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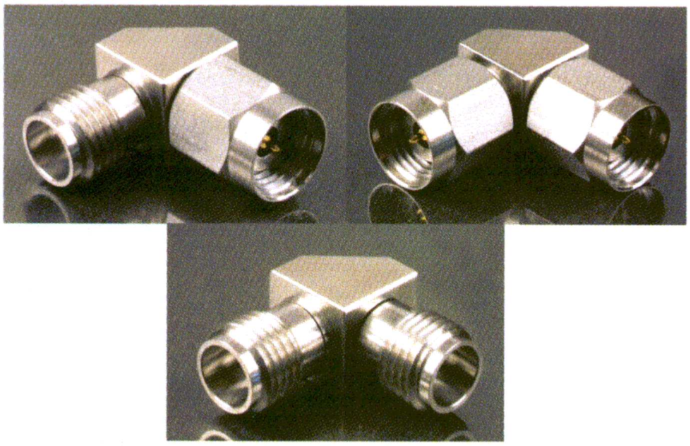

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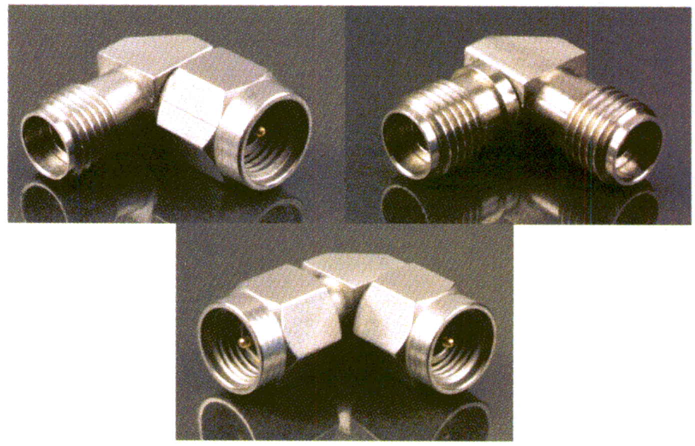

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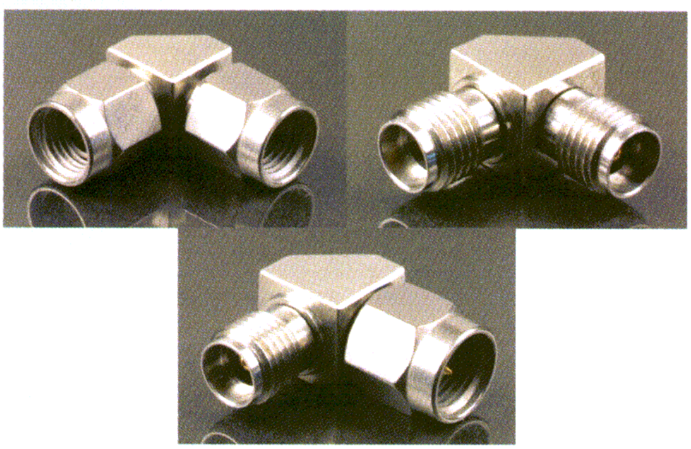

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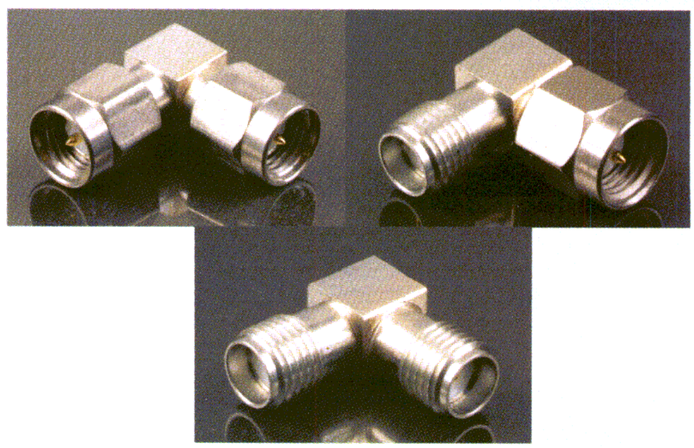

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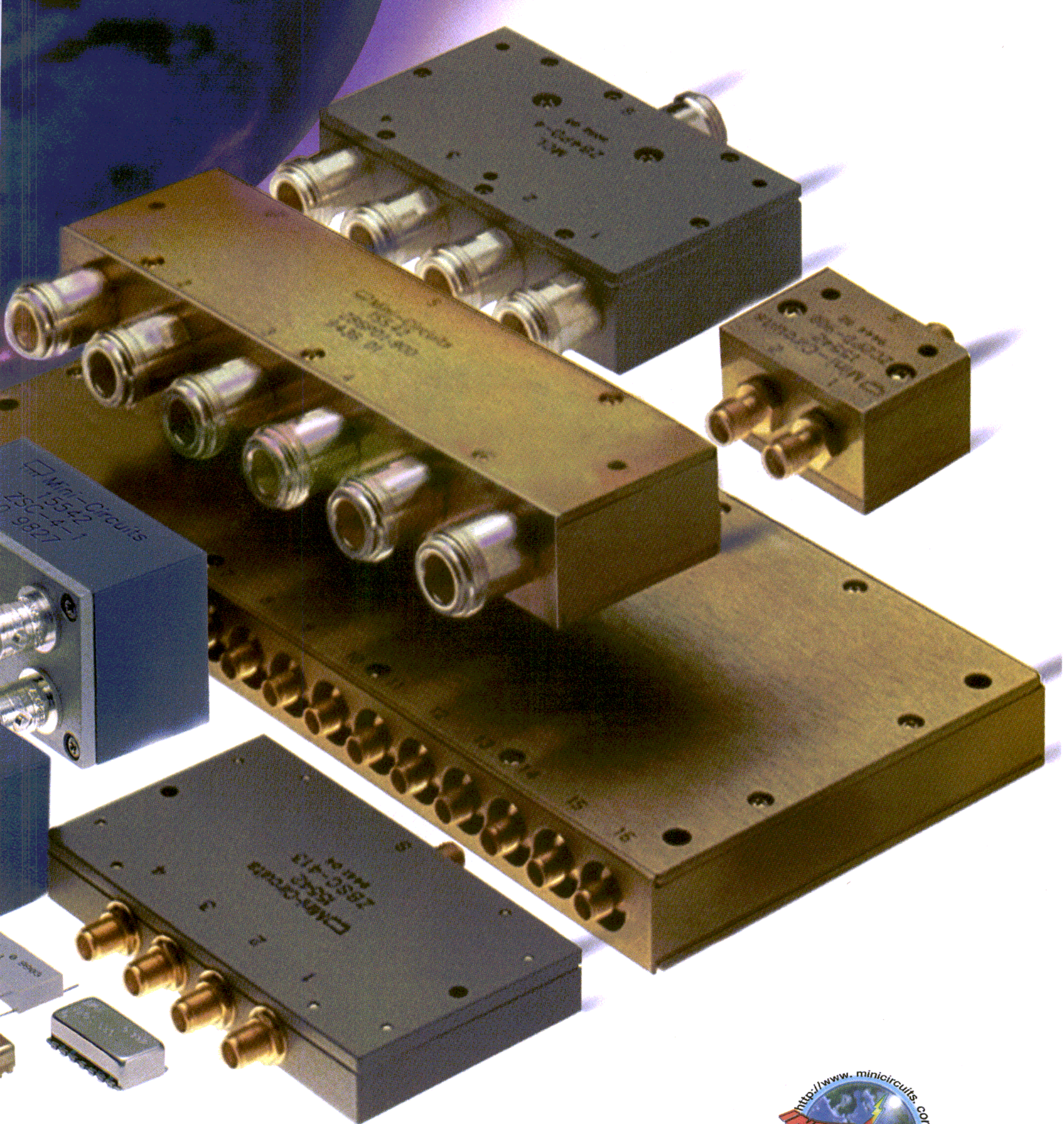
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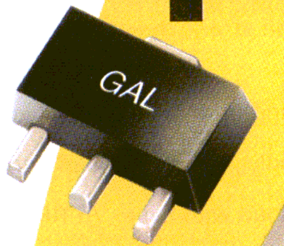

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GAL-3	DC-3000	22.4	19.1	±1.7	12.5	3.5	25	127	35	3.3	.99
• GAL-6F	DC-4000	12.1	11.6	±0.3	15.8	4.5	35.5	93	50	4.8	1.29
• GAL-4F	DC-4000	14.3	13.4	±0.5	15.3	4.0	32	93	50	4.4	1.29
• GAL-51F	DC-4000	18.0	15.9	±1.0	15.9	3.5	32	78	50	4.4	1.29
• GAL-5F	DC-4000	20.4	17.4	±1.5	15.7	3.5	31.5	103	50	4.3	1.29
• GAL-55	DC-4000	21.9	18.5	±1.7	15.0	3.3	28.5	100	50	4.3	1.29
• GAL-52	DC-2000	22.9	17.8	±2.5	15.5	2.7	32	85	50	4.4	1.29
• GAL-S66	DC-3000	22	17.3	±2.4	2.8	2.7	18	136	16	3.5	.99
GAL-6	DC-4000	12.2	11.8	±0.3	18.2	4.5	35.5	93	70	5.0	1.49
GAL-4	DC-4000	14.4	13.5	±0.5	17.5	4.0	34	93	65	4.6	1.49
GAL-51	DC-4000	18.1	16.1	±1.0	18.0	3.5	35	78	65	4.5	1.49
GAL-5	DC-4000	20.6	17.5	±1.6	18.0	3.5	35	103	65	4.4	1.49

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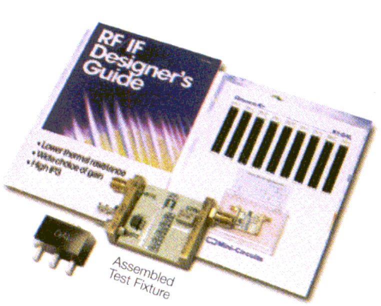

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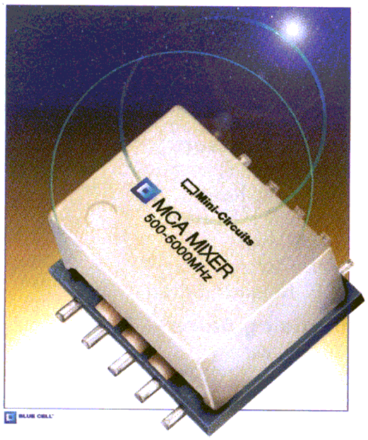

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TECHNICAL FEATURES

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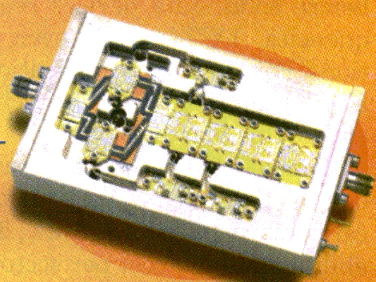
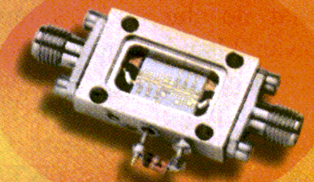
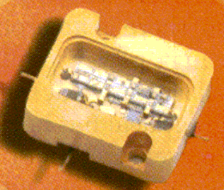
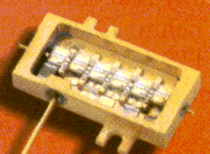
JCA34-P01	3.7-4.2	30	3	1	30	40	2.0:1	750
JCA56-P01	5.9-6.4	30	3	1	30	40	2.0:1	850
JCA78-P01	7.9-8.4	30	4	1	30	40	2.0:1	900
JCA812-P02	8.3-11.7	40	5	1.5	33	40	2.0:1	1700
JCA910-P01	9.5-10.0	30	4	1	33	40	2.0:1	1300
JCA1011-P01	10.7-11.7	30	4	1	30	40	2.0:1	950
JCA1819-P01	18.1-18.6	30	5	1	27	37	2.0:1	800
RADAR & COMMUNICATION BAND LOW NOISE AMPLIFIERS								
JCA23-302	2.2-2.3	30	0.8	0.5	10	20	2.0:1	80
JCA34-301	3.7-4.2	30	1	0.5	10	20	2.0:1	80
JCA56-502	5.4-5.9	50	1	0.5	10	20	2.0:1	160
JCA78-305	7.25-7.75	27	1.2	0.5	13	23	2.0:1	100
JCA910-305	9.0-9.5	27	1.4	0.5	13	23	1.5:1	150
JCA1112-305	11.7-12.2	27	1.5	0.5	13	23	1.5:1	150
JCA1415-305	14.0-14.5	26	1.6	0.5	13	23	1.5:1	160
JCA1819-305	18.1-18.6	22	2.0	0.5	10	20	1.5:1	160
JCA2021-600	20.2-21.2	30	2.2	1	13	23	1.5:1	240
TRI-BAND AMPLIFIERS (5.85 TO 14.5)								
JCA514-201	5.85-14.5	8	7	1.5	10	20	2.0:1	100
JCA514-300	5.85-14.5	14	6	1.5	10	20	2.0:1	150
JCA514-302	5.85-14.5	22	6	1.5	20	30	2.0:1	350
JCA514-400	5.85-14.5	25	6	1.5	10	20	2.0:1	250
JCA514-403	5.85-14.5	32	6	1.5	23	33	2.0:1	500
JCA514-501	5.85-14.5	35	6	1.5	16	26	2.0:1	375
JCA514-503	5.85-14.5	41	6	1.5	23	33	2.0:1	500
ULTRA-BROAD BAND AMPLIFIERS (2.0 TO 18 GHZ)								
JCA218-200	2.0-18.0	15	5	2.5	10	20	2.0:1	90
JCA218-300	2.0-18.0	23	5	2.5	10	20	2.0:1	110
JCA218-400	2.0-18.0	29	5	2.5	10	20	2.0:1	150
JCA218-500	2.0-18.0	39	5	2.5	10	20	2.0:1	180









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Letters

Editor:

I was somewhat disturbed by the comments on traveling wave tube amplifiers (TWTAs) in the article "Design and Test Considerations for Multicarrier LMDS Radios" (August 2001). Although the article made many excellent points, its representation of the trades between TWTAs and solid state power amplifiers (SSPAs) was inaccurate.

The choice between TWTAs and SSPAs depends on a number of factors. Generally, the higher the operating frequency and output power required from an amplifier, the more likely the best technological choice will be a TWTA or, its narrow band cousin, the klystron. The real question arises at lower output power levels. In the days before solid state devices reached the microwave frequencies, TWTAs were even used for receive applications. For the 1 watt application discussed in the article, a SSPA is probably the best choice. But at 10 watts and above, this may not true.

The article states that "traveling wave tube technology requires high to very high power, a large number of carriers per transmitter (i.e., 20 to 40 carriers), low reliability and high cost per sector. This method is also difficult to engineer for outdoor use and results in large prime power and challenging back-up requirements." What do the authors mean by "high to very high power"? I am particularly sensitive to this issue as there have been several papers on Ka-band SSPAs in the 16 to 30 watt power range in IEEE MTT publications that make similar statements. Contrary to these statements and common beliefs, TWTAs remain the most efficient linear power amplifiers. The reliability of TWTAs is comparable, if not superior, to SSPAs, as found by a recent NASA study.

Also, there is no problem operating TWTAs outdoors. Many high power amplifier vendors offer TWTA amplifiers specifically packaged for outdoor use.

A great deal has been written about the relative merits of power amplifiers. Proponents of SSPAs have tried to associate TWTAs with grided vacuum tube amplifiers (GVTAs). However, TWTAs operate on a completely different physical principle than that governing GVTAs. The only real similarity is that both make use of an electron beam. One of the most important differences is that TWTAs, as SSPAs, are still a developing technology. Improve-ments in TWTA parameters, such as efficiency, size and power level, are announced regularly. Because of the changes in the state of the art, it is important that comparisons are made between amplifiers of the same time frame. I am afraid some writers have compared the most recent SSPAs with TWTAs as they were more than a decade ago.

One of the biggest changes in TWTAs is the incorporation of linearizers in many modern TWTAs. Without linearization TWTAs must be operated at about half the power level of a comparably sized SSPA to achieve the same linearity. With linearization, TWTA linearity can be equal or superior to a similar power unlinearized SSPA. At higher output power levels TWTAs are less costly than SSPAs, even when TWTAs are linearized. The output level where this cost advantage begins is very much a function of frequency. At C-band, I found that a 300-watt SSPA (360 watts saturated) costs about 40 percent more than a linearized 400-watt TWTA. The SSPA is about four times the size of a comparable TWTA and requires more than twice the electrical power.

The cost advantage of a TWTA does decrease at lower power levels, but at 200 watts, the cost of an SSPA is still about 25 percent greater than a linearized TWTA. At Ka-band, these differences (cost, size, efficiency) are much greater. Even at 30 watts and in some highly linear applications, 5 or 10 watts, a Ka-band linearized TWTA will have the advantage.

Today, advanced TWTAs provide the highest efficiency and most compact microwave power amplifier technology available. In the future, even greater efficiency and smaller size is projected.

Dr. Allen Katz, President
Linearizer Technology, Inc.

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Editor:

I wish to thank the authors of the article "Design and Test conditions for Multicarrier LMDS Radios" for relating intermodulation products to LMDS link performance in a direct and informative fashion. At the same time, I'd like to clear up what appear to be misstatements regarding the relative performance characteristics of solid state and thermionic amplifiers.

The authors indicate that traveling wave tube technology is characterized by low reliability, high cost, difficulty in engineering for outdoor use and high power consumption. Those statements are inconsistent with a

growing body of empirical evidence. For the power levels required for LMDS architectures, TWTAs have been found to be two to four times more efficient than SSPA equivalents. A representative state-of-the-art 10-watt Ka-band SSPA from U.S. Monolithics requires a 130-watt power supply to produce 10 watts of utput at 1 dB compression. Intelsat, in one of the largest studies performed on the relative MTBFs of SSPAs and TWT equivalents, found the failure rate of the SSPA population to be about 15 percent higher than TWTs at frequencies as low as C-band. When packaged for outdoor use and integrated with appropriate power supply and supporting circuitry, SSPAs are on the order of $10,000 each in small quantities and about $4,000 each in 1,000-piece quantities. At that rate, LMDS SSPAs will reach the selling price of microwave ovens equipped with magnetrons at quantities far greater than the LMDS market will ever reach.

Perhaps the best solution for LMDS applications is a synergistic blend of solid state and thermionic technologies. A 100-watt transmit module operating from 6 to 18 GHz with an overall efficiency of 40 percent at band center has been demonstrated. The 50 dB gain of the module was equally balanced between the MMIC driver and a TWT output stage. The module represented a four-fold efficiency advantage over an SSPA alone, a 20 dB improvement in noise figure over a TWT alone and a ten-fold reduction in size relative to either SSPA or TWT.

Solid state devices are clear winners at lower frequencies and power levels, but as the number of junctions increases and frequency goes up, efficiency goes down and temperature increases exponentially. The crossover region for MMICs versus TWTs is on the order of tens of watts at tens of GHz, and it appears that it will stay there for quite some time. The optimum solution for an LMDS installation is a function of information capacity, distance, life cycle, cost, rain fade margins and reliability. There is no panacea amplifier.

Ken Schoniger
Consultant

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2. "Scanning the Technology Vacuum Electronics at the Dawn of theTwenty-First Century," *Proceedings of the IEEE*, Vol. 87, No. 5, May 1999.

Editors:

The article "Microwave Multiplexer Design Based on Triplexer Filters" (August 2001) fails to reference the original paper describing an exact theory for contiguous multiplexers having any number of channels. It consists of singly-terminated bandpass filters connected in series or parallel, plus an additional bandstop filter that matches the entire structure at all frequencies. The authors' paper describes the same arrangement for the case of a diplexer plus a bandstop filter. The original paper also describes how such multiplexers may be cascaded.

Although these authors have used the technique to design a 10-channel multiplexer with 6 MHz wide channels covering the 30 to 90 MHz band, they would not recommend widespread use, which would abandon the commonly used computer-aided design techniques based on combining channels without the complementary bandstop filter. Generally, it is more efficient to avoid the bandstop filter, and in particular there is no need to use it for the simple multiplexer described by the authors. The exact theory may be regarded as of interest in giving an analytic solution to a long-standing problem, i.e., it is of academic interest. It also gives a starting point for possible development of better techniques for synthesizing manifold multiplexers, with the bandstop filter replacing the usual end-short. In addition, the exact technique may find use where cascaded multiplexers provide a convenient solution to difficult problems, e.g., where two sets of contiguous multiplexers are separated by a guard band.

Ralph Levy
R. Levy Associates

References

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Information: Process Sciences Inc., Tel: 512-259-7071; Fax: 512-259-7073; Internet: www.process-sciences.com.

Besser Associates

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Georgia Institute of Technology

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El Segundo, CA . . . . .November 12, 2001
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El Segundo, CA . . . . .November 13-14, 2001
Information: Kimmel Gerke Associates, 1-888-EMI-GURU (364-4878), Internet: www.emiguru.com.

International Institute of Connector and Interconnection Technology (IICIT)

Basic Connector Technology
Orlando, FL . . . . .November 7-8, 2001
Basics of Fiber Optic Interconnection
Orlando, FL . . . . .November 9, 2001
Communications and Standards
Orlando, FL . . . . .November 9, 2001
Information: Suzanne Romeo, Tel: 1-800-854-4248; E-mail: sromeo@iicit.org; Internet: www.iicit.org.

Calendar

University of California at Los Angeles Extension
Optical Phase-Locked Loops
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Automatic Test Equipment Selection, Design and Programming
Los Angeles, CA . . . . .December 3-4, 2001
Design for Testability Built-In Self-Test
Los Angeles, CA . . . . .December 5-7, 2001
Information: UCLA Extension, Short Course Program Office, Tel: 310-825-3344; Fax: 310-206-2815.

TTi Technology Training Initiative (Tustin Technical Institute, Inc.)
Introduction to Mechanical and Structural Theory
Santa Barbara, CA . . . . . .November 8-9, 2001
Fixture Design for Vibration and Shock Testing
Santa Barbara, CA . . . . . .November 12-14, 2001
Environmental Testing Procedures
Santa Barbara, CA . . . . . .November 15-16, 2001
Environmental Test Specifications
Santa Barbara, CA . . . . . .November 19-20, 2001
Fundamentals of Vibration for Test Applications
Santa Barbara, CA . . . . . .November 26-28, 2001
Huntsville, AL . . . . .December 3-5, 2001
Mechanical Shock Techniques
Santa Barbara, CA . . . . . .November 29-30, 2001
Information: Brian P. Slatery, Tel: 805-682-7171; Fax: 805-687-6949; E-mail: brian@ttiedu.com; Internet: www.ttiedu.com.

Agilent Technologies
RF & Microwave Fundamentals
Winnersh, UK . . . . .December 4-6, 2001
Information: Tracey Bull, Tel: +44 118 9276741; Fax: +44 118 9276862; E-mail: tracey_bull@agilent.com.

University of Wisconsin at Madison
Engineering and Planning Telecommunication Local Loop Facilities
Madison, WI . . . . .December 4-7, 2001
Information: Katie Peterson, Tel: 1-800-462-0876; Fax: 608-263-3160; E-mail: custserv@epd.engr.wisc.edu; Internet: http://epd.engr.wis.edu.

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Deadline: December 1, 2001

Win With V8's

BRIEFS

• The name of the IMAPS Educational Foundation has been changed to the Dr. Sidney J. Stein Educational Foundation, to honor Stein's role as a founding member of the foundation and his years of service to the industry. The foundation helps universities recruit and support engineers and scientists in microelectronics, engineering and materials science technologies.

• Bechtel Telecommunications has opened a Training, Demonstration and Research (TDR) Laboratory at the company's headquarters in Frederick, MD. The TDR Lab is designed to showcase the company's simulation and evaluation capabilities for both wireless and wireline network design, including such protocols as VoIP, PSTN, LMDS, MMDS, 2G/3G wireless and IEEE 802.11 WAN.

• di/dt has announced the completion of a new manufacturing and headquarters facility in Carlsbad, CA. The 65,000-square-foot facility houses complete assembly and administrative activities for the DC/DC power conversion company.

• Invisix™, a mobile wireless systems integrator jointly owned by Motorola and Cisco Systems, has announced plans to open a new Competency Centre in Jaguariuna, Brazil. The location will focus on the development of both General Packet Radio Service (GPRS) and Code Division Multiple Access (CDMA) 2000 1X solutions.

• Qusion Technologies, a manufacturer of monolithically integrated optical components on indium phosphide, has opened a new fabrication facility at its headquarters in North Brunswick, NJ. The 8,000-square-foot facility includes a 3,000-square-foot class 1000 clean room and a 400-square-foot class 100 photolithographic clean room.

Submit information for our News section to Shannon O'Connor via E-mail at: amw@amwireless.com.

Research center to develop ICs

Atmel Corporation, Multilink Technology Corporation and United Monolithic Semiconductors (UMS) have announced a partnership with the University of Ulm in Ulm, Germany, to create a joint research and development group.

The Competence Center on Integrated Circuits in Communications will conduct research and development activities on integrated circuits for analog functions in wireless and high-speed optoelectronic systems, as well as semiconductor device modeling and device and circuit characterization. The research projects will make use of Atmel's silicon and silicon geranium processing facilities, as well as UMS' gallium arsenide processing.

Examples of planned projects include high linearity amplifiers for Code Division Multiple Access (CDMA) systems and front-end ICs for high-end, long-haul fiber optic communication systems operating at 40 GHz and beyond.

The three companies are supplying funding for the center, with the University of Ulm providing offices, laboratories and other infrastructure support.

Additional startup funding will come through the State of Baden-Wurttemberg's Ministry of Science, Research and Art. Also contributing through an ongoing grant program will be the Stifterverband fur die deutsche Wissenschaft, an industry-led scientific funding organization.

New thermoelectronic material increases cooling capabilities

Research Triangle Institute (RTI) has developed a new thermoelectronic material that is 2.4 times more efficient and responds 23,000 times faster than existing materials. The new material was introduced in the October 11, 2001, issue of *Nature* magazine.

A thermoelectronic module using one square centimeter of RTI's new material could provide 700 watts of cooling under a temperature gradient of 58 degrees F. Tiny dots of the material, applied specifically to hot spots on a microprocessor chip, would give better performance than cooling the whole chip and would consume less power.

The technology emerged from a U.S. Department of Defense initiative targeting thermoelectronic technology. Funding for RTI's research has been provided since 1993 through the Office of Naval Research and the Defence Advanced Research Projects Agency.

Applications for the advanced thermoelectronic material could include lower voltage electroholographic devices, proteomic chips for biotechnology and electrical conversion in hybrid automobiles.

RTI is an independent research organization based in Research Triangle Park, NC.

Procter & Gamble donation supports university research

Procter & Gamble has announced the donation of its proprietary Smart Power Management (SPM) technology to the University of Illinois at Chicago (UIC). The donation includes all patents and accompanying intellectual property.

The SPM development program will be administered through UIC's College of Engineering. The school plans to start up a faculty-led business to complete commercial development and deployment. According to Procter & Gamble, the company regularly donates technology such as the SPM program because it cannot fully develop all of the more than 27,000 patents it holds.

SPM technology consists of micro integrated circuits that deliver efficient power, enabling the development of smaller, lighter electronic devices as well as increasing the usable life of batteries.

News

BUSINESS AND FINANCE

RF Micro Devices to acquire RF Nitro Communications

RF Micro Devices has announced an agreement to acquire RF Nitro Communications of Charlotte, NC. Financial terms were not disclosed.

RF Nitro's operations include a four-inch wafer fabrication facility, scheduled for completion this year, that will be used both for production and as a technology incubator. The company also has design, assembly and test facilities for three compound semiconductor IC technologies, InGaP HBT, GaAs PHEMT and GaN HEMT, on both silicon carbide and sapphire substrates. The company is expected to operate under the name RF Micro Devices Charlotte.

RF Micro Devices, based in Greensboro, NC, manufactures proprietary radio frequency integrated circuits for wireless communications products and applications.

Motorola wins network contracts

Motorola has announced several contracts for network expansion services in China, India, Ukraine and the United States.

- Under two contracts with China Mobile Communications Corpora-tion, Motorola will expand the existing GPRS network in seven provinces, as well as the 800 MHz CDMA network in Hebei Province. The CDMA contract is valued at $34.8 million; the value of the GPRS network was not disclosed.
- A $178 million contract with Bharat Sanchar Nigam Ltd. calls for the deployment of a GSM communications infrastructure in each of the four states in southern India. A separate $70 million contract with Bharti Enterprises will result in a GSM network, including GPRS services, in western India.
- CST Invest Ltd. has awarded Motorola a contract to deploy a 3G CDMA 2000 1X network in Dnepropetrovsk, Ukraine. Financial terms were not disclosed.
- An agreement with Horizon PCS calls for the upgrade and expansion of the company's existing CDMA system in Pennsylvania, Ohio, Tennessee, Indiana and portions of New York and New Jersey. The value of the contract is approximately $30 million.

Motorola, based in Schaumburg, IL, provides semiconductors, integrated communications solutions, embedded electronic systems, components and network supply and infrastructure services.

C-MAC to acquire circuit switching lines from Nortel

Nortel Networks has reached an agreement for C-MAC Industries to purchase most of Nortel's digital multiplex system (DMS) manufacturing activities. In return, C-MAC will supply Nortel with systems integration, configuration and testing services for DMS products.

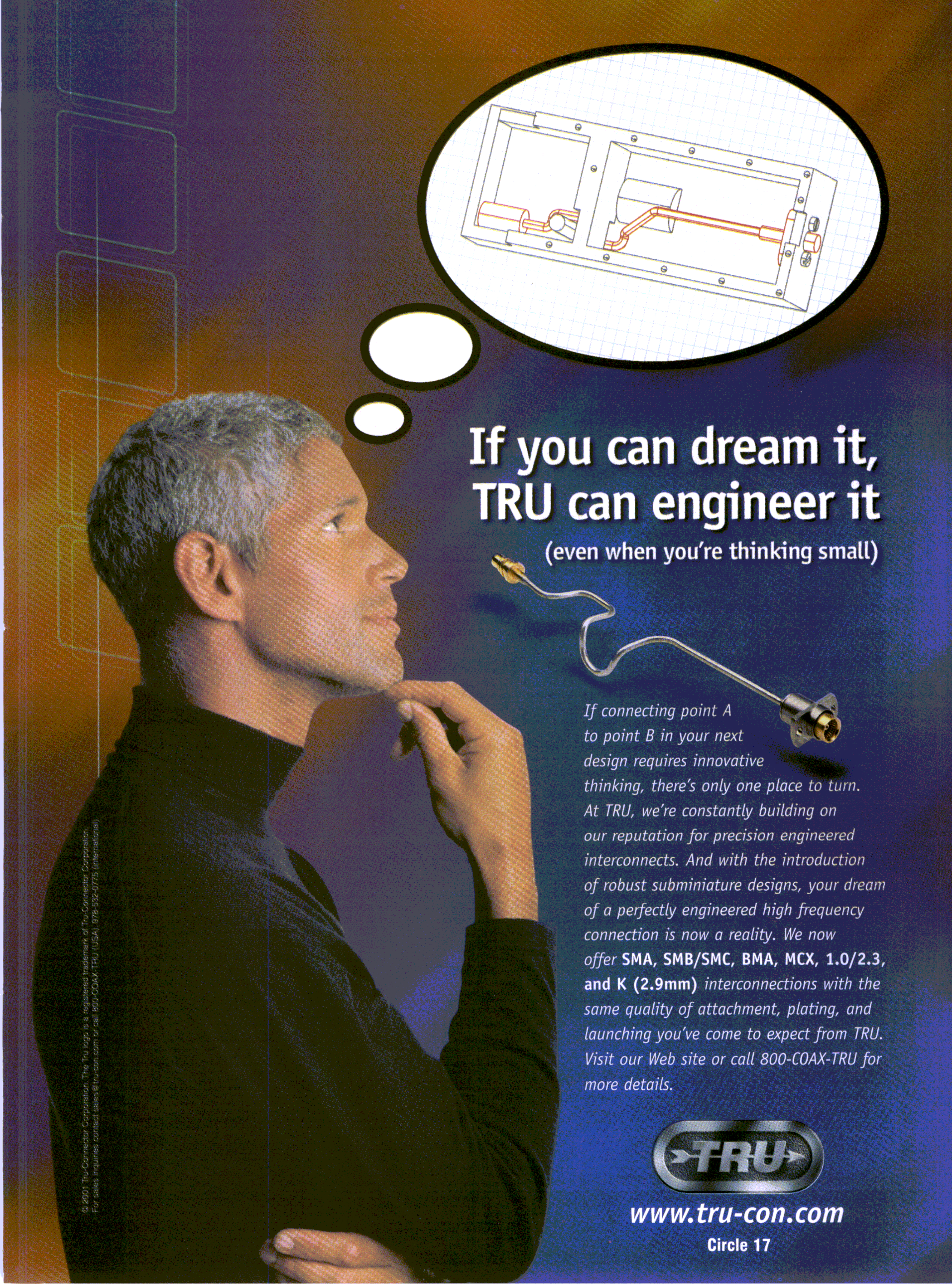
Under the agreement, C-MAC will acquire most of the DMS-related manufacturing activities performed at Nortel's Systems House in Research Triangle Park, NC, as well as similar activities from the Systems House in Monkstown, Northern Ireland.

C-MAC, based in Montreal, Quebec, Canada, manufactures integrated electronic solutions from components to full systems for communications, automative, instrumentation, defense and aerospace equipment markets. Nortel, based in Toronto, Ontario, Canada, provides networking and communications solutions and infrastructure.

Celestica acquires Omni

Celestica has announced the acqusition of electronics manufacturing services provider Omni Industries Ltd. of Singapore in a stock and cash transaction.

Celestica, based in Toronto, Ontario, Canada, provides electronics manufacturing, design, testing and distribution services.



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BUSINESS AND FINANCE

Harris receives Air Force contract

Harris Corporation has received a one-year, $24.3 million contract from the U.S. Air Force to provide operations, maintenance and support services for the Air Force Satellite Control Network (AFSCN) and Global Positioning System (GPS) ground network.

The contract, awarded through the Operational Space Services and Support (OSSS) program, is part of $50 million in agreements awarded to Harris. Options for five additional years could increase the total value to $202 million.

Under the contract, Harris Technical Services Corporation (HTSC) will provide antenna operation, GPS site software and hardware support, satellite orbital analysis, technical assistance and other services at several locations.

HTSC, based in Alexandria, VA, is a wholly owned subsidiary of Harris Corporation that provides engineering and technical support solutions for defense and government markets.

IlliCom using WaveRider system for network expansion project

WaveRider Communications has announced that IlliCom Technologies has purchased WaveRider's Last Mile Solution® system to expand its wireless network in east central Illinois. IlliCom will introduce broadband internet access using WaveRider's LMS3100 non-line-of-sight wireless products.

The LMS3100 system is designed to deliver broadband internet access at speeds of 1.4 Mbps using the 900 MHz spectrum. The system includes a wireless modem and indoor antenna that can be easily installed by subscribers.

WaveRider, based in Toronto, Ontario, Canada, provides data communications and wireless internet networking products.

Vyyo receives 3.5 GHz order

Vyyo has received an order from China Communications Systems Company Ltd. (China Comm) for the deployment of a 3.5 GHz broadband wireless network in China. Financial terms were not disclosed.

Under the agreement, China Comm will install Vyyo's system in the cities of Wuhan, Nanjing, Qingdao, Xiamen and Chonqing.

Vyyo, based in Cupertino, CA, supplies point-to-multipoint broadband wireless access systems for multiple frequencies.

Wideband Transformers: An Intuitive Approach to Models, Characterization and Design

By Chris Trask
Sonoran Radio Research

Wideband transformers constructed with high permeability ferrite and powdered iron magnetic materials are used extensively for impedance matching, power combining and power splitting, as well as other functions, for frequencies ranging from VLF to VHF. Despite this broad range of uses, a concise step-by-step procedure for characterizing wideband transformers and effecting a design making full use of the parasitic reactances is lacking.

This article presents such a procedure as well as discusses the development of the equivalent circuit of the wideband transformer. The new procedure aims to provide the designer with a means for obtaining an intuitive understanding of the limitations and compromises.

Ideal transformer model

In an ideal transformer, as shown in Figure 1, all the magnetic flux produced in one winding links all the turns of the second winding. Thus, the voltage and current ratios are [1]:

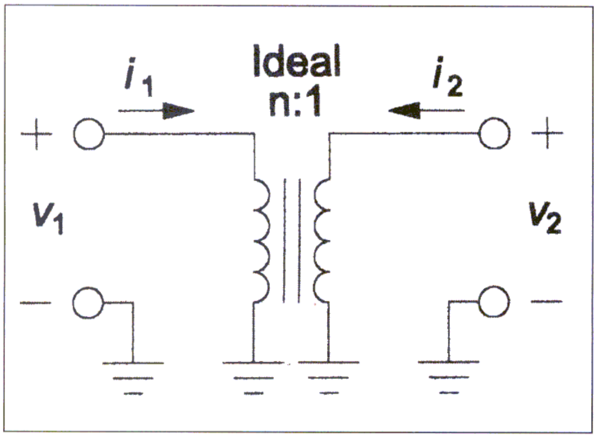
$$\frac{v_2}{v_1} = \frac{N_S}{N_P} = \frac{1}{n} \tag{1}$$

$$\frac{i_2}{i_1} = -n \tag{2}$$

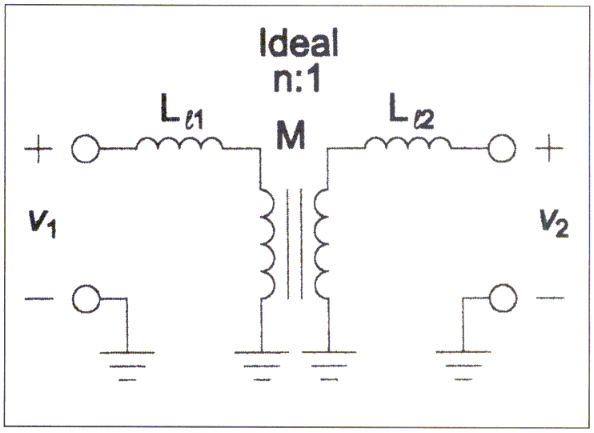
where N_P and N_S are the number of turns for the primary side and secondary side, respectively. The ideal transformer does not exist in the real world, but is instead affected by a number of parasitic elements.

Lossless transformer model

Actual transformers differ from the ideal transformer in several ways. A schematic of the equivalent model of a lossless transformer model is shown in Figure 2. The low-frequency performance is determined by the permeability of the core material and the number of turns on the windings [2]. The high-frequency performance is limited by the fact that not all of the



▲ **Figure 1. Ideal transformer model.**



▲ **Figure 2. Lossless transformer model.**

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	5	1.07 fc	1.26 fc	1.62 fc
	6	1.05 fc	1.18 fc	1.44 fc
	7	1.04 fc	1.14 fc	1.33 fc
	8	1.04 fc	1.11 fc	1.26 fc
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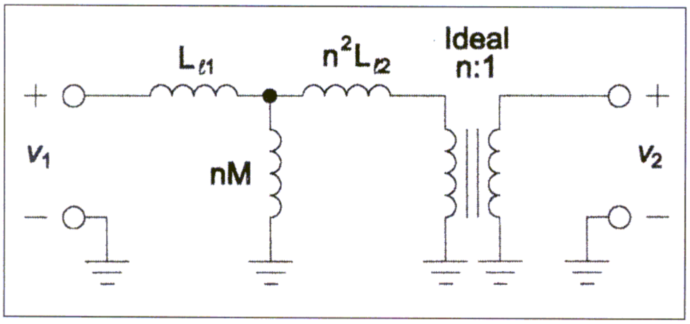
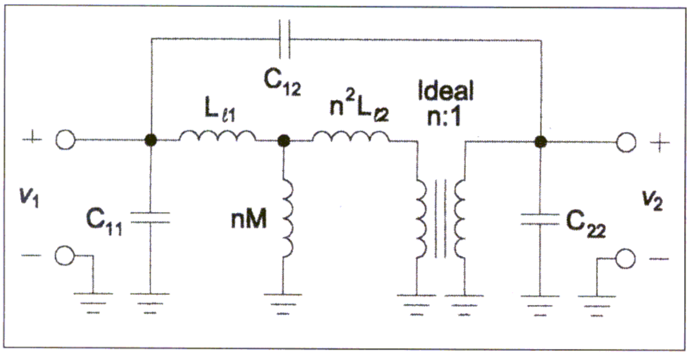


▲ **Figure 3. Lossless transformer model referred to primary side.**



▲ **Figure 4. Wideband transformer model.**

flux produced in one winding links to the second winding, a deficiency known as *leakage* [1].

Since the leakage flux paths are primarily in air, the resultant primary and secondary leakage inductances L_{l1} and L_{l2} are practically constant [3, 4]. A mutual inductance (M) between the two windings is a result of the linked flux in the transformer core. The nonlinearity of the magnetic characteristics of the core material is generally assumed to affect only the relation between the mutual flux and the exciting current. The mutual inductance M is determined by way of:

$$M = \sqrt{L_P L_S k} \tag{3}$$

where L_P and L_S are the primary and secondary winding inductances, which are measured at low frequency with an inductance bridge, leaving the opposite winding open. The coupling coefficient k is then determined by:

$$k = \frac{v_2}{v_1}\sqrt{\frac{L_P}{L_S}} \leq 1 \tag{4}$$

where v_1 and v_2 are easily measured across the primary and secondary terminals with a high impedance probe and a network analyzer, an RF sampling voltmeter, or an oscilloscope with the secondary unloaded and at a frequency that is much lower than the first resonant frequency of the transformer [5]. The value of k is always less than one, because perfect coupling between windings is unachievable. The coupling between windings can be improved by such techniques as twisting the wires together, which increases the interwinding capacitance, or the use of coaxial transmission line where the coupling fields are completely contained within the dielectric.

With the mutual inductance known, the leakage inductances L_{l1} and L_{l2} can be determined by way of [4]:

$$L_{l1} = L_P - nM \tag{5}$$

$$L_{l2} = L_S - \frac{M}{n} \tag{6}$$

It is often convenient to describe transformer models from the viewpoint of the primary winding. The model as such for the lossless transformer is described in Figure 3.

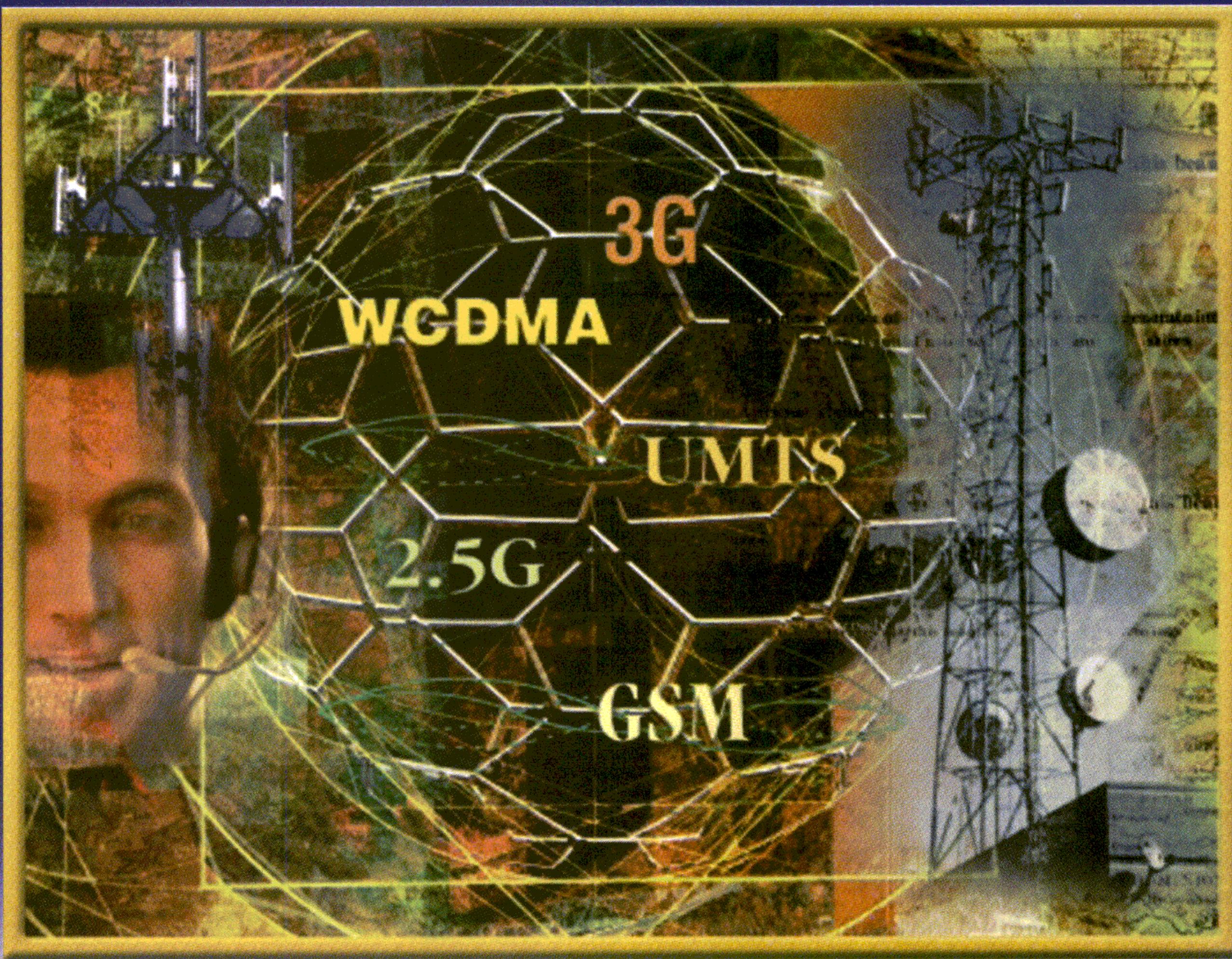
Lossless wideband transformer model

The transformer model that has been described so far is adequate for describing low-frequency transformers such as those used at audio frequencies where any parasitic resonances are far outside the frequency range of interest. At higher frequencies, however, these resonances, which are the result of stray capacitances, come into play. The three principle capacitances that define the wideband transformer are shown in Figure 4.

It is generally understood that the capacitances associated with wideband transformers are distributed, but it is inconvenient to model transformers by way of distributed capacitances per se; thus a single lumped capacitance is used.

In Figure 4, capacitor C_{11} represents the distributed primary capacitance, which is a result of the shunt capacitance of the primary winding. Likewise, C_{22} represents the shunt capacitance of the secondary winding. Some models depict these shunt capacitances as a single capacitor in parallel with the mutual inductance nM [6], but such models do not fully describe the resonances experienced in high-frequency transformers. Capacitor C_{12} is referred to as the interwinding capacitance [7], and is also a distributed capacitance. In conjuction with the distributed inductance of the windings, capacitor C_{12} can form a transmission line whose characteristic impedance can be controlled by way of the wire size, insulation thickness, and degree of twisting [8].

As with the earlier model of the lossless transformer, it is convenient to describe the lossless wideband transformer model from the viewpoint of the primary winding, as shown in Figure 5. Here, the three model capacitances are transformed in the following manner [4]:



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SXT-289	1800-2500	+24	+41	15.0	5.0	105

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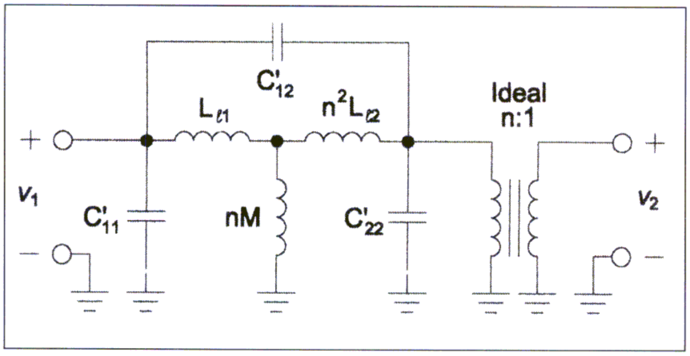
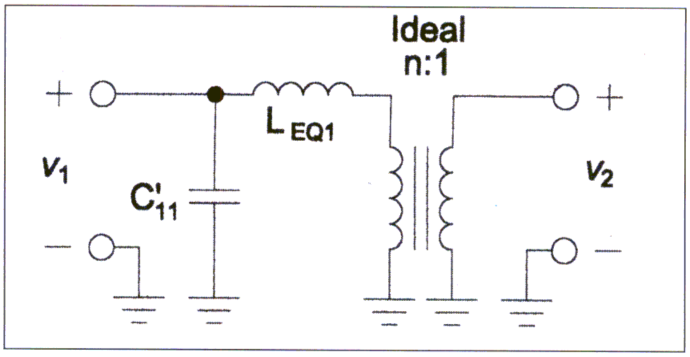


▲ **Figure 5. Wideband transformer model referred to primary side.**



▲ **Figure 6. Equivalent circuit for determining C'_{12} and C'_{22}.**

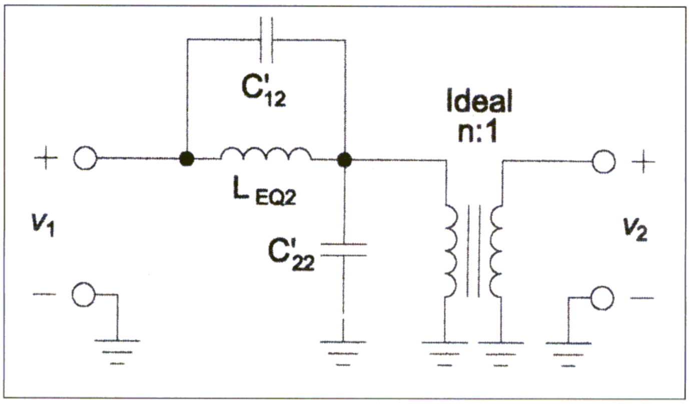
$$C'_{11} = C_{11} + C_{12}\left(1 - \frac{1}{n}\right) \tag{7}$$

$$C'_{12} = \frac{C_{12}}{n} \tag{8}$$

$$C'_{22} = \frac{C_{22}}{n^2} + C_{12}\left(\frac{1}{n} - 1\right) \tag{9}$$

In some circumstances, the capacitance C'_{11} may be hypothetically negative [4]. Furthermore, the capacitance C'_{11} is frequently ignored, since its effects are generally unimportant due to its position in the circuit [4]. However, at HF and higher frequencies it should be evaluated to determine its contribution to the high-frequency cutoff point.

Using the equivalent circuit shown in Figure 6, the interwinding capacitance C'_{12} can be determined by first measuring the parallel resonant frequency f_{12} of the unloaded transformer. Begin by connecting an appropriate signal generator to the primary winding of the transformer. Using a high impedance probe with a network analyzer, an RF sampling voltmeter, or an oscilloscope, observe the voltage at the secondary winding while adjusting the frequency of the generator. The resonant frequency f_{12} is the point where the voltage across the secondary winding reaches a null, or minimum.



▲ **Figure 7. Equivalent circuit for determining C'_{11}.**

$$C'_{12} = \frac{1}{L_{EQ2}(2\pi f_{12})^2} \tag{10}$$

where the equivalent transformer inductance L_{EQ2} is determined by:

$$L_{EQ2} = \frac{nML_{11}}{nM + L_{11}} + n^2 L_{12} \tag{11}$$

The output shunt capacitance C'_{22} can now be determined by measuring the series resonant frequency f_{22} at the input terminals of the transformer. With the generator still connected to the primary winding, adjust the frequency while observing the voltage across the primary winding. The resonant frequency f_{22} is the point where the voltage across the primary reaches a null, or minimum.

$$C'_{22} = \frac{1 - C'_{12} L_{EQ2}(2\pi f_{22})^2}{L_{EQ2}(2\pi f_{22})^2} \tag{12}$$

In a similar manner, the input shunt capacitance C'_{11} can be determined by measuring the series resonant frequency f_{11} at the output terminals of the transformer. Begin by connecting the signal generator across the seconday winding, leaving the primary winding open. The resonant frequency f_{11} is the point where the voltage across the secondary reaches a null, or minimum.

$$C'_{11} = \frac{1 - C'_{12} L_{EQ1}(2\pi f_{11})^2}{L_{EQ1}(2\pi f_{11})^2} \tag{13}$$

where the equivalent transformer inductance L_{EQ1} is determined by:

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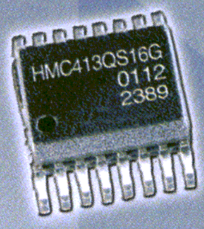
Market	Frequency (GHz)	Output Power P1dB (dBm)	Saturated Power (dBm)	Gain (dB)	PAE	Part Number
UNII & HiperLAN	5.0 - 6.0	26	+29	18	38 %	HMC406MS8G
		25	+29	15	28 %	HMC407MS8G
		29	+32	20	25 %	HMC408LP3
		23	+26	20	35 %	HMC415LP3
Wireless Local Loop	3.0 - 4.0	27	+30	21	45 %	HMC327MS8G
		28.5	+32	25	25 %	HMC409LP3
Cellular	1.5 - 2.3	27	+30	20	45 %	HMC413QS16G
MMDS	2.1 - 3.2	27	+30	20	32 %	HMC414MS8G

HMC327MS8G
- 3.0 - 4.0 GHz
- Output Power: 27 dBm
- Saturated Power: +30 dBm
- Gain: 21 dB

HMC406MS8G
- 5.0 - 6.0 GHz
- Output Power: 26 dBm
- Saturated Power: +29 dBm
- Gain: 18 dB



HMC413QS16G
- 1.5 - 2.3 GHz
- Output Power: 27 dBm
- Saturated Power: +30 dBm
- Gain: 20 dB

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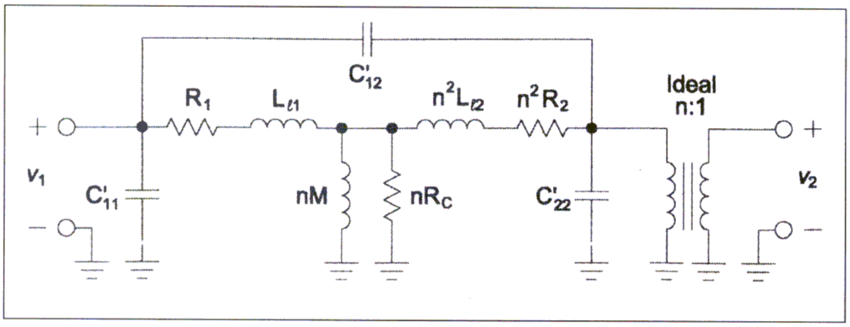


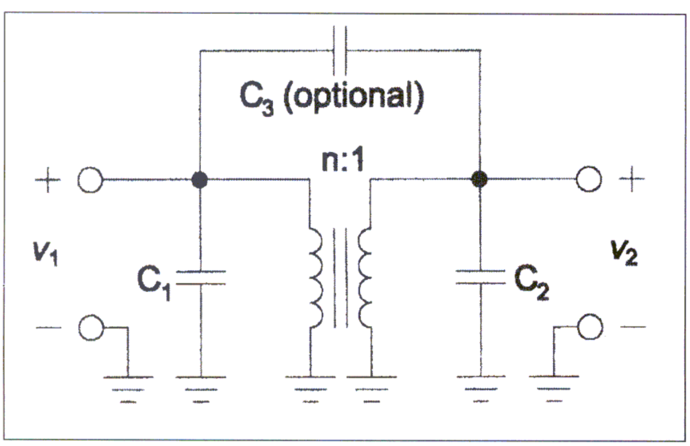
▲ **Figure 8. Complete wideband transformer model.**

$$L_{EQ1} = \frac{n^2 ML_{12}}{M + nL_{12}} + L_{11} \tag{14}$$

Complete equivalent wideband transformer model

No discussion about transformer models is complete without some mention of losses. Two principle loss mechanisms are involved in the wideband transformer: the resistive loss in the copper wires and the hysteresis loss in the ferromagnetic core material. Figure 8 depicts the complete model of the wideband transformer, which has been referred to the primary side earlier. The series resistance R_1 represents the loss associated with the wire in the primary winding. This resistance is nonlinear, increasing with frequency according to $\sqrt{\omega}$ because of the skin effect of the wire itself [5]. Similarly, the series resistance R_2 represents the loss associated with the wire in the secondary winding. Due to the short wire length used in wideband transformers having ferromagnetic cores, the contribution of the resistive loss to the total loss is small and is therefore generally omitted [5].

The shunt resistance R_C represents the hysteresis loss due to the ferromagnetic core [9], which increases with ω^2 or even ω^3, and is significant in transformers that are operated near the ferroresonance of the core material [5]. In general, this is not a problem provided that proper consideration is given to the selection of the core material.



▲ **Figure 9. Matching network components.**

Generally, the factors which affect the insertion losses of the wideband transformer at lower frequencies, attributed to the permeability of the core material and the amount of wire in the windings [2], have a negligible effect in the higher frequency regions [9]. At the same time, the factors associated with losses at higher frequencies, attributed to the skin effect of the wire and the ferroresonance of the core material, have a negligible effect at the lower frequencies [9]. At midband frequencies, losses are more likely to be a result of impedance mismatches.

Wideband transformer as a matching network component

Using a wideband transformer at HF and especially VHF frequencies without compensation is poor design practice. The series inductance alone is sufficient to reduce the high-end performance, and the capacitances do not make things much better. Therefore, we consider the equivalent series inductance of the transformer as being the series inductor of a 3-pole lowpass filter. We then add appropriate capacitors to the input and output and possibly one from the input to the output to complete a wideband matching network with minimal loss.

For the purpose of including the transformer inductance as an element in a wideband matching network, we first recognize that this equivalent inductance determines the maximum usable frequency ω_{max} for the matching network. Therefore, we begin by determining ω_{max} by:

$$\omega_{max} = \frac{L_{norm} R_S}{L_{EQ1}} \tag{15}$$

where R_S is the source impedance of the generator and L_{norm} is the normalized inductance of the lowpass filter section that will be used. A fairly complete listing of the protoype values for a wide range of filter approximations is shown in Table 1 [10, 11]. The added input and output capacitances C_1 and C_2 are determined by:

$$C_1 = \frac{C_{1norm}}{\omega_{max} R_S} - C_{11} \tag{16}$$

$$C_2 = \frac{n^2 C_{2norm}}{\omega_{max} R_S} - C_{22} \tag{17}$$

where C_{1norm} and C_{2norm} are the normalized input and output capacitances of the lowpass filter section, also listed in Table 1. Except for Bessel and Gaussian filter

sections, which are time delay filters [12], the prototype values for C_{1norm} and C_{2norm} are identical.

In some cases, it may be desireable to make an Inverse Tchebychev or elliptical filter section, such as in the case where additional close-in harmonic energy needs additional suppression, keeping in mind that the group delay variations in the vicinity of the cutoff frequency are more severe than for other filter types. In such a situation, an additional capacitance can be included from the input to the output of the transformer:

$$C_3 = \frac{nC_{3norm}}{\omega_{max} R_S} - C_{12} \tag{18}$$

Note that the transformer must be symmetrical, i.e., either noninverting unbalanced to un-balanced or balanced to balanced.

Design procedure summary

- Using an inductance bridge, measure the primary and secondary winding inductances L_P and L_S.
- Using a signal generator, and a high impedance probe with a network analyzer, a sampling RF voltmeter, or an oscilloscope, measure the input and output voltages v_1 and v_2 at an approriate midband frequency.
- Calculate the coupling coefficient k using Equation (4) and the mutual inductance M using Equation (3).
- Calculate the primary and secondary leakage inductances L_{l1} and L_{l2} using Equations (5) and (6), respectively.
- Calculate the equivalent inductances L_{EQ1} and L_{EQ2} using Equations (14) and (11), respectively.
- Continuing with the test setup described in (2), measure the transmission parallel resonant frequency f_{12}, the input series resonant frequency f_{11}, and the output series resonant frequency f_{22}.
- Calculate C'_{12}, C'_{22} and C'_{11} using Equations (10), (12) and (13), respectively.
- Calculate C_{12}, C_{11} and C_{22} using Equations (8), (7) and (9), respectively.
- Select a filter prototype from Table 1.
- Calculate the maximum usable frequency ω_{max} using Equation (15).
- Calculate the values for the input and output matching capacitors C_1 and C_2 using Equation (16) and (17), respectively.
- If required, calculate the value for the capacitor C_3 using Equation (18).

Filter Type	C_{1norm}	C_{2norm}	L_{norm}	C_{3norm}
Butterworth	1.000	1.000	2.000	
Bessel	1.255	0.192	0.553	
Gaussian	2.196	0.967	0.336	
Tchebyschev				
0.1dB	1.032	1.032	1.147	
0.5dB	1.596	1.596	1.097	
1.0dB	2.024	2.024	0.994	
Inverse Tchebyschev				
20dB	1.172	1.172	2.343	0.320
30dB	1.866	1.866	3.733	0.201
40dB	2.838	2.838	5.677	0.132
Elliptical (0.1dB Passband Ripple)				
20dB	0.850	0.850	0.871	0.290
25dB	0.902	0.902	0.951	0.188
30dB	0.941	0.941	1.012	0.125
35dB	0.958	0.958	1.057	0.837
40dB	0.988	0.988	1.081	0.057
Elliptical (0.5dB Passband Ripple)				
20dB	1.267	1.267	0.748	0.536
25dB	1.361	1.361	0.853	0.344
30dB	1.425	1.425	0.924	0.226
35dB	1.479	1.479	0.976	0.152
40dB	1.514	1.514	1.015	0.102
Elliptical (1.0dB Passband Ripple)				
20dB	1.570	1.570	0.613	0.805
25dB	1.688	1.688	0.729	0.497
30dB	1.783	1.783	0.812	0.322
35dB	1.852	1.852	0.865	0.214
40dB	1.910	1.910	0.905	0.154

▲ **Table 1. Matching section prototype values.**

Conclusion

Many considerations influence the design of circuits employing wideband transformers. It is not sufficient to conclude the design by simply choosing the wire and the core material to be used. The parasitic inductive and capacitive reactances that accompany the transformer must be considered in the overall circuit design in order to realize the best possible performance. The transformer models and equivalent circuits presented here should help the designer gain an intuitive understanding of the nature of these reactances and their impact on the circuits in which they are to be used. The procedure shown should give the designer a brief but thorough methodology for analyzing the transformers at hand and then completing the circuit design with the transformer parasitics incorporated in a broadband matching circuit with minimal losses. ■

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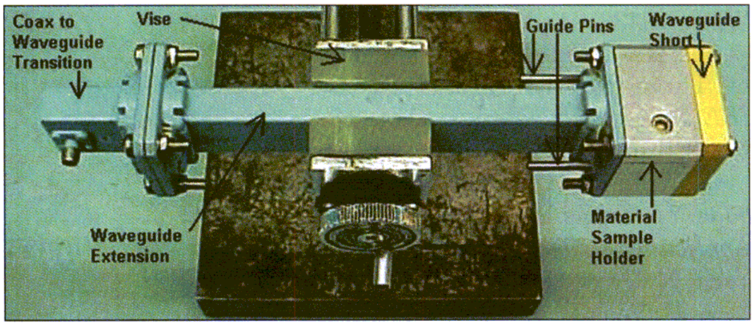
Author information

Chris Trask is a principal engineer at Sonoran Radio Research in Tempe, AZ. He also works as an RF and microwave consultant. He received bachelor of science and master of science degrees in electrical engineering from The Pennsylvania State University. He has published 13 papers and holds nine patents. He can be reached via E-mail: ctrask@qwest.net; or Tel: 480-966-7395.

Waveguide-Based Dielectric Measurements of Solid Low and Medium Dielectric Materials

By Travis J. Mensen and Bryan L. Hauck
Rockwell Collins

This article provides an overview of a method for characterizing the real part of the dielectric constant of solids as implemented at Rockwell Collins. This method is based on the waveguide reflection method originally developed by the National Institue of Standards and Technology (NIST).



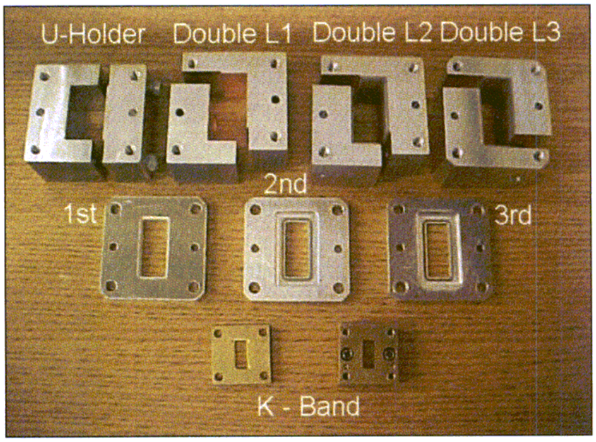
▲ **Figure 1. Complete setup of material sample holder.**

The Rockwell Collins method was developed using an Agilent 8510C for the network analyzer. The calibration kit is an X-band waveguide calibration kit from Maury Microwave. Agilent makes a software package, 85071C, that is based on algorithms designed by NIST for measuring the dielectric constant of materials in waveguide or coax. A computer with a minimum of 64 MB of RAM and a 400 MHZ processor is also used, and communication to the 8510C runs through the GPIB interface. Some of the smaller tools include a screwdriver, wrench, allen wrench (if needed), vise, scissors, copper tape, hardware and a caliber.

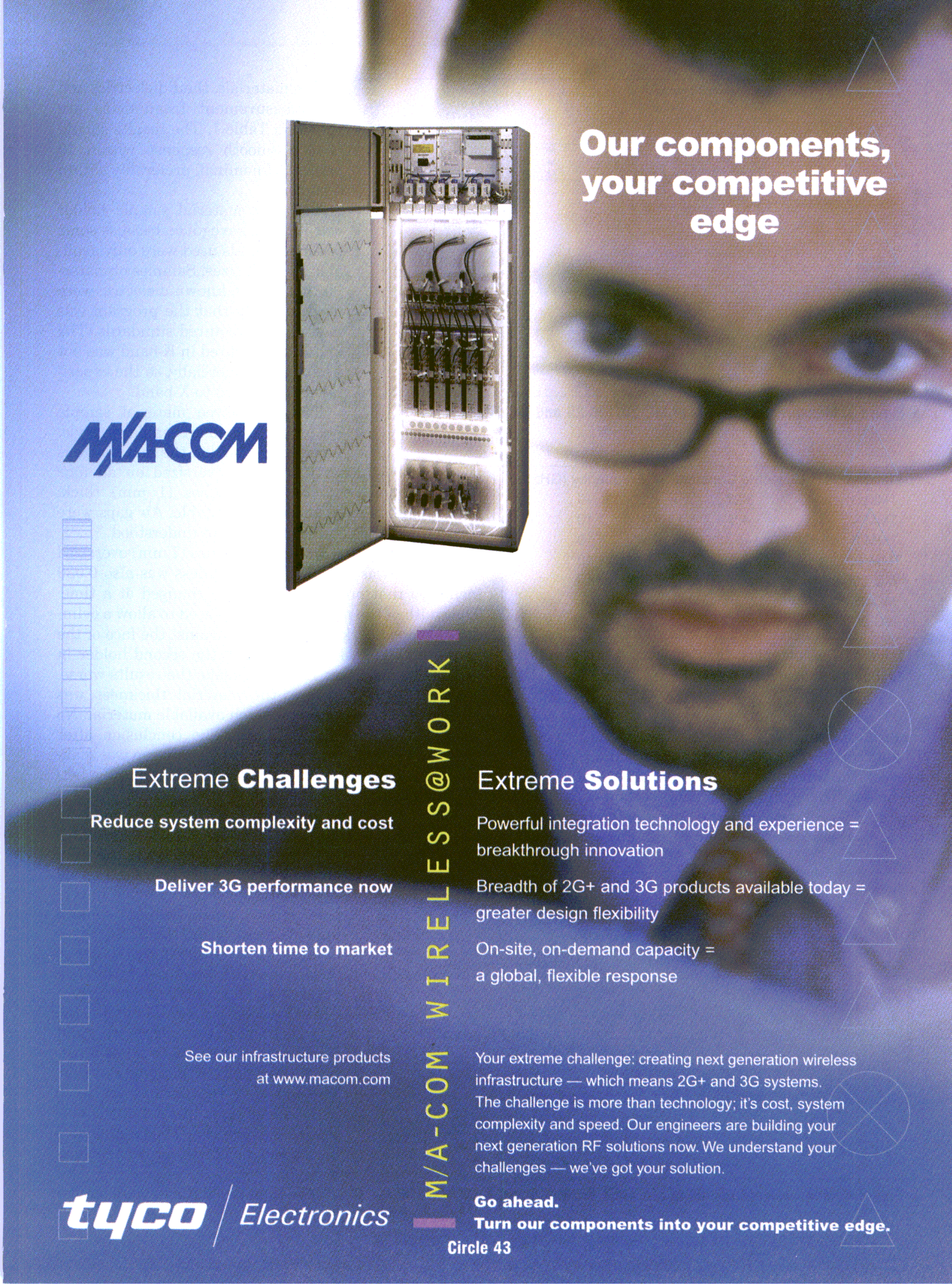
The process for measuring the dielectric constant is very delicate. The first step is to ensure that the waveguide pieces are precisely lined up using guide pins. A waveguide extension is used to prevent modal disruption. The extension is clamped into a vise for convenience and to prevent the cable from moving, which helps prevent phase changes during a measurement (Figure 1).

Starting the Agilent 85071C program loads the parameters into the Agilent 8510, which is necessary to prevent program failure when taking a measurement. The program does not load two parameters, Z_0 and waveguide delay. Z_0 should be set to 1 ohm for the waveguide method, and waveguide delay should be set for the frequency cutoff of the waveguide under the response menu of the Agilent 8510. The program and analyzer are now properly configured for calibration.

The sample holder must be included during the calibration so that it is calibrated out. As shown in Figure 2, the holder is two pieces, so a



▲ **Figure 2. Collection of material sample holders.**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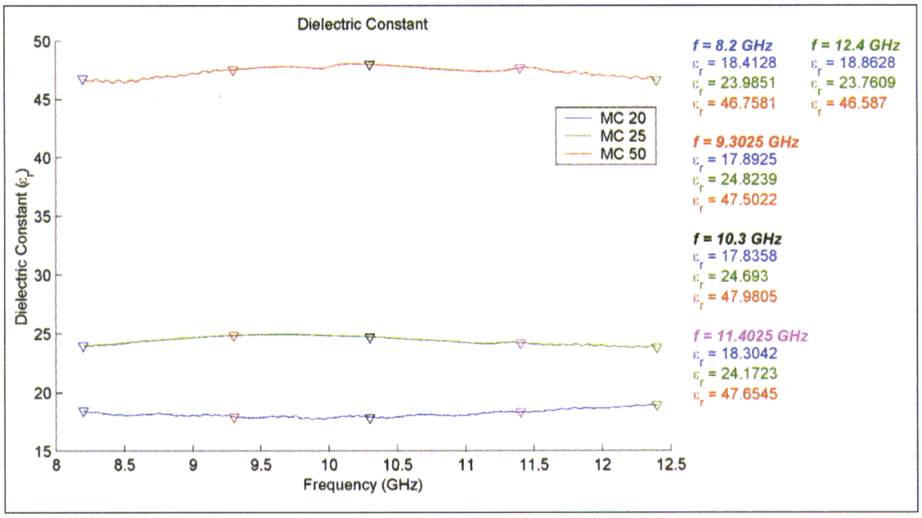


▲ **Figure 3. Measurement results of MC 20, MC 25 and MC 50.**

Material	Dielectric Constant	Frequency (GHz)
TMM 3	3.27	10
TMM 4	4.5	10
TMM 6	6	10
TMM 10	9.2	10
MC 20	19.4	10
MC 25	24.65	6
MC 50	48.13	1

▲ **Table 1. The materials, their dielectrics and the corresponding frequencies.**

small misalignment is possible between the two pieces. To help correct this problem, a copper shim tape is used between the waveguide extension and the material sample holder. The center of the shim is then cut to fit the waveguide opening. When the holder is tightened down, the copper shim conforms and fills in any cracks and/or air gaps.

The materials, their dielectrics and the measurement frequencies are shown in Table 1. The results should have a smooth response within 10 percent of nominal, preferably within 5 percent.

The first material sample holder was K-band waveguide, because some of the materials used were only available in small sizes. Samples of material that have a known dielectric were used to verify that the program was meeting the required standards. The material measured in K-band was not available commercially, so the experiment was moved to X-band.

The first three material sample holders in X-band are similar in style. For the first holder, the material was about 0.0295 inches (0.75 mm) oversized on all four sides. The pieces were 0.039 inches (1 mm) thick. Measurements were not very repeatable. Air gaps were present, but the significance was not understood.

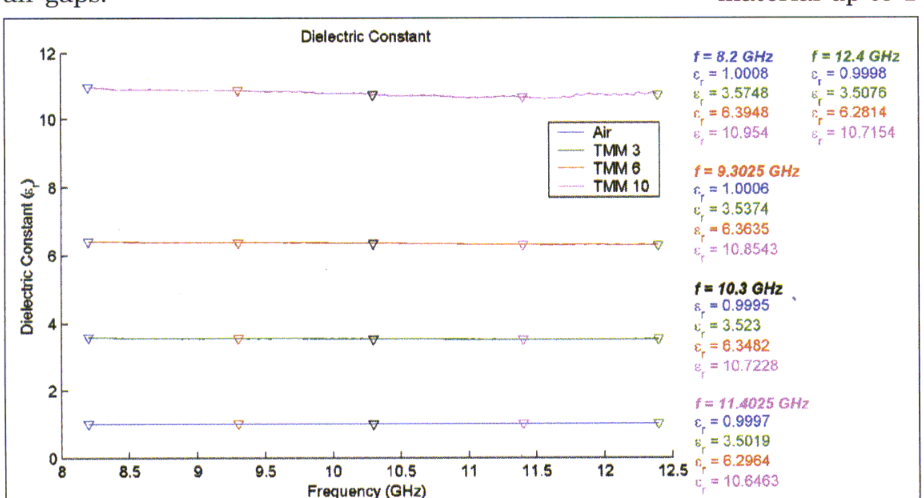
Next, a piece that was 0.2755 inches (7 mm) oversized on all sides was selected. The thickness was also 0.039 inches so that only one parameter changed at a time. The width of the material was increased to allow a cylindrical-shaped gasket to be placed against the face of the material. The gasket is shown in the second holder in Figure 2. The gasket seemed to make the results worse.

For the third fixture, the material thickness was changed to 0.125 inches to match available material. No significant difference was seen. The conclusion after these experiments was that the tiniest air gap was making a difference, so an attempt was made to build a better fixture.

The next two material sample holders are similar in style. Both are 1 inch in thickness and designed for material up to 1 inch thick. The U-holder was the first, designed 0.02 inches oversized on all sides, with the edges wrapped in copper tape. This holder showed improvement over the first three holders, with everything under a dielectric of 10 measured within tolerance, but there were still air gaps.

On the next fixture, double L1, the way that the fixture pieces were put together was changed to allow a better seal around the piece of material. The inside of the fixture was lined with copper tape, instead of the outside of the material. The results were better, but further improvement was desired.

The final two fixtures were similar



▲ **Figure 4. Measurement results of air: TMM 3, TMM 6 and TMM 10.**

to the first double L1. Double L2 is like double L3, except that double L2 was widened by one gasket width instead of two gasket widths. Double L2 explored the gasket being protruded inside the waveguide dimensions, but reflections resulted. The gasket was conductive and 0.028 inches thick and in a sheet form to line the entire inside of the material sample holder. The double L3 inside dimensions were barely over the dimensions of the waveguide, eliminating reflections. The material was somewhat bigger than the opening so that when the fixture was tightened down. The piece pressed into the gasket making an airtight seal. The resulting measurements were reasonably accurate up to a dielectric constant of 50.

Future tests are planned, using material with a higher dielectric constant and comparing results with various material thickness. These tests have identified problems and potential solutions. ■

Author information

Travis Mensen received his associate of applied science degree in electronics and telecommunications from Indian Hills Community College, Ottumwa, Iowa, in May of 2000. He is an engineering technician in the Advanced Technology Center of Rockwell Collins, Cedar Rapids, Iowa. His area of specialty is antenna technology and antenna metrology. His areas of interest include material parameter measurements of high dielectric materials at microwave frequencies, near-field and far-field antenna measurements and general antenna metrology. He may be reached via E-mail: tjmensen@rockwellcollins.com.

Bryan L. Hauck received a bachelor of science degree and master of science degree in electrical engineering from the University of Michigan, Ann Arbor, in 1995 and 1997, respectively. He is a senior electrical engineer in the Advanced Technology Center of Rockwell Collins. His interests include advanced technology for phased array antenna applications, material parameter measurements of high dielectric materials at microwave frequencies and antenna metrology. He is a member of the IEEE Antennas and Propagation Society, the IEEE Microwave Theory and Techniques Society, the Antenna Measurement Techniques Association and Sigma Xi, The Scientific Research Society. He may be reached via E-mail: blhauck@rockwellcollins.com.

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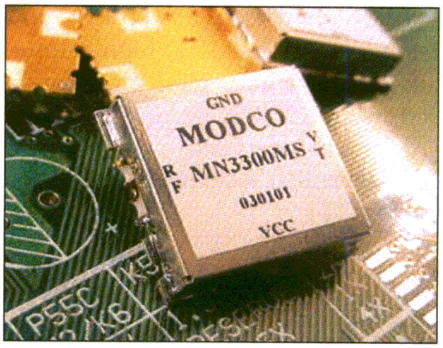

frequency. Offering up to 5 percent tuning bandwidths over a range of 2.4 to 4.2 GHz, the MN series operates on +5 VDC at less than 35 mA current. The MN series VCO is supplied in standard 0.5 × 0.5 × 0.13 inch surface-mount housing and is also available in UM series, offered in a 0.35 × 0.35 × 0.10 inch surface-mount package.

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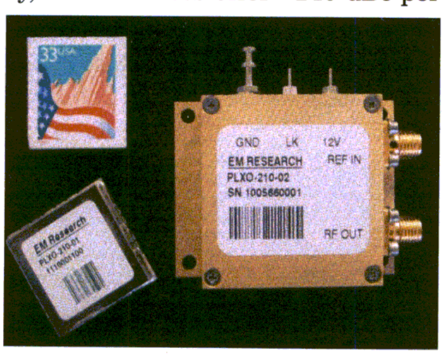


hertz phase noise at 10 kHz offset. Available in 1-inch square surface-mount or connectorized 1.5-inch square packages, the PLXO Series is ideally suited for use as high stability low phase noise reference (clock) oscillators.

EM Research, Inc.
Circle #164

High frequency oscillators

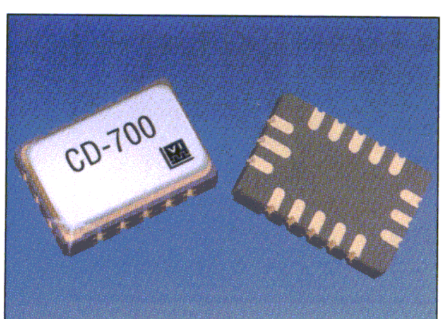
Champion Technologies has announced two new additions to its K1526/K1536 VCXO series, designated BLC and CLC. These devices offer a frequency range of 55.1 to 80 MHz, with frequency stability as

low as ±40 ppm in temperatures ranging from 0 to 70 degrees Celsius. Typically, these devices would be used in phase-locked loops, clock recovery, reference signal tracking and other applications.

Champion Technologies, Inc.
Circle #165

VCXO based phase-locked loop solution

Vectron International has introduced a new VCXO based phase-locked loop solution to simplify a variety of clock recovery and data retiming, frequency translation, clock smoothing and clock switching applications. The small size, versatile functionality and high performance of the CD-700 make it ideal for various applications in DWDM, switching, wireless bases-

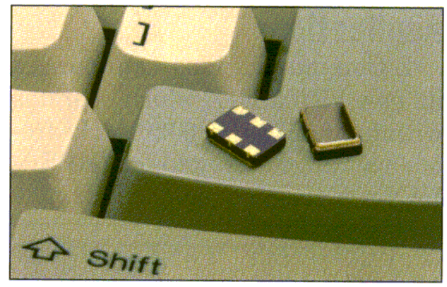


tation, ATM, SONET/SDH, xDSL, network communication, digital audio/video and PBX systems.

Vectron International
Circle #166

Surface-mount VCXO

M-tron has introduced the UVV Series surface-mount LVPECL/LVDS VCXO. Available in frequencies between 750 kHz and 650 MHz, these VCXOs provide either LVPEVL or LVDS compatible outputs in a standard 5 × 7 mm leadless ceramic package. The UVV products offer absolute pull ranges



from ± 25 to 100 ppm, –40 to +85 degrees Celsius operation with tri-state and tight (45/55 percent) symmetry. Low jitter and phase noise performance make the UVV ideal for SONET/SDH/ATM, optical carrier, DWDM, WDM, gigabit ethernet and fiber channel, PLL and clock recovery applications.

M-tron Industries
Circle #167

Module development kit

MTI-Milliren Technologies has announced a new development kit to be used in conjunction with its line of Stratum 3 and 3e synchronous timing modules. The development kit uses simple windows interface to allow design engineers access to the multiplicity of user-programmable features of the TM1 Series Stratum 3 and 3e network synchronization timing modules. Some of the features include six user programmable input reference frequencies and seven user selectable output reference frequencies from 2 kHz to 311.04 MHz.

MTI-Milliren Technologies, Inc.
Circle #168

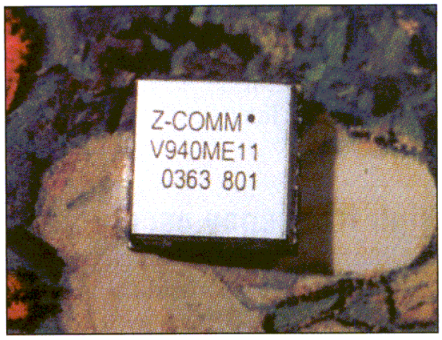
Ethernet clock oscillators

CTS has announced the addition of 125.000 MHz gigabit Ethernet clock oscillators to its CB3 family of 5 × 7 mm hermetically sealed, ceramic oscillators. The standard four-pad, 5 × 7 mm CTS surface mount oscillator packages are fully compatible with SMT pick-and-place assembly equipment and are supplied in standard tape-and-reel packaging.

CTS Corporation
Circle #169

VCO for U-NII band applications

Z-Communications announces its new V940ME11, which operates within the 5220 to 5420 MHz frequency range, is suited for HIPERLAN systems and delivers a single



sideband phase noise of –82 and –107 dBc per hertz at 10 and 100 kHz, respectively. The device also delivers 4 ±2 dBm into a 50-watt load with nominal 4.25 volt supply while drawing less than 25 mA.

Z-Communications, Inc.
Circle #170

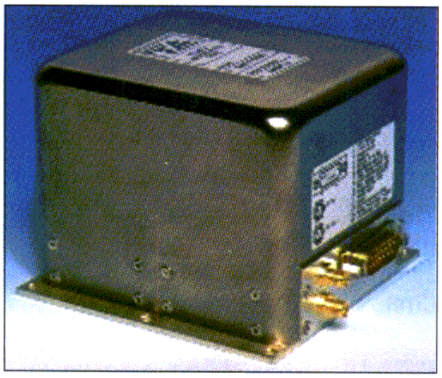
Synthesizers

Communication Techniques has introduced a series of synthesizers for the high data rate point-to-point and point-to-multipoint digital radio marketplace, as well as the satellite communications market. DRS series synthesizers are available for frequencies from 4 to 15 GHz in bands with step sizes from as low as 10 kHz. Other product features include low phase noise of –119 dBc per hertz at 100 kHz offset typical at 10 GHz, high immunity to phase hits, small size, bandwidths to 1200 MHz and external or internal reference.

Communication Techniques, Inc.
Circle #171

Rubidium atomic clock oscillator

TEMEX has announced the availability of a new Rubidium atomic clock, the RAFS series, approved by the European Space Agency for use in the Galileo Global Navigational System. The RAFS has a low temperature sensitivity of less than 1×10^{-13} per degree Celsius, required



for radio-navigation applications. The RAFS exhibits short-term stability of less than 5×10^{-14} per 10,000 seconds and drift of less than 3×10^{-13} per day.

TEMEX
Circle #172

AMPLIFIERS

SiGe:C integrated RF cascode amplifier

A cost-effective, high isolation cascode amplifier using silicon germanium: carbon (SiGe:C) technology has been introduced by Motorola's Semiconductor Products Sector. The MBC13916 amplifier is intended to help to simplify RF designs for 100 MHz to 2.5 GHz applications and is available in an SOT-343R miniature surface-mount package. This general purpose SiGe:C RF cascode amplifier is intended to be a similar but improved replacement for the MRFIC0916, which was introduced by Motorola in 1998.

Motorola, Inc.
Circle #173

Four-stage GaAs pHEMT buffer amplifier

Mimix Broadband has introduced a GaAs (gallium arsenide) monolithic microwave integrated circuit (MMIC) four-stage buffer amplifier. Using 0.15 micron gate length GaAs pHEMT device model technology, this gain block amplifier covers the 36 to 43 GHz frequency bands. The MMIC device has a typical gain of 25 dB, with gain increasing with frequency to compensate for other component roll-off factors common in 38 to 40 GHz systems. The device has a typical noise figure of 4 dB across the band.

Mimix Broadband
Circle #174

High power X-band solid state RF amplifier

Aethercomm has introduced its new solid state RF amplifier. The P/N SSPA 7.1-7.3-10 is a high power X-band SSPA used in the

DSCC exciter-transmitter subsystem. Features include 5 watts linear or pulsed RF power minimum, operation from 7.1 to 7.3 GHz, 0.2 dB gain flatness, 47 dB minimum small signal gain and an RF inhibit switch that gives 40 dB isolation minimum.

Aethercomm, Inc.
Circle #175

Gain block amplifiers

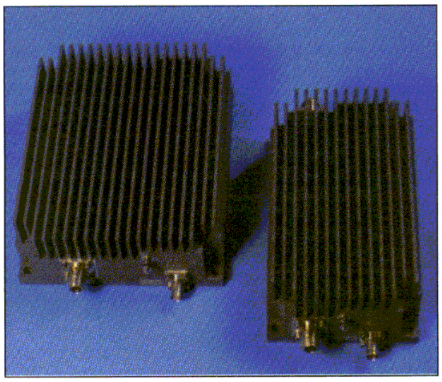
Anadigics has launched the AGB3301, the first device in a new product line of gain block amplifiers designed for use in environments requiring high linearity, low noise and low distortion. The AGB3301 is ideal for transmit and receive paths in 50-ohm wireless applications, such as cellular base stations, WLL and MMDS, all of which require highly linear components.

Anadigics, Inc.
Circle #176

Products

Multi-carrier IF line compensation amplifiers

Pascall Electronics has introduced its transmitting and receiving multi-carrier IF line compensation amplifiers to provide compen-



sation for cable loss in broadband microwave wireless systems. Pascall s line compensation amplifiers minimize the effects of cable loss and crossband slope without any significant degradation of the multi-carrier signal. The amplifiers are designed with a gain profile that compensates for the transmission characteristics of commonly used cables.
Pascall Electronics Limited
Circle #177

CABLES & CONNECTORS

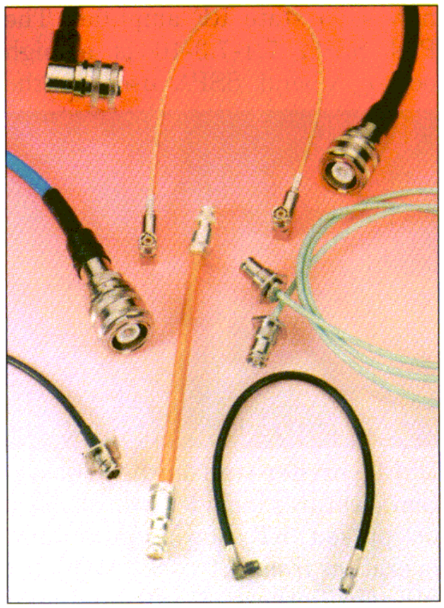
New low-loss connector

RF Connectors has released a new 7-16 DIN series coaxial connector designed for use with LMR-600fi low-loss cable from Times Microwave and WBC-600 low-loss cable from CommScope. The RFD-1604-2L2 is a 7-16 DIN male crimp

connector featuring a silver plated body and contact for intermodulation distortion reduction. The connector uses Teflonfi insulation for excellent dielectric performance.
RF Connectors
Circle #178

Coaxial cable assemblies

Custom fabricated standard and high power RF coaxial cable assemblies featuring various dielectrics, jacketing or plenum rated materials are being introduced by Tru-Connector. The custom cable assemblies are manufactured to



specification to meet special application and environmental requirements and can include various connector types, such as SMA, 7/16, C, LC, SC, HN, QDS and Triax.
Tru-Connector Corporation
Circle #179

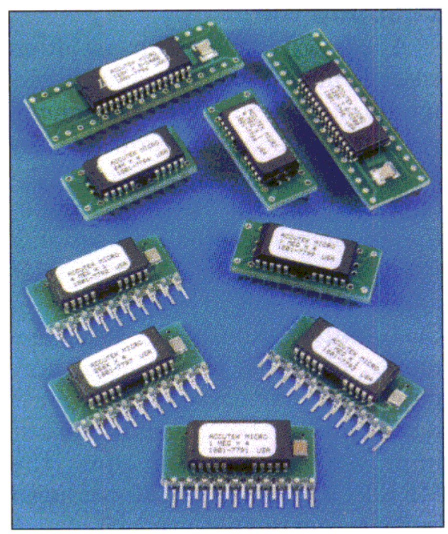
Telecom cable tool kit

The new TK-2 tool kit from Jensen is a complete set of tools specifically designed for cable sheath slitting, ring cutting, lacing, sewing, parallel drop wire slitting, switch or panel board terminating. The tool kit comes in a padded, zippered vinyl carrying case and is ideal for use in the central office or in the field. The kit includes a sheath stripper, sheath slitter, ring cutting tool, curved sewing needle,

straight seven-inch sewing needle, wire loop lacing needle, aerial drop wire slitter and a straight, flat metal sewing needle.
Jensen Tools, Inc.
Circle #180

DRAM and SRAM adapters

Accutek Microcircuit has introduced a line of DRAM and SRAM DIP and ZIP adapters that are pin-to-pin replacements for end-of-life ICs used in telecommunications, machine vision, VME, DSP, networking, data acquisition and high speed graphics applications. Configurations include 1 MEG × 1, 4 MEG × 1, 256 K × 4 and 1 MEG × 4 DRAM DIPs and ZIPs, a 64 K × 4



DRAM DIP and a 128 K × 8 SRAM DIP with 0.300 mil and 0.400 mil lead row centers.
Accutek Microcircuit Corporation
Circle #181

Coaxial connectors

Cristek Interconnects has introduced a coaxial connector with a hybrid combination of coaxial and signal contacts in one shell. Using

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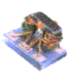

Measuring only 0.15"x0.15" square, the DBTC series from Mini-Circuits is quite simply the smallest 5 to 2000MHz directional coupler series on Earth! Available in 9 to 20db nominal coupling values, these patented 50&75 ohm couplers integrate Blue Cell™ design techniques for very flat response, low insertion loss, and multi-decade broad bandwidths. All-welded connections improve reliability, and automated production delivers high unit-to-unit performance repeatability. Preserve precious board space, and capital as well. Specify Mini-Circuits DBTC directional couplers...priced at only $1.99 each (qty. 25)!

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Coupling	Model	Freq. (MHz)	Ins. Loss (dB) Midband Typ	Directivity (dB) Midband Typ
9dB	DBTC-9-4	5-1000	1.2	18
10dB	DBTC-10-4-75	5-1000	1.4	20
12dB	DBTC-12-4	5-1000	0.7	21
13dB	DBTC-13-4	5-1000	0.7	18
13dB	DBTC-13-5-75	5-1000	1.0	19
		1000-1500	1.4	17
16dB	DBTC-16-5-75	5-1000	1.0	21
		1000-1500	1.3	19
17dB	DBTC-17-5	50-1000	0.9	20
		1000-1500	1.0	20
		1500-2000	1.1	14
18dB	DBTC-18-4-75	5-1000	0.8	21
20dB	DBTC-20-4	20-1000	0.4	21

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the Micro-D shell, designers can meet the demand for smaller, tighter packaging and weight while maintaining signal integrity. Cristek CCD connectors provide RF performance in standard Micro-D metal shells and uses RG 176, RG 179, RG 196 and RG 316 coax/power wire type.

Cristek Interconnects, Inc.
Circle #182

D and E connectors

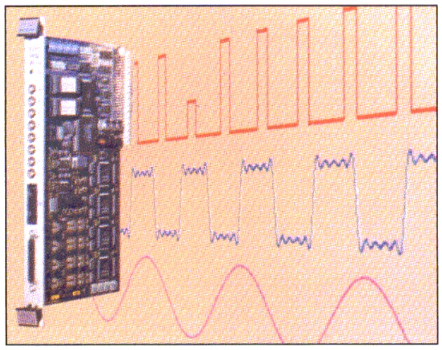
AVX now offers 2-mm hard metric type D and E male and female connectors. These connectors, designated the 8077 series, allow for a large number of signal pins in a small space, saving PCB real estate. Additionally, the high number of signal pins allows for use in higher speed applications.

AVX Corp.
Circle #183

TEST EQUIPMENT

Waveform generator

Highland Technology has introduced a new VME four-channel arbitrary waveform generator. The Model V370 was designed in coop-

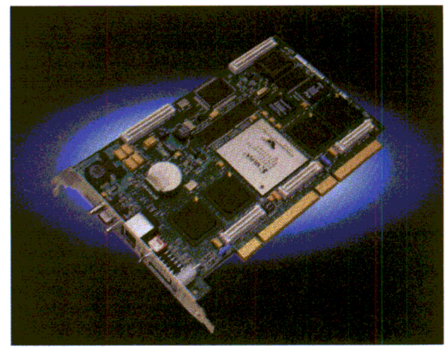


eration with major aircraft engine and power systems manufacturers to provide the features needed for accurate simulation of real-world sensor inputs. The V370 is ideal for simulation of sensor signals from complex rotating machines, generation of pure and distorted polyphase AC waveforms, generation of complex pulse trains with realtime control of relative pulse positions and amplitudes and shake table, servo or actuator drive.

Highland Technology
Circle #184

Analyzer and exerciser

VMETRO introduces an addition to its bus analyzer family, the PBT-



615 PCI-X bus analyzer and exerciser. Designed for debugging, testing and validating next-generation PCI-X boards and systems, the PBT-615 is a single-slot short card containing a complete logic analyzer, exerciser and protocol checker for the PCI-X and PCI bus.

VMETRO, Inc.
Circle #185

Compliance test system

IndexSAR has introduced an SAR compliance test system for automatically measuring specific absorption rate from mobile telephones and other wireless devices to the draft CENELEC and draft IEEE standards. Using a 6-axis, compact, industrial robot with probe positioning accuracy of ±0.04

mm, the system can measure SAR from mobile telephones and other wireless devices using an upright generic phantom head.

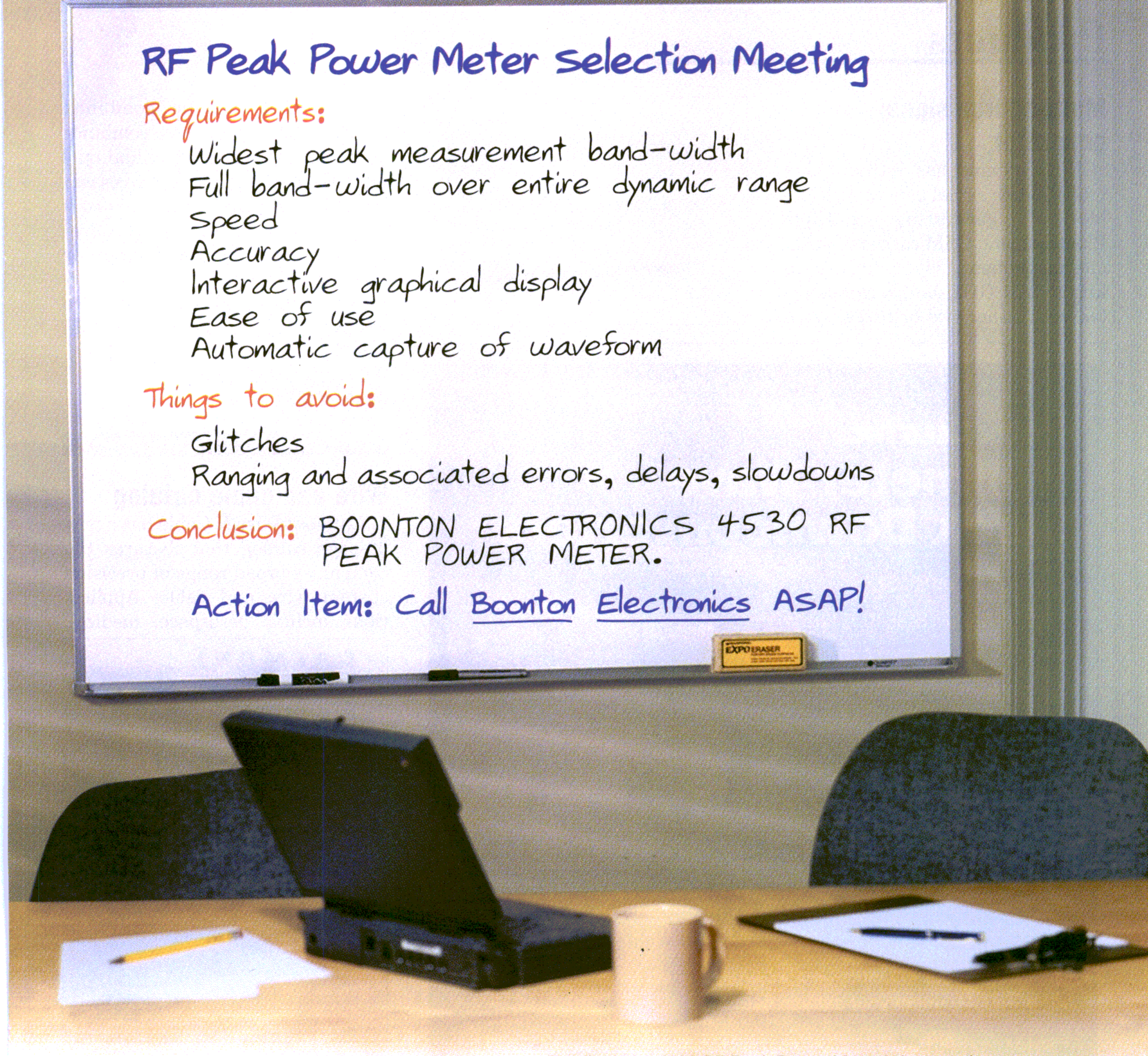
IndexSAR
Circle #186

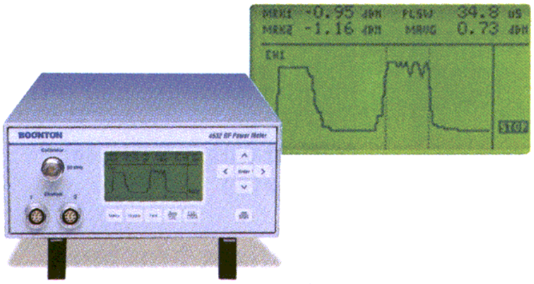
Remote monitoring analyzer

Morrow Technologies has launched its real-time, remote monitoring P9116 satcom spectrum

analyzer. The full-featured P9116 reduces both downtime and field service calls by allowing remote system monitoring from virtually any location via local area network, modem or the Internet. The analyzer was designed for ease of use and convenience.

Morrow Technologies Corporation
Circle #187

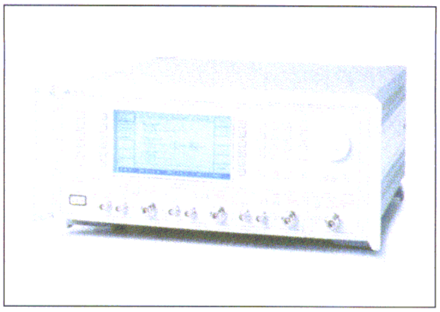

RF Peak Power Meter Selection Meeting
Requirements:
Widest peak measurement band-width
Full band-width over entire dynamic range
Speed
Accuracy
Interactive graphical display
Ease of use
Automatic capture of waveform
Things to avoid:
Glitches
Ranging and associated errors, delays, slowdowns
Conclusion: BOONTON ELECTRONICS 4530 RF PEAK POWER METER.
Action Item: Call Boonton Electronics ASAP!



Products

Multi-source signal generator

IFR Systems has introduced a multi-source signal generator capable of generating multiple Bluetooth and GSM carriers within one instrument. The 2026B, a 10 kHz to 2.51 GHz source, can be fitted with either two or three signal



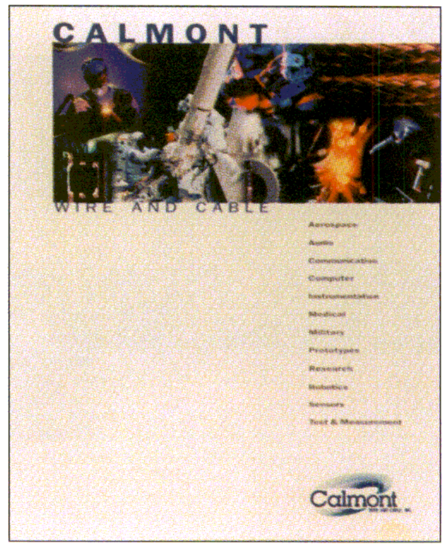
sources that can be combined together into a common combiner output or routed to individual outputs. Each of the signal sources can be frequency and amplitude-coupled so that common measurements, such as intermodulation or receiver selectivity, can be performed more quickly and more accurately than using individual signal generators.

IFR Systems, Inc.
Circle #188

PRODUCT LITERATURE

Wire and cable catalog

Calmont Wire & Cable has a new 30-page catalog that features the company s broad range of precision custom wire and cable. Applications include aerospace, medical,



robotics, sensors, military, instrumentation, computer and high-end audio. The catalog is subdivided according to construction, complexity and application, with particular emphasis to Calmont s high-flex, high temperature and ultra-miniature cables.

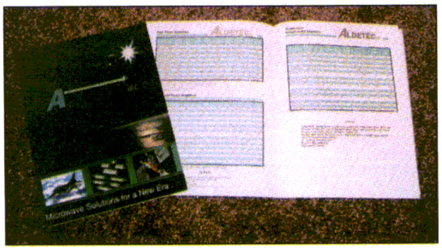
Calmont Wire & Cable, Inc.
Circle #189

Amplifiers brochure

A new 12-page short-form brochure from Aldetec describes the company s product line of RF and microwave amplifiers. Included in the product line are standard

Products

broadband amplifiers covering 500 MHz to 18 GHz, broadband power amplifiers with >2 watt P_{1dB} outputs, high power amplifiers featur-

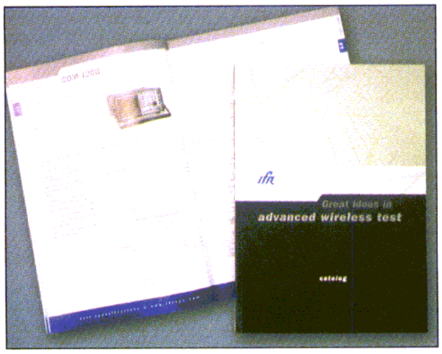


ing 10 watt P_{1dB} outputs to >14.5 GHz, low noise amplifiers featuring < 1 dB NF to > 15 GHz and temperature compensated amplifiers guaranteed over a range of –54 to +85 degrees Celsius.

Aldetec, Inc.
Circle #190

Wireless radio test sets catalog

IFR Systems has released its latest product catalog, covering specifications on the company's entire product offering. Included are radio



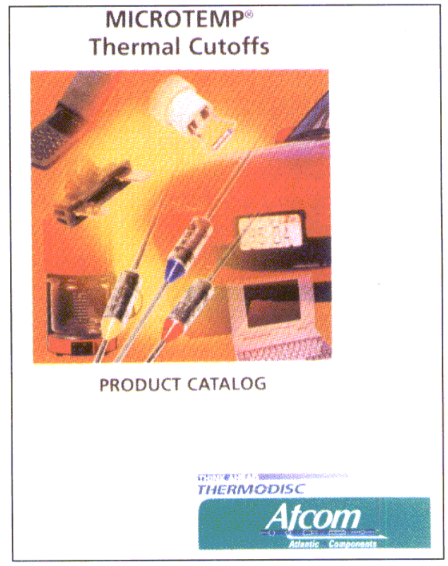
test sets that cover multiple standards and frequencies; signal sources that produce high-resolution, low phase noise carrier signals; and hand-held and desktop tele-communications testers.

IFR Systems, Inc.
Circle #191

Thermal cutoffs catalog

A new catalog describing a full range of thermal cutoffs is being offered by Atlantic Components. The MICROTEMP® Thermal Cutoffs Product Catalog features devices that protect against overheating by interrupting the electrical circuit when operating temperatures exceed their rated temperature. Different temperature and package configurations, along with application, installation, calibration and test procedures, are also provided in the catalog.

Atlantic Components
Circle #192



Low-Cost X-Band Gunn Oscillator for Educational and Research Purposes

By José-María Zamanillo, Constantino Pérez-Vega, Juan A. Saiz Ipiña and Miguel A. Solano
University of Cantabria

This article introduces a low-cost, high-performance X-band Gunn oscillator for educational and research purposes. The cavity was manufactured in standard waveguide WR90 with a UBR100 flange in order to make it compatible with waveguide circuitry available in a basic microwave laboratory. The ability to sweep the entire X-band (8 to 12 GHz) with constant output power and low phase noise, added to its low cost, makes this oscillator a broadly useful tool.

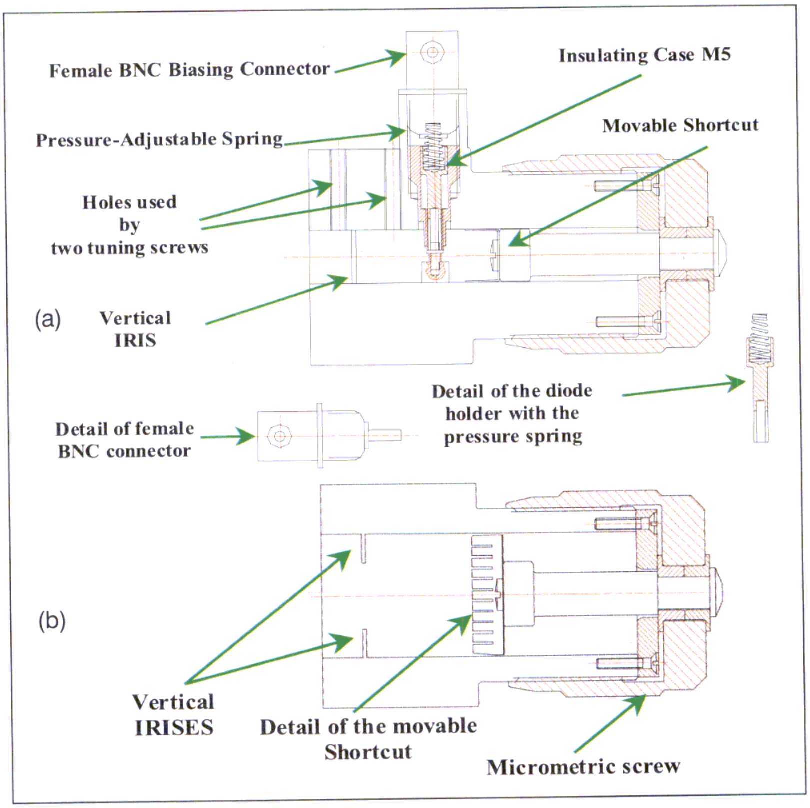


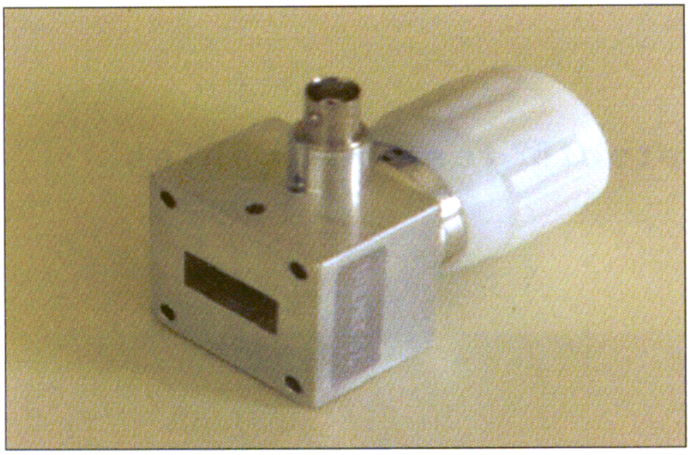
Figure 1. Mechanical drawing of the waveguide Gunn oscillator, showing (a) transversal view and (b) top view.

Introduction

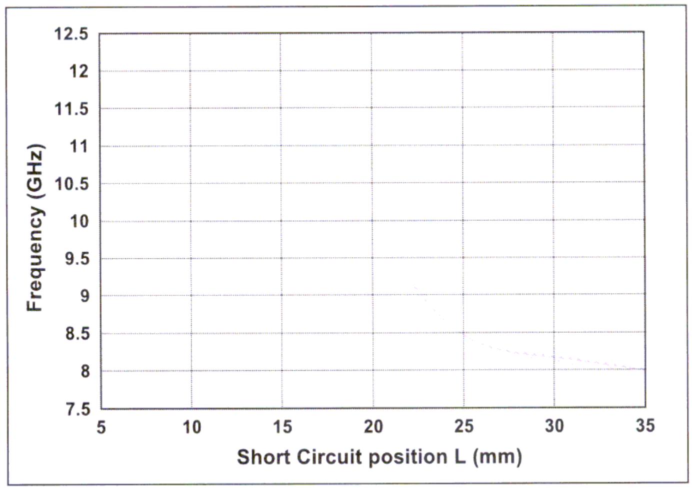
Given the limited resources of universities to equip microwave laboratories with the basic instrumentation required for educational uses, we decided to develop a high-performance, low-cost, robust microwave oscillator capable of sweeping the entire X-band (8 to 12 GHz) at constant output power with low phase noise. The oscillator would be suitable as a microwave source in a wide range of experiments. In addition, taking into account the fragility of Gunn diodes regarding their biasing, we designed an external protection circuit for the device. As a result, if the user reverses the supply polarity or exceeds the maximum working voltage, the Gunn diode will not be damaged.

Gunn oscillator

In a Gunn oscillator, the frequency of oscillation is determined by the resonant circuit, loaded by the impedance of the device. Experimental results and computer simulation [1–3] have shown that the device can be tuned over more than one octave bandwidth by an external cavity. The frequency is decreased as the inserted length of the tuning screw increas-



▲ **Figure 2. Photograph of the oscillator from Figure 1.**



▲ **Figure 3. Mechanical tuning characteristic of the configuration shown in Figure 4. Gunn device MA49156 V_{Bias} = 14.0 volts.**

es and the movable short circuit allows the user to change the frequency over a broader range.

Gunn oscillators are generally tuned over a wide frequency range by changing the reactance of the microwave circuit loading the device. The frequency of oscillation depends on the requirement that the susceptance of the device must resonate with the circuit susceptance.

When a Gunn device mounted in a microwave circuit is tuned mechanically by varying the sliding short-circuit position, the frequency of oscillation corresponds to a half-wavelength (or a multiple) of the cavity length. Deviations from this behavior can be attributed to either the mounting configuration of the device in the particular circuit, its package, or its finite capacitance. Because the tuning characteristic follows the half wavelength behavior, the loaded *Q* (quality factor) of the cavity increases. We decided to use a Gunn diode type MA49156 due to its excellent cost-performance ratio [4].

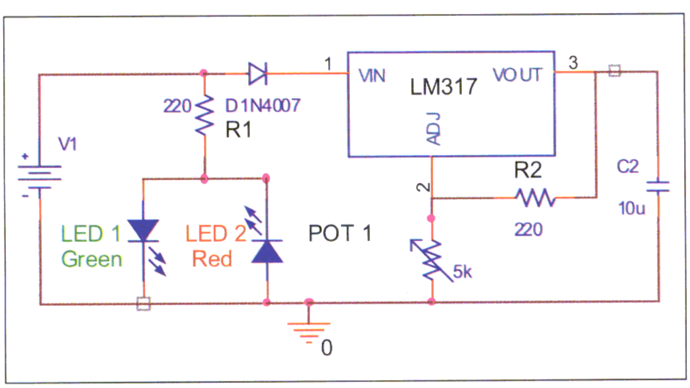
The mechanical drawing of the assembly is shown in Figures 1(a) and 1(b) for transversal and top views, respectively. Figure 2 shows the assembled oscillator with its WR-90 waveguide and UBR-100 flange.

Figure 3 shows the variation of frequency versus the length in millimeters of the movable short circuit from the diode position.

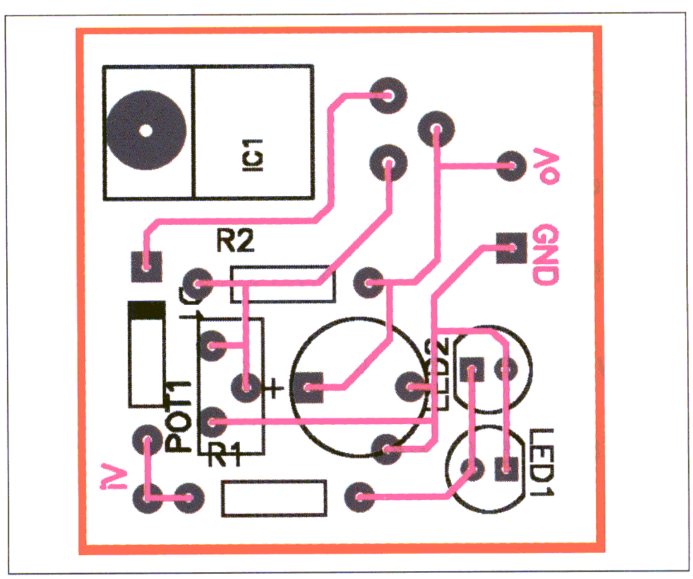
Protection circuit

The schematic for the external protection circuit is shown in Figure 4(a) and the PCB layout of the circuit in Figure 4(b). This circuit allows the biasing of the Gunn diode under a wide range of conditions. If the biasing polarity is reversed, the power diode D1N4007 does not conduct, and the output voltage is zero.

The programmable linear regulator LM317 is set to the optimum bias voltage (in this case 14 volts) for the Gunn diode by varying the value of the potentiometer POT1. If the input bias range exceeds the maximum operation voltage of the Gunn diode, the linear regulator adjusts the output voltage at 14 volts (or the maximum voltage set by POT1). Even if the DC supply voltage is increased to up to 40 volts — the maximum permissible voltage at the input of the LM317 — the Gunn diode will not be damaged.



▲ **Figure 4(a). Diagram of the protection circuit.**



▲ **Figure 4(b). PCB layout of the protection circuit.**

This oscillator generates a clean spectrum over the entire tuning range. The mean power obtained at these frequencies was 1.8 mW, with good phase noise characteristics. The observed output power over the entire X-band was fairly constant. We contemplate the extension of this cavity for higher power diodes. ■

References

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2. R. P. Owens and D. Cawsey, "Microwave Equivalent-Circuit Parameters of Gunn-Effect-Device Packages," *Transactions on Microwave Theory and Techniques 18.11*, November 1970.

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Author information

José-María Zamanillo received a bachelor of science degree and a Ph.D. in physics from the University of Cantabria, Spain, in 1988 and 1996, respectively. Since 1988, he has been devoted to education and research at the University of Cantabria, where he is an associate professor. He has been engaged in various European and Spanish projects, mainly in the fields of microwaves and device modeling. His research interests include linear and nonlinear modeling of GaAs MESFETs, HEMTs, (HBTs) and microwave active devices. He may be reached via E-mail: jose.zamanillo@unican.es; Tel: +34-942-200887; or Fax: +34-942-201488.

Constantino Pérez-Vega received a degree in electronics and communications engineering from the Escuela Superior de Ingeniería Mecánica y Eléctrica of México in 1965 and a Ph.D. in telecommunications engineering from the University of Cantabria, Spain, in 1997. He held several technical managing positions in the Mexican Government Radio and Television System since 1972 and was director of engineering at the Instituto Mexicano de Televisión, as well as technical advisor at the Dirección General de Radio, Televisión y Cinematografía of the Secretaría de Gobernación in México until 1988. He has been also dedicated to teaching since 1966 in México. Since 1989, he has been a professor at the University of Cantabria in the areas of television and communication systems.

Juan A. Saiz Ipiña received a bachelor of science degree and a Ph.D. in physics from the University of Cantabria, Spain. He has been an associate professor in the Communications Engineering Department at the University of Cantabria since 1990. His research interests include the fields of computational methods in electromagnetics, numerical analysis of waveguides and scattering problems.

Miguel A. Solano received a master of science degree in physics in 1984 and a Ph.D. in physics in 1991, both from the University of Cantabria, Spain. Since 1984, he has been with the University of Cantabria, formerly in the Electronics Department and now in the Communication Engineering Department, where he became an associate professor in 1995. In 1992, he worked at the Centre for Communication Research, University of Bristol, U.K., through a research grant from N.A.T.O. His current research activities include electromagnetic propagation in waveguide structures, with anisotropic and bianisoptropic media and numerical methods in electromagnetics.

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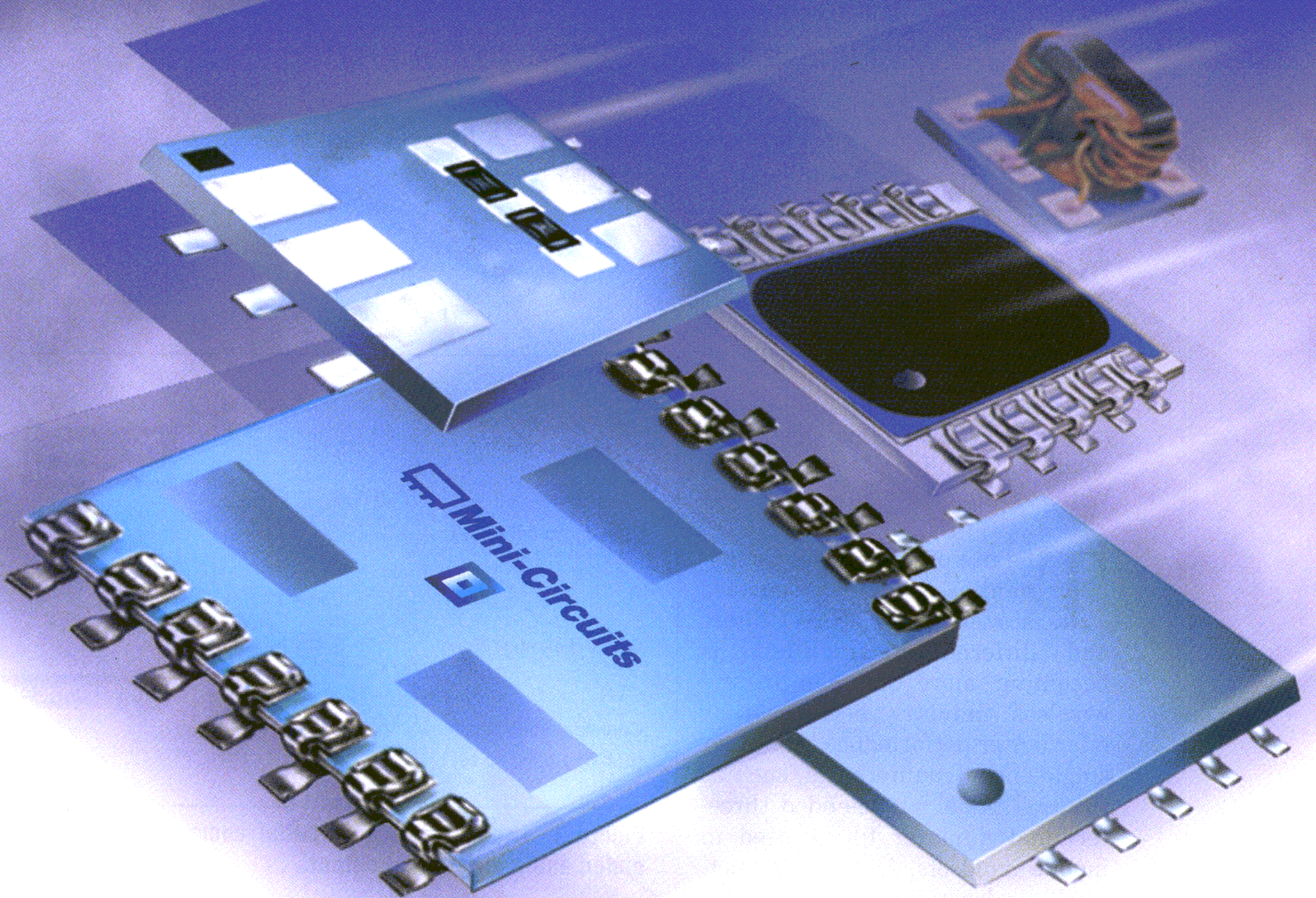

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2	0	SBB	5	800-2300	22-24	0.5-0.6	3.0-4.0	4.95
2	90	QBA	7	340-2400	21-28	0.25-0.80	3.0-7.0	6.95
4	0	SBD	1	1800-2600	20	1.0	8.0	9.95

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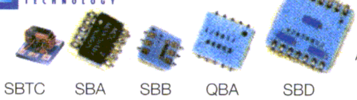

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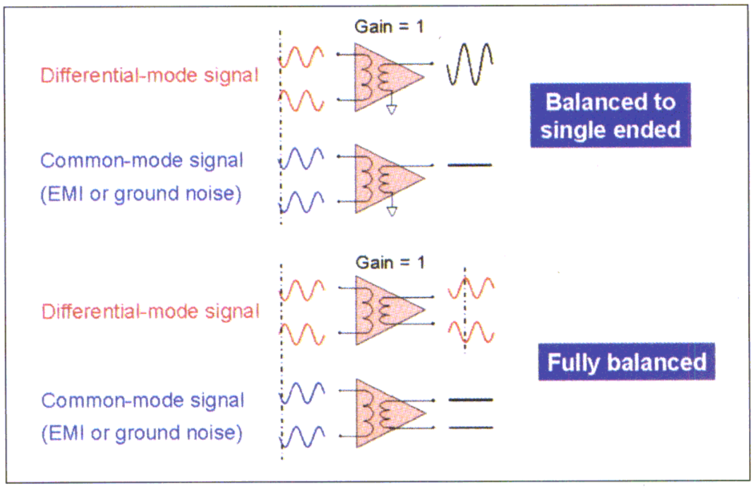
Comparing Differential Measurement Techniques

By Loren Betts
Agilent Technologies

Editor's note: This is part one of a two-part article. The second part will appear in the December 2001 issue of Applied Microwave & Wireless *magazine.*

The use of differential components, such as surface acoustic wave (SAW) filters and differential amplifiers is becoming more common in the wireless industry because they offer better performance than their single-ended counterparts, such as the traditional single-ended three-terminal duplexer filters used in mobile handsets. Most vector network analyzers have single-ended RF ports that cannot directly measure differential parameters, which makes accurate measurement of these components challenging.



▲ **Figure 1. Two types of differential components, with single-ended and a differential output.**

However, there are several alternative methods of obtaining the differential parameters needed to characterize these devices. This and a subsequent article discusses the challenges designers face in measuring the performance of differential components, and describes each of the most widely used techniques. Although each technique produces a specific level of accuracy that depends on the characteristics of the device to be tested, the "calculated mix-mode" method provides the most accurate device characterization and has the fewest drawbacks.

Figure 1 shows two types of differential components, one with a single-ended output and the other with a differential output. The differential port of the devices consists of a pair of these outputs. Differential components are unique in that signals are referenced not only to a common ground but to each other as well. The signals referenced to each other are called "differential mode" and the signals referenced to a common ground are called "common mode." Differential components can have both common mode and differential mode signals.

In most cases, the differential mode signals are anti-phase because their phase relative to each other is 180 degrees, which creates a virtual ground along the axis of symmetry of the device. At the virtual ground, the potential at the operating frequency does not change with time, regardless of the signal amplitude. Common-mode signals are induced at the terminals of the device with the same phase and amplitude relationship.

While a differential component has no perfor-

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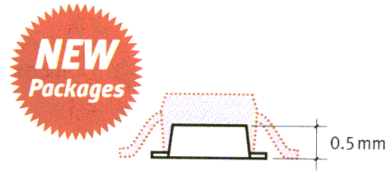

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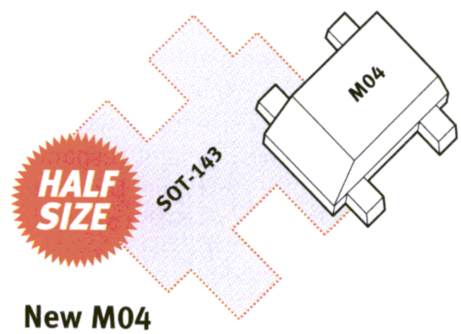

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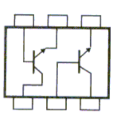

Part Number	Description	Q1 Spec	Q2 Spec
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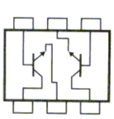
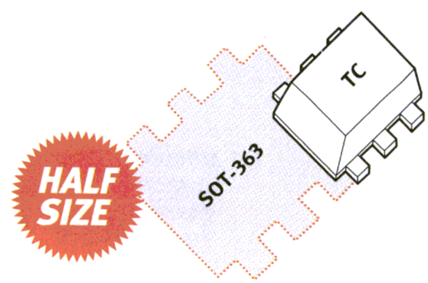
UPA814TC	Matched Die/Cascade LNA	NE688	NE688



Part Number	Description	Q1 Spec	Q2 Spec
UPA826TC	Matched Die/Osc-Buffer Amp	NE685	NE685
UPA840TC	Mixed Die/Osc-Buffer Amp	NE685	NE681



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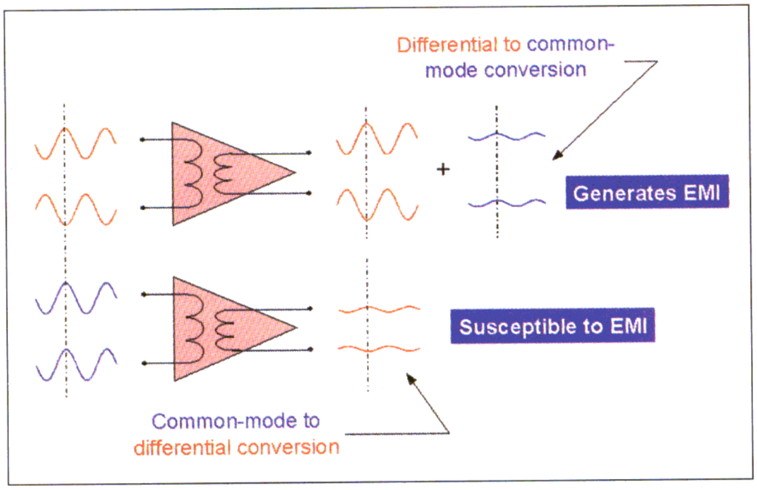
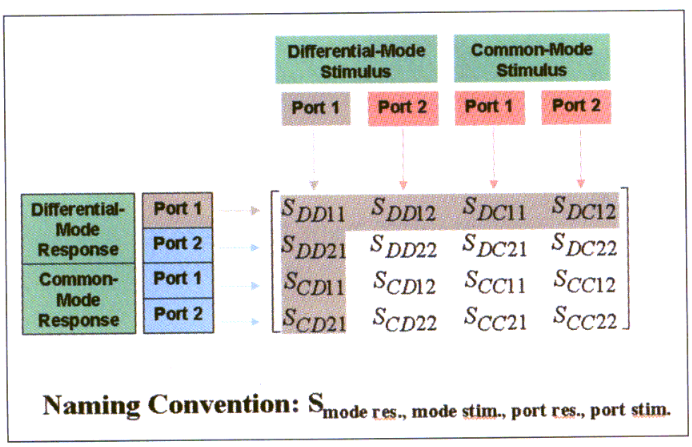


▲ **Figure 2. Mode conversion occurs when some of the differential signal is converted into the common mode and some of the common-mode signal is converted into the differential mode.**

mance advantage over a single-ended component when used in common mode, it exhibits significant benefits when used in differential mode because it will pass differential-mode signals and reject common-mode signals. For example, noise from a power supply will affect both terminals of the device equally with the same phase relationship. The device does not respond to these common-mode signals and attenuates them at the output. Since no device is ideal, some of the applied differential signal applied is converted into the common mode and some of the common-mode signal is converted into the differential mode. This is referred to as mode conversion (Figure 2), and is directly related to noise immunity of the device.

Measurement obstacles

Vector network analyzers are typically employed to measure RF components and are not designed for measuring differential parameters. Their RF ports are single-ended with common impedance values and cannot supply differential and common mode signals to the device. Single-ended devices have impedances of 50 to 75 ohms; differential components have no standard impedance values. To measure a four-terminal (two-port differential) component requires 16 S-parameters, but these single-ended S-parameters are not sufficient to accurately characterize a differential component operating in differential mode. As a result, measuring differential components accurately using a single-ended analyzer cannot be accomplished without applying some type of hardware or software conversion to the single-ended data. Two common approaches use either a balun, or perform some type of mathematical transform.

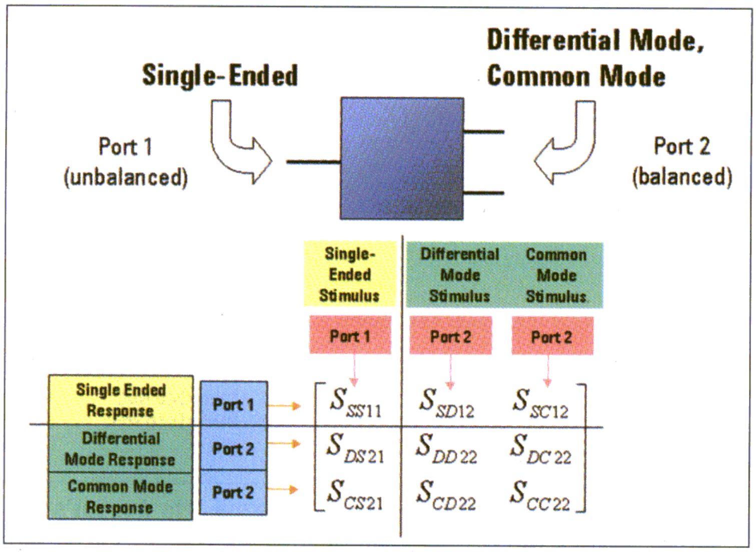


▲ **Figure 3. A mixed-mode *S*-matrix can be organized in a similar way to the single-ended *S*-matrix, in which each column represents a different stimulus condition, and each row represents a different response condition.**

The balun-based technique requires the balun to be placed between the differential port of the device and the single-ended port of the analyzer. The balun transforms the single-ended signal of the analyzer to a differential signal that is applied to the device. There are a number of problems with this approach. The measurement plane should be at the terminals of the device, but this is difficult to realize because there are no standard calibration standards for differential mode. As a result, a calibration must be performed at the single-ended input side of the balun. The analyzer is now measuring the performance of the differential component and balun as one device. The balun has finite return loss, insertion loss, amplitude balance, phase balance, and bandwidth, and becomes a major limiting factor in measuring the component. In addition, the measurement plane should also be at the terminals of the device, which cannot be accomplished with a balun in place. Also a balun will not pass common-mode signals, so none of the mode conversion parameters of the device can be measured. Finally, the balun method will only provide information about the differential mode of the component.

Another method called the mixed-mode S-parameter technique uses a mathematical transform to convert the single-ended data to differential parameters. It provides the common-mode and differential-mode parameters of the device, and is similar to single-ended measurement except that instead of stimulating a single terminal of the DUT, pairs of terminals are considered to be stimulated in either a differential (anti-phase) or a common (in-phase) mode. A physical differential/common mode stimulus to the device is not being provided. A single stimulus signal is actually supplied to each of the ports, and the response is measured. This single-ended data can then be transformed to mixed mode. The mixed-mode S-parameter technique seeks to determine (with a differential mode stimulus on a differential port) the corresponding differential and common mode responses on all of the device ports. For a common-mode stimulus, it attempts to determine the differential-mode and common-mode responses. A mixed-mode S-matrix can be organized in a similar way to the single-ended S-matrix, in which each column represents a different stimulus condition, and each row represents a different response condition (Figure 3). Unlike the single-ended example, the mixed-mode S-matrix not only considers the port but the mode of the signal at each port.

The naming convention for the mixed-mode S-parameters must include mode information as well as port information. Consequently, the first two subscripts in the matrix describe the mode of the response and stimulus respectively, and the next two subscripts describe the ports of the response and stimulus. The mixed-mode matrix fully describes the linear performance of a differential two-port network.



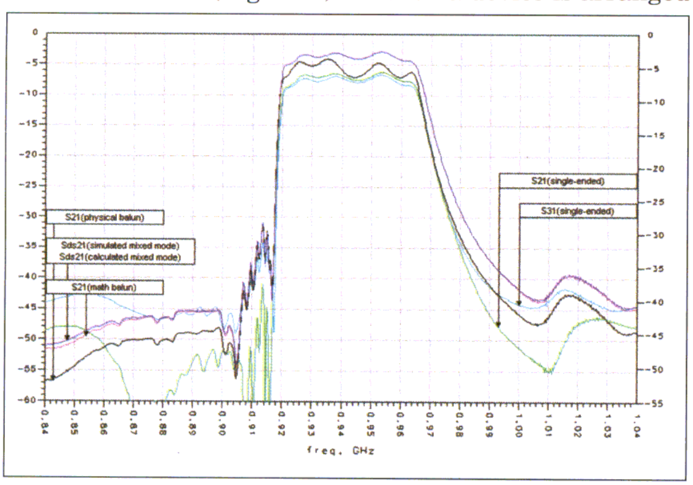
▲ **Figure 4. When the four-terminal matrix is converted to a three-terminal matrix, the differential and common modes on the differential ports and one mode on the single-ended port must be considered. The *S*-matrix for this configuration is arranged with the stimulus conditions in the columns and the response conditions in the rows.**

Differential measurement techniques

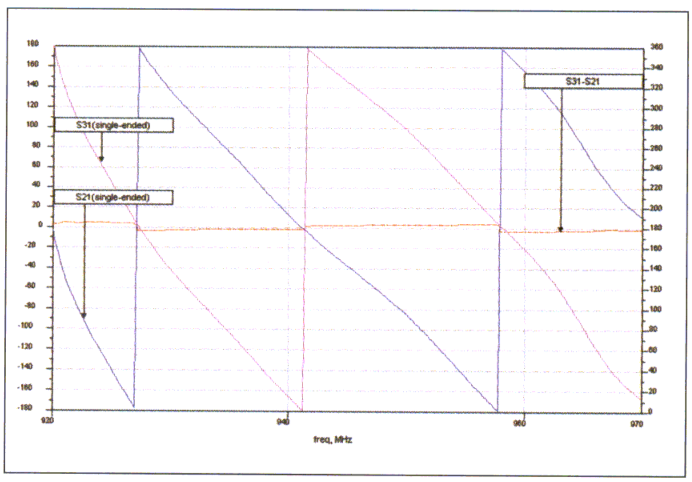
The techniques employed to measure a differential device in the following discussion are based on measurements made of a three-terminal SAW filter in an LTCC antenna switch module using an Agilent PNA Series vector network analyzer. The results are shown in Figures 5 and 6.

A simple extension of the mixed-mode concept can be applied to devices having a combination of differential and single-ended ports, as is the case with the SAW filter. The four-terminal matrix can be converted to a three-terminal matrix by removing the port 1 differential mode stimulus and response, as illustrated by the shaded row and column in Figure 3. In this scenario, the differential and common modes on the differential ports and one mode on the single-ended port must be considered.

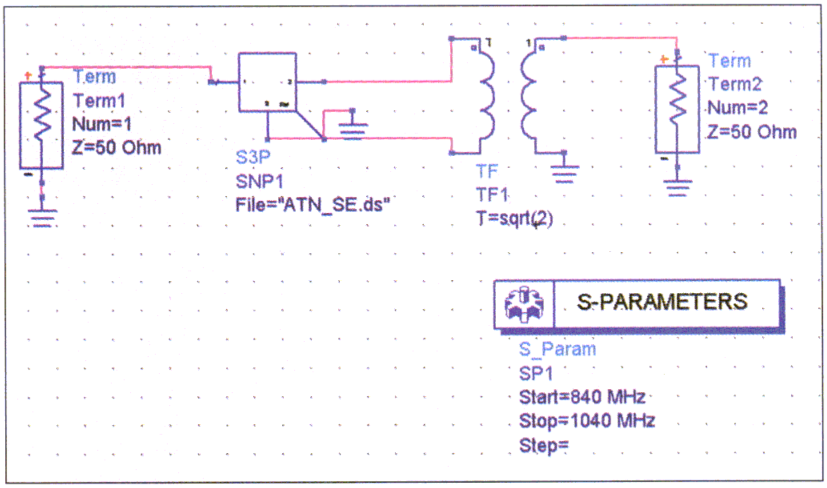
The S-matrix (Figure 4) for such a device is arranged with the stimulus conditions in the columns, and the response conditions in the rows. The mode on the single-ended port is referenced with an 'S' for single-ended instead of a 'C' for common mode because only one mode is available on this port. Two columns and two rows describe each differential port, and one column and one row describe each single-ended port. In this case, the four parameters in the lower right corner describe the four types of reflection that are possible on a differential port. The single parameter in the upper left describes the reflection on the single-ended port, and the other four parameters describe the differential and common



▲ **Figure 5. Offset between the values of S_{31} and S_{21} represents the overall balance between the two terminals that make up the differential port. This image also contains the comparison of the balun and mixed-mode methods.**



▲ **Figure 6. The phase difference between the two terminals (S_{31} and S_{21}) of the differential port measured single-ended as with the single-ended method.**



▲ **Figure 7. The single-ended data is imported into Agilent's Advanced Design System (ADS) simulation tool, where it can be transformed to differential data using a balun circuit component.**

mode transmission characteristics in the forward and revere directions.

Single-ended method

Measuring the differential device as a single-ended multiport device is easy yet time-consuming because multiple two-port measurements are needed to fully characterize the device. In addition, it can produce misleading results because the single-ended data may not give a representative indication of the performance of the device when it operates in one of its differential modes. This occurs because the single-ended data does not provide accurate information of differential performance. For example, S_{21} is the insertion loss measurement from the antenna terminal (terminal 1) to the rx+ terminal (terminal 2). It is not the same as the insertion loss measurement from the antenna to the differential port. In Figure 5, there is offset between the values of S_{31} and S_{21}, which represents the overall balance between the two terminals that make up the differential port. This offset may be caused by an asymmetrical device topology that will result in a decrease in differential mode performance. Ideally, S_{31} and S_{21} should have the same amplitude characteristics.

Delta method

The delta method measures the single-ended transmission phase characteristics of the device. The topology of most differential devices will constrain the electrical length of the two terminals comprising the differential port to give a 180-degree phase shift between them. This parameter is directly related to how well the device performs in differential mode. Figure 6 shows the phase difference between the two terminals (S_{31} and S_{21}) of the differential port measured single-ended as with the single-ended method. There should be 180 degrees of phase difference between the two terminals. The difference in phase shown in the figure is not exactly 180 degrees because of the asymmetries of the device. This method also does not yield insight into the full mixed-mode S-parameter matrix.

Physical balun method

A balun may be used to convert the single-ended port of the network analyzer to the differential port of the device, which transforms the impedance of the differential device to the impedance of the network analyzer. In this case, the differential port impedance is 100 ohms and the single-ended port impedance of the analyzer is 50 ohms. This method will provide some degree of accuracy about the differential characteristics of the device but does not provide information on common-mode performance. The accuracy of this method is also highly dependent on the calibration reference plane and the characteristics of the balun.

Mathematical 'ideal balun' method

The single-ended data may also be imported into a circuit simulator such as Agilent's Advanced Design System (ADS). This data can then be transformed to differential data using a balun circuit component in the simulator (Figure 7). As with the physical balun method, the common-mode performance of the device cannot be measured. The circuit component is an ideal balun, so the common-mode impedance is infinite (where the non-center tapped reflection coefficient equals +1). Any common-mode signals at the output of the device will reflect from the balun and possibly back to the output as an error signal, depending on the mixed-mode performance of the device. The same will be true when using a physical balun, but the reflection coefficient will differ depending on its characteristics. The mixed-mode performance of the device cannot be measured using a balun, so there is no way to determine what the error result may be. The same is true for the center-tapped balun (where the reflection coefficient equals –1).

Simulated mixed-mode method

A circuit simulator may also be used to measure mixed-mode parameters of the differential device (Figure 8). A center-tapped balun is used to perform the differential mode conversion and also provides the mechanism for the common-mode terms. The common-mode conversion occurs at the center tap of the balun where only common-mode signals will appear because of the characteristics of the balun. These common-mode signals are then terminated through a balun into a 25-ohm termination, which is the common-mode impedance of the SAW device. This configuration will allow all

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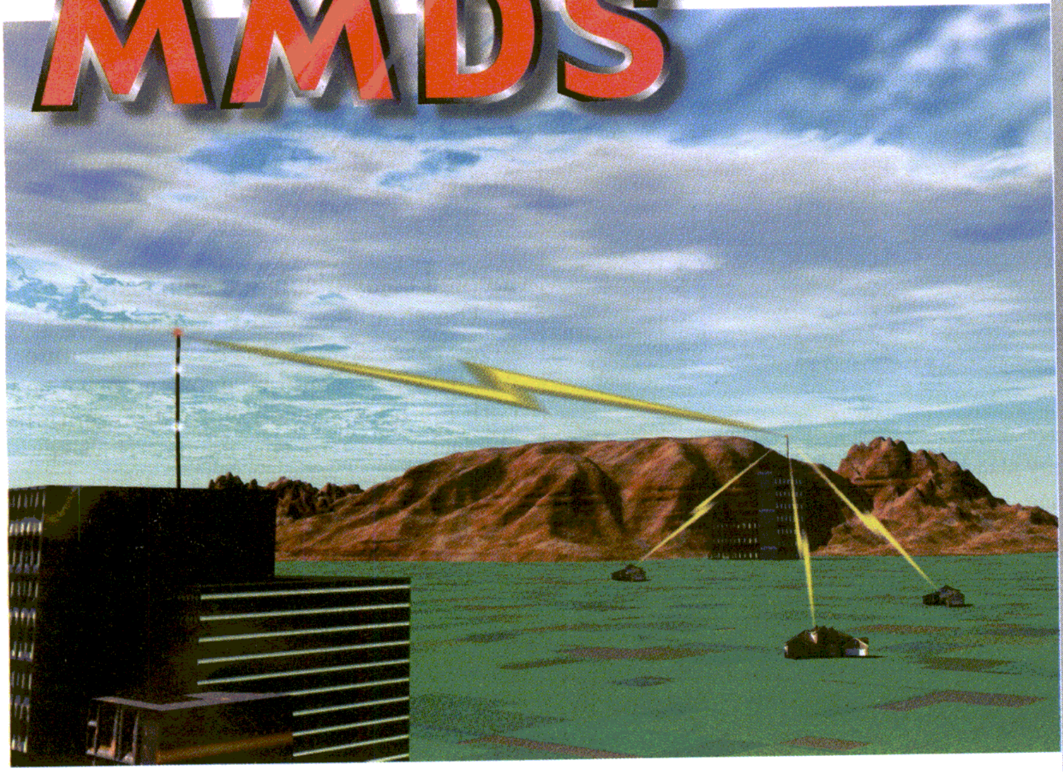

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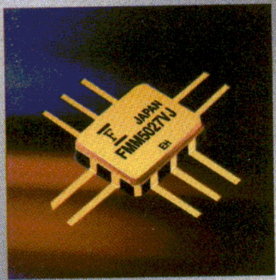

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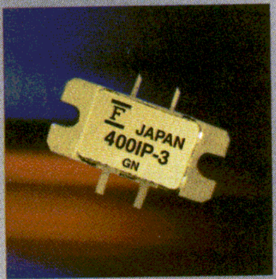
GL=19.0 dB (Typ.)



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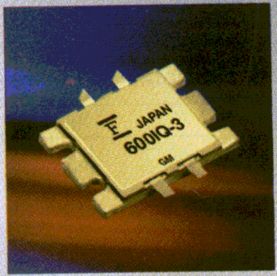

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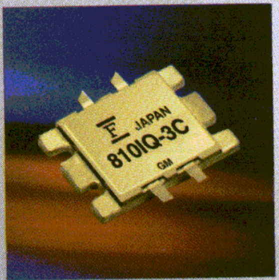

FLL810IQ-3C
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Pout=49.0 dBm (Typ.)
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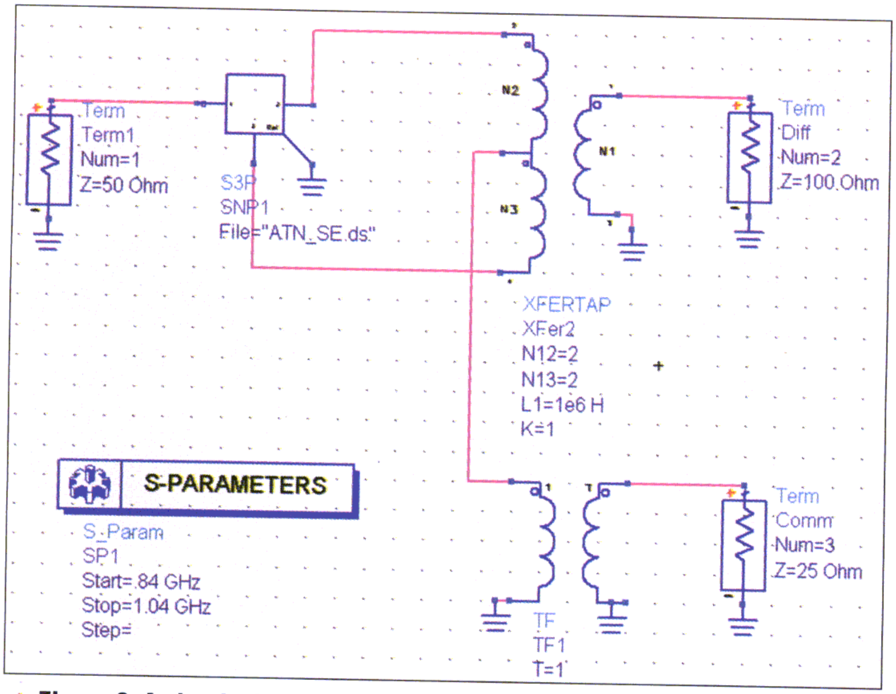


▲ **Figure 8. A circuit simulator may also be used to measure mixed-mode parameters of the differential device.**

the mixed-mode characteristics of the device to be measured. It also provides the appropriate terminations for differential and common-mode signals so that mode conversion terms do not cause errors like those produced by the balun method.

Calculated mixed-mode S-parameters method

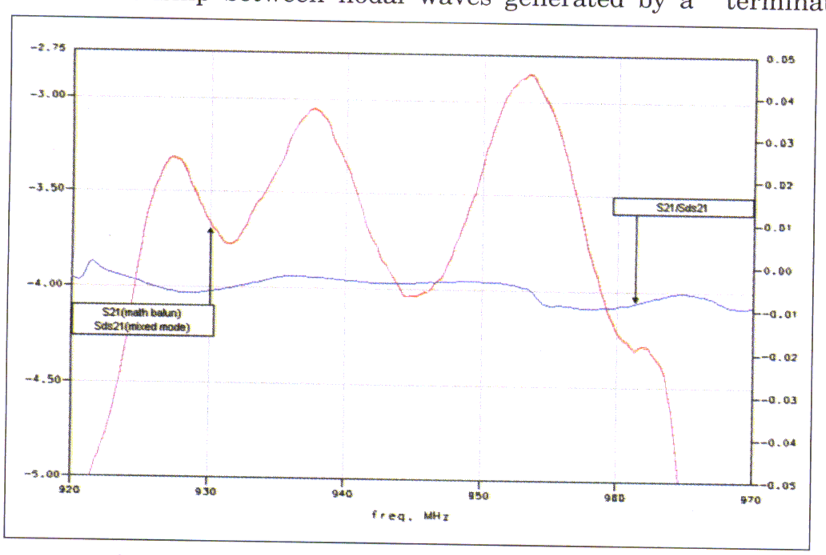
Bocklemann and Eisenstadt [1] have analyzed a method to convert the single-ended data to mixed-mode using mathematical algorithms. These algorithms show the relationship between nodal waves generated by a standard vector network analyzer and the associated common and differential waves that realize mixed mode *S*-parameters. This method is highly beneficial because of the quick and simple method of conversion. It does not require a circuit simulator and therefore can be performed in real time using a compiled math function library. For example, the mixed mode *S*-parameters of a differential device can be accurately measured in a manufacturing environment in which differential measurement speed and accuracy are of high concern.

Method comparison and conclusion

The simulated and calculated mixed-mode measurements have the same result, which would be expected because both methods provide all the differential characteristics of the device. However, the calculated mixed-mode method does not require a circuit simulator to perform the conversion. Figure 9 shows a comparison of the mathematical balun method versus mixed-mode *S*-parameters.

The error caused by mode conversion of the device is clearly seen in this measurement comparison, and is the reason why the balun method result is not accurate. Using calculated mixed-mode *S*-parameters, a differential device can be quickly and accurately characterized in all the modes of operation.

The second part of this article will provide a more rigorous analysis of the error effects of each measurement method, including issues such as improper third-port termination and fixturing. ■



▲ **Figure 9. The error due to the mode conversion of the device can be calculated by comparing the mathematical balun results to the mixed-mode results.**

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1. D. E. Bocklemann and W. R. Eisenstadt, "Combined Differential and Common-Mode Scattering Parameters: Theory and Simulation," *IEEE Transactions on Microwave Theory and Techniques*, Vol. MTT-43, July 1995.

Author information

Loren Betts joined Agilent Technologies in 1997 after graduating from the University of Alberta, Canada, with a bachelor's degree in computer engineering. He now works as an applications specialist with Agilent, focusing on network analyzer products. He is pursuing a master's degree in electrical engineering at Stanford University. He may be contacted via E-mail: loren_betts@agilent.com; or Tel: 707-577-2828.

Measurement of the Anisotropic Dielectric Constant of a Glass Bead in a Coaxial Test Fixture

By Bruce Bullard
Atmel Corporation

Hermetic connectors are used throughout the microwave industry, in applications ranging from extreme humidity and salt fog to vacuum systems. The most widely used seal is made from glass. For frequencies below 1 GHz, most glass is interchangeable. However, for higher frequencies where insertion loss and VSWR are factors, glass types behave differently. Some are easily fired and exhibit a very repeatable dielectric constant, but with high loss. Others have low loss but exhibit variations in the dielectric constant from lot to lot and as a function of firing schedules. This results in VSWR performance variation from lot to lot.

The solution to manufacturing a high performance hermetic connector is to identify a specific glass and then optimize the glass seal process to obtain low loss material and repeatable dielectric constant.

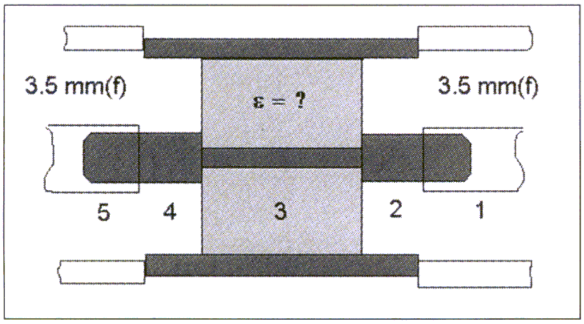
Measuring the dielectric constant poses a problem. The classical method [1, 2] places the bead in a resonant cavity, then uses the quality factor and changes in the resonant frequency to determine the properties of the bead. A drawback of this method is the bead is not measured in the configuration in which it is used. Also, when fired into the connector, the bead forms a compression seal with its surrounding body. This article will show that the compression influences the dielectric properties.

Another disadvantage of this method is that the bead should be fired into a regular shape. Since the beads are purchased in preformed shape (in this case a coaxial configuration), firing the bead into a different shape, one without an inner conductor, is cumbersome and requires undesired machining.

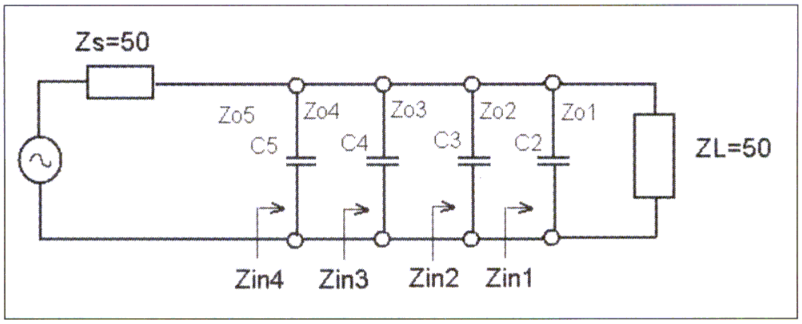
Ideally, the properties of the bead would be measured once it has been fired into the connector body. One method of achieving this is to fire a seal into a connector, perform a TDR measurement of the connector, and analyze the impedance of the connector around the glass seal. Drawbacks to this procedure are:

- The connector must be attached to a matched load for the measurement.
- The connector and matched load must operate to high frequencies (18 GHz minimum).
- The seal is on the order of 0.08 inches long. The resolution of two adjacent discontinuities by TDR is nearly equal to the width of the seal even for broadband measurements. If discontinuities adjacent to the seal are not minimized, they mask the impedance measurement of the glass.
- The TDR measurement is a broadband frequency measurement, which computes the impedance based on the response to all frequencies. Therefore, any variation with frequency is masked.
- The TDR method does not account for anisotropic dielectrics.

This article describes a method where a glass bead is fired into a coaxial test fixture configuration that is similar to the connector. The glass is assumed to be anisotropic following the firing. The parallel (to E field, $\varepsilon_{\parallel}$) relative dielectric constant is computed by comparing the theoretical VSWR to the measured VSWR. The perpendicular relative dielectric constant ($\varepsilon_{\perp}$) is computed by comparing the theoretical phase response to the measured phase response. The theoretical insertion phase is computed using established analytical methods [3, 4] for capaci-



▲ **Figure 1. Glass bead test fixture mated to 3.5 mm (*f*) connectors.**



▲ **Figure 2. Glass bead test fixture transmission line mode.**

tive discontinuities, combined with the theoretical phase response of a coaxial transmission line. Using the test fixture has the advantage that the same firing fixtures, cover gas, inner conductor, similar outer conductor, and heating schedule are used for firing the connector.

The bead is also under the same compression as when it is fired into the connector. These factors all contribute to minimizing a number of uncertainties in the firing process.

The test fixture

The glass bead is fired into the configuration shown in Figure 1. Four junction discontinuities are apparent when the fixture is mated to 3.5 mm (*f*) connectors. The dimensions of the fixture are determined based on dielectric constant estimates/values supplied by the manufacturer, or past experience. The following paragraphs and references [3, 4] provide the necessary analytical tools for design. The design should be optimized for the minimum possible VSWR, which reduces the phase measurement error.

$$
\begin{aligned}
ACBD_{eqv} &= \begin{pmatrix} A_{eqv} & B_{eqv} \\ C_{eqv} & D_{eqv} \end{pmatrix} \\
&= (C_5)(ABCD_4)(C_4)(ABCD_3)(C_3)(ABCD_2)(C_2)(ABCD_1) \\
&= \begin{pmatrix} 1 & 0 \\ j\omega C_5 & 1 \end{pmatrix} \begin{pmatrix} \cos\beta_4 l_4 & jZo_4 \sin\beta_4 l_4 \\ \dfrac{j}{Zo_4}\sin\beta_4 l_4 & \cos\beta_4 l_4 \end{pmatrix} \begin{pmatrix} 1 & 0 \\ j\omega C_4 & 1 \end{pmatrix} \\
&\quad \mathrm{X} \begin{pmatrix} \cos\beta_3 l_3 & jZo_3 \sin\beta_3 l_3 \\ \dfrac{j}{Zo_3}\sin\beta_3 l_3 & \cos\beta_3 l_3 \end{pmatrix} \begin{pmatrix} 1 & 0 \\ j\omega C_3 & 1 \end{pmatrix} \\
&\quad \mathrm{X} \begin{pmatrix} \cos\beta_2 l_2 & jZo_2 \sin\beta_2 l_2 \\ \dfrac{j}{Zo_2}\sin\beta_2 l_2 & \cos\beta_2 l_2 \end{pmatrix} \begin{pmatrix} 1 & 0 \\ j\omega C_2 & 1 \end{pmatrix} \\
&\quad \mathrm{X} \begin{pmatrix} \cos\beta_1 l_1 & jZo_1 \sin\beta_1 l_1 \\ \dfrac{j}{Zo_1}\sin\beta_1 l_1 & \cos\beta_1 l_1 \end{pmatrix}
\end{aligned}
$$

▲ **Equation 1.**

Device parameters

The transmission line model for the device is shown in Figure 2, and the parameters are tabulated in Table 1. The dielectric constant for section 3 is provided by the glass supplier, then refined using the following procedures. The glass dielectric is assumed to be anisotropic [5]. All dimensions are in inches; SI units are used in all calculations.

The ABCD parameters [6] are computed and converted to S parameters. The VSWR is the usual ratio of the reflection coefficient, and the phase response is the argument of the S_{21} term. The ABCD parameters for the lossless transmission line with the configuration described in Table 1 and Figure 2 are:

l_i = length of section

$$\beta_i = 2\pi f \sqrt{\mu_o \varepsilon_o \varepsilon_\perp}$$

$$Zo_i = \frac{59.96}{\varepsilon_\parallel} ln \frac{b_i}{a_i}$$

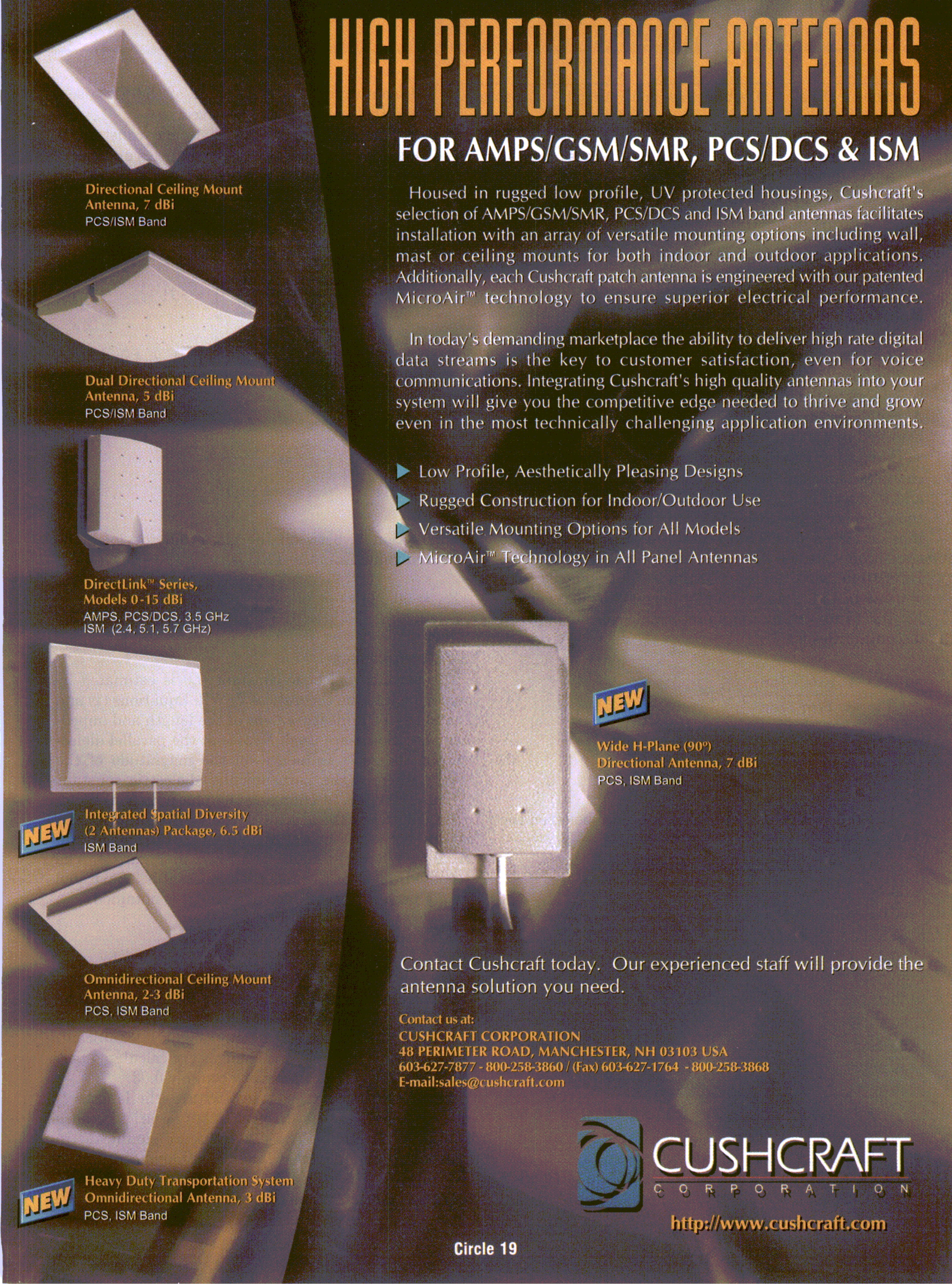
Equation 1 shows the resulting calculations.

The S_{11} and S_{21} parameters of the scattering matrix are:

$$S_{11} = \frac{A_{eqv} + \dfrac{B_{eqv}}{Z_o} - C_{eqv} - D_{eqv}}{A_{eqv} + \dfrac{B_{eqv}}{Z_o} + Z_o C_{eqv} + D_{eqv}} \tag{2}$$

$$\text{vswr} = \frac{1 + |S_{11}|}{1 - |S_{11}|}$$

$$S_{21} = \frac{2}{A_{eqv} + \dfrac{B_{eqv}}{Z_o} + Z_o C_{eqv} + D_{eqv}}, \quad Z_o = 50 \tag{3}$$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Section	Dia a (in)	Dia b (in)	Dielectric	Length (in)	Capacitance (in)
1	0.0598	0.1378	1	0	n/a
2	0.0360	0.1180	1	0.015	9.62
3	0.0235	0.1180	$\varepsilon_{\parallel}$, $\varepsilon_{\perp}$	0.085	1.88 $\varepsilon_{\parallel}$
4	0.0360	0.1180	1	0.015	1.88 $\varepsilon_{\parallel}$
5	0.0598	0.1378	1	0	9.62

▲ **Table 1. Glass bead fixture diameters.**

Measurement results

TDR measurement

The glass bead test fixture (Table 1) was manufactured and tested at room temperature using 7070 glass. A rough estimate of the dielectric constant can be obtained using the TRD method.

Using the HP8510 network analyzer's TDR mode, the characteristic impedance of the fixture was measured and is shown in Figure 3. The first high corresponds to the mismatch caused by the glass and the discontinuities on either side. The second high is a result of the transition between the glass seal and the load. The HP8510 network analyzer settings are: frequency span = 0.045 to 18.0 GHz, window = normal, low-pass mode. For this setup, the minimum resolvable spacing between discontinuities is:

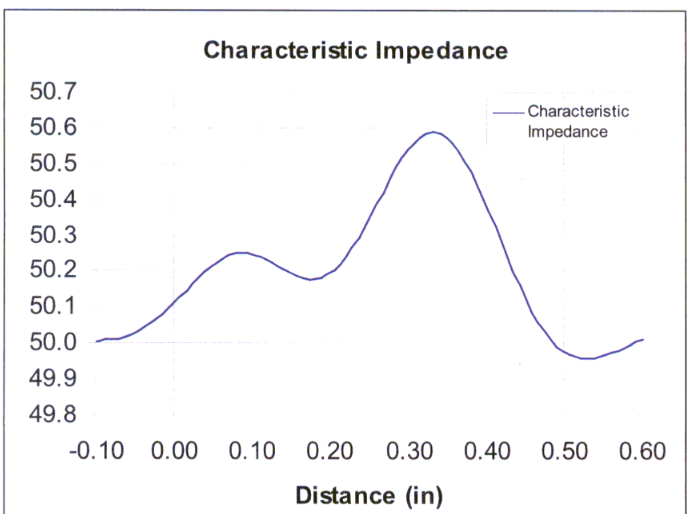
$$t_{\min} = \frac{1}{2} \cdot \frac{(0.60) \cdot (1.6)}{\text{Freq Span}} = 27 \text{ pS} \qquad (4)$$

$$l_{\min} = \frac{t}{2\sqrt{\mu_o \varepsilon_o \varepsilon_r}} \cdot \frac{100}{2.54} \approx 0.08 \text{ in} \qquad (5)$$

Note that with this setup, the difference between the discontinuities on either side of the glass seal and the seal cannot be resolved.

The estimate of the dielectric constant is made by ignoring the discontinuities and calculating the isotropic dielectric constant of the glass based on the impedance measured in the area, as well as the dimensions in that area. For section 3,

$$\varepsilon_r \approx \left[\frac{59.96}{50.25} \cdot ln\left(\frac{0.118}{0.0235} \right) \right]^2 = 3.71 \qquad (6)$$



▲ **Figure 3. TDR for the glass bead text fixture of Figure 2 and Table 1.**

Note that the impedance in sections 1, 2, 4 and 5 can skew impedance measurement. These values are:

$$Z_{o2} = Z_{o4} = 59.96\, ln \frac{b_2}{a_2} = 71.2\,\Omega \qquad Z_{o5} = 59.96\, ln \frac{b_5}{a_5} = 50\,\Omega \qquad (7)$$

For section 1, between the glass and the load, the measured value from the TRD is:

$Z_{o1} = 50.6\ \Omega$
$l_1 = 0.5$ in

VSWR performance

The parallel dielectric constant ($\varepsilon_{\parallel}$) is estimated by comparing it to the theoretical VSWR. Equations (1) and (2) and the TDR measurement of the length and impedance of section 1 are used. Note that the parallel dielectric has the most influence on the magnitude of the VSWR, whereas the perpendicular dielectric constant ($\varepsilon_{\perp}$) has the effect of changing the phase response and, to a small extent, the spacing of the VSWR nulls. A comparison between theoretical VSWR performance and actual VSWR performance is shown in Figure 4. The estimated value for the dielectric constant is 3.5 parallel and 3.5 perpendicular. Good agreement is obtained.

Phase response

The perpendicular dielectric is determined by comparing the measured phase response to the theoretical value calculated with Equations (1) and (3). The parallel dielectric is as calculated above, $\varepsilon_{\parallel} = 3.5$. The influence of the parallel dielectric is only slightly significant. Using an isotropic dielectric changes the calculated perpendicular dielectric constant by roughly 0.03.

When manufacturing the test fixture, the lengths l_2 and l_4 are relatively easy to control, but l_3 varies because of process variations. These lengths are significant in phase calculation. Lengths l_2 and l_4 are held constant at 0.015 inches. The measured phase and calculated perpendicular dielectric constant are listed in Table 2.

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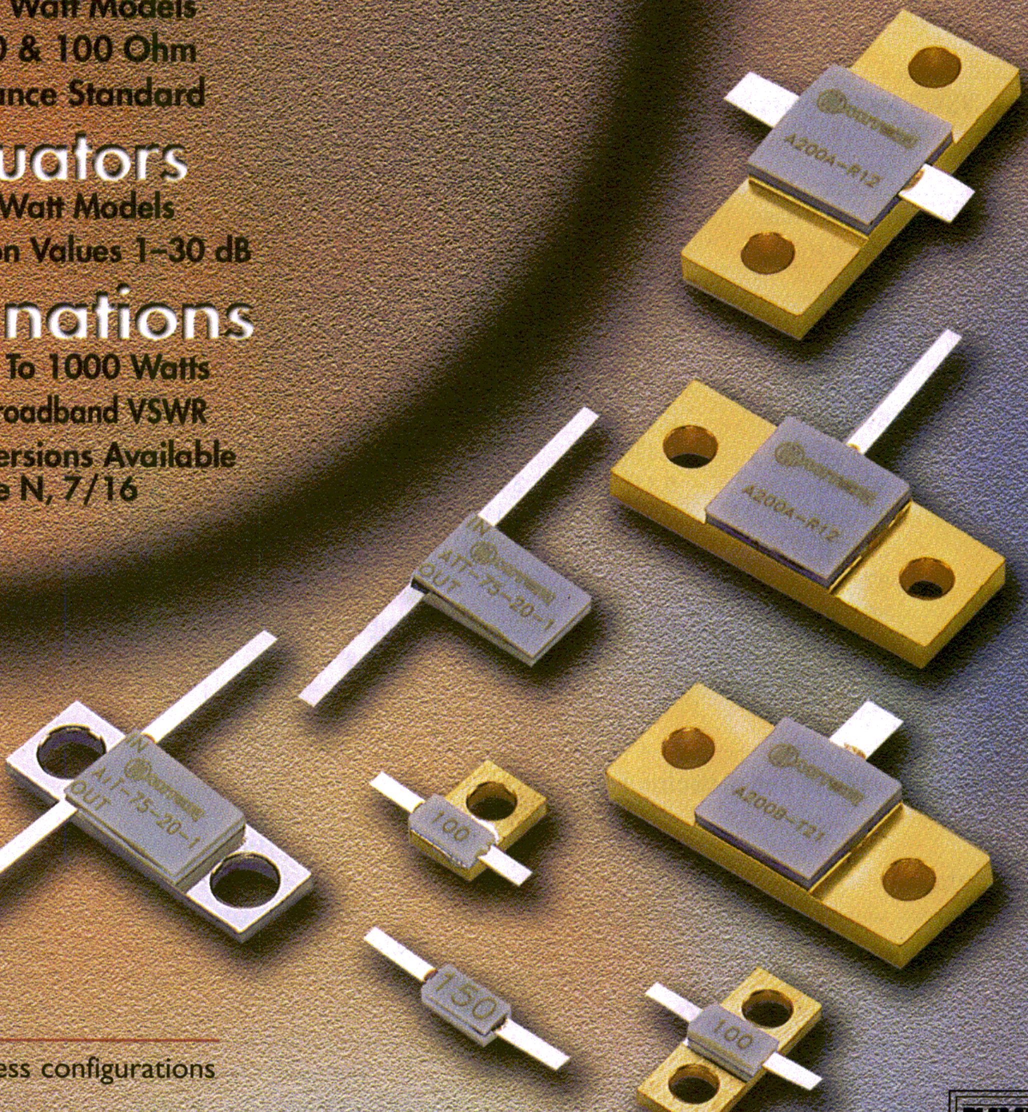

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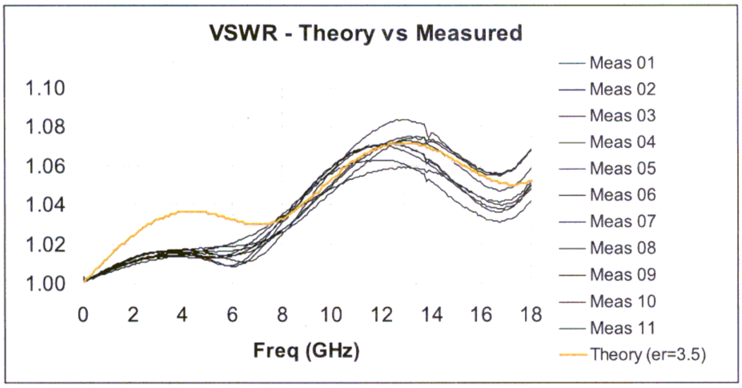

▲ **Figure 4. VSWR for the glass bead fixtures from Table 1.**

Sample	Measurement Phase (degrees)	Length (in) $l_2 + l_3 + l_4$	$\varepsilon_\perp$
1	–84.985	0.1135	4.02
2	–83.767	0.1110	4.12
3	–85.591	0.1125	4.19
4	–86.670	0.1140	4.16
5	–84.346	0.1110	4.19
6	–86.626	0.1150	4.07
7	–85.850	0.1150	3.98
8	–85.000	0.1120	4.17
9	–86.640	0.1150	4.07
10	–87.163	0.1160	4.03
11	–85.368	0.1130	4.11
Average = 4.10			
STD Dev = 0.07			
Δf = 13.343 GHz			

▲ **Table 2. Measured phase and dielectric constant.**

Published data

Several sources were found for the dielectric constant of 7070 glass. The published values are for an uncompressed dielectric or, in this case, the perpendicular dielectric constant ($\varepsilon_\perp$). Good agreement is obtained. The results are shown in Table 3.

Measurement errors

Calculation of the perpendicular dielectric constant based on phase response is subject to several sources of error. First, the fixture must be assembled as shown in Figure 1. The steps on either side of the glass seal must be relatively flush to the glass because capacitance of the junction is based on the dielectric constant of the glass, as well as the proximity of the step to the glass. The second error lies in the physical measurement of the fixture's dimensions. These can be reduced using standard measurement techniques for small dimensions, testing a large number of devices and taking the average dielectric constant. A third consideration is the accuracy of the capacitance estimate and its variance with frequency. The capacitance is estimated to within 3 percent, and the variance with frequency should not exceed 2 percent. Another error term is estimating the phase change per length of line from Equation 1. Specifically, the loss terms were ignored. The model is estimated to be within 0.5 percent.

The network analyzer measurement error is easily reduced. It has a worst-case phase measurement error of roughly ± 2.9 degrees [8]. Plugging this into Equations (1) and (2), the error in the dielectric measurement is roughly ± 0.3. (Keep in mind this error is the worst case, and typically the accuracy will be much greater.) A rough measurement of the network analyzer phase error has been made [9] and it was shown that most measurements are within ± 0.4 degrees. The reduction in the variance is accomplished by taking several phase measurements over varying frequencies.

VSWR can also generate phase errors. For the VSWR measured, 1.10:1 the phase error is estimated as [10]:

$$\phi_{err} = \pm \sin^{-1}|\rho_1 \rho_2| = 0.033° \qquad (8)$$

where

$$\rho_1 = 0.048$$
$$\rho_2 = 0.012$$

Ignoring the dimensional discrepancies, combining the capacitance errors, considering the errors in the model for phase/length and using a phase measurement error of ±0.4 degrees, the error in the dielectric measurement will be approximately ±0.08.

Because test fixture is similar, the errors from the capacitance estimates should be repeatable, as are the errors from the phase/length equations. To a lesser extent, because of the similarity of devices, one can argue that the errors in the phase measurement are also somewhat repeatable. These factors combine to yield an estimate of the dielectric constant that may have a mean error of ±0.08, but that has a low error variance.

Source	Frequency Range	Relative Dielectric Constant (room temperature)
Corning, Publish Spec	1 MHz	4.1
GBC Materials (supplier)	1 MHz	4.1
A. von Hipple	100 Hz to 10 GHz	4.0
A. von Hipple	25 GHz	3.9

▲ **Table 3. Published dielectric constant of 7070 glass [7].**

Verification using electromagnetic finite integration

A finite integration simulation/modeling program was made available by

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	Δf GHz	Dielectric Constant	Phase Response
CST Microwave Studio (phase is calculated from dielectric)	13.99	3.550	86.82°
ABCD Theory (dielectric is calculated from phase)	13.99	3.627	86.82°

▲ **Table 4. Comparison to finite integration analysis.**

CST America and used to confirm the accuracy of the theory presented. The device was used as defined in Table 1. The CST model was run with a isotropic dielectric of 3.55 instead of the measured $\varepsilon_r = 3.5_{\|}, 4.10_{\perp}$. For comparison, the ABCD model also uses an isotropic dielectric. The results are shown in Table 4.

Frequency dependent dielectrics

Frequency dependent dielectrics can be measured as described above. The goal is to narrow the frequency span to encompass the frequency of interest. Several measurements are made over varying frequency bands that surround the frequency of interest. Using this method, the phase measurement error can be reduced, thereby reducing the error in the measured dielectric constant.

Complex dielectric constant

The basic Equation (1) still applies. Beta becomes the complex propagation constant gamma, which takes into account conductor loss and dielectric loss. The cosine terms become hyperbolic cosines, the sine terms, hyperbolic sines and the j is dropped. Two good sources for proceeding are [11, 12].

Conclusion

The illustrated measurements show that the dielectric constant of 7070 glass is dependent on the compression the glass is under. In the direction parallel to the electric field, the relative dielectric constant ($\varepsilon_{\|}$) is reduced because of increased pressure, while the relative perpendicular component ($\varepsilon_{\perp}$), which is not under compression, maintains its typical value.

The key to designing low VSWR hermetic seals is to quantify the changes in the dielectric constant as a result of the compression seal. This method gives connector manufacturers tools for monitoring the firing process and adjusting when necessary. The tools and techniques required are basic and available in most microwave manufacturing facilities.

In addition, this method allows the designer the option of testing over frequency without setting up numerous resonate cavities. Although not as accurate as the resonate cavity, the method accounts for compression and anisotropic dielectrics and can be easily implemented. ■

Acknowledgements

The work detailed in this article was funded by Kaman Instrumentation. CST America provided a valuable analysis that helped greatly in confirming the theory.

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Author information

Bruce Bullard is a consulting engineer in the Silicon Germanium Circuit Evaluation Laboratory at Atmel Corporation. He is also a member of the Atmel Application and Integration Team for the W-CDMA Harmony Lite Chip. He has a master of science degree in electrical engineering from the University of Colorado at Colorado Springs. He may be reached via E-mail: bruce_b37@yahoo.com; or Tel: 719-540-1517.

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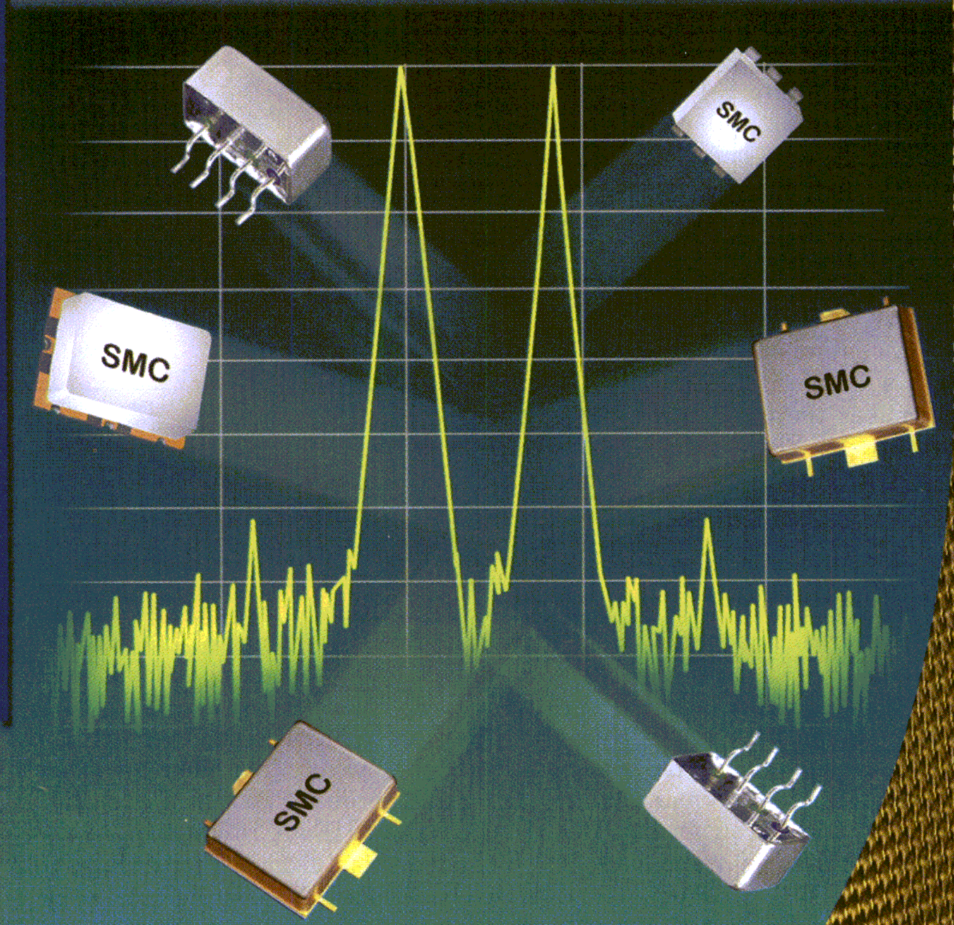

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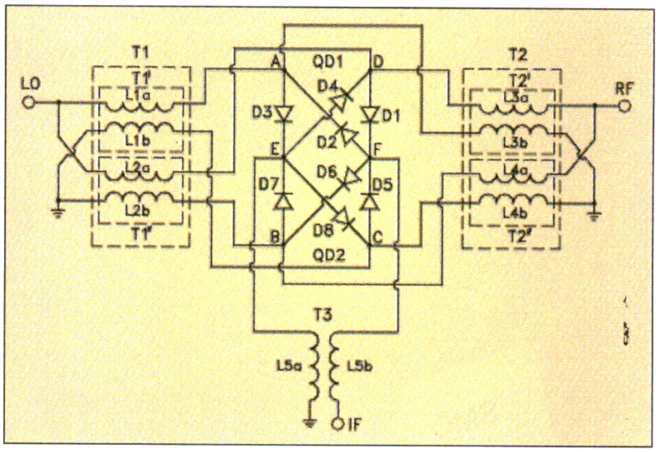
Circle 15

Mini-Circuits Introduces a New Low-Cost, Triple-Balanced LTCC Mixer

Double-balanced mixers are widely used in frequency translation applications. Some of the advantages of the double-balanced mixer are good L-R and L-I isolation and moderate R-I isolation. Triple-balanced mixers, also called double-doubly balanced mixers, have three separate baluns for LO, RF and IF ports. This structure helps provide wide bandwidth, good L-R, L-I and R-I isolations, VSWR, compression and IP3, all at the expense of 3 dB higher RF power. The only disadvantages are the number of components, resulting in a large size, and the complexity, which makes the design more expensive.



▲ **Figure 1. Schematic of the mixer.**

Mini-Circuits has designed a new low-temperature co-fired ceramic (LTCC) based mixer (patent pending) that is substantially reduced in size and complexity. It uses a combination of LTCC and ferrite based baluns and Schottky diode quads to realize a small, triple-balanced mixer.

Operating principle of the mixer

Figure 1 shows the schematic of the mixer. The mixer consists of an LO transformer T1, an RF transformer T2 and an IF transformer T3. Transformer T1 is comprised of transmission line transformers T1' and T1''. Similarly, T2 is comprised of T2' and T2''.

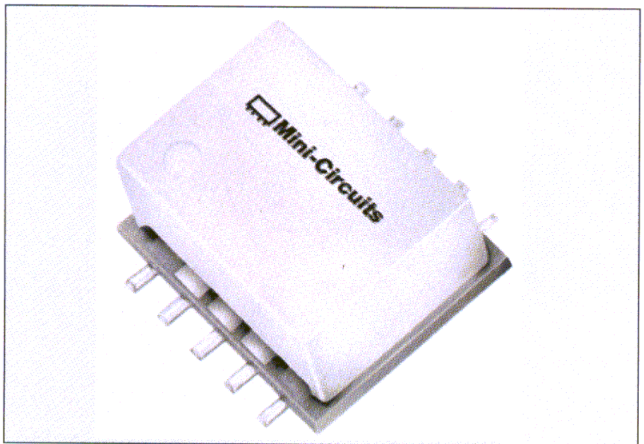
Transformer T3 is also a transmission line transformer. Quads QD1 and QD2 are in monolithic form to assure superior match.

During the positive half cycle of LO, diodes D1, D3, D6 and D8 are on. The resulting polarity of the RF signal at the IF port is of a particular polarity. During the negative half cycle of LO, diodes D2, D4, D5 and D7 are on the resulting polarity of the RF signal at the IF port is now reversed. This is equivalent to mathematical multiplication or mixing. The result is a sum and difference frequency of LO and RF signal. Usually, the difference frequency is used as the desired IF.

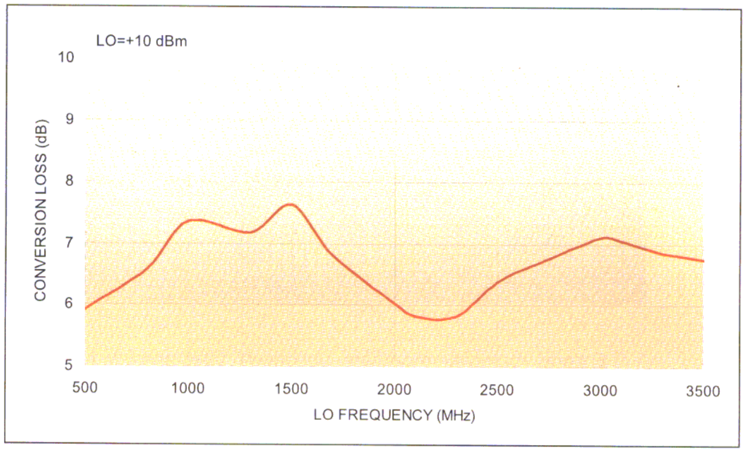
Interport isolation is achieved due to the balance of the circuit. In Figure 1, node E and F are at ground potential for LO signal. Hence, LO and IF are isolated.

The LO signal from T1 appearing at T2' terminal (so also at T2") is of the same polarity and magnitude. Hence, LO and RF ports are isolated from each other.

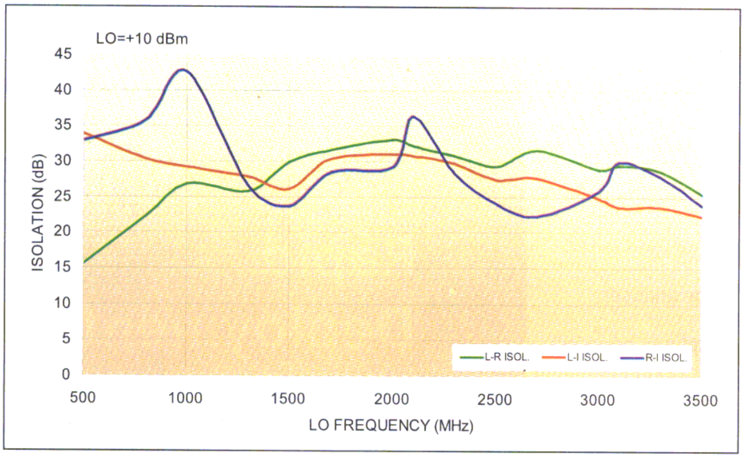
The RF signal appearing at the IF port is switched every half cycle. Using Fourier analysis, it can be shown that the RF signal component is not present at IF. This interport isolation is ideal for the usability of the mixer, because it eliminates or reduces the need for external filtering.



▲ **Figure 2. Photograph of the LTCC mixer.**



▲ **Figure 3. Conversion loss of the MCA-35LH mixer.**



▲ **Figure 4. L-I, L-R and R-I isolation of the MCA-35LH mixer.**

Construction of the LTCC mixer

For frequencies below 5 GHz, ferrite-based transformers are widely used as baluns. The result is a manual-oriented manufacturing process, which becomes a major handicap for large-scale manufacturing. In this new approach, low temperature co-fired ceramic (LTCC) has been used as the medium to realize two of the baluns to make it a compact mixer. Quads are used in die form to minimize size and cost.

The advantages of the new approach are:

- Small size (0.25 × 0.3 × 0.2 inches)
- Reduced manufacturing steps
- High repeatability due to integrated components
- Ease of mass production
- Low cost

Model number	Frequency (MHz)		Conversion Loss (dB)			L-R Isolation (db)		L-I Isolation@ center (db)		IP3 (dBm) Factor band	E Style*	Case	Price $ Qty. (10-49)
	LO/RF $f_L - f_U$	IF	Typ.	σ	Max.								
MCA-35LH	500-800	10-300	6.3	0.1	7.7	18	11	32	25	16	0.6	DM842	6.95
	800-1000	10-200	7.1	0.1	8.9	24	17	30	23	18	0.8		
	1000-1800	10-800	7.3	0.1	8.9	29	20	28	20	16	0.6		
	1800-2000	10-200	6.3	0.1	8.2	32	25	30	23	16	0.6		
	1800-2500	10-700	5.8	0.1	8.2	32	22	30	21	16	0.6		
	2000-3500	10-1500	6.5	0.1	8.7	29	17	26	16	17	0.6		
	500-3500	10-1500	6.9	0.1	8.9	•		•					
MCA-50LH	1000-1400	10-400	6.8	0.1	8.5	20	11	32	25	17	0.7	DM842	7.95
	1400-2000	10-600	6.6	0.1	7.7	25	20	32	25	17	0.7		
	2000-2600	10-600	7.8	0.2	9.9	25	18	30	24	18	0.8		
	2600-4500	10-1500	7.8	0.1	8.6	35	20	25	15	15	0.5		
	4500-5000	50-500	8.0	0.2	9.9	35	22	25	15	16	0.6		
	1000-5000	10-1500	7.3	0.2	9.9	•		•					

• See individual band specs.
E-Factor = [IP3(dbm) – LO Power (dbm)]/10
* Leadless version available in case style DN844

▲ **Table 1. Electrical specifications (LO = +10 dBm).**

Figure 2 shows the photograph of the new mixer.

Performance of the mixer

Figure 3 shows the conversion loss of the mixer MCA-35LH. The conversion is around 6.5 dB over 1800 to 2500 MHz. This mixer has a typical L-R, L-I and R-I isolation of 30 dB as shown in Figure 4. It also has excellent matching on all three ports with the following return losses:

LO port: 12 dB typical
RF Port: 10 dB typical
IF port: 20 dB to 500 MHz and 10 dB to 1000 MHz

Figure 5 shows the return loss of LO, RF and ports; Figure 6 shows that of IF port. Figure 7 shows the IP3 performance. The third order intercept point of this mixer is 16 dBm typically over its entire range. The 1 dB conversion point is typically at +16 dBm RF input. This mixer has useable performance over 500 to 3500 MHz.

A second mixer with optimized performance over 3000 to 4500 MHz with an IF BW of 10 to 1300 MHz and a useable range of 1000 to 5000 MHz has also been designed and designated model number MCA-50LH. The above mixers need 10 dBm LO power to operate.

Conclusion

A series of triple-balanced mixers have been designed using LTCC. These mixers are small in size, have repeatable performance, need less manufacturing steps and are less expensive. They are ideal for low, medium and high volume markets because they are easy to reproduce compared to ferrite based mixers. ■

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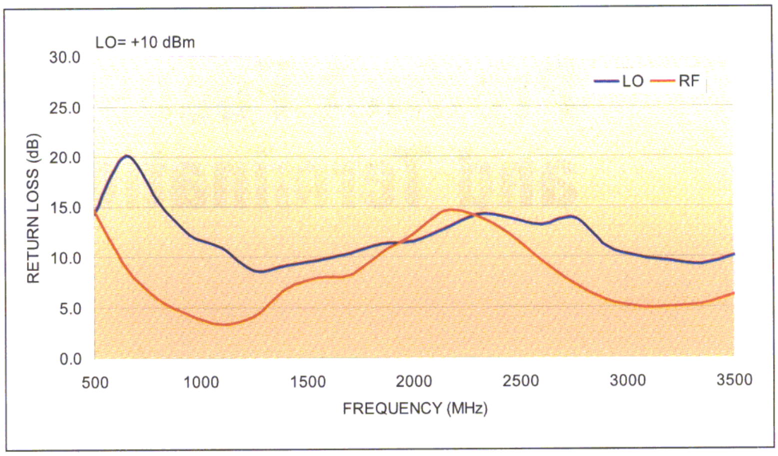
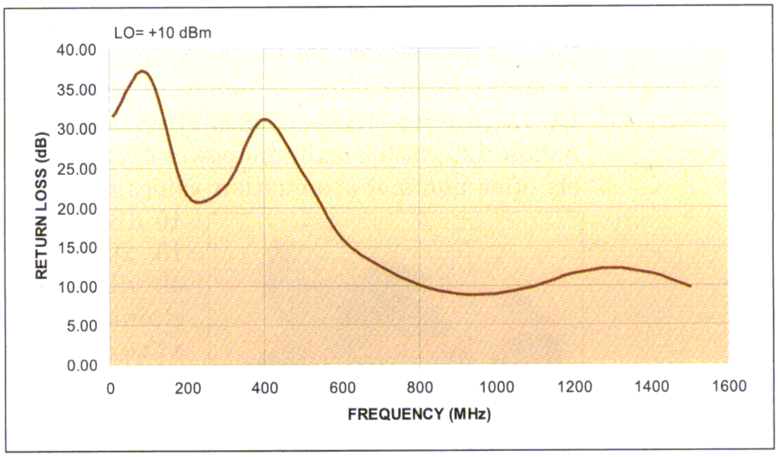
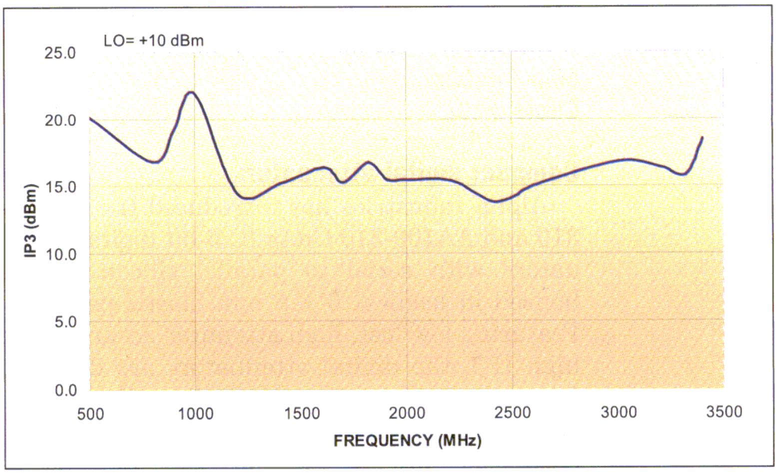


▲ **Figure 5. LO and RF return loss of the MCA-35LH.**



▲ **Figure 6. IF return loss of the MCA-35LH.**

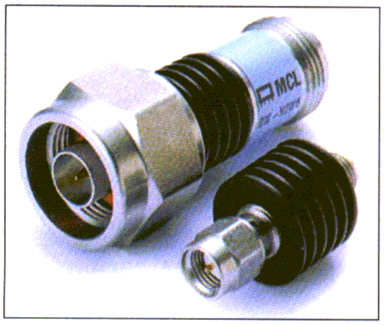


▲ **Figure 7. IP3 of the MCA-35LH.**

Product Focus — Attenuators and Terminations

Coaxial attenuators for DC to 18 GHz

Mini-Circuits offers a family of broadband DC to 18 GHz precision fixed attenuators that handle 125 watts maximum power. Fifteen models offer nominal attenuation values from 1 to 10 dB plus 12, 15, 20, 30 and 40 dB. These stainless steel attenuators exhibit excellent temperature stability, high repeatability and typical VSWR of 1.15:1. Applications include impedance matching and providing a 2 watt or 5 watt termination load for power amplifiers.



Mini-Circuits
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Compact digital attenuators

Alpha Industries has introduced the AA107-310 and AA109-310 GaAs IC 5-bit digital attenuators with serial to parallel silicon drivers housed in leadless 5 × 5 mm plastic packages. Featuring low loss, high attenutor accuracy and high IP3, the digital attenuators are designed for use in base stations, as well as for wireless data and wireless local loop applications.

Alpha Industries, Inc.
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Kit includes four attenuators

Bird Component Products has released a new attenuator kit, the 3-A-MFB-K1. The kit includes four attenuators, each rated at 3, 6, 10 and 20 dB values and with BNC male and female connectors. This cost-effective kit is designed for use by installers, technicians and engineers to determine exact dB values needed for transmit, receive or measurement systems, as an alternative to more costly variable attenuators. The kit is provided in a foam-lined, impact-resistant case.

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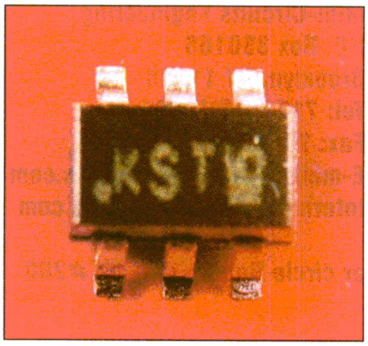
4-line array Schottky bus terminator

Diodes Incorporated has launched a new quad data line Schottky bus terminator. The QSBT40 is a high-density array designed to protect the sensitive in/out ports of Transistor-Transistor Logic (TTL) and Complementary Metal-Oxide Semiconductor (CMOS) integrated circuits from over-voltages caused by induced electrical transients. This SOT363 device is based on Diodes' application-specific multi-chip circuit high-density implementation. Four data line terminators are mounted in one surface-mount package, with



eight diodes arranged in four series-connected pairs. The packaging offers as much as a 10:1 board space reduction ratio as compared to other discrete Schottky diodes. Applications include LAN/WAN equipment, set-top boxes, Ethernet connections and T1 line cards.
Diodes, Inc.
Circle #196

Aluminum nitride attenuators

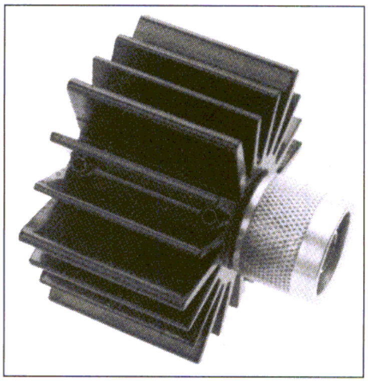
Anaren Microwave has introduced two RF power-brand aluminum nitride (AlN) attenutors, rated at 20 and 30 dB. The attenuators, designated RFP-100NXXAF and RFP-100NXXAE, handle 100 watts (average CW) and are designed for the DC to 2.5 GHz range. Specifications include thick film resistive elements, 99.99 percent pure silver leads and operating temperature range from –55 to +150 degrees Celsius. The RFP-100NXXAE also offers a copper nickel plated mounting flange per QQ-N-290 requirements. The units meet or exceed applicable portions of MIL-E-5400 and measure 0.350 × 0.225 inches.
Anaren Microwave Inc.
Circle #197

Low-cost coaxial terminations

Elcom Systems announces the availablity of a series of miniature 50-ohm coaxial terminations suitable for military and commercial equipment. The terminations are provided in BNC, TNC, N, SMA or SMB configuration, with connectors in male or female (SMB female only). Specifications include VSWR of 1.1:1 nominal and 1.25:1 maximum over a frequency range of DC to 2 GHz for BNC), 2.5 GHz for TNC and N, 4.2 GHz for SMA and 3 GHz for SMB. Disspation is 0.5 watts CW, 1 kW peak over the temperature range of –25 to +85 degrees Celsius. The terminations are designed with gold, silver or nickel plated connectors, with the resistor element mounted in a silver plated housing.
Elcom Systems, Inc.
Circle #198

10-watt terminations up to 10 GHz

Bird Component Products has increased the frequency range of its 10-T series 10-watt terminations. The series is now rated up to 10 GHz. Standard operating temperature is –40 to +40 degrees Celsius. VSWR is 1.1:1 maxiumum from DC to 2.4 GHz, 1.15:1 maximum from 2.4 to 6 GHz and 1.25:1 maximum from 6 to 10 GHz. Available connectors are N male and female. BNC and TNC connectors are also offered for applications with a maximum frequency range of 4 GHz or below.



Bird Component Products Inc.
Circle #199

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Very Low Power RF Transceiver Targets Demanding Applications

By Peder Martin Evjen
Chipcon AS

The new RF transceiver CC1000 from Chipcon is specifically designed to comply with stringent requirements for short-range RF applications. The CC1000 is a true single-chip UHF transceiver developed for very low power consumption and low voltage operation. Based on a 0.35 mm CMOS technology this is the only product currently available that offers a unique combination of low cost, high integration, high performance and flexibility, thus setting a new standard for short-range wireless communication devices.

Parameter	CC400	CC900	CC1000
Frequency range (MHz)	300-500	500-1000	300-1000
Range	Best	Best	Good
Power consumption	Good	Good	Best
Battery applications	Good	Good	Best
Supply voltage range	2.7-3.3V	2.7-3.3V	2.3-3.6V
Power down mode	Yes	Yes	Yes
Multi-channel systems	Yes	Yes	Yes
Frequency hopping	Yes	Yes	Yes
Frequency fine tuning	Yes	Yes	Yes
RSSI	No	No	Yes
Integrated bit synchronizer	No	No	Yes
Integration level	Good	Good	Best
Package	SSOP-28	SSOP-28	TSSOP-28

▲ **Table 1. Comparison of transceiver features.**

The CC1000 is designed primarily for FSK systems in the ISM/SRD bands at 315, 433, 868 and 915 MHz. However, it can easily be programmed for operation at other frequencies between 300 and 1000 MHz. The low power consumption and the high integration level make the CC1000 ideal for remote keyless entry, home automation, automatic meter reading and wireless alarm and security systems.

New technology, new applications

The introduction of CC1000 adds state-of-the-art technology to Chipcon's high performance CC400 and CC900 RF transceiver intergrated circuits. These two products target long-range and high sensitivity applications, features essential for applications such as automatic meter reading, home automation, point-of-sales systems, barcode scanners and industrial remote controls.

However, the ever-increasing demands for reduced size and low power consumption in battery applications led Chipcon to develop the very low power CC1000 transceiver. Table 1 summarizes and compares the features of these three chips.

The CC1000 operates from 2.3 to 3.6 V, making it ideal for battery applications using a single lithium manganese dioxide 3.0 V coin cell or a lithium thionyl chloride 3.6 V cell. The current consumption is 7.7 mA in receive mode and from 5.3 mA up to 28 mA in transmit mode, depending on the output power. Output power can be programmed in 1 dB steps up to 10 dBm at 433 MHz. In power-down mode, the device draws 0.2 mA, another important feature for battery applications. Remote keyless entry and tire pressure monitoring are two of the applications in the automotive industry that take advantage of the CC1000 benefits.

CC1000 comes in a TSSOP28 package, specified over the temperature range from –40 to +85 degrees Celsius. Table 2 summarizes the CC1000's technical specifications.

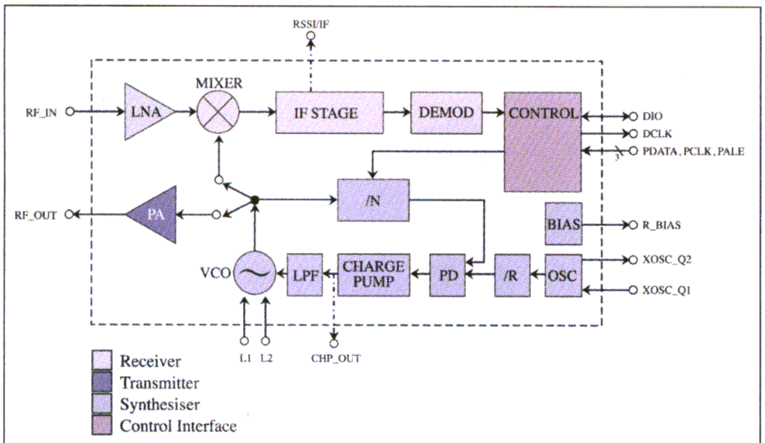
Specifications		Min	Typ (433/868 MHz)	Max	Unit
General:	RF Frequency Range	300		1000	MHz
	Data Rate	0.6		19.2	kBaud
TX Mode:	Output Power (programmable)	–20		10/5	dBm
	FSK Separation(programmable)	1		65	kHz
RX Mode:	Receiver Sensitivity, 1.2 kb/s		–107/–106		dBm
Power Supply:	Supply Voltage	2.3		3.6	V
	Current Consumption, RX		7,7/12		mA
	Current Consumption, TX, –20 dBm		5,3/8,6		mA
	Current Consumption, TX, –5 dBm		8/13,9		mA
	Current Consumption, TX, 0 dBm		11,6/16,4		mA
	Current Consumption, TX, 5 dBm		14,6/12,2		mA
	Current Consumption, TX, 10 dBm		27,8/-		mA
	Current Consumption, power down		0.2	1	mA

▲ **Table 2. Specifications of CC1000 at 433 and 868 MHz.**

A sensitivity of –109 dBm combined with high output power (10 dBm) makes it possible to build very low power battery operated systems offering a long and reliable communication range. The CC1000 has an integrated PLL that can be programmed in 250-Hz steps, making crystal temperature drift compensation possible as well as providing the options of multi-channel systems and frequency hopping protocols.

The CC1000's multi-channel possibilities are an advantage when designing equipment for operation under different regulatory restrictions and when developing solutions that can be used in different markets. Automatic frequency selection and frequency hopping protocols may be used to increase the reliability of the system as the frequency bands used for license-free short-range communications become more crowded.

The device offers an integrated bit synchronizer and RSSI and is easily interfaced to any micro-controller through 5 I/O lines.



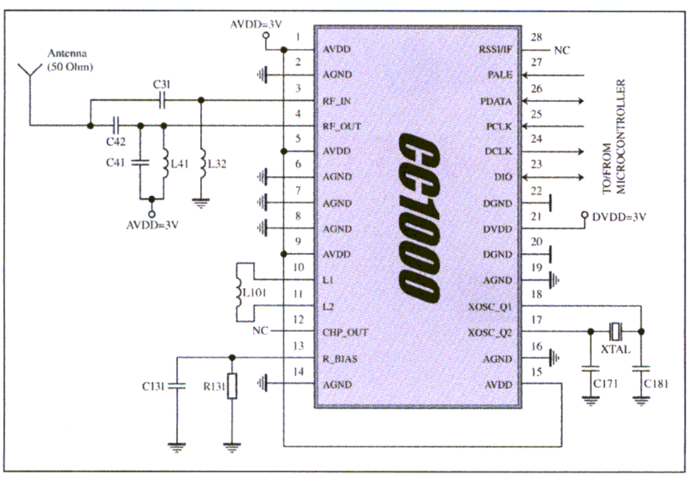
▲ **Figure 1. Block diagram of the CC1000.**

Application circuit

Few external components are required to make a complete transceiver, so the CC1000 is a good choice for lowest cost and smallest size applications. The application circuit is shown in Figure 2.

A highly integrated transceiver

The block diagram of the CC1000 (Figure 1) shows that all blocks needed to make a complete RF transceiver are integrated into the chip. Even with the high degree of integration, the CC1000 is very flexible.



▲ **Figure 2. Application circuit of the CC1000.**

Summary

The CC1000 transceiver by Chipcon is designed for demanding wireless automotive and consumer applications. It is supported by a development kit that provides a complete reference design with schematics, bill of materials and layout for the developer. Samples and complete development tools are available now. ■

Author information

Peder Martin Evjen is director of Technical Support for Chipcon AS.

For more information, contact:

ChipCon AS
Tel: +47-22958544
Fax: +47-22958546
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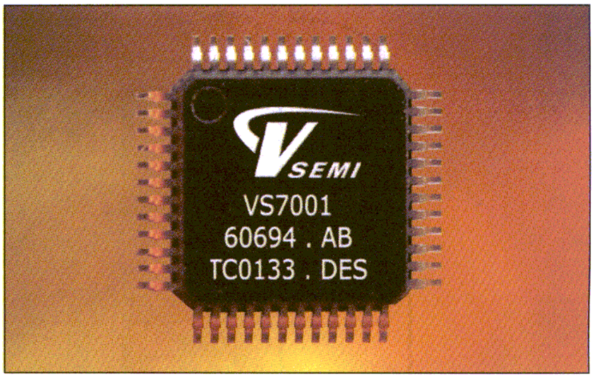
GPS Receiver Based On Pure CMOS Technology

Valence Semiconductor has introduced the VS7001 integrated circuit for GPS applications. The VS7001 is a fully integrated pure CMOS receiver, providing a clear path to a single-chip GPS system.

The VS7001 incorporates a fully integrated LNA front end, IF section, digital sampler and local oscillator synthesizer. It is fabricated in 0.35-micron pure CMOS that provides design advantages including low cost, robust performance, integration with industry-leading baseband technologies and low power consumption.

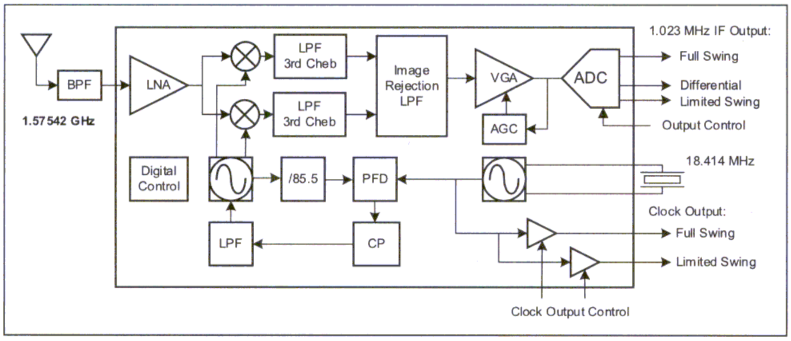
With typical total power consumption of 27 mW at 2.2V, the VS7001 receiver may be placed in battery-operated devices. Operational supply voltage range is 2.2 to 3.6 V. The receiver accommodates available baseband technologies including those offered by Sony and Mitsumi. It also eliminates the need for surface acoustical wave (SAW) IF filters, or multiple passive components, which other GPS systems use to attenuate image signal and noise.

Specifications for the VS7001 include input frequency of 1575.42 MHz (GPS L1), external LNA gain of 25 dB, noise figure of 8 dB and power dissipation of 0.027 W at 2.2 V. Clock frequency is 18.414 MHz, with crystal or TCXO clock input. Sampled IF output is 1.023 MHz. LO phase noise is –50 dBc at 10K and –80 dBc at 100K. Operating temperature range is –40 to +85 degrees Celsius.



▲ **Valence Semiconductor has introduced the VS7001 integrated circuit for GPS.**

The VS7001 is provided in a 48-pin TQFP package. Product samples are available now to prospective OEM partners. Full reference designs (supporting multiple basebands) and evaluation boards are also available. ■



▲ **Figure 1. GPS front end block diagram.**

For more information, contact:

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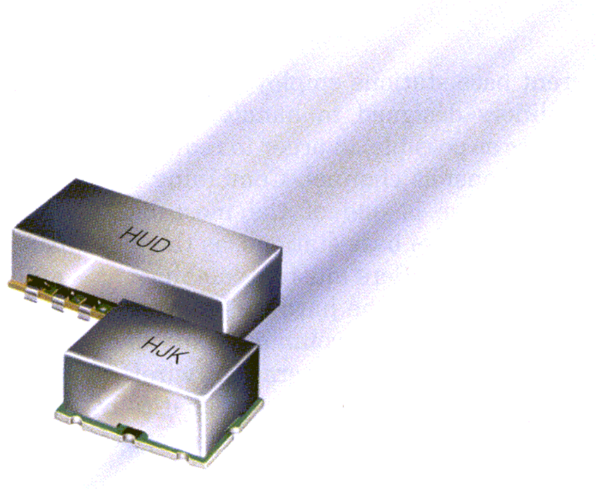

Best performance, best innovations, best value. Someone will always be on top. For example, while other companies offer *some* high IP3 mixers, at Mini-Circuits we've introduced an entire family of mixers incorporating unique patented technology to deliver the highest IP3 for low, medium, and high LO power levels. *IP3 as high as +38dBm!* And unlike other mixers, our HUD and HJK models require no DC power, which simplifies circuit layout and makes them very easy to use in 115 to 2300MHz applications. But while other suppliers are concentrating mainly on IP3 performance, Mini-Circuits goes beyond by providing very low conversion loss down to 6.7dB, high isolation up to 49dB, and 1dB RF compression point 3dB higher than LO power...all from a smaller, industry standard package fitting the same footprint as our popular SYM models. Nobody else has these high performance, high IP3 mixers. But even if they did, nobody but Mini-Circuits could offer them off-the-shelf at an amazing price from only $10.95 each!

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Typical Specifications:

	Model*	Frequency (MHz) RF	IF	LO level (dBm)	IP3 (dBm)	▲E Factor	Conv. Loss (dB)	Isolation (dB) L-R	L-I	Price $ea. (1-9)
	HJK-9	818-853	40-100	7	22	1.5	7.1	36	26	10.95
	HJK-19	1850-1910	70-130	7	21	1.4	8.0	30	24	10.95
	HJK-21	1850-1910	180-300	7	22	1.5	7.5	28	19	10.95
	HJK-9LH	818-853	40-100	10	27	1.7	6.7	37	27	12.95
	HJK-19LH	1850-1910	70-130	10	25	1.5	7.5	30	23	12.95
	HJK-21LH	1850-1910	180-300	10	25	1.5	7.2	28	19	12.95
	HJK-9MH	818-853	40-100	13	31	1.8	6.7	37	27	14.95
	HJK-19MH	1850-1910	70-130	13	30	1.7	7.4	30	23	14.95
	HJK-21MH	1850-1910	180-300	13	29	1.6	7.2	29	19	14.95
**	HJK-3H	140-180	0.5-20	16	37	2.1	8.0	44	44	16.95
	HJK-9H	818-853	40-100	17	33	1.6	6.7	35	31	16.95
	HJK-19H	1850-1910	70-130	17	34	1.7	7.7	28	22	16.95
	HJK-21H	1850-1910	180-300	17	36	1.9	7.6	28	25	16.95
**	HUD-3H	140-180	0.5-20	16	37	2.1	8.1	47	45	15.95
**	HUD-19SH	1819-1910	50-200	19	38	1.9	7.5	38	36	19.95

▲E Factor = [IP3 (dBm) − LO Power (dBm)] ÷ 10.
*Units protected under U.S. patents 5,416,043 and 5,600,169.
**Additional patents pending.
Size (L x W x H): HJK 0.500" x 0.375" x 0.23", HUD 0.803" x 0.470" x 0.250".

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Inline Couplers Split Signals to Multicarrier Power Amplifiers

Current base stations employ multicarrier amplifiers designed to handle as many carriers as possible, in order to reduce the cost of each transmit channel. Combining and dividing RF power is one of the most basic requirements in these base stations, as the outputs of multiple amplifiers must be combined at the antenna interface. The hardware employed to accomplish these power combining and division functions in second-generation wireless systems and beyond has decreased dramatically in size because of the need for lower effective radiated power in each sector.

The ILC-SH series of inline couplers from Merrimac Industries have reduced this size even further, to a fraction of the space required by competing technologies, while still handling up to 100 W CW in the operating range of 1.8 to 2 GHz (Figure 1).

The ILC-SH inline couplers are fabricated in the company s Multi-Mix process, the next generation of which, called Multi-Mix PICO, was announced this year. Multi-Mix is a multilayer technology in which fluoropolymer composite substrates are fusion-bonded together to form a multilayer structure of microwave circuits that are a fraction of the size of similar components realized with conventional fabrication technologies. Nearly any type of active or passive microwave element, from discrete semiconductors to MMICs, and plated-through via holes, can be contained within this structure to form a multifunctional circuit that is self-contained and requires no additional packaging.

Multi-Mix PICO advances Multi-Mix miniaturization by reducing the size of devices that can be fabricated by as much as 90 percent. The first Multi-Mix PICO devices introduced by Merrimac, quadrature hybrids and directional couplers, are the some of the smallest achievable with any commercially-available technology. A family of three-way power dividers has since been added to Multi-Mix PICO line, and they are typically 90 percent smaller than their nearest competitor.

ILC-SH inline couplers are also extremely small in comparison to current products that deliver the same functionality. Within a 0.2 × 0.56 × 13 inch package that weighs 0.023 ounces, the ILC-SH devices contain a complete inline coupler solution that handles RF input power of 10 W CW, and optionally 100 W.

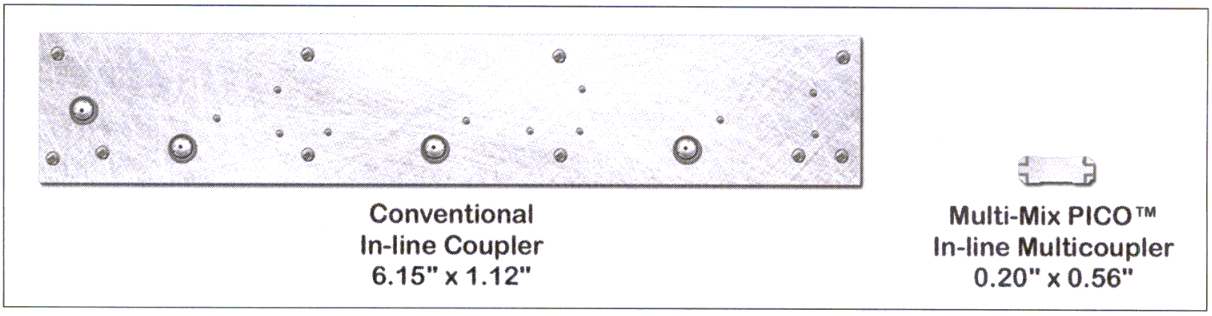

Figure 1. The Multi-Mix PICO ILC-SH inline coupler delivers dramatic size reduction compared to similar devices realized with conventional technology.

Operating frequency	1.8 to 2.0 GHz
Nominal coupling	4.8 dB
Insertion loss	0.25 dB typical
Amplitude balance	±0.6 dB typical
Isolation	at least 22 dB
VSWR	1.35:1 typical
Maximum input power1	10 W CW Optionally 100 W CW
Operating temperature range	–55 to +85 degrees C
Size	0.20 0.56 0.13 inches
Weight	0.023 ounces

** CW input power tested as a divider, internal load disspation 375 mW or less for 10-watt model and 1.5 W or less for 100-watt model*

▲ Table 1. ILC-3SH-2.0G inline coupler specifications.

Insertion loss is typically 0.25 dB, and isolation is at least 22 dB.

The ILC-SH series is a serial coupler network that has several advantages over corporate networks that employ Wilkinson power dividers in a parallel configuration. It allows any number of splits to be achieved, and produces a device with a wide bandwidth and lower insertion loss. The ILC-SH devices consist of serially connected couplers, each one with a different coupling value. Equal split loss is maintained at all coupled ports because the direct-path loss of previous couplers in the line is added. Each of the paths through a divider/combiner pair have equal insertion phase, but the individual path ports do not. The technique results in extremely accurate phase and amplitude balance, in the vector sum, which makes the ILC-SH inline couplers well suited for use in the multicarrier amplifier applications.

Three-way ILC-SH series devices are available on tape and reel from stock. Custom models in four-way, six-way and other configurations are available as well, with exact specifications tailored to customer requirements. ■

For more information, contact:

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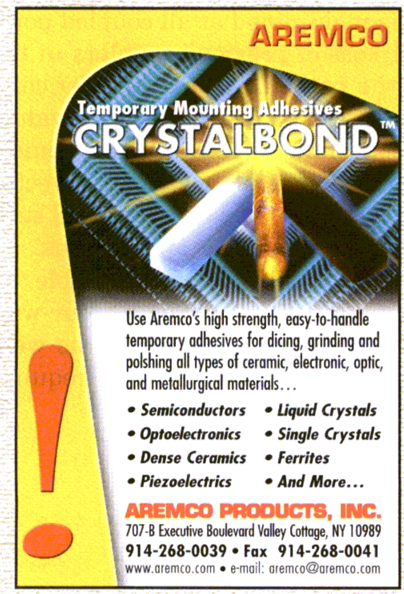

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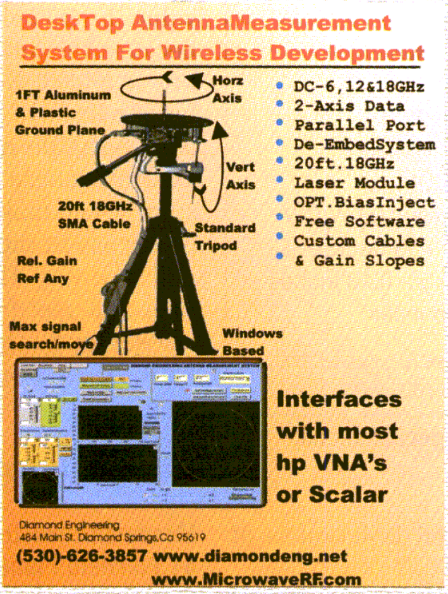

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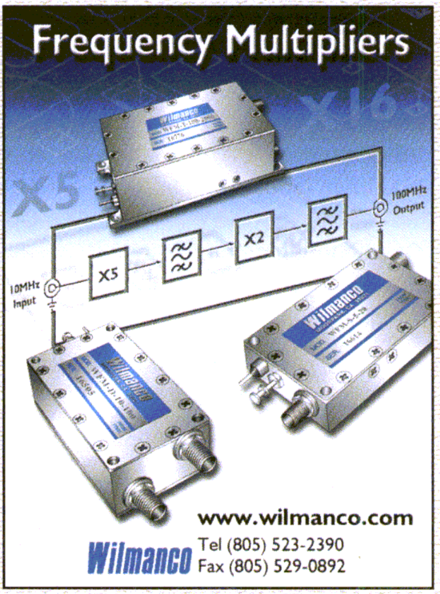

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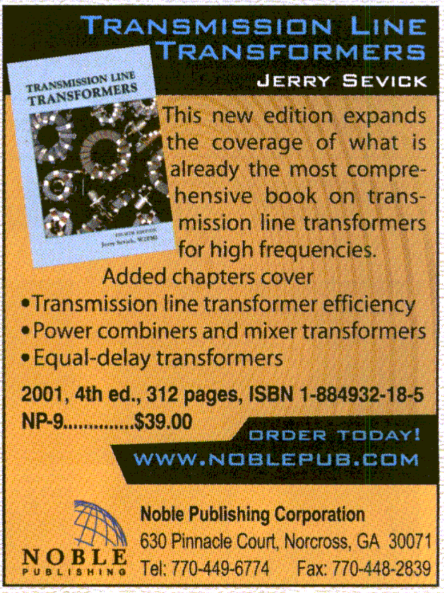

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technical direction and then by clearly articulating the corporate strategies and objectives that result from this process. In return, we hold all employees, and particularly our engineering team, accountable for delivering leading-edge solutions for our customers' needs.

Job satisfaction and career development opportunities are two additional areas of focus. Again, we have enhanced our personnel policies to be better aligned with the needs and expectations of the people to which they apply. Our engineers benefit from practical quality-of-life enhancements we have adopted, such as 25 three-day weekends per year, flexible work schedules to avoid stressful rush-hour commutes and a relaxed, casual environment conducive to improved productivity. We have begun tuition reimbursement for continuing education, improved orientation and training processes and career advancement options covering both management and technical opportunities. Our small-company environment also offers opportunities for creativity and innovation and allows employees to become involved in many aspects of the business, not just a narrow technical specialty. This provides the chance to form teams that strongly influence the success of the business.

This is not to say that compensation isn't important. Competitive salaries will always play a major role in attracting strong talent. In addition, despite the stock market decline of 1999-2001, stock options will remain an important component of compensation packages.

Over the next several months, Vari-L hopes to double its technical staff, adding engineers who are drawn to a well-rounded opportunity that offers strong potential for a successful career in conjunction with high quality of life. These objectives are not mutually exclusive; they are being achieved now. ■

Chuck Bland is president and CEO of Vari-L Company (www.vari-l.com), a developer of RF and microwave components for commercial wireless and military/aerospace customers. He has more than 25 years of international management, finance and operations experience, including senior executive positions with Owens Corning, where his assignments ranged from president of Africa/Latin America operations to president of Asia Pacific operations. Bland earned a B.S. in accounting and finance from Ohio State University and an MBA from the Sloan School at Massachusetts Institute of Technology.

Next month in Applied MICROWAVE & WIRELESS

On our cover

- **UltraRF** introduces the first in a range of high power amplifier modules designed to meet the linearity requirements of 3G wireless infrastructure systems..

New Wireless Technologies

- *"System-in-a-Package Solution for Short-Range Wireless Communications"*
  Christer Svensson and Shaofang Gong of **Bluetronics** describe a new module offering short design and prototype lead time, high performance and ease of upgrade.

Filters

- *"Transforms Aid the Design of Practical Filters"*
  Randall W. Rhea introduces several transforms useful for the design of filters using direct synthesis or lowpass prototype tables.

Frequency Synthesizers

- Product Focus featuring new frequency synthesizer products.

Also featured:

- *"Comparing Differential Measurement Techniques"*
  In the second part of this two-part article, Loren Betts of **Agilent Technologies** offers a rigorous analysis of the error effects of each measurement method, including such concerns as improper third-port termination and fixturing.

The Challenge of Finding Good Engineers

By Chuck Bland
Vari-L Company

In the 1990s, a robust economy and booming demand for wireless products and services were good news for RF engineers but challenging and often frustrating for companies trying to attract and retain those engineers. This increased demand occurred after a period of declining demand for RF and microwave engineers. The primary demand had previously come from military applications, and the shrinking defense budget of the late 1980s and early 1990s discouraged engineers from entering this discipline. In addition, RF and microwave engineering was not a commonly taught discipline in universities. This created a low supply.

The wireless boom altered this balance significantly. Engineers who once competed for a finite number of jobs were suddenly being pursued by employers competing to fill a seemingly endless need for engineers. In the second half of the decade, an unprecedented flow of investment capital into the wireless and technology sectors created more jobs and tightened the market even further. Established companies and a growing number of venture- and IPO-funded start-ups beefed up compensation packages with aggressive salaries and stock options, and for many companies the pursuit and retention of engineers was the determining factor in growth rates and product introduction. It was not uncommon with this seemingly relentless demand for an engineer to seek new opportunities with another company after a year or two of work.

Over the past couple of years, of course, the economy has slowed considerably, the wireless industry itself has come under pressure and many bubbles have burst. Scores of start-ups have shut down and investment capital has all but dried up, leaving many other companies on unsteady financial ground. As a result, while overall demand for talented engineers remains relatively strong, many engineers are re-examining career choices, objectives and priorities. With stock options underwater and layoffs looming, stability and quality of career and quality of life are re-emerging as key considerations in job selection. The horrific events of September 11 have only sharpened this focus as all of us examine what is important in our lives.

The pause in the economic and wireless boom cycle has also given employers an opportunity to re-examine their strategies for hiring and retaining engineers. Turnover can limit a company's performance in both good and bad times, adding cost, limiting new product innovation and disrupting cohesive teams that are critical for success. It's estimated that the cost of replacing an RF engineer is 160 percent of the cost of retaining one. That means a company will spend an additional $42,000 to identify, pursue, hire and train a new engineer to replace a departing one who earned $70,000. Even for small companies, such as one with 30 engineers and a 20 percent turnover rate, that represents nearly a quarter of a million lost dollars a year — not to mention the intangible costs associated with disruption of workforce and workflow.

At Vari-L, we have made several changes in our approach to attracting and retaining high-quality employees, including a growing number of RF engineers. These began with the arrival of a new executive management team in 2001. The basic tenet of our approach is enhancing the overall quality of life — work life and personal life — for all employees while remaining highly competitive and progressive in traditional key areas such as compensation, benefits, perks, advancement and creative freedom. This approach is of particular importance for us because we are aggressively expanding engineering staff at a time when many other companies in the wireless industry are contracting. We want to be sure new employees are coming for the right reasons, will become valuable contributors and will thus be more likely to remain with the company when the inevitable expansion cycle resumes.

The sports analogy is the prized free agent who cites the quality of his organization — the way the organization attends to the details and the ability of the team to win — as the determining factor in signing with an organization. We believe companies offering a good balance between quality of life, job fulfillment and compensation will have a decided advantage in attracting and retaining the best engineering talent.

In the case of Vari-L, our location in Colorado is an advantage because the state provides a variety of year-round, family-oriented outdoor activities commonly associated with the quality-of-life issue. Denver also has become a high-tech area with many telecommunications companies now located in the surrounding area, including Qwest, Level 3, Echostar, Spectralink and others.

Vari-L is also addressing less obvious issues that are often overlooked. For example, a lack of confidence in senior and mid-level management is consistently cited as one of the leading causes of employee turnover in the tech sector. Accordingly, we have increased the quantity and quality of communications between engineers, supervisors and executives. The small size of our company allows us to give engineers a voice in shaping our

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