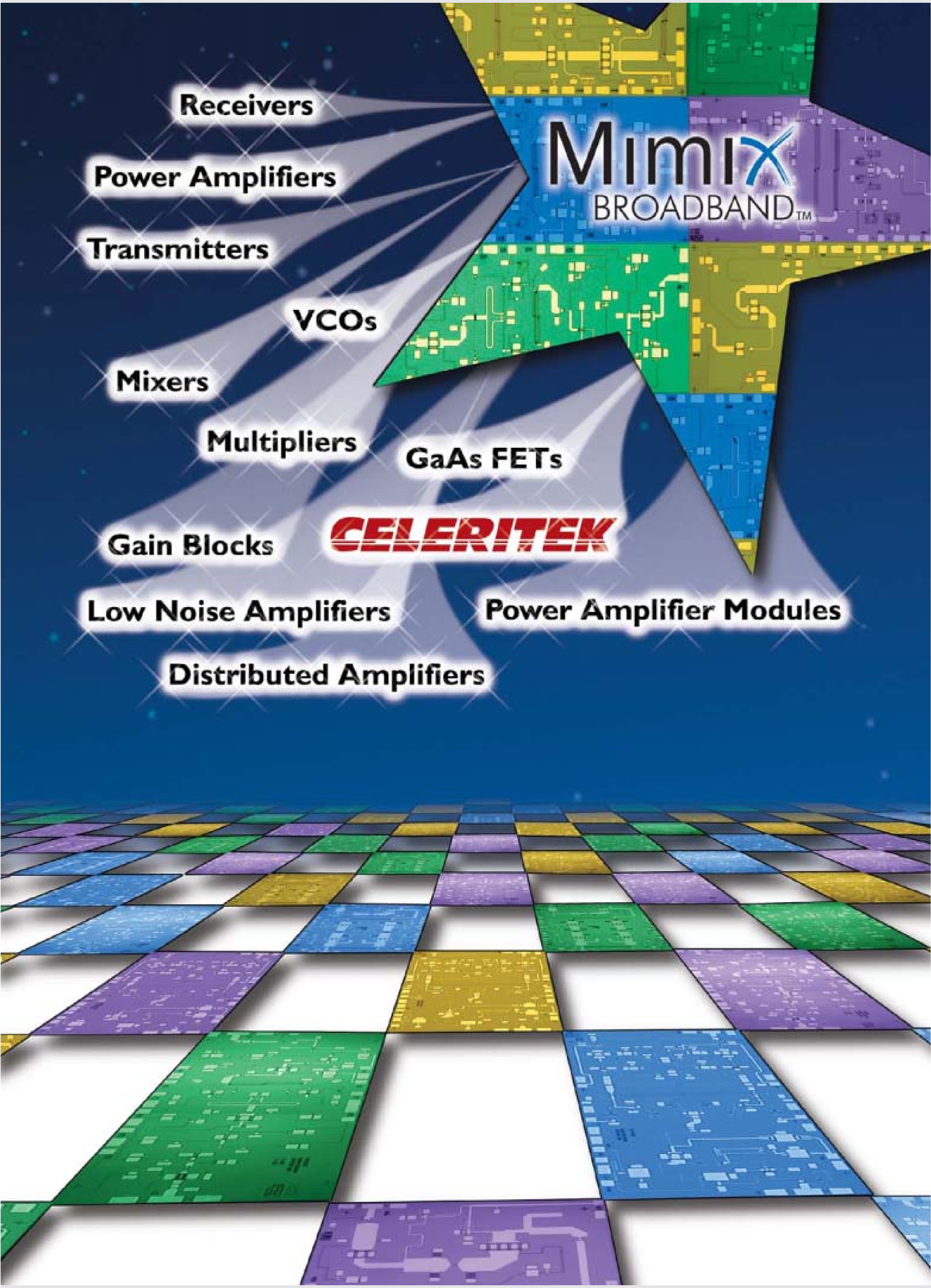
■ **Topics:** The OPERA-3d three-day course provides the user with structured guidance in 3D design. Theoretical aspects are covered in-depth to allow accurate use of the program's own designs, while "hands on" example models illustrate the variety of model creation tools and their efficient use.
■ **Site:** For location information, please visit url.
■ **Dates:** For date information, please visit url.
■ **Contact:** For more information, visit www.vectorfields.com.

Sonnet Training Class

■ **Topics:** This ½ day course is an introduction to the use of Sonnet's 3D planar electromagnetic simulation software. It will offer a hands-on technical training that is designed to bring new users up to speed and highlight new features to experienced users in the area of high frequency design. For more information, visit www.sonnetsoftware.com/support/training.asp.
■ **Site:** For location information, please visit url.
■ **Dates:** For date information, please visit url.
■ **Contact:** Yun Chase, Sonnet Software Inc., 100 Elwood Davis Road, North Syracuse, NY 13212 (315) 453-3096 or e-mail: chase@sonnetsoftware.com.

Integrated Nonlinear Microwave and Millimeter-wave Circuits

■ **Topics:** This workshop leverages its technical program to stimulate the discussion and promotion of new ideas in the field of nonlinear microwave and millimeter-wave circuits, both on active device characterization and modeling strategies, CAD analysis methods and design approaches. For more information, visit www.inmmic.org.
■ **Site:** Aveiro, Portugal
■ **Dates:** January 30–31, 2006
■ **Contact:** José Carlos Pedro, Instituto de Telecomunicações – Pólo de Aveiro, Aveiro, Portugal +351 234 377900, fax +351 234 377901 or e-mail: inmmic@av.it.pt.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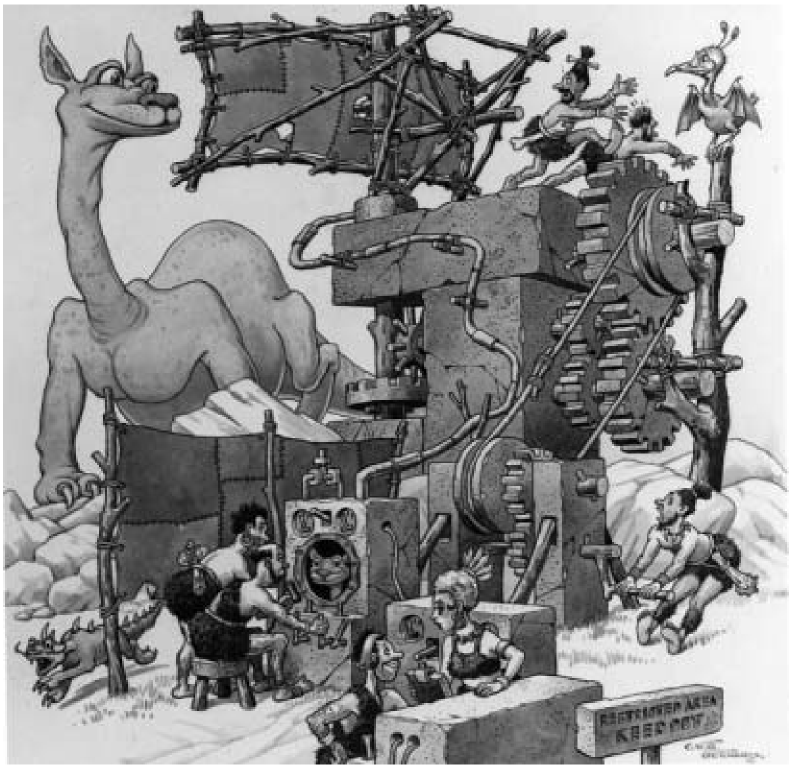
COVER FEATURE

PASSIVE COMPONENTS: A BRIEF HISTORY

The microwave industry is tied to the birth of radar. ***Figure 1*** is a whimsical look at the birth of radar as portrayed in a 1960s ad for Bomac tubes. In reality, radar was developed primarily at the MIT Radiation Laboratory (RadLab) during World War II. Much of the impetus came from the development of the magnetron in the UK, which was the first source of adequate microwave power and which the British provided to RadLab. There was also work going on at Bell Laboratories, Naval Research Labs and Harvard University.

Fig. 1 The birth of radar. (Courtesy of Varian Associates.)



In 1945, at the end of the war and the closing of the RadLab, there were very few active microwave devices. Power sources were limited to triodes, klystrons and magnetrons. The plasma TR tube was the only active switch and the only semiconductors were point contact germanium and silicon diodes. There were no varactors, PINs, bi-polar transistors, FETs, Gunn diodes, IMPATTS or Schottky diodes. Thus, almost all system functions were relegated to passive components. The development of those passive components was the genesis of the microwave industry as we know it today.

THE PIONEERS

The technical staff at the RadLab was a remarkable collection of highly skilled and motivated physicists and engineers. Much of their work, like that of Julian Schwinger, shown lecturing in ***Figure 2***, endures to this day. Seven of them went on to win Nobel Prizes and five became National Science Advisors. Of more importance to us, many took the entrepre-

HARLAN HOWE, JR.
Microwave Journal

▲ *Fig. 2 Julian Schwinger presents new theory at the RadLab. (Courtesy of MIT Museum.)*

neurial path and started their own companies when the RadLab closed. There were also many large companies that were manufacturing RadLab designed radars. They included RCA, Westinghouse, Raytheon, Sperry Gyroscope and Western Electric, not to mention Bell Laboratories. A number of the engineers in those organizations followed the same path, either joining or starting new companies. Just a few of the notable pioneers and their companies (not in any particular order) include:

Sigurd and Russell Varian – Varian Associates
Henry Riblet – Microwave Development Laboratories (MDL)
Ted Saad – Sage Laboratories
Harold Wheeler – Wheeler Labs (later Hazeltine)
Richard Walker – Microwave Associates
Tore Anderson – Airtron
Art Oliner – PRD
Bruno Weinschel – Weinschel Labs
Marion Hines – Microwave Associates
Joe Saloom – SFD Laboratories
George Southworth – Bell Labs
Bill Mumford – Bell Labs
Seymour Cohn – Stanford Research Institute and Rantec
Leo Young – Stanford Research Institute
George Matthaei – Stanford Research Institute

Following them was a second wave of younger but still significant contributors too numerous to list.

TRANSMISSION MEDIA

The concept of hollow tube waveguides goes back to Lord Rayleigh in 1897; however, the idea was not developed until George Southworth and Wilmer Barrow, after several years of independent research, held separate public demonstrations in 1936. When the rectangular waveguide was employed during WWII, many of the dimensions that became standard waveguide sizes were based on the availability of commercial extruded tubing used for decorative and architectural purposes. While much of the early component development was done in waveguide, its usage today is primarily for high power, millimeter-wave or very low loss applications, since smaller, lower cost techniques have evolved over the years and waveguide has limited bandwidth capability.

Heinrich Hertz demonstrated propagation in coaxial lines several years before Lord Rayleigh proposed hollow waveguides. He also demonstrated the concept of "skin effect," which showed that high frequency waves only penetrated the conductor to a very limited depth. Coaxial lines had the benefit of complete shielding compared to open wire systems in use at that time. Coaxial lines for short wave radio came into common use in the early 1930s, first as antenna lead-in cables and later as low loss, air dielectric line with bead supports. These air dielectric structures were the basis for the early development of coaxial components. Bandwidth is from DC to some upper frequency where the dimensions permit higher order modes beyond the fundamental TEM mode to propagate.

A variation on coaxial lines is a round or square conductor in between two extended ground planes. This was dubbed 'Slab-line.' This configuration allowed access to the center conductor from the side with minimal leakage. Hewlett-Packard produced one of the first coaxial slotted lines using this technique. Slab-line was the basis for many coaxial couplers, hybrids and filters, particularly for test equipment components where high performance was needed. It is still a common media for these components today, although the requirement for precision machining of the parts tends to raise the cost.

The idea of using flat printed circuits at microwave frequencies was first reported by Barrett & Barnes in 1951.[1] Bob Barrett was responsible for encouraging research contracts in the field. One of the contractors was Airborne Instruments Laboratory, which developed a printed line on a

Fig. 3 *A comparison of Tri-Plate Modules to waveguide. (Courtesy of Sanders Associates.)*

thin dielectric support suspended between two plates using air as the dielectric. They called it Stripline and registered that name as a trademark. Similar work was also under way at Sanders Associates. They used two boards, one with a pattern and the other as a cover, thus using the boards as the dielectric. They called it

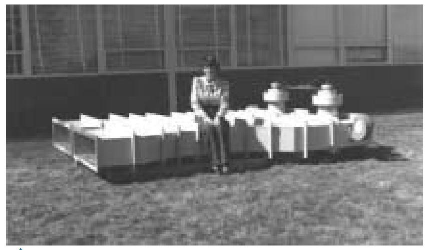


Fig. 4 *A high power duplexer using a sidewall hybrid in WR-2100 waveguide.*

Tri-Plate and registered that name as their trademark. They also introduced a product line of circuit modules using the technique, and in 1956 published the *Handbook of Tri-Plate Components*, which was an invaluable resource for early designers, such as myself. I suspect that they regretted printing it, since there was never a second printing or a second edition. ***Figure 3*** is a photo from the handbook of the products compared to a waveguide assembly.

In 1952, Grieg & Englemann[2] of the Federal Communications Research Laboratories published a paper describing a single board with a ground plane on one side and the pattern on the other. They called it Microstrip and also registered the name as a trademark. In due course of time, any printed double ground plane line became described by the generic term "stripline," the single ground plane became the generic term "microstrip" and the trademarks were ignored.

One of the problems with the new flat transmission lines was the launching connectors. The connectors in use at that time, such as UHF, N, BNC and TNC, were mechanically large and were not constant-impedance designs. The engineers at the Bendix Research Labs developed a small 3 mm connector they called the Bendix Real Miniature (BRM). This was later refined by Omni-Spectra as the Omni-Spectra Miniature (OSM®) and was subsequently produced by many sources as the 3 mm SMA. This development spurred the use of stripline and microstrip at much higher frequencies and spawned the whole new generation of high frequency connectors in common use today.

PASSIVE COMPONENTS

Hybrids

The workhorse of passive components is the 90° hybrid. It is used in many forms for mixers, switches, diplexing and duplexing, power division, phase shifting and matching reflective circuits. It is a four-port device that provides an even –3 dB split with a 90° phase difference at the output ports with the fourth port isolated. The waveguide sidewall and topwall hybrid was developed in 1950 by Henry Riblet at MDL. Tens of thousands have been sold as investment castings, along with cast bends and other components to be brazed or soldered into waveguide assemblies. Sidewall hybrids have even been fabricated from sheet metal for large waveguides such as the monster WR-2100 hybrid shown in ***Figure 4***. In stripline form, they can be done as direct coupled branch line circuits or as coupled line circuits. The branch arm type has limited bandwidth; however, the coupled line type operates over an octave for a single section. This can be extended by the use of multiple sections and by tandeming sections, thus permitting bandwidths as great as 20:1. Much of this

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Broadband Amplifiers by AML Communications

To Order Call: (805)388-1345 ext. 203

Model	Frequency (GHz)	Gain (dB)	Flatness (dB) max	NF (dB) max	P1dB (dBm) min	VSWR (In/Out)	DC Current @ +12/+15VDC
Broadband Low Noise Amplifiers							
AML016L2802	0.1 – 6.0	28	±1.25	1.3*	+7	2.0:1	190
AML48L3001	4.0 – 8.0	30	±1.0	1.2	+10	1.8:1	150
AML412L3002	4.0 – 12.0	30	±1.5	1.5	+10	1.8:1	150
AML218L0901	2.0 – 18.0	9	±1.0	2.2	+5	2.5:1	60
AML0518L1601-LN	0.5 – 18.0	16	±1.0	2.7	+8	2.2:1	100
AML0126L2202	0.1 – 26.5	22	±2.25	3.5*	+8	2.2:1	170
AML1226L3301	12.0 – 26.5	33	±2.0	2.8	+8	2.5:1	200
Broadband Medium Power Amplifiers							
AML0016P2001	0.01 – 6.0	21	±1.25	3.2*	+23*	2.0:1	480
AML26P3001-2W	2.0 – 6.0	28	±2.5	6	+33	1.8:1	1500
AML28P3002-2W	2.0 – 8.0	30	±2.0	5.5	+33	2.0:1	2000
AML218P3203	2.0 – 18.0	32	±2.5	4	+25	2.0:1	450
AML618P3502-2W	6.0 – 18.0	35	±2.5	4	+33	2.0:1	1850
Narrow Band Low Noise Amplifiers							
AML23L2801	2.8 – 3.1	28	±0.75	0.7	+10	1.8:1	150
AML1414L2401	14.0 – 14.5	24	±0.75	1.5	+10	1.5:1	130
AML1718L2401	17.0 – 18.0	24	±0.75	1.6	+10	1.8:1	150

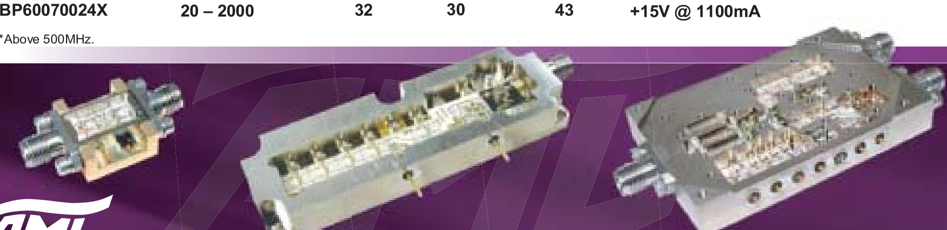
Low Phase Noise Amplifiers — **Phase noise (dBc/Hz) at offset**

Part Number	Frequency (GHz)	Gain (dB)	Output Power (dBm)	100Hz	1KHz	10KHz	100KHz
AML811PN0908	8.5 – 11.0	9	17	-154	-159	-167	-170
AML811PN1808	8.5 – 11.0	18	18	-152.5	-157.5	-165.5	-168
AML811PN1508	8.5 – 11.0	15	28	-145.5	-153.5	-158.5	-164.5
AML26PN0904	2.0 – 6.0	9	20	-150	-165	-165	-178
AML26PN1201	2.0 – 6.0	11	15	-155	-160	-160	-175

High Dynamic Range Amplifiers

Part Number	Frequency (MHz)	Gain (dB)	P1dB (dBm)	OIP3 (dBm)	DC
AR01003251X	2 – 32	21	32	52	+28V @ 470mA
AFL30040125	50 – 500	23	28	53	+28V @ 700mA
BP60070024X	20 – 2000	32	30	43	+15V @ 1100mA

*Above 500MHz.



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Model	Frequency (GHz)	Psat (dBm)	Psat (W)	P1dB (dBm)	Gain (dB)	DC Current(A) @ +12V
Broadband Microwave Power Amplifiers						
L0104-43	1 - 4	42.5	17.8	41.5	45	14
L0204-44	2 - 4	44	25	42.5	45	14
L0206-40	2 - 6	40	10	38.5	40	8.5
L0218-30	2 - 18	30	1	29	30	3
L0408-43	4 - 8	43	20	41.5	45	17
L0618-43	6 - 18	43	20	41.5	45	22
L0812-44	8 - 12	44	22	42	45	22
L1218-43	12 - 18	43	20	41.5	45	22
Millimeter-Wave Power Amplifiers						
L1826-34	18 - 26	34	2.5	33	35	4
L1840-27	18 - 40	27	0.5	26	30	2
L2632-37	26 - 32	37	5	36	38	10
L2640-27	26 - 40	27	0.5	26	30	2
L2630-37	26.5 - 30.5	37	5	36	38	10
L2732-35	27 - 32	35	2.8	33	35	6
L3040-30	30 - 40	30	1	29	35	4
L3236-36	32 - 36	36	4	35	40	12
L3640-36	36 - 40	36	4	35	40	10

High-Power Rack Mount Amplifiers

Model	Frequency (GHz)	Psat (dBm)	Psat (W)	P1dB (dBm)	Pac (kW)	Height (in)
C071077-52	7.1 - 7.7	52.5	170	51.5	1.8	10.25
C090105-50	9 - 10.5	50	100	49	1	8.75
C140145-50	14 - 14.5	50.5	110	49.5	2	10.25
C1416-46	14 - 16	46	40	45	0.35	5.25
C1820-43	18 - 20	43	20	41.5	0.25	5.25
C2326-40	23 - 26	40	10	39	0.25	5.25
C2630-40	26.5 - 30.5	40	10	39	0.25	5.25
C3236-40	32 - 36	40	10	39	0.25	5.25
C3640-39	36 - 40	39	8	38	0.24	5.25

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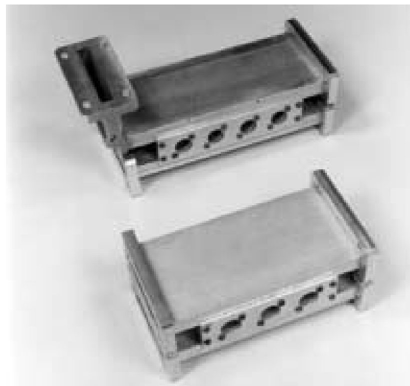


▲ *Fig. 5 High power branch arm couplers in ½ height waveguide.*

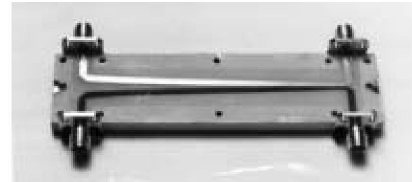
work was done by Joe Mosko at the USN Ordinance Test Station at China Lake, CA. A technique to improve the performance of these multi-section circuits by reducing the discontinuity between sections using non-uniform lines was developed by Carl Tresselt and reported in 1966.[3]

While realizing the coupled line hybrid was fairly simple in stripline, it was very difficult in microstrip. This problem was solved in 1969 by Julius Lange at Texas Instruments. He developed an interdigital coupler for microstrip,[4] which bears his name and is still used universally.

The companion to the 90° hybrid is the 180° hybrid. It provides an even –3 dB split at 0° phase at the colinear ports and a similar split with a 180° phase relationship from the difference port. It was invented in its waveguide form in the late 1930s as a bridge by W.L. Barrow, who called it a Magic Tee. It is generally fabricated as a casting, as is the 90° version. In stripline or microstrip, it is usually made as a 1½ λ circumference ring with ports at 0°, 90°, 180° and 270°, and is commonly called a "rat-race" because of its appearance. It is limited in bandwidth like the 90° branch hybrid. A broadband version for stripline was reported by DuHamel and Armstrong in 1965.[5] However, to the best of my knowledge, no one has solved the bandwidth problem in microstrip.

DIRECTIONAL COUPLERS AND POWER DIVIDERS

Branch arm waveguide directional couplers are frequently used in high power applications, like the ones shown in ***Figure 5***. However, most waveguide couplers are made by utilizing holes or slots in the common wall between two guides. There are a myriad number of these configurations, many of them named after the engineers who first developed them. The Bethe Hole Coupler was reported in the RadLab Series along with many other types. The first multi-hole coupler was described by Bill Mumford in 1944. This was followed by the Schwinger coupler and the Riblet & Saad coupler. Binomial distribution of a large number of holes resulted in very high directivity couplers, which are still used today in most test equipment, where size is not a consideration. TEM Mode couplers in coaxial lines, as well as stripline and microstrip had their start with a single section, quarter-wave coupler, first shown by Harold Wheeler in 1944. This type of coupler has a usable bandwidth of one octave. Many variations were reported through the years. The bandwidth can be extended by using multiple quarter-wave sections, either symmetrically or non-symmetrically. A drawback to multiple sections is decreased directivity due to the discontinuities at the many interfaces. This was improved by the use of non-uniform lines, as previously mentioned and by tapered lines such as the one shown in ***Figure 6***. Unlike stripline, side-coupled lines in microstrip suffered from reduced directivity due to the difference in propagation velocity between the even and odd modes. This problem has been addressed using dielectric overlays and a unique "Wiggly Line" coupler, described by Alan Podell in 1970.[6]



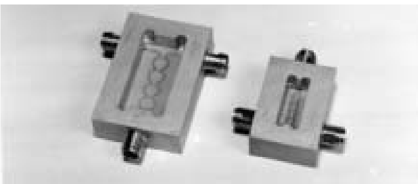
▲ *Fig. 6 An asymmetric tapered-line coupler.*

Power dividers as we know them today originated in 1960 with the introduction of the Wilkinson divider.[7] It was a single section equal N-way divider, matched at all ports with isolation between the output ports. It was limited to one octave bandwidth. In 1965, Parad and Moynihan introduced a similar structure for unequal division.[8] Seymour Cohn reported on a multi-

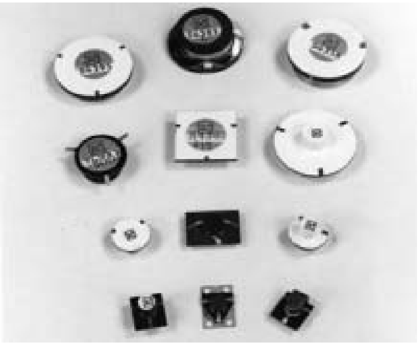
section design with multi-octave bandwidth in 1968.[9] ***Figure 7*** shows four section dividers built in microstrip.

FILTERS

More papers and books have been written on the subject of filters than any other passive device. In their review paper on the subject in the centennial issue of *MTT-S Transactions*, Ralph Levy and Seymour Cohn list 91 references and that was back in 1984.[10] However, the "Bible" on the subject was, and still is, *Microwave Filters, Impedance-Matching Networks and Coupling Structures* by Matthaei, Young and Jones. It was published by McGraw-Hill in 1964. It went out of print briefly but it was rescued and re-printed by Artech House Inc. in 1980. It is still available today and belongs on every engineer's bookshelf.[11]

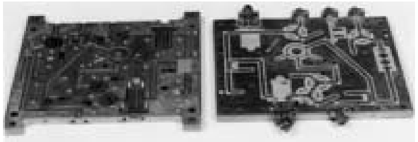


▲ *Fig. 7 Four section power dividers.*



▲ *Fig. 8 Early drop-in circulators.*

Filters have been made in every conceivable transmission line, from cavities to waveguide and all of the various TEM-mode configurations. Much of the early work was done during WWII at RadLab, Harvard University and Bell Labs, and most modern filter theory is still derived from this early work. There are so many contributors that I cannot distill the work into this short article. I suggest that readers interested in more detail go to the referenced Levy and Cohn paper and start from there.



▲ *Fig. 9 A Ferrotec stripline assembly with ferrites.*

FERRITES

The first demonstration of a microwave ferrite device occurred in 1949, but practical devices were yet to come. Early devices were based on the principle of Faraday rotation, followed by waveguide isolators of several types. The concept of the three-port circulator was presented by H.J. Carlin in 1954[12] and was refined for stripline by H. Bosma in 1961.[13] In the beginning all these devices were built as single components, mostly because of the large magnetic circuits that were needed to make them work, which made them difficult to integrate. With the increased use of stripline for subassemblies in the '60s there was a need to be able to integrate ferrites. Melabs (later acquired by Microwave Associates) introduced a line of drop-in circulators with the unofficial name of "Flying Saucers" because of their shape. ***Figure 8*** shows a selection of these circulators. The flying saucer had serious technical problems due in large part to the problem of maintaining ground plane continuity as well as the lack of a magnetic return path, which made them more susceptible to performance variation due to nearby ferrous objects and to temperature changes. In the late '60s, Ken Carr at Ferrotec (later acquired by Microwave Associates) came up with a technique for direct integration of ferrite circulators in stripline subassemblies. He used a suspended substrate line in a channeled construction. The ferrite disks were mounted in the channels, which were then lapped to provide uniform contact throughout the assembly. By laying out the circuit with alternate directions of rotation, the magnetic path of one circulator was returned through another, thus providing a shielded structure and temperature stability. ***Figure 9*** shows one of these early subassemblies, some of which are still being produced today. However, with the widespread use of microstrip for subassembly work, due to

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TC1.5-1+	1.5	D	.5-2200	2-1100	1.59
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TC3-1T+	3	A	5-300	5-300	1.29
TC4-1T+	4	A	.5-300	1.5-100	1.19
TC4-1W+	4	A	3-800	10-100	1.19
TC4-14+	4	A	200-1400	800-1100	1.29
TC8-1+	8	A	2-500	10-100	1.19
TC9-1+	9	A	2-200	5-40	1.29
TC16-1T+	16	A	20-300	50-150	1.59
*TC4-11+	50/12.5	D	2-1100	5-700	1.59
*TC9-1-75+	75/8	D	0.3-475	0.9-370	1.59

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TCML1-19+	1	G	800-1900	900-1400	1.09
TCM2-1T+	2	A	3-300	3-300	1.09
TCM3-1T+	3	A	2-500	5-300	1.09
TTCM4-4+	4	B	0.5-400	5-100	1.29
TCM4-1W+	4	A	3-800	10-100	.99
TCM4-6T+	4	A	1.5-600	3-350	1.19
TCM4-14+	4	A	200-1400	800-1000	1.09
TCM4-19+	4	H	10-1900	30-700	1.09
TCM4-25+	4	H	500-2500	750-1200	1.09
TCM8-1+	8	A	2-500	10-100	.99
TCM9-1+	9	A	2-280	5-100	1.19

* Step down transformer. TC+ and TCM+ Dimensions (LxW): 0.15" x 0.16" •Referenced to midband loss.

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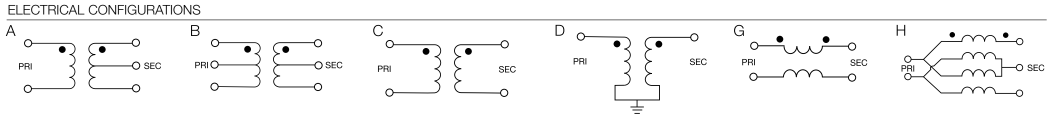

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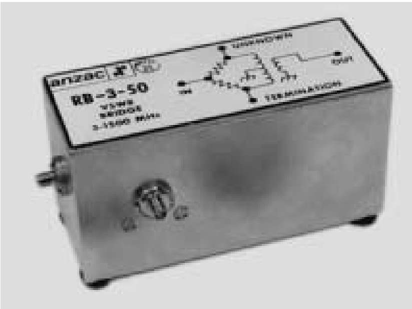

▲ *Fig. 10 An early transformer bridge.*

the ease of semiconductor integration, most ferrite components have reverted back to drop-ins, which have been vastly improved over the years.

MIXERS

Mixers date back to the RadLab. In most cases they were balanced types made by putting diode mounts onto either 90° or 180° hybrids in coaxial lines as well as waveguides. There were some single-ended mixers and at low frequencies some double-balanced mixers were made using transformers. Most of the improvements in mixer technology resulted in the use of new and better diodes to replace the early point contact devices. A major breakthough in microwave mixer design was the multi-octave, doubly balanced mixer introduced by Don Neuf at RHG Laboratories in the early 1960s. He built a 2 to 20 GHz mixer using orthogonal tapered microstrip baluns with a ring of four diodes in the third plane. It was difficult to fabricate, but nothing could touch it for performance. As monolithic ring diode assemblies came along, refinements were made and performance and manufacturability improved. This basic configuration is still the basis for most doubly balanced mixers today, although other configurations such as the star mixer are in use for circuits that require a high IF frequency.

TOROIDAL COMPONENTS

Much of the early work on ferrite toroidal transformers was done at Bell Labs for telephone use at low frequencies. Two of the early pioneers in bringing those concepts to the microwave world were Carl Sontheimer and Alan Podell at Anzac and Adams Russell (both companies later merged and were aquired by M/A-COM). Using time domain analysis, Sontheimer made some components with 500:1 bandwidths. ***Figure 10*** shows an early impedance bridge, which worked from 3 to 1500 MHz. As ferrite materials were improved, the upper working frequency for this class of device increased to 6 GHz and higher. However, it was Harvey Kaylie at Mini-Circuits who brought the technology to widespread use through manufacturing techniques and aggressive marketing that resulted in cost reduction and off-the-shelf availability.

CONCLUSION

This brief history is far from complete. I have not addressed rotary joints, electromechanical switches or tubular filters because very little historical material has been published and I do not have any personal experience in those areas. I will be happy to hear from anyone who can enlighten me. What I have done is to try to give a personal view from my areas of experience. For those who would like

a more complete and balanced history, there are a number of sources that I have listed below.

ADDITIONAL HISTORICAL MATERIAL

- Special centennial issue of *IEEE Transactions on Microwave Theory and Techniques*, Volume MTT-32, No. 8, September 1984.
- "Five Years at the Radiation Laboratory," originally presented to members of the RadLab in 1946, reprinted for the 1991 MTT-S International Microwave Symposium, Boston, MA.
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12. H.J. Carlin, "Principles of Gyrator Networks, *Proceedings Modern Advances in Microwave Techniques*, Polytechnic Institute of Brooklyn, November 1954, p. 175.
13. H. Bosma, "On the Principle of Stripline Circulation," *Proceedings of the IEEE*, Vol. 109B, 1961, p. 137.

Harlan Howe, Jr. *received his BS degree in optics from the University of Rochester in 1957. He has been actively engaged in the microwave industry for 48 years, first as a design engineer and then as an engineering manager. In 1990, he became the publisher/editor of* Microwave Journal. *He retired as publisher in 2001, but remains the editor. He is a Life Fellow of IEEE, past president of MTT-S and the recipient of an IEEE Third Millennium Medal in 2000 and the MTT-S Distinguished Service Award in 2005.*

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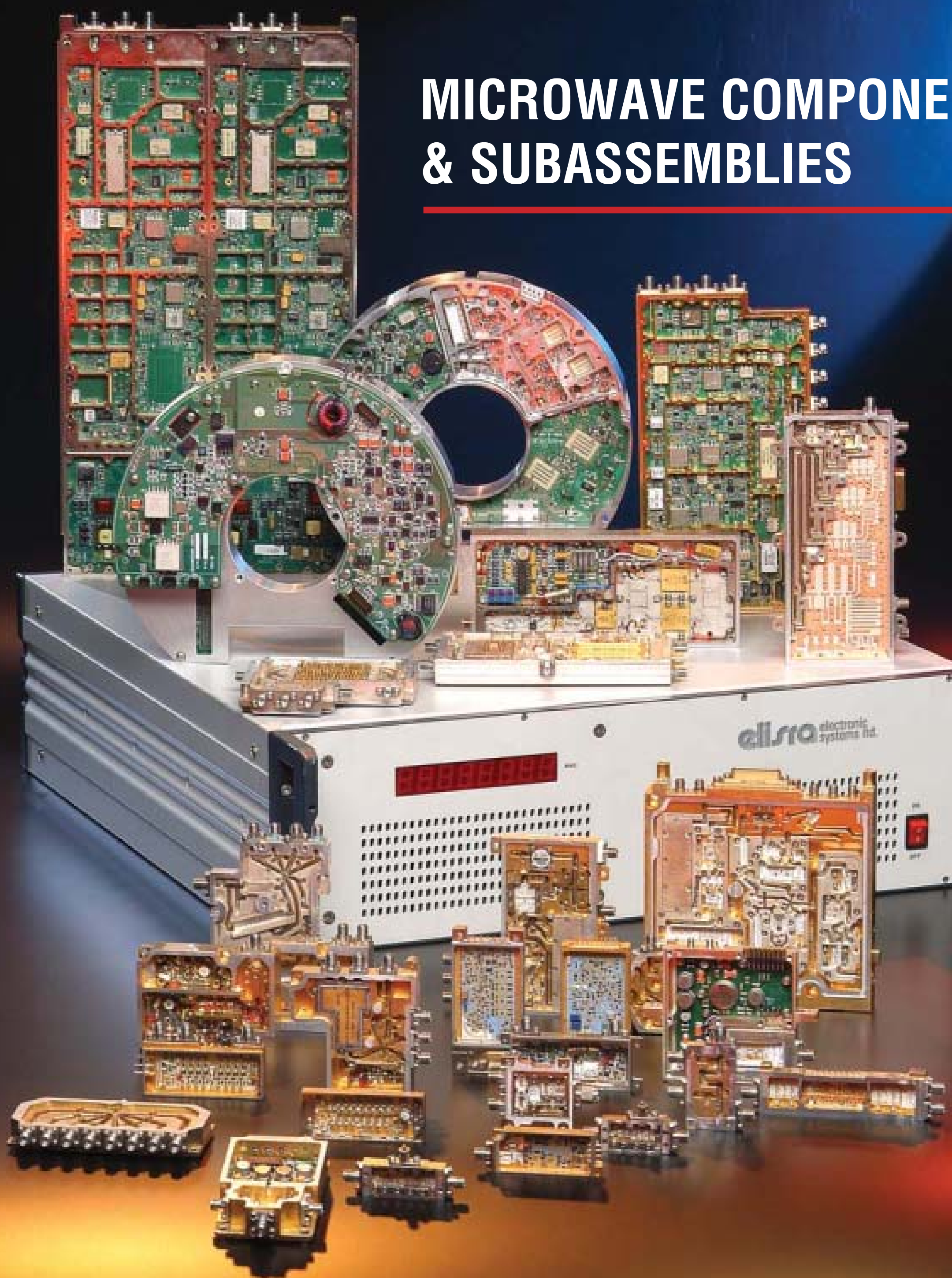

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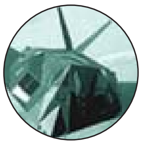


Lockheed Martin Deploys First Satellite Supporting Space-based GPS

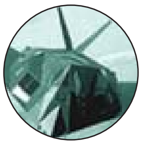
A satellite that will enable future air navigation enhancements has been deployed by Lockheed Martin. The payload is the first use of the GPS L5 civil navigation frequency for the Federal Aviation Administration (FAA), and supports a move toward satellite-based navigation to make air traffic management safer, more reliable and more accurate. It is also the first deployment of this technology for civil aviation use in North America. Lockheed Martin contracted with satellite service providers Telesat and PanAmSat to host navigational payloads it will operate as part of the FAA's Wide Area Augmentation System (WAAS) Geostationary Communications and Control Segment (GCCS) initiative. The successful launch of the Telesat Anik F1R satellite from the Baikonur Cosmodrome in the Republic of Kazakhstan is the first of the two GCCS payloads scheduled for deployment this year to support the program. The navigation payload of the Anik F1R satellite establishes a vital base for providing en route and vertical guidance to aircraft at thousands of North American airports and airstrips. Under the FAA's GCCS contract, Lockheed Martin and its teammates will provide ground uplink stations that receive global positioning system (GPS) correction and integrity data from the WAAS monitoring network and broadcast the data to the geostationary communications satellites for delivery to users. In aviation use, a WAAS broadcast message allows an improvement of GPS-base position accuracy from 30 m (100 ft) to approximately 7 m (27 ft). This improved accuracy enables instrument landing operations at previously unsupported airfields. Lockheed Martin and the FAA will perform 12 months of segment and system level integration and test, prior to the WAAS GCCS service going operational in September 2006. The FAA contract supporting these enhancements is valued at $314 M.

Harris Corp. Demonstrates Military Satellite Antenna Feed Prototype

Harris Corp. announced it has successfully completed multiple field demonstrations of a satellite antenna feed prototype that is capable of supporting selectable antenna polarization for orthogonal transmit and receive operations. The feed prototype achieved user data rates reaching an astounding 105 Mbps — more than 12 times the current satellite terminal capability — using a modified, Harris-built Lightweight High Gain X-band Antenna (LHGXA) and a modified AN/TSC-85C terminal operating over the recently launched XTAR-EUR commercial X-band satellite. The new selectable X-band polarization feed can be easily installed on existing LHGXA antennas and will be the standard offering on Harris' newest tactical satellite antenna, the Large Aperture Multiband Deployable Antenna (LAMDA). The 4.9 m LAMDA supports satellite communications connectivity over commercial C- and Ku-band satellites, as well as both military X-band satellites (DSCS, WGS, NATO and SKYNET) and the new XTAR satellites. More than 125 LHGXA are currently on the DoD inventory, providing reliable, rugged, high performance and highly mobile satellite communications connectivity to military personnel worldwide. The user-friendly, large-aperture antenna, with its 4.9 m (16 ft) diameter reflector, has the equivalent performance characteristics (that is G/T, EIRP) of a 20-foot reflector due to its shaped offset-fed design. XTAR-EUR, built by Space Systems/Loral (SS/L), was launched in February 2005 and carries twelve 72 MHz, high power X-band transponders that provide coverage from Eastern Brazil and the Atlantic Ocean, across all of Europe, Africa and the Middle East to Singapore. It is expected to provide services for nearly 20 years and is fielded by XTAR LCC, a new satellite communications company committed to serving the long-haul communications, logistics and infrastructure requirements of the US allied governments.

Northrop Grumman, Raytheon Closer to Flight Testing of New Radar Antenna

Northrop Grumman, working closely with Raytheon Co., has reached three milestones in a program to modernize the B-2 stealth bomber's radar system with an advanced, more reliable antenna. These achievements represent significant progress towards initial flight testing of the radar. Recently, the B-2 radar-modernization team passed a final design review by the US Air Force, delivered the first test model of the radar for integration, test and software development, and completed a suite of tests that proves the hardware and software work together as one subsystem. Northrop Grumman, prime contractor for the overall B-2 program, also leads the radar-modernization team that includes Raytheon, the radar system provider. This effort will replace the current antenna with an active, electronically scanned array (AESA) antenna. The final design review concluded in late May with the Air Force finding the radar system's design in compliance with engineering and performance requirements. In July, Raytheon delivered on schedule the first model of its APQ-181 radar. The APQ-181 is one of a series of revolutionary AESA radar systems that are replacing mechanically scanned antennas with faster and more reliable solid-state arrays. Following the radar delivery, the B-2 team completed initial radar-subsystem integration and acceptance testing ahead of schedule, paving the way for higher level systems integration and performance tuning. The current B-2 radar work is part of a $383 M system development and demonstration (SDD) contract awarded by the Air Force in 2004. During the SSD phase, Northrop Grumman and Raytheon are developing and testing the



radar and will install six systems on operational B-2 aircraft of the 509th Bomb Wing at Whiteman Air Force Base, MO. This phase will be followed by production to field the new radar and install the new antenna into the B-2 fleet.

Raytheon Delivers Non-lethal Sheriff Active Denial System

Raytheon Co. is helping transform US war fighting capabilities by delivering a short-range millimeter-wave directed energy non-lethal weapon to the Department of Defense's Full Spectrum Effects Platform (FSEP) program — also known as Project Sheriff — for the Office of Force Transformation (OFT). OFT, in partnership with the US Army's Futures Center and the Naval Surface Warfare Center (NSWC), is developing an operational prototype to provide forces fighting in complex urban environments with new options and a combination of lethal and non-lethal capabilities in responding to threats. By developing an operational package of vehicles that can be quickly outfitted and deployed in active operations, OFT will provide combat forces the opportunity to more quickly test tactics, techniques and procedures for new urban combat capabilities.

Applying streamlined processes and rapid prototyping tools to meet OFT's transformational rapid reaction goals, Raytheon Missile Systems designed, developed, fabricated and delivered the combat-hardened, non-lethal system for Sheriff in less than six months. An integrated Raytheon product team instituted innovative parallel scheduling in delivering the product on time and on budget successfully, while still adhering to a thorough quality review process. The breakthrough non-lethal capability Raytheon has delivered to FSEP will stop, delay, deter and turn back an adversary. It is intended to save lives by reducing unnecessary casualties and collateral injuries. It brings new operational flexibility and speed of light weapons to the war fighter. In the compressed urban environment, discriminating threats from non-threats is difficult and can require split second scalable response. The millimeter-wave energy beam can help discriminate the threat and assess the intent of an aggressor with a temporary reversible effect whose safety has been established and demonstrated in more than 12 years of testing by the Air Force Research Laboratory with sponsorship from the Joint Non-lethal Weapons Directorate. Raytheon is currently assisting NSWC with the integration of the short-range non-lethal system into a Stryker combat vehicle. Vehicle testing will take place at the Naval Surface Warfare Center, Dahlgren, VA; Quantico, VA; and Yuma Proving Grounds, AZ, before the system is turned over to officials at the Infantry Center at Ft. Benning, GA. ■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INTERNATIONAL REPORT

Richard Mumford, European Editor

Joint Venture Sets Up R&D Centre in India

In order to serve the fast growing needs of the Indian market for broadband connectivity, Alcatel and the Centre for the Development of Telematics (C-DOT), the Indian Government's telecom technology development centre, have announced a joint venture to design and develop products related to broadband wireless. In fact, the Indian Union Cabinet approved the joint venture within five months of the signing of the MoU between the two companies, typifying the Government's commitment to e-governance and other IT initiatives in rural areas. The new company, called CDOT Alcatel Research Centre, will be 51 percent owned by Alcatel and 49 percent by C-DOT.

The Chennai-based joint venture company will be a global research and development centre for broadband wireless products with a primary focus for rural, urban and suburban Indian markets, in compliance with the country's broadband requirement and spectrum allocation. This includes the end-to-end architecture and the radio access network based on Alcatel WiMAX 802.16e.

It is envisaged that the joint development of the latest broadband wireless technologies will facilitate the industrialization and volume production of the WiMAX solutions in India, thus helping the telecom-manufacturing sector. This will include the integration of the systems in the core network, the applications platform and the customer terminals adapted to the specifics of the targeted markets, as well as the piloting and showcasing of newly developed technologies through the setting up of a technology reality centre.

Innos Joins Investigation into Materials Imaging Technique

Innos, a UK research and development company delivering expertise in silicon, MEMS and nanotechnologies, has announced its involvement in an ongoing EPSRC-funded project led by Queen Mary College London. The project is investigating the development of a new imaging technique — Scanning Photo-induced Impedance Microscopy (SPIM). With support from the company's world-class clean room, SPIM measurements with good resolution and high sensitivity have been demonstrated.

The new SPIM imaging technique has potential in the investigation of the local dielectric properties of materials and biological specimen with good lateral resolution. Application areas include characterisation of smart materials and the development of new array technology for high throughput screening or sensing. It also can be used for investigation of cell-surface interactions, which are currently difficult to access with other techniques.

The best results obtained to date involved using thin silicon membranes from back etched silicon on insulator (SOI) wafers and thin, single crystalline silicon layers on silicon on sapphire (SOS) with a thin, thermally grown oxide; the latter even indicating the potential of submicrometer resolution.

Commenting on the close collaboration between the two companies, Dr. Steffi Krause from the department of materials at Queen Mary College said, "Innos has helped us to develop suitable semiconductor and insulator substrates that allow SPIM measurements with good resolution and high sensitivity." She added, "We have improved resolution and sensitivity by reducing the insulator impedance and testing alternative semiconductor substrates. As part of the ongoing work we will be focusing on testing amorphous silicon, which will be grown at Innos."

R&S Goes Extra Terrestrial in Korea

Rohde & Schwarz has been commissioned to supply terrestrial digital multimedia broadcasting (T-DMB) transmitter systems for the world's first T-DMB network in Korea. All of the various network operators involved in setting up the T-DMB transmitter networks in the country are being supplied with transmitter equipment by R&S. The system is used to transmit radio and TV programs as well as data to mobile receivers.

After a successful test phase with the company's air-cooled transmitters, the network operator Korean Broadcasting System (KBS) ordered liquid-cooled T-DMB transmitter systems with an output power of 1.4 and 2.3 kW for regular operation. Rohde & Schwarz is also supplying further transmitters with air and liquid cooling for Seoul Broadcasting System (SBS), MBC and YTN.

The complete T-DMB transmitter systems are designed to meet high fail-safety requirements as all have passive standby, while a special redundant system of combiners has been developed to increase reliability. The equipment is currently being supplied with installation expected to be completed by the end of 2005.

Ericsson Sets Sail on Maritime Venture

SeaMobile Inc. and Ericsson have signed a three-year contract under which the companies will work together to provide wireless voice and data services at sea. Ericsson will supply a complete wireless network, including hosted core infrastructure and GSM radio equipment.

Using this technology SeaMobile will offer global wireless communications at sea through leading suppliers of maritime satellite services. Radio base stations will be placed onboard ships, and communication will be secured through satellite links to the hosted core network on shore. Initial activities have included integration to Ericsson's hosted core network and Network Operations Centre.

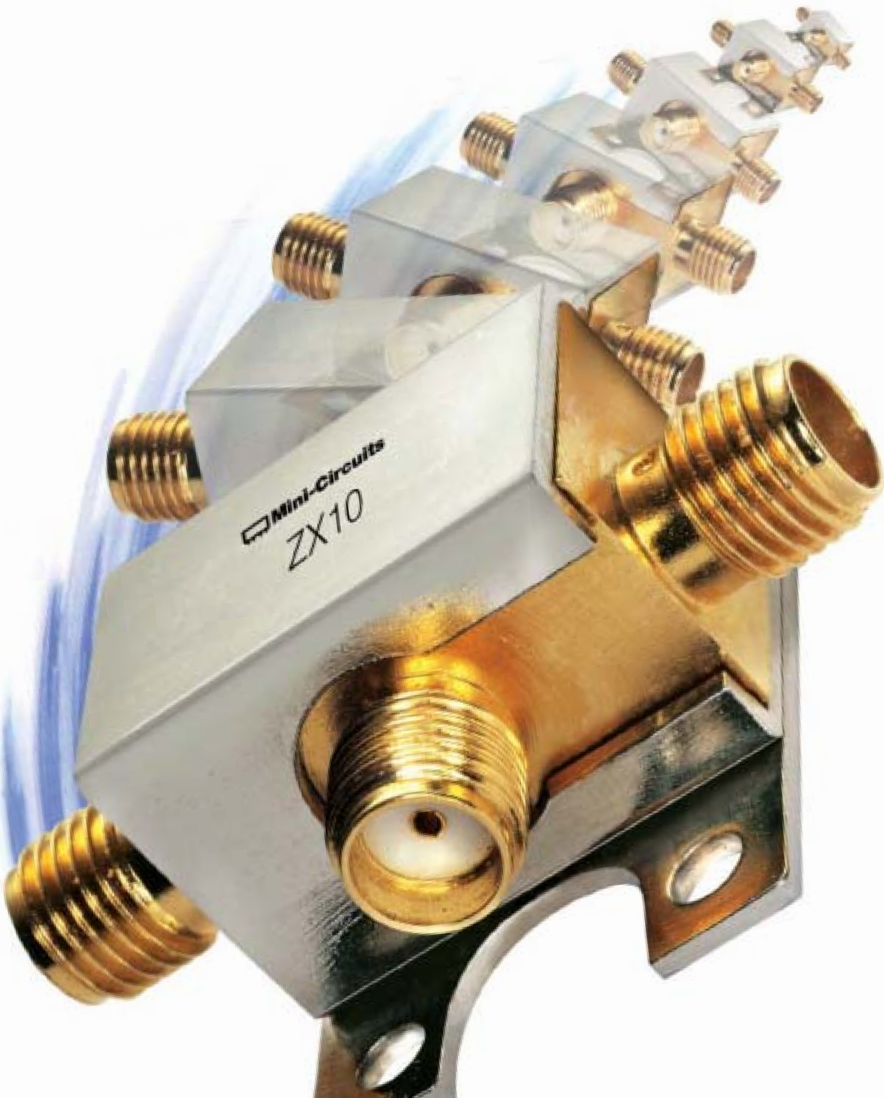
Commenting on the joint initiative SeaMobile president and CEO, William Marks, said, "The combined expertise of Ericsson and SeaMobile creates an unbeatable team in delivering advanced wireless services at sea." He continued, "Ericsson's position as the largest infrastructure provider for 3G and GSM technology is an excellent fit with SeaMobile. We're immediately providing a better wireless user experience at sea that will continue to evolve as we introduce services not currently available to the maritime industry."

Infineon Builds Research Centre in Bucharest

Having established its new Romanian subsidiary in April this year, Infineon has moved quickly to construct the Bucharest Research Centre. This centre specialises in the development of power semiconductors with analogical and digital functions (power-mixed-signal) for applications in various industries, particularly automotive. Thus, the company is endeavouring to meet the increasing demand for performing semiconductors in these industries. The research centre, a subsidiary of Infineon Technologies Austria AG, strengthens the Alliance of Research Centres in Automotive and Industrial Electronics, which currently has offices in Villach and Graz (Austria), Munich (Germany) and Padua (Italy).

Since its establishment in the spring of 2005, Infineon has managed to attract around 30 engineers. Most of them follow training courses in the research centres from Villach and Munich in order to acquire specific knowledge related to the development of semiconductors. Through 'intercultural training' specifically created by the company, the Romanian engineers will be trained over the next six to twelve months with regard to both their technical tasks and the company's culture.

On an area of about 2,200 m^2, Infineon is building offices, research laboratories and measurement technique laboratories. The company is also focusing on establishing close connections with the Polytechnic University of Bucharest, with the first steps towards the development of common master programmes having already been taken. The goal of such programmes is to complete the university curricula with subject matter that will see a high demand for research and development activities in the area of semiconductors in the near to medium future. ■



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Model	Frequency (GHz)	Isolation (dB)	Insertion Loss (dB)	Price $ea. (Qty. 1-24)
2WAY-0°			Above 3.0dB	
ZX10-2-12	.002-1.2	21	0.5	24.95
ZX10-2-20	.2-2	20	0.8	24.95
ZX10-2-25	1-2.5	20	1.2	26.95
ZX10-2-42	1.9-4.2	23	0.2	34.95
ZX10-2-71	2.95-7.1	23	0.25	34.95
ZX10-2-98	4.75-9.8	23	0.3	39.95
ZX10-2-126	7.4-12.6	23	0.3	39.95
4WAY-0°			Above 6.0dB	
ZX10-4-11	.8-1.125	20	0.6	38.95
ZX10-4-14	1.1-1.45	20	0.8	38.95
ZX10-4-19	1.425-1.9	20	0.75	38.95
ZX10-4-24	1.675-2.35	20	0.9	38.95
ZX10-4-27	2.225-2.7	20	1.0	38.95

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Small 4x4mm Package.

DAT SELECTION GUIDE

Specifications Typical at: T_{AMB} = 25°C

Model Series	Interface	Z (Ω)	Freq. MHz	Atten. dB	Steps dB	Bits	Price $ea. (Qty.10)
DAT-15R5-P ▲	Parallel	50	DC-4000	15.5	0.5	5	3.55
DAT-15R5-S ▲	Serial	50	DC-4000	15.5	0.5	5	3.55
DAT-15575-P ▲	Parallel	75	DC-2000	15.5	0.5	5	3.55
DAT-15575-S ▲	Serial	75	DC-2000	15.5	0.5	5	3.55
DAT-31-P ▲	Parallel	50	DC-2400	31.0	1.0	5	3.55
DAT-31-S ▲	Serial	50	DC-2400	31.0	1.0	5	3.55
DAT-3175-P ▲	Parallel	75	DC-2000	31.0	1.0	5	3.55
DAT-3175-S ▲	Serial	75	DC-2000	31.0	1.0	5	3.55
DAT-31R5-P ▲	Parallel	50	DC-2400	31.5	0.5	6	3.80
DAT-31R5-S ▲	Serial	50	DC-2400	31.5	0.5	6	3.80
DAT-31575-P ▲	Parallel	75	DC-2000	31.5	0.5	6	3.80
DAT-31575-S ▲	Serial	75	DC-2000	31.5	0.5	6	3.80

▲To specify Supply Voltage:
Add the letter (P) to model number for positive +3 volts.
Add the letter (N) to model number for Dual ±3 volts.
Example: DAT-15R5-PP or DAT-15R5-PN
Patent Pending

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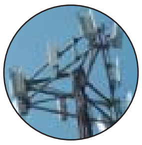

Wireless LAN Report Chronicles Major Technology Changes

Forward Concepts has announced the publication of an in-depth study of the WLAN and WiMAX equipment and chip markets. The new report, "Beyond Wi-Fi: 802.11n, VoWi and WiMAX," provides a comprehensive analysis of worldwide equipment and chip markets. The report also profiles the ODMs, OEMs and chip vendors serving these markets. Based on product reviews and interviews with numerous companies involved in the WLAN and WiMAX markets, the report provides detailed forecasts through 2009 of all market segments, including Access Points (AP) and base stations (WiMAX), Network Interface Cards (NIC), Wireless Router Gateways (WRG) and the chipsets enabling the entire WLAN and WiMAX networks. Units, average selling prices (ASP) and total revenues are forecast for every market segment. In spite of a predicted 23 percent average selling price drop, worldwide shipments of WLAN equipment products will increase six percent to the $5.2 B level in 2005. The report also predicts that WLAN equipment will continue growing at a higher rate in 2006 to the $5.9 B level as new IEEE 802.11n and VoWi-Fi equipment is introduced and the infrastructure for traditional Wi-Fi expands. The WiMAX and pre-WiMAX equipment market (including both 802.16d and 802.16e) is forecast to grow from $72 M in 2005 to just over $2 B in 2009, for an annual compound growth rate of 130 percent. The earlier views were that mobile WiMAX would be a threat to 3G, but now cellular equipment vendors such as Nokia are saying that WiMAX will be complementary. Preliminary analyses indicate that 802.16e data delivery costs can be significantly cheaper per megabyte than HSDPA or 1xEV-DO when provided as an overlay to a cellular network. Future WiMAX chipsets, beginning with 802.16d-compliant fixed-operation units, are beginning to ship in 2005 for estimated revenues of $5.4 M. However, mobility-capable 802.16e chipsets will begin sampling next year and the combined chip market is forecast to grow at a 209 percent compound annual growth rate to $489 M in 2009. Details of the report are available at www.fwdconcepts.com/wirelan5.htm.

Asia May Become World's Largest Mobile Market Within 10 Years

Asia may become the largest regional mobile telecommunications market in the world over the next five to ten years, reports In-Stat. In 2004, there were nearly 740 million mobile users in Asia (including Japan, Australia and New Zealand) with total revenue of roughly $180 B, the high tech market research firm says. By 2009, Asian mobile telecom revenues will reach over $260 B, according to In-Stat's forecast. India will be the fastest growing entity in this region, with a 32.8 percent compound annual growth rate (CAGR) in terms of subscribers and 31.1 percent by revenue, for the period from 2004 to 2009.

A recent report by In-Stat found the following:

- South Korean handset vendors like Samsung and LG are significantly gaining market share in both high end products and in developing markets.
- In the equipment area, Sony Ericsson is still winning within the region, with its advanced technologies and R&D facilities, extensive infrastructure, great product quality and successful marketing strategies.
- US equipment vendors like Motorola and Lucent have made significant headway in the Asian market because of their large new deployments of CDMA technology.

The report, "Asia Wireless Annual Report: Subscribers, Handsets and Infrastructure," covers wireless services, handset and equipment markets in 13 countries that make up the Asian market. It includes forecasts for subscribers regionally and by country, as well as revenue forecast by service, ARPU forecast and wireless equipment spending through 2009. Also included are vendor market shares, by country for 2004.

Fixed-mobile Technologies Level the Telecom Services Marketplace

The dream of using one telephone with one number whether at home, at work or on the street — and of networks smart enough to hand over a call in progress — is approaching reality. "Fixed-mobile convergence" is the buzz phrase for this telephonic utopia and it is being driven by a complex blend of threats and opportunities that are analyzed in a new ABI Research study, "Fixed-mobile Convergence: Comparative Business Plans, Implementation Scenarios and Capital Expenditure." The report, which includes a forecast of the market potential for FMC to 2010, outlines the technologies involved, explains the benefits for subscribers and analyzes business scenarios that make FMC attractive to operators. Pressured by VoIP operators such as Skype/eBay and Vonage, and anxious to reduce costs by bringing fixed and mobile businesses together, mobile operators, mobile virtual operators and integrated network operators (including France Telecom and British Telecom) are increasingly drawn to FMC. "The case for FMC rests on the availability of low cost, dual-use (cellular and WLAN) handsets," says the study's author, analyst Ian Cox. "The first models are nearing commercial launch and their prices should be competitive with conventional mobile handsets early in 2006. That will be the trigger for offering this service." FMC needs to be standards-based to ensure cross-network compatibility and this regulatory effort is well under way. The result: UMA, usable with existing mobile networks and IMS-capable for IP networks, will lead early in the

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consumer market, while SIP addresses the needs of the enterprise. "We expect FMC to take off sooner in Europe and Asia than in North America because of the greater prevalence of GSM and 3G services in those regions," notes Cox. "However, any operator using a suitable network can gain a competitive edge by early adoption of FMC, and ABI Research anticipates that up to a fifth of all broadband subscribers will take advantage of FMC's convenience and lower costs by 2010."

The Good, the Bad and the Ugly of 802.11n

The news and rumors surrounding attempts to establish an industry standard for the 802.11n Wi-Fi format paint a picture that changes with the viewing angle. The eagerly-awaited enhancement of the 802.11 wireless LAN standard, that promises wireless users throughput of 100 Mbps and more, has been bogged down in a two-camp battle over the shape of the specifications to be submitted for IEEE approval. That much is not news. But according to a new market review from ABI Research, the prospects for an agreed draft, which optimists had touted for as early as this month after encouraging IEEE announcements in July, are not looking good. "It was hoped that by now the two industry groups, WWiSE and TGn Sync, would have thrashed out a single proposal," says Philip Solis, senior analyst, who covers Wi-Fi semiconductors for ABI Research's Wi-Fi Research Service. "But we hear that four major companies — Broadcom, Intel, Atheros and Marvel, holding the lion's share of the Wi-Fi chipset market — have formed a third camp with the aim of writing a whole new proposal." In light of this development, ratification of a standard could be delayed until mid-2007, at the earliest. Cynical observers have called this an offensive gambit aimed at Airgo Networks, the small but energetic chipmaker that has rapidly been gaining ground in the consumer market. Senior analyst Sam Lucero, author of the Wi-Fi Research Service's equipment coverage, adds, "That interpretation may have some merit. If these companies, which have been slower bringing spatial multiplexing to market, can change the standard proposal drastically, Airgo would be forced to a fundamental redesign." ABI Research's Wi-Fi Research Service provides an overview of emerging Wi-Fi opportunities and the technology's future co-existence with cellular technologies and consumer electronics. It analyzes Wi-Fi semiconductor markets, Wi-Fi access points, adapters and switches, and explores the spread of Wi-Fi public hotspots. ■

GaAs T/R Diversity Switch DC–6 GHz

The SKY13267-321LF is a monolithic DPDT switch fabricated using Skyworks Solutions proprietary GaAs PHEMTs as the switching elements. This wideband switch offers very low insertion loss 0.8 dB type @ 5.2 GHz. The P_1dB is +30 dBm type @ 3V and RF signal paths within the SKY13267-321 are fully bilateral.

Skyworks Solutions 781-376-3000
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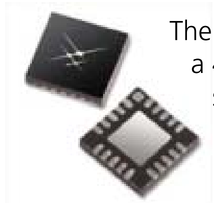
CMOS Switch DBS 4 x 2 Switch Matrix 250 MHz–2.15 GHz



The SKY13272-340LF is a 4-input to 2-output switch in a low cost QFN-20 4 x 4 mm package. The SKY13272-340LF enables 16 states, directing any of the four inputs to either of the two outputs. States are selected by four positive voltage control inputs. The switch can operate over the temperature range of -40 °C to 85 °C.

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250–2700 MHz Linear Power Amplifier Driver

The SKY65004 is a high performance, ultra-wideband amplifier with superior output power, linearity, and efficiency. The device uses low-cost, Surface-Mount Technology in the form of a 4-pin SOT package (-11 option) or lead (Pb)-free, 3-pin MCM package (-21 option, pin-compatible with the -11). The high linearity and superior ACPR/ACLR performance of the SKY65004 make it ideal for use in the driver stage of infrastructure transmit chains.

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The SKY65013, SKY65014, SKY65015, SKY65016 and SKY65017 are broadband devices with gain from13 to 21 dB, and output power (P_1dB) from 14 to 19 dBm. Skyworks gain blocks have the highest output third order intercept (IP3) and gain compression points (P_1dB) when compared to otherwise similar products that draw the same power supply current and operate over the same wide bandwidth.

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New Fall 2005 Product Selection Guide Brochure

Specifications for a broad portfolio of building blocks for mobile, infrastructure, medical, automotive, and many more applications requiring RF expertise.

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AROUND THE CIRCUIT

INDUSTRY NEWS

■ **Smiths Group** announced that it is expanding its interconnect business with the acquisition of **Millitech Inc.** A Massachusetts-based business, Millitech specializes in the design and manufacturing of millimeter-wave components, assemblies and integrated antenna systems for satellite communications, radar, passive imaging, space and remote sensing applications. Smiths acquired the business for $33.5 M, in cash.

■ **American Technical Ceramics Corp.** (ATC), a manufacturer of high performance electronic components, announced that it has entered into an agreement to purchase from **CTS Corp.** certain equipment and inventory used by CTS in the manufacture of low temperature co-fired ceramic products. It is anticipated that the closing of the transaction will occur in the second quarter of fiscal year 2006. ATC intends to transfer the equipment and inventory to its facility in Jacksonville, FL, where it will be used, among other things, in the production of LTCC products and certain of ATC's specialty multilayer capacitors. Terms of the transaction were not disclosed.

■ **RF Industries Ltd.** announced that it acquired, for cash and equity consideration, the assets of **Worswick Industries Inc.**, a privately held San Diego, CA-based manufacturer and supplier of custom and standard computer cable and wire harness assemblies. Terms of the acquisition were not disclosed.

■ **Andrew Corp.** has expanded its Geometrix® mobile location system product line with the acquisition of certain assets of **Nortel**'s wireless location business, augmenting a worldwide offering that supports innovative location-based services for commercial, consumer and public safety uses.

■ **Sensors Unlimited Inc.** accepted the $60 M cash offer to merge with **Goodrich Corp.** The definitive agreement has been approved by the boards of directors of Sensors Unlimited Inc. and Goodrich Corp. and is expected to close in the fourth quarter of 2005. The transaction is subject to approval by US regulatory agencies.

■ **Modelithics** has recently acquired a 0.3 to 6 GHz **Maury Microwave** noise parameter test system to expand the frequency range covered by the solid-state tuner-based noise parameter system that it currently uses for 2 to 26 GHz measurements. Modelithics also uses Maury's ATS load/source pull equipment for development and validation of its growing nonlinear transistor model library, as well as its custom measurement and modeling services.

■ **Anritsu Corp.** president Hiromichi Toda announced the opening of a sales and marketing office in Bangalore, India. The opening of the office allows the company to expand its business in India. With the opening, Anritsu now has offices in 17 countries, including the US and Japan.

■ **Aeroflex Inc.** announced the opening of a new sales office for its test and measurement group in Taipei City, Taiwan in the Xinyi District. This new office will become the nucleus for the sales and support of Aeroflex test products and systems in Taiwan and will help facilitate the strong growth being seen currently within the wireless manufacturing industry in this country. Allen Chen has been appointed as the new country manager for Taiwan. Chen comes to the company with seven years experience in the telecom industry and eight plus years in the IT industry.

■ **TestMart**, a marketplace operator and service provider for the test and measurement industry, announced an agreement with **Geotest-Marvin Test Systems Inc.**, a global supplier of PXI and PC-based test products, systems and solutions. The deal provides the US government and federal contractor marketplace with special pricing on select test instrumentation, instrumentation controllers and chassis, test development software and turn-key test systems.

■ **Digital Fountain Inc.**, a supplier of forward error correction technology for reliable communications, announced a major licensing deal with **Nokia**. The agreement will allow Nokia to incorporate Digital Fountain's advanced FEC technology in future Nokia products.

■ **Corning Gilbert**, a subsidiary of Corning Inc., announced that it has opened a new microwave customer service center in Vordingborg, Denmark. The customer service center will be operated out of Corning Gilbert's subsidiary, Corning Cabelcon.

■ **Elcoteq Network Corp.** announced the opening of Elcoteq Engineering Service Center in Richardson, TX. The mission of the new center is to provide customers an on-going source of state-of-the-art assembly and new product introduction manufacturing support.

■ **WJ Communications Inc.** announced that the company is expanding its global presence in strategic locations such as China to provide high quality local support to meet increased customer demand around the globe. In order to better serve WJ's growing base of customers throughout Asia, WJ has enhanced its sales and technical support capabilities at its Shenzhen office by the addition of two experienced and highly qualified individuals.

■ **Electro Rent Corp.**, with operations in North America and China, is establishing a wholly-owned operation to better serve customers in the European market. **ER Europe** has commenced operations and has access to the resources of Electro Rent Corp. As part of its plans, ER has acquired the operations of Everest ES, a Belgium equipment rental, sales and leasing business, which has been operating in an alliance with Electro Rent for several years. As part of the acquisition, David Saeys, the owner of Everest, becomes the general manager of ER Europe.

AROUND THE CIRCUIT

■ **Agilent Technologies Inc.** announced the availability of its 2005 releases of Advanced Design System and RF Design Environment. Both EDA software platforms contain Agilent's simulation technologies, announced in March, that help designers of wireless communications products get products to market faster. The new simulation technologies are fully integrated into the software and are shipping now to customers worldwide.

■ **Comarco Wireless Test Solutions** announced that high speed downlink packet access test capability is now available with its 3G scanning receiver, which comes in its three major product offerings, Seven.Five, Prizm and OEM Scanners. This capability may be purchased as an upgrade to already purchased products or as a new system.

■ **Applied Wave Research Inc.** announced a new integrated filter synthesis solution using **Nuhertz Technologies**' filter synthesis technology and is available immediately in both Microwave Office® and Analog Office™ design suites. High frequency circuit designers can now perform accurate filter synthesis quickly and easily from within the unified AWR design platform.

■ **TDK Semiconductor Corp.** announced that it has changed its name to **Teridian Semiconductor Corp.** The name change is effective immediately. As previously announced, Golden Gate Capital, in partnership with the US-based semiconductor management team, acquired the company from TDK-USA, a subsidiary of the Japan-based TDK Corp.

■ **Dynaco Corp.**, a fabricator of high reliability rigid-flex, flex and rigid printed circuit boards and assemblies for the military/aerospace market, announced that it has successfully achieved qualification to the Department of Defense Performance Specification MIL-PRF-31032, and the associated specifications MIL-PRF-31032/1 and MIL-PRF-31032/2 for rigid constructions, plus MIL-PRF-31032/3 and MIL-PRF-31032/4 for rigid-flex constructions.

■ **ANADIGICS Inc.** announced that the company has commenced production volume shipments of 3 by 3 mm cellular band CDMA power amplifiers to **Samsung Electronics** for several wireless handsets.

■ **RF Micro Devices Inc.** announced that it has shipped its 10 millionth POLARIS™ cellular transceiver. The company attributes the shipment milestone to continued strong sales of its EDGE transceivers.

■ **L-3 Communications Cincinnati Electronics** (CE) Space Electronics has been nominated to receive NASA's George M. Low award for quality and excellence. L-3 CE Space Electronics was nominated by Jet Propulsion Laboratory for the design development and production of three flight UHF transceivers that have served the Odyssey Mars Orbiter and the Mars Rovers Opportunity and Spirit.

CONTRACTS

■ **Symmetricom Inc.** announced that it has been selected as a supplier for the Advanced Technology Atomic Frequency Standard program (ATAFS), sponsored by the Department of Defense Joint Program Office. The ATAFS program will support the development of the next-generation atomic clocks for possible deployment on the Global Positioning System (GPS-III) satellite constellation. Under the contract, awarded by the **Space and Missile System Center** at Los Angeles Air Force Base and valued at $3.9 M, Symmetricom will develop high performance atomic clocks based on its proprietary optically-pumped cesium beam technology.

■ **EMS Technologies Inc.** announced that its defense & space systems division is performing on a contract from **Northrop Grumman Corp.**'s Norden Systems business unit to provide maintenance support and upgrades for the US Air Force's E-8C Joint Surveillance Target Attack Radar System (Joint STARS). The total value to EMS is estimated at US $2.1 M over 12 months. As part of the effort, EMS is upgrading phase shifter modules and control electronics to reduce power consumption and enhance accuracy. In addition, the contract includes support services.

■ **M/A-COM**, a business unit of Tyco Electronics and a provider of wireless radio frequency, microwave and mil-

limeter-wave components, announced that it has been chosen by the **Office of Naval Research** to lead a one-year, $1.5 M program, which will optimize the manufacturability of its patented high voltage multi-function self-aligned gate MMIC process.

■ **G.T. Microwave Inc.**, Randolph, NJ, was awarded a $1 M firm-fixed-price contract for 500 calibrated true position vector modulators. G.T. Microwave, a designer and manufacturer of state-of-the-art microwave integrated components, proposed the accepted plan to Lockheed Martin, an advance technology systems integrator. The vector modulators will be utilized to solve an antenna test application with 0.1-degree accuracy. G.T. has delivered the first production units on time in 14 weeks and continues scheduled deliveries.

■ **Astron Wireless Technologies** was recently awarded a Small Business Innovation Research (SBIR) Phase II award from the **Department of Defense** for development of an "Adaptive Bandwidth High Power RF Antenna." The objective is to develop a universal approach to the design of high power broadband antennas for the non-lethal utilization of RF energy. This development and production of a prototype system is based on the Phase I development of a 10:1 bandwidth basic Log-Conical antenna. During Phase II, the program will focus on creating a truly miniature high gain, high power antenna operating within the general operating band of up to 1000 MHz.

■ **MI Technologies** announced it has received an order from the UK's **Defense Science and Technology Laboratory** (DSTL) to provide upgrades to an existing outdoor far-field antenna test and measurement system. The equipment will be used in the design, development and validation of defense related RF projects. Under terms of the agreement, MI Technologies will install the outdoor test and measurement system at DSTL's facilities in Pershore, UK. The facility is scheduled to be operational in the winter of 2005.

■ **GigaBeam Corp.** announced that it has entered into an agreement with **Epsilon Lambda Electronics Corp.** pursuant to which GigaBeam will provide technology design and support for a millimeter-wave communications system for the US Navy.

FINANCIAL NEWS

■ **Merrimac Industries Inc.** reports sales of $7.6 M for the second quarter ended July 2, 2005, compared to $7.9 M for the same period in 2004. Net income for the quarter was $332,000 ($0.10/per diluted share), compared to a net income of $444,000 ($0.14/per diluted share) for the second quarter of last year.

■ **Applied Radar Inc.** announced that it has recently been awarded an additional $1.9 M in SBIR funding from the **Department of Defense**. The funding includes $1.4 M in new funding from the Air Force Research Laborato-

ry and the Defense Advanced Research Projects Agency to develop advanced digitally-controlled antenna systems, wideband digital receivers and exciters, and conformal integrated antennas for unmanned aerial vehicles.

■ **NEC Corp.** announced that it has reached a settlement with **Harris Corp.** in the patent infringement lawsuits filed against it on September 3, 2004, in the US District Court for the Northern District of California and the Federal Court of Canada, which claimed infringement of NEC's Digital Microwave Radio patents by NEC Corp. As a result of the settlement, Harris has agreed to take a royalty-bearing, non-exclusive license under NEC's Digital Microwave Radio patents. In addition, NEC agreed to dismiss all claims covered by the lawsuits filed on September 3, 2004, in the previously mentioned courts.

NEW MARKET ENTRY

■ **RFIC Solutions Inc.** is a new start-up fabless semiconductor company based in San Jose, CA and India. The company will specialize in RFIC and RF module design services using IC processes that include GaAs, PHEMT, SiGe, silicon, CMOS and BiCMOS. RFIC Solutions will offer reusable RF IP blocks such as LNA, PA, mixers, switches, transceiver front-end chips for WLAN, WiMAX, PCS and other RF systems. The company can be reached at US (408) 674-5045, India 91-712-2282129 or visit www.rficsolutions.com.

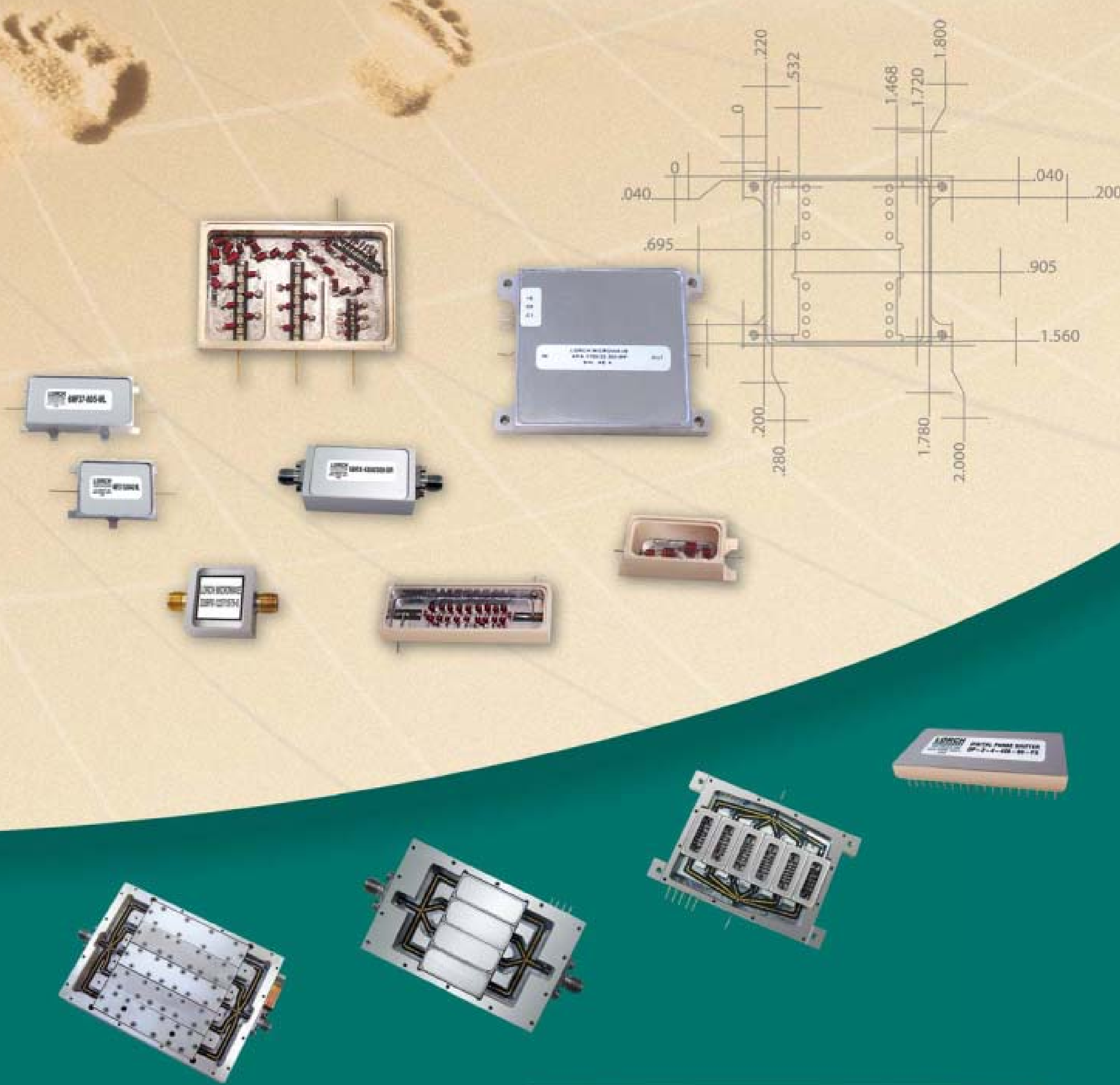
PERSONNEL

▲ Robin Southwell

■ Defence & Security Systems (DS), the defense pillar within EADS, has appointed **Robin Southwell** as chairman of DS UK, an appointment that is in addition to his current role as CEO of EADS UK. Previously, Southwell was CEO of AirTanker Ltd., a consortia of EADS, Cobham, Rolls Royce, Thales and VT Group. He has also held the position of group managing director of customer solutions and support at BAE Systems.

■ STMicroelectronics has made two major appointments. **Reza Kazerounian**, currently group vice president and general manager of the company's Smart Card division, has been promoted to corporate vice president, North America region, while **Marco Cassis**, currently vice president automotive and a board member of the Japanese subsidiary, STMicroelectronics K.K., has been promoted to corporate vice president of STMicroelectronics Japan. Kazerounian has twenty years of experience in the electronics industry and industry-wide acclaim for his contribution to the development of non-volatile memory and smartcard technologies. Cassis is a well-established member of ST Japan's senior executive team, with twelve years of experience in Japan and knowledge of the local business environment.

Whatever the Size of the Footprint, We Make it Fit



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▲ Ulf Meiners

■ **Ulf Meiners** has been appointed chief technical officer for UMS, which follows on from his position as joint managing director of UMS GmbH, where he took responsibility for all front-end wafer manufacturing. In his new role, Meiners will be responsible for all of the internal and external technical activities at both of the company's sites at Ulm in Germany and Orsay in France. He will also retain his joint management responsibility for UMS GmbH, where all manufacturing activities will be under the responsibility of **Jacques Bonnet**.

■ RF Monolithics Inc. (RFM) announced the appointment of **Wayne Stargardt** to the position of director of marketing, Wireless Systems, and **Duane Covell** to the position of director of OEM sales, Wireless Systems. In these positions, Stargardt and Covell will expand RFM's sales and marketing presence in emerging wireless systems markets. Stargardt has over 25 years of effective business experience in engineering, marketing and management. He was most recently vice president, sales and marketing at SensorLogic Inc. Covell has over 20 years of effective business experience in sales and sales management. He was most recently vice president North American sales for Plexus Corp.

REP APPOINTMENTS

■ **Locus Microwave**, a manufacturer of custom and standard RF amplifier products, announced the appointment of Tracy Alves of **CMI Technical Sales, South** for representation in North and South Carolina.

■ **IMS Connector Systems**, headquartered in Germany, has increased its worldwide sales and distribution network by enlisting **Extreme Components** as its new distribution partner in North America. The company's complete product range will be carried, including RF connectors and cable assemblies, components and antennas for mobile devices, as well as SMBA® (FAKRA) connectors for in-car communication. Targets for these products include telecommunications infrastructure, mobile phones and wireless data communication devices, measurement equipment, the automotive industry and various technology sectors.

■ **Reactel Inc.**, a manufacturer of RF and microwave filters, multiplexers, switched filter banks and subassemblies to the commercial, military, industrial and medical industries, announced the appointment of **KSA Electronics** as the company's representative in southern California. For more information about KSA, please visit www.ksa.com or telephone (619) 858-0770.

TRANSISTOR LC OSCILLATORS FOR WIRELESS APPLICATIONS: THEORY AND DESIGN ASPECTS, PART II

Part I of this tutorial article[1] gave a detailed discussion and analysis of the start-up and steady-state oscillation conditions for transistor LC oscillators with emphasis on CMOS devices. Part II presents both linear and nonlinear phase noise models for the parallel feedback and negative resistance oscillators. Each approach to the oscillator phase noise calculation has its own advantages and drawbacks. For example, the linear Leeson model for a parallel feedback oscillator expressed in an explicit analytical form is very simple and can explain the dependence of the oscillator phase noise on the resonant circuit loaded quality factor, signal power, active device noise figure and its low frequency flicker noise. However, it is not considering the effect of amplitude-to-phase conversion and higher order harmonics. In this case, the nonlinear Kurokawa model, developed for a negative resistance oscillator, demonstrates the explicit analytical relationship between the phase noise and oscillator stability margin, and shows the dependence of the oscillation frequency on the oscillation amplitude in a large-signal operation mode, in the form of derivatives of the device and circuit impedances. Also, the individual and joint effects of different nonlinear circuit elements, which will result in both amplitude and phase fluctuations, are discussed.

Part II presents both linear and nonlinear phase noise models for the parallel feedback and negative resistance oscillators.

PHASE NOISE MODELS FOR PARALLEL FEEDBACK AND NEGATIVE RESISTANCE OSCILLATORS

Leeson Model for a Parallel Feedback Oscillator

The simple Leeson linear noise model for a feedback oscillator, which was derived empirically, is based on the expectations that the real oscillator has two basic components.[2] The first component is caused by the phase fluctuations due to the additive white noise at frequency offsets close to the carrier, as well as due to the noise having a mixing nature resulting from the circuit nonlinearities. The second component is a result of the low frequency fluctuations or flicker noise upconverted to the carrier region because of the active device nonlinear effects.

ANDREI GREBENNIKOV
M/A-COM Eurotec Operations
Cork, Ireland

The phase noise at the input of the power amplifier is added to a signal as the sum of every bandwidth $\Delta f = 1$ Hz, each producing an available noise power at the input of the noise-free amplifier. Maximum power delivery can be achieved when the source internal impedance is conjugate-matched to the input impedance of the amplifier. As a result, only one-half of the root-mean-square noise voltage appears across the amplifier input and is equal to

$$e_{in} = \frac{e_n}{2} = \frac{\sqrt{4FkTR}}{2} = \sqrt{FkTR} \qquad (1)$$

where R is the equivalent resistance, which can be represented as the input resistance for the input root-mean-square noise voltage.[3] The input phase noise produces a root-mean-square phase deviation $\Delta\phi_{rms} = \Delta\phi/\sqrt{2}$ at each offset frequency $\pm f_m$ from the carrier f_0, as shown in ***Figure 1***, for which a total power-wise sum can be written for a small phase perturbation as

$$\Delta\phi = \Delta\phi_{rms}\sqrt{2} = \frac{e_{in}\sqrt{2}}{V_{in}} = \sqrt{\frac{FkT}{P_{in}}} \qquad (2)$$

where $V_{in} = \sqrt{2P_{in}R}$ is the signal voltage amplitude at the power amplifier input.

As a result, the double sideband spectral power density of the thermal phase noise in a frequency bandwidth $\Delta f = 1$ Hz can be written as

$$S_{\Delta\phi} = \left(\Delta\phi\right)^2 = \frac{FkT}{P_{in}} \qquad (3)$$

The Leeson model consists of an amplifier with a noise figure F and a resonator (or filter) in the feedback loop, as shown in ***Figure 2***.[3] The oscillator phase noise is modeled by assuming a noise-free power amplifier and adding a phase modulator at its input. Based on empirical predictions, the phase noise level of the oscillator at an offset frequency f_m from the carrier f_0 can be described by

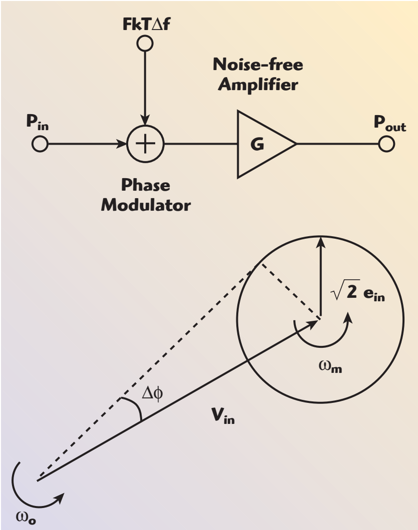


Fig. 1 Simplified feedback oscillator noise model.

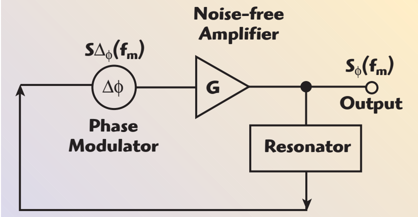


Fig. 2 Equivalent model of a parallel feedback oscillator.

$$S_\phi(f_m) = S_{\Delta\phi}(f_m)\left(\frac{f_0}{2Q_L f_m}\right)^2 \quad \text{for } f_m < \frac{f_0}{2Q_L} \tag{4}$$

$$S_\phi(f_m) = S_{\Delta\phi}(f_m) \quad \text{for } f_m > \frac{f_0}{2Q_L} \tag{5}$$

where $S_{\Delta\phi}(f_m)$ is determined, using Equation 3, as

$$S_{\Delta\phi}(f_m) = \frac{FkT}{P_{in}}\left(1 + \frac{f_c}{f_m}\right) \tag{6}$$

taking into account the effect of the signal purity degradation due to the low frequency flicker noise effect close to the carrier, described by the corner frequency f_c.

It should be noted that the empirical Leeson equation for $S_{\Delta\phi}(f_m)$ contains a multiplication factor of two in the numerator. Moreover, accurate agreement was achieved between the model and experimental results when the power density $S_{\Delta\phi}(f_m)$ was expressed in terms of the compressed (or large-signal) power gain G and output power P_{out} as $S_{\Delta\phi}(f_m) = 2GFkT/P_{out}$.[4] The parameter F in Equation 6 is associated with the active device noise figure and can be called an effective noise factor because, generally, it should represent the effects of the active device noise sources and the cyclostationary noise resulting from periodically varying processes in practical oscillators. Due to the inherent nonlinear nature of the active device, the effects of intermodulation between the wideband white noise and various harmonics of the fundamental frequency (for example, nonlinear transformation of the noise near the third harmonic downconverted to the near carrier region due to mixing effect with the second harmonic) must be included.[2] Also, the effect of low frequency noise modulation of the current, resulting in the reactance modulation of the input impedance of the circuit (for example, variation of the phase angle of the device forward transfer function versus emitter current), cannot be neglected.[5] Hence, it is impossible to calculate F accurately without taking into account the effect of the oscillator resonant circuit. Therefore, for such a linear model, the effective noise factor F as well as the corner frequency f_c can be considered more like fitting parameters, based on measured data.

The corresponding combined expression to calculate the normalized double-sideband phase noise power spectral density or the double-sideband noise-to-carrier ratio at the input of the feedback oscillator can be obtained from

$$S_\phi(f_m) = \frac{FkT}{P_{in}}\left(1 + \frac{f_c}{f_m}\right)\left[1 + \left(\frac{f_0}{2Q_L f_m}\right)^2\right] \tag{7}$$

which gives an asymptotic model showing generally a noise reduction of 9 dB/octave in the offset region with predominant low frequency 1/f noise, 6 dB/octave in the offset region due to feedback loop and 0 dB/octave representing the thermal or white noise spectrum.

The single sideband noise-to-carrier ratio at the input of the feedback oscillator can be described by

$$L(f_m) = \frac{1}{2}\frac{FkT}{P_{in}}\left(1+\frac{f_c}{f_m}\right)\left[1+\left(\frac{f_0}{2Q_L f_m}\right)^2\right] \quad (8)$$

whose idealized sideband spectral behavior for different values of the loaded quality factors is illustrated in ***Figure 3***. For the low Q_L case, there are regions with $1/f_m^3$ and $1/f_m^2$ dependencies for spectral power density close to carrier, as shown in (a). For the moderate QL case, (b), only a $1/f_m^3$ dependence exists as far as the intersection with the thermal noise floor. For the high Q_L case, (c), regions with $1/f_m^3$ and $1/f_m$ dependencies are observed near the carrier. Closest to the carrier, the $1/f_m^3$ phase noise behavior is a result of random frequency modulation of the oscillator by low frequency 1/f noise. In the region of $1/f_m^2$ phase noise behavior, the white noise causes random frequency modulation. The $1/f_m$ dependence is due to the mixing up of the 1/f noise with the oscillation frequency. Finally, the phase noise becomes constant, which is a result of the mixing up of white noise around the oscillation frequency.

To calculate the same phase noise spectral density at the oscillator output, it is necessary to replace the input power P_{in} by the power available at the output P_{out} and to multiply the numerator of Equation 8 by the power gain G. As a result, neglecting the effect of flicker noise and considering the case of $f_m << f_0$, one can obtain

$$L(f_m) = \frac{GFkT}{8Q_L^2 P_{out}}\left(\frac{f_0}{f_m}\right)^2 \quad (9)$$

where

$$G = \frac{1}{\left(1-\frac{Q_L}{Q_0}\right)^2} \quad (10)$$

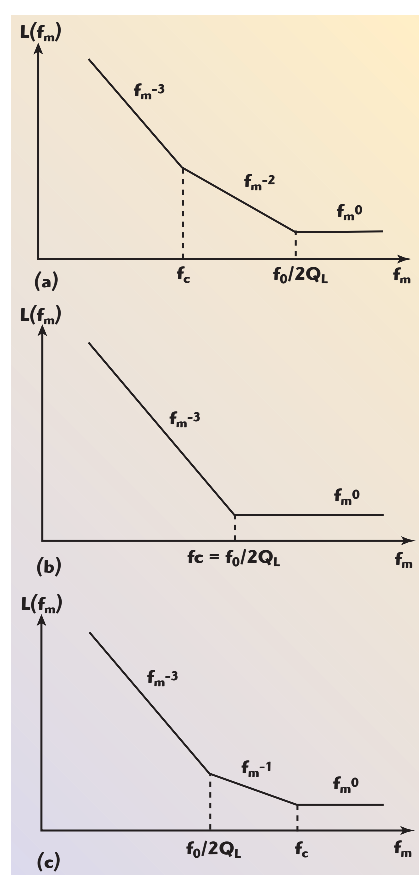


Fig. 3 Single sideband oscillator phase noise behavior.

is considered the transducer power gain and Q_0 is the unloaded quality factor.[6] From Equations 8 and 9, it follows that, to minimize the oscillator phase noise, it is necessary to reduce the noise figure F and to increase the input power P_{in} (or the output power P_{out} for a fixed power gain G of the amplifier) as much as possible. In addition, for frequency offsets inside the resonator bandwidth, it is desirable to maximize the oscillator loaded quality factor Q_L. However, the resonator insertion loss and loaded Q_L are interrelated, and one cannot arbitrarily increase Q_L without increasing the insertion loss, otherwise a larger power gain G is needed. The two competing effects result in an optimum loaded Q_L of approximately one half the unloaded Q_0 and insertion loss of approximately 6 dB.[7,8] Thus, the minimum noise occurs when $Q_L/Q_0 = 0.5$ resulting in

$$L(f_m) = \frac{2FkT}{P_{out}}\left(\frac{f_0}{2Q_L f_m}\right)^2 \qquad (11)$$

Note that the difference in the optimum noise performance predicted by different definitions of the output power (power dissipated in the resonant circuit or power available at the amplifier output) is small.[6]

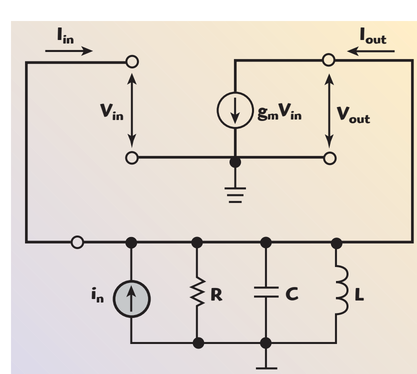
Now let us represent a parallel feedback oscillator model using the circuit schematic, with an active device and a parallel resonant circuit, shown in ***Figure 4***. Here, the active device is an ideal voltage-controlled current source with transconductance g_m. At the operating frequencies of $f << f_T$, where f_T is the transition frequency, such a simplified transistor model can describe the behavior of a FET device with the input gate-source capacitance C_{gs} (it can be included into the resonant circuit capacitance C), the gate-drain capacitance C_{gd} (its value is typically sufficiently small) and the drain-source resistance R_{ds} (its value is assumed to be infinite). The impedance of the parallel resonant circuit for $\Delta\omega << \omega_0$, where $\Delta\omega$ is the offset from the carrier, can be calculated from

$$\Delta\phi = \Delta\phi_{rms}\sqrt{2} = \frac{e_{in}\sqrt{2}}{V_{in}} = \sqrt{\frac{FkT}{P_{in}}} \qquad (12)$$

where $\omega_0 = 1/\sqrt{LC}$ is the resonant frequency and $Q_L = \omega CR$ is the loaded quality factor.

For a given open-loop voltage transfer function $H(j\omega) = g_m Z_L(j\omega)$, the closed-loop voltage transfer function $T(j\omega)$ can be written as

$$T(j\omega) = \frac{V_{out}(j\omega)}{V_{in}(j\omega)} = \frac{g_m Z_L}{1 - g_m Z_L} \qquad (13)$$



▲ *Fig. 4 Equivalent model of a feedback oscillator.*

By substituting Equation 12 into Equation 13 and using the steady-state oscillation condition corresponding to the Barkhausen criterion as $g_m R_L = 1$, the magnitude $T(f_m)$ of the complex transfer function $T(j\omega)$ can be obtained from

$$T(f_m) = \frac{V_{out}(f_m)}{V_{in}(f_m)} = \frac{f_0}{2Q_L f_m} \tag{14}$$

where

$f_m = \Delta\omega$

Since the noise current in is produced by the resistor R only, it can be transformed to the equivalent voltage noise with amplitude $V_{in} = \sqrt{8kTR\Delta f}$ at the device input. In this case, Equation 14 can be rewritten as

$$V_{out}^2(f_m) = 8kTR\Delta f \left(\frac{f_0}{2Q_L f_m}\right)^2 \tag{15}$$

As a result, the single sideband spectral noise power density in a frequency bandwidth $\Delta f = 1$ Hz, normalized to the total power dissipated in the oscillator $P = V_{out}^2/2R$, can be calculated from

$$L(f_m) = \frac{1}{2}\frac{V_{out}^2(f_m)}{V_{in}^2} = \frac{2kT}{P}\left(\frac{f_0}{2Q_L f_m}\right)^2 \tag{16}$$

where the factor ½ arises from neglecting the contribution of the amplitude noise since, for a totally uncorrelated noise, one half of the total noise power contributes to AM sidebands and the other half of the total noise power is converted into PM sidebands.[9] Equation 16 is similar to Equation 11 with only the difference in power definitions. Since P represents the total or DC power dissipated in the oscillator with an ideal lossless active device (for example, when the device operates in the switching class E mode), the power delivered to the load is PL = ηp, where η is the efficiency of the oscillator. The load resistance can represent a part of the resonant circuit resistance R. Thus, Equation 16 can be given in the form

$$L(f_m) = \frac{2\eta kT}{P_L}\left(\frac{f_0}{2Q_L f_m}\right)^2 \tag{17}$$

Linear Model for Negative Resistance Oscillator

Now consider the equivalent circuit of a simple single-resonant negative resistance oscillator shown in ***Figure 5***, where the available noise power is assumed to be totally from the active device. Here, R_n is the noise resistance associated with active device noise sources, the negative resistance R_{out} and the equivalent output capacitance C_{out} represent the device negative output impedance, L is the tank inductance and R_L is the load resistance. The derivation of the power spectral density will be based on the fact that the available noise power in the active device is amplified in a frequency selective way, resulting at resonance in the output power P_L being dissipated in the load resistance R_L.[10] This will happen in a steady-state condition when the values of the negative resistance and the load resistance are close to each other.

Then, assuming that $R_L + \Delta R = -R_{out}$ and defining the magnitude of the mean-square noise current flowing into the load from the mean-square noise voltage, one obtains

$$\overline{i_n^2} = \frac{\overline{e_n^2}}{(\Delta R)^2 + \left(\omega L - \frac{1}{\omega C_{out}}\right)^2} \tag{18}$$

Equation 18 can be rewritten in a common form as

$$\overline{i_n^2} = \frac{1}{\left(\frac{\Delta R}{R_L}\right)^2 + Q_L^2\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right)^2} \frac{\overline{e_n^2}}{R_L^2} \tag{19}$$

where

ω_0 = resonant frequency
Q_L = oscillator loaded quality factor at the resonant frequency

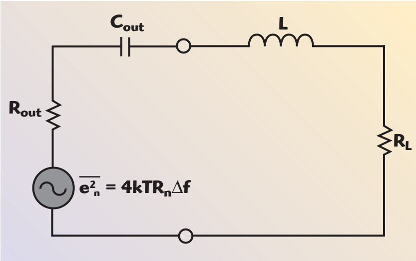
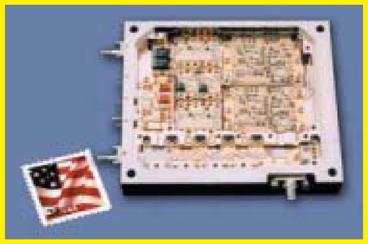
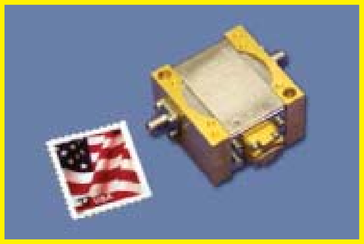


Fig. 5 Simplified negative resistance oscillator noise model.





By normalizing the power P_L dissipated in the load resistor R_L, Equation 18 can be rewritten through the spectral power densities by

$$S_\phi = \frac{S_{\Delta\phi}}{\left(\frac{\Delta R}{R_L}\right)^2 + Q_L^2\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right)^2} \qquad (20)$$

where $S_{\Delta\phi} = 4kTR_n/R_LP_L$ and $S_\phi = i_n^2R_L/P_L$ is the normalized power spectral density of the noise current across the load resistor R_L and $\Delta f = 1$ Hz.

Since at small offset frequencies $\omega_m = \omega - \omega_0$, close to the resonant frequency,

$$\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \cong \frac{2\omega_m}{\omega_0}$$

Equation 20 can then be rewritten in terms of the current and voltage power spectral densities as

$$S_\phi = \frac{S_{\Delta\phi}}{\left(\frac{\Delta R}{R_L}\right)^2 + \left(\frac{2Q_L\omega_M}{\omega_0}\right)^2} \qquad (21)$$

which is similar to the power spectral density at frequency offsets close to the resonant frequency for the parallel feedback oscillator. Equation 21 represents a Lorentz function corresponding to an exponential decay of the autocorrelation function in the time domain.[11] Since the total output power delivered to the load is equal to P_L,

$$\frac{1}{2\pi}\int_0^\infty S_\phi(\omega)d\omega = \left(\frac{\omega_0}{2Q_L}\right)^2 \frac{S_{\Delta\phi}}{2\Delta\omega_n} = 1 \qquad (22)$$

where

$$\Delta\omega_n = \frac{\Delta R}{R_L}\frac{\omega_0}{2Q_L}$$

is the Lorentzian linewidth (half-width at half-maximum), which is an oscillator spectrum linewidth characterized by the natural phase fluctuations, due to the thermal and shot noises of the oscillator. However, in a common case, due to the variation of the oscillator resonant circuit parameters, flicker noise, pushing or pulling effects, the effective spectrum linewidth widens, especially close to the resonant frequency.

By using a widely used definition of the loaded quality factor of the passive resonator in the form of

$$Q_L = \frac{\omega_0}{\Delta\omega_{3dB}} \qquad (23)$$

where $\Delta\omega_{3dB}$ is the full linewidth at half-maximum, one can write

$$2\Delta\omega_n = \frac{S_{\Delta\phi}}{4}\Delta\omega_{3dB}^2 = kT\frac{R_n}{R_L}\frac{\Delta\omega_{3dB}^2}{P_L} \qquad (24)$$

The normalized power spectral density can be expressed through the Lorentzian linewidth as

$$S_\phi(\omega_m) = \frac{2\Delta\omega_n}{\Delta\omega_n^2 + \omega_m^2} \cong \frac{2\Delta\omega_n}{\omega_m^2} \qquad (25)$$

showing a simple linear relationship between the Lorentzian linewidth and the oscillator phase noise spectrum at offset frequencies $\omega_m >> \Delta\omega_n$.[20] Substituting Equation 24 into

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Equation 25, and taking into account that $F = R_n/R_L$, results in the single-sideband noise to carrier ratio

$$L(f_m) = \frac{kTF}{2P_L}\left(\frac{f_0}{Q_L f_m}\right)^2 \qquad (26)$$

which is similar to the Edson noise formula.[10,12]

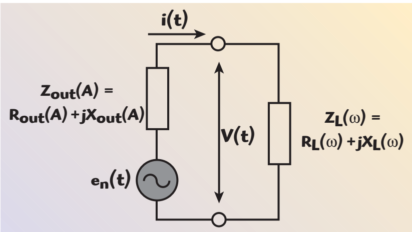

Fig. 6 Equivalent model of a one-port negative resistance oscillator.

Nonlinear Kurokawa Model for Negative Resistance Oscillator

The one-port negative resistance oscillator can generally be represented by the circuit shown in ***Figure 6***. A similar representation of an oscillator circuit can be obtained in terms of admittances. According to the Kurokawa model derivation, based on a small perturbation method, it is assumed that the active device output impedance $Z_{out}(A)$, with a negative real part, is a function of the oscillation amplitude and that the load impedance $Z_L(\omega)$ is a frequency dependent function.[22] By considering the sinusoidal current flowing through the active device with a slowly varying amplitude A(t) and phase $\phi(t)$ and assuming that the equivalent device noise voltage power can be given by $\overline{e_m^2} = 4FkTR_L\Delta f$, the single sideband noise to carrier ratio can be written as

$$L(f_m) = \frac{2FkT}{P_L}\left(1+\frac{f_c}{f_m}\right)\left(\frac{f_0}{2Q_L f_m}\right)^2 \bullet \left[1+\frac{q^2}{p^2+\left(\frac{2Q_L f_m}{f_0}\right)^2}\right] \qquad (27)$$

where

$$p = \frac{A_0 f_0}{2Q_L\left(\mathrm{Re}\,Z_L\right)^2} \bullet \left(\frac{\partial \mathrm{Re}\,Z}{\partial A}\frac{\partial \mathrm{Im}\,Z}{\partial f} - \frac{\partial \mathrm{Im}\,Z}{\partial A}\frac{\partial \mathrm{Re}\,Z}{\partial f}\right) \qquad (28)$$

is a function of the oscillator stability conditions, given in parentheses, characterizing the velocity of the establishment of the steady-state oscillators under small perturbations,

$$q = \frac{A_0 f_0}{2Q_L\left(\mathrm{Re}\,Z_L\right)^2} \bullet \left(\frac{\partial \mathrm{Re}\,Z}{\partial A}\frac{\partial \mathrm{Re}\,Z}{\partial f} + \frac{\partial \mathrm{Im}\,Z}{\partial A}\frac{\partial \mathrm{Im}\,Z}{\partial f}\right) \qquad (29)$$

is the parameter illustrating the dependence of the oscillation frequency on the oscillation amplitude in a large-signal mode of operation.

$$Q_L = \frac{f_0}{2\,\mathrm{Re}\,Z_L}\sqrt{\left(\frac{\partial \mathrm{Re}\,Z}{\partial f}\right)^2+\left(\frac{\partial \mathrm{Im}\,Z}{\partial f}\right)^2} \qquad (30)$$

is the oscillator loaded quality factor, P_L is the output power delivered to the load R_L, $Z = Z_{out} + Z_L$ is the overall circuit impedance.[14]

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From Equation 27, it follows that the oscillation becomes very noisy as one approaches the boundary of the stable region, that is, as the parameter p becomes close to zero. In addition, the greater the value of the parameter q, the higher the phase noise level is expected in the osclator spectrum due to amplitude-to-phase conversion. It should be noted that p = 1 when the oscillator circuit is adjusted to maximum power.[13]

For a particular case of a voltage-controlled oscillator (VCO), when it is assumed that $\partial ReZ/\partial f = 0$, Equation 27 can be rewritten as

$$L\left(f_m\right)=\frac{2FkT}{P_L}\left(1+\frac{f_c}{f_m}\right)\left(\frac{f_0}{2Q_L f_m}\right)^2+\frac{2kTR_e K_{VCO}^2}{f_m^2} \qquad (31)$$

where R_e is the equivalent noise resistance of the varactor, K_{VCO} is the oscillator voltage gain in Hz/V.[15]

Effect of Device and Circuit Nonlinearities

From the Kurokawa model for a negative resistance oscillator, it follows that the nonlinearity of the elements of the device equivalent circuit has a significant impact on the oscillator noise spectrum. The noise analysis, based on the nonlinear MESFET model with dominant nonlinear elements, shows that the gate-source capacitance is responsible for the conversion of low frequency noise into phase noise, whereas the amplitude noise is primarily determined by the nonlinear transconductance.[8,16] Compared to the transconductance and gate-source capacitance, the nonlinear drain-source conductance is of minor importance for the noise upconversion process.

As an example, consider an oscillator with a nonlinear output resistance dependent on the applied DC bias voltage and on the amplitude of the self-sustained oscillations. The basic oscillator circuit with the nonlinear negative output resistance R_{out}, capacitance C, inductance L, load resistance R_L and noise current $i_n(t)$ is shown in ***Figure 7***. The electrical behavior of such an oscillator, in terms of the voltage v(t) across the capacitance, can be represented by a second-order nonlinear differential equation

$$LC\frac{d^2v}{dt^2}+\frac{L}{R_L}\frac{dv}{dt}+v+L\frac{di}{dt}=e_n\left(t\right) \qquad (32)$$

where

$$e_n\left(t\right)=L\frac{di_n\left(t\right)}{dt} \qquad (33)$$

is the equivalent noise voltage and

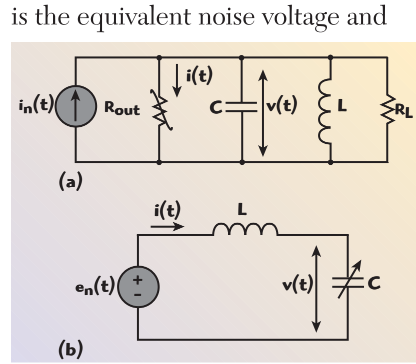
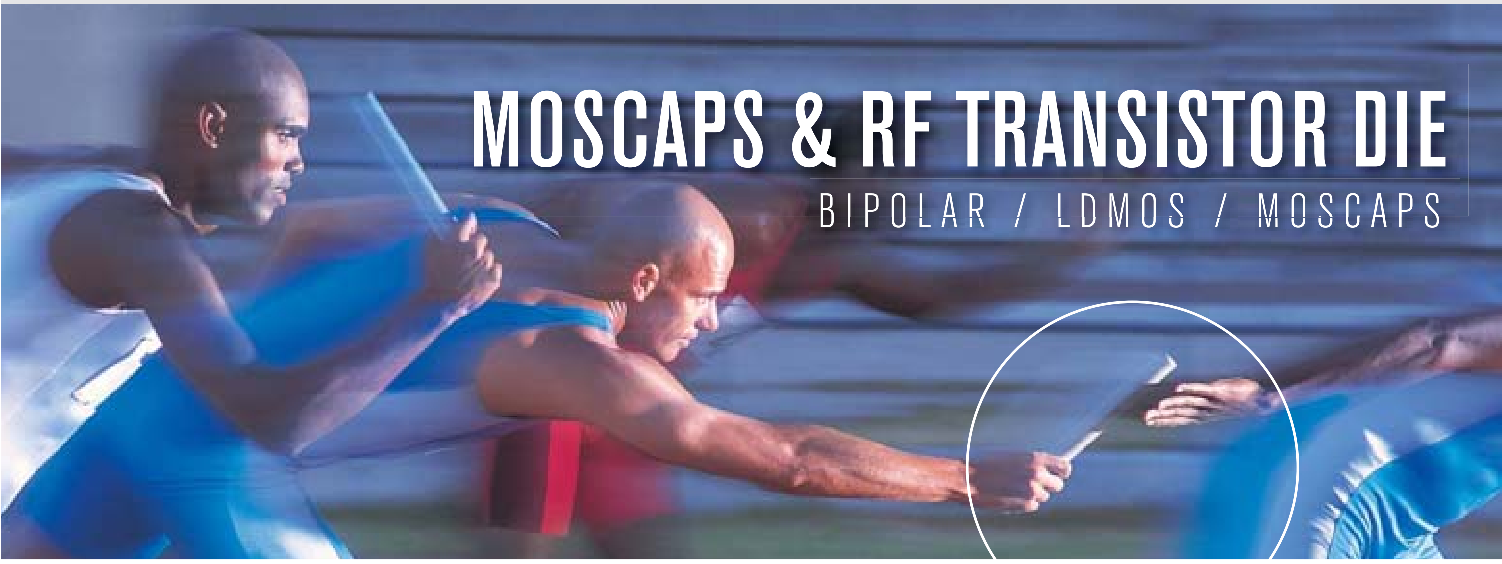


Fig. 7 Second-order nonlinear oscillation systems.



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MOS1-15	15	160	30	Gold	Gold	No
MOS1-36	36	160	30	Gold	Gold	No
MOS1-55	55	160	30	Gold	Gold	No
MOS1-80	80	160	30	Gold	Gold	No
MOS1-130	130	160	30	Gold	Gold	No
MOS7-115	115	185	30	Gold	Gold	YES
MOS7-190	190	185	30	Gold	Gold	YES
MOS4-25	25	200	30	Gold	Gold	No
MOS4-25	25	200	30	Gold	Gold	No
MOS4-38	38	200	30	Gold	Gold	No
MOS4-55	55	200	30	Gold	Gold	No
MOS4-115	115	200	30	Gold	Gold	No
MOS4-215	215	200	30	Gold	Gold	No
MOS8-14	14	215	30	Gold	Gold	No
MOS8-34	34	215	30	Gold	Gold	No
MOS8-41	41	215	30	Gold	Gold	No
MOS8-68	68	215	30	Gold	Gold	No
MOS8-200	200	215	30	Gold	Gold	No
MOS45-25	25	230	30	Gold	Gold	No
MOS45-35	35	230	30	Gold	Gold	No
MOS45-215	215	230	30	Gold	Gold	No
MOS2-19	19	240	30	Gold	Gold	No
MOS2-225	225	240	30	Gold	Gold	No
C16A60	16	160	60	Gold	Gold	No
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$$i = I_0 + \frac{v}{R_{out}} + \sum_{k=2}^{\infty} G_k v^k \qquad (34)$$

represents a power series expansion where I_0 is the DC current and G_k are the small coefficients.

In a steady-state operation mode, when the active device compensates for the losses in the load resistance, that is $R_{out} + R_L = 0$, Equation 32 can be rewritten as

$$LC\frac{d^2v}{dt^2} + v + L\frac{d}{dt}\left(\sum_{k=2}^{\infty} G_k v^k\right) = e_n(t) \qquad (35)$$

Seeking the general solution of the inhomogeneous differential equation as the superposition of the general solution of the homogeneous (noise-free) and specific solutions of Equation 35 as

$$v(t) = V(t)\cos\left[\omega_0 t + \phi(t)\right] + e_n(t) \qquad (36)$$

and applying a van der Pol approach for the slowly time-varying amplitude V(t) and phase ϕ(t), allows rewriting Equation 35 in the form

$$2\omega_0 \frac{dV}{dt}\sin(\omega_0 t + \phi) + 2\omega_0 V\frac{d\phi}{dt}\cos(\omega_0 t + \phi) = \frac{1}{C}\frac{d}{dt}\left(\sum_{k=2}^{\infty} G_k v^k\right) \qquad (37)$$

where $\omega_0 = 1/\sqrt{LC}$, and it is assumed that $e_n(t)$ is a small, slowly time-varying, low frequency noise voltage, for which

$$LC\frac{d^2e_n(t)}{dt^2} << e_n(t)$$

As a result, substituting Equation 36 into the right-hand side of Equation 37 and using trigonometric identities yields

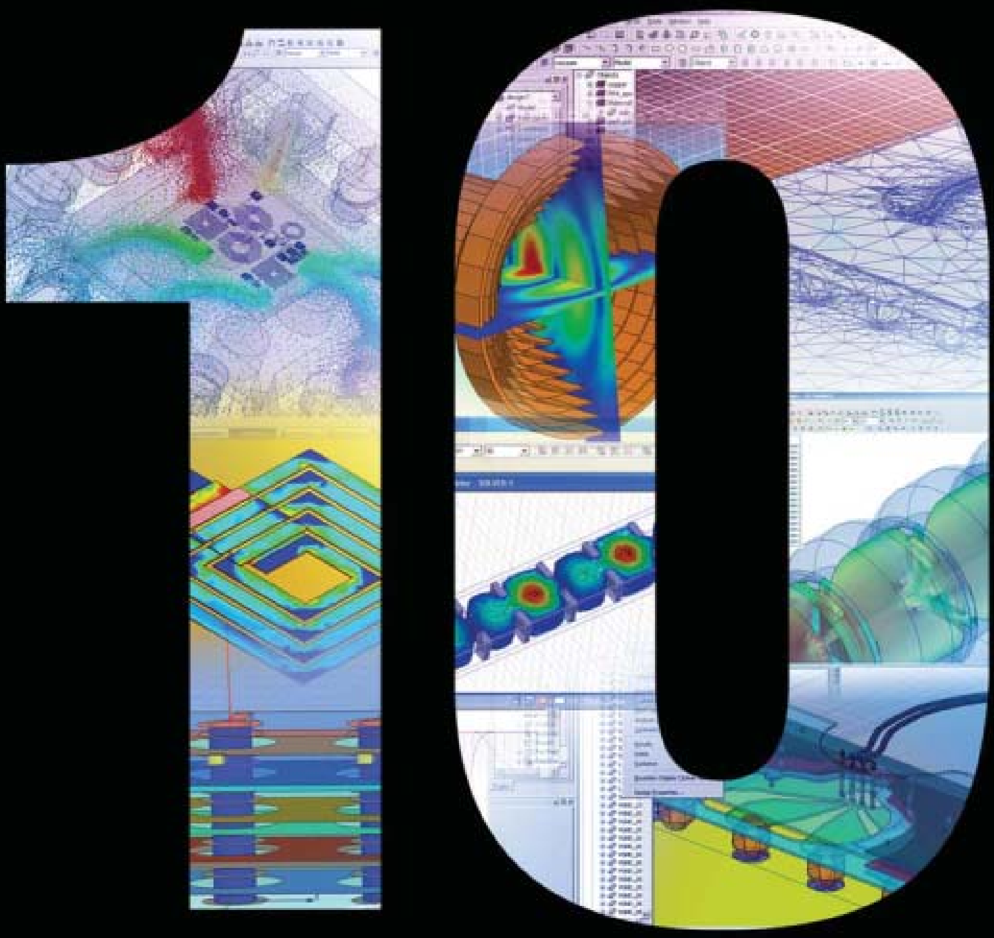
$$\frac{d\phi}{dt} = 0 \qquad (38)$$

which means that the nonlinear output resistance has no impact on the phase fluctuations. However, the amplitude fluctuations are not equal to zero because all factors on the right-hand side of Equation 37 have first-order sine components. Thus, the resistive type of nonlinearities alone would cause amplitude noise only, since the reactive elements determining the oscillation frequency remain constant.

Now consider a varactor-controlled oscillator with the varactor as a nonlinear element whose capacitance depends not only on the applied DC bias voltage but also on the amplitude of the self-sustained oscillations. The basic VCO circuit consists of the varactor with a nonlinear capacitance C, an inductance L and a noise voltage $e_n(t)$, as shown in the figure.[26] The voltage $e_n(t)$, can represent all the noise coming from both inside and outside the circuit, including any thermal noise from the resistors, flicker noise from the active device and noise from the power supply. The electrical behavior of the oscillator can be described by

$$v + L\frac{di}{dt} = e_n(t) \qquad (39)$$

$$i = \left(C + v\frac{dC}{dv}\right)\frac{dv}{dt} \qquad (40)$$



where the nonlinear term vdC/dv is included in Equation 40.

By expanding a nonlinear capacitance C into a power series

$$C = C_0 + \sum_{k=1}^{\infty} C_k v^k \tag{41}$$

with small coefficients C_k, substituting Equation 40 into Equation 39 and applying an asymptotic perturbation procedure with decomposition of the perturbed and unperturbed equations, the first-order differential equation for phase fluctuations with the slowly time-varying noise voltage e_n can be derived as

$$\frac{d\phi}{dt} = -\frac{\omega_0}{C_0}\left[C_1 e_n + C_2\left(\frac{3}{4}V^2 + 3e_n^2\right) + C_3 e_n\left(3V^2 + 4e_n^2\right) + \ldots\right] \tag{42}$$

where V is the voltage amplitude across the varactor.[17] Note that the nonlinear capacitance has no impact on the amplitude noise of the oscillator.

From Equation 42, it follows that:

- The first-order capacitance nonlinearity described by the coefficient C_1 contributes to the upconversion of the low frequency noise $e_n(t)$ to the sideband noise near the carrier ω_0.
- The second-order nonlinearity described by the coefficient C_2 generates a phase noise, due to both amplitude-to-phase conversion and low frequency noise upconversion.
- The higher order nonlinearities described by coefficients C_k, k = 3, 4, 5, ..., cause a more complicated noise behavior of the oscillator based on hybrid upconversion and amplitude-to-phase conversion due to cross terms of V and e_n.

In the case of a single-frequency LC oscillator, the main contributor to the phase noise is the nonlinear collector capacitance of the bipolar device or the gate-source capacitance of the FET device.

In a general case, the equivalent circuit of the active device is very complicated, including both nonlinear intrinsic and linear parasitic external elements. This means that it is difficult to evaluate analytically the impact of each nonlinear element on the upconversion mechanism. Moreover, the joint effect of different nonlinear circuit elements will result in both amplitude and phase fluctuations. ***Figure 8*** shows the nonlinear MESFET equivalent circuit with input and output current noise sources. As can be determined from numerical calculations, the phase noise can be significantly reduced by linearizing both the transconductance g_m and the gate-source capacitance C_{gs}; in other words, both nonlinearities are important contributors to the phase noise.[18] The amplitude noise also depends on the capacitance and transconductance nonlinearities. However, the capacitance nonlinearity will not affect the output current if the series gate resistance R_g is set to zero. The nonlinearities of the gate-drain capacitance C_{gd} and drain-

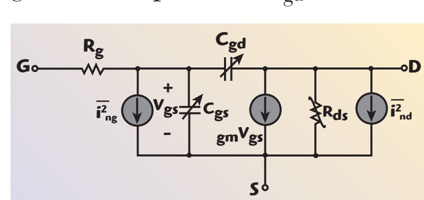


Fig. 8 Equivalent circuit of a MESFET device including noise sources.

source resistance R_{ds} have negligible effect on the amplitude and phase noise.

CONCLUSION

Both linear and nonlinear phase noise models for the parallel feedback and negative resistance oscillators are discussed in detail, with demonstrations of their advantages and drawbacks. The linear Leeson model for a parallel feedback oscillator is very simply expressed in an explicit analytical form, which can explain the dependence of the oscillator phase noise on the resonant circuit loaded quality factor, signal power, active device noise figure and its low frequency flicker noise. However, it cannot explain the effect of an amplitude-to-phase conversion and higher order harmonics. At the same time, the nonlinear Kurokawa model developed for a negative resistance oscillator demonstrates the explicit analytical relationship between the phase noise and oscillator stability margin, and shows the dependence of the oscillation frequency on the oscillation amplitude in a large-signal operation mode in the form of derivatives of the device and circuit impedances. Also, the individual and joint effects of different nonlinear circuit elements, which will result in both amplitude and phase fluctuations, are analyzed and discussed. ■

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AFS2-04000800-20-TC-2	4–8	17-22	1.00	2.0	2.0:1	2.0:1	+5	70
AFS3-04000800-18-TC-4	4–8	25-30	1.00	1.8	2.0:1	2.0:1	+8	100
AFS2-02000800-40-TC-2	2–8	14-19	1.50	4.0	2.0:1	2.0:1	+5	70
AFS3-02000800-30-TC-4	2–8	22-27	1.50	3.0	2.0:1	2.2:1	+8	150
AFS2-08001200-30-TC-2	8–12	12-16	1.00	3.0	2.0:1	2.0:1	+5	70
AFS3-08001200-22-TC-4	8–12	24-28	1.00	2.2	2.0:1	2.0:1	+8	100
AFS4-12001800-30-TC-8	12–18	22-26	1.00	3.0	2.0:1	2.0:1	+8	250
AFS4-06001800-35-TC-8	6–18	22-26	1.00	3.5	2.0:1	2.0:1	+8	250
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AFS3-00100200-25-27P-6	0.1–2	34	1.50	2.5	2.0:1	2.5:1	+27**	275
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AFS3-01500250-06-13P-6	1.5–2.5	38	1.00	0.60	1.8:1	1.8:1	+13	150
AFS3-01700190-06-13P-6	1.7–1.9	38	0.50	0.60	1.5:1	1.5:1	+13	150
AFS3-01800220-06-13P-6	1.8–2.2	38	0.50	0.60	1.5:1	1.5:1	+13	150
AFS3-02200230-06-13P-4	2.2–2.3	38	0.50	0.60	1.5:1	1.5:1	+13	150
AFS3-02300270-06-13P-6	2.3–2.7	36	0.50	0.60	1.5:1	1.5:1	+13	150
AFS3-02700290-06-13P-6	2.7–2.9	32	0.50	0.60	1.5:1	1.5:1	+13	150
AFS3-02900310-06-13P-6	2.9–3.1	32	0.50	0.60	1.5:1	1.5:1	+13	150
AFS3-03100350-06-10P-4	3.1–3.5	29	0.50	0.60	1.5:1	1.5:1	+10	150
AFS4-03400420-10-13P-6	3.4–4.2	40	0.50	1.00	1.5:1	1.5:1	+13	200
AFS3-04400510-07-S-4	4.4–5.1	30	0.50	0.70	1.5:1	1.5:1	+10	100
AFS3-04500480-07-S-4	4.5–4.8	30	0.50	0.70	1.5:1	1.5:1	+10	100
AFS3-05200600-07-10P-4	5.2–6	30	0.50	0.70	1.5:1	1.5:1	+10	100
AFS3-05400590-07-S-4	5.4–5.9	30	0.50	0.70	1.5:1	1.5:1	+10	100
AFS3-05800670-07-S-4	5.8–6.7	30	0.50	0.70	1.5:1	1.5:1	+10	100
AFS3-07250775-06-10P-4	7.25–7.75	30	0.50	0.60	1.5:1	1.5:1	+10	100
AFS3-07900840-07-S-4	7.9–8.4	30	0.50	0.70	1.5:1	1.5:1	+10	100
AFS4-08500960-08-S-4	8.5–9.6	32	0.75	0.80	1.5:1	1.5:1	+10	125
AFS3-09001100-09-S-4	9–11	26	0.50	0.90	1.5:1	1.5:1	+10	100
AFS4-09001100-09-S-4	9–11	32	0.75	0.90	1.5:1	1.5:1	+10	125
AFS4-10951175-09-S-4	10.95–11.75	32	0.75	0.90	1.5:1	1.5:1	+10	125
AFS4-11701220-09-5P-4	11.7–12.2	32	0.75	0.90	1.5:1	1.5:1	+10	125
AFS2-12201280-14-5P-2	12.2–12.8	14	0.75	1.40	1.4:1	1.5:1	+5	80
AFS4-12201280-13-12P-4	12.2–12.8	25	1.50	1.30	2.0:1	2.0:1	+12	200
AFS4-12701330-15-10P-4	12.7–13.3	30	0.75	1.50	1.5:1	1.5:1	+10	175
AFS4-13201400-16-10P-4	13.2–14	30	0.75	1.60	1.5:1	1.5:1	+10	175
AFS4-14001450-15-10P-4	14–14.5	30	0.75	1.50	1.5:1	1.5:1	+10	175
AFS4-20202120-25-8P-4	20.2–21.2	24	1.00	2.50	1.5:1	1.5:1	+8	175
AFS4-21202400-28-10P-4	21.2–24	23	1.00	2.80	2.0:1	2.0:1	+10	100
OCTAVE BAND AMPLIFIERS								
AFS3-00120025-09-10P-4	0.12–.25	38	0.50	0.9	2.0:1	2.0:1	+10	125
AFS3-00250050-08-10P-4	0.25–0.5	38	0.50	0.8	2.0:1	2.0:1	+10	125
AFS3-00500100-06-10P-6	0.5–1	38	0.75	0.6	2.0:1	1.5:1	+10	150
AFS3-01000200-05-10P-6	1–2	38	1.00	0.5	2.0:1	2.0:1	+10	150
AFS3-01200240-06-10P-6	1.2–2.4	34	1.00	0.6	2.0:1	2.0:1	+10	150
AFS3-02000400-06-10P-4	2–4	32	1.00	0.6	2.0:1	2.0:1	+10	125
AFS3-02600520-10-10P-4	2.6–5.2	28	1.00	1.0	2.0:1	2.0:1	+10	125
AFS3-04000800-07-10P-4	4–8	28	1.00	0.7	2.0:1	2.0:1	+10	125
AFS3-08001200-09-10P-4	8–12	26	1.00	0.9	2.0:1	2.0:1	+10	125
AFS3-08001600-15-8P-4	8–16	28	1.00	1.5	2.0:1	2.0:1	+8	100
AFS4-12001800-18-10P-4	12–18	28	1.50	1.8	2.0:1	2.0:1	+10	125
AFS4-12002400-30-10P-4	12–24	24	2.00	3.0	2.0:1	2.0:1	+10	85
AFS3-18002650-30-8P-4	18–26.5	18	1.75	3.0	2.2:1	2.2:1	+8	125
MULTIOCTAVE BAND AMPLIFIERS								
AFS3-00300140-09-10P-4	0.3–1.4	38	1.00	0.9	2.0:1	2.0:1	+10	125
AFS2-00400350-12-10P-4	0.4–3.5	22	1.50	1.2	2.0:1	2.0:1	+10	80
AFS3-00500200-08-15P-4	0.5–2	38	1.00	0.8	2.0:1	2.0:1	+15	125
AFS3-01000400-10-10P-4	1–4	30	1.50	1.0	2.0:1	2.0:1	+10	125
AFS3-02000800-09-10P-4	2–8	26	1.00	1.0	2.0:1	2.0:1	+10	125
AFS4-02001800-23-10P-4	2–18	25	2.00	2.3	2.0:1	2.0:1	+10	175
AFS4-06001800-22-10P-4	6–18	25	2.00	2.2	2.0:1	2.0:1	+10	125
AFS4-08001800-22-10P-4	8–18	28	2.00	2.2	2.0:1	2.0:1	+10	125
ULTRA WIDEBAND AMPLIFIERS								
AFS3-00100100-09-10P-4	0.1–1	38	1.00	0.9	2.0:1	2.0:1	+10	125
AFS3-00100200-10-15P-4	0.1–2	38	1.00	1.0	2.0:1	2.0:1	+15	150
AFS1-00040200-12-10P-4	0.04–2	15	1.50	1.2	2.0:1	2.0:1	+10	50
AFS3-00100300-12-10P-4	0.1–3	32	1.00	1.2	2.0:1	2.0:1	+10	125
AFS3-00100400-13-10P-4	0.1–4	28	1.00	1.3	2.0:1	2.0:1	+10	125
AFS3-00100600-13-10P-4	0.1–6	28	1.25	1.3	2.0:1	2.0:1	+10	125
AFS3-00100800-14-10P-4	0.1–8	28	1.50	1.4	2.0:1	2.0:1	+10	125
AFS4-00101200-22-10P-4	0.1–12	30	1.50	2.2	2.0:1	2.0:1	+10	150
AFS4-00101400-23-10P-4	0.1–14	24	2.00	2.3	2.5:1	2.5:1	+10	200
AFS4-00101800-25-S-4	0.1–18	25	2.00	2.5	2.5:1	2.5:1	+10	175
AFS4-00102000-30-10P-4	0.1–20	20	2.50	3.0	2.5:1	2.5:1	+10	125
AFS4-00102650-42-8P-4	0.1–26.5	22	2.50	4.2	2.5:1	2.5:1	+8	135

Note: Noise figure increases below 500 MHz in bands greater than 0.1-10 GHz.

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SAMPLING IF FILTERS AND THE RETURN OF THE SUPERHETERODYNE RECEIVER

Wireless systems-on-a-chip (SoC) have seen very high levels of integration in the last few years because the market is demanding reduced cost and low component count. Further to this trend, the integration of multiple standards has been introduced into the SoC. This integration strategy has caused superheterodyne receiver architectures to fall out of favor, thereby sacrificing performance. This article promotes analog front ends that are suitable for integrating into a SoC to relieve requirements on standard digital techniques and A/D converters.

A new technology called a "Sampling IF Filter" (SIF) will be discussed and it will be shown how it is more suitable for integrating filters onto a SoC. This is an enabling technology, which will re-popularize the superheterodyne receiver architecture for wireless SoC applications. How this technology can facilitate receiver design in general will also be discussed. To place the sampling IF filter in context, the current status of on-chip filters will be reviewed.

EXISTING FILTER TECHNOLOGIES

Three well-known filter technologies have been used for integrating IF filters on chip. The oldest is the active resistor capacitor (RC) filters. Another old technology for on-chip IF filters is the switched capacitor (SC) filter. The transconductor capacitor (g_m–C) filter has been used in more recent radio designs. Each approach has its own strengths and weaknesses, which often makes selection of the most appropriate approach difficult.

Active Resistor Capacitor Filters

These are typically constructed from bi-quad sections, where each bi-quad has a network of resistors, capacitors and op-amps. The bi-quad sections are then cascaded to produce the desired filter response. The major issue with this approach is that both the resistor and capacitor tolerances on-chip are not tight enough. This can yield a large range of RC time constants and consequently a large error in the frequency response. On-chip, the ratios of like devices such as resistor-to-resistor and capacitor-to-capacitor can be better controlled within a few percent, but the variations in time constants can vary by ±30 percent in a typical process. In an attempt to improve the frequency accuracy, SC filters were introduced.

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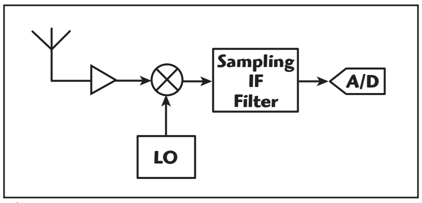
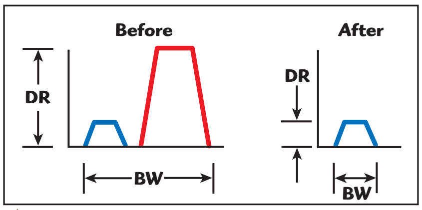


▲ *Fig. 1 Sampling IF filter.*



▲ *Fig. 2 Reduced dynamic range and bandwidth.*

SC Filters

SC filters are constructed by substituting the resistors in an active RC filter with switches and capacitors. These SC filters have very precisely defined passband characteristics, because the time constants associated with the frequency response depend only on the capacitor ratios and the clock frequency. A serious drawback with using SC filters in an IF stage is the danger of aliasing interfering signals. Using SC filters for an IF filter can result in noise problems because of the difficulty in obtaining low noise samples of the input signal. Since the op-amps used in SC filters must have greater bandwidth than the signal they are processing, they tend to consume too much power for high data rate applications that are battery operated. The noise problem can be solved by using larger capacitors, although this worsens the problems with op-amp bandwidth. These noise and aliasing issues are further compounded in a low power, low clock-rate system. In some ways, these filters are superior to g_m–C filters (see below) in that they are not sensitive to variations in temperature and process. This means that a working SC filter can be designed to achieve first silicon success, unlike the g_m–C filters that usually require several passes.

g_m–C Filters

g_m–C filters are constructed by replacing the inductor in an LC filter with a capacitor and a gyrator made with a transconductance amplifier. Typically, g_m–C filters provide lower noise and lower power in a given process than SC filters but suffer in two problem areas where SC filters perform better. The first problem is that, typically, a trade-off is required between power and linearity. The second is the trade-off between linearity and bandwidth. Any attempts at improving linearity force the designer to use multi-transistor g_m elements that reduce the available bandwidth by introducing extra parasitic poles. Since the time constants in a g_m–C filter depend on two independent process variables (g_m and C), they tend to have poorly controlled passband frequency response characteristics, unless a process calibration loop is included. This poorly controlled frequency response can adversely affect the chip yield and result in a re-spin of the chip to produce the desired transfer function (the transfer function of a filter is the gain or attenuation versus the input frequency).

SAMPLING IF FILTERS

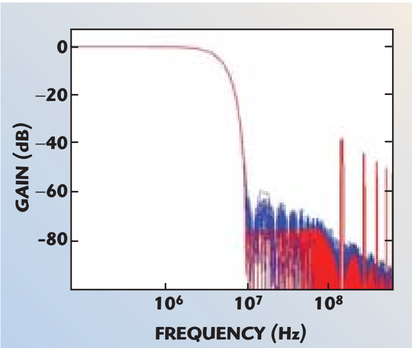
This article highlights a new filter technology called "Sampling IF Filter" (SIF) (see ***Figure 1***). This technology combines the low power, low noise properties of a g_m-C filter with the precisely controlled passband and process independence of a SC filter. The product has an automatic gain control (AGC) stage, an anti-alias filter, a channel select filter and a sampler in a single unit. This low power technique can be part of an integration strategy. Currently, designs of on-chip filters are up to 900 MHz and can be used to replace an off-chip SAW filter. As the technology lends itself to being field programmable, a sampling IF filter can replace several off-chip filters. The sampling IF filter is also process and temperature tolerant. This can help in lowering costs at the foundry as it reduces the number of fabrication runs and, because of this, manufacturers can get the product to market faster, enabling them to engage customers early on. Further, as the channel selection filtering is performed before the A/D converter as part of the sampler, this method substantially reduces both speed and resolution requirements specified for the A/D converter (see ***Figure 2***). In turn, this provides additional power savings by allowing the use of a much simpler A/D converter. Without filtering before the A/D, the bandwidth and dynamic range need to accommodate both the signal and any interfering signals. By sampling and filtering prior to the digital stage, the bandwidth and the dynamic range of the A/D can now be reduced. As the requirements for the A/D have been reduced, the A/D is no longer a high performance component and can easily be integrated as an add-on to the sampling IF filter. As stated earlier, the disadvantages of SC filters are that they are noisy and have high power consumption, although their transfer functions are stable over temperature and process variations. It was stated previously that g_m–C filters offered low power consumption, but their drawback is that they are less stable over the same variations. The sampling IF filter presented here represents a valuable breakthrough, because it is superior to the strengths of both of these filter technologies.

Field Programmable

The sampling IF filters can be designed for high RF image rejection without any requirement for digital correction techniques. Sampling IF filters are very versatile and can be applied to receivers in WiMAX 802.16, Bluetooth, GSM, WLAN (Wi-Fi) 802.11a, 802.11b, 802.11g, 802.11h and 802.11n, software-defined radios (SDR), multi-mode radios, pagers, and cable modem products. In multi-mode radio applications, there is a great advantage to making the sampling IF filter programmable, as all off-chip filters can now be integrated.

Process and Temperature Independence

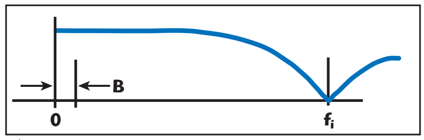
Changes in process or temperature in an RC filter change the time constants, which change the transfer function, making it difficult to achieve a higher order filter. In a sampling IF filter, process and temperature variation will only change the gain, which has no effect on the transfer function because the SIF is embedded inside an AGC loop, which will correct any gain error. With this last gain error corrected, the SIF is completely insensitive to any process and temperature varia-



▲ *Fig. 3 Effect of mismatch.*

tions. Automatic gain control can be added at any stage of the sampling IF filter. The AGC loops are digital and therefore can be very fast. Packet type architectures usually require a fast AGC because it is necessary that the AGC level be set before the training sequence or packet information becomes available for demodulation.

The red line, shown in ***Figure 3***, represents the response of the filter with no process or temperature variations included. The blue lines represent the effects of component mismatch and all other circuit and temperature variations. The frequency response of the passband is unaffected by any process or temperature variations. The stopband attenuation of the aliases is also maintained even with component mismatch. The sampling IF filter technique is less susceptible to process variations than either of the two incumbent on-chip filter technologies.

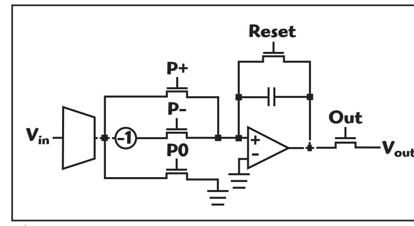


▲ *Fig. 4 Frequencies and bandwidth in the common anti-alias filter building block.*

The SIF filter is composed of different stages that can be assembled for the desired application. Some example stages of a sampling IF filter that can be used are discussed below.

The Common Anti-alias Filter Building Block

A common anti-alias filter building block, used in sampling intermediate frequency filters, is the finite impulse response (FIR) filter. This filter function is produced by the continuous time integration of the incoming signal over a minimum integration interval, T_i. This integrate-and-dump filter has a transfer function of sinc (f/f_i), where $f_i = 1/T_i$ is the integration frequency. The overall SIF filter has a passband B (see ***Figure 4***). Signals that will alias into this passband are those signals that will occur at frequencies that are multiples of f_i. After sampling, the amount of rejection for these signals is approximately 20 log (f_i/B). When the system architecture requires continuous-time pre-filtering in addition to the common sampling IF filter attenuation of 20 log (f_i/B), this pre-filtering can be done with passive elements or g_m–C filters. Here is a case where low tolerance g_m–C filters can be useful, because precision in the filter transfer function is not required. The only requirement is to supply substantial attenuation at frequencies close to f_i and its multiples.



▲ *Fig. 5 Block diagram of an enhanced anti-alias filter.*

Enhanced Anti-alias Filter by Extended Integration Intervals

By integrating over several periods of f_i, to produce a lower output frequency, f_s, the common anti-alias filter can be made more selective. This enhanced selectivity can be created either at DC, or, in the case of poly-phase filters, at certain discrete intermediate frequencies. For poly-phase filters, the center frequency is set by $F_{cAA} = N_1\,(f_i/N_2)$, where F_{cAA} is the center frequency of the common anti-alias filter, N_1 is a number used to select a channel or band of frequencies and N_2 is the number of phases in the poly-phase filter. If N_1 is made programmable and N_2 is greater than four, this will give some flexibility in the range of intermediate frequencies. Making f_i programmable further increases the flexibility of the design to create multiple transfer functions that can be applied to various wireless applications. ***Figure 5***

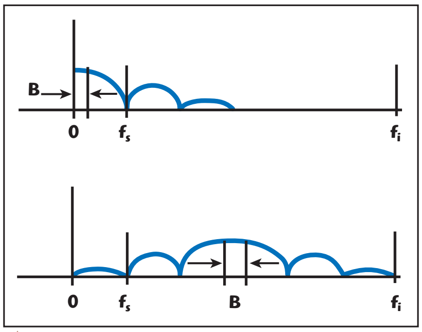


Fig. 6 Frequencies and bandwidth.

shows a simplified circuit diagram of the enhanced anti-alias filter. Typically, this method employs differential circuitry although the figure shows a single ended version to simplify the demonstration. In a differential circuit, the gain of –1 shown can be implemented by crossing a pair of wires to invert the polarity of the current. ***Figure 6*** shows some possibilities of different transfer functions that can be created by using this type of enhanced anti-alias filter. By manipulating N_1 and N_2, the useful bandwidth B can be moved from DC to a range of IF frequencies.

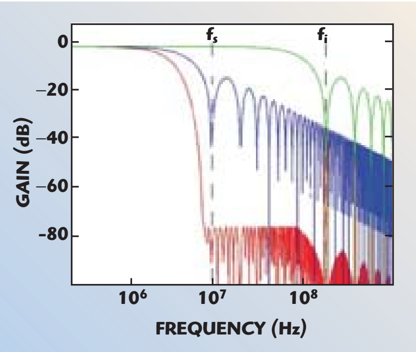


Fig. 7 Frequencies and bandwidth.

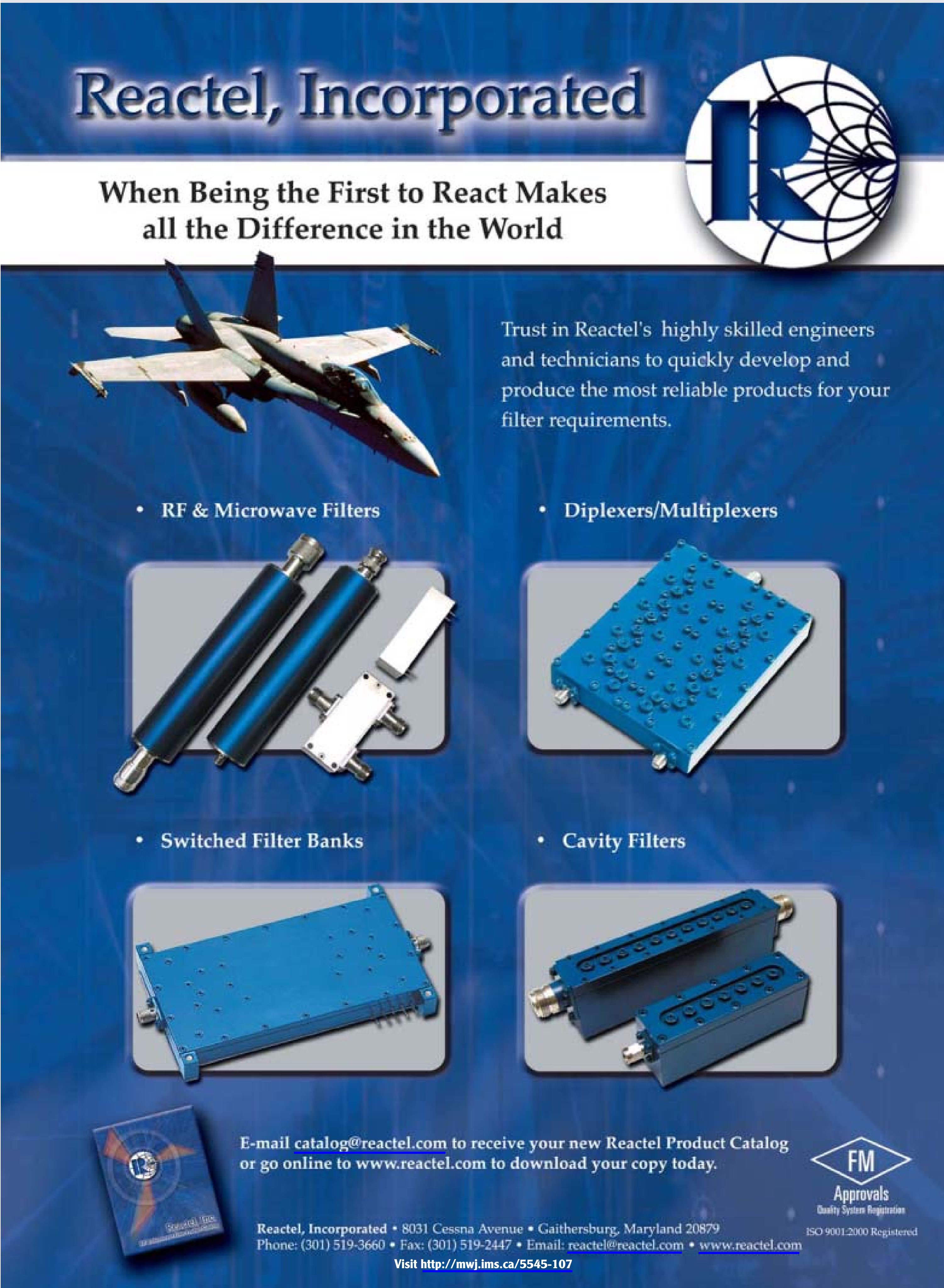
Enhanced Anti-alias Filter by Extended Integration Intervals and Resistor-weighted Tap Coefficients

While the enhanced anti-alias filter provides better anti-alias attenuation than the minimal anti-alias filter, a resistor tap coefficient can be used to provide better attenuation at multiples of the sampling frequency f_s. One of the reasons for having greater attenuation at multiples of f_s is to prevent aliasing when the output of the filter is sampled to the next stage.

This further enhanced anti-alias filter provides additional attenuation between the frequencies f_s and f_i. An example is shown in ***Figure 7***. In this example, the green line shows the built-in common anti-alias filter discussed previously, with its first notch at f_i. The notch occurs at 160 MHz because the integration interval is 6.25 ns. The blue line shows the transfer function resulting from extending the integration interval, as discussed in the previous section, to 16 integration intervals or 100 ns. This gives an output sampling frequency of 10 MHz and a deep notch at 10 MHz. By using 62 resistor-weighted tap coefficients, the transfer function can be created as shown by the red line. The increased attenuation, shown by the red line, is necessary to prevent signals from 10 MHz to f_i from aliasing into the desired signal bandwidth B. The further enhanced anti-alias filter can be created at DC or any frequency up to $f_i/2$. The impulse response is finite (FIR) and has a linear phase. To save area and power when higher levels of filtering are required, additional stages can be added to the anti-alias filter. This allows for the sampling frequency to be reduced in smaller steps with sufficient anti-alias filtering at each stage. This lower sampling frequency will save power in subsequent stages of the filter, the A/D converter and in the Digital Signal Processor (DSP). The idea is to lower the sampling frequency, which in turn lowers the number of tap coefficients required, while achieving the same desired filtering. A benefit of using the sampling IF filter is that filtering is performed on the analog side of the system instead of after the A/D.

Other IF Stages

One advantage of any sub-sampled technique is that a desired alias can be selected to down-convert to baseband. The advantage of the sampling IF filter technique is that other aliases can be easily rejected while not disturbing the desired alias. Previously, there was a high cost to down-conversion and filtering of IF stages. However, with this unique technique, IF down-conversion occurs as part of the sampling IF filter. As there is no longer an added cost for these down-conversion stages, the architectures can be re-examined to gain a competitive advantage. A process-indepen-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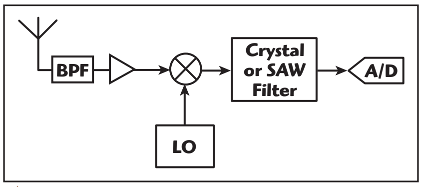
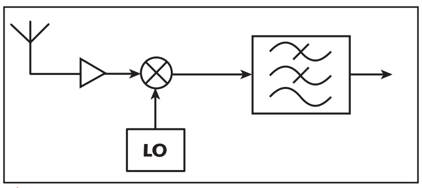


▲ *Fig. 8 Single IF receiver.*



▲ *Fig. 9 Zero IF receiver architecture.*

dent technique for integrating IF stages on chip will be discussed, using a sampling IF filter. This technique permits the integration of multiple IF stages on-chip enabling designers to migrate back to superheterodyne receiver architectures. The sampling IF filter technique offers further improvements by allowing for significant relaxation of the RF front-end requirements by using a higher IF frequency. These relaxed RF front-end requirements can be exploited for further cost reductions. Before looking at superheterodynes, a review of single IF receivers and zero IF receivers is presented.

A REVIEW OF RECEIVER ARCHITECTURES

To fully appreciate the value of the sampling IF filter, the commonly used SoC receiver architectures will be reviewed and their strengths and weaknesses discussed. Several years ago, as SoC integration began, low noise amplifiers, mixers and many of the basic components were integrated. However, there were still plenty of off-chip components, including frequency synthesizers, voltage-controlled oscillators, power amplifiers and filters. Eventually many of these were integrated leaving just the filters to integrate. Finally, when the industry required on-chip filters, there was no available filter technology for high frequencies. This forced the use of alternative architectures that employed low frequency filters that could be constructed from the types discussed earlier.

Single IF Receivers

A simplified block diagram of a single IF or heterodyne receiver is shown in ***Figure 8***. A band-select filter removes any out-of-band signals from the incoming RF signal received by the antenna. An image-reject bandpass filter (BPF) removes the image frequency from the LNA output prior to down-conversion to the desired intermediate frequency by the mixer. Typically, a crystal or SAW filter anti-aliases to allow A/D conversion, then the rest of the receiver is digital. Single-IF architectures have a trade-off between sensitivity and selectivity. Sensitivity relates to the ability of the receiver to distinguish extremely weak signals that are very close to the ambient noise floor. Selectivity is the ability to distinguish these weak signals from other interfering signals that are close in frequency. The potential source of interfering signals can be external or unintentionally generated by other parts of the system. If the IF is high, then the image appears far away from the desired signal band and can easily be suppressed by a bandpass filter. However, the high IF cannot be fed directly into the A/D converters for digital demodulation. If the IF is low, then channel selection becomes easier, but now proper suppression of the image becomes harder to achieve, because a high Q-factor RF selection filter is needed, which is very difficult to design (Q being the ratio of the center frequency to the 3 dB bandwidth). The use of an image-reject mixer can theoretically reduce the selectivity requirements of the image-reject filter, but this becomes difficult with conventional techniques if more than 40 dB of image rejection is required from the mixer.

Zero IF Receivers

A simplified block diagram of a homodyne receiver is shown in ***Figure 9***. The zero IF (ZIF) radio receiver uses a direct-conversion architecture. This means that it utilizes a single mixer stage and avoids any costly IF stages. In this architecture, the received signal is frequency converted directly to baseband. Therefore, a low pass filter is utilized to remove any out of band signals, thus avoiding the requirement for high Q bandpass filters. Many ZIF radio de-

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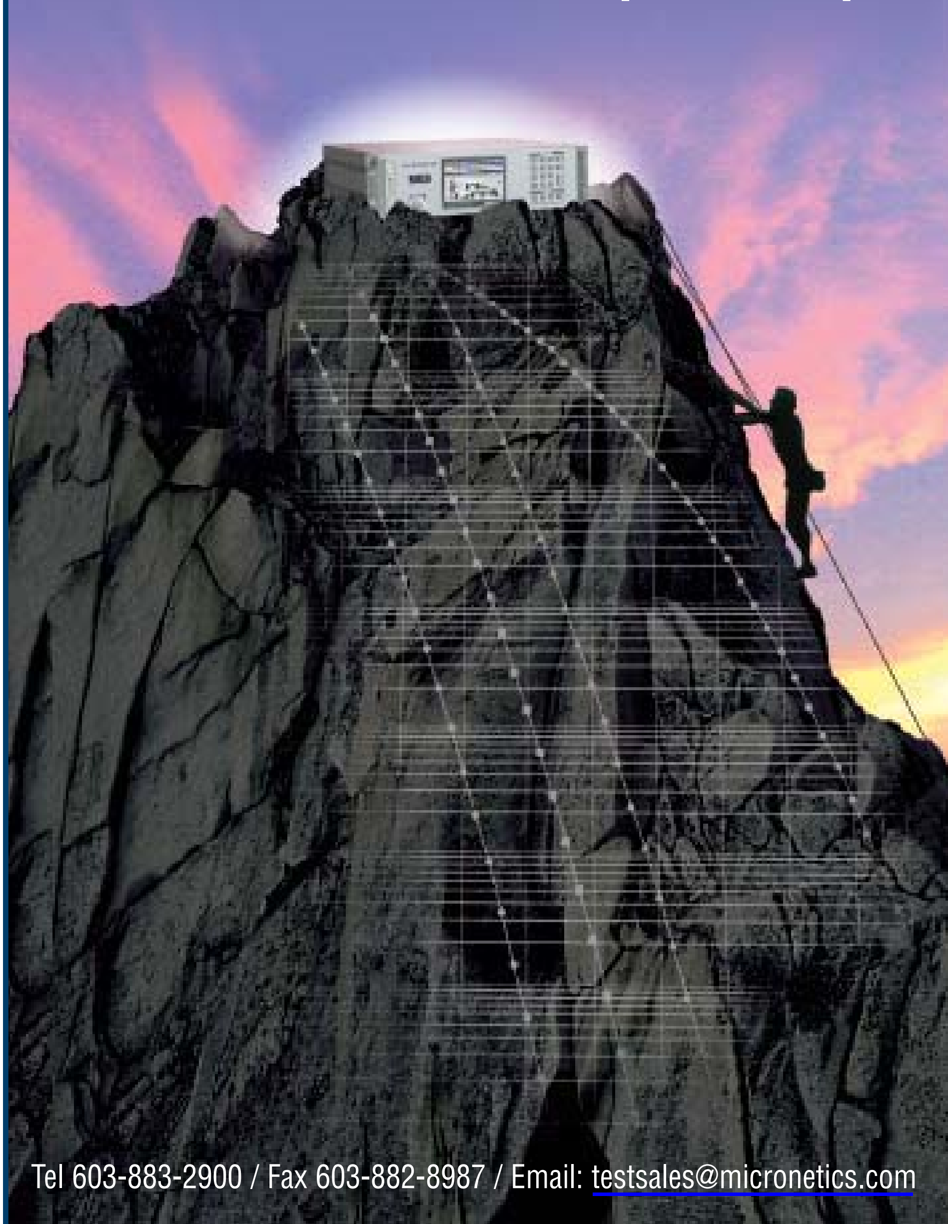

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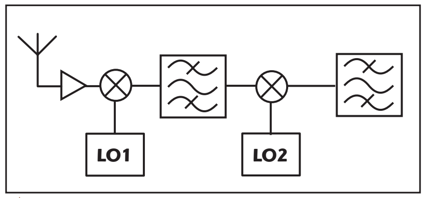


Fig. 10 Superheterodyne receiver architecture.

signs also integrate the low noise amplifier (LNA), voltage-controlled oscillator (VCO) and the baseband filters on a single die. Single-chip ZIF transceivers have been used in cellular applications and are currently being introduced in WLAN radio designs.

Some disadvantages of the ZIF receiver architecture are DC offset and flicker noise. DC offset is generated by any LO leakage entering the receive path. This LO leakage mixes with itself creating a DC component in the signal chain that affects the receiver performance and can saturate the RF stages. Preventing this leakage requires careful attention to the layout of the IC to prevent coupling of the LO or other interfering signals through substrate coupling or mutual inductance of supply lines. Flicker noise, or 1/f noise, is a low frequency noise generated in the transistors, which can corrupt signals in the receiver chain. Flicker noise is a major concern with ZIF architectures, because the desired signal is down-converted to the same low frequency where transistors have their highest noise level. This problem is most severe in pure CMOS processes. Performance in narrowband applications suffers greatly when ZIF architectures are applied. Another issue with ZIF receivers is that most of the gain is at a single baseband frequency in the receive path. Having a high gain at a single frequency can cause instability in the amplifier.

Superheterodyne Receivers

Superheterodyne receivers were once the workhorses of microwave receiving systems (see ***Figure 10***). They were the most popular and the highest quality receiver architecture for many years. Recently, systems architects have moved away from two-stage superheterodyne receivers in favor of single intermediate frequency (IF) or direct conversion architectures (ZIF). The main reason for reducing the number of IF stages in a SoC chip has been to avoid off-chip filters and their drivers in order to obtain a lower cost, power and pin count. The alternative of integrating the off-chip filters was not viable because of the lack of technology for on-chip high IF filters with repeatable transfer functions.

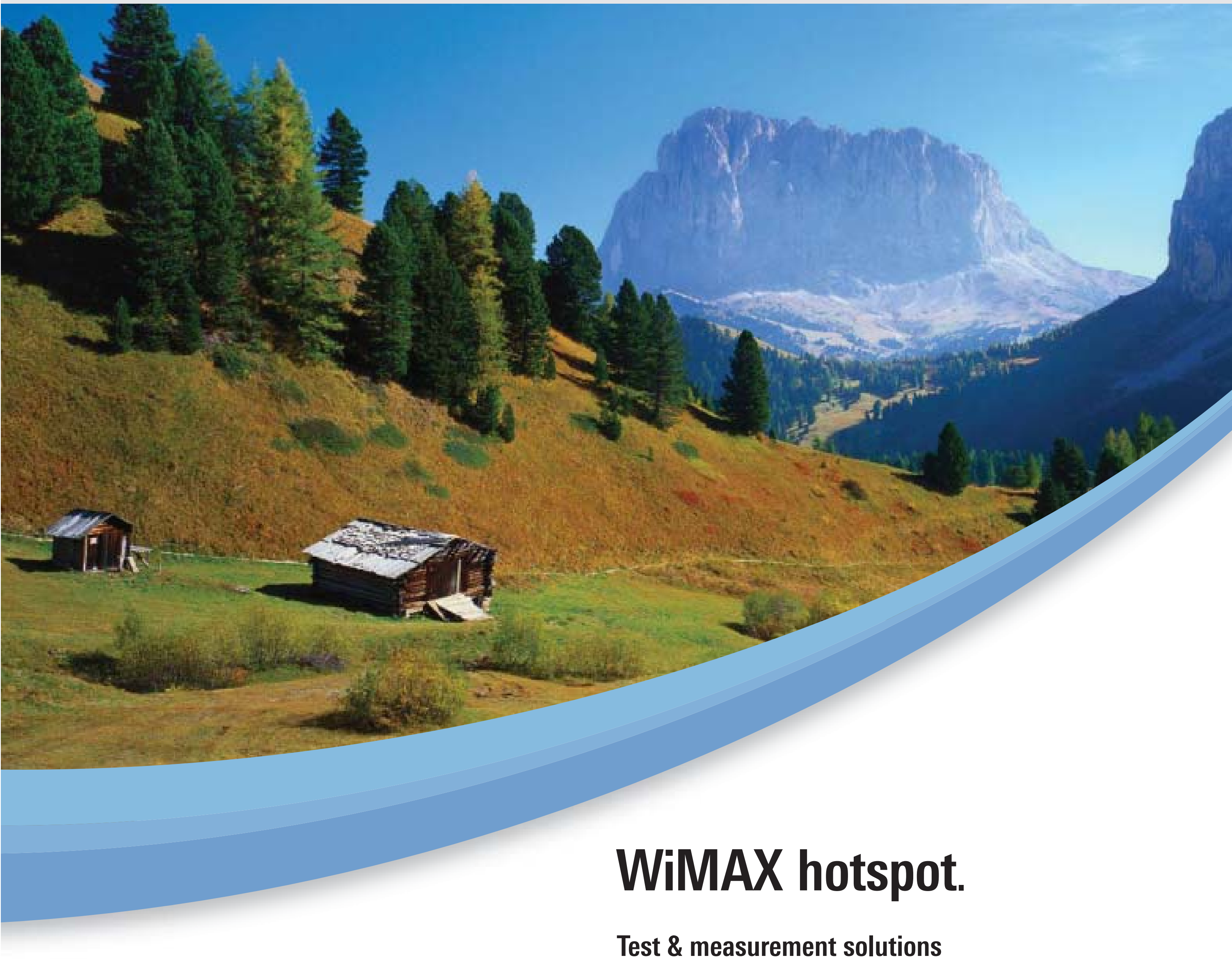
Advantages of the Superheterodyne

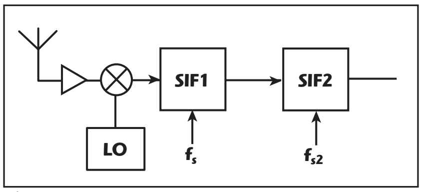
Superheterodyne is usually the preferred method of designing wireless communications receivers. This architecture utilizes a dual down-conversion process having two separate IF frequencies. These two separate IF mixer stages can simplify the filters at each stage. They also eliminate the conflict that exists between sensitivity and selectivity when using single IF or ZIF architectures. Having a high frequency first IF and hence an image frequency that is far away from the RF frequency means that a simple RF filter network is sufficient to reject interfering signals at the image frequency. Similarly, the first IF stage filter can also be a simple filter network as its function is only to eliminate image frequencies from the second LO. The second IF filter is required for channel selection only and operates at a low frequency. This means that high selectivity can be achieved with low Q components.

This architecture achieves more than simplifying the filters; it also improves the total system noise figure because simpler filter stages have less insertion loss. This improvement in system noise figure can typically be several dB. In a multi-mode receiver, the improvement in noise budget allows for more creativity on the part of the designer. For example, this improvement could be transferred to the Tx/Rx switching network, meaning that cost improvements could then be realized in both the filters and the switch. In a superheterodyne, this improvement in noise figure and sensitivity is achieved without any compromise in selectivity.

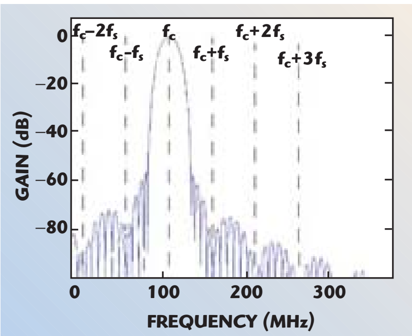
Disadvantages of the Superheterodyne

Careful selection on the frequency plan ensures that the system will be optimized for selectivity and sensitivity. However, the available off-chip filter technologies usually dictate the



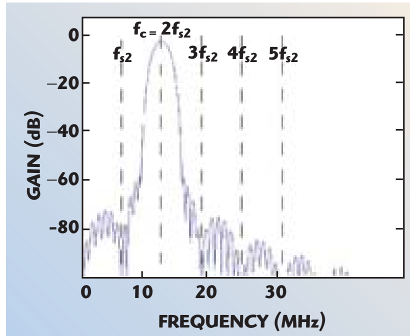


▲ *Fig. 11 SIF-based superheterodyne architecture.*



▲ *Fig. 12 Bandpass SIF transfer function for a superheterodyne first IF stage.*

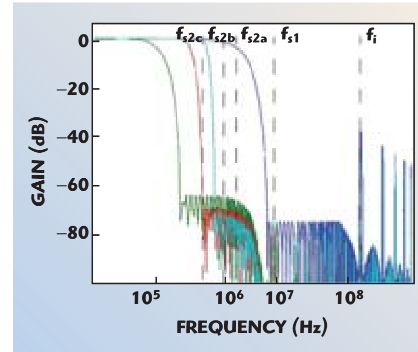
frequency plan. Designers are constantly scouring the market for low cost and easily available filters, and build their systems around the frequencies of these components. This off-chip conflict is more restrictive when one tries to integrate the filters on-chip.



▲ *Fig. 13 Bandpass SIF transfer function for a superheterodyne second IF stage.*

On-chip SIF-based Superheterodyne Architecture

An on-chip SIF-based superheterodyne architecture is shown in ***Figure 11***. In this architecture, the SIF1 and SIF2 stages each provide a filter and sampled output. The sampling at the output of each stage uses aliasing to down-convert a desired frequency band centered at f_c. ***Figure 12*** shows the SIF first stage transfer function, where f_c =113 MHz and f_s = 50 MHz. This causes the frequency components of the input signal near 113 MHz to alias to 13 MHz at the output of the first sampling IF filter stage. The frequencies near f_c are passed by the filter. Frequencies near $f_c \pm kf_s$, where $k \neq 0$, which would also alias to 13 MHz at the filter output, are highly attenuated. This way, only the desired alias of the input signal is selected for down-conversion by sampling (aliasing).



▲ *Fig. 14 Examples of programmable FIR filters.*

Figure 13 shows the second stage transfer function, where f_c = 13 MHz and f_{s2} = 6.5 MHz. As a result, the 13 MHz output from the first stage filter is further filtered and down-converted to a baseband output having in-phase and quadrature components for A/D conversion and further digital processing. In summary, the on-chip SIF-based superheterodyne architecture has all of the advantages previously enjoyed by the traditional superheterodyne, while permitting on-chip filter integration.

PROGRAMMABLE FILTER TRANSFER FUNCTIONS

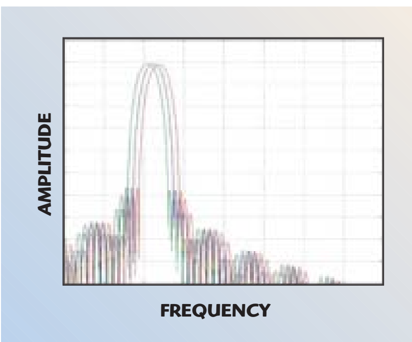
When the specifications call for different data bandwidths or data rates, the programmable FIR filter can create several predetermined filter transfer functions on the fly (see ***Figure 14***). The blue frequency response represents the first enhanced anti-alias filter stage. This was the red line in the figure shown in the enhanced anti-alias filter by extended integration intervals and resistor-weighted tap coefficients section. Any one of the other three frequency responses could be programmed into a second enhanced anti-alias filter stage.

The examples shown are used to indicate the variety of filter options that are available in a single stage.

TABLE I

PARAMETERS FOR THE PROGRAMMABLE FILTER IN FIGURE 14

Parameter	*Green*	*Red*	*Cyan*
3 dB bandwidth (kHz)	170	600	1200
Stopband attenuation (dB)	65	70	72
Transition bandwidth (kHz)	330	400	600
Output sampling frequency (MHz)	1	2	3



▲ *Fig. 15 Programmable FIR transfer functions.*

Each filter transfer function can have its own bandwidth, stopband attenuation and shape factor as part of each enhanced anti-alias filter stage. As the filtering requirements of the filter become more demanding on all three parameters, additional enhanced anti-alias filter stages need to be added. The three additional transfer function examples were chosen to be suitable for three different sampling frequencies at the output of the programmable FIR stage. To recap, the blue line is a single-stage enhanced anti-alias filter and the remaining three lines are made from a two-stage enhanced anti-alias filter. The parameters for the programmable filter are shown in ***Table 1***.

Figure 15 shows three transfer functions of a programmable enhanced anti-alias filter. In this case, the anti-alias filter is a bandpass filter selecting the desired alias. This example highlights how each stage of an anti-alias filter can have programmable transfer functions.

CONCLUSION

This article has shown that a sampling IF filter is a viable alternative to both SC and g_m–C filters, because of its advantages in power consumption and process/temperature tolerance. When compared to an SC filter, this technique exhibits lower noise, while avoiding the aliasing problems and is more stable over process and temperature than a g_m-C filter. Eliminating an AGC stage and reducing the bandwidth and resolution requirements of the A/D and digital processing realize further power savings. Superheterodyne wireless architectures are once again a viable option for the system designer due to the unique enabling capabilities of the sampling IF filter. ■

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A New Class of Asymmetrical Directional Couplers for Power/Antenna Control Applications

A new class of asymmetrical directional couplers, convenient for power and antenna control applications, is proposed. The couplers can be fully or partially integrated into a printed circuit board (PCB). The couplers operate over –20 to –40 dB coupling coefficients and are always theoretically compensated. The compensation is achieved by the proper displacement of a tuning ground plane with respect to the edge of the PCB. Experimental results are presented for a suspended multilayer microstrip-to-stripline –30 dB coupler, a coaxial line-to-microstrip –20 dB coupler and a DC block integrated together with a –30 dB coupler in the suspended multilayer configuration.

Integration of RF power, RF small signal, digital, DC and supervision circuits is still one of the main goals of modern radar, satellite and wireless communications technology. Today's solutions of close-to-antenna devices use a stack approach for integration of RF filters with power splitters/combiners and multilayer printed circuit boards (PCB) containing the rest of the circuitry.

One device that is difficult to integrate is a directional coupler mounted close to an antenna for the purpose of monitoring the transmitted and reflected powers. The requirements for such a coupler are of high importance: very low insertion loss, very good matching of the main line (carrying the power), sufficient power handling and a directivity better than 20 to 26 dB in both directions. Because the two coupled ports of the coupler are used simultaneously, there is no possibility to "improve" the directivity by tuning one of the coupled line ports. A modular approach is often implemented. This dictates that a high quality coupler would be placed separately, outside the integrated units, and connected utilizing cables or special transitions. This, of course, adds to the manufacturing cost as well as increases the number of steps in the process.

There are few known solutions allowing the partial integration of the coupler into the

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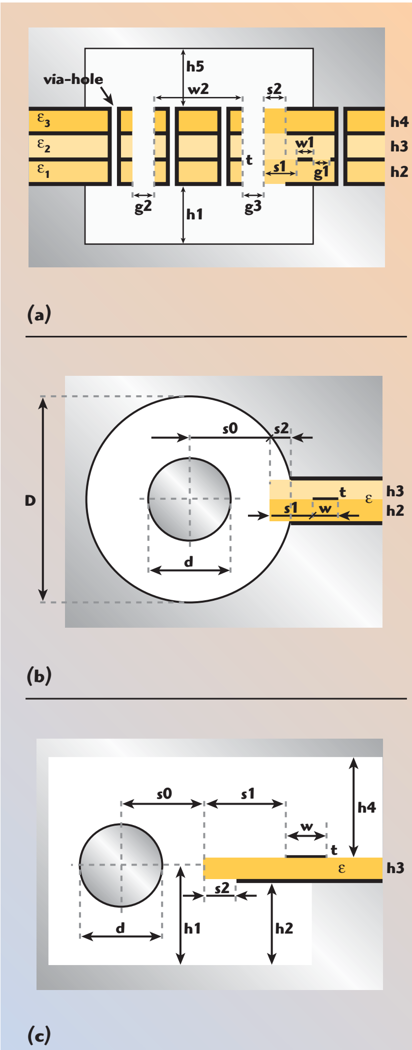
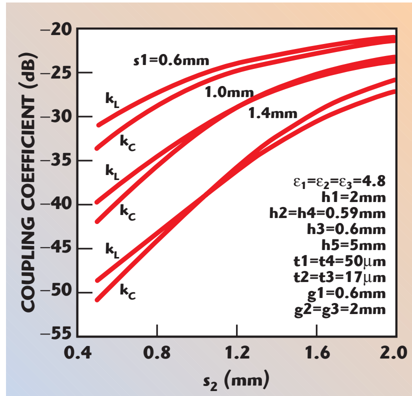


▲ *Fig. 1 Three cross-sections of novel asymmetrical directional coupler structures.*

PCB. Four of them are patented[1-4] and another one is offered by Teppati and Ferrero.[5] Most of them are of the coaxial-to-stripline or coaxial-to-microstrip configurations, where the printed line is placed on top of the coaxial line. Persson[4] uses a multilayer PCB configuration where both the main and coupled line are milled and suspended. The techniques used to compensate the coupler in order to achieve perfect matching and high directivity over a wide frequency range are not discussed in these publications. It is known[6,7] that, assuming the validity of quasi-static approximation, asymmetrical couplers in inhomogeneous dielectric media can be compensated if the inductive and capacitive coupling coefficients are equalized and the coupled lines are terminated with the proper impedances:



▲ *Fig. 2 Inductive and capacitive coupling coefficients versus s_2 for the fully embedded structure.*

$$k_L = k_C \tag{1}$$

and

$$Z_{Ti} = Z_i \quad \text{for } i = 1,2 \tag{2}$$

where

$$k_L = \frac{L_m}{\sqrt{L_1 L_2}} =$$

inductive coupling coefficient

$$k_C = \frac{C_m}{\sqrt{C_1 C_2}} =$$

capacitive coupling coefficient

Z_{Ti}, $i = 1,2 =$ characteristic impedances of terminating lines

$$Z_i = \sqrt{\frac{L_i}{C_i}},\ i = 1,2 =$$

characteristic impedance of line i in the presence of line j ($j = 1,2,\ j \neq i$)

L_i, C_i, $i = 1,2 =$ self-inductance and self-capacitance per unit length of line i in the presence of line j ($j = 1,2,\ j \neq i$), respectively

$L_m, C_m =$ mutual inductance and mutual capacitance per unit length, respectively

The scope of this work was to identify a directional coupler structure having high power handling in the main line, being fully or partially embedded into the PCB, achieving a coupling coefficient in the range –20 to –40 dB, and being always theoretically compensated. Three examples of possible solutions are presented in ***Figure 1***.[8] The first structure shown is fully embedded into the PCB. The main line is suspended over the chassis and is composed of four printed strips connected by via-holes. The coupled line is hidden inside the PCB and can be treated as a stripline. The dielectric material is milled out in the vicinity of the main line and protrudes past the edges of the tuning ground planes (at a distance s_2). The other structures shown are composed of a coaxial line (carrying the power) and a stripline or microstrip, as the coupled line, placed beside the coaxial line. In these structures, only the coupled line is embedded into the PCB. The dielectric material protrudes past the edge of the tuning ground plane(s), towards the center conductor of the coaxial line at a distance s_2.

There are many possible modifications of the coupler topology using the proposed concept. Other cross-section shapes and locations of the main line, as well as modifications of the coupled line are possible. Generally, the structure should contain the low loss main line (embedded into the PCB, or not), the coupled line (embedded into the PCB), tuning ground plane(s) and the dielectric material of the PCB protruding past the edge of the tuning plane(s), towards the main line. These last two features are necessary to assure theoretical compensation of the coupler.

Areas of achievable coupling coefficients under the compensation conditions, dependence of the coupler directivity on the deviation from these conditions and structural parameters for realizing a –30 dB coupler in a variety of dielectric substrate permittivity are presented in the analysis section. Experimental results are presented for two –30 dB and –20 dB directional couplers and for a DC block integrated with a –30 dB coupler.

ANALYSIS

The analysis of the proposed structures was performed using a static 2-D solver.[9] The transversal geometrical dimensions of the 50 Ω matched directional couplers were calculated as a function of the dielectric substrate protrusion distance s_2, and changes to the inductive k_L and capacitive k_C coupling coefficients were observed. Most of the calculations were carried out for the FR-4 Matsushita, halogen-free, substrate mate-

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rial. The results of the calculations for the fully embedded structure are presented in ***Figure 2***. It is clear that this structure can be compensated in the range of coupling coefficients from –40 to –20 dB by changing the distance s_1 between the main line and the coupled line along with adjusting distance s_2. The widths of the main (w2) and the coupled (w1) lines vary from 5.5 to 5.4 mm, and from 1.5 to 0.5 mm, respectively.

The results shown in ***Figure 3*** were obtained for the structure using a coaxial main line and a stripline coupled line with a modified outer ground enclosure. It is shown that even "pure air" and "pure stripline" coupled transmission lines can be compensated in spite of large difference between the values of effective dielectric permittivity of the two orthogonal modes propagated in the structure. This feature distinguishes the asymmetrical coupled lines from the symmetrical lines. The latter ones are compensated if values of these permittivities are equal. Results of computations for the third structure, using a coaxial main line and a microstrip coupled line, are shown in ***Figure 4***. The coupling level can be adjusted in this structure not only by changing the distance between the coupled lines ($s_0 + s_1$) but also by changing the suspension level h_2 of the PCB — the higher this suspension, the weaker the coupling. The effect of deviations from the compensation conditions on the directivity of the coupler is also shown. Tighter tolerances on the distance s_2 are needed for higher suspended PCBs.

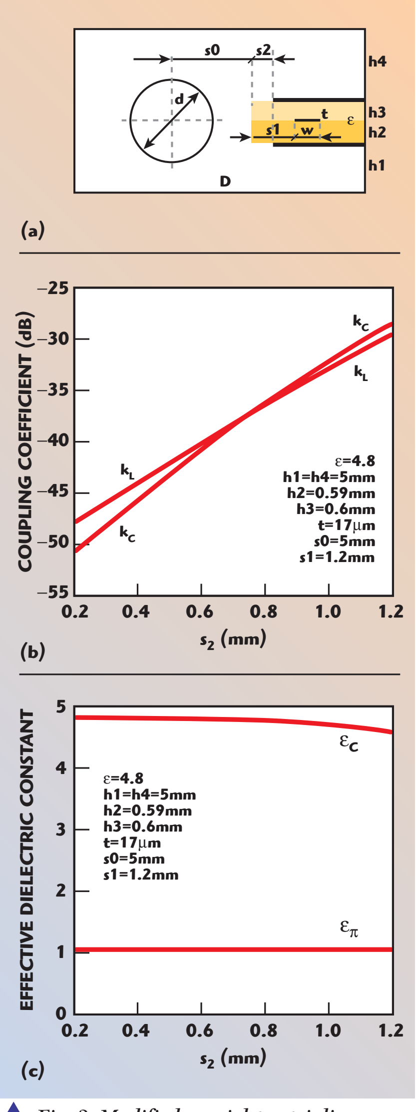

Fig. 3 Modified coaxial-to-stripline structure; (a) geometry, (b) inductive and capacitive coupling coefficients and (c) effective dielectric constant for two orthogonal modes.


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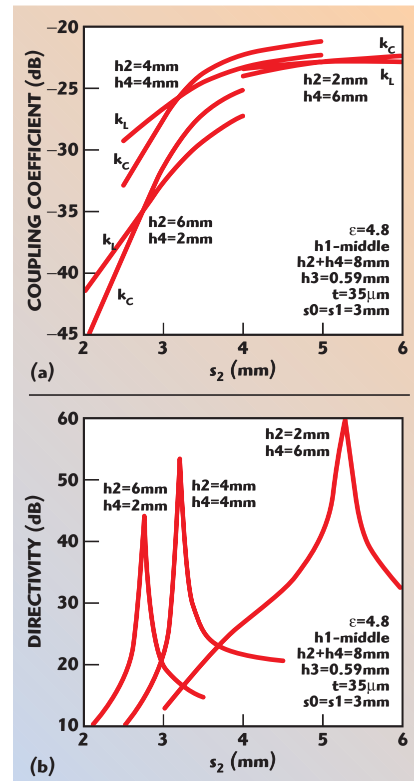
The structural dimensions of a –30 dB, 50 Ω matched, compensated coupler, with varying dielectric permittivity of the microstrip substrate, are given in ***Figure 5***. These curves are very convenient for a practical realization using any chosen PCB dielectric material. The dielectric material protrusion distance s_2 varies more dramatically in the region of low dielectric constant.

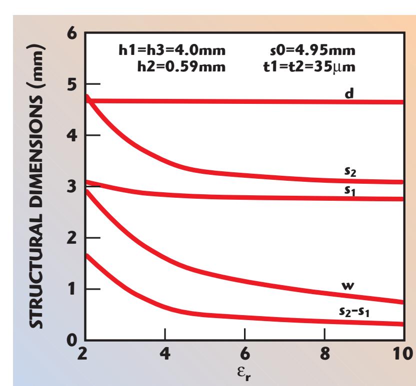
It has been shown that coupling coefficients in the range of –20 to –40 dB are easily achievable with the proposed structures. These coupling levels can be realized with convenient structural dimensions, meaning that the proposed structures are not sensitive to dimensional tolerances.

EXPERIMENTAL RESULTS

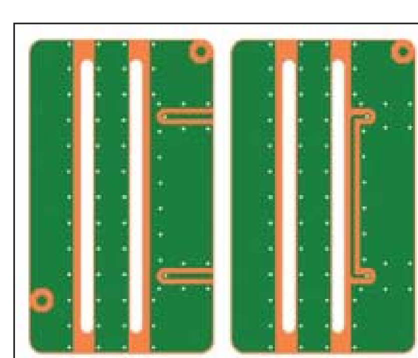
A –30 dB directional coupler was designed and manufactured in the fully embedded configuration. Layouts of the first and third PCB layers



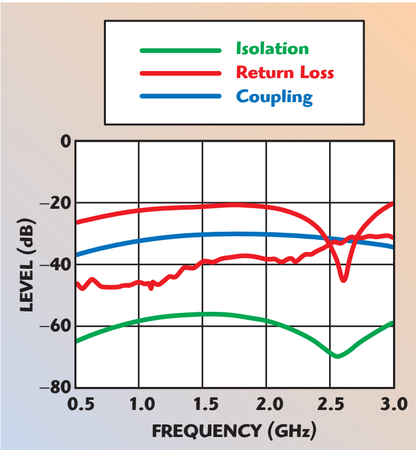
▲ *Fig. 4 Coaxial-to-microstrip structure; (a) coupling coefficients and (b) corresponding directivity of the coupler.*



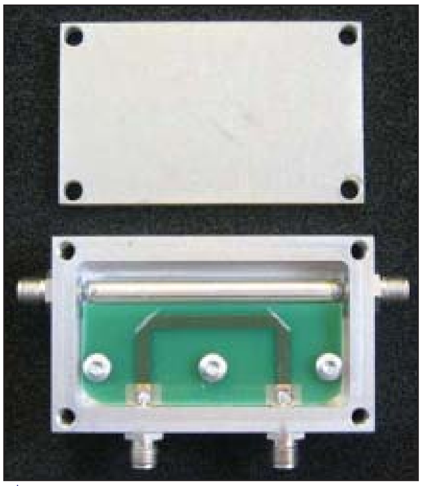
▲ *Fig. 5 Dimensions of the 50 Ω matched coaxial-to-microstrip coupler versus dielectric constant of the PCB.*



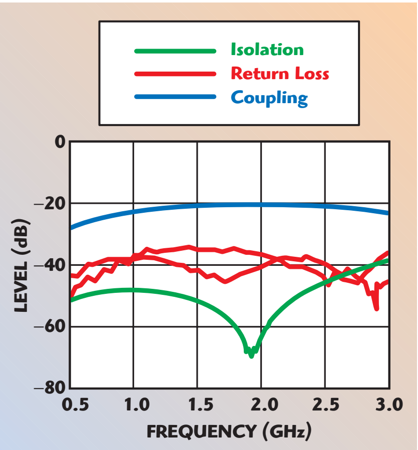
▲ *Fig. 6 Layout of the first and third layers of the –30 dB fully embedded directional coupler.*



▲ *Fig. 7 Measured responses for the -30 dB directional coupler.*



▲ *Fig. 8 The -20 dB coaxial-to-microstrip directional coupler.*

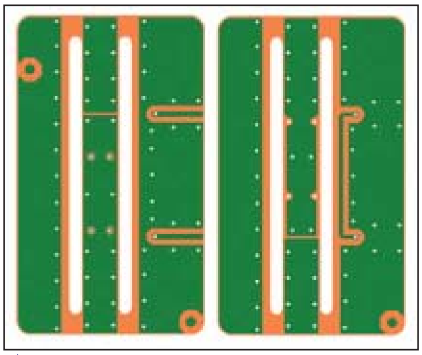


▲ *Fig. 9 Measured response of the -20 dB directional coupler.*

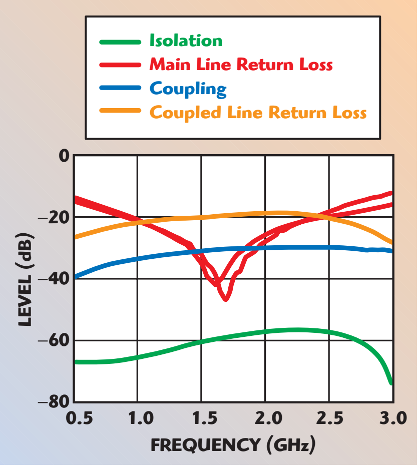
are shown in ***Figure 6***. The copper pattern on the second and the fourth layer is similar to the first one, except for the microstrip lines connecting the coupled line to the edge of the PCB. The measured results are presented in ***Figure 7***. The coupling level at the mid-band frequency is exactly as predicted in the design. The main suspended line is perfectly matched — the return loss is better than 30 dB. The directivity exceeds 25 dB in the entire frequency band presented, and follows the return loss of the rather poorly matched coupled line, which is affected by the stripline to microstrip and microstrip to SMA connector transitions.

A –20 dB directional coupler was designed and manufactured utilizing the coaxial to microstrip configuration. A photograph of this coupler and the measurement results are shown in ***Figures 8*** and ***9***, respectively. Small manual adjustments have been done to obtain optimal chamfering of the 90° microstrip bends. This has led to improvements in the microstrip line matching and also in isolation. The directivity of the coupler is better than 20 dB for frequencies up to 2.7 GHz and exceeds 30 dB in the 1.4 to 2.3 GHz frequency range. The visible directivity degradation above 2.7 GHz can be affected by the dispersion of the microstrip line, which is printed on a 1.7 mm thick substrate.

The fully embedded configuration is very convenient for further integration of other devices/circuits, commonly used close to the antenna port, such as a DC block or a lightning protection circuit. Using this configuration, a DC block was designed together with the power and antenna control –30 dB directional coupler. Layouts of the first and the third PCB layer are shown in ***Figure 10***. The slots and isolated via-holes in the main line are easily visible. The measurement results are shown in ***Figure 11***. The DC block is very well matched and has an insertion loss below 0.1 dB in the frequency band from 0.6 to 2.5 GHz. The center frequency for the DC block is shifted down 0.2 GHz compared to expectations because of the static design. The measured coupling value is less than the designed one by 0.3 dB. The directivity of the coupler is better than 25 dB throughout the whole frequency band presented, despite the poorly matched coupled line.



▲ *Fig. 10 Layout of the first and third layers of a DC block integrated with a fully embedded -30 dB directional coupler.*



▲ *Fig. 11 Measured response of the -30 dB directional coupler with an integrated DC block.*

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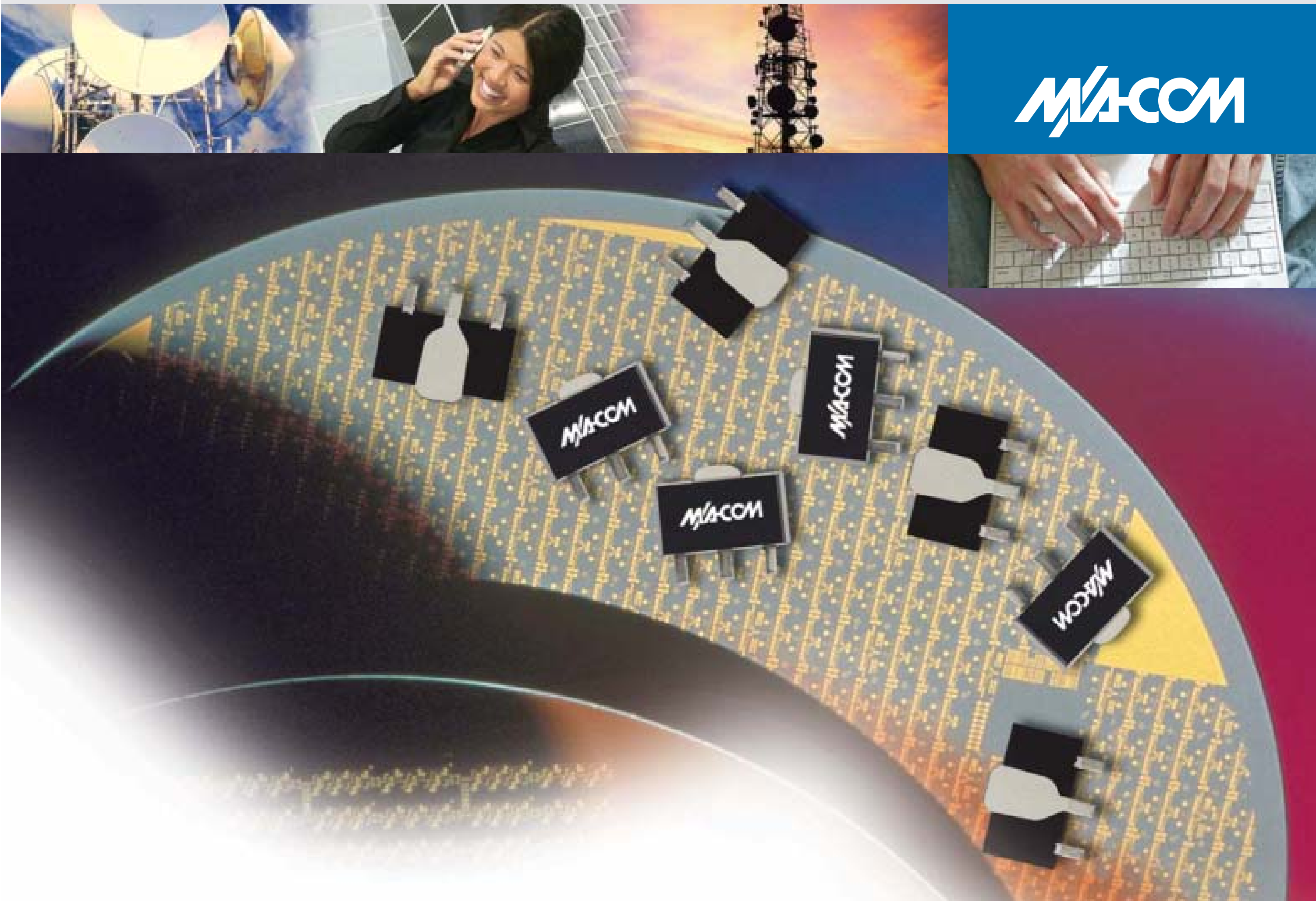

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MAAMSS0056	SOIC-8EP	2100	24	27	40 1dBm/Tone – 1 MHz	190
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These measurement results are remarkably good, which means that the quasi-static design can be used as a reliable starting point for a full-wave design. Some corrections should be considered, especially for higher frequency designs, due to discontinuities.

CONCLUSION

The various structures of directional couplers investigated in this article seem to be very promising for the sampling of forward and reflected RF power simultaneously. They allow the integration of RF power and small-signal parts of a device into a single unit. The required range of coupling coefficients from –20 to –40 dB is easily realizable. Very low insertion loss of the main line, wide bandwidth and good directivity are easily achieved. The insertion loss of the main line is additionally diminished due to the shorter length of the coupler compared to air solutions. This length is reduced approximately by a $(1 + \sqrt{\varepsilon_{effc}})$ factor, where ε_{effc} is the effective dielectric constant for the mode associated with the coupled line, embedded in the dielectric material of the PCB, or printed on a dielectric board. The quasi-static design, presented in the article, makes an effective design possible based on the equalization of inductive and capacitive coupling coefficients. ■

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VCO Part Number	Frequency (MHz)	Vtune (Vdc)	Kvco (MHz/V)	$\varnothing_N$@10KHz (dBc/Hz)	Output Power (dBm)	2nd Harmonic (dBc)	Pulling (MHz)	Pushing (MHz/V)	Vcc (Vdc)	Icc (mA)	Operating Temp (°C)
V150ME03	100 to 200	0 to 12.5	10	-111	7 ± 5	-10	<1	<1	12.0	26	-40 to 85
V220ME01	200 to 239	0.5 to 4.5	14	-120	7.5 ± 2.5	-22	<0.5	<0.5	5.0	16	-40 to 85
CLV1277A	1213 to 1341	0.5 to 4.5	38	-108	2.5 ± 2.5	-15	<1	<1	5.0	22	-40 to 85
CRO2155A*	***1960 to 2350***	***1 to 14***	***40***	***-106***	***7 ± 2***	***-10***	***<2***	***<0.5***	***6.0***	***27***	***0 to 85***
CRO2780A*	***2650 to 2910***	***0.5 to 15***	***20***	***-111***	***3 ± 3***	***-10***	***<0.5***	***<0.5***	***10.0***	***34***	***-40 to 85***
CRO2880A	2760 to 3000	0 to 15	18	-110	12.5 ± 2.5	-20	<1	<1	10.0	29	-40 to 85
V950ME07	3900 to 6000	0 to 20	126	-80	4.5 ± 4.5	-14	<36	<14	5.0	21	-40 to 85
CRO4500A	4499 to 4501	0.5 to 4.5	12	-104	2 ± 2	-15	<1	<2	5.0	20	-20 to 70

PLL Part Number	Frequency (MHz)	Step Size (kHz)	Output Power (dBm)	$\varnothing_N$@ 10KHz (dBc/Hz)	$\varnothing_N$@ 100KHz (dBc/Hz)	2nd Harmonic (dBc)	Ref Sup (dBc)	Lock Time (msec)	Vcc (Vdc)	Icc (mA)	Operating Temp (°C)
PCA1445C	1444 to 1446	1000	5 ± 2	-120	-140	-20	-59	3	5.0	40	-40 to 85
PCA1550A	1500 to 1600	1000	1.5 ± 2.5	-103	-124	-15	-70	3	5.0	40	-40 to 85
PSA2000C*	***1970 to 2030***	***100***	***2 ± 2.5***	***-107***	***-128***	***-15***	***-70***	***2.5***	***5.0***	***30***	***-40 to 85***
PCA3040C*	***3040 to 3040***	***1000***	***3 ± 3***	***-112***	***-132***	***-8***	***-60***	***1***	***5.0***	***35***	***-40 to 85***
PSA3330C	3305 to 3335	125	0 ± 3	-106	-130	-12	-70	1	5.0	35	-40 to 85
PSA3500A	3400 to 3600	1000	0 ± 3	-85	-109	-15	-70	2	5.0	40	-40 to 85
PSA3707C	3675 to 3738	250	0 ± 3	-105	-128	-15	-70	2	5.0	40	-40 to 85
PSA4202C*	***4144 to 4260***	***250***	***0 ± 3***	***-96***	***-119***	***-12***	***-70***	***1***	***5.0***	***40***	***-40 to 85***

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EXPERIMENTAL INVESTIGATION OF A POWER DIVIDER BASED ON MICROSTRIP AND METAMATERIALS WITH L-C LUMPED-ELEMENTS

One-dimensional metamaterials, using L-C lumped-elements, have a broad left-handed passband, with anti-parallel phase and group velocities. Power dividers, with symmetric and asymmetric structures, are proposed. They are composed of conventional microstrip lines and composite right-/left-handed transmission lines with L-C lumped-elements. The asymmetric power dividers can have different frequency bands for each output port. The simulated results, obtained from the circuit models, agree well with the experiments.

Metamaterials,[1,2] with simultaneous negative permittivity and permeability, are promising materials for new types of microwave components. In metamaterials, the waves are propagating with anti-parallel phase and group velocities, as demonstrated by backward waves.[3] Recently, an extended transmission-line approach to metamaterials, low loss and broadband structures was proposed[4–6] and studied by different groups.[7–10] A novel coupled-line directional coupler, using left-handed (LH) transmission lines, was also proposed and studied.[11–14] A symmetric LH/LH structure, used in a backward coupler with an arbitrary coupling level and broad bandwidth, has been fully explained.[13]

In this article, a symmetric right-/left-/right-handed (RH/LH/RH) power divider, a symmetric LH/RH/LH power divider and an asymmetric LH/RH/LH power divider are proposed, which consist of conventional microstrips and composite right-/left-handed (CRLH) transmission lines with L-C lumped-elements. The power dividers have multiple-port outputs, broad bandwidths and arbitrary coupling levels. If the loaded L-C elements in

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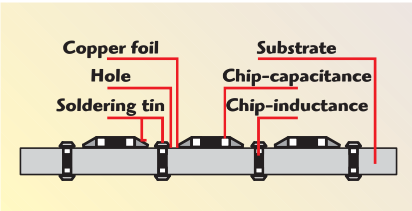
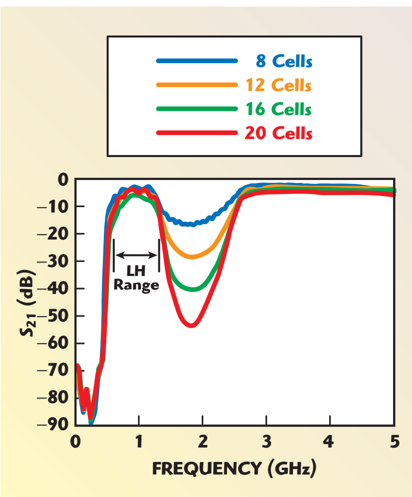


▲ *Fig. 1 Structure of the CLRH transmission line.*

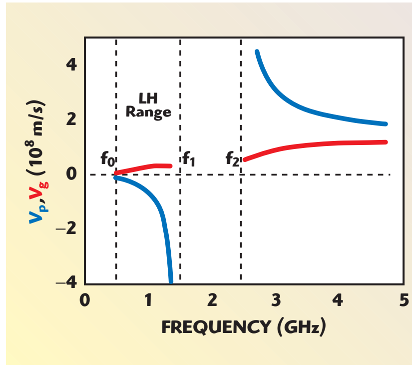


▲ *Fig. 2 S_{21} of the CLRH line for different numbers of cells.*

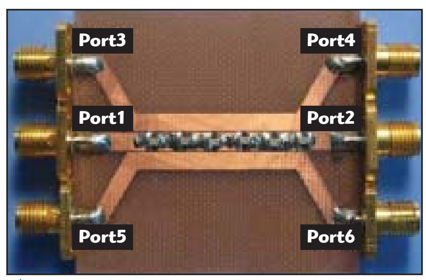
the two CRLH lines have different values, the LH/RH/LH power divider can have different frequency bands for each output port. At the same time, the lumped-elements offer some significant advantages: the circuit is more compact in size, the material parameters can be tuned and the method of fabrication is easy. This article presents first the characteristics of one-dimension metamaterials. A symmetric RH/LH/RH (or LH/RH/LH) power divider and an asymmetric LH/RH/LH power divider are then described and their simulated performance, obtained from the circuit models, are compared to experimental results.

THE CHARACTERISTICS OF ONE-DIMENSION METAMATERIALS

L-C lumped-elements are loaded on a conventional microstrip to fabricate the CRLH transmission line. ***Figure 1*** shows the structure of the CRLH transmission line. The LH line is fabricated on a 1.6 mm thick FR-4 substrate with a dielectric constant ε = 4.75, and the 50 Ω microstrip transmission-line segments are 2.945 mm wide. All the measure-



▲ *Fig. 3 Phase and group velocities in the passband.*

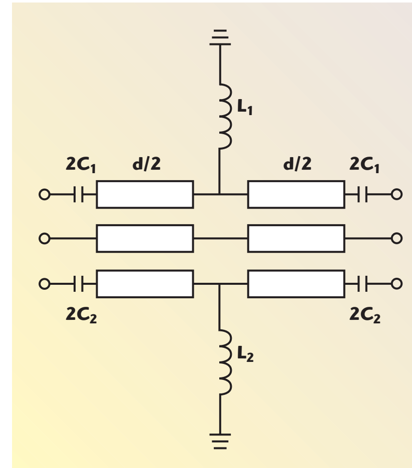


▲ *Fig. 4 The six-cell trisecting power divider with a symmetric structure (RH, LH, RH).*

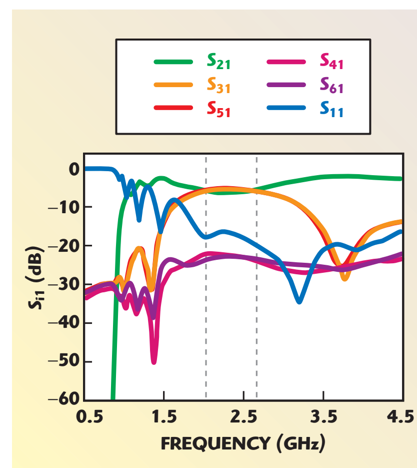
ments were carried out with an Agilent 8722ES vector network analyzer.

Figure 2 shows the S_{21} parameter of the CRLH line with a different number of cells. Each cell consists of a microstrip loaded with lumped-elements, C = 5.1 pF, L = 4.7 nH; the length of each cell is d = 7 mm. As the number of cells is increased, the attenuation in the bandgap increases. In the range considered, the attenuation in the bandgap is approximately 2 dB/cell. The group velocity and phase velocity can be obtained from the S-parameters and their phase. ***Figure 3*** shows the group velocity and phase velocity in the passband of a line with 20 cells.

By comparing the two figures, it can be observed that the left-handed range with anti-parallel phase and group velocities is beginning at the cut-off frequency ($f_0 = 1/4\pi\sqrt{CL}$) of the high pass structure to the bandgap; the right-handed range with parallel phase and group velocities is above the bandgap. The bandgap can also be found (from $f_1 = 1/2\pi\sqrt{CL_0 d}$, to $f_2 = 1/2\pi\sqrt{LC_0 d}$, where C_0 and L_0 are the distributed parameters of the microstrip) corresponding to a zero averaged refractive index between the left-handed range and the right-handed range, as first indicated by



▲ *Fig. 5 Unit cell circuit model.*



▲ *Fig. 6 Measured results for the six-cell trisecting power divider with a symmetric structure (RH, LH, RH).*

Reference 15. Such zero–$\bar{n}$ bandgap differs fundamentally from the usual bandgap induced by the Bragg scattering. That is, it is independent of scaling, and is insensitive to the disorder, incident angle and polarization.[15,16]

THE SYMMETRIC AND ASYMMETRIC POWER DIVIDERS

The Symmetric Trisecting Power Dividers

The trisecting power divider with a symmetric structure (RH/LH/RH) is shown in ***Figure 4***. It consists of two conventional microstrips (ports 3 to 4 and ports 5 to 6) and one CRLH transmission line (ports 1 to 2) between the microstrips. When a signal is applied to port 1, the power is coupled to ports 3 and 5 equally, and ports 4 and 6 are isolated. The unit cell circuit model used for simulation is shown in ***Figure 5***. The measured

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MFC4151-12	410 to 510	0.5 to 15	-112	+12
MFC6170-5	610 to 790	0.5 to 5	-113	+5
MFC7995-5	790 to 950	0.5 to 15	-114	+5
MFC8192-5	810 to 920	0.5 to 5	-106	+5
MFC81100-5	810 to 1000	0.5 to 10	-105	+5
MFC102110-5	1020 to 1100	0.5 to 5	-106	+5
MFC-S-1000	1000 to 2100	1 to 18	-99	+12
MFC138165-5	1380 to 1650	0.5 to 24	-102	+5
MFC170195-5	1700 to 1950	0.5 to 10	-104	+5
DCRO Series				
DCRO127175-5	1270 to 1750	0.5 to 18	-107	+5
DCRO128177-12	1280 to 1775	0.5 to 24	-112	+12
DCRO175260-5	1750 to 2600	0.5 to 15	-96	+5
DCRO204235-8	2040 to 2350	0.5 to 24	-109	+8
DCRO219250-8	2190 to 2500	0.5 to 24	-106	+8
DCRO243298-5	2430 to 2980	0.5 to 15	-101	+5
DCRO250300-10	2500 to 3000	0.5 to 24	-107	+10
DCRO270400-8	2700 to 4000	0.5 to 18	-93	+8
DCRO285345-5	2850 to 3450	0.5 to 24	-98	+5

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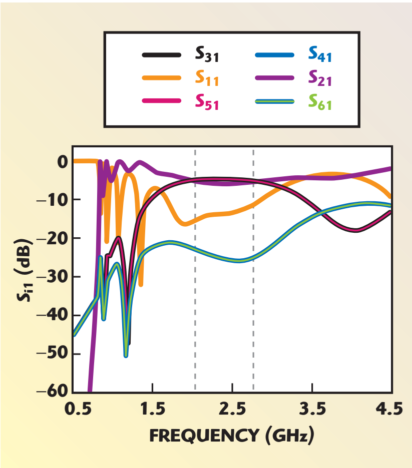
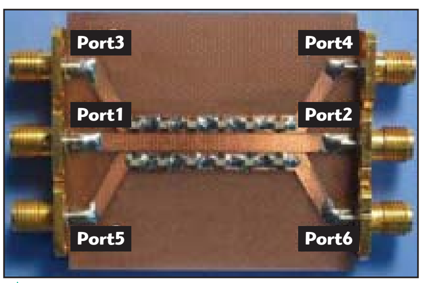
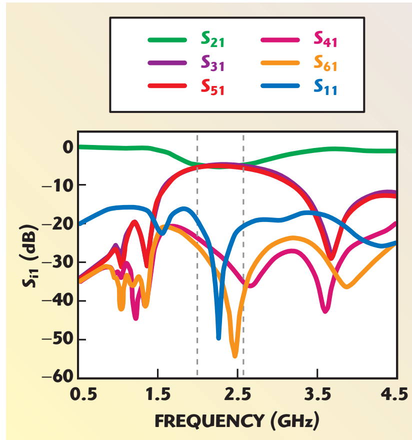


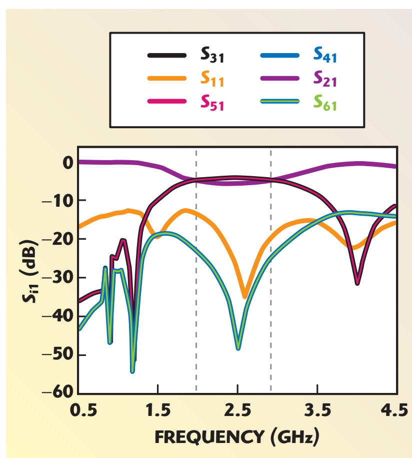
▲ *Fig. 7 ADS simulated results for the six-cell trisecting power divider with a symmetric structure (RH, LH, RH).*



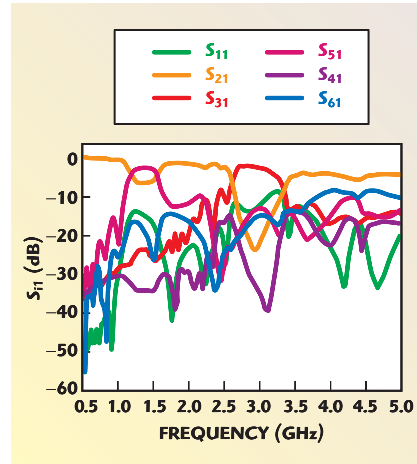
▲ *Fig. 8 The six-cell trisecting power divider with a symmetric structure (LH, RH, LH).*



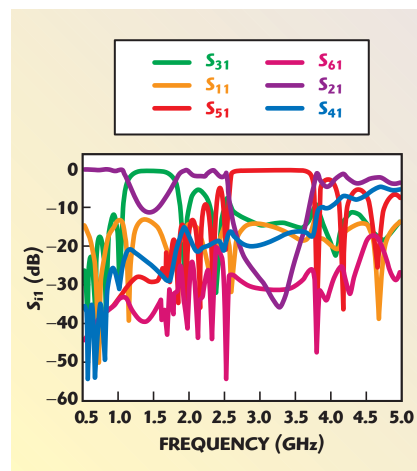
▲ *Fig. 9 Measured results for the six-cell trisecting power divider with a symmetric structure (LH, RH, LH).*



▲ *Fig. 10 Simulated results for the six-cell trisecting power divider with a symmetric structure (LH, RH, LH).*



▲ *Fig. 11 Measured S-parameters for the asymmetrical power divider.*



▲ *Fig. 12 Simulated S-parameters for the asymmetrical power divider.*

S-parameter results, shown in ***Figure 6***, are in good agreement with the simulated results shown in ***Figure 7***. From 2.1 to 2.7 GHz, the coupled power (ports 3 and 5) and the through power (port 2) are 4.77±1 dB, with a fractional bandwidth of approximately 25 percent.

The symmetric structure power divider can also be designed with a microstrip between two CRLH lines (LH/RH/LH). ***Figure 8*** shows the LH/RH/LH symmetric structure power divider. The measured S-parameter results shown in ***Figure 9*** are in good agreement with the simulated results shown in ***Figure 10***. These results are similar to the ones for the RH/LH/RH power divider.

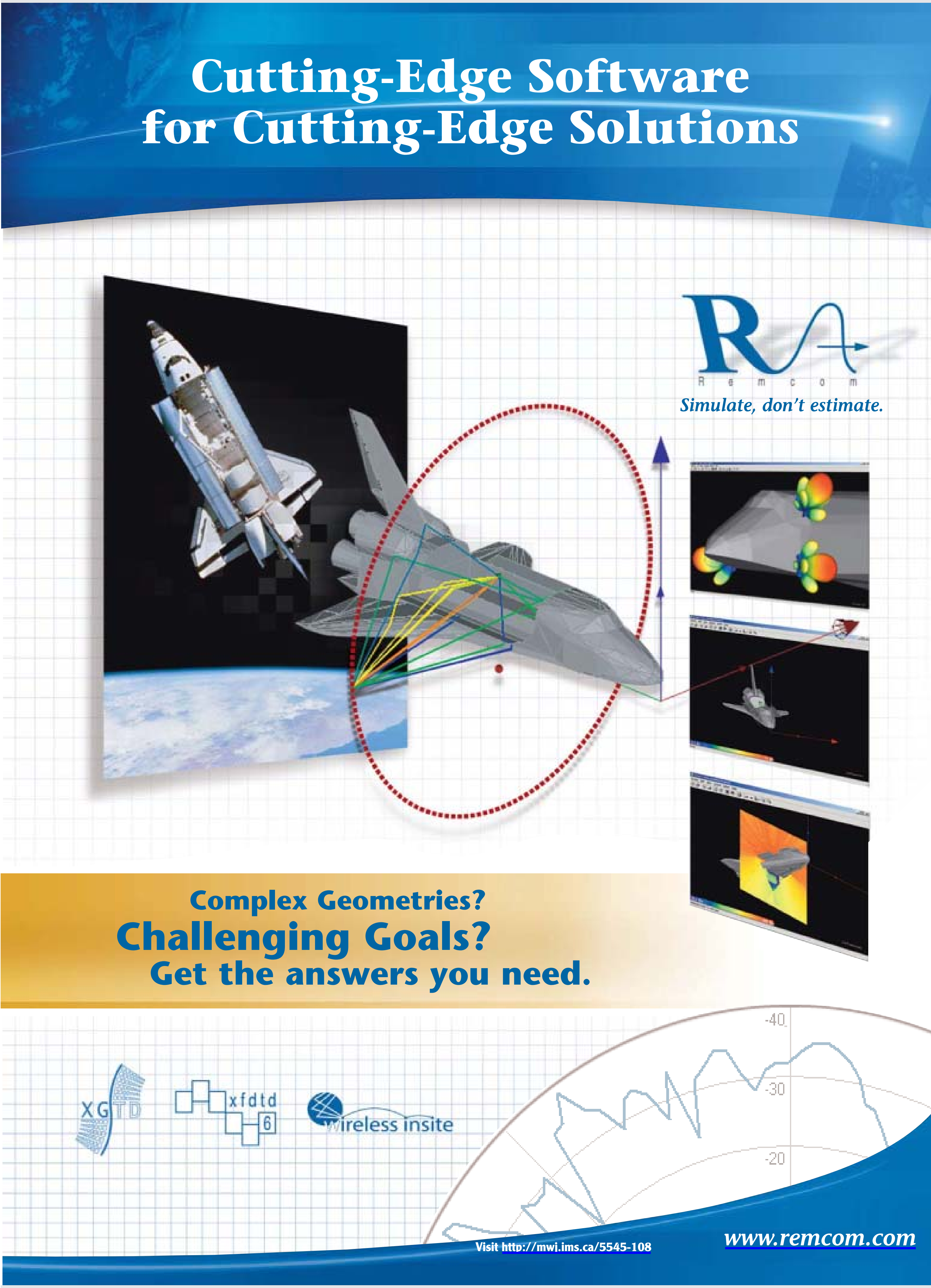
The Asymmetric Power Dividers

The asymmetric power divider has the same physical structure as the symmetric LH/RH/LH power divider, but the loaded L-C elements in the two LH transmission lines have different parameters. The asymmetric power dividers can have different frequency bands for each output port.

An asymmetric power divider was simulated and fabricated with loaded elements C1 = 5.1 pF, L1 = 8.2 nH in one CRLH transmission line (ports 3 to 4) and C2 = 1.0 pF, L2 = 1.8 nH in the other CRLH transmission line (ports 5 to 6). The gap between the lines is 0.2 mm and the coupled length is 100 mm. The CRLH transmission line consists of 20 cells with a length of 5 mm, which is much longer than for the symmetrical power dividers. As a signal is input to port 1, the power couples to port 3 and port 5 in different frequency bands, and ports 4 and 6 are isolated. The measured S-parameter results shown in ***Figure 11*** are in good agreement with the simulated results shown in ***Figure 12***.

DISCUSSION

This power divider works on the basis of a mixed conventional microstrip and composite right-/left-handed backward-wave directional coupler.[14] Here, the composite right-/left-handed transmission line is called one-dimension metamaterial. Compared with the traditional microstrip coupler, the difference is obvious. The coupler based on metamaterials has a higher coupling and a larger bandwidth. The coupling between a microstrip line and a one-dimension metamaterial line is much tighter than the coupling between two traditional microstrip lines. A possible mechanism may be that the group velocity in this metamaterial and between a microstrip line and a one-dimension metamaterial line is much slower than in a normal media. Shadrivov, et al.[17] supposed that there are vortexes between the LH and RH lines, resulting in a lower effective coupling length and lower group velocity. In this case, more energy can be transferred be-



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tween the microstrip line and the one-dimension metamaterial line, so that it shows a tighter coupling than in a traditional microstrip coupler.

CONCLUSION

One-dimensional metamaterials, using L-C lumped-elements, have a broad left-handed passband, with anti-parallel phase and group velocities. Symmetric and asymmetric structure power dividers were made, which show superior characteristics: broadband, multi-ports and arbitrary coupling level. The simulations and measurements all show their superior characteristics. ■

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Hongqiang Li *received his B.Sc. and M.Sc. degrees from Nankai University in 1991 and 1994, respectively. In 1997, after obtaining his PhD degree from the Institute of Physics, Chinese Academy of Science, he joined the physics department of Tongji University, where he is now a full professor. His research interests include photonic crystals, left-handed materials and metamaterials, and wave propagation in inhomogeneous media.*

Hong Chen *obtained his B.Sc. degree in physics from Shanghai Fudan University in 1982 and his PhD degree in condensed matter physics from Shanghai Jiaotong University in 1986. He has been with the Pohl Institute of Solid State Physics at Shanghai Tongji University since 1986, where he is currently a full professor. His current research interests include photonic bandgap materials and metamaterials.*

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MAKING THERMAL RESISTANCE MEASUREMENTS WITHOUT TEST DIODES OR THERMAL STAGES

Thermal resistance is a measure of the amount by which the junction temperature rises for a given level of power dissipation in a transistor

$$\theta = \frac{\left(T_{junction} - T_{external}\right)}{P_{diss}} \quad (1)$$

Thermal resistance is important since it determines how much power a device and package combination can safely dissipate. Thermal resistance measurements normally require the measurement of a test diode and a thermally controlled environment. This article describes an approach that uses the gate or base of the device under test, rather than a separate test structure, and additionally can produce results without the need of a thermal chuck or oven. A single pulsed-measurement instrument and the device itself are used to generate different temperature points by setting different internal power dissipations. The approach is straightforward enough to be performed on a desktop computer. The test device can be a standard packaged part in the form that will be used in a circuit, including any heat-sinking arrangements.

The use of pulses, which are short compared to the thermal time constants involved, allows measurements to be made rapidly enough for the thermal equilibrium to remain undisturbed. The whole diode characteristic is obtained, which allows temperature to be deduced. This is in contrast to the traditional approach where a single point is measured, which requires careful calibration in temperature-controlled ovens or using other means of setting a constant thermal environment. The novelty of this method over the long accepted technique of using the temperature dependence of the forward diode voltage at a fixed current is that a complete diode characteristic is fitted. There are two significant advantages to this approach. First, the temperature can be deduced without the need of a pre-calibration procedure. Second, as part of the fitting process, the series resistance can be fitted, which eliminates the effects of parasitic resistance that may change with temperature. Examples of the technique, applied to both a FET and a bipolar device, are given.

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THERMAL RESISTANCE

Thermal resistance is an important figure of merit for how well a transistor package and heat-sink combination is able to dissipate heat and thus to what power level the device may safely be run. Data sheets for commercial packaged parts give typical values for the general case. However, from the design engineer's point of view, the thermal resistance of the device, as used in the designed circuit, is required. One option is to calculate an estimate, based on the data sheet value and data for the heat sink used, the thermal paste and so on. A more straightforward approach is to directly measure it, although such a measurement has previously required a means of measuring junction temperature, which in turn has required an additional sampling diode and a calibration procedure using some means of uniformly elevating the device's temperature to a known value. This procedure is described in References 1 and 2, and runs to approximately thirty pages. The method described in this article requires only one temperature point, room temperature, thus avoiding the need for ovens or thermal chucks. Additionally, the forward conduction characteristics of the device itself are used to monitor the actual junction temperature.

MEASUREMENT OVERVIEW

The concept of thermal resistance assumes a linear relationship between the power dissipated in a device and its junction temperature. It is assumed that power dissipation has remained constant for long enough for thermal equilibrium to be attained. At zero power dissipation, the junction temperature should be room temperature (that is zero heating effect). An estimate of thermal resistance can, in theory, be made by measuring the junction temperature at just one elevated temperature point. In practice, better results will be obtained by making junction temperature measurements at several dissipated power levels and plotting a straight line through the results.

Thus, to measure the thermal resistance, the two variables to be measured are power dissipation and junction temperature. Power dissipation is simply obtained from the voltage drop across the device and the current through it. Junction temperature is more challenging and is discussed in the next section.

MEASURING JUNCTION TEMPERATURE

Since it is not possible to measure the junction temperature directly, an indirect approach must be used. The method used the effect the temperature has on the forward conduction characteristics of a diode. The test diode may be a separate device included in the test cell, close to the FET or bipolar under test, but it is also possible to use the gate or base as the test diode. In this article, it is assumed that there is no external test diode, so the base or gate of the device under test (DUT) must be used.

Though many large-signal models assume a diode-like behaviour for the gate of a FET or HEMT, or the base-emitter of a bipolar, in real devices, the behaviour is affected by the collector or drain terminal and may be far from being diode-like. For the method to work, the measurement of the diode current must be made with the collector of drain potential set to give a close approximation of the diode behaviour on the base or gate port. Measurements are thus best taken with zero, or close to zero voltage on the collector or drain port.

During the measurement, there will be almost no power dissipated in the device while the collector or drain is set to zero. The measurement must be performed fast enough so that there is no significant cooling of the junction during the measurement. In this work, a pulsed-measurement instrument (DiVA) was used to perform the measurements using repeated pulses, each of which was fast enough to prevent cooling from taking place. In brief, the measurement process is to set a range of different junction temperatures using different bias points for the device under test. At each bias point, the base or gate diode characteristics, with zero volt on the collector or drain, are measured using sufficiently fast pulses. The data is then analysed to obtain an estimate of the junction temperature for different power dissipations. The resultant points should lie on a straight line, the slope of which will give the thermal resistance.

OBTAINING THE JUNCTION TEMPERATURE FROM MEASUREMENT

The characteristics of a diode may be expressed in simple terms as

$$i = I_0(T)\left[e^{\frac{qV}{mkT}} - 1.0\right] \qquad (2)$$

where

q = electron charge
k = Boltzmann's constant
m = ideality factor
V = voltage across the diode (after accounting for any series resistance voltage drop)
T = temperature

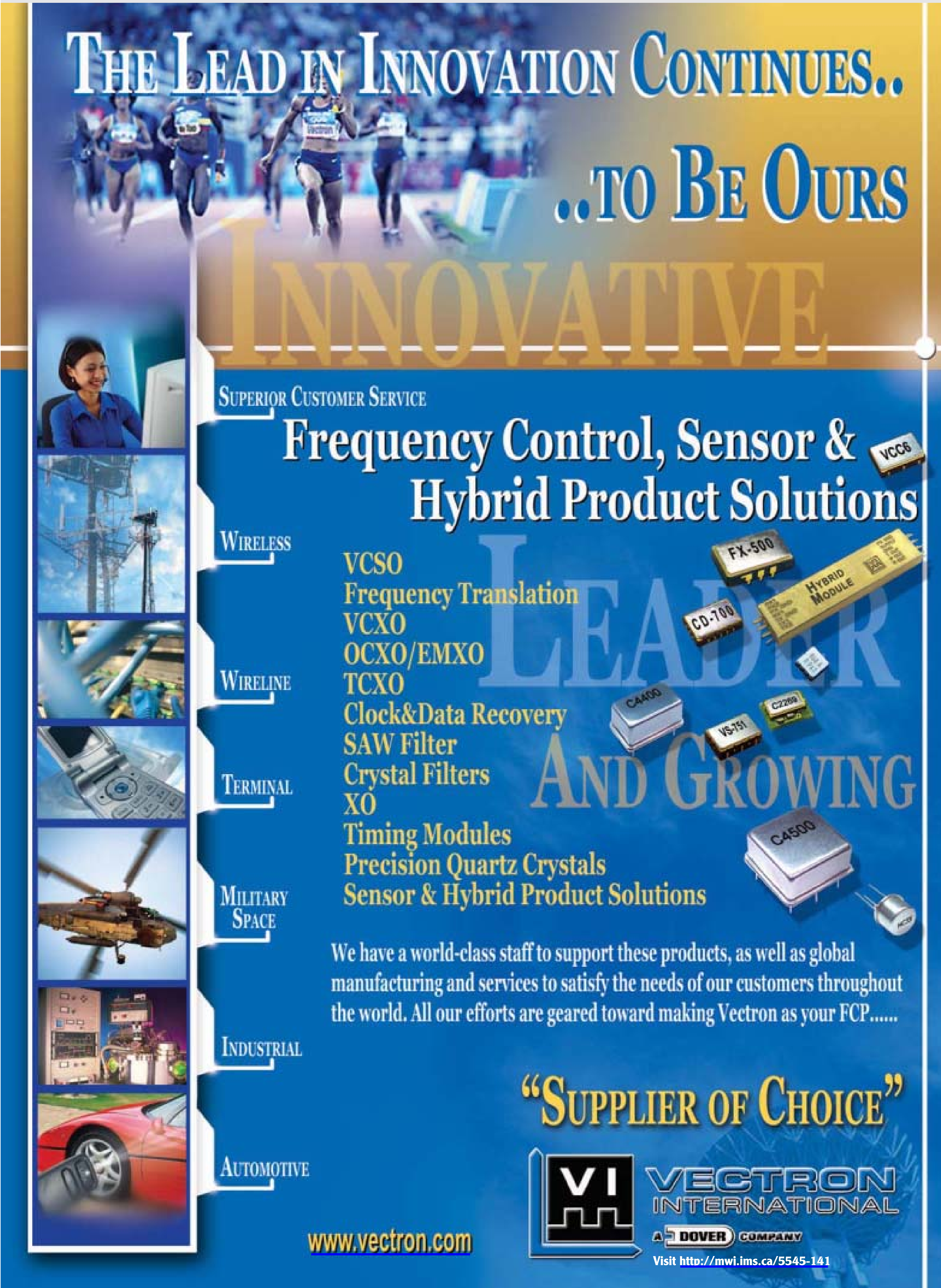
For any fixed temperature, $I_0(T)$ is fixed.

Fitting the above equation to measured data at a fixed temperature (value of T) yields two parameters. One is $I_0(T)$ and the other is the product m·T. The first, $I_0(T)$, is a complicated function of temperature so the value of T is best obtained from the second parameter. The value of m is obtained by fitting the IV characteristics of the diode to the results with no junction heating (zero bias current) where T is known to be the room temperature. The same value of m can then be used to determine T at other temperatures, where there is power dissipation, giving rise to an elevated junction temperature.

DETERMINING A SUITABLE PULSE LENGTH

It is important to get the right pulse length for the measurement of the base or gate diode characteristics. If the pulse length is set too long, then there will be some cooling of the junction during the measurement itself, leading to erroneous results. If the value is set too short, then effects arising from cable capacitance or inductance may arise and trapped charges within the device itself may also affect the results.[3]

To determine a suitable pulse length for each sample device, the pulsed-measurement instrument was used to measure a transient current. The results of the transient measurement showed if there were rapid changes as well as slower thermal effects in the transient behaviour. The pulse length was set to be longer than the rapid changes arising from trapped charge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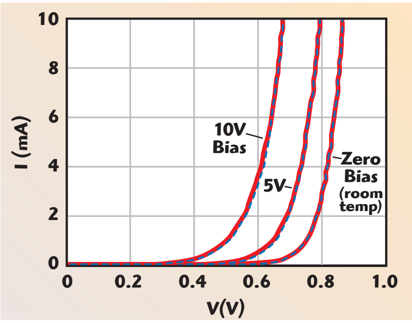


▲ *Fig. 1 Measured diode characteristics at three power dissipations for a BF480 (Si bipolar transistor).*

TABLE I

PARAMETERS FOR THE MEASURED BF480 BIPOLAR TRANSISTOR

V_{CE} bias (V)	0	5	10
I_C bias (mA)	0	36.9	40.9
P_{diss} (mW)	0	184	409
mT	570.5	751.3	924.9

TABLE II

CALCULATED JUNCTION TEMPERATURE FOR THE BF480 DEVICE

P_{diss} (mW)	184	409
$T_{junction}$ (K)	396	488
$\triangle T_{junction}$ (K)	95	187

movement, but still short enough to have no significant cooling during the pulse. Note that for completeness, it should be stated that some thermal effects can have very short time constants. This occurs when small devices are thermally isolated, perhaps by an isolation ring, so the volume to be heated is very small. But even in such cases, though the rise in temperature is rapid, the cooling may be a much slower process and so the method described in these notes may still be used to determine a relationship between dissipated power and junction temperature.

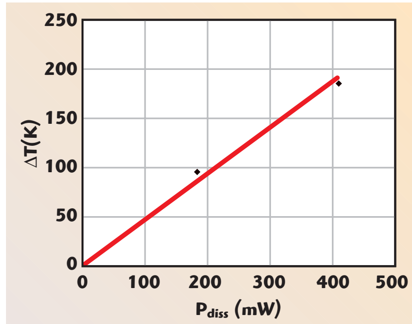
A SI BIPOLAR EXAMPLE

To illustrate the method, a BF480 silicon NPN bipolar transistor was used. For this sample device, a pulse length of five microseconds was selected. The diode characteristics at zero volts were measured for the case of zero heating and at two higher levels of heating (power dissipation). ***Figure 1*** shows the three measured characteristics along with the diode characteristic curves that were fitted to them. The parameters for the different cases are given in ***Table 1***. For the zero power dissipation case, the temperature can be taken to be room temperature, which in this case was 28°C or 301 K. From this, the value of m can be obtained

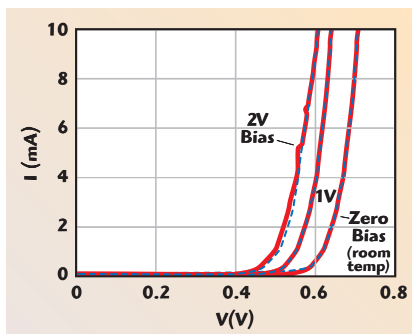
$$m = \frac{570.5}{301} = 1.895 \qquad (3)$$

Using the value of m, the junction temperature, for the cases when there is power dissipated in the device, can be calculated. The results are shown in ***Table 2***.

Figure 2 shows a plot of the junction temperature rise versus power dissipation together with a straight-line fit. The slope of the line gives the thermal resistance between junction and case. The calculated value of thermal resistance is 467 K/W. The safe operating maximum power and junction temperature for a BF480 taken from the data sheet assume a thermal resistance of 714 K/W. Given that the data sheet must allow for the worst case of all samples plus a safety margin, the two values of thermal resistance are consistent with each other.



▲ *Fig. 2 Junction temperature rise versus power dissipation for the BT480 device.*



▲ *Fig. 3 Measured diode characteristics at three power dissipations for CLY2, a GaAs FET sample.*

A GaAs FET EXAMPLE

As a second example, a GaAs FET was selected. The device is a CLY2 mounted in a brass jig. For this sample device, a pulse length of fifty microseconds was selected. (The CLY2 device is affected by trapped charges in deep levels so the pulse length must be long enough to enable such charges to redistribute.) As with the BF480 sample, the diode characteristics at zero volts were measured for the case of zero heating and at two higher levels of heating (power dissipation). ***Figure 3*** shows the three measured characteristics along with the diode characteristic curves that were fitted to them. The parameters for the different cases are given in ***Table 3***.

TABLE III

PARAMETERS FOR THE MEASURED CLY2 FET TRANSISTOR

V_{DS} bias (V)	0	1	2
I_D bias (mA)	0	594	587
P_{diss} (mW)	0	594	1174
mT	364.2	387.8	402.2

TABLE IV

CALCULATED JUNCTION TEMPERATURE FOR THE CLY2 DEVICE

P_{diss} (mW)	594	1174
$T_{junction}$ (K)	318	330
$\triangle T_{junction}$ (K)	19	31

For the zero power dissipation case, the temperature can be taken to be room temperature, which in this case was 26°C or 299 K. From this, the value of m can be obtained

$$m = \frac{364.2}{299} = 1.218 \qquad (4)$$

Using the value of m, the junction temperature for the cases when there is power dissipated in the device can be calculated. The results are shown in ***Table 4***.

Figure 4 shows a plot of the junction temperature rise versus power dissipation together with a straight-line fit. The slope of the line gives the thermal resistance between junction and case. The calculated value of the thermal resistance is 28 K/W. The data sheet for the CLY2 states that the thermal resistance is less than 110 K/W. The measured thermal resistance is quite a bit lower than the given maximum value; this may be due in part to the jig acting as a heat sink.

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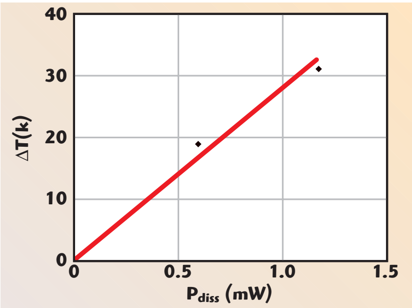
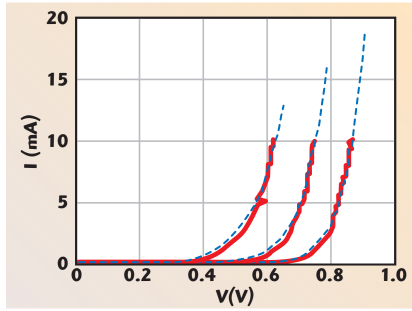


▲ *Fig. 4 Junction temperature rise versus power dissipation for the CLY2.*



▲ *Fig. 5 Comparison between calculated and measured characteristics of the BF480 device.*

METHOD ACCURACY

To check the experimental results, measurements could be made at elevated temperature, using an oven rather than power dissipation within the device itself. Such measurements have not yet been made but the work is ongoing. The software used to fit the diode characteristics allows for a series resistance, which removes some of the uncertainty in the measurements. Taking the measurements at zero collector or drain voltage eliminates collector or drain current, which may introduce a voltage offset. As shown in ***Figure 5***, the closeness of the fit to the diode characteristics gives some confidence that the base or gate is acting in a diode like manner under the measurement conditions.

CONCLUSION

This article describes a means of measuring thermal resistance using a pulsed-measurement instrument without the need for a separate test diode or the use of a temperature-controlled oven for calibration. The results obtained are reasonable but further work is needed to make direct comparisons with results obtained using a thermally controlled oven to make an estimate of the accuracy of the method. The method used makes it easy to measure an overall thermal resistance even for complicated packaging and heat-sinking arrangements. ■

ACKNOWLEDGMENT

The author would like to acknowledge the extensive input to this work, through technical discussions, of Dr. Peter Ladbrooke.

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James Bridge *obtained his degree in engineering science from Oxford University in 1982. He co-founded GaAs Code Ltd. in 1988, working on semiconductor device physics. In June 2001, he joined Accent Optical Technologies and now works as DiVA applications support manager in the optoelectronics division.*

A UWB Filter Using a Dual-mode Ring Resonator with Spurious Passband Suppression

A novel, compact, microstrip bandpass filter for ultra-wideband (UWB) radio systems is proposed. This filter has a 3 dB fractional bandwidth of 60 percent, with low insertion loss and sharp rejection. Based on a dual-mode ring resonator structure and with the addition of two stepped-impedance open stubs, it can provide a wide passband and two sharp stop-bands. A frequency rejection greater than –15 dB can be observed between 8 and 14 GHz with the proper position of tapped lines at the input and output. The filter is designed for full duplex systems in satellite communications.

The investigation of the use of ultra-wideband (UWB) has been one of the most controversial technologies of modern times. Its applications can be seen in diverse areas such as local area networks, position location searching, advanced imaging of the human body, etc. Microwave ring resonators are the components proposed for filtering, modulation and multiplexing/demultiplexing tasks in UWB integrated circuits. The basic concept for the microwave ring resonator was first proposed by P. Troughton in 1969.[1] It was used in the measurement of the phase velocity and dispersive characteristics of a microstrip line. A further study on microwave ring resonators, using transmission-line models, was proposed.[2] This model provided a T-network equivalent circuit to analyze the ring resonator structure. It has been proven that the microwave ring resonator can support two degenerate modes and offers bandpass characteristics. There have been many studies of dual-mode ring resonator bandpass filters.[3–5] However, these filters have narrow band characteristics and high insertion loss because of the coupling gap effect. L.H. Hsieh and K. Chang have proposed a ring resonator bandpass filter.[6] This filter provides a wide passband, sharp rejection and low insertion loss, but a spurious response is excited at $2f_0$. Also, the addition of two tuning stubs results in enlarging the area of the overall filter structure.

The idea presented in this article is based on a ring resonator structure, with the addition of stepped-impedance stubs and tapped input/output (I/O) lines, to construct a high performance bandpass filter. The tapped I/O

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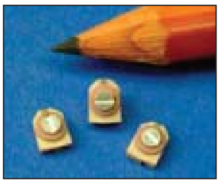

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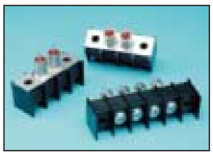

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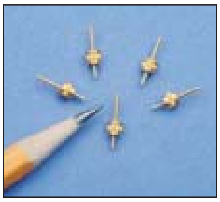

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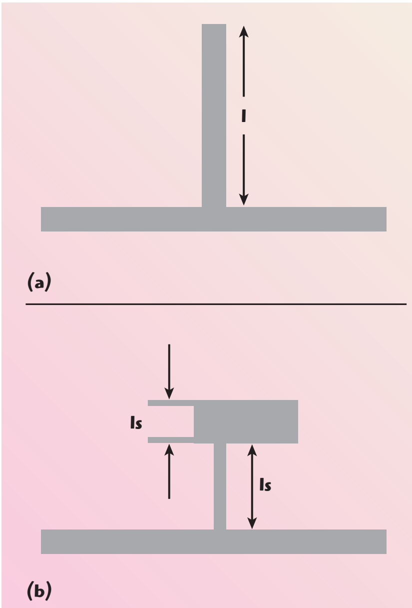
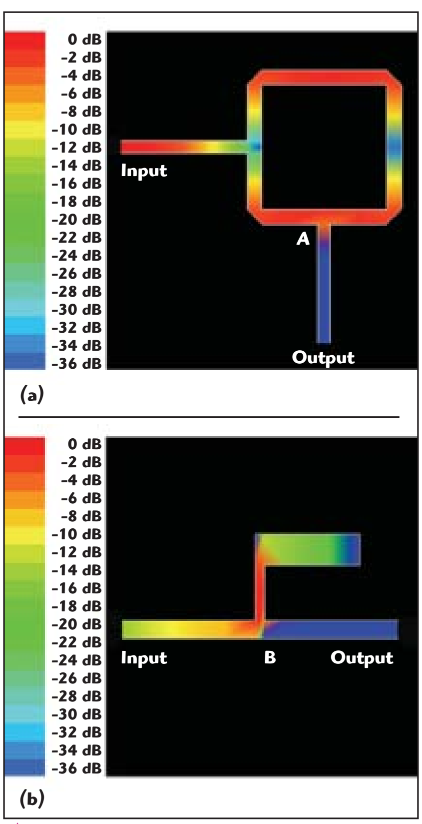
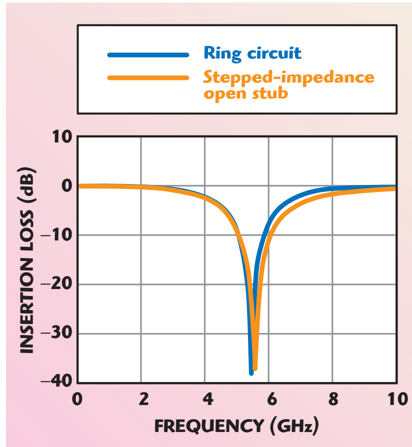


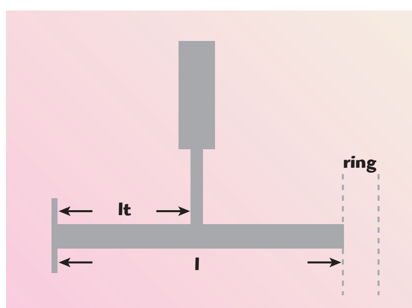
▲ *Fig. 1 Band-stop circuits; (a) an open stub and (b) a stepped-impedance stub.*



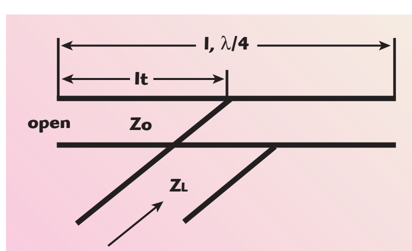
▲ *Fig. 2 Simulated electric current at the resonant frequency for (a) a ring resonator and (b) a stepped-impedance stub.*



▲ *Fig. 3 Simulated insertion loss of the band-stop filters.*



▲ *Fig. 4 Tapped-line resonator.*



▲ *Fig. 5 Tapped-line resonator equivalent circuit.*

lines may have an independent extra transmission zero in the stop-band without requiring complex coupling between resonators.[7,8] Without altering the passband response, the tapped I/O line can be properly applied to both feeding ports in order to control the positions of the extra zero. This is a very useful feature in practical receivers for rejecting spurious response and enhancing the rejection level in the stop-band of a bandpass filter. The proposed structure can suppress the spurious passband and save more area of the overall filter structure without degrading the performance of the bandpass filter.

STEPPED-IMPEDANCE OPEN STUBS

The band-stop circuits, illustrated in ***Figure 1***, are designed to excite a band-stop response by adding two λ/4 open stubs on two sides (0° and 90°) of the ring resonator. Based on transmission line theory, a transmission line section having a length ($\beta l < \pi/2$) can be replaced by combining a short length ($\beta l < \pi/4$) of line of high characteristic impedance with a short length ($\beta l < \pi/4$) of line of low characteristic impedance. The latter can be referred to as a stepped-impedance structure. The EM simulated electric current distributions in the ring circuit and in a stepped-impedance open stub band-stop filter at the same fundamental resonant frequency are shown in ***Figure 2***. The simulated electric current shows that the minimum electric fields appear at positions A and B, which correspond to the maximum magnetic fields. Thus, both circuits provide band-stop characteristics by presenting a virtual ground at their outputs at the resonant frequency, as can be observed in their simulated insertion loss frequency responses, shown in ***Figure 3***. Additionally, the stubs can be easily implemented inside the ring resonator because their length l_s must be shorter than λ/8. A new filter is formed which can provide a more compact size.

Figures 4 and ***5*** show the tapped-line structure and its corresponding transmission line model. The position parameters of the tapped line can be obtained from[9]

$$l_t = \frac{2l}{\pi}\cos^{-1}\left(\sqrt{\frac{\pi}{2}\frac{Z_L}{Z_0}\frac{1}{Q_s}}\right) \qquad (1)$$

or

$$Z_L = \frac{2}{\pi}Z_0 Q_s \cos^2\left(\frac{\pi}{2}\frac{l_t}{l}\right) \qquad (2)$$

where

l_t = position of tapped-line point
Z_1 = impedance of the tapped line
Q_s = loaded quality factor
Z_0 = characteristic impedance of the transmission line

UWB FILTER USING DUAL-MODE RING RESONATOR WITH SPURIOUS PASSBAND SUPPRESSION

A novel bandpass microstrip filter based on a ring resonator with tapped I/O lines is shown in ***Figure 6***. The input and output ports are directly connected to the ring resonator at 180° and 270°. Two stepped-impedance tuning stubs are implemented within the resonator at 0° and 90°. The circumference l_r of the ring res-

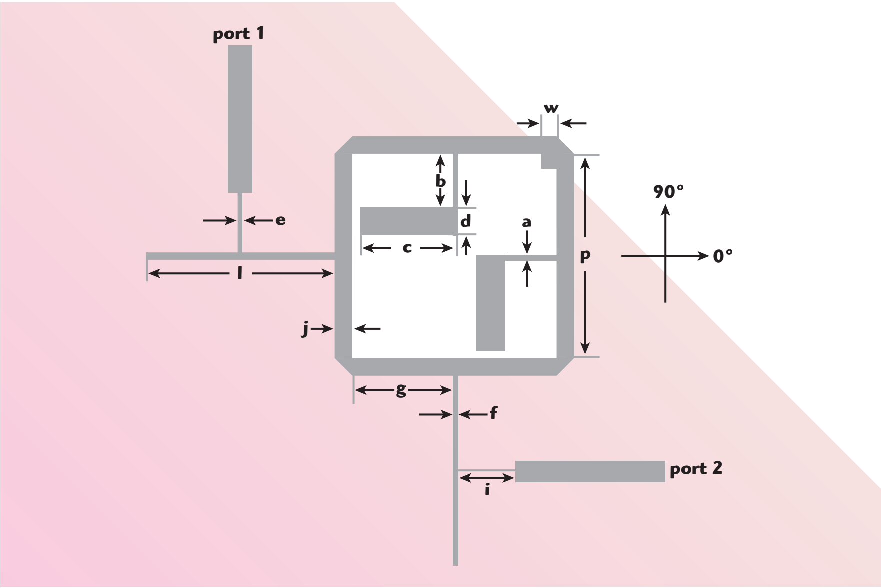


Fig. 6 *Layout of the designed UWB filter.*

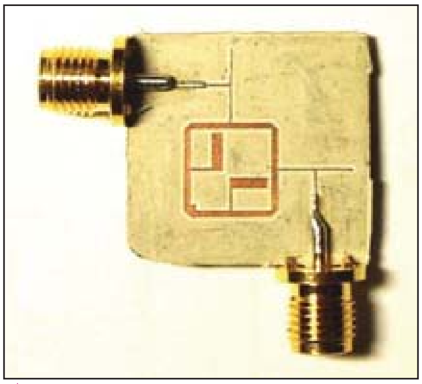


Fig. 7 *The fabricated UWB bandpass filter.*

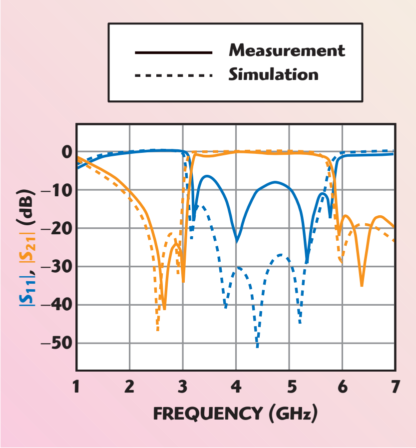


Fig. 8 *Passband performance of the designed UWB filter.*

onator is given by Equation 3, where n is the mode number, λ_g is the guided wavelength and a perturbation is positioned at 45°

$$l_r = n\lambda_g \quad (3)$$

The ring resonator and λ/8 stepped-impedance stubs are designed to resonate at 4.5 GHz and fabricated on an RT/Duroid 6010.2 substrate with a thickness h = 25 mil and a relative dielectric constant ε_r = 0.2. The dimensions of the filter are l_r = 31.4 mm, a = 0.3 mm, b = 1.8 mm, c = 3.3 mm, d = 1 mm, e = 0.3 mm, f = 0.3 mm, g = 3.2 mm, l = 3.5 mm, i = 1.9 mm, j = 0.6 mm, p = 7 mm and the width of perturbed square stub is w = 0.5 mm. The total size of the filter is less than 20 x 20 mm; a photograph of the prototype filter is shown in ***Figure 7***. The passband and stopband responses of the designed bandpass filter are shown in ***Figures 8*** and ***9***. The perturbation stubs can generate two transmission zeros or dual modes on both sides of the passband, located within 2.7 and 3.02 and 6.01 and 6.4 GHz. The filter has a 3 dB fractional bandwidth of 60 percent, an insertion loss less than 0.8 dB, two rejection bands greater than 18 dB within 2.46 to 3.22 GHz and 5.79 to 12 GHz, and attenuation rates for the sharp cutoff frequency responses of 177.7 dB/GHz at the lower edge of the passband and 181.8 dB/GHz at the higher edge. As shown, the spurious response of the designed filter can be considered as effectively suppressed. The rejection of the spurious response from 7 to 11.5 GHz is successfully suppressed to a level less than –18 dB by the effect of the tapped I/O lines. The edge of the transmission passband is related to the length c of the stepped-impedance stub, as shown in ***Figure 10***. The bandwidth and the center frequency of the proposed filter can be easily tuned over a 700 MHz range by adjusting the length of c.

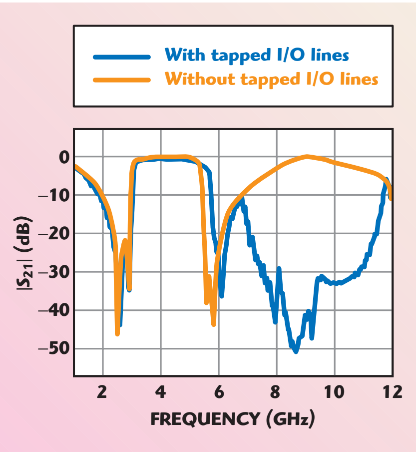


Fig. 9 *Spurious response of the designed bandpass filter.*

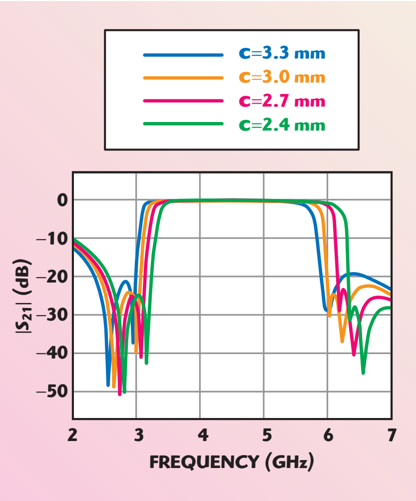


Fig. 10 *Simulated performance of the passband UWB filter as a function of the length c.*

The measured performance shows good agreement with the simulated results. The slight difference between the simulated and measured results may be due to a fabrication error, which can be improved by more precise fabrication technology.

CONCLUSION

This article presents a wideband microstrip bandpass filter with tapped I/O lines. Numerical simulations, using Zeland-IE3D, show good agreement with experiments. The proposed microstrip bandpass filter has the advantage of high performance, providing wider and deeper

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stop-band characteristics. The measured data for the fabricated bandpass filter also shows a fairly good insertion loss of approximately 0.7 dB. It can be integrated in an UWB radio system. ■

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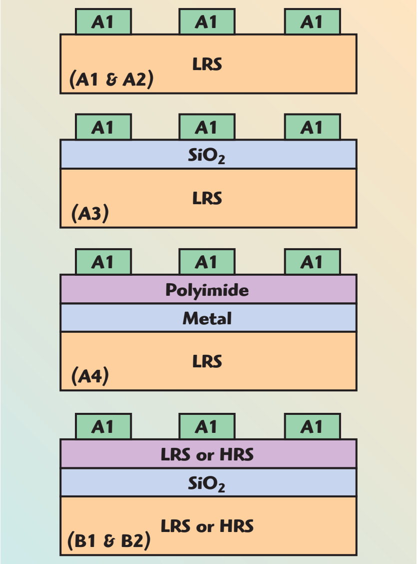
CPW TRANSMISSION INSERTION LOSS ON SI AND SOI SUBSTRATES

The transmission loss properties of coplanar waveguide (CPW), built on bulk silicon and low and high resistivity silicon-on-insulator (SOI) substrates, are investigated systematically in this article. The experimental results show that the insertion loss can be greatly decreased by isolating the CPW line and the substrate through a silicon oxide layer or using a metal shield and polyimide layer. The insertion loss of the CPW line on low resistivity (LR) SOI substrate is much lower than for the one made directly on the bulk silicon substrate with the same resistivity, decreasing by 45 percent. The transmission performance of the CPW line was improved by using a high resistivity (HR) SOI substrate.

Coplanar waveguides are widely used in MMICs as interconnects and matching networks. Designing low loss, multi-functional and highly integrable transmission lines is a key factor in obtaining high performance from MMICs in silicon technology.[1] Nevertheless, the low resistivity of a standard silicon substrate deteriorates the performance of on-chip passive components.[2,3] To overcome the problem of high dielectric loss of the low resistivity silicon (LRS) substrate, a thick SiO_2 layer[4] and a spin-coated thick polyimide layer[5] are utilized in microwave applications. A high resistivity silicon substrate is also used in the microwave region because of its low dielectric loss.[6,7]

Fig. 1 CPW structures.



In order to investigate the substrate effect on the transmission performance of the CPW line systematically, low resistivity Si (LRS) substrate, SOI substrate and HR SOI are employed in the experiments. The CPW structures fabricated on these substrates are shown in ***Figure 1***. The samples A1 and A2 are built directly on LRS substrates, A3 is fabricated on SiO_2, A4 on polyimide with a metal shield layer, and B1 and B2 are made on LR SOI or HR SOI substrates. The insertion losses of the CPW lines are greatly decreased when a shield metal layer and polyimide interlayer are on the LRS substrate. A novel HR SOI structure, with a buried oxide layer in silicon and a HR substrate, shows a low attenuation of the CPW lines.

EXPERIMENT

All the CPW lines were designed with a 50 Ω characteristic impedance. The metal is aluminum with a thickness of 0.6 µm. The width of the centerline is 0.1 mm and the gaps are 0.08 mm wide. The length of all the lines tested was 3 mm. Two categories of substrates were used: bulk LRS substrates with resistivities of 0.5 Ω cm and 20 Ω cm, and separation by implanted oxygen (SIMOX) SOI substrates with resistivities of 20 Ω cm and 1000 Ω cm. The buried oxide (BOX) thickness is 380 nm and the top silicon thickness is 200 nm. The samples, with different substrates and structures, are summarized

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TABLE I

CPW STRUCTURES

Substrate	Sample	Substrate Resistivity and Special Structure (Ω·cm)
Bulk silicon substrate	A1	0.5
	A2	20
	A3	20 with 1 µm SiO_2
	A4	20 with shield metal ground and 10 µm polyimide
SOI substrate	B1	20
	B2	1000

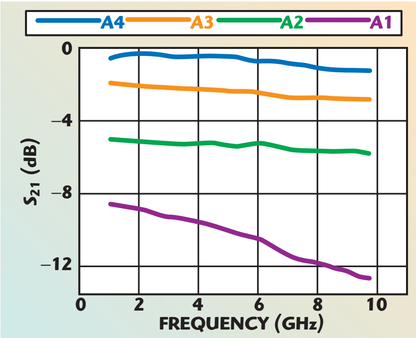
in ***Table 1***. The impact of the ground line width on the CPW transmission behavior was also investigated in this experiment. The ground line width was varied from 200 to 600 µm.

The CPW transmission characteristics were measured with an Agilent HP8722D vector network analyzer and probe station. The probe was calibrated before the measurements.

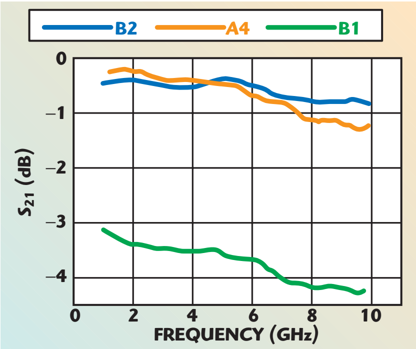
RESULTS AND DISCUSSION

CPW on Bulk Silicon Substrates

The measured results taken on the structures A1 and A2 are shown in ***Figure 2***. The attenuations of A1 and A2 are 2.9 dB/mm and 1.7 dB/mm at 2 GHz, respectively. The insertion loss of CPW on low resistivity silicon is high, which may be caused mainly by the substrate loss. There is a high concentration of free electron carriers in bulk silicon substrates. The electromagnetic coupling produces an induced current in the bulk silicon substrate, which induces an alternate polarization of the dielectric molecules and collision of crystal lattices, resulting in high losses. In addition, when the resistivity of the substrate increases to 20 Ω cm, the attenuation of the CPW line decreases to 1.7 dB/mm. Increasing the resistivity can reduce the substrate loss.



▲ *Fig. 2 Measured losses of the CPW lines on A structures.*



▲ *Fig. 3 Measured losses of the CPW lines on B structures.*

In order to isolate the passive components from the lossy substrate, a layer of silicon dioxide, approximately 1 µm in thickness, is deposited on the LRS substrate (sample A3). The curve A3 shows that the attenuation is 0.7 dB/mm at 2 GHz, 65 percent lower than that of A2. The insulating layer of silicon dioxide can prevent the propagation of the electromagnetic waves in the substrate, which results in a lower insertion loss.

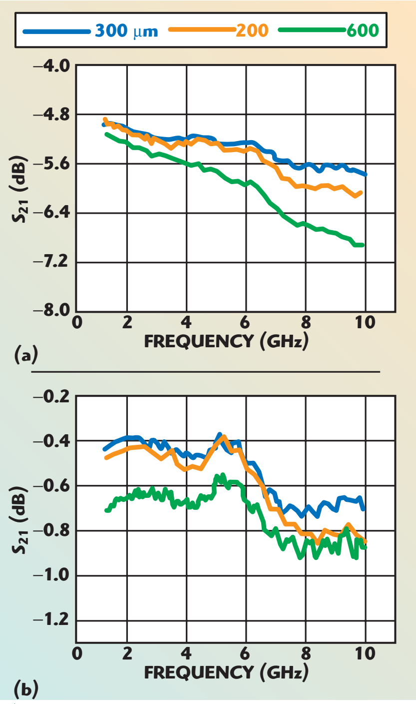
To improve the behavior of CPW on low resistivity silicon, a 0.6 mm thick ground plane, made of aluminum, is introduced in the LRS, on which a 10 µm thick polyimide layer is deposited (sample A4). The curve A4 shows that the attenuation is 0.06 dB/mm at 2 GHz. The ground plane completely shields the electromagnetic fields from the lossy bulk silicon substrate. The attenuation is greatly reduced compared to sample A2.

CPW on SOI

The CPW transmission performance on SOI substrate was also investigated in the experiment. They were fabricated directly on SOI substrates with resistivities of 20 Ω cm and 1000 Ω cm (samples B1 and B2). The measured results are shown in ***Figure 3***. At 2 GHz, the attenuation for B1 is 1.1 dB/mm and 0.13 dB/mm for B2. The attenuation of the CPW line made on SOI is 45 percent lower than that of the bulk silicon of the same resistivity, 20 Ω cm. The buried oxide layer in SOI offers a complete isolation between the top silicon and the bulk substrate and eliminates the substrate current injection path.

A high resistivity substrate can be selected for SOI to further reduce the CPW loss at high frequency, which is impossible for bulk technology due to latch up concerns. When the SOI substrate resistivity increases to 1000 Ω cm, the attenuation decreases to 0.13 dB/mm, as shown in curve B2, reduced by 88.2 percent, which is comparable to the structure with a shield layer (0.07 dB/mm). With HR SOI, the substrate loss can be reduced and the performance of the passive component is improved greatly. The results of A4 are also shown for convenient comparison. The attenuation of B2 is a little higher than that of A4 between 2 and 5 GHz, but when the frequency increases, the attenuation of B3 is lower than A4, which means that high resistivity SOI has the advantage in higher frequency range applications.

The transmission line attenuation is attributed to the conduction and dielectric losses.[8] The CPW lines used here are 0.6 µm thick, which is only about one skin depth, that is to say, only 1/e of the current can pass through the CPW lines, which will increase their loss. If the CPW lines are made thicker, the attenuation can be further reduced. In addition, if some



▲ *Fig. 4 Insertion loss of CPW lines with different ground line widths; (a) on LRS substrate and (b) on HR SOI substrate.*

isolation is added between the SOI substrate and the CPW line, the behavior of the CPW can be improved. Still, the HR SOI has great potential in improving the performance of passive components at high frequencies.

The Impact of the Ground Line Width on CPW Lines

Figures 4a and ***b*** show the attenuation of the CPW lines with ground line widths of 200, 300 and 600 µm on Si and SOI substrates, respectively. The Si substrates are 20 Ω cm and the SOI are 1 kΩ cm. The attenuation of the CPW line is the lowest when the ground line is 300 µm wide and largest when the ground line is 600 µm. The electromagnetic energy radiation declines when the ground line width increases. When the ground line width increases to a certain extent, however, a higher mode appears, which will increase the loss of the CPW lines again.

CONCLUSION

The transmission loss properties of CPW on bulk silicon and LR and HR SOI substrates were investigated systematically in this article. By isolating the CPW line and the bulk Si substrate with SiO_2 or a metal shield and polyimide layer, the performance of the CPW line is greatly improved. An SOI substrate with a buried oxide layer in silicon can eliminate the current path and improve the performance of passive components. With the choice of a high resistivity substrate, the insertion loss of CPW lines on HR SOI was reduced and demonstrated the prospect of attracting applications in the gigahertz field. ■

ACKNOWLEDGMENT

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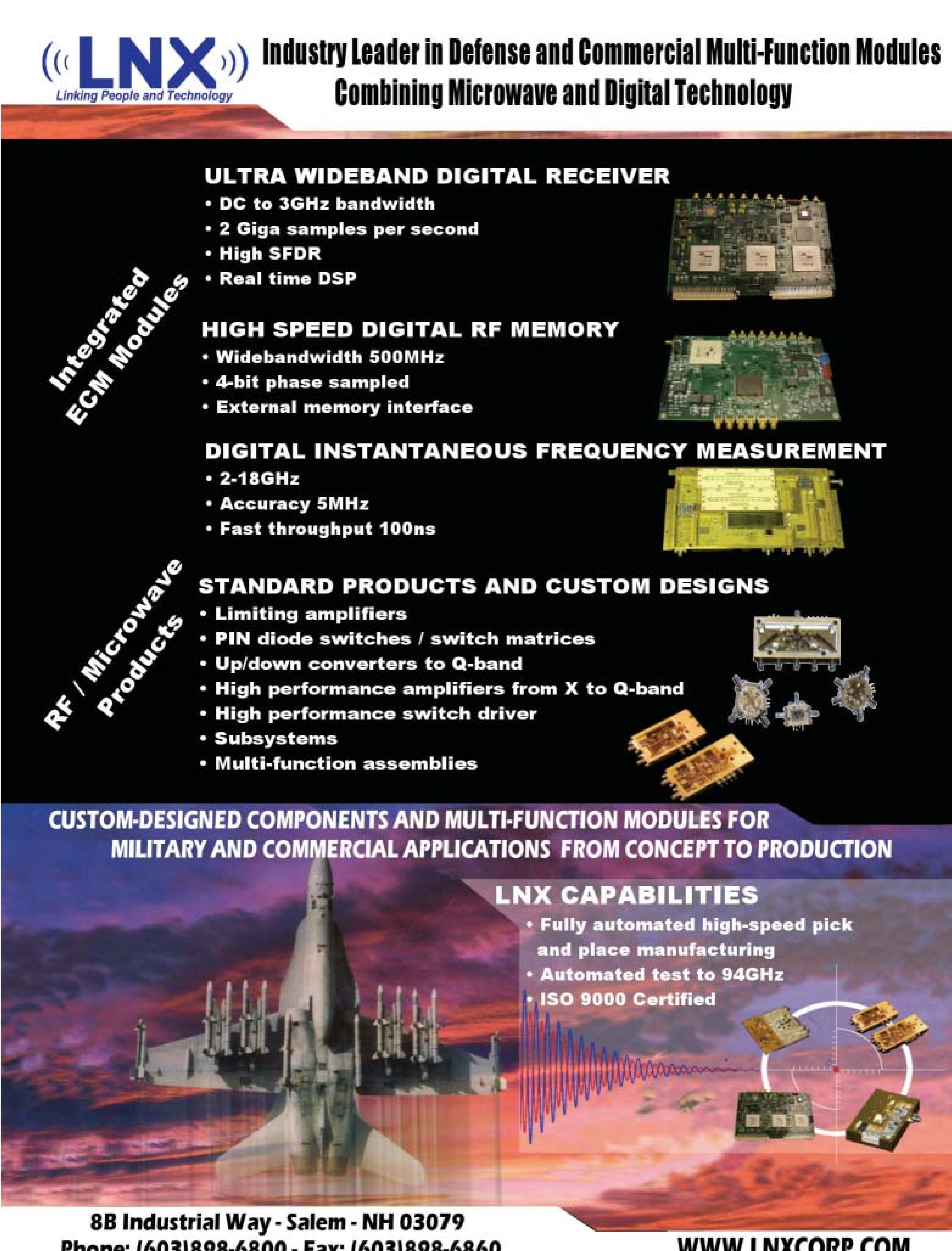

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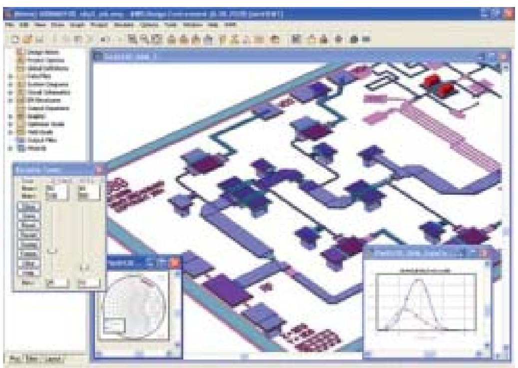

BREAKING THE EDA INTEROPERABILITY BARRIER

Electronic design automation (EDA) tools have traditionally been developed as separate sets of tools offering little or no interoperability with third-party tools. As the design of next-generation communications products evolves and becomes increasingly complex, the need for a more integrated tool flow is elevated, and it has become obvious to electronic designers that the closed, single solution is no longer a viable alternative. Tool interoperability has become so important that a panel including most of the biggest EDA vendors in the industry was convened at the 2005 IEEE International Microwave Symposium to discuss the issue. One panel member described the present EDA landscape as a feudal system in which each fiefdom has its own RF and microwave simulation software and does not want to integrate with third-party tools that compete with its own products. Perhaps the most significant presentation was made by industry guru Jim Rautio, president and CEO of Sonnet Software, who predicted that by the end of the decade, the relative success and failure of EDA vendors will be determined primarily by the relative success and failure of their interoperability efforts.

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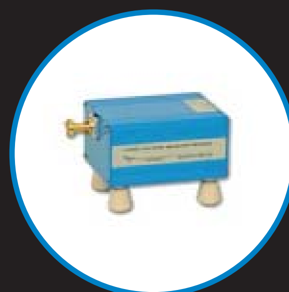

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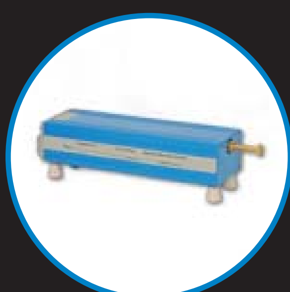

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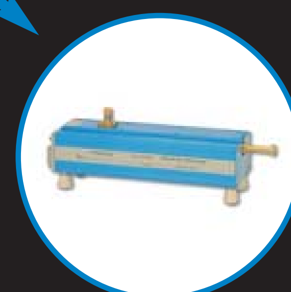

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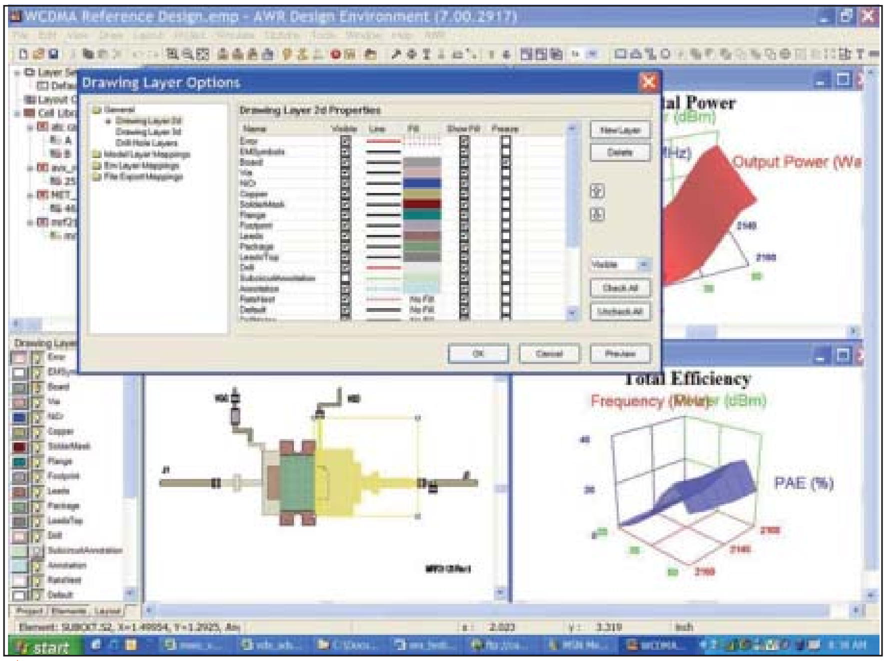

▲ *Fig. 1 The new EMediacy editor provides combined EM and layout editors.*

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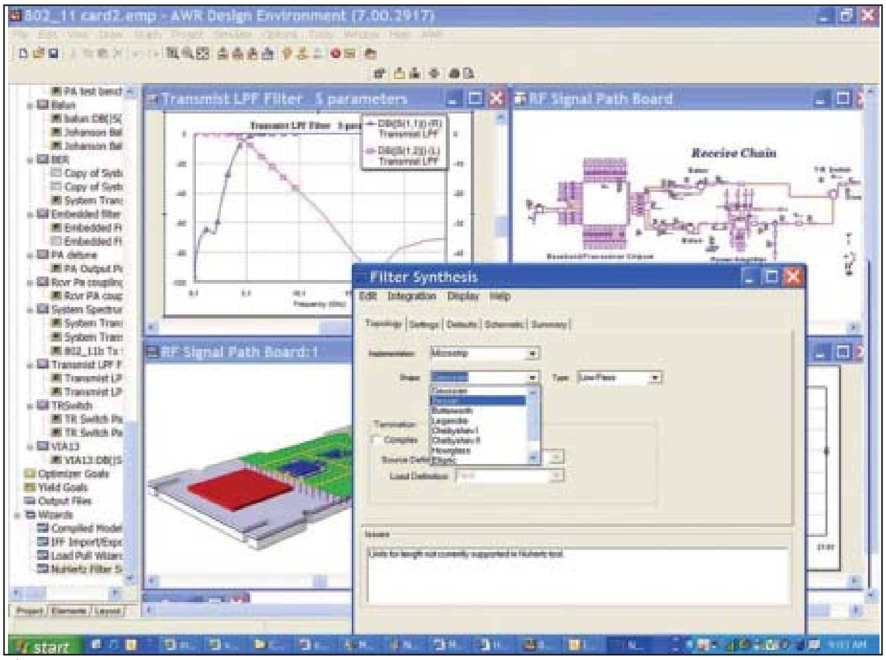
For the first time, the separate EM and layout editors that accomplish the creation of artwork are combined into one (see ***Figure 1***), reducing design time by removing one editor that engineers need to learn, as well as the associated tasks required when combining the two original editors. In combination with the schematic EXTRACT feature, EM simulation is now immediately available within the circuit design flow, rather than being a disconnected and disjointed task in a top-down design process. The EMediacy™ editor changes the fundamental way designers access EM simulation, extraction and physical analysis by embedding them as logical extensions of circuit design without sacrificing their role in verification.

INTEGRATED FILTER SYNTHESIS SOLUTION

AWR has answered the needs of customers for integrated filter synthesis by integrating into the Microwave Office 2006 design suite the industry-leading Nuhertz Technologies' filter synthesis technology (see ***Figure 2***). High frequency circuit designers can now perform accurate filter synthesis quickly and easily from within the unified AWR design platform. The feature offers complete synthesis capability for passive, transmission line, active, switched capacitor and digital filters, as well as two graphical user interfaces (GUI), one



▲ *Fig. 2 Nuhertz Technologies' filter synthesis integrated into Microwave Office.*

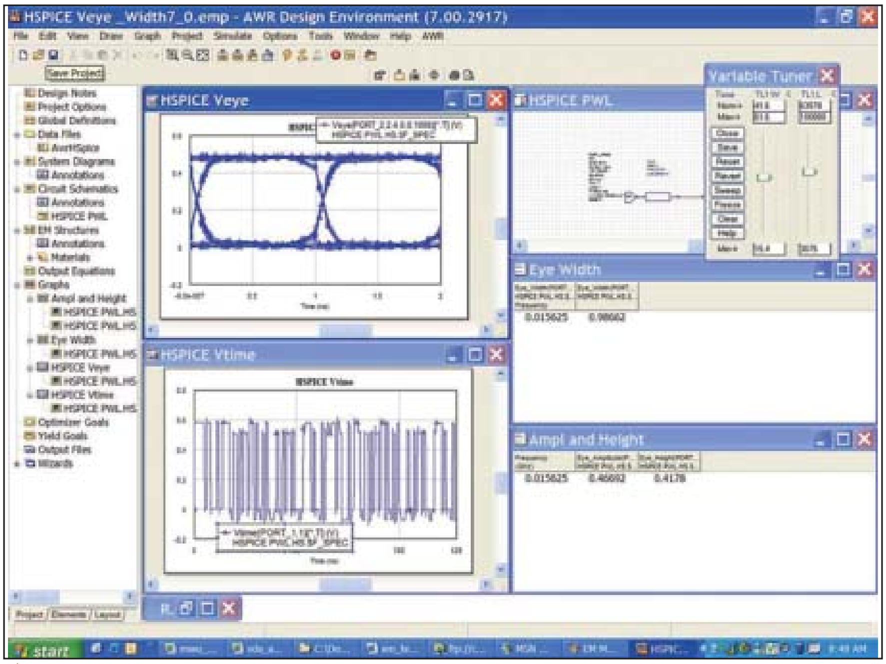


Fig. 3 Faster HSPICE simulation is offered in the Microwave Office 2006 suite.

for the power user who requires advanced options and capabilities, and one for the mainstream user who needs ease-of-use. The intuitive, easy-to-use, wizard-like GUI walks the occasional filter designer quickly through the required data steps for the specification and synthesis of all the popular filter topologies and implementations, including diplexers. For power users, one mouse-click leads to the comprehensive Nuhertz GUI, which enables the designer to quickly generate data for detailed analysis, design centering and manufacturing trade-offs. Synthesized schematics and related analyses can then be stored and viewed directly in the AWR design platform.

NEW EM SOCKET II SUPPORT VISUALIZATIONS FOR INTEGRATED THIRD-PARTY TOOLS

As part of its ongoing commitment to providing customers with greater flexibility and choice in their design methodology, AWR has created the EM Socket™ open standard interface, which enables users to access a broad variety of EM tools from leading vendors, without leaving the Microwave Office design environment. All EM manipulation and visualization features that were a valuable part of AWR's EMSight™ technology are now part of the EM Socket II interface to third-party tools. This enables EM Socket integrators like Sonnet, Zeland, Optimal, etc., to access current animations and E-field display.

OPEN ACCESS TO PROPRIETARY AWR XMODELS

AWR now provides open access of its proprietary Xmodels technology to third-party EM analysis software vendors who wish to integrate with Microwave Office. AWR's Xmodels are a group of discontinuity models that use the results of full-wave EM solutions of the parameterized discontinuity in order to estimate the electrical performance of the discontinuity. These models are a result of ongoing internal research and development at AWR in order to provide designers with the most accurate discontinuity models at a computational speed adequate for tuning, optimization and yield analysis. The AWR Xmodels have proven to be an efficient and reliable complement to full-wave EM simulation that not only enable the performance of various other important circuit analyses within Microwave Office, but also offer significant improvements in the accuracy of circuit simulations.

NEW SWITCH VIEWS FEATURE

A new view-switching feature enables multiple schematic/electrical views for the same layout, simplifying linear/nonlinear/system simulation and layout-vs.-schematic check (LVS). It is often important to model or simulate a component in multiple ways, depending upon the focus of the problem at hand. Switch Views helps users to accomplish this task by enabling them to associate more than one schematic with the same piece of the layout. This is especially useful, for example, when an S-parameter file for an inductor is the model and an LVS representation is needed. Microwave Office 2006 now permits two simultaneous views: an S-parameter view for simulation and an inductor symbol for LVS for the schematic associated with same layout.

NEW SIMULATION FILTERS AND FOLDER MANAGEMENT FEATURE

New simulation filters have been added that limit the scope of the project on which the simulators should focus, without disabling individual graphs or schematics. Currently, focusing on a particular simulation subset of a design requires disabling a number of graphs, schematics and EM structures. With the new Microwave Office simulation filters, engineers will be able to specify which parts of the design should be simulated. This streamlines the simulation process by focusing the sub-tasks, which may only need one or two of the analyses contained in a much larger, complex project.

Often the analyses of a circuit design are spread over a large number of graphs and schematics that are hard to keep track of. Project folders can now be created to group and manipulate related schematics and graphs in the design for quick reference. The Microwave Office 2006 design suite also provides new time-domain waveform measurements for use with harmonic balance and time-domain simulators. Eye diagrams and time-domain waveforms in general diagrams can now be broken down into certain well-known parameters. These parameters can be explicitly measured and used for optimization, design centering and yield analysis. Other enhancements to the software include: electrical rule check (ERC)/DRC enhancements and new advanced frequency sweep (AFS) capability for faster EMSight results

(3 to 5 time speed improvement) over a large bandwidth. AFS provides fast frequency sweep capability that enables EMSight to calculate a much more detailed response for broadband simulations by using a finite and smaller number of frequency points. Faster HSPICE simulation with support for larger arbitrary n-port devices is now offered as well (see ***Figure 3***). Previously, designers were limited to the size of S-parameter files exported to HSPICE. The limit is now much higher, and the speed of the HSPICE core engine has been enhanced to the point where some tuning can actually be done in real time. Reporting of simulation warnings and errors has also been improved. Information, warnings and errors are now available in a tabbed dialog box for faster and easier disposition and resolution. The errors are also reported during simulation rather than after, enabling poorly formulated test cases to be canceled earlier in the simulation process rather than discovering errors at the end.

A new license file-sensitive installer that simplifies administration of features has been included in the Microwave Office 2006 design suite. The installer for AWR products now reads the license file and assists the installer in identifying what configurations of AWR products should be accessible through the Start menu or desktop icons.

THE FUTURE OF EDA

AWR was founded on the concept of providing a superior solution through best-in-class technologies, tool interoperability and open third-party tool interfaces. The company continues to develop its own robust design flow in the Microwave Office 2006 design suite, which includes industry-leading technology such as the APLAC simulation engine. At the same time, its open architecture enables third-party tools to be easily integrated into the design flow, thus enabling new solutions to be created to address emerging design challenges and new applications. The Microwave Office 2006 design suite includes the most innovative integration technologies in the industry, such as the EM Open Socket™ interface. The software embodies AWR's commitment to the open design platform philosophy and its leadership position in providing designers with the flexibility to integrate tools of choice at different stages of the design process for a superior design methodology.

PRICING AND AVAILABILITY

AWR will release the Microwave Office 2006 design suite to customers in Q1 2006. A beta version is available immediately for customers with support contracts. The product supports Windows 2000, XP and Linux. US list prices for yearly, time-based licenses range from $5000 to $40,000 depending upon configurations. For more information, contact AWR at info@appwave.com.

Applied Wave Research Inc.,
El Segundo, CA (310) 726-3000,
www.appwave.com.

RS No. 301

A 125 W Pulsed Ku-band Power Amplifier

Advanced Ku-band radar systems used in airborne military platforms need reliable, small size and weight, high power amplifiers. Sophia Wireless Inc. has developed a solid-state power amplifier, model MPC8-1520, which meets these requirements. Its key features include:

- Operating frequency: 16 to 17 GHz
- Internally pulsed RF
- DC storage capacitor bank
- Monitor and control processor
- Forward/reverse power monitor
- Bias/thermal protection sensors

PRODUCT DESCRIPTION

The MPC8-1520 solid-state power amplifier provides an unparalleled size, weight and power combination. The SSPA features 125 W of power, 250 ns DC switching, 47000 µF capacitive DC storage, 45 dB gain with temperature compensation, forward and reverse power monitor with RS422 readout and a monitor and control processor, including bias sequencing control. The features and options are ideal for advanced Ku-band radar systems in military airborne platforms such as terrain-following and multi-mode radar systems, as well as navigation systems. The amplifier size is 11.1" × 4.5" × 3.1" and its weight is less than 5.1 lbs in its indoor configuration.

PERFORMANCE FEATURES

The MPC8-1520 solid-state power amplifier provides advanced radar systems with 125 W of RF power while the RF output power noise level during off periods is –172 dBm/Hz. The 250 ns DC switch (TTL controlled with a coaxial trigger) and the internal 47000 µF capacitor bank reduces the thermal management requirements and enables excellent power efficiency. A temperature compensation circuit minimizes gain and power variations. A forward and reverse power monitor simplifies the system power settings and reverse power protection. The safety controls include thermal and bias protection sensors to ensure reliable performance from initial system integration to harsh field environments.

Sophia Wireless Inc.
Chantilly, VA

TABLE I

MPC8-1520 SSPA SPECIFICATIONS

RF Parameters	Specification
Frequency band (GHz)	16.0 to 17.0
RF output power (W)	125
RF output power noise (during off periods) (dBm/Hz)	–172
F&R power monitor (15 dB range) (dB)	±0.25
Pulse width (μs)	0.5 to 400
Pulse repetition frequency (PRF)	100 Hz to 400 kHz
Cycle time (μs)	2.5 to 10,000
Duty cycle (%)	0 to 20
Linear gain (dB)	45 min.
Gain flatness (16.1 to 16.7) (dB)	±1 max.
Gain variation over temperature (dB)	2 max.
Non-harmonics/spurious (dBc)	–60
Noise figure (dB)	12
Input VSWR	1.8:1
Output VSWR	1.8:1
Monitor and Control Parameters	**Specification**
Thermal shutdown control (°C)	+85
Base plate temperature monitor (°C)	±5
Summary fault monitor	GUI viewer software
DC Power Parameters	**Specification**
+8 V DC current (current scales with duty cycle)	25A @ 20% duty cycle
–10 V DC current	100 mA
Environmental/Mechanical Parameters	**Specification**
Operating temperature (base plate) (°C)	–40 to +70
Storage temperature (°C)	–54 to +125
DC switch trigger	SMA female
RF input/output connectors	SMA female/WR-62
DC power connector	Mil-C-22992, QWLD, Shell 22-22
Monitor and control connector	Mil-DTL-38999, series 3, shell 11-35
Outline dimensions (includes options 01, 05, 06, 07)	11.1" × 4.5" × 3.1"
Weight (lbs)	< 5.1

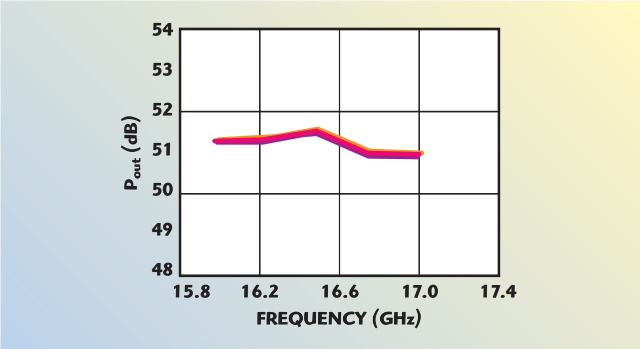


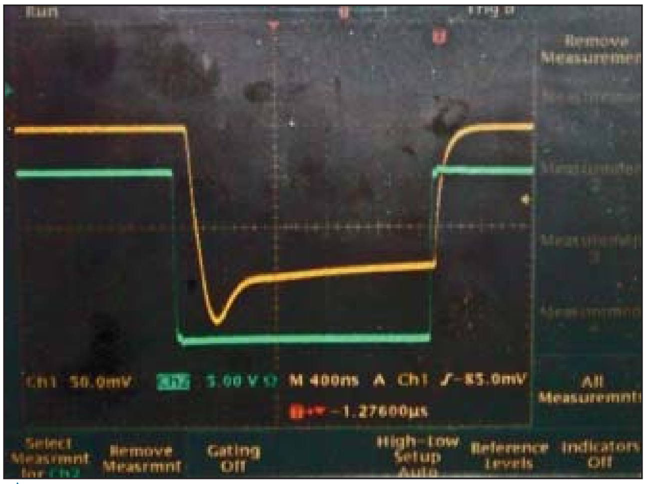
Fig. 1 MPC8-1520 SSPA output power vs. frequency.

The specifications of the MPC8-1520 solid-state power amplifier are given in ***Table 1***.

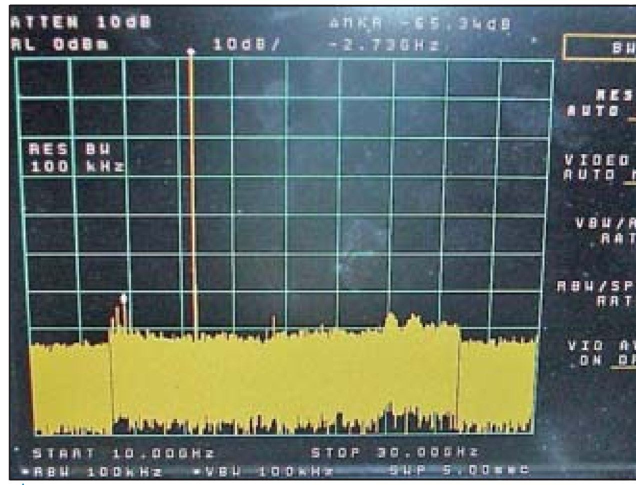
The output power of the amplifier as a function of frequency is shown in ***Figure 1***. An example of the pulse rise and fall times is shown in ***Figure 2***. The level of non-harmonics related spurious is less than –60 dBc, as shown in ***Figure 3***.

MONITOR AND CONTROL FUNCTIONS

The monitor and control processor interfaces with a host computer. All voltage, current, power and temperature sensors are read through this interface, including forward and reverse power monitoring. The RS422 port



▲ *Fig. 2 Pulse response of the MPC8-1520 amplifier.*



▲ *Fig. 3 Spurious levels in the output power of the MPC8-1520 SSPA.*

communications run at 9600 or 38400 baud, 8-bit word, no parity and 1 stop bit.

CONFIGURATION OPTIONS

The MPC8-1520 solid-state power amplifier comes in two configurations. An indoor, module-based component, as shown in the photograph, and an outdoor, system-based enclosure with integrated cooling. The specifications for the outdoor unit are identical to the ones shown for the indoor unit, except for the DC power parameters, which are +28 V DC, 235 W at 20 percent duty cycle, and the operating temperature range, which is –40° to +50°C ambient temperature. The size of the outdoor unit is 10.3" × 7.5" × 5.5" and its weight is less than 15 lbs. Additional options include AC/DC power supplies.

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RS No. 304

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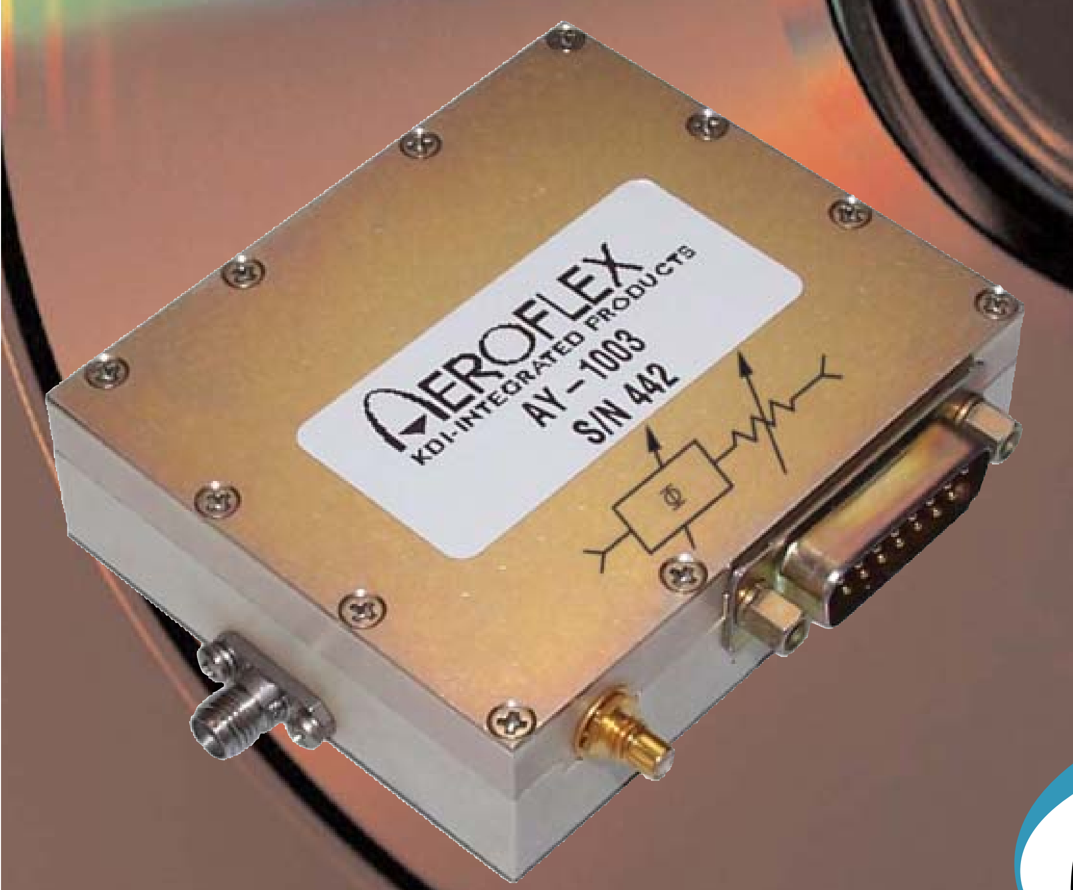

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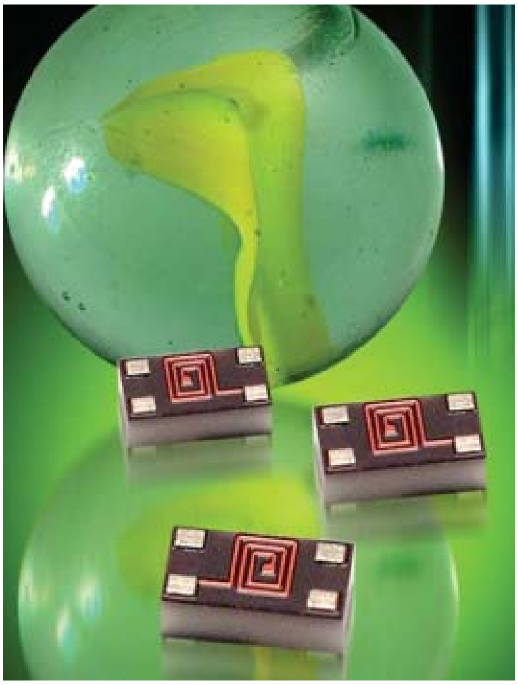


A HARMONIC LOW PASS FILTER WITH A LEAD-FREE LGA TERMINATION

AVX Corp. is now offering the LP0603 harmonic low pass filter with lead-free land grid array (LGA) termination. The ITF LGA low pass filter is based on thin-film technology and provides a miniature part with excellent high frequency performance and rugged construction for reliable automatic assembly. The ITF low pass filters are offered in a variety of frequency bands compatible with various types of high frequency wireless systems. The filter's design also results in an inherent low profile, self-alignment during reflow, excellent solderability and better heat dissipation.

APPLICATIONS

Harmonic low pass filters are used to reduce and eliminate unwanted harmonics in telecommunications applications such as mobile communications, satellite TV receivers, GPS, vehicle location systems, wireless LANs and WiMAX. While discrete components such as capacitors and inductors can be used to meet the same ends, it would be necessary to use a minimum of four or five of those components to achieve the same results as one LP0603. For handheld and portable equipment, such as cell phones, this can take up significant space, which is at a premium. The LP0603 takes the place of multiple components and also offers electrical performance unmatched in any other low pass filter available in a similar package size. The LP0603 would be found on the outside of a cell phone's power amplifier to eliminate unwanted harmonics generated by the amplifier. For example, in Wi-Fi applications, where a piece of equipment might be operating at 900 MHz, the amplifier will normally generate unwanted second and third harmonics. The LP0603 filters out the unwanted harmonics to prevent interference in the operation of the components and therefore the end product.

ELECTRICAL CHARACTERISTICS

The AVX's LP0603 has a lower insertion loss than most other low pass filters on the market, while the attenuation levels are higher. At a

AVX CORP.
Myrtle Beach, SC

Visit http://mwj.ims.ca/5545-135

TABLE I

ELECTRICAL CHARACTERISTICS OF THE LP0603 HARMONIC LOW PASS FILTERS

Model No.	Frequency Band (MHz)	I.L. (dB) typ.	I.L. (dB) max.	VSWR (max)	Attenuation (dB) $2F_0$	Attenuation (dB) $3F_0$
LP0603A0902ANTR	890 to 915	0.35	0.50	1.4	25	14
LP0603A0947ANTR	935 to 960	0.35	0.50	1.4	25	17
LP0603A1747ANTR	1710 to 1785	0.35	0.50	1.4	25	17
LP0603A1842ANTR	1805 to 1880	0.35	0.50	1.4	27	15
LP0603A1880ANTR	1840 to 1920	0.35	0.50	1.4	25	17
LP0603A1950ANTR	1920 to 1980	0.35	0.50	1.4	27	15
LP0603A2140ANTR	2110 to 2170	0.35	0.50	1.4	27	17
LP0603A2442ANTR	2412 to 2472	0.35	0.50	1.4	25	17

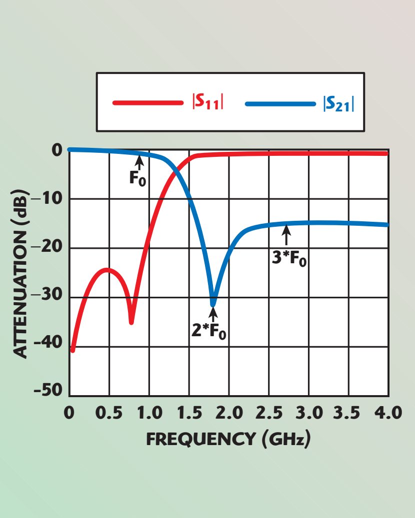
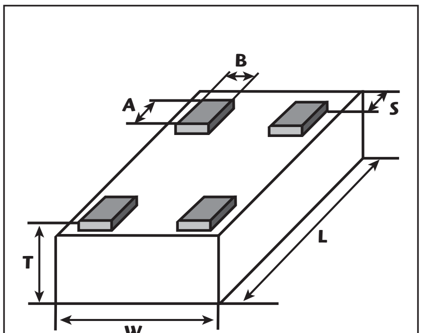


Fig. 1 Electrical characteristics of the LP0603A0902ANTR filter.



L	1.6±0.1 (0.063±0.004)	A	0.25±0.05 (0.01±0.002)
W	0.84±0.1 (0.033±0.004)	B	0.20±0.05 (0.008±0.002)
T	0.60±0.1 (0.024±0.004)	S	0.05±0.05 (0.002±0.002)

Dimensions in mm.

Fig. 2 LP0603 dimensions.

frequency of 902 GHz, for example, the LP0603 shows an insertion loss of –0.35 dB and attenuation at the second harmonic of –25 dB. ***Table 1*** shows the electrical characteristics of the available LP0603 harmonic low pass filters that are guaranteed over the –40° to +85°C operating temperature range. ***Figure 1*** shows the typical frequency response for a LP0603A0902ANTR harmonic low pass filter.

MECHANICAL CHARACTERISTICS

In using its thin-film technology in constructing the filter, AVX uses pure metals and controls the thickness and physical dimensions of the layers in a way that is not possible with MLCC technology and that allows for a much higher level of uniformity and consistency from component to component. The land grid array offers several advantages: an inherent low profile packaging, self-alignment during reflow soldering, excellent solderability, low parasitics and better heat dissipation. The package dimensions are shown in ***Figure 2***. The nickel/lead-free solder coating is compatible with automatic soldering technologies such as reflow, wave soldering, vapor phase and manual. The finish parts are 100 percent tested for electrical parameters and visual characteristics. Each production lot is evaluated on a sample basis for static humidity, at 85°C, 85 percent RH, for 160 hours, and for endurance at 125°C, IR for four hours.

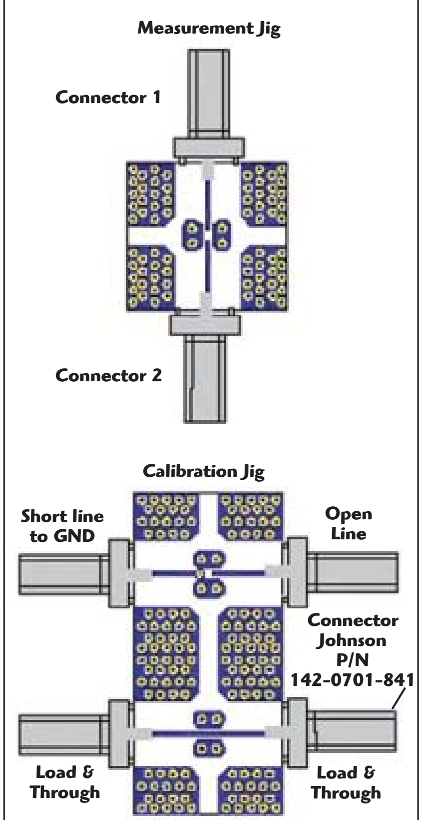


Fig. 3 Layout of the test jigs.

LOW PASS FILTER TEST JIGS

There are specific jigs designed for testing the LP0603 LGA low pass filters using a vector network analyzer. They consist of a dielectric substrate, having 50 Ω microstrips as conducting lines and a bottom ground plane located at a distance of 0.127 mm from the microstrips. The substrate used is Neltec's NH9338ST0127C1BC (or similar). The connectors are SMA females. Both a measurement jig and a calibration jig are provided. The calibration jig is designed for a full two-port calibration and consists of an open line, a shorted line and a through line. The load calibration can be done by a 50 Ω SMA termination. ***Figure 3*** shows the layout of the jigs.

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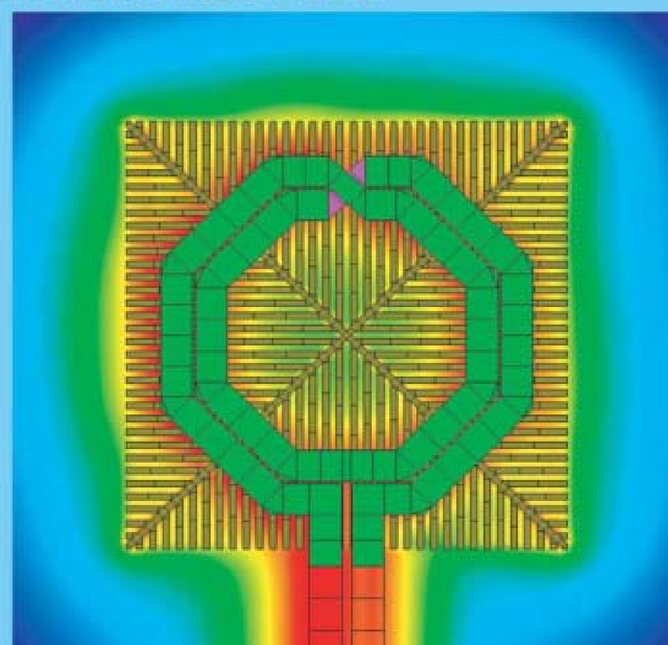
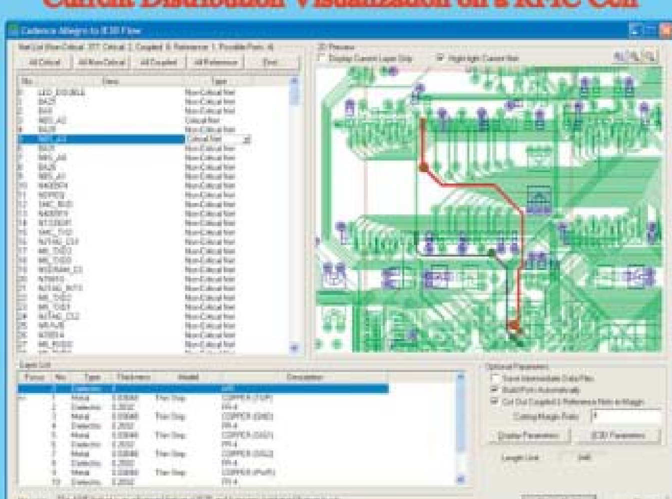
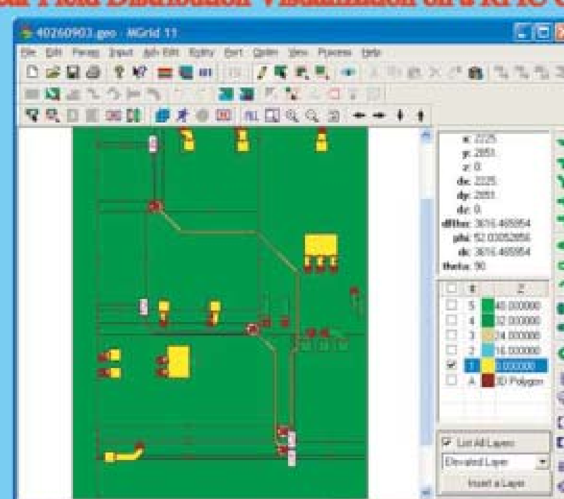


Near Field Distribution Visualization on a RFIC Cell



Critical Nets and Reference Nets Defined on a PCB Layout



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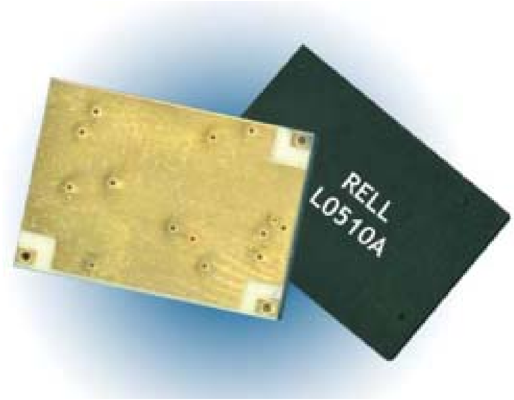

A LOW NOISE AMPLIFIER FAMILY FOR WIRELESS SYSTEM APPLICATIONS

A new product family of low noise amplifiers (LNA) has been introduced that is specifically designed to increase a wireless system's performance and reliability. The Super LNA Series LNAs are optimized to meet the design goal of achieving the lowest noise figure and best return loss simultaneously over a very wide bandwidth. These products reduce production cost and increase reliability without compromising system performance.

The Super LNA Series amplifiers cover a frequency range of 200 to 2600 MHz, which includes the VHF, UHF, CDMA, GSM, GPS, DCS, PCS, WCDMA, UMTS and MMDS bands. Typical applications include receivers, wireless data communications, and test and measurement. ***Table 1*** lists the current models that make up the RLA series, with custom models readily available to meet customer requirements.

KEY FEATURES

Among the key features that make these amplifiers stand out are their very wide bandwidth, making them excellent for broadband RF and microwave systems, and test and measurement applications. The new amplifiers also feature an ultra-low noise figure that provides a wide dynamic range. In addition, the LNAs exhibit superior gain flatness and are unconditionally stable within their specified frequency bands making them ideal for wideband applications. All matching and decoupling components are designed in, thus there is no need for external tuning or matching, thereby reducing overall component costs and rendering the device specification the true performance of the amplifier. The amplifiers are internally matched to 50 Ω and present low input and output VSWRs. Finally, they are packaged in a lead-free surface-mount package that is well suited to volume production.

A TYPICAL EXAMPLE

As an example, the model RLAS1722A is an ultra-low noise figure, wideband amplifier

RICHARDSON ELECTRONICS LTD.
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4 **PRINCIPAL JOB FUNCTION:**
Select one category from the following list that most closely describes your principal job function.

DESIGN & DEVELOPMENT ENGINEERING
03 ☐ Engineering
02 ☐ Management

ENGINEERING SERVICES
(evaluation, QC, reliability, standards, test)
05 ☐ Engineering
04 ☐ Management

01 ☐ GENERAL AND/OR CORPORATE MANAGEMENT

RESEARCH & DEVELOPMENT
07 ☐ Engineering
06 ☐ Management

MANUFACTURING & PRODUCTION
09 ☐ Engineering
08 ☐ Management/Supervision

10 ☐ ENGINEERING SUPPORT
(draftsman, lab assistant, technician)

11 ☐ PURCHASING & PROCUREMENT

12 ☐ APPLICATIONS ENGINEERING, SALES AND MARKETING

13 ☐ EDUCATORS

14 ☐ OTHER PERSONNEL *(explain)*

5 **PRIMARY END PRODUCT OR SERVICE**
Select a primary end product (or service performed) from the following list that most closely describes the end product of the company in which you work.

06 ☐ Communications Systems & Equipment
17 ☐ Cellular Systems & Equipment
26 ☐ WLAN, WiFi
10 ☐ Test & Measurement Equipment
27 ☐ Semiconductor, RFICs, MMICs, etc.
11 ☐ Active Components (including Power Supplies, Subsystems)
12 ☐ Passive Components (including Antennas, Devices, Subsystems)
16 ☐ Government/Military: Research, Design & Engineering
01 ☐ Radar Systems
04 ☐ Navigation, Telemetry Systems, GPS
08 ☐ Data Transmission, Computer Systems
28 ☐ Software Development
05 ☐ Electronic Warfare Systems
03 ☐ Ground Support Equipment, Aircraft/Missile
02 ☐ Weapons Control, Ordnance, Fusing Systems
13 ☐ Materials, Hardware
15 ☐ Industrial/Academic/R&D Laboratories, Consultants
14 ☐ Industrial/Commercial Control, Processing Equipment
29 ☐ Medical Equipment
20 ☐ Consumer Electronics
07 ☐ CATV Broadcast Systems
18 ☐ Automotives/transportation
19 ☐ Security/identification
09 ☐ Laser/Electro-Optical Systems, Equipment
21 ☐ Other *(please specify)*

USER
22 ☐ Government/Military
23 ☐ Industrial/Commercial
24 ☐ Technical Library
25 ☐ Other *(please specify)*

6 **YOUR WORK IS PRIMARILY:**
(check all that apply)
01 ☐ Below 1 GHz
02 ☐ 1–8 GHz
03 ☐ 9–18 GHz
04 ☐ 19–26.5 GHz
05 ☐ 26.6–40 GHz
06 ☐ Above 40 GHz
07 Other *(please specify)*

7 **PLEASE ESTIMATE THE ANNUAL VALUE OF PURCHASES THAT YOU INFLUENCE.**
06 ☐ $500,000 or more
05 ☐ $300,000 to $499,999
04 ☐ $100,000 to $299,999
03 ☐ $50,000 to $99,999
02 ☐ $10,000 to $49,999
01 ☐ less than $10,000

8 **IS YOUR WORK PRIMARILY:**
01 ☐ Commercial
02 ☐ Military

9 **WHICH OF THE FOLLOWING PRODUCTS DO YOU RECOMMEND, SUPPORT OR AUTHORIZE TO PURCHASE**
(check all that apply)

AMPLIFIERS AND OSCILLATORS
01 ☐ Amplifiers (Low Noise)
02 ☐ Amplifiers (Power)
03 ☐ Tubes or Tube Amplifiers
04 ☐ Solid State Oscillators

07 ☐ ANTENNAS & ACCESSORIES
13 ☐ CAD SOFTWARE OR SERVICES

CABLE AND CONNECTORS
16 ☐ General Purpose
17 ☐ Precision or Laboratory

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21 ☐ Switches (Solid State)
22 ☐ Attenuators & Phase Shifters

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27 ☐ Attenuators & Terminations
28 ☐ Filters
29 ☐ Resistors, Capacitors & Inductors
30 ☐ Isolators & Circulators

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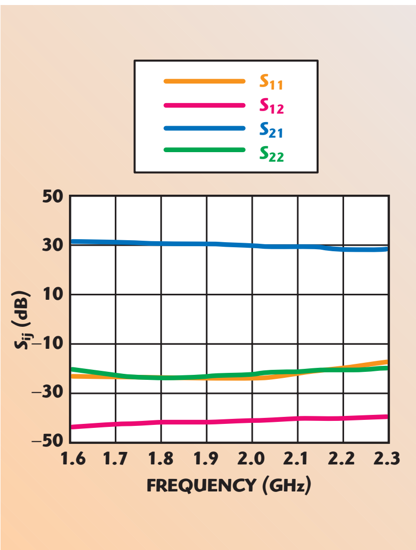
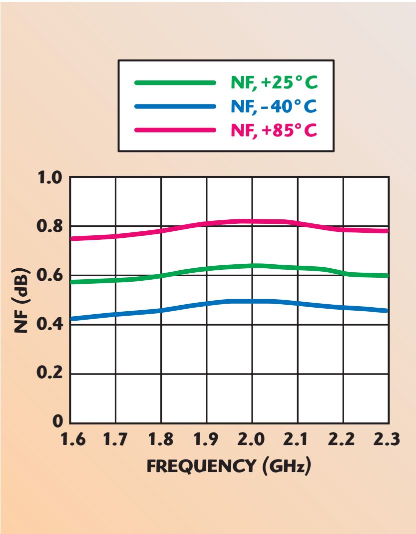
TABLE I

THE SUPER LNA SERIES

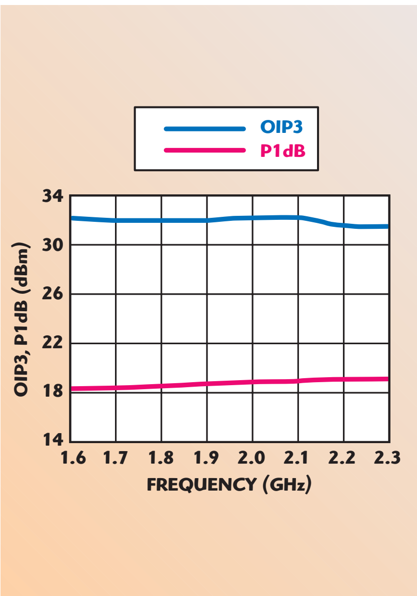
Part No.	Frequency (MHz)	Gain (dB)	Gain Flatness (dB)	NF (dB)	VSWR	P1dB (dBm)	DC (V)	DC (mA)	Package Size (")
RLAS0205A	200 ~ 550	43	±0.75	0.50	1.18	18	5	90	0.5 × 0.35 × 0.08
RLAS0510A	500 ~ 1000	38	±1.2	0.40	1.22	19	5	90	0.5 × 0.35 × 0.08
RLAS1216A	1200 ~ 1600	31	±1.0	0.50	1.17	10	3.3	40	0.5 × 0.35 × 0.08
RLAS1722A	1700 ~ 2200	30	±1.0	0.55	1.17	19	5	90	0.5 × 0.35 × 0.08
RLAS2026A	2000 ~ 2600	26	±0.75	0.60	1.25	12.5	5	55	0.5 × 0.35 × 0.08



▲ *Fig. 1 Small-signal performance at 25°C.*



▲ *Fig. 2 Noise figure performance at full temperature.*



▲ *Fig. 3 P1dB and OIP3 at room temperature.*

covering the 1700 to 2200 MHz frequency range. The amplifier offers a typical 0.55 dB noise figure, 20 dB input and output return losses, and 30 dB of gain. Its output P1dB is 19 dBm and its output third-order intercept (OIP3) is 32 dBm over the DCS, PCS, UMTS and 3G frequency bands. It is unconditionally stable and features a 50 Ω input and output impedance. ***Figures 1***, ***2*** and ***3*** show the RLAS1722A amplifier's S-parameters, noise figure and output characteristics, respectively, over its full operating frequency range.

The amplifier is powered by a single +5 VDC supply and draws typically 90 mA. There are built-in DC blocks at the input and output ports and built-in temperature compensation circuitry. It is supplied in a 0.5" × 0.35" × 0.08" SMT package and boasts a > 600,000 hour MTBF. The amplifier is designed to meet MIL-STD-202, MIL-STD-883 and MIL-STD-810F. A connectorized version is also available.

The other amplifier models in the series feature similar performance characteristics within their respective operating frequency ranges. Although other manufacturers are producing similar performance amplifiers, the new RLAS series amplifiers offer a distinct cost advantage over other competitive offerings.

CONCLUSION

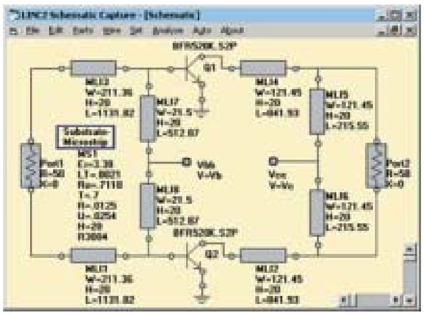
A new series of Super Low Noise Amplifiers has been described that covers the 200 to 2600 MHz frequency range and offers outstanding performance over a very wide bandwidth in an unconditionally stable, low cost SMT package. These new amplifiers are ideal for wireless system applications, including receivers, data communications, and test and measurement.

Richardson Electronics Ltd.,
LaFox, IL
(800) 737-6937,
(630) 208-3637,
e-mail: rwc@rell.com,
www.rfwireless.rell.com.

RS No. 305


Mini-Circuits
ZRL
SERIES
150MHz-3.5GHz

SOFTWARE UPDATE



DESIGN SOFTWARE

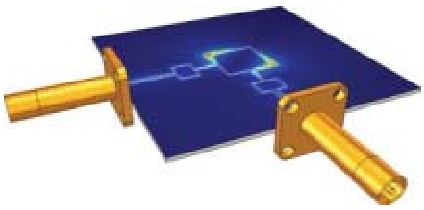
The LINC2 software combines high performance RF and microwave circuit design, synthesis, simulation and optimization into a single integrated program. This software provides a suite of design tools for the exact synthesis of a wide variety of active and passive circuits. LINC2's comprehensive amplifier synthesis tool produces circuit schematics for single and multi-stage amplifiers, low noise amplifiers, balanced amplifiers and differential (push-pull) amplifiers. The LINC2 impedance matching tool provides lumped and distributed circuit topologies for both broadband and narrowband applications, including balanced and unbalanced configurations. LINC2 filter synthesis automatically designs both differential and single-ended filters.

Applied Computational Sciences,
Escondido, CA (760) 612-6988, www.appliedmicrowave.com.
RS No. 311

EMISSIONS SOFTWARE

The model SW1006 is the latest version of the company's radiated susceptibility, conducted immunity and precompliance emissions software that automatically performs both calibration and immunity testing in full compliance with IEC 61000-4-3, 4-6, MIL-STD 461/462 RS103, CS114 and RTCA/DO160 Section 20 specifications. The software also supplies the user with selectable test parameters and a threshold mode for pre-compliance investigation of equipment susceptibility, as well as closed loop leveling. Pre-compliance emission testing can be done with the use of a spectrum analyzer and either a pre-amp or LISN.

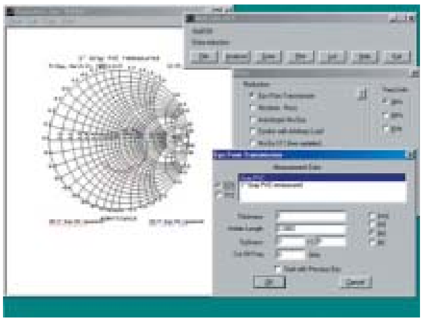
AR Worldwide • RF/Microwave Instrumentation,
Souderton, PA (215) 723-8181, www.ar-worldwide.com.
RS No. 312



MODELING AND SIMULATION

Version 3.2 of COMSOL Multiphysics has new features designed to boost productivity throughout the entire modeling and simulation process. The software now reads geometry files created with all major CAD packages. It introduces COMSOL Script, a standalone product featuring command-line modeling. The graphical user interface encourages the use of a consistent system of engineering units, and a moving-mesh feature allows a model to simulate moving parts and parametric geometries. For the easy importing of CAD drawings for modeling, a suite of optional CAD-import modules read a wide range of CAD and mesh file formats. Also, improved solvers handle models with millions of degrees of freedom to calculate the answers faster than before.

COMSOL AB,
Stockholm, Sweden +46-8412-9500, www.comsol.com.
RS No. 313

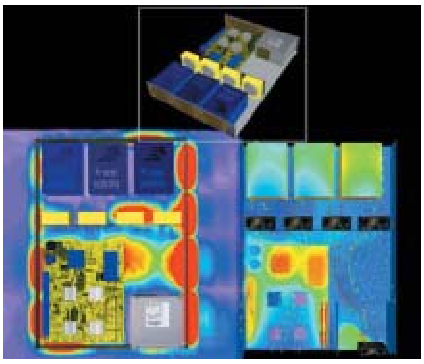


SOFTWARE PACKAGE

The MU-EPSLN™ is a versatile software package for Microsoft Windows and Macintosh OS X that performs all of the major functions associated with collecting and processing data to determine material constitutive properties, such as permeability and permittivity. The program controls common network analyzers made by Agilent, Anritsu, and Rohde & Schwarz, and the program leads the user through calibration and measurement steps. A variety of data processing options are available to determine mu, epsilon and other parameters. A time domain option is available, which allows the S-parameters to be gated.

Damaskos Inc.,
Concordville, PA (610) 358-0200, www.damaskosinc.com.
RS No. 314

SOFTWARE UPDATE



INTEGRATED ANALYSIS ENVIRONMENT

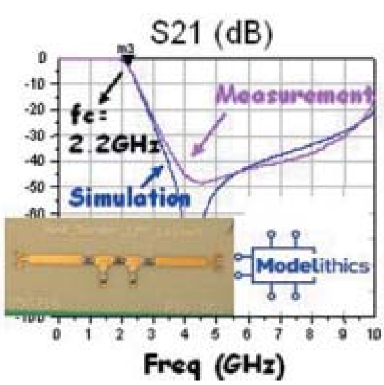
Flotherm and Flo/EMC Version 6 of the company's integrated analysis environment for physical design of electronics features improved communication between thermal and electromagnetic compatibility simulation. The latest Flotherm and Flo/EMC software packages further reduce the time required for integrated simulation by automatically generating two different meshes, one optimized for thermal simulation and one for EMC simulation. Version 6 makes it possible for users to optimize thermal and EMC meshes independently. With the recent release, users can now access the Web-based "SmartParts3D" library directly, utilizing compact models and reducing the time required for modeling and analysis.

Flomerics Inc.,
Marlborough, MA (508) 357-2012, www.flomerics.com.
RS No. 315

DEVICE CHARACTERIZATION SOFTWARE

The release of ATS 4.00 device characterization software features over 50 new capabilities, including an advanced multi-dimensional sweep plan, an electrothermal memory characterization tool and migration of ATN noise parameter extraction algorithms. In addition, ATS 4.00 was subject to enterprise-class QA and rigorous regression procedures to ensure stable and reliable operation.

Maury Microwave Corp.,
Ontario, CA (909) 987-4715, www.maurymw.com.
RS No. 316



FILTER DESIGN KITS

This filter simulation and design kit includes complete know-how for common low pass, high pass and bandpass filters. Fully customizable Global Models™ for all components, and layouts for EDA software are included. These generate highly accurate simulations that enable first-pass design success that can be validated using the pre-assembled sample filters included. Simply modify the design goals and create similar filters using blank boards and extra part samples included to meet specific needs.

Modelithics Inc.,
Tampa, FL (813) 866-6335, www.modelithics.com.
RS No. 317

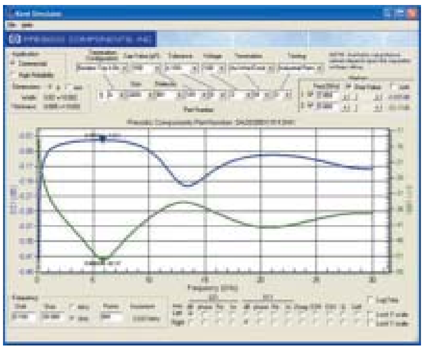
GRAPHICAL DESIGN AND DEVELOPMENT SOFTWARE

LabVIEW 8 is a graphical design and development environment for custom test, measurement and automation applications. The new release addresses accelerating product development, globalization of design and manufacturing, increasing design complexity, and other RF and communications industry challenges. LabVIEW 8 introduces distributed intelligence, a suite of powerful features for engineers to design, distribute and synchronize intelligent custom devices and systems. The new release streamlines these jobs with shared variable communication technology and tightly integrated target management.

National Instruments Corp.,
Austin, TX (888) 280-7645, www.ni.com/labview.
RS No. 318


USAF
U.S. AIR FORCE
OK

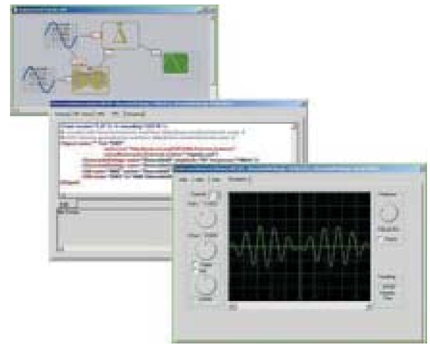
Software Update



CAPACITOR MODELING

The KENT SIMULATOR is designed for single layer capacitor modeling. The KENT SIMULATOR is free and can be downloaded from the company's Web site. Once the download is complete, users simply double click on the .exe file to install the simulator. The KENT SIMULATOR obtains commonly needed RF capacitor parameters for patented buried single layer ceramic capacitors. All device parameters are derived from a series transmission line model developed by Dr. Gordon Kent. The graph presents parameter data for a selected capacitor part number for a user defined frequency range up to 30 GHz. Data saved in the S2P format is compatible with and easily imported into many microwave circuit design programs.

Presidio Components Inc.,
San Diego, CA (858) 578-9390, www.presidiocomponents.com.
RS No. 319



IEEE 1641 TEST SOFTWARE

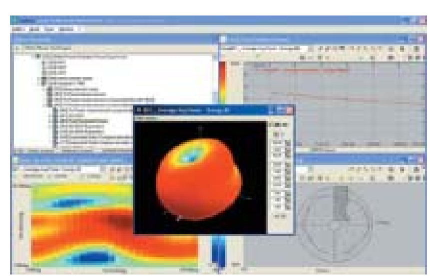
The user-friendly software tools called newWaveX have been designed for design and test engineers. The three products available are: a downloadable 30-day evaluation version; newWaveX Lite, a version for signal design and simulation and XML export of signal definitions; and the full version of newWaveX, for comprehensive test program development and compliance verification that includes import/export of XML definitions. Significant is that newWaveX encapsulates IEEE 1641 (Signal and Test Definition), a new standard which expresses tests in terms of signals applied to the unit under test. It enables test development through an easy to use drag and drop graphical development environment, and real-time test simulation through DirectX technology.

Racal Instruments Group,
Wimborne, Dorset, UK +44 (0)1202 872800,
www.racalinstrumentsgroup.co.uk.
RS No. 320

ASSEMBLY BUILDER PROGRAM

This interactive "High Speed Cable Assembly Builder" program at www.samtec.com/cable_builder makes it as easy as a "click" to create valid Samtec High Data Rate Cable part numbers and order samples. This program allows specification of assemblies on 0.5 mm, 0.635 mm, 0.8 mm, 0.050" (1.27 mm), 2 mm and 0.100" pitch. Standard options such as cable length, wiring configuration and screw features for rugged applications are easily specified with the program.

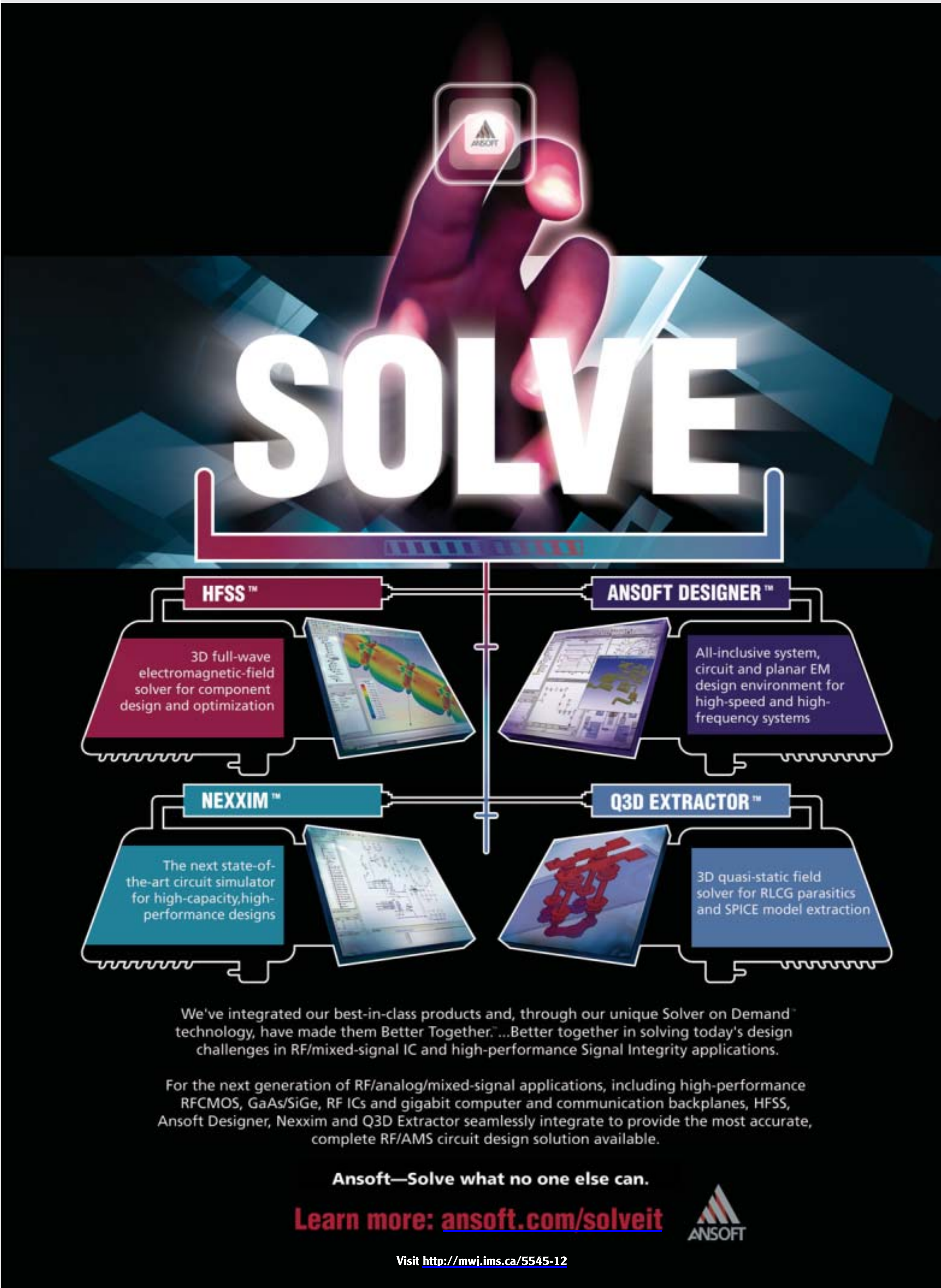
Samtec Inc.,
New Albany, IN (800) 726-8329, www.samtec.com.
RS No. 321



ANTENNA TEST RANGE SOFTWARE

SatEnv is a powerful software package for acquisition, processing and visualization of measurement data from antenna test ranges. The software provides unlimited possibilities for data acquisition set up, and includes a full set of analysis tools. Advanced macro and scripting capabilities provide a high level of automation for data processing, data export and plot generation. Equipment drivers are easily added, making SatEnv an attractive upgrade for older test ranges with outdated control software. Available plug-in computation modules include near-field to far-field transforms, SAR evaluation, spherical and planar back/forward near-field propagation, and ray tracing algorithms.

Satimo,
Courtaboeuf, France +33 1 6929 0247, www.satimo.com.
RS No. 322


SOLVE
HFSS™
3D full-wave electromagnetic-field solver for component design and optimization
ANSOFT DESIGNER™
All-inclusive system, circuit and planar EM design environment for high-speed and high-frequency systems
NEXXIM™
The next state-of-the-art circuit simulator for high-capacity,high-performance designs
Q3D EXTRACTOR™
3D quasi-static field solver for RLCG parasitics and SPICE model extraction
We've integrated our best-in-class products and, through our unique Solver on Demand™ technology, have made them Better Together.™...Better together in solving today's design challenges in RF/mixed-signal IC and high-performance Signal Integrity applications.
For the next generation of RF/analog/mixed-signal applications, including high-performance RFCMOS, GaAs/SiGe, RF ICs and gigabit computer and communication backplanes, HFSS, Ansoft Designer, Nexxim and Q3D Extractor seamlessly integrate to provide the most accurate, complete RF/AMS circuit design solution available.
Ansoft—Solve what no one else can.
Learn more: ansoft.com/solveit
ANSOFT
Visit http://mwj.ims.ca/5545-12

SOFTWARE UPDATE

CLUSTER COMPUTING SOLUTION

The emCluster™ computing solution is available with the 10.53 release of Sonnet Suites Professional and reduces analysis time and shortens design cycles. This newly added emCluster module can split a project and assign individual analysis frequencies to available resources across an IT cluster environment significantly reducing the completion time of the EM analysis to just a fraction of what it would have taken with a single computing resource. In addition to speeding up the analysis time, emCluster intelligently schedules analysis jobs to minimize idle time of valuable computing resources.

Sonnet Software Inc.,
North Syracuse, NY (315) 453-3096, www.sonnetsoftware.com.
RS No. 323

CARRIER-TO-INTERFERENCE ANALYSIS

With this release, the accompanying Log View™ software utility can now display both the radio's "carrier" sweep as well as the "interference" sweep simultaneously. Markers can be set to the carrier's modulation bandwidth and the software will automatically compute the "total channel power" of both the carrier and the interference, normalized to the carrier's bandwidth, and calculate the carrier-to-interference figure, in dBm, to indicate the real world fade margin at the site. Because of its simple interface, the Analyze-R™ can be used by non-technically trained personnel and is designed for highly portable field use.

XL Microwave Inc.,
Oakland, CA (510) 428-9488, www.xlmicrowave.com.
RS No. 324

2006 IEEE Radio Frequency Integrated Circuits Symposium San Francisco, California June 11-13, 2006

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Electronic Paper Submission/Communication: Technical papers must be submitted in an electronic form via the RFIC-2006 web site **www.rfic2006.org** - **Hard Copies are not accepted.**

Technical Areas: Papers are solicited describing original work in RFIC design, system engineering, system simulation, design methodology, RFIC circuits, fabrication, testing and packaging to support RF applications in areas such as, but not limited to:

- **Cellular System IC's and Architectures:** GSM, EDGE, TDMA, CDMA, 3G, WCDMA, GPS
- **Wireless Data System IC's and Architectures:** WLAN, Bluetooth, 802.1x, Telemetry, RFID
- **Wide Band Communication System IC's and Architectures:** UWB, MMDS, CATV, TV Tuners
- **Optical System IC's and Architectures:** OC-48, OC-192, OC-768, Gigabit Transceivers
- **Small-Signal Circuits:** LNA's, Mixer's, VGA's, Active Filters, Modulators
- **Large-Signal Circuits:** Power Amplifiers, Drivers, Advanced TX circuits
- **Frequency Generation Circuits:** VCO's, PLL's, Synthesizers
- **RFIC & Device Technology:** IC Technologies, Packaging, Modules, RF Test and Characterization
- **RFIC Modeling and CAD:** Device and behavioral modeling, Design Methodology

Technical Format: The technical sessions will be held for three days from Sunday through Tuesday. Workshops will be on Sunday. Several invited sessions and talks will take place during the conference.

Microwave Week 2006: The RFIC 2006 will be in conjunction with the IEEE MTT-S International Microwave Symposium (IMS). Microwave Week 2006 will continue with the International Microwave Symposium and Exhibition, and the Microwave Historical Exhibit.

Guest Program: Northern California is a wonderful place to visit and the weather in June is quite pleasant. See the Golden Gate Bridge, Fisherman's Wharf, China Town, Golden Gate Park, Coit Tower, Muir Woods, the Wine Country, the Pacific coast, Alcatraz and Treasure Islands and many other sites. Many of these will be destinations in our guest program.

Electronic Submission Deadlines

Technical Paper Summaries in PDF format:	**2 January 2006**
Final Manuscripts for the Digest and CD-ROM:	**6 March 2006**
All submissions must be made through the RFIC2006 portal:	**www.rfic2006.org**
ALL SUBMISSIONS MUST BE IN PDF FORM	**Hard copies not accepted**

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http://www.rfic2006.org

SSCS EDS

NEW WAVES: Passive and Control Components

Blocking Capacitor

The GZ capacitor series is specifically designed to address DC blocking issues from 15 kHz to 40 GHz in optical transceivers. The technology incorporates small footprints to conserve board space and is designed to match 0.015" and 0.020" microstrip widths. The assemblies combine high capacitance Ni-Au terminated MLCs for low frequency coverage. Specially configured maxi dielectric single layer ceramics facilitate conventional surface mounting. Most applications will experience resonance-free performance of < 0.5 dB through at least 26.5 GHz. Price: $1.00 to $4.00, depending on quantity. Delivery: four to six weeks.

AVX Corp.,
Myrtle Beach, SC (843) 448-9411,
www.avx.com.

RS No. 216

5 W Power Divider/Combiner

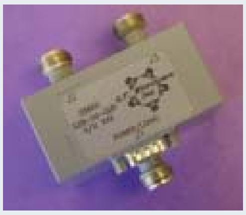
The model series 152-085-XXX of 50 Ω power divider/combiners are available in two-, three- and four-way configurations. These 5 W average (at the common port) power divider/combiners are for telecommunication infrastructure and test system applications. The two- and four-way models operate at a frequency range from 800 to 2700 MHz and the three-way model operates from 800 to 2000 MHz. RF connectors are N female. Delivery: available from stock.

BroadWave Technologies Inc.,
Franklin, IN (317) 346-6101,
www.broadwavetech.com.

RS No. 217

PIN Diode Switch

The model S2B-29-0JB is a SP2T high power reflective PIN diode switch that operates from 20 to 1000 MHz. Across the entire band, VSWR is less than 1.3, insertion loss is less than 1 dB and the isolation is greater than 40 dB. The switch is capable of handling up to 100 W CW of RF power with cold switching. With TTL compatible logic, the switching speed is less than 5 µs.



G.T. Microwave Inc.,
Randolph, NJ (973) 361-5700,
www.gtmicrowave.com.

RS No. 218

Diode Detectors

The model DZM020BB is a zero-bias Schottky diode detector that operates from 100 kHz to 2 GHz. These diodes feature a flat frequency response of 0.1 per 100 kHz and matched input for good VSWR of 1.3. This model is ideal for lab testing, power monitoring and level circuits. It offers a maximum input power of 200 mW CW and is available in either negative or positive output polarity.

Herotek Inc.,
San Jose, CA (408) 941-8399,
www.herotek.com.

RS No. 219

Double-balanced Mixers

The models HMC260LC3B, HMC292LC3B and HMC329LC3B are passive MMIC double-balanced mixers that exhibit conversion loss as low as 7.5 dB, and input IP3 as high as +21 dBm, while on-chip balun structures enable LO-to-IF and LO-to-RF isolations as high as 40 dB. Each mixer operates with an LO drive level as low as +9 dBm, and offers wide IF bandwidth of DC to 8 GHz. These 3×3 mm SMT packaged mixers are ideal for upconverter and downconverter applications in VSAT, microwave radio, test equipment and industrial/automotive sensor applications from 14 to 32 GHz.

Hittite Microwave Corp.,
Chelmsford, MA (978) 250-3343,
www.hittite.com.

RS No. 220

High Power Notch Combiner

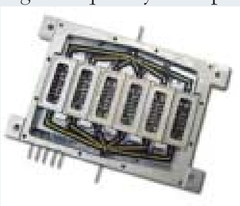
This joint tactical radio system (JTRS) is a broadband data, voice and video acquisition system designed to provide seamless real-time communications among war fighters. This JTRS is a low loss high power notch combiner product designed to combine various transceivers to one antenna, while mitigating RF interferences with joint tactical information distribution system (JTIDS) platforms. This module is a three-port device. The low channel passes RF energy from DC to 920 MHz and the high channel passes from 1270 to 2000 MHz, while JTIDS signals are attenuated by 45 dB.

K&L Microwave,
Salisbury, MD (410) 749-2424,
www.klmicrowave.com.

RS No. 221

Low Profile Switch Bank

The model 6IFA-7250/9750-MP is a six channel high frequency low profile switch bank. This unit utilizes a +5 V power supply at 75 mA maximum. The switching speed is 50 ns and the filter's bandwidths are < 10 percent. Typical VSWR is 2.0 and the maximum insertion loss is 10 dB at band centers. The unit features a 2.6 × 2.0 × 0.390 tall package with 2-56 heli-coil mounting.



Lorch Microwave,
Salisbury, MD (410) 860-5100,
www.lorch.com.

RS No. 222

Electromechanical Switches

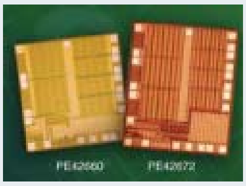
These electromechanical switches utilize a non-molded cavity design that achieves a high performance. The switches are available in a variety of mechanical and electrical configurations tailored to meet a system's requirements. The DPDT transfer switch offers a miniature design with SMA connectors and is available up to 18 GHz. Options available include latching or failsafe actuator, suppression diodes and indicators. Delivery: one to two weeks ARO.

Microwave Communications Laboratories Inc.,
Saint Petersburg, FL (727) 344-6254,
www.mcli.com.

RS No. 223

RF Switches

The model PE42672 SP7T and model PE42660 SP6T are RF switches that have been released on the HarP-enhanced UltraCMOS process and are designed for quad-band GSM and GSM/WCDMA handset applications. The former is said to be the world's first monolithic SP7T switch with an on-board CMOS decoder. This highly integrated solution simplifies and lowers the cost of RF designs by reducing the overall part count by as many as six devices and 13 wire bonds. The PE42660 switch is drop-in compatible with the PE4263 GSM handset switch that was released at the end of 2004. Both devices have good RF performance levels offering exceptional linearity (PE42672: $2f_o$ –85 dBc and $3f_o$ –79 dBc; PE42660: $2f_o$ –88 dBc and $3f_o$ –85 dBc); IP3 better than +70 dBm; 1.5 KV ESD tolerance; 2.75 V operating voltage and ultra-low power consumption.



Peregrine Semiconductor,
San Diego, CA (858) 731-9400,
www.psemi.com.

RS No. 224

Band Reject Filter

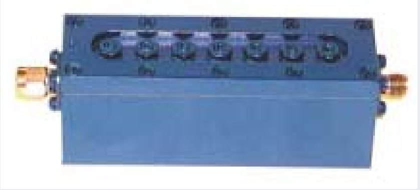
The model 5BR5R65G-2R5G-CD-SFF is a band reject filter that offers an extended pass-band performance up to four times the notch

frequency. The suspended substrate technology utilized allows these filters to perform with good extended frequency ranges. This model offers greater than 60 dBc rejection from 5.4 to 5.9 GHz while maintaining less than 2 dB of insertion loss from DC to 4.5 GHz and 7 to 20 GHz.
Planar Filter Co.,
Frederick, MD (301) 662-5019, www.planarfilter.com.

RS No. 225

Reallocated 2 GHz BAS Filters

This line of filters is intended for the reallocated 2 GHz BAS spectrum. These units will help users to upgrade systems from the old 17 MHz bandwidth standard to the new 12 MHz bandwidth standard. These units feature insertion loss of less than 2.5 dB and rejection of greater than 25 dB at center frequency ±12 MHz. The company can manufacture these units at any center frequency from 1995.5 to 2652.5 MHz to help customers upgrade equipment.



Reactel Inc.,
Gaithersburg, MD (301) 519-3660, www.reactel.com.

RS No. 226

Hermetic Electromechanical Switches

The RSMH series is a hermetic electromechanical switch that offers dependability of its design in a hermetic laser welded package. All seals are glass-to-metal or metal-to-metal with no epoxy used. These SPDT switches are sealed in a dry environment and will operate at –55° to +85°C in the most severe conditions. This series is available in break before make, latching or failsafe configurations.

Renaissance Electronics Corp.,
Harvard, MA (978) 772-7774,
www.rec-usa.com.

RS No. 227

Four-way Power Divider

The model WPD-50/4N is a 50 Ω, four-way broadband Wilkinson power divider that covers an 800 to 2700 MHz frequency range. This divider offers a minimum of 20 dB (25 dB typical) of port-to-port isolation with an insertion loss of 1 dB nominal above the theoretical split. This model features an aluminum enclosure with N female connectors on all ports. Applications include antenna sharing, in-building systems and test and verification lab environments. Delivery: available from stock.

Trilithic Inc.,
Indianapolis, IN (800) 344-2412, www.trilithic.com.

RS No. 228

PIN Switches

The model CP10T-77308030-D2 and model CP4T-77305030-D2 are PIN diode-based SP10T and SP4T switches that operate at the RF frequency range between 75 to 78 GHz. The SP10T and SP4T switches offer 10 dB and 6 dB maximum insertion loss and greater than 30 dB isolation, respectively. The bias conditions for the switches are +5 VDC/160 mA, –5 VDC/0 mA for the SP10T version and +5 VDC/50 mA, –5 VDC/0 mA for the SP4T version, respectively. Size: SP10T: 2.75" × 1.7" × 1" and SP4T: 1.4" × 1.7" × 1".

WiseWave Technologies Inc.,
Torrance, CA (310) 539-8882, www.wisewave-inc.com.

RS No. 229

COMPONENTS

DC to 18 GHz Adapters

These broadband adapters virtually cover type-N/SMA and SMA-F to SMA-M interconnection needs within the DC to 18 GHz band. The adapters offer a passivated stainless steel construction that withstands tough environments.

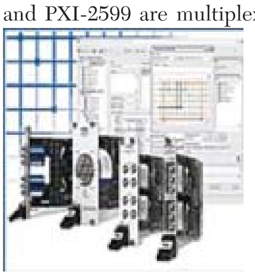
The 50 Ω adapters offer high performance features such as flat response and good VSWR. Price: $8.95 each (1–49).

Mini-Circuits,
Brooklyn, NY (718) 934-4500,
www.minicircuits.com.

RS No. 230

RF PXI Switch Modules

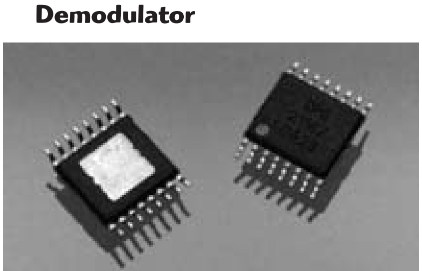
The models PXI-2596, PXI-2597, PXI-2598 and PXI-2599 are multiplexers, SPDT relays and transfer switch modules designed for routing RF or microwave signals in automated test applications. These modules offer 26.5 GHz switching in multiple PXI configurations for a complete switching solution on one platform. Engineers can now use these modules and the latest version of the company's NI Switch Executive switch management software for PXI-based RF and microwave test applications. The switches, combined with the new per-path calibration capability of NI Switch Executive 2.1, offer modularity and programming flexibility for communication test systems.



National Instruments Corp.,
Austin, TX (800) 531-5066, www.ni.com.

RS No. 231

Direct Quadrature Demodulator



The model SRQ-2116Z is a high linearity, silicon germanium direct quadrature demodulator designed for direct conversion and low IF cellular base station and WiMAX receivers. The SRQ-2116Z is available in the industry-standard TSSOP-16 package. This device features high second- and third-order intermodulation suppression, high LO-RF isolation and good quadrature accuracy. Samples are available now and volume quantities will be available in Q1 2006.

Sirenza Microdevices,
Broomfield, CO (303) 327-3030,
www.sirenza.com.

RS No. 252

Type N Male Connector

The model EZ-600-NMC-2 (3190-1387) is a no-solder EZ two-piece clamp-style connector for LMR-600. This connector features a combination hex/knurl coupling nut that allows tightening by hand or with a wrench. This model offers a slimmer design for easier handling and installation. It is assembled with two 15/16" wrenches and is compatible with the standard ST-600C prep tool. This connector is ideal for spread spectrum and ISM band applications up to 5.8 GHz and higher. Price: $20.90.

Times Microwave Systems,
Wallingford, CT (203) 949-8400,
www.timesmicrowave.com.

RS No. 232

AMPLIFIERS

mm-wave IC Expansion

The family of millimeter-wave integrated circuits has been expanded to include low cost, surface-mount amplifiers operating in the 20 to 40 GHz frequency range. The new product family consists of seven devices. The AMMP-6231 is a high performance, low noise amplifier ideally suited for 18 to 31 GHz receive chains. The AMMP-6345 and AMMP-5040 are driver amplifiers for 20 to 45 GHz broadband applications. The AMMP-5024 is a traveling-wave amplifier operating from 100 kHz to 40 GHz. Two others, AMMP-6425 and AMMP-6430, are high performance 1 W power amplifiers for use in frequencies from 17 to 33 GHz. Finally, the AMMP-6130 is a frequency multiplier with integral driver amplifier operating in the 30 GHz satellite band.

Agilent Technologies Inc.,
Semiconductor Products,
Palo Alto, CA (800) 235-0312,
www.agilent.com.

RS No. 233

Detector Log Video Amplifier

The model LVD-812-50-ICW-LD1205 is an 8 to 12 GHz detector log video amplifier (DLVA) that offers a 42 dB dynamic range and typical frequency flatness of 0.4 dB. The VSWR is 2.5 and the TSS is –43.5 dBm typical. The CW-immunity is to –10 dBm while the logging range is –42 to 0 dBm and the log slope is 50 mV/dB. This DLVA was developed as a form, fit and function replacement for DLVAs no longer in production by other manufacturers. Size: 5.126" × 2.50" × 1".

American Microwave Corp.,
Frederick, MD (301) 662-4700,
www.americanmicrowavecorp.com.

RS No. 234

Broadband Amplifiers

The model AMP100G3-20-20ES is a broadband amplifier that operates at a frequency range from 100 MHz to 3 GHz and has a minimum of 20 dB gain. This amplifier features a noise figure below 3 dB and a P1dB of at least +18 dBm. This model is equipped with N(f) input and output connectors and draws less than 200 mA in DC current. The unit contains a built-in bias tee that allows the +15 V input voltage to be injected into the output connector. These broadband amplifiers are ideal for wireless applications. Delivery: available from stock.

Amplical Corp.,
Verona, NJ (201) 919-2088,
www.amplical.com.

RS No. 235

Booster Amplifier Accessories

This group of accessories add more versatility and functions to its battle-tested model KMW1030 booster amplifier. This model is a lightweight, manpack unit for tactical radio, developed for use in the toughest, most demanding applications. It operates in the 30 to 512 MHz frequency bands including SATCOM, and is available in 12 and 20 W configurations. This model can be supplied to operate from 12 or 24 VDC and boost 1 to 5 W of radio power to 20 W.

AR Worldwide • Modular RF,
Bothell, WA (425) 485-9000,
www.ar-worldwide.com.

RS No. 236

Low Noise Amplifiers

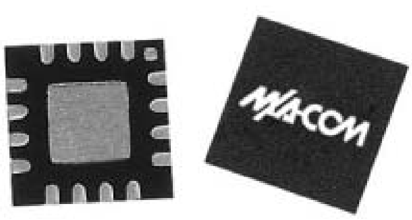
This complete line of low noise amplifiers (LNA) includes several products. The wideband ultra low noise amplifiers operate in a frequency range from 3.1 to 11 GHz and offer a noise figure of 1 to 1.5 dB. The wideband low and medium power LNAs cover a frequency range from 0.5 to 18 GHz and feature a noise figure of 4 to 4.5 dB and P1dB of 10 to 12 dBm. The wideband high power LNAs operate in a frequency range from 2 to 18 GHz and offer a

power out of 30 dBm. The wideband and high gain LNAs operate in a frequency range of 2 to 18 GHz and feature a power out of 17 dBm and gain of 70 dB.
Empower RF Systems Inc., Inglewood, CA (310) 412-8100, www.empowerrf.com.
RS No. 237

Solid-state Power Amplifiers

The S71000 Ku-band series of solid-state power amplifier (SSPA) modules range from 10 to 100 W of output power. These SSPAs provide good performance, reliability and cost effectiveness, all in a slim and compact package.
Locus Microwave Inc., State College, PA (814) 861-3200, www.locusmicrowave.com.
RS No. 238

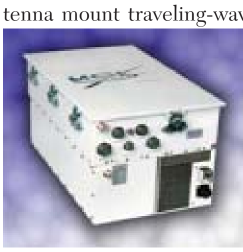
Power Amplifier



The model MAAPSS0076 is a RoHS-compliant 1880 to 1930 MHz DECT power amplifier for applications that require dual power modes, high gain and small size at a low cost. This amplifier offers a wide voltage operating range and is a dual mode power amplifier that maximizes system performance while reducing DC power consumption. The MAAPSS0076 is a three-stage power amplifier designed for digitally enhanced cordless telephone applications and is available in a lead-free 3 mm 12-lead PQFN plastic package. Price: $0.54 (10,000).
M/A-COM Inc., Lowell, MA (800) 366-2266, www.macom.com.
RS No. 253

TWT Amplifier

The model MT4400 is a weather-resistant, antenna mount traveling-wave tube (TWT) amplifier available for DBS-band applications at 750 and 500 W. The MT4400 is also available for C-, X- or Ku-band applications at 750 W. Tri-band is available upon request. Features include an advanced thermal design, rugged construction for extreme environments, optional handheld controller for complete local monitoring and control, prime power interfaces to a wide variety of voltages and frequencies, and field replaceable modules for serviceability.



MCL Inc., Bolingbrook, IL (630) 759-9500, www.mcl.com.
RS No. 239

Low Noise Amplifier

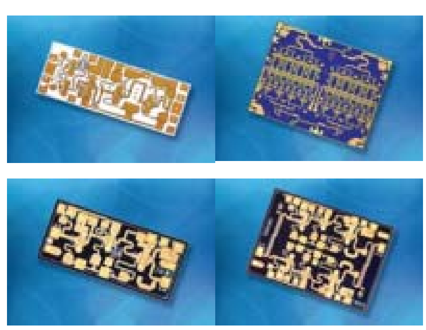
The model PEC-12-50M40G-4R0-15-SFF is a low noise amplifier that operates in a frequency range from 50 MHz to 40 GHz. This amplifier has been developed to provide good gain flatness over an ultra-wide bandwidth. The model offers 12 dB of gain while providing a low midband noise figure of 4 dB. Production units are now available.

Planar Electronics Technology, Frederick, MD (301) 662-5019, www.planarelectronicstechnology.com.
RS No. 240

Successive Detection Log Video Amplifier

The model SDLVA-30M90-80 is a successive detection log video amplifier (SDLVA) designed to operate over the frequency range of 52 to 68 MHz. The input dynamic range is from –14 to –84 dBm while the logging range is –80 to 0 dBm typical. The limited IF output is 0 dBm, while the maximum RF input power is 0 dBm. The logging rise and fall times are 500 ns maximum, the slope is at 55 mV/dB and the propagation delay is 10 ns typical. Size: 3.75" × 1.50" × 0.50".
Planar Monolithics Industries Inc., Frederick, MD (301) 631-1579, www.planarmonolithics.com.
RS No. 241

Wideband mm-wave Amplifiers



Four high performance wideband millimeter-wave amplifiers with industry-leading power bandwidth performance have been launched. The TGA4521 is a wideband millimeter-wave driver amplifier covering the 32 to 45 GHz frequency range, which has +25 dBm saturated output power and +24 dBm 1 dB compressed output power. The TGA4522 is a balanced version of the TGA4521, offering a higher output power capability with +27.5 dBm saturated and +27 dBm 1 dB compressed performance at 38 GHz. The TGA4046 is a balanced high power amplifier MMIC for Q-band applications, particularly military and commercial satellite uplink communications with a saturated output power of +33 dBm and 1 dB compressed power of +32 dBm. Finally, the TGA4040 is a medium power amplifier/frequency multiplier MMIC for a variety of applications including military and commercial satellite communications, electronic warfare, digital radio and instrumentation.
TriQuint Semiconductor Inc., Hillsboro, OR (503) 615-9000, www.triquint.com.
RS No. 243

Power Amplifiers

The model QPN-94042730-01 and model QPN-94043025-02 are W-band power amplifiers that operate from 92 to 96 GHz and offer 25 dB of gain, minimum. Two saturated output power levels are offered: QPN-94042730-01 at 27 dBm and QPN-94043025-02 at 30 dBm. DC requirements are +5 V at 2.0 A for 0.5 W output and 4.0 A for the 1 W output. Sizes: 2" × 1.4" × 1" for 0.5 W and 2.2" × 1.7" × 1.4" for 1 W, with WR-10 inputs and outputs. Fast modulation is available.
QuinStar Technology Inc., Torrance, CA (310) 320-1111, www.quinstar.com.
RS No. 242

ANTENNA

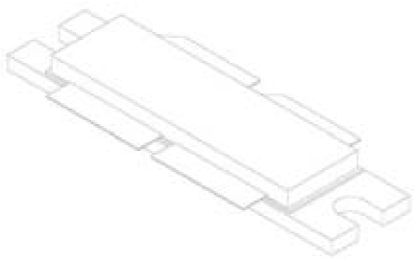
Two-piece Grid Antenna

The model GS2-54-N is a high gain 24.8" × 16.5" two-piece grid antenna that was designed for use in the licensed Public Safety communications frequency band of 4.94 to 4.99 GHz as well as the unlicensed frequency bands from 5.25 to 5.85 GHz. The main reflector is precision die cast aluminum. The mount is designed to mate to a 2" OD pipe and an optional mast adapter kit is available for mast pipe sizes up to 4.5" diameter. Mount hardware used for assembly is stainless steel. The input for the feed is a type N female connector.
mWAVE Industries LLC, Gorham, ME (207) 857-3083, www.mwavellc.com.
RS No. 244

NEW PRODUCTS

DEVICE

LDMOS Power Transistor



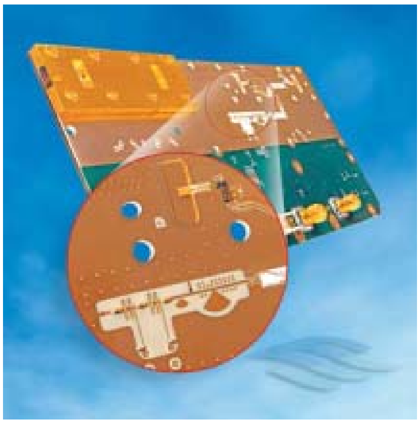
The model PD21120R6 is an internally matched 120 W, 2170 MHz, 28 V RF LDMOS power transistor. This device was designed for push-pull WCDMA/UMTC applications. It provides over 14 dB of gain at 2170 MHz, with 22 percent efficiency during 20 W average power conditions, and 13 dB of gain with 48 percent efficiency during 120 W P1dB CW conditions. The device is capable of handling a 10:1 VSWR.

Peak Devices Inc.,
Boulder, CO (720) 406-1221,
www.peakdevices.com.

RS No. 245

HARDWARE

Packageless Transceiver



This commercial introduction of an integrated transceiver board solution is designed to eliminate 30 percent of the typical mechanical packaging cost from conventional modules, while preserving 100 percent of electrical performance. The proprietary packaging technique used is called Epsilon Packaging™ In addition to cost reduction, Epsilon Packaging significantly reduces the size and weight of a typical transceiver subassembly. This is accomplished by replacing heavy-weight metal mechanical items with plated FR-4 and injection molded metallized plastics.

Endwave Corp.,
Sunnyvale, CA (408) 522-3127,
www.endwave.com.

RS No. 254

INTEGRATED CIRCUIT

SoC ZigBee Solution

The CC2430 family is a system-on-chip (SoC) ZigBee solution providing on-chip programmable flash memory as well as a certified ZigBee software stack, all on a single silicon die. It is based on the company's SmartRF03 technology platform in 0.18 µm CMOS and is available in a 7 ×7 mm, 48 pin package. The CC2430 SoC family comprises three products – CC2430-F32, CC2430-F64 and CC2430-F128 – the difference being the flash configurations of 32, 64 and 128 kBytes with each configuration having 8 kBytes of RAM and other powerful supporting features.

Chipcon AS,
Oslo, Norway (+47) 22 95 85 44,
www.chipcon.com.

RS No. 246

MATERIAL

Prepregs and Laminates

The TLG-2.9 and 3.0 are low loss bromine-free prepregs and laminates designed with a low dissipation factor of Df = 0.0030 to 0.0038 at 10 GHz – thickness dependent. These "green" materials are designed for military, RF and high speed digital markets, where a homogeneous dielectric constant throughout the laminate is desirable. The TLG-2.9 and 3.0 were designed and beta tested on thick backpanels with the intent to improve the ease of fabrication in difficult designs.

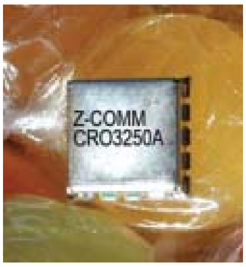
Taconic,
Petersburgh, NY (518) 658-3202,
www.taconic-add.com.

RS No. 255

SOURCE

Coaxial Resonator Oscillator

The model CRO3250A is an L-band coaxial resonator oscillator (CRO) that utilizes voltage-controlled oscillator (VCO) technology and is geared for the point-to-multi point radio market. This model utilizes a coaxial resonator, which greatly increases the Q of the VCO



therefore optimizing the performance by delivering low phase noise performance in a low cost, compact package. This model is designed for narrowband operation and offers a tuning bandwidth within 0.5 to 4.5 VDC of control voltage with an average tuning sensitivity of 13 MHz/V for easy integration with PLL ICs. Size: 0.50" × 0.50" × 0.22". Delivery: stock to four weeks.

Z-Communications Inc.,
San Diego, CA (858) 621-2700,
www.zcomm.com.

RS No. 247

TEST EQUIPMENT

Digital Spectrum Monitoring

A new test capability, the PN95-DSM digital spectrum monitoring (DSM) option has been added to the expanding PN9500 phase noise test system. The PN9500 with DSM performs a much wider range of high frequency measurements within one modular system. The DSM option allows the user to view the spectrum of a signal in the same way as a spectrum analyzer and also integrates specific processes that are ideal for radar testing. The PN95-DSM measures power level, power variation, adjacent channel power and harmonic distortion, and more measurements are possible by setting markers. The user can perform narrowband spectrum analysis, but can also opt to measure any wideband spectrum from DC to 18 GHz.

Aeroflex Inc.,
Plainview, NY (516) 694-6700,
www.aeroflex.com.

RS No. 248

Compact Spectrum Analyzer

This compact spectrum analyzer (CSA) is the first in a series of intuitive, low cost spectrum analyzers that enable technicians, and R&D and manufacturing engineers to make complicated RF measurements with speed, ease and confidence. Available in 3 and 6 GHz models, the CSA incorporates a full-featured, general-purpose spectrum analyzer with an internal VSWR bridge and tracking generator. This en-

New Products

ables users to easily and accurately characterize single and dual-port devices. The built-in measurement help leads users through each measurement task step by step and its modern connectivity simplifies tasks such as remote control, data transfer and firmware upgrade. Also, its convenient form factor, low weight and battery operation make it suitable for installation and maintenance of communications systems.
Agilent Technologies Netherlands B.V.,
Amstelveen, The Netherlands
+31 20 547 2000, www.agilent.com.
RS No. 249

Signal Generator Enhancements

Pulse modulation enhancements to the MG3690B signal generators result in more precise and convenient simulation of pulsed signals used in civilian and military radar applications. These enhancements provide more narrow leveled pulses, increased resolution when using the internal pulse generator, and include higher frequency internal waveform generators to simulate modulated signals. The signal generator's pulse modulation performance responds to emerging needs of radar systems, especially those operating in the 1 to 2 GHz L-band such as air traffic control, traffic and collision avoidance systems, joint tactical information distribution system and other distance measuring equipment.
Anritsu Ltd.,
Luton, Bedfordshire, UK
+44 1582 433433, www.anritsu.com.
RS No. 250

Vector Network Analyzer

The R&S ZVA24 is introduced as the high end model of the vector network analyzer family, comes with two or four test ports and is designed to operate in the frequency range from 10 MHz to 24 GHz. Characteristics include a dynamic range of more than 145 dB, an IF bandwidth of up to 1 MHz and a measurement speed of 3.5 µs per test point. It has versatile measurement capabilities and high flexibility, making it suitable for balanced measurements as well as complex measurements on frequency-converting active components such as amplifiers, mixers or frequency converters for mobile radio, WLAN, SAT or other RF applications. The vector network analyzer offers ample functionality even in its basic version, which can be expanded with a variety of options.
Rohde & Schwarz GmbH & Co. KG,
Munich, Germany +49 89 4129-13779,
www.rohde-schwarz.com.
RS No. 251

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NEW LITERATURE

DIGITAL DEBUG SOLUTIONS CD-ROM

This CD-ROM will assist design engineers with measurement insight into tough digital debug problems with information about tools that can help minimize project risk and drive products to market faster. The CD also provides quick access to application notes, product data sheets and web links.
Agilent Technologies Inc.,
Palo Alto, CA (800) 829-4444,
www.agilent.com.

RS No. 200

PRODUCT DATA SHEET

This data sheet details the company's dual-band CDMA power amplifiers that reduce time to market and PCB space. An overview, features, product descriptions and performance data are also provided.
ANADIGICS Inc.,
Warren, NJ (908) 791-6000,
www.anadigics.com.

RS No. 201

AMPLIFIER BROCHURE

The SuperPower Amps brochure provides a detailed overview of the company's microwave amplifiers for EMC and wireless testing. This brochure contains information on the many "Gigs Galore" microwave offerings from AR, along with specifications on the ST, T and S series amplifiers that comprise these offerings.
AR Worldwide • RF/Microwave Instrumentation,
Souderton, PA (215) 723-8181,
www.ar-worldwide.com.

RS No. 202

PRODUCT DATA SHEET

This data sheet details the company's end launch/waveguide to coax adapters. The end launch adapters operate in standard waveguide sizes from 10 to 40 GHz. Waveguide to coax adapters are made of aluminum with standard cover flanges.
Microwave Development Laboratories (MDL),
Needham, MA (781) 292-6680,
www.mdllab.com.

RS No. 208

PRODUCT BROCHURE

This brochure features the company's UFX-EbNo series of precision generators. Features, ratios, a simplified functional block diagram, specifications and ordering information are also provided.
Noise Com,
Parsippany, NJ (973) 386-9696,
www.noisecom.com.

RS No. 203

RF SWITCH DATA SHEET

This data sheet provides detailed information on the company's wireless UltraCMOS™ switches that extend performance to 3 GHz. The switches feature high isolation, high ESD tolerance and low insertion loss and are ideal for wireless applications.
Peregrine Semiconductor Corp.,
San Diego, CA (858) 731-9400,
www.psemi.com.

RS No. 204

MILLIMETER-WAVE PRODUCTS

This brochure highlights the company's components, subsystems and integrated assemblies, engineering services and custom products that serve virtually every application and market including communications, radars and sensors, military applications, test, measurement and instrumentation, and industrial and scientific applications.
QuinStar Technology Inc.,
Torrance, CA (310) 320-1111,
www.quinstar.com.

RS No. 205

SPECIALTY MATERIALS BROCHURE

This product capabilities brochure offers an array of specialty materials including high frequency circuit materials, laminates, photoimageable covercoats, high performance foams, busbars, EL lamps and drivers, elastomer components, nitrile floats and nonwoven materials.
Rogers Corp.,
Rogers, CT (860) 774-9605,
www.rogerscorporation.com.

RS No. 206

PRODUCT SELECTION GUIDE

This selection guide highlight's the company's various products from the recent acquisitions of Amplifonix, FSY Microwave, Magnum Microwave, Q-bit, Salisbury Engineering and Radian Technologies. The updated guide details the features, benefits and performance characteristics of each product type.
Spectrum Microwave Inc.,
Palm Bay, FL (321) 727-1838,
www.specwave.com.

RS No. 207

TEST CABLE BROCHURE

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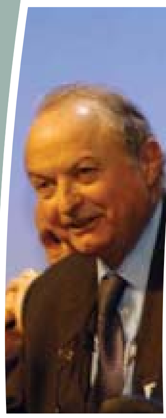

THE BOOK END

Planar Microwave Engineering: A Practical Guide to Theory, Measurements and Circuits

Thomas H. Lee

Cambridge University Press
880 pages; $75
ISBN: 0-521-83526-7

This book covers everything one needs to know to design, build and test a high frequency circuit. Chapter 1 provides a short history of RF and microwave circuits. Chapter 2 introduces some definitions and basic concepts. Chapter 3 provides a brief introduction to the Smith chart and S-parameters, which are staples of classical microwave design. Chapter 4 presents a number of impedance matching methods along with a brief explanation of the Bode-Fano limit. Chapter 5 surveys a number of popular connectors, their domain of application and the proper ways to care for them. Cables and their characteristics are discussed as well. Chapter 6 examines the characteristics of lumped passive elements at microwave frequencies. Simple circuit models are presented. Chapter 7 introduces the most common way of building microwave circuits: microstrip. Chapter 8 presents several methods for making impedance measurements, ranging from time-domain reflectometry to vector network analysis. Chapter 9 is devoted to microwave diodes, following which Chapter 10 describes numerous mixers. Chapter 11 presents a survey of the many types of transistors that have been developed until now. Chapter 12 considers how to squeeze the most out of whatever transistor technology is used. Chapter 13 discusses noise models and presents the theory of noise matching. Chapter 14 describes the principles underlying noise figure measurements. Chapter 15 describes how to produce controlled instability to build oscillators, while frequency synthesizers are the subject of Chapter 16. Chapter 17 analyzes the important subject of phase noise, where it comes from and how to reduce it. Chapter 18 describes phase noise measurement methods. Chapter 19 describes spectrum analyzers, oscilloscopes and probes. Chapter 20 presents numerous ways to implement power amplifiers at RF and microwave frequencies. Chapter 21 shows how to get power into and out of the air, with emphasis on microstrip antennas. Finally, Chapters 22 and 23 focus on the design of passive filters.

"This book covers everything one needs to know to design, build and test a high frequency circuit."

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Jolyon M. de Freitas

Artech House
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ISBN: 1-58053-759-6

This collection of predesigned linear phase digital filters has been prepared to serve the category of users whose research or development applications require some form of digital filtering, but do not have the time to explore the very wide range of digital filter design techniques available. The aim of this working handbook, therefore, is to provide a collection of digital filters whose characteristics can be quickly determined, compared and applied immediately, avoiding the sometimes intricate design process. In Chapter 2, a description is given of how the filters presented were designed. Some basic implementation algorithms are described. Chapters 3 to 5 give the filter coefficients as well as their performance features in a data sheet format. Each filter is treated as an individual product with its associated data sheet. Some preliminary material is given at the beginning of each chapter that provides information on terminology and definition of terms used to characterize the filters. The preliminary material in Chapter 3 (Low Pass Filters) contains discussions that are largely applicable to the other two chapters. As such, the introductory material in Chapters 4 (High Pass Filters) and 5 (Bandpass Filters) are short as they are unique to their description and characterization. Chapter 6 provides coefficients for first- and second-order differentiating filters and shows how to manage the noise and avoid its amplification. Again, the coefficients are viewed as individual products with specific properties. The limitations of their use are also discussed. Chapter 7 (Hilbert Transformers) is very different from any of the other chapters. It gives the Hilbert coefficients, but within the context of FM/PM demodulation. Complementing the book, a CD-ROM is included, which contains an archive of filter coefficients and design functions written in MATLAB.™

"...a collection of digital filters whose characteristics can be quickly determined, compared and applied immediately..."

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Wireless Technologies 2005

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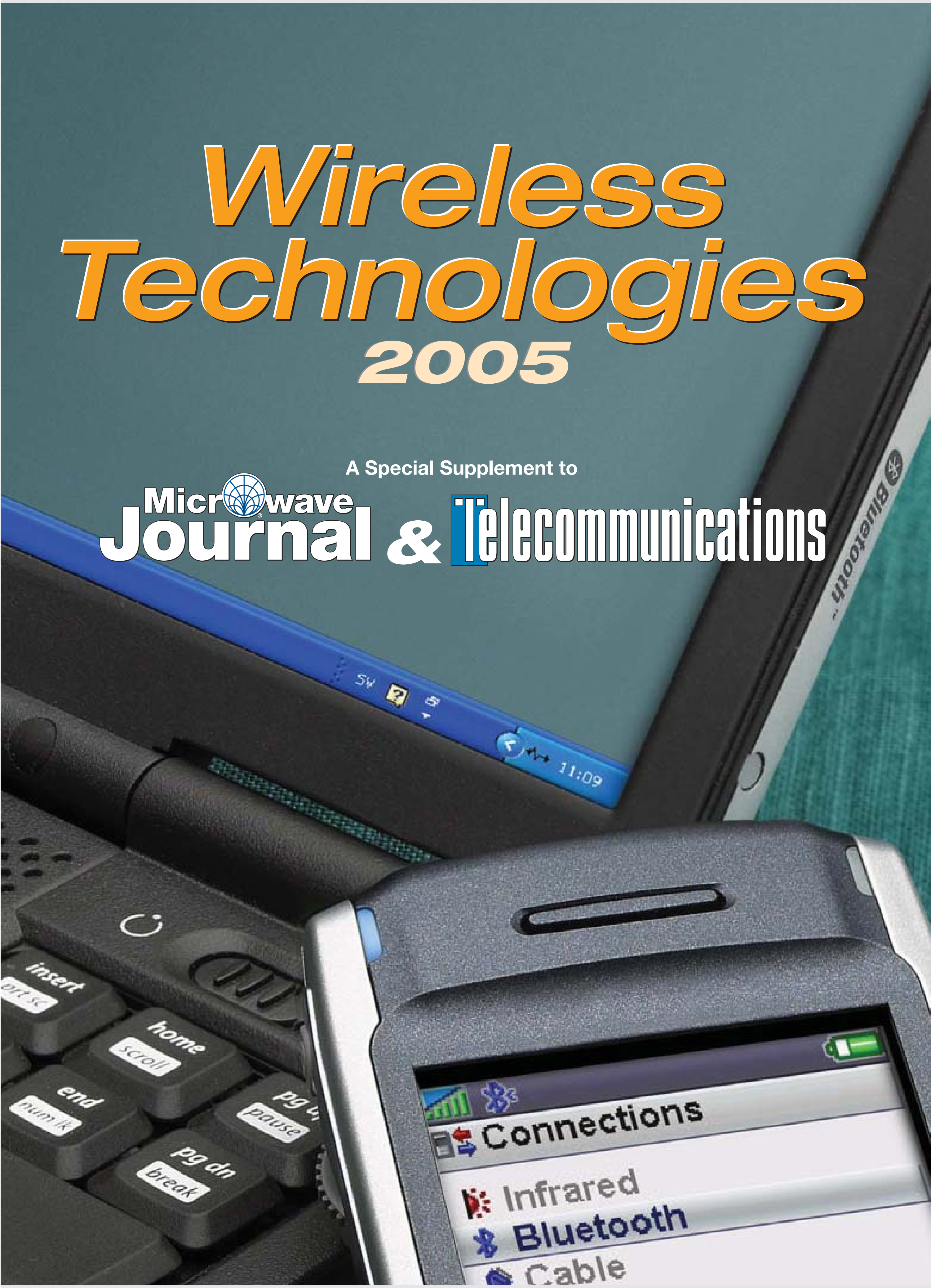

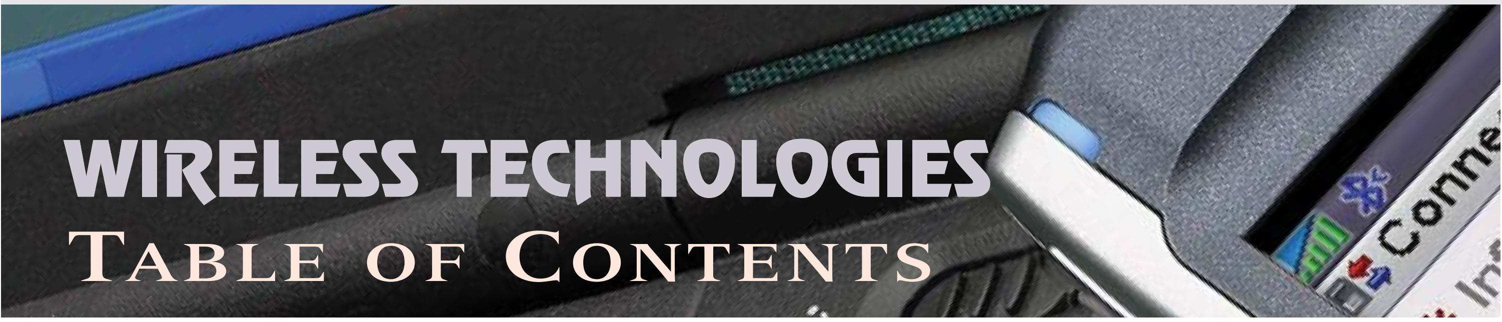

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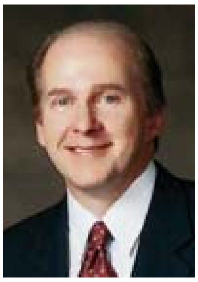
Wireless Technologies – Viewpoints from Two Markets and Two Continents

The editors of *Microwave Journal* and the Americas and International issues of *Telecommunications* interviewed key figures in their markets concerning the interaction between suppliers and users of equipment for wireless communications. Their comments are presented below. Because of space limitations, some answers have been edited.

Jerry D. Neal
Executive VP and Co-Founder
RF Micro Devices



MWJ: *As a supplier to the wireless systems market, what information would you like to see from the systems providers?*

JN: Working closely with our customers and sharing information is key to RFMD's success. Through close collaboration with handset manufacturers, we gain valuable insight into their plans for future handset designs, which enables us to anticipate the features that will differentiate future products. This helps to determine form factors and the types of additional functionality such as Bluetooth® technology, GPS, WLAN and multimedia capability that will be included in handsets and allows us to develop the RF solutions our customers will need.

MWJ: *How have evolving requirements of the communications sector affected your product mix/plans?*

JN: Cell phones are evolving into wireless handheld computing devices that incorporate a variety of complex components. Handset manufacturers are looking to their key RF suppliers to solve this complexity and provide more complete systems solutions to accelerate handsets to market. Once suppliers of PA MMICs, today we supply highly integrated PA modules that incorporate advanced functionality such as power control, switch technology and other passive components. As they incorporate more features such as cameras, web browsing, FM tuners and streaming video, a big challenge is extending battery life. We now include DC-to-DC converters for power management in our modules to extend battery life. As RFMD evolves into a provider of systems solutions, we are increasing the number of systems and signal processing engineers on our staff.

MWJ: *What new technologies do you foresee impacting your product lines?*

JN: We continue to optimize our proven GaAs HBT process and we are adopting new technologies, including lithium tantalite, pHEMT, indium gallium phosphide and custom silicon processes. We're using lithium tantalite to build SAW filters, which handset manufacturers previously purchased from other suppliers. We are incorporating these filters into our modules, which provides our customers with a more complete solution, reducing their design time and streamlining their supply chains. We are also incorporating switches into our products and are developing new pHEMT process technology. We're also working with silicon for our transceiver products, *Bluetooth*® solutions and even our PA products. More specifically, we're working with CMOS and silicon germanium, since as we move further into the digital arena and signal processing, the need for silicon technologies will increase.

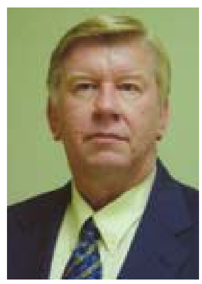
MWJ: *Is there a significant difference between domestic and overseas markets?*

JN: Fundamentally, the domestic and overseas markets share some similarities. No matter the market or the geographic location, semiconductor manufacturers must be agile and have the ability to adapt and anticipate the needs of all handset markets. As a global supplier, we have established offices and operations all over the world to support our customers. Today, over 80 percent of RFMD's business comes from overseas. As the cell phone market continues to evolve, one of the biggest near-term growth drivers is providing components for low cost handsets designed for the emerging low tier markets for developing nations, such as China, India, Bangladesh, Indonesia and the Philippines.

MWJ: *How much of your communications products and solutions have dual usage for military and commercial applications?*

JN: RFMD is primarily focused on high volume, commercial products. Very few of our products are suitable for military use. Some of our technologies, however, are applicable to military applications. For instance, while the military has used GPS technology for years, we believe there may be a need for low cost GPS military applications in the future. Also our work in gallium nitride (GaN) technology may have the most immediate application because it has potential broad uses in radar and other imaging technology.

Charles A. McCauley
Vice President
and General Manager
Renaissance Electronics Corp.



MWJ: *As a supplier to the wireless systems market, what information would you like to see from the systems providers?*

CM: We would encourage the systems providers to share sufficient information to allow us to understand both the technical objective as well as the market conditions, in order to provide an economical solution in a timely manner. This sharing of information needs to be reciprocal. We as a supplier must be willing to share our capabilities and work with the customer in order to find the optimal solution.

MWJ: *How have evolving requirements of the communications sector affected your product mix/plans?*

CM: As more companies are branching out to new and different technologies, integrating systems has become more difficult and important. We have seen everything from making products that can be used simultaneously for CDMA, TDMA and MSM, to products which are capable of handling the 800, 900, 1800 and 1900 MHz spectrums. The telecommunications market is trying to find new ways to save both space and money, while being able to use every available technology and every kilohertz of bandwidth at their disposal. The bandwidth driver for 3G, 4G, data and video will be the most prevalent of demands.

MWJ: *What new technologies do you foresee impacting your product lines?*

CM: The two most significant are nano-switching and cryogenic reception. Also, we foresee new technologies and revolutionary uses for optical technologies that will indirectly affect our company.

MWJ: *Is there a significant difference between domestic and overseas markets?*

CM: There is and there isn't. Even though the actual technologies and deployment are very different, the end goal of all the companies is very similar. While a device may not function in another market, a similar product would be required in almost every market. All the markets are always concerned with coverage, capacity and economics.

MWJ: *How much of your communications products and solutions have dual usage for military and commercial markets?*

CM: Many of our applications have dual usage. While quite often the frequency ranges will overlap for military and commercial applications, both have uses for high power (transmit) and low power (receive) functions. The major difference between the two applications is that military customers usually have the performance specifications as their primary concern while commercial customers usually have price and economics as their main focus. Our primary methodology, in order to optimize engineering productivity, is to design units that fit both market places.

Bill Flerchinger
Strategic Product
Planning Manager
Agilent Wireless Division

MWJ: *As a supplier to the wireless systems market, what information would you like to see from the systems providers?*

BF: The two primary areas of information from systems providers relate to their interconnectivity and emerging wireless technology needs. In the case of interconnectivity we need to understand requirements for physical connections (like GP-IB, LAN, USB, etc.) as well as the software interconnectivity (drivers, OS, etc.), both today and in the future.

Understanding what new wireless technologies are needed and when they will be deployed is essential. The acceleration of digital wireless technologies continues at an unprecedented rate for broadband wireless access, wireless data access, high mobility cellular, specialized mobile radio and military applications. There is literally no segment of the wireless market that is not moving to new digital access technologies or to 2nd, 3rd and even 4th generation technologies. Agilent has or is developing new test solutions to support this evolution. We have engineers serving on all the major standards boards and committees to help us ensure that our customers designs and products meet those standards.

MWJ: *How have evolving requirements of the communications sector affected your product mix/plans?*

BF: We have a comprehensive set of test solutions for every stage of the wireless device life cycles and for each major digital communication technology. We continue to build on very early successful products and extend those solutions to evolving technology and customer needs. This has helped the industry, since most of our products can be upgraded to new capabilities. These include power supplies, oscilloscopes, logic analyzers, signal generators, vector signal analyzers, spectrum analyzers, network analyzers and one-box testers. We are also driven to design for future needs. Specifically, to accommodate higher data rates, increasing system bandwidths and higher frequency ranges by taking advantage of the latest technology available to provide industry leading test solutions for the next generation of systems.

MWJ: *What new technologies do you foresee impacting your product lines?*

BF: The evolution of existing high mobility cellular standards like GPRS, EGPRS, cdma2000, 1xEV DO release 0 and W-CDMA will continue. We continue to enhance our products for these technologies. Additionally, there are several new technologies that will have new focus in the coming year. These include HSDPA, HSUPA, 1xEV DO release A, TD-SCDMA, Flash OFDM and others. Broadband wireless access and wireless data access technologies like Bluetooth®, UWB (Ultra-wideband 802.15.3a), WLAN (802.11x), WiMax (802.16) and ZigBee® (802.15.4) will continue to keep our attention.

One of the areas that I see having a significant impact on the industry is the "converging wireless device." This is the personal electronic device that everyone will carry with them, which is a cell phone with multiple high mobility technologies plus other features like GPS, Bluetooth®, WLAN, FM stereo, digital multimedia (DVB-H or DMB) receivers, etc. It has tremendous time-to-market and cost pressures and will continue to be an opportunity to efficiently integrate, validate and verify the capabilities both in development and production environments.

MWJ: *Is there a significant difference between domestic and overseas markets?*

BF: There are many differences between domestic and overseas markets. However, with the globalization that has taken place over the last decade, the differences are becoming less and less. Still there are unique needs in the wireless market around the globe. Examples include language localization, customer use models and differences like Christmas or Lunar holidays that drive customer buying patterns. There are also emerging technologies that are geographically focused. For example, the Chinese TD-SCDMA 3G cellular standard, DMB (Digital Multimedia Broadcast) currently being deployed in Korea and Japan, MediaFLO initially US only and DVB-H being trailed in Europe. With our strong global support, delivered locally, we can better meet the unique needs of our customers no matter where they are in the world.

MWJ: *How much of your communications products and solutions have dual usage for military and commercial applications?*

BF: Many of Agilent's design, verification and test products see dual usage. In some instances it is the same product but with different software or hardware configurations. In other instances the products are specifically targeted to specific market segments. We have test solutions for both.

THE TCS PERSPECTIVE

Vanu Bose
Founder and CEO
Vanu Inc.

TCS: *As a wireless systems builder, what would you like to see from the component and subsystem industries/segments?*

VB: We see five key subsystems in a software radio system: the antenna, power amplification, RF up/down conversion, A/D and D/A conversion and the processing engine.

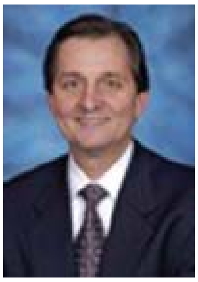
We would like to see more commoditization of components through the use of standardized interfaces. The antenna/PA interface is fairly well standardized at this point, as is the RF/PA interface. The RF-digital interface is just evolving. There are two competing industry standardization efforts, OBSAI and CPRI, but neither one is flexible enough to support multiple standards and incorporate the necessary control and management interfaces to integrate with infrastructure equipment from multiple vendors. An open RF-digital interface with sufficient flexibility and control is essential for reducing the cost of wireless infrastructure systems.

Another feature that would provide significant benefit in the RF section would be tunable narrowband RF filters, perhaps based on MEMS technology or some other new technology. This would significantly increase the performance, capacity and interference rejection of infrastructure systems. The A/D and D/A technology has made significant progress in the last ten years, in particular in the spurious free dynamic range of high speed A/D converters and the functionality of digital up/down converters. While they are available from multiple vendors, they will need to incorporate the open digital-RF interface in order to drive volumes up and costs down in the future. The processing platform portion of the supply chain is in the best shape, with a wide variety of processing platforms available from FPGA to DSPs to GPPs.

TCS: *How have evolving requirements of the communications sector affected your product mix/plans?*
VB: A major goal of a software radio system is to increase the pace of change in wireless systems by replacing fork lift upgrades with software downloads. We view changing requirements as a good thing, and something that our technology can help to address.
TCS: *What new technologies do you foresee impacting your product line?*
VB: RF MEMS technology, in the form of tunable high-Q RF filters, has the potential to impact both the performance and cost of the system. Also, improvement in MCPA technology to improve efficiency and lower overall system cost will be important.
TCS: *Is there a significant difference in the requirements of local and outside (N. America) markets?*
VB: Large established international markets such as Europe have similar requirements as the US Tier I market. However, international requirements for new network builds in emerging countries such as India and Africa are very different. Certainly cost is a driving factor in these markets, especially since many of the areas have low population density. In some ways these requirements are similar to rural America, but more extreme in terms of cost sensitivity. Other requirements that differ are environmental factors and the quality and availability of infrastructure services such as power and telecom. Power is often unreliable, and regulation of frequency and voltage is not as tight. Telecom services are often unavailable in many areas where coverage is desired, and a cell phone is often the first phone that people in these areas have ever had. The lack of telecom infrastructure requires creativity in backhaul solutions, including unlicensed and licensed wireless as well as satellite.
TCS: *How much of communications products and solutions have dual usage as in commercial and military?*
VB: As the military moves more towards leveraging commercial off-the-shelf (COTS) technologies, more and more commercial communication products will find use in military applications. For example, we recently announced that we successfully completed and demonstrated a prototype mobile GSM cellular base station with satellite backhaul for secure, rapid, field deployable applications for the military. We were awarded the development contract from the US Army Communications Electronics Command Research, Development and Engineering Center (CERDEC).

The prototype extended the capabilities of our commercial base station to address the mobility and encryption required to support dismounted soldiers on the battlefield. The Anywave Base Station, built on commercially available hardware, speeds development and deployment times to the field for the military and speeds time to market for commercial deployments. The small form factor and remote management capabilities will enable application to a number of new market opportunities for secure, mobile communications.

Adrian Nemcek
President, Networks Business
Motorola Inc.



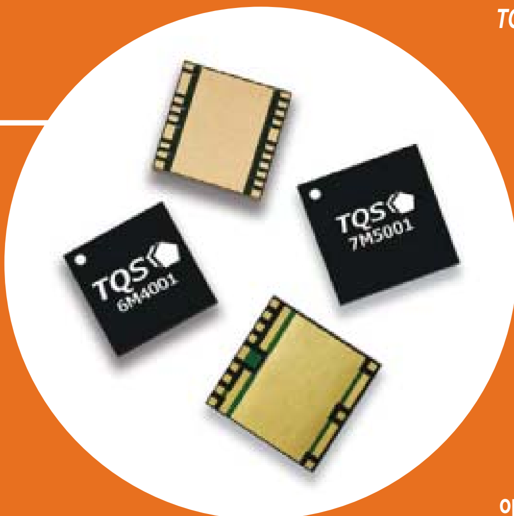
TCS: *As a wireless systems builder, what would you like to see from the component and subsystem industries/segments?*
AN: As the general purpose computing and telecommunications industries continue to align on technologies, telecommunications network infrastructure is being based increasingly on COTS and Free Open Source Software (FOSS) components. In many cases, we don't see these aspects of our products to be points of differentiation for our customers. We do see prudent leverage of this area to offer advantages in time to market and economies of scale. We would like for suppliers into the COTS and FOSS environment to continually evolve their solutions to include carrier grade standards. Further, we would like for more standardization bodies to drive standards and profiles in this area. This applies to platform and board level standards such as ATCA as well as software component standards such as Carrier Grade Linux (CGL) and Service Availability Forum (SAF). In particular, we would like to see the industry move up on the food chain with increased standardization to facilitate a greater ecosystem of Carrier Grade network management solutions.
TCS: *How have evolving requirements of the communications sector affected your product mix/plans?*
AN: The telecommunications sector is rapidly moving down a path of convergence. Integration across Internet, wireless and wireline technology is driving greater demand among end users and operators for delivery of new innovative voice, data and video services at faster speeds and lower costs.

Motorola is meeting these demands by making investments in the access technologies that are crucial today and will continue to gain momentum in the future — DSL, fiber, WiMAX, cellular, cable. We foresee carriers will continue the rapid adoption of IMS to provide a common session control and service/application foundation across these networks.

We are committed to delivering seamless mobility solutions that leverage IP-based technologies and allow operators to deliver compelling, new end-user services across a converged services network. Motorola's Networks portfolio continues to leverage our heritage in cellular and core networks, as well as our focus on next generation network solutions and services such as "push-to" technology, enhanced multimedia messaging, wireless broadband and WiMAX, MVNO capabilities and other service delivery platforms.

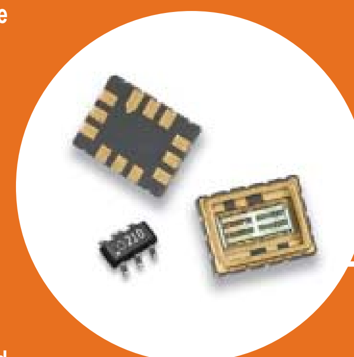
The rapid growth of these new technologies, increasing complexity of networks and the competitive pressures to have networks operating at peak performance has also led many operators to enlist Motorola's experience in providing managed services.

Through a converged services network, operators can integrate multiple access technologies such as cellular, WiFi and broadband technologies, while offering their customers a consistent service experience across all the environments cost effectively. Users now can have a single device and a single service



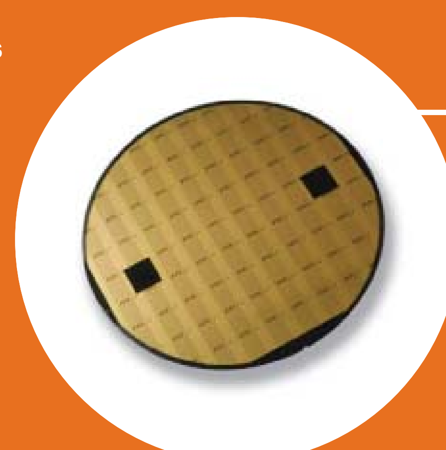
TQM6M4001 – GSM / GPRS: TriQuint designed its GSM / GPRS quad-band transmit (Tx) module to be the smallest, most highly integrated product available. Design variants (TQM6M4002 / 4003) ensure multiple transceiver compatibility; a dual-band product is also available. TriQuint continually shrinks its transmit modules to fit the newest GSM handset designs that pack more features into smaller spaces. Our Tx modules are only 6x6x1.1mm – 44% smaller than competitors, while still offering superior cost effectiveness, a wide range of optimization alternatives and zero SMDs – all with no external matching required.

TQM7M5001 / 5002 – EDGE PA Modules: Our new EDGE PAMs offer high performance in very small, low cost packages. The 5001 features a linear PA; the 5002 is ideal for polar designs. These modules operate in dual modes: as a saturated GSM power amplifier module with internal power control, and as a linear or polar EDGE PAM. Both products' ACPR substantially exceeds ETSI standards over temperature and voltage.



TQM7M6001 – WCDMA PA Module: This new module is a dual-band WCDMA (UMTS) PAM in a small form factor – only 4x4x1.1mm. It supports cost-effective GSM / EDGE / WCDMA compressed mode architectures with only one antenna. This module optimizes idle current consumption for better talk time and to power all those features that make 3G the next leap in customer satisfaction.

CDMA Handsets: TriQuint Semiconductor's CDMA product portfolio offers comprehensive RF Tx front-end solutions. Our innovative PAs, PA modules and highly integrated Tx modules offer better cost structures and performance tailored to individual customer designs. The world's largest manufacturers rely upon TriQuint's exceptional CDMA SAW filter portfolio. Our impressive CDMA zero-IF filter families are designed and validated to support slim-line design. TriQuint also offers Tx and Rx filters for all bands in all popular designs for complete manufacturing flexibility.



Wireless Network Base Stations: TriQuint is bringing new base station filter products to market that reduce size and increase performance for superior network cost efficiency. The latest SAW IF filters for WCDMA (UMTS) network base stations are 50% smaller than previous generations while offering rugged, dependable performance in a smaller, convenient size. We are also developing new base station active components including low-noise amplifiers and power transistors providing the highest efficiency levels available. In millimeter wave components, our newest compact driver amps and HPAs for digital radio, point-to-point radios and related applications are built with TriQuint's proven 0.15µm gate power pHEMT process. See us for the newest designs, including solutions for cost-sensitive markets including WiFi, WLAN and WiMAX.

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provider for all their communications needs, with the ability to personalize services. Additionally, users can experience a diverse portfolio of applications seamlessly whenever they need them and wherever they are at an affordable price.

TCS: *What new technologies do you foresee impacting your product line?*

AN: IP-based networks (internal and end-to-end voice over IP and real-time);
Media over IP (wireless and wireline)
IP Multi-media Subsystem;
Peer-to-Peer applications;
Mesh technologies for wireless;
Increased penetration of fixed interfaces to residences (DSL, fiber and cable);
Ever increasing performance of wireless broadband standards (WiMAX, 3GPP HSUPA, 3GPP2 HSDPA);
Security;
Content management;
Intelligence/capability in the edge/consumer device.

IMS and IP-based technologies will have a huge impact on Motorola's products and solutions as well as the industry as a whole.

• IMS is designed to support both wireless and wireline networks and is independent of the access technology. IMS will enable a rapid deployment of IP-based services across a network unlike the slow rollout of services tied to a legacy circuit-based switch.

• Motorola has developed an IP-based architecture for its CDMA RAN product line to provide a number of benefits such as enhanced connectivity options out to the cell sites, greater system capacities and enhanced services.

• An all IP-based architecture Carrier Access Points (CAP) is being developed first for the Motorola wi4 802.16e systems. Peer-to-peer connectivity at the RF sites eliminates the requirement for expensive base site controllers or radio network controllers. The CAP architecture supports the movement to an IMS core, which is a key enabler for seamless mobility services.

The demand for wireless broadband continues to drive the market.

• Motorola's unlicensed spectrum Canopy product line has enjoyed three years of solid substantial growth providing broadband data and VOIP services to homes and small businesses. As WiMAX emerges, the 802.16e licensed spectrum systems will deliver these same fixed based services and even more with full mobility for voice and broadband data services. Motorola is helping to drive these solutions within the industry with our MOTO wi4 solutions.

3G wireless technologies will also continue to gain momentum worldwide.

• On the UMTS front, the new standards incorporating HSDPA and HSUPA will enable a more cost-effective, high-speed data service and VoIP service.

TCS: *Is there a significant difference in the requirements of local and outside markets (assuming you mean domestic and international)?*

AN: Yes and no: We're operating in a global market. Users' increased demand for broadband, new services and seamless mobility is occurring across the globe and Motorola has a portfolio of solutions geared to specific operator environments whether they are wireless, wireline, incumbent, Greenfield, serving either metropolitan or rural areas.

1) Operators are looking to make their bundle sticky (basically providing FMC across as many devices in as many environments as possible). The idea is to make it more attractive to consumers to purchase a bundle rather than individual services. This is a developed market perspective and the drive is for compelling services/features that span the access technologies and devices.

2) In emerging markets, the focus is on cost (Opex and Capex) and the most efficient means to deploy basic telecommunications services to large populations with low ARPU.

TCS: *How much of communications products and solutions have dual usage as in commercial and military?*

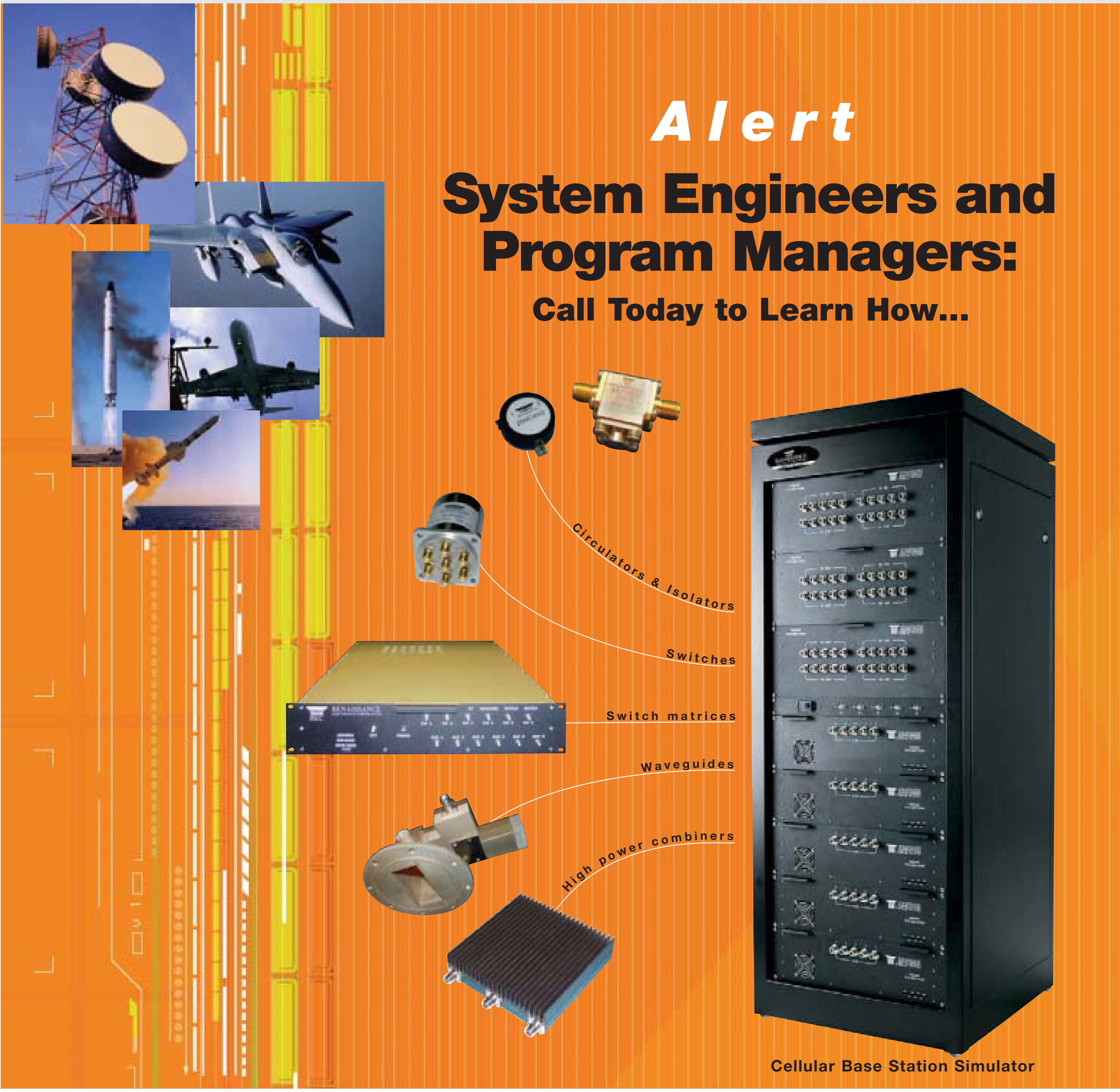
AN: We often see successful communications technologies developed in one domain gain adoption in the other. Both CDMA and mesh technologies were originally developed for military applications. From there, CDMA became a foundational technology in the growth of the world's commercial cellular service. Motorola's mesh technology enjoys wide commercial deployment providing a robust, scalable solution that reduces backhaul costs and enables peer-to-peer networking.

Increasingly, we are seeing the military adopt commercial technologies or slight extension on commercial technologies. This is true of common basic components (computing, protocols, etc.) as well as the basic RF technologies and standards. Costs and efficiencies will continue to drive this trend. Further, government organizations are looking to leverage public carrier networks in their solutions of the future and to find ways to allow their disparate networks to interoperate.

Today we also see public safety employees and the military use cellular communications as a backup and supplement to their dedicated and privately managed communication systems. The ongoing development of broadband data systems will provide an alternative service for voice and data for these groups while IMS systems being developed will provide IP connectivity between private and public systems to enable new coverage and service solutions for voice and data.

THE EUROPEAN PERSPECTIVE

The dynamic of the European wireless technologies market has changed dramatically since the height of the telecoms/Internet boom. Following the sell off of 3G licenses for large sums, the reality of the implementation of this technology has not matched original expectations and other competitive technologies have come to the fore. There is the continued emergence of WLAN, while more long-term effort is being put into the development of new technologies such as ultra-wideband (UWB) systems and WiMAX. In the telecommunications sector Europe is facing competition from the Far East with the emergence of low end products. Conversely though, the increased demand from developing countries for low cost products is a potential market that could be exploited. The wireless telecommunications market is international and thus, communication standards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must offer the potential of ubiquitous coverage worldwide. In the defence sector, however, there is more of a tendency for sourcing from European suppliers, particularly in reaction to US export regulation.

Jean-Luc Etienne
General Manager
Chelton Telecom & Microwave/Sales
France

MWJ: *As a supplier to the wireless systems market, what information would you like to see from the systems providers?*

J-LE: This very demanding market has forced many suppliers to define their approach in order to cope effectively. For example, delivery times are often shorter than the manufacturing cycle time, especially for components like ours. Consequently, Chelton Telecom & Microwave has established a strategy of working closely with its customers to understand their requirements and to respond to their need for flexibility. Also, market forecasting needs to be accurate to limit financial risks and to create a responsive supply chain. We realise that the unpredictability of the market can make forecasting difficult but we need as much information about business development as possible to cater for sudden variations in predicted forecasts.

Also, as technology is becoming more and more complex, system providers often look to high end companies like CTM that have the facility to satisfy these complex specifications and that have the capability to respond to integrated solutions such as modules or sub-systems. Through our divisional setup and with easy access to research and development, we can meet these requirements and are encouraging system providers to openly discuss and exchange ideas about their technical needs with our engineers.

MWJ: *How have evolving requirements of the communications sector affected your product mix/plans?*

J-LE: For a long time, cost reduction has been a major driving force in the telecom business and in the good times when demand was increasing this could be accepted. But now, when there is no increase in volume we can no longer make savings through better negotiations with the supplier or organise production better and must find alternative means such as cheaper technical solutions. Also, due to systems improvements, we have to be ready to accept tighter specifications to satisfy the trend for component development without any additional cost.

Meanwhile, Asian competitors are emerging with low end products. It is very difficult to compete with these companies, even with low cost, off shore manufacturing plants, as our overheads are at the western European level! Thus, our strategy, in the wireless market, is to concentrate on low cost, high end technology products. In this context, the ability of R&D to innovate is fundamental in order to present road maps for competitive and effective products. To achieve success, our plans include partnerships with suppliers to develop new materials and with universities to extend our R&D vision and capabilities.

MWJ: *What new technologies do you foresee impacting your product lines?*

J-LE: In the telecom field, time is a sensitive parameter, as the end customer needs to respond to its marketing plan to cover the increasing demands of subscribers. Therefore, great effort has gone into our R&D to produce SMD version isolators in order to follow the trend for pick and place automated systems. The second important factor is the production of low cost products and hopefully this technology also leads to achieving parallel low profile, low weight solutions, that will satisfy our customers.

For telecommunication satellites the ongoing development of new satellite spacebuses like Alphabus has driven our customers to develop high power payloads. Our market awareness and experience has led us to undertake concurrent engineering, design, testing and qualification of high power vacuum solutions. In the radar field, more and more radars are using active antenna with phased arrays, thus decreasing the need for waveguides. And, in the context of MMIC generalisation, discrete elements, mainly used for power handling limitation, are emerging. The idea is to integrate these discrete elements into modules, for example combining limiting diodes with ferrite-based components.

MWJ: *Is there a significant difference between domestic, European and International markets?*

J-LE: Yes, definitely, but it is not so true in the telecom field where the players tend to be globally based. Ideally CTM aims, "To be a domestic supplier in each country." However, operating internationally requires an efficient and reactive worldwide sales and support network. Achieving this target is my main mission, having just joined the company's management team.

In the military and space market, being a domestic supplier still offers a clear advantage, for proximity of service, confidentiality and even easier communication (speaking the same language in particular). On the other hand, Europe is being viewed as being more and more domestic. Through the construction of the European defence sector, European customers are clearly expressing their preference for European suppliers following the reinforcement of US export regulation. This evolution in the HiRel market is impacting on our strategy to assume a leading position in Europe.

MWJ: *How much of your communications products and solutions have dual usage for military and commercial applications?*

J-LE: Most of the technologies developed for communications products can be used for military applications after adjustment to be compliant with military operating conditions. This is especially true for L-band and S-band systems, for which the frequencies are not so different from cellular network frequencies. For X-band applications there are also common types of products with radio link systems. As an example, our circulators are being used for IFF application (1,030 to 1,090 MHz), which was initially developed for GSM BTS (925 to 960 MHz).

For military applications quantities are far less than those for commercial ones, so the investment for industrialization is not at the same level. Nevertheless, even if prices are higher in the military market, the technology, together with our expertise, enables our customers to achieve cost reduction targets along with having confidence in using high quality products.

Graham Martin
Business Development Director
Chipcon, Norway

MWJ: *As a supplier to the wireless systems market, what information would you like to see from the systems providers?*

GM: We would like to see greater transparency as to future frequency requirements, especially when new frequencies or bands are being considered. It would also be interesting to receive feedback on the requirements of the RF link (receiver sensitivity, output power, blocking performance, etc.), particularly which trade-offs are acceptable considering that very often a significant increase in RF link performance comes at a price, be it cost, power consumption, etc. The system power budget requirements and the required or preferred choice of microcontroller and memory technology are also significant in complete system on chip solutions.

In wireless mesh networking solutions it is important to understand the different requirements of the individual markets and again what trade-offs are possible. Many additional features are possible such as increased network security, increased network speed, faster network response times or enhancements such as over the air download. They often come at an additional cost (e.g., through increased system memory requirements) though, which may be acceptable to military, industrial or professional systems but not to certain consumer applications. Other helpful input would include future mechanical and environmental requirements. The reason being that due to the design cycle of new RF systems such decisions usually have to occur one to two years before we actually start delivering volume production, so it is important to receive such information at a very early stage.

MWJ: *How have evolving requirements of the communications sector affected your product mix/plans?*

GM: The requirements of the communications sector has lead us to focus on achieving high RF link reliability, low cost, low power and smaller sizes. In particular, the evolving and rapidly growing deployment of various wireless communications systems means that we have to put more emphasis on ensuring robust reliable wireless communication in very busy environments (interference). Two-way systems with messaging acknowledgment are increasing in popularity as well as systems with (adaptive) frequency hopping or DSSS, along with good receiver sensitivity and blocking properties.

Battery operation and increased miniaturization are driving size and power requirements. Therefore, complete RF system on chip solutions (radio, microcontroller, memory, peripherals on one IC) are helping to decrease size, power consumption and costs. In response, Chipcon has recently announced the fourth generation of products to meet all of these requirements.

MWJ: *What new technologies do you foresee impacting your product lines?*

GM: Low power, low cost and robust wireless sensor networking will play a major role in our future product lines based on IEEE 802.15.4 radio standards and ZigBee. This technology will be far reaching as it can be used in multiple applications including homeland security, home and building automation, health care, asset management, transportation, etc. Also significant is the utilisation of low power, low cost, robust RFICs in the ISM bands, which will replace older, simpler technologies such as most 27/49 MHz solutions and simple SAW filter solutions.

MWJ: *Is there a significant difference between domestic, European and International markets?*

GM: We see no significant difference in geographical requirements other than the variations in local frequency regulations in the sub 1 GHz ISM band. The differences are more in relation to market areas — consumer, industrial, automotive, etc.

MWJ: *How much of your communications products and solutions have dual usage for military and commercial applications?*

GM: All of our communications products can be used in commercial and military applications, although additional screening/selection would be necessary for certain military scenarios.

Wolfgang Bosch
CTO
Filtronic ICS, UK

MWJ: *As a supplier to the wireless systems market, what information would you like to see from the systems providers?*

WB: I would like to see better visibility of market trends in system design and solutions, particularly in relation to alternative competing technologies. This would be very helpful when making internal technology decisions. Trends for new frequency bands and their deployment by region and the overall effect of WiMAX on the cellular rollout plans are of great interest too. It would be useful for our subsystem designers to have a closer involvement earlier in the system design process, thus enabling them to make the appropriate system trade-offs at an earlier stage. Especially as designing to black box customer specifications, electrical and mechanical requirements is often very challenging and may adversely impact performance, time to market, cost and manufacturability. However, the inclusion of subsystems suppliers earlier in the design process would require the system providers to reveal some of the closely guarded IP in their systems.

MWJ: *How have evolving requirements of the communications sector affected your product mix/plans?*

WB: In general there is an increasing demand for higher data capacity solutions that require high yielding technologies, innovative designs and an increased level of integration, which has led to newly developed product lines matching the complex demands of emerging standards. In addition, the development of new communications systems and the growth of existing ones, necessitate the development of equipment that caters to several standards. As a result, although similar in function, the product mix is characterized by equipment development for a number of standards. There is also a noticeable increase in demand for products targeting developing countries with a greater price pressure. This has driven the product mix towards equipment at frequency bands suitable for these regions but at a lower price.

MWJ: *What new technologies do you foresee impacting your product lines?*

WB: Any new technology that impacts on the deployment of the communication systems, for which Filtronic designs RF equipment, will impact our product lines. For example, the deployment of emerging WiMAX and related systems will significantly affect our technology mix and our products. Specific technologies that impact on Filtronic's current product portfolio are emerging power semiconductors, DSP and implementation technologies that enable intelligent, highly efficiency and linear power amplifiers. Another area to watch is room temperature 'supercooled' filter technology that would be significant for our current product line.

MWJ: *Is there a significant difference between domestic, European and International markets?*

WB: As far as our product portfolio is concerned there is little difference between these markets and, in order to be successful, communication standards must offer the potential of ubiquitous coverage worldwide. As such, equipment developed for these systems address a worldwide market.

MWJ: *How much of your communications products and solutions have dual usage for military and commercial applications?*

WB: Typically our communications products are customized and are therefore specific to each application and do not find dual usage in military and commercial markets. However, the technology, design and manufacturing techniques and the expertise we have developed have a high degree of transportability and capabilities developed for one market often find application in others. ■

Measuring VSWR and Gain in Wireless Systems

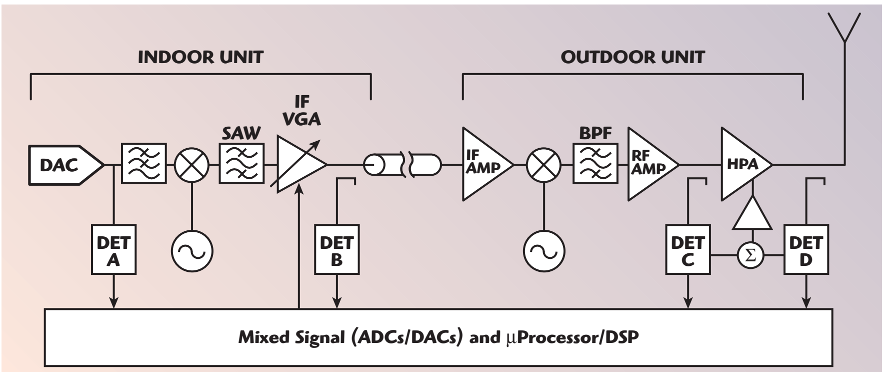
EAMON NASH
Analog Devices, Wilmington, MA

Measurement and control of gain and reflected power in wireless transmitters are critical auxiliary functions that are often overlooked. The power reflected back from an antenna is specified using either the voltage standing wave ratio (VSWR) or the reflection coefficient (also referred to as return loss). Poor VSWR can cause shadowing in a TV broadcast system as the signal reflected off the antenna reflects again off the power amplifier and is then rebroadcast. In wireless communications systems, shadowing will produce multi-path-like phenomena. While poor VSWR can degrade transmission quality, the catastrophic VSWR that results from damage to coaxial cable or to an antenna can, at its worst, destroy the transmitter. The gain of a signal chain is measured and controlled as part of the overall effort to regulate the transmitted power level. If too much or too little power is transmitted, the result will be either violation of emissions regulations or a poor quality link. The reflection coefficient is calculated by measuring the ratio between forward and reverse power. Gain, on the other hand, is calculated by measuring input and output power. The high commonality of hardware used to measure gain and VSWR can reduce overall component count. This article will focus on techniques that can be used to perform these in-situ measurements in wireless transmitters.

A TYPICAL WIRELESS TRANSMITTER

Figure 1 shows a typical wireless transmitter. It consists of mixed-signal base band circuitry, an up-converter (which generally includes one or more intermediate frequencies or IFs), amplifiers, filters and a power amplifier. These components may be located on different PCBs or may even be physically separated. In the example shown, an indoor unit is connected to an outdoor unit with a cable. In such a configuration, both units may be expected to have well defined, temperature-stable gains. Alternatively, each unit might be expected to deliver a well-defined output power. There are two different approaches to the ultimate goal of delivering a known power level to the antenna: power control or gain control.

With power control, the system relies on being able to precisely measure the output power (using detector D in this example). Once the output power has been measured, the gain of some component in the



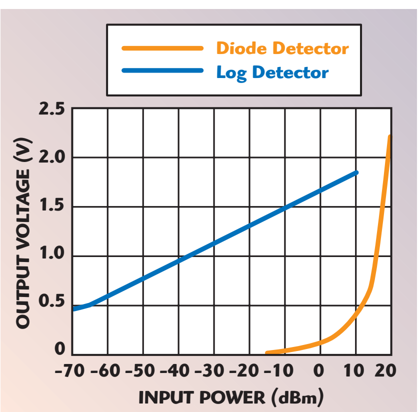
▲ *Fig. 1 Power control versus gain control.*

system (in this case, it might be the IF VGA) is varied until the correct output power is measured at the antenna. It is not necessary to know the gain of the circuit or the exact input signal amplitude; it is just a matter of varying the gain or input signal until the output power is correct. This approach is often (incorrectly) referred to as automatic gain control or AGC. To be correct, it should be referred to as automatic power control or APC since it is power not gain that is being precisely regulated.

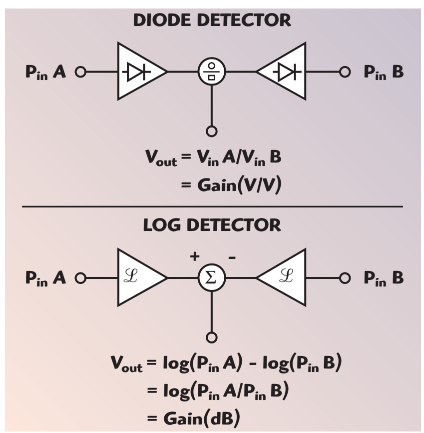
Gain control takes a different approach. Here, at least two power detectors are used to precisely regulate the gain of the complete signal chain or a part thereof. A precise input signal is then applied to the signal chain. A number of factors ultimately determine which approach is used. Power control requires only one power detector and makes sense in a non-configurable transmitter whose components are fixed. For example, power could be measured at the output of the RF HPA but adjustments would be made using the IF VGA. Gain control, on the other hand, may make more sense in a reconfigurable system whose components come from different vendors. In the example, the input power and output power of the HPA are being measured (using detectors C and D) so the gain can be regulated independent of the other blocks in the circuit. Note that the power/gain control loops can be all analog or microprocessor based. Gain control would be less practical in the example since the two required detector signals (detectors A and D) are physically remote from one another. A more practical approach would be to independently control the gain of the indoor and outdoor units.

RF DETECTORS

Until recently, most RF power detectors were built using a temperature-compensated half-wave rectifying diode circuit. These devices deliver an output voltage that is proportional to the input voltage over a limited dynamic range (typically 20 to 30 dB). As a result, the relationship between output voltage and input power in dBm is exponential (see ***Figure 2***). While the temperature stability of a temperature-compensated diode detector is excellent at high input powers (+10 to +15 dBm), it degrades significantly as the input drive is reduced. A log detector, on the other hand, delivers an output voltage proportional to the log of the input signal over a large dynamic range (up to 100 dB). The temperature stability is usually constant over the complete dynamic range. A log-responding device offers a key advantage in gain and VSWR measurement applications. In order to compute the gain or the reflection loss, the ratio of the two signal powers (either OUTPUT/INPUT or REVERSE/FORWARD) must be calculated (see ***Figure 3***). An analog divider must be used to perform this calculation with a linear-responding diode detector, but only simple subtraction is required when using a log-responding detector (since log (A/B) = log (A) – log (B)). A dual RF detector has an additional advantage compared to a discrete implementation. There is a natural tendency for two devices (RF detectors in this case) to behave similarly when they are fabricated on the same silicon wafer. Both devices will have similar temperature drift characteristics, for example. At the summing node, this drift will cancel to yield a more temperature-stable result.



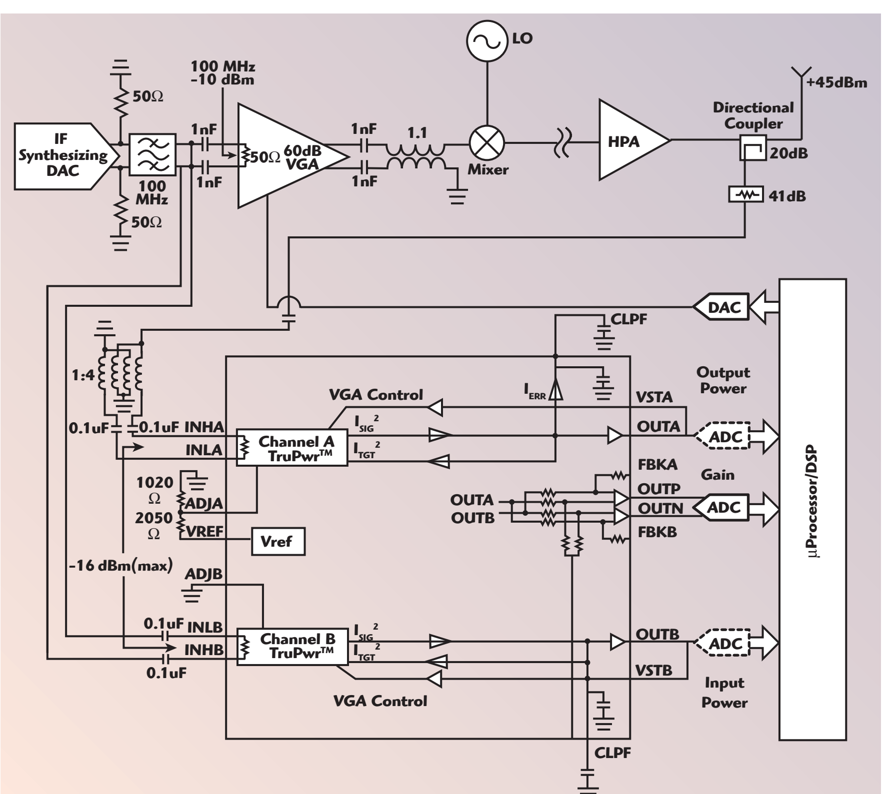
▲ *Fig. 2 Transfer functions of diode and log detectors.*



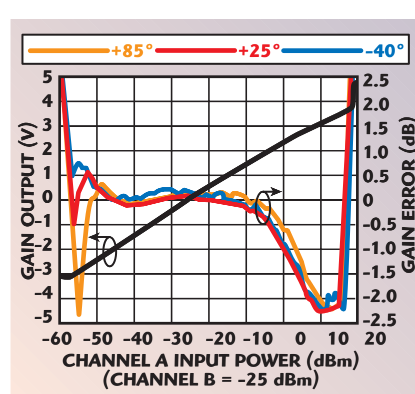
▲ *Fig. 3 Calculating the gain using diode and log detectors.*

GAIN MEASUREMENT EXAMPLE

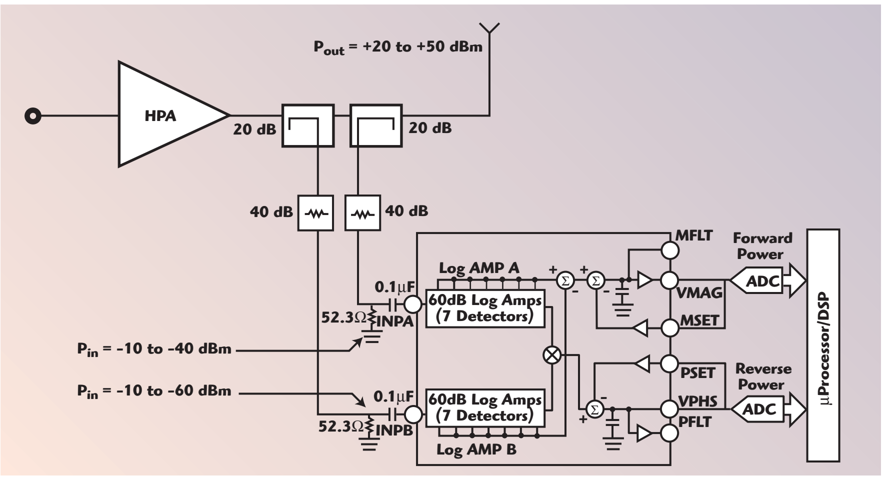
Figure 4 shows a transmitter whose gain is regulated using a dual power detector. The simplified transmit signal chain shown consists of a high performance IF-synthesizing DAC, a VGA, a mixer/upconverter and a high power amplifier. High performance DACs, such as the AD9786 and AD9779 that run at sampling frequencies up to 500 MSPS and beyond, are capable of synthesizing intermediate frequency outputs (100 MHz in this example). The output of the DAC is Nyquist filtered using a bandpass filter before being applied to an ADL5330 variable gain amplifier. Conveniently, the amplifier accepts a differential input that can be tied directly to the output of the differential filter. This, in turn, is tied to the DAC output. The VGA output is converted from differential to single-ended using a balun transformer, and is then applied to the ADL5350 mixer. After appropriate filtering (not shown), the signal is amplified and transmitted at a maximum output power level of 30 W (approximately +45 dBm).



▲ *Fig. 4 Gain control using a dual rms-responding log detector.*

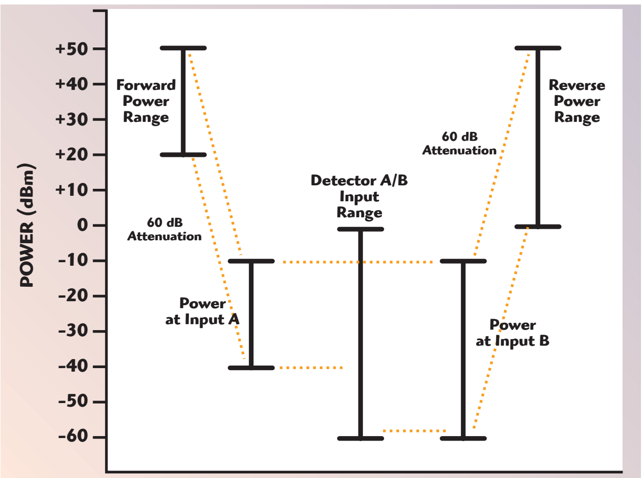


▲ *Fig. 5 Gain transfer function of a dual rms-responding log detector.*



▲ *Fig. 6 Return loss measurement using a dual log detector.*

The gain of the signal chain is measured by detecting the power at the DAC output and at the output of the HPA. The gain is then regulated by adjusting the gain of a VGA. At the DAC and PA outputs, a sample of the signal is taken and fed to the detectors. At the HPA output, a directional coupler is used to tap off some of the power going to the antenna. The transfer function of the AD8364 dual detector (see ***Figure 5***) shows that at the output frequency used (2140 MHz in this case), the detector has the best linearity and the most stable temperature drift at power levels below –10 dBm. Thus, the power coming from the directional coupler (+25 dBm max) must be attenuated before being applied to the detector. If maximizing the detector dynamic range is not critical to the application, the attenuation can be conservatively set at 41 dB so that the detector sees a maximum input power of –16 dBm. This still leaves about 34 dB of useful dynamic range over which the gain can be controlled. To detect the input power level at the DAC output, a directional coupler is impractical at this low frequency. In addition, directional coupling is not necessary since there will be little or no reflected signal at this point in the circuit. Furthermore, the power being delivered to the VGA is –10 dBm, so the power to be delivered to the detector is only 6 dB lower. Since the detector has an input impedance of 200 Ω and the VGA has an input impedance of 50 Ω, it quickly becomes clear that the two devices can simply be connected in parallel. With the same voltage present at both inputs, the 50 to 200 Ω impedance ratio will result in a convenient 6 dB power difference. Where high measurement precision is required, care must be paid to the temperature stability of the power detectors. This issue is further complicated if the temperature drift characteristics of the detectors change with frequency. The dual detector shown provides temperature compensation nodes. The temperature compensation is activated by connecting a voltage to the ADJ pins of each detector (this voltage can be conveniently derived using a resistor divider from the 2.5 V on-chip reference). No compensation is required for the low frequency input (ADJB is grounded), while a 1 V compensation voltage is required at ADJA to minimize temperature drift at 2.1 GHz. While the focus of the application circuit is gain measurement, it should be noted that input power and output power can also be measured. The outputs of the individual detectors are available and can be separately sampled. Because the detectors are log responding, their outputs can be simply subtracted to yield gain. This subtraction is performed on chip and the gain result is delivered as a differential voltage. The full-scale differential voltage is approximately ±4 V (biased up to 2.5 V) with a slope of 100 mV/dB. Digitizing with a 10-bit ADC with an LSB size of ~10 mV (±5 V full scale), 0.1 dB measurement resolution is achievable.



▲ *Fig. 7 Level planning for VSWR measurement using a dual log detector.*

VSWR MEASUREMENT EXAMPLE

A dual log detector can also be used to measure the reflection coefficient of an antenna. In ***Figure 6***, two directional couplers are used, one to measure the forward power and one to measure the reverse power. As in the previous example, additional attenuation is required before applying these signals to the detectors. The AD8302 dual detector has a measurement range of ±30 dB. The level planning used in this example is graphically depicted in ***Figure 7***. In this example, the expected output power range from the HPA is 30 dB, from +20 to +50 dBm. Over this power range, reflection coefficients from 0 dB (short or open load) up to –20 dB should be able to be accurately measured. Each of the AD8302's detectors has a nominal input range from 0 to –60 dBm. In this example, the maximum forward power of +50 dBm is padded down to –10 dBm at the detector input. When the HPA is transmitting at its lowest power level of +20 dBm, the detector sees a power of –40 dBm, still well within its input range.

The power from the reverse path is padded down by the same amount. This means that the system is capable of measuring reflected power up to 0 dB. This may not be necessary if the system is designed to shut down when the reflection coefficient degrades below a certain minimum (such as 10 dB), but it is permissible because the detector has so much dynamic range. For example, when the HPA is transmitting +20 dBm, the reverse path detector will see an input power of –60 dBm if the antenna has a return loss of 20 dB. The application circuit provides a direct reading of return loss, but no information is provided about the absolute forward or reverse power. If this information is required, the dual detector used in the gain control would be more useful because it would provide a measure of absolute forward and reflected power along with the reflection coefficient. The dual log detector used in the return loss measurement also provides a phase output. Because of the large gain in the main signal path of a progressive compression log amp, a limited (amplitude saturated) version of the input signal is a natural by product. These limiter outputs are multiplied together to yield a phase-detected output with a range of 180° centered around an ideal operating point of 90°. In a VSWR application, this information constitutes the phase angle of the reflected signal (with respect to the incident signal) and may be of use in optimizing the power delivered to the antenna.

AMPLIFIER GAIN MEASUREMENT USING A SINGLE LOG DETECTOR AND AN RF SWITCH

Figure 8 shows an alternative approach to gain measurement, which is also applicable to VSWR measurement. In this application, measuring and controlling the gain of a PA is desired. The PA in the example is running at 8 GHz and has an output power range from +20 to +50 dBm. This is a fixed-gain PA, so the output power is adjusted by changing input power. Two directional couplers are used to detect input and output power. However, there is only a single log detector so the two signals are alternately connected to the detector using a single-pole, double-throw RF switch. The AD8317 detector has a 0 to –50 dBm input range at this frequency. To measure the gain, the input and output powers are alternately measured and digitized. The results are then simply subtracted to yield gain. Once the gain is known, the digital control loop is completed by making any necessary adjustments to the gain of the PA via a bias adjustment. The level planning for this example is shown in ***Figure 9***. Attenuation is used so that the two input power levels at the RF switch are close together and within the input range of the detector.

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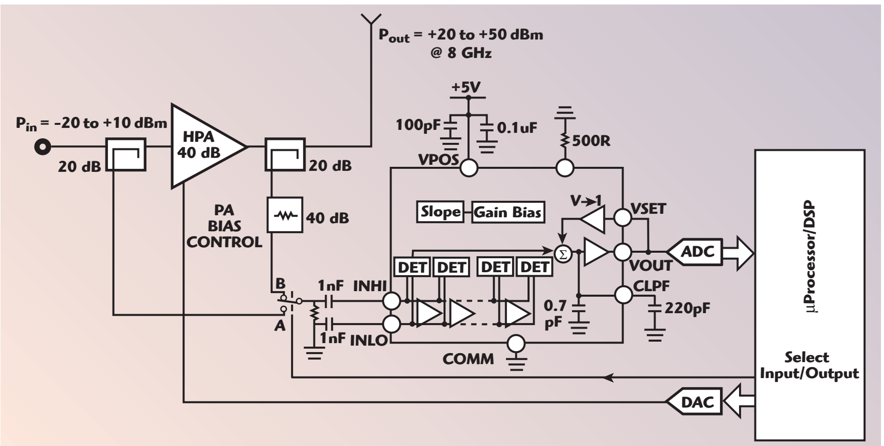

Fig. 8 Gain measurement using a single log detector.

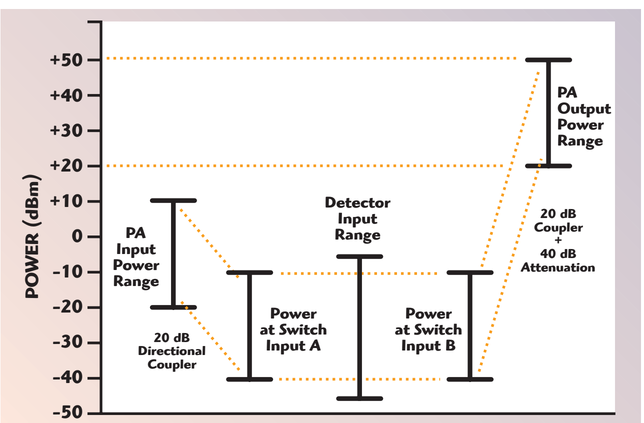

Fig. 9 Level planning for gain measurement using a single log detector.

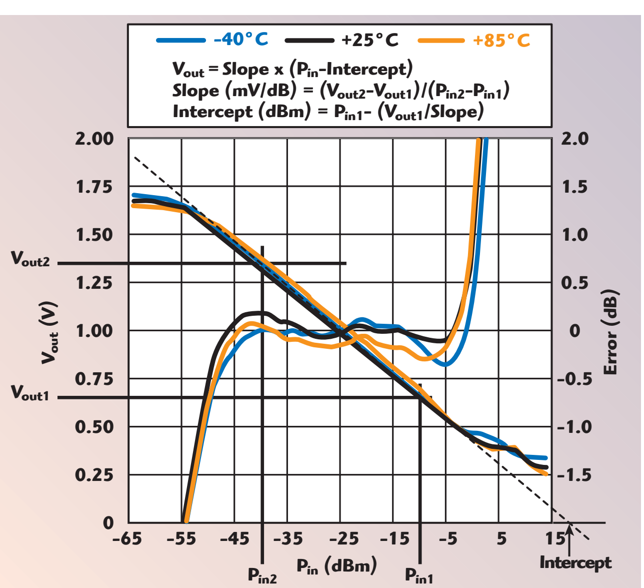

Fig. 10 Calibrating a log detector.

PRECISE GAIN MEASUREMENT WITHOUT FACTORY CALIBRATION

In addition to reducing component count, this gain measurement method has a number of interesting features. Because the same circuit is being used to measure input and output power, it is possible to make precise, temperature-stable gain measurements without ever calibrating the circuit. A look at the nominal transfer function of a log detector will help in understanding why (see ***Figure 10***).

$$V_{OUT1} = SLOPE \bullet (P_{IN1} - INTERCEPT)$$

To figure out the unknown, P_{IN}, the equation can be rewritten as

$$P_{IN1} = (V_{OUT1}/SLOPE) - INTERCEPT$$

Since gain is the difference in the measured input powers (the different attenuation levels of the two paths still have to be factored in), it can be written as

$$GAIN = (V_{OUT1} - V_{OUT2})/SLOPE$$

Therefore, the intercept of the detector is not required to calculate the gain. Even though the slope of a detector will change from device to device and over temperature, if V_{out1} and V_{out2} are close to each other (it can be done with good level planning and because of the finite input range of the detector), a typical value for the slope can be taken directly from the datasheet and used in the above calculation.

OUTPUT POWER MONITORING

In the gain measurement using a single log detector, the power is measured in order to calculate gain, so the system shown can also be used to monitor the output power. However, this cannot be done precisely without factory calibration. To calibrate the circuit, the antenna must be temporarily replaced by a power meter. The output power and detector voltages are then measured at two points within the linear range of the detector. These numbers would then be used to calculate the slope and intercept of the detector. For optimum precision, the detector includes a temperature compensation pin. A resistor is connected between this pin and ground to reduce the temperature drift to approximately ±0.5 dB at the frequency of operation (8 GHz in the example shown). As a result, it is not necessary to do any additional calibration over temperature.

CONCLUSION

Because of their linear-in-dB transfer function, log amplifiers can be easily used to measure gain and return loss. When dual devices are used, very high measurement precision is achievable. In some cases, this can be achieved without factory calibration. In all cases, careful power level planning is necessary so that the power detectors are driven at power levels that offer good linearity and temperature stability. ■

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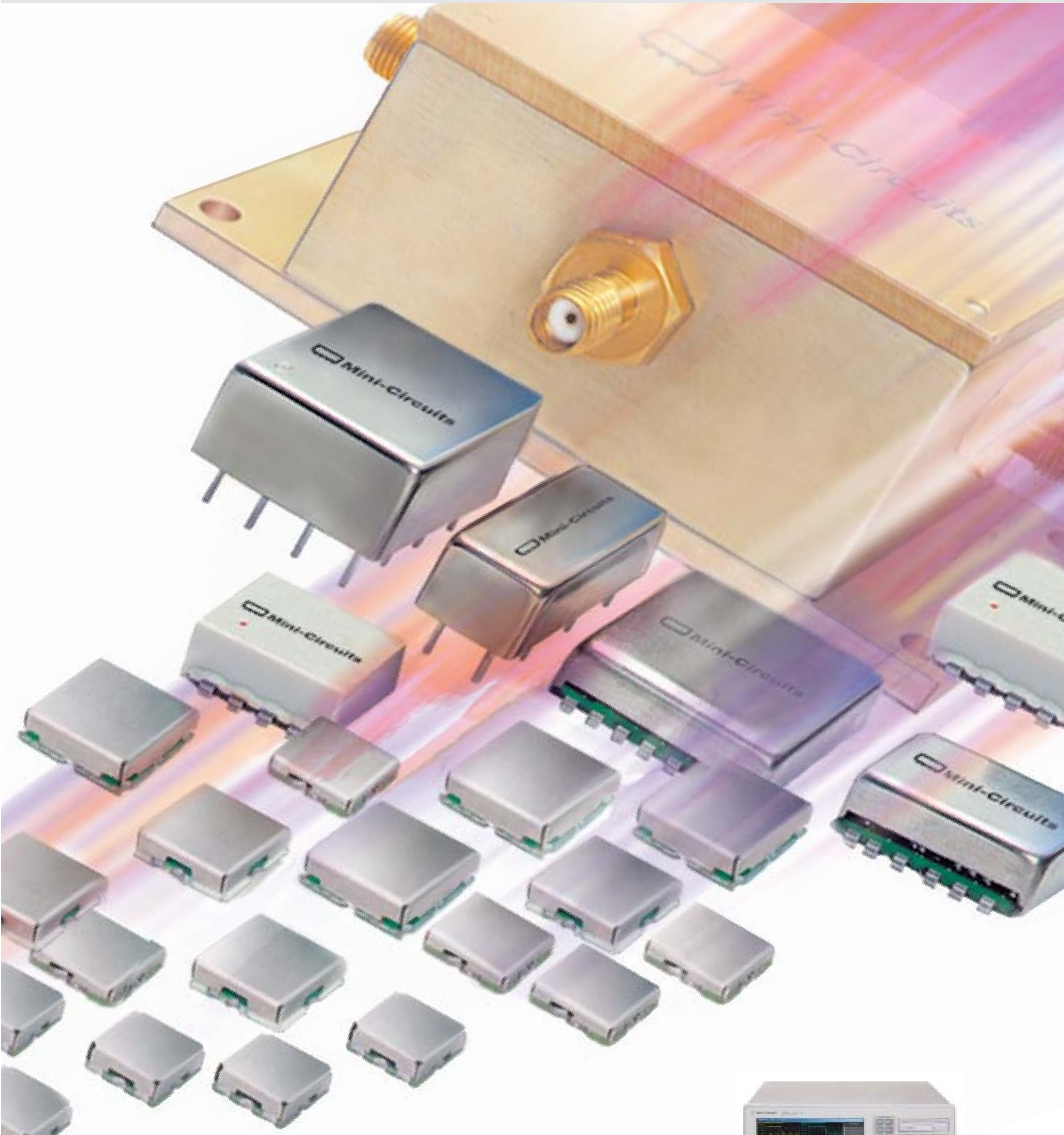

Mini-Circuits

UHF RFID Industry Growth Powered by RF Technology

TOM CAMERON
WJ Communications Inc., Stittsville, Ontario, Canada

RF identification (better known by the acronym RFID), while not a new technology, has experienced a recent resurgence due to new applications in the supply chain management, coupled with advances in the technology. The identification of an object using RF interrogation has been in use almost as long as radar but until recently the technology has not received much attention in the commercial sector due to the high cost of implementation. In the early 1990s, low cost inductively coupled tags saw mass deployment in applications such as access control. Today, UHF RFID is undergoing a revolution driven by the disruptive changes in technology, which are allowing low cost tag designs, and readers that are moving from cumbersome systems to integrated circuits. This has resulted in the passage of new standards, the assignment of global frequency bands in the UHF range and deployment of pilots worldwide.

RFID systems operate at several popular frequencies worldwide. Inductive loop systems operate at low frequency (LF) at either 125 or 134 kHz; HF systems operate worldwide at 13.56 MHz. Systems at these frequencies are very robust and insensitive to obstructions such as the human body. However, they are limited by the reach of the inductive loop and hence work to approximately 1 meter range. UHF systems are in the radiative domain and provide a much longer reach, typically 3 to 10 m for a passive tag and 100 m for an active or battery assisted tag. These systems operate in the ISM bands between 860 and 960 MHz and also at 2.4 GHz worldwide.

The RFID systems attracting the most industry attention recently are the passive backscatter systems operating in the UHF band. This is partially due to the efforts of EPC-Global to standardize the technology while simultaneously being driven by applications in supply chain management. In late 2003, Wal-Mart delivered a mandate to its top-100 suppliers, requiring them to provide RFID tags on all pallets and cases or cartons delivered to their distribution centers by January 2005. Other large retailers in the US such as Target and large European retailers such as Metro and Tesco joined Wal-

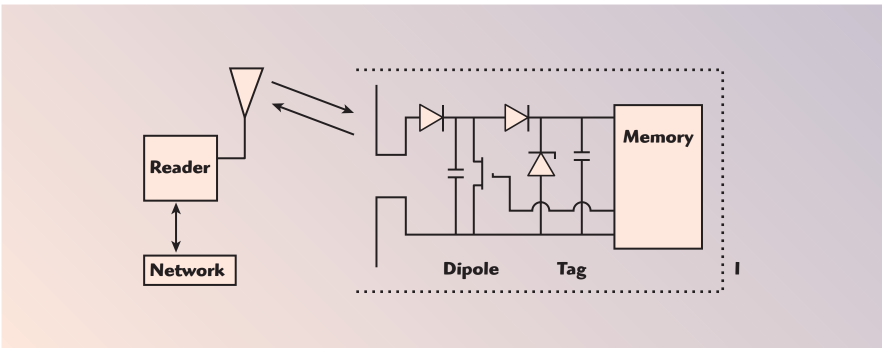


Fig. 1 *Block diagram of a typical UHF RFID system.*

Mart. In conjunction with these large retailers, the US Department of Defense (DoD) has initiated the adoption of UHF RFID, further fueling the commercial interest in this technology. The supply chain industry predicts that, as the cost of tags works its way down to just a few cents, billions of tags and millions of readers will be deployed, enabling identification of every mobile item. According to Venture Development Corp. (VDC), the $1.7 B global RFID market in 2004 will grow to $5.9 B in 2008. Such staggering growth will be fuelled both by the ratification of standards and technological breakthroughs at all levels of the RFID infrastructure.

The focus of this article is the challenge that faces the RF engineer in the RFID industry. Similar to the cellular phone industry in the late 1980s, the RFID infrastructure must evolve to reduce cost and size while adding network functionality and user driven applications. In the same manner that bulky mobile phones of yesteryear have become miniature personal computing devices, the coming years will see RFID readers reduced to inexpensive, ubiquitous network interface devices that are a part of our daily lives. As the reader technology becomes implemented in high density IC processes and the power consumption is reduced through new architectures, the readers will become fully mobile, resulting in new applications inconceivable beyond the supply chain.

RFID systems, whether passive, active or semi-passive, all consist of the same common components, a reader and a tag (see ***Figure 1***). The tag memory contains data formatted to identify the object. The reader communicates with the tag to read and write data to and from the tag's memory. The simplified block diagram of a tag shows its major components. The tag contains circuits to both rectify DC power from the incoming RF signal as well as to detect and extract the information modulated on the signal. The antenna load is a controlled resistance, illustrated as a transistor that changes the impedance of the dipole, enabling the backscatter. The backscatter technique results in a very low cost tag but a more complicated reader to achieve the necessary read range.

The tag IC is mounted on a carrier — known as a strap — and subsequently bonded to the antenna to form the fully assembled tag. Several examples of tags are shown in ***Figure 2***, illustrating the variety of antenna configurations available. Given the sensitivity of RF signal launch and propagation to environmental factors, much effort is being spent on efficient, low cost antennas and substrate materials. While the expectation is that a tag will cost a few cents in very large volume production, today that cost may still be an order of magnitude from the projected target. The cost of tags has been driven down by several factors. The chip itself is very small, enabled by modern CMOS technology. The mounting of the die on a carrier has been made very inexpensive and capable of large volume by either flip-chip or by other innovative techniques such as the Alien Technologies Fluidic Self Assembly process.

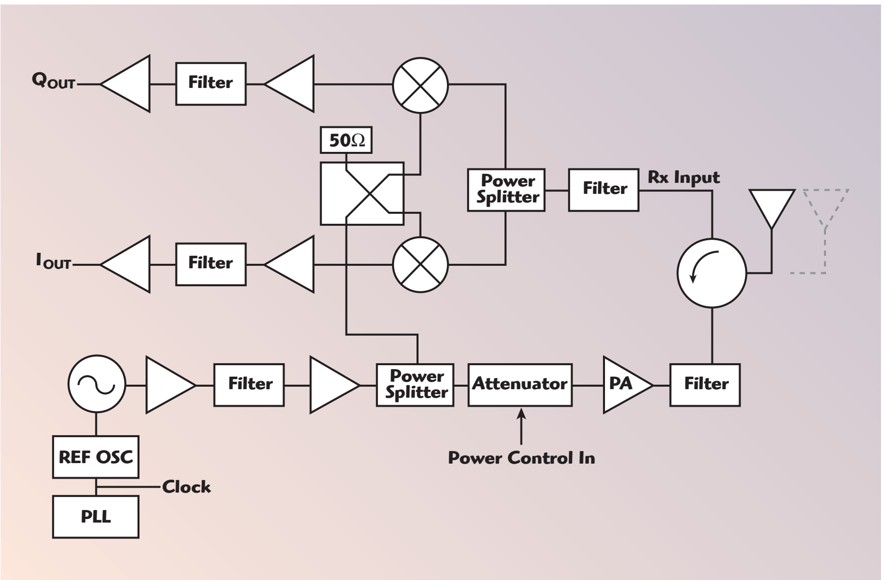
Fig. 2 *Examples of UHF passive RFID tags.*

TABLE I

FREQUENCY BANDS ASSIGNED TO UHF RFID BY REGION

Region/Country	*Frequency (MHz)*
Europe	865–868
South Korea	866–869 and 923–925
US and Canada	902–928
Japan	952–954

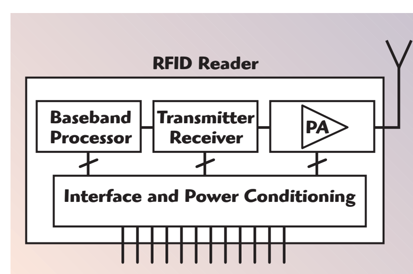
To completely understand the UHF RFID landscape, one must be aware of the various frequencies of operation and the protocols. The frequency bands, as signed internationally by region, are listed in ***Table 1***. While these are the presently known bands, several other countries are in the process of allocating frequencies internationally in the ISM bands for UHF RFID. In addition to frequency bands being assigned, data protocols are also being standardized. This is a tremendous step towards unifying all the disparate systems being used around the world and for helping to drop the overall cost of these systems. Generally, in North America, EPC standards are predominant. These protocols have been named Class 1 and Class 0



▲ *Fig. 3 Generic RFID reader of RF architecture.*

▲ *Fig. 4 WJ Communications' SR2200 portal reader.*

▲ *Fig. 5 PCMCIA RFID reader engine.*

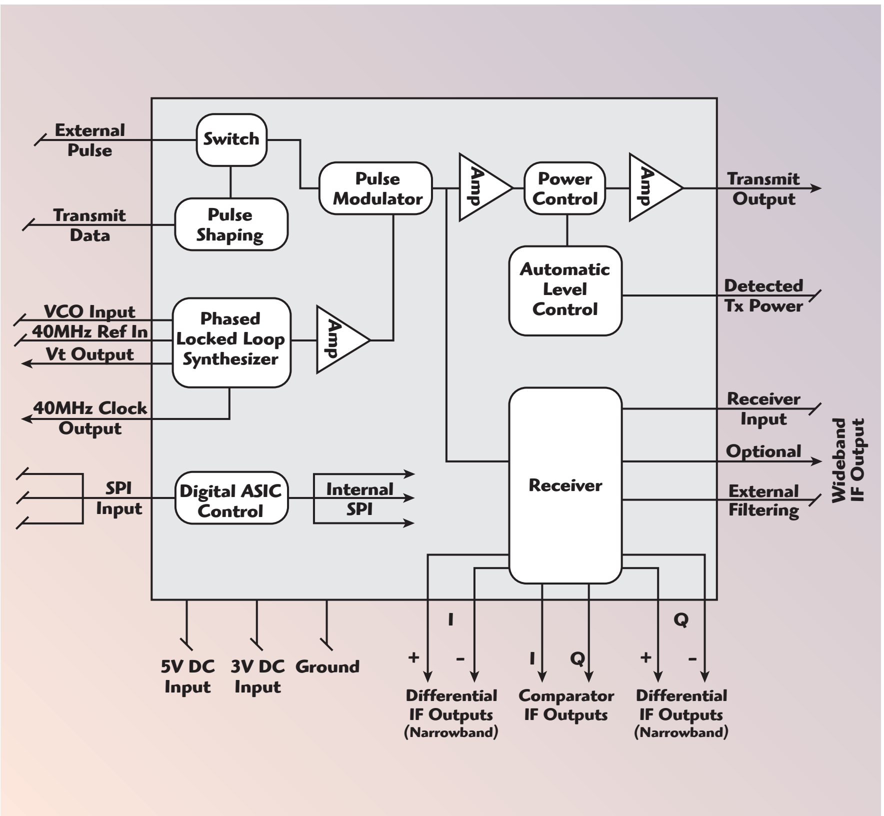


▲ *Fig. 6 Chipset-based RFID reader architecture.*

(Class 0+ is an addendum of Class 0), each having its own merits. Recently, the second generation of class 1 has been ratified by EPC-Global, known as Class 1 Gen 2; it is expected that the North American industry will converge on this standard. In Europe and internationally, the ISO-18006 standard has been the protocol of choice. Class 1 Gen 2 is presently being considered for adoption into the ISO standards process and it is expected that in the future the EPC and ISO protocols will converge into a single worldwide standard. While it is not necessary to describe the various protocols here, it is important to understand that there are several protocols to be supported, all requiring different reader hardware. The designer must be aware of this and understand the unique specifications of both the protocol and air interface being implemented. All of these issues add to the complexity of the reader design.

RFID readers, otherwise known as transponders, communicate with the tags to program them and read their data. Due to the simplistic nature of the tags, most of the link budget gains come from the reader and antenna, making today's readers complex, bulky and expensive. The reader transmits at powers up to 4 W EIRP and the reader must be capable of discerning a received signal of the order of –90 dBm, in an environment susceptible to fading, multipath and interference. In addition to this link budget challenge, an ideal reader should be capable of reading several protocols in real time at any of the UHF ISM bands, while frequency hopping to meet regulatory masks. This is quite a challenge for the radio designer. Today's RFID readers are both simplistic and elegant in nature. The general architecture, as shown in ***Figure 3***, utilizes direct conversion on both the up- and down-links. The transmit chain consists of a modulator of some sort, whether IQ or direct modulation of the carrier. The receive chain is a homodyne receiver with filtering defined in the channel, filtering defined in the baseband or the digital circuitry. When designed to operate for two or more protocols, the readers must be agile, hence they are software-or firmware-controlled radios. The direct conversion architecture can only be realized with a very clean synthesizer and low coupling between the transmit and receive chains. An example of an RFID reader is shown in ***Figure 4***. The SR2200 reader from WJ Communications is indicative of product offerings available for portal applications. The SR2200 is a multi-protocol UHF reader supporting Class 0, 0+ and Class 1, and is Class 1 Gen 2 ready. It has four pairs of transmit and receive connectors to support eight antennas. This type of reader is used at dock doors and around conveyors to quickly and accurately identify large number of items passing through.

In order for readers to realize mass deployment, the size and cost must decrease over the coming years and their functionality must increase. Today, the supply



▲ *Fig. 7 WJ Communications' UHF RFID transceiver chip functional block diagram.*

chain is focused on pallet and carton-level tagging but as item-level tagging becomes practical, pressure will increase on the evolution of reader technology to ensure that applications such as smart shelves are economically practical. This scale of economy can only be accomplished through integration. It is predicted that the RFID reader hardware will evolve in a manner similar to the cellular phone and WiFi hardware. Already, reader form-factors are being reduced based on available semiconductor technology such as the PCMCIA reader engine shown in ***Figure 5***. This engine is very compact and can transmit 1 W on two antenna ports. Unlike the larger portal reader, the PCMCIA reader transmits and receives on the same port, so two fully duplexed antennas may be used. This type of product is suitable for small form factor applications such as hand-held readers or RFID label print-

Continued on page 42

A WiMAX Design Library for System-level RF/DSP Co-design

AGILENT TECHNOLOGIES,
EESOF EDA DIVISION, SANTA ROSA, CA

The IEEE 802.16-2004 standard, generally referred to as WiMAX, specifies air interfaces for broadband wireless access (BWA) systems. The standard is expected to energize the BWA industry and create opportunities to deploy systems in applications that were previously cost-prohibitive. WiMAX enables multiple services in a wireless metropolitan area network (WMAN), such as wireless backhaul for telecommunications, high bandwidth/high reliability remote connectivity, E1/T1 replacements for small- and medium-sized businesses, and residential "wireless DSL" for broadband Internet at home.

WiMAX supports fixed broadband wireless access for both licensed and unlicensed spectra in the 2 to 11 GHz range. The mandatory PHY mode is 256-point FFT orthogonal frequency division multiplexing (OFDM). The WiMAX Forum certifies equipment supporting the OFDM PHY model.

WiFi 802.11a and 802.11g also use OFDM and have established an excellent performance record for robust wireless networking. WiFi uses 64-point OFDM. The much larger number of carriers for WiMAX helps it achieve greater range, because a receiver using 256 OFDM can tolerate delay spreads of up to 10 times that of systems using 64 OFDM. Also, 256 OFDM provides good non-line-of-sight capability.

THE WIMAX DESIGN EXPLORATION LIBRARY

With competition heating up for WiMAX-related products, system designers are looking into EDA tools that can help them design products that achieve the best power performance at the least cost. This is challenging, especially as designers look for optimal system performance. Such efforts require a good understanding of the system design and the ability to optimize individual system block specifications. System blocks contain both analog/RF and DSP components. The WiMAX design exploration library provides preconfigured simulation

TABLE I

SIGNAL-TO-NOISE RATIO, BIT ERROR RATE AND FRAME ERROR RATE RESULTS FOR AN EXAMPLE RECEIVER

Receiver SNR	BER	FER
4	0.000	0.000
0	0.001	0.429

setups, signal sources and fully coded BER analysis for simulation of the circuitry used in BWA designs. It speeds the development cycle by allowing system designers to analyze a system's performance before all of its components are designed. It works within the ADS 2005A environment and with the Agilent Ptolemy simulator to streamline design and verification of OFDM-based, last-mile service designs. The WiMAX library can also be imported into Agilent's RF design environment (RFDE), allowing RFIC designers to access WiMAX test benches within the Cadence Virtuo Custom IC platform through links developed as part of the ongoing alliance between Agilent Technologies and Cadence Design Systems. Transmitter measurements performed for both uplink and downlink subframes include EVM, constellation, CCDF, spectrum mask, waveform and spectral flatness.

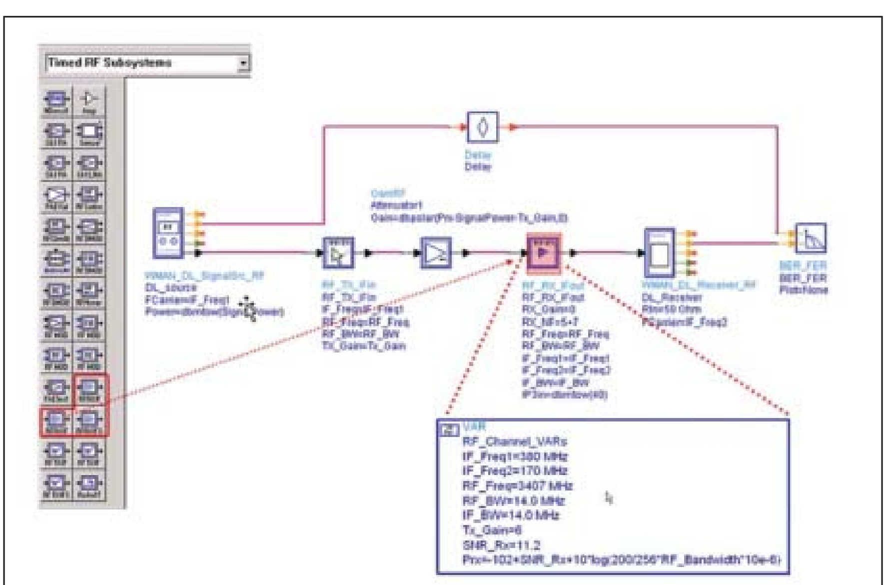
Receiver measurements performed for both uplink and downlink subframes include receiver sensitivity, BER and PER in AWGN, BER and PER in fading channel, and adjacent channel rejection.

The fixed wireless access propagation channel model is included in the WiMAX Design Exploration Library. The multi-path fading is modelled as a tapped-delay line representing 6 SUI (Stanford University Interim) channel models.

RECEIVER SENSITIVITY TEST BENCH

One of the test benches in the WiMAX Design Exploration Library is the receiver sensitivity measurement. The standard dictates a BER/FER limit based on a receiver signal-to-noise ratio (SNR) and a maximum noise figure (NF) of 12 dB.

The test bench can introduce non-idealities due to the architecture of the circuits in the RF or DSP section. The measurement results can then easily identify circuit architectures that meet the standard requirements. For example, by changing the receiver SNR, one can realize the minimum received power needed to meet overall system frame error rate (FER). ***Table 1*** shows the result of a sweep of a receiver SNR based on a double conversion receiver architecture. The RF and IF parameters shown in ***Figure 1*** can easily be changed to adapt to the specific data rate of the WiMAX system. Furthermore, designers can select a different RF_RX_IF architecture to perform trade-off analysis. Timed components in Agilent Ptolemy are time-based signals that carry I, Q, Δt (time step resolution) and FCarrier information. This powerful signal representation is based on the timed synchronous data flow (Timed SDF) engine in Agilent Ptolemy. In this example, a double conversion receiver versus low IF architectures can be studied using Agilent Ptolemy built-in components from Timed RF subsystems selection available in Agilent Ptolemy, within the receiver test bench of the WiMAX design exploration library.

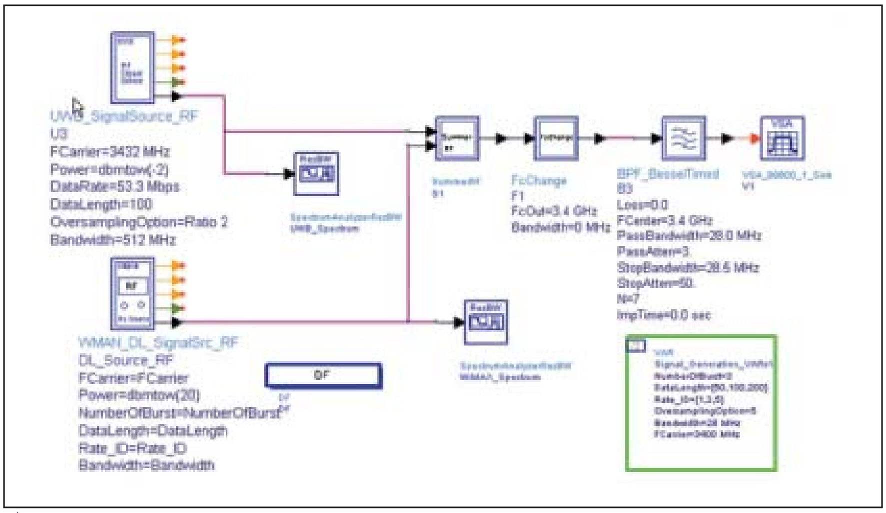


▲ *Fig. 1 RF architecture trade-off analysis in the WiMAX design exploration library.*

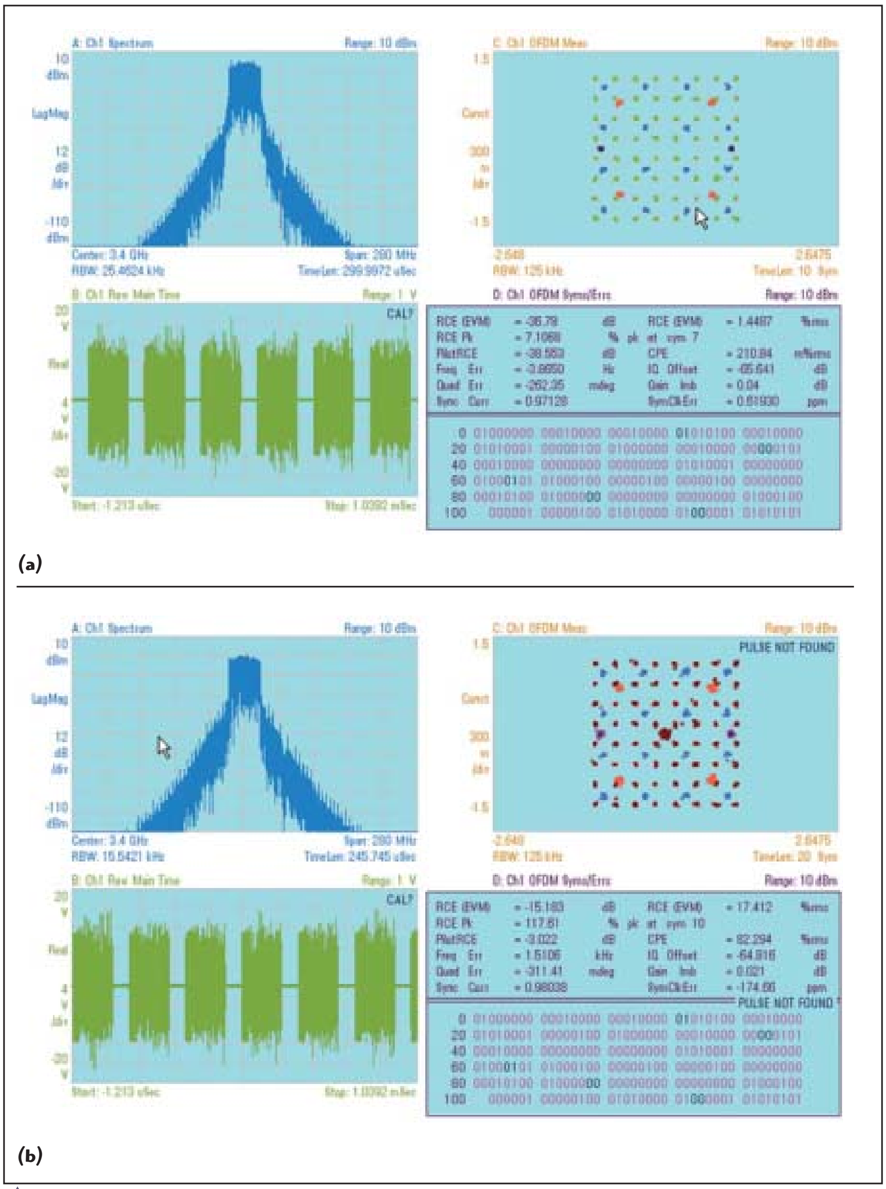
UWB INTERFERENCE ON WMAN SIGNAL

The actual operating environment of WiMAX transmitters and receivers includes transmitters based on other standards. Transmitters using the UWB standard are a potential source of interference with WiMAX, and one with which designers must be concerned. The interference of UWB signals in the 3.4 GHz band is of special concern when using the 802.16-2004 technology. Recent industry studies show that the UWB signal must be detected, and possibly moved to another RF frequency, to avoid destructive interference with fixed broadband wireless devices.[1]

The flexibility of Agilent Ptolemy Design Libraries allows designers to set up interference signals from various sources at different frequencies and signals. With the software that is available with Agilent's vector signal analyzer (VSA) instrumentation, designers can quickly test the conditions that are destructive within the simulation environment.[2] The power level setting for UWB interference, centered at 3432 MHz, can be set from the variables indicated on the UWB_Signal Source_RF component, as shown



▲ *Fig. 2 UWB interference on the WiMAX signal using VSA software running inside the Agilent Ptolemy simulator.*



▲ *Fig. 3 VSA software analysis with the UWB interference signal set at –9.9 dBm (a) and –2 dBm (b).*

in ***Figure 2***. After summing the UWB interference with the WiMAX signal, the signal centered at 3.4 GHz will be filtered and then analyzed with the VSA software running in Agilent Ptolemy.

The results shown in ***Figure 3*** indicate the increase in EVM due to the power increase in the UWB signal, from –9.9 to –2 dBm. The RF and base band filters from the Agilent Ptolemy filter design library can be selected to determine the filter characteristics that provide optimal interference rejection. Trade-offs between interference rejection and receiver sensitivity can be evaluated quickly with these simulations.

CONCLUSION

The main challenge facing communications system designers is performing RF architecture selection, optimization and verification concurrently with digital base band design to make intelligent trade-off decisions and not over-design the system. Agilent Ptolemy provides a unique capability where timing synchronization (Timed SDF) enables digital base band models to be co-simulated with high fidelity RF behavioral models. Designers need to use this environment to catch problems early in the design cycle to prevent unnecessary hardware iterations later. The WiMAX Design Exploration Library offered with ADS 2005A enables designers to achieve easy design performance analysis of their RF and DSP components for WiMAX system designs.[3] ■

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Model	Passband (MHz)	fco, (MHz) Nom. (Loss 3dB) Typ.	Stopband (MHz) (Loss >20dB) Min.	No. Of Sections	Price $ ea. Qty. 10
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LFCN-95	DC-95	165	220	7	3.99
LFCN-105	DC-105	180	250	7	3.99
LFCN-120	DC-120	195	270	7	3.99
LFCN-225	DC-225	350	460	7	2.99
LFCN-320	DC-320	460	560	7	2.99
LFCN-400	DC-400	560	660	7	2.99
LFCN-490	DC-490	650	800	7	2.99
LFCN-530	DC-530	700	820	7	2.99
LFCN-575	DC-575	770	900	7	2.99
LFCN-630	DC-630	830	1000	7	2.99
LFCN-800	DC-800	990	1400	5	1.99
LFCN-900	DC-850	1075	1275	7	1.99
LFCN-1000	DC-1000	1300	1550	7	1.99
LFCN-1200	DC-1200	1530	1865	7	1.99
LFCN-1325	DC-1325	1560	2100	5	1.99
LFCN-1400	DC-1400	1700	2015	7	2.99
LFCN-1450	DC-1450	1825	2025	7	2.99
LFCN-1500	DC-1500	1825	2100	7	2.99
LFCN-1525	DC-1525	1750	2040	7	2.99
LFCN-1575	DC-1575	1875	2175	7	2.99
LFCN-1700	DC-1700	2050	2375	7	1.99
LFCN-1800	DC-1800	2125	2425	7	2.99
LFCN-2000	DC-2000	2275	3000	5	1.99
LFCN-2250	DC-2250	2575	2900	7	1.99
LFCN-2400	DC-2400	2800	3600	5	1.99
LFCN-2500	DC-2500	3075	3675	7	1.99
LFCN-2600	DC-2600	3125	3750	7	1.99
LFCN-2750	DC-2750	3150	4000	7	1.99
LFCN-2850	DC-2850	3300	4000	7	1.99
LFCN-3000	DC-3000	3600	4550	7	1.99
LFCN-5000	DC-5000	5580	6800	7	1.99
LFCN-6000	DC-6000	6800	8500	7	1.99
LFCN-6700	DC-6700	7600	9300	7	1.99
HFCN-650	850-2490	650	480	7	1.99
HFCN-740	900-2800	740	550	7	1.99
HFCN-880	1060-3200	880	640	7	1.99
HFCN-1200	1380-4600	1180	940	7	1.99
HFCN-1300	1510-5000	1300	930	7	1.99
HFCN-1320	1700-5000	1320	1060	7	1.99
HFCN-1500	1850-5500	1550	1250	7	1.99
HFCN-1600	1950-5000	1600	1290	7	1.99
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HFCN-1810	2250-4750	1810	1480	7	1.99
HFCN-2000	2410-6250	2000	1530	7	1.99
HFCN-2100	2500-6000	2100	1530	7	1.99
HFCN-2275	2640-7000	2275	1770	7	1.99
HFCN-2700	3000-6500	2500	1800	7	1.99
● HFCN-5500	6000-11500	5500	4500	5	2.99

LFCN = Low Pass, HFCN = High Pass

LFCN Models: U.S. Patent #6,943,646 except LFCN-800,-1325,-2000 & -2400.

*For applications requiring DC voltage to be applied to the input or output, *add suffix letter "D" to model number* (DC resistance to ground is 100 megaohms min.) *and add $0.50 to unit price.*

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RF/IF MICROWAVE COMPONENTS

393 Rev R

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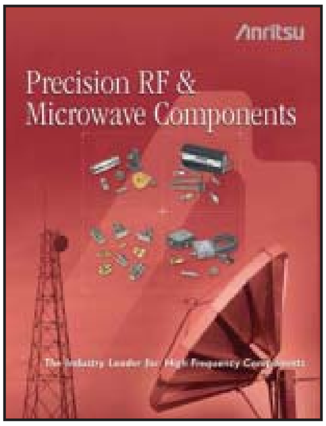


Components Catalog

This catalog features the company's design and production of precision RF and microwave components. The company has created and trademarked the V and K connectors and continues to develop innovative connector products. In addition, the company also manufactures assemblies and components to meet specific customer requirements. Also offered are a wide range of high speed devices designed to meet the electrical requirements of next generation optical network communications.

Anritsu Co.,
Morgan Hill, CA (408) 778-2000, www.anritsu.com.

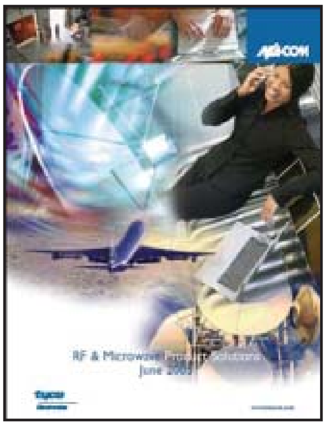
RS No. 325

Product Data Sheet

The Microwave Office 2006 Design Suite data sheet details the new features and benefits offered in this latest release of the company's RF/microwave EDA software suite. The data sheet describes the integration into the AWR design environment of several new technologies, including the APLAC foundry-approved RF simulation technology, a new EM editor, EMediacy, and new simulation filters and folder management features that streamline the design process.

Applied Wave Research Inc.,
El Segundo, CA (310) 726-3000, www.appwave.com.

RS No. 326

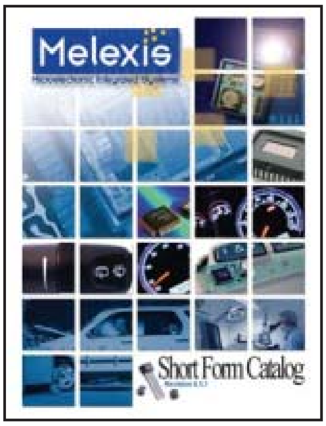


Product Solutions Guide

This product solutions guide contains information on over 1000 standard microwave and RF products. It contains parameters for each part number organized by product category. RoHS compliant products and off-the-shelf distribution products are noted for easy reference. Most standard products can be up-screened to meet defense and space applications or parts can be customized to meet certain specifications.

M/A-COM,
Lowell, MA (800) 366-2266, www.macom.com.

RS No. 327



Short Form Catalog

This short form catalog features the company's most popular standard product devices that are currently available. The company produces and delivers semiconductor/IC products that enable customers to achieve product development goals. Products include programmable sensors, sensor interface devices and microcontrollers that provide customers with competitive advantage in system cost, capability and flexibility.

Melexis Inc.,
Concord, NH (603) 223-2362, www.melexis.com.

RS No. 328

Amplifier Product Catalog

This 74-page AFS series amplifier product catalog includes technical specifications and typical test data of the company's wide array of low noise and high power amplifiers. In addition, the catalog covers many special amplifier designs such as surface-mount, variable gain, dual output, temperature-compensated, cryogenic, high intercept, limiting and pulse modulated power gated designs.

MITEQ Inc.,
Hauppauge, NY (631) 436-7400, www.miteq.com.

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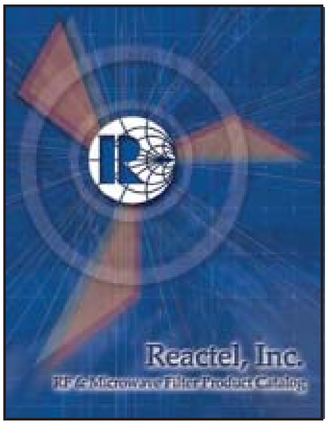
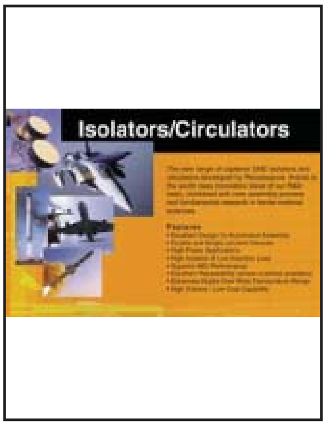


RF and Microwave Filter Catalog

This full line catalog features high reliability filters, multiplexers and switched filter banks covering DC to 50 GHz for wireless, commercial and military applications. The catalog is available for download from the company's Web site or e-mail catalog@reactel.com for a free copy. The company is ISO-9001:2000 certified, and welcomes a clients most difficult filter requirement.

Reactel Inc.,
Gaithersburg, MD (301) 519-3660, www.reactel.com.

RS No. 330



Product Catalog

This catalog features the company's isolators and circulators that provide performance able to handle the most stringent environmental conditions. These isolators and circulators are offered in different product configurations from 88 MHz to 40 GHz solving today's requirements for lower insertion loss, higher power, greater isolation and low intermodulation products.

Renaissance Electronics Corp.,
Harvard, MA (978) 772-7774, www.rec-usa.com.

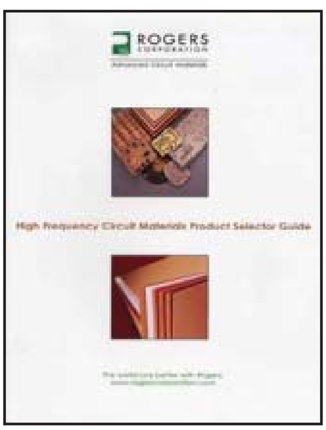
RS No. 331

Product Brochure

This product brochure highlights the company's complete line of high performance, low power Bluetooth® system-on-chip solutions. The company offers feature-rich HCI software, upper layer protocol stack software, all major profiles and the development tools that are needed to speed a design to market.

RF Micro Devices Inc.,
Greensboro, NC (336) 678-5570, www.rfmd.com.

RS No. 332

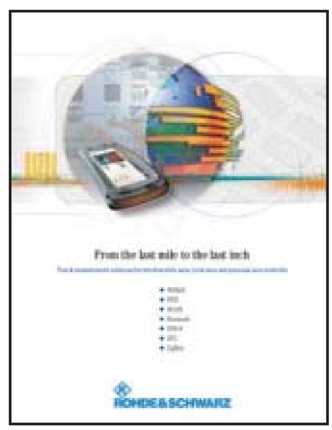


Product Selector Guide

This product selector guide highlights the company's extensive selection of high frequency circuit materials for many types of applications. The materials range from low loss, others are temperature stable, some are rigid while others are suited for low cost commercial applications. The information contained in this guide assists a reader in designing with the company's laminates.

Rogers Corp.,
Rogers, CT (860) 774-9605, www.rogerscorporation.com.

RS No. 333



Product Brochure

This 15-page brochure features the company's test and measurement solutions for wireless wide area, local area and personal area networks. The brochure is divided into sections on the WiMAX, RFID, WLAN, Bluetooth®, DVB-H, GPS and ZigBee standards. Customers thus have at a glance an overview of the company's test and measurement instrument and system solutions for each of these standards.

Rohde & Schwarz GmbH & Co. KG, München, Germany
+49 (89) 41 29-0, www.rohde-schwarz.com.

RS No. 334

Handset Brochure

This brochure features RF front-end components for GSM/EDGE, GPRS, WCDMA (UMTS) and CDMA handsets. The complete RF product line includes PAs, PHEMT switches, receivers, duplexers, triplexers, SAW filters and modules. The company delivers highly integrated RF handset and network solutions for quicker design turns, higher performance, lower part count and lower solution cost.

TriQuint Semiconductor,
Hillsboro, OR (503) 615-9000, www.triquint.com.

RS No. 335

Product Brochure

This product brochure features the company's three product families for RF, microwave and millimeter-wave test solutions in active/passive device modeling, parameter extraction, noise and power characterization test systems, DC and RF pulsed IV test systems, load-pull and noise test systems, on-wafer test, solid-state active and passive mechanical tuners, and synthetic instruments 10 MHz to 50 GHz ATE manufacturing test systems.

Auriga Measurement Systems,
Lowell, MA (978) 441-1117, www.auriga-ms.com.

RS No. 336

ers. The use of the standard PCMCIA connector also moves the technology into the realm of network devices such as WiFi access points. One could envision a mesh network of RFID readers embedded in WiFi devices to transfer the data to the network. The removal of cabling would greatly simplify the deployment of an RFID system.

Further reduction in the size and cost of RFID readers can only be accomplished through integration. With the use of highly integrated circuits, one can see how the PCMCIA form factor can be reduced to compact flash (CF) or perhaps secure digital card (SD) format. It is through this evolution that UHF RFID readers will move into low power mobile applications. It is expected that the initial chip sets will consist of three ICs, as shown in ***Figure 6***, following the approach used for early 802.11 chipsets. A digital processor will contain all of the baseband signal processing and control functions. The analog transceiver will perform the translation between the baseband and RF signals, and finally a separate power amplifier will enable flexibility in the design. As technology and design techniques evolve, it may be possible to integrate the entire system onto a complete IC but today the requirements do not lend themselves to a single technology.

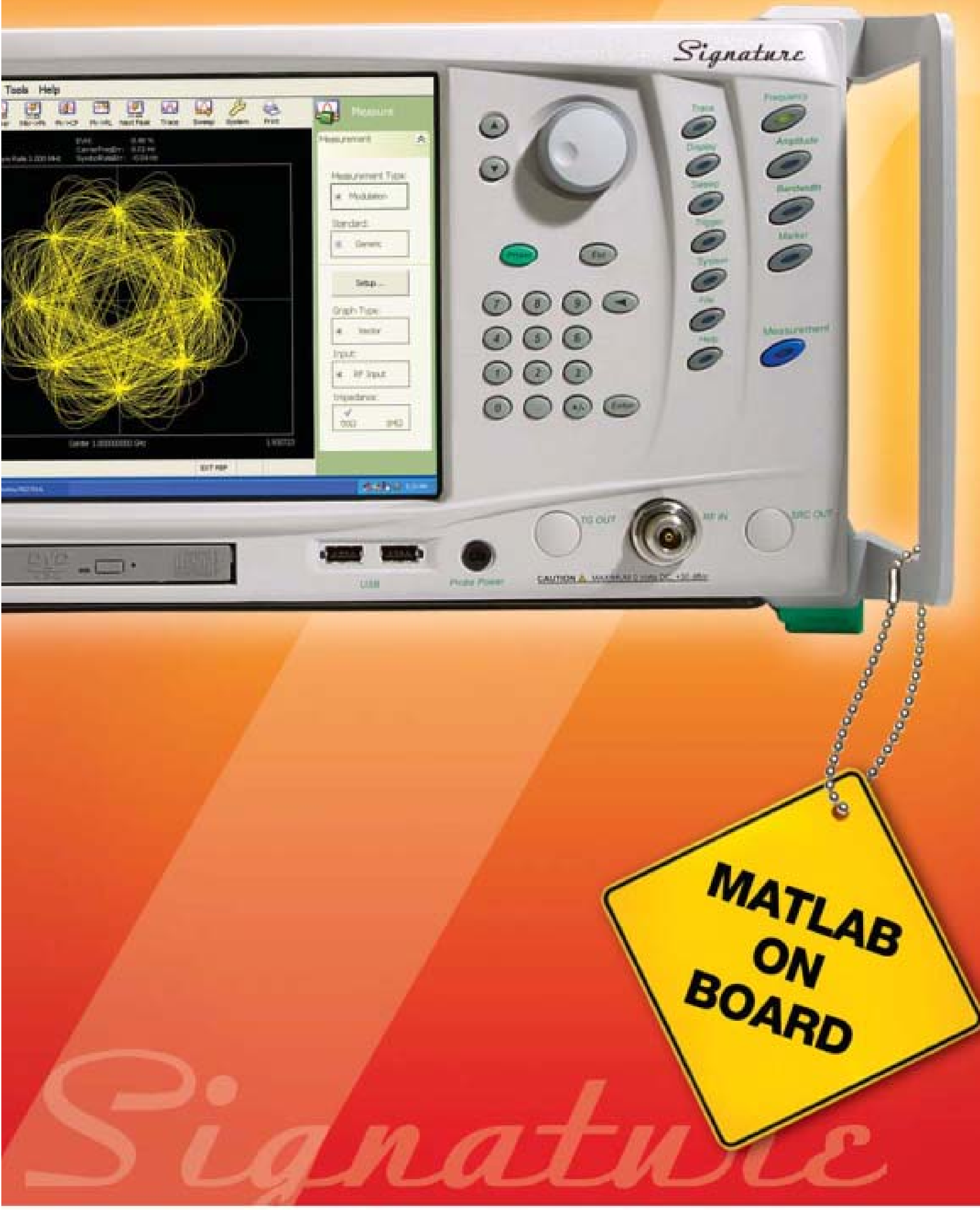
The baseband processor will, of course, be implemented in CMOS, the most cost-effective technology for this type of circuit. The transceiver is ideally suited for SiGe BiCMOS, as it enables the synthesizer to be integrated with the demanding linear functions in the signal path. Finally, the power amplifier will most likely be implemented with InGaP HBTs, which have become the industry standard for powers in the 1 W range. It may be possible to integrate the baseband and transceiver functions all in BiCMOS but the cost, risk and performance tradeoffs must be considered. It may also be possible to implement the power amplifier in SiGe, but it will depend on the requirements and capabilities of the selected semiconductor process. An example of transceiver chip architecture is shown in ***Figure 7***. The transceiver includes pulse shaping, frequency synthesis, modulation, automatic level control, up/down conversion and some RF power amplification. This chip, and others currently in development, represents a major step forward toward the fully integrated RFID reader.

UHF passive RFID represents one of the most promising growth prospects in the RF industry in recent years, but there is much research and development to be completed. In this article, semiconductor solutions have been proposed but there are also innovations needed in radio architecture, antenna design, mechanical packaging and signal processing to mention a few, that will drive the RFID industry toward maturity. As the technology evolves toward enabling low power mobile readers, many applications and markets will in turn develop. ■

Tom Cameron *received his PhD degree in electrical engineering from the Georgia Institute of Technology, Atlanta, GA. He joined WJ Communications Inc. in 2004 and is currently the regional sales manager for Europe. He is the co-inventor of seven patents and author of numerous technical papers and articles.*

1-800-Anritsu

Insight Inside



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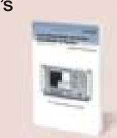

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